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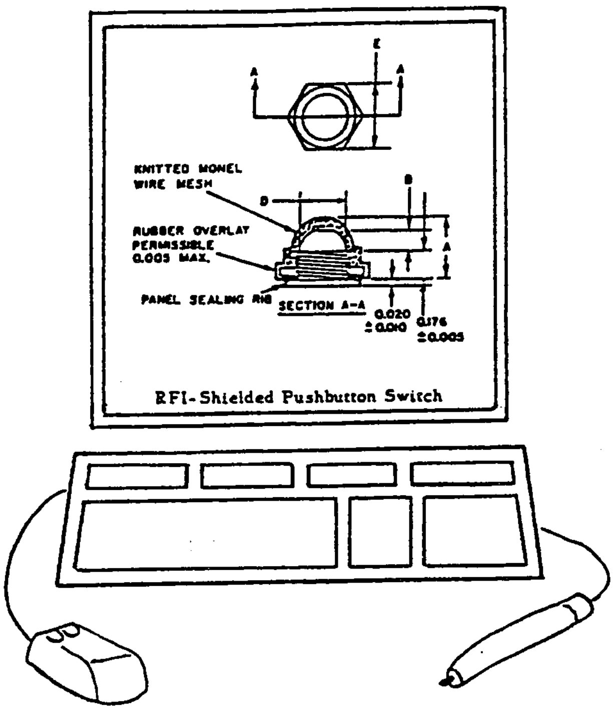
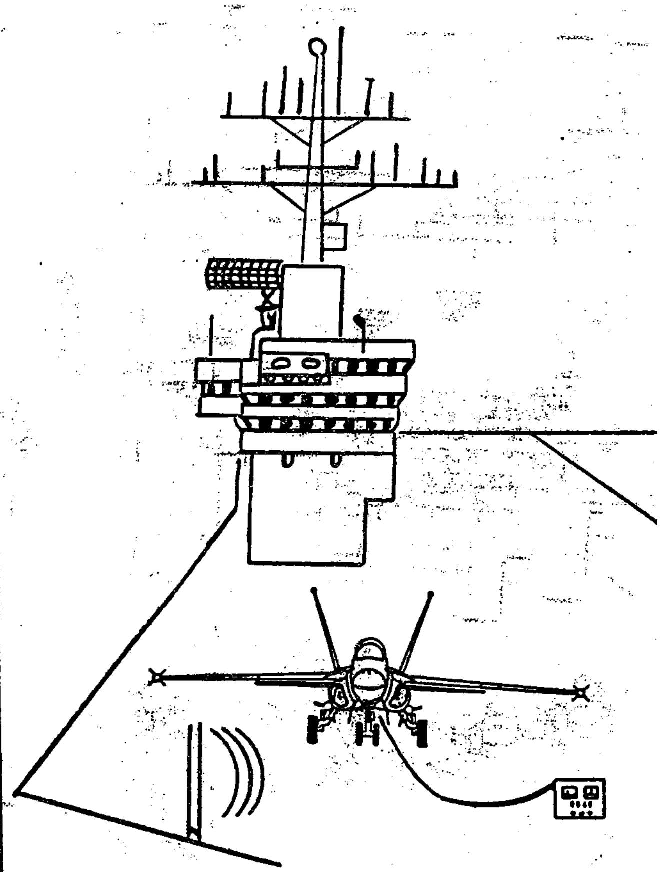


ELECTROMAGNETIC COMPATIBILITY DESIGN GUIDE FOR AVIONICS AND RELATED GROUND SUPPORT EQUIPMENT

NAVAL AIR SYSTEMS COMMAND



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DESIGN GUIDE
AD 1115, EMC Design Guide for Avionics and Related GSE

ORGANIZATION OF SUBMITTEE	CITY, STATE & ZIP

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1.0 INTRODUCTION

1.1 PURPOSE

The purpose of this publication is to provide Electromagnetic Compatibility (EMC) guidance in connection with the design and development of Navy Avionics and Related Ground Support Equipment (GSE). In terms of this purpose, it is appropriate to define the terms EMC, Avionics, and GSE, and then to indicate those aspects of the definitions which this design guide addresses.

EMC is defined in MIL-STD-463 as "the capability of electronic equipments or systems to be operated with a defined margin of safety in the intended operational environment at designed levels or efficiency without degradation due to interference." The definition is established in an Electromagnetic (EM) context, so it can be interpreted in terms of an Electromagnetic Environment (EME) rather than a physical or tactical environment. Physical or tactical environments identify where systems or equipments are located, but do not generally include descriptions of EM emissions or susceptibility effects.

Avionics is defined for this document using a non-military definition. Simply put, Avionics is the black box enclosures that the electronics go into for aeronautic and astronautic vehicles. This design guide will address avionics on Navy aircraft (Aeronautics), in particular.

GSE is defined in OPNAVINST 4790.2 as "... all equipment on the ground required to make a weapons system, support system, subsystem, or end-item of equipment operational in its intended environment. This includes all equipment required to install, launch, arrest (except Navy shipboard and shore-based launching and arresting equipment less missile target launching equipment), guide, control, direct, inspect, test, adjust, calibrate, appraise, gauge, measure, assemble, disassemble, handle, transport, safeguard, store, actuate, service, repair, overhaul, maintain, or operate the system, subsystem, end item, or component...." Specifically excluded from this GSE definition are handtools, house-keeping items, office equipment, common production tools, and ancillary equipment such as head-sets and microphones.

1.2 OBJECTIVE

It is probably not necessary to justify a statement that representative Navy shipboard and shore EMEs are becoming more complex and of higher intensity. Many examples can be cited of the increased use of EM emitters within or in the vicinity of shore maintenance shops, or near base flight-lines, or on ships of the fleet. These equipments have output levels ranging from fractions of milliwatts to tens of megawatts, with waveforms having noise-like properties as well as those of a very complex coherent nature. They often are developed to intentionally radiate these power levels within selected portions of the radio spectrum, but equally often may be emitting EM energies into parts of the spectrum as by-products of some non-radiation or limited radiation application (spurious emissions).

In a like manner, more equipments are appearing in these same operational locations that have relatively sensitive EM degradation thresholds. When these thresholds are exceeded (due to either spurious or intended emissions), they can result in reduced equipment performance capabilities. This, in turn, can cause a reduction in the overall effectiveness of the mission in which that equipment is used.

Avionics and related GSE operate in these environments. They are subject to the diverse energy levels and waveforms created by the fleet and other proximate activities. They also can influence the performance of other environmental equipments, should they couple power to susceptible devices. A very critical consideration throughout the life-cycle of Avionics and related GSE must therefore be its ability to operate computably in this environment. That is the objective of this design manual. More specifically, this publication is intended to provide Avionics/GSE engineering and management personnel with an awareness that a potential of interference exists between Avionics/GSE and other EM equipment. It is hoped to convey the philosophy that, if this potential is not recognized and taken into account during Avionics/GSE design, degradation to testing or operational equipment and systems (or both) can result.

The Electromagnetic Interference (EMI) problems encountered by GSE, and the interference reduction techniques that must be employed to achieve EMC, are often unique to this class of equipment (with the noted exception that the interference reduction techniques are also applicable to Avionics). This uniqueness on the part of GSE is brought about by features of the GSE itself, typically small, portable, and with limited Radio Frequency (RF) circuitry, and by the specialized EMEs in which both Avionics and GSE may operate (e.g., high narrow-band electric fields). A major objective of this design guide is to provide the Avionics/GSE designer with EMC guidance that concentrates on those areas of control dictated by these types of equipments and environments.

The design guide provides extensive technical data and information to assist Avionics and related GSE design and development engineers in creating a product that will be compatible with its operational environment. While it is not possible to anticipate all of the interface situations that can arise during the course of equipment development, the guidelines, approaches and techniques covered in this manual are expected to furnish cognizant engineers that have system EMC responsibility with the necessary tools for identifying, attacking and resolving EMI problem areas.

It is hoped that use of this design guide will benefit Avionics/GSE hardware during all stages of the design/development cycle. Recognition of the characteristics of a proposed piece of Avionics/GSE, and the features of its expected operational environment, can influence choices of conceptual approaches and point out design tradeoffs dictated by EMC considerations. Early identification of EMI system reduction requirements can prevent extensive hardware redesign and its attendant additional costs. Application of the correct EMI reduction measures can result in a piece of Avionics/GSE that is compatible with its environment, and meets its other performance and physical (size, weight, volume) objectives.

1.3 ORIENTATION

An individual who has worked previously in the area of interference suppression of electrical/electronic equipment and systems is likely to comment that there are already several EMC design guides available on this subject. He would be correct; there are not only several good compilations of EMC design techniques available, but there have been innumerable books and technical papers prepared on various facets of equipment EM interaction and interference reduction, and in such related areas as Electromagnetic Pulse (EMP) effects, EM hazards to ordnance, and biological effects of EM radiation. A few of the more extensive unclassified EMC design guides, handbooks, and manuals published by the government in recent years are listed in Table 1-1.

TABLE 1-1

Sample of EMC Information Sources

Electromagnetic Compatibility, DH1-4
Air Force Systems
Series 1-0 General
Third Edition, Revision 2, January 1976

Design Principles and Practices for
Controlling Hazards of
Electromagnetic Radiation to Ordnance
(HERO Design Guide)
NAVSEA OD 30393, First Revision,
15 September 1974

Engineering Design Handbook
Hardening Weapon Systems Against RF Energy
AMC Pamphlet AMCP 706-235
Headquarters, US Army Material Command
February 1972

Military Standardization Handbook
Electro-Magnetic (Radiated) Environment Considerations
for Design and Procurement of Electrical and
Electronic Equipment
MIL-HDBK-235-1, Unclassified; -2, Confidential; -3,
Secret 23 June 1972

Electromagnetic Compatibility Manual
NAVAIR 5335
Naval Air Systems Command
Washington, DC 1972

Electronics Installation and Maintenance Book
Electromagnetic Interference Reduction
NAVSHIPS 0967-000-0150
Naval Ship Engineering Center June 1970

Many of the interference reduction techniques covered by the listed publications are also applicable to the types of EMI situations that an Avionics/GSE design/development engineer might encounter; therefore these techniques have been incorporated into this publication as well. However, additional material has been provided in this guide to take into account two unique aspects of EMC control in GSE.

First, the class of avionics related GSE performs selected operational functions within the Navy, and constitutes a specialized group of devices as far as interference suppression requirements are concerned. This design guide recognizes the somewhat homogeneous nature of this group of equipments, and concentrates on potential EMI problem areas and interference reduction techniques that are most directly applicable to GSE. Second, the EMEs (particularly the shipboard environments) in which avionics related GSE must operate are often very specialized. This design guide pays particular attention to these environments and the unique EMI problems they can create.

Since this guide is directed toward experienced system designers, fundamental or basic engineering discussions have been avoided. The reader is referred to the documents listed in Table 1-1 (as well as those contained in the chapter appendices) for expansions of many of the topics covered in this publication.

1.4 SCOPE

The eleven chapters of this publication are intended to provide technical and management personnel concerned with Avionics/GSE design and development with the necessary information to evaluate, analyze and resolve potential EMI problems, and to maintain system-to-environment compatibility throughout the life of the equipment.

The General Scope of the Chapters are as follows:

Chapters 1 through 8 - EMC control through problem awareness and design.

Chapters 9 through 11 - EMC control through EMC program planning and testing.

The Specific Scope of the remaining chapters are as follows:

Chapter 2, Sources and Coupling of EM Energy - An unclassified discussion of the environments in which GSE must operate, with particular attention to radiated and conducted energies on flight and hangar decks, along base flightlines, and in shops and rework facilities.

Chapter 3, GSE Design Considerations - A summary of general EMC design guidelines; factors influencing GSE design; cost, maintenance, and safety considerations; examples of GSE and avionics design.

Chapter 4, Shielding - A discussion of applicable shielding techniques, with emphasis on case and cable shielding, and the cable/connector interface.

Chapter 5, Bonding - A summary of bonding techniques of interest to GSE designers.

Chapter 6, Grounding - A summary of single-point and multiple-point grounding, including the grounding of shielded cables.

Chapter 7, Filtering - A detailed discussion of the various types of filters that can be employed to reduce conductive EMI, and their applications in GSE design.

Chapter 8, Tables and Nomographs - A compilation of additional tables, charts, nomographs and other design aids useful to Avionics/GSE design engineers in investigating EMI effects.

Chapter 9, EMC Management Program - Guidance in organizing and implementing effective interference regulations, through EMC Control Plans.

Chapter 10, Specifications and Standards - A summary of government EMC-related specifications and standards applicable at component, subsystem, equipment, and system levels.

Chapter 11, Test Requirements, Plans and Techniques - Guidance on EMI regulations/test requirements and how to implement them through EMC test plans and techniques. In particular, the information in this chapter on Ground Plane Testing will help prepare Avionics/GSE designers for this shipboard environment simulation type testing. This chapter includes a delineation of the differences between MIL-STD-461A/B/&C.

The major EMI problems previously encountered by the GSE community have centered around the effects of the operational environment on GSE performance. As a result, the emphasis of the design guide is in this direction, rather than on possible GSE interference to the environment, or internal interference effects within particular GSE. This is not to imply that these other areas have been excluded from this guide, but merely points out that the material in the guide concentrates on coupling problems known to have been encountered in the past.

1.5 SPECIFICATION USAGE

Specifications have been referenced throughout this design guide, and although the revisions cited are current at the time of publication, they may be out-of-date at the time of reading. Therefore, the reader is cautioned to check for the latest revision of any specification of particular interest.

2.0 SOURCES AND COUPLING OF EM ENERGY

2.1 THE ENVIRONMENT

The multitude of man-made electrical and electronic equipments create almost the total operational EME in which GSE/Avionics must operate. Except in very rare cases, non-man-made sources that contribute to the environment (such as solar noise, galactic noise, atmospheric noise, etc.) can be disregarded.

From an EM point of view, these environments can be quite complex. As stated in Chapter 1, they may contain discrete or very narrow-band sources (such as the outputs of a communication transmitter), or sources with relatively complex modulation techniques that tend to produce a broad-band output (such as radars), or impulsive or noise like sources whose emissions cover a wide frequency range (such as gasoline engines or fluorescent lights). They may include civilian as well as government equipments.

The frequency band of the EM spectrum representing environments of concern to GSE/Avionics extends from a few Hertz to tens of Gigahertz. International radio regulations and national rules and procedures allocate the band between 9 kHz and 300 GHz for communications and navigation purposes. Frequency allocations and assignments for mobile radio, radar, radio astronomy, space telemetry, radiosonde, and other diverse types of equipments for commercial and U.S. government use are made in conformance with these rules. Table 2-1 is a brief summary of the spectrum allocated to various services with special emphasis on U.S. government allocations.

Many other types of emissions exist that are not strictly regulated. These could influence the performance of GSE. They occur at the lower end of the spectrum, and include emissions from power-lines, ignition systems, motor brushes, arc welders, and many others.

The designer of GSE/Avionics has very limited control over the EM environment. The equipment that is developed, whether a portable unit to be operated on a carrier flight deck, a computer in an aircraft or a large testing device connected to 3-phase, 60 Hz power mains at a base installation, must operate satisfactorily in the EM fields created by other equipments, without constraints to those other equipments. In addition, the emissions from the GSE/Avionics must be controlled so that equipments in the operational environment will not be degraded.

In order to design avionics or GSE to meet these constraints, a good understanding of these environments is necessary. The remainder of this chapter will discuss the general characteristics of the electromagnetic environment (EME). Specific features of environments in which GSE/Avionics equ ts may be expected to operate will be identified.
These discussions are limited in detail so that the design guide can be maintained as an unclassified document. More complete descriptions of

Table 2-1

Allocation of the Radio Spectrum

FREQUENCY kHz	SERVICE
9 - 14	RADIONAVIGATION
14 - 1995	FIXED, MARITIME MOBILE
19.95 - 20.05	STANDARD TIME
20.05 - 59	FIXED, MARITIME MOBILE
59 - 61	STANDARD TIME
61 - 70	FIXED, MARITIME MOBILE
70 - 90	FIXED, MARITIME MOBILE Radiolocation
90 - 110	RADIONAVIGATION
110 - 130	FIXED, MARITIME MOBILE Radiolocation
130 - 190	FIXED, MARITIME MOBILE
190 - 200	AERONAUTICAL RADIONAVIGATION
200 - 275	AERONAUTICAL RADIONAVIGATION Aeronautical Mobile
275 - 285	AERONAUTICAL RADIONAVIGATION Aeronautical Mobile Maritime Radionavigation (Radiobeacons)
285 - 325	MARITIME RADIONAVIGATION (Radiobeacons) Aeronautical Radionavigation (Radiobeacons)
325 - 335	AERONAUTICAL RADIONAVIGATION (Radiobeacons) Aeronautical Mobile Maritime Radionavigation (Radiobeacons)
335 - 405	AERONAUTICAL RADIONAVIGATION (Radiobeacons) Aeronautical Mobile
405 - 415	RADIONAVIGATION Aeronautical Mobile
415 - 435	MARITIME MOBILE AERONAUTICAL RADIONAVIGATION
435 - 495	MARITIME MOBILE
495 - 505	MOBILE (Distress and calling)
505 - 510	MARITIME MOBILE
510 - 525	AERONAUTICAL RADIONAVIGATION (Radiobeacons) MARITIME MOBILE (Ships only)
525 - 535	AERONAUTICAL RADIONAVIGATION (Radiobeacons) MOBILE
535 - 1605	BROADCASTING
1605 - 1615	MOBILE
1615 - 1625	BROADCASTING
1625 - 1705	Radiolocation
1705 - 1800	FIXED, MOBILE, RADIOLOCATION
1800 - 1900	AMATEUR
1900 - 2000	RADIOLOCATION
2000 - 2065	FIXED, MOBILE
2065 - 2107	MARITIME MOBILE
2107 - 2170	FIXED, MOBILE
2170 - 2173.5	MARITIME MOBILE (ship, telephony)

FREQUENCY kHz	SERVICE
2173.5 - 2190.5	MOBILE (Distress and calling)
2190.5 - 2194	MARITIME MOBILE (Ship, telephony)
2194 - 2495	FIXED, MOBILE
2495 - 2505	STANDARD TIME
2505 - 2850	FIXED, MOBILE
2850 - 3025	AERONAUTICAL MOBILE (R)
3025 - 3155	AERONAUTICAL MOBILE (OR)
3155 - 3230	FIXED, MOBILE except aeronautical mobile (R)
3230 - 3400	FIXED, MOBILE except aeronautical mobile Radiolocation
3400 - 3500	AERONAUTICAL MOBILE (R)
3500 - 4000	AMATEUR
4000 - 4438	MARITIME MOBILE
4438 - 4650	FIXED, MOBILE except aeronautical mobile (R)
4650 - 4700	AERONAUTICAL MOBILE (R)
4700 - 4750	AERONAUTICAL MOBILE (OR)
4740 - 4850	FIXED, MOBILE except aeronautical mobile (R)
4850 - 4995	FIXED, MOBILE
4995 - 5005	STANDARD TIME
5005 - 5060	FIXED
5060 - 5450	FIXED, MOBILE except aeronautical mobile
5450 - 5680	AERONAUTICAL MOBILE (R)
5680 - 5730	AERONAUTICAL MOBILE (OR)
5730 - 5950	FIXED, MOBILE except aeronautical mobile (R)
5950 - 6200	BROADCASTING

