ENGINEERING DESIGN HANDBOOK

ELECTROMAGNETIC COMPATIBILITY

HQ, USA MATERIEL DEVELOPMENT AND READINESS COMMAND MARCH 1977

Best Available Copy

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ENGINEERING DESIGN HANDBOOK
ELECTROMAGNETIC COMPATIBILITY

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ABBREVIATIONS FOR NONTECHNICAL TERMS

AAWS	air-to-air weapon system
AGE	aerospace ground equipment
AR	Army Regulation
ASARC	Army Systems Acquisition Review Council
ASWS	air-to-surface weapon system
BOI	basis of issue
BOIP	basis-of-issue plan
C-E	communication-electronic
CFP	concept formation package
CM/CB	counter mortar-counterbattery system
COMM	communications system
CONUS	continental U.S.
CSTA	combat surveillance and target acquisition system
CTP	coordinated test program
CRDA	Chief, Research, Development, and Acquisition (DA)
C/S	Chief of Staff
DA	Department of the Army
DCP	development concept paper
DCSOPS	Deputy Chief of Staff, Operations and Plans (DA)
DoD	Department of Defense
DoDD	Department of Defense Directive
DP	development plan
DPM	defense program memorandum
DSARC	Defense Systems Acquisition Review Council
DT	development test
ECAC	Electromagnetic Compatibility Analysis Center (DoD)
ECM	electronic countermeasures
EM	electromagnetic
EMC	electromagnetic compatibility
EMCAB	EMC Advisory Board
EMCP	electromagnetic compatibility program
EMI	electromagnetic interference
EWS	electronic warfare system
FAS	Frequency Assignment Subcommittee (IRAC)
FCC	Federal Communications Commission
FM	Field Manual
FS	frequency supportability
ING	International Notification Group (IRAC)
IOC	initial operational capability
IP	initial production
IPR	in-process review
IRAC	Interdepartment Radio Advisory Committee
ISM	Industrial, Scientific, and Medical
JCS	Joint Chiefs of Staff
J/FP	Joint Frequency Panel (MCEB)
LCMM	*see*: LCSMM
LCSMM	Life Cycle System Management Model (formerly LCMM)
MCEB	Military-Communications Electronics Board
MIJI	meaconing, intrusion, jamming, and interference
MIL-STD	Military Standard
OSD	Office of the Secretary of Defense
OT	operational test; Office of Telecommunications
OTP	Office of Telecommunications Policy
Pam	pamphlet
PM	program memorandum; project/product manager
QQPRI	qualitative and quantitative personnel requirements information
R&D	research and development
RFI	radio frequency interference
SAWS	surface-to-air weapon system
SIGSEC	signal security
SPS	Spectrum Planning Subcommittee (IRAC)
SSWS	surface-to-surface weapon system
TDS	tactical data system
T&E	test and evaluation
TOE	table of organization and equipment
TSC	Technical Subcommittee (IRAC)
USACC	US Army Communications Command (formerly USASTRATCOM)
VT	variable time (fuze)

ABBREVIATIONS FOR UNITS AND TECHNICAL TERMS

A	ampere
ac	alternating current
AGC	automatic gain control
AM	amplitude modulation
C	coulomb
CW	continuous wave
cm	centimeter
dB	decibel
dBm	dB with reference to 1 milliwatt
dBW	dB with reference to 1 watt
dB[ ]	decibel with reference to the quantity in brackets
dc	direct current
deg	degree
EMF	electromotive force
EHF	extremely high frequency
ELF	extremely low frequency
ft	foot
F	farad
FM	frequency modulation
GHz	gigahertz
H	henry
HF	high frequency
Hz	hertz
IF	intermediate frequency
J	joule
K	Kelvin
kHz	kilohertz
km	kilometer
kV	kilovolt
kW	kilowatt
log	logarithm
ln	logarithm, natural
LF	low frequency
MF	medium frequency
MHz	megahertz
MV	megavolt
m	meter
mA	milliampere
mH	millihenry
ms	millisecond
mV	millivolt
mW	milliwatt
Ω	ohm
pF	picofarad
PRF	pulse repetition frequency
PRR	pulse repetition rate
rad	radian
RF	radio frequency
rms	root mean square
s	second
sr	steradian
T	tesla
UHF	ultra-high frequency
V	volt
VHF	very high frequency
VLF	very low frequency
VSWR	voltage standing wave ratio
VTVM	vacuum-tube voltmeter
W	watt
Wb	Weber

PREFACE

The Engineering Design Handbook Series of the US Army Materiel Development and Readiness Command is a coordinated series containing basic information and fundamental data useful in the design and development of Army materiel and systems. The handbooks are authoritative reference books of practical information and quantitative facts helpful in the design and development of Army materiel. The purpose of this particular Handbook is to take the wealth of information accumulated by the Army over a period of years on the subject of electromagnetic compatibility and make this information available to the designer.

In support of the design information, the Handbook contains discussion of electromagnetic field coupling mechanisms, nonlinear circuit effects, and statistical concepts necessary to an understanding of the phenomena involved. To some extent the Handbook can be considered to be an update of the *Interference Reduction Guide* prepared by Filtron Co., Inc., in 1964 for the US Army Electronics Laboratories. Much of that material which is of current value has been retained in this document although rearranged in many places and extensively edited.

The Handbook on Electromagnetic Compatibility was prepared at the University of Pennsylvania, Philadelphia, PA, for the Engineering Handbook Office, Research Triangle Institute, Research Triangle Park, NC. Contributing authors included: R. M. Showers, F. Haber, R. S. Berkowitz, J. B. Butler, and L. Forrest, Jr. Sachs-Freeman Associates, Hyattsville, MD, assisted with Chapter 6. Technical guidance was provided by a committee with representatives from the US Army Electronics Command, Ft. Monmouth, NJ; US Army Aviation Systems Command, St. Louis, MO; Picatinny Arsenal, Dover, NJ; US Army Tank-Automotive Command, Warren, MI; and US Army Missile Command, Redstone Arsenal, AL. Members of this committee were: Mr. J.J. O'Neil, Chairman; Mr. John Snyder; Mr. A. Grinoch; Mr. Benjamin Ciocan; Mr. Leon Riley; and Mr. Basil L. DeNardi.

The Engineering Design Handbooks fall into two basic categories, those approved for release and sale, and those classified for security reasons. The US Army Materiel Development and Readiness Command policy is to release these Engineering Design Handbooks in accordance with current DOD Directive 7230.7, dated 18 September 1973. All unclassified Handbooks can be obtained from the National Technical Information Service (NTIS). Procedures for acquiring these Handbooks follow:

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CHAPTER 1
INTRODUCTION

1-1 BACKGROUND INFORMATION

The development of devices utilizing electrical or electromagnetic phenomena in their operation has been so extensive that few components or systems used for military or civilian purposes exist that do not depend upon them. However, the very nature of the properties associated with the phenomena which enable them to perform their purposes may also produce undesired effects on other equipment having different purposes. Thus, radio signal F_1 (see Fig. 1-1) which serves to provide communications to receiver R_1 is quite capable of degrading the response of receiver R_2 to signal F_2. In a similar way, electrical circuits in industrial applications W may degrade nearby receivers either directly by radiation E_1, E_2, by conduction along a common power line I, or by combined conduction and then radiation E_3. Magnetic inductive coupling between a signal line and a power line is represented by H. Other examples are:

a. Effects produced by an electrical system of a vehicle on nearby communications and in the same vehicle

b. Mutual interaction of two geographically close radars

c. Undesired interaction of receivers and transmitters due to spurious outputs or responses

d. Failure of computer or control equipment because of power line harmonics or switching transients.

These undesired interactions can take place at all levels; i.e., between systems, between subsystems of the same system, between equipments in the same subsystem and between components in the same "black box".

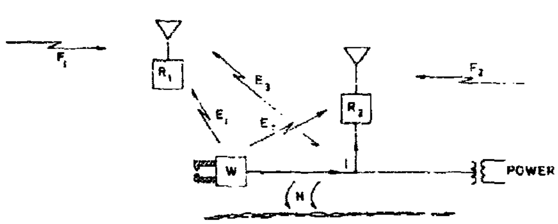

Figure 1-1. Forms of Interference Coupling

In some cases the causes of system degradation can be recognized and distinguished readily as in the case of ignition or continuous wave interference with amplitude modulated radio or television transmissions. In other cases, such as in control circuits, it may be much more difficult to identify the cause of a system failure.

Electromagnetic Compatibility (EMC) is the ability of C-E equipments, subsystems, and systems to operate in their intended operational environments without suffering or causing unacceptable degradation because of unintentional electromagnetic radiation or response.

Electromagnetic Interference (EMI) is the phenomenon resulting when electromagnetic energy causes unacceptable or undesirable responses, malfunction, degradation, or interruption of the intended operation of electronic equipment, subsystem or system. The term radio noise is reserved for emanations other than signals.

The rapid advance of technology has increased not only the number of electrical equipments, but also their complexity. Thus, the effectiveness of performing almost any single ultimate function is now usually dependent on the efficient performance of many other functions. Increasing density of equipments to perform these functions greatly increases the possibility of degradation in performance by undesired electromagnetic interactions.

Examples of recent technological developments which have increased the possibility of interactions include:

a. Miniature and integrated circuits which result in high component packing densities

b. Increased component susceptibility because of higher sensitivity

c. Use of wider bandwidths in equipment design.

Possible consequences of performance degradations caused by electromagnetic interference include:

a. Mission abort

b. Message inaccuracies

c. Message repeats or delays

d. Navigational errors

e. Failure of logistic support

f. Reduction of system availability

g. Movement surveillance under combat conditions.

1-2 THE SIGNAL CONCEPT OF EMC

One can consider that all instances of electomagnetic incompatibility result in the degradation of the ability of a system to transmit information from one intended point to another as shown on Fig. 1-2. An information source generates a message which is then encoded in some convenient form for transmission through a suitable medium. After transmission, the signal is received at some other point where the message is decoded and forwarded to the end user which may be either a human being or some device. This representation is characteristic of the usual communication system in which the signal channel is either a path through which a radio wave propagates — such as the troposphere or ionosphere surrounding the earth, or a pair of wires such as in telephone transmission, or a coaxial cable as in closed circuit television — and normally is the place at which interference is coupled into the system. Also shown is how interference can enter into the system at other places — such as at the information source, the encoder, or decoder — depending on the nature of the interference source and its location. Fig. 1-2 also can represent a control system except that a feedback loop (not shown) from destination to information source is required.

1-3 ELECTROMAGNETIC INTERFERENCE CONTROL

In order to control electromagnetic interference, the various possible electromagnetic interactions must be clearly understood. As previously illustrated, the "coupling" may be by way of many different paths and include such phenomena as conduction, induction, and radiation. These phenomena themselves are relatively easy to understand when the configurations are simple and all the parameters are known. However, configurations are seldom simple and, as a result, the interactions are usually quite complex. Furthermore, analysis often must be carried out under conditions in which there is considerable uncertainty as to the values of the various parameters involved.

Techniques for dealing with these interactions have been studied explicitly for more than 35 yr. Data have been compiled on emmission characteristics of various equipments capable of producing interference voltages, on the susceptibility of sensitive equipments, and on the coupling factors between them. Methods of controlling these interactions by (a) suppression of interference generating mechanisms at the source, (b) reduction of coupling by electrical filtering, shielding, circuit separation, and other wiring techniques, and (c) reduction of susceptibility by circuit linearization, filtering, and circuit design techniques have been documented. Methods of efficiently examining potentially interfering situations by modeling and computer analysis also have been developed. In recent years it has been clear that a need exists for bringing most of this information together and organizing it to make it readily accessible to the design engineer and others concerned with electromagnetic compatibility problems.

1-4 PURPOSE OF HANDBOOK

The purpose of this handbook is to assemble information pertinent to the establishment of a condition of electromagnetic compatibility for any equipment or system, with itself, and with the environment in which it is to be placed when performing its designated function, and to define procedures to assure that an optimum design is obtained. The major

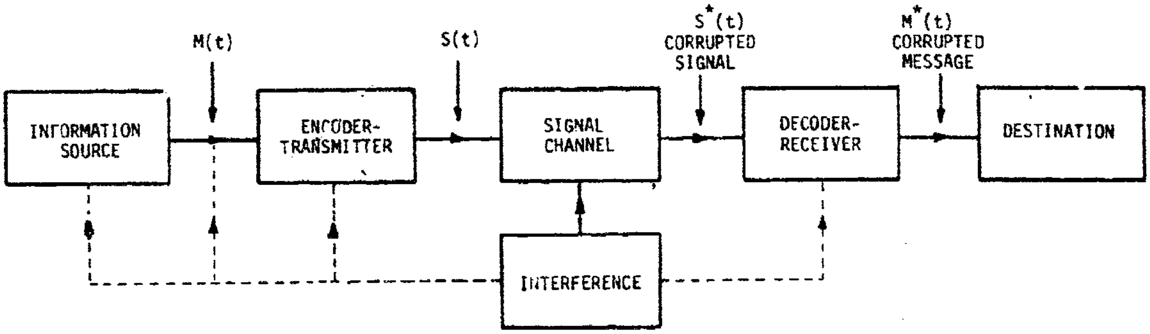

Figure 1-2. Communication System and Its Environment

emphasis is on those aspects of the problem of interest to the equipment design engineer and the system engineer. Special attention is paid to the early stages of development when EMI problems can be solved in an efficient way.

The handbook also will be of use to engineers concerned with assuring the compatibility of equipment with its environment after installation.

1-5 SCOPE OF HANDBOOK

Material included in this handbook is of general and specific interest to the US Army. It pertains to Army equipments used in the active battlefield areas as well as in support areas in an otherwise civilian environment. The treatment includes a review of fundamental electrical science and engineering phenomena, an understanding of which is required for application of design processes for achieving electromagnetic compatibility. The interactions involved are those which can result in either temporary or permanent degradation of performance of such equipment. Hazards resulting from exposure of biological materials and of munitions to electromagnetic radiation are treated elsewhere (Ref. 1) as are effects of the electromagnetic pulse (EMP) (Refs. 2, 3, and 4).

The major effort is directed toward the equipment and subsystem level, i.e., the design and installation of equipment or subsystem to perform specific functions within systems and which must be compatible with themselves as well as compatible with nearby located equipment, subsystems, or systems.

Interactions will be considered that occur between equipments and subsystems which are located no further apart than several miles. The frequency range covered is from 30 Hz to 30 GHz, although in some cases the effects of constant magnetic fields will be considered.

1-6 ORGANIZATION AND USE

1-6.1 ORGANIZATION

The material in this handbook is organized in accordance with the outline that follows. For details, consult the table of contents.

Chapter 1-Introduction

Chapter 2-EMC/EMI Requirements and Procedures. General background information is supplied on the philosophy of EMC control procedures as they have been developed in the past, with emphasis upon the particular characteristics of specific environments. Methods of implementing various types of EMC control programs are discussed. Specific topics include:

a. Signal environment classification
b. Approaches used by various organizations
c. Specifications
d. EMC program organization.

Chapter 3-EMC Phenomena and Analysis. A detailed discussion of the various elements that enter into the analysis of electromagnetic compatibility interactions is presented. This chapter gives information on the basic phenomena involved in such form that quantitative analysis can be carried out with the highest possible accuracy. Included are:

a. Source models:
   (1) Conducted and radiated
   (2) Narrowband and broadband
b. Source characteristics:
   (1) Natural
   (2) Man-made
c. Susceptibility:
   (1) Admission mechanisms
   (2) Masking and error induction
d. Coupling phenomena:
   (1) Induction field
   (2) Radiation field
   (3) Conduction
   (4) Grounding.

Chapter 4-EMC/EMI Design Techniques. Practical aspects of EMC/EMI control are treated in detail with quantitative data on the various methods of reducing interactions. The chapter makes reference to basic models and data given in Chapter 3. Topics considered are:

a. Emission control
b. Susceptibility control
c. Coupling control
d. Filters
e. Wiring and cabling
f. Shielding
g. Grounding and bonding.

Chapter 5-Applications to Specific Devices. Electromagnetic interference and control characteristics of specific equipments in electrical and electronic categories are discussed. In the power category emission is emphasized, while electronic equipments are discussed both from an emission and susceptibility standpoint. Cross-references are made to earlier portions of the handbook. Paragraph headings include:

a. Rotating electrical machines
b. Power distribution
c. Power control
d. Electronic power supplies
e. Vehicles and other engine driven equipment
f. Receivers
g. Transmitters
h. Radar equipment
i. Antenna interaction control

j. Infrared equipment
k. Aircraft
l. Aerospace ground equipment
m. Special circuit considerations
n. Telemetering.

Chapter 6-Systematic Prediction and Analysis. This chapter presents a discussion of analysis and EMC prediction techniques with particular reference to large and complex configurations. The evaluation of the wide variety of possible interactions, and the effects on circuits of multiple parameter variations, generally requires the use of a large scale digital computer. The chapter describes how methods used can be classified and presents a description of the major models. The information is presented in a fashion to enable the user to select the appropriate method. Included are:

a. Intrasystem vs intersystem evaluation
b. Data files
c. Application of data files and models
d. Description of programs
e. Model utilization.

Chapter 7-Measurements. The applications of various types of measurement equipment and associated test techniques to achieve EMC are discussed. Limitations due to uncertainty factors in applications are described. Techniques which will find general usefulness in the laboratory in the early stages of equipment development are covered, along with techniques specially designed for measurements according to specifications. Included are:

a. Test requirements
b. Instrumentation
c. Test facilities
d. Measurement techniques
e. EMC test facilities.

1-6.2 USE OF HANDBOOK

This handbook is organized for use by engineers representing a wide variety of technical experience. Generally, it assumes the reader has, as a minimum, the equivalent of a Bachelor's degree in Electrical Engineering, but other persons with experience in specialized aspects of EMC will find it useful.

If one is specifically concerned with a given type of equipment, it is best to start with the appropriate discussion in Chapter 5. The engineer concerned with circuit development will find Chapter 4 the most useful, while one concerned with early development stages of a given equipment or system should make extensive reference to Chapter 2.

The index at the end of the handbook provides immediate identification of all elements in the handbook which pertain to particular compatibility problems.

Other general and specific handbooks and manuals are available which the reader may find useful in certain applications. These are Refs. 5 to 12.

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CHAPTER 2
EMC/EMI REQUIREMENTS & PROCEDURES

2-1 INTRODUCTION

This chaper is concerned with the general concepts underlying the procedures for controlling electromagnetic interference and achieving electromagnetic compatibility in Department of the Army and related programs, and provides detailed information on procedures to be used. The importance of the system viewpoint is emphasized, but the viewpoint of the subsystem/equipment designer is treated also.

2-2 THE ELECTROMAGNETIC ENVIRONMENT

Contributors to the electromagnetic environment can be divided into three main classes: (1) natural radio noise, (2) signals which are generated purposely to convey information, and (3) spectral components generated incidentally to the functioning of various electrical and electronic devices, and generally classed as man-made noise.

2-2.1 THE NATURAL RADIO-NOISE ENVIRONMENT

Natural radio noise originates in atmospheric disturbances and varies with location on the surface of the earth, time of day, and time of year. Usually the level varies only slowly with position in a given geographical area, and at frequencies in the VHF range and below it provides a background level which sets the maximum useful sensitivity of a receiver. Methods of estimating the expected level are discussed in par. 3-1.2.2.

2-2.2 THE MAN-MADE NOISE ENVIRONMENT

2-2.2.1 Noise Levels

Individual contributors to man-made (incidental) noise are discussed in detail in Chapter 3. Since most sources are significant only in their immediate vicinity, this environmental level can be expected to vary rapidly with distance. However, studies have shown that the expected level can be related to the population density in the surrounding area. Fig. 2-1 shows levels published by the Department of Commerce (Ref. 1) which show a difference between urban and rural areas of about 18 dB. The quantity F_a, called the noise factor, is defined by Eq. 3-19.

2-2.2.2 Intrasystem and Intersystem Compatibility

A special type of man-made noise environment is that created by various parts of a single electronic system. As discussed in Chapter 1, in most electrical and electronic systems, one part of a system may create a "hostile" environment for another part. This concept of intrasystem compatibility can be distinguished clearly from intersystem compatibility involving interactions between antennas of systems widely separated geographically.

In many cases the distinction is not so clear cut, especially where different systems are located close together physically, and indeed where such systems — such as radar, navigation, and communications — can be considered to be parts of a larger system, e.g., an aircraft which carries these systems. The relationships are illustrated on Fig. 2-2 in which intersystem interactions are illustrated with solid lines, while intrasystem interactions are shown dotted. At the higher "system" levels undesired coupling usually involves a highly sensitive antenna or transmission line and the coupling is by radiation. At the lower intrasystem levels the coupling is between cabinets or cables, or indeed between circuit elements by conduction or induction.

At the lowest circuit levels, relatively simple functions are performed. Criteria for determining the effectiveness with which the interactions take place are relatively easy to establish in terms of subsystem operational requirements. As one moves up the system "hierarchy", it becomes more difficult to establish firm criteria for performance because the function performed becomes more complex and it becomes increasingly difficult to control, or even to predict, the significant environmental conditions.

At the equipment level and below, usually it is possible to isolate any device from uncontrollable factors of the environment. This renders the design problem essentially deterministic; it is that of providing the required number of noninteracting (intrasystem compatible) communication paths of adequate reliability, and doing so at a minimum cost. Trade-offs can be definitive, and results can be evaluated reasonably well. At the upper levels, one must consider the natural and signal environments in which the system operates which are less controllable, so that it is

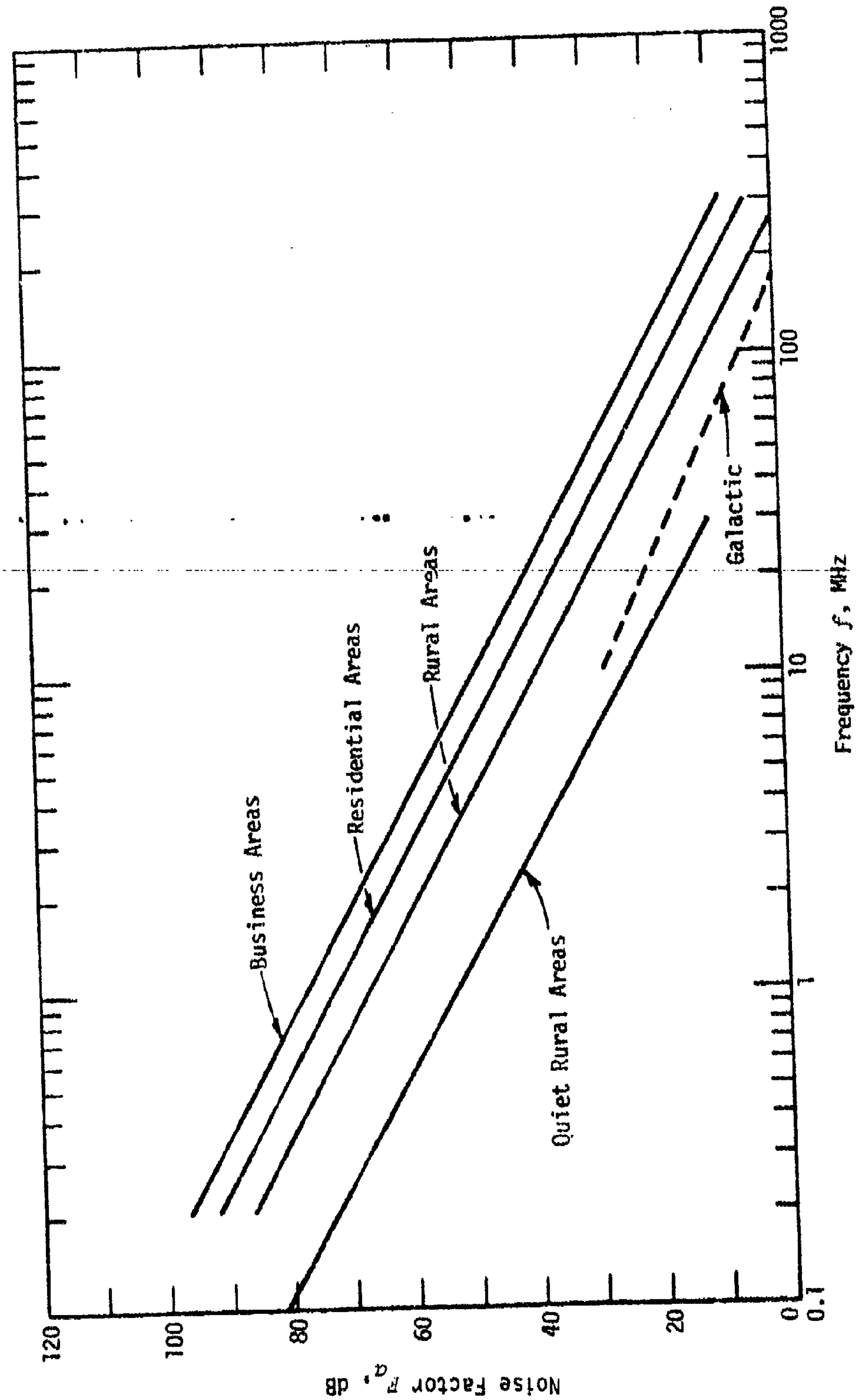


Figure 2-1. Estimates of Median Values of Man-made Noise Expected at Typical Locations

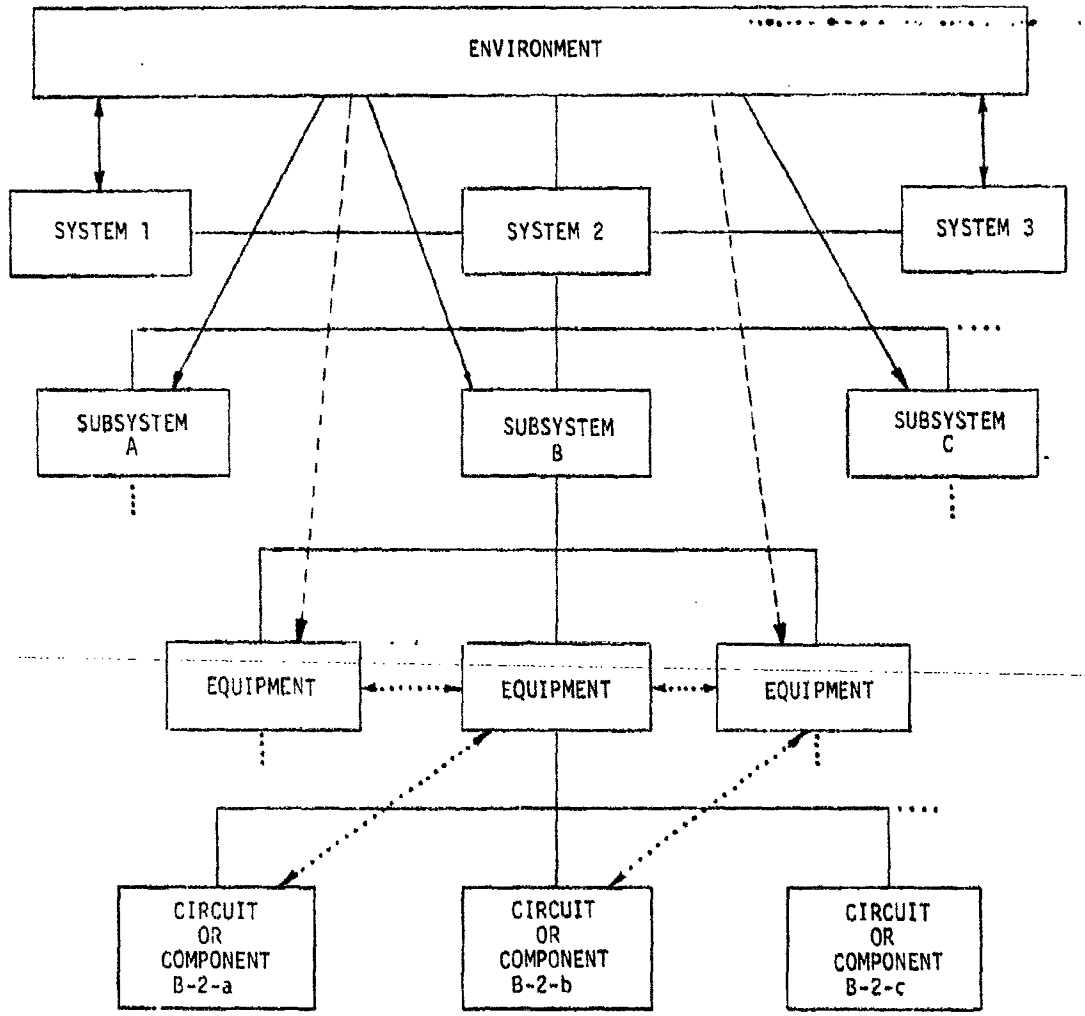


Figure 2-2. Model of System and Its Enviornment

necessary to formulate a statistical design, using statistical trade-off relationships. Thus, to do an adequate job of intersystem compatibility engineering, it is necessary to anticipate and describe the variable aspects of the interaction with the electromagnetic environment in which the system will be expected to operate, as illustrated on Fig. 2-2 with solid arrows.

2-2.3 THE SIGNAL ENVIRONMENT

Radio signals are the strongest components of the electromagnetic environment in almost any location. In contrast to natural and incidental radio noise, they occupy relatively narrow segments of the radio spectrum. For a given receiver, since only one of the signals is the desired signal at a particular time, the others are potential sources of interference, especially those of high signal strength or those located close to the desired signal in frequency.

Rough estimates of the field strength E due to a transmitter at a distance r meters can be made using the free-space transmission formula (see Eq. 3-145).

$$E = \frac{\sqrt{30 P_t G_t}}{r}, \text{ V/m} \qquad (2\text{-}1)$$

where

P_t = total radiated output power, W

G_t = transmitting antenna gain, dimensionless

Fig. 2-3 shows the variation of field strength with r for constant values of P_tG_t.

In using Eq. 2-1, the reader is cautioned that the rate of fall-off of E with distance may be greater or less depending on propagation conditions, see par. 3-3.2.

The signal environment in which military equipment and systems operate consists of two major elements, the military part and the nonmilitary part. The military part is created by the overall military system and thus can be controlled to a large extent. The nonmilitary is characteristic of the area and only limited control may be exercised over it.

2-2.3.1 Nonmilitary Signal Environments

Any environmental classification scheme must be considered to be partially conceptual, rather than absolute, since as one moves physically from one location to another, characteristics change gradually. However enough areas can be classified distinctly as one type or another to make classification useful. In this category we classify environments as rural, residential, commercial, industrial, and urban on the presumption that man-made radio noise levels will increase significantly as we pass from one to the other.

Rural areas can be considered to have a population density less than 500 persons per square mile and, except for isolated instances, are at least 25 miles from fixed transmitters having more than a kilowatt of output power. They have no substantial industrial activity. Radio-noise levels are determined by atmospheric noise levels except in the immediate vicinity of high voltage power lines, electric fences, or heavily travelled roads. Radio signals of all kinds are at relatively low levels and the ability to receive desired signals generally is limited more by low signal strengths than by interference from undesired signals.

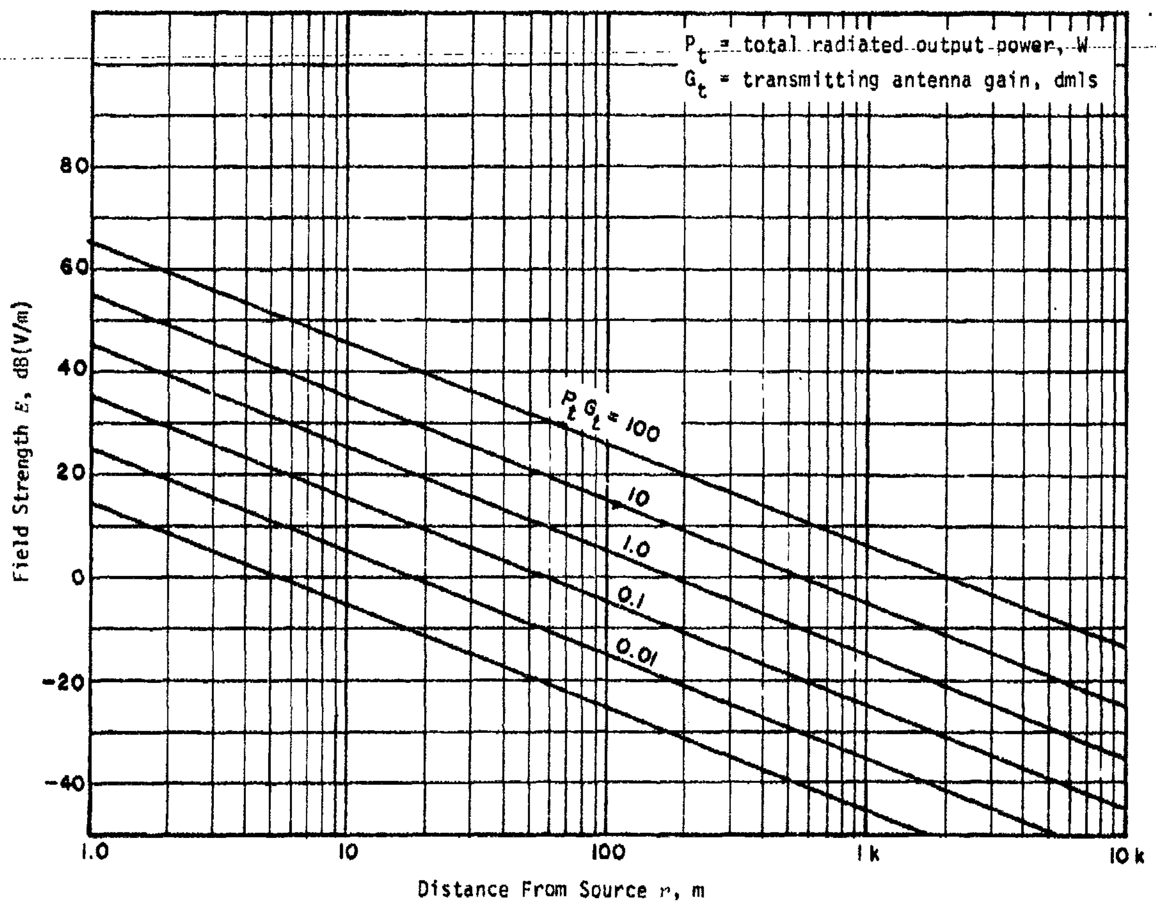


Figure 2-3. Field Strength as a Function of Distance of Transmitter

At the other extreme, urban areas are characterized by high population densities, which may range from 15,000 to more than 70,000 persons per square mile, and relatively high ambient noise levels due to numerous industrial and commercial area as well as densely populated residential areas. In the VHF and UHF bands, local ignition interference can be of considerable importance. Broadcast signals, because of their local origin, are usually relatively strong and the signal strengths in the vicinity of the antenna towers can be very large. In the VHF and UHF bands, however, the effects of tall buildings may cause serious reflections which will degrade performance locally and distort normal field strength contours.

2-2.3.2 Military Environments

This classification includes environments which are characteristic of peace-time operations as well as operations which are likely to be found only under conditions of conflict.

2-2.3.2.1 Ground Station

This classification includes a typical Army post. It usually includes an array of transmitters having a wide range of output power and corresponding receivers. In some cases, particularly at HF where high power may be used, receiving antennas are separated from transmitting antennas by distances of the order of miles in order to avoid interaction.

At VHF and UHF, line-of-sight multichannel links use low power (a few watts) and may use the same reflector for transmission and reception, and reflectors for separate links may be mounted on the same tower. Troposcatter links use high power with a highly directional antenna which may be used for both transmitting and receiving. Furthermore, high powered radars may be located on the site. Frequently such sites have industrial types of equipment in use.

2-2.3.2.2 Battlefield

In the battlefield environment, radio communications provide the principal means of exercising command and control. At the same time, radar, special weapons systems, and telemetry units are expected to function with a high degree of reliability.

A theater army area of operations may have tens of thousands of emitters of electromagnetic energy organized into many radio nets. The tactical situation determines the land mass occupied by this army, and expansion or contraction of the area will increase or decrease the interference level together with the desired signal strength.

The relatively dense concentration of emitters, as well as the possibility of intentional jamming, makes it essential that incidental equipment noise be as low as possible in order to enhance transmission quality and minimize adjacent channel and other spurious emission interference.

The heavy demand for radio frequencies makes it necessary to use the same frequency simultaneously in various nets. To prevent co-channel and adjacent channel interference, a trained frequency management officer assigns frequencies on a priority basis, utilizing data on geography, restrictions on time usage, power, bandwidths, and types of antennas. (See DA Pamphlet No. 105.2* and ST 24-2-1 for details, Ref. 2).

2-2.3.2.3 Ship and Aircraft

In many respects these environments are similar to that of the ground station. A principal difference is one of scale. Although very large ships can be considered to be a complete ground station on a mobile platform, the largest Army watercraft are likely to carry typical radar and communications equipment. Since antennas and various items of electrical equipment are in permanent locations, to some extent the installation can be designed to optimize the compatibility situation. However, the flexibility available is small and the consequences of any signal degradation whatsoever are likely to be serious. Below steel decks or within an aircraft fuselage, one has good examples of what are usually defined as intrasystem compatibility problems. In these locations potentially interacting equipments and their cables may be in close proximity, and antenna coupling is almost nonexistent. Above decks or external to the fuselage, both types of interaction are possible.

2-2.3.2.4 Missile

This item is placed in a unique classification because of the fact that it must operate with high reliability through rapidly changing external environments. At low altitudes it may pass quite close to urban areas containing relatively high-powered transmitters of both military and broadcast types. At high altitudes the missile is within line of sight of large areas of the earth and, therefore, large numbers of signals have the potential of affecting it. In addition, because of the high density of electronic components, its intrasystem environment is especially severe.

2-2.3.2.5 Equipment

This is the environment that can be found within a typical cabinet or metallic enclosure of some device.

*DA Pamphlets and AR's referred to in this paragraph are described in Table 2-6.

It is an extreme example of an intrasystem environment. Frequently, such an enclosure contains many separate electrical and electronic parts which must interact in a specified way in order to perform the useful function for which the equipment was designed. At the same time, these components may interact in ways that are not intended, and the unintended interactions may be caused by emission levels introduced into the inside of the equipment by the penetration of external fields or currents through the walls of the container. By proper design, such penetration phenomena can be controlled.

At the component level, environment can be controlled frequently by proper physical placement of components or by the use of filters and shields to isolate them from undesired interactions with other components. In other cases specific circuit designs can be used to reduce their susceptibility to local fields or to reduce the fields which they generate.

2-2.3.2.6 Other Categories

Other categories which are not explicitly covered may be considered to be combinations of the preceding. An example is the airport which may be compared with a military ground station. It has receivers, transmitters, radars, and, at least in civilian airports, is usually surrounded by commercial and industrial areas with high radio noise levels.

The environmental characteristics discussed in this paragraph are produced by electromagnetic phenomena which are analyzed explicitly in later chapters. In the treatment, methods of determining steps to be taken to avoid a particular effect or to reduce interactions that are experienced will be included. For reference purposes, Table 2-1 shows where some of the major types of source of interference contribute most importantly in each of the major types of environment. Also, reference is made to paragraphs in Chapter 3 which give quantitative information of use in a system or equipment analysis. Clearly, before such an analysis can be carried through, the design engineer must have familiarity with the theory and practice of EMC/EMI engineering. Further discussion of prediction techniques is given in Chapter 6.

2-3 SPECTRUM ENGINEERING

The radio spectrum is a limited resource (Ref. 3). The uses of the spectrum are expanding and, at any given time, the demand exceeds the supply. Furthermore, some services are best carried out in certain frequency ranges and demand in these ranges continues to grow. By international agreement and by decisions of national regulatory bodies, parts of the spectrum are allocated to particular services. These are further subdivided into channels whose width is sufficient to accommodate the signals typical of the service. Thus a principal microwave radio relay allocation is a band between 3700 MHz and 4200 MHz subdivided into 24 channels with adjacent channels separated by 20 MHz. Each channel is, in turn, able to accommodate 600 multiplexed voice channels. Electromagnetic compatibility requires a minimal geographic spacing between users assigned to the same channel, and also between users assigned to adjacent channels. In order to fit as many microwave relay users as possible into a geographical region, advantage is taken of the rejection afforded to unwanted signals by antenna directivity, and also by the

TABLE 2-1. IMPORTANT CONTRIBUTORS TO VARIOUS ELECTROMAGNETIC ENVIRONMENTS

Noise Source	Type of Environment					This Handbook Paragraph References
	Rural	Urban	Intersystem	Intrasystem	Equipment	
Atmospheric	X	X				3-1.2.2
Transmitters ISM	X	X	X	X		3-1.3.1.2
Spurious RF Generation			X	X	X	3-1.3.3.3
Ignition	X	X	X			3-1.3.2.4
Switching Transients				X	X	3-1.3.2 3-1.3.3.2
Powerline Harmonics				X	X	3-1.3.3.1

use of orthogonal polarizations. Allocations are sometimes shared by different services where the expectation of mutual interference is small. Channel assignments sometimes are made with the provision that no interference shall result to other users who have priority, or that transmission will occur during a specified time only.

The process by which assignments are made takes account of the need for the channel; the number of channels available; the noise level, both natural and man-made; the characteristics of receivers; the range and area to be covered; the location, whether fixed or mobile; the kind of signal to be used (the modulation and bandwidth); and the time during which it is required. An assignment specifies the allowable transmitted power, the limits of spatial coverage by the antenna, as well as frequency, time, and place.

Frequency allocations are made internationally through the International Telecommunications Union and, in the USA, by the Federal Communications Commission and the Office of Telecommunications Management. Because of the shortage of spectrum space, no development of Department of Army equipment requiring spectrum space can be initiated without a formal action regarding frequency allocation (see AR 705-16).

Intrasystem compatibility problems usually are ignored in frequency allocation procedures, it being assumed that by proper design techniques system degradation due to this cause can be avoided. To the designer, however — especially where systems are colocated, such as on a vehicle or at a command post — the considerations for dealing with intrasystem and intersystem compatibility may be quite similar.

2-4 ACHIEVING ELECTROMAGNETIC COMPATIBILITY

Achieving electromagnetic compatibility can be approached from the concept of either (1) a detailed evaluation of interactions between the elements, emitters and susceptors, on a pair by pair basis; or (2) from general requirements. In the latter case the designer attempts to define a so-called "typical" or "nominal" environment, and then designs his equipment so as to neither degrade the environment by raising levels of radio noise above the "nominal" level (at a specified distance from the equipment) nor to be susceptible to those levels. In general, neither of these two approaches can be used in isolation.

To some extent these approaches can be associated with two of the basic types of specification, i.e., the "system" specification and the "equipment" specification (see par. 2-5.2). The latter establishes equipment limits which will effectively "control" the equipment contribution to environmental levels. The "system" specification establishes system performance requirements which in turn require a design that directly controls interactions between individual components of equipments.

2-4.1 THE SYSTEM APPROACH

The main fact to be recognized in system design is that each system is unique and therefore EMC requirements are likely to be unique. It is the overall system requirements that determine what the component equipment EMC requirements are and the extent to which standard requirements will be suitable.

Just as with equipment design, these requirements must be determined initially before manufacture. Thus one must obtain, early in the design and development stages, an adequate description of the ultimate circumstances under which a specific system is used. Then it is possible to initiate design of the system components in accordance with the derived requirements. To do this, the designer must have a reliable interference prediction process readily available. This means having adequate models of emitters, susceptors, and the couplings between them. The engineer also must have the same type of information pertinent to equipment design including data on interference reduction circuits and techniques and on cost effectiveness trade-offs.

In practice the entire process is usually not as complicated as this description might lead one to believe. It is probable that only a limited number of parameters involved in the system design will exhibit critical cost-effectiveness trade-offs. Many of the possible interactions can be controlled effectively by design techniques which do not, in themselves, inherently lead to very high costs. The system designer must be able to identify those elements of the system which are critical with regard to cost and performance, and must deal with these in a first-order analysis with properly assigned priorities.

2-4.2 REVIEW OF APPROACHES TO EMC

Different approaches to EMC control are exemplified by those used by the Federal Communications Commission, industry, and the Department of Defense. Since the technologies involved are interrelated, the discussion is relevant to a variety of typical EMC problem areas.

2-4.2.1 The Federal Communications Commission

In carrying out its responsibilities for regulating radio communication, the Federal Communications Commission has been faced with the very practical question of how to write the rules so that they can protect effectively the various services under its cognizance, while at the same time not unduly burden those who provide the services. To assure this, the Commission makes effective use of the public hearing procedure. In this procedure, the Commission publishes a notice of proposed rule-making and invites comments from those involved in providing the service, the public at large making use of the service, and industry which manufactures the equipment which will have to meet the requirements. After some study, a set of limits is proposed. These limits are then discussed at great length. After duly taking into account all the evidence before it, the Commission finally issues specific rules and regulations. In doing this, the Commission makes use of both the system approach and the environmental approach.

The environmental approach can be illustrated by the method used to control interference to radio broadcasting. Here, one can look upon the procedure as protecting a defined field strength at a specified distance from an interference source. The field strength is a function of the service involved, while the distance at which protection is to be obtained is a function of the source, and where it is used. For example, sources arising in industrial environments where receiver antennas are not likely to be located are permitted to radiate more energy than those which appear in homes near receivers or receiver antennas. Also taken into consideration are such factors as whether danger to life is involved, economic aspects (which are a function of the state of the art), and the possibilities of alternate means of obtaining the same type of service. In many respects the rule-making procedure is a way of establishing a *de facto* cost-effectiveness trade-off relationship.

In other cases, the Commission adopts an equipment interaction approach in that it establishes standards for various types of transmitting equipment and allocates frequencies based upon the assumption that the equipment meets those standards. For example, transmitter harmonic levels are controlled along with antenna patterns. A person applying for a transmitter license must show that his equipment will not produce interference with other services already established in the same area.

In order to simplify FCC rule-making procedures, unintentional emitters have been classified in three general categories (Refs. 4, 5). Examples are given in Table 2-2. The major features of the regulations are summarized in Tables 2-3, 2-4, and 2-5.

2-4.2.1.1 Incidental Radiation Device

A device that radiates radio frequency energy during the course of its operation, even though the device is not intentionally designed to generate radio frequency energy, is classified as an incidental radiation device (Ref. 4, part 15).

TABLE 2-2. EXAMPLES OF ELECTRONIC DEVICES REGULATED BY FCC RULES PARTS 15 AND 18

Regulated by Part 15		Regulated by Part 18
Restricted Radiation Devices RF Energy Purposely Generated	Incidental Radiation Devices RF Energy Unintentionally Generated	Industrial, Scientific, and Medical
1. Radio Receivers 2. Carrier Current Systems a. "Campus" b. Telephone c. Industrial 3. Low-power Communications a. Wireless Microphones b. Garage Door Openers c. Phonograph Oscillators 4. Field Disturbance Sensors 5. Class I TV Devices	1. Fluorescent Lights 2. Electric Appliances 3. Electric Motors 4. Electric Shavers 5. Ignition Systems 6. Defective Insulators	1. Industrial Heaters a. Induction b. Dielectric 2. Medical Diathermy 3. Miscellaneous a. Epilators b. Ultrasonic c. Microwave Ovens d. RF Neon Signs

TABLE 2-3
TECHNICAL REQUIREMENTS FOR PART 15 (INCIDENTAL AND RESTRICTED RADIATION) DEVICES

TYPE OF EQUIPMENT	FREQUENCY OF RADIATION, MHz	FIELD STRENGTH LIMIT, μV/m	ADDITIONAL REQUIREMENTS	REFER TO SECTION**
Incidental Radiation Device* (Subpart B)	Any frequency	None	In the event that harmful interference is caused, the operator shall promptly take steps to eliminate the harmful interference.	15.31
Radio Receivers (30-890 MHz) (Subpart C) TV broadcast	0.45-25	None	Power line RF voltage limitation: 100 μV/m	15.63
All other	0.45-9	None	Power line RF voltage limitation: 100 μV/m	15.63
	9-10	None	Power line RF voltage limitation: 100-1000 μV/m*	15.63
	10-25	None	Power line RF voltage limitation: 1000 μV/m	15.63
	25-70	32 at 1000 ft	Applicable to All Receivers (30-890 MHz)	
	70-130	50 at 1000 ft	Certification required for each model receiver.	15.69
			Identify each certified receiver with seal or label.	15.70
	130-174	50-150 at 1000 ft		15.71
			Any measurement procedure acceptable to the FCC may be used. The following standards are considered acceptable:	15.75
	174-260	150 at 1000 ft		
	260-470	150-500* at 1000 ft	Power line RF voltage: IEEE STD 213 Radiation: IEEE STD 187; also IEC 10b/10ba and EIA RS 378	
	470-1000	500 (350 for TV rcvr, see Sec. 15.63) at 1000 ft	TV sensitivity: IEEE STD 190	
Door Openers (receivers)	25-70	32	Measurements according to FCC document T7001.	
	70-200	50		
	200-1500	50-500*		
	over 1500	500		
Low Power Communication (Subpart E)			Max. Input to Final, W / Max. Antenna Length, ft	
Operation Under Antenna, and Power Limitations	0.16-0.19	None	1.0 / 50*	15.203
	0.51-1.6	None	0.1 / 10*	15.204
	26.97-27.27	None	0.1 / 5	15.205
Operation under Field Strength Limitation	0.010-0.490	2.4/f(in MHz) at 1000 ft		15.202
	0.510-1.6	2.4/f(in MHz) at 1000 ft		15.202
	70-1000	receiver radiation limit applies	Limit operation automatically to 1 s on followed by 30 s off	15.211
	above 1000	500 at 100 ft	Limit operation automatically to 1 s on followed by 30 s off	15.211
Telemetering and Wireless Microphone	88-108	50 at 50 ft in 200 kHz band 40 at 10 ft outside of band	Channel width is limited to 200 kHz.	15.212

* including transmission line

** References are to Part 15 or Part 18, as applicable, of the FCC Rules and Regulations.

(cont'd)

TABLE 2-3 (cont'd)

TYPE OF EQUIPMENT	FREQUENCY OF RADIATION, MHz	FIELD STRENGTH LIMIT, μV/m	ADDITIONAL REQUIREMENTS	REFER TO SECTION**
Biomedical Telemetering Devices	174-216	150 at 100 ft	Limit applies only to the fundamental frequency. Harmonic and other spurious emissions outside of 174-216 MHz shall be suppressed by at least 20 dB (subject to change in Docket 19846).	15.215
Door Openers (transmitters)	70-130	125 at 100 ft	The emission permitted on 73-75.4, 108-118, 121.4-121.6, 242.8-243.2, 265-285, 328.6-335.4, 406-410, 608-614, 960-1215 (all in MHz) and other bands. See Sec. 15.215.	15.215
Field Disturbance Sensors (Subpart F)	Any Frequency	15 at 157/f(in MHz) ft	Limit applies to the fundamental, any harmonic, and other spurious frequencies.	15.305
	915	50,000 at 100 ft	Band limited to ±13 MHz	15.307, 15.309
	2450	50,000 at 100 ft	Band limited to ±15 MHz	15.307, 15.309
	5800	50,000 at 100 ft	Band limited to ±15 MHz	15.307, 15.309
	10525	250,000 at 100 ft	Band limited to ±25 MHz	15.307, 15.309
	24125	250,000 at 100 ft	Band limited to ±50 MHz	15.307, 15.309
			Applicable to All Field Disturbance Sensors Certification Measurement	15.303, 15.317

** References are to Part 15 or Part 18, as applicable, of the FCC Rules and Regulations.

TABLE 2-4. FREQUENCIES FOR INDUSTRIAL, SCIENTIFIC, AND MEDICAL (ISM) EQUIPMENT FOR WHICH NO RADIATION LIMITS ARE SPECIFIED

Frequency, kHz	Tolerance, kHz	Frequency, MHz	Tolerance, MHz
13,560	± 6.78	915	± 13
27,120	± 160	2,450	± 50
40,680	± 20	5,800	± 75
		24,125	± 125

There are no regulations specifying constraints on radiation from devices in this category other than if harmful interference does occur, immediate steps must be taken to eliminate it. (This latter regulation is applied to all incidental device categories.)

2-4.2.1.2 Restricted Radiation Device

A restricted radiation device is one in which the generation of small amounts of RF energy is intentionally incorporated into the design, but does not require licensing.

To control interference from restricted radiation devices, unless otherwise stated, the limit field strength is 15μV/m at a distance of $\lambda/(2\pi)$ where λ is the wavelength corresponding to the operating frequency. The design must be in accordance with the best engineering practice, and operate with the minimum power required to produce the desired result. It may be noted that the distance of $\lambda/(2\pi)$ is the distance from a dipole at which the variation in field strength with distance changes from an inverse cube relationship (induction field) to an inverse first power relationship (radiation field). Also at a fixed distance

TABLE 2-5. TECHNICAL SPECIFICATIONS FOR PART 18 (INDUSTRIAL, SCIENTIFIC, AND MEDICAL SERVICES) DEVICES

TYPE OF EQUIPMENT	FUNDAMENTAL FREQUENCY OF RADIATION, MHz	FIELD STRENGTH LIMIT OUTSIDE OF ISM BANDS	ADDITIONAL REQUIREMENTS	REFER TO SECTION*
Medical Diathermy	13.56, 27.12, 14.68	25 μV/m at 1000 ft		18.141
(Subpart E)	915, 2450, 5800, 24135	Reduce radiation to greatest extent practicable.	Reduce bandwidth of emissions to the greatest extent practicable	18.142
	Any Other Frequency	15 μV/m at 1000 ft	15 μV/m at 1000-ft limit also applies to radiation at the fundamental	18.142
			Applicable to All Medical Diathermy Equipment Use FCC radiation measurement procedure. Certification or Type Approval. Certificate must be renewed every 3 yr. Equipment operated on off-ISM frequencies requires a rectified and filtered power supply.	 18.143 18.144 18.141 18.142
Ultrasonic (Subpart C)	Up to 0.490	24/f(in MHz) μV/m at 1000 ft	In predominantly residential areas and on frequencies below 0.490 MHz the radiation limit may be increased as the square of the generated power to 500 W. The limit, however, is not permitted to exceed 10 μV/m at 1 mi.	18.72
	0.490-1.6	24/f(in MHz) μV/m at 100 ft	Applicable to All Ultrasonic Equipment Use FCC radiation measurement procedure. Power line RF voltage: below 0.49 MHz-1000 μV above 0.49 MHz-200 μF	 18.78 18.72(e)
	Over 1.6	15 μV/m at 100 ft	Certification, or Type Approval.	18.80 18.73
Industrial Heating (Subpart C)	Below 5725	10 μV/m at 1 mi	Power line radiation limit: 10 μV/m at 50 ft at points 1 mi or more from the equipment. Use FCC radiation measurement procedure.	18.102 18.107
	Above 5725	Reduce radiation to greatest extent practicable.	Applicable To All Industrial Heaters Operation not permitted on: 0.490-0.510 MHz, 2.170-2.194 MHz, and 8.354-8.374 MHz Certification. Periodic Inspection.	 18.102(b) 18.113 18.105
RF Stabilized Arc Welders (Subpart F)	Any Frequency	10 μV/m at 1 mi	Power line radiation limit: 10 μV/m at 50 ft at points 1 mi or more from the equipment. Use FCC radiation measurement procedures. Measure quasi-peak using an instrument equivalent to ANSI specifications C63.2 to C63.4 Certification, or Type Approval.	18.5 18.107 18.181 18.182 18.181
Miscellaneous (Subpart H)	These rules apply to ISM equipment other than medical diathermy, industrial heating, ultrasonic and RF Stabilized arc welders in which RF energy is applied to materials to produce physical, biological, or chemical effects such as microwave ovens, ionization of gases, mechanical vibrations, hair removal, and acceleration of charged particles which do not involve the use of radio receiving equipment. Requirements are the same as for medical diathermy except that for equipment other than that used in predominantly residential areas, the radiation limit may be increased as the square root of the generated power to 500 W. The limit, however, is not permitted to exceed 10 μV/m at 1 mi.			

* References are to Part 18 of the FCC Rules and Regulations.

this limit decreases with frequency, which corresponds to the behavior of atmospheric radio noise levels.

2-4.2.1.3 Radio Receivers

Radio receivers, including television broadcast receivers, are treated as a special case of restricted radiation devices.

For frequencies above 25 MHz the radiation from these devices is limited to specific field strengths measured at specific disances. The limit increases somewhat with frequency corresponding partly with state of the art, and partly with increasing cosmic noise levels. At 25 MHz and below, a limit is placed on voltages conducted to the power lines since this is considered to be the most likely source of coupling from such devices.

2-4.2.1.4 Low Power Communication Devices

A low power communication device is a restricted radiation device (Ref. 4) exclusive of those employing conducted or guided radio frequency techniques used for transmission of signs, signals (including control signals), writing, images, and sounds or intelligence of any nature by radiation of electromagnetic energy (e.g., wireless microphone, phono oscillators, garage door openers, radio control models).

The design and operation of low power communication devices are more closely regulated than are those of the previously discussed devices. The general rule requiring the immediate elimination of harmful interference applies here as in all other categories. Allowed bands of operation have been established (see Table 2-3) and, in the case of telemetry transmitters and wireless microphones, maximum channel widths have been established. Limits of radiated energy, in-band and out-of-band, and within-channel and out-of-channel, have been established for all but one band. A limit to the energy which may be conducted to the power line has been established in the 510 - 1600 kHz bands. In the 26.97 - 27.27 MHz band where no radiation limit has been directly established and as an alternative method in the 160 - 190 kHz and 510 - 1600 kHz bands, device design limits are used which effectively serve to limit the device radiation. An upper limit is set for input power to the final amplifier and the length of the antenna (5 ft for the 26.97 - 27.27 MHz band) or the combined length of the antenna and its associated transmission line. In addition, the level of power input to the final amplifier lying outside of the band of operation (e.g., if operation is on the 160 - 190 kHz band, for all frequencies above 190 kHz or below 160 kHz) must be suppressed at least 20 dB below the level of the unmodulated carrier. An additional design limitation prohibits class B operation.

Field disturbance sensors are classed as restricted radiation devices but, because of their special function, have special limits assigned in the GHz range and above.

2-4.2.1.5 Industrial, Scientific, and Medical Equipment (ISM)

The ISM category of equipment (Ref. 5) includes devices which use radio waves for industrial, scientific, medical, or other purposes which are neither used or intended to be used for communication.

Operating frequencies have been allocated, and allowable tolerances established for which there are no radiation limits (Table 2-4). Operation is permitted at other frequencies but radiated emissions are limited (Table 2-5). These limits are defined in terms of radiated field strength at a specified distance, and apply to radiated harmonics as well as spurious radiation. In addition, conducted emission limits have been established for ultrasonic equipment.

Medical diathermy equipment, when operated on frequencies other than those assigned for ISM operation, are subject to a radiated emission limitation in terms of a maximum field strength measured at a specified distance. There are, as well, some design requirements which will suppress spurious and harmonic frequency radiation; for example, a rectified and filtered plate voltage must be used.

Industrial heating equipment is subject to a limitation as to its radiated field strength at a specified distance when operated below 5725 MHz and on any frequency other than those established for ISM equipment. Above 5725 MHz the regulations only specify that radiation must be suppressed as much as possible. Though there are no specific limits to conducted emissions to the power lines from industrial heating equipment, it is required that sufficient filtering be provided to limit radiation from power lines. A limit to power line radiation is provided in terms of radiated field strength at a distance. These limits represent compromises between practical state-of-the-art considerations and use in specialized areas.

2-4.2.1.6 Licensing of Test Facilities

Unless a Government agency assumes responsibility for running tests, test facilities set up by contractors require FCC licenses or at least temporary authorization if substantial radiation may occur, either in connection with scientific studies or for testing of communication equipment including transmitters and antennas. Requirements are covered in

Parts 5 (Ref. 9) and 15 (Ref. 4) of the Rules and Regulations. At frequencies below about 10 MHz the criterion for obtaining FCC approval is considered to be whether emission levels will exceed 15 μV/m at a distance of $\lambda/(2\pi)$ where λ is the wavelength of the radiation.

In the usual case, a contractor will not require a formal radio station license (Ref. 9, par. 5.52) but will be able to qualify for a special temporary authorization (par. 5.56) as an experimental radio service (Research or Developmental) (par. 5.2).

Informal applications for special temporary authorizations for Government contractors should:

a. Be prepared as a letter, in duplicate with the original signed and sent to the Federal Communications Commission, Washington, DC 20554.

b. Contain the following information:

(1) Name and address

(2) Need for special action

(3) Type of operation to be conducted

(4) Purpose of operation

(5) Time and date of proposed operation

(6) Class of station, call sign of station, and nature of service

(7) Location of proposed operation

(8) Equipment to be used, including name of manufacturer, model, and number of units.

(9) Frequency(s) desired

(10) Plate power input to final radio frequency stage

(11) Type of emission

(12) Antenna height (FCC Form 401-A if par. 5.55 so requires)

(13) Full particulars as to the purpose of the request.

c. Include:

(1) FCC Form 44A in triplicate "Supplemental Information for Applications in the Experimental Radio Service Involving Government Contracts".

(2) If the application involves communications essential to a research project:

(a) A description of the nature of the research project being conducted

(b) A showing that communication facilities are necessary for the research project involved

(c) A showing that existing communication facilities are inadequate.

If a Government agency assumes test responsibility, it must obtain permission from the Interdepartment Radio Advisory Committee (see par. 2-6.4.5).

2-4.2.2 Industry Standards

Industry in the United States is well aware of EMI phenomena and has devoted much effort to its control. In many instances the interference from particular devices is controlled in accordance with limits established by individual companies. In those cases in which the interactions have been sufficiently significant to justify a joint industry effort, standards have been published. These limits generally are based upon a carefully examined cost-effectiveness trade-off. Primary concern has been with interference to radio and television broadcasting, although recently there has been a growing concern about establishing electromagnetic compatibility between industrial and commercial use devices. Industry has devoted most of its efforts to establishing measurement techniques and instrumentation, as exemplified by the following standards published by the American National Standards Institute:

a. C63.2-1963, Radio-Noise and Field-Strength Meters, 0.015 to 30 Megacycles/Second, Specifications for.

b. C63.3-1964, Radio-Noise and Field-Strength Meters, 20 to 1000 Megacycles/Second, Specifications for.

c. C63.4-1963, Radio-Noise Voltage and Radio-Noise Field Strength, 0.015 to 25 Megacycles/ Second, Low-Voltage Electric Equipment, and Nonelectric Equipment, Methods of Measurement of.

Industry groups have established voluntary limits as follows:

a. Society of Automotive Engineers, Inc.: J551, "Measurement of Electromagnetic Radiation from Motor Vehicles (20-1000 MHz)."

b. National Electr[illegible] Manufacturers Association: "Semiconducto[illegible] Dimmers for Incandescent Lamps," WD2-197[illegible]

c. Radio Technic[illegible] Committee for Aeronautics: DO-138, "Environmen[illegible] Conditions and Test Procedures for Airborne El[illegible]ronic/Electrical Equipment and Instrument."

d. International Special Committee on Radio Interference (CISPR):

(1) Publication 11: "Limits and methods of measurement of radio interference characteristics of industrial, scientific and medical (ISM) radio-frequency equipment (excluding surgical diathermy apparatus)," 1st edition, 1975.

(2) Publication 12: "Limits and methods of measurement of radio interference characteristics of ignition systems of motor vehicles and other devices," 1st edition, 1975.

(3) Publication 13: "Limits and methods of measurement of radio interference characteristics of sound and television receivers," 1st edition, 1975.

(4) Publication 14: "Limits and methods of measurement of radio interference characteristics of

household electrical appliances, portable tools and similar electrical apparatus," 1st edition, 1975.

(5) Publication 15, "Limits and methods of measurement of radio interference characteristics of fluorescent lamps and luminaires," 1st edition, 1975.

2-4.2.3 Department of Defense EMC Program

The Department of Defense has established an integrated Electromagnetic Compatibility Program (EMCP) to ensure electromagnetic compatibility of all military communication-electronic (C-E) equipments, subsystems, and systems during conceptual, design, acquisition, and operational phases (see DoD Directive 3222.3). Its objectives are to: (a) achieve compatibility of all equipments, (b) to obtain built-in design compatibility rather than remedies added after the fact, and (c) use common approaches and techniques in C-E material programs. To accomplish this it directs action and assigns responsibility in the following EMC program areas: Standards and Specifications, Measurement Techniques and Instrumentation, Education for EMC, Data Base and Analysis Capability, Design, Concepts and Doctrine, Operational Problems and Test and Validation.

DoD recognizes that frequency management has a strong influence on compatibility and has assigned responsibilities for the use of the radio frequency spectrum. Specific spectrum management functions are delegated to individual departments. Departmental membership in the Interdepartmental Radio Advisory Committee (IRAC) and liaison with FCC provides interdepartment, international, and national (non-Government) coordination. DoD directives address themselves to spectrum management, and EMC programs and equipment. Typical examples are:

a. DoDD 4630.1, Programming of Major Telecommunications Requirements

b. DoDD 4630.5, Compatibility and Commonality of Equipment for Tactical Command and Control and Communications

c. DoDD 4650.1, Management and Use of the Radio Frequency Spectrum

d. DoDD 5000.3, Test and Evaluation

e. DoDD 5100.35, Military Communications-Electronics Board (MCEB).

2-4.2.4 Department of The Army EMC Program

The Department of the Army has established an EMC Program consistent with the Department of Defense EMC Program. It directs that electromagnetic compatibility of materiel be controlled throughout each phase (concept, validation, full-scale development, cost production, and deployment) of the Life Cycle System Management Model (LCSMM) (par. 2-7), and recognizes three principal factors which must be addressed by the EMC Program:

1. Environmental Geometry: The identification of all emitters and receptors, the spatial relationships among the emitters and receptors, and the particulars regarding terrain and propagation

2. C-E Materiel Characteristics: All features which influence the degree of performance achievable under given electromagnetic environmental operating conditions

3. Frequency Assignments: The distribution of available spectrum resources.

A number of regulations and pamphlets, as shown in Table 2-6, describe policies and procedures related to matters of direct concern in EMC programs related to Army materiel procurements.

2-5 THE EMC PROGRAM REQUIREMENTS

MIL-HDBK-237 (Ref. 6) provides criteria for establishing, managing, and evaluating an EMC program for electronic, electrical, and electromechanical equipments, subsystems, and systems. It provides EMC guidance to those responsible for the program so as to increase the probability of achieving intrasystem and intersystem compatibility.

2-5.1 EMC PROGRAM GOALS

In broad terms, the goals of any EMC program are (Ref. 6):

a. Assure efficient integration of engineering, management, and quality assurance tasks, as related to EMC.

b. Assure the efficient integration of EMC with all other system performance factors and disciplines affecting system effectiveness and cost, such as reliability, maintainability, vulnerability, and safety.

c. Assure the integration of the engineering functions — such as design, development, and test — as related to EMC.

d. Assure intrasystem and intersystem design and operational electromagnetic compatibility.

e. Assure continuous traceability of EMC requirements and design alternatives throughout the program, so that the sources and impact of design changes and deficiencies in equipment and subsystems, and the impact of contractual performance requirements are promptly determined, accurately identified, and properly communicated.

f. Permit timely and optimum redefinition of EMC requirements in response to changes directed

TABLE 2-6. ARMY REGULATIONS AND DA PAMPHLETS CONTAINING EMC GUIDANCE

NO.	TITLE	PURPOSE	EMC PROVISIONS	RELATION TO OTHER DIRECTIVES
10-5	Organizations and functions, Dept. of the Army	Provides specific responsibilities of the general and special staff	Responsibilities for the EMCP frequency management and, commonality of C-E eqmt	Implemented in AR's 11-13, 15-14, 70-1, 70-27, 105-3, etc.
11-13	Army Electromagnetic Compatibility Program	Provides policies and procedures for the EMCP	Policy guidance for EMCP and program of activities associated with the LCSMM	Implements DoDD 3222.3 and AR 10-5 for EMC. Relates to AR's 105-16, 105-2, 105-3, 105-16, 105-24
DA Pam 11-13	Army Electromagnetic Compatibility Program Guide	To assist developers in electromagnetic compatibility considerations and frequency supportability during life cycle system management	Describes the processes necessary to comply with AR 11-13; indicates sources of EMC analytic support in areas of analysis, test, and measurement	Supplements AR 11-13 with emphasis on intersystem EMC decisions/actions
DA Pam 11-25	Life Cycle Systems Management Models for Army Systems	Provides step by step procedures for systems acquisition process	Check lists for IPR and ASARC reviews include EM requirements	Implements AR's 1000-1, 15-14, and 11-25
15-14	Systems Acquisition Review Council Procedures	Provides procedures for ASARC reviews	Responsibilities for ASARC review process. Check lists for each ASARC include EM considerations.	Implements AR 1000-1
70-1	Army Research, Development, and Acquisition	Establishes responsibilities, policy, and general procedures for Army R&D	Responsibilities to R&D. Need to coordinate RF spectrum and EM requirements as early as possible.	Cites AR's 11-13 and 105-16
70-10	Test Evaluation during Development and Acquisition of Materiel	Describes objectives, concepts, responsibilities, policies for T&E	Provision for EMC testing in DT and OT, and outline test plans for CTP	Implements DoDD 5000.5, and AR's 1000-1 and 70-1, Relates to AR 71-3
70-27	Outline Development Plan/Development Concept Paper/Program Memorandum	Prescribes procedures for preparation and content of DP/DCP/PM	EMC considerations are provided in outline for Systems Summary in App. B.	Implements AR's 1000-1 and 15-14
71-1	Army Combat Developments	Establishes policy and procedures for combat development activities including guidance for preparation of future requirements	Responsibilities for combat development and EM considerations in OCO's and requirement operational characteristics	Implements AR 1000-1 in the area of combat developments and provides guidance for AR 71-3
71-2	Basis of Issue Plans	Establishes policy and procedures for basis of issue plans for new equipment	Establishes coordination requirements for EMC.	Relates to AR's 11-25, 71-1, 310-31
71-3	User Testing	Prescribes policies and procedures for operational testing	EMC content of operational tests	Implements AR 71-8 and DODD 5000.3
DA Pam 105-2	Management of the Electromagnetic Spectrum	Provides information background on spectrum management at the international, national, and DOD levels and a description of the Army EMC program	Spectrum management aspects of EMC	Implements and provides background for AR's 105-16, 105-24, 105-63, and 105-67
105-3 (C)	Reporting Meaconing, Intrusion, Jamming, and Interference (MIJI) of Electromagnetic Systems	Establishes procedures for reporting and evaluating information concerning incidents of MIJI of US military electromagnetic equipment or systems	MIJI reporting concerned with electromagnetic emissions. Responsibilities are assigned, procedures decribed, and format presented.	Responds to JCS tasking in SM528-70

(cont'd)

TABLE 2-6 (cont'd)

NO.	TITLE	PURPOSE	EMC PROVISIONS	RELATION TO OTHER DIRECTIVES
105-16	Radio Frequency Allocations for Equipments under Development, Production, and Procurement	Prescribes procedures for obtaining RF spectrum allocations	EMC considerations associated with spectrum allocation	Relates to AR's 11-13, 105-24, 105-63, 105-67; and DA PAM 105-2
105-22	Telecommunications Requirements Planning, Developing, and Processing	Prescribes policies, responsibilities, and procedures for telecommunications requirements within DA	Specifies responsibilities for compliance with the EMCP and makes special provisions for inclusion of SIGSEC requirements	Relates to AR's 11-13, 105-1, 530-2, and 530-4
105-24	Radio Frequency and Call Sign Assignments for Army Activities within the Continental United States	Assigns responsibilities and establishes procedures for frequency assignments in CONUS	EMC considerations associated with frequency assignments	Relates to AR 105-16 and DA PAM 105-2
105-63	Army Electromagnetic Spectrum Usage Program	Provides instructions for reporting usage of assigned frequencies in 4-30 MHz band	Provides data for RF utilization file	Relates to AR's 105-16 and 105-24, DA PAM 105-2
105-67	Electromagnetic Compatibility Program — Reporting of US Military Electronic Equipment Environmental Data	Prescribes reporting of C-E equipment in CONUS to ECAC EM environment file	Provide data for ECAC file	Joint regulation implementing part of DODD 3222.3. Relates to AR's 11-13, 105-16, 105-24; and DA PAM 105-2
1000-1	Basic Policies for Systems Acquisition by the Department of Army	Prescribes basis policies and goals for systems acquisition	None specifically	Implements DODD 5000.1. Provides guidance for all regulations and directives governing R&D and systems acquisition

by the procuring authority, or problems identified through performance measurements, if required.

g. Provide inputs to, and assure compatibility with, all interfacing management and information systems employed in the project.

2-5.2 EMC PROGRAM RESPONSIBILITIES

In order to meet EMC program goals, the EMC program must (Ref. 7, par 15.1.2):

a. Establish a single, authoritative coordinated organization responsible for all EMC program matters.

b. Designate specific responsibility for the preparation of EMC control plans, test plans, reports production specifications, and other associated documentation. The relationship of other organizational elements which provide inputs for these documents to the organization of primary EMC responsibility must be clearly defined.

c. Designate responsibility for EMC testing at all levels. This should include all equipment/subsystem level testing activity — such as development, qualification, and engineering — as well as systems level testing activity such as qualification acceptance and compliance.

d. Require EMC representation during the various program reviews and design reviews. If the review is performed by a designated team, one of the team members will be an EMC engineer.

e. Designate responsibility for preparation, approval, and submission of all requests for deviation from EMC requirements or specifications.

2-5.3 PROGRAM ORGANIZATION

In establishing a suitable program, the following must be considered:

a. Possible conflicts in requirements between EMC documents and other applicable standards and specifications for the specific procurement involved.

b. Requirements for establishing an EMC advisory board

c. Requirements for specific EMC analyses

d. Coordination of EMC requirements with other engineering requirements such as those associated with electromagnetic hazards, electronic warfare, security compromising by electromagnetic emissions, susceptibility to electromagnetic pulse, measurements, lightning and static protection

e. System EMC requirements

f. Subsystem and equipment emission and susceptibility control

g. Need or advisability of tailoring EMC limits for the specific application.

2-5.4 DETAILED REQUIREMENTS

In addition to the broadly applicable requirements already mentioned, materiel programs are bounded and defined by requirements initiating the program. As the program progresses, the initial requirements, as well as additional standard and tailored requirements, are used to regulate program activities.

Documentation placing mandatory requirements on program management includes general military standards and specifications (systems and equipment) as well as requirements specific to EMC/EMI. The most important documents are identified in Table 2-7.

2-6 EMC PROGRAM PLANNING

EMC programs should be directed by a System Program Plan which in turn usually contains the following EMC plans (Ref. 6):

a. EMC Program Plan

b. EMC Control Plan

c. EMC Test Plan.

It should be noted that, for the purpose of establishing suitable programs, systems have been divided into two classes, major systems and nonmajor systems (see par. 2-7). Generally, the detail described in the paragraphs that follow for these three plans can be considered to apply to the first class. Certain details can be omitted in a system of the second class.

2-6.1 EMC PROGRAM PLAN

To achieve the greatest EMC engineering benefits, management and engineering personnel must establish the necessary EMC program early in the program life cycle. The EMC program plan, which is usually prepared as a separate part of a proposal or contract work statement, documents clearly defined tasks and milestones, and is made completely compatible with the overall program plan. Appendix A outlines the contents of a typical EMC program plan. Fig. 2-4 (Ref. 7) shows typical interfaces in an EMC program between various appropriate documents and functions.

2-6.2 CONTROL PLAN

2-6.2.1 General

The EMC control plan, sometimes referred to as EMI control plan or EMC development plan, is the key EMC technical document in the system program. It is prepared as early as possible after initiation of a project and is the source for all EMC technical information, and defines or refers to all of the requirements, directives, and control mechanisms, including organization, specification, design criteria, and test plans.

A control plan is required by MIL-STD-461, MIL-E-6051 and is listed in AR 70-27, *Research and Development Plan/Development Concept Paper/Program Memorandum*. The content for the EMI control plan is shown in Appendix B.

It is sometimes desirable to modify the completely integrated control plan approach by generating separate supplemental EMC documents such as specifications, design criteria, and various test plans, and referencing these documents in the master EMC Control plan. In this way it is possible to make changes to, and reissue, the supplemental documents without changing the entire control plan.

2-6.2.2 Design Instructions

In any materiel design program it is necessary to issue instructions to the design groups in the first stages of work. These instructions should take the form of a separate design instruction document or a preliminary control plan embodying mechanical, wiring, and circuit design instructions. The use of design instructions (in the form of specifications when available) will assure meeting all requirements and promoting uniformity of design. Early issuance is especially important, since early design decisions can commit the program to poor EMC practices. As design, analysis, production, and development testing progress, the design instructions must be modified, amplified, and reissued. Some of the EMC areas which should be covered by initial design instructions include:

a. Mechanical design: including choice of metals and hardware, corrosion control procedures, and types of construction

b. Electrical bonding and grounding: electrical interfaces with other equipment, subsystems, or other systems, electrical power returns, and conductor shields

c. Shielding: encompassing equipment and subsystem case shielding, shielding provision of the

TABLE 2-7. EMC AND RELATED SPECIFICATIONS

NUMBER	TITLE	SCOPE
MIL-STD-188	Military Communications System Technical Standards	Overall technical design standards for military communication systems. It is used in the development of new equipment and the procurement of production models of standard equipment. The objective is to enable the design, installation, and operation of military systems to be accomplished with a minimum of equipment interface problems of system/equipment incompatibility. The broad areas included are overall tactical system planning, tactical transmission systems, switching systems, and instruments, telegraph and data transmission, interface standards, and methods of measurement. The general requirement for EMC, specifically MIL-STD-461, is cited. Exceptions to MIL-STD-461 do exist in this standard.
MIL-STD-202	Test Methods for Electronics & Electrical Component Parts	Uniform methods for testing electronic and electrical component parts, including basic environmental tests to determine resistance to deleterious effects of natural elements and conditions surrounding military operations, and physical and electrical tests. This standard applies only to small parts weighing up to 300 lb or having an rms test voltage up to 50,000 V unless otherwise specified.
MIL-STD-220	Methods of Insertion Loss Measurement	A method of measuring, in a 50-ohm system, the insertion loss of single- and multiple-circuit radio-frequency filters at frequencies up to 1 GHz
MIL-STD-285	Attenuation Measurements for Enclosure, Electromagnetic Shielding, for Electronic Test Purposes, Method of	A method of measuring the attenuation characteristics of electromagnetic shielding enclosures used for electronic test purposes
MIL-STD-449	Radio Frequency Spectrum Characteristics, Measurement of	Uniform measurement techniques that are applicable to the determination of the spectral characteristics of transmitters, receivers, antennas, and system couplers
MIL-STD-454	General Requirements for Electronic Equipment	Components and construction details for the design and construction of electronic equipment for the Department of Defense. It includes in one document, under suitable subject headings, the fundamental design requirements of 13 general electronic specifications. It is updated biannually through the cooperative efforts of Government and industry. It references MIL-STD-461, 462, and 469.
MIL-STD-461	EMI Characteristics, Requirements for Equipment	Requirements and test limits for the measurement and determination of the emission and susceptibility characteristics of electronic, electrical, and electromechanical equipment and subsystems which
(cont'd)		

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
		are used independently or become part of other subsystems or systems. The requirements and test limits are applicable, to the extent specified, in the individual equipment or subsystem specification contract. Accordingly, when invoking this standard, the requirements and limits it contains must be analyzed to verify its suitability and applicability for the specific procurement. When it is known that the equipment or subsystem will encounter worst case operational conditions which are more or less severe than the levels contained in this standard, the individual specifications may modify the limits in this standard.
MIL-STD-462	EMI Characteristics, Measurement of	Measurement techniques to determine emission and susceptibility characteristics of electrical, electronic, and the electromechanical equipment, as required by MIL-STD-461. Setups for each test are given in block diagram form, along with details on conducting each measurement.
MIL-STD-463	Definitions and Systems of Units, EMI Technology	Abbreviations, multiplying symbols, frequency spectrum designations, and definitions of terms applicable to EMI along with their approved symbols.
MIL-STD-469	Radar Engineering Design Requirements, EMC	Engineering design requirements to control the spectral characteristics of all new radar systems operating between 100 MHz and 40,000 MHz in an effort to achieve EMC and to conserve the frequency spectrum available to military radar systems. The design requirements and criteria are not intended to inhibit the free and unrestricted approach to research related to the development of new radar systems which promise an increase in effectiveness. It is recognized that certain requirements stated in this document are not applicable to all types of radar systems. Where this is true, the intent of the requirements shall be applied with the best engineering judgment and approval by the procuring agency.
MIL-STD-633	Mobile Electric Power Engine Generator Standard Family Characteristics Data Sheets	Detailed information on the physical and electrical characteristics and logistical data on the DoD approved family of mobile electric power engine generator sets.
MIL-STD-704	Electric Power, Aircraft, Characteristics and Utilization of	Characteristics of electric power supplied to airborne equipment at the equipment terminals and the requirements for the utilization of such electric power by airborne equipment.
MIL-STD-831	Test Reports, Preparation of	Format and content criteria to be used in the preparation of test reports covering tests on systems, subsystems, equipments, components, and parts.

(cont'd)

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
MIL-STD-833	Minimization of Hazards of Electromagnetic Radiation to Electroexplosive Devices	Criteria to be applied to the design of electroexplosive devices (EED's) and their application in systems. The purpose of this standard is to minimize the hazards of electromagnetic radiation to EED's. Applies to the design, selection, and application of EED's and their firing circuits for all new development programs of systems that use EED's.
MIL-STD-1275	Electrical Circuit, 28 Volt DC Transient Characteristics for Military Vehicles	Limits of transient voltage characteristics and steady state limits of the 28 V dc electric power circuits of military vehicles.
MIL-STD-1310	Shipboard Bonding and Grounding Methods for Electromagnetic Compatibility	Shipboard construction and equipment installation requirements and the practices necessary to minimize the electromagnetic interference (EMI) environment aboard Naval ships.
MIL-STD-1337	General Suppression System Design Requirements for Portable Electric Hand Tools	Electromagnetic interference design requirements for portable electric hand tools (metal or insulated encasements or a combination thereof) having functional or double insulation.
MIL-STD-1377	Effectiveness of Cable, Connector, and Weapon Enclosure Shielding and Filters in Precluding Hazards of Electromagnetic Radiation to Ordnance; Measurement of	Provides a weapon developer or designer with shielding and filter effectiveness test methods for determing whether the particular weapon design requirements of MIL-P-24014 have been properly implemented. It is not intended to be a substitute for full-scale electromagnetic hazards evaluation tests of the weapon system, but rather an aid in developing a weapon system with a high probability of successfully passing such environmental tests.
MIL-STD-1385	Preclusion of Ordnance Hazards in Electromagnetic Fields; General Requirements for	General requirements to preclude hazards resulting from ordnance having EED's when exposed to electromagnetic fields. The nominal frequency range covered by this standard is from 10 kHz to 40 GHz.
MIL-STD-1512	Electroexplosive Subsystems, Electrically Initiated Design Requirements and Test Methods	Uniform design and qualification requirements and test methods for the design, development, and acceptance of all electroexplosive subsystems and components.
MIL-B-5423	Boots, Dust and Water Seal (for Toggle and Push-Button Switches and Rotary-Activated Parts), General	General requirements for molded silicone-rubber boots for use on toggle and push-button switches and rotary-actuated parts such as rotary switches, variable resistance, capacitors, inductors, and transformers. The boots protect the switch-actuating mechanism from sand, dust, water and other contaminants, and seal the panel on which the switches are mounted.
MIL-C-5 (cont'd)	Capacitors, Fixed, Mica-Dielectric, General Specification for	General requirements for molded, dipped, and potted mica dielectric, fixed capacitors intended primarily for use in high-stability, low-loss radio-

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
		frequency applications such as tuned circuits. This is a graded specification covering ranges in temperature coefficient, capacitance, tolerance temperature, and vibration.
MIL-C-7078	Cable, Electric, Aerospace Vehicle, General Specification for	Electric cable for use in aerospace vehicles and other applications for which its performance characteristics are suitable. The types of cables covered are unshielded-unjacketed, jacketed, shielded, and and shielded and jacketed.
MIL-C-11693	Capacitors, Feedthrough, Radio Interference, Reduction, AC & DC (Hermetically Sealed in Metallic Cases), General Specification for	General requirements for established reliability (ER) and non-ER capacitors designed for operation with alternating current (ac) and direct current (dc), paper, metallized paper, and metallized plastic dielectric radio-interference reduction, feedthrough capacitors, hermetically sealed in metal cases, for use primarily in broadband, radio-interference suppression application. Capacitors meeting the established reliability requirements specified herein have a maximum failure rate of 1%/1000 h. This failure rate is established with a 90% confidence limit based on the life test parameters specified and are maintained at a 10% producer's risk. An acceleration factor of 5:1 has been used to relate the life test data obtained at 140% of rated voltage (ac or dc) at the applicable high test temperature to the rated voltage at the applicable high test temperature. Styles CZ20, CZ25, CZ32, and CZ33 contained herein are of a metallized construction and should be used only in circuitry in which high values of insulation resistance are not essential, and in which occasional momentary breakdowns can be tolerated.
MIL-C-12889	Capacitors, Bypass, Radio Interference Reduction, Paper Dielectric, AC and DC (Hermetically Sealed in Metallic Cases), General Specifications for	Performance and general material requirements for bypass radio-interference reduction, alternating current (ac) and direct current (dc), paper-dielectric capacitors, hermetically sealed in metallic cases, for use primarily in broadband, radio-interference suppression application. In addition, this specification indicates the ambient test conditions within which the capacitors must operate satisfactorily and reliably. These capacitors are suitable for operation over a temperature range of −55° to +85°C.
MIL-C-13909	Conduit, Metal, Flexible; Electrical, Shielded	Shielded, electrical, flexible metal conduit. The conduit consists of a core of flexible metal tubing with a covering of wire braid for use in military applications.

(cont'd)

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
MIL-C-25200	Cable Assemblies, Special Weapons, Electrical, General Requirements for	Detailed requirements for the design, manufacture, and testing of Air Force special weapons cable assemblies, including interconnecting cables.
MIL-C-45662	Calibration of Standards	Establishment and maintenance of a calibration system to control the accuracy of the measuring and test equipment used to assure that supplies and services presented to the Government for acceptance are in conformance with prescribed technical requirements.
MIL-E-6051	System EMC Requirements	Overall EMC requirements for systems, including control of the system electromagnetic environment, lightning protection, static electricity, bonding, and grounding. It is applicable to complete systems, including all associated subsystems and equipments. Requirements for the overall integrated EMC program for the system. This includes the necessary approach, planning, technical criteria, and management controls, and is based on the requirements in this specification, other specifications, statement of work, and other applicable contract documents.
MIL-E-7080	Electric Equipment, Aircraft, Selection and Installation of	Requirements for the installation and selection of electric equipment in piloted aircraft. Electric equipment includes electric power generation and utilization equipment, and control and protective devices.
MIL-E-45782	Electrical Wiring, Procedures for	Fabrication and termination of shielded electrical harness and cable assemblies and the wiring of electrical and electronic circuits of subassemblies used in missile systems.
MIL-F-15733	Filters, Radio Interference, General Specifications for	General requirements for current-carrying filters, ac and dc, for use primarily in the reduction of radio noise.
MIL-F-18327	Filters; High Pass, Low Pass; Band Suppression, and Dual Functioning, General Specifications for	General requirements for passive frequency-selective networks over the frequency range of 0 to 50 MHz for use in electronic and communication equipment. It does not cover filters weighing more than 50 lb or requiring rms test voltage ratings greater than 5000 V. Also, it does not cover interference reduction, brute-force filters; pulse-forming or resistive-capacitance networks; IF and RF transformers and coils.
MIL-F-19207	Fuseholders, Extractor Pose Type, Blown Fuse, Indicating and Nonindicating, General Specifications for	Enclosed, panel mounted, extractor post type electrical fuseholders, both blown fuse indicating and nonindicating.

(cont'd)

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
MIL-I-6181	Interference Control Requirements, Aircraft Equipment	Design requirements, interference test procedures, and limits for electrical and electronic aeronautical equipment to be installed or in closely associated with aircraft. The types of tests are: a. Interference tests; measurement of conducted and radiated emissions b. Susceptibility tests.
MIL-R-9673	Radiation Limits, Microwave and X-radiation, Generated by Ground Electronic Equipment (as Related to Personnel Safety)	Requirements for the preparation and submission by a contractor of data describing and defining radio-frequency power density and X-ray characteristics for ground electronic systems, subsystems, equipments, components, and end items procured by USAF. Furnishes guidance regarding permissible levels of exposure to X ray and provides for submission of data.
MIL-W-6858	Welding, Resistance: Aluminum, Magnesium, Nonhardening Steels or Alloys, Nickel Alloys, Heatresisting Alloys, and Titanium Alloys: Spot and Seam	Requirements for resistance spot and seam welding of the following nonhardening materials: Group (a) - Aluminum, aluminum alloys, and magnesium alloys Group (b) - Steels, austenitic and ferritic and precipitation hardening steels, nickel and cobalt base alloys Group (c) - Titanium and titanium alloys.
MIL-HDBK-162	United States Radar Equipment	Technical and functional descriptions, installation considerations, and reference data for operational ground, airborne, and shipborne radar equipment used within the Department of Defense. Radar equipment described is limited to the "radar set" or "end item" level. The Joint Electronics Type Designation System (JETDS) was used as a guide to select the type of radar equipment. With a few exceptions, only "P" and "S" "Type of Equipment" indicator letter codes are included.
MS-25384	Plug, Fuel Nozzle, Grounding	Details the design of electrostatic discharger jumper, fuel nozzle-to-aircraft.
MS-33645	Receptacle Installation, Fuel Nozzle Jumper, Aircraft	Details the method of installing the grounding receptacle to be used with the electrostatic discharger jumper described in MS-25384.
SAE-ARP-936	Ten-μF Capacitor	Requirements of a special purpose 10-μF feedthrough capacitor to be used in series with the power line to an electrical or electronic device during EMI tests.
SAE-ARP-958	Measurement of Antenna Factors	Method and technique for the checkout and calibration of electromagnetic interference measurement antennas. This applies to conical logarithmic

(cont'd)

TABLE 2-7 (cont'd)

NUMBER	TITLE	SCOPE
		spiral antennas as described in USAF drawings: 62J4040 200 to 1000 MHz; 62J4041 1 to 10 GHz
SAE-J551	Measurement of Electromagnetic Radiation from Motor Vehicles (20 to 1000 MHz)	Test procedures for the measurement of electromagnetic radiation from motor vehicles and limits to insure that their operation does not seriously interfere with radio communications and other electronic equipment. This covers all emissions from all sources, except short duty cycle equipment in the frequency range 20-1000 MHz.
ANSI C1	National Electrical Code	Electric conductors and equipment installed within or on public and private buildings or other structures, industrial substations, mobile homes, recreational vehicles, and other premises; conductors that connect the installations to a supply of electricity; and other outside conductors on the premises.

system structure, shielding and twisting of conductors, anticipated electromagnetic and electrostatic environments

d. Circuit design for transient control: encompassing suppression of transients from inductive sources, suppression of contactor transients, and surge limiting within the system power profile.

e. Circuit design for radiated signal control: encompassing spurious signals from intentional radiation sources, unintentional radiation sources, control of response to radiated signals, projected intentional radiated environment created by the system including frequency allocation, antenna location, antenna patterns, and signal levels.

f. Interference susceptibility prediction methods

g. Cable and/or conductor routing configuration, which should encompass conductor separation and isolation, and location of equipments and subsystems

h. Any special system EMC considerations which bear upon design.

2-6.3 TEST PLANS

Test plans may include facilities and instrumentation required, description of test samples, test procedures, test reports, and enumeration of tests such as equipment, subsystem, development, production, and operational tests. Test plans may be issued as supplements to the EMC Control Plan or as part of the plan itself.

Separation into supplements is often desirable, since many of the groups having primary interest in testing — i.e., quality assurance, system test, production test — are not associated organizationally with design engineering, and may have no need for the overall program documents. In addition, the final test plan usually cannot be defined until design and analysis have progressed to the point where firm equipment designs are called out to allow exact tests to be specified.

In drawing up plans for tests it should be recognized that they can be conducted at three levels of system development as follows:

a. Subsystem level. Tests are required to demonstrate that subsystems (power supplies, transmitters, receivers, elect. generators, vehicles, etc.) meet criteria given in MIL-STD-461.

b. System level. Tests are required in accordance with MIL-E-6051 to demonstrate that the system will operate without EMI.

c. Intersystem level. Tests are required to demonstrate that the system will operate with other military and civilian systems without EMI during training and battlefield operations. General requirements are called out in AR 11-13, and detailed requirements may be obtained from the system project manager, from the US Army Training and Doctrine Command

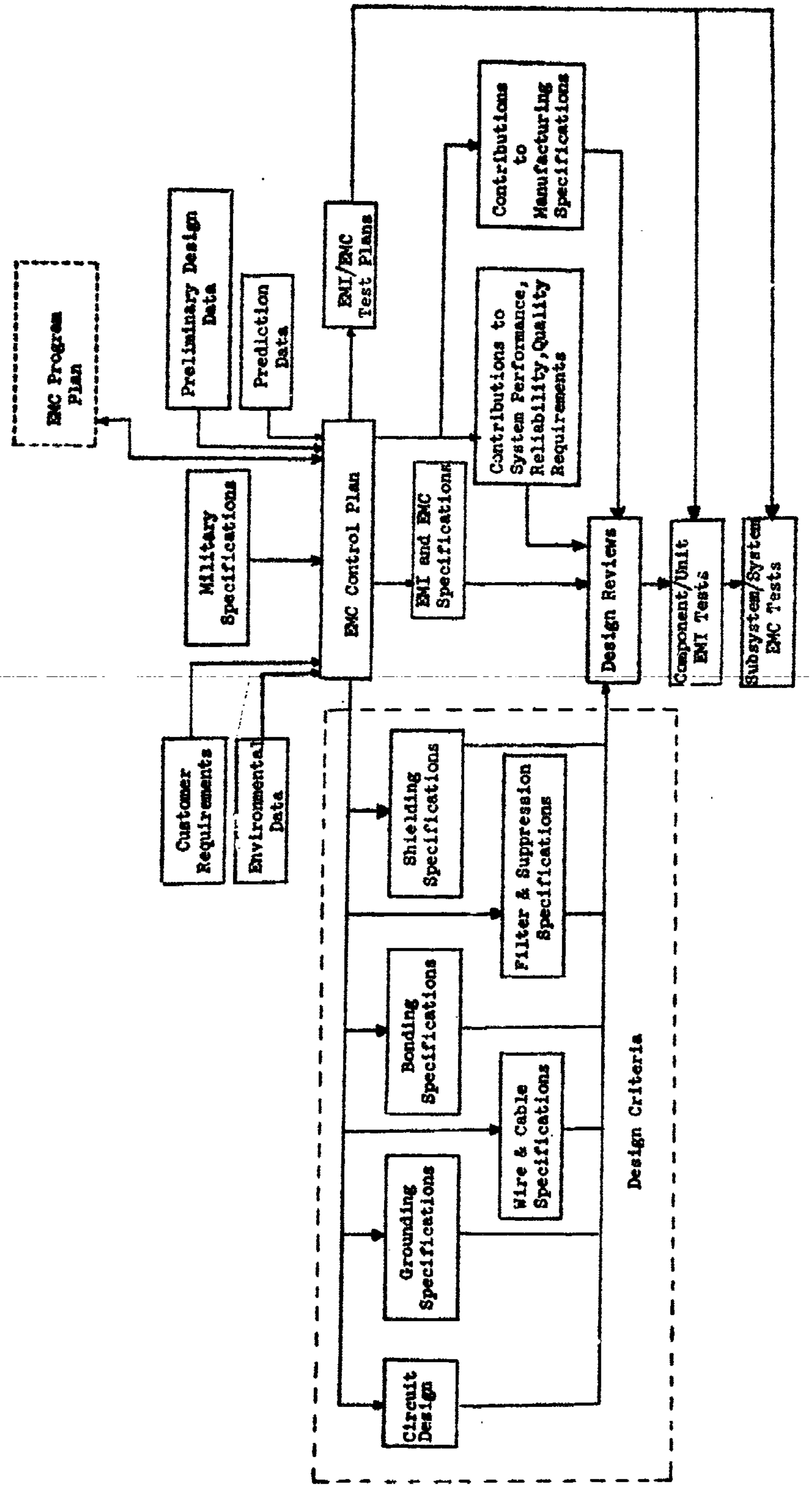


Figure 2-4. Typical EMC Functional Interfaces

(TRADOC), or from other Department of the Army agencies.

The content of an EMI test plan is shown in Appendix C. For further details, see Ref. 6 and AR 70-10.

2-6.4 FREQUENCY ALLOCATIONS

One of the most important program life cycle activities of communication-electronic materiel development programs is the requesting and obtaining of frequency allocations and assignments.

2-6.4.1 Initiating the RF Allocation Request

For C-E equipments "Application for RF Allocation" form DD 1494 must be submitted at specific life cycle management model (LCSMM) phases (see par. 2-7) as follows:

Phase	Action Required
Conceptual	Apply for experimental RF spectrum allocation
Validation	Apply for developmental RF spectrum allocation
Full-scale development	Apply for frequency assignments

(See AR 15-14, AR 71-1 for details.)

2-6.4.2 Request Processing

In the Department of the Army (DA) the requests follow the flow shown in Fig. 2-5. Form DD 1494 goes via appropriate command channels to the Deputy Chief of Staff for Operations and Plans (DCSOPS) who either processes the request at that level or passes it on to the Spectrum Planning Subcommittee (SPS) of the Interdepartment Radio Advisory Committee (IRAC) for action. (See AR 105-16 for details.)

The Federal Communications Commission (FCC) frequency planning is shown on the right side of Fig. 2-5. FCC line responsibility is to Congress, while Department of Defense (DoD) is to the President, and liaison between DoD spectrum management and the non-Government spectrum management (FCC) is via FCC representation of IRAC.

2-6.4.3 DCSOPS Partial Function Summary

In accordance with AR 10-5, the Deputy Chief of Staff for Operations and Plans (DCSOPS) has Army General Staff responsibility for radio frequency management throughout the Department of the Army, as follows:

a. Army policy and procedures for processing RF allocation requests

b. Advising on frequency matters during all stages of Army materiel development

c. Department of the Army staff management of the use of the electromagnetic spectrum

d. Coordination and processing of Army RF allocations with the Joint Frequency Panel of the Military Communications-Electronics Board (MCEB)

e. Management of the Army Electromagnetic Compatibility Program.

Further details may be found in AR 105-16.

2-6.4.4 SPS Partial Function Summary

The SPS is responsible for carrying out those functions that relate to planning for the use of the electromagnetic spectrum in the national interest, including the apportionment of spectrum space for the support of established or anticipated radio services, as well as the apportionment of spectrum space between or among Government and non-Government activities.

The SPS consists of a representative appointed by each of the following member departments and agencies:

Agriculture
Air Force
Army
Atomic Energy Commission
Coast Guard
Commerce
Federal Aviation Admin.
General Services Admin.
Health, Education & Welfare
Interior
Justice
National Aeronautics & Space Admin.
Navy
State
Treasury
United States Information Service

Liaison between the SPS and the FCC is effected by a representative appointed by the Commission to serve in that capacity.

2-6.4.5 IRAC Function Summary

IRAC reports to the Assistant Director for Frequency Management, OTP. It consists of a representative appointed by each of the departments and agencies listed in par. 2-6.4.4, except Commerce. Its mission is to assist the Director of OTP in the discharge of his responsibilities pertaining to the use of the electromagnetic spectrum, especially in assigning frequencies to US Government radio stations and in

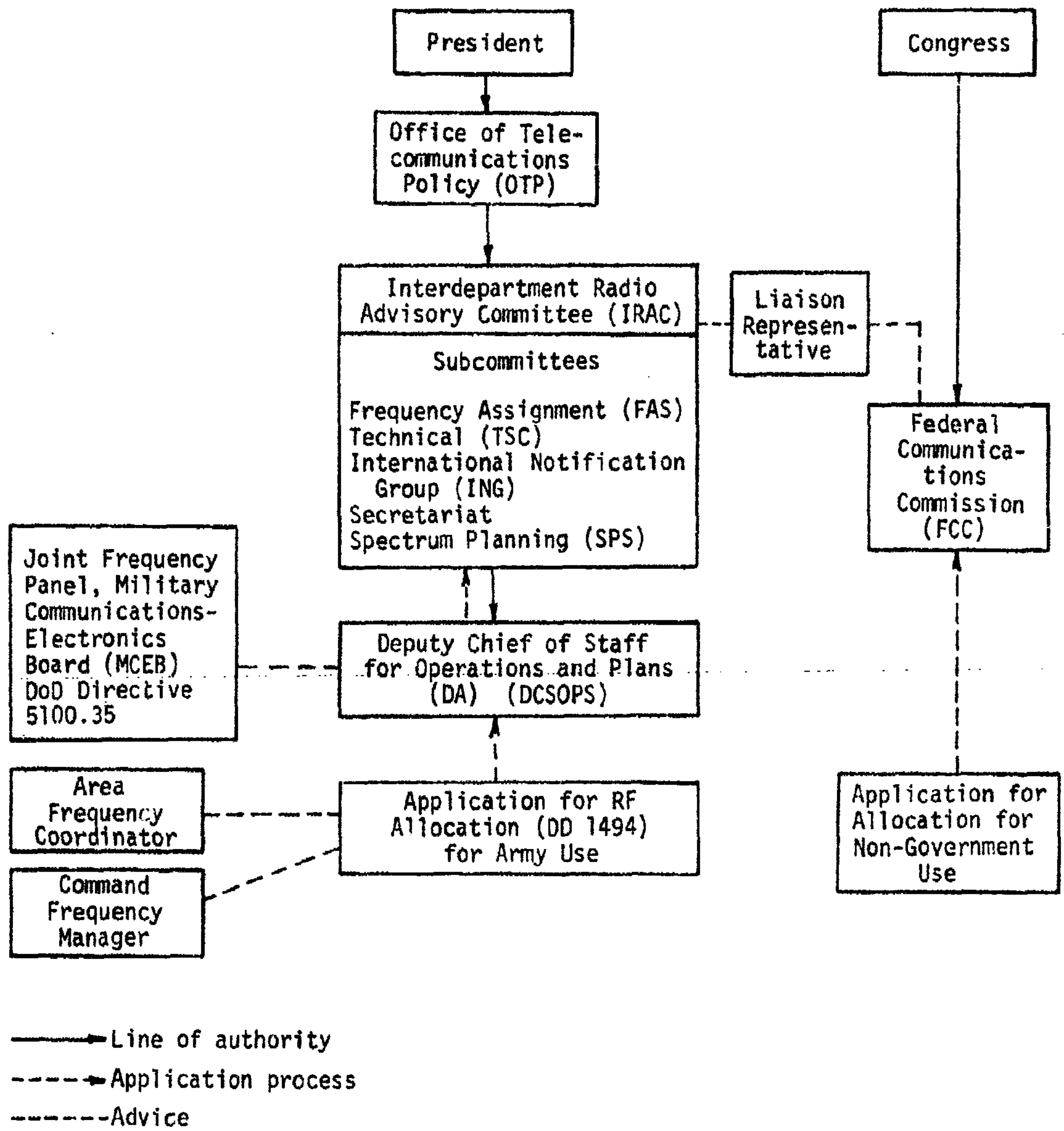


Figure 2-5. Authority and Processing of RF Spectrum Applications

developing and executing policies, programs, procedures, and technical criteria pertaining to the allocation, management, and use of the spectrum. In addition to the SPS, the IRAC's substructure consists of the Frequency Assignment Subcommittee (FAS), the Technical Subcommittee (TSC), the International Notification Group (ING), and the Secretariat.

2-6.4.6 OTP Function Summary

Subject to the authority and control of the President, the Director of OTP develops and sets forth plans, policies, and programs with respect to telecommunications that promote the public interest, support national security, sustain and contribute to the full development of the economy and world trade, strengthen the position and serve the best interests of the U.S. in negotiations with foreign nations, and promote effective and innovative use of telecommunication technology, resources, and services. Agencies consult with the Director of OTP to ensure that their conduct of telecommunication activities is consistent with the Director's policies and standards.

2-6.5 INTERFERENCE PREDICTION

An integral part of materiel system planning is appropriate scheduling of interference prediction. Serious interference possibilities must be foreseen and predicted as early as possible in a system life cycle. The EMC Control Plan defines and schedules the prediction work, and funds must be budgeted for analysis and prediction.

2-6.5.1 Analysis Procedures

Interference possibilities are predicted and described by modeling emitters and susceptors, identifying the coupling modes, and calculating the transmission. Although the process is not difficult for a single source and receiver, it becomes geometrically more difficult as sources and receivers increase in number. Typical procedures may include:

a. Theoretical analysis:

(1) Model the system (see Chapters 3, 4, and 5 for emitter, susceptor, and coupling data).

(2) Analyze the system using:

(a) Rule of thumb or comparison with other systems

(b) Hand calculation

(c) Computer analysis (see Chapter 6).

b. Simulate all or parts of the system.

c. Measure each piece of the system, and do a semitheoretical analysis based on measured results.

d. Observe the actual system during operation.

e. Use combination of two or more of the preceding techniques.

f. Recommend corrections (see Chapter 4).

2-6.5.2 Applications of Interference Prediction

Each phase of a system life cycle requires appropriate analysis and prediction. Concept formulation activity must not only predict problems to be encountered in later phases, but must use analytical techniques to effect a judicious balancing and modification of system design, operating environment, and system requirements. Succeeding phases must include analysis leading to prediction of the expected EMI status of the next phase as well as the identification of correction measures, if required.

Testing and analysis in the later phases of the program should verify and expand the prior analysis and corrections. If an unforeseen interference occurs once a system has been deployed, sufficient analytical and test data should be available from the development program to pinpoint quickly the source of trouble and allow remedies to be applied offering the best trade-off between cost and effectiveness.

The application and frequency of use of prediction techniques, and the resulting decisions regarding corrections, are important program management functions directly involving system cost effectiveness.

Programs which do not fit the system life cycle management model, such as research programs or beyond the state-of-the-art developments, also benefit from the use of EMI prediction techniques. The impact of the EMI environment on such an advanced program, the impact of the program on potential susceptors, as well as the normal internal program EMI problems should be analyzed and described. Advanced programs that have been appropriately analyzed, and for which prediction, correction, and test data are available — when converted to operational hardware through the system life cycle — tend to maintain a high system effectiveness and experience few expensive EMI fixes.

2-6.6 COST-EFFECTIVENESS CONSIDERATIONS

Cost-effectiveness considerations must be part of every program decision at every phase in the system cycle. In order to maintain a proper balance between the most elegant design and the greatest ease of fabrication, between meeting a high standard of performance on the one hand and maintaining a frugal and easily fabricated design on the other, it is necessary to seek a reasonable trade-off between program cost and program performance. The third variable of the program triumvirate, time, customarily is converted to dollars and included in the cost element. Fig. 2-6 shows typical relationships in a cost-effectiveness model.

EMC control decisions are particularly susceptible to the influence of cost-effectiveness trade-offs. Analysis, tests, and corrections account for considerable program expenses, and every effort must be expended to apply the exact amount of expense needed to achieve an appropriate measure of EMC.

Typical elements of the use of cost effectiveness include:

a. Setting the "tightness" of requirements beyond those which are mandatory in designated specifications and standards, par. 2-5.4.

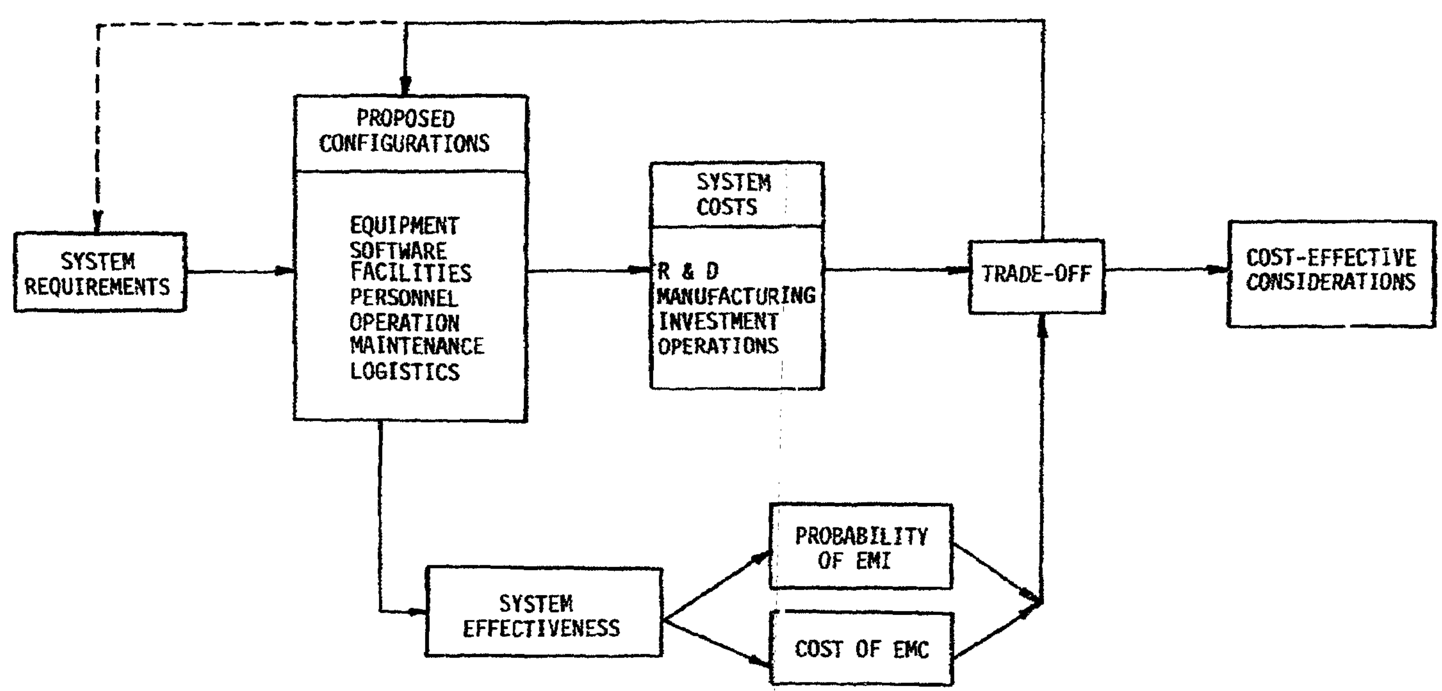


Figure 2-6. Cost-Effectiveness Model for Selecting a Configuration

b. Establishing the size of the EMC program within the system program

c. Selecting EMC design criteria

d. Selecting the most appropriate prediction/analysis method

e. Selecting the best solution (design, component, manufacturing, installation) to an EMI problem.

To establish the risk of EMI failure in terms of dollars, it is necessary to do a risk analysis which will establish potential EMI failure modes, cost penalty of the failure, and the probability of failure. Once the analysis is complete and an appropriate matrix prepared, the cost penalty multiplied by the probability of failure will give the dollar risk for the potential failure.

The task of establishing EMC program costs is not difficult when costs are limited to such readily measurable items as engineering labor, technician labor, consultants, test time, equipment, facility costs, and overhead. However, costs must also be estimated and included for such induced expenses as:

a. Labor applied by engineering, test, and quality assurance personnel, other than EMC personnel, to EMC activities.

b. Impact of EMC plans and reports on other program costs

c. Design review time devoted to EMC

d. Impact of EMC tests on other program costs.

Mathematically, the trade-off point exists when the incremental risk cost equals the incremental EMC program cost. However, in the face of the many unpredictable program elements, the trade-off should be taken only as a guide, and the final decision regarding the magnitude of the EMC program should be made by the Program Manager.

2-7 EMC PROGRAM IMPLEMENTATION

Major and nonmajor system life cycles require that actions be carried out, decisions made, documentation prepared, and program milestones met. EMC activities and decisions are important life cycle functions. Specific EMC decisions to be made during the LCSMM phases must be identified, described, and scheduled to meet the over-all program requirements. EMC decision points may be selected and defined by systematically analyzing the program as shown in Fig. 2-7. A detailed presentation of the EMC Decision process is given in Ref. 8.

2-7.1 EMC DECISIONS DURING SYSTEM LIFE CYCLE (REF. 8)

The discussion of life cycle actions and EMC decisions which follows takes up each LCSMM phase in turn as shown in Figs. 2-8 and 2-9. The distinction between major and nonmajor systems is made where an important difference exists.

2-7.1.1 Conceptual Phase

The conceptual phase of the LCSMM begins with approval of a required operational capability by Headquarters, Department of the Army (AR 15-14, AR 71-1).

For all tactical comand, control, and communication equipments, DoD Directive 4630.5 requires that DA coordinate the requirement with the other services and provide a record of this coordination to the Joint Chiefs of Staff (JCS).

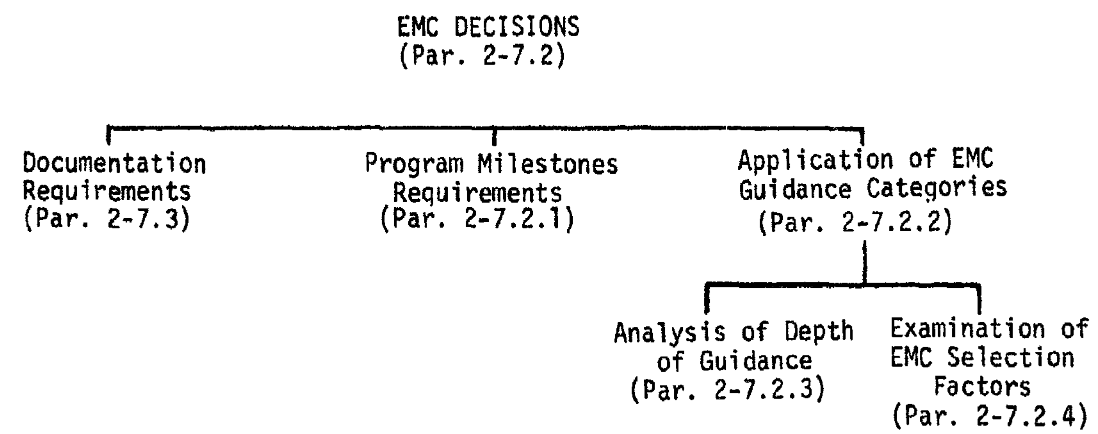
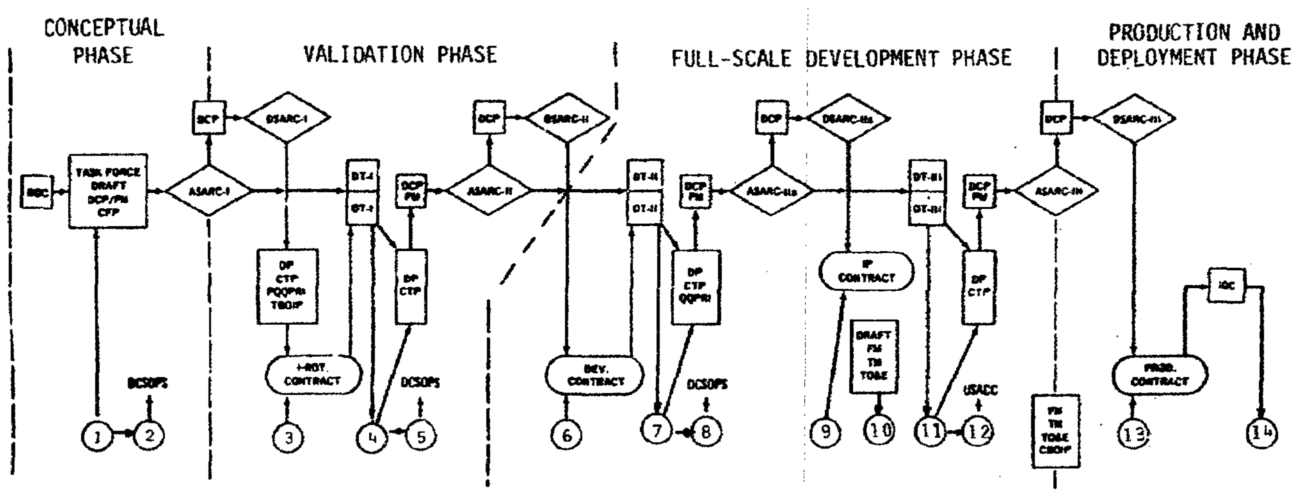


Figure 2-7. EMC Decision Process



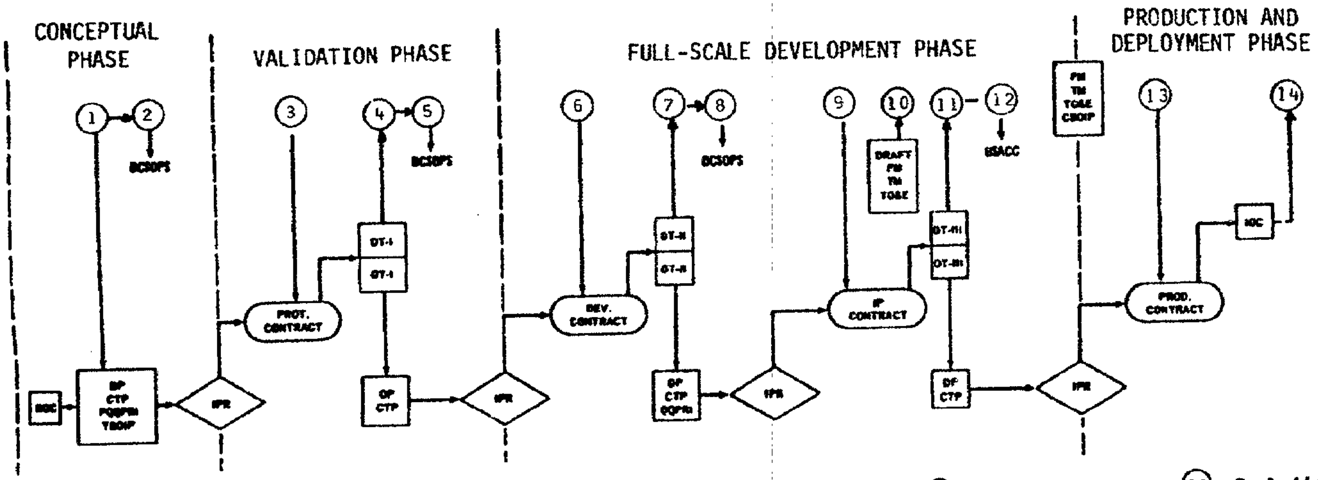
1. Preliminary equipment characteristics and FS
2. Experimental RF allocation
3. Prototype specifications
4. Verify prototype performance
5. Developmental RF allocation
6. Development specifications
7. Verify development equipment performance
8. Operational RF allocation
9. IP specifications
10. Approve training material and TOE
11. Verify IP performance
12. Operational frequency assignment
13. Production specifications
14. Verify production equipment performance

Abbreviations:

ASARC - Army Systems Acquisition Review Council
CBOIP - Complete basis of issue plan
CFP - concept formulation package
CTP - coordinated test program
DCP - development concept paper
DP - development plan
DCSOPS - Deputy Chief of Staff, Operations and Plans
DSARC - Defense Systems Acquisition Review Council
DT - Development test
FM - field manual
IOC - initial operational capability
IP - initial production
OT - operational test
PM - program memorandum
PQQPRI - preliminary qualitative and quantitative personnel requirements information
QQPRI - qualitative & quantitative personnel requirements information
RF - radio frequency
ROC - required operational capability
TBOIP - tentative basis of issue plan
TM - technical manual
TO&E - table of organization & equipment
USACC - U S Army Communications Command

○ Actions and Decisions, EMC
□ Principal Documents
◇ Decision Review Milestones

Figure 2-8. Flow Chart — EMC Decisions/Actions in the LCSMM for Major Systems



(1) Preliminary equipment characteristics and FS

(2) Experimental RF allocation

(3) Prototype specifications

(4) Verify prototype performance

(5) Developmental RF allocation

(6) Development specifications

(7) Verify development equipment performance

(8) Operational RF allocation

(9) IP specifications

(10) Approve training material and TOE

(11) Verify IP performance

(12) Operational frequency assignment

(13) Production specifications

(14) Verify production equipment performance

Abbreviations:

- ASARC - Army Systems Acquisition Review Council
- CBOIP - Complete basis of issue plan
- CFP - concept formulation package
- CTP - coordinated test program
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- DP - development plan
- DCSOPS - Deputy Chief of Staff, Operations and Plans
- DSARC - Defense Systems Acquisition Review Council
- DT - Development test
- FM - field manual
- IOC - initial operational capability
- IP - initial production
- OT - operational test
- PM - program memorandum
- PQQPRI - preliminary qualitative and quantitative personnel requirements information
- QQPRI - qualitative & quantitative personnel requirements information
- RF - radio frequency
- ROC - required operational capability
- TBOIP - tentative basis of issue plan
- TM - technical manual
- TO&E - table of organization & equipment
- USACC - US Army Communications Command

○ Actions and Decisions, EMC

□ Decision Review Milestones

◇ Principal Documents

Figure 2-9. Flow Chart — EMC Decisions/Actions in the LCSMM for Nonmajor Systems

In the case of major programs, and if the program requires Office of Secretary of Defense (OSD) approval, the principal action required during the conceptual phase is the preparation of a Development Concept Paper (DCP); otherwise a Program Memorandum (PM) (AR 15-14) is required. The DCP or PM, prepared by a special task force convened at the direction of the Chief of Staff, must define mission profiles and bands of performance and identify critical issues associated with EMC. The DCP or PM is used as the basis for the first Army Systems Acquisition Review Council review (ASARC-I) (and the first Defense Systems Acquisition Review Council review (DSARC-I), if required), which terminates the conceptual phase of the LCSMM.

For nonmajor systems the principal action required is preparation of the Development Plan (DP) by the materiel developer to include system and subsystem characteristics and a specific paragraph on EMC which requires quantitative statements of EMC requirements and plans for achieving these requirements (AR 70-27). This action occurs later in the LCSMM for major systems. The conceptual phase for nonmajor systems terminates with the feasibility in process review (IPR).

The major milestones in the conceptual phase for C-E equipments which are designed to use the electromagnetic spectrum are preliminary selection of the frequency band, including frequency supportability considerations; type of modulation, channelization, and other principal characteristics of the system; and submission of an application for an experimental RF spectrum allocation. These actions are required for evaluation of EMC feasibility for the IPR, ASARC, or DSARC review and to satisfy DoD Directive 4630.5 on compatibility and commonality of equipment.

For non-C-E systems and C-E systems which are not designed to utilize the radio frequency spectrum, a determination of system technical characteristics is required to a level of detail adequate to permit evaluation of potential unintentional receptor susceptibility or unintentional radiation interference.

Thus, in the conceptual phase, there are two critical EMC decisions:

*1. Select preliminary equipment characteristics and ascertain frequency supportability.

2. Apply for experimental RF spectrum allocation (see par. 2-1.1).

2-7.1.2 Validation Phase

The validation phase begins for major programs after ASARC-I approval to begin prototype development — for nonmajor systems after the feasibility IPR.

Principal activities during this phase include completion of the development plan (DP) (if major system), prototype development, first development test (DT-I), the initial operational test (OT-I), and the validation IPR (nonmajor systems) or ASARC-II (major systems).

Major milestones include award of the prototype development contract; preparation, coordination, and approval of test plans for DT-I and OT-I; review of test reports; and ASARC/DSARC review (IPR for nonmajor systems) of the prototype development program to determine whether to proceed with full-scale engineering development.

EMC considerations or actions during the validation phase include preparing the EMC-related portion of the equipment performance specifications for the prototype equipments, developing plans for the EMC portion of development and operational tests, reviewing results of EMC testing as part of DT-I and OT-I, and verifying that potential EMC problems have been averted or can be expected to be resolved during engineering development. Prior to ASARC-II/DSARC-II, the application for a developmental frequency allocation must be made. Thus, three critical EMC decisions are required in the validation phase:

3. Determine prototype equipment specifications

4. Verify prototype equipment performance.

5. Apply for developmental RF spectrum allocation (see par. 2-1.1).

2-7.1.3 Full-scale Development Phase

The full-scale development phase begins after an ASARC-II/DSARC-II or validation IPR decision to enter into full-scale development. Principal activities in this phase are the engineering development of the system or equipment, development testing (DT-II), operational testing (OT-II) and, for major systems/programs, ASARC-IIa/DSARC-IIa; for nonmajor systems, development acceptance IPR, low-rate initial production (IP), development testing DT-III, operational testing (OT-III), and finally production approval by ASARC-III/DSARC-III or production validation IPR.

Major milestones include award of the engineering development contract, approval of test plans for DT-II and OT-II, review of test reports for DT-II and OT-II, approval for initial production by ASARC-IIa/DSARC-IIa or development acceptance IPR, award of IP contract, approval of test plans for DT-III and OT-III, review of DT-III and OT-III test reports, and approval of full-scale production

*Numbers correspond to EMC Decisions, Tables 2-7 and 2-8.

ASARC-III/DSARC-III or production validation IPR.

EMC considerations and actions include preparing EMC portions of equipment development specifications, preparing EMC portions of test plans for DT-II and OT-II, reviewing EMC test results, verifying that EMC performance of developmental equipment is satisfactory, preparing EMC portions of equipment specifications for initial production, preparing EMC test requirements for DT-III and OT-III, reviewing DT-III and OT-III test results, and verifying that the system or equipment is ready for production from an EMC viewpoint.

During the initial production (IP) phase of the program, applications must be made for operational RF spectrum allocations based on previously ascertained FS data. Prior to DSARC-III, which authorizes full-scale production, DA approval of organization-equipment authorization document and training material — including table of organization and equipment (TOE), technical manuals (TM), and field manuals (FM) — is required in order to provide time for publication and use prior to introduction of the equipment to the field.

Application for frequency assignments, also keyed to current inputs on FS feasibility, should be made immediately after ASARC-III/DSARC-III or the production validation IPR.

Thus, there are seven critical EMC decisions/actions in the full-scale development phase of the LCSMM:

*6. Determine developmental equipment specifications.

7. Verify developmental equipment performance.

8. Apply for operational RF spectrum allocations.

9. Determine IP equipment specifications.

10. Approve TOE and training material.

11. Verify IP equipment performance.

12. Apply for frequency assignments.

2-7.1.4 Production and Deployment Phase

The production and deployment phase of the LCSMM begins after ASARC-III/DSARC-III or production validation IPR decision to enter into full-scale production.

Principal activities in this phase include the production of the system or equipment; publication of field manuals, technical manuals, tables of organization and equipment; resident training; introduction of the system or equipment into the field; unit training; operational use of the system or equipment; and finally disposal.

Major milestones include award of the production contract; publication of FM, TM, and TOE; achieving initial operational capability (IOC); identifying a requirement for a new or revised system of equipment; and finally replacement or disposal.

EMC considerations and actions include preparing EM portions of the production equipment specifications and verifying performance by review of operational performance reports and reports of interference or electronic warfare (meaconing, intrusion, jamming, interference: MIJI).

The following are the critical EMC decisions in the production and deployment phase of the LCSMM:

13. Determine production equipment specifications.

14. Verify operational equipment performance.

2-7.1.5 Summary

Of the fourteen EMC decisions/actions identified as critical, it is clear that all of them apply to both C-E and non-C-E equipment with the exception of those concerned with radio frequency spectrum allocation and frequency assignment.

Timing of the decisions has not been fixed precisely; in fact, the timing will vary and must be established as a part of the development plan. Thus, in the conceptual phase, selection of preliminary equipment characteristics will be made on the basis of performance requirements stated in the required operational requirements and must be made prior to the determination of feasibility in ASARC-I/DSARC-I or the feasibility IPR. EM decisions on equipment specifications must be made prior to award of prototype, development, initial production, or production contract bid solicitation. Obtaining FS supporting data for experimental, developmental, and operational RF spectrum allocations must be accomplished early enough to permit timely applications, planning for tests, and to be considered as a factor in equipment parameter selection. Verification of equipment performance in the validation, full-scale development, and production and deployment phases of the LCSMM depends upon DT and OT test results and must be accomplished prior to ASARC or IPR action.

2-7.2 GUIDANCE CATEGORIES

EMC decisions during the LCSMM are determined by analysis of program milestones and program document requirements, as well as by an evaluation of a group of EMC guidance categories

*Numbers correspond to EMC Decisions, Tables 2-7 and 2-8.

which serve as a standard grouping of EMC information (see Fig. 2-7). The information in the guidance categories is based upon EMC selection factors which make use of comparisons of similar programs and upon evaluations of depth of guidance required.

2-7.2.1 Program Milestones

A typical relationship between EMC decisions, LCSMM phase, and LCSMM program milestones is shown in Table 2-8. It is apparent that EMC

TABLE 2-8. EMC DECISIONS IN RELATION TO LCSMM MILESTONES

LCMM PHASE	LCMM MILESTONES	EMC DECISIONS
Conceptual	Future Requirements Approval	1. Select preliminary equipment characteristics and ascertain frequency supportability.
	ASARC-I, Feasibility IPR	2. *Apply for experimental RF spectrum allocation.
Validation	Prototype Contract DT-I OT-I	3. Determine prototype equipment.
		4. Verify prototype equipment performance.
		5. *Apply for developmental RF spectrum allocation.
	ASARC-II, Validation IPR	
Full-scale Development		6. Determine developmental equipment specifications.
	Development Contract DT-II OT-II	
		7. Verify development equipment performance.
	ASARC-IIa, Development Acceptance IPR	
		8. *Apply for operational RF spectrum allocation.
		9. Determine IP equipment specifications.
	IP Contract	
		10. Approve TOE and training material.
	DT-III OT-III	
		11. Verify IP equipment performance.
	ASARC-III Production Validation IPR	
		12. *Apply for frequency assignments.
Production and Deployment		13. Determine production specifications.
	Production Contract Publication of FM, TM, TOE Initial Operational Capability	
		14. Verify operational equipment performance.
	Requirement for new/modified materiel Disposal	

*Not required for non-C-E equipment; frequency assignment must be obtained for all equipment which requires an RF allocation before any radiation is permitted.

decisions must be made in the proper time sequence to support LCSMM activities primarily associated with EMC. It is not so apparent that other system decisions, such as system power levels, are not primarily EMC oriented but may be influenced heavily by EMC decisions and therefore must be considered in the decision process.

2-7.2.2 EMC Guidance Categories

Fourteen categories of EMC information have been identified to assist in EMC program decisions. These guidance categories are:

1. *C-E System Feasibility and Performance Requirements.* Supports preliminary evaluations and analyses of the ability of candidate C-E equipment and system concepts to achieve the system performance specified by requirements, such as required operation characteristics, within given technical, mission, and environmental constraints.
2. *Command and Organizational Principles.* Assessments of the impact of EMC factors on the development of organization and command principles of a combat force.
3. *System Operational Factors.* Equipment and system EMC performance data to assist in evaluating system operational performance.
4. *Economic Assessment.* Cost comparisons, life cycle costing analyses, cost-benefit analyses, and similar economic assessments in connection with equipment or system development.
5. *Electromagnetic Environment Evaluation.* Available frequency spectrum resources and RF power flux densities and/or EM field distributions as functions of frequency, time, and space coordinates.
6. *Natural Environment Evaluation.* Terrain, meteorological, and other physical factors which can affect the performance of C-E equipment.
7. *Hazard Evaluation.* Potential hazards to electrical equipment, munitions, electroexplosive devises, personnel, and ecology due to specified electromagnetic emissions.
8. *Equipment and Performance Characteristics.* Circuit, equipment, and subsystem performance parameters.
9. *Conformance to or Waivers of EMC Standards or Specifications.* Determination that equipments or equipment designs conform or do not conform to the EMC portions of military standards or specifications; and determination of whether to request waivers of specific equipment requirements.
10. *Spectrum Signatures.* Time and frequency-dependent amplitude response and phase characteristics of an electromagnetic emitter.
11. *Measures of System Effectiveness.* Determination of how well a specified combination of equipments, links, or networks provides a system function for C-E support of deployed forces.
12. *Site Survey and Selection.* Electrical, physical, and topographic characteristics of one or more sites.
13. *EMC Training Data.* Technical data, designs of training equipment, analysis techniques, computational aids, and curricula to ensure properly balanced emphasis on EMC in all formal training courses on concept, doctrine, operation, and maintenance of C-E equipment subsystems, and systems, including preparation and evaluation of MIJI reports.
14. *MIJI Report Analysis.* This category of EMC guidance information consists of evaluations of the sources and types of reported interference.

Table 2-9 shows EMC guidance categories being used to select and define EMC decision points in a typical program.

2-7.2.3 Depth of Guidance

Preparation of material for each EMC guidance category entails gathering information, and the effort needed in each case depends on a number of circumstances and requirements. The depth of EMC guidance required in the various guidance categories, in terms of needed effort and resulting information, can be characterized by:

a. Total time and resources expended.

b. Complexity of required imput parameters and associated analytic techniques employed.

c. Level of detail of EMC information obtained.

d. Resolution and accuracy of EMC information obtained.

The following circumstances must be considered:

a. State of finalization of C-E system parameters and of input data

b. Detail, resolution, and accuracy demanded by the documentation, decision, or action that the EMC guidance information is intended to support

c. Availability of prior substantiating or related EMC guidance information

d. Confidence in the available analytic techniques, in terms of the intended application. For example, the accuracy with which the characteristics of the EM environment are determined has a different payoff for a hazard investigation than for a computer-simulated EMC study.

e. Ready availability of inexpensive analytic techniques. For example, a developer may use a computer-aided EMC analysis methodology which provides much greater accuracy and detail than he

TABLE 2-9. EMC GUIDANCE CATEGORIES THAT MAY BE REQUIRED FOR EMC DECISIONS/ACTIONS

EMC DECISIONS/ACTIONS WITHIN LCSMM PHASES	EMC GUIDANCE CATEGORIES													
	1	2	3	4	5	6	7	8	9	10	11	12	13	14
Conceptual phase:														
1. Select preliminary equipment characteristics and ascertain frequency supportability	X	X	X	X	X	X		X			X			
2. Apply for experimental RF spectrum allocation	X		X		X	X		X		X		X		
Validation phase:														
3. Determine prototype equipment/specification			X	X			X	X	X		X			
4. Verify prototype equipment performance					X	X	X	X	X	X	X			
5. Apply for developmental RF spectrum allocation	X		X		X	X		X		X		X		
Full scale development phase:														
6. Determine developmental equipment specifications				X			X	X	X		X			
7. Verify development equipment performance					X	X	X	X	X	X	X			
8. Apply for operational RF spectrum allocation					X	X		X		X		X		
9. Determine IP equipment specifications							X	X	X		X			
10. Approve TOE and training material													X	
11. Verify IP equipment performance					X	X	X	X	X	X	X	X		
12. Apply for frequency assignments								X		X		X		X
Production and deployment phase:														
13. Determine production specifications							X	X	X		X			
14. Verify operational equipment performance					X	X	X	X	X	X	X	X		X

requires, when the model is readily available and costs less than simpler analytic techniques.

Within the preceding rationale, the depths of guidance will, in general, increase with succeeding LCSMM phases. The three depths of guidance are somewhat arbitrarily defined as follows:

a. Basic. Guidance is needed to select tentative C-E equipment characteristics. Application of planning factors and rules of thumb is sufficient. The EMC guidance information provided to the developer is expressed in broad, tentative, and general terms.

b. Detailed. Guidance is needed for firmer decisions on C-E equipment characteristics. Prototypes or development models are available. Theoretical data are validated with the use of computer-aided techniques or measurements on simplified development models.

c. Very Detailed. Guidance is needed to aid in making final decisions concerning equipment parameters, BOI, and operational use. Precise analytic techniques are required including comprehensive and detailed error and sensitivity analyses.

2-7.2.4 EMC Selection Factors

EMC selection factors are used to influence the selection of EMC guidance categories and the depth of detail of the information required within a guidance category. They are judgment aids rather than precise rules for determining the required EMC guidance information. The four EMC selection factors are:

A. Functions of the C-E portion of a system (func)

B. System type

C. Basis of issue (BOI), site selection, and deployment

D. Evolutionary versus technologically new development (evol/new).

Each factor is described in the paragraphs that follow.

2-7.2.4.1 Factor A, Functions of the C-E Portion of a System (func)

This EMC selection factor is concerned with the overall functions of the system under development, and particularly the functions of its C-E portions. The following fourteen C-E system functions are established, with abbreviations as indicated:

1. Surface-to-surface weapon system (SSWS)
2. Surface-to-air weapon system (SAWS)
3. Air-to-surface weapon system (ASWS)
4. Air-to-air weapon system (AAWS)

5. Counter mortar-counterbattery system (CM/CB)
6. Communication system (comm)
7. Electronic warfare system (EW)
8. Tactical data system (TDS)
9. Navigation and IFF system (nav/IFF)
10. Simple combat surveillance and target acquisition system (simple CSTA)
11. Variable time fuzes (VT fuze)
12. Complex combat surveillance and target acquisition system (complex CSTA)
13. Unintended C-E or non-C-E receptor unit (unint rec)
14. Unintended C-E or non-C-E emitter (unint emit).

The application of Factor A (func) is primarily to establish the over-all general pattern of EMC support requirements, rather than for specifying the detailed data to be obtained in each EMC guidance category. The general pattern of EMC support requirements for a development that involves one of the first 12 system functions should be in line with that of previous experience with developments that involve this function. Prior Army developments of C-E equipments with the same function have patterns of EMC support in the various EMC guidance categories that can be used as a checklist.

2-7.2.4.2 Factor B, System Type

The systems involved in the 14 system functions listed under EMC selection Factor A can be classified into five distinctive types. Each of these types includes system functions that are technically similar. The relationships between these system types and the previously defined functions are shown in Table 2-10, along with descriptions of the major composition of each system type in terms of system subfunctions.

The usefulness of these broad definitions of system types is that the developer can now rapidly establish a framework within which to specify detailed EMC guidance requirements in terms of technical system characteristics. From prior experience with these system types, the significance of high emitted power requirements, directional versus omnidirectional

TABLE 2-10. RELATION OF SYSTEM TYPES TO SYSTEM FUNCTION AND COMPOSITION

SYSTEM TYPE	SYSTEM FUNCTION	SYSTEM COMPOSITION
I	(a) SSWS (b) SAWS (c) ASWS (d) AAWS	Weapons Sensors Transmission/reception Processors
II	(e) Comm (f) CM/CB (g) EW (h) TDS (i) Nav/IFF (j) Simple CSTA (k) VT fuze	Baseband generator Data entry (modulation) Transmission/reception Demodulation Display/readout
III	(l) Complex CSTA	Complex platform with sensors, communications, processors (multiple site)
IV	(m,n: non-C-E) Unint rec/emit	Munitions Humans Animals Ecological systems Other non-C-E systems
V	(m,n: C-E) Unintended emission Unintended reception Spurious emanation	C-E system

antenna, high versus low information rate requirements, high versus low operating duty cycles, reception requiring high signal/noise contrast ratios, electroexplosive devices and other power-density sensitive components, etc., can be readily recognized. The technical characteristics and performance requirements can then be reflected in narrowing the selection of guidance categories and determining the depths of guidance required as an input for EMC decisions.

Systems in the same type classification generally share the need for the application of similar planning, analysis, measurement, and test methodologies, even if the data resulting from other EMC support activities are not directly applicable to the developer's needs. This can offer the developer broader insights into the applicability of available techniques, and their payoffs and costs than can reference to the more restricted range of choices available through selection Factor A.

2-7.2.4.3 Factor C, Basis of Issue (BOI), Site Sel . . n, and Deployment

The BOI, site selection, and deployment determine the impact that an equipment will have on the EM environment in the field army. A sufficiently large number of equipments densely deployed in the mission area tend to be self-jamming. The actual deployment and use of the equipment, as specified by doctrine reflected in TOE unit organization descriptions and DA field manuals, also impact on EMC analysis requirements.

The BOI, site selection, and deployment factor has a major impact on planning for the use of interference models and may be summarized in terms of the complexity with which the various EMC analysis techniques are employed. When a substantial number of C-E radiators or receivers are deployed in the mission area, and their performance may be affected by many friendly or hostile equipments in the same area, the developer must consider the use of analytical techniques which will take these wide-area effects into account. The need for a full computerized simulation of the field army and the hostile threat is not necessarily implied by this requirement, for rules of thumb or planning factors may be applicable to this case, depending upon circumstances considered in par. 2-7.2.3, "Depth of Guidance".

However, the BOI for a new C-E equipment may result in a sparse distribution compared to other radiators which already have been fielded. In this case, some EMC information categories may either be omitted entirely, or may be broad and general. Hazard evaluations and economic assessments are examples of categories which may be omitted or receive more general treatment if the radiators have modest BOI's and emit low powers.

2-7.2.4.4 Factor D, Evolutionary vs Technologically New Development (evol/new)

An equipment or a system development may be evolutionary, i.e., it may involve only changes in equipment characteristics such as weight and cube, or reliability and ease of operation and maintenance. Operational changes may consist of modest improvements in communication or radar ranges, quality or resolution of received signals, or more suitable pulse repetition frequencies. In contrast to such evolutionary development, a development may be technologically new. A technologically new development may result in the capability to satisfy a new mission requirement, or it may include radical innovations in circuit design or components which result in substantial improvements in performance.

In the case of an evolutionary development, there may be relatively minor and predictable EMC problems, requiring only a few EMC guidance information categories. Opportunities for using the results of previously conducted EMC analyses may be considerable if the changes incorporated in the new equipment are primarily internal upgradings to reflect improved component technology. However, for a technologically new development, the effects of EMC guidance requirements must be assessed individually for each information category. The opportunities for use of the prior EMC analytical results on similar systems, as in the case of Factors A and B, will be substantially fewer than for evolutionary development cases.

2-7.3 PROGRAM DOCUMENTATION

LCSMM flow charts, Figs. 2-8 and 2-9, show the time phasing for the preparation of typical material development documents. In order to have the appropriate EMC data available for the documents, EMC decision points must have been achieved, and appropriate EMC guidance category information collected on schedule. Table 2-11 shows the EMC guidance category information required for the principal material development documents of a typical program.

TABLE 2-11. EMC GUIDANCE CATEGORY INFORMATION REQUIRED FOR THE PRINCIPAL MATERIEL DEVELOPMENT DOCUMENTS

EMC GUIDANCE CATEGORY	MATERIEL DEVELOPMENT DOCUMENTS									
	DP	BR*	DCP	CFP	BOIP	CTP	QQPRI	Spectrum Alloc. Request	FM	TM
1. C-E system feasibility and performance requirements	x	x	x	x		x			x	
2. Command and organizational principles	x	x	x	x	x	x	x	x	x	
3. System operational factors	x	x	x	x	x	x	x		x	x
4. Economic assessment	x		x	x						
5. Electromagnetic environment evaluation	x	x	x	x	x	x		x	x	x
6. Natural environment evaluation	x	x				x		x	x	x
7. Hazard evaluation	x	x				x		x	x	x
8. Equipment and performance characteristics	x	x	x	x		x		x	x	x
9. Conformance to or waivers of EMC standards or specifications	x					x				x
10. Spectrum signatures								x	x	x
11. Measures of system effectiveness	x		x	x		x				
12. Site survey and selection								x	x	x
13. EMC training data	x						x		x	x
14. MIJI report analysis	x						x		x	x

*BR = basic requirements such as operational capability, materiel development objectives, requirement operational characteristics.

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APPENDIX A
OUTLINE OF CONTENT OF EMC PROGRAM PLAN (REF. 6)

A-1 INTRODUCTION

1. Purpose of plan
2. Scope of program
3. Description of system, subsystem, or equipment.

A-2 ORGANIZATION AND RESPONSIBILITIES

1. Contractor responsibilities
   1.1 Interface with procuring activity
   1.2 Key personnel
   1.3 Line and functional organization
   1.4 EMC advisory board (EMCAB)
   1.5 Facilities
2. Subcontractor responsibilities
   2.1 Interface with prime contractor
3. Coordination between prime contractor and subcontractors (how, when, and with whom?)
4. Reporting of problems to procuring activity and how such problems are to be handled within company.

A-3 EMC MILESTONES

A-4 APPLICABLE EMC DOCUMENTS AND REQUIREMENTS

1. Military
2. Company
3. Other.

A-5 EMC ENGINEERING ACTIVITIES

1. Application of applicable documents
2. Design review and schedules
3. Degradation criteria
4. Safety margins
5. Radiation hazard considerations
6. Frequency allocations.

A-6 EMC DESIGN CRITERIA

1. Design techniques to preclude EMI (bonding, grounding, shielding, cable separation, etc.)
2. Precautions to preclude spurious emanations, responses and unwanted resonances
3. Precautions to conserve frequency spectrum
4. Consideration of operational electromagnetic environment
5. Utilization of suppression techniques.

A-7 PREDICTION OF PROBLEM AREAS

1. Description of prediction and analysis techniques to be employed
2. Identification of operational problems anticipated
3. Proposed solutions for problems.

A-8 EMC TESTING FOR SYSTEMS, SUBSYSTEMS, AND EQUIPMENT

1. Engineering development
2. First article
3. Acceptance
4. Integration
5. Spectrum signature.

A-9 DOCUMENTATION AND SCHEDULE

1. Control plan
2. Test plan
3. Test report
4. Charter for EMCAB.

APPENDIX B
CONTENT FOR EMI CONTROL PLAN (REF. 6)

B-1 MANAGEMENT

The EMI Control plan shall include the specific organizational responsibilities, lines of authority and control, and the implementation plan, including milestones and schedules. In addition, the detailed EMI and EMC requirements to be imposed on associated suppliers and subcontractors for vendor items, test requirements to be placed on independent testing laboratories, a complete listing of all EMI and EMC requirements in supplier's procurement documentation, and additional requirements for GFE shall be enumerated.

B-2 SPECTRUM CONSERVATION

The program is employed to minimize emission spectrum and receiver bandwidths and to control oscillator frequencies, pulse rise times, harmonics, side bands, and duty cycles within the constraints of the equipment or subsystem. Specified design parameters shall be defined. A completed copy of DD Form 1494, shall be included for information purposes only.

B-3 EMI MECHANICAL DESIGN

The control plan shall describe the material and construction to be used to provide the inherent attenuation to electromagnetic emissions and susceptibilities while still meeting the contract end item specification requirements. Specific data shall include but need not be limited to the following:

a. Type of metals, castings, finishes, and hardware employed in the design

b. Type of construction, such as compartmentalizing, filter mounting and isolation of other parts, type and characteristics of filtering used on openings including ventilation parts, access hatches, windows, meter faces and control shafts, and type of attenuation characteristics of RF gaskets used on all internal and external mating surfaces

Shielding and design practices employed for determining shielding effectiveness

d. Corrosion control procedures.

B-4 ELECTRICAL AND ELECTRONIC WIRING DESIGN

The proposed electrical and electronic wiring design, cable separation, and routing to minimize emission and susceptibilities shall be described in accordance with the classification procedure required by the contract. Grounding philosophy shall be described in detail and methods of shielding and routing of cables shall be enumerated. Equipments or subsystems, consisting of a number of black boxes shall have interconnecting cabling diagrams supplied.

B-5 ELECTRICAL AND ELECTRONIC CIRCUIT DESIGN

The EMI control and suppression techniques which will be applied to all parts and circuitry, whether capable of generating undesirable emanations or suspected of being susceptible to the fields and voltage levels specified by the contract, shall be described fully. The specifically required design data shall include but not be limited to the following:

a. Choice of parts and circuitry, the criteria for use of standard parts and circuitry, and bonding and grounding techniques

b. Justification of selected filter characteristics including type and attenuation, technical reasons for selecting types of filters (for example, absorptive versus nonabsorptive filters) and specific circuit applications

c. Part location and separation based on orientation of EM fields for reduction of emissions, susceptibility, or both

d. Indicate valid technical reasons for selection of pulse shape. Pulse shapes utilized shall minimize the electromagnetic spectrum employed consistent with achieving design performance.

e. Location of critical circuits and decoupling techniques employed for each

f. Shielding and isolation of critical circuits.

B-6 ANALYSIS

Prediction or analysis techniques employed to determine adequacy of supplier's conclusions shall be included. Specific aspects of the mechanical, electrical, and electronic design to be included are as follows:

a. Adequacy of mechanical construction, and an analysis of the shielding afforded by the proposed designs over the specified frequency range and energy level

b. Complete frequency matrix of all frequencies associated with receivers and transmitters (see Fig. B-1), expected spurious responses of receivers at input signal levels and at frequency range(s) specified in the contract, and expected spurious outputs of items such as transmitters, local oscillators, and frequency synthesizers. Spurious responses f_{sp} will be determined by use of the following equation:

$$f_{sp} = \frac{pf_{lo} \pm f_{if}}{q} \qquad \text{(B-1)}$$

where

p = all integers, including zero, representing harmonics of the local oscillator (lo)

q = all integers, except zero, representing harmonics of the spurious signal

f_{lo} = local oscillator frequency

f_{if} = intermediate frequency

The frequency matrix is to used to formulate a subsystem test matrix of source and victim equipments. The test matrix will be included in the subsystem test plan.

c. Worst case analysis of multivibrators, switching (single and repetitive) and logic circuits, and clock and strobe signals.

d. Analysis of circuitry, subassemblies, and total equipment, or subsystem, including cabling and loads for:

(1) The prediction of susceptibility to internally and externally generated fields and voltages, whether below or above the limits specified in the contract

(2) The prediction of emissions, whether below or above the limits specified in the contract.

e. Subsystem analysis for mobile or fixed installations with two or more antennas shall include a description of radiation characteristics from antennas including fundamental and spurious energy, and discuss minimizing antenna coupling and isolation achieved by placement and location of antennas.

B-7 PROBLEM AREAS

Plans for potential EMI and EMC interface problems shall be presented, including the procedures for defining problems, formulating solutions, implementing and testing the solutions, and documentation procedures.

B-8 UPDATING

The method of updating the control plan shall be indicated.

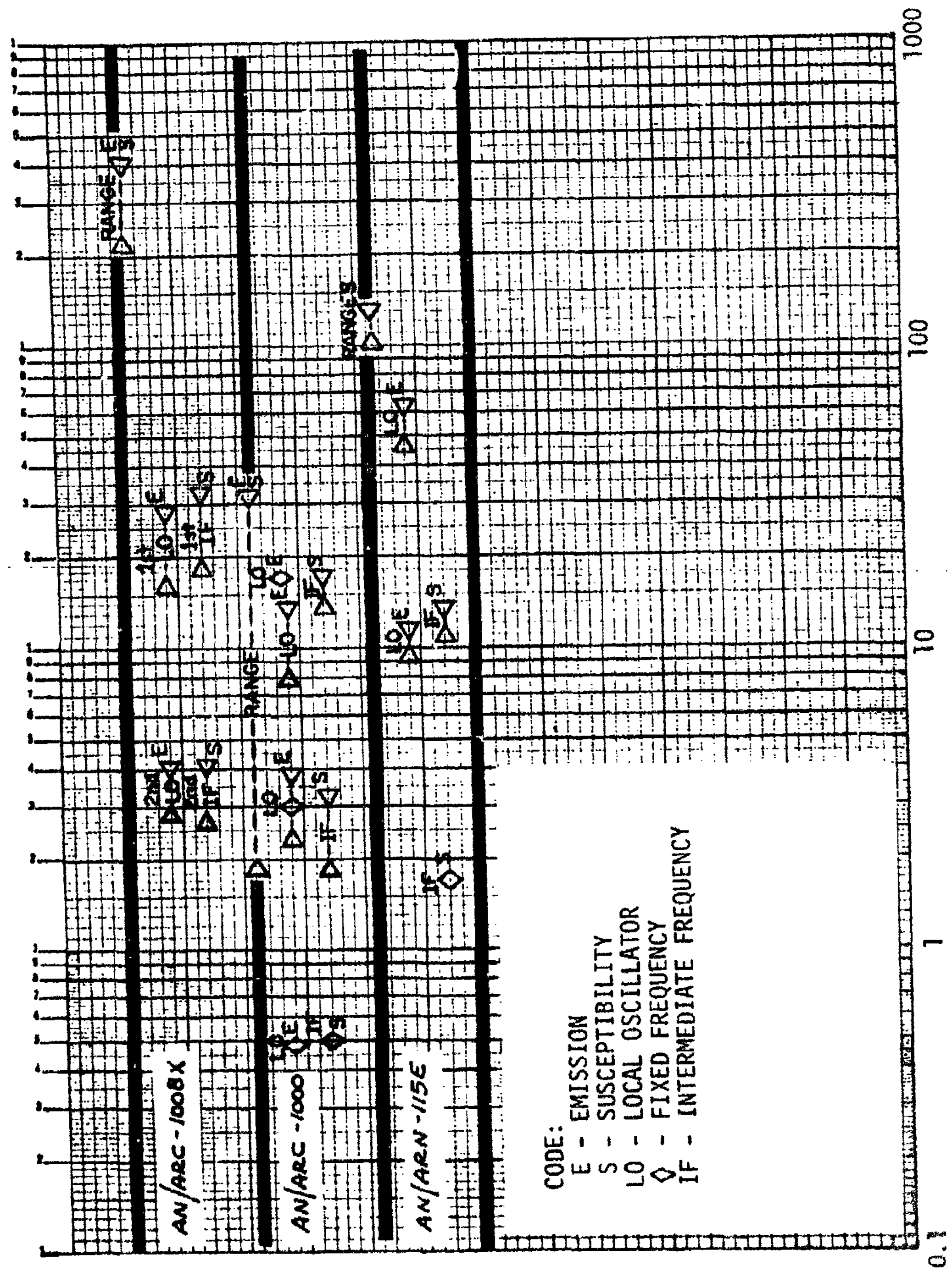


Figure B-1. Sample Frequency Matrix (Ref. 6)

APPENDIX C
CONTENT FOR EMI TEST PLAN (REF. 6)

C-1 INTRODUCTION OR SCOPE

The following shall be included:

a. An opening statement indicating the purpose of the plan and its relationship to the overall EMC program for the equipment and subsystem

b. A table listing all the tests to be performed, the paragraph number of the plan, and the corresponding test method of the basic standard.

C-2 APPLICABLE DOCUMENTS

Documents shall be listed as follows:

a. Military (standards, specifications, etc.)

b. Company (any in-house documents for calibration or quality assurance)

c. Other documents (Society of Automotive Engineers procedures, drawings, etc.)

C-1 TEST SITE

The following shall be included:

a. Description of test facility. shielded enclosure (size, power availability, filters, attenuation characteristics of room to electric, magnetic, and plane waves)

b. Description of ground plane (size and type) and methods of grounding or bonding test sample to the ground plane in order to simulate actual equipment installation

c. Spot check measurements of the ambient electromagnetic emission profile of the test facility, both radiated and conducted, to determine ambient suitability.

C-4 TEST INSTRUMENTATION

Instrumentation to be used shall be described as follows:

a. When matching transformers or band-reject filters are used, their characteristics must be described.

b. Bandwidth of the measurement instrumentation shall be specified.

c. List of test equipment.

d. Scanning speed used to drive EMI measuring equipment.

e. Monitoring equipment utilized during measurements.

C-5 TEST SAMPLE SET-UP

A description of the test sample shall include the actual physical layout of equipment under test, depicting position of test sample and feedthrough capacitors or line impedance stabilization networks on the ground plane, lead dress, bond straps, real or simulated loads, electrical or mechanical, and any test sets employed in the test. (Notes may be used to indicate height above ground plane for leads).

C-6 TEST SAMPLE OPERATION

a. Modes of operation for each test and operating frequency

b. List of control settings on the test sample

c. List of control settings on any test sets employed or characteristics of input signals

d. Test frequencies at which oscillators, clocks, and similar equipment may be expected to approach test limits

e. Performance checks initiated to designate the equipment as meeting minimal working standard requirements

f. Circuits, outputs, or displays to be monitored during susceptibility testing shall be enumerated, as well as the criteria for monitoring for degradation of performance

g. Normal, malfunction, and degradation of performance criteria (i.e., change in output spectrum, change in $(S + N)/N$ ratio, loss of synchronization, change in output waveform) for susceptibility testing shall be described.

C-7 TEST PROCEDURE

Test procedures to be employed to demonstrate compliance with the contractual requirements shall be fully described as a minimum and the following shall be included:

a. Block diagram depicting test setup for each test method

b. Test equipment used in performance of the test and methods of grounding, bonding, or achieving isolation for the measurement instrumentation

c. Detailed step-by-step procedures enumerating probing of the test sample, placement and orientation of probes and antennas, frequency range of test, selection of measurement frequencies, detector functions used, data to be recorded, frequency of recording data, and the units of recorded data

d. During susceptibility testing, the actual modulation characteristics of the interfering signal (amplitude, type, degree, waveform) shall be specified.

C-8 SUBSYSTEM TESTS

The test matrix similar to Table C-1 shall be included in subsystem test plans and enumerate potential:

a. Receiver to receiver interactions

b. Transmitter to receiver interactions

c. Transmitter to active and passive devices (magnetic devices).

C-9 DATA TO BE RECORDED

The data to be recorded shall be outlined as follows:

a. Sample data sheet

b. Sample test log

c. Sample graphs.

TABLE C-1. SAMPLE TEST MATRIX (Ref. 6)

I. Receiver-to-receiver interaction:

Source receiver	Victim receiver	Source receiver frequency, MHz	Victim receiver frequency, MHz	Test frequency, MHz	Interaction
AN/ARC-100BX	AN/ARC-1000	230.00	17.1	17.1	1st LO of Source with Victim
AN/ARC-100BX	AN/ARC-1000	238.80	25.1	25.1	1st LO of Source with Victim

II. Transmitter-to-receiver interaction:

Source transmitter	Victim receiver	Source transmitter frequency, MHz	Victim receiver frequency, MHz	Test frequency, MHz	Interaction
AN/ARC-1000	AN/ARC-100BX	3.85	225.95	3.85	Source frequency with 2nd IF of victim
AN/ARC-1000	AN/ARC-100BX	22.500	232.50	22.500	Source frequency with IF of victim
AN/ARC-1000	AN/ARN-115E	22.000	110.00	110.00	5x source frequency with victim
AN/ARC-1000	AN/ARN-115E	29.000	116.000	116.00	4x source frequency with victim

CHAPTER 3
EMC PHENOMENA

In this chapter models are developed for representing electromagnetic phenomena which are fundamental to electromagnetic compatibility. The parameters of the equivalent circuits of sources and susceptors are all functions of frequency over the range from below the power frequency to about 30 GHz. Because it is a practical impossibility to identify details completely, e.g., of impedances over this range, actual data on specific devices are limited.

Furthermore, many EMC phenomena vary more or less randomly from moment to moment, or from frequency to frequency (in the case of broadband phenomena). It is practical to describe these variations only in a statistical sense by using appropriate distribution functions or their parameters such as means, variances, or other measures.

3-0 LIST OF SYMBOLS

A = articulation index; effective area of circuit, m²; attenuation factor (plane earth or surface wave), dimensionless; local oscillator frequency, Hz; amplitude, dimensionless; amplitude, V or A; area of a loop, m²

A_{eff} = effective aperture, m²

A_r = effective area of receiving antenna, m²

A_t = effective area of transmitting antenna, m²

dA_s = incremental area of the source, m²

a = loop radius, m; bandwidth constant; saturation voltage magnitude; radius of earth, m

a_f = bandwidth constant

a_n = constants in power series expansion; constants in Fourier series expansion

a_o = fraction of output current resulting from saturation, dimensionless

B = magnetic flux density, T; bandwidth, Hz

ΔB = bandwidth of an ideal bandpass filter; bandwidth over which radio noise power is uniformly distributed, Hz

$B_e = \int_0^\infty \frac{|H(f)|^2}{|H(f_o)|^2} df$, effective noise power bandwidth of a network, Hz

$B_i = \int_0^\infty \frac{|H(f)|}{|H(f_o)|} df$, effective impulse bandwidth of a network, Hz

B_n = narrow bandwidth, Hz

B_w = wide bandwidth, Hz

B_3 = 3-dB bandwidth, Hz

B_6 = 6-dB bandwidth, Hz

b = $e/(kT)$ = (40 V⁻¹ at 300 K); height to ground plane, m; interference amplitude; magnitude of output current at saturation, A

b_n = constants in Fourier series expansion

b_f = bandwidth constant

C = capacitance, F

C_c = stray capacitance (between wires), F

CNR = carrier-to-noise ratio, dimensionless

CIR = carrier-to-interference ratio, dimensionless

c_o = speed of light, 3.00×10^8 m/s

D = desired component at the output of a mixer, V; distance between wires, in.; antenna diameter, m

D_u = ratio of upper decile to median value F_{am}, dimensionless

D_l = ratio of median value F_{am} to lower decile, dimensionless

d = duration, s; diameter of conductor, in.; distance of separation of parallel wires; loop diameter, m; helix diameter, m

dB( ) = level in decibels expressed with respect to quantity in parentheses

E = voltage; electric field strength, V/m

E_i = interfering voltage, V

E_o = generated noise, V

E_{pk} = peak voltage, V

E_S = Thevenin equivalent source voltage

E_{sat} = voltage at saturation, V

e = electronic charge, 1.602×10^{-19} C; induced electromotive force, V; naperian base

e_{om} = peak value of output voltage, V

e_2 = electromotive force induced in circuit 2 by circuit 1, V

erfc = complementary error function

F = noise figure, dimensionless

$\bar{F}$ = average noise figure, dimensionless

$F(f)$ = spot noise figure, dimensionless

F_a = noise figure (isothermal), dB

F_{am} = median value of F_a, dB above kTb

f = frequency, Hz

Δf = Doppler frequency shift, Hz

$f(u)$ = normalized surface transformer impedance function, dimensionless
$f(x)$ = probability distribution
f_a = frequency of desired signal, Hz
f_b = frequency of interfering signal, Hz
f_o = fundamental frequency, Hz; center frequency; reference frequency, Hz
f_s = signal carrier frequency, Hz
f_{SD} = desired signal frequency, Hz
f_{SI} = interfering signal frequency, Hz
G = power gain, dimensionless
$G(f)$ = available power gain, dimensionless; transfer function, dimensionless
$G(\theta, \varphi)$ = gain pattern of antenna, dimensionless
G_r = receiving antenna gain, dimensionless
G_t = transmitting antenna gain, dimensionless
$g(t)$ = transconductance (time varying), mho
g_c = conversion transconductance, mho
g_m = transconductance, mho
g_n = nth order conversion transconductance, mho
H = magnetic field strength, A/m
$H(f)$ = clearance, voltage or current transfer function, dimensionless
H_n = nth Fresnel zone clearance, m
H_r = component of magnetic field in r-direction, A/m
H_θ = component of magnetic field in θ-direction, A/m
$H_R(f)$ = relative response, dimensionless
H_n^k = Hankel function of the kth kind, order n
h = antenna height (distance to ground plane) m; Planck's constant 6.626×10^{-34} J·s; height above ground, in.; height of obstruction, ft
$h(t)$ = impulse response, s^{-1}
h_o = minimum effective antenna height, m
I = undesired or interfering component, V; current, A
$I(n)$ = current at harmonic n, A
$I(t)$ = interference
I_C = common-mode current, A
I_D = differential-mode current, A
I_{dc} = direct current, A
I_e = equivalent leakage radiation current, A
I_L = current in loop (antenna); interference current, A
I_M = measured current, A
$I_0(x)$ = 0th order Bessel function of the first kind
I_{pk} = peak current, A
I_{rms} = rms value of current at antenna terminals, A
I_s = reverse saturation current, A
i = instantaneous current level, A
$i_d(t)$ = desired component located on center frequency, A
i_L = load current, A
i_{line} = line current, A
i_n = noise current, A
i_{rms} = rms current, A; rms spectrum amplitude, $A/Hz^{1/2}$
i_1 = current in circuit 1, A
j = integers designating various signals; $\sqrt{-1}$
K = voltage amplification of interfering signal relative to that of desired signal, dimensionless; ratio of gains from receiver input to mixer input at the undesired signal frequency and the desired signal frequency, dimensionless
k = Boltzmann constant 1.38×10^{-23} J/K
k_a = modified earth radius, m
k_2, k_3 = parameter in coaxial cable leakage equation
L = inductance, H; luminance, $W/(sr{\cdot}m^2{\cdot}Hz)$; length of cable, m; loss associated with distance, dB
$L(f)$ = cable noise figure, dimensionless
l = length of cable, ft; integer in series expansion; dipole length, m
M = mutual inductance, H
M_m = magnetic dipole strength, $A{\cdot}m^2$
m = integer in series expansion
m_x = mean of x
N = number of turns (loop or inductor)
$N(f)$ = noise power spectral density per unit solid angle and per unit area at source, $W/(sr{\cdot}m^2{\cdot}Hz)$
$N_e(f)$ = equivalent noise power spectral density at the input contributed by the network, V^2/Hz
$N_i(f)$ = power spectral density (mean square per unit bandwidth) of shot noise current, A^2/Hz
N_o = power spectral density of white noise, V^2/Hz
$N_v(f)$ = power spectral density (mean square per unit bandwidth) of the noise voltage across a resistor, V^2/Hz
$N_1(f)$ = available input noise power spectral density at frequency f, V^2/Hz
$N_2(f)$ = available output noise power spectral density at frequency f, V^2/Hz
n = order of harmonic; integer in series expansion; number of braided strands; ratio of propagation constants, dimensionless
$n(t)$ = added interference, V
P = power
$P(x)$ = cumulative probability (distribution) of x

P_a = average desired signal power, W or dB
P_b = average interfering signal power, W or dB
$\vec{P}_D$ = power density, Poynting vector, W/m^2
P_I = power of an unmodulated interfering signal, W or dB
P_0 = received power expected in free space, W
P_r = received power, W
P_t = radiated power, W
p = helix pitch (twisted pair), m; order of local oscillator harmonic
$p(x)$ = probability density function of x
q = $\pi d/p$, helix parameter, dimensionless; order of signal harmonic
R = reciprocity; resistance, Ω; earth's reflection coefficient, dimensionless
$R(t)$ = bipolar rectangular waveform, A
R_a = radiation resistance of antenna, Ω
R_c = contact resistance, Ω
R_L = load resistor, Ω
R_o = dc resistance per unit length of outer conductor (coaxial cable), Ω/m
r = distance, cm or m; true radius of earth, m; loop radius, cm; distance from axis of helix, m
S = signal level, V
$S(f)$ = spectrum amplitude; s, V·s, or I·s
$S(t)$ = signal input to the phase detector; unit amplitude sampling waveform, dimensionless
dS = noise power spectral density per unit area at the observer, W/(m^2·Hz)
S_E = equivalent surface transfer impedance, Ω
S_I = power spectral density of an interfering signal
S_p = unit voltage pulse spectrum amplitude, V·s
S_s = unit voltage step spectrum amplitude, V·s
S/I = signal-to-interference ratio, dimensionless
$(S/N)_n$ = ratio of the rms spectral density of average speech to rms spectral density of the noise at the output in the nth band, dB
T = absolute temperature, K; thickness of outer conductor, m; length, s; period of waveform, s
$T(\theta, \varphi)$ = brightness temperature, K
T_a = antenna temperature, K
T_e = equivalent network temperature, K
T_o = thermal temperature of the earth, K; reference temperature, K
T_s = system temperature, K; sky temperature, K; solar noise temperature, K
T_{sr} = system temperature referenced to receiver input, K
T_1 = equivalent temperature of source, K; reference temperature, K
t = time, s
t_s = time at saturation, s
V = voltage, usually fixed, V
$V(f)$ = voltage Fourier Transform, V·s
V_D = peak amplitude of the desired signal at the receiver input, V
V_d = ratio of the rms voltage to the average voltage of the noise envelope, dB
V_{dm} = median value of V_d, dB
V_{dn} = V_d for narrow bandwidth, dB
V_{dw} = V_d for wide bandwidth, dB
V_I = peak amplitude of the interfering (undesired) signal at the receiver input, V
V_i = interfering voltage, V
V_L = load voltage, V
V_{pk} = peak voltage, V
V_{SAT} = voltage at saturation, V
V_{SD} = peak amplitudes of the envelope of the desired signal at mixer input, V
V_{SI} = peak amplitude of the envelope of the undesired signal at mixer input, V
Var(x) = variance
$VSWR$ = voltage standing wave ratio, dimensionless
v = applied voltage, V
$v(t)$ = time varying voltage, V
v_n = noise voltage, V
$v_n(t)$ = random noise voltage, V
v_o = amplifier output voltage, V
v_{rms} = rms voltage, V; rms spectrum amplitude, V/Hz$^{1/2}$
$v_s(t)$ = signal amplitude modulation function, V
W_n = articulation index parameter
x = input to nonlinear device
$x(t)$ = signal voltage waveform, V
$x_o(t)$ = oscillator voltage, V
$x_s(t)$ = signal voltage, V
$Y(f)$ = admittance
y = output of nonlinear device, frequency integration variable
$y_I(t)$ = interference signal
y_1 = interference cross-modulation component, A
y_s = desired cross-modulation component, A
$Z(f)$ = impedance, Ω
Z_L = load impedance, Ω; line impedance, Ω
Z_M = measured load impedance, Ω
Z_s = Thevenin equivalent source impedance, Ω; generator impedance, Ω
Z_t = surface transfer impedance, Ω/m
Z_0 = impedance of free space, 377 Ω

z = z-coordinate; polarization constant; mutual impedance, Ω

α = parameter of cable braid (coaxial cable), dimensionless; earth's absorption coefficient, dimensionless

β = phase constant, $2\pi/\lambda$, m^{-1}

γ = propagation constant, m^{-1}; carrier-to-noise ratio, dimensionless

Δ = envelope of filtered random noise divided by the rms value of the envelope, dB; $4\pi h_1 h_2/(\lambda r)$, dimensionless

δ = depth of penetration (skin depth), m; phase difference, rad

ϵ = error; permittivity, F/m

ϵ_0 = permittivity of free space, 8.85×10^{-12} F/m

ϵ_r = relative dielectric constant, dimensionless

ζ = displacement of the center conductor of a coaxial cable from the true center

η = wave impedance = $\sqrt{\mu/\epsilon}$, Ω

η_0 = $\sqrt{\mu_0/\epsilon_0}$, characteristic impedance of free space, Ω

θ = co-latitude angle in spherical coordinate system, rad; angle between reflected wave and ground level, rad

λ = wavelength, m; random phase angle, deg

μ = permeability, H/m

μ_0 = permeability of free space, $4\pi \times 10^{-7}$ H/m

σ = standard deviation; conductivity, mhos/m

σ^2 = mean square value of noise, V^2

σ_x^2 = variance of x

τ = time constant, s; deviation from unperturbed zero crossing time, s; switching time, s; duration of nonsaturation interval, s

φ = longitude angle in spherical coordinate system, rad; phase angle, rad

$\varphi(f)$ = phase vs frequency characteristic of IF amplifier, rad

φ_s = phase angle, rad

ϕ_i = power spectral density of an interfering signal, W/kHz

ψ = grazing angle with respect to the earth's surface

ω = angular frequency, rad/s

ω_{if} = intermediate angular frequency, rad/s

ω_o = local oscillator angular frequency, rad/s

ω_s = angular frequency of signal carrier, rad/s

ω_{SD} = desired angular frequency, rad/s

ω_{SI} = angular frequency of undesired signal, rad/s

Ω = solid angle, sr; pointing angle, sr

$d\Omega$ = increment of solid angle, rad

$\langle\ \rangle$ = mean value

3-1 SOURCE MODELS

3-1.1 GENERAL COMMENTS

For the purpose of simplifying analysis and prediction it is convenient to use "linear" mathematical models for representing various circuit elements. This is so in spite of the fact that nonlinear mechanisms are frequently responsible for interference generation and susceptibility. The usual procedure is to use the linear model and modify it as necessary to account for observed phenomena. Thus a typical source of conducted interference is represented as an electromotive force of specified value in series with a defined impedance. Although it should be recognized that such a representation is an approximation to the true model, fortunately it does appear to have general validity.

Emitters usually are identified as narrowband or broadband, in accordance with the nature of the spectrum of the emitted energy. Since these are relative terms, the dividing line between them is not sharp; one of the best bases for distinction is the acceptance bandwidth of any susceptible device (susceptor) whose operation may be degraded by the emission. If the bandwidth of the dominant energy is less than that of the susceptor, the emission is "narrow-band", otherwise it is "broadband". Emission arising from modulated sine wave oscillations are usually narrow-band, while those arising from switching action are usually broadband.

3-1.2 NATURAL NOISE SOURCES

Sources of natural noise include thermal noise of resistors and warm body radiators, shot noise in electronic devices, atmospheric noise, cosmic noise emanating from the sun and from sources in outer space, and triboelectric (frictional charging and discharging) noise. The statistical properties of noise from these sources range widely from spectrally flat Gaussian noise, to sporadically impulsive noise. In some cases the statistical parameters remain essentially constant in time. In instances where they are not constant, variations with time of day, season, and year may be given.

3-1.2.1 Electronic Noise

3-1.2.1.1 Sources of Electronic Noise

Associated with the real part of the impedance seen at a pair of terminals of a passive electrical network, is a Gaussian noise process called thermal or Johnson noise, which arises as a consequence of thermal agitation of the charge carriers in conductors.

3-1.2.1.2 Levels and Power Spectral Density of Electronic Noise

The open circuit noise voltage v_n across the terminals of a resistor is a Gaussian random process (Ref. 1) with a power spectral density* $N_v(f)$ given by

$$N_v(f) = \frac{4hfR}{\exp[hf/(kT)] - 1}, \text{V}^2/\text{Hz} \qquad (3\text{-}1)$$

where

k = Boltzmann constant = 1.38×10^{-23} J/K
T = absolute temperature, K
h = Planck's constant, 6.626×10^{-34} J·s
f = frequency, Hz
R = resistance, Ω

Below about 100 GHz, the power spectral density is uniform with frequency and is given by

$$N_v(f) = 4kTR, \text{V}^2/\text{Hz} \qquad (3\text{-}2)$$

If the resistor is connected to a network with voltage transfer function $H(f)$ the effective bandwidth B_e is given by

$$B_e = \int_0^\infty |H(f)|^2 df / |H(f_o)|^2, \text{Hz} \qquad (3\text{-}3)$$

*As is common in the literature, the term "power spectral density" refers to the mean square voltage or current per unit frequency. Later the term "available noise power spectral density" is used and has the dimension power/frequency.

where f_o is the center frequency of $H(f)$, and the expected or mean square value of the output voltage is

$$\langle v_n^2 \rangle^* = 4kTRB_e|H(f_o)|^2, \text{V}^2 \qquad (3\text{-}4)$$

For analysis purposes the resistor can be represented by the equivalent circuits shown in Fig. 3-1 where R is a noiseless resistor, $v_n(t)$ is a random noise voltage with an rms spectrum amplitude given by

$$v_{rms} = \sqrt{4kTR}, \text{V/Hz}^{1/2} \qquad (3\text{-}5)$$

and $i_n(t)$ is a random noise current with rms spectrum amplitude given by

$$i_{rms} = \sqrt{\frac{4kT}{R}}, \text{A/Hz}^{1/2} \qquad (3\text{-}6)$$

The maximum available noise power spectral density from the resistor R is obtained when its load is matched, i.e., the load resistor R_L equals R. The power density is then kT. Within an effective bandwidth B_e, the maximum power is kTB_e in watts.

Shot noise occurs whenever current can be identified as due to the motion of individual electrons. Such is the case in a solid-state device where current flow is a result of the random generation of carriers, or in a vacuum diode where current flow is due to the motion of electrons from cathode to anode. The

*The notation $\langle\ \rangle$ indicates the expected or mean value of the quantity enclosed by the symbol.

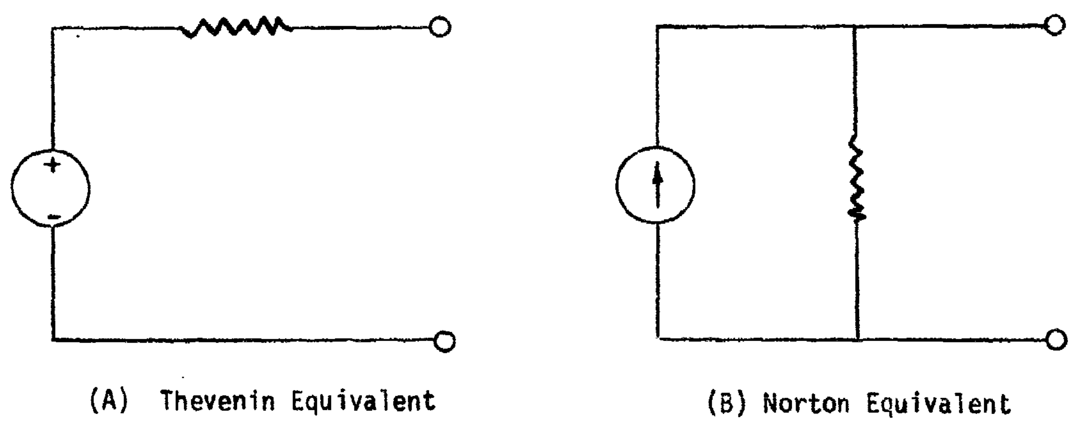


Figure 3-1. Noise Source Equivalent Circuits of a Resistor

current is comprised of a collection of pulses, each having an area equal to the electronic charge. A current so constituted will have a randomly fluctuating component with a power spectral density $N_i(f)$ given by

$$N_i(f) = 2eI_{dc} \text{ ,A}^2/\text{Hz} \quad (3\text{-}7)$$

where

e = electronic charge = 1.6×10^{-19}, C
I_{dc} = average (or direct) current of the pulses, A

Eq. 3-7 is correct at frequencies which are low relative to the reciprocal of the pulse widths.

If such a current source feeds a noiseless network with a current transfer function, $H(f)$, the output mean square value is

$$\langle i_n^2 \rangle = \int_0^\infty 2eI_{dc}|H(f)|^2 df$$
$$= 2eI_{dc}B_e|H(f)|^2 \text{ ,A}^2/\text{Hz} \quad (3\text{-}8)$$

where the effective bandwidth B_e is defined in Eq. 3-3.

3-1.2.1.3 Noise Figure and Noise Temperature

The noisiness of active and passive networks due to shot, thermal, and other noise sources, usually is measured by the noise figure $F(f)$ of the network, defined by

$$F(f) \equiv \frac{N_2(f)}{N_1(f)G(f)} = 1 + \frac{N_e(f)}{N_1(f)}$$
$$= \frac{T_1 + T_e}{T_1} \text{ , d'less} \quad (3\text{-}9)$$

where

$N_2(f)$ = available noise power spectral density at the network output at frequency f, V²/Hz
$N_1(f)$ = available power spectral density at the network input at frequency $f = kT_1$, V²/Hz
$G(f)$ = available power gain of the network, dimensionless
$N_e(f)$ = equivalent noise power spectral density at the input contributed by the network $= kT_e$, V²/Hz
k = Boltzmann constant = 1.38×10^{-23}, J/K
T_1 = equivalent temperature of the source, K
T_e = network equivalent temperature, K

The noise figure as written if Eq. 3-9 is a function of frequency, and is sometimes called the shot noise figure. If it varies significantly with the frequency one uses the average noise figure $\bar{F}$ given by

$$\bar{F} = \frac{\int_0^\infty N_2(f)\,df}{\int_0^\infty N_1(f)G(f)\,df} \text{ , d'less} \quad (3\text{-}10)$$

Or, using Eq. 3-9

$$\bar{F} = \frac{\int_0^\infty F(f)N_1(f)G(f)\,df}{\int_0^\infty N_1(f)G(f)\,df} \text{ , d'less} \quad (3\text{-}11)$$

The noise figure $F_{12}(f)$ of a cascade of two networks in which the individual networks have noise figures F_1 and F_2, and available power gains $G_1(f)$ and $G_2(f)$, respectively, can be written

$$F_{12}(f) = F_1(f) + \frac{F_2(f) - 1}{G_1(f)} \text{ , d'less} \quad (3\text{-}12)$$

Defining T_{e1}, T_{e2}, and T_{e12} as temperatures corresponding to $F_1(f)$, $F_2(f)$, and $F_{12}(f)$,

$$T_{e_{12}} = T_{e_1} + \frac{T_{e_2}}{G_1(f)} \text{ , K} \quad (3\text{-}13)$$

Frequently, the first network in a system is a lead-in cable which is somewhat lossy. It can be shown that for a matched cable

$$\frac{\text{input power}}{\text{output power}} = L(f) > 1 \text{ , d'less} \quad (3\text{-}14)$$

where $L(f)$ is the noise figure of the cable. The overall noise figure of the cable plus the rest of the system (assumed to be the second network) is

$$F_{12}(f) = L(f) + L(f)[F_2(f) - 1]$$
$$= L(f)F_2(f) \text{ , d'less} \quad (3\text{-}15)$$

Or, in terms of temperature with $T_{e_1} = T_1$

$$T_{e_{12}} = [L(f) - 1]T_1 + L(f)T_{e_2}$$
$$= [L(f) - 1]T_1 + L(f)[F_2(f) - 1]T_1$$
$$= [L(f)F_2(f) - 1]T_1 \text{, K} \quad (3\text{-}16)$$

Eq. 3-15 shows that the cable attenuation is reflected directly as a proportional increase in the noise figure.

In receiver applications, the first element in the system is an antenna which is immersed in a noise emitting environment. The noise may arise from the natural sources — atmospheric, terrestrial, and cosmic — or it may arise from the man-made causes to be discussed in par. 3-1.3. The noise entering the system may be accounted for through an antenna temperature T_a. The overall system temperature of a receiver connected to a transmission line and antenna is obtained by augmenting Eq. 3-16 by T_a. That is, the system temperature T_s, referred to the input of the transmission line, is

$$T_s = T_a + [L(f) - 1]T_1 + L(f)T_{e_2}\ ,\mathrm{K} \quad (3\text{-}17)$$

The system behaves as would one in which the antenna radiation resistance was at temperature T_s, and the system was otherwise noiseless. Sometimes it is preferable to reference the temperature to the receiver input. In this case the system temperature is denoted T_{sr}, and it is given by Eq. 3-17 divided by $L(f)$, or,

$$T_{sr} = \frac{T_a}{L(f)} + \left[1 - \frac{1}{L(f)}\right]T_1 + T_{e_2}\ ,\mathrm{K} \quad (3\text{-}18)$$

The system behaves as would one in which the resistance looking toward the transmission line from the receiver input terminals was at temperature T_{sr}, and the system was otherwise noiseless.

3-1.2.2 Atmospheric Noise

3-1.2.2.1 Origin

At frequencies below 20 MHz, the natural electrical noise associated with thunderstorm activity throughout the world plays a dominant role in radio communication. Often, noise of this origin is great enough to make all other sources of noise negligible.

Unfortunately, atmospheric noise is highly changeable with time and place, and in its short term properties may not be represented by any simple random process. One model of atmospheric noise suggested by physical phenomena is a low level Gaussian noise representing distant effects, plus a high level impulse process to represent local effects. However, simple impulse processes, such as the Poisson process in which pulses occur at random, do not accurately represent the local activity. The actual phenomena occur in bursts, so that there is interdependence among adjacent impulses.

3-1.2.2.2 Probability Distribution of Envelope

The first order probability distribution of the envelope in a 200-Hz bandwidth is shown on Fig. 3-2 (Ref. 3). Because the noise is non-Gaussian, the shape of the probability distribution depends on the receiver bandwidth. Empirical techniques for converting these results to other bandwidths are shown in Fig. 3-3.

The distribution depends on a parameter given by the ratio of the rms value to the average value of the envelope, and denoted V_d. For Gaussian noise, the instantaneous value of the envelope has a Rayleigh probability density, and V_d is equal to 1.05 dB. Fig. 3-2 shows the probability distribution of the instantaneous noise envelope normalized to the rms value of the envelope (denoted Δ) for various V_d.

Predictions of the value of V_d are also available in CCIR Report 322 (Ref. 3), and a typical estimate is shown in Fig. 3-4. The estimate is of a quantity designated V_{dm}, where the subscript m signifies median value; i.e., this is the value estimated to be exceeded half the time. Generally, V_d falls off with increasing frequency, suggesting that at higher frequencies the noise as seen in the 200-Hz receiver band becomes more nearly Gaussian. Examination of the range of estimated median values of V_d shows it to be from around 2 to 14 over all frequencies below 20 MHz, and for its entire time history.

Conversions to other bandwidths using Fig. 3-3 are illustrated by example. Suppose an estimate of the envelope distribution is sought at 1 MHz in a band of 2 kHz for the time block and season appropriate to Fig. 3-4; the median value of V_d is about 7 in a 200-Hz bandwidth. In a bandwidth of 2 kHz, implying a bandwidth ratio of 10 (B_w/B_n in Fig. 3-3), the value of V_d in the wide bandwidth (V_{dw} in Fig. 3-3) is obtained by reading the ordinate corresponding to the intersection of $B_w/B_n = 10$ and the curve labeled V_{dn} (subscript n stands for narrow) = 7. A value of V_{dw} (subscript w stands for wide) = 16 is obtained. The proper probability distribution curve for this case is the one marked $V_d = 16$ in Fig. 3-2.

Note the following in Fig. 3-3: (1) if V_d is 1.05 dB in any noninfinitesimal bandwidth, the input noise is Gaussian and V_d is 1.05 for any bandwidth, and (2) as bandwidth is decreased, V_d decreases, tending to approach 1.05 dB; i.e., decreasing bandwidth makes the noise more nearly Gaussian.

3-1.2.2.3 Spatial and Temporal Variations — Long Term Properties

The distributions discussed in par. 3-1.2.2.2 are determined from the instantaneous noise voltages

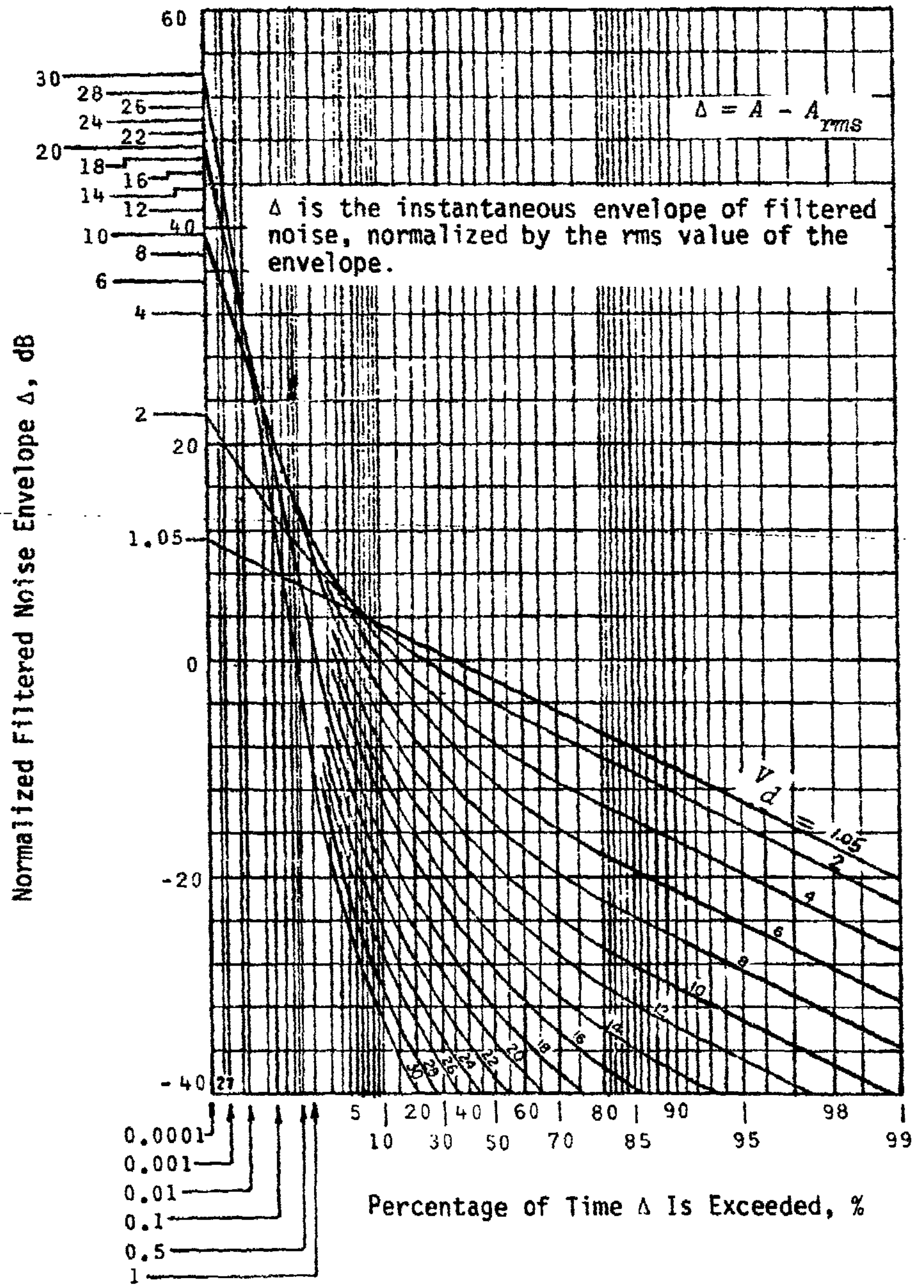
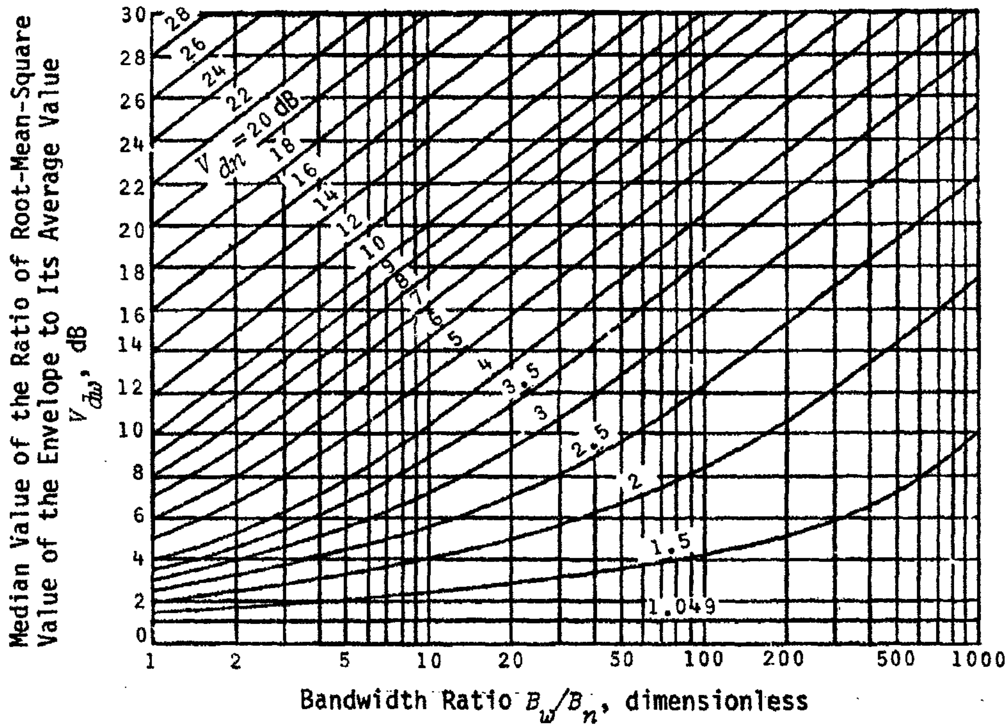


Figure 3-2. Amplitude Probability Distribution of the Noise Envelope (Ref. 3)



Corresponding values of V_{dw} and V_{dn} are read along the appropriate line for the bandwidth ratio.

B_w = the wider bandwidth

B_n = the narrower bandwidth

Figure 3-3. Conversion of V_d in One Bandwidth to V_d in Another Bandwidth

measured over relatively short intervals of time, typically about 0.25 h. The atmospheric noise intensity exhibits important long term variations, and variations with location. To represent these variations, Fig. 3-5 shows isothermal lines, or lines of equal noise temperature, covering the earth. The isothermals are given in terms of a noise figure F_a

$$F_a = 10 \log (T_a/T_o) \tag{3-19}$$

where

T_a = antenna temperature of a short vertical antenna above a highly conductive earth, K

T_o = reference temperature, 288 K

F_a can be converted to field strength in a 1-kHz bandwidth by

$$E = F_a - 65.5 + 20 \log f(\text{MHz}), \text{dB}(\mu\text{V/m})^* \tag{3-20}$$

CCIR Report 322 (Ref. 3) contains 24 figures such as Fig. 3-5, each for a specified 3-month season, and for a 4-h block in each day of the season. Fig. 3-5, for example, gives the noise temperature in the hours 0000-0400, during the months of June, July, and August, in the Northern Hemisphere, and during the months of December, January, and February, in the

* dB ( ) = level in decibels expressed with respect to the quantity in parentheses; in this case, microvolt per meter.

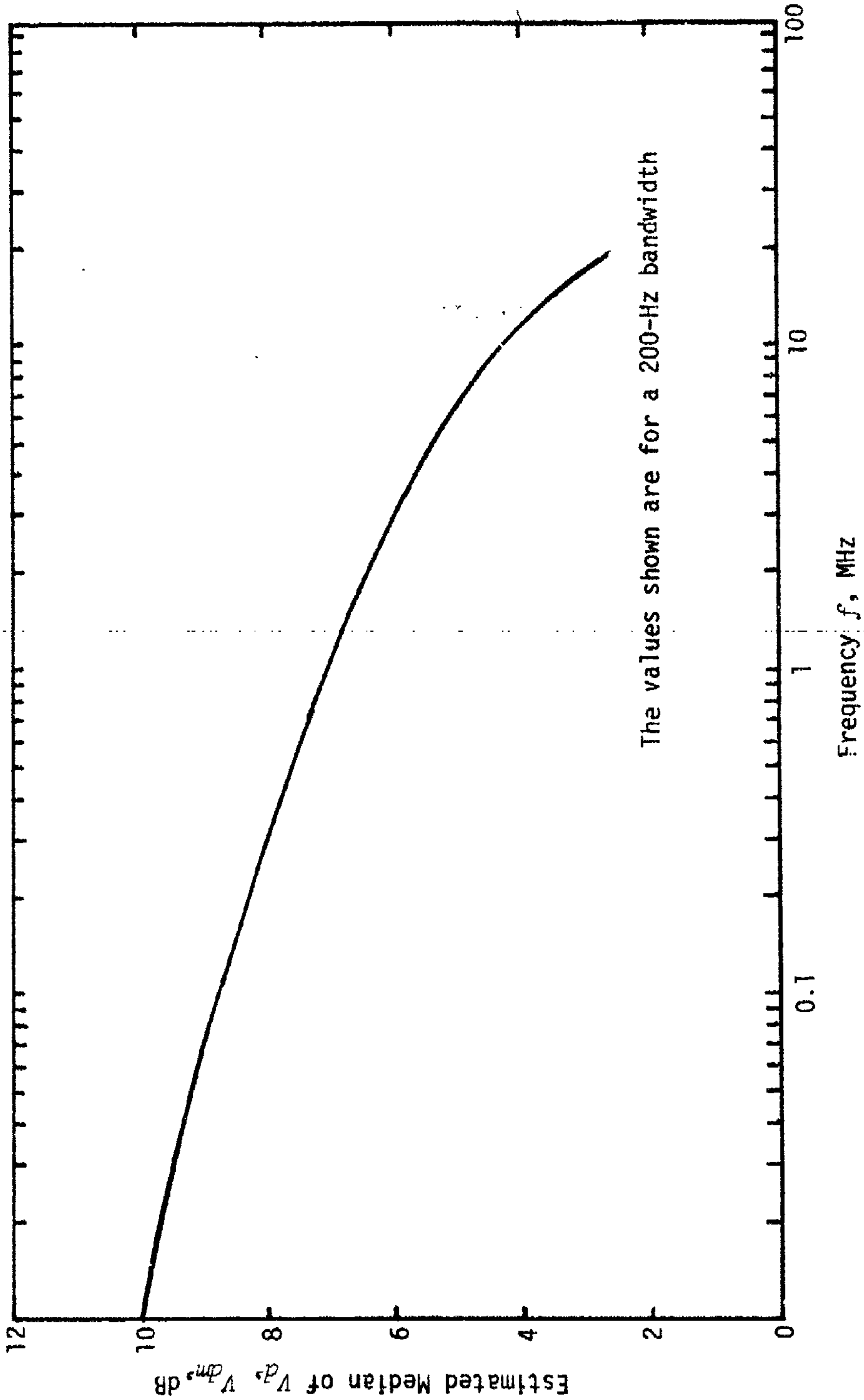


Figure 3-4. Noise Variability (Summer: 0000-0400 h)

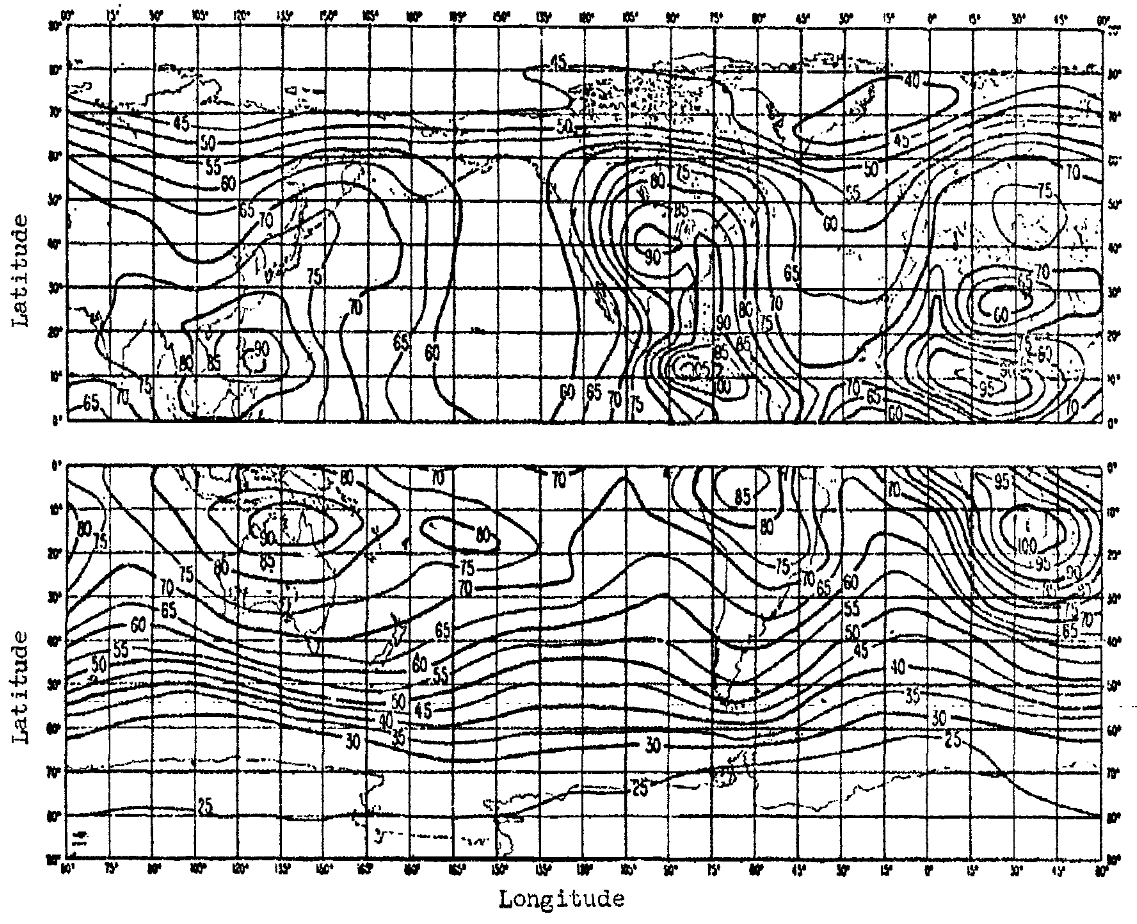


Figure 3-5. Expected Values of Atmospheric Radio Noise F_{am}, dB Above $kT_o b$ at 1 MHz (Summer: 0000-0400 h)

Southern Hemisphere. By organizing the blocks in this fashion, the variations within a time block are kept low. Within each time block, there are about 90 days per yr, and about 360 h. The predictions shown in Fig. 3-5 are based on measurements of F_a over a 0.25-h interval in each hour. The value plotted is F_{am}, where m implies the median value of the observed data over the time block. Other parameters of the empirical distributions of F_a, such as the upper and lower deciles, are given in CCIR Report 322, but are not reproduced here. The F_a statistics are short term averages, as are the V_d statistics mentioned earlier.

The data of Fig. 3-5 apply at a frequency of 1 MHz. The results vary with frequency, and to represent this we show Fig. 3-6 abstracted from CCIR Report 322. Fig. 3-6 is used with Fig. 3-5 to obtain an estimate of noise figure at any frequency below about 20 MHz. For example, if one requires the estimated noise figure at the time applicable to these figures at 100 kHz, and at 60° N. Lat., 45° E. Long., one first enters Fig. 3-5 at the proper position, and finds the noise figure at 1 MHz to be about 50 dB. Then one enters the 50-dB curve on Fig. 3-6 and reads 105 dB at 100 kHz. Note that F_a is a measure of power spectral density, and for moderate bandwidths is not dependent on bandwidth. Fig. 3-6 is plotted only to 10 kHz at the low end. Atmospheric noise levels tend to peak in this region, usually around 2 to 5 kHz. More detailed information on atmospheric noise in this range of frequencies may be found in Watt and Maxwell (Ref. 4).

Fig. 3-6 also shows estimates of man-made noise in

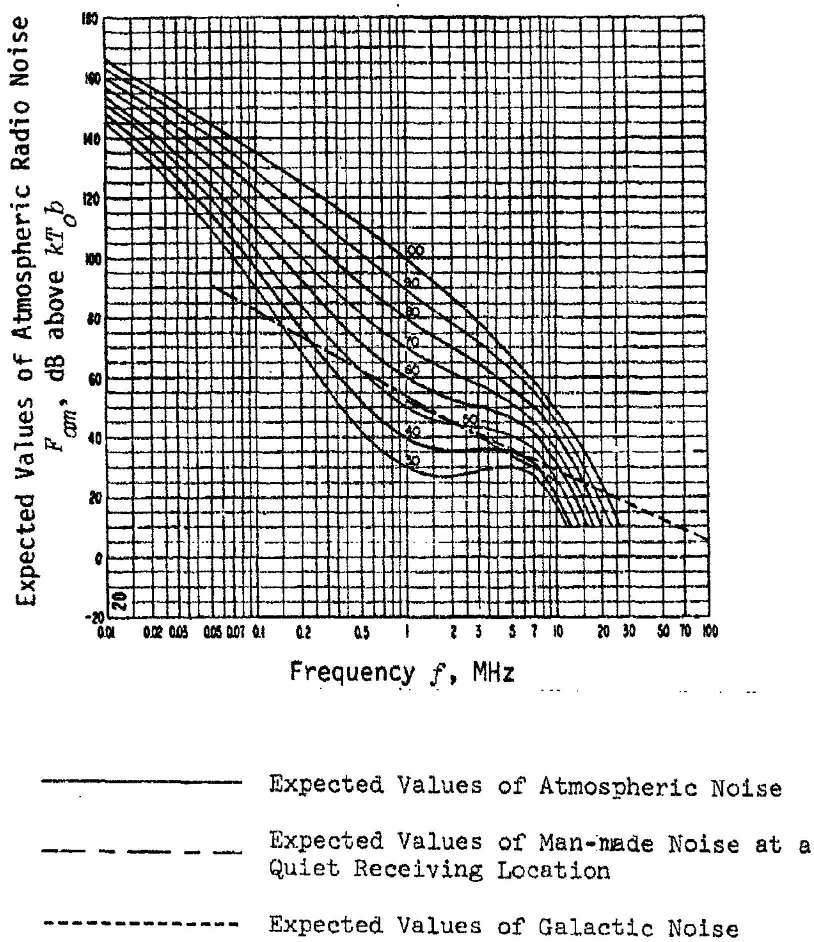


Figure 3-6. Variation of Radio Noise With Frequency (Summer: 0000-0400 h)

relatively quiet areas, as well as galactic noise. From these it is seen that below about 10 MHz, atmospheric noise will predominate at virtually all locations on earth. Galactic noise becomes most significant above 20 MHz. Man-made noise does not appear to be significant here, but that is only because the curve shown applies to quiet locations. In areas of industrial concentration and much auto traffic, man-made noise may be as much as 20 dB greater. Typical levels are shown on Fig. 2-1.

3-1.2.2.4 Atmospheric Pulse Properties — Lightning

Communication systems operating in the HF range will — by choice of signal power level, and by circuit design — be tolerant of atmospheric noise up to some design level. By this means it will be possible to guarantee satisfactory operation for a specified fraction of time. When there are thunderstorms in the immediate vicinity of the receiver, however, the implusive fields generated by lightning will exceed any reasonable quality design level.

Lightning discharges may take place between clouds, within clouds, and from cloud to ground. The latter is the most significant. Watt (Ref. 5, pp. 454-5) makes a distinction between short discharges, in which a single stroke occurs lasting perhaps 4 ms, and long discharges in which there are multiple strokes each lasting about 100 ms, with a stroke separation of about 40 ms. Between strokes there is a small but continuing trickle of charge. Typical measurement results are shown in Fig. 3-7 (Refs. 4 and 6). Field strengths of more than 70 V/m occur with a duration of the high amplitude part of about 200 μs at 1 mi from the discharge.

The multiple stroke phenomena are typified by the current waveform shown in Fig. 3-8 (Ref. 7). Shown there is a succession of 5 strokes, with an average separation of about 70 ms. The distribution of the number of peaks in a discharge is shown in Fig. 3-9 (Ref. 8).

The pulse separations between strokes are seen to be sufficiently long to allow a typical receiver used in the HF band, with a bandwidth from several hundred to several thousand Hertz, to recover to a quiescent state. The pulses will appear as separate discharges, though there may be overlap between discharges occurring simultaneously at different places in the thunderstorm. The phenomenon, as seen at a receiver output, will then appear to be a quasi-random succession of pulses with random amplitudes. Because of the clustering nature of the pulses, the process is not of the nature of Poisson noise where pulses occur totally at random. The probability density of the time between pulses of atmospheric noise is found to be of the form shown in Fig. 3-10 (Ref. 5). It shows a tendency for the time spacing to be around 20 ms. A Poisson process shows a smooth decrease as time spacing increases.

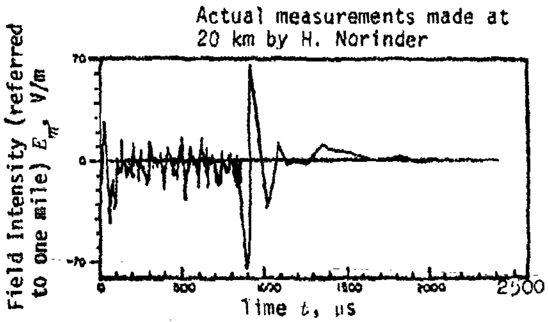


Figure 3-7. Waveform of Effective Radiated Electric Field (Refs. 4 and 6)

3-1.2.2.5 Antenna and Brightness Temperature

The noise power spectral density $N(f)$ per unit solid angle, and per unit area at the source (Fig. 3-11) is given by Planck's radiation law

$$N(f) = \frac{2hf^3}{c_o^2}\{\exp[hf/(kT)] - 1\}^{-1}, \text{W/(sr·m}^2\text{·Hz)} \tag{3-21}$$

where c_o is the speed of light, or by its approximate

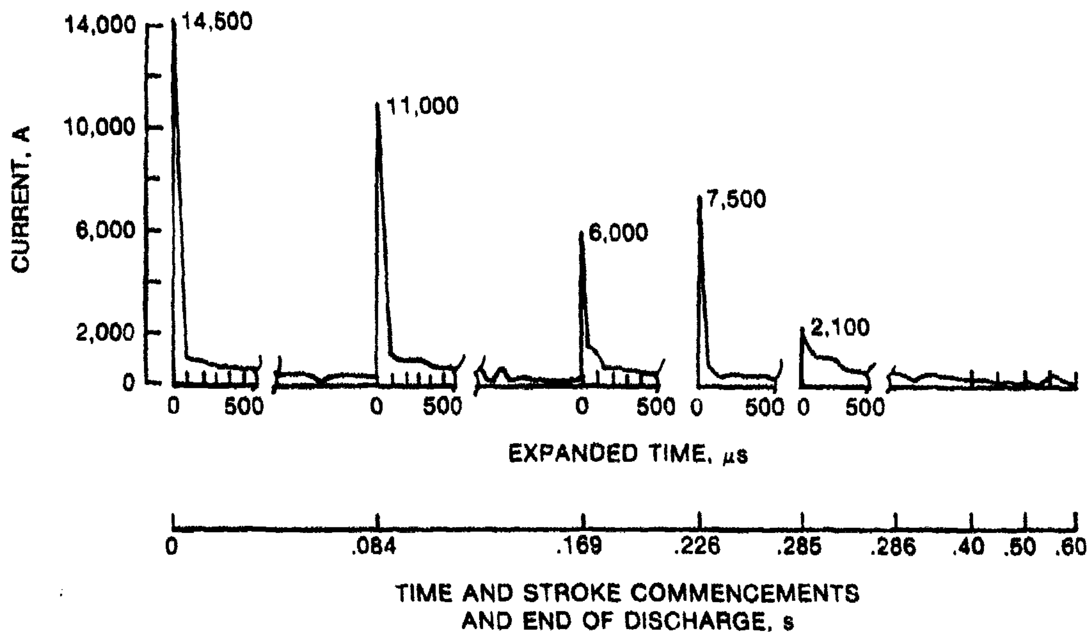


Figure 3-8. Lightning Stroke, Complete Discharge Current (Ref. 7)

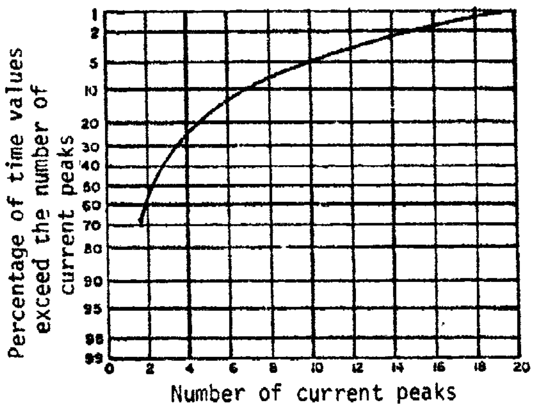


Figure 3-9. Cumulative Distribution of the Number of Current Peaks in an Individual Lightning Discharge (Ref. 8)

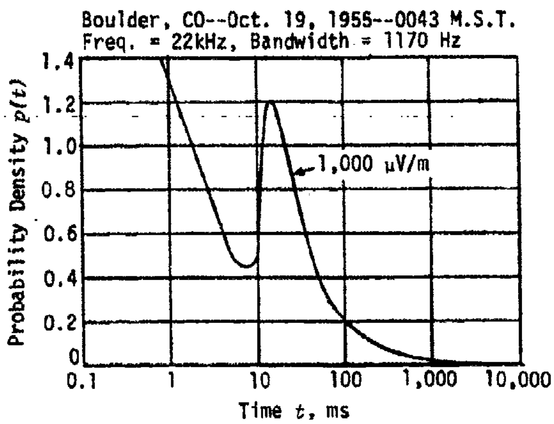


Figure 3-10. Probability Density of the Time Between Pulses of the Atmospheric Noise Envelope

version at low frequencies, known as the Rayleigh-Jeans law

$$N(f) = \frac{2kT}{\lambda^2}, f \ll \frac{kT}{h} \tag{3-22}$$

where λ is the wavelength.

In an increment of solid angle $d\Omega$, measured at the observer's position as seen in Fig. 3-11, the area at the source is $r^2 d\Omega = dA_s$, so that the received power spectral density per unit solid angle due to dA_s is $Nr^2 d\Omega$ in W/(sr·Hz). The noise power spectral density dS per unit area at the observer arising from dA_s is

$$dS = N d\Omega, \quad \text{W/(m}^2\cdot\text{Hz)} \tag{3-23}$$

The luminance L is defined as

$$L = \frac{dS}{d\Omega} = N = \frac{2kT}{\lambda^2}, \text{W/(sr}\cdot\text{m}^2\cdot\text{Hz)} \tag{3-24}$$

As the pointing angle Ω changes, the brightness, L varies so that

$$L(\theta, \varphi) = \frac{2kT(\theta, \varphi)}{\lambda^2} \tag{3-25}$$

where

$T(\theta,\varphi)$ = brightness temperature, K

and θ and φ are the angular dimensions in a spherical coordinate system; θ measures the co-latitude angle and φ measures the longitude angle.

An antenna with gain pattern $G(\theta,\varphi)$ will have an antenna temperature T_a given by

$$T_a = \frac{1}{4\pi}\int_{\varphi=0}^{2\pi}\int_{\theta=0}^{\pi} G(\theta, \varphi)\, T(\theta, \varphi) \sin\theta \, d\varphi d\theta, \text{ K} \tag{3-26}$$

and the available output noise power spectral density of the antenna is KT_a*.

If the antenna has an impulse pattern at the point (θ_0, φ_0) then, since by definition,

$$\frac{1}{4\pi}\int_{\varphi=0}^{2\pi}\int_{\theta=0}^{\pi} G(\theta, \varphi) \sin\theta \, d\varphi d\theta = 1$$

$$T_a = T(\theta_0, \varphi_0), \text{ K} \tag{3-27}$$

i.e., the antenna temperature is the same as the source temperature at the point at which the antenna is aimed. Eq. 3-26 is therefore an average temperature of the source, weighted by the antenna gain pattern.

The brightness temperature has become a standard for quantifying the emissive properties of radiating noise sources, whether or not they satisfy the radiation laws Eqs. 3-21 or 3-22. For real sources the brightness temperature will vary with frequency as well as with angular position.

3-1.2.2.6 Temperature of Earth, Sea, and Atmosphere

One component of the radiation received by an antenna arises from the electromagnetic emissions of hot bodies. Fig. 3-12 (Ref. 9) shows a polar plot of brightness temperature as a function of angle relative to the vertical at 3 GHz for various conditions. If the antenna is above typical ground with vegetation, the

* If the antenna is movable the gain pattern is a function of antenna position.

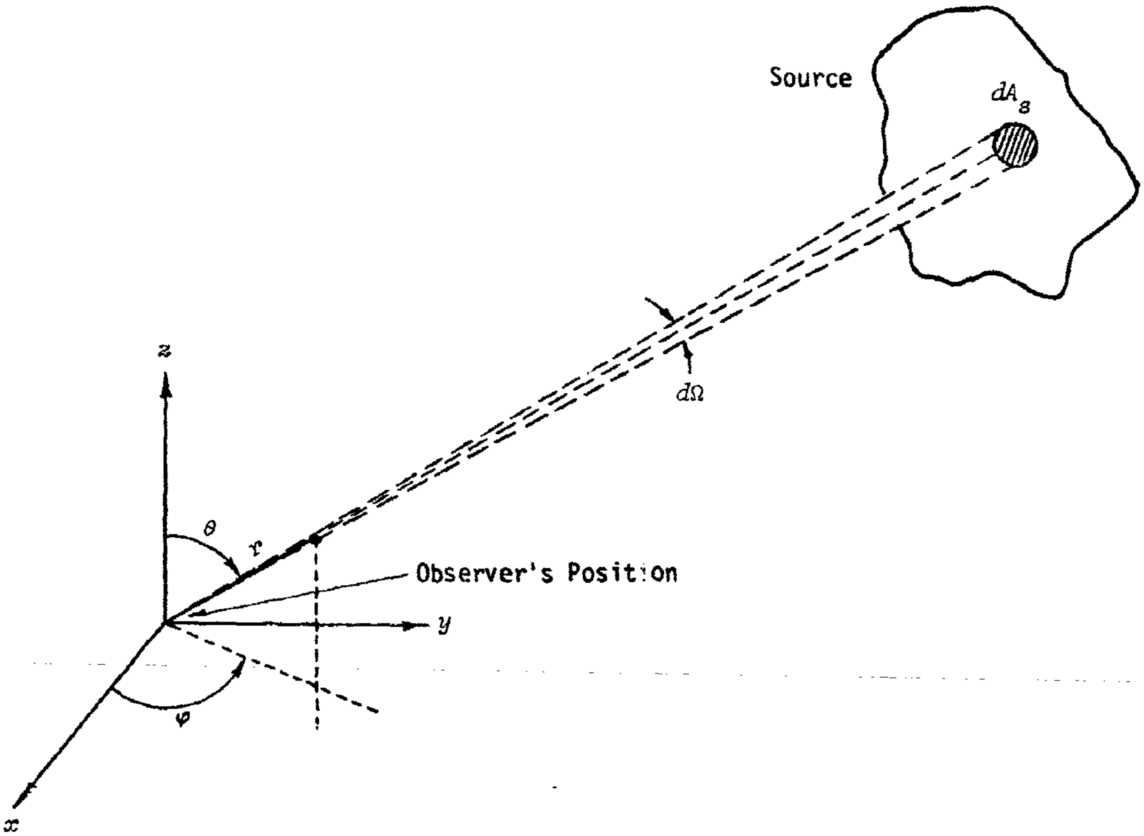


Figure 3-11. Increment of Solid Angle Measured at the Observer's Position

applicable parts of the figure are the part labeled T_s (representing sky temperature) and the dotted curve C. The curves T_s and B apply if the antenna is above calm sea. Finally, if the earth is a hypothetically perfect reflector, the appropriate curves are T_s and A. If θ is the angle from the vertical, the antenna temperature of a sharply directed beam antenna is

$$T_a(\theta) = \alpha(\theta)\,T_o + R(\theta)\,T_s(\pi - \theta), \frac{\pi}{2} < \theta < \frac{3\pi}{2} \quad (3\text{-}28)$$

$$= T_s(\theta) \qquad , -\frac{\pi}{2} < \theta < \frac{\pi}{2} \quad (3\text{-}29)$$

$$\alpha(\theta) + R(\theta) = 1 \quad (3\text{-}30)$$

where

$T_s(\theta)$ = sky temperature, K

T_o = thermal temperature of the earth, K

$\alpha(\theta)$ = earth's absorption coefficient, dimensionless

$R(\theta)$ = earth's reflection coefficient, dimensionless

When $-\pi/2 < \theta < \pi/2$, the antenna sees only the emissive sources above earth. These may be extraterrestrial sources such as the sun and the stars. $T_s(\theta)$ arises from the process of energy absorption and reradiation by components of the atmosphere. In the radio frequencies, oxygen and water vapor are the main absorbers, with water vapor being predominant. Fig. 3-13 (Ref. 9) shows a typical set of curves of antenna temperature for various angles θ, relative to the vertical, as a function of frequency. The peaks are due to the effects of water vapor. The variation with θ comes about because for small values of θ, the antenna looks into a thickness of atmosphere equal only to the height of the atmosphere (effectively about 10,000 ft), while at $\theta = \pi/2$ the antenna looks along the horizon seeing a much greater depth of atmosphere. At VHF and below, the contribution due

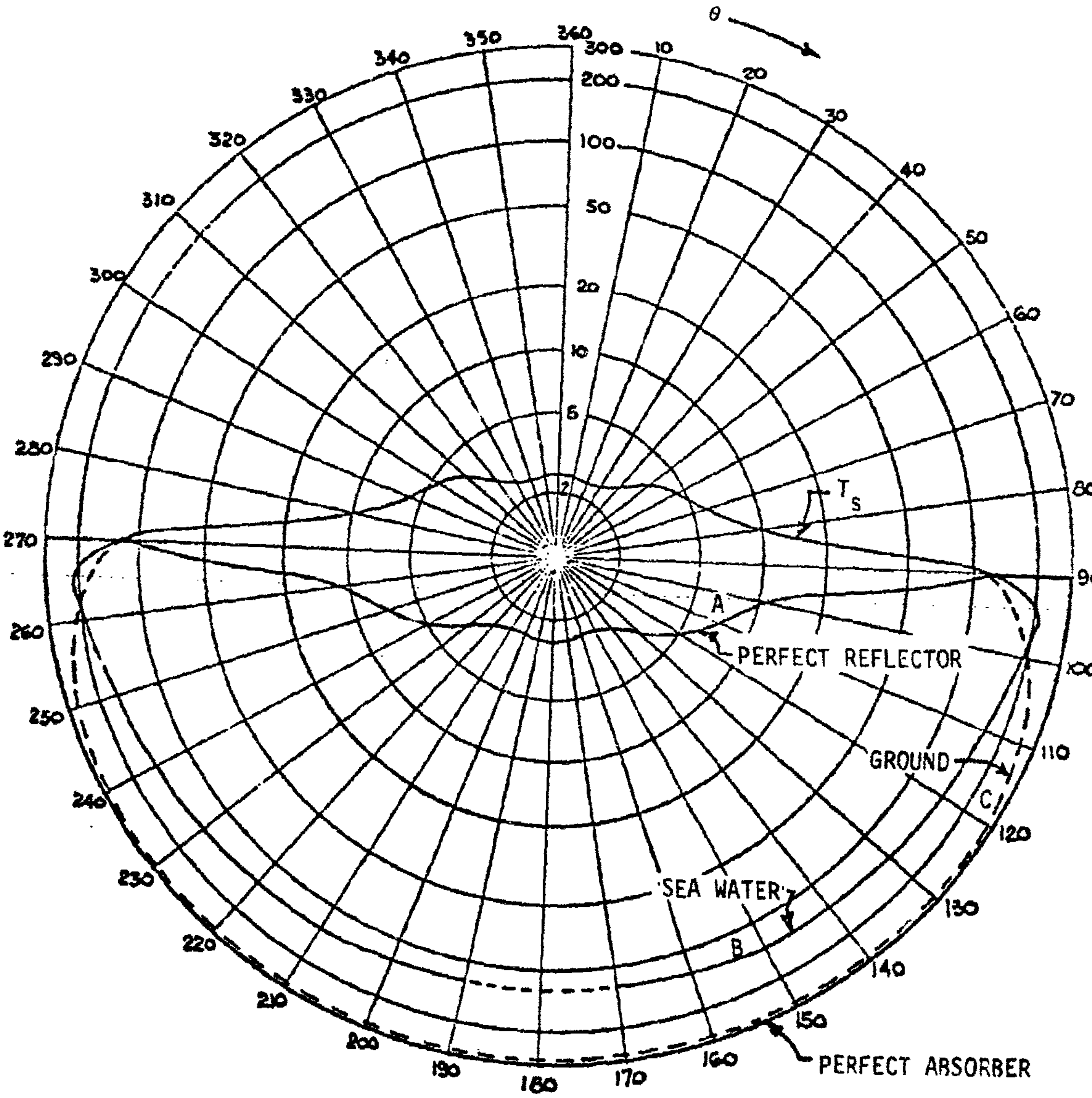


Figure 3-12. Brightness Temperature Distributions, Frequency—3 GHz (Ref. 9)

to these causes is negligible in nearly all applications.

When the antenna looks at the earth itself, it sees the sky sources partly reflected from the earth with contribution $R(\theta)T_s(\pi - \theta)$, and a part of the temperature of the earth $\alpha(\theta)T_o$. Because $\alpha(\theta)$ is close to unity over earth with vegetation, the brightness temperature seen by the antenna is very nearly the thermal temperature of the earth, as is indicated by Fig. 3-12. Because the sea is a poor absorber [$\alpha(\theta) \approx 0.5$], the brightness temperature seen by the antenna is much less. If the earth were a perfect reflector, the brightness temperature seen by the antenna looking at the earth would be a mirror image of the sky temperature. An antenna which in omnidirectional, or which sees symmetrically above and below the horizon, will, over ground, see a temperature of about

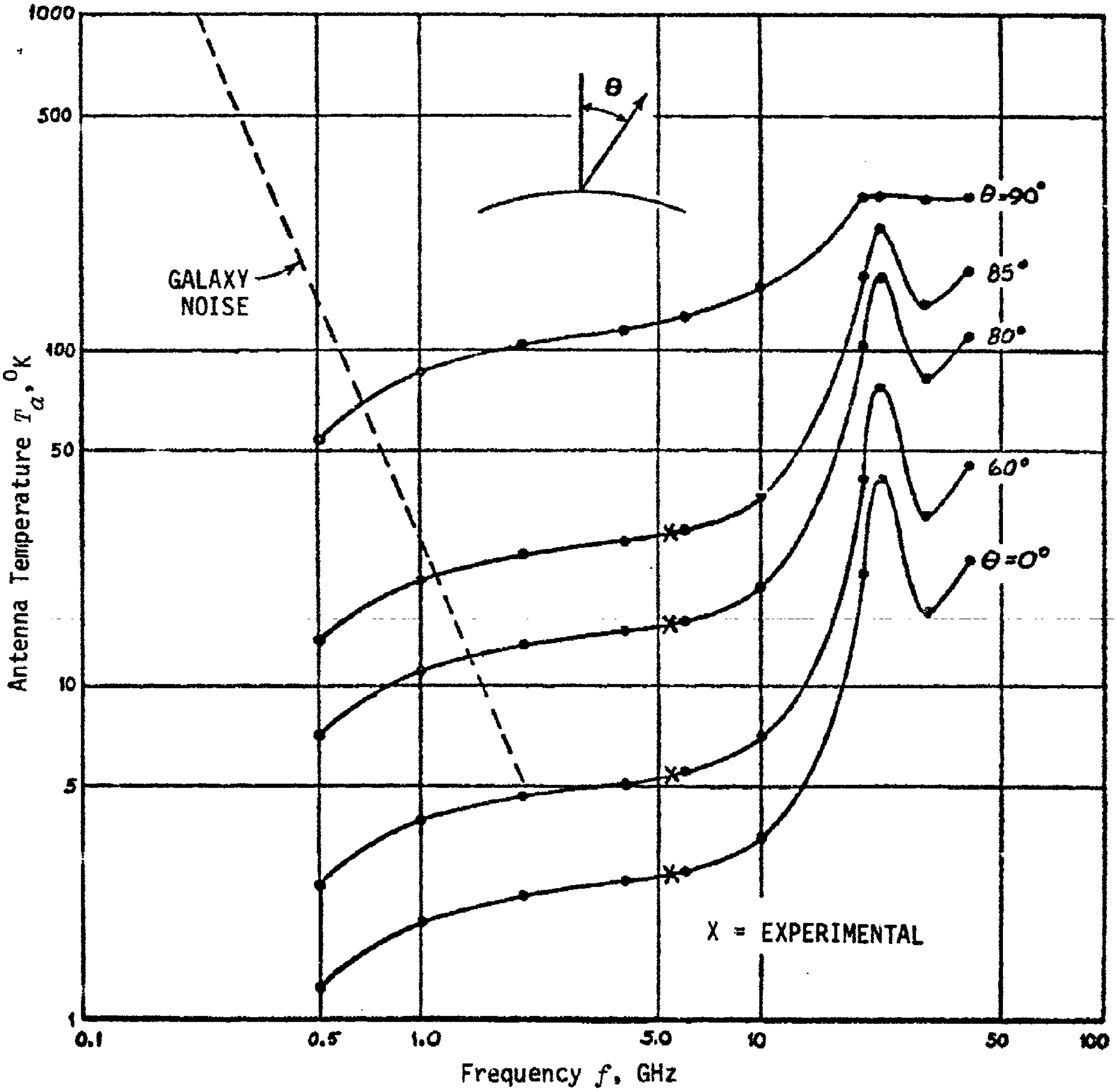


Figure 3-13. Antenna Temperatures Due to Oxygen and Water Vapor (Standard Summer Atmosphere)

150 deg. This is a consequence of the antenna seeing the ground in half of its pattern and the sky in the other half.

3-1.2.3 Extraterrestrial Noise

A receiving antenna on earth will be exposed to radiation emanating from sources outside the earth, as well as from the terrestrial sources discussed in par. 3-1.2.2.5. The sun and stars are the primary sources, with a minor contribution from nonemitters that act as reflectors. The temperature of the receiving antenna is given by Eq. 3-26, where the brightness temperature $T(\theta,\varphi)$ is the temperature seen by an ideal antenna, with an impulse pattern in the direction (θ,φ). The brightness temperature determined in this manner by an antenna on earth will include the effect of absorption in the atmosphere, and will, at some frequencies, poorly represent the source temperature.

3-1.2.3.1 Cosmic Noise

Cosmic noise refers to the electromagnetic emissions from bodies outside our own solar system. While some individual radio stars such as Cassiopeia A and Taurus are observable, the largest emissions come from our own galaxy. Fig. 3-12 shows the approximate brightness temperature estimated by Hogg (Ref. 9) looking into the center of our galaxy, for frequencies beyond 100 kHz. Fig. 3-6 shows the noise level estimated in CCIR Report 322 (Ref. 3) (refer to par. 3-1.3.2.3 for conversion from antenna noise figure to antenna temperature). These curves do not exactly correspond if they are linearly extrapolated; however, they are both based on an inverse square relationship between temperature and frequency. The curves show that significant effects due to cosmic noise are to be expected in the frequency range from about 20 MHz to about 500 MHz, with atmospheric noise dominating below 20 MHz, and various other noise sources such as thermal emission from the earth, and receiver noise, dominating above 500 MHz. When the antenna points away from the galactic center, the noise level falls off, and at the galactic poles might be 20 dB less.

3-1.2.3.2 Solar Noise

The sun is a significant contributor to receiver system noise, particularly when the antenna main lobe is directed close to or at the sun. A substantial contribution will occur, at times, even when side or back lobes are aimed at the sun. A table of the quiet sun brightness temperatures is given in Table 3-1 abstracted from Panter (Ref. 10). Because the solar disc subtends a solid angle of about 6×10^{-5} sr on earth (about 0.5 deg across the face of the disc), a receiving antenna with gain G across the disc (assumed to be approximately constant across the disc) will have a temperature contribution due to solar noise given by

$$T_a = \left(\frac{6 \times 10^{-5}}{4\pi}\right) G(\theta, \varphi)\, T_s \text{ , K} \qquad (3\text{-}31)$$

where T_s is the solar noise temperature as given in Table 3-1. At 1000 MHz, for instance, with $G(\theta, \varphi) = 1000(30$ dB) the antenna temperature arising from this source alone would be 1800 K. If a side lobe of 20 dB below the main lobe were pointing at the sun, a contribution of 18 K would be made to the antenna temperature.

During periods of high solar activity temperature increases of 10 to 20 dB may be found. (See, for example, Ref. 2, p. 27-4 Fig. 4 which shows such an increase when measurements are made with a dipole antenna.) Intense activity takes place at isolated regions on the solar disc, so that the brightness temperature within these regions is even greater. The regions of intense activity have varying lifetimes, change in size, and move. While the steady solar noise is unpolarized, the short intense bursts give rise to circularly polarized waves. Table 3-2 classifies the observed phenomena and their properties.

TABLE 3-1
SOLAR BRIGHTNESS TEMPERATURE
(Ref. 10)

Frequency f, MHz	Brightness Temperature T_s, K
100	10^6
200	9×10^5
300	7×10^5
600	4.6×10^5
1,000	3.6×10^5
3,000	6.5×10^5
10,000	1.1×10^4

3-1.2.4 Other Natural Noise Mechanisms

3-1.2.4.1 Triboelectric Noise

Triboelectric noise is noise generated as a consequence of frictional separation of charge. Noise impulses are generated when the charges separate, and when the accumulated charge on a body leaks off. Categorizing this kind of noise under natural phenomena is somewhat arbitrary, since the important manifestations of such noise are in applications involving machinery.

Frictional noise may occur between metals, between a metal and a nonmetal, or between nonmetals. The degree of seriousness is determined by the amount of friction encountered. It may occur between any two moving surfaces or the contact points of two charged surfaces. Examples of these phenomena are:

a. *Belt noise*. A charge between a dielectric (nonconducting) belt and its pulleys

b. *Bearing noise*. Occurs between bearings and their lubricant or housing

c. *Tire and track noise*. Appears between the tires or tracks of moving vehicles or tanks and the road

d. *Gear noise*. Results from two gears moving against each other.

Static charges on belts are a common occurrence in industry. The charges develop on both power-transmission and conveyor belts. Low temperatures

TABLE 3-2. SOLAR EMISSIONS

Class Characteristic	Basic Component (Thermal)	Slowly Varying Component	Short Term Interference	Unpolarized Burst: Outburst	Unpolarized Burst: Isolated Burst
Wavelength range	Unlimited	3-60 cm	1-15 m	1 cm-15 m	50 cm-50m
Duration		Weeks or months	Hours or days	Minutes	Seconds
Polarization	Random	Trace of circular	Strongly circular	Random	Random
Place of origin	Whole sun	Number of small areas	Small area above sunspot	Small area rapid movement	Small area
Associated optical features		Sunspots and others	Large sunspots	Flare	Unknown
Remarks	Constant over years	27-day component	With or without numerous bursts	No certain distinction	No certain distinction

appear to be more favorable for the accumulation of charge, although the effect may become serious in dry atmospheres at any temperature. Noise arising from vehicular tires is quite evident in mobile transmitters, and is very pronounced when a vehicle is traveling on paved roads. When two gears made of similar metals mesh, little noise is generated; when two gears of dissimilar metals mesh, noise in the form of pulses at the beginning of motion appear. The electrolytic action between two different metals causes an electrical discharge to occur as the gears mesh.

Frictional (Ref. 11) charge separation is encountered in aircraft. It occurs between the aircraft and particles in the air, or between the aircraft and nonconductive liquids such as fuel flowing into the tanks of the aircraft. The phenomenon is of importance in dry weather, when considerable charge can be accumulated without the opportunity to dissipate itself slowly through leakage. Large amplitude noise is generated when a substantial charge is accumulated, giving rise to high field intensities at sharp points on the aircraft exterior, and attendant corona discharge.

Frictional charging can take place on the ground, for instance in a dust storm, or in the air under a variety of conditions. The particles involved include snow, ice crystals, sand, dust, smoke, and exhaust system particles from lead aircraft in tight formation flying. The type of particle present controls the resulting polarity of charge, to some extent.

Dry snow impinging upon aircraft almost always produces a negative charge, as has been demonstrated in the laboratory as well as in flight. Air temperature controls the rate of charge, with the rate becoming maximum at about −10°C. Snow conditions prevail in any weather at intermediate altitudes, to produce severe interference. Fine ice crystals encountered at high altitudes, or the form described as ice spicules composing cirrus clouds, occurring generally above 30,000 ft, produce equally severe interference at all seasons.

Sand, dust, smoke, and exhaust particles generate charges, but the polarity depends upon aircraft finish air temperatures, and atmospheric charge centers The most commonly used aircraft paints and waxe lead to a negative charge on the aircraft at air tem peratures between −5° and −15°C. On the othe hand, a finish using titanium dioxide (TiO_2) or a pig ment of colloidal silica in cellulose nitrate, generate a positive charge on the aircraft at air temperatur between 0° and −10°C. The polarities cannot b predicted reliably at temperatures higher or low than those indicated. Experience has shown th clean bare aluminum is the most neutral materi over the widest temperature range for all types of pa ticles encountered. Even nonconductive surfaces su as radomes develop a local charge accompanied by a type of discharge known as streamering, unless they are protected with a thin, somewhat conductive coating, that is properly grounded to the aircraft metal structure.

Helicopters, while hovering in snow, sand, or dust, may experience frictional charging. Potentials from − 200,000 V to + 60,000 V have been recorded on different occasions which — depending upon the type of aircraft, the windborne particle material, and the atmospheric field gradients — can reach 5,000 V/m or more causing incipient lightning conditions. To neutralize these potentials, active static discharges in the form of high voltage power supplies capable of delivering at least 100 μA at either polarity as the occasion demands have been installed in helicopters.

3-1.2.4.2 Precipitation Static

Closely related to the frictional charge separation on aircraft discussed previously, is the phenomena of precipitation static—also referred to as p-static—which arises in wet weather conditions. Two different kinds of phenomena are encountered which give rise to charge separation and discharge. These are:

a. Frictional charge separation as the aircraft passes through uncharged raindrops, leaving the raindrop positively charged and the aircraft negatively charged. The rate of charge accumulation depends here on the density of raindrops, being directly proportional to the weight of material struck.

b. Charge induction in the airframe arising when the aircraft passes through the electric field of weather makers. It has been explained that, as a raindrop falls through the air, the friction between the air and the raindrop causes a physical separation of a negatively charged mist, which remains hanging in the cloud, thus leaving the falling drops positively charged. This is the fundamental mechanism whereby clouds are said to acquire charge. When the aircraft passes through the field between these regions, charge reorientation takes place on the aircraft by the mechanism of static induction. Corona discharges into the air occur at sharp points of the airframe, and into dielectrics which are part of the airframe. This is a consequence of the large fields that build up at sharp points on the surface of a charged conductor. Furthermore, different parts of a cloud, and different clouds, carry different amounts of charge; hence, fields exist within and between them and induce charge on the airframe.

While the breakdown potential gradient at sea level may be 30,000 V/cm, its value may be reduced to lower values at higher altitudes. Both the air density and the surface density of charge are parameters that influence the breakdown potential gradient. Air density obviously decreases as the altitude above sea level increases. It also becomes less at high-lift areas, such as along the upper surface of wings and at propeller blades. Surface density of charge depends upon the geometry of the material immersed in the field.

Once breakdown has occurred, a discharge will follow. Impulsive and continuing discharges, rich in both AM and FM components and their harmonics, are generated in the 0.1- to 400-MHz range, and occasionally beyond. Direct pickup by a receiving antenna may result in a received level as high as 100 μV/kHz; indirect pickup by way of the conduction of interference from the source to receivers via antennas, power lines, or other electric circuits in the aircraft may also result in significant effects. Special consideration is required for communication and navigation equipment in the low-frequency (LF) and medium-frequency (MF) bands because of their particular susceptibility to p-static interference. For manned aircraft, the problem is especially severe because this type of interference occurs during times of low visibility, when the pilot must fly by instruments, using voice communication and radio navigation aids.

A number of remedies for reducing p-static have been developed. One obvious method is to minimize corona by avoiding sharp points. Another method is to shield and ground sufficiently to minimize pickup. A third method is to use static dischargers, active or passive.

Further information on triboelectric noise and p-static will be found in Refs. 12 through 14.

3-1.3 MAN-MADE SOURCES

Man-made sources of electromagnetic energy may be intentional, such as radio transmitters, or incidental, as in the case of ignition noise or local oscillator radiation*. In considering levels of electromagnetic energy at a given location, it is important to recognize the fact that, while an incidental emitter can be a significant source, intentional radiation can be equally or more important.

A breakdown of man-made sources, listing examples of both broadband and narrowband types, is given in Table 3-3 (Ref. 15). It has been further subdivided to give a general indication of whether each source is active for long periods of time (continuous), for short periods (intermittent), or for very short and

* This description is not entirely consistent with that of FCC which places these two devices into two separate classes, incidental radiation devices and restricted radiation devices — see Chapter 2.

TABLE 3-3. TYPICAL MAN-MADE INTERFERENCE SOURCES (Ref. 15)

Broadband			Narrowband	
Transient	Intermittent	Continuous	Intermittent	Continuous
Mechanical function switches Motor starters Thermostats Timer units Thyratron trigger circuits	Electronic computers Motor speed controls Poor or loose ground connections Arc welding equipment Electric drills	Commutation noise Rectifiers Ignition systems Arc and vapor lamps Pulse generators Pulse Radar transmitters Sliding contacts Teletype-writer equipment Voltage regulators	CW-Doppler radar Radio transmitters and their harmonics Signal generators, oscillators, and other types of test equip-ment Transponders Diathermy equipment	Power-line hum Receiver local oscillators

infrequent time periods resulting from solitary-event switching action (transient). Extremely short duration transients are considered to be impulsive. Transients from repetitive switching action — such as occur in SCR's, printers, and fluorescent lamps — produce waveforms containing harmonics of the switching rate. Harmonics also may result from the application of a sine wave to a nonlinear device.

3-1.3.1 CW Sources

The waveforms included in this category are:

a. Sine waves, pure or modulated (except those modulated with very narrow pulses)

b. Periodic signals, whose harmonics are spaced far enough apart that no two of them appear in the acceptance bandwidths of potential susceptors.

3-1.3.1.1 Transmitters

Potentially interfering signals generated by transmitters include intentional emissions, harmonically related spurious emissions, and nonharmonically related spurious emissions.

The minimum bandwidth requirements for transmitter emissions are determined by the characteristics of the functional signal and the modulation method employed. In an amplitude-modulated wave, the width of the spectrum occupied is twice that of the highest modulation frequency. In frequency modulation, the bandwidth is approximately equal to twice the sum of the highest modulating frequency and the peak frequency deviation. With pulse modulation, the bandwidth is a function of the rise time associated with the pulse. Table 3-4 (Ref. 16) summarizes the bandwidth requirements of a variety of

TABLE 3-4. TYPICAL BANDWIDTH REQUIREMENTS (Ref. 16)

Transmission	Information Bandwidth, Hz	AM, Hz	Single-sideband, Hz	Narrowband FM ($m_f > 0.5$), Hz	Wideband FM ($m_f = 10$), Hz
Telegraph					
Morse Code 100 words per min	0-120	240	120	240	2,400
Speech					
High fidelity	40-15,000	30,000	15,000	30,000	300,000
Typical broadcast program	100-500	10,000	5,000	10,000	100,000
Long distance telephone quality	250-3500	7,000	3,500	7,000	70,000
Intelligible but poor quality	500-2000	4,000	2,000	4,000	40,000
Television					
Standard 525-line picture interlaced, 30-cycle repetition rate	$60\text{-}4.5 \times 10^6$	9×10^6	4.5×10^6	9×10^6	90×10^6
Pulse					
1 μs long	$0\text{-}1 \times 10^6$	2×10^6	1×10^6	2×10^6	20×10^6

typical signals for each of the applicable types of modulation.

Spurious emissions are signals emitted at frequencies outside the "necessary" bandwidth of the generating source, the levels of which may be reduced without affecting the quality of the intentional transmission. Spurious emissions include:

a. Harmonics of the transmitter fundamental frequency f_o

b. Nonharmonically-related outputs

c. Sideband splatter and noise.

Fig. 3-14 shows a simplified block diagram of a communication transmitter that uses conventional vacuum tubes, a master oscillator, and a multiplication scheme to generate the fundamental output. A typical output spectrum is shown in Fig. 3-15 (Ref. 16). In addition to the normal harmonic outputs, there are outputs that occur at a multiple of a frequency lower than the fundamental. These outputs usually are harmonics of the master oscillator used to generate the transmitter carrier frequency.

Fig. 3-15(B) shows the output spectrum that may be obtained from a system that uses a klystron for the final power amplifier. In this case, the output spectrum consists only of the fundamental, and harmonics of the fundamental. The output spectrum from a typical radar transmitter with a magnetron output tube is shown in Fig. 3-15(C). In this spectrum, in addition to the harmonics of the fundamental, there appear additional spurious outputs that do not bear any definite frequency relationship to the fundamental or to each other. These are usually attributed to multimoding phenomena associated with the multiple-cavity resonator that is used in the device. Although measures are taken to suppress the undesired modes, it is not possible to eliminate them completely.

Sideband splatter occurs in amplitude-modulated transmitters when the modulation exceeds 100 percent, and when the carrier is cut off on the negative modulation peaks. In single-sideband transmitters, splatter is caused most often by overdriving the power amplifier, so that it operates in a nonlinear region. With frequency-modulated transmitters, over-modulation causes the frequency swing to exceed the design maximum system deviation. Transmitters also radiate broadband noise. Although the level is quite low, it is sometimes high enough to produce interference with colocated receivers.

The emissions which have been discussed can be radiated not only from the antenna but also from the transmitter equipment cabinet itself, and also can be conducted along any signal or power lines connected to the cabinet.

Figure 3-14. Basic Parts of CW Transmitter

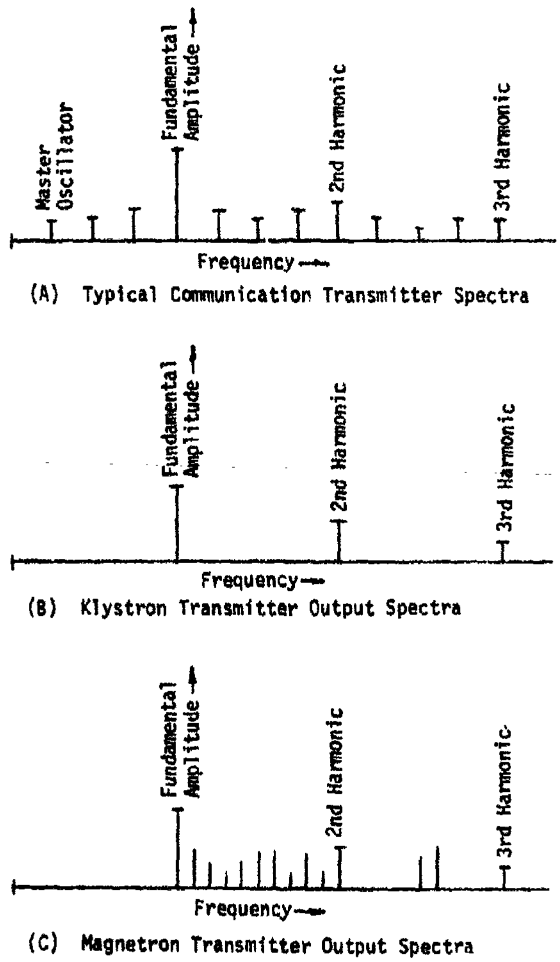


Figure 3-15. Typical Transmitter Output Spectra

When the transmitter cabinet is located close to the radiating antenna, direct leakage from the cabinet may be no more troublesome than that from the antenna. When the transmitter is in a shielded enclosure or a metallic building which offers shielding from the antenna, direct leakage from the transmitter cabinet may be the primary source of interference and must be controlled where sensitive equipment is present in the enclosure.

3-1.3.1.2 ISM Devices

A class of equipment that generates RF energy for noncommunication purposes but for which the levels can be as large as, or larger than, those used for communications is known as "Industrial, Scientific, and Medical" (ISM) equipment. ISM equipment generates sine wave signals at frequencies in the range from 10 kHz to 30 GHz, usually for the purposes of cleaning, heating, or plasma stabilization. Equipment designed for civilian use is subject to special limitations on emissions by Part 18 of FCC Rules and Regulations as described in Table 2-5.

Considerable data exist on measured emission levels from such devices. Pearce and Bull (Ref. 17) measured levels of emissions from wood gluers, plastic welders, and preheaters which appear at numerous frequencies between 30 MHz and 1 GHz. At a distance of 1000 ft the levels varied from 0 to about 60 dB (μV/m).

Garlan and Whipple (Ref. 18) measured emissions from arc welders, also at 1000 ft. In this case the emissions are broadband, but on any one emitter vary considerably with frequency over the range from 0.5 MHz to 30 MHz. Levels varied from −5 to about 38 dB[μV/(m·kHz)].

3-1.3.1.3 Local Oscillator Emissions

Local oscillators, used for heterodyning or detection in receivers, are potential sources of emissions via conduction on the power line or radiation either directly from the chassis or via a connected antenna.

In digital computers, clock oscillators are used to generate repetitive pulses to trigger logic circuits in synchronism. The associated emissions can be rich in harmonics, and can be radiated and/or conducted.

Some typical conducted and radiated narrowband emissions from receivers, transmitters, and a computer are shown in Figs. 3-16 and 3-17 (Refs. 19 and 20). The emissions shown below 10 kHz in Fig. 3-16 are power frequency harmonics, not related to local or clock oscillator emissions.

3-1.3.2 Switching Transients

Transients occur whenever electrical power suddenly is applied to or removed from a load. The power may be direct or alternating and the switching

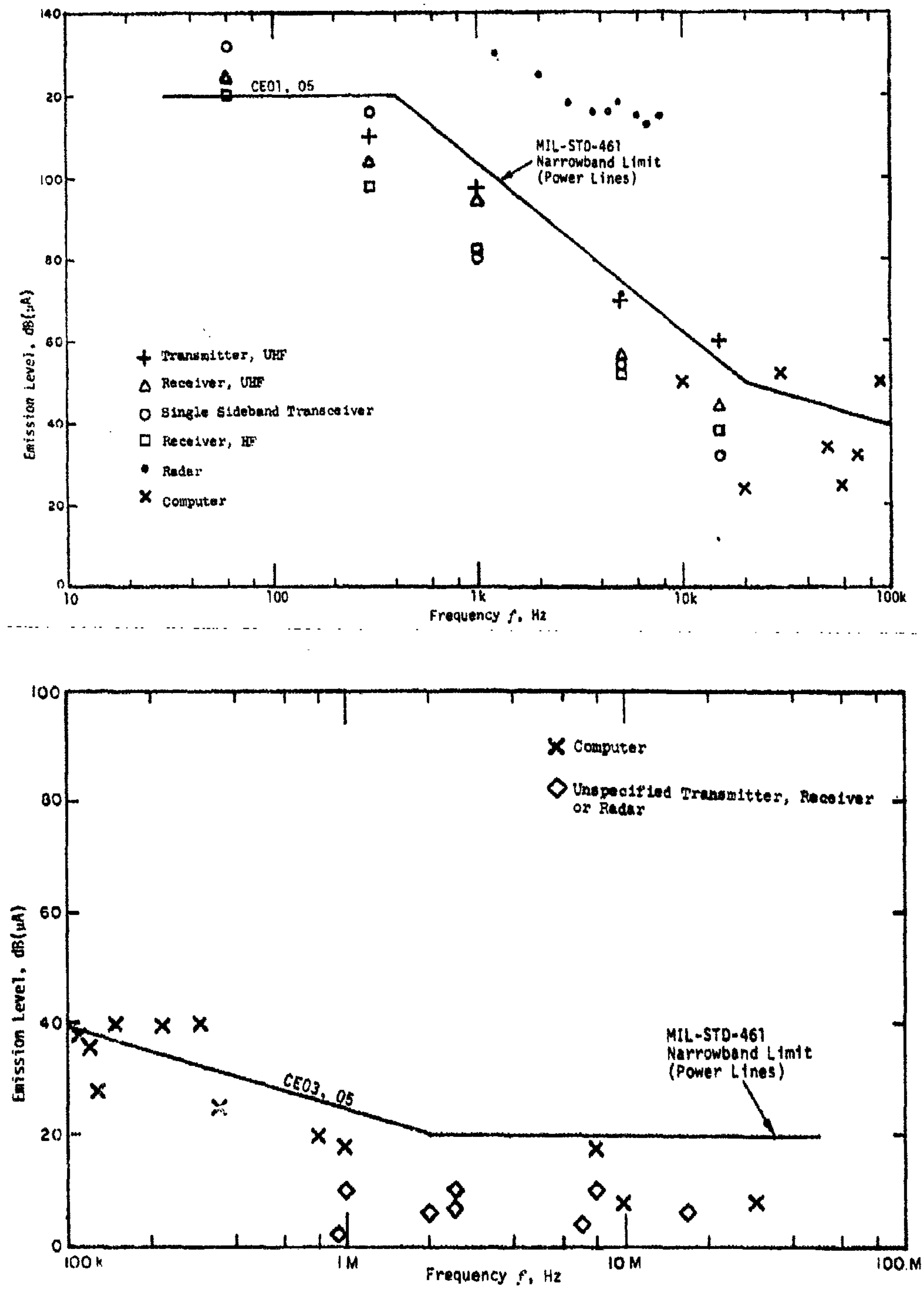


Figure 3-16. Typical Conducted Narrowband Emissions (Ref. 19)

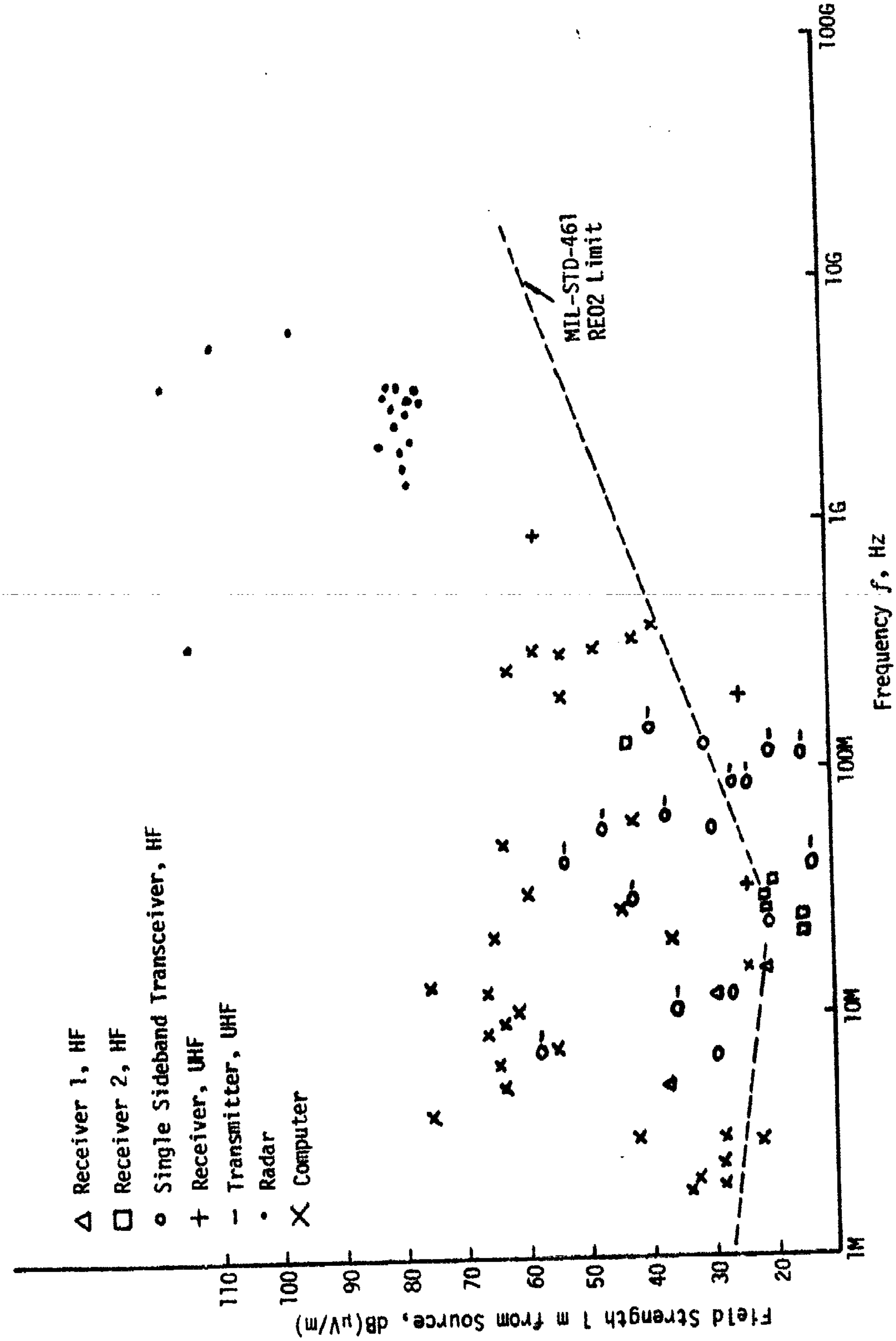


Figure 3-17. Typical Narrowband Electric Field Levels (Ref. 20)

action may be intentional, as in the case of a mechanical or electronic switch; or it may be unintentional, as in the case of intermittent contact due to a faulty mechanical connection between two parts of a circuit. The prototype switching action is to initiate or interrupt a steady current or voltage. A simplified circuit is shown in Fig. 3-18. In this figure, Z_s represents the generator impedance, and Z_L the load impedance. At high frequencies the lengths of the lines connecting the switch S to the source and the load are important.

3-1.3.2.1 Switching Action

At low frequencies if the impedances are purely resistive, the switch can be considered to function instantly, to establish (upon closing) a voltage across the load impedance or a current through it. The mathematical model for this action is the unit step (see Fig. 3-19) which has a spectrum amplitude $S(f)$

$$S(f) = 1/(\pi f)\ ,\ \mathrm{s} \qquad (3\text{-}32)$$

For a step of amplitude A (A nondimensional)

$$S(f) = A/(\pi f)\ ,\ \mathrm{s} \qquad (3\text{-}33)$$

For a step of amplitude V volts or I amperes, the spectrum amplitude is

$$S(f) = V/(\pi f)\ ,\ \mathrm{V{\cdot}s} \qquad (3\text{-}34)$$

or

$$S(f) = I/(\pi f)\ ,\ \mathrm{A{\cdot}s} \qquad (3\text{-}35)$$

The spectrum for step S_s of one volt is plotted on Fig. 3-20. Also shown on this figure are formulas showing how this spectrum is modified for an amplitude of A volts; i.e., for a 10-V step the line shown should be raised 20 dB. By changing the ordinate dimensions to dB (μA/MHz), the same figure can be used for a current step.

An actual switch and circuit differs from the ideal prototype in that the current or voltage is not immediately initiated or interrupted due to imperfect switching action, or because the circuit in which the contact is placed may have reactive or unmatched loads. Frequently the equivalent load or the connecting transmission line may be considered to contain an equivalent series inductance.

On closing such a circuit, the voltage $v(t)$ across the load resistor rises exponentially according to the relation

$$v(t) = V[1 - \exp(-t/\tau)]\ ,\ \mathrm{V} \qquad (3\text{-}36)$$

where

$\tau = L/R$, s
L = inductance, H
R = resistance, Ω
V = final voltage, V
t = time, s

The spectrum amplitude is given by the expression

$$S(f) = \frac{V}{\pi f[(2\pi f\tau)^2 + 1]^{1/2}}\ ,\ \mathrm{V{\cdot}s} \qquad (3\text{-}37)$$

Note that at frequencies for which $2\pi f\tau \ll 1$, the spectrum is identical to that of the step. For some purposes, this spectrum can be approximated by that of a linear rise of voltage to the maximum value V. It is given by the expression for S_r shown on Fig. 3-20 (Ref. 15). Its envelope (substitute 1 for sin $\pi f\tau$) is plotted for $t = 0.1$ μs, $A = V = 1$ V. Note that for frequencies above $1/(\pi t)$, the spectrum amplitude falls off as $1/f^2$ rather than $1/f$, and that in this range, it is 6 dB above that for an exponential rise with $\tau = t = 0.1$ μs.

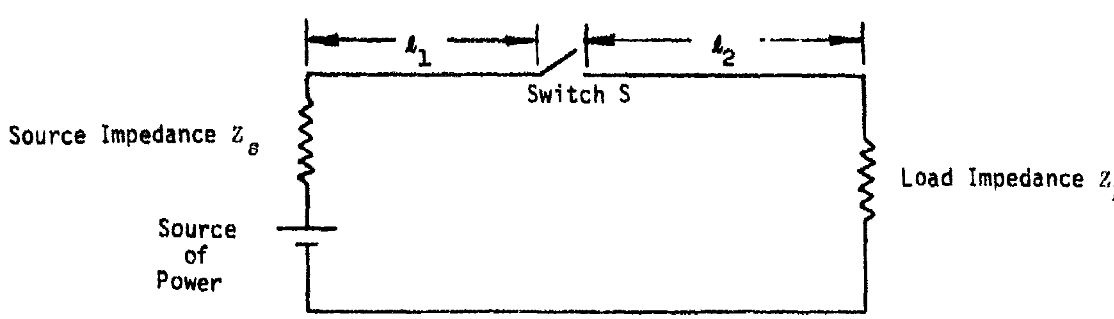


Figure 3-18. Simplified Switching Circuit

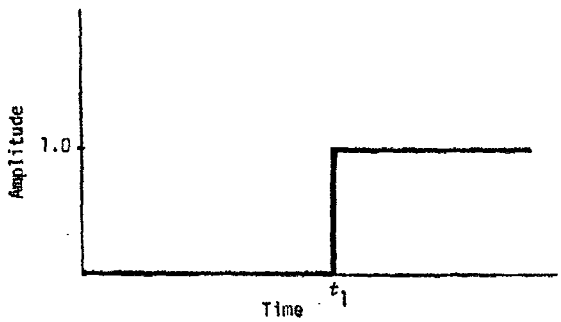


Figure 3-19. A Unit Step at $t = t_1$

For comparison purposes the spectrum of a trapezoidal pulse having amplitude unity, rise and fall times t, and duration d is shown on Fig. 3-20 (see formula and plot for S_p). In the portion of the spectrum where the amplitude varies as $1/f$, the spectrum of the unit step is 6 dB below that of the unit amplitude pulse. The dependence on the rise time for both is similar, but the pulse spectrum levels off at frequencies below $1/(\pi d)$.

3-1.3.2.2 Arcing Phenomena

When a switch opens, the current tends to change rapidly from a finite value to zero, and the large resulting di/dt can produce a large instantaneous

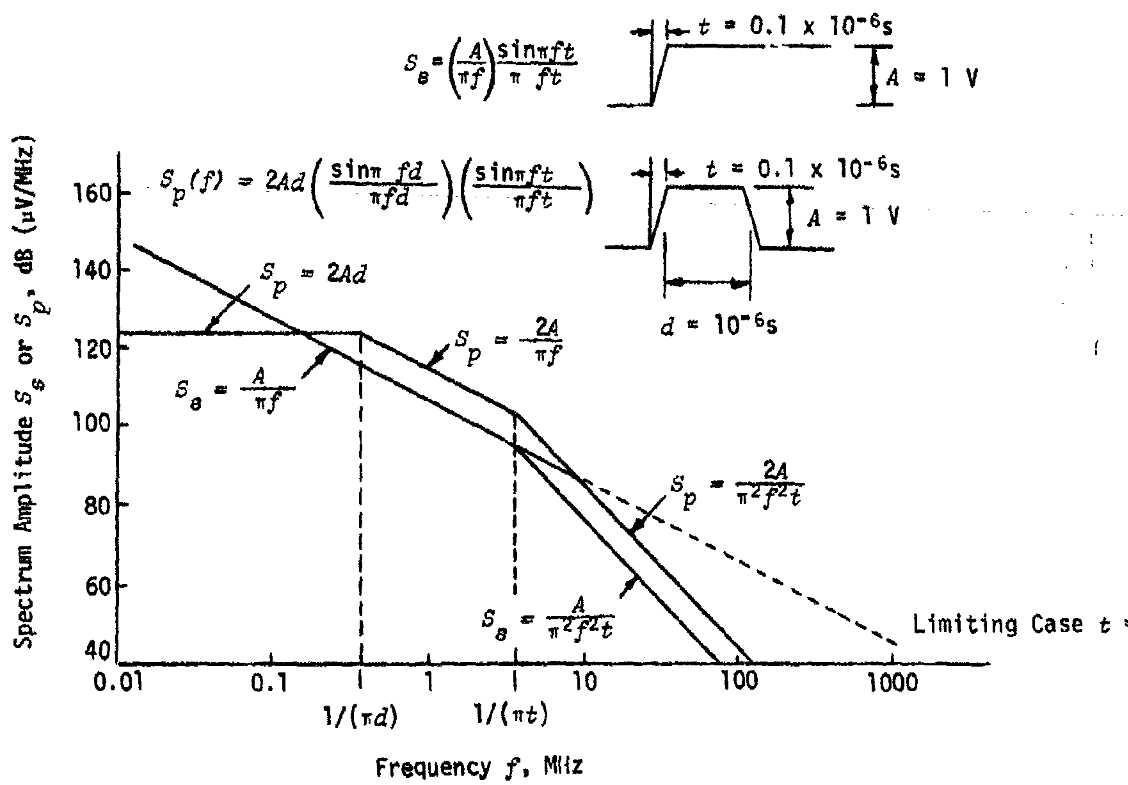


Figure 3-20. Interference Level for a 1-V, 1-μs Trapezoidal Pulse and 1-V Unit Step (0.1 μs rise time) (Ref. 1!

voltage pulse across any circuit inductance. The large voltage easily can cause an arc to be struck across the switch contacts, so that current continues to flow. Rüdenberg (Ref. 21) shows that if the contacts open in such a way that $1/R_c$ decreases linearly with time — R_c is the contact resistance — to a value 0 in time τ, then the peak voltage V_{pk} appearing across the contacts (see Fig. 3-21(A)) is

$$V_{pk} = \frac{V}{1 - L/(R_c \tau)}, \text{V} \qquad (3\text{-}38)$$

where

τ = switching time of contact, s
L = circuit inductance, H
R_c = contact resistance (initial value), Ω

For typical circuits, τ is greater than L/R_c and V_{pk} can be substantially larger than V.

If the circuit had distributed an intentional capacitance across the inductance, as shown in Fig. 3-21(B) (Ref. 21), the arcing can be reduced or effectively eliminated. In this case the voltage may become oscillatory, and the maximum voltage V_{max} is

$$V_{max} = I(L/C)^{1/2}, \text{V} \qquad (3\text{-}39)$$

With resistances in the branches (Fig. 3-21(C)) oscillation is eliminated if $R_1 = R_2 - (L/C)^{1/2}$.

With practical snap switches, on "make" the contacts may bounce apart several times before finally settling together; while on "break", the voltage spike generated may cause an arc to develop across the contacts which may be extinguished and restruck several times before the contacts are far enough apart to prevent it. The net effect of arcing and contact bounce is that the switch opening or closing is accompanied by not one but several rapid changes in current, and the overall switching transient voltage is actually a series of several pulses.

Typically, bounces and arcs may occur on the order of milliseconds apart, while the duration of the voltage "spikes" may be of the order of microseconds (see Fig. 3-22, Ref. 22). Such a "burst" of interference may be serious particularly in digital systems, where several data bits may be corrupted.

The effects of multiple arcs or oscillations on the emission spectrum is to increase the spectrum amplitude at the arc or oscillation frequency. The low frequency portion of the spectrum is determined only by the magnitude of the current or voltage change. Enhancement at one frequency can be expected to reduce the spectrum at higher frequencies.

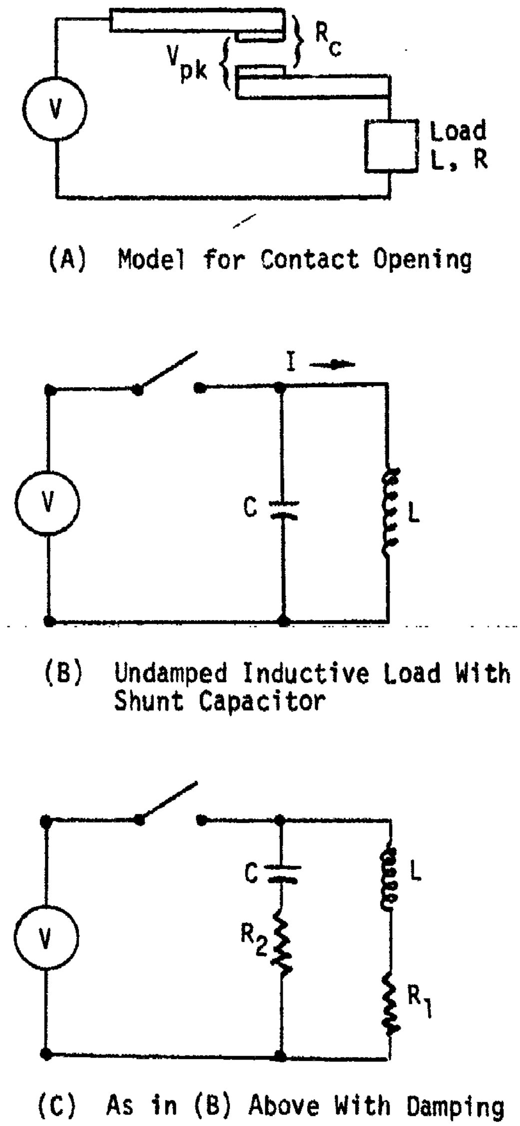
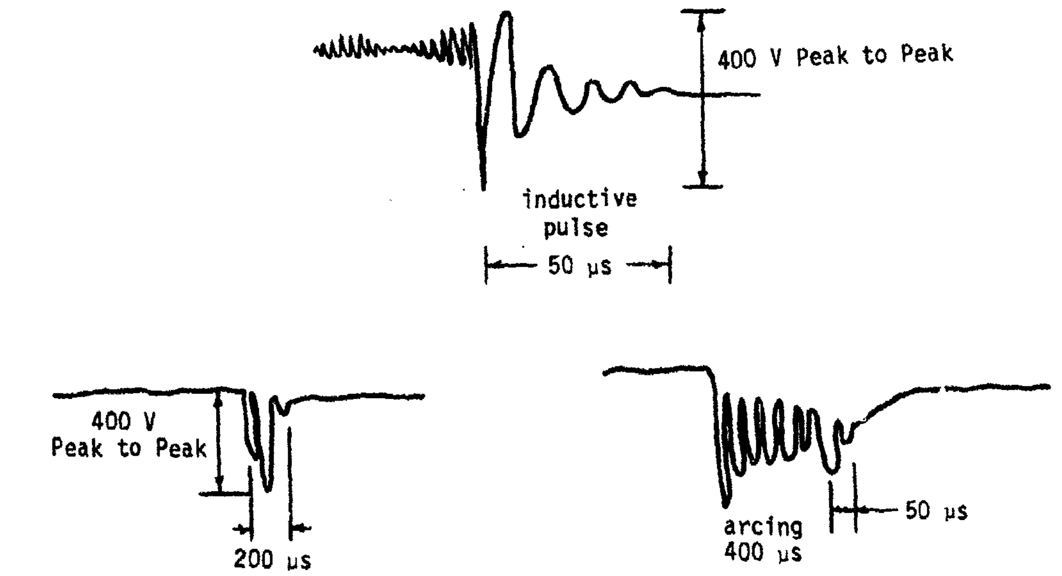


Figure 3-21. Contact Arc Suppression (Ref. 21)

Transients on power lines can be particularly severe, due to the large currents and inductances that may exist in a given circuit. Observed levels of such transients are as high as several hundred volts on a 110-V line (Ref. 23). Observed duration may vary widely, on the order of 1-100 μs, but fractional microsecond pulses also occur. The dominant component of power line inductance may be that of the line transformer, which may be of the order of 10-100 μH.

(A) EMI* Created by Toggle Switch



(B) EMI Generated by Electronic Switch (C) EMI Generated by Six Pole Relay Opening

*EMI = electromagnetic interference

Figure 3-22. Typical Switching Transients (Ref. 22)

Although spikes on power lines usually are generated by the switching of differential mode current, they may be coupled readily into common mode or ground loop circuits, thereby considerably increasing the effectiveness of the power line as a radiator. This situation is discussed further in par. 3-3.3.3.

Transients are generated in circuits containing relays both by opening and closing of the contacts of the relay itself, and by operation of the activator coil, usually by other contacts. Transients also appear in circuits containing solenoid valves. The relay coil or solenoid, as an inductance, generates a voltage spike when deactivated. Although the current in the coil may be small compared with that switched by the contacts of the relay, the voltage spike generated by the coil circuit may be comparable in level, due to the large inductance of the coil compared with power circuit inductance. Transients associated with relay or solenoid operation, other than those caused by the coil, are similar to those associated with mechanically operated switch circuits.

Representative spectra are shown on Figs. 3-23 and 3-24 (Ref. 24). On Fig. 3-23, peak conducted EMI levels measured in a circuit containing a relay coil are shown as a function of the magnitude of steady current after contact. The levels are seen to be a function of the current level and indeed, except at very low currents, directly proportional to the current. At high currents, the rate of fall off of spectrum is approximately inversely proportional to frequency, in accordance with theory, as is its magnitude (see Fig. 3-20). At lower currents the increase in spectrum level with frequency is not understood, but may be caused by decreased damping of the circuit inductance by increased load resistance.

Fig. 3-24 shows corresponding results for contact "break". Note that here the spectrum is better behaved, probably because the load resistance has less effect in an open circuit.

Measured levels of EMI radiated from the circuit containing the relay coil are shown on Fig. 3-25 (Ref. 24). No attempt is made to provide a radiation model

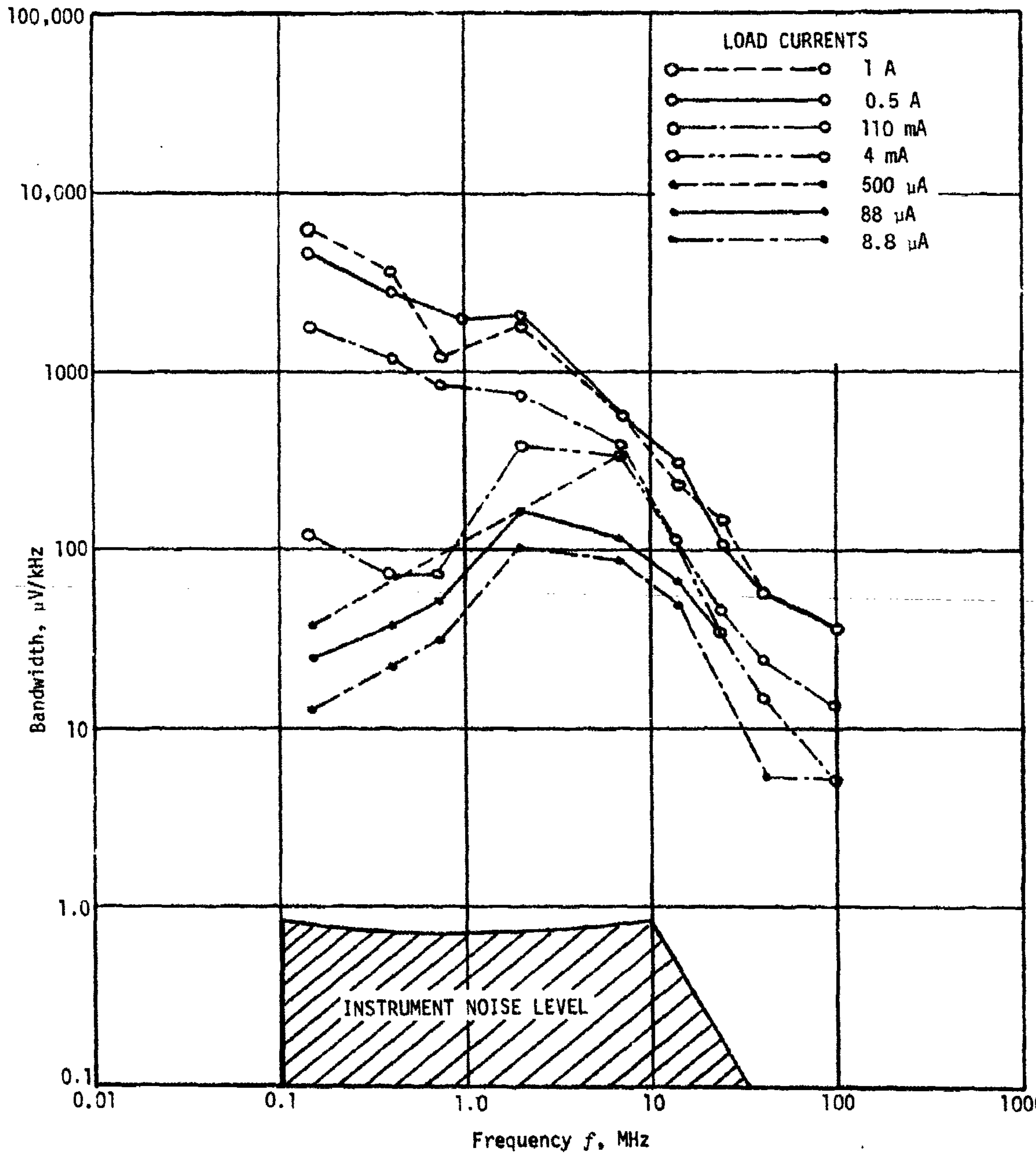


Figure 3-23. Make Contact Conducted EMI Associated With Current Magnitudes in an Electromagnetic Relay Contact Circuit (V_{load} = 6.0 V dc; Z_{load} = R) (Ref. 24)

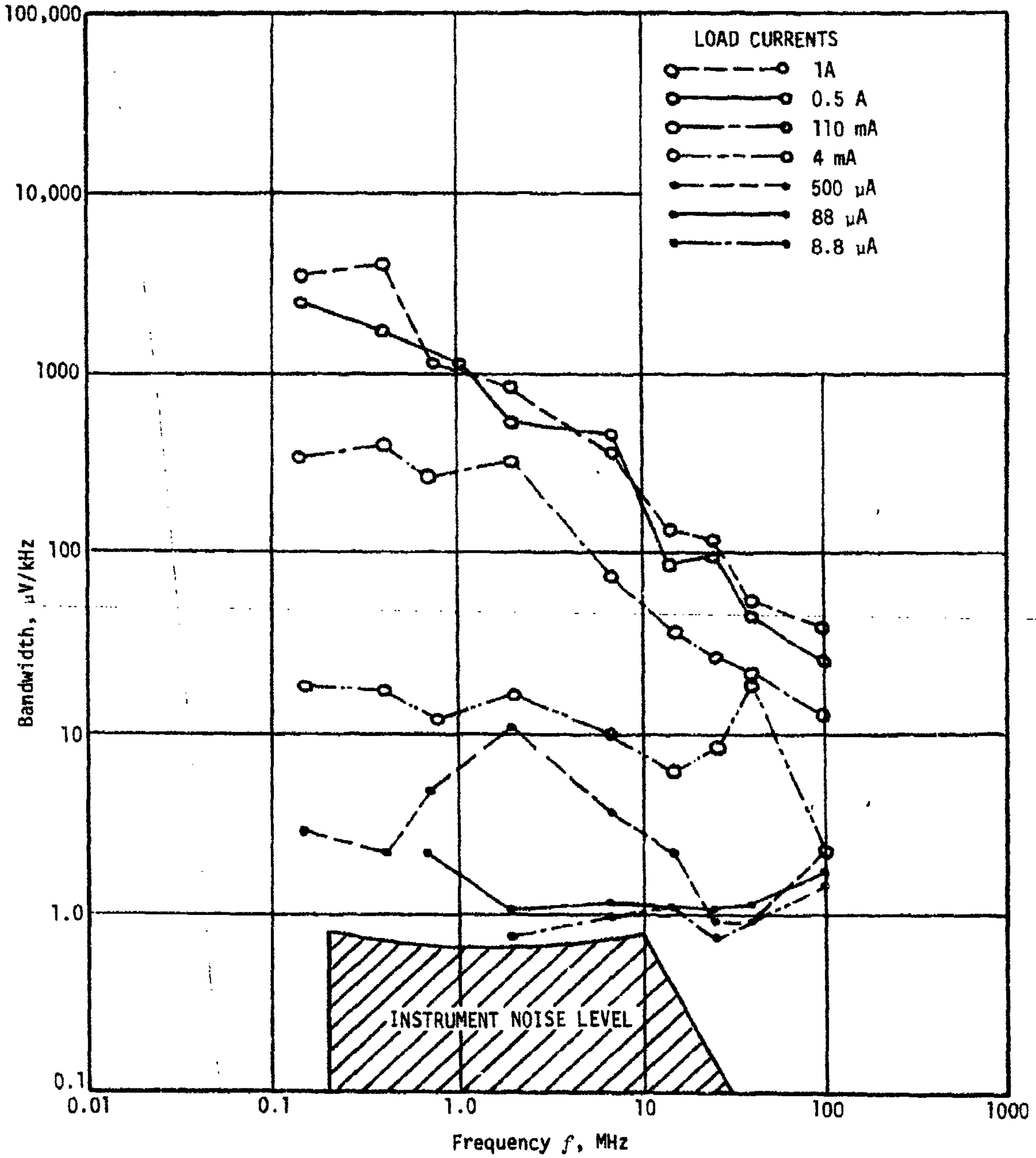


Figure 3-24. Break Contact Conducted EMI Associated With Current Magnitudes in an Electromagnetic Relay Contact Circuit
(V_{load} = 6.0 V dc; Z_{load} = R) (Ref. 24)

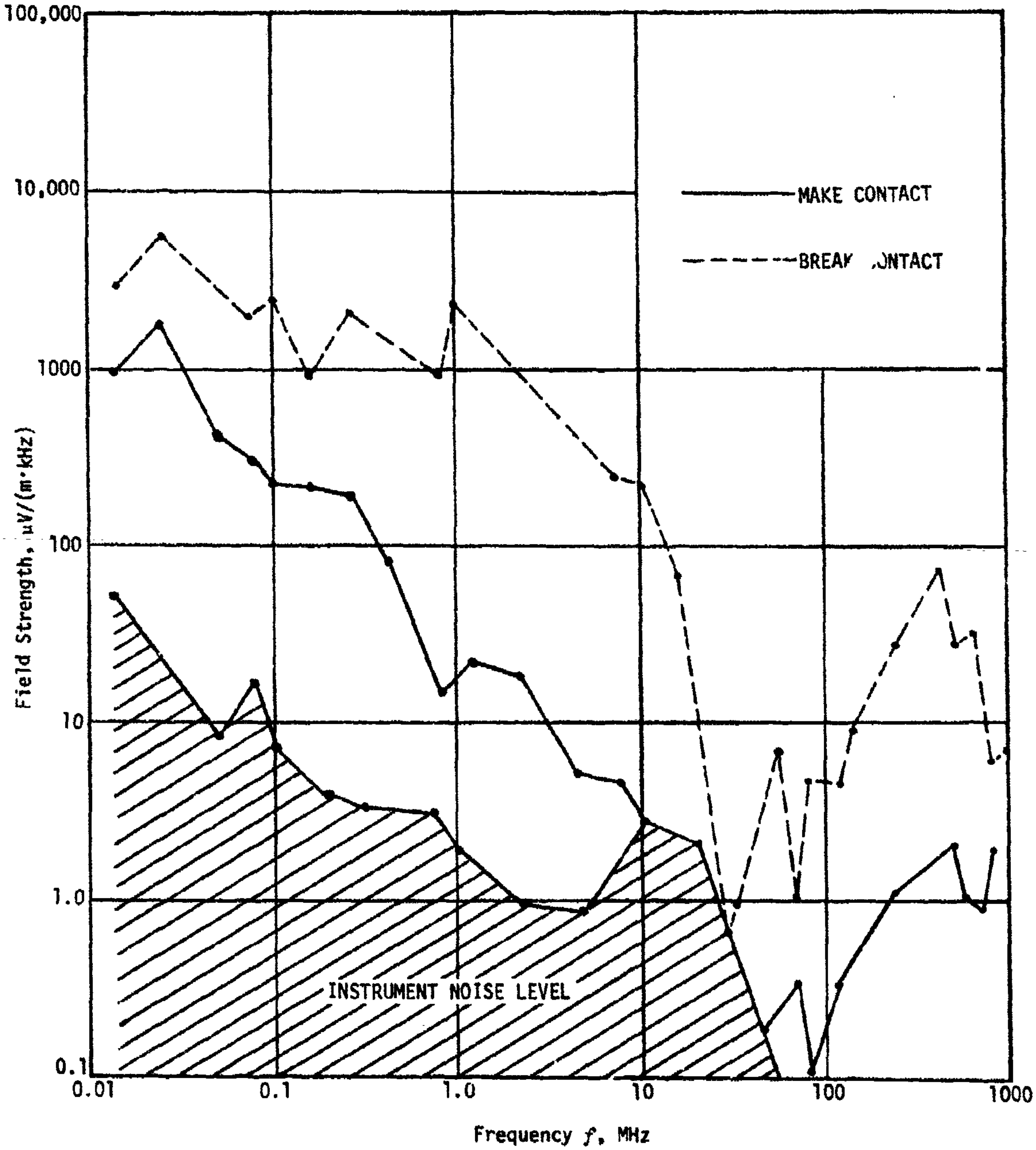


Figure 3-25. EMI Radiated from a Conventional Electromagnetic Relay Coil (no relay cover) (Ref. 24)

here because of uncertainty as to the actual configuration of the test circuit. Other data on conducted and radiated spectra from various relays or solenoid operated devices are shown in Table 3-5 (Ref. 22).

The principles of switching transients just discussed apply in alternating as well as direct current circuits, but in alternating current circuits the magnitude of the transient and contact arcing is very much a function of the time during a cycle that circuit interruption begins. The magnitude of the expected transient is of the same order for equal maximum voltage or current. However, the transient itself may be affected significantly by saturation effects in magnetic circuits caused by large inrush currents. Ref. 21 should be consulted for details.

3-1.3.2.3 Repetitive Switching

3-1.3.2.3.1 Rotating Machines

Because a commutator is, effectively, a device which performs an automatic and repetitive switching function in which one of the principal elements is the inductance of the machine windings, it can be expected to generate EMI in the same way. Actually, the brush used in commutation acts as a varying contact resistance in a way corresponding to the discussion leading up to Eq. 3-1.

Fig. 3-26 (Ref. 25) shows experimental data for a commutator type of generator.

Fig. 3-27 (Ref. 25) shows that the effective ac impedance of the equivalent noise source for a dc generator is independent of the load, although it varies rapidly with frequency. The dotted curve on Fig. 3-27 shows that on the basis of open circuit voltage and impedance measurements, the output noise level from this device can be predicted with reasonable accuracy for any load. Similar results are obtained for ac generators measured as noise sources.

Fig. 3-28 (Ref. 26) shows conducted noise voltage measured on a dynamotor in the frequency range 150 kHz to 15 MHz.

3-1.3.2.3.2 Gaseous Discharge Lamps

Gaseous discharge lamps operate on the principle of an arc breakdown, a phenomenon that occurs twice during each ac cycle.

Fig. 3-29 (Ref. 27) shows the electromagnetic interference (EMI) voltage from a fluorescent lamp in a limited frequency range, measured directly across the lamp terminal. Fig. 3-30 (Ref. 27) shows, however, that the impedance of the lamp (including its ballast) is so large that very little conducted noise voltage results. However, in the frequency region of impedance resonance, the lamp exhibits fairly sustained oscillations (in the neighborhood of 12 kHz). This oscillation might create serious radiated EMI in some circumstances.

Radiated EMI measurements on a different fluorescent lamp in a higher frequency range are shown on Fig. 3-31 (Ref. 28). In making these measurements, the electric antenna was placed 3 ft from the lamp. Electric fields from fluorescent lamps usually can be eliminated entirely by using glass with a conductive coating and whose surface is electrically bonded to an otherwise complete metal enclosure.

TABLE 3-5. EMI FREQUENCY SPECTRA OF ELECTROMECHANICAL SWITCHES (Ref. 22)

Switch Type	EMI Frequency Spectra
Latching power relay	15 kHz to 400 MHz radiated 1 MHz to 25 MHz conducted (coil lines) 30 kHz to 25 MHz conducted (contact lines)
Power transfer relay	15 kHz to 150 kHz, 25 MHz to 400 MHz radiated 150 kHz to 25 MHz conducted
Switching relay	130 kHz to 60 MHz radiated 600 kHz to 12 MHz oscillation conducted on coil lines 87 to 150 kHz oscillation conducted on coil lines
Solenoid-operated valves	1 to 8 kHz conducted
Power contactors	150 kHz to 25 MHz conducted
Coil-operated coaxial switch	150 kHz oscillation conducted on coil lines

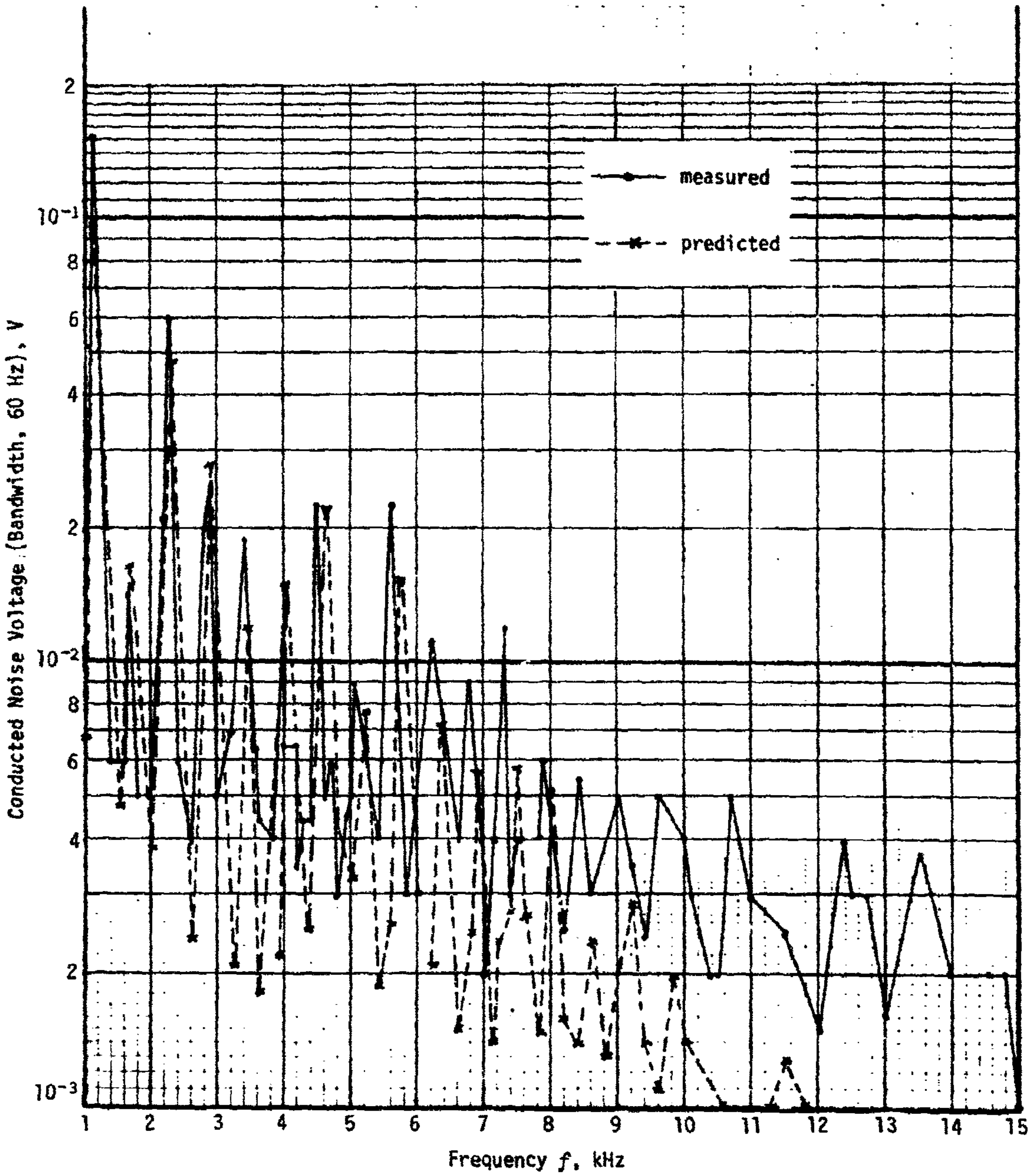


Figure 3-26. Conducted Noise Voltage from a 0.5-kW dc Generator at Full Load (1725 rpm, 115 V dc, 4.35 A, R = 2.76 Ω) (Ref. 25)

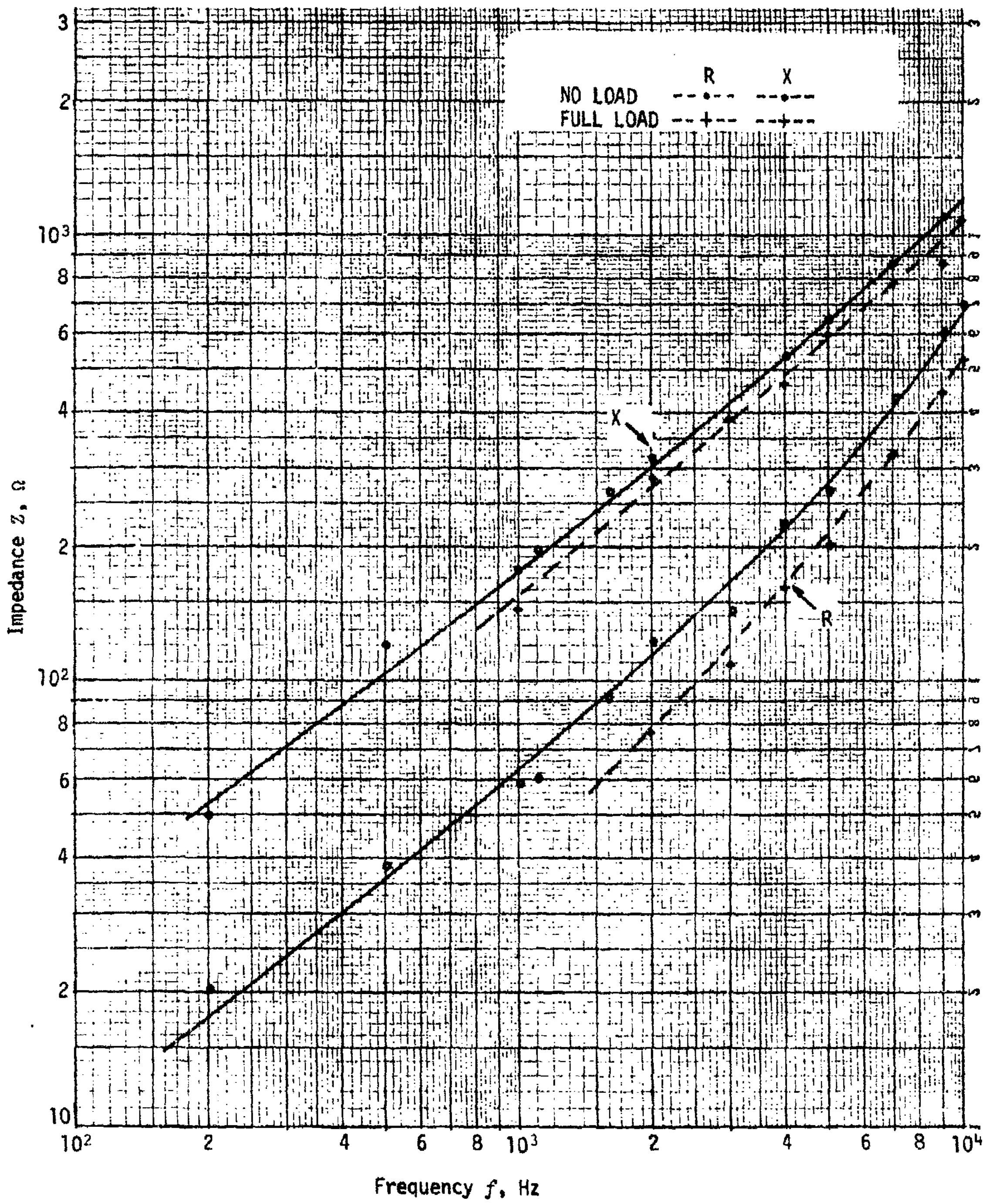


Figure 3-27. Resistive and Reactive Components of EMI Source Impedance of 0.5-kW dc Generator Showing Variation With dc Load Current (Ref. 25)

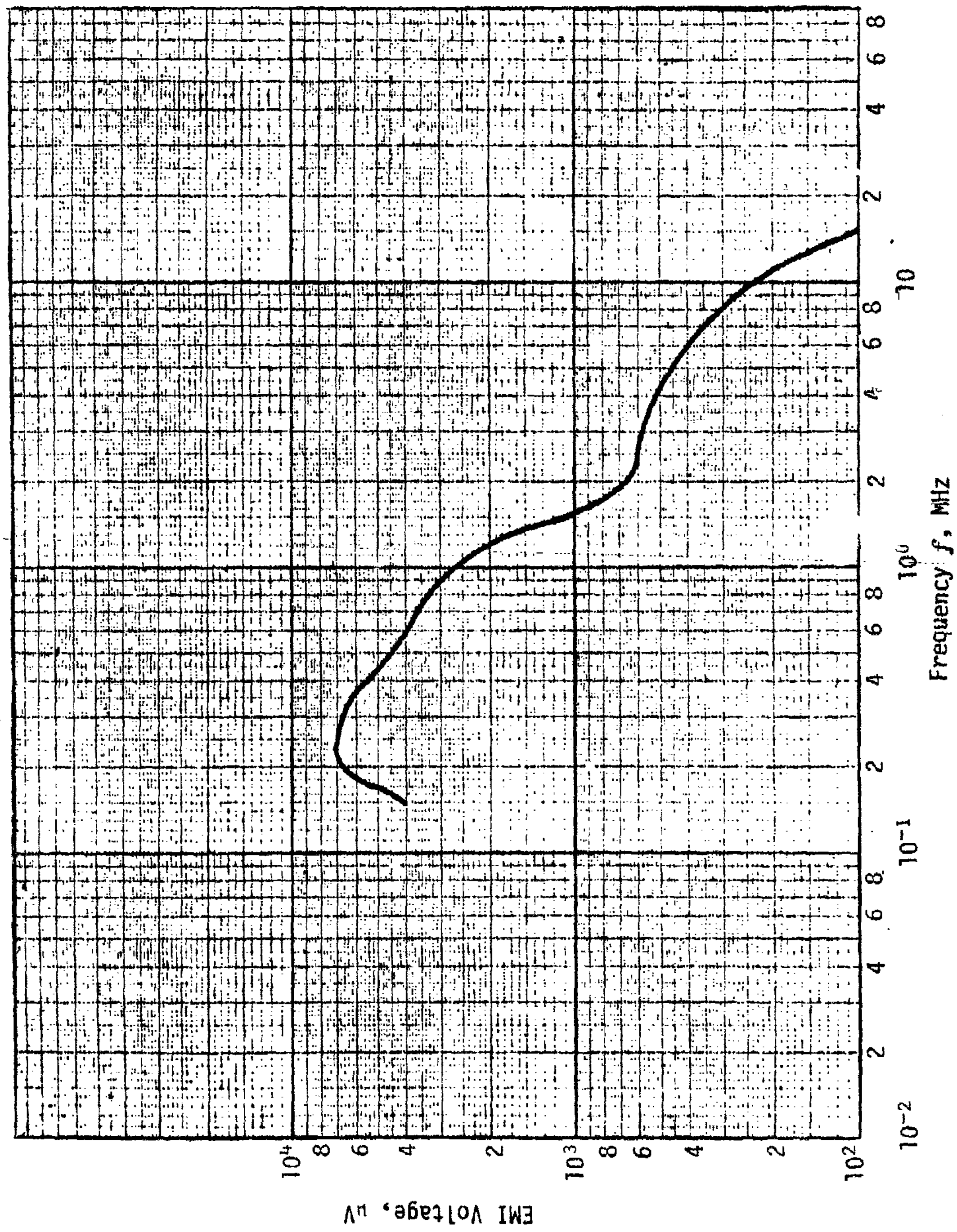


Figure 3.28. Dynamotor (14 V dc - 250 V dc) Conducted EMI Voltage Line to Ground (Ref. 26)

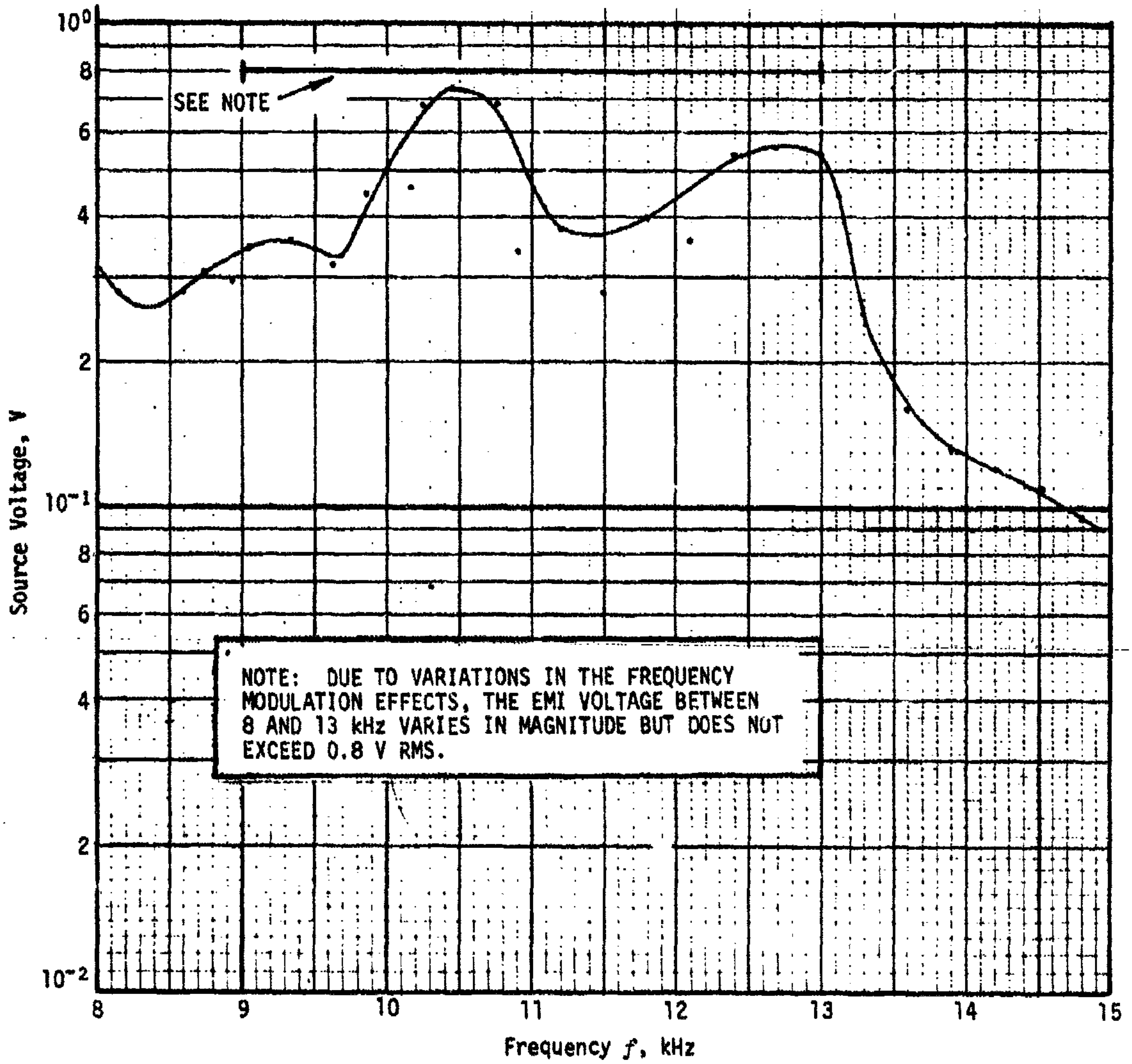


Figure 3-29. EMI Source Voltage of Fluorescent Lamp (Ref. 27)

3-1.3.2.4 Automotive Ignition

Data on a number of incidental radiating devices have been summarized by Myers (Ref. 29). He reports the work of Newell on a simulated ignition system as shown here in Fig. 3-32 (Ref. 30). The comparison between measured and theoretical current spectra shows remarkable agreement, the peak at about 20 MHz being due to a resonance in the system. The extension of this model to predict radiated values is almost impossible, because of the extreme dependence on the amount of shielding contributed by the metal body of the vehicle.