FREQUENCY kHz	SERVICE
6200 - 6525	MARITIME MOBILE
6525 - 6685	AERONAUTICAL MOBILE (R)
6685 - 6765	AERONAUTICAL MOBILE (OR)
6765 - 7000	FIXED, Mobile
7000 - 7100	AMATEUR, AMATEUR SATELLITE
7100 - 7300	AMATEUR
7300 - 8100	FIXED
8100 - 8815	MARITIME MOBILE
8815 - 8965	AERONAUTICAL MOBILE (R)
8965 - 9040	AERONAUTICAL MOBILE (OR)
9040 - 9500	FIXED
9500 - 9900	BROADCASTING
9900 - 9995	FIXED
9995 - 10005	STANDARD TIME
10005 - 10100	AERONAUTICAL MOBILE (R)
10100 - 10150	AMATEUR
10150 - 11175	FIXED, Mobile except aeronautical mobile (R)
11175 - 11275	AERONAUTICAL MOBILE (OR)
11275 - 11400	AERONAUTICAL MOBILE (R)
11400 - 11650	FIXED
11650 - 12050	BROADCASTING
12050 - 12230	FIXED
12230 - 13200	MARITIME MOBILE
13200 - 13260	AERONAUTICAL MOBILE (OR)
13260 - 13360	AERONAUTICAL MOBILE (R)
13360 - 13410	RADIO ASTRONOMY
13410 - 13600	FIXED, Mobile except aeronautical mobile (R)
13600 - 13800	BROADCASTING
13800 - 14000	FIXED, Mobile except aeronautical mobile (R)
14000 - 14250	AMATEUR, AMATEUR SATELLITE
14250 - 14350	AMATEUR

FREQUENCY kHz	SERVICE
14350 - 14990	FIXED, Mobile except aeronautical mobile (R)
14990 - 15010	STANDARD TIME
15010 - 15100	AERONAUTICAL MOBILE (OR)
15100 - 15600	BROADCASTING
15600 - 16360	FIXED
16360 - 17410	MARITIME MOBILE
17410 - 17550	FIXED
17550 - 17900	BROADCASTING
17900 - 17970	AERONAUTICAL MOBILE (R)
17970 - 18030	AERONAUTICAL MOBILE (OR)
18030 - 18068	FIXED
18068 - 18168	AMATEUR, AMATEUR-SATELLITE
18168 - 18780	FIXED, Mobile
18780 - 18900	MARITIME MOBILE
18900 - 19680	FIXED
19680 - 19800	MARITIME MOBILE
19800 - 19990	FIXED
19990 - 20010	STANDARD TIME
20010 - 21000	FIXED, Mobile
21000 - 21850	BROADCASTING
21850 - 21924	FIXED
21924 - 22000	AERONAUTICAL MOBILE (R)
22000 - 22855	MARITIME MOBILE
22855 - 23000	FIXED
23000 - 23200	FIXED, Mobile except aeronautical mobile (R)
23200 - 23350	AERONAUTICAL MOBILE (OR)
23350 - 24890	FIXED, MOBILE except aeronautical mobile
24890 - 24990	AMATEUR, AMATEUR-SATELLITE
23000 - 23200	FIXED, Mobile except aeronautical mobile (R)
23200 - 23350	AERONAUTICAL MOBILE (OR)

FREQUENCY kHz	SERVICE
23350 - 24890	FIXED, MOBILE except aeronautical mobile
24890 - 24990	AMATEUR, AMATEUR-SATELLITE
24990 - 25010	STANDARD TIME
25010 - 25070	LAND MOBILE
25070 - 25210	MARITIME MOBILE
25210 - 25330	LAND MOBILE
25330 - 25550	FIXED, MOBILE except aeronautical mobile
25550 - 25670	RADIO ASTRONOMY
25670 - 26100	BROADCASTING
26100 - 26175	MARITIME MOBILE
26175 - 26480	LAND MOBILE
26480 - 26950	FIXED, MOBILE except aeronautical mobile
26950 - 26960	FIXED
26960 - 27230	MOBILE except aeronautical mobile
27230 - 27410	FIXED, MOBILE except aeronautical mobile

FREQUENCY MHz		SERVICE
27.41	- 27.54	LAND MOBILE
27.54	- 28	FIXED MOBILE
28	- 29.70	AMATEUR, AMATEUR-SATELLITE
29.70	- 29.80	LAND MOBILE
29.80	- 29.89	FIXED
29.89	- 29.91	FIXED, MOBILE
29.91	- 30	FIXED
30	- 30.56	MOBILE, FIXED
30.56	- 32	LAND MOBILE
32	- 33	FIXED, MOBILE
33	- 34	LAND MOBILE
34	- 35	FIXED, MOBILE
35	- 36	LAND MOBILE
36	- 37	FIXED, MOBILE
37	- 37.5	LAND MOBILE
37.5	- 38	RADIO ASTRONOMY
38	- 38.25	FIXED, MOBILE, RADIO ASTRONOMY
38.25	- 39	FIXED, MOBILE
39	- 40	LAND MOBILE
40	- 42	FIXED, MOBILE
42	- 46.60	LAND MOBILE
46.60	- 47	FIXED. MOBILE
47	- 49.60	LAND MOBILE
49.60	- 50	FIXED, MOBILE
50	- 54	AMATEUR
54	- 72	BROADCASTING
72	- 73	FIXED, MOBILE
73	- 74.60	RADIO ASTRONOMY
74.60	- 74.8	FIXED, MOBILE
74.8	- 75.2	AERONAUTICAL RADIONAVIGATION
75.2	- 75.4	FIXED, MOBILE
75.4	- 76	FIXED, MOBILE
76	- 88	BROADCASTING (Television)
88	- 108	BROADCASTING (FM broadcasting)

FREQUENCY MHz		SERVICE
108	- 117.975	AERONAUTICAL RADIONAVIGATION
117.975	- 121.9375	AERONAUTICAL MOBILE (R)
121.9375	- 123.0875	AERONAUTICAL MOBILE (R)
123.0875	- 123.5875	AERONAUTICAL MOBILE
123.5875	- 137	AERONAUTICAL MOBILE (R)
137	- 138	METEOROLOGICAL-SATELLITE, SPACE OPERATION, SPACE RESEARCH
138	- 144	FIXED, MOBILE
144	- 146	AMATEUR, AMATEUR-SATELLITE
146	- 148	AMATEUR
148	- 149.9	FIXED, MOBILE
149.9	- 150.05	RADIONAVIGATION-SATELLITE
150.05	- 150.8	FIXED, MOBILE
150.8	- 156.2475	LAND MOBILE
156.2475	- 157.45	MARITIME MOBILE
157.45	- 161.575	LAND MOBILE
161.575	- 161.625	MARITIME MOBILE
161.625	- 161.775	LAND MOBILE
161.775	- 162.0125	MARITIME MOBILE
162.0125	- 173.2	FIXED, MOBILE
173.2	- 173.4	FIXED, Land mobile
173.4	- 174	FIXED, MOBILE
174	- 216	BROADCASTING (Television)
216	- 220	MARITIME MOBILE, Radiolocation, Fixed, Land Mobile, Aeronautical Mobile
220	- 225	FIXED, MOBILE, Radiolocation
225	- 328.6	FIXED, MOBILE FREQUENCY SERVICE
328.6	- 335.4	AERONAUTICAL RADIONAVIGATION
335.4	- 399.9	FIXED, MOBILE
399.9	- 400.05	RADIONAVIGATION-SATELLITE
400.05	- 400.15	STANDARD TIME

FREQUENCY MHz	SERVICE
400.15 - 401	METEOROLOGICAL AIDS SPACE RESEARCH METEOROLOGICAL-SATELLITE Space Operation
401 - 402	METEOROLOGICAL AIDS SPACE OPERATION Earth Exploration-Satellite Meteorological-Satellite
402 - 403	METEOROLOGICAL AIDS Earth Exploration-Satellite Meteorological-Satellite
403 - 406	METEOROLOGICAL AIDS
406 - 406.1	MOBILE-SATELLITE (EPIRB)
406.1 - 410	FIXED, MOBILE, RADIO ASTRONOMY
410 - 420	FIXED, MOBILE
420 - 450	RADIOLOCATION
450 - 460	LAND MOBILE
460 - 470	Meteorological-Satellite
470 - 512	BROADCASTING, LAND MOBILE
512 - 608	BROADCASTING
608 - 614	RADIO ASTRONOMY
614 - 806	BROADCASTING
806 - 902	LAND MOBILE
902 - 928	RADIOLOCATION
928 - 935	FIXED
935 - 941	LAND MOBILE
941 - 960	FIXED
960 - 1215	AERONAUTICAL RADIONAVIGATION
1215 - 1240	RADIOLOCATION, RADIONAVIGATION-SATELLITE
1240 - 1300	RADIOLOCATION
1300 - 1350	AERONAUTICAL RADIONAVIGATION Radiolocation
1350 - 1400	RADIOLOCATION, Fixed, Mobile

FREQUENCY MHz	SERVICE
1400 - 1427	RADIO ASTRONOMY, EARTH EXPLORATION-SATELLITE, SPACE RESEARCH
1427 - 1429	FIXED, MOBILE except aeronautical mobile SPACE OPERATION
1429 - 1435	FIXED, MOBILE
1435 - 1530	MOBILE
1530 - 1535	MARITIME MOBILE-SATELLITE Mobile
1535 - 1544	MARITIME MOBILE-SATELLITE
1544 - 1545	MOBILE-SATELLITE
1545 - 1559	AERONAUTICAL MOBILE-SATELLITE
1546 - 1610	AERONAUTICAL RADIONAVIGATION RADIONAVIGATION-SATELLITE
1610 - 1626.5	AERONAUTICAL RADIONAVIGATION
1626.5 - 1645.5	MARITIME MOBILE-SATELLITE
1645.6 - 1646.5	MOBILE-SATELLITE
1646.5 - 1660	AERONAUTICAL MOBILE-SATELLITE (R)
1660 - 1660.5	AERONAUTICAL MOBILE-SATELLITE (R) RADIO ASTRONOMY
1660.5 - 1668.4	RADIO ASTRONOMY, SPACE RESEARCH
1668.4 - 1670	METEOROLOGICAL AIDS, RADIO ASTRONOMY
1670 - 1700	METEOROLOGICAL AIDS, METEOROLOGICAL-SATELLITE
1700 - 1710	FIXED, METEOROLOGICAL- SATELLITE
1710 - 1850	FIXED, MOBILE
1850 - 1990	FIXED
1990 - 2110	FIXED, MOBILE
2110 - 2200	FIXED
2200 - 2300	FIXED, MOBILE, SPACE RESEARCH
2300 - 2310	RADIOLOCATION, Fixed, Mobile
2310 - 2390	RADIOLOCATION, MOBILE, Fixed
2390 - 2450	RADIOLOCATION

FREQUENCY MHz		SERVICE
2450	- 2483.5	FIXED, MOBILE, Radiolocation
2483.5	- 2500	RADIO-DETERMINATION SATELLITE
2500	- 2655	BROADCASTING-SATELLITE, FIXED
2655	- 2690	Earth Exploration-Satellite, Radio Astronomy, Space Research
2690	- 2700	RADIO ASTRONOMY, EARTH EXPLORATION SATELLITE, SPACE RESEARCH
2700	- 2900	AERONAUTICAL RADIO NAVIGATION, METEOROLOGICAL AIDS Radiolocation
2900	- 3100	MARITIME RADIONAVIGATION Radiolocation
3100	- 3500	RADIO LOCATION
3500	- 3700	AERONAUTICAL RADIONAVIGATION, RADIOLOCATION
3700	- 4200	FIXED, FIXED-SATELLITE
4200	- 4400	AERONAUTICAL RADIONAVIGATION
4400	- 4990	FIXED, MOBILE
4990	- 5000	RADIO ASTRONOMY, Space Research
5000	- 5250	AERONAUTICAL RADIONAVIGATION
5250	- 5350	RADIOLOCATION
5350	- 5460	AERONAUTICAL RADIONAVIGATION, RADIOLOCATION
5460	- 5470	RADIONAVIGATION, Radiolocation
5470	- 5600	MARITIME RADIONAVIGATION, Radiolocation
5600	- 5650	MARITIME RADIONAVIGATION METEOROLOGICAL AIDS, Radiolocation
5650	- 5925	RADIOLOCATION FREQUENCY SERVICE MHz
5925	- 7125	FIXED, FIXED-SATELLITE. MOBILE
7125	- 7190	FIXED
7190	- 7235	FIXED, SPACE RESEARCH
7235	- 7250	FIXED

FREQUENCY MHz		SERVICE
7250	- 7300	FIXED-SATELLITE, MOBILE-SATELLITE Fixed
7300	- 7450	FIXED, FIXED-SATELLITE
7450	- 7550	FIXED, FIXED-SATELLITE, METEOROLOGICAL-SATELLITE, Mobile Satellite
7550	- 7750	FIXED. FIXED-SATELLITE, Mobile Satellite
7750	- 7900	FIXED
7900	- 8025	Fixed, FIXED-SATELLITE, MOBILE SATELLITE
8025	- 8175	EARTH EXPLORATION-SATELLITE, FIXED, FIXED-SATELLITE, Mobile Satellite
8175	- 8215	EARTH EXPLORATION-SATELLITE, FIXED, FIXED-SATELLITE, METEOROLOGICAL SATELLITE, Mobile Satellite
8215	- 8400	EARTH EXPLORATION-SATELLITE, FIXED, FIXED-SATELLITE, Mobile Satellite
8400	- 8500	FIXED, SPACE RESEARCH
8500	- 9000	RADIOLOCATION
9000	- 9200	AERONAUTICAL RADIONAVIGATION, Radiolocation
9200	- 9300	RADIOLOCATION, MARITIME RADIONAVIGATION
9300	- 9500	RADIONAVIGATION, Meteorological Aids, Radiolocation

FREQUENCY GHz		SERVICE
9.5	- 10.55	RADIOLOCATION
10.55	- 10.6	FIXED
10.6	- 10.68	EARTH EXPLORATION-SATELLITE SPACE RESEARCH
10.68	- 10.7	RADIO ASTRONOMY, EARTH EXPLORATION-SATELLITE SPACE RESEARCH
10.7	- 11.7	FIXED, FIXED-SATELLITE
11.7	- 12.2	FIXED-SATELLITE, Mobile except aeronautical mobile
12.2	- 12.7	FIXED, BROADCASTING SATELLITE
12.7	- 13.25	FIXED, FIXED-SATELLITE, MOBILE
13.25	- 13.4	AERONAUTICAL RADIONAVIGATION Space Research
13.4	- 14	RADIOLOCATION, Space Research, Standard Frequency
14	- 14.2	RADIONAVIGATION, Space Research
14.2	- 14.4	FIXED-SATELLITE
14.4	- 14.5	Mobile, Fixed
14.5	- 14.7145	FIXED, Mobile, Space Research
14.7145	- 15.1365	Fixed, MOBILE, Space Research
15.1365	- 15.35	FIXED, Mobile, Space Research
15.35	- 15.4	RADIO ASTRONOMY, EARTH EXPLORATION SATELLITE, SPACE RESEARCH
15.4	- 15.7	AERONAUTICAL RADIONAVIGATION
15.7	- 16.6	RADIOLOCATION
16.6	- 17.1	RADIOLOCATION, Space Research
17.1	- 17.2	RADIOLOCATION
17.2	- 17.3	RADIOLOCATION, Earth Exploration-Satellite, Space Research
17.3	- 17.7	Radiolocation
17.7	- 18.6	FIXED, FIXED-SATELLITE, MOBILE
18.6	- 18.8	EARTH EXPLORATION-SATELLITE SPACE RESEARCH

FREQUENCY GHz		SERVICE
18.8	- 19.7	FIXED, FIXED-SATELLITE, MOBILE
19.7	- 20.2	FIXED-SATELLITE, Mobile-Satellite
20.2	- 21.2	FIXED-SATELLITE, MOBILE-SATELLITE Standard Frequency
21.4	- 22	FIXED, MOBILE
22	- 22.21	FIXED, MOBILE except aeronautical mobile
22.21	- 22.5	FIXED, MOBILE except aeronautical mobile, RADIO EXPLORATION SATELLITE, SPACE RESEARCH
22.5	- 22.55	FIXED, MOBILE
22.55	- 23.55	INTER-SATELLITE, FIXED, MOBILE
23.55	- 23.6	FIXED, MOBILE
23.6	- 24	RADIO ASTRONOMY, EARTH EXPLORATION SATELLITE, SPACE RESEARCH
24	- 24.05	AMATEUR, AMATEUR-SATELLITE
24.05	- 24.25	RADIOLOCATION, Earth Exploration Satellite
24.25	- 25.25	RADIONAVIGATION
25.25	- 27	FIXED, MOBILE, Earth Exploration Satellite, Standard Frequency
27	- 27.5	FIXED, MOBILE, Earth Exploration Satellite
27.5	- 29.5	FIXED, FIXED-SATELLITE, MOBILE
29.5	- 30	FIXED-SATELLITE, Mobile-Satellite
30	- 31	FIXED-SATELLITE, MOBILE-SATELLITE Standard Frequency
31	- 31.3	Standard Frequency
31.3	- 31.8	RADIO ASTRONOMY, EARTH EXPLORATION SATELLITE, SPACE RESEARCH
31.8	- 32	RADIONAVIGATION
32	- 33	RADIONAVIGATION, INTER-SATELLITE
33	- 33.4	RADIONAVIGATION
33.4	- 36	RADIOLOCATION

FREQUENCY GHz		SERVICE
36	- 37	FIXED, MOBILE, EARTH EXPLORATION SATELLITE, SPACE RESEARCH
37	- 38.6	FIXED, MOBILE
38.6	- 39.5	FIXED, MOBILE, FIXED-SATELLITE
39.5	- 40.5	FIXED-SATELLITE, MOBILE-SATELLITE
40.5	- 42.5	BROADCASTING-SATELLITE, BROADCASTING, Fixed, Mobile
42.5	- 43.5	FIXED, FIXED-SATELLITE, MOBILE except aeronautical mobile RADIO ASTRONOMY
43.5	- 45.5	FIXED-SATELLITE, MOBILE-SATELLITE
45.5	- 47	MOBILE, MOBILE-SATELLITE RADIONAVIGATION, RADIONAVIGATION-SATELLITE
47	- 47.2	AMATEUR, AMATEUR-SATELLITE
47.2	- 50.2	FIXED, MOBILE, FIXED-SATELLITE
50.2	- 50.4	EARTH EXPLORATION-SATELLITE FIXED, MOBILE, SPACE RESEARCH
50.4	- 51.4	MOBILE, FIXED-SATELLITE MOBILE-SATELLITE
51.4	- 54.25	EARTH-EXPLORATION-SATELLITE SPACE RESEARCH, RADIO ASTRONOMY
54.25	- 58.2	EARTH EXPLORATION-SATELLITE FIXED, INTER-SATELLITE, MOBILE SPACE RESEARCH
58.2	- 59	EARTH-EXPLORATION-SATELLITE SPACE RESEARCH, RADIO ASTRONOMY
59	- 64	FIXED, INTER-SATELLITE, MOBILE RADIOLOCATION
64	- 65	EARTH EXPLORATION-SATELLITE SPACE RESEARCH, RADIO ASTRONOMY
65	- 66	EARTH EXPLORATION-SATELLITE SPACE RESEARCH, Fixed, Mobile