Fig. 3-33 (Ref. 31) shows the range of field strength (both horizontal and vertical polarization) as a function of frequency for 21 vehicles and also shows the limits of SAE Standard J551. These data were taken in 1969 at a distance of 10 m.

When vehicles are moving, a spread in levels even larger than shown on Fig. 3-33 will be experienced. Fig. 3-34 (Ref. 32) shows the distribution function for vehicles on a 7-lane freeway when measured by an antenna 51 ft from the nearest lane of traffic. In this case a power measurement was made. Although the average value of measured power was only 18 dB

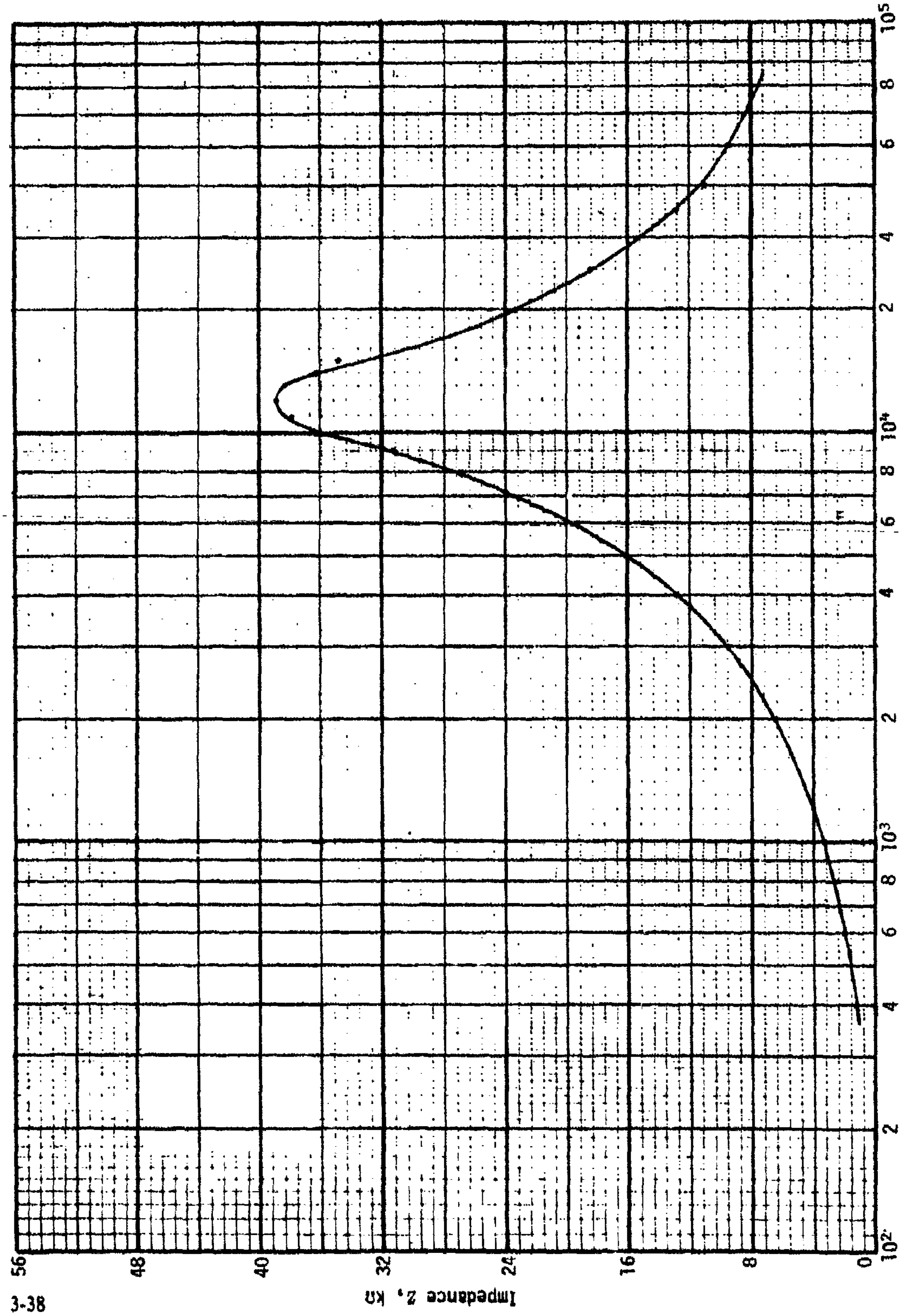


Figure 3-30. Source Impedance of the Fluorescent Lamp Fixture (Ref. 27)

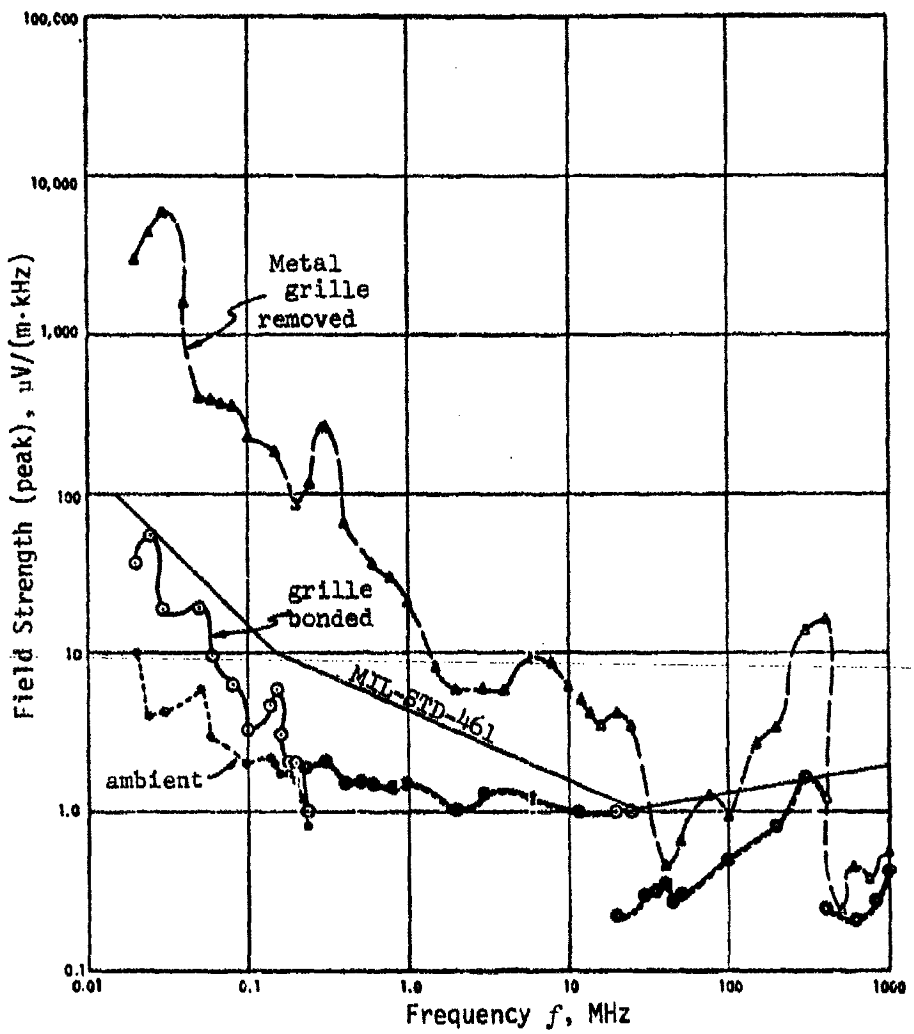


Figure 3-31. Broadband Radiated Interference (Peak) (Ref. 28)

above kT_0B, the value exceeded 1% of the time was about 50 dB above kT_0B.

Military vehicles for tactical use are suppressed to lower levels than SAE Standard J551 (see par. 5-6.3.2). This is because they frequently carry communications equipment which must be usable while the vehicle is moving or standing with engine operating. Furthermore, because equipment mounted on the vehicle or connected to its power supply is potentially susceptible, levels of conducted interference must be controlled. Even if tactical vehicles do not carry communication equipment, they must be suppressed because they can be located close to communication installations.

3-1.3.2.5 Semiconductor Switching

Fig. 3-35 shows a source of interference arising from a switching transient (Ref. 113) in a semiconductor diode. The source of interference is a pulse generated at diode turnoff by stored carriers in the junction.

3-1.3.2.6 High-voltage Power Lines

EMI generated by power lines is of two types, namely:

a. That due to various forms of corona which generally has a random waveform, but which is modulated at the power frequency or multiples of it. This

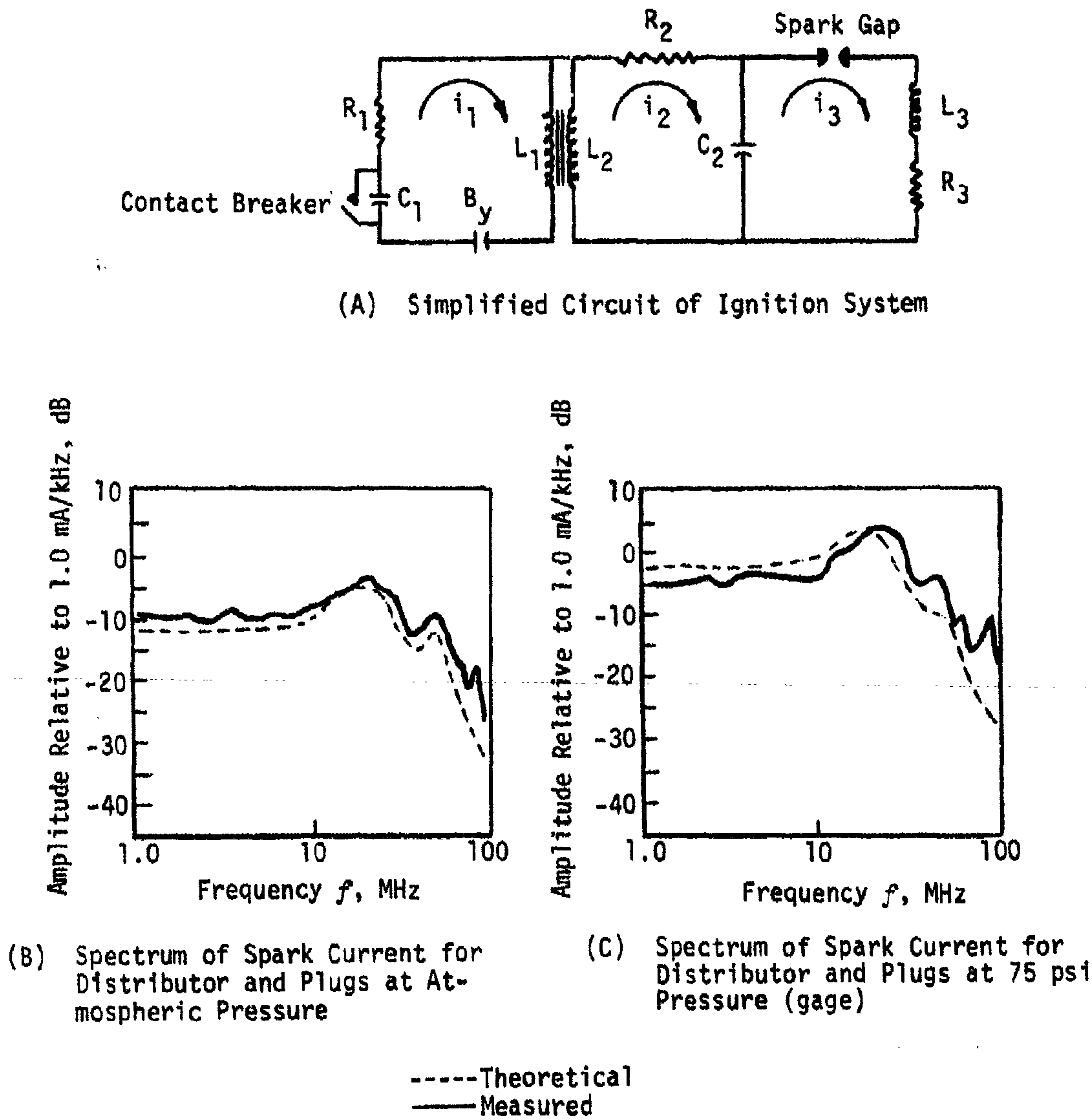


Figure 3-32. Spark Current Generation and Current Spectra (Ref. 30)

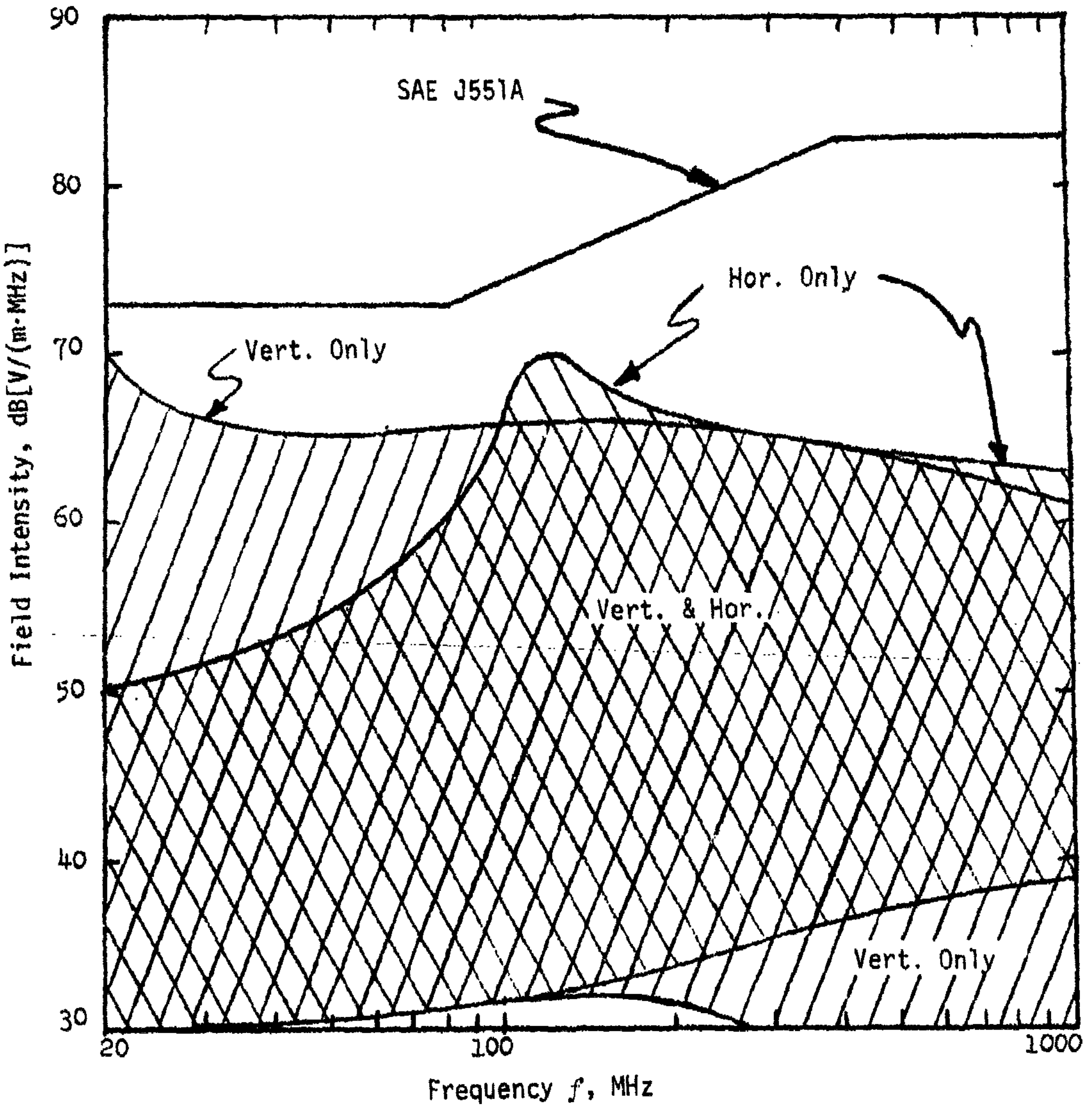
source has a high duty cycle and is usually not important at frequencies above 100 MHz.

b. That due to localized static discharges across insulators or other forms of line hardware which is more impulsive in character. This type of interference has substantial components at frequencies of the order of 100 MHz and does not propagate on the line.

Measured data compiled by Skomal (Ref. 33) are shown on Fig. 3-36, for a variety of line voltages and distances from the line, under fair weather conditions. In fog, mist, or rain, levels may be 10 to 20 dB higher.

3-1.3.3 Nonlinear Phenomena

The term nonlinear phenomena could be interpreted to include almost any mechanism of EMI generation. For example, almost all switching involves some nonlinear action in the arcing that must always occur. Here the term is reserved for those



(Range of Radiation from Individual Vehicles
Measurement Distance 10 m)

Figure 3-33. Individual Vehicle Electromagnetic Radiation (Ref. 31)

devices in which there is a relatively well defined nonlinear relation between current and voltage. The devices of major interest are those in which ac (sine waves) or periodic pulses are applied.

Nonlinear impedances appear in electric circuits by several mechanisms. They may be intentionally included in the circuit — as in the case of rectifiers, mixers, and frequency multipliers — or they may arise unintentionally, as in the case of saturation in electronic tubes, transistors, and magnetic materials.

A nonlinear device excited by a single-frequency sine wave signal will produce harmonics of that sine wave. If the nonlinearity is quite severe, the signal produced may be characterized by abrupt changes in

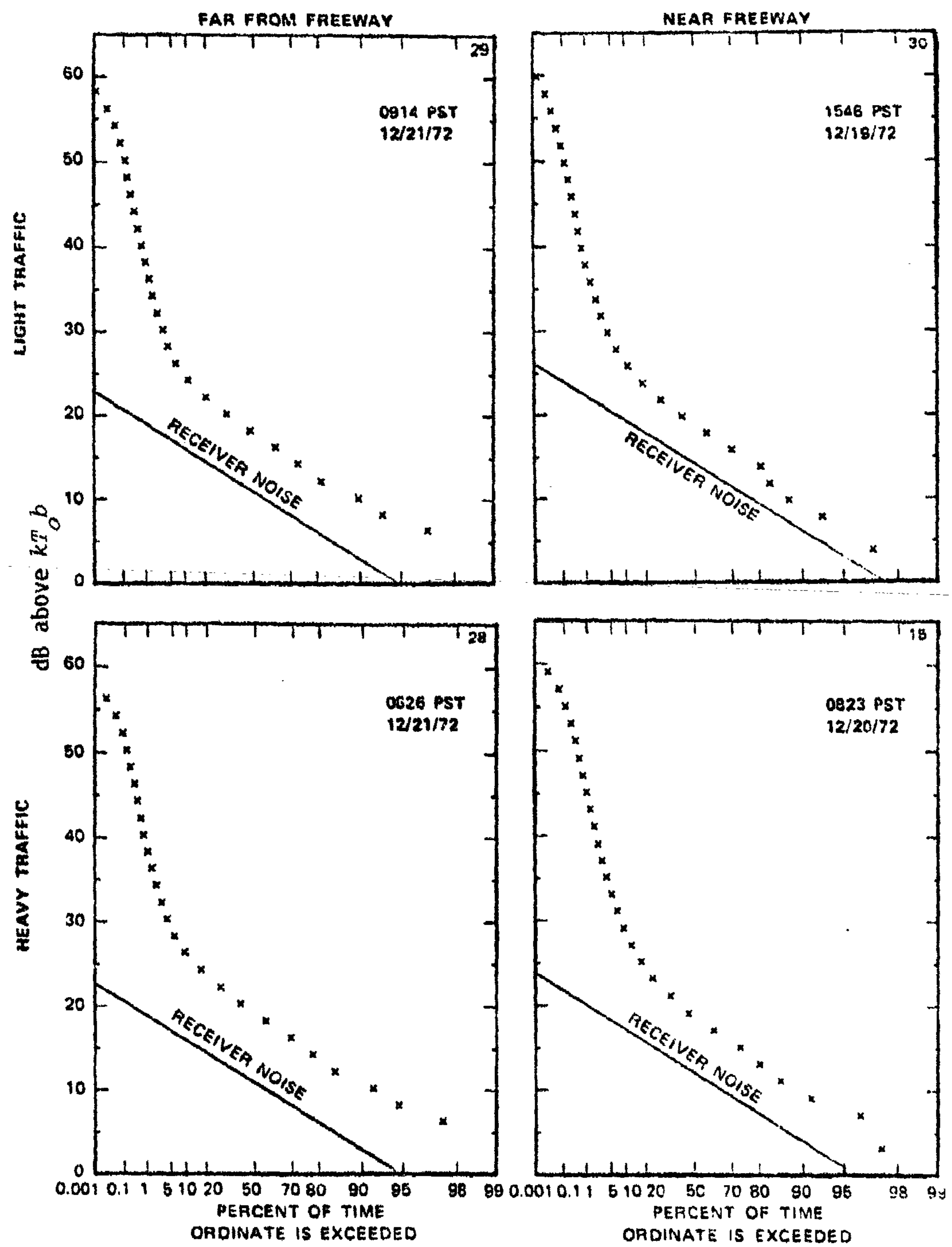
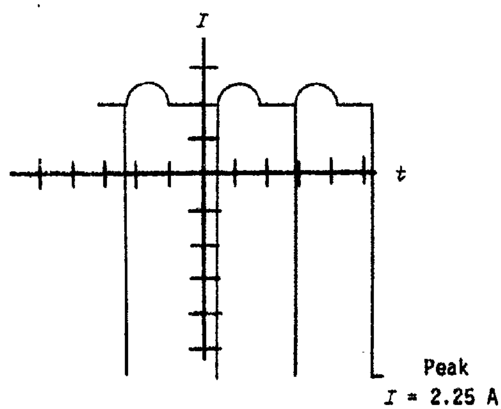
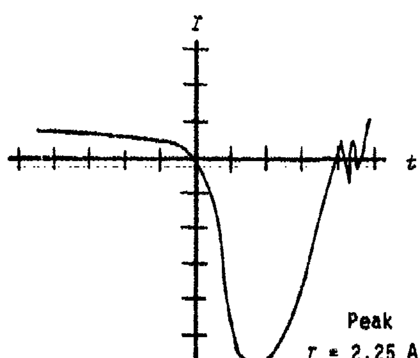


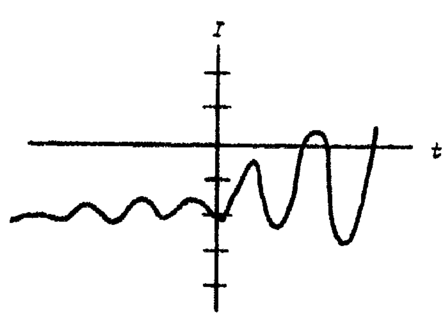
Figure 3-34. Average Probability Distributions for Freeway Traffic, 24.11 MHz (Ref. 32)



(A) Current Wave: Four Conducting Cycles of 1N345 at 880 mA



(B) Same as (A) With Sweep Rate Increased to 10 μs/cm



(C) Same as (A) With Sweep Rate Increased to 1 μs/cm

Figure 3-35. Oscillograms of Reverse Current Transients for a 1N345 Diode at 880-mA dc Load (Ref. 20)

level, and may appear as a succession of switching transients rather than a sine wave corrupted by harmonics. These concepts are illustrated in Figs. 3-37 and 3-38. The nonlinearities represented in Figs. 3-37(A) and (B) are typical of the "switching" type of nonlinearity, i.e., the slope of the output-input curve increases as input increases in magnitude (although not necessarily symmetrically as shown). The outputs of such nonlinearities are quite rich in harmonics; indeed, in the extreme case of Fig. 3-37(C) the level of the harmonics can be constant up to quite high frequencies. Examples of devices having this kind of nonlinearity to a lesser or greater degree are mixers, detectors, rectifiers, magnetic core inductors, and devices utilizing gas discharges, such as fluorescent lamps. Another type of nonlinearity is the "saturating" type shown in Fig. 3-38. In this type of nonlinearity, the output waveform is "clipped" and in the extreme case, represented in Fig. 3-38(B), becomes a square wave with fundamental frequency equal to the frequency of the sine wave input. The harmonics of a square wave fall off inversely with frequency. An example of a device with saturating nonlinearity is the linear amplifier when operated at too-high an input level.

Harmonic generation by nonlinear phenomena also creates problems due to the mixing of signals presented to the nonlinear device. Sometimes this is intentional, as in mixers, converters, and detectors. But intermodulation and cross-modulation are undesired phenomena in circuits designed to be linear, primarily radio frequency amplifiers.

3-1.3.3.1 Power Frequency Harmonic Generation

The primary sources of harmonics in ac electric power lines and distribution systems are rectifiers and magnetic saturation in transformers.

3-1.3.3.1.1 Rectifiers

The load and line currents for ideal single-phase, unfiltered half-wave and full-wave rectifiers are shown in Fig. 3-39. For the half-wave rectifier the magnitudes of the harmonic currents $I(n)$ for both line and load are given by

$$I(n) = \begin{cases} \dfrac{I_{pk}}{2}, \quad n = 1, \quad \text{A} & (3\text{-}40) \\[2ex] I_{pk}\dfrac{2}{\pi(n-1)(n+1)}, \; n \text{ even}, \; \text{A} & (3\text{-}41) \end{cases}$$

where

I_{pk} = peak value of current, A
n = order of harmonic

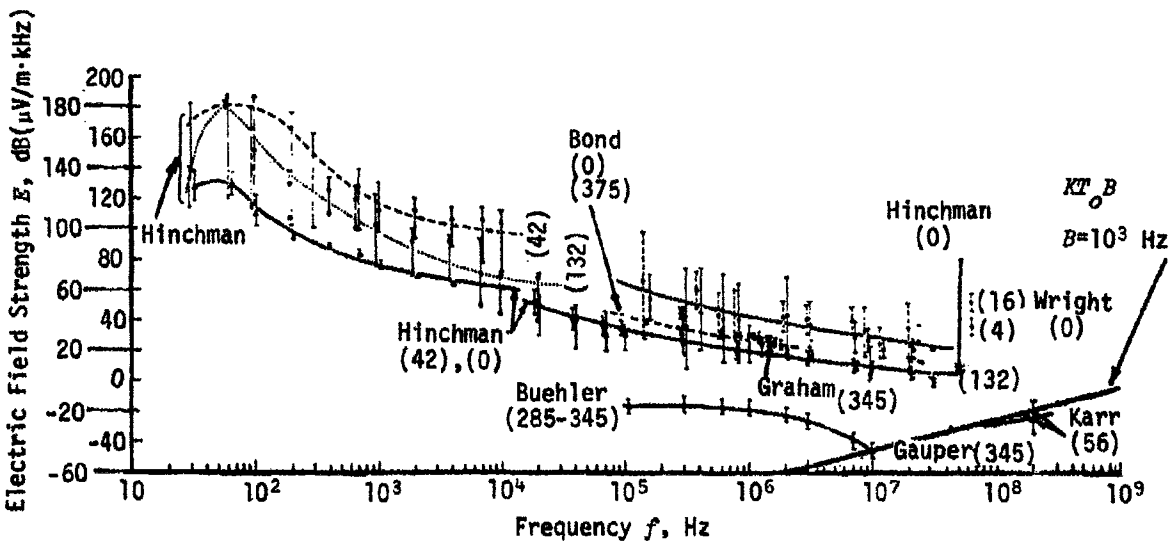


Figure 3-36. Radiated EMI from High-voltage Power Lines (*Key*: (Line voltage in kV as: (4), (16), (42), (66), (132), (345), (375). Observation point locations: ● beneath conductor, ◇ 25 ft laterally from line, ■ 50 ft to 100 ft laterally from line, ○ 100 ft laterally from line, △ 2500 ft laterally from line, ▲ 5000 ft laterally from line. Vertical lines show the range of measured values.)(Ref. 33)

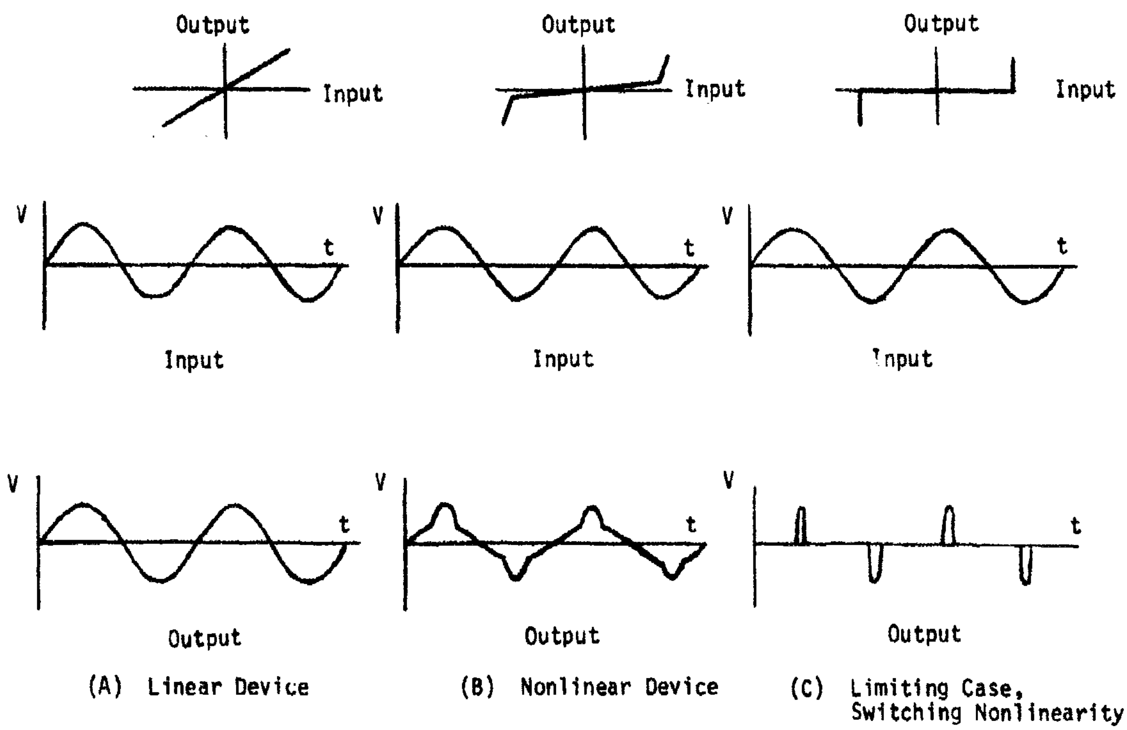


Figure 3-37. Waveforms Resulting from Switch-type Nonlinearities

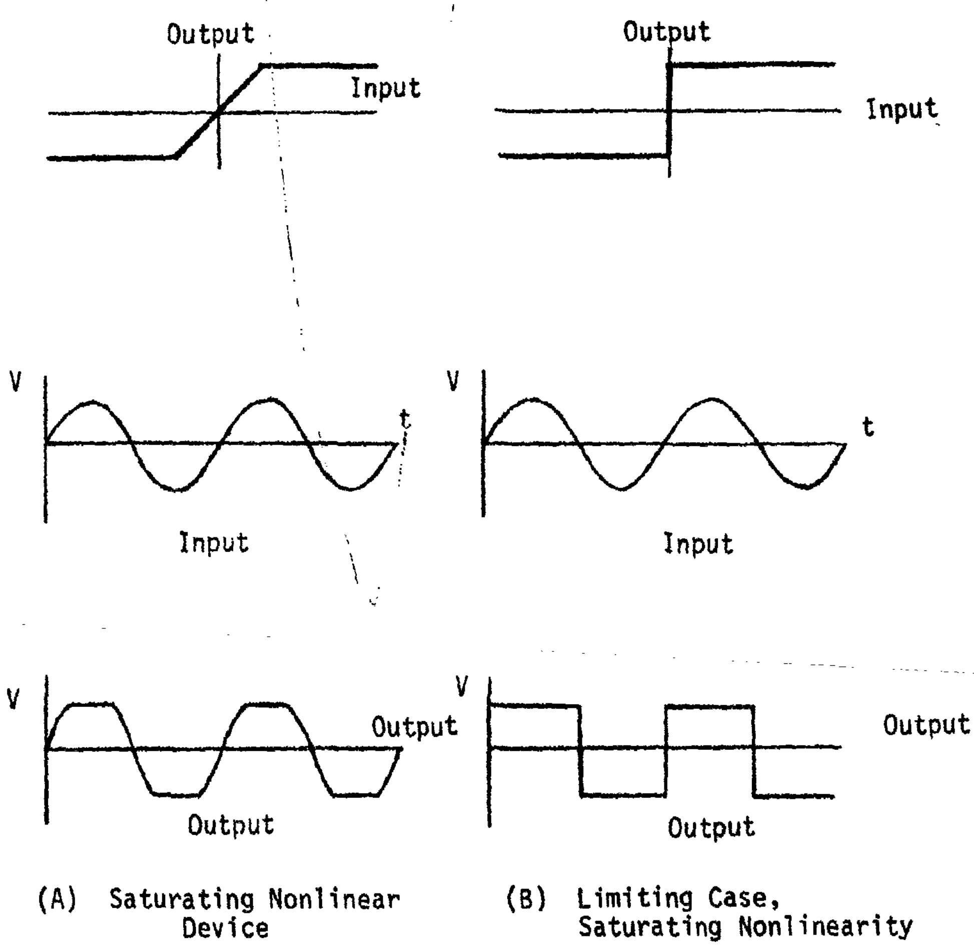


Figure 3-38. Waveforms Resulting from Saturating Nonlinearities

For the full wave rectifier the harmonics of the load current $I(n)$ are given by

$$I(n) = \frac{4I_{pk}}{\pi(2n^2 - 1)}, \text{A} \tag{3-42}$$

Note that the line current for the full-wave rectifier is sinusoidal, so that there would be no harmonics generated on the power line by an ideal full-wave, single-phase rectifier. The same conclusion holds for a full-wave bridge rectifier.

For most purposes the harmonic content of the load currents of simple rectifiers is too large for satisfactory use. To reduce these harmonics, filters are inserted between the rectifier output and its load, thus smoothing the waveform. However, generally these filters increase line harmonics, which are usually the harmonics of most concern from an EMC point of view. The effects of these filters are discussed in detail in par. 5-5.

For multiphase power supplies, analysis of line current is more complex than for single-phase supplies; however, multiphase circuits may be used to advantage to minimize ripple in the dc output as well as harmonic emissions on the power line. By proper choice of circuit, the rectifier may be designed inherently to cancel out certain harmonics. This is discussed in detail in par. 5-5.4.2.

3-1.3.3.1.2 Transformers and Inductors

Transformers and inductors using magnetic materials for cores have nonlinear current-voltage characteristics due to saturation effects in the core. In

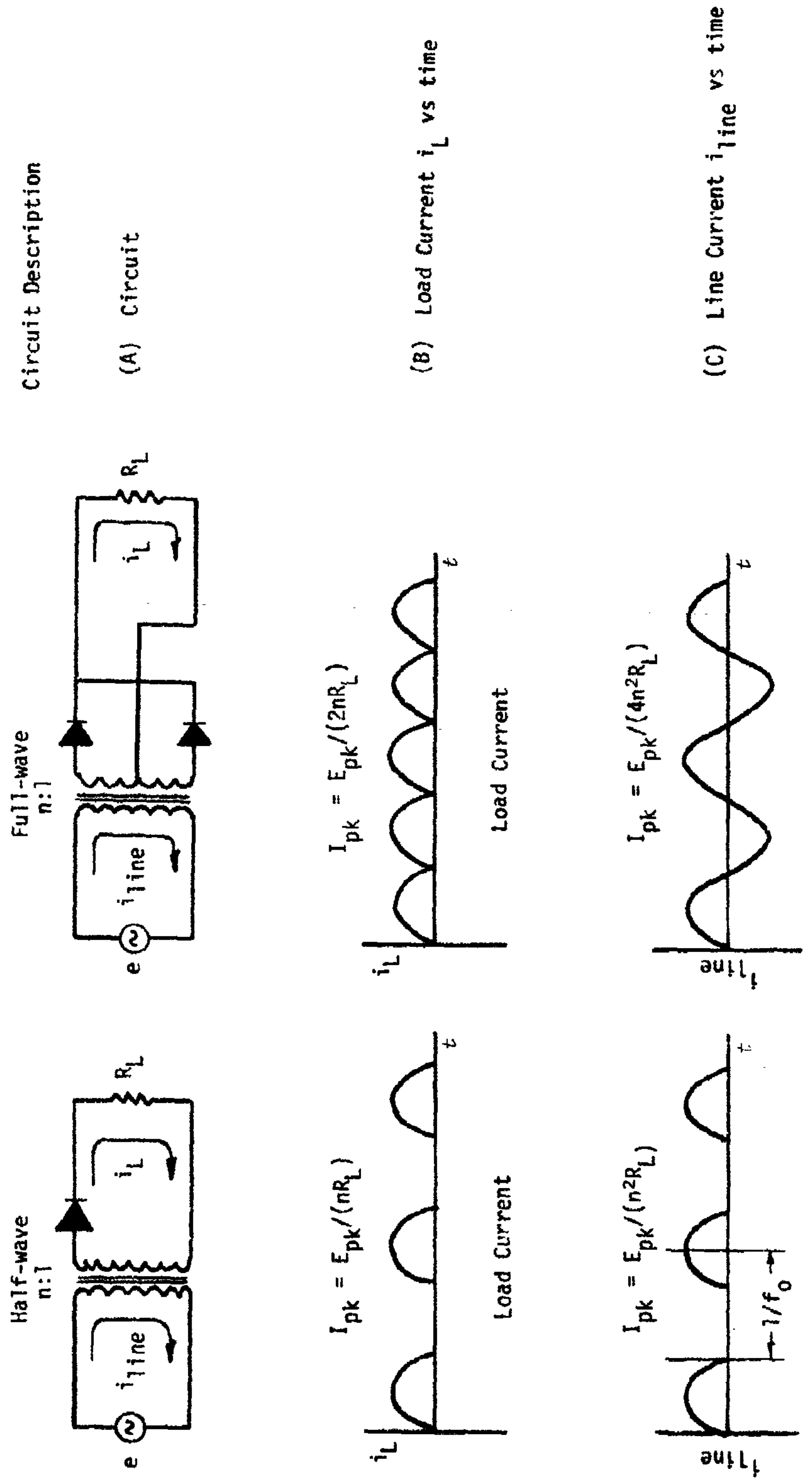


Figure 3-39. Ideal Half- and Full-wave Rectifiers

power circuits, the forcing function is generally a sinusoidal voltage, so that the current waveform is distorted. This distortion results in generation of harmonics of the power frequency which may be coupled into other circuits by induction or by common impedances.

The mechanism by which this occurs is that shown in Fig. 3-40. The magnetic field strength H is proportional to the current I, while the magnetic flux density B is proportional to the voltage E. As the applied voltage is increased beyond the saturation value E_{sat}, then, because of the smaller increment of magnetic flux per unit current increment, the current must increase by a relatively large amount to generate sufficient back emf in the inductor to equal E.

In well designed power transformers the third harmonic current generated at rated voltages will not exceed a few percent of full load current, and harmonics of higher order will decrease rapidly as the order increases.

3-1.3.3.2 External Mechanisms of Interference Generation

Nonlinear effects have been observed at radio frequencies in structures in the vicinity of transmitter

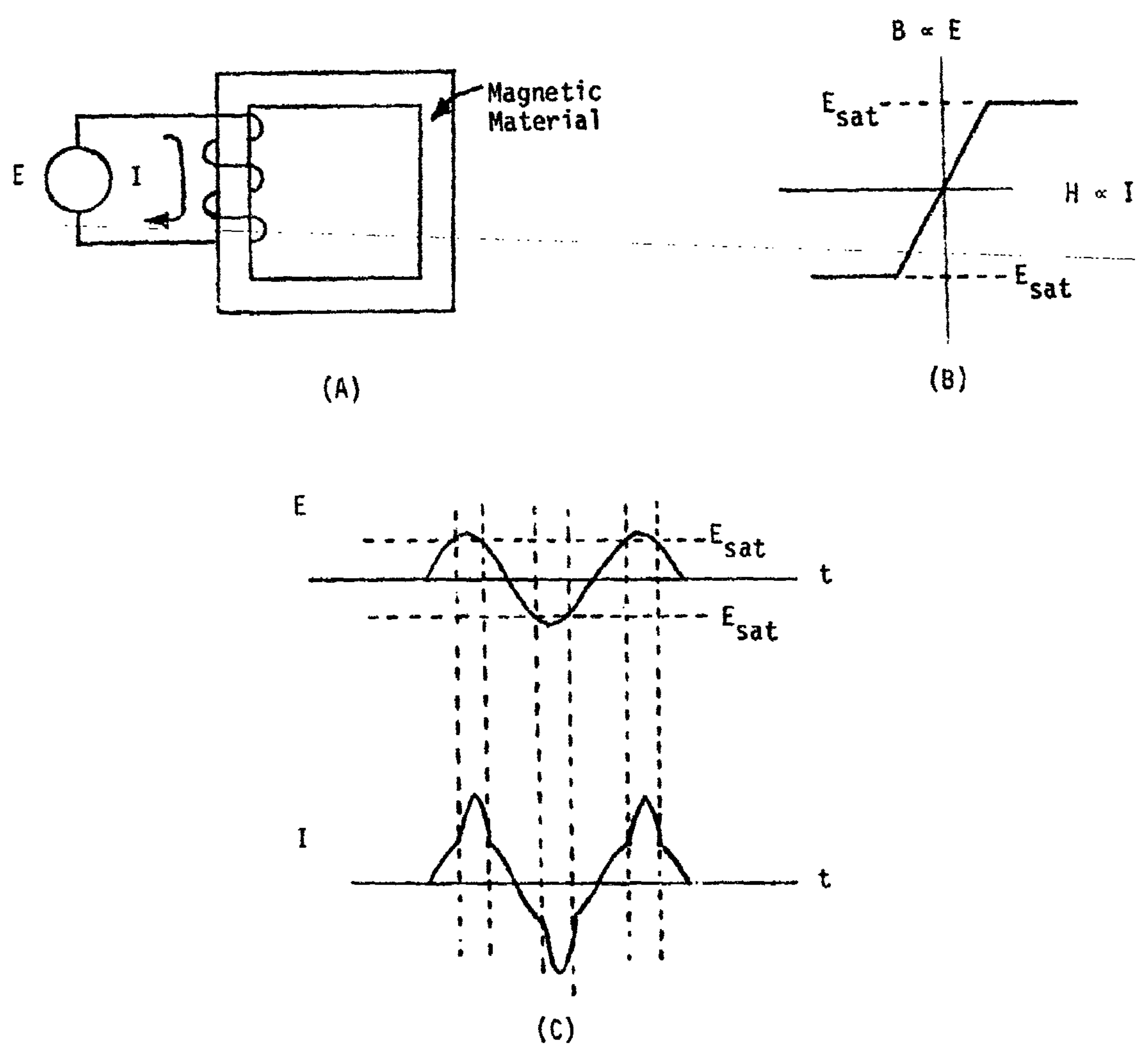


Figure 3-40. Generation of Harmonics in Magnetic Materials

antennas, resulting in the generation of spurious frequencies due to intermodulation. The phenomenon is known to arise in metals having magnetic properties, corroded metals, and corroded joints between metals—particularly where the joint is loose. The essential conclusions of a study on this matter (Ref. 34) were as follows:

a. The nonlinear phenomena found invariably had an odd symmetrical voltage-current characteristic; i.e., the power series expansion around the origin has only odd powers.

b. Welded joints are far superior to riveted or bolted joints. Turnbuckles and cable clamps are potential sources of nonlinearity.

c. The effect appears in magnetic materials such as steel, nickel, and mu-metal. Coating magnetic material with conducting material such as copper significantly reduces the effect.

d. Rough surfaces and oxide surfaces on steel exhibit the effect as do copper oxide formations on copper. Cleaning and polishing such metals are effective in reducing the effect.

e. The effect frequently is observed near sharp corners and bends. The interference level developed at the input of a receiver depends on the energies received from the transmitters by the nonlinear device, the efficiency of the conversion to the undesired frequency, and the re-radiation properties of the structure involved. If a receiver antenna is close enough to the structure, the interference might indeed be substantial. But if the receiving antenna was, for instance, 10 ft away, the interference very likely would be of negligible magnitude.

3-2 SUSCEPTIBILITY

3-2.1 MASKING AND ERROR INDUCTION

Measures of functional impairment, arising as a consequence of interference, depend on the purpose of the system. In a speech communication system, intelligibility is an important factor, and measures based on subjective articulation tests are used to assess its effects. Other criteria may be used in some cases, e.g., criteria based on speaker recognizability. Systems using visual displays — TV, radar displays, facsimile, graphic, and alphanumeric readouts — require a variety of measures, ranging from amorphously defined observed quality preferences to rigorously defined probabilistic quantities. Digital data systems invariably use average probability of error over all symbols used, but in many cases it is necessary also to know the transition probabilities for the different error pairs, and whether errors occur at random intervals or in bursts. Analog data systems typically make use of the mean square error as a basis for assessing system quality, though other quantities such as peak error and average absolute error sometimes are needed.

Training and fatigue also affect performance. Trained operators frequently can read signals correctly in interference in which an untrained operator would find them hopelessly lost. However, the concentration required to do this will tire the trained operator more quickly than will receiving under interference-free conditions. Resentment at the intrusion of an interferer also can be expected to have an effect on the quality of reception. These psychological factors have not been quantified adequately, and are only mentioned here as a caveat.

Performance measures usually are expressed as one or more curves giving the relationship between the quality measure, and the noise level for various levels of signal. In some instances, quality is related to the signal-to-noise ratio, and one curve is adequate for any signal level. The forms of these relationships sometimes vary smoothly and gradually, and sometimes abruptly, exhibiting a sharp threshold. Typically, where synchronization is affected by interference, loss of synchronism means total loss of output. Sharp thresholds also are characteristic of interference reducing modulation systems.

Ultimately in system evaluation a functional relationship between impairment and cost must be identified, a relationship which is often elusive. In a digital system, where the error probability is the quality measure, one might assess cost in terms of that required to provide the higher transmitting power needed over that needed in the absence of impairment. But then other costs must be accounted for; for example, the increased power may result in an increase in interference with another system.

3-2.1.1 Speech Systems

Audio systems intended for speech communication are evaluated in terms of syllable, word, or sentence intelligibility, as determined by psychoacoustic tests. The degree of intelligibility, called the articulation, is the ratio of the number of language elements currently understood to the total number of elements used. The results of articulation tests depend on the language itself, the class of talkers, what words (or other language elements) are used in the test, and the class of listeners.

The most common tests make use of lists of words. These words are read to a listener through the system being examined. An easy test to grade is the Rhyme test (Ref. 35) in which single syllable words are read to a listener who is given the word ending and must supply the initial sound (letter). Another test uses the Phonemically Balanced (PB) word list. Each successive group of 50 words is balanced to contain a proportioning of phonemes typical of that found in the English language. Listeners are informed that all words used are valid English words, and not nonsense syllables. The listener writes the entire word on this score sheet, and scoring can be based on the number of words or the number of phonemes in error. Both of the tests described here are used for diagnostic purposes to determine which phonemes are apt to be misinterpreted and whether or not a member of one class of phoneme (e.g., a stop sound) is apt to be interpreted as a member of another class (e.g., a fricative). A discussion of these procedures will be found in Drucker (Ref. 36).

The use of articulation tests is costly and time consuming. In their place one may use an analytical method that gives estimates of articulation, using predetermined spectral properties of speech and the spectral properties of the interference. The method is based on the spectral distributions shown in Fig. 3-41 (Ref. 37), and a derived quantity called the articulation index (Refs. 37 through 40). The curves pertaining to speech show spectral density levels of sound pressure at one meter in front of a talker using a raised voice; the raised voice is 6 dB above the level of a normal speech. Three curves corresponding to peak speech levels, average levels, and minimum levels are shown. Peaks are 12 dB above the average level, and minimums are 18 dB below the average level. The frequency ranges from 200 to 6100 Hz, and the abscissa is divided in such a way that equal width intervals correspond to equal contributions to the articulation index. If all of the area between the peak and minimum curves is uncontaminated by interference, is not filtered away, and does not fall below threshold or above overload, the articulation index is 100%. If part of this region is obscured, the loss of articulation index is determined by the area of the obscured portion. The abscissa is marked off in frequencies which are the centers of 20 critical bands, each band contributing 5% to the articulation index. The articulation index A in a particular application is given by

$$A = \sum_{n=1}^{20} 0.05 W_n \tag{3-43}$$

where

W_n = articulation index parameter

= [(speech peak of nth band)

− (noise rms value in nth band)]

÷ 30 (3-44)

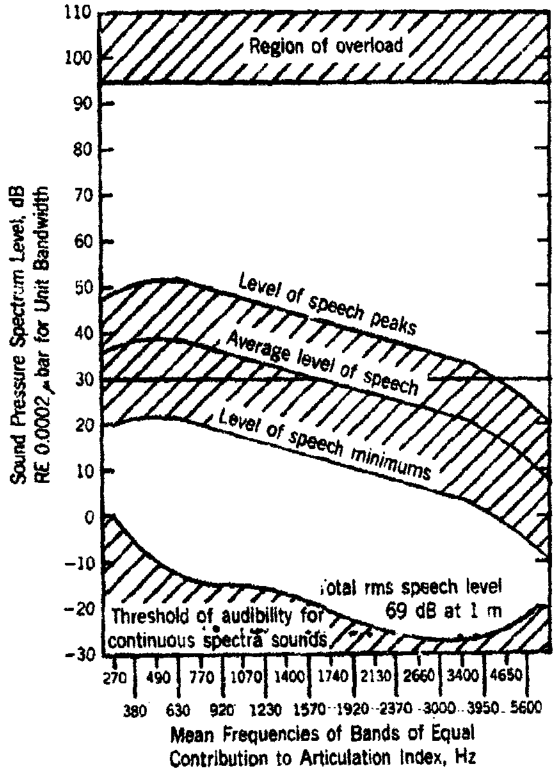


Figure 3-41. Plot on a Spectrum Level Basis, of (1) the Speech Area for a Man Talking in a Raised Voice; (2) the Region of "Overload" of the Ear of an Average Male Listener; and (3) the Threshold of Audibility for Young Ears (All curves are plotted as a function of frequency on a distorted frequency scale.) (Ref. 37)

Both terms in the numerator of W_n are measured in dB relative to 0.0002 μbar in a 1-Hz band. The speech peak is taken from the curve of Fig. 3-41. The factor 30, in the denominator corresponds to the 30-dB range between peaks and minimums. W_n is set equa to zero if Eq. 3-43 is negative, and it is set equal to unity if Eq. 3-43 is greater than unity. By assuming a speech communication system in which norma volume levels are set so that one is never belov threshold or above overload, and where the speecl band at the output is unchanged from that shown i Fig. 3-43, Eq. 3-44 can be rewritten

$$W_n = \frac{(S/N)_n + 12}{30}, \; -12 < (S/N)_n < 18 \tag{3-45}$$

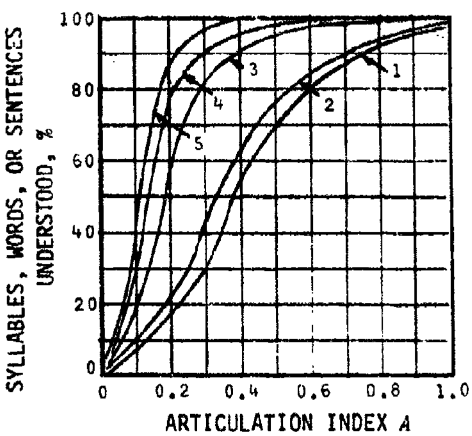
where

$(S/N)_n$ = ratio of the rms spectral density o average speech to rms spectral density c the noise at the output in the nth banc dB

Shown superimposed on Fig. 3-41 is a flat spectral density of about 31 dB, which is an assumed level for an interfering noise signal whose total rms value over the band from 200 to 6100 Hz equals the total rms level of the average speech curve. This is 69 dB in a band of 5900 Hz or an average of 31.3 dB in a band of 1 Hz. By using Eqs. 3-44 and 3-45 and Fig. 3-41, the articulation index — corresponding to a unity ratio of total signal power of average speech to total noise power — is found to be about 40%. The articulation index at other levels of total signal-to-noise ratios is obtained by drawing a horizontal line which lies above or below the horizontal line drawn on Fig. 3-41 by an amount equal to the total signal to noise ratio, and making the calculation indicated by Eqs. 3-43 and 3-45.

The articulation index is related to syllable, word, or sentence articulation. By using Fig. 3-42, taken from Kryter (Ref. 40), sentence articulation is found to be about 95% if the articulation index is 40%. Word articulation for a PB test is on the order of 90%.

THESE RELATIONS ARE APPROXIMATE. THEY DEPEND UPON TYPE OF MATERIAL AND SKILL OF TALKERS AND LISTENERS.



1. PS words (1000 different words)
2. Non-sense syllables (1000 different syllables)
3. Test vocabulary limited to 255 PB words
4. Sentences
5. Test vocabulary limited to 32 PS words

Figure 3-42. Several Experimental Relations Between Articulation Index and Speech Intelligibility (Ref. 40)

If this method is applied to noise consisting of a pure audio tone, it can reside in only one of the 5% articulation index bands, and it would degrade the articulation index to no less than 95% no matter what its amplitude is. The intelligibility loss, however, does depend on the noise amplitude. Results of masking by pure tones are given by Stevens, et al. (Ref. 41) and Christman, et al. (Ref. 42). By using PB word lists, Ref. 43 shows that at a speech level of 69 dB (which corresponds to the total rms speech level for average speech at 1 m from a loud talker, see Fig. 3-41) the articulation is as shown in Table 3-6. The results vary with speech level and the frequency of the masking tone. At a speech level of 69 dB, 90% articulation is obtained with noise levels ranging from 87 dB at 100 Hz to 114 dB at 1000 Hz. At lower speech levels, the spread in noise levels over this frequency range is smaller.

The Christman report also gives results for the random noise masking of speech. In general, random noise is much more effective than are sine waves. Typically, for 75-dB speech, random noise requires about 30 dB less power to produce the same loss of intelligibility as do sine waves.

3-2.1.2 Visual Display Systems

Visual displays cover a wide range of types that need very different quality measures. In television, verbal descriptions of quality or preference are used. In radar, measures such as increased time for detection are used.

Measurements of observer reaction to television pictures affected by interference have been made by Weaver (Ref. 43) on 405-line (3-MHz bandwidth) and 625-line (5-MHz bandwidth) systems. Gaussian random electrical noise with a bandwidth appropriate to the two video systems was used. The addition of signal and noise was carried out in the video circuits (i.e., not in RF circuits). Essentially, flat electrical noise which was passed through a differentiator giving a 6-dB-per-octave decrease of attenuation was inserted. Observer reaction was rated on a

TABLE 3-6. ARTICULATION AS FUNCTION OF SIGNAL-TO-NOISE LEVEL (SIGNAL LEVEL 69 dB)

NOISE LEVEL, dB	S/N	ARTICULATION
115	−46	0.35
105	−36	0.57
95	−26	0.82
85	−16	0.92
75	−6	~1.0

scale of 1 through 6, the scale elements having the associated descriptions:

1. Imperceptible
2. Just perceptible
3. Perceptible but not disturbing
4. Somewhat disturbing
5. Objectionable
6. Unusable.

The results obtained are indicated in Fig. 3-43. The signal-to-noise ratio is based on the peak-to-peak signal amplitude, and the rms value of the noise. For the US standard 525-line system, it is reasonable to assume the results would fall somewhat to the right of a curve midway between the 405 line and 625 line curves shown here. For example, if "just perceptible" noise is acceptable (rating 2) about 38 dB (peak-to-peak signal)/(rms noise) is indicated as being required for flat noise.

A more recent set of tests on US Standard 525-line television using a 7 step quality scale has been reported by Cavanaugh (Ref. 44). This scale is similar to the Weaver scale, with the addition of a level in the middle described as "impairment but not objectionable". The objective of this study was to devise a noise weighting curve as a function of frequency which accounts for subjective effects, and which suits both monochrome and color television. It was with such a weighting curve, which falls off 14 dB from low frequencies to 4.5 MHz, that the tests were made. The noise spectra used were those typically found in the Bell System transmission networks plus two flat spectra noises, one going to 4.5 MHz and the other to 200 kHz. The data show that for flat noise to 4.5 MHz, more than half of the observers judged the noise just perceptible or imperceptible with a 49 dB (peak-to-peak)/(rms noise) signal-to-noise ratio for both color and monochrome pictures. With a signal-to-noise ratio of about 43 dB, more than half the observers judged the pictures to have at most some impairment, but not to be disturbing.

Results of tests of the effect of co-channel and adjacent-channel interference on television reception are reported by Allnatt, et al. (Ref. 45). These tests were scored using a 5 point scale. With co-channel interference, the observer reaction was influenced strongly by the degree of frequency offset between the desired and undesired signal, certain offsets giving much improved results. In the worst case of frequency offset, a 44-dB signal-to-interference ratio was needed to achieve a 50 percent favorable

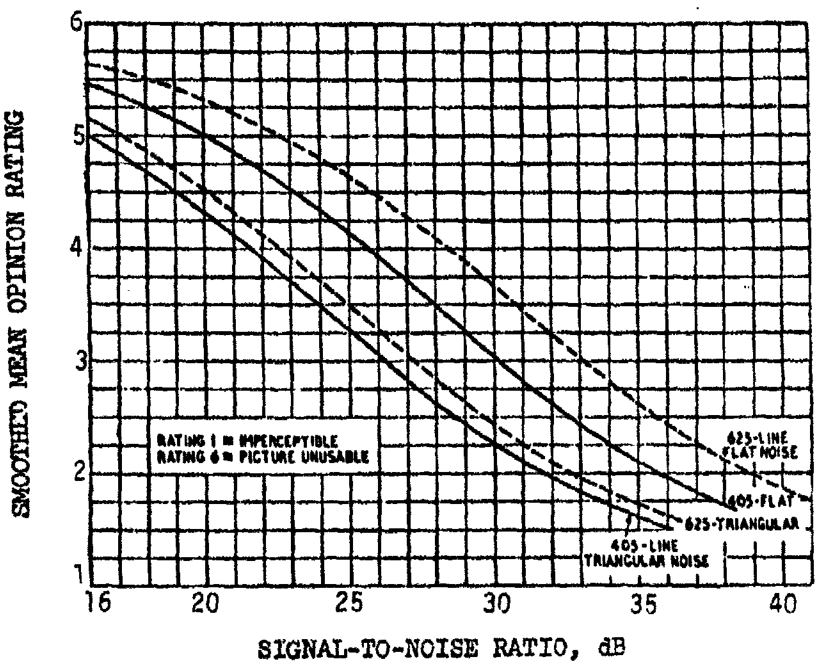


Figure 3-43. Relation Between Signal-to-Noise Ratio and Mean Opinion on a 6-point Scale

response. In the best case, 26 dB was needed. No substantial difference is found between the two standards used in these tests, 625 lines and 405 lines. One set of adjacent channel tests was made with the idea that adjacent channel interference is frequently of the kind where the audio information in a channel falls into the picture band of the adjacent channel. The interfering signal was placed accordingly 1.5 or 2.5 MHz below the vestigial sideband picture carrier. The undesired signal was either CW, or an FM or AM wave modulated with a 1-Hz sinusoid. Under worst conditions, an 8-dB signal-to-interference ratio resulted in a favorable reaction from half or more of the observers. The interference, in this case, was measured in terms of its rms value, and the values of both signal and interference were measured at the receiver input. In another set of adjacent channel tests with the interference above the desired signal frequency, the interference was due largely to the picture components of the undesired signal. In this case, with the two carriers separated by 7 MHz, a worst case result of 14-dB signal-to-interference ratio for 50 percent or better favorable reaction was found. It should be pointed out that adjacent channel results depend heavily on the receiver selectivity characteristic. Results found in one application only should be applied in other applications with caution.

Subjective interference with target location on a radar PPI (plan position indicator) has been reported by Hudson and Limburg (Ref. 46). Interference typical of unwanted signals from nearby radars was mixed with one or more desired target signals, and the increased time required for detection by trained operators was measured. Interference conditions used in this test are shown in Fig. 3-44 (Ref. 46) (these categories have been established by the Rand Corp. and are specified in Ref. 47), and the decrease in detection range (which can be related to increase in detection time) as a function of target velocity is shown in Fig. 3-45. Values of S/I to obtain the five interference conditions are given by Katz (Ref. 48).

Effects of electrical noise in systems used for transmission of printed and written documents will be found in the literature. Schlaepfer, et al. (Ref. 49) discusses the effect on systems using data compressing run-length codes which obliterate several lines of information when an error occurs. A literature review of legibility of displays is given by Shurtleff (Ref. 50).

3-2.1.3 Digital Systems

The error probability in coherent and differentially coherent phase-shift keyed (PSK and DPSK) digital transmission systems operating in noise and interference has been calculated by Rosenbaum in several papers (Refs. 51, 52, and 53). The result for coherent binary PSK is reproduced in Fig. 3-46 (Ref. 51); results for multiphase coherent and differentially coherent PSK will be found in the references cited. The curves shown are based on ratios of carrier-to-noise (CNR) and ratios of carrier-to-interference (CIR) where carrier, noise, and interference are observed at the input to an ideal phase detector followed by a threshold device that determines which binary alternative was sent. The RF-IF filter is accounted for indirectly by specifying the mean square value of the noise at the phase detector input; thus the signal input $S(t)$ to the phase detector is written

$$S(t) = \cos[2\pi f_s t + \varphi_s(t)] \qquad (3\text{-}46)$$

where

f_s = signal carrier frequency, Hz

$\varphi_s(t)$ = signal phase which, in the binary case, is a random rectangular wave shifting 0 deg or 180 deg every t seconds

The interference $I(t)$ is written in simplified form here as

$$I(t) = b\cos(2\pi f_i t + \lambda) \qquad (3\text{-}47)$$

where

b = interference amplitude

λ = random phase angle uniformly distributed in 2π

Any phase modulation which may exist in the interference has no effect because of the randomness assumed for λ. The noise is assumed to be a Gaussian random process, with mean square value of σ^2, σ depending on the RF-IF bandwidth. If it is assumed that the signal is sent in a sequence of instantaneous phase changes lasting the baud length T, and that the noise input to the receiver is a white Gaussian random process with power spectral density (PSD) of N_0 in V²/Hz, the optimum RF-IF filter is the matched filter with impulse response $h(t)$ given by

$$\left.\begin{aligned} h(t) &= \frac{2}{T}\cos[2\pi f_s(T-t)] \quad, 0 < t < T, \mathrm{s}^{-1} \\ &= 0,\ t \text{ elsewhere} \end{aligned}\right\} \qquad (3\text{-}48)$$

The output of this filter is observed at intervals of T seconds to measure the phase, the exact instants being the ends of each baud interval. At these instants the magnitude of the signal is maximum relative to the root-mean-square value of the noise. The mean

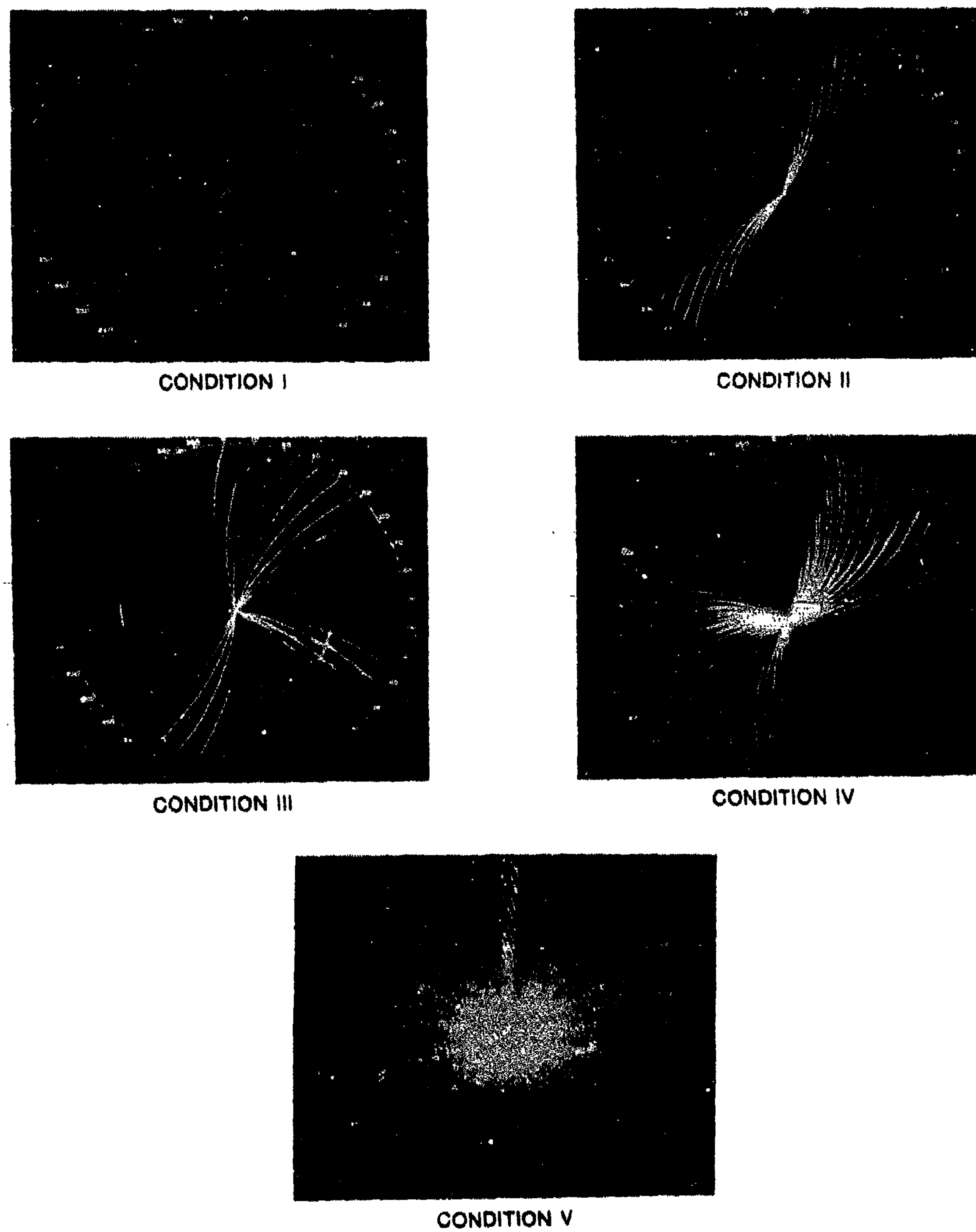


Figure 3-44. Interference Conditions of PPI Display (Ref. 45)

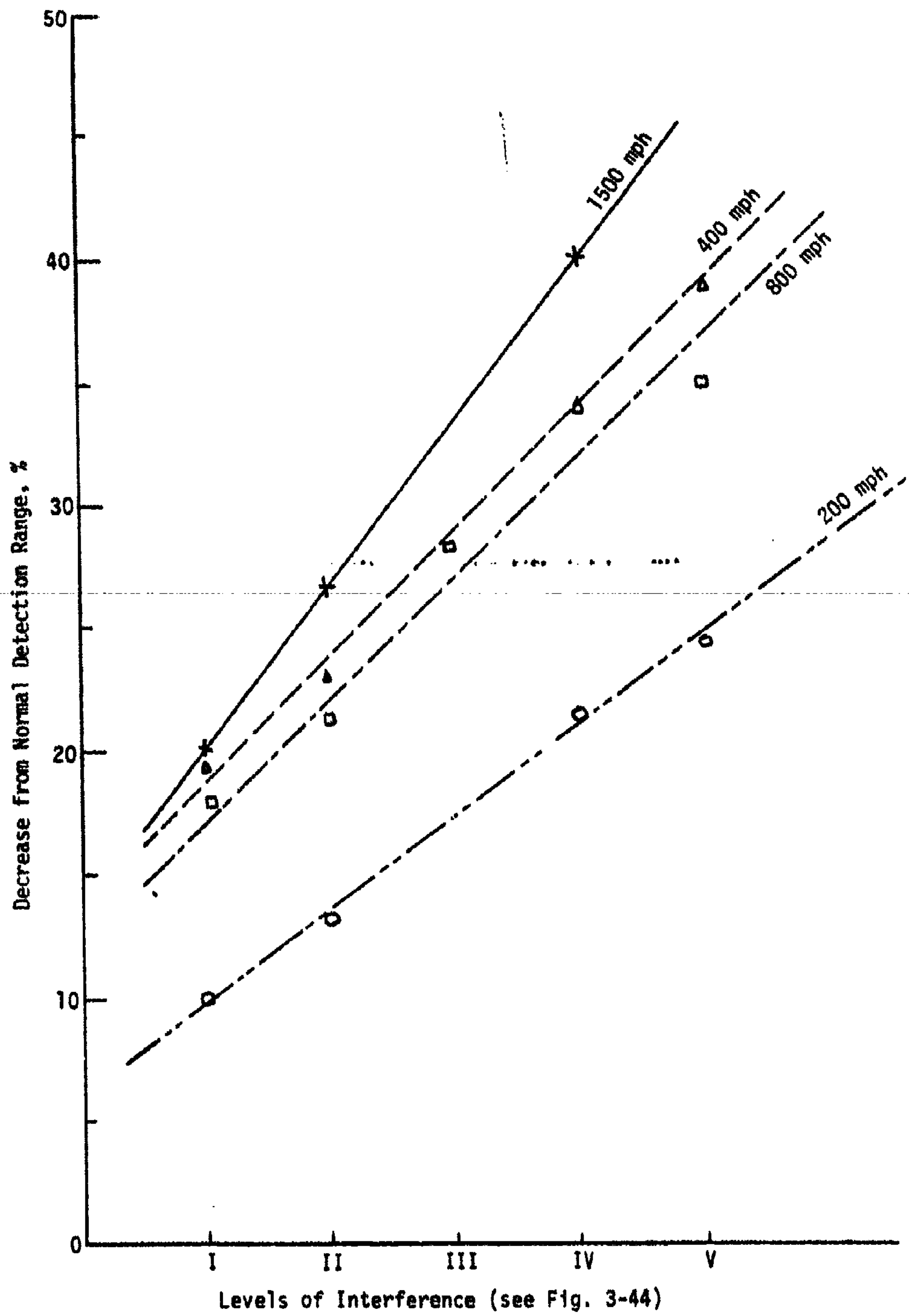


Figure 3-45. Search Radar Detection Range vs Interference Level (Ref. 48)

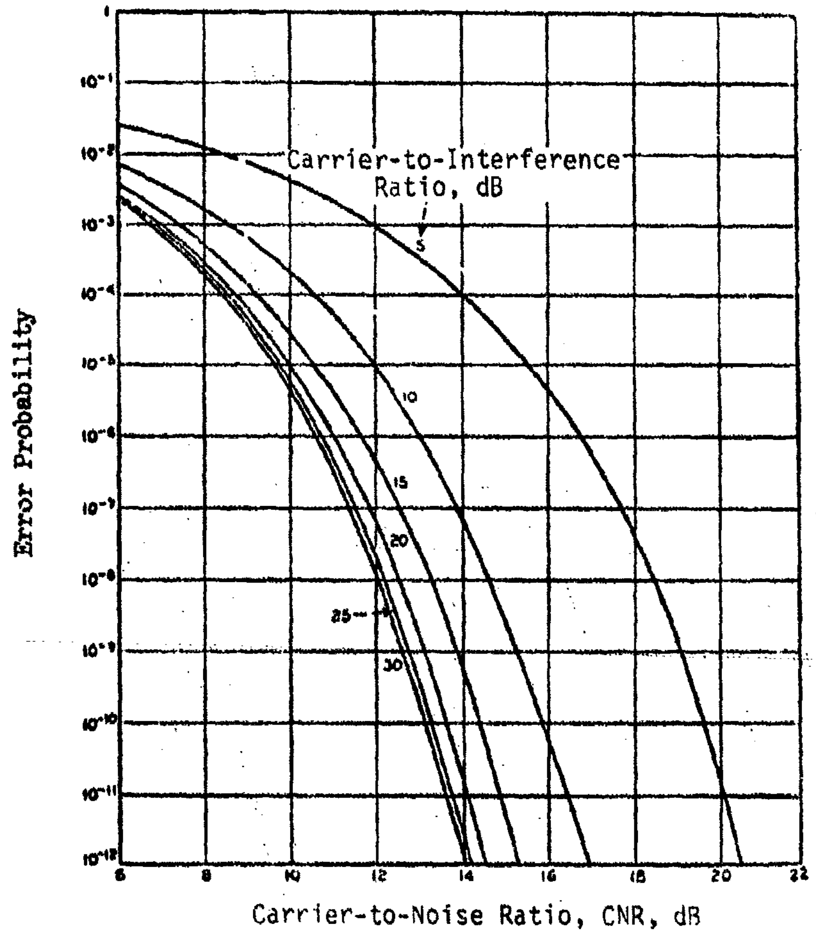


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Figure 3-46. Binary Error Probability vs CNR — Coherent Detection

square value σ^2 of the noise at these instants of observation (and at other times as well) can be shown to be

$$\sigma^2 = \frac{2N_0}{T}, \text{V}^2 \qquad (3\text{-}49)$$

where

N_0 = power spectral density of white noise, V^2/Hz

In terms of the effective bandwidth B_e of the receiver, the mean square value is given by

$$\sigma^2 = 2N_0B_e, \text{V}^2 \qquad (3\text{-}50)$$

The gain of the RF-IF amplifier defined by Eq. 3-48 is normalized to unity at the signal frequency and $1/T$ is the effective bandwidth of the matched filter.

The RF-IF filter, matched or not, affects the interference component of the input. If this component is a pure sinusoid, the amplitude will only change by a constant amount; b in Eq. 3-47 is the amplitude after the RF-IF filter. If the interference component contains phase modulation, the RF-IF output fluctuates in amplitude — the degree of fluctuation depending on the relative bandwidths of the interference and the RF-IF filter. For the purposes here, assume that the interference bandwidth is small relative to the bandwidth of the RF-IF amplifier, so that fluctuation is negligible.

The carrier-to-noise ratio (CNR) in dB in Fig. 3-46 is

$$10 \log (1/\sigma^2) = 20 \log (1/\sigma) \text{ , dB} \qquad (3\text{-}51)$$

since the carrier level is taken to be unity at the phase detector input. The carrier-to-interference (CIR) level in dB is

$$20 \log (1/b) \text{ , dB} \qquad (3\text{-}52)$$

The effect of a given mean square value of interference at the phase detector input is not as severe as the effect of random noise of the same mean square value. This is generally true of all the systems analyzed in the papers cited and is shown specifically for the binary case by example. Suppose both CNR and CIR are 15 dB. From Fig. 3-46 the error rate is found to be somewhat higher than 10^{-12}. If the interference were random noise, the CIR would be infinite and the CNR would be decreased by 3 dB to a value of 12 dB. Fig. 3-46 contains no curve for CIR = ∞, but it is adequate to assume CIR = 30 dB. The error probability shown is about 10^{-8}, a substantial increase. The increase in error probability is not so dramatic at lower CNR; it will in fact be found that for low values of CNR, the effect of random noise is only somewhat greater than it is for interference.

Since analysis of the effects of arbitrary interference on the many kinds of systems apt to be found is neither known nor easy to compute, the foregoing discussion suggests that the effect of interference be approximated by assuming it to be random noise of the same power. As a rule, this approach will result in a pessimistic estimate of error probability. In cases where the interference is broadband relative to the bandwidth of the RF-IF amplifier, the output of the latter will look more nearly like Gaussian noise to the interference input, and as a result the approximation will be close. If the interference is a composite of several independent signals, as happens in a heavily used frequency division multiplex system where interference is the result of intermodulation, the composite is approximately Gaussian.

Analyses of error probability in Gaussian noise are available for virtually all known communication systems. Fig. 3-47, abstracted from Ref. 54, gives error probability in terms of CNR for several commonly used binary communications systems. CNR is measured at the output of the RF-IF amplifier at instants spaced at baud intervals, for which it is maximum. For the nonbinary case, in addition to the results in noise and interference given by Rosenbaum (Refs. 51 and 52), results in Gaussian noise will be found in Sakrison (Ref. 55), Lindsey (Ref. 56), and Viterbi (Ref. 57).

The discussion thus far has been limited to the single digit error probability. For some purposes, such as where error control coding is used, it is necessary to know the error burst properties; i.e., the probability of a number of errors in a block of digits. If the probabilities of error in successive digits are independent, the burst error statistics can be obtained from the single digit error probabilities. Where they are not independent, joint probabilities of error in different digit positions are required. Independence exists in the case of white Gaussian noise but not generally otherwise. In the case of narrowband interference where there is a high degree of correlation in the interference from one signal digit to the next, there is a tendency for multiple errors. It is worth pointing out that in cases of high correlation, it is possible, in principle, to improve the signal detection process by tracking the undesired interference and subtracting in from the received signal. Also, some systems such as differentially coherent PSK have an inherent tendency to generate double errors in pairs of digits because the detection process compares the signals received in the adjacent digit positions.

3-2.1.4 Acceptance Ratios

The results cited in pars. 3-2.1.1 through 3-2.1.3 were based on signal and interference levels as they appear to the recipient. In the case of digital systems, signal and noise levels were those found at the input of the device making the digital decision. It is more directly useful to know the relationship between output quality and signal and interference, at the points of entry of signal and interference. This requires that signal and interference be traced through the device from their points of entry to the output, and then related to the interference effect at the output. This process is difficult to carry out accurately, except for interfering signals that are simple deterministic waveforms. Typical interfering waveforms which are modulated random processes are not, as a rule, analytically tractable. Where particular kinds of interference are expected, and points of entry are known, experimental determination of the acceptable ratio of signal and interference levels is in order.

Attempts at determining such ratios by analytical methods have been made by Schwartz (Ref. 58) and by Tierzza, Mayher, and Pressman (Ref. 59). Acceptance ratios for various radio communication systems exposed at the antenna to co-channel interference by other systems of a similar kind are given in Table 3-7. This table, originally published in a NASA handbook (Ref. 60), is based on Refs. 58 and 59 and on

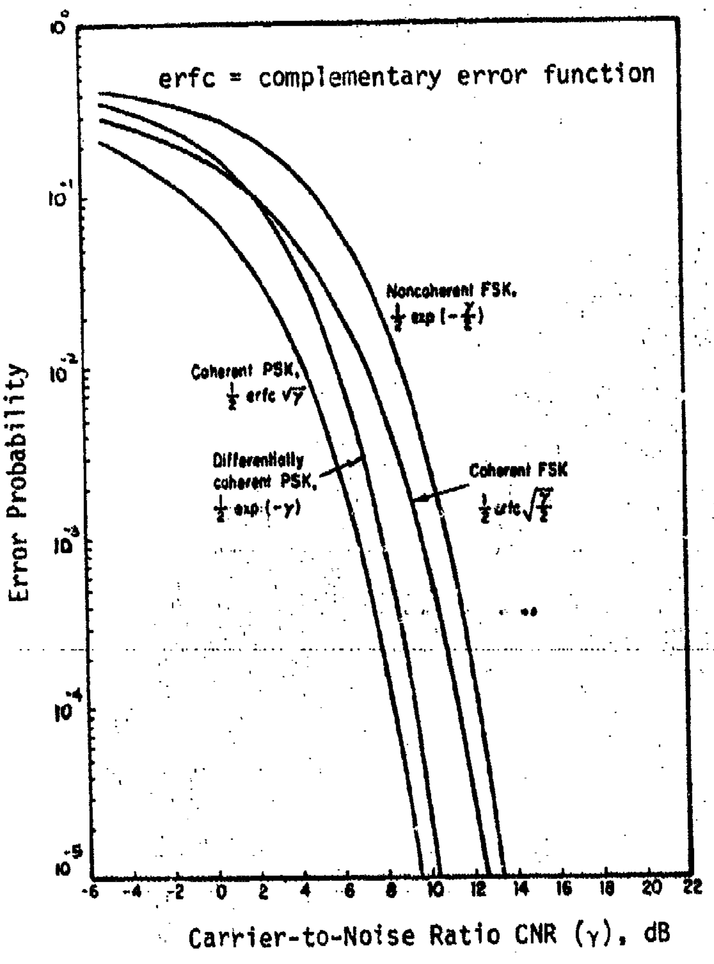


Figure 3-47. Error Rates for Several Binary Systems

results obtained informally from Georgia Institute of Technology (entries in Table 3-7 from this source are denoted G). The figures shown were determined using rough approximations and intuition wherever no other approach would be manageable. The percent figures given in each box represent the "acceptability" criterion at the output, and are to be interpreted loosely as the degree of intelligibility. The figures in dB in each box are the acceptance ratios in rms units except for interference to pulsed radar (denoted "Pulse" in the Table) for which the ratios are of peak amplitude levels. Furthermore, the criterion for interference on the display was assumed to be equal signal and noise peaks. Users of Table 3-7 are urged to consult the original sources to assure themselves that the figures given are adequate for their purposes.

3-2.1.5 Synchronization Error

Timing information for detection of pulsed and digital signals is obtained from the received waveform by filtering or tracking a component whose frequency is related to the digital rate. The component is extracted frequently using a phase-locked loop which, in its steady tracking condition, is equivalent to a narrow-band filter centered on the received sinusoid.

The output v_o of the tracker-filter (when referred to the receiver input level) is of the form

$$v_o(t) = S \cos \omega_d t + n(t) \quad , \text{V} \qquad (3\text{-}53)$$

where

S = the input signal amplitude, V
ω_d = 2π × digital data rate, s^{-1}
$n(t)$ = added interference, V
t = time, s

With no interference, the zero crossings of the cosine wave provide perfect timing information. In the presence of noise, the zero crossings are not perfectly periodic and the deviations from unperturbed zero crossing times τ contribute to the phenomena of jitter in the detection process.

If $n(t)$ is flat and Gaussian over its filter bandwidth, the mean square timing error is

$$\left\langle \tau^2 \right\rangle = \frac{2 N_o B_e}{S^2 \omega_d^2} \qquad (3\text{-}54)$$

where

τ = deviation in zero crossing time from unperturbed value, s
N_o = input noise power spectral density, V^2/Hz
B_e = effective noise power bandwidth, Hz
S = input signal level, V

Narrowing the filter-tracker bandwidth will reduce the jitter. However, it also results in sluggishness of the tracker thus making it slow in acquiring the correct timing, and slow to respond to variations in the transmitted timing. The transient behavior of phase-locked loops is discussed in specialized books by Viterbi, Lindsey (Refs. 57, 61), and in less detail in many general books on communications.

The effect of jitter is an increase in bit error rate. This occurs because the decision on the signal transmitted in one signal interval is being made by sampling the signal only partly in that interval and partly in a neighboring interval.

In systems depending on coherence and timing, noise and interference which substantially disrupt synchronism destroy the signal virtually completely. In many systems there are different levels of synchronism which must be maintained. The system may depend on RF coherence such as coherent phase shift keying and single-sideband transmission for nonvoice signals. It may require bit timing and various higher levels of timing. Pulse code modulation, for example, requires bit and word timing. The conventional TV system depends on line timing (horizontal synchronism) and frame timing (vertical synchronism). Protection of synchronizing circuits is of great importance and, in a well-designed receiver, total failure of these circuits in noise should come only after degradation due to the normal effect of additive noise is large.

3-2.2 ADMISSION MECHANISMS

Devices susceptibility to undesired inputs are classified according to the mechanism of unwanted signal intrusion as follows:

a. Linear intrusion via normal input terminals
b. Nonlinear intrusion via normal input terminals
c. Intrusion through ports not intended as signal inputs.

In this paragraph the first two items are treated in detail. The third mechanism is covered in par. 3-3. The discussion that follows is oriented to a typical communication receiver; however, the principles can be applied to any device for which a normal signal acceptance bandwidth can be defined, and which also may contain nonlinear elements.

The block diagram of Fig. 3-48 shows the essential elements of a receiver. The acceptance band of the receiver and the spectrum of an adjacent-channel signal are both shown in Fig. 3-49. In the linear intrusion mode, the receiver acts as a normal bandpass filter that accepts any input containing frequency components in the receiver passband, as indicated in Fig. 3-49. Unwanted inputs, whose spectrum is centered at or near the frequency to which the RF filter is tuned, originate from communication systems or from other noise sources that cause interference. In the second mode, called the nonlinear intrusion mode, unwanted signal energy that lies outside the normal passband of the receiver acts on a nonlinear element in such a way as to enable the receiver to accept undesired signals. The RF filter in Fig. 3-48 is a preselector network that limits the frequency band of energy passing through the succeeding active elements. These elements nearly always have some residual nonlinear properties that play a significant role when the input amplitudes are large. When the RF filter is inadequate to limit large out-of-band inputs to a satisfactory low level, the nonlinear devices (vacuum tubes, transistors, and diodes) will generate frequency components not originally present. Interference can occur when these new frequency components are within the passband of the portion of the receiver following the nonlinear device. Phenomena typical of nonlinear intrusion include single spurious response, multiple spurious response (intermodulation), and sideband transfer (cross modulation).

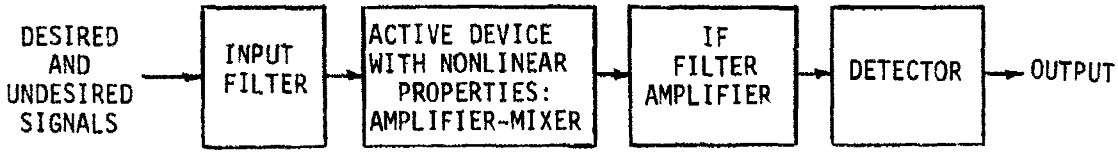


Figure 3-48. Block Diagram of Basic Receiver Elements

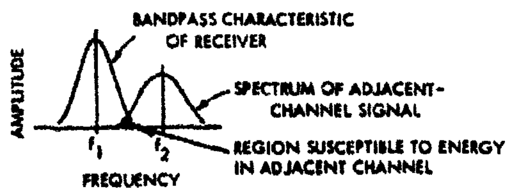


Figure 3-49. Interference Produced by a Signal in an Adjacent Channel

3-2.2.1 Linear Intrusion

Unwanted signals fall into one of the following categories:

a. Broadband noise arising from natural or man-made sources

b. Signals from communication or noncommunication sources designed to act as generators at or near the frequency to which the receiver is tuned, and confined to a limited bandwidth.

3-2.2.1.1 Broadband Noise

Broadband noise usually is characterized as random or impulsive. Random noise can be best described by its power spectral density (Ref. 62). The mean-square value of the noise v_o admitted through an amplifier of a receiver is

$$\langle v_o^2 \rangle = \int_0^\infty N_i(f)\,|H(f)|^2 df \text{ , V}^2 \quad (3\text{-}55)$$

where

$N_i(f)$ = power spectral density of the noise at the receiver input, V^2/Hz

$H(f)$ = complex transfer characteristic from the receiver input to the IF amplifier output, dimensionless

$H(f)$ may be expressed in terms of an amplitude $|H(f)|$ and a phase characteristic $\varphi(f)$, according to

$$H(f) = |H(f)| \exp[j\varphi(f)] \text{ , d'less} \quad (3\text{-}56)$$

If the noise spectral density is flat over the passband, and if the center frequency of the IF amplifier is f_o, the mean-square value of the IF amplifier output is, using Eq. 3-3,

$$\langle v_o^2 \rangle = N(f_o)\,|H(f_o)|^2 B_e \text{ , V}^2 \quad (3\text{-}57)$$

If impulses applied at the input are spaced sufficiently in time so that they do not overlap in passing through the receiving device, they are best described in terms of their Fourier spectrum.

The instantaneous output voltage $v_o(t)$ of an amplifier to an input that has a Fourier transform $V(f)$ in volt-seconds is

$$v_o(t) = \int_{-\infty}^{\infty} V(f)\,H(f) \exp(j2\pi ft)\,df \text{ , V} \quad (3\text{-}58)$$

If $V(f)$ is virtually constant over the passband at a value $V(f_o)$, then using the amplitude and phase notation for $H(f)$, this reduces to

$$v_o(t) = 2V(f_o) \int_0^\infty |H(f)| \cos[2\pi ft + \varphi(f)]\,df \text{ , V} \quad (3\text{-}59)$$

If $|H(f)|$ has even symmetry around the center frequency f_o and $\varphi(f)$ has odd symmetry and is linear around f_o, the peak value of the output is

$$V_{pk} = 2V(f_o)|H(f_o)|B_i = V(f_o)|H(f_o)|B_i \text{ , V} \quad (3\text{-}60)$$

where

$V(f_o)$ = spectrum amplitude at f_o, V·s

$$B_i = \int_0^\infty \frac{|H(f)|}{|H(f_o)|}\,df \quad (3\text{-}61)$$

, the effective impulse bandwidth of the IF amplifier, Hz

From the given relations, and a knowledge of the input signal level, the signal-to-noise ratio at the input to any detector used in a susceptible device can be determined. The output signal-to-noise ratio depends upon the detection process used in the susceptible receiver or device, and the exact waveforms of the modulating signal and the noise.

3-2.2.1.2 Interference from Sources Intended as Generators

3-2.2.1.2.1 Co-channel Interference

The term "co-channel interference" designates interference that involves communication systems that have been assigned equal, or nearly equal, carrier frequencies. Co-channel frequency assignments ordinarily are made when the probability of the simultaneous encounter of signals from the two systems is insignificant. Such systems are separated physically by large distances or do not operate at the same time. Sometimes co-channel interference will arise because of unusual propagation conditions or because co-channel sources operate under conditions for which they are not intended to operate.

The term "adjacent-channel interference" designates interference between communication systems that have been assigned neighboring channels. Channel-spacing policy varies, but the term "adjacent" is used to mean channel separation by a frequency difference greater than the average of the two signal bandwidths. Fig. 3-49 illustrates interference of this kind; energy on the skirt of the adjacent-channel signal spectrum is shown overlapping the bandpass characteristic of the receiver. Although, in the typical case, the receiver skirt sensitivity is low, compared to the in-band sensitivity, receivers located close to an adjacent-channel transmitter can be exposed to very large magnitudes of unwanted signals.

Estimates of interference arising from linear intrusion can be made in several ways without excessive numerical complexity by treating the unwanted signal as: (1) a pure sinusoid, (2) a broadband waveform perfectly centered in the band, (3) a broadband waveform whose center frequency is sufficiently removed from the frequency to which the receiver is tuned so that the unwanted spectrum is nearly constant over the receiver band, or (4) a band-limited waveform falling on a small portion of the receiver selectivity curve. The first of these is useful for the estimation of both co-channel and adjacent-channel effects. The second is appropriate for the evaluation of co-channel interference, and the third is appropriate for the determination of adjacent-channel interference. In the third case, the bandpass filter is exposed to a portion of the one sideband of the unwanted signal. The effect is not much different from that produced by thermal noise having equal mean-square value. Therefore, the spectral density of the unwanted signal is estimated at the center of the reception band and, as in the case of nonperiodic broadband noise previously discussed, the mean-square value of IF amplifier output is determined by use of Eq. 3-56. The fourth way of estimating interference arising from linear intrusion is illustrated in par. 3-2.2.1.2.3.

3-2.2.1.2.2 Receiver IF Channel Interference

Penetration of unwanted signals that are centered at one of the IF channels within a receiver is a somewhat different mechanism involving linear phenomena. For instance, a large-amplitude signal centered at the frequency of one of the IF amplifiers may manage to pass through the input selective RF circuits to the IF amplifier in question. Once there, it proceeds through the rest of the receiver in the normal manner. To overcome this difficulty, the selectivity of the input RF circuit and/or stray paths to the sensitive circuits must be controlled. As would be expected, the most susceptible frequency is that of the first IF amplifier, but consideration needs to be given to all succeeding IF amplifiers as well.

3-2.2.1.2.3 Adjacent-channel Interference

The level of interference arising from an unwanted signal in a band adjacent to the desired band can be estimated in the manner that follows.

Assume the receiver selective circuits consist of n identical single-tuned parallel RLC circuits, with bandwidths a small fraction of the center frequency f_o. In terms of the 3-dB bandwidth, B_3* of the passband, the relative response $H_n(f)$ is

$$H_n(f) = \left[1 + \frac{4(f-f_o)^2\,(2^{1/n}-1)}{B_3^2}\right]^{-n/2}, \text{ d'less} \qquad (3\text{-}62)$$

When the receiver is tuned to the frequency $f_o = f_c$, the power P_i of an unmodulated interfering signal at frequency f_b at the output of the selective circuits is

$$P_i = |H_n(f_b)|^2 P_b \text{ , W} \qquad (3\text{-}63)$$

* The 3-dB bandwidth of n-identical, single-tuned circuits in cascade is given by

$$B_3 = B\sqrt{2^{1/n}-1}$$

where B is the 3-dB bandwidth of one single-tuned circuit (Ref. 63). Similarly, the 6-dB bandwidth is given by

$$B_6 = B\sqrt{4^{1/n}-1}$$

where

P_b = average interfering signal power, W

If the desired signal is concentrated near the band center and has a mean square value of P_a, the output signal-to-interference power ratio S/I is

$$\frac{S}{I} = \frac{P_a}{|H_n(f_b)|^2 P_b}, \text{ d'less}$$

$$= \frac{P_a}{P_b}\left[1 + \frac{4(f_b - f_a)^2 (2^{1/n} - 1)}{B_3^2}\right]^n, \text{ d'less} \quad (3\text{-}64)$$

If $n = 3$, the frequency separation is only one bandwidth $[(f_b - f_a)/B_3 = 1]$ and both signals are of equal power, $S/I = 4$. If the separation is equal to 2 bandwidths, $S/I = 27$.

If the interfering signal is single-sideband modulated with broadband noise with the power P_b distributed uniformly over a bandwidth ΔB centered about the interfering frequency f_b, the power spectrum of the undesired signal and the receiver response curve are shown in Fig. 3-50 where the power spectral density of the interfering signal is ϕ_i, W/kHz. With the receiver tuned to the frequency f_a, the power of the unwanted interference that gets through the passband is given by

$$P_i = \int_0^\infty \phi_i |H_n(f)|^2 df, \text{ W} \quad (3\text{-}65)$$

Since ϕ_i is assumed constant over the interval*

$$\left(f_b - \frac{\Delta B}{2}\right) < f < \left(f_b + \frac{\Delta B}{2}\right)$$

and is assumed equal to zero elsewhere, the interfering power P_i is

$$P_i = \phi_i \int_{\left(f_b - \frac{\Delta B}{2}\right)}^{\left(f_b + \frac{\Delta B}{2}\right)} \left[1 + \frac{4(f - f_a)^2 (2^{1/n} - 1)}{B_3^2}\right]^{-n} df, \text{ W} \quad (3\text{-}66)$$

By an appropriate change of variable, the integral in Eq. 3-66 can be reduced to a standard integral (e.g., Ref. 2, p. 44-20) of the form

$$c\int_{a_f}^{b_f} (ay^2 + 1)^{-n}\, dy \quad (3\text{-}67)$$

where

$$c = \phi_i B_3$$

$$a = 4[2^{1/n} - 1]$$

$$b = 1$$

$$y^2 = \frac{(f - f_a)^2}{B_3^2}$$

$$dy = \frac{df}{B_3}$$

$$a_f = \frac{f_b + \frac{\Delta B}{2} - f_a}{B_3} = \text{a constant}$$

$$b_f = \frac{f_b - \frac{\Delta B}{2} - f_a}{B_3} = \text{a constant}$$

* It will sometimes be sufficiently correct to assume that the entire unwanted signal band is attenuated uniformly by the relatively flat portion of the tail of the receiver bandpass characteristic; i.e., it will be possible to let

$$H_n(f) = H_n(f_b)$$

for all values of f in the band ΔB.

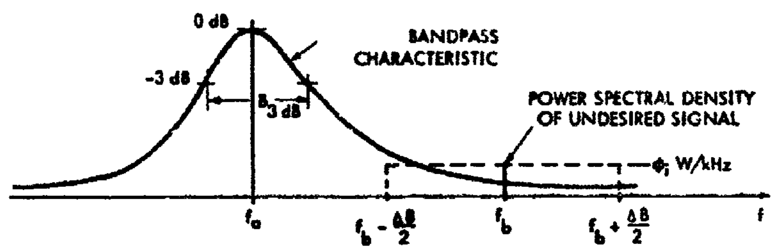


Figure 3-50. Unwanted Signal Spectrum and the Receiver Bandpass Characteristic

If, for example, the selective circuits consist of three single-tuned stages, i.e., $n = 3$, $a \approx 1.0$. For $n = 3$, the expression for P_i, from Eq. 3-66, is

$$P_i = \phi_i B_3 \left\{ \frac{a_f}{4(a_f^2+1)^2} + \frac{3}{4}\left[\frac{a_f}{2(a_f^2+1)} + \frac{1}{2}\mathrm{Tan}^{-1}(a_f)\right] - \frac{b_f}{4(b_f^2+1)^2} - \frac{3}{4}\left[\frac{b_f}{2(b_f^2+1)} + \frac{1}{2}\mathrm{Tan}^{-1}(b_f)\right]\right\} \quad (3\text{-}68)$$

Following the assumption that the unwanted signal power is uniformly distributed over its band, the quantity ϕ_i is obtained from the average signal power by

$$\phi_i = \frac{P_b}{\Delta B} \quad (3\text{-}69)$$

and the signal-to-interference ratio is then

$$\frac{S}{I} = \frac{P_a}{P_i} \quad (3\text{-}70)$$

As a numerical example, suppose

$$P_a = P_b = -115\ \text{dBm}\ (= 3 \times 10^{-15}\ \text{W})$$
$$\Delta B = 30\ \text{kHz}$$
$$B_3 = 30\ \text{kHz}$$

and

$$f_b - f_a = 25\ \text{kHz}$$

then

$$a_f = 1.33,\ b_f = 0.33 \text{ and from Eq. 3-68}$$
$$P_i = (3 \times 10^{-15})(0.267) = 8 \times 10^{-16}\ \text{W}$$
$$= -121\ \text{dBm}$$

thus the value of P_i turns out to be -121 dBm or 6 dB below the desired signal power.

3-2.2.2 Nonlinear Intrusion

Nonlinear effects arise as a result of inadequate rejection of the unwanted signal in the input filter circuits of the receiver, followed by some nonlinear process in an electronic device. Less common nonlinear admission mechanisms include the effects of imperfect joints between conductors prior to the receiver filter circuits.

3-2.2.2.1 Spurious Responses

Spurious responses of a receiver result from: (1) nonlinearity in an early stage, which gives rise to harmonics of incoming signals; (2) nonlinearity in the mixer, which results in oscillator and signal harmonics; and (3) frequency multiplication in the local oscillator and its related circuits. At each frequency to which the receiver is tuned, a specific set of possible spurious response frequencies exists, and each of these sets has its own level of significance (Ref. 64).

The nonlinear device — such as a transistor, diode, or vacuum tube — has an input-x output-y characteristic that may be specified by the power series

$$y = \sum_{n=0}^{N} a_n x^n \quad (3\text{-}71)$$

where

a_n = a constant

The mixing operation occurs with the simultaneous application of a signal voltage $x_s(t)$, and oscillator voltage $x_o(t)$

$$x_s(t) = v_s(t)\cos(\omega_s t + \varphi_s)\ ,\text{V} \quad (3\text{-}72)$$

$$x_o(t) = A\cos\omega_o t\ ,\text{V} \quad (3\text{-}73)$$

where

$v_s(t)$ = signal amplitude modulation function, V
ω_s = angular frequency ($2\pi f_s$) of signal carrier, rad/s
f_s = signal carrier frequency, Hz
φ_s = phase angle, rad
ω_o = local oscillator angular frequency, rad/s
A = local oscillator amplitude, V

Then $x = x_s - x_o$ and

$$y(t) = \sum_{n=0}^{N} a_n[x_s(t) + x_o(t)]^n = \sum_{n=0}^{N} a_n \times \sum_{n=0}^{N}\binom{n}{k} x_s^k(t)\cos^k(\omega_s t + \varphi_s)A^{(n-k)}\cos^{(n-k)}\omega_o t \quad (3\text{-}74)$$

The periodic cosine functions can be expanded (Refs. 65 and 66) in a Fourier series so that, when Eq. 3-74 is written as a sum of individual cosine terms, the result contains all frequencies

$$q\omega_s \pm p\omega_o$$

where

$q = 0, 1, 2, ..., N$
$p = 0, 1, 2, ..., N$

Whenever one of these frequencies coincides with the intermediate angular frequency ω_{if}, a potential spurious response exists; i.e., any input frequency ω_s that satisfies the equation

$$\omega_s = \frac{|p\omega_o \pm \omega_{if}|}{q}, \text{ rad/s} \quad (3\text{-}75)$$

with any combination of the signs, leads to a potential spurious frequency.

Frequently, the amplitude of the signal component $x_s(t)$ is small compared to that of the oscillator component $x_o(t)$, and terms involving $x_s^k(t)$ can be ignored when $k > 1$. The significant portion of Eq. 3-74 is, then,

$$y_1(t) = \sum_{n=1}^{N} na_n x_o^{(n-1)}(t)x_s(t) = g(t)x_s(t) \quad (3\text{-}76)$$

The quantity $g(t)$ is the transconductance as a function of time when an oscillator voltage, $x_o(t)$ is applied, i.e., the transconductance is the derivative of Eq. 3-71, or

$$g = \frac{dy}{dx} = \sum_{n=1}^{N} na_n x^{(n-1)} \quad (3\text{-}77)$$

so that with $x = x_o(t) = A\cos\omega_o t$ as in Eq. 3-73,

$$g(t) = \sum_{n=0}^{N} na_n A^{(n-1)}\cos[(n-1)\omega_o t]$$
$$= \sum_{n=0}^{N-1} g_n \cos^n \omega_o t \quad (3\text{-}78)$$

where

g_n = nth order conversion transconductance, mho

The last sum on the right of Eq. 3-78 is the form that is obtained when $\cos[(n-1)\omega_o t]$ is expanded and all terms of the same harmonic are collected. Thus, for Eq. 3-76, using Eq. 3-72,

$$y_1(t) = v_s(t)\sum_{n=0}^{N}\frac{g_n}{2}\{\cos[(\omega_s - n\omega_o)t + \varphi_s]$$
$$+ [(\omega_s + n\omega_o)t + \varphi_s]\} \quad (3\text{-}79)$$

i.e., frequencies $\omega_s \pm n\omega_o$ will be obtained. The quantity $g_n/2$ is the conversion transconductance corresponding to the nth oscillator harmonic. If $g(t)$ is a cosine function at the frequency ω_o (i.e., if g versus x is a straight line over the region of oscillator swing), then the only output frequencies are $\omega_s \pm \omega_o$. From the viewpoint of minimizing interference, those electronic devices which can operate close to this ideal over a portion of the frequency range should be restricted to that range. However, designers frequently do not, or cannot easily, control the level of oscillator voltage applied to the mixer. Maximum conversion transconductance at the desired frequency (i.e., with $n = 1$) is obtained with a large oscillator input, and this often results in more than a proportionate increase in the conversion transconductance at undesired frequencies (i.e., with $n > 1$). Also, the output of a variable-frequency oscillator is rarely constant over an appreciable range of frequencies; the conversion gain generally varies over the band.

When the mixer is a diode, as it often is in microwave receivers, the mixing of the signal with a harmonic of the local oscillator ordinarily cannot be avoided. The signal voltage is multiplied by a square-wave switching function containing all odd harmonics of the oscillator frequency, so that harmonic mixing with all odd oscillator harmonics is unavoidable. The current-voltage characteristic i of the diode given by

$$i = I_s[\exp(bv) - 1], \text{ A} \quad (3\text{-}80)$$

where

I_s = reverse saturation current, A
b = a constant which, in theory, is $e/(kT)(\approx 40\ \text{V}^{-1}$ at 300 K)
v = applied voltage, V
e = electronic charge, 1.602×10^{-19} C
k = Boltzmann constant, 1.38×10^{-23} J/K
T = temperature, K

The transconductance is

$$g_m = \frac{di}{dv} = bI_s\exp(bv), \text{ mho} \quad (3\text{-}81)$$

so that when the applied voltage is

$$v = A\cos\omega_o t, \text{ V}$$

The time varying transconductance is

$$g_m(t) = bI_s \exp(bA\cos\omega_o t)$$
$$= bI_s\left[I_0(bA) + 2\sum_{n=1}^{\infty} I_n(bA)\cos n\omega_o t\right] \quad (3\text{-}82)$$

In this expression, $I_n(bA)$ is the modified Bessel function of the first kind of order n ($n = 0, 1, 2, ...$) and of argument (bA) (Ref. 65). The conversion transconductance, as defined by Eq. 3-79, is

$$\frac{g_n}{2} = bI_sI_n(bA)\ \text{, mho} \quad (3\text{-}83)$$

From Bessel function theory it will be found that for values of (bA) apt to be used here (ranging around 10), values of $I_n(bA)$ up to about $n = 6$ are of the same order of magnitude; i.e., harmonic conversion will be very significant for the sixth harmonic of the oscillator. For higher values of the argument (bA), the value of $I_n(bA)$ becomes significant for even greater values of n.

The interference-to-signal ratios are not calculated easily, because the gain or loss of amplitude of the undesired signal between the input and the point at which the nonlinear effect takes place is not ordinarily known. Furthermore, the oscillator level and harmonic and subharmonic content, as a function of the frequency to which the receiver is tuned, ordinarily are not known. It is more common to measure the intensity of the spurious responses than to calculate them. The usual procedure is to set the tuning control to three frequencies in each band, one at the center and the others in the vicinity of the band extremes. With the receiver fixed at each frequency, an input signal is applied from a generator that is tuned through the frequencies of potential response. A desired signal, modulated or unmodulated, may be applied simultaneously. The observed quantity is the ratio of signal input voltage to interference voltage required to give a stated output. This ratio may depend on the input level. The intensity of the spurious responses, nevertheless, can be calculated in many cases, especially when only approximate values are needed. An example of calculated spurious responses intensities follows.

Let $N = 3$ in Eq. 3-71. The collection of signals produced in the nonlinear device is found by expanding Eq. 3-74 to obtain

$$\left.\begin{aligned}
y(t) = {} & a_o + a_1 v_s(t)\cos(\omega_s t + \varphi_s) \\
& + a_1 A\cos\omega_o t \\
& + \frac{1}{2}a_2[v_s^2(t) + A^2] \\
& + \frac{1}{2}a_2 v_s^2(t)\cos 2(\omega_s t + \varphi_s) \\
& + \frac{1}{2}a_2 A^2\cos 2\omega_o t \\
& + a_2 v_s(t) A\cos[(\omega_s - \omega_o)t + \varphi_s] \\
& + a_2 v_s(t) A\cos[(\omega_s + \omega_o)t + \varphi_s] \\
& + \left[\frac{3}{4}a_3A^3 + \frac{3}{4}a_3 A v_s^2(t)\right]\cos\omega_o t \\
& + \left[\frac{3}{4}a_3 v_s^3(t) + \frac{3}{4}a_3 A^2 v_s(t)\right] \\
& \quad \times\cos(\omega_s t + \varphi_s) \\
& + \frac{1}{4}a_3 v_s^3(t)\cos 3(\omega_s t + \varphi_s) \\
& + \frac{1}{4}a_3 A^3\cos 3\omega_o t \\
& + \frac{3}{4}a_3 v_s^2(t)\ A\cos[(2\omega_s - \omega_o)t + 2\varphi_s] \\
& + \frac{3}{4}a_3 v_s^2(t)\ A\cos[(2\omega_s + \omega_o)t + 2\varphi_s] \\
& + \frac{3}{4}a_3 v_s(t)\ A^2\cos[(\omega_s - 2\omega_o)t + \varphi_s] \\
& + \frac{3}{4}a_3 v_s(t)\ A^2\cos[(\omega_s + 2\omega_o)t + \varphi_s]
\end{aligned}\right\} \quad (3\text{-}84)$$

From the collection of components in Eq. 3-84 consider the terms for which the angular frequency of the interfering or undesired signal, now written ω_{SI}, is see Eq. 3-75, and ω_{if} is an intermediate angular frequency.

$$|\omega_{SI} - 2\omega| = \omega_{if}\ \text{, rad/s} \quad (3\text{-}85)$$

The receiver is tuned to the desired component, whose frequency is

$$\omega_{SD} - \omega_o = \omega_{if}\ \text{, rad/s} \quad (3\text{-}86)$$

where ω_{SD} is the desired angular frequency.

The intensity of a spurious response at frequency ω_{SI} when receiving a desired signal of frequency ω_{SD} is determined from the coefficients of the appropriate frequency terms given in Eq. 3-84. Suppose for example that $\omega_{SD} = 140$ MHz, $\omega_o = 110$ MHz, and $\omega_{if} = 30$ MHz. Then, the frequency of the interfering signal $\omega_{SI} = 2\omega_o \pm \omega_{if}$ has the two values of 250 MHz and 190 MHz. From Eq. 3-84, the desired component D at the output of the mixer has an angular frequency of $\omega_s - \omega_o$ and is given by

$$D = a_2 v_s(t)\, A \cos\left[(\omega_s - \omega_o)t + \varphi_s\right] \text{ , V} \qquad (3\text{-}87)$$

The peak voltage of this component is

$$D = a_2 A V_{SD} \text{ , V} \qquad (3\text{-}88)$$

where

V_{SD} = peak amplitude of $v_s(t)$, the envelope of the desired signal at the mixer input, V

The undesired or interfering component I has angular frequency satisfying $\omega_{if} = |\omega_{SI} - 2\omega_o|$ and is given by

$$I = \frac{3}{4} a_3 v_s(t)\, A^2 \cos\left[(\omega_s \pm 2\omega_o)t + \varphi_s\right] \text{ , V} \qquad (3\text{-}89)$$

Its peak value is

$$I = \frac{3}{4} a_3 V_{SI} A^2 \text{ , V} \qquad (3\text{-}90)$$

where

V_{SI} = peak amplitude of $v_s(t)$, now the envelope of the undesired signal at the mixer input.

At the output of the mixer, the ratio of the level of the desired signal to that of the interfering or undesired signal is, therefore,

$$\frac{D}{I} = \frac{4a_2 V_{SD}}{3a_3 V_{SI} A} \text{ , d'less} \qquad (3\text{-}91)$$

However, the quantity of interest is the ratio of the voltage of the desired signal to that of the interfering signal at the receiver RF input rather than at the input to the mixer.

The voltages given in Eq. 3-91 may be converted to voltages at the input of the receiver by taking account of the RF gain of the receiver. Let K be the voltage amplification of the interfering signal relative to that of the desired signal resulting from gain in the RF circuits preceding the mixer. Then, if V_D and V_I (in place of V_{SD} and V_{SI}) designate respectively the peak voltage of the desired and undesired signals at the input of the receiver, and $V_{SI} = KV_I$, the required ratio becomes

$$\left(\frac{D}{I}\right)_{input} = \frac{4a_2 V_D}{3a_3 AKV_I} \text{ , d'less} \qquad (3\text{-}92)$$

K represents the ratio of gains from receiver input to mixer input at the undesired signal frequency and the desired signal frequency. For example, if two single-tuned circuits each of bandwidth B are used in the RF stage, then, to an approximation good for large separation of center frequencies relative to the bandwidth

$$K = \left[\frac{B}{2(f_{SI} - f_{SD})}\right]^2 \text{ , d'less} \qquad (3\text{-}93)$$

where

f_{SI} = desired signal frequency, Hz

f_{SD} = interfering signal frequency, Hz

With $f_{SI} = 190$ MHz, $f_{SD} = 140$ MHz, and $B = 15$ MHz; $K = 0.0225$. If $a_2 = 2.6 \times 10^{-5}$ A/V² and $a_3 = 8 \times 10^{-7}$ A/V³ (Ref. 66, Section 2.11), and local oscillator level $A = 10$ V, from Eq. 3-92.

$$\frac{D}{I} = \left(\frac{4 \times 2.6 \times 10^{-5}}{3 \times 8 \times 10^{-7} \times 10 \times 2.25 \times 10^{-2}}\right)\frac{V_D}{V_I}$$

$$= 192.5\frac{V_D}{V_I}$$

For equal values of desired and undesired components, $D/I = 1$, and the input ratio must be

$$\frac{V_I}{V_D} = 192.5$$

That is, for equal signal and interference at the mixer output, the interfering signal at the receiver input has to be 20 log(192.5) = 45.7 dB above the level of the desired signal.

Fig. 3-51 shows a plot of Eq. 3-75 relating f_{SD} and f_{SI} for several values of m and n for a receiver covering a range from 100 to 200 MHz, and having an oscillator frequency f_o set 30 MHz below the tuned frequency. Measured or computed values of strength of response can be indicated on the diagram at appropriate points, as shown in one case. Or, for each spurious response line on Fig. 3-51, a corresponding curve can be plotted as shown in Fig. 3-52, showing the relative response at each tuned frequency.

3-2.2.2.2 Intermodulation and Cross-modulation

Intermodulation in receivers results when two or more unwanted signals are present simultaneously at the device input, generating a new composite undesired signal in the desired band. Cross-modulation is the transfer of information from an undesired carrier onto the desired one. In either case, nonlinearity in a circuit near the receiver input is usually the cause.

Intermodulation is the more important of these mechanisms. It becomes especially important when a range of frequencies is subdivided into separate communication channels, and when a number of closely spaced channels must be used simultaneously. Then, two unwanted signals of the form

$$x_j(t) = v_j(t)\cos[\omega_j t + \varphi_j(t)] \text{ , } j = 1, 2 \tag{3-94}$$

give rise to a component of the form (see Eq. 3-74)

$$y_I(t) = v_1^2(t)v_2(t)\cos[(2\omega_1 - \omega_2) + 2\varphi_1(t) - \varphi_2(t)] \tag{3-95}$$

where

$y_I(t)$ = interference signal generated

When they are applied together to a device with a third degree nonlinearity, the component $y_I(t)$ is significant because $2\omega_1 - \omega_2$ is not too different from

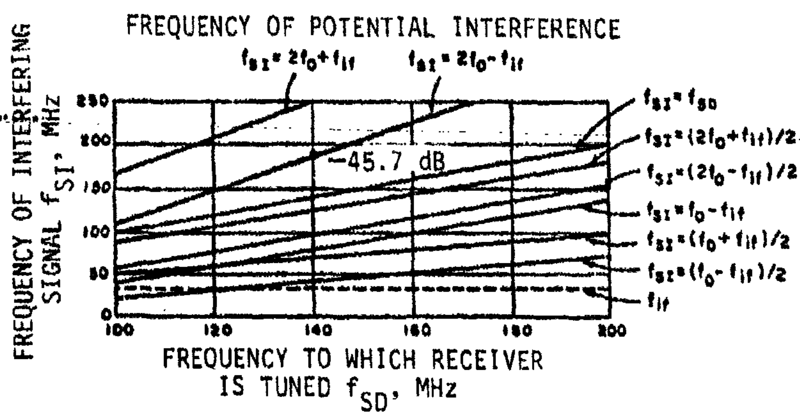


Figure 3-51. Possible Spurious Responses in a High-frequency Receiver With Local Oscillator Frequency f_o Set 30 MHz Below Desired Frequency

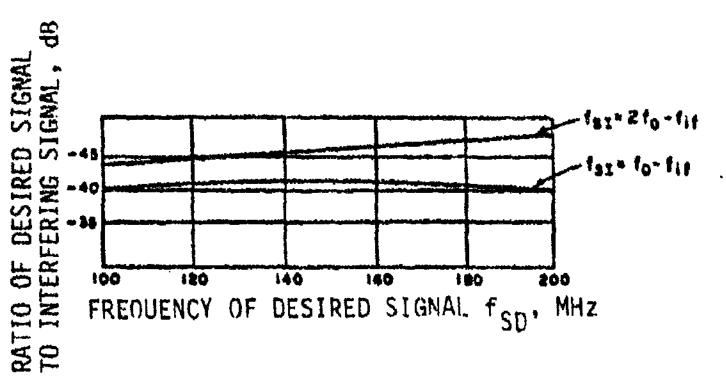


Figure 3-52. Relative Signal-to-Interference Response for Equal Outputs

either frequency ω_1 or ω_2 if the two are not far apart. For instance, if $f_1 = 100$ MHz, and $f_2 = 101$ MHz, the new unwanted frequency is 99 MHz.

Similarly with three channels at frequencies f_1, f_2, and f_3, intermodulation products having frequencies near to but not coincident with the original generating frequencies are

$$
\begin{aligned}
&f_1 + f_2 - f_3\\
&f_1 - f_2 + f_3\\
&2f_1 - f_2\\
&2f_1 - f_3\\
&2f_1 - f_1\\
&2f_2 - f_3\\
&2f_3 - f_1\\
&2f_3 - f_2
\end{aligned}
$$

The components at these frequencies are obtained by expanding

$$\left[\sum_{j=1}^{3} V_j(t)\cos(\omega_j t + \varphi_j)\right]^3$$

The even-degree terms in Taylor's series expansion for the output-input characteristic of the nonlinear element also give rise to intermodulation components, but they are all far from the range of frequencies in question. Though the third-degree term is generally the most important, the fifth-degree term may have to be accounted for also. Possible interference components due to a fifth-degree nonlinearity are of the form

$$
\begin{aligned}
&f_1 + f_2 + f_3 - f_4 - f_5\\
&2f_1 + f_2 - f_3 - f_4\\
&f_1 + f_2 + f_3 - 2f_4\\
&2f_1 + f_2 - 2f_3\\
&3f_1 - f_2 - f_3\\
&3f_1 - 2f_2
\end{aligned}
$$

The components at these frequencies are obtained by expanding

$$\left[\sum_{j=1}^{5} V_j(t)\cos(\omega_j t + \varphi_j)\right]^5$$

These are only representative forms; the subscript on these frequencies may be permuted in any way among the assigned frequencies. Thus, if there are 10 frequencies, $3f_7 - 2f_{10}$ or $f_1 - f_2 + f_3 - f_4 + f_5$ are frequencies that can be significant. Techniques for channel selection to avoid interference are given by Babcock (Ref. 67) and also by Beauchamp (Ref. 68).

Tests of susceptibility to intermodulation in actual receivers have been described in detail. McLenon (Ref. 69) applied signals to commercial grade receivers to give potential intermodulation at 5.1 MHz. He obtained a resultant equivalent interference carrier level of 0.5 μV for inputs ranging from 0.01 V to 0.1 V. The highest input was required in a receiver that had two tuned circuits before the first amplifier tube.

A sample calculation of the magnitude of intermodulation interference is now given. Third-degree nonlinearity is assumed. Carrying out an expansion similar to that given in Eq. 3-74, but with $x_s(t)$ and $x_o(t)$ replaced by two incoming signals $x_1(t)$ and $x_2(t)$ as given by Eq. 3-94 and with $n = 3$, one of the output interference components is

$$y_i(t) = \frac{3a_3}{4} v_1^2(t) v_2(t)\cos[(2\omega_1 - \omega_2)t] \qquad (3\text{-}96)$$

The tuned frequency of the receiver is $2\omega_1 - \omega_2 = \omega_s$. A desired signal, $x_s(t) = v_s(t)\cos\omega_s t$, entering the receiver at the same time produces an output term determined by the first-degree term (with coefficient a_1) of the Taylor series. This desired signal term is then $y_s(t) = a_1 v_s(t)\cos\omega_s t$.

The signal-to-interference voltage ratio S/I is defined as the ratio of the coefficients of the desired signal term and the undesired signal term as given in Eq. 3-96; i.e.,

$$S/I = \frac{4a_1 v_s(t)}{3a_3 v_1^2(t) v_2(t)}\text{, d'less} \qquad (3\text{-}97)$$

If, for simplicity, $v_1(t)$, $v_2(t)$, and $v_s(t)$ are taken to be the constants v_1, v_2, and v_s, respectively, and the two unwanted signal amplitudes are assumed equal so that $v_1 = v_2$, then the signal-to-interference ratio is unity when

$$v_1 = \left(\frac{4a_1 v_s}{3a_3}\right)^{1/3}\text{, V} \qquad (3\text{-}98)$$

For example, if

$$
\begin{aligned}
v_s &= 10 \times 10^{-6}\ \text{V}\\
a_1 &= 5 \times 10^{-3}\ \text{mho}\\
a_3 &= 5 \times 10^{-5}\ \text{A/V}^3
\end{aligned}
$$

then v_1 = 0.11 V. At VHF, an interfering signal of this magnitude could be produced by a 50-W transmitter with a spacing of about 150 ft between the transmitting source and the receiving antenna.

The intermodulation interference has amplitude v_1 at the input to the nonlinear element, but its amplitude at the antenna terminals will be greater than this value. However, if the selectivity of the input circuit is not sufficient to cause much attenuation to the unwanted signals, the unwanted input signal voltage can be about the same value.

In the case of cross-modulation arising from a third-degree nonlinearity, the interference component y_1 — again using Eq. 3-94 and expanding in a form such as Eq. 3-74 with $v_2(t)\cos[\omega_2 t + \varphi_2(t)]$ viewed as the desired signal — is

$$y_1 = \frac{3a_3 v_1^2(t) v_2(t)}{2}\cos[\omega_2 t + \varphi_2(t)] \text{ , A} \qquad (3\text{-}99)$$

This component contains a mixture of sidebands from the unwanted v_1 and the wanted v_2 signals. Since the desired component y_s in this case, is

$$y_s(t) = a_1 v_2(t)\cos[\omega_2 t + \varphi_2(t)] \text{ , A}$$

the signal-to-interference voltage ratio, defined as the ratio of the coefficients of $y_s(t)$ and $y_1(t)$ with $v_1(t)$ and $v_2(t)$ constant at v_1 and v_2, respectively, is

$$S/I = \frac{2a_1}{3a_3 v_1^2} \text{ , d'less} \qquad (3\text{-}100)$$

Only unwanted signals of large amplitude will make this ratio significant. When the signal-to-interference ratio is unity,

$$v_1 = \left(\frac{2a_1}{3a_3}\right)^{1/2} \text{ , V} \qquad (3\text{-}101)$$

for cross-modulation. For example, if the values of a_1 and a_3 given previously are used,

$$v_1 = 8.1 \text{ V}$$

3-2.2.2.3 Desensitization

Desensitization refers to a reduction in overall receiver gain, or sensitivity, or both, without other discernible effects when a large unwanted signal enters the receiver. Possibly the most important desensitization mechanism results from limiting in the input-output transfer characteristic such as shown in Fig. 3-38(A).

As indicated in Fig. 3-53, the desired signal denoted $x(t)$ (units of voltage are assumed for input) is assumed small compared with the undesired pure sinusoid denoted $A\cos\omega t$. The latter drives the device into its limiting state on both positive and negative peaks, and during limiting the desired signal is obliterated. The output current $i(t)$ is approximated by

$$\begin{aligned} i(t) &= \frac{b}{a}[x(t) + A\cos\omega t] \text{ , } |A\cos\omega t| < a \text{ , A} \\ &= \pm b \text{ , } |A\cos\omega t| \geq a \text{ , A} \end{aligned} \qquad (3\text{-}102)$$

where

a = magnitude of input voltage at which the output current saturates, V

b = magnitude of output current at saturation, A

This can be written

$$i(t) = \frac{b}{a}[x(t) + A\cos\omega t]S(t) + R(t) \text{ , A} \qquad (3\text{-}103)$$

where $S(t)$ is a unit amplitude sampling waveform, and $R(t)$ is a bipolar waveform of amplitude b, as shown in Fig. 3-53. These waveforms have periodicities determined by the period of the unwanted sinusoid, and they can be expanded in Fourier series

$$S(t) = a_o + \sum_{n=1}^{\infty} a_n \cos\left(\frac{n4\pi t}{T}\right) \text{ , } T = \frac{2\pi}{\omega} \text{ , d'less} \qquad (3\text{-}104)$$

$$R(t) = \sum_{n=1}^{\infty} b_n \cos\left(\frac{n2\pi t}{T}\right) \text{ , A} \qquad (3\text{-}105)$$

Thus

$$i(t) = \frac{b}{a}\left[x(t) + A\cos\omega t\right]\left[a_o + \sum_{n=1}^{\infty} a_n \cos 2n\,\omega t\right] + \sum_{n=1}^{\infty} b_n \cos n\,\omega t \text{ , A} \qquad (3\text{-}106)$$

Assume the spectrum of the desired signal $x(t)$ is as shown in Fig. 3-54, not overlapping the frequency of

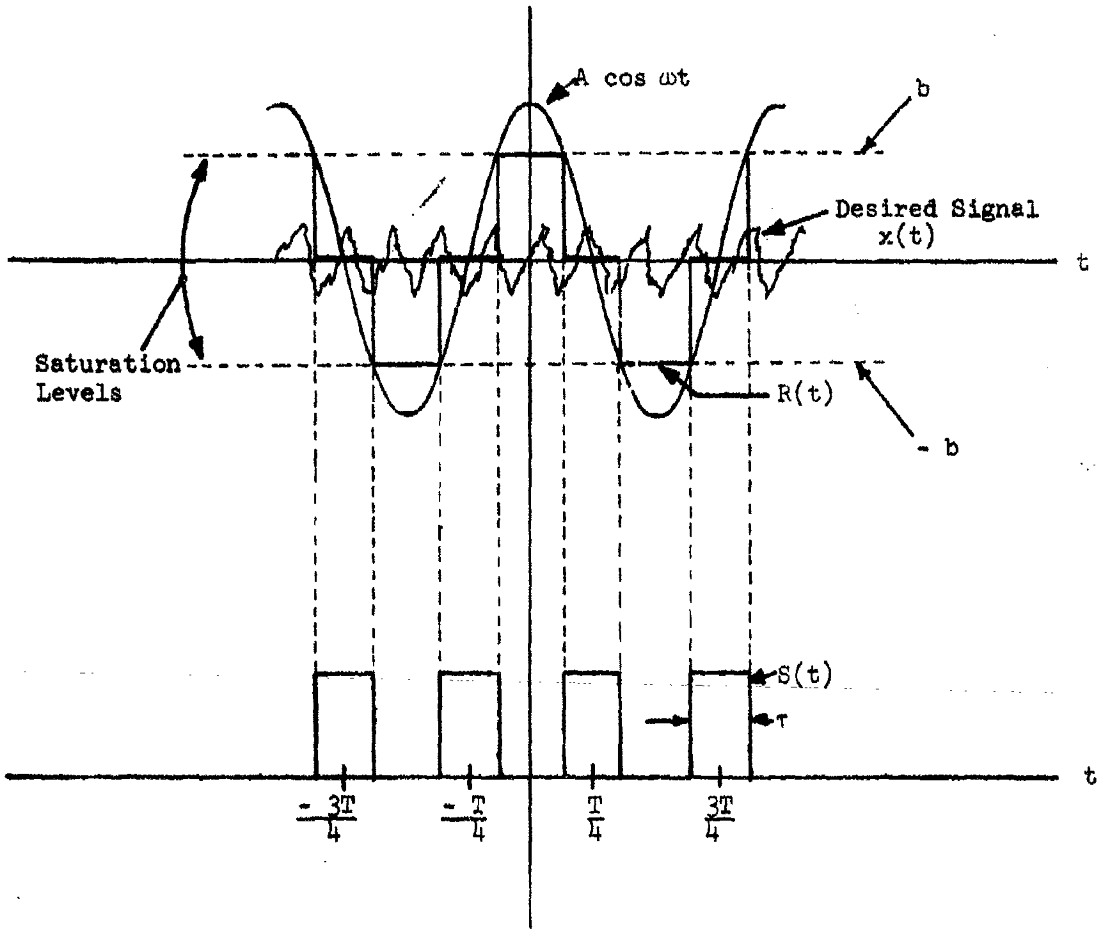


Figure 3-53. Waveform Diagram for Desensitization Analysis

the unwanted sinusoid but being close to it (if it were not close, the sinusoid would be strongly attenuated before reaching the nonlinear device). Examining Eq. 3-106 it will be found that of the many components only one will be located on the desired center frequency ω_o and that component is

$$I_d(t) = \frac{b}{a} a_o x(t) \ , A \qquad (3\text{-}107)$$

A zonal filter which sees only components around ω_0 therefore will see Eq. 3-107 only; it is affected only by the dc component of the sampling waveform. Before determining the value of this dc component, which will determine the amount of desensitization, it is pointed out that if the undesired signal is modulated it is still possible for the output to be of the form of Eq. 3-107, i.e., only attenuated and not affected by the modulation on the undesired signal, provided that the bandwidth of the undesired signal is not so wide as to produce spectral overlap from the limited unwanted signal to the wanted one.

Defining τ as the duration of the nonsaturation interval, it is determined from

$$A^2 \cos^2 \omega t_s = a^2 \ , A \geq a \qquad (3\text{-}108)$$

or

$$\cos 2\,\omega t_s = \left(\frac{2a^2}{A^2} - 1\right) \ , A \geq a \qquad (3\text{-}109)$$

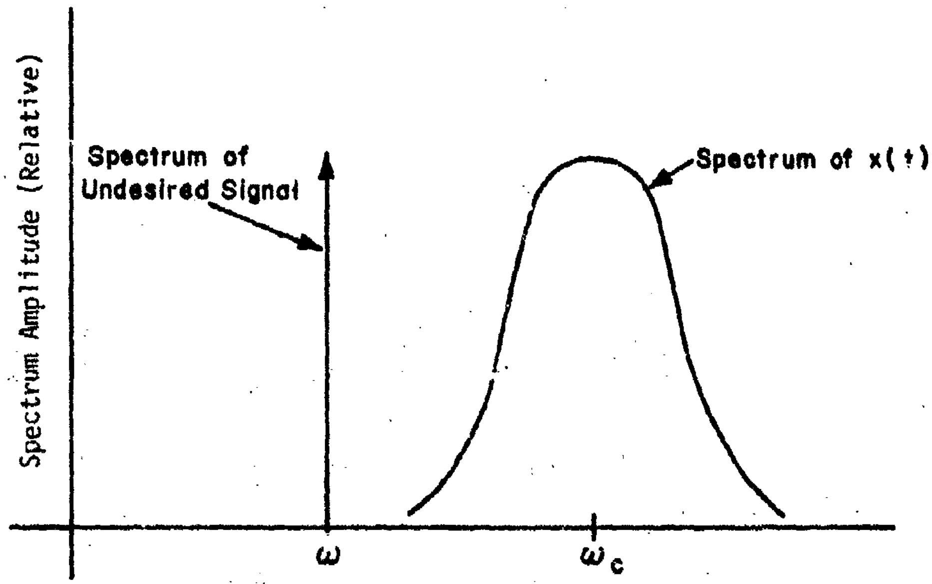


Figure 3-54. Spectra of Undesired and Desired Signals

where t_s is the instant at which the sinusoid amplitude is equal to the saturation level. The nonsaturation time τ is

$$\tau = \frac{T}{2} - 2t_s = \frac{T}{2} - \frac{T}{2\pi}\text{Cos}^{-1}\left(\frac{2a^2}{A^2} - 1\right), \quad A \geq a \text{ , s} \qquad (3\text{-}110)$$

The value of a_o is

$$\begin{aligned} a_o &= \frac{2\tau}{T} = 1 - \frac{1}{\pi}\text{Cos}^{-1}\left(\frac{2a^2}{A^2} - 1\right), \ A \geq a \text{, d'less} \\ &= 1 \qquad , A \leq a \text{, d'less} \end{aligned} \qquad (3\text{-}111)$$

so that

$$\begin{aligned} i_d(t) &= \frac{b}{a}\left[1 - \frac{1}{\pi}\text{Cos}^{-1}\left(\frac{2a^2}{A^2} - 1\right)\right]x(t) \ , A \geq a \text{ , A} \\ &= \frac{b}{a}x(t) \qquad , A < a \text{ , A} \end{aligned} \qquad (3\text{-}112)$$

The form of a_o, the fraction of output current remaining from saturation, which represents the desensitization effect, is shown in Fig. 3-55 as a function of a/A.

A mechanism related to the one just described is based on the change of input impedance of the electronic device over the period of the interference waveform. Typically, a normally reversed bias input junction of a solid-state amplifier may be forward biased part of the time by interference creating a virtual short circuit to the desired signal source.

Other important mechanisms are associated with the AGC system, its filters, and bias generating RC filters.

The AGC voltage is determined by the carrier level at the detector input, and any signals present there will affect it. Should interference of large amplitude penetrate the receiver as far as the detector, it will contribute to the AGC voltage whether or not it produces a discernible detected output.

In systems having RC networks for bias generation or in those having AGC filter networks in the early stages, overload will cause a change in bias and a reduction in gain. The bias is sustained for an interval of time depending on the RC time constant and the peak value of the undesired signal.

Pulsed signals of low duty cycle, such as radar emissions, can be especially troublesome in active circuits with bias or AGC filters because of their large

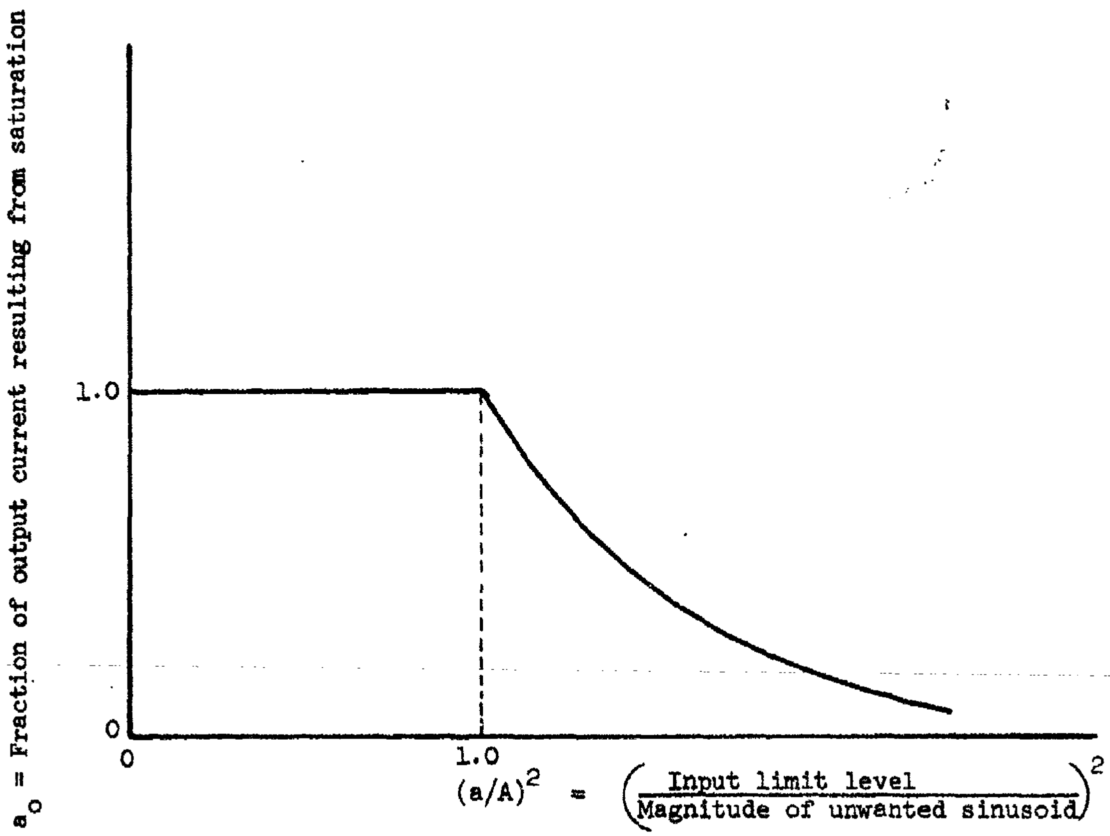


Figure 3-55. Desensitization Curve

peak amplitudes. Microwave radar interference to low-frequency communication systems, by overload of an early receiver stage, is not uncommon — largely because the lumped tuned circuits at the input are virtually useless as filters of microwave energy. Once the unwanted pulse appears at the input amplifying device, it will overdrive the input during peaks and charge filter capacitors; thus creating a long duration bias.

The capture phenomena encountered in envelope detection systems, in FM systems and in receivers using frequency tracking are, in effect, desensitization mechanisms when the undesired signal is unmodulated. When it is modulated, the modulation on the undesired signal appears at the output.

In par. 3-2.2.2.1, it was shown that diode mixers naturally act as harmonic mixers to create spurious responses. Diode mixers are also subject to desensitization effects (Ref. 70). The effect is found to arise in microwave receivers (e.g., radar receivers) where the mixer is the first electronic device following the input terminals. It can be shown that the conversion transconductance (defined below Eq. 3-79) is altered by the presence of a large unwanted signal. A more important effect, however, appears to be associated with impedance mismatch; the effective output impedance of the mixer at the IF frequency is altered by the unwanted signal. If the input impedance of the IF amplifier is matched to the impedance of the mixer in the absence of unwanted signals, it will become unmatched when the unwanted signal appears. Tests reported (Ref. 69, p. 28) show a drop in conversion efficiency by approximately 3 dB for an unwanted sinusoid whose amplitude is equal to that of the local oscillator signal (Fig. 3-56); the greater the local oscillator power fed to the mixer, the larger will be the magnitude of the unwanted signal that can be tolerated. It was shown previously, however, that with increasing local oscillator inputs, the harmonic conversion transconductance g_n becomes significant

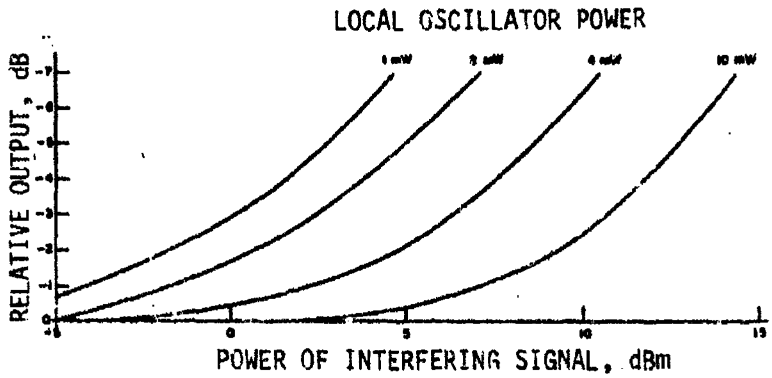


Figure 3-56. Tests Results of Conversion Loss in a Crystal Mixer

for higher values of n. Therefore, a compromise is needed between high local oscillator power (to minimize desensitization potential) and low local oscillator power (to minimize spurious response potential).

It is evident that with adequate filtering prior to the active elements in the receiver, the effects of nonlinearity in these elements can be reduced. Ideally, the bandwidth of circuits ahead of a potentially nonlinear element should be equal to the bandwidth of the IF amplifier, but this generally is impractical and difficult to accomplish. Unwanted signals whose frequency is relatively near that of the desired band therefore will not always be easy to reject in the RF amplifier. Where such interference is expected, it is desirable to use input circuits with large dynamic ranges to avoid such effects as overload and desensitization. However, sharp rejection filters (wave traps) have been devised (Refs. 71 and 72) especially for rejecting fixed-frequency unwanted signals in an adjacent channel.

3-3 COUPLING PHENOMENA

The coupling of the interference signal from the source to the susceptor occurs in two basic ways: (1) by way of a mutual impedance, and (2) by radiation. It also can be coupled by a combination of both methods. For example, interference can be radiated from one equipment, picked up on interconnecting cables of another equipment, and thereafter conducted into the equipment enclosure. Conversely, interference can be conducted to the outside of the cabinet or enclosure of the source by cables, and then radiated.

Conductive coupling occurs when a susceptible circuit shares a common circuit path with the interference source as illustrated schematically in Fig. 3-57 (Ref. 11). A current flowing in circuit 1 directly produces a voltage in circuit 2. The magnitude of the mutual impedance z is the ratio of the open circuit voltage of circuit 2 — with all other sources of voltage in circuit 2 removed — to the current in circuit 1.

The common impedance may be any circuit element, including structural elements. Typical examples are:

a. Common ground return impedances, including chassis grounds and cabinet bonds and ground straps

b. Common power supply impedances, including distribution cables and decoupling networks.

Inductive coupling occurs when two circuits are located physically close to each other so that there is a common flux linkage between them, even though there may be no direct interconnection, as shown in Fig. 3-58 (Ref. 11). The electromotive force e_2 induced in circuit 2 by circuit 1 is given by

$$e_2 = M\left(\frac{di_1}{dt}\right), \text{V}$$

where

M = coefficient of mutual inductance, H

i_1 = current in circuit 1, A

Likewise, circuit 2 induces an emf in circuit 1.

In a similar way, mutual electric coupling can occur between two circuits because of distributed capacitance, as shown in Fig. 3-59.

The coupling here is quite similar to that shown in Fig. 3-57(B), and calculations can be made using that circuit at low frequencies. At high frequencies, i.e., when the length of circuit over which inductive

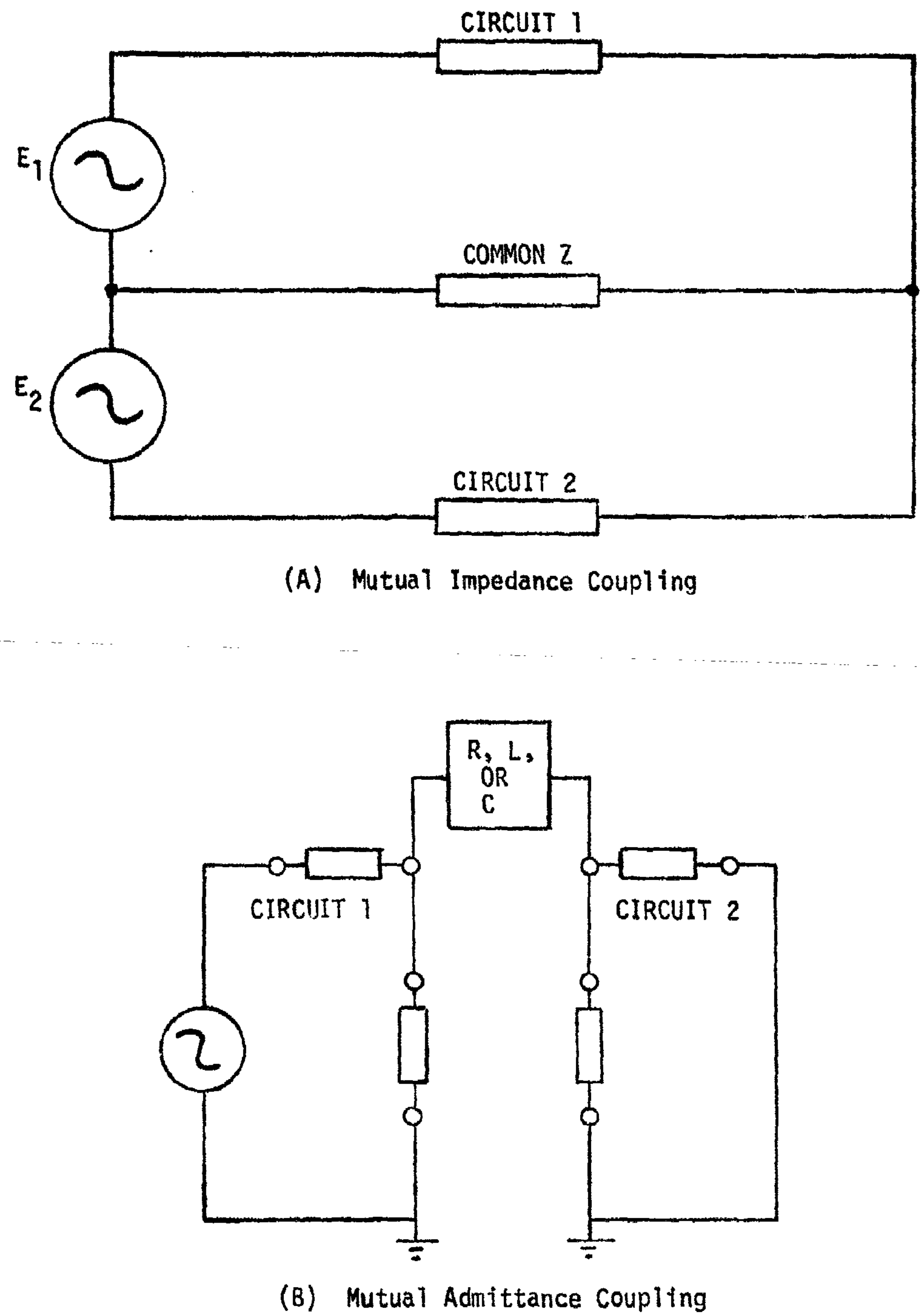


Figure 3-57. Mutual Conductive Coupling

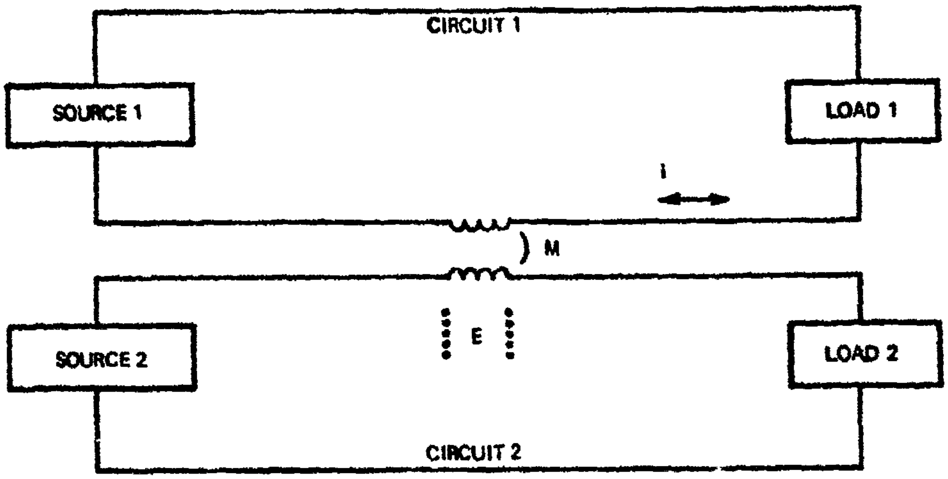


Figure 3-58. Mutual Inductance Coupling

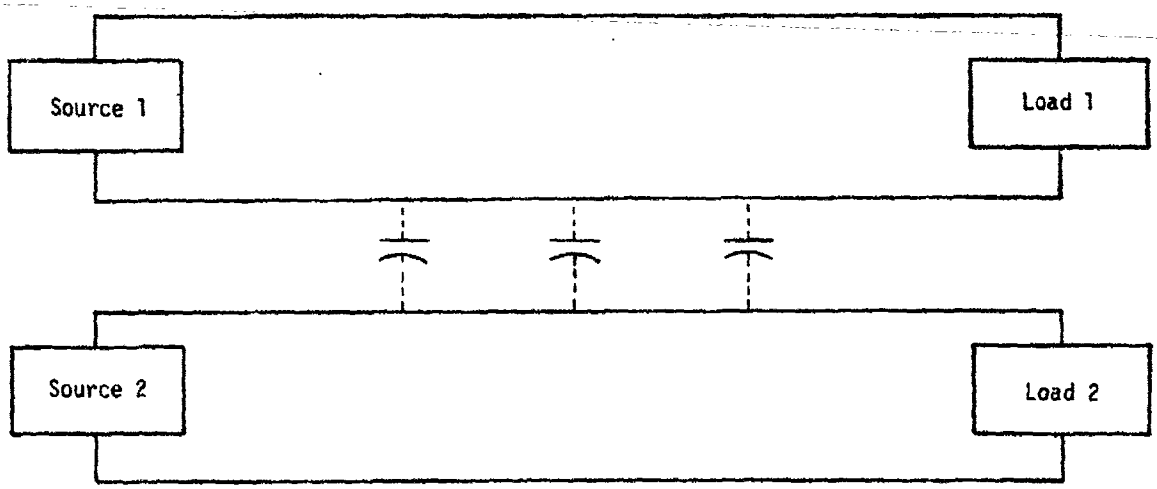


Figure 3-59. Mutual Capacitance Coupling

coupling occurs approaches about a sixth of a wavelength, the distributed nature of the coupling path must be taken into account.

Radiative coupling (as distinguished from induction field coupling) usually occurs when the circuits involved are widely separated physically when measured in terms of the wavelength of the emission, and in terms of the physical dimensions of the objects.

3-3.1 INDUCTION FIELD COUPLING

At distances from the source that are small compared with approximately one-sixth of a wavelength, the field is dominated by the static or induction field, also called the near field. The characteristics of this component of the field are such that:

a. At distances greater than the source dimensions, but small compared with about one-sixth of a wavelength, it diminishes rapidly with distance from the source — e.g., in the case of sources which may be represented as dipoles, as the reciprocal of distance cubed.

b. It may be predicted employing "static" models for the source, e.g., Ampere's law for magnetic fields and Laplace's equation for electric fields.

For internal circuit-to-circuit coupling, such as occurs within an equipment cabinet, stray mutual inductances or capacitances can be as important as conductive coupling. Outside of a cabinet or cables, and below about 100 kHz, magnetic induction is usually much more important than electric induction because it is relatively easy to provide electric shielding. Above this frequency, and up to about 10 MHz, both electric or magnetic field effects may be important. Above 10 MHz, it is usually not significant to differentiate between electric and magnetic field effects, but it is customary in the literature to refer to the electric field component.

3-3.1.1 Magnetic Field Coupling

The mutual inductance between two circuits depends upon the geometry not only of the circuits involved, but any others in the vicinity. Except in the case of multiconductor cables where the several conductors show a mutually symmetrical configuration, it is customary in EMC work to consider multiple circuit arrangements on a pair by pair basis. This is done because of the necessity of simplifying analysis, and also because it is adequate in most cases.

3-3.1.1.1 Magnetic Fields from Devices or Cabinets

Magnetic fields can arise from equipment or cabinets enclosing equipment containing large inductors, which may or may not have magnetic cores. At spacings close to such devices, sensitive circuits can exhibit susceptibility. A convenient model for such sources is the magnetic dipole.

3-3.1.1.1.1 Dipole Properties

For an infinitesimally small magnetic dipole source, the field components at a distance r are given by (Ref. 73)

$$H_\theta = \frac{M_m\beta^3}{4\pi}\left(-\frac{1}{\beta r} + j\frac{1}{(\beta r)^2} + \frac{1}{(\beta r)^3}\right)\sin\theta\ ,\ \text{A/m}$$

$$H_r = \frac{2M_m\beta^3}{4\pi}\left(\frac{1}{(\beta r)^3} + j\frac{1}{(\beta r)^2}\right)\cos\theta\ ,\ \text{A/m}$$

$$E_\varphi = \frac{M_m\beta^3\eta}{4\pi}\left(\frac{1}{\beta r} + j\frac{1}{\beta^2 r^2}\right)\sin\theta\ ,\ \text{V/m} \tag{3-113}$$

where

H_r and H_θ = respectively, the components of the magnetic field in the r- and θ-directions (see Fig. 3-60)

E_φ = component of the electric field in the φ-direction

β = phase constant $= \dfrac{2\pi}{\lambda} = \dfrac{2\pi f}{c}$, 1/m

f = frequency, Hz

λ = wavelength, m

c = speed of light, 3×10^8 m/s

M_m = magnetic dipole strength, A·m²

$j = \sqrt{-1}$

η = wave impedance = $2\pi\ 60\Omega$ for free space

For $\beta r \ll 1$, the magnetic field strength is dominated by the terms varying as $(1/r)^3$, and in the direction of maximum θ ($\theta = \theta/2$) its magnitude is given by

$$H_r \approx \frac{M_m}{2\pi r^3}\ ,\ \text{A/m} \tag{3-114}$$

For example, if $r < 20$ m, and $f < 1$ MHz, it is satisfactory to model the magnetic field strength as falling off as $(1/r)^3$.

The magnetic flux density B is given in terms of the magnetic field strength by

$$B = \mu H \approx \frac{\mu M_m}{2\pi r^3}\ ,\ \text{T*} \tag{3-115}$$

μ = permeability, H/m

If the flux density B' is measured at a distance r', then the flux at any other distance within the induction field is found by

$$\frac{B}{B'} = \left(\frac{r'}{r}\right)^3 \tag{3-116}$$

i.e., the $(1/r)^3$ relationship of Eq. 3-115 is used.

3-3.1.1.1.2 Flux Density from a Loop

If the maximum dimension of a source is comparable in size with the distance at which the field is to be calculated, a better model is a current loop. The equations for the field from a loop are relatively complex off the axis of the loop; however, at any distance the field off axis lies between its values on axis and in the loop plane.

* T = tesla = weber/meter²

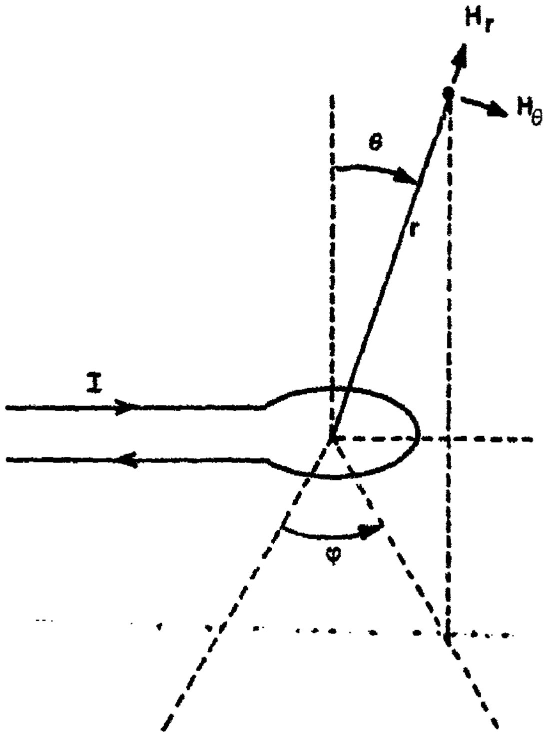


Figure 3-60. Coordinate System for Magnetic Source (loop lies in plane normal to θ axis)

Far from the loop, the loop acts like a magnetic dipole of moment $N\pi d^2 I_L/4$, so that the field strength is, by Eq. 3-114

$$H = \frac{NI_L d^2}{8r^3}, r \gg d \text{ , A/m} \qquad (3\text{-}117)$$

where

N = number of turns
d = loop diameter, m
I_L = loop current, A

Fig. 3-61 (Ref. 74) shows the variations of the magnetic field strength as a function of distance measured from the center of the loop, normalized with respect to the dipole moment, in a direction in the plane of the loop, and in a direction along the axis of the loop. At large distances, the field varies inversely as the cube of the distance in both directions, with the field along the axis being 6 dB higher than the field in the loop plane. At close distances, the field is lowest along the axis of the loop, and approaches an asymptotic value that depends upon the loop radius, being (for loops of equal dipole strength) lower for loops having the larger radii. For a direction in the plane of the loop, the value of the field increases rapidly without limit as the position approaches the location of the actual loop conductor. Fig. 3-62 (Ref. 74) shows the same information as shown in Fig. 3-61, except normalized with respect to the loop radius a.

As a practical matter, it should be recognized that: (1) in the usual device it is generally not possible to get closer to the effective loop than the loop radius itself, and (2) any susceptible device has an effective pick-up area of finite size and therefore responds to the average field strength over that area. Because of these factors it is considered reasonable to expect that the effective value of the field at any distance r will lie between the two straight lines shown on Fig. 3-62, which are drawn coincident with the actual field lines at large distances. In any event, one would not expect it to exceed the upper curve. Thus, a prediction of coupling based upon the upper curve should provide a conservative estimate of any possible interaction effects due to a source of this type.

For some purposes it may be more useful to have the data on Fig. 3-62 in terms of the distance measured to the edge of the loop. Fig. 3-63 (Ref. 75) show plots of the asymptotic relations as one moves away from the loop. The line with slope = $1/r$ gives the flux density from a long straight wire (see par. 3-3.1.1.2). Note that this figure gives values of the magnetic flux density B as compared with field strength on Figs. 3-61 and 3-62 (Refs. 74 and 76). In free space one can convert from one of these forms to the other by means of the relationship

$$B[\text{dB}(\mu\text{T})] = H[\text{dB}(\text{A/m})] + 2 \qquad (3\text{-}118)$$

where

[dB( )] indicates that the preceding quantity is to be expressed in decibels with respect to the unit in parentheses.

3-3.1.1.1.3 Experimental Data

Fig. 3-64 (Ref. 74) shows field strength vs distance for a configuration of a steel box surrounding a 500-W isolation transformer.

In addition to showing the approximate levels to be expected, these results show that, for the source used, an inverse cube-distance relationship (allowing for a 6-dB spread) applies over practical distances for measurements beyond the loop radius.

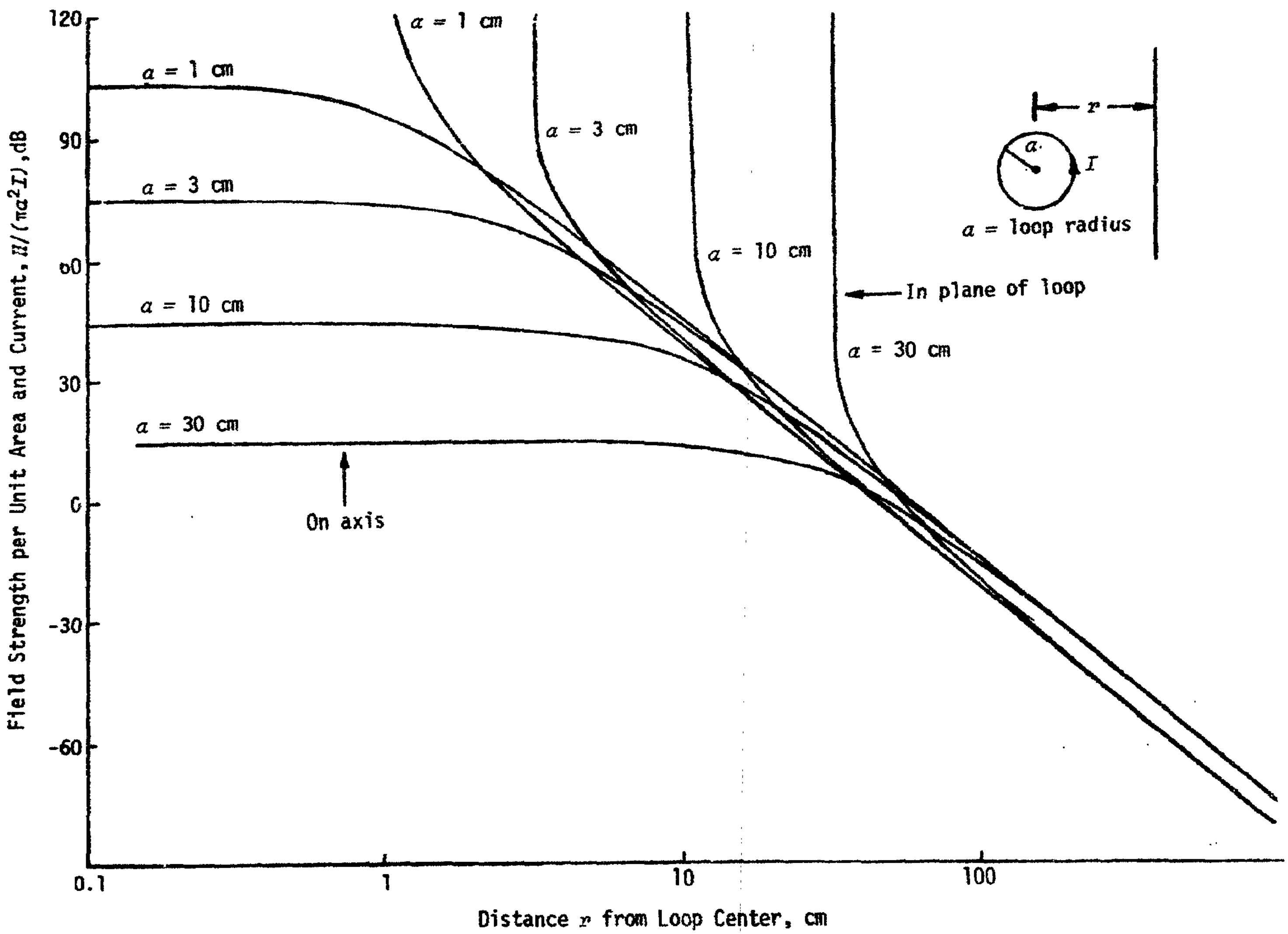


Figure 3-61. Magnetic Field Strength from Circular Current Loop on Axis and in Plane or Loop (Ref. 74)

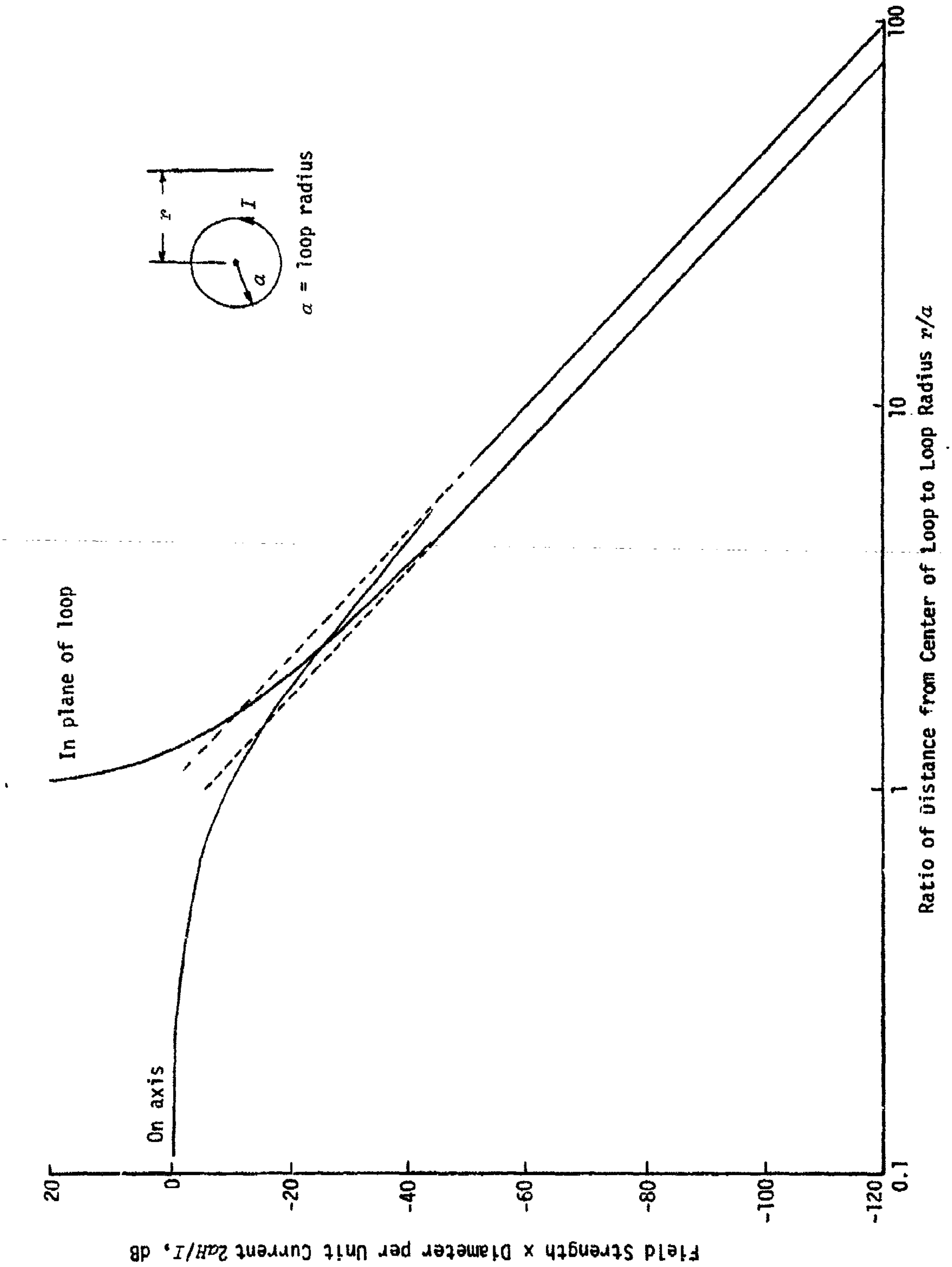


Figure 3-62. Normalized Magnetic Field Intensity from Circular Current Loop on Axis and in Plane of Loop (Ref. 74)

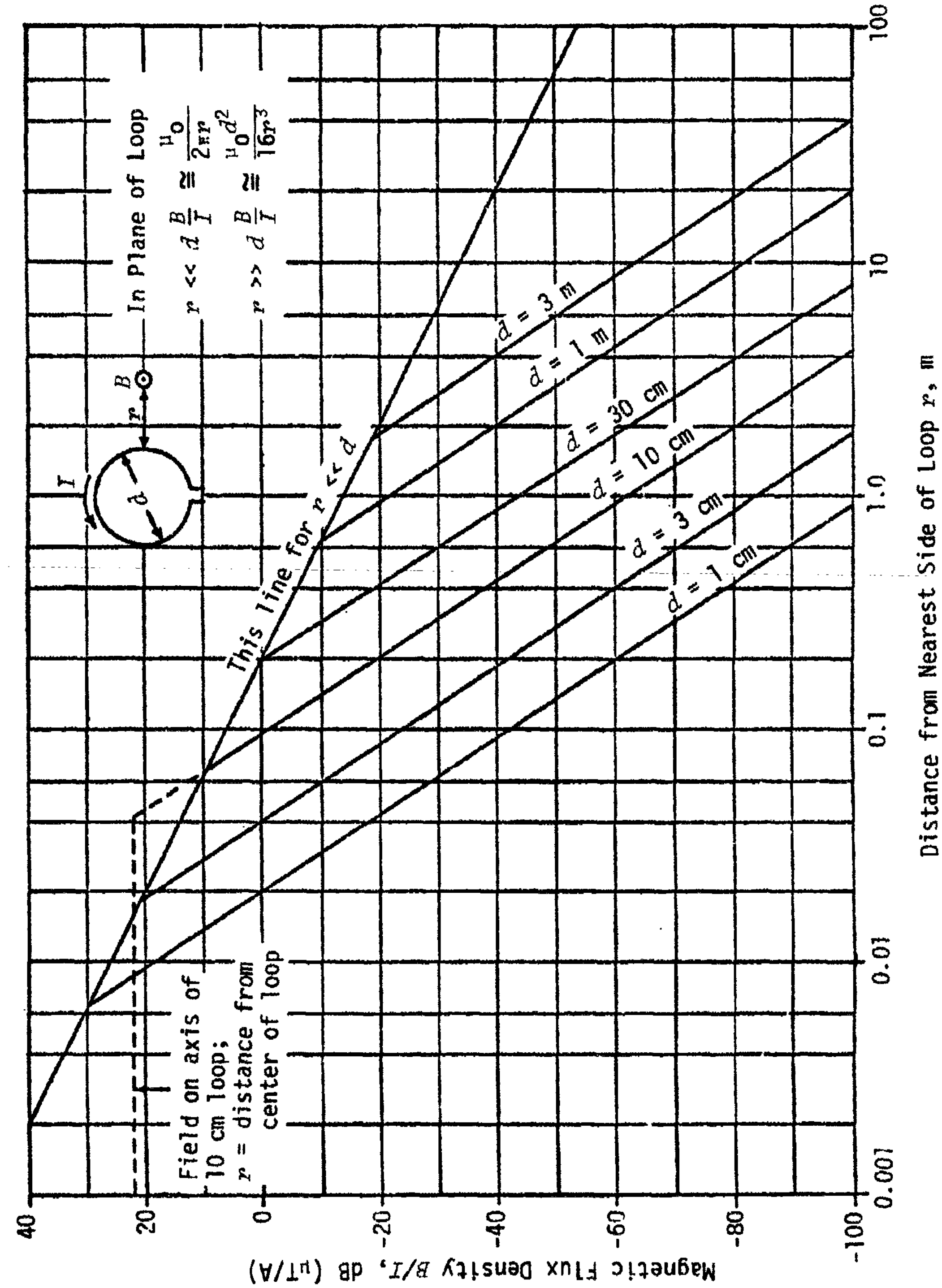


Figure 3-63. Maximum Flux Density from a Circular Loop

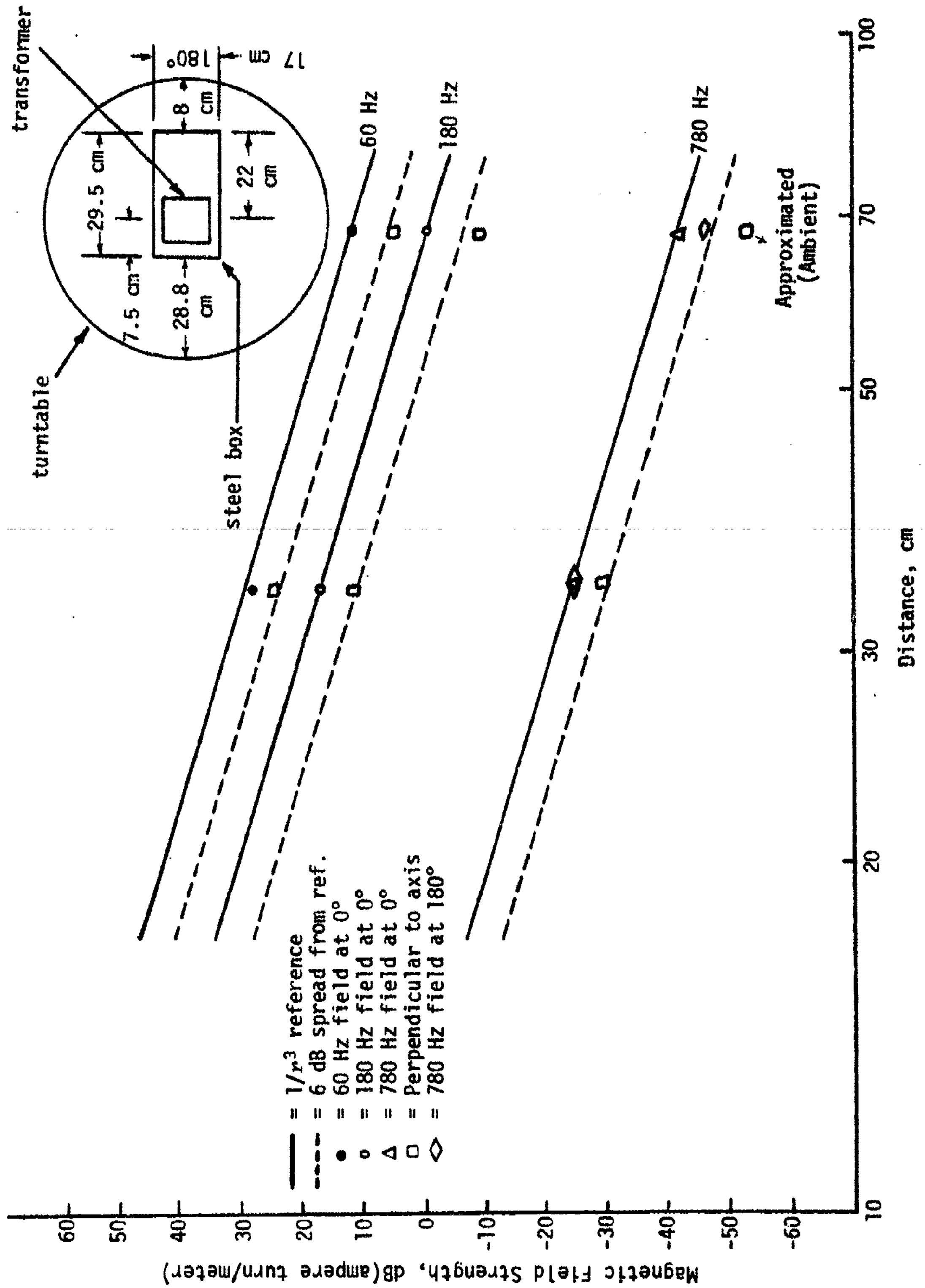


Figure 3-64. Magnetic Field Strength of 500-W Transformer

The effective strength of any such source is probably described most correctly in terms of its effective dipole strength M_m. Taking 20 × log of Eq. 3-114 yields

$$M_m[\text{dB(ampere-turns-meters}^2)] = 16 + H_{max}[\text{dB(ampere-turns/meter)}] + 60 \log r \quad (3\text{-}119)$$

where r is the distance from the center of the dipole in meters. The data on Fig. 3-65 were measured at distances of 0.5 m and 1.0 m for two transformers, and show the levels of the effective dipole strength at harmonic frequencies up to 1020 Hz.

The dipole model has validity except in a situation where there are two or more similar dipoles located together but oppositely oriented, i.e., effectively a "quadripole". Then field cancellation effects take place and the dipole model becomes invalid. On the other hand, due to cancellation, the resultant field will be reduced considerably over what one would expect if the generating fields are aiding. A prediction based upon a single dipole is a conservative prediction.

Devices with a large number of turns, such as cathode ray tube deflection yokes or a large area such as a power distribution switchboard, will have large effective dipole strengths.

3-3.1.1.2 Magnetic Fields from Wires and Cables

The magnetic flux density B from a long, straight, isolated wire surrounded by free space (or other nonmagnetic mat¬rial) is given by the relation

$$B = \frac{\mu_0 I}{2\pi r}, \text{T} \quad (3\text{-}120)$$

where

I = current in the wire, A

r = distance from the wire to the point at which B is calculated, m

μ_0 = permeability of free space, $4\pi \times 10^{-7}$ H/m

The direction of the magnetic flux density is perpendicular to the wire and uniform circumferentially around the wire (see Table 3-8).

The isolated, single wire cannot exist in reality, because there must be a return path for the current. However, at points very close to the wire compared with the distance to the return path, Eq. 3-120 is valid.

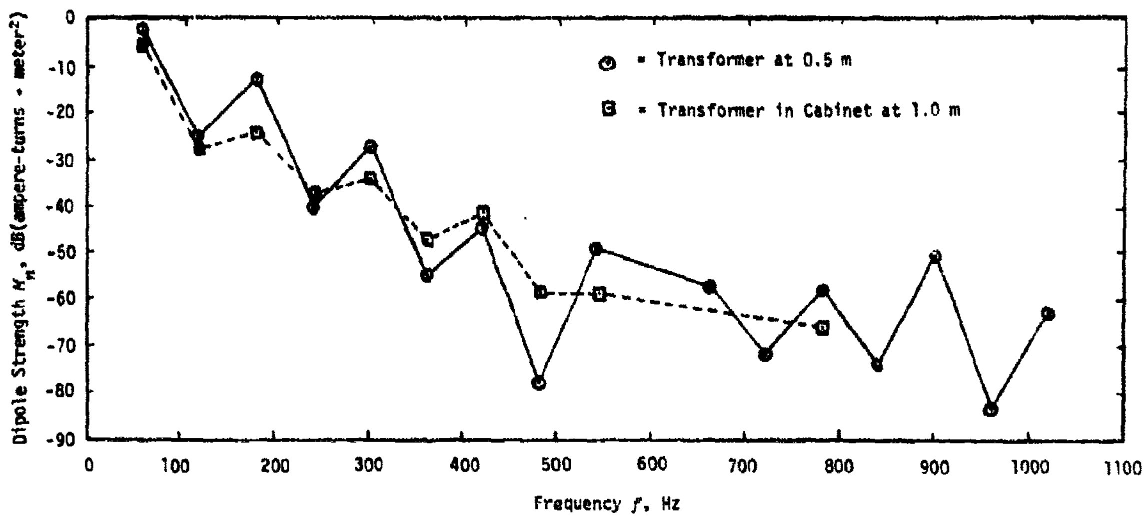
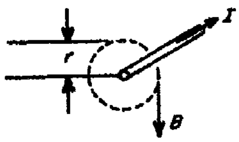


Figure 3-65. Dipole Strength vs Frequency

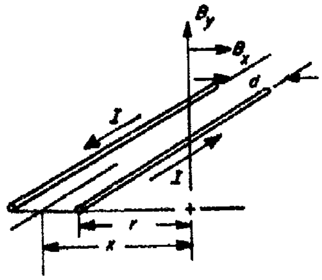
TABLE 3-8. LOW-FREQUENCY MAGNETIC FIELDS FROM WIRES AND CABLES

SINGLE WIRE



$$B = \frac{\mu_0 I}{2\pi r}$$

PARALLEL PAIR

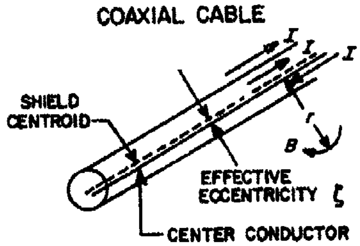


$$B_x = \frac{\mu_0 I y}{2\pi}\left[\frac{1}{\left(x+\frac{d}{2}\right)^2+y^2} - \frac{1}{\left(x-\frac{d}{2}\right)^2+y^2}\right]$$

$$B_y = \frac{\mu_0 I}{2\pi}\left[\frac{x-\frac{d}{2}}{\left(x-\frac{d}{2}\right)^2+y^2} - \frac{x+\frac{d}{2}}{\left(x+\frac{d}{2}\right)^2+y^2}\right]$$

$$B_{max} = |B_y(y=0)| = \frac{\mu_0 I d}{2\pi r(r+d)}$$

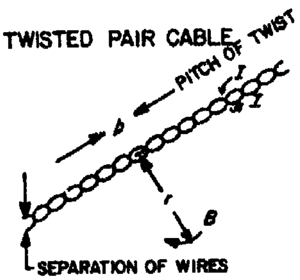
COAXIAL CABLE



$$B_{max} = \frac{\mu_0 I \delta}{2\pi r(r+\zeta)}$$

δ a function of frequency as given in Fig. 3-67.

TWISTED PAIR CABLE



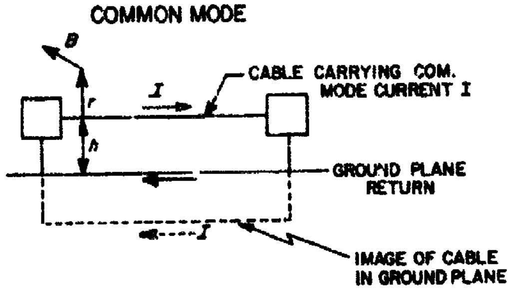
$$B_{max} = \frac{\mu_0 I}{pr}\, q I_0(q)\exp(-2\pi r/p)$$

$$q = \frac{\pi d}{p}$$

$I_0(x)$ = 0th order modified Bessel function of first kind.

Correction to B_{max} for parallel wire line of same spacing to obtain twisted pair B_{max} in Fig. 3-68.

COMMON MODE



$$B_{max} = \frac{\mu_0 I 2h}{2\pi r(r+2h)}$$

If there are n current-carrying wires, all long and straight, then the total magnetic field from the ensemble is the vector sum of the fields from each wire alone

$$B_{tot} = \sum_{i=1}^{n} B_i \qquad (3\text{-}121)$$

where

$$B_i = \frac{\mu_0 I_i}{2\pi r_i}, i = 1, \ldots, n$$

I_i = current in wire i, A

r_i = distance from wire i to point at which B_{tot} is calculated, m

The direction of B_i is tangent to a circle of radius r_i concentric with wire i, and is related to the direction of the current by a right-hand screw relation.

3-3.1.1.2.1 Parallel-wire Line (Low Frequencies)

The formulas for the flux density are shown in Table 3-8. The maximum value at a given distance r from the parallel-wire occurs in the plane of the two conductors. Near the line, the nearest conductor dominates, and the flux density can be estimated as that from a single-wire line, falling off as the reciprocal of distance from that wire. For $r \gg d$ the flux density falls off as the reciprocal square of distance.

The asymptotic relations are plotted in Fig. 3-66 (Ref. 75).

At frequencies for which distances of interest are comparable with the wavelength, the current distribution along the line is not uniform, transmission line effects become important, and the field computation becomes more complex. Also, a significant electric field component may exist.

3-3.1.1.2.2 Coaxial Cables

At low frequencies, a coaxial cable carrying a current will have an external field much like that of a two wire line if the outer conductor is not exactly concentric with the inner conductor, but is displaced by a small distance ζ due to manufacturing tolerances, or other causes. The fields external to the cable are identical to those from a parallel-wire line spacing ζ. The maximum magnetic flux density from such a cable will be

$$B = \frac{\mu_0 I \zeta}{2\pi r(r + \zeta)}, \text{T} \qquad (3\text{-}122)$$

where r is taken from the center of the cable (see Table 3-8).

Values of ζ for several common coaxial cable types are shown as functions of frequency in Fig. 3-67 (Ref. 76). As frequency increases beyond 10 kHz, the effective ζ for larger cables decreases due to skin effect and redistribution of current on the inner wall of the outer conductor, such that the current distribution centroid moves closer to the center conductor.

At higher frequencies, radiation from the cable is caused by leakage due to imperfections in the shield and/or imbalance in the terminating circuits leading to generation of common mode currents. These topics are covered in later paragraphs.

3-3.1.1.2.3 Twisted-pair Cables

If a current-carrying conductor and its return path are two identical wires twisted about each other to form a double helix, the magnetic field emitted is reduced considerably from what it would be for a parallel-wire line of the same spacing. The reduction factor in dB is shown in Fig. 3-68 (Ref. 76).

The formula for the maximum magnetic flux density from a twisted pair line is approximately

$$B = \frac{\mu_0 I}{\sqrt{pr}} q I_0 q \exp(-2\pi r/p), \frac{1}{20} < q < \frac{2}{3}, \text{T} \qquad (3\text{-}123)$$

or

$$B = \frac{\mu_0 I d}{2\pi r(r + d)} \text{ close to cable, } r < \frac{p}{3}, \text{T} \qquad (3\text{-}124)$$

where

d = helix diameter, m

p = helix pitch, m

r = distance from axis of helix, m

q = $\pi d/p$, helix parameter, dimensionless

$I_0(x)$ = 0th order modified Bessel function of the first kind

3-3.1.1.2.4 Common-mode Generation of Magnetic Fields

The formulas in pars. 3-3.1.1.2.1 through 3-3.1.1.2.3 apply when the current in one conductor in a cable equals the return current in the second conductor. If this is not the case, then effectively a net current flows in the two conductors (taken as a unit). This current is called the common-mode current. At low frequencies, common-mode current flows usually because of the existence of an alternate return path through the ground plane. In this case, the cable and its image in the ground plane may be considered to be

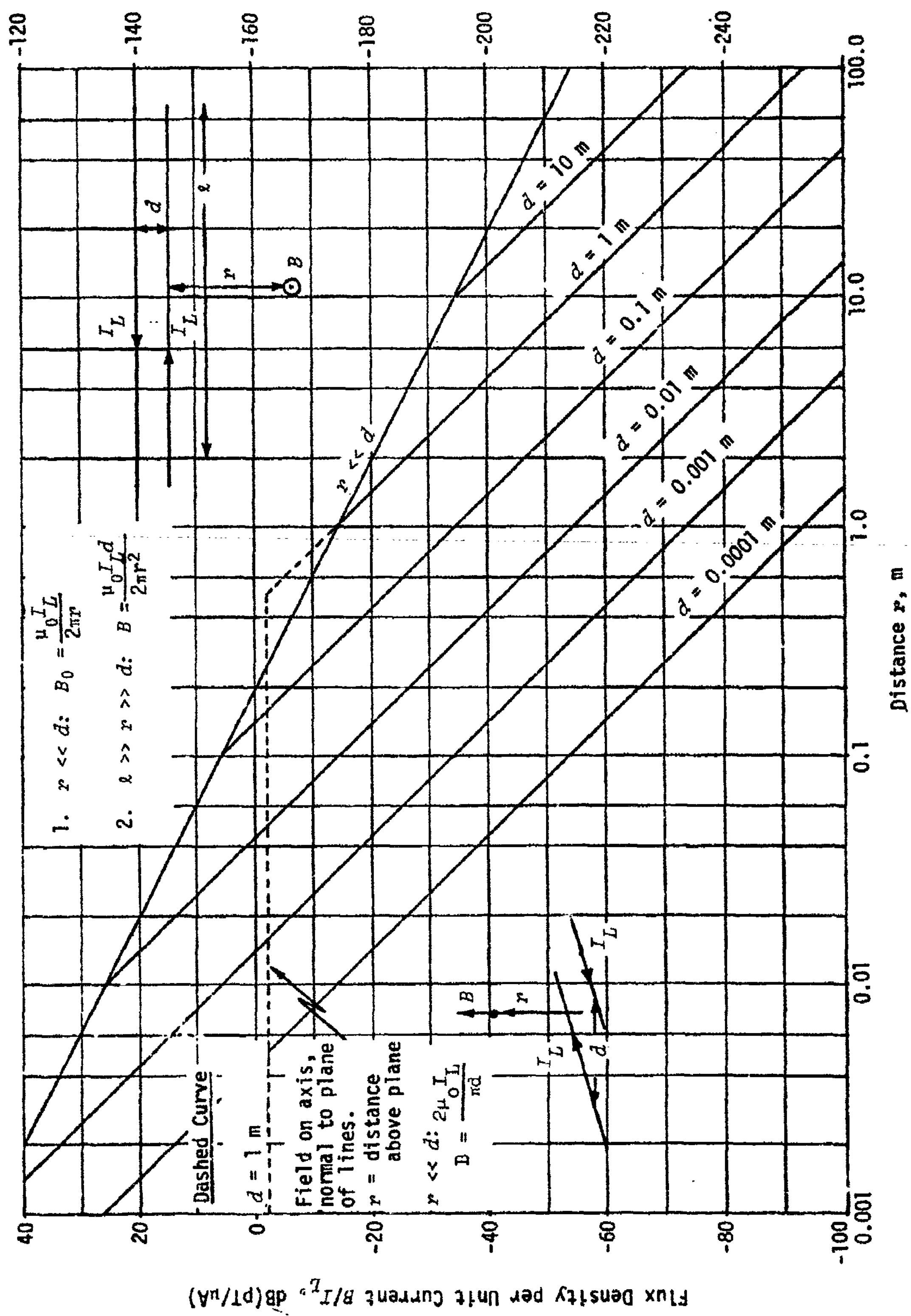


Figure 3-66. Flux Density from Parallel Conductors (Ref. 75)

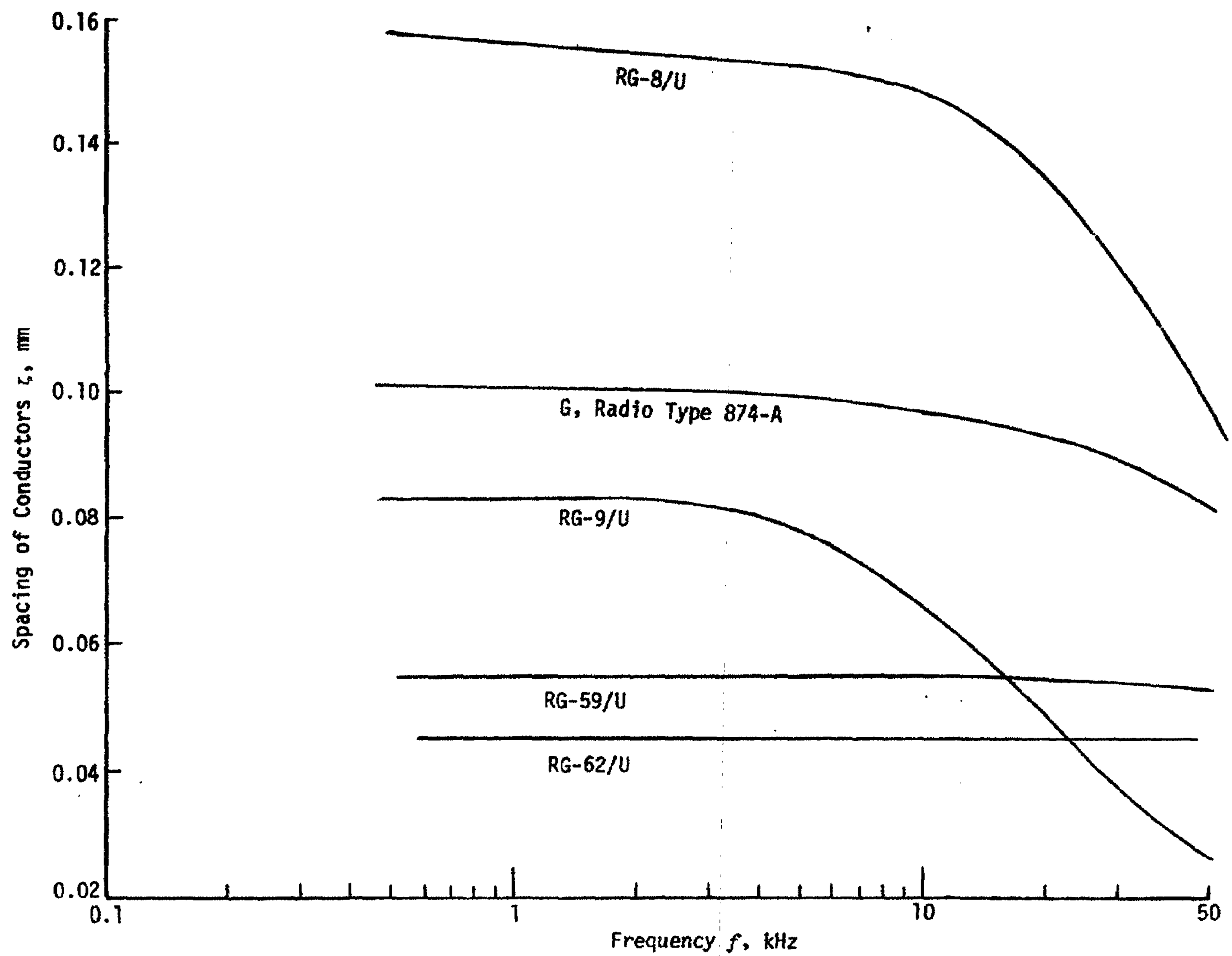


Figure 3-67. Spacing of Conductors in the Equivalent Two-wire Line as a Function of Frequency for Different Cables (Ref. 76)

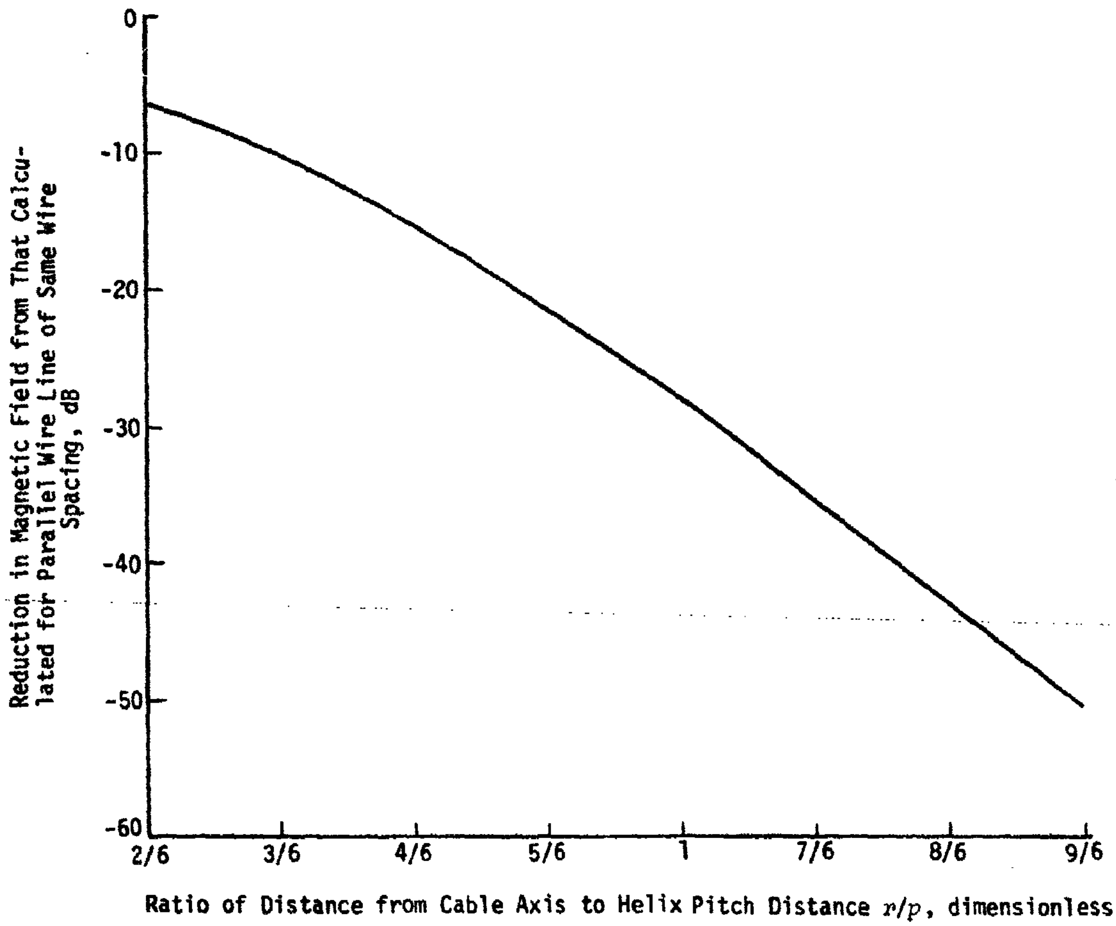


Figure 3-68. Correction Factor for Estimating Field from Twisted Wire Pair (Ref. 76)

an effective two-wire radiator, as shown in the bottom figure of Table 3-8, with the maximum field given by the two-wire model, or

$$B_{max} = \frac{\mu_0 I h}{\pi(h^2 + r^2)}, \text{T} \qquad (3\text{-}125)$$

where

h = distance between the cable and the ground plane, m

r = distance along the ground plane, m

Because the distance h is generally much greater than the conductor separation d or ζ in the two-conductor cable models previously discussed, the cancellation effect of the field from the cable image in the ground plane becomes significant only at a distance substantially greater than h. Thus, the common-mode field drops off much more slowly than the differential-mode field and can be an important source of EMI if it is not properly controlled.

The parallel-wire plot, Fig. 3-66 (Ref. 75), may be used to predict the magnetic flux density from all of the previously described cable types, if the parameter d is interpreted properly as in the following list (refer to Table 3-8 for meaning of symbols):

Cable Type	Value for d in Fig. 3-66
Single Wire	∞
Parallel Wire	d
Coaxial Cable	ζ (Fig. 3-67)
Twisted Pair	d (correction factor in Fig. 3-68 must be added to B)
Common Mode	$2h$

3-3.1.1.2.5 Leakage Fields of High Frequency

3-3.1.1.2.5.1 Solid Cables

Fig. 3-69 (Ref. 76) shows the coaxial cable geometry. The solution for the fields outside an infinitely long cable carrying a current $I \cos \omega t$ and whose outer conductor is several skin depths in thickness yields (Ref. 77)

$$\left.\begin{aligned} H_{\theta_3} &= \frac{Z_t I j\omega \epsilon_3 H_1^2(k_3 r)\exp(j\omega t - \gamma z)}{k_3 H_0^2(k_3 b)}, \text{A/m} \\ E_{z_3} &= \frac{Z_t I H_0^2(k_3 r)\exp(j\omega t - \gamma z)}{H_0^2(k_3 b)}, \text{V/m} \\ E_{r_3} &= \frac{Z_t \gamma I H_1^2(k_3 r)\exp(j\omega t - \gamma z)}{k_3 H_0^2(k_3 b)}, \text{V/m} \end{aligned}\right\} \quad (3\text{-}126)$$

where

$\gamma \approx -\omega^2\mu_1\epsilon_1$ = propagation constant, m^{-1}

$k_3^2 \approx -\omega^2(\mu_1\epsilon_1 - \mu_3\epsilon_3) = -(\beta_1^2 - \beta_3^2)$

$k_2^2 \approx -j\omega\sigma_2\mu_2$

ω = angular frequency, rad/s

$\beta = 2\pi/\lambda$, m^{-1}

λ_i = wavelength in medium i, m

μ_i = permeability in medium i, H/m

σ_i = conductivity in medium i, mhos/m

ϵ_i = permittivity in medium i, F/m

H_n^k = Hankel function of the kth kind, order n

r, θ, z = cylindrical coordinates of system

Z_t = surface transfer impedance, Ω/m

$$Z_t = \frac{E_z|_{r=b}}{I} = \left[\frac{k_2}{\sigma_2}\right]\left[\frac{1}{\pi a}\sqrt{\frac{a}{b}}\left(\frac{j}{\exp[-jk_2(b-a)]}\right)\right], \Omega/\text{m} \quad (3\text{-}127)$$

I = current carried on center conductor, A

An examination of Eq. 3-126 reveals that the external near fields all depend in the same manner on the characteristics of the shield which are embodied in the surface transfer impedance. Increasing the conductivity σ_2, permeability (μ_2), inner and outer radii (a and b), and thickness of the shield results in a reduction of the magnitudes of the external fields. The external fields are also dependent on the permittivity of the inner line ϵ_1, decreasing with an increase in ϵ_1. It should be noted, however, that an increase in ϵ_1 reduces the line power for the same current.

Eqs. 3-126 are rather complex to evaluate numerically, however they all exhibit the same dependence on the surface transfer impedance Z_t which can be expressed as

$$|Z_t| = \frac{2\sqrt{2}\,T}{\delta} R_o \exp(-T/\delta), \Omega/\text{m} \quad (3\text{-}128)$$

where

δ = depth of penetration $= \sqrt{\dfrac{2}{\omega\sigma_2\mu_2}}$, m

T = thickness of outer conductor, $b - a$, m

R_o = dc resistance per unit of length of outer conductor, Ω/m

This can be written in terms of a normalized surface transfer impedance function $f(u)$ where

$$|Z_t| = R_o f(u), \Omega/\text{m} \quad (3\text{-}129)$$

where

$$u = \frac{2T}{\delta}$$

$f(u)$ is shown on Fig. 3-70 (Ref. 78). Fig. 3-71 (Ref. 79) shows the surface transfer impedance for 1/8-in. OD copper, mild steel, and stainless steel as a function of frequency.

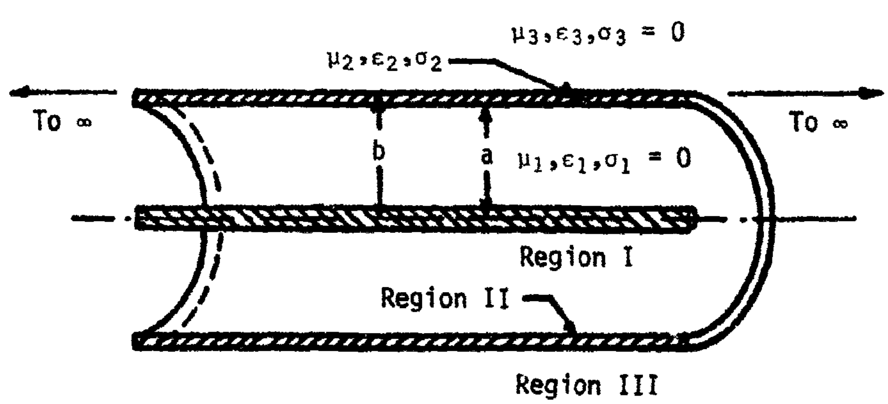


Figure 3-69. Coaxial Cable Geometry and Parameters (Ref. 76)

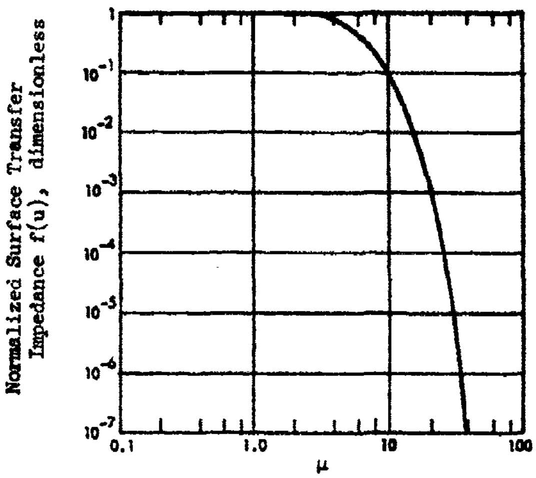


Figure 3-70. Plot Showing Frequency Dependence of Normalized Surface Transfer Impedance for Solid Shields (Ref. 78)

3-3.1.1.2.5.2 Leakage from Braided Shields

Electromagnetic leakage fields, leaking through single gaps and apertures in conductive shields, can be interpreted as emanating from fictitious magnetic and electric dipoles located in the gaps or apertures (Ref. 80).

The equations which are given for leakage from solid cables are valid for the braided shield if one uses an "equivalent surface transfer impedance" in place of the transfer impedance given by Eq. 3-127.

$$Z_t = \left(\frac{Z_0}{n\lambda}\right)(0.3\cos\alpha)\left(\frac{g}{a}\right)^4 \times (1+\tan^4\alpha-\sin^4\alpha)\,,\ \text{for } \frac{g}{a}<1\ ,\Omega/\text{m} \quad (3\text{-}130)$$

where

λ = wavelength in free space, m

Z_0 = impedance of free space, 377Ω

n = number of braid wires or braid wire strands which form the diamond shaped mesh apertures

g, a, and α are shown in Fig. 3-72

Fig. 3-73 (Ref. 79) shows the general shape of the curve for a braided cable.

The transfer impedance, Eq. 3-130, increases linearly with frequency. This is observed at the higher frequencies where the transfer impedance behaves as an inductance. At the lower frequencies the resistive component of the transfer impedance is dominant; the curve is flat with respect to frequency, then decreases as frequency increases as in the case of a solid conductor until the inductive component becomes dominant. Results from Ref. 79 are shown on Figs. 3-73, 3-74, and 3-75 for RG-59 C/U, RG-223U, and RG-213/U cables, respectively, as measured with a triaxial tester. Figs. 3-76 through 3-80 show results taken from Ref. 78 for RG-62B/U single shield, RG-216/U double shield; RG-58/U triax, and RG-12/U with the armor: (a) floating, and (b) connected at both ends.

3-3.1.1.2.5.3 Induced Fields

When the distance $r \ll L$, where L = the actual length of the cable, the infinite cable approximation can be used.

For the case of $|k_3 r| \ll 1$, Eqs. 3-126 reduce to

$$H_{\theta_3} = \frac{-2Z_t I\omega\epsilon_3 \exp(j\omega t-\gamma z)}{\pi k_3^2 r\left\{1+j\frac{2}{\pi}\left[\ln\left(\frac{2}{k_3 b}\right)-0.577\right]\right\}} \quad \text{A/m} \quad (3\text{-}131a)$$

$$E_{z_3} = \frac{Z_t I\left\{1+j\frac{2}{\pi}\left[\ln\left(\frac{2}{k_3 r}\right)-0.577\right]\right\}\exp(j\omega t-\gamma z)}{\left\{1+j\frac{2}{\pi}\left[\ln\left(\frac{2}{k_3 b}\right)-0.577\right]\right\}}, \quad \text{V/m} \quad (3\text{-}131b)$$

$$E_{r_3} = \frac{-2Z_t I\gamma j \exp(j\omega t-\gamma z)}{\pi k_3^2 r\left\{1+j\frac{2}{\pi}\left[\ln\left(\frac{2}{k_3 b}\right)-0.577\right]\right\}}, \quad \text{V/m} \quad (3\text{-}131c)$$

To give some idea of the magnitudes involved in the expansion, we have the following example: for $\omega = 10^6$, $\epsilon_1 = 2\epsilon_3$, k_3 of the order of 10^{-3} and $k_3 r \le 0.1$, r can be as large as 100 m.

For the case of the distances of observation of the order of the outer radius of the coaxial cable

$$\left.\begin{aligned} H_\theta &= \left(\frac{j\omega\epsilon_3}{k_3}\right) Z_t I\left[\frac{1}{k_3 r\ln\left(\frac{2}{k_3 b}\right)}\right], \text{A/m} \\ E_{r_3} &= \left(\frac{\gamma}{k_3}\right) Z_t I\left[\frac{1}{k_3 r\ln\left(\frac{2}{k_3 b}\right)}\right], \text{V/m} \\ E_{z_3} &= Z_t I\left[\frac{\ln\left(\frac{2}{k_3 r}\right)}{\ln\left(\frac{2}{k_3 b}\right)}\right], \text{V/m} \end{aligned}\right\} \quad (3\text{-}132)$$

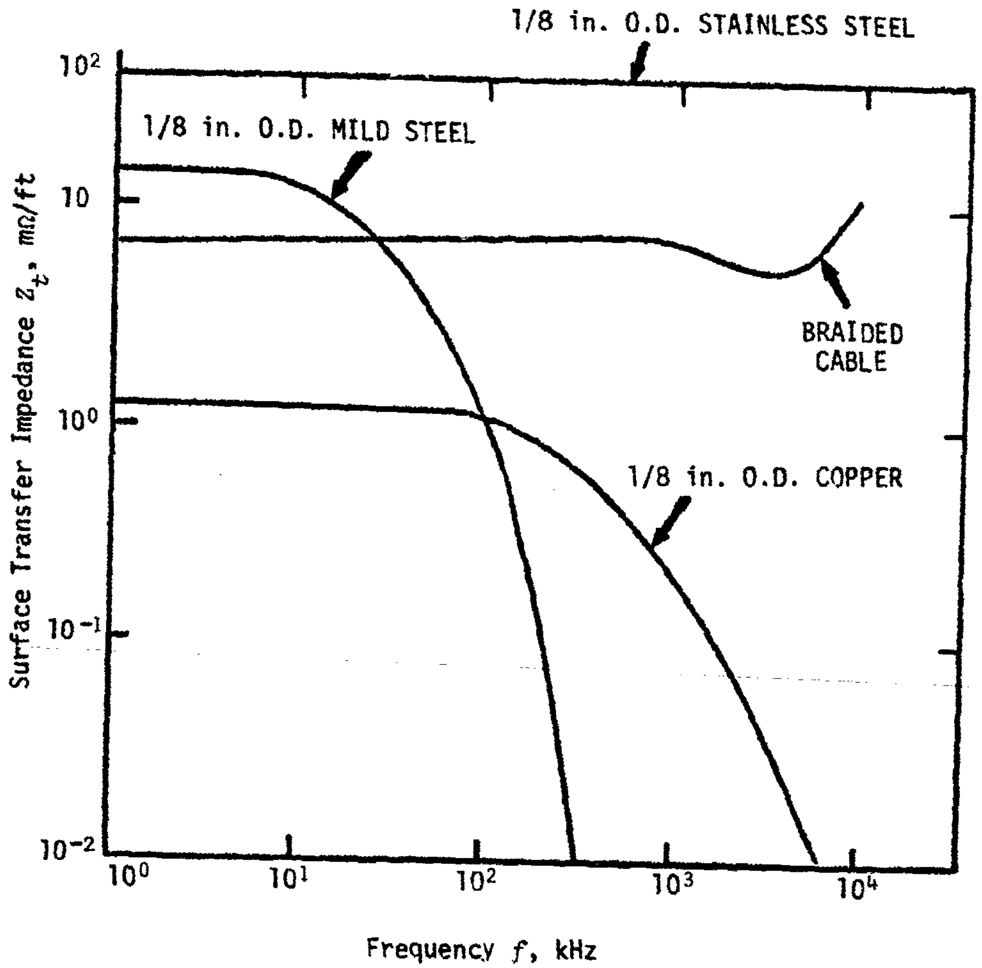


Figure 3-71. Surface Transfer Impedance

3-3.1.1.2.5.4 Radiation from Cables of Finite Length

In the far-field, leakage emanating from cables of finite length (Ref. 76) may be thought to be caused by an equivalent leakage radiation current I_e flowing along the axis of the cable. This equivalent leakage radiation current is obtained from the angular magnetic field leakage component close to the cable for infinitely long cables.

$$I_e = \frac{2\pi j\omega\epsilon_3 Z_t I \exp(j\omega t - \gamma z)}{k_3^2 \ln\left(\frac{2}{k_3 b}\right)}, \text{A} \qquad (3\text{-}133)$$

Thus, the evaluation of the leakage fields from cables of finite length can be treated by solving for the fields of an antenna with a traveling wave type current distribution. A typical radiation pattern is shown in Fig. 3-81 (Ref. 80). However, the possibility of a standing wave must be considered.

3-3.1.1.3 Magnetic Induction Susceptibility

A magnetic field generates a voltage in the susceptor circuits by Faraday's law, which for sinusoidal signals takes the form

$$V = \omega AB, \text{V} \qquad (3\text{-}134)$$

where

A = effective area of circuit, m²
B = magnetic flux density normal to plane of area A, T
$\omega = 2\pi f$ = radian frequency, rad/s

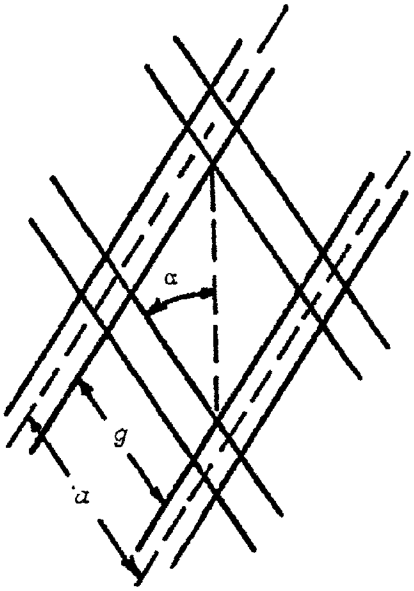


Figure 3-72. Dimensional Notation for Braided Shield

Eq. 3-134 may be used to predict the interference voltage generated in cables as well as circuits, providing the effective area is known or can be determined from the cable geometry.

For cables, the effective area A is found as in the following table (Ref. 75). (Refer to Table 3-9 for definitions of symbols.)

Cable Type	A
Parallel-wire	$d \times$ length
Coaxial	$\xi \times$ length
Twisted Pair	projection of helix ≈ 0.318 pd (independent of length)
Common Mode	$2h \times$ length

Eq. 3-134 is plotted in Fig. 3-82 (Ref. 75) as V/B versus f with A as a parameter. For all cables except the twisted pair, the voltage obtained from the plot is the voltage induced per meter of length, per unit flux density. For twisted pair the induced voltage is independent of length.

3-3.1.2 Electric Coupling

Two circuits with a mutual capacitance were illustrated on Fig. 3-57(B). If the capacitance is distributed, the equivalent circuit may be difficult to define and computations may not be accurate because of uncertainty as to how to assign proper capacitance values. Capacitance effects can be quite serious in cases of internal coupling, so the designer must have some way of estimating their magnitudes. Formulas for standard configurations can be found in handbooks on radio engineering (Ref. 2).

For cables, electric coupling is probably most important between unshielded wires in a common cable harness. The capacitance per unit length depends not only on wire sizes and spacing between any given pair, but also on the presence of other wires in the cables.

Fig. 3-83 (Ref. 81) illustrates a simple model for the electric coupling in a cable. The interfering voltage E_i couples through stray capacitance C_c to produce voltage E_x on an adjacent cable. The interfering cable and the adjacent cable have stray capacitances to ground, C_a and C_b. Each cable has its system loads, Z_1 and Z_2, and Z_3 and Z_4, across which the stray capacitances appear. If Z_x is a high resistance load R_x

$$\frac{E_x}{E_i} = \frac{C_c}{C_c + C_b}\left\{\frac{R_x^2}{R_x^2 + \left[\frac{1}{2\pi f(C_c + C_b)}\right]^2}\right\}^{1/2} \quad (3\text{-}135)$$

The plot of E_x/E_i against frequency is shown in Fig. 3-84 (Ref. 81). When Z_x contains inductive reactance, resonances may cause variations in coupling with frequency. Such effects are likely to be noticeable at the higher frequencies and are likely to exhibit broad resonances due to loading in dielectrics and connected circuits. When Z_x contains capacitive reactance, it is equivalent to an increase in C_b.

The capacitance between two parallel wires above a ground plane can be calculated using the formulas

$$C = \frac{(7.35 \times 10^{-12})(l)\left[\log\left(\frac{S_{12}}{D}\right)\right] K_{eff}}{\left[\log\left(\frac{4h}{d}\,\frac{1}{\sqrt{2 - \sqrt{\frac{d}{D}}}}\right)\right]^2 - \left[\log\left(\frac{S_{12}}{D}\right)\right]^2}, \text{F} \quad (3\text{-}136)$$

where

$$K_{eff} = 1 + \frac{\left(\frac{d_1}{d}\right)^2 - 1}{\frac{1}{2}\left(\frac{d_1}{d} + \frac{D}{d}\right)^2 - 1}(\epsilon_r - 1)$$

$$S_{12} = \sqrt{D^2 + 4h^2}\,,$$

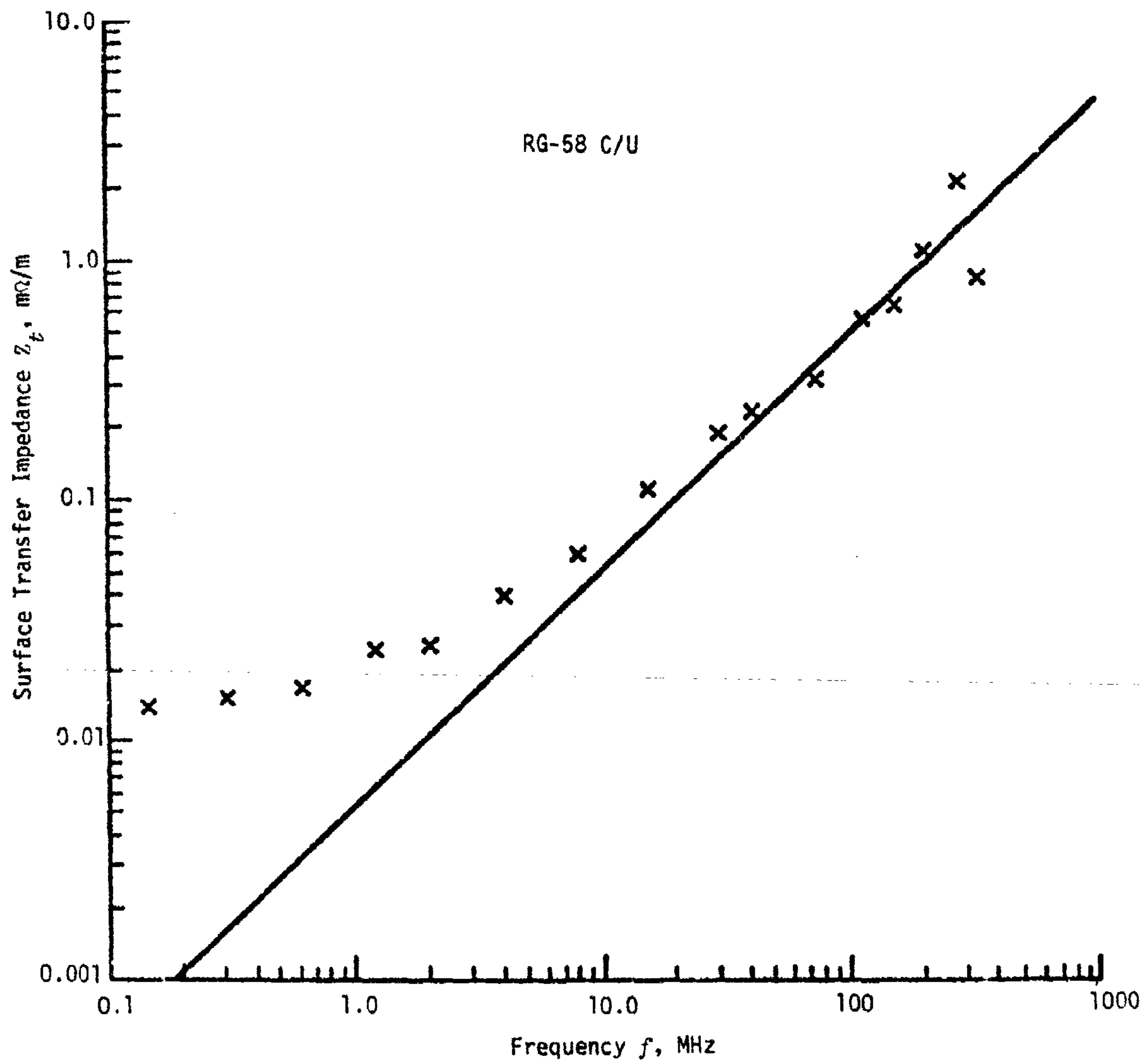


Figure 3-73. Surface Transfer Impedance of RG-58 C/U Cable (Ref. 79)

l = length of wires, ft
D = separation of wires, in.
h = height above ground plane, in.
d = diameter of the wire conductor in.
d_i = diameter of the wire including insulation, in.
ϵ_r = relative dielectric constant of the wire insulation, dimensionless

The factor $(2 - \sqrt{d/D})^{-1/2}$ in the capacitance equation has been determined experimentally.

3-3.1.3 Combined Electric and Magnetic Coupling, Low Frequency Case

At low frequencies, the effects of electric and magnetic coupling may be approximated by lumping these parameters, as shown on Fig. 3-85 (Ref. 81). It is seen that the total mutual inductance is considered to be located between two capacitors, each of which is one-half of the static coupling capacitance between the two lines. This approximate summation has

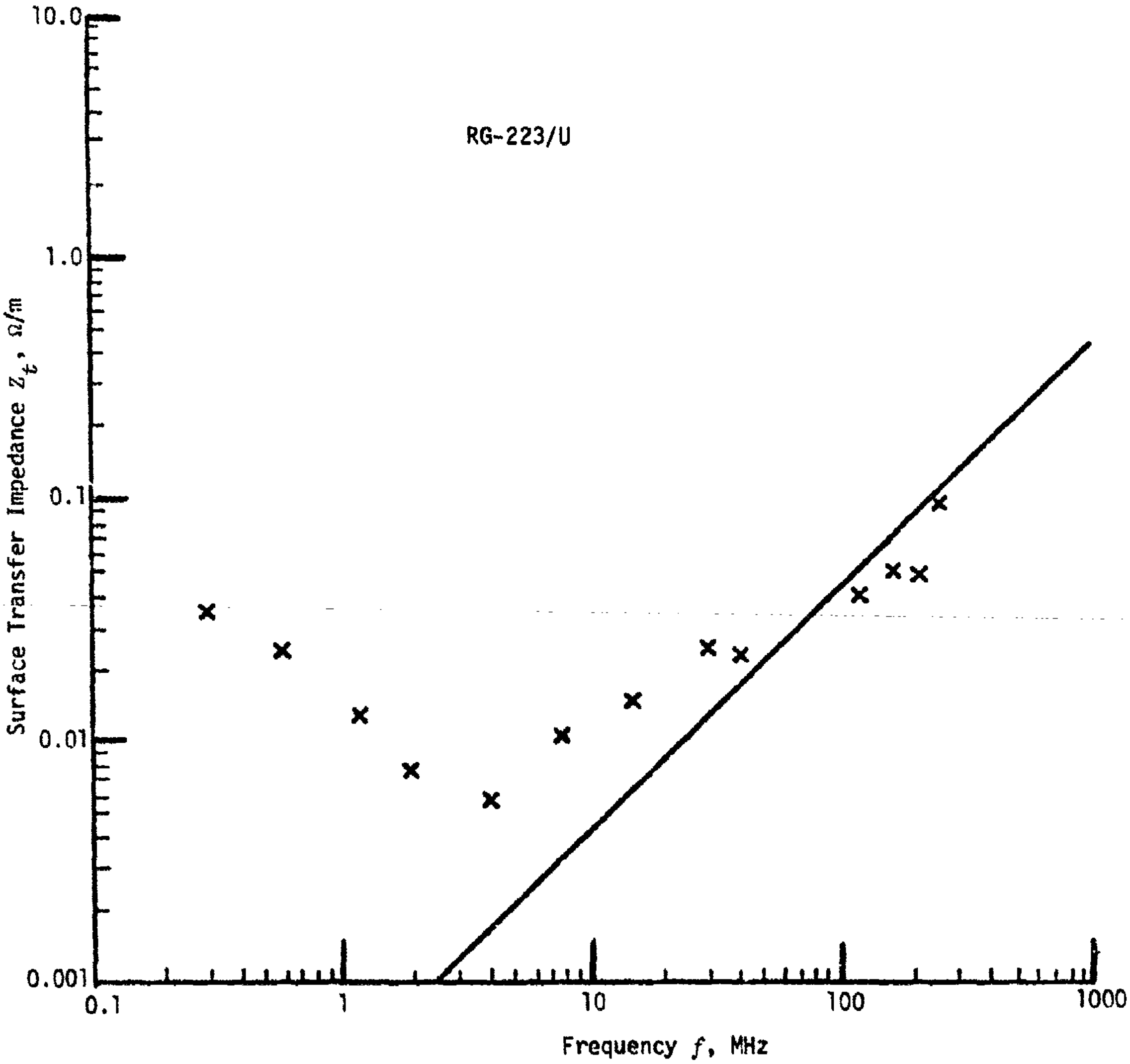


Figure 3-74. Surface Transfer Impedance of RG 223/U, Formerly RG-55/U, Cable (Ref. 79)

validity only at frequencies well below those for which the line is a quarter wavelength long.

Note from the formulas in Fig. 3-85 that at the end of the susceptible circuit adjacent to the noise generator E_o, the coupled-in voltage E_{2G} is the sum of the electric and magnetic components. At the opposite end of the susceptible circuit, the coupled-in voltage is shown as the difference of the electric and magnetic components. For the approximation made here, the two components of E_{2L} are in phase opposition. Rather than depend upon a difference voltage, the larger of the two components is assumed to be the coupled-in interference. As frequency increases, it is common to assume that voltage transfers are proportional to frequency until unity transfer is reached, and then the transfer is approximated at unity for all higher frequencies.

Fig. 3-86 (Ref. 81) shows a plot of the approximate voltage transfer ratio, E_{2L}/E_o which has a 6 dB/octave rise up to unity. Typical experimental data are plotted with the approximation. It will be noted that the approximation results in slightly more

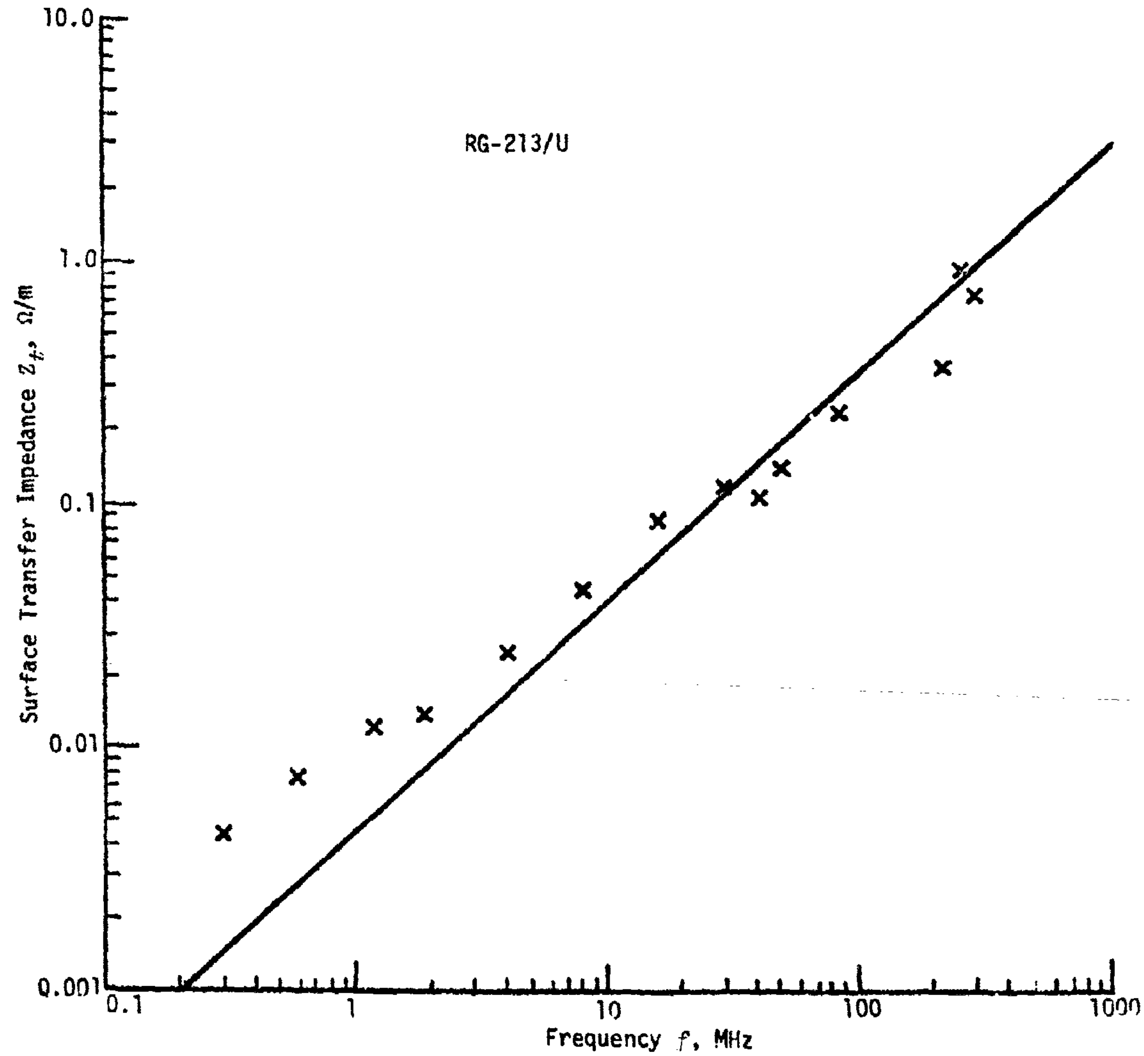


Figure 3-75. Surface Transfer Impedance of RG-213/U, Formerly RG-8A/U, Cable (Ref. 79)

coupled-in interference than is shown by the experimental data. This usually will be experienced; however, in situations of light circuit loading, resonances in the wiring may produce an inversion, so that experimental data may rise above the approximation in a narrow frequency band.

For a considerable variation in loading in the vicinity of 300 Ω, $K_G = 2K_L$ (K_G, K_L defined on Fig. 3-85) is a very good approximation.

3-3.1.4 Solutions at Higher Frequencies

General solutions of the coupling between two cables are available only in formal notation. In some cases specific solutions have been worked out (Refs. 79 and 114).

3-3.1.5 Multiconductor Coupling

The general configuration of n conductors with distributed capacitance and inductance coupling is difficult to analyze. Although no general solution has been found, various approximate solutions appear in the literature. Ref. 82 discusses lumped mutual capacitances and conductances.

3-3.2 RADIATION FIELD COUPLING

When source and susceptor are separated by distance large compared with (1) one-sixth of a wavelength, and (2) the maximum dimensions of either, the coupling can be considered to be by means of the radiation field. At these distances the field generated

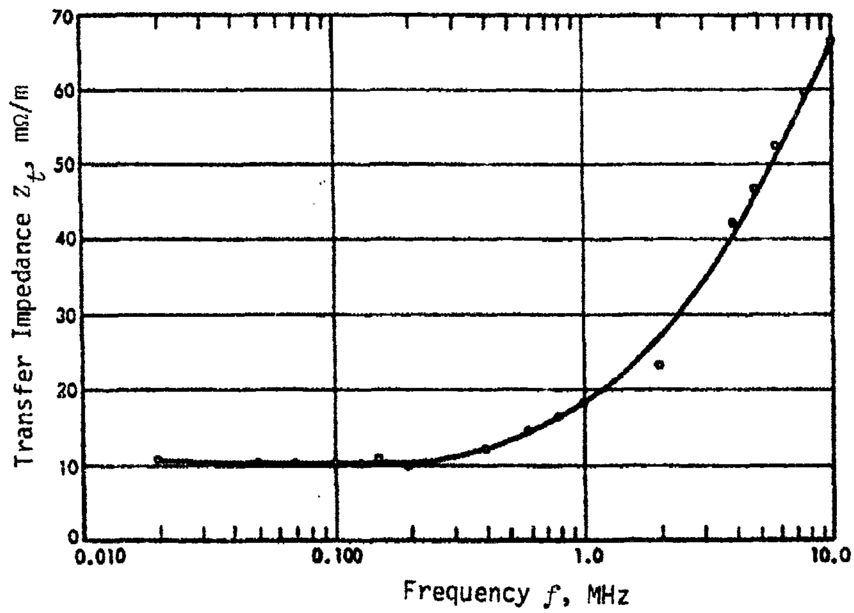


Figure 3-76. Transfer Impedance vs Frequency, RG-62B/U, Formerly RG-13A/U (Ref. 78)

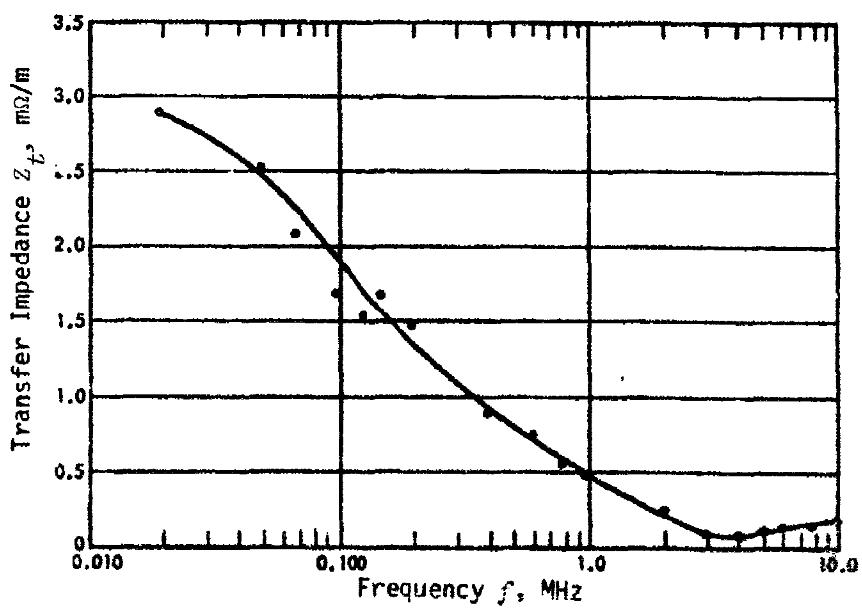


Figure 3-77. Transfer Impedance vs Frequency, RG-216/U, Formerly RG-13A/U (Ref. 78)

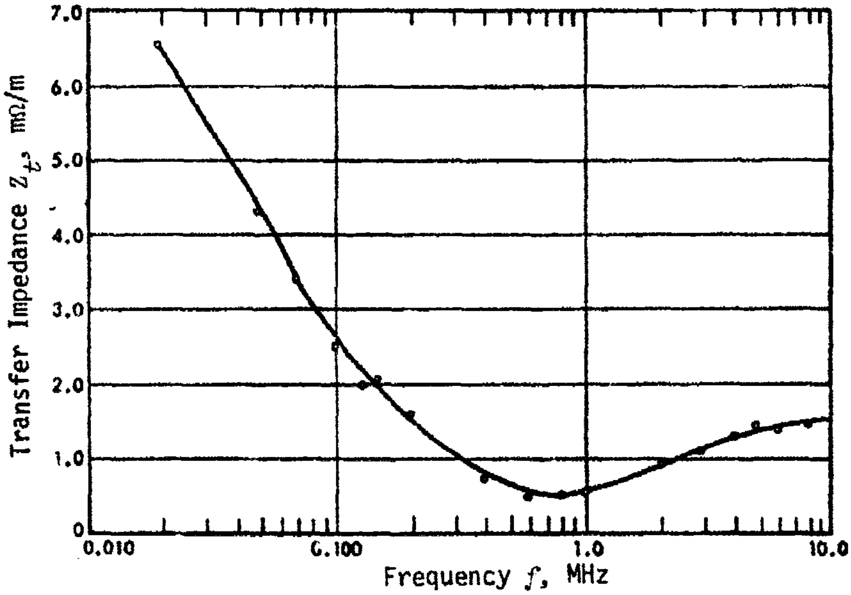


Figure 3-78. Transfer Impedance vs Frequency, RG-58 A/U Triax (Ref. 78)

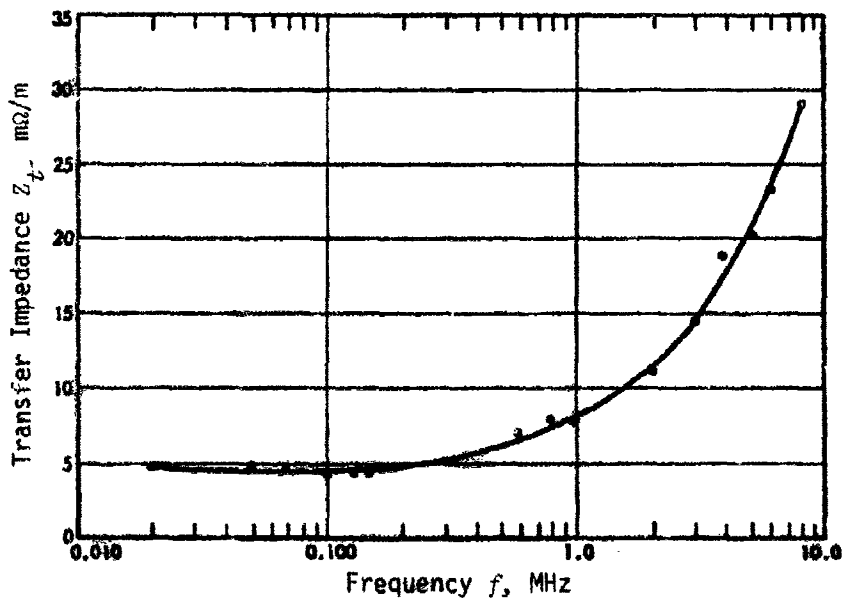


Figure 3-79. Transfer Impedance vs Frequency, RG-12/U Armored (armor floating) (Ref. 78)

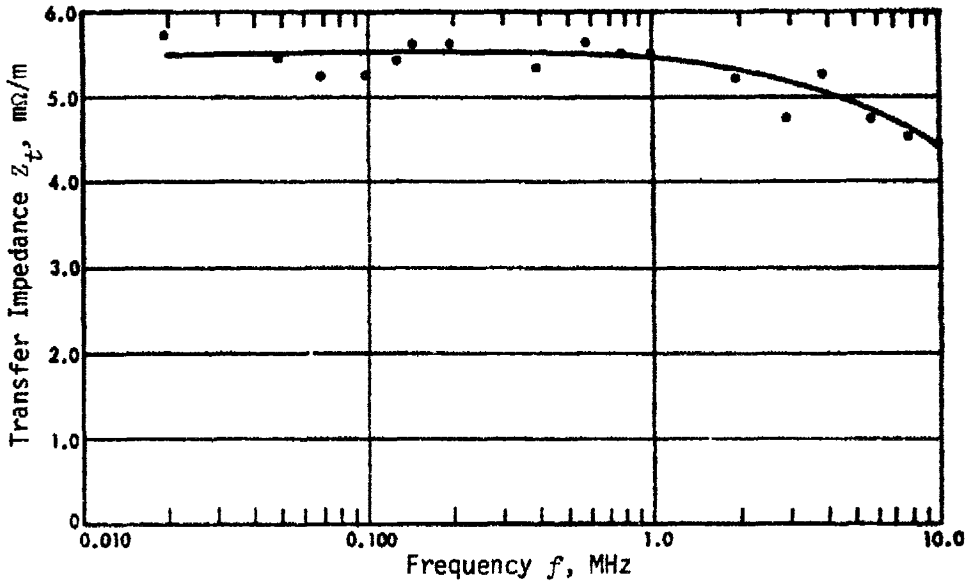


Figure 3-80. Transfer Impedance vs Frequency, RG 12/U Armored (armor bonded to cable shield at ends) (Ref. 78)

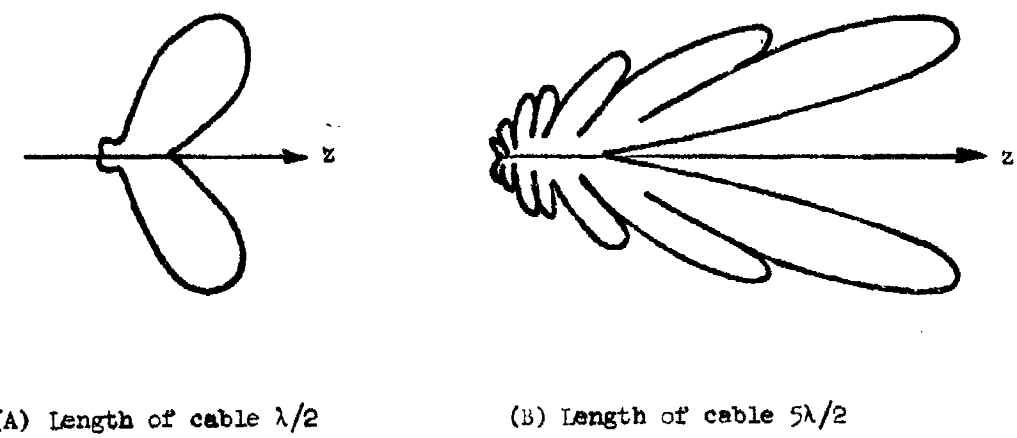


Figure 3-81. Patterns for Field from Leakage Currents (cable in z-direction) (Ref. 80)

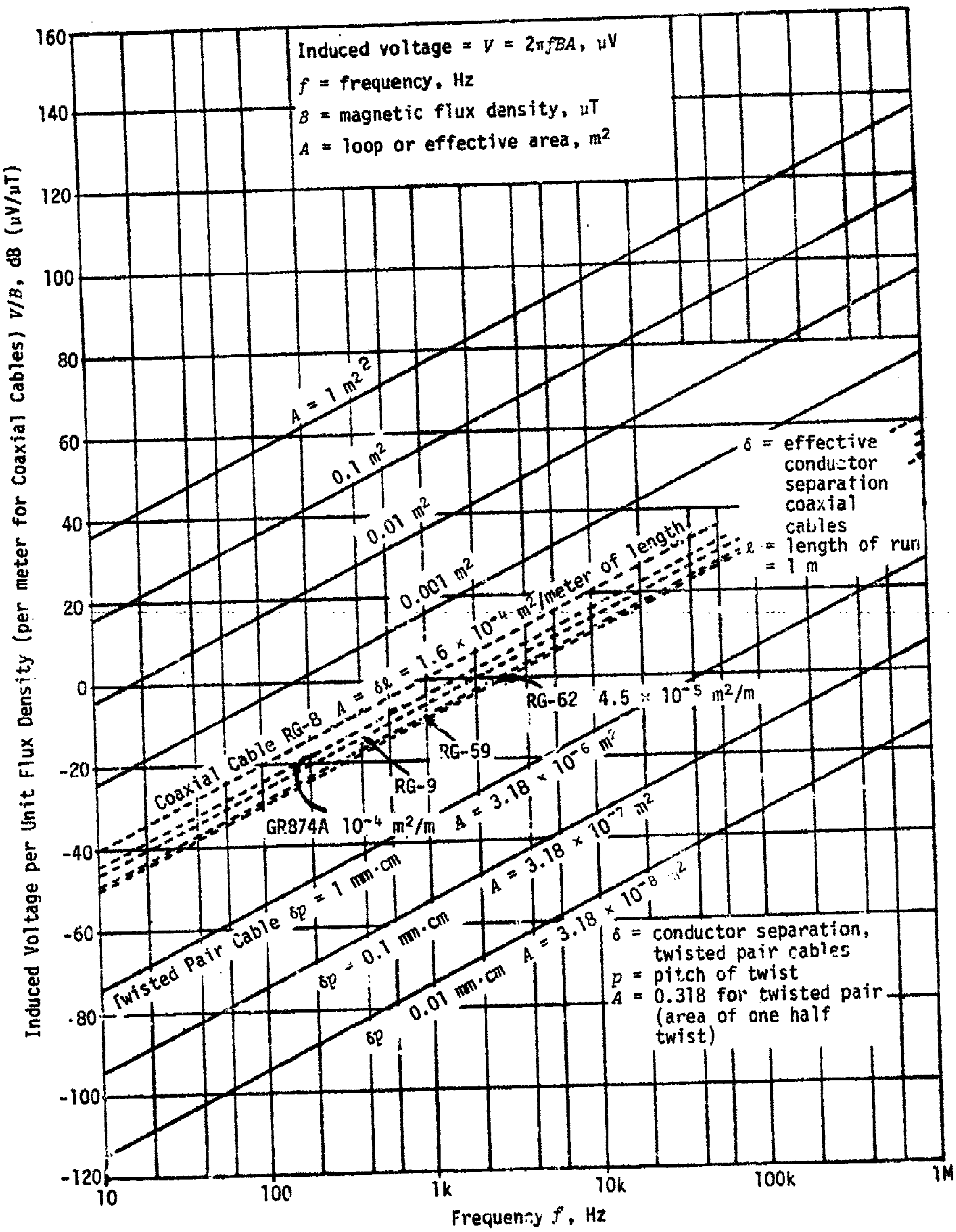


Figure 3-82. Cable and Loop Susceptibility (Ref. 75)

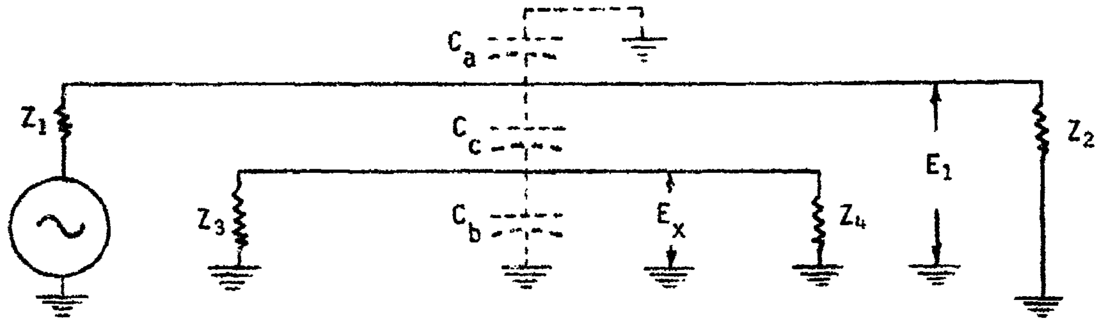


Figure 3-83. Model for Electric Coupling

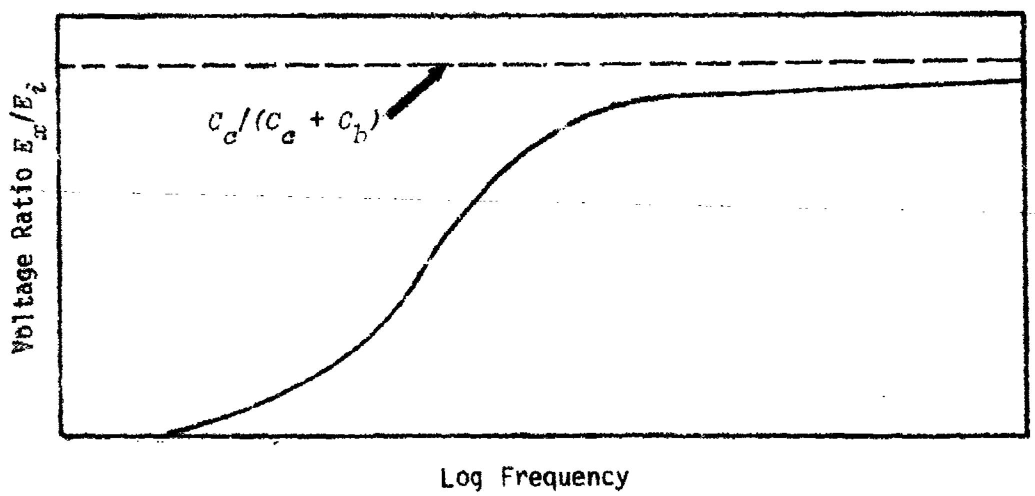


Figure 3-84. Frequency Dependence of Voltage Ratio for Electric Coupling (See Fig. 3-83 for definitions of $C_b + C_c$) (Ref. 81)

by the emitter resembles that of a plane wave, but the field components fall off inversely as the distance from the source. The presence of the susceptor produces negligible reaction on the emitter.

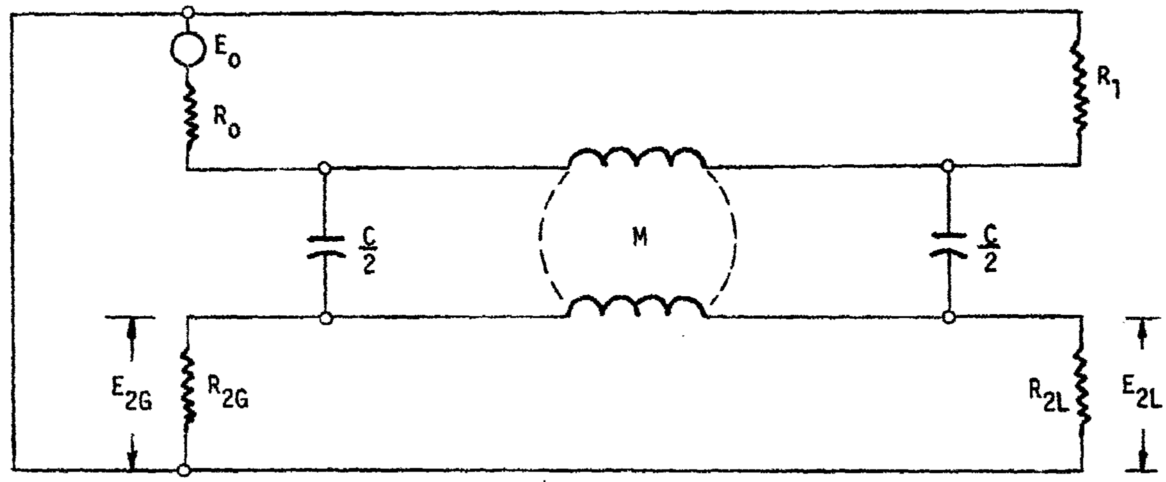
The treatment here is based upon antenna theory, because devices which are emitting or receiving energy by "radiation" are acting essentially as antennas. An equipment cabinet at a frequency at which its length is a multiple of a quarter wave length has the potential to be as effective an antenna as a device that is intentionally designed for the purpose.

3-3.2.1 The Elementary Dipoles

The elementary electric and magnetic dipoles may be considered to be the prototypes for all antennas or radiating objects. The radiating properties of objects of finite size can be synthesized from assemblies of such dipoles.

3-3.2.1.1 The Magnetic Dipole

Equations for the field of an infinitesimal magnetic dipole, as a function of distance from the dipole are given in Eqs. 3-113, par. 3-3.1.1.1. Specific components of these fields vary inversely with the distance, the distance squared, and the distance cubed. The magnitudes of the several components of any one field vector are equal to a distance of $r = \lambda/(2\pi)$; at closer distances, in the reactive near field region (see Glossary), the $(1/r)^2$ and $(1/r)^3$ terms are dominant; at greater distances in the "radiation" field, the $1/r$



ELECTRIC COMPONENT

MAGNETIC COMPONENT

$$\frac{E_{2G}}{E_0} = \left[\frac{R_1}{R_1 + R_0} \times \frac{R_2}{X_c}\right] + \left[\frac{X_M}{R_1 + R_0} \times \frac{R_{2G}}{R_{2G} + R_{2L}}\right] = K_G f$$

$$\frac{E_{2L}}{E_0} = \left[\frac{R_1}{R_1 + R_0} \times \frac{R_2}{X_c}\right] - \left[\frac{X_M}{R_1 + R_0} \times \frac{R_{2L}}{R_{2G} + R_{2L}}\right] = K_L f$$

$$R_2 = \frac{R_{2L}\,R_{2G}}{R_{2L} + R_{2G}}$$

Figure 3-85. Circuit and Equations Representing Electric Coupling and Magnetic Coupling Between Parallel Lines (Ref. 81)

terms are dominant. Phenomena due to the $(1/r)^3$ term were discussed in par. 3-3.1.1.1.

In the radiation field

$$E_\varphi = \left(\frac{\eta_0 \beta^2 I A}{4\pi r}\right) \sin\theta \text{ , V/m} \quad (3\text{-}138)$$

$$H_\theta = -\left(\frac{\beta^2 I A}{4\pi r}\right) \sin\theta \text{ , A/m} \quad (3\text{-}139)$$

where

A = area of a loop (assumed to consist of only a single turn) carrying a current of I amperes, m². The product AI is the magnetic dipole moment.

r = distance from source, m

$\eta_0 = \sqrt{\mu_0/\epsilon_0}$ = characteristic impedance of free space, Ω

$\beta = 2\pi/\lambda$, m^{-1}

θ, φ = space coordinates defined on Fig. 3-60.

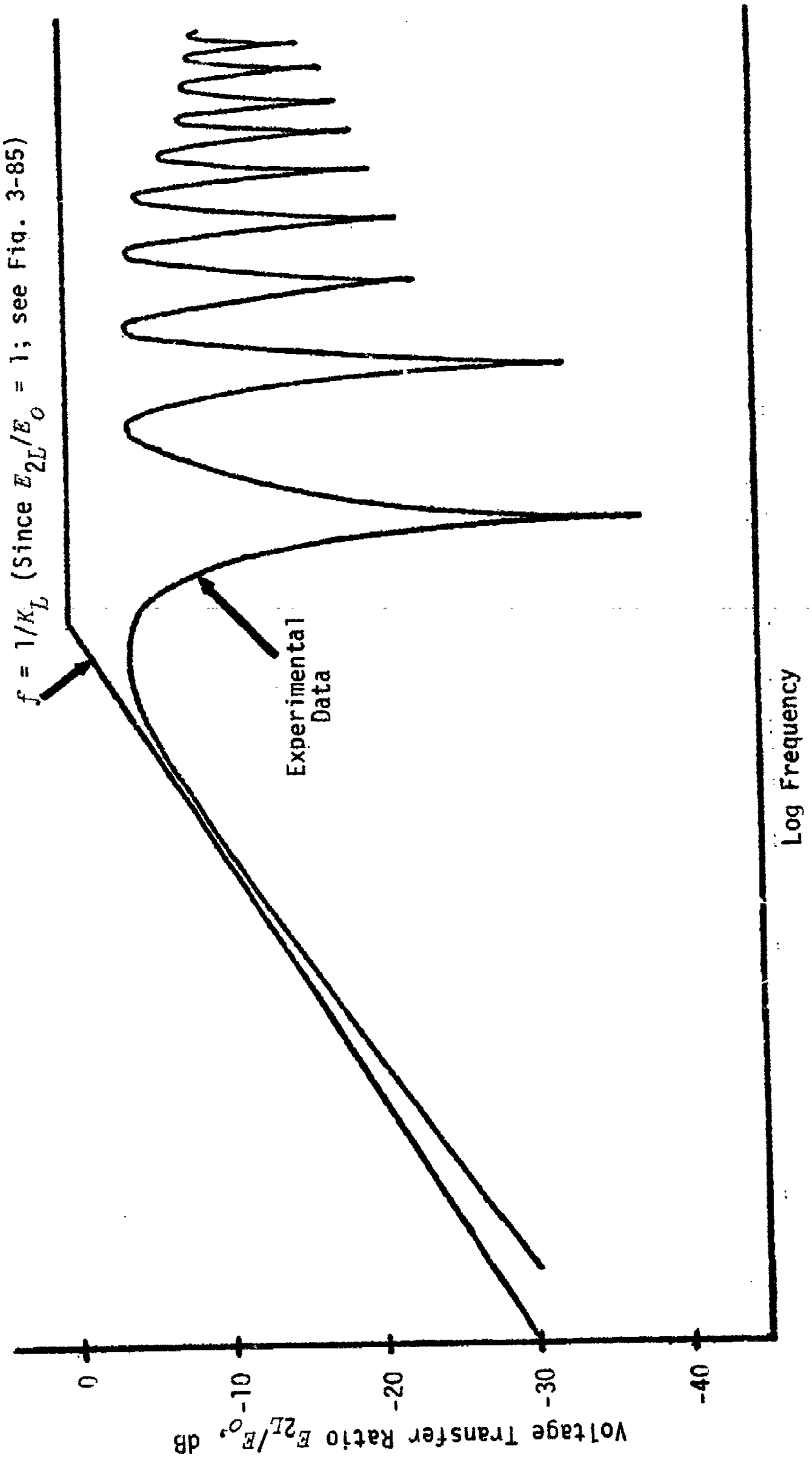


Figure 3-86. Approximate Signal Coupling in Wiring

The power density, defined as the rate of energy flow through unit area, is obtained from the Poynting vector $\vec{\mathbf{P}}_D$

$$|\vec{\mathbf{P}}_D| = |\vec{\mathbf{E}} \times \vec{\mathbf{H}}| = E_\varphi H_\theta = \frac{E_\varphi^2}{\eta_0}$$

$$= \eta_0 H_\varphi^2 = \frac{\eta_0 \beta^4 l^2 A^2}{16\pi^2 r^2} \sin^2\theta \text{ , W/m}^2 \quad (3\text{-}140)$$

3-3.2.1.2 The Electric Dipole

The fields produced by an elementary electric dipole are given by (Ref. 73)

$$\left.\begin{aligned} E_r &= \frac{Il\beta^3}{\omega 2\pi\epsilon}\left(-j\frac{1}{\beta^2 r^2} + \frac{1}{\beta^3 r^3}\right)\cos\theta \text{ , V/m} \\ E_\theta &= \frac{Il\beta^3}{\omega 4\pi\epsilon}\left(-\frac{1}{\beta r} - j\frac{1}{\beta^2 r^2} + \frac{1}{\beta^3 r^3}\right)\sin\theta \text{ , V/m} \\ H_\varphi &= \frac{Il\beta^2}{4\pi}\left(\frac{1}{\beta r} + j\frac{1}{\beta^2 r^2}\right)\sin\theta \text{ ; A/m} \end{aligned}\right\} (3\text{-}141)$$

where

E_r, E_θ = r and θ components of the electric field, respectively, V/m

H_φ = angular component of magnetic field, A/m

I = dipole current, assumed in direction of θ-axis (Fig. 3-60), A

l = dipole length, assumed small compared with r and located at origin of coordinates, m

ω = angular frequency,

and other terms are as defined in par. 3-3.1.1.1.1, Eq. 3-113.

From these relations, the Poynting vector yields

$$|\vec{\mathbf{P}}_D| = E_\theta H_\varphi = \left(\frac{\eta_0 \beta^2 I^2 l^2}{16\pi^2 r^2}\right) \sin^2\theta \text{ , W/m}^2 \quad (3\text{-}142)$$

In this field region, the main factor that distinguishes different emitters is their radiation pattern. The equations for the infinitesimal dipole can be used for finite size dipoles if their dimensions are substantially less than a quarter wave length.

3-3.2.2 Antenna Gain

As the frequency increases, the relative size of any given object increases when measured in terms of wavelengths. In general, as the size increases in this way the possible radiation patterns become more complex. To obtain directivity, antennas usually are designed using dipole arrays for frequencies up to at least 100 MHz, and "aperture" antennas above that. An aperture antenna is usually a continuous metal surface. Its properties can be computed in various ways, including representing it as an extensive dipole array.

Detailed properties of arrays or apertures are discussed in par. 5-10. One of their significant properties is that of antenna gain.

Antenna gain, sometimes called directive gain, is the ratio of the power required at the input of a reference antenna to the power supplied to the input of a given antenna to produce, in a given direction, the same field at the same distance. When not otherwise specified, a stated gain is the gain in the direction of the main radiation lobe. Two types of reference antenna have been used in the past: (1) the half-wave dipole, and (2) the isotropic radiator. The first has been used primarily at frequencies where dipole types are prevalent. The second is commonly used for antennas operated in the microwave region of the spectrum. Even though it cannot be realized in practice, the isotropic antenna is a convenient reference, and is used in this handbook. An alternate term used to express the same property is effective aperture. The effective aperture A_{eff} is given in terms of the gain G by

$$A_{eff} = \frac{\lambda^2 G}{4\pi} \text{ , m}^2 \quad (3\text{-}143)$$

This usually is given for the maximum gain; i.e., $G = G_{max}$

It is customary to define the boundary separating the Fresnel (close) and Fraunhofer (far) regions of an antenna as

$$r = \frac{2D^2}{\lambda} \text{ , m} \quad (3\text{-}144)$$

where

r = distance measured from the antenna whose greatest dimension normal to the direction of r is D.

It is only at locations in the Fraunhofer region that one can count on the electromagnetic field component magnitudes (E and H) varying inversely with the distance from the antenna. For an aperture antenna in the microwave region, the critical distance can be hundreds or even thousands of meters, whereas for an elementary dipole it is effectively $\lambda/6$.

In the Fraunhofer region, the electric field streng at a distance r from a lossless antenna with gain G

(ratio) and radiated power P_t (in watts) is given by the expression

$$E = \frac{\sqrt{30\,P_t G_t}}{r}, \text{V/m} \qquad (3\text{-}145)$$

This formula is useful in estimating the field strength due to any given source at the location of a possible susceptor, if both its radiated power and its effective gain in the direction of the susceptor are known. This relation is plotted on Fig. 2-3.

In the Fresnel region, the field components cannot be calculated accurately using Eq. 3-145. Correction factors for certain types of antennas are given in par. 5-10.

3-3.2.3 Characteristics of Simple Antennas

The effectiveness of an antenna as a radiator can be specified in terms of its radiation resistance R_a in accordance with the relation

$$P_t = R_a I_{rms}^2 \text{ , W} \qquad (3\text{-}146)$$

where

I_{rms} = rms value of the current at the antenna terminals, A

At frequencies below 10 MHz, most incidental radiators can be considered to be electrically short. As a consequence, they radiate poorly. The short linear antenna has low radiation resistance, but has a high input impedance (corresponding to a capacitance of a few picofarads for typical configurations). For comparable dimensions, the loop also has a low radiation resistance, but has a relatively low input impedance (an inductance of several microhenries). Table 3-9 summarizes formulas for radiation resistance and gain for these types and the half-wave dipole.

As one moves in frequency above 10 MHz, the dimensions of equipment are such as to permit almost any metallic structure to become an efficient radiator at some frequency. To avoid this phenomenon, special attention should be paid to: (1) avoiding excitation from stray or leakage currents and fields, and (2) installation using proper grounding and bonding techniques as discussed in par. 4-7.

3-3.2.4 Field Susceptibility

Undesired electromagnetic fields can be coupled into sensitive circuits either directly, by means of an antenna as in the case of a receiver or an exposed input terminal, or indirectly through power or control cables.

Just as a changing magnetic field induces a voltage in a closed loop (see par. 3-3.1.1.3), an electric field

TABLE 3-9. CHARACTERISTICS OF SIMPLE ANTENNAS

Antenna	Radiation Resistance, Ω	Gain	Effective Length, m
Short Linear (above ground plane) $l \ll \frac{\lambda}{4}$	$20\pi^2\left(\frac{l}{\lambda}\right)^2$	1.5 (1.7 dB)	$\frac{l}{2}$
Half-wave Dipole $l = \frac{\lambda}{2}$	73	1.64 (2.14 dB)	$\frac{\lambda}{\pi}$
Circular Loop diam $\ll \frac{\lambda}{4}$	$3.8 \times 10^{-5}(Nf^2r^2)^2$	1.5 (1.76 dB)	$6.57 \times 10^{-[illegible]}$ – $(Nfr^{[illegible]})$

l = antenna length, m
λ = wavelength, m
r = loop radius, m
f = frequency, MHz
N = number of turns

will induce a voltage in an exposed metallic conductor. The effective length of the conductor is defined as the ratio of the open circuit voltage appearing at its terminals to the component of the electric field parallel to its length.

Effective length is used frequently to describe vertical antennas over a ground plane. For lengths l, which are small compared with a quarter wavelength, the effective length is $l/2$. This definition is useful at low frequencies for estimating currents appearing on the outer surface of cabinets and cables due to external fields, but the accurate estimation of field susceptibility levels requires careful analysis of the exact configuration, including neighboring metallic structures.

Receiving antennas can also be characterized by a gain G_r applicable to the far field. Gain is a reciprocal property of an antenna, i.e., the gain of an antenna used for receiving will be the same as when used for transmitting. At the higher frequencies where the structure size is comparable with, or more than, a wavelength, the power received P_r can be estimated in terms of the power transmitted, and the transmitter and receiver antenna gains G_t and G_r (both expressed as ratios), respectively, by

$$P_r = \frac{\lambda^2}{(4\pi r)^2} G_t G_r P_t \text{ , W} \qquad (3\text{-}147)$$

For isotropic antennas, the ratio of power received to power transmitted is, in free space,

$$\frac{P_r}{P_t} = \frac{\lambda^2}{(4\pi r)^2} \qquad (3\text{-}148)$$

A nomograph for this relation is given on Fig. 3-87 (Ref. 83), where the loss is given in dB. For antennas with specified gains, the loss can be reduced by the corresponding number of dB for each antenna. In terms of effective areas, A_t and A_r

$$\frac{P_r}{P_t} = \frac{A_t A_r}{\lambda^2 r^2} \qquad (3\text{-}149)$$

where

A_t = effective aperture of transmitting antenna, m^2

A_r = effective aperture of receiving antenna, m^2

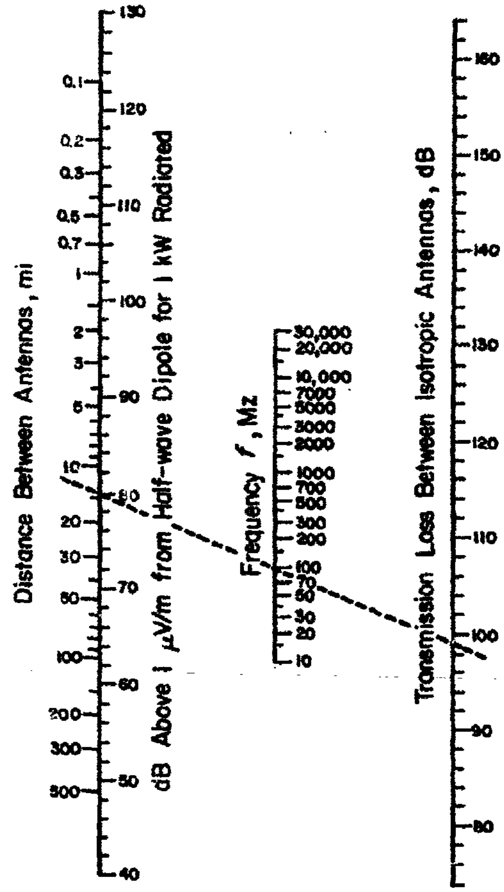


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Figure 3-87. Free Space Transmission (Ref. 83)

3-3.2.4.1 Propagation Effects (Refs. 83 and 115)

The concept of free space transmission assumes that the atmosphere is perfectly uniform and nonabsorbing, and that the earth is either infinitely far away, or its reflection coefficient is negligible. In practice, the effects of the earth, the atmosphere, and the ionosphere need to be considered. The treatment here is confined to phenomena of significance only over distances of tens of miles, so that ionospheric and tropospheric scatter effects are ignored.

3-3.2.4.2 Transmission Within Line of Sight

The presence of the ground modifies the generation and the propagation of radio waves so that the received power or field strength is ordinarily less than would be expected in free space, see Fig. 3-87. The effect of plane earth on the propagation of radio waves is given by

Direct Wave	Reflected Wave	"Surface Wave"	Induction Field and Secondary Effects of the Ground

$$\frac{E}{E_0} = 1 + Re^{j\Delta} + (1 - R)Ae^{j\Delta} + \ldots.. \quad (3\text{-}150)$$

where

E_0 = free space electric field, V/m
E = electric field due to a given source, V/m
R = reflection coefficient of the ground
A = "surface wave" attenuation factor, dimensionless
$\Delta = \dfrac{4\pi h_1 h_2}{\lambda r}$, dimensionless
$h_{1,2}$ = antenna heights measured in same units as the wavelength and distance
r = distance from source, m

The parameters R and A vary with both polarization, and the electrical constants of the ground. For near grazing paths, R is approximately equal to -1, and the factor A can be neglected provided both antennas are elevated more than a wavelength above the ground (or more than 5 to 10 wavelengths above sea water). Under these conditions, the effect of the earth is independent of polarization and ground constants, and Eq. 3-150 reduces to

$$\frac{|E|}{|E_0|} = \sqrt{\frac{P_r}{P_0}} = 2\sin\left(\frac{\Delta}{2}\right) = 2\sin\left(\frac{2\pi h_1 h_2}{\lambda r}\right) \quad (3\text{-}151)$$

where

P_0 = received power expected in free space, W

More precisely, the reflection coefficient $R = |R| \exp(j\varphi)$, assuming a plane earth has magnitude $|R|$ and phase shift φ which are typically of the form in Fig. 3-88 (Ref. 84). The curves apply to the case of a vertically polarized plane wave whose normal makes an angle ψ, called the grazing angle, with the reflecting earth's surface. The quantity n is the ratio or propagation constants of the region above the earth

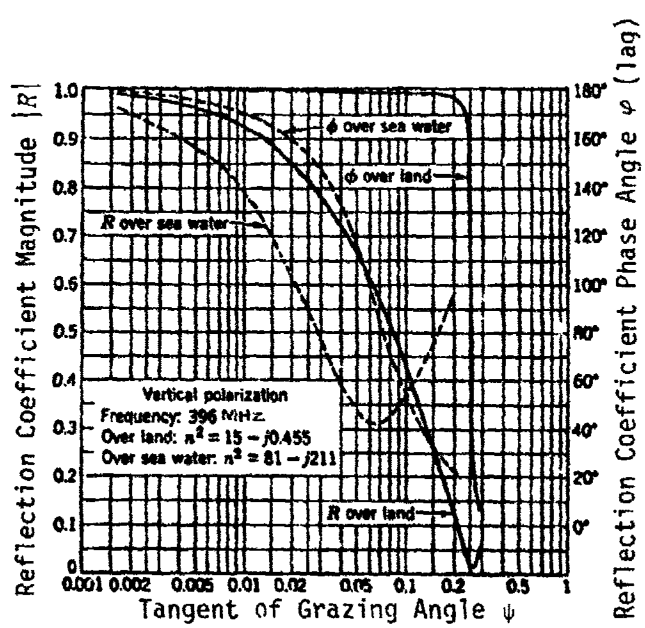

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Figure 3-88. Values of R and φ as Functions of $\tan\psi$ (Ref. 84)

(taken to be empty space) and the region below the earth's surface. This quantity is given by

$$n' = \epsilon_r - j\,18 \times 10^{-3}\,\sigma/f \quad \text{, dimensionless} \qquad (3\text{-}152)$$

where

ϵ_r = relative dielectric constant, dimensionless

σ = conductivity of the reflecting region, mho/m

f = frequency, Hz

The above-ground region is assumed to have $\epsilon_r = 1$ and $\sigma = 0$. The curves in Fig. 3-88 are plotted for typical values of permittivity and conductivity as found in sea water and wet earth and indeed show that for small grazing angles $|R| \doteq 1$ and $\varphi \doteq 180$ deg.

For horizontal polarization, the reflection coefficient magnitude tends to be closer to one than it is with vertical polarization and the phase shift at reflection is close to 180 deg for all conditions.

It is emphasized that the foregoing applies to unobstructed plane earth. Rough earth, violent seas, and obstructions complicate the interaction causing multiple reflections (scattering) and time variation. Refs. 85-88 deal with measurements and models of propagation in the presence of irregular terrain and structural obstacles in an urban environment. Inone and Akiyama (Ref. 89) report typical data of line-of-sight transmission over sea paths.

Eq. 3-151 is the sum of the direct and ground reflected rays. As a function of h_2, the height of the receiver above ground, the field pattern shows an oscillatory lobe structure. The first maximum occurs when the difference between the direct and ground reflected waves is a half-wavelength. In most applications (except air to ground) the principal interest is in the lower part of the first lobe; i.e., where $\Delta/2 < \pi/4$. In this case, $\sin(\Delta/2) \approx \Delta/2$ and the transmission loss ratio over plane earth is given by:

$$\frac{P_r}{P_t} = \left(\frac{h_1 h_2}{r^2}\right)^2 G_t T_r \qquad (3\text{-}153)$$

This relation is plotted in Fig. 3-89 (Ref. 83) for isotropic antennas. Fig. 3-89 is not valid when the indicated transmission loss is less than the free space loss shown in Fig. 3-87, because this means that Δ is too large for this approximation.

Although the transmission loss shown in Eq. 3-153 and in Fig. 3-89 has been derived from optical concepts that are not strictly valid for antenna heights less than a few wavelengths, approximate results can be obtained for lower heights by using h_1 (or h_2) as the larger of either the actual antenna height, or the minimum effective antenna height shown in Fig. 3-90 (Ref. 83). The error that can result from the use of this artifice does not exceed ± 3 dB, and occurs where the actual antenna height is approximately equal to the minimum effective antenna height.

Frequently the amount of clearance (or obstruction) is described in terms of Fresnel zones. All points from which a wave could be reflected with a path difference of one-half wavelength from the boundary of the first Fresnel zone; similarly, the boundary of the nth Fresnel zone consists of all points from which the path difference is $n/2$ wavelengths. The nth Fresnel zone clearance H_n at any distance r_1 is given by

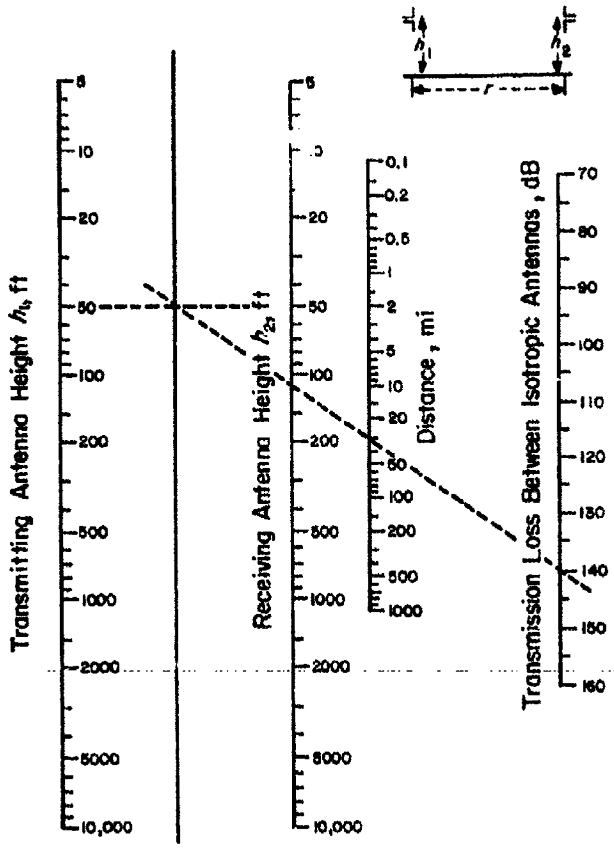
$$H_n = \sqrt{\frac{n\lambda r_1(r - r_1)}{r}} \text{, m} \qquad (3\text{-}154)$$

Although the reflection coefficient is very nearly equal to -1 for grazing angles over smooth surfaces, its magnitude may be less than unity when the terrain is rough. The classical Rayleigh criterion of roughness indicates that specular reflection occurs when the phase deviations are less than about $\pm(\pi/2)$, and that the reflection coefficient R occurs when the phase deviations are less than unity when the phase deviations are greater than $\pm(\pi/2)$. In most cases this theoretical boundary between specular and diffuse reflection occurs when the variations in terrain exceed 1/8 to 1/4 of the first Fresnel zone clearance. Experimental results with microwave transmission have shown that most practical paths are "rough", and ordinarily have a reflection coefficient in the range of 0.2 to 0.4. In addition, experience has shown that the reflection coefficient is a statistical problem and cannot be predicted accurately from the path profile (Ref. 90).

3-3.2.4.3 Miscellaneous Effects

This paragraph describes some miscellaneous effects of line of sight transmission that may be important at frequencies above about 1 GHz. These effects include variation in angles of arrival, maximum useful antenna gain, useful bandwidth, the use of frequency or space diversity, and atmospheric absorption.

On line-of-sight paths with adequate clearance, some components of the signal may arrive with variations in angle of arrival of as much as 0.5 deg to 0.75 deg in the vertical plane, but the variations in the horizontal plane are less than 0.1 deg (Refs. 91 and 92). Consequently, if antennas with beamwidths less than about 0.5 deg are used, there occasionally may be some loss in received signal because most of the incoming energy arrives outside the antenna beam-



Notes: 1. This chart is not valid when the indicated power is greater than the free space power shown on Fig. 3-87.
2. Use the actual antenna height or the minimum effective height shown on Fig. 3-90, whichever is the larger.

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Figure 3-89. Transmission Loss Over Plane Earth (Ref. 83)

width. Signal variations due to this effect usually are small compared with the multipath fading.

Multipath fading is selective fading, and it limits both the maximum useful bandwidth and the frequency separation needed for adequate frequency diversity. For 40-dB antennas on a 48-km path, the fading on frequencies separated by 100 to 200 MHz is essentially uncorrelated regardless of the absolute frequency. With less directive antennas, uncorrelated fading can occur at frequencies separated by less than 100 MHz (Refs. 93 and 94). Larger antennas (narrower beamwidths) will decrease the fast multipath fading and widen the frequency separation between uncorrelated fading, but at the risk of increasing the long term fading associated with the variations in the angle of arrival.

When ground reflections are controlling, optimum space diversity requires that the separation between antennas be sufficient to place one antenna on a field strength maximum while the other is in a field strength minimum. In practice, usually the best spacing is not known, because the principal fading is

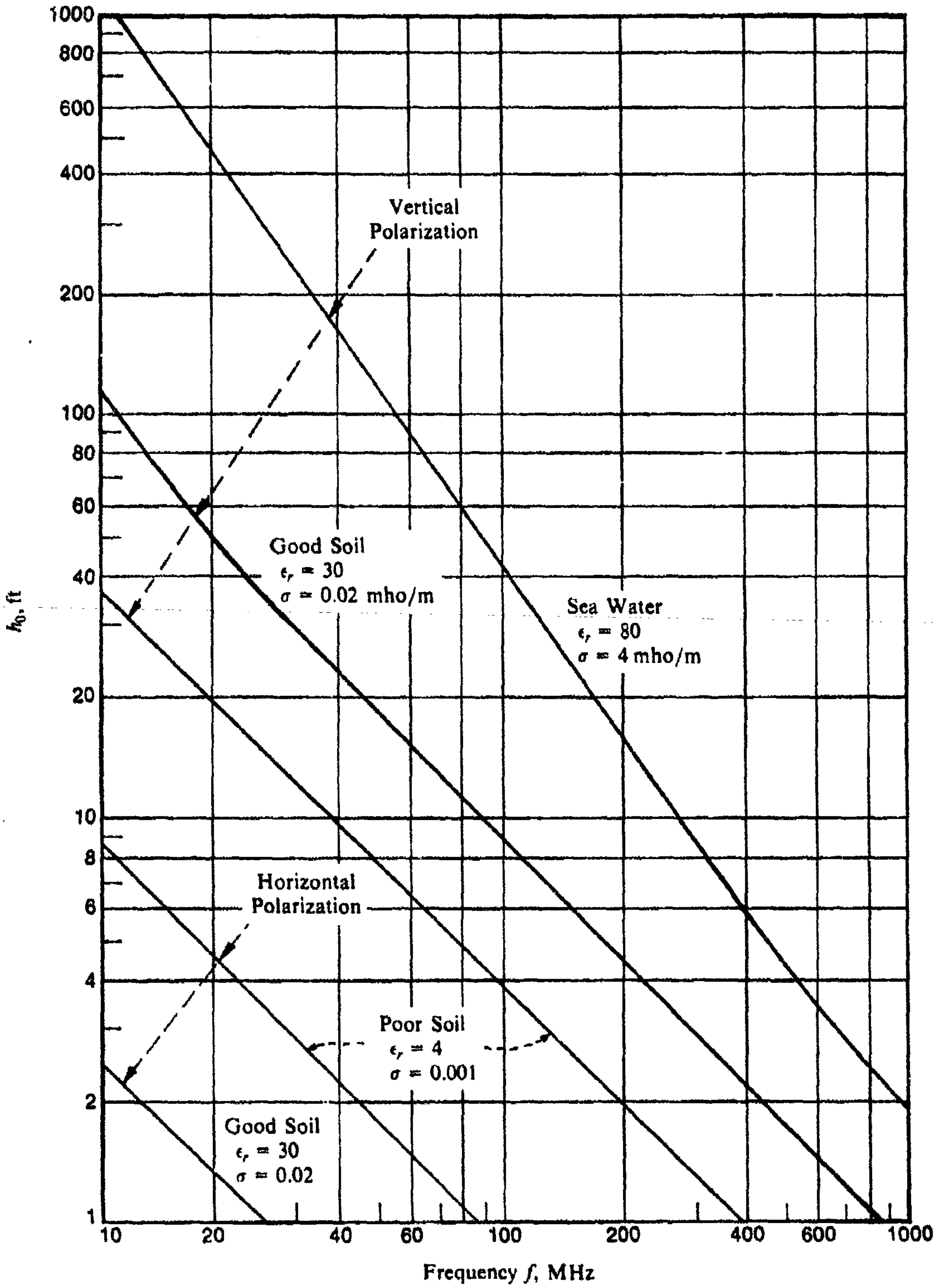

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Figure 3-90. Minimum Effective Antenna Height

caused by multipath variations in the atmosphere. However, adequate diversity usually can be achieved with a vertical separation of 100 to 200 wavelengths.

At frequencies above 5 to 10 GHz, the presence of rain, snow, or fog introduces an absorption in the atmosphere which depends on the amount of moisture and on the frequency. During a rain of cloudburst proportions, the attenuation at 10 GHz may reach 3.1 dB/km and at 25 GHz it may be in excess of 15.5 dB/km (Ref. 95). In addition to the effect of rainfall, some selective absorption may result from the oxygen and water vapor in the atmosphere. The first absorption peak due to water vapor occurs at about 24 GHz, and the first absorption peak for oxygen occurs at about 60 GHz.

3-3.2.5 Tropospheric Transmission Beyond Line of Sight

Though the scope of this handbook limits the region of concern around potential sources and receptors to the order of 10 miles, beyond line-of-sight transmission is of significance in this region. The distance to the horizon for an antenna 20 m off the ground is about 10 mi. Furthermore, the direct path between relatively close sources and receptors often is obscured by buildings, hills, and other obstacles.

Energy can be transmitted beyond the line of sight by three principal phenomena: reflection, refraction, and diffraction. The reflection mechanism of greatest significance for beyond-the-horizon-transmission is ionospheric reflection. This mechanism, which is effective at HF frequencies and below, is used mainly for long distance transmission. Reflections from extended conductive surfaces at close range are also important, but the mechanism is essentially the same as that covered in par. 3-3.2.4.2 (the ground reflected wave).

3-3.2.5.1 Refraction

The dielectric constant of the atmosphere normally decreases gradually with increasing altitude. The result is that the velocity of transmission increases with the height above the ground and, on the average, the electromagnetic energy is bent or refracted toward the earth. Provided the change in dielectric constant is linear with height, the net effect of refraction is the same as if the waves continued to travel in a straight line but over an earth whose modified radius k_a is

$$k_a = \frac{a}{1 + \dfrac{a\, d\epsilon_r}{2\, dh}}, \text{ m} \qquad (3\text{-}155)$$

where

a = true radius of earth, m

$\dfrac{d\epsilon_r}{dh}$ = rate of change of relative dielectric constant with height

Under certain atmospheric conditions the dielectric constant may increase ($0 < k_a/a < 1$) over a reasonable height, thereby causing the waves in this region to bend away from the earth. Since the earth's radius is about 6371 km, a decrease in dielectric constant of only 7.85×10^{-4} per km of height results in a value of $k_a = 4/3$ which is commonly assumed to be a good average value (Ref. 96). When the dielectric constant decreases about four times as rapidly, or by about 3.1×10^{-4} per km of height, the value of $k_a = \infty$. Under such a condition, as far as propagation is concerned, the earth then can be considered flat, since any ray that starts parallel to the earth will remain parallel.

When the dielectric constant decreases more rapidly than 3.1×10^{-4} per km of height, radio waves that are radiated parallel to, or at an angle above the earth's surface, may be bent downward sufficiently to be reflected from the earth. After reflection, the ray is again bent toward the earth, and the electromagnetic energy appears to be trapped in a duct or waveguide between the earth and the maximum height of the wave path. This phenomenon is variously known as trapping, duct transmission, anomalous propagation, or guided propagation (Refs. 97 and 98).

Duct transmission is important because it can cause long distance interference with another station operating on the same frequency; however, it does not occur often enough nor can its occurrence be predicted with enough accuracy to make it useful for communication services requiring high reliability.

An important mechanism of beyond-the-horizon transmission at VHF and above is that of tropospheric scatter (Refs. 99 and 100). The phenomenon described earlier in this paragraph is based on a smooth variation of refractive index with altitude. It generally is believed that there are irregular spatial and temporal variations of refractive index in the atmosphere which have the effect of multiple reflecting or refracting bodies turning back some of the impinging energy to earth. The mechanism is used for long-range transmission, usually exceeding 50 mi and extending to about 700 mi.

3-3.2.5.2 Diffraction Over a Smooth Spherical Earth and Ridges

As an object passes between a source and receiver so as to interrupt the direct path between them, a radio signal at the receiver does not immediately drop to zero because of the phenomenon of diffraction. The magnitude of the loss caused by the obstruction increases as either the distance of separation or the frequency is increased, and it depends to some extent on the antenna height (Ref. 101). The loss resulting from the curvature of the earth is indicated by Fig. 3-91 (Ref. 83) as long as neither antenna is higher than the limiting value shown at the top of the chart. This loss is in addition to the transmission loss over the plane earth obtained from Fig. 3-89.

When either antenna is as much as twice as high as the limiting value shown on Fig. 3-91, this method of correcting for the curvature of the earth indicates a loss that is too great by about 2 dB, with the error increasing as the antenna height increases. An alternate method of determining the effect of the earth's curvature is given by Fig. 3-92 (Ref. 83). This method is approximately correct for any antenna height, but it is theoretically limited in distance to points at or beyond line-of-sight, assuming that the curved earth is the only obstruction.

Fig. 3-92 gives the loss relative to free-space transmission (and hence is used with Fig. 3-87) as a function of three distances: r_1 is the distance to the horizon from the lower antenna, r_2 is the distance to the horizon from the higher antenna, and r_3 is the distance beyond the line-of-sight. In other words, the total distance between antennas is $r = r_1 + r_2 + r_3$. The distance to the horizon over smooth earth is given by

$$r_{1,2} = \sqrt{2k_a h_{1,2}}\ , \text{m} \qquad (3\text{-}156)$$

where

$h_{1,2}$ = appropriate antenna height, m

k_a = effective earth's radius, m

The preceding discussion assumes that the earth is a perfectly smooth sphere, since the results are critically dependent on a smooth surface and a uniform atmosphere. The modification in these results caused by the presence of hills, trees, and buildings is difficult or impossible to compute, but the order of

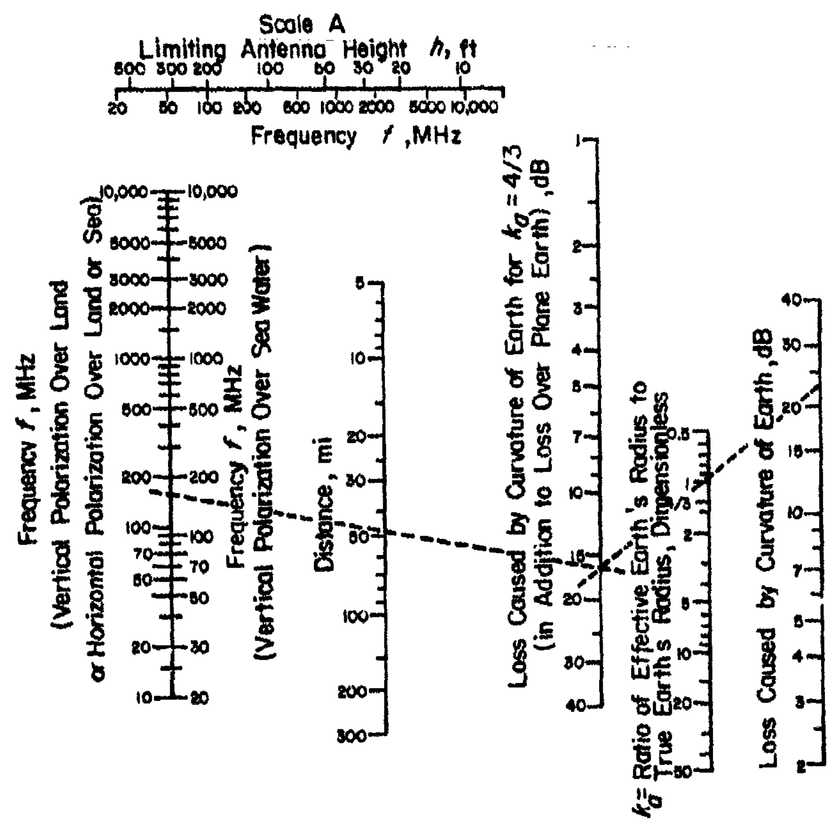

Figure 3-91. Diffraction Loss Around a Perfect Sphere (Ref. 83)

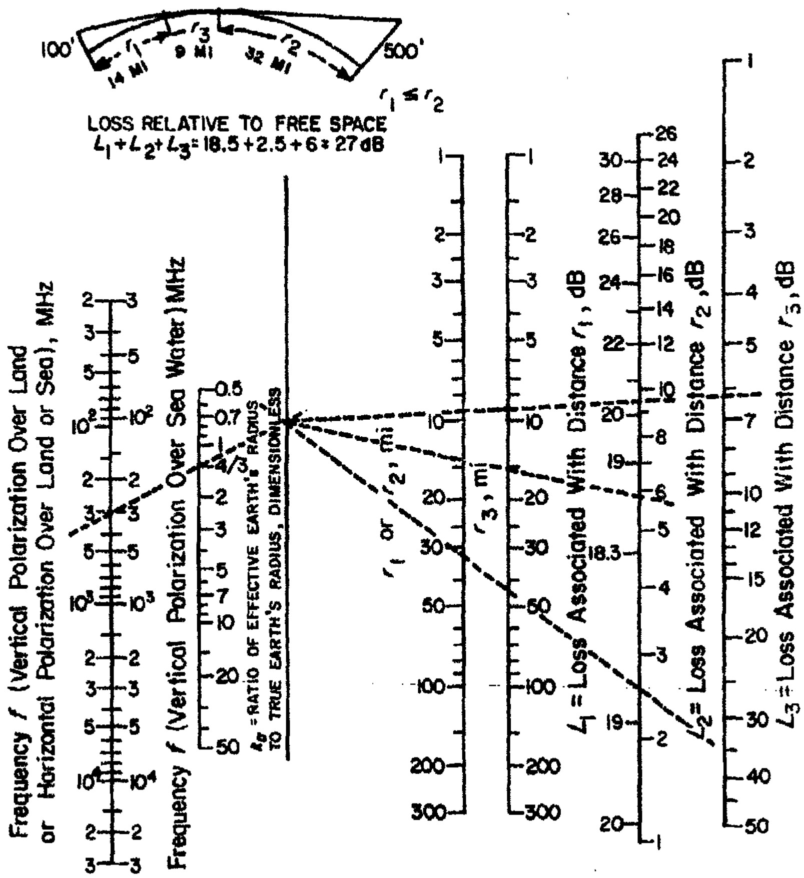

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Figure 3-92. Diffraction Loss Relative to Free Space Transmission at All Locations Beyond Line-of-sight Over a Smooth Sphere (Ref. 83)

magnitude of these effects may be obtained from a consideration of the other extreme case, which is a propagation over a perfectly absorbing knife edge.

The diffraction of plane waves over a knife edge or screen causes a shadow loss whose magnitude is shown on Fig. 3-93 (Ref. 83). The height of the obstruction *h* is measured from the line joining the two antennas to the top of the ridge. It will be noted that the shadow loss approaches 6 dB as *h* approaches 0 (grazing incidence), and that it increases with increasingly positive values of *h*. When the direct ray clears the obstruction, *h* is negative; and the shadow loss approaches 0 dB in an oscillatory manner as the clearance is increased. In other words, a substantial clearance is required over line-of-sight paths in order to obtain "free-space" transmission. The knife-edge diffraction calculation is substantially independent of polarization provided the distance from the edge is more than a few wavelengths.

At grazing incidence, the expected loss over a ridge is 6 dB (Fig. 3-93), while over a smooth spherical earth (Fig. 3-92) indicates a loss of about 20 dB. More accurate results in the vicinity of the horizon can be obtained by expressing radio transmission in terms of path clearance measured in Fresnel zones as shown in Fig. 3-94 (Ref. 115). In this representation the plane earth theory and the ridge diffraction can be represented by single lines, but the smooth sphere theory requires a family of curves with a parameter *M* that depends primarily on antenna heights and frequency. The big difference in the losses predicted by diffraction around a perfect sphere and by dif-

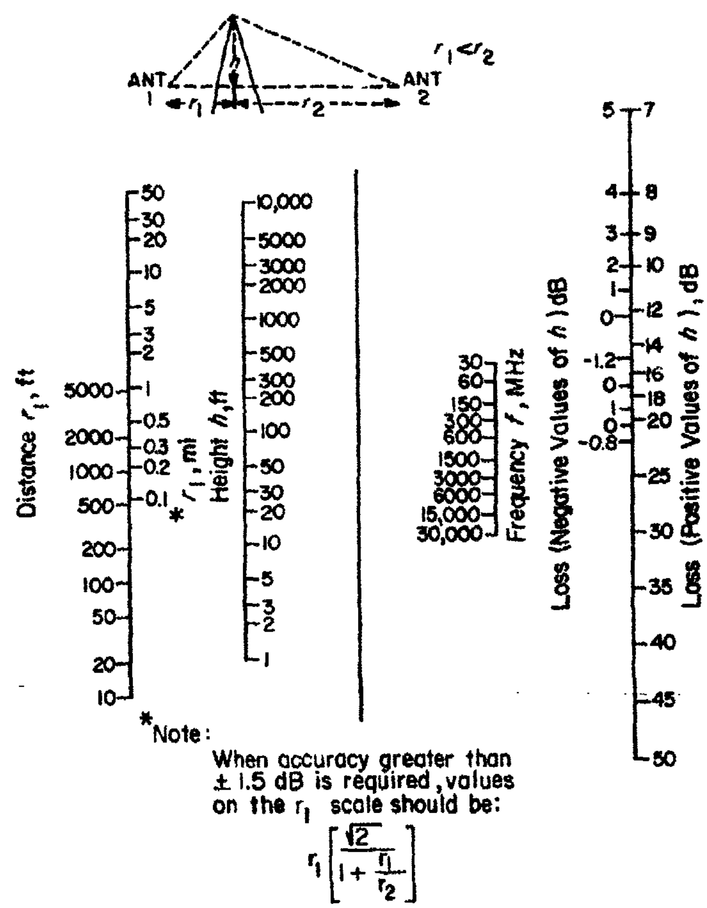

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Figure 3-93. Knife-edge Diffraction Loss Relative to Free Space (Ref. 83)

fraction over a knife-edge indicates that diffraction losses depend critically on the assumed type of profile. A suitable solution for the intermediate problem of diffraction over a rough earth has not yet been obtained.

3-3.2.5.3 Effects of Nearby Hills — Particularly on Short Paths

The experimental results on the effects of hills indicate that the shadow losses increase with the frequency and with the roughness of the terrain (Ref. 102).