FREQUENCY GHz		SERVICE
66	- 71	MOBILE, MOBILE-SATELLITE RADIONAVIGATION, RADIONAVIGATION SATELLITE
71	- 74	FIXED, FIXED-SATELLITE, MOBILE SATELLITE, MOBILE
74	- 75.5	FIXED, FIXED-SATELLITE, MOBILE
75.5	- 76	AMATEUR, AMATEUR-SATELLITE
76	- 81	RADIOLOCATION
81	- 84	FIXED, MOBILE, FIXED-SATELLITE
84	- 86	FIXED, MOBILE
86	- 92	EARTH EXPLORATION-SATELLITE RADIO ASTRONOMY, SPACE RESEARCH
92	- 95	FIXED, MOBILE, FIXED-SATELLITE RADIOLOCATION
95	- 100	MOBILE, MOBILE-SATELLITE, RADIONAVIGATION, RADIONAVIGATION-SATELLITE, Radiolocation
100	- 102	EARTH EXPLORATION-SATELLITE SPACE RESEARCH FREQUENCY SERVICE
102	- 105	FIXED, MOBILE, FIXED-SATELLITE
105	- 116	EARTH EXPLORATION-SATELLITE RADIO ASTRONOMY, SPACE RESEARCH
116	- 126	FIXED, INTER-SATELLITE, MOBILE EARTH EXPLORATION-SATELLITE, SPACE RESEARCH
126	- 134	FIXED, INTER-SATELLITE, MOBILE RADIOLOCATION
134	- 142	MOBILE, MOBILE-SATELLITE, RADIONAVIGATION, RADIONAVIGATION-SATELLITE, Radiolocation
142	- 144	AMATEUR, AMATEUR-SATELLITE
144	- 149	RADIOLOCATION
149	- 150	FIXED, FIXED-SATELLITE, MOBILE

FREQUENCY GHz	SERVICE
150 - 151	EARTH EXPLORATION-SATELLITE FIXED, MOBILE, FIXED-SATELLITE SPACE RESEARCH
151 - 164	FIXED, MOBILE, FIXED-SATELLITE
164 - 168	EARTH EXPLORATION-SATELLITE RADIO ASTRONOMY, SPACE RESEARCH
168 - 170	FIXED, MOBILE
170 - 174.5	FIXED, INTER-SATELLITE, MOBILE
174.5 - 176.5	EARTH EXPLORATION-SATELLITE FIXED, MOBILE, SPACE RESEARCH INTER-SATELLITE
176.5 - 182	FIXED, INTER-SATELLITE, MOBILE
182 - 185	EARTH EXPLORATION-SATELLITE RADIO ASTRONOMY, SPACE RESEARCH
185 - 190	FIXED, INTER-SATELLITE, MOBILE
190 - 200	MOBILE, MOBILE-SATELLITE, RADIONAVIGATION, RADIONAVIGATION SATELLITE
200 - 202	EARTH EXPLORATION-SATELLITE FIXED, MOBILE, SPACE RESEARCH
202 - 217	FIXED,FIXED-SATELLITE, MOBILE
217 - 231	EARTH EXPLORATION-SATELLITE RADIO ASTRONOMY, SPACE RESEARCH
231 - 235	FIXED, FIXED-SATELLITE, MOBILE Radiolocation
235 - 238	EARTH EXPLORATION-SATELLITE FIXED, FIXED-SATELLITE, MOBILE SPACE RESEARCH
238 - 241	FIXED, FIXED-SATELLITE, MOBILE Radiolocation
241 - 248	RADIOLOCATION
248 - 250	AMATEUR, AMATEUR-SATELLITE
250 - 252	EARTH EXPLORATION-SATELLITE SPACE RESEARCH
252 - 265	MOBILE, MOBILE-SATELLITE, RADIONAVIGATION, RADIONAVIGATION-SATELLITE
265 - 275	FIXED, FIXED-SATELLITE, MOBILE, RADIO ASTRONOMY
275 - 300	FIXED, MOBILE
300 - 400	(Not allocated)

the environments are in MIL-HDBK-235() entitled "Electromagnetic (Radiated) Environment Considerations for Design and Procurement of Electrical and Electronic Equipment"[2].

2.2 ENVIRONMENT SOURCES

2.2.1 General Signal Characteristics

EM signals are often classified as being narrowband or broadband types. Since "narrow" and "wide" are relative terms, many signals cannot be categorized this conveniently, but the narrowband/broadband division is useful for understanding and analyzing signal effects.

A narrowband signal is obviously one that occupies a very small portion of the radio spectrum. Such a signal is a continuous sine wave, since the energy in a sinusoidal wave is concentrated at a single frequency. Most communications transmitters in which the carrier is modulated by voice sideband, or in which suppressed-carrier techniques are employed, are classed as narrowband emitters. Their emission is usually confined to only tens or hundreds of kilohertz of bandwidth. Single-channel AM, FM and SSB transmitters fall in this category, as do some multiplex analog and digital systems.

Other types of emissions can be classed as narrowband emissions. The harmonic outputs of narrowband communication transmitters are narrowband. Prime power signals and their harmonics, communication receiver local oscillators, fire-control system CW illuminators, radar altimeters, and many other functional man-made sources have narrowband emissions.

In contrast to narrowband emissions, a broadband signal may disperse its energy across tens or hundreds of Megahertz or more. This type of distribution results from signals that are composed of narrow pulses having relatively short rise and fall times. The randomness with which the pulses occur and the individual characteristics of the pulses involved dictate the spectral distribution that will result.

Broadband signals can be further divided into random and impulsive sources. Random sources consist of closely spaced EM impulses that are not clearly distinguishable from one another. The impulses are frequent and overlap, with sharp peaks exceeding the average level. This type of output may be intentionally generated by a noise jammer, may be unintentionally produced by a communication or radar transmitter at frequencies far removed from its operating frequency, or may be contributed by nature in such forms as galactic and solar noise. Impulse sources are characterized by sharp pulses that are relatively infrequent and clearly separated. This type of output may be generated by internal combustion engine ignition systems, power line discharges, motor brush sparking, gas tube charges, and other electrical or electromechanical devices.

The frequency spectrum associated with either random or impulse energies is generally extremely broad. The magnitude of broadband emissions is generally specified in terms of microvolts per MHz. Random and impulse interference can be distinguished by the response of a receiver. For random interference, both peak and average levels are proportional to the square root of the receiver bandwidth. For impulse interference, peak voltages are proportional to the receiver bandwidth, whereas average voltages are independent of the receiver bandwidth.

There are a variety of natural and man-made sources of broadband interference. Natural interference is referred to as static. It produces the familiar noises that can sometimes be heard in a radio receiver. This natural static originates in thunderstorms, galactic sources, and the sun. Man-made broadband noise arises from numerous electrical devices. It is especially significant in cities and industrial areas. Man-made noise may also be particularly severe aboard ships.

Table 2-2 provides a list of examples of narrowband and broadband man-made interference sources [3]. The table has been subdivided to give a general indication of whether the source involved is in operation for long periods (continuous), for short periods (intermittent), or for very short and infrequent periods, (transient). The level of those sources on this list that can cause interference to GSE/Avionics is discussed later in this design guide.

2.2.2 Pulse Signal Characteristics

Pulsed radar is one type of broadband source that is often encountered in a GSE/Avionics topside environment. Pulsed emissions contain spectrum components that can encompass a very wide frequency range.

Figure 2-1 illustrates this effect. The figure shows a typical time representation of a pulse train, and its corresponding power spectrum as might be observed on an ideal spectrum analyzer. In this case, the envelope of the pulse in the time domain is a rectangle. The resultant frequency domain presentation follows a shape of:

$$\frac{\sin^2 x}{x^2}$$

Fourier analysis techniques can be employed to describe the spectrums of pulse signals having a variety of shapes. A usual way of describing the spectrum is in power in a given bandwidth (such as dBm per kilohertz). It is often convenient to describe the envelope of the spectrum (see Figure 2-1(B) and not the detailed lobe structure of the spectrum output.

The pulse spectrum envelope analytical representations of the rectangular pulse case discussed above, as well as the trapezoidal, cosine-squared, and rectangular with linear pulse compression cases are summarized in Table 2-3. As explanation, pulse compression is a radar technique to improve range resolution. It involves intentionally shifting the carrier frequency of the radar during its pulse interval.

TABLE 2-2
TYPICAL MAN-MADE INTERFERENCE SOURCES

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 Electric typewriters 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

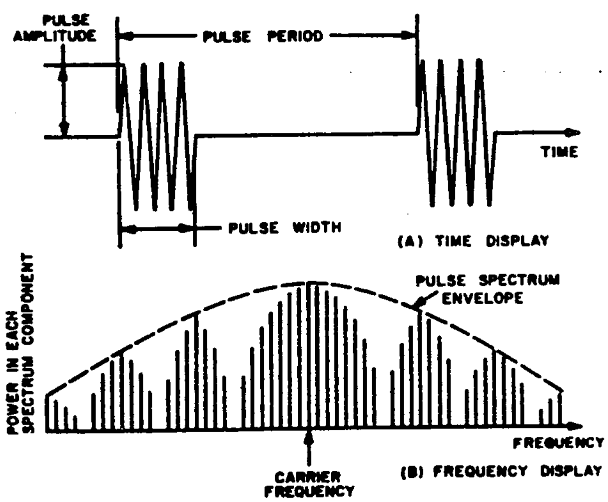
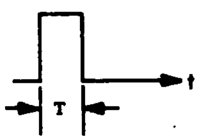
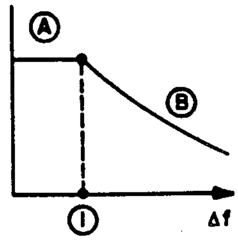
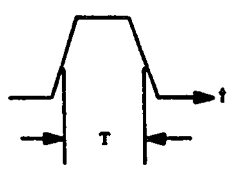
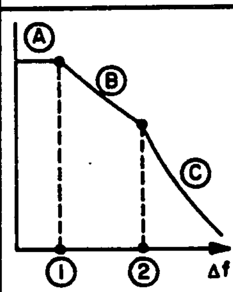
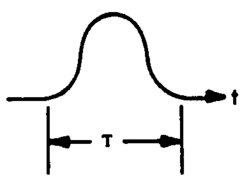
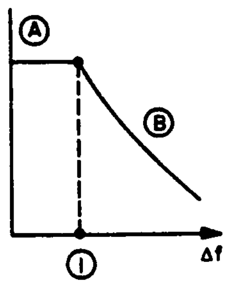
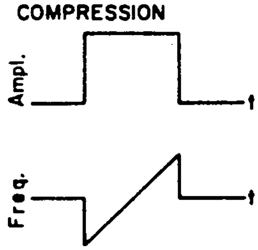
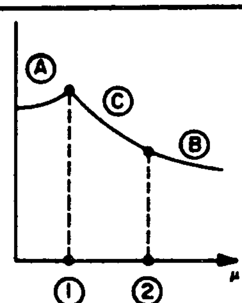


Figure 2-1
Pulse Spectrum Representations

TABLE 2-3 REPRESENTATIVE PULSE SPECTRUMS

PULSE CHARACTERISTICS	SPECTRUM ENVELOPE (positive side only shown)	POWER SPECTRAL DENSITY VALUES	FREQUENCY INTERCEPT VALUES
RECTANGULAR 		A $P = \hat{P}\tau^2/T$ B $P = \hat{P}/\pi^2 T \Delta f^2$	1 $1/\pi\tau$ P = Power per unit bandwidth, in watts per kilohertz $\hat{P}$ = Peak pulse power, in watts τ = Pulse width, in sec. T = Pulse period, in sec. f = Frequency displacement from carrier frequency, in kHz
TRAPEZOIDAL 		A $P = \hat{P}\tau^2/T$ B $P = \hat{P}/\pi^2 T \Delta f^2$ C $P = \dfrac{\hat{P}\left(\frac{1}{\delta_1} + \frac{1}{\delta_2}\right)}{4\pi^4 \Delta f^4}$	1 $1/\pi\tau$ 2 $\frac{1}{2\pi}\left(\frac{1}{\delta_1} + \frac{1}{\delta_2}\right)$ δ_1 = Pulse risetime, usec. δ_2 = Pulse decay time, usec.
COSINE-SQUARED 		A $P = \hat{P}\tau^2/T$ B $P = \dfrac{\hat{P}}{4T\pi^2\Delta f^2(1-4\ ^2\Delta f^2)^2}$	1 $\dfrac{\sqrt{\frac{9}{4} - \frac{\pi^2}{6} + 48\pi^2\tau^3}}{12\pi\tau^3}$
LINEAR PULSE COMPRESSION 	 $\mu = \dfrac{2\Delta f\tau}{D}$	A $P = \dfrac{\hat{P}\tau^2}{TD} \cdot \left(1 + \frac{1}{\pi\sqrt{D}} \frac{2}{1-\mu^2}\right)^2$ B $P = \dfrac{4\hat{P}\tau^2}{\pi^2 D^2 T} \cdot \left(\dfrac{1}{\left[\mu - \frac{1}{\mu}\right]^2}\right)$ C Log-linear interpolation	1 $\mu = 1 - \sqrt{\frac{2}{D}}$ 2 $\mu = 1 + \sqrt{\frac{2}{D}}$ D = Pulse compression ratio

Other pulse spectra data is provided in Figure 2-2 [3]. In this figure, the spectrum falloff curves have been normalized on the basis of area under each pulse. In terms of spectral occupancy, the advantage gained by increasing pulse rise and decay times is evident. Pulses having shapes approximating a Gaussian curve are relatively efficient in conserving spectrum space. For any pulse shape, emission levels at frequencies close to the carrier depend primarily on the area under the pulse. At frequencies further from the carrier, the levels depend on the number and steepness of the pulse slopes.

Actual spectrums generated by pulsed systems or transient signals often are considerably different from those created using Fourier spectrum synthesis techniques. Figure 2-3 is a spectrum analyzer photograph of a magnetron output for a 4.5 microsecond, 300 pulses-per-second pulse modulation signal [5]. Note its asymmetry and the breakup of lobes on the high side of the spectrum. Such differences are due to combinations of pulse amplitude and carrier frequency changes that actually occur and noise outputs that exist in the pulse generators. In the analysis, these are not taken into account and approximations are made of the actual waveshapes. Measurements of operational environment characteristics provide a much more reliable indication of the environment than attempting to describe the environment analytically.

2.2.3 Signal Transmission

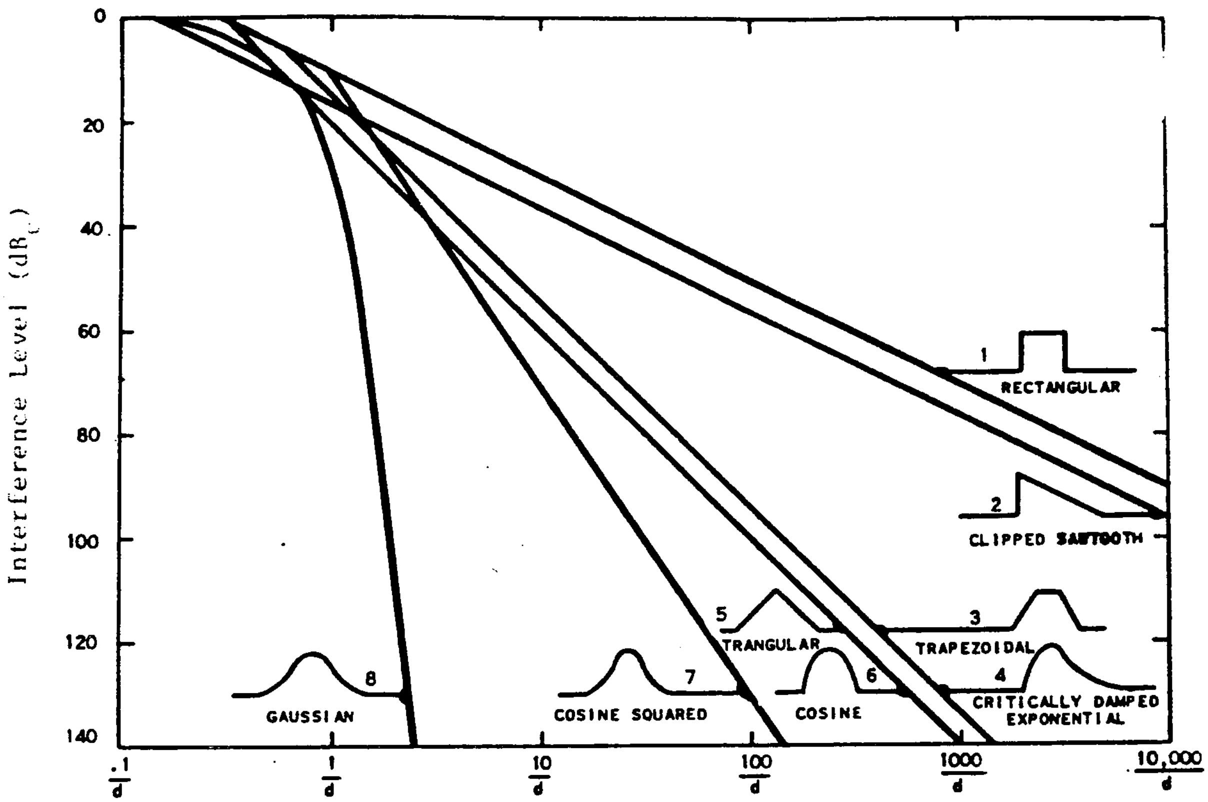
Interference signals may be transferred from one point to another by conduction, by radiation, or a combination of these. Conducted interference is transferred over common interconnections between a source and a receiver, and its propagation conforms to conventional circuit and transmission-line theory. Radiated interference is transferred via an EM field produced by a signal source, and its distribution conforms to field theory relationships. Figure 2-4 shows the conceptual differences between conducted and radiated paths [6].

Some transfer of signals may occur over a combination of conducted and radiated paths. For example, a source might couple energy onto a power cable by radiated means. The power system could then transfer this energy to various equipments tied to the power system by conduction. Similarly, coupling between two cables and the propagation of energy into and out of the cables may involve both radiation and conduction effects.

2.3 RADIATED ENERGY

2.3.1 The Radiation Field

The term radiated interference is commonly used to mean any interfering signal transferred through a medium by an EM field. The radiation field represents the energy which escapes from a source and spreads according to the laws of wave propagation.



A = Peak Amplitude of Pulse (Volts)
d = Average Pulse Duration (sec)
Ad = Area under Pulse
2Ad = Energy From Pulse on Both Sides of Carrier Frequency

Figure 2-2 Interference Levels for Various Pulse Shapes

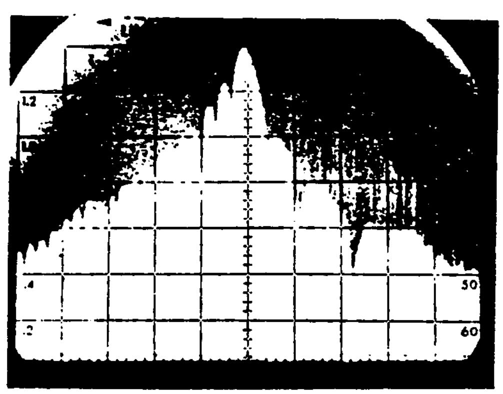


Figure 2-3 Spectrum Analyzer Photograph of Magnetron output when modulated by 4.5 μs, 300 pps pulse. Ordinate is linear in frequency; abscissa in logarithmic in level.

SOURCES

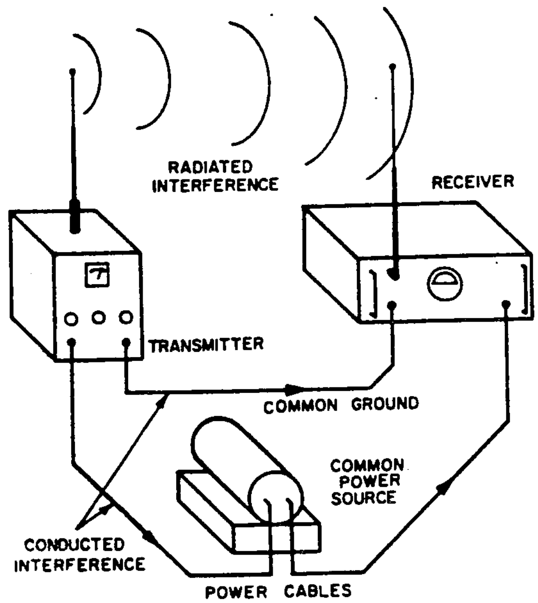


Figure 2-4 Propagation of Conducted and Radiated Interference

The radiation field has two components, an electric field (E field) and a magnetic field (H field). A varying electric field is always accompanied by a magnetic field. Conversely, a varying magnetic field is always accompanied by an electric field. The ratio of electric-to-magnetic fields is the wave impedance.