A summary of the available empirical data is shown on Fig. 3-95 (Ref. 115). The roughness of the terrain is represented by the height h shown on the profile at the top of the chart. This height is the difference in elevation between the bottom of the valley and the elevation necessary to obtain line of sight from the transmitting antenna. The right hand scale of Fig. 3-95 indicates the additional loss above that expected over plane earth. Both the median loss and the difference between the median and the 10 percent values are shown. For example, with variations in terrain of 152 m (500 ft), the estimated median shadow loss at 450 MHz is about 20 dB, and the shadow loss exceeded in only 10 percent of the possible locations between points A and B is about 20 + 15 = 35 dB. This analysis is based on large-scale variations in field strength, and does not include the standing wave effects which sometimes cause the field strength to vary considerably within a few centimeters.

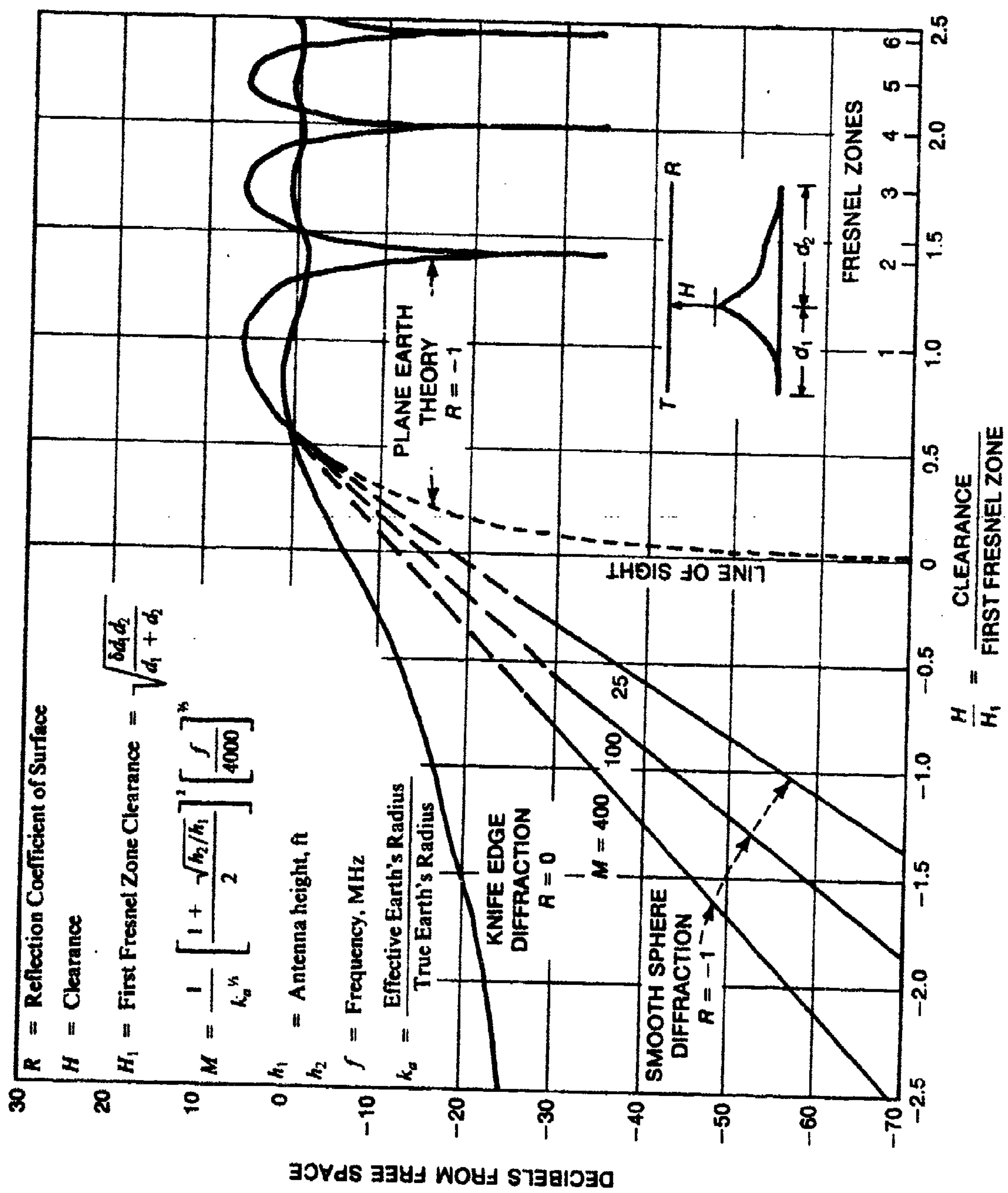

Figure 3-94. Transmission Loss vs Clearance (Ref. 115)

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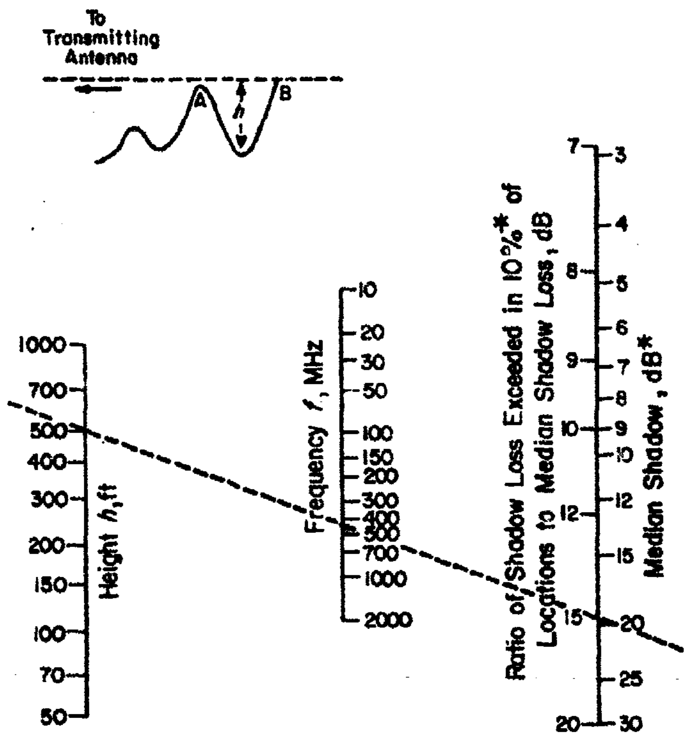

Figure 3-95. Field Intensity for Vertical Polarization Over Sea Water for 1-kW Radiated Power from a Grounded Whip Antenna as a Function of Roughness of Terrain (Ref. 115)

3-3.2.5.4 Effects of Buildings and Trees

The shadow losses resulting from buildings and trees follow somewhat different laws from those caused by hills. Buildings may be more transparent to radio waves than the solid earth, and ordinarily there is much more back scatter in urban areas than in open country. Both of these factors tend to reduce the shadow losses caused by the buildings but, on the other hand, the angles of diffraction over or around the buildings are usually greater than for natural terrain.

Typical values of attenuation through a brick wall are from 2 to 5 dB at 30 MHz and 10 to 40 dB at 3 GHz, depending on whether the wall is dry or wet. At frequencies in the UHF range, attenuation near windows can be quite negligible.

When an antenna is surrounded by moderately thick trees and is below tree-top level, the average loss at 30 MHz is usually 2 or 3 dB for vertical polarization, and negligible with horizontal polarization. However, large and rapid variations in the received field strength may exist within a small area, resulting from the standing-wave pattern set up by reflections from trees located at a distance of several wavelengths from the antenna. Consequently, several near-by locations should be investigated for best results. At 100 MHz the average loss from surrounding trees may be 5 to 10 dB for vertical polarization and 2 or 3 dB for horizontal polarization. The tree losses continue to increase as the frequency increases, and above 300 to 500 MHz they tend to be independent of the type of polarization. Above 1 GHz, trees that are thick enough to block vision are roughly equivalent to a solid obstruction of the same overall size.

3-3.2.6 Medium and Low Frequency Ground Wave Transmission

Wherever the antenna heights are small compared with the wavelength, the received field strength is ordinarily stronger with vertical polarization than with horizontal, and is stronger over sea water than over poor soil. In these cases the "surface wave" term in Eq. 3-150 cannot be neglected. This use of the term "surface wave" follows Norton's usage and is not equivalent to the Sommerfeld or Zenneck "surface waves".

The parameter A is the plane earth attenuation factor for antennas at ground level. It depends upon the frequency, ground constants, and type of polarization. It is never greater than unity, and decreases with increasing distance and frequency, as indicated by the following approximate equation (Ref. 83 and 103):

$$A \approx \frac{-1}{1 + j\dfrac{2\pi d}{\lambda}(\sin\theta + z)^2}, \text{ d'less} \qquad (3\text{-}157)$$

where

$$z = \frac{\sqrt{\epsilon_o - \cos^2\theta}}{\epsilon_o}, \text{ for vertical polarization}$$

$$z = \sqrt{\epsilon_o - \cos^2\theta}, \text{ for horizontal polarization}$$

$\epsilon_o = \epsilon - j60\sigma\lambda$
θ = angle between reflected ray and the ground,
= 0 for antennas at ground level, deg
ϵ = dielectric constant of the ground relative to unity in free space
σ = conductivity of the ground, mho/m
λ = wavelength, m
d = distance from source, m

In terms of these same parameters, the reflection coefficient R of the ground is given by (Ref. 104).

$$R = \frac{\sin\theta - z}{\sin\theta + z}, \text{ dimensionless} \qquad (3\text{-}158)$$

When $\sin\theta \ll |z|$, the reflection coefficient approaches -1; when $\sin\theta \gg |z|$ (which can happen only with vertical polarization), the reflection coefficient approaches $+1$. The angle for which the reflection coefficient is a minimum called the pseudo-Brewster angle, and it occurs for $\sin\theta = |z|$.

For antennas approaching ground level the first two terms in Eq. 3-150 cancel each other (h_1 and h_2 approach zero and R approaches -1) and the magnitude of the third term becomes

$$|(1 - R)A| \approx \frac{2}{\left(\dfrac{2\pi r}{\lambda}\right)z^2} = \frac{4\pi h_o^2}{\lambda r} \qquad (3\text{-}159)$$

where

h_o = minimum effective antenna height shown in Fig. 3-90
$= \left|\dfrac{\lambda}{2\pi z}\right|$
z = as defined for Eq. 3-157

The surface wave term arises because the earth is not a perfect reflector. Some energy is transmitted into the ground and sets up ground currents which are distorted relative to what would have been the case in an ideal perfectly reflecting surface. The surface wave is defined as the vertical electric field for vertical polarization, or the horizontal electric field for horizontal polarization which is associated with the extra components of the ground currents caused by lack of perfect reflection. Another component of the electric field associated with the ground currents is in the direction of propagation. It accounts for the success of the wave antenna at lower frequencies, but it is always smaller in magnitude than the surface wave as previously defined. The components of the electric vector in three mutually perpendicular coordinates are given by Norton (Ref. 105).

In addition to the effect of the earth on the propagation of radio waves, the presence of the ground also may affect the impedance of low antennas, and thereby may have an effect on the generation and reception of radio waves (Ref. 104). As the antenna height varies, the impedance oscillates around the free space value, but the variations in impedance are usually unimportant provided that the center of the antenna is more than a quarter-wavelength above the ground. For vertical grounded antennas (such as are used in standard AM broadcasting) the impedance is doubled and the net effect is that the maximum field strength is 3 dB above the free space value instead of 6 dB as indicated in Eq. 3-151 for elevated antennas.

Typical values of the field strength to be expected from a grounded quarter-wave vertical antenna are shown in Fig. 3-96 for transmission over poor soil, and in Fig. 3-97 for transmission over sea water. These charts include the effect of diffraction and average refraction around a smooth spherical earth as discussed in par. 3-3.2.5, but do not include the ionospheric effects. The increase in signal obtained by raising either antenna height is shown in Fig. 3-98 for poor soil, and in Fig. 3-99 for sea water.

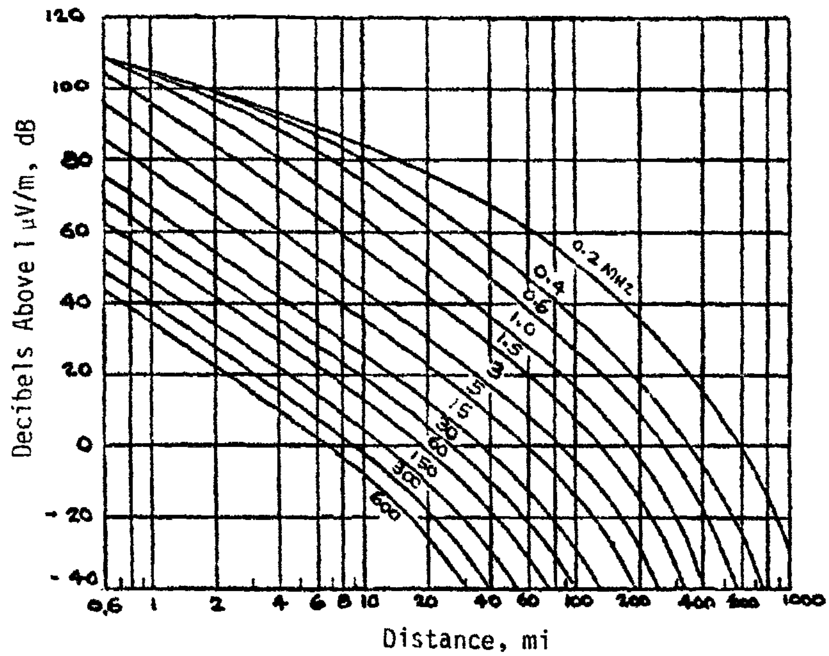

Figure 3-96. Field Intensity for Vertical Polarization Over Poor Soil for 1-kW Radiated Power from a Grounded Whip Antenna

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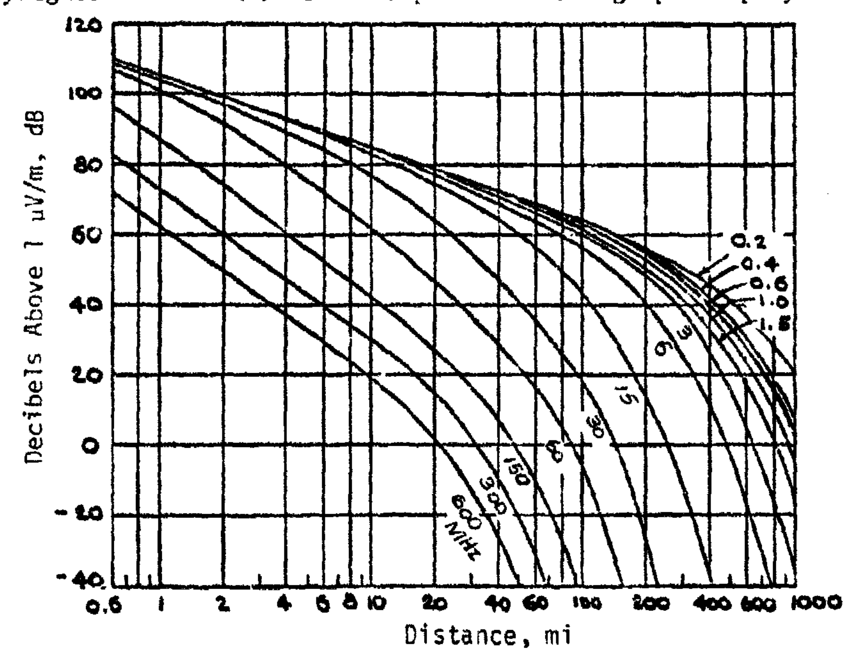


Figure 3-97. Field Intensity for Vertical Polarization Over Sea Water for 1-kW Radiated Power from a Grounded Whip Antenna

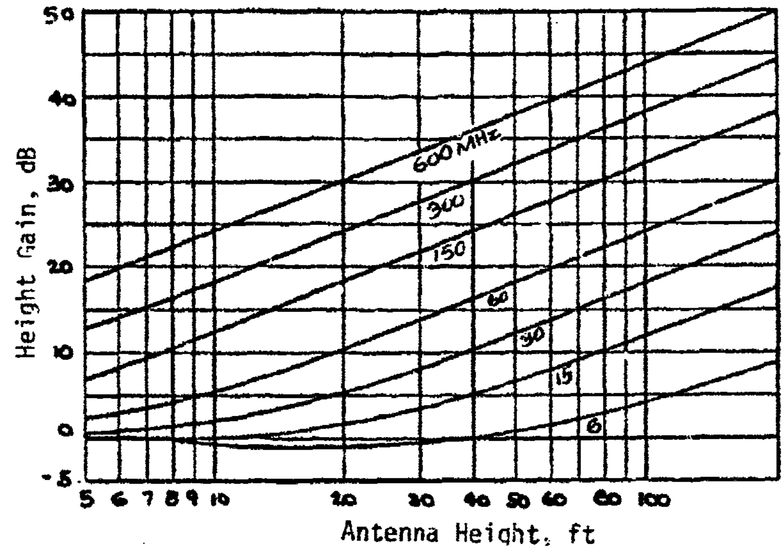


Figure 3-98. Antenna Height Gain Factor for Vertical Polarization Over Poor Soil

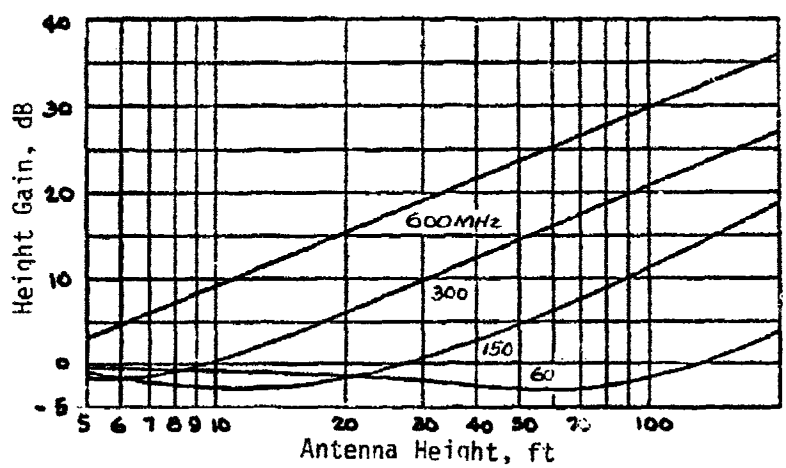


Figure 3-99. Antenna Height Gain Factor for Vertical Polarization Over Sea Water

3-3.3 CONDUCTIVE COUPLING

3-3.3.1 Introduction

In concept, direct or conductive coupling paths are readily modeled in terms of a simple coupled circuit as shown in Fig. 3-100 (Ref. 106). These coupling paths may exist in a variety of forms, especially in complex circuits; however, it is convenient to identify three types:

a. Common power supplies
b. Common ground return paths
c. Signal lines.

Signal line coupling can occur in two modes, known as differential and common modes. In the differential mode, the interference is unintentionally coupled into the signal line directly from the circuit to which it connects, and is propagated along the line in a manner similar to that of the desired signal. Common mode coupling, which is discussed in par. 3-3.3.3, can occur wherever circuits which are interconnected by a signal cable are at different reference potentials. This mode of coupling is quite similar to that classified as common ground return path coupling, which is discussed in par. 3-3.4.

Coupling through a common power supply is probably the most common form of conductive coupling. Usually it occurs between equipments through ac or dc main power lines, or between circuits in the same equipment through a common electronic rectifier dc power source. In the latter case, the model of Fig. 3-101 (Ref. 106) applies, and the coupling impedance is the output impedance of the supply. Methods of calculating the coupling and of reducing it through use of filters are discussed in par. 4-3.1.1.4.

Coupling through the main or primary ac or dc power line is the most difficult situation to analyze because of uncertainties as to the impedances of various items connected to the line, and its length. Furthermore the line actually consists of several conductors which may be connected in a variety of ways at the distribution transformer. For these reasons it can be treated properly only from a statistical approach.

3-3.3.2 Powerline Coupling

A general model of power line conducted interference is shown in Fig. 3-100. In this figure the interference source represents an equipment which generates undesired signals (interference) on the line; the power line system, a standard single-phase, three-wire, 60-Hz, power line with switching and protective circuits; and the susceptor, an equipment whose performance may be degraded by the interference from the source.

There are several features of this diagram that should be noted:

a. The square blocks represent cabinets. In effect, the presence of a local ground plane shown on the figure requires that to represent accurately the interference source and the susceptible device one needs a four-terminal equivalent circuit, one of the terminals being the cabinet itself. If there is no direct connection between the cabinet and the ground plane, as is frequently the case, it is necessary to recognize that there is in effect at least a capacitive coupling between them. This coupling can play a major role in the interference properties of either a source or a susceptor.

b. The power line system itself may or may not attenuate signals coupled through it, depending upon the frequency and the ways in which the source and susceptor are connected. Although connecting the susceptible device to the power line system can affect the impedance seen by the interference source, in individual cases, it is best considered as having little

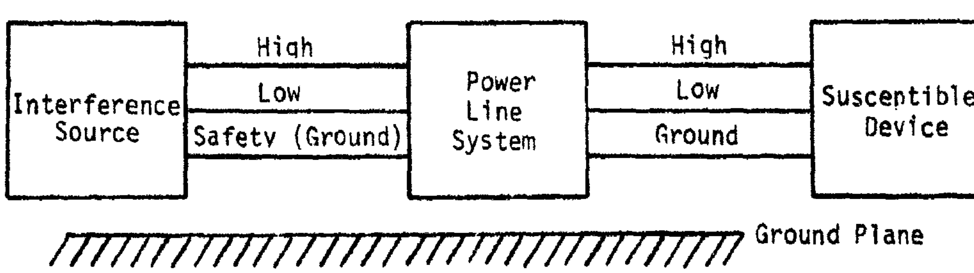


Figure 3-100. Power Line Conducted Interference Model (Ref. 106)

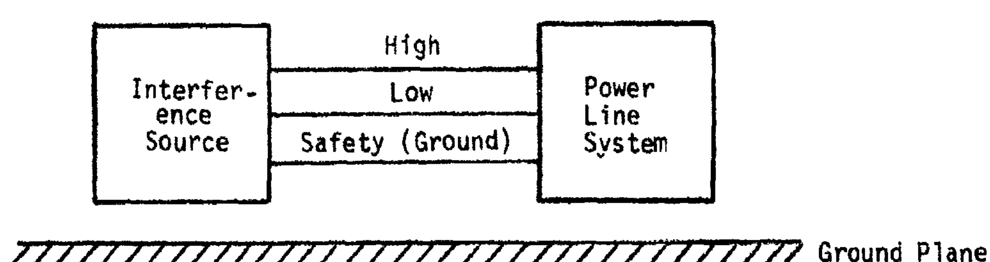


Figure 3-101. Model of Interference Source and Its Load (Ref. 106)

effect in a statistical sense. That is, the average impedance as seen by the interference source is assumed independent of the connection of the susceptor.*

c. For most circuits, at frequencies below about 10 Hz, the low and safety ground lines can be considered to be connected together, and indeed connected to the cabinet. At interference sources and susceptors, usually the safety ground is connected to the cabinets, effectively reducing them to three terminal devices. At a distribution panel where the low line is connected also to the neutral or safety line, a two-terminal representation may be adequate. On the other hand, at frequencies above about 500 Hz, the impedances as seen from any of the three lines — high, low and safety — when looking towards either the power system, the source, or the susceptor, will be substantially affected by the impedance of the individual conductors themselves and it is difficult to distinguish between these lines on the basis of impedance. An exact analysis at high frequencies is extremely difficult to carry out because of the complexity of the equivalent circuit.

In order to simplify the analysis of such circuits, it is common to approximate them in terms of their two-terminal characteristics. At low frequencies, each of the power conductors is considered to act independently against a neutral if one exists; otherwise, they are considered to act against each other. At frequencies above about 10 MHz, each of the line conductors can be considered to act against the cabinet in the case of a source or susceptor, or against the ground plane in the case of the power distribution system. At intermediate frequencies a transition region must be recognized in passing from one model to the other.

* This assumption may not be accurate in certain cases, for example, if the succeptor has a power line input filter containing capacitors which may dominate the line impedance in a specific frequency range.

3-3.3.2.1 Source and Line Models

For the purpose of analyzing the spurious voltage output impressed on the line by the interference source of Fig. 3-100, one considers only the source and the load presented to it by the power line system, as shown in Fig. 3-101.

3-3.3.2.1.1 Two-terminal Representation

The source and line are represented as shown in Fig. 3-102 (Ref. 106). E_S and Z_S are the Thevenin equivalent voltage and impedance of the source, respectively, at the frequency of interest, while Z_L is the line impedance at this frequency.

The power generator is not shown, but it is understood in this and subsequent discussions that the interference source is drawing its normal power current from the line.

Conducted emission measurement techniques presently in use provide the magnitude of the current I_M flowing from the source through a known load Z_M. Typical techniques are

a. Short-circuit current technique: $Z_M = 0$

b. Line impedance stabilization network technique (Refs. 107 to 109).

The ratio of the voltage appearing across the terminals of a susceptible device connected to a line of impedance Z_L, to the current through an impedance Z_M is

$$\left|\frac{V_L}{I_M}\right| = |Z_L| \cdot \left|\frac{Z_S + Z_M}{Z_S + Z_L}\right| \qquad (3\text{-}160)$$

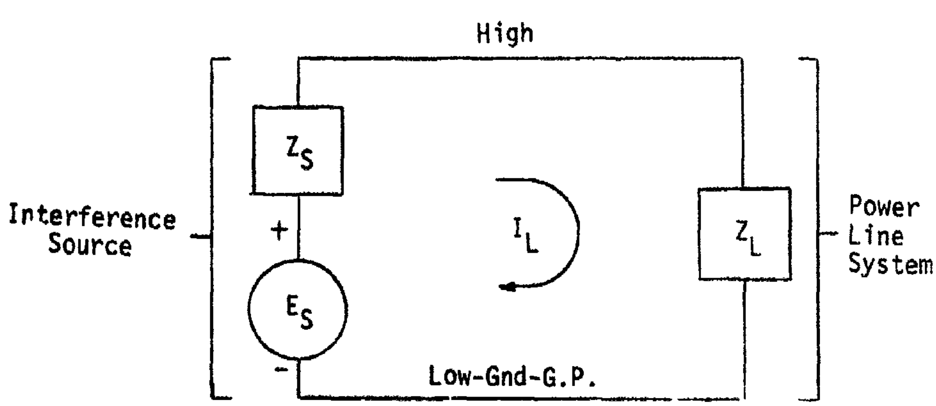


Figure 3-102. Two-terminal Representation

Thus if Z_M matches the line impedance, the voltage measured across Z_M is equal to the voltage that the susceptible device will experience.

Typical values of line impedance as a function of frequency are shown in Fig. 3-103 (Ref. 110). The separate curves are for different outlets in a laboratory where the measurements were made between the line conductor and earth (presumably the safety ground). This figure shows that the impedance at frequencies below about 200 kHz increases linearly with frequency as would be simulated by an inductance. Above about 500 kHz the impedance oscillated above and below a nominal value of about 50 Ω.

Other measurements have been made which show similar results. The solid curve on Fig. 3-103 corresponds to an inductance of about 30 μH. Evidence indicates that most power circuits have equivalent inductances ranging from about 20 μH to 50 μH, with the lower values for those with the higher current ratings and the higher values typical for the home and office or small laboratory.

3-3.3.2.1.2 Statistical Approach

Fig. 3-103 shows that, as a practical matter, it is not possible to know the exact impedance of the line to which a given source will be connected, especially since the impedance measured at any frequency and at any given time can be modified substantially by the connection of any additional device to the line in the vicinity of the point of measurement. Thus the prediction of the current or voltage produced by a given source can be determined only in a statistical sense. In setting a limit for the current measured from a given source, one selects a value which insures that when the source is connected to a potentially susceptible load (receiver), its susceptibility threshold is likely to be exceeded by more than a given probability, say, one or two percent. To do this one must have statistical data on the values of impedances likely to be experienced. Resistance and reactive components are discussed separately.

3-3.3.2.1.3 Resistance Distribution

A two-parameter, one-sided distribution which seems to be suitable for a wide variety of data is the gamma density, shown in Fig. 3-104 (Ref. 106), and given by

$$f(x) = \begin{cases} 0 & , \text{for } x < 0 \\ \dfrac{c^{b+1}}{\Gamma(b+1)} x^b e^{-cx} & , \text{for } x > 0 \end{cases} \tag{3-161}$$

where

$b > -1$

$c > 0$

$\Gamma(b+1)$ = gamma function

$$\Gamma(b+1) = \int_0^\infty y^b e^{-y} dy = b\,\Gamma(b) \tag{3-162}$$

The mean m_x and variance σ_x^2 of this distribution are given by

$$\langle x \rangle = m_x = \frac{b+1}{c} \tag{3-163}$$

$$\mathrm{Var}(x) = \sigma_x^2 = \frac{b+1}{c^2} \tag{3-164}$$

where

σ_x = standard deviation of x

Solving for b, and c, we find

$$b = \left(\frac{m_x}{\sigma_x}\right)^2 - 1 \tag{3-165}$$

$$c = \frac{m_x}{\sigma_x^2} \tag{3-166}$$

The function $f(x)$ may be fit to any set of resistance data by computing the data mean and standard deviation, and thence b and c from Eqs. 3-165 and 3-166, respectively.

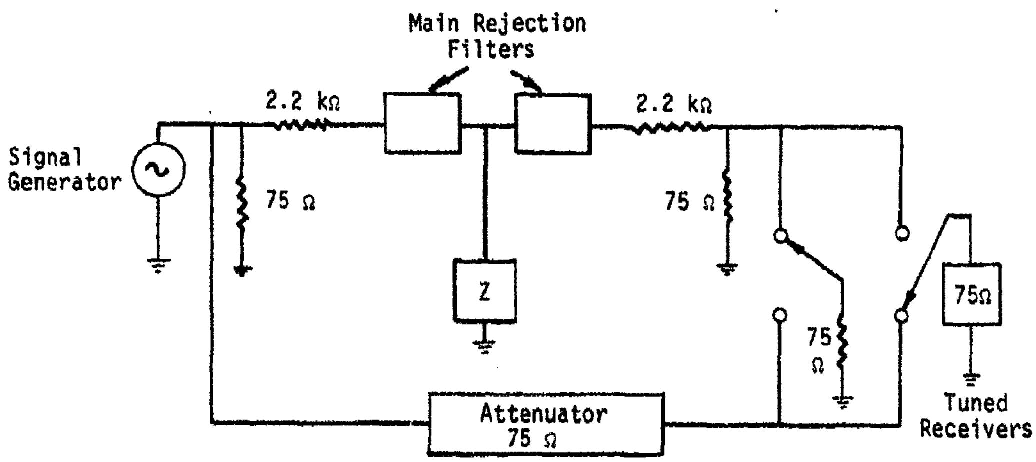
Typically, the resistance data are found to have a standard deviation greater than the mean, i.e., the data have a large spread, with most of the values concentrated at the low end. Therefore, the appropriate form of the assumed distribution is the curve in Fig. 3-104 for $b < 0$ or $m < \sigma_x$, viz., the one having a vertical asymptote at zero.

3-3.3.2.1.4 Reactance Distribution

Since reactance is equally likely to be positive or negative, at least at frequencies above about 1 MHz, a suitable distribution is one that is symmetrical about its mean value. The Gaussian density function, which has a good fit with experimental data, is shown in Fig. 3-105 (Ref. 106) and given by

$$f(x) = \frac{1}{\sigma_x \sqrt{2\pi}} \exp[-(x-m_x)^2/(2\sigma_x^2)] \tag{3-167}$$

Values of m_x and σ_x determined from the reactance data may be substituted directly into Eq. 3-167 to obtain the assumed distribution for reactance.



Circuit for Measuring Impedance of Mains

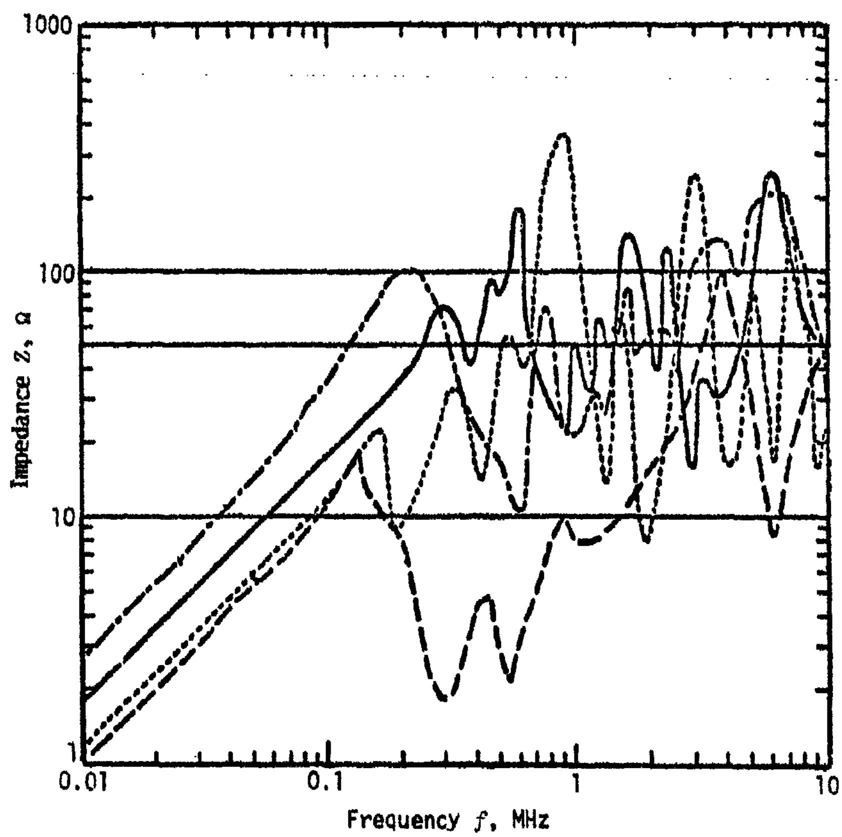


Figure 3-103. Impedance Between Live and Earth Connections at Ring Main Sockets in ERA Laboratory (Measurements made at four suitably disposed points in the building.)

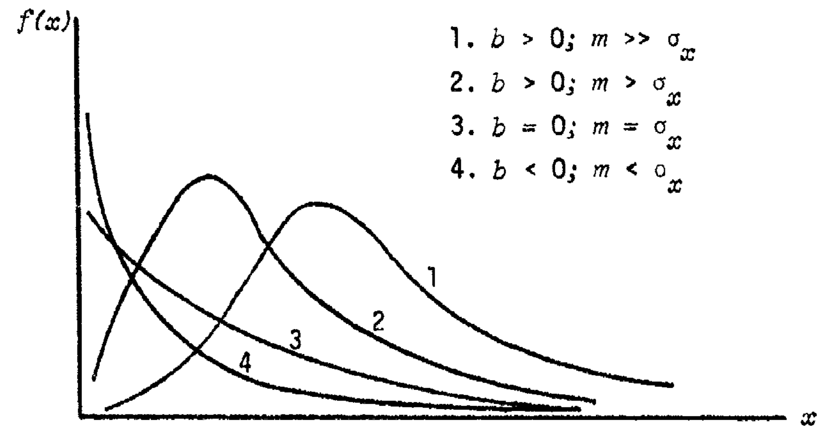


Figure 3-104. Gamma Probability Density for Several Parameter Values (Ref. 106)

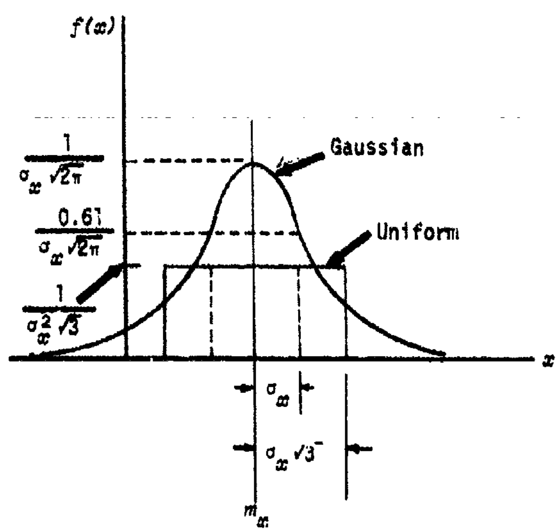


Figure 3-105. Comparison of Uniform and Gaussian Probability Density Functions With Same Mean and Standard Deviation (Ref. 106)

Typical means and standard deviations for impedance data in the frequency range 5-30 MHz are given in Table 3-10 (Ref. 106). These data were used to compute the statistics of $|V_L/I_M|$ given by Eq. 3-160 for the short-circuit (SC) measurement case, $Z_M = 0$. In this case, $|V_L/I_M|$ is simply the magnitude of the parallel combination of Z_L and Z_S

$$\left|\frac{V_L}{I_M}\right|_{SC} = \frac{|Z_L|\cdot|Z_S|}{|Z_L + Z_S|} \quad (3\text{-}168)$$

3-3.3.2.2 Typical Characteristics

3-3.3.2.2.1 *5-30 MHz*

Fig. 3-106 (Ref. 106) shows the distribution for the ratio defined by Eq. 3-168 for interference injected into the power line by various receivers when connected to the power line by power cords of both 20 cm length and 6 ft length. The 20 cm length represents the shortest possible cord length between the power duct and the receiver. The 6 ft length represents the length of a flexible cord connection. Pertinent data for this figure are:

Average V_L/I_M	66.7 Ω
Standard deviation	116.9 Ω
Minimum computed value	1.90 Ω
Maximum computed value	3970.7 Ω

3-3.3.2.2.2 *0.4-4.9 MHz*

Fig. 3-107 (Ref. 106) shows data quite similar to those for both high and low lines.

Separate calculations were made for the neutral or safety ground line in this frequency range because of its consistently low impedance values. Here the data lie to the left of those for the high and low lines. The lower impedance at low frequencies results from the diminishing inductive impedance as frequency is lowered.

3-3.3.2.2.3 *50 kHz to 200 kHz*

In this frequency range, one is almost always below the first maximum or "resonant" frequency of the impedance vs frequency curve. Two conclusions

TABLE 3-10. TYPICAL MEAN AND STANDARD DEVIATION OF IMPEDANCES, 5-50 MHz (Ref. 106)

	Resistance		Reactance	
	Mean m_x, ohm	Std. Dev. σ_x, ohm	Mean m_x, ohm	Std. Dev. σ_x, ohm
Source	25.58	44.44	− 0.95	106.60
Line	27.21	36.75	42.65	66.83

are obvious: (1) the spread in values of V_L/I_M on Fig. 3-108 is less than that for higher frequencies, and (2) the safety ground line curves lie considerably to the left of the high and low line curves.

From each of these figures, the probability that a given line voltage will be exceeded can be determined if a specified short circuit current is measured from a given source. For example, if the susceptibility of all equipments is required to be above a certain level — for example, 1 V — one can determine where the limit current should be set in order that the probability of interference does not exceed, for example, 2.5%. This statement implicitly assumes that all equipments become susceptible when the voltage on any of the conductors of the power line exceeds the specified limit. Actually, there will be a distribution of susceptibility voltages for groups of equipments on each of the respective lines, even though the lowest susceptibility on any of the three conductors may be at least 1 V. A more exact calculation of the probability of interference would have to take this distribution of susceptibility voltages into account.

3-3.3.3 The Common-mode Concept

The common-mode current is important for two reasons: (1) it usually appears directly in a ground return path which may be shared with several circuits and, (2) it can be a primary source of an induction magnetic field in the vicinity of cables on which it exists. The relationships between common-mode currents, differential-mode currents, and the actual line currents are explained by Fig. 3-109 (Ref. 106). This shows a susceptible device B connected to a power source A by a two-conductor cable, both devices located on a ground plane. For this system, the common-mode current is defined by $I_C = I_1 + I_2$, where the direction of the current must be taken into account and the differential-mode current, is defined by $I_D = (I_1 - I_2)/2$. The differential-mode components of the current are equal in magnitude on the two conductors and in opposite directions. In the absence of a ground or return circuit, the common-mode current must be zero. The definition of common mode current can be extended to cables containing more than two conductors, i.e., to be the net current (taking sign into account) carried by all conductors. If it is other than zero there must be a return path for the current — usually through a ground return conductor or "ground plane".

The configuration of the ground return path may have a significant effect on the amount of common-mode current that can flow in a given connection. At frequencies below several hundred kHz where capacitive reactances between the cabinet of a device and the "ground" plane may be high, the absence of a direct connection to the ground plane can reduce the common-mode current to quite small values. For significant common-mode current, both the source and the susceptor must be coupled through fairly large capacitors such as with line to ground filters. At higher frequencies, the capacitive reactance becomes low — and indeed since the power line itself can be represented as an inductance to common-mode currents, it is possible that at a given frequency series or parallel resonance will occur — and the impedance to common-mode current flow will vary rapidly with frequency.

Experimental data demonstrate these effects. The common-mode impedance of the power line of a receiver connected by a large ground strap to a copper-covered test bench is shown in Fig. 3-110. These measurements were made by using a two-current-probe method. In the frequency range between about 800 kHz and 25 MHz, the impedance is inductive, increases with frequency approximately linearly, and has a relatively constant phase angle near 90 deg. Fig. 3-111 shows a similar impedance measurement in which the large ground strap was removed and effectively replaced by an antenna connection. This antenna connection was a coaxial cable connected between the receiver and a connector, the outer shell of which was conductively connected directly to the ground plane. The total length of this

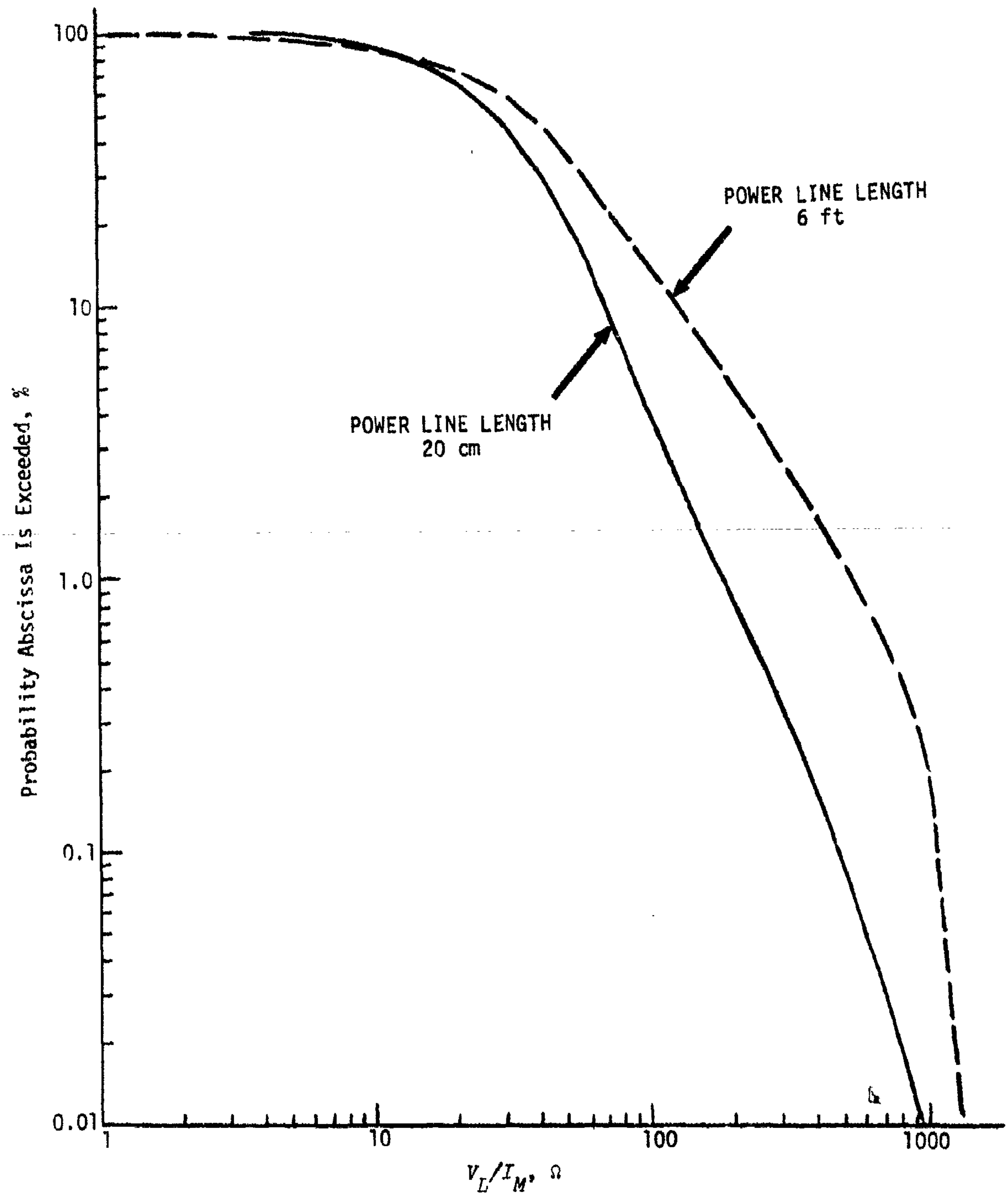


Figure 3-106. Distribution of V_L/I_M, 5- to 30-MHz Typical Power Lines (Ref. 106)

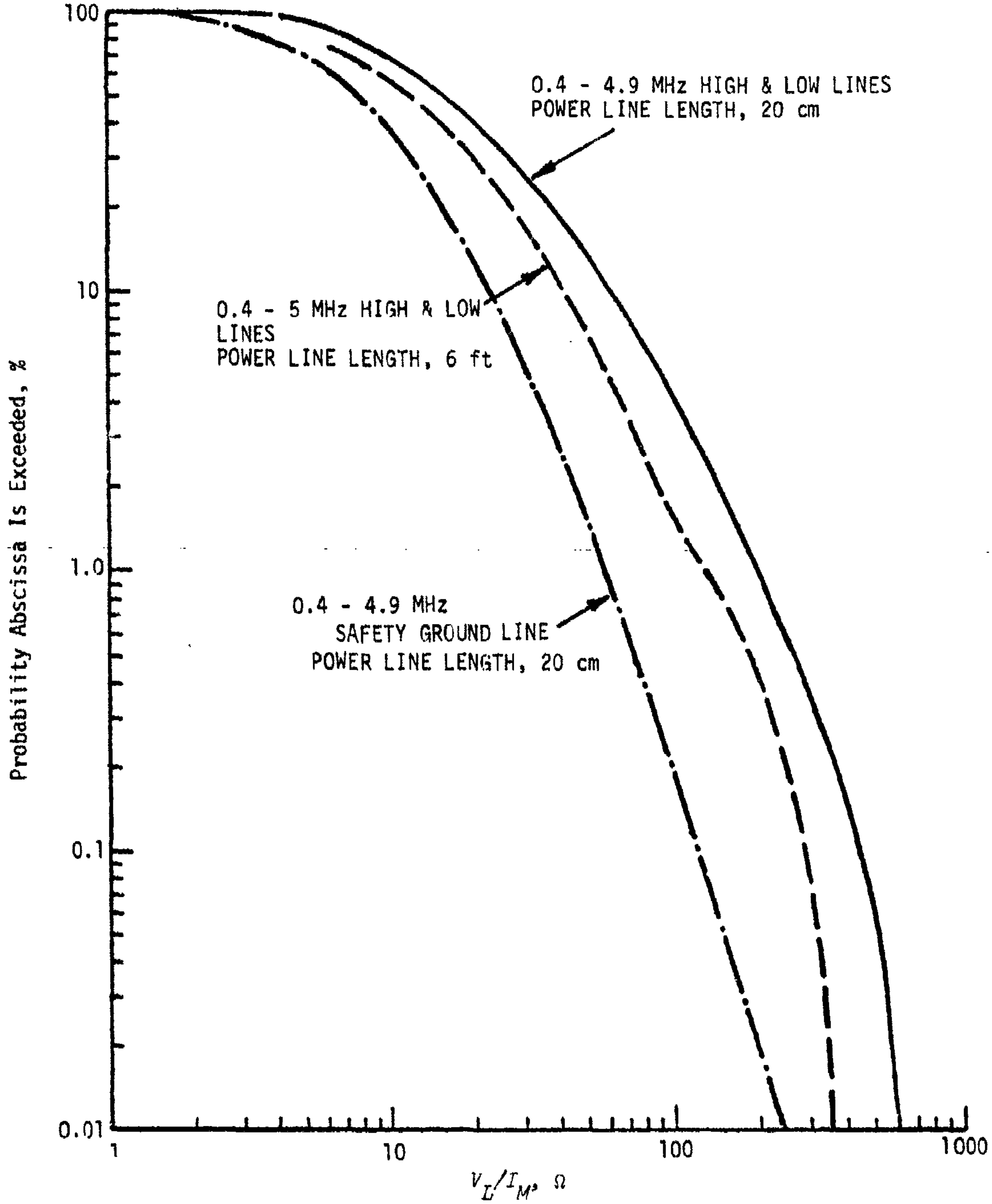


Figure 3-107. Distribution of V_L/I_M, Typical Power Line (Ref. 106)

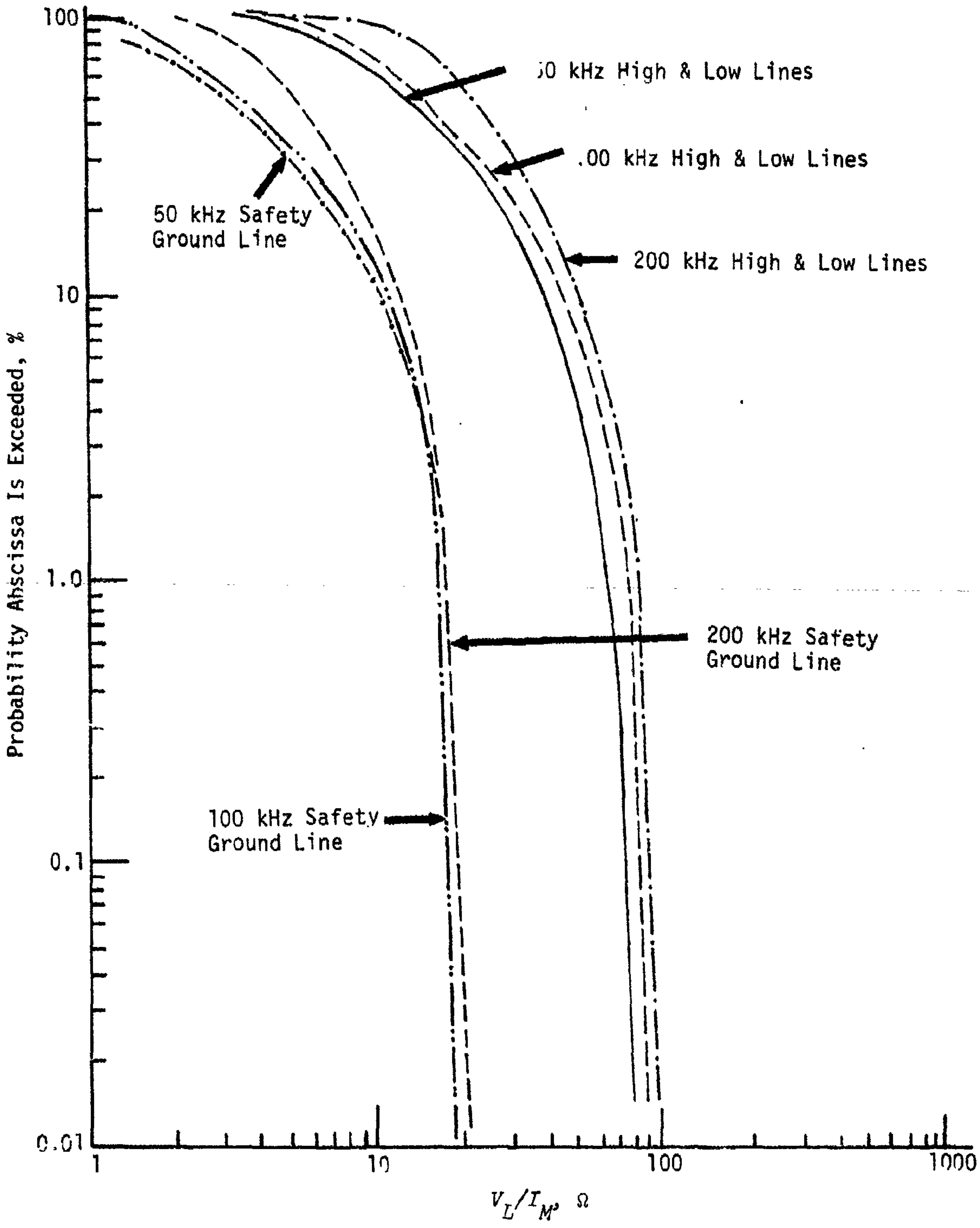
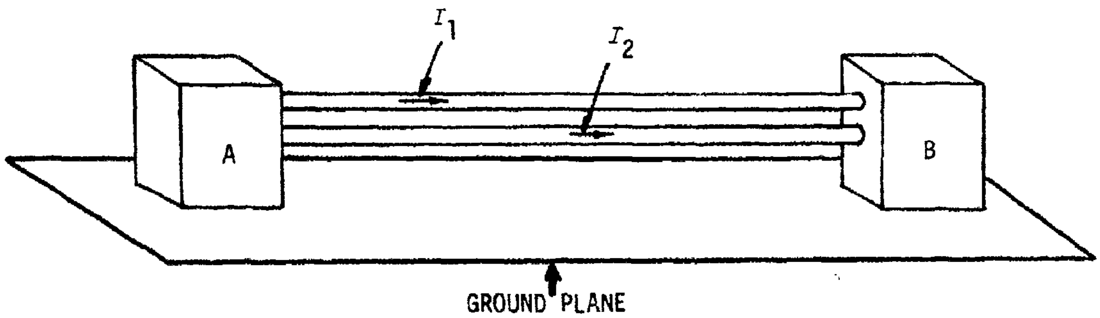


Figure 3-108. Distribution of V_L/I_M 50-, 100-, and 200-kHz Typical Power Lines



Common-mode Current: $I_C = I_1 + I_2$

Differential-mode Current: $I_D = \dfrac{I_1 - I_2}{2}$

Figure 3-109. Common-mode Model (Ref. 106)

lead was about 3 ft. In this frequency range the impedance is generally higher in Fig. 3-110 than in Fig. 3-111, but passes through a resonance at approximately 15.5 MHz. The phase angle of the impedance shown in Fig. 3-112 shifts from 90 deg positive to about 90 deg negative at the resonant frequency. The resonant impedance is of the order 5000 Ω.

Figs. 3-113 and 3-114 show common-mode impedances for another receiver in the grounded and ungrounded conditions. With the ground (Fig. 3-113) the reactance is inductive and increases linearly with frequency. When ungrounded, the impedance is capacitive at low frequencies and passes through resonance at about 11 MHz.

3-3.4 GROUNDING

3-3.4.1 General

The name "ground" originates from the use of the term in connection with the general practice of providing a path for the return of lightning discharge current which could otherwise produce potentials hazardous to both personnel and equipment. Its use has been extended to include return paths and reference potentials for power and signal currents, whether or not an actual connection to earth exists. The basic principle is to maintain all portions of the system — electrical, mechanical, or structural — at the same reference potential by providing low-impedance paths at all frequencies for current returns throughout the system. In some systems, reference potential variations on the order of volts may be tolerated. In others, a few microvolts of variation may be quite intolerable.

In designing for electromagnetic compatibility, the objective is to prevent any electromagnetic field, voltage, or current generated or used at one point of the system from being transferred through a common ground impedance to other units with whose operation it could interfere. Sometimes this can be achieved best by segregating grounding connections among the various parts of the system. For example, there may be a signal ground, a control ground, a power ground, and a safety ground.

3-3.4.2 Static and Structural Grounds

Static and structural grounds include all those conductive parts of the system which are not designed especially to carry current. These parts may be mechanical strength members and mechanical parts and enclosures that act as static shields, cable shields, shafts of rotating machinery, or control shafts.

They may serve a twofold purpose: (1) to prevent the build up of a static charge on structural members, and (2) to prevent electric fields from penetrating into areas where they might degrade equipment performance. For the purpose of preventing static charge from building up on structural members, all that is required is a continuous connection to a ground reference plane which itself usually is connected to earth. This arrangement is used in lightning protection systems where the purpose is to return any discharges to earth in such a way as to avoid high potentials on any equipment since these may be hazardous to personnel or the equipment itself. Methods of obtaining good connections to earth are discussed in Refs. 16 and 111. The same techniques are

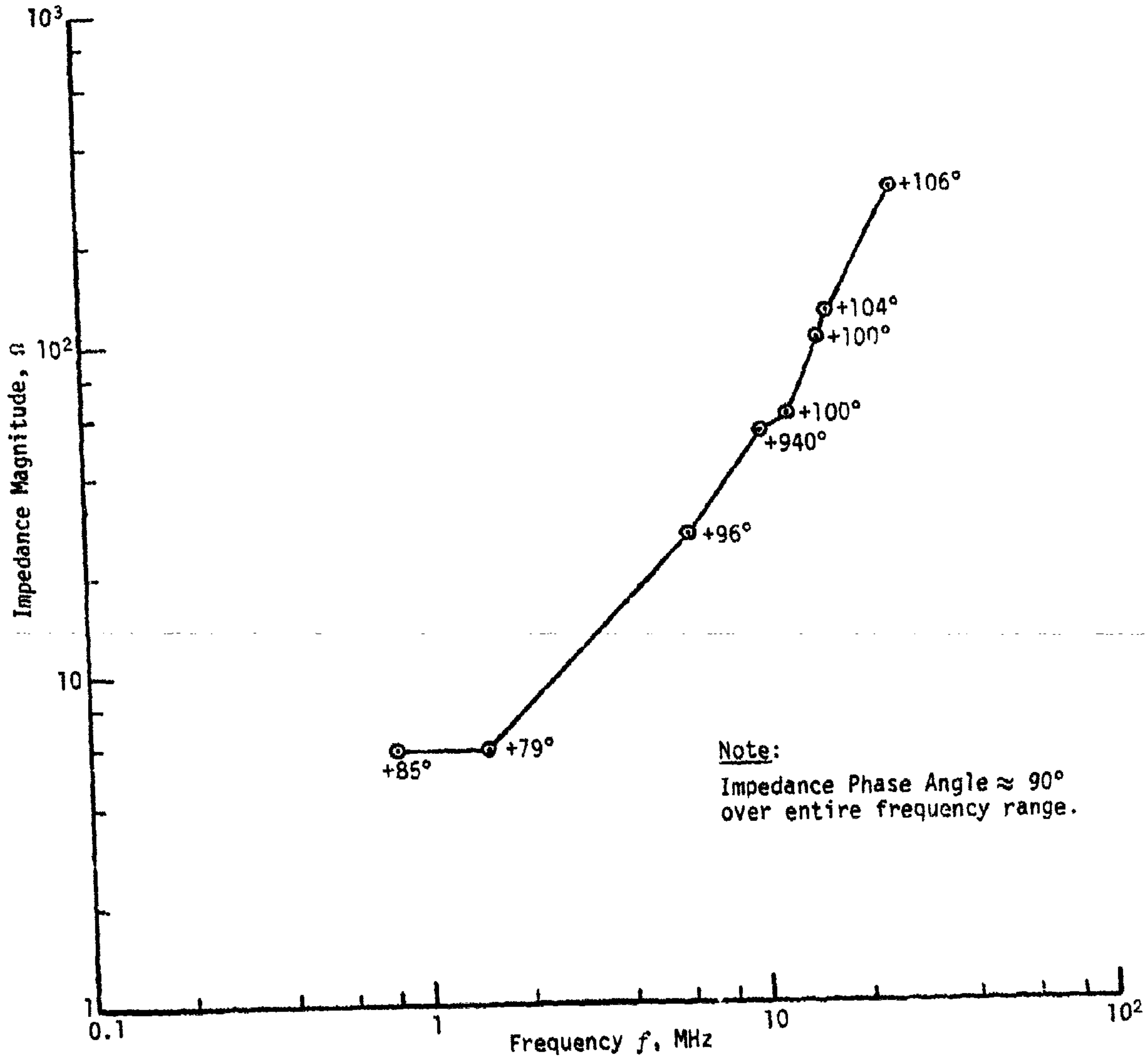


Figure 3-110. Common-mode Impedance of Receiver Bonded to Ground Plane

used where equipments or systems are exposed to high level radio transmitter fields which otherwise could subject the equipment to high radio frequency potentials. For the second purpose, a complete enclosure of the circuit to be protected is required, and the structure becomes a "shield". The properties of shields are discussed in par. 4-6.

3-3.4.3 Power System Ground

In the U.S., the National Electrical Code requires that the prime power system neutral be grounded at the service entrance equipment, and at no other point.

Equipment that is powered by such a system should have, in addition to the necessary high and low lines, a separate conductor connected to power system neutral. This conductor should not normally carry current. Its presence is necessary to protect personnel and equipment in the event of a short circuit to the equipment case. The grounding conductor and associated bonds must have cross-sectional areas sufficiently large to carry the required current.

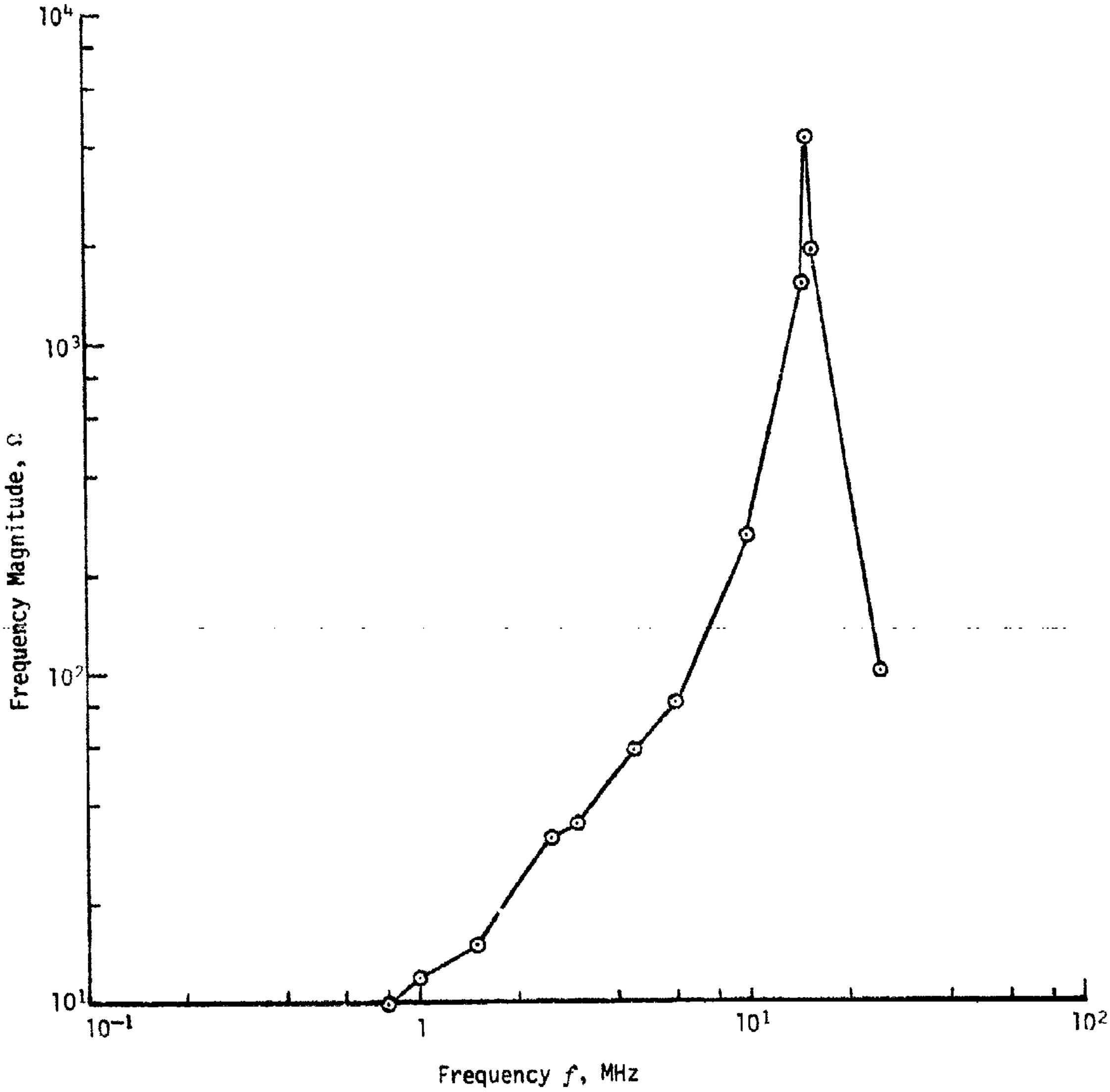


Figure 3-111. Common-mode Impedance of Receiver Grounded by Means of 3-ft Coaxial Antenna Cables

3-3.4.4 Ground Planes

Since the performance of electrical systems generally is dependent on a number of related and interacting functions, acting in a precise way, their interactions must be controlled and carefully limited to those designed into it. Unintentional interactions between various circuits can degrade the design performance seriously. The use of common potential planes or ground planes is frequently necessary for two possible reasons:

a. *Convenience*. Circuit construction can be simplified by returning circuit elements to the nearest appropriate point on the chassis.

b. *Circuit efficiency*. At the higher frequencies the lengths of leads must be short in order to keep the self-inductance of the leads to acceptable values.

An ideal ground plane is an equipotential surface having zero impedance. Practical ground planes may be conductive metal surfaces or a network of wires. The dimensions of the ground plane can be critical if

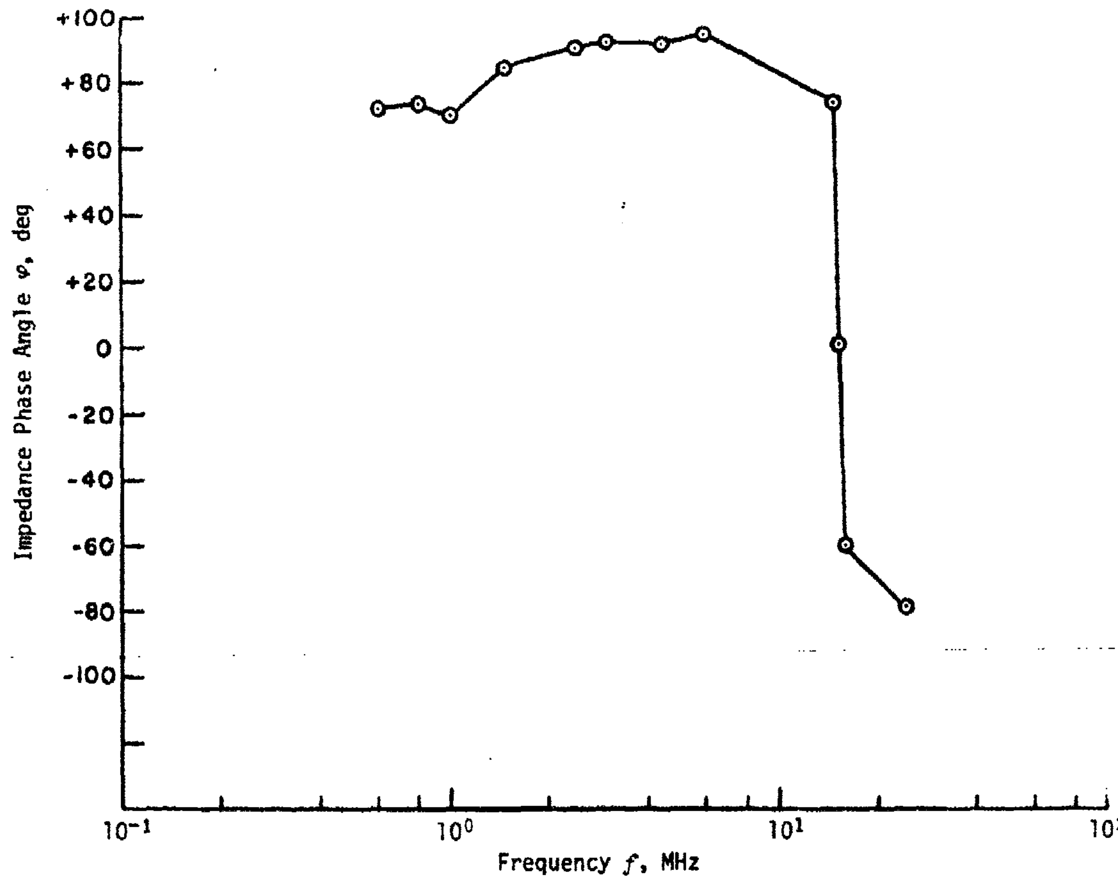


Figure 3-112. Phase Angle of Common-mode Impedance of Receiver Connected as for Fig. 3-111

they are comparable with system wavelengths. In many cases, a specific ground plane may not be easy to identify.

Circuit arrangements with respect to a ground plane may be of several types (Ref. 111): (1) a floating ground system as shown in Fig. 3-115, (2) a single-point ground system as is shown in Fig. 3-116, and (3) a multi-point ground system as is shown in Fig. 3-117.

3-3.4.4.1 Floating Ground System

The floating ground is most feasible at low frequencies where the circuits can indeed be isolated, and any stray capacitance between the circuits and the ground plane is sufficiently low to limit circulating currents. The inductive coupling shown can inhibit such currents. To the extent that such currents can be controlled, the system itself is not subject to the quality of the ground plane.

3-3.4.4.2 Single-point System

This system is used where a ground is essential, such as to provide a power supply return where the circuits operate at frequencies such that the lengths of the several connections to the single tie point are small fractions of a wavelength. This arrangement also tends to make the system performance independent of currents flowing in the ground plane.

3-3.4.4.3 Multipoint System

Where the various parts of a system are physically separated by distances which are a substantial fraction of a wave-length (say $\lambda/10$)* of signals used by the system or which can couple into it, a multipoint grounding system must be used. In this case a continuous conducting sheet or plane will provide the lowest possible impedance path between components.

* In critical cases $\lambda/50$ has been recommended.

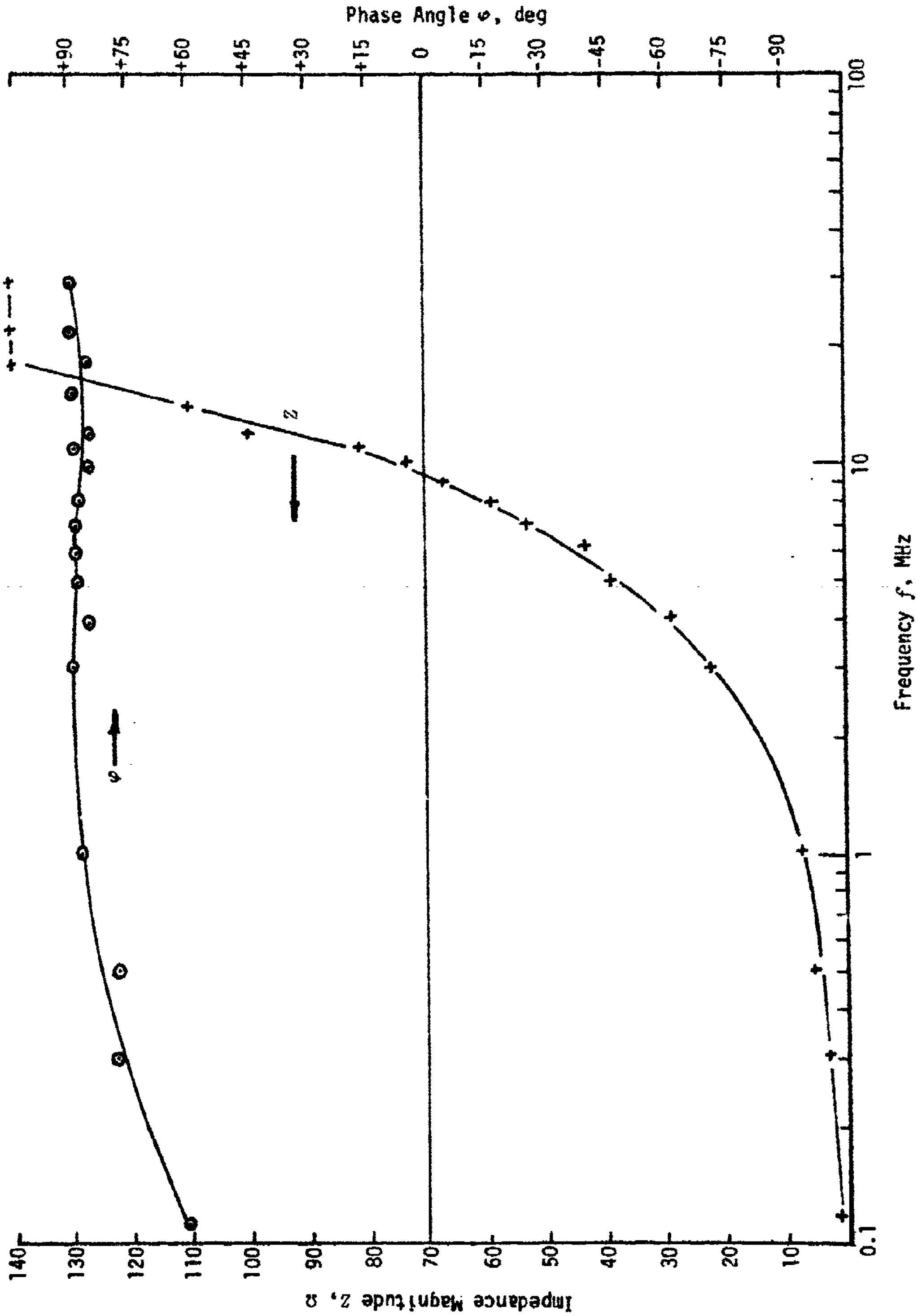


Figure 3-113. Impedance Magnitude and Phase vs Frequency; BC-348, Common Mode Grounded Shielded Enclosure Test Bench

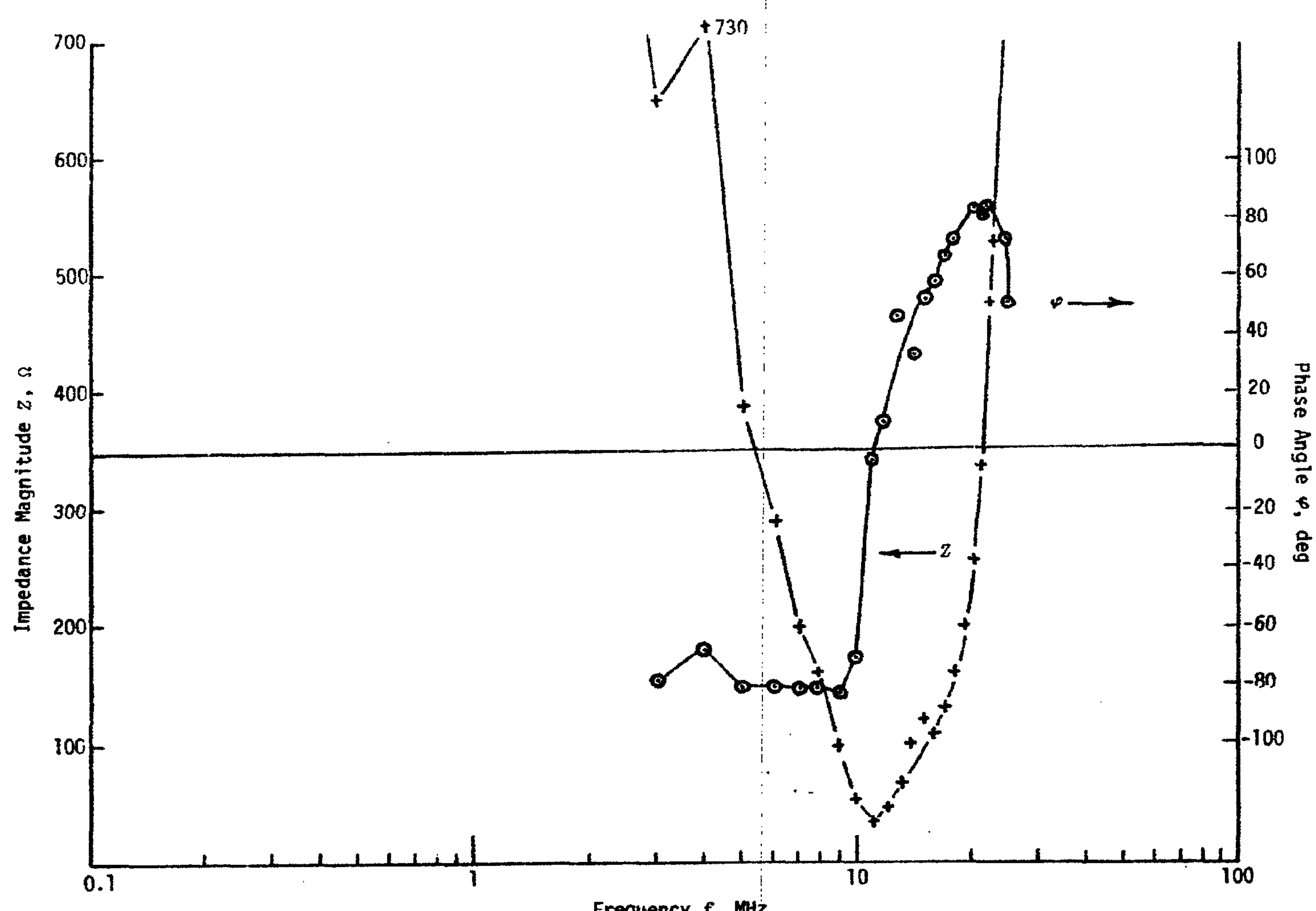


Figure 3-114. Impedance Magnitude and Phase vs Frequency; Common-mode — Receiver Ungrounded

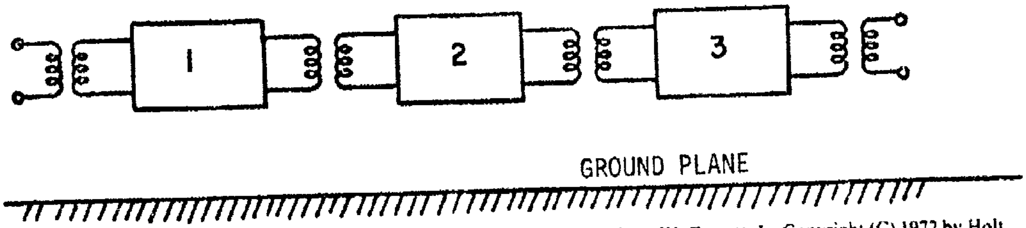


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Figure 3-115. Floating Ground Plane

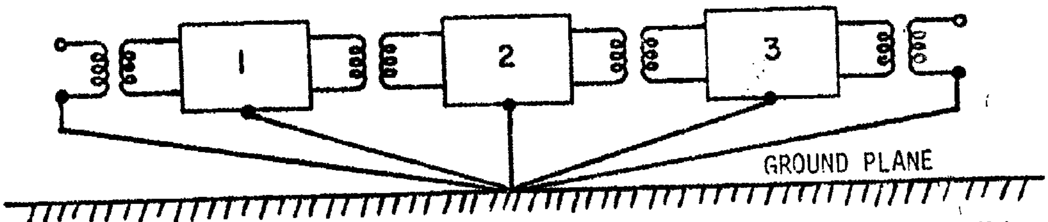

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Figure 3-116. Single-point Ground System

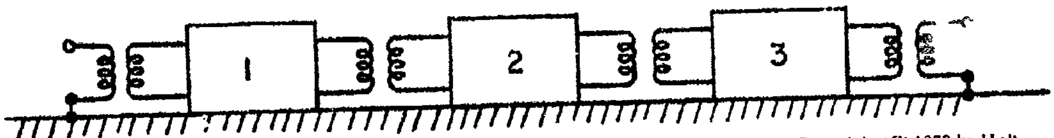

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Figure 3-117. Multipoint Ground System

To increase the effectiveness of multipoint grounds, each stage or unit of the system should be wired with a single point ground as shown in Fig. 3-118 (Ref. 111). Thus the two stages have their own single-point grounds. High currents associated with the output stage circulate only in this stage, and do not couple into the low level input.

In some multipoint ground systems, it may be advisable to cut holes or slots in the ground plane to act as baffles and thus channel ground currents to proper points.

Where the circuits are separated by long distances so that cables are required for interconnection, several factors must be considered. The length of the cable may be such as to be resonant. If so, the cable can act as an efficient radiator of energy; likewise, it can receive energy radiated from some other source. The possibility of such radiation can be reduced substantially by connecting the cable to the ground plane at a large number of points.

3-3.4.4.4 Balanced Coupling Circuits

Fig. 3-119 (Ref. 111) shows the arrangement of a balanced coupling circuit. Current flowing in the loop formed by the signal cable and the ground will induce voltages across the twisted pair signal leads. If these leads are perfectly balanced, the voltages (also known as "common" mode voltages) on these leads separately will produce equal and opposite effects in the receiver; hence they tend to cancel. The circuit is then said to have a high common-mode rejection ratio. Current in the shield of the cable can be minimized by connecting only one end of the shield to the ground as shown, but then a substantial potential may appear between the end of the shield and the receiver input leads. The balance of the circuit should also help to reduce the effects of this, however. In particularly sensitive circuits, the balance may be adjustable to obtain a maximum common-mode rejection.

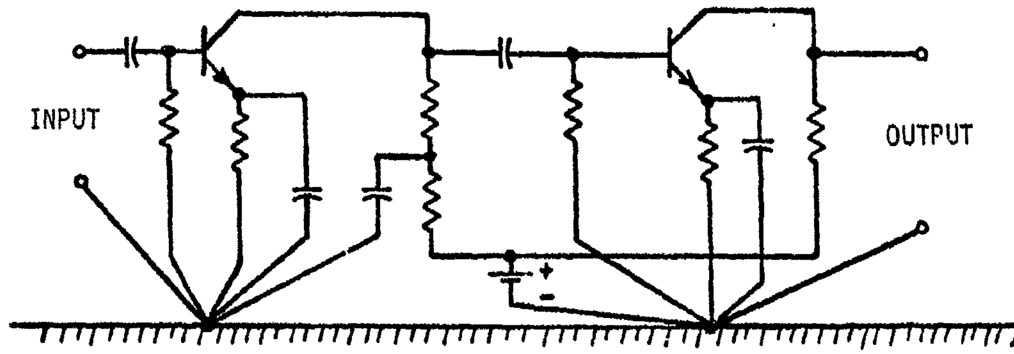

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Figure 3-118. Multipoint Ground Detail

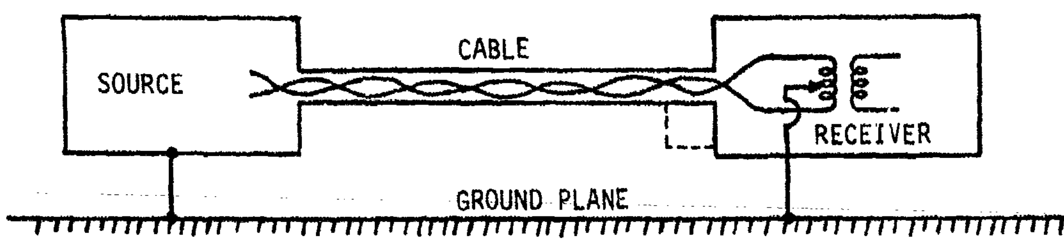

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Figure 3-119 Balanced Coupling Circuit (Ref. 111)

The use of balanced circuits can be made most effective where conductive coupling is not required in the original circuit, such as when the signal is ac, and transformer coupling can be used. Even with dc, a balanced system can be used, but it may be less satisfactory. In some cases, dc to ac inverters have been inserted in signal circuits to avoid this difficulty. Details of common mode voltage rejection circuits and their analysis are given in Ref. 112.

3-3.4.4.5 Ground Loops

Ground loops are most likely to arise where parallel grounds are possible because of conflicting requirements. For example, where coaxial cable is required for low loss or distortionless signal transmission, the outer conductor may appear in parallel with the power supply ground. The appearance of a magnetic field in the loop formed by the parallel grounds may result in a current induced at the frequency of the magnetic field. If the magnetic field arises from switching or pulsed currents, it may have a broad frequency spectrum. To reduce the effects of such loops, the power grounds should be run close to the signal cable, or triaxial cable may be used so that the signal carrying portion of the cable may be insulated from such loop currents.

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CHAPTER 4
EMC DESIGN TECHNIQUES

In order to obtain electromagnetic compatibility, the design engineer has a variety of techniques at his disposal. These can be defined as lying in two broad divisions: (1) initial design, in which one attempts to select components and circuit arrangements which cause the minimum amount of undesired interaction and, where this is not possible or does not result in an optimum design, (2) isolation, in which circuits that would mutually interact in an undesirable way are prevented from doing so by interposing an electrical barrier. The isolation can be obtained by filters, shields, or physical spacing. As an example, one can (1) design a circuit so that the spurious frequencies are not generated in the first place, or (2) design a circuit in which the frequencies are generated but are prevented from reaching a susceptible circuit by the interposition of an appropriate filter.

In this chapter, initial design considerations are treated in the early paragraphs covering emission control, susceptibility control, and coupling control. Isolation techniques are treated in the later paragraphs covering wiring and cabling, filtering, shielding, and grounding and bonding. The emphasis is on circuit design, but the principles and techniques apply to obtaining electromagnetic compatibility between the larger units such as specific equipments or systems. For the latter, design details may be found in various paragraphs of Chapters 5 and 6.

4-0 LIST OF SYMBOLS

A = area, m²; absorption power loss, dB; penetration loss, dB; amplitude, V
A_a = aperture attenuation, dB
A_{L2} = shielding factor, susceptible circuit unshielded (transient), dimensionless
A_{SS} = shielding factor, emitter and susceptor shielded (transient), dimensionless
A_{S1} = shielding factor emitting circuit shielded (transient), dimensionless
a = width of conductor between holes, m
a_{c2} = shielding factor, susceptible circuit shielded (ac), dimensionless
a_{L2} = shielding factor, susceptible circuit unshielded (ac), dimensionless
a_{S1} = shielding factor, emitting circuit shielded (ac), dimensionless
a_{S2} = shielding factor, susceptible circuit shielded (ac), dimensionless
B = re-reflection loss, dB; loss factor due to multiple reflections in shield, dB
B_a = correction factor for aperture reflections, dB
b = length of a rectangular hole, m; width of conductor, in.
C = capacitance, F; mutual capacitance, F
C_b = emitter bypass capacitor
C_d = decoupling capacitor
c = thickness of conductor, in.
D = depth of aperture, in.; distance between conductors, in.; damping coefficient, dimensionless
d = diameter of conductor, in.; damping factor (dimensionless); diameter of circular aperture, m
E = electric field strength, V/m
E_L = load voltage, V
E_{L1} = load voltage with filter, V
E_{L2} = load voltage without filter, V
E_S = source voltage, V
E_{S1} = input voltage, V
e = voltage, V
e_c = common-mode voltage, V
e_d = differential-mode or balanced voltage, V
e_f = feedback signal, V
e_g = interference generator voltage, V
e_s = source voltage, V
e_x = difference between input and feedback signals, V
e_2 = voltage induced in the susceptible circuit, V; noise voltage, V
f = frequency, Hz
f_c = center tuned frequency, Hz
f_n = notch frequency, Hz
f_o = cutoff frequency, Hz; center frequency, Hz
G = relative conductance, dimensionless; gain without feedback, dimensionless
G_f = gainwidth feedback, dimensionless
g = conductivity relative to copper, dimensionless
H = magnetic field strength, A/m
$H(f)$ = voltage transfer function, dimensionless
h = height above ground plane, m
I_f = fault current, A
I_L = complex load current, A

I_S = complex source current, A
IL = insertion loss, dB
i = current, A
i_1 = source current, A; emitting circuit current, A
i_2 = current induced in the susceptible circuit, A; current induced in victim circuit, A
K = parameter ratio for twin T-filter, dimensionless
K_1 = correction factor for number of openings per unit square — antenna is far from shield compared with distance between holes in shield, dB
K_2 = correction factor for penetration of the conductor at low frequency, dB
K_3 = correction factor for coupling between closely spaced shallow holes, dB
K_ω = ratio of shield impedance to impedance of magnetic field, i.e., Z_s/Z_H, dimensionless
k = gyrator constant, Ω^2; ratio of aperture characteristic impedance to impedance of an incident wave, dimensionless
L = self-inductance, H; filter inductance, H; conductor length, m
L_C = net self-inductance of a shielded susceptible wire, H
L_S = self-inductance of shield, H
L_1 = foil inductance, H; self-inductance of source circuit, H
L_2 = self-inductance of victim circuit, H
l = length, m
M = mutual inductance, H
N = number of holes per square inch, $1/\text{in.}^2$
P = power density, W/m^2
P_1 = incident power density, W/m^2
P_2 = transmitted power density, W/m^2
p = helix pitch, cm; ratio of wire diameter to skin depth, dimensionless; ratio of conductor width to skin depth, dimensionless
Q = resonant circuit quality factor, dimensionless
Q_u = unloaded Q, dimensionless
R = resistance, Ω; reflection power loss, dB
R_a = aperture reflection loss, dB; source resistance of emitting circuit, Ω
R_b = load resistance, Ω
R_c = terminating resistance (nearest the emitting circuit source), Ω
R_d = circuit terminating resistance (nearest the emitting circuit load), Ω
R_E = electric field reflection loss, dB
R_f = emitter resistor, Ω
R_M = magnetic field reflection loss, dB
R_S = resistive component of shield conductor, Ω; shunt resistance, Ω
R_s = short circuit resistance; source resistance, Ω
R_1 = resistance of metalized foil, Ω; first boundary reflection loss, dB
R_2 = second boundary reflection loss, dB
r = distance from source to shield, in. or m
r_o = output impedance, Ω
SE = shielding effectiveness, dB
TE_{on} = transverse electric mode (in waveguide)
t = time, s; material thickness, m or in.
V_{cb} = collector bias voltage, V
V_d = magnitude of interference voltage, V
$V_{d(max)}$ = maximum magnitude of interference voltage, V
V_{EE} = emitter bias voltage, V
v = voltage, V
v_c = carrier signal, V
v_{cc} = predistorted carrier signal, V
v_o = output voltage, V
v_L = load voltage, V
v_s = modulating signal, V
v_{ss} = predistorted modulating signal, V
W = diameter, in.; width of rectangular aperture, in.; width of center conductor, m or in.
X = internal reflection correction factor, dimensionless
Z_a = impedance of a rectangular waveguide well below cutoff, Ω
Z_b = ground impedance, Ω
Z_c = collector impedance, Ω
Z_d = decoupling impedance, Ω
Z_g = interference generator impedance, Ω
Z_{gp} = ground plane impedance, Ω
Z_i = input impedance, Ω
Z_L = load impedance, Ω
Z_o = impedance at resonance, Ω
Z_s = source impedance, Ω; impedance of metal (shield), Ω/m^2; aperture characteristic impedance, Ω/m^2
Z_w = impedance of incident wave, Ω
Z_ω = impedance of magnetic field, Ω

α = waveguide attenuation, neper
β = fraction of output fed back to the input, dimensionless
δ = helix conductor separation, mm
ϵ = permittivity, F/m
ϵ_0 = permittivity of free space = $10^{-9}/(36\pi)$ F/m
λ = wavelength, m
μ = permeability, H/m
μ_0 = permeability of free space, $4\pi \times 10^{-7}$ H/m

μ_r = permeability relative to free space, dimensionless
ρ = resistivity, Ω·cm or Ω·m
σ = complex propagation constant; conductivity, mho/m
τ = characteristic rise time, s
χ = multiplying factor, dimensionless
ω = angular frequency, rad/s

4-1 EMISSION CONTROL

The emission of electrical energy that can cause interference may be directly associated with the generation of the desired signal, or indirectly as a secondary effect associated with providing equipment operational control, switching, ignition, or power.

4-1.1 SIGNAL DESIGN

An electrical signal is used to convey information. As such, it will occupy a limited but necessary part of the frequency spectrum. In the interest of minimizing interference, such a signal should occupy no more spectrum space than absolutely necessary for the information rate required.

The minimum bandwidth required is dependent on the baseband signal. Voice at baseband requires about 3 kHz, while analog data requires a bandwidth dependent upon the expected rate of change of the variable being transmitted (current or voltage). Digital systems required a bandwidth somewhat greater than the maximum pulse rate.

Digital signals are particularly easy to analyze. As was shown in par. 3-1.3.2.1, the simple step function has an infinite spectrum. By limiting the rise time of the step function, the spectrum energy is reduced, as shown on Fig. 3-20. Similarly, the spectrum of a pulse is critically dependent upon its shape. Fig. 4-1 shows the spectra of several pulse types (including rectangular, triangular, trapezoidal, raised cosine, and Gaussian) for pulses of equal area. It is seen that the Gaussian-shaped pulse occupies the least spectrum.

Where signal transmission is required over long distances, utilizing wire or radio transmission media,

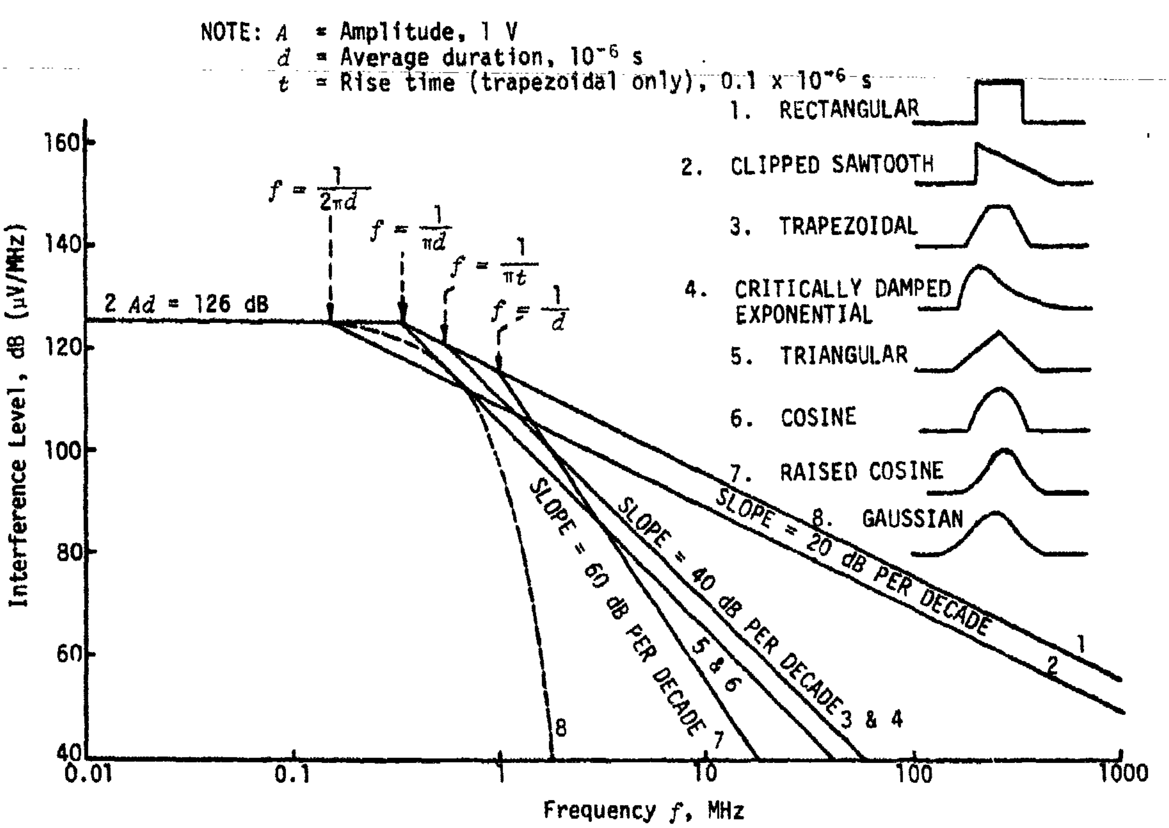


Figure 4-1. Interference Levels for Eight Common Pulses

translation in the frequency spectrum and modulation may be necessary. In addition, analog-to-digital conversion may be desired. The bandwidth considerations in the choice of modulation are discussed adequately in texts on communication systems (Ref. 1) and are not repeated here, but in the early stages of system design they are important considerations, not only from the point of view of emission of interference by the system, but also from that of susceptibility of the system to external signals.

Except for signal design considerations, interference generation in most emitters results from secondary effects associated with the normal functioning of the device. Hence, interference control measures applied at the design stage call for minimizing undesired effects without degrading the efficiency with which the device in question performs its intended function.

4-1.2 MECHANICAL SWITCHES

As discussed in par. 3-1.3.2, mechanical switches cause sudden changes in current and produce energy over a broad frequency spectrum. For many functions the changes in current are much more rapid than required, and considerable interference results from arcing and restriking phenomena at the switch contacts caused by circuit inductance. The arcing and restriking phenomena can be minimized by using fast-acting swtiches. Otherwise, suppression techniques — such as application of filters, nonlinear devices, and shielding, described in par. 5-3.1 — must be applied.

4-1.3 DIODES

These circuit elements operate much as switches, and indeed are capable of very short effective switching times. The broad frequency spectrum can be controlled in the same way as for mechanical switches. Extremely sharp impulses may arise from minority carrier storage (see par. 3-1.3.2.5).

The typical diode used in rectifier or detector circuits switches at zero voltage. The Zener diode switches from off to on at some negative voltage and, because of the avalanche effect associated with breakdown, it can produce a very rapid rise in current. This can be controlled by series circuit resistance and reactance.

4-1.4 TUNNEL DIODES

These devices have an inherent negative resistance characteristic and usually are used in circuits for obtaining high switching speeds, or in conjunction with reactances to generate sustained oscillations at particular frequencies. In a circuit that has multiple resonant frequencies (because of multiple or distributed reactances), oscillations can be generated at frequencies other than the intended one. Such behavior can be avoided by using elements having a minimum of distributed reactance, short leads, and small series parasitic damping resistors in circuits not part of the main resonant path.

4-1.5 TRIODES AND TRANSISTORS

When operated in a switching mode, transistor and semiconductor controlled rectifiers (SCR's) may produce shorter rise times and larger currents than diodes. The emission spectrum can be controlled by similar techniques in which the suppression elements are placed between the emitter and the collector. For use in amplifiers, modulators, or detectors, vacuum or semiconductor devices should be operated without excessive voltages or currents in order to limit harmonic or other spurious frequency generation (see par. 3-2.2). One of the first steps to be taken is to select the operating point on transistor or tube transfer characteristics to optimize performance. For amplifiers one should maximize the linear range, while for modulators and detectors usually a square-law characteristic is optimum.

4-1.6 POWER AMPLIFIER DESIGN

Where it is desired to obtain large power output, it is expedient to operate transistors or vacuum tubes in a "large signal" mode. Inherently, such operation is likely to generate large amounts of harmonics. The types of operation have been divided into broad classes known as Class A, B, and C (Ref. 2). In Class A operation, the plate or collector current flows throughout a complete ac cycle and usually is considered to generate the lowest harmonic levels of all three classes. In Class B operation, it flows only over one-half cycle in any one tube or transistor, but two of these elements are used connected back to back so that current flows in the output circuit over the full cycle. With balanced circuit elements there is a tendency for even harmonics to be cancelled, but odd harmonics can be quite large without careful design. In Class C operation, current flows in the active element only over a small fraction of the cycle. Large amounts of harmonics are generated which must be filtered in order to prevent them from reaching the output of the device. Class C amplifiers are used primarily at radio frequencies where the signals to be amplified are narrow-band and where sharply-tuned, high-Q filters are available to control levels of harmonic frequencies.

4-1.7 LINEARIZATION TECHNIQUES

Reduction of harmonic generation in amplifiers can be achieved by use of feedback and compensating nonlinearity (Ref. 3).

The use of negative feedback to improve linearity is relatively well known (see Fig. 4-2). By using the technique, one reduces the gain of the stages over which it is applied from the gain without feedback G to

$$G_f = \frac{G}{1 + G\beta}\text{, dimensionless} \tag{4-1}$$

where β is the fraction of the output voltage v_o fed back to the input. This gain reduces to $1/\beta$ when $G\beta$ is much larger than 1.

The advantage of negative feedback is that it permits the amplifier output stage to be operated at a large signal level, with distortion much reduced from what is obtained without it. Fig. 4-3 shows how feedback can be applied to reduce distortion in the output of a modulator by connecting a demodulator in the feedback path.

An example of the use of predistortion of the applied signal to overcome nonlinearity is shown in Fig. 4-4. A parametric diode modulator was used. The predistortion network contains a compensating diode. Through this technique, considerable improvement in modulator performance may be obtained.

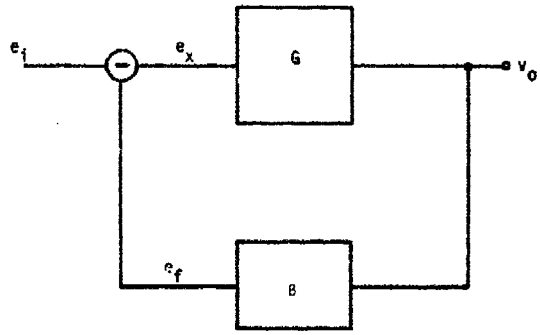


Figure 4-2. Feedback Network

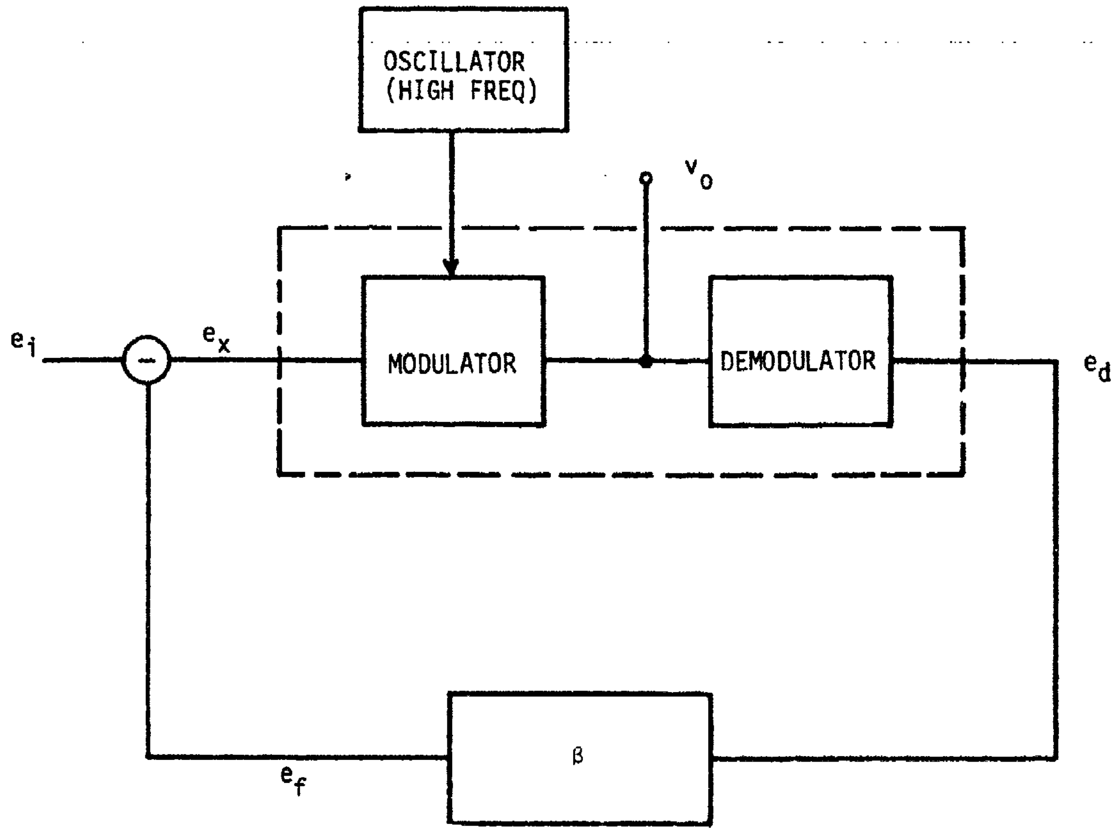
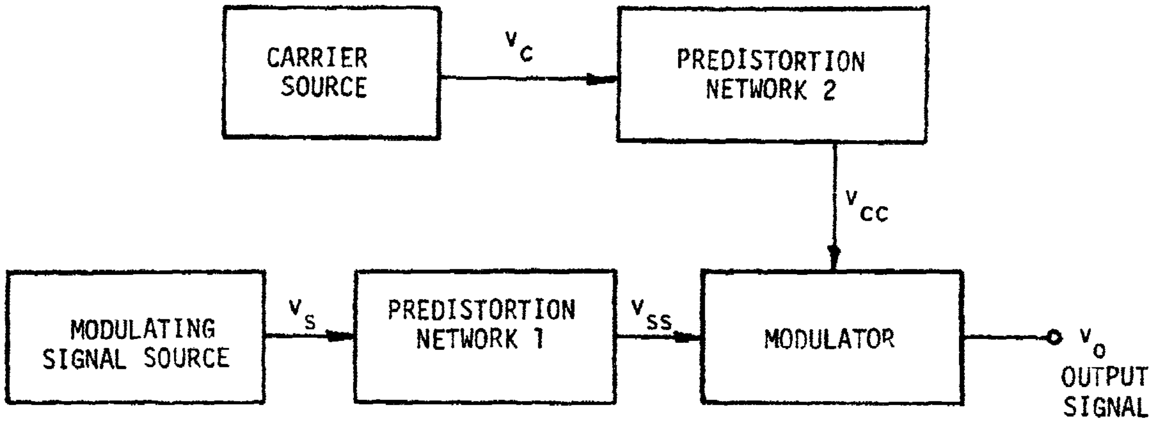


Figure 4-3. Feedback Around Modulator



v_c = carrier signal
v_s = modulating signal
v_o = output signal
v_{ss} = predistorted modulating signal
v_{cc} = predistorted carrier signal

Figure 4-4. Predistortion Block Diagram

4-1.8 BALANCED CIRCUITS

As shown in par. 3-3.3.3, common-mode currents in cables are more effective in producing external fields than differential-mode currents. To reduce common-mode emissions, balanced circuits should be used. With such circuits, balanced output cables consisting of twisted-pair conductors can be used to maintain the balance.

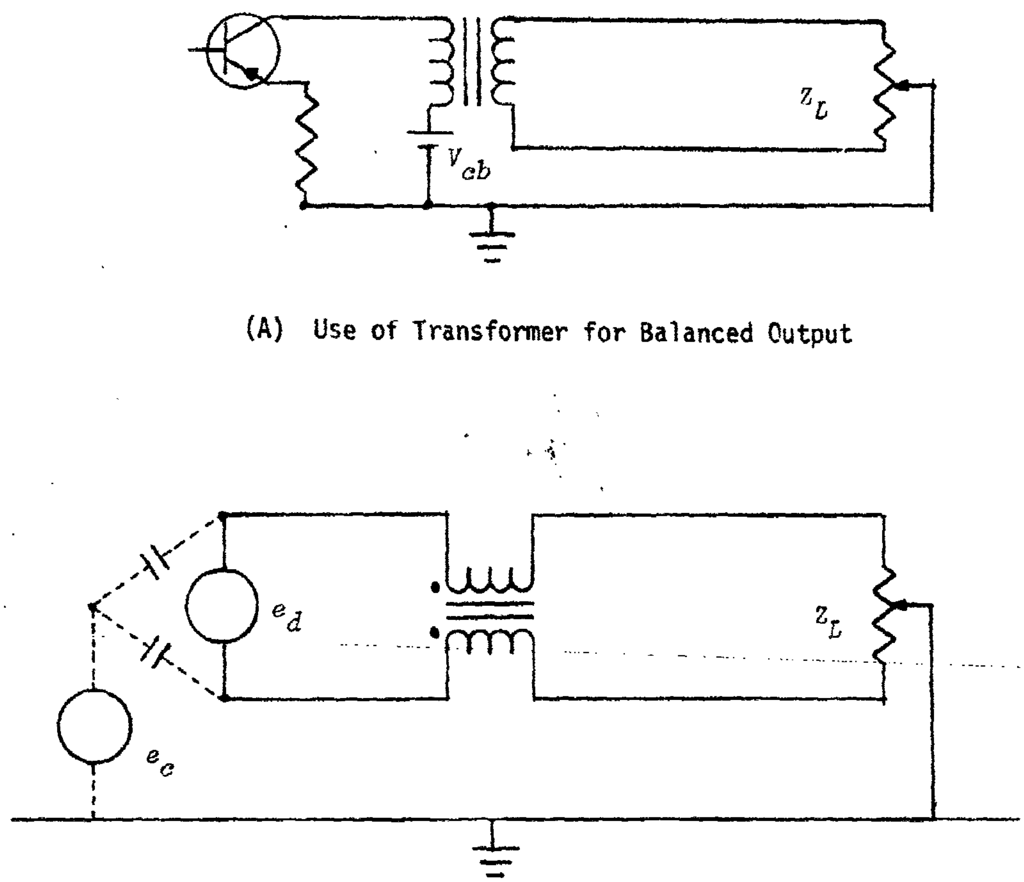
Where single ended circuits are used, a balanced output can be obtained using a transformer or a "balun", as shown in Fig. 4-5 (A). At frequencies up to at least 100 kHz, transformers are available for this purpose. Some are designed specially to provide a high degree of balance, and have internal shields to minimize capacitance between primary and secondary windings, which could couple common-mode voltages into the output circuit.

Where sensitivity to common-mode signals is of special concern, it may be appropriate to insert a common-mode choke into the output line. A common-mode choke (Ref. 4) is essentially a one-to-one transformer in which the windings are connected in series with the lines to the load as shown in Fig. 4-5(B). The windings are polarized and the device presents an impedance $j\omega L$, where L is the self-inductance of one of the windings, to the flow of currents excited by the common-mode voltage e_c. To be effective, the coefficient of coupling between the two windings should be as near unity as possible (Ref. 5).

4-2 SUSCEPTIBILITY CONTROL

If undesired energy does appear in a circuit, malfunction can occur as a result of changes in operating characteristics such as stage gain or triggering voltage, or as a result of blocking or spurious signal generation. In addition, high level undesired signals can produce permanent performance degradation because of burnout. The energy can be considered to enter the circuit in three ways: (1) via signal leads, (2) via control or power leads, and (3) via radiation effects of local fields.

Susceptibility to local fields can be reduced by minimizing wiring loop areas and lengths of exposed wires. Quantitative relations between induced voltages and field levels are given in pars. 3-3.1.1, 3-3.1.2, and 3-3.2.4. If the field susceptibility cannot be



(A) Use of Transformer for Balanced Output

(B) Circuit With Common-mode Choke

V_{cb} = collector bias voltage
e_d = differential mode or balanced voltage
Z_L = load impedance
e_c = common-mode voltage

Figure 4-5. Circuits for Controlling Common-mode Currents

reduced adequately in this way, shields must be used (see par. 4-6).

The most important circuit design technique for limiting susceptibility of signal circuits is to maintain linearity over as broad a range of input voltage or current as possible. Techniques for improving linearity have been discussed in connection with emission control techniques (par. 4-1.7). Just as these techniques reduce spurious frequency generation and subsequent emission, they also limit response of a signal-receiving device to unintended signals. The techniques include setting the operating point, using negative feedback, and compensating nonlinearities.

4-2.1 SELECTIVITY

Susceptibility to frequencies outside of the normal range of a given device can be reduced by the addition of selective circuits or filters, especially ahead of the first nonlinear circuit element. Filters also can be used to reject specific frequencies such as image or intermediate frequencies. These techniques are discussed in more detail in par. 5-7 on receivers.

The effects of undesired signals at frequencies close to the desired one can be reduced by use of automatic frequency control (AFC). The function of this circuit is to compensate for shifts in receiver local oscillator frequency which otherwise would result in receiver mistuning.

Other specialized techniques for reducing susceptibility include limiting and noise pulse blanking.

4-2.2 LIMITING

The limiter makes use of the relatively sharp cutoff characteristic of a diode (or the base of a transistor, or grid of a vacuum tube) by which the element switches from a nonconductive state to a conducting state when the bias voltage reaches zero. By proper circuit design, this technique can be used to limit voltages in following amplifier stages to those near the maximum for which the circuit is designed. The limiter is most effective in reducing the effects of large short-duration transient voltages. Since the circuit involves rectification of large signals, the circuit should provide for rapid discharge of any accumulated charges.

Silencers are used to reduce the output of a receiver to an inaudible level during the time when high interfering voltages are applied to minimize interference with the received message. They are used also on devices having other than an audio output.

4-2.3 COHERENT AND MATCHED FILTER DETECTION

Coherent detection is distinguished from envelope detection in that it makes use of the reference or carrier phase to detect signals modulated in amplitude or phase. The reference phase is obtained from the signal source, usually in the form of a small unmodulated sinusoid sent along with the information signal. The technique has been shown to have some advantage in detecting signals in random noise and to be useful in rejecting some forms of co- and adjacent channel interference.

The matched filter or correlation detector also is used to improve interference rejection (Ref. 6).

4-3 COUPLING CONTROL

As pointed out in par. 3-3, coupling is of two basic types: conducted, and radiated (including induced).

4-3.1 CONDUCTIVE COUPLING

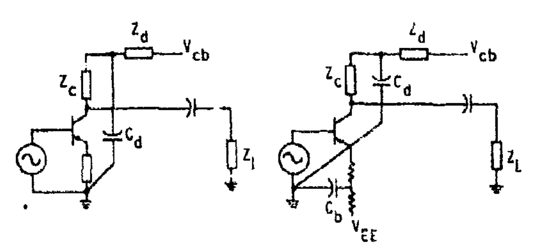
The general principles underlying conductive (impedance) coupling are discussed in par. 3-3. It can be reduced by inserting filters which are frequency selective to permit the desired signal to pass, but attenuate any undesired signal which is presumably of a different waveform and frequency. In certain circumstances, where the desired signal is much stronger than an undesired signal, coupling control might be effected by inserting an attenuator which would be nonfrequency selective. It would reduce the undesired signal to a level which would render it harmless, while at the same time passing the desired signal with sufficient amplitude to allow it to perform its function.

4-3.1.1 Decoupling

Electromagnetic energy in any given functional electronic circuit can be unintentionally conductively coupled to other stages using the same bias power supplies. It is frequently necessary to decouple the bias supplied on a stage-by-stage basis.

4-3.1.1.1 Power Output Stages

For high-current stages, such as a transistor power output stage, the collector current which does not flow through the load should be bypassed to the same point as the emitter as shown in Figs. 4-6(A) and (B). This technique reduces signal current through the power source V_{cb} and hence the coupling to other circuits connected to that supply. In addition, the current through the wiring may induce voltages in other wires. A series impedance Z_d is provided to



(A) Emitter Returned to Ground

(B) Emitter Returned to Bias Supply V_{EE}

Figure 4-6. Output Stage Decoupling

raise the impedance of the path through the power supply which shunts C_d. There will be a frequency below which the high impedance of C_d will force the current to flow through the power supply. The value of C_d and Z_d should be selected so that this frequency is much lower than any frequency likely to be impressed on the input to the stage.

The signal current that flows through Z_L is the desired output of the device. This current also should be returned to the emitter with a minimum of disturbance to other circuits. If both the emitter and the load are connected to the chassis, the return current can provide a voltage drop in the chassis impedance which might interfere with other circuits. For low-frequency circuits, where the signal system is grounded at only one point, the best return path is a wire which is twisted with the other wire to Z_L. At high frequencies, the capacitance of the lead which connects to Z_L will cause currents to flow to the chassis along its length and to adjacent wires. This current, too, must get back to the emitter. A shield around the cable will provide a defined return current path if the shield is connected to the emitter return point. If both emitter and load are grounded, a percentage of the return current will flow on interchassis grounds. Twisted pairs inside of shields and transformer coupling should be considered for decreasing this coupling. Fig. 4-6(B) shows the emitter bypass capacitor used with two-supply biasing. With the connection shown, the transistor current flows through the emitter bypass capacitor C_b. It might seem preferable to connect C_d to the emitter side of C_b so that the collector current does not pass through the emitter bypass capacitor. However, that connection allows disturbances on the power supplied to be coupled into the base emitter signal loop, hence the method shown in Fig. 4-6(B).

4-3.1.1.2 Tuned Circuits

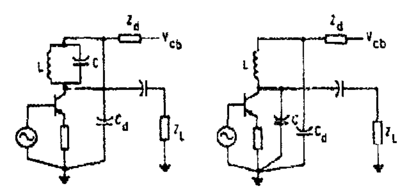
Interference to tuned output currents can be minimized by connecting any tuning capacitor across the tuning coil rather than between collector and ground. The two configurations are shown in Figs. 4-7(A) and (B). With the capacitor across the coil (Fig. 4-7(A)) the current through the decoupling capacitor at resonance is $v_o/(Q_u \omega L)$, where Q_u is the unloaded Q (resonant circuit quality factor) of the tank circuit, ωL is the impedance of one arm of the tank at resonance, and v_o is the output voltage. If the capacitor is connected instead to ground (Fig. 4-7(B)), the current through C_d is $v_o/(\omega L)$, which is higher by a factor Q_u, which may be approximately 100. Hence, for easier decoupling the former connection is preferred. In many cases, most of the current is through distributed capacitance to the chassis and for those cases C_d will have to be chosen large enough to handle the full tank current.

4-3.1.1.3 Emitter Followers

Fig. 4-8 shows an emitter-follower stage with the collector bypassed to the emitter ground. However, if $Z_L \ll R_f$ (typical case when a separate emitter supply is used), the current through the chassis can be reduced by returning C_d to the point at which the load current is returned to the chassis.

4-3.1.1.4 Interstage Decoupling

The interstage coupling of a pair of transistors is shown in Fig. 4-9. The second stage transistor is represented by its input impedance Z_i, and its base biasing resistors by R_1 and R_2. The function of this stage is to amplify the input signal represented by e_s and supply maximum current in Z_i and minimum current to the impedances in common with other circuits. At the same time, disturbances on the supplies or in the chassis impedance should supply minimum current to Z_i. The emitter is shown bypassed to the input signal ground to return the base current signal directly to the driving source without going through the chassis impedance. The ground point of C_d has



(A) Tuning Capacitor Across Inductor (B) Tuning Capacitor Returned to Ground

Figure 4-7. Tuned Output Stage Decoupling

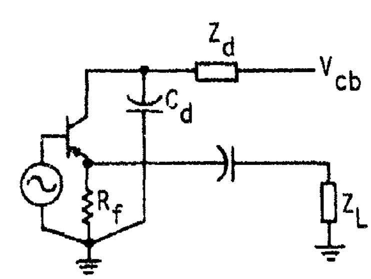


Figure 4-8. Emitter-Follower Decoupling

conflicting requirements. In the connection shown, all of the transistor current flows through the chassis so that coupling to other stages may exist. If C_d is connected back to the first-stage ground, this chassis current is reduced by the amount of current that flows through Z_c and R_2. If a current exists in the chassis due to some other source, however, the configuration of Fig. 4-9(A) minimizes the amplification of this undesired current by the stages which follow. This is demonstrated in equivalent circuits shown in Figs. 4-9(B) and (C). The transistor has been replaced by its output impedance r_o. The circuit of Fig. 4-9(A) is shown in Fig. 4-9(B), where e_g and Z_g represent the interference source, e.g., chassis currents. Any current which passes through Z_i must pass through r_o, which is usually a high impedance. However, in Fig. 4-9(A), which represents the circuit with C_d returned to the emitter ground, the current through Z_i can flow through R_1 and Z_c which are usually much lower impedances than r_o.

4-3.1.1.5 Flip-Flops

When flip-flops change state, the current required from the supply changes momentarily. The generated pulse is rich in high-frequency components and can couple into wires adjacent to those carrying the supply current. These transients should be kept on the flip-flop board with a decoupling network and by locating all circuit elements as close as possible.

Some computer circuits use a reference voltage to bias clamping diodes to obtain constant-level pulses. The resulting pulse currents in the reference supply and wiring are a potential source of interfering signals. A decoupling network consisting of an inductor and resistor, and shunt capacitor usually will enable control of high-frequency current in the reference supply and wiring.

4-3.1.1.6 Switching Power Supplies

Power supplies may use a square-wave chopper to generate the high voltage from the low voltage input. The sharp transient currents associated with the chopper may have rise times in the order of 1 μs. Filtering and shielding may be required. Usually, a transformer is incorporated which enables input and output circuits to be isolated so that the power supply might have the configuration shown in Fig. 4-10. Switching power supplies are discussed also in par. 5-5.

4-3.1.1.7 Sensitive Audio Amplifiers

Since the amplifier itself usually is grounded, in order to avoid coupling to various currents flowing in the ground plane, it is necessary to use a transformer to isolate the input from common-mode voltages. With a reasonably well-balanced primary winding, the rejection of common-mode voltages can be as high as 120 dB. The signal wiring should use twisted pairs, and single-point grounding should be used within stages of amplification.

4-3.2 INDUCTIVE COUPLING

Coupling, through induction, is an inherent property of all closely spaced circuit configurations. The coupling can be reduced by isolation of an emitter and a susceptor by: (a) spacing by sufficient distance, (b) judicious arrangement of components, or (c) insertion of appropriate shields between the circuits. Pars. 3-3.1 and 3-3.3 cover the mathematical models used for inductive and capacitive coupling.

At the circuit level, inductive coupling usually takes place between wires and cables. Cables, which consist of pairs of wires carrying currents in both directions, should be twisted, or should be of the coaxial type to minimize emission and susceptibility

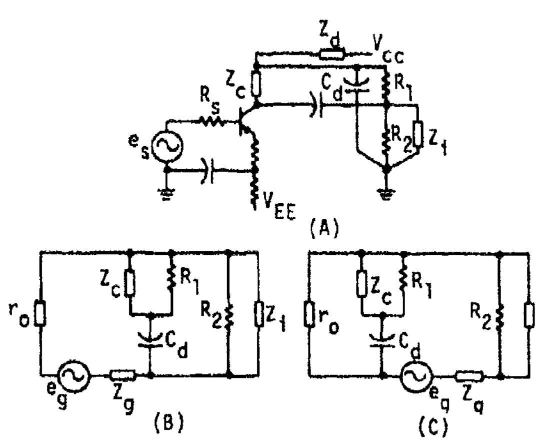


Figure 4-9. Interstage Decoupling

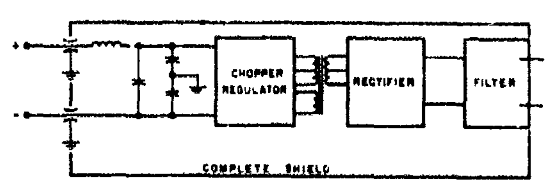


Figure 4-10. Typical Final Power Supply Configuration

characteristics. In many cases, single conductors are used. Their placement can be critical in some circuits.

4-3.2.1 Mutual Impedance

Inductive coupling can be looked upon as a form of impedance coupling. Fig. 4-11 shows the mutual inductance per centimeter between two parallel wires, and Fig. 4-12 shows the mutual capacitance. From those figures, the voltages and currents which can be coupled unintentionally between co-located circuits can be estimated at the various frequencies of concern. It can be noted that the coupling falls off rapidly as the distance is increased. The presence of a nearby ground plane contributes to a more rapid fall off than otherwise. Where circuit loops can be well defined, the area subtended by the loops should be as small as possible.

These formulas may not be accurate for circuit dimensions and spacings larger than about 1/10 of a wavelength since the coupling can be influenced by resonant conditions. At such frequencies, special care is necessary to avoid undesired coupling, and shields must be used extensively.

Also, it should be noted that where there are many different circuits in close proximity, there will be multiple coupling paths between any two points. In such complex circuits, special care must be taken to reduce unwanted coupling paths to the minimum possible.

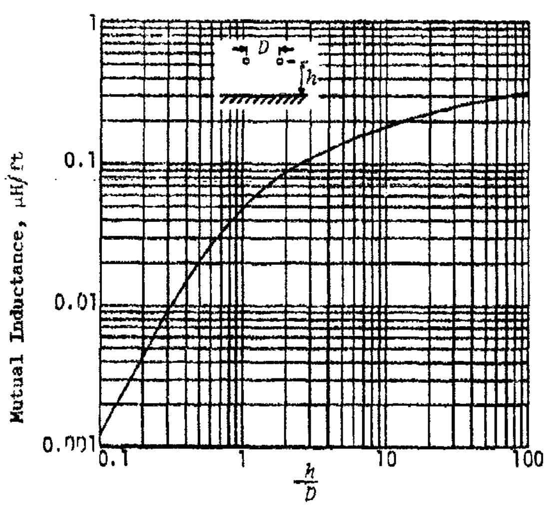


Figure 4-11. Mutual Inductance Between Two Wires Over a Ground Plane

4-3.2.2 Transient Coupling

Where the circuit contains pulses, approximate analysis can be made in terms of rates of change of voltage and current. The current i induced in a circuit through a mutual capacitance C is given by

$$i = C\left(\frac{dv}{dt}\right), \text{A} \tag{4-2}$$

and the effective series voltage v induced in a closed circuit through mutual inductance M is

$$v = M\left(\frac{di}{dt}\right), \text{V} \tag{4-3}$$

From a knowledge of the victim circuit impedances, the current resulting from Eq. 4-2 or the voltage from Eq. 4-3, which appears at the susceptible device, can be estimated. This estimate is valid only if the total energy transferred between the two circuits

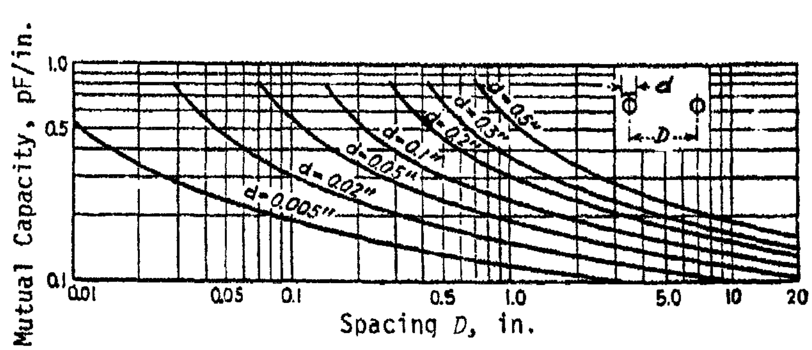


Figure 4-12. Capacity per Inch of Two-wire Line With Various Wire Diameters as a Function of Spacing

is a relatively small part of the energy contained in the source transient, and the voltage and current time rates of change are known. If the coupling is not small, a complete coupled circuit model must be established and solved.

Items that produce short rise time pulses are silicon-controlled rectifiers (SCR's); vacuum contacts and gaps; dc motor and generator commutators; insulation failures that produce spark-over; and nearly all gaseous discharges such as sparks, glow discharges, and arcs. Common gaseous discharge devices are arc welders, neon lights, and fluorescent lamps. Some devices that produce heavy fields are multiturn elements, often with partial magnetic cores, such as relays and other magnetic actuators. Also, these devices almost always produce high conducted interference unless equipped with suppression components (e.g., diodes) (see par. 5-6).

4-3.2.3 Constant Magnetic Fields

When the magnetic field is unchanging or changing slowly, it can interfere with devices responsive to constant magnetic fields. Well-known examples are a magnetic compass, a Hall-effect device, a cathode-ray display tube, and superconductors. The magnetic field can interfere also with devices which normally operate with an alternating field, but whose magnetic core is subject to saturation. Small RF or AF reactors and transformers with open ferrite or powdered iron slug cores are examples. Other devices such as solid-state rectifiers, transistors, lasers, and other light-emitting devices can be affected by strong fields.

Permanent magnetic fields may appear near cabinets or supports made of steel as a result of the manufacturing process. This magnetism may directly polarize nearby components.

4-3.3 RADIATIVE COUPLING

At frequencies above about 100 MHz, it is usually meaningless to differentiate between electric and magnetic field components, since both can be expected to exist simultaneously. The rate of fall off of coupling with distance is not faster than inverse distance. Because the wavelength is of the order of circuit dimensions, the voltages or currents induced may be quite frequency sensitive due to resonance effects and, as at lower frequencies, can cause bias to appear across diodes, transistors, tube grids, and other circuit elements. The circuits can exhibit nonlinear properties at sufficiently large field levels, and can generate beat frequencies between the interference and the signal frequencies in the circuit along with harmonic frequencies of the interfering signal.

4-4 WIRING AND CABLING

4-4.1 INTRODUCTION

The term wiring refers to interconnections among elements of one chassis or device enclosed in a single cabinet, whereas cabling refers to methods of interconnecting separate chassis or cabinets.

4-4.2 DESIGN CONSIDERATIONS

During initial design stages, consideration should be given to proper location of equipment and wiring to minimize interference coupling between transmission paths. Sensitive components should be kept as far as possible from units that may be sources of electrical interference. Cabinet panels or partitions should be used to separate or shield these components.

Power leads, control wiring, and other cabling to sensitive equipment should not be close to any interference source or leads that may be carrying interference because of the inductive coupling that can exist between wires. If it is necessary for sensitive signal leads to pass near interference-carrying leads, relative orientation should be as nearly at right angles as possible to minimize the magnetic coupling effect. The distribution of power through multiple lines, from a primary power source to the components of a piece of equipment, is recommended to reduce component interaction. The signal circuits should be separated from ac power circuits and any other circuits that can transfer interference to them.

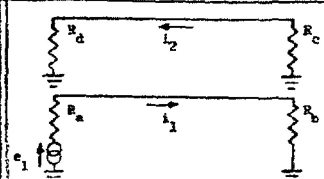
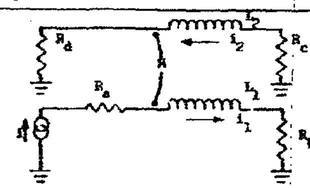
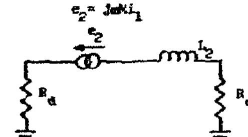
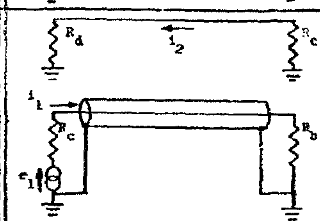
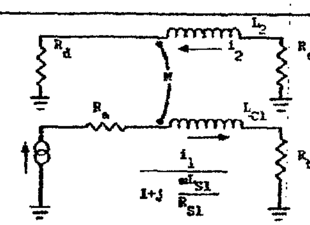
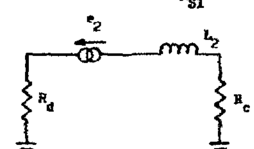
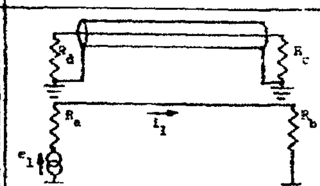
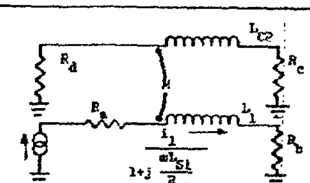
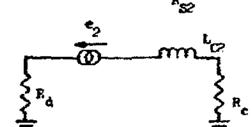
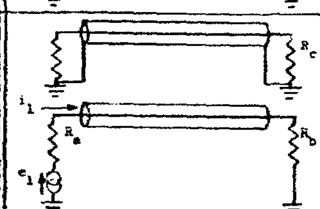
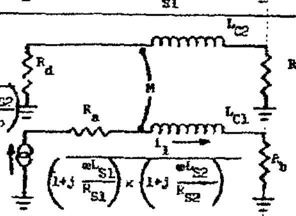
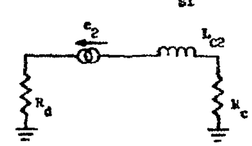
The use of shielded hookup wire inside a chassis helps to prevent internal interference coupling to sensitive circuits. At the ends where connections are made, the shield braid should be pared back for minimum length to keep the shielding as complete as possible. It should be bonded directly to the chassis at its end and at convenient points along its length. Leads that run side by side, or cross over each other, should have their shields bonded together.

4-4.3 WIRES OVER A GROUND PLANE

4-4.3.1 Magnetic Coupling

Where the important coupling is magnetic, at low frequencies, one can use lumped impedance models. Results for parallel lines located over a ground plane with and without shields are given in Table 4-1 (Ref. 7). The coupling is given in terms of the voltage

TABLE 4-1. SUMMARY OF OPEN WIRE AND SHIELDED WIRE INDUCED INTERFERENCE

Interference Case	Schematic	AC Equivalent Circuit	AC Equivalent Circuit Of Victim Loop	Output Voltage
(1) Open Wire-To-Open Wire			$e_2 = j\omega M i_1$ 	AC rms: $V_d = 2\pi f M i_1 \frac{R_d}{R_c+R_d} a_{12}$ Peak Transient: $V_{cMAX} = \frac{MI_1}{T}\frac{R_d}{R_c+R_d} A_{12}$
(2) Shielded Wire-To-Open Wire		$\frac{e_1}{1+j\frac{\omega L_{S1}}{R_{S1}}}$  $\frac{i_1}{1+j\frac{\omega L_{S1}}{R_{S1}}}$	$e_2 = j\omega M i_1 \frac{1}{1+j\frac{\omega L_{S1}}{R_{S1}}}$ 	AC rms: $V_d = 2\pi f M i_1 \frac{R_d}{R_c+R_d} a_{S1} a_{12}$ Peak Transient: $V_{dMAX} = \frac{MI_1}{T}\frac{R_d}{R_c+R_d} A_{S1}$
(3) Open Wire To-Shielded Wire		$\frac{e_1}{1+j\frac{\omega L_{S1}}{R_{S1}}}$  $\frac{i_1}{1+j\frac{\omega L_{S1}}{R_{S1}}}$	$e_2 = j\omega M i_1 \frac{1}{1+j\frac{\omega L_{S2}}{R_{S2}}}$ 	AC rms: $V_d = 2\pi f M i_1 \frac{R_d}{R_c+R_d} a_{S2} a_{C2}$ Peak Transient: $V_{dMAX} = \frac{MI_1}{T}\frac{R_d}{R_c+R_d} A_{S2}$
(4) Shielded Wire-To-Shielded Wire		$\frac{e_1}{\left(1+j\frac{\omega L_{S1}}{R_{S1}}\right)\left(1+j\frac{\omega L_{S2}}{R_{S2}}\right)}$  $\frac{i_1}{\left(1+j\frac{\omega L_{S1}}{R_{S1}}\right)\times\left(1+j\frac{\omega L_{S2}}{R_{S2}}\right)}$	$e_2 = j\omega M i_1 \frac{1}{1+j\frac{\omega L_{S1}}{R_{S1}}} \frac{1}{1+j\frac{\omega L_{S2}}{R_{S2}}}$ 	AC rms: $V_d = 2\pi f M i_1 \frac{R_d}{R_c+R_d} a_{S1} a_{S2} a_{C2}$ Peak Transient: $V_{dMAX} = \frac{MI_1}{T}\frac{R_d}{R_d+R_c} A_{SS}$

across a resistor in the victim loop in terms of the current in the interference source line. By dividing these two quantities, one obtains an effective transfer impedance between the two lines under specified conditions of loading. The table also gives magnitudes of the transient voltages which are generated in the victim lines under the various conditions when the current in the source line is a simple exponential rising with a characteristic risetime τ. The values of mutual and self-inductance can be obtained from Figs. 4-11 and 4-13, respectively. For shielded conductors the appropriate parameters are obtained from Figs. 4-14 and 4-15. L_2 is the self-inductance of the susceptible wire, L_{S1} and L_{S2} the self-inductance of the shields on wires 1 and 2, respectively, and L_{C2} is the net inductance of the shielded susceptible wire, i.e., the difference in the inductance of the wire taken by itself and the inductance of the shield. R_{S1} and R_{S2} are, respectively, the resistive components of the shield conductors.

In the formulas in Table 4-1, it is assumed that the shields are connected to the ground plane at both ends, and that the ground plane is impedanceless. Generally, these formulas can be used in the frequency range from about 1 kHz up to several hundred kHz, and of course only for those situations in which the cable is terminated with an impedance that is low compared with the characteristic impedance. In these formulas no account is taken of the effect in coaxial cables which occurs because the center conductor is not precisely in the center of the cable, and which is discussed in par. 3-3.1.1.2.2. At frequencies above about 100 kHz, braid leakage and skin effect become important and calculations should be made in terms of transfer impedance as discussed in par. 3-3.1.1.2.5 (Ref. 8).

4-4.3.2 Electric Coupling

The discussion in par. 4-4.3.1, takes no account of electric field coupling. Such coupling can be significant with high impedance circuits or where long interconnections and highly sensitive circuits are involved. In such cases, simple braided shields can be very effective. For low frequencies, where ground circulating currents should be avoided, the shield should be grounded at one end only.

4-4.3.3 High Frequency Considerations

At very high frequencies, the dimensions of wiring required to provide significant effective radiation or pickup become quite small. For instance, at 100 MHz, a 1 in. length of wire formed into a loop can be a significant impedance. Return circuit dimensions on wire shield ground connections, bypass capacitor ground leads, and similar wiring should be as short as possible. It is imperative that the wiring of filters be

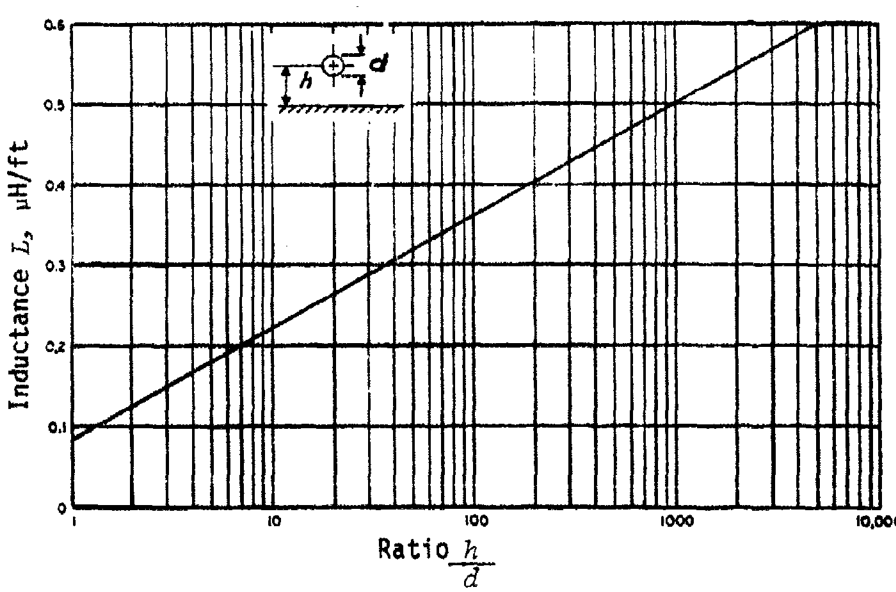


Figure 4-13. Inductance of a Wire Over a Ground Plane

carefully controlled to prevent inductive coupling between input and output of the filter, which will make it ineffective and single-point grounding is not practical. Where long connections are necessary, shielded cables must be used and all connectors must be electromagnetically tight to reduce radio frequency currents external to the shield. In order to reduce such coupling, double shielding may be necessary in radio frequency cabling (see par. 3-3.1.1.2.5.2).

4-4.4 TWISTED PAIRS

In par. 4-3.2, it was emphasized that wherever it is possible to isolate return currents, twisted pairs or coaxial leads should be used for signal transmission. The same principle applies to leads from power supplies, both direct voltage and ac, such as used with filament supplies where twisted leads are commonly used. The fields from twisted wire pairs can be estimated from Figs. 3-66 and 3-68 and the susceptibility of twisted pairs to those fields can be estimated from Fig. 3-82, if the spacings of the wires and their pitch distances are known. Values of these parameters for a variety of cables now in use are given in Table 4-2 (Ref. 9).

4-4.5 POWER WIRING

The three phases of each delta-connected transmission system, and the three phases and the neutral wire of each four-wire, wye-connected transmission system, should be twisted to form one cable. Twisting the wires together effectively cancels the electric and magnetic fields produced by the 120-deg phase-differential voltages and currents for either type of connection. Inclusion of the neutral wire in the four-wire, wye-connection effectively cancels the magnetic field produced by the in-phase third-harmonic currents. These third harmonic currents are generated when the iron cores of transformers or motors are driven to near saturation or operate in the nonlinear portion of the magnetization curve.

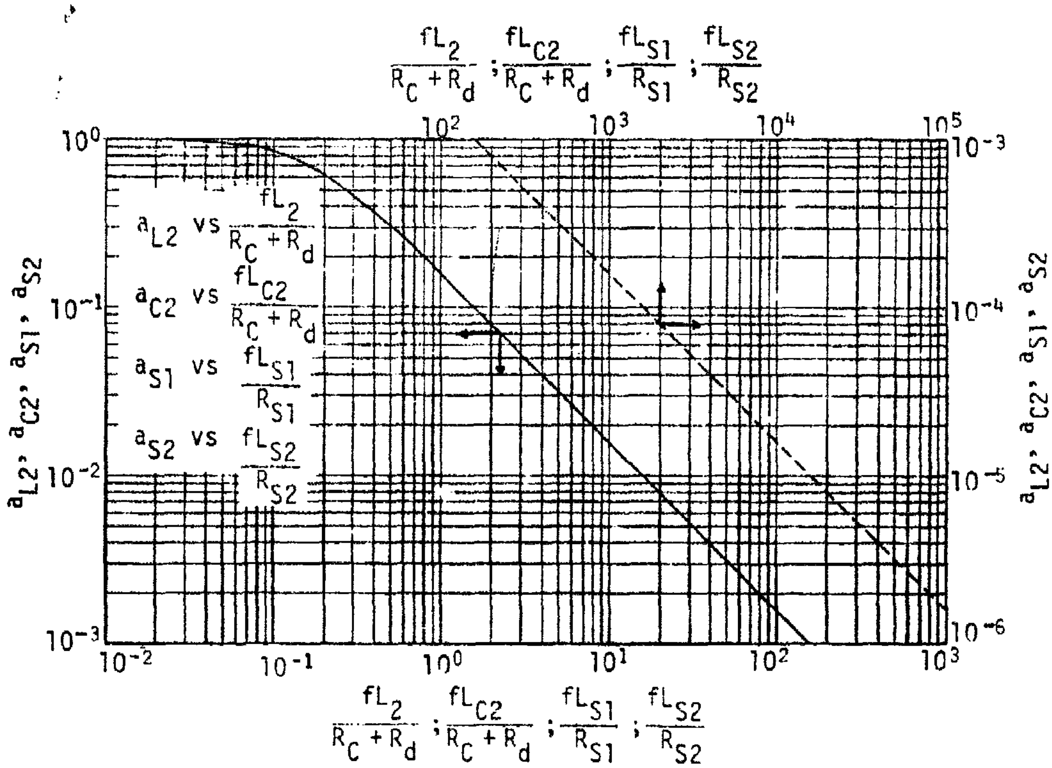


Figure 4-14. AC Inductive and Shield Attenuation Factors

4-4.5.1 Separation of Motor Loads From Signal Equipment Loads

Since a common source of interference is commutation in dc motors and slip-ring friction in ac motors, low-level signal equipment power lines should be separated as much as possible from the power lines to such equipment.

4-4.5.2 Separation of Utility Lines From Signal Equipment Loads

Likewise, because of the varied types of equipment which can be connected to general utility lines, they should be separated from signal equipment lines.

4-4.5.3 Placement of Conduit and Wireways

Installation plans should identify high- and low-level cables for special routing and segregation (see par. 4-4.11). The use of metal conduit for signal or power lines may be necessary. Where cables and wireways are tied into junction boxes, separate boxes should be used for power and signal-equipment lines. If not, separate tie points and internal shielding between lines are needed. Open overhead wireways should be avoided, particularly in the vicinity of RF sources. Closed wireways and conduit, suitably grounded, are recommended.

4-4.6 CABLE TYPES

The principal types of cables that are available include (Ref. 10) unshielded single and multiple conductor, shielded single wire, shielded multiconductor, coaxial, unshielded twisted-pair, and shielded twisted-pair. Cables are also available with multiple shields, in many different forms, and with a variety of physical and electrical characteristics. Proper selection and application of appropriate cables for specific design requirements are highly important in preventing, controlling, and eliminating interference.

Cables are generally specified or identified according to:

a. Size
b. Number of conductors
c. Characteristic impedance
d. Attenuation
e. Shielding (single, double, triple)
f. Power rating

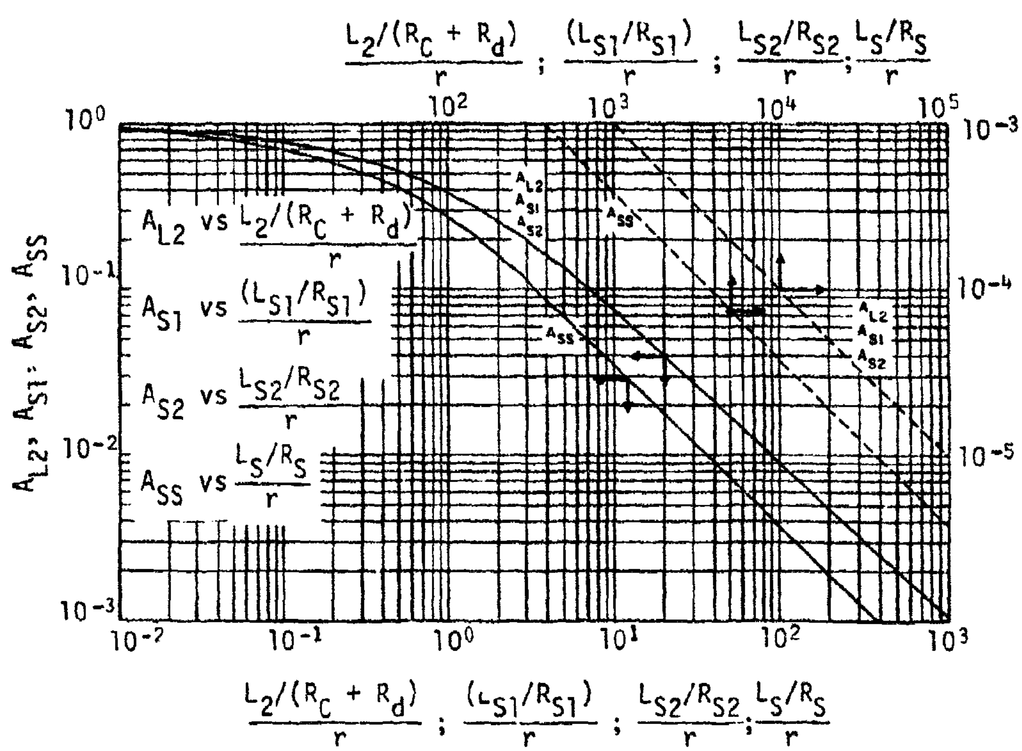


Figure 4-15. Transient Inductive and Shield Attenuation Factors

g. Maximum operating voltage including hi-pot testing

h. Type of jacket (standard black, standard gray, low-temperature black, polyethylene, fiberglass, or armor)

i. Type of dielectric (for example, polyethylene or Teflon, solid or spiral ribbon, pressurized, or unpressurized).

An excellent source of information concerning the characteristics of wire and cable is available (Ref. 11). It contains details of the more frequently used cable types including:

a. Dimensions

b. Materials used and their important characteristics

c. Power carrying characteristics

TABLE 4-2. SPACINGS AND PITCH DISTANCES FOR CABLES (Ref. 9)

Cable Type	Helix Conductor Separation δ, mm	Helix Pitch p, cm	Area A, $m^2 \times 10^{-4}$
DSGA 3	1.58	7.87	3.94
DSGA 4	1.80	8.64	4.95
DSGA 9	2.39	11.43	8.68
DSGA 14	2.74	13.21	11.52
DSGA 23	3.18	15.24	15.39
DSGA 30	3.58	17.15	19.52
DSGA 40	3.89	18.67	23.08
DSGA 50	4.24	20.32	27.41
DSGA 60	4.72	22.61	33.96
DSGA 75	5.18	24.89	41.02
DSGA 100	5.74	27.18	49.61
DSGA 125	6.43	30.99	63.31
DSGA 150	7.06	33.78	75.85
DSGA 200	8.05	38.68	98.86
DSGA 250	8.84	42.42	119.24
DSGA 300	9.50	45.72	138.11
DSGA 400	10.95	52.58	183.08
2SJ 22	0.84	3.81	1.02
2SJ 20	0.92	3.81	1.11
2SJ 18	1.07	3.81	1.30
2SJ 16	1.14	6.35	2.31
2SJ 14	1.32	6.35	2.67
DCOP 1	0.74	2.54	0.60
DCOP 1-1/2	0.99	5.08	1.60
DCOP 2	0.99	5.08	1.60
TSP	0.76	6.35	1.54
2SWA	0.76	6.35	1.54
2SWF	0.76	6.35	1.54
2SA	0.76	6.35	1.54
2SU	0.76	6.35	1.54
TTHFWA	0.79	5.72	1.43
TTRSA	0.99	6.35	2.00
2A40	1.02	3.81	1.23
2AU40	1.02	3.81	1.23
2SWU	1.22	7.62	2.95
2U	0.64	3.81	0.77
TTSU	0.97	5.72	1.75
TPNW	0.61	5.08	0.99
RG-22B/U	1.14	10.80	3.92
RG-111A/U	1.14	10.80	3.92
RG-130/U	2.39	10.80	8.92
RG-108A/U	1.41	6.35	2.10

d. Voltage ratings

e. Capacitance per unit length

f. Characteristics impedance

g. Transfer impedances.

Also included are applicable equations and discussions of important considerations in selecting cable and wire types.

4-4.7 CONNECTORS

Cable connectors are made in many styles for a multitude of power, signal, control, instrumentation, transducer, audio, video, pulse, and radio-frequency applications. They are made to fulfill special functions, and may be required to be hermetically sealed, submersion proof, and weatherproof. They are manufactured in the straight type, angle type, screw-on type, bayonet twist and lock type, bayonet screw-on types, barrier type, straight plug-in type, and push-on type (Table 4-3). In addition to having design features which will meet resistance to damage occurring during insertion and extraction under field conditions, and have a reliable and adequately low pin-contact-resistance, a connector should have shielding effectiveness characteristics which will not degrade that of the cable with which it is used.

To assure that the connector adequately shields the circuits passing through it, the connector shell must have a conductive finish, and there must be no break in the shielding through the connector/cable combination through which unwanted fields may enter. Where power circuits pass through a connector, the connector shield at the interface of the two connector halves must make positive contact before the power contacts mate, and maintain contact until after the power contacts break. The entire periphery of the shield of the cable being terminated must be bonded to the shell of the connector around the entire periphery of the cable entryway (see Fig. 4-16 (Ref. 10)). This should be done by soldering or metal forming; "pig-tailing" or bonding with conductive epoxy is unsatisfactory. Likewise, the entire periphery of bulkhead connectors must be bonded to the bulkhead or chassis in which it is mounted.

To assure that a good high conductivity path exists between mating pins, they should be plated with a material which has good conductivity and resists tarnishing and corrosion. Two layer plating, hard gold over ductile nickel (elongation of not less than 5%) has been found to be very effective over a copper alloy base metal. Where the base metal is a nickel iron alloy, as often used in hermetically sealed conductors, copper is plated on the contacts as a preliminary step. A fine microfinish will increase corrosion resistance and reduce friction as well.

MIL-STD-454, *Standard General Requirements for Electronic Equipment:* Requirement 10 sets forth requirements for connectors to be used in military equipment. This document should be used to locate specifications applicable to connectors which are under consideration for inclusion in equipment design.

4-4.8 CABLE APPLICATION

4-4.8.1 General

The choice of cable is dictated by the operating signal or power level, frequency range, susceptibility level, and physical isolation. While it is not feasible to

TABLE 4-3. CONNECTOR APPLICATION SUMMARY

Connector Series	Coaxial Cable Size	For RG-/U Cables	Disconnect Style	Voltage Rating	Characteristic Impedance	Freq. Range	Method of Assembly
N	Medium & Large	5,6,8,9, 10,11,12, 13,14,17, 18	Screw-on type	500 V peak	50 ohm 70 ohm (constant)	up to 10 GHz	Manual
GR-874	Medium & Large	8,9,29,55, 58,58A,59, 62,116	Push-on type	1500 V peak	50 ohm	up to 7 GHz	Manual
C	Medium & Small	8,9,10,12, 14,55,58	Bayonet Lock type	1000 V peak	50 ohm	—	Manual
UHF	Medium & Small	8,9,10,11, 12,13,55, 58,62,63, 65,71	Screw-on type	500 V peak	(nonconstant)	up to 200 MHz	Manual
LC	Large	17,18	Screw-on type	5000 V peak (modified to 10 kV)	50 ohm	—	Manual
BN	Medium & Large	8,9,10,17 18	Screw-on type	5000 V peak	50 ohm (constant)	—	Manual
BN	Small	55,58,59, 62,71	Screw-on type	250 V peak	(nonconstant)	up to 200 MHz	Manual
BNC	Small	55,58,59, 62,71	Bayonet lock type	250 V peak	50 ohm (constant)	up to 10 GHz	Manual & Crimp-on
Subminiature	Subminiature	174	Screw-on & Push-on types	—	50 ohm 75 ohm (constant)	—	Crimp-on

set specific rules for cable selection without making an analysis of signal levels and waveforms, the following general rules are suggested:

a. Use unshielded wire for external power circuits such as 115 V ac or 28 V dc. An exception to this rule is when the power source itself generates substantial EMI, such as with an inverter or an alternator rectifier that is not adequately filtered at the source; generally, leads to a power source regulator must be shielded (see par. 5-6). Ac line wires should be twisted.

b. Use shielded wire for multiple-ground, audio frequency, or power circuits. Single-conductor, single-shield cable can be used for low-frequency instrumentation applications using ground return circuits. It is effective when signals to be transmitted are of moderate levels, and a good low-impedance system ground is available.

c. Use twisted-pair for audio-frequency circuits grounded at a single point and for internal power circuits.

d. Use shielded twisted-pair for single-point ground circuits and multiple-ground circuits where maximum low-frequency isolation is required.

e. Use coaxial cable for transmission of RF pulses, high-frequency applications, and where impedance match over a broad frequency range is critical,

f. Special measures may be required with short rise time, for which solid copper transmission lines or triaxial cables are used (Fig. 4-17).

4-4.8.2 Multiconductor Cables

Multiconductor cables are of many types. They may contain groups of unshielded lines, coaxial lines, or multiple conductor shielded combinations. For maximum effectiveness of the shields it is essential that each be insulated individually. The various conductors within a single shield should be used preferably in pairs and should operate at comparable voltage and current levels.

Initially unshielded cable may be shielded by routing it through continuous flexible or rigid metallic conduit. For economical reasons, rigid conduit is generally of aluminum; but from the standpoint of shielding high-level power circuits, galvanized steel conduit is more effective at power frequencies. Continuous, high-μ materials are also available for use with audio-frequency signal cables.

Flexible conduits for high- and low-voltage shielding usually consist of flexible metal hoses over which are wound one or more layers of braid. Nonconducting coverings sometimes are used over the braid. These coverings provide watertightness and/or added mechanical protection. If applied tightly, they may decrease contact resistance between wires comprising the braid, thereby improving shielding effectiveness. Such coverings should be reasonably rugged and not subject to physical and chemical attack by substances with which they come into contact. They should maintain their desirable characteristics over the anticipated range of operating temperatures.

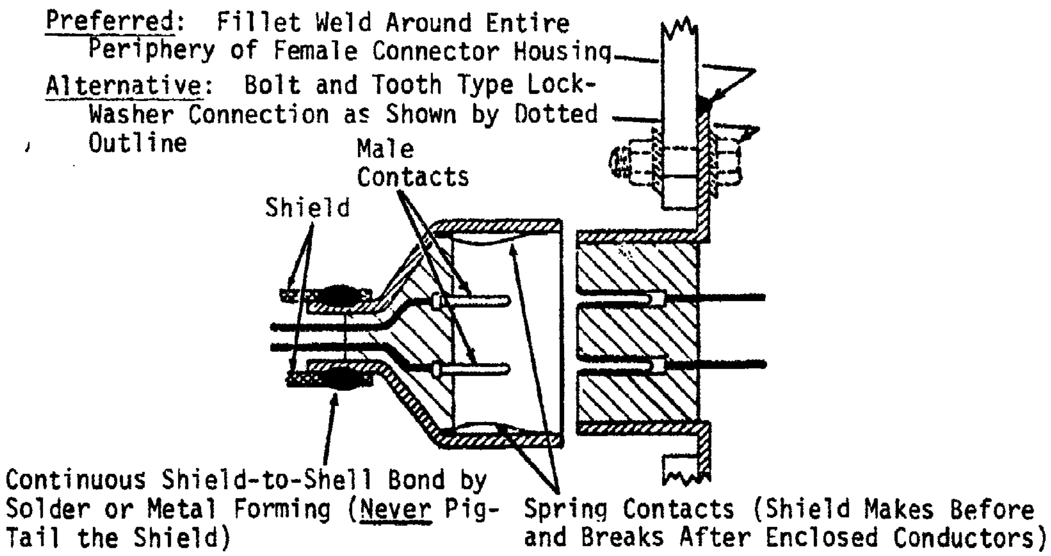


Figure 4-16. Shield Termination for Connectors (Ref. 10)

Shielded conduit is used for many diversified purposes, such as:

a. To shield wires and cables electrically that otherwise would radiate interference

b. To provide a channel through which wires and cables may be pulled or pushed for installation or replacement in inaccessible places

c. To protect insulated wires and cables against mechanical damage, for example, chafing and abrasion

d. To keep foreign matter (moisture, oil, grease, gasoline) away from electrical conductors or their insulation

e. To facilitate dissipation of heat for protection of insulation.

To be effective, a flexible shielding conduit should be:

a. An effective shield against electrical interference over the entire range of frequencies under consideration

b. Reasonably flexible and capable of being bent to a small radius

c. Rugged enough to withstand considerable abuse and prolonged vibration without serious impairment of either its electrical or mechanical properties

d. Watertight and airtight. The coverings used with it should be immune to attack from lubricants, coolants, antifreeze, and fuels.

e. Capable of withstanding ambient temperatures likely to be encountered.

4-4.9 SHIELD GROUNDING

In accordance with the discussion of grounding principles in par. 3-3.4.4, shields of balanced circuits operating at frequencies up to about 100 kHz should be grounded at only one end. Furthermore, the shield should be insulated so that the single point grounding practice can be observed. On the other hand, if the cable length is more than about 1/10 of a wavelength for any frequency of concern and an established ground "plane" exists, the shield should be bonded periodically to limit the possibility of large currents or voltages being induced on the external surface of the shield. MIL-STD-1310 requires that cable armor be bonded at each metallic bulkhead or equipment space penetration. On the other hand, shielding conduits need be bonded only at points 3 to 5 ft from each end. Bonding should be as direct as possible (see par. 4-7.2). The outer shield of coaxial cables must be grounded to the chassis at both ends; "pig-tail" type connections should be avoided. On the assumption that grounding techniques have been employed, the following are suggested as guidelines for good signal cable practice:

a. Shields should not be used for signal return circuits.

b. All signal circuits, including signal ground returns, should be individually shielded and have insulating sleeves or coverings over the shields. Balanced signal circuits should use twisted pair or

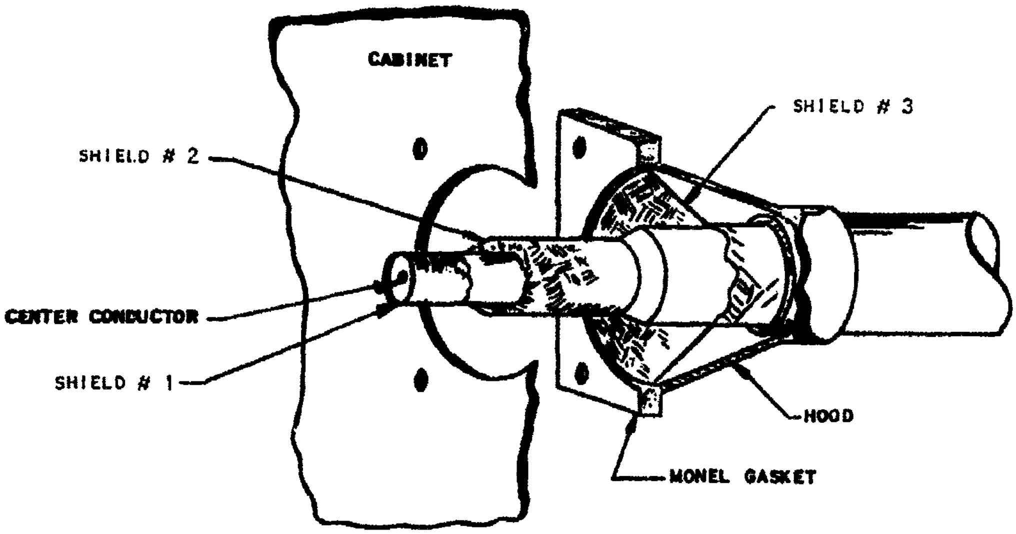


Figure 4-17. Triaxial Cable Application

balanced coaxial line with a common shield. Where multiconductor twisted pair cables are used which have individual shields as well as a common shield, all shields should be insulated from one another within the cable.

c. Coaxial cables should be terminated in their characteristic impedance and grounded at both ends.

d. When electronic and electrical systems are distributed over a large area, multipoint shield grounding is usually effective for interference control at the higher radio frequencies. The multipoint approach allows short ground connections, permits a low-impedance ground return circuit, and improves the effectiveness of filter installations. At low frequency in low-level systems with audio or servo amplifiers, single-point grounding may be necessary since, when a shielded cable in a sensitive circuit is grounded at both ends, power-frequency currents in the ground plane can induce audio-frequency interference.

e. Coaxial cables carrying high-level energy should not be bundled with unshielded cables or shielded cables carrying low-level signals.

4-4.10 CONNECTOR GROUNDING

Where shields are grounded at both ends of a cable, the objective is to obtain a continuous equipment enclosure shield as shown on Fig. 4-18. To accomplish this, wires and coaxial cables must be terminated properly at the connector. The connector itself must be grounded to its mounting by a clean metal-to-metal contact. A shielded cable should not be run into a completely sealed box and grounded internally; rather the shield should be run well inside the connector and be bonded to the connector shell. The arrows on Fig. 4-19 show the path that any signal or interference, that is picked up on the outer surface of the shielding, must follow to return to ground. The currents around the loop generate a field in the enclosed box, as do coupling loops used with resonant cavities. Fig. 4-20 illustrates the correct method of treating the shield.

For a low-impedance RF connection, the shortest length of connecting strap or jumper that is mechanically practical should be used, and the bonding procedures in MIL-STD-1310, MIL-B-5087, and MIL-E-45782 should be followed (see also par. 4-7.2). Great care must be taken at connectors if impedance characteristics and shielding integrity are to be maintained. A shielding shell should be used to shield the individual pins of a connector; a well-designed connector has a shielding shell enclosing its connecting points (Fig. 4-21). The shell of multipin connectors should be connected to the cable shield. Coaxial lines should terminate in shielded pins. The use of pigtail connections for the outer conductor of coaxial lines is undesirable since it permits RF leakage. Shields should be grounded on both sides of a connector to avoid discontinuity. If this is not possible, the shield should be carried across the connector through a connector pin. Grounding a number of conductor shields by means of a single wire to a connector ground pin should be avoided, particularly if the shield-to-connector or connector-to-ground

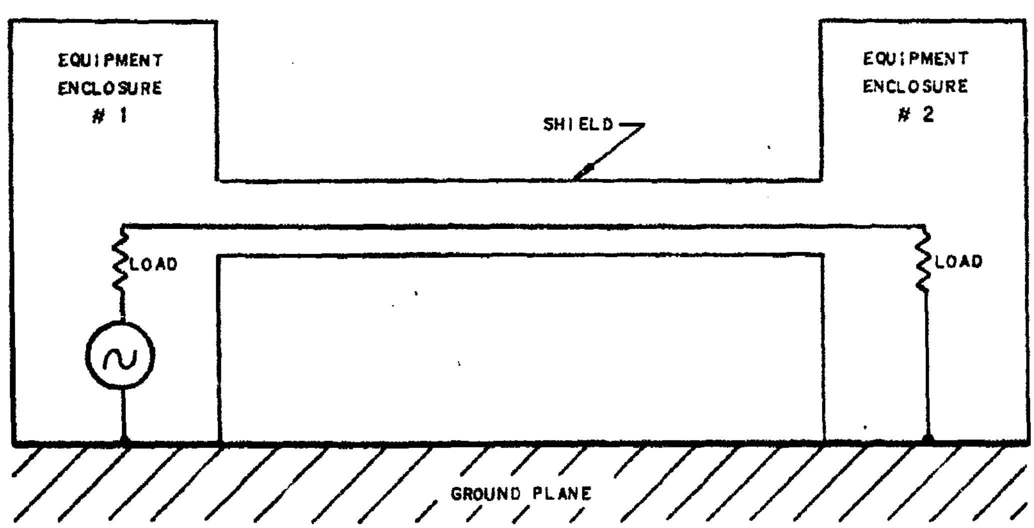


Figure 4-18. Continuous Equipment Enclosure Shield

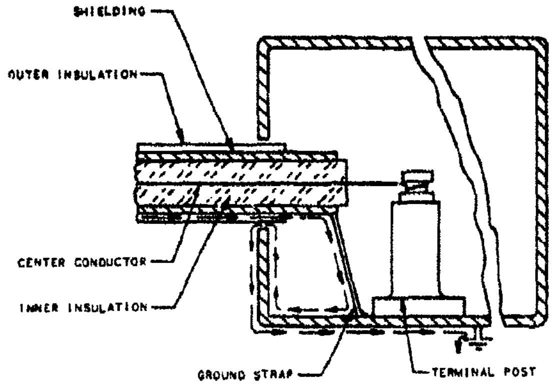


Figure 4-19. *Incorrect* Method of Introducing Shielded Cable

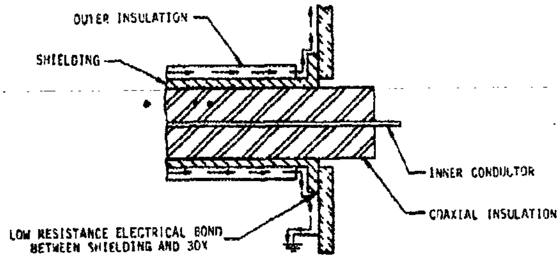


Figure 4-20. Correct Method of Introducing Shielded Cable

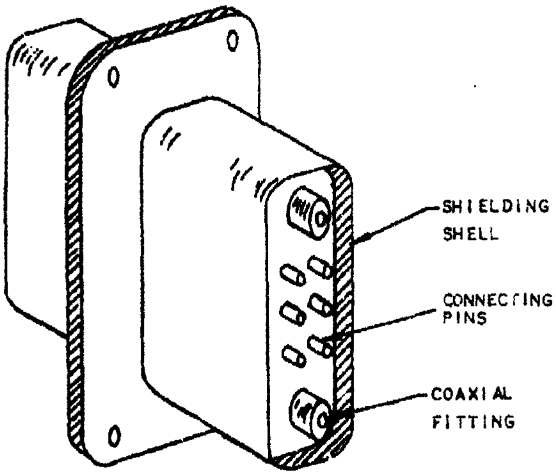


Figure 4-21. Connector With Shielding Shell Enclosure For Connecting Points

lead length exceeds 1 in., or where different circuits that may interact are involved. Such a ground lead is a common impedance element across which interference voltages can be developed and transferred from one circuit to another.

For grounding the outer shield on a cable, a backshell type of connector with a continuous contact around the shield is the most satisfactory. This may be accomplished by means of tapered rings whereby pressure is applied to a continous metal-to-metal contact between shield and connector. Typical arrangements are shown schematically in Fig. 4-22. In Fig. 4-22(A) the shield is flared and held between a pair of tapered rings. In Fig. 4-22(B) a compressible "iris" is used to clamp the shield.

If the cable also contains wires which are shielded individually, the individual shields can be brought out of the cable assembly just before the connector — bent back over the outer cable shield and held in the same manner as the outer shield.

In a similar arrangement an elastomer pressure seal can be made with a compressible "O"-ring contacting a neoprene or other waterproofing jacket located over the shield.

Fig. 4-23 shows a backshell of the type used for fitting over an unshielded connector. Here again the outer shield, which may be separate from the cable, is fastened completely around the shield by means of a crimping or soldered ring or other means. The mating surface with the jack is rendered EMI-proof by the use of an EMI gasket (see par. 6-4.5.4). Also, an EMI gasket is shown between the jack and the panel on which it is mounted.

Also note that connectors are available in which small low-pass filters are incorporated so as to act in series with individual pin connections (see par. 4-5.5.3).

Electric plugs and rece cles usually are mounted on the front or rear of equipment chassis, or on the mounting base. If ric receptacles are on the front of the case, the plu hould be separate units. Shield grounds should be m e in accordance with Fig. 4-24. If electric plugs and ceptacles are placed at the rear of the case, at least one unit should be attached securely to the case or chassis; the other should either be separate from or attached securely to the mounting base. Methods of grounding cable shields shown on Fig. 4-25 are *not* recommended.

To prevent discontinuity of the shield because of possible disconnect at intermediate connectors, shields should be grounded to the structure on both sides of the connector. Where this is not possible, the ground should be carried across the connector, or through a conductor pin, to ensure continuity.

Coaxial fittings should be kept tight at all times, not only to provide a good impedance match but to eliminate loose connections that might result in possible rectification of interference energy at the fittings. Shielding or bonding clamps that may be a part of the fittings also should be kept tight. Soldered fittings are recommended, particularly at terminations of shielding and braid.

When any high level pulse lead passes through a connector in a way that involves a discontinuity in the coaxial structure of the shielded lead (for example, when a pulse lead center-conductor and shield are attached to separate pins of a connector), an extremely low-impedance circuit should be provided for the ground lead of the shield in the connector. If this is not done, the entire shield along the lead may radiate or conduct interference. A suitable low-impedance connection can be obtained by the use of an extremely heavy pin, or by the use of several pins in parallel, preferably distributed circumferentially about the pin attached to the pulse lead center-conductor. The pulse lead and nearby susceptible leads should be shielded to prevent transfer to other leads.

4-4.11 CABLE SEGREGATION AND HARNESSING

Cable routing should be planned in order to isolate sensitive (susceptibility prone) wires and cables from possible interference carrying conductors. Traditionally, this has been done by arranging cables

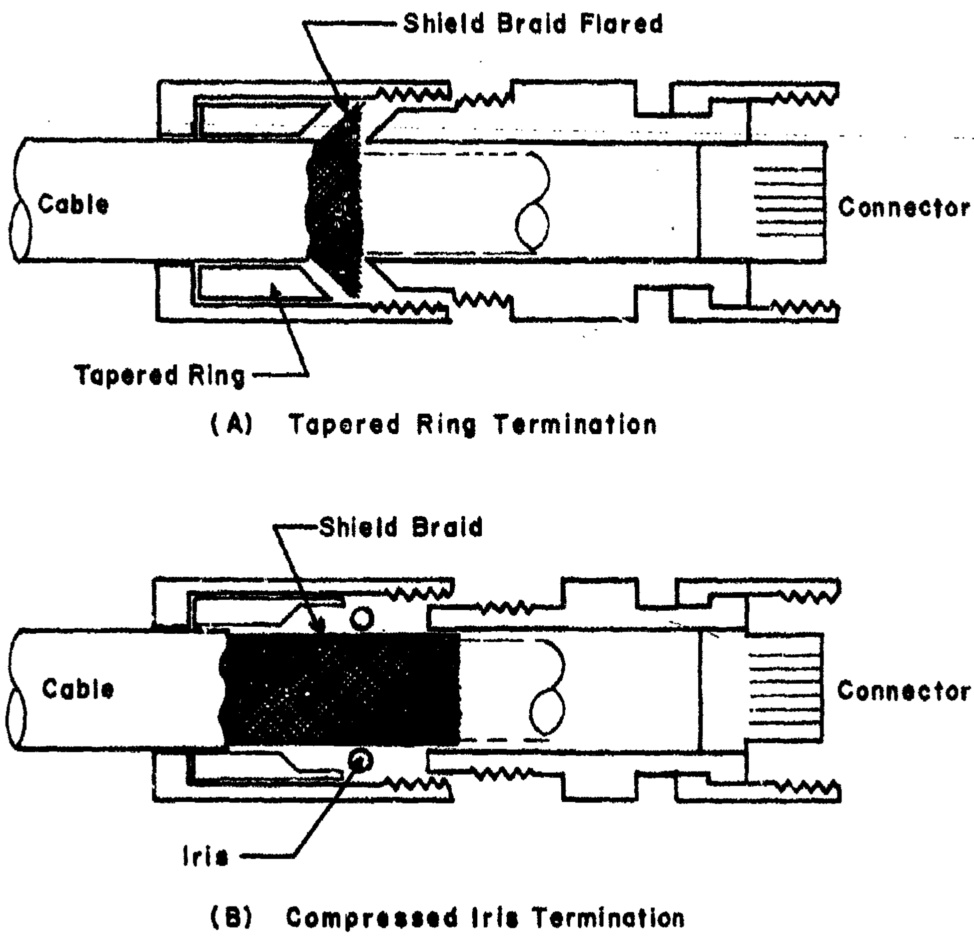


Figure 4-22. Connector Backshell Arrangement for Terminating Cable Shields

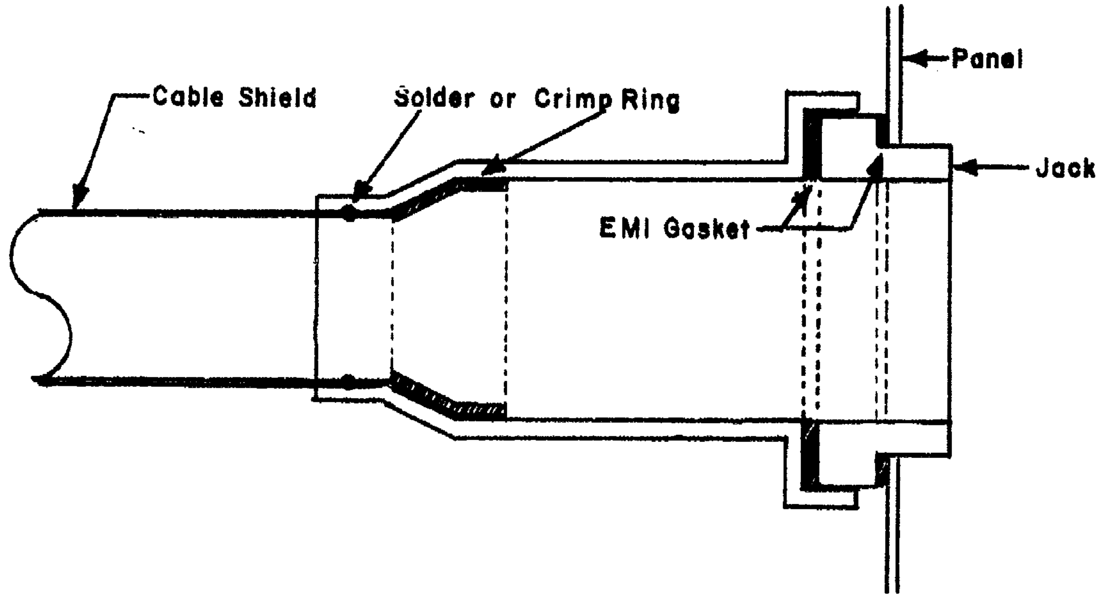


Figure 4-23. Backshell Arrangement for Terminating Cable Shields

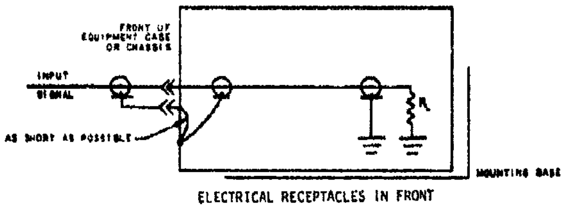


Figure 4-24. Cable Shield Bonding

into compatible groups. MIL-STD-461 identifies the following five major categories:

a. Power wiring including main power distribution circuits

b. Secondary power including low voltage power and lighting circuits, servo and synchro circuits, and secondary dc power with voltages up to 5000 V

c. Control wiring including that to relays or other intermittent operating devices involving switching

d. Sensitive wiring including circuits carrying signals such as audio, digital data, analog control, and demodulator output.

e. Susceptible wiring including wiring to electro-explosive, antiskid, spoiler actuator or "safety of mission" devices; and coaxial cables to receiving antennas, fire warning, fuel quantity, and liquid oxygen indicator devices.

A typical installation is shown on Fig. 4-26. In some installations, rows and tiers of cable trays are used to facilitate the desired physical cable separation. Detailed planning is required prior to installation to ensure that the best overall routing is achieved.

While arrangements of this kind are feasible, they must be used with care, since: (a) it is not always possible to classify easily a given cable into only one of these groupings; and (b) one must establish appropriate minimum spacings between all of these cable groups in the installation, and it may not be possible to maintain these spacings everywhere.

Another classification system (Ref. 10) includes 7 wire classes and 9 circuit types as shown in Table 4-4. Each circuit is assigned a classification on the basis of its being most similar to one of the following classes:

a. Power and Control Circuits (Class I):

(1) Dc Power Circuit. A dc circuit using current more than 2 A.

(2) Dc Control Circuit. A dc circuit which uses less than 2 A.

b. Dc Reference Circuit (Class II). A dc circuit requiring critical tolerances on the voltage or current.

c. Ac Circuits (Class III and IV). Any circuit which is supplied by ac power sources.

d. Ac Reference Circuit (Class III and IV). An ac circuit in which a single-phase line is used to supply critical voltages or frequencies.

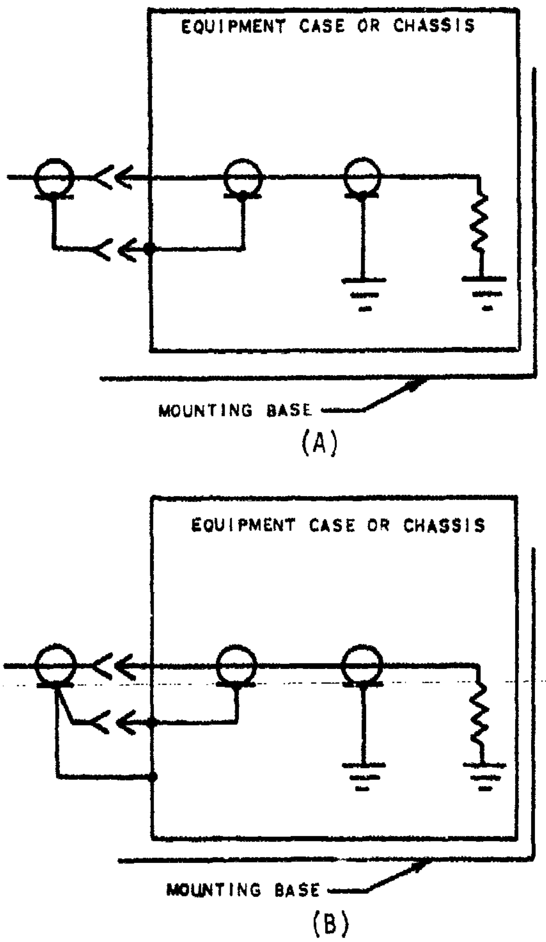


Figure 4-25. Poor Cable Bonding

e. AF Susceptible Circuits (Class II). Those circuits whose performance may be degraded in the presence of an undesired audio signal. The voltage or current of these circuits will normally be less than 1 V rms or 200 mA rms. Direct current reference circuits are considered as audio-susceptible circuits.

f. AFI Circuits (Class III or IV). Those circuits operating below 15 kHz and whose amplitudes will normally be greater than 1 V rms or 200 mA rms.

g. RF Susceptible Circuits (Class V). Those circuits whose performance may be degraded in the presence of an undesired RF signal.

h. RFI Circuits (Class VI) are as follows:

(1) Narrowband Circuits. Those RF-interference circuits in which the signal levels exceed the following values with respect to 50 ohms:

(a) −45 dBmW at 150 kHz reduced to 20 dB per decade of frequency to −75 dBmW at 5 MHz

(b) −75 dBmW from 5 to 25 MHz

(c) −75 dBmW at 25 MHz increased to 10 dB per decade of frequency to −45 dBmW at 1 GHz

(d) −45 dBmW above 1 GHz

(2) Broadband Circuits. Those RF-interference circuits which conduct pulse information or transient disturbances such as caused by relay actuation, switch contacts, or clock pulses.

(3) Antenna Circuits (Class VII). Those RF circuits connecting subsystems or equipment to system antennas.

In layout wiring and routing, the maximum practical separation must be maintained in accordance with the general rules described in the paragraphs that follow. A 2-in. separation between wires in different classes is a minimum design goal.

Route dc power and control circuits (Wire Class I) in a separate wire group. The ac power and control circuits are divided into groups when the system uses more than one source of electrical power. The following general rules apply:

a. The ac wiring supplied by different power sources must never be routed together.

b. The ac wiring from different ac power sources must never be connected to a subsystem or equipment, unless there is a specific design and installation.

c. Route reference and susceptible circuits separately from power and interference circuits except as follows:

(1) Reference dc and audio-susceptible circuits may be routed together in the same way provided that proper isolation is maintained through shields, shield terminations, and connectors.

(2) Reference ac circuits may be routed with ac power circuits if the rules governing multiple power sources are not violated, and they are not classified as susceptible.

RF-susceptible circuits may be routed with Class II circuits if the RF-susceptible circuit is not a source of interference to the other circuits.

Where a specific layout is given, computer programs are available to identify specific cases where coupling will be sufficient to cause operational degradation (see Chapter 6). To use the programs, one must put into the computer all cable and location data. While this procedure can provide high accuracy, it is more involved than the use of the cable categorization technique.

4-5 FILTERS

4-5.1 INTRODUCTION

Though proper circuit design has been emphasized as an appropriate means for controlling interference,

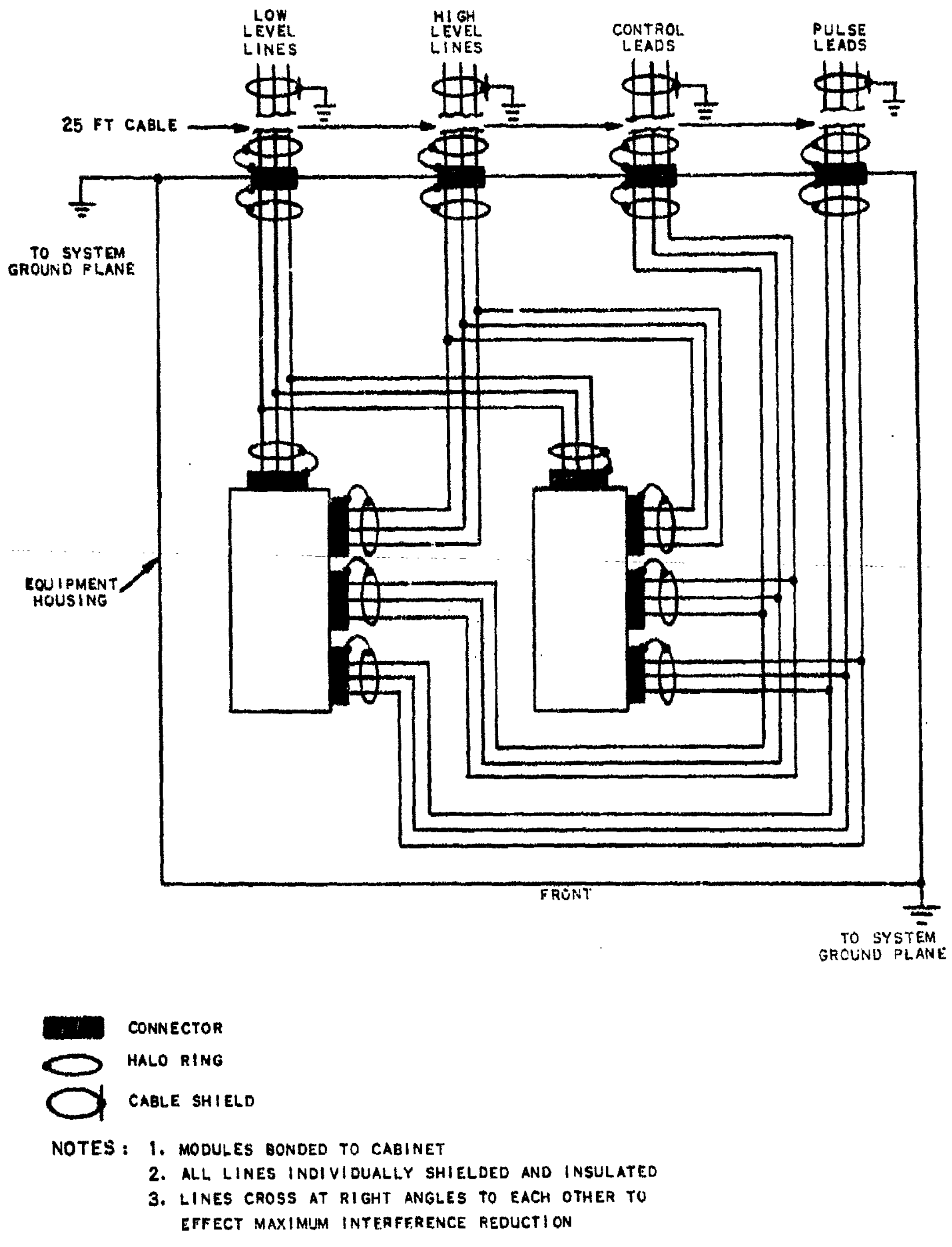


Figure 4-26. Typical Cabling Methods

it is not always possible or advisable to control it in this manner. Thus, for technical as well as economic reasons, filters should be viewed as legitimate tools of interference control.

Filter theory is highly developed. A theoretical survey on lumped element filters is given in Chapters 7 and 8 of Ref. 12, and transmission line, waveguide, and cavity filters will be found in Chapters 22 and 23 of Ref. 12. A filter is a linear two-port device modeled by the block diagram shown in Fig. 4-27. The source, input, and load voltages E_S, E_{S1}, and E_L and the source and load currents I_S and I_L are complex quantities representing sinusoids at frequency f. Given E_S, Z_S, and Z_L, the behavior of the filter is determined from the voltage transfer function $H(f)$

$$H(f) = E_L(f)/E_{S1}(f)\text{, dimensionless} \quad (4\text{-}4)$$

and the input impedance seen looking into the input terminals of the filter.

Manufacturers normally give filter characteristics for fixed source and load impedances, usually 50 ohms as measured in accordance with MIL-STD-220. However, actual circuit impedances can vary widely, especially with frequency, and will influence performance significantly. The consequences of this are discussed in par. 4-5.2.1.6.

The transfer function is a complex quantity specifying the gain, which is the ratio of the magnitudes of E_L and E_{S1}, and the phase shift from input to output. The reciprocal of the gain is attenuation, a quantity frequently used to characterize the behavior of a filter. Because the output is not fully determined by the transfer function but is affected by the input impedance as well, a preferred filter characteristic is the insertion loss defined as

$$IL = 20 \log \left(\frac{E_{L1}}{E_{L2}}\right), \text{ dB} \quad (4\text{-}5)$$

where E_{L1} and E_{L2} are the magnitudes of the load voltage, respectively, with and without the filter in the circuit. The insertion loss and the attenuation are identical if the input impedance of the filter and the load impedance are equal. The passband of a filter is the frequency range in which there is little or no attenuation. The stopband is the frequency range in which attenuation is desired.

Filters can be classified grossly according to the relative positions of the passband and stopband in the frequency spectrum. There are four classes — low-pass, high-pass, bandpass, and band-reject — and the discussions to follow deal with these classes. Attenuation as a function of frequency for each of the classes is shown in Fig. 4-28 (Ref. 13).

Filter synthesis for specified behavior using lumped parameters has been, and still is, difficult. In EMC work, precise attenuation characteristics generally are not demanded and an adequate approach is to search through a collection of known structures to obtain one that will perform the task. Sometimes a simple bypass capacitor or series inductor will be adequate. Otherwise, the more elaborate L, T or Pi (π) structure is required.

Certain guidelines are helpful in deciding what type of filter circuit to apply in any given instance. For example, if it is known that the filter will connect to relatively low impedances in both directions, then a circuit which contains series filter elements may be

TABLE 4-4. CIRCUITS AND WIRE CLASSES (Ref. 10)

CIRCUIT TYPE	WIRE CLASS
DC Power and Control	I
DC Reference	II
AC Power and Control	
(a) Left Hand Bus	III
(b) Right Hand Bus	IV
AC Reference	
(a) Left Hand Bus	III
(b) Right Hand Bus	IV
Audio-Frequency-Susceptible	II
Audio-Frequency Interference	
(a) Left Hand Bus	III
(b) Right Hand Bus	IV
RF-Susceptible	V
RF-Interference	VI
Coaxial Cables	VII

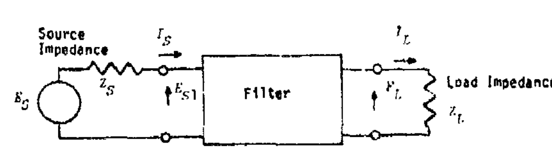


Figure 4-27. Linear Two Port Model of Filter

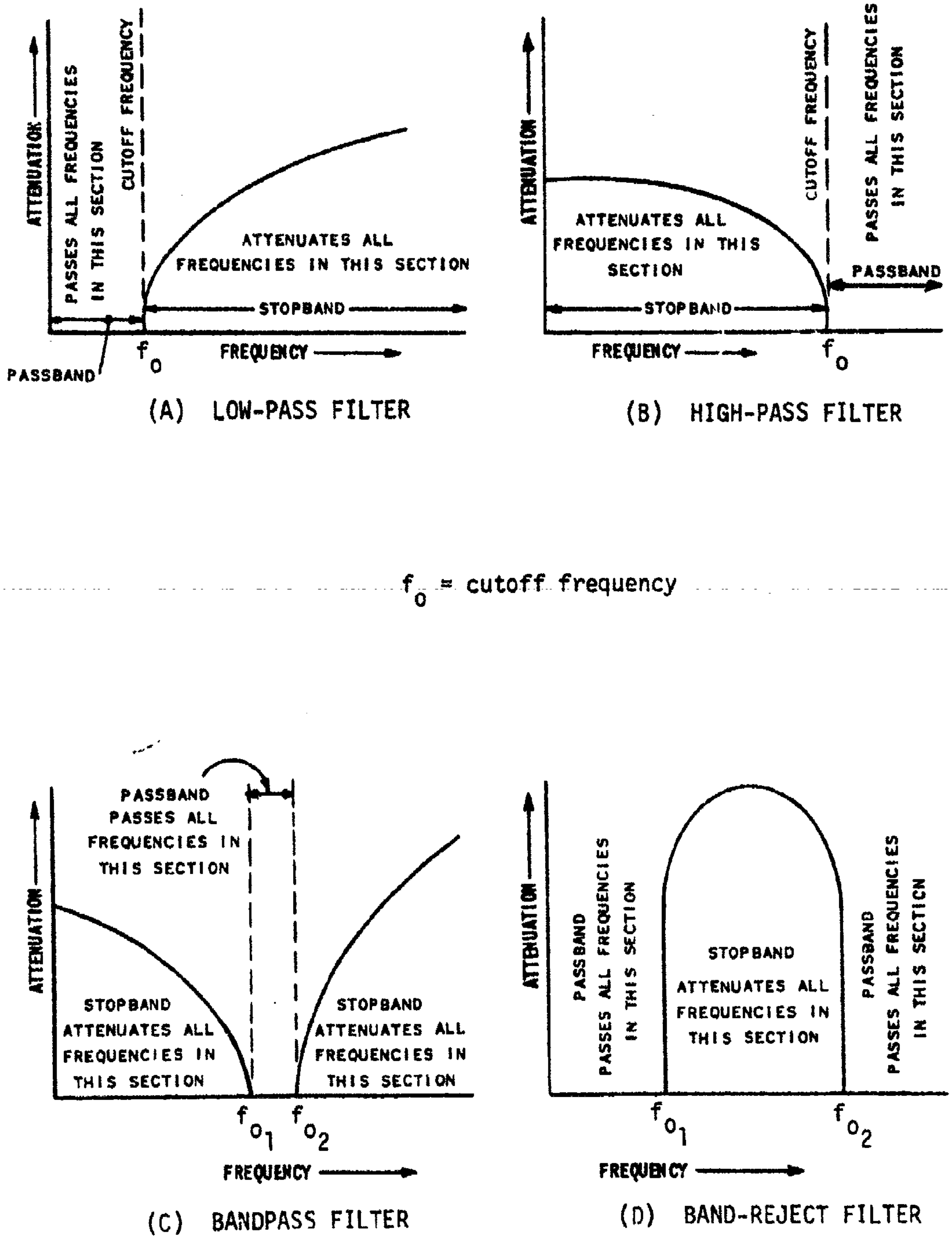


Figure 4-28. Filter Attenuation as a Function of Frequency

best (a T-circuit, for instance). Conversely, a high-impedance system calls for a π-filter. If the filter is connected between two severely mismatched impedances, then an asymmetric filter circuit, such as two-L-section elements, can be used in which the series element faces the low-impedance side of the system.

4-5.2 LUMPED ELEMENT FILTERS (Ref. 13)

4-5.2.1 Low-pass Filters

Low-pass filters are commonly used on ac or dc powerlines. They are used also in amplifier circuits and transmitter output circuits to attenuate harmonics and other spurious signals.

4-5.2.1.1 Shunt Capacitor Filters and General Capacitor Characteristics

In the ideal resistance terminated circuit of Fig. 4-29, the insertion loss IL of a shunt capacitor is:

$$IL = 10\log[1 + (f/f_o)^2] \text{ , dB} \tag{4-6}$$

where

$f_o = 1/(\pi RC)$, Hz
f = frequency, Hz
R = source or load resistance, Ω
C = filter capacitance, F

An actual capacitor contains both series resistance and inductance in the leads and connections to the active element, and shunt resistance due to dielectric losses, as shown in Fig. 4-30. The inductance causes a capacitor to exhibit a self-resonant frequency above which the capacitor behaves like an inductive reactance as illustrated in Fig. 4-31. Note the effect of changing capacitor leadlength on this self-resonant frequency.

Metalized paper capacitors, while small in physical size, offer poor RF bypass capabilities because of high resistance contact between the leads and the capacitor metal film. They are subject to dielectric punctures which self-heal by burning away the metal film. The standard wound aluminum foil capacitor may be employed as a radio frequency bypass in the frequency range up to 20 MHz.

Mica and ceramic capacitors of small values are useful up to about 200 MHz. If the plates are round as in a ceramic disc, the capacitor will remain effective to higher frequencies than one of square or rectangular construction. A ceramic capacitor element is affected by operating voltage, current, frequency, age, and ambient temperature. The amount the capacity varies from its nominal value is determined by the composition of the ceramic dielectric which can be adjusted to obtain a zero or negative temperature. In obtaining one characteristic, other characteristics may become undesirable for certain applications. For example, when the dielectric composition is adjusted to produce minimum size capacitors, the voltage characteristic may become so negative that its capacity is reduced by 50% at full operating voltage, and high ambient temperature may cause an additional sizable reduction. In addition, for some materials, the dielectric constant of the materials used may decrease with life by 25% of the original value.

Capacitors of short-lead construction and feed-through capacitors are designed to reduce inherent end-lead inductances. Fig. 4-32(A) (Ref. 14) shows that connecting to the short lead has a tendency to distribute the lead inductance between the two circuits connected to the capacitor and reduces the inductance exclusively in series with it. This principle is carried to the extreme in the feed through arrangement shown in Fig. 4-32(B). An equivalent circuit is shown in Fig. 4-32(C). Feedthrough capacitors come in a variety of shapes and sizes. Fig. 4-32(D) shows one example which can be mounted directly in a hole in a partition or cabinet wall. A mounting of this type is essential to obtain maximum usefulness of this type

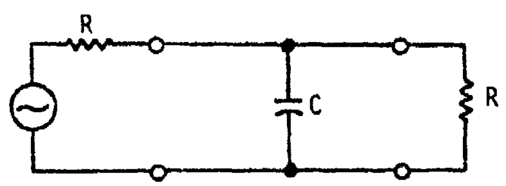


Figure 4-29. Capacitor Low-pass Filter

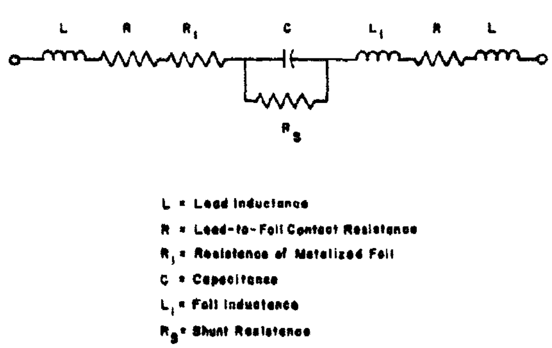


Figure 4-30. Metalized Capacitor Equivalent Circuit

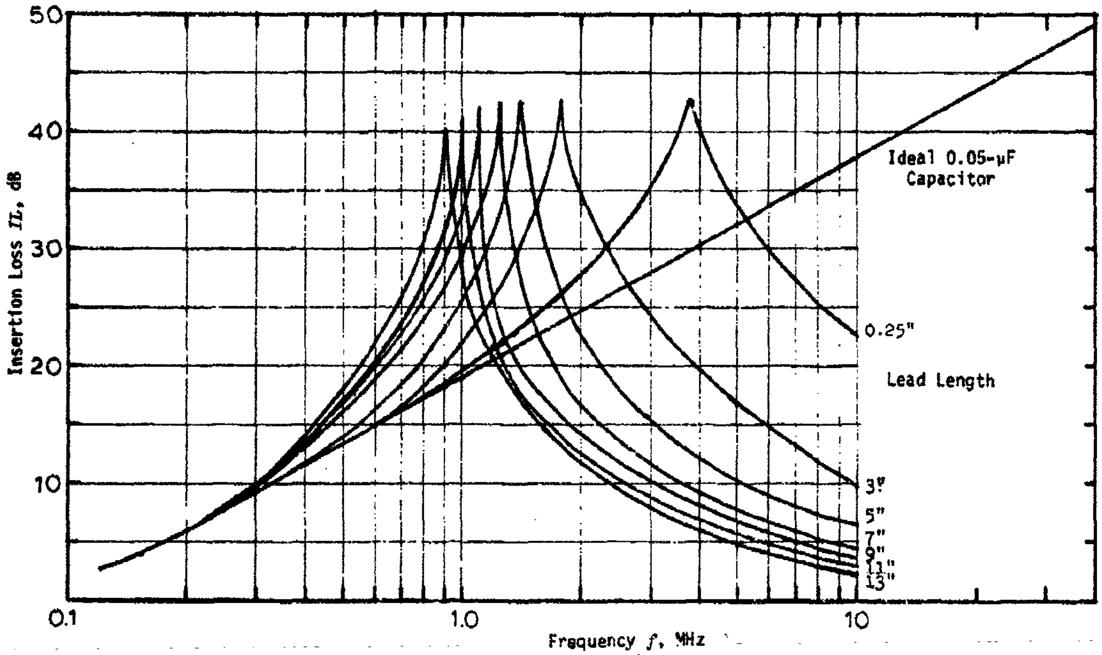


Figure 4-31. Insertion Loss of a 0.05-μF Aluminum Foil Capacitor

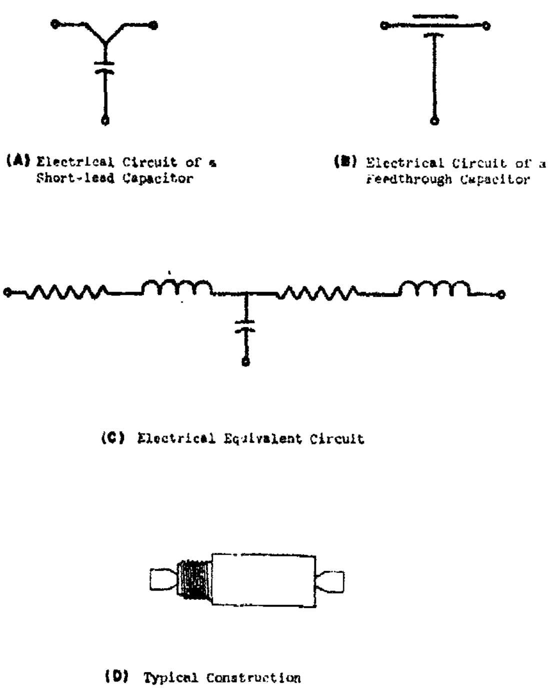


Figure 4-32. Three-terminal Capacitor

of component since then the separate circuits on input and output sides are otherwise isolated from each other.

The short-lead capacitor is suited ideally for EMI suppression in the frequency range of 1 to 1000 MHz. Feedthrough capacitors are available with a resonant frequency well above 1 GHz. The feedthrough current rating is determined by the stud diameter. Fig. 4-33 indicates the construction details of a feedthrough unit.

Metalized Mylar capacitors offer compact design and good reliability. Their dissipation factor is very low, and lead length generally can be kept short to improve high frequency performance.

Wet-type electrolytic capacitors are used for dc filtering and sometimes used in EMI filters. They are single polarity devices, but their high dissipation factor or series resistances make them poor RF filters. An RF bypass capacitor should be placed across the output of dc supplies using electrolytics. The dissipation factors of electrolytic capacitors increase, and their capacitances decrease with age.

Tantalum electrolytic capacitors provide a large value of capacitance in a small space. They are sensitive to over-voltages and are damaged by reverse polarity. The dissipation factor is considerably higher

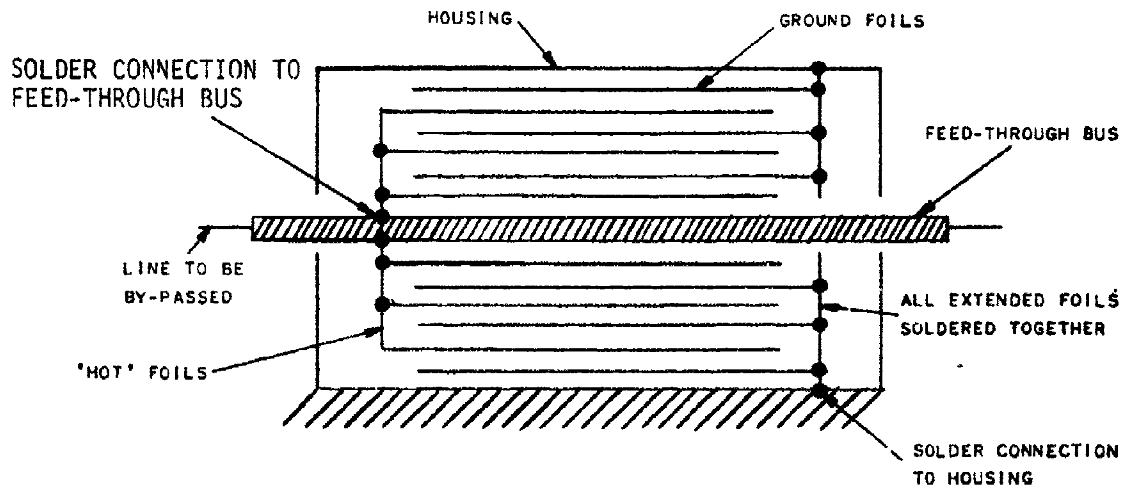


Figure 4-33. Typical Feedthrough Capacitor Construction

than for Mylar or paper capacitors, and high frequency characteristics are poor.

Capacitors for 120 V ac applications should be rated at 400 V dc to be suitable for ac use. A unit of Mylar and foil or of paper-Mylar and foil is recommended. The dissipation factor is low, and high frequency performance is good. For 240 V ac applications, an oil-impregnated paper and foil unit is recommended.

4-5.2.1.2 Series Inductor Filters, and General Inductor Characteristics

Another simple form of low-pass filter is an inductor connected in series with the interference carrying conductor (Fig. 4-34). Its theoretical insertion loss IL is given by the relationship:

$$IL = 10 \log \left[1 + \left(\frac{f}{f_o}\right)^2\right], \text{ dB} \qquad (4\text{-}7)$$

where

$f_o = R/(\pi L)$, Hz

L = filter inductance, H

In practice, inductors contain distributed capacitance between turns and exhibit a self-resonant frequency. Above self-resonance, it appears as a capacitive reactance, with the interwinding capacitance being dominant.

Filter inductors are usually toroidal, wound on cores of powdered iron, molybdenum permalloy, or ferrite material. The size of the core is determined by required inductance and current rating. The magnetic flux should not drive the core to more than 50% of magnetic saturation. Distributed capacitance effects may be reduced by a careful arrangement of turns. In some cases, two or more coils wound on separate cores are connected in series to raise the self-resonant frequency.

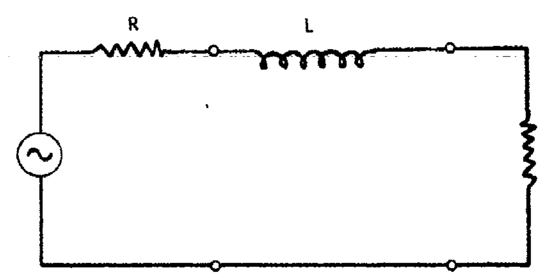


Figure 4-34. Inductor Low-pass Filter

4-5.2.1.3 Low-pass L-Section Filters

In single-element filters, out-of-band falloff rate is only 6 dB per frequency octave (20 dB per decade). A two element filter in an L configuration has a falloff rate of 12 dB/octave.

The two possible low-pass arrangements are shown in Fig. 4-35. The theoretical insertion loss for the L-

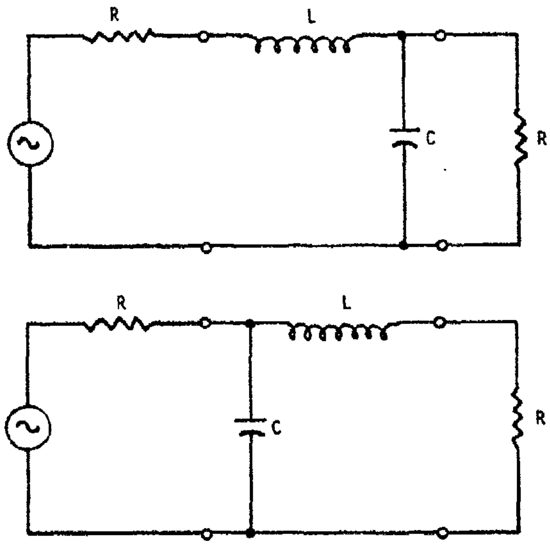


Figure 4-35. Low-pass L-Section Filter

section filter is independent of the direction of inserting it into the line, if source and load impedances are equal and is given by

$$IL = 10\log\left[1 + \left(\frac{f}{f_o}\right)^2 D^2/2 + \left(\frac{f}{f_o}\right)^4\right], \text{dB} \quad (4\text{-}8)$$

where

$D = (1 - d)/\sqrt{d}$

$d = L/(CR^2)$ = damping ratio, d'less

$f_o = 1/(\pi\sqrt{2LC})$, Hz

When source and load impedances are not equal, the greatest insertion loss will be achieved when the capacitor shunts the higher impedance. The "damping ratio" d relates the magnitudes of the filter elements to the magnitude of the source and load impedance. If it is equal to 1 (ideal damping), the squared-frequency term cancels from the insertion loss equation and the most abrupt transition from the passband to the stop band is produced. Corresponding to a Butterworth filter design, Eq. 4-8 is plotted in Fig. 4-36 for the case $d = 1$, where it can be compared with the single element C- or L-filter. Commercially available L-section low-pass filters can maintain an adequate rejection level to 1 GHz.

4-5.2.1.4 π-Section Filter

The π-section filter, shown in Fig. 4-37 has high insertion loss over a wide frequency range and moderate space requirements.

The theoretical insertion loss with source and load resistance R is:

$$IL = 10\log\left[1 + \left(\frac{f}{f_o}\right)^2 D^2 - 2\left(\frac{f}{f_o}\right)^4 D + \left(\frac{f}{f_o}\right)^6\right], \quad \text{dB} \quad (4\text{-}9)$$

where

$D = (1 - d)/(3\sqrt{d})$

$d = L/(2CR^2)$ = damping factor, dimensionless

$f_o = \frac{1}{2\pi}\left(\frac{2}{RLC^2}\right)^{1/3}$, Hz

The shape of the insertion-loss curve is highly dependent on the damping factor as shown in Fig. 4-36. Again the sharpest cutoff characteristic occurs for $d = 1$.

An actual insertion loss curve of a π-section filter has a slope of approximately 18 dB per octave beyond the cut-off frequency f_o, but it will ultimately level off. Typically, a filter with cutoff at 50 kHz will level off to about 80 dB insertion loss at about 1 MHz. The high-frequency performance can be improved by internal shielding within the filter case. However, the π-circuit is very susceptible to oscillatory ringing when excited by a transient.

4-5.2.1.5 T-Section Filters

A T-section filter (see Fig. 4-38) has an insertion loss (with source and load resistance R) given by

$$IL = 10\log\left[1 + \left(\frac{f}{f_o}\right)^2 D^2 - 2\left(\frac{f}{f_o}\right)^4 D + \left(\frac{f}{f_o}\right)^6\right], \quad \text{dB} \quad (4\text{-}10)$$

where

$D = (1 - d)/(3\sqrt{d})$

$d = R^2C/(2L)$ = damping factor, dimensionless

$f_o = \frac{1}{2\pi}\left(\frac{2R}{L^2C}\right)^{1/3}$, Hz

The major disadvantage of the T-type filter is the requirement for two inductors, which under some circumstances may cause a size penalty. The equations for the insertion loss of a T-circuit and a π-circuit, as given by Eqs. 4-9 and 4-10, are seen to be identical. The filters have three modes of response. When d equals 1, the response is optimally damped and is the

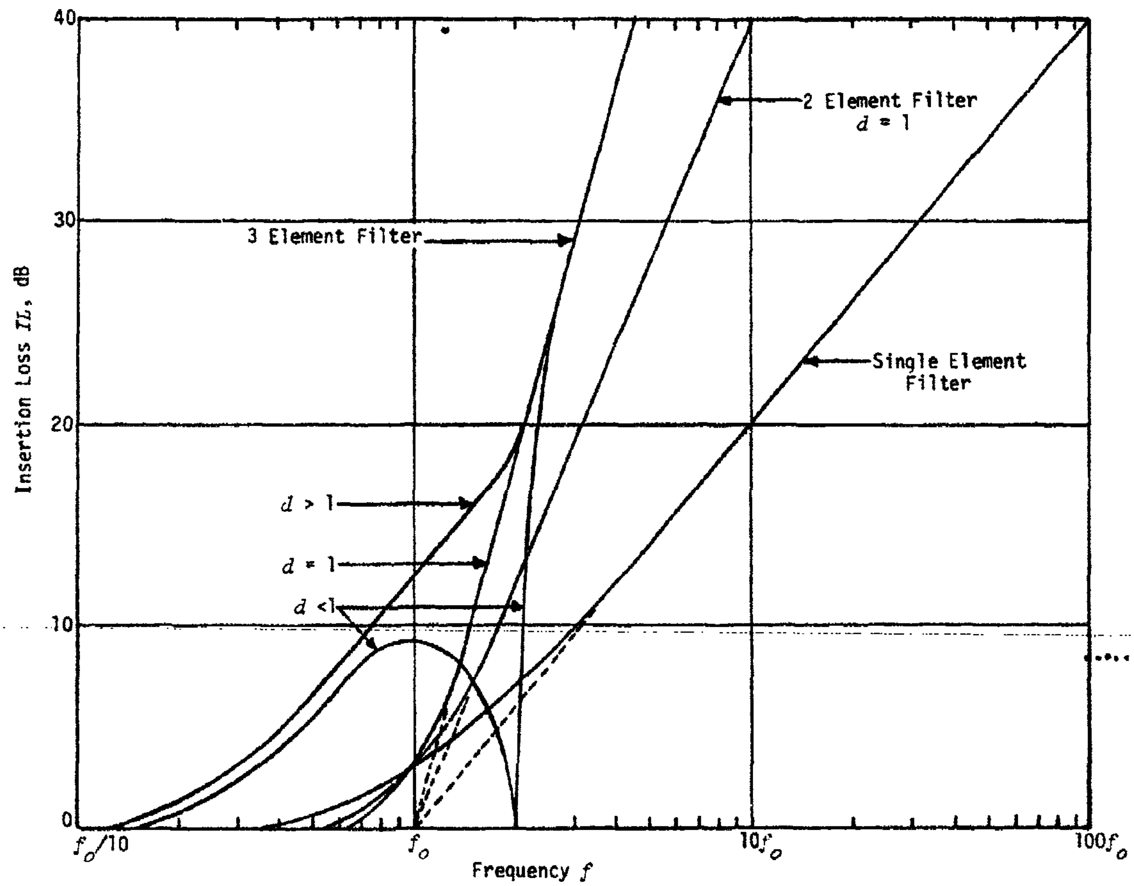


Figure 4-36. Response Modes of Filters

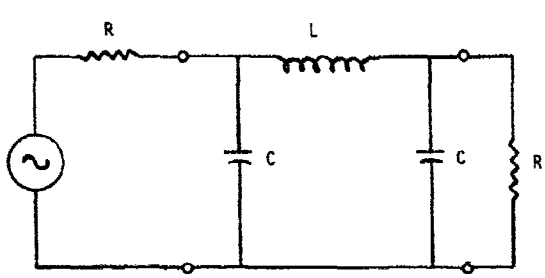


Figure 4-37. Low-pass π-Filter

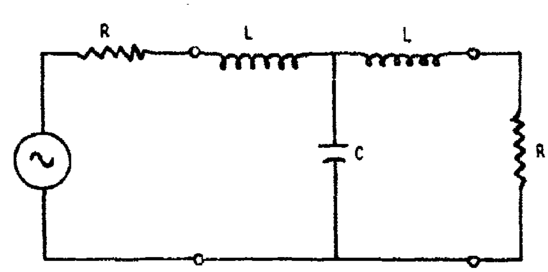


Figure 4-38. Low-pass T-Filter

ideal (Butterworth) response curve. When d is less than 1, the curve has a maximum in-band loss of:

$$IL = 10 \log \left(1 + \frac{4D^3}{27}\right), \quad \text{dB} \qquad (4\text{-}11)$$

at the frequency where $f = \sqrt{D/3}$. A minimum loss point will also occur at the frequency where $f = \sqrt{D}$

4-5.2.1.6 Multiple Section Filters

All of the filters just described can be cascaded. The main objectives of cascading are (1) when similar filters are cascaded, to obtain a higher rate of increase in attenuation in the stop band over that obtainable with a single section; and (2) when dissimilar filters are cascaded, to provide good attenuation characteristics over a broad frequency range. For example, with a simple shunt capacitor filter, two capacitors commonly are used — one of which is effective at low frequencies, and the other at high frequencies.

In addition, the performance of a multiple section filter is less affected by terminating impedances as discussed in the next paragraph.

4-5.2.1.7 Effects of Filter Terminations

In actual use — especially with power line filters — the source and load impedances are unequal, may be reactive, and vary with frequency. Furthermore, impedances can vary with time at any one location.

When the terminating impedances are primarily reactive, resonance with reactance in the filter may occur, producing an insertion gain at or near the frequencies of resonance. These are usually in the pass band or near the cutoff frequency in the stop band. When the terminating impedances are resistive but either very much higher or lower than the characteristic impedance of the filter, the filter elements themselves may exhibit resonance because of inadequate damping and likewise produce insertion gains at particular frequencies. Fig. 4-39 shows curves for filters of single T-, L-, and π-sections and two T- and L-sections under conditions of a 0.001-ohm source resistance and a 3.33-ohm load resistance. Note that each filter has negative insertion losses at various frequencies.

When the external impedances are not under the designer's control, nor are they fully known, it may be advantageous to use filter structures which are less sensitive to external influences, such as multisection filters. Another alternative is to use lossy lumped filter elements or lossy distributed elements.

4-5.2.1.8 Lossy Filters

Although these filters are not strictly "lumped" element filters, they are covered here because of the similarity of their characteristics to those of lumped filters. Usually they are constructed using lossy ferrite cores in the inductances or coaxial transmission lines with ferrite or other lossy dielectrics. As an example, Fig. 4-40 (Ref. 15) shows a low-pass filter made of ferrite tube with conducting silver coatings deposited on the inner and outer surfaces to form the conductors of a coaxial transmission line. Lossy ferrite beads encircling a wire produce an effective series inductance of the order of microhenries and series resistance varying from fractions of an ohm below 1 MHz to tens of ohms at 100 MHz (Ref. 14). More or longer beads provide additional series inductance and resistance. High amplitude power frequency or other currents may cause some reduction in the suppression effect due to ferrite saturation.

Improvement of high frequency rejection characteristics of a conventional low-pass filter may be obtained by employing a conventional reactive filter in cascade with a lossy line section. This arrangement can provide an overall characteristic having both a rapid cutoff slope and a high-stop band attenuation. An example of the improvement in stop-band attenuation that can be gained by preceding a reactive filter with a lossy line section is illustrated in Fig. 4-41, which shows the performance of a reactive low-pass filter constructed with lumped constant elements. The rapid cutoff at 400 MHz is followed by a high attenuation region between 400 MHz and 3 GHz, but at frequencies above 3 GHz the attenuation is greatly reduced. If the same low-pass filter is preceded by a section of lossy coaxial line, the attenuation characteristic is altered to that shown in Fig. 4-42. The addition of the lossy section has increased the passband attenuation only slightly, but the stopband attenuation has been increased to greater than 60 dB.

Still another form of ferrite filter that extends the ferrite bead concept is the filtering connector. Lossy filters are built directly into a male connector assembly, as shown in Fig. 4-43(A). The insertion loss characteristic is shown in Fig. 4-43(B).

A variety of materials in other forms is available for suppressing EMI. These materials may not be approved for certain military applications, especially where exposure to weather or humidity is severe, such as with ignition systems. When contemplating their use one should make certain of the acceptability. An example is tubing that can be slipped over standard wire and cable, and can provide shielding from low frequency electric and magnetic fields, and will not cause dc or power frequency losses. Representative data are shown in Fig. 4-44.

4-5.2.2 High-pass Filters

Although not as common as the low-pass type, high-pass filters also have application in EMI reduction. In particular, such filters have been used to remove ac power line frequencies from signal channels and to reject particular low frequency environmental signals.

High-pass filters can be designed by inverting the high-pass filter response requirements, so that they become requirements on a low-pass filter. Low-pass filters meeting this new requirement can be readily transformed back into the high-pass filter of interest.

The low-pass filter transforms into a high-pass filter by replacing each coil with a capacitor, and vice versa.

This method is based on the fact that inductors and capacitors are inverse elements. The impedance of an inductor L in a given branch at a frequency f_a has the same magnitude as that of a capacitor C at a frequency f_b if $2\pi f_a L = 1/(2\pi f_b C)$. Thus, if a low-pass filter has been designed as shown in Fig. 4-45(A), it can be converted to a high-pass filter as shown in Fig. 4-45(B). Each element in (A) has been inverted in (B) in such a way that $LC = 10^{-12}$. As a result, filter B will have attenuation at frequency f_b equal to that of filter A at frequency f_a,
where

$$2\pi f_b = \frac{1}{2\pi f_a LC} = \frac{10^{12}}{2\pi f_a} \tag{4-12}$$

Courtesy of Cornell-Dubilier, EMC Compatibility Solution Bulletin 220-2/72/7.5M (MSD).

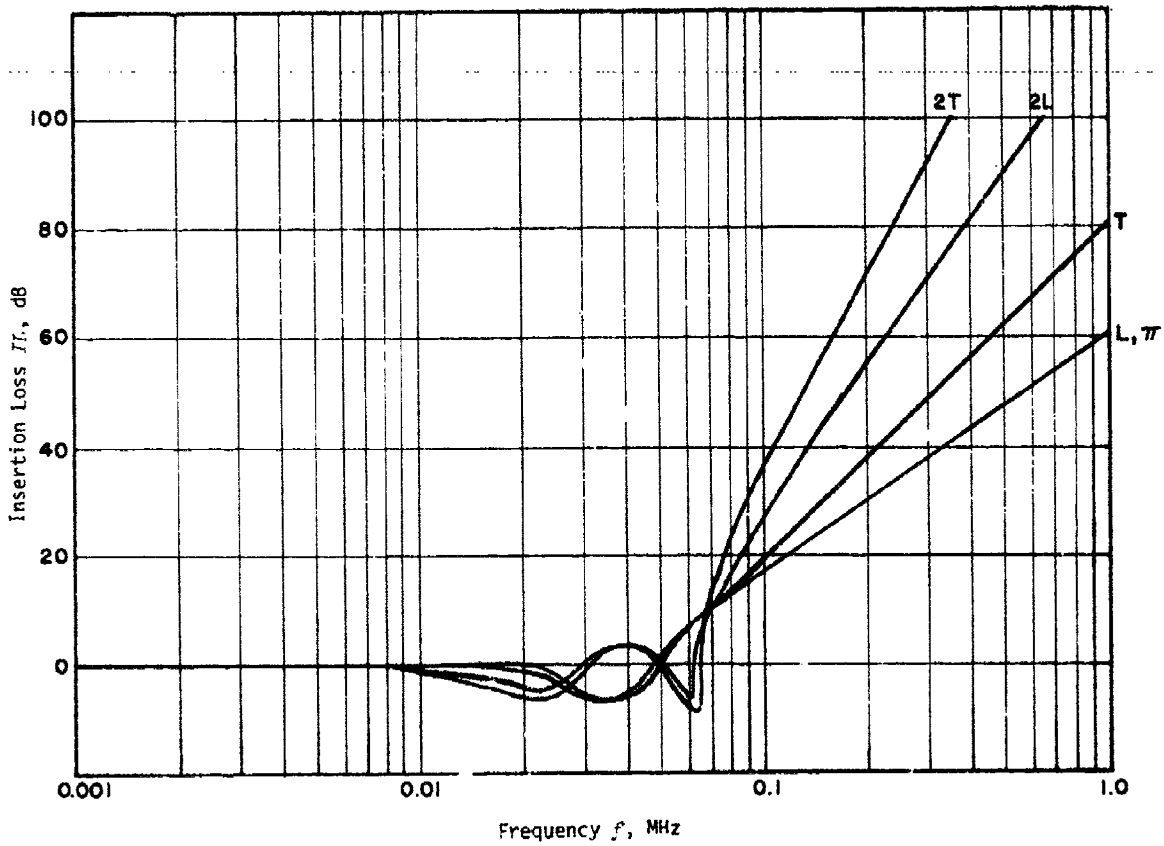


Figure 4-39. Insertion Loss for Mismatched Filters (Source Resistance = 0.001 Ω, Load Resistance = 3.33 Ω)

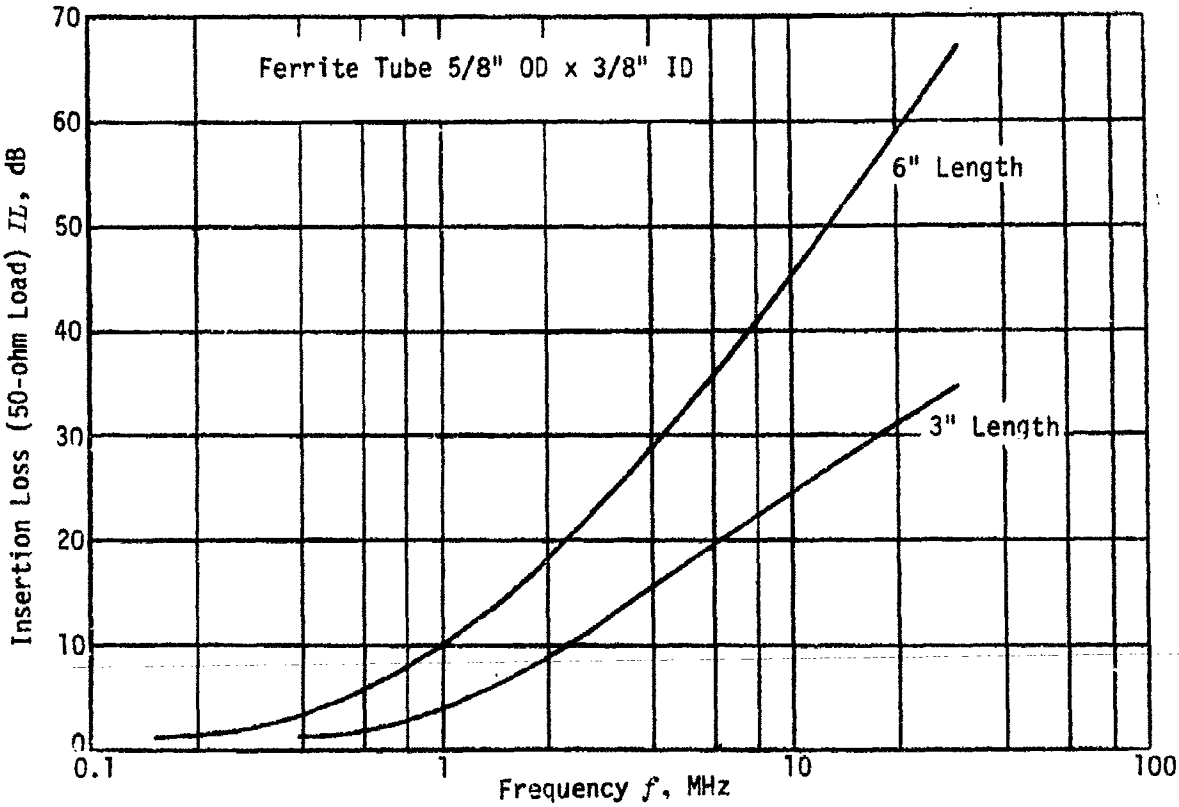


Figure 4-40. Insertion Loss of a Ferrite Tube Low-pass EMI Filter (Ref. 15)

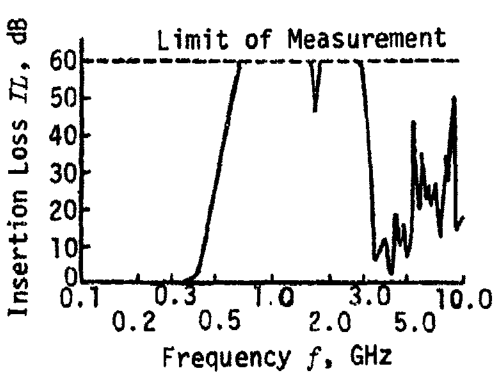


Figure 4-41. Typical Low-pass Filter Loss Characteristics, Low-pass Filter Only

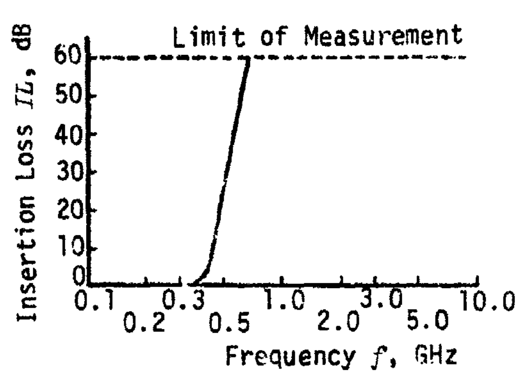
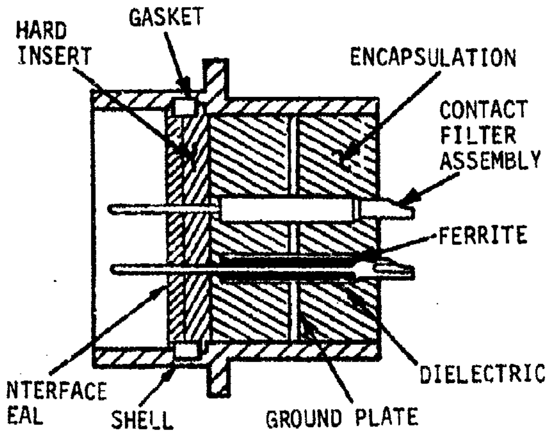
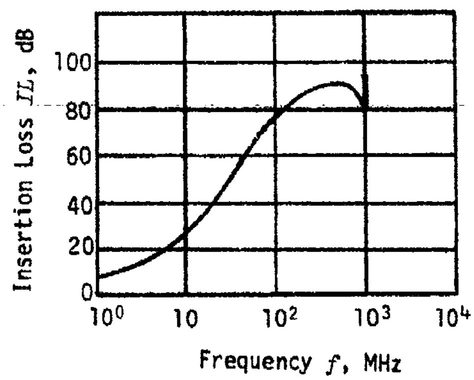


Figure 4-42. Typical Low-pass Filter Plus Lossy Filter Section



(A) Construction Details



(B) Typical Attenuation, dB

Figure 4-43. Typical Characteristics of Lossy Connector

For example, the cutoff frequency of filter (A) is approximately 10 kHz. The cutoff frequency of filter (B) is therefore approximately 2.5 MHz.

4-5.2.3 Bandpass Filters

Each low-pass filter also can be the basis for defining a unique family of symmetrical bandpass filters with known characteristics and vice versa. Thus, the requirements of a bandpass filter (a filter designed to pass a given frequency band and reject signals outside that band) can be established readily by use of the following transformation procedure:

a. Convert the desired bandpass filter requirements into low-pass filter requirements. The low-pass prototype has the same 3-dB bandwidth and insertion loss as the bandpass filter.

b. In the case of single section L, T, and π filters having equal input and output resistances, select a low-pass filter with the required attenuation using the two- and three-element filter design equations discussed in the Handbook.

c. Establish the filter element values in the manner previously described using the RF 3-dB bandwidth value.

d. Resonate each L and C at the required bandpass center frequency.

As an example of this procedure, consider a bandpass filter requirement of a center-tuned frequency, f_c = 1.0 MHz, and a skirt roll-off rate of at least 15 dB/octave at the required input and output resistance. There is to be no ripple in the pass-band, i.e., response should be monotonic at all points. Bandwidth is to be 100 kHz between the −3 dB points.

A three-element, Butterworth, low-pass π-network is selected as the prototype low-pass filter. Eq. 4-9 is satisfied for L and C values of 160 μH and 0.03 μF, respectively, using a cutoff frequency of 100 kHz, a damping factor of unity, and R = 50 Ω.

The L and C values are next resonated at 1.0 MHz using the relationship

$$f_c = \frac{1}{2\pi\sqrt{LC}}, \text{ Hz} \tag{4-13}$$

The result is a 150-pF capacitor in parallel with L, and a 0.8-μH inductor in series with C. The final filter configuration is shown in Fig. 4-46.

Note that the RF filter response is log-frequency symmetrical. That means that the response on a logarithmic frequency axis at some displacement above f_c is a mirror image of the equivalent displacement below f_c; i.e., if the attenuation at a frequency χf_c is N dB, then the attenuation will be the same at f_c/χ — χ is a multiplication factor. For the example given, the bandpass filter cutoff frequencies are shown in Fig. 4-46 as 0.95 and 1.05 MHz.

Note that in this transformation technique after transforming the low-pass filter section into a bandpass filter, a capacitor is added across each coil of a

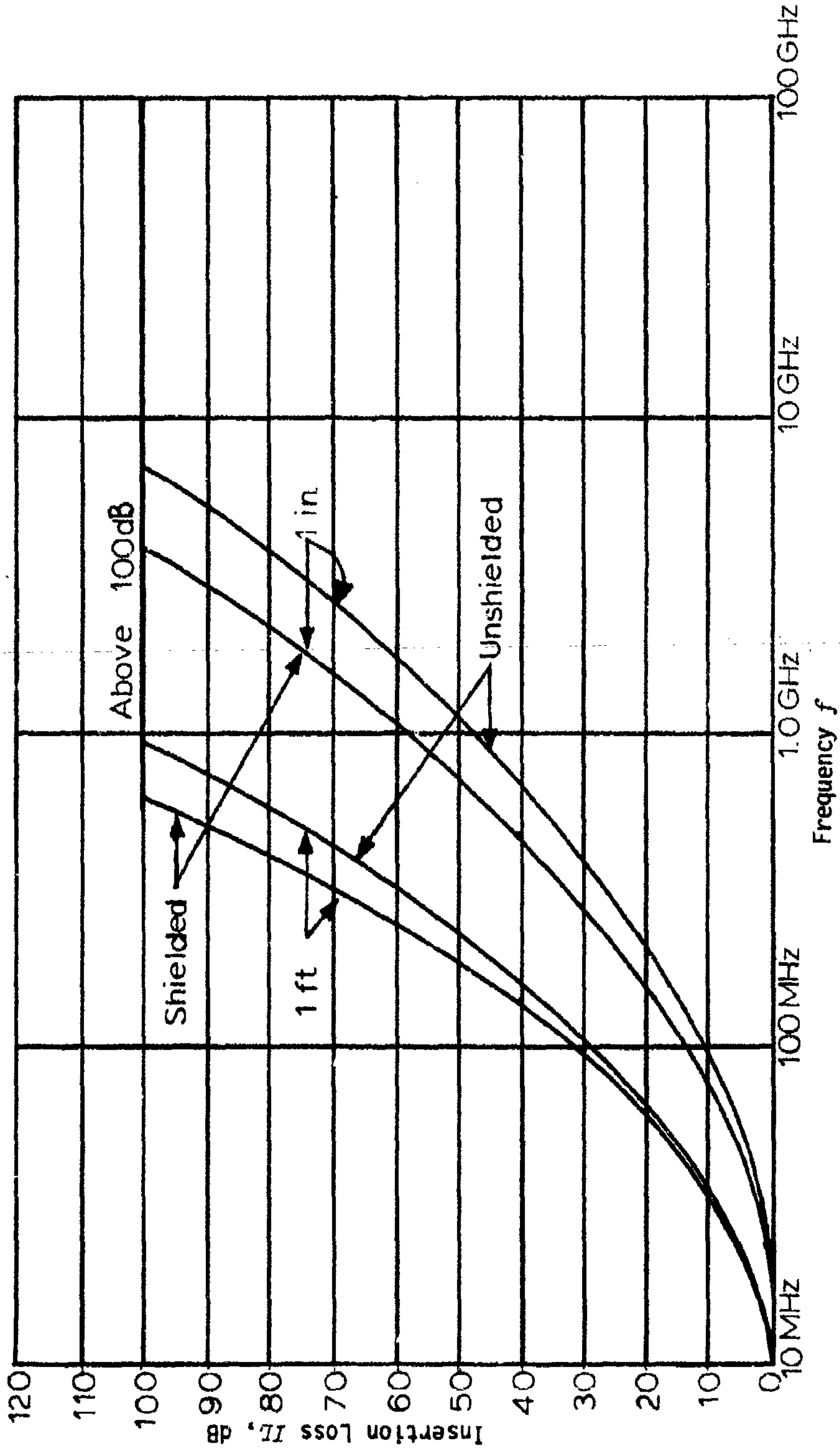


Figure 4-44. Typical Characteristics of Lossy Line Suppressant Tubing

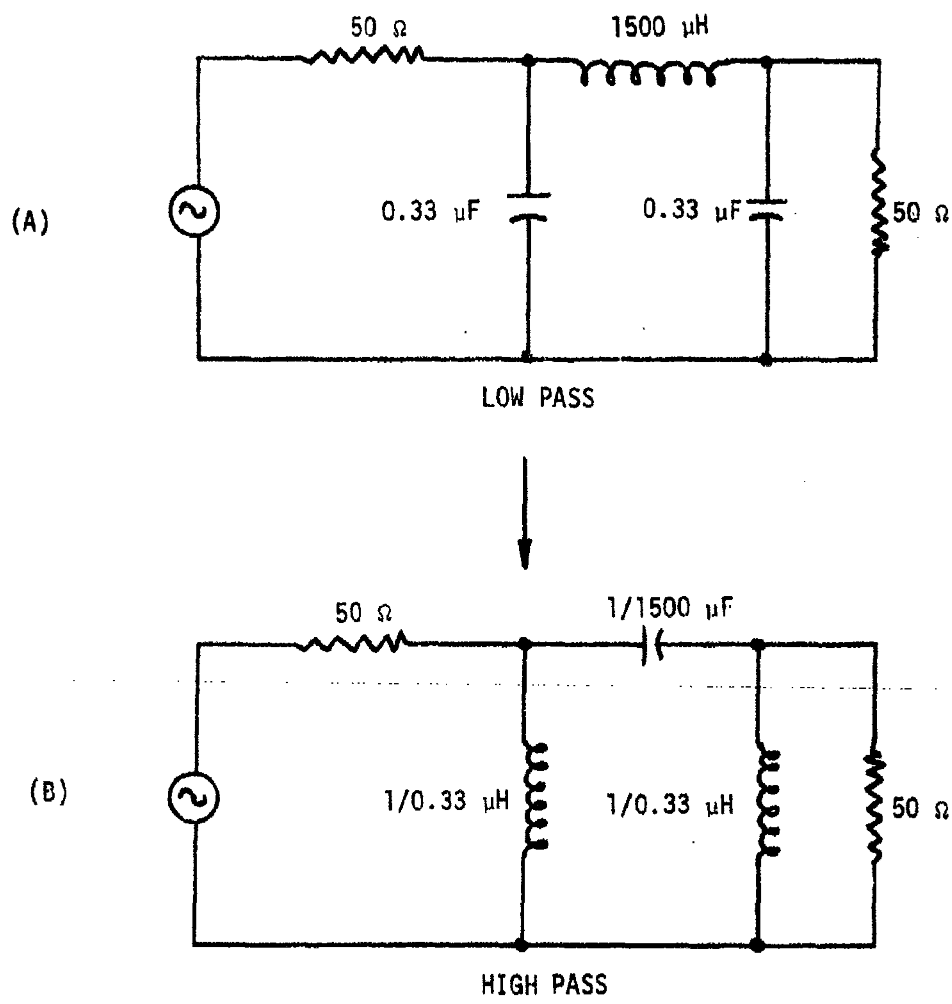


Figure 4-45. Low-pass to High-pass Transformation

size to resonate the coil at f_c. A coil is added in series with each capacitor of the low-pass filter of a size to resonate the capacitor at f_0.

The preceding approach is conceptually the same for multiple-section bandpass filters, or for filters whose input and output impedances differ, but the process becomes much more complicated. Tabular techniques are available to simplify the process under these circumstances. They are based on first converting the filter requirements to per-unit values (per cycle, per ohm of input impedance, etc.), designing the low-pass filter on that basis, and then converting back to the bandpass equivalent.

Butterworth filters have a maximally flat bandpass response. If some ripple within the passband can be tolerated, then a steeper descent into the attenuation band can be obtained. Tchebyscheff filters have a greater roll-off rate than Butterworth filters for the same number of components, and generally are used in bandpass designs where bandpass ripple can be tolerated. Tabular approaches to the design of Tchebyscheff bandpass filters are also available (Ref. 16).

4-5.2.4 Band-rejection Filters

Band-rejection filters attenuate a specific narrow band of frequencies. This type of device normally is used for series rejection between the interference source and the load. An alternative is to use a bandpass configuration that shunts the interference to ground. A notch filter or wavetrap is a filter that rejects a specific frequency. It may take the form of a lumped-constant inductor-capacitor circuit, or it may be a shorted quarter-wave coaxial or waveguide stub, or a crystal or ceramic filter lattice. The inductive characteristics of capacitor leads and foil can be planned so that the capacitor acts as a self-contained wavetrap. For frequencies below about 1 MHz, a twin-T resistor-capacitor filter often is found to be an acceptable configuration.

The simplest types of wavetraps are parallel or series resonant circuits such as those shown in (A) and (B) of Fig. 4-47. The configuration of Fig. 4-47(A) will give a high impedance at the resonant frequency, while the configuration of Fig. 4-47(B) provides a low impedance at resonance. The disadvantages of these circuits are that their skirt falloff rates are low (6 dB/octave), and they do not present a good impedance match to either the source or load. Band reject performance can be improved by using parallel and series-tuned elements in L, π, or T configurations, also illustrated in Fig. 4-47(A) to (D).

The details on the design of the types of band-reject filters are available in many filter textbooks and handbooks, and will not be discussed here. However, some additional comments are considered appropriate on one particular type of notch filter, because of its wide use in this type of application.

The twin-T notch filter shown in Fig. 4-48 is useful as a band-reject filter in the lower frequency ranges. At low frequencies, the twin-T filter can achieve a circuit Q on the order of 100, which would not be economically feasible for a wavetrap or inductance-capacitance type filter at the same frequency. Shunting effects reduce its usefulness at high frequencies. The notch frequency f_n is determined by (Ref. 17)

$$f_n = \frac{0.1592\,K}{(R_1 R_2 C_1 C_2)}, \quad \text{Hz} \tag{4-14}$$

Three special cases are of interest: The case when the twin-T parameter ratio $K = 1$ gives the symmetrical form of Fig. 4-49. With $K = 0.5$, a circuit with three equal resistances as shown in Fig. 4-50 is obtained. In Fig. 4-51, with $K = 2$, three equal capacitances result. It should be pointed out that the twin-T notch filter parameters must be accurately selected to obtain attenuation at the null frequency. Getting the best possible null requires careful balancing in network tuning; a convenient way to do this is by use of trim capacitors or potentiometers.

4-5.3 ACTIVE FILTERS

EMI filters made of passive elements are sometimes bulky and heavy. Active filters, using transistors, can provide large values of equivalent L and C without excessive size and weight. Moreover, the low impedance levels existing at low frequencies in power lines can be accommodated more easily with active devices.

Active filters for a dc line may contain capacitors as storage elements, series regulators to create a high impedance path, and shunt regulators in combination with high gain feedback systems for cancellation of interference. In the case of ac lines, cancellation is

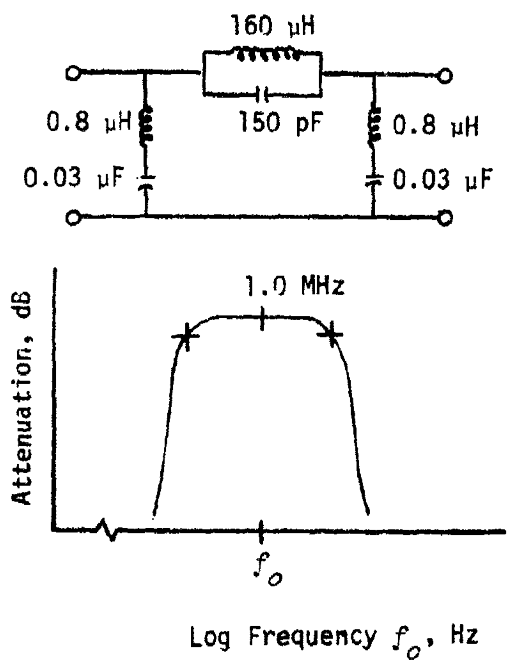


Figure 4-46. Example of Bandpass Filter Design

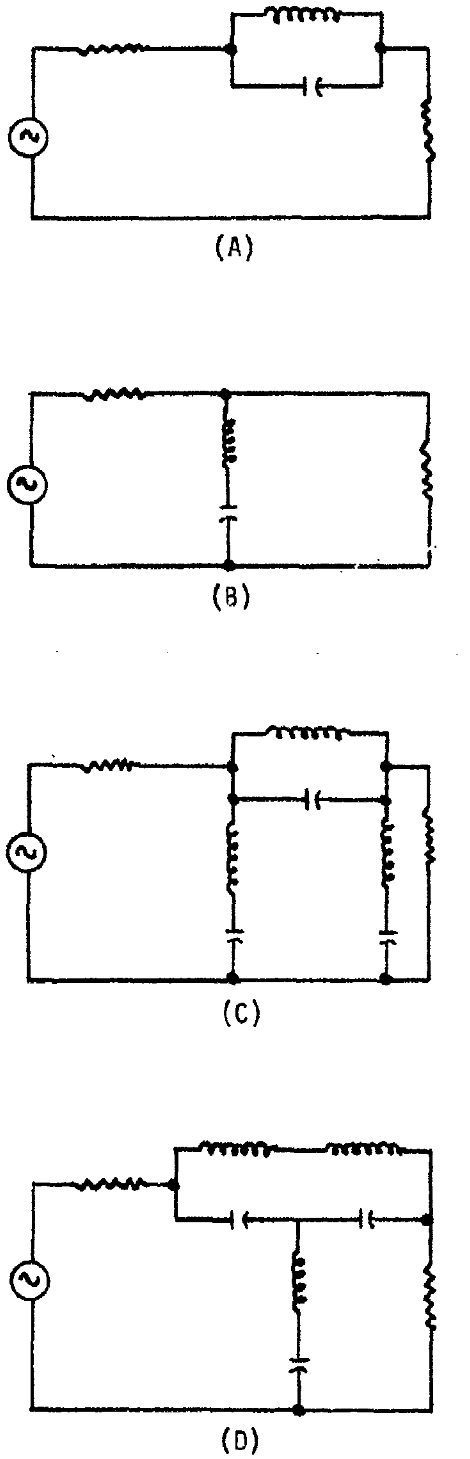
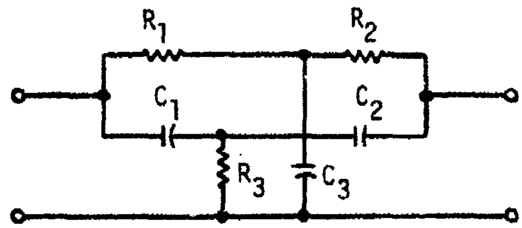


Figure 4-47. Band-reject Filter Configurations

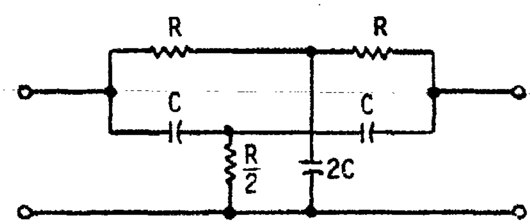


$$K = \frac{C_1 + C_2}{C_3} = \frac{R_1 R_2}{R_3(R_1 + R_2)} \qquad \omega = \sqrt{\frac{K}{R_1 R_2 C_1 C_2}}$$

$$R_3 = \frac{R_1}{2K} \qquad f = \frac{0.1592\ K}{\sqrt{R_1 R_2 C_1 C_2}}$$

$$C = \frac{C_1 + C_2}{K}$$

Figure 4-48. Twin-T Notch Filter

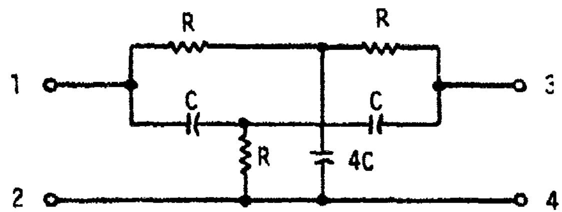


$K = 1$

$\omega = 1/(RC)$

$f = 0.1592/(RC)$

Figure 4-49. Twin-T Network With $K = 1$

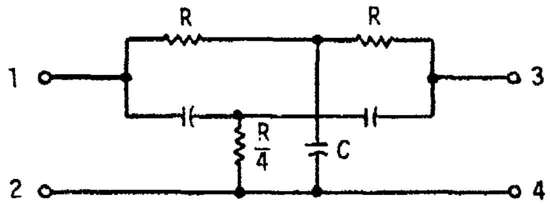


$K = 0.5$

$\omega = 1/(RC\sqrt{2})$

$f = 0.1125/(RC)$

Figure 4-50. Twin-T Network With $K = 0.5$



$K = 2$

$\omega = \sqrt{2}/(RC)$

$f = 0.2251/(RC)$

Figure 4-51. Twin-T Network With $K = 2$

a most effective way to minimize interference. In contrast to conventional regulators used in regulated power supplies, these filters must not regulate the amplitude of the power to be passed.

Active ac power line interference filters pass, with high efficiency, only a narrow band about the power frequency. Voltage attenuation values of approximately 30 dB can be obtained even at very low load and source impedance levels; two filters may be cascaded for higher attenuation values.

The power line filter scheme shown in Fig. 4-52(A) uses phase cancellation and operates as follows. The input is fed into an ac-coupled amplifier through a notch filter, which is tuned to the power line frequency. The amplified interference signal, without the power line fundamental frequency, and with opposite polarity, is returned in series to the source through the transformer. All signals except the fundamental are attenuated by the gain of the amplifier. Within a limited range, a separate control circuit can provide automatic tuning of the notch filter to powerline frequency and correct any change in the filter tuning itself. The voltage attenuation curve for a 220-V, 20-A unit is shown in Fig. 4-52(B).

When inserted in the line, the filter introduces the equivalent of a small inductance of 700 μH at the pass frequency. The change in output voltage at full current rating caused by this inductance is negligible, i.e., 1 V for a load power factor approaching unity and up to 6 V in the worst case for zero power factor. This type of filter can handle large interference voltages, but with some loss of efficiency. Efficiencies of 90% have been realized for 200-V, 20-A filters.

Two devices that have been used are two-ports known as the gyrator and the negative-impedance converter (NIC). When loaded with a capacitance C, the input impedance of a gyrator is that of an inductance of value $L = Ck$ where k is a constant characteristic of the gyrator. Gyrators at low frequencies are fabricated using conventional amplifying elements with feedback; a total of four such elements is needed (Ref. 18).

Negative-impedance converters have the property of presenting a driving point impedance at one port, which is the negative of the terminating impedance connected to the other port. A practical realization of an NIC is obtained using two conventional amplifying elements (Ref. 18).

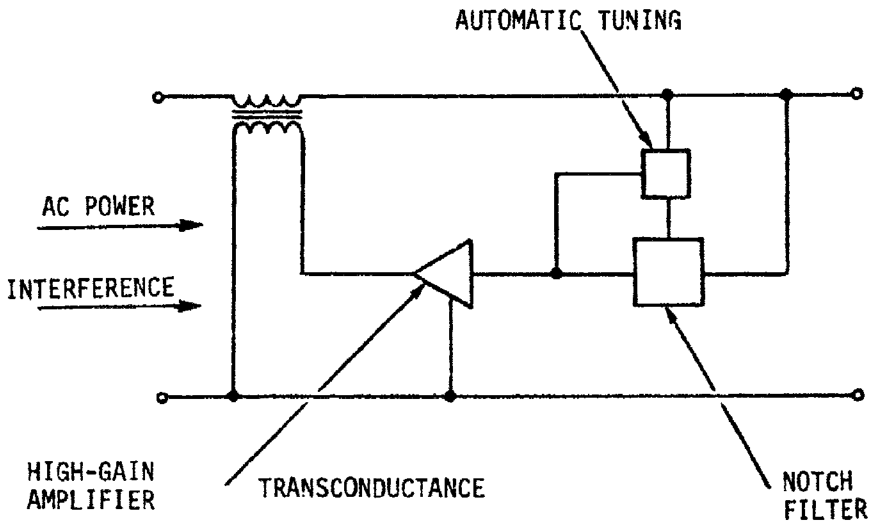
Integrated and hybrid circuit technology has enabled the construction of small active filters for use in low-level circuits to perform typical low-pass, high-pass, or bandpass filtering such as required for various circuit purposes. Multiple pole Butterworth, Tchebysheff, and Thomson characteristics are available and, in some cases, are adjustable. Although it is possible to construct such filters with relatively large power handling capacity, they are presently most useful — in so far as EMC/EMI control is concerned — when they can be applied in low-level circuits.

4-5.4 MICROWAVE FILTERS

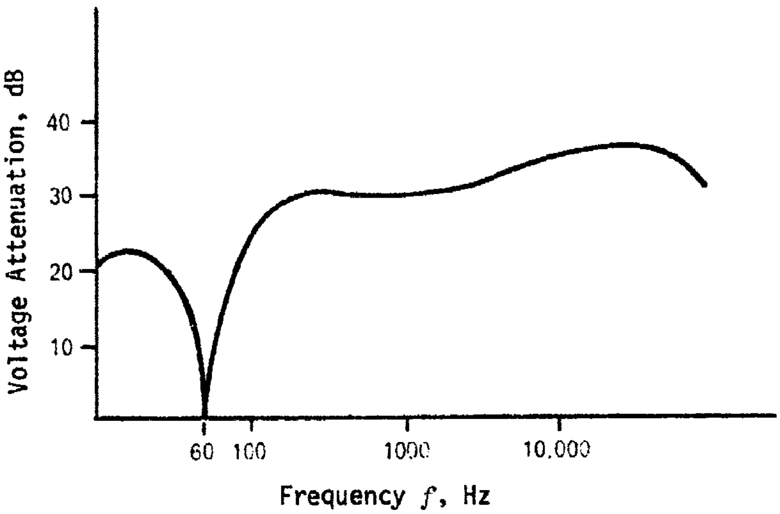
Microwave filters usually are of coaxial, stripline, or waveguide construction. The coaxial technique is applicable up to 4 GHz, while waveguide elements normally are used at higher frequencies. An important consideration is choosing the filter structure to be used is the power-handling capacity. Waveguide structures generally are capable of handling considerably more power than either of the other two. A survey of microwave filters useful in interference control (Ref. 19) shows the application of all these structural types to low-pass, high-pass, band-pass, and band-elimination filtering. The use of electrically tunable ferromagnetic devices, such as the YIG (Yttrium Iron Garnet) crystal, also is mentioned there.

4-5.4.1 Stripline Filters (Ref. 20)

The relative ease of fabricating stripline circuits and their inherent low losses make them attractive for microwave applications. Fig. 4-53 shows an example of the stripline configuration. A single copper strip forms the center conductor. This strip is embedded between two dielectric sheets which are in turn covered by two copper plates considered to be at ground potential. Filters are constructed using combinations of series lines, open or shorted shunt lines, and series capacitors (formed by transverse slots). Shunt lines are formed by lengths of stripline at right



(A) CIRCUIT DIAGRAM



(B) ATTENUATION CHARACTERISTIC

Figure 4-52. Active Filter for Powerline Interference

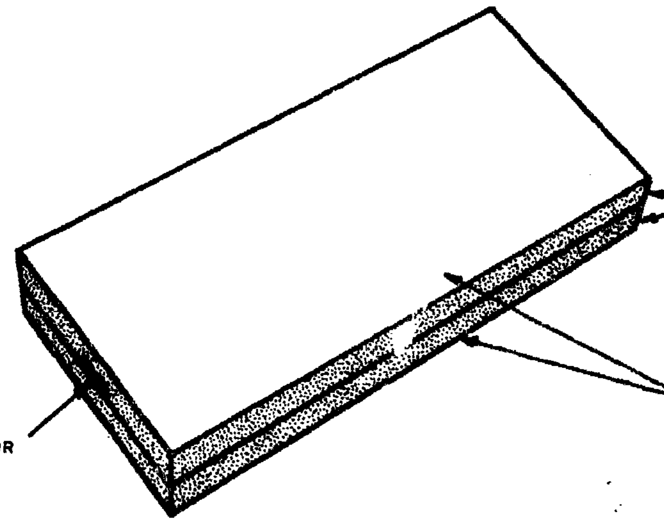


Figure 4-53. Strip Transmission Line Construction

angles to the main line. Short lengths of open-circuited line appear capacitive; short lengths of short-circuited line appear inductive. A method of producing shorted shunt lines is shown in Fig. 4-54. A common problem encountered in the use of stripline techniques is the propagation of higher-order modes. Such modes can be excited by any unintentional tilt of the center conductor, yielding narrow spurious pass responses in a rejection band. These can be eliminated by loading the line with resistor cards, powdered iron slugs, or screws located so as to absorb energy from the higher modes without affecting the main transverse electromagnetic (TEM) lines. In addition to preventing the propagation of undesired higher-order modes by electrical means, the preceding methods provide a mechanically rigid structure.

Fig. 4-55 shows the stripline center conductor pattern and the low frequency equivalent circuit for a T-type filter. The value of the inductive elements may be varied by altering the width W of the center conductor and hence the characteristic impedance. The capacitive element in the lumped model is produced by a transverse line of the appropriate electrical length. Note that the transverse element is placed symmetrically about the center conductor totally to one side. Splitting the shunt capacitive reactance into two equal sections on either side of the line reduces the necessary physical length of the stubs and therefore places spurious responses farther away from the cutoff frequency. By combining the basic π- and T-sections, more complex filters may be fabricated. The center conductor configuration for a multiple section low-pass filter and its equivalent low-frequency network are shown in Fig. 4-56.

There are two types of spurious responses generally encountered in this type of stripline filter. One type of response occurs at a frequency where the spacing between shunt elements is equal to a half wavelength and at integer multiples thereafter. Another type occurs when the shunt element length equals a half wavelength or an integer multiple thereof. In carefully designed stripline filters, these responses occur far above the pass-band. If it is found necessary, these responses may be eliminated by cascading the filter with a low-pass filter having a cutoff frequency slightly lower than the first spurious response, and having no coincident spurious responses.

Impedance matching networks may be fabricated readily in stripline. Purely resistive matching can be accomplished by linearly or exponentially tapering the width of the center conductor. Fig. 4-57 shows a line linearly tapered to match two different characteristic impedances Z_1 and Z_2. More complex matching may be accomplished by including shunt reactive elements of the proper electrical length.

4-5.4.2 Waveguide Filters

All waveguides act as high-pass filters with cutoff frequencies determined by the shape and size of the

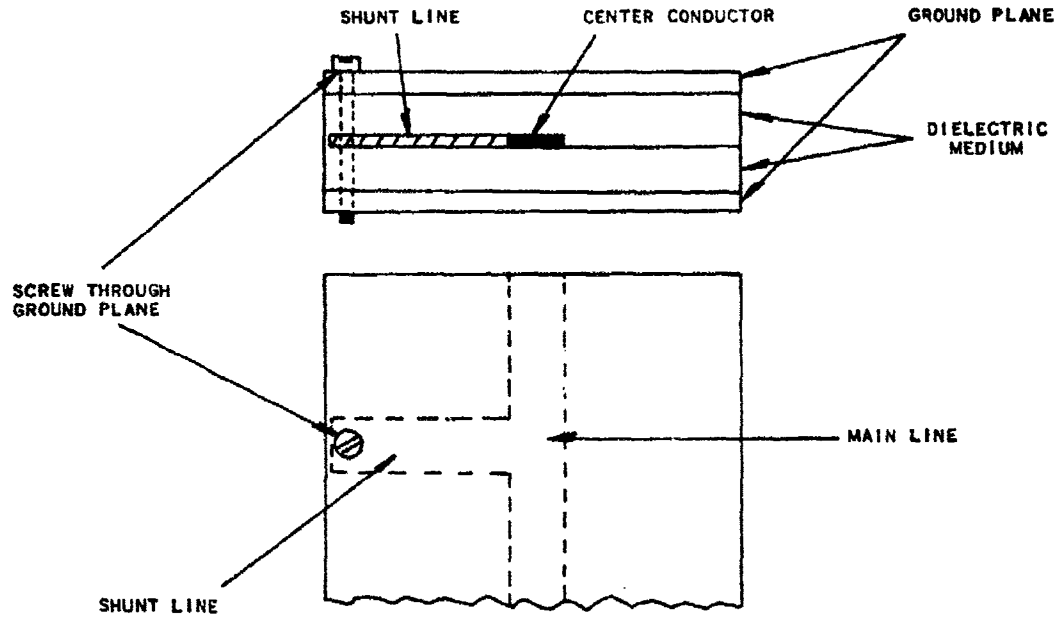


Figure 4-54. Shorted Shunt Line Construction

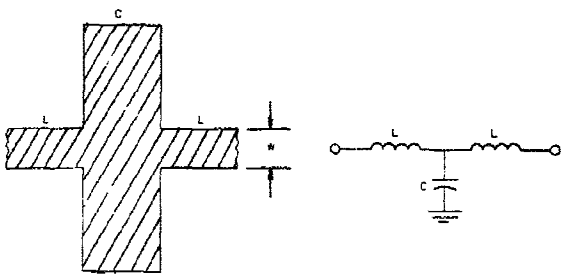


Figure 4-55. Symmetrical T-Section Filter and Equivalent Network

waveguide, and the mode of transmission. Application of the principle of the waveguide below cutoff as a filter in the design of equipment enclosures to provide openings for control shafts, meters, and ventilation is discussed in par. 4-6.5.2.2.

4-5.4.2.1 Wide-band Reflective Waveguide Filters

The serrated-ridge type waveguide provides high stop-band insertion loss over a wide frequency range. Values of pass-band losses of less than 0.1 dB, and of stop-band losses greater than 50 dB over a two-to-one frequency range, easily are attained with these waveguides. The power-handling capacity of this type of waveguide filter is only about 1% of the unperturbed waveguide. If carefully evacuated, such filters are able to operate at up to half the power-handling capacity of corresponding unridged waveguides at one atmosphere of air pressure. This type of filter also may be useful as part of a hybrid filter, where the power is divided into several branches. A similar corrugated waveguide filter (Fig. 4-58) has been designated the waffle-iron filter because of its construction. As in serrated, ridged waveguides, only the dominant mode propagates in the stop-band; therefore, there is no multimode problem. This filter is very compact, easily fabricated, and has a power-handling capacity of about 3% of the corresponding uncorrugated waveguide. The same filter, when evacuated, is capable of handling 100% of the power that the unperturbed waveguide can handle at one atmosphere of air pressure. Fig. 4-59 shows the characteristics of this filter.

4-5.4.2.2 Reactive Mode Devices

There are numerous reactive devices that can be used in filter designs to increase the stop-band insertion loss for a particular mode. Decreasing the height dimension of the waveguide in the filter will reflect most of the TE_{0n} modes and the degenerate modes,

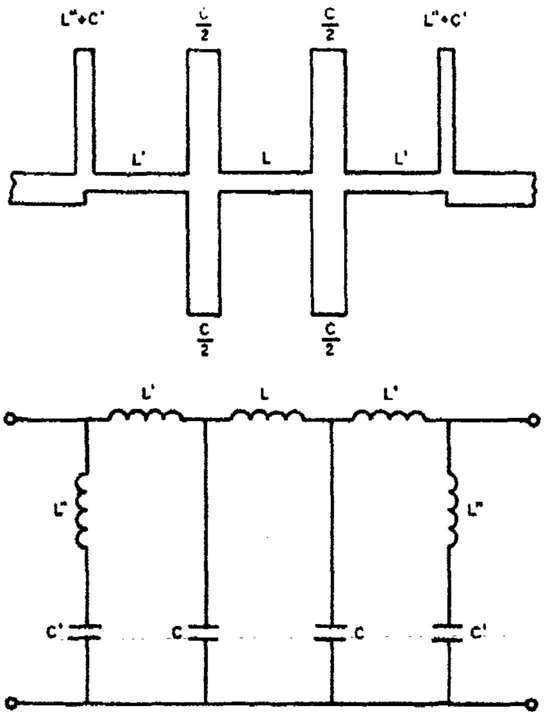


Figure 4-56. Low-pass Filter and Equivalent Network

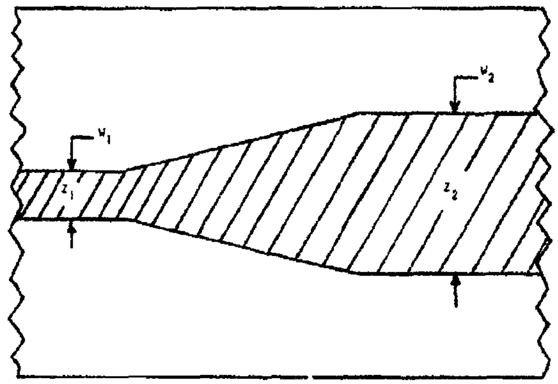


Figure 4-57. Linear Taper Matching of Strip Lines

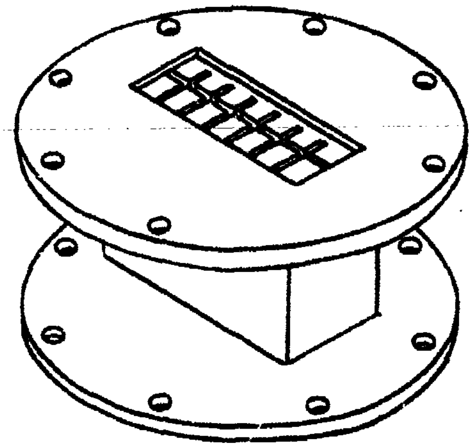


Figure 4-58. Waffle-iron Waveguide Filter

thereby providing high insertion loss (50 dB or more). However, the power-handling capacity is decreased in the same proportion that the height dimension is decreased. The same insertion loss for the same modes can be attained by placing a septum in the waveguide, parallel to the *y-z* plane, which does not disturb the propagation of the dominant mode. By tapering and rounding the edges, as shown on Fig. 4-60, the gradient field enhancement at the front and back edge of the septum becomes negligible. Thus, a thin septum has little effect on the power-handling capacity. It can be constructed either in the form of a series of rods projecting across the filter, or in the form of bullets projecting out from the narrow walls (Fig. 4-61).

4-5.4.2.3 Tuned Cavities

Another reactive device that is useful for increasing the stop-band loss is the tuned cavity. Fig. 4-62 shows a section of such a cavity attached to a waveguide section. A few cavities can add 30-dB loss over a narrow frequency range. Placement of these cavities away from the regions of maximum electric field enables the section to maintain 100% power-handling capacity.

4-5.4.2.4 Ferrite Filters

Ferrite materials, properly located in a waveguide, provide attenuation by absorbing undesired energy, thus providing a nonreflective filter. By placing a thin slab of ferrite material across the broad walls of a section of waveguide and biasing the ferrite with magnetic fields that control the resonant frequency, it is possible to absorb large amounts of microwave power over a selected frequency range. A substantial power loss may be experienced at the fundamental frequency in attempting to obtain extra broad stop-band loss. Ferrite slabs, placed on the waveguide

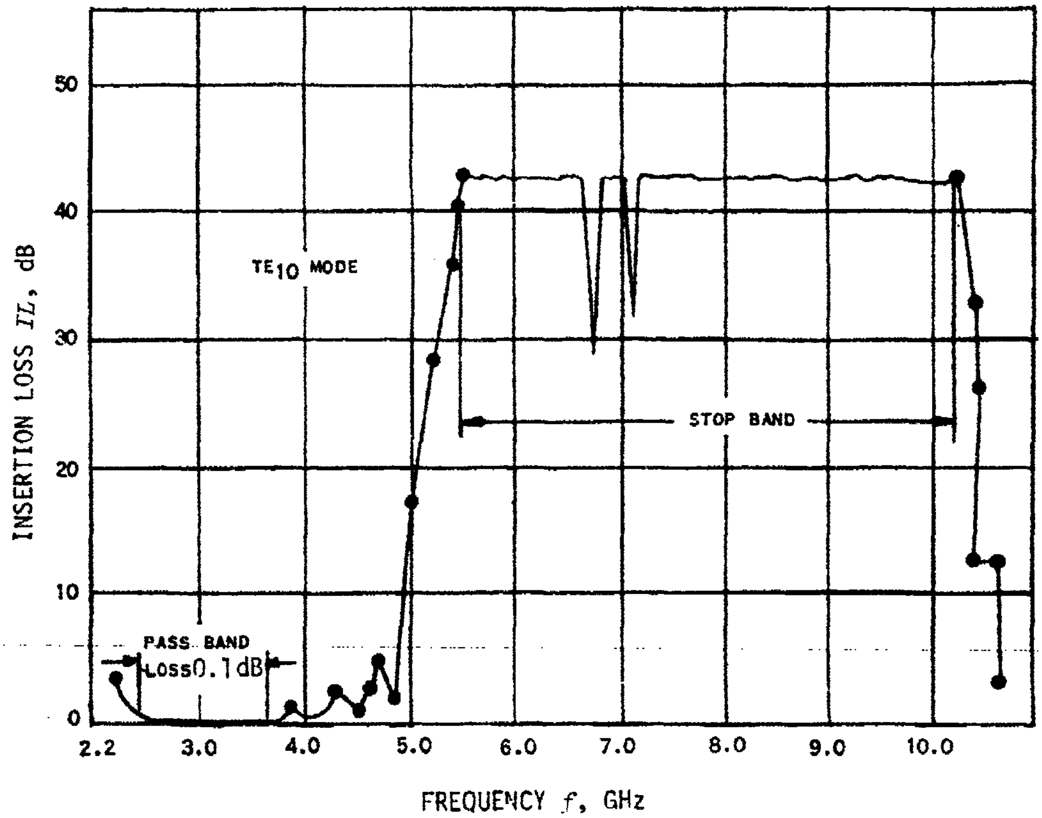


Figure 4-59. Characteristics of Waffle-iron Waveguide Filter

broad walls, will not absorb energy from all modes with equal efficiency because the intensities of the RF magnetic fields often vary considerably from mode to mode. In such filters, it is often necessary to reduce the height dimension of the waveguide to generate a sufficient magnetic field across the ferrite slab. Then, a sizable amount of the energy in the narrow wall modes is reflected instead of absorbed, and the power-handling capacity is lowered. Excellent characteristics of 0.2 dB to 0.3 dB pass-band loss and 30 dB to 50 dB stop-band loss can be achieved using ferrites at powers in the range of 5 MW and higher.

4-5.4.2.5 Absorbing Mode Filters

There are various filter techniques that can be used to absorb certain modes. Among these techniques is the placement of a thin resistance film perpendicular to the narrow wall and parallel to the broad wall of a waveguide, as shown on Fig. 4-63. Such a film effectively absorbs many of the modes that have a narrow wall component of electric field, without affecting the dominant mode. A narrow slot, filled with absorbing material and placed along the broad wall of the waveguide in a manner similar to a slotted line, will absorb the energy from all modes which have a current path perpendicular to the slot.

4-5.5 FILTER INSTALLATION AND MOUNTING TECHNIQUES

4-5.5.1 General

Proper installation of filters is necessary to achieve good results. Input and output wiring must be separated — particularly for good high-frequency performance — because radiation couples input energy directly to output wiring, thus nullifying the filter. Input and output terminal isolation is accomplished most easily by using a filter that mounts through a bulkhead or chassis. Where bulkhead mounting isolation is not feasible, shielded wiring

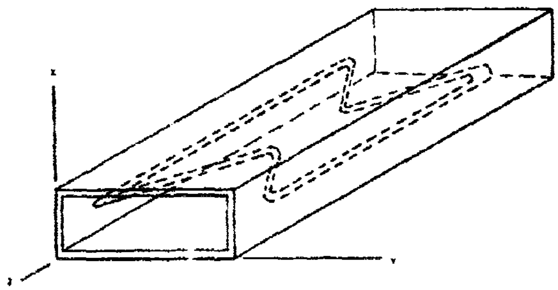


Figure 4-60. Septum Section

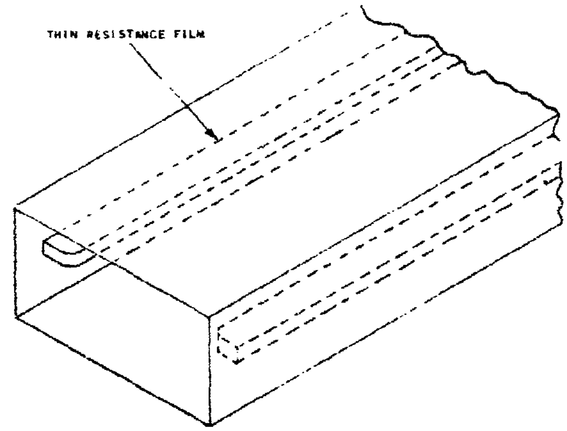


Figure 4-63. Narrow Wall Mode Absorber Strips

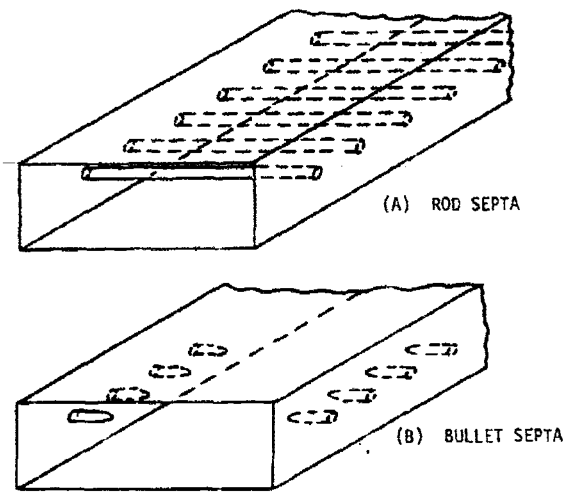


Figure 4-61. Rod and Bullet Septa

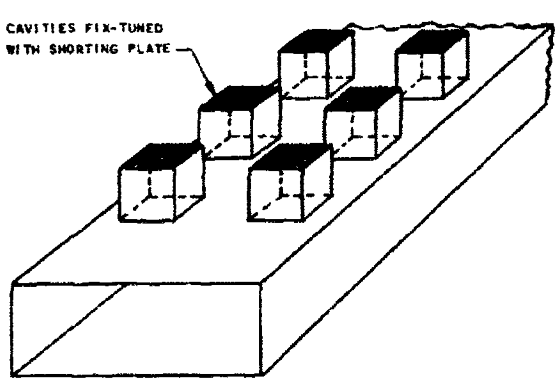


Figure 4-62. Tuned Cavity Section

should be used. When a filter is connected between a line and ground, the impedance of the ground connection can have a significant effect on the filter performance, as shown in Fig. 4-64. When a ground impedance Z_b exists, the filter case is raised above ground and the current through the first shunt capacitor C_1 divides at the junction of the two capacitors and Z_b. Some current flows through Z_b, depending upon the impedance. The remaining current flows through the second shunt capacitor C_2 to the load, thus compromising filter performance.

Although the location of the filter depends on the individual application, it should be installed as close as possible to the interference source.

4-5.5.2 Chassis Mounting

In general, any of the following five methods can be employed to mount grounded filters on a chassis:

a. Tabs on the filter body
b. Screws or bolts on the filter body
c. A flange on the filter body
d. A clamp on the filter body
e. A feedthrough stud for bulkhead mounting.

Fig. 4-65 shows these mounting techniques.

A typical filter installation is shown on Fig. 4-66, where the filter is integral with the interference source — in this case, a dc motor. In Fig. 4-66(A), the bulkhead mounting principle is used, and the filter input and output circuits are completely isolated. Fig. 4-66(B) shows how direct input-output coupling can reduce the effectiveness of the filter. Fig. 4-67 illustrates two incorrect methods of mounting a filter, both ineffective at high frequencies. In Fig. 4-67(A) the input and output leads are physically crossed. In

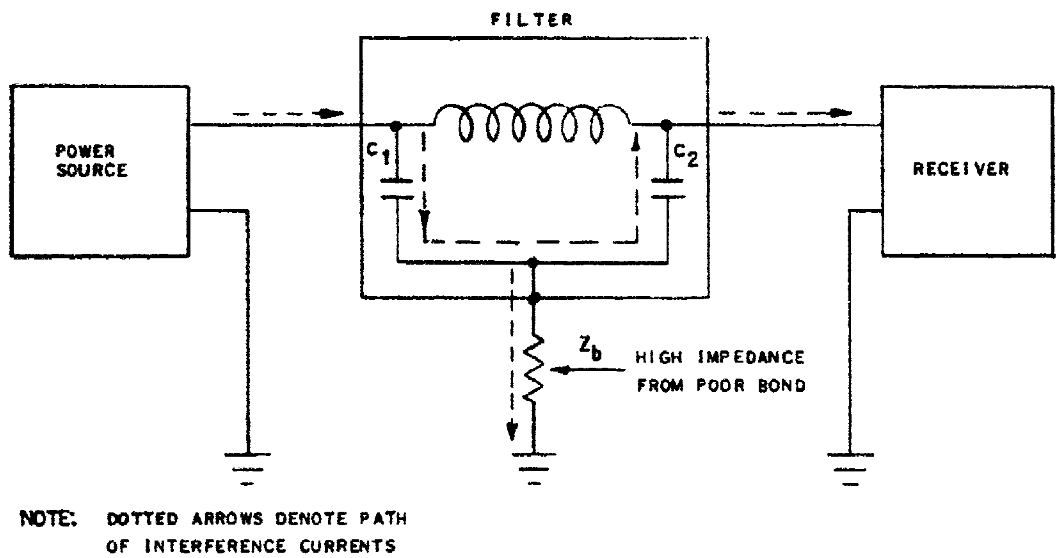


Figure 4-64. Effect of a Poor RF Ground Bond on Filter Effectiveness

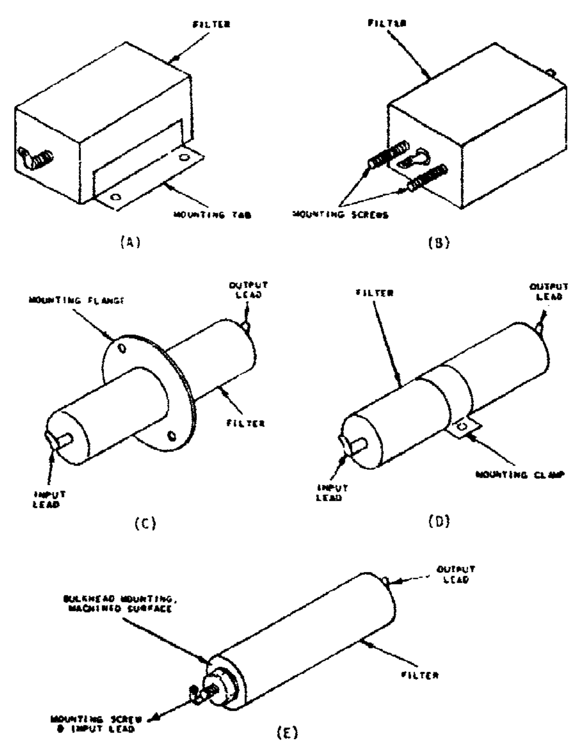


Figure 4-65. Typical Filter Mounting Techniques

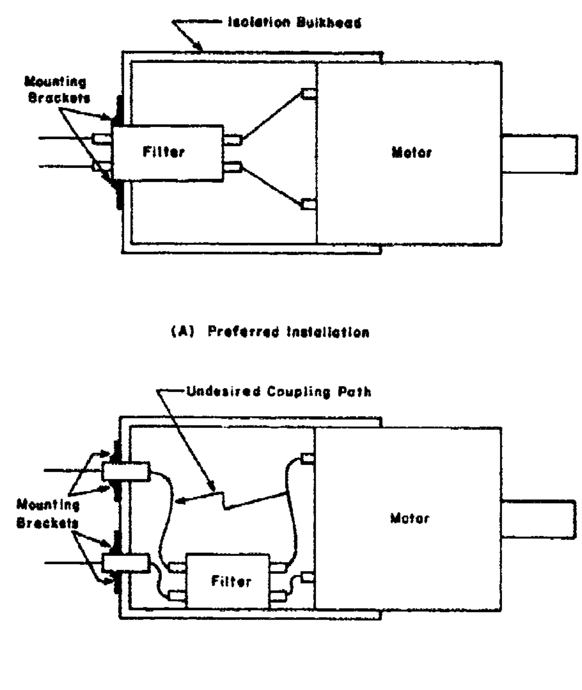


Figure 4-66. Filter Installation

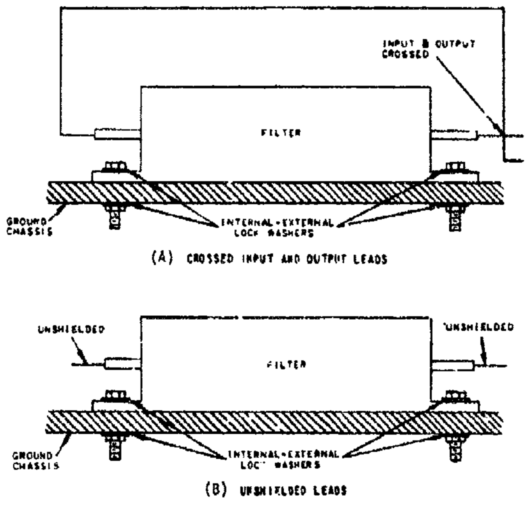


Figure 4-67. *Incorrect* **Filter Mounting Methods**

Fig. 4-67(B) isolation between input and output circuits is not complete due to lack of shielding on the leads — although there is the advantage of ease of assembly, and significant isolation can be obtained up to about 5 MHz. Proper installation of power line filters is shown on Fig. 4-68.

4-5.5.3 Connector Mounting

As mentioned in par. 4-5.2.1.6, very small low-pass filters which are quite effective for frequencies above about 1 MHz commonly are installed in pins on connectors. These may take the form of simple shunt capacitors between the wire and ground or more complex L-, T-, or π-arrangements. The series inductance is a ferrite-type material which also has loss which contributes to the attenuation. These filters are effective particularly in avoiding coupling of undesired energy from one equipment to another through the cable, or radiation of this energy from the cable.

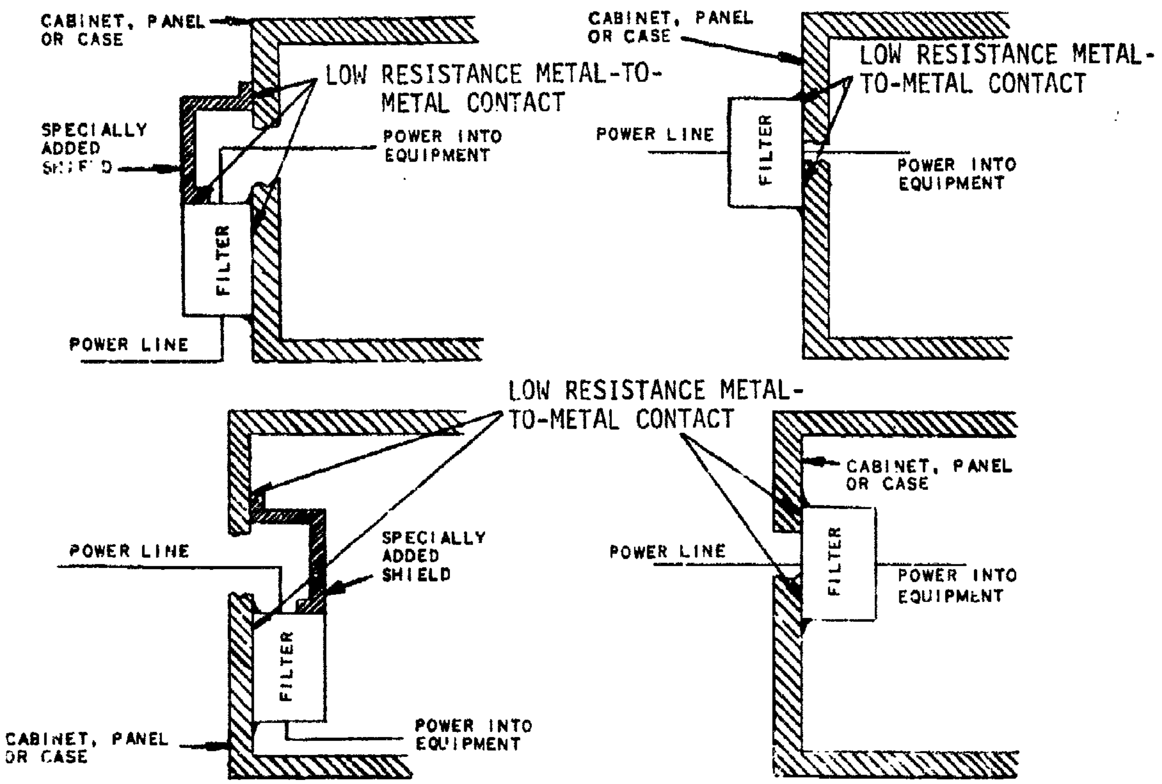


Figure 4-68. Installation of Power Line Filters

4-6. SHIELDING (Refs. 20, 21)

4-6.1 INTRODUCTION

It is often necessary to provide shielding to contain interference generated by particular electronic assemblies, or to protect such assemblies from external electromagnetic fields, especially when unshielded off-the-shelf items must be integrated into an electronic complex. The procedures for establishing shielding design are similar for any type of enclosure. All seams should be designed to prevent leakage (par. 4-6.5.3), and power circuits should be filtered (par. 4-5.5). Cable entry into the enclosure is critical (par. 4-7.1.3), access doors and removable panels should have gaskets (par. 4-6.5.4), and ventilating apertures should be covered with screening (par. 4-6.5.2).

Because shield shapes deviate from configurations readily analyzed, such as infinite planes or circular cylinders, theory gives attenuation values that are only approximations of those actually obtained. Where the shield entirely surrounds the object to be protected, the theoretical value is considered to be quite reliable. The usual procedure is to select a shielding material which, if perfect, would provide an adequate amount of protection in the application involved and then to design the penetrations so that the effectiveness is not degraded to any appreciable extent by them.

4-6.2 THEORETICAL CONSIDERATIONS

The properties of electromagnetic shields are very dependent on frequency. At frequencies below several kilohertz, the effectiveness of magnetic shields can be calculated using a quasi-static theory in which the determining factor is the reluctance of the path followed by lines of magnetic flux. Such shields are discussed in par. 4-6.6.1. At higher frequencies, and for nonmagnetic materials at all frequencies, a transmission line model is used in which skin depth plays a major role; the incident field is considered to be partially reflected and partially transmitted at the outer surface of the shield, and attenuated in passing through it (see Fig. 4-69) (Ref. 20). At the inner surface, partial transmission and reflection can occur again.

The shielding effectiveness can be defined in three basic ways: (1) by the ratio of the power densities in the field at the point in question before and after the shield is inserted, (2) by the corresponding ratio of electric fields, or (3) the corresponding ratio of magnetic fields. It is possible also to define it in terms of the ratio of field component of the wave incident on the shield front surface to that radiating from the other side. These definitions can produce different results in particular circumstances which are dependent on configuration, type of incident wave, angle of incidence, etc.

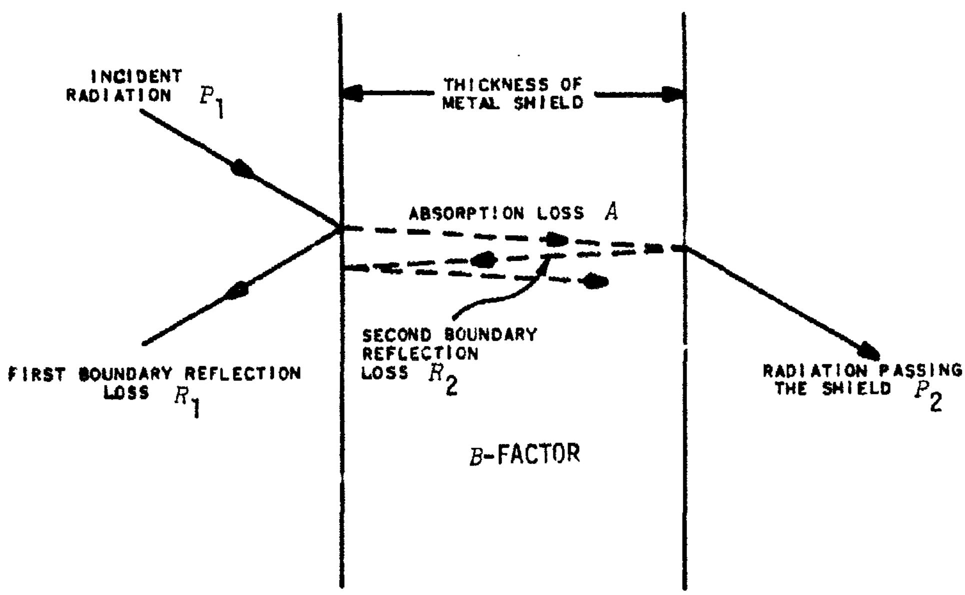


Figure 4-69. Metal Shielding Effectiveness (Ref. 20)

For a plane wave normally incident on a plane surface the definitions are unambiguous, giving the same result for all three definitions, i.e., expressing the shielding effectiveness SE in dB.

$$SE = 20 \log \left(\frac{E_1}{E_2}\right) = 20 \log \left(\frac{H_1}{H_2}\right)$$

$$= 10 \log \left(\frac{P_1}{P_2}\right), \text{ dB} \quad (4\text{-}15)$$

where

E_1, H_1, P_1 = incident electric field strength, magnetic field strength, power density, respectively

E_2, H_2, P_2 = corresponding quantities of field exiting from the shield

For the purpose of calculation, it has become conventional to express the shielding as the sum (in dB) of three parts as follows:

$$SE = R + A + B \text{ , dB} \quad (4\text{-}16)$$

where

$R = R_1 + R_2$ = reflection power loss of the first and second boundaries, dB

A = absorption power loss, dB

B = factor due to multiple reflection within the shield, which can be neglected if A is greater than 10, dB

A plane wave representation is a good approximation for flat surfaces if the source is remote from the surface. For such a source, the impedance of the incident wave Z_w, defined as the ratio E_1/H_1, is closely equal to the impedance of free space $\sqrt{\mu_0/\epsilon_0} = 377$ Ω. For dipole sources located closer to the source than $\lambda/(2\pi)$, the wave impedance varies with distance and is higher or lower than 377 Ω, depending on whether the source is of electric or magnetic type. Magnetic fields occur in the vicinity of coils or small loop antennas. Because reflection losses for magnetic fields are small for most materials, magnetic shielding depends primarily on absorption losses. Electric fields are stopped readily by metal shields because large reflection losses easily are obtained.

Absorption loss represents the reduction in signal due to dissipation as it proceeds through the body of a shield; for conventional materials, it is independent of the type of radiator emitting the signal. For thin shields, the reflection loss may be the most important loss.

The reflection loss R is given in terms of the impedance of the incident wave Z_w and that of the metal Z_s by

$$R = 20 \log \left[\frac{(Z_s + Z_w)^2}{4 Z_s Z_w}\right], \text{ dB} \quad (4\text{-}17)$$

where

$$Z_s = (1 + j)\sqrt{\frac{\mu_r f}{2g}} \times 3.69 \times 10^{-7}, \Omega \quad (4\text{-}18)$$

or

$$|Z_s| = \sqrt{\frac{\mu_r f}{g}} \times 3.69 \times 10^{-7}, \Omega \quad (4\text{-}19)$$

Thus, for a low impedance field

$$R = 20 \log \left[\frac{0.462}{\sqrt{\frac{fg}{\mu_r}}\, r} + 0.136 \sqrt{\frac{fg}{\mu_r}}\, r + 0.354\right], \text{ dB} \quad (4\text{-}20)$$

For a plane wave field

$$R = 168 - 10 \log \left(\frac{\mu_r f}{g}\right), \text{ dB} \quad (4\text{-}21)$$

For a high impedance field

$$R = 354 - 10 \log \left(\frac{\mu_r f^3}{g} r^2\right), \text{ dB} \quad (4\text{-}22)$$

where

μ_r = permeability relative to that of free space, dimensionless

g = conductivity relative to copper, dimensionless

f = frequency, Hz

r = distance from source to shield, in.

These relations are plotted in Figs. 4-70, 4-71, and 4-72 respectively (Ref. 21). The absorption loss A is given by (Fig. 4-73)

$$A = 3.34 \sqrt{\mu_r f g}\, t \text{ , dB} \quad (4\text{-}23)$$

where

t = material thickness, in.

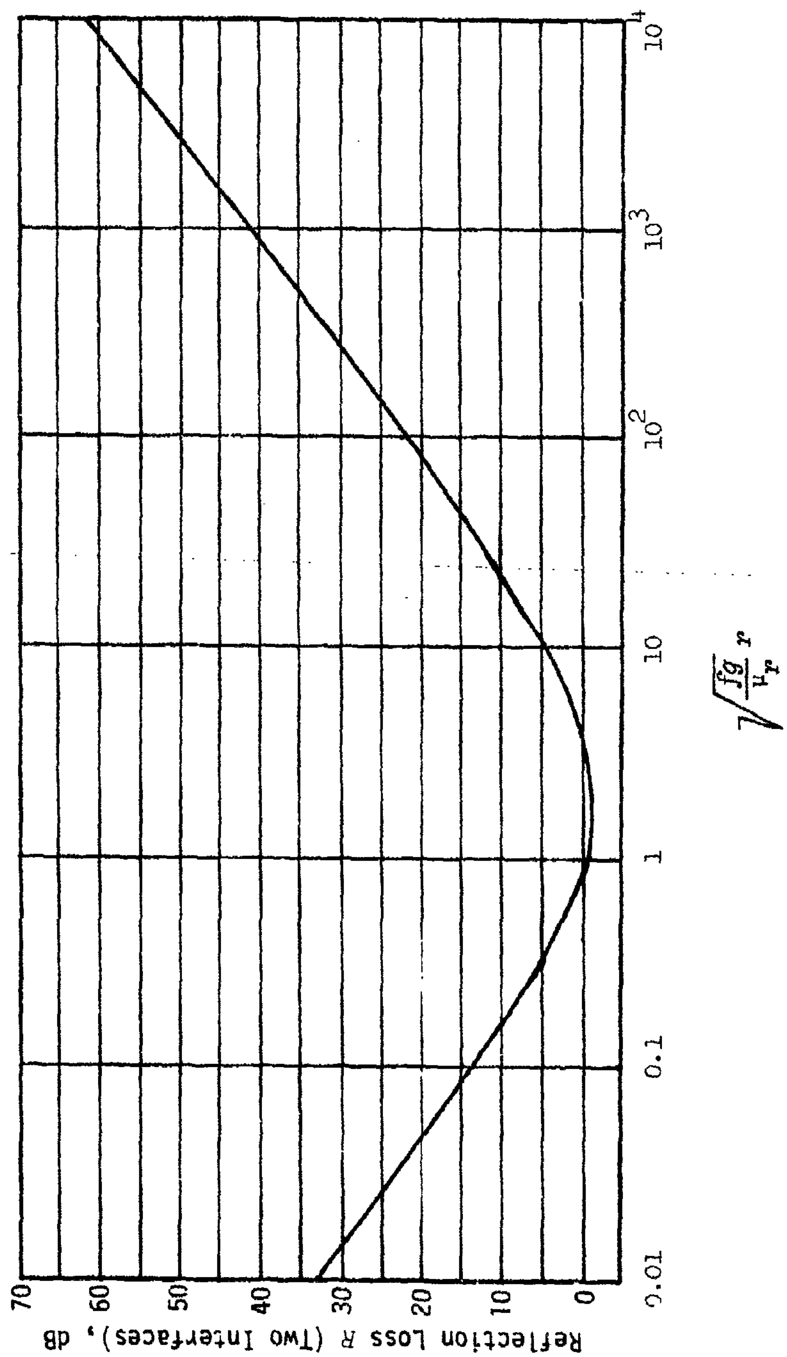


Figure 4-70. Reflection Loss for Low-impedance Source

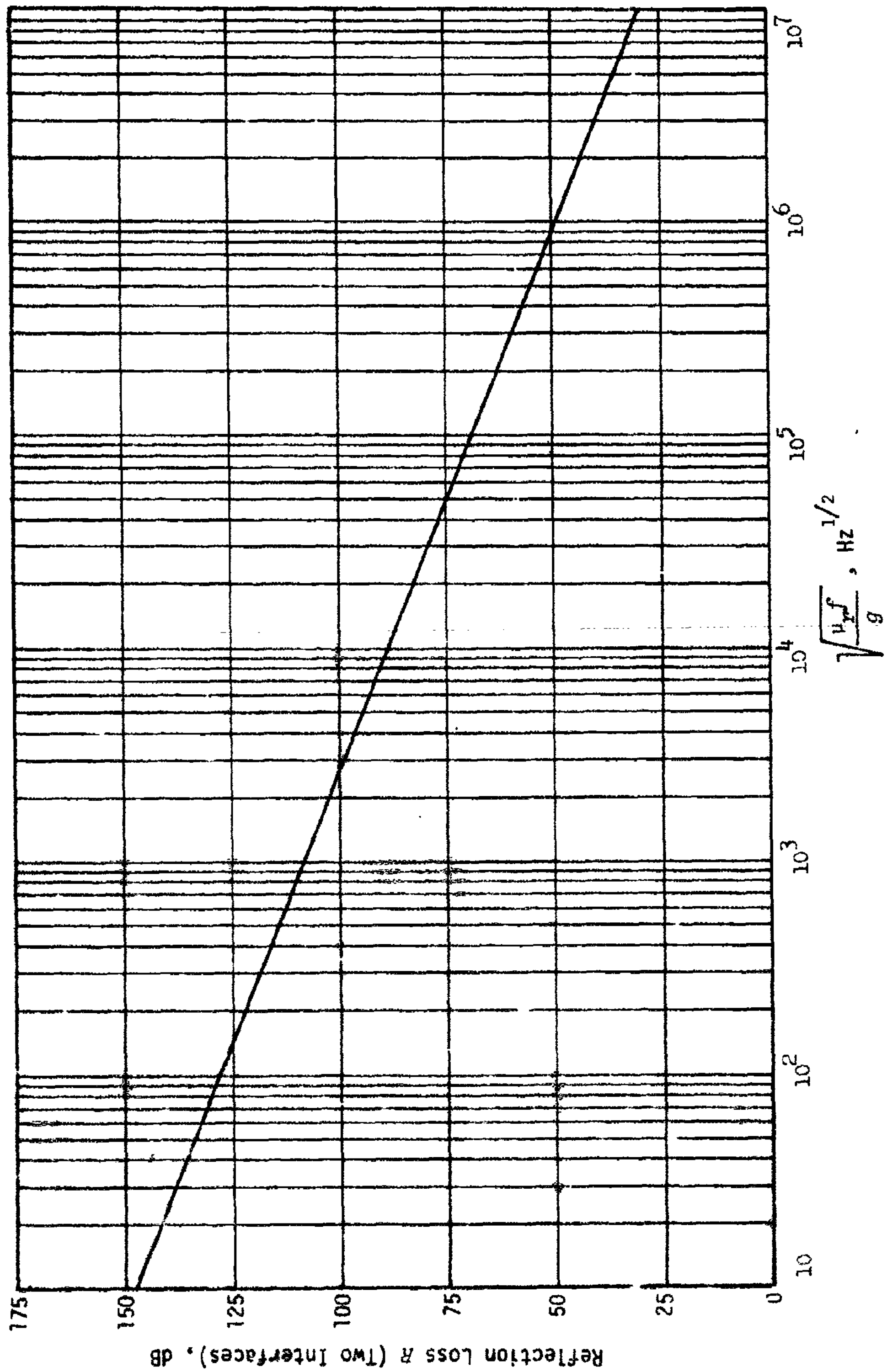


Figure 4-71. Reflection Loss to a Plane-wave

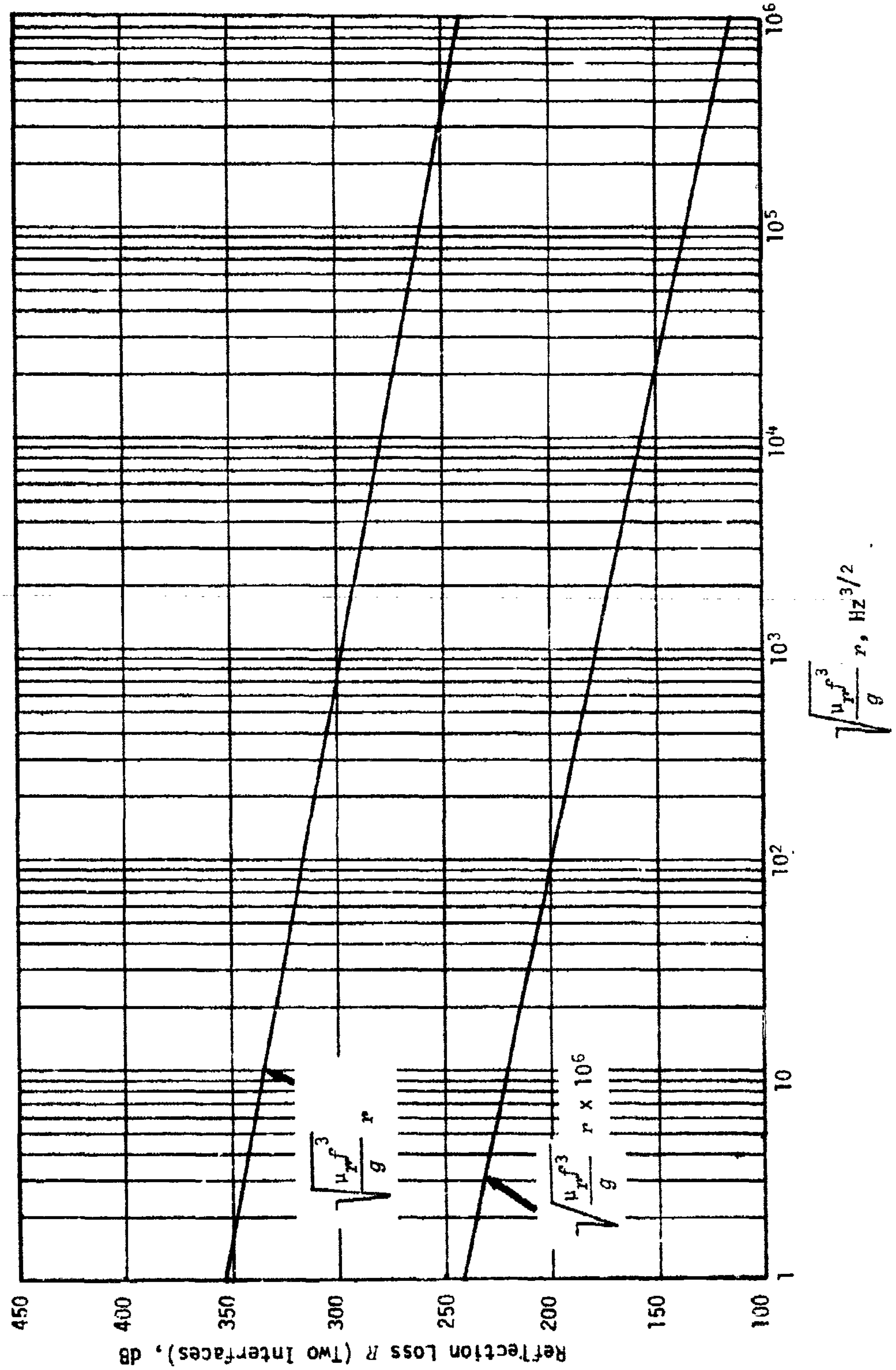


Figure 4-72. Reflection Loss for High-impedance Source

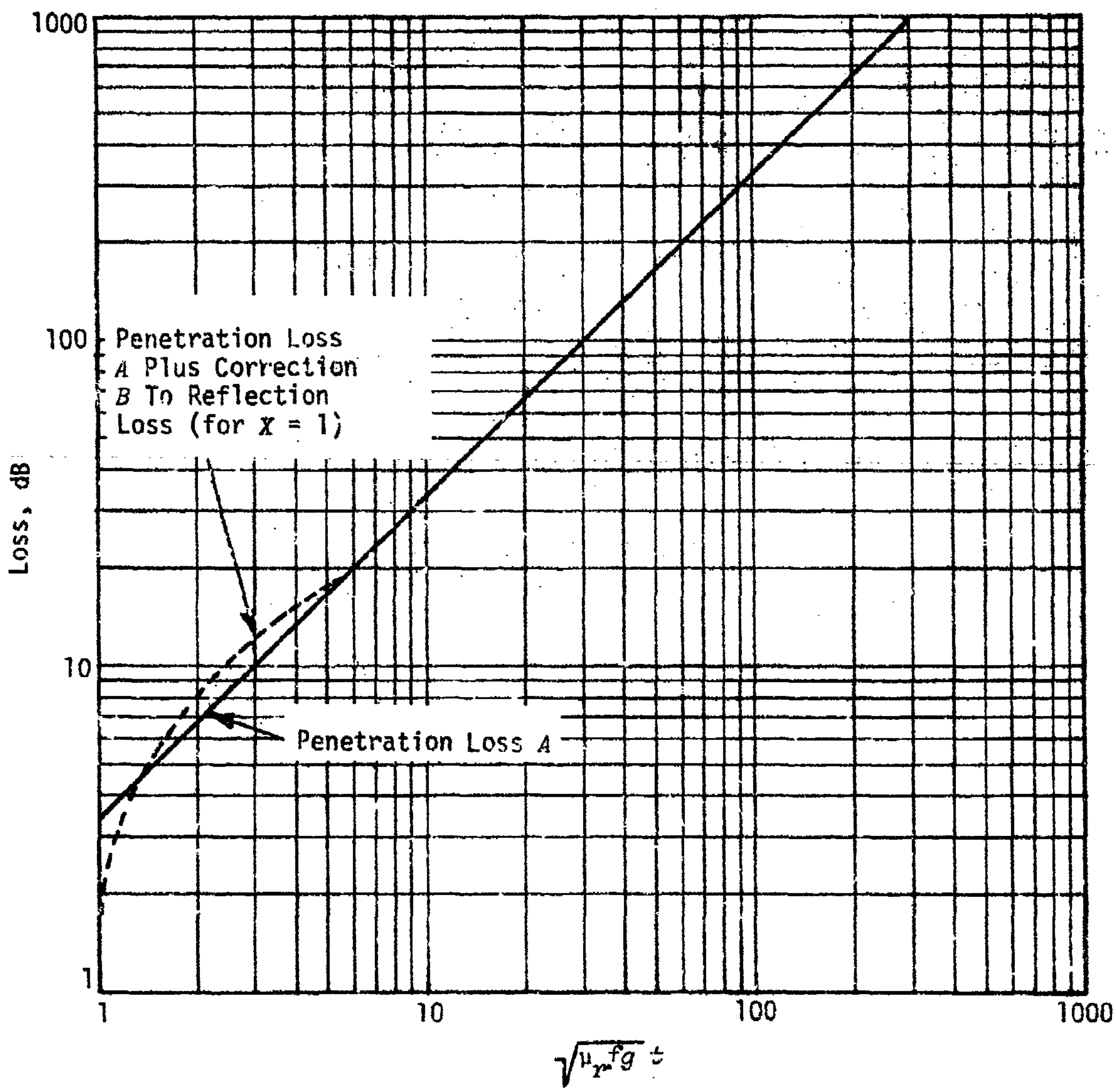


Figure 4.73. Absorption-loss Curve

The re-reflection loss B is given by

$$B = 20 \log\left[1 - X \times 10^{-A/10}(\cos 0.230 A - j \sin 0.230 A)\right], \text{dB} \qquad (4\text{-}24)$$

The correction factor X is given on Fig. 4-74 and is very nearly equal to 1, except in the case of low-frequency shielding against magnetic fields. A is given by Eq. 4-23. It should be noted that the value of μ_r is relative to free space and the value of g is relative to that of copper. Typical values of g and μ_r (at 150 kHz) are given in Table 4-5.

From these charts and figures one can determine the minimum combination of thickness, permeability, and conductivity which will have the effectiveness required. The solution of the shielding equation has been put into nomograph form (Refs. 28 and 29). Fig. 4-75 gives the absorption loss. As an example of its use, it shows that 10 mils of copper will produce an attenuation of 10 dB at 100 kHz. For materials other than those listed, the attenuation can be calculated in terms of appropriate values of $\mu_r g$, the product of the relative permeability and the conductivity. Also note that for magnetic materials, low frequency values of μ_r are assumed. Therefore, as frequency increases, adjustments in the point on the right-hand scale must be made to account for decreased effective values of μ_r.

Fig. 4-76 shows the magnetic field reflection loss, Fig. 4-77 electric field reflection loss and Fig. 4-78 plane wave reflection loss. Figs. 4-79 and 4-80 are used to account for re-reflection losses B for incident magnetic fields. Initially, Fig. 4-79 is used as shown for obtaining the appropriate value of K_ω, defined as the ratio of the shield impedance Z_s to the impedance of the incident magnetic field Z_ω and given by

$$K_\omega = \frac{Z_s}{Z_\omega} = \frac{1.3}{\sqrt{\frac{gf}{\mu_r}}\, r} \qquad (4\text{-}25)$$

TABLE 4-5. ABSORPTION LOSS OF METALS AT 150 kHz

Metal	g Relative Conductivity	μ_r Relative Permeability (at 150 kHz)	Absorption Loss (at 150 kHz) dB/mil
Silver	1.05	1	1.32
Copper, annealed	1.00	1	1.29
Copper, hard drawn	0.97	1	1.26
Gold	0.70	1	1.08
Aluminum	0.61	1	1.01
Magnesium	0.38	1	0.79
Zinc	0.29	1	0.70
Brass	0.26	1	0.66
Cadmium	0.23	1	0.62
Nickel	0.20	1	0.58
Phosphor-bronze	0.18	1	0.55
Iron	0.17	1000	16.9
Tin	0.15	1	0.50
Steel, SAW 1045	0.10	1000	12.9
Beryllium	0.10	1	0.41
Lead	0.08	1	0.36
Hypernick	0.06	80,000	88.5*
Monel	0.04	1	0.26
Nu-Metal	0.03	80,000	63.2*
Permalloy	0.03	80,000	63.2*
Stainless steel	0.02	1000	5.7

*Obtainable only if the incident field does not saturate the metal

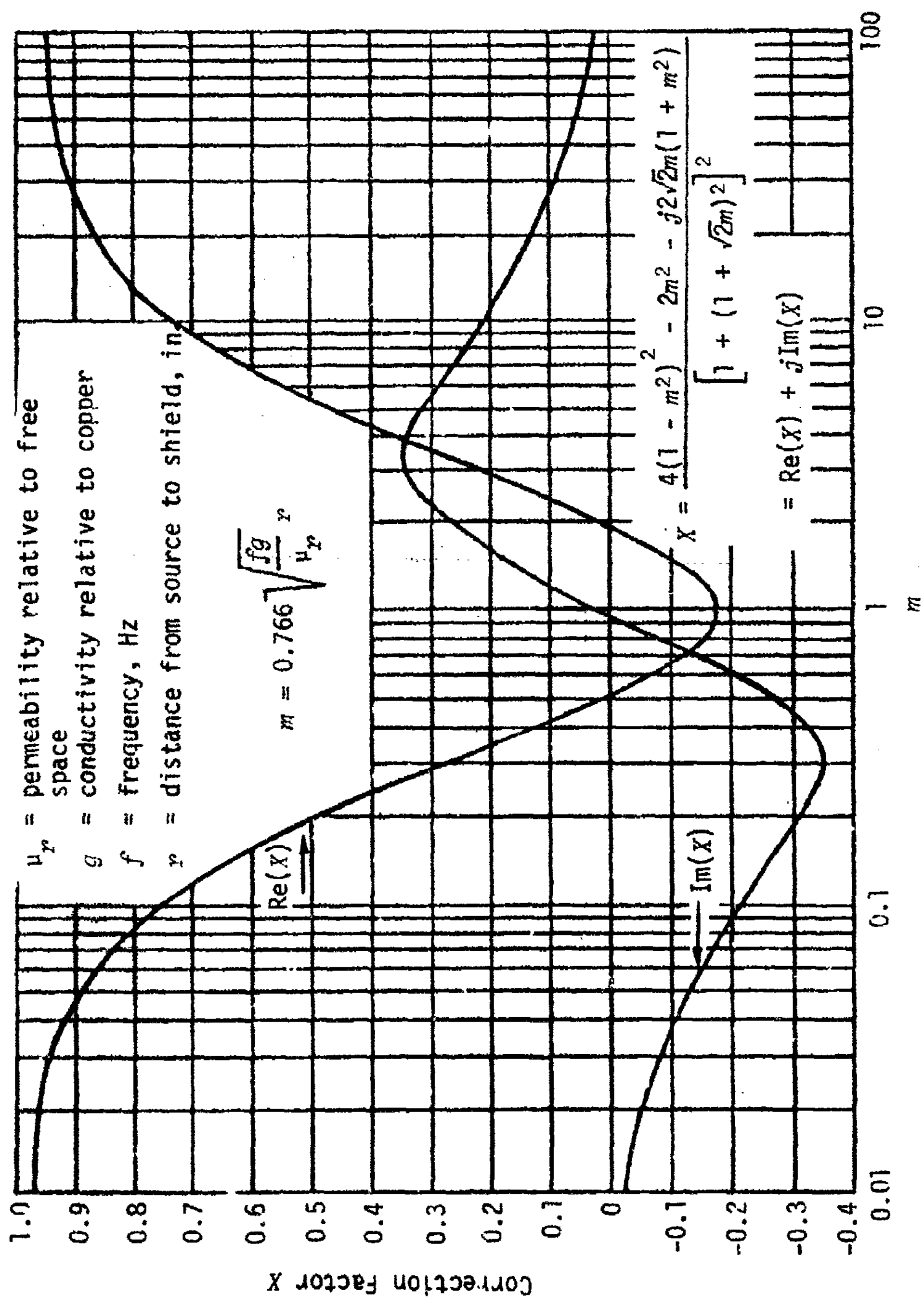


Figure 4-74. Correction Factor in Correction Term for Internal Reflections

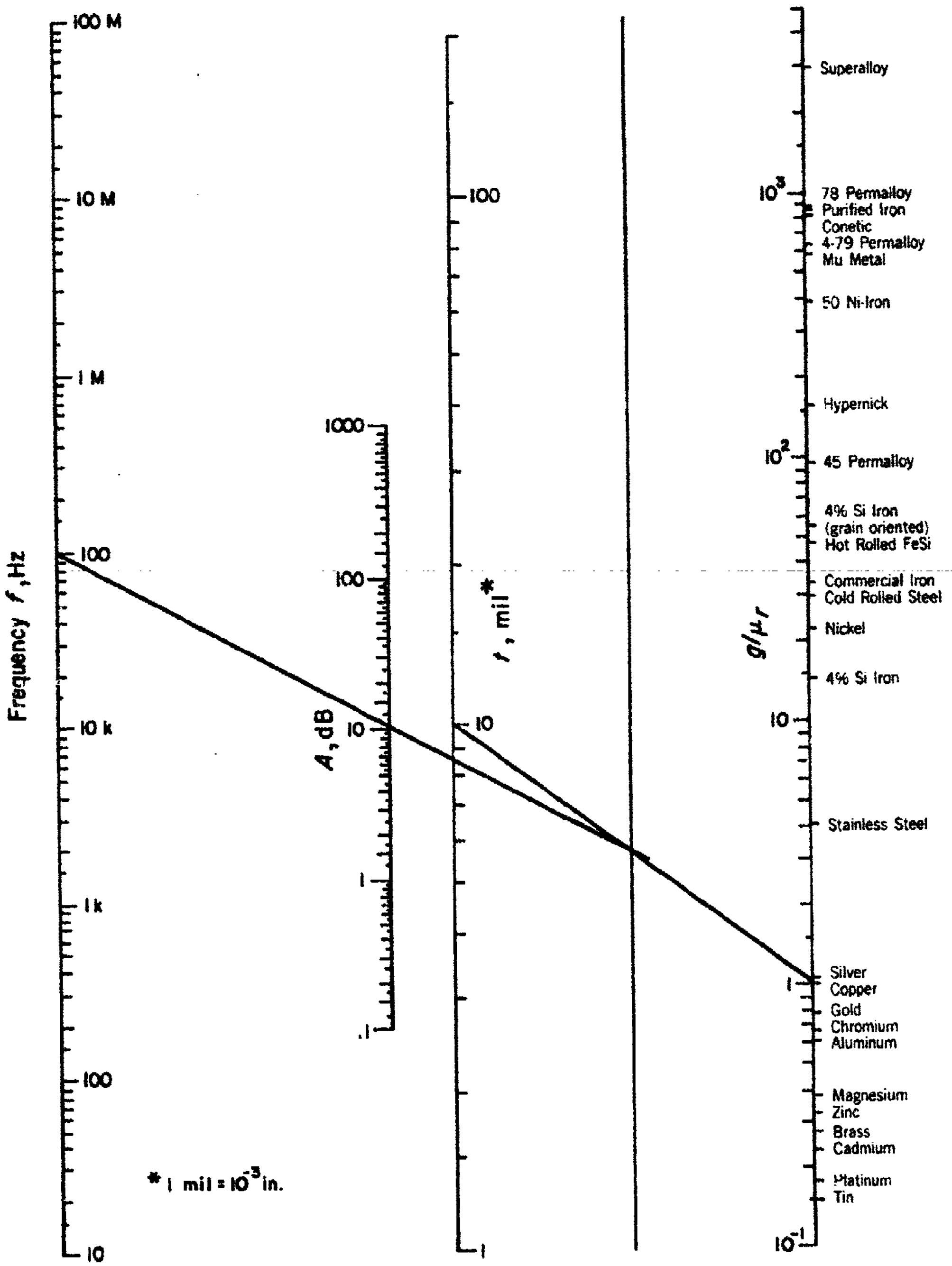


Figure 4-75. Absorption Loss *A*

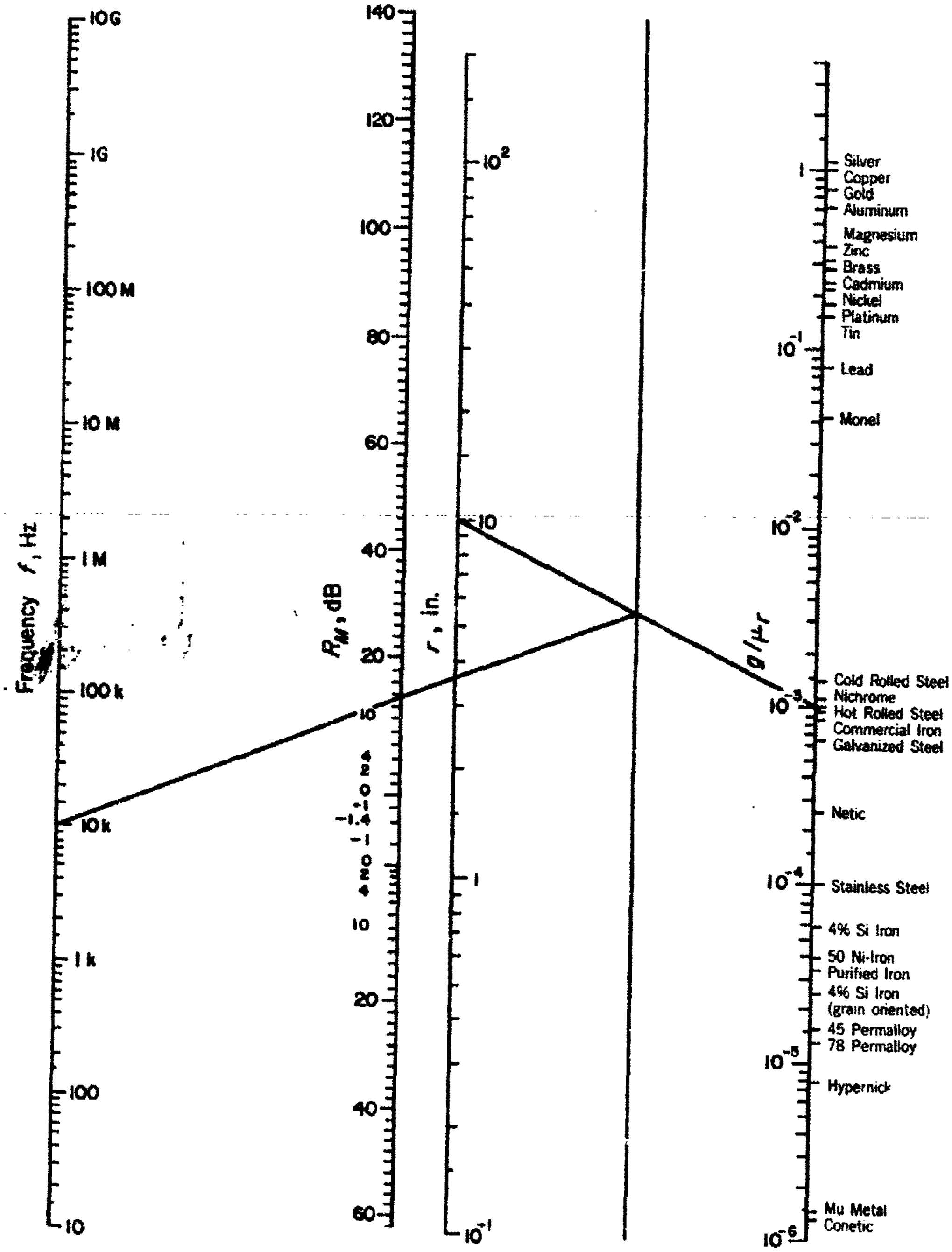


Figure 4-76. Magnetic Field Reflection Loss R_M

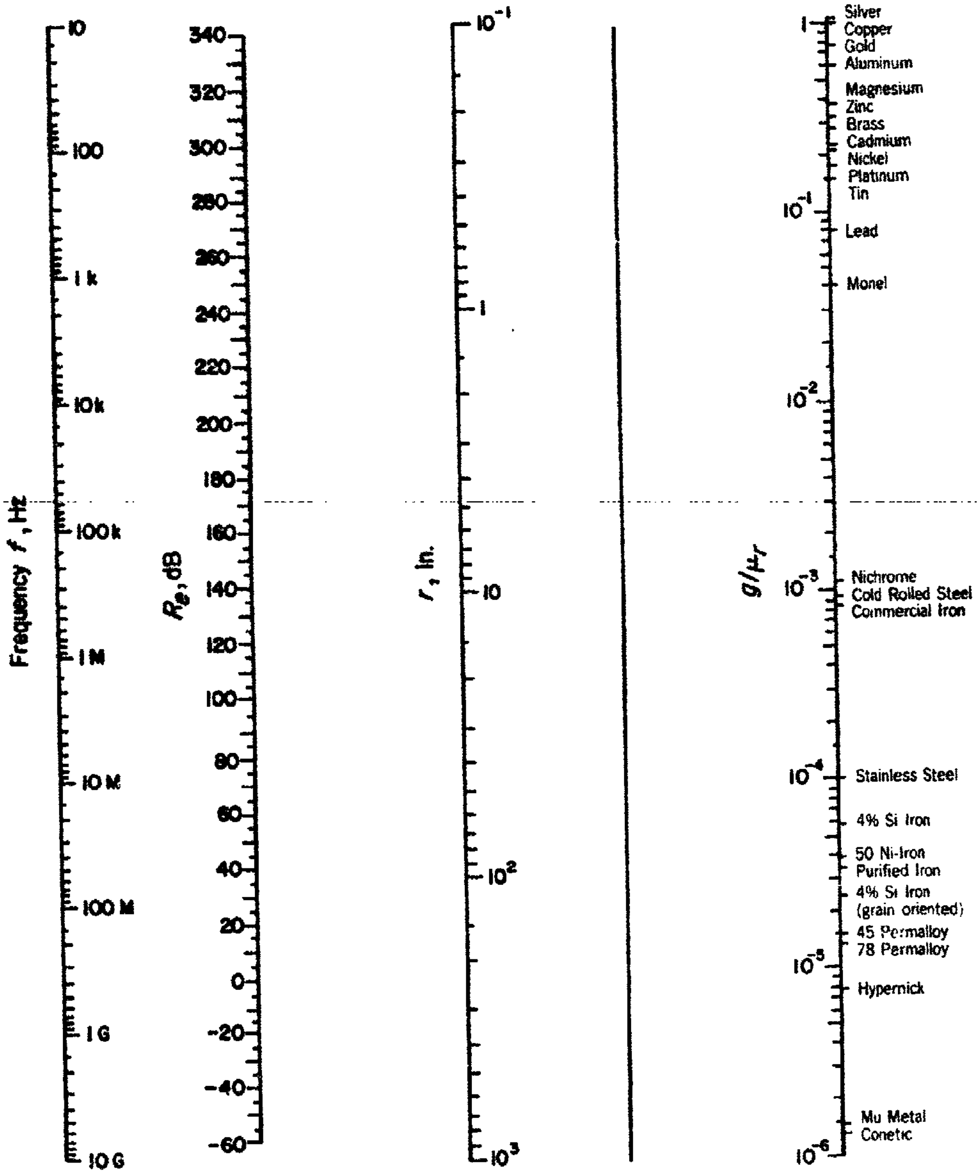


Figure 4-77. Electric Field Reflection Loss R_E

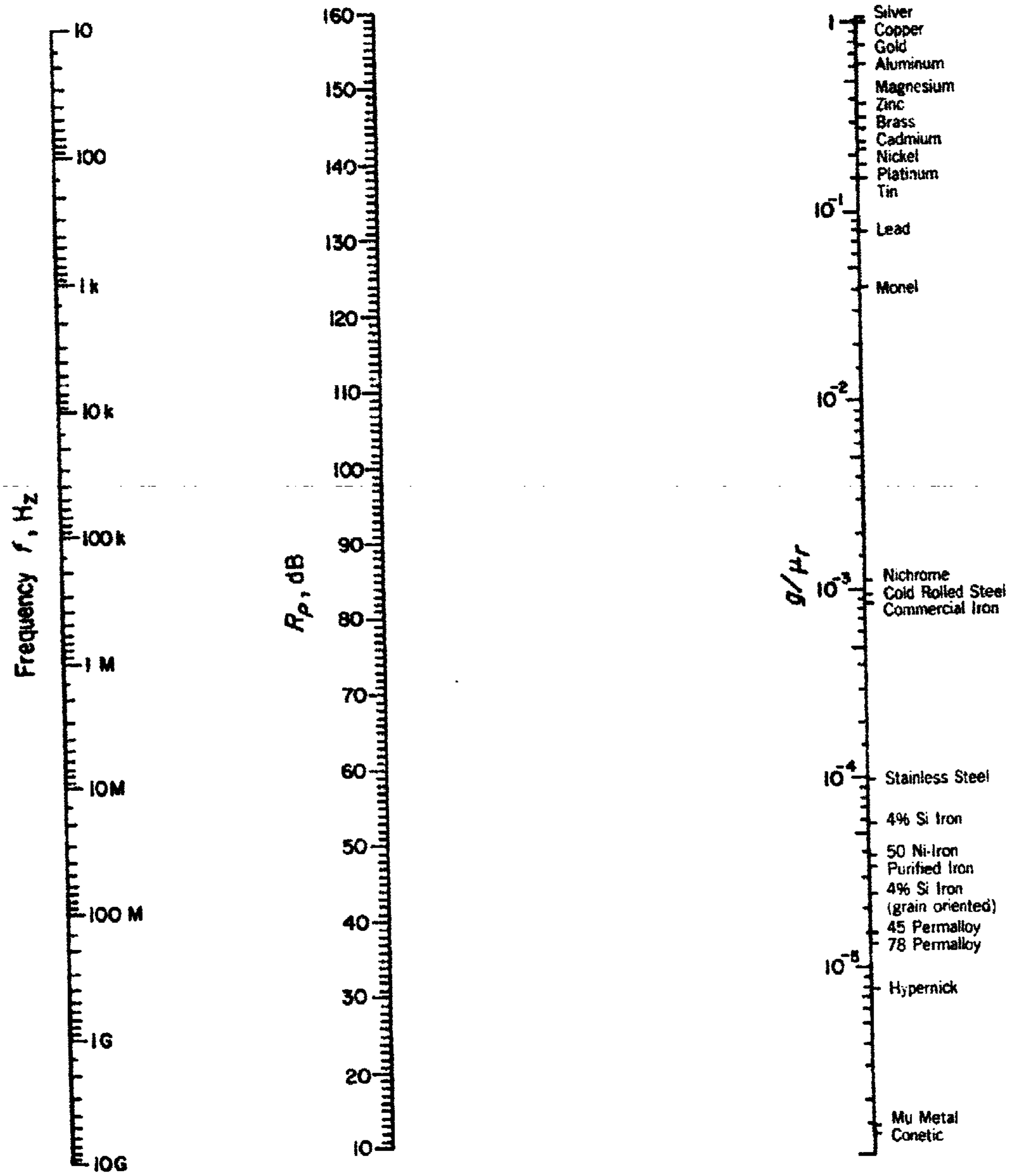


Figure 4-78. Plane Wave Reflection Loss R_P

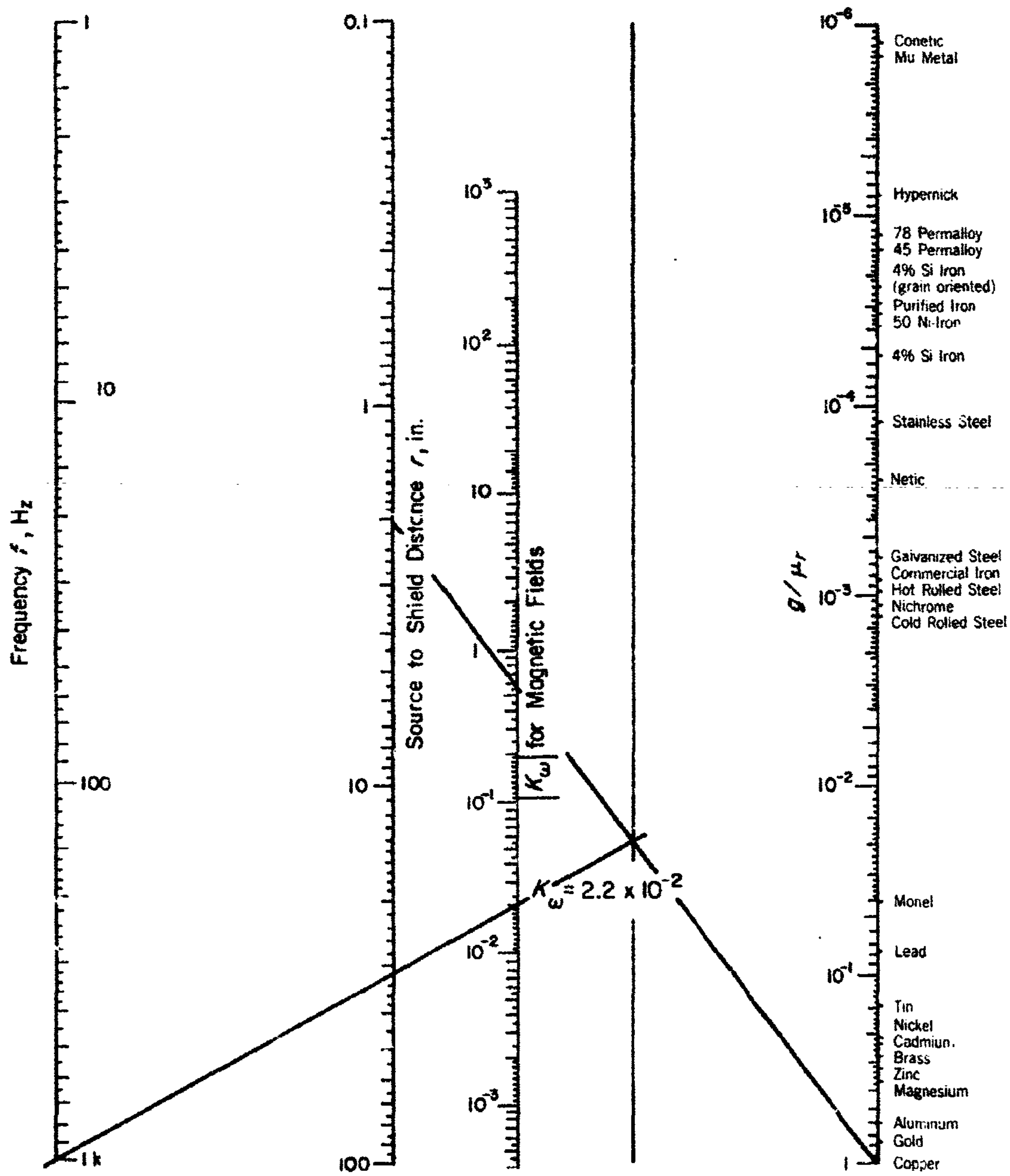


Figure 4-79. Chart for Computing K for Magnetic Field Secondary Reflection Loss

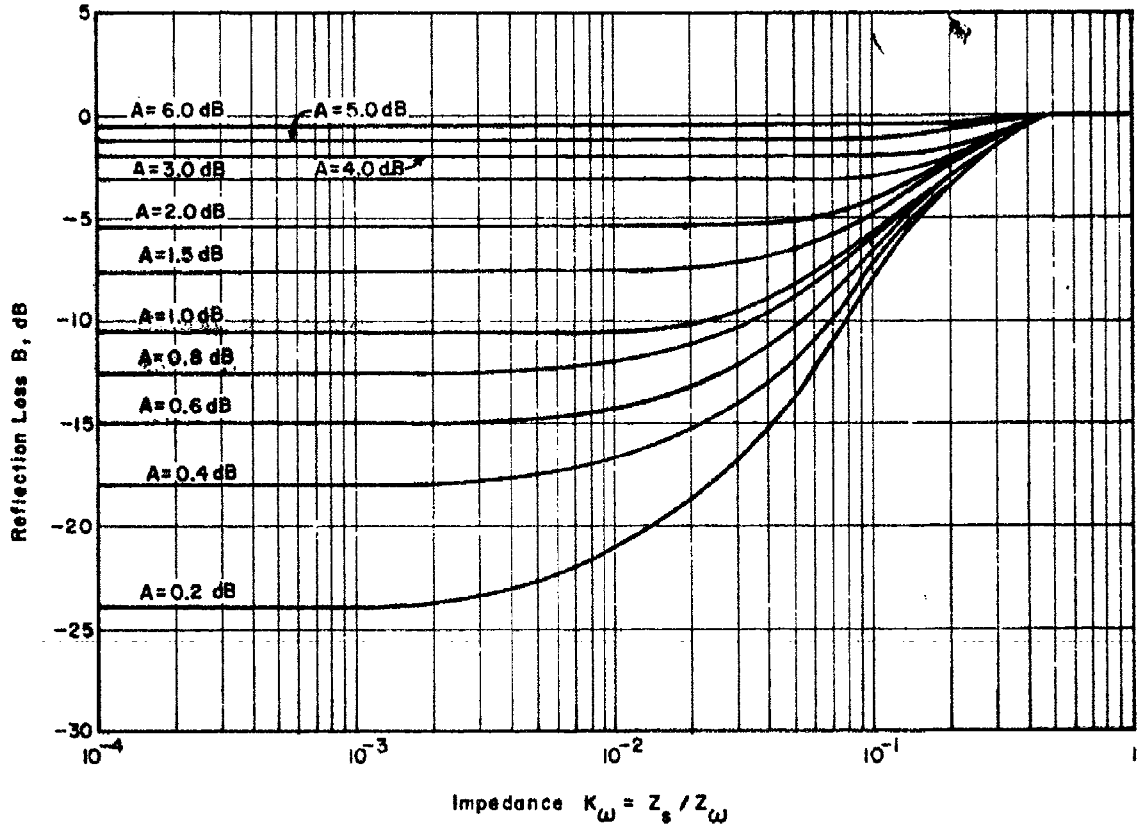


Figure 4-80. Chart for Computing Secondary Losses for Magnetic Fields

Then, Fig. 4-80 is used to obtain the value of B for the appropriate value of absorption loss A.

In some applications double shields are used. The calculations of the properties of double shields are fairly complex. If there is an air space between the surfaces of double shields, the total shielding effectiveness may be estimated by doubling that for a single shield. However, in critical cases a more careful examination of the matter is advisable since reflection effects between the surfaces may be substantial in the area between the shields. In the double shield, however, the effects of penetrations by leads become quite serious and one must be extremely careful in the design in order to obtain the maximum use of the double thickness of material.

4-6.3 DESIGN DATA

The next three paragraphs give data on absorption, reflection, and combined absorption and reflection losses as a function of frequency for various types of materials and thicknesses and for various types of incident wave. Par. 4-6.3.4 considers the effect of re-reflection (important for thin shields), and par. 4-6.3.5 gives examples of calculations of shielding effectiveness.

4-6.3.1 Absorption Loss

For a given metal the absorption loss in dB is proportional to the thickness. Values per mil are given in the third column of Table 4-5 for a frequency of 150 kHz.

Table 4-6 gives values of g and μ_r and that of absorption loss for copper, aluminum, and iron as a function of frequency f. Note that the absorption loss of Hypernick, at 150 kHz, is 88.5 dB per mil. It is cautioned, however, that the high permeability is useful only if the incident field is not of sufficient intensity to saturate the metal.

4-6.3.2 Reflection Loss

The reflection loss R to an electric field source that exists approximately 12 in. from the shield can be calculated using Eq. 4-22.

$$R = 354 + 10\log[g/(f^3\mu_r r_1^2)] \text{ , dB} \quad (4\text{-}26)$$

Values for copper, aluminum, and iron obtained using parameters values in Table 4-6 are given in Table 4-7. The reflection loss to a magnetic field source, located 12 in. from the shield, can be calculated using Eq. 4-20 and for a plane wave field by Eq. 4-21. Actual values are also listed in Table 4-7 for copper, aluminum, and iron.

4-6.3.3 Combined Losses

The combined absorption and reflection shielding effectiveness of copper, aluminum, and iron in magnetic fields is summarized in Table 4-8. Similarly, the comparative shielding effectiveness of copper and

TABLE 4-6. ABSORPTION LOSS OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS AT 60 Hz TO 10,000 MHz

Frequency	Copper		Aluminum		Iron		Absorption Loss, dB/mil		
	g	μ_r	g	μ_r	g	μ_r^a	Copper	Aluminum	Iron
60 Hz	1	1	0.61	1	0.17	1000	0.03	0.02	0.33
1000 Hz	1	1	0.61	1	0.17	1000	0.11	0.08	1.37
10 kHz	1	1	0.61	1	0.17	1000	0.33	0.26	4.35
150 kHz	1	1	0.61	1	0.17	1000	1.29	1.0	16.9
1 MHz	1	1	0.61	1	0.17	700	3.34	2.6	36.3
15 MHz	1	1	0.61	1	0.17	400	12.9	10	106.0
100 MHz	1	1	0.61	1	0.17	100	33.4	26	137.0
1500 MHz	1	1	0.61	1	0.17	10	129.0	100.0	168.0
10,000 MHz	1	1	0.61	1	0.17	1	334.0	260.0	137.0

[a] Other values of μ_r for iron are: 3 MHz, 600; 10 MHz, 508 and 1000 MHz, 50

TABLE 4-7. REFLECTION LOSS

	Electric Field[a], dB			Magnetic Field[a], dB			Plane Wave[b, d], dB		
	Copper	Aluminum	Iron	Copper	Aluminum	Iron	Copper	Aluminum	Iron
60 Hz	279	—	241	22	—	−1	150	148	113
1000 Hz	242	—	204	34	—	10	138	136	100
10 kHz	212	—	174	44	—	8	128	126	90
150 kHz	177	175	—	56	54	19	117	114	79
1 MHz	152	150	116	64	62	28	108	106	72
15 MHz	117	115	83	76	74	42	96	94	63
100 MHz	92	90	64	84	82	56	88	86	60
1500 MHz	c	—	c	c	—	c	76	74	57
10,000 MHz	c	—	c	c	—	c	68	66	60

[a] For signal source 12 in. from shield. Wave impedance much greater than 377 Ω. (For distances much greater or smaller than 12 in., recalculate the reflection loss using the formulas given in text.)
[b] If penetration loss is less than 10 dB total, reflection loss must be corrected by use of B-factor.
[c] At these frequencies, the fields approach plane waves with an impedance of 377 Ω.
[d] Signal source greater than 2λ from the shield.

iron for plane waves is summarized in Table 4-9; and of copper, aluminum, and iron for electric fields in Table 4-10. The curves of absorption and reflection loss for copper and iron are shown on Figs. 4-81 and 4-82 in electric, magnetic, and plane wave fields. The curves are plotted for signal sources that are 1 ft from the shield. As frequency increases, the 1-ft distance becomes a greater portion of a wavelength; as the electrical distance from the source increases, the radiation approaches a plane wave. Therefore, the three curves converge as frequency increases. Even for the poorest reflection loss curve for iron at 10 kHz, *A* is 4.35 dB per mil of thickness. A shield of iron 0.03 in. thick at this frequency would therefore have a theoretical shielding effectiveness of more than 130 dB.

TABLE 4-8

SHIELDING EFFECTIVENESS IN MAGNETIC FIELD (WAVE IMPEDANCE MUCH SMALLER THAN 377 Ω) OF SOLID COPPER, ALUMINUM, AND IRON SHIELDS FOR SIGNAL SOURCE 12 IN. FROM THE SHIELD AT 150 kHz TO 100 MHz

Frequency, MHz	Copper (10 mils)					Aluminum (10 mils)					Iron (10 mils)				
	A (dB)	+	*R* (dB)	=	*SE* (dB)	*A* (dB)	+	*R* (dB)	=	*SE* (dB)	*A* (dB)	+	*R* (dB)	=	*SE* (dB)
0.15	13	+	56	=	69	10	+	54	=	64	169	+	19	=	188
1.0	33	+	64	=	97	26	+	62	=	88	363	+	28	=	391
15	129	+	76	=	205	100	+	74	=	174	1060	+	42	=	1102
100	334	+	84	=	418	260	+	82	=	342	1370	+	56	=	1426

TABLE 4-9

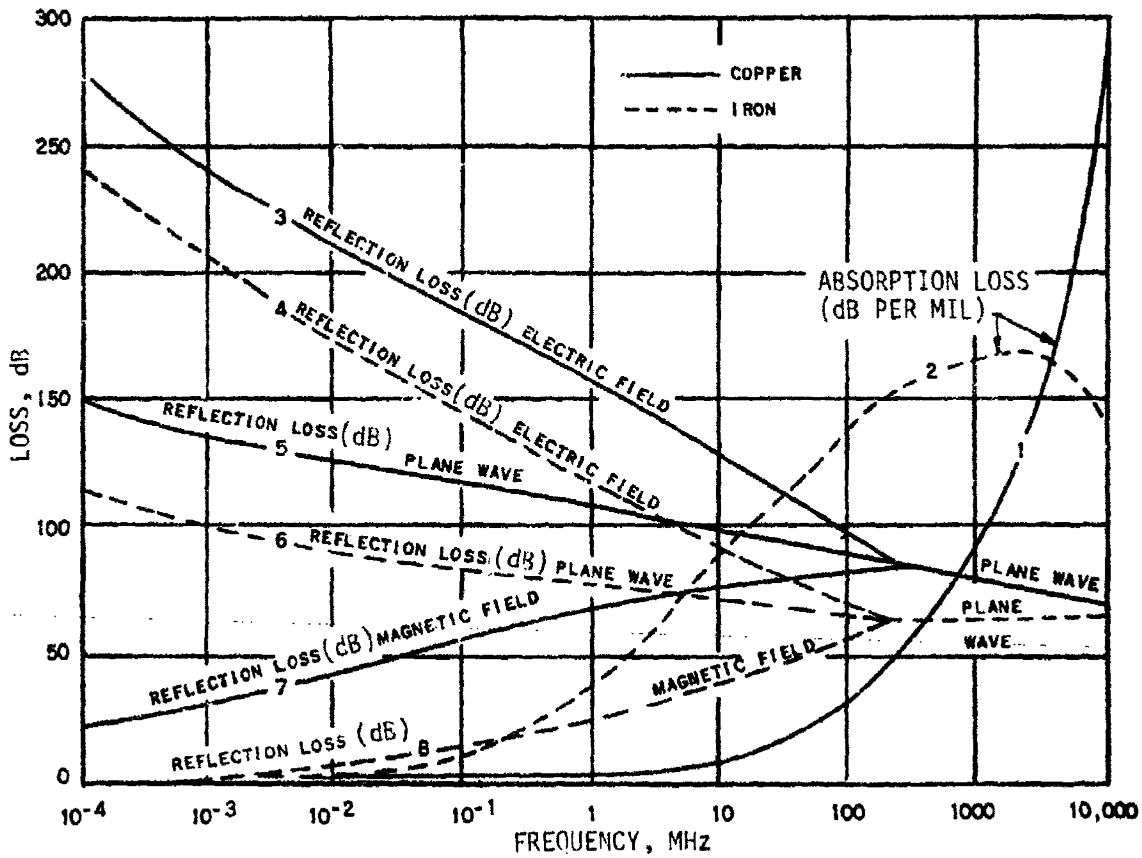
SHIELDING EFFECTIVENESS IN PLANE WAVE FIELD (WAVE IMPEDANCE EQUAL TO 377 Ω) OF SOLID COPPER AND IRON SHIELDS FOR SIGNAL SOURCE GREATER THAN 2 λ FROM THE SHIELD AT 150 kHz TO 100 MHz

Frequency, MHz	Copper (10 mils)					Iron (10 mils)				
	A (dB)	+	*R* (dB)	=	*SE* (dB)	*A* (dB)	+	*R* (dB)	=	*SE* (dB)
0.15	13	+	117	=	130	169	+	79	=	248
1.0	33	+	108	=	141	363	+	72	=	435
15	129	+	96	=	225	1060	+	63	=	1123
100	334	+	88	=	422	1370	+	60	=	1430

TABLE 4-10

SHIELDING EFFECTIVENESS IN ELECTRIC FIELD (WAVE IMPEDANCE MUCH GREATER THAN 377 Ω) OF SOLID COPPER, ALUMINUM AND IRON SHIELDS FOR SIGNAL SOURCE 12 IN. FROM THE SHIELD AT 0.15 MHz TO 100 MHz

Frequency, MHz	Copper (10 mils)					Aluminum (10 mils)					Iron (10 mils)				
	A (dB)	+	*R* (dB)	=	*SE* (dB)	*A* (dB)	+	*R* (dB)	=	*SE* (dB)	*A* (dB)	+	*R* (dB)	=	*SE* (dB)
0.15	13	+	176	=	189	10	+	175	=	185	169	+	139	=	308
1.0	33	+	152	=	185	26	+	150	=	176	363	+	116	=	479
15.0	129	+	116	=	245	100	+	115	=	215	1060	+	83	=	1143
100	334	+	92	=	426	260	+	90	=	350	1370	+	64	=	1434



1 AND 2 = ABSORPTION LOSS PER MIL THICKNESS OF METAL
3 AND 4 = REFLECTION LOSS - ELECTRIC WAVES
5 AND 6 = REFLECTION LOSS - PLANE WAVES
7 AND 8 = REFLECTION LOSS - MAGNETIC FIELDS

SHIELDING EFFECTIVENESS = REFLECTION LOSS + ABSORPTION LOSS

SE = R + A

FOR COPPER: ELECTRIC FIELD; SE = CURVE 3 + CURVE 1 × "t" (THICKNESS IN MILS)*
MAGNETIC FIELD; SE = CURVE 7 + CURVE 1 × "t" (THICKNESS IN MILS)*
PLANE WAVE; SE = CURVE 5 + CURVE 1 × "t" (THICKNESS IN MILS)*
FOR IRON: ELECTRIC FIELD; SE = CURVE 4 + CURVE 2 × "t" (THICKNESS IN MILS)*
MAGNETIC FIELD; SE = CURVE 8 + CURVE 2 × "t" (THICKNESS IN MILS)*
PLANE WAVE; SE = CURVE 6 + CURVE 2 × "t" (THICKNESS IN MILS)*

* IF THE SHIELD IS ELECTRICALLY THIN (A LESS THAN 10 dB), THEN CALCULATE THE B-FACTOR AND INCLUDE IN THE SHIELDING EFFECTIVENESS EQUATION (SE = R + A + B)

Figure 4-81. Shielding Effectiveness in Electric, Magnetic, and Plane Wave Fields of Solid Copper and Iron Shields (for Signal Source 12 in. from Shield at 100 Hz To 10 GHz)

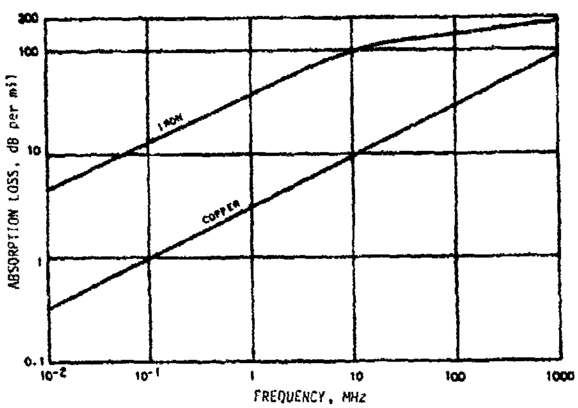


Figure 4-82. Absorption Loss for Copper and Iron, dB per mil

4-6.3.4 RE-REFLECTION LOSS

If the shield is electrically thin (A less than 10 dB), then the B-factor should be calculated and included in the shielding effectiveness equation ($SE = R + A + B$). The B-factors for copper and iron in an electrical field, magnetic field, and for a plane wave are shown in Table 4-11.

4-6.3.5 EXAMPLES OF SHIELDING EFFECTIVENESS CALCULATIONS

To illustrate the application of the data, three examples of shielding effectiveness calculations are given.

a. *Example No. 1.* The results of shielding effectiveness calculations for copper 7 mils thick, and steel 1 mil and 50 mils thick, in electric, magnetic, and plane wave fields at a distance of 165 ft are plotted on Figs. 4-83, 4-84, and 4-85 and summarized in Tables 4-12, 4-13, and 4-14. The shielding effectiveness is calculated by using Eq. 4-16: $SE = R + A + B$. R can be calculated for several distances for the electric, magnetic, and plane wave fields (Tables 4-15, 4-16, and 4-17, respectively). The values for steel and copper are plotted on Figs. 4-86 and 4-87. For examples, at 10 kHz, for steel in an electric field at a distance of one mile, Fig. 4-86 shows 98 dB of reflection loss. At 150 kHz, for copper in a magnetic field at a distance of 24 in., Fig. 4-87 shows 62 dB of reflection loss. The absorption loss A can be calculated for each thickness of metal. In Table 4-18 the values for steel (1 mil and 50 mils thick) and copper (7 mils thick) are given. These values are plotted on Fig. 4-88. The B-factors are summarized in Table 4-19.

b. *Example No. 2.* The shielding effectiveness of steel (1 mil thick) at 30 Hz in an electric field 165 ft distant from the source is determined as follows:

$$SE = R + A + B$$

where

R = 203 dB (from Table 4-16)
A = 0.2 dB (from Table 4-18)
B = −27.0 dB (from Table 4-19)

SE = 176.2 dB.

Obviously a shield can be designed so that there is an appropriate division between reflection loss and absorption loss. The reflection loss in the electric field decreases inversely with frequency and approaches the plane wave reflection loss curve, while the reflection loss in the magnetic field increases inversely with frequency and approaches the plane wave reflection curve (Fig. 4-81). The absorption loss of iron at 10 MHz is about 70 dB greater than the magnetic field reflection loss, but, at 100 kHz, it is a few dB less than the reflection loss. These factors should be fully considered when designing metal shields. For example, if shielding against electric fields at frequencies of about 10 kHz is necessary, the tremendous reflection loss (over 200 dB on curve 3 of Fig. 4-81) should be used rather than the absorption loss alone. To summarize, shielding effectiveness is the result of $R + A + B$ together, and not any single part. At frequencies as low as 60 Hz, absorption loss and reflection loss become negligible for magnetic fields so that, for lower frequencies, very thick metallic barriers may be necessary to shield against magnetic fields.

c. *Example No. 3.* Based on the examples given in Table 4-20, a metallic barrier made of iron with a thickness of 300 mils must be provided to obtain a shielding effectiveness of 100 dB at 60 Hz for magnetic fields. For copper and iron, reflection and penetration losses are small for magnetic fields at low frequencies. Since magnetic materials, such as Mu-metal, have high permeability at low frequencies, and therefore high absorption loss, they are most effective as shields. The resulting increase in absorption loss is obtained at the expense of a reduction in reflection loss. Following are general rules for selection of shielding materials:

a. Good conductors such as copper, aluminum, and magnesium should be used for high-frequency shields to obtain the highest reflection loss.

b. Magnetic materials such as iron and Mu-metal should be used for low-frequency shields to obtain the highest absorption loss.

c. Any structurally sound shielding material will usually be thick enough for shielding electric fields at any frequency.

d. To provide a given degree of shielding, reference to the curves of absorption loss permit quick estimates of the required metal and thickness. In most applications, it is necessary to reduce an interference field to some specified level. This shielding requirement can be determined by measuring the interference field without the shield and comparing it with the specification limits.

4-6.4 MULTIPLE SHIELDING

4-6.4.1 General

The shielding requirements necessary to protect against magnetic field transference may be difficult to achieve at low frequencies. In the normal power-frequency range, for example, copper must be very thick to serve as a practical magnetic shield. Mu-metal and similar type high-permeability alloys provide good shielding for weak fields; multiple magnetic shielding is recommended for strong fields.

TABLE 4-11. B-FACTORS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF SOLID COPPER AND IRON SHIELDS

Shield Thickness, mils	60 Hz	100 Hz	1 kHz	10 kHz	100 kHz	1 MHz
Copper, $\mu_r = 1, g = 1$, Magnetic Fields						
1	−22.22	−24.31	−28.23	− 9.61	−10.34	−2.61
5	−21.30	−22.07	−15.83	− 6.98	− 0.55	+0.14
10	−19.23	−18.59	−10.37	− 2.62	+ 0.57	0
20	−15.35	−13.77	− 5.41	+ 0.13	− 0.10	
30	−12.55	−10.76	− 2.94	+ 0.58	0	
50	− 8.88	− 7.07	− 0.58	0		
100	− 4.24	− 2.74	+ 0.50			
200	− 0.76	+ 0.05	0			
300	+ 0.32	+ 0.53				
Copper, $\mu_r = 1, g = 1$, Electric Fields and Plane Waves						
1	−41.52	−39.31	−29.38	−19.61	−10.33	−2.61
5	−27.64	−25.46	−15.82	− 6.96	− 0.55	+0.14
10	−21.75	−19.61	−10.33	− 2.61	+ 0.57	0
20	−15.99	−13.92	− 5.37	+ 0.14	− 0.10	
30	−12.73	−10.73	− 2.90	+ 0.58	0	
50	− 8.81	− 6.96	− 0.55	+ 0.14		
100	− 4.08	− 2.61	+ 0.51	0		
200	− 0.62	+ 0.14	0			
300	+ 0.41	+ 0.58				
Iron, $\mu_r = 1000, g = 0.17$, Magnetic Fields						
1	+ 0.95	+ 1.23	− 1.60	− 1.83		
5	+ 0.93	+ 0.89	− 0.59	0		
10	+ 0.78	+ 0.48	+ 0.06			
20	+ 0.35	+ 0.08	0			
30	+ 0.06	− 0.06				
50	0	0				
Iron, $\mu_r = 1000, g = 0.17$, Electric Fields and Plane Waves						
1	−19.53	−17.41	− 8.35	− 1.31		
5	− 6.90	− 5.17	+ 0.20	0		
10	− 2.56	− 1.31	+ 0.36			
20	+ 0.16	+ 0.54	0			
30	+ 0.58	+ 0.42				
50	+ 0.13	0				

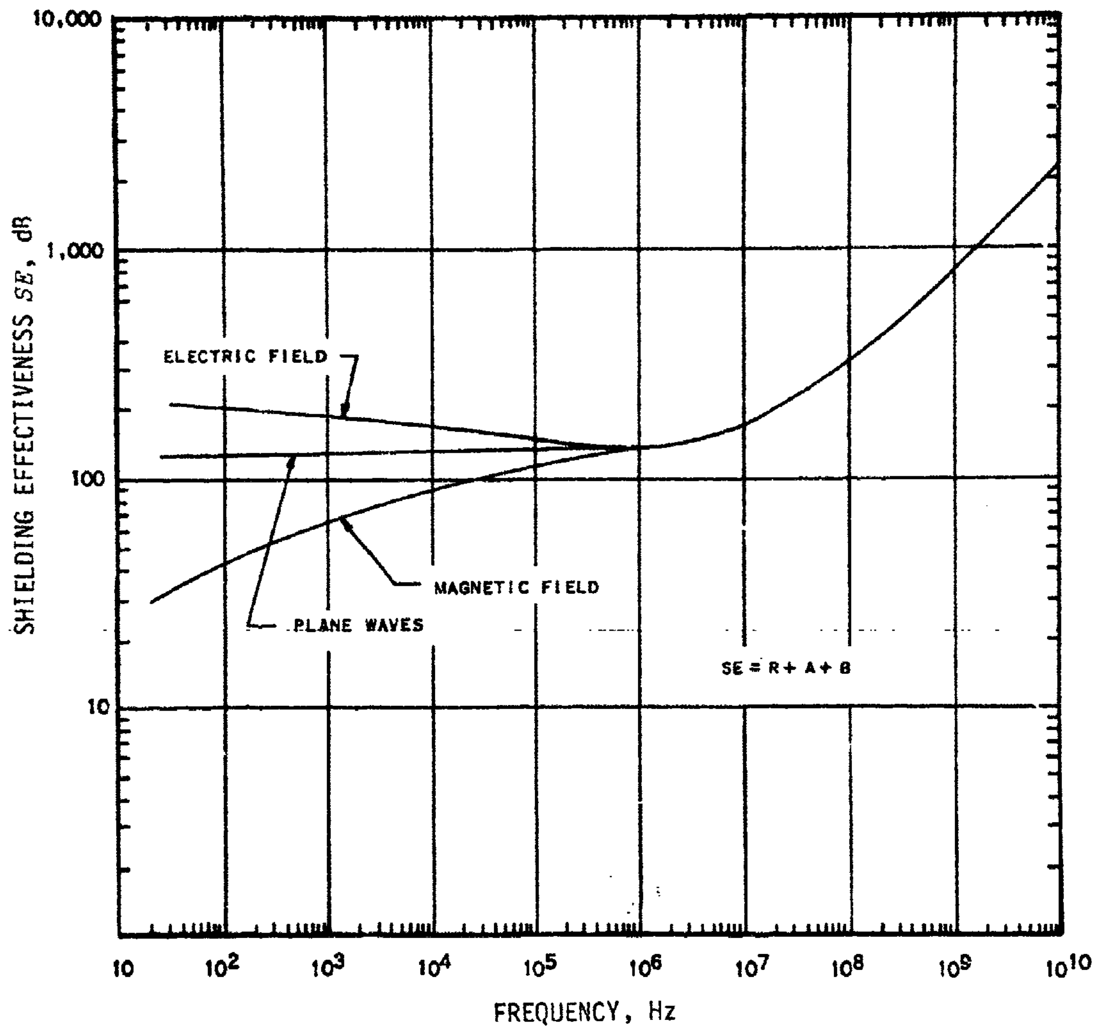


Figure 4-83. Shielding Effectiveness in Electric, Magnetic, and Plane Wave Fields of Copper Shields (7 mil Thickness) for Signal Sources 165 ft from the Shield

Power transformers and audio transformers, mounted near each other, may require multiple shielding. At high frequencies, the shielding effectiveness is good because of reflections from the surface and rapid dissipation of the field by penetration losses. At low frequencies, it is also possible to obtain reflection losses in magnetic materials. When much of the usefulness of shielding is due to reflection losses, two or more layers of metal — separated by dielectric materials, and yielding multiple reflections — will provide greater shielding than the same amount of metal in a single sheet. Copper, Mu-metal, iron, conetic and netic type materials, and other metals — some with excellent electric-field reflection loss and some with excellent magnetic field absorption-loss properties — can be used effectively in combination. In many applications, it is possible to reduce shielding effectiveness requirements for the overall equipment housing by employing suppression techniques within the equipment. The recommended techniques make use of component shields, filters at the source of the undesired signal interference, partial

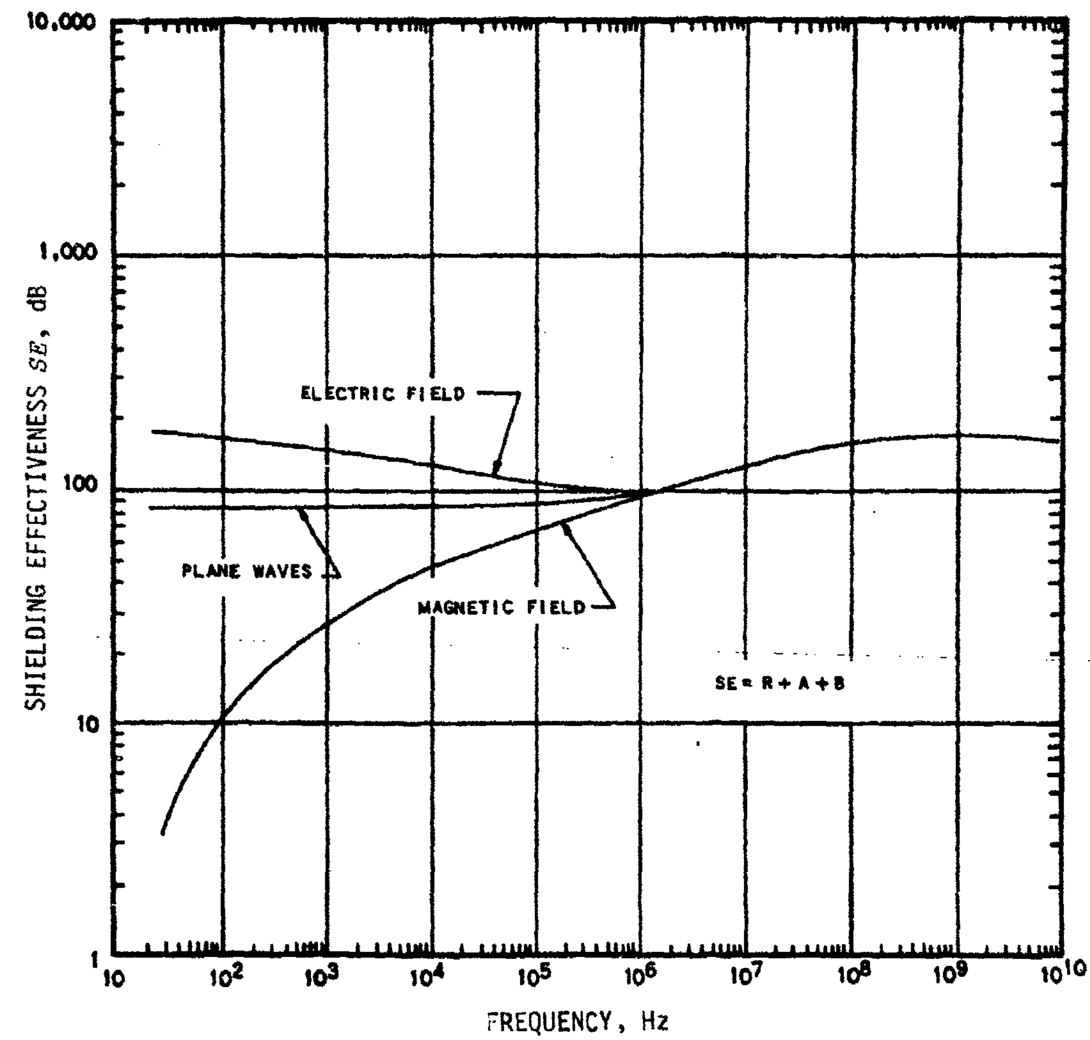


Figure 4-84. Shielding Effectiveness in Electric, Magnetic, and Plane Wave Fields of Steel Shield (1 mil Thickness) for Signal Sources 165 ft from the Shield

shields, isolation of circuits by decoupling, short leads, and the ground plane as the ground return lead.

4-6.4.2 Mutliple Shielding Applications

An effective method of handling a cabinet cover problem by multiple shielding is shown on Fig. 4-89(A). The cover shield is in the form of a sandwich, the center of which is insulating material. The inner conducting surface of the lid makes spring contact with the inner side of the shield, while the outer conductor of the lid makes spring contact with the other side of the shield. Such an arrangement is several hundred times as effective as a simple, all-metal lid with spring fit, and is satisfactory under many circumstances that otherwise would require wing-nut clamping arrangements.

The most important source of fields in a signal generator is the oscillator circuit coil. The effectiveness of the shielding system is increased greatly if the coil is enclosed in an auxiliary shield placed inside the main shield. In some cases, the entire tuned circuit — or the oscillator tube and associated RF tuned circuits and chokes — are placed in a separate shield

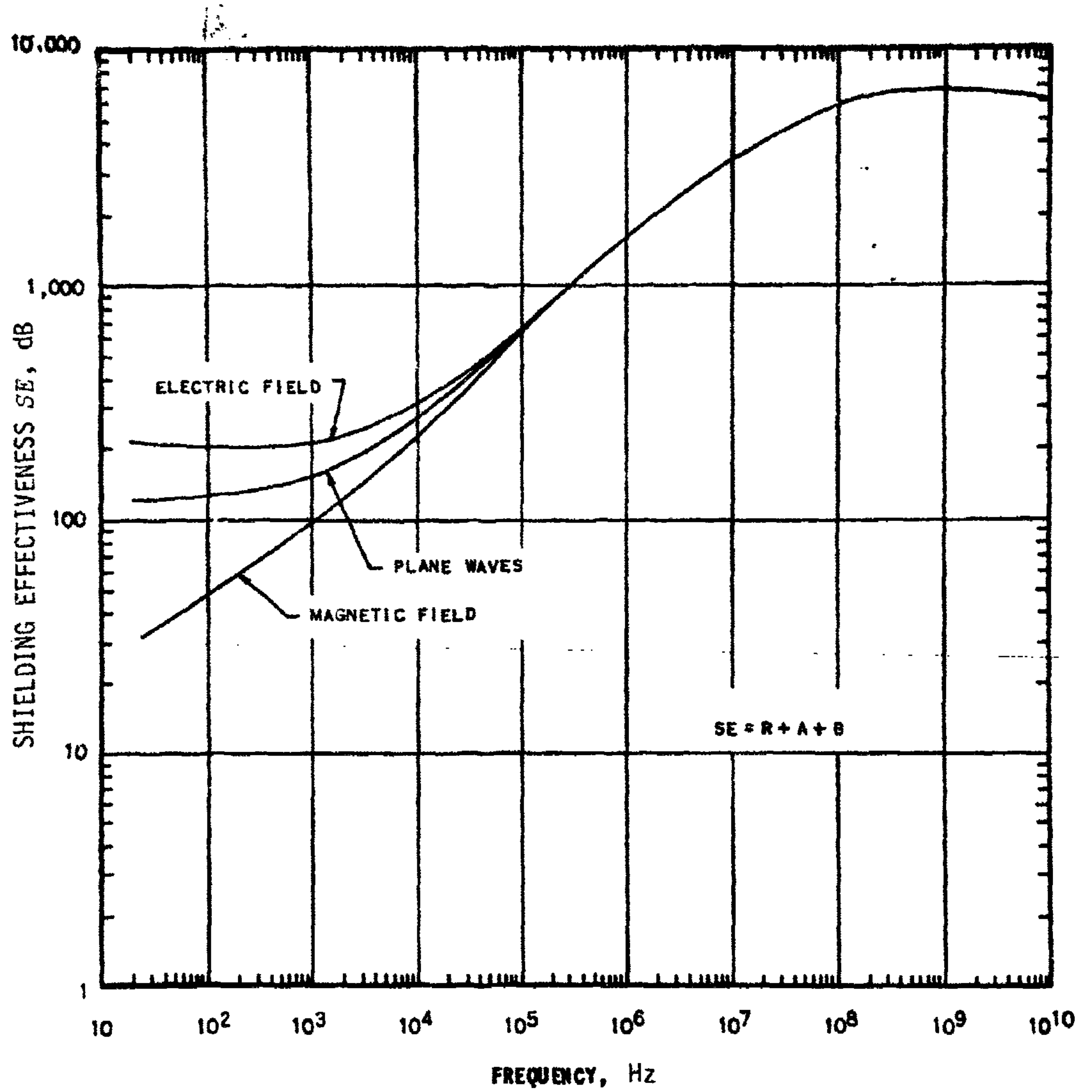


Figure 4-85. Shielding Effectiveness in Electric, Magnetic, and Plane Wave Fields of Steel Shield (50 mil Thickness) for Signal Sources 165 ft from the Shield

that is within the main shield. In general, refinements such as filters for leads or single-point grounding are not used in such a coil or tuned-circuit shield. If the shielding that results with such an arrangement is not adequate, the main shielding container is placed inside an outer shield. As shown on Fig. 4-89(B) the inner shield is insulated from the outer shield except for a single connection between the two. This arrangement precludes the possibility of currents circulating around a loop completed between the shields. In such an arrangement, leads passing through both inner and outer shields commonly are provided with additional filtering located in the space between the shields. Shafts that extend from the outside to the inner compartment should be of nonconducting material to avoid introduction of additional electrical connections between the shields.

4-6.5 IMPERFECTIONS IN SHIELDS

Imperfections can be classified as of three types: apertures, seams, and joints. Apertures are placed intentionally in otherwise closed shields to permit access via shafts or wiring or to provide ventilations and prevent moisture accumulation. Frequently,

TABLE 4-12
SHIELDING EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF COPPER SHIELD (7-MIL THICKNESS) FOR SIGNAL SOURCE 165 FT FROM THE SHIELD AT 30 Hz to 10GHz

Frequency	Plane Wave, dB	Electric Field, dB	Magnetic Field, dB
30 Hz	122	213	32
60 Hz	122	207	39
100 Hz	122	202	42
500 Hz	123	189	57
1 kHz	123	183	63
10 kHz	123	163	83
50 kHz	123	149	98
150 kHz	124	140	108
1 MHz	131	—	—
3 MHz	144	—	—
10 MHz	172	—	—
15 MHz	187	—	—
100 MHz	322	—	—
1000 MHz	818	—	—
1500 MHz	981	—	—
10 GHz	2408	—	—

TABLE 4-13
SHIELDING EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF STEEL SHIELD (1 MIL THICKNESS) FOR SIGNAL SOURCE 165 FT FROM THE SHIELD AT 30 Hz to 10 GHz

Frequency	Plane Wave, dB	Electric Field, dB	Magnetic Field, dB
30 Hz	85	175	4
60 Hz	86	171	6
100 Hz	86	166	10
500 Hz	86	152	21
1 kHz	86	146	26
10 kHz	86	125	46
50 kHz	87	113	61
150 kHz	89	105	73
1 MHz	98	—	—
3 MHz	110	—	—
10 MHz	136	—	—
15 MHz	142	—	—
100 MHz	164	—	—
1000 MHz	287	—	—
1500 MHz	186	—	—
10 GHz	164	—	—

screens are placed over apertures to reduce the loss in attenuation due to the aperture. Seams occur as a result of the process of manufacture of the shield, and joints occur because: (a) parts of separate containers are placed in contact to form the completed shield, or (b) because it is necessary to provide access doors.

TABLE 4-14
SHIELD EFFECTIVENESS IN ELECTRIC, MAGNETIC, AND PLANE WAVE FIELDS OF STEEL SHIELD (50 MIL THICKNESS) FOR SIGNAL SOURCE 165 FT FROM THE SHIELD AT 30 Hz to 10 GHz

Frequency	Plane Wave, dB	Electric Field, dB	Magnetic Field, dB
30 Hz	121	211	31
60 Hz	123	208	39
100 Hz	125	205	46
500 Hz	138	204	73
1 kHz	151	211	91
10 kHz	249	289	210
50 kHz	455	481	430
150 kHz	725	741	709
1 MHz	1465	—	—
3 MHz	2311	—	—
10 MHz	3801	—	—
15 MHz	4140	—	—
100 MHz	5338	—	—
1000 MHz	11850	—	—
1500 MHz	6547	—	—
10 GHz	5338	—	—

These imperfections can reduce the shielding effectiveness expected from the formulas and tables given in par. 4-6.3, which must be looked upon as upper limits in any given configuration. Thus, the imperfections may dominate the overall shielding effectiveness unless carefully controlled.

4-6.5.1 Apertures

Shielding Effectiveness *SE* formulas are given first, followed by discussion of their derivation and application.

4-6.5.1.1 Formulas for Shielding Effectiveness

$$SE = A_a + R_a + B_a + K_1 + K_2 + K_3 \quad , \text{dB} \quad (4\text{-}27)$$

where

A_a = aperture attenuation

$$= 27.3 \frac{D}{W} \text{ for rectangular aperture, dB} \quad (4\text{-}28)$$

$$= 32 \frac{D}{d} \text{ for circular aperture, dB} \quad (4\text{-}29)$$

D = depth of aperture, in.
W = width of rectangular aperture, in.
d = diameter of circular aperture, in.
R_a = aperture reflection losses

$$= 20 \log\left(\frac{|1+k|^2}{4|k|}\right), \quad \text{dB} \quad (4\text{-}3$$

TABLE 4-15. TOTAL REFLECTION LOSS IN ELECTRIC FIELD (WAVE IMPEDANCE MUCH GREATER THAN 377Ω) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

Frequency	dB Loss[a]					
	Steel Distance From Source To Shield			Copper		
	24 in.	165 ft	1 mi	24 in.	165 ft	1 mi
30 Hz	241	203	173	282	243	213
60 Hz	233	195	165	273	235	205
100 Hz	226	188	158	266	228	198
500 Hz	205	167	137	245	207	177
1 kHz	196	158	128	236	198	168
10 kHz	166	128	98	206	168	138
50 kHz	145	107	77[b]	185	147	117[b]
150 kHz	131	92	—	171	132	—
1 MHz	108	69[b]	—	146	108[b]	—
3 MHz	91	—	—	132	—	—
10 MHz	79	—	—	116	—	—
15 MHz	75	—	—	111	—	—
100 MHz	56[b]	—	—	86[b]	—	—

[a] If absorption loss is less than 10 dB, the total reflection loss must be corrected by use of *B*-factor.
[b] At these frequencies, the wave impednace approaches that of plane waves ($r > \lambda$), and the values for plane wave reflection loss should be used.

TABLE 4-16. TOTAL REFLECTION LOSS IN MAGNETIC FIELD (WAVE IMPEDANCE MUCH SMALLER THAN 377Ω) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

Frequency	dB Loss[a]					
	Steel Distance From Source To Shield			Copper		
	24 in.	165 ft	1 mi	24 in.	165 ft	1 mi
30 Hz	−1	24	53	25	63	93
60 Hz	−1.4	26	56	28	66	96
100 Hz	−1.2	30	59	30	69	99
500 Hz	1.4	36	66	37	76	106
1 kHz	3.2	39	69	40	79	109
10 kHz	11	49	79	50	89	119
50 kHz	18	56	86[b]	57	96	129[b]
150 kHz	23	60	—	62	100	—
1 MHz	32	70[b]	—	70	109[b]	—
3 MHz	37	—	—	75	—	—
10 MHz	43	—	—	80	—	—
15 MHz	46	—	—	82	—	—
100 MHz	60[b]	—	—	90[b]	—	—

[a] If penetration loss is less than 10 dB, the total reflection loss must be corrected by use of *B*-factor.
[b] At these frequencies, the wave impedance approaches that of plane waves ($r > > \lambda$), and the values for plane wave reflection loss should be used.

TABLE 4-17
TOTAL REFLECTION LOSS IN PLANE WAVE FIELD (WAVE IMPEDANCE EQUALS 377Ω) AT BOTH SURFACES OF SOLID STEEL AND COPPER SHIELDS

	dB Loss[a]	
Frequency	Steel	Copper
30 Hz	113	153
60 Hz	110	150
100 Hz	108	148
500 Hz	101	141
1 kHz	98	138
10 kHz	88	128
50 kHz	81	121
150 kHz	76	116
1 MHz	70	108
3 MHz	66	103
10 MHz	61	98
15 MHz	60	96
100 MHz	58	88
1000 MHz	51	78
1500 MHz	56	76
10 GHz	58	68

[a] If penetration loss is less than 10 dB, the total reflection loss must be corrected by use of B-factor.

$k = \dfrac{Z_a}{Z_w}$ = ratio of aperture characteristic impedance Z_a in Ω/cm^2 to impedance of the incident wave Z_w in Ω/cm^2

= $W/(3.142r)$ for rectangular apertures and magnetic fields (4-31)

= $d/(3.682r)$ for circular apertures and magnetic fields (4-32)

= $jfW\, 1.7 \times 10^{-4}$ for rectangular apertures and radiated fields (4-33)

= $jfd\, 1.47 \times 10^{-4}$ for circular apertures and radiated fields (4-34)

f = frequency, MHz

r = distance from signal source to shield, in.

B_a = correction factor for aperture reflections, dB (B_a becomes insignificant when A_a is more than 10 dB.)

$$= 20 \log \left| 1 - \frac{(k-1)^2}{(k+1)^2} 10^{-A_a/10} \right| \text{, dB} \quad (4\text{-}35)$$

K_1 = correction factor for the number of openings per unit square when the test antennas are far from the shield in comparison to the distance between holes in the shield.

$$= 10 \log \left(\frac{1}{AN} \right) \text{, dB} \quad (4\text{-}36)$$

A = area of each hole, in.^{2}

N = number of holes per in.^{2}

K_2 = correction factor for penetration of the conductor at low frequencies

$$= -20 \log \left(1 + \frac{35}{p^{2.3}} \right) \text{, dB} \quad (4\text{-}37)$$

p = ratio of the wire diameter to skin depth for screening

= ratio of the conductor width to skin depth between holes for perforated sheets, dimensionless. (Copper skin depth = $\dfrac{2.6 \times 10^{-3}}{\sqrt{f}}$ in.)

K_3 = correction factor for coupling between closely spaced shallow holes

$$= 20 \log \left(\frac{1}{\tanh \left| \dfrac{A_a}{8.686} \right|} \right) \text{, dB} \quad (4\text{-}38)$$

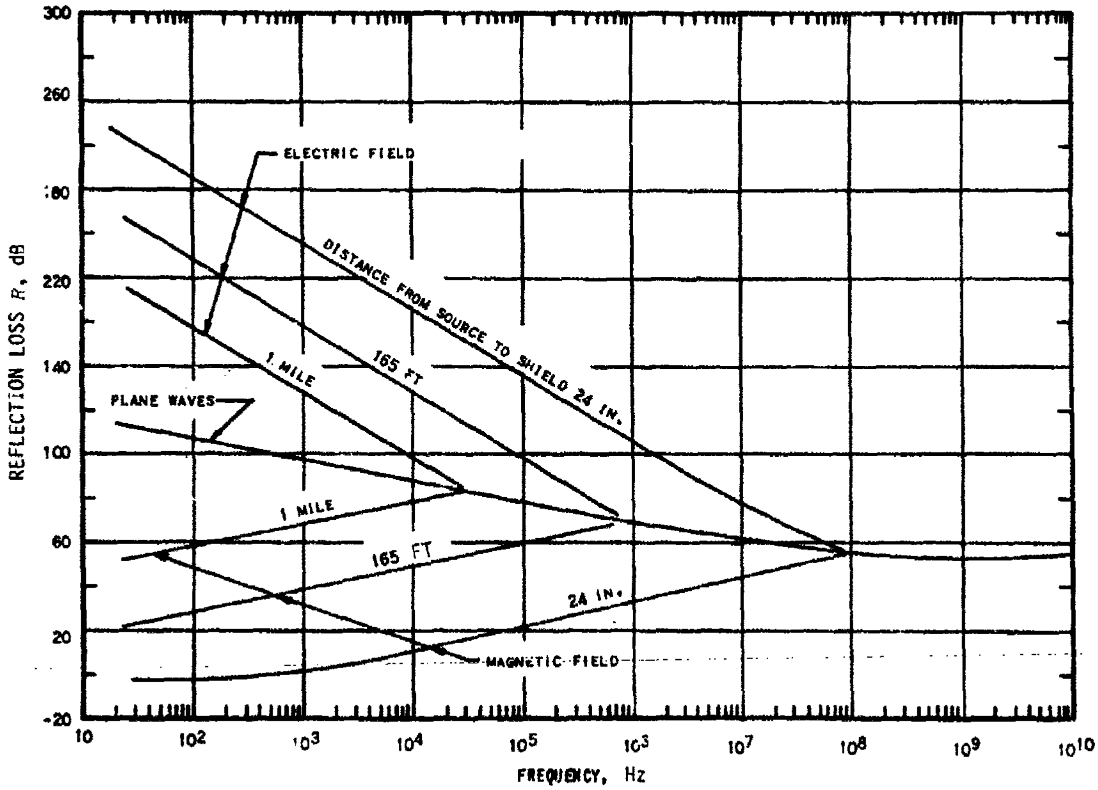


Figure 4-86. Total Reflection Loss at Both Surfaces of a Solid Steel Shield

4-6.5.1.2 Shielding Effectiveness Formula Derivations

Shielding effectiveness formulas follow:

a. *Attenuation calculations.* In Eq. 4-27, A_a represents the attenuation as the wave passes through the aperture which, for frequencies below cutoff in rectangular guides, is given as:

$$A_a = 8.68\pi \frac{D}{W} \sqrt{1 - \left(\frac{f}{f_o}\right)^2}\ , \text{dB} \qquad (4\text{-}39)$$

where f and f_o are the frequency under consideration and the cutoff frequency, respectively. It will be observed that W is always that hole dimension perpendicular to the E-field. Eq. 4-39 evaluated for frequencies well below cutoff, provides Eq. 4-28; and a similar procedure for circular guides provides Eq. 4-29.

b. *Reflection calculations* R_a. in Eq. 4-30 the reflection losses are calculated as a function of the ratio of the waveguide characteristic impedance below cutoff to the incident wave impedance. The characteristic impedance Z_a of a rectangular waveguide well below cutoff is:

$$Z_a = \frac{j\omega\mu_0 W}{\pi}\ , \Omega \qquad (4\text{-}40)$$

The impedance Z_w of the wave, emitted by a small loop source at points close to the source, compared to a wavelength, is:

$$Z_w = j\omega\mu_0 r\ , \Omega \qquad (4\text{-}41)$$

Taking the ratio of Eq. 4-40 to Eq. 4-41, we have:

$$k = \frac{W}{\pi r}$$

which is the equation for k given by Eq. 4-31 for magnetic fields. Both Eqs. 4-40 and 4-41 are in MKS units, but, in taking the ratio, W and r may be

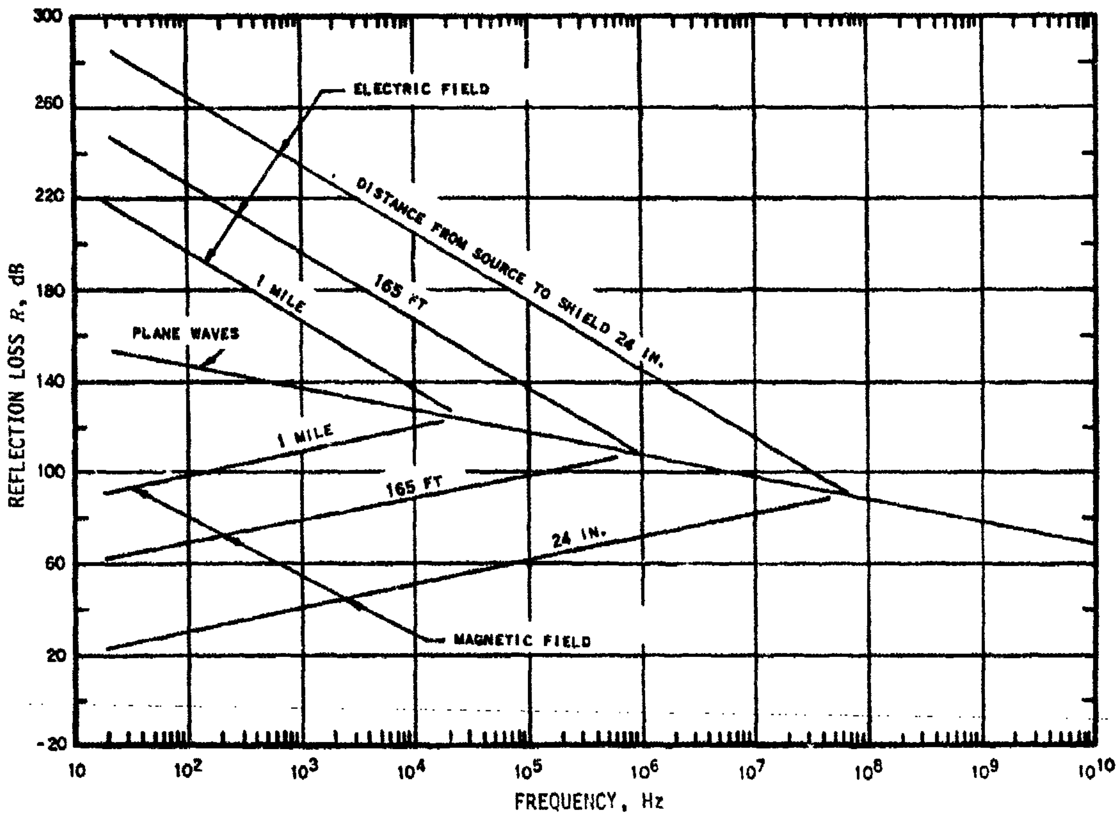


Figure 4-87. Total Reflection Loss at Both Surfaces of a Solid Copper Shield

expressed in either inches or meters. Similar procedures provide Eq. 4-32, 4-33, and 4-34.

c. *Corrections to reflection claculations* B_a. This factor is given for metal shields as:

$$B = 20\log\left|1 - \frac{(k-1)^2}{(k+1)^2}e^{-2\sigma t}\right|, \text{ dB} \quad (4\text{-}42)$$

where σ and t are, respectively, the complex propagation constant and the thickness of the shield in MKS units. In a waveguide below cutoff, the phase constant approaches zero and the propagation constant becomes equal to the attenuation constant, so that $\exp(-2\sigma t)$ becomes equal to the reduction in signal intensity in nepers for twice the depth of the waveguide. The expression $\exp(-2\sigma t)$, therefore, may be expressed in decibels and is equal to $10^{-A/10}$ which converts Eq. 4-42 to Eq. 4-35.

d. *Corrections for the number of openings that must be considered* K_1. It is obvious that when electromagnetic signals pass through a metal shield by penetration of openings, the amount of power transferred from one side of the shield to the other is a function of the number of openings. Not so obvious is the fact that if insertion loss tests are performed on the shield, the results will be a function of the distance between the antenna and the surface of the shield, assuming the shield to be centrally located between the antennas. If small antennas of approximately the same size as the openings, or smaller, are used for the test and are located on each side and adjacent to one of the openings, the measured shielding efficiency will be that of the opening itself. On the other hand, if the antennas are located at a considerable distance from the shield in comparison to the distance between holes in the shield, the measured shielding effectiveness will be equal to that of a single opening plus the ratio (in decibels) of the total wall area illuminated by the radiator to the total opening area located in the illuminated region. If the openings are evenly distributed, this ratio is a constant, since any change in the wall area illuminated will cause a similar change

TABLE 4-18. ABSORPTION LOSS OF STEEL AND COPPER SHIELDS AT 30 Hz to 10 GHz

Frequency	dB Loss		
	1 Mil Steel	50 Mil Steel	7 Mil Copper
30 Hz	0.2	9	0.13
60 Hz	0.3	13	0.18
100 Hz	0.3	17	0.23
500 Hz	0.7	37	0.52
1 kHz	1.0	53	0.74
10 kHz	3.2	161	2.34
50 kHz	7.5	374	5.23
150 kHz	13	649	9
1 MHz	28	1395	23
3 MHz	45	2245	40
10 MHz	75	3740	74
15 MHz	82	4080	90
100 MHz	106	5280	234
1000 MHz	236	11800	740
1500 MHz	130	6490	905
10 GHz	106	5280	2340

in hole area. Therefore, the minimum shielding effectiveness becomes that of the single opening plus the correction factor of Eq. 4-36. At intermediate points, the shielding effectiveness lies between the two values. Both values are of practical importance. The shielding effectiveness near the shield is important for protection of sensitive equipments which may be placed close to the walls (data cables, etc.). It is especially important in cases where radiation hazards may exist, so that personnel are protected who otherwise may unknowingly enter a strong field existing in the vicinity of a poorly designed seam. Calculations of shielding effectiveness in interior parts of a structure are important since, if advantage is taken of added shielding effectiveness, a considerable reduction in cost of shield construction can be achieved.

e. *Low frequency corrections K_2 for screen-type shields.* Numerous tests have shown that the high-frequency shielding effectiveness of screening materials can be satisfactorily approximated by assuming that the openings in the screen are equivalent in size to the openings in a flat perforated metal sheet, and that the depth of the openings is equal to the wire diameter. At low frequencies, when the skin depth becomes comparable to the radius of the wire, a considerable loss in shielding effectiveness occurs. This may be considered from the viewpoint that the apertures, as waveguides, are made wider and shorter by a skin depth. However, this approach runs into difficulty when the skin depth becomes equal to or greater than the wire radius, since this is the borderline region where leakage through the metal itself must also be considered. To maintain reasonable simplicity for calculation purposes, test results for a variety of copper screen shields were plotted as a function of skin depth, and an empirical equation (Eq. 4-37) was derived for the correction factor. This correction factor also may be satisfactory for perforated sheet metal, but no corroborative tests have been made.

f. *Corrections K_3 for closely spaced shallow openings.* When apertures in a shield are closely space, and the depth of the openings is small compared to the width, the shielding effectiveness has been found to be greater than otherwise would be expected. This is interpreted as being a result of coupling between adjacent holes, which becomes important when the attenuation through the openings is small. By considering two such adjacent holes subjected to an electromagnetic field, aligned as shown in Fig. 4-90, it appears that current induced on the conductor between the holes can flow into one side of a hole and return immediately via the adjacent hole — in effect, merely encircling the conductor. Since the current is the same in closely spaced holes, this is equivalent to placing a low resistance short circuit at the end of each hole considered as a waveguide. The impedance

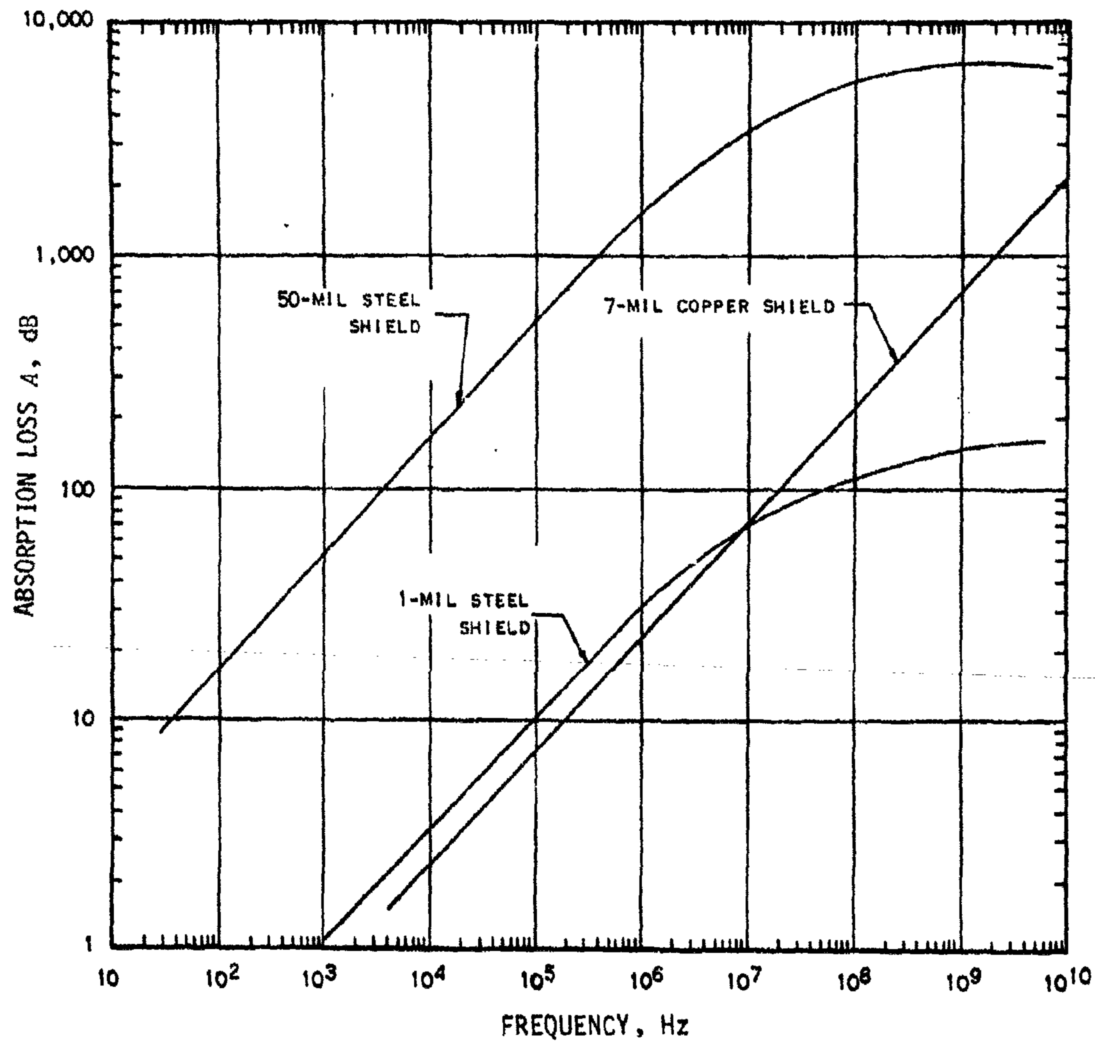


Figure 4-88. Absorption Loss for Steel and Copper Shields at 30 Hz to 10,000 MHz

of the short may be approximated for rectangular holes as the surface impedance presented by the surface of the conductor between the holes:

$$Z_L = \sqrt{\frac{j\omega\mu}{\sigma}}\left(\frac{a}{b}\right), \ \Omega \qquad (4\text{-}43)$$

where

$\sqrt{\frac{j\omega\mu}{\sigma}}$ = intrinsic impedance of the metal, Ω

$\omega = 2\pi f$, s^{-1}

μ = permeability of the metal = 1.26×10^{-6} H/m for copper

σ = conductivity of the metal = 5.8×10^7 mho/m for copper

a = width of conductor between holes (Fig. 4-90), m

b = length of rectangular hole (Fig. 4-90), m

The characteristic aperture impedance given by Eq. 4-40 is $j\omega\mu W/\pi$ at all frequencies being considered. The ratio of Z_L to Z_a is therefore:

$$\frac{Z_L}{Z_a} = \sqrt{\frac{j\omega\mu}{\sigma}}\left(\frac{a}{b}\right)\left(\frac{\pi}{j\omega\mu W}\right) = \frac{a\pi}{bW}\sqrt{\frac{1}{j\omega\mu\sigma}}, \quad \text{d'less} \qquad (4\text{-}44)$$

TABLE 4-19. CALCULATED *B*-FACTORS FOR STEEL AND COPPER SHIELDS FOR SIGNAL SOURCES 165 FT FROM THE SHIELD

Electric Field and Plane Wave				Magnetic Field			
Frequency	50 Mil Steel, dB	1 Mil Steel, dB	7 Mil Copper, dB	Frequency	50 Mil Steel, dB	1 Mil Steel, dB	7 Mil Copper, dB
30 Hz	−1.14	−27.0	−30.7	30 Hz	−1.07	−20.0	−30.7
60 Hz	$A > 10$ dB	−24.7	−27.9	60 Hz	$A >$ dB	−20.5	−27.9
100 Hz		−22.7	−25.8	100 Hz		−19.6	−25.8
500 Hz		−16.0	−19.0	500 Hz		−15.6	−19.0
1 kHz		−13.3	−16.1	1 kHz		−13.3	−16.1
10 kHz		− 5.6	− 7.5	10 kHz		− 5.6	− 7.5
50 kHz		− 1.7	− 3.1	50 kHz		− 1.7	− 3.1
150 kHz		$A > 10$ dB	− 1.2	150 kHz		$A > 10$ dB	− 1.2
			$A > 10$ dB				$A > 10$ dB

At frequencies as low as 10 kHz, the expression under the radical is smaller than 10^{-6}, showing that for all reasonable opening dimensions and any frequency of interest Z_L is much smaller than Z_a.

In accordance with transmission line theory, the input impedance of the guide Z_i may be calculated by:

$$Z_i = Z_a\left(\frac{Z_L \cosh\alpha + Z_a \sinh\alpha}{Z_a \cosh\alpha + Z_L \sinh\alpha}\right), \Omega \quad (4\text{-}45)$$

where α is the waveguide attenuation in nepers. Since, for the express condition being investigated, α is small, and since $Z_L \ll Z_a$, Eq. 4-45 simplifies to:

$$Z_i = Z_a\left(\frac{\sinh\alpha}{\cosh\alpha}\right) = Z_a \tanh\alpha \,, \Omega \quad (4\text{-}46)$$

The ratio of the intensities of the reflected wave to the transmitted wave, when the attenuation is large, is equal to:

$$\frac{(Z_w + Z_a)^2}{4 Z_w Z_a}, \text{ dimensionless} \quad (4\text{-}47)$$

When the attenuation is small, the ratio becomes:

$$\frac{(Z_w + Z_i)^2}{4 Z_w Z_i}, \text{ dimensionless} \quad (4\text{-}48)$$

For all practical purposes, the wave impedance Z_w is always much larger than either Z_a or Z_i, and the ratio of Eqs. 4-48 to 4-47 reduces to:

$$\frac{Z_a}{Z_i} = \frac{1}{\tanh\alpha}, \text{ dimensionless} \quad (4\text{-}49)$$

Eq. 4-49, in dB, provides the correction factor for closely spaced shallow holes as given by Eq. 4-38. It will be observed that, since Z_a is always larger than Z_i for A_a equal to 10 dB or less, the correction factor is always positive and increases the shielding effectiveness.

4-6.5.2 Aperture Screening

An equipment enclosure that requires inlet and/or outlet apertures for ventilation or pressurization should be designed with appropriate metallic coverings placed over the apertures. Although louvered openings generally are used for cooling air circulation, they are extremely poor for RF integrity because of their long, narrow gaps. In descending order of attenuation properties, the following materials should be used: waveguide below cut-off panels, perforated metal sheet, woven metal mesh, and knitted metal mesh. Solid metal covers for apertures designed for temporary access are discussed in par. 4-6.5.4.

TABLE 4-20
EXAMPLES FOR CALCULATING SHIELDING EFFECTIVENESS OF SOLID-METAL SHIELD

	10 kHz—10 Mils					
	Magnetic Field		Electric Field		Plane Wave	
	Copper	Iron	Copper	Iron	Copper	Iron
Reflection	44.2	8.0	212.0	174.0	128.0	90.5
Absorption	3.6	43.5	3.3	43.5	3.3	43.5
B-Factor	−2.6	0	−2.6	0	−2.6	0
Total Loss, dB	45.2	51.5	212.7	217.5	128.7	134.0

	60 Hz—Magnetic					
	1 Mil		10 Mils		300 Mils	
	Copper	Iron	Copper	Iron	Copper	Iron
Reflection	22.4	−0.9	22.4	−0.9	22.4	−0.9
Absorption	0.03	0.33	0.26	3.34	7.80	100.0
B-Factor	−22.2	+0.95	−19.2	+0.78	+0.32	0
Total Loss, dB	0.23	0.38	3.46	3.22	30.52	99.1

	10 kHz — 30 Mils — Magnetic		1 kHz — 10 Mils — Magnetic	
	Copper	Iron	Copper	Iron
Reflection	44.20	8.0	34.2	0.9
Absorption	10.02	130.5	1.06	13.70
B-Factor	+0.58	0	−10.37	+0.06
Total Loss, dB	54.80	138.5	24.89	14.66

	10 Mils—Copper					
	150 kHz			1 MHz		
	Electric	Plane Waves	Magnetic	Electric	Plane Waves	Magnetic
Reflection	176.8	117.0	56.0	152.0	108.2	64.2
Absorption	12.9	12.9	12.9	33.4	33.4	33.4
B-Factor	+0.5	+0.5	+0.5	0	0	0
Total Loss, dB	190.2	130.4	69.4	185.4	141.6	97.6

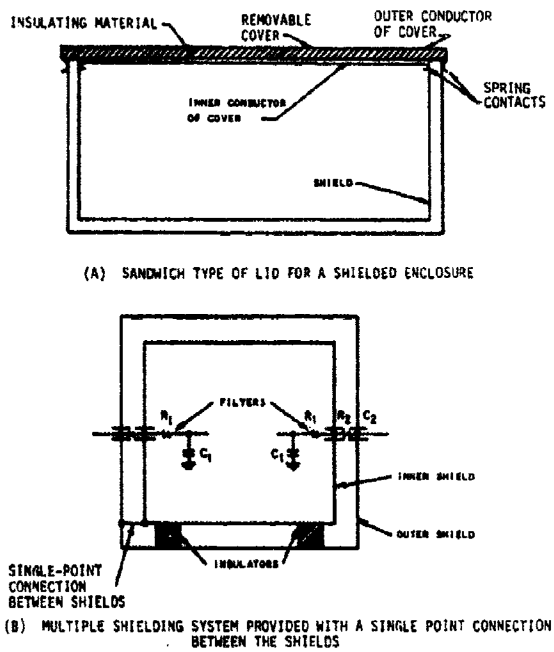


Figure 4-89. Multiple Shielding Applications

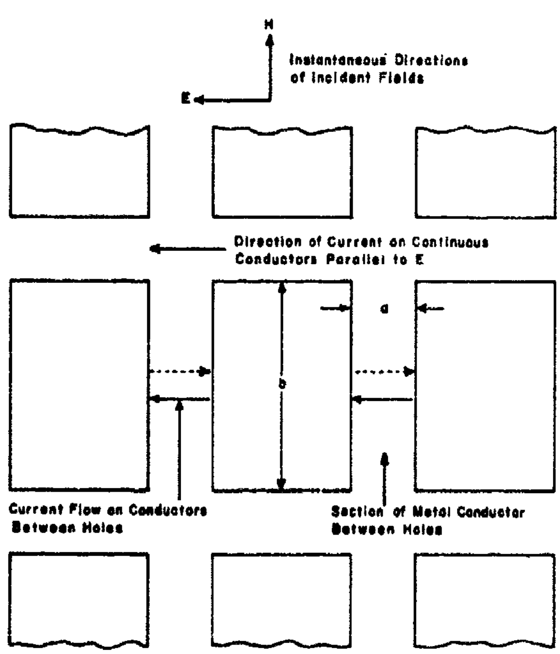


Figure 4-90. Currents Induced on Perforated Metal Sheet

4-6.5.2.1 Screening

A single or double layer of copper or brass screen of No. 16 or 22 gage wire, should have openings no greater than 1/16 by 1/16 in. A mesh less than 18 by 18 (wires to the inch) should not be used. The mesh wire diameter should be a minimum of 0.025 in. (No. 22 AWG). If more than a nominal 50 dB of attenuation is required, the screening should have holes no larger than those in a 22 by 22 mesh made of 15 mil copper wires.

The attenuation of an electromagnetic wave by a mesh is considerably less than that afforded by a solid metal screen. The principal shielding action of a mesh is due to reflection. Tests have shown that mesh with 50% open area and 60 or more strands per wavelength introduces a reflection loss very nearly equal to that of a solid sheet of the same material. The mesh construction should have individual strands permanently joined at points of intersection by a fusing process so that permanent electrical contact is made, and oxidation does not reduce shielding effectiveness. A screen of this construction will be very effective for shielding against electric (high-impedance) fields at low frequencies because the losses will be caused primarily by reflection. Installation can be made by connecting a screen around the periphery of an opening (Figs. 4-91 and 4-92). The results of shielding effectiveness measurements for screening are shown on Figs. 4-93 through 4-98. Table 4-21 lists mesh, wire, and aperture sizes for various screening materials.

The screening materials given in Table 4-22 and on Figs. 4-99 and 4-100 exhibit low-shielding effectiveness — probably because of crossover discontinuities in mesh. A comparison between No. 16 aluminum mesh and No. 12 copper mesh (Fig. 4-99) reveals a 5 dB differential, and both have an attenuation that is comparable to that of 60 mils perforated steel shown in Fig. 4-100. All are in the 50 dB range, with the perforated material having a greater percentage of open area and, therefore, less resistance to air flow. Because mesh, for effective shielding, rarely has more than an open area of 50%, the size of apertures must be increased correspondingly for effective ventilation. Mesh should be easily removable; it should be attached with screws or bolts. These should be in sufficient number to ensure high-pressure contact along a continuous line completely around the edge. Contact surfaces should be cleaned thoroughly each time the mesh is removed.

In the 40-dB range, the 30-mil, 1/4-in. spacing galvanized steel mesh and the 0.037-in. aluminum perforated sheet (Fig. 4-100) are comparable to No. 10,

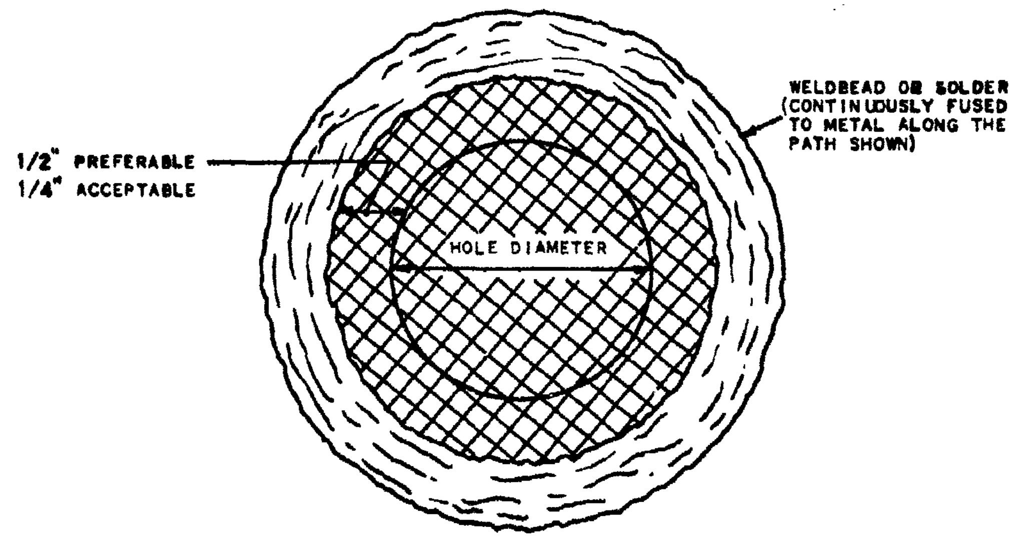


Figure 4-91. Typical Welded Screen Installation Over a Ventilation Aperture

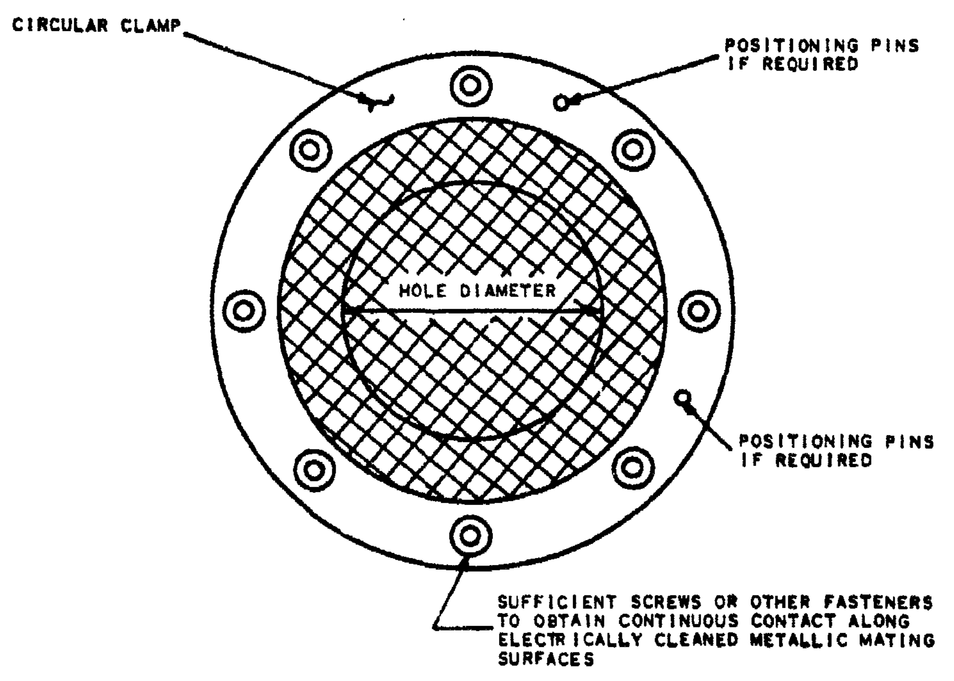


Figure 4-92. Typical Clamped Screen Installation Over a Ventilation Aperture

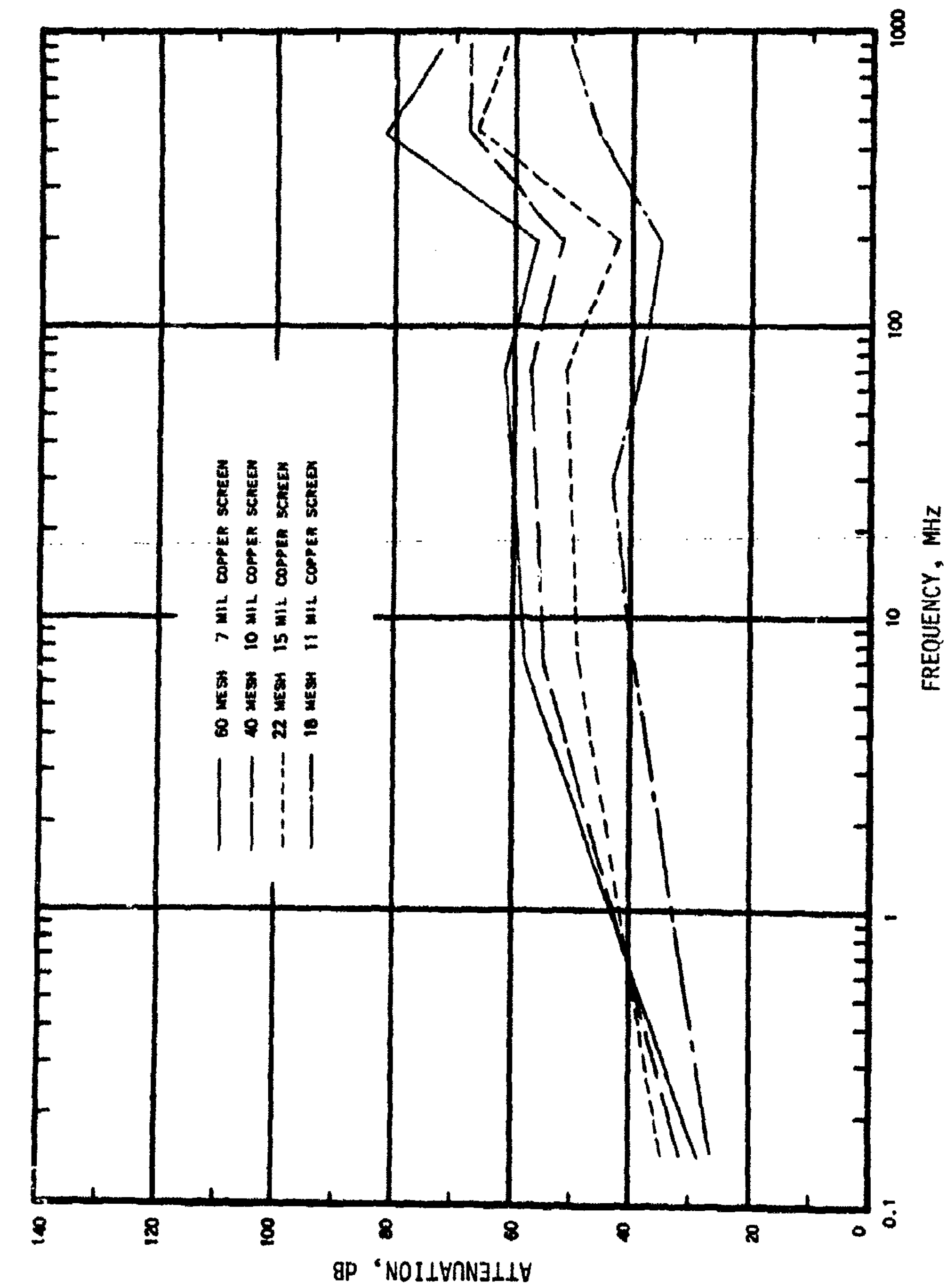


Figure 4-93. Shielding Effectiveness of Various Copper Screens

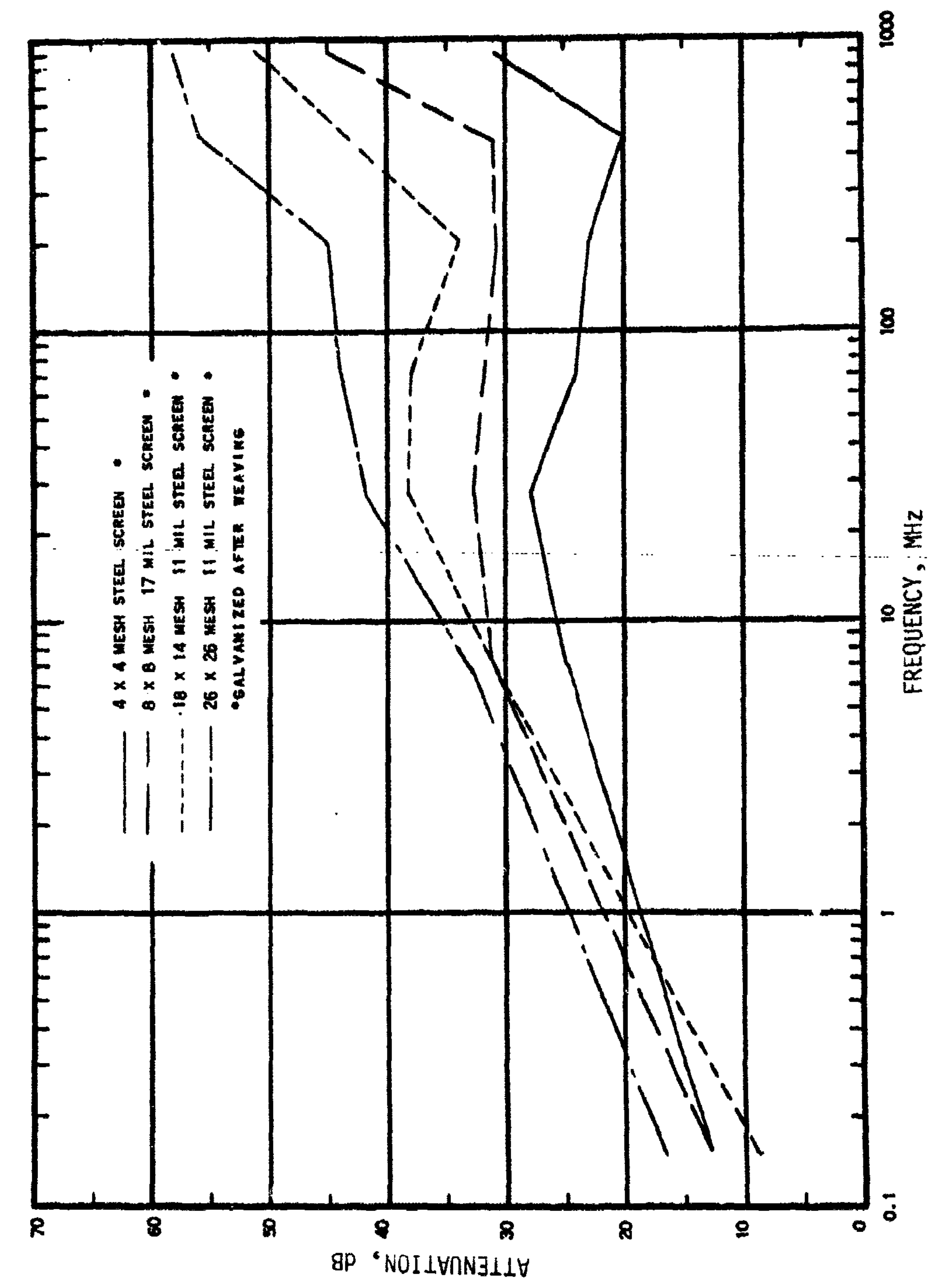


Figure 4-94. Shielding Effectiveness of Various Galvanized Steel Screens

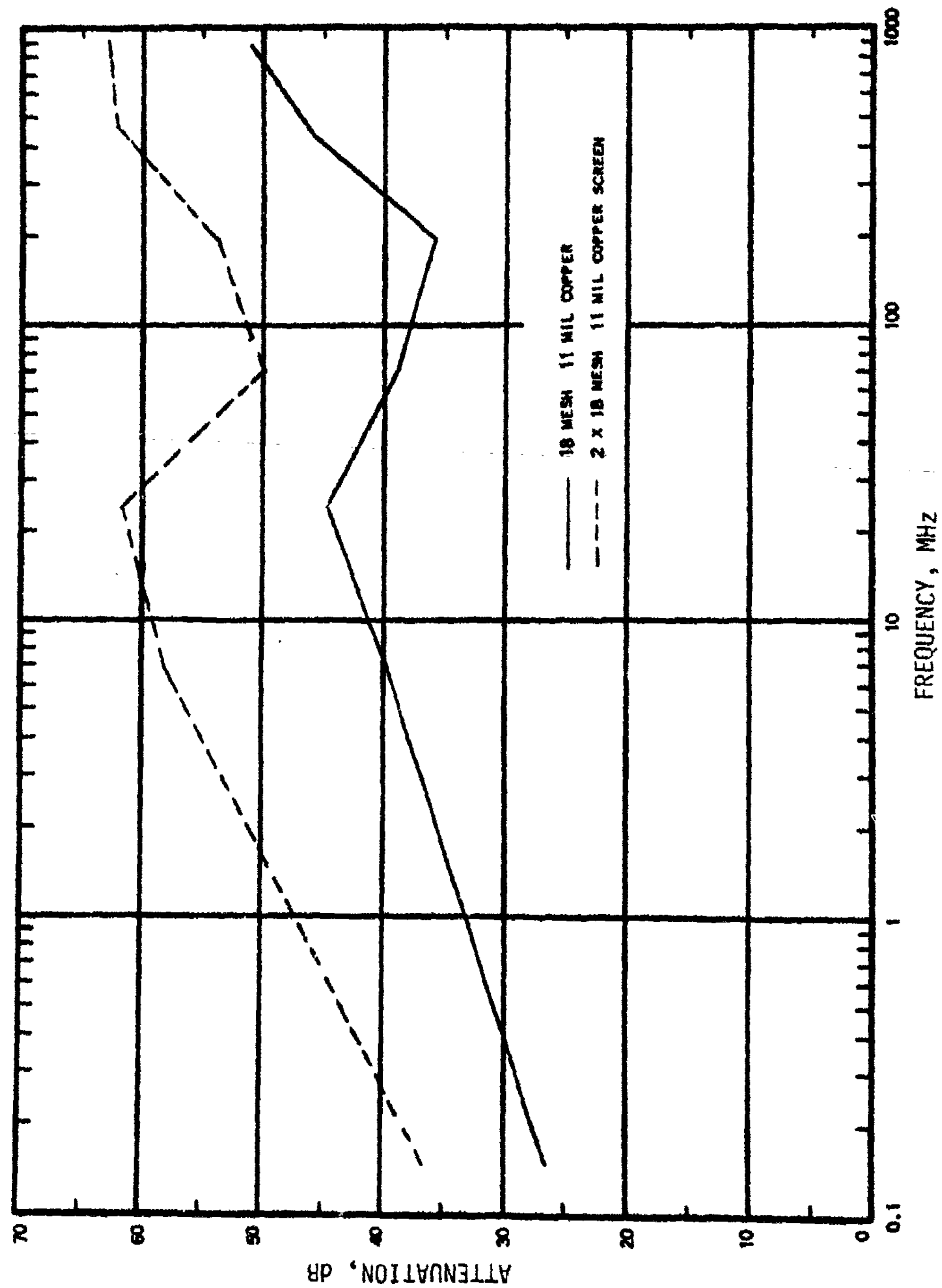


Figure 4-95. Comparison of the Shielding Effectiveness of Single and Double 18 Mesh Copper Screening

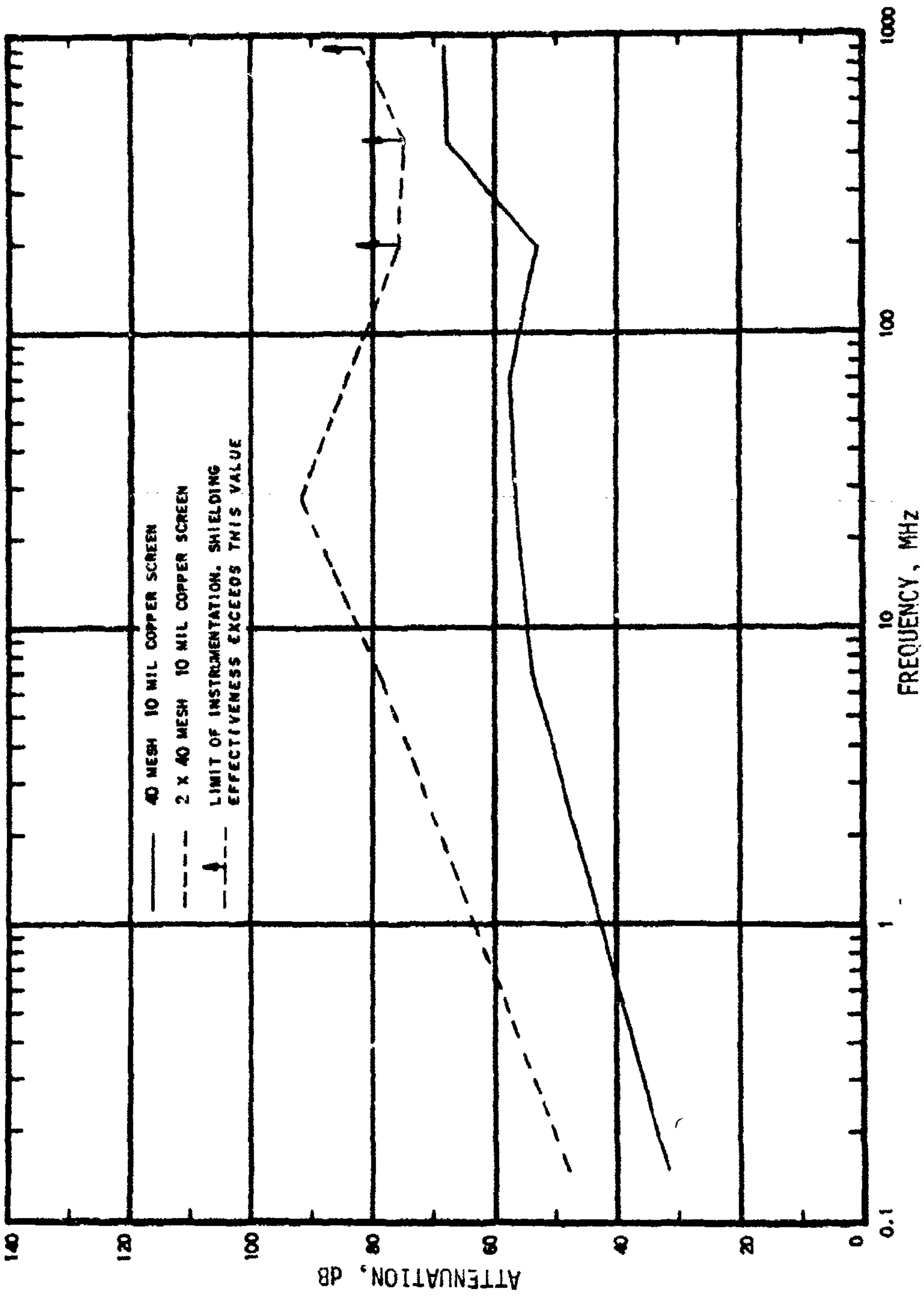


Figure 4-96. Comparison of the Shielding Effectiveness of Single and Double 40 Mesh Copper Screening

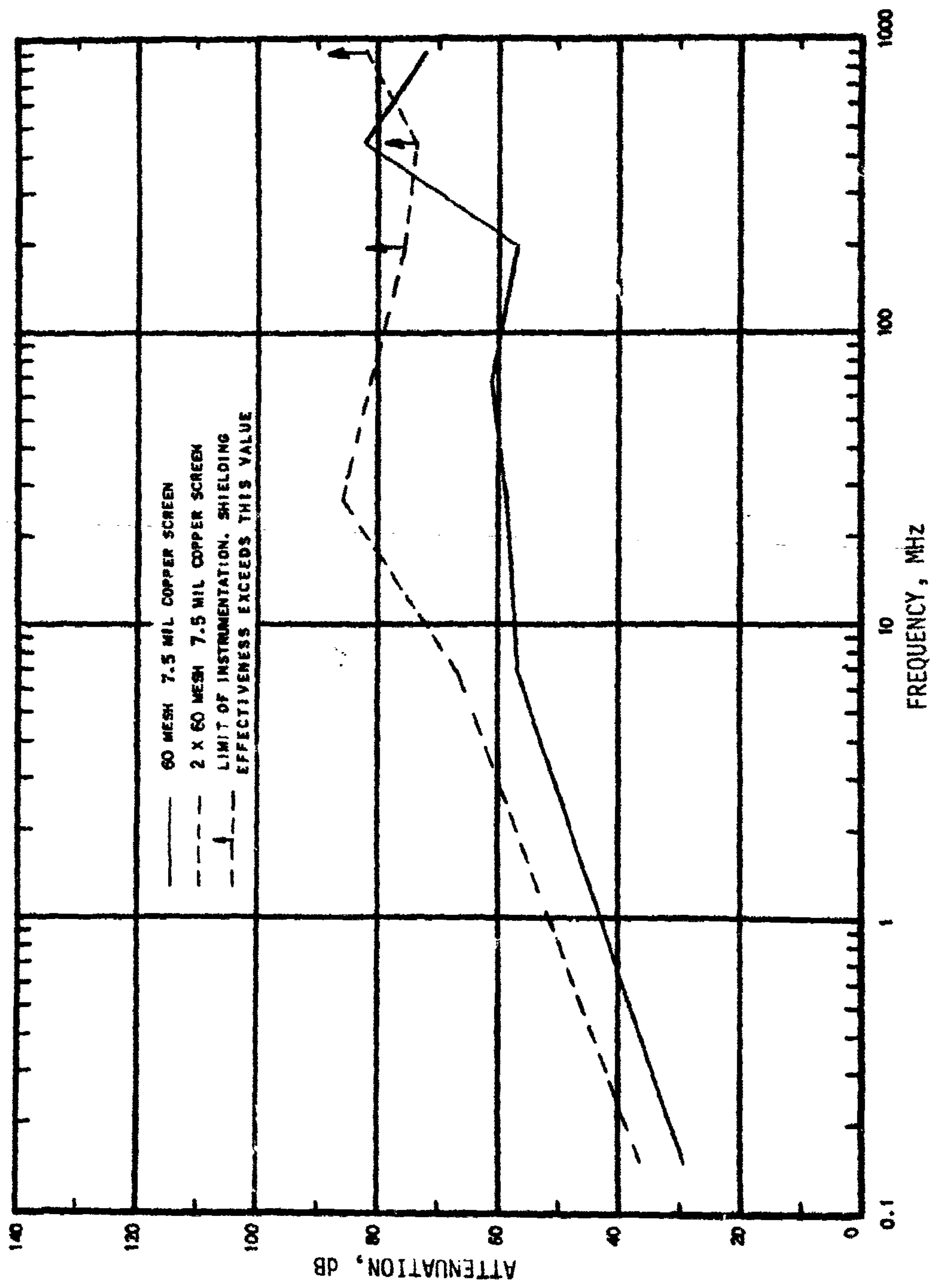


Figure 4-97. Comparison of the Shielding Effectiveness of Single and Double 69 Mesh Copper Screening

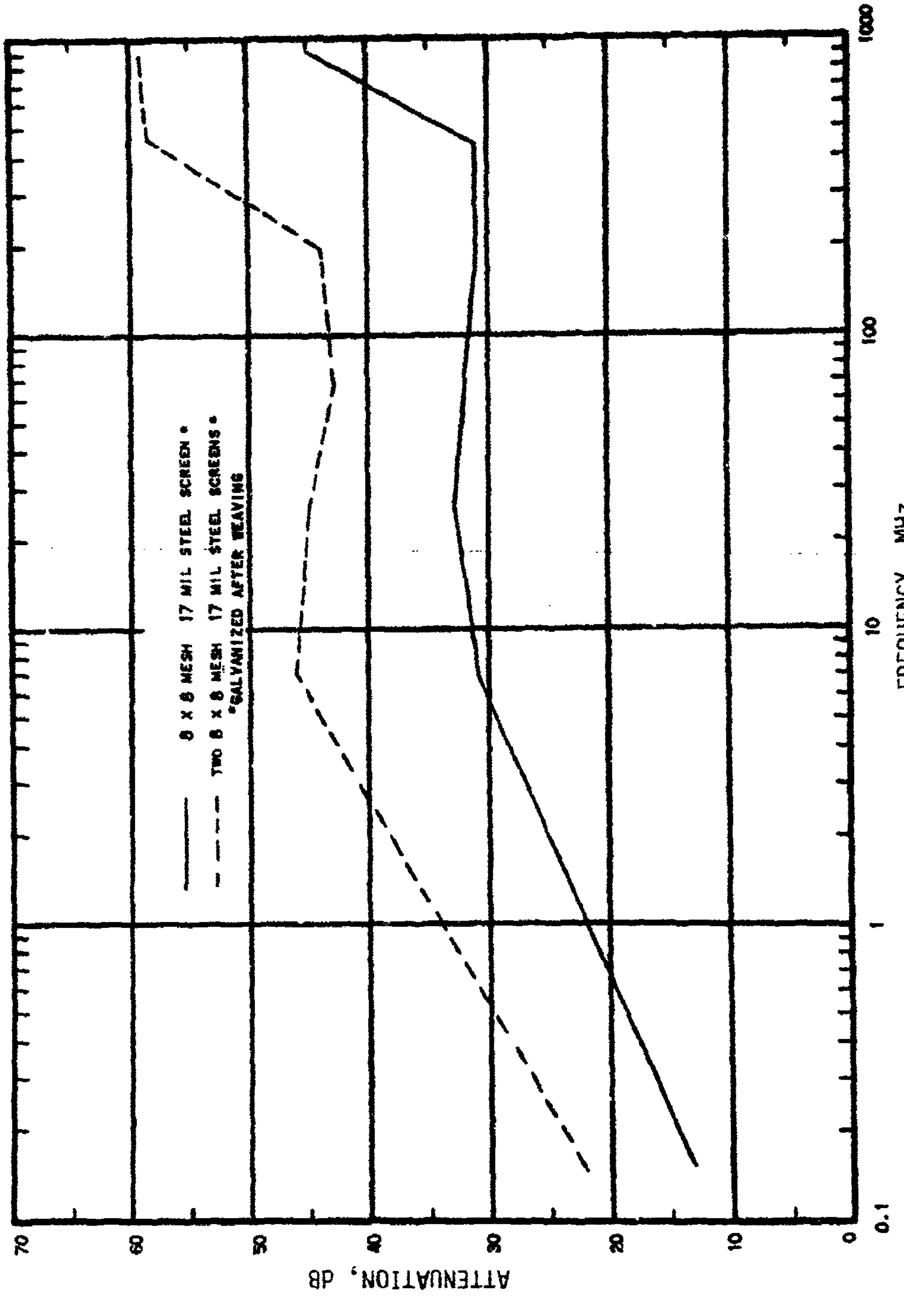


Figure 4-98. Comparison of the Shielding Effectiveness of Single and Double 8 Mesh Galvanized Steel Screening

TABLE 4-21
MESH, WIRE, AND APERTURE SIZES FOR TYPICAL SCREENING MATERIALS

Mesh	Wire Diameter, in.	Size of Opening, in.
8 × 8	0.028	0.097
8 × 8	0.032	0.093
8 × 8	0.035	0.090
8 × 8	0.047	0.078
8 × 8	0.063	0.062
10 × 10	0.025	0.075
10 × 10	0.032	0.068
10 × 10	0.035	0.065
10 × 10	0.041	0.059
12 × 12	0.018	0.065
12 × 12	0.023	0.060
12 × 12	0.028	0.055
12 × 12	0.035	0.048
12 × 12	0.041	0.042
14 × 14	0.017	0.054
14 × 14	0.020	0.051
14 × 14	0.025	0.046
14 × 14	0.032	0.039
16 × 16	0.016	0.0465
16 × 16	0.018	0.0445
16 × 16	0.028	0.0345
18 × 18	0.017	0.0386
18 × 18	0.020	0.0356
18 × 18	0.025	0.0306
20 × 20	0.014	0.0360
20 × 20	0.016	0.0340
20 × 20	0.020	0.0300
22 × 22	0.015	0.0305

18 mils monel mesh (Fig. 4-99). The galvanized mesh has the largest percentage of open area.

Fig. 4-99 shows test results of materials. The 40-count copper-nickel 36% open sample has a range greater than 90 dB. Such electroformed mesh materials are excellent as RF screening agents for panel meters, pilot lamps, counters, and other similar aperture discontinuities, or as simple inner-unit shields where small volumes of air transmission are necessary.

Figs. 4-101 and 4-102 illustrate the measured magnetic field shielding effectiveness of screens of the various designs and metallic compositions listed in Table 4-23 for the range of 0.15 to 1000 MHz.

4-6.5.2.2 Waveguide-below-cutoff Devices

In many cases, shielding screens introduce excessive air resistance, and sometimes greater shielding effectiveness may be needed than they can provide. In such cases, openings may be covered with specially designed ventilation panels (honeycombs) with openings that operate on the waveguide-below-cutoff principle. When an insulated control shaft passes through a waveguide attenuator, the control function can be accomplished with almost no interference leakage. A sample panel is shown on Fig. 4-103. Honeycomb-type ventilation panels in place of screening:

a. Allow higher attenuation than can be obtained with mesh over a specified frequency range.

b. Allow more air flow without pressure drop for the same diameter opening.

c. Are less easily damaged, and are therefore more reliable.

TABLE 4-22
MESH, WIRE, AND APERTURE SIZES FOR SCREENING MATERIALS TESTED

Metal	Mesh Size	Wire Diameter, mils
Monel	No. 10	18
Copper	No. 12	20
Aluminum	No. 16	20
Galvanized steel	1/4 in. × 1/4 in.	30
Galvanized steel	1/2 in. × 1/2 in.	30
Perforated steel[a]	1/8 in. diameter holes	on 3/16 in. centers
Perforated aluminum[a]	1/4 in. diameter holes	on 5/16 in. centers
Perforated aluminum[b]	7/16 in. diameter holes	on 5/16 in. centers
Aluminum honeycomb[c]	1/4 in. segregated cells	

[a] 60 mils thick
[b] 37 mils thick
[c] 1 in. thick

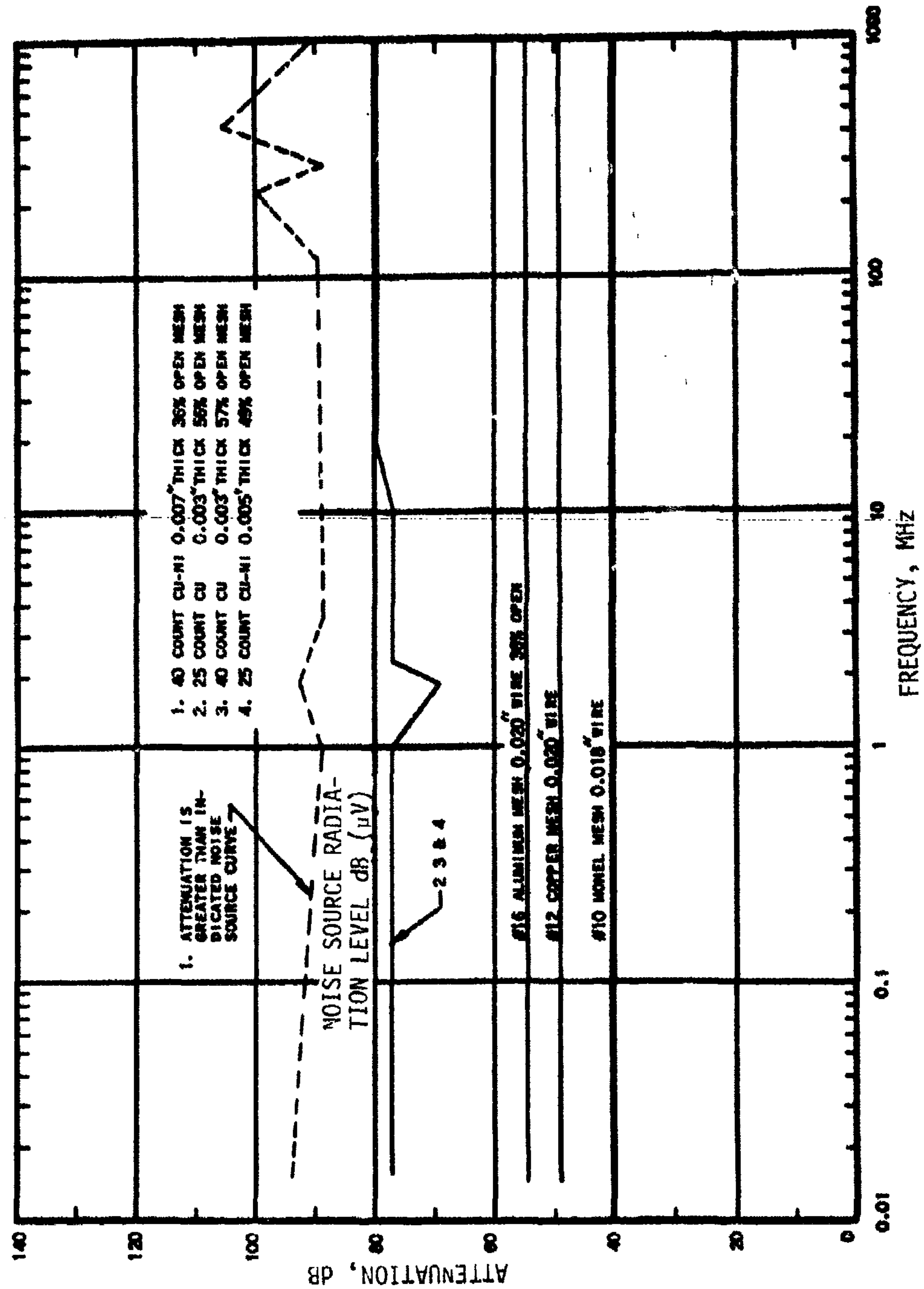


Figure 4-99. Attenuation vs Frequency Curves for Various Screens

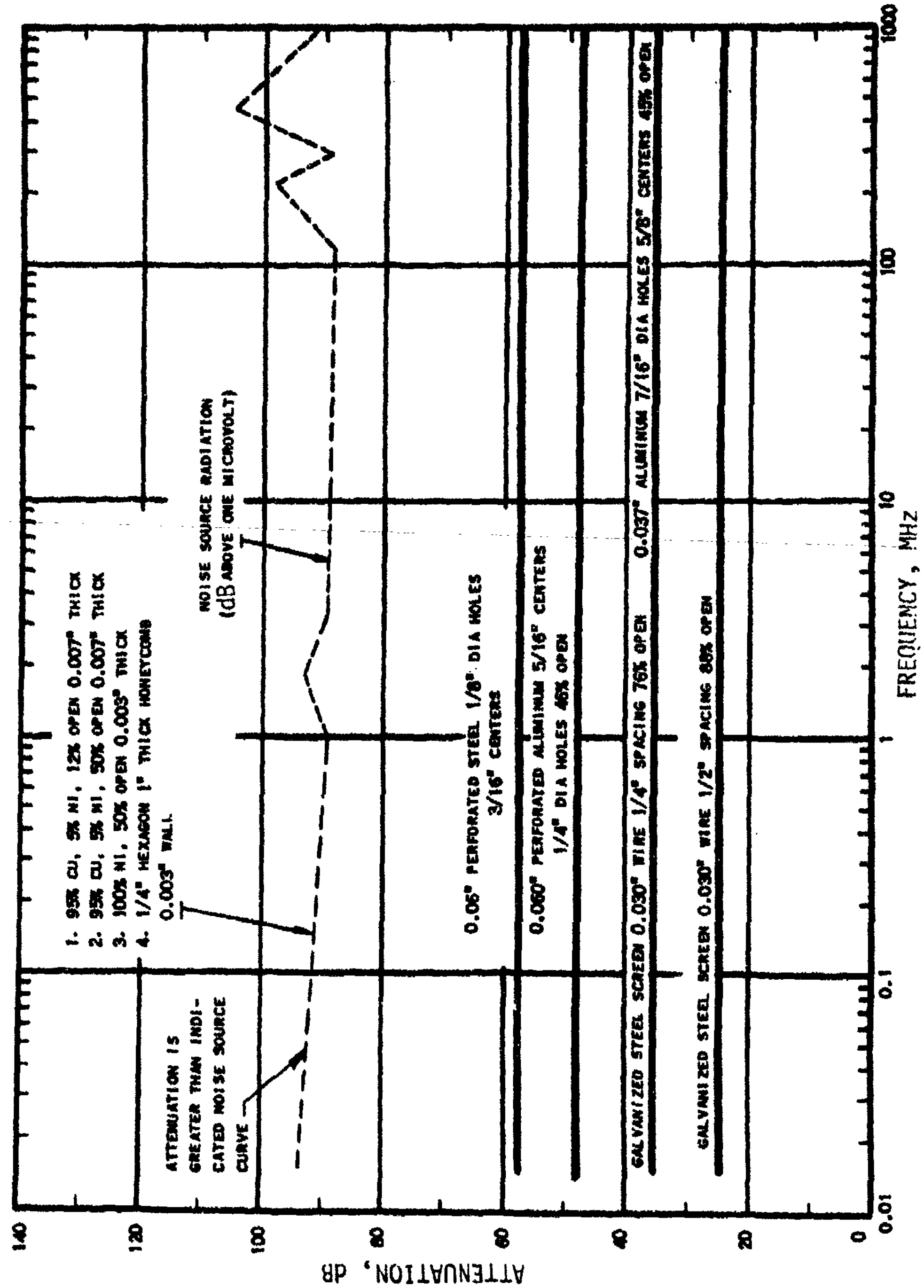


Figure 4-100. Attenuation vs Frequency Curves for Various Screens and Honeycomb

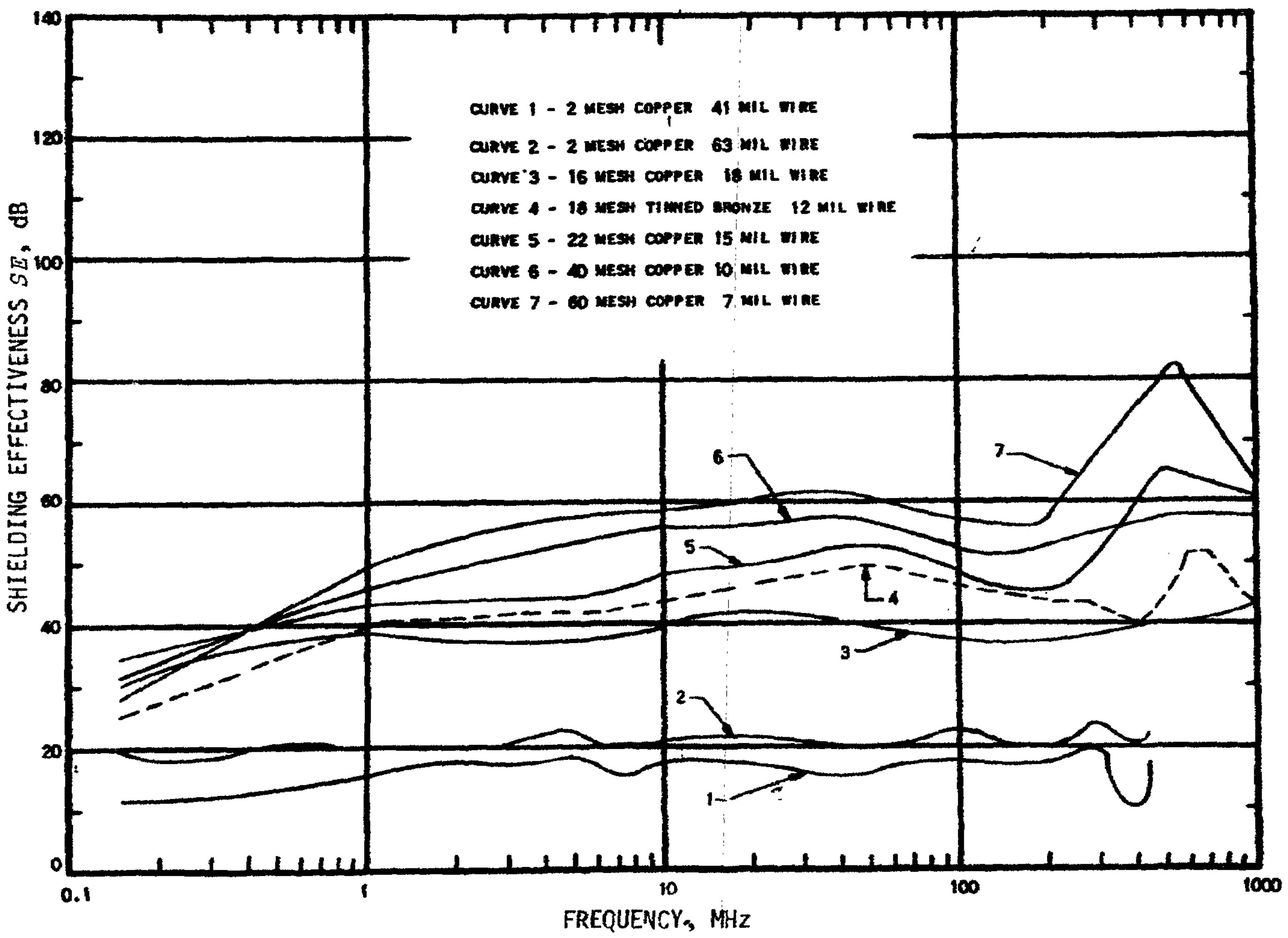


Figure 4-101. Minimum Shielding Effectiveness of Copper and Bronze Screening in a Magnetic Field

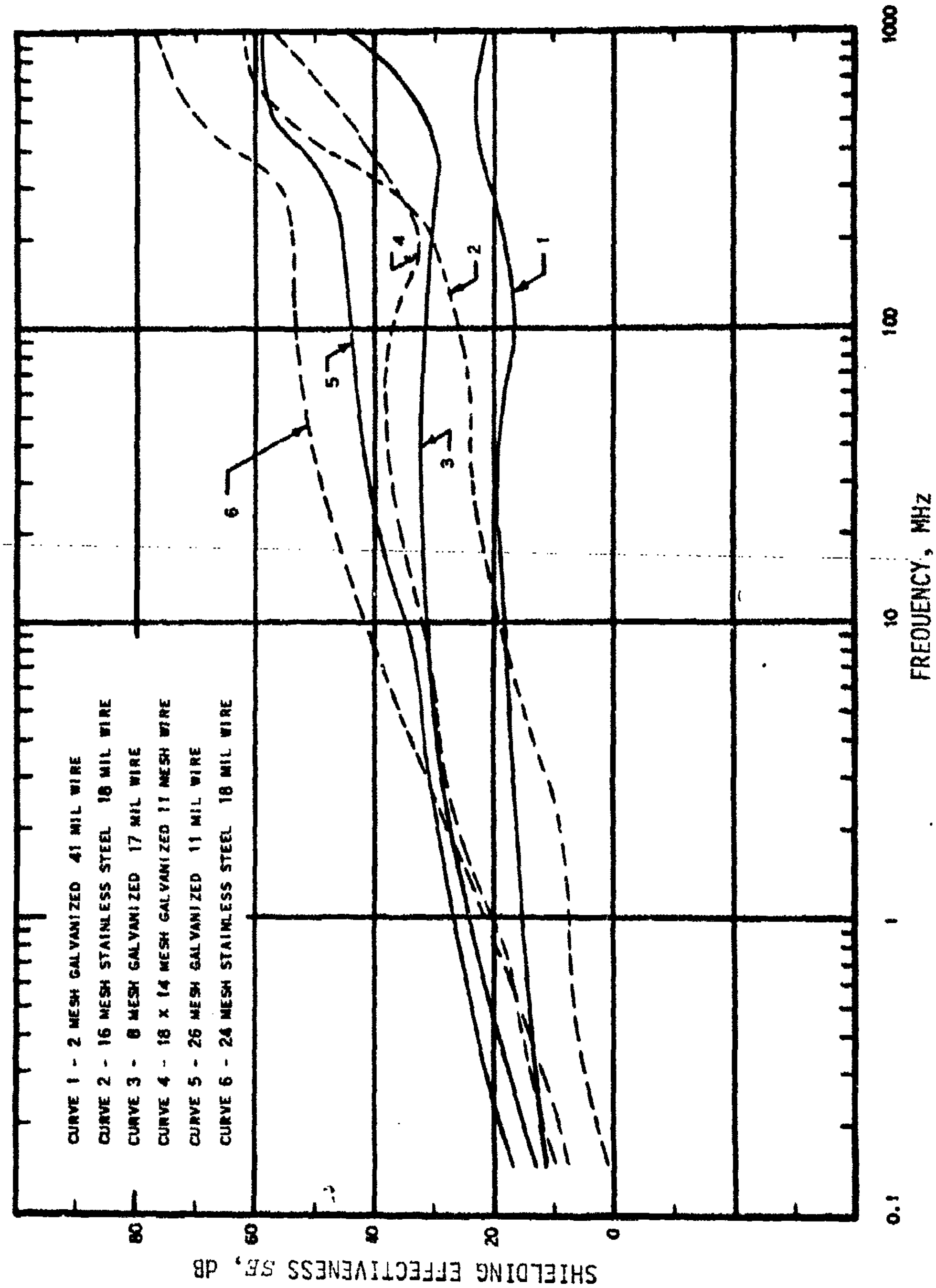


Figure 4-102. Minimum Shielding Effectiveness of Steel Screening in a Magnetic Field

TABLE 4-23
COPPER, BRONZE, AND STEEL SCREENING USED IN MAGNETIC FIELD SHIELDING EFFECTIVENESS TEST MEASUREMENTS

Metal	Mesh, wires/in.	Wire Diameter, mils
Copper	2	41
	2	63
	16	18
	22	15
	40	10
	60	7
Tinned Bronze	18	12
Galvanized Steel	2	4
	8	17
	18 × 14	11
	26	11
Stainless Steel	16	18
	24	18

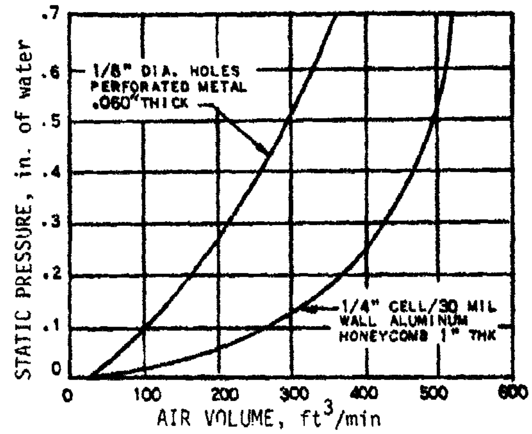


Figure 4-104. Air Impedance of Perforated Metal and Honeycomb

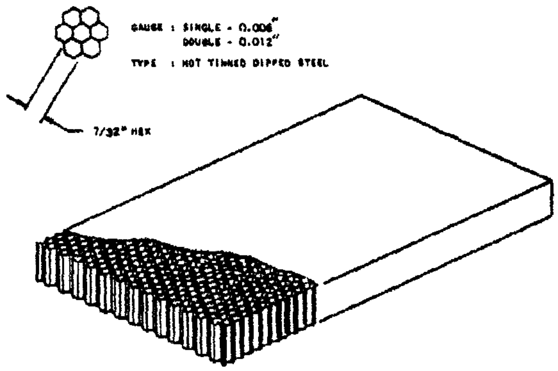


Figure 4-103. Honeycomb Type Ventilation Panel

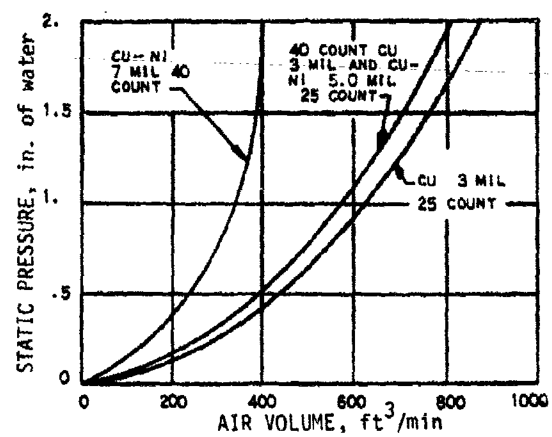


Figure 4-105. Air Impedance of Copper and Nickel Mesh

d. Are less subject to deterioration by oxidation and exposure.

A comparison of air impedance characteristics of honeycomb and screen materials is shown on Figs. 4-104 and 4-105. Panels of honeycomb vary in thickness from 3/4 in. to 2-3/8 in., depending upon the attenuation desired. Honeycomb panels can achieve attenuations to 135 dB, above 10 MHz.

When frequencies above 1000 MHz are to be highly attenuated, ventilation openings must be designed as waveguide attenuators. To obtain an opening of sufficient size to admit the required volume of ventilating air, tubes should be placed side by side until sufficient air flow is achieved.

The attenuation of the waveguide operating below cutoff frequency, given by Eq. 4-39, is plotted on Fig. 4-106 for a rectangular waveguide, and Fig. 4-107 for a circular waveguide, both for a D/W ratio of 1. For ratios other than 1, the value in decibels, obtained from the curve, must be multiplied by D/W to arrive at the correct value of attenuation. For example, a shielding effectiveness of over 100 dB can be obtained at 10,000 MHz with a 0.25 in. diameter tube, 1 in. in length, or a 1/2 in. diameter tube, 2.25 in. long.

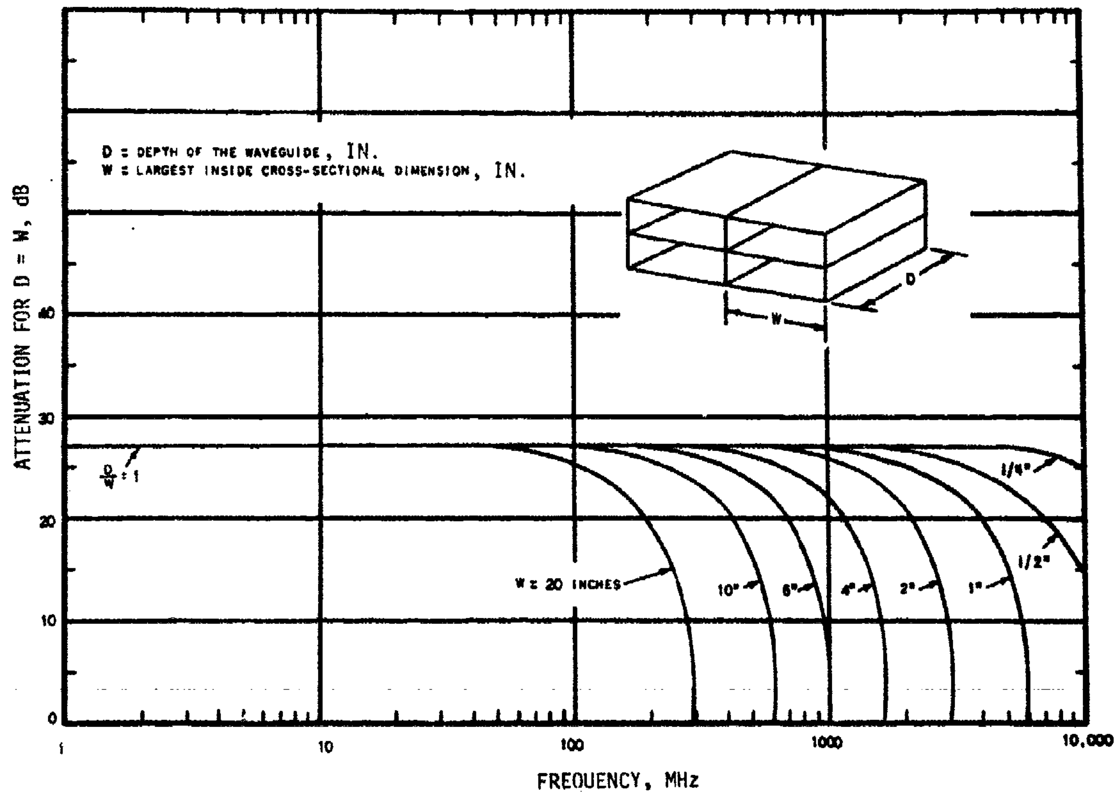


Figure 4-106. Attenuation — Rectangular Waveguide

Openings 1 in. or more in diameter would have little or no attenuation at 10,000 MHz. For a rectangular waveguide attenuator, the cutoff frequency f_o is given by

$$f_o = \frac{5910}{W}, \text{ MHz} \tag{4-50}$$

where

W = largest inside cross-sectional dimension, in.

For a circular guide

$$f_o = \frac{9044}{W}, \text{ MHz} \tag{4-51}$$

where:

W = diameter, in.

Figs. 4-106 and 4-107 indicate the frequency range over which any particular opening size may be useful. It is assumed that the conductivity and electrical thickness of the metal walls between openings are sufficient at frequencies as low as 1 MHz. The flat characteristic of the curves continues to the lowest frequency at which this condition is met.

Control shafts can be brought out of a shielded equipment enclosure through a waveguide below cutoff attenuator. The shaft must be made of nylon, Teflon, or other dielectric material. Fig. 4-108(A) shows a metal shaft which is shielded by a metal insert under the knob but which must be grounded to the panel of the equipment by means of a spring washer to complete the shield. Fig. 4-108(B) shows a dielectric shaft mounted within a metal tube. The tube acts as a waveguide below cut-off attenuator, and no further shielding is required.

4-6.5.3 Enclosure Seam Design

The design of seams requires that joints be arc-welded, bolted, spot-welded, or treated to produce

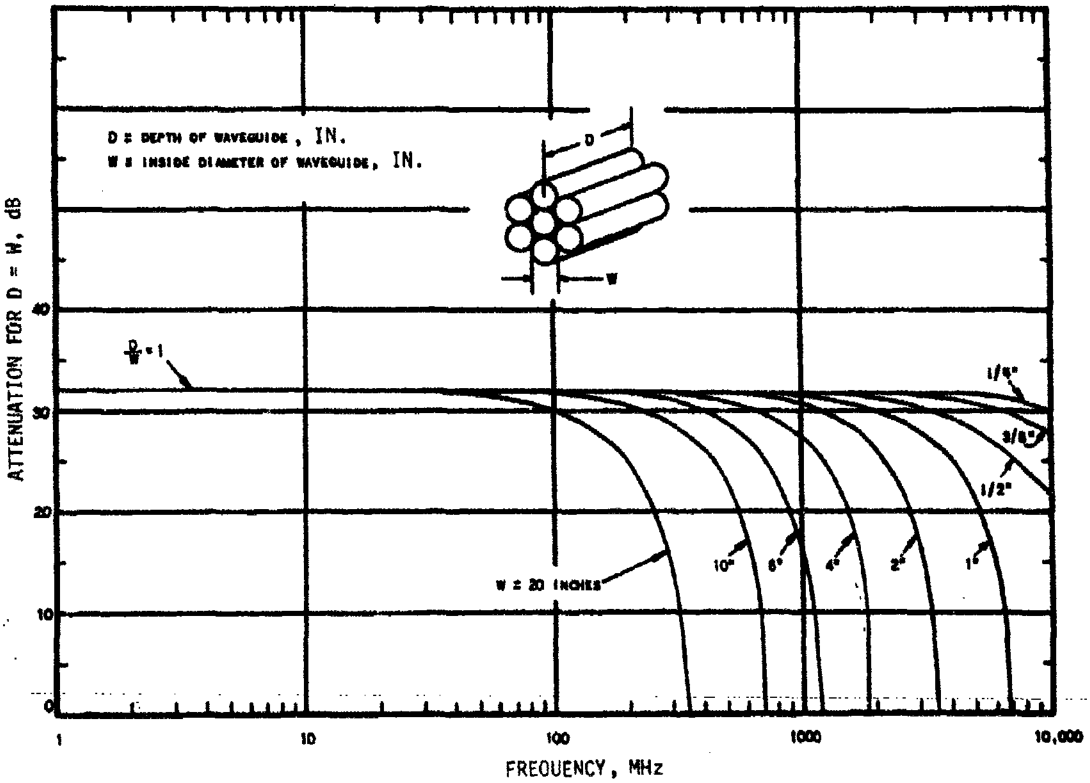


Figure 4-107. Attenuation — Circular Waveguide

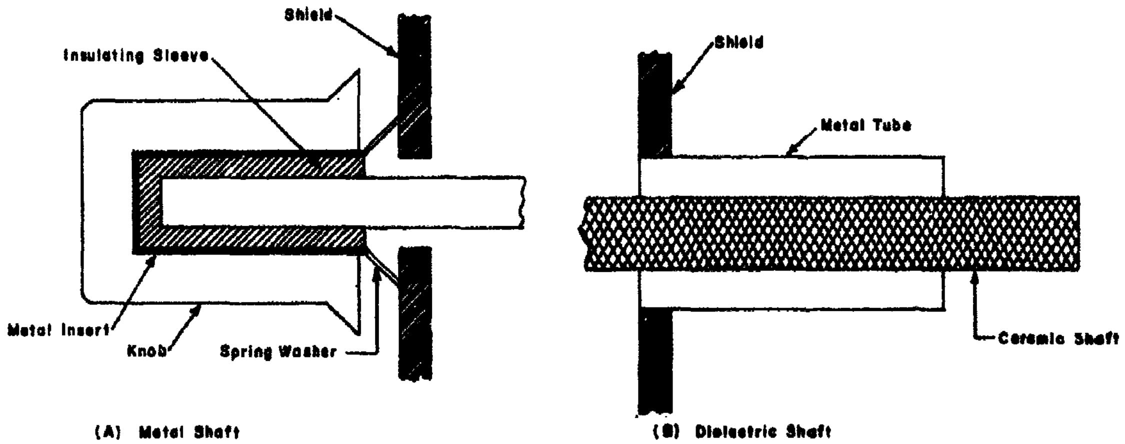


Figure 4-108. Shaft Arrangements

continuous metallic contact. If a material with comparatively poor conductivity is used, the depth of material through which signals must pass should provide a shielding efficiency similar to that of the case material itself. Fig. 4-109 illustrates a well-designed seam, using solder along the edges, with a conductivity of 1/10 that of copper. Complex seams are openings that incorporate metallic folds or flanges as part of the design (Fig. 4-110). In general, these types of seams provide better shielding efficiency than simpler types.

4-6.5.4 Joints in Shields

4-6.5.4.1 General

When it is necessary to join several parts of a complete shield, it is important that a continuous metal-to-metal contact be maintained along the joint. When the pressure is maintained by screws or bolts, a sufficient number of them must be used to ensure high pressure at the points farthest away from any screw or bolt. Lack of stiffness of mating members produce distortion, as shown in Fig. 4-111, and insufficient pressure for preserving good electrical contact. The design of these joints can be simplified by employing conductive gaskets (electronic weather stripping), especially where close fabrication tolerances need not be maintained. Even with such gaskets, it is difficult to maintain a uniform pressure distribution over a large surface contact area unless hold-down screws are very closely spaced. Recommended spacings are given in Fig. 4-112. Gasket materials include metallic textile gaskets and knitted wire mesh which are available in many different materials such as tin-coated copper-clad steel, monel, aluminum, and silver-plated brass. They can be combined with or imbedded in, rubber or plastic to serve as water, air, and oil seals, as well as interference shields. Other materials in common use are conductive elastomers, and beryllium copper finger stock. To make a joint of this nature effective, the covers must be sufficiently rigid to preclude bowing between screws. To prevent electrolytic action, reference to metal compatibility tables (see Table 4-31) should be made for proper selection of materials. These tables should be prepared for gasket purposes since lists based on actual performance differ from theoretical handbook lists. Numerous combinations of finger, mesh, and elastomeric gasket designs are available commercially so that any sealing problem can be resolved successfully with proper design of mating surfaces. Gasket materials shall comply with applicable requirements of MIL-P-11268.

4-6.5.4.2 Groove Gaskets

There are two general types of compressible gaskets: (1) the flat gasket, and (2) the groove gasket. The groove gasket usually has either a circular or "D" cross section and is installed as shown on Fig. 4-113. The best shielding will occur in a situation in which the gasket makes good contact with the bottom of a groove and the cover. Suggested groove dimensions for a given gasket size are available from gasket manufacturers and should be followed. Gaskets that are too large for a given slot don't make good contact on the bottom because of side friction and bulge out over the edges of the groove on top. Figs. 4-114 and 4-115(B), (C), and (D) illustrate arrangements for use of the groove type of conductive gasket.

4-6.5.4.3 Flat Gaskets

These gaskets are placed between abutting flanges as shown in Fig. 4-115(A). Various types are shown in Fig. 4-116. They may be held in place by sidewall friction, an attachment fin, or positioned by a shoulder as shown on Fig. 4-117. Soldering is generally undesirable because of wicking action by the gasket, making it stiff. Bonding cement, even conductive type, generally is a poorer conductor than the gasket and creates a "hole" wherever it is used. In the sidewall friction method the friction may be too low to hold a gasket in by itself if the gasket is the proper size for the slot. Periodic indentation along the side walls will overcome this problem. These indentations need to be used only every several inches and are very effective in holding a gasket in place and still allow for easy replacement.

4-6.5.4.4 Resiliency

A rule of thumb that applies specifically to fluid gaskets can also be applied to conductive gaskets: the greater the compressibility, the greater the sealability. This principle is illustrated by Fig 4-118, which depicts a simulated joint and three gaskets. Gasket *a* is one-half the height of *b* and *c*, and is very resilient; gasket *c* has the same resiliency as *a*; and gasket *b* is harder than *a*. For simplicity, assume gaskets *a* and *c* have twice the resiliency of *b*. They compress 50% under the force *F* applied to the joint, while *b* compresses only 25% down to 75% of original height. Fig. 4-118(A) shows gasket *a* compressed to 50% at the point of maximum compression; this is not sufficient to seal the joint fully. Gasket *b* is then inserted (Fig. 4-118(B)), but, because it compresses

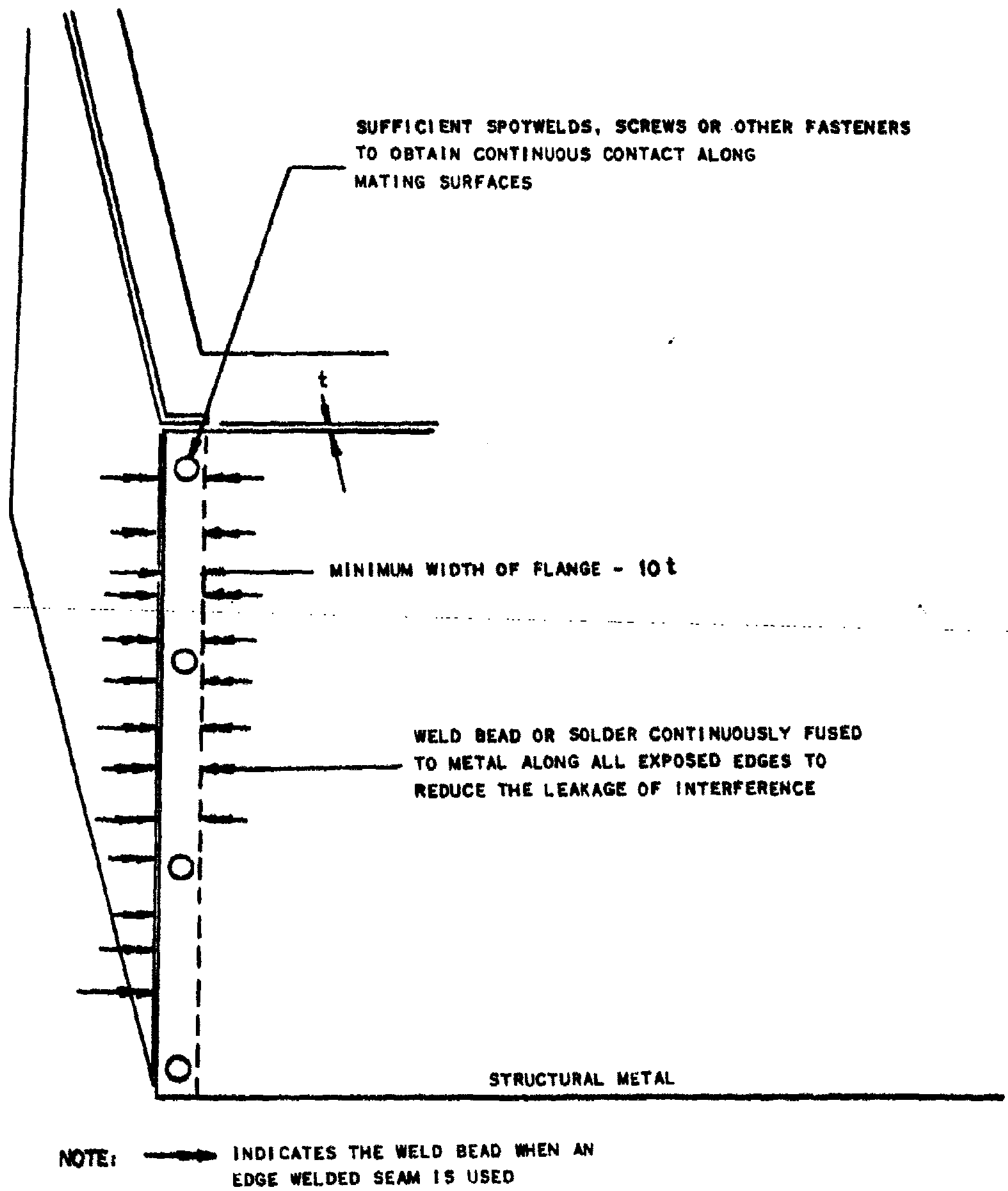


Figure 4-109. Seam Design for Minimum Interference

only 25% under the same force, its greater height does not result in more sealability ($0.5\,h = 0.25 \times 2h$). In Fig. 4-118(C), gasket *c* is compressed 50%, the same percentage as *a*, because they are equally resilient. Because *c* is twice as thick as *a*, the same percentage of compression results in twice the actual compression. In the example in Fig. 4-118(C), this is sufficient to effect a seal. Thus, the gasket must be compressible enough to conform to the irregularities of both surfaces under the applied force. However, an additional requirement is that the contact pressure must be high enough to make adequate contact, even in the presence of any possible nonconducting corrosion films.

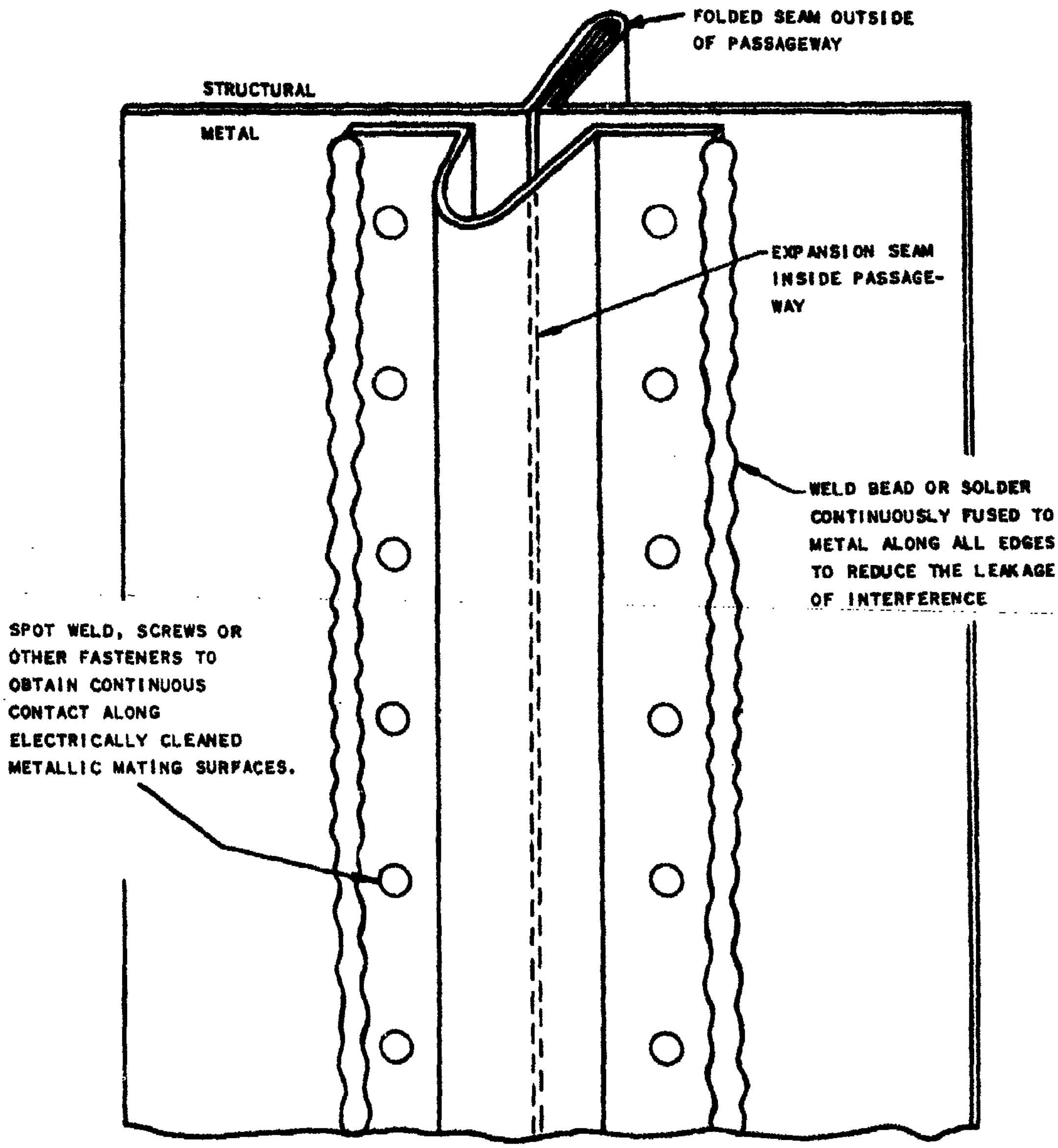


Figure 4-110. Vertical Expansion Joint, an Example of a Complex Seam

4-6.5.4.5 Joint Classification

Joints vary not only in their degree of misfit, but also in the frequency and manner in which they are opened and closed during the life of the equipment. Most resilient metallic gaskets will take some set when compressed, and so will not return completely to their initial state when released. A similar set will occur with the second compression cycle.

Joints are classified as follows:

1. *Class A, permanently closed.* After initial closure, a Class A, permanently-closed joint is opened only for major maintenance or repair. Feed-through mounted interference filters and many wave-guide joints are examples of Class A joints.

2. *Class B, fixed position.* In a Class B joint, the relative positions of mating surfaces and gasket are

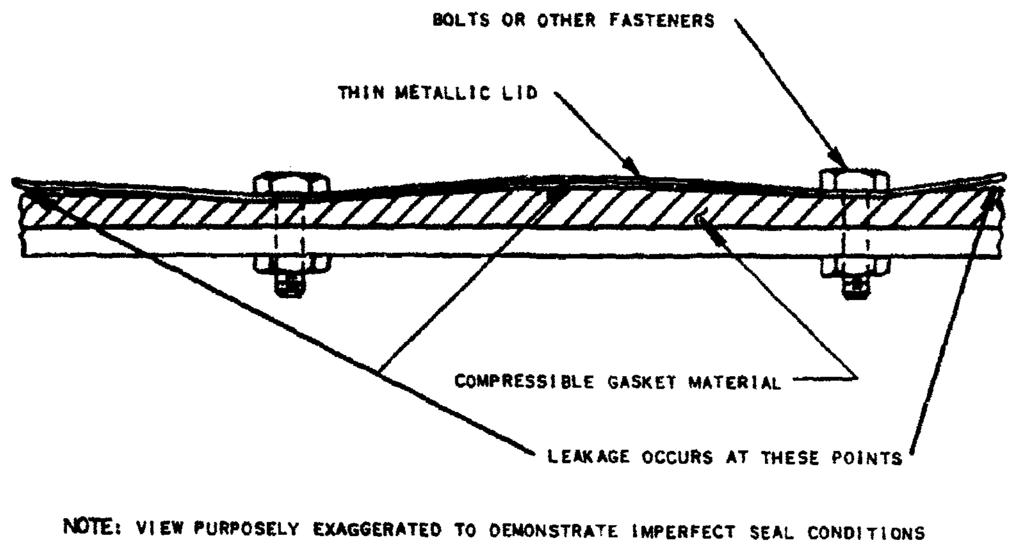


Figure 4-111. *Improper* **Gasket Application**

always the same. For example, any point on the edge of a door will close against its equivalent point on the jamb after every opening. In a Class B joint, the gasket is compressed and released from the same operating height at the same point on the gasket many times during its life. Hinged lids and doors, and rack and panel installations are typical of Class B joints.

3. *Class C, completely interchangeable.* A Class C joint is one in which mating surfaces and shielding materials are completely interchangeable, and/or the relative positions of two mating surfaces and gaskets may change. In a Class C joint, the gaskets may be compressed to several different operating points many times. Symmetrical cover plates and waveguide choke flanges are examples of Class C joints.

4-6.5.4.6 Insertion Loss

Fig. 4-119 depicts insertion loss, in dB, of a typical resilient metal gasket, plotted against applied pressure measured as shown schematically in Fig. 4-120 (Ref. 30). Insertion loss gradually improves with increasing pressure until a limit is reached. Values much above 50 psi generally do not offer much shielding improvement, but generally can be obtained without difficulty in typical enclosure joints. Insertion loss data should be included with each EMC/EMI control plan.

4-6.5.4.7 Gasket Characteristics

Various materials have been used to combine resiliency and conductivity. These characteristics can be combined by using several methods and materials. Gasketing materials, such as woven metal-neoprene combinations, are usually less effective than mesh gaskets. Metal fingers are less effective than mesh gaskets because of the few points of contact and the openings between fingers. They are useful for sliding contacts where other materials cannot be used. Gasketing materials composed of conductive elastomeric materials are especially useful in cases where an air or dust seal also is required. Some of the more common materials are tabulated in Table 4-24 and described in the following paragraphs:

a. *Knitted wire mesh* is made of many interlocked loop-shaped springs; it combines springiness with flexibility and cohesion. These properties are retained when the mesh is compressed and make it dense enough to be an efficient shielding material. These types of gaskets should not be used on magnesium structures.

b. *Beryllium copper gaskets* are 10-mil strips of beryllium copper, made by puncturing thin sheets in both directions with a pointed tool. The resultant sharp raised points do an excellent job of making good contact with both sides of the joint. Such a gasket can be imbedded in rubber to give a good pressure seal as well as a good RF seal. It also can be used

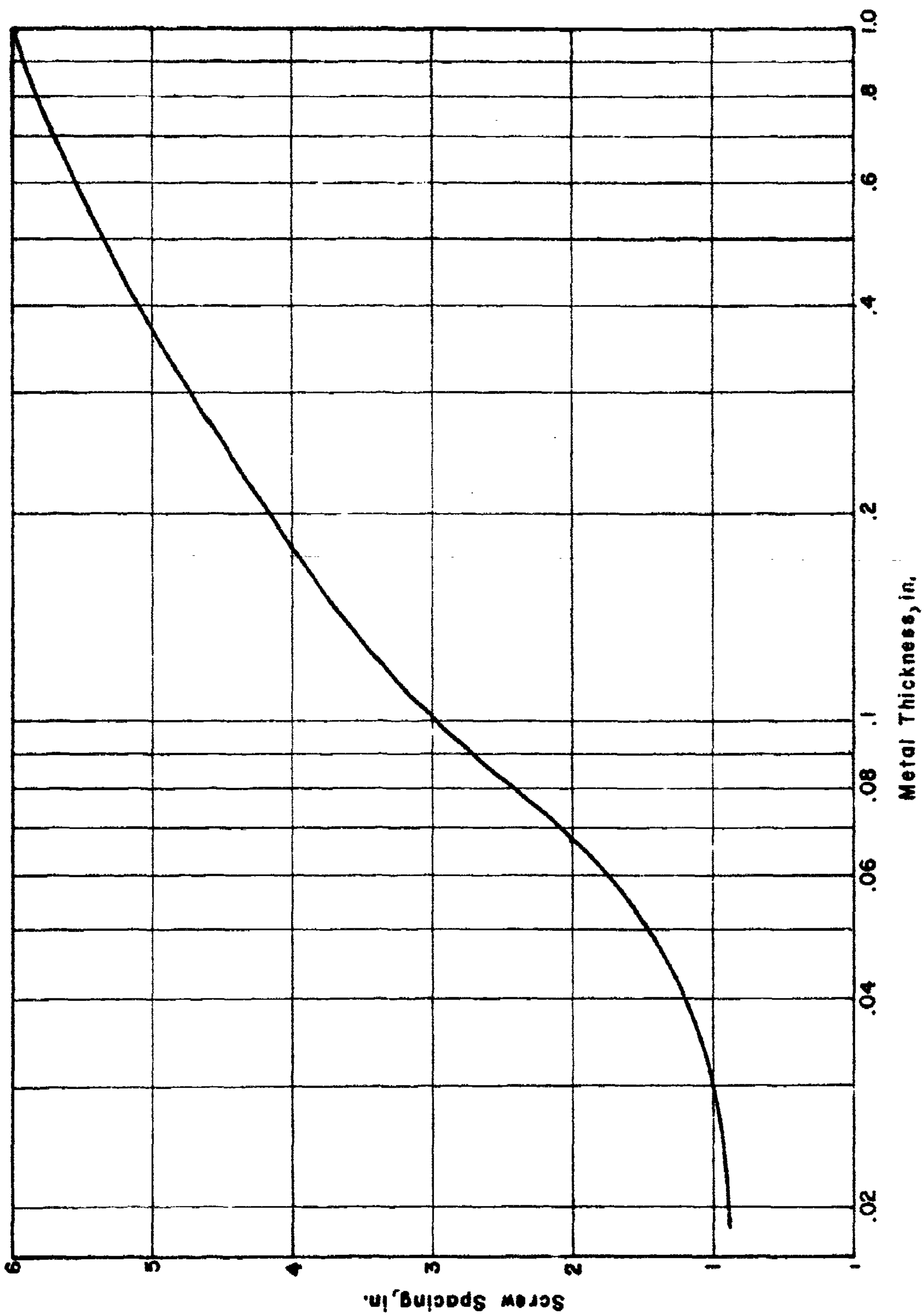


Figure 4-112. Recommended Screw Spacing To Cause a Minimum of Buckling of the Metal

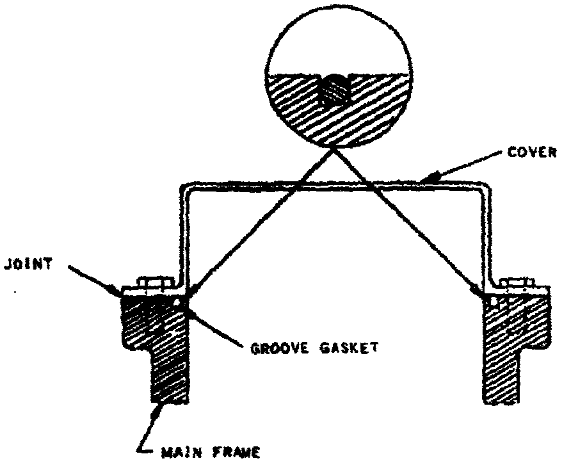


Figure 4-113. Flat-flange Type Joint With Groove Gasket

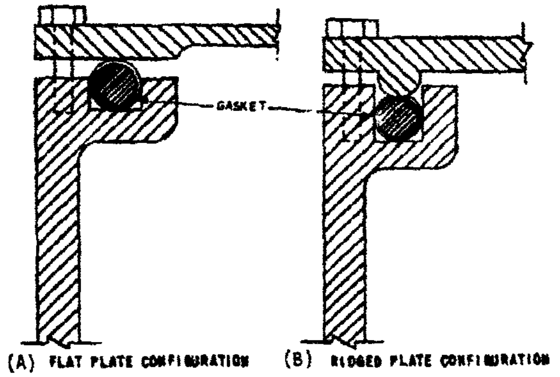


Figure 4-114. Typical Conductive Gasket Applications — Flat Plate and Ridged Plate Configurations

where surfaces are anodized or corroded because it has the ability to cut through the films, making good metal-to-metal contact.

c. *Imbedded wire gaskets* have many wires, each shaped like an open V rotated 90 deg from the plane of the mating surfaces and imbedded in silicone rubber; they make very efficient conductive gaskets. The wires in the gaskets have the ability to puncture any oxide film on mating surfaces, and the bends in the wires allow the gaskets to compress while the ends of the wires establish contact. Because of the small amount of metal in this type of gasket, leakage through the vertical wires limits its effectiveness. This

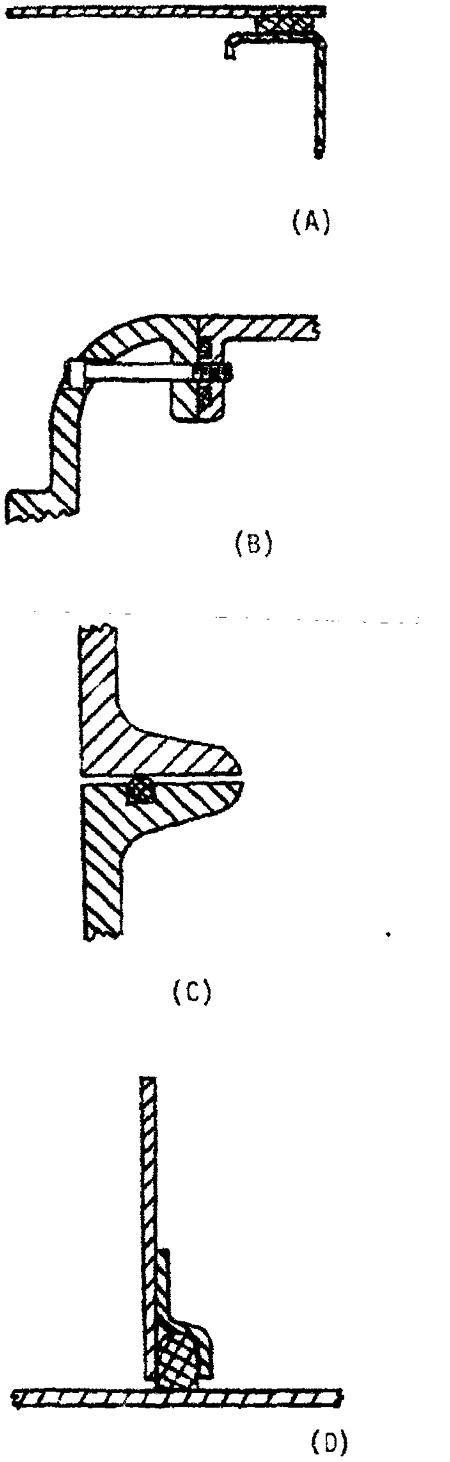


Figure 4-115. Typical Conductive Gasket Applications

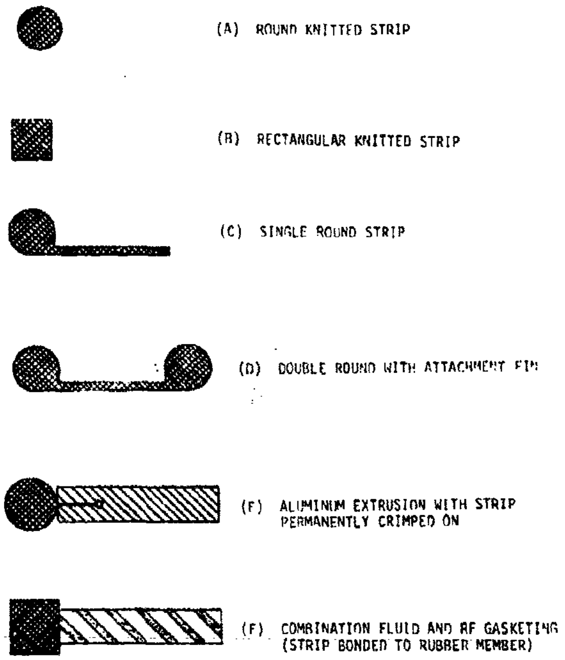


Figure 4-116. Types of "Flat" Gaskets

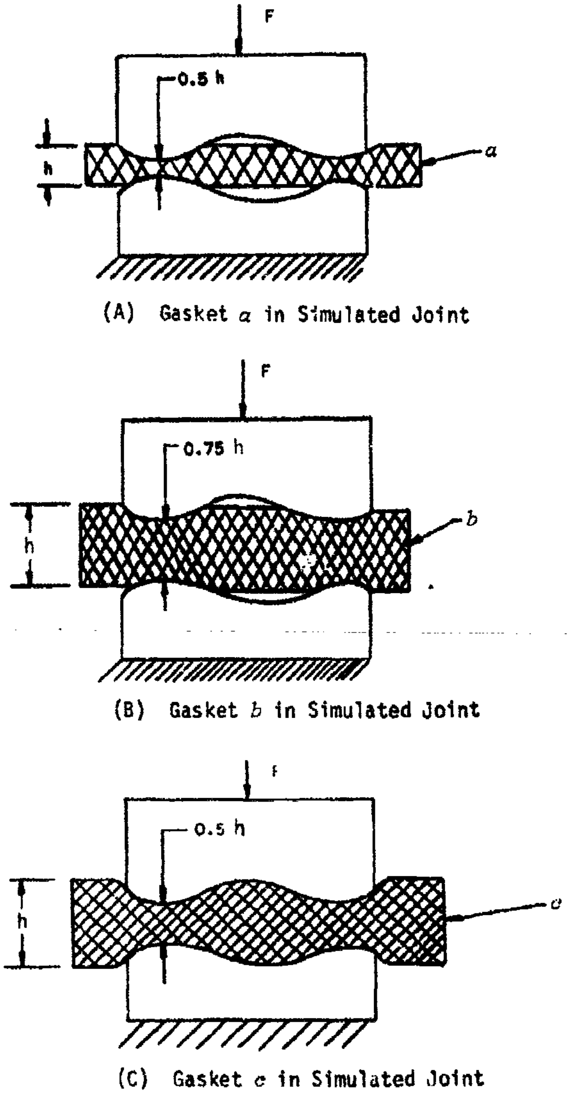


Figure 4-118. Sample Gaskets in Simulated Joints

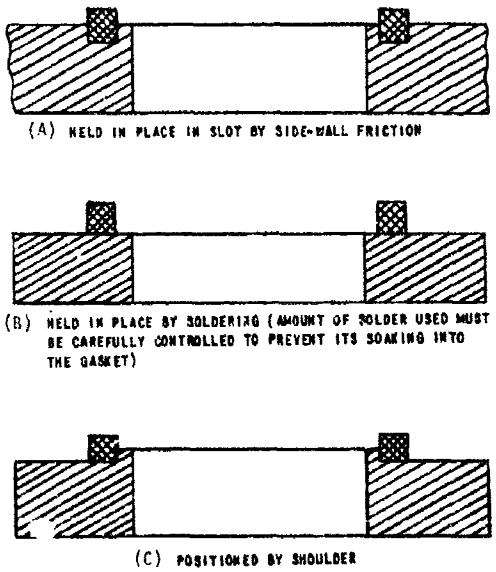


Figure 4-117. Typical Gasket Mounting Methods

type of gasket should not be used on magnesium structures.

d. *Woven aluminum mesh or screen impregnated with neoprene* is ground off to expose peaks of the aluminum mesh on both sides. One example of this type of material uses 16-gage aluminum screen.

e. *Soft metals* such as copper or lead have been used as conductive gaskets.

f. *Metal over rubber.* By covering rubber with metal foil, or wire, the resiliency of rubber is combined with the conductivity of metal.

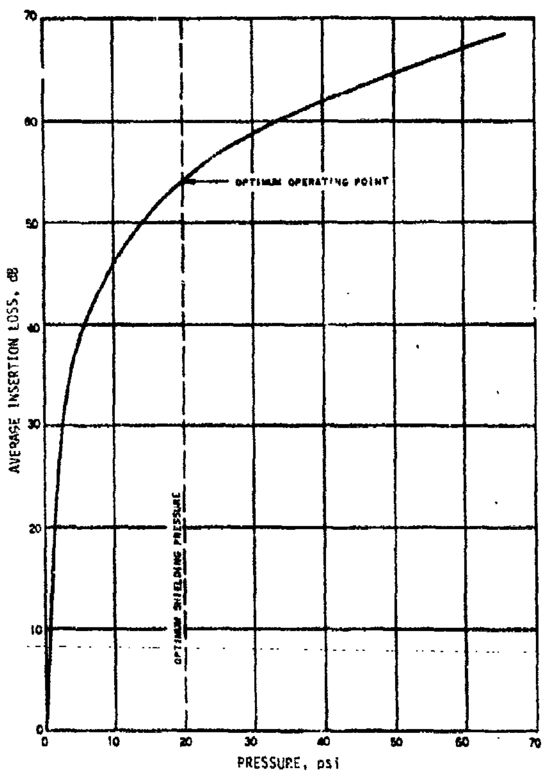


Figure 4-119. Insertion Loss vs Pressure for Resilient Metal Gasket

g. *Conductive elastomers* made of conducting particles embedded in polymer compounds have proven to be very effective, particularly at high frequencies. In general, they require greater pressure than metal mesh for satisfactory compression and necessitate strong, well-designed mating surfaces. Such materials should have a resistivity at least as low as 0.001 ohm·cm to achieve reasonable shielding effectiveness at frequencies as low as 0.15 MHz. However, extreme galvanic action occurs when these materials are coupled with aluminum and magnesium structures — requiring special and unique treatment of the mating surfaces. Unless specifically required by the procuring activity, silver conductive elastomer materials should not be used or coupled with magnesium structures. For aluminum mating surfaces, baked or iridite treatment per MIL-F-14072 should be applied prior to a light (0.001 to 0.002 in.) coating of silver epoxy paint. Detailed surface treatment and finishes required for use with silver conductive elastomer should be included in the equipment specification by the procuring activity.

8. *Serrated contacting fingers* have resiliency achieved by having cantilever springs which make wiping contact between the two surfaces (Fig. 4-121).

9. *Convoluted spiral* material is wound from thin flat spring material which offers a series of low impedance bond paths which readily conform to the unevenness of a joint.

Table 4-25 is a guide to selection of RF gaskets based on mechanical suitability.

4-6.5.4.8 Qualification of Gasket Material

Gasket material should comply with MIL-P-11268 and withstand the equipment environmental requirements of MIL-STD-810. Gasket material should comply and be tested in accordance with Federal Test Method Std. No. 406 and Federal Test Method Std. No. 601, as applicable for the characteristics shown in Table 4-26, as a minimum. These data are submitted as part of the EMI/EMC control plan.

4-6.5.4.9 Panels

Fig. 4-122 illustrates an equipment panel sliding into a case with an internal flange. If the panel also holds the chassis, then the entire weight of the equipment, exclusive of the case, is borne by the flange. Therefore, it is very desirable that there be a rigid, positive step on any gasketing material used in the joint between the panel and the case flange. The gasket shown is ideally suited for this application. It has an aluminum extrusion to which is attached a resilient gasketing material. The panel comes to a stop at the thickness of the extrusion. Because the uncompressed thickness of the material is larger than the thickness of the extrusion, it operates under pressure.

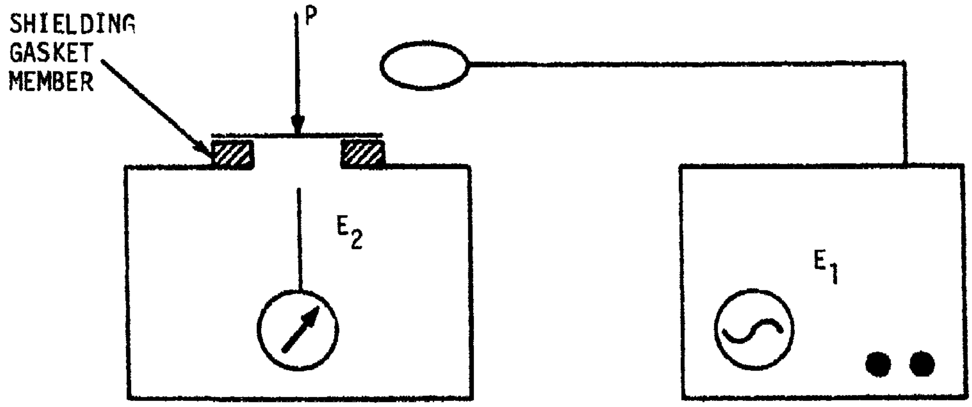
4-6.5.4.10 Connectors

Often, AN-type connectors are mounted on bulkheads and must maintain both electromagnetic radiation and fluid seals. In such cases, conductive elastomer sheet material, or sintered metal impregnated with neoprene, are suitable. The connectors are made to close tolerances and usually are mounted on a surface sufficiently rigid so that the small amount of compression of this material is adequate. These gaskets are often so small that no other material could be used to make a successful combination sealing gasket.

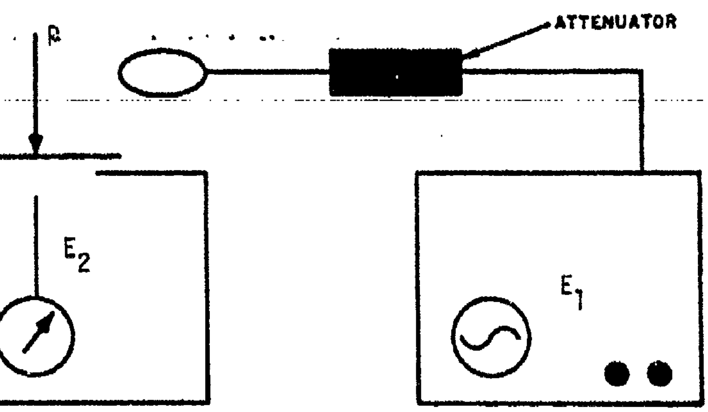
4-6.5.4.11 Pressure Seals

Some applications require seals to be both pressure and interference tight. Fig. 4-123(A) shows one type

Courtesy of Magnetic Metals Co.



(A) E_2 Obtained for Varying Values of P With Gasket Shielding Material. E_1 Held Constant.



(B) E_2 Repeated for Varying Values of P With No Shielding Gasket Material Present. Insertion Loss Measured by Attenuator. E_1 Held Constant.

Figure 4-120. Technique for Measurement of Insertion Loss of Resilient Metal Gasket

of gasket; Fig. 4-123(B) shows another. A conductive gasket is mounted directly to rubber, or other elastic material, to make a combination pressure and interference seal. This combination seal can be held in place by bonding the rubber to one of the surfaces, or by use of the screw holes in the gasket. Conductive elastomer gaskets provide excellent pressure seals and can be obtained in various materials that meet outgassing and corrosive atmosphere requirements. In some equipment there may be a requirement for a pressure differential of only 1 or 2 in. of water, and for cooling air to be ducted so as not to leak out indiscriminately at joints. A combination of airtight and interference-tight gasket material to meet this requirement can be made by knitting one or two layers of mesh over a neoprene or silicone material (Fig. 4-124). It will keep rain out, but not water under pressure. It also represents a compromise as an interference shield, since there is much less metal in this structure than in an all-metal gasket.

Pressure seals at waveguide joints may be obtained with either a flat gasket (for flat flanges) or

TABLE 4-24
CHARACTERISTICS OF CONDUCTIVE GASKETING MATERIALS

MATERIAL	CHIEF ADVANTAGES	CHIEF LIMITATIONS
Compressed knitted wire	Most resilient all-metal gasket (low flange pressure required). Most points of contact. Available in variety of thicknesses and resiliencies.	Not available in sheet form (certain intricate shapes difficult to make). Must be 0.040 in. or thicker.
Beryllium copper gasket or equivalent	Best break-through on corrosion films.	Not truly resilient. Not generally reusable.
Imbedded wire gasket or equivalent	Combines fluid and conductive seal.	Requires 0.25 in. thickness and 0.5 in. width for optimum shielding.
Aluminum screen impregnated with neoprene	Combines fluid and conductive seal. Thinnest gasket. Can be cut to intricate shapes.	Very low resiliency (high flange pressure required).
Soft metals	Cheapest in small sizes	Cold flows, low resiliency.
Metal over rubber	Takes advantage of the resiliency of rubber.	Foil cracks or shifts position. Generally low insertion loss yielding poor RF properties.
Conductive elastomer	Combines fluid and conductive seal.	Relatively high cost.
Contact fingers	Best suited for sliding contact.	Easily damaged. Few points of contact.
Convoluted spiral.	Can provide conduction at forces as low as one pound per linear inch. Diameter of one inch can be obtained.	Not available in sheet form. (Many intricate shapes cannot be made.) Smallest diameter is 3/64 in.

groove gaskets (for choke flanges). In both cases, gasket construction should be adequate to withstand the expected temperature rise at the joint when carrying high power. Its conductivity should be high enough so that it does not increase the heat generated, and the material should not be subject to deterioration from the gas used to pressurize the waveguide.

4-6.5.5 Cable Shielding

As discussed in pars. 3-3 and 4-4, interference may be transferred from one circuit or location to another by interconnecting cabling. The interference may be radiated from a cable or transferred into a cable from external fields. Once interference has been transferred by radiation or common-impedance circuit elements into a cable circuit of an electronic or electrical complex, it can be conducted through interconnecting cables to other elements of the complex. Also, because of cable proximity in cable runs or elsewhere, intracable and/or intercable crosstalk may occur as a result of electromagnetic transference between cables. The cables may be of the commercially available type, prefabricated, or specially assembled from a group of individually insulated conductors.

Electrical cables may be unshielded, individually shielded, shielded as pairs or groups, or shielded as a whole by a single shield. When a shield is used, i

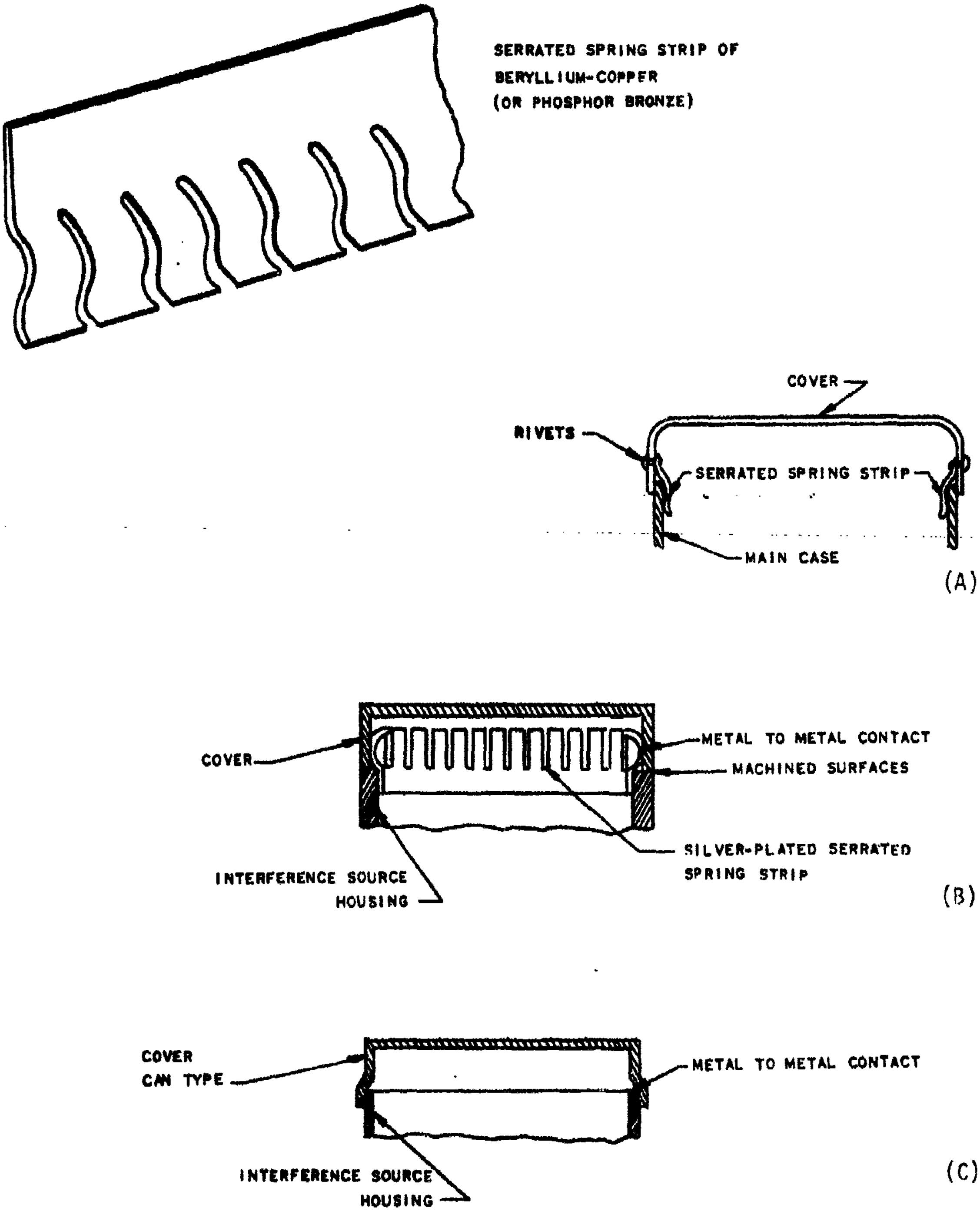


Figure 4-121. Metal-to-Metal Contact Techniqu of Cover and Case for Interference Shields

TABLE 4-25. GUIDE TO CHOICE OF TYPE OF RF GASKET BASED ON MECHANICAL SUITABILITY

1, WELL SUITED; 2, CAN BE CONSIDERED; 3, NOT SUITED

		MESH STRIPS ○	MESH STRIPS □	MESH STRIPS ○_	MESH STRIPS ○_○	EXTRUSION GASKET	COMBINATION STRIP	FORMED RF GASKETS	BE CU GASKETS OR EQUIVALENT	COMBINATION GASKETS	WOVEN ALUM. + NEOPRENE	MESH OVER RUBBER	CONDUCTIVE RUBBER[i]	FINGERS	CONVOLUTED SPIRAL ○	CONVOLUTED SPIRAL ○_
ATTACHMENT METHOD	HELD IN A SLOT OR GROOVE BY TIGHT SIDEWALL FIT	1	1	3	3	3	2	1	3	2	3	1	1	3	3	3
	NONCONDUCTIVE SPOT-BONDING[a]	2	1	2	2	3	2	1	3	2	2	1	1	3	2	1
	NONCONDUCTIVE BONDING AWAY FROM RF GASKET PORTION[b]	3	3	1	1	3	1	3	3	1	3	3	3	3	3	1
	BOND RF GASKET PORTION WITH CONDUCTIVE ADHESIVE[c]	2	1	2	2	3	2	1	3	2	2	1	1	3	2	1
	SCREW, SPOTWELD OR RIVET	3	3	1[d]	1[d]	1	3	3	2	3	3	3	3	1	3	1
	SOLDER	3	2	2	2	3	3	2	2	3	3	3	3	1	3	3
	POSITION-BY-BOLTS THROUGH BOLT HOLES	3	3	2	2	1	1	2	1	1	1	3	1	3	3	1
	PRESSURE-SENSITIVE ADHESIVE BACKING	3	3	3	3	3	1	3	3	1	3	3	3	3	3	1
OTHER GASKETING FUNCTIONS	COOLING AIR TIGHTNESS	3	3	3	3	1	1	3	3	1	1	1	1	3	3	1
	RAIN TIGHTNESE	3	3	3	3	1[e]	1	3	3	1	1	1[e]	1	3	3	1
	PRESSURE TIGHT	3	3	3	3	3	1	3	3	1	1	3	1	3	3	1
PRESSURE AVAILABLE	0 - 5 PSI[f]	1	1	1	1	1	1	1	2	1	3	1	1	1	1	1
	5 - 50 PSI	1/2	1	1/2	1/2	1	1	1	1	1	1	1/2	1	2	1	1/2
	OVER 50 PSI	2	2	2	2	1	2	2	1	2	1	2	1	3	1	2
TOTAL JOINT UNEVENESS[g]	LESS THAN 0.002	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
	0.002 TO 0.030	1	1	1	1	1	1	1	1	1	3	1	2	1	1	1
	0.030 TO 0.060	1	1	1	1	1	1	1	2/3	1	3	1	2/3	1	1	1
	OVER 0.060	1	1/2	1	1	2	2	1/2	3	2	3	1	3	1	1	1
SPACE AVAILABLE, WIDTH	LESS THAN 0.060[h]	1	1	3	3	3	3	2	3	3	1	3	1	3	1	1
	0.060 TO 0.500	1	1	1	2	2/3	2	1	2	2	1	1	1	2	1	1
	0.500 TO 1.50	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
SPACE AVAILABLE, THICKNESS	LESS THAN 0.030	3	3	3	3	3	3	3	2	3	1	3	2	2	3	3
	0.030 TO 0.060	1	1	2	2	3	2	1	1	2	3	2	1	1	1	3
	0.060 TO 0.090	1	1	1	1	2	2	1	1	2	3	2	1	1	1	1
	OVER 0.090	1	1	1	1	1	1	1	3	1	3	1	1	1	1	1
TYPE OF JOINT	COMPRESSION ONLY	1	1	1	1	1	1	1	1	1	1	1	1	1	1	1
	COMBINED COMPRESSION & SLIDING	2	2	2	2	2	2	2	3	2	3	2	2	1	2	2
	SLIDING ONLY	2	2	2	2	2	3	2	3	3	3	2	3	1	2	2

a - NONCONDUCTIVE SPOT BONDING: A NONCONDUCTIVE ADHESIVE CAN BE USED DIRECTLY UNDER AN RF GASKET IF IT IS USED ONLY IN 1/8-TO 1/4-INCH DIAMETER SPOTS, 1- TO 2-INCHES APART.

b - A NONCONDUCTIVE ADHESIVE CAN ALWAYS BE USED CONTINUOUSLY IF IT IS USED UNDER THE ATTACHMENT OR RUBBER PORTION OF COMBINATION STRIP AND GASKETS, BUT NOT UNDER THE RF GASKET ITSELF.

c - A CONDUCTIVE ADHESIVE CAN BE APPLIED CONTINUOUSLY UNDER AN RF GASKET.

d - WITH BACKING STRIP OVER ATTACHMENT FINS.

e - IF MESH-OVER-RUBBER VERSION OF EXTRUSION GASKET IS USED.

f - EVALUATION IS ONLY FOR MECHANICAL SUITABILITY. PRESSURE MAY BE HIGH ENOUGH TO GIVE SUFFICIENT INSERTION LOSS.

g - EVALUATION BASED ON SPACE BEING AVAILABLE TO USE THICK ENOUGH GASKET.

h - INCLUDING SPACE FOR ATTACHMENT METHOD INTEGRAL TO MATERIAL CONSIDERED.

i - EVALUATION IS NOT BASED ON ELECTRICAL SUITABILITY OF CONDUCTIVE RUBBER WHICH IS GENERALLY POORER THAN OTHER MATERIALS LISTED.

TABLE 4-26
GASKET MECHANICAL TEST REQUIREMENTS

Parameter	Method	
	Std. No. 406	Std. No. 601
Brittleness	2051	5311, 5321
Compression	1021	12151, 3321, 3331, 12141, 3311, 12131, 5411
Durometer	1082, 1083	3025, 3021, 5511
Elongation	1011, 1012 1013, 1063	4121, 10311, 11021, 13031
Tear Strength	1121	4211, 4221
Tensile Strength	1101, 1012 1013, 1063	4111, 11011, 13021 6111, 6121, 4131
Volume Resistivity	4042	9111

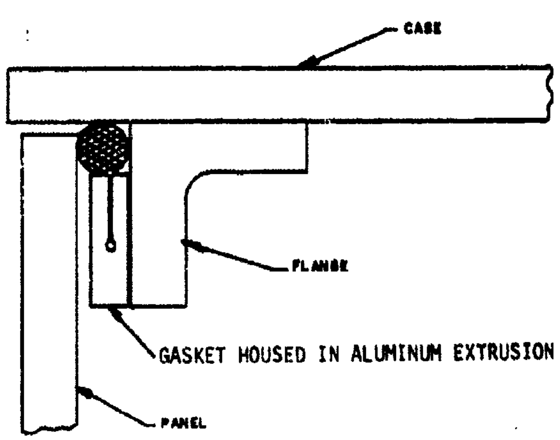


Figure 4-122. Typical Conductive Gasketing Application

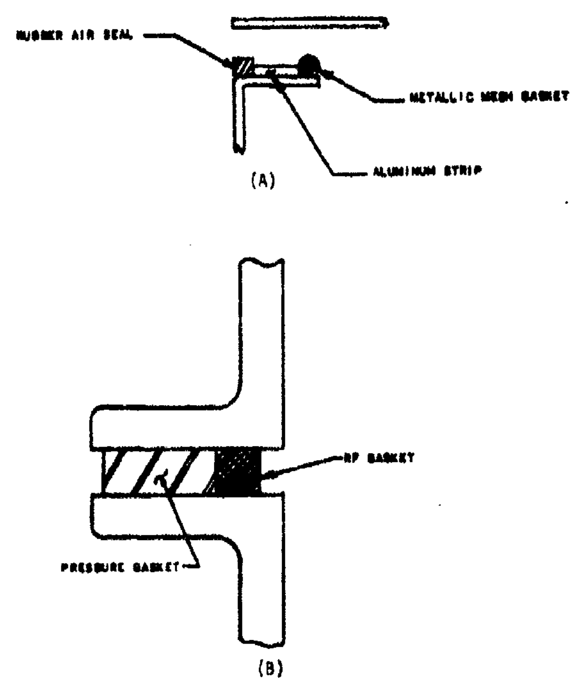


Figure 4-123. Combination Pressure and Interference Seals

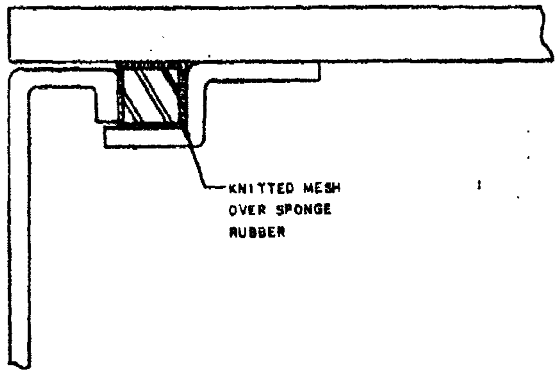


Figure 4-124. Raintight Conductive Gasket Seal

generally consists of a braid or conduit. The purpose of such shielding is to attenuate radiated interference and susceptibility. Proper care should be taken during installation to insure that shielding integrity is maintained. The choice of cable to be used depends upon the characteristic impedance desired, amount of signal attenuation permitted, environment within which the cable must exist, and characteristics of the signal to be transmitted. Signal isolation between different cable circuits is a function not only of cable shielding and the character of signals being distributed, but also of the physical separation of high-level and low-level circuits as discussed in par. 4-4.5. The effectiveness of a shield is a function of the conductivity of the metal, strand sizes, percentage of coverage, and size of openings. Multiple layers of shielding, separated by dielectric material (except at connectors), are much more effective than a single layer of shielding. Although leakage power may be a very small percentage of transmitter power, if the power being carried by a cable is large, this percentage might cause a considerable interference problem to sitive circuits.

4-6.6 MAGNETIC SHIELDING TECHNIQUES

At low frequencies, otherwise well-engineered circuits can be affected seriously by interference from spurious magnetic fields set up by neighboring components. Experience has shown that these effects can be minimized by judicious orientation of components on the chassis. There are many instances, however, where this cannot be accomplished because of space limitations. In addition, many low-level applications can be upset by the earth's magnetic field. This can be critical in amplifier circuits where the extraneous signal may be amplified thousands of times. For example, the field generated by the driving motor of a tape recorder may be of sufficient magnitude to dominate the signal completely unless the record and playback heads are well-shielded. The electron beam of a cathode-ray tube can be bent by the field of a transformer used in the circuit. A magnetic compass is of such strength as to affect a radar scope 6 ft away. Because of such effects, input transformers handling very low-level signals must be protected from any spurious fields. Some relief can be obtained in the design of transformers by using a humbucking construction, where some of the fields are bucked out by splitting the windings on the magnetic core structure. Another solution is the use of a core structure in which air gaps between lamination pairs (where fringing of flux develops) are contained within the electrical coil. In general, the solution of most of these problems lies in the use of a magnetic shield around the affected component.

4-6.6.1 Materials

At low frequencies, attenuation occurs because the magnetic shield has a high permeability and therefore provides a low-reluctance path for the magnetic flux. The high μ-materials generally have the same basic constituents of 80% nickel and 20% iron. An 80% nickel alloy is difficult to draw and form. The best way to handle these alloys is by the polyform process — casting a matrix in a die or mold to the desired shape. The matrix is rotated and cooled while the magnetic shielding material is sprayed on it to the desired thickness. When complete, subsequent grinding, drilling, or other machining is carried out; then the matrix is removed, leaving an enclosure that is adequate for electronic assemblies. The enclosure will have internal dimensions exactly matching the matrix and therefore will be superior to a fabricated can, particularly when sharp corners are required. To achieve optimum efficiency and add to ductility, annealing usually follows this stage.

The shielding effectiveness is somewhat dependent on the position of the shield surface with respect to the orientation of the magnetic field, and upon the structure and treatment of the material. For example annealing of high permeability materials usually increases shielding effectiveness, while physical shock decreases it. Fig. 4-125 gives data which are considered representative for 1/32 in. thicknesses of various materials (Ref. 22).

Practically all magnetic materials are subject to magnetic saturation effects at high flux densities of the order of 1 tesla as shown on Fig. 4-126. These high flux densities are likely to be experienced in practice only near conductors carrying hundreds of amperes, or near devices purposely designed to produce high fields such as electromagnets. Large electrical motors, generators, and transformers are likely to produce moderate fields because of their enclosed magnetic circuit construction, but next to the cables connected to these devices and associated switchboards the 60-Hz fields may be large. These fields can be controlled by using relatively low-μ materials — which will not saturate — to enclose switchboards and for conduits for the cables. When weight is an important consideration, the choice of material must be made with care.

Multiple shields may be used for two reasons:

a. Where large attenuation is required, two thin layers of material separated by a space filled with air, or a low-permeability material will produce more shielding effectiveness than the same amount of material in a single shield.

b. If the field is large enough to cause saturation of high-μ material, a low-μ material is used on the high field side to reduce the field to a value that will not saturate the high-μ material.

Laminates of magnetic and high-conductivity materials sometimes are used; the former for shielding of a magnetic field, and the latter for shielding of an electric field. To insure maximum permeability from these materials, they must be prepared metallurgically with great caution.

4-6.6.2 Structures

Structurally, magnetic shields fall into two broad classes: (1) those produced by deep drawing from flat blanks, and (2) those formed and welded. Because of difficulties involved in deep drawing nickel-iron alloys, drawn shields usually are confined to smaller sizes; and because nickel-iron alloys work harden

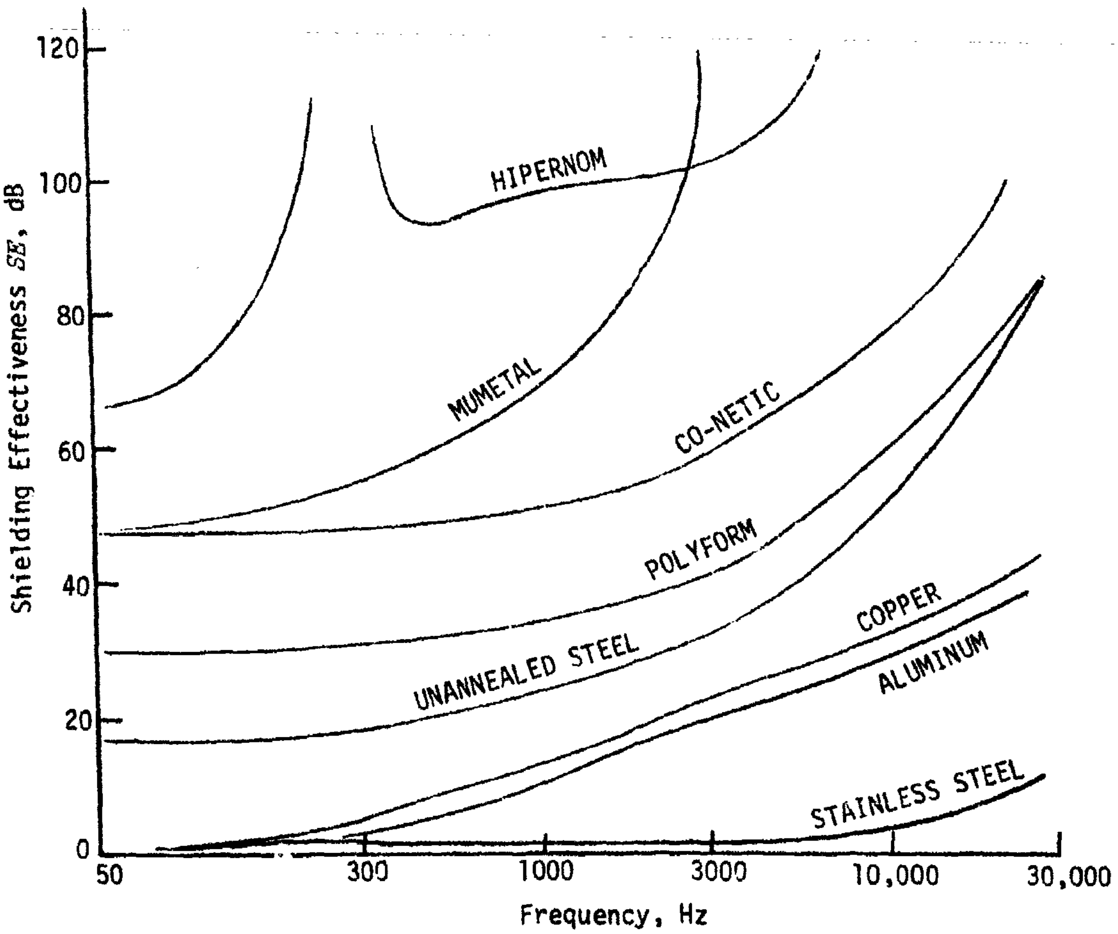


Figure 4-125. Comparison of Shielding Effectiveness vs Frequency for the Various 1/32-in. Thick Shield Cans

very rapidly in the drawing process, generous radii must be provided to prevent tearing. Shields for such items as transformers, cathode-ray tubes, and photomultiplier tubes are fabricated from flat, unannealed sheets of metal. The material is bent on brakes or rolls and the joints are overlapped and spot welded. All holes and slots are pierced prior to the forming operation. It has been found that a 3/8-in. overlap of material is sufficient to prevent any penetration by the extraneous field. Spot welding at intervals of 0.5 in. is adequate to seal the joint.

Placement of joints in the shield surface affects the shield effectiveness. In a cathode-ray tube shield, for example, joints in the axial direction of the tube have little degrading effect on shielding effectiveness, but joints normal to the axis reduce the effectiveness of the shielding. Fig. 4-127 shows examples of these constructions, and also the developed blanks from which they are formed. The blank used to form the shield with the axial joint (Fig. 4-127(A)) has a maximum flux path; the normal joint in the second example (Fig. 4-127(B)) divides the blank into two parts that have lower permeance, resulting in decreased shielding effectiveness. If space and mounting difficulties are no problem, costs can be cut by eliminating expensive layouts, as shown on Fig. 4-128. Here, a simple frustrum of a cone that follows the contour of the tube is substituted for a three-piece construction (Fig. 4-128(A)). In addition to the economy realized through simpler construction, annealing costs can be

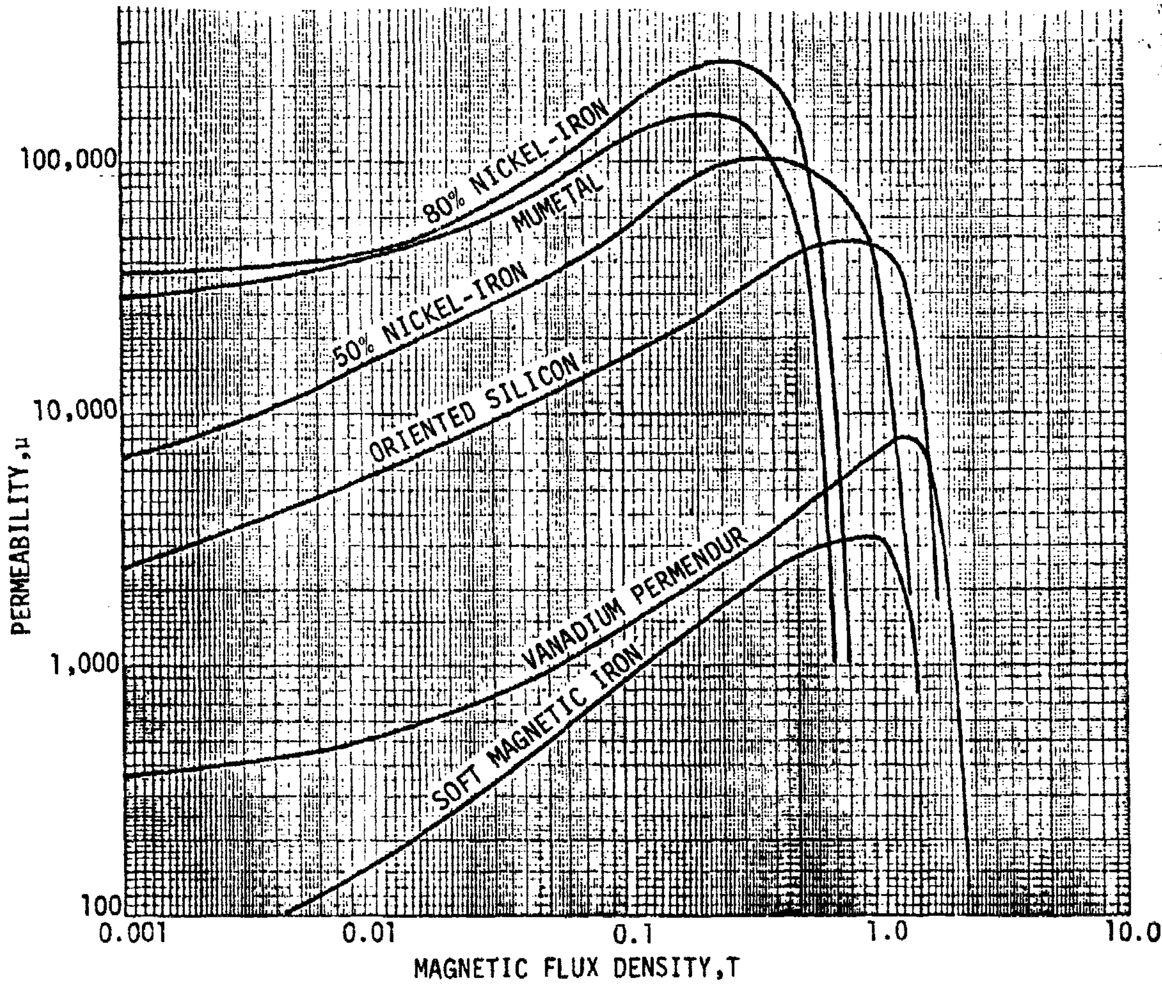


Figure 4-126. Typical Curves Showing Effects of Magnetic Saturation on Permeability of Various Materials

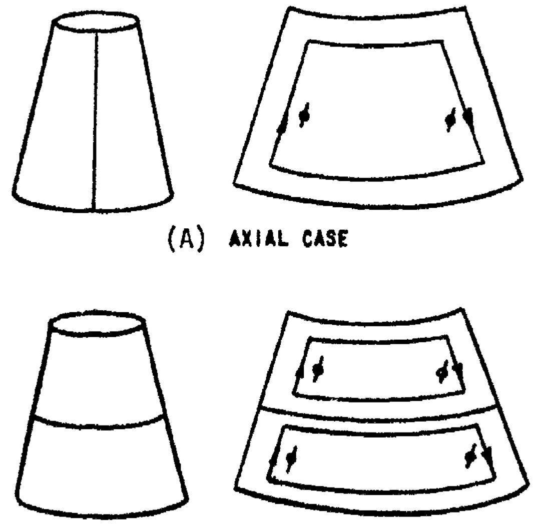


Figure 4-127. Effect of Joint Orientation on Flux Paths

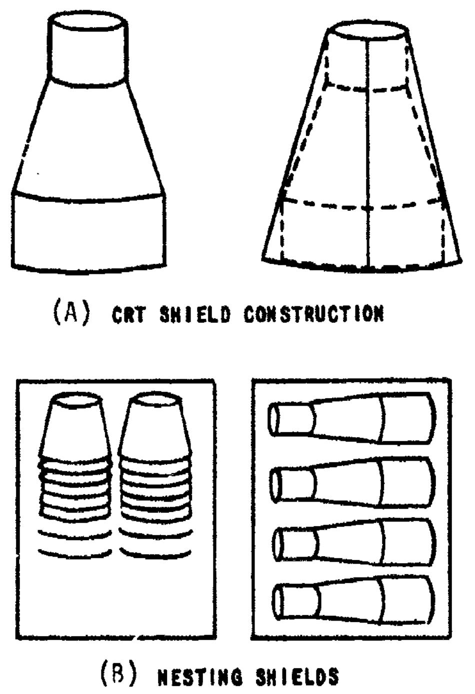


Figure 4-128. Shield Economies

greatly reduced by nesting the cones in the annealing box, as shown on Fig. 4-128(B).

The question often is raised as to the possibility of using annealed Mu-metal sheets to eliminate the much higher cost of annealing in the final form. Unfortunately, the permeability of Mu-metal is adversely affected by cold work of any kind; therefore, the effectiveness of this shield would be inferior to one made in the recommended manner and then given a final heat treatment. This is particularly true because most shields are produced from fairly heavy-gage material (0.025 to 0.035 in. thick).

4-6.7 SHIELD PENETRATIONS

For good shielding construction, it is necessary for all items that penetrate the shielding, such as pipes and conduits, to be electrically bonded to the shielding at the point of entrance by soldering, brazing, or welding. Handles, latches, screw heads, nails, and other metal projections that pierce the shield should be brazed or soldered to the shield, all breaks should be bonded in continuous seams. These precautions prevent the antenna effect, which can occur when a metal element projects through the shielding and acts as a receiving antenna on one side, picking up radiated energy and reradiating signals in the opposite direction on the other side. At high frequencies, such isolated hardware is comparable to, and can radiate as, a waveguide probe. Minimal impedance between such a projection and the shielding will eliminate the antenna effect.

Data output openings — such as those for direct view storage tubes, cathode-ray tubes, and meters — represent a large discontinuity in an equipment case. Meters and other visual readout devices often present difficult interference protection problems, particularly at high frequencies, because, at present, materials that are both optically transparent and conductive are ineffectual shields. If fine knitted mesh or conductive glass is used for display shielding, it may not furnish the degree of attenuation required. In most cases, this problem is overcome by installing a shield around the rear of the readout device and filtering or bypassing all leads entering and leaving it, as shown on Fig. 4-129. A number of semitransparent shielding materials are available for application to data output and display devices. Included are copper mesh screening, perforated metal, conductively coated glass, and conductively coated plastic. The optical and electrical transmission properties of some of these materials are summarized in Table 4-27. A comparison of relative levels at three frequencies indicates that, for virtually all of the conductively coated materials, the ratio of electrical transmittance to optical transmittance is relatively high. For example, a 30-micron-thick gold film on plastic yields an electrical transmittance of 0.16% (at 5.9 GHz) and an optical transmittance of 24%. The electrical transmittance decreases with frequency; at higher frequencies, considerably higher electrical transmittance (poorer shielding qualities) can be postulated. The 20-mesh copper screen, while showing the same order of electrical transmittance as the gold film at 5.9 GHz, has approximately twice the optical transmittance. In addition, the shielding effectiveness of copper mesh improves with decreasing frequency, although at higher frequencies, comparison of the two materials would favor the copper mesh. Economically, copper mesh is more practical because deposition of thin metal films is an expensive process compared with using copper mesh.

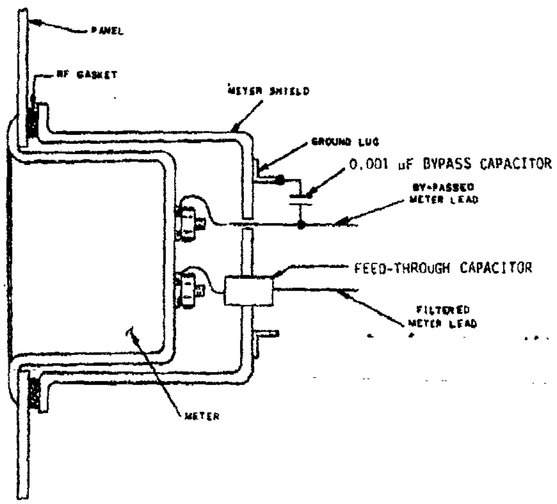


Figure 4-129. Meter Shielding and Isolation

4-6.7.1 Direct-View Storage Tubes

The application of magnetic shielding to direct-view storage tubes practically eliminates the effects of stray magnetic fields. A double shield is required to stop these fields effectively. It may consist of one shield of Netic-type alloy and one shield of Co-Netic-type alloy. Principal Netic-type alloy material characteristics are high-flux capacities and extremely low retentivity. Co-Netic-type alloy materials are more effective for low-intensity, low-frequency problems. The inner shield should be of Co-Netic type material or an equivalent; the outer shield, Netic type material or an equivalent. The Co-Netic-type shield should be 0.025 in. thick; the Netic-type shield, 0.062 in. thick. In this arrangement, the dc magnetic field is reduced by a factor of 1000, and the ac magnetic field, tested at 60 Hz, is reduced by a factor of 30,000.

TABLE 4-27
MICROWAVE AND OPTICAL PROPERTIES OF SEMITRANSPARENT SHIELDING MATERIALS

Material	Microwave Transmittance, Percent			Optical Transmittance, Percent
	5.9 GHz	9. GHz	18.8 GHz	
Gold film about 11 μ thick on plastic (300 ohms/square)	23	10	0.8	49
Gold film about 30 μ thick on plastic (12 ohms/square)	0.16	0.1	0.01	24
Gold film about 75 μ thick on glass (1.5 ohms/square)	0.04	0.01	0.004	3.2
Copper mesh (20 per in.)	0.1	0.2	0.2	50
Copper mesh (8 per in.)	1.0	1.3	2.5	60
Lead glass (X-ray protective, 1/4 in. thick)	30	25	16	85
Lucite (3/16 in. thick)	80	50	25	92
Libby-Owens-Ford Electrapane glass, with conductive coating about 150 μ thick (120 ohms/square)	16	16	16	85
Libby-Owens-Ford Electrapane glass, with conductive coating about 300 μ thick (70 ohms/square)	9	10	8	80
Corning heating panel glass, with conductive coating about 1.5 μ thick (15 ohms/square)	1.6	1.2	0.08	45

4-6.7.2 Cathode-Ray Tubes

Cathode-ray tube openings represent one of the largest discontinuities and are of a type most difficult to treat. The disruption of the shield is such that, if not considered in the initial design stages, it may become extremely difficult to achieve an acceptable final product. Treatment of the opening is made additionally difficult by the requirement for unrestricted transparency (Fig. 4-130).

4-6.7.3 Indicating and Elapsed Time Meters

Meter movements, when shielded, are not affected by ac or dc magnetic fields. The most common source of interference to unshielded meters are transformers, motors, generators, current-carrying cables or buses, solenoids, and other components producing magnetic fields. Shielding against interference from such items permits unrestricted use of meters in otherwise difficult environments.

4-6.7.4 Fuse Holder and Indicator Lamp Openings

In electronic equipment, both active and spare fuses may be sources of radiation because they can act as antennas. The lack of shielding in most fuseholders permits internal high-frequency interference to propagate through the opening in an otherwise well-shielded panel. One solution is to group all fuseholders together; a shield of solid metal with wire mesh gasketing then may be used to surround the fuse cluster. This approach can be used for indicating lamps, provided screening or special conducting glass is substituted for the solid-metal fuse holder shield.

4-6.7.5 Switching Devices

Components, such as solenoids or other devices involving high inrush currents or incorporating switching devices that normally develop high amplitude transients, can prove a source of difficulty in an interference-free design, particularly where space is at a premium.

Relay coils can produce strong local fields which may require a magnetic shield. Otherwise, the magnitude of the current switched may produce transients on the wires or cables carrying this current.

4-6.7.6 Cables

Interconnecting wires and cables may require shielding, either because they are carrying large currents or are connected to sensitive devices. The shields may be made of metal braid, or it may be a solid metal tube (see par. 3-3.1.1.2.5). In the latter case corrugated construction will produce considerable flexibility in a tube having as good a shielding effectiveness as a rigid tube of the same material and thickness. Shields for application on motor vehicles ignition and other circuits are discussed in par. 5-6.3.2.2.

Another form of cable shield is heat shrinkable tubing and boots of various types containing an internal conductive coating which is effective for electric field shielding. Where such materials are permitted, they are convenient because of their flexibility in application.

For low frequency applications, solid tubing of magnetic materials of both the rigid and flexible types are available. For use on cables carrying high currents, special layer type construction may be necessary to avoid saturation.

An adequate magnetic shield often is developed by wrapping a continuous layer of annealed tape around the cable. A typical application may involve shielding a cable of approximately 0.5 in. diameter, which has to be flexible in the final assembly. Annealed mumetal tape 0.001 in. thick and 0.25 in. wide wrapped in two layers should prove suitable. The first layer can be spaced approximately 0.125 in. between convolutions, with the second layer overlapping the first layer to cover the gaps between turns. The assembly can be covered with a protective rubber coating and may be flexed without losing its shielding effect. A form of shielded cable using four counterspiral-wound bands of foil, Netic, Co-Netic or their equivalent, is also recommended. This construction is shown on Fig. 4-131. The strips can be from 0.25 in. to 1 in. wide. To minimize leakage between unavoidable gaps, it is necessary to wind the material so as to permit spiral positioning along the length of the cable, with each following layer consisting of another spiral in the opposite direction. Successive layers of the tape wound in this manner insure a minimum of gaps and permit flexibility. Such spiral-wound shielded cables are commercially available. A design engineer, who encounters the need for a shield of this nature, can procure the tape in foil form and, for evaluation purposes, fabricate a prototype shield for his own cables. A total of four wraps, or multiples of four, may be necessary for cables carrying appreciable current. For conductors carrying currents greater than 2 A, the first two layers should be Netic S3-6 foil or its equivalent; the remaining layers should be Co-Netic AA foil or its equivalent. Netic and Co-Netic type foils, or their equivalents, are available from 0.002 to 0.007 in. in thickness and in various widths. Also, these alloys provide a simple method of shielding transformers and small reactors; the foils are wrapped carefully around them with the necessary number of layers to provide the desired attenuation level. After wrapping, the cable can be

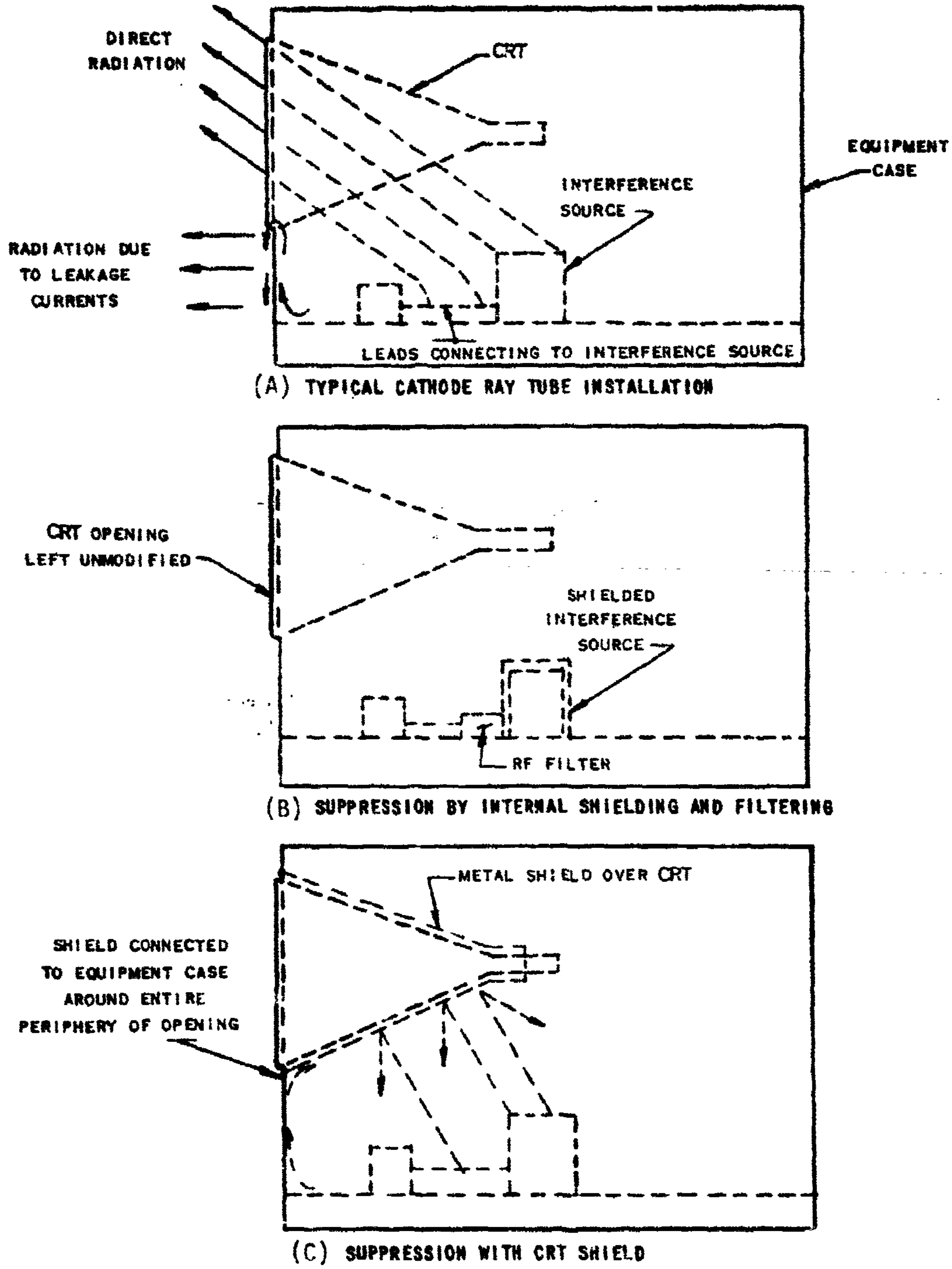


Figure 4-130. Treatment of Cathode-ray Tube Openings

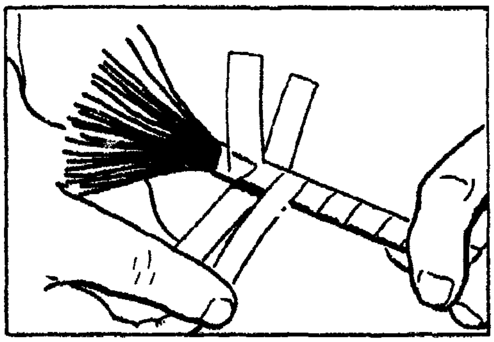


Figure 4-131. Shielding Cable With Four Bands of Foil

potted or encapsulated to prevent unraveling of the foil. The Netic-type foil and its equivalent exhibit the ability to carry relatively high flux densities; the Co-Netic-type foil and its equivalents have the property of offering maximum attenuation. When an interfering field is of sufficient intensity to saturate the Co-Netic-type foil partially, and therefore limit the realizable attenuation, Netic-type foil can be placed between layers of Co-Netic foil. The Netic-type foil, under these circumstances, acts as a buffer for the Co-Netic-type foil. These foils are extremely thin and cannot function effectively in fields that approach their saturation levels. The extreme versatility of these materials should help the engineer to develop many applications where their use will save space, improve signal-to-noise ratios, and considerably reduce interference.

4-6.7.7 Conductive Surface Coatings

Highly conductive surface coatings specifically formulated for shielding applications are readily available. They are fine, silver-based lacquers that adhere excellently to metal, plastic, ceramic, wood, and concrete. When applied to a nonconductor, the surface resistivity is substantially less than 1 ohm/in.[2] Successive coats further decrease the surface resistivity, and some types can be brushed on or sprayed. These coatings often are used to improve the RF integrity of metal housings or screen rooms. This is done by applying a surface coating to joints, seams, and contacting surfaces. The material is sufficiently fluid so that it readily flows into cracks. Metal-to-metal contact is improved significantly by applying such coatings. The surface coating fills slight irregularities and makes such intimate contact with exposed metal that even corroded joints can be made into greatly improved interference seals. Spraying of the coating can drive the silver particles into crevices and other hard-to-reach places. For dilution, or as a cleanup solvent, methyl ethyl ketone (MEK) may be used. In a typical shielded box or enclosure of complete metal construction, application of this type of coating to all contact surfaces improves insertion loss by a value of about 30 dB over the frequency range from 15 kHz to 10,000 MHz. Improvement is greater at the high-frequency end of this frequency spread. Thin coating material will not fill gross voids at joints or seams; a caulking compound must first be used in such applications. When these coatings are to be used on a metal surface; grease, oil, wax, paint, dirt, and other nonconductive films first must be removed with a solvent or cleansing agent, or by grinding, buffing, or machining the surface to be coated. When electrical contact is established with the metallic base, the coating is applied to the prepared surface by brush or spray. The coating should air-dry to less than 1 ohm per in.[2] surface resistivity in 1 h. Additional coats can be applied as desired. Some of the conductive surface coating and caulking applications are illustrated on Fig. 4-132 and summarized in Table 4-28. The conductive caulking compounds have the consistency of putty and can be applied by hand or with an air-activated gun. The thickness of deposition is controlled easily. Some of the compositions do not harden or set and therefore do not permit joint break but do allow flexibility; others cure to a hard resin surface and, still others, to a rubbery consistency. Elevated temperature cures can be used and are recommended. The joint should be maintained under pressure. Excess compound, squeezed from the joint, can be recovered and reused. Conductive caulking compounds have some adhesivity, but should not be relied on to hold a joint together. Most of these compounds are not cements; however, there are available conductive epoxys with great holding power.

4-7 GROUNDING AND BONDING

The principles underlying grounding methods were discussed in par. 3-3.4. The techniques for implementing those principles are discussed in this paragraph, which covers certain wiring practices, methods of obtaining a satisfactory "earth ground", and bonding.

4-7.1 GROUNDING

4-7.1.1 Grounding Connections

A ground point is the physical location where a circuit, piece of equipment, or system is connected to

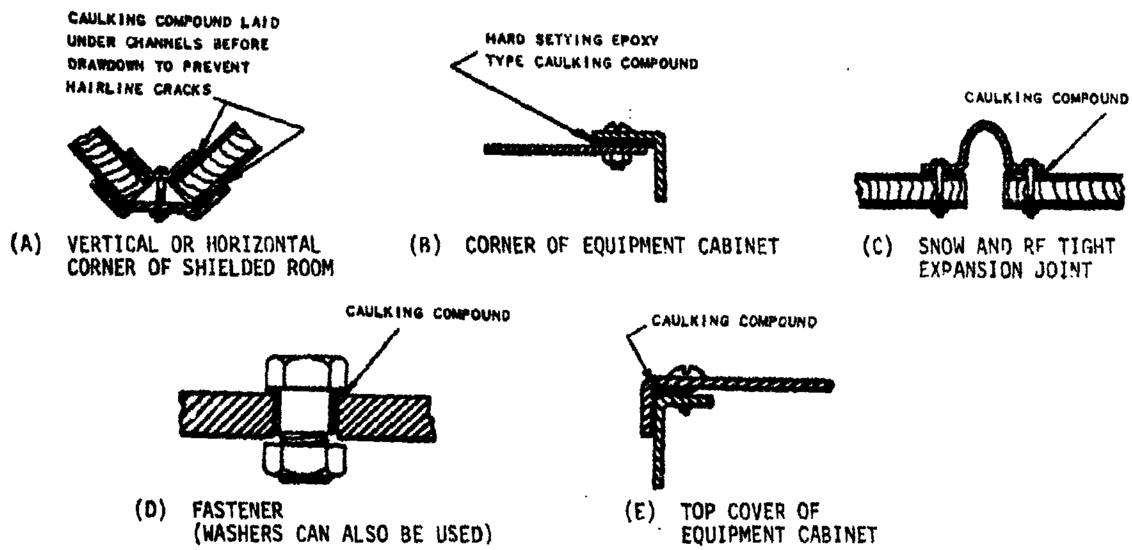


Figure 4-132. Typical Conductive Caulking Applications

TABLE 4-28. CONDUCTIVE SURFACE APPLICATIONS

APPLICATION	PREVIOUS METHOD	PRESENT METHOD	ADVANTAGES
Seams and static joints in rooms, containers, and enclosures	Welding or soldering	Caulking compounds	Ease of application; reliable performance; no damage to metal surfaces, flex resistance; no hairline cracks
	Knurled surfaces on clamped joints; metal strips for overlapping seams	Caulking compounds	Bolt spacing increased; no hairline cracks; lighter weight joints; accommodates expansion, contraction, warping, and vibration; weathertight seam
Sealing of doors, access hatches, container covers, and flanges	Woven wire mesh strip; copper finger stock	Gaskets Strip gaskets	RF and pressure seal with a single material; low-sealing pressures; small-gap spacings; good compression set; high attenuation
	Oriented wire mesh embedded in rubber strip	Strip gaskets	Reusable; reliable performance; high attenuation
	Machined metal seal with molded O-ring, knurled contact surface; wire screen filled with rubber	Seal gaskets	No scarring of flange faces; lower cost; reusable; low-sealing pressures
Sealing of fasteners, bolts, nails, screw heads	Welding or soldering	Seal caulking compounds Seal gaskets	Ease of application; reliable performance; no damage to to metal surfaces; easily disassembled

From P. J. Higgs, "An Investigation of Earthing Resistance", IEE Journal, Vol. 68, p. 736, Feb. 1930.

the ground plane. The impedance of a ground connection is a function of the conductor length, size, the shape of its cross section, and wiring techniques. If the ground connection is made improperly, it may be inadequate for satisfactory operation of the circuit and may, in fact, be more detrimental to the control of interference than if there were no ground connection. Grounding the RF portion of the spectrum is difficult and complex. Its complexity varies in direct proportion to the operating frequency. Several factors contribute to this:

a. Every wire has a definite inductance.

b. As radio frequencies increase, inductive reactance causes circuit impedances to increase.

c. The resonant frequencies of even small inductances acting in conjunction with circuit capacitance often fall within the operating frequency of the circuit.

d. As radio frequencies increase, skin effect becomes an important consideration.

A low-impedance ground connection requires the ground leads to be as large in cross section and short as possible, and to be securely bonded directly to the ground plane. A representative ground lug connection and its equivalent circuit is shown on Fig. 4-133. It shows how two separate circuits connected to the same ground point (Fig. 4-133(A)) can be inductively coupled together (Fig. 4-133(B)). Fig. 4-134(A) illustrates a typical, albeit *improper*, method of connecting power and signal grounds at a connector. As the frequency increases, the inductance of the ground jumper can become appreciable and, if the power or signal circuit contains high-frequency interference currents, they may be conducted through the ground pin into the external wiring. In contrast, Fig. 4-134(B) shows the proper method of installing a ground to avoid conducting the interference through the connector.

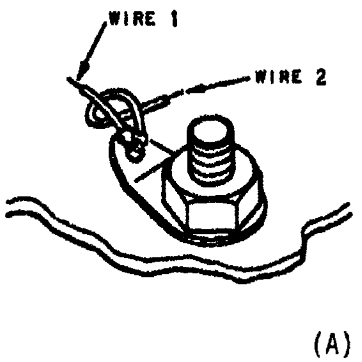
4-7.1.2 Chassis Grounds

4-7.1.2.1 Distribution of Chassis Potential

A typical plot of the chassis-potential of a ground plane carrying relatively large currents is shown on Fig. 4-135. On this figure, the dark areas surround ground lugs or shields in high-voltage and/or high-current areas where the power is adequate to present a signal on the physical ground plane. A potential plot, if available, aids in determining the location of low potential or equipotential points that can be used in grounding small-signal-sensitive circuits such as grid bias resistors. A matter of an inch in the location of ground points can make a difference of several millivolts in the potential. It may be advantageous at times to run a longer ground wire to find a point of lower potential on a ground plane.

4-7.1.2.2 Circuit Considerations

Each electronic circuit contributes its own ground currents. Any ground return path that goes around corners or crosses other return paths may cause intercircuit or interstage coupling (Fig. 4-136). The magnitude of this coupling is dependent upon impedance



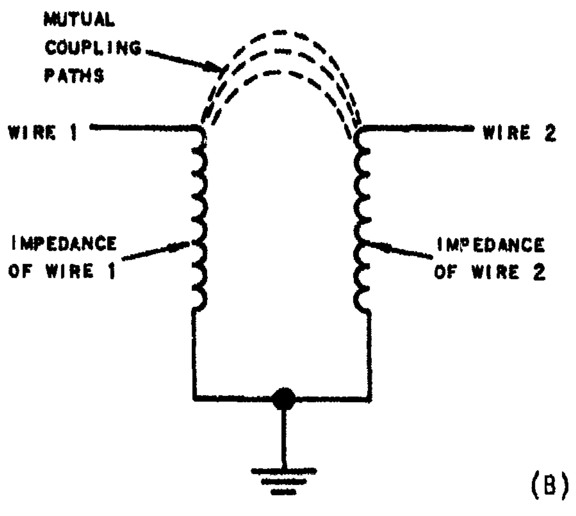
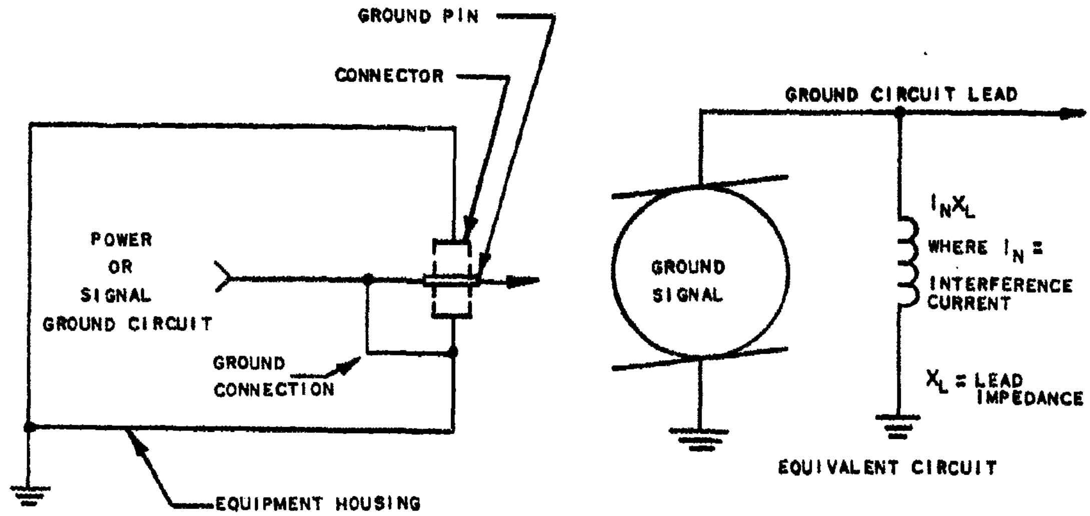
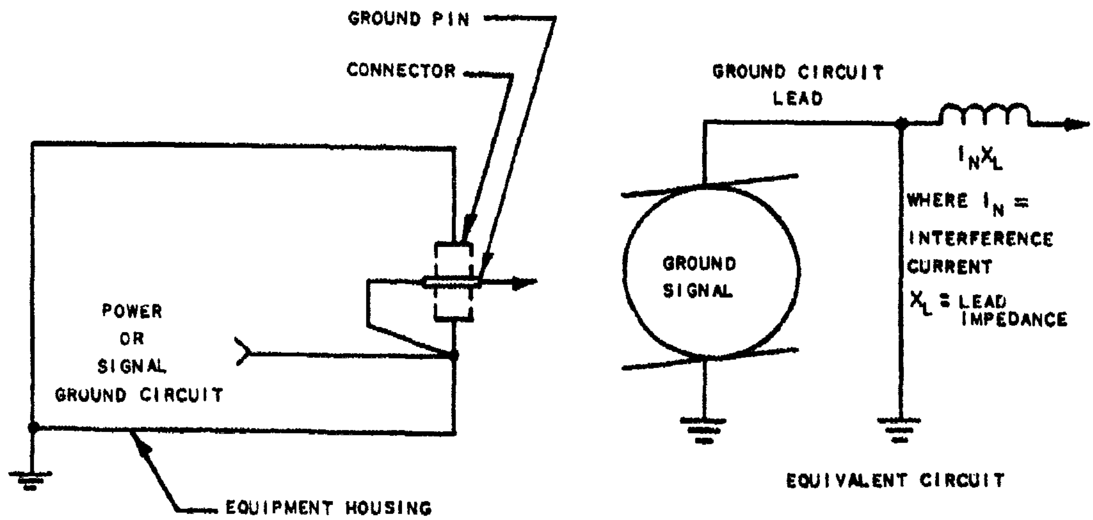


Figure 4-133. Ground Lug Connection and Equivalent Circuit



(A) METHOD COMMONLY EMPLOYED IN MAKING GROUND CONNECTIONS IN ELECTRICAL AND ELECTRONIC EQUIPMENT (*NOT* RECOMMENDED)



(B) PROPER METHOD OF GROUNDING FOR MAXIMUM INTERFERENCE REDUCTION

Figure 4-134. Grounding Methods

between the circuit ground points, and the current in the ground plane (Fig. 4-137). The current I of a low-impedance circuit produces a potential IZ_{gp} in a high-impedance circuit. Circuit components should be arranged so that ground return paths are short and direct and have the fewest possible crossings.

The effect of ground potential can be nullified by electrically isolating circuits with isolation transformers as shown on Fig. 4-138. This method is especially effective at audio and low radio frequencies.

4-7.1.2.3 Shield Grounds

Grounds for apparatus housed within a shield should be arranged so that the return conductor also is housed within the shield and the shield itself is not used as a return conductor. In this way, the current

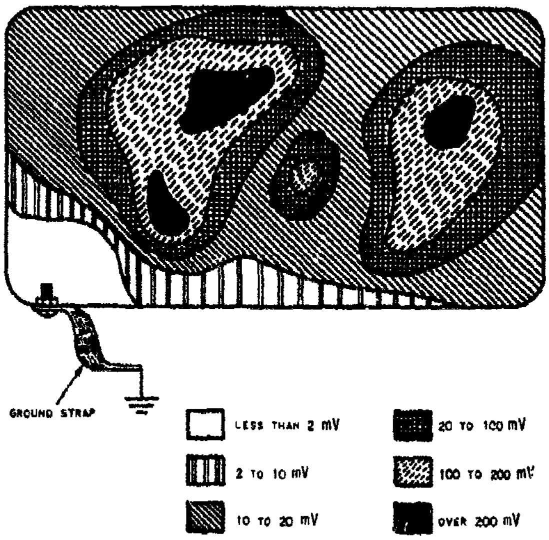


Figure 4-135. Typical Ground Potential Graph at a Specific Frequency

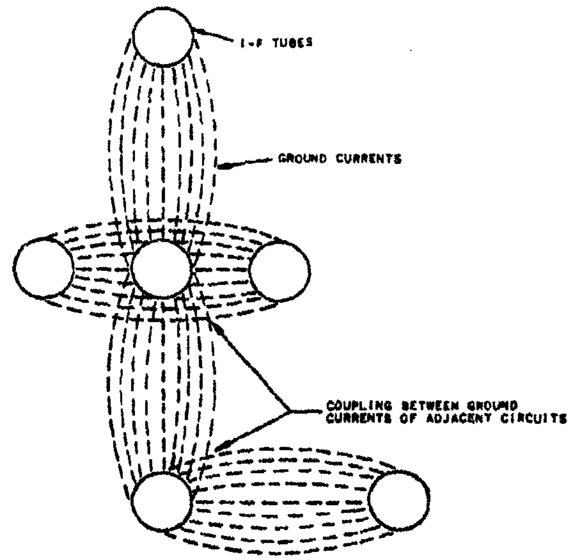


Figure 4-136. Example of Intercircuit Coupling

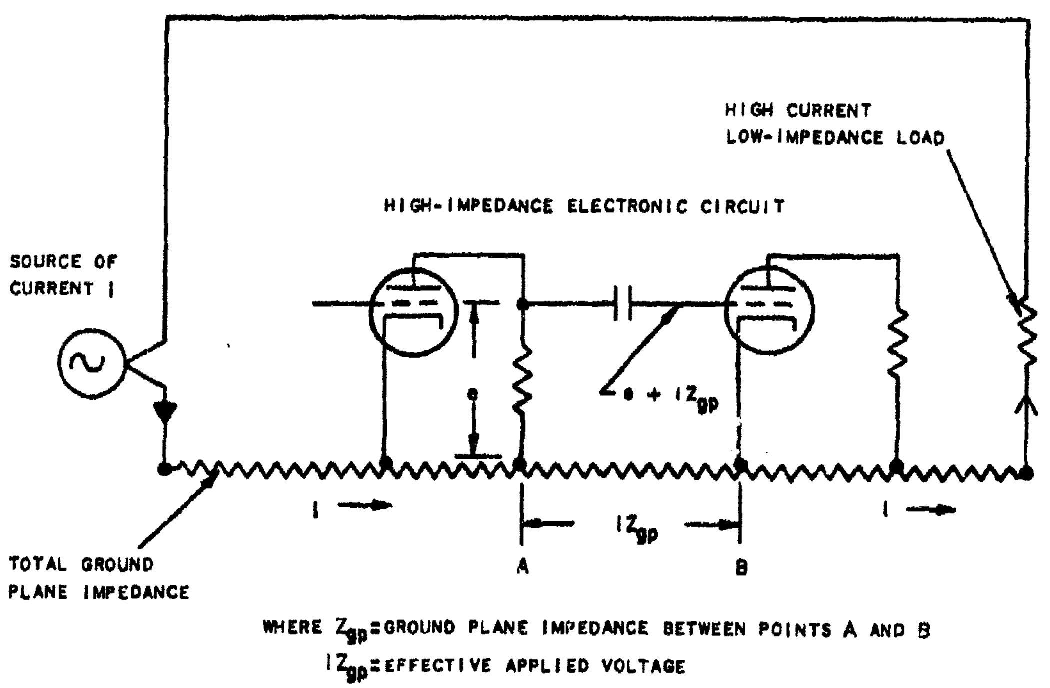


Figure 4-137. Effect of Circulating Currents on a Typical Current

that flows through the shield is reduced to a very small value, and the tendency for energy to leak out through holes or joints is minimized. The ideal grounding system makes use of a ground bus or ground plate within the shield which is insulated from the shield except at a single point (ground point) as illustrated on Fig. 4-139.

4-7.1.2.4 Printed Circuit Boards

For maximum shielding and isolation, shields on printed-circuit boards should be grounded directly to the main chassis, independent of any grounds located on the printed-circuit board. In the case of a shield for an RF transformer located on a printed-circuit board, grounding the shield to the printed-circuit board permits the conduction of stray currents into the printed-board circuits, thus causing secondary interference problems. The correct way of grounding this type of shield is directly to the main chassis, as illustrated on Fig. 4-140.

4-7.1.3 Cable Grounding

Electrical compatibility in a complex electrical or electronic system is dependent in many cases on the treatment of the shielding and the grounding of the shields of interconnecting leads. Injudicious application of a grounded shield to a wire may cause coupling problems that otherwise would not exist. Grounding of the shields may be accomplished with single-point or multipoint connections as discussed in par. 3-3.4.

a. *Single-point shield grounding.* For multilead systems, each shield may be grounded at a different physical point provided individual shields are insulated from each other. Single-point grounding is more effective than multipoint shield grounding only

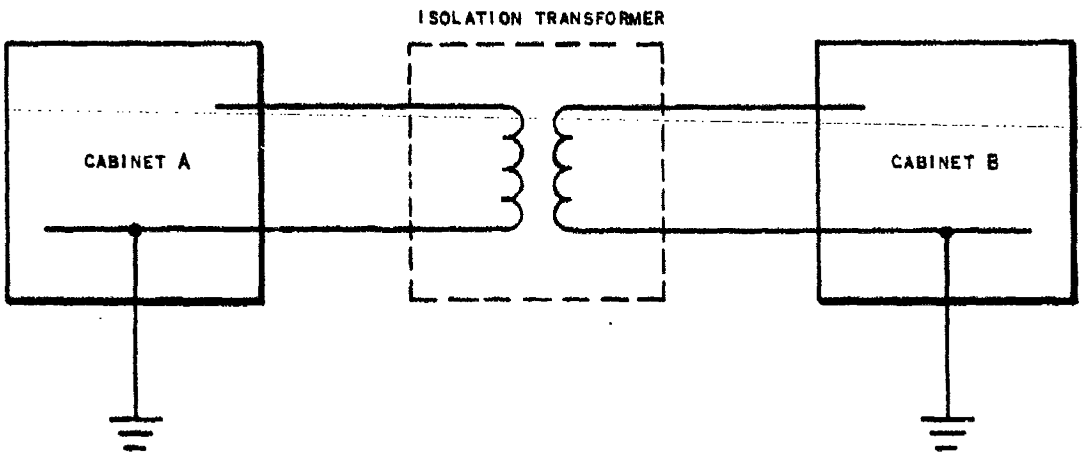


Figure 4-138. Isolation Transformer Technique for Minimizing Ground Potential

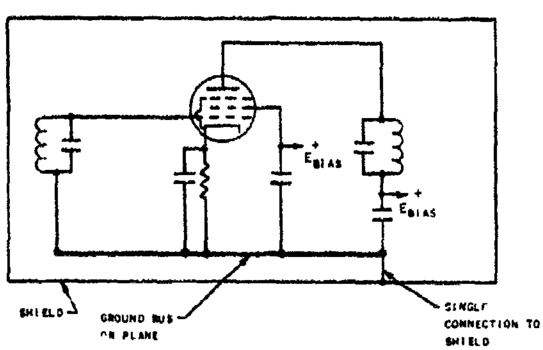


Figure 4-139. Single-point Ground Bus Arrangement

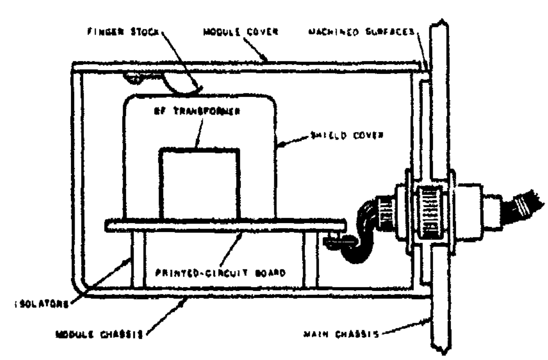


Figure 4-140. Direct Use of Chassis for Good Ground

for short shield lengths. Single-point grounding is ineffective in reducing magnetic or electrostatic coupling when conductor-length-to-wavelength (L/λ) ratios are greater than 0.15 where the wavelength is that of the highest frequency to be used (or the highest frequency interference to be expected) on the wire or in the system.

b. *Multipoint shield grounding.* For L/λ ratios greater than 0.15, multipoint grounding at intervals of 0.15λ is recommended, since the shield can act as an antenna at frequencies for which L is a multiple of $\lambda/4$. When grounding the shield at intervals of 0.15λ is impracticable, shields should be grounded at each end. Multipoint shield grounding is effective in reducing all types of electrostatic coupling, but is subject to failure if large ground currents exist. In general, multipoint shield grounding is effective at frequencies greater than about 100 kHz, and single-point shield grounding is effective at frequencies below 100 kHz because of the ground currents. Where a circuit can respond at both low frequencies where single-point grounding should be used, and at high frequencies where multipoint grounding should be used, the direct ground can be used at one point and a capacitor used to connect the other points to ground. The size of the capacitor depends on the frequencies involved.

4-7.1.4 Static Grounds

Metallic parts not otherwise returned to "ground", which may pick up static charges or are of sufficient dimension to receive and re-radiate energy from local radio transmitters, should be connected to the ground plane (see par. 3-3.4.2).

4-7.1.5 Power Supplies

4-7.1.5.1 General

The power ground and signal ground should be isolated from each other throughout the chassis to minimize the possible coupling of singals from any one type of ground line to any other type of ground line. The application of the following techniques often can avoid potential problems:

a. Incorporate, where possible, individual ground paths for ac voltages, dc voltages, and signals.

b. Connect a ground path to the largest conductor (lowest impedance) by as direct a route as possible.

c. Use several arterial ground paths to the supply common point, as opposed to one super ground bus.

d. Avoid multiended ground buses or lateral ground loops.

e. Have as few series connections (solder joints, connectors) as possible in a ground bus, and make sure that they are good, solid electrical connections.

4-7.1.5.2 Separation of AC Neutral from Frame Ground

A prime source of interfering ground currents is from ac neutral distribution through parallel frame ground paths (common-made currents). They can be avoided by isolating all ac neutral sources from frame ground. All ac power sources within the immediate equipment area should be floated and referenced to frame ground at one point only. Power sources are defined as secondary ac transmission to any part of the equipment area to include lighting, heating, air conditioning, utility outlets, communication equipment, etc. Floating a circuit means using closed circuit transmission with a single point as a ground reference with no dependence on earth ground or frame ground to complete any part of the neutral return path. All ac power circuits must be floated to one location. No ac neutral should be connected to frame or earth ground at any other point. Ac neutral bus bars should not be connected to the frame ground bus bars in any major item or equipment area except in one location. All power (input or interconnecting) cables (portable or transportable) should be provided with a green insulated conductor intended only for grounding noncurrent-carrying metal parts of equipment. This equipment ground wire (EGW) should be terminated at both ends in the same manner as the other conductors. The EGW should not be connected to the power return circuit (neutral), and the neutral wires should not be used as EGW's. Wire color coding is recommended. All neutral power conductors should be white; all EGW should be green. The following are examples:

a. Single phase two-wire ac service:
 (1) black (hot)
 (2) white (neutral)
 (3) green (ground)

b. Floating single phase two-wire ac service:
 (1) grey (hot)
 (2) grey (hot)
 (3) green (ground)

NOTE: Ground wire is essential; 3-wire cable required.

c. Three phase ac circuits:
(1) black (hot)-L_1 (Phase A)
(2) red (hot)-L_2 (Phase B)
(3) blue (hot)-L_3 (Phase C)
(4) white (neutral)
(5) green (ground)

NOTE: Ground wire is essential. Five-wire cable for WYE connection and 4-wire cable for DELTA connection are required.

d. Single phase three-wire ac service (such as 220 V):
   (1) black (hot)
   (2) red (hot)
   (3) white (neutral)
   (4) green (ground).

NOTE: Ground wire is essential; 4-wire cable required.

4-7.1.5.3 Marine Craft Bonding and Grounding Methods

MIL-STD-1310, *Shipboard Bonding and Grounding Methods for Electromagnetic Compatibility*, outlines shipboard construction and equipment installation requirements, shipboard bonding methods, and the practices necessary to minimize the electromagnetic interference (EMI) environment aboard marine craft. These requirements should be adhered to when communication, radar, electronic, and electromechanical equipments are installed in Army marine craft.

4-7.1.5.4 Ground Studs

A ground stud usally is required on individual equipments and provides the electrical ground connection to the chassis or frame. It is secured mechanically to insure low resistance joints by: (1) soldering to a spot-welded terminal lug or to a portion of the chassis or frame that has been formed into a soldering lug; or (2) by use of a terminal on the ground wire which then is secured by a screw, nut, or lock washer. The ground stud should be of a size to allow electrical connection of size AWG-10 wire. All hardware used for grounding or other electrical connections should be made from copper or copper alloys. Terminal lugs should be tin-plated or hot tin dipped. Paint, varnish, lacquer, etc., should be removed from the vicinity of the fastening point to insure metallic contact of the two surfaces. Corrosion protection should be provided for all ground connections. Internal or external lock washers should not be used on any grounding or other screw-type electrical connections. Washers should not be located between the metal plate and terminal lug or other part being grounded, so that they will not interfere with the full and direct contact between these two members. Neither locking terminal lugs nor self-locking nuts should be used for grounding. Flat washers should be inserted next to any part having insufficient contact area with its adjacent part.

4-7.1.6 Earth Ground

4-7.1.6.1 General

The earth can be used as a means of dissipating excessive charges caused by man-made and natural interference that may injure personnel who are operating equipment, and damage the equipment involved. Connections to the earth, called grounds or ground connections, are made to pipes of buried water systems, driven rods, buried metal plates, or buried wire. When buried pipe systems are not available, driven rods are considered the most satisfactory substitutes. Low resistance grounds are essential. Ground resistance is affected by the ground rod resistance, lead connection, contact resistance between ground rod and soil, and by the type of soil. Type of soil will have the greatest affect on ground resistance. Numerous kinds of soil prevent a simple classification. Study of certain types of soil reveals a definite trend in resistivity. Soil can be roughly classified into one of the types given in Table 4-29.

TABLE 4-29. SOIL CLASSIFICATION CHART

Type of Soil	Resistance in ohms of Ground Connection (One 5-ft rod, 5/8 in. diameter)		
	Average	Minimum	Maximum
Fills, ashes, cinders, brine waste	14	3.5	41
Clay, shale, gumbo, loam	24	2	90
Same, with varying proportions of sand and gravel	93	6	800
Gravel, sand, stones with little clay or loam	554	35	2,700

Temperature is another variable that affects ground resistance. There will be only a slight change in ground resistance for temperatures above 32°F. Fig. 4-141 shows how resistance increases with decrease in soil temperature.

Soils of the same type can differ greatly in resistance due to variations in moisture content. There is a wide variation in the moisture content of soils, from 10% during dry seasons to about 35% during wet seasons. The approximate average is from 16 to 18%. The effect of moisture content on resistance of soil can be seen from Table 4-30.

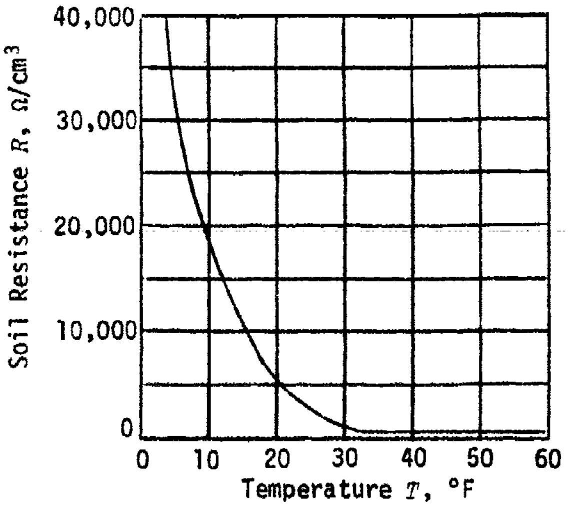


Figure 4-141. Variation of Soil Resistance With Temperature

TABLE 4-30. EFFECT OF MOISTURE CONTENT ON SOIL RESISTANCE

Moisture Content, % by weight	Resistivity, Ω/cm³	
	Top Soil	Sandy Loam
0	over 1 billion	over 1 billion
2.5	250,000	150,000
5	165,000	43,000
10	53,000	18,000
15	19,000	10,500
20	12,000	6,300
30	6,400	4,200

4-7.1.6.2 Grounding in Subzero Zones

Two methods of grounding are presented:

a. Method A. A good earth ground is difficult to achieve beneath snow and ice; however, the following method usually will produce satisfactory results. Place a grounding rod horizontally in a narrow trench in the snow as deep as possible. Extend lines of No. 6 gage or heavier copper wire radially from the rod for about 10 ft. Use chemical aids along the rod and cables, and bury the entire system in the snow. For general effectiveness and anticorrosion qualities, the main usable chemical aids rank as follows:

(1) Magnesium sulphate
(2) Copper sulphate
(3) Calcium chloride
(4) Sodium chloride
(5) Potassium nitrate.

b. Method B. Ground rods driven vertically in soil normally have a relatively high resistance at depths of less than 5 ft. A shallow circular trench about 18 in. in internal diameter around the ground rod should be dug about 1 ft deep and lined with the chemical aids. The area around the ground rod should be kept moist (Ref. 31).

4-7.2 BONDING

4-7.2.1 General

A bond is an electrical union between two metallic structures to provide a low impedance path between them. Bonding is a process used to achieve a ground connection or the establishment of a ground plane, as discussed in pars. 3-3.4 and 4-1.4. The structures involved may be housings, subassemblies, or components such as the chassis of an electronic assembly or the frame of an electrical machine.

4-7.2.2 Types of Bonds

It should be noted that the term "bond" refers to both the mechanical interface between the joined conductors, as well as the conductor bonding jumper or strap used to interconnect two separate structures. Where the separate structures are placed in immediate contact, the bond is "direct"; otherwise, it is indirect.

4-7.2.2.1 Direct Bonds

Direct bonds include permanent metal-to-metal joints formed of machined metal surfaces, or with conductive gaskets held together by lock-threaded devices, riveted joints, tie rods, or pinned fittings driven tight and not subject to wear or vibration. The best bonded joint is formed by welding, brazing, or sweating. Soldering is not a good method of direct bonding because soldered joints have appreciable contact resistance. Good metal-to-metal contact must be maintained for the life of the joint, and precautions should be taken to seal the joint against moisture that would cause galvanic corrosion. Conductive epoxy adhesives are used in certain applications. Dissimilar metals in direct contact should be avoided. Screw threads never are considered adequate bonding surfaces. In particular, sheet-metal type screws are inadequate for use in bonding. If two structural members are held together by screws, the impedance between them is usually comparatively high unless good direct contact is maintained. MIL-STD-1310 and MIL-B-5087 give details on methods of preparing bonding joints.

4-7.2.2.2 Indirect Bonds

When a direct bond is not practical, an indirect bond is necessary. A good indirect bond is one that presents a low impedance throughout the interference spectrum, and retains its usefulness for an extended period of time. An indirect bond is usually a bond strap or jumper. Bond jumpers are short, round, braid conductors for application where the interference frequency is below a few megahertz. They generally are used in low-frequency devices, and where the development of static charges must be prevented. Bond-straps are either solid, flat, metallic conductors, or woven braid configurations where many conductors are effectively in parallel. Solid-metal straps are preferred generally for the majority of applications. The most significant feature of a bond-strap is its resiliency which is determined by its material and thickness. Beryllium copper and phosphor bronze are used often. Under conditions of severe vibration, a corrugated strap often proves useful in preventing excessive damping and in achieving maximum service life. Braided or stranded bond-straps generally are not recommended because of several undesirable characteristics. Oxides may form on each strand of nonprotected wire and cause corrosion. Because such corrosion is not uniform, the cross-sectional area of each strand of wire will vary throughout its length. The nonuniform cross-sectional areas (and possible broken strands of wire) may lead to generation of interference signals within the cable or strap. Broken strands may act as efficient antennas at high frequencies, and interference may be generated by intermittent contact between strands. They also have higher self-inductance than solid straps.

4-7.2.3 Bonding Impedance

The equivalent circuit of a bond which takes into account its series resistance R and the distributed capacitance C between the two structures being bonded together is shown in Fig. 4-142 along with a typical plot of the magnitude of its impedance as a function of frequency. It is seen that resonance occurs at some frequency which may be of the order of tens of megahertz, but depends upon the physical characteristics of the bond. At frequencies above this first resonance, the behavior is not easy to describe, but the equivalent circuit may resemble that of an open-ended transmission line, and multiple resonances may occur. Also, it should be noted that when resonance occurs, the bond is capable of radiating an appreciable amount of energy.

The sharpness of the resonant peak in Fig. 4-142 is dependent upon the resistance of the bonding connection. The effective resistance of any conductor is a function of frequency due to skin effect. Fig. 4-143 shows the skin depth in copper and the variation of resistance and inductive reactance (at frequencies where it exceeds the resistance) with frequency for several conductors of circular cross section, and 10 cm (4 in.) length. At 1 MHz the resistive impedance is substantially less than the inductive reactance for all wire sizes. By using a conductor in the form of a strip with rectangular cross section, both the inductive reactance and the resistance can be decreased significantly.

The inductance L of a bond strap (Ref. 24) is

$$L = 0.00508\, l\left[2.303 \log\left(\frac{2l}{b+c}\right)\right] + 0.5 + 0.2235\left(\frac{b+c}{l}\right),\ \mu\text{H} \quad (4\text{-}52)$$

where

l = length, in.
b = width, in.
c = thickness, in.

The corresponding inductive reactance is shown on Fig. 4-144. Reductions of about an order of magnitude are possible compared with wires of equal cross section. The relatively high impedance at high frequencies illustrates that the bond strap is not a substitute for a direct bond. A rule of thumb for achieving minimum bond strap inductance is that the length-to-width ratio of the strap should be 5:1 or less. This ratio determines the inductance — the major factor in the high-frequency impedance of the strap.

The bond strap or jumper may be held mechanically by means of bolts, rivets, welding, brazing, or sweating. Tooth type lockwashers are used with bolt fasteners to insure no deterioration of metal-to-metal contact of bond strap connections. Fig. 4-145 shows a typical bond strap bolted into position.

4-7.2.4 Bond Measurements

It is conventional to check the quality of a bond by measuring its dc or low-frequency resistance. Such a test is satisfactory in production where it is desired to determine the quality of the mechanical work performed. Good bonds should have a resistance between 0.1 and 2.5 mΩ, depending on the magnitude of the ground currents expected and the sensitivity of the equipment. To check the quality of a completed bond at high frequencies, it is necessary to use an impedance bridge. Particular care should be taken to identify the frequency at which the first resonance appears. In hazardous areas, under conditions of fault current, the bond resistance must be below a specified value for safety reasons. Fig. 4-146 shows the maximum permitted resistance according to the National Electrical Code (Ref. 25).

4-7.2.5 Bond Design

4-7.2.5.1 Physical Requirements

The mechanical stresses to which bonds are subjected should not be of such a magnitude as to cause damage. However, bonds should remain accessible for maintenance and inspection and not interfere with the intended operation of the equipment either by reducing accessibility or by limiting its movement. When bonding jumpers are used, their lengths are critical and must be as direct and as short as possible (see par. 4-7.2.3). A high area of contact requires that the surfaces to be joined be flat, with no surface irregularities.

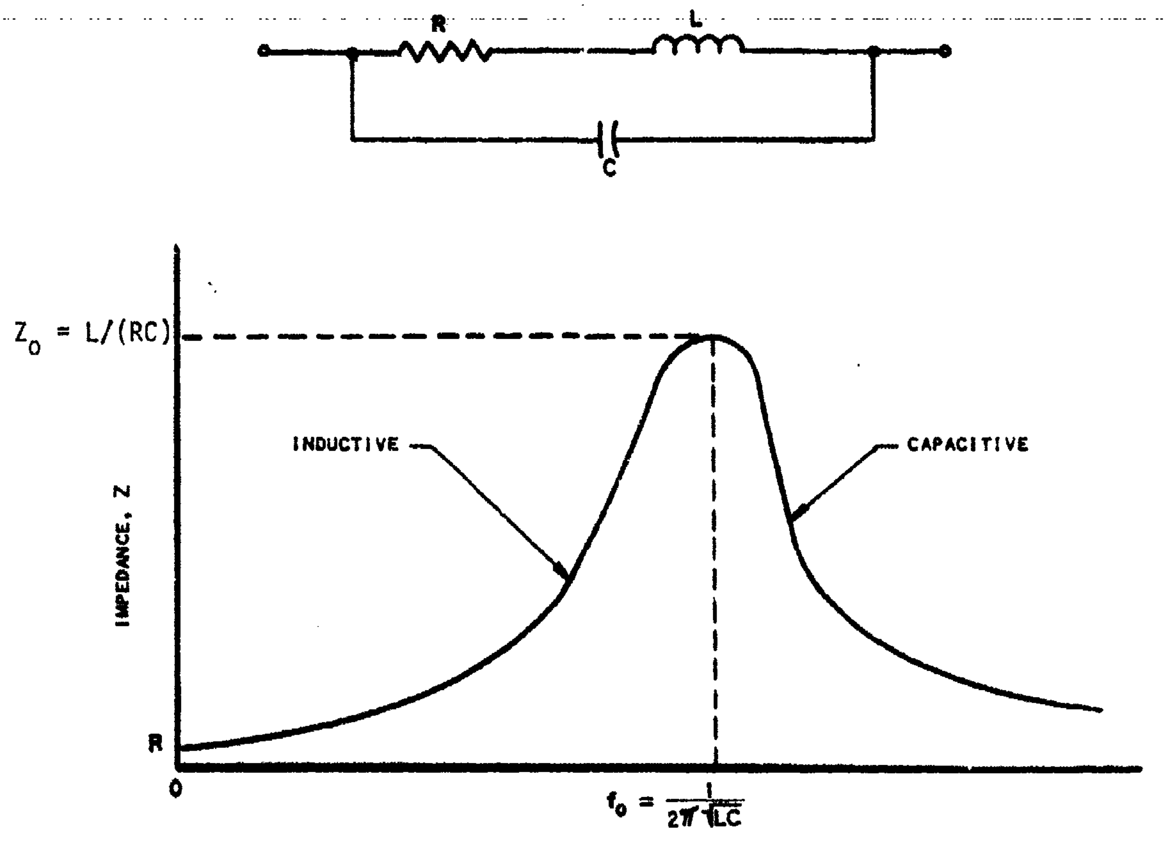


Figure 4-142. Bonding Strap Impedance Characteristics

4-7.2.5.2 Choice of Materials

The choice of material for a given bonding application usually is dictated by consideration of the metals being bonded and the environment within which the bond must function. When joining dissimilar metals, corrosion becomes an important consideration. Factors contributing to corrosion are the proximity of metals in the electromotive series and the amount of moisture present. Corrosion is attributed to two basic electrochemical processes: galvanic and electrolytic corrosion.

Several methods can be employed for minimizing or preventing corrosion and its adverse effects on bonding. One is to use metals low on the activity table, such as copper, lead, or tin (Table 4-31). Where members of the electrolytic couple are widely separated on the activity table, it is sometimes practical to use a plating such as cadmium or zinc, which helps to reconcile the dissimilarity (Refs. 26 and 27). Thin, bimetallic plates, formed by mechanical bonding of dissimilar metals, cold-flowed together under high pressures, have been used to interconnect two structural units of dissimilar metals. Where bimetallic plates are to be used, the junctures of the two metals normally are covered with a protective coating, such as grease or polysulphate, to exclude moisture and retard corrosion. This coating reduces the area of metal exposed to an electrolyte. If bonding is such that corrosion is likely to occur, the bond should be designed as a replaceable element such as a jumper, plate, separator, or washer.

Acceptable contact surface materials that may be used to fasten bonding jumpers to structures are indicated in Table 4-32. The arrangement of the metals listed in this table is in the order of their decreasing galvanic activity when exposed to an electrolyte. The screws, nuts, and washers to be used in making the connections are indicated as Type I, cadmium or zinc plated, or aluminum; and Type II, passivated stainless steel. Where neither type of securing hardware is indicated, Type II is preferred from a corrosion standpoint.

The possibility of galvanic and/or electrolytic action necessitates extreme care in assembling joints that serve as bonds. Surfaces should be absolutely dry

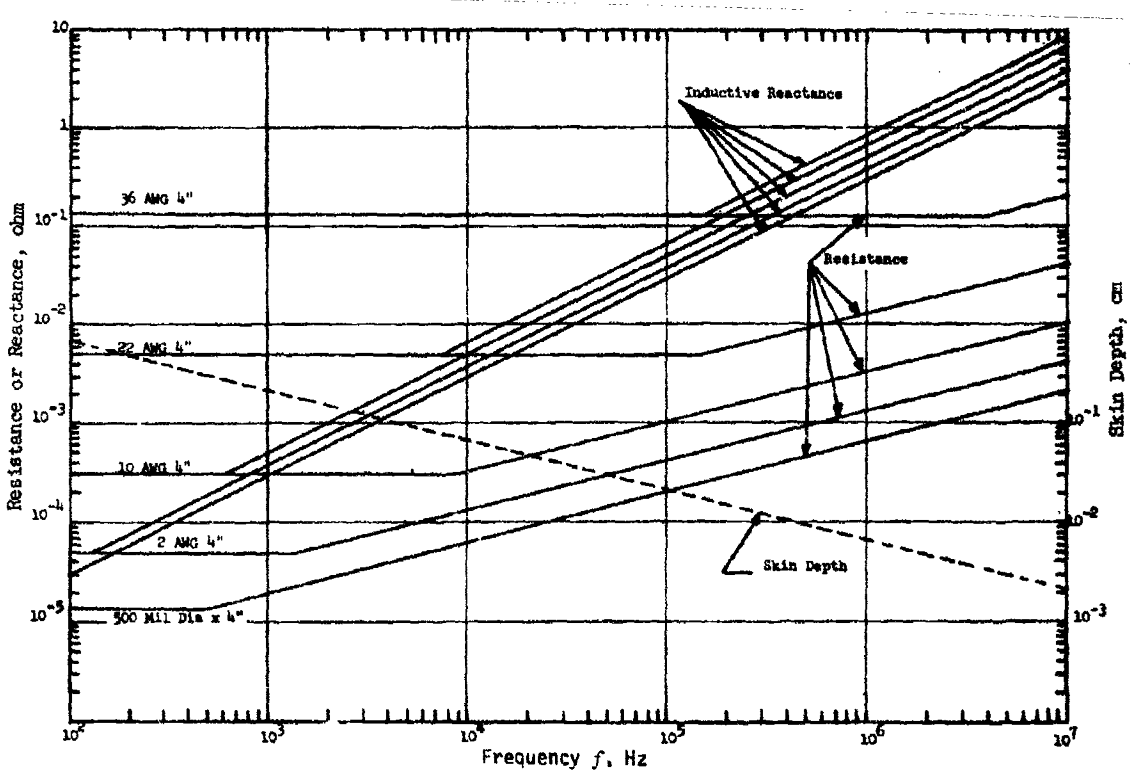


Figure 4-143. Skin Depth in Copper, and Resistance and Reactance Variation With Frequency for Conductor of Circular Cross Section

before mating, and should be held together under high pressure to minimize the possibility of moisture entering. The use of number 7/0 garnet finishing paper or equivalent is recommended to remove paints, anodic films, and oxides from surfaces. Care must be taken not to remove excessive metal under the protective finish. Abrasives, such as emery cloth or sandpaper, cause corrosive action because their particles embed themselves in the metal; therefore, they should not be used. The contact area should be brushed clean; it should be about 1.5 times greater than the area necessary for actual mounting. After a joint (free of moisture) is assembled, the periphery of the exposed edge should be sealed in accordance with the applicable requirements of MIL-F-14072.

The finishes which are currently available fall into two categories: nonconductors and conductors. The nonconductors — such as anodized films, Dow 8 and HAE — are excellent for corrosion protection and wearability but, being nonconductors, cannot be

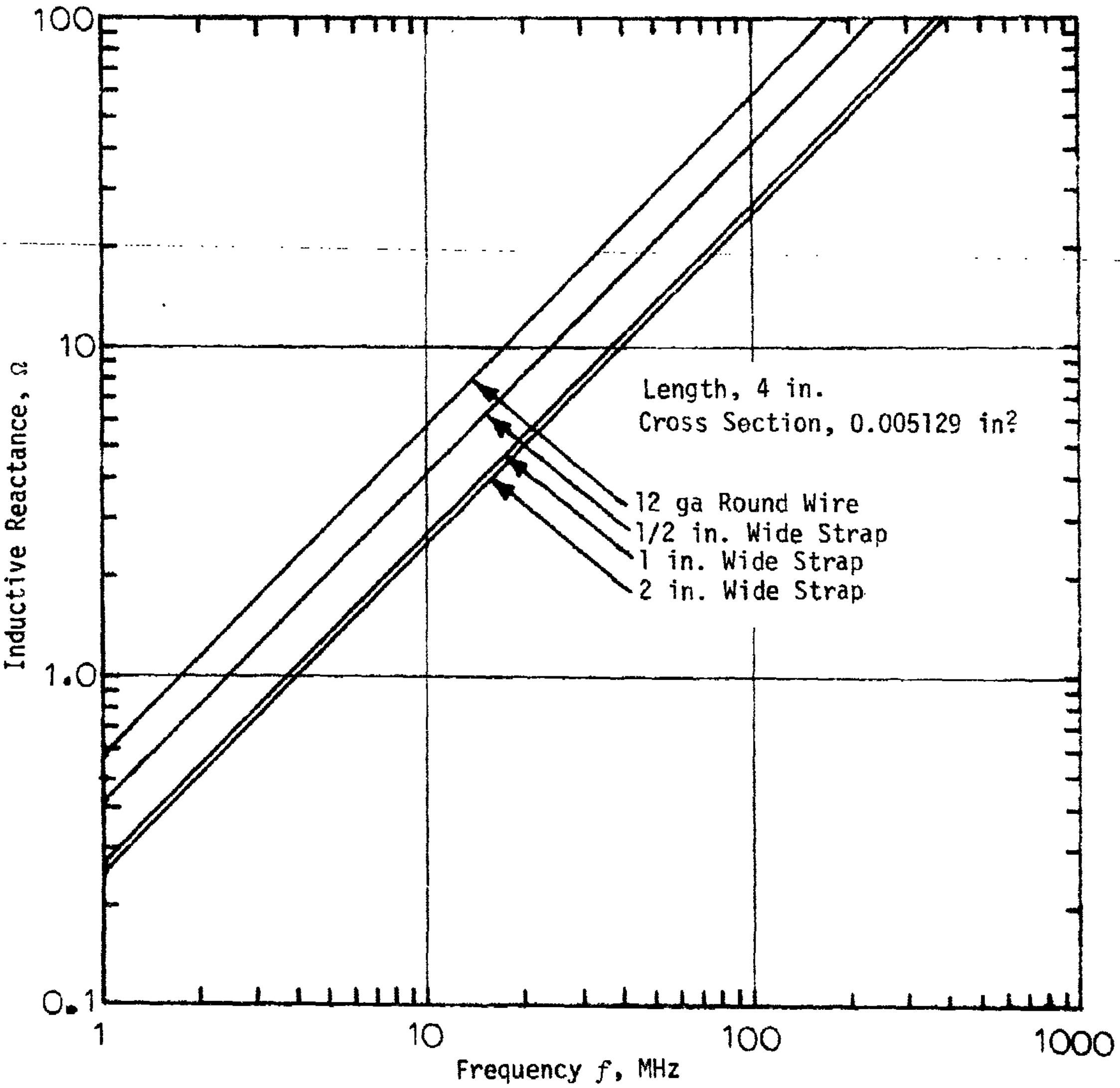


Figure 4-144. Inductive Reactances of Bond Strap and a Bond Jumper Having Constant Cross-sectional Area

applied to surfaces used for bonding. The conducting films — such as Iridite 14-2, Dow 15, Alodine, and Oakite 36 — offer fair corrosion protection (much better than bare metal), but do not afford as much wear resistance as the anodized films. The bond and adjacent metal should be covered with protective finishes after the bond has been assembled and its resistance checked.