In the vicinity of a source of EM energy, the ratio of electric-to-magnetic fields is strongly dependent upon the point where it is observed. The "loop" formed by a conductor carrying large current variations into a load from a low-impedance source, with the return current flowing through a ground plane is one common source (Figure 2-5A). This produces a strong H field and a relatively weak E field, unless the source voltage is exceptionally high. In the vicinity of this loop, the field impedance is very low, because E is small relative to H, E/H is low. As the distance from the loop increases the H field weakens more rapidly than the E field. Their ratio changes with distance until a free-space value for E/H of 120 π is reached. Beyond this distance, the field impedance remains constant, at 120 π ohms unless disturbed by interfering objects or a change in medium. For practical applications, 377 ohms is used instead of 120 π ohms.

Figure 2-5B shows a conductor connecting a high-voltage, low-current ac-generator to its load. Here again, the return current flows through the ground plane. In this example, the H field is weak, because the current is low. The E field is strong, because the voltage per meter (the voltage gradient) between the conductor and the ground plane is high. In the vicinity of this source, the field impedance is very high, because E is large relative to H, E/H is high. Away from this source, however, the E field weakens faster than does the H field, and E/H approaches 120 π ohms.

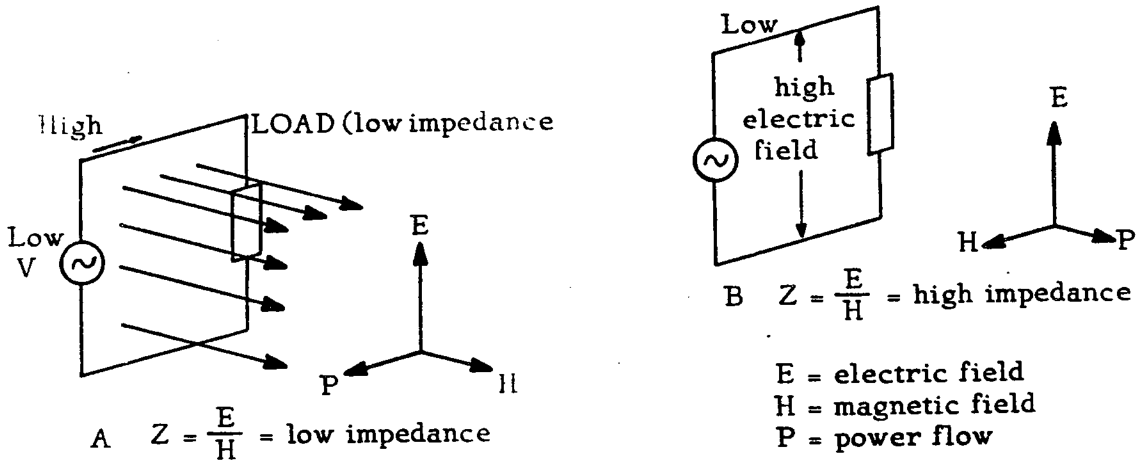
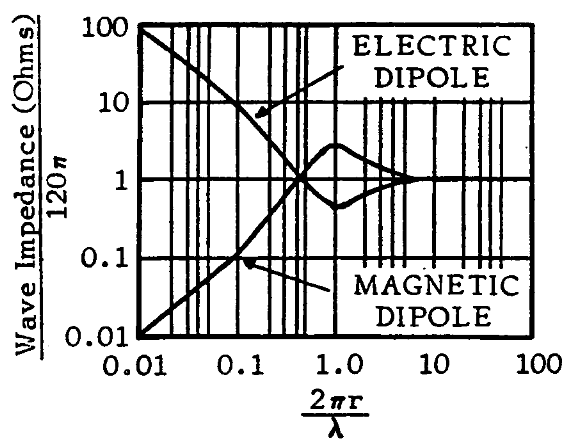


Figure 2-5 Sources and Their Fields

The distance at which the wave impedance approaches 120 π ohms is approximately one wavelength.

Figure 2-6 shows how the wave impedance for the cases of Figure 2-5 varies with distance r from the source. At distances beyond one wavelength, λ, from the source, the values are close to the free space value of 120 π ohms.



where: r = distance between source and monitoring point in same units as λ

Figure 2-6 Wave Impedance vs. Distance From Dipole Source

For the radiation field, the ratio of electric-to-magnetic field is equal to the impedance of the medium in which the wave propagates. The impedance of free-space is 120 π ohms or approximately 377 ohms. In the radiation field, emissions are essentially a plane wave. The electric- or magnetic-field intensity is an inverse function of the distance from the source. The field intensity at any point in a radiation field is also a function of the size and orientation of the radiator.

In order for any body to be an efficient radiator, its electrical dimensions must be at least of the order of one-half wave-length. Some sources have copious amounts of energy within the low frequency spectrum, but little of it may be radiated because of short, ineffective antenna lengths. The performance of a device that must operate in the vicinity of the radiating source is a function of the directivity of the energy as it leaves the source, the losses incurred in propagating to the device, and the susceptibility of the device to the particular characteristics of the energy.

The problem of determining the radiated signal level at a point in space some distance from a source through analysis, is not a simple task for several reasons. It is often necessary to establish the effective gain of the antenna in the direction of the observer when the observer is not in the antenna mainbeam. In addition, the characteristics of the antenna change with frequency. The effective gain of the antenna at a frequency of interest outside of the design range of the antenna (such as at the third harmonic of the source) may have to be considered if a device in the EME may be susceptible to that frequency. The difficulty may be further compounded because many antennas either physically or electrically change their directional characteristics (such as a scanning antenna) so that the effective gain in a given direction varies as a function of time. And finally, if there are any conducting materials in the neighborhood of the antenna, such as ship superstructure, these materials can introduce reflection effects that may alter the radiation field. When the radiator is not a well-defined structure like an antenna, but is an equipment case, or a cable located in a harness, then the problem of determining the radiated level at some point in space becomes even more complex.

With very limited exceptions, it is necessary to perform measurements to obtain even approximate indications of expected levels. Measurements only give an indication of the levels and cannot be considered exact. Everything in the field perturbs the field and each item introduced changes the field further. This includes the measuring equipment, GSE, aircraft, and people. Since these items are not constant, neither is the field.

2.3.2 Distinction Between Induction and Radiation Fields

The mathematical description of the total fields around a conductor carrying a varying current is very complicated and cannot be solved except in very simple cases. However, the fields at a remote point due to the uniform current flow in a short, thin wire (or a very small, circular loop) have three types of terms. These terms vary with $1/r^3$, $1/r^2$ and $1/r$ (where r is the distance from the source to the point of observation). The relative importance of the various terms depends upon the ratio of r to the wave-length of the radiation, as shown in Figure 2-7.

From Figure 2-7, it can be seen that, when r/λ is much greater than $1/2\pi$, the radiation ($1/r$) field is predominant. When r/λ is equal to $1/2\pi$, the induction components ($1/r^2$, $1/r^3$) and the radiation ($1/r$) field are equal. When r/λ is much smaller than $1/2\pi$, the radiation ($1/r$) term is negligible.

A generally accepted rule is that the radiation ($1/r$) field of the antenna main beam is where r is greater than $2D^2/\lambda$, where D is the maximum aperture dimension of the antenna. D is thus the diameter of a circular aperture or the diagonal of a rectangular aperture. For interference analysis purposes, it appears that r can be considerably less than this value (perhaps as low as $r \leq D^2/\lambda$) before significant error is introduced by using the $1/r$ relationship [6]. As one moves out of the antenna main beam, D becomes much smaller (because it is actually a function of the effective antenna aperture in the direction of the point of interest), and the $1/r$ relationship becomes valid at much closer separations, but the gain of the antenna decreases.

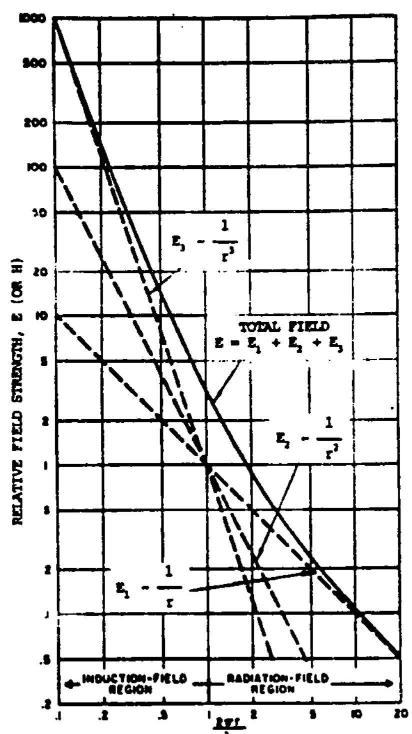


Figure 2-7 Variation of Field Strength With Distance For Small Dipole Antenna

2.3.3 Nonlinear Environmental Effects

The characteristics of high RF power and close equipment proximity in a shipboard environment have resulted in selected elements of the ship acting as unique EM interference sources. These sources of interference are located topside. They consist of those portions of the ship that, under high signal conditions, act as nonlinear electrical devices. An example of a nonlinear device is an amplifier when the input is overdriven. The output does not increase linearly (1:1). A single signal is compressed. Multiple input signals produce distortion products.

Environmental nonlinear sources aboard ship consist of nonlinear resistance junctions created at metallic mating surfaces. These junctions may be formed by salt or chemical deposition, by corrosion, or through a number of other means. The nonlinear junction acts like a mixer device and has the potential of creating distortion called intermodulation. Some GSE/Avionics may have to operate in environments containing contributions from these sources.

Intermodulation produces sum and difference products during transmissions involving two or more signals of different frequencies. If n frequencies are transmitted, the frequencies of intermodulation products formed in a nonlinear element can be determined by the relationship:

$$f_i = pf_1 \pm qf_2 \pm rf_3 \pm \ldots \pm zf_n$$

where

f_i is an intermodulation frequency,

p, q, r...x are integers,

and the sum (p + q + r +...+ z) defines the intermodulation order. Due to the symmetrical nature of the nonlinearities, odd-order intermodulation products and harmonics are much more pronounced than even order. Therefore, third-order intermodulation products and the third harmonic are normally the strongest signals generated by environmental nonlinearities.

The effects of nonlinear environmental sources have been most pronounced in the HF (2-30 MHz) frequency range. This phenomenon may change as more powerful emitters are produced at higher frequencies.

As the number of transmitters in the environment is increased, the number of possible intermodulation products generated rises rapidly. With three signals being radiated, 10 third-order products can be generated. With ten signals being radiated, approximately 800 third-order products can be generated.

In a study of this effect, metallic lifelines were found to be one of the major nonlinear environmental sources [7]. The junction was formed either at the turnbuckles or the eyelets in the line. The line acted as a fairly efficient radiating element.

Another series of sources were associated with the deck expansion joints on an AVT (World War II carrier). Included was a vertical copper boot bridging the gap created by the joint. Also included were cables and other structures crossing underneath the joint's covering plates. Other sources included mooring and anchor chains, slap-down plates, pipe and bracket joints, hatch hinges, life-raft hangers, ladders, armored cables, antenna guy wires, guard rails, booms, metallic frames and supports, gang planks, and similar structures.

2.3.4 Propagation Effects

Radiation involves the transmission of EM energy through space. The propagation characteristics are affected by the phenomena which occur in the transmission medium. Since the space above the earth's surface is used as the transmission medium, the radiated energy is affected by its characteristics.

The various modes of propagation discussed in this section are applicable primarily to situations where an interference source is a considerable distance away from its potential victim. This is not the normal situation with most GSE. EMI situations with GSE normally involve close-in equipments and devices. This summary is intended as introductory material for the limited number of cases where the effects of distant equipments are of concern.

Propagation, particularly sky wave propagation (waves reflected from the ionosphere), is a complex subject concerned with many variables. These include (1) the focusing effect of the ionospheric layers, (2) fading due to multipath propagation, (3) polarization fading, (4) loss of energy due to ground and ionospheric absorption, and (5) losses due to dust, water, gasses, and the dielectric of the media.

In the LF and VLF frequency bands (2-300 kHz), wave propagation takes place via ground-wave. At these frequencies, the ionosphere absorbs rather than reflects any radiation. EM radiation is propagated along the earth's surface. Ground wave paths may be used over long distances.

The MF band (0.3-3 MHz) is transitional in nature. The ground-wave is the primary propagation mode at the lower frequencies. Some skywave propagation occurs at the higher frequencies.

HF band (2-30 MHz) propagation may be either by ground or sky waves. The mode of propagation depends on the equipment physical configuration, frequency, and ionospheric conditions. In the HF band, ground-wave propagation proves the most reliable, but the distance coverage is limited to approximately 100 miles. On the other hand, skywave propagation has an unlimited range, depending upon the parameters outlined above.

As frequencies increase above 30 MHz, the energy reflected from the ionosphere decreases and eventually disappears. Above the HF band, transmission depends mainly upon free-space-propagation. In the VHF and UHF bands, 30 MHz to 300 MHz and 300 MHz to 3 GHz, respectively, propagation beyond the horizon is possible due to atmospheric scattering.

For frequencies above 3 GHz, skywave propagation is not practicable due to the extremely short wavelengths involved. Specifically, no focusing effect of the ionospheric layers occurs, and absorption losses due to dust, matter and gases become more pronounced.

The general characteristics of the most commonly encountered propagation modes are listed in Table 2-4, along with information on effective frequencies and ranges, fading and other variational characteristics, and link reliability.

2.3.5 The GSE/Avionics Radiated Environments

The major topside sources of potential radiated interference to GSE/Avionics come from the array of communications and radar equipments aboard ship. Interference from nearby ships must also be considered in the design of GSE/Avionics. If operation of the GSE/Avionics is expected while in port, shore based emitters are also of concern.

A study of the complement of electronic radiators aboard a particular ship will provide an immediate indication of those frequency bands where high field strengths will probably be encountered. Distances and direction from an antenna are useful in determining probable areas of high field strengths. Use caution, the direction of maximum energy from a particular source may not correspond to the optical direction to the source. This is due to super-structure masking and reflections from masts, stacks, and enclosed above-decks spaces and the deck itself. Additionally, the GSE may have to operate in the induction field of an antenna because of the dimensional restriction and general structure of the ship.

Field strengths on carrier flight decks and weather decks of other type ships could exceed 10 volts/meter in communication equipment operating bands with levels greater than 200 volts/meter from equipments operating in the HF band. The specific levels depend on location, transmitter power, antenna gain, reflections, and masking. Flight deck average power density levels in acquisition and tracking radar bands can be greater than 10 milliwatts per square centimeter. Peak power density levels of hundreds to thousands of times these values are possible. These signals will have the characteristics of either plane waves or high-impedance waves, so that electric field rejection techniques in GSE/Avionics design will apply.

The EMEs on the hangar deck and in avionics support equipment spaces can be expected to be somewhat lower than the flight deck environment. Nevertheless, these levels must also be taken into account when developing GSE/Avionics that must operate in this location. Levels in the HF range can exceed 1 volt/meter. Microwave levels could reach one milliwatt per square centimeter.

The major documents currently used for EMI control are MIL-STD-461C and MIL-STD-462. MIL-STD-461 defines conducted and radiated emission and susceptibility limits for a wide range of electrical/electronic equipment. MIL-STD-462 describes the test procedures to be used. These standards are usually prescribed in GSE/Avionics design contracts or agreements.

SOURCES

TABLE 2-4 [8]

SUMMARY OF CHARACTERISTICS OF SELECTED PROPAGATION MODES

PROPAGATION MODE	FREQUENCY (MHz)	DISTANCE (mi)	MECHANISM	FADING	DIURNAL VARIATIONS	SEASONAL VARIATIONS	AVAILABILITY RELIABILITY
Line-of-sight links	No definite limits. Most useful above 50	Up to 30-50 per hop; no fixed upper limit on system length	"Free-space" (1/r) propagation, modified by multipath	1. Fast, due to multipath 2. Slow, due to ray bending	Day: Std. atmos. Night: [illegible] get temp. inversion fading	Less fading in winter	Approaching [illegible]
Obstructed-path links	No definite limits. Most useful (est.) 50-5000	Up to 200 per hop (est.)	Knife-edge diffraction	Similar to that for line-of-sight paths, but perhaps slightly less severe	±3 dB for no reflection multipath; +10 dB for reflection multipath		Approaching 100%
Tropospheric Scatter	40-10,000	<700-800	Scattering by blobs or reflection from layers in the atmosphere	1. Fast, due to multipath in atmospher. 2. Slow, due to change in average atmospheric properties	Slightly stronger in morning	Hourly median varies with $\sigma \cong 3$ to 8 dB about yearly median. Signal strongest in summer, weaker in winter.	Approaching 100%
Ionospheric Reflection	0.03-80 Most useful: 0.08-0.2; 2-30	No limit	Reflection from layers of high ionization density in the ionosphere; coupled with reflection from the ground for ranges greater than about 2500 miles.	1. Rapid multipath fading 2. Polarization fading ≅ 3 dB 3. Skip fading (when f is close to Max. Unable Frequency 4. Absorption	Max. Usable Frequency varies greatly being lower at night, highest at noon. Absorption highest during day, noise highest during night.	Max. Usable Frequency, absorption, and noise levels shift over the earth's surface with the seasons.	Typically ≅ 90% or less, considering absorption, noise, multipath and skip fading, and shift of Max. and Lowest Usable frequency. Ranges from about 60% in the auroral zones to about 90% in the equatorial zone.
Ionospheric Scatter	30-60	< 1400 Most useful: 600-1200	Scattering by D layer of ionosphere	1. Fast due to multipath (.2-5 Hz) 2. Slow due to variation in height of reflecting layer	10-12 dB variation. Strongest at noon; weakest at 8 p.m.	Hourly median varies with $\sigma \cong 6$-8 dB about yearly median. Signal strongest in June, July, weakest in Sept.	≅ 95%
Meteor-burst	25-75	200-1000	Electron trails due to ionized meteor material	Extremely rapid (up to ≅ 500 dB/sec)	Maximum possible information rate may vary by 3 to 1; Max. at 6 a.m., Min. at 6 p.m.	Meteor rate may vary by 10 to 1. Max in Summer, Min. in Spring (Northern Hemisphere)	≅ 95% availability. When mode is available, duty cycle is approx. 5 to 10%.
Surface Wave	0.010-0.300 for long range (up to ≅ 500 mi or more). Up to 3 short ranges	< 1200 (greatest over sea water, least over poor soil	Wave guided by interface between earth and air	Multipath with skywaves can occur at ranges beyond about 400 miles	Only very slight due to slight variations of earth's electrical constants (e.g., due to freezing)	Only slight due to slight variation of earth's electrical constants	≅ 95% (suffers from atmospheric noise, especially at low latitudes)
Active Satellites	100-10,000	No limit with multi-satellite system	Receiver-Transmitter in satellite	1. Movement of antenna beam 2. Doppler shift 3. Faraday fading (≅ 0, f >1000 MHz)	None except due to motion in orbit	None except due to motion in orbit	Approaching 100%
Passive Satellites	1000-10,000	No limit with multi-satellite system	Passive reflector	1. Movement of antenna beam 2. Doppler shift	None except due to motion in orbit	None except due to motion in orbit	Approaching 100%
ELF and VLF	Most useful: 0.01-0.03	No limit	Ducting within the waveguide formed by the ionosphere and the surface of the earth	Very slow fading due to 1. Diurnal or Abnormal changes in ionospheric conditions 2. Multipath interference with surface waves	Slightly stronger at night	Slightly stronger in winter	90% to 99.9%

MIL-STD-461C limits often are not compatible with the environment in which the GSE/Avionics must operate. This is particularly true in the case of susceptibility requirements for carrier flight deck equipment. These levels should be established based upon the intended emission levels that can be found in MIL-HDBK-235-2 (NAVY)[2]. Although conducted susceptibility tests for signal lines and broadband radiated susceptibility tests are addressed for EMP testing (CS10, CS11 & RS05), there may be a need to address similar tests in MIL-STD-461C to simulate more commonly encountered EMEs.