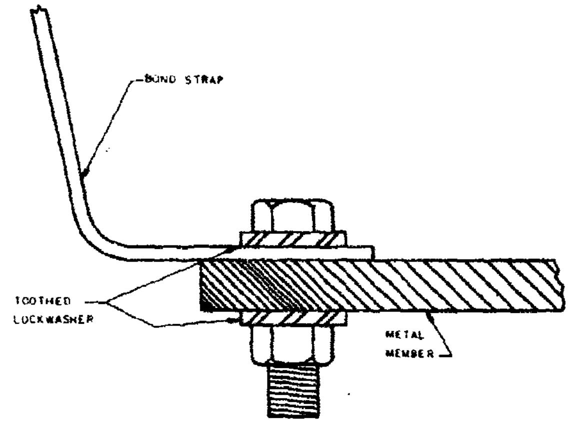


Figure 4-145. Recommended Bond Strap Bolting Installation

4-7.2.5.3 Conductive Adhesives

Direct conductive permanent bonds sometimes can be made between two surfaces with a high viscosity plastic cement. The user of such cements must decide upon the relative importance of such factors as high adhesive strength, low resistivity, cure cycle, ease of

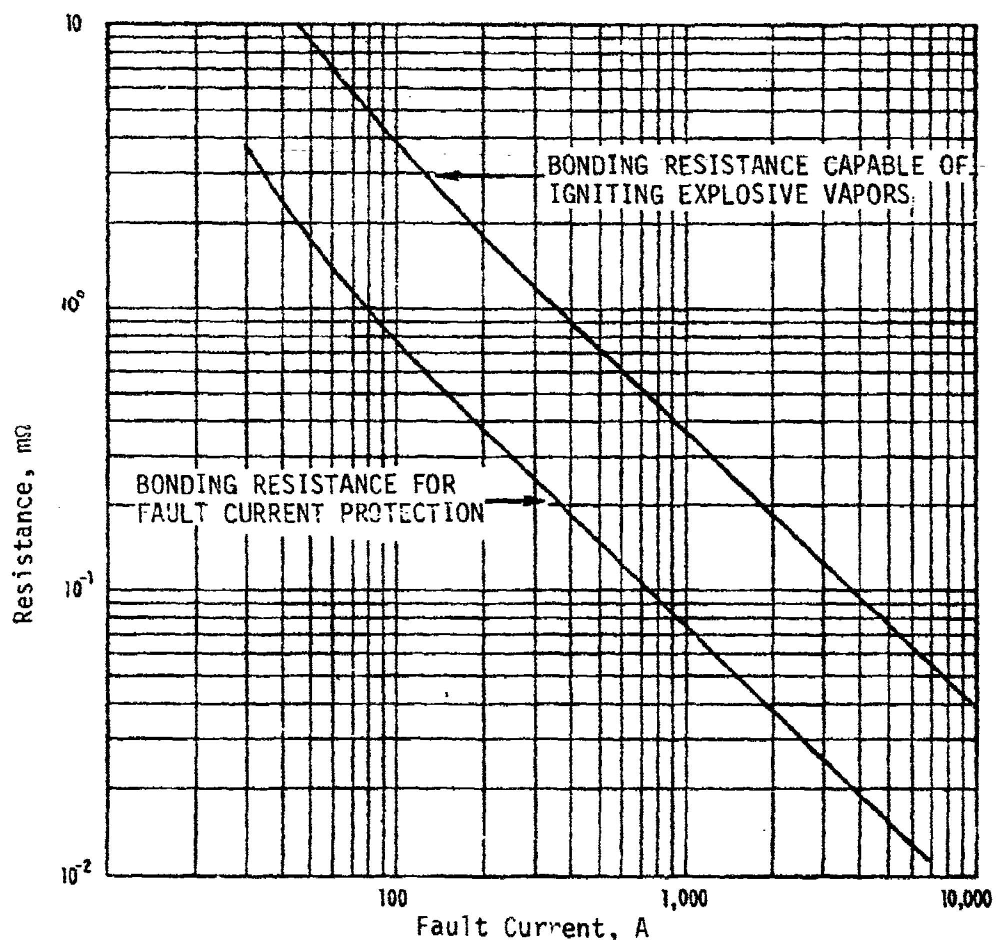


Figure 4-146. Fault Current vs Maximum Allowed Resistance for Bonding Between Equipment and Structure

TABLE 4-31
ELECTROMOTIVE FORCE SERIES OF COMMONLY USED METALS*

METAL	ELECTRODE POTENTIAL, V
Magnesium	−2.375
Beryllium	−1.700
Aluminum	−1.670
Zinc	−0.7628
Chromium	−0.740
Iron	−0.441
Cadmium	−0.402
Nickel	−0.230
Tin	−0.1406
Lead	−0.1263
Copper	+0.346
Silver	+0.7996
Platinum	+1.200
Gold	+1.420

*Select dissimilar metals so that if corrosion occurs, it will be in the replaceable components — such as grounding jumpers, washers, bolts or clamps — rather than structural members or equipment enclosures. When two different metals are in contact, the one higher in the electromotive-force series will be more affected by corrosion than the other. The smaller mass (generally the more easily replaceable) should therefore be made of the higher metal; for example, cadmium-plated washers are recommended for use with steel surfaces since cadmium is lower than steel.

handling, and cost. Adhesives may be used as alternatives (subject to specific requirements) to soldering, welding, and mechanical joining operations. Most of the commercially available materials are based on epoxy resins, either two-component low-temperature cure or one-component oven-cure systems.

The resistance of these two adhesives depends on the conductive materials with which they are loaded. The usual materials are: carbon, resistivity, $\rho = 10$ ohm·cm; silver and gold, $\rho = 5.7 \times 10^{-3}$ to 18×10^{-6} ohm·cm. For silver, both platelet and metal powder coated with silver are used.

The platelet form is ideal where a thin glue line is needed, whereas gritty powders make better contact with the metal surfaces to which they are applied and generally offer lower resistivity. At least 60 to 70% by weight of metallic content is needed to provide an optimum combination of adhesion and conductivity. As a rule, the greater the metal content, the less will be the adhesion; and the less the metal content, the greater the adhesion.

Bond shear strength values of 288 kg/cm² are being claimed by at least one conductive adhesive manufacturer. This is coupled with the simultaneous claim of 2×10^{-3} ohm·cm resistivity when fully cured. This particular epoxy is coarse in texture and not intended for delicate electronic wiring work but rather is used for gross metal-to-metal joints, heat conduction, and RF shielding. Finer work is obtainable with 225 kg/cm² adhesives having resistivities of 2×10^{-4} ohm·cm after curing.

Joint strength tests have indicated greatest strength in shear and least strength in peel when a length of copper wire was attached by adhesive to an etched pad area in a phenolic laminate. It was found that those conductive adhesives with the highest joint strengths also had the least conductivity.

4-7.2.5.4 Conductive Pastes

Several types of paste, caulking, and sealing compounds are available for such applications as pipe and conduit threads, shielded room or enclosure seams, removable cover plane seams, expansion joints, and fastener hardware caulking. Resistivity varies with:

a. Type of metal loading use in the paste
b. Configuration of the metallic content
c. Condition of the metal surfaces upon which the paste is applied
d. The pressure applied to the joint after application of the paste.

Manufacturers should be consulted for details of applications and limitations.

4-7.2.6 Bonding Applications

4-7.2.6.1 Shock Mounts

A frequent application for which indirect bonding is the only suitable type is that involving shock-mounted equipment. The designer should consider the degree of relative motion to be expected between two surfaces to be bonded, the characteristics of the materials involved, and the frequency range over which the bonding is expected to be effective. A typical shock mount is shown on Fig. 4-147. The application of a bond-strap to a vehicle engine is shown on Figs. 4-148 and 4-149. The resiliency of the bonded mount should be determined by characteristics of the mount, not of the bond-strap. The strap should not significantly dampen the shock mount and, where necessary, it should be corrugated to withstand severe and continued vibration. In the VHF range and higher, two bond straps across each shock mount should be used.

TABLE 4-32. METAL CONNECTIONS

Metal Structure (Outer Finish Metal)	Connection for Aluminum Jumper	Screw Type*	Connection for Tinned Copper Jumper	Screw Type*
Magnesium and magnesium alloys	Direct or magnesium washer	Type I	Aluminum or magnesium washer	Type I
Zinc, cadmium, aluminum and aluminum alloys	Direct	Type I	Aluminum washer	Type I
Steel (except stainless steel)	Direct	Type I	Direct	Type I
Tin, lead, and tin-lead solders	Direct	Type I	Direct	Type I or II
Copper and copper alloys	Tinned or cadmium-plated washer	Type I or II	Direct	Type I or II
Nickel and nickel alloys	Tinned or cadmium-plated washer	Type I or II	Direct	Type I or II
Stainless steel	Tinned or cadmium-plated washer	Type I or II	Direct	Type I or II
Silver, gold, and precious metals	Tinned or cadmium-plated washer	Type I or II	Direct	Type I or II

*Type I is cadmium- or zinc-plated, or aluminum; Type II is stainless steel. Where either type is indicated as acceptable, Type II is preferred from a corrosion standpoint.

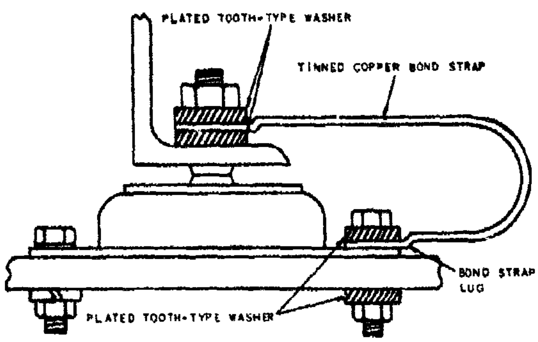


Figure 4-147. Typical Shock Mount Bond

4-7.2.6.2 Rotating Joints

Frequently, it is necessary to bond shafts of rotating machinery to prevent accumulation of static charges. Bonding generally is accomplished by use of a slip ring and brush assembly, or a phosphor-bronze finger riding directly on the shaft.

4-7.2.6.3 Tubing Conduit

The outer surfaces of long spans of conduit or shielded cable may be high-impedance paths for interference currents from external sources. To minimize this possibility, such spans should be properly bonded to structures at both ends and at several intermediate points. Ordinary clamps cannot be used to bond flexible conduit since the required pressure on a comparatively small surface area of the conduit may be sufficiently high to compress or collapse it. To overcome this, a flared split-sleeve is fitted around the flexible conduit. This sleeve distributes the high pressure of the bonding clamp over a large area, thereby exerting low pressure on the conduit (Fig. 4-150(A)). Contact further is improved by soldering the sleeve to the conduit, material permitting, through several holes in the sleeve provided for this purpose. Fig. 4-150(B) illustrates a method for bonding rigid conduit to a structure through supporting attachments. The conduit or tubing to which bonding clamps are attached should be cleansed of paint and foreign material over the entire area covered by the clamps. All insulating finishes should be removed from the contact area before assembly only and conductive screws, nuts, and washers should be used to attach contacting parts.

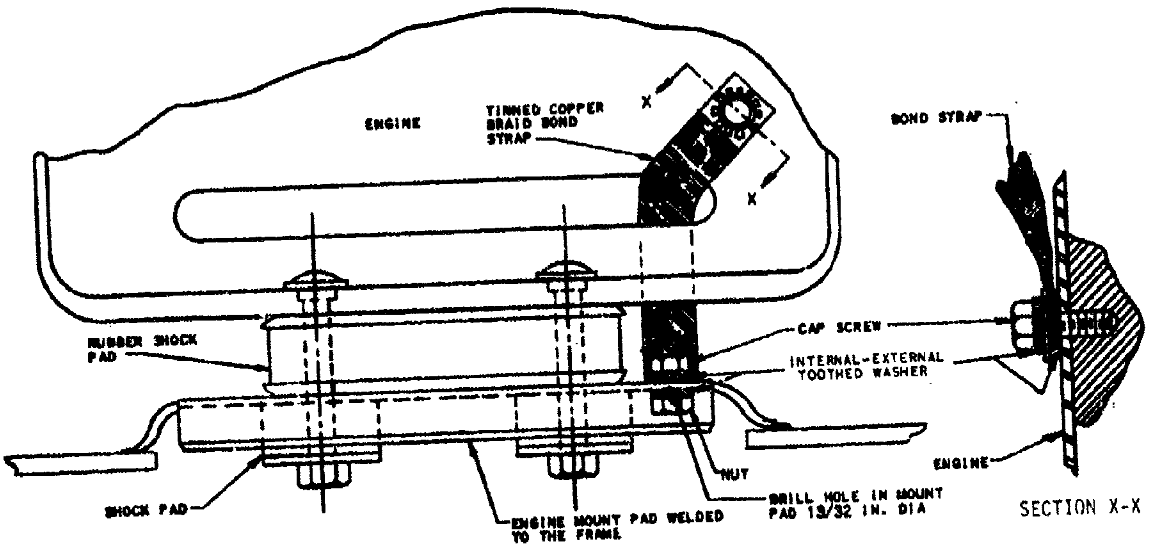


Figure 4-148. Bonded Engine Shock Mount — Front

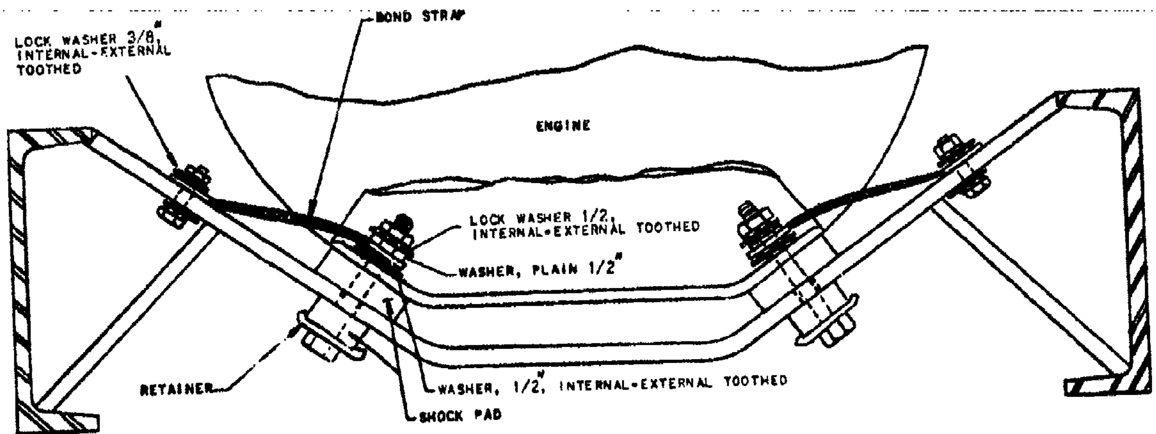


Figure 4-149. Bonded Engine Shock Mount — Rear

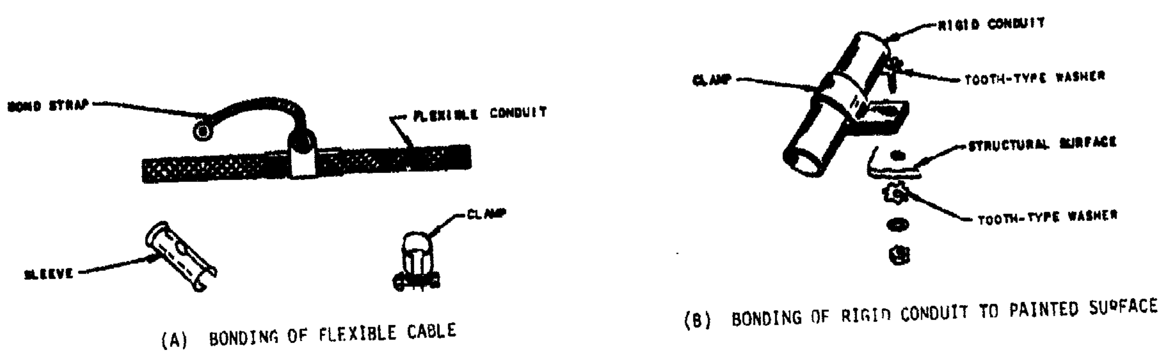


Figure 4-150. Cable and Conduit Bonding

4-7.2.6.4 Hinges

Where hinges must be used, it is necessary to accomplish bonding by suitable means. Fig. 4-151 shows a typical configuration for bonding hinges.

4-7.2.6.5 Cable Trays

Cable trays should be used as part of the overall system bonding scheme. Each section of each tray should be bonded to the following section to provide a continuous path (Fig. 4-152). The trays also should be connected to equipment housings by wide, flexible, solid bond-straps. A typical example of cable tray bonding is shown on Fig. 4-153.

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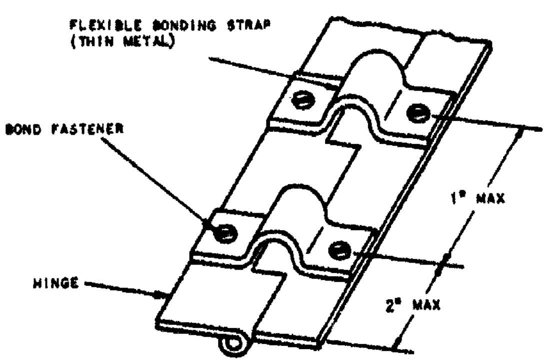

Figure 4-151. Bonding of Hinges

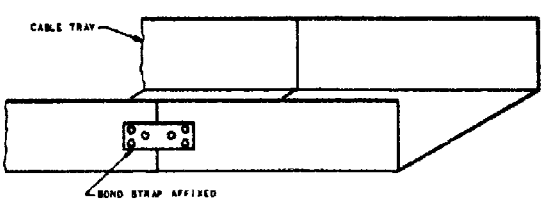

Figure 4-152. Cable Tray Section Bonding

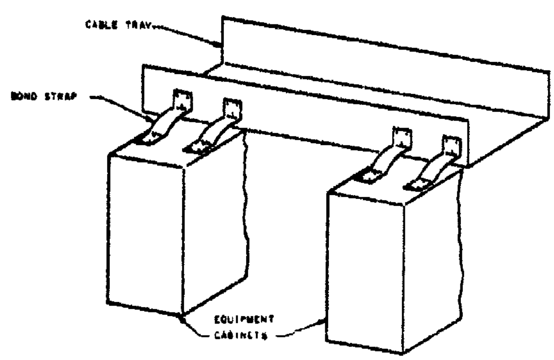

Figure 4-153. Equipment Cabinets Bonded to Cable Tray

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CHAPTER 5
APPLICATIONS TO SPECIFIC DEVICES

All military equipment which is capable of emanating or is susceptible to electromagnetic energy requires interference reduction treatment in order to assure that it will be electromagnetically compatible with other equipment or the environment in which it is expected to operate. To assure compatibility the equipment must meet requirements given in various military standards as outlined in Chapter 2, especially conducted and radiated limits given in MIL-STD-461, in accordance with test procedures in MIL-STD-462. In general, requirements exist over the frequency range from 30 Hz to 40 GHz.

In this chapter the objective is to give examples showing how the various general techniques discussed in Chapter 4 are applied in specific types of circuits and devices, and also to describe any specialized techniques that may be appropriate.

5-0 LIST OF SYMBOLS

A = constant
A_r = effective area of receiving antenna, m^2
A_t = effective area of transmitting antenna, m^2
A_1, A_2 = constant, V·s
a = wire diameter, m; the ratio of the amplitudes of an undesired signal to that of the desired signal, dimensionless
a_n = coefficient in series expansion
B = bandwidth, Hz
B_e = effective bandwidth, Hz
B_m = moment bandwidth of a modulating waveform, Hz
BER = bit error rate, s^{-1}
C = capacitance, F
c_o = speed of light, m/s
D = antenna diameter, m
D_Y = antenna dimension (Y-axis), m
D_Z = antenna diameter (Z-axis), m
$DPSK$ = differential phase shift keying
d = diameter; distance from a single wire; horizontal distance from power transmission line, m; distance from antenna, m
E = electric field strength, V/m; ac voltage magnitude, V; signal "energy" per pulse (radar), V^2·s
E_c = critical field gradient, electrostatic field, V/m
E_L = line phase voltage, V
E_0 = field strength at angle 0, V/m
E_s = secondary phase voltage, V
E_ψ = field strength at angle ψ, V/m
e = electronic charge, C
e_c = common-mode voltage, V
e_d = differential-mode voltage, V
f = frequency, Hz
Δf = ratio of frequency spacing of desired and undesired signals; Hz; relative air density, dimensionless
f_c = 3-dB filter bandwidth, Hz
f_d = binary data rate, Hz
f_F = fundamental frequency, Hz
f_n = frequency of nth harmonic, Hz
f_r = tuned receiver frequency, MHz
f_s = carrier frequency, Hz
$G(G,\varphi)$ = gain of antenna, dimensionless
G_r = receiver antenna power gain, dimensionless
G_t = transmitter antenna power gain, dimensionless
H = magnetic field strength, A/m; height of source from ground, m
$H(f)$ = filter transfer function, dimensionless
h = height from power transmission line, m; height of conductor from ground plane, m
I = current, A
I_c = induced cavity current, A
I_d = output current, A
I_L = line current, A
I_o = direct current (in a tube), A
I_{pr} = peak reverse current, A
I/S = interference-to-signal ratio
i = current, A
i_n = current of nth harmonic, A
$i_1(0^-)$ = value of the interrupted supply current, A
i_2 = secondary current (coupled coils), A
J_n = nth order Bessel function of the first kind
K = constant in propagation equation, V; phase conversion factor, deg/dB

K_B, K_F = empirical constants (microstrip), dimensionless
k = an integer; coefficient of coupling (coupled coils); attenuation
L = inductance, H; loss factor, dimensionless
L_1 = inductance of primary coil (coupled coils), H
L_2 = secondary inductance (coupled coils), H
L_λ = aperture diameter measured in wavelengths, dimensionless
l = length, m; dipole length, m
m = electron mass, kg
N_o = noise power spectral density; V^2/Hz; neutral point in a rectifier circuit
$NBFM$ = narrow-band frequency modulation
n = order of harmonic; ratio of primary to secondary phase voltage, dimensionless
n_1 = number of turns, coil primary
n_2 = number of turns, coil secondary
P = power, W
P_i = power input, W
P_o = power output, W
P_r = power received, W
P_t = power transmitted, W
PCM = pulse code modulation
p = atmospheric pressure, cm of Hg
Q = quality factor of a network, dimensionless
R = resistance, Ω; distance to the boundary between the near-field and far-field (antennas), m; signal-to-noise ratio; reflection loss, dB
R_D = resistance of a diode, Ω
R_1 = resistance of primary coil (coupled coils), Ω
R_2 = secondary resistance (coupled coils), Ω
r = conductor radius, cm; range; separation distance, m
S = spacing between conductors, m
SE = shielding effectiveness, dB
s = spacing between input and output cavities, m
$s(t)$ = signal level, function of time, V
T = spacing of pulses in time, s
T_D = propagation time per unit length, ns/ft
T_{rr} = reverse recovery time, s
t = time, s; temperature, °C
t_o = line propagation time, ns
t_r = rise time, s

V = voltage, V
$V(f)$ = Fourier spectrum, V·s
$V_B(t)$ = voltage of coupled line at the sending line, V
$V_F(t)$ = voltage of coupled line at the receiving line, V
V_o = beam accelerating voltage, V
$V_o(t)$ = voltage at the receiving end of the driving line at line t (microstrip), V
V_s = supply voltage, V
V_1 = peak gap voltage at buncher (klystron), V
$v(t)$ = voltage, function of time, V
v_n = voltage of nth harmonic, V
v_1 = voltage, primary winding, V
v_2 = voltage, secondary winding, V
W = conductor width, m
$WBFM$ = wide band frequency modulation
x = instantaneous input; bunching parameter
x_a = input (phase a)
x_b = input (phase b)
$x_1(t)$ = information signal
$x_2(t)$ = carrier signal
y = instantaneous output
y_a = intermodulation component; output of a nonlinear device (phase a)
y_b = cross-modulation component; output of a nonlinear device (phase b)
Z = impedance, Ω
Z_o = characteristic impedance, Ω

α = firing angle, deg
Δ = rise time of a trapezoidal pulse, s
δ = difference frequency between desired and undesired signals, Hz; relative air density, dimensionless
η = intrinsic standoff ratio, dimensionless
θ = elevation angle, rad; beam angle (antenna), rad; half of the conduction angle, rad; line length, rad
λ = wavelength, m
σ = radar cross section, m^2; VSWR of load, dimensionless
τ = pulse duration, s
φ = azimuth angle, rad; phase
φ_c = carrier phase, rad
ψ = direction of horizon, rad
ω = angular frequency, rad/s; excitation frequency, rad/s
ω_c = angular carrier frequency, rad/s

5-1 ROTATING ELECTRICAL MACHINES

5-1.1 BRUSH PHENOMENA

Any rotating machine with sliding contacts is a potential source of interference. If the contacts are used on continuous "slip" rings, the interference generated is a result of varying contact resistance caused by varying brush pressure or surface conditions on the ring. It is likely to be random in nature and have energy only in the low audio frequency portion of the spectrum.

If the contact surface is that of a commutator, the associated switching and arcing processes cause rapid current and voltage changes that generate interference energy throughout a wide frequency range as discussed in par. 3-1.3.2.

Brushes and brush leads are the most likely components from which interference can be radiated or transferred. Provision should be made in the original design of motors or generators for installation of capacitors at the brushes (see Fig. 5-1). In very critical cases it may be necessary also to shield the brush leads. Brush-generated interference may be minimized by proper consideration of the following:

a. Brush pressure. Increasing brush pressure generally reduces generated interference at all frequencies since it tends to reduce contact resistance and variation of that resistance. In addition, in cases where the slip ring or commutator is not precisely concentric or there is wear in shaft bearings, increased brush pressure will minimize contact "bounce" phenomena. The brush mounting arrangement should be designed to provide some damping without reducing the effective force exerted on the contact.

However, as the brush pressure is increased the rate of wear of the brush and the slip ring or commutator increases necessitating frequent brush replacement.

b. Current density. Generated interference decreases with decreased current density. As the current density is increased, the heat generated at the brush surface is increased, possibly causing the formation of an oxide film on the sliding metal surface. Rapid variations in the sliding contact resistance, resulting from irregularities in this oxide film, produce high-frequency transients. To offset the heat increase, a somewhat larger brush-surface area than otherwise necessary can be used.

c. Brush resistivity. Generally, low resistivity brushes may be expected to generate less interference than those of higher resistance. However, when used with a commutator, the resistance of the brush should match the requirements for good commutation. The design engineer should select material of the lowest resistivity that satisfies the other requirements of good functional performance. When used with slip rings, a wide choice of brush material is available because no switching action is involved. Low-resistance brushes are available with silver, copper, or cadmium-impregnated graphite.

5-1.2 COMMUTATION

5-1.2.1 Design Considerations

Dc motors and generators are serious offenders in generating interference because they require commutators for their operation. The bars of a commutator, rapidly sliding past the contacting brushes, produce a switching action that causes voltage transients, or pulses. The objective of interference reduction design techniques is to provide a smooth transition from one value of impedance to another while switching between each armature coil. The techniques include use of interpoles, compensating windings, increased number of armature coils and commutator bars, laminated brushes, and commutator plating. A discussion follows:

a. Interpole Windings. The best way to improve commutation is by adding interpole windings. Interpoles counterbalance the self-induction of the armature coils during the commutation period, and also reduce the induced voltage in the armature coils resulting from the coil-cutting fringing-flux during the commutation period. The use of properly designed interpoles produces a rapid change in the armature-coil current at the beginning of the commutating period, reducing the steepness of the transient at the end of the commutating period.

b. Compensating Windings. To a lesser degree, compensating windings produce the same effect as interpoles and, in addition, help to prevent field distortion. They also assist in reducing cross flux produced by the armature coils. The use of interpoles and compensating windings lessens critical brush positioning requirements with respect to the commutator, and provides electromotive forces in the coils under commutation which oppose the electromotive forces of self- and mutual-induction in these coils.

c. Increased Number of Armature Coils. Increasing the number of coils on the armature (thereby increasing the number of commutator segments, or bars) reduces interference by reducing the current broken per bar and the reactance voltage per coil. The largest number of armature slots, in which the coils are uniformly distributed with respect to the commutator bars, should be used, and the armature

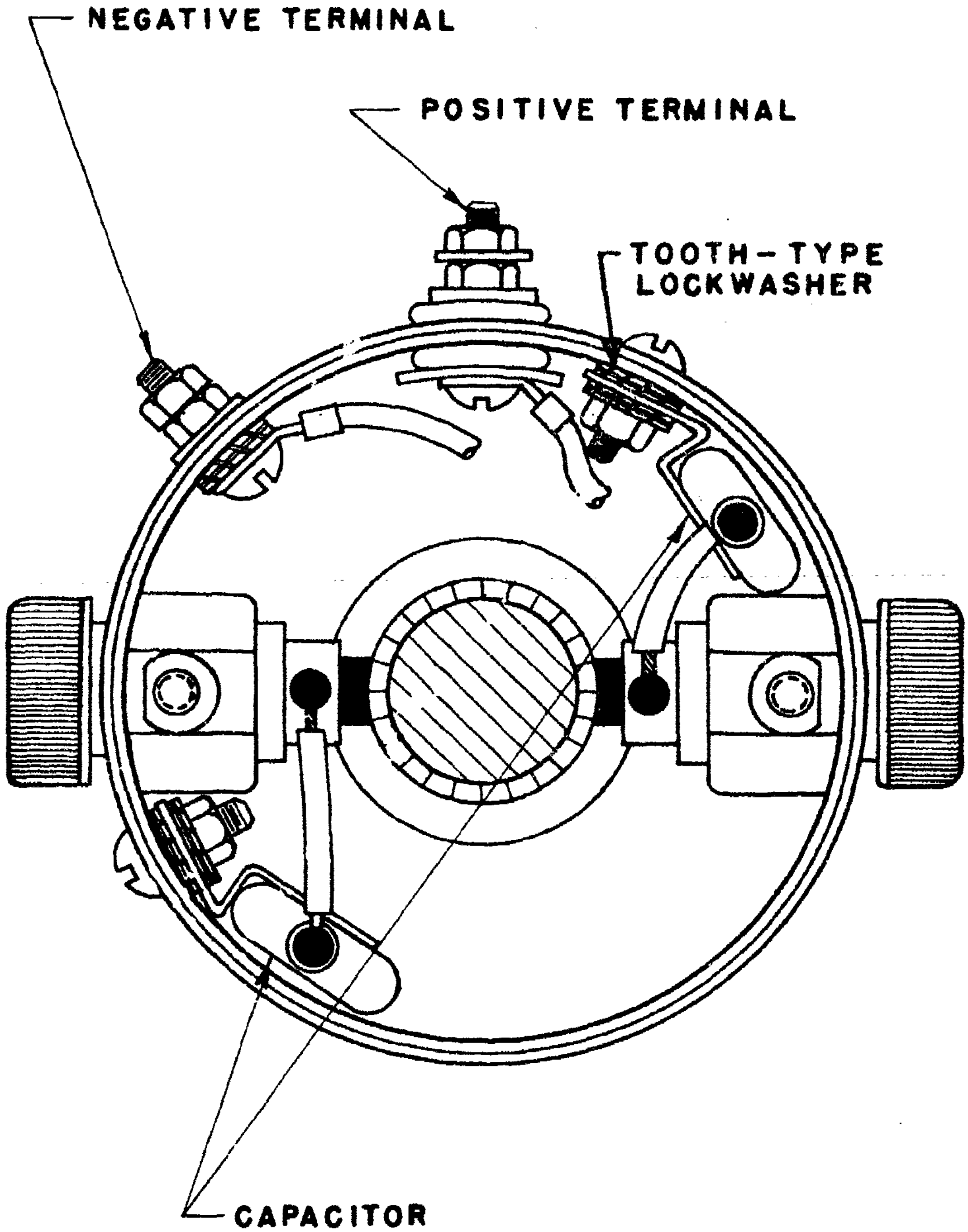


Figure 5-1. Installation of By-pass Capacitors

slots should be as shallow as possible. The use of short-pitching windings reduces interference by reducing the reactance voltage of each coil.

d. Laminated Brushes. The break transients, resulting from the switching action of the commutator, can be smoothed out through the use of laminated brushes. These consist of brush materials of different resistivity cemented together by nonconducting glue which provides insulation between adjacent brush segments. The operation of laminated brushes is indicated in Fig. 5-2 — having the successive segments of the brush increase in resistance avoids the sharp current drop after the brush leaves the commutator segment. A more linear coil-current reversal results, thus reducing the break transients. Circulating currents, resulting from the self-inductance of the coil under commutation and from the coil-cutting fringe flux from the pole pieces, must flow through the entire length of two brush laminations. The total resistance of this length is much greater than that presented by a direct path across the face of the brush (as would occur with a solid brush). Therefore, circulating currents are reduced early in the commutation period, and desirable division of current through the two adjacent commutator bars is achieved.

Good commutation can be achieved over a fairly wide range of brush positions, relative to the magnetic neutral, so that brush positioning becomes less critical and less dependent upon armature current. The design of laminated brushes should include two or, at most, three laminations. The following criteria should be incorporated in the design:

(1) The thickness of the leading-edge lamination of a two-lamination brush should be about 90% of the total thickness, and its resistivity should be as high as allowable for heat dissipation.

(2) The resistivity of the trailing-edge lamination should be about 15 times that of the leading edge. This lamination should be thick enough to preclude mechanical weakness.

(3) A brush with varying resistance characteristics from the leading edge to the trailing edge can be manufactured without the use of insulating separators and will act somewhat like a laminated brush.

e. Commutator Plating. A copper commutator, in contact with a carbon or graphite brush, develops, after several hours, a layer of copper oxide mixed with carbon particles from brush wear. This copper-oxide film has a resistance of higher value at the brush used as cathode than at the one used as anode,

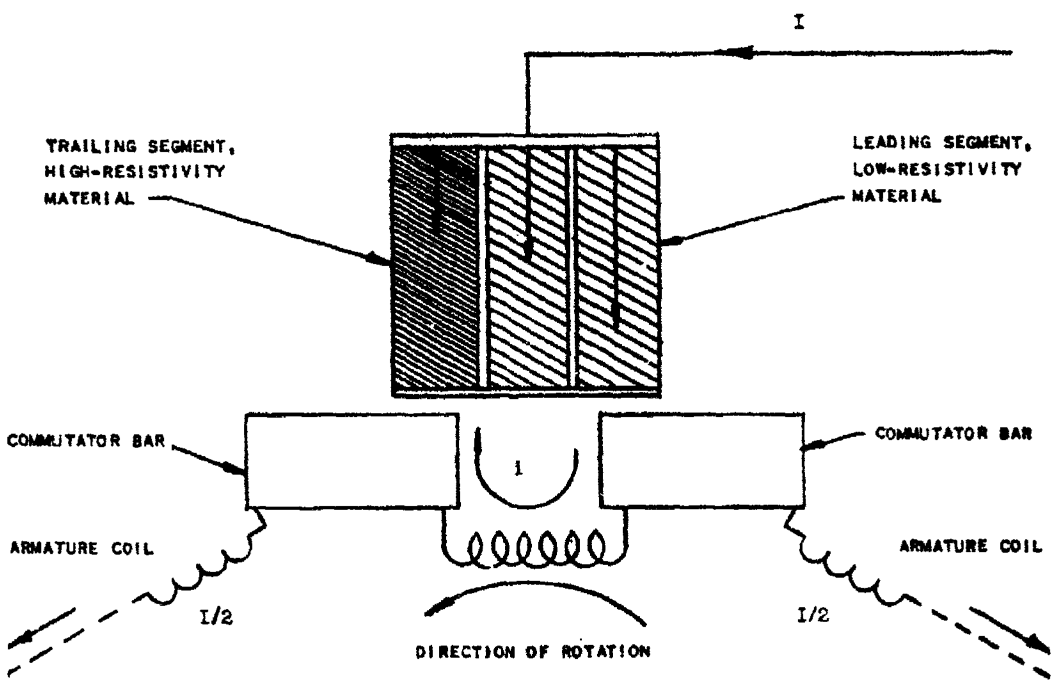


Figure 5-2. Commutation of an Armature Coil by Laminated Brushes

which upon breakdown passes current in discontinuous surges. Approximately 10 times as much interference may result from the cathode brush as from the anode brush. Plating the copper commutator with chromium to a thickness of about 1 mil reduces the interference level from a cathode brush to that of a relatively quiet anode. In addition, the hard chromium surface prevents threading and grooving of the commutator. Wear-rate and sliding friction of many brush materials on chromium are of the same order of magnitude as those for copper.

5-1.2.2 Suppression

The interference produced by a motor or generator can be controlled by filtering and shielding. The most effective and economical technique is the installation of capacitors at the brushes. If the machine frame is grounded, the interference can be bypassed to the frame (see Fig. 5-1). The lead from the brush to the capacitor should be as short as possible. A good value for such a capacitor is 0.1μF and it should have an appropriate voltage rating. An additional capacitor can be installed at the output (armature) terminal. The preferred installation is a feedthrough capacitor at the generator housing as shown in Fig. 5-3(A). Fig. 5-3(B) illustrates a less satisfactory mounting of a bypass capacitor at the armature terminal. The installation of capacitors reduces the interference appearing externally on the armature, field terminals, and wiring.

To prevent the radiation of interference from within the machine, the machine housing should be designed to provide maximum shielding effectiveness. It may be necessary to apply screening, which is bonded to the housing, over all ventilation openings. In addition, commutator and brush inspection covers should be machined as closely as possible, and should be wide enough to cover the inspection opening adequately, with sufficient overlap to ensure good contact. The cover should have closely spaced bolts — 2 in. to 6 in. depending upon application — to assure that a metal-to-metal surface is maintained between cover and housing. Interference gasketing can be installed around the periphery of the opening. After removal of an inspection cover, all contact surfaces on the cover and the generator should be thoroughly cleaned before the cover is put back into position.

Finally, there should be metal-to-metal contact between the three sections of the generator frame: the two end-plates and the main housing. Where necessary this may be accomplished by the bonding practices discussed in par. 4-7. Since the shaft provides an exit path for interference, the interference should be bypassed directly to the generator housing by grounding. To do this, the shaft may be grounded to the housing by a brush, riding on a special slip ring (or riding directly on the shaft). This grounding also will eliminate bearing interference (bearing static or shaft current) that results from periodic discharge of static electricity generated in the bearing, and also leakage due to eddy currents induced in the shaft and the housing by the flux lines in the machine. These currents also can be caused by certain combinations of armature segments per pole, air gap and permeability inequalities, rotor eccentricities, insulation leakage, or stray electric fields.

On some dc motors an adjustable speed control is included in which the field leads are connected to an externally-mounted rheostat. This arrangement can permit interference, generated inside the motor, to be conducted out of the housing unless capacitors are connected to these leads to bypass such interference to ground and mounted either within the housing or, preferably, through the housing if of a feedthrough-type.

5-1.3 ALTERNATORS AND SYNCHRONOUS MOTORS

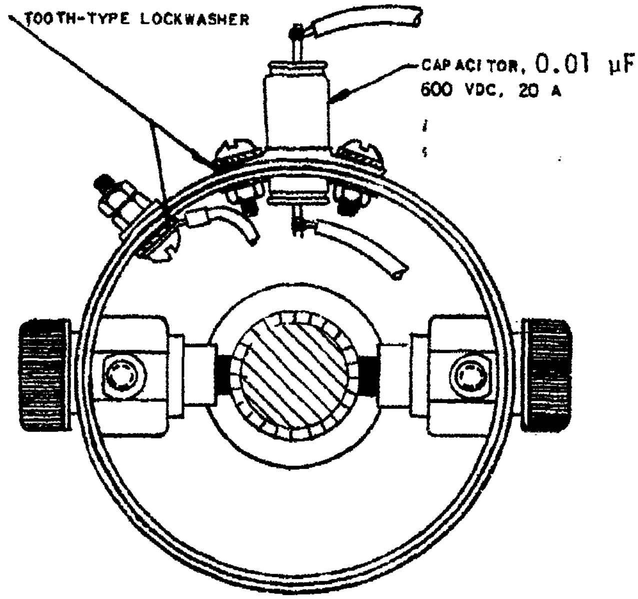
5-1.3.1 Design Considerations

Alternators and synchronous motors have slip rings rather than commutators. Interference can arise from the brushes and from the generation of harmonics. Brush interference is minimal because most alternators and synchronous motors have stationary armatures and rotating fields; heavy power currents need not be supplied to the rotor. Only the much smaller field currents are supplied through the brushes. Because commutation need not be considered in the selection of brushes, the design engineer is permitted a much wider choice in brush pressure, size, and material than for commutator machines.

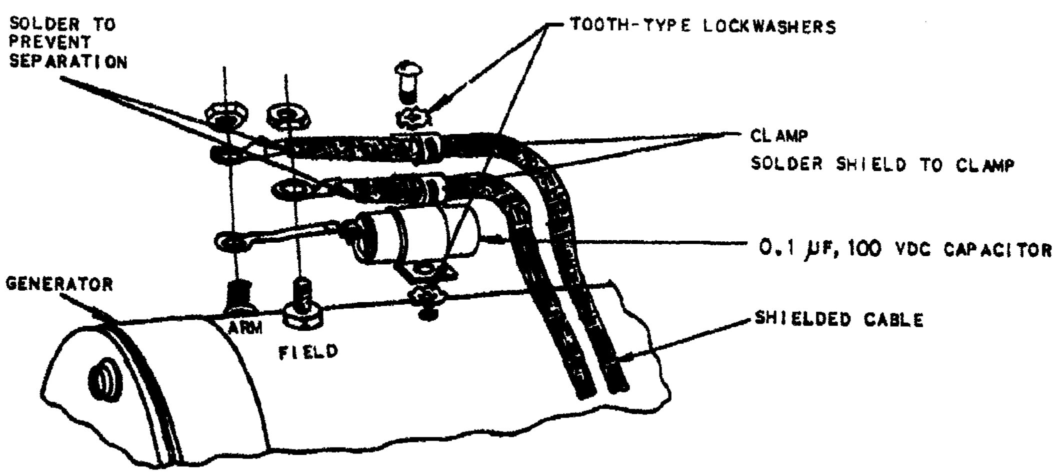
If the alternator is not intended to be used with nonlinear (rectifying) loads, the machine designer can minimize the generation of harmonic currents and voltages by giving special attention to the following:

a. Flux distribution. The most important factor determining the waveform of the generated voltage is the distribution of the magnetic flux around the periphery of the armature. Sinusoidal distribution, which produces the least amount of harmonics, may be achieved by chamfering the pole tips or skewing the pole faces.

b. Symmetry. In a perfectly symmetrical machine, there are no even harmonics. Special care must be exercised to construct identical pole pieces, to make the yoke and armature perfectly symmetrical, and to



(A) PREFERRED INSTALLATION OF FEED-THROUGH CAPACITOR



(B) MOUNTING OF BYPASS CAPACITOR AT ARMATURE TERMINAL
IN A DC GENERATOR

Figure 5-3. Mounting of By-pass Capacitors

produce a perfectly uniform winding on the armature.

c. External connections. In a three-phase alternator, the third harmonic and its multiples tend to cancel at the terminals of a delta connection or a wye connection with floating neutral.

d. Distribution factor. The distribution factor should be chosen to eliminate the lowest harmonic not eliminated by any of the devices mentioned in pars. b or c.

e. Tooth ripples. The generation of tooth ripples is greatly decreased by skewing, through one slot pitch, either the pole shoes or the armature slots. Tooth ripples may be eliminated altogether by making a number of armature-slots per pole-pair an odd number. The chord factors for the harmonics that are contained in the tooth ripples are then reduced to zero.

If the alternator is intended to be used with nonlinear loads, it is normally not possible to filter the lower order harmonics generated, and these will appear in the windings of the machine regardless of how well it is designed otherwise. The nature of the interference generated by rectifying action and means of suppressing it are treated in par. 5-6.

5-1.3.2 Suppression

Slip ring and brush material design considerations for the commutator surface in dc machines apply equally well here. The effects of brush bounce, due to vibration or irregularities of armature motion, can be minimized by the use of two or more brushes per slip ring.

The exciter, which is essentially a dc generator, is correspondingly a potential source of interference. The techniques of shielding, plating of the commutator, the use of proper brushes and brush pressure, and the application of bypass capacitors are applicable to the exciter. Conducted interference generated directly by the alternator can be reduced by the use of bypass capacitors installed at terminal outlets.

As in the case of the dc generator, shielding of the alternator is incorporated in the design of its housing. Low-impedance paths between sections of the housing, provisions for bonding, and screening of all ventilating louvers and shaft grounding must be carefully maintained. The alternator terminal outlets can be prevented from radiating interference by the installation of capacitors of the two-terminal type just inside the terminal strip, or of the feedthrough-type in the terminal strip.

5-1.4 INDUCTION MOTORS

The primary source of interference in a single-phase induction motor is the starting mechanism. The starting winding is in series with a switch (or capacitor and switch) that is closed when power is off. When the motor reaches approximately 80% of its rated speed, the switch is opened (either by centrifugal force or by a solenoid coil) and a single pulse of interference is generated. To reduce interference, the switch should be placed in a shielded housing and the leads leaving the housing should be filtered.

5-1.5 PORTABLE FRACTIONAL-HORSEPOWER MACHINES

Portable fractional-horsepower machines include such equipment as portable electric drills and saws. Power is furnished by high-speed, lightweight, ac-dc universal motors, or ac electric motors. Interference generation occurs because of commutation, and the suppression techniques of par. 5-1.2.1 apply. In some portable ac-dc machines, restrictions of size and shape prevent the installation of capacitors at the brushes, and it is more feasible and economical to mount the capacitors in other parts of the equipment. Shielding may be used to insure that no interference couples back into the leads before they leave the unit.

5-1.6 SPECIAL-PURPOSE MACHINES

Special-purpose rotating machinery includes a variety of equipment; the most important of these are rotary inverters, dynamotors, motor generators, and generators for electric arc-welding equipment.

5-1.6.1 Rotary Inverter

A rotary inverter, which converts dc to ac, is basically a dc motor with added taps on the armature winding; slip rings are connected to these taps to provide the ac output. Interference is generated by both the ac and dc functions — commutator and brush action in the motor, and brush action and harmonics in the alternator. Fig. 5-4 illustrates an interference-reduction design technique for an inverter. The schematic diagram shows two feedthrough capacitors bypassing interference from the output leads of the alternator. The dc lead from the motor is filtered by a feedthrough capacitor and shielded to prevent radiation from the terminal on the hot side of the capacitor. The ac output leads do not require shielding because the interference generated by the alternator is much less severe than that generated by the dc motor.

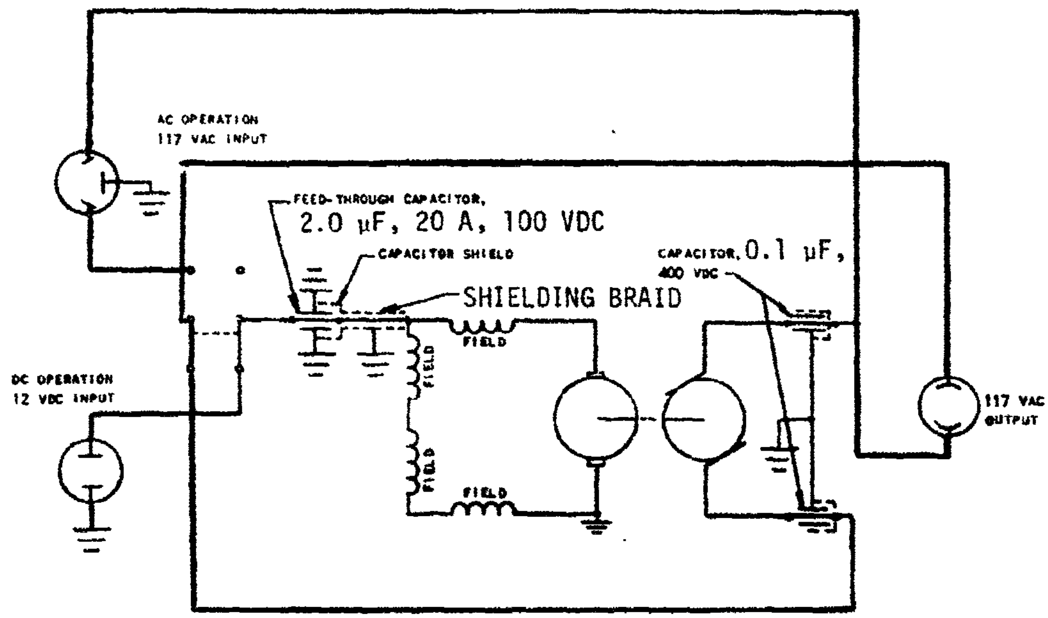


Figure 5-4. Interference Reduction Design Technique for Rotary Inverter

5-1.6.2 Dynamotor

A dynamotor (a combination dc motor and generator with a single magnetic field) has an armature with two separate windings and two separate commutators, one at each end of the armature. It transforms low-voltage dc to high-voltage dc, or vice versa. The two commutators make this machine a particularly prolific source of interference. The suppression techniques for dc generators and motors apply to the dynamotor. Fig. 5-5 illustrates a dynamotor, with feedthrough capacitors bypassing interference to the housing on both the input and output leads.

5-1.6.3 Electric Arc Generators

Generators for electric-arc equipment require special attention only in that the supply power to the arc, which generates the interference, should be located away from communication equipment, and in buildings with good shielding characteristics. The leads from the generator to the welding electrodes can become very effective interference radiators and should be adequately shielded.

5-2 POWER DISTRIBUTION

Except at voltages greater than at least 30 kV, power distribution circuits are more likely to be secondary sources of interference than primary sources. Thus, they form part of the coupling path from an interference source to a susceptor. The sources involved can include any device connected to the distribution circuit including the following:

a. Machinery. Motors and generators
b. Motor-driven Appliances:
   (1) Electric office machines
   (2) Electric motor-driven household-type appliances
   (3) Laundry equipment
   (4) Motor-driven tools
c. Arc-producing Devices:
   (1) Electric razors
   (2) Welders
   (3) Gaseous rectifiers
   (4) Spark-ignition-oil burners
   (5) Fluorescent and other gas-filled lamps
d. Communication Devices:
   (1) Radar equipment
   (2) Communication transmitters
   (3) Radio and television receivers
   (4) Teletype machines
e. Switching Devices:
   (1) Neon or flasher signs
   (2) Calls systems (bells, buzzers, etc.)
   (3) Thermostats
   (4) Electric heaters

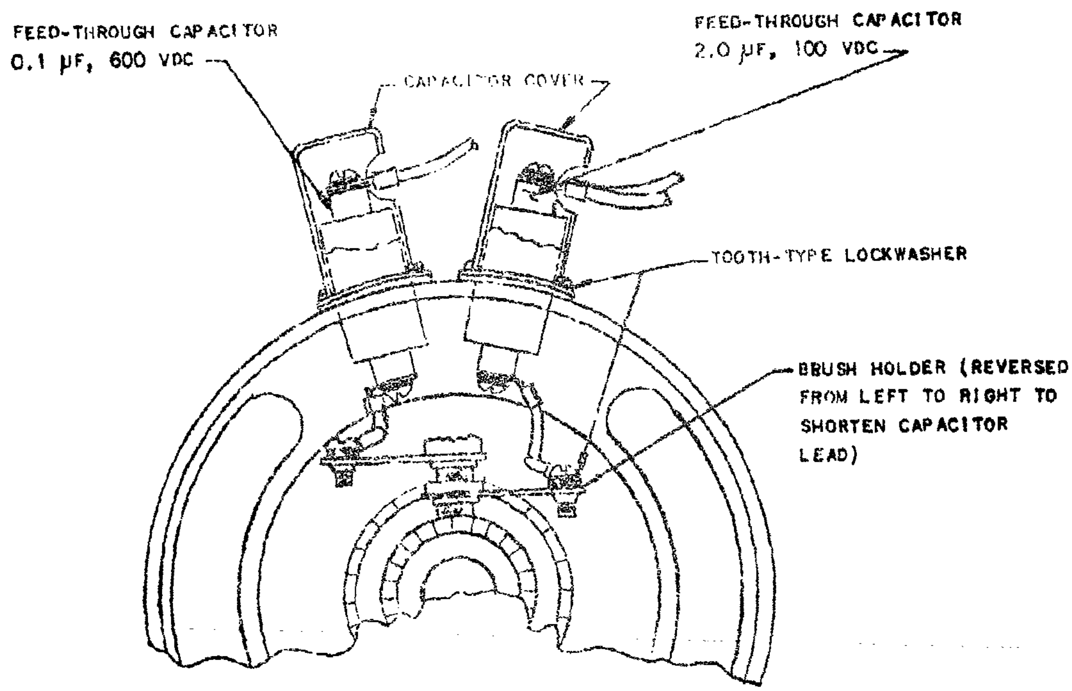


Figure 5-5. Interference Reduction Design Technique for a Dynamotor

f. Industrial Equipment:
   (1) Induction heating equipment
   (2) Medical diathermy equipment
   (3) X-ray machines
   (4) Ultraviolet lamps.

These devices can produce interference of varying kinds. Perhaps the most common kind is the switching transient which may be very short and have a broad frequency spectrum. When arising primarily from equipment turn on and off, it may occur relatively infrequently. A more detailed discussion of such transients appears in par. 5-3. Another kind is harmonics of the power frequency. Where they appear, their magnitude (in terms of conducted current) decreases at least as rapidly as $1/n$ where n is the order of the harmonic. Otherwise, the type of interference can be characterized by the type of source, e.g., a device generating a specific frequency in connection with its operation is likely to couple energy onto the power line at that frequency.

5-2.1 LOW VOLTAGE DISTRIBUTION

5-2.1.1 Direct Voltage Distribution

Direct voltages are used for distributing electrical power where the primary source itself is direct such as a storage battery, or where there are extensive requirements for direct voltage for operation of motors or relays. Such lines usually are considered to be indirect sources through conduction or induction. The prime sources of the interference on the lines are commutation or switching (discussed in par. 5-1) and relay transients. In addition, it is common to use either rotary converters or static inverters to convert low direct voltages to higher alternating or direct voltages. In the latter case, the frequency of inversion and its harmonics will appear on the primary direct voltage line.

5-2.1.1.1 Conductive Coupling

If the source of direct voltage is a battery, the battery is usually of such a low impedance that little coupling will occur between several sources connected directly to the battery terminals. However, if there is a substantial length of wire from the battery to the point at which different loads are branched, conductive coupling can occur by virtue of the mutual impedance between the two circuits contained in the length of line from the branching point to the battery. This impedance can be obtained from Fig. 4-13 if an appropriate correction is made for the actual length of the wire involved. At frequencies where the total length of line is more than about 1/6 of a wavelength, the impedance must be calculated on the basis of transmission line formulas, taking into account the spacing between the two distribution wires (or in the case of the ground return, the spacing between the wire and the ground plane) and the actual length of line. In addition, filters can be used for decoupling in accordance with the discussion in par. 4-5.

5-2.1.1.2 Inductive Coupling

Magnetic or electric inductive coupling between the distribution line and any lines or circuits near it can occur in accordance with discussion in par. 4-4, and the methods mentioned there can be used to control it. Where it is possible to isolate the power return circuit so that there is a return wire, it should be twisted with the main power conductor in order to reduce the magnitude of the induced fields. In some applications, however, where weight is an important factor such as in an aircraft where the fuselage is used for power return currents, this technique is not available. Indeed, the tendency to use common bundles of cables for various functions results in high degrees of coupling between power circuits and signal or control circuits which can be reduced if the signal or control circuit can be carried with twisted or coaxial lines. Otherwise, the only cure to a potential compatibility problem is physical separation of the circuits, or the use of high quality filters to avoid the possibility of any coupling from the interference source to the distribution circuit.

5-2.1.2 Alternating Voltage Distribution

Similar principles apply for alternating voltage distribution as for direct voltage distribution, except that it is not common with ac distribution for the return current to flow through the ground plane. Typically, for a single-phase circuit, both conductors are carried from the source to the load in the same wire bundle. In recent years, it has been common practice for these lines to be twisted, which reduces the possibility of substantial induction from such lines. An exception is the case of extremely large high-current lines having a large cross section, relatively large spacing (of the order of 0.5 in. or more) and a large pitch distance (see Fig. 3-68).

Multiphase circuits may have a grounded or ungrounded neutral. With a grounded neutral, any unbalanced load produces a return current in the ground plane. This current effectively flows in a parallel line made up of the main power distribution cable and its image in the ground plane. Such a configuration is therefore quite capable of producing a substantial field. Where induction from such a field would cause difficulties, it is better to return the neutral in an additional conductor in the basic power cable. This would give the same effects of cancellation that are obtained in a twisted pair of wires carrying equal currents in opposite directions.

In many circuits using ac distribution it is common practice to connect filters between each line and ground for interference suppression purposes. Typically, the filters contain capacitors which can shunt power frequency currents into the ground plane. Balanced filters on three-phase systems theoretically will not produce a net ground current, but manufacturing tolerances on filter components are such that actually substantial currents will exist. The current has been defined in Chapter 3 as a common-mode current and will induce magnetic fields in accordance with the parallel-wire line concept. Such currents can only be avoided by connecting the filters line-to-line instead of line-to-ground or by inserting isolation transformers between each equipment with filters and the main power supply.

5-2.1.3 Switching Transients

Transients appear on power lines primarily as a result of switching action resulting from turning connected devices on and off. The basic theory of such transients is discussed in par. 3-1.3.2. The spectrum associated with any given transient is related to its rise and fall times, and its strength is measured in volt-seconds or in ampere-seconds. It should be noted that the most serious transients do not arise necessarily from high power equipment. Frequently, devices which contain a large amount of equivalent series inductance may generate the most serious transients (Ref. 1). Because of the wide variation in the characteristics of individual transients, the results of measurements must be presented statistically. Figs. 5-6, 5-7, and 5-8 give statistical data on the distribution of transient disturbance amplitude, durations, and rise times. It is seen that a transient may have an amplitude of many hundreds of volts, a duration of several hundred nanoseconds and a rise time t_r of a few nanoseconds. Given specific values of these parameters, the spectrum amplitude can be obtained from Fig. 5-9.

These pulses contain substantial components at frequencies in the tens of megahertz range. The use of rubber covered cable will introduce attenuation factors that will change the pulse shapes significantly in so far as these components are concerned. Fig. 5-10 shows how the rise and fall times are affected for a ten-conductor cable, and Fig. 5-11 shows the attenuation (in terms of transmittance) of such a cable

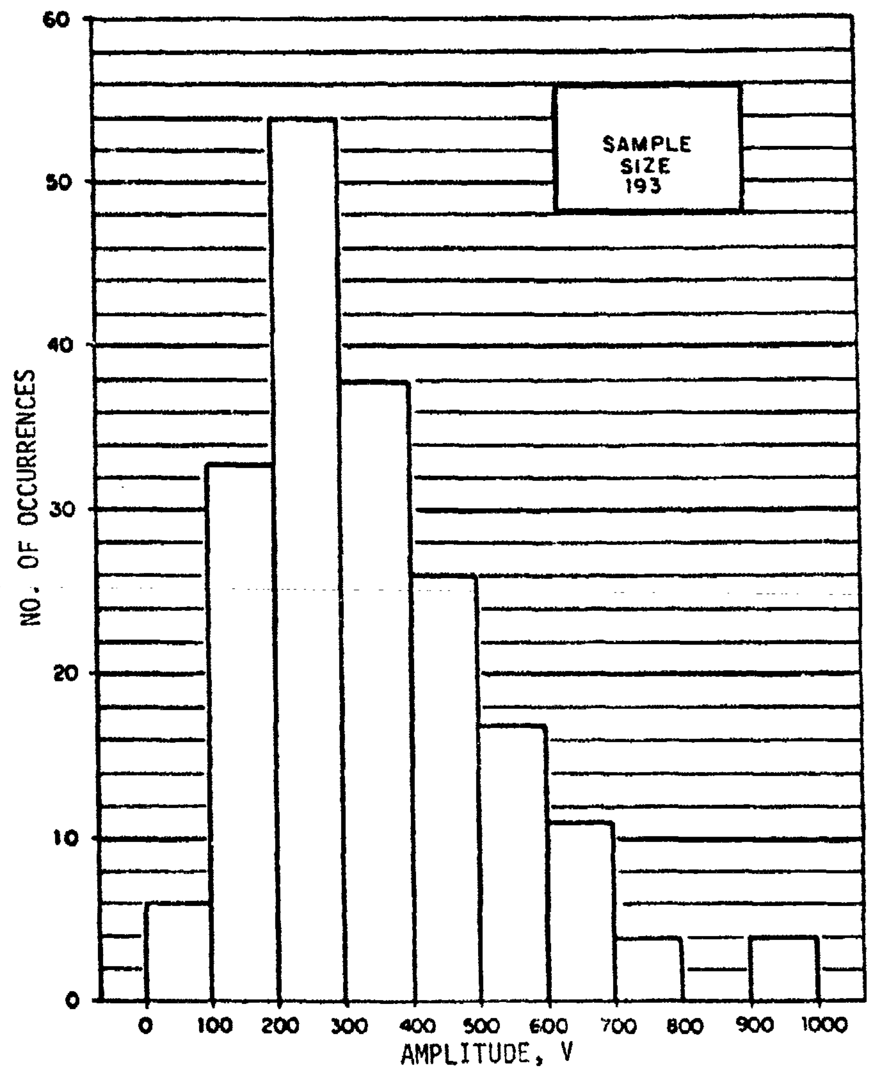


Figure 5-6. Power Line Transient Distribution of Amplitude

as a function of frequency and length of cable. Fig. 5-12 shows corresponding results for conductors in a 3-in. conduit. For these curves, it can be seen that one would not expect the higher frequency component to be carried along the power conductors for substantial distances, so that they must be considered only to the extent that they arise locally (for example, within the immediate premises).

5-2.2 HIGH VOLTAGE LINES

Although similar principals apply to high voltage lines as for low voltage lines, the major sources of interference are high voltage breakdown either in the form of high field corona or across insulators. Such sources are mentioned in par. 3-1.3.2.6 and may occur directly on the wires of the distribution circuits or on the hardware connected to them, such as the transformers and switchgear.

5-2.3 RADIO-NOISE FROM HIGH VOLTAGE TRANSMISSION LINES (Ref. 2)

Corona generated radio-noise on power lines is attenuated very quickly with distance. Figs. 5-13 and

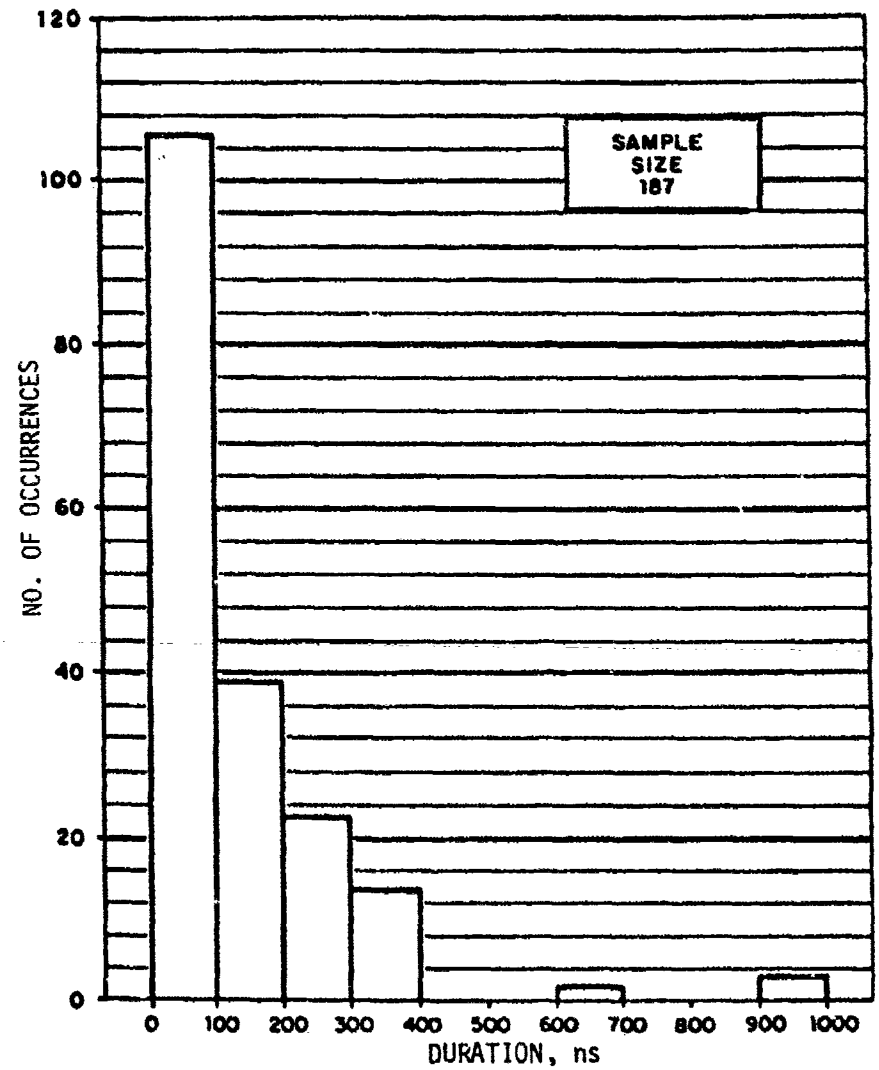


Figure 5-7. Power Line Transient Distribution of Duration

5-14 show the levels (quasipeak) from typical 225-kV and 750-kV power lines having the triangular construction configurations shown. The level at 100 m from the line is quite low in normal weather conditions, but is likely to be at least 20 dB higher under conditions of heavy rain. Corona radio-noise levels are relatively significant in the broadcast and HF bands but decrease rapidly at frequencies above about 1.0 MHz, as shown on Fig. 5-15. At frequencies above 100 MHz, the levels appear to be virtually insignificant. Most of the measurements taken during on-site tests have been at frequencies in the broadcast band and below. These tests have shown that levels of corona noise can also fluctuate greatly from day to day as indicated on Figs. 5-13 and 5-14. Conditions of humidity and wind can cause large and rapid fluctuations of noise levels. These levels also tend to change over long periods of time as the transmission line ages; newly installed lines tend to be noisier than older lines.

Corona always will emanate from local points of sufficiently high field intensity, i.e., from points of surface discontinuity at which a high potential gradient exists. For this reason, corona discharges

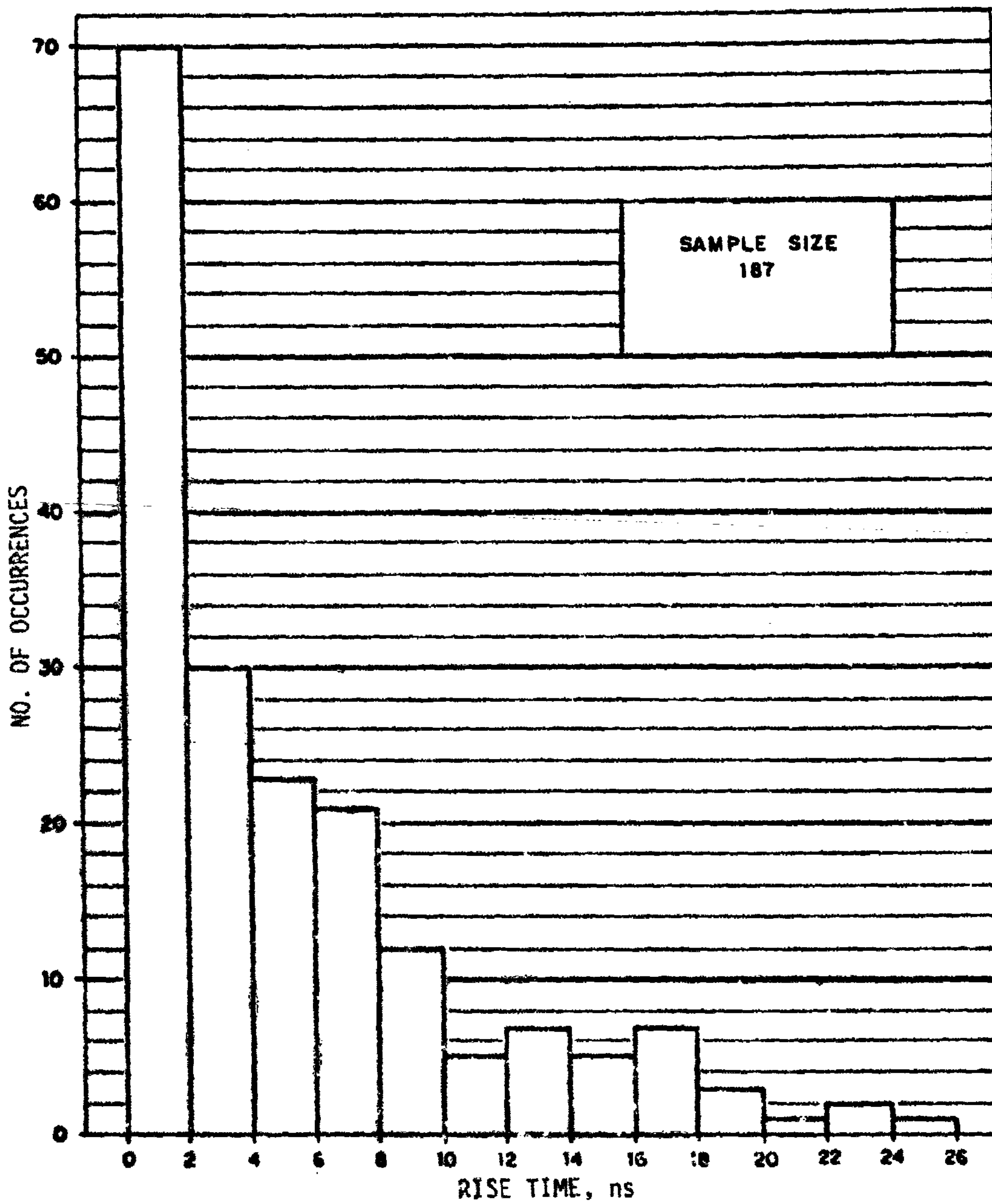


Figure 5-8. Power Line Transient Distribution of Rise Time

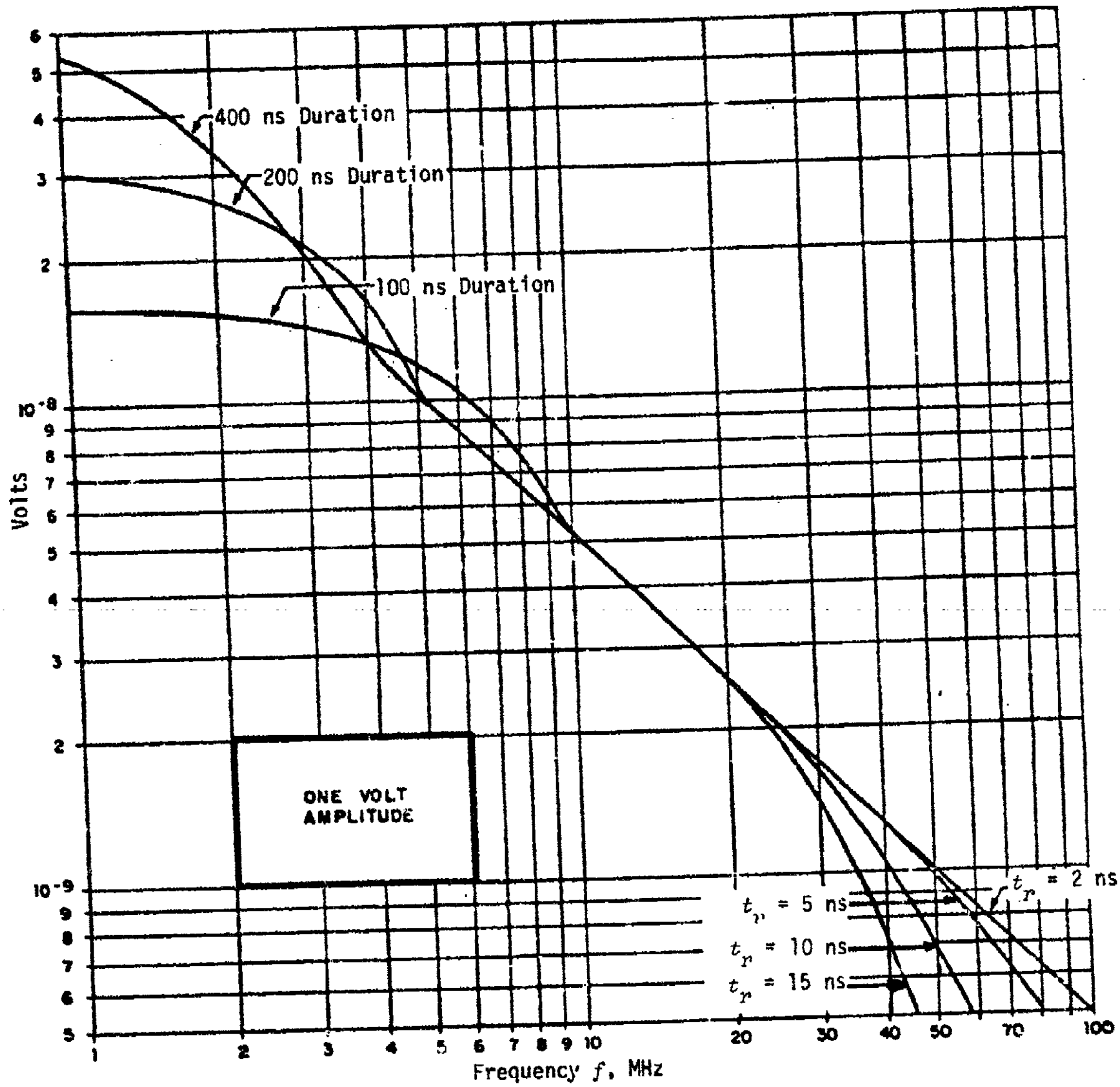


Figure 5-9. Triangular Pulse Spectrum

take place at the sharpest points on the surface of a conductor. The mechanism of discharge may be explained as follows. Incidental ionization, which is always present, provides a supply of electrons in the vicinity of the conductor. At negatively charged points of high-potential gradient, these electrons are accelerated away from the conductor by the strong field and, in fact, possess sufficient energy to ionize the surrounding air molecules. An avalanche effect then results. However, the region of ionization remains confined to the space surrounding the discharge point. The positive ions created by the discharge are attracted to the negatively charged corona point, thereby reducing the potential gradient and quickly quenching the corona discharge. The discharge current is therefore pulse shaped and lasts only for a short time, ranging between 0.1 and 0.5 μs.

The discharge from a corona point tends to repeat itself such that a continuous series of recurrent pulses is generated. The repetition rate of these pulses is a function of the potential gradient at the point where the corona is formed. Generally, pulse repetition

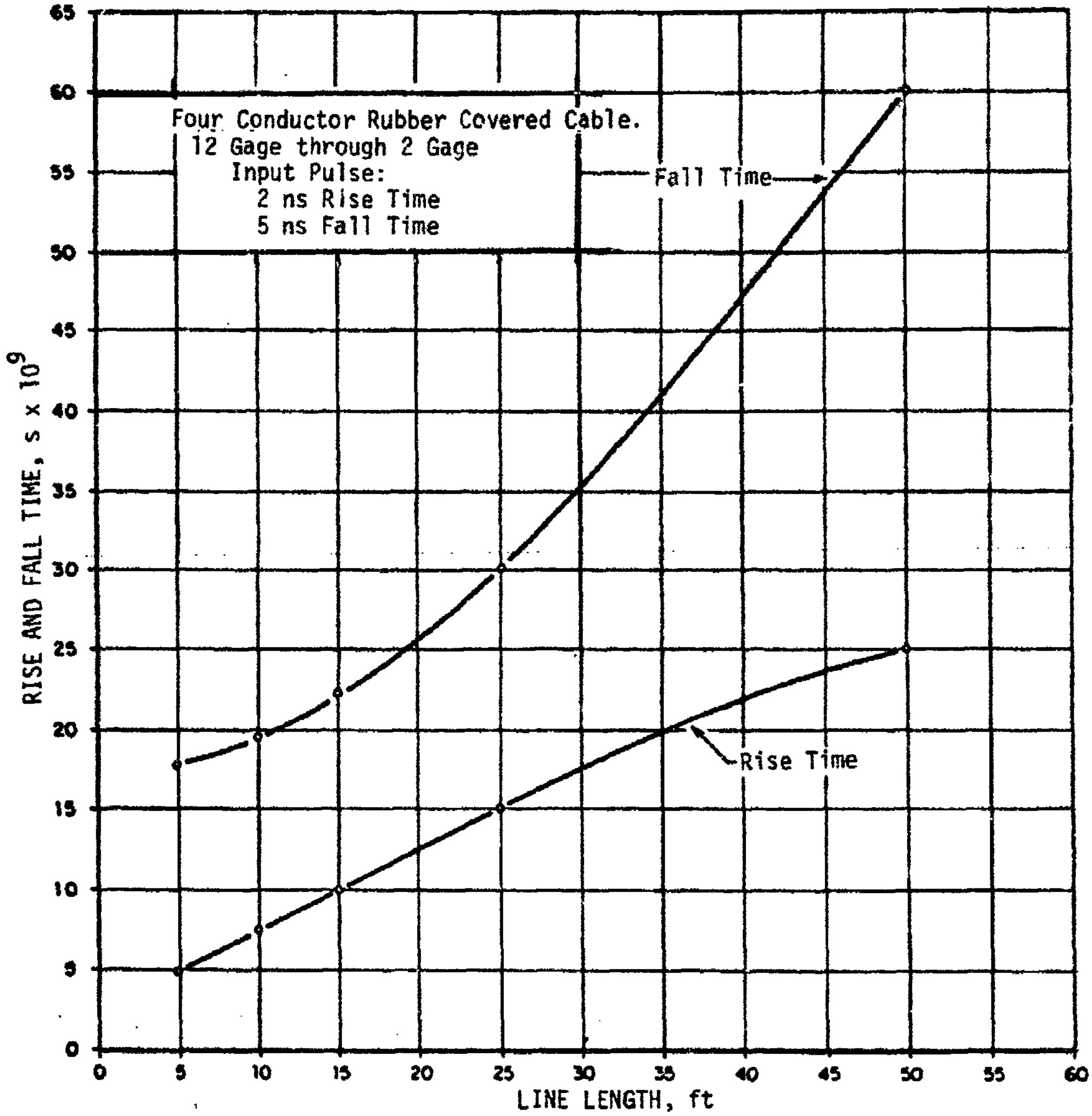


Figure 5-10. Power Cable Degradation of Rise Time and Fall Time

rates are of the order of 1 MHz or greater, depending on the line voltage. Whereas the discharge current from a single corona point is roughly periodic, the total discharge current on a line containing numerous corona spots must be treated as a set of randomly occurring pulses. To this extent, the noise generated by a power line is similar in effect to shot noise.

Corona discharge from points that are at a high positive potential have been found to be somewhat different from those at a negative potential. The potential gradient required for the formation of corona is somewhat higher, and the current is not impulsive in nature. The pulses obtained are of greater amplitude and lower repetition frequency than those for negative corona.

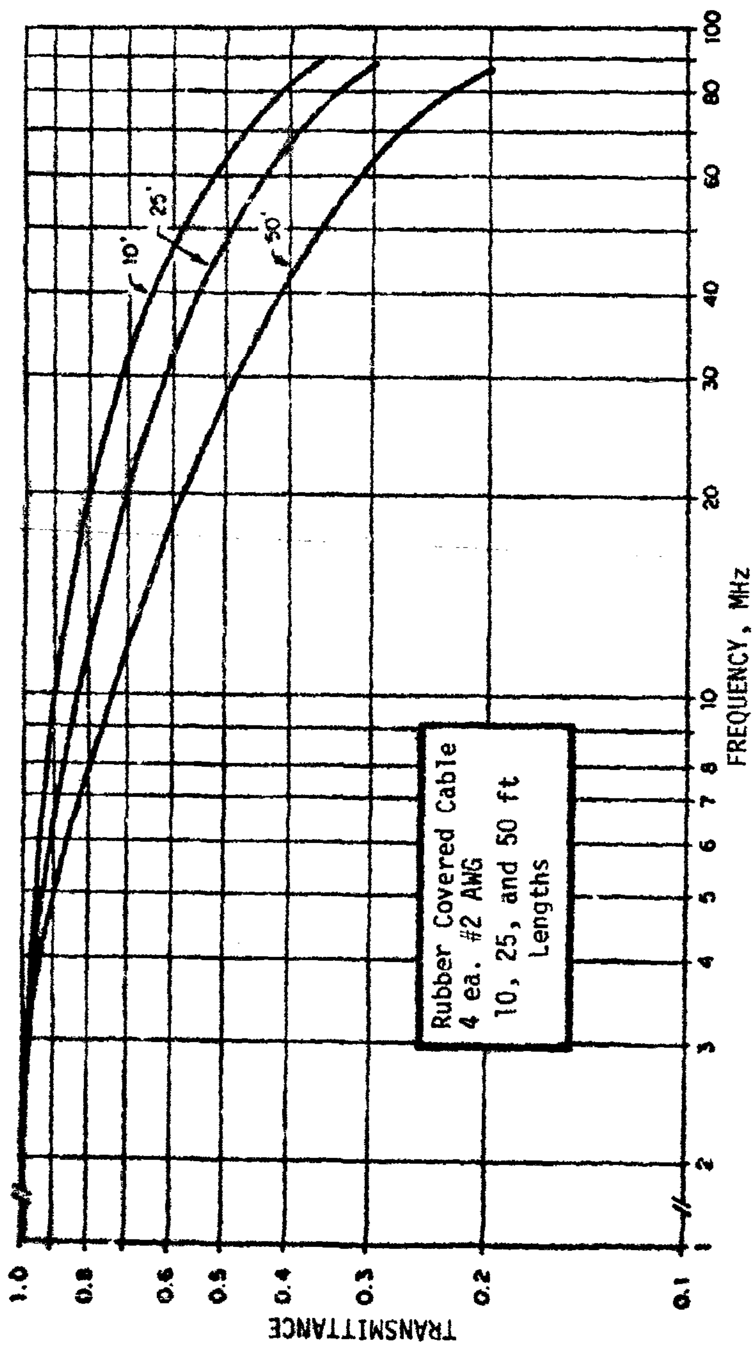


Figure 5-11. Transfer Function-Frequency Domain

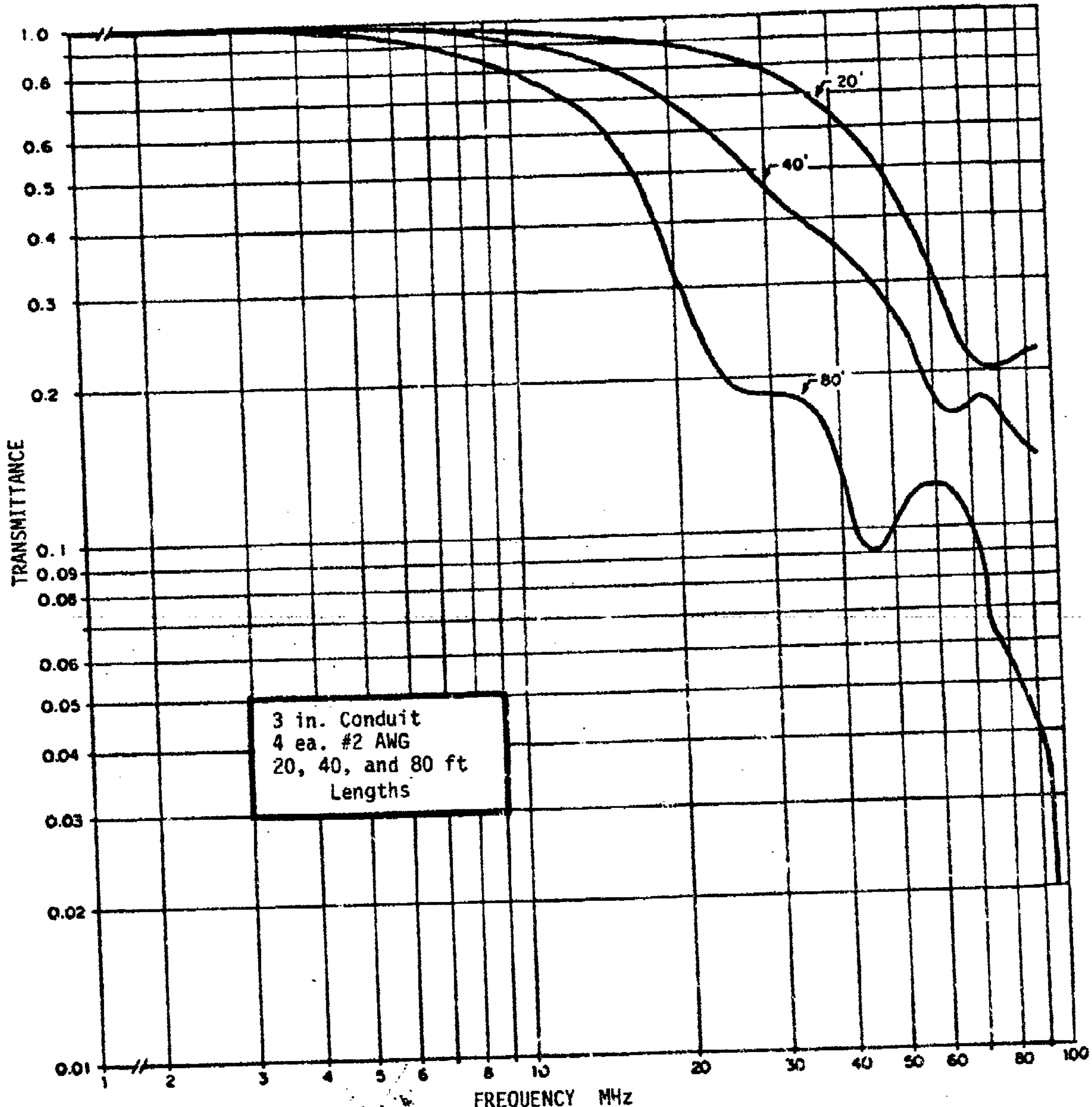


Figure 5-12. Typical Multiwire in Conduit Transfer Function

The initiation of corona occurs at a critical field gradient E_c usually considered to be given by Peek's well-known formula

$$E_c = 31\delta\left(1 + \frac{0.308}{\sqrt{\delta r}}\right), \text{ kV/cm} \quad (5\text{-}1)$$

where

r = conductor radius, cm

δ = relative air density

$= \dfrac{3.92\,p}{273 + t}$

p = atmospheric pressure, cm of Hg

t = temperature, °C

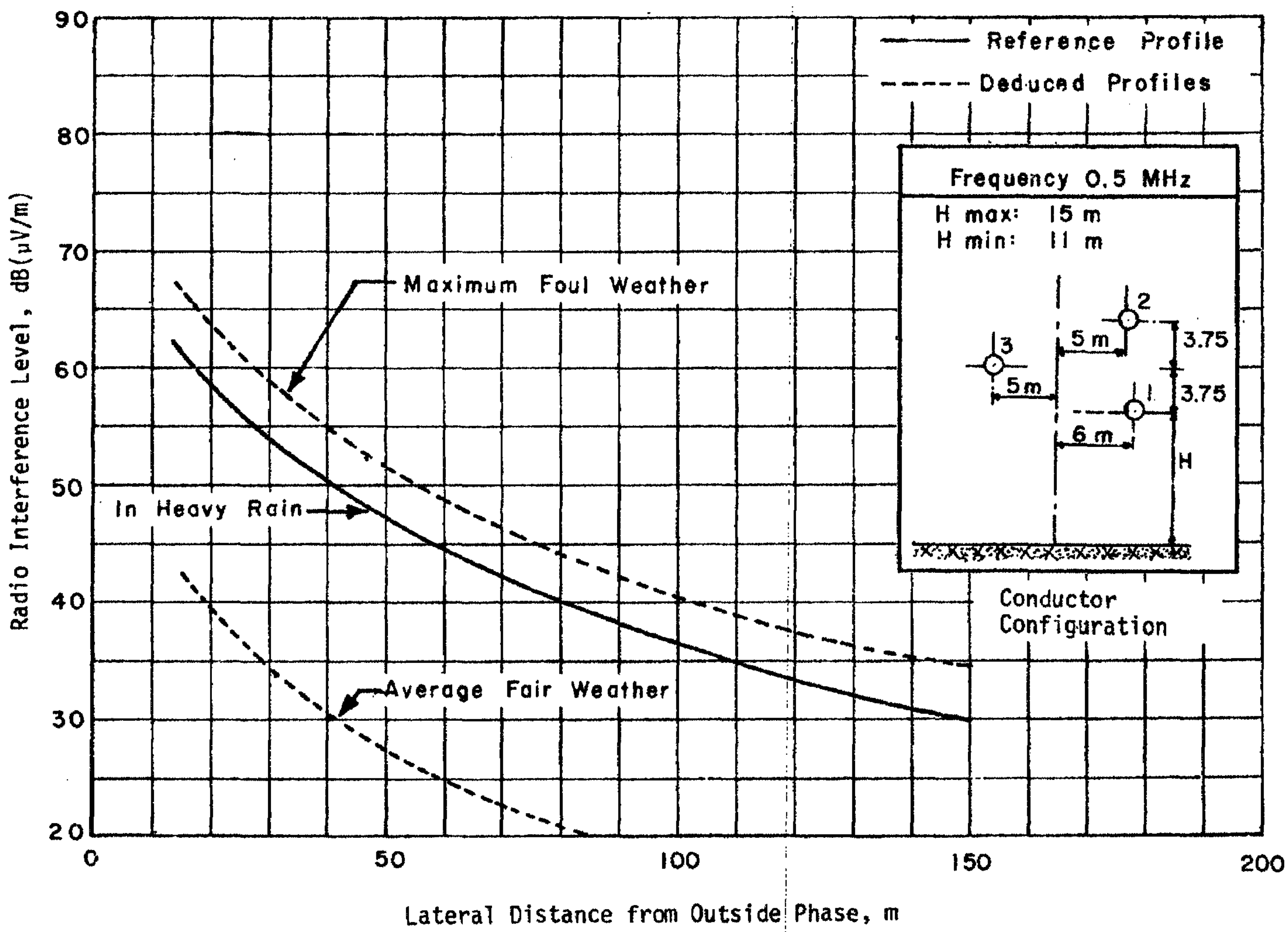


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Figure 5-13. Predicted Radio Interference Profile for 225-kV Power Line

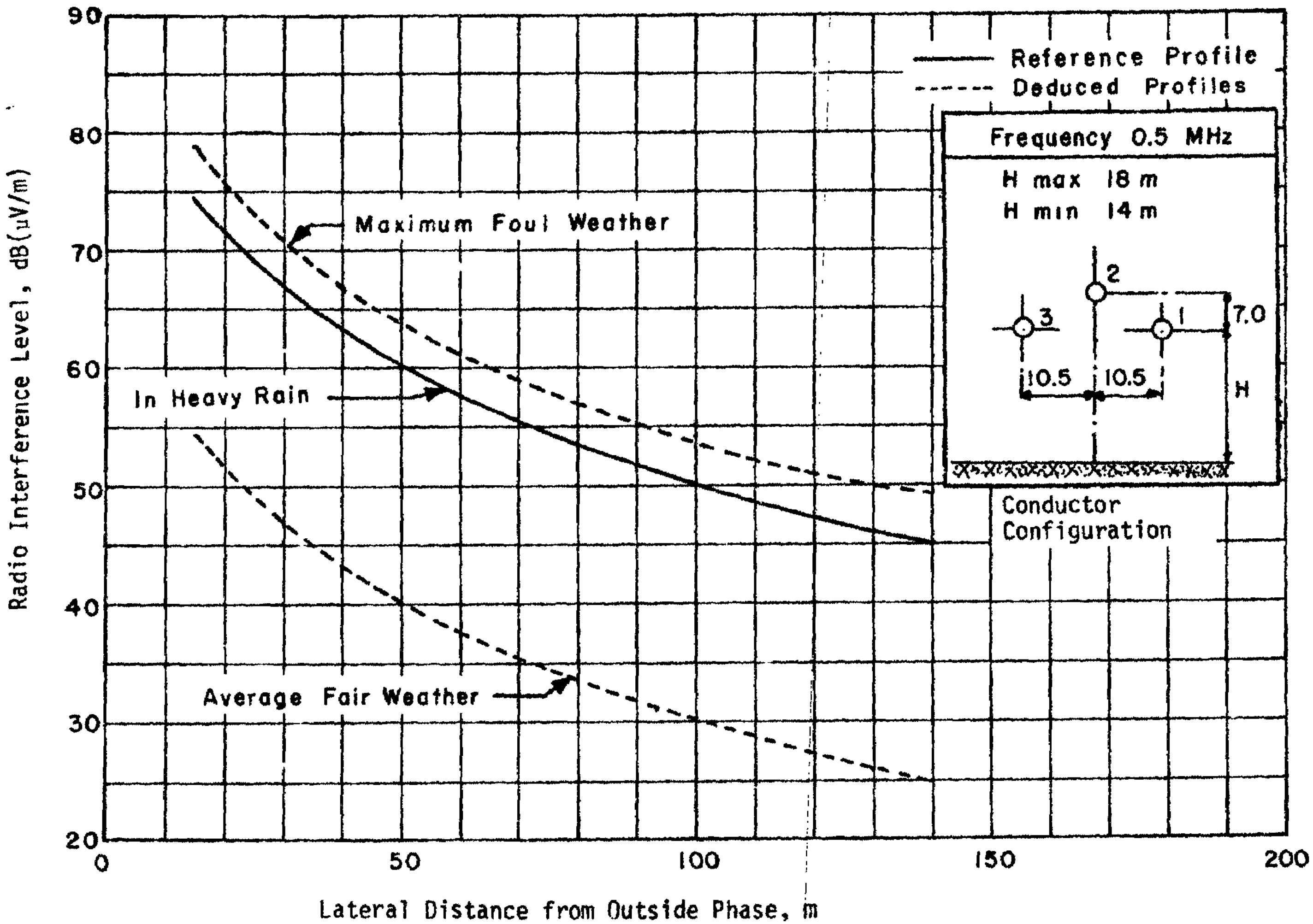

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Figure 5-14. Predicted Radio Interference Profile for 750-kV Power Line

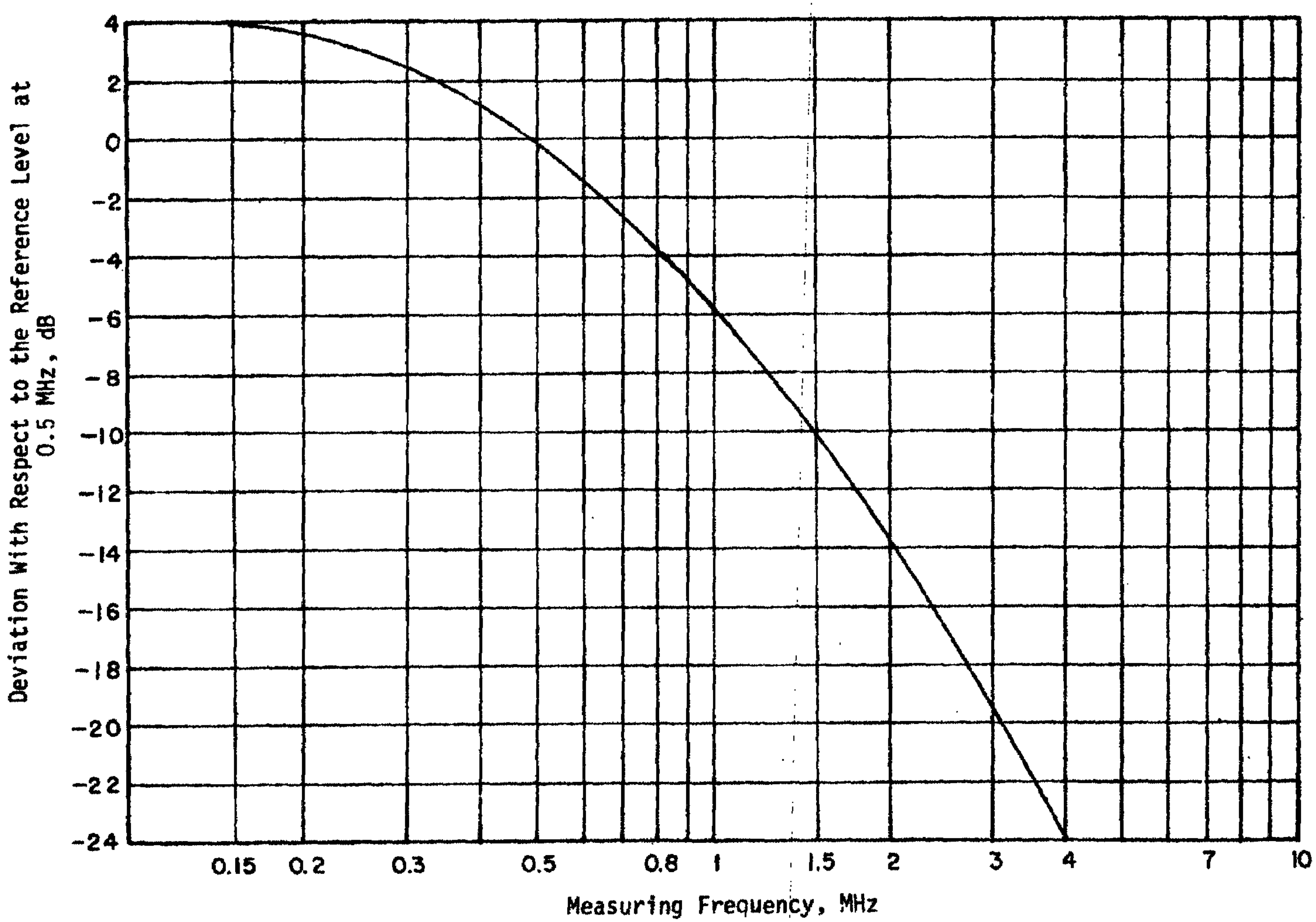

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Figure 5-15. Typical Frequency Spectra of the Interference Fields of High Voltage Power Lines

In actual practice, radio interference occurs at theoretical gradients well below this because the line conductors are far from smooth due to surface irregularities produced in manufacturing and handling and because of contamination and water droplets (in rainy weather). These irregularities cause local gradients above the theoretical values.

Fig. 5-16 shows the variation of relative interference level as a function of the maximum calculated gradient based upon empirical data on actual lines. The relative effect of rain and conductor radius is shown. Most lines operate under conditions between the shaded areas.

5-2.4 GAP TYPE DISCHARGES

These discharges occur as the result of a potential build up between conductors with a sudden discharge as the result of insulation breakdown. Because of the highly impulsive nature of the discharge, it has a frequency spectrum much broader than that of the usual corona type discharge. These kinds of gap-type discharges are:

a. Sparks between two metal parts

b. Microsparks between a metal part and an electrically charged insulating surface

c. Surface discharges on insulators

d. Internal discharges in voids within insulating materials.

Sparks may occur between poorly contacting metallic parts carrying high voltage, e.g.,

a. Between caps and pins in insulator strings

b. At the connections of the strings to the tower or the line-conductor

c. Between spacer and subconductor of a bundle

d. Between parts of a counterweight or vibration damper.

Poor contacts between metallic parts can be caused by corrosion, dust, or other forms of contamination. Dry weather improves the isolation of metallic parts and is therefore favorable for the occurrence of sparks. Movement of an insulator string by the wind sometimes causes a burst of interference, especially with lightly loaded strings.

Sparks caused by electrostatic induction have been observed between a metal plate bearing a warning notice and the steel tower on which it was bolted.

The typical frequency spectrum of a corona current pulse and an example of the spectrum of a gap-type discharge is given in Fig. 5-17.

When an overhead line has been correctly designed to prevent radio interference in the frequency range around 1 MHz, interference problems above 30 MHz will be due to faulty equipment.

5-2.4.1 Propagation of Interference (Ref. 2)

At frequencies above 30 MHz the propagation mechanism of the interference along a high voltage line becomes quite different from that at frequencies around 1 MHz.

The attenuation of the line will rise with frequency, and the length of line — over which the distributed and local noise sources can be assumed to add up — will diminish. The result will be that the contribution of the distributed corona noise sources on the conductors to the total interference field will become less, and the local noise sources in the towers become more important and can be identified.

Whereas, the interference power is transmitted along the line at frequencies around 1 MHz, at the higher frequencies, it is radiated in directions perpendicular to it.

Radiation from a source above the ground and in a direction perpendicular to the line can be described by the addition of a direct wave and of a wave reflected by the ground. Exact calculation of the field strength caused by a particular source is possible but will be complicated. However, qualitative considerations show that the field strength E at a height h and a distance d (both in meters) from the line is mainly governed by a relation of the form:

$$E = \frac{2K}{d} \sin\left(\frac{2\pi hH}{\lambda d}\right), \text{ V/m} \qquad (5\text{-}2)$$

where

H = height of the source, m

λ = wavelength, m

K = a constant for distances farther than $d = \sqrt{H^2 + h^2}$ from the line, whose values depends on the power, the radiation efficiency and the directivity of the source, V

A graph showing a typical relationship between field strength and distance is shown in Fig. 5-18. It can be seen that the field strength passes through a series of minima and maxima as the distance from the line increases.

The values of the maxima in field strength decrease by 6 dB for every doubling of the distance, while the last maximum is found at a distance of $d = 3.1\ Hh/\lambda$. Beyond that distance the field strength decreases rapidly at a rate of 12 dB for every doubling of the distance.

5-2.4.2 Passive Interference (Ref. 3)

In the vicinity of high-voltage line with steel towers, an appreciable distortion of the field of an incident electromagnetic wave can be observed at frequencies of about 1 MHz and higher. The distortion

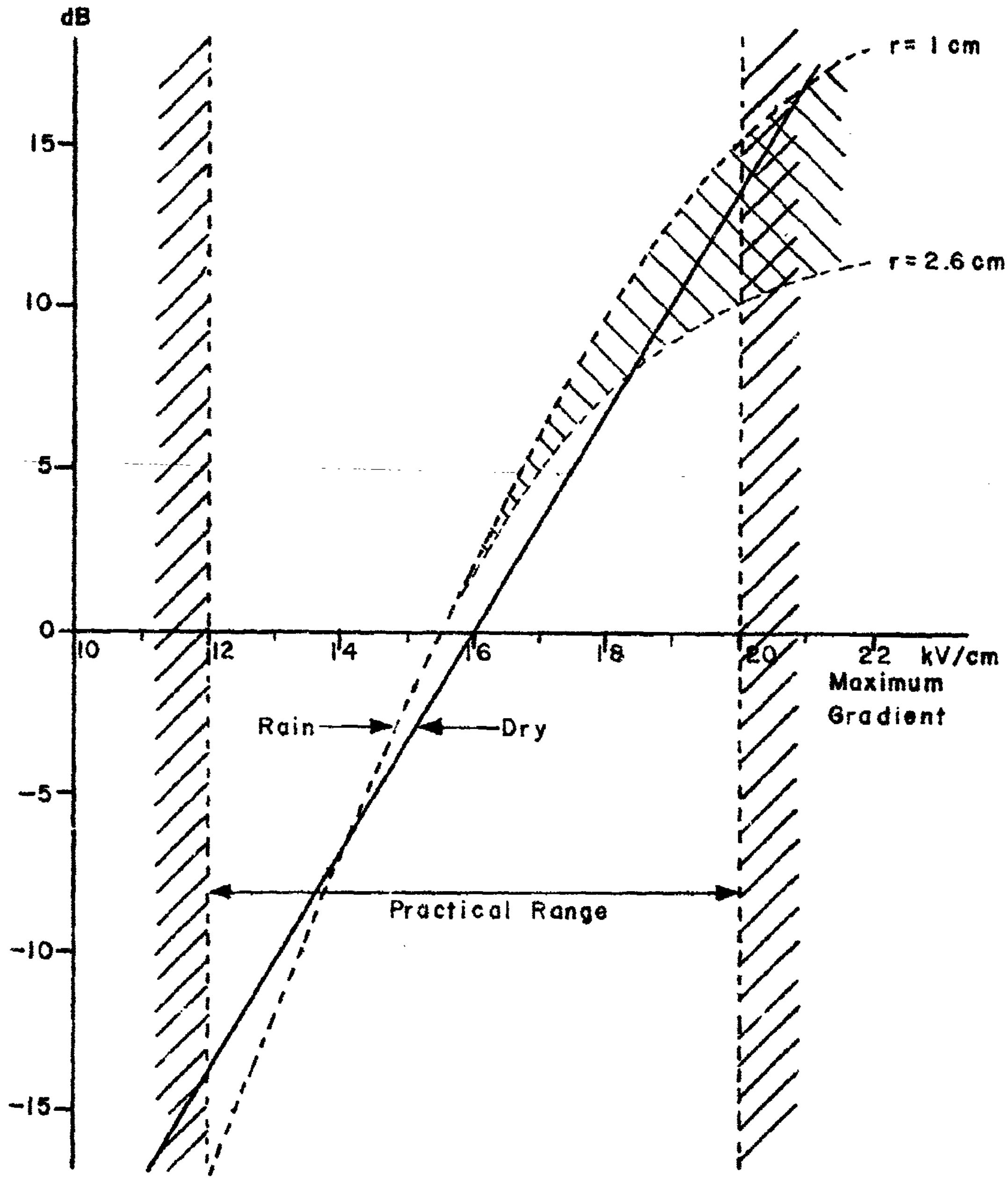


Figure 5-16. Radio Noise Level vs Maximum Gradient

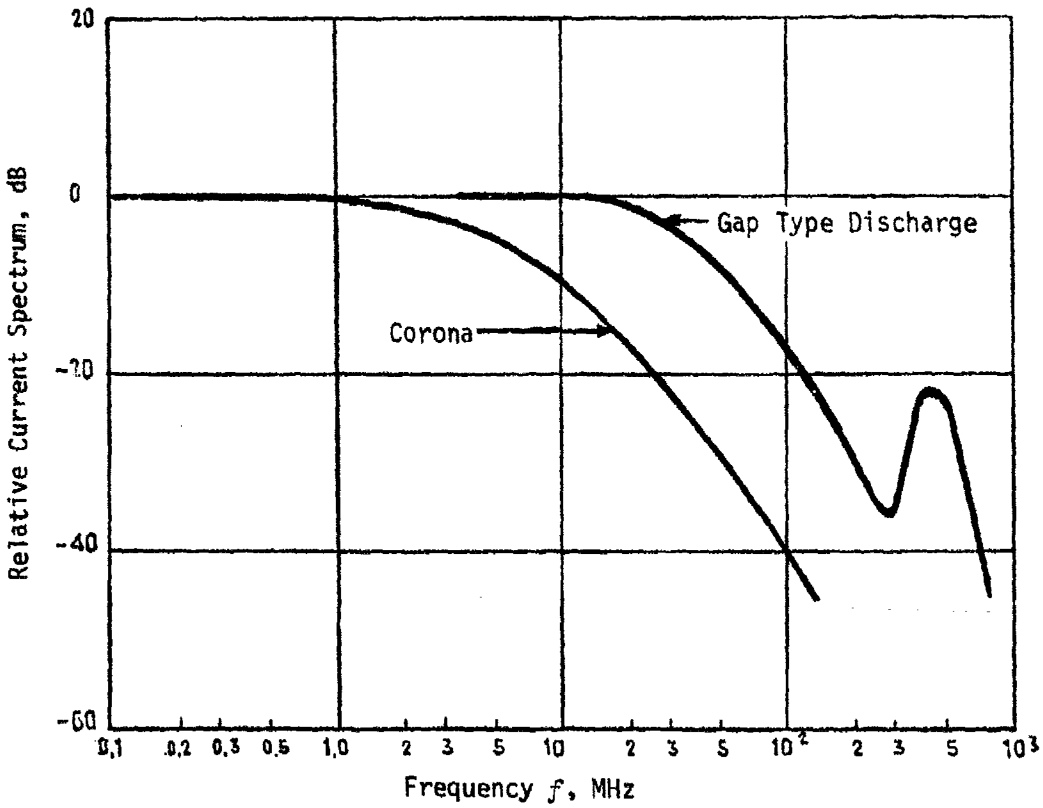


Figure 5-17. Relative Current Spectrum of Corona and of a Gap Discharge

of the field can be described as a suppression of the original incident wave and of waves that are reflected by the conducting parts of the line. This superposition of waves near the line is perceptible in two ways:

a. The incident and reflected waves form an interference pattern of standing waves with maxima and minima which make the total local field strength dependent on the position relative to the line.

b. The signals modulated on the reflected waves have a time delay with respect to the signal modulated on the incident wave, causing envelope distortion.

Experience has shown that degradation from reflections by overhead lines mostly occurs when the receiver is on the transmitter side of the line, and more so when the incident wave is perpendicular to the line.

The reflected wave can be measured separately from the incident wave by means of a directional antenna. The ratio of the incident field strength to the field strength of the reflected wave, expressed in dB, is called the reflection loss.

The level of the reflected wave can be surprisingly high, and in extreme cases it can be even higher than the level of the incident wave. To explain this it should be remembered that, due to its height, the line is in a much more favorable position to pick up the incident wave than the usual receiving antenna.

The reflection loss R at a distance d from a single wire of diameter a (in meters) increases with frequency to a value given by:

$$R = -20 \log \sqrt{\frac{a}{d}}, \text{ dB} \qquad (5\text{-}3)$$

Therefore it can be found that at UHF, the reflections from an overhead line are less than those at VHF.

In case of bundled conductors, the frequency dependence of the reflection loss is more complicated because of the possible reflections between the

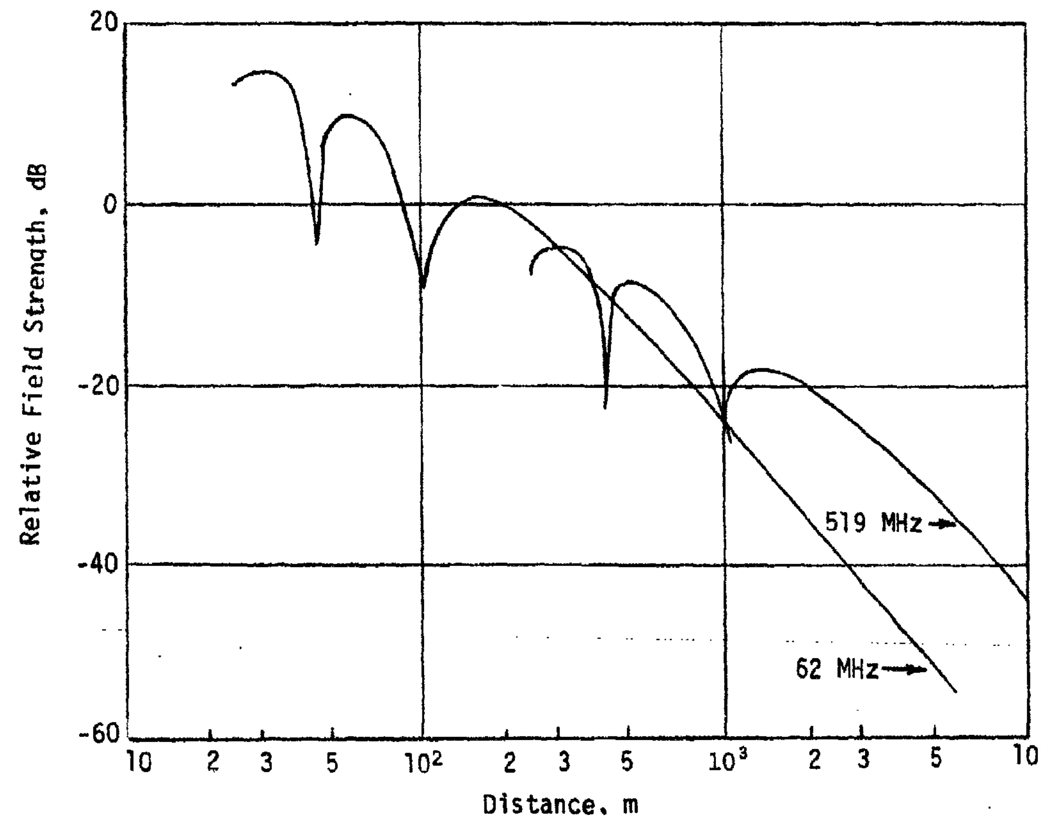


Figure 5-18. Relative Interference Field Strength as a Function of the Distance from the Line

constituting conductors. Fig. 5-19 shows the calculated reflection loss of 3-conductor bundle for horizontally polarized waves as a function of the direction of reflection. At 62 MHz the reflection loss is practically independent of the direction of reflection, whereas at 519 MHz the reflection loss increases markedly depending on the direction of reflection. Additional data are given in Ref. 3.

5-2.5 INTERFERENCE LEVEL AND QUALITY OF RECEPTION (Ref. 2)

Table 5-1 shows typical values of signal-to-noise ratio required for various subjective reception qualities when interference generated by power lines degrades amplitude modulated signals. For other types of modulation, e.g., frequency modulation, a quite different performance may be expected (see par. 3-2.1.4).

5-3 POWER CONTROL

Power flow may be controlled through (1) switching in which currents and voltages are changed suddenly in steps, or (2) continuous variation using amplifiers, phase control devices, or other means.

5-3.1 SWITCH POWER CONTROL

The step action in opening or closing a switch or contactor gives rise to broad spectrum interference, as discussed in par. 3-1.3.2, whose magnitude is dependent on the magnitude of the current or voltage which is initiated or interrupted. Where these quantities are large, relays and remote controlled circuit interruptors may be used.

5-3.1.1 Interference Generation

Switch-generated interference can be associated with three mechanisms of generation: (1) high-voltage gaseous discharges (which may give rise to a

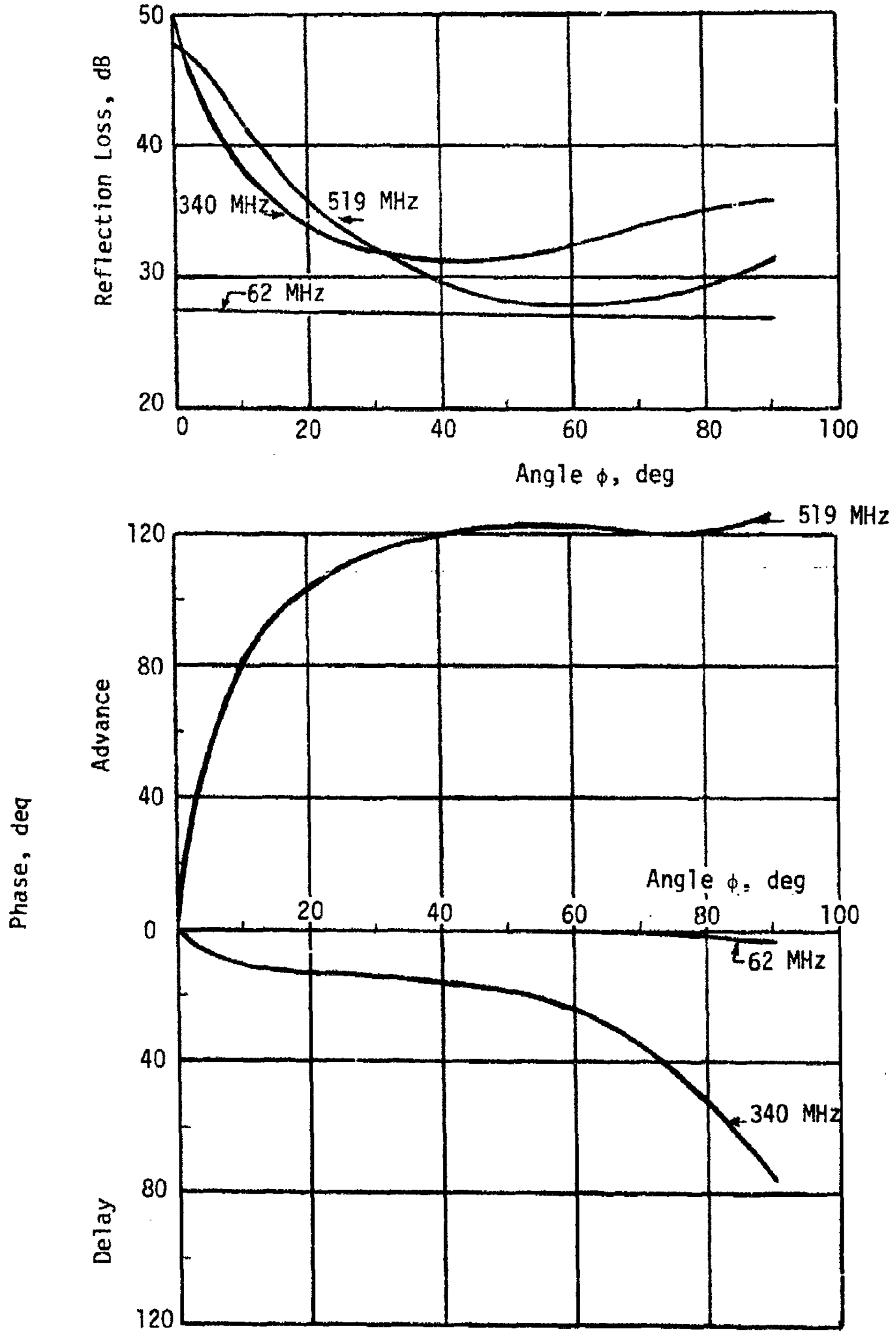


Figure 5-19 Reflection Loss and Phase of a Three-conductor Bundle as a Function of the Direction of Reflection

TABLE 5-1. RECEPTION QUALITY AS A FUNCTION OF SIGNAL-TO-NOISE RATIO

Signal/Noise Ratio Scales		Code	Reception Quality Subjective impression
dB	Ratio		
30	32	5	Interference not audible
24	16	4	Interference just perceptible
18	8	3	Interference audible but speech perfectly received
12	4	2	Unacceptable for music but speech intelligible
6	2	1	Speech understandable only with great concentration
0	1	0	Spoken word unintelligible, noise swamps speech totally

saw-tooth type of waveform), (2) low-voltage high-field breakdown (bridging), and (3) current changes between the two steady-state values. The high-voltage discharge, when present, produces more interference than the other causes combined throughout the entire frequency range from 15 kHz to 1000 MHz. There is a broad peak in the spectral distribution of the interference caused by high voltage breakdown that occurs in the region of a few megahertz for common values of circuit parameters. The bridging has a negligible effect at the lowest frequencies, but increases in relative importance with frequency until, at 1000 MHz, it is nearly as great a source of interference as the high-voltage breakdown. The change of current between the two steady-state values yields its greatest contribution to interference at the low-frequency end of the spectrum (Fig. 5-20).

5-3.1.2 Interference Reduction

In reducing interference, first consideration should be given to elimination of the high-voltage discharge. This is accomplished by employing an interference reduction device that prevents the voltage across the switch gap from exceeding the value required to initiate a glow discharge (approximately 300 V). To eliminate the formation of bridges, it is necessary to prevent the electric field between the contacts from exceeding a critical value (approximately 5×10^6 V/in. for untreated contacts). Bridge elimination may be accomplished upon switch opening by mechanically increasing the speed of separation of the contacts, and by electrically decreasing the rate of build-up of the potential across the contact gap. It is impossible to alter the changing load current between the two steady-state values without affecting normal operation of the circuit; however, interference may be contained effectively in the regions of the switch and load side of the circuit by inserting filters in the external power supply leads. A good interference reduction circuit should:

a. Retard the build-up of voltage across the gap during the initial period of contact separation to minimize bridging reclosures.

b. Limit the peak voltage across the switch gap, upon opening, to eliminate gaseous discharges.

c. Limit the surge of current through the switch, upon opening or closing, to minimize sharp wavefront transients.

In selecting a suitable circuit, it is usually necessary to make a trade-off between interference reduction and other considerations such as the allowable rise or decay time of the load current, ease of installation of the circuit, physical size of the interference reduction components, and adaptability to the power supply. The circuits that are used usually resemble simple types of low-pass filters, i.e., they consist of capacitors placed across the contacts and/or inductors in series with them to limit the rate of rise or fall of current through the external circuit. Resistors are placed at appropriate locations to absorb the energy stored in these elements. The values of these elements must be chosen so that they perform the necessary suppression function without disturbing the control function being performed in any significant way.

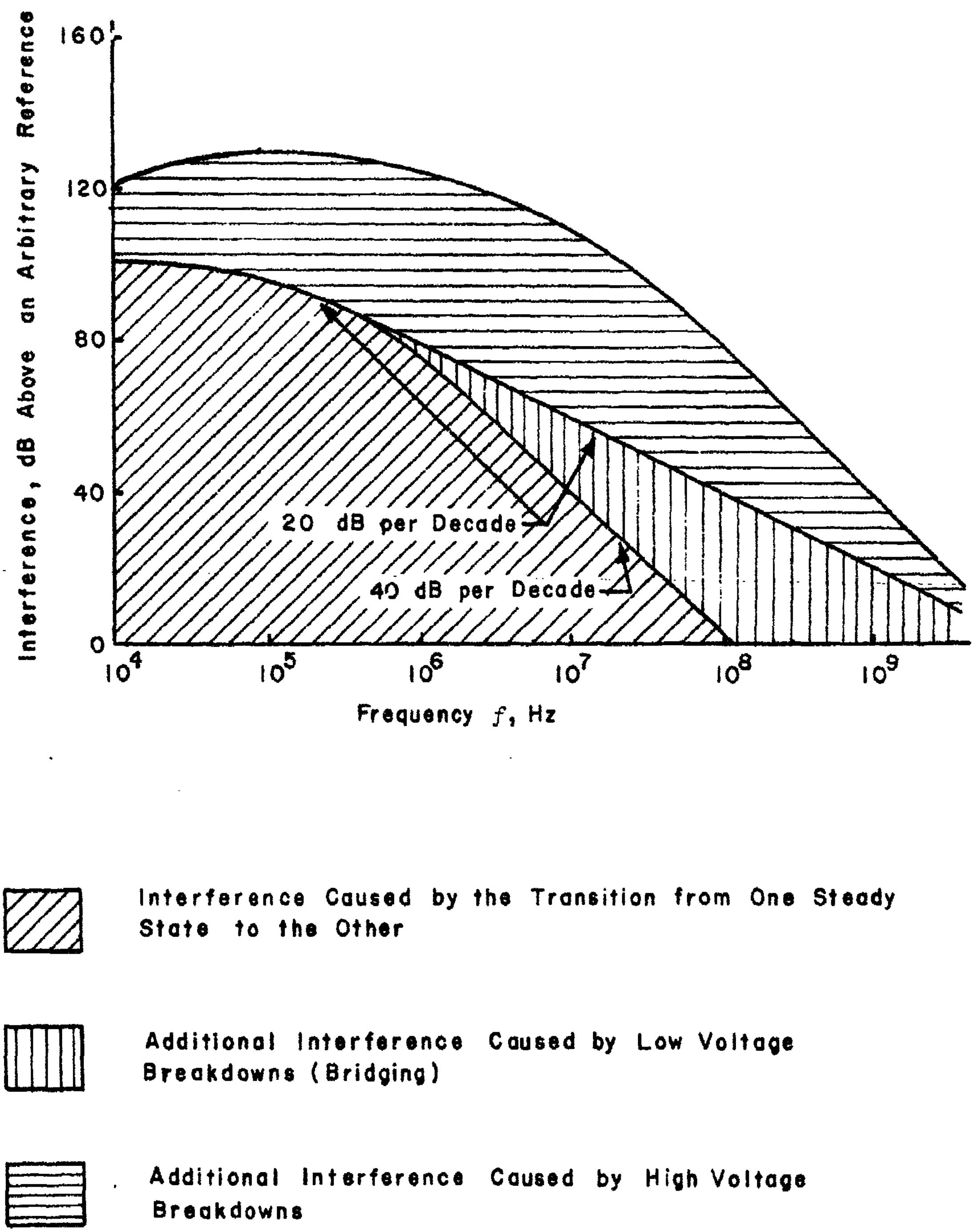


Figure 5-20. Spectra of Switching Interference Phenomena

For example, the value of a capacitor C placed across the switch is determined by the current value and the load inductance L; supply voltage and frequency of switching not being important factors. It is necessary that such a capacitor be able to store temporarily all of the energy previously contained in the load inductance without the voltage across the capacitor exceeding the glow discharge value of approximately 300 V ($C > L\,(1/300)^2$, (from Eq. 3-39).

5-3.1.2.1 Use of Diodes and Varistors

Devices with nonlinear resistance-voltage characteristics, such as diodes and varistors, are useful components for interference reduction. A diode has low forward resistance and high reverse resistance. A varistor conducts well at high voltage but not at low voltage. The function of either a diode or a varistor in an interference reduction application is to provide an alternate path for the induced current to that through the contact.

The diode need dissipate only a fraction of the total energy stored in the load since most of it is dissipated within the load resistance itself.

A discussion follows:

a. Dc Circuits. An effective interference reduction element is the diode shown on Fig. 5-21. The diode is inserted in the circuit so that its polarity opposes that of the impressed voltage. The i^2R_D loss (R_D is the resistance of the diode) through the shunt circuit is small. For maximum interference reduction, the diode should be installed as close to the inductive element as possible. Negligible current flows through the diode under steady-state conditions but, when the switch is opened, the diode provides a low resistance circuit through which the inductive current may flow. The peak voltage appearing across the switch is thereby limited to the sum of the battery voltage and the forward drop of the diode. By choosing a diode with a low forward resistance, the voltage across the switch is essentially just that of the battery. This circuit is very effective in providing the complete elimination of all high-voltage, saw-tooth type discharges; it is ineffective in retarding the gap voltage build-up. A second method of using a diode is shown on Fig. 5-22. This diode should possess a sharp knee in its voltage-current curve (Zener diode) at a voltage value that is somewhat greater than the supply voltage, so that the steady-state current through the load will be essentially zero when the switch is open.

b. Ac Circuits. Components consisting of two similar Zener diodes placed back-to-back are commercially available as a single unit. Their combined current voltage characteristic is shown on Fig. 5-23. When placed across the load, as shown in Fig. 5-24, the back-to-back unit is inferior to the load-shunt diode interference reduction circuit (useful in dc circuits) in that it does not limit the overshoot in the gap voltage to as small a value. Bridging is therefore more prevalent. When the contacts are opened, one diode, depending on the polarity of the source at the instant switching occurs, will limit the peak driving voltage of the coil. The stored energy of the coil is dissipated in the diode and in the coil itself.

c. Varistors. A varistor could be used in place of the diode in Fig. 5-21. However, the circuit shown in Fig. 5-21 results in lower power consumption and heating because it conducts at a lower induced voltage. A varistor can be used in place of the diode in Fig. 5-22, but it may produce higher current drain

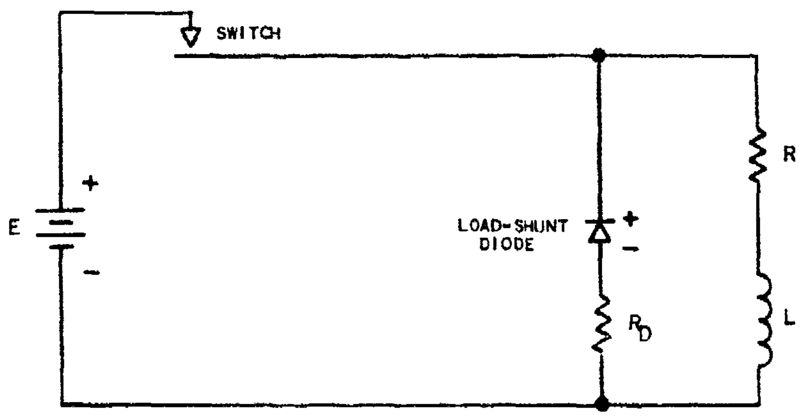


Figure 5-21. DC Switching Interference Reduction by Load-Shunt Diode

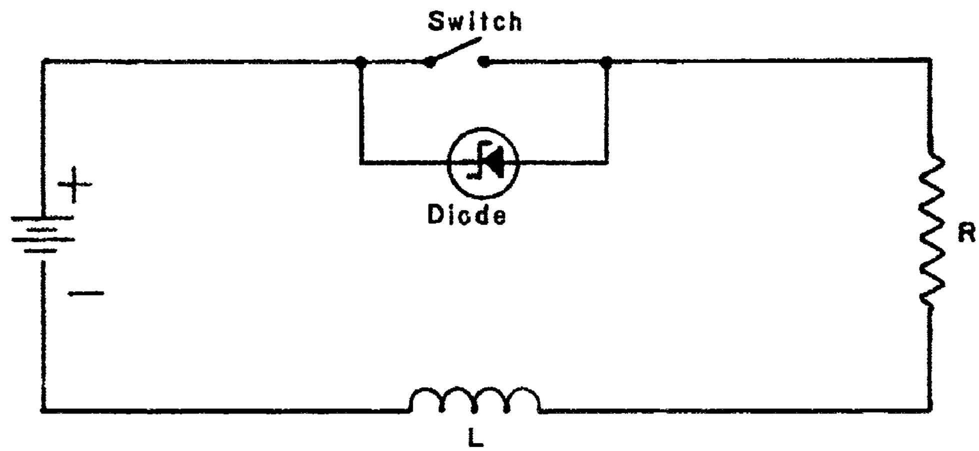


Figure 5-22. DC Switching Interference Reduction by Switch Shunt Diode

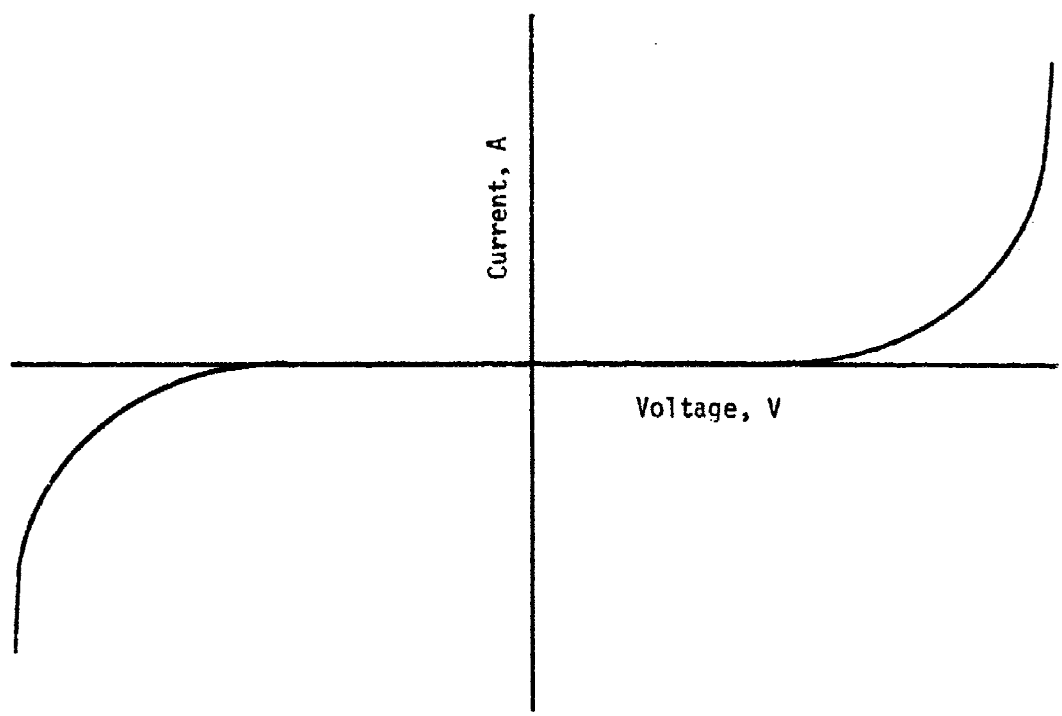


Figure 5-23. Current vs Voltage Characteristics of Two Back-to-Back Zener Diodes

when the switch is off than will the diode. A varistor can be used in place of the back-to-back diodes shown in Fig. 5-24.

d. Application Guidelines. When designing interference reduction circuits using diodes or varistors, the following design guidelines should be considered:

(1) When using a single diode, its current-rating should be equal to the load current when repeated cycling is necessary.

(2) For intermittent use, a rating equal to one-half of the current is usually sufficient.

(3) Breakdown-voltage ratings of diodes should exceed the supply voltage by at least 20%. For selenium diodes, a low safety factor is normally satisfactory.

(4) Back-to-back diodes may be used for ac circuits. The breakdown-voltage must exceed the supply voltage.

(5) When using varistors, the characteristic curve resistance knee must be above the supply voltages, and have adequate heat dissipating capacity.

5-3.1.2.2 Interference Reduction Circuits

A variety of circuits made up of resistors, capacitors, and inductors can be used with or without the nonlinear elements discussed in par. 5-3.1.2.1 as follows:

a. Series Capacitor and Resistor. Contact erosion arises from the use of capacitors and can be alleviated by the addition of a series resistor (Fig. 5-25). In a circuit containing inductance, the voltage that appears across the switch immediately upon opening is equal to the product of this resistance and the interrupted current; therefore this resistance should be low. On the other hand, a large value of resistance is desirable to minimize the contact erosion upon closing. While the *IR* drop does allow some bridging

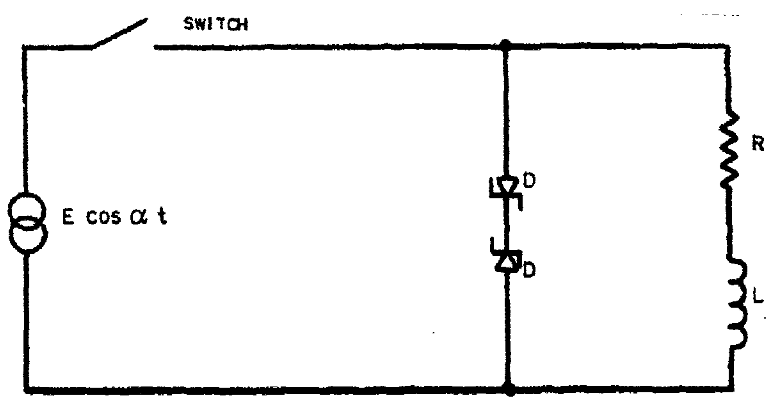


Figure 5-24. AC Switching Interference Reduction by Two Diodes

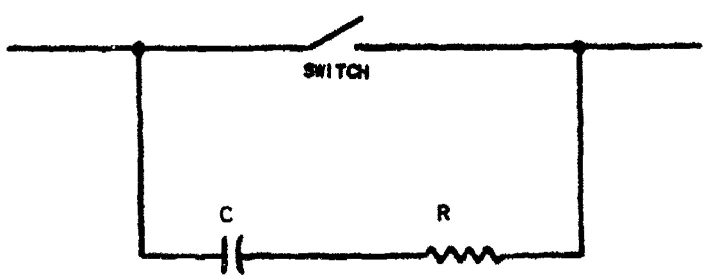


Figure 5-25. Series Capacitor and Resistor Circuit to Reduce Contact Frosion

to occur, a sufficiently large value of capacitance will eliminate high-voltage breakdown. The values of resistance and capacitance should be chosen empirically from the results of test data to minimize interference.

b. Series Capacitor and Nonlinear Resistance. The series capacitor and resistor can be improved by shunting the resistor with a diode, as shown on Fig. 5-26. This action permits the achievement of the desired low-resistance value for a switch opening and the desired high-resistance value for switch closing. For maximum effectiveness, it is necessary that the capacitor discharge completely during the closed interval of the switch, or bridging will occur when the switch is again opened. Therefore, it is necessary to maintain the linear resistance in shunt with the diode to assure that the initial condition is maintained.

c. Typical Resistance-Capacitance Circuit. A very effective circuit consists of a resistor placed in series with the load circuit and a capacitor in parallel with the series combination of resistor and switch (Fig. 5-27). This circuit is similar to the series capacitor-resistor combination of Fig. 5-25. Because the resistor must carry the normal load current with a negligible voltage drop, there is a practical limit to the maximum value of the resistance. In practice, this circuit is much more effective in reducing interference than the series capacitor-resistor unit. Whatever disturbance is produced is largely confined to the switch-resistor capacitor loop. This circuit not only alters the phenomena occurring at the gap, thereby functioning as an interference reducer, but also provides containment for the interference which is produced.

d. Special Interference Reduction Circuits. Special interference reduction circuits are those that deviate from the basic circuit of switch, supply, and load — all connected in series. Several types of reduction circuits are presented:

(1) Coupled Coils. The effect of inductance in series with the switch can be reduced by providing an alternate closed path through which the collapsing

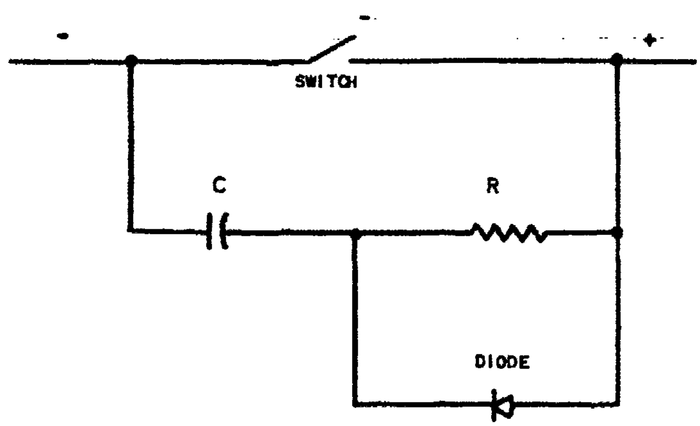


Figure 5-26. Interference Reduction by Series Capacitor and Nonlinear Resistance

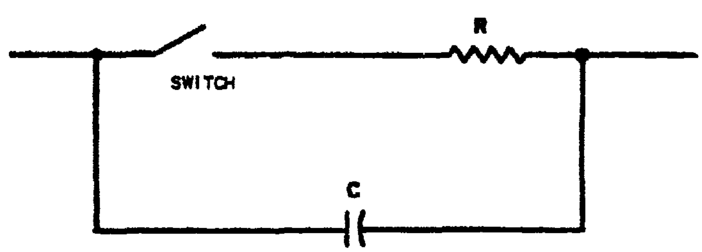


Figure 5-27. Interference Reduction by Series Resistance and Parallel Capacitance

magnetic field can cause the current to circulate (Fig. 5-28). In such a circuit, the higher the value of the coefficient of coupling k the more rapidly is the stored energy absorbed. A transient analysis of this circuit shows the voltage v_1 across the primary winding to be given by:

$$v_1(t) = -R(n_1 n_2)^2 i_1(0^-) \exp(-Rt/L_2), \text{V} \qquad (5\text{-}5)$$

where

$i_1(0^-)$ = the value of the interrupted supply current, A

n_1 = number of turns, primary, dimensionless

n_2 = number of turns, secondary, dimensionless

t = time, s

$$R = R_1 + \left(\frac{n_1}{n_2}\right)^2 R_2$$

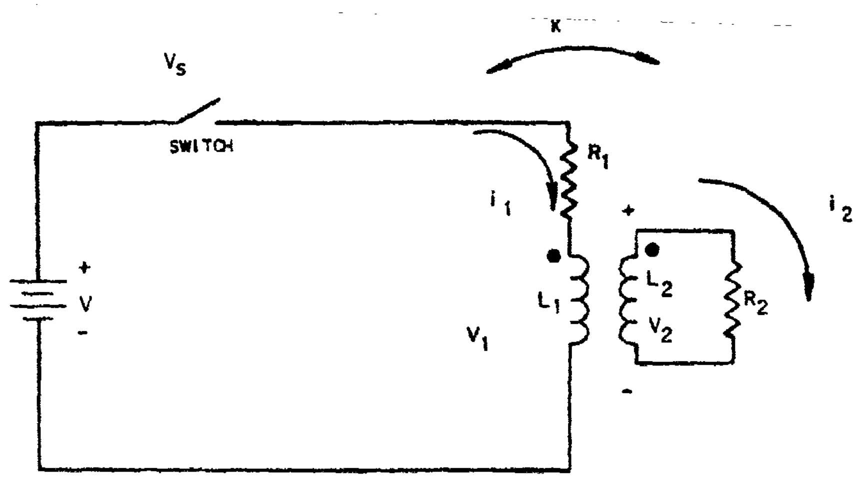
R_1 = resistance of primary coil, Ω

R_2 = resistance of secondary coil, Ω

The corresponding expression for the current i_2 in the secondary loop is given by:

$$i_2(t) = (n_1 n_2) i_1(0^-) \exp(-Rt/L_2) \qquad (5\text{-}6)$$

Eq. 5-5 shows that, if the factor $R_2(n_1/n_2)^2$ is made to approach zero (either through the choice of secondary resistance or of turns ratio), the voltage across the primary winding will approach zero, and the voltage across the switch will be that of the supply. This circuit exhibits the characteristic of a resistance load. When the switch is closed, the current in the primary circuit rises more rapidly as the term $R_2(n_1/n_2)^2$ is decreased. This requirement, however, is in direct opposition to the requirements for maximum interference reduction upon opening of the switch. Nevertheless, it should be possible to select a compromise value for this term which would make the opening and closing interference levels equal. Another possibility is to replace the resistor R by a diode as shown



R_1 = RESISTANCE OF RELAY COIL, WHICH IS THE LOAD OF SWITCH S

L_1 = INDUCTANCE OF RELAY COIL, WHICH IS THE LOAD OF SWITCH S

R_2 = RESISTANCE OF COUPLING COIL

L_2 = INDUCTANCE OF COUPLING COIL

Figure 5-28. Interference Reduction by Coupled Coils

on Fig. 5-29 to present a low-resistance value to the opening transient current and high resistance to the closing transient current.

(2) Bias Batteries. In dc inductive circuits, reduction of arcing at contacts can be achieved by employing bias batteries, or their equivalent, in addition to nonlinear resistances or diodes. The primary advantage of these circuits is that the diodes need to dissipate only an incidental portion of the inductively-stored energy; the diodes may therefore have low electrical ratings and be of small physical size. Six variations of this type of circuit are shown on Fig. 5-30. Circuits (A) and (B) are essentially the same — one employs a tapped inductive load and the other a load with a closely coupled second winding. The dc source in each case also serves as the bias battery. Circuit (C) employs a simple load winding but requires a separate bias battery. Circuit (D) is similar to (C), except that the tapped winding allows greater freedom in selection of the bias voltage. Circuit (E) is applicable when the switching occurs at a reasonably rapid and constant rate so that the repeated transient currents through the rectifier to build up a nearly constant bias voltage across the parallel resistor-capacitor combination and eliminate the need for a separate battery. Circuit (F) is adapted for loads that are supplied from an ac source.

(3) Composite Interference Reduction Components. Interference reduction can be obtained by using simple components in combination. Two composite arrangements are shown on Fig. 5-31. As each of the simple suppressors has some deficiency, it may be desirable to consider whether, when used together, each can supplement the other to improve overall performance. Frequently, however, it is easier to use an interference filter instead of a composite reduction circuit.

As mentioned in par. 5-3.1, to prevent the voltage across the switch from having an excessive peak value during opening of the contacts, the capacitor may have to be large. The required value of capacitance can be calculated from the circuit parameters. If, for instance, the load current is 3 A, the load inductance is 10 H, and 300 V is the maximum allowable voltage, then the calculated value of capacitance is:

$$C = L(I/V)^2 = 10\,(3/300)^2 = 10^{-3}\ \text{F} = 1000\ \mu\text{F} \tag{5-7}$$

The value of the capacitance C on Fig. 5-31 is also a factor in determining the low-voltage breakdown or bridging. For the larger capacitance values, the voltage build-up across the contacts upon switch opening is slower and, consequently, the electric field intensity within the gap is reduced. This consideration is a second factor in the selection of the value of this series capacitor. Under some circuit conditions, and with a fast opening switch, the value so determined will be less than the capacitance required to limit the peak switch voltage. Composite Circuit (B) is a modified configuration of Circuit (A).

5-3.1.2.3 Summary

Table 5-2 compares the ability of several circuits to retard the build-up of gap voltage, limit the peak gap voltage, and minimize sharp wave-front transients. Table 5-3 compares the general interference reduction characteristics of components. The factors affecting the determination of component values are summarized in Table 5-4.

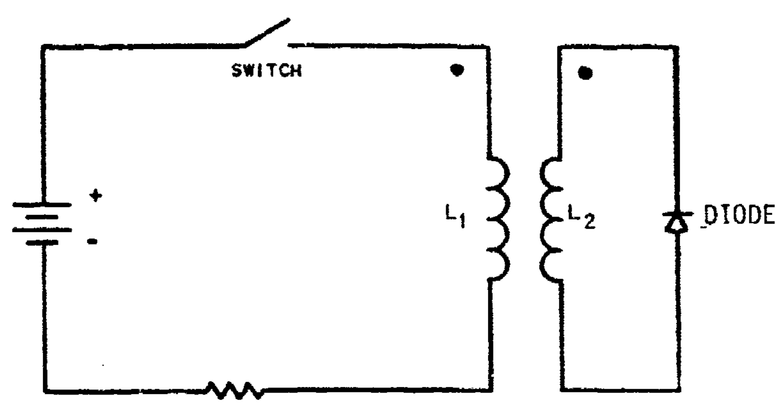


Figure 5-29. Interference Reduction by Coupled Coil With Diode

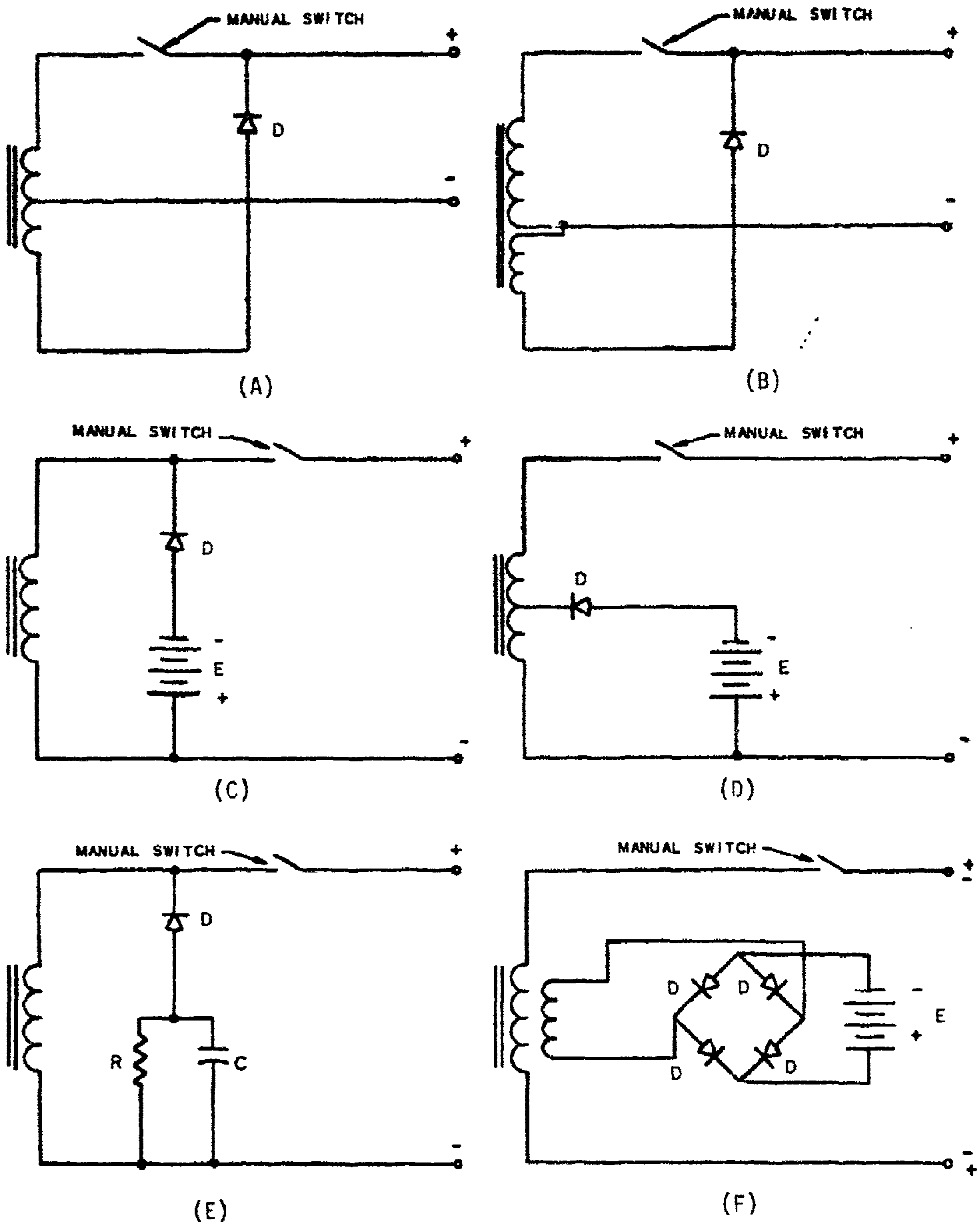
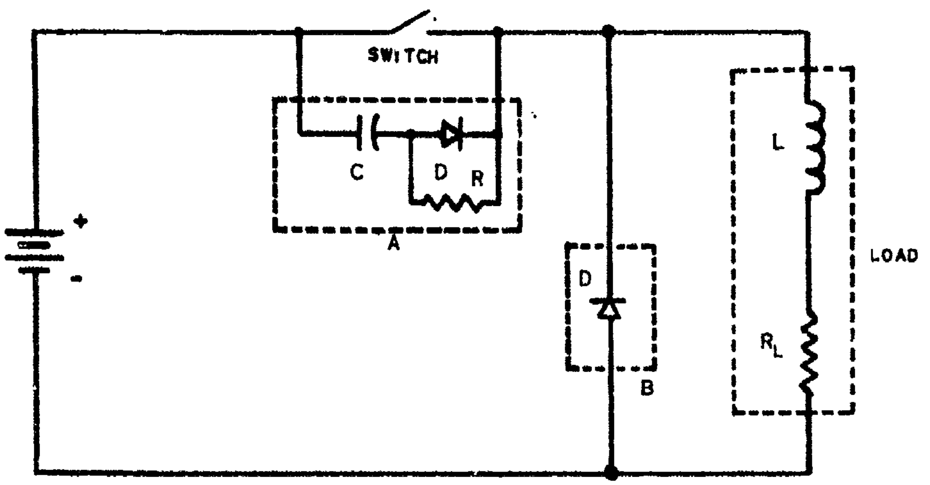
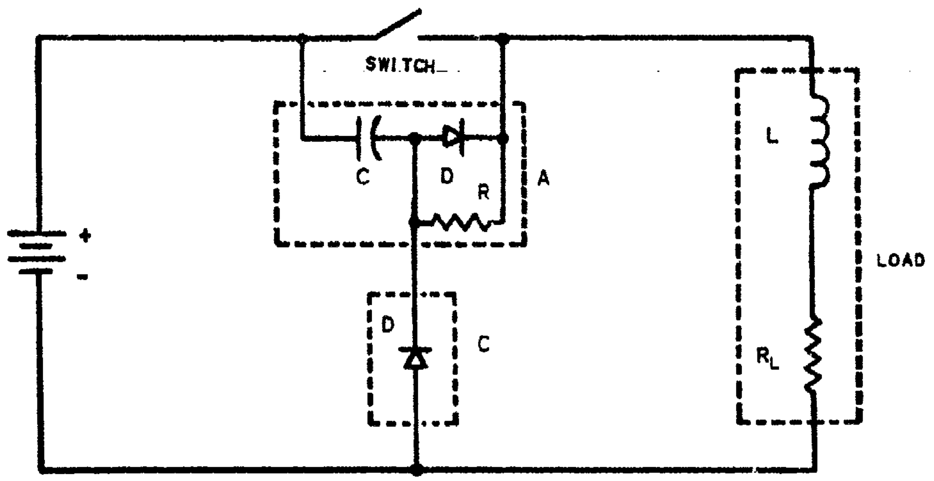


Figure 5-30. Interference Reduction Circuits Employing Bias Batteries



(A) COMPOSITE CIRCUIT A



(B) COMPOSITE CIRCUIT B

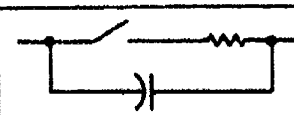
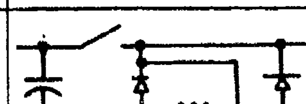
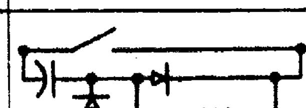
Figure 5-31. Composite Interference Reduction Components

5-3.2 CONTINUOUS POWER CONTROL

The use of simple on-off type controls is not practical for controlling large amounts of power. Furthermore, for many purposes, such as control of traction equipment or in industrial type processes, continuous control of power output is required. This can be achieved with special electromechanical equipment such as motor generator sets and amplidynes. The interference generation characteristics of these types of equipments were discussed in par. 5-1.

The development of semiconductor devices has resulted in extensive application of them in recent years in power control devices. They can be applied in various ways with ac power circuits. Typical techniques are shown in Fig. 5-32 and include, among others, phase angle control, symmetrical phase angle

TABLE 5-2. COMPARISON OF INTERFERENCE REDUCTION COMPONENTS

Components	Placement	Requirements: 1 Retard Build-up of Gap Voltage	Requirements: 2 Limit Peak of Gap Voltage	Requirements: 3 Minimize Sharp Wave Front Transients
Capacitor	Load	G	A[c]	P
	Switch	G	A[c]	P
Linear Resistor	Load	P	A[d]	A
	Switch	P	A[d]	G
Semiconductor Diode	Load	P	G	A[b]
	Switch[a]	P	G	
Back-to-back Diodes	Load	P	G	A[b]
	Switch[a]	P	G	A[b]
Capacitor and Diode 	Load	Capacitor is superfluous		
	Switch	G	A[c]	G
Series R Shunt C		A[d]	A[c]	G
Coupled Secondary		P	G[d]	A
Diode and Battery 	Load	P	G	A[b]
	Switch	P	G	G
Composite Circuit		G	G	A
Composite Circuit		G	G	G

G = Good A = Intermediate P = Poor
[a] = Diode must have knee at voltage greater than that of supply
[b] = Determined by inherent shunt capacitance of diode
[c] = Capacitance must be sufficiently large
[d] = Resistance must be sufficiently small

TABLE 5-3 GENERAL INTERFERENCE REDUCTION CHARACTERISTICS OF COMPONENTS

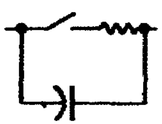
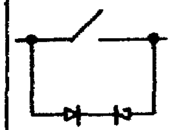
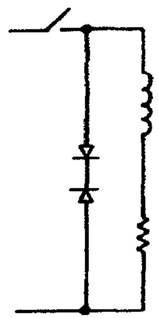
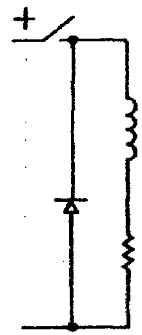
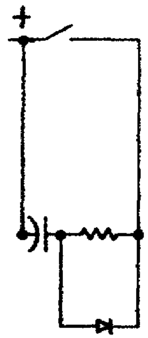
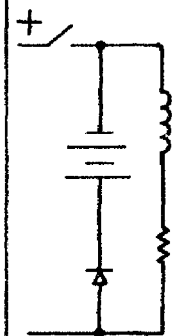
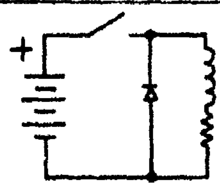
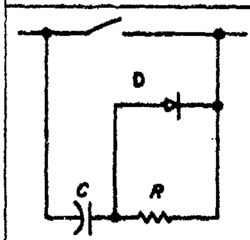
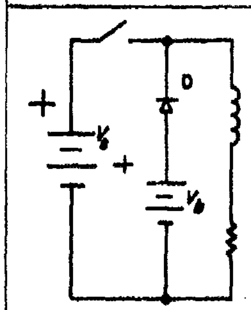
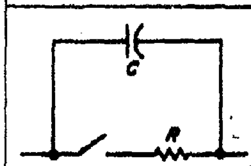
Suppressor Feature	Circuit					
						
Maximum Interference Reduction	Best	Good	Good to Poor	Poor	Good	Poor
Minimum Decay Time	Poor	Good	Good	Very Good	Good	Best
Ease of Adding to Circuit	Fair	Best	Best	Good	Fair	Poor (Need for Battery)
Size of Components For Small $LI^2/2$	Small	Small	Small	Small	Small	Depends on Bias Battery
Size of Components for Large $LI^2/2$	Large	Moderate	Moderate	Small	Large	

TABLE 5-4. DETERMINATION OF COMPONENT VALUES

Suppressor Circuit		Supply Voltage V_s	Load Current I	Load Inductance L
	Across Switch	Voltage for reverse knee must exceed supply	Reversed diode must dissipate all of stored energy. Select on heat basis. If duty cycle is high, current rating may be less than normal rectifier rating. Knee < 300 V.	
	Across Load	Same	Same	Knee < (300 − V_s)
		Same	Diode must dissipate only a fraction of stored energy. Diode with rectifier, average-current rating equal to load current is conservative. If duty cycle is low, normal rating of diode may be increased by a factor up to ten.	
	D	No effect	Should have low forward resistance. $R_f \le V_s/I$	In practice, no effect. Diode, selected on rf basis, could carry load current indefinitely.
	C	No effect	$C \ge L(I/300)^2$ to eliminate sawteeth	
	R	No effect	Must completely discharge C during switch-closed interval. $R = T^*/(5C)$	
	D	Voltage for reverse knee must exceed: $V_b + V_s$	R_f must be less than: $\frac{300 - V_b - V_s}{I}$	If V is large (to reduce decay time), diode selected on basis of rf could carry load current indefinitely. No effect of L.
	V	V_b should be maximum possible and still keep gap voltage less than 300 V: $V_s + V_b + IR_f < 300$ V		
	C	No effect	$C \ge L\,(I/300)^2$ to eliminate sawteeth	
	R	No effect	R = value which produces maximum allowable voltage regulation	No effect

*T = duration of closed interval

control, burst control, and symmetrical burst control. It should be noted that methods of power conversion from ac or dc sources, using rectifiers and inversion techniques, are discussed in par. 5-5.

Because control methods using semiconductor elements generally produce waveform distortion, they automatically create significant levels of harmonic or subharmonic voltages and currents in circuits coupled to them directly by conduction, or indirectly by induction. Furthermore, these generally undesired currents can result in overheating of some of the circuit components.

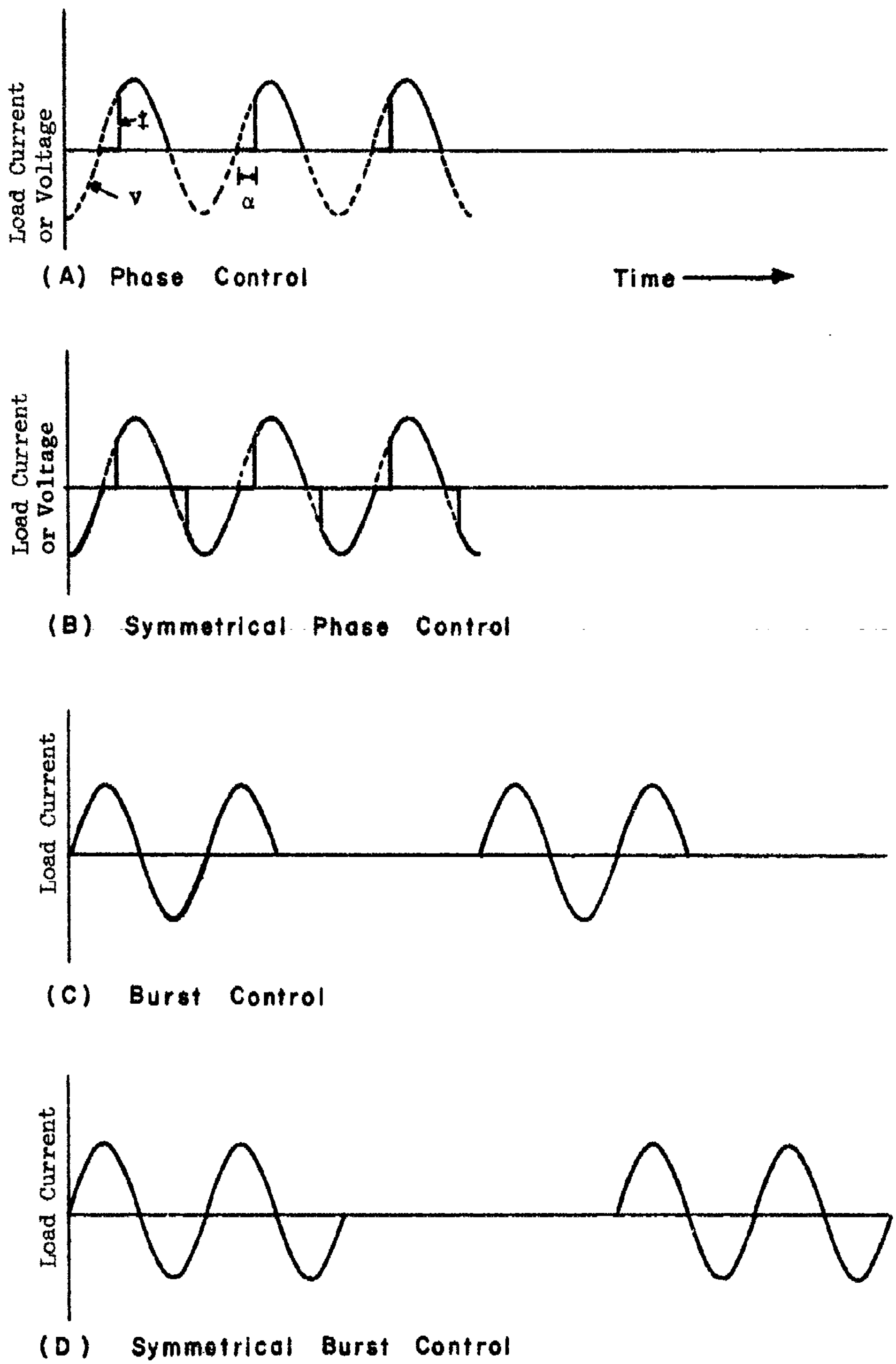


Figure 5-32. Typical Power Control Waveforms

5-3.2.1 Distortion Levels (see Ref. 4)

The burst firing control technique will produce periodic voltage fluctuations on the power line at the burst frequency which in turn can cause a visual flicker effect from lamps connected to the line. The subjective effect of flicker is shown on Fig. 5-33. It is seen that 2% or less voltage fluctuations can cause objectionable flicker. The use of higher power frequencies, such as 400 Hz, may be more satisfactory because the flicker rate is likely to be above the frequencies at which vision is sensitive.

The voltage harmonic levels can be computed from a knowledge of the current harmonic levels and the impedance of the power source. The equivalent circuit for a typical source in the audio-frequency range is an inductance and a resistance in series as shown in Fig. 5-34. In this circuit V represents the generated electromotive force producing the power frequency (60 Hz) line voltage. L is an inductance, and R is the dc resistance which have values dependent on the size of the power source and interconnecting power line conductors.

Values of L and R can vary from 50 μH and 0.01 Ω, respectively, for circuits capable of carrying several hundred amperes to values as much as 10 times these for a typical low current household distribution circuit (15 or 20 A).

The current i shown on Fig. 5-34 represents harmonic currents generated in nonlinear loads which can be expected to flow back through L and R (rather than through the other loads since the equivalent impedance of the connected loads is usually higher than the impedance of the source) and thereby produce the harmonic voltages appearing as distortion on the line. On a 60-Hz system for a current of i_n amperes at the nth harmonic of 60 Hz, the voltage v_n would be

$$v_n = i_n \sqrt{R^2 + (2\pi 60 Ln)^2}\ , \text{V} \tag{5-8}$$

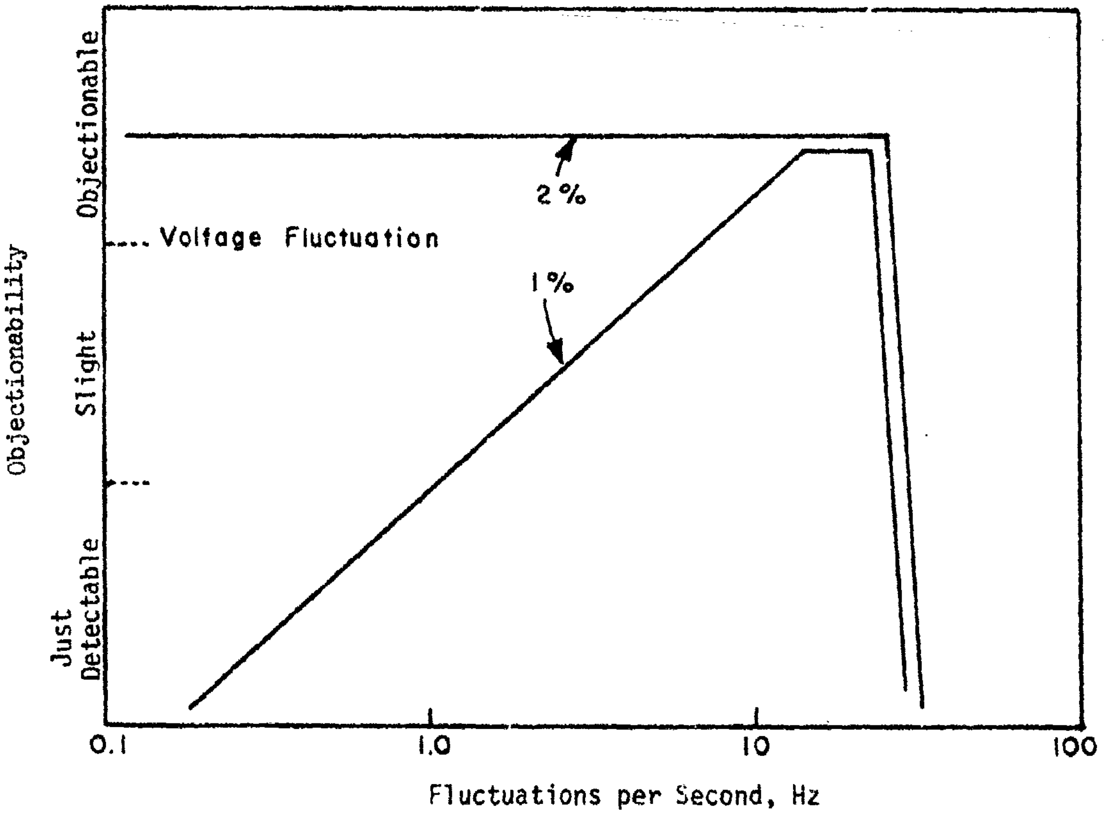


Figure 5-33. Objectionability of Lamp Flicker as a Function of Frequency

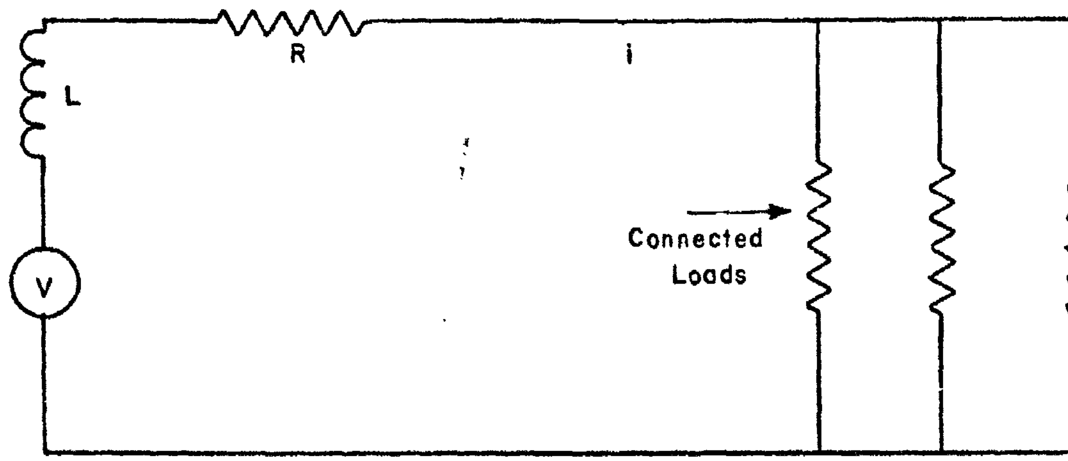


Figure 5-34. Typical Equivalent Circuit for a Power Source at Low Frequencies

For an unsymmetrical phase control arrangement, Fig. 5-35 shows the variation of the 3rd, 4th, and 5th harmonic levels in terms of the peak current as the firing angle (shown as α of Fig. 5-32) is varied. Fig. 5-36 shows the corresponding result for a symmetrical arrangement.

In addition to flicker, high harmonic content can cause other compatibility problems at power frequencies. Examples are:

a. Metering. The flow of direct current and of harmonic currents through a watt-hour meter can cause it to show incorrectly the watt-hours consumed. Errors up to 3% have been observed.

b. Power Factor. The power factor will be affected because of phase shifts between the supply frequency current and voltage. A phase controlled circuit with a resistance load will have a lagging power factor. For high firing angle (low loading), the power factor approaches zero. Where low-pass type filters are used in conjunction with the device to reduce line harmonic content, the capacitance inserted across the line may tend to improve the power factor, but also may cause it to change to leading.

5-3.2.2 Radio Interference Generation

The sharp rise in current with phase control of semiconductor devices can produce much higher values of radio interference than gas tube vacuum circuits. Estimates of the spectrum level at the point of the device in the circuit can be made using the formulas in par. 3-1.3.2. The levels produced can be reduced using appropriate filters as discussed in pars. 4-5 and 5-1. Levels of conducted interference are shown on Fig. 5-37 for a suppressed and unsuppressed 600-W lamp dimmer (suppression consisted of an L-C filter).

The types of filter suggested for use with SCR's are shown in Figs. 5-38 and 5-39. The circuit in Fig. 5-38 confines the high-frequency current to flow in the circuit close to the SCR, whereas in the circuit in Fig. 5-39, some of the current flows in the load. Depending on the nature of the load and the current path through it, the latter arrangement may produce induction and radiation fields in the vicinity.

The burst firing technique, because it tends to switch at the time the voltage or current is zero, does not produce the high rates of change. Consequently radio interference levels are orders of magnitude lower than with phase control as shown on the bottom curve of Fig. 5-37.

5-3.2.3 Susceptibility of SCR Circuits

SCR units are susceptible to pulse triggering from undesired transients. They may appear on either the gate or anode supply voltage as a result of coupling from other SCR's or other transient-producing devices. Efforts should be made to suppress such transients. When they appear in the voltage supply, a capacitor divider may be used to protect the circuit as shown for the unijunction transistor (UJT) circuit shown on Fig. 5-40 (Ref. 5). The capacitors should have values satisfying

$$\frac{C_1}{C_1 + C_2} = \eta \quad \text{, dimensionless} \qquad (5\text{-}9)$$

where η is the intrinsic stand-off ratio for the UJT.

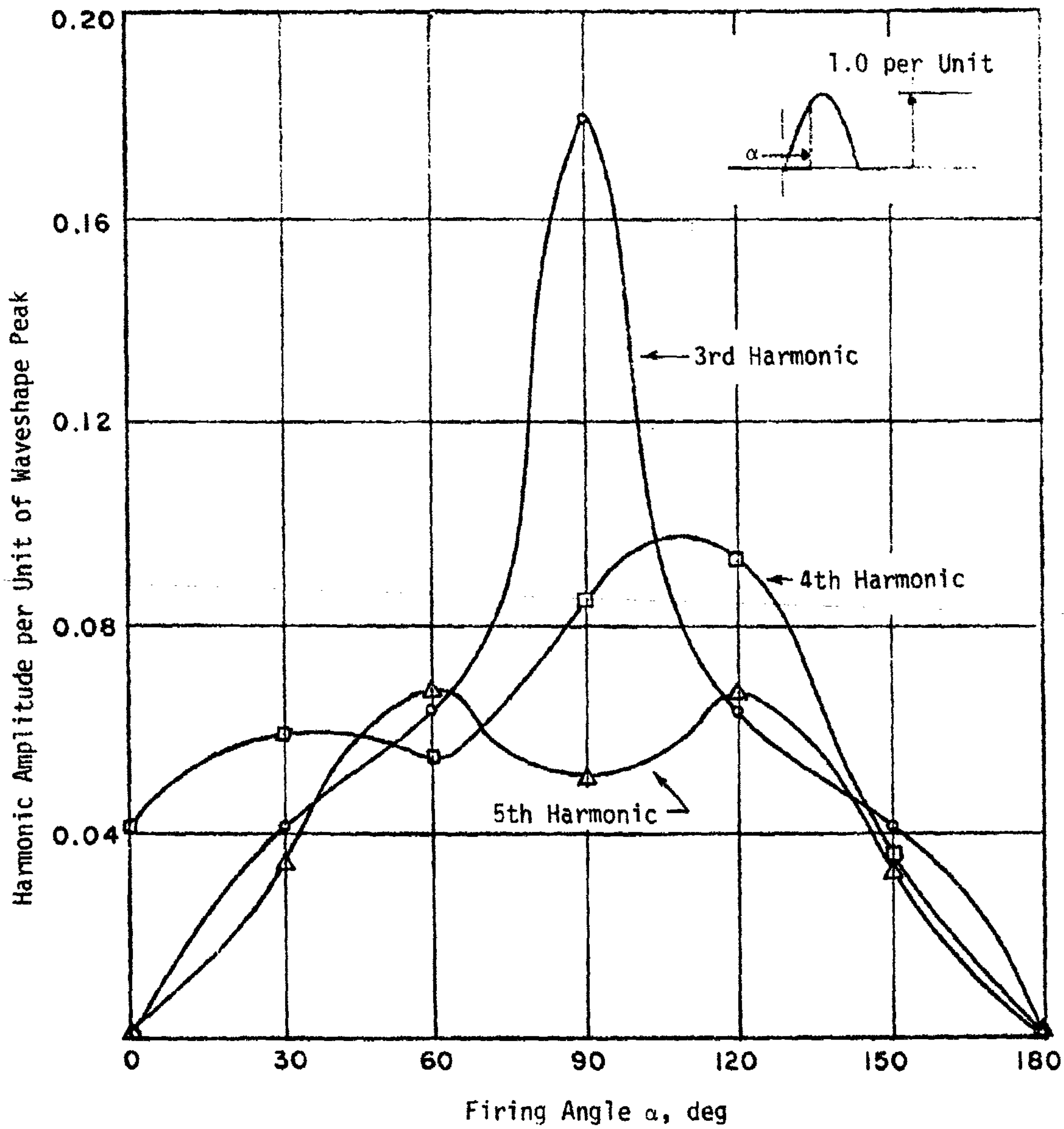


Figure 5-35. Harmonic Content vs Function of Firing Angle (Unsymmetrical Phase Control)

5-4 LIGHTING

5-4.1 INTRODUCTION

Two types of lamps are in general use — incandescent and gaseous discharge. Incandescent lamps normally are not considered to be a source of interference. In par. 3-1.3.2.3.2 a brief description was given of the radio noise from a gaseous discharge lamp. It arises from the switching phenomenon, random current conduction effects in the gas, and from oscillations that take place because the gas breakdown phenomenon can exhibit negative resistance characteristics.

5-4.2 FLUORESCENT LAMPS

The fluorescent lamp is an arc discharge that can be maintained at a relatively low voltage, but must be

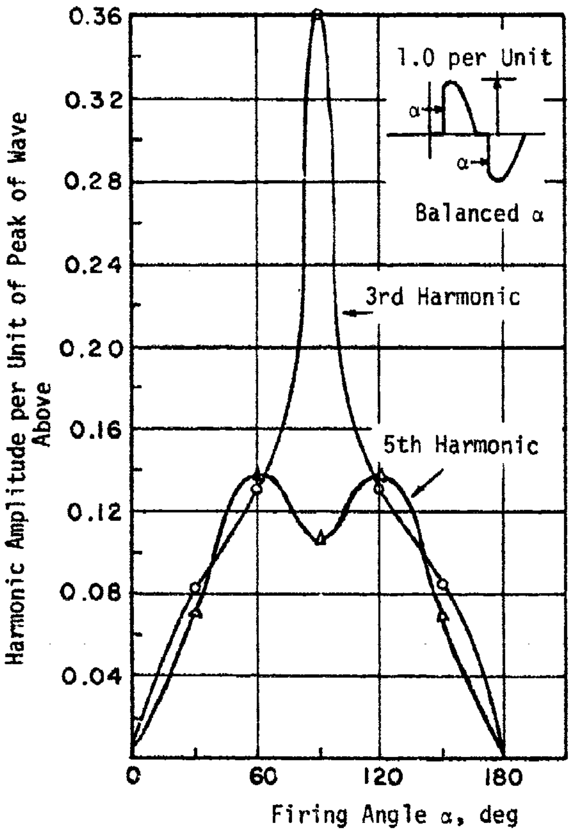


Figure 5-36. Harmonic Content vs Firing Angle (Symmetrical Phase Control)

restruck with a high voltage each time the lamp is turned on (if supplied from direct current), or on each half cycle (if supplied by alternating current). The arc is struck by means of a ballast which is effectively a large inductance which resists any sudden change in current (see Fig. 5-41).

Three types of lamps are used (Ref. 6). The switch start and rapid start types have filaments to heat cathodes which aid in arc initiation by emitting electrons. The switch start circuit contains a "starter" as shown in Fig. 5-41(A) which when closed initially enables the filament to heat and then opens after a few seconds causing the arc to strike.

A rapid start circuit is shown in Fig. 5-41(B) in which initially the tapped inductor causes the filaments to heat. After the arc strikes, which may occur in a few tenths of a second, the voltage across the tapped inductor drops and the filaments are operated with reduced current.

The instant start lamp has no filaments as shown in Fig. 5-41(C). Initially the combination of L_1 and C resonates to produce a high voltage to strike the arc. After the arc strikes, the resonant condition is no longer satisfied and the voltage across the lamp reduces.

Often forms of these basic circuits are used for purposes of improving efficiency, power factor, and cost. In a lamp for direct current, a series resistor is used to limit the current after the arc is initiated. Fig. 5-42 shows a two-lamp ballast.

The current through a lamp on ac supply will be nearly sinusoidal because of the presence of the ballast which causes the load to be primarily inductive. However, at each current zero the restriking phenomenon will cause a voltage transient which will appear across the lamp. Since this occurs on each half cycle, one can expect relatively strong levels of harmonics of the power line frequency to appear in the circuit in addition to radio frequency components. Because the ballast itself is a high impedance, the levels of harmonic current appearing on the line will be relatively small.

To reduce radio frequency components it is conventional to place a capacitor across the bulb terminals as shown with dotted lines in Fig. 5-41. The design of the ballast and the effects of distributed capacitance in its windings can be important in the overall radio interference characteristics of the lamp.

In addition to arc formation, another source of radio noise is oscillations produced by the nonlinear characteristics of the arc discharge (see par. 3-1.3.2.3.2). Such oscillations have been observed extensively in the frequency range around 10 kHz but experimental evidence indicates that similar phenomena can occur at higher frequencies of the order of 1 MHz.

5-4.2.1 Emission Levels

For an alternating voltage lamp supply, the switching transient generated on each half cycle for reignition is a major source of interference from low frequencies up to the megahertz range. Voltage levels measured at the terminals of the lamp connected to the power line can be expected to be of the order of 50,000 to 100,000 μV at harmonics in the vicinity of 100 Hz and vary inversely with the frequency reducing to a few microvolts at 100 kHz. Because of the instability of the arc, the interference will be broadband in nature so that higher order harmonics are difficult to separate.

The oscillations that appear can produce quite high voltages ranging from about a volt in the frequency range around 10 kHz (see Fig. 3-29), down to several

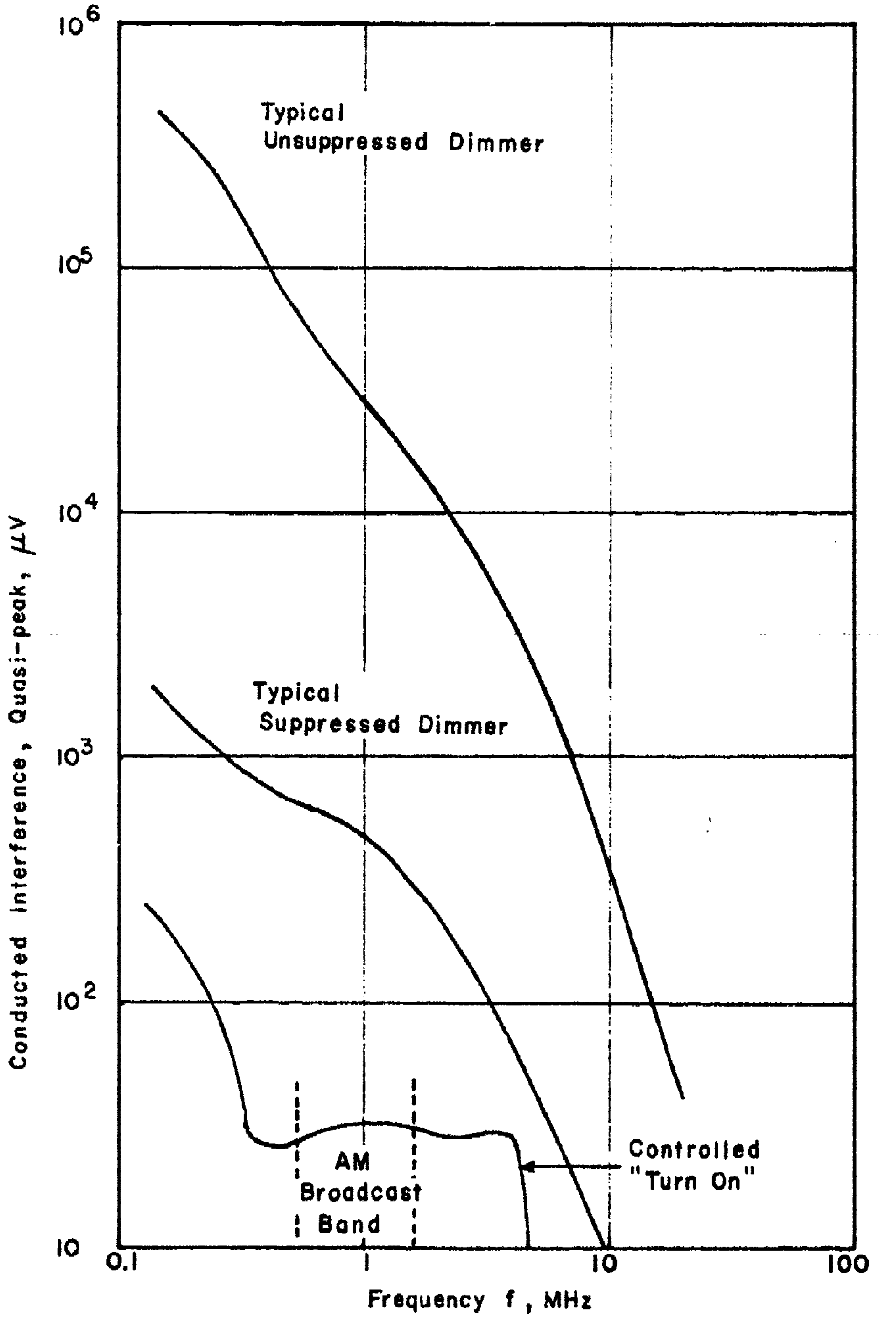


Figure 5-37. Conducted Interference for Light Dimmer

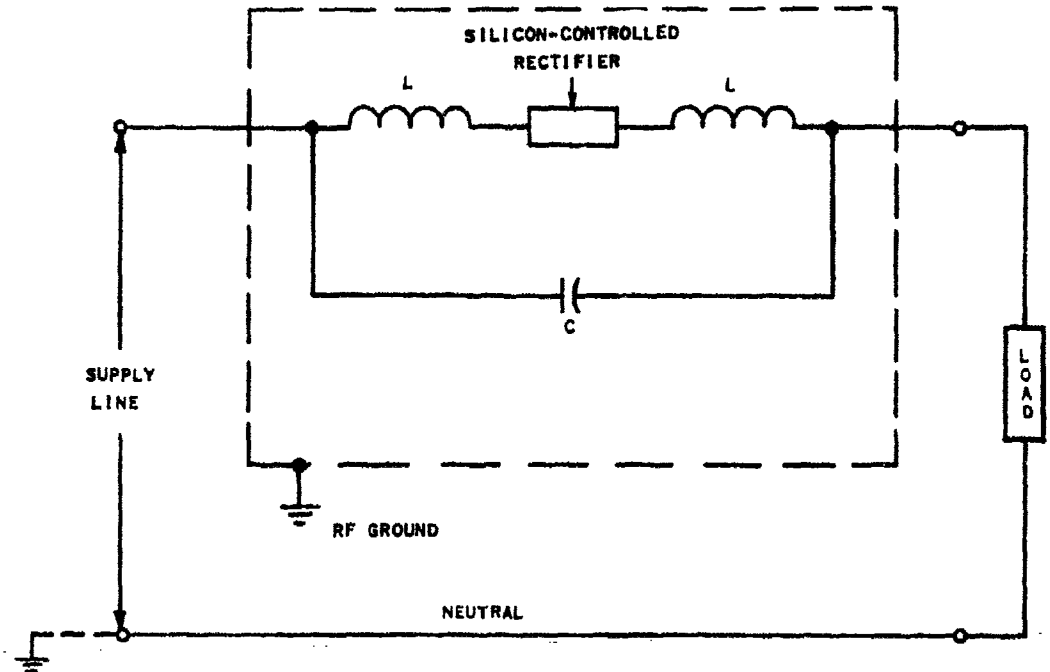


Figure 5-38. Conducted RFI Suppression With RF Ground Inaccessible

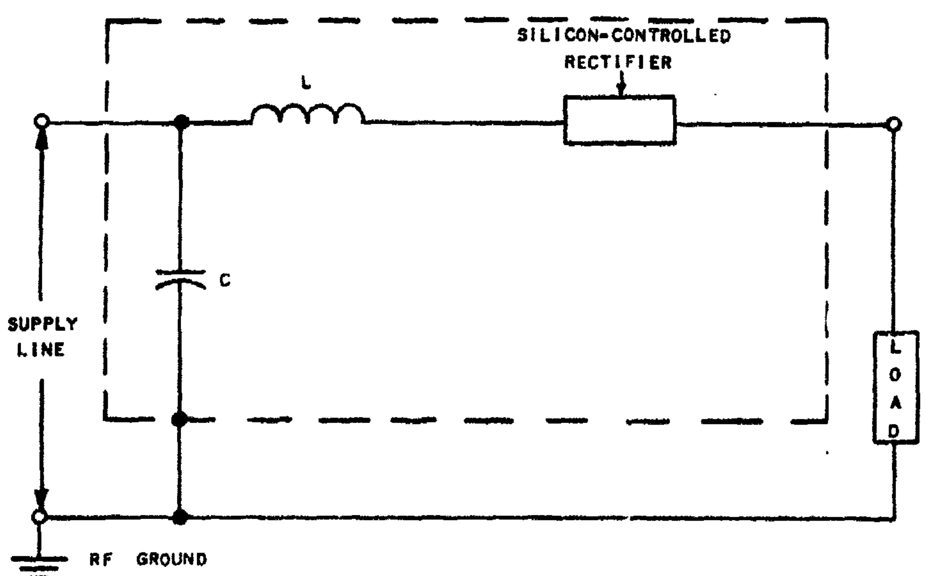


Figure 5-39. Conducted RFI Suppression With RF Ground Accessible

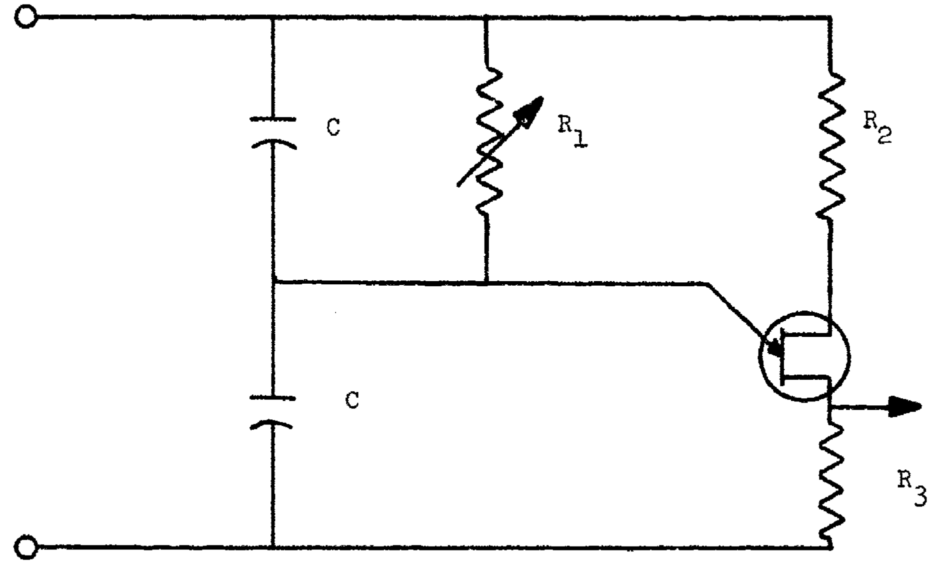


Figure 5-40. Decoupling Against Supply Transients

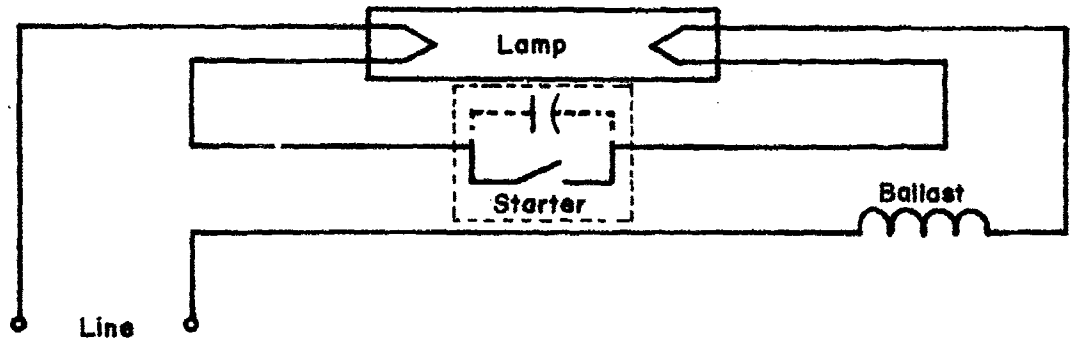
millivolts in the range around 1 MHz as shown in Fig. 5-43 and several hundred microvolts at 2 MHz as shown in Fig. 5-44. Oscillations in the vicinity of 10 kHz are usually accompanied by a high impedance at that frequency due to a self-resonant characteristic of the ballast, such as shown in Fig. 3-30. It is likely that peaks in emission at frequencies in the neighborhood of 1 MHz occur because of electronic resonance phenomena in the arc discharge itself. Radiated emission characteristics of fluorescent lamps arise because of two phenomena: (1) the magnetic field associated with the ballast, and (2) the electric field associated with the arc discharge in the fluorescent tube. Because of the high impedance of the ballast, the ballast currents themselves are pretty much restricted to low frequencies. Substantial magnetic fields can be measured at 60 Hz and at the first few odd harmonics of 60 Hz. The effective dipole strength at 60 Hz is on the order of 1 ampere-turn meter squared and at harmonic frequencies can be expected to be lower in accordance with an inverse harmonic number squared law. The magnetic flux density will vary inversely with the cube of the distance from the center of the ballast and can be computed quantitatively using Eqs. 3-114 or 3-115.

The electric field components are generated directly as a result of the high voltage which appears across the fluorescent tube on ignition. It is of a broadband type and, as shown in Fig. 3-31, is of the order of 100 μV·m/kHz of bandwidth at 1 MHz when measured at a distance of approximately 3 ft. It varies inversely with the frequency. The level falls off rapidly with distance and generally can be assumed to obey an inverse cube law at distances from the lamp which are larger than the lamp dimension itself. Furthermore, the level of emission is considerably dependent upon the construction of the lamp fixture. If the lamp is fully enclosed within a metal container, except for the minimum opening necessary for illumination, levels much lower than the levels shown in Fig. 3-31 may be obtained. Furthermore, the emission level will depend upon one's position with respect to the lamp opening. Usually because the lamp is backed by a metal box or reflector, levels on the back side will be much lower than those on the front side (Ref. 7).

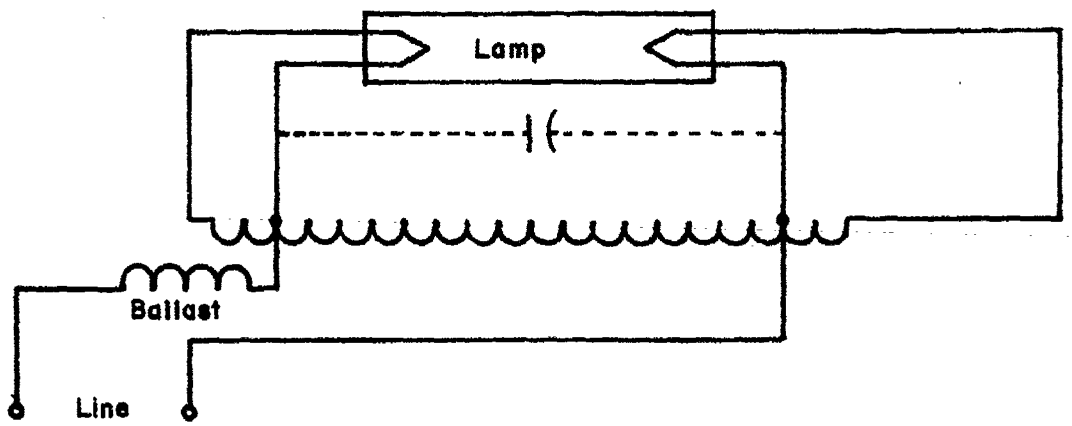
5-4.2.2 Interference Reduction

5-4.2.2.1 Radiation

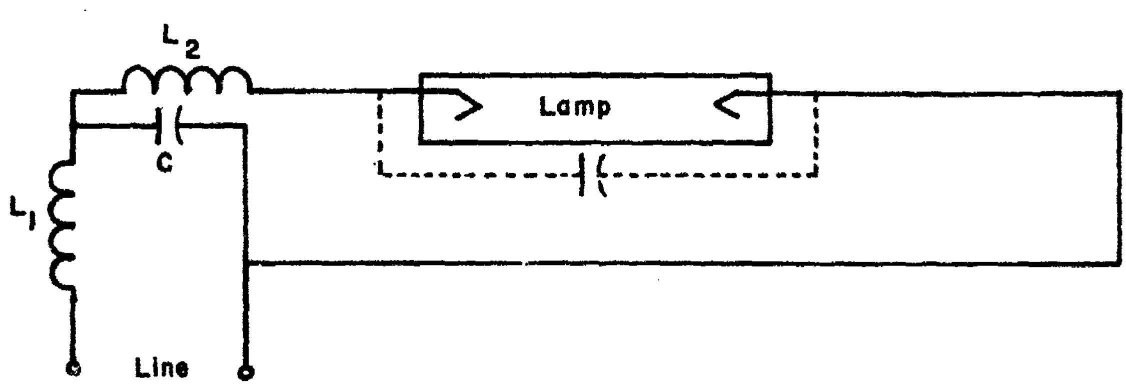
The fluorescent lamp fixture is a significant source of interference only when it is located close to a sensitive receiver or receiving antenna. At distances more than a few meters, direct emission in the form of an electric field is not likely to cause difficulty. Where interference reduction is necessary, the two techniques usually used are: (1) a metal grille or louver which essentially surrounds the lamp with a complete metal enclosure, and (2) conducting (transparent) glass or plastic. With regard to the former, openings in the



(A) Start Switch Circuit



(B) Rapid Start Circuit



(C) Instant Start Circuit

Figure 5-41. Fluorescent Lamp Circuits

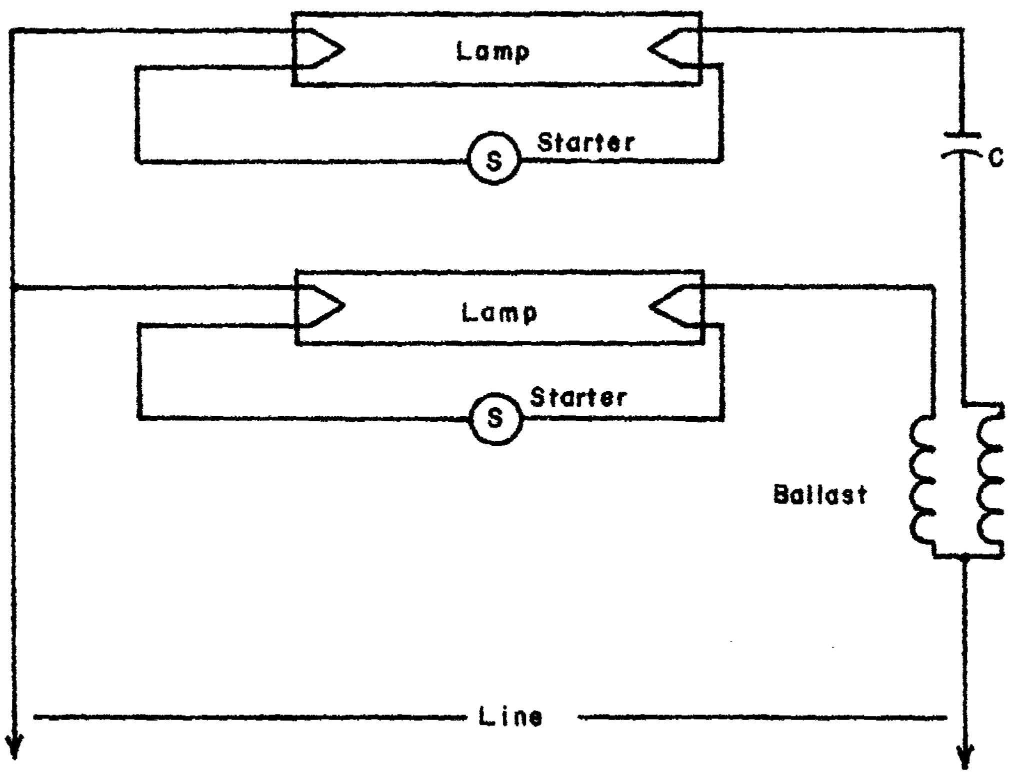


Figure 5-42. Dual Lamp Fluorescent Fixture

grille are not significant sources of leakage or radiation at the frequencies emitted by the lamp. The expected attenuation can be estimated using formulas in Chapter 4 for screens or, if of sufficient thickness, those for waveguide below cutoff devices (see par. 4-6.5.2). The conducting technique probably permits greater light transmissibility. The effectiveness of a grille is indicated in Fig. 3-31. Corresponding attenuations can be obtained with conducting material.

5-4.2.2.2 Conduction

Although the levels of emission currents on the power line are usually small, they can produce interference to nearby equipments. To control emissions in the broadcast frequency range, the ballasts should be constructed to have a minimum distributed capacitance in order that substantial impedances to current flow will be obtained. Differential-mode currents can be controlled by inserting filters at the line terminals. A simple line-to-line capacitor may be sufficient. The flow of common-mode currents can be reduced by using balanced arrangements of lamps and the associated ballast, and by carefully controlling the distributed capacitance to the fixture. If such techniques are not adequate, it is possible to insert common-mode filters or chokes in series with the power-line connection.

Grounding of the lamp fixture is probably best done through the ground wire or conduit of the power cable. Where an environmental (or local) ground is available, grounding to it can increase common-mode currents, and thus increase local induced or radiated fields.

5-5 ELECTRONIC POWER SUPPLIES

5-5.1 INTRODUCTION

The most common form of electronic power su ply is one used to convert the power from an alte nating or direct voltage to a different direct voltag Usually, the ac supply uses a transformer and a rec fier. For converting from direct voltage to dire voltage, a switching system is used which converts t direct voltage to alternating voltage, which may

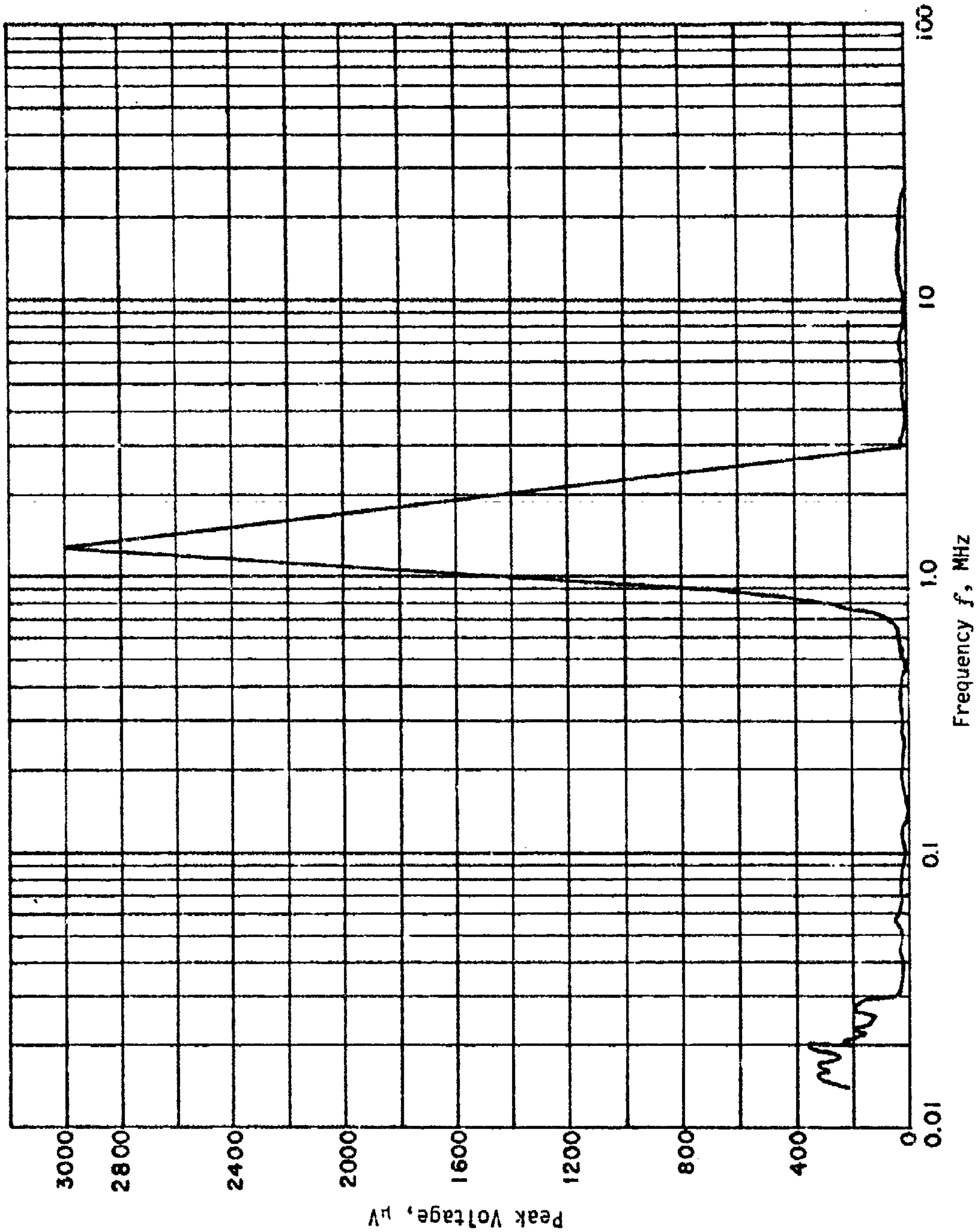


Figure 5-43. Conducted Interference from Fluorescent Lamps

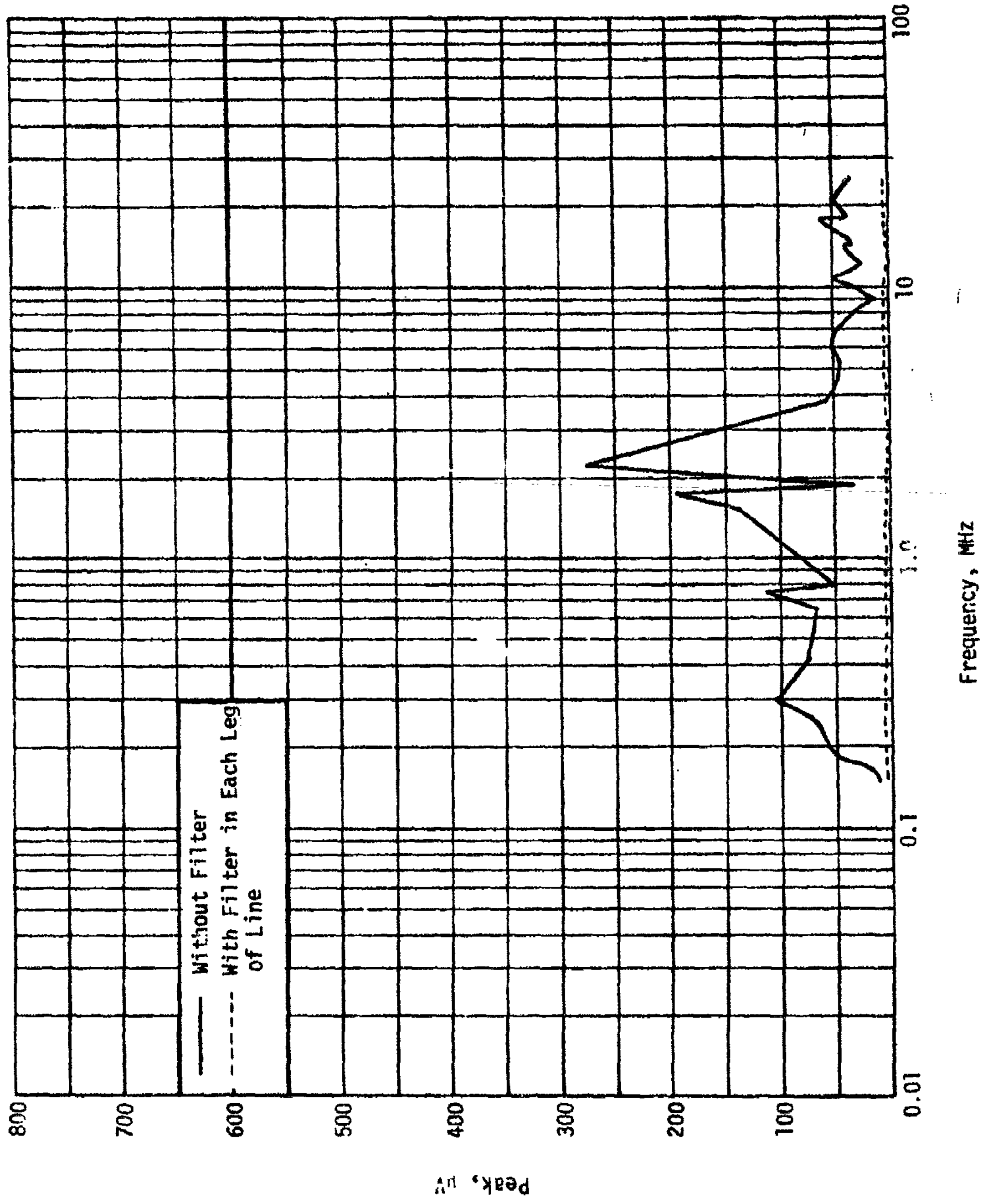


Figure 5-44. Conducted Interference from Three-lamp Fluorescent Fixture, With and Without Filter

transformed, and then a rectifier is provided to convert it back again. Inverter power supplies of this type frequently operate at frequencies in the kHz range. Indeed, some ac supplies are now being built with inverters. Because the operation of power supplies depends upon switching action, they are potentially prolific sources of interference. The basic theory of the generation of an emission spectrum from such switching action has been presented in pars. 3-1.3.2.1 and 3-1.3.3.1.1. In the circuits where the switching actions take place, the spectrum of the currents that flow is extensive, and locally such currents are capable of creating magnetic and electric fields of substantial magnitudes. Here, concern is not so much with the internal currents, voltages, and fields as it is with those that appear at the terminals of the power supply. These may be considerably less than those that appear internally due to filtering that takes place either automatically as a result of circuit design, or intentionally by interference reduction or power supply ripple filters. It is assumed that the designer will limit any fields which are created within the supply by placing appropriate shields around the circuits involved.

Interference from power supplies can occur in two ways (Fig. 5-45): (1) it may be generated within the power supply itself or (2) it may be generated in one circuit or equipment and transferred through the common impedance of the power supply to other circuits or equipments. Interference control measures include filtering, shielding, circuit planning, and selection of components. The use of well-regulated

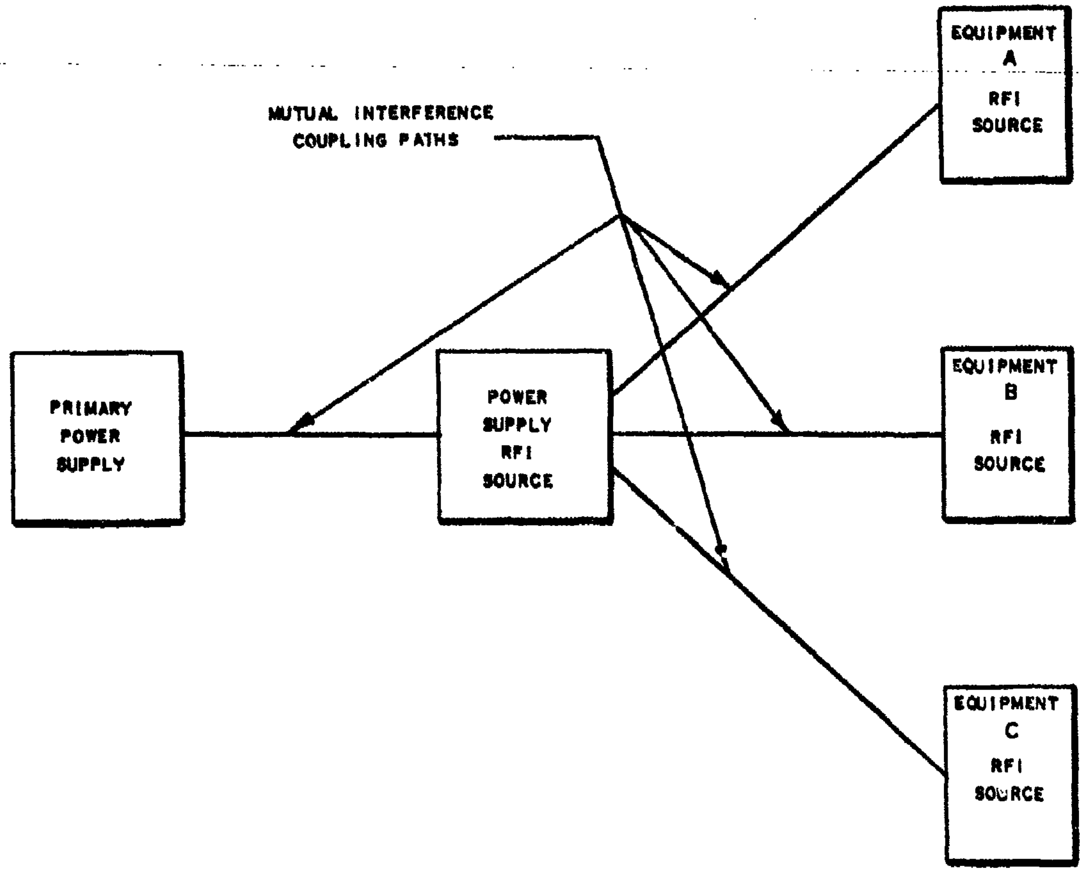


Figure 5-45. Mutual Interference Coupling

supplies helps to achieve very low common impedance coupling between circuits, which will substantially reduce this mode of interference transfer.

5-5.2 CIRCUIT ARRANGEMENTS

When a common power supply must serve many circuits, the design should isolate interference sources from susceptible circuits by means of filters (Fig. 5-46) in accordance with their individual power requirements and spectrum characteristics. When the power supply is an interference source (Fig. 5-46(A)), its lines should be filtered. When the equipment is the interference source, the equipment lines should be filtered (Fig. 5-46(B)).

Minimizing coupling path circuits is an important design consideration. All circuits employing a common power supply should be located as close together as practicable so that the number of suppression components and amount of shielding is minimized, and long lengths of shielded cable can be eliminated. Fig. 5-47 illustrates a power supply that is isolated from its load. The interference control measures consist of two shielded cases, filter, shielded cables, and an internal case partition.

5-5.3 COMPONENTS

5-5.3.1 Vacuum and Gas Tubes

Some components normally used in power supply circuits — such as diodes, and thyratrons — are significant sources of interference. Power-frequency harmonic interference generation is discussed in par. 3-1.3.3.1. Magnetic field interference is generated by the associated transformers and choke coils. Gas-tube and RF circuits usually produce high levels of interference in power supplies. When gas diodes and thyratrons (or silicon-controlled rectifiers) are used, interference arises from two distinct effects: (1) the steep voltage and current wave-fronts associated with the firing (ionization) cycle of the device, and (2) any plasma oscillations that occur during the discharge. The external effects of both types of interference can be minimized by shielding the circuit and by filtering.

5-5.3.2 Radio Frequency Components

Radio-frequency power conversion circuits contain free-running oscillators and sharply resonant circuits. The oscillator output is amplified, and rectified and filtered. Stray RF energy radiates from the wiring, transformer, and other components, and is conducted along input and output wires. It must be confined to the power supply itself by shielding and filtering.

5-5.3.3 Semiconductor Generated EMI

5-5.3.3.1 General

Sudden step changes in current caused by semiconductor or mechanical switching generate a broad energy spectrum. This spectrum, once produced, can be conducted and radiated throughout the entire system. It can be reduced to practical limits, the amount of reduction depending on factors such as cost, space, weight, and conflicting specifications. Semiconductors used in power conversion circuits include diodes, thyristors, SCR's, diacs and triacs.

5-5.3.3.2 Diode Recovery Times

Reverse recovery (par. 3-1.3.2.5) is a phenomenon common to all p-n junction diodes. Reverse recovery time T_{rr} occurs immediately after forward bias is removed and is a measure of the time required to remove or "sweep out" the minority carriers from the n- and p-regions and reunite their covalent bonds. T_{rr} for switching diodes ranges from 20 ns to 20 μs and is proportional to the magnitude of peak reverse current I_{pr}. It has been shown that I_{pr} ranges from greater than five times forward current to values less than forward current. The reverse current pulses, once produced, are then conducted back to the power source. The physical characteristics of the device, such as size and diffused impurities, can reduce reverse recovery time. Circuit conditions which reduce T_{rr} are low forward current, large reverse voltages, low impedances, and parasitic circuit capacitance.

5-5.3.3.3 Forward Recovery

Forward recovery transients have an interference generating effect similar to reverse recovery transients. Forward recovery time occurs immediately after a forward drive pulse is applied and is a measure of the settling time of the p-n junction to a steady-state conduction level. The most common methods of reducing forward transients and the resulting EMI are: (1) shape the driving pulse for maximum tolerable rise time, (2) keep the junction slightly forward biased, (3) reduce drive pulse amplitude to lowest level to maintain reliable switching action, and (4) select a device with fast recovery characteristics.

5-5.3.3.4 SCR Recovery

Silicon-controlled rectifiers (SCR's) can produce reverse recovery transient currents similar to p-n junction diodes. The reverse recovery transients in SCR applications are intensified by increased forward current and large junction area. T_{rr} for a typical SCR is of the order of microseconds. In full-wave

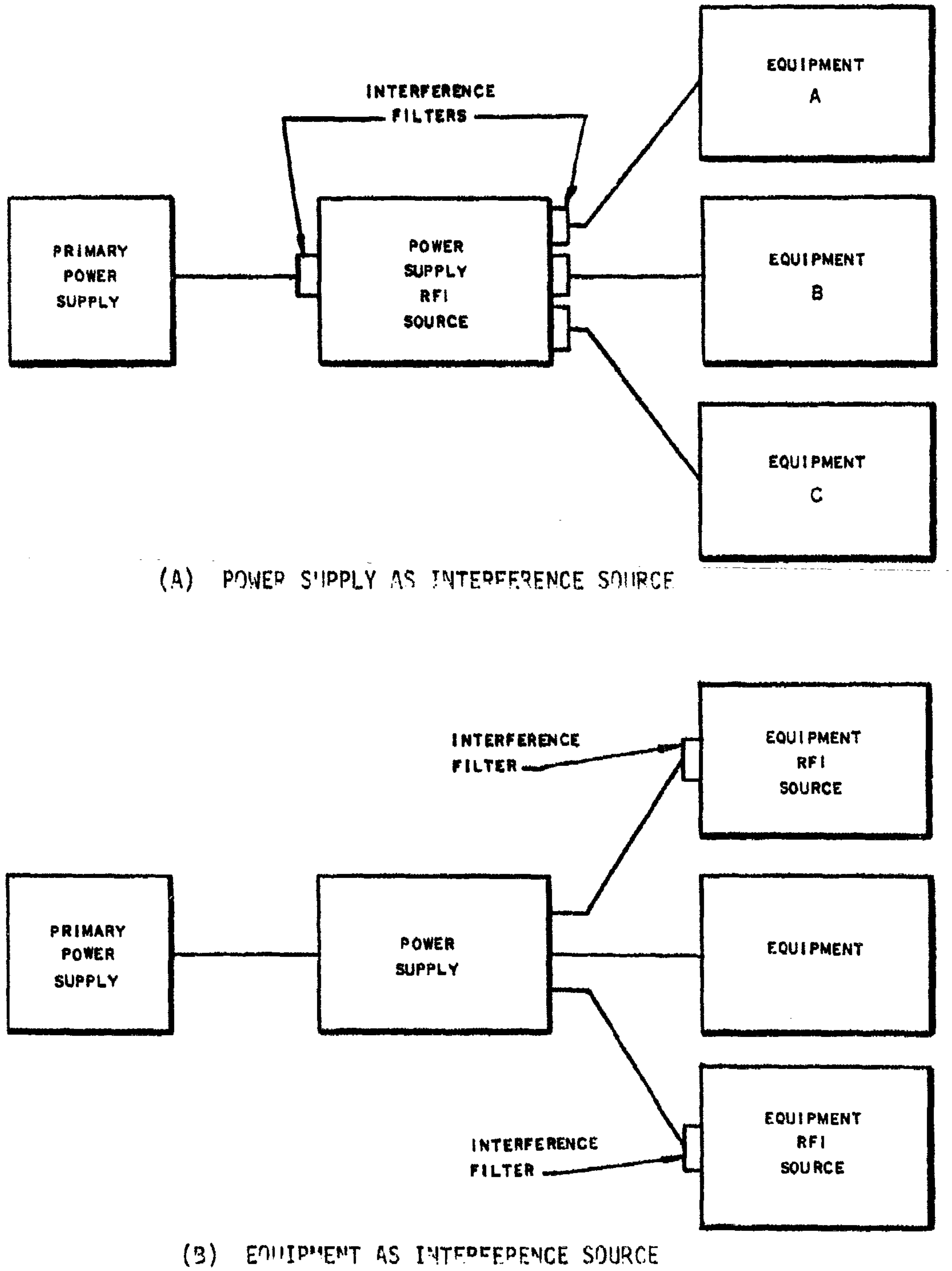


Figure 5-46. Power Supply Filtering

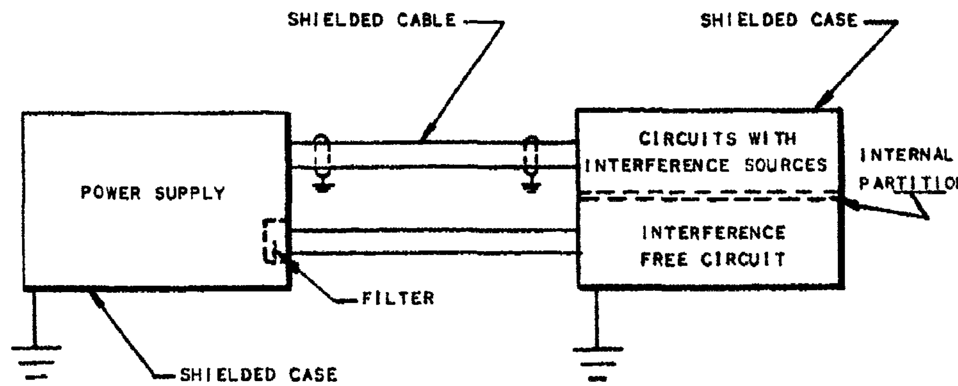


Figure 5-47. Power Supply Isolation

SCR rectifier applications, the reverse recovery current has to be carried by the complementary SCR's. When SCR's are connected in series, mismatch of reverse recovery times can alter the reverse voltage distribution enough to exceed the maximum reverse voltage rating of the SCR's. Consideration of load current and surge current sharing is necessary when parallel SCR operation is required.

5-5.3.3.5 SCR Turn On

Perhaps a more important source of EMI in SCR's is the large and sudden rate of rise *di/dt* of forward current when the device turns on, especially when phase control is used. This phenomenon can be controlled by adding inductance in series with the SCR as shown in Figs. 5-38 and 5-39, par. 5-3.2.2. Since the most significant transient occurs at turn on, where current is low, the inductors can be designed so as to saturate at the maximum currents to be carried without degrading their effectiveness.

5-5.3.4 Transformers

Transformers produce transients that conduct and radiate EMI to the load and back into the power distribution system. The nonlinear hysteresis loop allows various amounts of residual flux to remain in a transformer core when line voltage is removed. When line voltage is reapplied with a polarity that tends to increase the residual flux, the core saturates resulting in excessive current being drawn from the supply. This magnetizing current transient is impressed on the secondary of the transformer and also back through the power distribution system. Line voltage interruptions can create magnetizing in-rush current transients as well as at initial turn on. Controlled line voltage switching is the most effective method to eliminate magnetizing current transients. Filtering can reduce the generated EMI, but physical size limits high power low frequency filtering. Electrostatic and electromagnetic shielding are other methods that could be used to reduce EMI radiated from transformers.

5-5.3.4.1 Electrostatic Shielding of Transformers

Electrostatic shielding is accomplished by enclosing the secondary of the transformer by a thin metal conducting shield, or shields, usually made of copper or aluminum. The shields are insulated from one another and are grounded in ways depending upon the application. This does not hamper the inductive coupling of the primary and secondary, but does reduce the interwinding capacitances, thus stopping the direct flow of high-frequency currents between primary and secondary.

5-5.3.4.2 Electromagnetic Shielding of Transformers

Stray magnetic fields outside a transformer may be a source of hum in high-gain electronic systems. There are four methods for reducing these fields: (1) use of a wound toroidal-core construction, (2) reduction of flux density in the core, (3) constructing the transformer case of a high magnetic permeability material, and (4) enclosing the transformer in a magnetic shield (see par. 4-6.6).

The last of these techniques generally is the most practical, and can be quite effective where space is available to accommodate the shield, and adequate transformer cabling can be maintained.

5-5.3.5 Relays (see par. 3-1.3.2.2)

Relays generate EMI in both the switched circuit and control circuit. The inductive coil induces transients that can adversely affect other relays and low-voltage components in the control circuit. In some applications, relays can be replaced by semiconductor switches with controlled forward current rise time and noninductive control circuits.

5-5.4 SUPPLY LINE HARMONICS (Ref. 8)

5-5.4.1 Single-Phase Power Supply Rectifier Circuits

The conducted EMI caused by a rectifier can be determined by developing the Fourier Series of the ac line current. Unfortunately, the exact solution is quite tedious and time consuming, particularly when the rectifier load is complex. Therefore, it is expedient to use simplifying approximations. Two approximations that have been used are (1) the "flat-topped" approximation for the line current in rectifier circuits having a resistive or inductive load, and (2) the "impulse" approximation for rectifier circuits having a capacitive load. The results of these calculations are summarized in Table 5-5 and shown in Fig. 5-48. The calculations assume ideal diode and transformer characteristics. Although both assumptions are reasonable, it must be remembered that an additional line current, the magnetizing current, will be present in the primary winding of any transformer. The magnetizing current has two components — one component to supply transformer losses and another to provide the magnetomotive force necessary to set up the flux in the transformer core. The loss component is sinusoidal while the flux component is cosinusoidal with a prominent third harmonic due to the nonlinearity of the *B*-*H* curve of the core material. The powerline distortion caused by rectifiers under different load conditions is shown in Fig. 5-49.

TABLE 5-5. SINGLE-PHASE RECTIFIERS

ORDER OF HARMONIC	RELATIVE DISTORTION IN LINE CURRENT, dB*							
	HALF-WAVE RECTIFIER				BRIDGE RECTIFIER			
	L	*R*	*C*-LOAD APPROX	*R*-LOAD APPROX	*L*	*R*	*C*-LOAD APPROX	*R*-LOAD APPROX
1	0	+4	+6	+3	0	+10	+6	+2
2	None	−3.5	0	−4	None	None	None	None
3	None	None	−3.5	−6.5	None	None	−3.5	−7.5
4	None	−17.5	−6	−10	None	None	None	None
5	None	None	−8	−11	None	None	−8	−12
6	None	−25	−9.5	−13.5	None	None	None	None
7	None	None	−11	−14	None	None	−11	−15
8	None	−30	−12	−16	None	None	None	None
9	None	None	−13	−16.5	None	None	−13	−17
10	None	−34	−14	−18	None	None	None	None
11	None	None	−15	−18	None	None	−15	−18.5
12	None	−37	−15.5	−19.5	None	None	None	None
13	None	None	−16	−19	None	None	−16.5	−20
14	None	−40	−17	−21	None	None	None	None
15	None	None	−17.5	−20.5	None	None	−17.5	−21.5
16	None	−42	−18	−22	None	None	None	None
17	None	None	−18.5	−21.5	None	None	−18.5	−22.5
18	None	−44	−19	−23	None	None	None	None
19	None	None	−19.5	−22.5	None	None	−19.5	−23.5
20	None	None	−20	−24	None	None	None	None

*Relative to dc Output

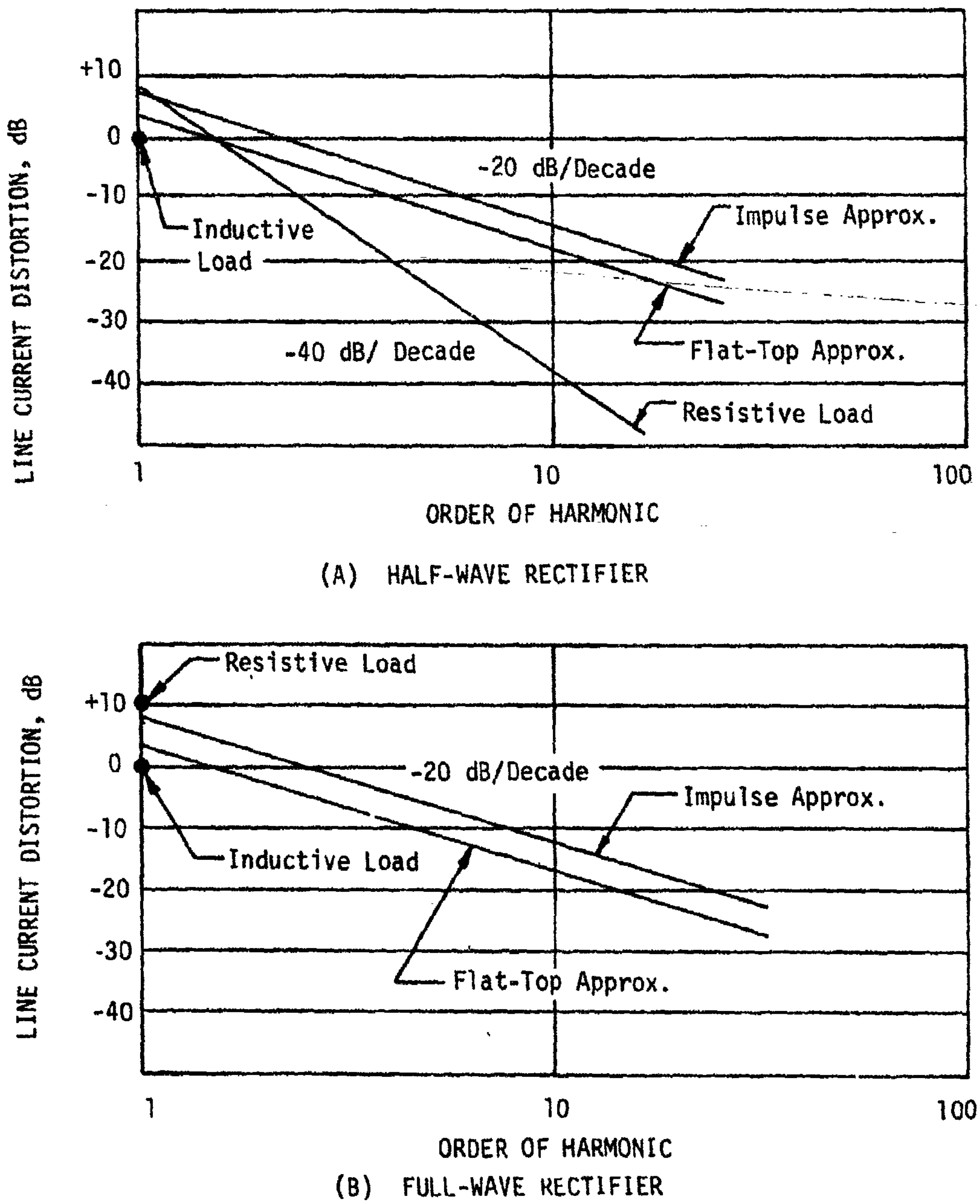


Figure 5-48. Current Spectrum for Single-phase Rectifiers

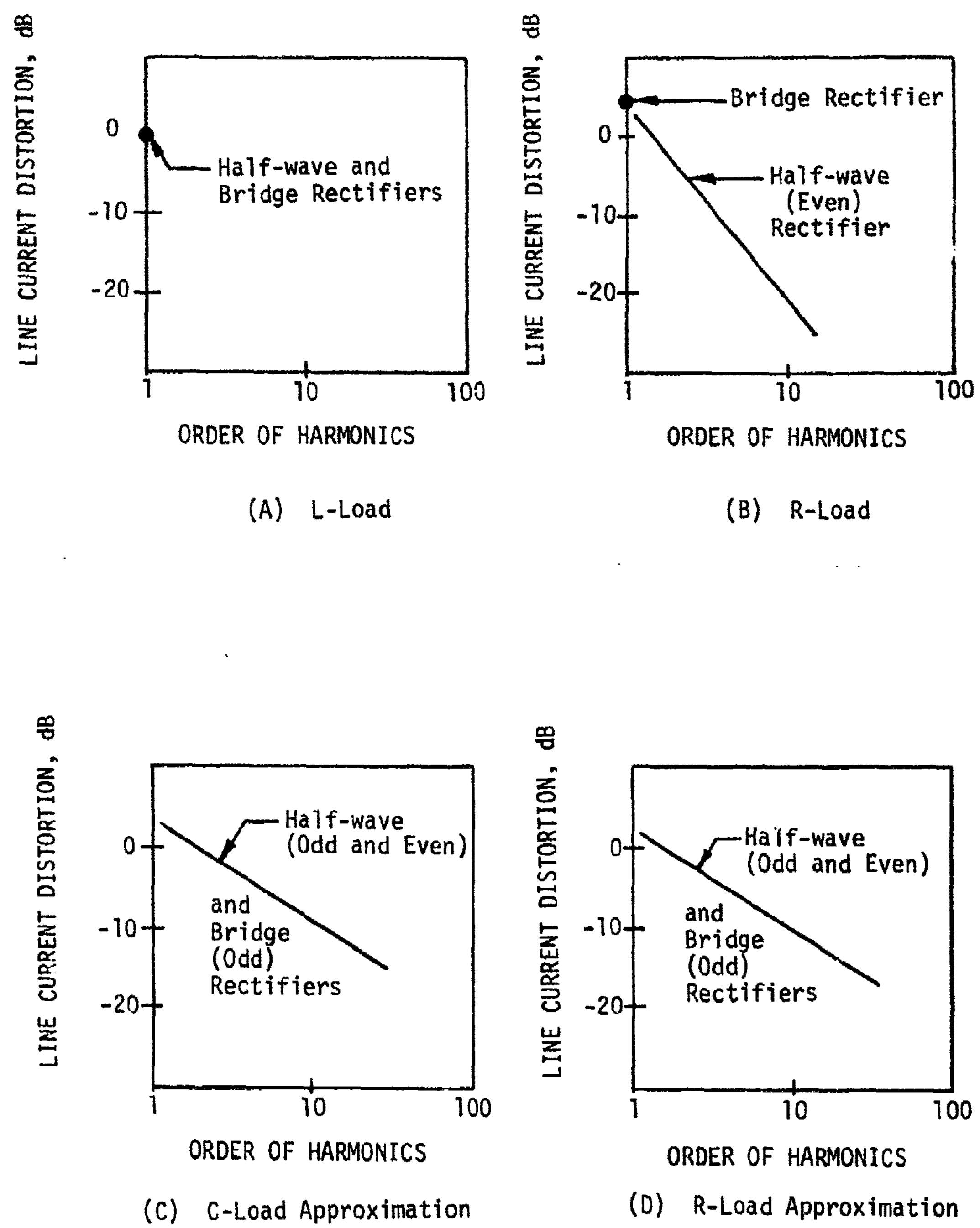


Figure 5-49. Current Spectrum for Single-phase Rectifiers Under Different Load Conditions

5-5.4.1.1 Half-wave Rectifiers

Of special interest is the half-wave rectifier as shown in Fig. 5-50. The current in the primary winding of a transformer whose secondary winding supplied a half-wave rectifier cannot be determined by the principle of superposition. The primary current is not the summation of the secondary current plus the nonload exciting current. The primary current during the interval when the secondary current is zero is an exciting current, but it is not the nonload exciting current. Since the average value of the secondary current is not zero, the transformer core is subjected to a magnetic bias that increases the exciting current in a manner similar to that in a saturable reactor (refer to Fig. 5-50). Obviously, half-wave rectifier circuits should not be used without adequate determination that excessive EMI will not result.

5-5.4.1.2 Full-wave Rectifiers

The full-wave rectifier shown in Fig. 5-51 has twice the ripple frequency in its output, thus reducing the smoothing filter requirements to achieve a given ripple output. This circuit features a minimum of dc transformer core polarization because the current flow in each half of the center-tapped secondary tends to cancel.

5-5.4.1.3 Bridge Rectifiers

It is of some significance that the bridge rectifier shown in Fig. 5-51 operating into either a resistive or an inductive load causes no appreciable line current harmonics. In practical cases, a filter capacitor is used which produces harmonics. The power supply designer should give careful consideration to the economics of using a choke-input filter sufficient to maintain the load inductive.

5-5.4.2 Multiphase Rectifier Circuits

5-5.4.2.1 Harmonic Content of Multiphase Rectifier Circuits

The harmonics in the primary of a transformer supplying a multiphase rectifier circuit are fewer in number, higher in frequency, and lower in amplitude than the harmonics generated by a comparable single-phase rectifier. Multiphase rectifiers can be analyzed in the same manner as were the single-phase circuits and the same approximations can be applied. The results of such an analysis are summarized in Table 5-6. Table 5-7 shows that a 12-phase rectifier causes a comparatively minor disturbance to the line current with no harmonics being generated lower than the 11th. A 6-phase rectifier shown in Fig. 5-52 generates a fifth harmonic, while a 3-phase rectifier shown in Fig. 5-53 generates a second, fourth, and fifth harmonic. Moreover, the 24-phase circuit generates no harmonics until the 23rd and 25th, while the 36-phase circuit doesn't generate any harmonics lower than the 35th (see Table 5-7).

5-5.4.2.2 Multiphase Rectifier Designs

A system consisting of six 6-phase rectifiers can be converted into a 36-phase system by supplying three rectifier transformers with delta-connected and three with wye-connected ac windings, and by the addition of four 10-deg phase shifting autotransformers, so that a 10-deg phase shift is obtained between the windings of successive rectifiers. Such systems are rather common in the electrochemical industry. A 24-phase rectifier, common in industrial power systems, includes a transformer with two or three primary windings that accept a 3-phase potential and multiple secondary windings, with rectifiers connected in four 6-phase groups displaced 15 electrical degrees. Such a system requires 24 rectifier diodes instead of 6, but it produces much less EMI and less output ripple.

5-5.5 INTERFERENCE CONTROL METHODS

5.5.5.1 Filtering

Filters, as shown in Fig. 5-54, may be used at the ac line terminals of a rectifier to reduce the flow of harmonic currents into the ac system. Such a filter usually consists of reactors connected in series with the ac line, and capacitors of resonant shunts connected between the line terminals on the rectifier side of the series reactors. The capacitors, or the resonan shunts, provide a low-impedance path for the har monic currents, while the impedance of the series re actors opposes the flow of these currents into the a system (see par. 4-5.2.1).

5-5.5.1.1 Conventional *L-C* Filters

A number of manufacturers provide high currer power line filters; however, there is often a wide di crepancy between the actual in-circuit performan of these filters and the manufacturer's rated pe formance. These discrepancies can be so great as result in insertion gain, rather than loss, at specif frequencies within the range of 14 kHz to 10 GH Most power-line filter problems can be traced to o of the following: (1) the use of marginal or su standard components, such as under-size induct cores which saturate at much less than the rated c rent; and (2) impedance mismatch caused by the u of 50 ohms for source and load impedance in t filter design criteria rather than full rated load i pedance and more realistic source impedance o

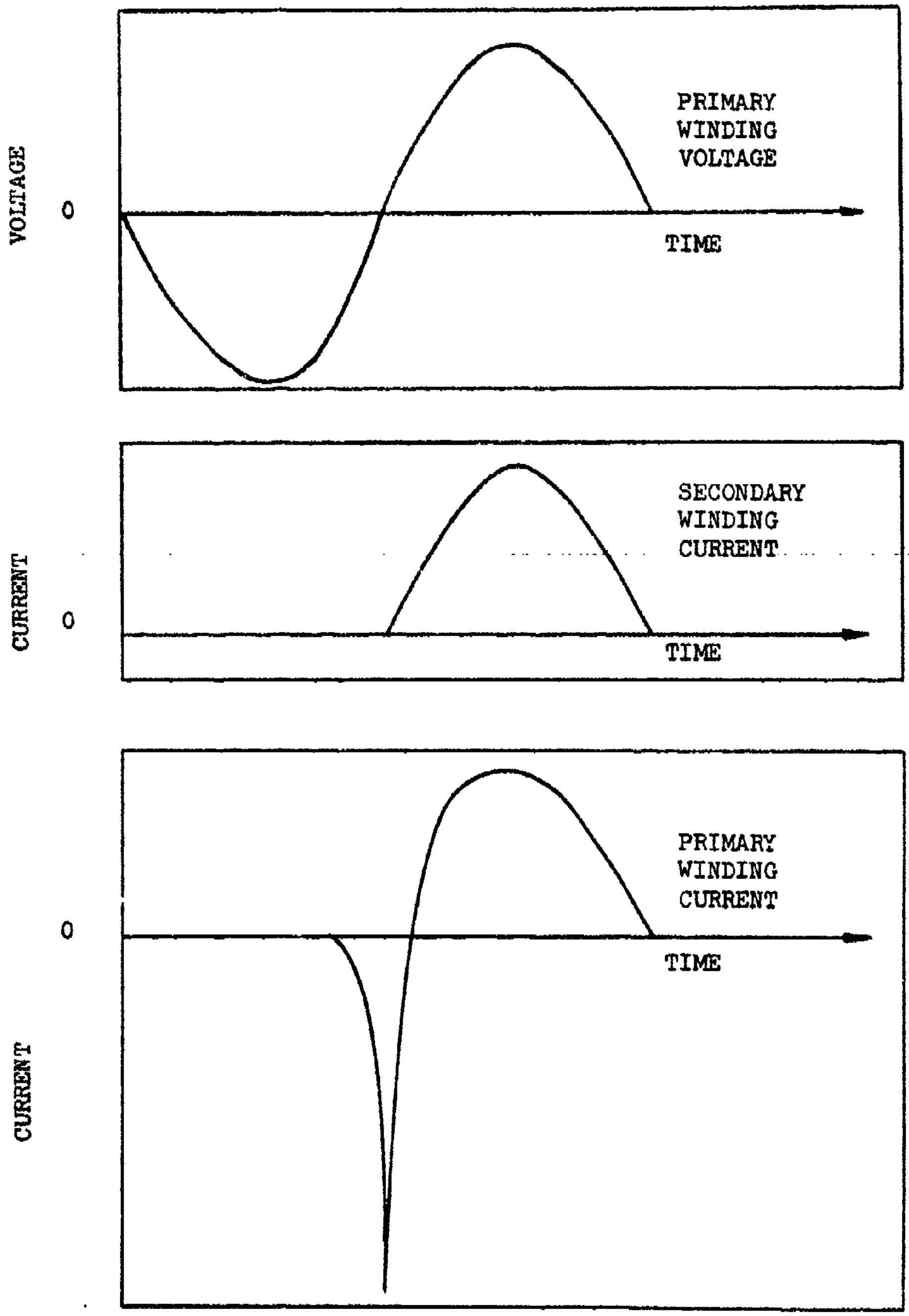
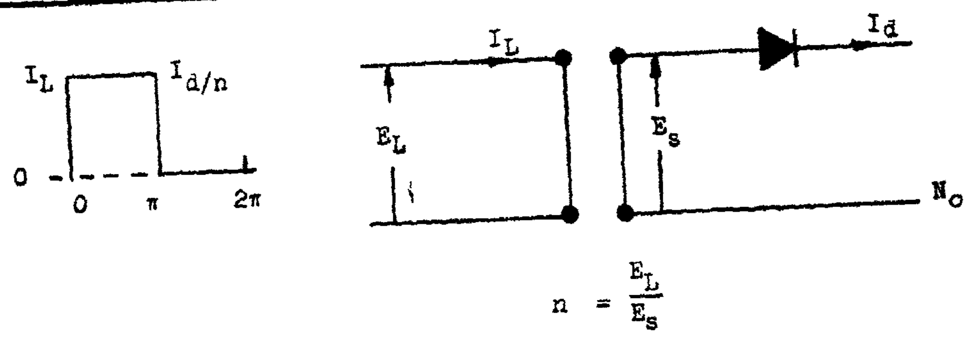
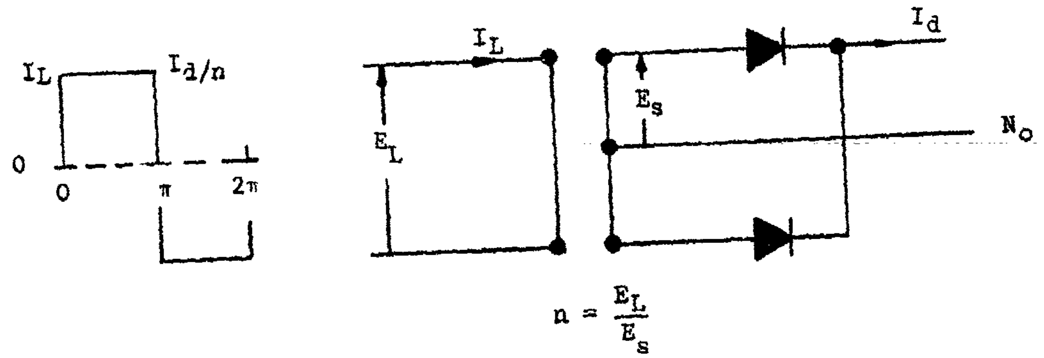


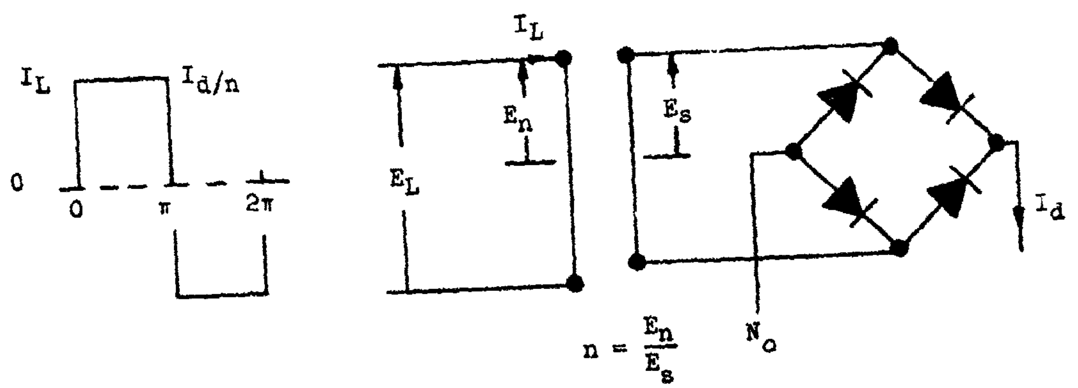
Figure 5-50. Waveforms in the Half-wave Rectifier Circuit



(A) HALF-WAVE RECTIFIER



(B) FULL-WAVE RECTIFIER



(C) BRIDGE OR DOUBLE-WAY RECTIFIER

Figure 5-51. Single-phase Rectifiers

TABLE 5-6. THREE-PHASE RECTIFIERS

RELATIVE DISTORTION IN LINE CURRENT, dB*						
ORDER OF HARMONIC	Δ-Y BRIDGE		ZIG-ZAG BRIDGE		Y-GENERATOR BRIDGE	
	R-LOAD APPROX	*C*-LOAD APPROX	*R*-LOAD APPROX	*C*-LOAD APPROX	*R*-LOAD APPROX	*C*-LOAD APPROX
1	+5.5	+6	−2		+1	+1
2	None	None	−5.5		None	None
3	None	None	−11.6		None	None
4	None	None	None		None	None
5	−8	−8	−16		−13	−13
6	None	None	−14.5		None	−25.5
7	−11	−11	−19		−16	16
8	None	None	None		None	None
9	None	None	−21.5		None	−29
10	None	None	−19		None	None
11	−15	−15	−23		−20	−20
12	None	None	None		None	−31.5
13	−16.5	−16	−24.5		−21.5	−21.5
14	None	None	−22		None	None
15	None	None	−25.5		None	None
16	None	None	None		None	None
17	−19	−18.5	−26.5		−24	−24
18	None	None	−24		None	−35
19	−20	−19.5	−27.5		−25	−25
20	None	None	None		None	None

*Relative to dc Output

fraction of an ohm to a few ohms at frequencies below 100 kHz. One filter specified to provide 100 dB at 14 kHz (with 50-ohm load) did not reach this attenuation level in the actual circuit until 185 kHz. Measurements on other filters at nondesign loads are shown on Fig. 5-55. These show a severe change in the pass-band of the filter, with reduced attenuation at the lower frequencies.

Obviously, the power supply designer must make his own analysis of the ability of a given power line filter to reduce harmonics to meet an EMI requirement, taking into account characteristics of the circuit into which it is to be inserted. Common power-line filters are the Butterworth and the m-derived types. The elliptic design has the advantage of providing a rapid transition from stop-band to pass-band and may be worth serious consideration; however, it usually will be accompanied by additional size and weight. The effect of saturation of the filter inductors is a widening of the pass-band. For example, an increase of approximately 4 times (from 10 kHz to 43 kHz) in the breakpoint of the typical power line filter was noted.

5-5.5.1.2 Compatible Lossy Filters (see par. 4-5.2.1.6)

The compatible lossy filter is a lumped parameter approximation of a lossy transmission line. Typically, the number of sections in a ladder filter are greatly increased and the shunt capacitance is provided by low-Q ($Q < 1$) ceramic capacitors. Compatible filters are not so dependent on impedance matching as are the conventional class of filters; thus, they have more dependable attenuation characteristics under a variety of source and load impedance conditions. Difficulties encountered with this type of filter include devaluation of ceramic capacitors with applied voltage and pronounced reduction of the dielectric constant at only moderately high temperature (125°C).

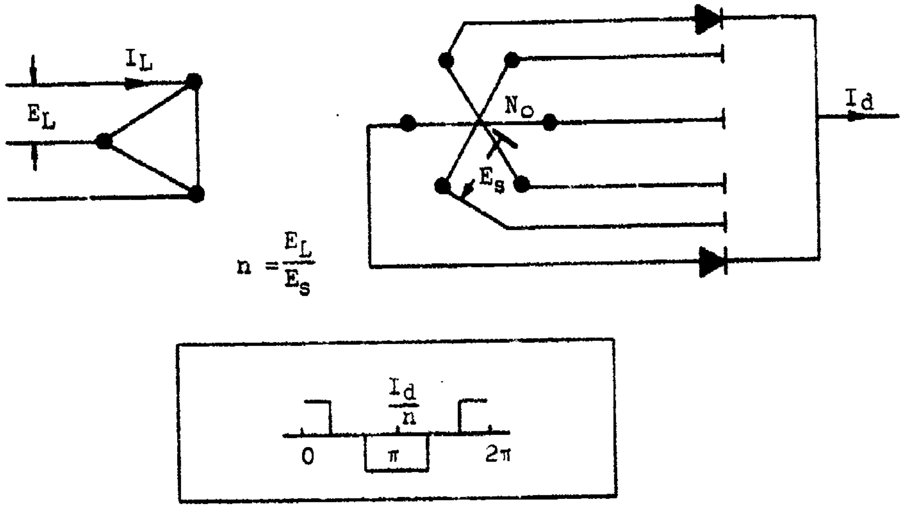
5-5.5.1.3 Common Mode Filtering

Because common mode currents (par. 3-3.3.3) usually are so much more effective in generating EMI fields than differential mode currents, in many applications line-to-ground filtering should be avoided in favor of line-to-line filtering.

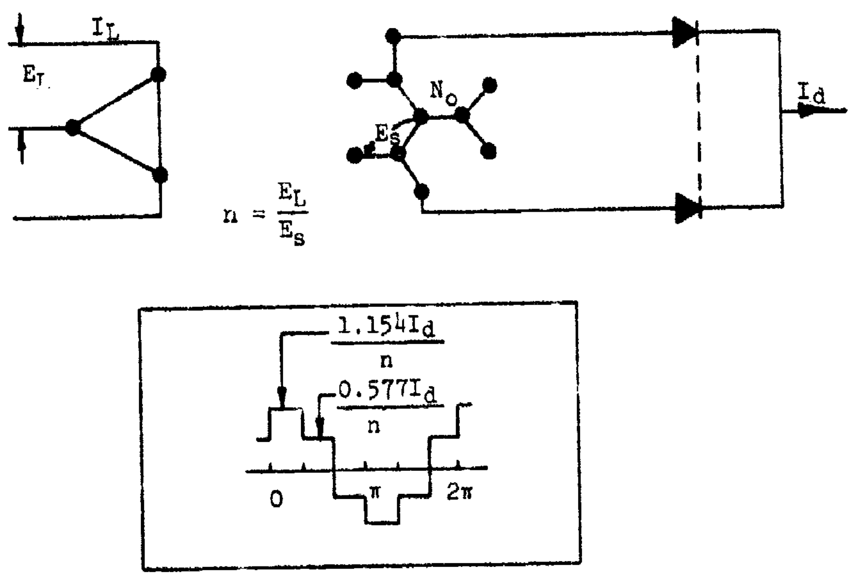
TABLE 5-7.
LINE CURRENT HARMONICS IN MULTIPHASE RECTIFIERS

ORDER OF HARMONICS	NUMBER OF PHASES					FREQUENCY, Hz
	3	6	12	24	36	
2	X					120
3						180
4	X					240
5	X	X				300
6						360
7	X	X				420
8	X					480
9						540
10	X					600
11	X	X	X			660
12						720
13	X	X	X			780
14	X					840
15						900
16	X					960
17	X	X				1020
18						1080
19	X	X				1140
20	X					1200
21						1260
22	X					1320
23	X	X	X	X		1380
24						1440
25	X	X	X	X		1500
26	X					1560
27						1620
28	X					1680
29	X	X				1740
30						1800
31	X	X				1860
32	X					1920
33						1980
34	X					2040
35	X	X	X		X	2100
36						2160
37	X	X	X		X	2220

Fig. 5-56 shows an equivalent circuit for a device in which interference voltages are generated and coupled to a two wire power source (shown within dotted lines) located over a ground plane G. The voltage source for differential mode currents is represented by e_d, and that for common mode currents by e_c; Z_1, Z_2 and Z_3 are appropriate values for the impedances of the equivalent circuit. In most two wire power supplies the differential mode voltage e_d is the dominant one, especially at harmonic frequencies up to high orders (several hundred), hence filtering of the line is best accomplished with the capacitor C connected line to line. If two separate capacitors are used, connected line to ground (as shown in Fig. 5-57), increased current flows in the ground plane — especially if the capacitors are not equal (typical capacitors of equal nominal value may be different by as much as 25%), or if the equivalent impedances Z_1 and Z_2 are not connected to the same point of the ground plane. The inductors L impede the flow of

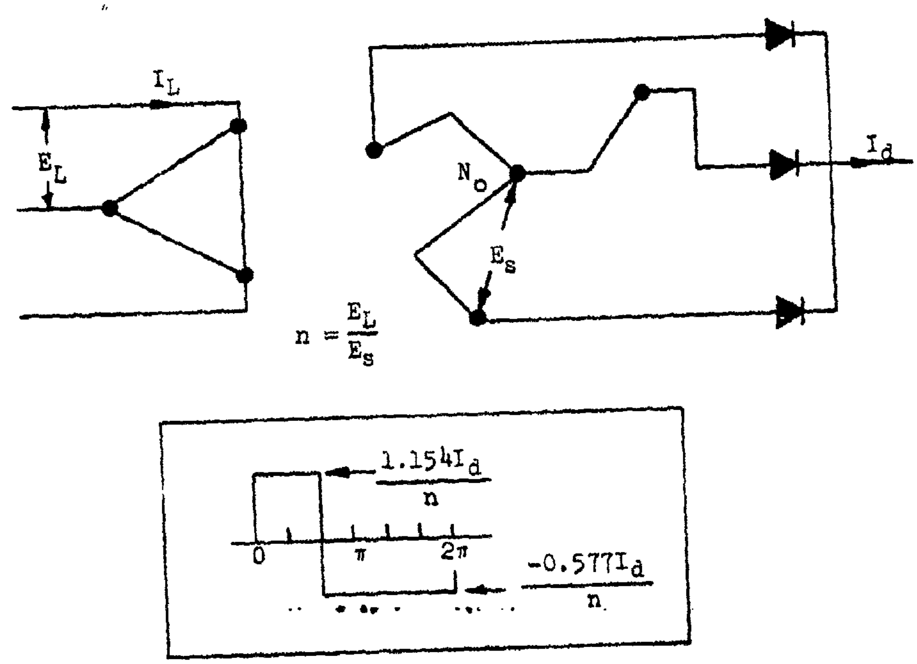


(A) DELTA/STAR

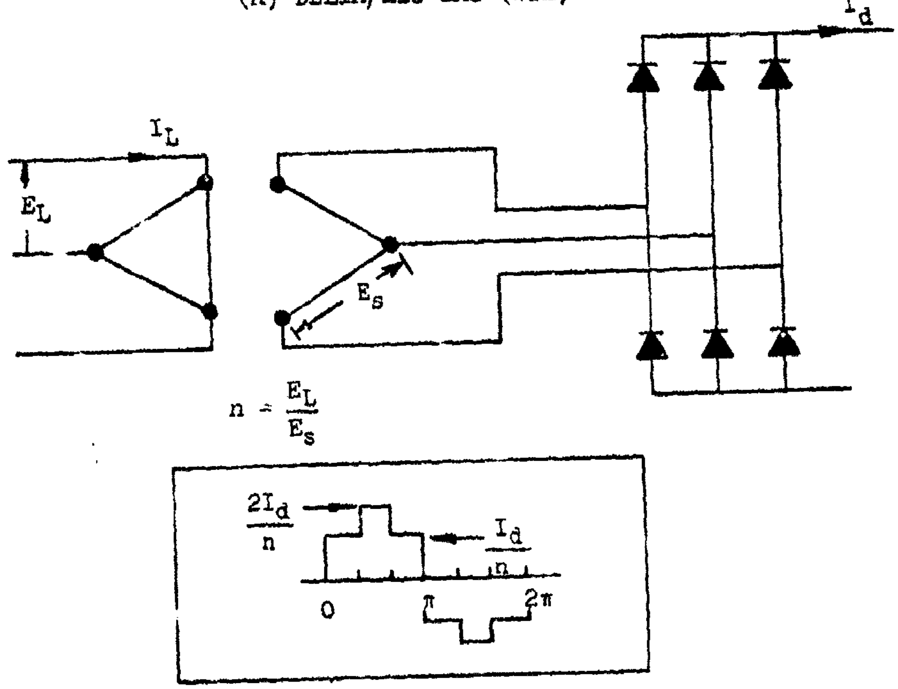


(B) DELTA/DOUBLE FORK

Figure 5-52. Six-phase Rectifiers

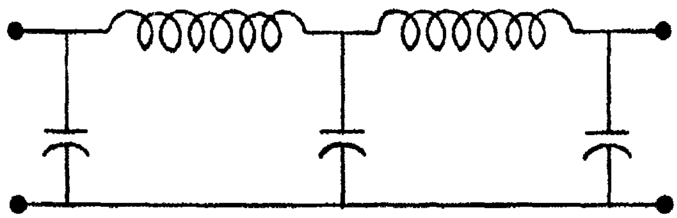


(A) DELTA/ZIG-ZAG (WYE)

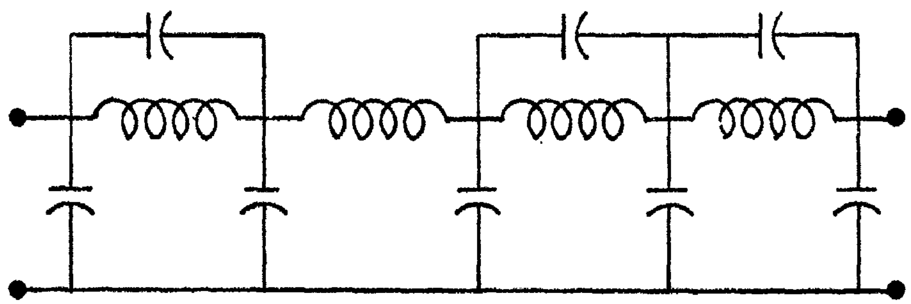


(B) DELTA/DELTA

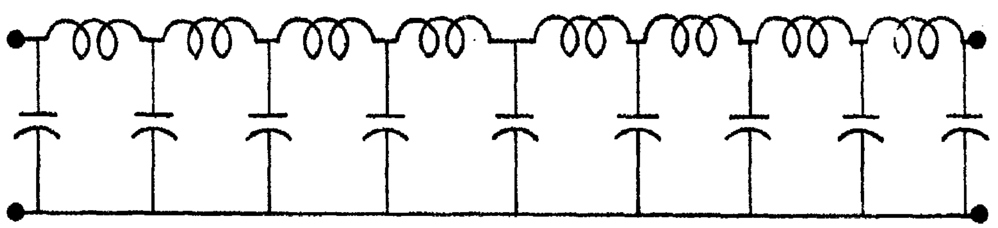
Figure 5-53. Three-phase Rectifiers



(A) Simplified Filter



(B) Typical Filter



(C) Compatible Lossy Version of the Simplified Filter

Figure 5-54. Power-line Filters

both common mode and differential mode currents. In some common mode filters the two inductors are coupled together with mutual inductance so as to maintain balance in the currents in each line (equal and opposite) as much as possible. Such an inductance arrangement is known as a common-mode choke (see par. 4-1.8 and Fig. 4-5(B)) and can present very high impedance to common-mode currents.

A symmetrical filter of this type will not be necessary if the device is well isolated from ground so that Z_3 has a very high value. In that case the currents in lines A and B will tend to be balanced and an unsymmetrical filter may be satisfactory and, of course, less expensive.

On the other hand, if the common-mode voltage e_c is high, which is likely only at the higher frequencies (above 1 MHz), or if for some reason the internal device connections are significantly asymmetrical, some line to ground filtering may be necessary. Thus low

1. 50-ohm source
   50-ohm load

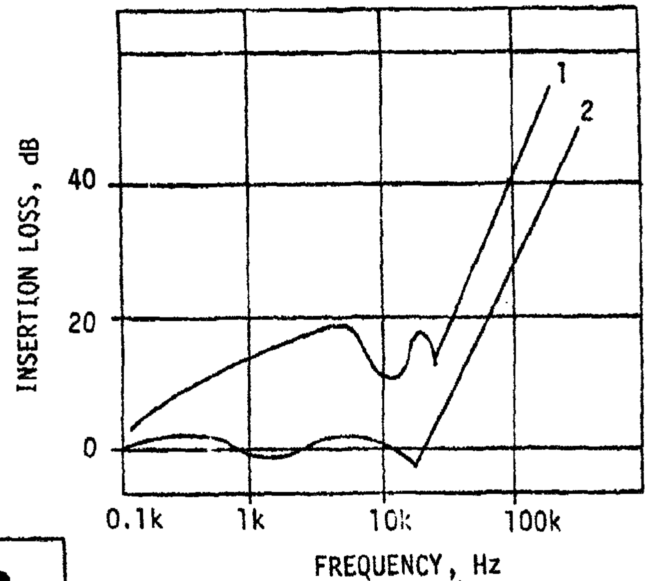
   $L_1 = L_2 = 30\ \mu H$

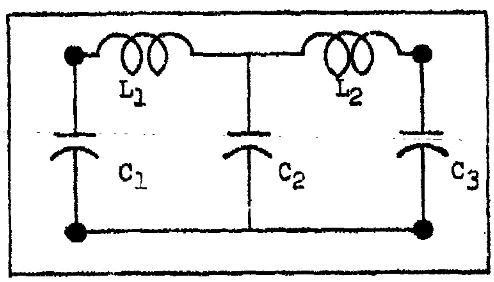
   $C_1 = C_2 = C_3 = 0.2\ \mu F$

2. 1-ohm source
   50-ohm load

   $L_1 = L_2 = 30\ \mu H$

   $C_1 = C_2 = C_3 = 0.2\ \mu F$





3. 50-ohm source
   50-ohm load

   $L_1 = L_2 - 1.6\ \mu H$

   $C_1 = C_2 = C_3 = 0.2\ \mu F$

4. 1-ohm source
   1.2-ohm load

   $L_1 = L_2 = 1.6\ \mu H$

   $C_1 = C_2 = C_3 = 0.2\ \mu F$

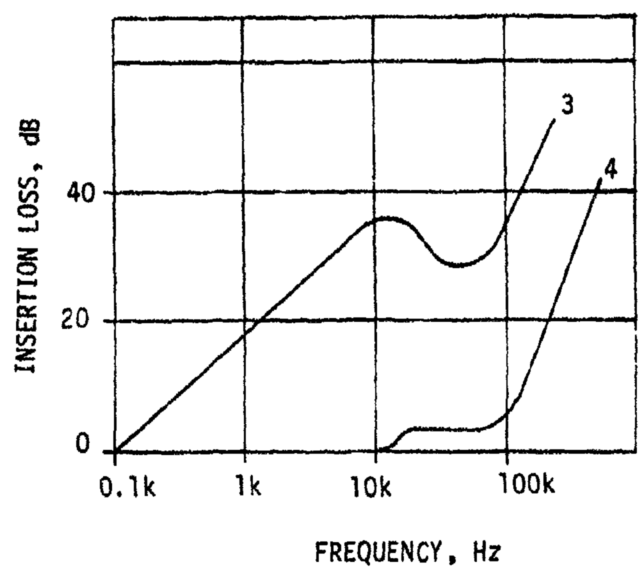


Figure 5-55. Effect of Different Load and Source Impedances on Insertion Loss

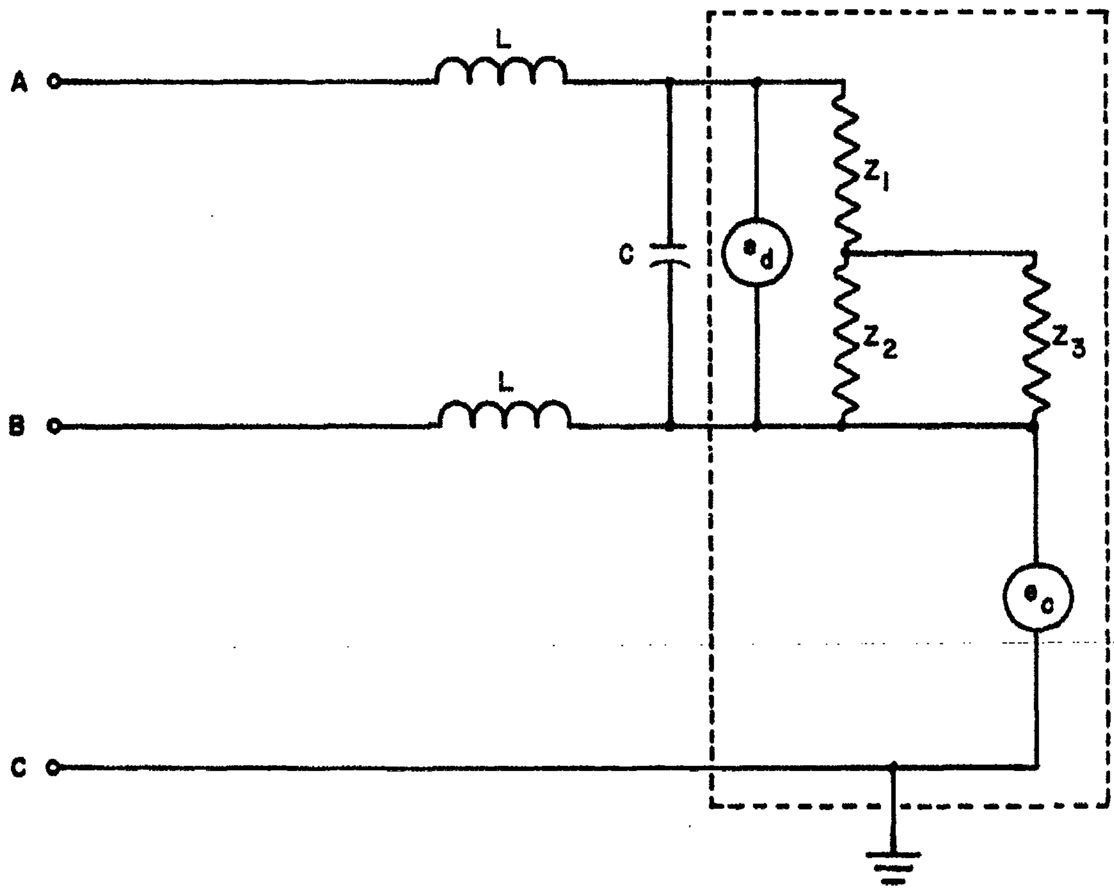


Figure 5-56. Common Mode Filtering

value capacitors (say, 0.01μF) may be required between each line and ground to obtain effective interference reduction.

In asymmetrical power sources, such as used with direct current in which the return conductor is the ground plane, only asymmetrical filters can be used.

5-5.5.1.4 Filter Installation

In arranging and mounting filters and filter components, good practice as described in par. 4-5.5 should be followed. In particular, spurious capacitive coupling between input and output leads should be prevented by judicious use of electric shields, enclosures, and feed-through capacitors. Inductors, because they may carry relatively high currents at power supply frequencies, should be enclosed in magnetic shields to prevent stray magnetic field coupling.

5-5.5.2 Shielding

Shielding (see par. 4-6) offers major wide-band protection from the power supply when it and the susceptible circuit cannot be separated far enough to attenuate the radiation sufficiently. At frequencies above 100 kHz, 40 dB of attenuation of magnetic fields can be obtained easily with a single shield. With careful design and construction, values as high as 70 dB can be obtained with a single shield. Double shielding can produce much greater shielding effectiveness *SE*, as high as 120 dB. Fig. 5-58 illustrates typical expected shielding effectiveness for single and double thicknesses of high permeability material. Note that the *SE* decreases as the frequency decreases but that it approaches asymptotic values. The shielding effectiveness for an *E*-field source is usually higher than that for an *H*-field source (see Tables 4-12, -13, and -14). Depending

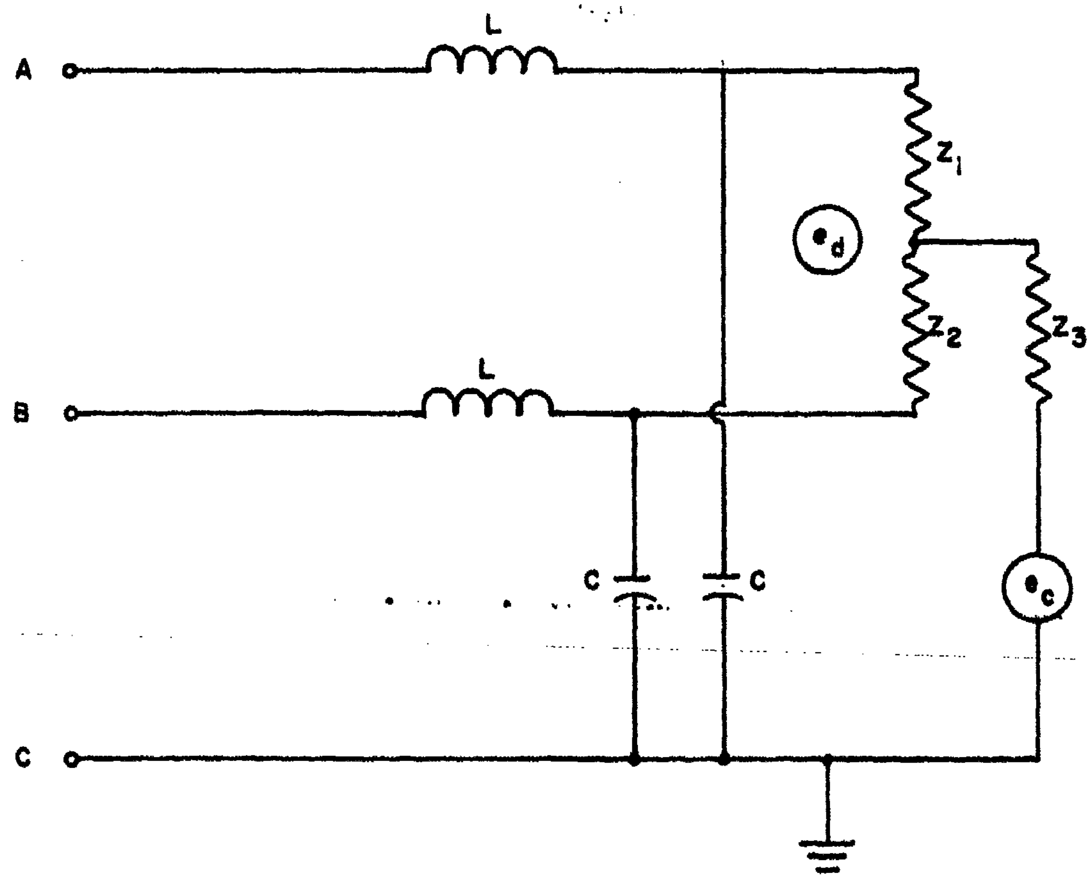


Figure 5-57. Line-to-ground Filtering

upon the material and the configuration of the shield, the *E*-field shielding effectiveness may either decrease or increase as frequency approaches zero (see Figs. 4-77, -78, and -79).

5-5.5.3 Relay Transient Suppression

Relays are frequently used for switching power circuits both on the primary (ac) and secondary (dc) sides of the supply. Since the relay control circuits are usually operated at remote locations, the transients generated will be carried on the interconnecting wiring and can couple into nearby susceptible circuits unless the relays are properly suppressed. Methods of suppression discussed in par. 5-3 are applicable here.

5-5.5.4 Voltage Regulators

5-5.5.4.1 Linear Voltage Regulators (Ref. 8)

Figs. 5-59 and 5-60 show the elements in series and shunt types of linear voltage regulators. Negative feedback is used to provide precise control of output voltage. In the frequency range where the feedback is effective, such a supply will have a very low output impedance and therefore will reduce the coupling that might otherwise be expected in a conventional supply.

Whenever feedback is present in an active circuit, it is potentially unstable. To determine whether oscillation will occur, as well as the frequency of oscillation, the entire circuit must, of course, be modeled with high-frequency parameters. To determine if the system will be stable, the Routh-Hurwitz method, the Nyquist plot, and the Bode plot or other methods could be used to analyze the system.

5-5.5.4.2 Prevention of Oscillation

Careful packaging is one of the most important steps in preventing oscillation. Without careful component layout and minimization of lead lengths, it is very difficult to prevent oscillation without seriously compromising the ripple attenuation characteristics at the regulator. The designer of a regulated circuit can consider a number of possible approaches to prevent oscillation, such as reducing transistor gain at high frequency, bypassing the filter capacitor at the output of the unregulated supply with a capacitor with short leads; and using a relatively low value, low inductance bleeder resistor connected from the output of the regulated supply to the ground side of tfhe load.

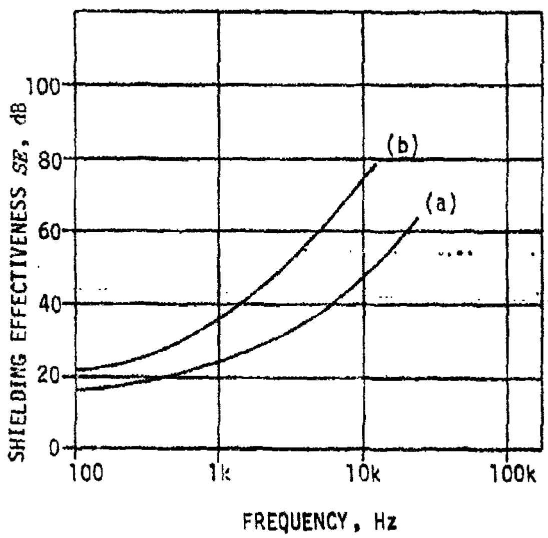


Figure 5-58. Magnetic Shielding Effectiveness for (A) A Single Thickness, and (B) a Double Thickness of Material With High Permeability

5-5.5.4.3 Switching Regulators

The switching regulator shown in Fig. 5-61 performs the voltage regulation function by means of a switching transistor that chops the unregulated input voltage in such a manner as to maintain the average output voltage level constant. The output of the regulator then consists of a series of rectangular pulses. A low-pass filter is used to smooth these pulses to a relatively constant dc level. A portion of the output voltage is compared with a reference voltage, and an error voltage is generated. The error voltage is fed back to a voltage-controlled multivibrator. The duty cycle of the pulse output from the multivibrator depends upon the magnitude and phase of the error voltage. The output of the multivibrator controls a driver circuit that, in turn, controls the switching transistor. In this manner, voltage regulation is achieved by what is, in essence, a pulse-duration modulation process. For highest efficiency, the switching transistor of the regulator must switch as fast as possible between cutoff and saturation. This method of voltage regulation can result in efficiencies as high as 95%. In contrast, the linear

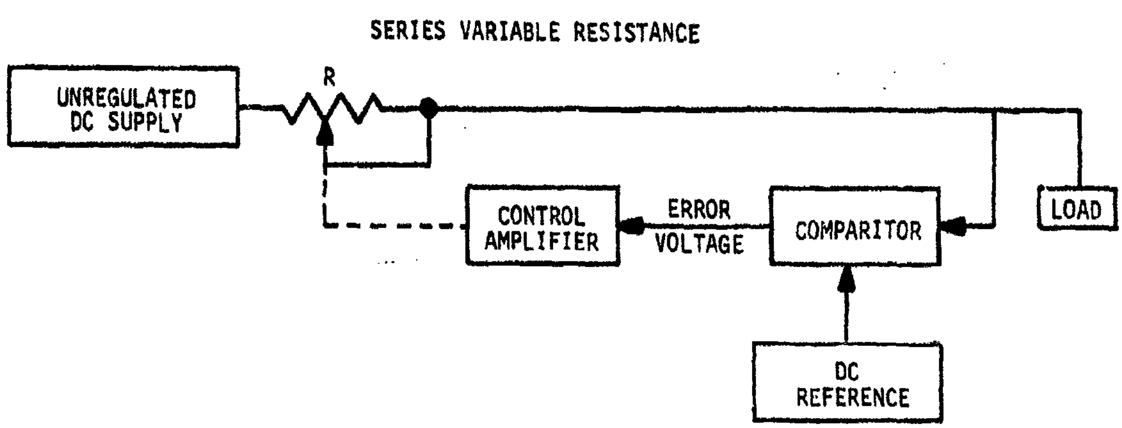


Figure 5-59. Series Regulator Block Diagram

voltage regulator is in a continuously "on" position, usually resulting in greater power losses and resultant efficiencies of only 20% to 40%. However, efficiencies of 80% or 90% can be obtained with linear voltage regulators.

5-5.5.4.4 Applications

Switching regulators are used in applications where space and weight are limited, and high efficiency is needed. Because of the high efficiency, smaller power supplies and less cooling hardware are required for the regulator itself since power dissipation is much less than for linear regulators. However, as its name suggests, the switching regulator operates on the principle of switching circuitry with its efficiency dependent on its rectangular wave characteristics. An ideal rectangular wave generates harmonics which decrease at a rate of 20 dB/decade beyond the frequency of oscillation. If the radio interference generated by the switching regulator will cause problems, it must be suppressed. However, radio interference suppression results in a decrease of efficiency since filters must be used and filters dissipate power. By predicting whether or not the radio interference caused by the unsuppressed switching regulator will create any interference problems, the degree of suppression can be determined. Efficiencies of 60% to 80% can still be achieved by building the switching regulator with radio interference suppression as a part of the circuit.

5-5.5.4.5 Transient Suppression

If the radio interference generated by the switching regulator is predicted to cause problems, there are

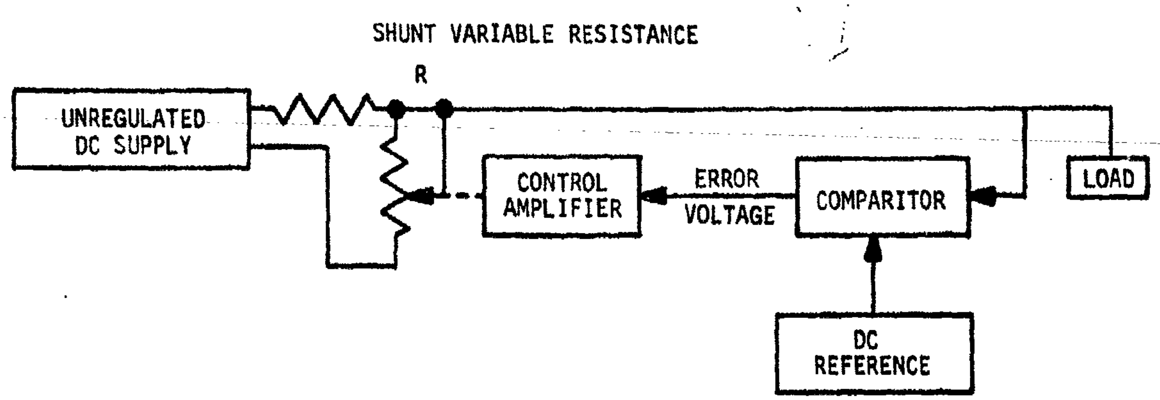


Figure 5-60. Shunt Regulator Block Diagram

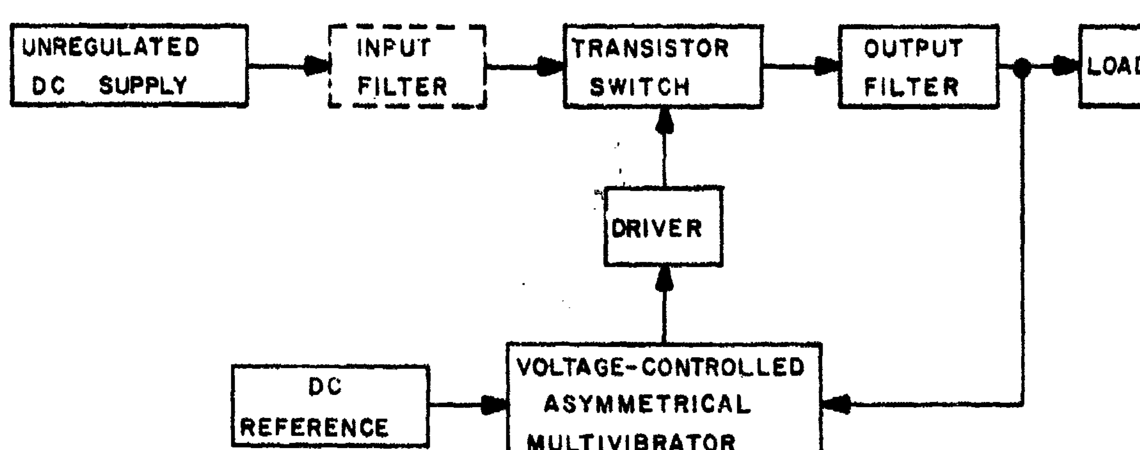


Figure 5-61. Switching Regulator Block Diagram

various methods available to suppress the radio interference and still maintain relatively high efficiency. Several different types of circuits can be incorporated into the circuits of the regulator. The following are some examples of circuits that might be incorporated:

a. A filter consisting of an inductor and a shunt capacitor could be placed on the input line from the unregulated supply to attenuate the ripple current which would otherwise be reflected back into the line (Fig. 5-62).

b. Inductors can be placed on both input lines to allow the regulator to float. Also, a transformer can be connected in series which permits the input to remain constant (see Fig. 5-63).

c. Two low impedance capacitors in series, center-tapped to ground, can be placed across the line along with two high impedance inductors bifilar wound on the same core in series with the shunt capacitor. This filter attenuates ac currents and protects against transients (see Fig. 5-64).

d. Since the rectangular wave produced by the switching transistor causes noise due to the harmonic content of the wave, another technique for radio interference reduction is to round the corners of the pulse at the transistor. By rounding off the rectangular pulse, the drop-off of the harmonics can be reduced from 20 dB/decade to approximately 80 dB/decade or greater. One possible circuit which can be used is shown in Fig. 5-65(A). Care must be taken in any type of rounding circuit since increased transistor dissipation may occur. Transistor safe operating area is the main consideration in the design of any switching regulator. Inductive load lines or excessive collector voltage, even for tens of nanoseconds, can result in transistor destruction.

e. Further protection against interference in the higher frequency range can be added by the use of ferrite beads. Ferrite beads are relatively small and can be slipped onto component leads. No dc losses are introduced.

f. In some situations, voltage transients may reach the input of the switching regulator. These voltage transients may be fed through the regulator and cause a malfunction or destruction of the load. A suppressor circuit using a Zener diode or varistor to limit the transient can be designed and placed on the input to the regulator (see Fig. 5-65(B)). Such circuits must be designed to handle transients of the maximum

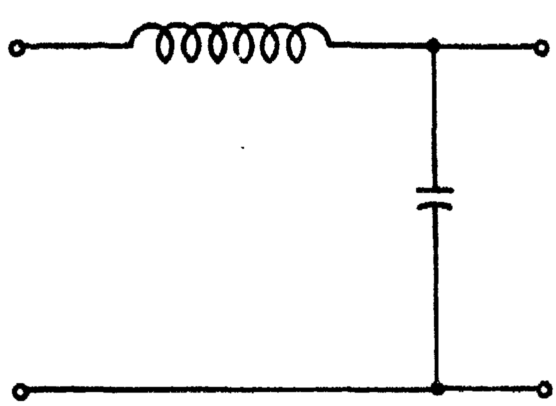


Figure 5-62. ***L-C*** **Transient Filter**

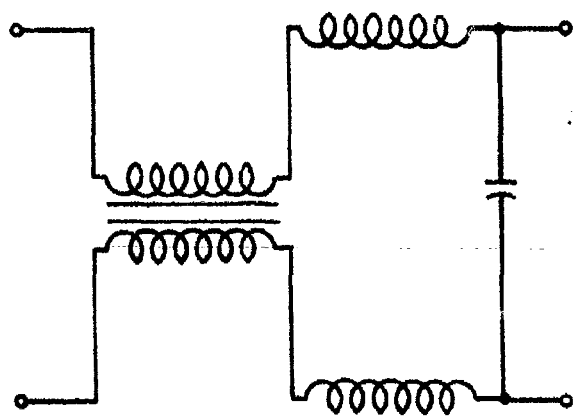


Figure 5-63. Balanced *L-C* Transient Filter

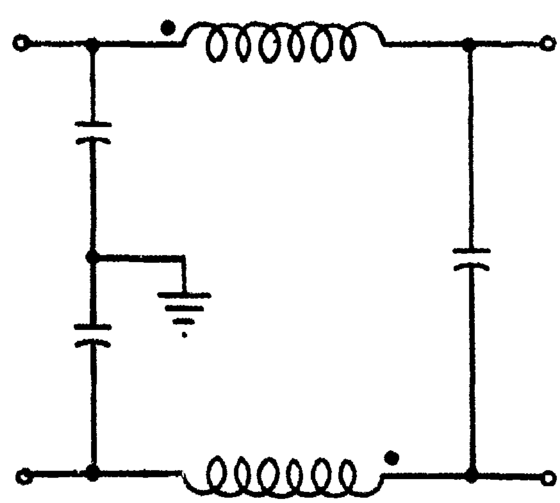
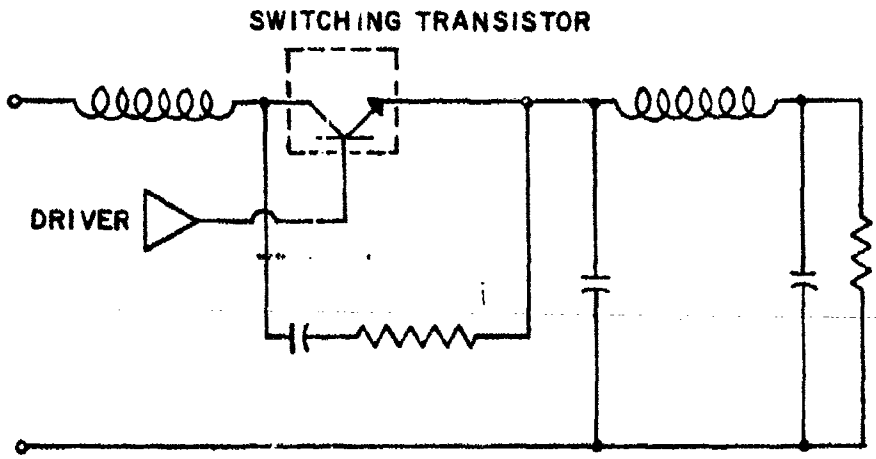


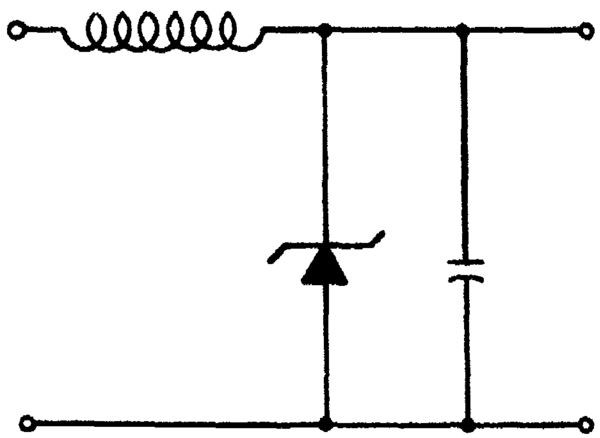
Figure 5-64. Bifilar Transient Filter

magnitude and duration that can occur; otherwise, the nonlinear element may be destroyed. Varistors have relatively high energy handling capabilities although their "knee" is not sharp when compared with a Zener diode.

g. Abrupt switching due to inductor core saturation also should be considered. Inductor cores should be made of material which has "soft" saturation characteristics. If abrupt core saturation occurs, excessive peak currents are developed in the switching transistors. Material with high permeability, low losses, and high electrical resistivity — such as powdered molybdenum-permalloy cores — should be considered. With this type of core material, excessive



(A) Pulse Rounding Circuit



(B) Transient Clamp

Figure 5-65. Transient Suppression Circuits

currents will cause only a gradual increase in switching frequency since permeability reduction is gradual with excessive currents.

h. It is important to note that any diodes or capacitors which run to ground should be grounded separately from other circuit components. Large current transients flow through these diodes and capacitors, which can develop large voltages across relatively short lengths of wire. Layout is one of the most important considerations in reducing interference. Stray capacity to ground can cause noise currents to circulate throughout the unit, and careful control of these currents is necessary to prevent interference with other circuits. Short lead lengths and careful component placement (packaging) can avoid these problems.

5-6 VEHICLES AND OTHER ENGINE DRIVEN EQUIPMENT

5-6.1 INTRODUCTION

The term "vehicles" covers a wide range of equipments; however, common sources of interference permit generalizations in the application of suppression techniques. Vehicles generally have ignition systems, charging circuits, starting circuits, switching devices, horns, and windshield wipers as potential interference sources. Tactical vehicles, in addition to the interference-producing components common to all vehicles, have a variety of accessory equipments — such as gun-tracking mechanisms, ventilating blowers, personnel heaters, and auxiliary engine generators — that complicate the suppression problem. For example, basic armored vehicles have been modified as communication centers and as complex weapon systems. These systems employ electrical drive servos, complete radar fire control, and numerous other potential sources of interference. Typical elements of military electrical and electronic systems that are associated with or part of vehicles are shown in Table 5-8 (Ref. 9).

When missiles are mounted on vehicles, the possibility of inadvertent launching due to EMI must be considered carefully, since: (1) missile launching is usually done with an electric initiator-electroexplosive device (EED), and (2) EMI could upset the missile control systems that are dependent on ratio, radar, and computer circuits. Protection against this kind of hazard is covered in Ref. 10. Thus the complexity of the interference reduction applications is dependent upon the equipments installed aboard the vehicle.

The discussion in this paragraph applies also to devices normally not mounted on vehicles but which utilize engine generators for which the suppression measures discussed are appropriate, such as motor generator sets.

5-6.2 INTERFERENCE GENERATION

The sources of interference in vehicles may be categorized as:

a. Ignition systems including spark plugs, ignition coils, breaker points, and magnetos

b. Rotating machinery including dc generators, motors, and ac alternators

c. Switching devices including ac and dc regulators, solenoids, and relays

d. Static electrical discharges, including those from tank treads

e. Intentional emitters including communications transmitters, radar transmitters, and local oscillators of receivers when their emission appears at undesirable places as the result of conductive, inductive, or radiated coupling

f. Miscellaneous sources including electrical sending units for gages, instruments, starting systems, turn signals, horns (some of which are of short duration interference), windshield wipers, malfunctioning warning devices, and personnel heaters

g. Special tactical equipment.

The levels of interference produced by ignition were discussed in par. 3-1.3.2.4. Interference currents and voltages produced by the other devices are similar to those which have been discussed in Chapter 3 and in pars. 5-2 and 5-3.

5-6.3 SUPPRESSION AND CONTROL TECHNIQUES

In determining required suppression techniques three applications should be distinguished: (1) engine drive equipment such as motor generator sets which may or may not be mounted on vehicles, (2) vehicles for general military use (not tactical) which usually are suppressed to SAE STD J551, and (3) technical vehicles which are suppressed to MIL-STD-461 and also usually are waterproofed.

5-6.3.1 Nonignition Equipment

Except for the suppression of the ignition system, radio-interference suppression techniques for use on vehicles are those described in Chapter 4.

Proper bonding and the installation of bypass capacitors will suppress most sources adequately. When necessary, more sophisticated filters, shielding, and special physical layouts may be used. Shielding of various circuits is necessary to attenuate high-frequency interference. However, it is necessary that

TABLE 5-8. TYPICAL ELEMENTS OF MILITARY ELECTRICAL SYSTEMS (Ref. 9)

A. COMMUNICATION SYSTEMS
  1. Radio systems
    a. Transceivers
    b. Receivers
    c. Antennas
  2. Intercom systems
    a. Amplifiers
    b. Control Sets

B. AUTOMOTIVE SYSTEMS
  1. Engine starting system
    a. Starter motor
    b. Starter relay
    c. Slave receptacle
    d. Neutral start interlock
  2. Engine ignition system
    a. Distributor
    b. Coil
    c. Spark plugs
    d. Magneto
    e. Glow plug
  3. Vehicle operational control systems
    a. Steering
    b. Mode selection
    c. Gear selection
    d. Fuel pumps & controls
    e. Electric winch
    f. Ramp control
    g. Electrohydraulics
    h. Master switch
  4. Vehicle lighting system
    a. Interior lights
    b. Exterior lights
    c. Trailer receptacle
  5. Vehicle instrumentation system
    a. Temperature, RPM, pressure, speed, and voltage
    b. "ON" indicators
    c. Malfunction warnings
    d. Instrument lighting

C. AUXILIARY ELECTRICAL POWER SYSTEM
  1. Standby power for vehicle systems
  2. Power for electronic equipment
  3. Power for welders
  4. Power for hand tools

D. ENVIRONMENTAL & DAMAGE CONTROL SYSTEMS
  1. Vehicle damage control systems
    a. Fire suppression
    b. Bilge pumps
  2. Vehicle environmental control systems
    a. Personnel & coolant heaters
    b. Coolers
    c. Ventilators
    d. Windshield defrosters
    e. Windshield wipers
    f. Engine winterization systems
    g. Chemical, biological, and radiological protection

E. MISCELLANEOUS SYSTEMS
  1. Suspension systems
    a. Optical sensor
    b. Sonic sensor
    c. Laser sensor
    d. Electrohydraulic actuators
  2. Electric drive systems
    a. Traction motors
    b. Motor controller
    c. Actuators

F. WEAPON SYSTEMS
  1. Armament systems
    a. Charging system
    b. Ammunition feed system
    c. Casing and link ejection
  2. Fire control systems
    a. Vision devices
    b. Laser range finder
    c. Ballistic computer
    d. Weapon drives, azimuth and elevation
    e. Stabilization
    f. Search lights
  3. Firing systems
    a. Safety interlocks
    b. Weapon firing circuits
  4. Supporting systems
    a. Power distribution
    b. Ventilating
    c. Instruments and indicators
    d. Lighting
  5. Missile systems
    a. Target tracking
    b. Missile tracking
    c. Missile command computer
    d. Missile guidance data link

these items be bonded and shielded properly as discussed in par. 4-7. Efficient use of existing shielding — such as afforded by housings, dust covers, or partitions — is an excellent means of augmenting a suppression system.

In the design of the physical layout, the source or circuits in which interference is present should be separated as far as practicable from susceptible parts of the same equipment and from external antenna circuits of any kind.

5-6.3.1.1 Rotating Machinery

The techniques of suppressing rotating machinery mounted in vehicles are those described in par. 5-1.1.

Those principles especially apply to large separate ac generators used for auxiliary power supplies to operate large electrical or electronic equipments and also to the generator used with the vehicle engine for battery charging.

A typical vehicular alternator schematic is shown in Fig. 5-66. The major sources of EMI in this example are the slip rings supplying the rotating field winding and the harmonic content of the rectified alternator output. As described in par. 5-1.1, shielding of the alternator is accomplished by its housing, and the original design of the alternator housing should incorporate the principles of good shielding. Low-impedance paths between sections of the housing, adequate bonding, and screening of ventilating louvers should all be provided if the overall suppression system is to be effective. As in dc generators, a means of escape from the shield for interference is the alternator shaft. It can be prevented by the application of a brush which rides on a special slip ring or directly on the shaft.

Most alternators have a stationary armature and a rotating field. This field can be provided by a permanent magnet which requires no field current, or by a winding on the rotor which requires slip rings and associated brushes carrying relatively little current. Because commutation need not be considered in the selection of brushes, a much wider choice in brush pressure, size, and material is permitted.

Capacitors are usually the most effective suppression components and the most economical. These serve to bypass interference due to brush action on the slip rings as well as reducing the rectifier harmonics from the armature.

Currently available commercial vehicular alternators are designed to provide 12 V dc and are suppressed to meet SAE STD J551. Military requirements dictate the use of 24-V dc alternators. Two are available — one is a 60-A alternator-rectifier-regulator system, and the other is a 100-A alternator-rectifier-shielded cable-regulator-EMI filter system. In order to obtain high electrical capacity and low emission, the military alternators are larger and heavier than their commercial counterparts.

5-6.3.1.2 Switching Devices

This paragraph discusses, as an example, the suppression of an electromechanical voltage regulator. Solid-state voltage regulators also are used in military vehicles. In addition, there are switches and relays serving various purposes, which are also discussed in par. 5-3.

Electromechanical voltage regulators are installed commonly in vehicles, aircraft, watercraft, locomotives, power units, and a variety of other equipments. They are used to regulate both dc and ac generators. A wiring diagram of a typical vibrating-type voltage regulator used in battery-charging systems is shown in Fig. 5-67. The regulator has two functions which produce radio interference — the make-and-break action of the contact points with the rapid variations of impedance resulting from arcing,

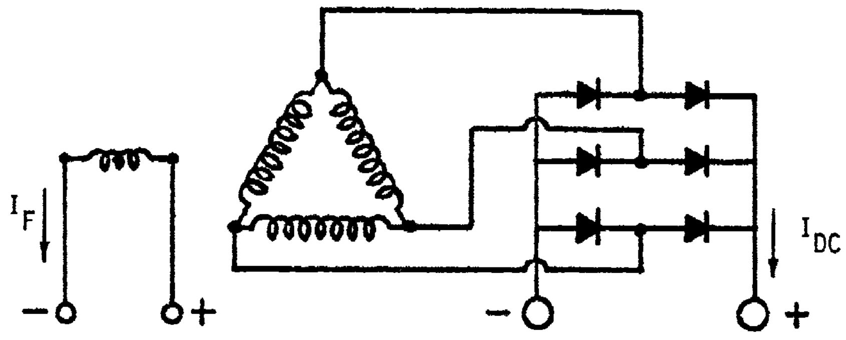


Figure 5-66. Typical Vehicular Alternator Schematic

and the switching action of the contact points which causes a transient due to the collapse and surge of voltage. The resistors in the circuit aid in suppressing both of these sources of interference (similar to the resistor-suppressors in an ignition circuit); however, they do not eliminate it entirely. To suppress the residual interference, shielding and capacitors are used.

Fig. 5-68 shows a voltage-regulator installation illustrating the application of shielding and capacitors where waterproofing is not necessary. The field and armature leads to the generator are covered with shielding braid to prevent radiation of interference created by the generator as well as by the regulator. A capacitor is connected to the armature terminal of the regulator. The interference present at the "B" terminal will be residual interference generated by the regulator, and this is bypassed by an additional capacitor. The lead to the battery, although rather long in most installations, generally does not require shielding.

5-6.3.1.3 Static Electric Discharges

Parts of the vehicle that can become electrically insulated from each other through corrosion of metal parts or shock mounting using nonconducting materials must be bonded with conducting material in accordance with the discussion in par. 4-7. A typical bonding installation is shown on Fig. 5-69. Usually, bonding is not required between welded parts, but such a provision may be necessary on panels supporting electrical devices even though bolted directly to the chassis of the vehicle.

Track static is particularly difficult to deal with since bonding straps across the contacting parts are not possible. The magnitude of the interference generated is highly dependent on the type of terrain over which the vehicle is moving and on atmospheric conditions. Since vehicle mounted communication systems operate at VHF or alone where the static induced interference will be directional, it is possible to construct the vehicle to provide shielding between it and the vehicle antennas.

5-6.3.2 Ignition System Suppression

A typical spark ignition system is shown in Fig. 5-70. EMI results from steep transients (see par. 3-1.3.2.2) initiated by the firing of the spark plugs, and arcing of the distributor cap and breaker points. These transients may result in interference being radiated by any of the ignition system components or the interconnecting wiring. The spectrum of this radiation extends into the microwave region (see par. 3-1.3.2.4).

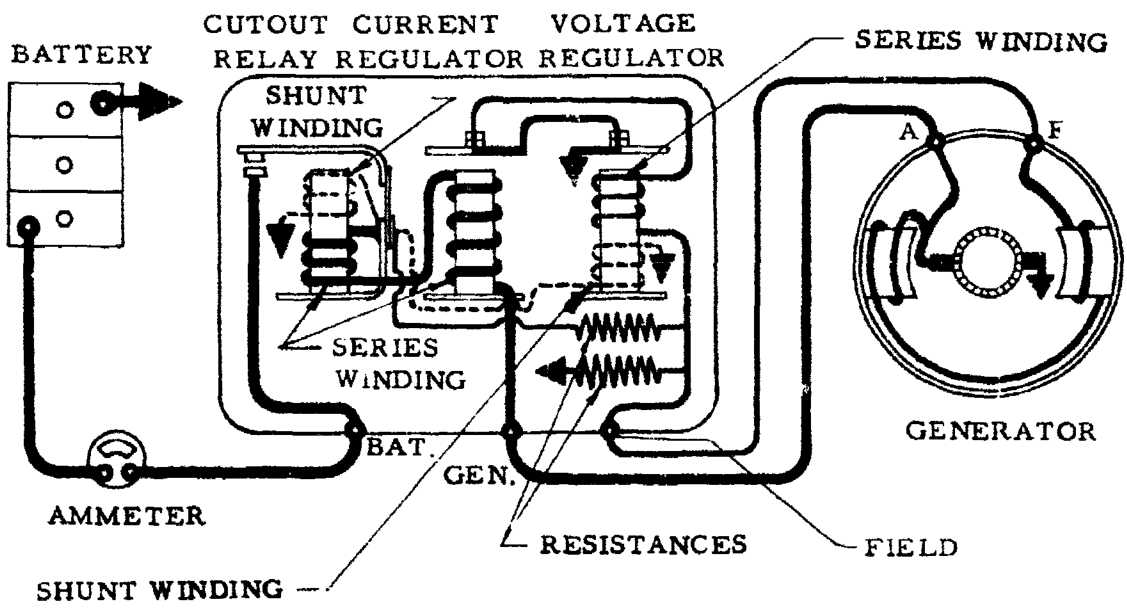


Figure 5-67. Vibrating Type Voltage Regulator

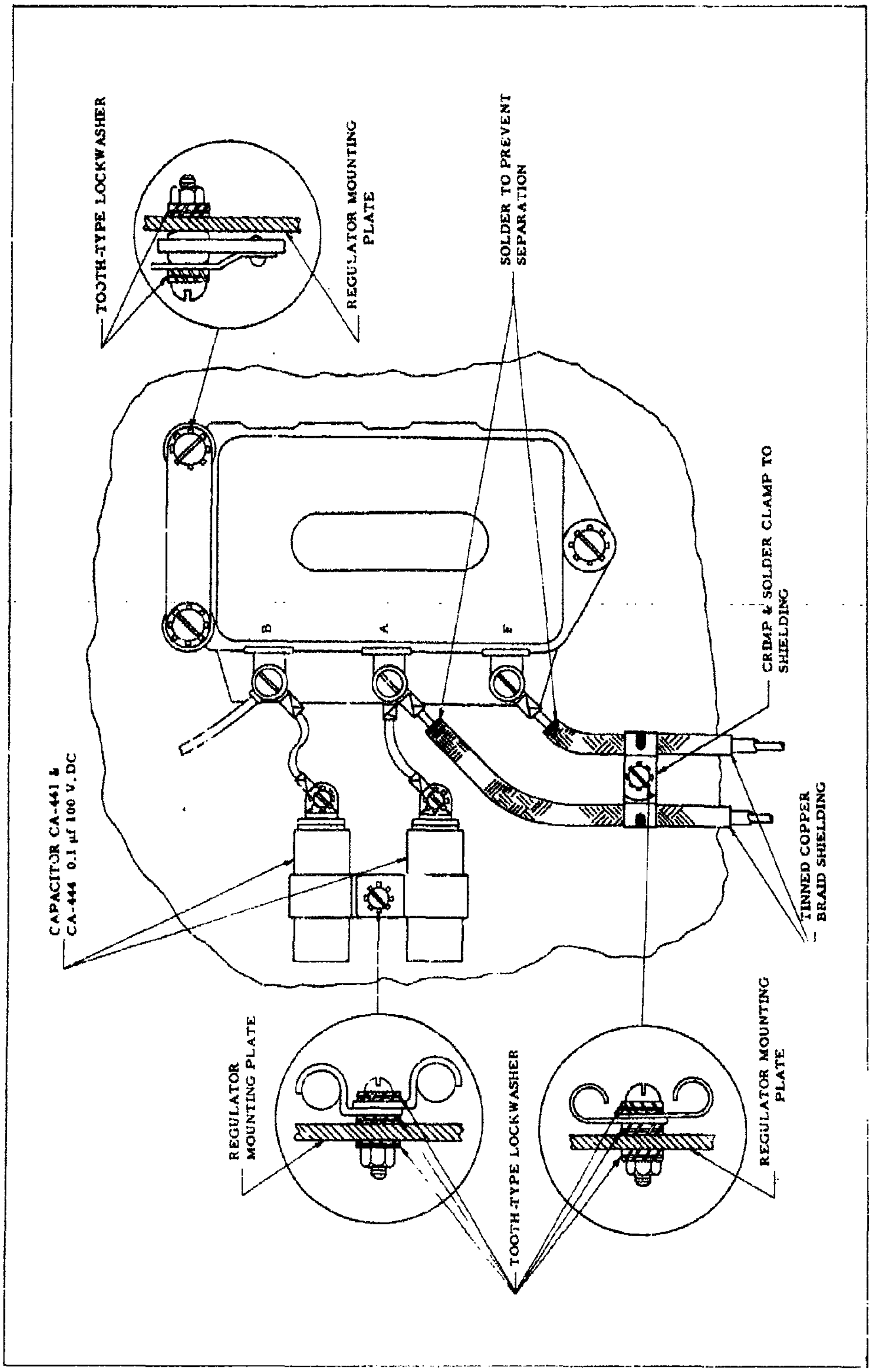


Figure 5-68. Suppression Applied to a Voltage Regulator

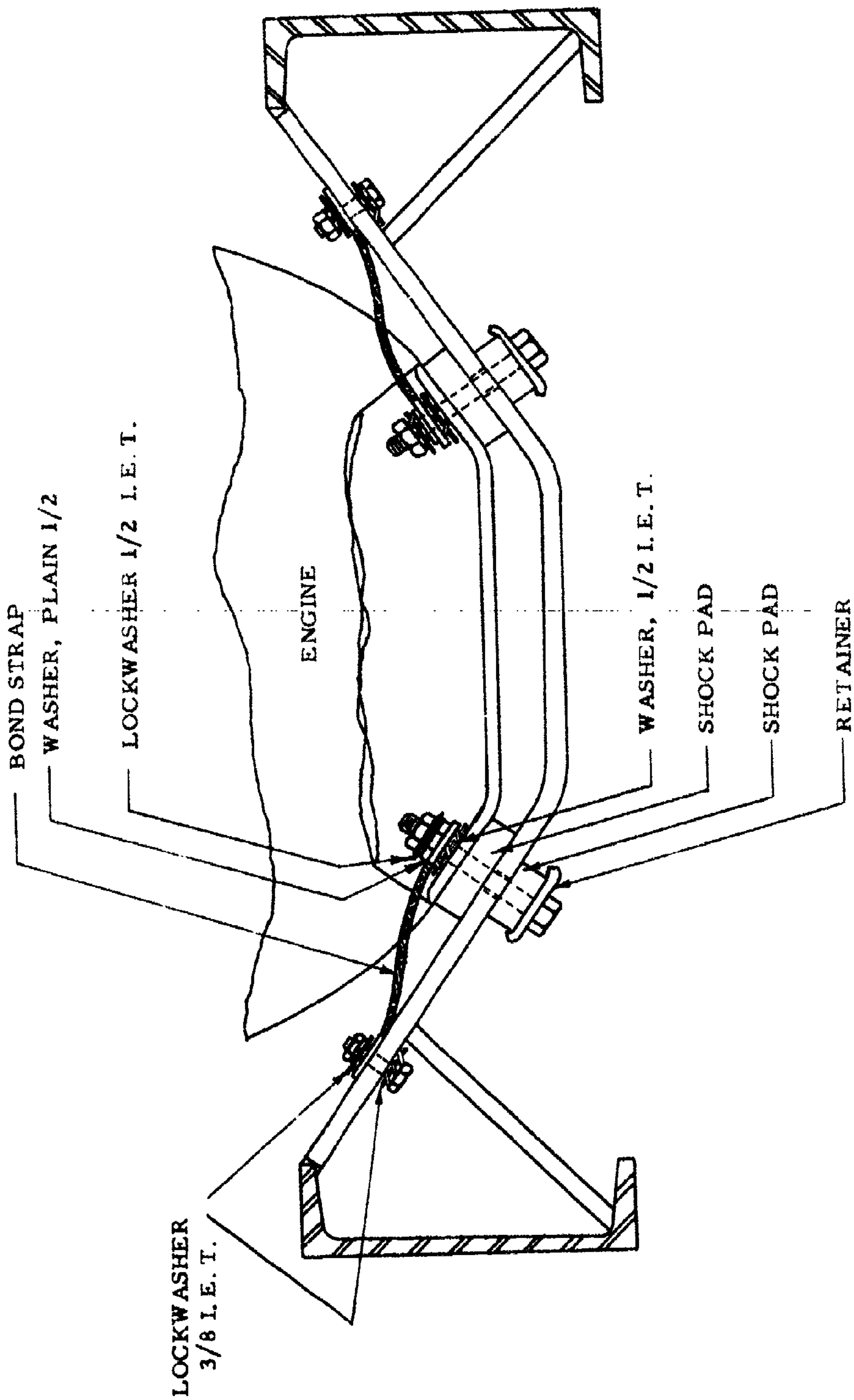


Figure 5-69. Bonding of Shock-mounted Engine Rear Mount

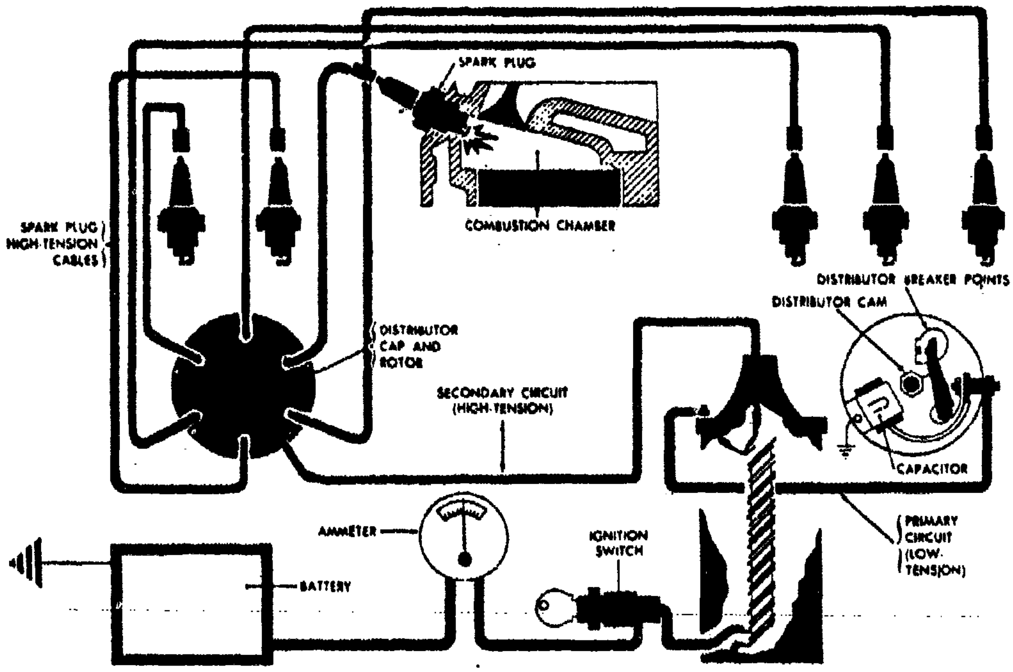


Figure 5-70. Battery Spark Ignition System

A typical electronic ignition system eliminates one of the arcing points by substituting an impulse generator for the breaker point but, as shown in the examples that follow, does not necessarily decrease the EMI generated.

Fig. 5-71 (Ref. 11) shows the EMI generated by a 1-1/4-ton commercial truck having an electronic ignition system (with a commercial 12-V alternator) designed to meet the requirements of SAE STD J551. Note that it did not meet the requirements of MIL-STD-461. Fig. 5-72 (Ref. 12) shows the EMI generated by a commercial truck having a breaker point nonelectronic-type ignition (with a commercial 12-V alternator). Although there are differences in the detail of the curves in Fig. 5-71 and Fig. 572, there are no remarkable differences in the overall interference levels found.

5-6.3.2.1 Resistor Suppression

Suppression resistors serve to limit the rise time of the ignition transient and damp oscillations in the circuit. A resistance value of 10 kΩ has been found (Ref. 13) to suppress effectively both spark plugs and distributor gaps without degrading ignition performance. The transients generated by the breaker points, however, cannot be suppressed by means of a series resistor since this degrades system performance. Some suppression of both conducted and radiated interference can be achieved using a bypass capacitor connected to the circuit near the breaker points and connected to ground.

A suppression resistor should be located physically close to the gap where the transient is generated. Even a short length of conductor between suppressor and arc may radiate considerable interference. In commerical vehicles built to meet SAE Standard J551 the resistance is distributed along the length of the ignition cable.

Suppression of the spark plug transient may be achieved by using integrally suppressed spark plugs. These units contain resistors internal to the plug and may be shielded or unshielded. A shielded unit is shown on Fig. 5-73. Suppression of the distributor gap is achieved by the use of an integrally suppressed distributor having 10-kΩ resistive element mounted in the distributor cap (usually in the central tower) or on the rotor.

Because of the broad ignition transient spectrum, the resistance of suppressors must not change with frequency. Wire wound resitors are not satisfactory in this service.

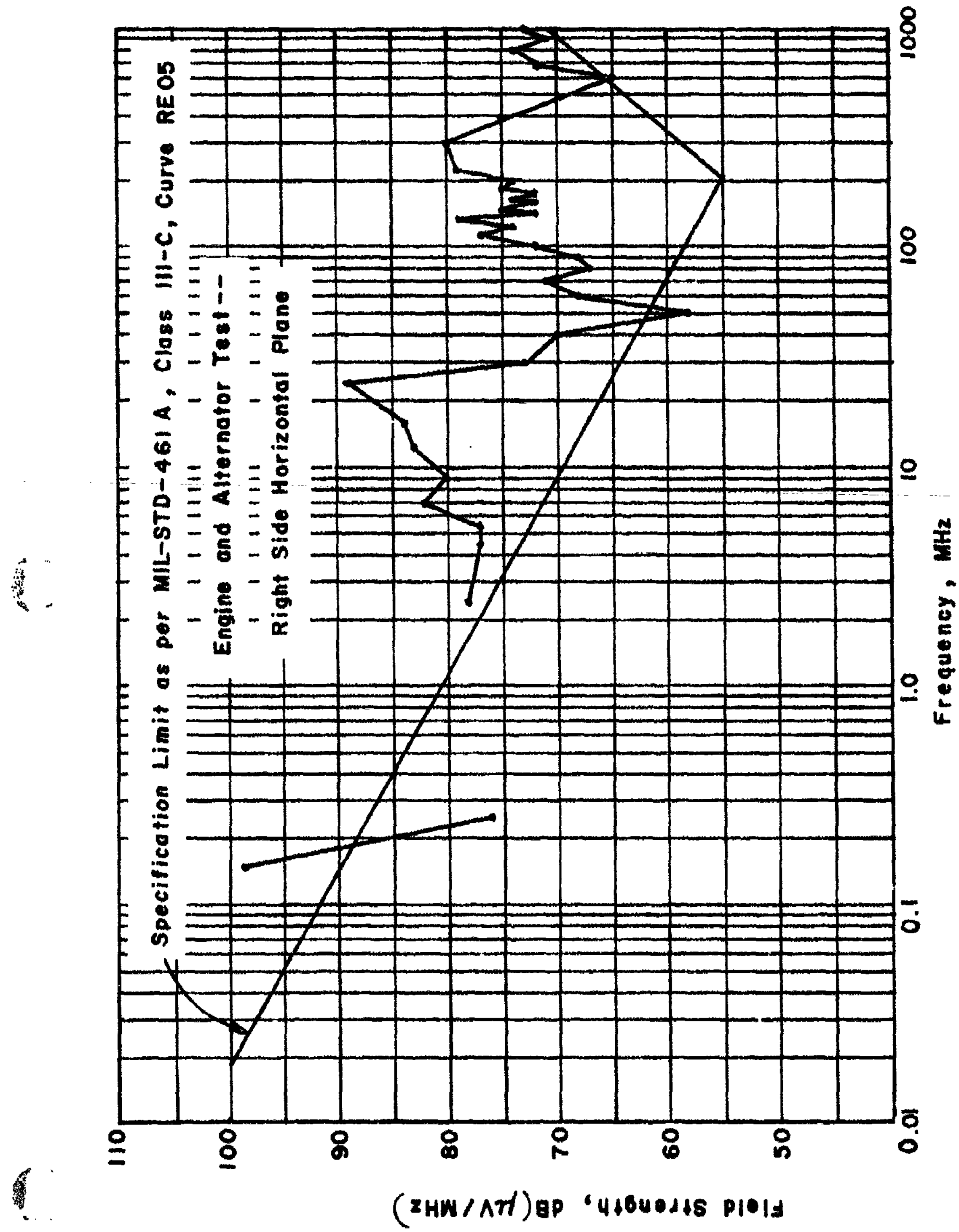


Figure 5-71. Emission from a Commercial Truck Having a Semiconductor Ignition System

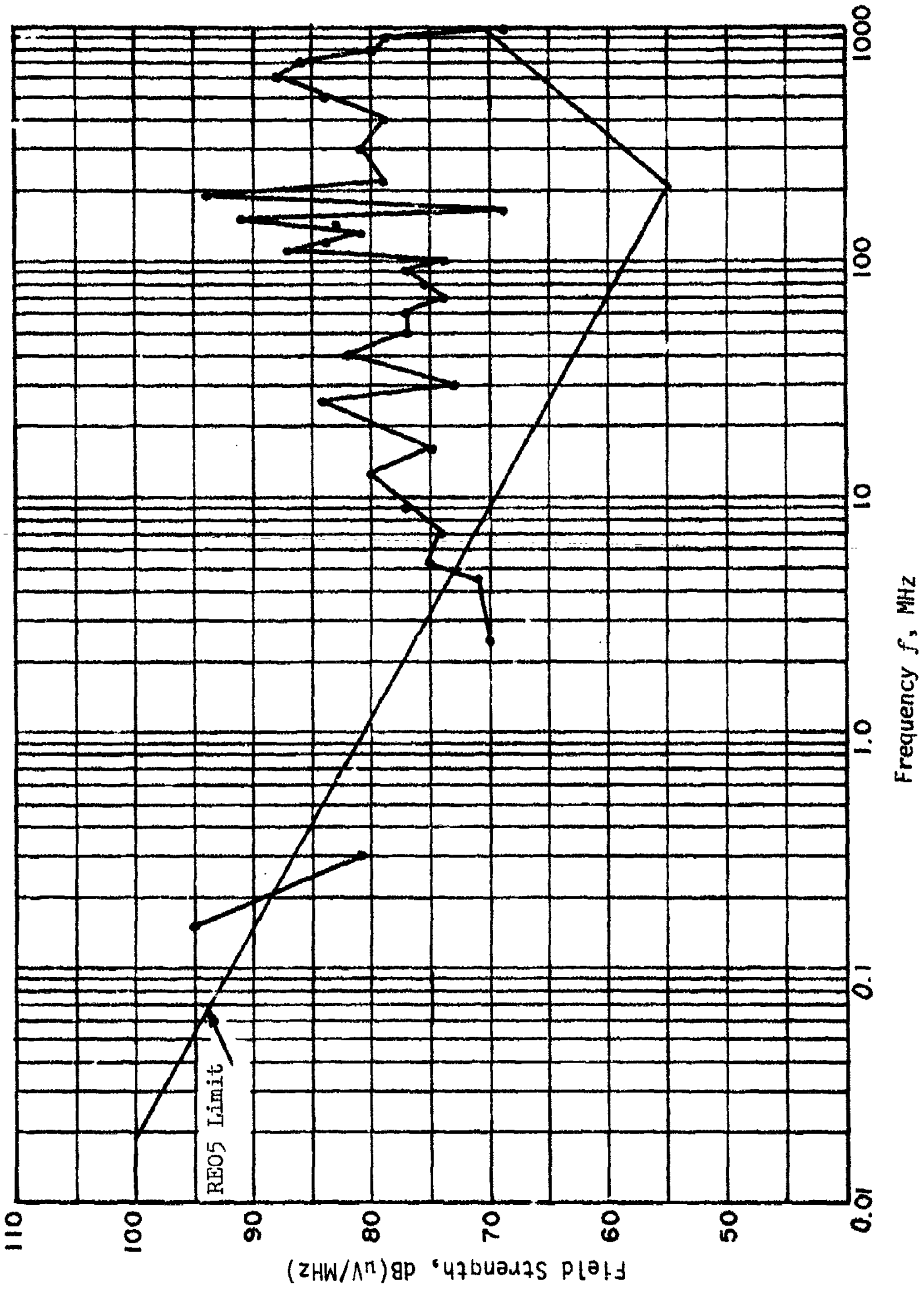


Figure 5-72. Emission from a Commercial Truck Having a Breaker-point Ignition System

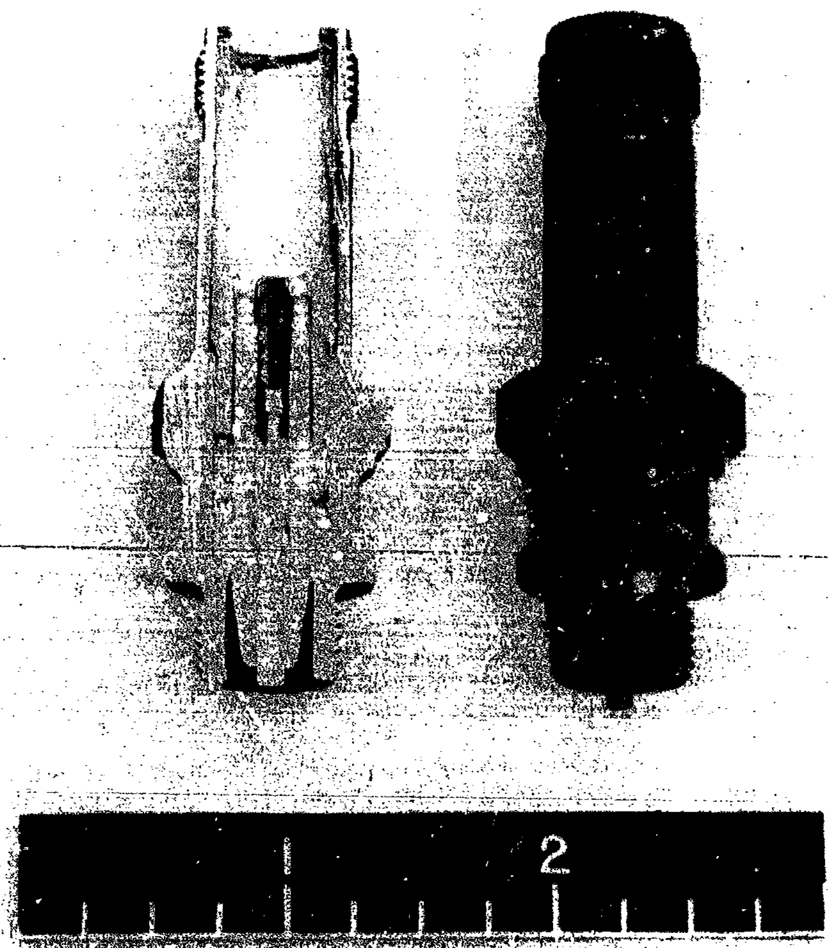


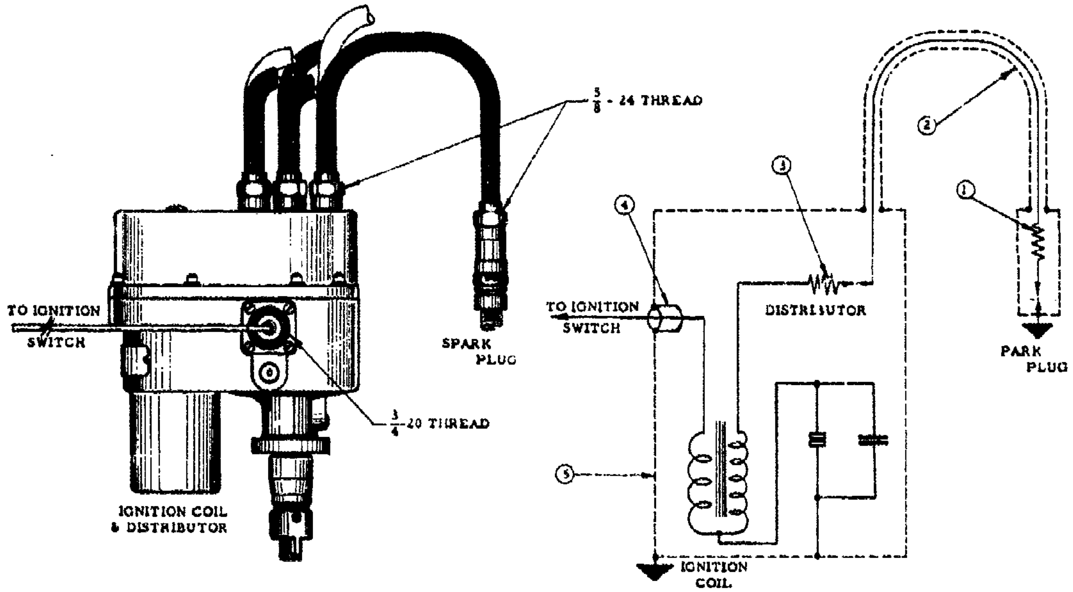
Figure 5-73. Integrally Shielded and Suppressed Spark Plug

5-6.3.2.2 Shielding

For military vehicles which are used in critical operational areas, suppression must be to MIL-STD-461. To obtain this degree of suppression requires shielding of the complete system including components and exposed wiring from the ignition coil forward. The shielding should be sufficient to contain the broad spectrum of interference, and generally will provide a waterproofed system.

An "ignitor" consisting of an integrally suppressed, self-shielded distributor and spark generator may be used. It has the advantage of minimizing the possibility of interference leakage since it is shielded as a unit and requires little maintenance to retain its suppression characteristics. Such an installation is shown in Fig. 5-74. It should be mounted on, and carefully bonded to, the engine block. Bonding practices are discussed in Chapter 4.

Flexible shielding hose gives high percent coverage and is effective in preventing the radiation of interference from ignition wiring. The shielding is made of strip metal formed either into spiral bellows or into



1. 10,000-ohm resistor built into a shielded spark plug whose fitting makes a good electrical contact with the high-tension cable shield end-fittings.
2. Braided shielding covering the high-tension cable. The end-fittings make good electrical contact with the spark plug shield and the distributor cap shield.
3. 10,000-ohm resistor built into the distributor rotor as close as possible to the electrode.
4. A feed-through capacitor in the ignition coil lead to the ignition switch mounted in the wall of the ignition unit housing.
5. The ignition coil mounts in a well in the distributor housing which is grounded to the engine. A metal cover makes good electrical contact with the housing. The cover has fittings that mate with the high-tension cable shield end-fittings.

Figure 5-74. Integrally Suppressed and Shielded Coil and Distributor (Ignitor)

some other kind of spiral that allows interlocking of adjacent strips. Either it may be soldered at the seams or allowed to provide sliding action between turns. For more effective shielding, it may be covered with one or more layers of woven metal braid. The hose must be used in waterproof applications or where severe abrasion can occur, e.g., exposed wiring on a bull-dozer.

Flexible shielding hose has disadvantages which necessitate holding its use to a minimum. Two of these are:

a. It is expensive.

b. It requires the use of considerable quantities of strategic copper, brass, and/or bronze.

The addition of resistor-suppressors in the secondary circuit, which reduces the steep transients, makes it possible to substitute tinned-copper braid shielding for flexible conduit shielding. Fig. 5-75 shows a waterproof assembly using conduit for the shield. Details on this structure are given in Military Standards MS 51010 and 51011. In addition, flexible metal hoses of both braided and solid metal construction are available which meet military requirements.

Tests performed to determine the effectiveness of tinned-copper braid shielding show that braided shield ignition cables provide adequate shielding when installed in conjunction with integrally shielded and suppressed spark plugs and ignition units.

Special shielding materials are available which can be applied to ignition or other wires after an engine is assembled. They usually consist of flexible material

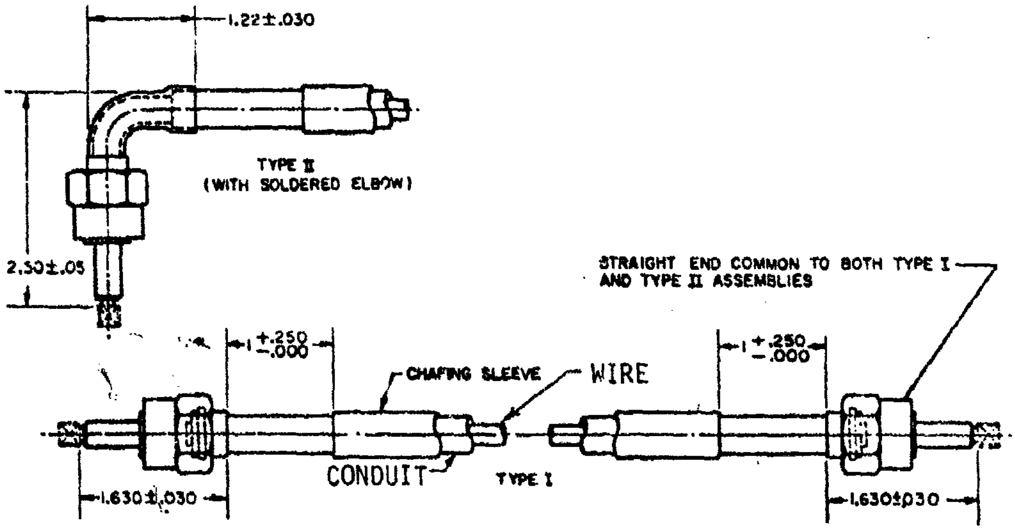


Figure 5-75. Waterproof Spark Plug Lead and Conduit Assembly

having magnetic or conducting properties which can be wrapped or laid over the wires, and the overlapping joints fastened to provide good electrical contact and in some cases a moistureproof or waterproof joint. While these materials are effective in reducing radiated interference, usually they are not approved for permanent installation.

The low-tension wiring is less important than the high-tension wiring in the radiation of interference from the ignition system; however, it still necessitates careful consideration because of the interference resulting from the make-and-break contact action of the breaker points.

The battery supply lead to the coil does not need shielding because of the capacitor installed at the coil. Since transients in the lead from the breaker points to the coil cannot be suppressed without interfering with the operation of the system, shielding of this lead usually is necessary. A single layer of tinned-copper braid shielding is adequate in most installations.

In addition to vehicles designed specifically for military use, the Department of the Army makes use of commercial type vehicles which it modifies to meet appropriate specifications. As purchased, these vehicles are suppressed to the extent required by SAE STD J551, the applicable industrial specification. This was the case with the two vehicles described in par. 5-6.3.2. Fig. 5-76 (Ref. 11) presents results of a different test on the truck with the electronic ignition system from that shown on Fig. 5-71, showing that it passed SAE STD J551A. The military requirements of MIL-STD-461 were met on this vehicle by installing a shielded ignition system, substituting a 24-V alternator which itself met MIL-STD-461, and by redesigning the windshield wiper and heater motor circuits to bring them into compliance with MIL-STD-461. To make it meet MIL-STD-461, the truck with the standard breaker-point ignition system (data shown on Fig. 5-72) was modified in a similar way.

Magneto ignition suppression is accomplished in the same manner as in battery ignition systems. Shielded and suppressed spark plugs should be used. The spark plug cables should be shielded and the shields made continuous with those of the spark plugs and the magneto shield. A suppressor resitor should be installed between the secondary winding of the magneto and the distributor, and the stop switch lead should penetrate the magneto shield via a feedthrough capacitor.

Fig. 5-77 shows in summary form a typical suppression system for a tactical vehicle. It should be noted that the lead length from the radio junction box to the battery should be as short as possible.

5-6.4 MISCELLANEOUS ENGINE-DRIVEN EQUIPMENTS (Ref. 14)

Engine-driven equipments include a variety of installations such as cranes, concrete mixers, road graders, and road rollers. Since these equipments are in the support category, they usually do not meet the

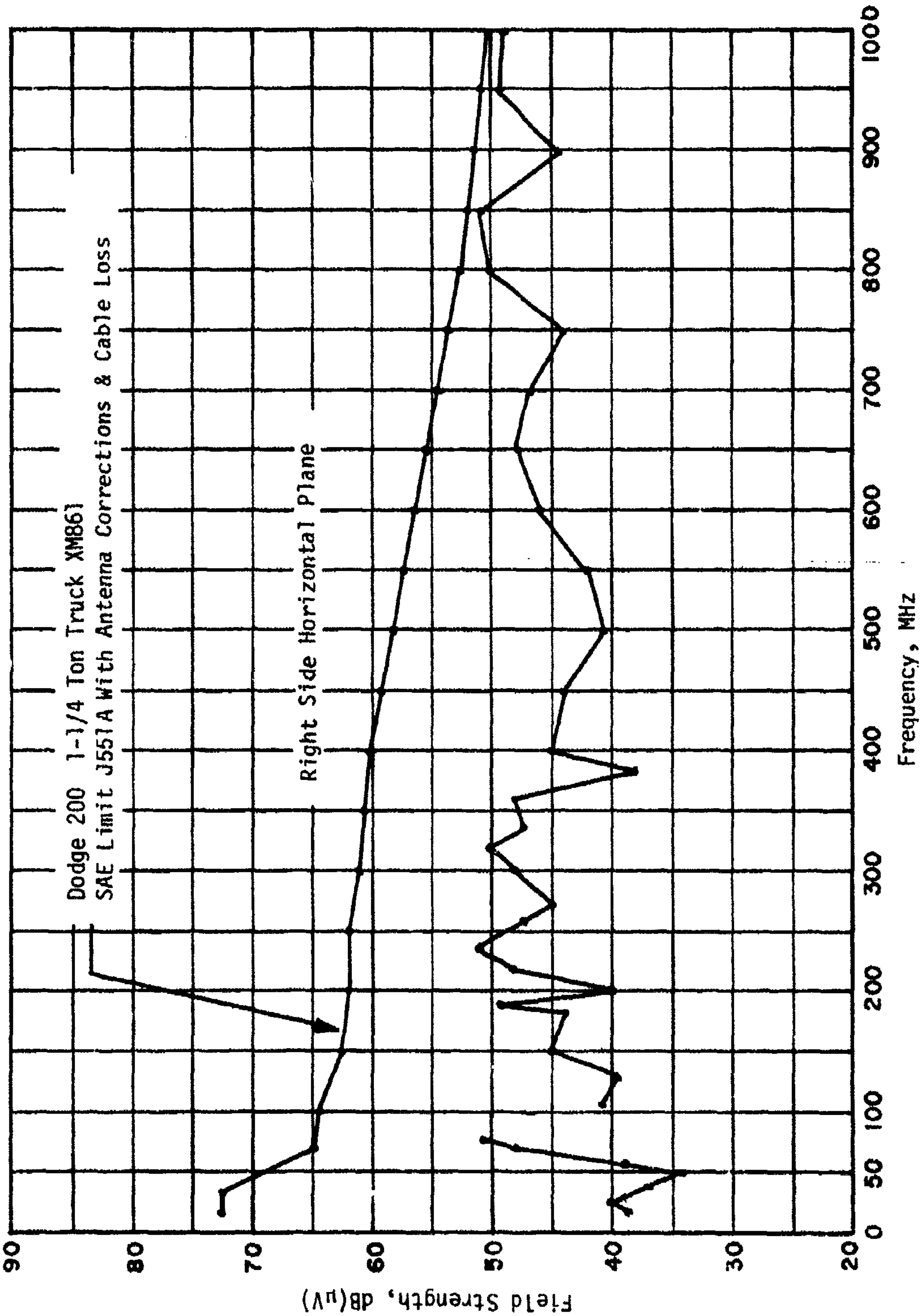


Figure 5-76. Emission from a Commercial Truck Having Semiconductor Ignition

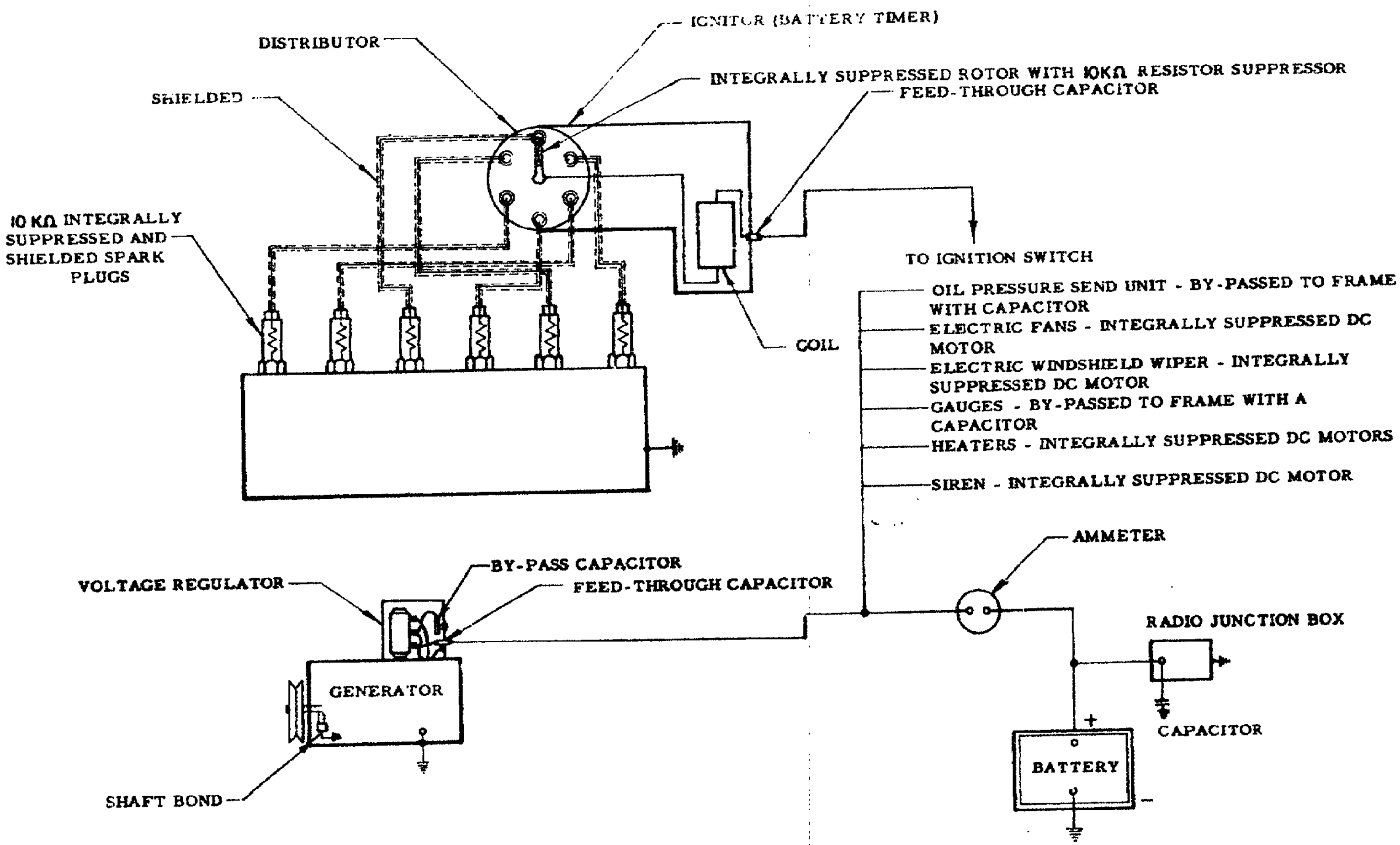


Figure 5-77. Schematic Diagram of a Typical Suppression System for Tactical Vehicles

requirements for tactical vehicles and are not waterproofed. As an example, typical suppression on a 20-ton crane consisting of a truck driven by a 6-cylinder gasoline engine is described. Another example is a material handling crane operated by electric motors and mounted on a truck bed.

5-6.4.1 Truck Suppression

The system employed for suppression on the truck engine follows:

a. Each spark plug is integrally shielded and suppressed.

b. The distributor and ignition coil are enclosed in an ignition coil shield assembly which is a part of an ignitor.

c. Each high-tension lead is shielded with tinned-copper braid terminating at the spark plugs and distributor assembly coil shield with appropriate threaded fittings.

d. A 10,000-ohm resistor-suppressor is inserted in the high-tension coil-distributor lead in the distributor shield.

e. The low-tension lead from the ignition coil within the distributor is shielded by the ignition housing, terminating at each end with appropriate threaded fittings.

f. The low-tension lead from the ignition switch and the distributor is shielded with tinned-copper braid terminating at the coil shield assembly and at the instrument panel by means of clamp connections. Cadmium-plated tooth-type lockwashers are used for bonding each clamp connection.

g. For systems in which the regulator is in a unit separate from the alternator, the armature and field leads are individually enclosed in tinned-copper braid shielding terminating in a soldered receptacle and/or connector bonded at the regulator receptacle housing and generator by means of tooth-type lockwashers.

h. The battery lead from the voltage regulator to the ammeter is shielded with tinned-copper braid terminating at each end in clamp connections and grounded at the voltage regulator mounting bracket and at the instrument panel by means of plated tooth-type lockwashers.

i. Bypass capacitors (0.1 μF, 100 V dc, typical) mounted with tooth-type lockwashers are used between:

(1) The primary terminal of the ignition coil and the ignition coil shield

(2) The armature terminal of the battery charging generator and the generator housing

(3) The armature terminal of the regulator to the regulator bracket

(4) The battery terminal of the regulator to the regulator bracket.

j. An approved braid bond strap and tooth-type washer, or tooth-type washers alone, or conductive gasket material are used between:

(1) Ignition assembly and engine

(2) Generator end plates and generator mounting brackets

(3) Generator mounting bracket and engine

(4) Generator end plate and generator adjusting bracket

(5) Generator adjusting bracket and engine

(6) Engine and front cross member

(7) Hood top section and solid portion of the hood.

5-6.4.2 Material Handling Crane

Although electrically operated from the vehicle battery, this device illustrates techniques that are effective in interference reduction (Ref. 15), namely:

a. Contact surfaces on the motor, reel, and relay housing were sandblasted and cadmium plated.

b. Relay mounting brackets were sanded to remove rust, and tooth-lock washers were used on all mounting bolts and contact terminals.

c. A shielding housing was installed around the control lines filter and mounted over the point of entrance of the control lines connector jack. A similar arrangement was provided for the main power filter.

d. The boom limit-switch was rewired to the output of the control lines filter.

e. The main power cable and the control cable were wrapped with a metallic tape which was bonded to the housing, using an electrical type fitting.

f. Anodized aluminum boxes were made to shield the brake solenoids, and the solenoid ground lead was terminated inside the box. The dc power lead to each solenoid was shielded between the main housing and the shielding box. The shield was grounded at each end.

g. Correspondingly, the load limit switch cable was shielded.

h. A 4.0-μF, 50-V dc capacitor was installed across each of the relay coils.

5.7 RECEIVERS

5-7.1 INTRODUCTION

Receivers act as both interference emitters and as susceptors.

5-7.2 EMISSION

The likely frequencies of emission are the frequencies of internally generated signals such as those associated with generation of the local oscillator signal, synchronizing signals in pulse and digital systems, the intermediate frequency (since the desired signal will be amplified to a fairly high level in the IF amplifier), the audio or video frequencies of the output signal, and power line harmonic frequencies. The local oscillator waveform is generally of substantial amplitude, and because it is injected at a point in the receiver not far from the antenna terminals, it will likely couple to the antenna and be radiated.

At the antenna terminals MIL-STD-461A allows 34 dB(μV) for narrowband emission and 40 dB(μV/MHz) for broadband emissions, from 10 kHz to 12.4 GHz. On tests on one radar receiver (Ref. 11), only one emission frequency was found, the local oscillator at 1380 MHz, at a level of 33 dB(μV). Modern receivers using frequency synthesizers are potential sources of emission at the many frequencies entering into the synthesizing process, both above and below the oscillator output frequency.

5-7.3 SUSCEPTIBILITY

5-7.3.1 Admission Ports

Unwanted signals may enter a receiver by conduction through power and control wiring and by radiation or induction to the case, the external cabling, or the antenna.

5-7.3.2 Nonantenna Inputs

Limits specified in MIL-STD-461A for power line susceptibility range between 1 V and 3 V from 30 Hz to 400 MHz. Measured data on a number of receivers (Ref. 16) are shown in Fig. 5-78. The levels are seen to be lower than the limit for most of the types tested by significant amounts which illustrates that care in receiver design is necessary to meet the limit.

Limit levels on conducted susceptibility of control circuit terminals of receivers are not specified in MIL-STD-461A except for the test of squelch circuits (Test CS07).

Receiver susceptibility to fields impinging on case and cables can be significant because of inadequate shielding. As discussed in Chapter 6, ventilation holes, lids, shaft passages, and connectors must not impair the integrity of the shield.

5-7.3.3 Antenna Inputs

Mechanisms of receiver susceptibility to unwanted signals are discussed in detail in par. 3-2. Fig. 5-79 shows the essential elements of a typical receiver connected to an antenna via an external unit identified as a coupling device. The RF bandpass preselector is mainly intended to provide some suppression against the mass of unwanted signals far from the receiver tuned frequency to which the active circuits of the RF amplifier otherwise would be exposed. Most of the receiver selectivity is obtained in the IF amplifier. Strong unwanted signals that are not sufficiently attenuated by the preselector will produce interference. Intermodulation, cross modulation, and desensitization occur because of nonlinearity in the stages preceding the mixer. Spurious and image responses arise in the mixer.

Nonlinear effects also may occur in the coupling network preceding the receiver itself. Solid-state devices used in filters, circulators, and switches — e.g., ferrites and Yttrium-Iron-Garnet (YIG) devices — behave nonlinearly for large inputs.

Desensitization caused by high amplitude pulsed signals was described in par. 3-2.2.2.3. Radar emissions, because of their intensity, can penetrate lower frequency receivers and appear at the input of active devices with sufficient magnitude to overload the input of the active device, decreasing the gain of the input tuned circuit for the duration of the pulse. The desired signal is thereby amplitude modulated at the pulse rate, causing a cross modulation from the undesired pulse carrier to the desired signal.

An associated mechanism is that of detection in the first active device. The sidebands of the detected signal may contain energy in the receiver passband. Broadband signals of this kind may have bandwidths to about 15 MHz or even more, depending upon the parameters of the interfering radar. By taking into account possible sideband distortion, this mechanism should be considered potentially significant through the HF band.

The mechanisms described in the two preceding paragraphs are significant in receivers operating well below the microwave frequency of the radar because of the inability of the input circuits to filter signals adequately very far from their design center frequencies. Ordinary HF tuned circuits can pass microwave frequencies through stray coupling paths. If a receiver is expected to operate in such an environment, low-pass filters suitable for rejecting microwave signals may be required at the receiver input.

The suppression mechanism known as the capture effect, in frequency modulation receivers is well documented (Ref. 17). In such a receiver it is possible to reject channels operating near the desired signal even though the amplitude of the undesired signal is a large fraction of the amplitude of the desired one. It is

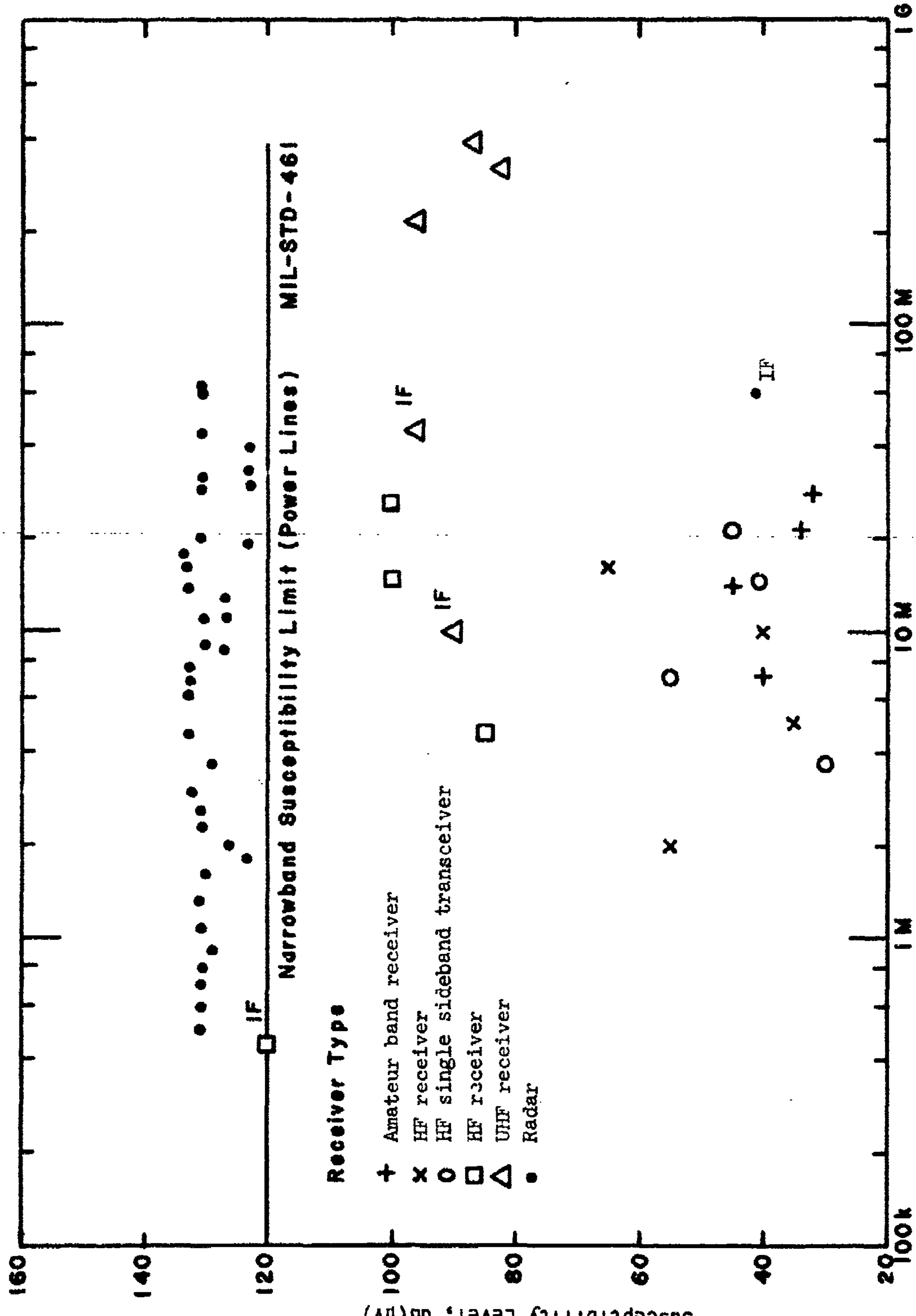


Figure 5-78. Receiver Power Line Susceptibility

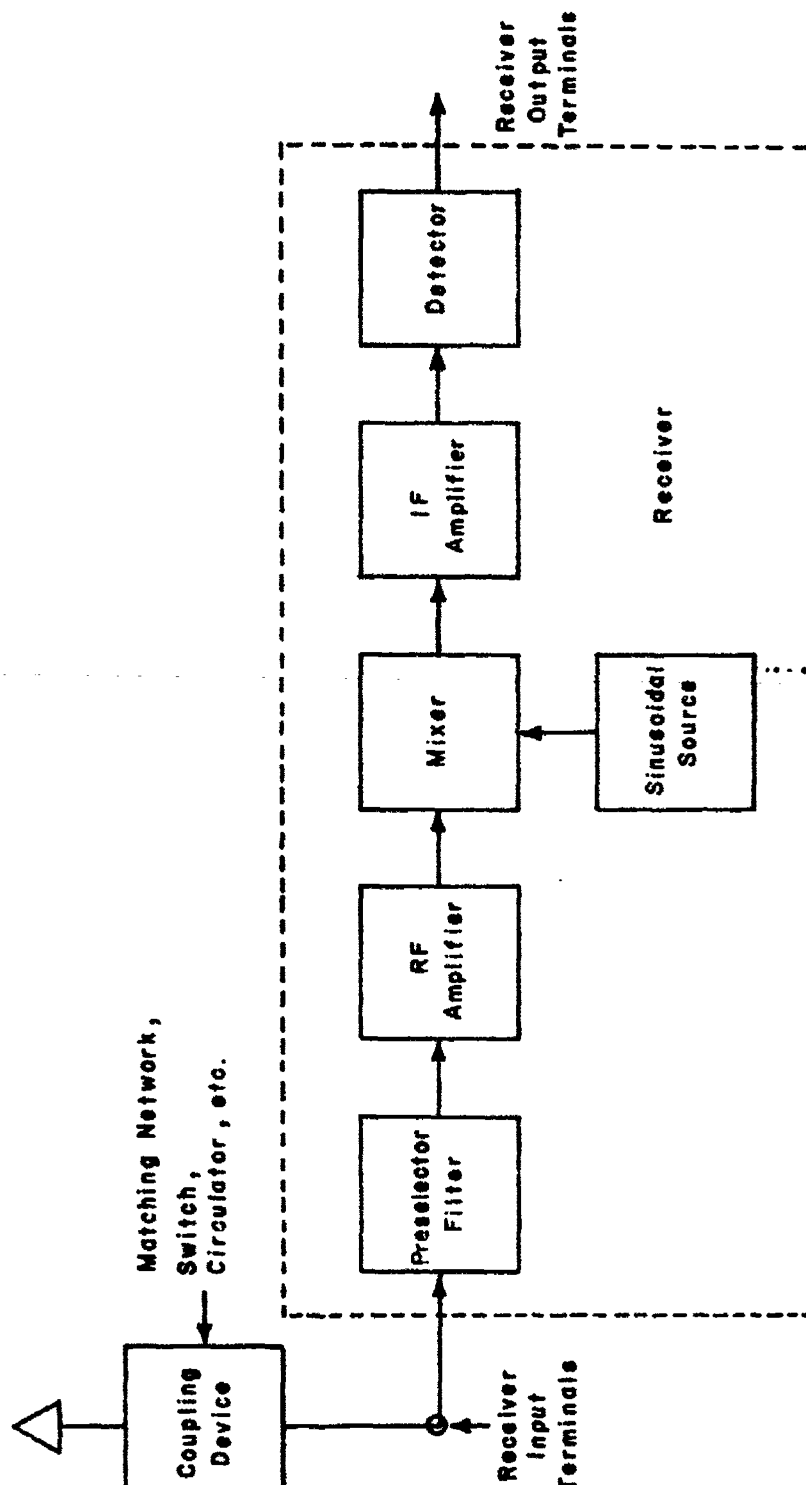


Figure 5-79. Receiver Block Diagram

necessary to have large bandwidth in the discriminator to achieve the rejection. An estimate of the required bandwidth B is

$$B = \frac{2a\delta}{1 - a^2}, \text{ Hz} \qquad (5\text{-}10)$$

where

a = $\dfrac{\text{amplitude of undesired signal}}{\text{amplitude of desired signal}}$

δ = magnitude of difference frequency between desired and undesired signals, Hz

If the interference-to-signal ratio, $a = 0.9$, the bandwidth $B = 9\delta$. If δ is the adjacent channel separation frequency, typically 200 kHz in broadcast FM, a discriminator bandwidth of close to 2 MHz is indicated.

5-7.4 INTERFERENCE REDUCTION TECHNIQUES

The degree to which the receiver is capable of withstanding undesired signals entering by these mechanisms can be improved by proper design, usually at some expense in a desired property and at a premium cost. While not much can be done to eliminate the effects of signals which spectrally overlap the desired signal (though the effect can, in principle, be minimized by the use of an optimum detector), elimination of nonoverlapping signals is a matter of adequate filtering and improving the degree of linearity. The superheterodyne, in fact, is intended to make it easy to provide an adequate and approximately constant selectivity characteristic by the use of a fixed tuned IF amplifier. Alternatively, one may look to active devices with minimal nonlinearity. As a rule, one can expect to find an operating point on an active device where the nonlinear effect is minimal. Also, as is discussed in par. 3-2.2, mixers with oscillator input limited to the mixer square-law region would have no spurious responses other than the image response. However, any measure which reduces interference susceptibility usually will lower conversion efficiency to the desired signal and a trade-off will have to be made between interference rejection capability and receiver sensitivity. One may look for optimal operating conditions with whatever active element is being used, and to linearization schemes which seek to cancel the important nonlinear effects. Techniques in the latter category have been demonstrated in laboratory studies but there appears to be no practical implementation of such schemes (see par. 4-1.7).

Typically, in communication receivers the RF amplifier bandwidth varies with tuning, and the local oscillator output to the mixer is also dependent on tuning. The spurious response levels as well as their frequencies therefore vary with tuning. Also it has been found that substantial statistical variation occurs in these responses (Ref. 18), the differences arising among different receivers of the same design (Ref. 19). The variability is especially severe for responses involving high degree nonlinearity and undesired inputs passing through the RF stages far from the tuned frequency. Then fixed estimates of spurious response levels are not dependable and a statistical measure must be used. Typically, mean levels of spurious response might be 80 dB with a standard deviation of 10 dB.

The intermodulation and cross modulation limit levels are specified at 66 dB above a standard reference input for each undesired signal input. The spurious response limit level is specified at 80 dB above a standard reference input for frequencies in the receiver tuning range, but outside the 80-dB bandwidth points on each side of the tuned frequency, and 1 mV outside the receiver tuning range. Tests of a high quality radio receiver tuned to 3.0 MHz are shown in Fig. 5-80 (Ref. 16). The numbers identifying the measured points on the plot represent the harmonic orders, the first digit representing the oscillator harmonic order, the second digit representing the input signal harmonic order, and the sign indicating whether the response arises as a sum or difference in the mixer. Responses are seen to be in the range of about −83 to −120 dB relative to the input reference level.

Tests of intermodulation and cross modulation on the same receiver show that with the receiver tuned to 3.001 MHz, intermodulation products were found in the vicinity of the tuned frequency as follows for 3rd order nonlinearity:

Frequency Difference, kHz	Power of Each Signal, dBm
13	−58
25	−50
49	−37

The standard reference signal was −109 dBm so that the inputs are 51 dB, 59 dB, and 72 dB, respectively, above the reference signal. The frequency difference is the difference between the frequencies of the two inputs simultaneously applied with the power shown.

Earlier in this paragraph mechanisms were described which involve inputs far out of the received band. Such inputs sometimes penetrate because the input circuits are poorly selective far from their design center frequencies. To illustrate this point,

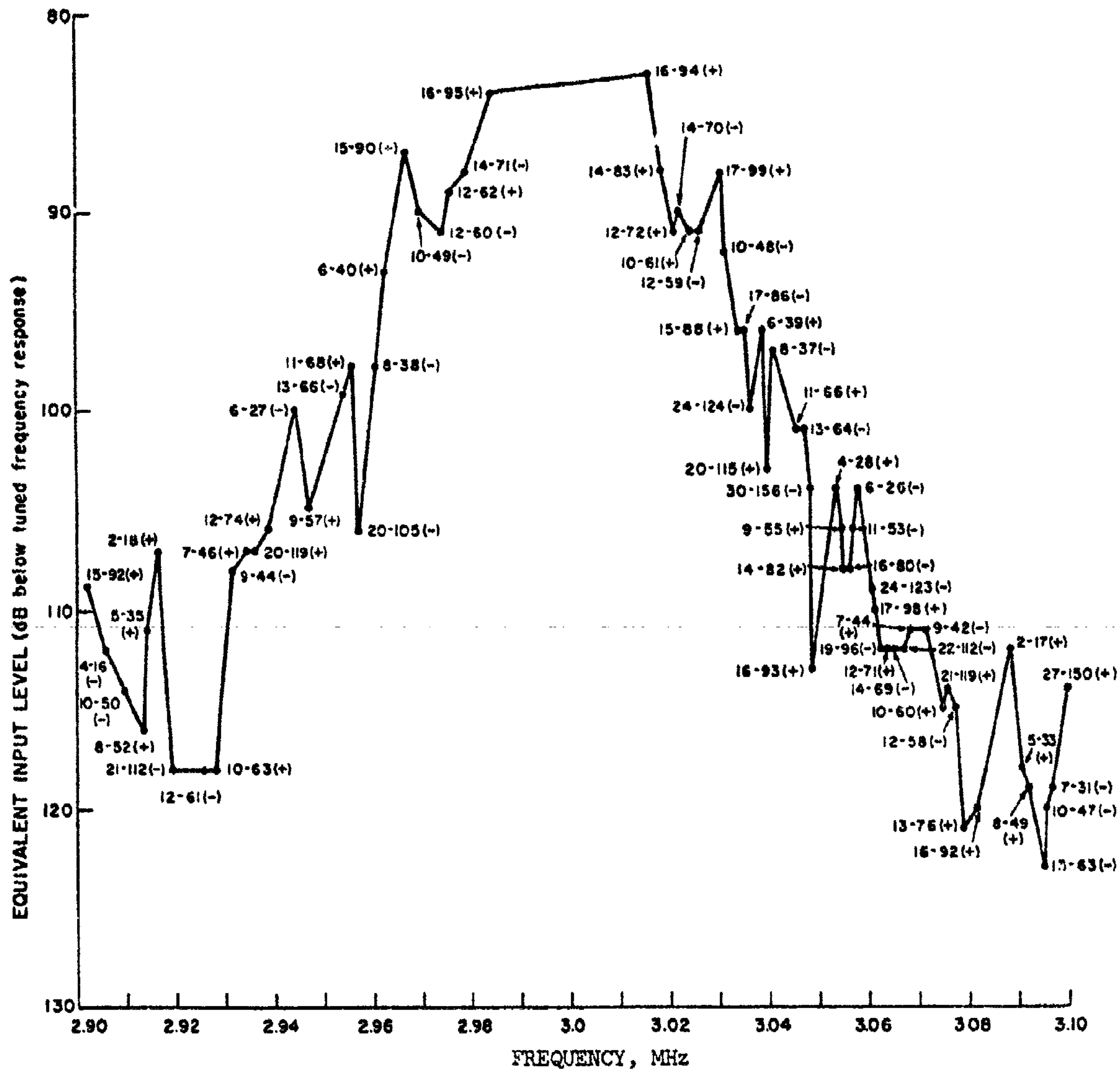


Figure 5-80. Spurious Responses of Receiver Tuned to 3.0 MHz (Ref. 16)

Fig. 5-81 shows the relative response of the RF circuits of one receiver tuned to 70 MHz. At 435 MHz the attenuation relative to that at 70 MHz is only about 20 dB because of stray parameters in the tuned circuits. Phenomena of this kind pose a major problem in interference control. The designer usually will focus on the primary properties of a circuit and be satisfied when he achieves it. Secondary effects, particularly when far outside the band, generally are overlooked, and virtually every circuit will behave unexpectedly far outside its design range.

5-7.5 MILLIMETER WAVE RECTIFIERS

Receivers for use in the millimeter wave range (about 10 to 100 GHz) differ in EMC characteristics from receivers for use in the microwave range only because of internal noise, linearity, and directivity properties. The active electronic devices used are scaled down klystrons and travelling wave tubes, and solid-state devices such as IMPATT diodes, Schottky barrier diodes, and tunnel diodes.

A receiver for use in K_a-band (≈35 GHz) makes use of an IMPATT diode oscillator and a balanced

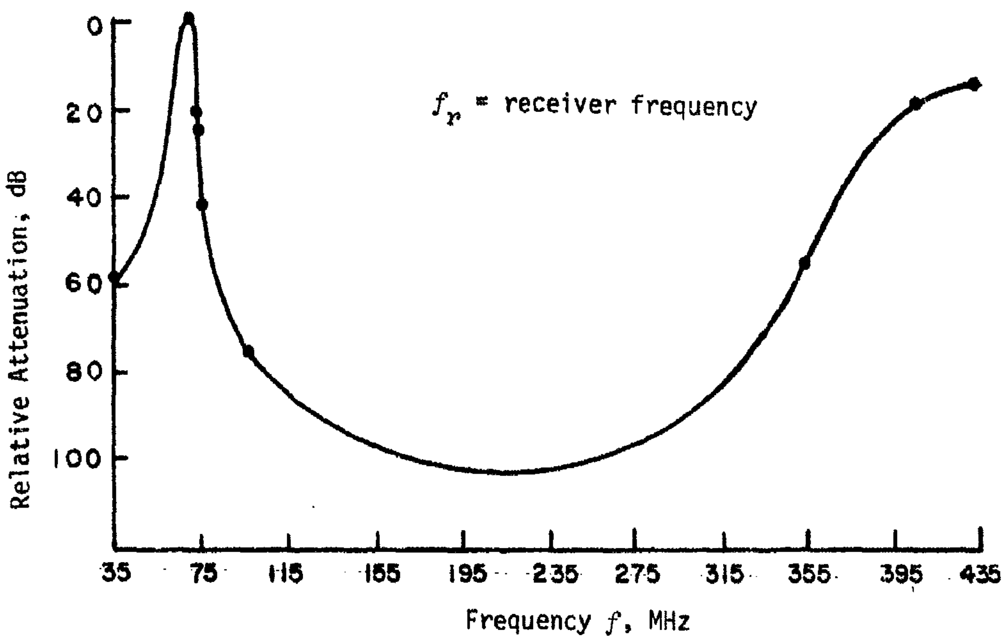


Figure 5-81. RF Amplifier Characteristic, f_r = 70 MHz

mixer at the front end to achieve a noise figure of 10 dB (Ref. 21). Without special frequency stabilization and with a single-ended mixer, a noise figure of about 40 dB would have been obtained. Frequency stabilization is used in a repeater for a millimeter wave transmission system which achieves noise figures of better than 9 dB (Ref. 22). Spurious response sensitivity of a K_a-band balanced mixer using IMPATT diodes are reported to be −40 dB relative to the desired input (Ref. 23), a value said to be typical of balanced mixers. The relative intermodulation behavior of TWT amplifiers designed for microwave operation and millimeter wave operation is shown on Fig. 5-82. For operation near saturation, the 3rd order intermodulation products of the millimeter wave and the microwave amplifiers are not substantially different. Below saturation the millimeter wave device is better, attributed largely to the difference in tuning systems in the two cases.

It is also found that conducted susceptibility in millimeter wave systems occurs at higher levels than at microwave frequencies because of the decreased cable coupling and, because of the narrower beamwidths easily achievable at millimeter wave frequencies, it is easier to colocate systems without fear of interaction.

5-7.6 INTERFERENCE EFFECT IN DIGITAL RECEIVERS

In par. 3-2.1 the effects of the admitted signals in terms of signal masking and error induction were discussed from the viewpoint of theory and principles. Data on the effects of interference on digital receivers were obtained by simulation (Ref. 24). The simulated receiver was that used in the AN/GRC-103 for pulse code modulation-frequency modulation (PCM-FM). The local oscillator was not modeled to conform to the actual receiver oscillator but the mixer nonlinearity was modeled to generate harmonics as does the actual mixer. The timing recovery was modeled so that the actual timing jitter is approximated in the simulated system. Figs. 5-83 and 5-84 show results obtained with various kinds of interference for PCM-FM and 4 phase, differential phase shift keyed (4ϕ DPSK) systems, respectively. The interference is assumed added to the signal at the antenna terminals. The quantity of Δf represents the ratio of the frequency separation between the two signals (desired and undesired) to the data rate. For both cases illustrated it is to be noted that the theoretical error rates for Gaussian noise obtained by analysis are substantially lower than those obtained in the

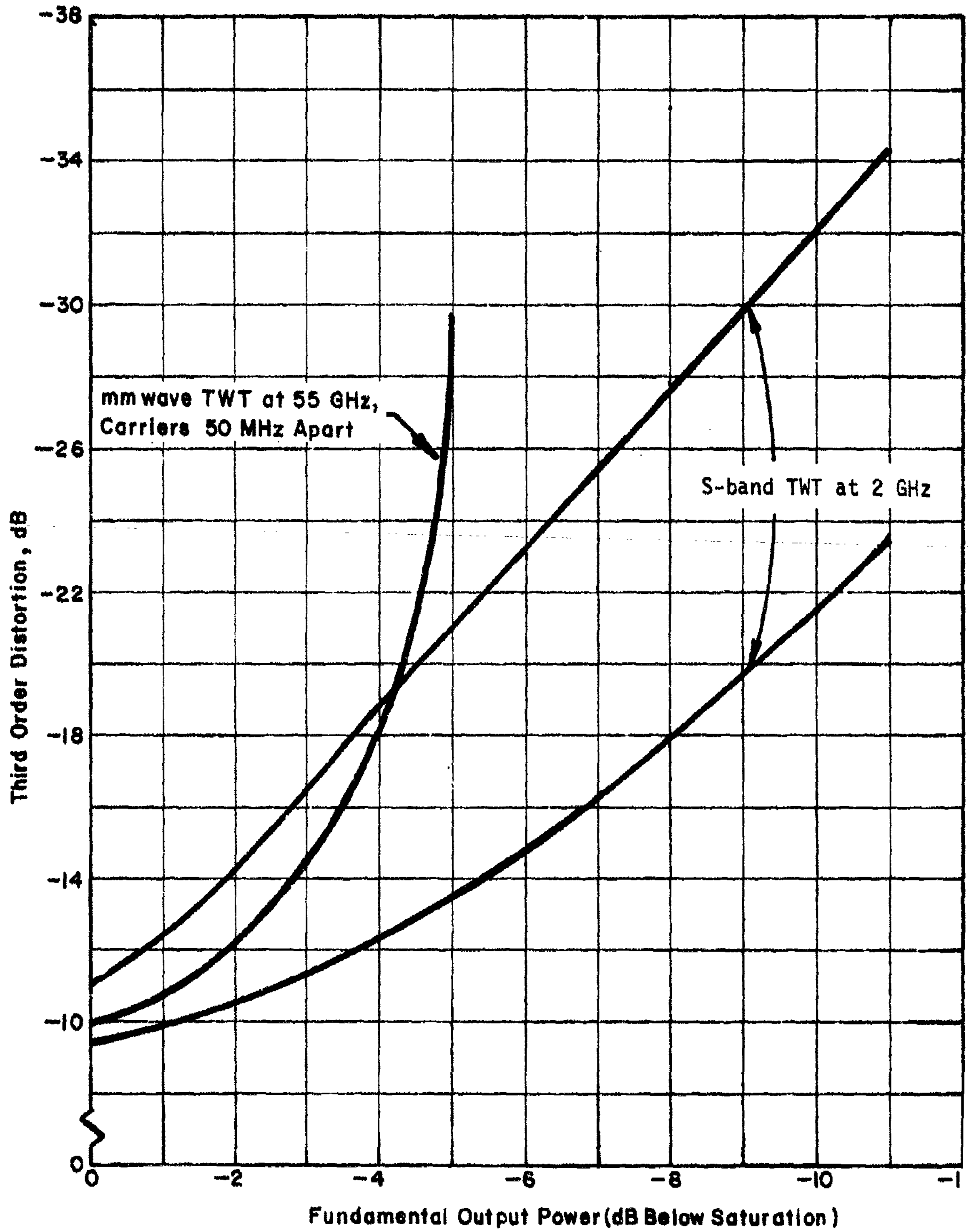


Figure 5-82. Comparison of Third Order Distortion of a Microwave Amplifier vs a High Power Millimeter Amplifier

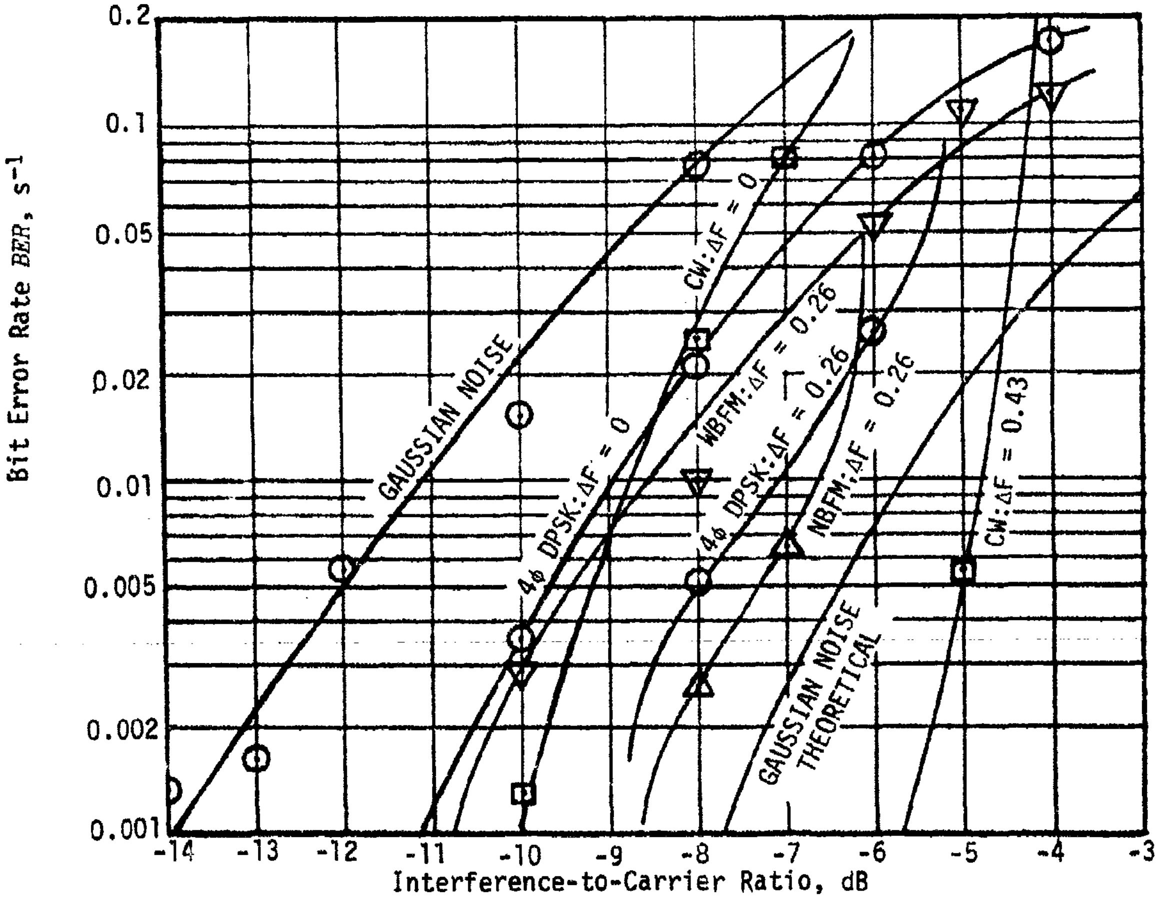


Figure 5-83. Susceptibility of 4ϕ DPSK to Antenna-coupled Interference

simulation; in the case of PCM-FM it is worse by about 3 dB and for 4ϕ DPSK it is worse by 5 to 6 dB. Among reasons given for this is that the theoretical case assumes an ideal detector, ideal bit timing recovery, and ignores effects of pre-detection and post-detection filters. In short, the results suggest that actual receivers fall substantially below the theoretical potential.

The results using various other kinds of interference generally are found to degrade the error probability less than does Gaussian noise of the same level. The interference types presented are Continuous Wave (CW), Narrow-band FM (NBFM), Wide-band FM (WBFM), as well as interference similar to the desired signal.

5-8 TRANSMITTERS

5-8.1 INTRODUCTION

In par. 5-8 emphasis is on those considerations directly related to the generation of radio frequency energy having the proper modulation and power level necessary to transmit the desired signal. Associated with the basic transmitter are circuits and devices capable of generating interference as discussed in other portions of this handbook. Examples are power supplies, motors for driving ventilating fans, power switches, and relays. These can be controlled by the usual techniques of filtering, shielding, and bonding. In applying these techniques, special care may be

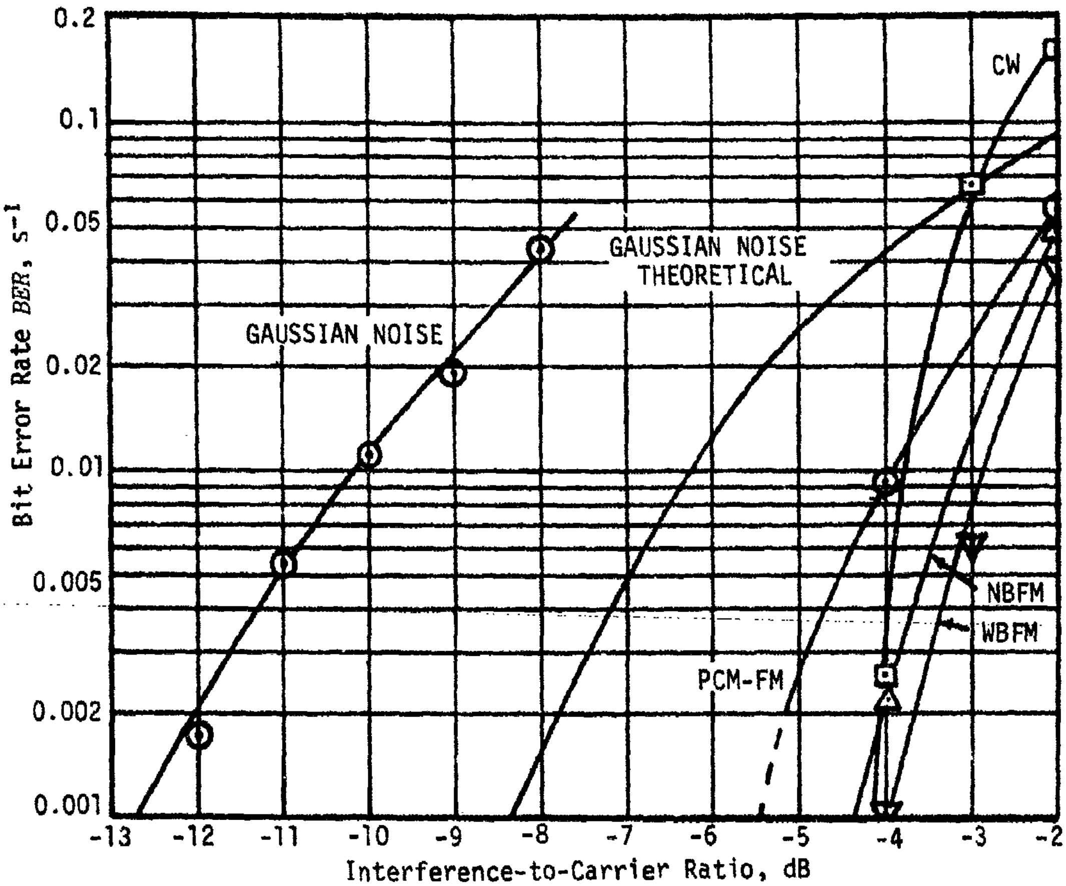


Figure 5-84. Probability of Bit Errors for PCM-FM With Continuous Antenna-coupled Cochannel Interference (Δf = 0.24)

necessary because of the numerous frequencies present in various circuits especially where frequency multiplication occurs. Individual circuits should be isolated as much as possible with shields and filters to avoid stray coupling. In particular, power supply output lines should be decoupled where several stages are fed from the same supply, and separate circuits should be isolated by compartmentalization.

Filters may be required on all power, control, and monitoring leads where they penetrate the case or cabinet. To protect against radio frequency leakage, usually these filters may be of relatively simple low pass types. The components should be designed to work at the frequencies present. Where it is necessary to prevent broad band interference such as generated by switching transients, rectifiers, or commutator motors producing energy in the low kilohertz region of the spectrum, the filters may have to be more complex and have separate sections for low and high frequencies, respectively. In addition, apertures in cabinets for ventilation should be covered by screening, access door leakage prevented by metal gaskets, and indicating meters mounted in leakproof arrangements.

5-8.2 POWER AND BANDWIDTH LIMITING

From a compatibility viewpoint, the design of the transmitter should be guided by the general principle that transmitted power and bandwidth should equal, but not greatly exceed, that necessary to give the required receiver output quality. Well-designed systems such as radio relay systems take careful account of noise, interference, and fading and are built with bandwidths and power levels to give a specified received signal reliability and quality. Over-design may insure good operation for one system but deny it to another. By experience, it is found that a bandwidth of about 3.5 kHz is adequate for voice transmission at a signal-to-noise power ratio of about 30 dB (Ref. 25, p. 50). Short rise-time radar pulses, when not needed to achieve some specified objective, should not be used. In digital data transmission, the use of band-limited pulses may give rise to intersymbol interference. However, by use of the class of band-limited waveforms known as Nyquist waveforms (Ref. 25, pp. 448-453), instead of rectangular or triangular waveforms, one can have band-limiting and zero intersymbol interference.

The most readily recognized Nyquist waveform is one whose Fourier spectrum $V(f)$ is given by a constant A_1 over a bandwidth interval $2B$ Hz.

$$V(f) = \left.\begin{array}{ll} A_1, & -B < f < B, \text{ V·s} \\ 0, & \text{elsewhere} \end{array}\right\} \quad (5\text{-}11)$$

The corresponding time-domain waveform $s(t)$ is

$$s(t) = 2A_1B\frac{\sin(2\pi Bt)}{2\pi Bt}, \text{ V} \quad (5\text{-}12)$$

which is zero at all points $t = k/(2B)$, $k = 0, \pm1, \pm2, \ldots$ The pulses last indefinitely but they are zero at periodic instants. The receiver measures the amplitude of each waveform by sampling at these instants. Successive signal pulses may be spaced by

$$T = \frac{1}{2B}, \text{ s} \quad (5\text{-}13)$$

as shown in Fig. 5-85 without encountering intersymbol interference.

The pulses of Eq. 5-12 are not convenient to generate. A more practical pulse is one whose Fourier spectrum falls off slowly at the edge of the band. It turns out that many waveforms have the Nyquist property of no intersymbol interference. To have it, the spectum must be such that the sum of all shifted repetitions of the spectrum, the shift being the reciprocal of the pulse spacing, is constant. This situation is illustrated in Fig. 5-86. Note that these waveforms occupy a band ranging from $1/(2T)$ for the waveform given by Eqs. 5-11 and 5-12 to $1/T$. A common choice is the raised cosine function for which

$$V(f) = \left.\begin{array}{ll} A_2[1 + \cos(\pi f T)], & \dfrac{-1}{T} < f < \dfrac{1}{T}, \text{ V·s} \\ 0, & \text{elsewhere} \end{array}\right\} \quad (5\text{-}14)$$

which occupies a band twice the minimum.

Signal pulse shaping can be accomplished using simple resistor-capacitor (R-C) filters, typically one or more isolated R-C circuits in cascade. The transfer function of such a filter is of the form

$$H(f) = \frac{1}{\left[1 + j\left(\dfrac{f}{f_c}\right)^n\right]}, \text{ dimensionless} \quad (5\text{-}15)$$

f_c = 3-dB bandwidth of each filter section
n = number of sections in cascade

In a recent study (Ref. 26) a data filter of this kind was assumed for simulating the spectral output of a digital transmitter. Shown in Figs. 5-87 and 5-88 are the digital signal spectra of a data sequence comprised of 0's and 1's in the order 00[illegible]101101101010 before and after such a filter. The plots are of the discrete Fourier transform (using the Fast Fourier

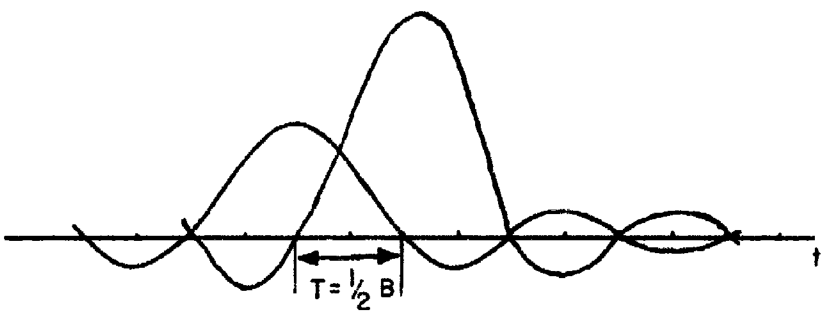


Figure 5-85. Successive Nyquist Pulses

Transform Algorithm) with 0's represented by −1, and 1's represented by +1. For Fig. 5-88 the binary data rate f_d was $2f_c$ and n was 2. Filtering in this way results in substantial reduction of sideband energy, though not as much as with the Nyquist waveform, and some intersymbol interference must be expected.

5-8.3 DESIGN CONSIDERATION

Tailoring the deliberately generated band is a first step in limiting potential interference. The next step requires controlling emissions which arise because of nonideal characteristics of real electronic devices. Two serious sources of unwanted components are the transmitter modulator and the final amplifier. In the case of the former, nonlinear distortion in the baseband amplifier will cause spectral broadening commonly called sideband splatter, and, when the baseband signal is modulated onto the RF carrier, the process — usually carried out in a nonlinear device — may give rise to further broadening of the spectum. For the sake of efficiency, the final amplifier often is operated in a nonlinear mode (typically Class C for amplitude and frequency modulation), generating harmonic currents into the final tank circuit. Though the tuned tank provides a filtering effect, some residual unwanted output is inevitable. Some types of microwave amplifiers are inherently based on impulse feeding a tuned cavity, implying strong harmonic drive to the tuned cavity and consequent harmonic output.

Less common sources of unwanted outputs may be found at intermediate points in a transmitter system. The audio or video signal, prior to modulation, may find a leakage path out of the transmitter. In systems wherein a low-frequency sinusoid is generated initially and then multiplied to the required output frequency, the intermediate frequencies may leak out of the transmitter. The possibility of parasitic oscillations also exists at various points. These are not common in competently designed, low-frequency transmitters but are common in microwave transmitters. Another mildly troublesome source is oscillator interference. The effect of such interference is similar to that of sideband splatter, but the level is generally lower.

There are times when the transmitter is not, by itself, responsible for the spurious output. The output of nearby transmitters may enter a transmitter, usually through its antenna, and mix with the desired signal to form an unwanted output component. Again, the effect depends on nonlinearity in the final amplifier, in the transmission system between the final amplifier and the antenna, or in the antenna itself. The effect produced is either cross-modulation, wherein the information sidebands of the undesired signal appear with the desired signal, or intermodulation, wherein a third signal, containing some version of the information sidebands of both signals, is formed.

The nature of these mechanisms is developed in greater detail in the paragraphs that follow.

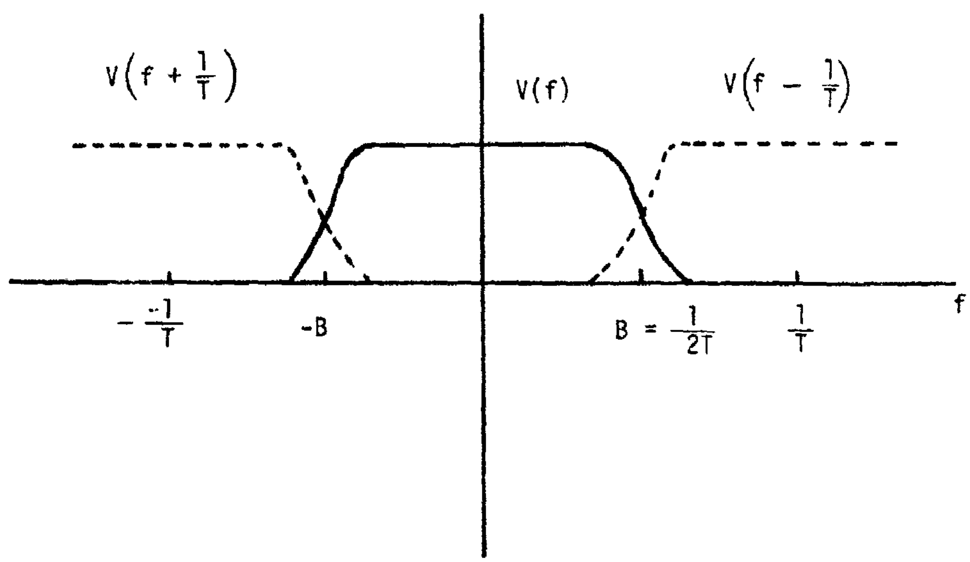


Figure 5-86. Spectrum of Nyquist Pulse Showing Constancy of Repeated Spectrum

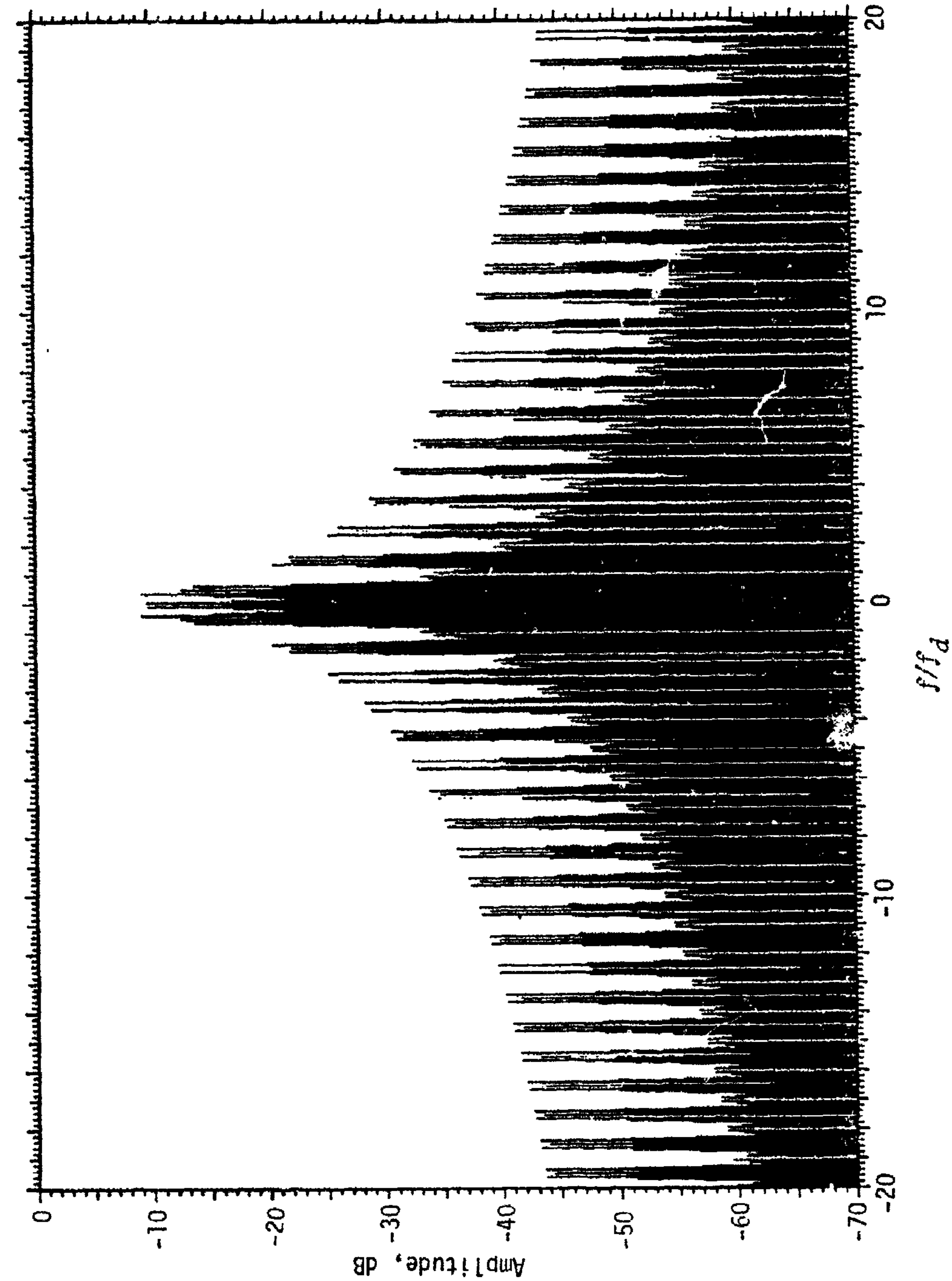


Figure 5-87. Spectrum of Unfiltered Data Sequence

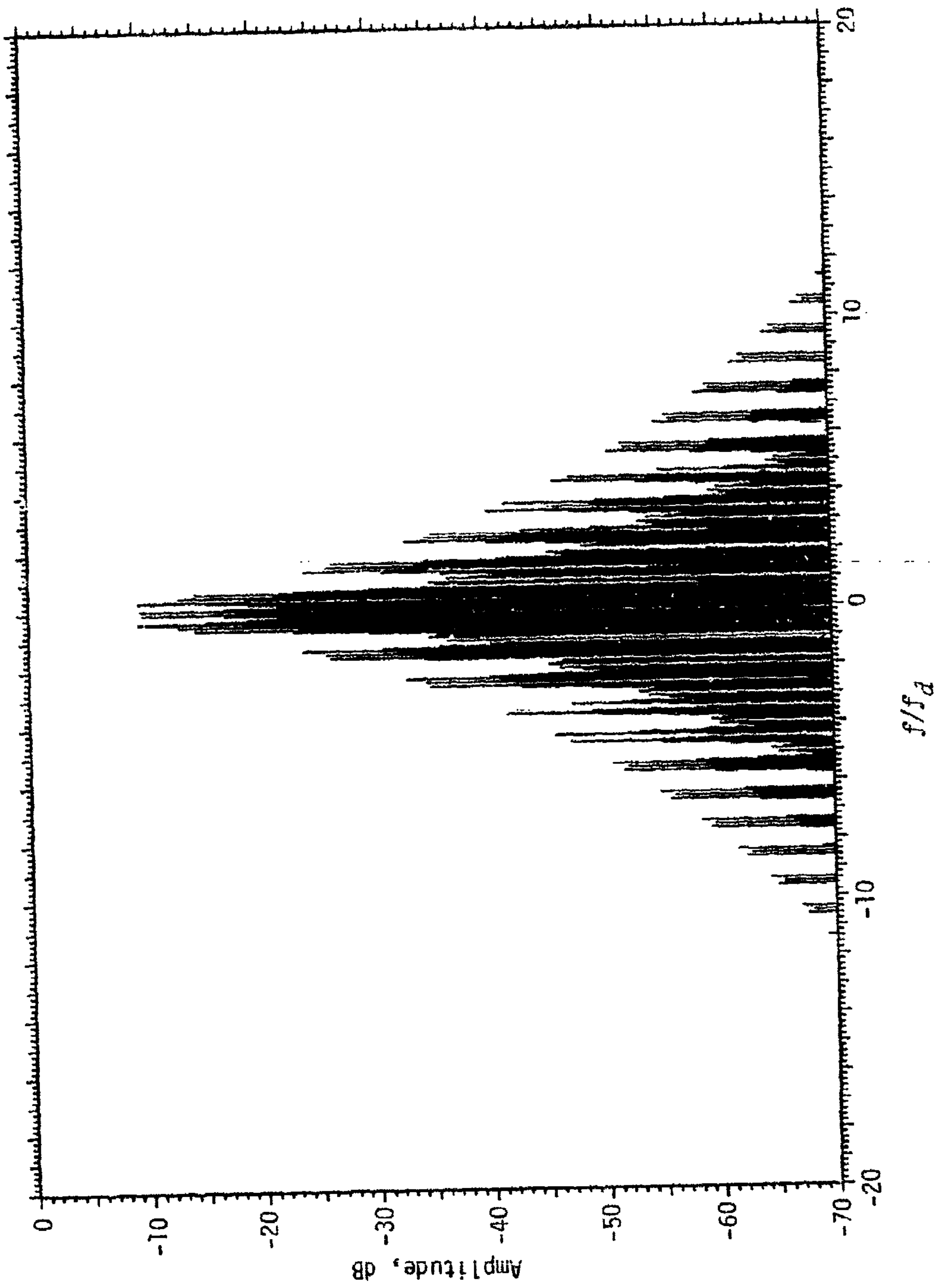


Figure 5-88. Spectrum of Filtered Data Sequence

5-8.4 SIDEBAND SPLATTER AND ITS SUPPRESSION

Sideband splatter refers to the generation of spectral components close to the carrier frequency but beyond those ideally required for the modulation process. It is of particular importance in linear modulation systems such as amplitude modulation (AM), single-sideband (SSB), and double-sideband (DSB) systems. In these systems, as for narrow-band FM, the minimum bandwidth is determined by the baseband signal and, if its spectrum is broadened, so is the spectrum of the transmitted signal. In broad-band frequency modulation (FM) systems the transmitted bandwidth is largely a function of the deviation rather than the baseband bandwidth and the effect of splatter is less significant.

5-8.4.1 Mechanism

With linearly-modulated signals, the principal mechanism of modulation is multiplication of an RF carrier by the baseband. This operation ordinarily is carried out indirectly using mixing in a nonlinear device as indicated in Fig. 5-89.

Ideal nonlinear elements are the square-law device and the linear diode. To obtain the required result without distorting the desired sidebands, the amplitude of the sinusoidal carrier applied to the linear diode must be much larger than that of the information waveform. The ideal characteristics are, however, only approximately realizable. A typical nonlinear device can be described by the output-input characteristic

$$y = \sum_{n=0}^{N} a_n x^n \qquad (5\text{-}16)$$

where

y = instantaneous output

x = instantaneous input

Input and output are either current, or voltage, or one current and the other voltage; the coefficients a_n must have the units necessary for consistency. Thus, if y is in amperes, x is in volts; then, a_n has units of ampere/(volt)^n. The degree of significant nonlinearity is assumed here not to exceed that represented by N. The output of the nonlinear device for the input $[x_1(t) + x_2(t)]$ with $x_2(t)$, as given in Fig. 5-89, is

$$\left.\begin{aligned} y &= \sum_{n=0}^{N} a_n (x_1 + x_2)^n \\ &= \sum_{n=0}^{N} a_n \sum_{k=0}^{n} \binom{n}{k} x_1^{(n-k)} \cos^k(\omega_c t + \varphi_c) \end{aligned}\right\} \qquad (5\text{-}17)$$

The nonlinear element is followed by a narrow-band filter centered at the carrier frequency ω_c, so that only terms containing $\cos(\omega_c t + \varphi_c)$ are passed. By expanding $\cos^k(\omega_c t + \varphi_c)$ and retaining only the $\cos(\omega_c t + \varphi_c)$ terms,

$$y = \sum_{n=1}^{N} \sum_{k=1,3,5\ldots}^{n} a_n \binom{n}{k} x_1^{(n-k)} \times \frac{k!}{2^{k-1}\left(\frac{k-1}{2}\right)!\left(\frac{k+1}{2}\right)!} \times \cos(\omega_c t + \varphi_c) \qquad (5\text{-}18)$$

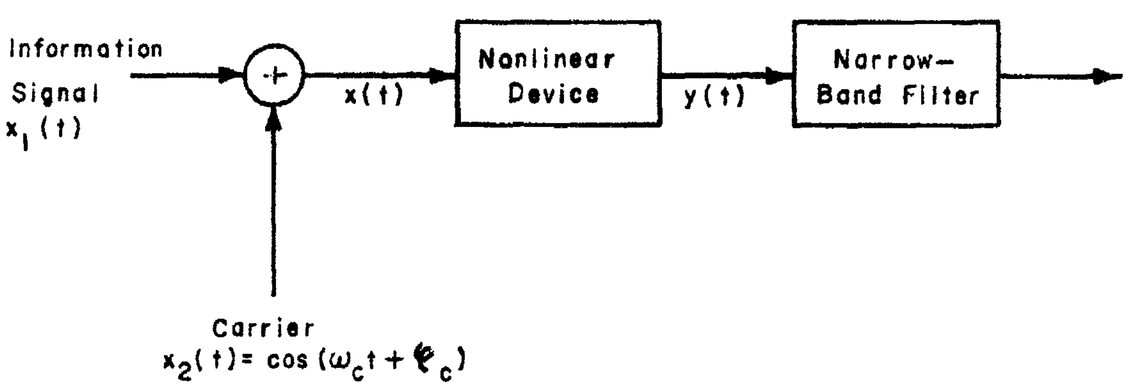


Figure 5-89. Basic Elements of Product Modulators

The x_1 terms higher than the first degree are distortion terms which extend the sidebands to $(N - 1)$ times the frequency band of the modulating signal. For instance, if $N = 7$,

$$
\begin{aligned}
y = \Bigg[&\left(a_1 + \frac{3a_3}{4} + \frac{10a_5}{16} + \frac{35a_7}{64}\right) \\
&+ \left(2a_2 + \frac{12a_4}{4} + \frac{60a_6}{16}\right)x_1(t) \\
&+ \left(3a_3 + \frac{30a_5}{4} + \frac{210a_7}{16}\right)x_1^2(t) \\
&+ \left(4a_4 + \frac{60a_6}{4}\right)x_1^3(t) \\
&+ \left(5a_5 + \frac{105a_7}{4}\right)x_1^4(t) \\
&+ 6a_6x_1^5(t) + 7a_7x_1^6(t)\Bigg]\cos(\omega_c t + \varphi_c) \qquad (5\text{-}19)
\end{aligned}
$$

If coefficients a_n are zero for $n \geq 3$, i.e., if the device is purely square law, the output is exactly as required.

The component $x_1^2(t)\cos(\omega_c t + \varphi_c)$ in Eq. 5-19 contributes sideband energy covering a band equal to four times the modulating frequency rather than two times, as in the ideal case; the $x_1^6(t)\cos(\omega_c t + \varphi_c)$ component contributes energy in a band twelve times the original modulating frequency. Thus, even when the signal input $x(t)$ is not nonlinearly distorted by the baseband-forming circuits, it may become distorted in the modulation process. Similar observations can be made in the case of FM, where the signal waveform is required to vary linearly in the instantaneous frequency of a carrier. One mechanism of frequency modulation makes use of product modulators just as does AM, except that the carrier phase is different from that in AM (this mechanism generates a low deviation FM signal which must be frequency multiplied to get a high deviation signal (see Ref. 27)). Here, too, the product modulator will generate unwanted sideband if the square law does not hold. If the frequency modulation is obtained by voltage variation of capacity, or by varying pulse duration of frequency or a relaxation oscillator, similar phenomena will be observed. The frequency is never a perfectly linear function of the applied voltage and some nonlinear distortion is unavoidable.

In the case of AM, one also must be careful to avoid overmodulation. When this occurs, the carrier is cut off and the sideband splatter components will be very large.

Measured and calculated results of excessive sideband output are reported by Firestone, et al. (Ref. 28). In an AM transmitter having an audio-modulating signal of 3 kHz and 54.7% modulation, the spectrum distribution shown in Fig. 5-90 was obtained. The total harmonic distortion amounted to 3.3%, which is not an excessive figure. A receiver operating in an adjacent channel 10 to 15 kHz from the interfering signal will receive splatter components of the order of 60 dB below the level of the carrier component. If, as may readily happen with a nearby transmitter, the carrier power of the transmitter produces at its frequency an input voltage of about 100 mV at the receiver, the level of unwanted sidebands will correspond to about 100 μV. This often will be far greater than the level of a desired signal at the sideband frequency.

5-8.4.2 Control of Sideband Splatter

To minimize splatter, it is desirable to choose nonlinear elements with characteristics as nearly ideal as possible. For a given device intended to be used as a square-law modulator, usually it will be possible to find empirical operating conditions which result in a characteristic that is nearly square law so that the higher order coefficients are small. Furthermore, the final tuned circuits in the transmitter also will act to reduce the level of the unwanted components. As a rule, however, the circuit Q cannot be made high enough to completely eliminate the splatter components. An effective technique to eliminate the unwanted sidebands associated with even powers of $x_1(t)$ in Eq. 5-18 is to use a balanced modulator.

5-8.5 HARMONIC GENERATION AND SUPPRESSION

A pure sinusoid passing through a nonlinear device generates components at frequencies which are an integral multiple of the applied sinusoid frequency. Modulated signals similarly applied give rise to components concentrated around integral multiples of the center frequency of the applied signal. The modulation sidebands may remain unchanged or they may become distorted. In some devices the nonlinearity is incidental to normal linear amplification as in class A tube and transistor amplifiers. In other devices, nonlinearity is a consequence of linear amplification by impulsive re-enforcement of a wave. RF amplifiers, of the class B or C type, pulse a tank circuit throughout a portion of the sinusoidal cycle and are therefore in this category. Klystrons and magnetrons are also in this class. Oscillations continue to

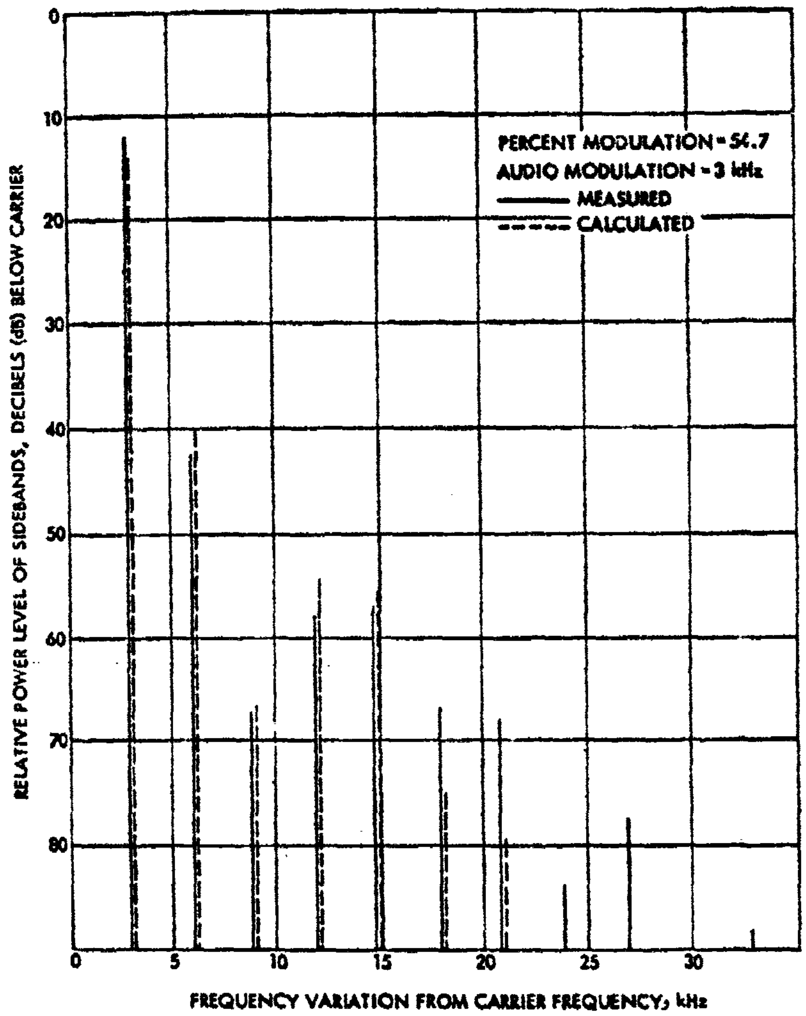


Figure 5-90. Calculated and Measured Results of the Typical Spectrum Distribution of Sidebands Produced by Amplitude Modulated Transmitters

build up until the output is limited by saturation and cutoff of the active element; hence, they also function in this manner. In other cases, the harmonics are an unwanted byproduct of a desired nonlinear function. It was pointed out earlier that the modulator may distort the modulating signal; it also may distort the RF signal. The ideal frequency multiplier yields only the desired harmonic but this is unusual. Many unwanted harmonics are present as well as the original input which becomes an unwanted component.

It often will be possible to find modes of operation for the active devices which minimize harmonic amplitudes, but this occurs usually at the cost of efficiency. The extent to which such efforts are successful depends on the amount of filtering used. Low-level circuits in the transmitter, incorporating many frequency selective circuits, are unlikely to result in significant harmonic output from the final stage. Some mechanisms for minimizing harmonic content are now described in more detail.

5-8.5.1 Balanced Modulator

Modulators for AM signals and signals related to AM are multipliers. The modulation function is performed perfectly when the sum of the low-frequency signal and a carrier is delivered to a square-law device. If $x(t)$ is the low-frequency signal and $\cos \omega_c t$ is the carrier, then the output of a square-law device is the squared sum of these input signals:

$$[x(t) + \cos \omega_c t]^2 = x^2(t) + 2x(t)\cos \omega_c t + \frac{1}{2} + \frac{1}{2}\cos 2\omega_c t \tag{5-20}$$

The desired term is $2x(t)\cos \omega_c t$. The term $x^2(t)$ is a distorted low-frequency signal which is readily rejected in the RF filter following the modulator; the dc term, 1/2, is similarly rejected. Some energy at the second harmonic of the carrier $(1/2)\cos 2\omega_c t$ may trickle through, especially if the output circuit is not highly selective. The use of a properly designed balanced modulator shown in Fig. 5-91 eliminates transmission of second harmonics. If each nonlinear device has an output-input characteristic given by the power series

$$y = a_0 + a_1 x + a_2 x^2 \tag{5-21}$$

where y is the output, x is the input, and x_a and x_b have values shown in Fig. 5-91, the total output may be written

$$\begin{aligned} y_a - y_b &= a_1(x_a - x_b) + a_2(x_a^2 - x_b^2) \\ &= 2a_1 x(t) + a_2[x(t) + \cos \omega_c t]^2 \\ &\quad - a_2[-x(t) + \cos \omega_c t]^2 \\ &= 2a_1 x(t) + 4a_2 x(t)\cos \omega_c t \end{aligned} \tag{5-22}$$

No second harmonic term is present, assuming perfect balance, and the term involving $x(t)$ alone is rejected in the output tank circuit. This arrangement, however, leaves second harmonics in the output if the degree of nonlinearity is not limited to square law. If, for instance, a term a_3x^3 were added to Eq. 5-21, the output would contain a term

$$6a_3x_1(t)\cos^2\omega_c t = 3a_3x_1(t) + 3a_3x_1(t)\cos 2\omega_c t \tag{5-23}$$

which has a second harmonic modulated component. [The rejection of splatter components involving even powers of $x_1(t)$ can be illustrated also by adding a_3x^3 and expanding as in Eq. 5-23.]

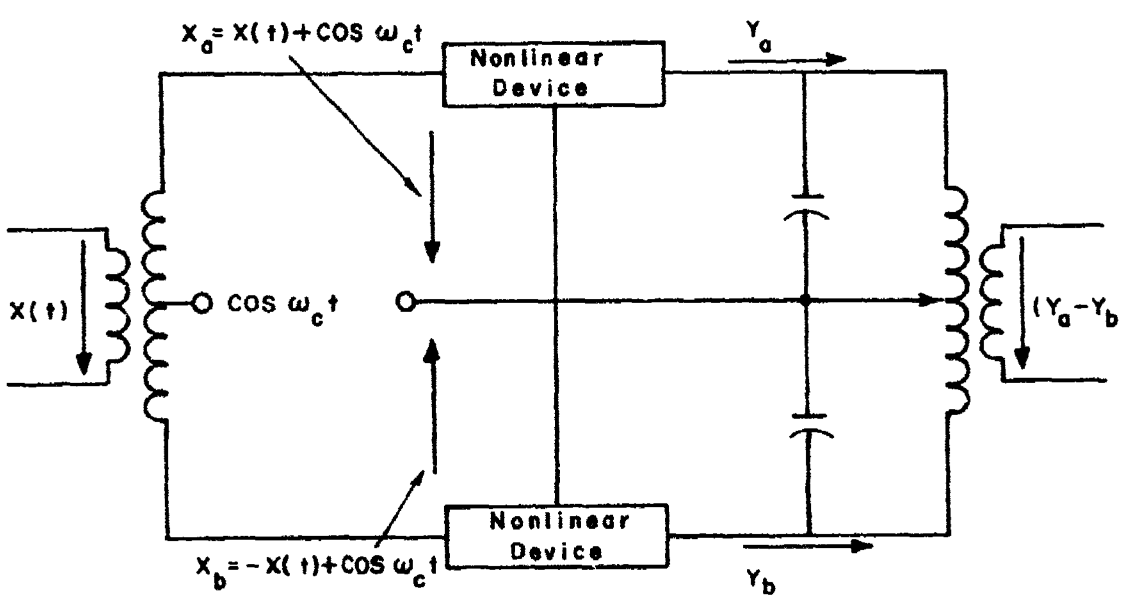


Figure 5-91. Schematic Diagram of a Balanced Modulator

5-8.5.2 Amplifier Linearity Control

It is common practice to use crystal controlled oscillators operating at moderate frequencies (i.e., 30 MHz and below) and to employ frequency multiplication to obtain the needed output carrier frequency. In such systems electronic devices biased at or below cutoff (e.g., in class B and C amplifiers) are the source of the carrier output, which is rich in harmonics. The current pulse is approximately of the form of the cap of a sinusoid as shown in Fig. 5-92 and given by the expression

$$i(t) = \begin{cases} A(\cos \omega t - \cos \theta), & \\ \quad \text{for } (2n\pi - \theta) \le \omega t \le (2n\pi + \theta), & \\ \quad n = 0, \pm 1, \pm 2, \ldots, & \\ 0 & \text{, otherwise} \end{cases} \tag{5-24}$$

The Fourier expansion of this wave is given by

$$i(t) = \frac{A}{\pi(1 - \cos \theta)} \left\{ \sin \theta - \theta \cos \theta + \sum_{n=1}^{\infty} \left[\frac{\sin(n+1)\theta}{n+1} + \frac{\sin(n-1)\theta}{n-1} + \frac{2 \sin n\theta \cos \theta}{n} \right] \cos n\omega t \right\} \tag{5-25}$$

Often it is found that the sinusoidal cap is itself somewhat distorted in passing through the amplifier, so that higher levels of harmonics appear. Squared sinusoidal caps, given by the expression

$$i(t) = \begin{cases} A^2(\cos \omega t - \cos \theta)^2, & \\ \quad \text{for } (2n\pi - \theta) \le \omega t \le (2n\pi + \theta), & \\ \quad n = 0, \pm 1, \pm 2, \ldots, & \\ 0 & \text{, otherwise} \end{cases} \tag{5-26}$$

may be better approximations of the magnitudes of current pulses from some electronic devices than the undistorted caps illustrated in Fig. 5-92 (Ref. 29). The relative magnitudes of the first three harmonics for the undistorted and square sinusoidal caps are shown in Fig. 5-93 as a function of conduction angle 2θ. In general, the larger the conduction angle, the smaller will be the amplitude of the harmonic in the output.

It is, in principle, possible to estimate the harmonic output of the transmitter using the levels determined by Eq. 5-25 or Fig. 5-92 together with the response characteristic of the tuned circuits following the source of harmonics. Such a procedure is described in Ref. 28. As a rough approximation, a single-tuned circuit with a $Q = 10$ will attenuate the second, third, and fourth harmonics approximately 24, 30, and 33 dB, respectively. Doubling the Q will increase the attenuation in each case by about 6 dB.

In the usual case, the final amplifier will contribute the major amount of harmonic output. The reasons for this are that ordinarily the final amplifier is driven hard in order to get as much efficiency as is possible and the tuned circuits following the final amplifier have limited selectivity, particularly at the lower frequencies.

Frequently, final amplifiers are operated as class C amplifiers. In principle, the harmonic output from class C amplifiers is not different from that of the frequency multipliers previously discussed. As the conduction angle is increased, the harmonic output decreases, but so does the efficiency. To reduce interference and obtain the effect of a linear amplifier, it is obviously advantageous to use a push-pull class B final amplifier. It may even be advantageous to use class C amplifiers in push-pull since this would tend to cancel the even harmonics. The amplitudes of odd

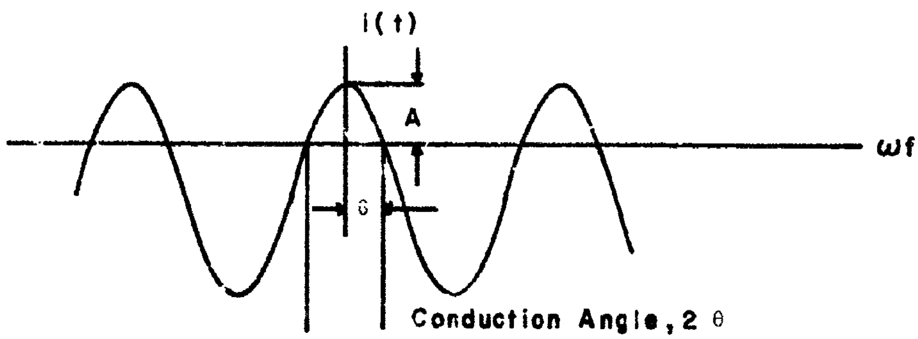


Figure 5-92. Cosinusoidal Cap Waveform

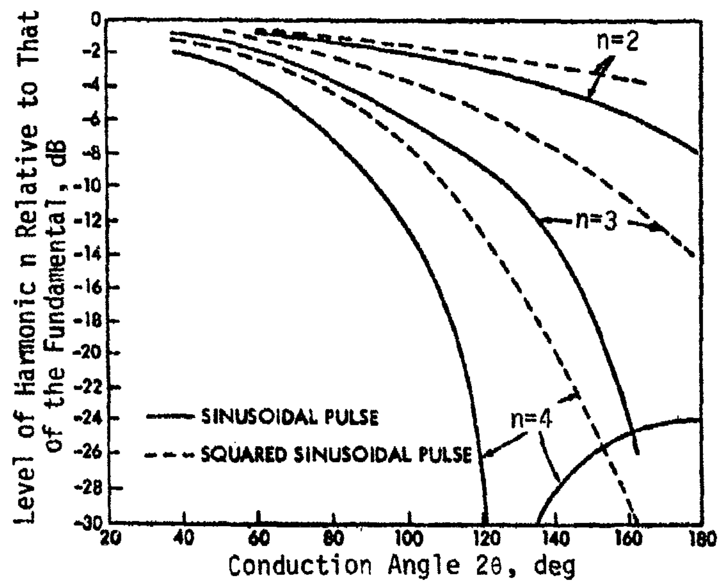


Figure 5-93. Relative Intensity of Harmonics for Two Pulse Shapes as a Function of Conduction Angle

harmonics relative to that of the fundamental would be unaffected by tubes in push-pull, but these harmonics are attenuated readily by tuned circuits in the output.

The harmonic outputs of 14 representative communication transmitters are shown in Fig. 5-94 (Ref. 30). Note that some transmitters produce harmonics whose levels are within 30 dB of that of the fundamental. For such harmonic content, a 50-kW transmitter will emit a 50-W signal at the harmonic frequency — in this case hardly insignificant. Transmitters made in accordance with MIL-STD-461 will have to satisfy requirements of Test CE-06. This limits harmonic and spurious emissions to less than −44 dBW in all cases, and for transmitters with power in the range 20-40 dBW, the limit of harmonic and spurious emission is −60 dBW.

5-8.5.3 Microwave Circuit Design

Amplifiers for microwave frequencies may have significant harmonic output. The klystron, which is used extensively, generates harmonics through the bunching process. The current induced in the catcher cavity I_c has a harmonic intensity given approximately by (Ref. 31).

$$I_c(n) = 2I_o J_n(nx) \text{ , A} \tag{5-27}$$

where

n = order of the harmonic
I_o = direct current in the tube, A
J_n = nth order of Bessel function of the first kind

and

$$x = \frac{\omega s V_1}{(e/m)^{1/2}(2V_o)^{3/2}} \text{ , bunching parameter, dimensionless}$$

in which

ω = frequency of excitation, rad/sec
s = spacing between input and output cavities, m
V_1 = peak gap voltage at buncher, V
e = electronic charge, C
m = electronic mass, kg
V_o = beam accelerating voltage, V

The bunching parameter value of $x = 1.84$ yields maximum efficiency. At maximum efficiency, the pulses of induced current are very high. To avoid large harmonic outputs, the cavity Q must be high.

For loaded circuits, values of Q ranging from 500 to 1000 are obtainable. By using such values of Q, it should be possible to attenuate harmonic outputs more than 60 dB below the fundamental, even with equal fundamental and second harmonic current in the tube. Measurements reported on a particular

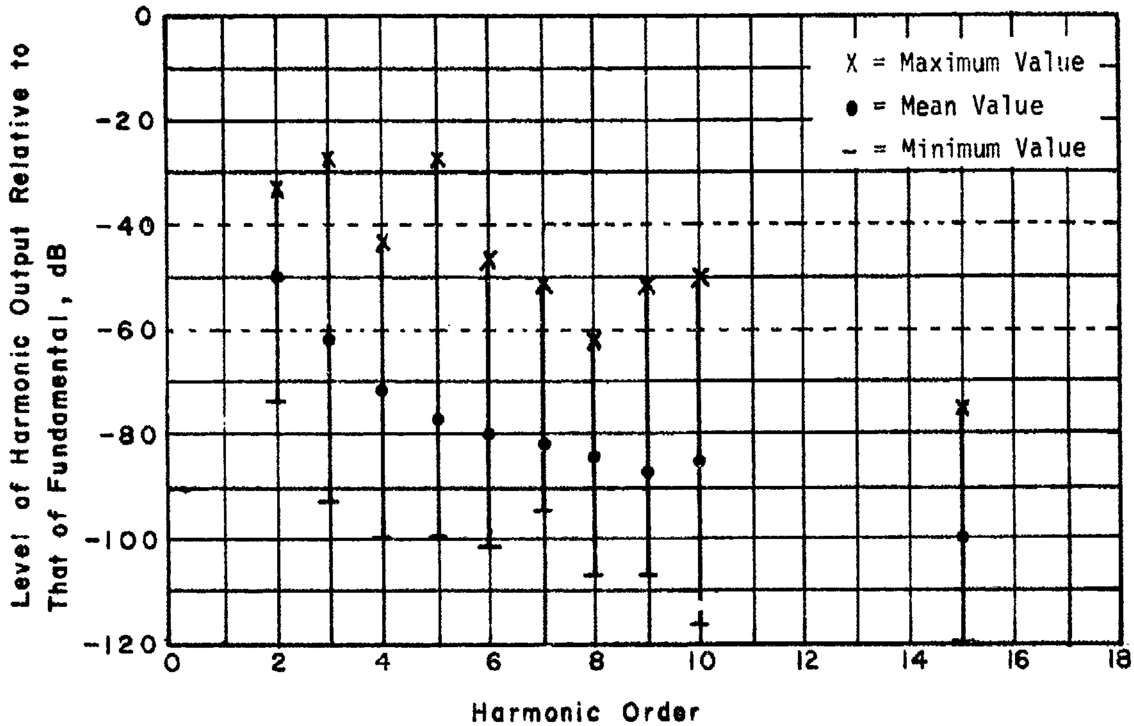


Figure 5-94. Typical Values of Harmonic Emissions Determined from Measurements on 14 Radio Transmitters

pulsed klystron (Ref. 32) are reproduced in Fig. 5-95. Variation of beam voltage, which alters the bunching parameters, is seen to alter the second-harmonic content markedly. However, at the value of beam voltage for which the second harmonic is minimum, the second harmonic power is only about 40 dB below the fundamental output.

The traveling wave tube (TWT) appears to be a good choice as a microwave power amplifier, generating little interference. The operation of the TWT depends upon uniform re-enforcement of a traveling wave and does not involve impulsive re-enforcement of a field, as in a klystron. Yet, reports (Ref. 31) indicate that, in particular cases, the second-harmonic output in TWT's may range from 20 to 40 dB below the fundamental. These conditions exist only at maximum efficiency when the tube is driven heavily. The input-output characteristic for specific traveling wave tubes is found to be linear over a given range of inputs: beyond the linear range, saturation is reached as shown in Fig. 5-96 which shows two curves, one giving output power P_o vs input power P_i, the other giving amplitude to phase conversion factor K vs input power P_i, as measured using a typical TWT employed in radio relay application. The output power vs input power curve saturates beyond a certain point, and operation into this region will give rise to observable harmonic content. The amplitude-to-phase conversion factor curve suggests another mechanism by which unwanted components are generated. In effect, amplitude variations at the input give rise to phase variations at the output, thereby spreading the output spectrum. In multicarrier systems, such as in radio relay applications, the phase effects as well as the amplitude effects give rise to the generation of intermodulation components (Ref. 33).

High-power magnetrons generally are used in radars as sources of pulsed RF energy, coupled directly to the antenna. In the magnetron, the energy of a rotating cloud of electrons is imparted to the field in tuned cavities on the periphery of rotation. The mechanism which selects electrons emitted from the cathode, where they are roughly in proper phase, tends to maintain them in bunches as in the klystron.

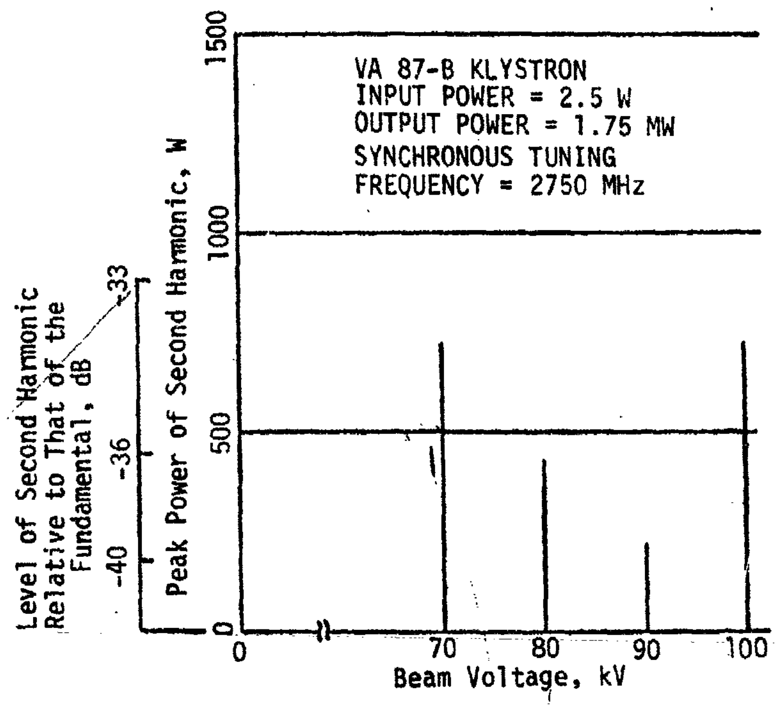


Figure 5-95. Harmonic Output as a Function of Beam Voltage for VA 87-B Klystron

The spatial distribution of charge is not sinusoidal, and, again, pulses rather than sinusoids are induced in the cavities. The magnetron cavities also can be excited in a harmonic mode so that a large number of output components are possible in addition to the easily identifiable harmonics. Harmonic level and frequency are both difficult to predict. A value of −40 dB with respect to the fundamental has been quoted by Tomiyasu (Ref. 32) for the second harmonic, but the third harmonic was said to reach −20 dB.

Cridlan (Ref. 34) presents measured data on the spurious outputs of L- and C-band radars. It is not made clear whether the radars use magnetrons or klystrons, but it is pointed out that both of these devices generate nonharmonic spurious emissions. The L-band radar is shown to have many spurious components each of the order of 10-W EIRP spread across the band from its fundamental to its second harmonic frequency. The C-band radar is shown to have 10- to 20-W components, both above and below its fundamental, not at a harmonic frequency. A point to be observed about magnetrons is that the output characteristics depend on the load (known as the phenomena of "pulling") and that the spurious emissions may vary therefore with antenna sweep.

From the foregoing, it is evident that harmonic intensities can be reduced by operating oscillators and amplifiers over linear regions at reduced efficiencies, canceling nonlinear components in balanced circuits, and filtering. The first of these involves an increase in power consumption and heat generation. As a result, more effort must be spent on heat removal. Balanced circuits and filtering involve additional circuits: the balanced push-pull circuit eliminates only even-order harmonics. Filtering is, by far, the most practical method for suppressing harmonics. Common forms use the wave trap or bypass principle. In the case of a transmission-line system, the filter elements may take the form of shunt or series stubs. One application of such a filter to remove second harmonics is shown in Fig. 5-97.

The difficulty with wave-trap techniques is that power at the harmonic frequency is reflected back to the generator, causing an impedance mismatch. In theory, one way of avoiding this difficulty is to use an isolator between the generator and the filter. Unfortunately, most isolators are designed to operate at the fundamental frequency, and their performance at harmonic frequencies is usually unknown. This is especially true at the higher microwave frequencies. If

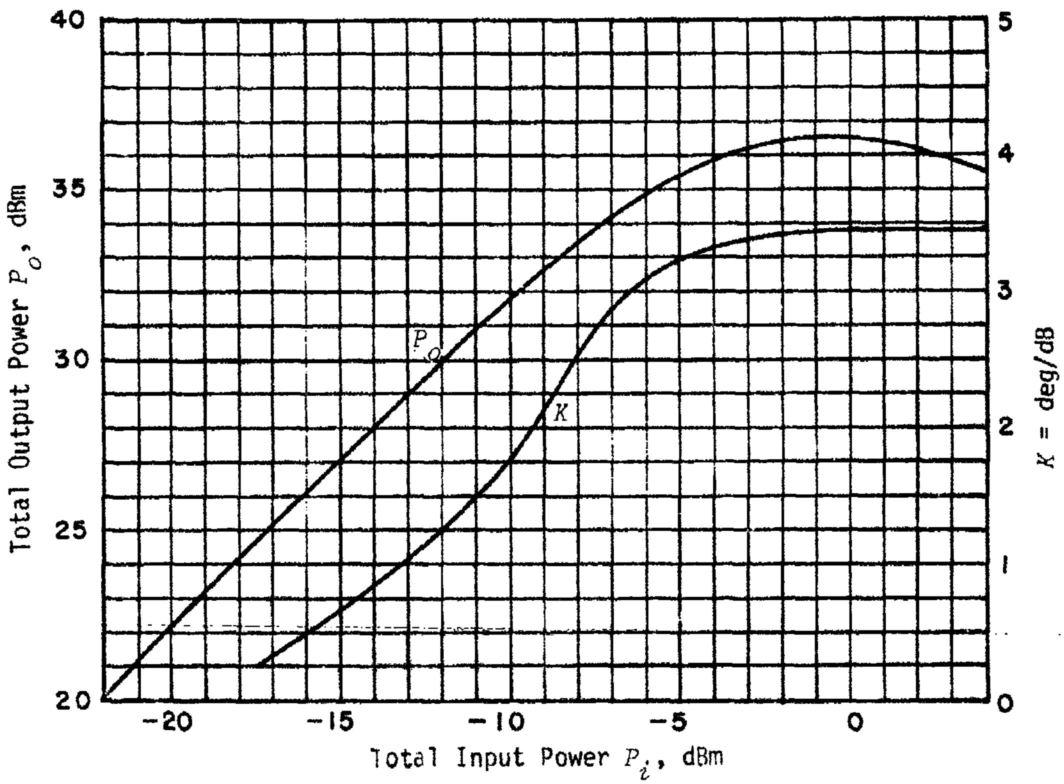


Figure 5-96. Single Carrier Power Transmission Characteristic and AM-PM Conversion Factor for a Typical TWT

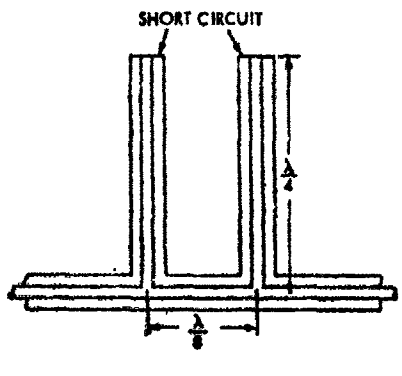


Figure 5-97. Diagram of Two Short-Circuited Stubs (each 1/4 wave-length long and spaced 1/8 wavelength at the fundamental frequency) Arranged to Attenuate the Second Harmonic

an isolator is used, it must be able to dissipate the power of all reflected components (fundamental plus harmonics) without exceeding its rating.

It is possible to use a ferrite circulator to suppress harmonics. For example, the configuration in Fig. 5-98 could be used to divert the reflected energy into a resistive load. Unfortunately, the circulator has bandwidth problems as does the isolator. Another harmonic filtering-absorbing technique directionally couples the unwanted signals into a second wave-guide, where they are absorbed.

A simple method for harmonic absorption, which also affords some reduction of the level of harmonics in the main load, can be achieved merely by coupling to the main waveguide a smaller waveguide whose cutoff frequency is below the fundamental but above that of all harmonics.

Additional filtering techniques are given in Refs. 25 and 35.

5-8.6 TRANSMITTER INTERFERENCE

The spurious sidebands discussed in par. 5-8.4 can be reduced through careful design of the modulator, but it frequently is found that a significant background interference level still exists around the transmitter carrier center frequency. This results from oscillator noise modulating the carrier in amplitude, frequency, or phase. The interference is associated mainly with the oscillator itself, which is quite noisy compared to amplifiers, but some interference is produced in the amplifiers as well. It is possible for corona to form, or for arcing to occur, at high-voltage points in the final amplifier. Subsequent filtering in the output tank circuit results in a pair of noise sidebands on either side of the carrier frequency; spurious levels are difficult to estimate accurately.

Measurements of noise power are appropriate though they are difficult to make because of the presence of the much larger power in the carrier. Some measurements have been reported on VHF transmitters (Ref. 36), and the results of tests on a 45-MHz and a 160-MHz transmitter are shown in Fig. 5-99. The basic crystal frequencies were 3.75 MHz and 6.67 MHz, respectively, and three or four frequency-multiplier stages were used before the final amplifier. The tests were performed on an unmodulated transmitter, but the effect of modulation on the noise sidebands was not determined. Data, not presented here, was also obtained for the power amplifier alone, without the exciter and its multipliers; the resulting

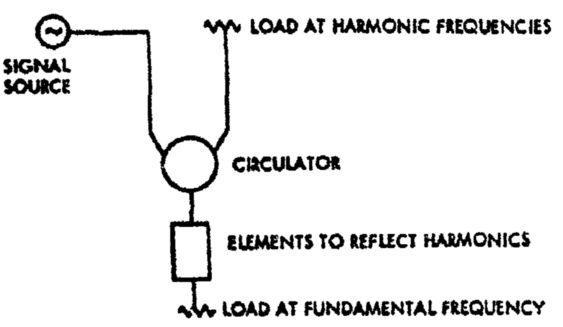


Figure 5-98. Diagram Illustrating the Use of a Circulator to Separate Harmonics

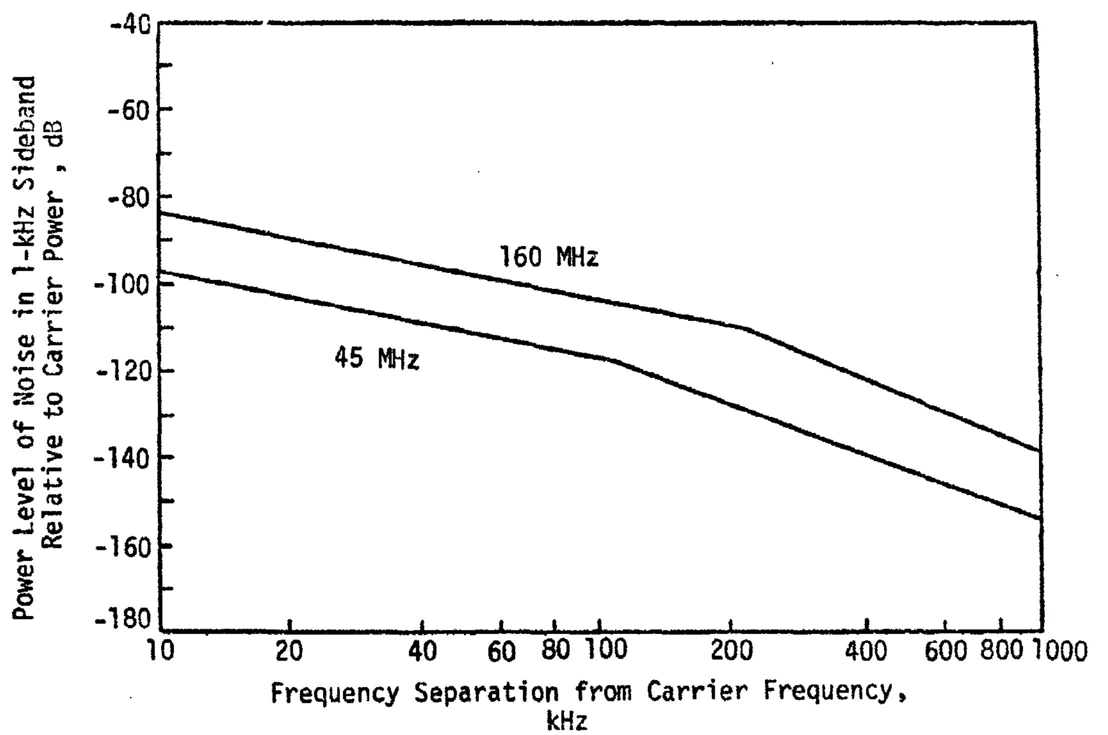


Figure 5-99. Measured Values of the Sideband Noise Level of Unmodulated 45- and 160-MHz Transmitters as a Function of Frequency Separation from the Carrier Frequency

From the Post Office Electrical Engineers Journal, UK.

noise was generally about 20 dB below that shown in Fig. 5-99. The noise levels are quite low in these cases, but the noise interference injected into a receiver can be quite significant when transmitter and receiver antennas are very close to each other.

5-8.7 INTERMODULATION AND CROSS MODULATION

The phenomena of intermodulation and cross-modulation imply the mixing of two or more signals in a nonlinear element in such a way that multiplicative mixtures of the two signals result. This occurs in receiver input circuits, as well as in transmitters, and sometimes in a nonlinear element in the channel. As far as transmitters are concerned, the process involves the reception of an unwanted signal by the transmitting antenna, which conducts it back to the output plate of the final amplifier, where it is mixed with the transmitted signal. Therefore, the process is of greatest significance when both the unwanted signal and the nonlinear product are within the passband of the final amplifier. This can occur if the nonlinear device has a characteristic of odd degree. For example, suppose the output signal has a component that depends on the third power of the output voltage. Two components of voltage at the output are the output signal voltage itself $x_1(t)$

$$x_1(t) = v_1(t)\cos[\omega_1 t + \varphi_1(t)] \tag{5-28}$$

and the unwanted signal voltage $x_2(t)$

$$x_2(t) = v_2(t)\cos[\omega_2 t + \varphi_2(t)] \tag{5-29}$$

The sum $x_1(t) + x_2(t)$, when applied to a nonlinear device having a cubic term, will result in a number of intermodulation components, one of which is of the form

$$y_a(t) = v_1^2(t)\, v_2(t)\cos[(2\omega_1 - \omega_2)t + 2\varphi_1(t) - \varphi_2(t)] \tag{5-30}$$

and cross modulation components, one of which is of the form

$$y_b(t) = v_1^2(t)\, v_2(t)\cos[\omega_2 t + \varphi_2(t)] \tag{5-31}$$

The center frequency of the intermodulation component differs from either ω_1 or ω_2; but, if the two frequencies are close to one another, the frequency of the new term is not too different from ω_1. The modulation on the term whose frequency is $(2\omega_1 - \omega_2)$ is a combination of the modulations on each of the causative signals. The cross-modulation component has a frequency equal to one or the other of the original signals, and its amplitude is a combination of the original amplitude coefficients of both mixed signals. The intermodulation term is usually more serious since it occurs in an entirely new frequency band. The cross-modulation term, given in Eq. 5-31, implies that the unwanted signal is reradiated with a distorted envelope. As a rule, the amplitude of the reradiated signal is much smaller than that of the original unwanted signal, defined in Eq. 5-29, and the reradiated signal might be viewed as a small new splatter or noise component.

5-8.8 OTHER SPURIOUS OUTPUTS

Parasitic oscillations occur in both low- and high-frequency amplifiers. In low-frequency devices, they result from stray external and internal capacitances and inductances that form spurious feedback loops. The cures are often very simple, sometimes involving inserting resistors in the leads to the grid and plate of the amplifier to increase losses of the spurious tuned circuit. The methods of handling these situations are well described in standard reference books (Ref. 37).

At microwave frequencies, corresponding phenomena exist. Often, the phenomena are completely internal to the tube. At some frequencies, transit-time effects are sources of negative resistance. If a tuned circuit somewhere in the structure is coupled to the negative resistance, parasitic oscillations occur. Various additional mechanisms of microwave tube spurious emission are referenced and surveyed in Ref. 32. These include harmonic and unharmonic generation in magnetrons; harmonic generation, diode oscillation, and drift tunnel oscillation in klystrons; and harmonic generation, diode oscillation, band edge, and second passband oscillation in travelling wave tubes.

5-9 RADAR EQUIPMENT

5-9.1 INTRODUCTION

Par. 5-9 is concerned with radar systems including equipment for modulation, transmission, and radiation of electromagnetic energy (usually pulses), and reception, amplification, detection, processing, and display of echoes received from desired targets. The simplest types of radars provide target range by measuring time of occurrence of the echoes. Search radars provide angular information, as well, by scanning a narrow antenna beam over a sector. Tracking radars provide accurate real-time position data by setting up suitable servo loops for continual determination of range gate position as well as angular beam pointing directions. Radar outputs can be used to display a visual picture of the target environment or can provide direct inputs to fire-control equipment for direction of artillery or rocket projectiles to a moving target location. Radar systems produce emissions that can interfere with other equipments (such as communication, navigation, and other radar equipments) and can be sensitive to external sources of interference. The objective here is to describe the harmful effects which can occur and techniques by which they can be reduced.

5-9.2 RADAR EMISSION CHARACTERISTICS

A determining factor for the power output of radar emission is the requirement for discernible echoes from targets at long ranges. This imposes the use of orders of magnitude more signal power than required by other types of systems (e.g., communication or navigation) over similar ranges. For line-of-sight transmission between two locations, the signal power P_r is given by (see Eq. 3-147)

$$P_r = \frac{\lambda^2}{(4\pi r)^2}\left(\frac{G_t G_r P_t}{L}\right), \text{ W} \qquad (5\text{-}32)$$

where

P_t = transmitted power, W
G_t = transmitter antenna gain, dimensionless
G_r = receiving antenna gain, dimensionless
λ = wavelength, m
r = range, m
L = loss factor, dimensionless

For a monostatic radar, however, (transmitter and receiver co-located) the signal power received is given by:

$$P_r = \sigma \frac{\lambda^2}{(4\pi)^3 r^4}\left(\frac{G_t G_r P_t}{L}\right), \text{W} \qquad (5\text{-}33)$$

where σ is the radar cross section in square meters (Ref. 38) of the "point target" at range r. Typical power requirements as determined by Eq. 5-32 may be in the several megawatt range, producing field strengths of the order of a hundred volts per meter at distances from the radiating antenna of a few hundred meters.

5-9.2.1 Frequency Band Utilization

Current practice in spectrum use of radar systems is (Ref. 38) summarized in Table 5-9. The first column gives the designations commonly used for different frequency bands. The second column gives the range of frequencies associated with each band. Corresponding wavelength ranges are given in the third column (obeying the relationship $\lambda = c/f$ where f is frequency and c is the speed of light).

The data on wavelength are important in relating the approximate beam directivity to antenna size. For beam angles θ less than about 0.1 rad, the width of the angle is related to the antenna effective diameter D by the relation:

$$\theta \approx \lambda/D \text{ , rad} \qquad (5\text{-}34)$$

where λ is the wavelength. Thus, in the SHF band, assuming $\lambda \approx 3$ cm, $D = 3$ m, the beamwidth θ will be approximately 1/100 rad or 0.57 deg. In general, the lower frequencies (up to about 1-2 GHz) are used for long-range search applications, providing good moving-target indication and relative immunity from weather effects. The higher frequencies (up to about 100 GHz) are used where light weight is important and also where accurate tracking is required.

5-9.2.2 Time-Frequency Characteristics of Radar Pulse Waveforms (Ref. 38)

Interference standards (Ref. 39) place a limit on the "emission bandwidth" of a given radar. The shape of

TABLE 5-9. RADAR FREQUENCY BANDS

Band Designation	Frequency Range	Wavelength Range
HF (high frequency)	3-30 MHz	10-100 m
VHF (very-high frequency)	30-300 MHz	1-10 m
UHF (ultra-high frequency)	0.3-3 GHz	10-100 cm
SHF (super-high frequency)	3-30 GHz	1-10 cm
EHF (extremely-high frequency)	30-300 GHz	1-10 mm

Note: Above designations officially have superseded the following radar band designations referred to frequently in the recent radar literature: L, 1-2 GHz; S, 2-4 GHz; C, 4-8 GHz; X, 8-12.5 GHz; K_u, 12.5-18 GHz; K, 18-26 GHz; K_a, 26.5-40 GHz.

the pulse modulation waveform used has a direct effect on the amount of energy emitted outside any prescribed band. This results in a set of requirements on system design distinct from those imposed by the desired radar performance. Typical functions of a radar equipment reduce to the following set of measurements or decisions:

a. Determine the presence of an echo in a given interval of time (range resolution cell).

b. Determine the time of occurrence of a given echo response (range accuracy).

One figure of merit for performance of these functions is the signal-to-noise ratio R

$$R = \frac{E}{N_o}\ , \text{ dimensionless} \qquad (5\text{-}35)$$

where

N_o = noise spectral density, V^2/Hz
E = signal "energy" per pulse defined by:

$$E = \int_{-\infty}^{\infty} |v(t)|^2 dt = \int_{-\infty}^{\infty} |V(f)|^2 df \ , V^2\text{-s} \qquad (5\text{-}36)$$

Note than in Eq. 5-36 $v(t)$ is the waveform of a single pulse (time domain); $V(f)$ is its spectrum (frequency domain).

Additional waveform characteristics are also significant in the time measurement implied by (b); in this case, not only accuracy of measurement is involved but also the ability of the system to resolve closely-spaced multiple targets.

The performance of a radar system with respect to the range accuracy and range resolution requirements that are frequently used as the basis of system design can be expressed as a direct function of the moment bandwidth B_m of the modulating waveform given by

$$B_m = \left[\frac{\int_o^{\infty} f^2 |V(f)|^2 df}{\int_o^{\infty} |V(f)|^2 df}\right]^{\frac{1}{2}}, \text{ Hz} \qquad (5\text{-}37)$$

In achieving this moment bandwidth, the behavior of the spectrum $V(f)$ at frequencies far removed from a desired B_m will have very little effect on the value of B_m but may be paramount in determining the extent to which the radar may interfere with systems using contiguous frequency regions. Thus, the requirement for minimization of interference effects adds a new dimension to the waveform design problem.

The energy ratio in Eq. 5-35 depends directly on the product of peak power by effective duration of the radiated waveform. Since peak power is an important limitation on available transmitting tubes (i.e. magnetrons and klystrons), energy increase to achieved desired range must be achieved by corresponding increase in the duration of the pulse waveform. Usually it is necessary to use waveforms of high intrinsic bandwidth to provide the necessary range accuracy and resolution capability. Hence, the use of high time-bandwidth product waveforms, together with suitable matched filters at the receiver for pulse compression, is common practice in radar system design.

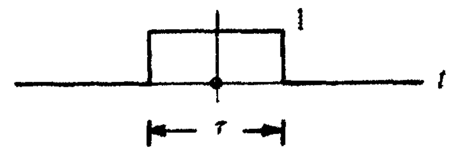
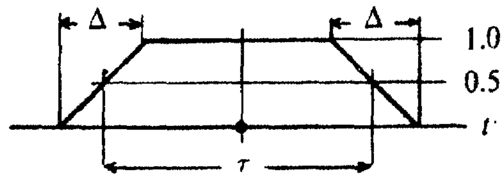
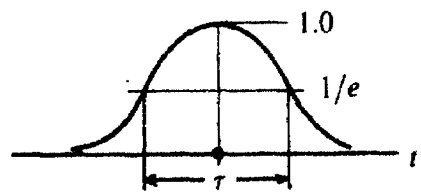
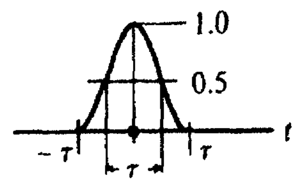
Four representative pulse waveforms are shown in Table 5-10. The rectangle (a) is the simplest one used in practice and will be shown to have the least favorable out-of-band spectral characteristics. The gauss (c) is the optimum with the trapezoid, and raised cosine (d) representing compromises between desired performance and ease of realization. The first column gives their shapes and defines a "duration" parameter τ for each. The third column gives the mathematical expression $v(t)$ for each, while the last column gives the spectrum $V(f)$ (the Fourier transform of $v(t)$) for each. The energy spectral density $|V(f)|^2$ can be readily calculated to determine the relative intensity at different frequencies.

The spectra given in Table 5-10 can be interpreted as follows (Refs. 38 and 40). If a repetitive train of pulses is generated, the resultant Fourier series coefficients will have magnitudes proportional to $|V(f_n)|$ where f_n is the nth harmonic of the pulse repetition frequency. Further, if the pulse or pulses are used to modulate a carrier of frequency f_o, the resultant spectra will be the displaced values $V(f-f_o)$.

Expressions for the energy effective bandwidth B_e and moment bandwidth B_m of the representative waveforms are given in Table 5-11. Note that the moment bandwidth of the simple rectangular pulse is infinite. This is caused by the discontinuities in the waveform, which make it hard to realize physically as well.

A listing of numerical bandwidth characteristics for these waveforms is given in Table 5-12. In addition to the effective and moment bandwidths B_e and B_m discussed, there are included the commonly used 3-dB and 6-dB bandwidths frequently used as references (the frequencies at which $|V(f)|$ is 3 or 6 dB below the maximum $|V(0)|$) as well as the 40-, 60-, 80-, and 100-dB bandwidths, important for assessing the relative attenuation of emitted energy in frequency channels adjacent to an assigned carrier

TABLE 5-10. FOUR REPRESENTATIVE WAVEFORMS

Name	Picture	Time Function $v(t)$	Frequency Function $V(f)$
(a) rectangle		$\mathrm{rect}(t/\tau)$	$\tau\,\mathrm{sinc}(f\tau)$
(b) trapezoid		$\frac{1}{\Delta}\,\mathrm{rect}(t/\tau) * \mathrm{rect}(t/\Delta)$	$\tau\,\mathrm{sinc}(f\tau)\,\mathrm{sinc}(f\Delta)$
(c) gauss		$\exp(-4t^2/\tau^2)$	$\frac{\tau\sqrt{\pi}}{2}\exp(-\pi^2 f^2\tau^2/4)$
(d) raised cosine		$\frac{1}{2}\left(1+\cos\frac{\pi t}{\tau}\right)\mathrm{rect}\left(\frac{t}{2\tau}\right)$	$\frac{\sin(2\pi f\tau)}{2\pi f(1-4f^2\tau^2)}$

t = time; τ = waveform "duration", s; Δ = trapezoid "risetime", s; f = frequency, Hz
$\mathrm{rect}(t/\tau)$ = rectangular pulse function = 1 for $|t| < \tau/2$; 0 for $|t| > \tau/2$
$\mathrm{sinc}(f\tau) \equiv \frac{\sin \pi f\tau}{\pi f\tau}$ (maximum) (unity) at $f = 0$; zero at $f = n/\tau$ (n an integer $\neq 0$)

$*$ ≡ convolution operator, i.e., $v_1(t) * v_2(t) \equiv \int_{-\infty}^{\infty} v_1(s)v_2(t-s)ds$
s = dummy variable

TABLE 5-11
ENERGY AND MOMENT BANDWIDTHS OF REPRESENTATIVE WAVEFORMS
(Normalized with respect to duration parameter τ)

Waveform	τB_e, dimensionless	τB_m, dimensionless
(a)	$\frac{1}{2}$	∞
(b)	$\frac{1}{2}\left(1 - \frac{\Delta}{3\tau}\right)$	$\frac{1}{\pi}\sqrt{\frac{\tau}{2\Delta}}$
(c)	$\frac{1}{\sqrt{2\pi}}$	$\frac{1}{\pi}$
(d)	$\frac{3}{8}$	$\frac{1}{2\sqrt{3}}$

frequency band and in meeting required EMC specifications (Ref. 39).

For example, consider a 1-μs pulse transmission. To determine the minimum frequency separation required to have the emitted spectrum 60 dB below the carrier frequency value, the following values of B_{60} should be used:

a. rectangular pulse: 318 MHz
b. trapezoidal pulse: 31.8 MHz
c. gaussian pulse: 1.7 MHz
d. raised cosine: 3.4 MHz

Since the "duration" values defined in Table 5-10 are somewhat arbitrary, magnitudes of the different bandwidth measures relative to the corresponding 3-dB bandwidths are given in Table 5-13.

The preceding results are directly applicable to radars using simple pulse transmission. For long-range, high-resolution applications, however, it is necessary to use pulse waveforms of high intrinsic time duration — frequency bandwidth products (rather than the $B\tau \approx 1$ relation which holds for simple pulses) (Ref. 38, Chapter 3, and Ref 40). In the case of waveforms made up of sequences of contiguous subpulses with a random phase coding, the preceding results can be applied directly, now referring to subpulse waveform rather than overall waveform. For frequency-modulated (chirp) pulses, it is necessary to perform a separate calculation or make a corresponding measurement to ensure that specifications will be met.

5-9.2.3 Spurious Emissions

Transmitter tubes used in radar equipment include those which can be classified as either crossed-field tubes or linear-beam tubes (Ref. 38, Chapter 7). The crossed-field category is based on the use of high strength electric and magnetic fields at right angles to each other, and includes magnetron high-power oscillators as well as variously named crossed-field amplifiers (CFA) that achieve amplification with a nonresonant structure. Linear-beam tubes, such as klystrons and travelling-wave tubes (TWT) use a high-energy electron beam, formed by electrostatic lenses or a collinear magnetic field that interacts with suitably injected RF energy in a structure whose dimensions are related to the operating wavelength.

Tubes are available with peak power up to several megawatts. Average powers can be as much as tens or hundreds of kilowatts. Thus, spurious responses several tens of decibels below the carrier frequency power can still constitute significant interference to nearby receivers tuned to nearby frequencies.

In most radar systems, it is found economical and practical to feed the antenna, in the transmit mode, with energy from a single high-power tube or pair of tubes rather than using a large number of parallel low-power tubes to obtain the desired high energy output. However, solid-state technology currently under development (Ref. 38, Chapter 30) has the potential of altering this picture radically. In addition to high-power transistors and transistor frequency multipliers, there are a number of new bulk-effect oscillators that can generate several watts of microwave energy. Such components form the basic building blocks of several large phased-array antenna radar systems (Ref. 41). Such systems are characterized by the use of a large array of identical elements independently controlled to permit electronic steering of the emitted and received beams.

TABLE 5-12
FREQUENCY BANDWIDTH CHARACTERISTICS FOR REPRESENTATIVE WAVEFORMS, NORMALIZED WITH RESPECT TO DURATION PARAMETER τ

Feature point	(a) rect	(b) trap	(c) gauss	(d) raised cos
τB_e	0.5	0.483	0.399	0.375
τB_m	∞	0.712	0.318	0.289
τB_3	0.442	0.440	0.374	0.360
τB_6	0.602	0.600	0.529	0.499
far-frequency voltage ratio bound**	$\dfrac{1}{\pi x}$	$\dfrac{\tau}{\pi^2 x^2 \Delta}$	$\exp(-\pi^2\lambda^2/4)$	$\dfrac{1}{2\pi x(4x^2-1)}$
τB_{40}	31.8	10.065	1.366	1.637
τB_{60}	318.3	31.831	1.673	3.438
τB_{80}	3183.1	100.658	1.932	7.366
τB_{100}	31831.0	318.310	2.160	15.851

Notes:

1. B_k = frequency at which attenuation is k dB with respect to that at 0 frequency
2. $x = \tau f_k$
3. For trapezoid,
   $\dfrac{\Delta}{\tau} = 0.1$ assumed.
4. $$B_e = \frac{\int |v(f)|^2 df}{|v(0)|^2}$$
5. $$B_m^2 = \frac{\int f^2 |v(f)|^2 df}{\int |v(f)|^2 df}$$

**Formulas given here are used in subsequent rows of table.

This flexibility carries with it the requirement for more extensive measurement procedures (Ref. 42) to define spurious emissions that can be present, since such emissions can vary considerably with respect to space, time, and frequency — depending on the particular operation modes employed.

5-9.2.3.1 Harmonics

Estimates of the relative magnitudes of harmonics of the carrier frequency that can be expected range from −25 to −30 dB below the fundamental for the 2nd and 3rd harmonics to −70 to −100 dB for the first half dozen or so harmonics. Typical measurements on a number of tubes recently gave results as shown in Table 5-14. The wide variation depends on the effective filtering performed in the transmission path between transmitting tube and antenna, which can be designed to control this effect.

5-9.2.3.2 Adjacent Band Spurious Noise

Problems may arise with CFA's or TWT's due to adjacent band spurious oscillation modes. In fact, significant spurious emissions of this type have been measured from most of the high powered tubes used as radar transmitters. Typical results are shown in Fig. 5-100, (Ref. 34). It is found that this type of emission changes in an anomalous manner as tube adjustments are made, compounding the problem of interference minimization in any specific situation.

TABLE 5-13

FREQUENCY BANDWIDTH CHARACTERISTICS FOR REPRESENTATIVE WAVEFORMS NORMALIZFD TO 3dB BANDWIDTHS

Ratio	Waveform			
	(a)	(b)	(c)	(d)
B_e/B_3	1.13	1.10	1.07	1.04
B_m/B_3	∞	1.62	0.85	0.80
B_6/B_3	1.36	1.36	1.41	1.39
B_{40}/B_3	71.0	22.9	3.65	4.55
B_{60}/B_3	720.0	72.3	4.47	9.55
B_{80}/B_3	7200.0	229.0	5.17	20.5
B_{100}/B_3	72000.0	723.0	5.78	44.0

5-9.2.3.3 Transmitter Stability Considerations

High-power pulsed oscillator tubes have a carrier frequency of emission that is highly dependent on control electrode voltages. A small ripple in power supply voltage can cause significant changes in oscillator frequency. Changes in frequency on the order of 1 part in 10^4 can be experienced in magnetron oscillators due to a 1% change in high-voltage power supply voltage. Thus, at 10 GHz (X-band) frequency, changes up to 1 MHz can be expected from nominal. This problem is avoided in most sophisticated systems that require pulse-to-pulse coherence for proper operation (such as MTI systems) (Ref. 38, pp. 5-12 to 5-19) by using a separate coherent crystal-controlled oscillator with an amplifier at the output of the system.

5-9.3 RADAR SUSCEPTIBILITY

5-9.3.1 Effects of Interfacing Signals

The front end of a radar equipment is usually a highly sensitive receiver connected to a highly directional antenna for reception of echoes of distant targets. Due to the wide dynamic range of possible desired echo returns, the system can be sensitive to a

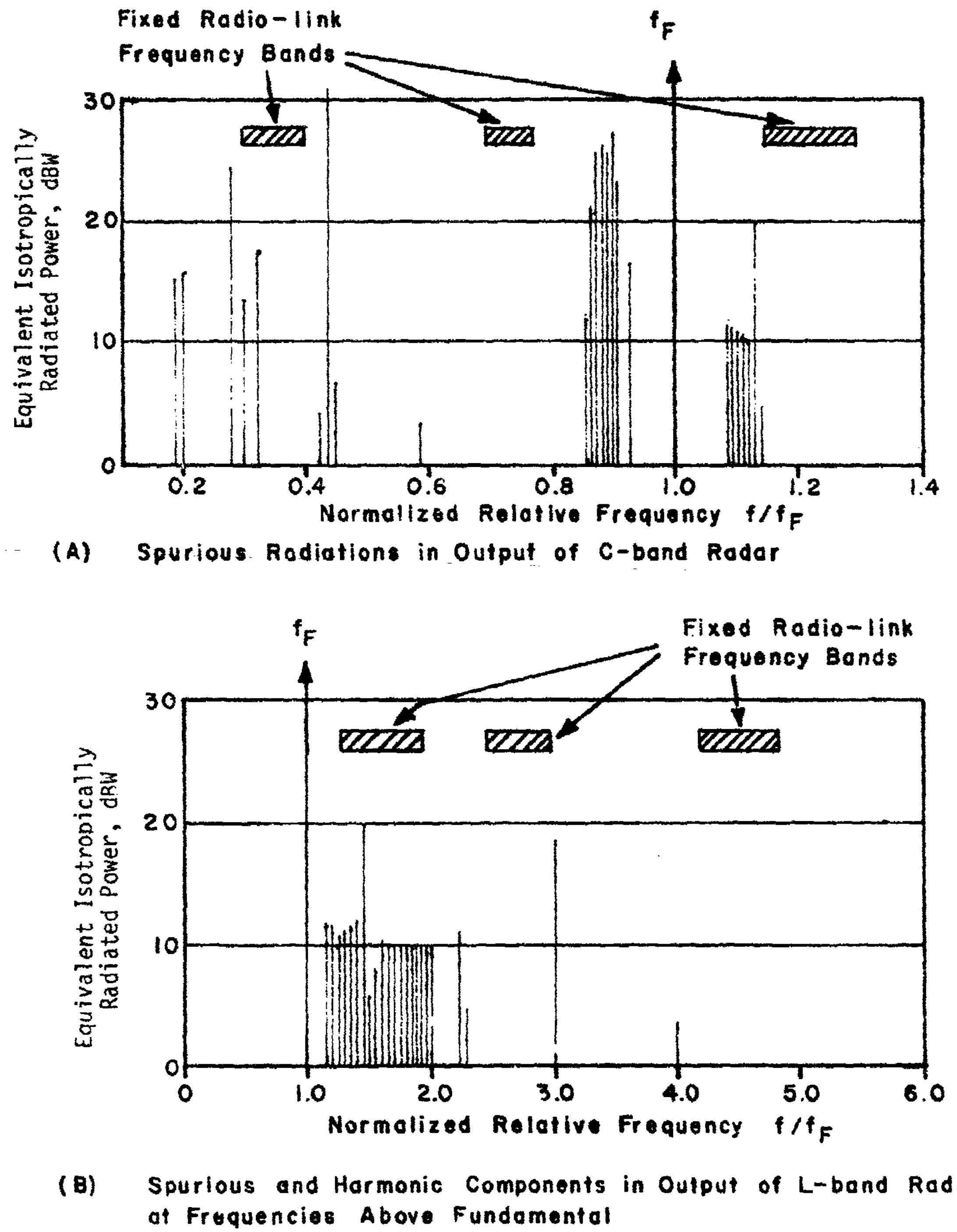
TABLE 5-14

SUMMARY OF RADAR HARMONIC LEVELS* (DECIBELS BELOW FUNDAMENTAL)

Harmonic / Tube type	2	3	4	5	6	7	8	9	10
Magnetron:									
Mean	78.1	71.7	77.1	86.0	87.2	91.9	99.3		
Range, high/low	57/103	45/100	62/93	67/114	76/96	81/96	83/114		
Number of samples	77	59	34	23	17	7	5		
Klystron:									
Mean	71.3	78.2	76.9	73.9	82.3	87.2			
Range, high/low	38/119	57/105	56/101	59/111	73/89	72/97			
Number of samples	44	27	21	8	7	4			
Tetrode and triode:									
Mean	83.2	76.0	99.6	90.0	96.7	100.2	106.2	97.2	100
Range, high/low	74/93	72/81	93/108	79/98	83/108	93/112	98/113	93/100	100/100
Number of samples	14	13	10	11	9	9	3	4	2

*Systems with waveguide transmission lines only.

(Courtesy of Department of Defense, Electromagnetic Compatibility Analysis Center)



(f_F = fundamental frequency)

Figure 5-100. Measured Characteristics of Typical Radars

wide variety of other signal sources. The effects of such undesired signals can be listed as follows:

a. Equipment degradation. This includes burnout of crystal detectors (Refs. 38 and 43) by strong enough pulses that enter the system at a point prior to sufficiently frequency-selective circuits that would attenuate them. Other effects such as overloading of amplifiers causing destruction of active elements may also be present.

b. Desensitization. The interfering signals that normally would be filtered out by receiving filters in later stages may be strong enough to drive leading stages into nonlinear regions of decreased gain so as to prevent full amplification of desired echoes.

c. Display reduction. In an oscilloscope display of echoes (e.g., A-scope, B-scope, or PPI), desired echo indications may be masked by an interference pattern.

d. Erroneous responses. Especially in automatic detection and tracking radars, interference may cause:

(1) False alarms—false target indications that can cause incorrect decisions and overloading of signal processing capacity (Refs. 44 and 45)

(2) Missed targets—failure to detect targets when required

(3) Erroneous target position estimates (e.g., angle estimates provided by a monopulse tracker).

5-9.3.2 Sources of Interfering Signals

The effects mentioned in par. 5-9.3.1 on system performance can be caused by a wide range of possible signal sources. These are listed in the paragraphs that follow together with a discussion of their main effects.

5-9.3.2.1 The Radar's Own Transmitter

For most pulse radar equipments, the receiver and transmitter share the same antenna. The decoupling necessary is achieved partially by the time separation of echoes from the transmitted pulse. However, protection of receiver circuits from the high power signals radiated calls for special circuits, known as duplexers (Refs. 38 and 46). Isolation of the order of 60 to 80 dB is required. Devices in common use include (a) gas-discharge microwave switches called transmit-receive (TR), and antitransmit-receive (ATR) tubes in various waveguide configurations, generally in conjunction with quadrature hybrid junctions to provide required broadband directional characteristics; (b) ferrite circulators of the Faraday rotational, differential-phase-shift, or junction types; and (c) diode devices such as varactor limiters or differential phase shifters. For higher power applications the gas discharge or ferrite-circulator duplexers are preferred.

A related type of interference is that caused by intraradar coupling of the pulse modulation waveform used. This involves high voltages and/or currents such as those required to drive a magnetron. Electric and/or magnetic fields generated or resultant pulses conducted via ground wires or power leads may be sufficient to cause deterioration of sensitive receiver components. Careful design procedures are required to avoid this danger.

5-9.3.2.2 Undesired Echoes

Echoes of undesired targets constitute passive types of interference. The following list includes the more significant cases; usually the resulting echo patterns are readily identifiable so as to distinguish them from other types of interference:

a. Ground or sea clutter necessitates the use of MTI (moving target indication) circuits when concerned with detection of aircraft, vehicles, or personnel at low elevation angles.

b. Clouds, rain, or snow produce echoes, an effect which becomes more significant at the higher microwave frequencies.

c. "Angels" or reflections from birds, insects, or dust may be significant (Ref. 47).

d. Multipath effects become important for radars at significant heights; the ground-bounce-path causing destructive interference with the direct-path echo, leading to difficulty in properly locating targets at low elevation angles.

e. Chaff or clouds of metallic strips may be emitted by an adversary. These usually will be effective over a given band of frequencies and will lead to false detections until enough temporal, velocity, and spatial separation exists for desired targets.

5-9.3.2.3 Environmental Fields

The radar receiver must operate in the real-world environment, subject to fields produced by ignition of internal combustion engines, welding machines, etc. Such fields might act directly on the receiver circuits, or enter through the power lines as well as possibly being picked up by the antenna and RF circuits. The effect of such interference on radar performance and displays can be quite severe. Techniques for control of such phenomena by shielding, use of filters, and proper grounding are described in other areas of this handbook. The effect of such fields on a given radar system would be highly dependent on the site selected for the radar as well as the method of supplying power; power line filters may be required.

The remaining sources of EMI to be discussed here are active signal sources generating waves picked up by the antenna of the radar.

5-9.3.2.4 Natural Sources

The sun and stars emit noise-like energy that can interfere with the detection of weak targets, see par. 3-1.2.3. The effect is to increase the effective system noise temperature by amounts varying from a few to several tens of decibels, depending on the antenna gain and pointing direction with respect to the sources. These effects can be predicted reasonably well, including the effects of solar storms (sun spots) in various parts of the RF spectrum.

5-9.3.2.5 Communication and Navigation Signals

These signals usually are confined to frequencies far removed from the radar bands, and can be considered of only marginal importance. An exception is over-the-horizon, HF radar systems (in the frequency band 2-30 MHz) which must have specially designed operating procedures to select a "quiet band" for operations as well as to cope with daily and yearly variation cycles of the ionospheric propagation medium.

5-9.3.2.6 Other Radars

These are perhaps the most important source of interference to a radar system since they can be expected to be present at frequency bands near the given radar operating frequency. Pertinent characteristics of radar emitters have been mentioned in par. 5-9.2, both inband and out-of-band. It should be pointed out here that there will be a complex time dependence of the interference caused by a radar due to the usual low-duty-cycle pulsed waveform as well as the directional variations of a radar pointing direction (e.g., rapid scanning of a search radar or dynamic tracking of a moving body) (Refs. 38 and 48).

5-9.3.2.7 Electronic Countermeasures (ECM)

Electronic countermeasures involve deliberate emissions by an adversary of noise-like signals (barrage jammers) or pulses synchronized with the transmitted pulses ("spoofing") or other signals designed to confuse. See Engineering Design Handbooks AMCP 706-411 through -416 for a detailed treatment of EMC.

5-9.4 INTERFERENCE CONTROL

Techniques exist for controlling the most significant interfering effects to enhance operation of radar equipments in the presence of interference and also to minimize the effects of radar emissions on other equipments. The interference control techniques to be described are concerned with the design of system components as well as the techniques employed for shaping and processing the signals.

5-9.4.1 Directional Selectivity

The objective is to confine both the emitted signal and the receiver sensitivity to a preselected target direction. This objective is complicated by the fact that the pointing direction changes as the radar scans over a sector to search for targets, or tracks a moving target.

The determining factor is the angular radiation (or sensitivity) pattern of the antenna being used. Usually, by reciprocity, the patterns (gain versus angles) are similar if not identical when the antenna is used for radiation or reception of microwave energy. Although the required gain determines the width of the main lobe of an antenna, certain steps can be taken to control specific side lobes and back lobes as discussed in par. 5-10.6.3, which describe techniques to suppress radiation through cancellation.

5-9.4.2 Time Selectivity

Time is important at several levels with respect to the operation of a radar system. Significant from the interference point of view are the scan-time required by a search radar to cover a given angular sector and the time between pulses during which echoes of a transmitted pulse may be received.

Concerning scan-time, it may be important to synchronize the scanning of nearby searching radar systems so that main-beam energy will not interfere directly when either radar points at the other.

Proper use of the time relations from pulse-to-pulse are of greater control importance. While target echoes will occur approximately equal times from successive transmitted pulses, interfering pulses in general will occur at differing times in the pulse recurrence frequency (PRF) interval. Suitable gating associated with pulse-to-pulse integration for signal enhancement can (Ref. 38) help suppress pulses at a different repetition rate from the radar; use of staggered (unequal) times between pulses also can help this technique. For radars in the same installation, it may be useful to synchronize the PRF's so that interfering pulses occur too early in the pulse-to-pulse interval to affect echoes. Another technique is to provide a receiver blanking signal to prevent radiated pulses from producing false alarm indications in a neighboring radar receiver.

An important technique used extensively is sensitivity time control (STC). This is based on the fact that the energy in an echo from a given size target will vary inversely as the fourth power of the range. STC attempts, by introducing a time variable attenuation in the radar receiver synchronized with the pulse repetition period so that attenuation is maximum for close targets (minimum delay) and is zero for targets at maximum range (maximum delay), to equalize the displayed echo amplitude at all ranges. This reduces the effect of clutter echoes on displays and tends to reduce active interference sensitivity at short observation ranges.

5-9.4.3 Frequency Selectivity

The concept here is to use knowledge of the location of the interfering signal in the RF spectrum in the operation of a radar. By careful control of the operating frequency, a suitable location in the tunable band may be found to minimize interfering effects. Rapid accurate tunability of transmitter and receiver circuits is important here. It is important to design the receiver IF circuits to match the transmitter spectrum so that adjacent frequency energy will be rejected as much as possible. Provision of RF circuit selectivity is important for rejection of signals at the image frequency.

A useful technique is that of frequency agility wherein different carrier frequencies are used on successive pulses. Coherent combination of successive echoes received by changing the receiver tuning, correspondingly, will tend to enhance the target echo-to-interference ratio. This method is particularly useful for minimizing ground clutter effects.

Another technique designed primarily for elimination of clutter echoes is MTI (Ref. 38, Chapter 17). This is a form of pulse-to-pulse processing that desensitizes the receiver for echoes which remain the same so that only echoes having a Doppler frequency shift with resultant carrier phase variations from pulse-to-pulse are accepted. This technique is not too helpful against active interference sources.

However, the pulse-Doppler technique for moving target detection can have important benefits from the point of view of interference suppression. Here, a burst of closely spaced pulses is transmitted; processing consists of passing range-gated echoes into a bank of relatively narrow-band filters. Only targets with well-defined Doppler frequency shifts will give significant outputs from appropriate filters. Interfering energy that does not have a well-defined pulse-to-pulse dependence will tend to be suppressed.

5-9.4.4 Amplitude Selectivity

Radar receivers are inherently quite sensitive to wide ranges of interfering signals. Even though directional, time, and frequency selectivity can be used to reduce interfering signal effects significantly, a dominant consideration must be the fact that the attenuation of interfering signals with distance varies only as $(1/r)^2$ (r = distance to interfering source), whereas desired target attenuation varies as $(1/r)^4$. Hence, interfering radar pulses (as well as jamming signals) can be many decibels larger than desired echo signals. Use of logarithmic amplifiers and limiters can help to reduce transient effects of large signals. Techniques such as Instantaneous Automatic Gain Control (IAGC) and Detector Balanced Bias (DBB) (Ref. 49) behave the same way, reducing saturation difficulties and permitting effective recognition and processing of target echoes when they occur. The Lamb Noise Silencing circuit is useful prior to the normal bandwidth IF filters. Its limiter at this point serves to decrease significantly the energy of a strong off-frequency pulse so that spectral sidelobe energy in the passband of the receivers will be decreased correspondingly.

5-9.4.5 Waveform Selectivity

The concern here is regarding techniques for designing waveform characteristics, together with corresponding processor characteristics, to enhance response to desired target echoes and simultaneously discriminate against interfering signals. This need is closely related to the objective of obtaining high resolution capability, i.e., the ability of the radar to distinguish echoes from targets that are closely spaced in time (range) or in frequency (range rate). Such high-resolution capability can be quite helpful in minimizing system effects of interference.

On recognition that radar waveforms are of necessity wide-band in the frequency domain, steps should be taken to limit the spectral utilization to no more than required for a specified resolution. Methods of shaping a transmitted pulse (see par. 5-9.2.3) are useful in this regard. The general principle of minimizing waveform time discontinuities is important to achieve corresponding minimization of spectral spread. For example, the use of a squared cosine or Gaussian-like pulse envelope can give significantly less spread to adjacent frequency bands than a simple rectangular or even a trapezoidal waveform envelope. It should be recognized, however, that such techniques are most important for the simplest radar waveforms where the significant energy bandwidth is roughly the reciprocal of the pulse duration. A variety of techniques exist for making a radar receiver output sensitive to specific features of the desired target echo waveforms and discriminate against other dissimilar interfering (pulse) waveforms. For pulse radars such techniques can be classified as those concerned with the interpulse periods and those concerned with the pulse waveform itself.

Simple pulse-to-pulse integration, either post-detection (incoherently) or coherently (e.g., by delay-line processing of IF data), will enhance desired target echoes at the expense of interfering pulses occurring at a different repetition rate. Combining such integration with pulse-to-pulse frequency agility as previously mentioned is a related technique with the same objective. Random or pseudo-random variation of interpulse period also can be useful in this regard.

Pulse-length selection is a simple technique of processing received pulses to determine their duration and then rejecting all pulses whose duration is not close enough to that of the desired target echoes (as determined by the own-radar transmitted waveform). A drawback of this technique is its dependence on an IF bandwidth somewhat larger than the reciprocal of the pulse duration so as to permit the required estimate to be made accurately, thus permitting more noise into the system and limiting the maximum range achievable for weak targets. Application of this technique should be limited, in any event, to pulses greater than some energy threshold if possible.

A significant degree of interference immunity is provided in the processing implicit in the use of high resolution pulse-compression (Ref. 38, Chapter 2) waveforms. To this end, waveforms with high time-bandwidth products are used, such as FM pulses ("chirp") or pseudo-random digitally coded pulse trains. Such waveforms are necessary to increase total energy per pulse for peak-power limited output tubes (to obtain desired long-distance range performance) by increasing pulse duration without decreasing range resolution. By use of specific IF or coherent video-matched filters, an output pulse can be obtained, the duration of which is proportional to the reciprocal bandwidth rather than the duration of the desired echo waveform. Interfering pulses will be greatly attenuated unless their behavior is a close replica of the own-radar waveform.

Techniques have been extensively developed for generation and matched filter detection of chirp pulses, involving oscillators with voltage-controlled frequency as well as dispersive delay lines (with delay a specified, usually linear, function of frequency). Pseudo-random digital coding of pulse waveforms involving phase-modulation of a large number of "subpulse" segments is achievable by use of digital circuits including solid-state components such as shift registers. Matched filters using compact surface acoustic wave technology are now available and are incorporated into systems. Use of digital logic components is also important since it provides the capability of sensitizing the receiver to specific waveform features, rejecting interfering waveforms which differ in modulation structure even though the same frequency band is occupied.

5-10 ANTENNA INTERACTION CONTROL

5-10.1 INTRODUCTION

Antennas are designed to meet certain performance specifications with respect to gain (or directivity), bandwidth, and physical characteristics (so as to be compatible with their locations). These characteristics are of direct concern also in evaluating EMC situations. However, in addition, the EMC engineer is concerned with characteristics of an antenna outside of its design operating band. Knowledge of such characteristics is limited. Most is known about the simple dipole which can be treated fairly well analytically. In the more complex antennas, especially those of the aperture type, the characteristics must be determined experimentally.

Interactions through antenna coupling between equipments operating at different design frequencies occur as follows:

1. Intermodulation can take place in a transmitter output stage because of energy coupled to it from a nearby transmitter.

2. A receiver can respond to a strong off-frequency emission from a nearby transmitter.

3. A receiver can respond to a spurious output of a nearby transmitter or receiver (local oscillator radiation) at the frequency to which the receiver is tuned.

5-10.2 INTERACTION COMPUTATIONS

In estimating the interaction of two antennas one can begin with the formula for the ratio of power received P_r to the power transmitted P_t (see Eq. 3-149).

$$\frac{P_r}{P_t} = \left[\frac{\lambda^2}{(4\pi r)^2}\right] G_r G_t = \frac{A_r A_t}{\lambda^2 r^2} \qquad (5\text{-}38)$$

where

r = separation distance, m
λ = wave length, m

$G_r G_t$ = respective power gains of the receiving and transmitting antennas with reference to an isotrope, dimensionless

$A_r A_t$ = corresponding effective areas, m^2

These formulas apply strictly only for free-space propagation conditions. Appropriate corrections should be made to account for reflecting objects in the path between antennas (see par. 3-3.2.4.1) and, if closely spaced, for near-field corrections (see par. 5-10.6.2).

The gains or effective areas must be known or estimated at the frequency of interaction under consideration and for the direction of the interconnecting path, which in general will not be that of the main lobe of either antenna.

5-10.3 ANTENNA PARAMETERS

In this paragraph, representative characteristics of various antennas are given, both at the design frequency and at frequencies off design to the extent that information is available. To define the performance of an antenna in free space at a given frequency, one must know its impedance at the feed point, its directive radiation pattern, and its polarization discrimination.

In the design band, the antenna performance is well defined. Polarization discrimination may be of the order of 30 dB or more. Antenna standing-wave ratios are usually less than two, so that mismatch loss is inconsequential. The antenna pattern usually has only one major lobe and the pattern description may be given by azimuth and elevation cuts or by specifying gain in the direction of transmission, half-power beamwidths, sidelobe levels, and front-to-back ratio. In more modern specifications, antenna noise temperature may be given.

Outside of the design band it has been very unusual to specify or control antenna performance and, as a consequence, variability can be expected. The patterns may have several major lobes, and the gain in the direction of maximum gain as well as in other directions of interest (along the horizon for instance) must be known. Polarization discrimination may be very small. Impedance variations are usually quite large and the resulting mismatch loss must be considered. Otherwise, out-of-band emission or susceptibility, especially at frequencies far from the design frequency of any particular device, can be controlled by the use of filters or otherwise frequency-selective circuits in the antenna feeder coupling circuits or in electronic circuits nearest the antenna.

5-10.4 PRESENTLY AVAILABLE PERFORMANCE DATA

In this paragraph data on parameters — such as gain, pattern, polarization, and impedance — important in antenna interaction control are given to the extent they are available.

5-10.4.1 Thin Dipoles

Thin dipoles are linear antennas having length-to-diameter ratios (l/d) greater than 10. Theoretical expressions for impedance and patterns as a function of frequency are given in the literature (Refs. 50, 51). Very little measured data are available. Characteristics are as follows:

a. Polarization Discrimination. The polarization discrimination of thin symmetrical dipoles is of the order of 30 dB or more and is independent of frequency. If the dipole is unsymmetrical, the discrimination is less than 30 dB, cannot usually be determined theoretically, and must be measured for each antenna.

b. Impedance:

Dipole resistance and reactance are functions of frequency and dipole length to diameter ratio (l/d). Fig. 5-101 modified from Polk's original data (Refs. 51 and 52) shows the VSWR (with reference to 50 ohms) of a typical dipole antenna for a frequency range of 5 to 1. For thinner dipoles (higher l/d), the peaks of the curve are sharper and occur at higher standing wave ratios. The data easily can be reduced to transmission loss for a specified transmission line (Ref. 53) and can be extended to frequency ranges greater than 5 to 1 by calculation. However, it should be noted that the impedance can be affected significantly by nearby ground planes or other conducting objects.

Fig. 5-102, redrawn from Refs. 51 and 53, show plots of thin dipole patterns as a function of dipole length to wavelength ratio (l/λ). The figures apply to center-fed dipoles as well as vertical dipoles above an infinite perfectly conducting ground. The pattern is shown for only one quadrant (it is symmetrical about the coordinate axis). The true pattern would be obtained by rotating the plot about the antenna axis. It can be seen that as frequency increases (increasing l/λ), the main lobe eventually breaks up, giving two main lobes ($l/\lambda = 0.875$ or 1) or two main lobes and a multiplicity of side lobes. The direction of maximum gain departs from the horizon ($\psi = 0$).

As the length to diameter ratio changes, the number and position of the lobes remains unchanged. However, the depth of the nulls (which are zero for

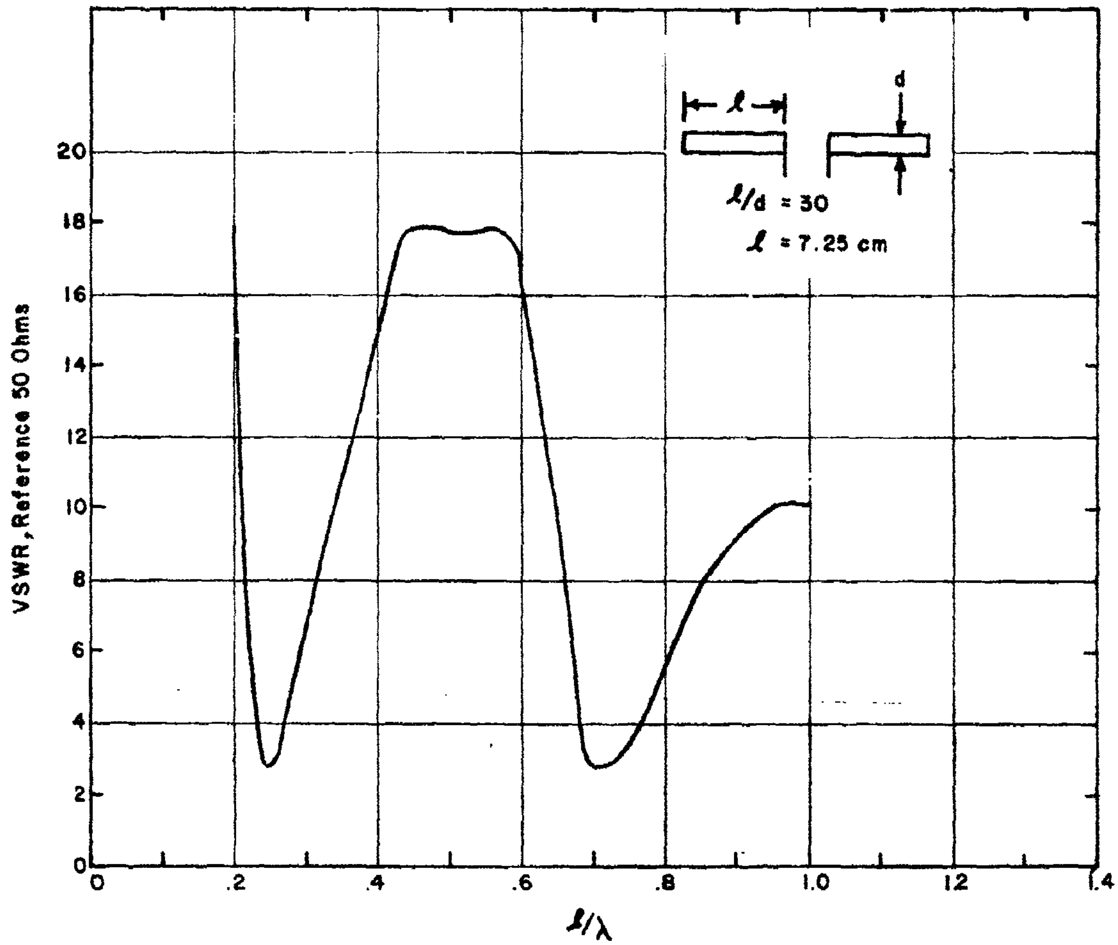


Figure 5-101. Impedance of Center-fed Dipole

infinitely thin dipoles) becomes less deep as l/d increases and the amplitude of the side lobes increases (Ref. 37). Measured and calculated patterns show excellent agreement.

If a vertical dipole is mounted above a ground plane having finite conductivity, the patterns change from those shown. The main lobe, in general, is tilted above the horizon. The tilting is only in the vertical plane, the horizontal plane patterns remaining circles if the ground plane is isotropic. In this case one must resort to measured patterns for each ground conductivity of interest.

c. Gain:

The gain of a thin vertical dipole antenna, calculated from the patterns of Fig. 5-102 is shown in Fig. 5-103. It is given in the direction of the horizon (ψ = 0) as well as in the direction of maximum gain. In the direction of the horizon, it will be noted that the gain oscillates between zero and 1.5 dB with decreasing amplitude as frequency increases (l/λ increases). The average gain is about 1 dB.

In the direction of maximum gain, the gain oscillates with increasing amplitude as frequency increases. The minima of this curve coincide approximately with the solid line curve given by Terman (Ref. 37). It will be noted that the average of the maximum gain will be higher than the value given by Terman.

The direction of maximum gain for the thin vertical dipole antenna is given by the dotted curve of

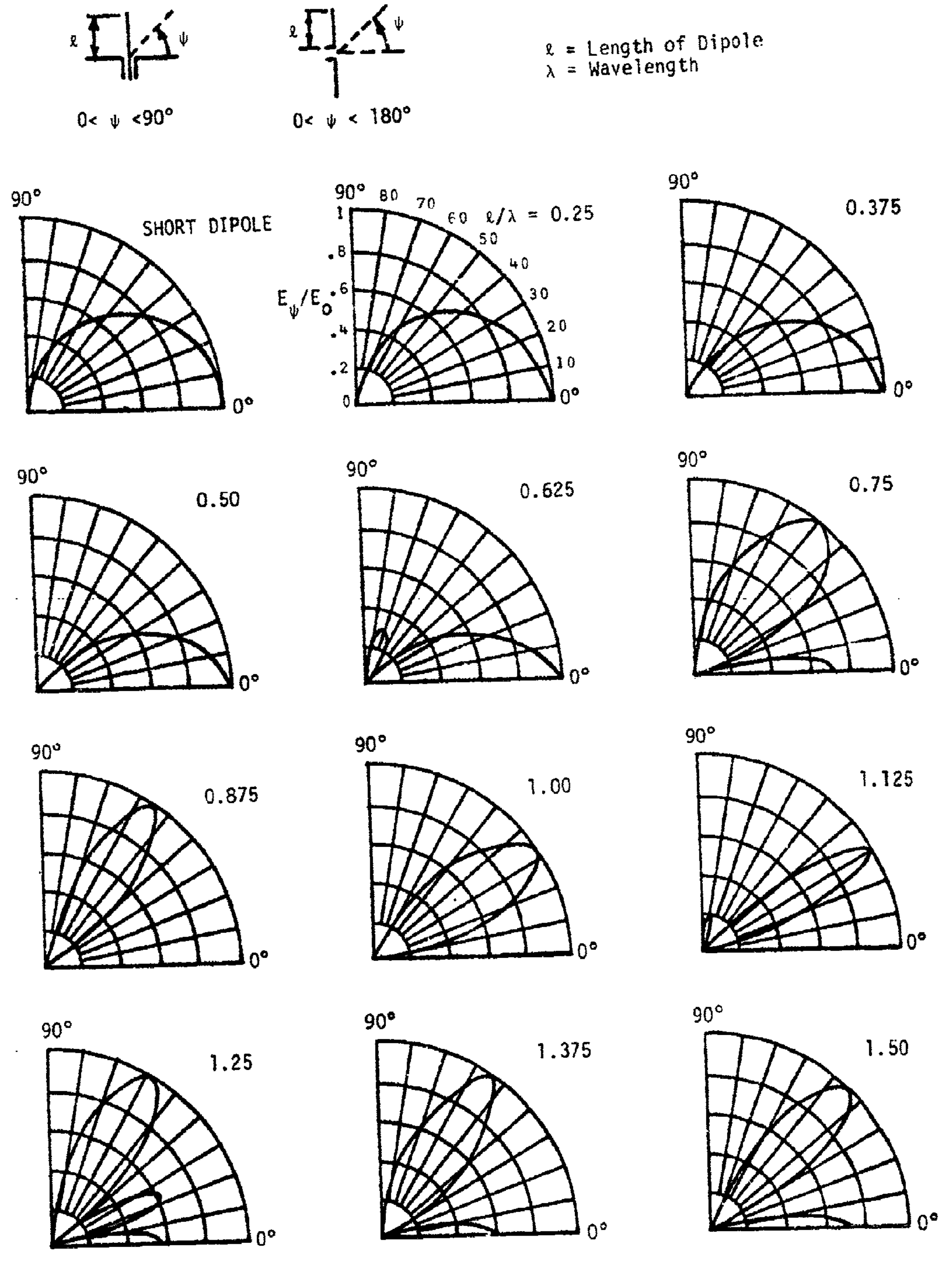


Figure 5-102(A) Far-field Radiation Patterns of Thin Dipoles (l/λ = 0 to 1.5)

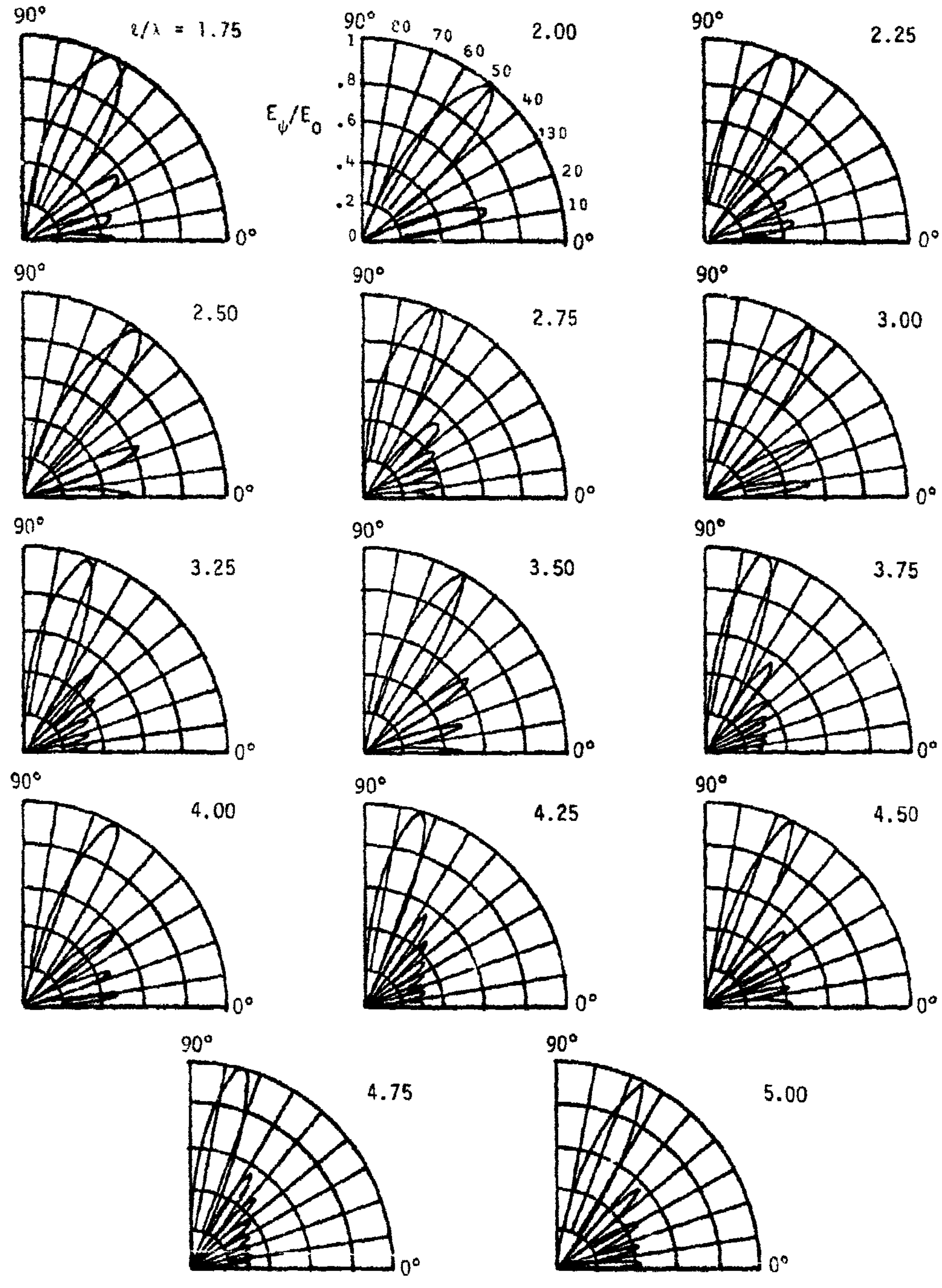


Figure 5-102(B) Far-field Radiation Patterns of Thin Dipoles (l/λ = 1.75 to 5)

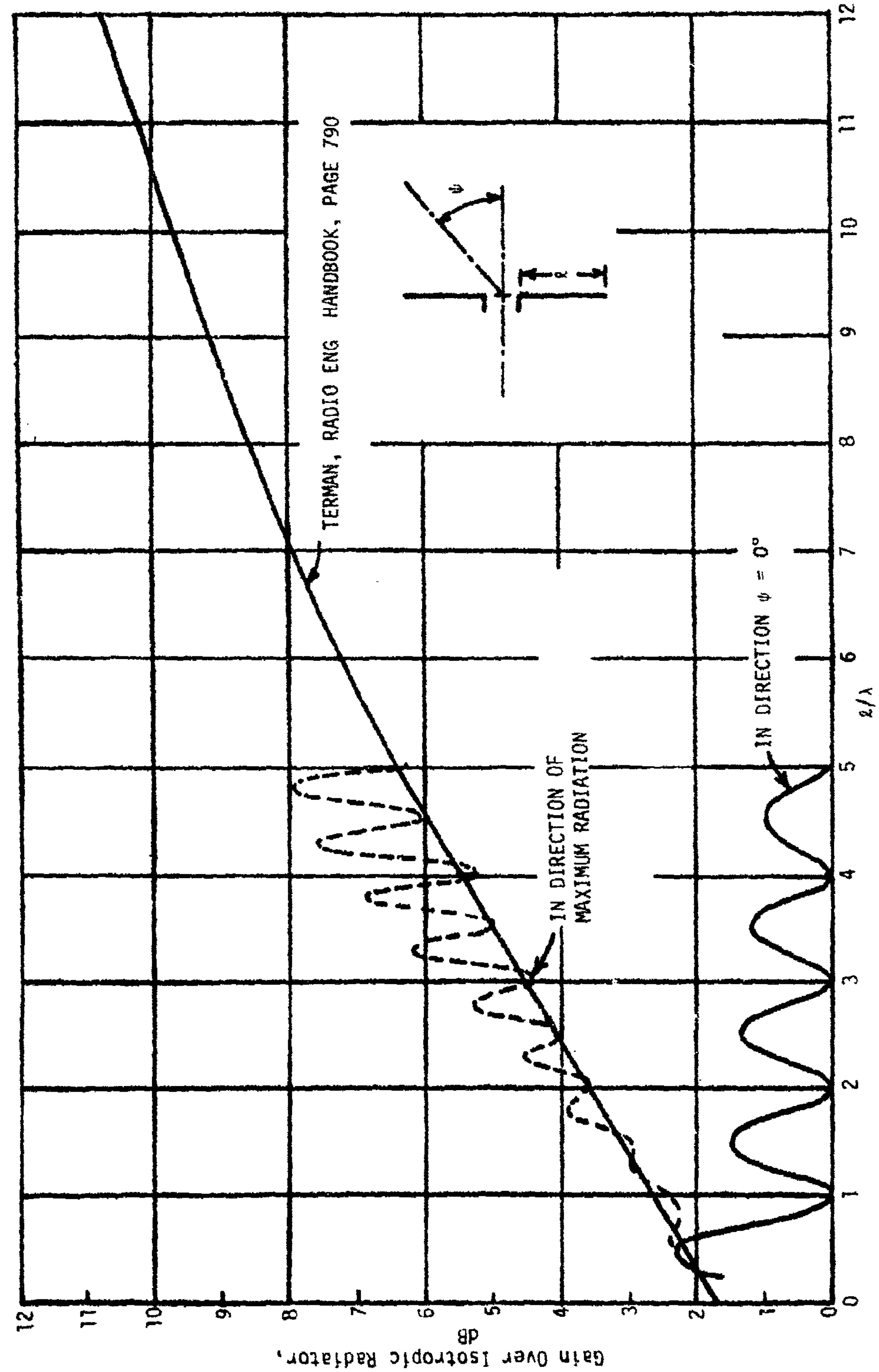


Figure 5-103. Gain of Center-fed Dipole

Fig. 5-104. Here it will be noted that as frequency increases, the direction of maximum gain approaches 70 deg.

5-10.4.2 Dipole Arrays

Polar directive amplitude diagrams for arrays of 2 and 16 identical antennas spaced at equal distances along a straight line with equal phase differences introduced between the currents in adjacent elements are available (Ref. 50). The patterns of these collections are for fundamental frequency operation only. Off-frequency data cannot be obtained readily from these collections.

Measured patterns, over a 5 to 1 frequency range, for collinear cylindrical antennas, and V dipoles for various length to diameter ratios are given by Reich (Ref. 54). These data have not been reduced to gain and direction of maximum gain. Consequently, data on these antennas outside of their design band must be obtained by measurement.

5-10.4.3 Disc-Cone Antennas

The disc-cone antenna may be used as a balanced radiator (biconical antenna), as a single-ended radiator above a semi-infinite ground plane (disc-cone) or a hemispherical ground plane, or it may be used as a biconical V antenna. The radiator itself may have a conical cap, a hemispherical cap, or be open-ended. The half angle of the cone may vary from 0 deg (thin dipole) to 47 deg with most values between 30 and 47 deg. The rod disc-cone antenna is an approximation of the solid disc-cone antenna. In-band and out-of-band performance of the biconical horn antenna are summarized (Refs. 54 and 55):

a. Polarization. No information is found in the literature on the polarization discrimination of the biconical antenna as a function of frequency. This probably will require experimental determination because of complications due to the feed structure.

b. Gain. Since the directivity patterns of disc-cone antennas are somewhat similar to those of the thin dipole, frequency increases the direction of maximum gain changes. This is shown by the solid curve on Fig. 5-104. The gain of the antenna in the direction of maximum gain and along the horizon is shown in Fig. 5-105 for 47-deg solid cone over a hemispherical ground. The maximum gain tends toward 3.5 dB at angles between 50 deg and 60 deg while the gain along the horizon falls at first to a minimum of 0.5 dB for $l/\lambda = 0.5$ and then increases approximately linearly with frequency. Similar gain curves for the biconical antenna have not been obtained as yet.

c. Patterns:

Directivity patterns for the biconical antenna, conical antenna above a semi-infinite ground plane, and conical antenna above a hemispherical ground plane are shown in Fig. 5-106. These are measured patterns. Agreement between measured and computed patterns is relatively good.

Measured directivity patterns for the biconical V antenna, over the same frequency range of about 5 to 1, are given by Reich (Ref. 54).

The patterns for the 30-deg solid disc-cone antenna, given in Refs. 52 and 54, are not too different from those for the 47-deg solid disc-cone given in Fig. 5-105. The 30-deg solid disc-cone patterns may be expected to apply to the rod-disc-cone type of antenna, such as the AT-197/GR, up to a length to wavelength ratio of about 2.

d. Impedance:

The VSWR of the 30-deg rod-disc-cone, the solid 47-deg cone above finite ground plane, and 47-deg cone above hemispherical ground plane are given in Fig. 5-107. Excellent agreement (Ref. 55) has been obtained between theoretically computed and measured impedance for the solid disc-cone antenna over the frequency range.

The agreement between the measured impedance of the rod-disc-cone and the solid cone is fairly good up to a frequency of about 1 GHz. Beyond that the VSWR rises rapidly and erratically. In this latter region, the spacing between the tips of the rods becomes comparable with the wavelength and one should not expect good agreement.

5-10.4.4 Exponential Horns

The horn antenna tends to have a uniform gain over a broad frequency range. Fig. 5-108 shows the maximum insertion gain (including mismatch loss) over a frequency range of 10 to 1 for both design polarization (vertical) and cross-polarization. Between 0.91 and 1.82 GHz — the cutoff frequencies for TE_{10}, TE_{01} and TE_{20} modes — the gain increases linearly with frequency, as one would expect for an aperture whose illumination remains constant. Below 0.91 GHz, the gain drops rapidly to zero because of waveguide cutoff of the dominant propagating mode. The gain is seen to be relatively constant over a frequency range of 7 or 8 to 1. Other characteristics are:

a. Polarization. The gain of the cross-polarized component (horizontal) is about 35 dB below design polarization in the design band. Outside of the design band, the polarization rejection becomes progressively less. Beyond a frequency range of 7 to 1, the

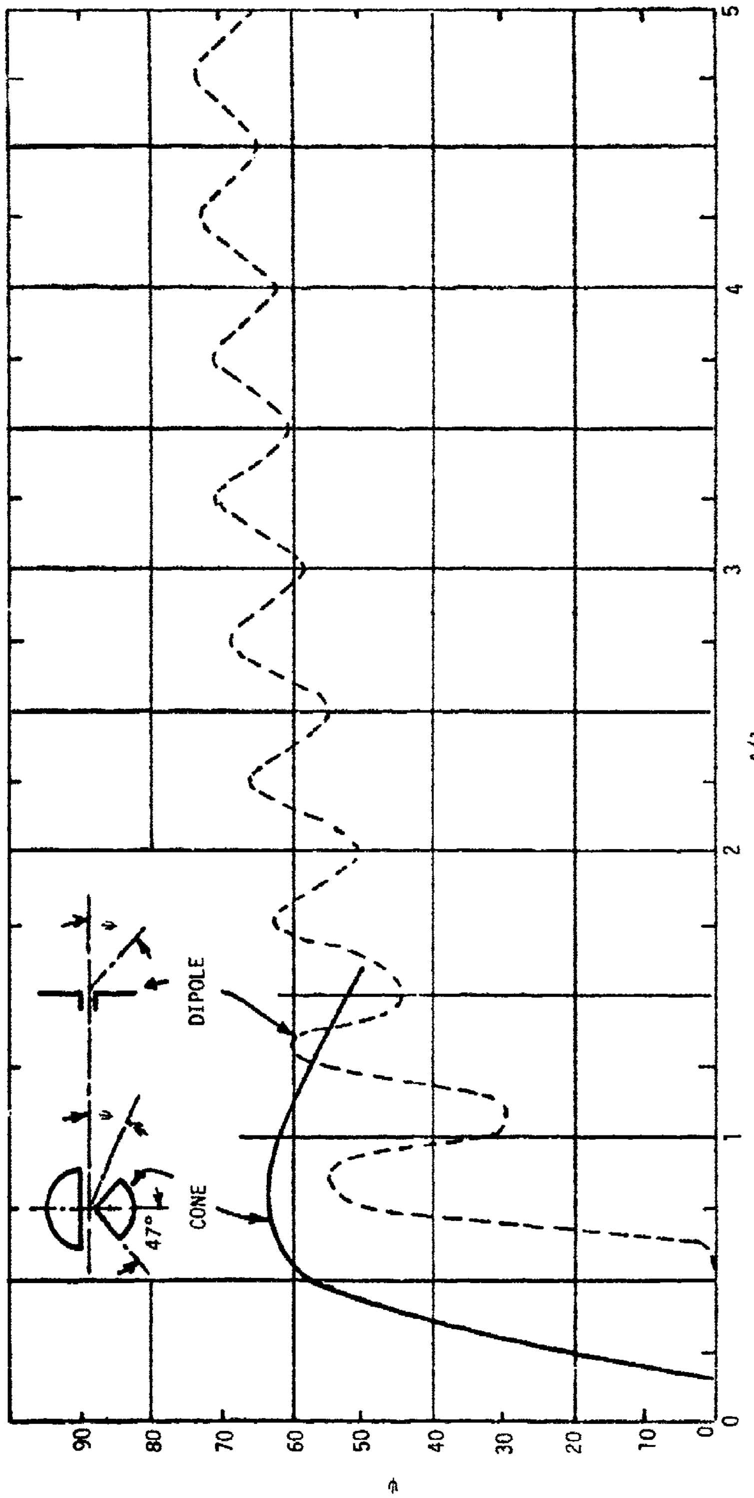


Figure 5-104. Direction of Maximum Radiation Intensity for Center-fed Dipole and for Solid Cone Over Hemispherical Ground

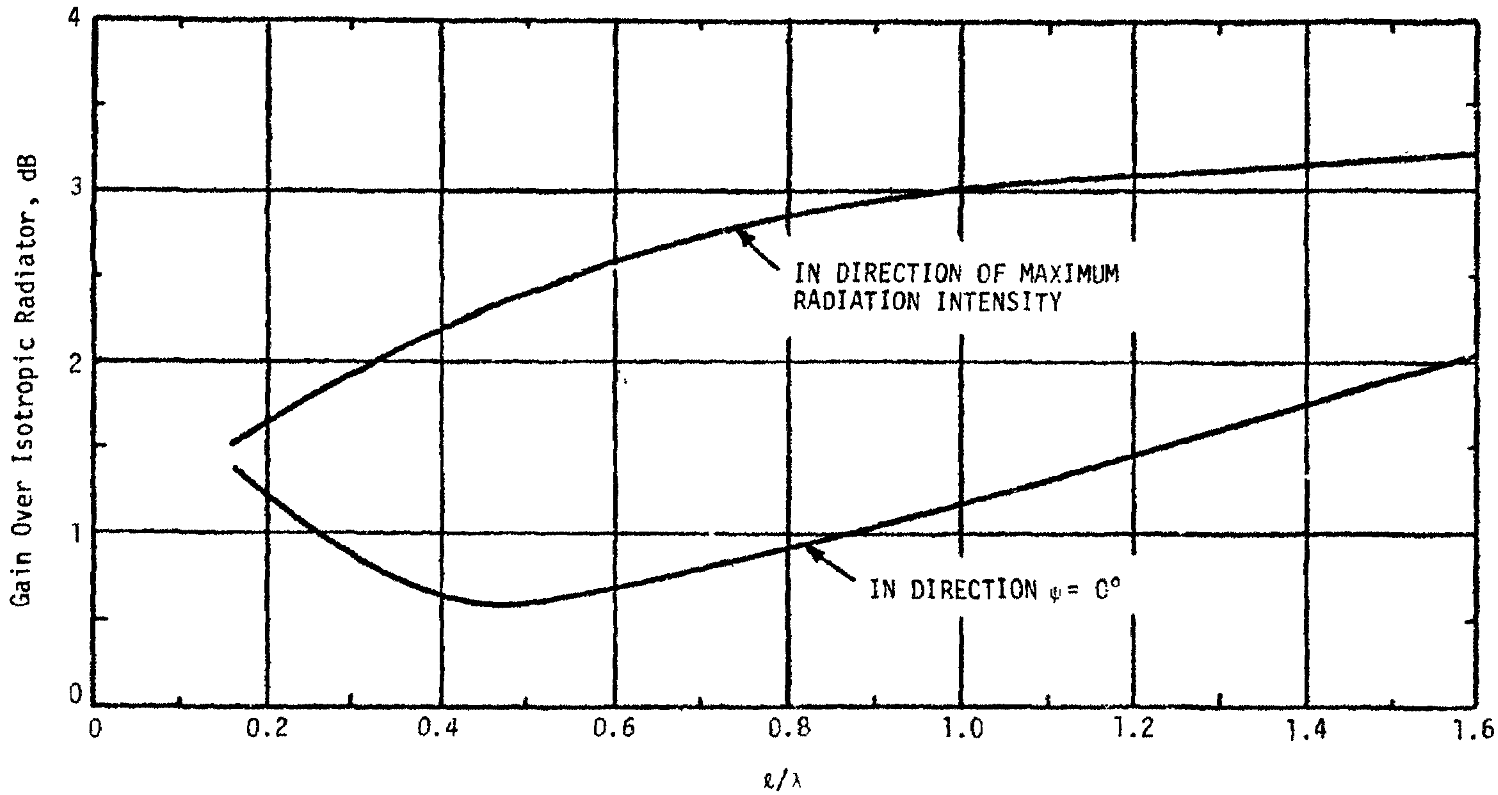


Figure 5-105. Gain of 47-deg Solid Cone Above a Hemispherical Ground

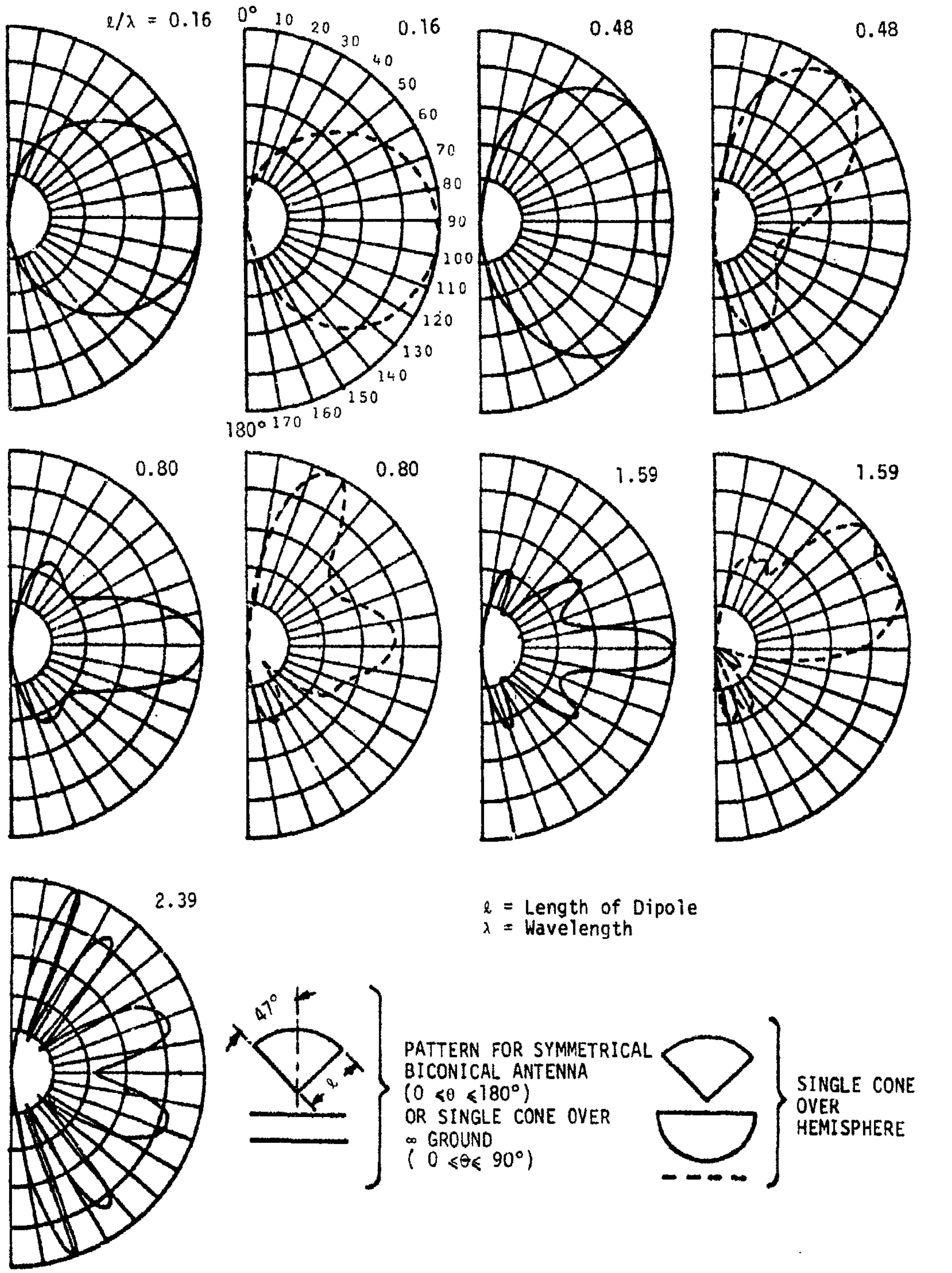


Figure 5-106. Far-field Radiation Patterns of Conical Antennas Above Infinite and Hemispherical Ground

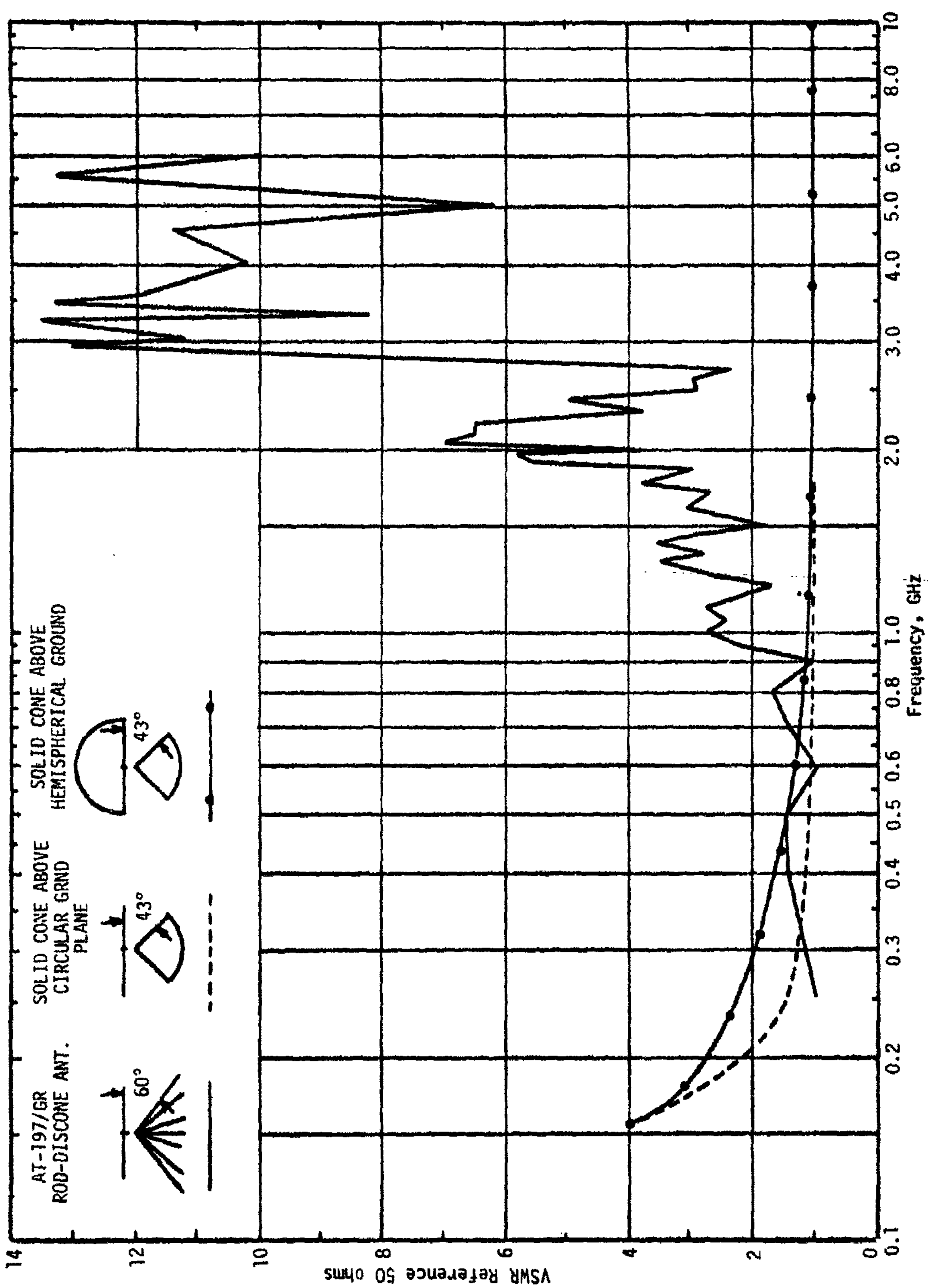


Figure 5-107. VSWR of Disc-cone Antennas

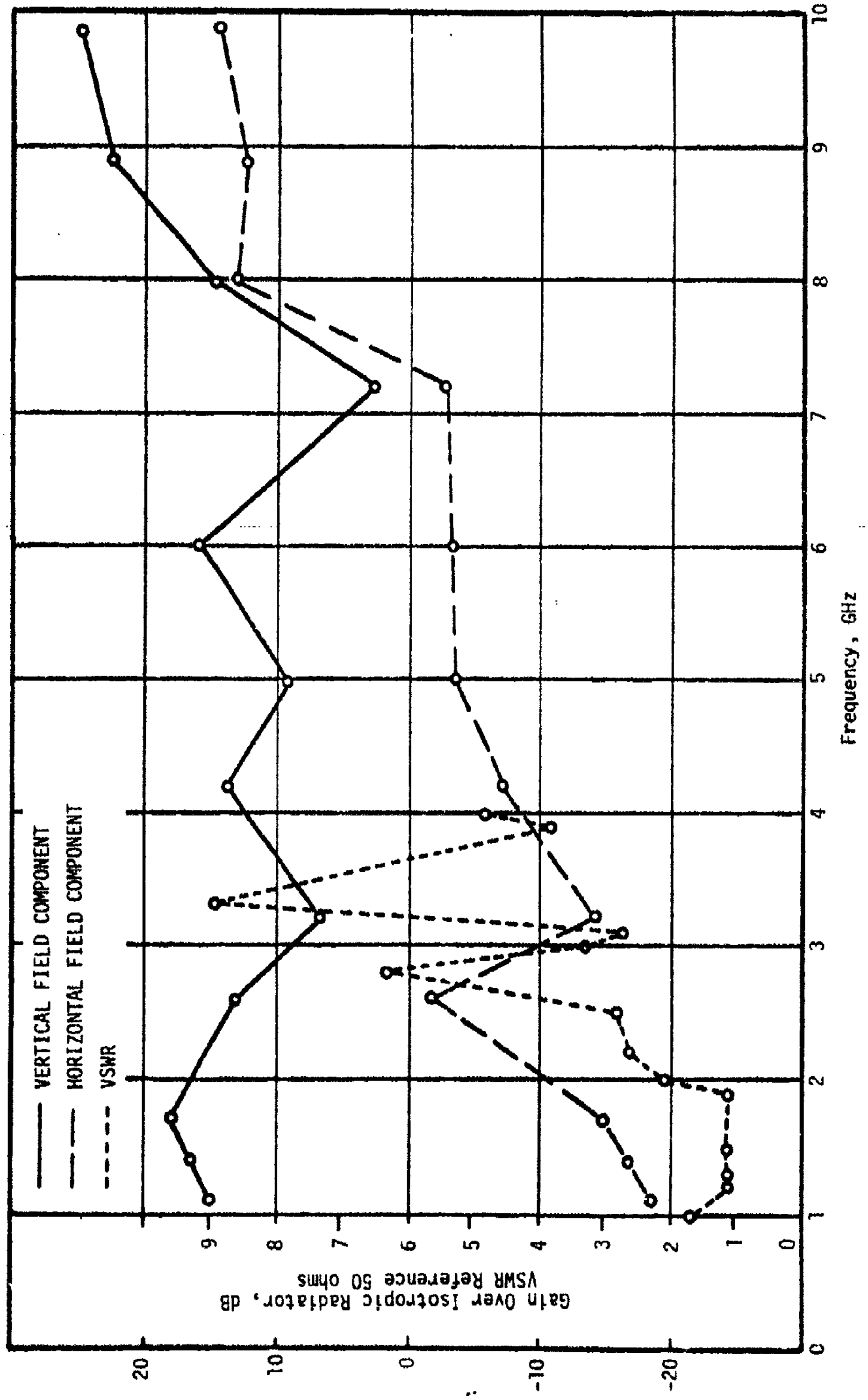


Figure 5-108. Gains nd VSWR of Horn Antenna With Transducer

desired and cross-polarized components have essentially the same gain. This is probably due to the modes that are excited by the coaxial to waveguide transducer which is used with the horn.

b. Impedance. The dotted curve on Fig. 5-108 shows the measured variation of standing wave ratio for the horn and its waveguide to coaxial transducer. No attempt was made to measure VSWR beyond 4 GHz because of the extremely large and rapid variations of frequency. It will be noticed, however, that the VSWR is less than 2 and relatively constant over the frequency range of 1 to 2 GHz.

c. Patterns:

Typical measured far field patterns are shown in Fig. 5-109(A) for vertical polarization at 1, 2, and 4 GHz. Note that they get narrower as frequency increases. For the third harmonic, however, the pattern is broader than for the second and splits into three beams. The splitting is probably caused by the TE_{20} mode as shown. The measurement system noise level for these patterns was 55 dB below maximum gain.

Patterns for the horizontally polarized component are shown in Fig. 5-109(B). Since the patterns are normalized, the true gain must be obtained from Fig. 5-108. The dotted portions of the patterns are below the recording system noise level.

d. Mode Patterns:

Some of the pattern distortions become clear when one examines antenna patterns as a function of excitation mode. Fig. 5-110 from Ref. 56 illustrates three patterns for an "S" band exponential horn. The solid curves are experimental and theoretically calculated patterns for dominant mode transmission at design frequency. Considering the approximations involved, there is good agreement between these curves.

The dashed curves are for the same horn excited with dominant mode at the second harmonic of the design frequency. The agreement between computed and measured performance is not as good in this case. It will be noted, however, that the pattern does become narrower, indicating higher gain as one would expect for a constant aperture illumination.

The third set of curves, dotted in Fig. 5-110, are for the horn excited with a TE_{20} mode transducer at the second harmonic frequency. The transducer had a mode purity of the order of 30 to 35 dB. Note that the main lobe has split up into two identical lobes with a deep null along the antenna axis. The half-power beamwidth of the split beam is wider than that for the dominant mode second harmonic beam but narrower than dominant mode design beam.

5-10.4.5 Horn Parabola

The horn parabola antenna is an electromagnetic horn capped by a sector of a paraboloidal-shaped reflector. The paraboloidal reflector converts the spherical wave emanating from the feed horn to a plane or uniphase front at the antenna aperture, thus insuring maximum gain and minimum side lobe radiation. Fig. 5-111, from Ref. 57, illustrates a typical pattern at 3.74 GHz showing the excellent performance of the antenna. Side and back lobes are even lower at higher frequencies. Table 5-15, compiled from Ref. 58, gives the antenna performance over a 3 to 1 frequency range. It can be seen that the impedance is constant, and gain does increase with frequency.

5-10.5 DESIGN IMPROVEMENTS

5-10.5.1 Thin Dipoles and Disc-Cones

Because of the nature of these antennas, out-of-band performance cannot be controlled significantly by design. Hence, tuned circuits or other forms of filters — which take into account the dipole impedance characteristics — must be applied as necessary in the transmission system or circuits coupled to it.

5-10.5.2 Dipole Arrays and Aperture Antennas

At the design frequency, side and back lobes can be reduced by smoothly reducing the amplitude of the feed to the outermost dipoles of the array. Cosine squared tapers give lower side lobe levels than linear tapers. This treatment may have a similar effect on off-frequency patterns but is increasingly unreliable as the deviation from the design frequency varies.

5-10.6 GAIN CHARACTERISTICS OF APERTURE ANTENNAS

For system performance prediction purposes, knowledge of the geometry of the simpler types of antennas such as the dipole, disc-cone, log periodic, and helical types is adequate to make fair estimates of the coupling that will be obtained. With aperture types, the prediction would be subject to statistical uncertainty.

5-10.6.1 Statistical Description

A typical pattern is shown in Fig. 5-112. The irregular nature of the side and back lobe region is characteristic and will vary in detail from one antenna to another, even for antennas of the same type. The statistical properties of this pattern are shown in Fig. 5-113. To simplify prediction work it is common

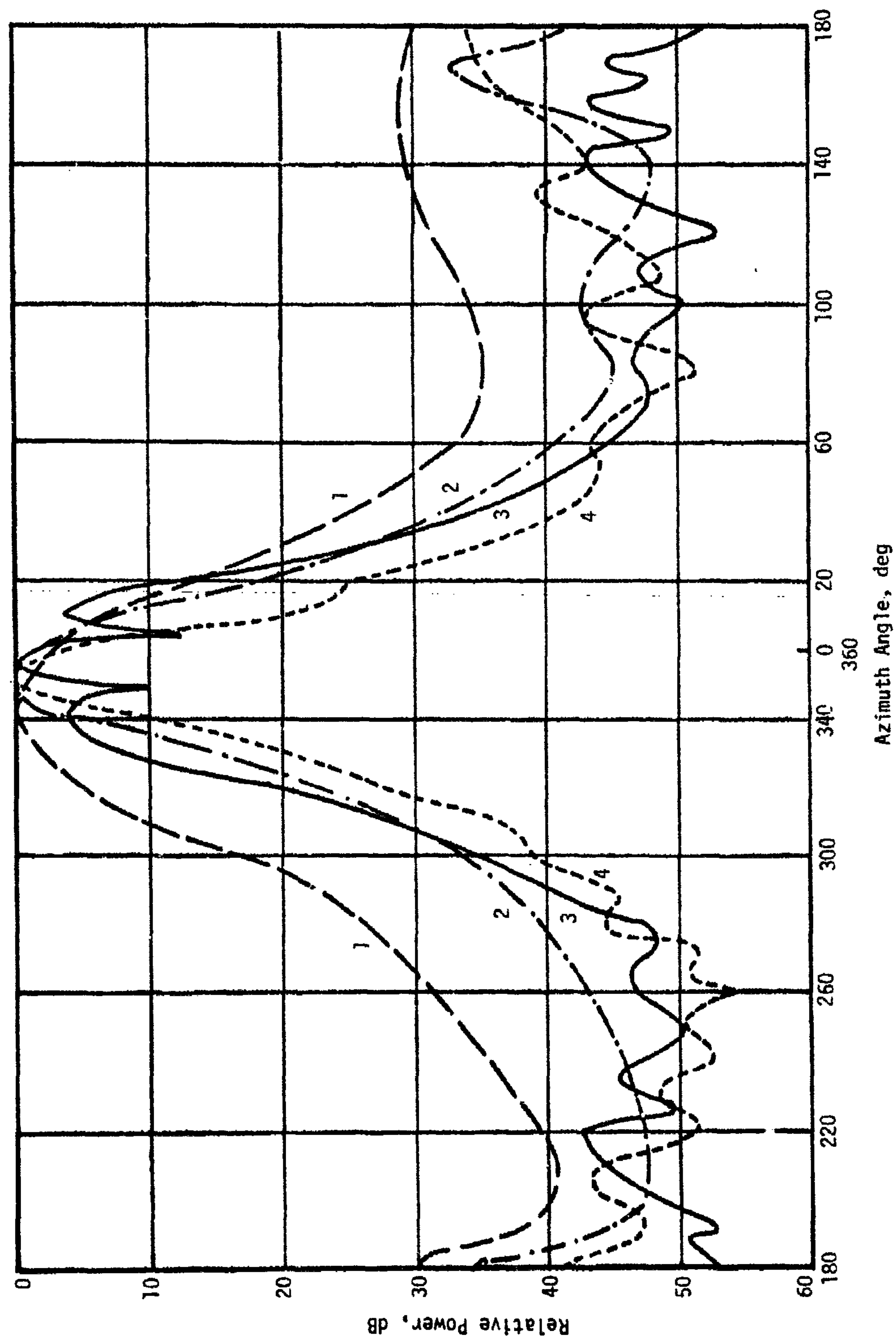


Figure 5-109(A) Exponential Horn Azimuth Patterns, Vertical Polarization

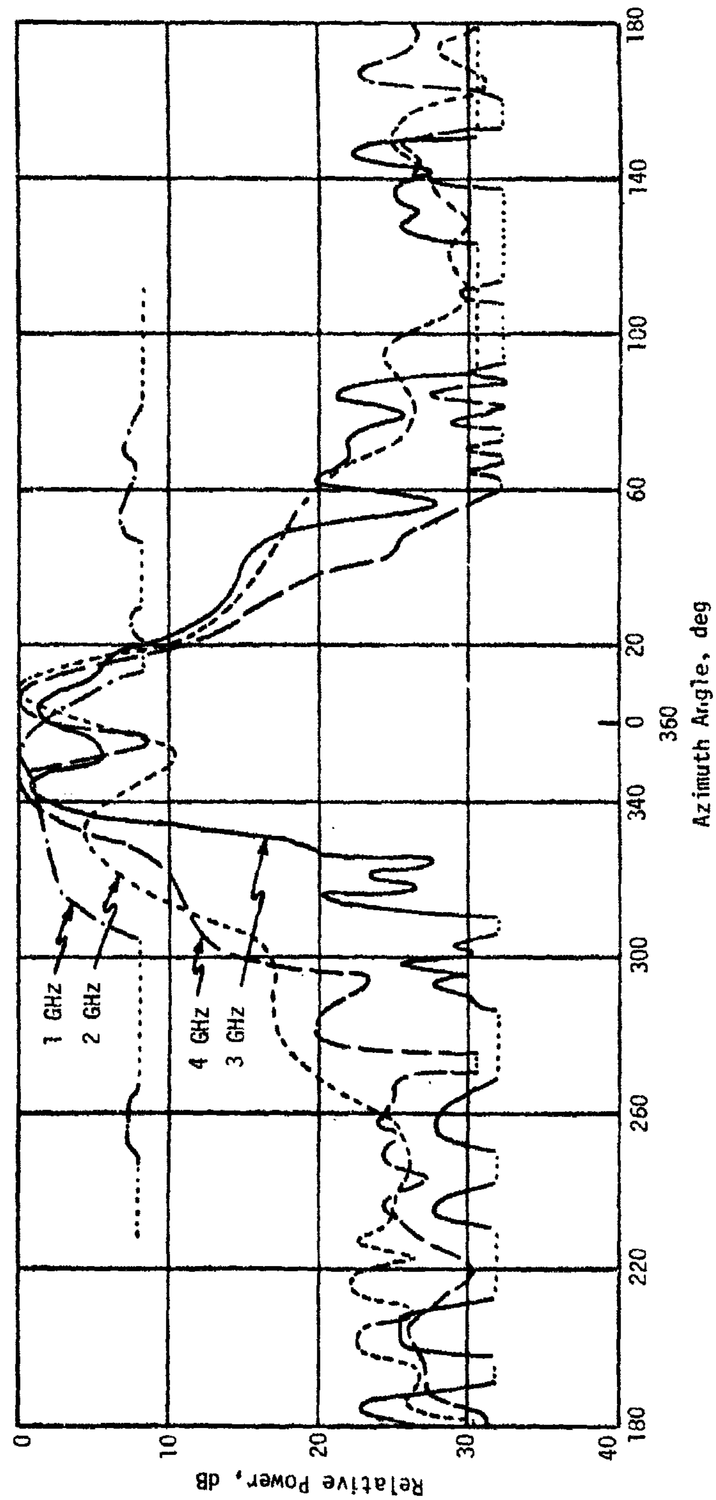


Figure 5-109(B) Exponential Horn Azimuth Patterns, Horizontal Polarization

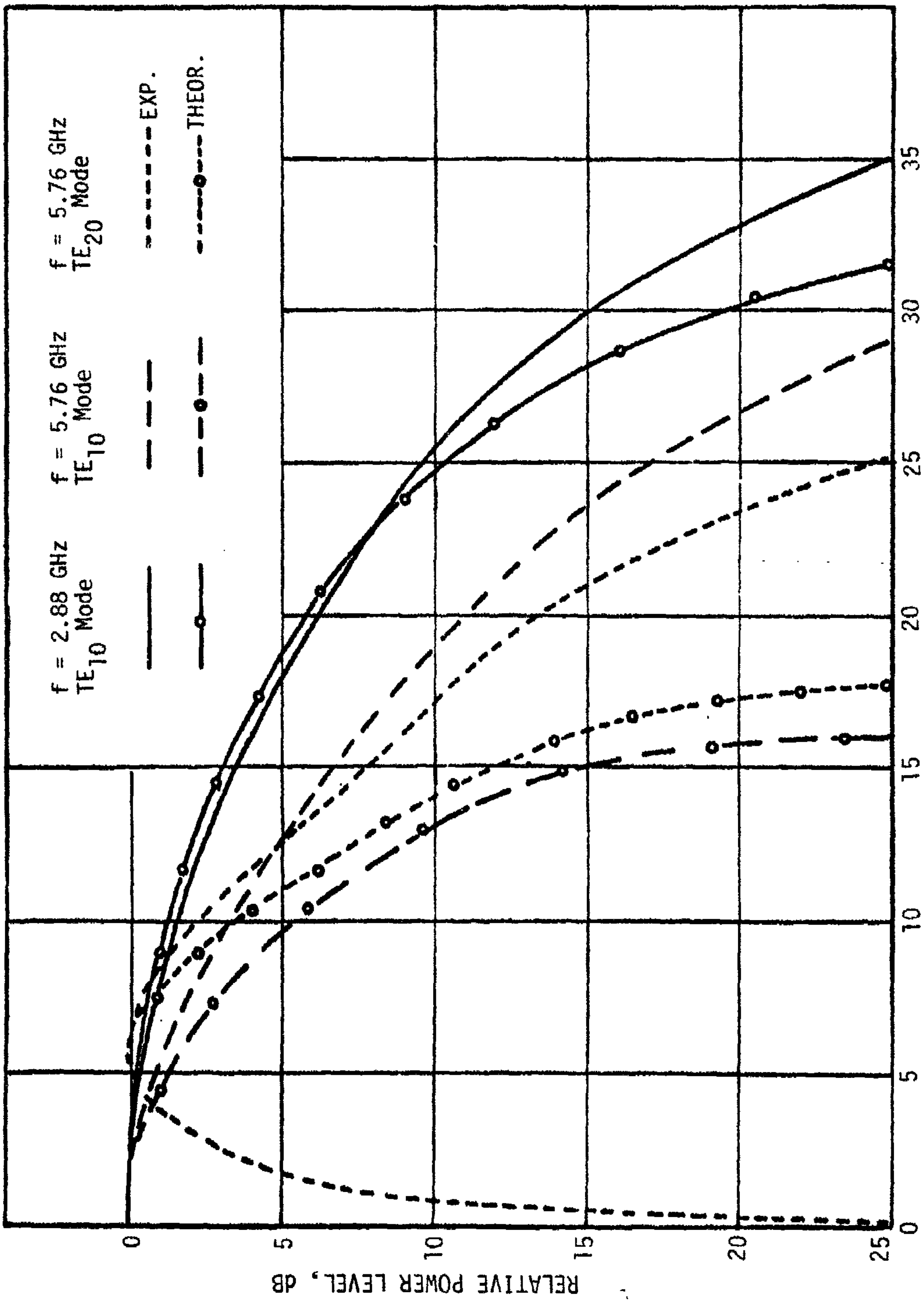


Figure 5-110. Horn Patterns as Function of Frequency and Excitation Mode

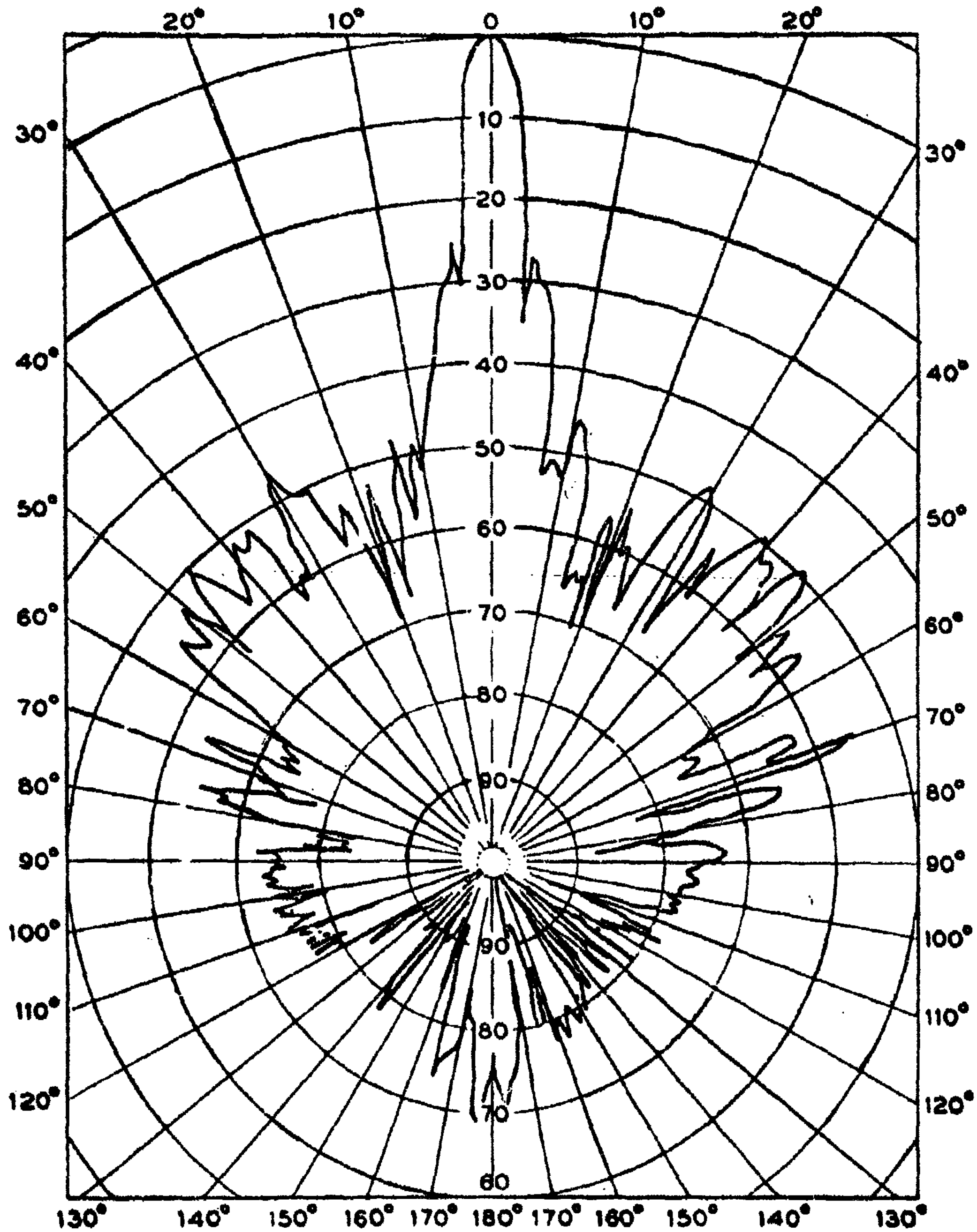


Figure 5-111. Measured Pattern in dB of Horn Parabola Antenna at 3.74 GHz (Ref. 57)

TABLE 5-15
HORN PARABOLA CHARACTERISTICS
(From Ref. 58)

Frequency, GHz	3.7 to 4.2		5.925 to 6.425		10.7 to 11.7	
Polarization	Vert.	Horiz.	Vert.	Horiz.	Vert.	Horiz.
Midband gain dB above isotrope	39.6	39.4	43.2	43.0	48.0	47.4
Front-to-back ratio, dB	71	77	71	71	78	71
Half-power beamwidth						
Azimuth, deg	2.5	1.6	1.5	1.25	1.0	0.8
Elevation, deg	2.0	2.13	1.25	1.39	0.75	0.88
VSWR	1.015	1.0055	1.020	1.008	1.007	1.01
Crosstalk between antennas spaced 20 ft apart						
Ant. back to back, dB	140	122	140	127	139	140
Ant. side to side, dB	81	89	120	122	94	112
Cross polarization discrimination*, dB	46		51		53	

*Only within the main lobe of the pattern.

to use a three-level approximation as shown in Fig. 5-114.

The statistical representation is especially useful in cases where the interacting antennas are rotating or otherwise have varying orientation with respect to each other. The three-level representation is useful if the orientations are fixed or vary only in restricted ways.

Empirical data on typical values of the levels and their expected standard deviations in terms of the gain of the main beam have been tabulated (see Ref. 19).

5-10.6.2 Near Field Considerations

In antenna field prediction work it is convenient to define the boundary between the Fresnel (or near-field) and the Fraunhofer (or far field) regions of an antenna. The boundary R is considered given by

$$R = \frac{2D^2}{\lambda}, \text{ m} \tag{5-39}$$

where D is the antenna diameter and λ is the wavelength (all dimensions in the same units). In the Fraunhofer region an antenna in free space produces an energy density that varies inversely with the square of the distance (see, e.g., Eq. 3-147). In the Fresnel region, the mean energy density has a lower value than predicted by the far-field relation but the actual value may vary around the mean as the distance changes (Ref. 59). Correction factors to be applied to the inverse distance squared formula have been calculated for on-axis fields of rectangular apertures with various types of illumination as shown in Figs. 5-115 to 5-119 (Ref. 60). Additional data on Fresnel region patterns are given in Ref. 19.

5-10.6.3 Design Considerations

Aperture antenna patterns can be specified in terms of two angles, azimuth angle φ and elevation angle θ (Fig. 5-120). Assume an aperture in the Y-Z plane centered at the origin, with radiation generally in the O-X direction. D_Y and D_Z are the approximate dimensions of the aperture in the corresponding directions. Angles φ and θ define the direction of the line OP.

In the far-field (Ref. 38, Chapters 9, 10, and 11), most of the energy radiated (or greatest sensitivity)

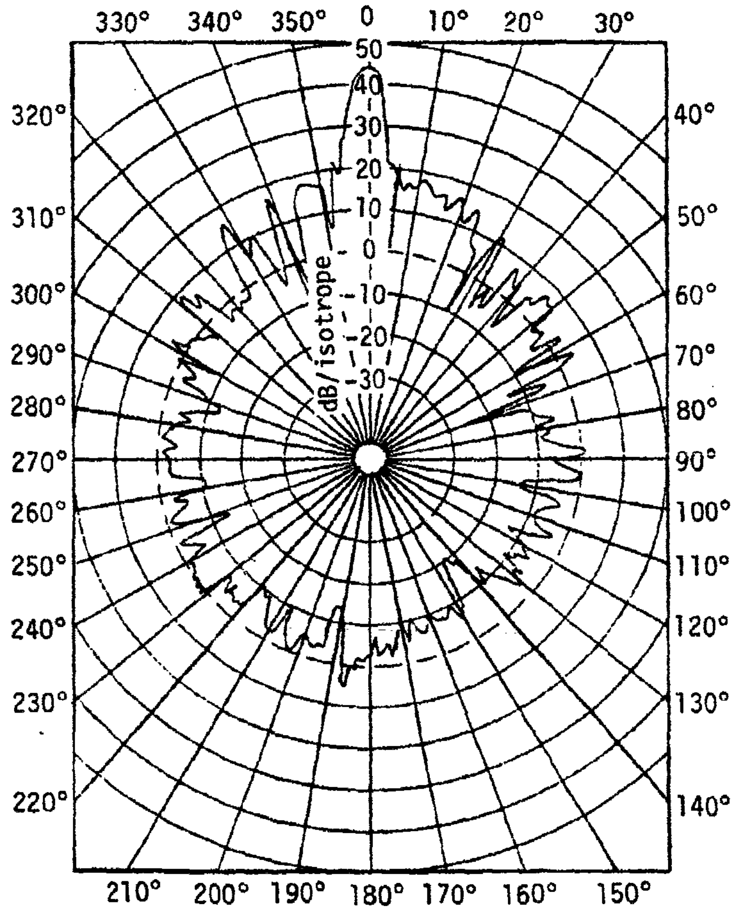


Figure 5-112. Typical Directional Antenna Pattern

will be in a solid angle, generally in the *O-X* direction of approximate dimensions $\Delta\varphi_B$, $\Delta\theta_B$, given approximately by:

$$\Delta\theta_B \approx \frac{\lambda}{D_Z}, \text{rad} \quad \text{and} \quad \Delta\varphi_B \approx \frac{\lambda}{D_Y}, \text{rad} \tag{5-40}$$

where λ is the wavelength of the signal radiated or received. The main beam will tend to be pencil-like if D_Y and D_Z are approximately equal, or fan-like if they are unequal (e.g., $\Delta\varphi_B \ll \Delta\theta_B$ if $D_Z \ll D_Y$). More precisely, the gain $G(\theta,\varphi)$ — a measure of radiation or sensitivity in the direction (θ,φ) — is given by the square of the magnitude of the two-dimensional Fourier transform of the excitation or sensitivity in the aperture plane (with suitable normalizations, coordinate conversions, and account taken of polarization induced by the direction of excitation currents or fields). From this fact, two conclusions are reached regarding interference characteristics inherent in antenna design:

a. The antenna patterns can be expected to have significant sidelobes in directions other than the pointing directions. Generally, these will be smaller and smaller as angular separation from pointing direction increases. However, for high-power, high-sensitivity radar systems, significant interference may exist between noncooperating equipments.

b. Gain in pointing direction will be maximum when amplitude and phase of illumination across the aperture are uniform. This will provide a minimum mainlobe-beamwidth, but appreciable sidelobes. For a rectangular aperture, the sidelobes may be only 13 dB below the mainlobe.

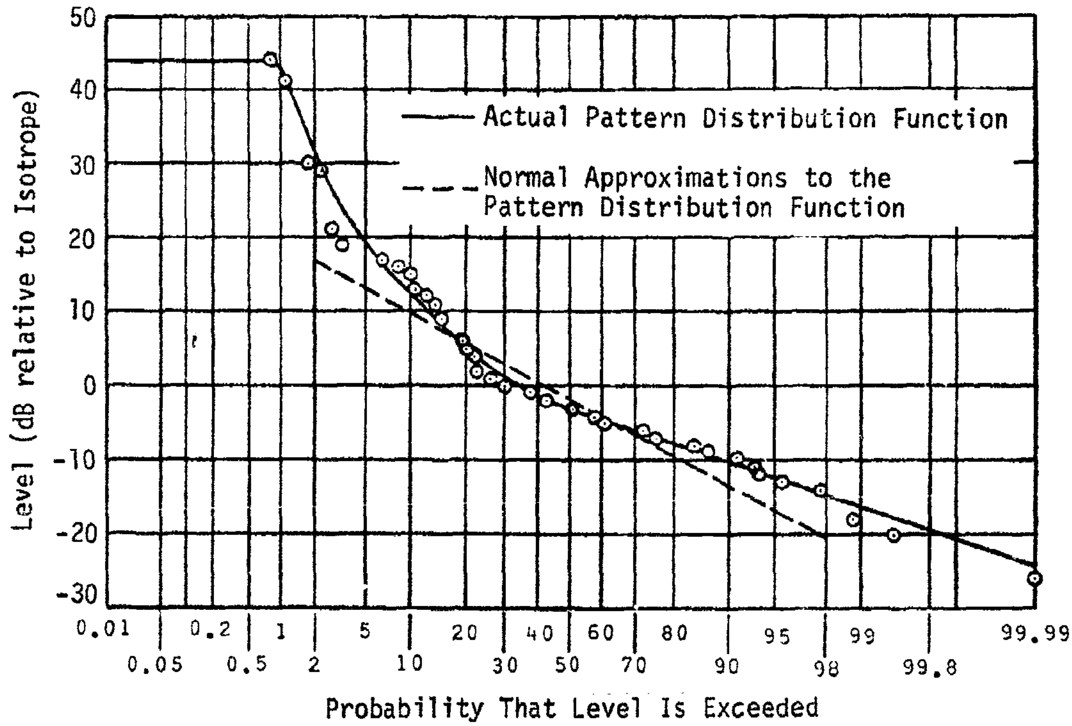


Figure 5-113. Pattern Distribution Function for Typical Antennas

Directional selectivity control of interference involves the following specific alternatives:

a. Decrease dimensions of main beam.
b. Decrease sidelobes.
c. Decrease back radiation (sensitivity).
d. Use auxiliary elements.
e. Use properly placed shielding.

Items a, b, and c are pertinent for both suppression of generated interference and reduction of receiver interference susceptibility. They are concerned intimately with the physical design of the antenna — the type of illumination used, the precision of its fabrication, the size and shape of reflector or lens used (usual in mechanically-steered antennas), and the quality and uniformity of modules in phased array systems. Items d and e are specific techniques concerned with reducing receiver sensitivity to signals from undesired directions.

In light of the preceding discussion, Items a and b generally are conflicting in that steps taken to decrease the main beamwidth will usually cause sidelobes to increase. As a result, "optimum" compromise designs have been devised (Ref. 38, Chapter 9) involving amplitude and/or phase taper of aperture illumination (or sensitivity) for sidelobe reduction without excessive main beam spreading. In mechanically-steered antennas, this can be achieved by suitable shaping and placement of illuminating (waveguide) feedhorn with respect to reflecting dish or dielectric lens. Also, proper shaping of the reflecting surface can provide desired beam shapes such as the cosecant-squared antenna pattern used in air-search radars for providing echo variation effectively independent of range for constant altitude targets. In phased array, electronically-steered systems (Ref. 38, Chapter 11), the necessary tapers of antenna illumination can be achieved by insertion of suitable attenuation and/or phase shifts in the signal channels associated with each element of the array.

Item c can be achieved by careful design and placement of the illuminator with respect to a reflecting dish; also by adding suitable wavetraps (Ref. 19) near the circumference of the antenna surface to prevent diffraction of significant energy to the reverse direction.

Item d (auxiliary elements located near the main antenna) can be used to suppress signals received from certain directions. The Sidelobe Blanking Technique (Ref. 38) uses a small broadbeam antenna with a separate processing channel with decision circuits

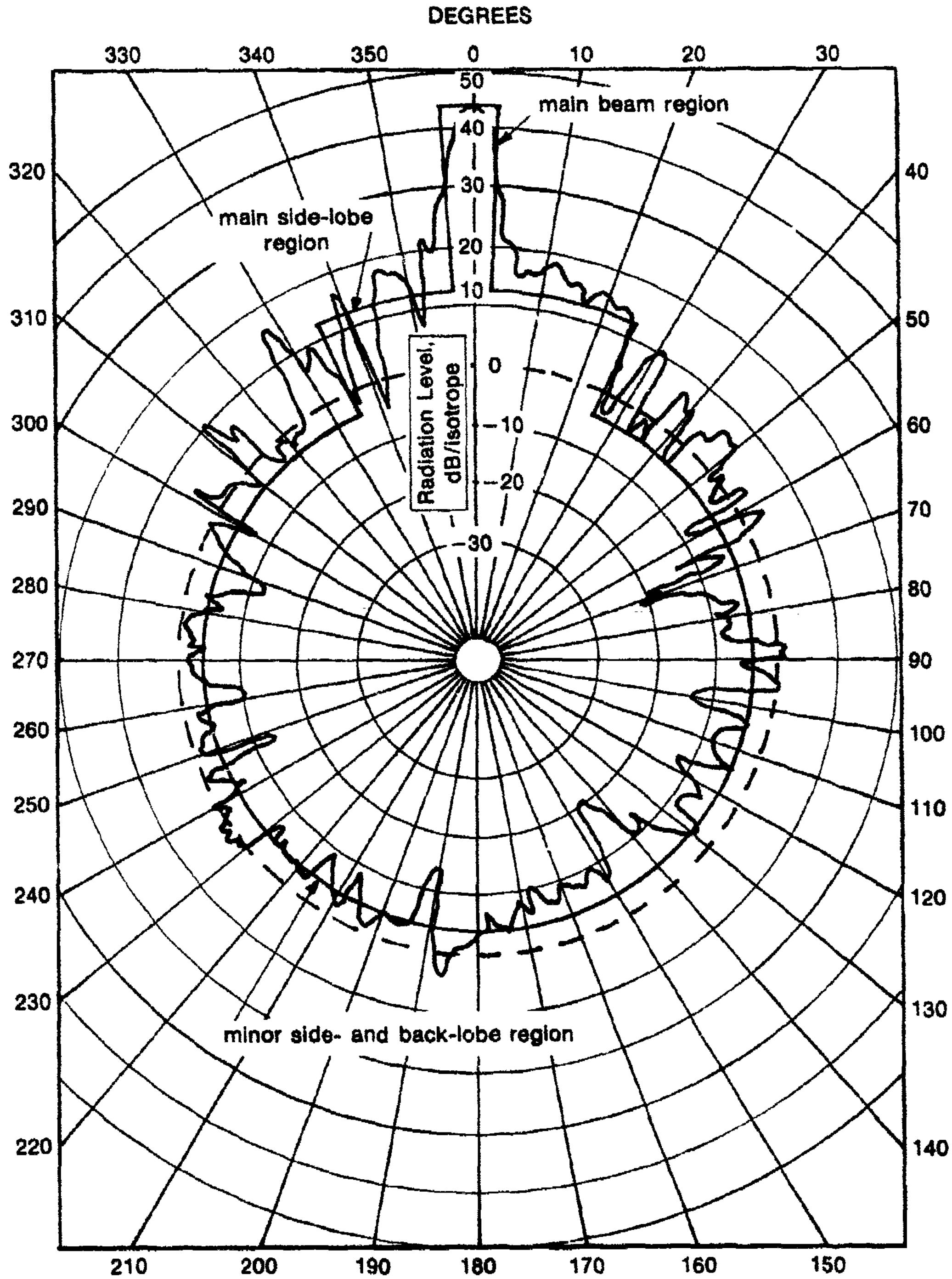


Figure 5-114. Three-sector Representation of Antenna Pattern

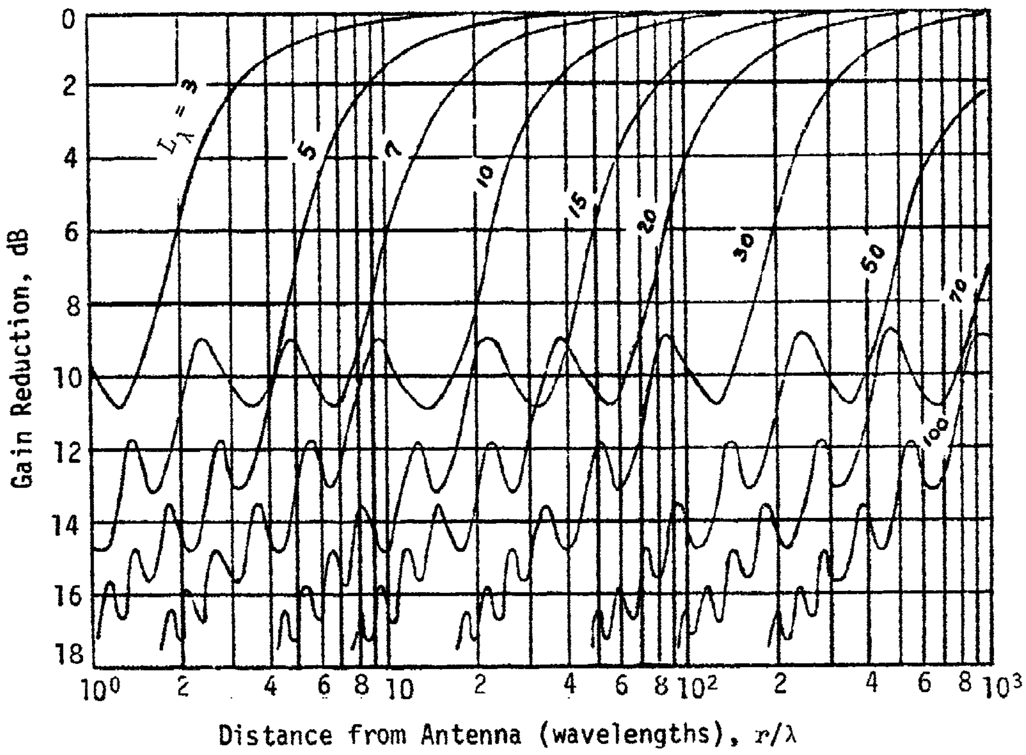


Figure 5-115. Fresnel Region Gain Correction for Uniform Illumination (L_λ = aperture diameter of antenna measured in wavelengths, dimensionless)

which reject apparent echoes for which the amplitude of the received signal in the auxiliary channel is too large compared to that received in the mainbeam, as would be the case for an interfering signal entering a sidelobe of the mainbeam. The Coherent Sidelobe Cancellation Technique forms a signal to subtract from the main beam signal at RF or IR with amplitude and phase adjusted to cancel out all interfering signals from a given direction in the sidelobes. The amplitude and phase adjustments may be performed adaptively if the interfering signal is of sufficient amplitude and duration.

Item e, use of shielding, is important when high-powered transmitted signals originate physically near sensitive receivers. Strategically placed shielding can help to reduce such effects. In predicting the interfering effects present, it is important to make use of the near-field pattern corrections described in par. 5-9.6.2.

5-11 INFRARED EQUIPMENT

5-11.1 INTRODUCTION

The region of the electromagnetic spectrum which is occupied by infrared radiation extends from about 1 mm on the low-frequency end up to about 1μ on the high-frequency end. This is divided into three regions termed near infrared (NIR), intermediate infrared (IIR), and far infrared (FIR). These regions differ primarily in the nature of sensitivity of detectors. In the IIR region, photosensitive semiconductors of lead sulfide, lead telluride, lead selenide, and germanium materials are used. In the FIR region, the bolometer is used to detect and measure the radiation. This device absorbs the incoming radiation and converts it into heat which causes an electrical resistance to increase in temperature. The radiation is measured as a change in temperature. In the NIR region, photographic plates and photoemissive cells are used and

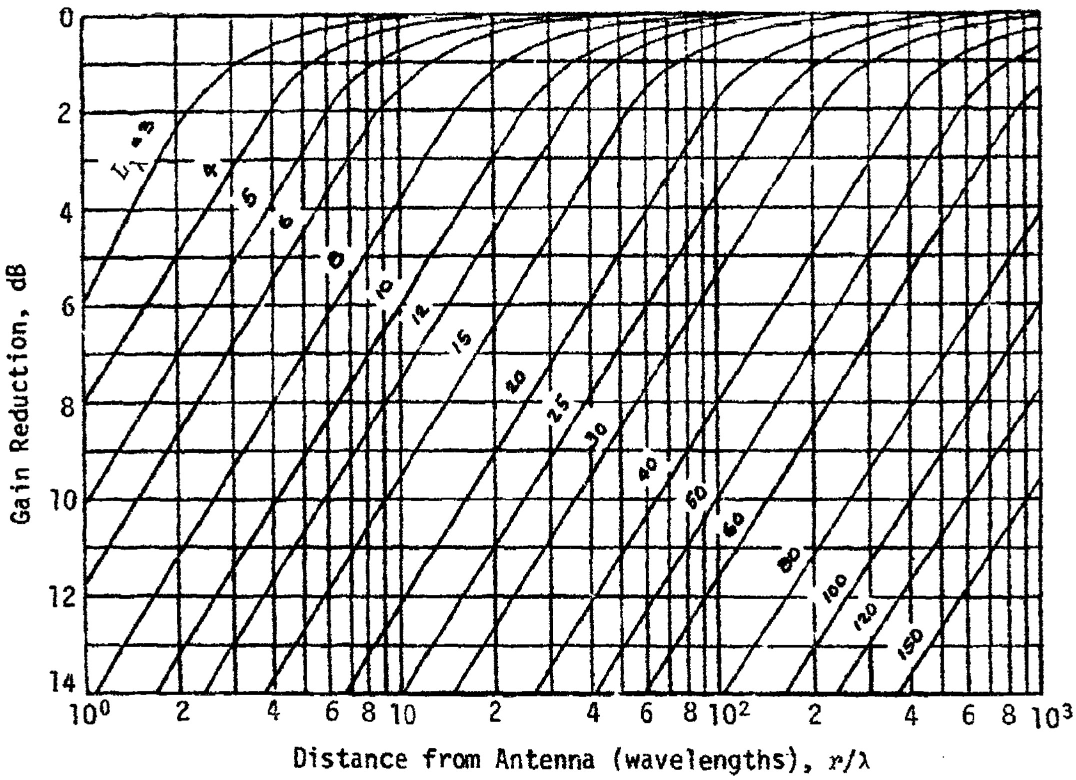


Figure 5-116. Fresnel Region Gain Correction for Cosine Illumination
(L_λ = aperture diameter of antenna measured in wavelenghts, dimensionless)

some phosphors are satisfactory. Each type of detector has specific sensitivities and usefulness. For more information on this subject see Ref. 61.

5-11.2 EQUIPMENT TYPES

Infrared equipment can be divided into two types. One is the relatively self-contained type such as represented by the infrared viewer. In this device one end of the device is pointed in the direction of the object to be viewed, and one looks into the other end and sees a corresponding visual image. In operation the infrared rays are focused into an image on a sensitive screen which converts the infrared energy into electrical signals, which can then be converted back to visible light by means of a cathode-ray type device. Such units can be relatively self-contained, except possibly for the power supply.

The other type of device is exemplified by the diagram which is shown in Fig. 5-121 (Ref. 61). In the extreme case each of the blocks shown on this figure could be a separately contained device interconnected to the others with wires and cables. Clearly, in this device there is much more overall exposure to the effects of undesired electromagnetic radiation.

5-11.3 EMISSION CHARACTERISTICS

Normally, one would not expect an infrared device to be a significant emitter of electromagnetic energy. However, to the extent that it has motors and servo indicators for the purposes of control of the orientation of the sensor, it is possible for electromagnetic fields to exist in the immediate vicinity which could affect nearby sensitive devices. Another possible source of energy would be in a power supply designed to convert low voltage to high voltage required for some of the viewing tubes. If an inverter is used, it would have the possibility of radiating or conducting interference on interconnected wires at the inversion frequency and at its harmonics (see par. 5-5.5.4.4).

Both of these phenomena are discussed in other areas of this handbook and are not treated further here.

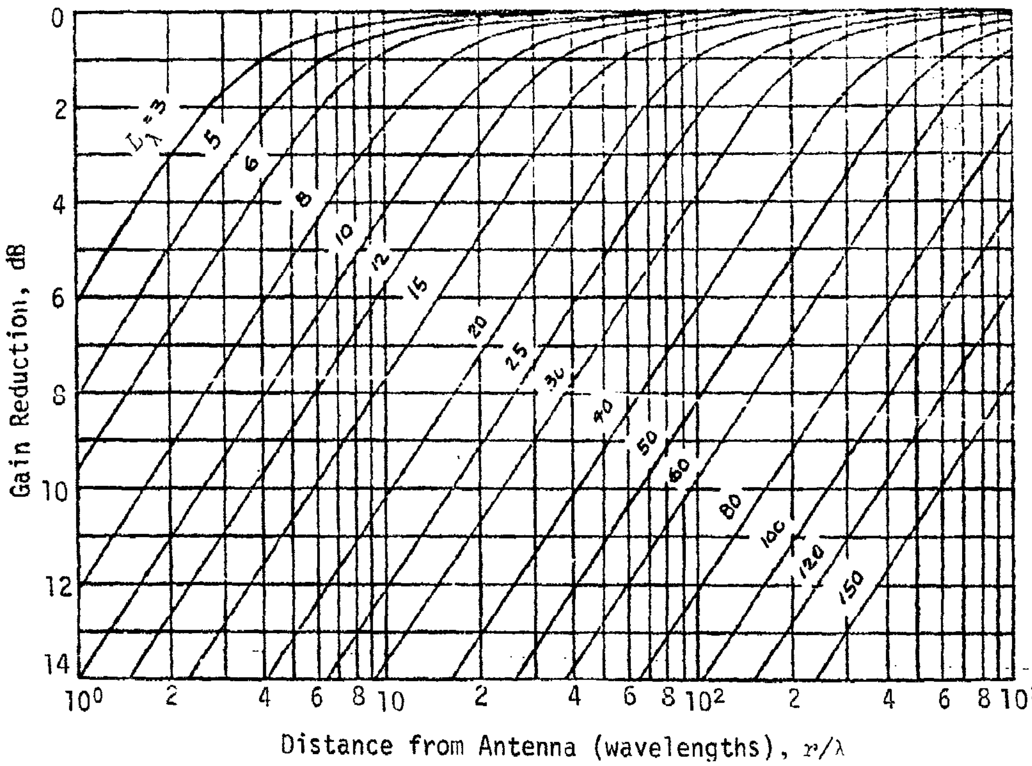


Figure 5-117. Fresnel Region Gain Correction for Cosine Squared Illumination
(L_λ = aperture diameter of antenna measured in wavelengths, dimensionless)

5-11.4 SUSCEPTIBILITY

Susceptibility to electromagnetic fields can occur in two specific ways: (1) by virtue of penetration of the aperture through which the infrared rays pass to the infrared detector, and (2) through interconnecting wires and cables such as required for the system shown on Fig. 5-121. The treatment of the coupling to wires and cables is not significantly different than the treatment of other types of systems containing electrically interconnected units. One should be cautious about harmonics and transients on the power line coupling into the device to upset the stability of the power supply, the servo devices, and cause interference in the output display as the result of direct interference with electrical signals from the sensor or from the signal processor to the display. Normally, the cables from the sensor to the signal processor will be the most sensitive cables because usually they operate at a lower voltage level than other cables. In a typical situation, one examines the voltage levels on each of the cables and determines their susceptibility levels in view of the functions that are performed; takes into consideration whether the signals are analog or digital and their voltage levels; estimates the coupling to sources of interference in the vicinity; and then takes the necessary steps to improve shielding, filtering, and cabling.

Interference directly through the infrared sensor aperture is a slightly different problem from those that have been treated elsewhere in this handbook. By its very nature, the aperture must be transparent to infrared rays, and any practical window material that is used for this purpose also will admit electromagnetic waves. These electromagnetic waves could penetrate into the electrical circuits of the device if it does not have adequate metal barriers between the infrared optical part and the electrical circuits. However, since it is possible to insert such barriers, penetration beyond the infrared detector should not normally be of any consequence. Thus, the susceptible

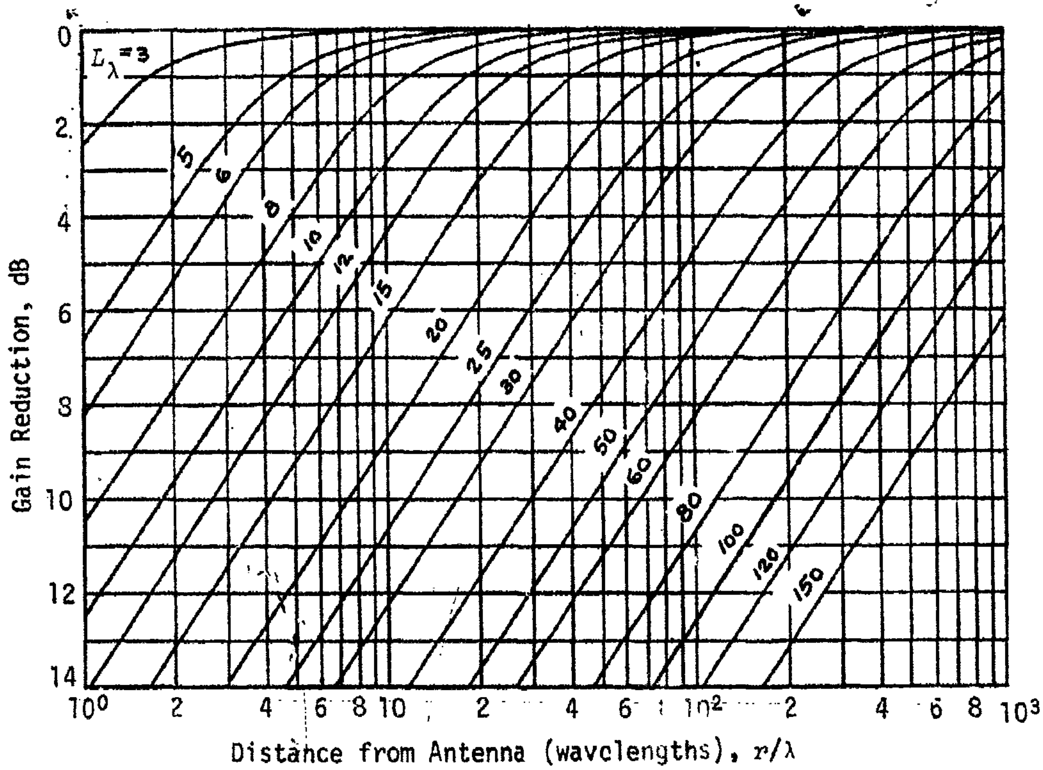


Figure 5-118. Fresnel Region Gain Correction for Cosine Cubed Illumination
(L_λ = aperture diameter of antenna measured in wavelengths, dimensionless)

element is the infrared detector on which the electromagnetic radiation can impinge directly or on conductors connected to it. Since the infrared detector is frequently a semiconductor or nonlinear electrical element, partial detection of the electromagnetic radiation can occur at the infrared sensor and be passed on to the rest of the electrical circuit. The level of radiation required to produce disturbing effects in the detector will depend upon the material used as detector and may vary considerably from one type of detector to another. Data on the sensitivity of such elements are not readily available at the present time.

5-11.5 CONTROL AND SUPPRESSION TECHNIQUES

Control of susceptibility of the infrared portion of infrared equipment can be accomplished in three basic ways: (1) by the use of conducting glass, (2) the use of fine screen mesh in the aperture cover, and (3) the use of filters.

The technique of covering a transparent material with a thin metallic coating which will provide a substantial conductivity to the surface of the transparent material without seriously degrading its transparency has been mentioned in pars. 3-1.3.2.3.2 and 5-4. The procedures for preparing these materials are well established commercially and by estimating the strength of fields to be experienced and the sensitivity of the circuits in the detection section of the device, one can estimate the attenuation which is required. Attenuation is directly related to the conductivity of the coating placed on the window material. In a similar way, one can design a screen mesh to attain a certain attenuation of the radiation. The screen mesh must be extremely fine and should not occupy a very significant percentage of the total effective window. Design formulas are given in par. 4-6.5.2.

Filters can be used on the wire connections in order to prevent the propagation of the currents on the wires connected between the sensor and the interconnected electronic circuits. These filters can be simple shunt capacitors of relatively small size, such as is characteristic of filters used on pins in connectors.

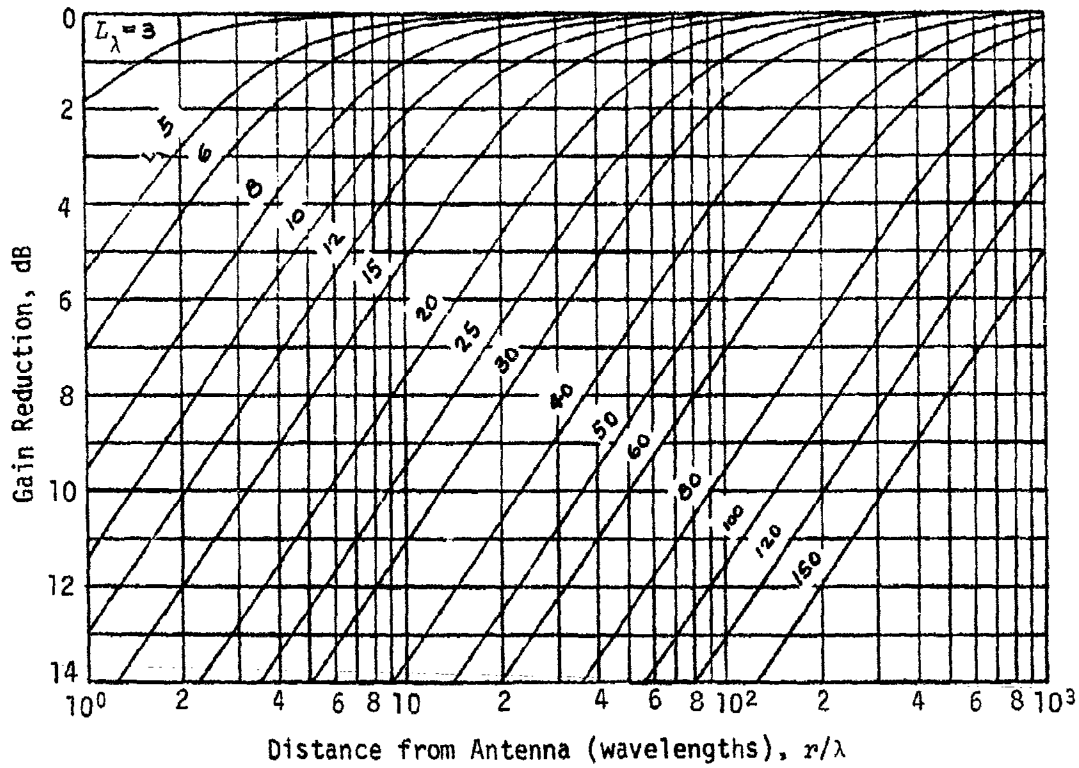


Figure 5-119. Fresnel Region Gain Correction for Cosine Fourth Illumination (L_λ = aperture, diamter of antenna measured in wavelengths, dimensionless)

The most serious radiation that can cause difficulties with this type of equipment is that which arises from radar systems. As discussed in par. 4-6.5, the ability of electro-magnetic radiation to penetrate the aperture depends upon the wavelength of the radiation. For typical apertures used in infrared devices, appreciable penetration does not occur below frequencies in the gigahertz range. When apertures are the order of a wavelength in dimension or larger, radiation can pass through them relatively freely (unless protected by a conducting coating or screen mesh). Thus, by using the formulas available in Chapter 4, one can estimate as a function of the incident power density and frequency, the amount of power that could be converted into a voltage or current in the internal leads in the device.

5-11.6 INTERFERENCE CRITERIA

The level of electromagnetic radiation which will produce degradation of infrared system performance will depend upon the levels at which sensitive circuits operate in the infrared device. In most cases, these devices are designed to have a maximum sensitivity. Normally, this means that circuits are designed to be sensitive to any type of detected signal which falls above the internal noise level of the sensor itself. As is well known, the ultimate limit is determined by the internal thermal and excess noise characteristic of such circuits (see par. 3-1.2.1). The thermal noise depends upon the operating temperature of the device (related to the so-called resistance noise) and the equivalent noise resistance of the first amplifying device. Such devices operate close to the thermal noise limit but may be some significant number of decibels above that. Also, excess noise may characterize the infrared detector, especially if it has a bias current flowing through it as part of the detection process. The figures that one uses with specific detection arrangements are available in standard treatments of the theory of infrared sensors (Ref. 62).

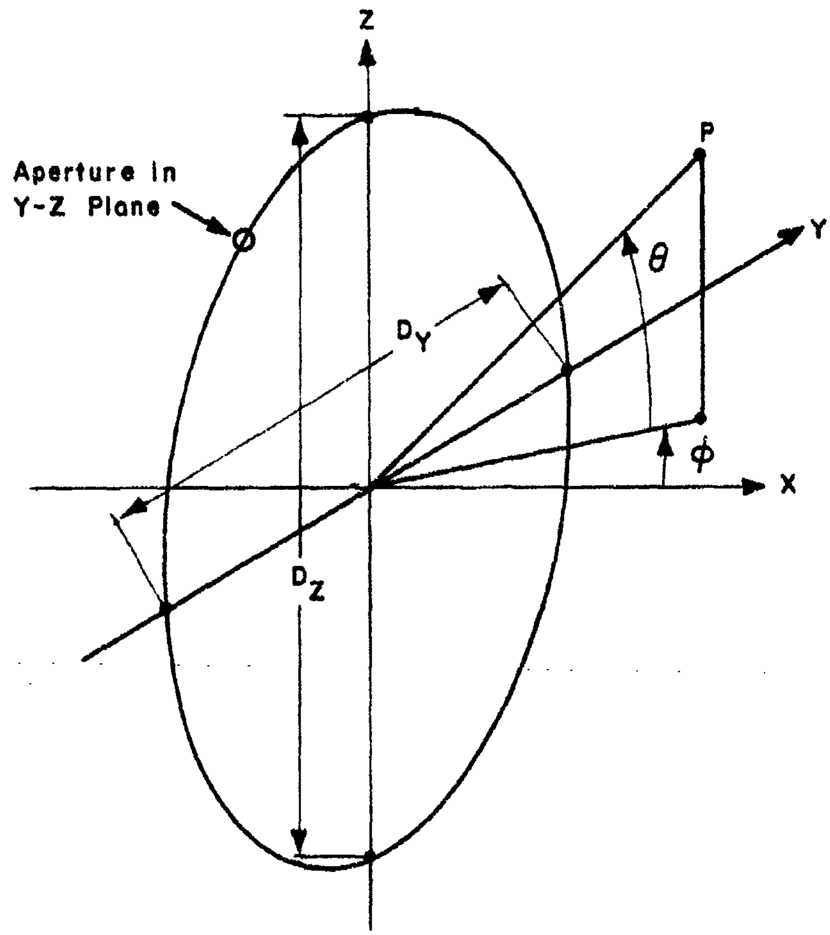


Figure 5-120. Simplified Aperture Antenna Geometry

5-12 AIRCRAFT

5-12.1 INTRODUCTION

Aircraft range in size from small fixed-wing observation types through the variety of helicopters to complex fighter-bombers containing highly sophisticated electronic systems. As with any electrical and electronic system, to produce an aircraft with satisfactory characteristics, EMC control must be started during the conceptual phase of the aircraft design, carried through all of the system and subsystem designs, and eventually be proven in the final aircraft tests. The airframe manufacturer locates and designs the equipment installation, does his own cabling and harnessing, but looks to others to supply such subsystems as propulsion, communication, radar, armament, and instrumentation. Thus, the prime aircraft contractor must not only control the entire system, but must direct and control the development and manufacture of a number of subsystems. Much of the material presented in other areas of this handbook is directly applicable to these subsystems.

EMC is a requirement both internal to the aircraft hull and external to it. In par. 5-12 the major concern is with the internal region. The primary concern in the external region is with interactions between various antennas, which are controlled by techniques of antenna placement, frequency assignment, filters at receiver inputs and transmitter outputs, and specialized techniques such as radar blanking. Models to be used for estimating coupling between various types of antennas on aerospace vehicles are discussed in Ref. 63, and computer programs are available such as discussed in par. 6-4.10.

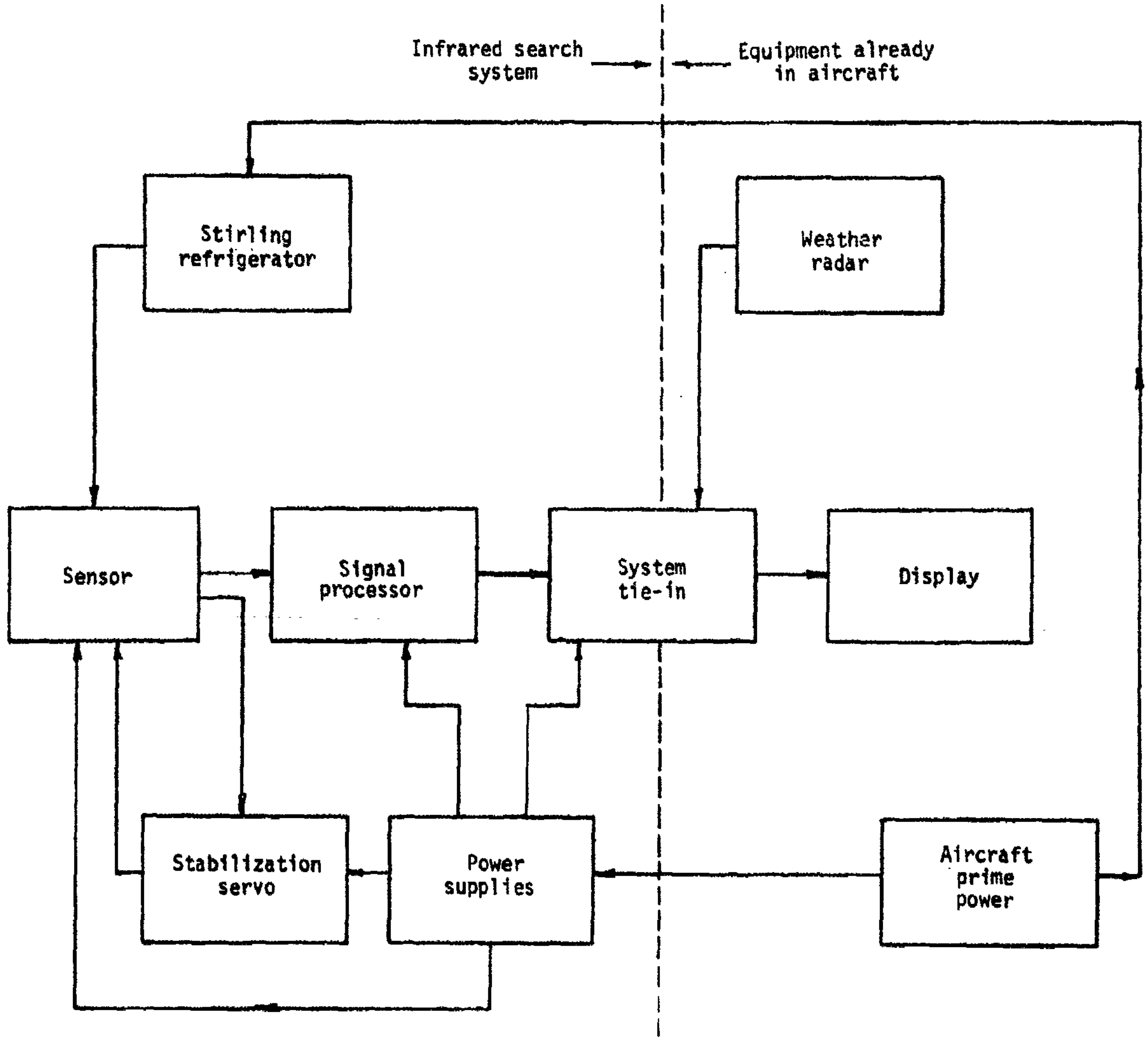


Figure 5-121. Infrared Search System and Its Interconnections With the Aircraft

Fig. 5-122 shows typical helicopter equipment and systems, their locations, and major wire run arrangements. The frequency ranges of communication and associated electronic equipment are shown in Fig. 5-123. Table 5-16 lists typical aircraft equipment.

5-12.2 EMISSION CHARACTERISTICS

Although each element of aircraft cabling or equipment is a potential source of EMI and should be considered in system analysis, experience has shown that particular equipments are of special concern.

5-12.2.1 Power Systems

The electrical power supply system deserves special attention because any undesirable emission impressed on it appears throughout the aircraft by conduction and induction.

A typical power supply may consist of two separate ac generating units providing 115/208-V, 3-phase, 400-Hz ac and 28-V dc. Each system consists of a generator, a control unit feeding the ac distribution system, and a transformer-rectifier to supply dc (see Fig. 124). When the main generators are not on the line, power may be provided by an auxiliary

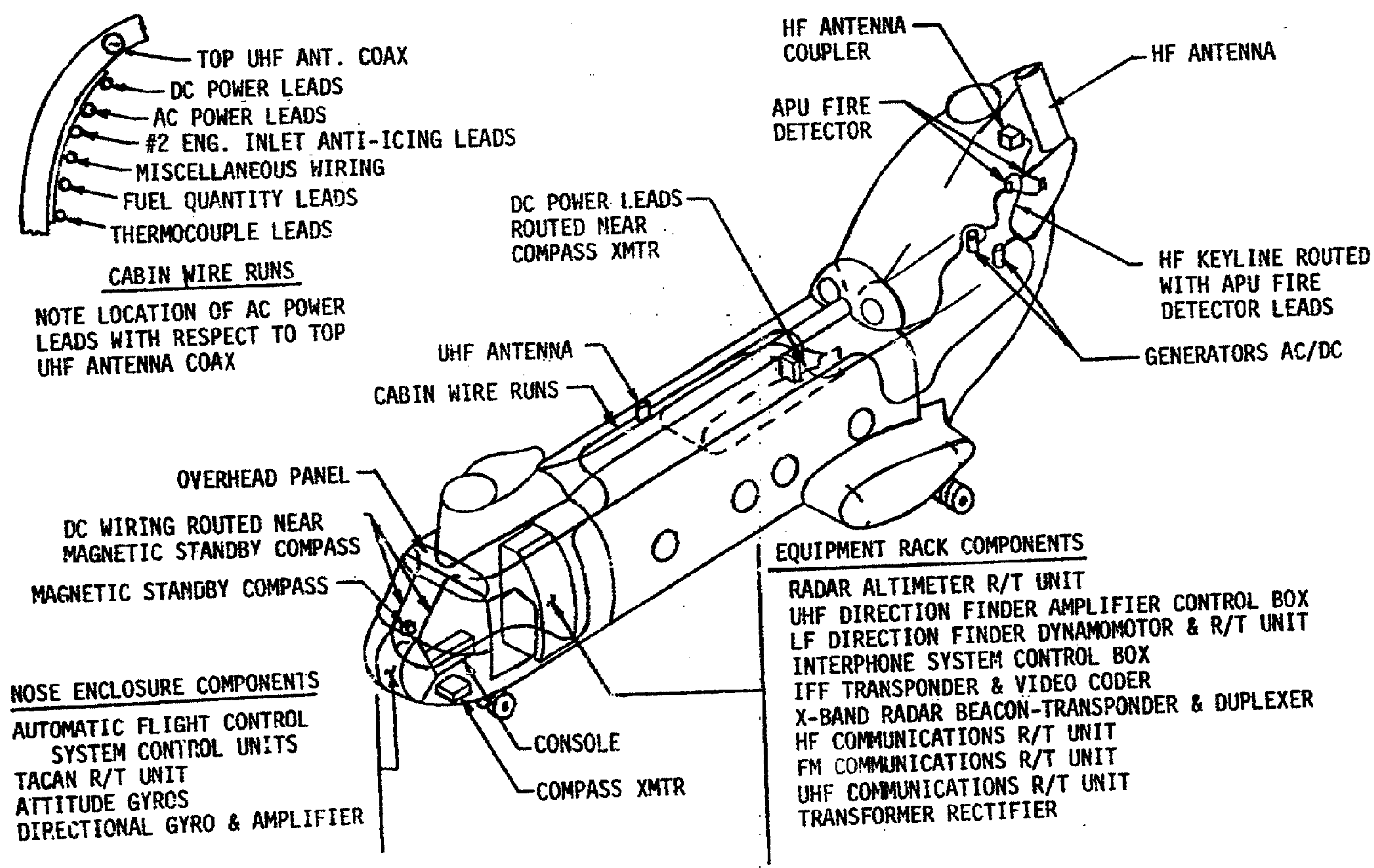


Figure 5-122. Typical Component Locations and Major Wire Runs

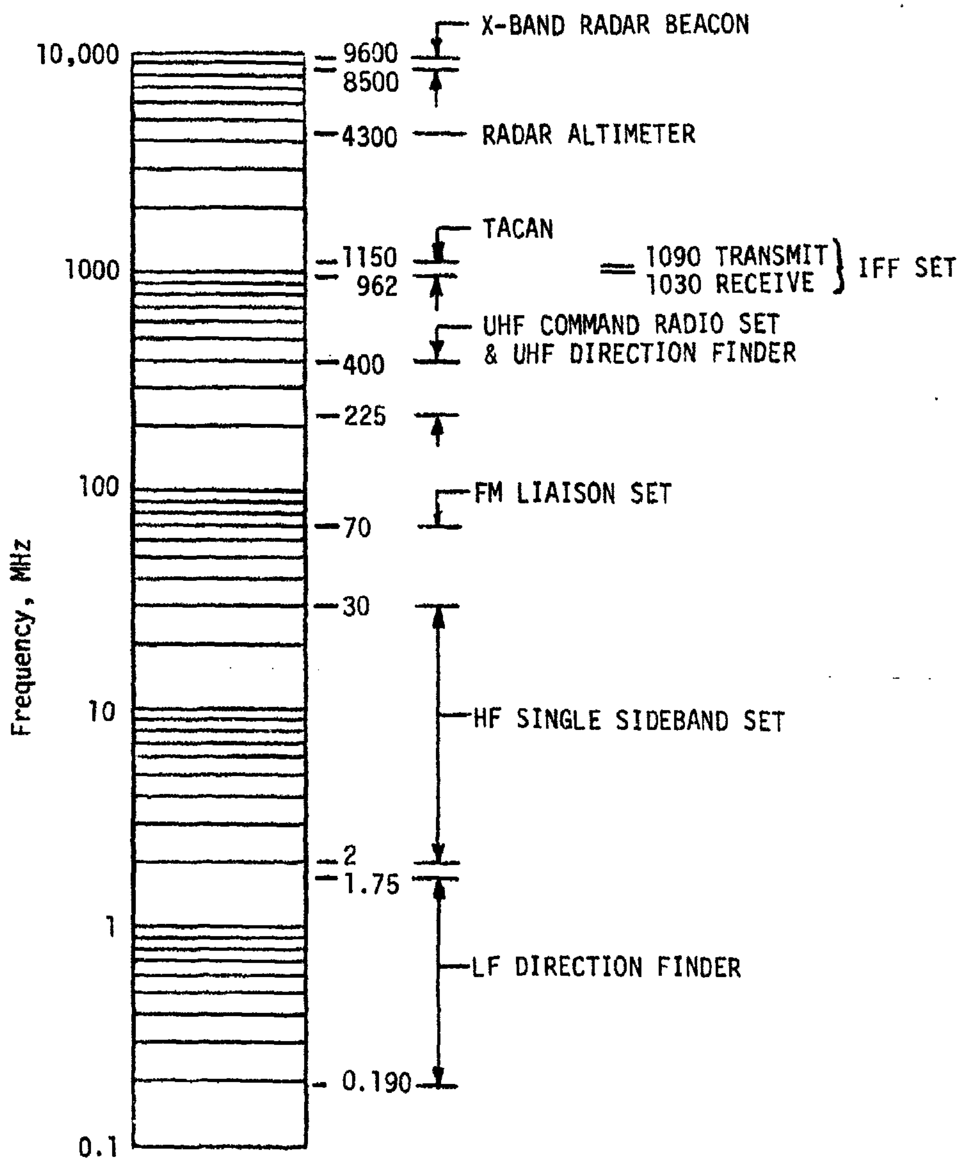


Figure 5-123. Typical Frequency Ranges of Communication and Associated Electronic Equipment

power unit (APU), ac from the APU-driven generator and dc from a transformer-rectifier. It is important to note that quality of regulation usually will be much poorer when the APU is functioning and this may affect power dependent equipment.

Usually the power system voltage is distorted. Ac systems contain substantial levels of harmonics and are subject to random transient variations. Some of the harmonics are contributed by the prime power source. Other harmonics and transients are due to load changes as equipments are switched on and off as discussed in par. 5-2.

These effects can be particularly severe in aircraft because of the limited size of the generators. The levels of voltage distortion are proportional to the magnitudes of the harmonic currents generated in

TABLE 5-16
LIST OF TYPICAL AIRCRAFT EQUIPMENTS

Eng. Pwr. Management System	Radar Altimeter
Engine Controls	Tacan
Ignition Systems	LF Direction Finder
Fuel Quantity System	UHF Direction Finder
Transformer Rectifier	Interphone System
AC-DC Generators	IFF Set
Elec. Sys. Control Panels	X-Band Radar Beacon
Eng. Fire Detection System	HF Communications
APU Fire Detection System	FM Communications
Environmental Control Unit	UHF Communications
Engine Anti-icing System	Electronic Flight Control Sys.
Interior Lighting	Exterior Lighting
Auxiliary Power Unit	Navigation Devices
Mag. Standby Compass	Ordnance
Attitude-Heading Reference Sys.	ECM Equipment

connected equipment and the combined impedance of the generator and intervening power cable. The requirements for the generating equipment are called out in MIL-STD-704 which provides limits for ripple (dc), harmonics (ac), and transients. Because of nonlinear connected loads which feed harmonic currents into the line, it is not unusual to find distortion levels well above specification limits for the equipment itself.

The extensive use of relays and solenoids in remote control of power to various systems adds additional EMI in the form of short duration spikes or relatively long duration surges.

5-12.2.2 Radio Systems

The presence of several relatively high-powered transmitters, together with their cables or waveguides in a restricted space, presents a potential for radiated EMI. Any design, manufacturing, or installation flaws are likely to influence susceptible equipments that are in proximity.

The HF single sideband radio system is generally the most difficult source of radiation to control. Because of its power level and the use of relatively nondirective antennas in this frequency range, substantial fields will be present in the entire aircraft during transmission. The system typically provides two-way communication on any one of 28,000 operating frequencies spaced at 1-kHz increments through the frequency range of 2 to 30 MHz. The basic components are receiver-transmitter, HF control panel, antenna tuner, and a monopole antenna.