Figure 2-8 presents sample E-field data of a CV hangar bay EME [9]. The information in the figure is not representative of worst case conditions. It is provided only to show the general trend of the environment at that location. Each dot represents the maximum level narrowband signal located over the frequency band in the vicinity of the dot. The level of the maximum data point in the HF range corresponds to a level of 0.6 volts/meter. The MIL-STD-461C narrowband radiated emission limit (test RE02) is also shown in Figure 2-8 for comparative purposes. The RE02 limit of MIL-STD-461 should be considered in specifying GSE/Avionics.

Similar data is provided in Figure 2-9 for an avionics shop environment. Note the generally lower levels involved when compared with Figure 2-8. The level of the maximum data point in this case is 200 microvolts/meter.

Radiated narrowband magnetic field data in the avionics workshop is plotted in Figure 2-10. For the case illustrated, it will be noted that GSE compliance to the radiated susceptibility threshold of MIL-STD-461C, test RS01, would not necessarily assure satisfactory performance in this environment, due to signal levels exceeding the threshold. The RS01 limit of MIL-STD-461 should be considered in specifying GSE/Avionics.

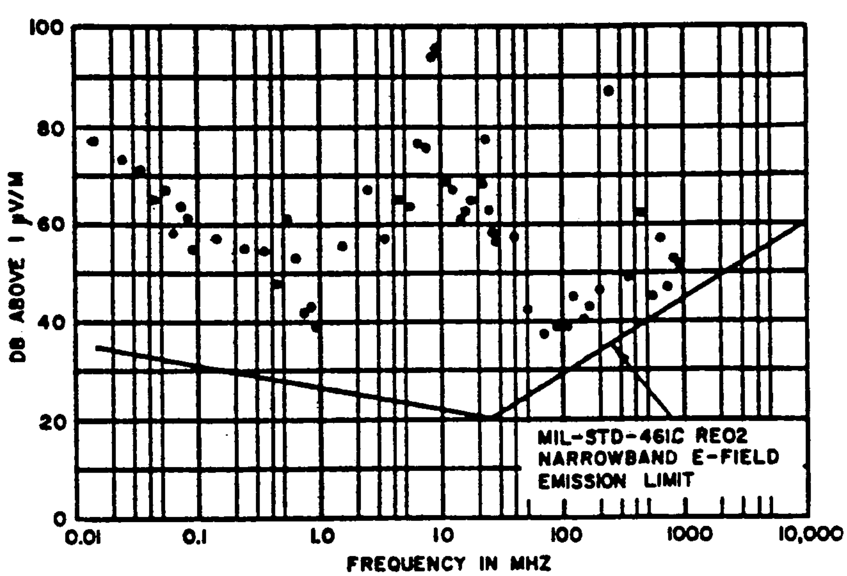


Figure 2-8 Narrowband Radiated Emissions Measured in Forward Hangar Bay Aboard CV-62, Port Side, Amidships Sections.

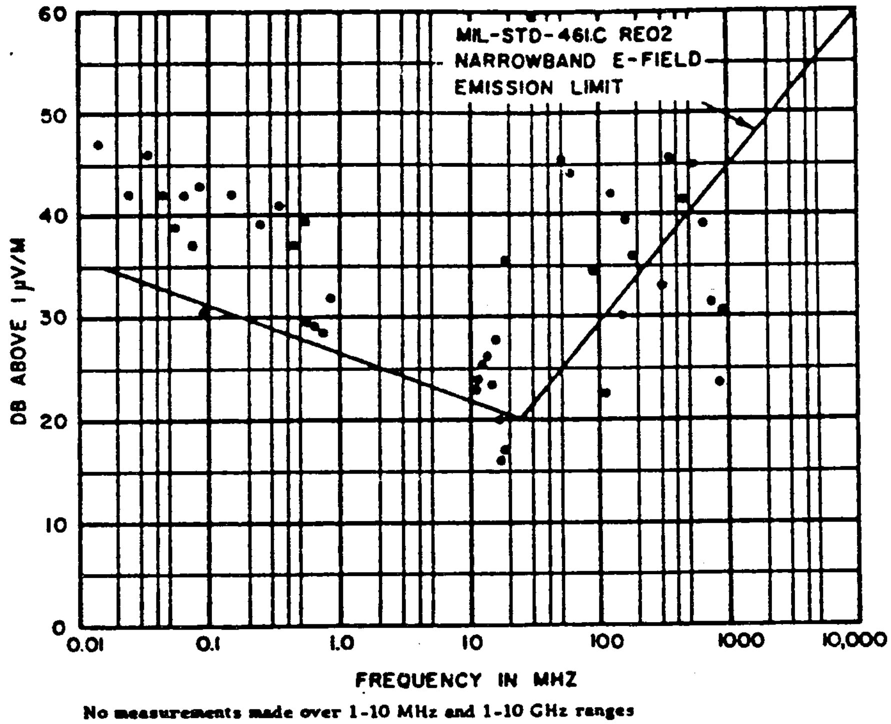


Figure 2-9 Narrowband Radiated Emissions Measured in Avionics Workshop #1 Aboard CV-62

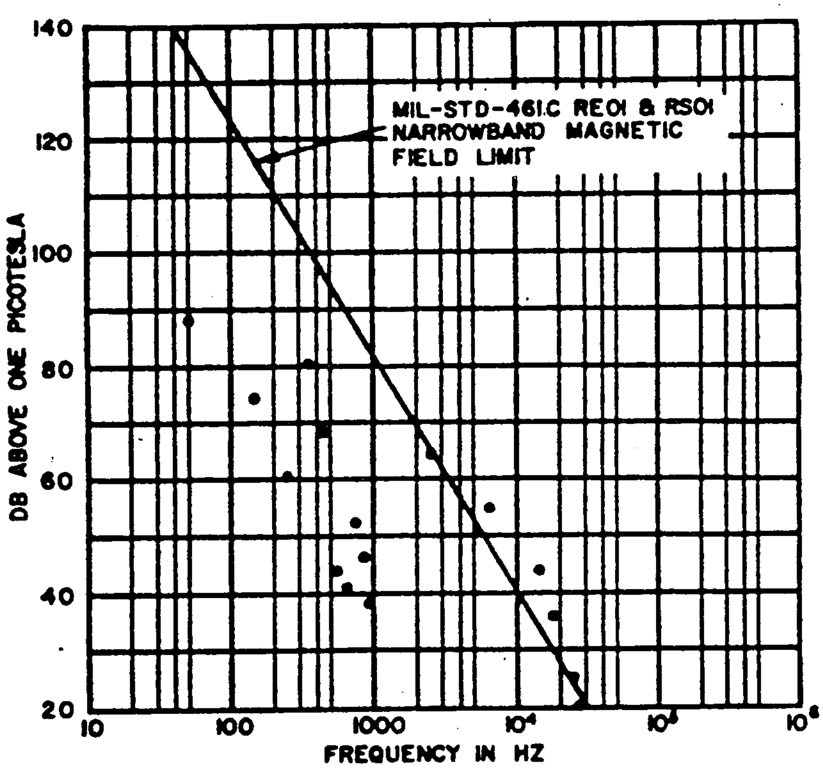


Figure 2-10 Narrowband Radiated Emissions Measured in Avionics Workshop #1 Aboard CV-62

SOURCES

2.4 CONDUCTED ENERGY

2.4.1 Conducted Routes

Conducted interference is the interference that propagates through a conductor such as wiring or any conducting structure. The conductor can directly transfer interference from one point in a circuit to another. It can also be the pickup mechanism for converting radiated interference to conducted effects.

Conducted interference requires a complete circuit path between an interference source and a victim. This may be formed by wiring, chassis, structure, or mutual inductances or capacitances. A common path for conducted interference is a ground plane. Conducted interference is especially severe when several electronic devices are connected to a common power source that has insufficient RF isolation. This mode of conducted interference may be transferred to each device because of mutual impedance in the ground path.

A conduction current may flow between circuits whenever there is a direct connection and a return path between two circuits. The return path may be another conductor, a mutual capacitance or inductance, or a common ground return. In general, the magnitude of resulting current depends on the difference of potential to ground between the points of exit and entry in the exciting circuit and on the total loop impedance between these points. Positions along the line where interference can enter and exit from the system become important when the dimensions of the circuits involved become large compared with the wavelength of interest. This situation may occur when the prime power distribution system is involved.

There are three predominant routes for conduction currents. They are power supply leads, control and accessory cables, and ground returns. These paths may introduce interference to GSE/Avionics. Particularly susceptible are those that must operate from power external to the GSE.

Signals can also be coupled onto equipment wires and cables by inductive or capacitive coupling. Inductive or capacitive coupling can occur between adjacent unshielded or inadequately shielded wires or cables, or internal wiring in equipment, where the return paths for the system may be through the same ground systems. Circuits are said to be mutually coupled whenever currents or voltages in one circuit induce corresponding voltages or currents in another circuit.

Inductive coupling occurs when a conductor is present in an EM field set up by interference energy in a nearby conductor. It is important to note that mutual inductance exists only between complete circuits. Each is a complete loop. The return portion of the loop of the circuit may be far removed from the inductive field. In such a case, it is common practice to speak of the voltage induced into a wire.

Analysis shows that inductive coupling varies in a complicated fashion dependent upon the geometrical configuration of the circuits. At fairly close separations, mutual inductive coupling is comparable to ordinary transformer action. A varying current in one loop of wire sets up a varying field around that loop. The varying field induces a varying voltage and resulting current in a second loop of wire.

There are two predominant methods of inductive coupling of interference, through transformer and through parallel wire runs. Because of core losses, a transformer normally will pass low frequency ac signals and acts as a low-performance filter against high frequency interference.

The parallel wire arrangement represents a special case of a transformer consisting of single turn primary and secondary windings. When two such lines are feeding different pieces of equipment from different sources, interference voltages present in either line can be induced into the other line. This situation enables interference to affect any susceptible part of a system. For example, primary power leads located near a B+ lead may induce ac hum into the system via the dc circuit.

Capacitive coupling is the linking of one circuit with another by means of the capacitance existing between them. The degree of coupling depends on the voltage and the frequency of the interference as well as the value of the interconnecting capacitance. At low frequencies, an extremely large capacitor is required to couple the signal; at radio frequencies, only a small capacitance is required. It should be noted that a complete path is also necessary for this type of coupling. It is normally the common ground connection for both leads which provides the return path.

Multiconductor cables act as especially good capacitive couplers at radio frequencies. If interference is present on a single lead of a multiconductor cable, the interference may appear on all conductors of the cable.

2.4.2 The GSE Conducted Environment

The major mechanism of transmitting conducted EMI energy to GSE is via the prime power distribution system. This type of interference is a function of the types and designs of other equipments that are also operating from these lines, and their locations on the lines.

Some indication of the conducted interference currents that might be encountered aboard ship is shown in Figure 2-11. The data represents the average of the three current maximums measured on a three-wire 60 Hz power system terminating in an avionics shop. Significantly greater current levels were measured on individual lines. Similar data taken on the same power buses at other locations aboard this same ship indicate that the currents are location-dependent, particularly over the lower frequencies.

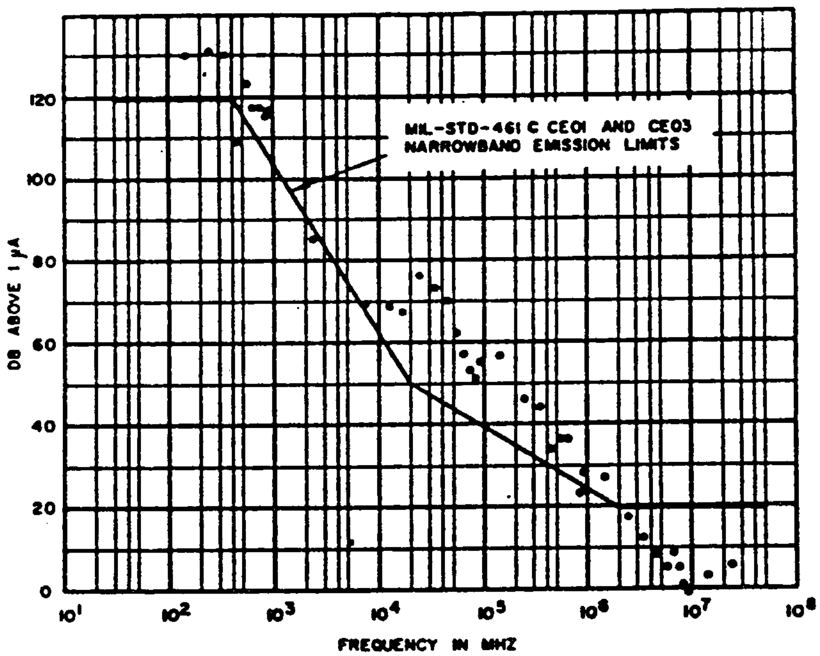


Figure 2-11 Narrowband Conducted Emissions Measured in Avionics Workshop #1 Aboard CV-62

2.5 COMBINED EFFECTS

Commonly, interference to GSE/Avionics results from a combination of radiated and conducted effects.

The wavelengths associated with the frequencies in the HF band (2-30 MHz) vary from 100 meters to 10 meters. The aircraft structure may act as a reasonably good antenna, because aircraft dimensions at these frequencies are a significant portion of a wavelength. This antenna-like behavior produces an RF voltage between the carrier deck and the aircraft structure. The voltage potential from aircraft to deck in a 100 v/m field may be as high as 250 volts RF [10]. Figure 2-12 shows the relative voltage distribution induced on a F6A aircraft [11].

The voltage pattern on an aircraft varies with such factors as radiation frequency, aircraft type, configuration, and orientation with respect to the antenna. Attaching a ground strap from the aircraft body to the deck does not eliminate the problem. This only references one point of the aircraft to the deck. The electrical length of the rest of the aircraft is still an appreciable fraction of a wavelength. A series of ground straps along the aircraft structure could cut down the effective electrical length presented by the aircraft to the HF. This is impractical due to the difficulties in achieving a low impedance RF bond between the aircraft and the deck.

SOURCES

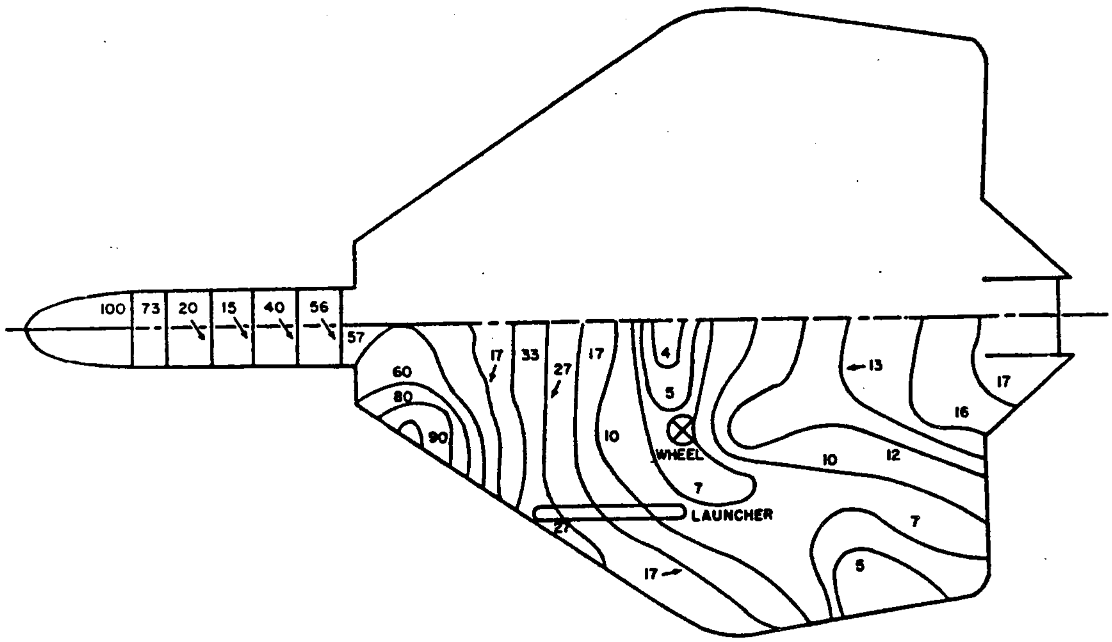


Figure 2-12 Voltage Distribution Induced On An F6A Aircraft

The whip antennas located on the carrier island and around the periphery of the flight deck are the primary sources of the fields producing the voltage differential. Since these whip antennas are non-directional and can be operated in a vertical or horizontal position, the emitter causing the potential to develop may not be obvious. Aircraft undergoing organizational level maintenance may be parked in close proximity to these antennas.

Figure 2-13 diagrams the coupling mechanisms involved. Many avionic systems use aircraft ground in their circuits. The aircraft structure serves as a ground return for the power supply in most aircraft. This "ground" may have an RF voltage with respect to the deck "ground." A GSE item placed on the carrier deck and connected by cabling to the aircraft avionics will appear as a relatively low impedance to deck to the RF energy on the aircraft. RF currents will flow down the cable through the GSE into the deck. RF currents may flow due to capacitive coupling even if the avionics circuit is isolated from aircraft ground or if there is no dc path to the deck.

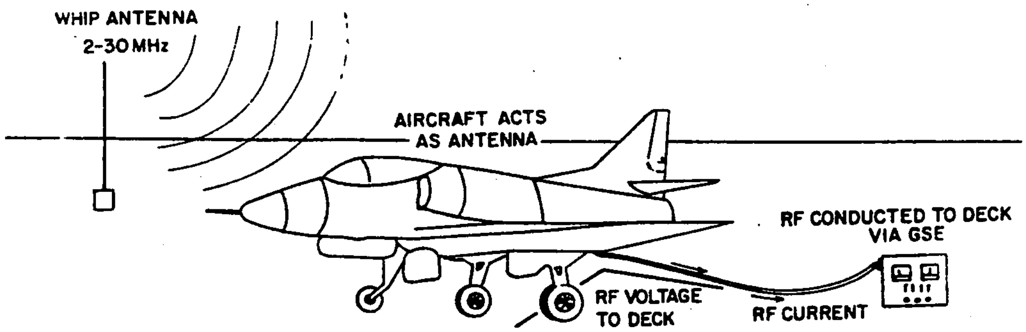


Figure 2-13 HF Coupling Mechanism Involved

Shielded cables, well terminated at both the aircraft and the GSE, provide a low impedance path to the deck. The RF currents will be confined to the outer portion of the shield due to skin effect. No coupling to the circuit lines will occur. Discontinuities in the shielding such as poor bonds, anodized connectors, or inadequate cable braid will cause increased impedance to the RF currents flowing on the shield. Coupling to GSE and avionics systems may then occur because of voltage drops at these discontinuities.