5-12.2.3 Environment EMI

Interference originating external to the aircraft is of three types:

a. Precipitation Static. Dust, snow, precipitation, and engine charging static can cause severe interference in radio receiving systems. Noise from corona discharges from the sharp edges of the aircraft couple into the receiving antennas. Precipitation static occurs when the aircraft is operating in precipitation containing ice crystals, while engine charging static is caused by the action of heavy positive ions and free electrons in hot engine exhaust gases. Helicopters generate appreciable static while hovering near the ground in clouds of dust or snow. Various additional types of static have been identified and described. Refs. 64 and 65 should be consulted.

b. Lightning:

Lightning strikes on aircraft generally impact the upper portion of the wing or fuselage, or in the case

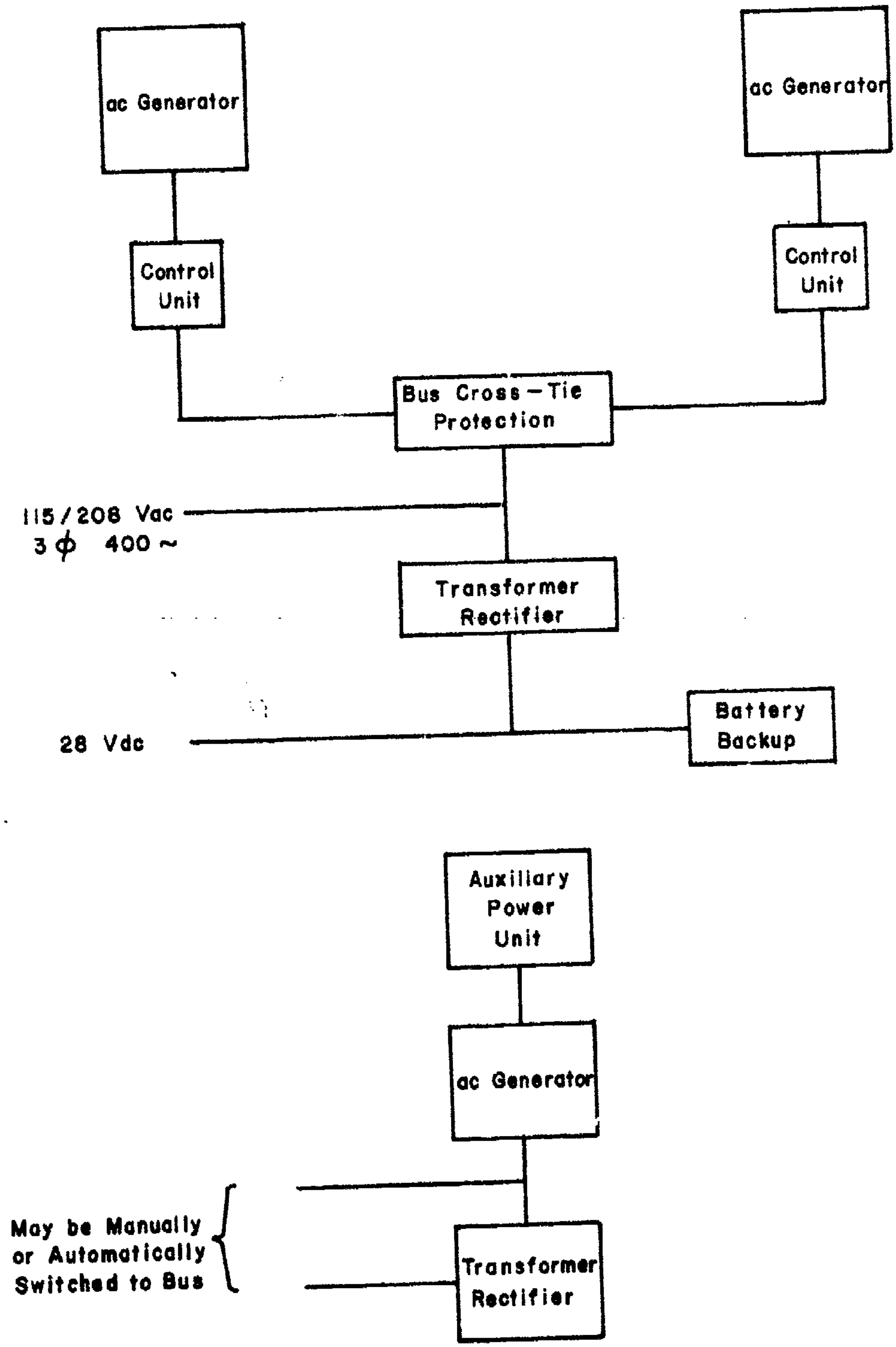


Figure 5-124. Typical Aircraft Power Unit

of a helicopter, a rotor blade. From the point of initial impact, the current seeks a path through the ship to an exit point in the lower portion, probably near the landing gear.

Since the current caused by the strike is extremely high, it is essential that the current be channeled through preplanned conduction and major structural elements (with no sparking) and not diverted to electrical systems and circuits.

c. Intentional and Unintentional Emitters. Aerospace Ground Equipment (AGE) and airport equipment may radiate, induce, or conduct interference to the aircraft while it is on the ground. Radiation encountered in flight, especially near emitters such as powerful ground based radars, may cause severe disturbances. Fields as high as 100 V/m can be experienced on the external surface.

5-12.3 SUSCEPTIBILITY

5-12.3.1 General

All electrical systems in an aircraft can malfunction if subjected to appropriate and sufficient interference. Obviously, susceptible equipment includes all those with antenna inputs and instruments dependent on sensitive input transducers such as used for flight control.

All equipment power requirements must be examined to make sure that raw aircraft power — controlled only by its own supervisory system but perturbed by switches, relays, and solenoids and occasionally jolted by lightning strokes or static discharges — will allow the equipment to function as required.

Dc equipments and subsystems inherently seem to be more susceptible to interference than ac equipments, probably because input filtering may be less effective than on ac circuits. The aircraft intercommunication system, which will have several stations and a comprehensive wiring network, integrates the audio facilities of the aircraft and has been found to be vulnerable particularly to HF radio.

5-12.3.2 Examples

Systems such as navigation equipment, attitude heading reference systems, and electronic flight control systems may contain integrated circuits. These circuits are not only sensitive to spikes but the circuit elements are themselves potential detectors due to their nonlinear characteristics and can be expected to react to high frequency radiation. Computers and other equipment containing logic and decision making circuits are sensitive to interference. Analog circuits are particularly vulnerable because of their relatively slow recovery time. Typical examples of equipment susceptibility that have occurred are (Ref. 66):

a. The auxiliary power unit fire (APU) warning light located in cockpit illuminated when the HF communication system was keyed because the APU fire detector leads picked up interference from the HF communications system control leads.

b. The fuel quantity indicators reacted erratically when the HF communication system was keyed, also due to coupling; in this case between the fuel quantity indicator signal lead and the HF system keyline.

c. The radar altimeter momentarily deflected when the cabin heater was turned on because of a sudden change in line voltage.

d. The magnetic standby compass in the cockpit acted erratically when certain dc loads were in use because of coupling directly between the compass and dc leads.

e. The compass transmitter in the wing sent erroneous signals to the directional gyro system during normal aircraft operation due to direct coupling between the transmitter and the dc power leads.

f. Computer programs performed erratically because of transients in the power supply.

g. Sensors, accelerometers, and telemetry subsystems were inaccurate because of susceptibility to line harmonics.

h. The guidance system was inaccurate because of transients in the power supply.

5-12.4 INTERFERENCE SUPPRESSION

EMC in aircraft is constrained by the limited envelope of the aircarft and by the priority space allocation given to structural integrity, crew function, and equipment function necessary for mission accomplishment. Thus, electrical equipment cannot be bulky, and in some cases must be specifically packaged to conform to the space available. The aircraft envelope, of course, produces real constraints on separation of wires and cables.

Aircraft range and payload are direct functions of aircraft weight. The weight constraint coupled with the need for high reliability for flight safety makes it necessary to perform difficult trade-offs to select redundant equipment. In this engineering climate, it is impossible to be lavish with shielding.

In many cases, the location of equipments, which are already constrained by the size of the aircraft, are further constrained by the necessity of functional location. For example, displays and controls must be located at specific crew stations, and generators are located on the engines.

5-12.4.1 Program Organization

The aircraft as a system is particularly well suited to EMC program organization and implementation described in Chapter 2. The aircraft design naturally flows through the Life Cycle System Management Model (LCSMM) stages, and the EMC organization is a viable part of an organization that includes mechanical, hydraulic, and electrical design; stress groups; weight groups; aerodynamics and equipment experts; and a myriad of manufacturing quality and test personnel.

EMC program plans, design criteria, and test plans all support the aircraft design and manufacture. EMC requirements must be established and placed upon all appropriate subcontractors.

5-12.4.2 Design Criteria

Much of the equipment installed by the airframe manufacturer is, in essence, "off the shelf". Although designed for aircraft use, it is not specifically tailored to the aircraft in question. To counterbalance this situation, most aircraft are one of a line of successful ships, and many of the EMI problems have been solved previously or at least identified clearly enough so that the latest model can profit by prior experience. Because of this hereditary growth, the problem of installing an absolutely new piece of avionic equipment in a currently successful aircraft should be approached with great caution. In particular:

a. Wiring and cabling, and the resulting coupling, should be a major aircraft design consideration.

b. Grounding and bonding must be adequate, well installed, and free from degeneration due to stress or corrosion.

c. Antenna interaction must be minimized starting at the design stage.

d. Particular attention must be paid to EMI caused by HF, susceptibility of dc equipments, and susceptibility of intercommunication systems.

e. Whenever possible, EMC should be achieved without the use of excess shielding, filtering, or suppression diodes.

Wire and cable routing is perhaps the most important of these design considerations. Grouping of like signals with like signals and then appropriate separation of groups such as shown in Fig. 5-122 or as discussed in par. 4-4.11 can be planned by the EMC engineer and issued to the designer in the form of a set of specific directions for categorizing and installing wire bundles. Such initial design discipline will limit coupling calculations to only the most difficult cases.

The EMI engineer should issue grounding and bonding instructions (see par. 4-7) not only to designers involved in equipment installation but to structural designers who interface with the equipment designers. In addition, at an early design stage attention should be given to protection from precipitation and engine exhaust static as well as lightning.

5-12.5 INERTIAL NAVIGATION EQUIPMENT

5-12.5.1 Introduction

Modern aircraft rely upon sophisticated inertial navigation systems for many operational maneuvers. These systems may be of the strapped down type, in which accelerometers and gyroscopes are mounted to the vehicle frame, or the type using a stable platform to carry the accelerometers and stabilizing gyroscopes. Most systems in actual use are of the variety having a stable platform.

5-12.5.2 System Description (Ref. 67)

Fig. 5-125 shows a block diagram of a conventional aircraft inertial navigator. The platform consists of a gyro-stabilized cluster of accelerometers whose outputs are fed to a computer. The computer, which may be the central computer of the aircraft or a special-purpose inertial-navigation computer, calculates the aircraft position and velocity from the outputs of the two accelerometers. The computer also calculates the gyro-precession signals, which maintain the stable element in the desired orientation relative to the earth. A vertical accelerometer sometimes is added in order to smooth the indication of altitude, as measured by a barometric altimeter.

The inertial navigator also contains platform-stabilization servos, a display-and-control panel, power supplies for the platform and computer, and often a battery to protect the computer against power transients. The system may be packaged in one or more containers.

Typically, an inertial system consists of three parts: (1) the stable platform, (2) the computer, and (3) the output indicator and control unit appearing in the cockpit. These units are interconnected with cables. The platform and computer units may be quite close together and in the newer systems may, in fact, be contained in the same enclosure.

5-12.5.3 Emissions

By its nature, inertial guidance equipment is not likely to have high levels of emissions. The strongest emissions would be from motor drives and synchros which can produce significant fields only in their immediate vicinity, and signal lines interconnecting equipments which may carry moderate level analog

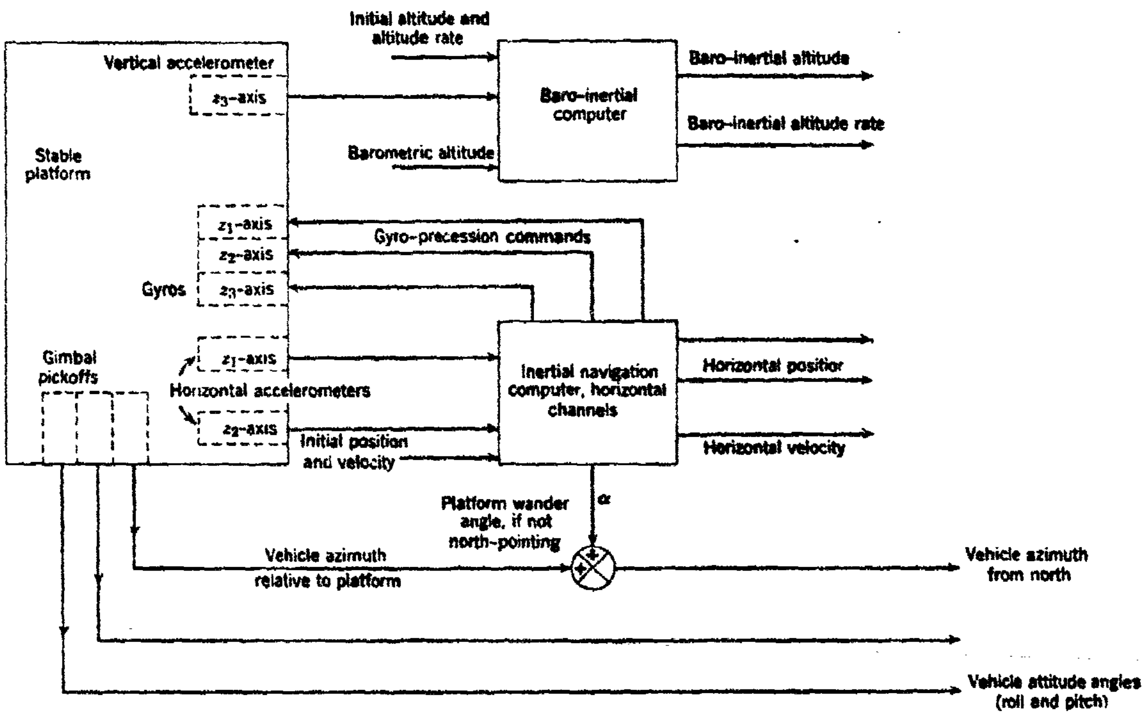


Figure 5-125. Block Diagram of an Aircraft Inertial Navigator

or digital information. These could be significant emitters only to very closely spaced coils and cables.

5-12.5.4 Susceptibility

Because an inertial system is a precision instrument, care must be exercised in its design to reduce its susceptibility to power supply transients, noise voltages, and variations of load current, or the power supply must prevent such variations from reaching the instrument. In some cases the power source may be separate from the main electrical power supply for the other subsystems. It is common to have both a precision electrical power source for use with the precision instruments and a less accurate but higher power source for less critical use throughout the aerospace system. By keeping these two supplies separate it is possible to provide the heating and other less critical loads with a large power source while at the same time using an independent precision ac and dc source for use with the precision instrument.

Cables carrying analog data, such as to and from synchros, could be susceptible because of the precision which may be required. Cables carrying digital data will usually not be susceptible unless the logic level is especially low (see par. 5-14.3)

5-12.5.5 Interference Control

Methods of compatibility control involve standard methods of cable separation and filtering. Filters may be needed in the power supplies, especially where a separate supply is not used. Since circuit sensitivity to interference can vary significantly, depending on the function performed, a careful analysis may be necessary to identify the most critical circuits. Shields around cables are avoided, if possible, to minimize weight.

Shields are used in the platform unit. Most gyros of gimbaled type have ferromagnetic elements (e.g. electric motor laminations) and dc conductors (torquer lead-wires) on the gimbal assembly, and usually also an electrically conductive rotor. The earth's field or a stray field of artificial origin can have two effects. One is magnetic attraction on the iron parts and dc circuits, and the other (usually the less serious) is eddy-current induction effect in the rapidly spinning conductive rotor.

The disturbing effect of fields exterior to the gyro can be reduced to a negligible value by surrounding the instrument with shields of highly permeable material such as permalloy or "Mumetal".

5-12.6 FLIGHT CONTROL EQUIPMENT

5-12.6.1 Introduction

Flight control equipment is that associated with the direct adjustment of aircraft surfaces for purposes of maneuvering the aircraft. This control may be direct through mechanical control, or indirect through a servosystem arrangement. The latter is especially required for large or high speed aircraft where substantial power is required to effect the required adjustments. In some cases the control equipment constitutes small computers in order to stabilize the flight. The computers have been called stabilizers, auto-pilots, stability augmenters, and dampers. This type of subsystem, Fig. 5-126, senses one or more airframe motions and a control signal to oppose this motion. The resulting signal is used to operate the servo actuator which in turn produces the appropriate surface motion.

A typical pitch-axis stability augmentor system is shown in Fig. 5-127 (Ref. 67). This system includes a pitch rate gyro for damping and a normal accelerometer which is used to achieve desirable handling qualities as well as pilot input and attitude gyro and path control inputs. The yaw axis stability augmenter system is similar, with the rate gyro replaced by a yaw axis rate gyro and the normal accelerometer replaced by a lateral accelerometer. The roll axis stability augmenter is also similar to the pitch-axis stability augmenter but no accelerometer is included.

6-12.6.2 Sensors

The most common sensors used with flight control equipment are:

a. Gyroscopes. The rate gyros are conventional single-degree-of-freedom gyros incorporating a rotor driven by a synchronous motor, gimbal structure; and torsion bar providing spring restraint; viscous or magnetic damping; and a microsyn pickoff. They are considerably less accurate and less expensive than inertial gyros. A typical rate gyro (Ref. 67) has an output of ±3 V rms at 400 Hz and is powered by 115 V, 400 Hz for the motor and 26 V, 400 Hz for the sensor.

b. Accelerometers. The accelerometers used to measure the lateral, longitudinal, and normal accelerations of an aircraft are usually simpler and less accurate than those required for inertial navigation systems (Ref. 68). They consist of a spring loaded proof mass enclosed in a fluid-filled case, the fluid providing needed damping. A variable reluctance pickoff, whose armature is mounted on the proof mass, provides the output signal. A typical accelerometer has an output of ±3 V rms, 400 Hz and is powered by 115 V, 400 Hz.

c. Angle of Attack Sensors:

Most frequently the angle of attack of an aircraft is measured by a pivoted vane projecting into the airstream whose axis of rotation is horizontal, and normal to the centerline of the aircraft. Pickoff is usually accomplished by a potentiometer, synchro, or digital encoder.

Angle of attack also can be sensed by means of static parts located above and below the fuselage or above and below a probe projecting into the airstream, differencing their pressures and correcting for static defects.

d. Other Sensors:

In addition, sensors may be used to measure temperature (resistance thermometer), pressure (diaphragm or bellows with strain gage bridge or potentiometer pickoff), or position of control elements or forces on them. For the latter, synchros, potentiometers, strain gages, or inductive pickoffs may be used.

In some cases, null balancing systems such as shown in Fig. 5-128 may be used to provide readings of improved accuracy from the sensors.

5-12.6.3 Actuation

Electromechanical and hydraulic actuators are commonly used in flight control equipment. Electromechanical actuators are used where the torques to

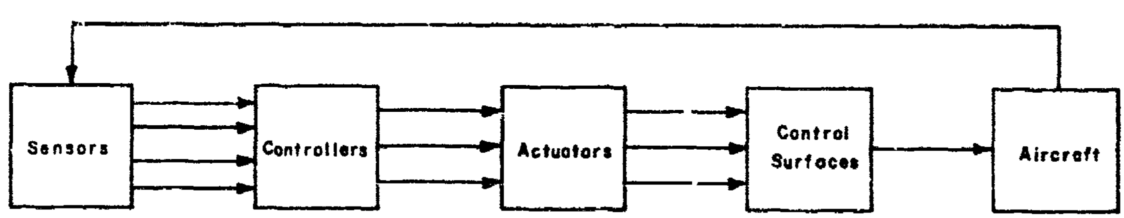


Figure 5-126. Flight Control System Overall Block Diagram

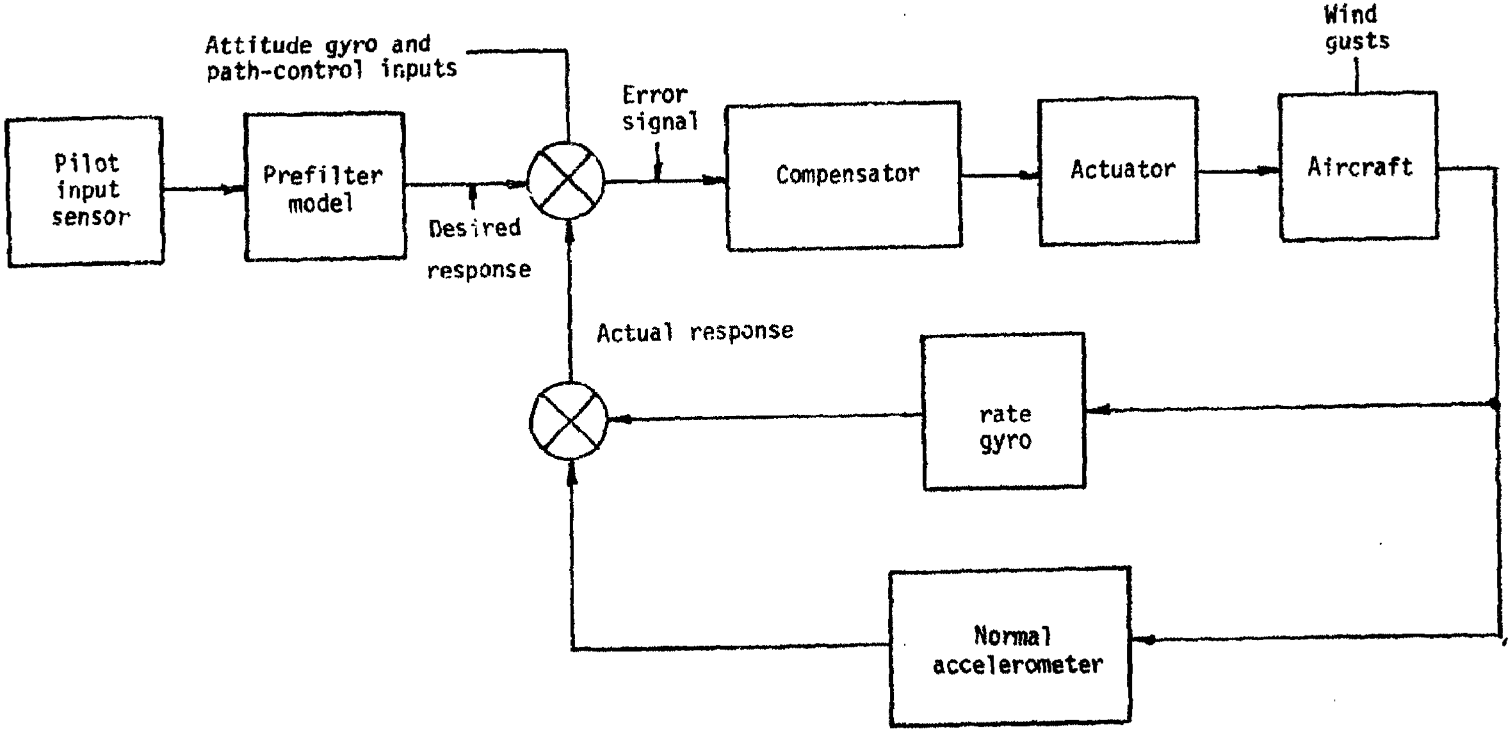


Figure 5-127. Typical Pitch-axis Augmentation Function (Ref. 67)

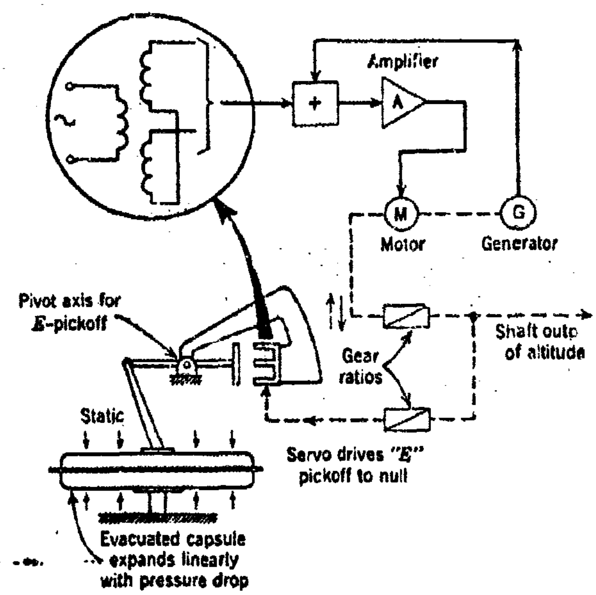


Figure 5-128 Displacement Follow-up Absolute-pressure Transducer

be supplied are not high and a moderate frequency bandwidth is adequate. Hydraulic actuators are used where high levels of force or a wide frequency bandwidth is necessary to effect satisfactory control.

5-12.6.4 Emission

Flight control equipment is not likely to be a significant source of emission. Since direct and alternating (400 Hz) power is supplied to both sensors and actuators, these and the associated wirings can produce local magnetic fields due to switching transients or alternating currents. Switching transients may be of special concern from circuits controlling hydraulic valves with solenoids.

5-12.6.5 Susceptibility

Actuators are unlikely to be directly susceptible to local fields since control voltage levels can be quite high. However, this may not be true of some of the sensor circuits.

The flight control subsystem has elements throughout the entire vehicle; therefore information is transmitted over long cables into which signals may be conducted or radiated. Transients that result when switches close, heaters turn on, power subsystem loads change, etc., can be propagated into the flight control subsystem and appear as a signal. Such a transient actuates the subsystem just as a signal from a gyro or an accelerometer would, and a spurious motion of the control surfaces can result in an uncoordinated maneuver of the vehicle; accordingly, interference in a sensor circuit can have serious consequences. Evaluation of the need for reduction of susceptibility in such circuits will depend upon the level of the sensor voltages and the sensitivity of the control system to swell variations of the sensor voltages. Further discussion of sensor systems is given in par. 5-15.

5-12.6.6 Control Techniques

The control of compatibility of flight control systems in the aircraft requires the application of more or less standard techniques of cable separation, grounding, bonding, shielding, and filtering. Sensor outputs should be balanced twisted pairs wherever possible so that common-mode voltages due to ground plane currents do not produce spurious sensor output voltages. Devices producing magnetic fields (actuators or sensor control motors) should be separated physically from sensitive devices or shielded (if the added weight is acceptable). Circuits containing solenoids or relay coils should be isolated physically and filtered to reduce transients.

5-12.6.6.1 Use of Fiber Optics

In order to reduce EMC problems and weight associated with multiconductor cabling, especially on aircraft, fiber optic systems have been developed (Ref. 69). One hundred foot fiber-optic cables were used to connect an elevator actuator with its control computer on an otherwise "fly-by-wire" system. More discussion of fiber optics systems is given in par. 7-4.4.

5-13 AEROSPACE GROUND EQUIPMENT (AGE)

5-13.1 INTRODUCTION

AGE, sometimes called ground-support equipment (GSE), consists of diverse equipments which must be used to check out the function of aircraft and complex missile systems on the ground. The type of equipment may range from small battery-operated test sets designed to check out a single function through truck or van mounted systems to permanently installed systems.

A typical list of aircraft subsystems which may be checked by AGE includes:

a. Armament
b. Control surfaces
c. Instruments
d. Engine instruments
e. Flight instruments
f. Heating, ventilating, and deicing
g. Ignition
h. Engine control
i. Lighting
j. Electrical power
k. Radio
l. Radar
m. Special electronics, IR, navigation, etc.
n. Warning and emergency.

AGE may be used in the field, on flight lines, and at base shops as well as at formal test and rework facilities.

Preflight AGE must be rugged, portable, largely self-contained, and self-checking. It may be capable of operating from its own power sources, sources of the equipment under test, or on locally available power. Shop testing equipment may be fixed or portable, and is in general more complex. For example, shop testing equipment may be computer directed.

AGE designers traditionally use as much "off-the-shelf" military and commercial ground equipment as possible because usually it is produced in limited quantities, and does not require airborne type reliability. Thus, it may incorporate standard trucks, vans, power supplies and regulators, batteries, meters, switches, amplifiers, receivers, transmitters, and instrumentation.

Despite the extensive use of standard equipment, AGE is unique in two respects: (1) it must interface with the specific subsystem to be tested (mechanically and electronically) and then suitably stimulate and measure the characteristics of the equipments under test, and (2) the AGE equipment, both standard and special, must be so packaged that it will be functional in the sometimes severe environment within which it must operate.

5-13.2 EMISSION CHARACTERISTICS

A substantial portion of the emissions from AGE comes from the standard equipment incorporated in the test system. Motors, generators, vehicle ignition, and power control switches all contribute to interference generation. See pars. 3-1.3, 5-1, 5-3, 5-6 for specific emission characteristics.

Special problems can arise from the cabling used with the equipment and from the special stimuli used to check out the designated subsystems. AGE often is connected to the aircraft or missile and to base power by lengthy and complex cable systems.

The special signals generated to stimulate equipment under test may be radiated or conducted. If radiated, the levels are usually at low levels and should produce no special problems. If conducted, they also will be at low levels, but where the cables convey a large number of such signals, interaction within the cable is possible but can be avoided with care in selection of cables and allocation of signals to conductors within the cable.

5-13.3 SUSCEPTIBILITY

If the equipment is connected to base power, large transients may appear on the lines. Furthermore, if the equipment under test is grounded, the transients or power frequency harmonics can produce significant ground loop currents which can couple either conductively or inductively into sensitive test circuits.

Furthermore, in a typical location there may exist high levels of field strength due to nearby radars, communication transmitters, welders, and other sources.

For many equipments the test signals will be of moderate level and not exhibit susceptibility under normal conditions. However, the use of long cables into which substantial voltages and currents can be induced by high fields requires a careful evaluation in each instance. Therefore, each item of equipment should be examined with respect to its cabinet susceptibility and all signal and power leads connected to it.

In addition to the test equipment, the characteristics of the equipment under test and its configuration should be examined to determine if the method of connection to AGE causes any special problems. In particular, the connection of a cable to an aircraft, either through a connector mounted on the outside of the aircraft or to one inside, can introduce voltages or fields internally that would not exist otherwise. As mentioned previously, common-mode currents could be troublesome, especially if the aircraft itself has a ground connection.

5-13.4 EMC CONTROL

In AGE EMC design decisions, cost-effectiveness should be a consideration. Where the production volume is small, it may be advantageous to over-design rather than to carry out an extensive EMC analysis.

Some account should be taken of where the equipment involved is to be mounted, such as within a permanent enclosure, on a vehicle, or exposed.

In addition to the use of the techniques of shielding, filtering, and grounding for EMC control, as discussed in Chapter 4, special attention should be given to the selection of the cables used in the design stage.

To avoid the effects of long cables, balanced output and input circuits should be used wherever possible and filters should be inserted in these circuits to reject signals other than those desired. Where high sensitivity circuits are involved, signals should be amplified before transmission through the cable wherever possible.

Where the test equipment is located inside a truck or van, the equipment itself may be protected from high levels of environmental fields by the van enclosures or specially constructed shielding enclosures. The usual practices of bonding equipment cabinets and racks should be followed, and any external connections to power supplies should be filtered where the lines enter the vehicle.

To protect test cables from external high-frequency fields, it may be necessary to use shielded cables. The shield probably will require a good connection to the vehicle under test and also to the test equipment on the vehicle or enclosure where it is located.

A specification for AGE for NASA installations is available (see Ref. 70).

Two matters concerning safety should be examined. One is the personnel shock and hazards associated with the interconnection of separate vehicles or test equipment by means of cables. Isolation transformers may be advisable where base power is used. The other is protection against lightning wherever the AGE is exposed. The following specifications and standards should be consulted:

a. MIL-STD-833 *Minimization of Hazards of Electromagnetic Radiation to Electroexplosive Devices*

b. MIL-STD-1377 *Effectiveness of Cable, Connector, and Weapon Enclosure Shielding and Filters in Precluding Hazards of Electromagnetic Radiation to Ordnance, Measurement of*

c. MIL-STD-1385 *Preclusion of Ordnance Hazards of Electromagnetic Fields; General Requirements for*

d. MIL-STD-1542 *Electromagnetic Compatibility (EMC) and Grounding Requirements for Space System Facilities*

e. MS-25384 *Plug, Fuel Nozzle, Grounding*

f. MS-33645 *Receptacle Installation, Fuel Nozzle Jumper, Aircraft*

g. MS-90298 *Connector, Receptacle, Electric, Grounding*

5-14 SPECIAL CIRCUIT CONSIDERATIONS

5-14.1 INTEGRATED CIRCUITS

The advent of semiconductor devices has enabled the construction of circuits having very high densities of circuit elements, including resistors, capacitors, diodes, and transistors. The high densities are particularly characteristic of digital circuits. While analog circuits can be made with the same densities in most applications, the number of elements required on a single chip is much less.

The analog circuit can be more susceptible to internal EMC problems than the digital circuit because of the switching threshold which is characteristic of the digital circuit. However, with fast rise time circuits considerable care in design must be exercised (see par. 5-14.3). The internal EMC problems arise from (1) the close spacing of components and their connections, (2) the parasitic or unwanted elements that appear in the circuit as a result of particular methods of construction used, and (3) the need for terminal connectors of small size to match the circuit board sizes.

5-14.1.1 Emission

Because of their small size and consequent low power requirements, integrated circuits themselves are unlikely to be significant sources of emission. However, interconnections between circuit mounting boards can be sources of both conducted and radiated EMI unless adequate filtering and shielding is used. The techniques described in detail in Chapter 4 apply to these circuits.

5-14.1.2 Susceptibility

Because of the small size of integrated circuit packages, the primary cause of radiated susceptibility arises from leads connected to them. Fig. 5-129 (Ref. 71) shows predicted electric field susceptibility for a typical amplifier. In this case linear susceptibility was defined as the field level necessary to produce an output voltage equal to that due to internal circuit noise. Nonlinear susceptibility is that field necessary to cause sufficient rectification to displace the operating point of the circuit by 1% from the normal value. In the calculations the area, occupied by the circuit and leads, was assumed to have a length of 0.5 in. and a width of 0.2 in. This configuration is modeled as an electric dipole at low frequencies and a fixed area at frequencies above about 1 GHz. At frequencies in this range the susceptibility is assumed to be due to nonlinear phenomena. For larger areas such as arising from wiring on circuit boards the levels of susceptibility could be as much as 40 dB less.

For digital circuits a noise immunity threshold of 0.5 V was assumed. In this case the susceptibility threshold for electric fields is shown in Fig. 5-130 for effective areas of 1 cm^2, 5 cm^2, and 25 cm^2.

Susceptibility of integrated circuits to conducted EMI has been measured in the gigahertz frequency range (Ref. 72). The data show that at 0.22 GHz analog circuits begin to show susceptibility at 10^{-5} W and this increases to about 10^{-1} W at 9.1 GHz. Digital circuit susceptibility may be two orders of magnitude higher, at least at 0.22 GHz.

Because of their small size, integrated circuits can be particularly susceptible to destruction from high levels of EMI or to static discharges as a result of handling. In some cases the circuits themselves contain self-protecting diodes to bypass high currents or charges around the active transistors (Ref. 73).

5-14.2 ANALOG CIRCUITS AND DEVICES

Analog circuits include those carrying signals of any kind where a parameter of the signal such as amplitude, frequency, or phase varies in direct proportion to the value of the quantity being transmitted. Analog circuits may operate at low audio (including dc) to provide control information, at audio frequencies for speech transmission, or at radio frequencies where modulation and demodulation are necessary. In all cases, superimposed interference will provide a distortion of the originally transmitted message signal to an extent which is directly dependent on the interference-to-signal ratio (I/S).

There is no simple universal way of measuring the signal degradation due to such interference for all analog signals. In some cases the significant degradation effect can be directly proportional to the level of interference. In others, as with frequency modulation signals, the degradation may exhibit a sharp threshold, in which values of I/S up to a value close to unity exhibit very little degradation, but beyond that there is almost total loss of information. These phenomena are discussed in some detail in par. 3-2.

Analog circuits may operate at levels varying from a few nanovolts to hundreds or thousands of volts. At the lowest levels (below about 1 mV) they can be expected to be extremely susceptible to external fields, whereas at higher levels (above about 1 V) they will be relatively insensitive and more likely to be sources of interference to other circuits. The mechanisms of emission and susceptibility have been discussed in some detail in Chapter 3 and are not discussed further here.

5-14.2.1 Electronic Instruments

Analog instrumentation includes devices used for displaying the value of some quantity as registered by a proportional displacement of some type of indicator, usually on a D'Arsonval type meter or a cathode-ray tube.

5-14.2.1.1 Emission

Instruments are not likely to be significant emitters except as follows:

a. Cathode-ray displays may use high magnetic fields for beam deflection. By its nature the deflection yoke has an air core and exhibits high leakage flux. The frequency of operation is usually below a few kilohertz, so that there can be coupling to audio frequency circuits in the vicinity.

b. Devices which contain an internal generator can produce significant conducted or radiated emissions. Typical is the sweep oscillator used in cathode ray oscilloscopes, the frequency reference oscillator used in frequency counters, and the local oscillator used for frequency conversion in transmitters or receivers. Transformers and high inductance coils or chokes also may radiate magnetic fields as may any solenoids or motor drives.

5-14.2.1.2 Susceptibility

5-14.2.1.2.1 Input and Signal Circuits

Usually, the input circuit is the most susceptible part of any instrument because it operates at the lowest level. Standard methods of controlling susceptibility include use of balanced circuits at low frequencies and coaxial systems at high frequencies

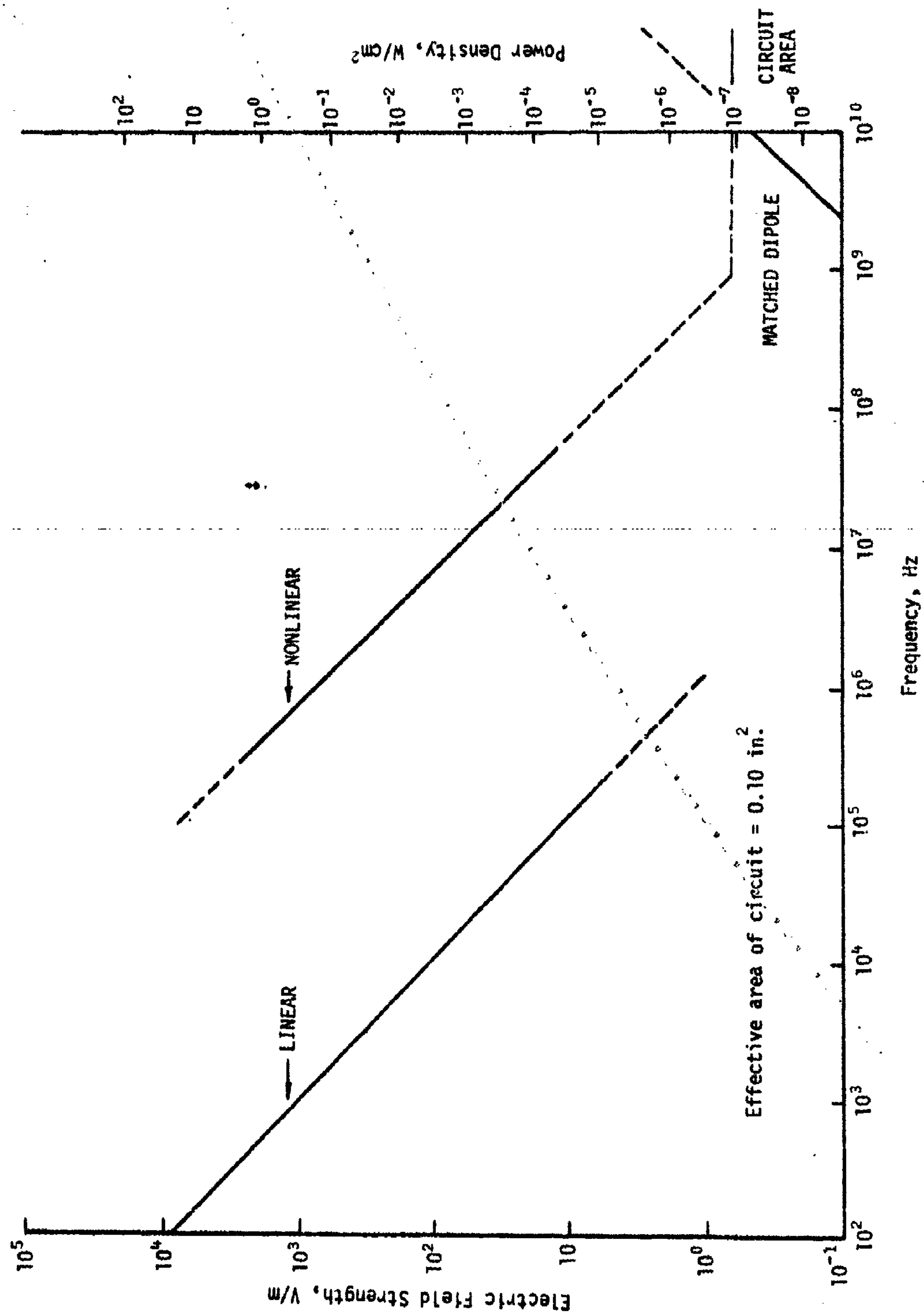


Figure 5-129. Amplifier Integrated Circuit CW Susceptibility

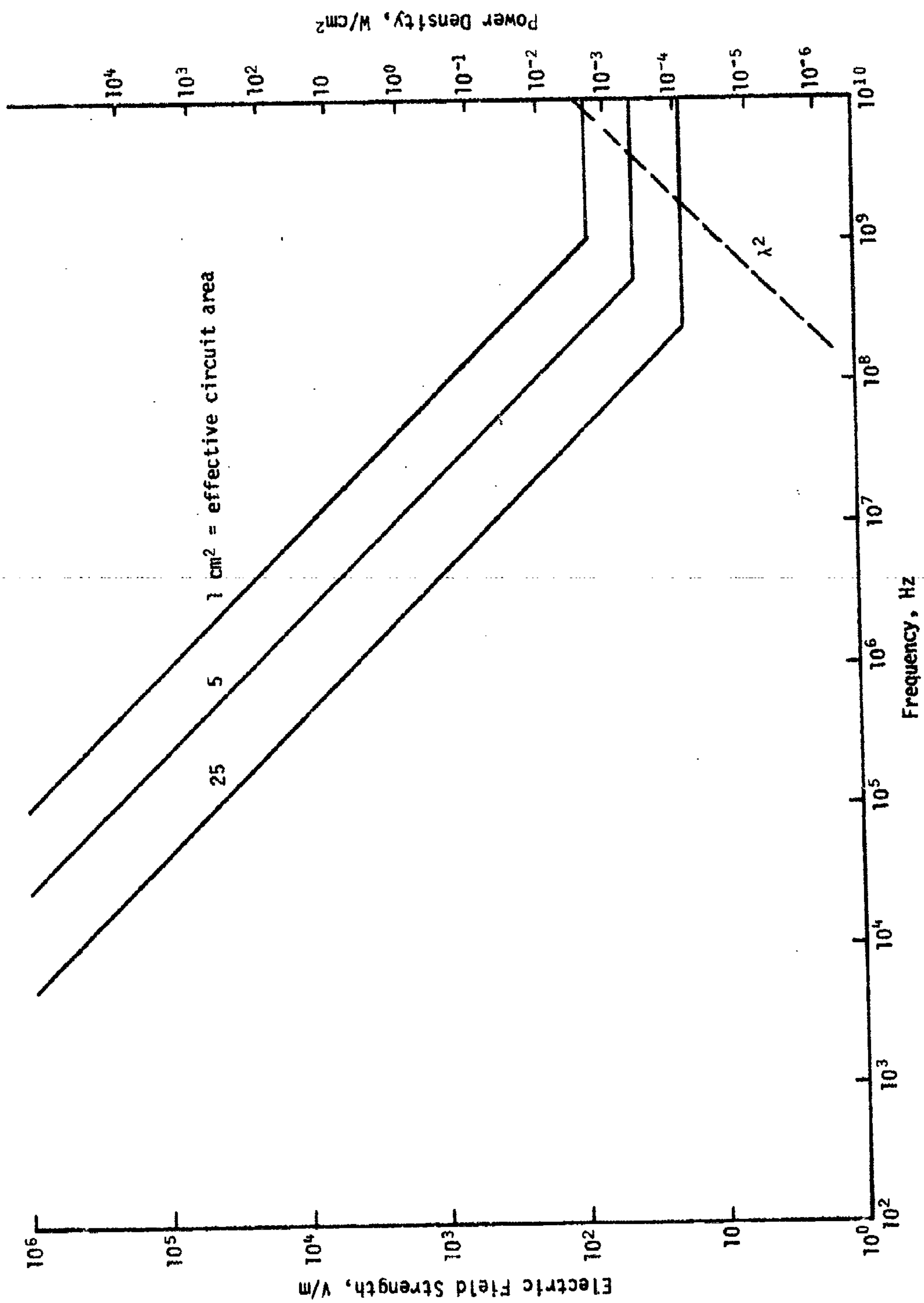


Figure 5-130. Digital Integrated Circuit CW Susceptibility (Predicted)

using small effective loop areas, and shielding. Special care is required for circuits connected to long input cables to avoid coupling to external fields and where high accuracy is involved.

5-14.2.1.2.2 Other Circuits

If the input circuits are properly protected, other parts of the instrument may have significant levels of susceptibility. Power and control cables can provide admission paths for conducted interference as can apertures in the cabinet for radiated interference. If the instrument is to be used in high field regions such as near electromagnets or radar systems, indicating meters should have special shielded and filtered construction. Also, cathode ray tubes may require magnetic shields to protect against the earth's field or local magnetic fields.

For industrial process instruments, the peak-to-peak ripple and total noise level in any circuit should be less than 0.25% of the maximum signal or 10 mV for a 5-V signal (Ref. 74).

5-14.2.2 Synchros

Synchros, which are extensively used for transmitting position or control information, operate at relatively high voltages, but because they operate on a null principle, any undesired coupling into the null circuit will limit the accuracy of the indication obtained. The tolerable level can be calculated in terms of the sensitivity of the synchros (approximately 1 V/deg) and the accuracy desired. In addition, the operation is dependent upon the reference circuit, usually the 60- or 400-Hz power supply. Substantial levels of harmonic voltages appearing on this circuit can introduce significant errors. Where power supply waveform purity cannot be maintained, a special power supply should be used to supply the reference circuit.

5-14.2.3 Analog Computers

Computer circuits require special consideration because of their accuracy requirements. Normally, they are designed to operate with as large a signal range as possible and still maintain linearity. By using feedback, such as in the well-known operational amplifier, the signal range can be maximized while satisfying linearity requirements. Since the range may be several volts, an induced level of a few millivolts can cause significant error. While this is not a highly sensitive level, if the computer must operate in high field regions, special protection means may be required.

5-14.3 DIGITAL DATA SYSTEMS

5-14.3.1 Introduction

Digital data techniques are used for communication, data processing, indicator, and control systems. Frequently, a digital data processing unit functions as a central control and interfacing unit for all radio data link, navigation, flight stabilization target detection, acquisition, tracking, and weapon control systems. Modern requirements for high data rates lead to systems having short pulse duration and wide bandwidth which are subject to degradation from interference on power, control, or signal leads. Digital data equipments, which contain large numbers of solid-state switching circuits, are also capable of emitting conducted or radiated interference. The emission and susceptibility level of digital data equipments depends upon design, construction, and installation.

Binary circuits are capable of restoring pulses or levels that have become partially degraded in level or shape by EMI. In such circuits, it is necessary only for the digital data information level to reach a value sufficient to satisfy a switching threshold or comparator circuit, and the pulse or level will be restored thereby to a full-amplitude, relatively noise-free condition. However, if EMI exceeds a certain threshold, it can be quantized into a full-level bit error.

5-14.3.2 Emission

The predominant interference generated by digital equipment can be attributed to the repetitive operation of a multitude of switching circuits having fast rise times, whose switching operations are synchronized by clock-timing logic. The clock frequency and its harmonics dominate the generated EMI spectrum.

Other major interference sources are the operations involving the basic oscillator, time pulse distributors, register counters, drums, and any other fixed repetitive functions.

Interference can be expected at the fundamental and harmonic frequencies of these operations during normal computer functioning. Other computer operations may become sources of interference under execution of certain instruction sequences in some programs. The program cycle itself, if sufficiently short in duration, falls into this category.

In one series of tests on a computer using solid-state logic (Ref. 75), the computer was divided into five test groups:

a. Master clock timing and basic operations

b. Central processing system and programmed operations

c. Core storage element

d. Drum system and associated peripheral unit processing circuits

e. Typewriter, card reader, and associated peripheral unit processing circuits.

For each of these groups the antenna was positioned 3 ft from the computer frame. Results are shown in Fig. 5-131. The values indicated are the maximum for each frequency. The specific frequencies of emission correlate well with computer timing frequencies.

5-14.3.3 Susceptibility

The noise margin and noise immunity parameters of gating circuits used in digital devices are measures of the gate susceptibility to noise signals. Noise margin is defined as the magnitude in volts of pulse noise, which, when appearing at the input of a digital gate and riding on the worst-case logic level, will cause a significant reaction at the output of the gate. Dc noise margin, as in wide, slow rise time noise pulses, is the difference between the worst-case logic voltage level and the worst-case switching threshold voltage of the gate. Transients of ac noise margin, as in fast rise time noise pulses, are in most cases less than the dc noise margin. Noise immunity is defined as

$$\text{Noise immunity} = \frac{\text{worst case noise margin}}{\text{maximum logic voltage swing}} \times 100\% \qquad (5\text{-}41)$$

5-14.3.3.1 Susceptibility Mechanisms

Digital circuit susceptibility coupling mechanisms include:

a. Induction Coupling:

Induction coupling in a digital circuit may occur in two ways: (1) internally within the device itself, or (2) externally via cables and connectors of associated circuits.

Internal coupling occurs as a consequence of parasitic mutual capacitance and inductance inherent in the semiconductor device. The effective isolation between input and output of semiconductor devices (the circuit elements most commonly used in digital data systems) may be poor, permitting undesirable coupling and crosstalk between two gates in the same integrated circuit package. Where there are unused terminals, a decision must be made as to whether they can be allowed to float or whether they should be tied to some bias voltage or to ground.

External coupling is due to mutual capacitance and inductance of interconnecting lines and conductors. The effect is to cause differentiated leading and trailing edges of input signals to appear on the output (or other inputs) of a gate even though the gate inputs did not satisfy the switching logic level. A coupled signal of sufficient magnitude and width may be amplified and shaped by a string of gates so that it results in a logic signal that causes an error.

Of special concern are magnetic materials used as information elements in digital data processing systems. Magnetic tapes, discs, or drums may be used as input/output or storage devices; and magnetic toroids are frequently used in shift register, decoder, buffer storage, and memory circuits. These magnetic materials may be easily influenced by extraneous magnetic fields so as to erase the data stored or it can bias the magnetic materials toward an "all ones" or "all zeros" state. The portions of digital data processing systems that use magnetic materials must therefore be protected from internal and external fields.

b. Conduction Coupling:

Large switching transients in the power and ground circuits of digital circuits may be directly coupled into sensitive digital circuits. Because a number of circuits may be packaged to form one module, power supply decoupling (see par. 4-3.1.1) may be necessary within the module. When the circuits are grouped into modules and each module contains power supply decoupling, the currents to be evaluated for EMI are:

(1) Ground currents that leave the module and flow into the system ground

(2) Signal currents that flow in decoupling capacitors mounted in the module.

High speed switching circuits may place a large transient current demand upon power supplies during transition from one logic state to the other. Knowledge of the rate of change of these currents will allow the system designer to calculate, approximately, the tolerable impedance of the bypass capacitor and the system ground.

The path from the module to system ground must be a low impedance. If the module is a printed circuit card, it is almost always necessary to use more than one of the connector pins as a ground return. The common ground return for groups of modules also should be designed for low impedance.

Susceptibility of a digital circuit to disturbances in the power and ground systems may be determined by introducing a disturbing pulse between the ground pin of the module and the system ground, or between the power pin and the system power. If the logic circuit under test is one of a string of gates, the width and amplitude of the disturbing pulse may be varied to determine the minimum pulse that will result in the propagation of an erroneous logic signal.

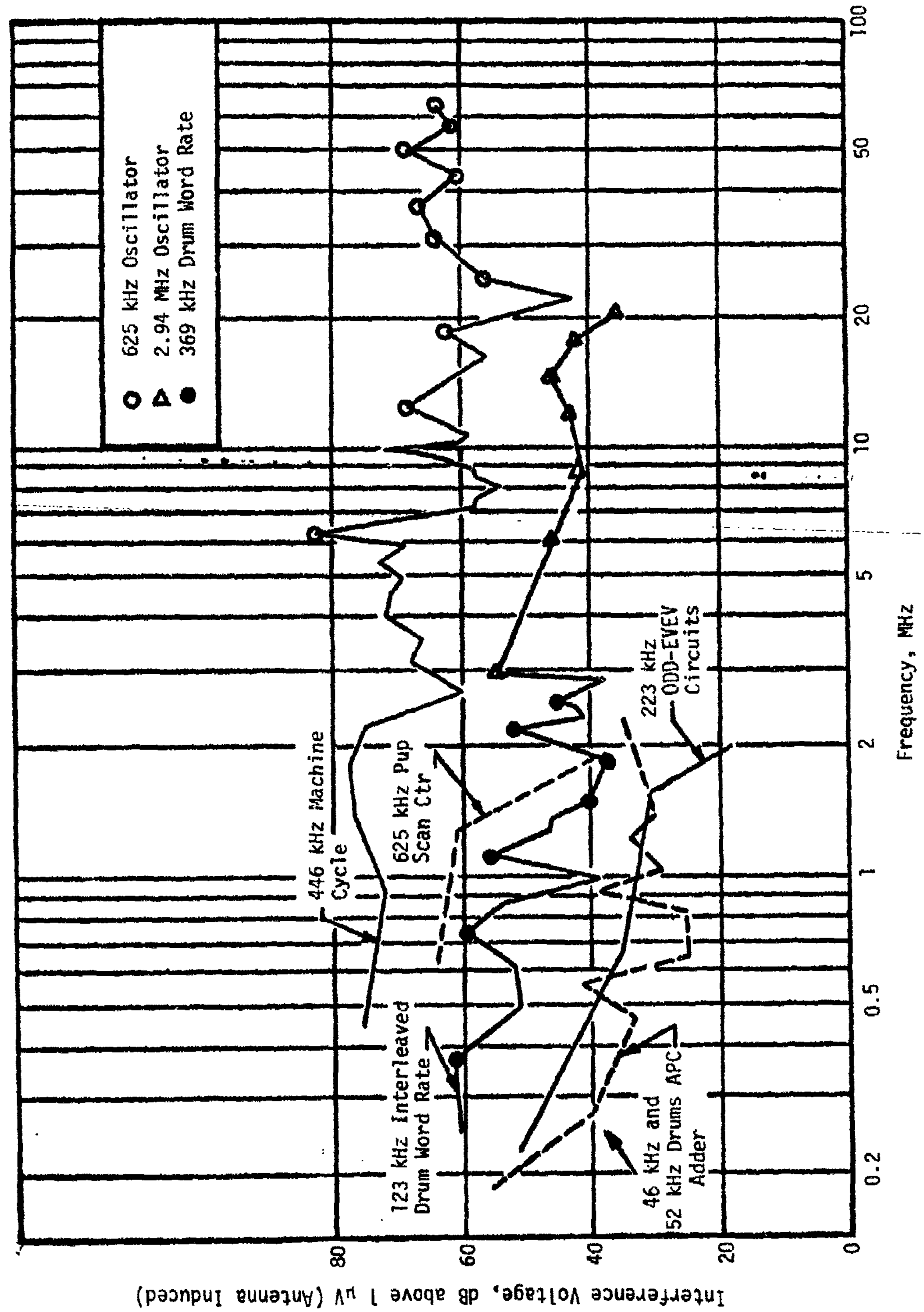


Figure 5-131. Composite CW Radiated Interference Spectra (Antenna 3 ft from each computer frame)

c. Radiation Coupling. Strong fields from nearby intentional emitters — such as radio stations, or industrial, scientific, or medical equipment — can penetrate shields of cables or equipment housings. If they are of sufficient strength, such fields may cause interference either by direct interference with circuit signals (if their frequencies are sufficiently close to clock and switching frequencies) or by causing bias voltage changes due to rectification in active circuits.

5-14.3.3.2 Measured Susceptibility Levels

A discussion of measured susceptibility follows:

a. Vacuum Tube Computer. A computer was exposed to the radiation from a radar simulator (Ref. 76). This system had about 50,000 vacuum tubes and 170,000 diodes. Tests were performed in the frequency range of 450 to 2900 MHz with various combinations of pulse duty cycle and signal polarization, and exposures were taken on each unit of equipment and front, back, and side. Results showed that:

(1) Many circuits of the computer do not malfunction under electromagnetic field radiation as high as 400 V/m. Circuits in this category are high-level circuits such as flip-flops, gates, and/or pulse amplifiers, relay drivers, level inverters, and cathode followers.

(2) Low-level circuits of the computers (whose normal input is within the range of 50 mV to 2 V peak to peak) did malfunction when subjected to moderate field strengths.

(3) Sense amplifiers of the memory element malfunctioned at the field strength of 16.4 V/m peak, the tuning fork oscillator of the output section malfunctioned at 40.8 V/m peak, and drum read amplifiers of the drum section malfunctioned at 44.8 V/m peak.

(4) Data conversion receivers of the input section malfunctioned at 52.1 V/m peak.

(5) Flux amplifier of the output section failed at 98.7 V/m peak.

(6) Generally, circuits are more susceptible to radiation near the low end of the 450-2,900 MHz frequency band than to the higher frequencies in this band.

(7) Variations in susceptibility due to variations in pulse width and PRF are slight.

(8) Polarization of the signal is significant, but varies widely by computer unit. Generally, the computer is more susceptible to vertically polarized signals.

In actual usage, the susceptibility data levels of malfunction must be identified with the specific parameters which characterize the radar. The sense amplifier of the memory element was determined to be the circuit most sensitive to electromagnetic radiation.

b. Solid-state Computer (Ref. 75):

A solid-state computer was also subjected to pulse-modulated high-frequency radar signals produced by the portable radar simulator.

The resulting computer susceptibility levels varied from 64 V/m (peak) to over 400 V/m peak. The 64 V/m (peak) susceptibility level occurred in the peripheral unit processor. This unit along with the central processing system yielded the most susceptible areas. These units predominantly contain basic logic circuits which have high frequency characteristics due to their design for fast rise and response times. Consequently, they are more susceptible to the high frequency characteristics of the radar pulses than other circuits.

The core storage and drum units, which house special circuits, including core and drum information amplifiers (both of which were most susceptible in the vacuum tube computer), were not as susceptible as circuits in the central processing system and the peripheral unit processor. The core and drum amplifiers amplify low-frequency signals and therefore differ in design from high-speed pulse circuits. As a result, they do not so readily accept the radar pulses. Also it was noted that for all units, susceptibility decreased as frequency increased.

Of the four units tested, the drum unit was the least susceptible, being below 400 V/m (peak) at only one test frequency. An important factor contributing to this effect is the method of packaging. This unit, although not RF shielded, was designed in such a manner that all circuits are drawer-mounted and housed in a metallic enclosure. In one instance, one of the drum circuit drawers was pulled out and the susceptible level decreased from 240 V/m (peak) to 145 V/m (peak).

5-14.3.4 Interference Reduction

As with other types of equipments the effects of radiation on digital equipment can be reduced, in some cases, by shielding the susceptible circuits, or by redesign.

Typical of the circuit changes which can reduce the susceptibility of sensitive circuits are:

a. RF filtering of the interference at the circuit input

b. Reduction of the circuit rectification characteristics

c. AF filtering of the rectified interference pulses.

In typical digital data equipments, some circuits have an inherent immunity to malfunction under intense electromagnetic influence. Relatively invulnerable circuits include high-level circuits such as flip-flops, line drivers, relay drivers, level inverters, and emitter followers. On the other hand, low-level circuits whose normal input is within the range of 50 mV to 2 V peak to peak, tend to malfunction when subjected to interference of moderate field intensities. Circuits in this category include tape readers and magnetic core memory-sensing amplifiers. Digital data systems employing radio data links are most susceptible to degradation in the radio circuit itself rather than in the logic circuits that follow the quantization process. The EMI control measures discussed in par. 5-7 for receivers apply to digital data receivers.

In comparing vacuum tube circuits with solid-state circuits, an important contribution to the reduction of the drum information amplifier susceptibility in the solid-state computer was a circuit change replacing the old input transformer with two stages of push-pull amplification in order to eliminate in-phase noise. Susceptibility reduction in the memory core information amplifier was achieved by a drastic reduction in length of conductors in the high gain differential amplifier stage. This reduction was made possible by using transistors and subsequently compacting circuits.

However, solid-state basic circuits are more susceptible than the vacuum tube type. The vacuum tube basic circuits were not susceptible at field strengths of 400 V/m (peak) or better; however, levels below 100 V/m (peak) were found to cause malfunctions in the solid-state circuits. Two important factors contribute to this phenomenon. One is the better high frequency handling capability of the new circuits. The other is the 10 to 1 reduction in the operating signal level of the circuits, thereby causing a reduction in the ratio of signal level to radar level for any given radar environment. Fig. 5-132 shows comparative data.

5-14.3.5 Logic Design

In addition to other techniques of controlling susceptibility, special logical design techniques can be used to reduce the susceptibility to impulse type interference, especially if it occurs infrequently, namely:

a. Parity error check. Data words are transmitted with a known parity which is checked on reception. The more sophisticated systems can correct for simple errors.

b. Redundancy. Each word is sent twice, either in series or parallel channels.

c. Gating. The circuit status is read only over a short period of time (the gate period) during each operation cycle. In this way, spurious signals due to interference are prevented from causing errors unless they occur simultaneously with the gate.

5-14.4 MICROWAVE CIRCUITS

Microwave circuits have only a limited number of significant EMC characteristics. This is because microwave energy is propagated over only limited distances within equipment cabinets. It is coupled to an antenna by means of coaxial cable or waveguide using as short a length as possible considering the application.

In the design of microwave systems, one makes use of several circuit components not in common use at lower frequencies. Included are:

a. Directional couplers
b. Hybrid junctions
c. Isolators
d. TR switches
e. Circulators.

One of the most significant functions of these elements is to permit the use of a single antenna both for transmission and reception. Their characteristics are important in determining the overall performance of a system, especially with respect to frequencies out of the design band. Unfortunately, very little specific information on such characteristics is available in published form (Ref. 77). Where they are used, the number of interacting elements can be quite large and analysis is performed using matrix methods such as is common for treating a number of antennas having mutual coupling.

5-14.4.1 Emission

The only sources of emission at microwave frequencies are the devices used to generate and amplify microwave energy. High-power devices for use in radar systems were discussed in par. 5-9.

5-14.4.1.1 Tunnel Diode

Tunnel diode oscillators are able to deliver milliwatts of power and thus are used in microminiaturization programs. Both output amplitude and frequency are dependent on the nonlinearity of the diode V-I curve.

If the tunnel-diode (T-D) oscillator is fed to a mismatched load through a length of transmission line, discontinuities may appear in its tuning curve. There may be jumps or skips in frequency and power output. The effect is due to the formation of cusps or loops in the admittance curve seen by the negative

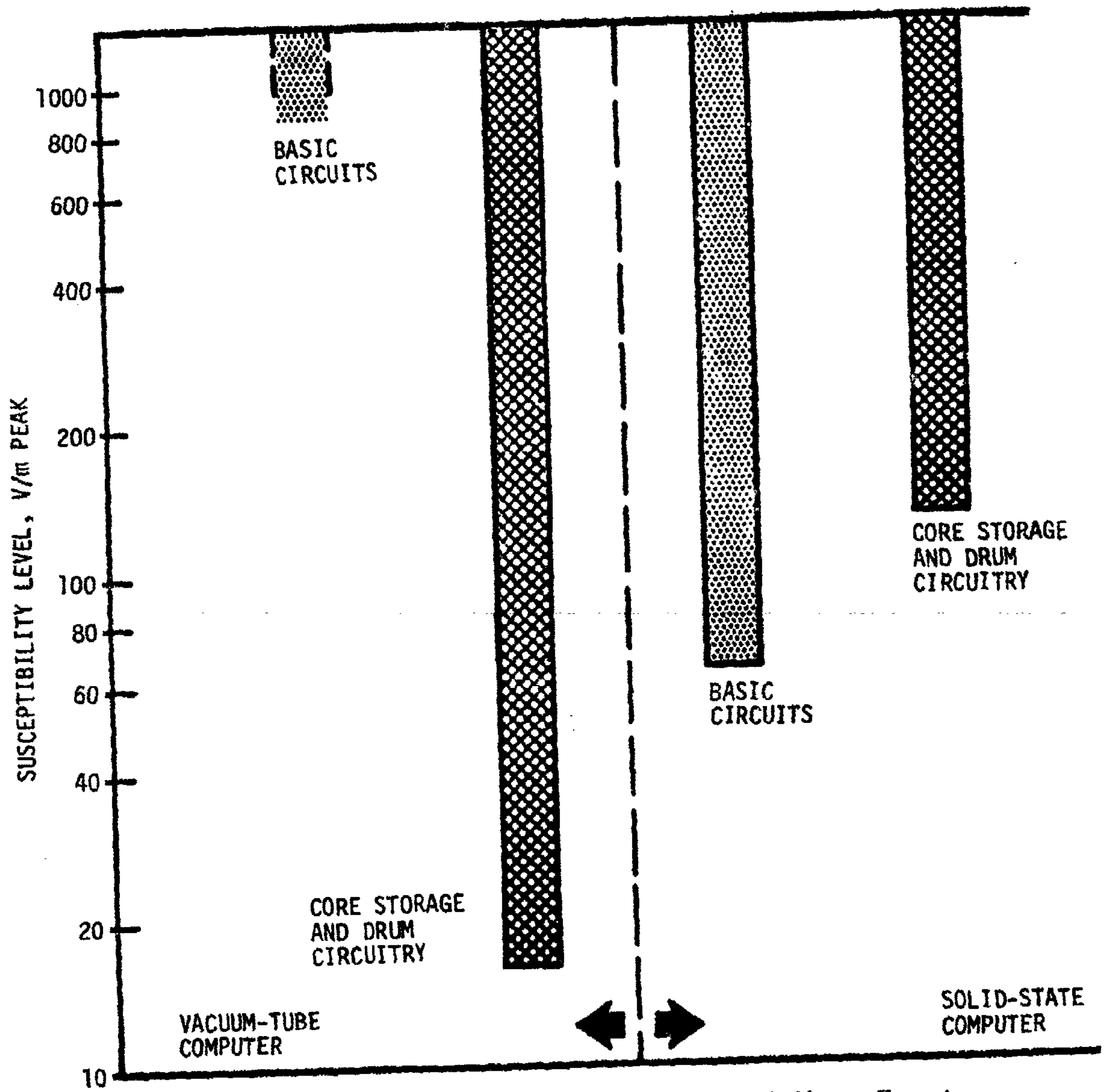


Figure 5-132. Susceptibility Comparison of Vacuum Tube and Solid-state Computer

conductance, and is called oscillation hysteresis. The effect is avoided if the load is perfectly matched to the characteristic impedance of the transmission line. The condition for no oscillator hysteresis is given by:

$$\sigma < \sqrt{1 + \frac{2Q}{\theta}}, \text{ dimensionless} \quad (5\text{-}42)$$

where

σ = VSWR of load, dimensionless

Q = resonance Q for line, dimensionless

θ = line length, rad

The oscillator frequency is sensitive to diode bias as shown in Fig. 5-133, while output power vs bias voltage is shown in Fig. 5-134. The second harmonic output vs bias is shown in Fig. 5-135. All other outputs were -60 dBm or lower.

Other types of small semiconductor oscillators have been developed including the Gunn and Read IMPATT diodes. Little specific EMI data on these devices are available, but general design information can be obtained from Ref. 78.

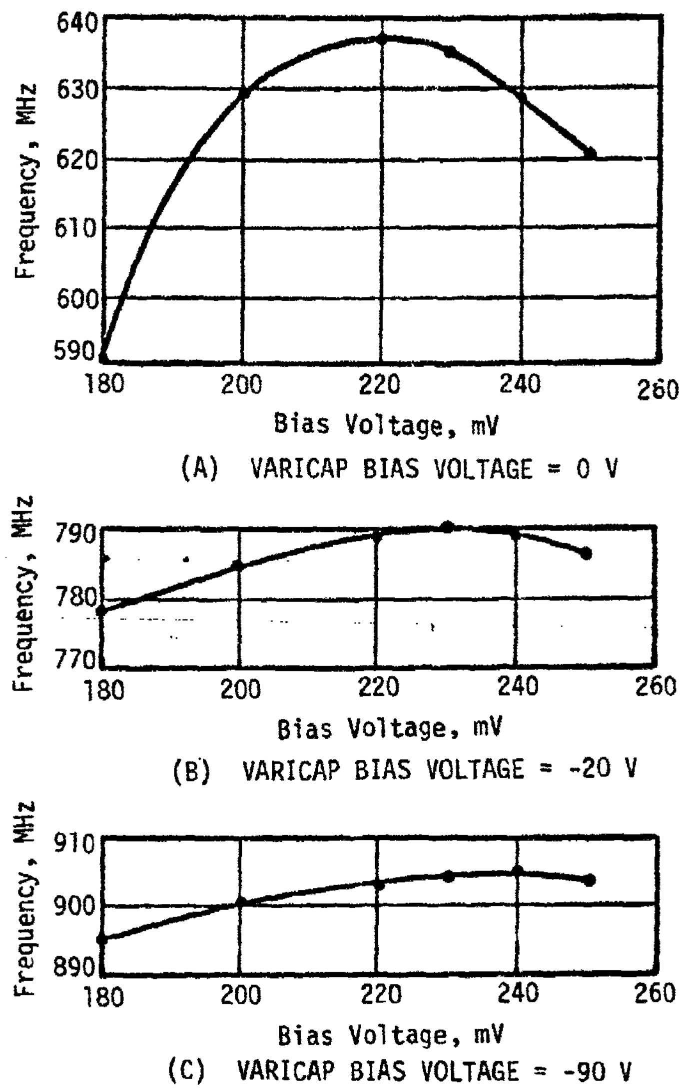


Figure 5-133. Fundamental Frequency vs Bias Voltage of Tunnel Diode for Tunnel-diode Oscillator

5-14.4.2 Susceptibility

Susceptibility at microwave frequencies takes place either in active devices used for amplification or in detectors, or through coupling between transmission line structures used for interconnecting circuit elements.

5-14.4.2.1 Masers

The maser provides an amplifier with an extremely low noise level. The traveling wave maser (TWM) has a higher stability than the cavity maser and is the type in most general use.

Interference analysis of traveling wave masers was carried out both experimentally and theoretically in Ref. 79. The conclusions of this study were:

a. Masers are the most sensitive amplifiers in all the frequency bands for which they have been developed. A noise temperature of 10 K is typical.

b. Saturation of the maser is predictable and is a function of average input power and maser gain as shown in Fig. 5-136.

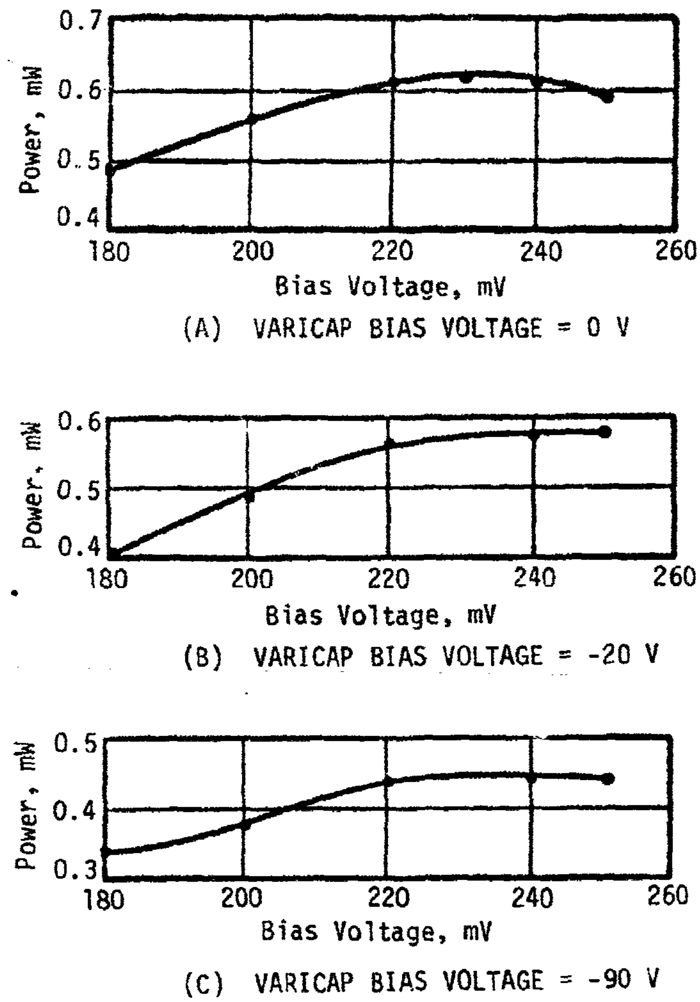


Figure 5-134. Total Output Power vs Bias Voltage of Tunnel Diode for Tunnel-diode Oscillator

c. Desensitization is predictable and is a function of average input power.

d. Cross-modulation effects are predictable on the basis of the input energy of the interfacing signal, the maser saturation characteristic, and the gain-recovery characteristic. When interfering signal power results in an input power greater than maser saturation level, then:

(1) If the interference amplitude modulation (AM) period is greater than maser recovery time, the amplitude of desired signal varies proportional to gain of maser.

(2) If the interference AM period is less than maser gain recovery time, no cross-modulation is observed, but gain desensitization is proportional to total average input power.

(3) If the interference AM period is approximately equal to gain-recovery time, then cross-modulation is a function of the ratio of the interference AM period to maser gain-recovery time.

e. Intermodulation is nonexistent based on the fact that no harmonic signal generation is predictable or measurable.

f. The gain stability of the maser is dependent on the stability of the externally applied dc magnetic field.

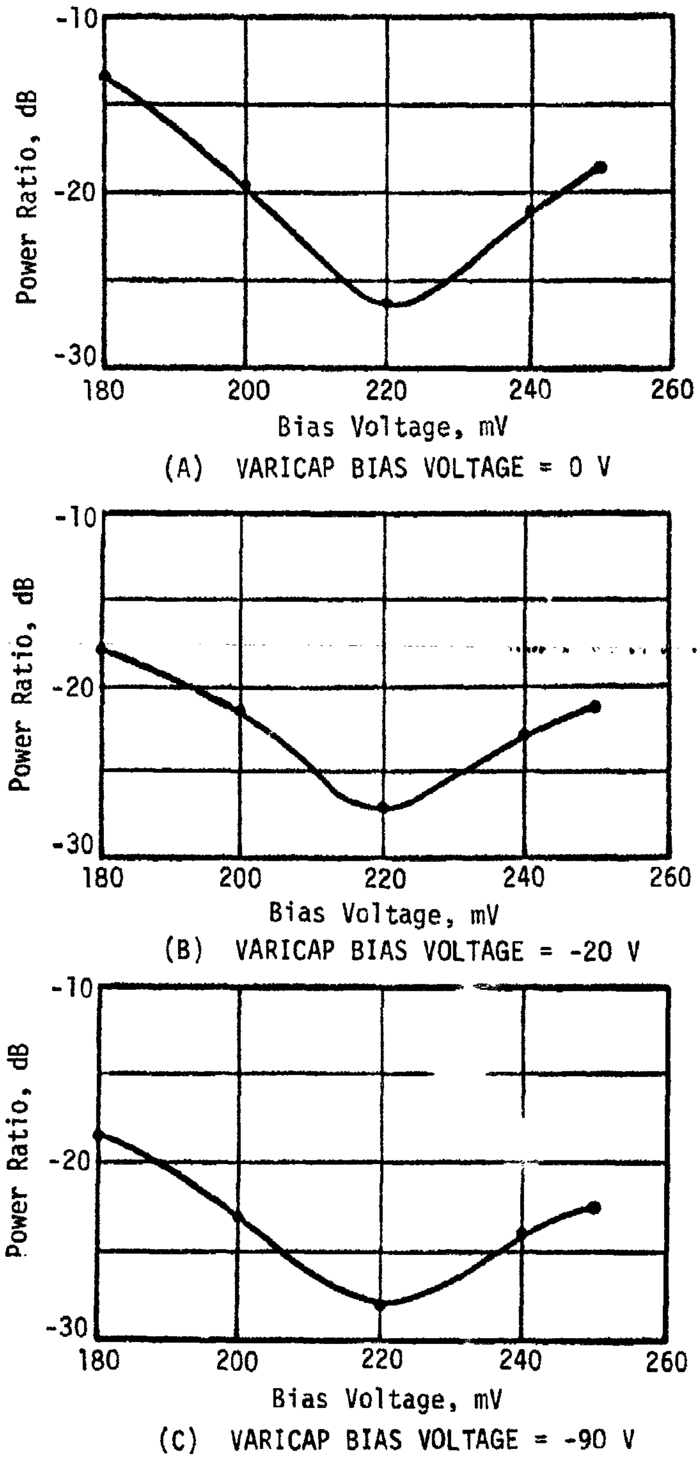


Figure 5-135. Ratio of Second-harmonic Power to Fundamental Power vs Bias Voltage of Tunnel Diode

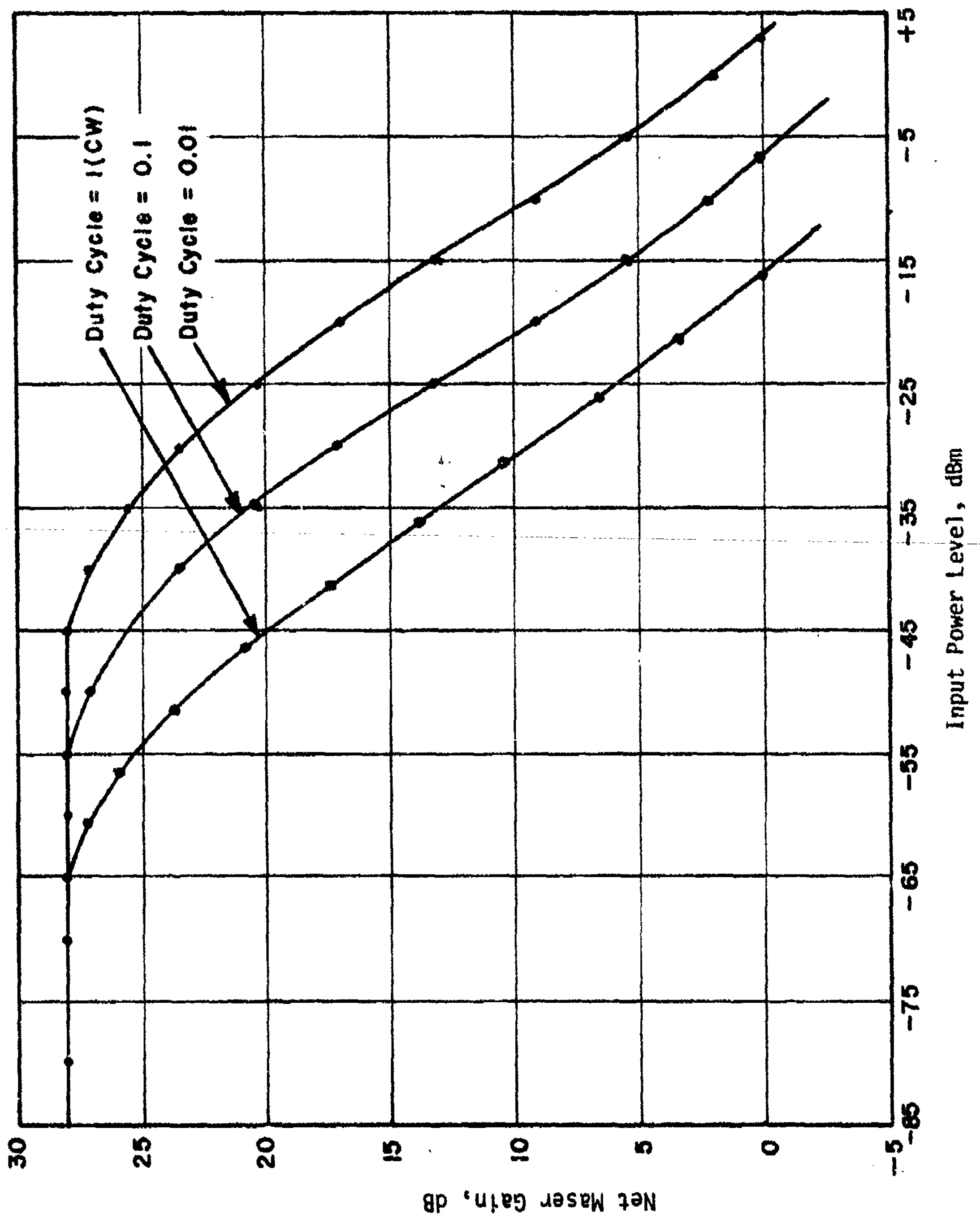


Figure 5-136. Gain Saturation as a Function of Duty Cycle

g. Gain-recovery time of a maser is dependent on maser material and bath temperature. Typical gain-recovery times range from 50 to 200 ms.

h. The characteristics of the TWM, because of its high degree of output to input isolation, are relatively unaffected by moderate impedance mismatches to external circuits.

The behavior of the TWM characteristic is given by a power output versus a power input curve rather than the V-I curve characteristic of vacuum tubes or transistors. The transfer impedance of the maser is equal to a constant even at saturation levels, and the equivalent V-I characteristic is linear.

5-14.4.2.2 Parametric Amplifiers

The parametric amplifier also is a low noise, wide bandwidth signal frequency amplifier. Its noise performance is between that of a maser and good tunnel diodes, traveling wave tubes, and conventional diodes. It can be shown that a time-varying capacitance such as obtainable with a semiconductor diode excited by a suitable "pump" frequency will give linear amplification of small amplitude signals.

The interference characteristics of these devices are saturation, desensitization and cross-modulation, intermodulation and gain stability, and impedance effects. Refs. 79 and 80 give a detailed experimental and theoretical analysis of parametric amplifiers under CW and pulsed conditions.

The two most important interference characteristics in parametric amplifiers are gain-change effects and generation of additional frequencies.

The measurement of the cross-modulation of a 1.4-GHz parametric amplifier indicated that:

a. When the sum of the powers of the desired and interfering signals is less than the saturating power level, the cross-modulation is proportional to the interfering power level.

b. For a constant gain, when the sum of the powers of the desired and interfering signals is less than the saturating power level, the percentage of cross-modulation is independent of the desired signal input power.

c. When the level of the interfering signal exceeds the saturating power level, the power required for a given amount of cross-modulation increases by the same amount as the decrease in gain for the desired signal.

Fig. 5-137 illustrates the saturation characteristic; and cross-modulation performance versus frequency is shown in Fig. 5-138, while cross-modulation versus input interference signal level is shown in Fig. 5-139.

Fig. 5-140 shows small signal gain vs interference for several frequencies. Fig. 5-141 illustrates power required for a fixed reduction in small signal gain (20 dB).

The parametric amplifier exhibits no unusual effects from large pulse-power inputs or moderate input powers at its idler frequency; however, pump-power radiation from input and output ports is observed. Appropriate filtering is therefore necessary.

5-14.4.2.3 Tunnel Diodes

The block diagram of a typical tunnel diode amplifier is shown in Fig. 5-142. The circulator is used to isolate the amplifier input and output, and the directional filter masks the out of band characteristics of the circulator.

The noise figure of tunnel diode amplifiers is better than conventional mixers. They have rather low dynamic range, however. The dynamic range varies with the product of the differences between the peak and valley currents and voltages. It can be increased by using matched diode pairs in push-pull as shown in Fig. 5-143. This arrangement also provides optimum noise performance. A typical saturation curve for a single diode amplifier is shown in Fig. 5-144.

The desensitization and cross-modulation characteristics of tunnel-diode amplifiers can be related directly to the amplifier saturation curve. A typical desensitization curve is shown in Fig. 5-145. Cross-modulation is shown in Figs. 5-146 and 5-147.

The resistive cutoff frequency of the tunnel diode limits the maximum gain-bandwidth product and the minimum noise figure obtainable. The amplifier is also more susceptible to interfering signals when operating near the resistive cutoff frequency.

5-14.4.3 Compatibility Control

There are two means of controlling compatibility other than adjustment of active device and antenna characteristics (as discussed in pars. 5-10 and 5-14.4.2): (1) use of filters, and (2) control of leakage from transmission lines.

5-14.4.3.1 Transmission Lines

Because the skin depth is so small at microwave frequencies, there will be no significant penetration of solid conductor enclosed transmission lines such as coaxial cables or waveguides except at discontinuities such as at joints or penetrations for tuning adjustments.

At waveguide flanges the leakage depends upon mechanical alignment. Levels of field strength 130 dB below that in the guide have been measured (Ref. 81).

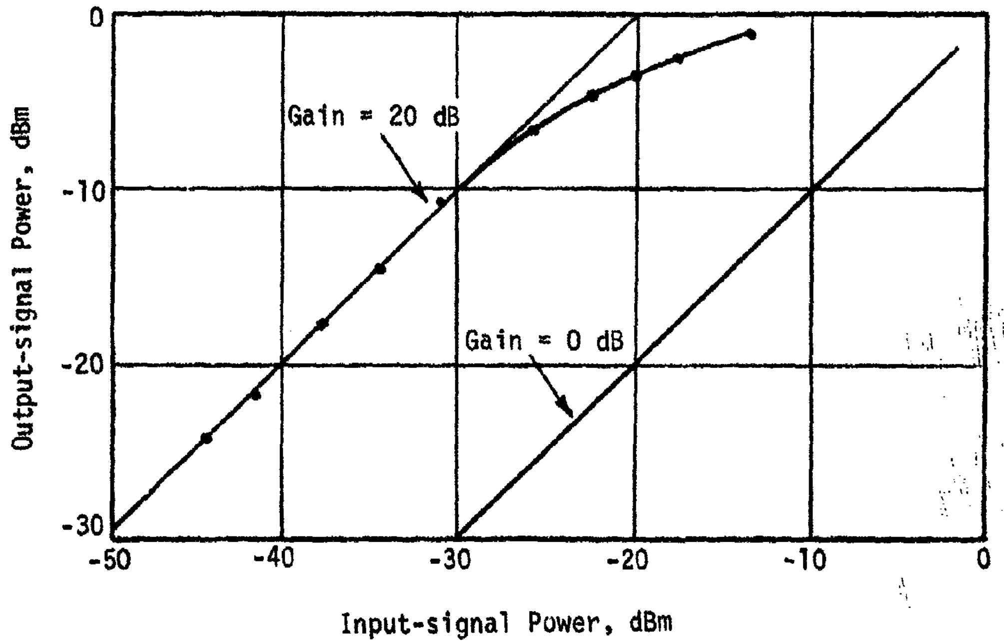


Figure 5-137. Saturation Characteristic of Microwave Parametric Amplifier

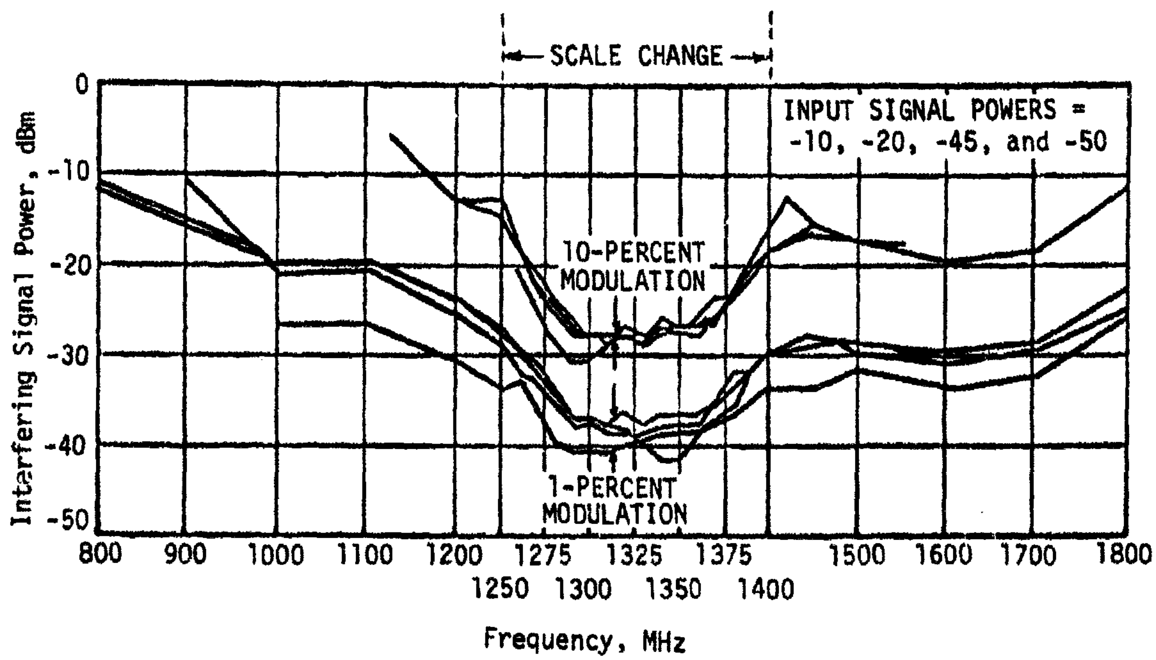


Figure 5-138. Cross-modulation Measurements

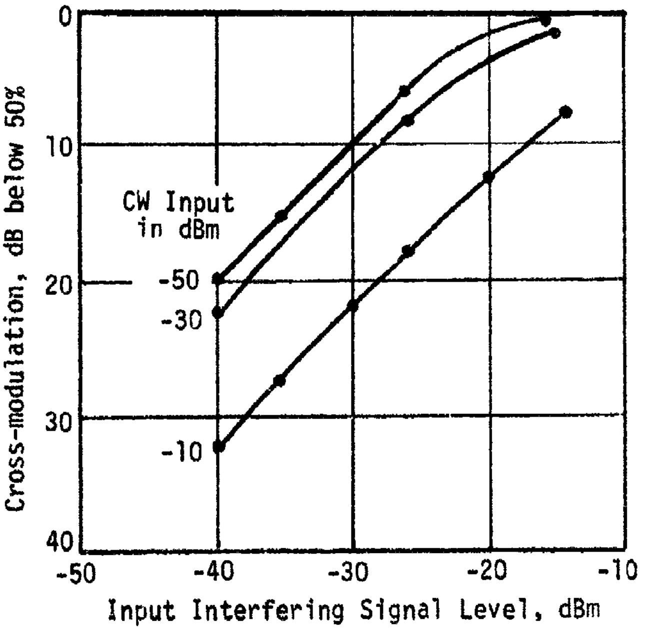


Figure 5-139. Cross-modulation vs Input Power

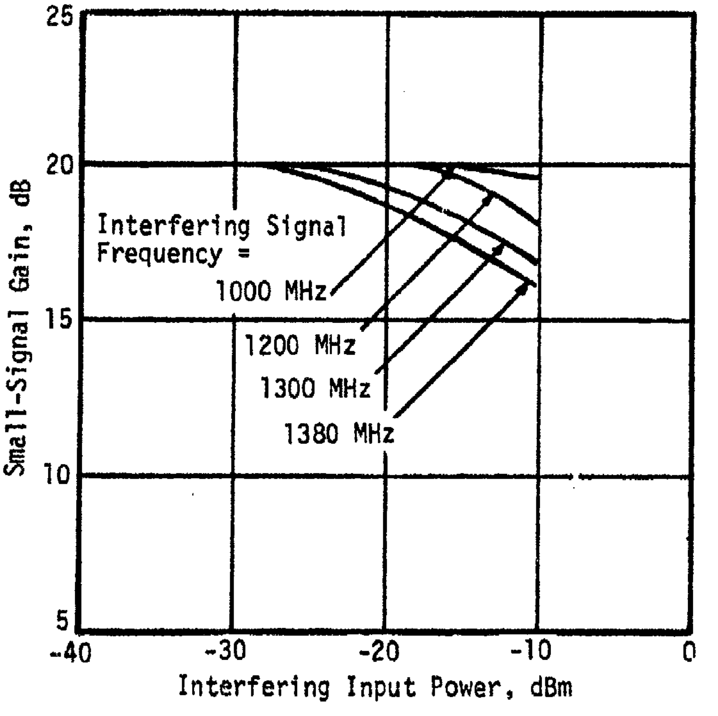


Figure 5-140. Small Signal Gain vs Interfereing Input Power

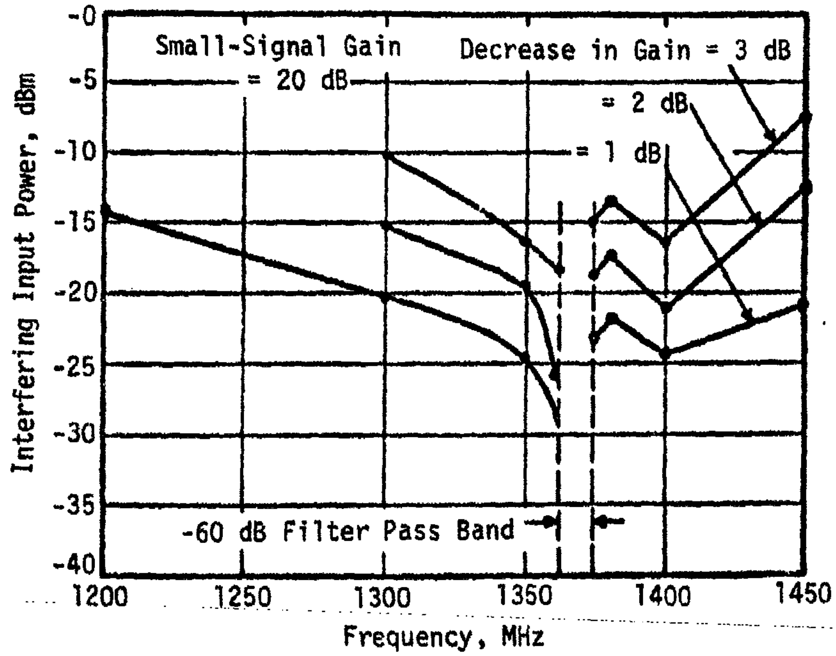


Figure 5-141. Reduction of Gain as a Function of Interfering Power and Frequency

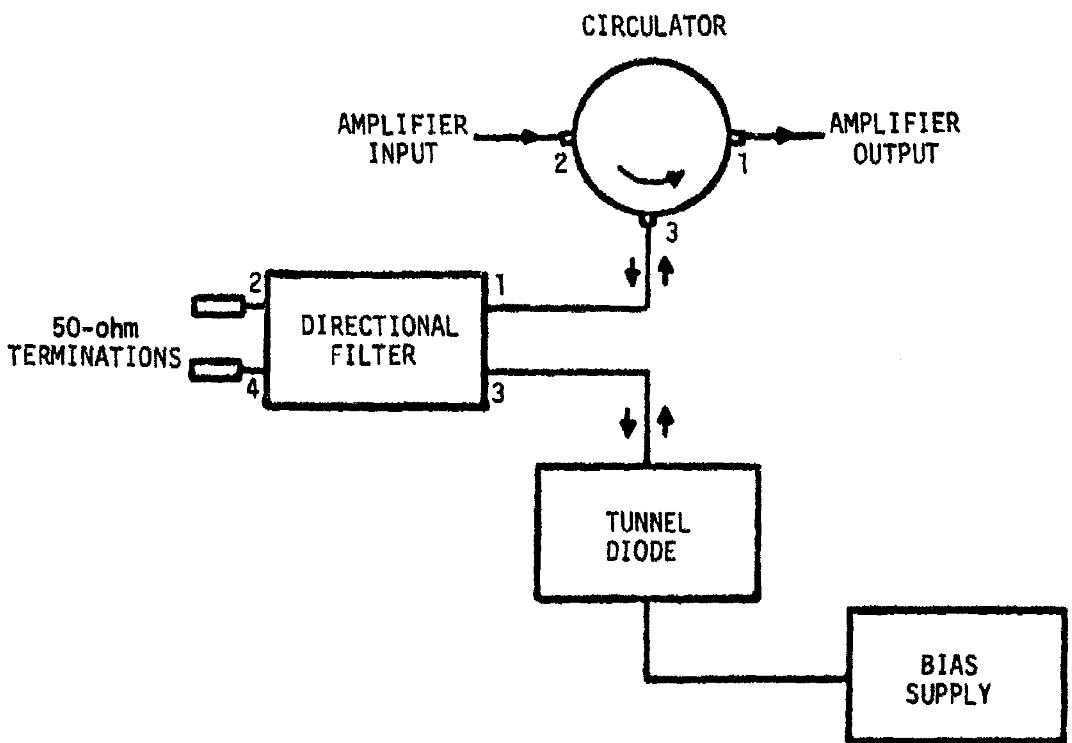


Figure 5-142. Block Diagram of Tunnel-diode Amplifier

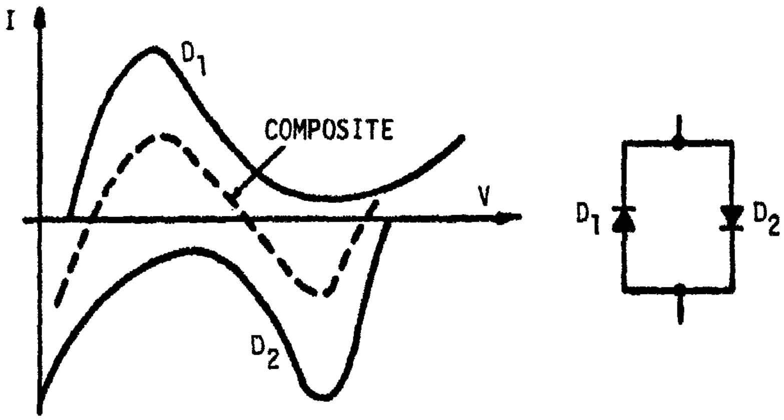


Figure 5-143. Method for Extending Tunnel-diode Dynamic Range Using Inverted Pairs

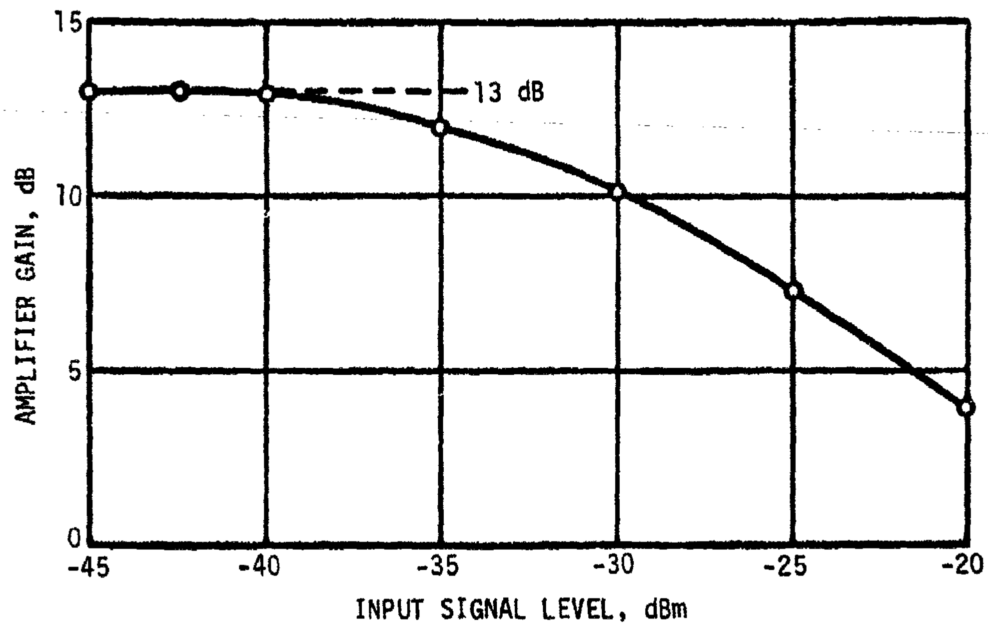


Figure 5-144. Saturation Characteristics of Tunnel-diode Amplifiers

In critical cases the leakage can be controlled by properly designed gaskets (see par. 4-6.5.4).

Flexible coaxial cables in which the outer conductor is made of metal braid is subject to leakage at the crossovers of the braid wires. Few data are available on the magnitude of this leakage in the microwave range, but the theory of such leakage is discussed in par. 3-3.1.1.2.5.2, along with some limited experimental data on transfer impedance in the microwave ranges. For use at microwave frequencies, it is common to use double-braided cable where flexibility is a requirement.

Within device cabinets it is common to use microstrip which consists of a single ribbon conductor laid over a "ground" plane and insulated by means of a flat dielectric. In such a structure the major portion of the energy, up to at least 90%, flows in the region directly under the ribbon conductor. The remainder

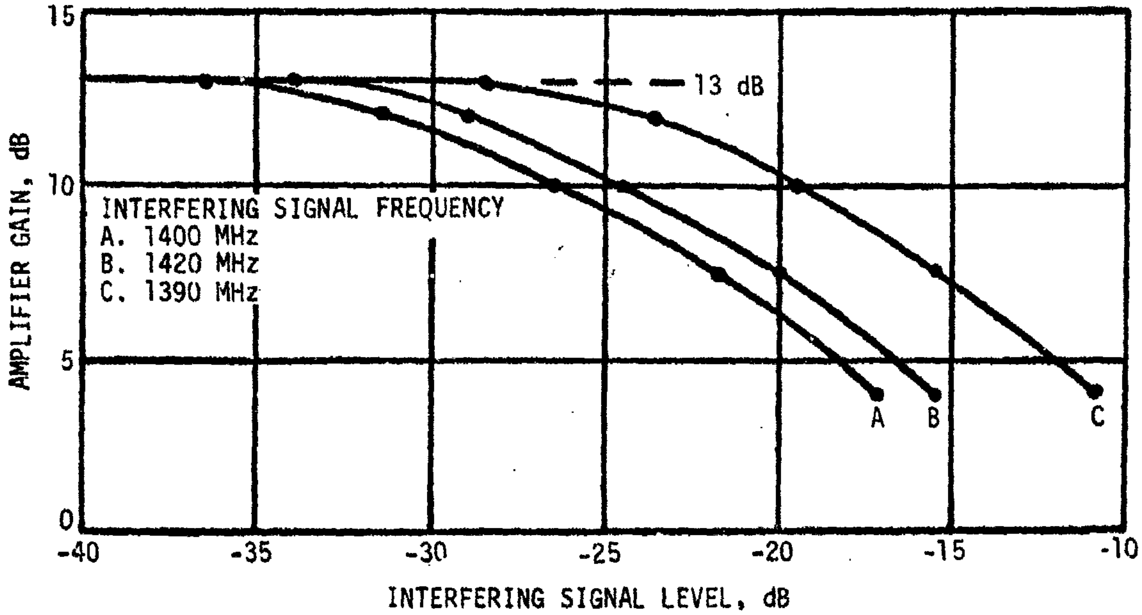


Figure 5-145. Desensitization Characteristics of Tunnel-diode Amplifiers

can be considered to constitute a leakage field and can couple to other lines in its vicinity. The magnitude of the coupling can be estimated very roughtly by using Figs. 4-11 and 4-12 for the mutual capacitance and inductance of wires over a ground plane. Accurate, theoretical values for the coupling do not appear to have been worked out. Results are given by Kaupp (Ref. 82) in terms of two empirical constants K_B and K_F by the formulas

$$V_B(t) = K_B V_o(t + t_o) - K_B V_o(t - t_o) \text{ , V} \quad (5\text{-}43)$$

$$V_F(t) = K_F l \frac{dV_o(t)}{dt} \text{ , V} \quad (5\text{-}44)$$

where

$V_o(t)$ = voltage at the receiving end of the driving line at time t, V

$V_B(t)$ = voltage of the coupled line at the sending end, V

$V_F(t)$ = voltage of the coupled line at the receiving end, V

$t_o = T_D$ = line propagation time, nsec

T_D = propagation time, nsec

l = coupled length of lines, ft.

Values of K_B and K_F are shown in Fig. 5-148. Both lines are assumed terminated in their characteristic impedance. It should be noted that K_B is larger than K_F and that the coupled voltage can easily be 5-10% of the driving voltage for close line spacings.

5-14.4.3.2 Filters

Filters are commonly used in transmitter outputs to prevent the emission of harmonics and spurious frequencies that are outside the band of frequencies of the desired signal. Details on the characteristics of such filters are given in par. 4-5.4. The design of such filters can be difficult for waveguide applications because of the possibility of higher order modes in the waveguide.

For use with microstrip internal to the cabinet, similar techniques can be used (Ref. 83).

5-15 TELEMETERING

5-15.1 INTRODUCTION

Telemetering is a term applied to the transmission of data from one point to another, usually by means of an electrical circuit. The points between which the information is transmitted are considered to be located remotely so that there is no direct visual communication between the two points. Many systems operate with a radio link in which case the basic data must be modulated onto a radio carrier, transmitted,

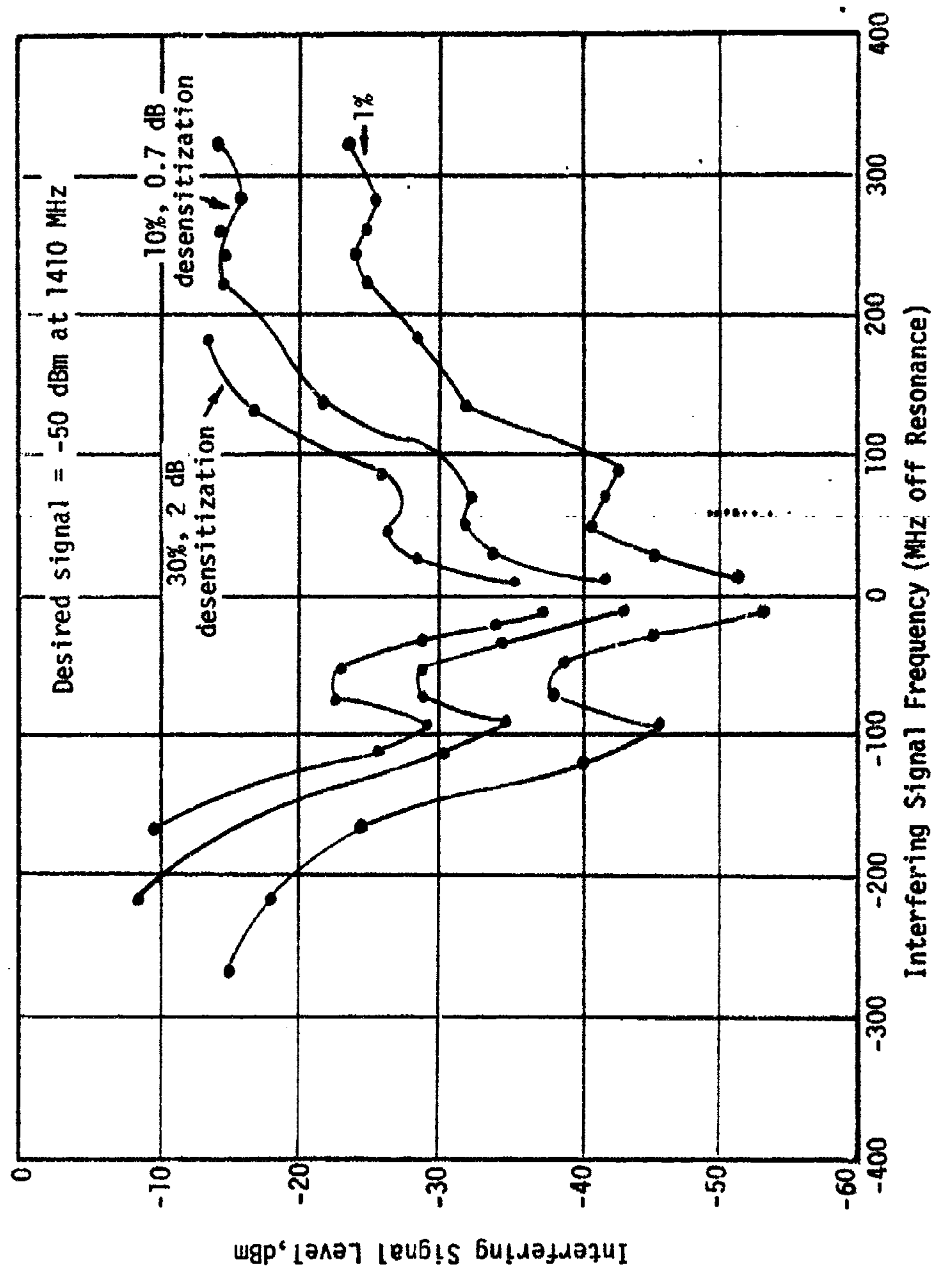


Figure 5-146. Cross-modulation Characteristics of Tunnel-diode Amplifier

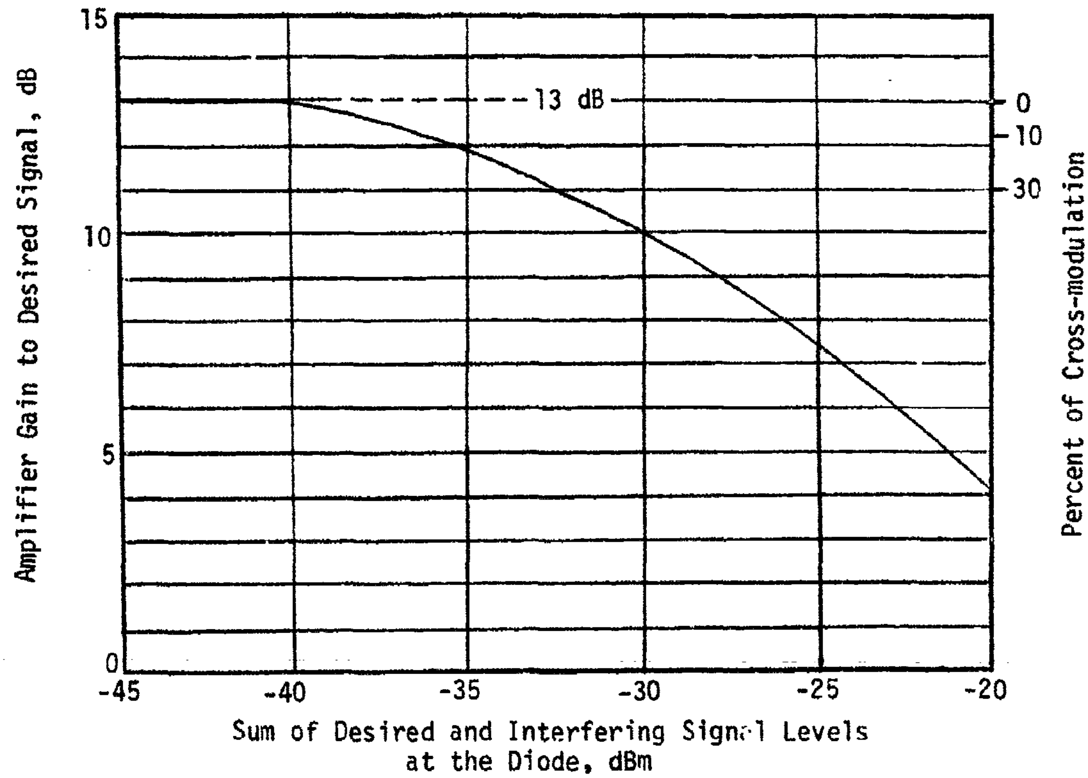


Figure 5-147. Summary of Interference Characteristics of Tunnel-diode Amplifier

detected, and then coupled to an output device or indicator. A general diagram for a telemetering system is shown in Fig. 5-149 (Ref. 84). Most telemetry systems operate with a commutator so that various kinds of data can be multiplexed onto a single channel. A corresponding commutation system must be used at the receiver in order to separate the information into the various output channels. In some cases, the transducers will be followed directly by signal-conditioners which convert the signal from the transducer into a form more suitable for modulation of the subcarrier oscillators.

The sensors of transducers can be of various types as shown in Table 5-17.

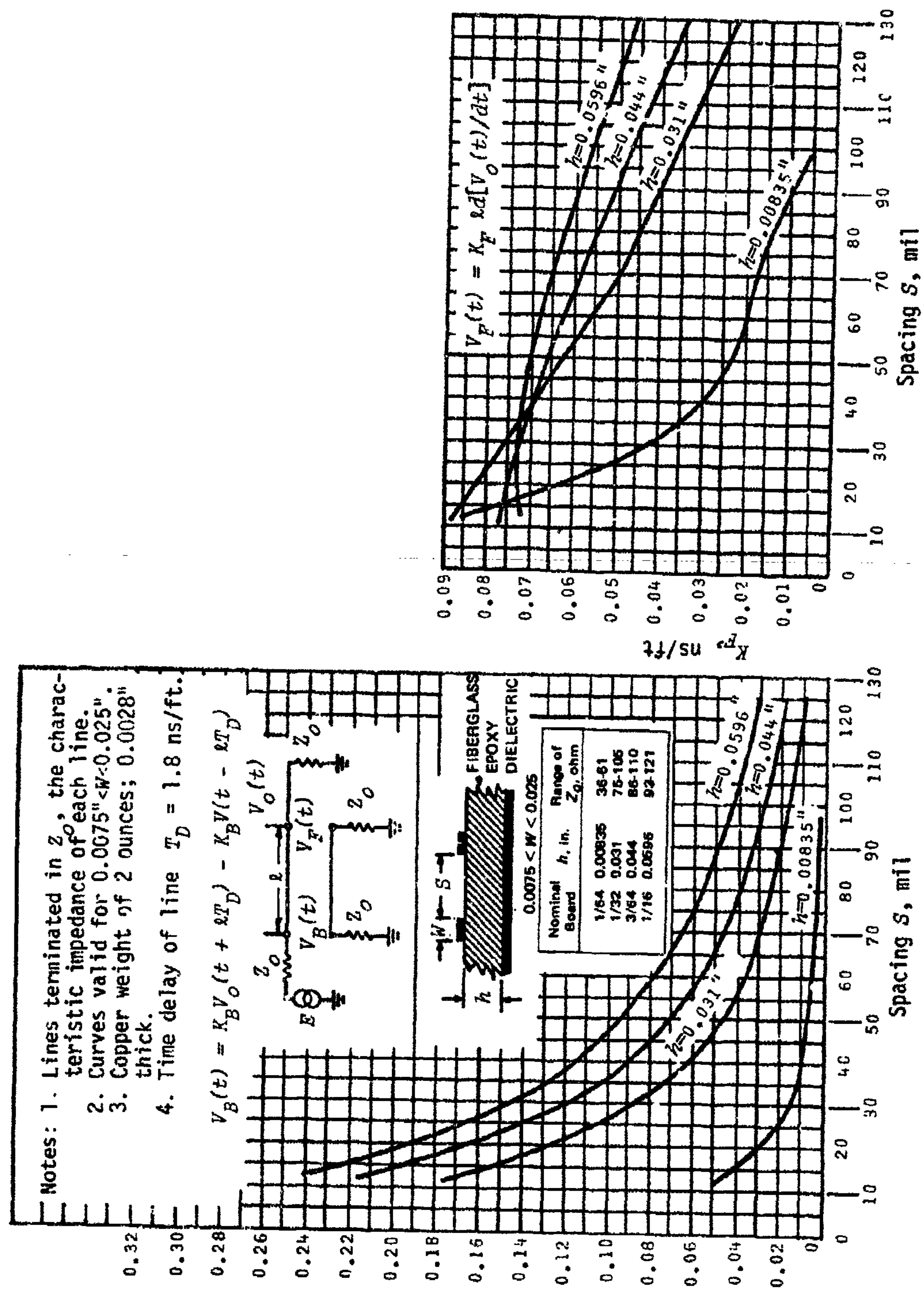
5-15.2 INTERFERENCE SOURCES

Telemetering equipment can produce interference from the following:

a. Bias supply on a passive transducer
b. Subcarrier oscillator
c. Commutator
d. Brush arcing of commutator drive motor
e. Radio transmitter
f. Radio receiver
g. Power supplies.

Interference from the radio transmitter and receiver is of the type discussed in Ref. 85 and in pars. 5-7 and 5-8, and will not be further discussed here. Brush arcing is discussed in par. 5-1 and power supplies in par. 5-5. Interference from the subcarrier oscillator is also similar to that from a receiver local oscillator, the only difference being that the subcarrier oscillator frequency may be lower than that used with most receivers and transmitters. Interference can be both radiated and conducted, and is controlled in the same fundamental ways.

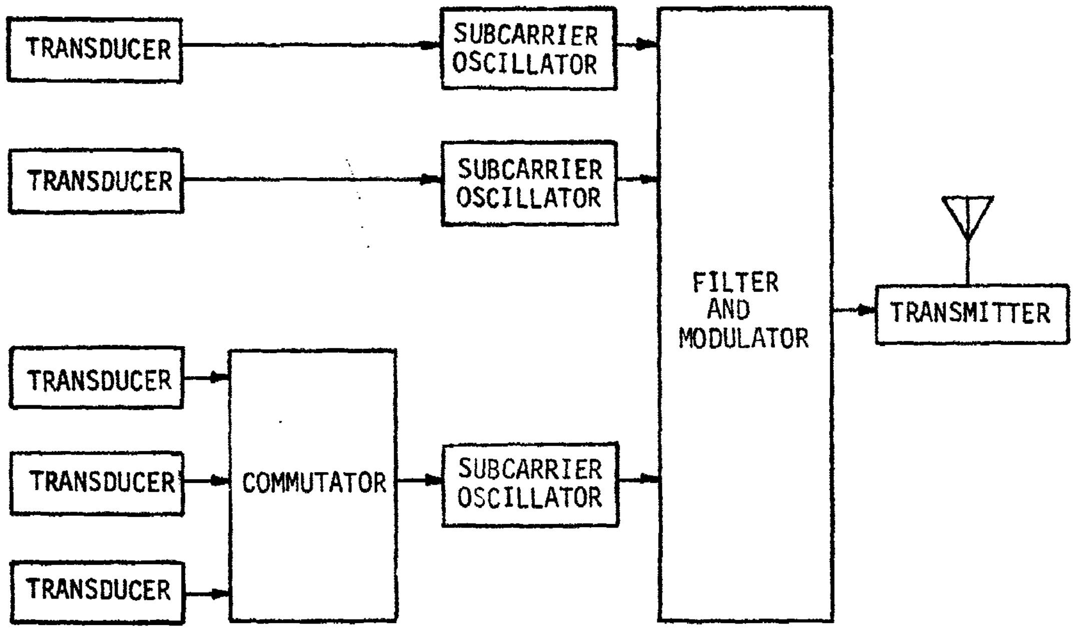
Interference from the transducers can occur only where a voltage or current bias is required. For example, a variable resistance transducer will have either a direct current or alternating current impressed upon it. If the effective loop area of the circuit in the vicinity of a transducer is large, the transducer itself is a potential source of interference to other transducers or devices located close to it. Correspondingly, a variable conductance, differential transformer, or



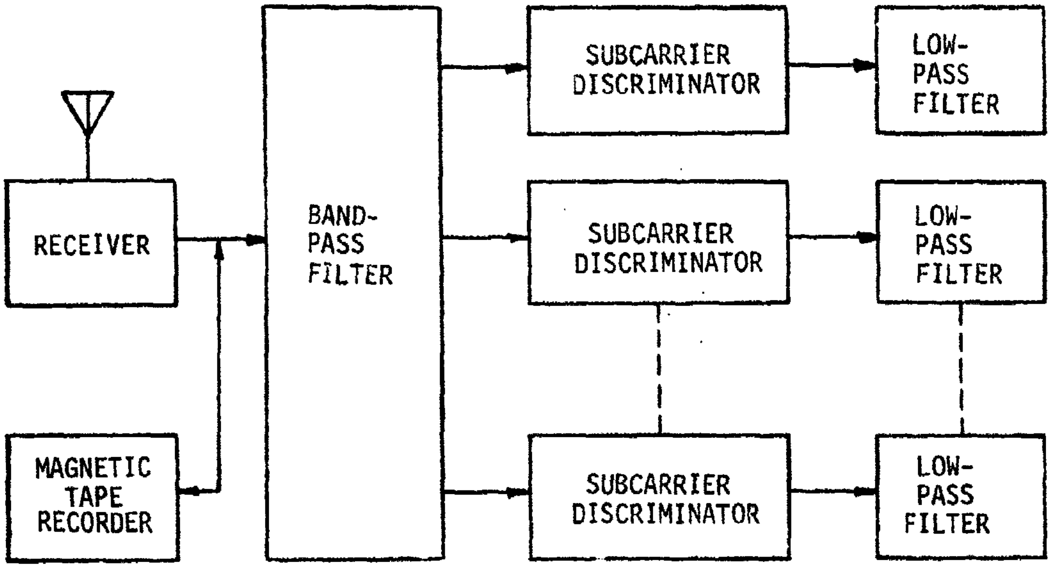
(A) Back Crosstalk

(B) Forward Crosstalk

Figure 5-148. Crosstalk Constants as a Function of Line Spacing



(A) Transmitter Block Diagram



(B) Receiver Block Diagram

Figure 5-149. FM-FM Telemetry System (Ref. 75)

TABLE 5-17. ACTIVE AND PASSIVE TRANSDUCERS

Active (voltage- or current-generating transducers)	Passive (variable-parameter transducers)
1. Piezoelectric	1. Variable resistance
2. Photoelectric	a. Nonmechanical resistance change
3. Thermoelectric	b. Change in internal structure
4. Magnetoelectric	c. Mechanically variable resistance
5. Electronic	2. Variable capacitance
6. Electrochemical	3. Variable inductance
7. Radioactive	4. Differential transformer
	5. Magnetostrictive

magnetostrictive type of device can produce significant levels of magnetic fields in their vicinity and again can couple into closely located circuits. Generally, the magnitudes of such fields will be low and with a reasonable amount of care they can be eliminated as significant sources of interference.

The commutator can be of either electronic or mechanical type. The mechanical type may require relatively large loop areas for the circuit connections to the commutator segments and thus are capable of radiating magnetic fields. However, the current levels at which such commutators usually operate is low and unlikely to cause significant levels of interference especially at frequencies in the 10 kHz range.

Variations in contact resistance can produce undesired transients in the data circuits themselves. Fundamentally, this is a circuit design problem and not a problem in electromagnetic compatibility. Measurements of ac generated noise may be in the region of 10 to 15 mV. The same magnitude can be expected in changes in dc level from one circuit to the next.

With the electronic switching technique, semiconducting elements are pulsed in sequence to provide a conducting path for each channel in its turn. Rates of commutation may be in the megahertz range. Such frequencies can produce radiation in a broad frequency range, especially considering that rectangular shaped waveforms are used. Again, the voltage levels required are relatively small so that it should be possible to provide adequate shielding and filtering to prevent significant interference.

5-15.3 SUSCEPTIBILITY

The radio circuits associated with telemetry are susceptible in the same way as the radio circuits discussed in par. 5-7. Special consideration may be given to the type of modulation. It is common to use pulse-duration modulation, pulse-width modulation, pulse-position modulation, pulse-code modulation, and pulse-amplitude modulation. Each of these has its own special performance criteria (Ref. 84). To the extent that they are forms of analog or digital data transmission, the performance as a function of signal-to-noise ratio can be presented in standard form. The most susceptible circuits in these systems are those associated with the sensory element and the radio receiver input circuit since these operate at the lowest levels. For some of the sensing elements, operation can occur at relatively high signal levels, thus eliminating this as a potential source of system degradation; however, in cases where the desire is to minimize weight such elements may operate at relatively low levels. The electrical connection to the sensor may be either balanced or unbalanced, the latter with a ground return as part of the sensor circuit. With the balanced circuit, the most sensitive part of the system will be that in the immediate vicinity of the sensor or in terminal boxes, since it is in these regions that the two conductor pairs are likely to be separated with the largest spacing. The susceptibility can be computed easily based upon the effective loop area. If the system is unbalanced, then the entire region between the sensor and the signal conditioner, if it exists, or the subcarrier oscillator or commutator could be susceptible to currents in the ground planes or common-mode currents in nearby wiring. Here, again, the susceptibility can be estimated directly in terms of models of wire lines as presented in par. 3-3.1.1.

5-15.4 CONTROL TECHNIQUES

The principal methods of interference coupling are through common impedance in the ground circuit, magnetic field coupling, antenna-to-antenna coupling, and RF pickup in transducer leads if long (> 20 ft) and exposed to nearby transmitter fields. In the latter case shields grounded at both ends may be required, otherwise such shields are grounded at only one end to prevent coupling through "ground loops". Another method of reducing susceptibility to RF fields is by placing filters in the signal conditioning or commutating circuits. Such filters are most effective where the data rates to be transmitted are low.

It is sometimes necessary to place a transducer or cable near a source of a strong magnetic field such as a motor or solenoid. Magnetic shielding material can be used to provide the necessary isolation.

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CHAPTER 6
SYSTEMATIC PREDICTION AND ANALYSIS

6-0 LIST OF SYMBOLS

A = parameter in shielding factor equation, m^{-1}
A_e = maximum effective aperture, m^2
A_1 = $0.32a_{21}$
ALS = antenna lobe selection
$ANT\ LB$ = antenna lobe
AS_j = sensitivity level of susceptor j, V
AS'_j = total tolerable input of a susceptor j to interference from priority No. 2 generators, V
$a(f,d)$ = power attenuation function
a_{w1} = height of wire 1 above ground plane, m
a_{w2} = height of wire 2 above ground plane, m
asm = adjustment safety margin
B = magnetic flux density, Wb/m^2; factor in magnetic field conversion equation
B_l, B_u = lower and upper limits, respectively, of receiver band pass, Hz
B_1 = constant in electric field conversion, m^{-1}
B_2 = constant in electric field conversion, $m^{-1} \cdot s^{-1}$
B_3 = constant in electric field conversion, $m \cdot s^{-1/2}$
B_4 = nondimensional
C_g = shunt capacity to ground, F
C_{L2} = lumped and distributed capacitance to ground for wire 2, F
C_{s1w2} = capacitance between wire 2 and and the shield of wire 1, F
C_T = total capacitance to ground, F
C_{w1s1} = coupling capacitance between wire 1 and its shield, F
C_{w1w2} = wire-to-wire capacitance between generator circuit and susceptor circuit, F
C_{w2g} = wire-to-ground capacitance, F
$C_1 \ldots C_8$ = constants in propagation formula, dimensionless
$CEAR$ = Comparative EMI Analysis Routine
CW = continuous wave
D = directivity of power radiated by wire
D_{ij} = distance of susceptor from generator wire for radiation coupling; separation between antennas i and j, m
D_r = reciprocal square of first resonant frequency of generator wire and its shield, s^2
d = conductor separation; distance to image in ground plane; propagation distance, m; reference distance
d_{ij} = separation between wires i and j, m
d_{12} = spacing between wires 1 and 2, m
E = electric field strength, V/m
$(FH)_R$ = highest frequency to which receiver can be tuned, Hz
$(FH)_T$ = highest frequency to which transmitter can be tuned, Hz
Fh_{s1} = propagation factor for field through holes of shield
$(FL)_R$ = lowest frequency to which receiver can be tuned, Hz
$(FL)_T$ = lowest frequency to which transmitter can be tuned, Hz
f = frequency, Hz; reference frequency, Hz; tuned frequency of receiver or transmitter, MHz
f_c = base frequency; crystal frequency, Hz
f_i = interfering transmitter frequency, Hz
f_{IF} = intermediate frequency, Hz
f_{LO} = local oscillator frequency, Hz
f_r = receiver tuned frequency, Hz
f_s = frequency of spurious emission, Hz
f_t = transmitter tuned frequency, Hz (MHz where noted)
f_1, f_2 = interfering frequencies, Hz
G = generator circuit
G_d = directive gain of dipole, dB

$G_i(f)$ = magnitude of generator i circuit output Fourier spectrum, V/Hz
G_r = receiving antenna gain, dB
G_t = transmitting antenna gain, dB
G_α = directive gain
GM = mean antenna coupling between transmitter and receiver, dB
GMA = mutual antenna gain, dB
$GR1$ = main beam gain of susceptor antenna, dB
$GT1$ = main beam gain of transmitting antenna, dB
$g(f, f_r)$ = receiver power selectivity function for tuned frequency f_r
H = magnetic field strength, A/m
HAR = harmonic level, dBm
$h(f, f_t)$ = transmitter power spectral density function at tuned frequency f_t, W/Hz
h_1 = desired portion of transmitter power spectral density function, W/Hz
I_1 = current in generator wire, A; current in wire 1, A
I_{N1} = net current, in generator wire 1 plus return current in its shield, A
IL = received interference level, dB
I_{s1} = current induced into wire 1 shield, A
INR = receiver output interference to noise ratio, dB
$INT\ MARGIN$ = receiver interference level relative to receiver interference threshold, dB
IPR = Initial Processing Routine
ISF = Intrasystem File
J = factor which describes what fraction of the system specification each subsystem will be allowed to generate

$$KAE = \frac{\text{voltage induced in susceptor}}{E\text{-field at susceptor antenna}}, \text{m}$$

$$KAH = \frac{\text{voltage induced in susceptor}}{H\text{-field at susceptor antenna}}, \Omega\cdot\text{m}$$

KE = voltage to E-field transfer function for antenna source, m^{-1}
KE^* = voltage to E-field transfer function for wire source, m^{-1}
KED = E-field fall-off exponent
KH = current to H-field transfer function for antenna source, m^{-1}
KH^* = current to H-field transfer function for wire source, m^{-1}
KHD = H-field fall-off exponent
k = width of guard band, Hz
L_f = off-frequency rejection, dB; bandwidth mismatch loss, dB
L_p = propagation path loss between transmitter and receiver, dB
L_{s1} = self-inductance of shield around wire 1, H
L_{w2} = self-inductance of circuit 2, H
$LPROP$ = path loss between antennas, dB; path loss to receiver, dB
l_w = wire length, m
l_{w1} = length of wire 1, m
l_{w2} = length of wire 2, m
l_{12} = common length of wires 1 and 2, m
M = mutual inductance of two circuits, H; mean ratio of expected signal and threshold level
m = harmonic number from transmitter
N = roll-off rate exponent, total receiver noise power, dBW
n = total noise power, W
P = interference margin, dBW
P_D = power density, W/m^2 or dBm/m^2
P_d = incident power density, dBW/m^2
P_i = probability of failure due to interference
$P_j(f)$ = magnitude of the voltage transfer function describing the frequency response of susceptor j, dimensionless
P_n = probability of failure due to noise
P_s = probability of successful operation
P_t = transmitter output; interfering transmitter power, dBW (except where noted)
P_t' = power, W
P_1 = desired output of transmitter, $10 \log p_1$, dBW
$PCT\ OVLAP$ = percentage of receiver tuning range overlap
PLO = local oscillator power present at antenna terminals of interfering receiver, dBm
POP = power output, peak, kW
p = oscillator harmonic number
p_1 = desired output of transmitter, W
q = signal harmonic number
R = signal-to-noise ratio, dB
R_{a2} = parallel combination of R_{g2} and R_{L2}, Ω

R_{g1} = source resistance of generator circuit, Ω
R_{g2} = source resistance of susceptor circuit, Ω
R_j = susceptor circuit j
R_{L1} = load resistance on generator wire 1, Ω
R_{L2} = load resistance of susceptor circuit, Ω
R_{os1} = dc resistance of wire 1 shield, Ω
R_r = radiation resistance, Ω
R_s = receiver sensitivity, dBW
R_{s1} = resistance of the shield on wire 1, Ω
R_u = interference-to-noise ratio, dB
RPS = relative performance score
$RSENS$ = receiver sensitivity or interference threshold, dBm
r_o = frequency overlap ratio, dimensionless
r_{s1} = radius of shield 1, m
r_{w1} = radius of wire 1, m
r_{w2} = radius of wire 2, m
S = low frequency shielding factor, dimensionless; signal level, W
$\bar{S}_p$ = mean value of predicted distribution of $S/(I+N)$, where S = signal, I = interference, and N = noise, dB
S_1 = measured value of $S/(I+N)$, where S = signal, I = interference, and N = noise, dB
S_{S1} = parameter in H-field conversion equation
S_{T1} = twisting factor for wire 1 = fraction of total current that returns to generator by a path other than the twisted return wire, dimensionless
S_{T2} = twisting factor for wire 2 = fraction of total current that returns to source end by a path other than the twisted return wire, dimensionless
S_t = total shielding factor
SHL = spurious harmonic level
$SIGMA\ G$ = standard deviation of the antenna gain product, dB
$SIGMA\ P$ = standard deviation of path loss, dB
SGR = specification generation routine
SPS = system performance source
$\overline{SPS}$ = mean SPS value for a specific bin
SPS_m = measured value of system performance score
$SPUR$ = receiver in-band spurious response rejection, dB
SS = sidelobe of transmitter to sidelobe of receiver
$(S/N)_O$ = output signal-to-noise ratio, dB
$(S/N)_{OT}$ = output signal-to-noise threshold value, dB
$[S/(I+N)]_O$ = output signal-to-interference plus noise ratio, dB
$[S/(I+N)]_{OT}$ = output signal-to-interference plus noise threshold value, dB
s = signal power, W
$T_{ij}(f)$ = magnitude of the voltage transfer function from generator i to susceptor j, dimensionless
$TART$ = task analysis routine
TE = E-field to voltage transfer function for antenna receptor, m^{-1}
TE^* = E-field to voltage transfer function for wire receptor, m^{-1}
$TED = \dfrac{E\text{-field at susceptor}}{E\text{-field at 1 m from generator}}$, dimensionless
TET = field-to-voltage conversion factor, m
$TEV = \dfrac{\text{voltage induced in susceptor}}{E\text{-field at susceptor wire}}$, m
TH = H-field to voltage transfer function for antenna receptor, $\Omega^{-1}\cdot m^{-1}$
TH^* = H-field to voltage transfer function for wire receptor, $\Omega^{-1}\cdot m^{-1}$
$THD = \dfrac{H\text{-field at susceptor}}{H\text{-field at 1 m from generator}}$, dimensionless
THT = H-field transfer function, Ω·m
$THV = \dfrac{\text{voltage induced in susceptor}}{H\text{-field at susceptor wire}}$, Ω·m
TI = inductive coupling current to voltage transfer function for close coupled wires, Ω
TV = capacitive coupling voltage transfer function for close coupled wires, dimensionless
t = shield thickness, m
UPS = upper performance score
V = voltage induced on shield of cable, V
V_c = common-mode voltage, V

V_{ij} = voltage induced by generator i at a reference point in circuit of susceptor j, V
V_N = voltage induced into wire pair as result of twisted loop area, V
V_1 = generator voltage, V
V_2 = voltage appearing on susceptor wire, V
X = parameter in shielding factor equation, m^{-1}
X_1 = frequency function in electric field conversion equation
x = ratio of shield thickness t to field penetration depth δ, dimensionless
Z = common impedance (contained in the TI model), Ω
z_o = impedance in shielding factor equation, Ω
δ = field penetration depth into shield, m; receiver bandwidth, Hz
θ = angle with respect to dipole axis; direction of receiver relative to transmitter, rad
λ = wavelength, m
μ_0 = permeability of free space $= 4 \times 10^{-7}$ H/m
σ = standard deviation
Φ = flux linking circuit in which V_c is induced, Wb
$\langle\ \rangle$ = expected value

6-1 INTRODUCTION

Most electrical and electronic devices are used as parts of systems or subsystems that are sufficiently complex that a large number of possible electromagnetic interactions are possible. The interactions can take place not only between antennas, but also interconnecting cables and equipment cabinets, usually considered on a pair-by-pair basis. Consequently, some organized procedure must be used in order to carry through a complete analysis. Various procedures have been devised in accordance with the nature of the interactions which are being considered.

In order to carry out the detailed analysis, it may be convenient to use digital computer programs. Because of the specialized nature of most available programs, it is not possible to prescribe their use for general problem solving. They can be applied to those problems for which they have been designed, but their application to other problems must be approached with caution.

An important application of system prediction and analysis is the determination of any special requirements for individual systems. Trade-offs of physical separation, shielding, and filtering against emission and susceptibility limits are frequently possible so that EMC standard military requirements can be "tailored" for a given configuration. Thus, computers can be programmed to develop a self-consistent set of specifications for all equipments comprising any given configuration. In another form they can examine the consequences of granting waivers from the standard limits.

An important side benefit of automated analysis is the ready availability of data stored in files for use in analysis programs. Typical of such facilities are those of the Electromagnetic Compatibility Analysis Center (ECAC) which are described in par. 6-4.4.

6-2 PROCEDURES

6-2.1 GENERAL

The basic logic of the large-scale EMC analysis program typically proceeds as follows:

a. A possible susceptor of EMI is selected.

b. A possible source of EMI to that susceptor is selected.

c. The EM Energy at the susceptor from the source is determined over all possible coupling paths.

d. The process is repeated for all possible EMI sources, and a decision is made of the extent to which the susceptor performance is degraded.

e. The process is repeated for other susceptors of interest. Computations are made by assigning quantitative values to pertinent parameters which define each source, susceptor and coupling path.

Individual procedures will differ regarding which parameter values must be provided, and, accordingly, the extent to which they use experimental or theoretical relationships; the types of source, receptor and coupling path they can handle; and the accuracy obtained. The logic used to generate specification limits usually proceeds as follows:

a. The tolerable interference threshold of a particular susceptor is identified.

b. The unintentional or extraneous EM emissions at the susceptor from all possible sources are compared against the threshold.

c. The emission limits of the sources or the configuration parameters are adjusted to reduce the level of signal in excess of this threshold or, where feasible, the susceptibility limit of the susceptor is adjusted.

d. The process is repeated for other susceptors of interest.

From a problem processing-time standpoint, it is customary to eliminate quickly any source/susceptor combinations that obviously will not result in system degradation through a "culling" operation. It is carried out by using very simple but conservative representations of the emitters, coupling paths, and susceptors involved.

6-2.2 INTRASYSTEM VS INTERSYSTEM EVALUATION

The distinction between intrasystem and intersystem EMI evaluation is characterized most easily by the differences in coupling paths involved. Typically, the major paths dealt with in intersystem investigations are those between antennas, especially those in which the antennas are sufficiently far apart that radiation field coupling calculations are made.

The major coupling paths for intrasystem EMI investigations are between cables, between equipment cases, and between cables and cases. Common impedance effects, such as use of the power supply bus, or ground currents that couple from one circuit to another, also are considered. When coupling via antennas is involved, it usually has to do with an antenna illuminating a cable or case, or with antennas that are in the near field of each other.

6-2.3 NONLINEAR MODELING

Many effects of interference are due to the inherent nonlinear characteristics of the source and susceptor devices involved. Unfortunately, the parameters that most strongly influence these nonlinearities are usually uncontrolled, and as a result cause wide variations in EM effects.

The primary nonlinear system characteristics that are described in available analysis programs include:

a. Harmonic and spurious emissions of emitters

b. Spurious responses and desensitization effects in susceptors

c. Intermodulation and cross modulation effects

d. Signal processing through detectors.

The frequency translation effects of these nonlinearities are describable using simple equations (for example, see par. 3-2.2.2.1 for receiver spurious responses and par. 3-2.2.2.2 for intermodulation). Expressions for amplitude or power levels of nonlinear outputs are much more complex and usually are avoided in intrasystem EM models because they require a large amount of processing time. Instead, nonlinear effects are often treated using measured data.

6-2.4 STATISTICAL CONSIDERATIONS

Many of the factors that influence whether or not interference will take place are time-dependent — such as coupling path loss variations, orientations of scanning antennas, orientations and locations of mobile equipment, and changes in equipment parameters with time. Additional factors vary in a non-time-dependent manner, such as changes in subsystem characteristics as a function of dial settings, tuned frequency, or equipment serial number. System analysis programs often take into account one or more of these effects.

The advantage of including statistical factors in an EMC analysis is that a better understanding of the potential of interference can be provided. Thus, a conclusion that system degradation of a particular type can occur 20% of the time, or in 40% of the cases, is considerably more meaningful than to indicate only that degradation can take place. However, statistical input descriptions or data often are not available.

6-3 DATA FILES

In the application of any of these techniques one of the most serious difficulties one experiences is in getting adequate data on characteristics of equipment. Usually design data — such as frequency range, sensitivity, type of modulation, and antenna gain — can be obtained. For nondesign information such as on spurious emissions or susceptibility one usually must rely on test data such as accumulated through the application on MIL-STD-462 and MIL-STD-449.

6-3.1 USAMSSA

The C-E Branch, US Army Management Systems Support Agency (USAMSSA), Washington, DC, maintains an EMC data base primarily to support the conduct of detailed EMC/EMI analyses; (Ref. 1) support from the EMC data base is available to any military activity to aid in conceptual and/or Research and Development (R&D) projects. Some of the data base files and their content are described in the paragraphs that follow for a limited view of the type of data available for studies, plans, and projects. Special retrievals and displays can be provided from each of these files.

a. Tactical Deployment File. Large scale C-E deployments are developed and maintained to satisfy the analysis of C-E based problems related to defined force structures. Deployments include C-E equipments for both friendly and opposing forces. The deployments usually are centered around major

Army studies and provide the basic tool for analyzing the EMC implications of new equipments, organizations, and concepts. Deployments previously developed range in size from a complete theater/Communication Zone operation to a separate brigade. Once developed, deployments provide an ideal study base for determining the probable effect of introducing a new C-E component into a specified environment. The C-E deployments are maintained and stored on magnetic tapes. Map overlays and various narrative descriptions also are developed to supplement the magnetic tape. In addition to their use in detailed analysis, the deployments contain a host of statistics relating to the tactical disposition of C-E equipments in a combat posture that can be useful in special projects. Examples of data that can be extracted include:

(1) Equipment densities in organizational units, command posts, net types, vehicles, TOE's, frequency bands, and modulation types

(2) Other retrievals and displays: equipment distributions, frequency utilization data, net/frequency requirements, overlays, systems diagrams, and equipment characteristics.

b. Equipment Authorization File. The Equipment Authorization File (EAuF) contains information normally derived from DA Tables of Organization and Equipment (TOE), US Air Force, and opposing force authorizations. Additional information is available to describe the normal netting associated with the C-E equipments. Each record in the EAuF has information on one C-E component. The equipment complement of each TOE is entered in the file, and each component of the major C-E equipment item is associated with an operator (equipment user, i.e., pilot, CO, XO, etc.). Vehicle types serving as C-E platforms (e.g., fixed, manpack, ¼-ton truck, and Ov-1D aircraft) also are entered in the file.

c. Equipment Characteristics File. The Equipment Characteristics File (ECF) contains the nominal C-E equipment technical characteristics for each component entered in the EAuF. The technical characteristics of the equipment are entered in the file as a function of the equipment application or operational mode. This file contains the component frequency range, emission type, antenna type, pulse data (if pulsed equipment), transmitter power output, etc.

d. Equipment Netting File. The Equipment Netting File (ENF) is used to identify the various C-E equipment net types typically associated with tactical military operations. A unique number is used to identify a discrete type net. The ENF carries the operational characteristics of the nets, such as modulation type, frequency limits, and antenna polarizations.

e. Antenna Characteristics File. The Antenna Characteristics File (ACF) contains specific antenna data related to the production of deployment data. Each antenna is described in relation to the type of vehicles in which it can be mounted, antenna height, and the location on the vehicle when used with aircraft, etc. The purpose of the ACF is to reference the type of antenna used with the C-E equipments in the ECF and the EAuF.

f. Equipment Application File. The Equipment Application File (EAF) provides data on the use and purpose of the C-E equipment. Some examples of uses are mobile ground FM communications and air ground communications.

g. Frequency Allocation to Equipment File. The Frequency Allocation to Equipment File (FAEF) contains all current frequency allocation and actions issued by the Department of Defense Joint Frequency Panel. Each allocation for the Army, Navy, and Air Force is entered into the FAEF record in English with a maximum of ninety-five fields of information. This file is supported by two indexes. One ordered by frequency and one ordered by equipment nomenclature.

h. Army Equipment Records File. The Army Equipment Records File (AERF) is a consolidated equipment data base designed for retrieval of current Army C-E equipment characteristics. It contains information similar to that found in the FAEF but only for Army C-E equipments. Some of the data found in this file are technical characteristics, phase in/phase out, type classification, inventory objectives and planned procurement quantities, equipment unit cost, line number, and Federal Stock Number. This file is also supported by two indexes, similar to the FAEF.

6-3.2 ECAC

The Electromagnetic Compatibility Analysis Center (ECAC) maintains a data base to support its work in analysis and prediction, and as a service to DoD (Ref. 2).

To permit efficient control and usage in a wide variety of applications, the data are organized into a multiplicity of files and subfiles, namely:

a. The Environmental File:

The environmental file consists of many different subfiles that pertain to the location and operating characteristics of communication and electronic equipment. The principal use of environmental data is to identify equipments in a given environment which are potential sources of interference to a proposed victim receiver or which are susceptible to interference from a proposed transmitter. This information is used to assist the compatibility engineer in such areas as frequency assignment, planning, and site selection.

There are two types of environmental data files: frequency oriented files and equipment oriented files. Frequency files include records of frequency assignments made by the ITU, the FCC, the IRAC, and by various elements of the DoD. Frequency files within DoD constitute a consolidated system (Frequency Resource Record System) which is maintained by ECAC.

Equipment oriented files were developed for EMC analysis activities at the Center. The major equipment oriented file (E-File) has been generated from data on DD Form 1374, *Electronic Equipment Environment Data Form*. This form is no longer used by the Army. Instead, the Army collects corresponding data on *Frequency Assignment Request Form* DA 2212.

Examples of the types of information recorded in environmental files include equipment location, operating frequency, transmitter power, modulation type, emission bandwidth, equipment nomenclature, and operating agency. Table 6-1 lists the various environmental data files.

b. The Nominal Characteristics File:

The Nominal Characteristics File (NCF) is composed of information regarding the technical characteristics of communication-electronic equipments. Such items as transmitter power, receiver sensitivity, bandwidths, antenna gain, antenna beamwidth, pulse width, pulse repetition frequency, and modulation type are extracted from technical publications and entered into the automated data base via remote terminals.

In addition to containing equipment technical characteristics, the NCF has information which relates individual receiver, transmitter, and antenna components to the systems of which they are a part. With this file, the compatibility engineer can identify the technical characteristics of the individual transmitters, receivers, and antennas, and, in addition, the communication-electronic systems, of which they are a part.

c. Organization and Platform Allowance File (OPAF). The environmental data files (which are location oriented) do not adequately accommodate mobile emitters on ships or aircraft. The OPAF was established to identify the C-E equipment complement on ships, aircraft, missiles, satellites, and tactical ground units. Their is linkage between the OPAF and NCF so that the technical characteristics of equipment referred to in the OPAF can be obtained easily. If information concerning the specific ships or type of aircraft that may be operating in a geographic area is available, the OPAF can be used to determine the C-E equipment to be considered in an EMC Analysis. Table 6-2 lists the various OPAF subfiles.

d. Spectrum Allocation and Use File (SAUF). The SAUF is a file containing rules, regulations, and agreements concerning the use of each allocated band in the frequency spectrum. Although some of the major allocation documents are readily available in handbooks (e.g., ITU Tables, US Government Allocation Tables, and FCC Tables) there are many obscure documents that concern the use of the spectrum both in the U.S. and in other areas of the world. The SAUF is an extract from these documents stored in the computer so that all information concerning a particular band and/or geographic area can readily be selected. Tables 6-3 through 6-5 list the documents presently in the SAUF.

e. Topographic File:

The topographic file consists of a grid of ground elevations extracted from topographic maps and stored in compressed form in the computer. The data are used to construct topographic profiles between any two points of interest. Computer programs have been developed to generate line-of-sight coverage overlays from the topographic file. This capability is referenced in the analysis section of this document.

The topographic file is based on a spherical coordinate system and is designed to accommodate data of variable grid spacing. The majority of the data has a grid spacing of 30 s (this corresponds to approximately 0.5 mi between data points).

f. The Frequency Allocation File:

One of ECAC's continuing tasks is to support the Joint Frequency Panel (par. 2-6.4.3) in its frequency allocation mission. Support consists of a review of each application for frequency allocation submitted to the Joint Frequency Panel by the individual Military Services. The review process includes checks for consistency of the technical information, checks to insure the application adheres to the national and international rules and agreements concerning usage of the frequency spectrum, and an evaluation of the proposed equipment as a potential compatibility problem.

Applications for a frequency allocation are submitted to the Joint Frequency Panel on DD Form 1494, *Application for Frequency Allocation*, which provides room for 109 fields of technical and administrative information. Results of the deliberations of the Joint Frequency Panel are announced in Action

TABLE 6-1.
DESCRIPTION OF ENVIRONMENTAL DATA SUBFILES

SUBFILE	DESCRIPTION	UPDATE SCHEDULE	FREQUENCY BANDS, MHz	COVERAGE AREA	AGENCIES	FORMAT
1374 Data	Contains data on transmitters and receivers at fixed locations obtained from reports made on DD-1374 forms. The instructions on the form exclude reporting of tactical equipment and equipment in operation less than six months, such as experimental systems. The Army is no longer providing 1374 data.	New data received on continuing basis Annual update of all data in file	Primarily Government and shared bands	US&P AF-Europe	DOD FAA	84 word
NAVAIDS	Contains data on TACAN, ILS, VOR, Glideslope, Beacons, Air Ground Communications, and other Navigation Aids. Data are obtained from 1374 forms; from the IFR Handbook, from FCC Licenses, and any other available source. Although this file is often considered part of the "E-File", it is not merely a file of 1374 reported data.	Monthly from IFR Handbook Continuing input from FCC Licenses	0.2-0.415 108-136 225-400 960-1215	US&P (VOR-TACAN Worldwide)	DOD FAA FCC	84 word E-File
IFF	Contains data on IFF ground interrogators obtained from DD 1374's and augmented with data on tactical IFF systems from the IRAC file and systems under development by industry from the FCC file. Additional data fields have been added which are needed to describe IFF systems. The file is maintained apart from the standard E-File but records are linked to the corresponding records in the 1374 file.	Update from 1374 file every 3 months	1030 1090	US&P AF-Europe	DOD FAA FCC	84 word E-File
Commercial Microwave	Contains data which are extracted from the FCC licenses for microwave communication stations. This includes data on privately owned microwave stations used by industry as well as common carriers; however, data for the largest common carrier, the AT&T Corp., are maintained separately in the AT&T file. This is considered part of E-File.	Weekly	890-960 1850-1990 2110-2200 2450-2690 3700-4200 5925-6425 6525-6875 10700-11700 12200-12700	US&P	Non-Govt	84 word E-File
AT&T	The data in this file are obtained direct from AT&T. It contains high quality location and equipment data. It does not contain specific operating frequency but does have the range of frequencies used on each link.	Semi-annually	3700-4200 5925-6425 11200-11700	US&P	AT&T	28 word E-File

(cont'd)

TABLE 6-1. (Cont'd)

SUBFILE	DESCRIPTION	UPDATE SCHEDULE	FREQUENCY BANDS, MHz	COVERAGE AREA	AGENCIES	FORMAT
FCC	This is a file containing the frequency, power, emission and location of all non-Government licensed emitters.	Semi-annually	All Non-Govt Bands	US&P	Non-Govt	28 word E-File
IRAC	This is the Master Radio Frequency File for Government Stations which constitutes the record of frequency assignments approved by the Interdepartmental Radio Advisory Committee (IRAC). This file is the only source of data on some of the nonmilitary government agencies.	Monthly	All Govt bands	Primarily US&P	Govt Agencies	CFEF
ITU	This is the record file of frequency assignments that have been registered with the International Frequency Registration Board (IFRB) of the International Telecommunications Union (ITU). Classified military usage of the spectrum is not in the file.	Semi-annually	All bands	Worldwide		28 word
CINCPAC (Pacific) (Atlantic) CINCAL (Alaska) CINC-SOUTH (Panama) CINCEUR (Europe)	These are frequency assignment files for the overseas commands. They are the most reliable source of US Military environmental data outside of CONUS. Equipment data are available in varying degrees depending on the subfile.	Daily Message Input	All bands	As indicated by subfile	*All US Military	CFEF
AFC's** White Sands Arizona Western Eastern Gulf	These are frequency assignment files for the Area Frequency Coordinators (AFCs). They contain reliable data in the vicinity of test ranges for which the AFC's are responsible.	As reqd (Data are very current)	All bands	Vicinity of test ranges indicated by subfile	Determined by AFC	CFEF

*Although the "CINC Frequency Files" are primarily records of US Military uses of the spectrum, each command includes data on nonmilitary operations in his area of responsibility (e.g., CINCAL includes FAA; CINCSOUTH includes Panamanian Civilian; CINCPAC includes Korean Civilian).
**Area Frequency Coordinators

Memoranda which authorize, deny, or modify the allocation request in some way. At this time, approximately 4000 applications have been processed through the Joint Frequency Panel and provide a source of ready reference information to the compatibility engineer. Key data elements from the DD Form 1494 which are printed in the Frequency Allocation List (FAL) are stored in a file at ECAC.

In addition to the automated files which have been described, there is a wealth of data available in documents, listings, and informal correspondence. Data in this "nonautomated" library category include:

a. Spectrum signature reports that contain measured data on the emission and reception characteristics of selected C-E equipment

b. Frequency allocation applications (J-12 Applications) which are maintained in the ECAC Library

c. Listings of data from Frequency Assignment Files that are not in the automated data base

TABLE 6-2. ORGANIZATION AND PLATFORM ALLOWANCE FILES

SUBFILES	DESCRIPTION	UPDATE SCHEDULE	
WATER	Contains data on the C-E equipment complement on Navy and Coast Guard ships by type, class, hull number, and ship name. The C-E complements are listed by systems nomenclatures and do not include components of systems unless the component is operating independently from the system. The US Army watercraft are entered as typical C-E configurations by basic design number for each type of vessel. The file also contains actual C-E configurations for Commercial Dry Cargo and Tanker Maritime ships operating under the US Flag.	Navy Coast Guard Army Commercial Maritime	- Quarterly - Semiannually - Semiannually - As required
AIR	Contains Air Force, Army, Navy, and Marine Corps Aircraft by Type/Model/Series of aircraft together with their complete C-E configurations. Information as to what equipments may be installed on specific tail numbers within a type/model/series of aircraft is not available. The file also contains the quantities of the type/model/series of aircraft in the present inventory for each Service. The file contains the Typical C-E Configuration of Commercial Air Carrier Type Aircraft by type, model, and manufacturer. It does not contain information on Air Carrier Aircraft manufactured by Lockheed Corporation.	Air Force Navy Army Commercial Air Carrier	- Annually - Semiannually - Deponds on availability of publications - Annually
LAND	Contains the complement of equipment in US Army, US Marine Corps, and US Air Force Land Tactical units including mobile tactical units. The C-E configuration for the US Army units is entered as typical configuration for TO&E Company/Battery Level of Tactical Ground Units. The C-E configuration for the US Marine Corps and US Air Force Tactical Units is based on the actual on hand C-E equipment.	Army US Marine Corps US Air Force	- No fixed schedule (Update source being investigated) - Quarterly - Annually

d. Extensive information in Technical Reports which describe details of systems analyzed at ECAC

e. A Future Systems Index which serves as a cross reference to data available in the NCF and other "hard copy" sources concerning future systems.

The library data described are usually not provided on routine data requests. It can be made available by establishing a data consulting task.

6-4 DESCRIPTION OF PROGRAMS

6-4.1 INTRODUCTION

The paragraphs that follow provide brief descriptions of some of the programs used in EMC evaluations. While no specific attempt is made to divide them into exclusive categories, the programs considered to emphasize antenna coupling are described

TABLE 6-3.
INTERNATIONAL SOURCE DOCUMENTS

ISSUING AGENCY	SHORT TITLE	LONG TITLE	FREQUENCY BAND & GEOGRAPHICAL AREA	CLASS & DOC. DATE
Allied Radio Freqency Agency	NATO Peace-Alloc. Tables	NATO Frequency Planning Document Part I (Peace)	20 MHz — 40 GHz NATO Countries	CN Mar 72
Allied Radio Frequency Agency	NATO War-Alloc Tables	NATO Frequency Planning Document Part II (War)	20 MHz — 40 GHz NATO Countries	SN Mar 72
Department of State	North American Reg. B/C	North American Regional Broadcasting (TIAS 4460)	540 kHz — 1600 kHz North America	U 15 Nov 50
Department of State	US/Canadian Agreement	Agreement Between the United States and Canada	30.56 MHz — 36 GHz North America	U 1967
International Tele-communication Union	ICAO Facilities	International Civil Aviation Organization — List of Facilities	108-117.915 MHz 960-1216 MHz Sea Area (Except Australia & New Zealand)	U Oct 71
German Bundes Post Ministry	Microwave Constr. FTC-173-R	Channel Designations for Radio Frequencies in Germany	2-14 GHz Germany	U Aug 72
International Tele-communications Union	I.T.U. Radio Regulations	International Telecommunica-tions	10 kHz — 275 GHz Worldwide	U 1971

first and those emphasizing wire and cable coupling are described last.

Table 6-6 shows a comparison of the programs, which emphasizes explicit capabilities. In some cases, programs not listed for a particular capability may be adopted for the purpose through a special operational procedure, which may involve manual or automatic analysis in addition to the program itself.

Antenna-coupled (intersystem) electromagnetic interference problems usually occur when several communication, radar, or navigation systems must operate at the same time in a relatively small area such as a city, military base, industrial site, building, or vehicle, such as a ship or airplane, and must send or receive signals over considerable distances.

The principal factors that enter into the analysis are:

P_t = interfering transmitter power, dBW
G_t = transmitter antenna gain, dB
L_p = propagation path loss between transmitter and receiver, dB
L_f = off-frequency rejection, dB; plus bandwidth mismatch loss, dB
G_r = receiver antenna gain, dB
R_s = receiver sensitivity, dBW
f_r = receiver tuned frequency, Hz
f_t = transmitter tuned frequency, Hz

6-4.2 THE ALLEN MODEL

This program was developed by the US Army Electronics Command, Fort Monmouth, NJ, to analyze antenna coupled interactions in a field army deployment (Refs. 3 and 4).

A complex of receivers and transmitters is located at specific points throughout a geographic area. The statistics of the signal-to-noise and signal-to-interference ratios are then computed using the following models.

6-4.2.1 Power Level in Receiver

The basic point of reference in system performance is the level of the power in the IF amplifier of the receiver. Mathematically, this can be given as

$$S = \int_0^{\infty} g(f,f_r)\,a(f,d)\,h(f,f_t)\,df\ ,\mathrm{W} \qquad (6\text{-}1)$$

where

S = signal power, W
d = distance, m
f = frequency, Hz
$g(f,f_r)$ = receiver power selectivity function based on tuned frequency f_r as a parameter

TABLE 6-4.
NATIONAL SOURCE DOCUMENTS

ISSUING AGENCY	SHORT TITLE	LONG TITLE	FREQUENCY BAND & GEOGRAPHICAL AREA	CLASS & DOC. DATE
The Portugese ARFA Member	Portugal Alloc. Tables	Portugese Allocation Plan	27.5 MHz — 40 GHz Portugal	C 13 Jul 70
Ministry of National Defense & Ministry of Communications	Rep. of China Alloc. Tables	Chinese Table of Frequency Allocations	10 kHz — 40 GHz Republic of China	U 25 Aug 70
Republic of Korea	Korean Table Freq. Alloc.	Korean Table of Frequency Allocations	10 kHz — 40 GHz Korea	U Jul 71
Executive Office of the President	OTP (IRAC) Manual	The US Government Table of Frequency Allocations	10 kHz — 275 GHz US & Possessions	U May 75
Executive Office of the President	OTP (IRAC) Supplement	Office of Telecommunications Policy Manual of Regulations and Procedures for Radio Frequency Management IRAC Supplement)	Various Frequencies Canada & US & Possessions	C Jan 73
Board of Communications	Philippine Alloc. Tables	Philippine Table of Frequency Allocation	1605 kHz — 1525 MHz Republic of the Philippines	U 1973
Ministry of Posts & Telecommunications	Japan Allocation Tables	Principles of Frequency Allocation	10 kHz — 275 GHz Japan	U Jan 73
Federal Communication Commission	FCC Rules & Regulations	Federal Communications Commissions Rules & Regulations	10 kHz — 300 GHz US Civil	U May 75
German Bundespost Ministry	Fed. Rep. Ger-Alloc. Tables	Frequency Plan of the Federal Republic of Germany	10 kHz — 40 GHz Germany	C Jan 66
Australian Post Office	Australian Alloc. Tables	Australian Table of Frequency Allocations	10 kHz — 40 GHz Australia	U Feb 74
Department of Transport	Canada-Allocation Tables	Canadian Frequency Allocation Tables	27.5 MHz — 40 GHz Canada	U Jan 69
German Bundespost Ministry	CEFM-5A	Subject: Additional Radio Frequency Information for the Federal Republic of Germany	70-8200 MHz Germany	C Jul 64

$a(f,d)$ = power attenuation function which depends, among other things, on the transmission distance d as a parameter

$h(f,f_t)$ = transmitter power spectral density function, which depends on transmission frequency f_t as a parameter, W/Hz

Normally, the integration taken only over a finite frequency range defined by B_l (in Hz) for the lower RF frequency and B_u (in Hz) for the upper RF frequency. Over the bandwidth considered, the attenuation function a is considered to be independent of frequency. Thus, Eq. 6-1 becomes

$$S = a(f_t,d) \int_{B_l}^{B_u} g(f,f_r)\,h(f,f_t)\,df\ , \text{W} \quad (6\text{-}2)$$

The transmitter output spectrum h is considered made up of separate emissions consisting of the desired output emission and various spurious output emissions. Defining p_1 by the relation

$$p_1 = \int_0^{\infty} h_1\,(f,f_t)\,df\ , \text{W} \qquad (6\text{-}3)$$

where h_1 is the portion of h consisting of the desired output.

$$S = p_1 \, a(f_t, d) \int_{B_t}^{B_r} g(f, f_t) \left[\frac{h(f, f_t)}{p_1}\right] df \text{ , W} \quad (6\text{-}4)$$

and, the signal-to-noise ratio R can be written as follows:

$$R = 10 \log(s/n) = P_1 + G_r + G_t - L_p + 10 \log \int_{B_t}^{B_r} g(f, f_r) \left[\frac{h(f, f_t)}{p_1}\right] df - N \text{ , dB} \quad (6\text{-}5)$$

where

R = signal-to-noise ratio at the output of the IF amplifier, dB
N = $10 \log n$, total receiver noise power, dBW
P_1 = $10 \log p_1$, desired output of transmitter, dBW
n = total noise power, W
L_p = propagation loss, $10 \log a$, dB
G_r = receiver antenna gain, dB
G_t = transmitter antenna gain, dB

6-4.2.2 Expected Values

The propagation loss is assumed to be statistical in nature and dependent upon the characteristics of the terrain. It is assumed that the loss in dB is a normally distributed random variable for which the mean or expected value $\langle L_p \rangle$ and the standard deviation σ, respectively, can be defined by the functions f_1 and f_2:

$$\langle L_p \rangle = f_1 (f, d) \text{ , dB}$$
$$\sigma = f_2 (f, d)$$

By use of the relationships in Eq. 6-5, the expected value $\langle R \rangle$ of the signal-to-noise ratio and the standard deviation σ are:

$$\langle R \rangle = P_1 + G_r + G_t - f_1 (f_t, d) + 10 \log \int_{B_t}^{B_r} g(f, f_r) \left[\frac{h(f, f_t)}{p_1}\right] df - N \text{ , dB} \quad (6\text{-}6)$$

$$\sigma = f_2 (f_t, d) \quad (6\text{-}7)$$

A normal probability density function with these parameters exist for each signal detected by the receiver. If the signal is the intended transmission, then the expected value $\langle R \rangle$ of the carrier-to-noise ratio is

$$\langle R \rangle = \langle \text{Input Power} \rangle - N \text{ , dB} \quad (6\text{-}8)$$

If the signal is not the intended transmission but instead an interference signal, then the expected value $\langle R_u \rangle$ of the interference-to-noise ratio and the standard deviation σ_u are given by:

$$\langle R_u \rangle = \langle \text{Input Power} \rangle - L_f - N \text{ , dB} \quad (6\text{-}9)$$

and

$$\sigma_u = f_2(f_t, d_u) \quad (6\text{-}10)$$

where the subscript u denotes the "unintended" or "undesired" signal, and the rejection L_f is obtained from the RF or IF filter characteristic as appropriate.

By carrying out the integrations of these equations, one can determine the probability of a successful transmission between the desired transmitter and its intended receiver in terms of these statistical variables and the minimum acceptable values of carrier-to-noise ratio and carrier-to-interference ratio. The procedure is relatively straightforward and is not described in detail here. The model also takes into account the possibility of interference being received from more than one source on a given receiver.

6-4.2.3 Receiver Model

The receiver model has an RF amplifier, frequency converter or mixer, and IF amplifier. As mentioned earlier, the criteria for acceptable performance are in terms of the radio frequency energy in the IF amplifier. The RF and IF amplifiers are modeled in terms of bandpass filter characteristics. Frequencies received that are outside of the RF amplifier, or outside the IF amplifier after frequency conversion, are considered to be incapable of producing a significant level of interference in the receiver. Spurious response frequencies f_s are determined by means of the equation (see Eq. 3-75):

$$f_s = \frac{p f_{LO} \pm f_{IF}}{q} \text{ , Hz} \quad (6\text{-}11)$$

where

f_{LO} = local oscillator frequency
f_{IF} = intermediate frequency
p and q = integers.

Since there are a large number of combinations of p and q that can be used, information is fed into the

TABLE 6-5. US MILITARY SOURCE DOCUMENTS

ISSUING AGENCY	SHORT TITLE	LONG TITLE	FREQUENCY BAND & GEOGRAPHICAL AREA	CLASS. & DOC. DATE
Military Communications Electronics Board (MCEB)	MCEB-M 156-70	Subject: International Registration of Frequencies Used by US Military on Foreign Soil	Entire Spectrum Worldwide	S Apr 71
MCEB	MCEB-M 149-71	Subject: US Military Frequency Allotment Plan for the 138-150.8 MHz Band	138-150.8 MHz US Military/ Worldwide	C 14 Apr 41
MFEB	MCEB-M 166-71	Subject: Radio Frequency Coordination Procedure	1435-1535 MHz Worldwide	U 27 Apr 71
MUEB	MCEB-M 600-67	Subject: Augmentation of US Military Frequency Resources	Various Frequencies US Military Worldwide	S 29 Dec 67
MCEB	MCEB-M 180-68	Subject: Military Frequency Planning 4400-5000 MHz	4400-5000 MHz Worldwide	C 1 May 68
MCEB	MCEB-M 221-74	Subject: NATO Frequency Planning Information	Various Frequencies and Future Trends Information NATO Countries	CN Nov 74
MCEB	MCEB-M 114-33	Message Form-Subject: Unattended Sensor Radio Frequency Band Selection	138-153 MHz 162-174 MHz US Military Worldwide	C 15 Jan 73
MCEB	MCEB-M 280-68	Subject: Clarification of Policy Guidance on Frequency Provisions for Drone Control and Test Range Safety Control Devices	406-550 MHz 4400-5000 MHz US Military Worldwide	C Jul 68
MCEB	MCEB-M 415-68	Subject: Recommended Tuning Standards for Radio, Communications in the 14 kHz-30 MHz Band	14 kHz-30 MHz US Military Worldwide	U Aug 65
MCEB	MCEB-M 9-67	Subject: Frequency Allocation Plan for the 138-144 & 148-158.8 MHz Bands	138-158 MHz US Military US & Possessions	C Jan 67
MCEB	MCEB-M 92-65	Subject: Frequency Assignments Plan for Air/Space/ Ground Telemetering Operations	225-260 MHz 435-1540 MHz 2200-2300 MHz US Military Worldwide	U Jan 65
MCEB	MCEB 373-48	Subject: Frequency Plan for 225-400 MHz Band	225-400 MHz US Military Worldwide	C 9 Jan 69
MCEB	MCEB 445-33	Subject: Establishment of Parameters for Radar Equipment Development & Operations in the 2700-2900 MHz Aeronautical Radionavigation Band	2700-2900 MHz US Military Worldwide	C 30 Dec 63
Joint Frequency Panel	Army-Air Force Agreement	Army-Air Force Agreement Regarding Use of Frequencies in the Bands 162-173 and 406-420 MHz	162-174 MHz 406-420 MHz US Military, US & Possessions	S 12 Apr 71
MCEB	CCB-199/17	Aus-Can-UK-US Combined Functional Frequency Allocation Plan	27.5 MHz-275 GHz AUS/CAN/UK/US	S May 66
Department of Navy	OPNAVINST 2400.23	Subject: Discontinuance of MF/HF Double Side Band Radio Telephony Emiss-	1605-30,000 kHz US Military Worldwide	C 11 Aug 72

(cont'd)

TABLE 6-5. (Cont'd)

ISSUING AGENCY	SHORT TITLE	LONG TITLE	FREQUENCY BAND & GEOGRAPHICAL AREA	CLASS. & DOC. DATE
Joint Chiefs of Staff	JCS 160300Z Apr 71	Message: Interference Coordination of Maritime Mobile Bands	4 MHz-23 MHz Pacific Command	C Apr 71
Joint Frequency Panel	J/FP 232214Z Apr 71	Message: Exceptions in Ship/Shore Operations in the Maritime Mobile Bands	4-25 MHz Pacific Command	C Apr 71
MCEB	CINCPAC 282325Z Jun 71	Message: Frequency Diversity Policy	Certain Bands in the 1712-8400 MHz Range Pacific Command	U Jun 71
MCEB	COUSUK COMMANEX (69) 1	Subject: COUSUK COMMANEX (69) 1	Discrete HF Frequencies	S 12 Apr 71
Joint Chiefs of Staff	USCINCEUR ED100-6	USCINCEUR Spectrum Management (Policies & Procedures)	Certain Bands in the 10 kHz-1700 MHz Range US DoD Agencies & Organization-Europe	U 1 Apr 74
Joint Chiefs of Staff	US/Japan Sta. of Forces Agreement	Chapter Two of the Japanese Telecommunications-Electronics Agreement	Various Frequencies US Military/Japan	U 23 Aug 74
Joint Chiefs of Staff	USCINCEUR 240934Z Apr 75	Subject: Radio Frequency Engineering Information	7-8 GHz Military & Civil in Germany	C Apr 75
MCEB	JANAP 141 (F)	US Joint Military Radio Frequency Allocation Plan	25.0 MHz-300 GHz Worldwide	S 20 Jun 74
NATO	ARFA Policy Handbook (TACAN)	Subject: Use of 960-1215 MHz by NATO Countries	960-1215 MHz NATO Countries	CN Mar 72
MCEB	MCEB-M-335-74	Subject: Wartime Use of Non-Government frequencies	1800 kHz-148 MHz Military-US & Canada	C 10 Oct 74

program when it is set up as to which values of p and q should be used for particular receivers.

6-4.2.4 Frequency Spectrum Model

Computations of interference effects are made at discrete frequencies even though the source may have a broad spectrum range. This is done by dividing the spectrum up into discrete frequency bands and inserting into the input data single frequency components having energies equal to the total energy contained in the incremental frequency bands. While this is an approximation, if the incremental frequency bands are small enough, the total energy computed should correspond closely with the expected total energy.

6-4.2.5 Propagation Loss Model

Based upon the methods of Longley-Rice (Ref. 5), 15 different propagation loss formulas are available for use depending upon the propagation path; e.g., between antennas on the same vehicle, on the same site, air to ground, and air to air. These formulas are all of the form

$$\langle L_p \rangle = C_1 + C_2 \log f + C_3 \log d + C_4 d \text{ , dB} \quad (6\text{-}12)$$

$$\sigma = C_5 + C_6 \log f + C_7 \log d + C_8 d \quad (6\text{-}13)$$

where the values of constants $C_1 \ldots . C_8$ are assigned for the particular path (Ref. 4), and f and d are, respectively, a reference frequency and a reference distance.

6-4.2.6 Antenna Gain

Two-dimensional antenna patterns of the sector gain type are used. The gains may be functions of frequency.

6-4.2.7 Method of Analysis

The model consists of 12 different computer programs each one of which performs one or more special functions in setting up the mathematical representation or in carrying out various aspects of the

TABLE 6-6. COMPARISON OF INTRASYSTEM PROGRAMS

	Allen (6-4.2)	EMETF/IPM (6-4.3)	ECAC-B (6-4.4.5)	IPP-1 (6-4.5)	COSAM (6-4.6)	SEMCA (6-4.7)	SEMCAP (6-4.8)	ISCAP (6-4.9)	IEMCAP (6-4.10)	SIGNCAP (6-4.11)
Specification Generation/ Compliance				X			X		X	
Interference Analysis	X	X	X	X	X	X	X	X	X	X
Waiver Analysis							X		X	
Culling Routines		X	X	X	X	X				
Model Routines										
Sources										
CW/AM/FM/FSK	X	X	X	X	X	X	X	X	X	X
Pulse/Ramp/Step	X	X			X		X	X	X	X
Digital Modulation	X	X	X						X	
Noise	X	X	X		X				X	
MIL-STD Levels			X					X	X	
Transfer Functions										
Wire-Wire										
Inductive							X	X	X	X
Capacitive							X	X	X	X
Conductive							X	X	X	X
E-Field				X			X	X	X	
H-Field				X			X	X	X	
Case/Case							X	X	X	
Case/Wire							X	X	X	
Antenna/Antenna	X	X	X	X	X	X	X	X	X	
Susceptors										
Linear										
Single Filter	X		X	X	X		X	X	X	X
Multiple Filter	X				X		X	X	X	X
Nonlinear										
Spurious	X		X	X	X			X		X
IM/CM			X	X	X			X		X
Harmonics	X		X		X		X	X		
Threshold Criteria										
Interference Levels	X		X		X		X	X		
(S + I)/I Levels	X		X		X			X		
Bit Error Scores		X			X					
Statistical Considerations	X									
Probability Scoring	X	X			X					
Statistical Analysis	X	X		X	X					
Computer Considerations										
Computer*	5700		1110	635	1110		6600 360/70	1108	6600 613	635 370/155
Core Size (kilowords)	90		50	90	32		64	50	60	48
Typical Running Time (minutes)	variable	10			2		10			
Developer	ECOM	AEPG		RADC	ECAC	GE	TRW	SFA	Mac/ DAC	Signa- tron
Cognizant Agent	ECOM	AEPG		RADC	ECAC	NSEC	RADC	ESD	RADC	RADC

*Computer Manufacturers are: 6600 - CDC; 360/70 - IBM; 613 - Honeywell; 370/155 - IBM; 1108 - Univac; 635 - Honeywell; 1110 - Univac; 5700 - Burroughs

compatibility analysis itself. The functions of the separate programs and the interrelationship of the various routines are shown in Fig. 6-1.

The Y program handles interference prediction by the statistical terrain method. The other programs in the model provide for the synthesis of the problem before interference prediction begins and for the statistical analysis of the detailed results afterwards. All 12 programs have a common method of representing the requisite details of the communication system under study and a common "system data set".

The system data set consists of three separate data arrays or lists. The first of these is the "circuit list", which contains the requisite data for each of the simplex radio communication circuits contained in the overall system. Data pertaining to a single circuit include the transmitter type, transmitter identity or location, receiver type, receiver identity or location, transmission frequency, receiver priority, and the transmitter on-off configuration. The entries in the circuit list are always ordered with respect to circuit transmission frequency. The second list in the system

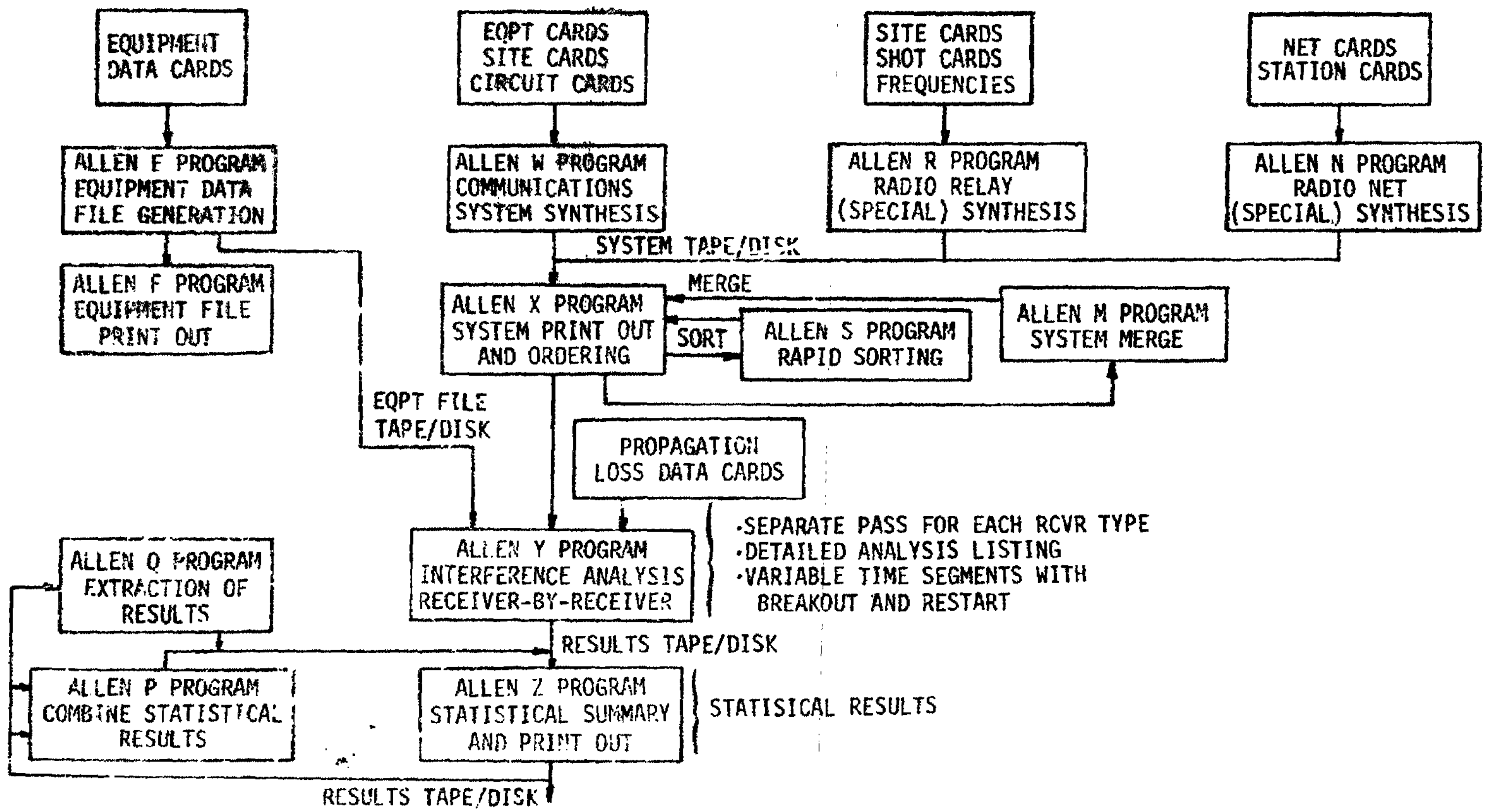


Figure 6-1. The Allen Model for the Synthesis and Analysis of Electromagnetic Compatibility Problems

data set is the "site list". It contains one entry for each significant geographical location associated with the communications system. Each entry consists of a unique site number, three-dimensional coordinates for the point, the total number of transmitters at the point, and the total number of receivers at the point. The site number provides the cross-reference between the site list and the circuit list: every circuit list entry refers to two site list entries, one corresponding to the transmitter site and one corresponding to the receiver site.

The third list is the equipment list which contains one entry for each different type of radio communication equipment used in the overall system. Each entry contains the unique designation of the equipment type, total numbers of transmitters and receivers for the given type, and other data relevant to the operating characteristics of the equipment.

The Y program analyzes the interference picture for only one radio receiver type in a single pass, regardless of the number of different types contained in the system. The investigator selects the receiver type to be analyzed in the pass by means of the master input card. The program then analyzes all receivers of the selected type found in the circuit list. The pass continues until the entire complement of receivers of the selected type has been examined or until the program is interrupted on the basis of the elapsed time on the computer. A separate pass must be scheduled for each radio receiver type for which interference prediction is desired.

The computer analysis scheme also provides for the selection in each pass of the types of transmitters to be included in the interference analysis. The program analyzes signals only from those transmitters whose equipment types have been explicitly selected at the beginning of the pass. All the remaining transmitters, if any, are presumed to be inoperative. The transmitter can be any signal source, friendly or enemy, intended or deliberate jamming.

The routine cycles through every receiver in the circuit list. If the receiver on the list is one of the selected equipment types, then a complete analysis is made of the intended and unintended signals to which it responds. A transmitter loop cycles through all possible signal sources. Those whose signals are rejected by the receiver by 235 dB or more are eliminated from further consideration. For those remaining, if any, the program generates implicitly the probability that the received signal exceeds the noise threshold of the receiver. Data for each of the 6 transmitters with the highest such probabilities are saved. No transmitter signal data are saved unless they exhibit a probability larger than 0.00003.

When the program has cycled through all possible signal sources the intended signal power distribution and the composite interference signal power distribution then provide the basis for calculation of the probability of successful operation P_s, the probability of failure due to noise P_n, and the probability of failure due to interference P_i. It should be noted that $P_s + P_n + P_i = 1$.

The power balance calculations for generating the expected carrier-to-noise ratio for the intended and unintended signals of a receiver use two large subroutines called the Input Power Procedure and the Rejection Procedure. This takes into account the various parameters in Eq. 6-5, such as geometry, antenna characteristics, propagation, and receiver selectivity.

6-4.2.8 Availability

For use of this program, contact should be made with Code DRSEL-NL-RY-5 at the US Army Electronic Command, Fort Monmouth, NJ.

6-4.3 EMETF/IPM

6-4.3.1 Introduction

EMETF is the Electromagnetic Environmental Test Facility of the US Army Electronic Proving Ground, Ft. Huachuca, AR. It has a test and validation mission (Ref. 1) derived from US Army Regulation 11-13, *Army Electromagnetic Compatibility Program*, to determine the "ability of communications-electronics equipments, subsystems, and systems, together with electromechanical devices, to operate in their intended operational environments without suffering or causing unacceptable degradation because of unwanted electromagnetic radiation or response." Since unwanted radiation could be either unintentional interference or intentional jamming, this responsibility embraces both susceptibility and vulnerability.

6-4.3.2 Operational Concept

In the performance of its mission the EMETF has developed a combination of automatic empirical testing facilities (see Chapter 7) to determine the degree of degradation that different types of interference or jamming cause to C-E and weapon systems and the conditions under which this degradation occurs. These empirically derived degradation data are used in conjunction with an Interference Prediction Model

(IPM) (Ref. 6) to provide a probability of satisfactory operation or system effectiveness for the equipments or systems in typical tactical environments (deployments). These tactical environments include continuously updated threat data of deployed opposing forces.

These analyses determine whether or not equipments or jammers in the environment can interfere with a particular equipment or system and the potential degradation in performance that each interfering equipment or jammer might cause. This potential degradation is then combined with other factors such as duty cycle, operational doctrine, scanning rate of rotating antennas, and similar operational and engagement parameters to derive a probability of satisfactory operation for the particular equipment or system.

6-4.3.3 Electromagnetic Compatibility Analysis

The Analysis Facility includes a program library and a system executive. The program library consists of computer programs and routines used in a variety of electromagnetic compatibility evaluations. By appropriate tailoring, the model can provide data on the probability of satisfactory operation of equipments in a specified tactical environment without any interference or jamming, with unintentional interference, with jamming including the unintentional degradation caused by the jammer, and the effectiveness of any Electronic Counter-Countermeasures (ECCM) techniques employed. The model can be tailored to provide an analysis of both the technical vulnerability of equipments or system and the operation vulnerability of the system in the tactical environment.

Deployment test beds are available from division site to an army in the field complete with Air Force support elements and opposing forces. Deployments vary from the current time frame to proposed concepts.

The model consists of interrelated mathematical models programmed basically in FORTRAN IV for the CDC 6000-Series Computer. The model mathematically simulates the electromagnetic environment of the tactical situation, simulates the C-E materiel operating characteristics, and predicts the performance of C-E equipments, subsystems, and systems. The mathematical models and the input data are derived from applied theory supported where possible by empirical data from actual physical measurements.

The model consists of two major sections: the Data Preparation section and the Interference Prediction section. There are many separate computer programs and routines within these two sections which are selected as appropriate for the specific task.

The Data Preparation section performs two major functions: (1) development and maintenance of a master data base, and (2) extraction and creation of a task test bed. The master data base is constructed from input data representing the deployment or usage of C-E materiel in hypothetical battle actions or scenarios.

It also includes technical data about the C-E equipments in the deployment, as well as scoring parameters relating to the satisfactory and unsatisfactory operation of the equipments. The task test bed consists of data selected from the master data base in accordance with the requirements of a specific task.

The Interference Prediction section consists of four modules: Link Selection, Interference Identification, Scoring, and Output.

The *Link Selection* module accepts data from the task test bed and selects links to be analyzed in accordance with criteria established in the test plan. The output of the Link Selection module is input to the Interference Identification module. Since the total number of equipments involved in a problem may be very large, a statistical sampling process is used within the Link Selection module to reduce the number of links formed by the C-E equipments in the deployment to a practical level for evaluation. The sample size is determined by the level of analytical precision required by the problem. The sampling process is weighted toward selection of those equipments having the greatest relative importance to the accomplishment of the assigned tactical mission.

In the *Interference Identification* module, the statistics of the desired and interfering signal levels at the input terminals of the receiver of the links being evaluated are calculated. These levels are a function of several factors, including transmitter RF output power, the power gains of the transmitting and receiving antennas, and the basic propagation path loss. The Interference Identification section identifies those transmitters which are potential interferers to each system being evaluated. This determination is based on transmitter power levels, transmitter duty cycles, and receiver characteristics. Duty cycle is defined as the percentage of time a transmitter actually is radiating. The output from the Interference Identification module is input to the Scoring module.

The *Scoring* module contains programs which compute the probability of satisfactory operation of the equipment being evaluated. The scores are presented in terms appropriate to the particular system being evaluated. The output from the Scoring module is input to the Output module.

The *Output* module accepts as input the link scores computed by the Scoring module and computes the system effectiveness (SE) of groups of equipments, categorized on the basis of experimental design criteria, in contributing to the success of the assigned military mission.

The principal outputs provide quantitative measures to predict how well the C-E equipments or systems will perform their designed functions, and whether the C-E equipments and systems will function in their intended operational environments. Typical outputs are:

a. The probability of satisfactory operation for individual C-E equipment in the absence of other C-E systems. This is a measure of communicability or operability.

b. The probability of satisfactory operation for individual C-E equipment in the presence of other C-E systems (compatibility). The difference between the operability and compatibility scores is a measure of compatibility.

c. The probability of satisfactory operation in the presence of enemy electronic countermeasures (vulnerability). The difference between compatibility and vulnerability scores is a measure of vulnerability.

d. The system effectiveness of preselected groupings of C-E systems under each of the preceding conditions.

e. Low-score analysis of links or groups of links in the deployment with scores falling below specified thresholds. These outputs are intended to provide quick recognition of C-E problem areas and provide insights to possible causes and corrective measures.

Both printed and plotted output data can be provided. All outputs are in camera-ready format, and can be tailored to the requirements of the specific task.

6-4.3.4 Computational Procedure (Ref. 7)

The steps followed in carrying out the computations are:

a. Receiver Characterization:

A receiver is chosen from the environment, and data concerning its location, assigned frequency, antenna type, and antenna orientation are stored in the computer memory. Then a search is conducted on a file of permanent receiver spectrum signature data pertaining to that particular type of receiver. When found, this information is stored in the memory of the computer. It consists of measured receiver characteristics such as frequencies acceptable to the receiver, selectivity, and sensitivity representing spurious responses and intermodulation characteristics of the receiver. They have been obtained from actual measurements made on the various receivers of interest. Predictions of spurious response frequencies are calculated using Eq. 6-11, and frequencies producing intermodulation interference are calculated from the following expressions:

$$\left.\begin{aligned} f_r &= f_1 - f_2 \text{, Hz} \\ f_r &= 2f_1 - f_2 \text{, Hz} \\ f_r &= 3f_1 - 2f_2 \text{, Hz} \end{aligned}\right\} \qquad (6\text{-}14)$$

where

f_1 and f_2 = interfering signal frequencies, Hz

f_r = receiver tuned frequency, Hz

The existence of cochannel and adjacent channel interference is determined from information contained in the file of permanent receiver data. Assuming a rectangular IF passband, the bandwidth δ is given and the corresponding adjacent channel attenuation is given. Signals whose frequencies f are such that:

$$f_r - \frac{\delta}{2} \le f \le f_r + \frac{\delta}{2} \text{, Hz} \qquad (6\text{-}15)$$

present the possibility of cochannel interference, and signals whose frequencies are such that:

$$k + f_r + \frac{3}{2}\delta > f > f_r + \frac{\delta}{2} + k \text{, Hz} \qquad (6\text{-}16)$$

or

$$-k + f_r - \frac{3}{2}\delta < f < f_r - \frac{\delta}{2} - k \text{, Hz} \qquad (6\text{-}17)$$

where $k \ge 0$ is the width of guard bands between channels, present the possibility of adjacent channel interference. These frequency ranges and corresponding sensitivity values are stored to be used later in the program to determine signal power level required to interfere with the receiver at that frequency.

b. Transmitter Characterization.

A transmitter is chosen from the environment, and data concerning its location, assigned frequency, antenna type, and antenna orientation are stored in the computer memory. Then a search is conducted on a permanent transmitter spectrum signature file to locate the data corresponding to that particular type of transmitter. Each frequency with associated power level now can be considered as a monofrequency transmitter. And, for simulation purposes, an actual transmitter is represented by its spectrum signature data as a composite of many monofrequency transmitters.

When the desired transmitter spectrum signature has been located in the file, the frequency of each monofrequency transmitter is compared against each spurious response of the receiver. If there is no match, the transmitter is discarded and another transmitter is chosen from the environment, and Steps 3, 4, and 5 of this paragraph are repeated until all transmitters have been examined. If there is a match, there exists the possibility of an interfering signal and a further investigation must be made. When all transmitters in the environment have been investigated, another receiver is chosen from the environment and Steps 1 through 5 are repeated until all receivers have been investigated.

c. Antenna Mismatch and Terminal Loss:

At the matching frequencies it is necessary to examine the power output of the transmitter and to modify it by appropriate loss factors before assessing its contribution to the electromagnetic environment at the receiver site.

The first modification to the transmitter power output is for antenna cable loss and impedance mismatch. This modification to the output power is necessary since many spectrum signature measurements are made into a resistive load. If the antenna and transmitter are matched at the fundamental frequency, the measurement of the fundamental into a resistive load is quite realistic. However, a mismatch occurs at other frequencies, namely, harmonics of the fundamental, so that if measurements of power at those frequencies are made into a resistive load, an unrealistic power output is reported.

A search is conducted on a file of antenna information to determine the terminal power loss. This information is stored in the form of a four-dimensional matrix of which the dimensions represent types of equipment, type of antenna, harmonic, and measured power loss. When the power loss has been located for both transmitter and receiver antennas, these numbers are subtracted from the transmitter power output to provide a more realistic power input to the antenna terminals.

d. Antenna Pattern Loss:

This portion of the program accounts for the loss of power due to the transmitter and receiver antenna.

A search is conducted on an antenna pattern file to find the antenna pattern information associated with both the transmitter and receiver antenna. This information has been compiled as the result of field measurements and appears in the form of a three-dimensional array of elements whose dimensions are azimuth angle, elevation angle, and loss. When the correct antenna pattern loss information has been extracted from the file, the angles that determine the orientation of the transmitter and receiver antennas are calculated and the antenna pattern loss factor is extracted from the array of elements and subtracted from the transmitter power output.

e. Propagation Loss. Propagation loss from transmitter to receiver is calculated on the basis of the Longley-Rice model (Ref. 5).

f. Receiver Function, Scoring Function, and Output:

At this point, the power level of the signal is compared with the corresponding receiver threshold-sensitivity value. If the signal strength is less than the receiver sensitivity, the signal is discarded, a new transmitter frequency is chosen, and the complete process is repeated. If the signal strength is greater than the receiver sensitivity, two things occur: (1) signal strength and transmitter-receiver identifying information are printed out to permit analysis, and (2) the frequency and power level of the signal are stored in the computer memory to be used in an interference scoring process (Ref. 8).

When all signals affecting a particular receiver have been determined, a test is made to determine if the receiver has been saturated due to the presence of the desired and undesired signals. The saturation level is determined from a measured "dynamic range". If the sum of the power input to the receiver exceeds the saturation point, a condition of complete interference is assumed, the demodulator and scoring portion of the program is omitted, and a new receiver is chosen from the environment for analysis.

If the saturation point has not been exceeded, the response of the demodulator to the signal in the IF stage is determined and a spectral analysis is made. The scoring process is accomplished by dividing the audio spectrum into increments of unequal width in frequency, but each of which contributes equally to the intelligibility according to the French and Steinberg theory. Signal-to-noise ratio in each band is computed and an articulation index determined (Ref. 9).

6-4.3.5 Availability

Information on the application of the IPM can be obtained from Cdr, US Army Electronic Proving Ground, Attn: Code STEEP-MT-M, Ft. Huachuca, AZ 85613.

6-4.4 ELECTROMAGNETIC COMPATIBILITY ANALYSIS CENTER (ECAC)

6-4.4.1 General

A tabulation of various analytical models which have been programmed for automated calculation are shown in Table 6-7. Table 6-8 lists programs that are used primarily to retrieve specific information from the data files described in par. 6-3.2

Representative applications of use of data files and models are given in the paragraphs that follow. Typical applications are grouped into three categories, which are illustrative only. There is considerable overlap of the items in each category, and there are variations in the types of problems encountered in each program. Many capabilities are applicable to frequency supportability (FS), radiation aspects of signal security (SIGSEC), and electronic counter-countermeasures (ECCM), in addition to EMC. Also, indicated parenthetically, are the technical EM areas applicable to each category.

Two of the more complex analysis programs available at ECAC are described in some detail in par. 6-4.4.5 (Model B) and par. 6-4.6.1 (COSAM).

6-4.4.2 Spectrum Utilization

Five examples of programs that relate to spectrum utilizaion are:

a. Frequency Band Allocation and Use. Determination of the frequency band to be used for a new C-E system required a study of the allocation rules for candidate bands and a survey of present and projected use of the band by other C-E systems. These studies not only identify the most appropriate band for a C-E system but are useful in support of a frequency allocation request (EMC, ECCM, SIGSEC, FS). Applicable data files and models are:

Data Files	Models
SAUF	Spectrum Occupancy Plots
NCF	Environment Analysis Systems
E File	

b. In-band EMC Assessment. An analysis of the EMC of the planned C-E system with those systems already operating in the band or planned for operation in the band frequently is conducted to preclude development of a system that is fundamentally incompatible with other systems in the band. These analyses consider the interaction on a one-to-one basis between the planned system and the others in the band, taking into account the emission and susceptibility characteristics of the systems involved (EMC, ECCM, FS). Applicable data files and models are:

Data Files	Models
E File	Emission Spectrum Models
NCF	Receiver Selectivity Models
Measured Equipment Characteristics Data	Degradation Models
OPAF	Propagation Models
Topographic	Antenna Pattern Models
	Intermodulation Models

c. Adherence to Standards and Specifications. A further requirement for a C-E system is confirmation that the system technical characteristics conform to applicable EM standards for the band and its design function. This type of evaluation is carried out initially in the processing of a frequency allocation request and further evaluated as the system design evolves to its final configuration (EMC, FS, ECCM, SIGSEC). Applicable data files are:

Data Files
Appropriate Specifications and Standards

d. Compatible Frequency Assignment Determination. A study is conducted to identify a compatible frequency assignment for a planned equipment deployment or a modification of an existing network. The ability to arrive at a compatible assignment is frequently a prerequisite for production and operational deployment of a system (EMC, FS). Applicable data files and models are:

Data Files	Models
E File	Frequency Assignment Models
Special Collections	Propagation Models
Topographic	
OPAF	

e. Evaluation of Frequency Assignment. A number of problems require the evaluation of a proposed frequency assignment for a network or an operational facility such as command post or an airfield. Capabilities have been developed which provide uniform and thorough evaluations of installations of this

TABLE 6-7 LISTING OF ECAC MODELS

1. SUBSYSTEM MODELS
   - Transmitter Emission Spectrum Synthesis
   - Transmitter Emission Spectrum Models
   - Fourier Transform of FM Trapezoidal Pulses (HTRANS)

   - Receiver Waveform Simulation Model (RWS)
   - Frequency Analyses Subroutine (FAS)
   - Receiver Response to a Family of Pulses
   - Butterworth and Chebyshev Digital Filters
   - Pulse Compression Matched Filter
   - Impedance Computation and Analysis Program (ZCAP)

   - Multiple Level Antenna Model (MLS)
   - Antenna Data Analysis Program (ANDATA)
   - Pattern Analysis Subroutine (PATAS)
   - Transmitter/Receiver Antenna Coupler (TRACE)

   - Intermodulation Analysis Model
   - Spurious Response Identification Model
   - Mixer Response Model
   - Adjacent Channel Interference Summary

2. PROPAGATION MODELS
   - MASTER PROPAGATION SYSTEM.
     - Terrain Integrated Rough Earth Model (TIREM)
     - Integrated Propagation System (IPS)
       - Smooth Curve Smooth Earth (SCSE)
       - Distance Free Space Spherical Reflection Field (SFSRF)
       - Simplified Theoretical Ground Wave (STGW)
       - Modified YEH Troposcatter (MYEH)
   - NLAMBDA (Nλ) Ground Wave Model
   - SKYWAVE HF PROPAGATION MODEL
   - COSITE COUPLING MODEL
   - PROPAGATION STATISTICS GENERATOR

3. ENVIRONMENT ANALYSIS SYSTEMS
   - Demand Analysis Programs
   - Model B
   - Pulse Density Model
   - Site Analysis Model (SAM)
   - Target Acquisition Model (TAM)
   - Power Density Display Program (PDDP)
   - IFF MARK X (SIF) Model (AIMS-PPM)
   - IFF MARK XII Model (AIMS-PPM)
   - Rapid Cull Model
   - Equipment Desnity Program

4. COSITE MODELS
   - Cosite Analysis Model (COSAM)
   - Airframe Communications Analysis (AVPAK)
   - Aircraft Ordinance RF Analysis (AVPAK 2)

5. SPECTRUM MANAGEMENT ANALYSIS MODELS (FREQUENCY ASSIGNMENT)
   - Multiple Channel Assignment Systems (MCAS)
   - LGS Angle Data Channel Assignment
   - Mutual Interference Table
   - Frequency Assignment Support Subroutine (GRAFAS)
   - FAA-ATC-VHF Frequency Model
   - Tactical Landing Force EMC Evaluation
   - Off-Frequency Rejection-Distance Calculation (OFRCAL)

6. DEGRADATION ANALYSIS MODEL
   - Degradation Analysis

7. STATISTICAL AND NUMERICAL ANALYSIS
   - Auto-Cross Correlation Analysis Model (ACCAM)
   - List Processing Routines for Digital Simulations
   - Generalized File Statistics Analyzer (Q63)
   - MATH-PACK
   - STAT-PACK
   - Random E-File Generator (REG)
   - E-File Equipment Statistics (X08K36)
   - Antenna Data Analysis System (ADAS2)
   - Topographic Data Scatter Diagram (SCATER)
   - Experimental Calculation-INR Distribution

TABLE 6-8
DATA BASE SELECT CAPABILITIES

On-Line E-File General Format and Print Program
On-Line E-File General Select Capability
On-Line E-File (SPACE)
The C&E Deployment System
Topographic Data File Select and Print Program (TOPSEL)
Profile Print Program
Operational Platform Allowance File (OPAF)
General Format and Print Program
Nominal Characteristics File (NCF) Equipment Selection Programs
Propagation Measurement Retrieval System (PMRS)

type (EMC, FS). Applicable data files and models are:

Data Files	Models
NCF	Frequency Assignment Models
Measured Equipment Characteristics Data	Cosite Models
Special Collection	

6-4.4.3 System EMC Analyses

Analyses are used to determine system technical characteristics, measurement requirements and evaluations as follows:

a. Theoretical Analysis of C-E System Characteristics. There is often a need to ascertain the emission and susceptibility characteristics of a C-E system in advance of the availability of hardware for assessment of adherence to specifications and standards, and for prediction of the impact of a proposed system on those already occupying the band. Models and techniques for accomplishing this are applied to proposed designs (FS, EMC, ECCM, SIGSEC). Applicable data files and models are:

Data Files	Models
NCF	Emission Spectrum Models
Measured Equipment Characteristics Data	Receiver Selectivity Models
	Degradation
	Intermodulation Models

b. Specification of Measurement Requirements. Specific requirements and techniques must be formulated for measuring the emission and susceptibility characteristics of C-E systems. These empirical data are needed to assess the adherence to specifications, to evaluate the EMC with other systems, and to provide an input to the enhancement of equipment modeling (EMC, SIGSEC, ECCM). Applicable data files are:

<u>Data Files</u>

Appropriate Specifications and Standards

c. Evaluation of Measured Characteristics. Measured data for C-E system emission and susceptibility must be evaluated to validate theoretical modeling of these characteristics and to determine the contribution of spurious and nondesign emission and susceptibility to the overall characteristics of the system (EMC, SIGSEC, ECCM). Applicable data files are:

<u>Data Files</u>

Appropriate Specifications and Standards

6-4.4.4 EMC Consultation and Guidance to Research, Development and Engineering

ECAC also will provide technical consulting as follows:

a. Recommend EM Provisions for Requests for Proposals (RFP's). Review of RFP's in the drafting phase has resulted in the incorporation of provisions that alert prospective bidders to the EMC considerations that must be made in the design and development of the equipment to be acquired (EMC, ECCM, SIGSEC).

b. EM Evaluation of Proposals and Equipment Specifications. The proposals and specifications are reviewed for consideration of all EM aspects in the equipment development cycle. Recommendations are provided to the Program Manager for measures to insure compatible operation of the systems being acquired (EMC, ECCM, SIGSEC, FS).

6-4.4.5 Model B

The Model-B system is a computer model for the processing of large environments in compatibility analysis. The program incorporates various models listed in Table 6-7 and is used for two basic purposes:

a. To assess the interference effects produced by a surrounding transmitter environment at a given receiver, or

b. To determine the interference effects produced when a new transmitter is introduced into a receiver environment.

The quantities desired as program outputs usually are either an interference-to-noise ratio R_u at the receiver or the power density P_d produced at some location by surrounding transmitters. R_u is obtained from the following equation (cf. Eq. 6-5):

$$R_u = P_t + G_t - L_p - L_f + G_r - R_s - 20\log(f_r/f_t),\ \text{dB} \qquad (6\text{-}19)$$

Each element of Eq. 6-19 is calculated with an individual program module, then they are summed for high-speed printer output.

A cull threshold can be specified by the user. Any coupling that produces R_u below the threshold is not included in the printed output.

The incident power density P_d produced at a given location by a transmitter is given by:

$$P_d = P_t + G_t - L_p + 20\log f_t - 38.5,\ \text{dBW/m}^2 \qquad (6\text{-}20)$$

where f_t is the transmitter frequency in MHz (low end of tuning range if range-tuned).

The user can specify a power density cull threshold. Those couplings that produce a power density value below the cull are not included in the printed output.

6-4.4.5.1 Environment File Processing

The input processor selects equipment records from any of the data files based on user input constraints.

Environment records with any or all of the following options can be selected:

a. Equipment type (transmitter or receiver)
b. Frequency range
c. Location (within a given radius of a selected point).

Certain fields in the data base may be missing for various reasons. Where a missing value may be assumed or calculated with a fair degree of accuracy, a procedure is used to replace it in the record. The following fields required in R_u and P_d calculations are recoverable:

a. Antenna structural height
b. Antenna gain
c. Receiver image response level
d. Transmitter harmonic suppression
e. Pulse width
f. Antenna pattern type
g. Location.

The two required parameters that follow are not contained in the data base. Fixed values are automatically assumed by the program.

a. Receiver spurious level
b. Emission spectrum or selectivity slope fall-off.

Missing data in some fields will result in the record not being processed, since recovery is not possible with the existing program. The equipment identification and the contents of the void field are printed for all such records. The fields follow:

a. Tuned frequency
b. Transmitter power
c. Transmitter bandwidth
d. Receiver sensitivity
e. Receiver bandwidth
f. Receiver IF frequency.

The equipment records that pass the input constraints are stored in the computer, and complete pairs (one transmitter, one receiver) are formed and analyzed in turn.

6-4.4.5.2 Analysis Models

6-4.4.5.2.1 Transmitter Power

This parameter P_t is carried in the equipment record in kilowatts and is converted to dBW for use in the R_u equation:

$$P_t = 30 + 10\log(P_t'),\ \text{dBW} \qquad (6\text{-}21)$$

where

P_t' = power, W

6-4.4.5.2.2 Transmitter and Receiver Antenna Gain

The Multilevel Synthesis Model (MLS) was developed from a statistical analysis of measured antenna pattern data. An antenna pattern is synthesized consisting of up to four attenuation levels (number of levels in a function of main beam gain), the widths of which are a function of the antenna horizontal beamwidth (see Fig. 6-2). Antenna A is a high gain fixed directional antenna oriented along a 155 deg azimuth, and Antenna B is a lower gain fixed directional antenna oriented at 35 deg azimuth.

The mutual gain between the two is found by adding the gain of level (3) for Antenna A and the gain of level (3) for Antenna B.

In the case of omnidirectional or rotating antennas the mainbeam gain is used for that antenna. Sector scanning antennas are synthesized in the same manner as fixed directional antennas, except that the scan limits are taken into account when determining which gain level to use.

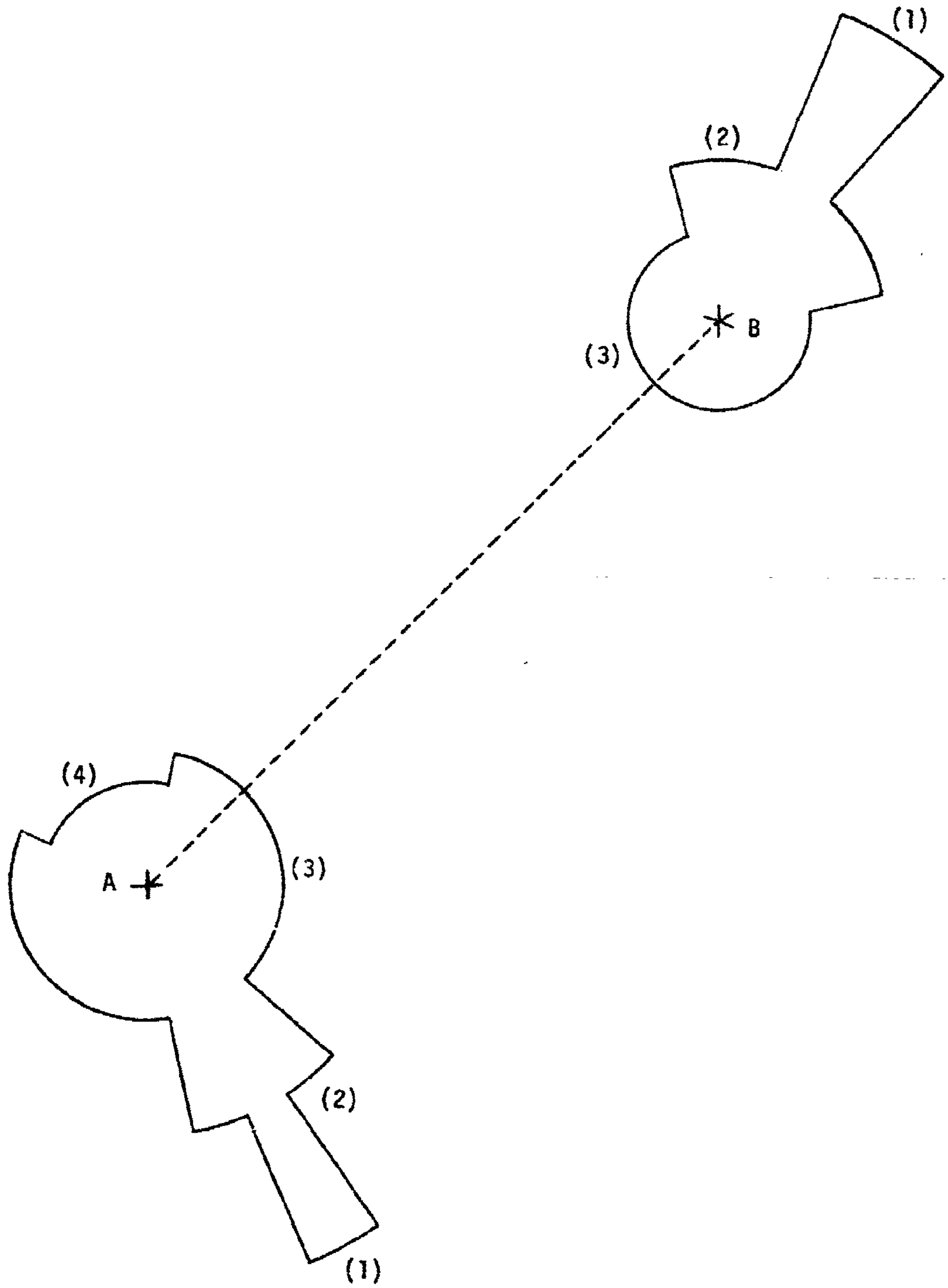


*Figure 6-2. Typical MLS Antenna Pattern Synthesis

6-4.4.5.2.3 Propagation Path Loss

There are two methods of path loss computation: one assumes that the earth is a smooth sphere with a 4/3 earth effective radius (smooth earth); the other includes the effects of intervening terrain (rough earth).

6-4.4.5.2.3.1 Smooth Earth Path Loss Model

The primary smooth earth model is the N-LAMBDA Program. This is a ground-wave model with a valid frequency range of 0.01 MHz to 20 GHz. It consists of three sections, with the choice of section depending on the path geometry for the given problem. The three paths are:

a. Line-of-sight analysis using the direct and reflected rays for propagation over a spherical earth with medium-height antennas

b. Plane earth surface wave analysis for low antennas

c. Beyond line-of-sight diffraction analysis over a spherical earth, with a check of possible tropospheric scatter propagation.

The program selects the proper mode from the path characteristics and returns a path loss.

6-4.4.5.2.3.2 Rough Earth Path Loss Model

The Terrain Integrated Rough Earth Model (TIREM) is used to obtain a terrain-dependent path loss. This model first extracts the topographic profile between the two equipments concerned from a digitized terrain file. It next analyzes the profile geometry and selects one of the following propagation modes:

a. Free space
b. Knife edge beyond line-of-sight
c. Effective double knife-edge
d. Rough earth diffraction
e. Troposcatter
f. Diffraction-troposcatter.

The propagation path loss is then computed using the equations for the selected mode.

6-4.4.5.2.4 Off-Frequency Rejection

The Frequency Analysis System (FAS) is used to calculate off-frequency rejection. It accounts for image and spurious responses in the receiver, and up to eight harmonics from the transmitter.

The module considers all the possible frequency interactions (such as fundamental-fundamental and harmonic-image) by integrating the emission spectrum and receiver selectivity curves to determine the total power coupled into the receiver for each case. This power is then converted to a loss below the co-channel value, which may be different from zero if the receiver bandwidth is less than the transmitter bandwidth. The interaction producing off-frequency rejection is selected.

6-4.4.5.2.5 Receiver Sensitivity

This parameter is available directly from the data files in dBm units.

6-4.4.6 Availability

Information may be obtained from the Army Deputy Director, Electromagnetic Compatibility Analysis Center, North Severn, Annapolis, MD 21402.

6-4.5 INTERFERENCE PREDICTION PROCESS-NUMBER 1

6-4.5.1 Introduction

The Interference Prediction Process-Number 1 (IPP-1) (Ref. 10) was developed at the Rome Air Development Center (RADC), Griffiss Air Force Base, NY. Its purpose is to analyze and predict potentially interfering situations among a proposed or existing deployment of transmitters and receivers. It is useful for:

a. Developing preliminary equipment or system requirements and specifications

b. Preparing test plans for specification compliance

c. Evaluating test results

d. Revising specifications or equipments for conditions of noncompliance

e. Evaluating systems in specific operating environments.

IPP-1 was programmed initially for the Burroughs 205 computer and the Univac 1103A computer. A more complete and detailed set of programs was then developed for used in the CDC 1604 computer.

6-4.5.2 Analysis Process

An interference margin P, which corresponds closely to the negative of the interference-to-noise ratio R_u (in dB), is calculated in terms of R_s, the receiver sensitivity at its input terminals; thus

$$P = (P_t + G_t + G_r - L_p - R_s)\ \text{, dBW} \qquad (6\text{-}22)$$

The actual interference analysis is divided into two parts: the rapid cull, as its name implies, uses some very simple criteria for eliminating all of the more obviously noninterfering portions of the problem. For all cases eliminated by the rapid cull, there is no probability of interference. For the remaining cases, there is at least a small probability that interference occurs. The frequency cull considers the frequency separation between potentially interfering transmitter outputs and receiver responses. It also considers the statistical distribution of all of the amplitudes involved and the criteria for interference to determine the significant probabilities of interference.

The basic portions of the analysis are carried out during the frequency cull. Major steps of the frequency cull analysis are outlined in the block diagram of Fig. 6-3. The analysis begins with a statistical expression for the transmitter output amplitudes and an independent tabulation of the specific transmitter output frequencies over the range which was determined from the rapid culling step. The transmitter amplitude function includes the statistical effects of the coupling system between the transmitter and the transmitting antenna. The transmitter amplitude function is modified successively by the pattern distribution function (PDF) for the transmitting antenna, the intervening propagation conditions, and the pattern distribution function for the receiving antenna. The output frequencies from the transmitter are unmodified during these analysis steps, but serve as references to the basic frequency range over which each of the factors must be considered.

The statistical description of the receiver response amplitudes includes the statistical effects of the coupling system between the receiving antenna and the receiver. At the next step in the analysis which begins at the lower left of Fig. 6-3, the frequency information is merged with the amplitude information. The receiver response frequencies are compared with the transmitter output frequencies to obtain the frequency difference between suspect output-response frequency pairs.

The frequency analysis is now joined with the amplitude analysis for the remaining steps. The energy distribution about each suspect transmitter output is considered along with the selectivity about each suspect receiver response to determine the amount of interfering energy that enters the receiver. The statistical distribution of the "interference margin" is then tabulated.

The program begins with three basic catalogs on magnetic tape. The catalogs are:

a. The Transmitter File, which contains all pertinent transmitter information that is required in performing the interference analysis. This information includes nomenclature and general transmitter data (frequency range, type of modulation, etc.), describing the multilist representation of the transmitter output envelopes, and possibly exact outputs from the transmitter which may be pertinent to the analysis.

b. The Receiver File, which contains all pertinent receiver information that is required in performing the interference analysis. This information is analogous to the information provided in the transmitter file.

c. The Antenna File, which contains general technical characteristics of the antenna (including nomenclature, height, length, frequency range, etc.) and detailed pattern distribution data.

Also provided initially to the computer when performing an analysis are such specific problem data as equipment coordinates, terrain characteristics between specific pieces of equipment, antenna orientation and modes of operation, and the specific tuned frequencies of each piece of equipment.

6-4.5.3 Data Outputs

The following illustration shows how this program can be used when installing a new transmitter into an existing equipment complex (Ref. 11). An equipment complex of three radar receivers and one UHF receiver is assumed. The deployment within the site is shown in Fig. 6-4. The transmitter and the receivers are designated T-1 and R-1 through R-4, respectively, and their characteristics are summarized in Table 6-9.

The results obtained for one of the transmitter-receiver pairs show potential interference as follows:

a. The fundamental output of the transmitter with the fundamental response of the R-1 receiver

b. The second harmonic of the transmitter with the fundamental response of the R-2 receiver

c. The fundamental output of the transmitter with the image response of the R-3 receiver.

6-4.5.4 Availability

The IPP-1 normally is used for internal work but may be used for external programs if time is available; Rome Air Development Center, Electronics System Division, Code RADC/RBCT, Griffiss AFB, NY 13441 should be contacted.

6-4.6 COSITE ANALYSIS MODEL

6-4.6.1 Introduction

The Cosite Analysis Model (COSAM) (Ref. 12) is an automated program designed to evaluate the electromagnetic compatibility of a single site that uses a large number of transmitting and receiving equipments in the 200- to 400-MHz band, with AM, FM, and SSB modulations. It can handle up to 50 single channel transmitters and 50 single channel receivers.

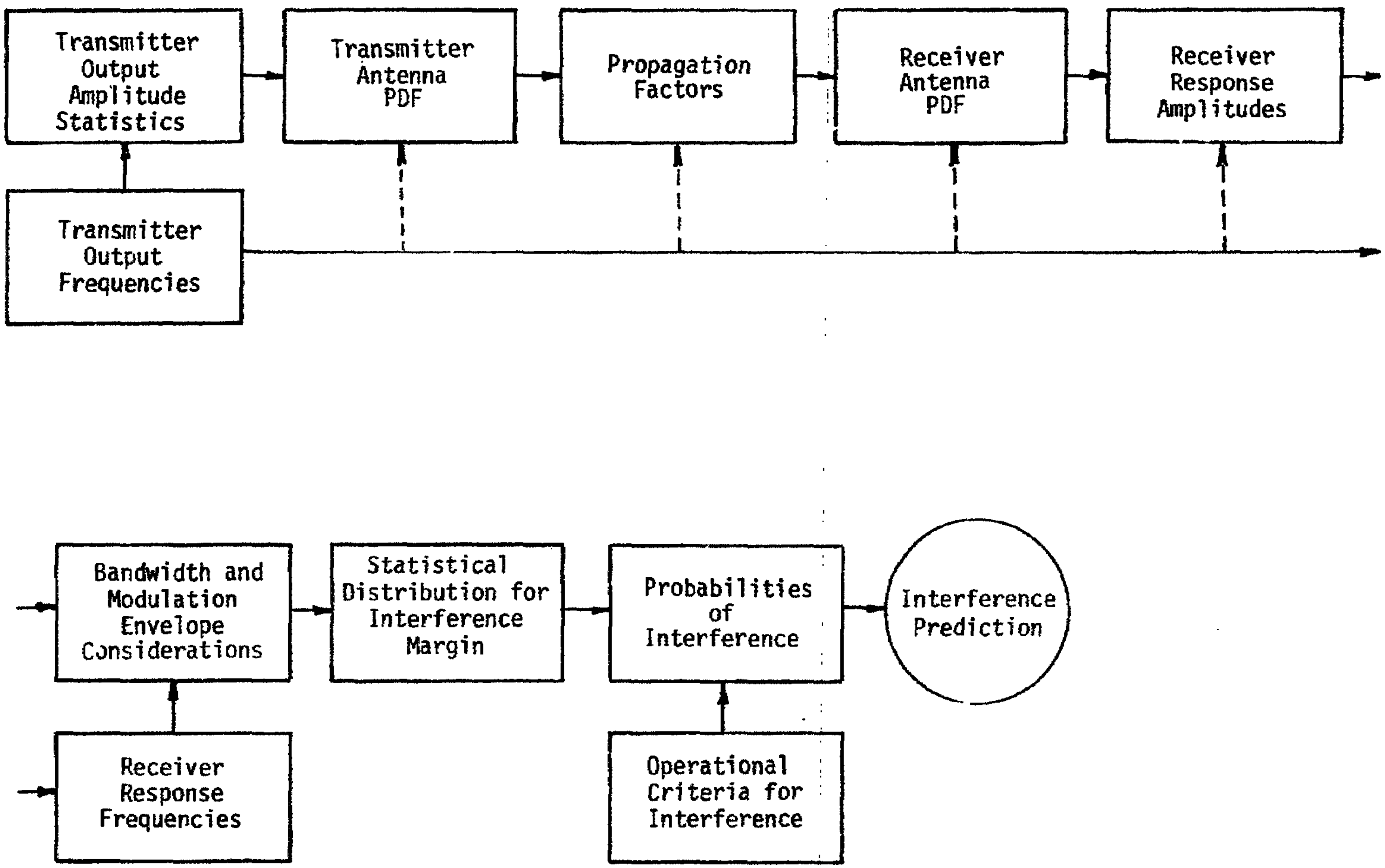


Figure 6-3. Frequency Cull Steps of Interference Prediction

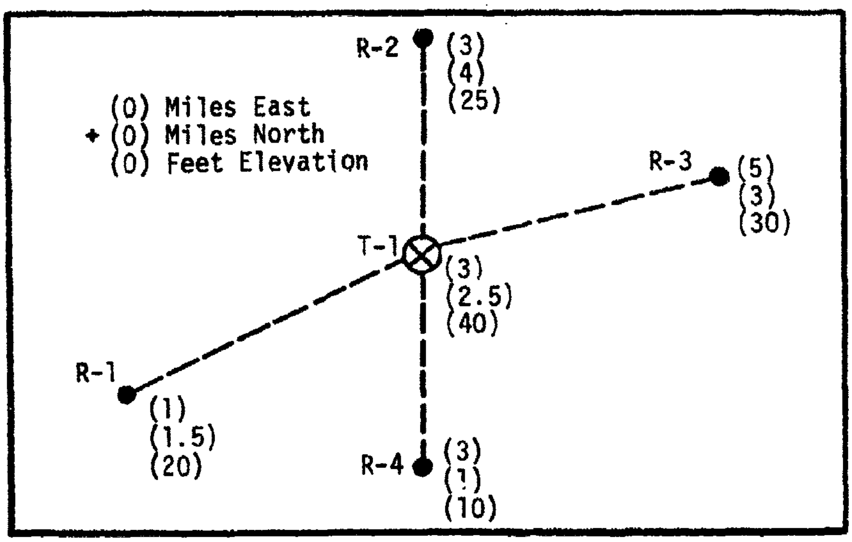


Figure 6-4. Coordinate System for Installing a Transmitter

TABLE 6-9. EQUIPMENT CHARACTERISTICS

Characteristic	T-1	R-1	R-2	R-3	R-4
Frequency Band, MHz	2800	2800-3100	5450-5825	2700-2900	225-399
Tuned Freq. f_t, MHz	2800	2830	5600	2740	275
Power or Sensitivity, dBm	+97	−100	−100	−102	−92
Antenna Gain, dB	35	39	33	32	7
Antenna Height, ft	40	29	25	30	10
Antenna Bearing, deg/North	*	*	45	325	**
East Coordinate, mi	3	1	3	5	3
North Coordinate, mi	2.5	1.5	4	3	1
Spurious Characteristics ($f > f_t$)					
Slope, dB/decade	−30	20	20	10	37
Intercept, dB above Nominal	−40	76	76	56	57
Standard Deviation, dB	10	12	12	12	12

*Denotes rotating antenna
**Denotes omnidirectional antenna

A cosite EMC problem implies (a) "close" proximity between transmitting and receiving antennas, and (b) relatively "large" signals impinging on receiver inputs (as well as transmitter outputs). In general, five types of interactions are considered: (1) adjacent signal, (2) spurious emissions, (3) spurious responses, (4) receiver intermodulation, and (5) transmitter intermodulation.

Because of uncertainties in defining the amplitudes of the interactions, a statistical approach is used in analyzing these interactions. Furthermore, several interactions are likely to occur simultaneously, requiring a method for assessing multiple affects.

The program is written in FORTRAN, and requires at least 32 words of memory. The program is run on a Univac 1110 computer at the DoD Electromagnetic Compatibility Analysis Center (ECAC), Annapolis, MD.

Inputs consist of a list of equipments involved and their operating frequencies, antenna types, and their location on the site. Applications have included UHF-AM transmitters (Ref. 13) and VHF-FM and AM transmitters and receivers. Antenna types are Yagi, helix, and corner reflector types as well as loaded and nonloaded whip antennas and vertical ground plane antennas. Equipment parameters are obtained from a combination of nominal characteristics and spectrum signature data. Various automated, semiautomated, and manual procedures are used to compute the parameters. Where appropriate, the parameters are described by means and standard deviations which have been derived from measured data.

Outputs consist of (1) performance scores derived from predicted S/N and $S/(I + N)$ output distributions, and (2) causes of significant degradation, if any, which are identified for each receiver in quantitative terms; i.e., mean values of effective input on-frequency power levels due to each significant interaction.

6-4.6.2 Program Models

The models used in the program are for adjacent signal, noise, spurious response and emission, and intermodulation.

6-4.6.2.1 Adjacent Signal Model

A review of measured data and practical experience indicates that beyond a certain deviation (depending on equipment type) from the receiver tuned frequency, adjacent signal interference is not significant. Therefore, a maximum deviation, e.g., 20 MHz, is postulated, above which serious operational degradation is not anticipated. For each receiver, then, all transmitters whose frequencies are within 20 MHz of the receiver tuned frequency are considered.

6-4.6.2.2 Noise Model

The noise model is used in conjunction with all other submodels. If no interaction is attributed to any specific transmitter or groups of transmitters, the primary interference mechanism is assumed to be noise. Consequently, a predicted noise interaction, in effect, indicates "no degradation" for all types of interactions.

6-4.6.2.3 Spurious Emission and Response

These responses are identified in frequency with respect to the tuned frequency. For transmitters spurious emission may arise from harmonics of a crystal frequency f_c. In this case the computer determines if the harmonics occur at spurious response frequencies and if so stores them in an interaction table.

In a corresponding way, significant receiver spurious response frequencies are computed in terms of receiver local oscillator and intermediate frequencies and stored.

6-4.6.2.4 Intermodulation

Transmitter or receiver intermodulation can occur when a victim transmitter or receiver tuned to the frequency f_r also is subjected to a strong signal from an interfering transmitter of frequency f_i related to f_r as discussed in par. 3-2.2.2.2.

An interaction table is developed to reduce or "cull" the problem. In the interest of conserving computer time, not every possible interaction is computed. Rules are established to eliminate interactions which are clearly insignificant. The interaction table lists, for each receiver, every transmitter or transmitter pair which may cause a significant interaction. Power levels are not considered at this point. The table may be printed prior to the final computation if desired.

6-4.6.3 Scoring Techniques

Three scores are provided based upon ratios of signal-to-noise $(S/N)_O$ and signal-to-interference plus noise, $[S/(I + N)]_O$ as illustrated in Figs. 6-5 and 6-6, where $(S/N)_{OT}$ and $[S/(I + N)]_{OT}$ are, respectively, output signal-to-noise and output signal-to-interference plus noise threshold value ratios (SINAD). The upper performance score *UPS* is the probability of providing "adequate" or "good" performance with no interference present. The system performance score *SPS* is the probability of adequate (or good) performance in the presence of interference.

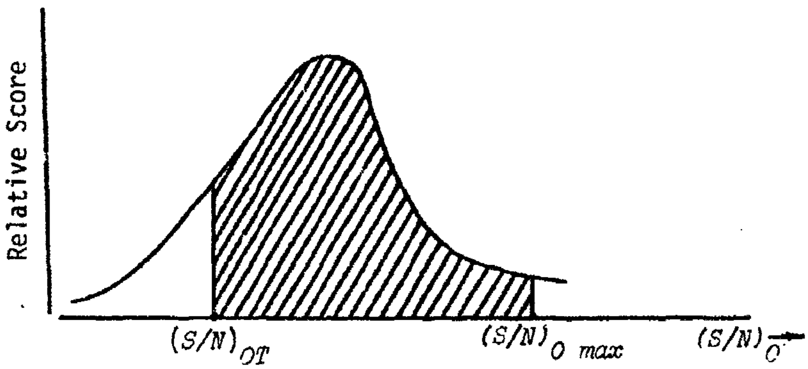


Figure 6-5. Upper Performance Score (UPS)

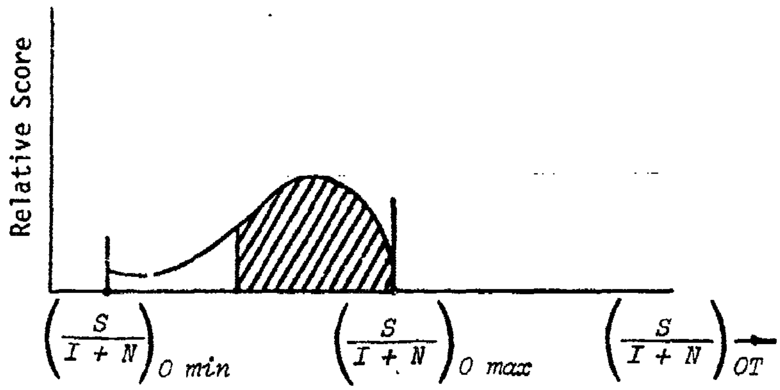


Figure 6-6. System Performance Score (SPS)

The relative performance *RPS* score *SPS*/*UPS* provides the user with a third measure which, in conjunction with the other scores, gives additional understanding of receiver performance. For example, if the *SPS* were 0.4, one would expect poor performance. However, if the *UPS* were also 0.4, *RPS* = 1.0 and an inadequate desired signal thus is identified.

6-4.6.4 Comparison of Measurements and Predictions

Measurements were made of numerous interactions at an installation involving 6 AM voice communications transceivers operating in the 225-400 MHz frequency range (Refs. 12, 13). Twenty-five frequency assignments were tested with 3 different desired signal levels. Major results were:

a. A comparison of 90 measured coupling values and associated predicted mean values resulted in an average difference of less than 1 dB, with a standard deviation of 3.4 dB.

b. The model bias (for all 436 interactions) was 1.55 dB, implying a small tendency toward prediction of too much interference. The standard deviation of the bias was 5.3 dB.

c. An evaluation of each of the groups of interactions identified as being due to one of the specified mechanisms indicated that, for 85% of the cases, the bias values for each group were less than 3.5 dB and that standard deviation values were less than 6.2 dB.

d. As noted in Fig. 6-7, 92% of all of the cases resulted in differences between measured SINAD values S_M and associated predicted mean values $\bar{S}_p$ of less than 10 dB.

e. Results of the "Bin Method", noted in Fig. 6-8, indicated that a confidence level of 90% can be assigned to a prediction of *SPS* ± 0.225. *SPS* is the average *SPS* value for a specified bin, and SPS_m is the measured *SPS* value associated with the same bin.

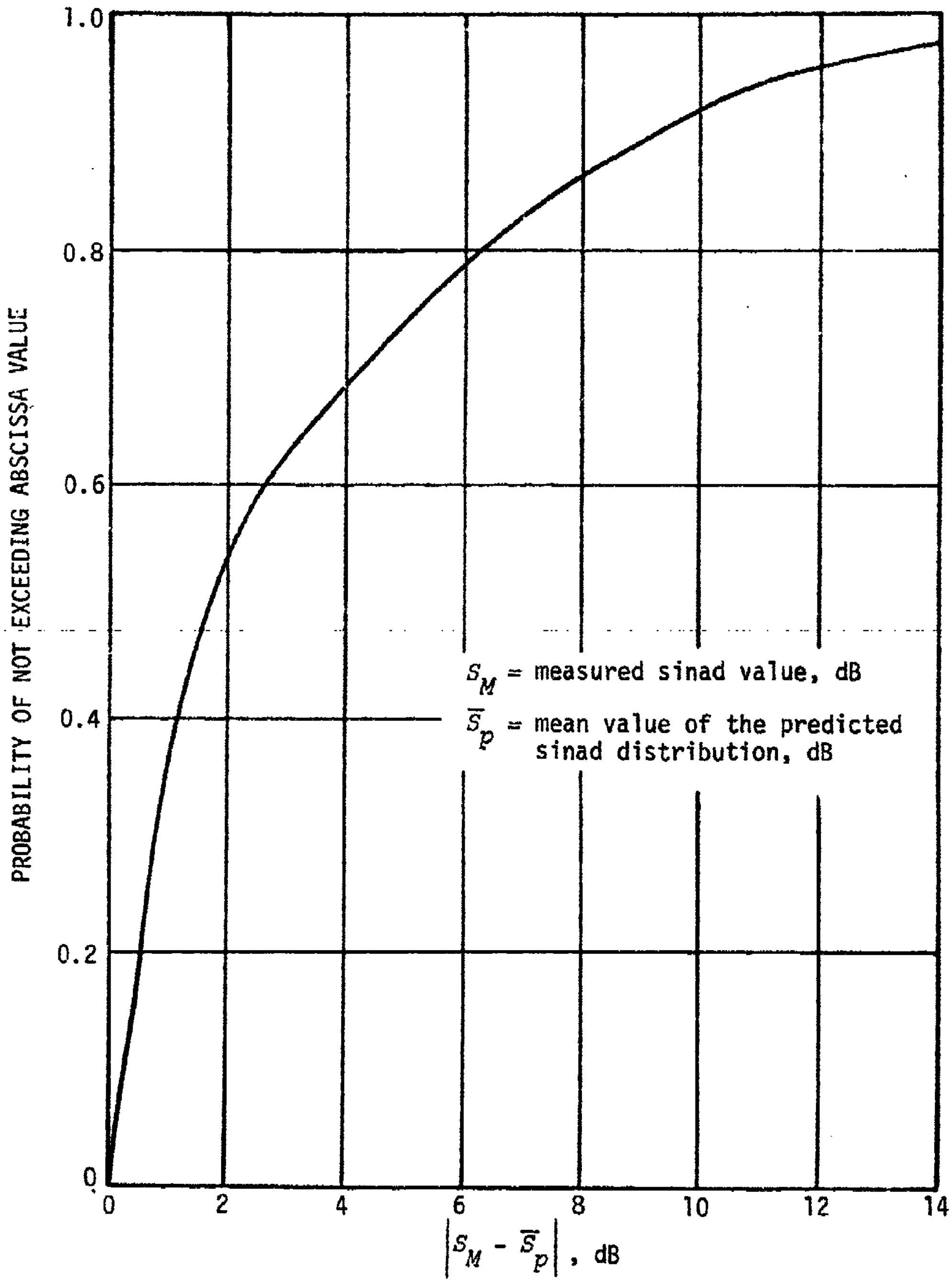


Figure 6-7. Cumulative Probability Distribution of $|S_M - \overline{S}_p|$

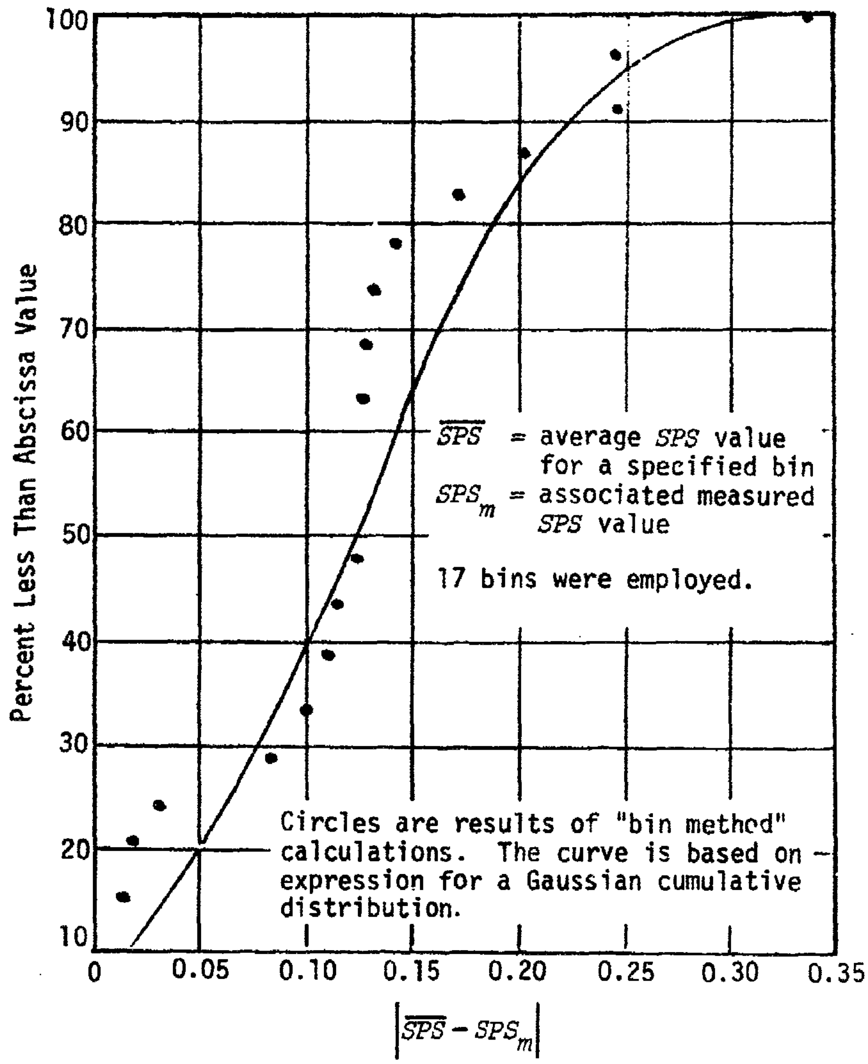


Figure 6-8. Cumulative Probability Distribution of $|\overline{SPS} - SPS_m|$

f. Results of the "interference condition method", using a 5-condition scale, indicated that results were within 1 condition for 76% of the cases and within 2 conditions for 92% of the cases.

g. By use of the results of the coarser "interference condition method", the probability of a prediction resulting in a gross error is less than 0.08. (A gross error is defined as a situation where a prediction will indicate good performance when a measurement indicates intolerable degradation, or the converse situation.)

6-4.6.5 Availability

Information may be obtained from the Army Deputy Director, Electromagnetic Compatibility Analysis Center, North Severn, Annapolis, MD 21402.

6.4.7 SHIPBOARD ELECTROMAGNETIC COMPATIBILITY ANALYSIS AND SHIPBOARD ELECTROMAGNETIC COMPATIBILITY ANALYSIS MICROWAVE (NAVSEA)

SEMCA is the acronym for Shipboard Electromagnetic Compatibility Analysis (Ref. 14). One program (cull 3) performs a basic analysis in which transmitter output spectra are processed through designated transmitter channel components (i.e., guides, antenna tuners, and antennas), the antenna coupling loss between all transmitting and receiving antennas is calculated, and the resulting spectrum that arrives at each receiving antenna is determined. From this information, plus data on the receiver RF channel components (i.e., filters, RF section, etc.), the potential interference spectrum at the first mixer of each receiver is determined. A second program (cull 4) then processes the resulting data through the modeled mixer and IF stage combinations to the detector.

A separate SEMCA program known as Frequency Selection is used to assign compatible transmitter and receiver center frequencies in the LF through UHF frequency range. Frequency assignment is based on a set of criteria to minimize potential interference. These criteria specify the separation between spectral lines emitted by the transmitters. The amount of separation is determined by whether the spectral lines correspond to fundamental or harmonic frequencies. If the user desires, he can insert transmitter and/or receiver frequencies, called priority frequencies, that are not to be considered as assignable frequencies.

Fig. 6-9 is a sample output sheet for the Frequency Selection Program. In the exercise of this program, the input (not shown) had called for ten transmitter frequencies to be assigned; however, there was only spectrum available for nine, as indicated by the output comment at the top of the sheet. Under the caption "Row 1" the 9 available frequencies are printed. The priority frequencies that had been included in the input data are printed next for reference purposes. Then the receiver frequencies are printed. A group of frequencies appears for each row of the input data, which was as follows:

a. List frequencies between 2 and 15 MHz that are separated from transmitter frequencies by 1.0 MHz. The results indicated 70 frequencies to be available with a channel separation of 20 kHz.

b. List frequencies between 15 and 22.5 MHz that are separated from transmitter frequencies by 2.0 MHz. The results indicated 73 frequencies to be available.

c. List frequencies between 22.5 and 30 MHz that are separated from transmitter frequencies by 2.0 MHz. The results indicated 176 frequencies to be available.

d. List frequencies between 2 and 15 MHz that are separated from transmitter frequencies by 0.5 MHz. The results indicated 91 frequencies to be available.

e. List frequencies between 15 and 30 MHz that are separated from transmitter frequencies by 0.5 MHz. The results showed 199 frequencies, the maximum the program will handle.

SEMCAM (Shipboard Electromagnetic Compatibility Analysis Microwave) is a computer program developed to aid in assessing the electromagnetic compatibility between microwave transmitting and receiving equipments employed on the topside region of Naval combatant ships. Some of the more important applications of the program are:

a. Topside design of new ships
b. Assess EMC impact of alternate equipment sites
c. Assess EMC merits of alternate antenna sites
d. Frequency assignment
e. Assess EMI "quick fix" merits
f. Aid in generation of EMC specifications for new equipment design.

The model, operating from a user-prepared list of transmitters, receivers and antennas, calculates the interaction between each microwave transmitter/receiver pair for various relative tuning situations and antenna orientations. Several output parameters, which relate to the degree of interaction and degradation, are produced. The two primary parameters are receiver output interference-to-noise ratio *INR* and mixer input "burnout" power. The *INR* parameter is an indication of the degree of receiver degradation. For example, for search radars the *INR* value, when processed through the radar signal processing model, leads to a performance degradation index which can be related to a reduction in detection range. For fire-control radars the *INR* value, processed through the radars signal processing model, leads to a performance degradation index which can be translated to a reduction in acquisition time, etc. The mixer "burnout" power can be used in several ways: (1) It permits determination of possible degradation of the mixer crystal diodes due to high power effects, and (2) the power level in the mixer can be used to assess the likelihood of spurious responses or receiver desensitization.

Significant features of the model are the antenna-to-antenna coupling functions. These functions are based on an elaborate, measured data base representing several thousand data samples. The coupling model accounts for effects such as obstacle blockage,

ONLY 9 FREQUENCIES ASSIGNED TO ROW 1

ASSIGNED TRANSMITTER FREQUENCIES FOR

ROW 1

4.120
6.472
9.414
14.118
16.053
18.252
20.751
23.591
26.818

PRIORITY TRANSMITTER FREQUENCIES

2.000
3.500
5.500
5.000
8.000
12.000

RX FREQS. -- LOWER AND UPPER FREQ. LIMITS OF CHANNELS

NUMBER OF RX CHANNELS AVAILABLE = 70 WITH SEPARATION 1.00E-02

10.420	11.000
13.010	13.110

NUMBER OF RX CHANNELS AVAILABLE = 73 WITH SEPARATION 2.00E-02

17.060	17.240
19.260	19.740
21.760	22.500

NUMBER OF RX CHANNELS AVAILABLE = 176 WITH SEPARATION 2.00E-02

22.500	22.580
24.600	25.800
27,820	30.000

NUMBER OF RX CHANNELS AVAILABLE = 91 WITH SEPARATION 5.00E-02

2.500	2.950
7.000	7.450
8.500	8.900
9.950	11.500
12.550	13.600
14.650	15.000

NUMBER OF RX CHANNELS AVAILABLE = 199 WITH SEPARATION 5.00E-02

15.000	15.550
16.600	17.750
18.800	20.250
21.300	23.050
24.100	26.300
27.350	29.900

Figure 6-9. Sample Output

out-of-band antenna operations, and dissimilar antenna polarization.

The entire microwave equipment complement of a large combat ship can be completely characterized with only a nominal amount of computer expenditure and time. Information on the use of these programs can be obtained from Naval Ship Engineering Center, Naval Sea Systems Command, Code 6174, Hyattsville, MD 20782.

6-4.8 SPECIFICATION AND ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM

SEMCAP is an acronym for Specification and Electromagnetic Compatibility Analysis Program (Refs. 15 and 16). It is a computer program for:

a. Analyzing the electromagnetic compatibility of a system

b. Developing black box EMC specification generation and susceptibility limits consistent with desired signal requirements

c. Evaluating waiver proposals in terms of their effects on system performance.

6-4.8.1 Coupling Models

Electromagnetic transfer functions are defined for wire-to-wire, wire-to-field, and field-to-wire configurations. To account for the effects of antennas within a system, antenna-to-antenna and wire-to-antenna transfer functions also are included (see Fig. 6-10).

The susceptor models include linear filters to control bandwidth, and an integration process to perform spectral density integration. Threshold levels are established that represent the expected malfunction level of given susceptor circuits.

6-4.8.1.1 Wire-to-Wire Coupling

The wire-to-wire, or close coupled, transfer functions account for capacitive and inductive couplings, as shown by parameters TV and TI, respectively, in Fig. 6-10. They relate the voltage induced in a load connected to a wire to the spectral voltages and currents present on the wire. These models take into account many wire configurations, and can include shielding and wire twist effects over a frequency range from 10 Hz to 10 GHz.

Wire-to-wire transfer is treated as two additive effects: "capacitive transfer" and "inductive transfer". This approach allows the shielding analyses for the two effects to be handled separately.

6-4.8.1.1.1 Capacitive Transfer

Close-coupled capacitive transfer models have been developed for unshielded wires over a ground plane, for shielded wires over a ground plane, and for two-wire circuits.

6-4.8.1.1.1.1 Unshielded Wires

For unshielded wires (Fig. 6-11(A)) the transfer function TV is (see Eq. 3-124)

$$TV = \frac{V_2}{V_1} = \frac{1}{1 + \dfrac{1}{2\pi f R_{a2} C_{w1w2}}} \qquad (6\text{-}23)$$

where

V_1 = generator voltage, V

V_2 = voltage appearing on susceptor wire relative to ground, V

$R_{a2} = \dfrac{R_{g2} R_{L2}}{R_{g2} + R_{L2}}$, Ω

R_{g2} = source resistance of susceptor circuit, Ω

R_{L2} = load resistance of susceptor circuit, Ω

C_{w1w2} = wire-to-wire capacitance between generator circuit and susceptor circuit, F

f = frequency, Hz

The calculation of C_{w1w2} is complex because of the proximity of the ground plane. However, it has been established that the equation with no ground plane effect is adequate for the proximities anticipated, and this relationship is used in the model. The equation is

$$C_{w1w2} = \left(\frac{3.5 \times 10^{-10}}{2\pi}\right)\left(l_{12} + 10^{-6}\right) \Big/ \ln\left(\frac{d_{12}^2}{r_{w1} r_{w2}}\right), \quad \text{F} \qquad (6\text{-}24)$$

where

l_{12} = common length of wires 1 and 2, m

d_{12} = spacing between wires 1 and 2, m

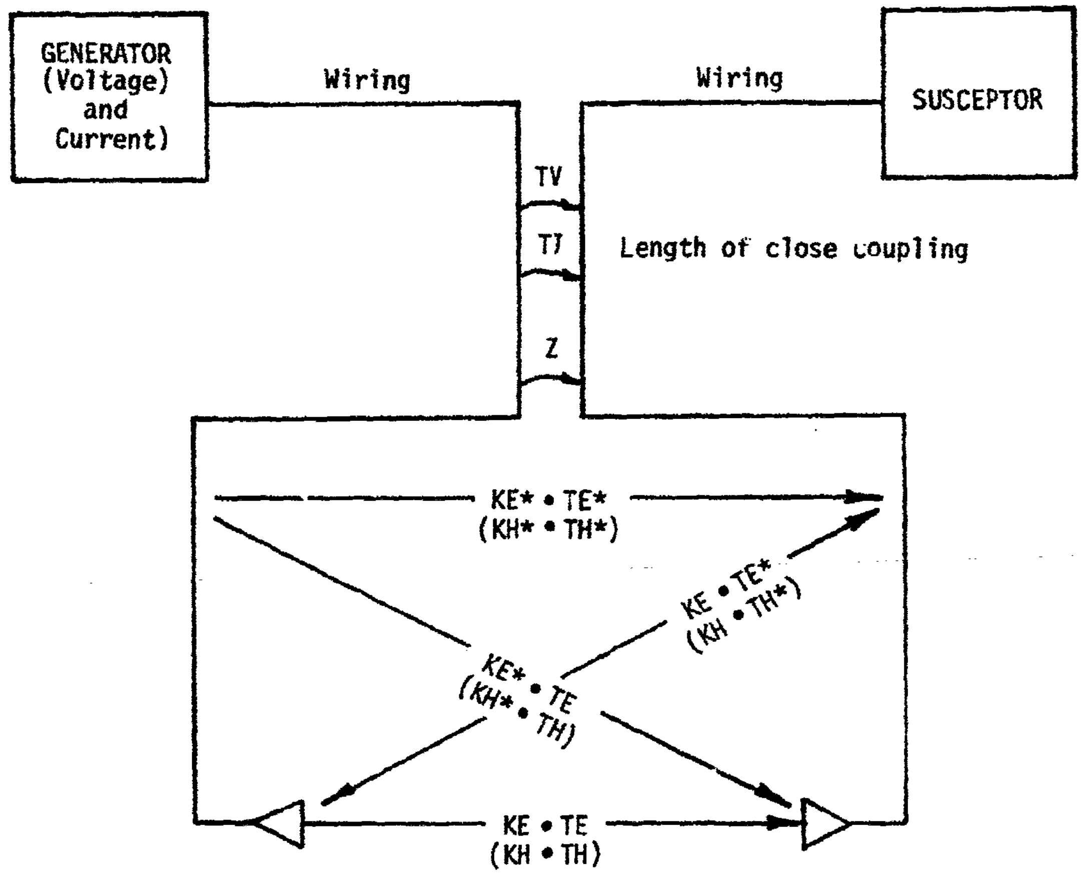
r_{w1} = radius of wire 1, m

r_{w2} = radius of wire 2, m

Eq. 6-23 is an approximation that avoids a square-root computation, thereby reducing computation time. The maximum error incurred is a factor of 1.4 (3.0 dB), and occurs only over a narrow frequency range in which the second term of the denominator is approximately equal to unity.

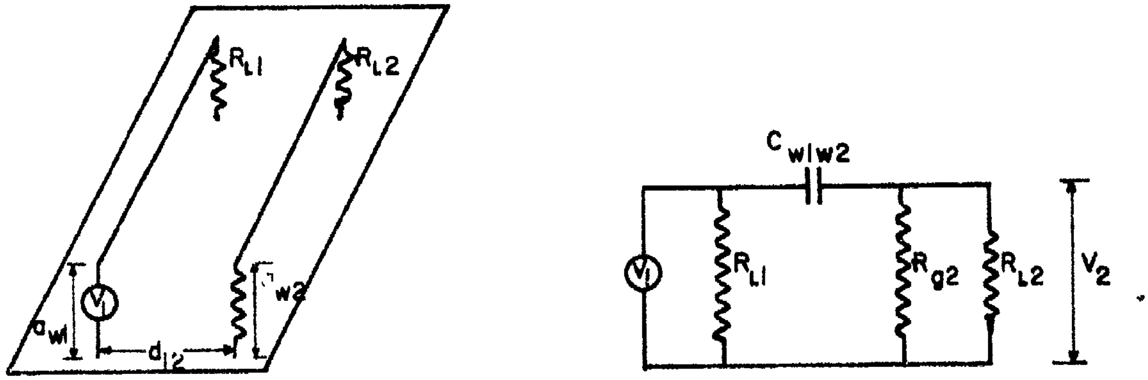
6-4.8.1.1.1.2 Two-Wire Circuits

The coupling between nontwisted two-wire circuits is given by the equivalent one-wire model (see Fig. 6-11(B)), as determined by the theory of images. The coupling between ground-referenced twisted-pair circuits is given by the one-wire model, using the net or average voltage of the twisted-pair generator. The

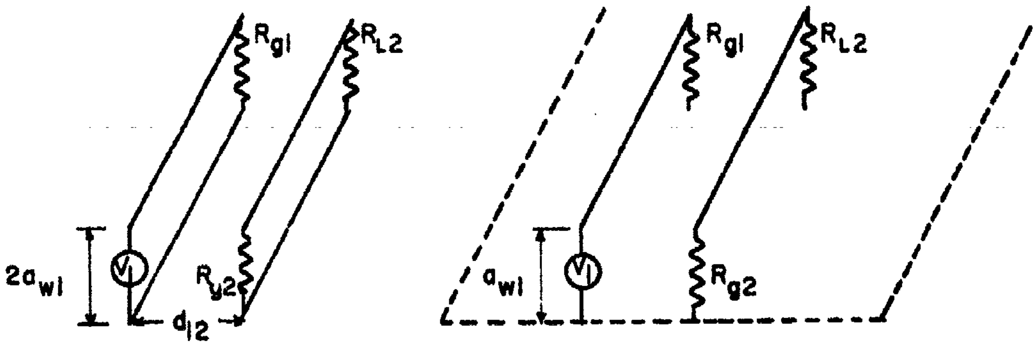


1. TV - Voltage Transfer Function
2. TI - Current to Voltage Transfer Function } close coupled wires
3. KE - Voltage to *E*-field Transfer Function (antenna source)
4. TE - *E*-field to Voltage Transfer Function (antenna receptor)
5. KE* - Voltage to *E*-field Transfer Function (wire source)
6. TE* - *E*-field to Voltage Transfer Function (wire receptor)
7. KH - Current to *H*-field Transfer Function (antenna source)
8. TH - *H*-field to Voltage Transfer Function (antenna receptor)
9. KH* - Current to *H*-field Transfer Function (wire source)
10. TH* - *H*-field to Voltage Transfer Function (wire receptor)
11. Z - Common Impedance (contained in the TI Model)

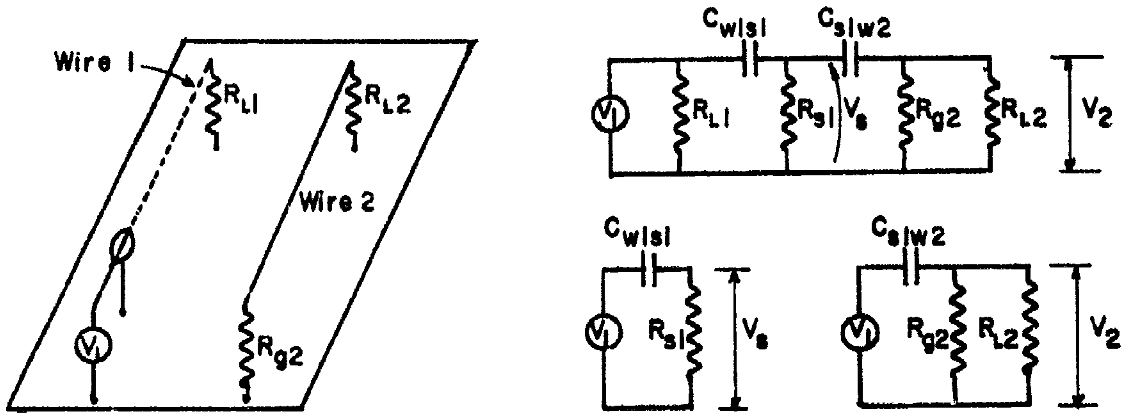
Figure 6-10. Transfer Function



(A) Capacitive Model for Unshielded Wires



(B) Capacitive Model for Two Wire Circuits



(C) Capacitive Model for Shielded Wires

Figure 6-11. Close-coupled Capacitive Transfer Models

voltage used is one-half the total voltage for an unbalanced pair, and zero for a perfectly balanced pair.

6-4.8.1.1.1.3 Shielded Wires

The capacitive models for shielded wires include both a nonresonant shielding effect and a resonant shielding effect. With reference to Fig. 6-11(C), C_{s1s1} is the coupling capacity between wire 1 and its shield; C_{s1w2} is the capacity between the shield and wire 2; and R_{s1} is the resistance of the shield on wire 1. When the impedance of C_{s1w2} is large compared with R_{s1}, the circuit may be decoupled as shown. The model uses the decoupled equivalent, making it equal to two unshielded coupling models in series. An unshielded coupling term also is added to account for unshielded portions of the wire. The low-frequency shielding factor S is given by Eq. 6-25, which has the same form as the unshielded transfer equation.

$$S = \frac{1}{1 + \dfrac{1}{2\pi f R_{s1} C_{w1s1}}}, \text{ dimensionless} \quad (6\text{-}25)$$

The shield resistance R_{s1} is approximated by

$$R_{s1} \approx R_{os1}\left[x + \frac{1}{1 + x(1 + 6x^2)}\right], \ \Omega \quad (6\text{-}26)$$

where

R_{os1} = dc resistance of wire 1 shield, Ω
$x = t/\delta$, dimensionless
t = shield thickness, m
δ = field penetration depth into shield, m

At frequencies at which the generator wire and its shield resonate, the shield will have minimum shielding effect. In this case, the RC product of Eq. 6-23 is multiplied by $[1 + D_r f^2 (1 + 6D_r^2 f^4)]$, where D_r is the reciprocal square of the first resonant frequency and is calculated based on the capacitance and inductance of the wire-shield circuit.

To model the attenuation of the electric field through the shield and its holes, the shielding factor S is also multiplied by $(e^{-x} + Fh_{s1})$. The first term represents the exponential decay of the field as it penetrates the metal, and Fh_{s1} represents the field propagated through the holes of the shield. Eq. 6-25 with R_{s1} replaced by Eq. 6-26, and the result multiplied by the resonating factor of the previous paragraph and the penetration factor described in this paragraph, defines the total shield factor S_1.

The derivation of the models for the case of a shielded susceptor wire, and for the case when both wires were shielded, follows the same procedures, and very similar results are obtained.

6-4.8.1.1.2 Inductive Transfer

Inductive transfer models have been developed for unshielded wires, for shielded wires, for twisted generator wires, and for twisted susceptor wires.

The basic inductive transfer model represents the case of two single-wire ground return circuits that are parallel over some common run length (see Fig. 6-12). The inductive transfer factor TI, which is the ratio of the voltage induced in R_{L2} to the current I_1 in the generator wire, is given by

$$TI = 2\pi f M\left(\frac{R_{L2}}{R_{L2} + R_{g2} + 2\pi f L_{w2}}\right), \ \Omega \quad (6\text{-}27)$$

where

M = mutual inductance of the two circuits, H

$$= \frac{\mu_0}{4\pi}\ln\left[\frac{(a_{w1} + a_{w2})^2 + d^2_{12}}{(a_{w1} - a_{w2})^2 + d^2_{12}}\right], \text{ H} \quad (6\text{-}28)$$

μ_0 = permeability of free space = 4×10^{-7} H/m
a_{w1} = height of wire 1 above ground plane, m
a_{w2} = height of wire 2 above ground plane, m
L_{w2} = self-inductance of the circuit², H

$$= \frac{1.25 \times 10^{-6}}{2\pi}(l_{w2})\ln\left(\frac{2a_{w2}}{r_{w2}} - 1\right), \text{ H} \quad (6\text{-}29)$$

l_{w2} = length of wire 2, m

The resistive terms include ground plane resistance (or spreading resistance due to the finite conductivity of the ground plane), and circuit lead resistance for both the source and load ends of the susceptor wire.

6-4.8.1.1.2.1 Shielded Wires

For shielded wires in which the shield is grounded at both ends, a current I_{s1} is induced in the wire 1 shield.

$$I_{s1} = -I_1\left(\frac{j2\pi f L_{s1}}{R_{s1} + j2\pi f L_{s1}}\right), \text{ A} \quad (6\text{-}30)$$

where

I_1 = current in wire 1, A
L_{s1} = self-inductance of the shield around wire 1, H

$$L_{s1} = \frac{1.25 \times 10^{-6}\, l_{w1}}{2\pi}\ln\left(\frac{2a_{w1}}{r_{s1}} - 1\right), \text{ H} \quad (6\text{-}31)$$

l_{w1} = length of wire 1, m
r_{s1} = radius of shield 1, m

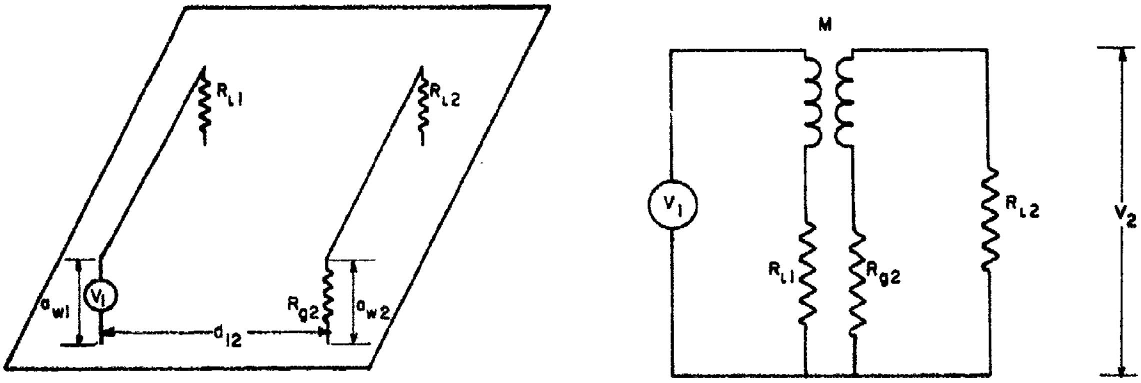


Figure 6-12. Close-coupled Inductive Transfer Model for Unshielded Wires

If the shield is ungrounded at one or both ends, no current flows through it, and there is no shielding effect.

When the wire current is forced to flow back on the shield instead of via the ground plane, the net current I_{N1} (in generator wire 1 plus return current in its shield) in the first circuit is given by the equation

$$I_{N1} = I_1 + I_{s1} = I_1 \left(\frac{R_{s1}}{R_{s1} + j2\pi f L_{s1}} \right), \text{A} \quad (6\text{-}32)$$

The magnitude of this current can be approximated by disregarding the orthogonality of the two denominator terms, as in the capacitive model. The shielding effect is given by the ratio of the net current to the wire current, which is the term in parentheses. The shield resistance, R_{s1}, is calculated using the same equation as in the capacitive model, Eq. 6-26, and L_{s1} is given by Eq. 6-31.

When both wires are shielded, the shielding effect is given by the product of the two shielding factors. When shields are terminated with pigtails rather than coaxially, the loop created by the pigtail wire allows the leakage of flux, which limits the amount of shielding. This effect is modeled by adjusting the shielding factor by a term which is the ratio of the area of the pigtail loop to the area of the shield loop.

6-4.8.1.1.2.2 Twisted Wires

Two models are used, one for an emitter and one for a susceptor:

a. Emitter. If a pair of wires is twisted, the field produced is a result of the net flow of current in one direction as seen externally. Thus, if the outgoing and return currents are equal, as occurs when there is no leakage to ground, no field is produced.

Let a twisting factor S_{T1} for wire 1 be defined as the multiplying factor to be applied to Eq. 6-27 as a result of the generator connection being a twisted pair rather than a single wire with ground return. The twisting factor therefore is the fraction of the total current which is common mode, i.e., current which returns in the ground plane rather than in the twisted return wire due to distributed or lumped capacitance to ground. This effect is given by Eq. 6-33 where C_g is the shunt capacity to ground and C_T is the total capacity to ground, i.e., C_g plus the wire-to-wire shunt capacity.

$$S_{T1} = \frac{2\pi R_{L1} C_g}{2\pi R_{L1} C_T + 1/f}, \text{ dimensionless} \quad (6\text{-}33)$$

where

R_{L1} = load resistance on generator wire 1, Ω

If a shield is present and grounded at the same end as the one where one lead of the twisted pair is grounded, then the distributed capacity to ground is eliminated.

b. Susceptor. For twisted susceptor wires, refer to Fig. 6-13. The voltage V_N induced in the twisted loop area is assumed to be negligible compared to the voltage induced by conversion from a common mode voltage V_c induced in the ground plane. The common mode voltage V_c is given by

$$V_c = 2\pi f \Phi, \text{V} \quad (6\text{-}34)$$

where

Φ = flux linking the circuit in which V_c is induced, Wb

Thus, the twisting factor S_{T2} for wire 2 is the ratio of the voltage V_2 (induced by V_c in the load resistance)

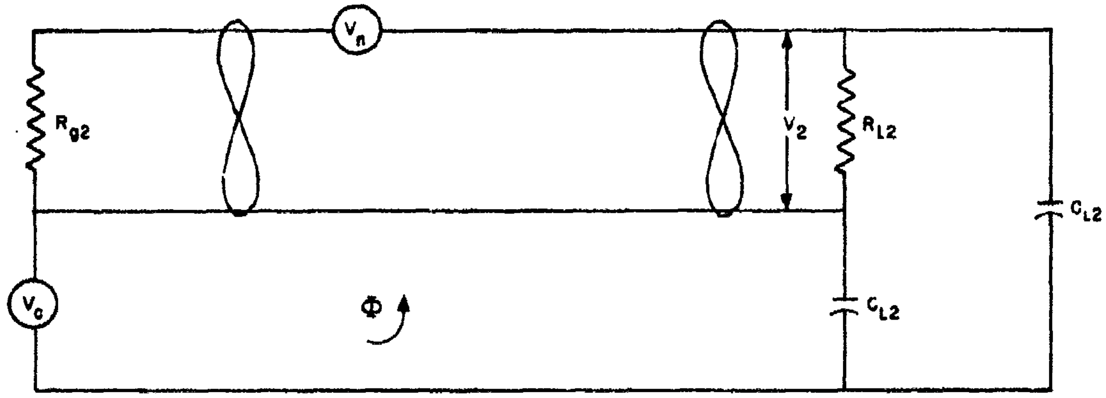


Figure 6-13. Simplified Equivalent Lumped Circuit or Twisted Susceptor Wires

to V_c. This is obtained by circuit analysis and is given by Eq. 6-35, which has the same form as for the case where the generator wire is twisted (Eq. 6-33).

$$S_{T2} = \frac{2\pi R_{a2} C_{L2}}{2\pi R_{a2} C_T + 1/f}, \text{ dimensionless} \quad (6\text{-}35)$$

where

$$R_{a2} = \frac{R_{L2} R_{g2}}{R_{L2} + R_{g2}}, \ \Omega \quad (6\text{-}36)$$

C_{L2} = lumped and distributed capacitance to ground for wire 2, F

C_T = C_{L2} in parallel with wire-to-wire shunt capacitance, i.e., total capacitance to ground, F

6-4.8.1.2 Field Coupling

The coupling of energy from a generator to a susceptor via an electromagnetic field is treated in two parts: the field produced by the generator (par. 6-4.8.1.2.1) and the voltage produced by the field in the susceptor or receiver (par. 6-4.8.1.2.2).

6-4.8.1.2.1 Generator Fields

The functions KH, KH^*, KE, and KE^* listed on Fig. 6-10 are used to convert from a current or voltage on a wire, shielded or unshielded, and on an antenna, to the magnetic field H and electric field E at a distance of 1 m from the wire or antenna. Formulas are available to establish coupling values for frequencies between 10 and 10^{10} Hz. Antenna models allow the approximation of antenna properties as calculated by a separate manual or computer analysis, or as obtained by testing.

6-4.8.1.2.1.1 Conversion of Voltage to *E*-Field

KE is a factor converting antenna terminal voltage to the E-field at a distance of 1 m from the antenna. The program assumes the conversion known at the band center of the antenna and it models the variation with frequency using an arbitrary characteristic similar to the form of the transfer function of a Butterworth filter (Ref. 15, par. 3-3.2.4.2.2). However, unlike filters for which out-of-band attenuation approaches infinity, the models assumed here have limited out-of-band attenuation. The conversion factor used at band center is determined by a separate computation or measurement assuming the point of field observation is at a back lobe.

KE^* is a factor converting voltage on a long wire above a ground plane to the E-field at a distance of 1 m from the wire. The physical model of the generator wire is similar to that shown in Fig. 6-11(A) or the unshielded generator case and to that shown in Fig. 6-11(C) for the shielded generator case. While the capacitance coupling computation in par. 6-4.8.1.1.1 determined the voltage on a nearby wire, the computation now is of the field near the generator wire. The field equation is given in the form

$$KE^* = (A + X)S_l, \ \text{m}^{-1} \quad (6\text{-}37)$$

S_l is a factor accounting for the effect of the shield and is determined in the manner described in par. 6-4.8.1.1.1.3 to account for the effect of shield when calculating capacitive coupling. The two terms in parentheses, A and X, account respectively for the vertical induction field (near the ground plane) and the radiation field components at the point of measurement. They are given by

$$A = \frac{2a_{w1}}{\ln(2a_{w1}/r_{w1})}, \text{ m}^{-1} \quad (6\text{-}38)$$

$$X^2 = \left(\frac{8.75\times 10^3}{l_{w1}\, z_o^2}\right)(1 + 1.32\times 10^{-8}\, l_{w1} f)$$

$$\times [1 + 0.4 \ln(1.0001 + 6.7\times 10^{-9}\, l_{w1} f)]$$

$$\times \left(1 + \frac{8.3\times 10^{15}}{l_{w1}^2 f^2}\right)^{-1}$$

$$\times \left(1 + \frac{2.9\times 10^{15}}{a_{w1}^2 f^2}\right)^{-1}, \text{ m}^{-2} \quad (6\text{-}39)$$

where

z_o = the magnitude of the impedance seen by the source of voltage v_1, Ω

f = frequency, Hz

6-4.8.1.2.1.2 Conversion of Current to *H*-Field

KH is a factor converting antenna input current to the *H*-field at a distance of 1 m from the antenna. As in the electric field case described in par. 6-4.8.1.2.1.1 the program assumes the conversion known at the antenna's band center and it models the variation with frequency using an arbitrary filter-like characteristic of the same form as for the electric field.

*KH** is a factor converting current on a long wire parallel to a ground plane to the induction *H*-field at a distance of 1 m from the wire in the vicinity of the ground plane. The field here is tangential to the ground plane in a direction perpendicular to the plane of the wire and its image. The physical model of the generator wire is similar to that shown in Fig. 6-11(A) for an unshielded wire, and to that shown in Fig. 6-11(C) for a shielded wire, with the shield grounded at both ends. The model also applies to the case of a twisted pair from source to load above a ground plane with ground at one end of the run. The field equation is of the form

$$KH^* = A_1 (S_{S1} + B) S_{T1}, \text{m}^{-1} \quad (6\text{-}40)$$

where $A_1 = 0.32\, a_{w1}$, S_{S1} accounts for the shield (= 1 if there is no shield or if shield is grounded only at one end), B accounts for the area of pigtail loops when a shield is used and when the shield is not grounded coaxially (= 0 if grounding is coaxial) and S_{T1} is the twisting factor when a twisted pair is used (= 1 if no twisting is used). The exact form of these factors, which depend on the geometry of the structure and on the frequency, will be found in Ref. 15, par. 3-3.2.5.1.

6-4.8.1.2.2 Field Reception Models

The field reception models contain two effects: (1) the field decay with distance, and (2) the voltage induced into wires and antennas. The susceptor model contains the voltage induced into both a wire and an antenna.

In the field decay models, the decay of *E*-field and *H*-fields with distance is treated in the same manner. The field from the generator is described at 1 m, and the decay to greater distances or the increase to shorter distances is described by a field decay exponent parameter associated with the generator. The appropriate decay rate must be determined outside the computer.

6-4.8.1.2.2.1 *E*-Field Transfer Function

TE and *TE** are the *E*-field to voltage transfer ratios, *TE* applying to the voltage appearing at the terminals of an antenna and *TE** applying to the voltage induced on a wire. These quantities give the ratio of voltage developed to the magnitude of the field at a point 1 m from a generator. The total voltage developed at a susceptor input per unit field (see Fig. 6-10 which shows, for example, a generator antenna illuminating both the susceptors antenna and its antenna lead-in wire) is the sum of voltages per unit field developed by antenna and connecting wire. The total voltage is given by

$$TET = TE + TE^* = TED(KAE + TEV), \text{ m} \quad (6\text{-}41)$$

where

$$TED = \frac{E\text{-field at susceptor}}{E\text{-field at 1 m from generator}}, \text{ d'less}$$

$$KAE = \frac{\text{voltage induced in susceptor}}{E\text{-field at susceptor antenna}}, \text{ m}$$

$$TEV = \frac{\text{voltage induced in susceptor}}{E\text{-field at susceptor wire}}, \text{ m}$$

TE and *TE** are, respectively, the susceptor antenna contribution and the susceptor wire contribution. The field decay vs distance function *TED* is

$$TED = (D_{ij})^{-KED}$$

$$= \frac{\text{E (at distance } D_{ij} \text{ from source)}}{\text{E (at 1 m from source)}} \quad (6\text{-}42)$$

where KED = *E*-field fall-off exponent.

The antenna reception function is *KAE*. It allows antenna properties to be specified by the user. In addition, a bulkhead factor (to account for shielding of wires by intervening conductive materials) is applied to fields from generator wires. The bulkhead factor is not applied to fields from generator antennas. The output of an E-field receiving antenna, *KAE* is a product of three factors. The first is the maximum effective height of the antenna (as a function of frequency). The second is a function giving the frequency characteristic of the antenna and is similar to that used to determine field emission from an antenna (par. 6-1.8.1.2.1.1). The third is a bulkhead attenuation factor which is set equal to 1 if the receptor is illuminated by an antenna and less than 1 when it is illuminated by a wire.

The E-field wire reception function, *TEV*, contains two effects; a low frequency capacitance-like response to vertical fields, and a high frequency response. The voltage is assumed to be induced over the length of the wire not sharing a common run length with the generator wire, i.e., the separation D_{ij} is too great for inductive or capacitive transfer. A bulkhead factor is here applied to fields from generator antennas, but not to fields from generator wires.

The low frequency induced voltage is derived by assuming that the wire-to-ground capacitance C_{w2g} represents the source impedance for the vertical field, resulting in the equivalent circuit shown in Fig. 6-14. This assumes a wire, such as the receptor wire of Fig. 6-11(A) or (C), with resistors R_{g2} and R_{L2} at the ends. The source voltage is the E-field times the height of the wire.

The high frequency part of the function is based on aperture theory. The voltage V induced in the load is given by Eq. 6-43, where A_e is the maximum effective aperture, P_d is the incident power density, and the term in parentheses represents the result of impedance mismatch in energy transfer to the susceptor load R_{L2} coming from a source with radiation resistance R_r.

$$V = \sqrt{A_e P_d}\left(\frac{R_r^{1/2}}{R_{a2}} + \frac{1}{R_r^{1/2}}\right)^{-1}, \text{ V} \quad (6\text{-}43)$$

The power density is assumed to be $E^2/377$, and the effective aperture of the wire of length l_w is given by

$$A_e = 2 \times 10^8 \, l_w/f \text{ , m}^2 \quad (6\text{-}44)$$

TEV then turns out to be of the form

$$TEV = \left[\left(B_1 + \frac{B_2}{f}\right)^{-1} + B_3\left(\frac{f}{X_1} + X_1\right)^{-1}\right] B_4 \text{ , m} \quad (6\text{-}45)$$

where B_1, B_2, B_3 and B_4 are constants depending on geometrical factors, and X_1 is slowly varying with frequency at high frequencies (Ref. 15, par. 3-3.2.5.3.3). The first term in the brackets represents the low frequency coupling, the second term represents the high frequency coupling, and the factor B_4 accounts for bulkhead shielding. It is equal to unity when no bulkhead intervenes. The program also allows shielding of the receptor wire to be accounted for. The form obtained is similar to Eq. 6-45, except for changes in the factors (Ref. 25, par. 3-3.2.5.3.4), which makes some of them dependent on frequency.

6-4.8.1.2.2.2 *H*-Field Transfer Function

The H-field to voltage transfer function is formed in a manner similar to the E-field function. It is comprised of terms *TH* and *TH**, respectively giving the

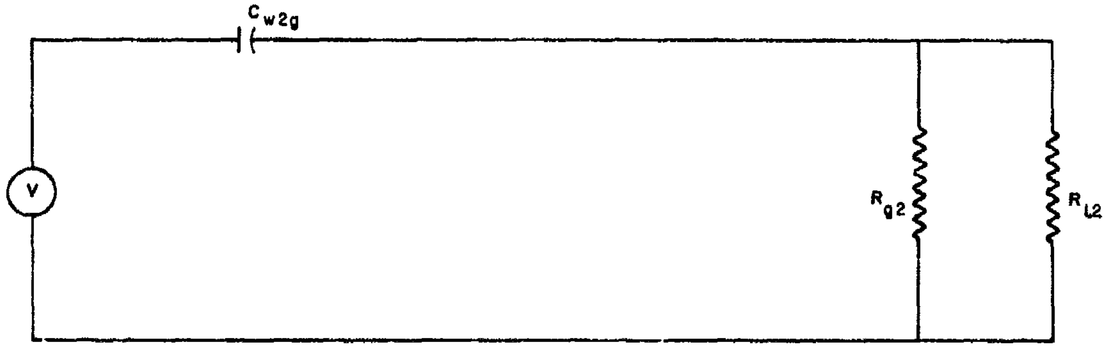


Figure 6-14. Equivalent Circuit for E-Field Wire Reception Model

voltage appearing at antenna terms and voltage induced on a wire, when immersed in an *H*-field. The total voltage per unit *H*-field 1 m from the generator is given by

$$THT = THD(KAH + THV)\ ,\ \Omega\cdot\text{m} \quad (6\text{-}46)$$

where

$$THD = \frac{H\text{-field at susceptor}}{H\text{-field at 1 m from generator}}\ ,\ \text{d'less}$$

$$KAH = \frac{\text{voltage induced in susceptor}}{H\text{-field at susceptor antenna}}\ ,\ \Omega\cdot\text{m}$$

$$THV = \frac{\text{voltage induced in susceptor}}{H\text{-field at susceptor wire}}\ ,\ \Omega\cdot\text{m}$$

The field decay vs distance function THD is

$$THD = (D_{ij})^{-KHD} = \frac{H\ (\text{at distance } D_{ij} \text{ from source})}{H\ (\text{at 1 m from source})} \quad (6\text{-}47)$$

where KHD = field fall-off exponent.

The antenna reception function is KAH. It allows antenna properties as determined outside the computer to be specified. In addition, a bulkhead factor (to account for shielding of wires by intervening conductive materials) is applied to fields from generator wires. The bulkhead factor is not applied to fields from generator antennas. KAH is a product of three factors as in the *E*-field case. The first is the maximum antenna field to voltage transfer (as a function of frequency). The second is a filter-like model giving the antenna frequency characteristic. The third is a bulkhead factor.

The *H*-field wire reception function, THV, is based on the assumption that the pick-up loop area is that portion of the wire which does not share a common run with the generator. A bulkhead factor is applied to fields from generator antennas, but not to fields from generator wires. The function THV for an unshielded wire turns out to be directly proportional to frequency, the multiplying factor being determined by the electrical and geometric parameters of the wire and the bulkhead factor, if applicable. If the receptor wire is shielded and/or it is a twisted pair, the effects can be accounted for in the multiplying factor which, however, puts a frequency dependence into the factor (Ref. 15, par. 3-3.2.5.4.3).

6-4.8.2 Model Utilization

The objective of the computer program is to produce interference generation and susceptibility limits based upon the functional requirements of any particular system in such a way that no intolerable malfunctions occur in the equipment.

The various emission and susceptance portions of the electronic system are divided into four broad classifications of two priorities. Priority No. 1 circuits are those which have generation and reception characteristics which are functionally required for system operation. Priority No. 2 circuits are those circuits whose generation and reception characteristics are unnecessary for system operation, and which might be undesirable from the system environment point of view.

For each priority classification there are generator and susceptor subclassifications whether the generator or susceptor characteristics are intentional or unintentional. A single physical circuit possibly could fall into all four of these classifications but, for purposes of analysis, it is treated as four separate entities.

The analysis philosophy is to (1) limit the interference generation of priority No. 2 generator circuits to a level which is compatible with the required reception properties of priority No. 1 susceptor circuits, and (2) limit the total environment incident on priority No. 2 susceptor circuits from both priority No. 1 and priority No. 2 generators to a tolerable level.

In the first step in the analysis the total interference from required (priority No. 1) sources incident upon each required (priority No. 1) susceptor is determined. Unless the required sources are incompatible with the required susceptors, a margin will exist that represents the difference between the malfunction level of the susceptors and the total interference incident. This margin or quota then may be allotted to the extraneous priority No. 2 generators.

6-4.8.2.1 Compatibility Analysis

An approximation to the voltage V_{ij} induced by a generator i at a reference point in the circuit of susceptor j is calculated using the following formula:

$$V_{ij} = \int_{f_1}^{f_2} G_i(f)\, T_{ij}(f) P_j(f)\, df\ ,\ \text{V} \quad (6\text{-}48)$$

where

$G_i(f)$ = magnitude of generator i output Fourier spectrum in V/Hz

$T_{ij}(f)$ = magnitude of the voltage transfer function from generator i to susceptor j, dimensionless

$P_j(f)$ = magnitude of the voltage transfer function describing the frequency response of susceptor j, dimensionless

Eq. 6-48 gives the peak response of a device to a spectral density function whose frequency components are all in phase and assumes the susceptor responds to the peak voltage, and the transfer functions have linear phase characteristics. It is considered to represent a worst-case assumption.

Fig. 6-15 illustrates relationships for voltage transfers between one generator and one susceptor. The transferred and bandwidth limited spectrum is integrated and can be compared with the susceptibility threshold AS_j. If it does not exceed AS_j, the two circuits represented are compatible with each other.

The process described by Fig. 6-15 is performed many times in several different modes.

6-4.8.2.2 Specification Development

Fig. 6-16 shows a number of priority No. 1 generators *G*1, *G*2, *G*3, *G*4, etc., representing circuits within a system.

System generators have both required and unrequired emission characteristics. As an example, a transmitter must transmit at some required frequency at some prescribed power level and over a given

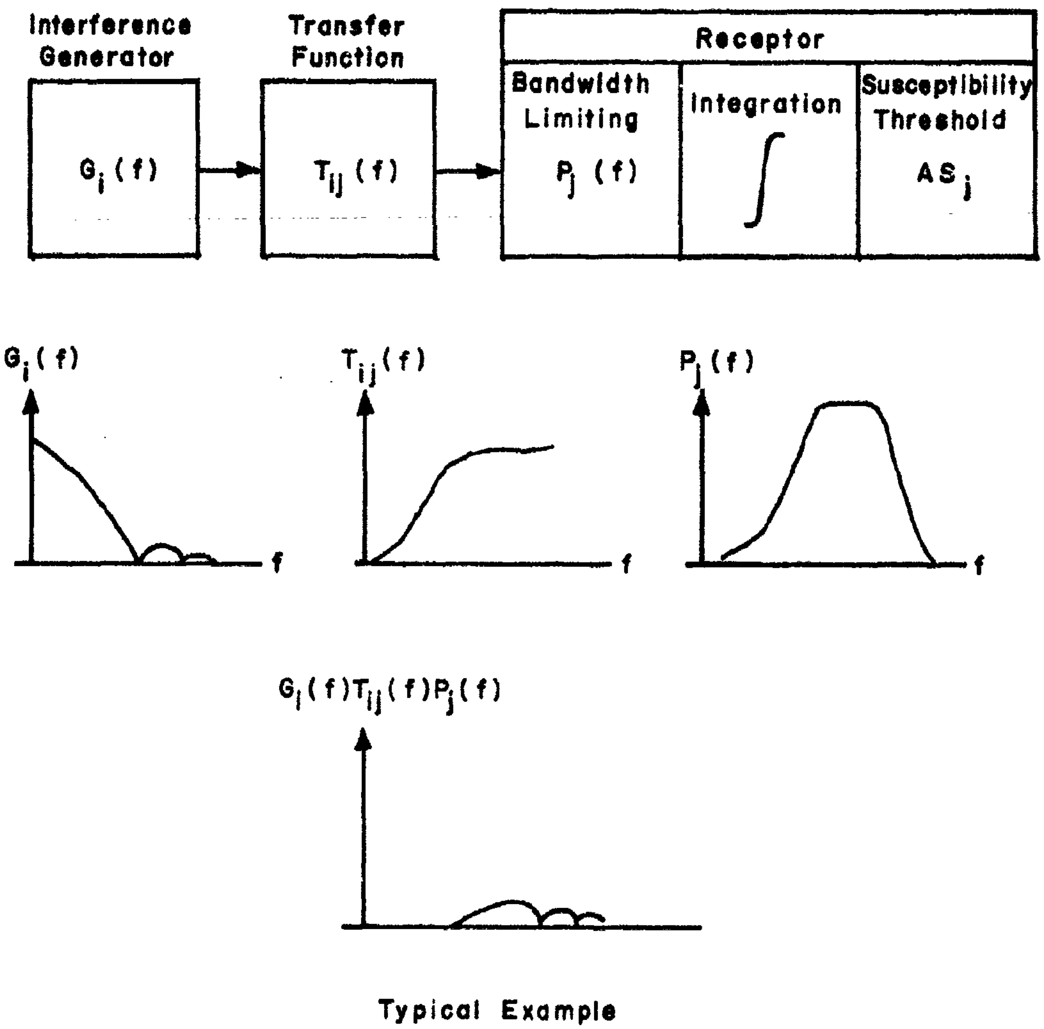


Figure 6-15. SEMCAP Compatibility Analysis

bandwidth. However, it may also transmit at harmonic or at other spurious frequencies. Since these harmonic and spurious outputs are not needed for proper system operation, these particular extraneous or undesirable emanations are treated separately.

At a susceptor shown here by R_j, the required spectral distributions from each of the modeled sources after transfer to the susceptor terminals are summed. The unrequired or priority No. 2 generators are combined in Fig. 6-16 in the box labeled "Extraneous Emitters". A total tolerable interference output AS_j is assigned to these priority No. 2 sources by use of Eq. 6-49

$$AS_j' = AS_j - \int \left[\sum_i G_i(f) T_{ij}(f) \right] P_j(f)\, df \text{ , V} \tag{6-49}$$

where

AS_j' = total tolerable input to susceptor j from all priority No. 2 generators, V

AS_j = the sensitivity level of susceptor j, V

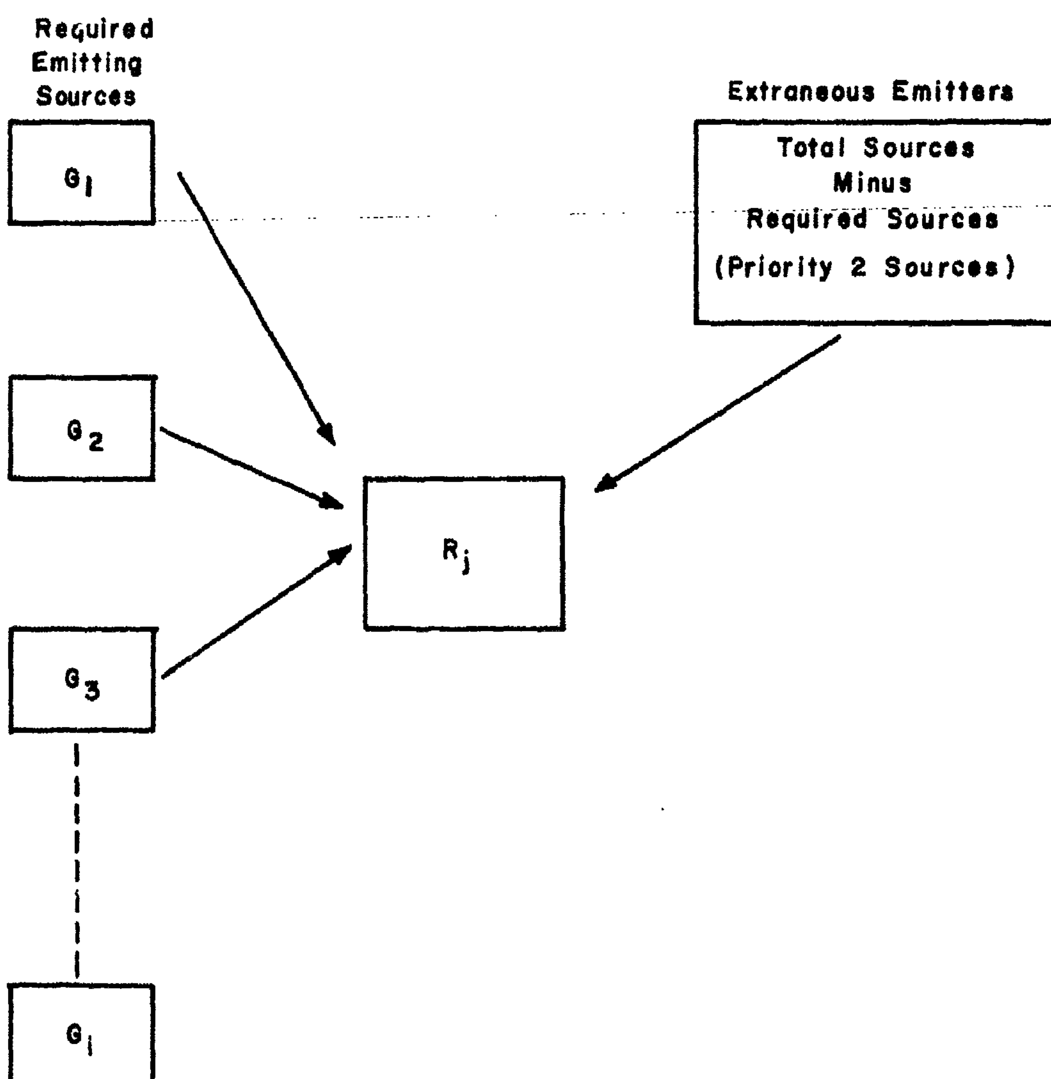


Figure 6-16. Interference Received by a Typical Susceptor

The quota of interference represented by AS'_j can be assigned to the sum total of all extraneous emitters. However, since AS'_j is a voltage which defines the extraneous interference after it has been integrated through the bandwidth of the susceptor in question, there are an infinite number of generation characteristics which, when passed through a transfer function, multiplied by a bandwidth limiting factor and integrated could equal AS'_j.

6-4.8.2.2.1 Generation Specification Routine

The routine used is shown in Fig. 6-17. A spectrum called the first order standard is defined which portrays the spectral shape and amplitudes of the emanations fr.m all the extraneous generators when few limitations have been imposed.

Each priority No. 2 susceptor is allowed to modify this initial spectrum iteratively until a spectrum is created which induces into that susceptor a voltage no greater than AS'_j the voltage allowable from the extraneous generators. This iterative modification process is designed to reduce the level of the first order standard primarily in those areas of the spectrum where the most interference is received, such as in the pass band of the susceptor. Up to four iterations have been found to be required to reduce the received voltage to within 2 dB of AS'_j depending upon the type of spectrum involved.

The first order standard, as modified by the first priority No. 2 susceptor, is stored. The second priority No. 1 susceptor is then allowed to modify the first-order standard until a spectrum compatible with that susceptor is created. This process is repeated for each susceptor to be examined. The final specification spectrum must be compatible with all priority No. 1 susceptors. A spectrum that is the minimum envelope of those individual spectrums, developed by this modification process, achieves this result. It is nowhere higher than the first-order standard and is lower in those regions of the spectrum where the interference was incompatible with one or more of the priority No. 1 susceptors of the system.

This spectrum represents the interference level which all the extraneous generators of the system acting together can be allowed to generate. At this stage it is a system interference specification with no margins present.

To create a specification with margins that can be applied to individual subsystems or smaller units, the minimum spectrum is divided by a J factor that describes what fraction of the system specification each subsystem will be allowed to generate. J must be based on an estimate of how the interference spectrums of each extraneous generator will combine, and how many are significant for all priority No. 1 susceptors.

An interference generation specification is created by the preceding process for each of four kinds of interference:

a. Conducted voltage
b. E-fields
c. Conducted current
d. H-fields.

6-4.8.2.2.2 Susceptibility Limits

Susceptibility limits are developed to control the response of extraneous susceptors. Any susceptor in the system is exposed to the two sources of the total environment, as shown in Fig. 6-18, i.e., the summation of the emissions due to black-box specification limits (extraneous emitters) and the summation of energy due to functional circuits. The sum of these two sources represents the total system environment. It also represents the environment in which the susceptor circuits are required to function.

Since the priority No. 2 generator levels have been established by the generation specification routine (see par. 6-4.8.2.2.1), only the priority No. 2 susceptor levels are considered to be adjustable at this time. If the effects of the total environment exceed AS'_j, the routine reduced priority No. 2 susceptibility of a receiver to meet the compatibility requirement. This is done on the assumption that the environment is created from either broadband or narrowband sources. Each susceptor is considered successively in this manner.

6-4.8.2.3 Waiver Evaluation

The SEMCAP computer program considers a request for a specification waiver in terms of its effect on system performance. The routine calculates the voltage induced into each priority No. 1 susceptor based upon the spectral density data used for the generator. If the received voltages do not exceed the allowed voltages, the generator is compatible with the priority No. 1 requirements of the system and a waiver can be granted. However, if the generator spectrum exceeds the specification by a large amount in some frequency ranges, the impact of the interference on priority No. 2 susceptibilities must be evaluated before a waiver can be granted.

To evaluate this situation a new susceptibility specification is created based on the inclusion of the new generator data. A determination then can be made whether the granting of the waiver would significantly change the susceptibility specifications.

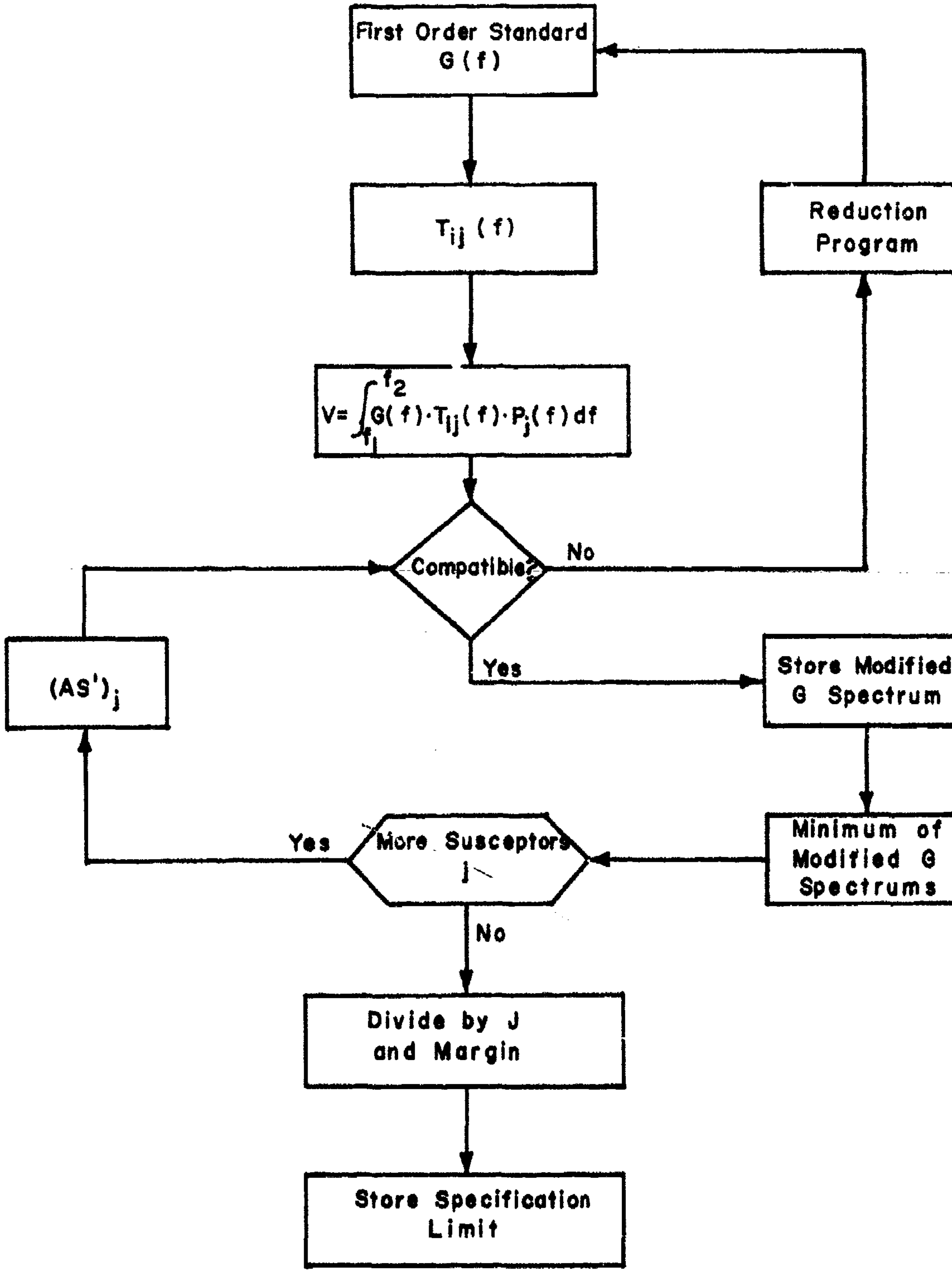


Figure 6-17. Interference Specification Limits Program

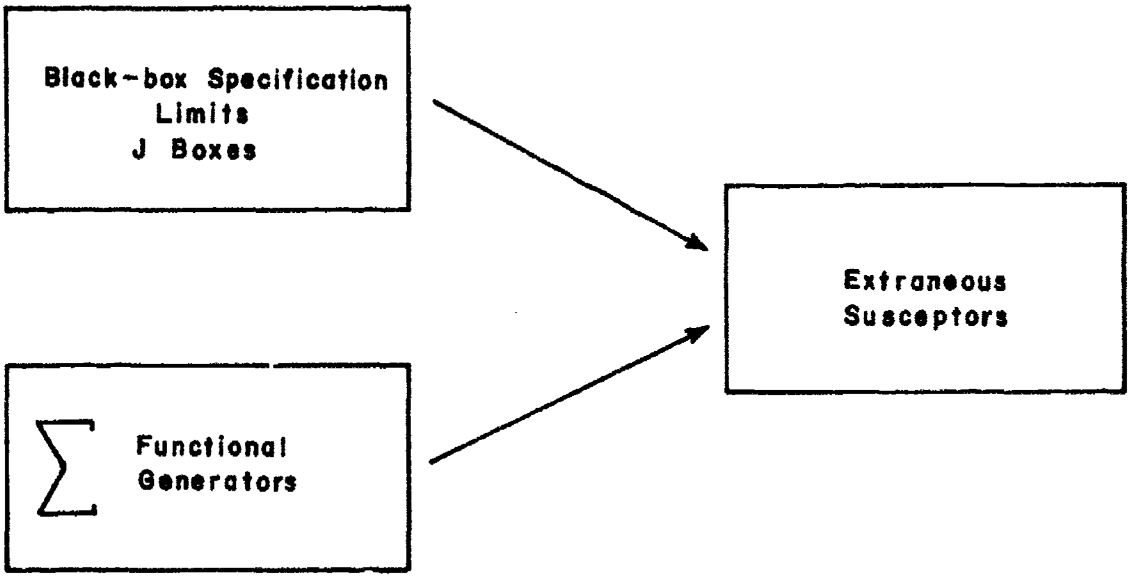


Figure 6-18. Susceptibility Limit Development

6-4.8.2.4 Use and Insertion of Test Data

At the time of specification development, the portion of the EMC environment from uncontrolled sources is assumed to have a set of levels equal to those of the specification limits. As test data are made available, this assumed set of levels is replaced by actual test data. Then at the completion of all black-box-level testing, a final compatibility analysis run is made, making maximum use of test data. Based on this final prediction, circuits whose operations are marginal are selected for monitoring during the system test.

6-4.8.2.5 Program Output

Fig. 6-19 is a sample SEMCAP output sheet. It is an example of a portion of a detailed receptor-by-receptor investigation. The quantitative details to establish how each generator interferes with a particular receptor are presented. The major part of the output identifies the fraction of tolerable voltage received from each generator via each of the four kinds of coupling. The corresponding dB margin also is printed, with negative margins representing interference levels above the receptor threshold. The generators are designated in the right-hand column by name. An "S" indicates that a specified, rather than a measured, generator output was used in the computation.

Flags call attention to potential problem areas and have the following meaning:

a. A = Almost an interference problem, +10 to +30 dB margin

b. M = Marginal interference problem, −10 to +10 dB margin

c. P = Probable interference problem, less than −10 dB margin

The flags in the left-hand column are based on the sum of all four types of interference from each generator. For the voltage fraction received from all generators, together with generators operating simultaneously, the corresponding margin is given at the bottom of the printout.

6-4.8.2.6 Program Availability

SEMCAP is currently operational on CDC-6600 and IBM 360/70 computers. It requires 64k (decimal) of core storage. Some of the programming is in machine language. It is able to handle 240 generators and 240 susceptors. One of the longest programs run to date, for the B-1 Bomber, involved 150 emitters and 120 susceptors. This program required 50 min on the CDC-6500 or 20 min on the IBM 360/370. A spacecraft analysis generally can be done in 5 to 10 min. The program was developed by TRW Systems,

RECEPTOR SUBSYSTEM 30, TERMINAL 2 DCA FROM AIA FILTERED US

THE FRACTION OF THE SENSITIVE VOLTAGE RECEIVED FROM EACH GENERATOR TERMINAL (I) IS

		CAPACITIVE		E FIELD			INDUCTIVE		H FIELD			
	I	RATIO	DB	RATIO	DB		RATIO	DB	RATIO	DB		
	1	1.67-08	155.5	2.22-10	193.1		1.82-02	34.8	3.09-04	70.2		IPAD CONTROL SIGNAL
	2	2.52-04	72.0	4,79-07	126.4		6.73-04	63.4	2.53-07	132.0		DISCRETE TLM UNSHIELD
	3	5.16-08	145.7	2.46-10	192.2		2.13-04	73.4	8.26-08	141.7		DISCRETE TLM SHIELDED
	4	3.32-10	189.6	1.84-05	94.7		0.00	400.0	4.50-06	106.9		LINK 1 RF SWITCH
	5	1.45-07	136.3	6.80-09	163.4		2.35-04	72.6	8.60-07	121.3		LINK 2 CONVERTER CONTR
	6	1.45-07	136.8	6.82-09	163.3		2.35-04	72.6	8.60-07	121.3		LINK 1 CONVERTER CONTR
M	7	1.50-03	56.5	2.78-11	211.1	M	1.23+00	-1.8	0.00	400.0		EPS PRIMARY BUS TRANS
	8	7.87-04	62.1	1.13-06	119.0		1.05-06	119.6	2.75-10	191.2		BUS MONITOR TRANSIENTS
M	9	6.48-04	63.8	1.16-11	218.7	M	5.18-01	5.7	0.00	400.0		4 V BUS TRANSIENTS
	10	6.38-14	263.9	2.08-08	153.6		0.00	400.0	3.09-09	170.2		HEATERS
	11	8.06-14	261.9	2.18-04	73.2		0.00	400.0	2.95-03	50.6	$	CAPE ENVIRONS 150KHZ
	12	3.97-15	288.0	2.88-05	90.8		0.00	400.0	1.45-04	76.8	$	CAPE ENVIRONS 1.5MHZ
	13	1.43-16	316.0	2.57-06	111.8		0.00	400.0	5.28-06	105.5	$	CAPE ENVIRONS 10MHZ
	14	2.00-15	294.0	5.18-05	85.7		0.00	400.0	8.39-05	81.5	$	CAPE ENVIRONS 243 MHZ
	15	6.30-15	284.0	3.38-05	89.4		0.00	400.0	3.44-04	69.3	$	CAPE ENVIRONS 1 GHZ
	16	1.11-14	279.1	1.67-05	95.5		0.00	400.0	9.31-04	60.6	$	CAPE ENVIRONS 2.7 GHZ
	17	1.62-14	275.6	1.34-05	97.4		0.00	400.0	3.05-03	50.3	$	CAPE ENVIRONS 9 GHZ
	18	3.62-12	228.8	1.48-10	196.6		0.00	400.0	8.88-09	161.0	$	RW TACH SIGNAL TO ACC
	19	2.80-11	211.1	1.52-09	176.3		0.00	400.0	0.00	400.0		SPIN/JET COMMANDS ACC
	20	2.45-12	232.2	9.67-12	220.3		0.00	400.0	1.06-08	159.5	$	REACTION WHEEL DRIVE
	21	4.10-12	227.8	1.78-10	195.0		0.00	400.0	0.00	400.0	$	REACTION WHEEL TACH
	22	1.42-14	277.0	2.19-12	233.2		0.00	400.0	1.09-09	179.3		VERNIER THRUSTERS
	23	1.42-14	277.0	2.19-12	233.2		0.00	400.0	1.09-09	179.3		ATTITUDE THRUSTERS
	24	1.53-12	236.3	3.42-11	209.3		0.00	400.0	4.19-07	127.6		DECODER COMMAND TO CEA
	25	1.36-08	157.3	2.64-11	211.6		8.06-05	81.9	2.90-07	130.7		DECONDER COMMAND TO EIA
	26	1.81-08	154.9	3.07-11	210.2		1.03-04	79.7	2.55-07	131.9		DECONDER COMMAND TO SPS
	27	8.06-17	321.0	7.02-12	223.1		0.00	400.0	7.05-12	223.0	$	LINK 3 COHERENT DRIVE
	28	3.42-14	269.3	1.90-13	254.4		0.00	400.0	2.23-08	153.0	$	PRN TO BASEBAND
	29	1.07-13	259.4	5.09-13	245.9		0.00	400.0	2.81-09	171.0	$	PCM SUBCARRIER DRIVE
	30	3.60-14	268.9	8.43-08	141.5		0.00	400.0	9.42-10	180.5	$	RF OUTPUT
	31	2.23-10	193.0	2.14-11	213.4		1.65-04	75.7	5.39-07	125.4	$	DIGITAL TLM
	32	2.88-08	150.8	2.50-14	272.0		1.01-04	79.9	0.00	400.0	$	128 KHZ CLOCK
A	33	1.43-05	96.9	3.21-10	189.9	A	1.05-01	19.6	0.00	400.0		TOR 9B BROADBAND
	34	1.43-05	96.9	3.21-10	189.9		4.62-03	46.7	0.00	400.0		TOR 9B BB TWISTED
A	35	3.39-07	129.4	3.21-10	189.9	A	3.34-02	29.5	0.00	400.0		TOR 9B BB SHIELDED

VOLTAGE FRACTION RECEIVED = 1.924032+00 DB MARGIN = -5.7 THIS RECEPTOR IS INCOMPATIBLE WITH THE SYSTEM.

Figure 6-19. Sample SEMCAP Output Sheet

Redondo Beach, CA, for the NASA Manned Spacecraft Center, Houston, TX. Currently TRW Systems, One Space Park, Redondo Beach, CA 90278, provides engineering services in connection with the use of the program.

6-4.9 ISCAP

6-4.9.1 General Description

ISCAP is an acronym for Intrasystem Compatibility Analysis Program (Ref. 17). The analysis section consists of six routines designed to investigate the principal sources of interference and modes of energy transfer which exist within an electronics-oriented system that has an identifiable electronic and physical configuration, e.g., a communication central or a tactical air control center. They are:

a. MIL-STD-469 (Ref. 18) Check of radar design parameters

b. MIL-STD-188 (Ref. 19) Check of communications equipment design parameters

c. Spectrum overlap and interference potential from transmitter fundamental, harmonic, and spurious emissions

d. Interference potential through receiver spurious, intermodulation, and cross-modulation responses

e. Interference potential of receiver local oscillator radiation

f. Interference potential due to radiation from cables, cases, and antennas coupling into nondesign receptors.

The computer program and supporting documentation provide the following capabilities:

a. Organization of system component elements into a four-level hierarchy, i.e., systems, equipment, major units, and minor units.

b. Multilevel data input, storage, and retrieval. This means that characteristics that apply to subordinate units within a system only need to be entered once at the highest level of ownership and not at all lower level units.

c. Analysis routines to investigate equipment conformance to military standards, and potential interference interactions based on design characteristics and specifications.

d. Run-time selection of analysis routines and suboptions through use of a common program driver.

e. Run-time identification of equipments and associated minor units which require investigation.

6-4.9.2 Data Base

The analysis program uses a common data base from which inputs to the various analysis routines are obtained. The data base is self-organizing and contains characteristic data keyed to the four-level hierarchy. It requires only one input file consisting of cards whose number depends on the complexity of the problem being considered. A minimum of 6 cards are required for any run, and the normal maximum length of the file is 15,000 cards. However, this limit can be extended to the limit of available core storage in the computer. Multiple input files can be processed in a single computer run.

Fig. 6-20 represents a simple example of the hierarchical organization of an electronic system. As one considers elements at lower points in the hierarchy, characteristics become more and more unique until, at the minor unit level, information is pertinent to a given specific item. Conversely, as one considers higher elements in the hierarchy, information becomes more general, applying to all subordinate elements. For example, location information for equipment No. 1 in Fig. 6-20 could apply to all subordinate elements. Hence, the computer program is designed to model a series of elements that represent emitters and receptors and, subsequently, to associate them with the proper higher level elements.

To conduct an analysis, characteristics of the elements within the four levels must be entered and recorded. This is accomplished through the use of attribute cards which contain data such as power output, antenna gain, location, wire/cable characteristics, transmitter/receiver tuning range, signal/susceptibility thresholds for receiver/control/signal circuits, etc.

Where measured or design data are not available, the analysis routines utilize limits specified by MIL-STD-461 (Ref. 20).

6-4.9.3 Military Standards Check

Two military standards checks are performed: (1) the MIL-STD-469 check of radar design parameters, and (2) the MIL-STD-188 check of communications equipment design parameters.

The MIL-STD-469 routine performs the following:

a. Radar emission bandwidth check for which measured, design, or synthesized spectrum data can be used

b. Radar spurious emission checks which use design, measured, or MIL-STD-461A data

c. Radar harmonic emission checks which use design, measured, or MIL-STD-461A data

d. Radar tunability check

e. Transmitter stability check

f. Receiver acceptance bandwidth check

g. Receiver image rejection check

h. Receiver spurious rejection check.

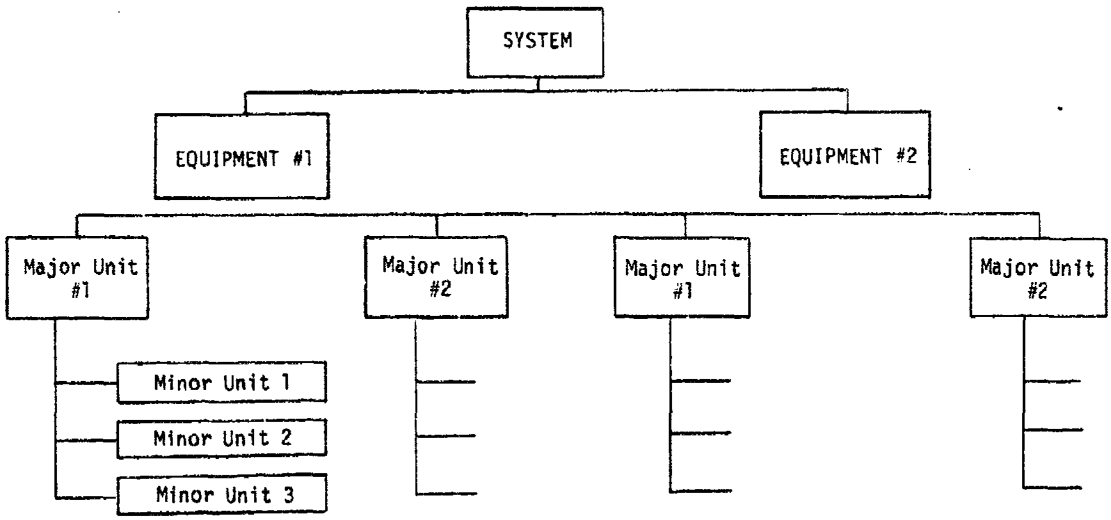


Figure 6-20. Simple System Organization

The MIL-STD-188 routine checks harmonic and spurious emission levels for global, tactical, and limited integration aircraft systems that fall within the frequency limits and modulation classes specified by MIL-STD-188.

6-4.9.4 Analysis of Transmitter Fundamental, Harmonic, and Spurious Emissions

This subroutine, as illustrated in Fig. 6-21, investigates frequency overlap between the fundamental and harmonics of intentional and spurious transmitter emissions, and the receiving bands of potentially susceptible devices. For those situations where overlap occurs, the possibility of the received signal exceeding the receiver sensitivity can be determined.

The frequency overlap check compares the tuning range of each fundamental and harmonic emission with a receiver tuning range, and identifies the percentage of the receiver range contained within the transmitter range of concern. When the interfering transmitter tuning range, or the mth harmonic of its range overlaps the upper end of the receiver tuning range, the overlap ratio r_o is

$$r_o = \frac{(FH)_R - m(FL)_T}{(FH)_R - (FL)_R}\text{, dimensionless} \quad (6\text{-}50)$$

where

$(FH)_R$ = highest frequency to which the receiver can be tuned, Hz

$(FL)_R$ = lowest frequency to which the receiver can be tuned, Hz

$(FL)_T$ = lowest frequency to which the transmitter can be tuned, Hz

m = harmonic number from transmitter

If the interfering transmitter tuning range, or the mth harmonic of its range, overlaps the lower end of the receiver tuning range,

$$r_o = \frac{m(FH)_T - (FL)_R}{(FH)_R - (FL)_R}\text{, dimensionless} \quad (6\text{-}51)$$

where

$(FH)_T$ = highest frequency to which the transmitter can be tuned, Hz

Similarly, if the interfering transmitter tuning range, or the mth harmonic of its range, fall totally within the receiver tuning range,

$$r_o = \frac{m(FH)_T - m(FL)_T}{(FH)_R - (FL)_R}\text{, dimensionless} \quad (6\text{-}52)$$

If the transmitter tuning range, or its harmonic, completely overlaps the receiver range the ratio is unity. Should the ratio be zero, the implication is that no transmitter power will get into the receiver and no magnitude or level computation is needed. If the ratio is small, interference might be avoided by a small change in frequency assignment.

The fundamental emission range and the ranges of the first 49 harmonics of a transmitter are processed in this way. Input spurious signals and their harmonics falling within the receiver tuning range also are identified. A user option provides for either a

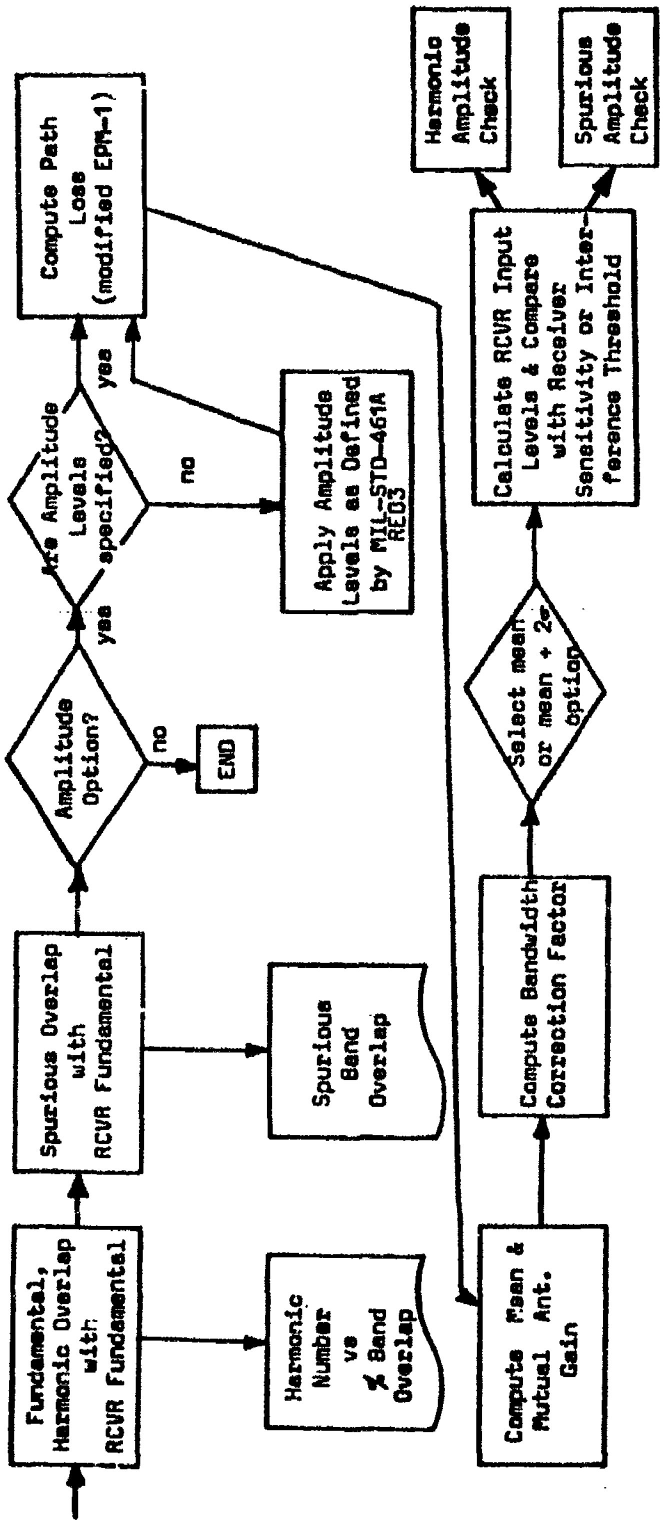


Figure 6-21. Simplified Logic for Spectrum Overlap and Interference Potential Check

"mean" computation (the relation between the expected received signal level and threshold), designated M or a "mean-plus-two-sigma" calculation (the relation between the value of signal level which is exceeded only about 2% of the time and the threshold), designated $M + 2\sigma$.

For an amplitude check, the program identifies the harmonic or spurious emission level of the transmitter for those cases which passed the frequency overlap check. Fundamental emissions that may be coupled into a receiver band are excluded from this analysis. A user option also is provided for either a "mean" or a "mean-plus-two-sigma" calculation.

The program also computes the propagation path loss between the emitter and receiver of concern. The mechanism for this computation is the EPM-1 Empirical Propagation Model reported by Frazier and Anderson (Ref. 21). This is a smooth earth, antenna height-dependent model which, in the version employed here, requires only four input parameters; namely, two antenna heights, a separation distance, and frequency. A standard deviation for the path loss also is determined. Fig. 6-22 provides a generalized illustration of the path loss model for use above 50 MHz.

The program then establishes the mean mutual antenna gain coupling value for the link of concern, and the standard deviation associated with this value. A bandwidth correction factor also is computed to account for the reduced coupling situations that occur when the emission bandwidth is large compared with the receiver bandwidth.

The output format of this evaluation is shown in Fig. 6-23, which is a sample run of the frequency overlap and transmitter harmonic/spurious interference checks.

The first section of the output summarizes several of the run input parameters for the emitter-receptor pair being examined, plus the mean antenna gain product *GM*, the standard deviation of the antenna gain product *SIGMA G*, and the standard deviation of the path loss *SIGMA P*. Other parameters in this section are:

ANTLB = antenna lobe
SS = sidelobe of transmitter to sidelobe of receiver
*GT*1 = main beam gain of transmitting antenna, dB
*GR*1 = main beam gain of susceptor antenna, dB
POP = power output, peak, kW
RSENS = receiver sensitivity or interference threshold, dBm

The next section of the output provides fundamental/harmonics checks and spurious/harmonic checks as follows:

a. The absolute level (dBm) of the harmonic (*HAR*) signal
b. Whether MIL-STD-461 was used to identify the harmonic level
c. The harmonic number resulting in spectrum overlap
d. The relative level (dB) of the harmonic signal
e. The percentage of receiver tuning range overlap (PCT OVLAP)
f. Which run options were used ("mean" or "mean-plus-two-sigma")
g. The interference level (dB) relative to the receiver interference threshold (INT MARGIN)

The spurious/harmonics checks are:

a. The absolute value (dBm) of the spurious/harmonic signal (*SHL*)
b. Whether MIL-STD-461 was used to identify the spurious/harmonic level
c. The harmonic number resulting in spectrum overlap
d. Frequency of spurious emission
e. The relative level (dB) of the spurious (*SPUR*)/harmonic signal
f. Which run options were used ("mean" or "mean-plus-two-sigma")
g. The interference level (dB) relative to the receiver interference threshold.

6-4.9.5 Receiver Spurious Responses, Intermodulation, and Cross-Modulation

6-4.9.5.1 Description of Analysis Routine

This analysis routine, shown in Fig. 6-24, investigates the possibility of the fundamental emitter antenna-radiated energy exceeding the sensitivity of a receiver spurious, intermodulation, or cross-modulation response. The reference levels are those in MIL-STD-461.

For each transmitter-receiver pair whose terminals are not intended to communicate with one another, the program determines whether spectrum overlap could occur within or outside of the receiver tuning range.

Using inputs describing receiver sensitivity, a calculation is made to determine whether or not the received interference signal will exceed the sensitivity

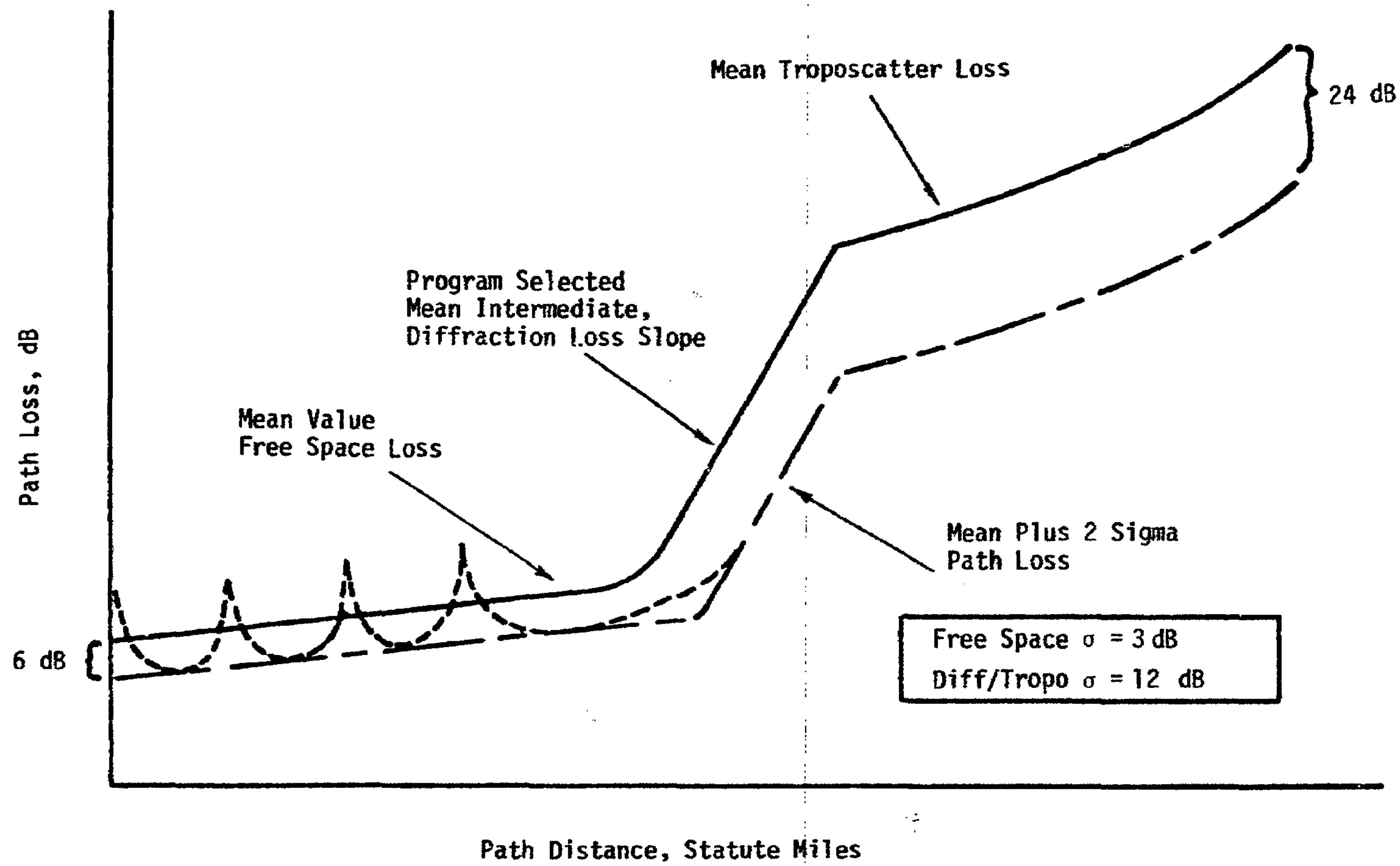


Figure 6-22. Representation of Mean and Mean + 2σ Path Loss Model (Modified EPM-1)

RECEPTOR			*EMITTER*			*ANTENNA COUPLING*							
NAME	LO FREQ MHZ	HI FREQ MHZ	NAME	LO FREQ MHZ	HI FREQ MHZ	ANT LB	GT1 DB	GR1 DB	GM DB	SIGMA G DB	SIGMA P DB	POP KW	RSENS DBM
UHF1	225.00	400.00	VHF-1	116.00	150.00	55	1.5	1.5	-10	4.2	0.0	0.020	-99

FUNDAMENTAL/HARMONICS CHECKS

SHL DBM	461 OPTION	HAR M	HAR LEVEL DB	PCT OVLAP	OPTION	INT MARGIN DB
-57	YES	2	70	38.9	M +2S	+30.4
-57	YES	3	70	29.7	M +2S	+30.4
-67	NO	2	80	38.9	M	+2.00
-67	NO	3	80	29.7	M	+2.00

SPURIOUS/HARMONICS CHECKS

SHL DBM	461 OPTION	HAR M	FREQ MHZ	SPUR LEVEL DB	OPTION	INT MARGIN DB
-57	YES	2	150.00	70	M + 2S	30.4

Typical output if mean value option was selected and input data was provided.

Figure 6-23. Harmonic-Spurious Overlap Amplitude Check Output Format

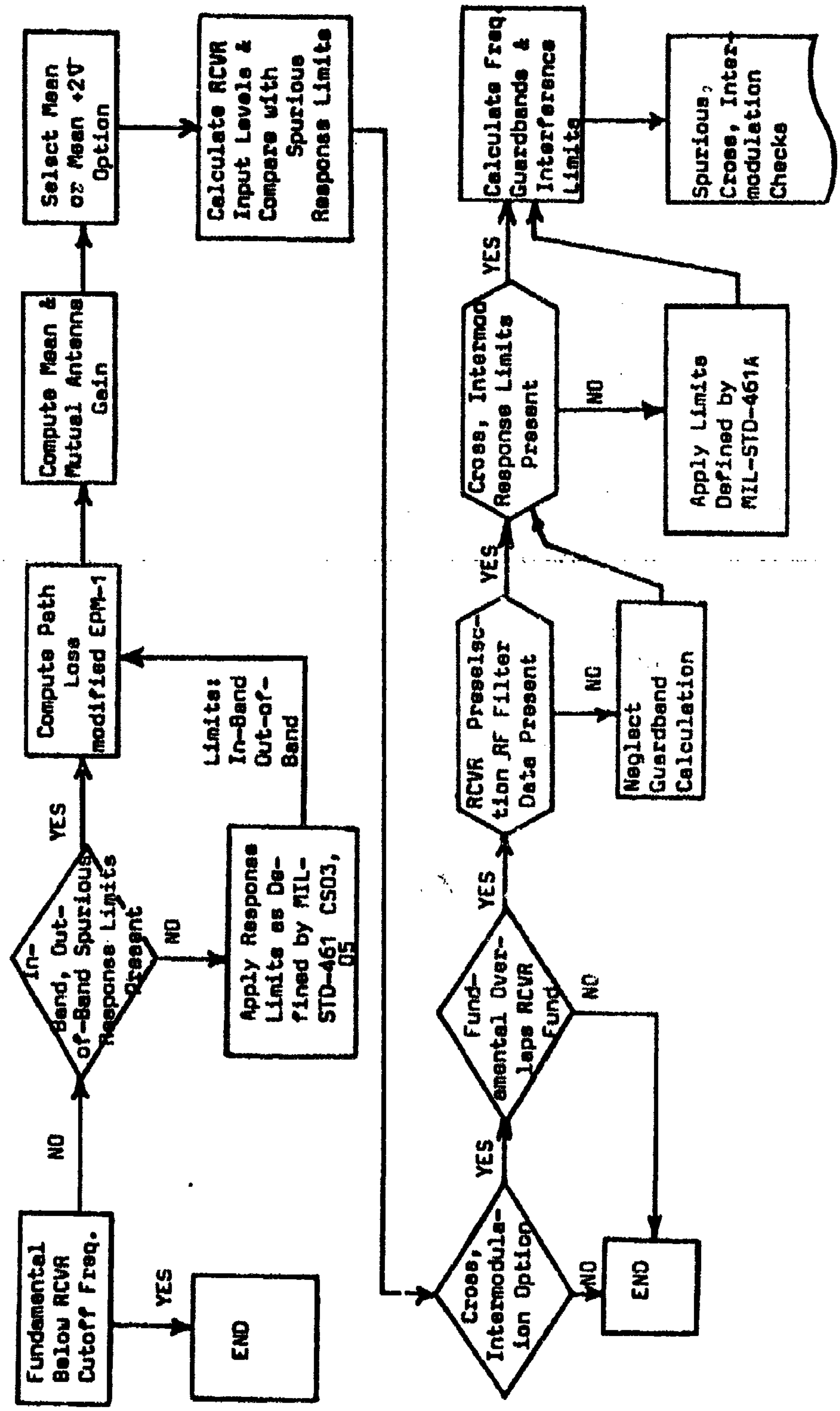


Figure 6-24. Simplified Logic for Receiver Analysis Checks

threshold. Eq. 6-53 is used to calculate the corresponding interference level IL.

$$IL = 10 \log POP + GM + LPROP$$
$$= RSENS - SPUR + 60 \text{ , dB} \qquad (6\text{-}53)$$

where

POP = peak power, kW
$LPROP$ = path loss between antennas, dB

6-4.9.5.2 Results of Receiver Analysis Routine Operation

The output format of the evaluation is shown in Fig. 6-25 which is a sample-run of the receiver spurious, intermodulation, and cross-modulation routine. The columns of significance are:

a. Receiver Spurious Check:

(1) The run option (mean, or mean + 2σ) relative to this test

(2) The interference level (dB) relative to the receiver sensitivity

(3) The receiver spurious level (dB) used in the calculation.

b. Cross-modulation Check:

(1) The interference level (dB) relative to the receiver cross-modulation threshold

(2) The guard band (MHz) required when using the specified preselection filter. In the example (first line), the guard band is 1.9 MHz and the interference margin is 56 dB. Thus a preselection filter having a greater than 56 dB rejection 1.9 MHz off the carrier frequency must be inserted in the receiving system.

(3) The cross-modulation threshold value specified by the user or by MIL-STD-461.

c. Intermodulation Check:

(1) The interference level (dB) relative to the receiver intermodulation threshold

(2) The guard-band (MHz) required when using the specified preselector filter

(3) The intermodulation threshold value specified by the user or by MIL-STD-461.

d. Data. This section contains the data used in the computations. ALS means "antenna lobe selection" and SS means "sidelobe to sidelobe" as before.

6-4.9.6 Description of Local Oscillator Radiation

The received interference level IL above the receiver sensitivity level, resulting from local oscillator radiation, is calculated by Eq. 6-54.

$$IL = PLO + GMA + LPROP$$
$$- RSENS \text{ , dB} \qquad (6\text{-}54)$$

where

PLO = local oscillator power present at antenna terminals of interfering receiver, dBm
GMA = mutual antenna gain, dB

The output format for this analysis routine is shown in Fig. 6-26, which is a sample run for the local oscillator overlap and amplitude checks.

6-4.9.7 Analysis of Cable, Case, and Antenna Coupling

A much simplified block diagram of this subroutine is presented in Fig. 6-27. The levels of signals emanating from cases and cables are based on MIL-STD-461 limits.

Emission limits for broad/narrowband conducted (CE01, CE02, CE03, CE04), radiated magnetic field (RE01), and radiated broad/narrowband electric field (RE02) are considered for all case and cable minor units. Transmitter/transceiver minor units having antenna ports use the transmitter peak power and main beam antenna gain for the calculation of effective radiated interference power.

Coupling calculations for wire and case combinations are based on the SEMCAP equations, as discussed in par. 6-4.8.1. The output format of this analysis routine is shown in Fig. 6-28. The first section of the output identifies the susceptor and generator pairs being examined; minor unit names and identification numbers are presented. The next section of the output presents data concerning the interference modes and contains four subsections: antenna generated interference, conducted emission, radiated E-field and radiated H-field.

The column headed "'Antenna Generated Interference" presents interference signals which are coupled into the susceptor wire port. Calculations are performed using the mid-band operating frequency of the transmitter which feeds the antenna port.

The column "Conducted Emission Mode" considers both narrow and broadband emissions and provides the level of those interference signals that are coupled through capacitive and inductive transfer. Calculations are made using the mid-band frequency of the wire information and the bandwidth.

Columns headed "Radiated E-Field" and "Radiated H-Field" present interference levels coupled into the receptor wire through E- and H-field transfer.

The third section of the output format is entitled "Interference Data", This section presents the total narrow and broadband interfering signals. These values represent the sum of entries for conducted and radiated modes. In addition, the interference levels are compared to the desired signal and safety margin.

RECEIVER CHARACTERISTICS FREQUENCY AND AMPLITUDE CHECKS

RX LIST		TX LIST			RX SPURIOUS		CROSS MODULATION			INTERMODULATION			DATA			
				OPT	INT MARG DB	SPUR DB	INT MARG DB	GRD BAND MHZ	THLD DB	INT MARG DB	GRD BAND MHZ	THLD DB	POP KW	RSENS DBM	GM DB	ALS
UHF-1		UHF-2		m	+42.0	80.0	+56.0	1.9	66	+56.0	1.9	66	0.020	-99	-10	SS
FL MHZ	225.0		225.0													
FH MHZ	400.0		400.0													
UHF-1		UHF-2		m	+22.0	00.0	+10	0.20	102	+52	1.8	70	0.020	-99	-10	SS
FL MHZ	225.0		225.0													
FH MHZ	400.0		400.0													

MIL-STD-461 LIMIT

Measured Limits

Cross Mod: 102 DB at ± 10 KHz

Intermod: 70 DB down at ± 500 KHz

Figure 6-25. Output Format for Receiver Characteristic Analysis Routine

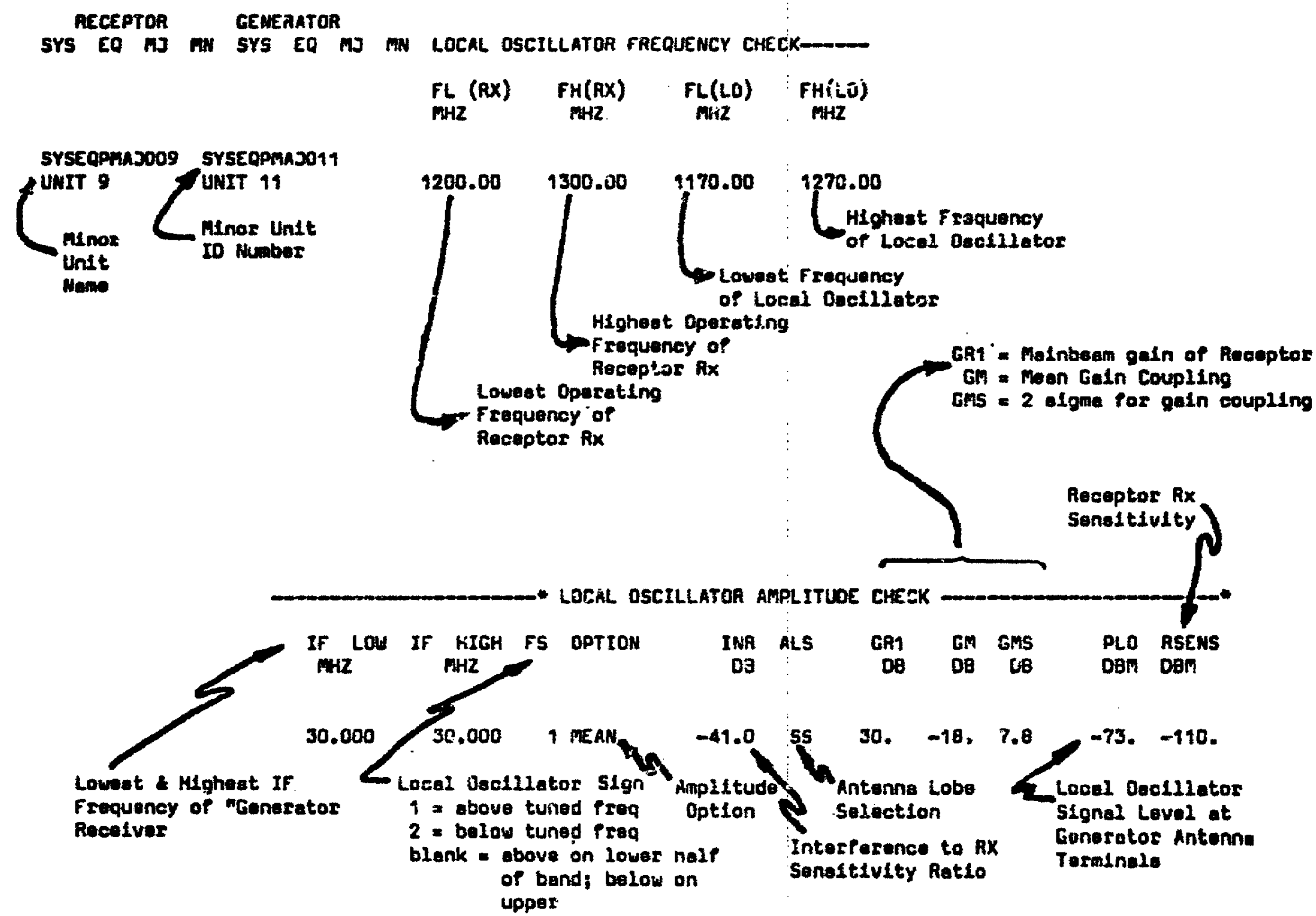


Figure 6-26. Output Format for Local Oscillator Analysis Routine

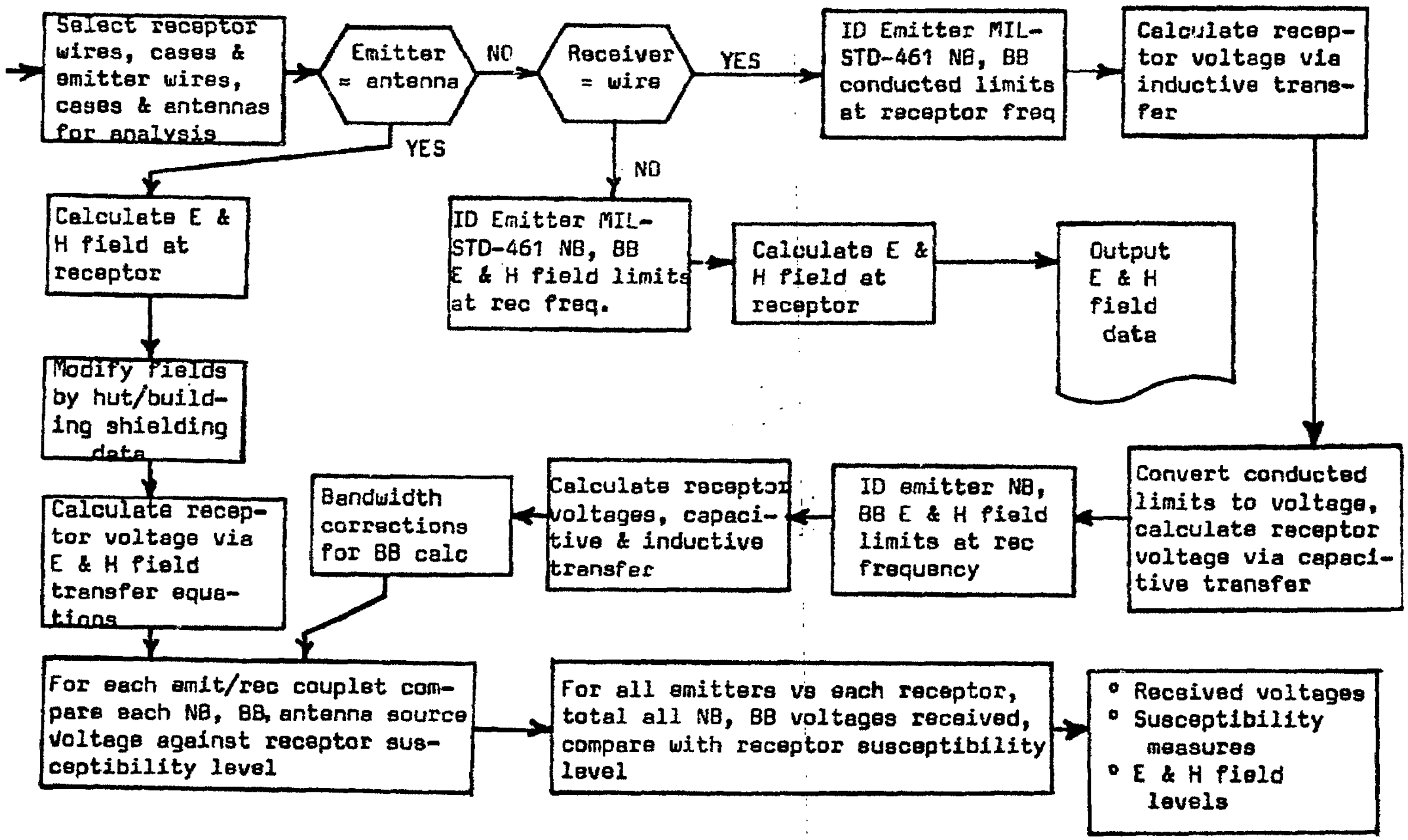


Figure 6-27. Simplified Block Diagram for Close-coupled Routine

```
                                      .----------------CLOSE COUPLING ANALYSIS RESULTS--------.

SUSCEPTOR/    .----------------------------------INTERFERENCE         MODES --------------------.
    GENERATOR
                                     CONDUCTED EMISSION          RADIATED E FIELD       RAD H
                  ANTENNA          --------------------------    ------------------
SYS EQ MJ MN     GENERATED         NARROWBAND      BROADBAND     NARROW-     BROAD-     FIELD
               INTERFERENCE        CAP     IND     CAP     IND    BAND        BAND
                   VOLTS           VOLTS   VOLTS   VOLTS   VOLTS  VOLTS       VOLTS      VOLTS
  1   1   1   1
HOT LNTOWER
PORTYP=WIRE

   1    1    1    2
  HF  TXMTR
  PORTYP=ANT         .5818158   .00000  .00000  .00000  .00000  .00000  .00000  .00000

                      TOTAL FOR SUSCEPTOR    1   1   1   1

                                  *--- INTERFERENCE       DATA ----*  * FIELD STR AT RC

                                  COUPLED SIGNALS  ACCEPTABILITY     E FIELD    H FIELD
NOTE:  PORTYP = PORT TYPE         ---------------  -------------    --------   -------

                                  NARROW-  BROAD   NARROW-  BROAD-   NAR   BRD
                                   BAND    BAND     BAND    BAND     BAND  BAND   AMPS/
                                  VOLTS    VOLTS    VOLTS   VOLTS    VOLTS VOLTS  METER

                                  .00000   .00000   .00000  -.56957  .0004  357   .000

                                  .00000   .00000   .00000  .00000
```

Figure 6-28. Output Format for Close-coupled Analysis Routine

The result of this comparison is presented under the "Acceptability" heading. The last section presents *E*- and *H*-field strength levels at the receptor wire or case.

After all generators have been played against the susceptor the total interference signal present on the wire is calculated. This interference signal then is used to calculate the total acceptability for the susceptor wire being examined. These data are found below the last generator entry for the susceptor being examined.

6-4.9.8 Implementation

The ISCAP program runs on a UNIVAC 1108. Information on its use can be obtained from the Electronic Systems Division (AFSC), Hanscom AFB, MA 01731.

6-4.10 INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM (IEMCAP)

6-4.10.1 Introduction

The Intrasystem Electromagnetic Compatibility Analysis Program (IEMCAP) (Ref. 22) was developed to provide an economical implementation of EMC analysis at all stages of a system life cycle, from conceptual studies of new systems to field modification of old systems. This is accomplished by performing the following tasks:

a. Provide a data base which can be continually maintained and updated to follow system design changes.

b. Generate EMC specification limits tailored to the specific system.

c. Evaluate the impact of granting waivers to the tailored specifications.

d. Survey a system for incompatibilities.

e. Assess the effect of design changes on system EMC.

f. Provide comparative analysis results on which to base EMC trade-off decisions.

6-4.10.2 Models

6-4.10.2.1 Emitters

Emitter models are incorporated in the program which correspond to common signal sources—single pulse, repetitive pulse, information-bearing pulse, and CW signals. Provision is made for user input of spectral densities for those types not modeled. The following sources are modeled:

a. Binary frequency-shift keying

b. Amplitude modulation with stochastic process

c. Angle modulation with stochastic process

d. Single sideband amplitude, phase, and frequency modulation with a stochastic process

e. Chirp radar spectra

f. Pulse code modulation/amplitude modulation—nonreturn to zero

g. Pulse position modulation

h. Pulse code modulation/amplitude modulation - biphase

i. Pulse amplitude modulation/frequency modulation

j. Pulse duration modulation

k. Single pulse

l. Pulse train (rectangular, trapezoidal, triangular, sawtooth, high frequency exponential, damped sinusoidal).

6-4.10.2.2 Susceptors

The basic approach is to accept input data on in-band sensitivity, along with a bandwidth parameter, and then to form a rectangular-shaped susceptibility function in the required spectral region that connects to nonrequired spectra defined by user-adjusted military specification interference limits. An RF receiver representation in the program will, in general, actually have a trapezoidal shaped susceptibility function (in-band) due to the skirt slopes of the normal selectivity curve. Specifically, the susceptibility of an RF port is assumed to be equal to the tuned sensitivity of the receiver, as provided in the input data, over the entire frequency range defined by the user-specified bandwidth. The susceptibility of a signal or control port is assumed to be equal to the operating level [dB(μA)] less 20 dB. This somewhat arbitrary susceptibility level is based on characteristics of common avionic equipments. If the user wants a higher or lower susceptibility level in the required range of a signal or control port, he need only specify a higher or lower operating level, respectively.

In the case of receivers where more is known about the details of the response curve than just the flat response previously discussed, the user can specify the known response curve by a discrete spectrum of up to ten frequencies with associated levels. Interference calculations using either the specified spectrum or the flat response function are conducted by weighting the received interference signal power by the ratio of the receptor to emitter bandwidths, when the former is the smaller value of the two.

The user has an additional option of augmenting either of the two preceding models with certain filter functions as described in par. 6-4.10.2.3.1. When thus used, the filter models relate the susceptibility level of a given susceptor to the level at the filter input terminals for compatibility assessment.

The combination of user-supplied susceptor spectra with the filter models in the program provides the user considerable flexibility.

6-4.10.2.3 Transfer Models

These models are used to compute the ratio between the energy output at an emitter port and that present at the input to a susceptor port. For example, the antenna-to-antenna transfer model computes the ratio of the energy output of a transmitter to the energy at the input of the receiver. Propagation models are incorporated in the program for free-space transmission, ground-wave propagation, and antennas mounted on aircraft bodies. Receiver models relate the energy spectrum at the receiver port to the response produced by that spectrum. This calculation is based on the sensitivity of the receiver versus frequency.

6-4.10.2.3.1 Filter Models

Seven filter models are used: single tuned, transformer coupled, Butterworth tuned, low pass, high pass, band pass, and band reject. The models represent filters as ideal, lossless networks, made up only of reactive elements (capacitors and inductors). The filter transfer models calculate the insertion loss in dB provided by a filter at a given frequency, i.e., the reduction is delivered power due to insertion of a filter. Provision is made for the entry of a minimum insertion loss to represent the actual dissipation at the tuned frequency or in the pass band.

6-4.10.2.3.2 Antenna Model

Antennas are categorized into two groups. The first group includes low gain antenna types such as a monopole, dipole, slot, or loop. The second group includes medium to high gain types, such as those using horns or parabolic reflectors.

All antennas in the first group are modeled analytically by trigonometric expressions. A dipole, for example, has a directive gain $G_\alpha \approx 1.6 \sin^2\theta$, where θ is the angle of the direction with respect to the dipole axis.

All antennas in the second group are modeled by a three-dimensional three-sector representation. Each sector subtends a solid angle in the unit sphere and has an associated quantized antenna gain.

6-4.10.2.3.3 Field-to-Wire Compatibility Analysis

Coupling from environmental electromagnetic fields onto wiring usually occurs via the fields entering through dielectric apertures in the system skin and coupling to immediately adjacent wires. Exposed wires are assumed to be adjacent to the aperture, and the amount of RF energy coupled depends on the aperture size and location. A transmission line model similar to those described in par. 6-4.8 is then used to compute the currents induced in the wires. Worst-case electromagnetic field vector orientation is determined and used for the calculation.

6-4.10.2.3.4 Wire-to-Wire Coupling

During the calculation of the coupling from an emitter port to a particular receptor port, a check is made to determine if any wires connected to any of the emitter ports are in the same bundle and run as wires connected to the receptor port. If there are such wires, the wire-to-wire coupling routine is called. This routine, similar to those of par. 6-4.8, computes the spectral voltages induced in the receptor circuit by the emitter circuit. These calculations are performed on a pair basis (only one emitter circuit considered to couple with the receptor circuit for each calculation). Each possible pair coupling is computed in turn, and the total coupling is calculated by summing all of the pair couplings without regard to phase.

6-4.10.3 Basic Analysis Approach

All intentional ports must generate and/or receive certain types of signals to perform their intended function. The signals or responses which intentionally are generated and coupled from port to port are called operationally required and cannot be altered without affecting system operation. In addition to the required signals, undesired outputs and/or responses may exist which are called operationally nonrequired. The two categories are essentially the same as the functional and extraneous categories of SEMCAP.

An incompatibility is said to exist when sufficient signal from an emitter port, or ports, unintentionally is coupled to a susceptor port and exceeds its susceptibility threshold. The limits for nonrequired signals are listed in the EMC specifications. Ideally, if all ports have no emissions and susceptibilities exceeding these limits, the system is compatible. An important task of IEMCAP is the generation of a set of specification limits tailored to the specific system under analysis.

The emissions and susceptibilities, both required and nonrequired, are represented by spectra. For each emitter, a two-component spectrum (broadband and narrowband) represents the power levels produced over the frequency range. The broadband component represents continuous emissions which vary slowly with respect to frequency, while the narrowband component represents discrete emissions which vary rapidly with respect to frequency. The broadband components are in units of power spectral density, and the narrowband components are in units of power.

For each susceptor, a spectrum represents the susceptibility threshold over the frequency range. The susceptibility level is defined as the minimum received signal which will produce a response at a given frequency.

For each intentional source or receiver port, a portion of the frequency range is defined as the required range. All signals within this range are required and cannot be adjusted. Outside this range, limits may be set for the maximum emission and minimum susceptibility levels. Within the required range, the spectrum is defined by a mathematical model of signal level versus frequency. This can be either from equations of the frequency domain representation of the signal or directly from a user-defined spectrum. Outside the required range, assumed levels are used for the port spectra.

During specification generation, if these assumed spectrum levels cause interference, they are adjusted for compatibility. By adjusting the spectra of emitters and susceptors for compatibility the maximum nonrequired emission and minimum susceptibility levels are obtained for a compatible system. To prevent overly stringent specifications from being generated, each spectrum has an adjustment limit.

Initially, the limits of military EMC specifications MIL-STD-461 and MIL-I-6181 are used. The levels may be relaxed or tightened from these if desired. These specifications are used for a variety of reasons. First, since they are widely used, most EMC engineers are familiar with them. Those portions of the port spectra not requiring adjustment remain at the usual specification levels. Also, if new equipments are added to a system containing existing equipment developed and tested to these specifications, the IEMCAP-generated specifications will be at the same general levels and not require radical changes in EMC design. This also facilitates adapting an equipment from one system to another system.

The general approach in performing the various tasks is two-fold. First, a susceptor is selected and its type, connection, wire routing, etc., is examined to determine if a coupling path exists. If a path exists, the received signal is computed at the susceptor and compared to the susceptibility level. In addition to the emitter-susceptor port pair analysis, the program also computes the total signal from all emitters coupled into each susceptor acting simultaneously. Details of the two analyses are contained in Ref. 22.

In determining the spectra and in other phases of the analysis, the termination impedance must be known. A wide variety of terminations can exist in the system, but for the most part they can be represented by a series resistor and inductor with a shunt capacitor.

Power transfer relationships are normalized to a 1 ohm impedance. The actual port impedances that are input to the program then are used to transform these power relationships appropriately.

6-4.10.4 Spectrum Representation

The fundamental computation carried out by IEMCAP is one giving the power into the susceptor and is essentially the same as Eq. 6-1 under the Allen Model (par. 6-4.2.1). Note that it is also similar to Eq. 6-48 under SEMCAP (par. 6-4.8.2.1) except that the latter uses source amplitudes and transfer function amplitudes rather than power spectral densities and power transfer functions.

A sampled spectrum technique is used in which each spectrum amplitude is sampled at various frequencies across the range of interest. Considering the requirement of MIL-STD-461 of 3 frequencies per octave from 30 Hz to 18 GHz, this requires approximately 90 sample frequencies. This is a reasonable resolution for EMC specifications in which limits of emission and susceptibilities are set and can apply over large regions of the spectrum. If greater resolution is desired, the user can specify individual frequencies. To avoid missing narrow peaks between sample frequencies, the spectrum is sampled in the interval half-way between the sample frequency and each of its neighboring sample frequencies. For emission spectra, the maximum in the interval is used; for susceptibility spectra, the minimum is used. This effectively quantizes the spectra with respect to the sample frequencies.

To minimize core memory and data file size requirements, a table of sample frequencies is defined for an equipment, and all spectra of ports within that equipment are quantized at these frequencies. This eliminates storing a table of sample frequencies for each port which saves 50,400 words of file storage, plus the input/output time required to store and retrieve the information.

The equipment frequency tables can be defined using two options. First, the user may specify the upper and lower frequency limits, the maximum number of frequencies (up to 90), and the number of frequencies per octave. The program then generates a table of geometrically spaced frequencies within the specified limits. Optionally, the user may specify the upper and lower frequency limits, the maximum number of frequencies, and a number of specific frequencies (up to the maximum number) of interest. The program then generates geometrically spaced frequencies to fill in the number of frequencies not specified. For example, if the maximum number of frequencies to be used is 90 and 10 are user specified, the program generates 80 geometrically spaced frequencies over the specified frequency range and inserts the 10 user frequencies at the appropriate places.

The range of frequencies covered by the analysis is controlled by the user. The program will accept any range from 30 Hz to 18 GHz but, if desired, the user may concentrate all 90 frequencies over a smaller interval within this range.

Each port is categorized by function into one of six types, each type having its own subinterval of frequencies within the overall frequency range. These subintervals, adapted from MIL-STD-461/462 ranges for the port function, are shown in Table 6-10. The nonrequired spectrum model routines will generate zero emission and susceptibility outside the subintervals.

The spectra and amplitudes are represented within up to 90 contiguous intervals across the frequency range of interest quantized to the sample frequencies. This representation allows for flexibility and program efficiency. It also allows the program to be divided into two sections, each running in 60k (decimal) of core memory. With this arrangement, the program is readily adaptable to a wide variety of computers, and machine-dependent techniques, such as overlaying, are not required. One section of the program contains the data management and spectrum model routines, and the other contains the analysis and transfer model routines. Each section is executed separately so that both are not in core at one time.

6-4.10.5 Logic Flow

Fig. 6-29 shows the basic functional flow through its two sections. These sections are executed independently with intermediate data storage on a number of disc or tape files known as working files. Depending on the analysis and the size of the system being analyzed, the program sections can be executed in succession or run separately.

The functions performed by each program section are described briefly in the paragraphs that follow. Details can be found in Ref. 22.

The first section called the Input Decode and Initial Processing Routine (IDIPR), is divided into three basic routines (Fig. 6-29). The Input Decode Routine (IPDCOD) reads and decodes the free-field input data from punched cards and checks the data for errors.

The program then proceeds into the Initial Processing Routine (IPR). This routine performs data management, interfaces with spectrum models, and generates the working files. Data defining the system and all of its components are stored on a magnetic disc or tape called the Intrasystem File (ISF). This file is a data base which is maintained by IPR, incorporating changes in the system design. For a given run, the system to be analyzed can be defined from punched cards only, from a previously created ISF, or from an old ISF updated by cards. IPR assembles and merges the data to be analyzed and writes these

TABLE 6-10
PORT EMISSION AND SUSCEPTIBILITY TESTS AND FREQUENCY RANGES

PORT FUNCTION	EMITTER		RECEPTOR	
	MIL-STD-462 Test(s)	Freq Range, Hz	MIL-STD-462 Test(s)	Freq Range, Hz
RF	CE06	14k-18G	CS04	14k-18G
Power	CE02/03	30-50M	CS01/02	30-400M
Signal	CE02/04	30-1G	CS02/04	30-10G
Control	CE02/04	30-1G	CS02/04	30-10G
EED	—	—	CS02/04	30-10G
Eqpt Case	RE02	14k-10G	RS03/04	14k-10G

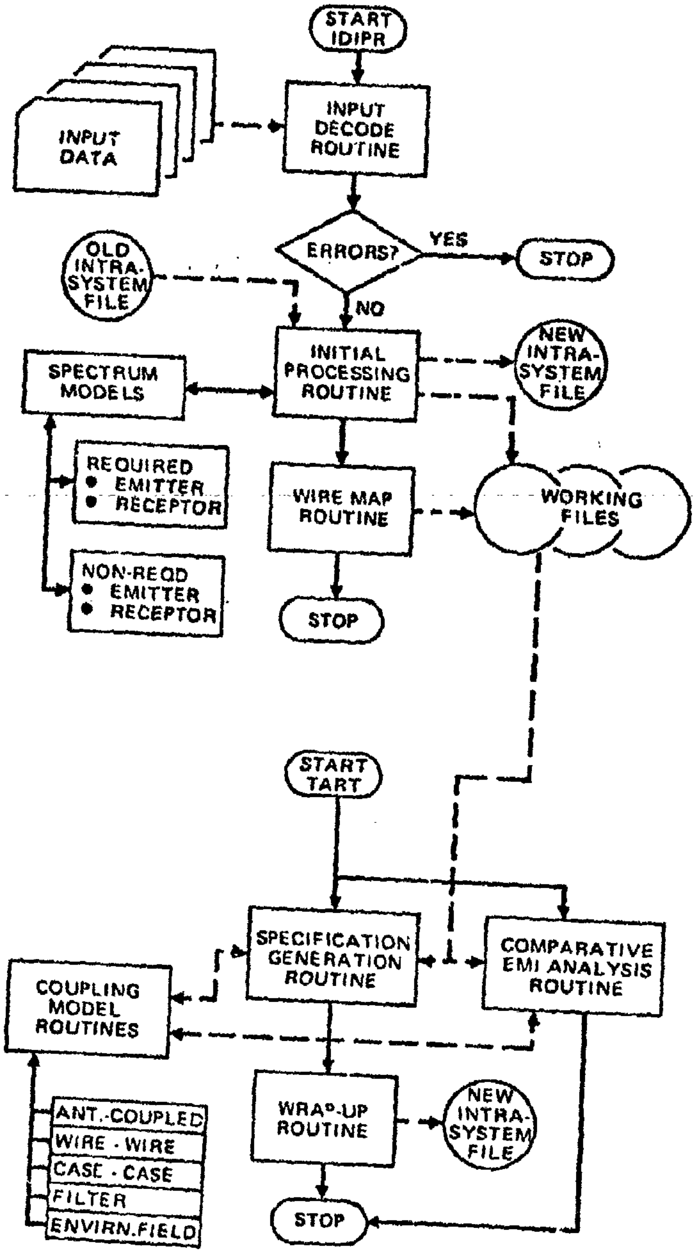


Figure 6-29. IEMCAP Functional Flow

data on a new ISF for future runs and on the working files for analysis. During this process, IPR interfaces with the spectrum math model routines. These use the user-specified port spectrum parameters to compute the spectrum amplitudes quantized to the equipment sample table frequencies for each port. From IPR, the program enters the wire map routine which generates cross-reference map arrays for use by the wire coupling math models during analysis. At this point, execution of IDIPR terminates. The computer can either stop or continue into the second section, depending on the job setup.

The second section called the Task Analysis Routine (TART) uses the data compiled by IDIPR to perform the desired analysis task. This is one of the four tasks summarized as follows:

a. Specification Generation. Adjusts the initial nonrequired emission and susceptibility spectra such that the system is compatible, where possible. A summary of interference situations not controlled by EMC specifications is printed. The adjusted spectra are the maximum emission and minimum susceptibility specifications for use in EMC tests.

b. Baseline System EMC Survey. Surveys the system for interference. If the maximum of the EMI margins over the frequency range for a coupled emitter-receptor port pair exceeds the user specified printout limit, a summary of the interference is printed. Total received signal into each receptor from all emitters also is printed.

c. Trade-off Analysis. Compares the interference for a modified system to that from a previous specification generation or survey run. The effect on interference of antenna changes, filter changes, spectrum parameter changes, wire changes, etc., can be assessed from this analysis.

d. Specification Waiver Analysis. Shifts portions of specific port spectra as specified and compares the resulting interference to that from a previous specification generation or survey run. From this, the effect of granting waivers for specific ports can be assessed.

TART is composed of two basic routines (Fig. 6-29). The Specification Generation Routine (SGR) performs the first task previously described and the Comparative EMI Analysis Routine (CEAR) performs the remaining three. These interface with the coupling math model routines to compute the transfer ratios between emitter and receptor ports.

6-4.10.6 System/Subsystem Specification Generation

The Specification Generation Routine (SGR) attempts to adjust the nonrequired portions of the port spectra, initially computed to IPR, to produce a compatible system. These spectra are limits. Thus, an emitter cannot generate outputs greater than the nonrequired spectrum levels, and a receptor cannot respond to received signals less than these levels or interference will result. For the analysis, each port is assumed to emit and respond at these levels. For each emitter-receptor port pair in the system with a coupling path between them, the received signal is computed using the assumed maximum emission levels. This signal is compared to the assumed minimum susceptibility levels over the frequency range, and where the susceptibility level is exceeded in the emitter nonrequired range, the emission levels are reduced such that the margin is equal to the user-defined adjustment safety margin (asm) or to the adjustment limit level, whichever is greater.

When each emitter has been adjusted in conjunction with each receptor, the receptor spectra are adjusted. The received signal from each emitter with a coupling path to a given receptor is computed using the adjusted emission spectra and summed. The susceptibility spectrum levels then are compared to this total signal, and where the level is exceeded in the nonrequired range, the susceptibility is raised such that the margin is asm, or to the adjustment limit level, whichever is less.

As a result of this process, a set or port spectra is generated which must be met if the system is to be compatible. These then, are EMC specification limits to which the equipment ports can be tested.

After the adjustment process, a number of port pairs may exist which are still incompatible. This is called unresolved interference and results from required emissions and responses, nonrequired spectra adjusted to their limits, and from nonadjustable spectra of previously procured equipments. Consequently, after susceptor adjustment, SGR recomputes the interference between adjusted emitters and adjusted susceptors. If the maximum of the EMI margin exceeds a user specified printout limit, the case is printed as unresolved interference along with a summary of the spectrum levels and the EMI margins.

6-4.10.7 Outputs

Outputs are provided for the three SGR phases: emitter spectrum adjustment, susceptor adjustment, and unresolved EMI. After these, the finally adjusted spectra are summarized for each port. An example summary of the adjusted emitter broadband and narrowband spectra is shown in Fig. 6-30. After a given susceptor port has been adjusted, SGR scans through

ADJUSTED EMITTER SPECTRA

EMTR -- SUBS = CNI EQPT 5 = TACAN PORT 2 = TACRF

RCPT -- SUBS = INGPD EQPT 0 = DDPYL PORT 4 = DBDR

PATH = ANT TO ANT

			NARROWBAND					BROADBAND					
FREQUENCY (HERTZ)	FREQ BASE	TRANSFER RATIO	ADJSTD EMI MGN	RCVD SIGNAL	ADJ EMTR SPCT LEV	ADJMT AMOUNT	SPCT LEV TO LIMIT	ADJSTD EMI MGN	RCVD SIGNAL	ADJ EMTR SPCT LEV	ADJMT AMOUNT	SPCT LEV TO LIMIT	BDWTH FCTR
1.19689E+04	EMTR	0.0	-94.0	6.0	6.0	0.0	30.0	-98.4	1.6	26.0	0.0	30.0	-24.4
2.03502E+04	RCPT	0.0	-94.0	6.0				-96.1	3.9				-22.1
2.53638E+04	EMTR	0.0	-94.0	6.0	6.0	0.0	30.0	-95.2	4.8	26.0	0.0	30.0	-21.2
3.98795E+04	RCPT	0.0	-94.0	6.0				-93.3	6.7				-19.3
5.34919E+04	EMTR	0.0	-94.0	6.0	6.0	0.0	30.0	-96.7	3.3	26.0	0.0	30.0	-22.7
7.56093E+04	RCPT	0.0	-94.0	6.0				-95.2	4.8				-21.2
1.13895E+05	EMTR	0.0	-94.0	6.0	6.0	0.0	30.0	-93.4	6.5	25.0	0.0	30.0	-19.5
1.43973E+05	RCPT	0.0	-94.0	6.0				-92.4	7.6				-19.4
2.39069E+05	EMTR	0.0	-94.0	6.0	5.0	0.0	30.0	-90.2	9.9	26.0	0.0	30.0	-16.2
2.76342E+05	RCPT	0.0	-96.1	3.9				-95.4	4.6				-15.6
5.85444E+05	EMTR	0.0	-34.1	-5.2	-5.2	-11.2	19.8	-28.1	.8	-4.0	-30.0	0.0	4.8
5.30413E+05	RCPT	0.0	-28.4	-5.2				-22.4	.8				4.8
1.00860E+06	RCPT	0.0	-10.4	-5.2				-4.4	.8				4.8
1.01889E+06	RCPT	0.0	-10.4	-5.2				-4.4	.8				4.8
1.86445E+06	EMTR	0.0	-10.4	-5.2	-5.2	-11.2	18.8	-4.4	.8	-4.0	-30.0	0.0	4.8
1.95413E+06	RCPT	0.0	-15.3	-10.1				-4.4	.8				4.8
2.25877E+06	EMTR	0.0	-16.5	-11.3	-11.3	-17.3	12.7	-4.4	.8	-4.0	-30.0	0.0	4.4
3.75078E+06	RCPT	0.0	-3.9	-4.7				-4.4	.8				4.8
4.00009E+06	RCPT	0.0	-9.3	-3.4				-4.4	.8				4.8
4.77517E+06	EMTR	0.0	-43.7	-1.5	-1.5	-7.5	22.5	-41.4	.8	-4.0	-30.0	0.0	4.8
5.03808E+06	RCPT	0.0	-52.8	-1.0				-49.2	2.6				4.9
7.19918E+06	RCPT	0.0	-97.4	2.5				-90.5	9.5				-3.0
1.08958E+07	EMTR	-2.3	-96.2	3.9	6.0	0.0	30.0	-79.2	20.9	26.0	0.0	30.0	-3.0
1.39180E+07	RCPT	-5.5	-39.5	.5				-81.1	18.9				-1.6
2.13414E+07	EMTR	-11.3	-105.3	-5.3	5.0	0.0	30.0	-85.0	15.0	26.0	0.0	30.0	.3
2.65223E+07	RCPT	-14.2	-108.2	-8.2				-87.0	13.0				1.2
4.51169E+07	EMTR	-21.3	-115.3	-15.3	6.0	0.0	30.0	-91.7	8.3	26.0	0.0	30.0	3.5
5.09069E+07	RCPT	-22.9	-116.9	-16.9				-92.8	7.2				4.1
9.53794E+07	EMTR	-31.4	-125.4	-25.4	5.0	0.0	30.0	-98.6	1.4	26.0	0.0	30.0	6.9
9.77103E+07	RCPT	-31.7	-125.7	-25.7				-98.8	1.2				6.9
1.87547E+08	RCPT	-40.5	-134.6	-34.5				-107.9	-7.9				6.2
2.81639E+08	EMTR	-41.6	-135.6	-35.6	6.0	0.0	30.0	-108.6	-8.6	26.0	0.0	30.0	7.0
3.59978E+08	RCPT	-49.7	-143.7	-43.7				-114.2	-14.2				9.5
4.25274E+08	EMTR	-52.1	-146.1	-46.1	5.6	0.0	30.0	-115.9	-15.9	26.0	0.0	30.0	10.3
6.90942E+08	RCPT	-59.0	-153.0	-53.0				-120.5	-20.5				12.4
8.00000E+08	EMTR	-61.1	-155.1	-55.1	5.0	0.0	30.0	-122.1	-22.1	26.0	0.0	30.0	13.9
9.01198E+08	EMTR	-62.9	-156.9	-56.9	6.0	0.0	30.0	-123.3	-23.3	26.0	0.0	30.0	13.5
9.62033E+08	EMTR	-63.9	-157.8	-57.9	5.0	0.0	30.0	-124.3	-24.3	26.0	0.0	30.0	13.8
1.02590E+09	EMTR REQD	-64.8	-30.8	70.0	134.8	0.0	30.0	-71.3	28.7	97.5	0.0	30.0	-4.8
1.13003E+09	EMTR REQD	-66.2	-31.4	58.6	134.8	0.0	30.0	-72.7	27.3	97.5	0.0	30.0	-4.0
1.21800E+09	EMTR	-67.3	-161.3	-61.3	6.0	0.0	30.0						
1.29000E+09	EMTR	-68.1	-162.1	-62.1	6.0	0.0	30.0						

Figure 6-30. Sample Output — SGR Adjusted Emitter Spectra

the emitters coupled to it and computes the margins. If the maximum margin exceeds the user specified printout limit, a summary is printed, as illustrated in Fig. 6-31.

Printouts also can be provided for baseline system EMI survey outputs, trade-off, and waiver outputs. Error printouts, input data card listings, and intra-system signature file reports can be obtained as part of the IDIPR printed outputs.

6-4.10.8 Availability

The program is written in FORTRAN IV for the CDC-6600 or the Honeywell 635. It is divided into two sections, each of which requires 60k (decimal) of core memory (par. 6-4.10.4). One section is for data setup and the other is for the actual program run. A 7- or 9-track tape is used, with ACSII II code.

The program, as well as technical assistance in its use, can be provided via Code RBC, Rome Air Development Center, Griffis Air Force Base, NY. Copies of the program are made available to those who provide a blank tape for program recording. The tape is recorded in card image format.

6-4.11 SIGNCAP I

6-4.11.1 Objectives

This nonlinear circuit analysis program (Refs. 23, 24, and 25) allows the engineer to determine the nonlinear transfer functions of an electronic circuit containing resistors, inductors, capacitors, transistors, vacuum tubes, and diodes. It uses a set of standard electric circuit elements, and can analyze networks made up of interconnections of these elements. The objective of determining these transfer functions and analyzing the networks is to provide a close prediction of the degradation caused by general classes of interfering signals.

A realistic receiver model must be nonlinear to account for such effects as desensitization, cross-modulation, intermodulation distortion, and spurious responses. Models that have been developed are nonlinear canonic models of portions of communication receivers that lead to a representation of the given receiver in terms of building blocks of known form, multipliers, and receiver-dependent parameters. The number of parameters required generally will be much smaller than the number of devices and components in the receiver. As a result, the canonic model approach leads to a practical means of analyzing and simulating the input-output behavior of a receiver as a nonlinear signal-processing black-box.

6-4.11.2 Data Inputs

Five types of inputs are needed to describe a given circuit. The input statements define the topology of the circuit, the circuit element values, the linear and nonlinear devices used in the circuit, the circuit excitation and the order of the analysis, and the desired output.

6-4.11.3 Analysis Process

The network is divided into segments which are automatically cascaded. The cascade equations include linear loading effects. Segmentation is done such that nonlinear interaction is small. There is no limit to the size of a network that can be analyzed.

Structurally, the nonlinear network problem is solved by forming both the nodal admittance matrix (Y matrix) for the entire network, and the first-order generator (current source) excitation vector, for all of the linear sources in the entire network. The generators can be located at any node in the network, and can have any desired frequency, amplitude, and phase. The usual procedure of premultiplying the generator vector by the inverse Y matrix results in the first-order nodal voltage vector for the network, the elements of which are the first-order transfer functions at all nodes in the network at the given excitation frequency. In the event that there is more than one generator at a given frequency, the first-order transfer function is the total transfer function due to the superposition of the generators, since the first-order transfer function is a linear function. The higher order functions are solved iteratively.

6-4.11.4 Data Outputs

The software available can analyze up to 50 node networks directly. The output of the program is the nonlinear transfer function of the desired order at all the nodes of the network.

6-4.11.5 Availability

The program is written in FORTRAN IV, and has been implemented on two computers, the IBM 1130 with 8k of core and the Honeywell 635. The IBM 1130 version can directly analyze networks containing up to 50 nodes. Larger networks can be analyzed by segmenting the network into subnetworks, and then employing a cascading program.

Copies of the program on punched cards can be obtained by contacting Code RBCT, Rome Air Development Center, Griffiss Air Force Base, NY.

UNRESOLVED INTERFERENCE

EMTR — SUBS = CNT EQPT 1 = UHFCO PORT 2 = COMLO

RCPT — SUBS = GHI EQPT 6 = IFF PORT 2 = IFFRF

PATH = ANT TO ANT

NOTE - R = IN REQD RANGE, I = INTERPOLATED VALUE

FREQUENCY (HERTZ)	FREQ BASE	TRANSFER RATIO	RECEPTOR SPCT LEV	RCPT LEV TO LIMIT	NARROWBAND EMI MARGIN	EMITTER SPEC LEV	EMTR LEV TO LIMIT	RCVD SIGNAL	BROADBAND EMI MARGIN	EMITTER SPEC LEV	EMTR LEV TO LIMIT	RCVD SIGNAL	BDWTH FCTR
1.53666E+04	ENTR	-8.6	103.0 I		-69.6	43.0	30.0	35.0					
1.66071E+04	RCPT	-8.6	103.0	30.0	-68.8	43.0 I		35.0					
1.87223E+04	ENTR	-8.1	103.0 I		-65.1	43.0	30.0	34.9					
3.12328E+04	RCPT	-8.1	103.0	30.0	-68.1	43.0 I		34.9					
5.87392E+04	RCPT	-8.2	103.0	30.0	-68.2	43.0 I		34.8					
6.14400E+04	ENTR	-8.2	103.0 I		-69.2	43.0	30.0	34.8					
1.18478E+05	RCPT	-8.4	103.0	30.0	-68.4	43.0 I		34.6					
1.22893E+05	ENTR	-8.4	103.3 I		-68.6	43.0	30.0	34.6					
2.03763E+05	RCPT	-8.7	103.0	30.0	-68.7	43.0 I		34.3					
2.45768E+05	ENTR	-8.7	103.1 I		-68.8	43.0	30.0	34.3					
3.90733E+05	RCPT	-9.6	103.2	29.0	-79.6	32.7 I		23.7					
4.91528E+05	ENTR	-9.2	114.1 I		-95.7	27.6	14.5	18.4					
7.34847E+05	RCPT	-9.5	133.0	0.0	-123.4	19.1 I		9.6					
9.83043E+05	ENTR	-9.8	133.0 I		-129.8	13.0	0.0	3.2					
1.38332E+06	RCPT	-10.2	133.0	0.0	-130.2	13.0 I		2.9					
1.56604E+06	ENTR	-10.6	133.0 I		-130.6	13.0	0.0	2.4					
2.59915E+06	RCPT	-11.0	133.0	0.0	-131.0	13.0 I		2.0					
3.93216E+06	ENTR	-11.8	133.0 I		-131.8	13.0	0.0	1.2					
4.88828E+06	RCPT	-12.3	133.3	0.0	-132.3	13.3 I		.8					
7.86432E+06	ENTR	-15.1	110.4 I		-112.5	13.0	0.0	-2.6					
9.18318E+06	RCPT	-16.9	103.0	30.0	-100.1	19.9 I		2.9					
1.57286E+07	ENTR	-23.3	103.0 I		-83.3	43.0	30.0	19.7					
1.72885E+07	RCPT	-24.5	103.0	30.0	-84.5	43.0 I		18.5					
3.14573E+07	ENTR	-32.5	103.0 I		-92.5	43.0	30.0	18.5					
3.25162E+07	RCPT	-32.9	103.0	30.0	-92.9	43.0 I		16.1					
6.11533E+07	RCPT	-42.3	103.0	30.0	-102.3	43.0 I		.7					
6.29146E+07	ENTR	-42.7	103.0 I		-102.7	43.0	30.0	.3					
1.15818E+08	RCPT	-52.8	103.0	30.0	-112.8	43.0 I		-9.8					
1.25829E+08	ENTR	-54.4	103.0 I		-114.4	43.0	30.0	-11.4					
2.16298E+08	RCPT	-64.6	103.0	30.0	-82.1	105.5 IR		40.9	-292.9	-112.2 IR		-109.9	-13.0
2.51654E+08	ENTR	-67.7	103.0 I		-47.7	123.0 R	30.0	55.3	-47.9	137.9 R	30.0	55.2	-13.6
4.95789E+08	RCPT	-75.1	103.0	30.0	-58.1	123.0 IR		44.9	-54.3	135.9 IR		44.8	-13.0
5.03316E+08	ENTR	-83.0	103.0 I		-63.6	123.0 R	30.0	40.6	-63.2	135.9 R	30.0	39.8	-13.0
7.65044E+08	RCPT	-94.0	103.0	30.0	-184.2	92.8 I		-1.2	-78.5	105.7 I		24.5	12.8
1.00663E+09	ENTR	-101.9	29.2 IR		-58.0	73.0	30.0	-28.8	-35.7	85.9	30.0	-7.5	8.5
1.03860E+09	RCPT	-102.5	23.0 R	30.3	-53.5	72.0 I		-30.5	-67.1	40.9 I		-44.1	8.5
1.09000E+09	RCPT	-104.2	103.0	30.0	-137.7	69.6 I		-34.6	-235.6	-34.8 I		-132.6	10.4
1.43081E+09	RCPT	-112.9	103.0	30.0	-195.3	57.6 I		-55.3	-679.0	-473.7 I		-575.0	11.6
2.01327E+09	ENTR	-124.1	103.0 I		-184.1	43.0	30.0	-81.1					
2.70596E+09	RCPT	-134.5	103.0	30.3	-194.6	43.0 I		-91.6					
4.02653E+09	ENTR	-149.7	103.0 I		-209.7	43.0	30.0	-106.7					
5.08907E+09	RCPT	-159.1	103.0	30.0	-219.1	43.0 I		-116.1					

Figure 6-31. Sample Output — SGR Unresolved Interference

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CHAPTER 7
MEASUREMENTS

7-0 LIST OF SYMBOLS

A = area of loop antenna, m^2
A_f = antenna factor, dB
a = loop diameter, m; interior dimension of an enclosure, m; envelope amplitude, V
a_p = relative peak response with respect to peak obtained with a slow sweep, dimensionless
B = magnetic flux density, T
B_e = effective noise power bandwidth, Hz
B_i = effective impulse bandwidth, Hz
B_3 = 3-dB bandwidth, Hz
b = interior dimension of an enclosure, m
C = capacitance, F
C_a = capacitance at the terminals of an antenna, F
c = interior dimension of an enclosure, m; speed of light, 3×10^8 m/sec
D = contact separation distance, m; diameter of larger antenna, m
d = distance, m; contact diameter, m; pulse duration, s; distance between loops, m; diameter of smaller antenna, m
E = electric field strength, V/m; peak pulse voltage, V
E_o = free space electric field strength, V/m
e = charge of an electron, C
F = sweep width (spectrum analyzer), Hz
f = frequency, Hz
f_o = tuned frequency, Hz
f_p = pulse recurrence frequency, Hz
G = antenna gain, dimensionless
G_r = receiving antenna gain, dB
G_t = transmitting antenna gain, dB
H = magnetic field strength, A/m
$H(f_o)$ = voltage gain transfer function, dimensionless; center frequency gain, dimensionless
h_1 = height of antenna number 1, m
h_2 = height of antenna number 2, m
I = current, A
I_1 = primary current, A
I_{dc} = direct current, A
i_n = noise current, A
L = inductance, H
L_a = self-inductance (loop antenna), H; primary leakage inductance, H
L_b = secondary leakage inductance, H
L_m = magnetizing inductance, H
l = length (antenna), m
l_e = effective length or effective height (antenna), m
m = integer
N = number of turns; noise level (used in signal plus noise-to-noise ratio), W; noise power spectral density at frequency of measurement, V^2/Hz
$N(f_o)$ = Gaussian random noise spectral density, V^2/Hz
$N_i(f)$ = spectral power density of noise current, A^2/Hz
N_1 = number of turns of transmitting loop, dimensionless
n = integer
P_r = power received, W
P_t = power transmitted, W
p = integer
$p(v_s)$ = amplitude probability of v_s
$p_a(a)$ = probability density of the output a of envelope detector, V
R = resistance, Ω; resolution, Hz; rms fluctuation of a mean square relative to a true mean square value, dimensionless
R_a = radiation resistance, Ω; primary resistive loss, Ω
R_b = secondary resistive loss, Ω
R_c = detector charge resistance, Ω
R_d = detector discharge resistance, Ω
R_H = Hall coefficient, m^3/C
R_m = core loss resistance, Ω
R_o = antenna loss resistance, Ω; load resistance, Ω; characteristic impedance of a transmission line, Ω
R_s = source resistance, Ω
r = separation distance, m
r_1 = radius of transmitting loop, m; distance between antennas, m
r_2 = radius of receiving loop, m; distance between receiving antenna and source antenna image, m
S = signal level (used in signal plus noise to noise ratio), W; scanning rate, Hz/s; impulse strength, V·s
$S(f)$ = spectrum amplitude, V·s

$S(f_o)$ = spectrum amplitude of the tuned frequency, V·s
$s_o(t)$ = RF-IF amplifier response to an impulse, V
T = averaging time, s; time to sweep through F (spectrum analyzer), s
t = thickness, m; time, s
V = voltage, V
$V(f)$ = Fourier spectrum, V/Hz
V_a = antenna coupled voltage, V
V_b = direct coupled voltage, V
V_e = root mean square value of random noise voltage, V
V_H = Hall effect voltage, V
V_o = output voltage, V
V_c = charging voltage, V
v_i = induced voltage, V
$v_s(t)$ = source voltage (function of time), V
Z_a = antenna impedance, Ω; loop impedance (loop antenna), Ω
Z_{in} = insertion impedance, Ω
Z_t = transfer impedance, Ω
γ = reflection coefficient, dimensionless
ϵ_0 = permittivity of free space, F/m
λ = wavelength, m
μ_0 = permeability of free space, H/m
ρ = resistivity, Ω·m
σ = impulse strength, V·s
ω = radian frequency, rad/s

7-1 INTRODUCTION

Measurements are made for three distinguishable purposes:

a. Determination of characteristics for use in EMC analysis and prediction (programs)

b. Determination of compliance with specifications

c. Determination of environmental characteristics (site surveys).

The first two purposes may be accomplished with similar facilities. The last item may require separate facilities. The following principles must be borne in mind in planning and implementing any tests:

a. The test must meet the program requirements in types of data and accuracy.

b. A given test setup must be specific to the data required.

c. The test must be cost effective.

7-2 TEST REQUIREMENTS

EMC test requirements, test plans, and test reports generated as part of the EMC Program (par. 2-6) prescribe the types of measurement needed.

Fig. 7-1 diagramatically shows applicable requirements evolving into program test plans, which are then implemented and intended to verify that the final design and/or hardware meet the specification and contractual requirements.

7-2.1 DEVELOPMENT TESTING

During the course of equipment design and development, or prior to equipment redesign, it is often necessary to conduct tests to obtain data on:

a. Equipment emission and/or susceptibility characteristics

b. Coupling properties of design configurations

c. Transfer functions of networks, filter, or shielding effectiveness of equipment enclosures

d. The electromagnetic environment of the intended equipment, subsystem, or system installation site.

7-2.2 VALIDATION TESTING

Validation or quality assurance testing is performed to ascertain that the design and hardware produced from that design meet the specified requirements. Equipment, subsystem/system specifications and the requirements such as MIL-STD-461 define acceptable EMC characteristics. Corresponding tests are described in MIL-STD-462 and the contractor prepared, Government approved, test plan.

Also included under validation testing are field tests, on-site or installation tests. Because the environment cannot be controlled, facilities are limited, or test time may be limited, plans for such tests should be carefully drawn up in advance to simplify test procedures as much as possible. To be considered are:

a. Test facility required instrumentation — type and characteristics

b. Scanning and identifying the spectrum for unacceptable environmental levels of EMI, both radiated and conducted

c. Checking of electronic systems for responses to undesired signals

d. Identification of internally generated signals which may cause EMI because of inadequate shielding, grounding, and bonding at the installation.

Test setups should be engineered as thoroughly as possible during the test planning stage (par. 2-6.3). All the requirements of the applicable specifications should then be considered in detail, and special interpretations may be sought when required.

Measurements of radiated emission and susceptibility share many common problems. The wide frequency range involved in EMI/EMC testing requires a variety of test equipment, facilities, and techniques to perform the measurements in accordance with

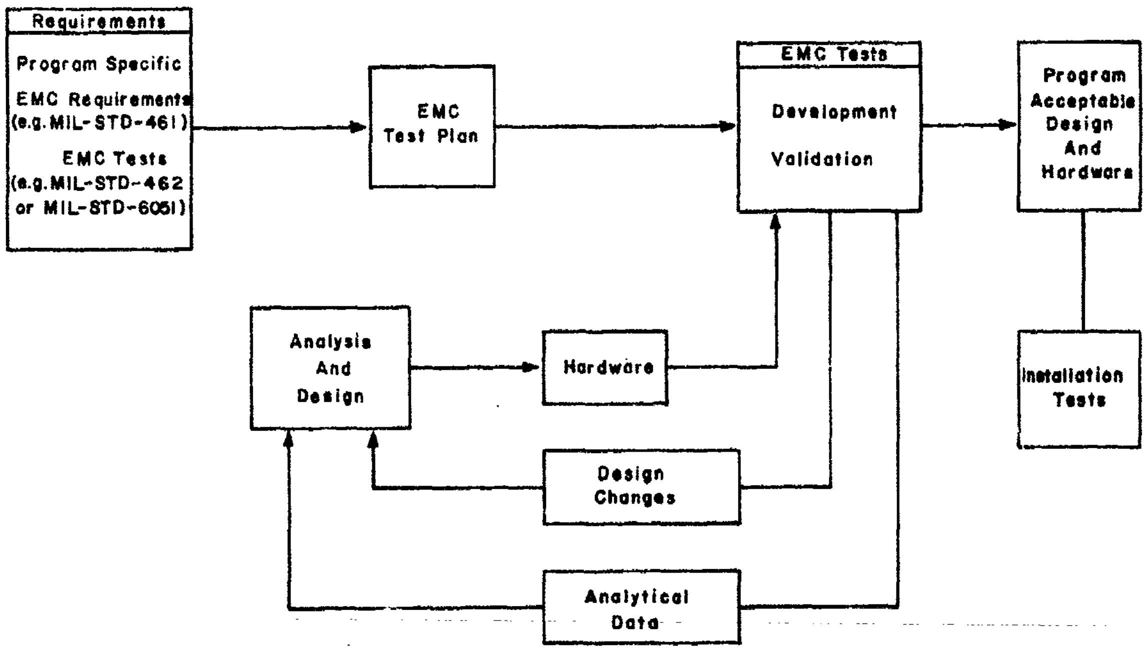


Figure 7-1. EMC Program Test Plan

prescribed requirements depicted in military standards and specifications.

7-3 INSTRUMENTATION

Instruments for test and measurement fall into two categories, namely: (1) standard instruments for testing according to prescribed methods, and (2) specialized instruments for laboratory analysis. In Category No. 1 are the measuring receivers, sensors, and sources invoked by such standards as MIL-STD-462; in Category No. 2 are such devices as auto- and cross-correlators, noise pulse counters, and probability density analyzers. The distinction is blurred somewhat in that standard EMI instruments are useful in laboratory analysis, and special purpose devices are sometimes written into compatibility test plans. The paragraphs that follow are devoted to those specific instruments and devices most often used in specification testing including field intensity receivers and automatic testing equipment.

7-3.1 MEASURING SYSTEM

A typical EMI measuring system, Fig. 7-2, is comprised of a calibrated measuring receiver and associated pickup sensors (Ref. 1). The receiver differs from conventional communication receivers in that it contains a variety of matching devices for coupling different sensors, attenuators, internal devices for calibration, detectors with special properties, and output displays. The RF amplifier, mixer, local oscillator, and some IF amplifiers are similar to circuits found in communications receivers, but differences may exist in the bandwidths used and in the amplitude range which the circuits are designed to accommodate. The tuning range of available equipment extends from about 30 Hz to more than 40 GHz. To some extent these instruments are standardized both here and abroad though the standards differ. Instruments covering 0.015 to 1000 MHz are standardized by the American National Standards Institute (Refs. 2 and 3) and the International Electrotechnical Commission (Refs. 4, 5, and 6). Suggested instruments for use in testing for military procurements are listed in MIL-STD-462.

The antennas used at low frequencies are either untuned loops or rods; at higher frequencies broadband antenna types, e.g., the biconical antennas and double-ridged horns, are used. Means are available for measuring noise voltages conducted on cables and power transmission lines. Each of these input devices requires its own coupling circuit into the noise meter and some instruments have these built in.

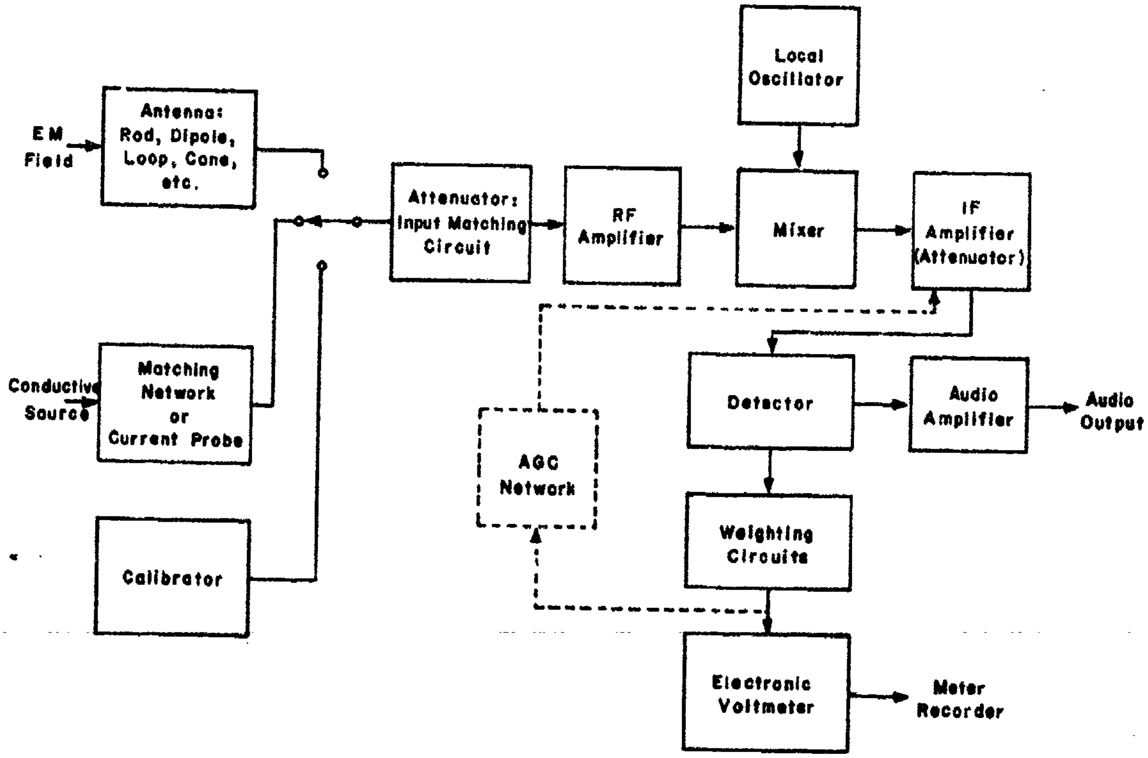


Figure 7-2. Block Diagram of an EMI Measuring System

The calibrator is a built-in source of either CW, pulsed CW, random noise, or very short pulses (impulses). The latter two, being broadband, provide energy for calibration over the frequency range of the instrument without the need for tuning. The latest instruments favor the impulse source. CW and pulsed sources require tuning if they are to provide calibration at any frequency. In some instances fixed frequency sources are used. In the microwave range (above 10 GHz) CW or pulsed CW calibrators are used.

To permit measurements over a wide range of input levels most instruments are equipped with step attenuators allowing three or four decades of attenuation. The first decade is usually found in the IF amplifier and the remaining sections are at the input. The RF-IF amplifier is designed to allow a large input amplitude range. This usually reflects itself in the design of the final IF amplifier to allow waveforms of large peak amplitudes to pass undisturbed. In some cases the IF amplifier is made approximately logarithmic by the use of automatic gain control which derives its control voltage from the detector output and in others, a logarithmic amplifier is used. The compression afforded serves to increase the allowable range of input amplitudes over that tolerated by a linear amplifier. The response of the instrument depends greatly on the bandwidth. Some, but not all, instruments will have essentially constant bandwidth, at least over each tuning band.

7-3.1.1 Measurement Functions

The output indications of the instrument depend on the nature of the waveform being measured, the RF-IF amplifier bandwidth, and the detector function.* It has become customary to think of instrument response in terms of three kinds of standard waveforms — namely, sine waves, low repetition rate periodic short pulses, and Gaussian random noise. Actual waveforms measured may be none of these but often will be near enough to one or the other to suggest similar behavior in the measuring instruments.

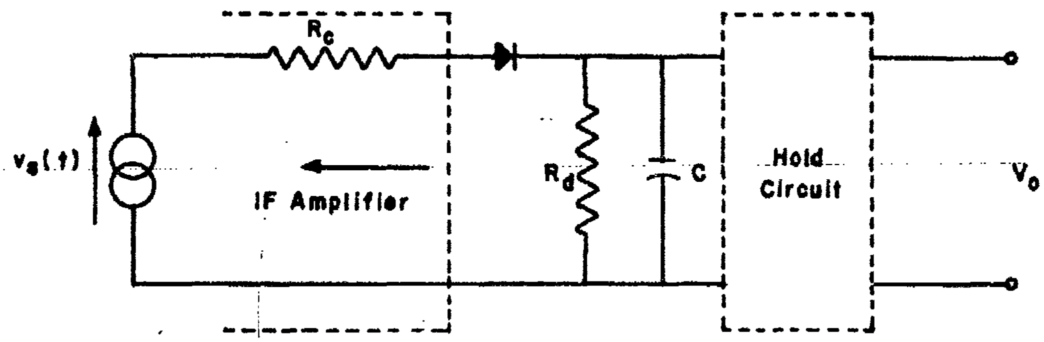
* A detailed treatment of the EMI measuring instrument, including design considerations, methods of calibration and use, and instrument performance testing will be found in Ref. 7.

7-3.1.2 Detector Functions

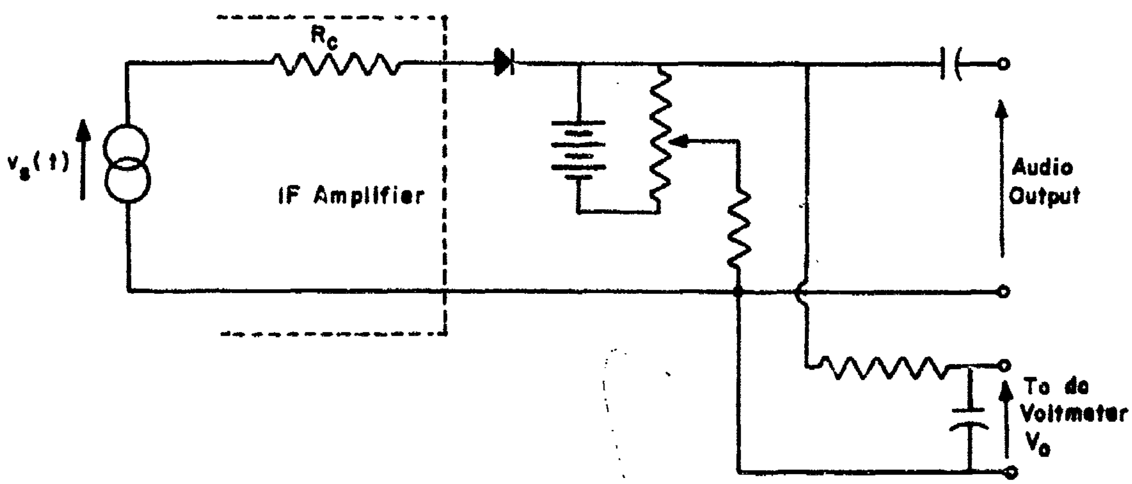
Instruments are equipped with one or more detectors which read some parameter of the applied waveform as it appears at the IF amplifier output. The detector functions include peak, quasi-peak, rms, and average of the envelope.

7-3.1.2.1 Peak Detectors

The peak detectors are either of the direct reading kind as in Fig. 7-3(A) or of the indirect (or slide-back) kind as in Fig. 7-3(B). In the direct reading detector, by making the discharge resistor as large as possible and the charging resistor as small as possible, the voltage to which the capacitor charges is very nearly the peak value of the source voltage v_s. These circuits have been made to operate very effectively for short pulse inputs so as to read the peak value of a single pulse effectively. On such instruments, to avoid rapid fluctuations in the output indicator when scanning in frequency or when observing rapidly fluctuating interference, the peak reading is held for a manually adjustable period. The output V_o is read by a high input impedance electronic voltmeter. In the alternative arrangement of Fig. 7-3(B), the voltage V_o is increased manually until the audio output, as heard in head phones or as indicated on a meter, is reduced to zero. At this point, the diode is just cut-off and V_o is equal to the positive peak of $v_s(t)$. V_o is then read on a voltmeter.



(A) Form of Direct-reading Peak Detector Circuit



(B) Form of Slide-back Peak Detector Circuit

Figure 7-3. Detector Circuits

The peak value is particularly useful when inputs are of an impulsive nature. Impulsive interference is typically of low repetition frequency compared with the bandwidths employed in the EMI instruments. When a short pulse is applied to the meter, the narrow-band IF filter develops an output burst whose width is of the order of the inverse of the IF amplifier bandwidth. If the inverse of the repetition rate is much less than the resulting output pulse width, then the impulses are, for practical purposes, isolated, nonoverlapping, transients. Under these conditions the reading obtained in the peak measuring circuits of Fig. 7-3 is given by (see Eq. 3-60)

$$V_o = S(f)\left|H(f_o)\right|B_i \text{ , V} \qquad (7\text{-}1)$$

where

$S(f_o)$ = spectrum amplitude of the pulse at the tuned frequency, V·s

$H(f_o)$ = gain of the instrument at its tuned frequency from the input to the detector output*, dimensionless

B_i = effective impulse bandwidth, Hz

7-3.1.2.2 Average Detector

The average detector, sometimes called the field intensity detector, is an envelope detector followed by an averaging circuit comprised of a series resistance and shunt capacitance. Time constants are chosen to follow the envelope variations allowed by the bandwidth of the IF amplifier. (The principle of envelope detection is treated in Ref. 8.)

When a pure sinusoid (or a frequency-modulated sinusoid) is applied to a detector of this nature the output is, in principle, the peak value of the sinusoid. An amplitude-modulated wave results in an output which is equal to the peak of the carrier alone. It is for this reason that the term "field intensity"** is used to identify this function; when measuring the field strengths of signals, readings independent of the modulation are obtained.

When an impulse is applied to the EMI instrument the area of the IF amplifier output envelope is given by

$$\text{Area} = S(f_o)H(f_o) \text{ , V·s} \qquad (7\text{-}2)$$

where

$S(f_o)$ = spectrum amplitude of the impulse at the tuned frequency f_o

$H(f_o)$ = center frequency gain.

If a sequence of impulses of repetition frequency f_p is applied, the average value of the output is

$$V_o = S(f_o)H(f_o)f_p \text{ , V} \qquad (7\text{-}3)$$

The response of the average detector to Gaussian random noise is determined by recognizing that the output a of the envelope detector preceding the averaging circuit is described by the Rayleigh distribution

$$\left.\begin{aligned} p_a(a) &= \left(\frac{a}{V_e^2}\right)\exp\left[-a^2/(2V_e^2)\right] \text{ , V} \qquad && a \geq 0 \\ &= 0, && a < 0 \end{aligned}\right\} \qquad (7\text{-}4)$$

where

a = envelope amplitude, V

V_e^2 = mean square value of the Gaussian noise entering the envelope detector, V^2

The averaging circuit develops an output V_o given by the mean value of a which is found to be

$$V_o = \int_0^\infty a p_a(a)da = V_e\sqrt{\frac{\pi}{2}} \text{ , V} \qquad (7\text{-}5)$$

In Chapter 3 (Eq. 3-57) the mean square value of the output $\langle v_o^2 \rangle$ of a filter to Gaussian noise is given by

$$\langle v_o^2 \rangle = V_e^2 = N(f_o)\left|H(f_o)\right|^2 B_e \qquad (7\text{-}6)$$

where

$N(f_o)$ = Guassian random noise spectral density, V^2/Hz

B_e = effective noise bandwidth, Hz

so that the average meter reading is

$$V_o = \sqrt{\frac{\pi}{2}N(f_o)H^2(f_o)B_e} \text{ , V} \qquad (7\text{-}7)$$

7-3.1.2.3 RMS Detectors

For root mean square detectors if the input is random noise, the reading is given directly by Eq. 7-5. For nonoverlapping repetitive impulses, the energy of a single pulse at the detector input is

$$\int_{-\infty}^{\infty} s_i^2(t)\, dt \text{ , } V^2\text{·s} \qquad (7\text{-}8)$$

* Efficiency of the detector is included in $H(f_o)$ as here defined. It would be determined by appling a sinusoidal input at frequency f_o and measuring the dc output voltage at the detector. $H(f_o)$ is then the ratio of output voltage to peak input voltage.

** The use of the term *field intensity* (FI) has been deprecated by the IEEE in favor of *field strength*. However, the term *intensity* is still used in connection with radio interference instrumentation.

where

$s_o(t)$ = RF-IF amplifier response to the impulse, V. By the Fourier Integral Energy Theorem (or Parseval's theorem, Ref. 9, Eq. 2-75) this is also

$$\int_{-\infty}^{\infty} s_o^2(t)\,dt = \int_{-\infty}^{\infty} V^2(f)H^2(f)\,df$$

$$= \frac{1}{2}S^2(f_o)H^2(f_o)B_e\,,\ \mathrm{V^2 \cdot s} \quad (7\text{-}9)$$

where

$V(f)$ = Fourier transform of $s_o(t)$, V/Hz

The average power at the detector input is given by the energy per pulse multiplied by the pulse repetition rate. Thus, the mean square output is

$$V_o^2 = \frac{1}{2}S^2(f_o)H^2(f_o)B_e f_p,\ \mathrm{V^2} \quad (7\text{-}10)$$

An rms detector reads V_o, the square root of this quantity.

Interference meters are calibrated to read the rms value of an input sine wave regardless of which detector is being used. That is, an appropriate factor automatically is inserted on each detector function to account for differences in output voltage when a sine wave is applied. Under these conditions the output readings are as summarized in Table 7-1.

TABLE 7-1. READINGS OF RFI METER CALIBRATED TO READ THE RMS VALUE OF AN INPUT SINE WAVE

Input	Detector Function		
	Peak	Average of Envelope	RMS
Sine wave; rms value = E	E	E	E
Impulses; Impulse Strength = σ	$\sqrt{2}\,\sigma B_e$	$\sqrt{2}\,\sigma f_p$	$\sigma\sqrt{2B_e f_p}$
Random noise power spectral density = N	—	$1.25\sqrt{NB_e}$	$\sqrt{2NB_e}$

B_i = effective impulse bandwidth, Hz
B_e = effective noise power bandwidth, Hz
f_p = pulse repetition rate, pps (assumed low enough to prevent overlapping)
N = noise power spectral density at frequency of measurement, $\mathrm{V^2 \cdot s}$
σ = impulse strength = area under impulse, V·s, Eq. 7-11.

7-3.1.3 Calibration and Methods of Use

Two methods of calibration are the direct reading method and the substitution method. In the direct reading method the instrument scale usually is calibrated in terms of the rms value of a sine wave. In the substitution method, the instrument is used as an indicator. A reading is obtained with the waveform to be measured, and a calibrated standard waveform source is then connected to the instrument input and its magnitude adjusted to duplicate the earlier reading. The source calibration provides the measured value.

Virtually all EMI meters, whether intended for direct or substitution measurements, have built-in calibration sources. For the direct reading instrument they may be viewed as aids for setting the gain of the instrument; for substitution measurements they provide the basis for the output readings. The waveforms used are impulses, sine waves, or Gaussian random noise, with impulses preferred since they can be constructed to give useful output to frequencies beyond 10 GHz.

One method of generating impulses is to discharge periodically a transmission line (Fig. 7-4) whose length determines the width of the pulse. Pulse widths of the order of 10^{-9} are in common use, with corresponding frequency spectra which exceed 1000 MHz. Under ideal conditions, with the line impedance R_o perfectly matched to the discharging load R and with instantaneous closing of the switch contacts, the output is a rectangular pulse whose height is 1/2 the voltage to which the line was charged by the dc source through the resistance R_s, and whose width is the time required [illegible] a disturbance to travel twice the length of the line [illegible] repetition rate, in some instruments, is main[illegible] constant at about 60 Hz; in others it is variable [illegible] several Hz to several thousand Hz.

The spectrum amplitude $S(f)$ of the pulse is (par. 3-1.3.2.1 and Fig. 3-20)

$$S(f) = 2V_c d\left(\frac{\sin\pi fd}{\pi fd}\right) = 2\sigma\left(\frac{\sin\pi fd}{\pi fd}\right),\ \mathrm{V \cdot s} \quad (7\text{-}11)$$

where

V_c = line charging voltage, V
d = pulse duration, s
σ = impulse strength, V·s

Impulse calibrators are used at low frequencies where the approximation $\sin(\pi fd)/(\pi fd) \approx 1$, and the spectrum is independent of frequency.

It has been found useful to mismatch deliberately the pulse forming line and the load into which it discharges. Then (Ref. 10) the spectrum is

$$S(f) = V_c d \left(\frac{\sin \pi fd}{\pi fd}\right) \frac{1-\gamma}{\sqrt{1 - 2\gamma \cos 2\pi fd + \gamma^2}}, \quad \text{V·s} \qquad (7\text{-}12)$$

where $\gamma = (R - R_0)/(R + R_0)$ is the reflection coefficient. By a proper choice of γ, the spectrum can be made flatter in the region just below $f = 1/d$ than the spectrum obtained with a matched line. Other methods of generating impulse like spectra are available (Refs. 11 and 12) including one using a pulsed sine wave.

Sources of random noise are generally temperature limited diodes connected to the circuit shown schematically in Fig. 7-5. The noise current i_n in the output load resistor has a power spectral density $N_i(f)$ given by Eq. 3-7

$$N_i(f) = 2eI_{dc}, \text{ A}^2/\text{Hz} \qquad (7\text{-}13)$$

where

e = charge of the electron = 1.602×10^{-19} C

I_{dc} = direct current flowing between anode and cathode, A

Random noise calibrators using gas-filled devices (thyratrons, neon bulbs, fluorescent tubes) also are used occasionally. Diodes have been used as internal calibrators in EMI meters only up to about 30 MHz. However, calibrated noise generators as separate devices are available for use up to the UHF region.

7-3.1.4 Summary

Table 7-2 summarizes the sensitivities of available EMI measuring instruments. The frequency ranges shown are representative of divisions used but most

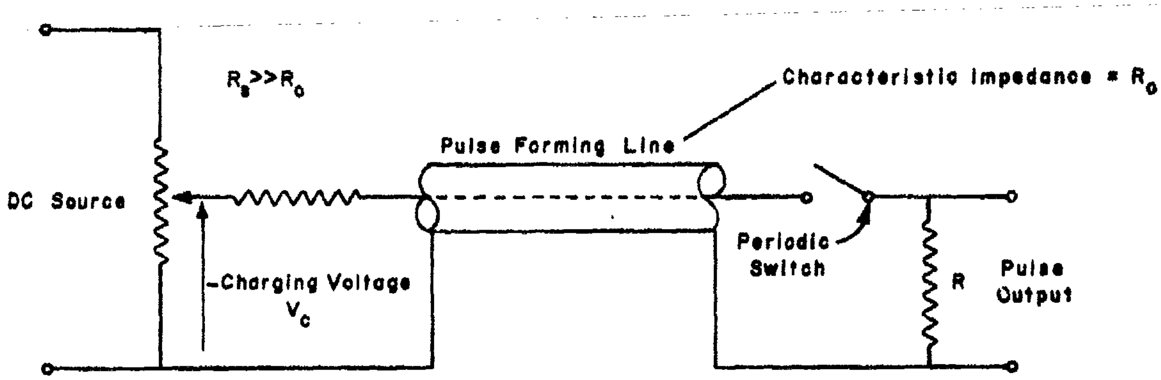


Figure 7-4. Simplified Schematic of an Impulse Generator

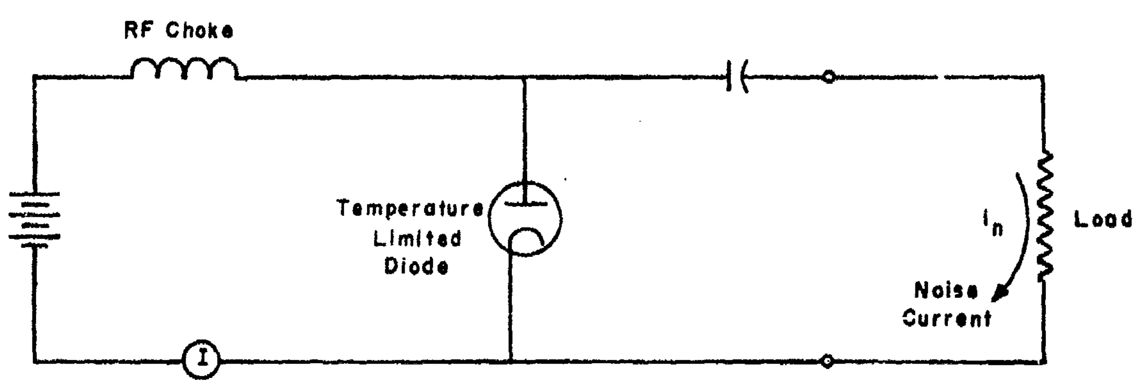


Figure 7-5. Simplified Schematic of a Random Noise Source

TABLE 7-2. TYPICAL INSTRUMENT SENSITIVITIES

Frequency Range	Bandwidth	Sensitivity*							
		Narrowband				Broadband			
		Conducted (50Ω)		Radiated		Conducted		Radiated	
		μV	dB(μV)	μV/m	dB(μV/m)	μV/MHz	dB(μV/MHz)	μV/(MHz·m)	dB[μV/(MHz·M)]
10 kHz-150 kHz	5 Hz 50 Hz 250 Hz 2.5 kHz	0.015	−36	5	14	56	35	10^4	80
150 kHz-30 MHz	100 Hz 10 kHz	0.05	−26	2.5	8	20	26	2000	66
30 MHz-1 GHz	10 kHz 100 kHz 1 MHz	0.20	−14	5	14	3	10	32	30
1 GHz-10 GHz	0.1 MHz 0.5 MHz 5.0 MHz	1.0	0	a	a	2	6	a	a
10 GHz-40 GHz	1.0 MHz 5.0 MHz	2.5	8	a	a	2	6	a	a

*Note: Sensitivities are typical. Narrowband is based upon the narrowest bandwidth listed in second column for each frequency range. Broadband is based on the widest bandwidth listed for each frequency range. Radiated sensitivities are based upon the poorest sensitivities (due to antenna effective height) to be expected over each frequency range. Sensitivities may be considerably better at other frequencies in the range.

a = values highly dependent on antenna gain characteristics.

actual instruments overlap these ranges. Some instruments contain a basic measuring unit and replaceable RF-IF amplifier heads. Virtually all instruments intended for EMI work contain average-of-envelope, and direct reading or slide-back peak detectors. In addition, true rms and quasi-peak detectors are available.

The narrowband sensitivity specified in Table 7-2 is the rms level of an input sinusoid at the tuned frequency, which equals the rms value of the internal receiver noise reflected to the input terminals. Correspondingly, the broadband sensitivity is the minimum detectable output from an impulse generator. The levels given are considered to be typical. Actual instruments can be expected to vary from these figures. The sensitivities for radiated measurements are based on the use of a 1-m rod below 30 MHz, and antennas described in par. 7-3.3.2.4 above that frequency. Newly designed instruments are largely transistorized and are therefore capable of battery operation. Such instruments are especially appropriate for portable field use.

7-3.2 WAVE AND SPECTRUM ANALYZERS

The designations "wave analyzer" and "spectrum analyzer" are reserved for instruments which have as their primary function the decomposition of waveforms into their spectral components and the measurement of these components. These instruments can be used to obtain useful spectral measures of nonperiodic, or random, waveforms.

7-3.2.1 Wave Analyzers

The wave analyzer is essentially a tunable narrowband filter followed by a suitable ac meter circuit as shown in Fig. 7-6. Commercially available instruments generally have an upper frequency limit in the order of 1 MHz.

The more frequently used wave analyzers are of the heterodyne type, as shown in Fig. 7-7, where a fixed tuned narrowband filter selects a spectral slice of the input after it has been frequency shifted in a mixer. Typically, an intermediate frequency of 100 kHz or higher is used. At IF amplifier frequencies crystal filters can be used to obtain the very narrow bandwidths usually desired. Instruments are found whose bandwidth can be varied in discrete steps; the wider bandwidth allows less critical tuning where high resolution is not needed. Minimum bandwidths of the order of 5 Hz and less are achieved. To help keep the measured signal in a narrowband filter even when its input frequency fluctuates, some devices use automatic frequency control.

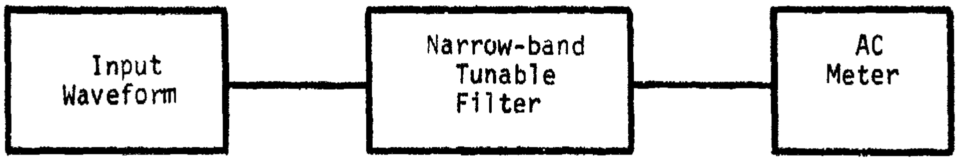


Figure 7-6. Basic Constituents of Wave Analyzer

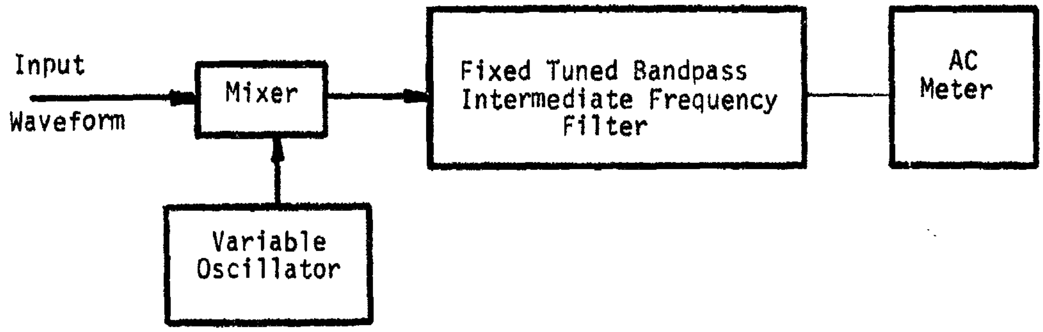


Figure 7-7. Basic Constituents of Heterodyne Wave Analyzer

Most available instruments use rectifier-filter detectors which actually measure the half or full wave average of the IF output. The meter is calibrated in terms of the root-mean-square value of a sine wave. When random noise is applied at this input, the root-mean-square value can be obtained by applying a correction factor obtained from Table 7-1. It should be recognized that the final detector output is a continuing computation of the average in a moving finite time interval. The running average so measured should not fluctuate excessively. When discrete spectrum signals are to be measured, the averaging interval should be at least ten times the period of the lowest frequency component measured for a fluctuation of about 0.5%. For continuous spectrum input, the fluctuation depends on the statistical nature of the input. For Gaussian noise, the rms fluctuation R in the mean square measurement relative to the true mean square value is given by (Ref. 1)

$$R = \left(\frac{1}{\sqrt{2TB_e}}\right)^{1/2}, \text{ dimensionless} \quad (7\text{-}14)$$

where

T = averaging time, s

B_e = effective noise power bandwidth of the filter, Hz

Values of TB_e of 70 would be needed for a relative fluctuation of less than 10%. This may be taken as a guide for other kinds of noise as well. Some commercial instruments make provision for adjusting the averaging time so as to optimize the trade-off between accuracy and observation time.

7-3.2.2 Spectrum Analyzers

The spectrum analyzer is basically a heterodyne wave analyzer that is automatically tuned and shows the spectrum on a visual display. The block diagram of Fig. 7-8 shows it essential elements. It usually is equipped with a manually variable marker oscillator which applies a known frequency to the input for calibration purposes. The mixer input has a bandwidth large enough to accommodate the range of frequencies to be displayed, while the IF amplifier bandwidth is small in order that spectrum variations with frequency will be resolved properly. Analyzers have variable sweep ranges, and variable IF amplifier bandwidths. Linear and square-law detectors sometimes are used and, occasionally, integration time is variable as well.

The resolving power of an analyzer is determined by the bandwidth of the IF amplifier. With a single sinusoid within the sweeping range of the analyzer applied to the input, the resolution R can be defined as the width of the response 3 dB below the maximum (see Fig. 7-9). For a slow sweep R is the 3-dB bandwidth of the IF amplifier. Two sinusoids of equal amplitude separated from one another by R hertz can just about be distinguished.

When the sweep rate is high the effect is to broaden the output shown in Fig. 7-9. The magnitude of this effect is developed in Ref. 13 where it is shown

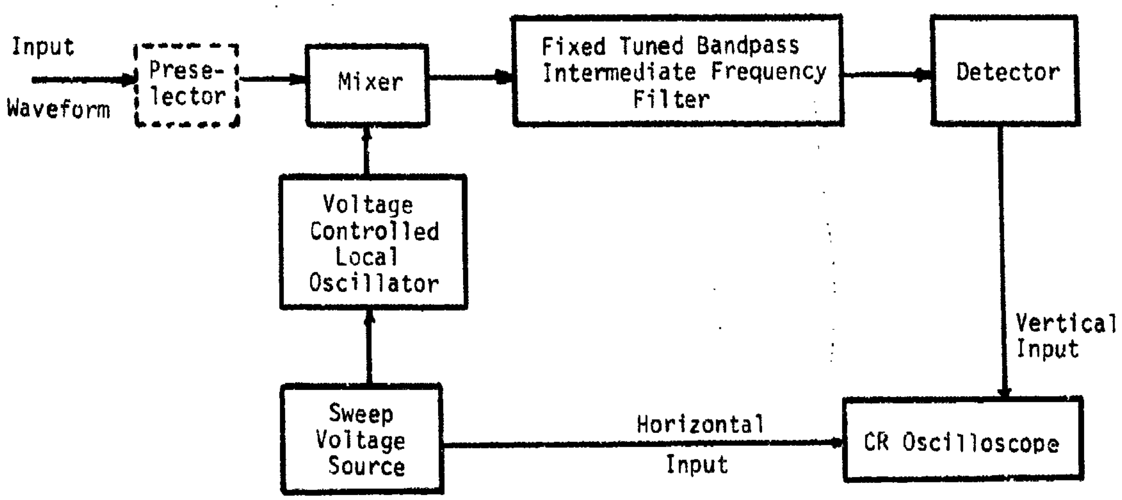


Figure 7-8. Basic Elements of Spectrum Analyzer

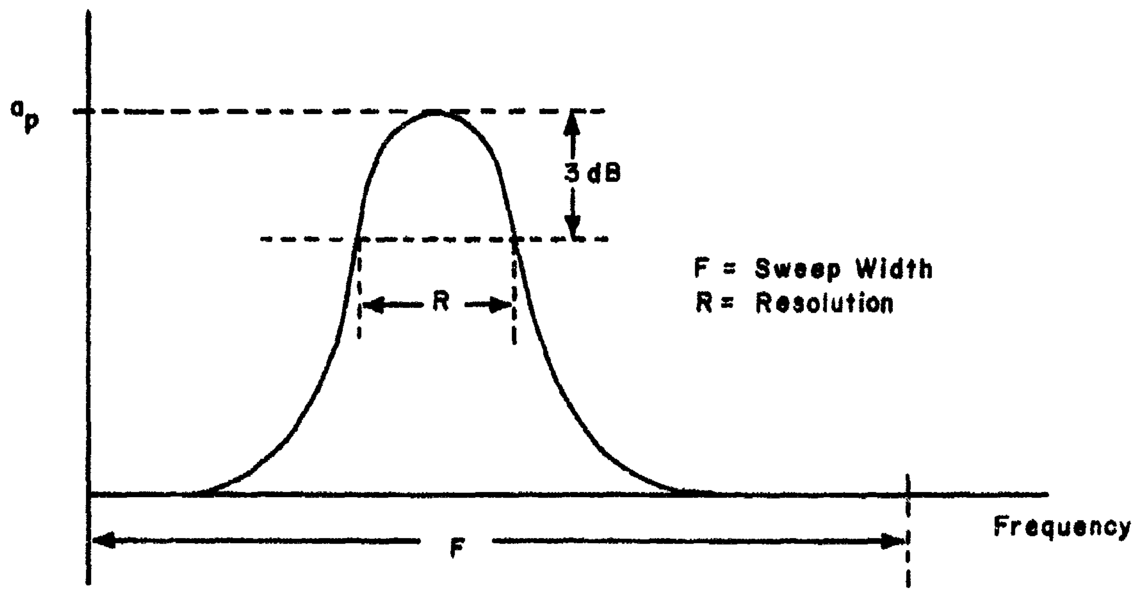


Figure 7-9. Response of a Spectrum Analyzer to a Sinusoid

that for a filter having a Gaussian shaped amplitude-frequency characteristic, and for a linear sawtooth sweep, the resolution R is given by

$$R = B_3\left[1+\left(\frac{2\ln 2}{\pi}\right)^2\left(\frac{F}{TB_3^2}\right)^2\right]^{1/2}, \text{Hz} \quad (7\text{-}15)$$

where

B_3 = 3-dB bandwidth of the IF amplifier, Hz
F = sweep width, Hz
T = time required to sweep through F, s

Furthermore, the peak of the response shown in Fig. 7-9 falls off according to

$$a_p = \left(\frac{B_3}{R}\right)^{1/2}, \text{dimensionless} \quad (7\text{-}16)$$

where a_p is the relative peak response with respect to the peak obtained with a slow sweep; it is equal to one when there is no resolution reduction. The relative resolution, R/B_3, is shown plotted in Fig. 7-10 as a function of $F/(TB_3^2)$.

Normally, the bandwidth determines the resolution. For R/B_3 close to unity, the fractional decrease in resolution defined by $(R/B_3 - 1)$, is approximately

$$\frac{R}{B_3} - 1 \approx 0.1\left(\frac{F}{TB^2}\right)^2, \text{dimensionless} \quad (7\text{-}17)$$

Assuming a fractional decrease of 10%, the sweep rate should be

$$\frac{F}{T} = B_3^2, \text{Hz}^2 \quad (7\text{-}18)$$

or less. The sweep frequency is then

$$\frac{1}{T} < \frac{B_3^2}{F}, \text{Hz} \quad (7\text{-}19)$$

For example, an instrument having an IF-bandwidth of 20 kHz and a sweepwidth of 5 MHz requires

$$\frac{1}{T} < \frac{(20\times 10^3)^2}{5\times 10^6} = 80\text{ Hz} \quad (7\text{-}20)$$

in order that the output be independent of sweep rate. On some analyzers a special light operates if sweep time is too high.

Because the emphasis on a spectrum analyzer normally is on speed of analysis, the circuit uses an envelope detector at the output of the IF. When random fluctuations are being observed, the fluctuations in output indication can be annoying. For this reason the analyzer usually contains provisions for insertion of a video filter, which permits integrating the output over a period of time, at the expense of scanning speed.

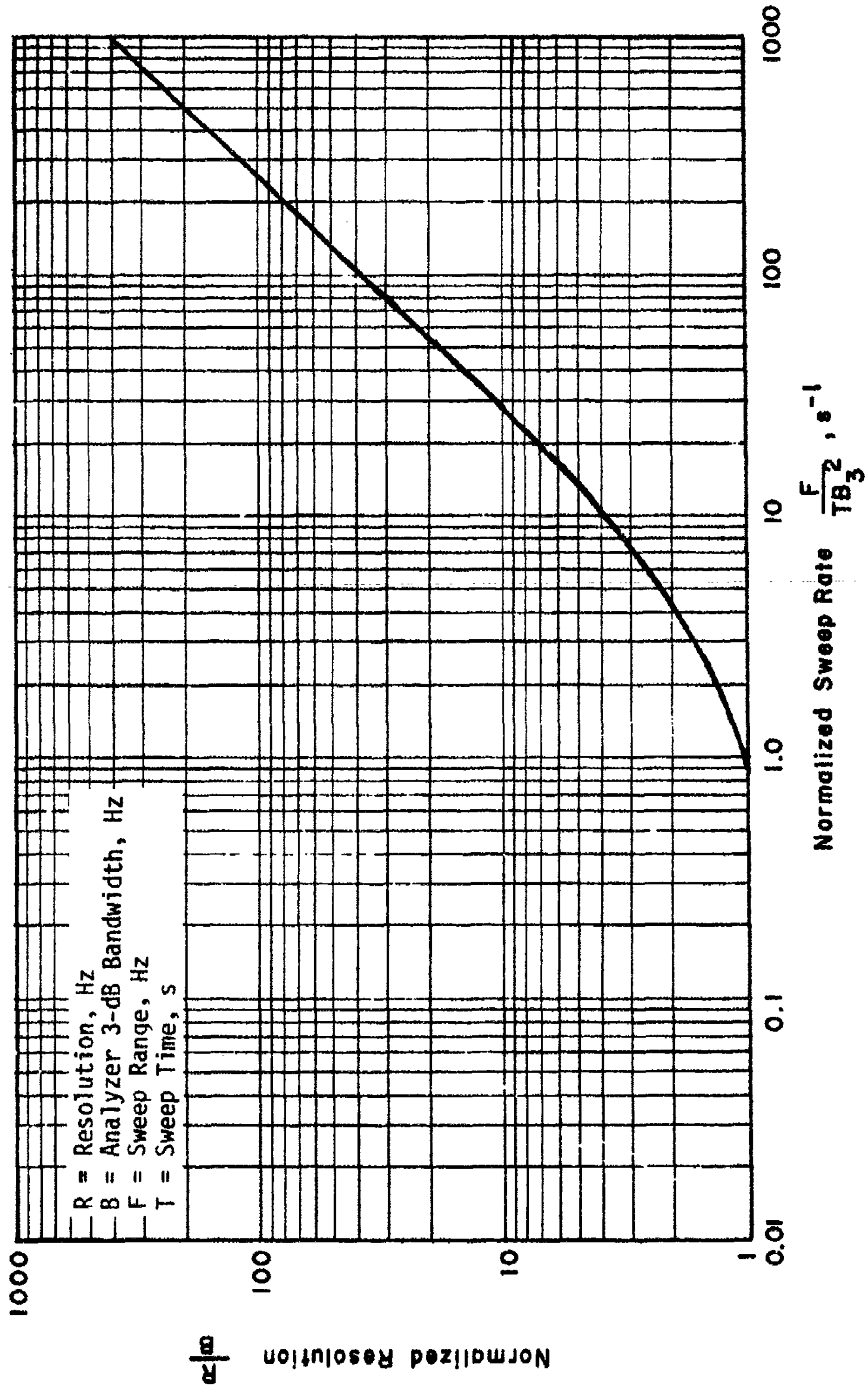


Figure 7-10. Normalized Resolution vs Sweep Rate

Courtesy of Polarad Electronics Instruments.

The integration time T required should be at least enough to give a satisfactory average as discussed under wave analyzers; i.e., $B_eT > 70$ for an output having rms deviation from the average of 10% (some commercially available equipment quote B_eT in the order of 5, but this will theoretically result in considerable fluctuation with noise inputs). It appears reasonable that the scan should sweep no more than one IF bandwidth in the integration interval T. Thus the scanning rate S in hertz per second is

$$S \leq \frac{B_e}{T}, \text{Hz/s} \tag{7-21}$$

Now, if the integration interval is chosen to be

$$T = \frac{70}{B_e}, \text{s} \tag{7-22}$$

then

$$S \leq \frac{B_e^2}{70}, \text{Hz/s} \tag{7-23}$$

For example, if B_e = 20 kHz

$$S \leq \frac{400 \times 10^6}{70} = 5.7 \times 10^6, \text{Hz/s} \tag{7-24}$$

If the sweep width is 5 MHz, the sweep duration required is $(5 \times 10^6)/(5.7 \times 10^6) = 0.88$ s.

The spectrum analyzer is subject to all the spurious response problems of receivers, namely (1) image response, (2) mixing with harmonics of the local oscillator, (3) mixing with harmonics of the input signal, (4) IF feedthrough, and (5) detection in the mixer. Because the receiver is to be swept over a wide range of frequencies, preselection is not ordinarily used. When the first IF amplifier frequency is substantially greater than the sweep range, preselection can be used to advantage. For instance, a typical first IF amplifier frequency is 160 MHz and a typical sweep for a receiver operating near 1 GHz is ± 50 MHz. The local oscillator would then sweep over the range 1.11 GHz to 1.21 GHz, so that inputs in the range 0.95 GHz to 1.05 GHz would be displayed. This assumes the local oscillator is working above the frequency of the incoming signal, but an acceptable IF signal will be developed also for inputs in the image range 1.27 GHz to 1.37 GHz. A single tuned preselector with 3-dB points at the end of the range 0.95 to 1.05 GHz will attenuate image frequencies by more than 14 dB. In general, the higher the ratio of IF amplifier frequency to sweep range, the more effective can the preselector be in eliminating most spurious responses.

Available instruments for spectral analysis cover a frequency range from about 1 Hz to 100 GHz. A typical unit covering the range from HF through SHF has a variable sweep width from several tens of kilohertz to several tens of megahertz. The resolution is variable over the range from about 1 to 100 kHz and the sweep repetition rate is variable from about 1 to 100 sweeps per second. Sensitivities at the low end of the frequency range can be expected to be about 1 μV; at the high end sensitivities of 10-50 μV generally are achieved. Instruments for the frequency range below HF and into the audio range have sweep widths from fractions of a hertz to the order of 100 kHz. Sensitivities of the order of 10 μV, and resolutions of fractions of a hertz to the order of 100 hertz are obtainable. Sweep repetition rates are of the same order as for the higher frequency instruments. Spectrum analyses are not recommended for use in performing radiated emissions or susceptibility tests.

7-3.3 ANTENNAS AND PROBES

7-3.3.1 Conductive Measurement Sensors

Measurements of emission into power conductors, and control and signal lines can be made in terms of voltage or current. Because the measured values depend on the impedance looking into the line from the equipment under test, it is essential that the impedance be standardized. This is commonly done using feedthrough capacitors, line impedance stabilization networks, or both. These devices are described in par. 7-5.2.1.1.

7-3.3.1.1 Impedance Standardization

The use of a feedthrough capacitor has the disadvantage that if the source impedance is inductive, the capacitor may resonate with it at a certain frequency. Under that condition, the current or voltage measured could be much larger than would be obtained in an actual circuit. The converse is also true, i.e., the source impedance may actually resonate with the circuit impedance producing much higher currents (voltages) than actually measured. Because of this situation, measured data taken with this technique should be examined with some care.

The line impedance stabilization network also can be subjected to the same criticism, at least at the lower frequencies where its impedance tends to be inductive. However, data taken with it are probably more representative of what will be obtained in an actual installation for the following reasons:

a. Its impedance characteristic is similar to that of a typical line (see par. 3-3.3.2).

b. The network does contain some damping which reduces resonant effects from those that could obtain without it.

Voltages can be measured directly by connecting the measuring instrument (input impedance 50 Ω) from the line to be measured and the ground plane in series with a decoupling capacitor (so that power frequency voltage is not impressed directly across the input terminals). Current is measureable with a current probe.

7-3.3.1.2 The Current Probe

The basic construction of a current probe is indicated in Fig. 7-11. The current to be measured is shown at the center of the figure surrounded by a toroid of high permeability material. The varying magnetic field in the toroid induces a voltage in the winding about the toroid which is measured with an EMI meter. Several current probes of this kind use split toroidal cores which can be opened and clamped around the wire to be measured.

The principle of operation is essentially that of a two-winding transformer in which the primary winding has a single turn.

The important properties of the current probe are (1) the ratio of output voltage to input current V_o/I, denoted transfer impedance; and (2) the insertion impedance Z_{in} which the instrument presents to the line being measured. A typical transfer impedance characteristic curve is shown in Fig. 7-12. Actual calibration curves are furnished by the manufacturer.

An equivalent circuit is shown in Fig. 7-13 (Ref. 14). L_a and L_b are primary and secondary leakage inductances, respectively; R_a and R_b are the resistive losses associated with these inductances. L_m represents magnetizing inductance, and R_m represents core losses. The turns ratio is 1:N, the primary being of 1 turn for the current probe. R_o includes the load presented by the measuring instrument.

To calculate the transfer impedance Z_t, replace the circuit on the left of Fig. 7-13, including the transformer, with a current generator of magnitude I_1/N. Then neglecting R_b as being small compared with R_o, and R_m (core losses)

$$Z_t = \frac{V_o}{I_1} = \frac{\omega L_m R_o}{N\sqrt{R_o^2 + \omega^2(L_b + L_m)^2}}, \Omega \quad (7\text{-}25)$$

where

ω = frequency, rad/s

At frequencies for which $\omega(L_b + L_m) \ll R_o$,

$$Z_t = \frac{\omega L_m}{N} R_o, \Omega \quad (7\text{-}26)$$

which is proportional to frequency. Fig. 7-12 shows this behavior below frequency $f_1 = R_o/[2\pi(L_b + L_m)]$.

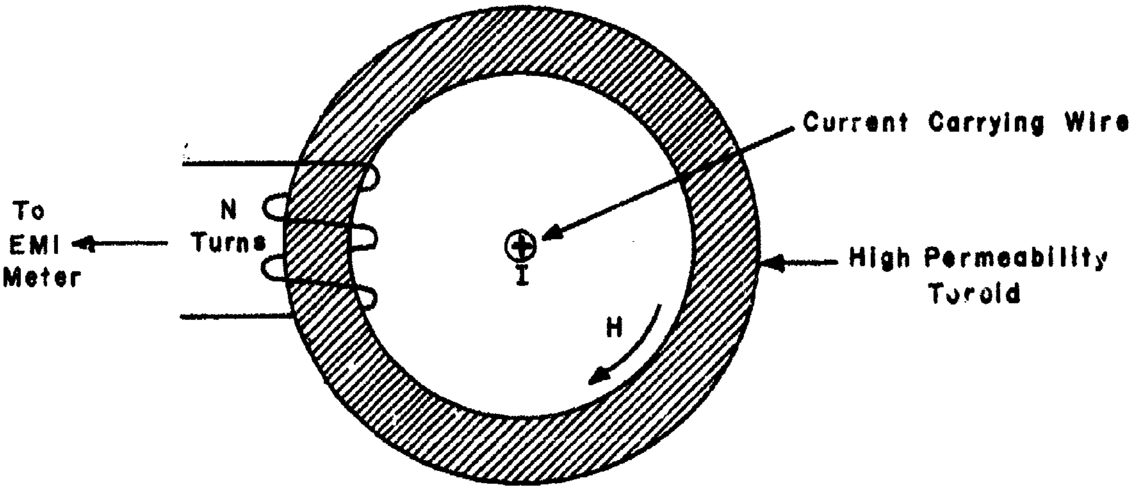


Figure 7-11. Current Probe

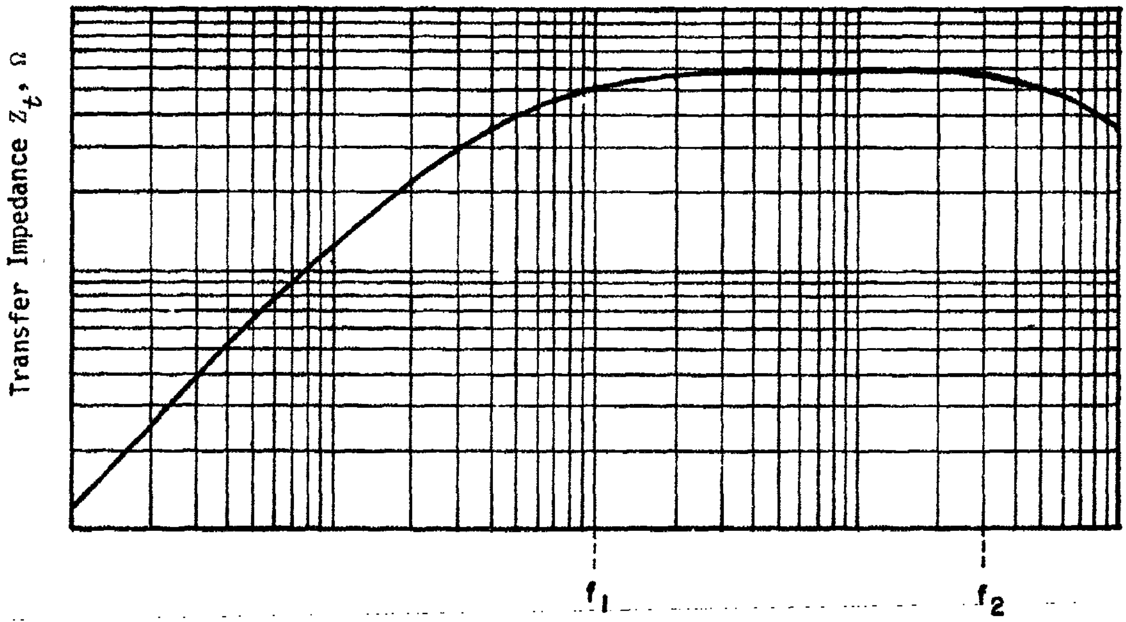


Figure 7-12. Transfer Impedance Characteristic of a Current Probe

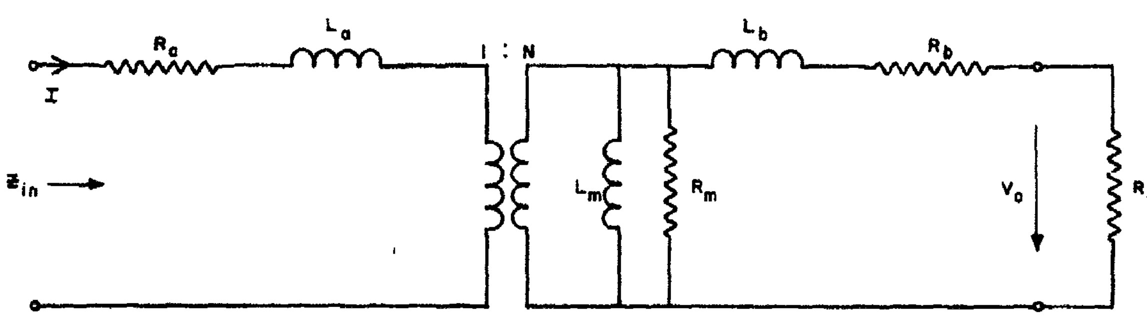


Figure 7-13. Equivalent Circuit of a Current Probe

When $R_o \ll \omega(L_b + L_m)$,

$$Z_t = \frac{R_o L_m}{N(L_b + L_m)}, \Omega \tag{7-27}$$

which is independent of frequency if $f_1 < f < f_2$, Fig. 7-12. As frequency increases gradually, the core loss term R_m becomes significant along with the distributed capacitance of windings and leads, and the transfer impedance begins to fall off ($f > f_2$, Fig. 7-12).

Due to core loss alone, at a frequency such that $\omega L_b \ll R_b$ and R_o,

$$Z_t = \frac{R_o R_m L_m}{N[R_m(L_m + L_b) + j\omega L_m L_b]}, \Omega \tag{7-28}$$

and 3-dB fall off occurs at

$$f_2 = \frac{R_m(L_m + L_b)}{2\pi L_m L_b}, \text{Hz} \tag{7-29}$$

The insertion impedance can be obtained by transforming the impedance on the secondary side of Fig. 7-13 to the primary side as shown on Fig. 7-14. At low frequencies, neglecting the shunting effect of R_o on L_m (and neglecting R_m and R_o),

$$Z_{in} = \omega\left(L_a + \frac{L_m}{N^2}\right), \Omega \tag{7-30}$$

In the mid-frequency range, R_o dominates the impedance and

$$Z_{in} = \frac{R_o}{N^2}, \Omega \tag{7-31}$$

At higher frequencies the leakage reactances L_a and L_b, core loss resistance R_m, and distributed capacitance all can affect the impedance and it can either rise or fall with frequency. Note that in the mid-frequency range it has the same general shape as the transfer impedance.

Typical values of mid-frequency transfer impedance are about 1 ohm with f_1 about 100 kHz. However, current probes are available with f_1's down to a few hundred hertz and up to about 100 MHz. Transfer impedances range up to 100 ohms. The ratio of f_2 to f_1 may be several orders of magnitude. Note that to be calibrated properly, the current probe must be operated with the proper terminating impedance, usually 50 ohms.

7-3.3.2 Antennas

Many types of antennas are used in various frequency ranges. Up to high frequencies, the rod and untuned dipole are used to measure the electric field, and the loop is used to measure the magnetic field. Dipoles, discones, and log periodic types are used at VHF and UHF; and aperture types are used at UHF and above. An important consideration in the choice of an antenna is its bandwidth, in particular, the range of frequency over which it can be used without being tuned or adjusted.

The use of active antennas should be avoided because of the possibility of spurious responses when subjected to strong fields.

7-3.3.2.1 Electric Dipole

A short (length less than $\lambda/4$) vertical rod, above a ground plane has an effective length l_e (see par. 3-3.2.4)

$$l_e = \frac{l}{2}, \text{m} \tag{7-32}$$

where l is the physical length of the antenna. The induced open circuit voltage v_i at the base of the antenna with respect to ground is

$$v_i = El_e, \text{V} \tag{7-33}$$

where

E = electric field strength, V/m.

The short rod has an impedance Z_a

$$Z_a = R_a + \frac{1}{j\omega C_a}, \Omega \tag{7-34}$$

where

R_a = radiation resistance, Ω

C_a = capacitance measured at the terminals, F

A 1-m rod (41 in. physical length), which is typical of EMI meter antennas, has a capacitance of about 10

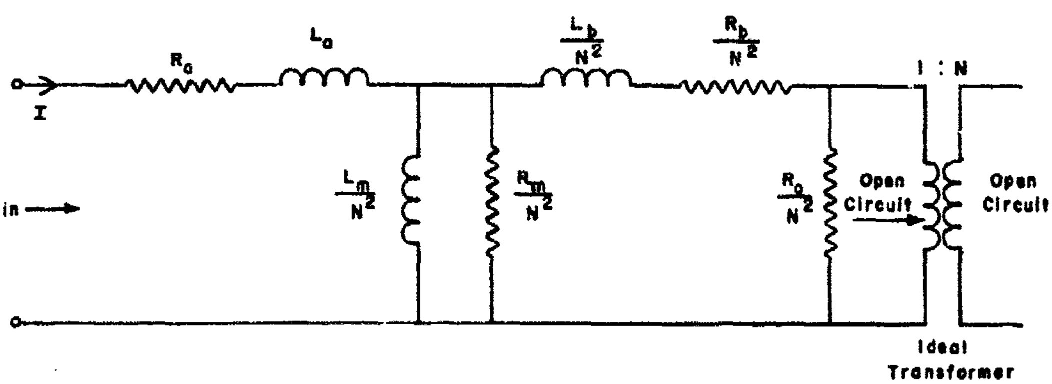


Figure 7-14. Equivalent Circuit of a Transformer Probe With Impedances Reflected to Primary Side

pF. The capacitance is absorbed in the tuning of the input circuits of the measuring instrument. The radiation resistance R_a is given by

$$R_a = 40\pi^2\left(\frac{l}{\lambda}\right)^2, \Omega \qquad (7\text{-}35)$$

λ being the wavelength of the electromagnetic field. The radiation resistance is quite small being about 6Ω for $l/\lambda = 1/8$. A 1-m rod functions in this manner up to 30 MHz.

7-3.3.2.2 Magnetic Antennas

7-3.3.2.2.1 Loop Antenna

A small loop in a varying magnetic field will develop an induced voltage v_i (par. 3-3.1.1.3)

$$v_i = \omega BNA, \text{V} \qquad (7\text{-}36)$$

where

ω = radian frequency of flux variation, rad/s
B = magnetic flux density normal to the loop plane, T
N = number of turns, dimensionless
A = area of the loop, m²

Unless the loop area is small, it is possible that an electric dipole mode will generate a significant component of current in the loop (Ref. 15). It will have negligible effect if the loop diameter is less than 0.01λ. The presence of an electric field effect can be checked by rotating the loop 180 deg in its own plane. If no change is observed, the effect can be neglected.

The loop impedance Z_a is given by

$$Z_a = R_a + R_o + j\omega L_a, \Omega \qquad (7\text{-}37)$$

where

R_a = 31,200 $NA/2$, the radiation resistance, Ω
R_o = loss resistance of the loop, Ω
L_a = loop self-inductance, H

There is also some distributed capacitance between turns and some capacitance due to the connecting cable which effectively shunts the source comprised of v_i and Z_a. The reactances generally become part of the input circuit tuning.

The effective height l_e of a loop is commonly taken to be the ratio of induced voltage to the electric field strength component of a plane wave. In such a field the electric E and magnetic H components are related by

$$E = \sqrt{\frac{\mu_0}{\epsilon_0}}\, H = 377\, H, \text{V/m} \qquad (7\text{-}38)$$

where

μ_0 = permeability of free space, $4\pi 10^{-7}$ H/m
ϵ_0 = permittivity of free space, $1/36\pi \times 10^{-9}$ F/m
H = magnetic field strength, A/m

Then the effective height (effective length) is

$$l_e = \frac{\omega NA}{c}, \text{m} \qquad (7\text{-}39)$$

where

c = speed of light in free space = 3×10^8 m/s

It must be emphasized that in near-field measurement the height as calculated by this relationship has no unique relationship with the actual electric field at the point of measurement.

7-3.3.2.2.2 Hall Effect Sensor

The Hall effect sensor is a device for measuring magnetic flux density in the HF range and below, using the principle of the Lorentz law for the force on moving charges in a magnetic field. The principal element in the sensor is a bar of semiconductor material, such as indium antimonide. The B-field to be measured impinges in the bar perpendicular to one face (see Fig. 7-15) and a direct current I_{dc} is passed through the bar perpendicular to the direction of B. As a consequence, electrons (or holes) in motion experience a force in a direction perpendicular to both B and I_{dc} and this manifests itself as an electric field perpendicular to both B and I_{dc} which is sensed externally as a voltage across two sides of the semiconductor. The voltage V_H is given by (Ref. 16):

$$V_H = R_H \frac{I_{dc}B}{t}, \text{V} \qquad (7\text{-}40)$$

where

R_H = Hall coefficient, m³/C
t = thickness of the bar, m
I_{dc} = direct current, A
B = magnetic flux density, T

Field probes using the Hall effect generally are comprised of a Hall sensor inserted in a gap in a ferrite flux collector. (For construction details see Ref. 17.) A probe using a 1-ft long, high permeability ferrite, with a Hall sensor current of 50 mA, was reported to have a transfer ratio of 0.3 volt/gauss (1 gauss = 10^{-4} tesla). Field detectability depends on the noise generated by the device. It is reported (Ref. 17) that for a unity signal-to-noise ratio at the output, as seen in a 1-Hz bandwidth, the flux density has to be about 10^{-8} gauss at frequencies of several hundred hertz and higher, and about 10^{-7} gauss at 10 Hz. The

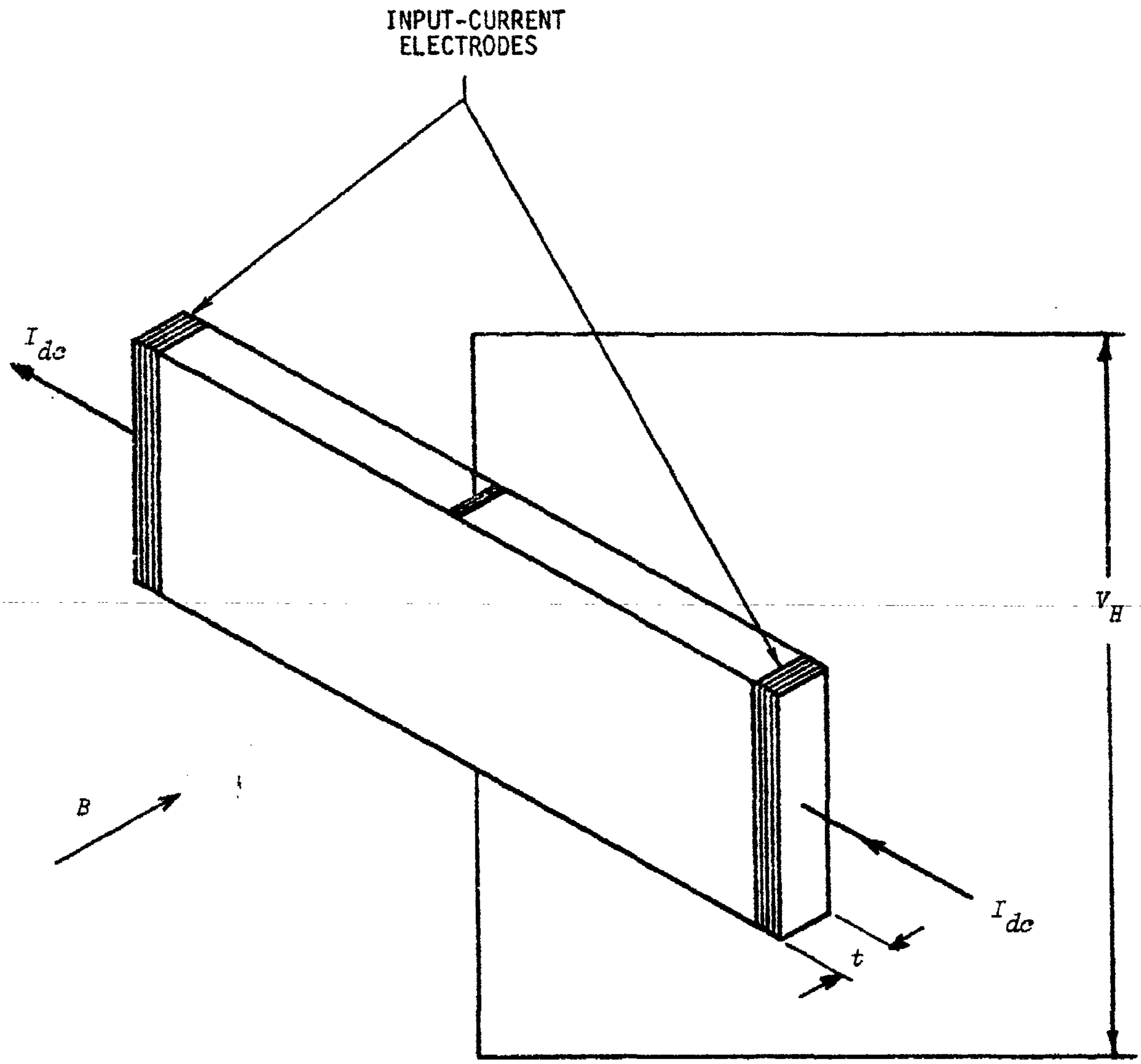


Figure 7-15. Schematic View of a Hall Effect Sensor

deterioration of the sensitivity is a consequence of the $1/f$ noise.

7-3.3.2.2.3 Variable-mu Sensor

The variable-mu sensor is another device for measuring magnetic flux density. Practical implementations of this device operate in the range 0.1 Hz to 50 MHz. It utilizes the property that the *B-H* curve of ferromagnetic material is not linear and that the μ can be controlled by an external field. A ferrite rod is excited by a biasing field to a region where μ vs the *B*-field is nearly linear. Then, external fields to be measured modulated the μ around the biased value. An inductor (see Fig. 7-16) wound around the ferrite has a variable inductance, depending on the instantaneous value of the μ and the inductance, being part of an oscillator circuit, causes a frequency modulated signal to be generated. The frequency deviation is therefore a measure of the magnetic flux density. A particular realization (Ref. 18) operates from 10 Hz

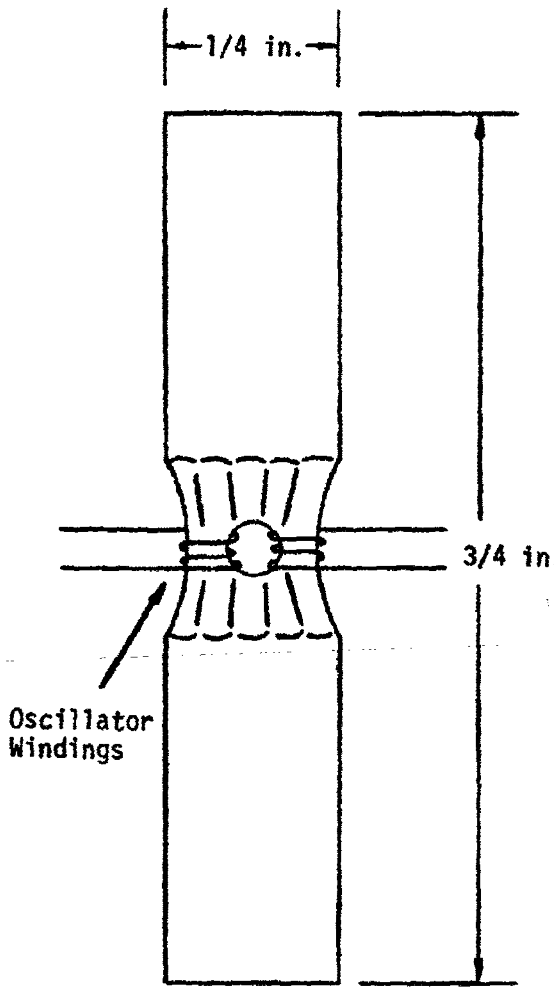


Figure 7-16. Basic Structure of Variable-mu Sensor

to 10 MHz and has a sensitivity of 5×10^{-4} gauss. In a 1-Hz bandwidth sensitivity of 10^{-5} μT has been achieved.

7-3.3.2.3 Half-wave Dipole

At frequencies in the order of 50 MHz and higher, it becomes practical to use the tuned half-wave dipole $l \approx \lambda/2$ which has an effective length

$$l_e = \frac{\lambda}{\pi}, \text{m} \qquad (7\text{-}41)$$

Its impedance is close to 72 Ω, resistive. This value is easy to match in an instrument input circuit, but the necessity of tuning or length adjustment is a distinct disadvantage. Furthermore, it requires a balanced input circuit on the instrument for best results. With care dipole antennas are capable of good field measurement accuracy and often are used as a basis for comparison with other antennas at frequencies above the HF range and into the UHF range where their sensitivity becomes low.

7-3.3.2.4 Broadband Antennas

For frequencies above 20 MHz, antennas capable of operating over frequency ranges of the order of 10:1 without requiring tuning adjustment are available. Such antennas are particularly useful for carrying out automatically swept measurement over a broad frequency range. MIL-STD-461 recommends use of a number of broadband antennas for radiated emission and susceptibility tests as follows:

a. 20-200 MHz : biconical with balun
b. 30-300 MHz : biconical with balun
c. 200 MHz - 1 GHz : conical log-spiral
d. 200 MHz - 2 GHz : double-ridged waveguide horn
e. 1-10 GHz : conical log-spiral antenna
f. 1-12.4 GHz : double-ridged waveguide horn
g. 12-18 GHz : horn-fed 18 in.-dish
h. 18-26 GHz : horn-fed 12 in.-dish
i. 26-40 GHz : horn-fed 18 in.-dish

In dealing with such antennas it is customary to specify the antenna factor A_f defined by

$$A_f = 20 \log(E/V) \text{ , dB} \qquad (7\text{-}42)$$

where

V = voltage measured at the input terminals of the EMI receiver, V
E = electric field at the antenna location, V/m

MIL-STD-462, Notice 3, which is called out for Army procurements, includes a biconical antenna with an improved bifilar balun for the range 30-300 MHz. It has an antenna factor ranging from about 5 to 30 as shown in Fig. 7-17. The 20-200 MHz biconical antenna listed is an earlier design with antenna factor ranging from about 9 dB to about 18 dB.

The conical log spiral, as the name implies, is a logarithmically spaced winding on a cone. It has an antenna factor as shown in Fig. 7-18 and is circularly polarized as is characteristic of spiral antennas.

An antenna type frequently used is the ridged guide antenna covering the band 200 MHz to 2 GHz. The structure is similar to that of a horn, using a double-ridged construction to get broader bandwidth (a decade range) than one gets with a conventional

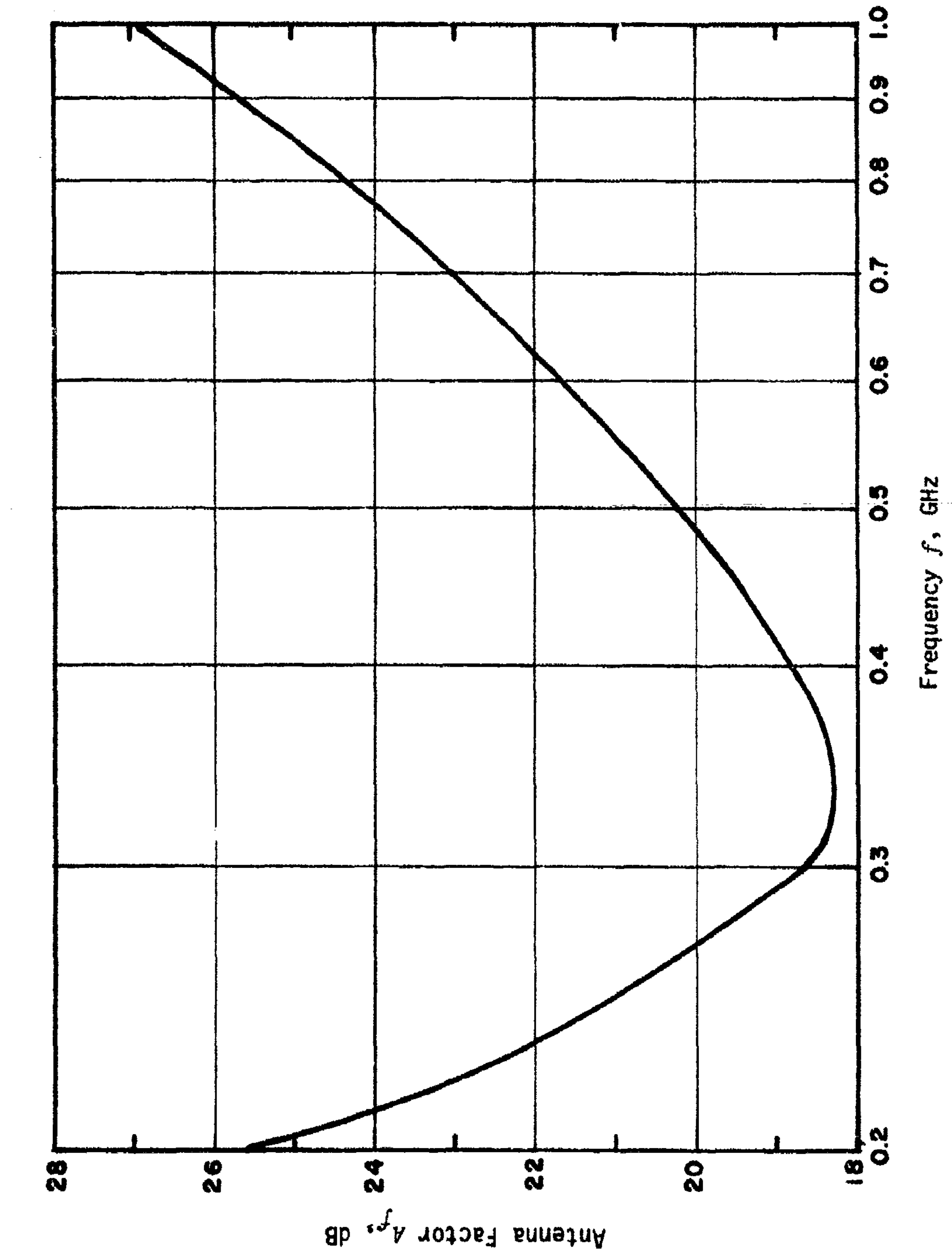


Figure 7-18. Antenna Factor for Log Spiral Antenna

horn. The antenna factor ranges from about 12 to 25 dB over the working frequency range. A similar structure for the 1-12 GHz range has an antenna factor with ranges from 20 to 40 dB. The polarization is linear.

The horn-fed parabolic dishes are relatively narrowband compared to the other antennas described, principally because the horn feed is relatively narrow band. Polarization is linear. The 12-18 GHz antenna has an antenna factor close to 19 dB; the 18-40 GHz antenna has an antenna factor close to 23 dB.

Radiated susceptibility measurements are performed using the same antennas as for emission test RE01. MIL-STD-461 specifies, however, that in the range 14 kHz — 25 MHz, where a 0.5-m rod ordinarily would be used, this antenna be used for susceptibility testing as well, except where fields in excess of 1 V/m are to be generated. In the latter case, antennas capable of handling the power are required such as the long wire and cage antennas and the parallel plate line, see par. 7-5.3.2.3. The long wire antenna and the parallel plate line are approved devices for generating high field intensity in the range 14 kHz to 30 MHz.

7-3.3.3 Calibration

Calibration of instruments with antennas or probes connected usually is not carried out in the user's laboratory. The National Bureau of Standards and the Army Metrology Center, Redstone Arsenal, AL, can provide this service.

Two methods of antenna calibration have been identified as the standard antenna and standard field methods (Ref. 19). The standard antenna method was used to calibrate a number of instruments having rod antennas under near perfect ground plane conditions (Ref. 20). The results showed very good agreement among instruments, probably showing that theoretical values of effective length for the rod antenna were close to actual values, and with theoretical calculations (Ref. 21). In these tests the source antenna was a rod located vertically over a ground plane. The current at its base was measured with a current probe.

Since at frequencies below 30 MHz the rod antenna has a very high reactive impedance, it cannot be attached to a cable — without using a matching circuit — without severe losses. For this reason the antenna normally is mounted directly on the EMI instrument case, and the case serves as a counterpoise. When the instrument is mounted off of a ground plane, relatively good accuracy can be obtained with operation from batteries contained in the case. Otherwise, an attached cable can cause considerable error in measured value.

Matching transformers have to be used with a rod antenna connected at the end of a cable; however, losses in sensitivity occur because of the reactive input impedance.

Loop antennas usually are calibrated with an induction field method (Ref. 19) in which the loop to be calibrated is placed parallel to a transmitting loop carrying a current of I amperes and located at a distance d from it. The effective value of the equivalent electric field E over the area of the receiving loop is then

$$E = \frac{60\,A N_1 I}{(d^2 + r_1^2 + r_2^2)} \sqrt{1 + \left(\frac{2\pi d}{\lambda}\right)^2}\,,\ \text{V/m} \quad (7\text{-}43)$$

where

A = area of transmitting loop, m^2
N_1 = number of turns of transmitting loop
r_1 = radius of transmitting loop, m
r_2 = radius of receiving loop, m
d = distance between loops, m
λ = the wavelength, m

The current in the transmitting loop usually is measured with an accurately calibrated thermoammeter.

Loops can be connected to cables because their impedance is low, cable loss is low, except at frequencies near 30 MHz, and their self-inductance still can be treated as part of the tuned circuit of the input stage. Special design techniques may be necessary to enable insertion of an attenuator in the input circuit.

MIL-STD-461 specifies a technique for determining the antenna factor for the antenna types invoked for use above 20 MHz. The technique is based on a method using identical antennas for source and receiver as described in the specification ARP-958 of the Society of Automotive Engineers (SAE). The test setup is shown in Fig. 7-19. After tuning adjustments have been made, the level in volts of the transmitted signal is determined with (1) the antennas coupling transmitter and receiver, and (2) a direct connection between transmitter and receiver (the antennas are disconnected for this step) and transmitter readjusted to give the same received level as in (1). The two levels of voltage are denoted V_a (antenna coupled) and V_b (direct coupled). The antenna gain G and antenna factor A_f are

$$G = \frac{4\pi r/\lambda}{V_b/V_a}\,,\ \text{dimensionless} \quad (7\text{-}44)$$

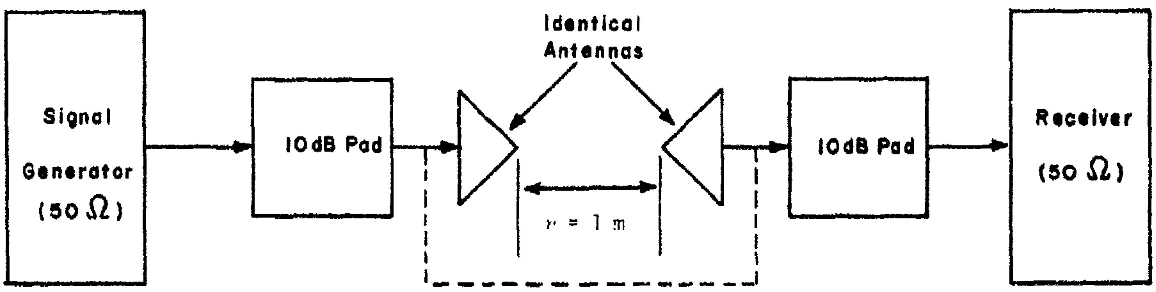


Figure 7-19. Setup for Measuring Antenna Factor

$$A_f = 20 \log\left(\frac{9.76}{\lambda\sqrt{G}}\right), \text{dB} \qquad (7\text{-}45)$$

where

r = distance between transmitting and receiving antennas, m

λ = wavelength, m

The distance r is shown to be 1 m in Fig. 7-19, but for the cavity-backed spiral, MIL-STD-461 specifies 4-ft separation.

7-3.4 AUTOMATIC INSTRUMENT SYSTEMS

Specifications such as MIL-STD-462 impose emission and susceptibility test requirements over frequency ranges extending from 30 Hz to 20 GHz. For example, test REO3 for measuring transmitter radiated spurious and harmonic emissions specifies testing over the range 10 kHz to 1 GHz, using a standard EMI meter using three antenna types to cover the range, and it specifies a spectrum analyzer with a variety of preselectors and antennas to cover the range above 1 GHz. The equipment switching, retuning, calibrating, measuring, and recording operations over such a wide range are time consuming when done manually. Systems programmed to perform these functions automatically or semi-automatically are currently available for both emission and susceptibility testing. The available systems are not designed to cover the entire range of audio frequencies through the microwave range in a single system.

Fundamentally, these systems are comprised of standard EMI measuring instruments which are mechanically or electronically sweep-tuned and which automatically switch in the proper tuning heads and antennas as needed. Output is displayed using an x-y plotter.

A block diagram of a typical automatic emission measuring system is shown in Fig. 7-20. The heart of the unit is the Programmer and Control Center. Here the program is stored for switching sensors, tuners, bands, attenuators, and matching networks; selecting sweep rates and ranges and carrying out tuning operations; changing bandwidths and detector functions; and for other operations according to the particular system. The control circuits generate the necessary instructions to the other elements of the system and provide the necessary signals and power to carry out the operations.

The sensor selector unit receives inputs from the control unit and connects the proper antennas, probes, line stabilization networks, and sensor coupling devices to conform with the program requirements and with the frequency being observed. If there are bandswitches on a sensor, it also is switched when the frequency requires it.

The RF unit typically contains (or is connected to) impulse calibrators, step attenuators, RF bandpass and/or band-elimination filters, and, where important, a low noise preamplifier. The calibrator output ultimately results in a set of calibrated lines appearing on the output display identified with amplitude level. Since the calibrator usually is coupled into the system at a point after the sensor, it cannot account for the sensor sensitivity. Sensor sensitivity, i.e., antenna factor, must be added to the calibration obtained when the calibrator output is injected at the input of the RF unit. In some cases, injection can be made at the antenna, but this normally will not account for antenna directivity. When current probes are used, they can be calibrated by direct injection into a circuit around which it is clamped.

The receiver system required to cover the wide range involved usually is comprised of several tuning heads and several IF amplifiers. These are switched in as the program requires, and the tuners are scanned

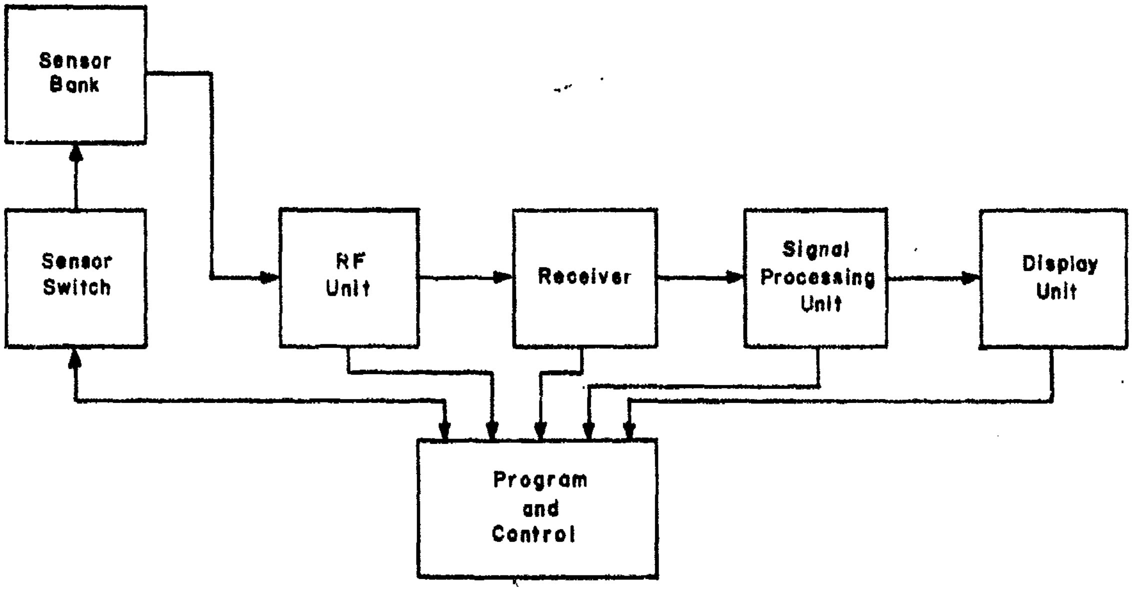


Figure 7-20. Block Diagram of Automatic Measuring Equipment

at speeds programmed by the operator. Typical programs allow portions of bands to be scanned, and scans to be repeated. Since conventional EMI receivers ordinarily are used, modifications are necessary to allow automatic tuning, and to make the signal conform to the frequency of the processing equipment which follows.

The signal processing unit is comprised of IF amplifiers and detector circuits. The amplifiers are generally logarithmic to accommodate the wide range of amplitudes expected. Bandwidth tailoring may be done here. The different detector functions may be switched in, or programmed to operate simultaneously to give multiple outputs. As a rule peak and average functions are performed. The peak circuit frequently is equipped with a variable peak holding duration and dump circuit to accommodate the response time of x-y plotters. Features which permit signal reading while writing are also included. The average circuit measures the average of the envelope of the IF output in a time interval appropriate to the IF bandwidth. Where instantaneous logarithmic compression is used in the IF amplifier, the measured function is actually the average of the logarithm of the envelope. Conversion to the true envelope average can be done if the waveform shape is known, or, for random waveforms, if the envelope probability density function is known. If the waveform is not known, a digital processing technique can be used to provide the correct average value.

The displays are typically an x-y plotter oscilloscope, and/or audio monitor. The x-y plot is sometimes a multiple line display showing many bands, one above the other. The ordinary display shows amplitude vs frequency, but some systems are capable of other kinds of signal analysis (e.g., probability density function determination) and these too are printed out. In some systems it is possible to program in the specification limit and have this printed out simultaneously, thus giving a quick indication of nonconformity with the specification requirements.

Automatic susceptibility testing involves programming a suitable collection of signal generators, and radiators or injectors. Modulation type (CW or pulse) and amplitude are programmable; drive signals as required by specifications are thereby synthesized rapidly. These systems also extract output information from the equipment under test and display it to show where in frequency, and to what extent, malfunction occurs.

7-4 TEST FACILITIES

7-4.1 GROUND PLANES

Ground planes, or more accurately ground surfaces, are used for two basic purposes: (1) establish a known reference potential for conduct of measurements, and (2) establish a surface of known reflectivity for radiated measurements.

The theory of such surfaces, as applied to conducted measurements, is discussed in par. 3-3.4, and methods of connecting to them in par. 4-7. In test arrangements they serve to minimize the possibility of spurious voltages being introduced on interconnecting cables by tying all equipment cabinets to a common reference potential. Without this common potential surface, uncontrolled potential differences between cabinets can exist as a result of common mode currents (see par. 3-3.3.3) flowing between cabinets, or currents induced on the cables from local radio frequency fields.

Except for open field tests, where it is impractical, the size of the surface should correspond with that required to accommodate the equipment under test and the test equipment. It should be of sufficient thickness to accommodate the largest currents which may flow, without producing potential differences which may either (1) adversely affect the operation of equipment under test, or (2) cause significant errors in measured voltages and currents.

The resistance R between two circular areas of contact on a flat conducting sheet is

$$R = \frac{\rho}{\pi t} \ln \left\{ \frac{D}{d} \left[1 + \sqrt{1 - \left(\frac{d}{D} \right)^2} \right] \right\}, \ \Omega \qquad (7\text{-}46)$$

where

ρ = resistivity, $\Omega \cdot$m

d = diameter of contact areas (assumed equal), m

D = distance between contacts, m

t = thickness of sheet, m

Fig. 7-21 shows values for this function for a 0.25-mm copper sheet, $\rho = 1.724 \times 10^{-8}$ $\Omega \cdot$m

The ground plane may have an important effect on radiation measurements. Fig. 7-22 shows that with a horizontal electric dipole, the effect of a ground plane on the field is accounted for in terms of an inverted image. For a perfectly conducting plane and at distances d large compared with the antenna heights h_1, h_2, the field E at point P (in the field region) at a distance r is given in terms of the free space field E_0 (Eq. 3-145) by (see Eq. 3-151)

$$\frac{E}{E_0} = 2 \sin \left(\frac{2\pi h_1 h_2}{\lambda r} \right) \qquad (7\text{-}47)$$

On the other hand, at distances where the inverse cube law applies (par. 3-3.2.1), if one is close to the antenna compared with the antenna height above the plane, the image can be neglected.

For a vertical electric antenna, the image is not inverted if the reflected ray does not approach grazing incidence. Then Eq. 7-47 is replaced with

$$\frac{E}{E_0} = 2 \cos \left(\frac{2\pi h_1 h_2}{\lambda r} \right) \qquad (7\text{-}48)$$

and the field is greatest near the plane. However, under conditions of grazing incidence, the image becomes inverted and Eq. 7-47 applies also in that case (Refs. 22 and 23). This latter condition is not likely to apply in most EMC test work.

For magnetic dipoles the effects of reflection are reversed, i.e., the image of a horizontal magnetic dipole is not inverted and a vertical magnetic dipole is inverted. It should be noted that while most conducting surfaces form images that are of strength closely equal to that of an original source, magnetic sources are fully reflected in a ground plane if the surface itself has good magnetic properties. Practically this means that for frequencies in the kilohertz range and below, magnetic images will be formed only if the thickness of the conducting plane is about twice the skin depth or it is of high permeability.

7-4.2 SHIELDED ENCLOSURES

A shielded enclosure is used to provide an environment free of extraneous electromagnetic radiation which would otherwise disturb EMC testing. The basic theory of the attenuation provided by conductors of which such enclosures are fabricated is discussed in par. 4-6. A variety of materials has been used depending upon the objectives, including screening and solid sheet material. Where screening is used, it is common to use a double-wall construction in order to obtain the required attenuation. The two walls can be either (1) completely insulated from each other except at the point where power leads penetrate the enclosure, or (2) in contact at the perimeter of panels (of the order of 3 ft × 8 ft or larger) which are bolted together to form the enclosure. The latter construction is sometimes referred to as a "cell" type and generally can be expected to have a lower attenuation than the isolated wall type. Materials which have a high permeability μ are used where high attenuation is required below 10 kHz. As discussed in par. 4-6, in any construction method special care is required in making seams or joints and in necessary openings.

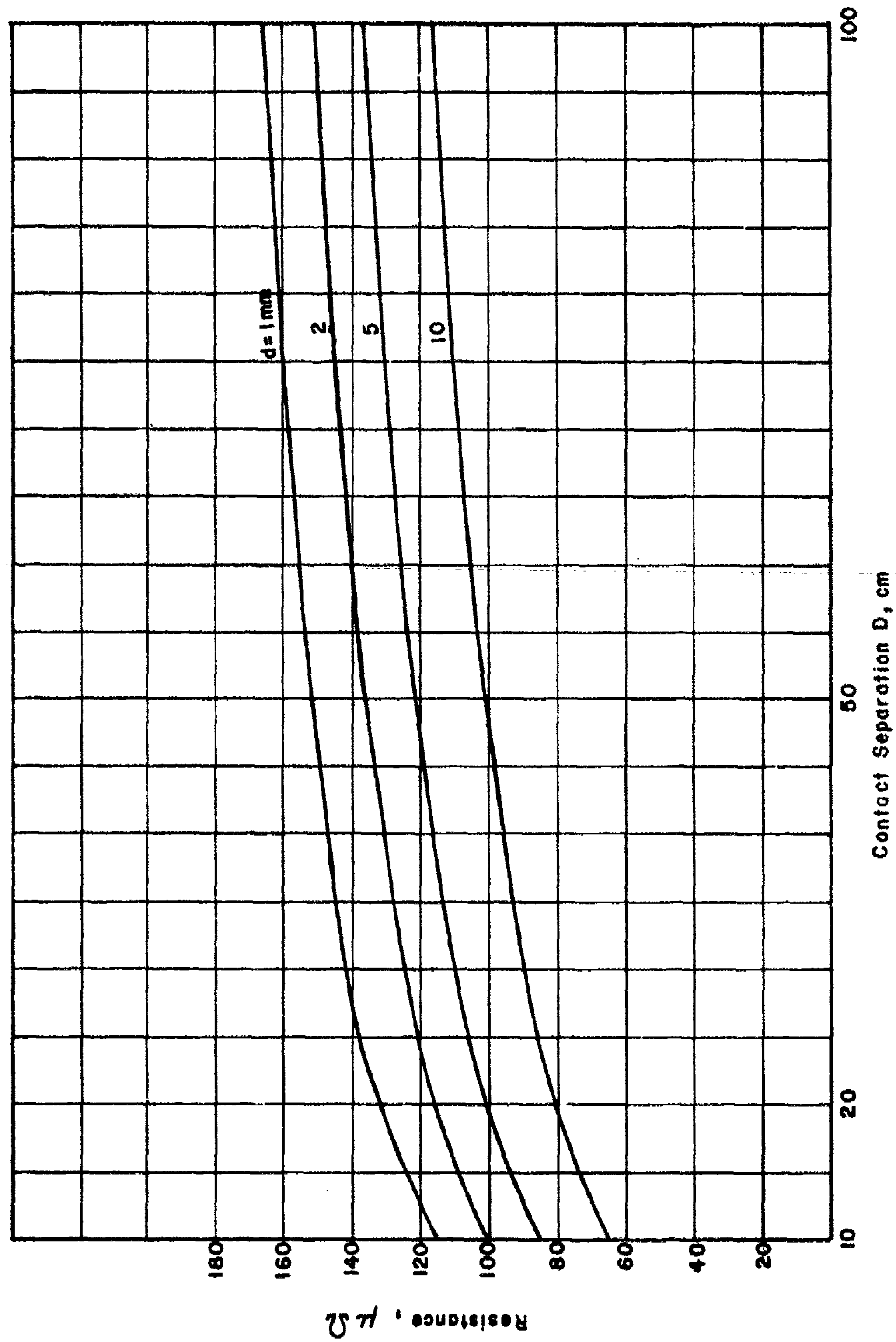


Figure 7-21. Resistance Between Two Circular Electrodes on 0.25-mm Copper Ground Plane

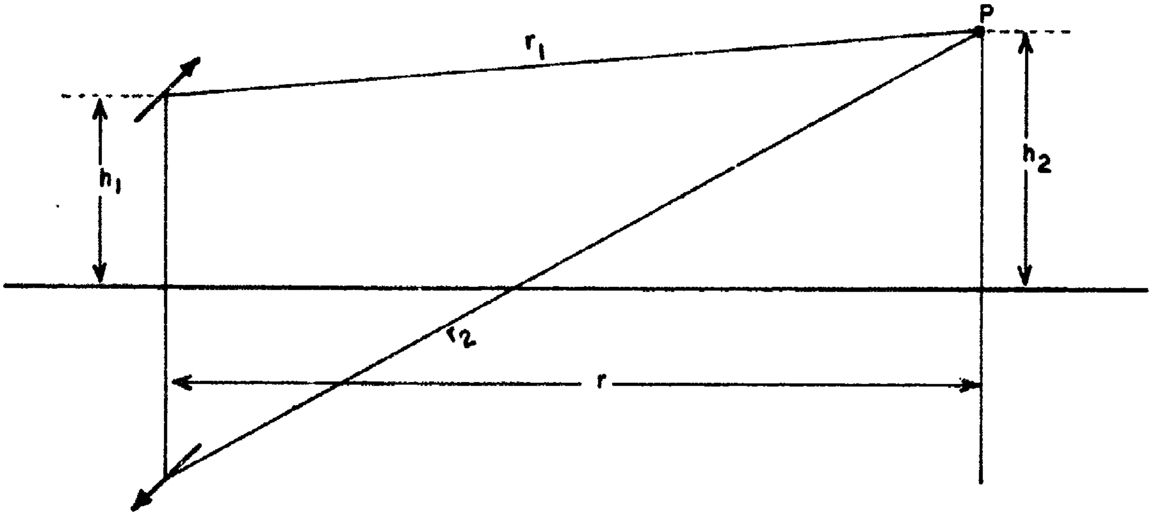


Figure 7-22. Effect of Ground Plane on Field from Horizontal Dipole

7-4.2.1 Attenuation

The attenuation provided by the shielded enclosure is probably best defined as the reduction in field at a given location when the location is surrounded by the enclosure. Measurement of the attenuation usually must be done with the enclosure in place with a source field antenna or other radiating device on one side of the enclosure and a receiving probe or antenna on the other. It can be quite dependent on the technique used, especially since leakage at joints or cable penetrations may be significant and the measured attenuation will depend on exactly where the measurement antennas are placed. Standard methods are available (Ref. 24, 25) and prescribe specific procedures for various segments of the frequency range, using loops at low frequencies, dipoles at intermediate frequencies, and horns at UHF.

The theoretical attenuation obtainable with various materials is given in par. 4-6 (see, e.g., Figs. 4-77, -78, -79). Attenuations of 100 dB or more are easily obtained for electric and radiation fields over the entire frequency range. For magnetic fields, typically, attenuation drops rapidly below about 1 MHz and in the audio frequency range may be only 10 dB or so unless material with high permeability is used.

7-4.2.2 Construction

Where the enclosure is designed to be movable, it is constructed of panels which are bolted together. To get good performance requires continuous low resistance metal-to-metal contact. For low-frequencies (below 10 kHz) the magnetic material should have an overlap to minimize magnetic field leakage at the joint. Where the enclosure is in a fixed position, joints can be soldered or welded so that they are as good as continuous material. Ventilating louvers should be covered by screen or waveguide below cutoff filters, as discussed in par. 4-6.5.2.

7-4.2.3 Arrangements for Testing

Usually, testing is confined to equipment within the enclosure itself. Then the only conductors brought into the room are the power lines, both ac and dc as required. These lines must be filtered over the entire frequency range for which the enclosure is designed or its effectiveness will be compromised. With care in design, suitable filters are available covering most of this range. The low audio frequencies are the most difficult to filter. Power is supplied for lighting and test purposes in conduit enclosed wiring. Incandescent lights are preferred in smaller rooms. Fluorescent lights can be used if adequately shielded and filtered.

For conducted emission measurements, a metal covered work bench is installed at table height, with power supply connections as needed. It should have facilities for bonding various items of equipment to it, and should be bonded itself to the enclosure periodically, at least every 30 in. One edge of it should run alongside a wall of the enclosure.

The enclosure also should have provision for connecting signal cables through the wall of the room so that auxiliary test equipment can be isolated from the test space. The connectors should provide for coaxial as well as twisted pair arrangements and, where necessary, should permit the possibility of passing multiconductor cables through the wall. The essential requirement is that such cables be shielded and that a good connection be made of this shield to the wall so that currents picked up on the shield external to the room cannot be conducted inside.

7-4.2.4 Radiated Emission Testing

An enclosure that does not contain absorbing material will reflect most of the electromagnetic energy that impinges on its wall. Hence, the field from a radiating source can be expected to be significantly different within an enclosure than outside of it. Indeed a room of rectangular shape and free of objects with significant electrical properties will resonate at frequencies f defined by the relation (Ref. 26, p. 23-19)

$$f = \sqrt{150\left(\frac{m}{a}\right)^2 + \left(\frac{n}{b}\right)^2 + \left(\frac{p}{c}\right)^2}\ , \text{MHz} \qquad (7\text{-}49)$$

where

a, b, c = interior dimensions of the enclosure, m

m, n, p = integers only one of which may be zero at a time

For an enclosure of cubical shape, 3 m in each direction, the lowest resonant frequency is approximately 70 MHz. Clearly, at this frequency and above the field will vary in an unpredictable way with distance from a source.

A reflecting room can be used at frequencies well below resonance under the following conditions:

a. The object under test and the test antennas have dimensions no more than about 1/3 of the minimum enclosure dimension and are located near the center of the room.

b. The receiving antenna is balanced and in the induction or near-field of the object under test.

The first condition is imposed in order to insure that the enclosure does not affect significantly the radiating impedance of either receiving antenna or emitting device. Because of the second condition the field emitted can be expected to fall off according to an inverse cube law (par. 3-3.2.1) and therefore fields from image sources reflected by the walls of the enclosure can be expected to be quite weak. Tests have shown that results obtained under these conditions are within a few dB of open field measurements (Ref. 27) made as the same distance from a room quite clear of reflecting objects. As the antenna dimension increases sufficiently so that the ends of the antenna come close to the walls of the room, the capacitance coupling produces significant errors at frequencies well below the first room resonance, say at 40 MHz (Ref. 28).

7-4.2.5 Recommended Arrangements

The design of any specific shielded enclosure will depend upon the use to be made of it; i.e., the size of the device to be tested, the electromagnetic environment surrounding it, size of necessary doors, and the power and ventilation or air conditioning requirements. However, for general test purposes the following design characteristics are recommended as a minimum:

a. Size: 20 ft × 20 ft × 10 ft
b. Shielding Effectiveness: Magnetic Field, 30 dB at 10 kHz and above
Electric Field, 100 dB at 30 Hz and above
c. Power: Single phase, 115 and 220 V, 60 and 400 Hz regulated with not more than 3% total harmonic distortion
d. Filters at Entrance: 100 dB, 10 kHz to 1 GHz
e. Ventilation: air conditioned
f. Ground Plane: Table height 3 ft wide along the longest side
g. Access Panel: For signal entry or exit.

7-4.2.6 The Anechoic Enclosure

An enclosure can be made nonreflecting by coating the walls with absorbing material. To be effective over a broad frequency range, experience has shown that the depth of material should be about one-fourth wavelength. For this reason, enclosures are practical only for frequencies above about 50 MHz. Sometimes the enclosure walls are made nonrectangular to reduce the number of resonant nodes. The effectiveness of the absorption is such that essentially free space field patterns can be obtained within the enclosure.

When the cost of installing such material over the walls of the entire enclosure is too great, partial installation of material will provide significant improvement over a reflecting enclosure, especially if it can be placed at locations at which energy is reflected from or towards a directive antenna, or placed around the antenna (Ref. 29).

7-4.2.7 Antenna Pattern Synthesis

In the past few years a technique of antenna pattern synthesis has been developed in which the field distribution over the antenna aperture is measured and the pattern is calculated by a computer. A small dipole type antenna physically scans the aperture and both the amplitude and phase of the detected signal are recorded (Ref. 30). An anechoic volume is required which is only large enough to contain the antenna under test and the scanning apparatus.

7-4.3 OPEN FIELD TESTS

Open field test sites are used when:

a. A suitable shielded or anechoic enclosure is not available perhaps because of equipment size.

b. At the lower frequencies, testing in the far field requires large antenna spacings.

c. It is desired to examine the effects of ground reflections on antenna patterns.

A suitable site requires a flat area of uniform soil characteristics free from reflecting objects within a distance to the equipment in use at least equal to the separation between the radiating and the receiving antennas, and preferably substantially greater than that, and it should be remote from areas having substantial industrial activity, dense population, or moderate or high power transmitters. Power lines to the equipment should be buried underground so as not to affect the electromagnetic properties of the site. The device under test is usually placed on a turntable for convenience in determining the radiation pattern, and in particular, for locating the direction of maximum radiation. For tests where ground reflections are significant, the height of the test object can be important as discussed in par. 7-4.1. Equipment can be placed in weather protective buildings made of nonconducting, low dielectric constant material where necessary.

Because of the difficulty of finding suitable test sites and the cost of acquiring and equipping them, open field testing usually is used only as a last resort.

7-4.4 ELECTRO-OPTICAL TECHNIQUES

Recent developments in fiber optics employing light for data transmission have enabled development of instruments in which connected wires have been eliminated. Arrangements using these instruments have substantial advantages in some tests involving radiated fields, where the connected wires could otherwise disturb the fields being set up or measured. The measured value is coded into a form suitable for transmission on the fiber optic cable, which itself does not disturb the field, to a remote detector which decodes the optic signal to indicate the measured value. Devices using this principle include (1) a 20-200 MHz antenna system for use in shielded enclosures (Ref. 31), and (2) an electric field probe (Ref. 32).

In another technique (Ref. 33) a small dipole probe whose impedance can be modulated by a light beam is used to measure the field distribution near an antenna. The modulation caused the reradiated component to be measurable even in the presence of a relatively strong antenna field and the absence of any wire connected to the probe avoids any significant distortion of the original field.

7-5 MEASUREMENT TECHNIQUES

7-5.1 INTRODUCTION

Just as in dealing with the analysis of EMC phenomena, in discussing measurement techniques, it is convenient to separate them into those used for (1) measuring the characteristics of emitters, (2) measuring characteristics of susceptors, and (3) measuring coupling factors between emitters and susceptors. Military standards are concerned primarily with the first two categories. The third category is measured only indirectly according to military standards, usually in the form of laboratory measurements since such measurements are of considerable interest in connection with gathering data necessary for analysis and prediction work in the early phases of equipment or system design. They are treated in par. 7-5.4.

Measurements in the first two categories are discussed in pars. 7-5.5 and 7-5.6. The emphasis is on the principles to be observed in making significant measurements rather than on details of the procedures which, in many cases, can be obtained from appropriate military specifications and standards. Individual test methods described in MIL-STD-462 are referenced wherever appropriate, but the discussion given here should not be interpreted as superseding that document. When called out, that document should be consulted for details on the test setup and procedure.

For reference purposes, Table 7-3 lists the tests described in MIL-STD-462. These cover conducted and radiated emission and susceptibility tests over the frequency range 30 Hz to 40 GHz.

Uncertainties in test procedures arise especially where there are large numbers of equipments that have to be tested together. As a guide, the equipment should be placed as nearly as possible in relative positions that simulate their installation arrangement, and generally the adjustable controls should be set in positions that will produce worst case emission or susceptibility. In other situations the device under test will be connected to a simulator to provide operationally realistic signals. In a similar fashion, auxiliary monitoring equipment may be connected to the device under test to provide a measure of its performance. It should be ascertained that such auxiliary equipment, and any standard test equipment — such as oscilloscopes, signal generators, and frequency meters — do not themselves produce interference not present with the equipment under test in normal operation.

Site survey measurements are discussed in par. 7-5.7.

7-5.2 MEASUREMENT ACCURACY

The accuracy of measurement of EMC parameters depends critically on the nature of the measurement, the instruments used and the test arrangement. Specific items to be considered are as follows:

a. Measuring Instrument. Measuring instruments have an accuracy usually considered to be about 2 dB for voltage measurements and 3 dB for field measurements, assuming quality instruments and recent calibration. Where substitution measurements are made,

TABLE 7-3.
LIST OF MIL-STD-462 EMI MEASUREMENT TESTS

Test	Title
	Conducted Emission (CE)
CE01	30 Hz—50 kHz, DC Power Leads
CE02	10 kHz—50 kHz, AC Power Leads
CE03	30 Hz—50 kHz, Control and Signal Leads
CE04	50 kHz—50 MHz, Power Leads
CE05	50 kHz—50 MHz, Control and Signal Leads
CE06	10 kHz—12.4 GHz, Antenna Terminal
CE07	1.5—65 MHz, Power Source, Tactical Vehicles
	Conducted Susceptibility (CS)
CS01	30 Hz—50 kHz, DC Power Leads
CS02	50 kHz—400 MHz, Power Leads
CS03	30 Hz—10 GHz, Intermodulation
CS04	30 Hz—10 GHz, Rejection of Undesired Signals (2 Generator Method)
CS05	Deleted — conflict with MIL-STD-188C
CS06	Spike, Power Leads
CS07	Squelch Circuits
CS08	Deleted
	Radiated Emission (RE)
RE01	30 Hz—30 kHz, Magnetic Field
RE02	14 kHz—1000 MHz, Electric Field — Broadband
RE03	10 kHz—40 GHz, Spurious and Harmonics, Radiated Technique
RE04	20 Hz—50 kHz, Magnetic Field
RE05	150 kHz—1000 MHz, Vehicles and Engine Driven Equipment
RE06	14 kHz—1 GHz, Overhead Power Line Test
	Radiated Susceptibility (RS)
RS01	30 Hz—30 kHz, Magnetic Field
RS02	Magnetic Induction Field Spike
RS03	10 kHz—400 MHz, Electric Field

better accuracy is obtainable with a stable substitution source if it has the same type of waveform as the interference being measured. For this reason, a stable internal impulse generator is potentially capable of enabling measurements of broadband interference with accuracies better than 2 dB. When measuring narrowband interference, however, the accuracy depends upon knowledge of the bandwidth. Variations of the bandwidth with time and tuned frequency can occur. They can be detected and accounted for by calibration with an accurate sine wave generator along with the impulse generator.

b. Impedance Matching. Measurements at frequencies above 30 MHz can be quite sensitive to cable lengths unless care is taken to match impedances at all terminations. Even relatively small standing wave ratios can cause substantial error because of interactions that can occur between sources and loads.

c. Waveform Purity. Spurious emissions and responses from communication equipment require measurement of low signal levels in the presence of large signal levels. Since spurious signals can be generated in any nonlinear device, care must be taken to insure that in any measurement a characteristic of the device in question is being measured. For example, a receiver tuned to a harmonic of the generator can respond either to a harmonic in the output of the generator or a harmonic generated in the receiver input circuit as a result of the high amplitude fundamental frequency at the receiver input. Thus, in measurements of harmonic and spurious output of a transmitter, care must be taken to insure that measured levels are not generated in the measuring receiver. Likewise, auxiliary equipment in the circuit such as power monitors and dummy loads should be checked to determine that they are operating linearly. Techniques for detecting false measurements are mentioned in par. 7-5.6.1.2.

d. Radiation Measurements: The presence of reflecting objects can seriously degrade the accuracy of radiation measurements. In open field measurements the effects of ground reflections can be explored by varying the antenna height and checking for conformity with Eq. 7-48. The effects of reflecting objects can be checked by displacing the antenna in various directions to determine if significant standing waves are present at the test location.

Antenna characteristics must be accounted for. In general, any antenna should be oriented for maximum sensitivity both in direction and polarization. If it is circularly polarized, it should be confirmed that incident radiation is not polarized in the reverse direction or if plane polarized, an appropriate correction factor should be used.

e. Voltage Measurements. Accurate voltage measurements usually require a well defined ground plane reference. Typical ground planes have significant impedance (see par. 7-4.1) so that areas of a plane near points to which large currents are returned, such as from a signal or pulse generator, should be avoided for use as reference potentials.

7-5.3 TEST EQUIPMENT

In addition to basic interference measuring instruments, certain other items of test equipment will be found to be very useful for EMC testing, and indeed will, in many cases, be essential to obtaining the most significant data. The equipment should be of good quality and be maintained regularly. The equipment complement could include the following:

a. Attenuators — 10-dB and 1-dB steps, 50 Ω

b. Signal generators providing continous coverage — 10 kHz to 40 GHz

c. Power amplifiers (up to 100 W at frequencies to several hundred megahertz)

d. Line impedance stabilization networks (with current carrying capacity up to at least 50 A)

e. Isolation transformers (up to 1 kW)

f. Audio power transformers (up to 100 W)

g. 10-μF feedthrough capacitors

h. Current probes

i. Adjustable filters (low-pass, high-pass and bandpass)

j. Notch filters

k. Wave analyzers (at least 0-50 kHz)

l. Audio oscillator

m. Impulse generator

n. Variable width pulse generator

o. Transient generator (see Fig. 7-28 for typical wave form).

7-5.4 LABORATORY TYPE MEASUREMENTS

Frequently measurements are made for exploratory purposes to determine fundamental equipment or component characteristics such as filter insertion loss and shielding effectiveness.

7-5.4.1 Probe Measurements

Four types of probes are used: voltage, current, electric field, and magnetic field.

Conductor emission data can be obtained using voltage or current probes. At frequencies below 1 MHz, instruments are available with high impedances (up to 100,000 Ω) so that usually direct connection to the circuit can be made without disturbing its operation because of loading effects. On lines carrying power, a blocking capacitor of appropriate size may be adequate to reduce the power frequency voltage to levels that will not overload the measuring instrument. If not, or if the line carries a strong desired carrier frequency not subject to emission restrictions, a notch filter may be used between the terminals of the circuit to be measured and the instrument. Current can be measured with a clip-on type current probe. Since the insertion impedance of the probe usually is less than 1 Ω, it is unlikely to disturb the measured current. However, note that the calibration of the probe (transfer impedance) is dependent upon its termination impedance (see par. 7-3.3.1.1), and if the measuring instrument or voltmeter does not have an input impedance of the proper value, a matching network must be constructed and taken into account in the calibration.

Field emission measurements in the immediate vicinity of equipments generally cannot be made accurately because of the interaction that takes place between the antenna and the equipment itself. An exception to this occurs when measuring a small magnetic source which is surrounded by nonmagnetic material. Such a source can be viewed as a magnetic dipole with a field having a highly predictable local distribution (see par. 3-3.1.1.1). It can be measured quite accurately with a small loop or Hall effect sensor. Furthermore, the presence of the human body in the vicinity does not disturb the field, so that an unknown magnetic source can be quite accurately located by probing, especially at the lower frequencies (below about 1 MHz.)

Electric field probes are less effective for several reasons. In the first place, at low frequencies electric fields due to equipment are likely to be weak because cabinets are conductors and will not support large field strengths. As the frequency increases, the wavelength decreases and impedances increase so that detectable field strengths appear. However, the presence of large amounts of metal such as in a tank turret causes multiple reflections which destroy any semblance of a "dipole" field. Furthermore, the probe, if on the end of a cable, will be insensitive because of poor impedance match and the field will vary as the body is moved in manipulating the probe. For these reasons, electric field probes are seldom used. In the frequency range up to about 30 MHz, probing with a loop of small size (about 6 in. diameter max) and a few turns will successfully locate the most significant sources of field leakage.

Susceptibility of circuits to induce voltages can be simulated by using a current probe. When a signal generator is connected to the probe terminals in place of the measuring instrument, the series induced voltage in the circuit it encloses is proportional to the current through the probe and is a function of frequency. Similarly, magnetic and electric field probes can be driven by a signal generator to produce local fields for susceptibility testing.

7-5.4.2 Filter Measurements

The standard test for filters is contained in MIL-STD-220 in which the filter is terminated in 50 Ω at each end. Since many EMI filters are placed in circuits, such as power lines, in which the termination is quite different from 50 Ω, special tests may be advisable for which there are no standard methods at the present time. By placing the filter in a circuit simulating actual termination impedances, its effectiveness can be measured by means of current or voltage probes. In setting up a test circuit, the impedance of the generator must be taken into consideration, and appropriate matching networks used as needed.

7-5.4.3 Shielding Effectiveness

Shielding effectiveness of enclosures or cabinets can be measured using MIL-STD-285 if the enclosure is sufficiently large. If not, special methods must be devised. Included are the inserting of a small battery operated transistor oscillator in the cavity and measuring the field outside of it. Otherwise, energy can be fed to small loops or electric dipoles in the enclosure, using well shielded coaxial cables. Because of the possibility of cable leakage, the enclosure can be placed against the wall of a shielded enclosure and the cable connected to it through an access connector in the wall of the enclosure.

Because the shields on most cables are quite effective providing high values of attenuation, measurement of the actual value of cable shielding effectiveness requires special equipment. Perhaps the most common method is that using a triaxial arrangement (Ref. 34) in which the cable under test is placed within a cylinder which together with the cable shield forms a second coaxial structure isolated from the first except for leakage through the shield of the inner cable. By measuring the voltage across the outer structure when the inner one is driven, the shielding effectiveness can be determined.

At frequencies where the triaxial structure becomes resonant, a quadraxial method has been devised (Ref. 35) in which provision is made for proper loading of the structure to avoid standing waves.

In addition, techniques using an absorbing clamp and a leakage field measurement technique have shown good results at frequencies above 30 MHz

(Ref. 36). Because of the specialized nature of SE test methods, they are not described here in detail. Those interested should consult the references.

7-5.4.4 Articulation Measurements

The principles underlying and general procedures for making articulation measurements are discussed extensively in par. 3-2.1.1. As mentioned, such tests can be carried out with a group of individuals or, in cases where the primary objective is to compare the relative merits of various types of similar equipment, by a simulation in which the elements of speech generally agreed as contributing to its quality are analyzed in analog and digital equipment in various forms. Where human subjects are involved the facility should include acoustic dead rooms, quality microphones, amplifiers, loud speakers, and reproducing facilities. The testing must be done carefully, with the speaking voice at a uniform level, with adequate diction, and controlled symbol enunciation speed. Detailed procedures will be found in Ref. 36 of Chapter 3. Facilities for such measurements are available at Ft. Huachuca (par. 7-6.1.3).

7-5.5 EMISSION CHARACTERISTICS

Measurement of emitter characteristics is complicated by the necessity to take into account both conduction and radiation, and the large variety of waveforms which are possible from one emitter to another. Current and voltage conducted measurements can be made, and field measurements can be made of either the magnetic component or the electric component and their various polarizations. Variations in waveform are identified by using narrowband and broadband measurements as discussed in par. 7-3.

7-5.5.1 Conducted Measurements

7-5.5.1.1 Power Line Measurements

As discussed in detail in par. 3-3.3, an interference source can be modeled as an active network with as many terminals as there are individual conductors or conductor terminals connected to it, plus possibly an extra terminal to account for the cabinet of the device itself. Because of the complexity of an exact equivalent circuit representation of such a source, no attempt is made to obtain it. Indeed, except where common-mode measurements are made on power circuits (see par. 3-3.3.3), it is customary to treat each terminal as independent of others and measure between it and a ground plane. The cabinet of the device being measured is bonded to the ground plane when the measurement is being made as well as all test equipment. In part, this procedure is included for safety reasons. Furthermore, the measure of impedance across the terminals of an active circuit is not a simple procedure and practically is never attempted as a routine matter. The values of impedance that such lines are expected to have can vary over a large range, especially at frequencies above 0.5 MHz (par. 3-3.3.2.1.2) and for this reason knowledge of the source impedance may not be critical. Likewise, the importance of identifying a specific value of measurement impedance may not be critical.

On power lines, measurements are made with either (1) a 10-μF feedthrough capacitor on dc lines at frequencies up to 50 kHz (CE01), or (2) a line stabilization network on ac lines from 10 kHz to 50 MHz (CE02, CE04).

The 10-μF capacitor (Ref. 37) must be capable of carrying full line current and is connected from line to ground in the line to be tested as shown in Fig. 7-23 to provide a low impedance path to ground. The current in the line is measured using a transformer having sufficiently low impedance on the line side to carry the line current, and a turns ratio sufficient to provide measurable voltages at the input to the RIFI meter for currents at the specification limits. The requirement on the 10-μF capacitor is that it should have a controlled equivalent leakage inductance. The effects of leakage inductance are described in par. 4-5.2.1. Fig. 4-31 shows the variation of impedance with frequency for typical capacitors. Measured insertion loss vs frequency is shown on Fig. 7-24 (Ref. 38) along with the specification requirement of SAE. These measurements were made in a 50-ohm network. It should be noted that power line source impedances typically are lower than that of the capacitor at frequencies below about 10 kHz; in fact, resonance with an inductive line impedance is possible near this frequency. The size of this impedance indicates that interference on the power source will not necessarily be bypassed. Hence, checks should be made to assure that measured currents originate in the device under test.

The circuit for the line impedance stabilization networks is shown in Fig. 7-25. Their purpose is to provide an effective power supply impedance corresponding to that which would be experienced in typical distribution circuits. Network (A), which is used from 10 kHz to 10 MHz, has an impedance dominated by the inductance below about 250 kHz and by the 50 Ω resistor above. Network (B) is used above 2 MHz. With these networks, interference current is measured with appropriate current probes on each power conductor. At the higher frequencies

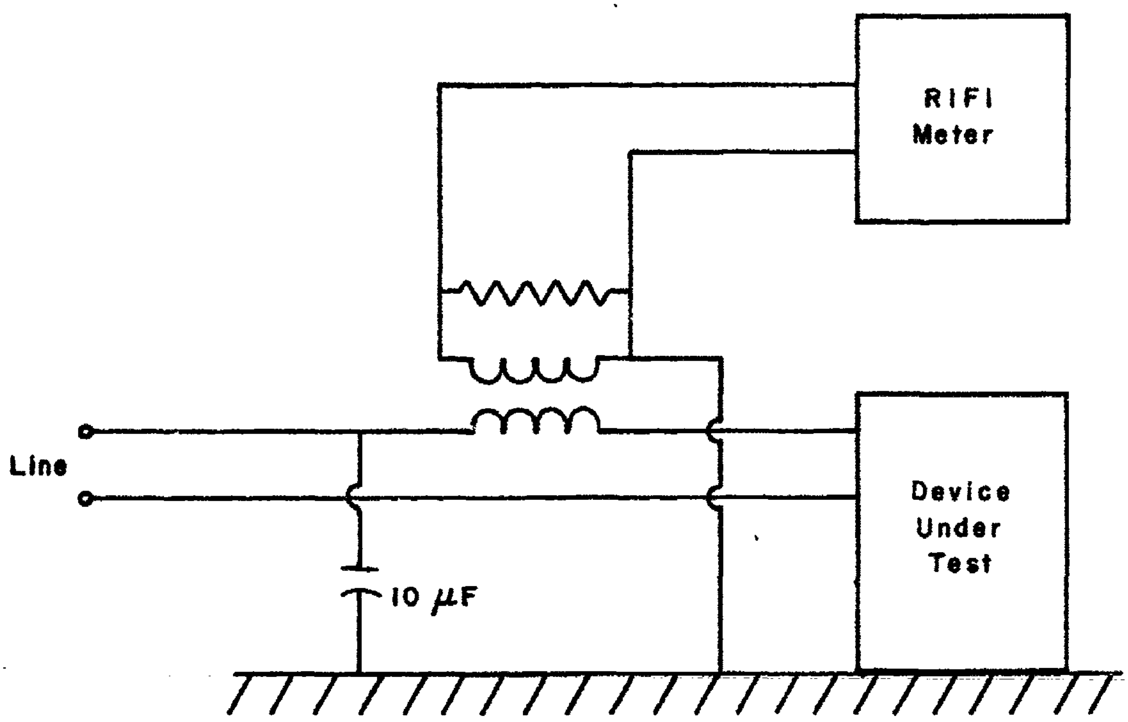


Figure 7-23. Measurement of Interference Current With 10-μF Capacitor

the probe is located along the conductor to find the position of maximum reading.

A special technique is used on vehicles where the impedance of the power supply is usually quite low (CE07). In this case, the voltage on the line is measured directly through an appropriate decoupling capacitor. Two sizes are used, one from 0.15 to 30 MHz, the other 20 to 1000 MHz in order to avoid the possibility of series resonance because of lead length distributed inductance.

7-5.5.1.2 Signal Lines

Measurements on equipment signal and control lines (CE03, CE05) are made with the lines terminated as in normal use or with dummy loads where that is not practical. In order to separate wires in a sheathed bundle, a special unsheathed length of lead is constructed to enable the current probe to be inserted around appropriate groups of leads. Any group of leads should not contain a pair carrying the same signal in opposite directions.

7-5.5.1.3 Antenna Terminals

Receivers and transmitters with power outputs less than 5 kW having removable antennas are tested from 10 kHz to 12.4 GHz at the antenna terminals (CE06). The upper frequency limit depends on the operating frequency of the device under test. In general, a network to reject the transmitter designed output frequency must be inserted in front of the measuring instrument in key-down position. At frequencies below 100 MHz the output can be measured directly across a dummy load. At frequencies above 100 MHz, directional couplers are used between the transmitter and dummy load or actual antenna to avoid effects of standing waves in the transmission line.

7-5.5.2 Radiation Measurements

Magnetic field H emission components usually are made at frequencies from 30 Hz to 30 kHz, electric fields E from 10 kHz to about 15 GHz. Above about 100 MHz antennas may respond to the electric field,

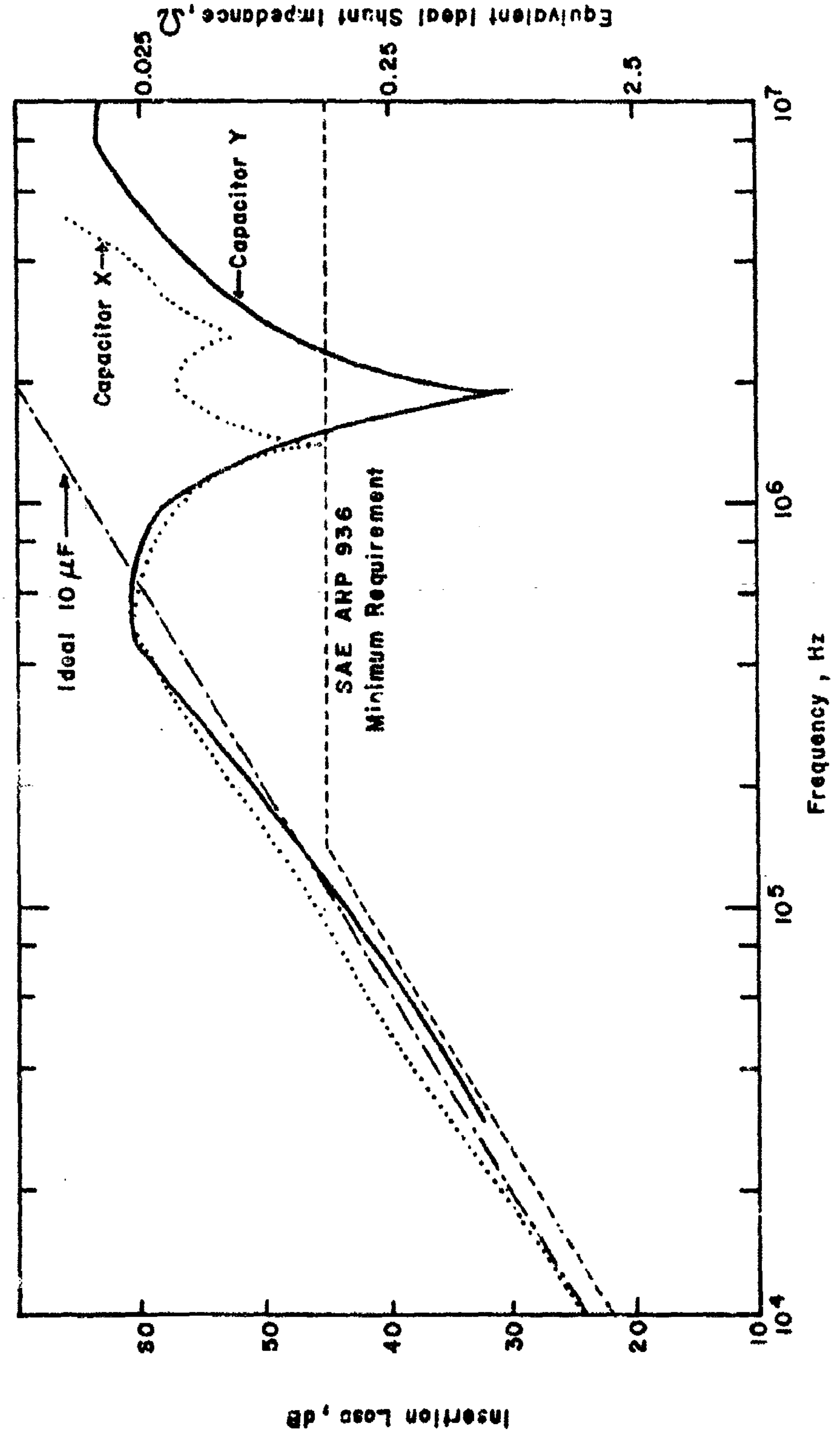
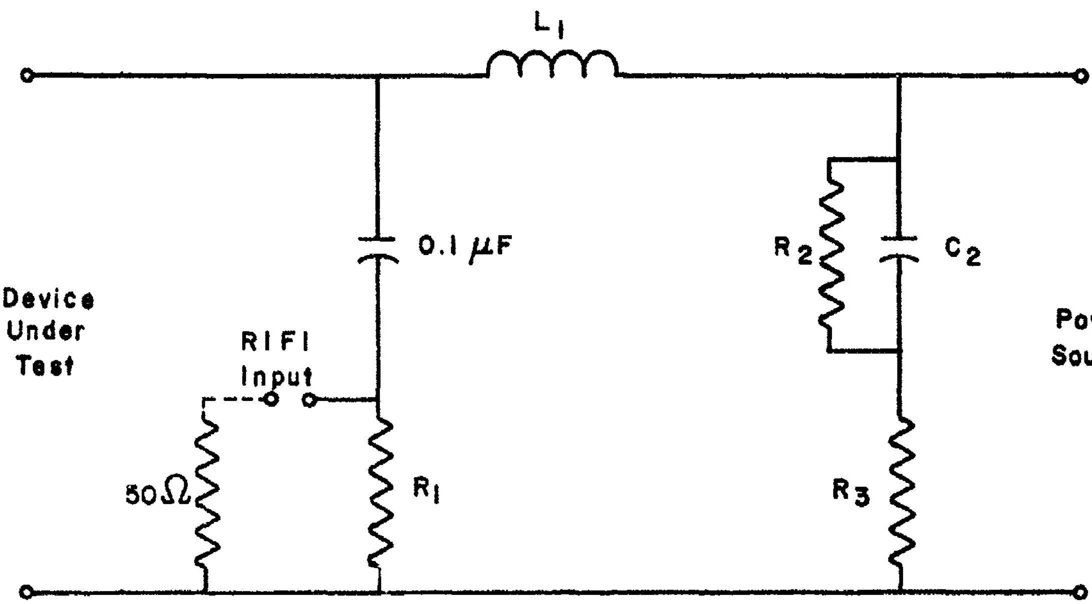


Figure 7-24. Insertion Loss of 10-μF Feedthrough Capacitors



Network (A)

L_1	56 µH
C_2	20 µF
R_1	1 kΩ
R_2	25 kΩ
R_3	1 Ω

Network (B)

L_1	5 µH
C_2	1 µF
R_1	5 kΩ
R_2	∞
R_3	0

Figure 7-25. Line Impedance Stabilization Networks

magnetic field, or both. Some caution should be observed in using these guidelines too rigidly. As has been noted (Ref. 39), in some cases interference control measures that have reduced values of field measured with an electric antenna actually have increased the interference. This can happen in an induction field region if a capacitor is applied to a circuit which reduces the potential across it (and also the associated electric field) but increases the current (and the associated magnetic field). This can occur in the frequency range from 10 kHz to 30 MHz. Furthermore, even at distances where the radiation field relations would be expected to apply, because of reflections giving rise to standing waves, measures of E and H would not be correlated.

7-5.5.2.1 Magnetic Field

This measurement is designed to protect devices and cables located close to one another. A loop of diameter 13.3 cm is located 7 cm from the cable or device enclosure, and moved about to obtain the position and orientation for maximum measured level (RE01). For some purposes it is conventional to express the level of magnetic field in terms of the equivalent value of electric field E for a plane wave; in other words (par. 7-3.3.2.2.1)

$$E = 377\,H\ ,\ \mathrm{V/m} \qquad (7\text{-}50)$$

where H is the magnetic field strength, A/m.

In another procedure (RE04) a loop or other form of magnetic loop sensor is placed at a distance of 1 m. Because of the increased distance, a lower field level is measured and also the measured level may be less sensitive to the exact position of the sensor. For purposes of prediction, the measurement at a distance of 1 m will provide the best basis for estimating the field at other distances (according to the inverse cube law, distance measured to center of radiation, see par. 3-3.2.1.1) except for distances close in, say less than 20 cm, where the 7-cm measurement distance is probably optimum.

7-5.5.2.2 Electric Field

7-5.5.2.2.1 Nonantenna Emitted

Measurements are made with an appropriate antenna at distances of 1 m for most equipment (RE02) but up to 50 m for transmission lines (RE06). Portable equipment and the measuring antenna are located at a height of approximately 1 m above a ground plane, which must underlie both the equipment under test and the measurement antenna (see Fig. 7-26). Note that the EMI meter is also located on the ground plane as are the line impedance stabilization networks which are used with the device under test. Furthermore the length of power line leads between the LISN's and the device under test is restricted, since these leads can contribute to the measured field. Antenna or other output terminals are connected to shielded dummy loads.

In the case of nonportable equipment, it is placed directly on the ground plane and the antenna counterpoise extended to join the ground plane as shown in Fig. 7-27.

Special techniques are necessary for the measurement of vehicles. In one test procedure (Ref. 40) the measurement distance is 10 m and both vertical and horizontal components of the field are measured. In method RE05, which applies to tactical vehicles and certain special purpose engine driven equipment, the distance is 2 m from the perimeter of the vehicle, and adjusted in height from 1 to 2 m for maximum. The biconical antenna, used between 25 and 200 MHz, is placed alternately to receive both horizontal and vertical polarization. Where the engine compartment has a top opening, the antenna is placed over the opening with the direction of maximum sensitivity directed towards the opening.

Figs. 7-28 and 7-29 show typical testing arrangements inside a shielded enclosure. Fig. 7-30 indicates the adaptation of the general procedures to a special situation.

7-5.5.2.2.2 Antenna Radiated Emissions

This technique is used where the output power is high or it is inconvenient to measure output power with the antenna disconnected. A receiving antenna, which may be directive, is placed at a convenient distance which will provide adequate sensitivity. At frequencies where aperture antennas are used, the separation should be adequate to insure operation in the far field of both antennas so that the antenna gains will be as close as possible to their calibrated values. Eq. 5-38 can be rewritten:

$$P_t = \frac{(4\pi r)^2 P_r}{\lambda^2 G_t G_r}, \quad \text{W} \tag{7-51}$$

where

P_t = power transmitted, W
P_r = power received, W
G_t and G_r = respective gains of transmitting and receiving antennas, dB
r = separation distance, m
λ = wavelength, m

and applies, provided that,

$$r > \frac{2D^2}{\lambda}, \quad \text{m} \tag{7-52}$$

where

D = diameter of larger antenna, m

and $D \gg d$, the diameter of the smaller antenna. If $d \geq 0.4D$, MIL-STD-462, Notice 3, RE03 requires

$$r = \frac{(D + d)^2}{\lambda}, \text{ m} \tag{7-53}$$

Antenna gain corrections that can be used if the far field distance is not used are given in par. 5-10.6.2. RE03 requires equipment sensitivity which will detect a spurious signal which is 80 dB below the fundamental transmitted power at a level which is at least 10 dB above the measuring receiver internal noise level.

During a measurement, the measuring antenna is adjusted in direction and elevation to maximize the measured level. Because of the presence of a high level desired output signal, it may be necessary to (1) place a rejection filter at the input to the measuring receiver to limit its value, and (2) place the measuring receiver in a shielded enclosure.

7-5.6 SUSCEPTIBILITY

In testing for susceptibility, voltages or fields are impressed on the device under test. The level is

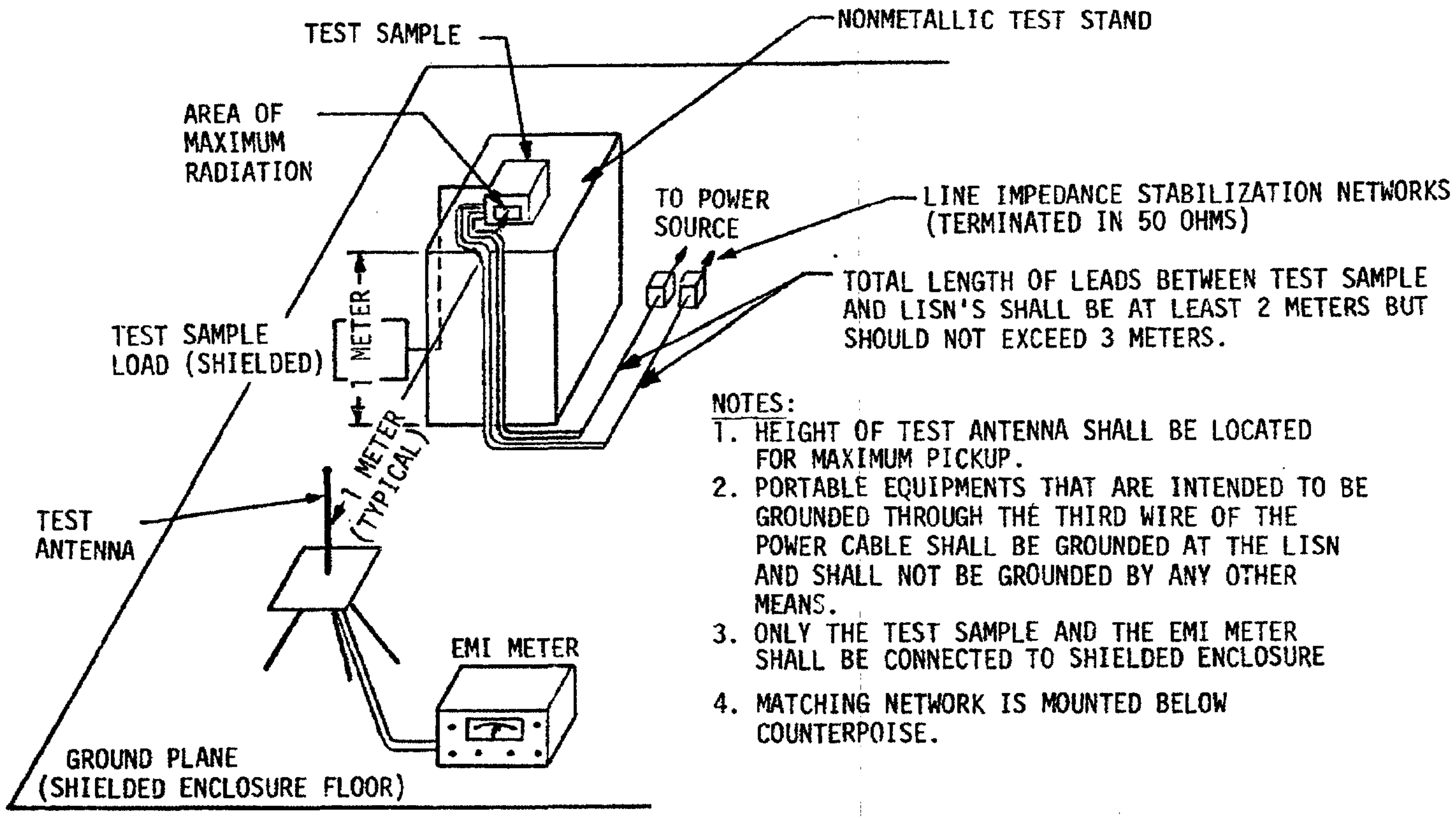
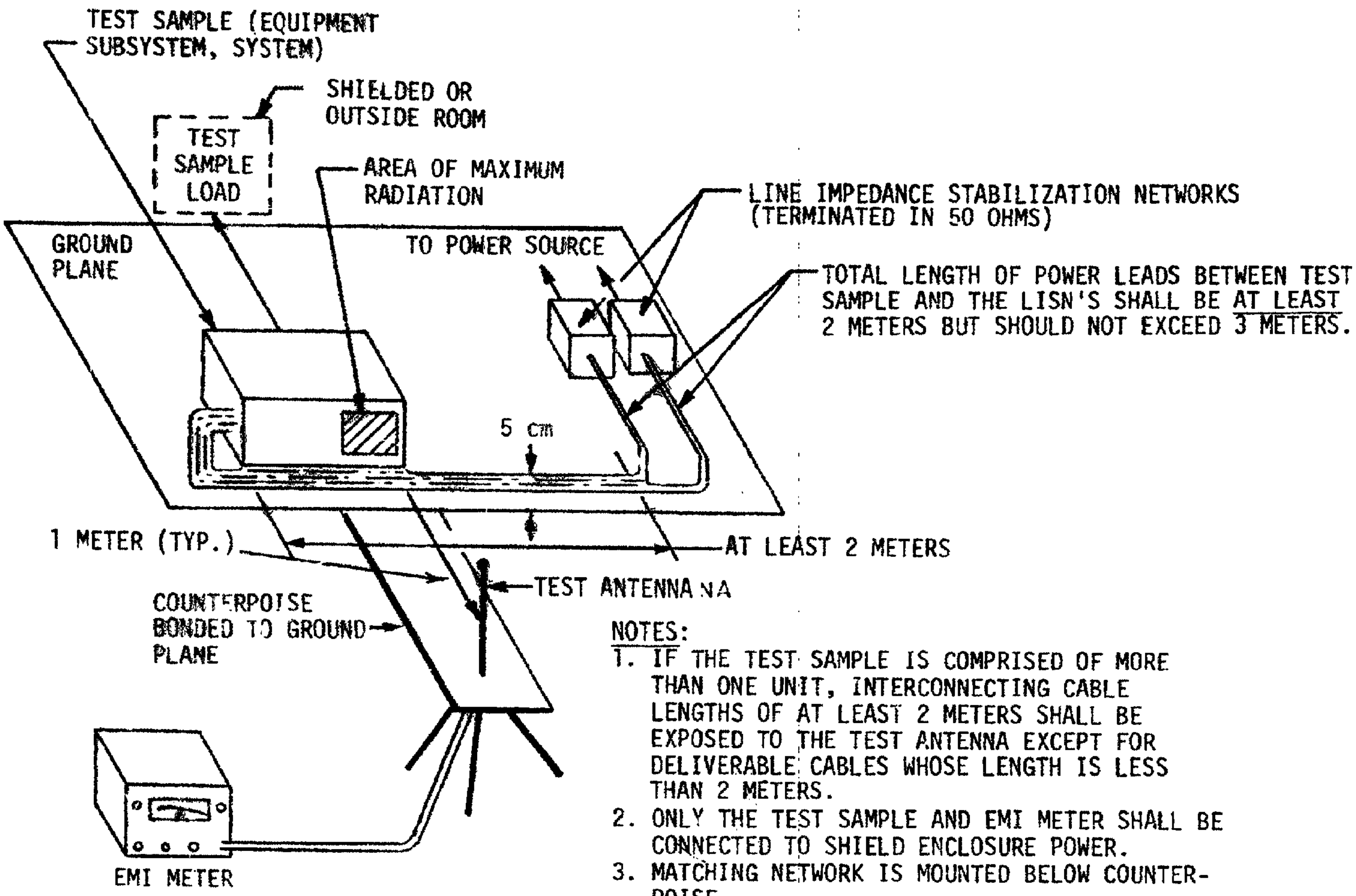


Figure 7-26. Typical Test Setup for Radiated Measurements on Portable Equipment



NOTES:

1. IF THE TEST SAMPLE IS COMPRISED OF MORE THAN ONE UNIT, INTERCONNECTING CABLE LENGTHS OF AT LEAST 2 METERS SHALL BE EXPOSED TO THE TEST ANTENNA EXCEPT FOR DELIVERABLE CABLES WHOSE LENGTH IS LESS THAN 2 METERS.
2. ONLY THE TEST SAMPLE AND EMI METER SHALL BE CONNECTED TO SHIELD ENCLOSURE POWER.
3. MATCHING NETWORK IS MOUNTED BELOW COUNTER-POISE.

Figure 7-27. Typical Test Setup for Radiated Measurements

Figure 7-28. Typical-Case Radiation Test Setup (Whip Antenna)

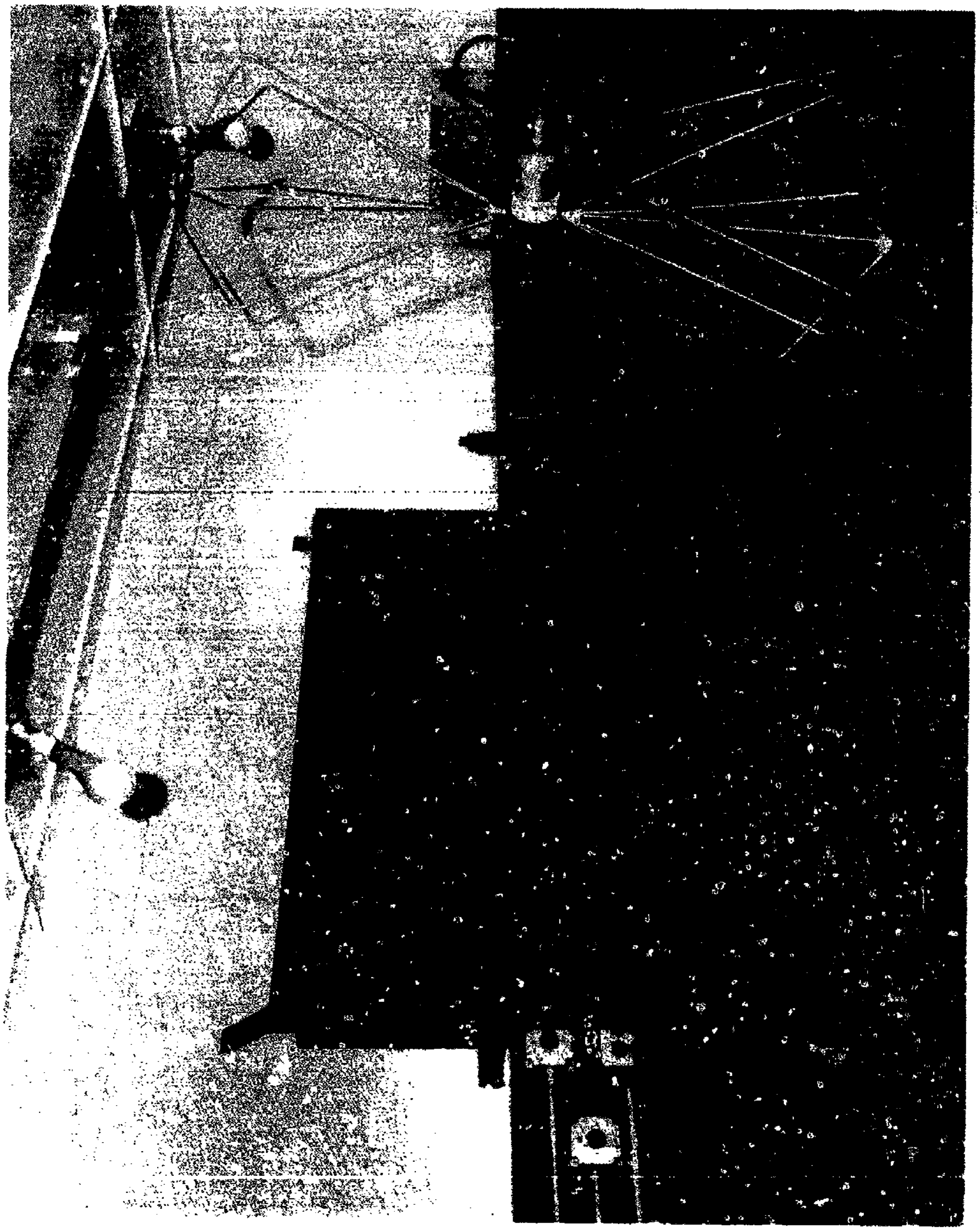


Figure 7-29. Typical Case Radiation Test Setup (Biconical Antenna)

Figure 7-30. Special-purpose Test Setup (Crane Motor)

gradually increased or decreased until a certain criterion is met. If the impressed voltage or field is at a frequency which enables direct detection, the criterion is usually a certain ratio of observed output (either audible or visible) due to the internal noise level. In some cases the observation can be made at a high ratio of observed signal-to-noise, say 20 dB, and the input level at another ratio, say 6 to 10 dB, calculated from it. In other cases the measurement may be made with a receiver bandwidth having one value and corrected for a standard bandwidth having another value.

Corrections for level are made proportionally. Corrections for bandwidth are made in accordance with the discussion in par. 7-3.1.2. For receiver sensitivity the correction is according to the square root of the bandwidth.

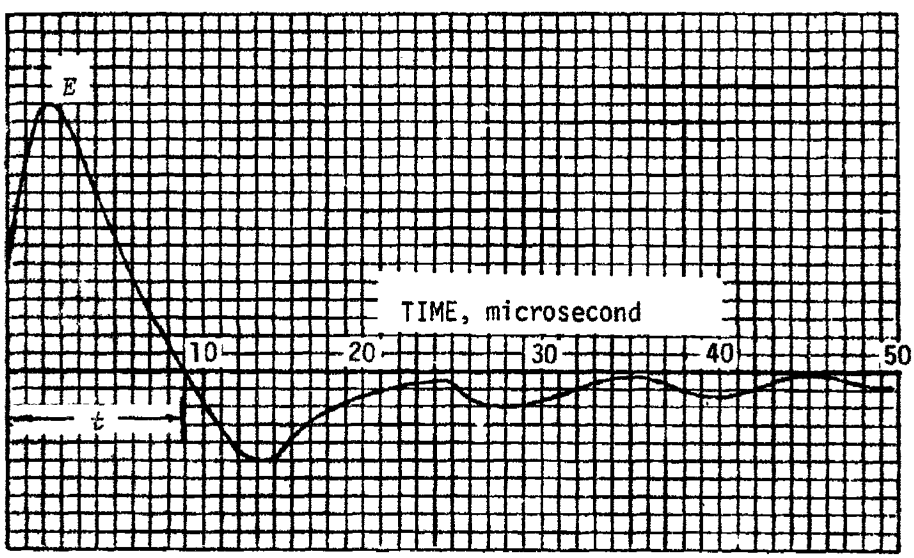
In those tests in which it is necessary to have an impressed desired signal in order to detect susceptibility to an undesired emission, the level and modulation of the desired signal can have significant effects upon the result. In most cases they should be set at the conditions approximating minimum performance, as specified in the procurement documents for the equipment under test.

For broadband tests one form of standard impulsive waveform is that shown in Fig. 7-31 which has an amplitude up to 100 V and a duration of the main transient of 10 μs. In special cases, pulses of larger magnitude (up to about 1000 V) and shorter duration (down to 10, or even 1, ns) may be appropriate.

7-5.6.1 Conducted Susceptibility

7-5.6.1.1 Power Line Susceptibility

Sine wave susceptibility can be measured by coupling a voltage in series with one of the power lines. In the case of dc lines, the voltage is coupled directly in series with the high line (CS01); otherwise it is applied to the 50-ohm terminal of the LISN connected into the circuit as for emission tests (CS02). In either case the injected current or voltage can be monitored with a current probe or tuned voltmeter placed line to line.



E = 2 times line voltage or 100 V, whichever is less

t = 10 μs

Figure 7-31. Spike Characteristics

For broadband tests the sine wave generator is replaced with an impulse generator, and the voltmeter is replaced with a calibrated oscilloscope, which is preceded by a line frequency rejection filter. The circuit is shown in Fig. 7-32.

7-5.6.1.2 Intermodulation

Because the signal sources used in intermodulation testing are themselves subject to nonlinear effects which may affect the measured result, special care is necessary in the experimental arrangement. Commonly, filters and attenuators are used as shown in Fig. 7-33 (CS03).

In this arrangement, the filters are tuned to the respective signal generator frequencies and in each case prevent frequencies from the other generator or from the receiver from affecting their output. Furthermore, they attenuate harmonics of the desired signal generator. The coupling network should have at least 20 dB insertion loss between signal generator terminals.

In use, one of the signal generators is modulated and the frequencies of each are adjusted according to the order of intermodulation of interest, as shown in Table 7-4 (see par. 3-2.2.2). One of the frequencies, say f_1, is usually adjusted as close to f_o as possible without producing a response at the output by itself. Then f_2 is adjusted in frequency and amplitude until a maximum intermodulation response is obtained. f_1 may be selected both above and below f_o. Normally, the 3rd and 5th order are the most significant, since for those orders both f_1 and f_2 are close to f_o. The intermodulation rejection is the difference between the level in dB to which the outputs of the two generators are adjusted and the level of one of the modulated generators when tuned to the receiver frequency when adjusted to provide standard output of the receiver, say $(S + N)/N = 10$ dB.

Care should be exercised to make sure the desired intermodulation response is being observed and not a spurious output of either generator or spurious response of the receiver. If the correct response is

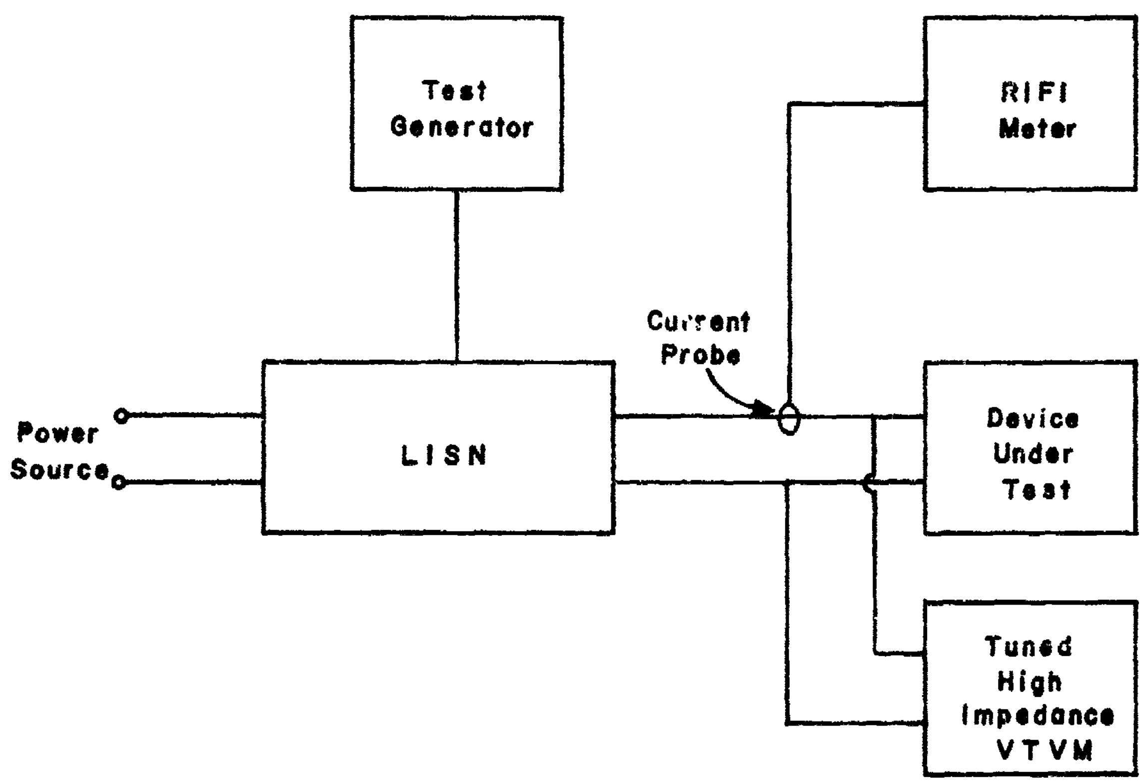


Figure 7-32. Power Line Conducted Susceptibility Tests Using LISN

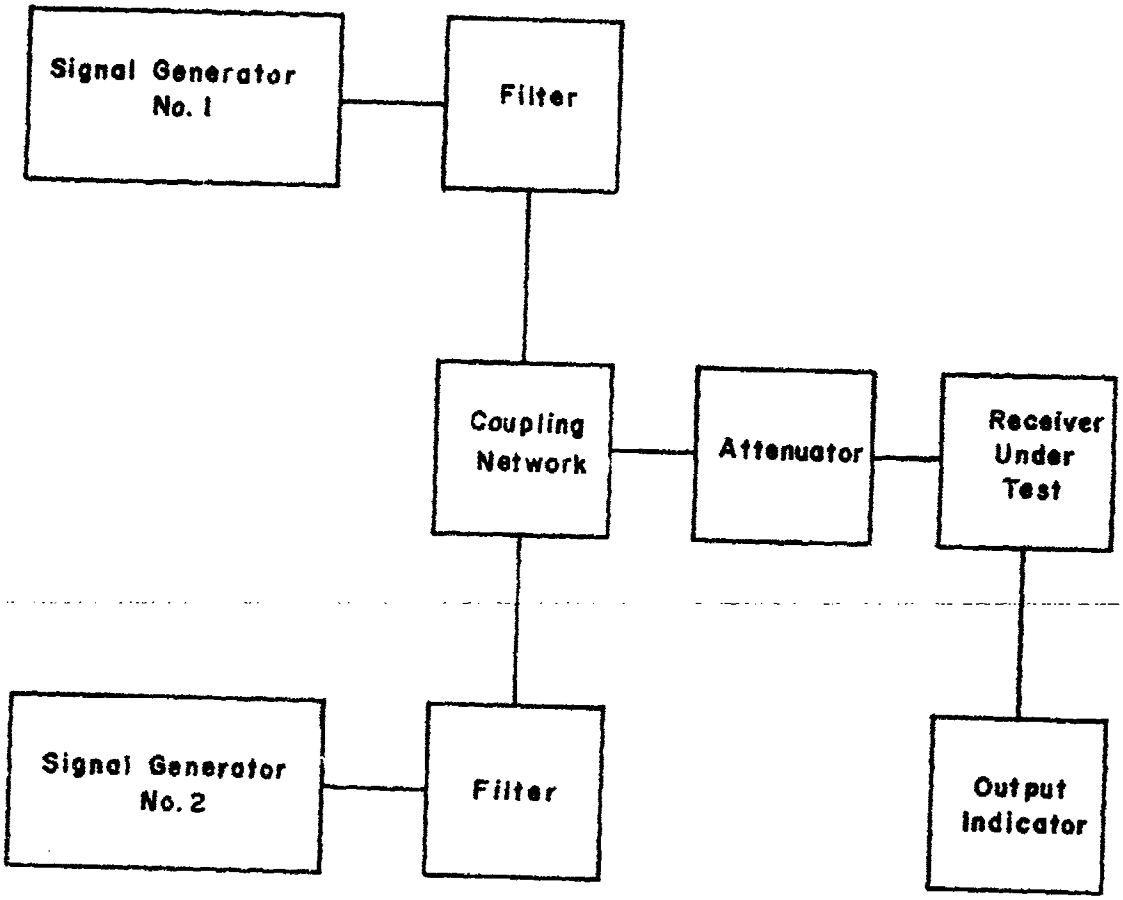


Figure 7-33. **Intermodulation Measurements**

TABLE 7-4.
TEST FREQUENCY RELATIONSHIPS VS MODULATION ORDER

Order	Relationship
2	$f_o = f_1 - f_2$
3	$f_o = 2f_1 - f_2$
4	$f_o = 2f_1 - 2f_2$
5	$f_o = 3f_1 - 2f_2$

being measured, a change in attenuation setting will not produce a linear change in the observed output level.

7-5.6.1.3 Cross Modulation

Cross modulation can be measured with the same equipment as is used for intermodulation. Signal generator (S.G.) No. 1 is tuned to the same frequency to which the receiver under test is tuned and its output is adjusted to produce an appropriate level at the receiver input, say 10 or 20 dB above the receiver sensitivity level. The interfering S.G. No. 2 is tuned to a frequency near the tuned frequency but at least two receiver bandwidths away from it. Initially with S.G. No. 2 turned off, S.G. No. 1 is modulated 30% at 400 or 1000 Hz and the output level control adjusted for a normal output level. Then with S.G. No. 1 unmodulated. the output of S.G. No. 2 with 30% modulation is adjusted for the same level in the output of the receiver as obtained with S.G. No. 1 alone. The level of S.G. No. 2 at the input of the receiver is noted. The test is repeated for other frequencies of S.G. No. 2 on both sides of the receiver tuned frequency. The results are plotted in terms of microvolts (on a logarithmic or dB scale) vs frequency. The test can be repeated at other output levels from S.G. No. 1 (see Ref. 41).

7-5.6.1.4 Spurious Response and Desensitization

Undesired receiver responses can be detected by applying a signal generator at the input and scanning the frequency range of interest (CS04). The signal generator should not have spurious output frequencies or they should be filtered as in Fig. 7-33. By adjusting the attenuator, one can distinguish between generator and receiver spurious since receiver spurious responses are not linearly related to input level.

To make quantitative measurements, a second signal generator adjusted to the receiver tuned frequency at a level which produces standard output can be applied at the input (see Fig. 7-33). The level of the first generator which causes noticeable change in the standard output level is the spurious response or desensitization level.

7-5.6.2 Radiated Susceptibility

As with emission measurements, radiated susceptibility testing is done with a magnetic field source at frequencies up to 30 kHz and with an electric field source at 14 kHz and above. The strip line techniques enable combined electric and magnetic field testing.

7-5.6.2.1 Magnetic Field

Usually the field is generated by a loop (12 cm diameter in RS01) which is used to scan the surface of an equipment to locate the point of maximum sensitivity. The field H along the axis of a concentrated loop of N turns can be calculated from (Ref. 42)

$$H = \frac{NIa^2}{2(a^2 + r^2)^{3/2}}, \text{ A/m} \qquad (7\text{-}54)$$

where

I = current, A
a = loop diameter, m
r = distance along the axis from the center of the loop, m

The magnitude of the field from a particular coil can be established either by measurement of the field at a specified distance using an appropriate magnetic sensor or by calculation in terms of the coil current, if the construction is sufficiently precise. The current through the coil can be measured by a current probe and tuned voltmeter or by measuring the voltage across a resistor in a series with the coil.

A pair of identical coils parallel to each other in a Helmholtz arrangement on the same axis and spaced by a distance equal to the radius of each coil will provide a quite uniform field

$$H = \frac{8NI}{125a}, \text{ A/m} \qquad (7\text{-}55)$$

7-5.6.2.2 Electric Field

The fields required may be established in several ways. Where space is available a standard antenna such as described in par. 7-3.3.2 can be fed from a signal generator of sufficient power to generate the field required at the test distance. The field value obtained should be checked with an antenna and receiver combination of known sensitivity, with the antenna located at the test position. Since, for large equipments, the field strength may vary significantly from one point to another, it may be necessary to change the relative location of the equipment and the source antenna during the test to insure that all parts are exposed to the same field strength.

7-5.6.2.3 Transmission Line Techniques

Two types of transmission line techniques have been used. In one a wire is stretched between opposite sides of a shielded enclosure (RE03). The enclosure then forms the outer conductor of a simulated coaxial structure. When the line is properly terminated, standing waves on the line are avoided and the ratio of the radial electric to the circumferential magnetic field strengths approximates the characteristic impedance of free space, at least close to the wire. The fields can be calculated in terms of the measured current I in the wire. Because of the effect of the walls, corrections must be applied to the calculated fields. One obtains:

$$E = 60I\left(\frac{1}{d} + \frac{1}{2d_1 - d} - \frac{1}{2d_2 + d}\right), \text{ V/m} \qquad (7\text{-}56)$$

where

I = current, A
d, d_1, d_2 = distances as shown on Fig. 7-34, m

The parallel plate transmission line is shown on Fig. 7-35. Except for fringing effects at the edges of the line, the electric field component is normal to the plates and the magnetic field parallel. When the line is properly terminated, the ratio of electric to magnetic field strengths is also that of free space or 377 ohms.

The upper frequency at which both these lines are usable is limited by the discontinuities in electrode shapes at the line ends. Generally, the parallel plate

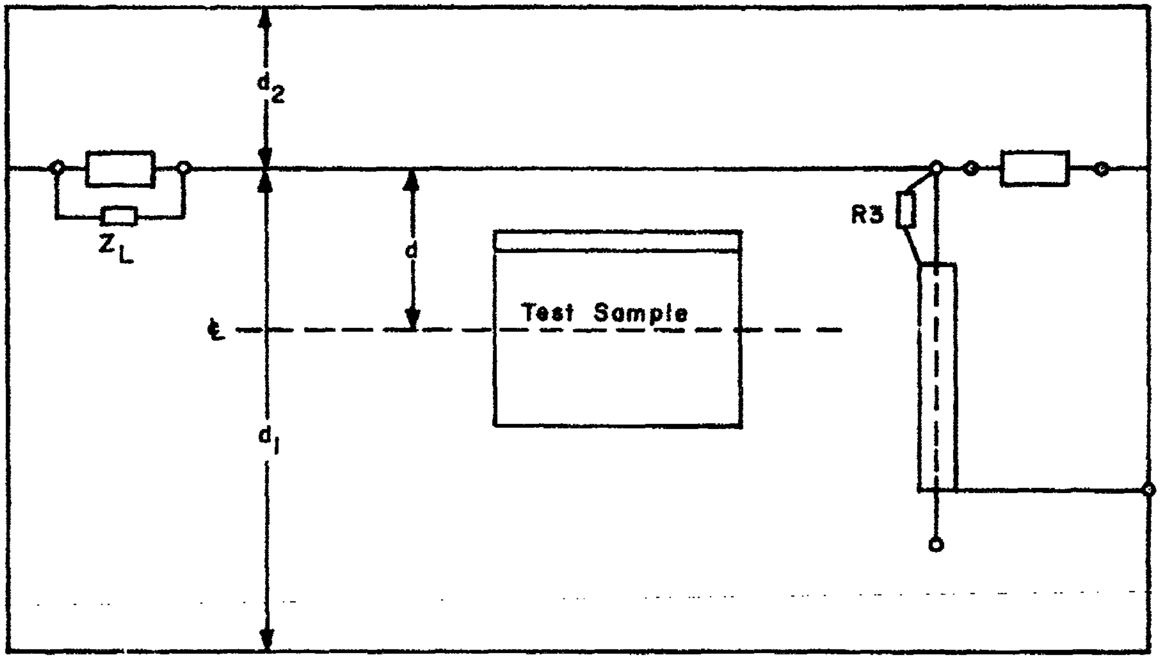


Figure 7-34. Long-wire Transmission Line

line can be used to about 50 MHz and the wire line to somewhat less.

A variation of the parallel plate line construction is that of the fully enclosed parallel plate line, or "TEM Cell" (Ref. 43) in which a center is completely surrounded by the outer enclosure. Fig. 7-36(A) shows the external appearance of the line with tapered sections at each end. Fig. 7-36(B) shows a cross section with the plate in the center. A uniform field test region is available both above and below the center plate. Because this line is fully enclosed no leakage field is produced, and therefore it is easier to match with the tapered sections than is the two-parallel-plate line. With this construction it is easier to obtain satisfactory operation up to the point at which the spacing of the plates approaches 1/2 wavelength and moding occurs.

In all of the field susceptibility testing techniques, it should be observed that equipment can be expected to be sensitive to polarization. Therefore, the object under test should be repositioned during the test so as to expose it to the most effective field polarization.

7-5.7 SITE SURVEYS

Surveys of the electromagnetic environment at particular locations are made to determine either (1) the suitability of a particular location for a future military installation or (2) whether fields at an already existing installation (usually due to a recent installation of new equipment) are sufficient to degrade operations in the vicinity. Most often, they are performed on the ground, but also are carried out in aircraft. On the ground, it is most expedient to use a vehicle on which all equipment, antennas, and power supplies are self contained. Automatic frequency scanning and recording are desirable unless the tests being conducted are of a specialized nature.

The requirements on characteristics of equipment used follow.

7-5.7.1 Frequency Range

Surveys can be conducted over the range 10 kHz to 40 MHz. For particular purposes, such as for plotting the fields from a particular transmitter or other source, only its fundamental frequency, its harmonics, and other expected spurious frequencies need be covered. If the purpose of the survey is to determine the suitability for installation of a particular receiver, frequencies to which the receiver may be susceptible should be surveyed. Also frequencies at which extraordinarily high field strengths exist due to local transmitters should be measured.

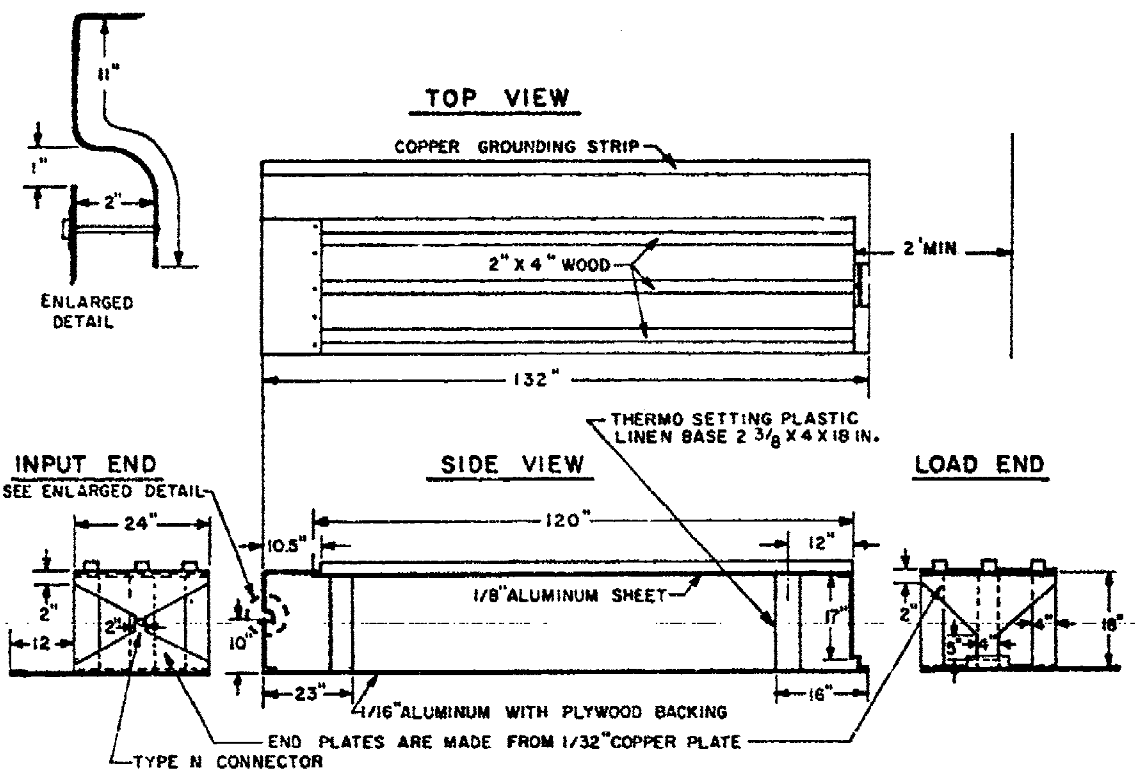


Figure 7-35. Parallel Strip Line for Radiated Susceptibility Tests (Top and Side View)

7-5.5.2 Sensitivity

For general purposes receiver sensitivity should be equal to the state of the art for calibrated broad tuning range receivers, i.e., a few dB at frequencies up to the tens of megahertz increasing to about 10 dB at 1 GHz. Lesser sensitivities would be satisfactory for surveys in connection with radiation hazards.

7-5.5.3 Antennas

Antennas should be broadly tunable to minimize the number required to cover the frequency range and for most purposes should be as nondirectional as required sensitivities permit. At frequencies up to 30 MHz both a vertical rod and loop can be used. The latter should be rotatable in its vertical plane.

Above 30 MHz the antennas should be plane polarized and orientable for both horizontal and vertical polarization.

Furthermore, the height of the antenna should be adjustable, preferably up to a height of 10 m so that points of maximum field strength can be identified.

7-5.5.4 Locations of Antennas

In general, test locations should be chosen as far as possible from reflecting objects. However, if the purpose of the survey is to determine hazards, numerous test locations — including some close to reflecting objects — may be appropriate. In such cases it is desirable to choose sufficient locations to enable plotting contours of constant field strength (from a particular source). The number required depends on the frequency of radiation of interest, being greater at the higher frequencies. For a detailed survey, measurement locations may be spaced as closely as every quarter wavelength or less.

On large sites that are distant from local sources, only a few measurement locations may be necessary.

7-5.5.5 Data Recording

Peak, average, and rms detectors are useful in survey work. Normally signals are measured with the average detector, and atmospherics are measured with the rms detector although sometimes both the

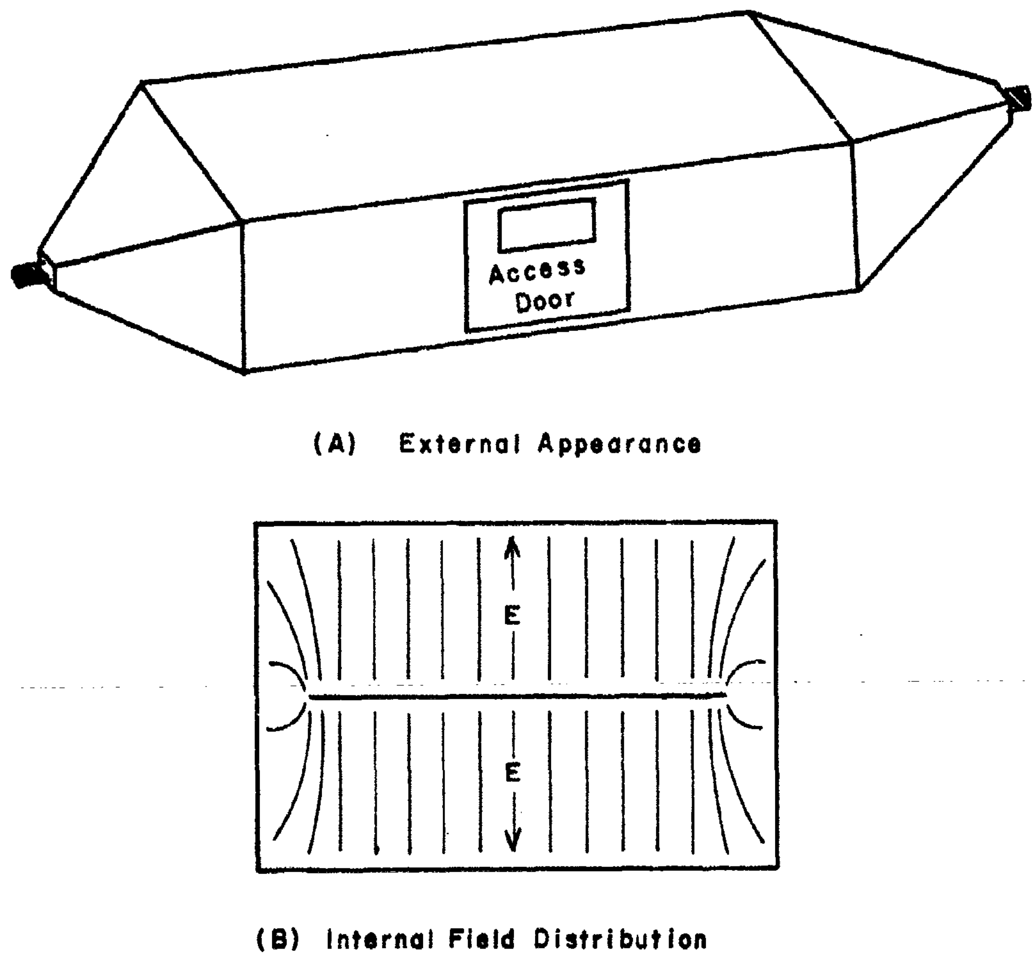


Figure 7-36. Enclosed Parallel Plate Line

rms and average are measured (see par. 3-1.2.2.2). The peak detector commonly is used for measuring man-made radio-noise, especially when it is intermittent, or highly impulsive. In all cases it is important to record the acceptance bandwidth of the measuring receiver so that corrections can be made in the effective level for receivers having different bandwidths. In this respect, measurements made on more than one detector may be helpful. In addition, measurements of variations in measured level with time should be recorded, if test time permits, along with antenna height.

7-5.5.6 Calibration

Instruments and antennas should be calibrated in accordance with pars. 7-3.1.3 and 7-3.3.3.

7-6 EMC TEST FACILITIES

Although "in-house" facilities may be used extensively in carrying out the test procedures discussed in this chapter, there may be times when the use of outside facilities may be advisable, especially if large numbers of tests and special facilities are required. Frequently, private test laboratories are available. In addition a number of Military Laboratories have set up test facilities with a variety of capabilities (Ref. 44). These may be available to individual contractors, especially in the case of unusual requirements, and are described briefly. Table 7-5 gives points of contact for each facility.

7-6.1 ARMY FACILITIES

7-6.1.1 Electromagnetic Interference Test Facilities

Location: US Army Electronics Command, Fort Monmouth, NJ

Instrumentation is available for implementing testing requirements of MIL-STD-461 and -462.

7-6.1.2 Materiel Testing Directorate

Location: Aberdeen Proving Ground, MD

EMI/EMC tests in accordance with MIL-STD-461 and other specifications can be performed at frequencies up to 40 GHz. Of three automated spectrum surveillance systems, one is mobile and used for the testing of noncommunication-electronic equipment including track and wheel vehicles, watercraft, air conditioners, and other engine drive equipment. The second system is transportable and is used for tests on communication-electronic equipment as well as noncommunication-electronic equipment. The third system is permanently installed in a screened room.

An EMI shielded enclosure, 94 ft × 60 ft × 28 ft, is used for EMI/EMC tests of large vehicles; i.e., tanks, tractor-trailers, shelters, construction and material handling equipment, and engine-generator sets.

7-6.1.3 Electromagnetic Environmental Test Facility (EMETF)

Location: US Army Electronic Proving Ground, Fort Huachuca, AZ

The EMETF has six interrelated experimental capabilities as follows:

a. The Instrumented Workshop (IWS)/Automatic Data Collection System contains facilities for conducting performance tests of electronic equipments under controlled electromagnetic interference conditions to investigate interference mechanisms, furnish information for designing interference suppression devices, establish intelligibility measurements, provide data for the further development of the Interference Prediction Model (IPM) and the development of methods of scoring the performance of Communication-Electronics (C-E) systems. (See par. 6-4.3)

b. The Weapon System Electromagnetic Environment Simulator (WSEES) is comprised of a computer programmed RF generator, and RF absorption room, a steerable platform, and other support equipment necessary to simulate complex electromagnetic environments and develop scoring data for systems operating in the microwave frequency bands, including radar systems.

c. The Systems Scoring Facility (SF) consists of two listener facilities (one mobile, one fixed) designed to measure the performance of voice communication systems used tactically by the Army in the field.

d. The Digital Scoring System (DSS) enables rapid evaluation of a variety of digital communication links and minimizes human intervention in obtaining the data needed.

e. The Spectrum Signature Facility (SSF) can provide measurements in accordance with MIL-STD-449.

f. The Field Facility (FF) conducts EMC tests of large equipments, antenna measurements, and any tests that require open-field measurements or that cannot be contained in the IWS.

7-6.1.4 Electromagnetic Radiation Effects Test Facility

Location: White Sands Missile Range, NM

Equipments are available with power output up to 50 kW to produce high intensity (200 V/m) electromagnetic fields at frequencies from 100 kHz through 18 GHz.

7-6.1.5 Missile Electromagnetic Effects Test Facility

Location: Redstone Arsenal, AL

This missile electromagnetic effects (EME) test facility is designed for rapid missile test and evaluation. Transmitter frequency coverage is from 100 kHz to 18 GHz with power levels from 1 kW to 10 kW. Transmitters are servo tuned for rapid and remote control in the 100 kHz to 350 MHz region. Seven fixed broadband transmitters are used above 350 MHz. The facility includes a dedicated computer which can provide transmitter test control, data acquisition, and data analysis.

The EME test facility can interface with the Army Missile Command Advanced Simulation Center (ASC) to evaluate missile hardware during simulated missions. Missile testing utilizing the ASC in conjunction with EME testing provides precise missile hardware performance evaluation.

TABLE 7-5. POINT OF CONTACT LIST

Parent Agency	Facility	Parent Organization	Address
Army	EMETF (Electromagnetic Environmental Test Facility)	US Army Test and Evaluation Command	Cdr US Army Electronic Proving Ground, Attn: Code STEEP-NT-MP, Fort Huachuca, AZ 85613
Army	EM Interference Test Facility	USAECOM	Cdr US Army USAECOM, Attn: AMSEL-GG-I Fort Monmouth, NJ 07703
Army	Materiel Testing Directorate	US Army Test and Evaluation Command	US Aberdeen Proving Ground Attn: STEAP-MT-G Aberdeen Proving Ground, MD 21005
Army	Electromagnetic Radiation Effects Test Facility, White Sands Missile Range	US Army Missile Test Evaluation Command	White Sands Missile Range NM 83002
Army	Missile Electromagnetic Effects Test Facility	US Army Missile Command	Cdr US Army MICOM, Attn: AMSMI-RTE Redstone Arsenal, AL 35809
Navy	EMC Section, Electronic Systems Branch Systems Eng. Test Directorate Naval Air Test Center	Naval Air System Command	Cmd, Naval Air Test Center Attn: SY80 Patuxent River, MD 20670
Navy	System Test Facility	Naval Electronics Laboratory Center	San Diego, CA 92152
Navy	Technical & Environmental Evaluation Division	Naval Electronics Laboratory Center	271 Catalina Blvd. San Diego, CA 92152
Navy	Communications Systems Branch	US Navy Research Laboratory	Washington, DC 20375
Navy	Systems Integration & Instrumentation Branch	US Navy Research Laboratory	Washington, DC 20375
Navy	Naval Avionics Facility	Naval Air System Command	Dept. of the Navy, Naval Avionics Facility 6000 E. 21st Street Indianapolis, IN 46218
Navy	Naval Ship Engineering Center, Norfolk Div.	Naval Ship Eng. Center	Naval Ship Engineering Center, Norfolk Div. Naval Station Norfolk, VA 23511
Navy	Naval Electronic Systems Test & Evaluation Facility	Naval Electronic Systems Command	Patuxent River, MD 20670

(cont'd)

TABLE 7-5. (Cont'd)

Parent Agency	Facility	Parent Organization	Address
Navy	Naval Surface Weapons Center Dahlgren Laboratory	Naval Materiel Command	Dahlgren, VA 22448
Navy	Naval Underwater Systems Center T & E Facility (NESTEF)	Chief of Naval Materiel	Newport, RI 02840
Air Force	EM Interference & Compatibility Branch	Aeronautical Systems Division	Wright Patterson AFB, OH 45433
Air Force	Antenna Proving Range Newport, NY	Rome Air Development Center	Griffiss AFB, NY 13441
Air Force	EM Test Facility Verona, NY	Rome Air Development Center	Griffiss AFB, NY 13441
Air Force	EM Interference & Analysis Facility	Electronics System Division	Griffiss AFB, NY 13441
Air Force	Richards Gebaur AFB	AF Communications Service	Richards Gebaur AFB MO 64030

7-6.2 NAVY FACILITIES

7-6.2.1 Naval Air Test Center (NATC)

Location: Patuxent River, MD

Personnel and specialized test equipment are available to perform lightning and EMC tests of complete aircraft weapon systems and of airborne avionic/electrical equipment to determine compliance with MIL-E-6051 and MIL-STD-461.

The laboratory has a main shielded area approximately 300 ft by 140 ft by 66 ft at the highest ceiling point, and an adjoining nosebay shielded area approximately 100 ft × 62 ft × 30 ft. The lightning test facility contains a surge voltage generator and a surge current generator.

7-6.2.2 Naval Surface Weapons Center

Location: Dahlgren, VA

The Naval Weapons Laboratory has equipment and facilities used to study HERO and high power RF effects. Two ground planes, each 240 ft × 100 ft, of 3/8-ft welded steel, are equipped with 21 ft diameter turntables with a capacity of 40 tons.

Shielded RF laboratories up to 2300 sq ft and an anechoic chamber, 8 ft × 8 ft × 16 ft, are available. A mobile research van can measure EME in the 2-32 MHz frequency band.

7-6.2.3 Naval Electronic Laboratory Center (NELC)

Location: System Test Facility, San Diego, CA

The facility has the capability of conducting spectrum surveys of the RF environment and of the RF environment internal to communication systems from 14 kHz to 1 GHz and automatically plotting the results.

7-6.2.4 Naval Electronic Laboratory Center (NELC) Technical and Environmental Evaluation Division

Location: San Diego, CA

Measurement capability is principally in the area of EMI testing of components and equipments and in evaluation of EMI properties of shipboard electronic equipments against requirements of MIL-I-16910.

7-6.2.5 US Naval Research Laboratory

Location: Washington, DC

The Communications Sciences Division has instrumentation on hand for EMC/EMI measurements over the LF through UHF bands and includes enclosures in the security vault for critical TEMPEST measurements.

7-6.2.6 Naval Avionics Facility

Location: Indianapolis, IN

A shielded enclosure is available which can be adapted for shielding effectiveness tests on materials, devices, cables and connectors, and filters.

7-6.2.7 Naval Ship Engineering Center, Norfolk Division

Location: Norfolk, VA

Two instrumented vans capable of measurements up to 40 GHz are available. Measurements can be made of spectrum signature, EM ambient levels, antenna patterns, and limited MIL-STD-461 measurements.

7-6.2.8 Naval Electronic Systems Test and Evaluation Facility (NESTEF)

Location: Patuxent River, MD

This laboratory is set up for measurement of spectrum signatures and the performance of technical investigations in support of ECAC and site survey investigations.

7-6.2.9 Naval Underwater Systems Center (NUSC)

Location: Newport, RI

The Atlantic Undersea Test and Evaluation Center (AUTEC), located in the Andros Islands area, is capable of monitoring and obtaining a bearing on any electromagnetic radiation.

7-6.3 AIR FORCE FACILITIES

7-6.3.1 Electromagnetic Interference and Compatibility Branch

Location: Wright-Patterson AFB, OH

The facility can perform EMC measurements on aeronautical systems.

7-6.3.2 Antenna Proving Range

Location: Newport, NY

Four antenna ranges, one of length 7000 ft, are located in an essentially interference-free area and used to measure antenna performance parameters (down to +0.25 dB sidelobes with over 40-60 dB dynamic range at higher frequencies, 360 deg in azimuth, and 180 deg vertical cut for antennas with up to 60-ft aperture width). A dark room range (15 ft × 12 ft × 22 ft) with walls covered with absorbing material is used to test antennas (3 ft max dia, 20 lb max) with a lower frequency limit of 1 GHz.

7-6.3.3 Electromagnetic Test Facility

Location: Verona, NY

The facility supports engineering evaluation and operational testing of ECCM, radar, communications, millimeter wave research, and optical surveillance techniques.

7-6.3.4 Electromagnetic Interference and Analysis Facility

Location: Griffiss Air Force Base, NY

Anechoic chambers up to 54 ft × 18 ft × 18 ft in size contain equipment to generate up to 50-kW peak power. Antenna gain and pattern measurements can be performed on any antenna whose physical dimensions are less than 6 ft overall.

7-6.3.5 Air Force Communication Service (AFCS)

AFCS facilities have the capability of making measurement in the field and can be used to analyze, identify, and initiate corrective action to prevent electromagnetic interference world-wide on communication electronic-and-meterological (C-E-M) equipments.

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GLOSSARY

The terms associated with EMC/EMI have been developed over many years by persons working in several specialized technical areas. Hence, various terms have broad as well as specialized meanings and consequent differences in usage. The terms included here are those directly related to use in the text of this Handbook. In so far as practical, the definitions given are based upon the following sources:

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2. *Advance Edition of International Electrotechnical Vocabulary, Publication 50 (902): Radio Interference*, 1973, International Electrotechnical Commission, Geneva, Switzerland
3. *IEEE Std-100-1972, IEEE Standard Dictionary of Electrical and Electronic Terms*, The Institute of Electrical and Electronic Engineers, Inc., NY, 1972.

Adjacent channel interference. Interference in which the extraneous power originates from a signal of assigned (authorized) type in an adjacent channel.

Ambient level. The level of radiated and conducted energy existing at a specified location and time when the test sample is de-energized. Atmospheric noise, signals (both desired and undesired from other sources) and the internal noise level of the measuring instruments all contribute to the "ambient level".

Antenna. A device employed as a means for transmitting or receiving radiated electromagnetic energy.

Antenna effective area. In a given direction, the ratio of the power available at the terminals of an antenna to the power per unit area of a plane wave incident on the antenna from that direction, polarized coincident with the polarization that the antenna would radiate.

Antenna effective length. The ratio of the antenna open circuit induced voltage to the strength of the field component being measured.

Antenna factor. That factor which when multiplied by the voltage at the input terminals of the measuring instrument, yields the electric field strength in volts/meter or the magnetic field strength in amperes/meter. This factor can include the effects of antenna effective length, and mismatch and transmission line losses.

Antenna gain. The ratio of the power required at the input of a reference antenna to the power supplied to the input of the given antenna to produce, in a given direction, the same field at the same distance. When not otherwise specified, the gain figure for an antenna refers to the gain in the direction of the radiation main lobe. In services using scattering modes of propagation the full gain of an antenna may not be realizable in practice and the apparent gain may vary with time.

Antenna induced voltage. The voltage which is measured at, or calculated to exist across, the open-circuited antenna terminals.

Antenna, isotropic. A hypothetical antenna that radiates or receives energy of all polarizations equally well in all directions. An isotropic antenna is a lossless point source used as the theoretical reference in describing the absolute gain of a real antenna.

Antenna pattern. A graph of the radial component of the Poynting vector at a constant radius as a function of the azimuthal or elevation angle.

Antenna, phased array. An array antenna whose beam direction or radiation pattern is controlled primarily by the relative phases of the excitation coefficients of the radiating elements.

Antenna terminal conducted interference. Any undesired voltage or current generated within a receiver, transmitter, or their associated equipment appearing at the antenna terminals.

Balanced circuit. A circuit in which two branches are electrically alike and symmetrical with respect to a common reference point, usually ground. *Note:* For an applied signal difference at the input, the signal relative to the reference at equivalent points in the two branches must be opposite in polarity and equal in amplitude (*See: Differential mode interference*).

Bandwidth, impulse. The peak value at the output of the circuit involved divided by the area of the impulse response envelope.

Bond, direct. An electrical bond in which the two metal surfaces are placed in intimate contact.

Bond, electrical. A low resistance current path between two metallic surfaces (normally between a metallic device and a metal surface at ground reference potential).

GLOSSARY (cont'd)

Bond, indirect. An electrical bond in which two surfaces are connected by a bond strap or other conductor.

Common-mode conversion. The process by which differential-mode voltages (or currents) are produced in a signal circuit by common-mode voltages (or currents) applied to the circuit.

Common-mode interference. Interference that appears between both signal leads and a common reference plane (ground) and causes the potential of both sides of the transmission path to be changed simultaneously and by the same amount relative to the common reference plane (ground).

Common-mode rejection ratio. The ratio of the common-mode interference voltage at the input terminals of the system to the effect produced by the common-mode interference, referred to the input terminals for an amplifier. For example,

$$CMRR = \frac{V_{CM}\text{ (root-mean-square) at input}}{\text{effect at output/amplifier gain}}$$

Compatibility, intersystem. The ability of electronic or electromechanical equipments or systems to operate in an electromagnetic environment without causing unacceptable degradation or malfunction to surrounding equipments or systems.

Compatibility, intrasystem. The ability of electronic or electromechanical components or subsystems to operate in their intended operational environment without suffering or causing unacceptable degradation or malfunction to related components or subsystems.

Conducted interference. Undesired electromagnetic energy which is propagated along a conductor.

Coupling. The effect of one system or subsystem upon another. (A) For interference, the effect of an interfering source on a signal transmission system. (B) The mechanism by which an interference source produces interference in a signal circuit.

Coupling, capacitive. That type of coupling in which the mechanism is capacitance between the interference source and the signal system; i.e., the interference is induced in the signal system by an electric field produced by the interference source. Capacitive coupling sometimes also is called *electric coupling.*

Coupling, conductive. That type of coupling in which the mechanism is conductance between the interference source and the signal system.

Coupling, inductive (or induction field coupling). The type of coupling in which the mechanism is mutual inductance between the interference source and the signal system; i.e., the interference is induced in the signal system by a magnetic field produced by the interference source. *Inductive coupling* also is called *magnetic coupling.* The term *inductive coupling* is used also in a more general sense to include coupling due to both magnetic and electric field components.

Coupling, mutual impedance. Coupling of two circuits through one or more common elements. The element is usually a resistor, capacitor, or inductor.

Coupling, radiative. That type of coupling in which the interference is induced in the signal system by electromagnetic radiation produced by the interfering source. *Note.* Where the interference source and susceptors can be modeled as antennas, radiative coupling occurs if the antenna separation is sufficient for the one with the smaller equivalent antenna to be in the radiating near-field region of the other or beyond. If not, the coupling is inductive coupling. The term radiative coupling also is used in a more general sense to include both types of coupling as opposed to "*conductive*" *coupling.*

Cross-coupling. Undesired signal coupling between two or more different communication channels, circuits, or components.

Cross modulation. A type of intermodulation in which modulation on an undesired signal is transferred to a desired signal.

Cross talk. An undesired signal disturbance introduced in a transmission circuit by mutual electric or magnetic coupling with other transmission circuits.

Decoupling. The reduction of transfer of interference energy from one circuit to another.

Desensitization. Reduction in receiver sensitivity due to the presence of a high-level off-channel signal overloading the RF amplifier, mixer stage(s), or causing AGC control action.

Differential mode interference. Interference that causes the potential of one side of the signal transmission path to be changed relative to the other side. *Note:* That type of interference in which the interference current path is wholly in the signal transmission path.

Electric field strength. The magnitude of the potential gradient in an electric field.

Electric induction. The process of generating charges, currents, or voltages in a conductor by means of an electric field.

Electromagnetic compatibility. The ability of C-E equipments, subsystems, and systems to operate in

GLOSSARY (cont'd)

their intended operational environments without suffering or causing unacceptable degradation because of unintentional electromagnetic radiation or response.

Electromagnetic interference. The phenomenon resulting when electromagnetic energy causes unacceptable or undesirable responses, malfunction, degradation, or interruption of the intended operation of electronic equipment, subsystem, or system.

Emission. Electromagnetic energy propagated from a source by radiation or conduction.

Emission, spurious. Any electromagnetic emission on a frequency or frequencies which are outside the necessary emission bandwidth, the level of which may be reduced without degrading the corresponding intended transmission of information. Spurious emissions include harmonic emission, parasitic emission, and intermodulation products, but exclude emission in the immediate vicinity of the necessary emissions bandwidth, which are a result of the modulation process for the transmission of information.

Environment, electromagnetic. Radiated or conducted electromagnetic emission levels which may be encountered by an equipment, subsystem, or system during its life cycle.

Environment, operational. The aggregate of all conditions and influences that may affect the operation of a composite system, vehicle system, and ground system and their respective subsystems and equipment.

Equipment. Any electrical, electronic, or electromechanical device, or collection of items intended to operate as an individual unit and perform a singular function. As defined herein, equipments include, but are not limited to, the following: receivers, transmitters, transceivers, transponders, power supplies, electrical office machines, hand tools, processors, test apparatus and instruments, and material handling equipment.

Far-field region. The region of the field of an antenna where the angular field distribution is essentially independent of the distance from the antenna. *Notes:* (1) If the antenna has a maximum over-all dimension D that is large compared to the wavelength, the far-field region commonly is taken to exist at distances greater than $2D^2/\lambda$ from the antenna, λ being the wavelength. (2) For an antenna focused at infinity, the far-field region is sometimes referred to as the *Fraunhofer region* on the basis of analogy to optical terminology.

Fraunhofer region. See: Far-field region.

Frequency, characteristic. A frequency which can be easily identified and measured in a given emission.

Fresnel region. The region (or regions) adjacent to the region in which the field of an antenna is focused (i.e., just outside the Fraunhofer region). See Note 2 of *Near-field region, radiating.*

Ground plane. A conducting surface or plane used as a common reference point for circuit returns and electrical or signal potentials.

Grounding, single-point. A scheme of circuit/shield grounding in which each circuit/shield has only one physical connection to ground, ideally at the same point for a given subsystem. This technique prevents return currents from flowing into the structure.

Image frequency. In heterodyne frequency converters in which one or two sidebands produced by beating is selected, the image frequency is an undesired input frequency capable of producing the selected frequency by the same process. The word "image" implies the mirror-like symmetry of signal and image frequencies about the beating oscillator frequency or the immediate frequency, whichever is higher.

Image rejection. The decrease in response of a superheterodyne receiver to the image frequency as compared with its response to the desired signal, usually expressed in decibels.

Image rejection ratio. The magnitude of the ratio of the amplitude of the image signal to that of the desired signal to produce the same receiver response. The ratio usually is expressed in decibels.

Immunity (to interference). The quality of a receiver enabling it to reject radio interference.

Impulse. An electrical pulse of short duration relative to a cycle at the highest frequency being considered. Mathematically, it is a pulse of infinite amplitude, infinitesimal duration, and finite area. Its spectral energy density is proportional to its volt-time area, and is uniformly and continuously distributed through the spectrum up to the highest frequency at which it may be considered an impulse. Regularly repeated impulses of uniform level will generate a uniform spectrum of discrete frequencies (Fourier components) separated in frequency by an amount equal to the repetition frequency.

GLOSSARY (cont'd)

Impulse bandwidth. The peak value of the response envelope at the output of a network corrected for the sine wave gain of the network at a state reference frequency (usually the frequency of maximum response) divided by the spectrum amplitude of an impulse applied at the input to the network.

Impulse strength. The area under the amplitude-time relation for the impulse.

Incidental radiation device. A device that radiates radio frequency energy during the course of its operation although the device is not intentionally designed to generate radio frequency energy.

Induced current (or voltage). The interference current (or voltage) appearing in a signal path as a result of coupling of the signal path with an interference field.

Insertion impedance.

Insertion loss (of a network in a transmission system). At a given frequency, the ratio of voltages appearing across the line immediately beyond the point of insertion, before and after insertion.

Interference, broadband. A disturbance that has a spectral energy distribution sufficiently broad that the response of the measuring receiver in use does not vary more than 3 dB when tuned ± two impulse bandwidths.

Interference, narrowband. An undesired emission for which the spectral energy falls entirely within the impulse bandwidth of the measuring instrument.

Intermediate frequency rejection ratio. The ratio of the voltage level of the desired signal to that of an incoming signal at the intermediate frequency for equal IF amplifier output. Both signals should be unmodulated sine waves. The ratio usually is expressed in decibels.

Intermodulation. The mixing of two or more RF signals in a non-linear element to produce signals at new frequencies which are sums and differences of the input signals or their harmonics. The non-linear element may be the output stage of a transmitter, the input circuits of a receiver, or some external device.

Johnson noise. The noise caused by thermal agitation (of electron charge) in a dissipative body. *Notes:* (1) The available thermal (Johnson) noise power N from a resistor at temperature T is $N = kT\Delta f$, where k is Boltzmann's constant and Δf is the frequency increment. (2) The noise power distribution is equal throughout the frequency spectrum, i.e., the noise power is equal in all equal-frequency increments.

Magnetic field strength. The magnitude of the magnetic field vector.

Magnetic induction. The process of generating currents or voltages in a conductor by means of a magnetic field.

Malfunction. A failure of a system or associated subsystem/equipment due to electromagnetic interference or susceptibility that results in loss of life, loss of vehicle, mission abort, or permanent unacceptable reduction in system effectiveness.

Near-field region, radiating. The region of the field of an antenna between the reactive near-field region and the far-field region wherein radiation fields predominate and wherein the angular field distribution is dependent upon distance from the antenna. *Notes:* (1) If the antenna has a maximum overall dimension which is not large compared to the wavelength, this field region may not exist. (2) For an antenna focused at infinity, the radiating near-field region sometimes is referred to as the Fresnel region on the basis of analogy to optical terminology.

Near-field region, reactive. The region of the field immediately surrounding the antenna wherein the reactive field predominates. *Note:* For most antennas the outer boundary of the region is commonly taken to exist at a distance $\lambda/(2\pi)$ from the antenna surface, where λ is the wavelength.

Noise (electrical). Disturbances of a random nature occurring in electrical circuits usually as a result of thermal effects and discrete electron flow (*See: Johnson noise*).

Noise, electromagnetic. An electromagnetic phenomenon, usually impulsive and random, but which may be of a periodic nature, and which does not correspond with any signal.

Noise figure. Of a linear system, at a selected input frequency, the ratio of (1) the total noise power per unit bandwidth (at a corresponding output frequency) delivered by the system into an output termination to (2) the portion thereof engendered at the input frequency by the input termination, whose noise temperature is standard (290 K at all frequencies).

Noise temperature. The temperature of a passive system having an available noise power per unit bandwidth equal to that of the actual terminals.

Open area. A site for radiated electromagnetic interference measurements which is open flat terrain at a distance far enough away from buildings, electric lines, fences, trees, underground cables, and pipe lines so that effects due to such are negligible. The

GLOSSARY (cont'd)

ambient electromagnetic level of the open area should be at least 6 dB below the applicable emission level.

Radio interference. Impairment of the reception of a wanted radio signal caused by an unwanted radio signal or a radio disturbance.

Radio noise (deprecated for use in the US Army). An electromagnetic noise in the radio-frequency range.

Reflection coefficient. The ratio of the phasor magnitude of the reflected wave to the phasor magnitude of the incident wave under specified conditions.

Restricted radiation device. A device in which the generation of radio frequency energy is incorporated intentionally into the design, exclusive of transmitters which require licensing, and exclusive of devices in which the radio frequency energy is used to produce physical, chemical, or biological effects in materials.

Shield. A housing, screen, or other object, usually conducting, that substantially reduces the effect of electric or magnetic fields on one side thereof, upon devices or circuits on the other side.

Shielded enclosure. A specially designed enclosure which affords attenuation to outside RF ambients thereby permitting measurements of electromagnetic emissions from the test sample to be measured without interference from undesired external electromagnetic radiators.

Shielding. An alternate term for shield frequently used to describe metallic braid applied over a flexible electrical cable or wiring.

Shielding effectiveness. For a given external source, the ratio of electric or magnetic field strength at a point before and after the placement of the shield in question.

Spectrum. The distribution of the amplitude (and sometimes phase) of the components of the wave as a function of frequency.

Spectrum amplitude. The magnitude of the spectrum (usually voltage, current, or field strength) per unit frequency bandwidth. For a time function $F(t)$, the spectrum amplitude is $2A(f)$, where $A(f)$ is defined by the expression:

$$F(t) = 2\int_0^{\infty} A(f)\cos[2\pi ft + \varphi(f)]df$$

and φ is the phase angle.

Spectrum signature. A presentation of output level from any device as a function of frequency. The output level may be in terms of power, field strength, current or voltage or several of these, depending upon the device and its electrical characteristics.

Spurious response. Any response of an electronic device, through its intended input terminal, to energy outside its designed reception bandwidth.

Subsystem. For the purpose of establishing EMI requirements either of the following shall be considered as subsystems. In either case, the devices or equipments may be physically separated when in operation and will be installed in fixed or mobile stations, vehicles, or systems. (A) A collection of devices or equipments designed and integrated to function as a single entity but wherein any device or equipment is not required to function as an individual device or equipment. (B) A collection of equipments and subsystems as defined in (A), designed and integrated to function as a major subdivision of a system and to perform an operational function or functions. Some activities consider these collections as systems; however, as noted, they will be considered as subsystems.

Suppression. The reduction or elimination of undesired noise by means of filtering, bonding, shielding, and grounding or any combination thereof.

Suppression component. A device used in the reduction of radio interference at its source. Common components are resistors, capacitors, inductors, filters, and shields.

Susceptibility. The degree to which an electronic equipment, subsystem, or system evidences undesirable responses when subjected to electromagnetic interference.

Susceptibility, conducted. A measure of the interference signal voltage required on power leads to cause an undesirable response or degradation of performance.

Susceptibility, electromagnetic. The ability of an electronic equipment to be degraded by unwanted electromagnetic disturbances.

Susceptibility, radiated. A measure of the radiated interference field required to cause equipment malfunction.

Susceptibility threshold. The signal level at which the test sample exhibits an undesirable response.

System. A composite of equipment, subsystems, skills, and techniques capable of performing or supporting an operational role. A complete system includes related facilities, equipment, subsystems, materials, services, and personnel required for its

GLOSSARY (cont'd)

operation to the degree that it can be considered self-sufficient within its operational or support environment.

Transients. Single-incidence impulses or pulses of low repetition rates generated by a switching action, by relay closures, or other cyclic events.

Transmission line. A material structure forming a continuous path from one place to another for directing the transmission of electric or electromagnetic energy along this path. The term transmission line(s) includes telephone lines, power cables, waveguides, coaxial cables, and other similar items.

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TABLE 3-7. COCHANNEL ACCEPTANCE

Interference / Source	AM VOICE		AM TELETYPE		AM ANALOG DATA		FM VOICE		FM TELETYPE	
AM VOICE	50%	0dB(S)	50%	7.5dB(S)	50%	0dB(S)	50%	3dB(S)	50%	10.5dB(S)
AM TELETYPE	99% 10%	10dB(E) 0dB(E)	99% 10%	10dB(S) 0dB(E)	99% 10%	10dB(E) 0dB(E)	99% 10%	10dB(S) 0dB(E)	99% 10%	10dB(S) 0dB(E)
AM FACSIMILE	50%	10dB(S)		17.5dB(S)	50%	10dB(S)		13dB(S)	*	22.2dB(S)
AM DIGITAL DATA	*	8.3dB(S)	*	8dB(S)	*	8.3dB(S)	*	6dB(S)	*	6dB(S)
AM ANALOG DATA	50%	10dB(E)	*	17.5dB(E)	50%	10dB(E)	*	13dB(E)	*	22.2dB(E)
FM VOICE	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)
FM TELETYPE	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)
FM DIGITAL DATA	90%	6dB(S)	90%	6dB(S)	90%	6dB(S)	90%	dB(S)	90%	6dB(S)
FM ANALOG DATA	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)
FDM	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)
SSB	90% 50% 10%	−12dB(G) −16dB(G) −20dB(G)	*	3dB(S)	50%	−16.5dB(S)	*	−1.9dB(S)	*	3dB(S)
DSB-SC	50%	−19dB(E)	*	0dB(E)	50%	−19.5dB(E)	*	−5dB(E)	*	0dB(E)
PULSE	*	0dB(S)	*	0dB(S)	*	0dB(S)	*	0dB(S)	*	0dB(S)
TV	99%	45dB(S)	99%	45dB(S)	99%	45dB(S)	99%	45dB(S)	99%	45dB(S)

* Percentages, which indicate output fidelity, are in some cases not specified in the original source. Bases for the acceptance ratios in these cases will be found in the appropriate references.

(E) Electromagnetic Compatibility Analysis Center (G) Georgia Institute of Technology (S) Schwartz, Ref. 5[?]

ACCEPTANCE RATIO (Adapted from Refs. 58 and 59)

FM TELETYPE		FDM		SSB		PULSE		RANDOM NOISE		FM ANALOG DATA		TV	
0%	10.5dB(S)	50%	−10dB(S)	50%	10dB(S)	50%	−44dB(S)	50%	10dB(S)	50%	10dB(E)	50%	0dB(E)
9% 0%	10dB(S) 0dB(E)	99% 10%	10dB(S) 0dB(E)	99% 10%	10dB(S) 0dB(E)	99% 10%	10dB(S) 0dB(E)	99% 10%	10dB(E) 0dB(E)	99% 10%	10dB(E) 0dB(E)	99% 10%	10dB(E) 0dB(E)
	22.2dB(S)	50%	0dB(S)	*	20dB(S)	*	−30dB(S)	50%	10dB(S)	50%	0dB(E)	50%	10dB(E)
*	6dB(S)	*	8dB(S)	*	6dB(S)	*	6dB(S)	*	12.7dB(S)	*	6dB(E)	*	8.3dB(E)
*	22.2dB(E)	50%	0dB(E)	*	20dB(E)	*	−30dB(E)	50%	10dB(E)	50%	0dB(E)	50%	0dB(E)
0%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(E)	50%	0dB(E)
.999% 0%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)	99.999% 10%	10dB(E) 0dB(E)
0%	6dB(S)	90%	6dB(S)	90%	6dB(S)	90%	6dB(S)	90%	10dB(S)	90%	6dB(E)	90%	6dB(E)
0%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)	50%	0dB(E)
0%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(S)	50%	0dB(E)	50%	0dB(E)
*	3dB(S)	*	−10dB(S)	90% 50% 10%	−2dB(G) −9dB(G) −16dB(G)	*	−60dB(S)	*	0dB(S)	*	−10dB(E)	50%	−16.5dB(E)
*	0dB(E)	*	−13dB(E)	90% 50% 10%	−7dB(G) −14dB(G) −17dB(G)	*	−63dB(E)	*	−3dB(E)	*	−13dB(E)	50%	−19.5dB(E)
*	0dB(S)	*	0dB(S)	*	0dB(S)	*	0dB(S)	*	0dB(S)	*	0dB(E)	*	0dB(E)
9%	45dB(S)	99%	45dB(S)	99%	27dB(S)	99%	20dB(S)	99%	45dB(E)	99%	45dB(E)	99%	45dB(E)

Schwartz, Ref. 58

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PAGES 3-59 AND 3-60 ARE NOT AVAILABLE.

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