In specifying shielding for GSE, caution must be exercised. Most shielding effectiveness measurement methods use radiated fields to measure the combined reflection and penetration loss. In the case described, where large currents exist on the surface of the shields, only the penetration loss of a shield is effective. The most informative measure of shielding effectiveness for this type of coupling is the Surface Transfer Impedance (STI) test described in MIL-STD-1377.

If threshold signal levels are low, interference may result even if good shielding techniques are used. In this case and in cases where proper shielding techniques cannot be readily implemented, filtering may be necessary.

Mounting GSE on the aircraft can provide one means of reducing the possibility of EMI in the HF range. The difference of potential between equipment and the aircraft is greatly reduced or eliminated. The mounting surfaces must be clean, and all interconnecting cabling must be kept as short as possible.

REFERENCES

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3.0 DESIGN CONSIDERATIONS

3.1 GENERAL GUIDELINES

Chapter 2 outlined the factors that must be considered in the design of Avionics/GSE to achieve EMC.

Following the definition of requirements that a given piece of Avionics/GSE must fulfill, its design normally begins with a delineation of the electrical circuits needed to interconnect its components. Such components may consist of black-box units, integrated circuits, or individual components or piece parts. After the electrical aspects of the design have been done, the mechanical configuration of the equipment is established and detailed.

3.1.1 Electrical Design

An early step in any electrical design is the selection of signal levels. In selecting signal levels, consideration must be given to the radiated EMEs in which the equipment must operate. If high level radiated EMEs are expected to be encountered, such as those found on a carrier flight deck, large signal levels and low impedances should be used in the design. If high level radiated EMEs are not to be encountered, signal levels should be the lowest that are consistent with the achievement of the needed signal-to-system noise ratio throughout the equipment. Generally low voltage, low impedance design is desirable from the viewpoint of minimizing capacitive coupling problems, but may aggravate inductive coupling problems if current flows are too high.

Because interference often has a different spectral content from that of the desired signals, much can be done by the proper use of filters to constrain signals to their desired paths within the equipment, such as through common power supply leads. Chapter 7 discusses filtering in detail.

It should be noted here that care should be taken to employ high reliability EMI suppression components. To illustrate the importance of using high reliability components consider a filter by-pass capacitor that fails by opening. If the capacitor is a part of the equipment power supply filtering circuit, its failure will immediately be evident because of improper equipment operation. If it is a part of a filter to keep interference on the power line from entering the Avionics/GSE, it may be a relatively long time before evidence indicating the capacitor has opened is identified. The designer should pay particular attention, during component selection, to those elements of the system whose breakdown would not be immediately observable.

After the basic circuit design has been accomplished and the necessary filtering has been specified, attention must be given to capacitive coupling of conducted energy that may occur in leads carrying currents from differing sources. Since even a short piece of wire has an impedance, voltages are developed along wires and can cause effects not intended by the designer. This is especially true when attempts are made to return signal and power circuits through a common lead. Chapter 6, Grounding, discusses this aspect of the design.

3.1.2 Physical Layout of Components

After the basic circuit design has been completed and grounding requirements have been established, attention must be given to the physical layout of the components. Capacitive coupling of conducted energy and inductive coupling of radiated energy are of most concern here, but inductive coupling of conducted energy also must be considered. Often the desired amount of physical separation between sources of energy and sensitive elements cannot be achieved within the physical size constraints of the equipment. To reduce this problem of close proximity, one can use analytical techniques to determine the fields emanated, and then manipulate the values of certain elements to reduce these fields. A summary of this analysis and selection process follows.

The analysis is based upon the simple model of two PC board traces connecting the input and output terminals of a source device to the input and output terminals of a load device. One determines the vector potential at any point (r, θ, ϕ) above the PC board and takes the spatial derivatives; these are the magnetic and electric field components.

The vector potential at (r, θ, ϕ) is dependent upon the contributing vector potentials of the current segments defined by the traces and the paths through the source and load elements. The short current segments through the source and load devices are assumed to be uniform and are modeled as Hertzian dipoles. The traces are modeled as transmission lines and their contributions are the Integral of the elemental vector potentials over the length of the traces. This entire analysis can be accomplished with the use of a FORTRAN program. Due to the complexity of the mathematics involved, formulas have not been presented. The reader is directed, instead, to Reference [1].

The following are some practical component selections to reduce emanations [2]:

(a) minimize source voltage to reduce E.
(b) minimize the length of the traces to reduce H.
(c) maximize the load impedance to minimize currents thus reducing H.
(d) minimize the ratio of (distance between traces/trace characteristic source impedance) to reduce H. Reducing trace characteristic source impedance is aided by minimizing the width of the trace.

Other significant factors with respect to the layout of components internal to the case are:

(A) Isolation (electrical and physical) of all incoming circuitry (power lines, signal lines, and signal processing circuitry, etc.) from all internal circuitry should be accomplished by shielding entry areas and circuitry and by utilizing π-type or T-type feed-through filtering for lines into the internal circuitry.

(B) Electrical and physical isolation of internal interference generating circuitry from other circuitry should be accomplished.

3.1.3 Exterior Considerations

When the piece of equipment must operate in the presence of a strong external source, the problem of close proximity to radiation arises again. This makes shielding the internal circuitry necessary. To accomplish this, special consideration must be given to the equipment case in addition to cables, connectors and any other components external to the case. Some of these design considerations are:

Equipment Cases

Significant factors with respect to case construction are:

(A) The use of material with good shielding properties.

(B) The use of welded or overlapped seams.

(C) Secure bonding of panels, cover plates and aperture covers to the case, using gaskets with adequately spaced and properly tightened screws or bolts.

(D) The cleaning of mating surfaces immediately prior to assembly to ensure good electrical contact and minimize corrosion. The subject of bonding is discussed in Chapter 5.

Cables

(A) Cabling penetrating the case should be shielded and the shield terminated in a peripheral bond to the case at the point of entry.

(B) Cable shield grounds should be maintained separate from any signal grounds or circuitry grounds.

(C) Cable shields should be bonded peripherally to adapter and connector shells; cable shields should not be "pig-tailed off" and run through on connector pins.

Connectors

(A) Connectors should be of the type which make peripheral shield (shell) ground before the pins mate during the process of connection. The pins should disconnect before the shield (shell) separates.

(B) Pins of connectors leading to electronic circuitry should be, wherever possible, female. Otherwise, they should be recessed male pins so as to exclude contact with any portion of the shell of the mating connector or with fingers of the operator.

(C) Connector backshells should be selected based upon empirical data from EMI tests of the unshielded cable assembly.

With the unshielded cable and no backshells on the connectors, a frequency sweep should be performed over the range called out by the specification used.

Next, the amount and type of shielding required to reduce radiation to acceptable levels should be calculated and appropriate shielding selected. At this point, a backshell type should be selected which will interface with the cable and connector and terminate the cable shielding without significantly degrading the shielding effectiveness of the overall assembly.

Finally, the prototype cable assembly should be tested in the system configuration.

Shielding techniques, including techniques applicable to cables, connectors, and backshells are discussed in Chapter 4.

Components External to the Case

(A) EMI boots should be installed on toggle switches.

(B) EMI rotary shaft seals should be incorporated.

(C) Screening and shielding of meter faces and indicators should be accomplished.

(D) Shielded lights should be used.

(E) Shielded fuseholders should be used.

(F) Application of waveguide-below-cutoff techniques should be employed.

In general, Avionics/GSE protection against a high-power EME should be included in the design of the equipment from its inception and directed toward the elimination and exclusion of energy which will affect the operation of the equipment from the standpoint of both radiated and conducted interference. To achieve this objective, the techniques just outlined should be used. Chapters 4, 5, 6, and 7 provide detailed information about the application of these techniques.

3.2 SAFETY

3.2.1 The Effects of High RF Fields

High ambient EM levels present on a flight, hangar, or weather deck not only introduce unique design requirements on the equipments involved, but they also can create operator and equipment hazards as well.

Two classes of safety problems can be identified. One type is due to arc discharges occurring between GSE and the flight/weather deck, and/or between GSE and an aircraft. The other type is caused by large currents being induced in GSE/Avionics.

The arc discharges result because aircraft (with their accompanying Avionics) and GSE are often isolated from the flight or hangar deck, or from each other, and build up different levels of electric potential. These potential differences may arise under normal circumstances because of the incompatibilities of equipment ground systems, or because of different charge accumulations on different assemblies, but the problem can be severely aggravated by a high level EME. Under these circumstances, when an attempt is made to ground the test equipment to the deck, or to connect a test cable to an aircraft, a breakdown of the gap between the two charged assemblies occurs, and an arc results. When test cables between these same units are removed, an arc can be drawn in the process. Considerable damage can be caused to GSE or an aircraft due to the resultant surge current effects, and injury to the equipment operator could take place. The transient might also be capable of triggering ordnance devices such as fuses or detonators.

Environmental electric fields can couple to cables, discontinuities in shields, and other elements of GSE and GSE-interfaces. While a cable is not considered as an antenna in the general sense, it can nevertheless be a reasonably efficient mechanism by which radiated fields are transformed into electric currents. The amount of interference signal that might appear at some point along the cable is dependent on the standing-waves induced on the cable; this in turn is a function of the frequency and level of the environmental signal, the orientation of the cable relative to that signal, the way in which the cable is terminated, and other factors.

A good example of the dangers of high-level environmental energy coupling into Avionics occurred in a test of an A-7E aircraft in a simulated flight-deck environment. While the A-7E was running on auxiliary power (see Figure 3-1), a fire broke out in the

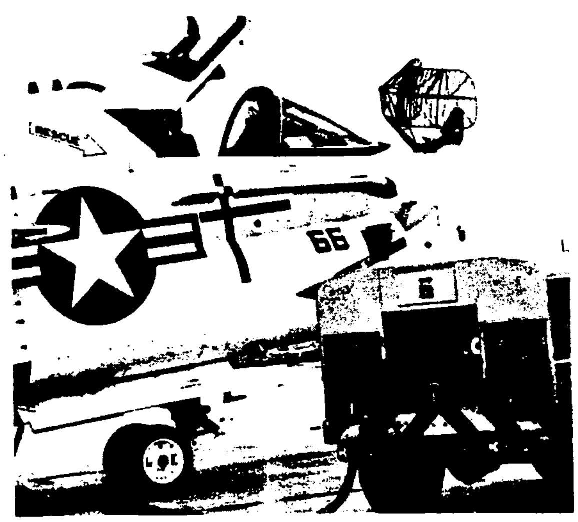

Figure 3-1 A-7E Aircraft on External Power, Simulated Flight Deck Environment

unit that switches the aircraft from external to internal power. Figure 3-2 shows the extent of fire damage to the switch. A subsequent analysis of the cause of the fire indicated that environmental energy was being coupled onto the auxiliary power cable to the aircraft, and the extraneous energy caused the relay to chatter and over-heat. Environmental magnetic fields can couple to any electrically conductive configuration that forms a closed loop. Such loops might include metal ferrules around plastic knobs, cable connector rings that are not screwed onto connectors, and many others. If the field intensity is high enough, and the loop is oriented appropriately, a high enough current can be generated in the loop to cause it to get hot.

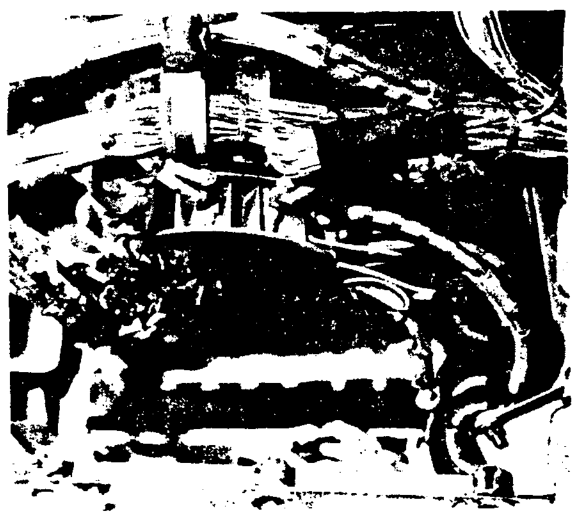

Figure 3-2 Wiring Harness Damage on Aircraft Caused by High-Level RF Radiated Energy

The transfer of environmental energy is not only affected by hardware deployed on the flight/weather deck, but by personnel as well. The human body also displays the characteristics of an antenna, and personnel location on the deck can influence the efficiency of the transfer path of RF energy to the susceptible portions of the GSE or Avionics.

3.2.2 Safety Procedures and Design Criteria

It is extremely important that proper procedures be followed in both the design and use of GSE to counteract the potential safety hazards indicated above. These procedures are as follows:

- o Prior to making any connections to an aircraft or weapon, it is important that recommended avionics maintenance procedures be followed that relate to grounding the unit to the carrier flight or hangar deck.

- When a test or umbilical cable is to be connected to a GSE, to a weapon launcher, or to other components of a weapons system, the GSE should if possible, first be electrically bonded directly to the launcher or other component by contacting surfaces. A short ground strap may be employed for this purpose. Weapon-launcher systems should be designed in such a way as to provide convenient connection of the test or umbilical cables between GSE, aircraft, weapon, and launcher, after good electrical grounding of the units involved is achieved.

- When a GSE and weapon must be mated before the weapon is racked in the launcher, a potential of RF arcing also exists. Mating of the GSE and weapon skin should be accomplished prior to connection of any cabling, using a separate ground strap if necessary.

- It may be advantageous to know the potential difference between a test equipment case and the skin of an aircraft or the frame of a launcher prior to making any electrical connections between the two units. This can be accomplished using a high impedance input vacuum-tube voltmeter or solid-state Volt-Ohm Meter (VOM) (at least 10 Megohms input impedance). If the potential difference is due to static charge accumulation, the voltmeter will probably dissipate the charge in the process.

- A proper filter or RF suppression device should be considered for all lines and hard contacts that must mate with aircraft or weather-deck connectors. This is particularly important when the external wiring can act as a receiving antenna and efficiently couple radiated energy. Also, the filters must suppress the effects of arc generation (including component burnout), and the generation of any signal that could act as a triggering or ordnance firing signal.

- Uniform twisting of test circuit leads reduces the effective area of the pickup loop created by the leads. In addition, the twisted wires cause cancellation of any current that may be induced in the leads. To be effective, this balancing condition must be maintained throughout the system wiring.

- When required, static discharge resistors should be placed between each line and system ground in the test circuit. These resistors should be matched and they should not be smaller than 100,000 ohms. In addition to providing a path for static potentials to be dissipated, the resistors will offer a high impedance between each line and ground, and help maintain balance to RF energy.

- Weapons test and firing circuits are almost always balanced to ground, and isolated from case ground, so that no direct path for RF energy exists during the handling or loading of weapons.[1] When GSE is developed to test this circuitry, the original firing circuit design philosophy must be maintained. It is important that the GSE contractor become aware of the specific precautions that controlled design, so that they are not violated when the test equipment is connected.

1. For more information on EMC design of weapons systems, refer to NAVSEA OD 30393 (HERO Design Guide).

- Test and umbilical cables and connectors between GSE and other equipment must be completely shielded, with 360 degree termination at both ends. The connector must be designed with the female portion on the aircraft or launcher and the male portion on the GSE. All connector pins must be recessed, and designed in such a way that the shield mates before the connector pins mate, and the pins unmate before the shield separates.

- Multiple and long test cables create a potential hazard, because of an increased possibility of levels of ground loop currents that can cause weapon system malfunctions. The number of cables should be kept to a minimum, and the cables should always be as short as possible. Flexible cable shielding must be given particular attention; see Chapter 4 on Shielding for further guidance.

- Access doors on the GSE that might have to be opened in the high-level environment (such as battery compartments, fuse panels, etc.) should be kept to a minimum. When it is necessary to have an access door, any cables that are exposed when the door is open must be shielded. No compartments or circuits that may be susceptible to the environment should be accessible through the access door.

- When external sections of the system are non-metallic, all circuitry behind these sections must be properly shielded and filtered. Non-metallic housings, such as fiberglass and plastic, do not afford any shielding protection, and these materials must be coated, with conductive finishes to obtain adequate shielding.

- As indicated elsewhere in the design guide, cable interfaces are particularly critical from the shielding effectiveness viewpoint. Special attention should be given to the uniformity of pressure applied to the shield around the connector periphery.

- Avoid using metal configurations that can act as shorted turns and thus generate high circulating currents. For example, avoid wrapping wires with labeling tags around cables, and use plastic labeling methods instead.

The above guidance is obviously somewhat general, and the GSE designer must tailor his specific requirements and approaches to meet the particular situation he encounters. His major objectives in the safety area must be (a) to assure there is no possibility of an operator shock hazard, (b) to be sure that the test equipment is not damaged by the operational environment, and (c) to assure that neither the GSE nor the equipment being tested is damaged or inadvertently triggered as a result of making or breaking connections.

As one example of options the designer has available, consider the question of grounding a piece of GSE to the flight/weather deck before initiating tests. The discussion thus far has emphasized performing this grounding operation so that a common ground is established for all equipments and the operator.

An alternative philosophy has occasionally been used. During the development of the AN/AWM-38 Armament Wiring Test Set, however, it was decided that it would be advantageous to isolate the test set from the flight deck, to avoid discharging the aircraft through the test equipment. This was accomplished by using non-conducting operator controls and rubber sleeves over switches and connectors, and by mounting the case on non-conducting legs. To isolate the test device from ground, the exterior surface of the equipment was sprayed with a special high-dielectric strength paint.

3.3 FACTORS INFLUENCING GSE EMC

The achievement of EMC in avionics related GSE involves a number of practical factors in both the planning and technical areas. These factors are summarized next.

First, the wide range of GSE functional objectives and test equipment complexities is obvious. Devices that are classified as GSE range from those that perform simple analog and digital checks, to others that include very complex signal processing and simulation tasks. The result of this diversification of objectives and technical approaches is that EMI reduction techniques must be tailored to the particular system being designed. In order to do this, the responsible engineer must not only be aware of the technical alternatives he has available (a major purpose of this design guide), but he must also be able to assess the extent to which operational GSE environments might affect performance of particular equipment.

Second, because of the particular environments in which GSE may operate, and because of EMI reduction techniques and devices that have been incorporated into existing GSE, some assessment of the potential areas of interference to this type of equipment can be made. The more prominent aspects of this assessment are summarized as follows:

- In general, designers of weather-deck and hangar-deck operated GSE and Avionics should concentrate on protecting the test equipment against high-power, narrow-band radiated electric fields in addition to the corrosive marine environment of salt water and salt spray. The fields of concern will be greatest in the HF and microwave ranges. Protection should center around maintaining the equipment case-integrity, and adequately shielding and filtering all external cables. Cable shielding is a particularly critical area, not only with regards to the EME aboard ships but also with regards to the marine environment. Problems arise when cables not qualified for use in a marine environment deteriorate. While conductive braid is not degraded by EM waves, it can be corroded by salt spray if not treated properly or covered. For those Avionics/GSE which use cables not qualified for use in marine environments, substitute cabling should be used. See the next section for information addressing the problem of corrosion.

- In general, GSE designers of other below-decks and shore-based equipment should concentrate on protecting the test equipment against conducted interference via the power line, as well as against broadband, radiated electric and magnetic fields. Signals below 100 MHz are predominantly of concern. Narrowband radiated electric fields may also be potential interference sources to some shore-based equipment. Protection should center around power line and signal/control cable filtering, including transient suppression. Case and cable shielding, and good grounding and bonding practices should be incorporated also.

- Many GSE need to operate within only a very restricted frequency range, and the application of filtering techniques outside of this range can solve most anticipated EMI problems. For example, equipment that performs continuity checks, High-Potential (HI-POT) tests, signal level checks and similar functions can often operate in a narrow frequency band ranging from dc to a frequency dictated by the sampling requirements of the system. Heavy filtering around the actual fault indicator device above this maximum frequency could preclude the need for extensive shielding. This is one of the reasons some GSE operate successfully on carrier flight decks without maintaining tight shielding integrity. For examples of this point, see comments on the AN/ASM-373 Airborne Torpedo Presetter Test Set and the AN/DRM-29 Transponder Set Test Set (section 3.4.1).

- The use of GSE often requires a change to the configuration of the equipment or system under test, with the result that the performance characteristics of the unit being tested may be different from those representative of normal operation. For example, access panels on aircraft often have to be open so that GSE cables can be connected to the aircraft. This may expose aircraft components and wiring, including the circuitry under test, to the external EME and allow radiated energy to couple into either the weapon electrical system, the GSE, or both. As a result, GSE designers must take into account the impact of mitigating the shielding or circuit integrity of the system being tested as part of the test equipment development program.

- Production quantities of GSE are almost always fairly small. In general, quantities above one or two hundred are rare. The major implication, as far as EMC design is concerned, is that it may be more cost-effective to overdesign the system suppression requirements, if this will lead to lower engineering costs. In other words, if there is a cost of $15 per unit to add a specific set of RF gasketing to a GSE enclosure, and 20 units will be built, it may not pay to spend design and test engineering time to evaluate whether the gasketing could be reduced in effectiveness or eliminated entirely. On the other hand, a different conclusion might result if a 1200 unit production run was involved. The specific circumstances involved must dictate whether or not EMC overdesign is the more reasonable approach.

- The only method currently available to assure that flight or hangar-deck GSE design is adequate from the operational environment standpoint is to use facilities that simulate the flight deck environment. The Naval Surface Warfare Center (NSWC) operates ground planes at NATC Patuxent River, Maryland and NSWC Dahlgren, Virginia, for simulating flight-deck EMEs. This is accomplished by illuminating the test item on the ground plane using representative shipboard radiators (See Chapter 11). Although military EMC specifications and standards such as MIL-STD-461/462 are usually imposed on GSE designs, the direct application of such a specification does not necessarily result in operational equipment compatibility.

- While not normally categorized as EMC, another factor that should not be overlooked by GSE designers is hardware and personnel safety. The different electric potentials that may be built up on the ship decks and superstructure, on aircraft, and on shipboard maintenance personnel can result in effects ranging from annoying transients when cables are connected or disconnected, to component burnout, to shocks to GSE operators. Chapter 3.2 of this guide provides design guidance in the safety area.

3.3.1 Corrosion Considerations

As indicated at the beginning of section 3.3, an Avionics/GSE engineer must be able to assess the extent to which operational environments might affect the performance of the equipment being designed. This is the case with controlling corrosion in an operational environment which includes salt water/salt spray. Corrosion can degrade the effectiveness of cable shielding, connectors and gaskets.

Specifications for corrosion control are called out on the design drawings of Avionics/GSE hardware. At present, many of the connector specifications called out on these drawings require cadmium plating of the connector base metal (typically steel). The problem here is that this is only a sacrificial coating to slow down corrosion of the base metal, it does not permanently protect the base metal. Other coatings include chromate wash and nickel flash coating.

Anti-corrosion technology regarding connectors and cables includes the use of composite connectors (such as the Deutsch DG123), molded cable assemblies or convoluted tubing, and films (such as Raychem GELTEK). Connectors will be qualified by specifications based on application.

The Deutsch DG123 connector is a conductive coated composite connector (see Section 4.4.3 for further information).

Molded cable assemblies can be manufactured to suit the application. One such manufacturer is Glenair. The molding compounds include neoprene, PVC, silicone, polyurethane, Viton and butyl. To cover the cable alone, Glenair also manufactures convoluted tubing made of annular Kynar or helical Teflon (see Figure 3-3 for an illustration of a molded cable assembly).

Raychem GELTEK is a silicon gel which can be used to impregnate various configurations of metal meshes and knits. According to reference [3]:

"The gel is very compliant, displaces moisture, and will deep out electrolytes that cause dissimilar metal corrosion in conventional EMI gaskets. Under compression, the gaskets metal structure makes electrical contact with the substrates by piercing through the gel. Because of its self-healing nature, the gel conforms around these contact points, sealing them from corrosive elements."

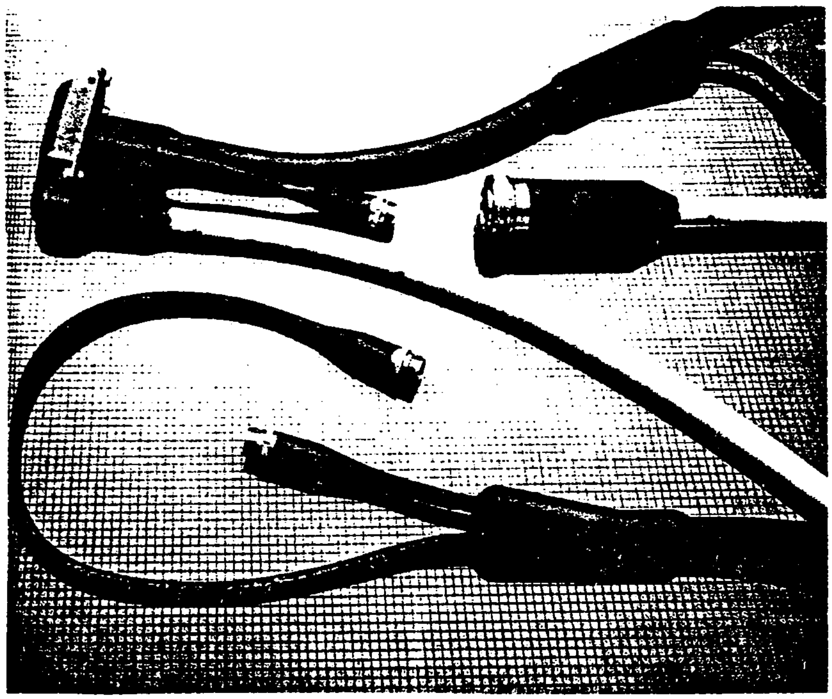

Figure 3-3 Glenair Molded Cable Assembly

Publications pertinent to corrosion control include:

NAVMAT P 4855-2, Design Guidelines for Prevention and Control of Avionic Corrosion.
MIL-STD-1344A, Test Methods for Electrical Connectors, September 1977.

Presently, one problem area of corrosion is with EMI gaskets on aircraft airframes (specifically the F/A-18). Although an airframe is a larger black box enclosure for electronics, the avionics design principles are the same.

The requirements for an EMI gasket on an airframe are essentially [3]:

(a) provide a highly conductive path for RF currents,
(b) possess a high resiliency to conform to any given joint,
(c) retain conductivity and resiliency over life of joint,
(d) provide corrosion compatibility with the joint surfaces in the presence of moisture and salt fog.

EMI gaskets for airframes are typically composites to ensure both a weather seal and an EMI seal. The problem is the corrosion of the conductive medium in the composites.

Some of the fillers used for composites include [3] graphite materials: carbon black, graphite fibers, metal powders: copper, zinc, nickel, aluminum, brass, and iron, aluminum flakes, stainless steel fibers, and metal coated inorganic fillers: silver coated glass spheres, aluminum coated glass fibers, and nickel coated mica. These fillers should be as close to the metal surfaces they join in the electrochemical series (see section 5.3).

One possible solution [3] is to protect the conductive particles of the gasket with an organic coating. Studies are presently under way by various government agencies and manufacturers to resolve these corrosion problems (see reference [3] for details of some of these studies).

Designers of Avionics/GSE should be aware of these ongoing studies and watch for the EMI gasketing products/techniques which emerge.

3.3.2 EMC Maintenance Considerations

Since the EMC requirements associated with a piece of Avionics/GSE are expected to be met over the equipment lifetime, some comments are appropriate on achieving these requirements.

As mentioned in section 3.1.1 high reliability suppression components should be selected for EMC applications. The main reason for this is that the time between failure of an EMC component and its replacement may be very long compared with the equivalent time for a component used for other purposes. An example of this is the fact that there are no scheduled periodic replacements of EMI suppression components on Navy aircraft, at present.

Many EMI suppression components are perishable, that is, they lose their effectiveness with use. For example, metallic gaskets may lose resilience after being compressed many times. The mesh in shielded rubber boots may break after continued use. Spring fingers may corrode or become blunt over a period of time. Even toothed lockwashers that have to bite into metal may become ineffective after continued applications.

Avionics/GSE design engineers must recognize these limitations and, where necessary and possible, provide feedback to those individuals on the program that are responsible for establishing equipment maintenance policies. Such inputs might range from controls on torque applied to gasketed seals, to keeping a log of gasketed compartment openings and closings, to periodic replacement of EMI suppression components in accordance with manufacturer recommendations.

3.3.3 Cost Benefit Considerations

Optimization of EMC design may not be practical or desirable when considering the benefits to be gained against the engineering costs involved. As pointed out earlier, the typical low GSE production volumes involved, and the frequent model or modification changes, would indicate that, in general, the greater benefit to be derived is from over-design of GSE from the EMC viewpoint rather than attempting to reduce EMI protective measures to a minimum consistent with estimates of expected EM environmental levels.

In many cases, present day operational EM environmental energy levels cannot be estimated accurately, and estimates of future conditions are subject to even greater uncertainty. For these reasons, in addition to those cited above, it is more often a better choice to incorporate added filtering than to consume additional engineering time and cost in attempting to reduce EMC hardware requirements to an optimum level.

This is not to suggest to the GSE designer that he apply every conceivable EMI fix he can think of in the development of a piece of equipment. To the contrary, with a good understanding of how his system must work, the way it must interface with the equipment under test, and the type of EME in which it must operate, the designer will be in a position of deciding whether or not various types of interference suppression should even be considered and which reduction approaches are appropriate to employ.

3.3.4 Microprocessors and Digital Systems

Microprocessors and digital systems include all equipments that incorporate linear or digital Integrated Circuits (ICs). EMC problems that might involve these equipments include performance degradation due to external circuit operation. EMC risk is predominantly in the performance degradation area, with broadband emissions a possible but unlikely threat.

Two basic types of performance degradation are possible. One type is called "upset" and includes such effects as reduced noise margins, logic state changes, and changes in output drive levels of analog and linear circuits. The second type of performance degradation is called "failure" and includes such effects as junction shorting, burnout, and metal failure.

Upset and failure occur at absorbed power levels of 10 μW - 10 mW and 10 mW - 10 W respectively. External power levels are highest at their source, and decrease exponentially, or $1/r^2$, with distance from the source. Metallic shielding can reduce radiated field levels by factors of 60 dB or more. Pin filters and transient suppressors can reduce conducted signal levels by an equal amount.

The development of IC technology has led to increased use of microprocessors as integral parts of avionics systems. Microprocessors are sometimes defined as one-chip processors although most are small multi-chip processors. Microprocessors usually include several hardware modules, including a processor module, Central Processing Unit (CPU) or Micro-processing Unit (MPU), one or more memory modules (RAM or ROM), and various Input/Output (I/O) modules. Overall, microprocessors can be characterized by their power requirements, noise immunity, impedance, and signaling or logic levels.

Microprocessor and digital circuit technology is extensive and rapidly changing. Currently, there are eight basic logic types: Emitter Coupled Logic (ECL), Diode Transistor Logic (DTL), Transistor-Transistor Logic (TTL), Metal Oxide Semiconductor (MOS), Current Mode Logic (CML), Integrated Injection Logic (I^2L), Resistor Transistor Logic (RTL), and High Threshold Logic (HTL) plus two basic linear circuit types (bipolar and bifet) that constitute various medium scale and large scale integrations (MSI and LSI). Integrated circuits have traditionally employed silicon technology, but now also employ gallium arsenide technology to gain increased switching speeds.

The densities of microcircuits has increased, increasing the coupling between individual circuits within a module and the risk of undesired broadband emissions on I/O signal leads. Increased density also increases the risk of performance degradation by coupling external signals picked up by I/O leads to internal circuits. Since 5 V is the standard logic supply voltage, internally generated noise will be relatively low so that interference risk is low. However, the susceptibility risk is high, especially in high ambient environments such as those created by shipboard radar, and in avionics maintenance areas. For this reason, shielding, pin filtering and good cable-to-connector bonding are important in digital systems design.

IC characteristics are given in Table 3-1. Included are switching speed, supply voltage and noise margin, as well as interference and susceptibility risks. In general, the noise generated by the circuits increases with switching speed. ECL circuits are an exception because they have a balanced configuration in which supply current to the gate is the same in the ON and OFF states, so that there is no change in current when the gate switches. Although slower, TTL circuitry is about 10 times as noisy as ECL circuitry.

Table 3-1

INTEGRATED CIRCUIT CHARACTERISTICS

CIRCUIT TYPE	SWITCHING SPEED	SUPPLY VOLTAGE	NOISE MARGIN	INTERF. RISK	SUSCEP. RISK
ECL	Very Fast	5	0.3	Low	High
DTL	Slow	5	1.0	Moderate	Moderate
TTL	Fast	5	1.3	High	Moderate
CMOS	Slow	16	7.0	High	Low

Performance degradation and hence susceptibility risk is inversely proportional to noise margin. This is the input noise voltage, riding on the worst-case logic level, that will cause the gate to trigger and cause latching or data loss. Susceptibility risk increases with switching speed because faster circuits can respond to narrower noise pulses. For this reason, and also because its noise margin is low, ECL circuitry has high susceptibility risk.

Digital computers currently in use in Naval avionics systems include the AN/AYK-14. This computer includes a CPU, memory, and various I/O modules, and is utilized in the F-18 aircraft and LAMPS helicopter.

There are several different types of upsets that can occur in microprocessors and digital systems. These include: Incorrect External Interrupt, Max Bus Interrupt, I/O Control Lose, ALU Malfunction, Incorrect Data Transfer, Memory Alteration, and Memory Loss.

In addition, external signals can cause circuit failures from which there is no recovery (burnout).

Two types of electrical transients can cause performance degradation in microcircuits. One is electrostatic discharge transients due to handling during maintenance or operation. The other is system transients generated within the digital system or its operating environment or by its external EME. Static discharge transients are usually high level, short duration transients with high source impedances (500 V, 150 μs, 1500 Ω typically). System transients are usually lower level, longer duration transients with lower source impedances (10 V, 10 μs, 100 Ω typically). Studies of microcircuit overstress tolerance have shown that microcircuits can be separated into sensitive vs. nonsensitive categories on the basis of being able to withstand 1000 V static discharge transients, and 50 V system transients, with durations and impedances as given above. In a test of 40 microcircuit types, representing all common technologies, very few circuits failed the discharge transient test at voltages up to 1000 V; however, several circuits failed the system transient test at 50 V. The circuits that failed the system transient test included high-speed TTL, first generation I^2L, certain NMOS devices, and certain analog devices. The screening procedure has been proposed for incorporation into MIL-STD-883, Test Methods and Procedures for Microelectronics.

Electromagnetic Vulnerability (EMV) design requirements for avionics systems must reflect the avionics environment, digital system susceptibilities, and potential coupling mechanisms. The environment will include high-level radar and communication signals from co-located or nearby transmitters. Coupling will be direct to I/O lines, or indirect via intercable coupling. Platforms will provide shielding in some cases, but cannot be depended upon to do so. This is especially true in a maintenance environment where equipments are opened and then operated. IC susceptibilities are a function of circuit type, but generally are not low enough to withstand external interference without protection. ICs and their interconnecting leads should be shielded, and shields should be circumferentially bonded to chassis. In some cases pin filters should be installed at the entry point to the chassis. High-level radar signals are particularly troublesome and should be protected against as a matter of course.

A possible way to reduce microprocessor and digital system susceptibility to externally generated interference is to include digital information processing techniques such as forward error correcting codes in the data bus and I/O design. Error-correcting techniques are used effectively in data communications, and would provide similar benefits if used for internal data transfers within digital systems. This is an area that is yet to be explored for protection of avionic systems in an EMV test environment.

EMV test environment requirement levels are defined in MIL-HDBK-235, and are applied to an avionics system at the discretion of the procuring activity. Contractors do not have a great deal of experience in meeting EMV requirements and thus must use an iterative process to generate a design. In general, the EMV design problem is solved by systematically isolating sensitive circuits from the high level EMEs. Because of weight

penalties, shielding requirements need to be determined precisely which is difficult to do because of the wide variation in component susceptibility levels and coupling factors (60dB and 40dB respectively). Since analysis techniques are not adequate, development tests must be performed.

In general, digital circuits are less susceptible to EMV environments than analog circuits. Digital circuits are more sensitive than analog circuits to pulsed signals, while analog circuits are more sensitive than digital circuits to steady Continuous Wave (CW) signals. Since pulsed signals are typically 30dB higher than CW signals, there are many choices to be made in selecting components. Environment component susceptibility, and coupling information are all very important to successful EMV design.

EMV susceptibility tests on microprocessors and digital systems are based on viewing these devices as components. The modified MIL-STD-461C susceptibility tests are usually specified.

As stated in MIL-STD-461C, the RS03 requirement is that "the test sample shall not exhibit any malfunction, degradation of performance, or deviation from specified indications, beyond the tolerances indicated in the individual equipment or subsystem specification, when subjected to a radiated electric field of 1 Volt/meter. Above 30 MHz, the requirement shall be met for both horizontally and vertically polarized waves; circular polarized waves are also acceptable."

In addition to the requirements of MIL-STD-461C, one additional RF susceptibility test may be conducted. No malfunction, undesirable response, or change in indication beyond the tolerance given in the equipment specification or approved test plan shall be produced in any equipment when 1000 mV rms RF signals in the 2-30 MHz band are coupled into the interconnecting wiring.

The RF signal is a 400 Hz carrier modulated 80 percent with the frequency and/or type of modulation to which the test sample is most susceptible. The RF signals are applied to all interconnecting wires or cables between component parts of subsystems, and to wires and cables connected to other equipment.

The RF signals are measured in a one-turn loop with an RF voltmeter as shown in Figure 3-4. The signal generator has an output impedance of 50 ohms or less and a maximum open-circuit voltage of at least 3 V rms. The test procedure is as follows:

a. With the current probe clamped around the wire bundles under test, set the signal generator for maximum output and slowly scan from 2 MHz to 30 MHz.

b. At those frequencies where a response is noted, move the current probe along the wire for maximum indication. Reduce the signal generator output and determine the threshold of susceptibility of the wire bundle, probe each wire in the bundle separately to determine the susceptible wire(s).

c. Record and include in the test report the threshold of susceptibility data (frequency, amplitude, and description of response) obtained in (b) above. Also include in the test report the corrective action required for each wire that does not meet the required level.

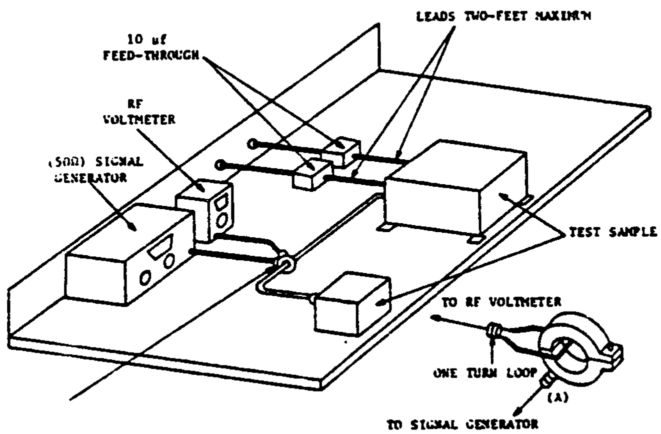

Figure 3-4 RF Signal Measurement

Wire bundles may be tested in sections if the current probe will not fit around the complete wire bundles. Chassis or circuit grounds not normally routed with the interconnecting wiring are measured individually and are not included in the wire bundle under test. Power ground return leads are not placed in the probe.

3.4 EXAMPLES OF AVIONIC AND GSE DESIGNS[2]

GSE, in order to perform their diverse checkout-associated missions, will have to operate in a variety of different physical environments. The GSE support functions may place individual equipments on aircraft or helicopter flight decks, on hangar decks, in avionics shops, or in a number of other

2. Much of this material is based on the Naval Avionics Facility, Indianapolis 1971 Annual Engineering Department Report.

shipboard spaces. Specific units may also be required to perform operations on base flightlines, at base shops, at test and rework facilities, or at other ground-based locations.

Chapter 2 discussed the EMEs which might exist in some of these locations, and intended to provide some understanding of several of the factors associated with the creation of these environments.

This chapter has presented the basic design guidelines applicable to avionics and GSE. Next are provided specific examples of such designs.

3.4.1 Flight or Hangar Deck Operation

GSE designed for flight or hangar deck operation, can be grossly categorized in relation to the system or systems to which they provide support. These systems include:

- Airborne guided weapons
- Weapons delivery systems
- Aircraft

Guided missile test sets provide checks of missile operating status or capabilities by direct measurements of selected missile parameters, or by simulations of a target and noting the missile response to that target. One such device is the AN/DSM-92A Guided Weapon Test Set shown in Figure 3-5.

This test set operates and monitors the WALLEYE weapon to provide a GO/NO-GO verification check of the weapon guidance section. The test set supplies all ac and dc power required to operate the weapon with the exception of the ram air turbine. A video indicator monitors the weapon video outputs; a GO/NO-GO type quality meter provides visual checks of the circuits under test; and various indicator lamps and switches duplicate the functions of the aircraft weapon controls. The test set has built-in self-test capabilities for checking the voltage and phase rotation of the input power and for checking the test set power supplies. An adapter cable is provided to check the video indicator using Aircraft Weapon Control Test Set AN/ASM-184. A power adapter cable is also provided to obtain three-phase power from the aircraft.

After initial versions of this test set were produced, it was found necessary to reduce the susceptibility of the test set to the radiated environment. This was accomplished by replacing the original test cables with shielded cables. In addition, it was found necessary to develop a modification kit that would reduce EM emissions from the AN/DSM-92 to meet MIL-STD-461.

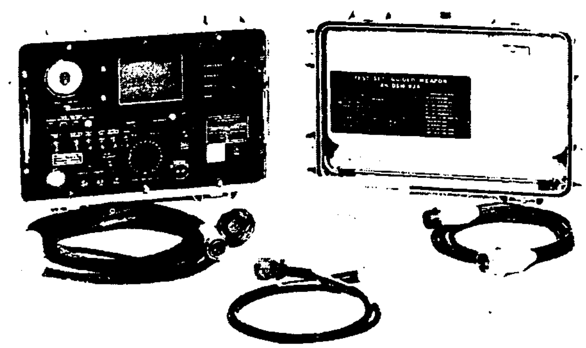

Figure 3-5 Guided Weapon Test Set AN/DSM-92A

Another example of a missile test set that is operated on flight and hangar decks is the AN/ASM-373 Airborne Torpedo Presetter Test Set. The equipment is designed to check the C-7821/A and C-8990/A torpedo depth controls installed in SH-3H and S-2F aircraft. The test set also checks the gyro cage and shackle release signals and the delay between them. The basis for test set operation is a stepper switch that sequences dc test voltages in 20 millisecond step intervals.

The major EMI suppression required on the AN/ASM-373 to assure its satisfactory performance in the operational environment included line filters on both the input and output connectors, and diodes across the stepper switch coil to reduce transients. Significantly, it was not necessary to maintain shielding integrity of the test set enclosure, since test equipment degradation occurred only due to environmental signals that were coupled onto its input and output cables. The front panel gasket is made of nonconductive rubber, front panel indicator lamps have plastic covers, and the front panel meter is not shielded.

Weapon delivery test sets provide checks of the status and capabilities of weapon launchers and ancillary devices. Several examples of such testers follow.

The Armament Wiring Test Set AN/AWM-38(V) is a portable field test set for performing HI-POT and circuit continuity tests on the Aircraft Armament Monitor and Control (AMAC) Systems. It is used in checking out the AMAC system prior to loading nuclear weapons aboard Navy aircraft. The check assures the aircraft readiness as a preflight test of the AMAC system. The test set is shown in Figure 3-6.

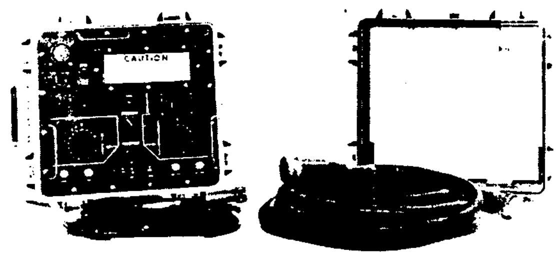

Figure 3-6 Armament Wiring Test Set AN/AWM-38(V)

In order for this unit to operate in a flight-deck environment without burning out transistors, it was necessary to use very tightly-braided shielded cabling, and to incorporate a "Magnaforming" technique for assuring shielding integrity between the cable and its connectors. Magnaforming will be discussed in the Shielding chapter of this design guide. In addition, filters were required to prevent damage to the test set due to transients introduced onto the cable when connecting to the AMAC systems.

The Aircraft Firing Circuit Tests Sets (AFCTS) AN/AWM-53 and AN/AWM-54 are portable, battery operated flight line test sets used to perform stray voltage and firing circuit tests. The AFCTS sets provide a visual indication of unsafe low-level voltages and of the presence or absence of firing current. The indications are furnished by a red NO-GO and a green GO indicator visible under all lighting conditions including red light. The AWM-53 is a hand held unit, while the AWM-54 is waist-mounted (see Figure 3-7).

During the stray voltage test, the AN/AWM-53 test set monitors the aircraft's firing circuit to determine that the ambient voltage of the firing circuit is not sufficient to detonate electro-explosive devices. The test set monitors voltages with frequencies from dc to 1 MHz.

During the firing circuit test, the test set monitors the aircraft's firing circuit to determine firing capability. The aircraft must be capable of producing a minimum firing current for a specified period of time before the test set will produce a GO indication.

During self test, the test set checks the threshold response of the stray voltage circuit; the function of the lamp and logic circuits; the charge status of the battery pack; and the continuity of the fuse, input circuitry, and the aircraft adapter cable.

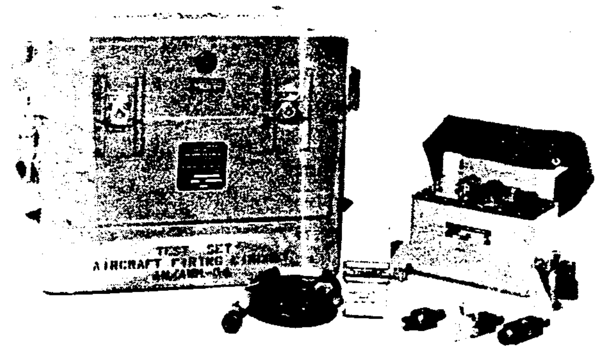

Figure 3-7 Aircraft Firing Circuit Test Set AN/AWM-54

The AN/AWM-54 differs from the AN/AWM-53 because it is designed to the requirements of MIL-T-21200, utilizes a stray energy sense circuit in lieu of a stray voltage circuit, and employs a resistance sensitive circuit for firing circuit testing in lieu of a current sensitive circuit.

Extensive EMI reduction techniques have been applied to these units. These include separately shielded battery packs, filtering of battery output leads, filters in signal leads, shielding of cables and good bonding of cable shields to connector shells, conductive gaskets around front panels and other covers, use of conductive finishes, incorporation of waveguide below cutoff recesses for lights, and many others.

Aircraft test sets provide the means for evaluating a wide variety of aircraft-associated components, equipments, and subsystems prior to takeoff. Equipment checkouts might range from responses from IFF transponders to turbine temperature testing, or from radar power output measurements to engine trim box operation. Two examples of these types of units follow.

The Transponder Set Test Set AN/DRM-29 (See Figure 3-8) is a portable flight-line test set to be used to test the Transponder Set AN/DRQ-4(V) and its associated wiring and antennas when installed in a target aircraft or drone. The test set is a portable hand-held battery-powered unit with a built-in battery charger.

A GO-NO-GO indication is given for the transponder set condition, and for all self-checks for the test set.

The test set transmits to the transponder set an FM signal in the 1.76 to 1.85 GHz band. A FM signal is then received from the transponder set in the 2.20 to 2.29 GHz band. The received signal is analyzed and a meter gives a GO/NO-GO decision on the transponder set performance.

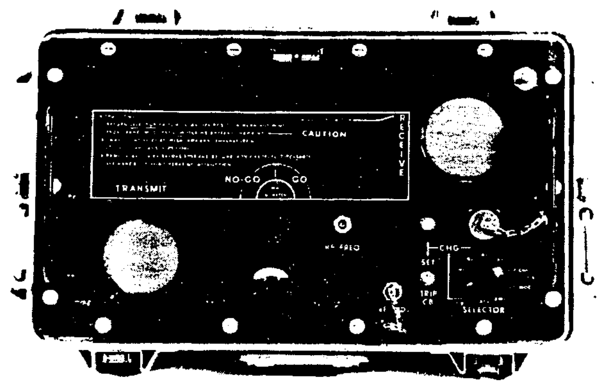

Figure 3-8 Transponder Set Test Set AN/DRM-29

A microstrip board containing spiral transmitting and receiving antennas, a 1760 MHz bandpass filter, delay lines, coupler, and 180 degree hybrid discriminator is used. Another board contains amplifier, oscillator, voltage regulator, and battery charger circuits.

The test set is completely self-contained, and this feature eliminates the need to consider power and signal cable filtering. It was found that the set was relatively insensitive to the flight deck radiated environment, thus no RF gasketing was employed, and case assembly using screw fasteners was considered adequate. However, portions of internal shields (particularly those on the front surface of the unit) have been solder-seamed, and microwave absorbent material has been used to cut down on internal reflections.

The TF34-2 Engine Trim Test Set is a solid-state device for trimming installed engines in the S3A aircraft. The unit provides monitoring of temperatures, speed and inlet guide vane angles during engine flight checkout. Particular attention has been given in design of the test set to interference from sources of EMR, and an environmentally sound case and shielded cables are employed.

RF gaskets are used extensively to provide a good electrical seal between the front panel of the test set and the box. Push button and toggle switches incorporate neoprene-filled conductive gaskets, and rotary switches including EMI shaft seals. Indicating lamps are EMI-protected and backed with neoprene-filled conductive gaskets, and LED displays are behind a wire screen. A thumbwheel switch for presetting parameters is located behind a door whose edges are sealed with a neoprene-filled conductive gasket. Cable shields are required to have 95% coverage.

The Multi-application Engine Vibration Test Set is used to amplify and display signals produced by self-generating Piezoelectric accelerometer or velocity type transducers attached to gas turbine engines. It incorporates considerable EMI suppression, including weather seal/conductive gaskets, conductive boots on toggle switches, and bulkhead-mounted power and signal line filters. Conductive mylar is used to shield the LED tachometer display window, while other indicating lights and meter dials are protected with wire mesh screens. All test set cabling is shielded.

3.4.2 Shop Testing

GSE designed for shop use, or for use at test and rework facilities, typically involve more comprehensive equipment or system checkout capabilities than is normally possible on hangar or weather decks because of space and time limitations. These units are normally ac powered, frequently non-portable, and sometimes contain or are associated with mechanical test fixtures. Compliance with MIL-STD-461C will generally insure EMC for this type of equipment.

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[2] R. Raut, "On The Computation Of Electromagnetic Field Components From A Practical Printed Circuit Board," IEEE International Symposium On Electromagnetic Compatibility, September 16-18, 1986, San Diego, CA, pp. 161-166.

[3] Naval Air Systems Command (AIR-5161), EMI Gasketing and Corrosion Colloquium, January 15, 1988, pp. 2-8 (and Attachment 6).

[4] D. Fernald, "Comparison of Shielding Effectiveness of Various Backshell Configurations," ITEM, 1984.

4.0 SHIELDING

4.1 GENERAL SHIELD DESIGN CONSIDERATIONS

Shielding has two main purposes: (a) to keep radiated EM energy confined within a specific region, and (b) to prevent radiated EM energy from entering a specific region. Thus, shielding is essentially a decoupling mechanism used to reduce radiated interactions between equipments, or between portions of a given equipment.

Shielding requirements, as far as representative GSE/Avionics design is concerned, are of greatest importance with regard to shipboard topside units. As indicated previously, a major consideration in the development of flight and hangar deck test equipment is the maintenance of case shielding integrity, and the protection of associated cables from the expected high-level EME and physical environment (salt spray).

The shielding effectiveness of an equipment or subassembly enclosure is a complex function of a number of parameters, the most notable of these being the frequency and impedance of the impinging wave, intrinsic characteristics of the shield materials, and the numbers and shapes of shield discontinuities. The equipment design process consists of establishing undesired signal levels on one side of a proposed shielding barrier, estimating tolerable signal levels on the other side, and trading off shield design options to achieve the necessary effectiveness level.

This chapter will identify many of these shield design options. It is subdivided into discussions of solid shields (including multiple wall and thin-film shields), non-solid shields, cable shielding and the maintenance of shielding integrity through a connector, and shield discontinuities (including seams, gasket requirements, the use of waveguide attenuators, conductive glass, and other aspects).

Throughout the discussion, it should be recognized that shielding, although an important technique for reducing EMI effects, is not the only technique available for this purpose. Application of shielding techniques should not be made without due regard to the roles that filtering, grounding, and bonding will play in the suppression program.

4.2 SOLID SHIELDING MATERIALS

4.2.1 Shielding Analysis

Interference signal attenuation by a solid shield is due to two distinct effects: (a) reflection (due to impedance mismatches) of the interference wave at the air-metal boundary as the wave strikes the metal surface, and reflection at the metal-air boundary as the interference wave emerges from the metal shield and (b) absorption of the interference wave in passing through the metal shield between the two boundaries. The first loss is generally called Reflection Loss and the second is called Absorption or Penetration Loss. The combined loss due to these two effects is considered the shielding effectiveness of the shield.

It is convenient to separate the initial reflections from both surfaces of the shield from subsequent reflections that may take place at these surfaces. These effects are called the Single Reflection Loss and Multiple Reflection Correction Term, respectively. Under circumstances when the absorption loss is greater than about 15 dB, the multiple reflection correction term can be ignored.

Using transmission line theory, the Shielding Effectiveness, S, of solid shielding materials is defined [1],[2],[3] as:

$$S \text{ (in dB)} = A + R + B \tag{4-1}$$

where

A = Absorption (Penetration) Loss by the material, in dB

R = Single Reflection Loss from both surfaces of the sheet, in dB

B = Multiple Reflection correction term, in dB

Magnetic shielding depends primarily on absorption losses, since reflection losses for magnetic fields are small for most materials. Electric fields are readily stopped by metal shields because large reflection losses are easily obtained. Absorption loss, is the same for electric and magnetic fields.

The absorption loss term, A, is independent of wave impedance, and can be described by the relationship:

$$A \text{ (in dB)} = 3.34\, d \sqrt{\mu f g} \tag{4-2}$$

where

d = shield thickness, in inches

g = material conductivity, relative to copper

μ = material permeability, relative to copper

f = frequency, in Hertz

For a given material, absorption loss, in dB, at a specific frequency is a linear function of material thickness. The characteristics of the material that influence this loss are its conductivity and permeability. Table 4-1 list values of conductivity and permeability relative to copper for a variety of useful shielding materials, and indicates calculated values of absorption loss at 150 kHz.

The single reflection loss term, R, is dependent upon wave-impedance, and can be described by the relationships:

TABLE 4-1 CHARACTERISTICS OF VARIOUS METALS USED FOR SHIELDS

Metal	Relative Conductivity g	Relative Permeability at 150 kHz u	Absorption Loss db/mil at 150 kHz
Silver	1.05	1	1.32
Copper-Annealed	1.00	1	1.29
Copper-Hard Drawn	.97	1	1.26
Gold	.70	1	1.08
Aluminum	.61	1	1.01
Magnesium	.38	1	.79
Zinc	.29	1	.70
Brass	.26	1	.66
Cadmium	.23	1	.62
Nickel	.20	1	.58
Phosphor-Bronze	.18	1	.55
Iron	.17	1,000	16.9
Tin	.15	1	.50
Steel, SAE 1045	.10	1,000	12.9
Beryllium	.10	1	.41
Lead	.08	1	.36
Hypernick	.06	80,000	88.5*
Monel	.04	1	.26
Mu-Metal	.03	80,000	63.2*
Permalloy	.03	80,000	63.2*
Steel, Stainless	.02	1,000	5.7*

* assuming material not saturated

$$R(\text{low impedance source}) = 20 \log \left[\frac{0.462}{r\sqrt{fg/\mu}} + 0.136\sqrt{fg/\mu} + 0.354 \right] \quad (4\text{-}3)$$

$$R(\text{high impedance source}) = 354 - 20 \log \left[r\sqrt{\mu f^3/g} \right] \quad (4\text{-}4)$$

$$R(\text{plane wave source}) = 168 - 20 \log\sqrt{\mu f/g} \quad (4\text{-}5)$$

where r = source to shield distance, in inches.

The corresponding Nomographs for equations (4-3), (4-4), and (4-5) may be found in Figures (8-16/8-17), (8-5) and (8-13/8-14) respectively. These equations are plotted in Figures 4-1 through 4-3. The amplitude of a wave reflected from solid shielding material depends upon the degree of mismatch between the impedance of the impinging wave, the air medium, and the impedance of the shield. It should be noted that the impedance of the impinging wave in the vicinity of the shield will be a function of the distance of the shield from the source and whether or not it is in the near field or far field. This is discussed a little later in this section.

The impedance of the shield is a complex function of the shield electrical parameters, shield thickness, and frequency of interest; in general:

(a) Impedance of a shield is <u>low</u> for materials of high conductivity.
(b) Impedance of a shield is <u>high</u> for materials of high permeability.

At radio frequencies of 100 kHz and above, up to 10 GHz or more, the thickness of the shield is generally great enough and the shield hole fraction small enough that the shield impedance can be approximated by that of the solid metallic surface of relative conductivity g and relative permeability μ. This impedance is:

$$2\pi\ (10^{-7})\sqrt{0.4\ f\mu/g} \qquad (4\text{-}6)$$

For a non-magnetic shield, $\mu \approx 1$. For most shield materials (aluminum, copper, silver, etc.) $g \approx 1$. Thus even at 10 GHz, the surface impedance is only on the order of 0.04 ohms.

In order that the reflected wave be as large as possible, or that the reflection loss be great, the shielding should have an impedance that is either very much greater than the wave impedance or very much less to maximize the mismatch. In shielding against plane waves (very high impedance), it is more practical to establish a mismatch by using shielding material having a very low impedance than it is to use very high impedance material, and that is why conducting material is always employed.

In the case of high or low impedance sources, it can be seen from equations 4-3 and 4-4 that the reflection loss obtained will be a function of the separation distance between the source and the shield. The equations have been derived on the assumption that these sources are point sources (that is, small compared to the wavelength of concern). The reason the loss is distance dependent is because we assume that the shield is in the near field ($r < 2D^2/\lambda$) where the impedance of the wave (E/H) impinging on the shield cannot be defined but must be approximated as a function of r.

In the case of a plane wave source, it can be seen from equation (4-5) that the reflection loss obtained will <u>not</u> be a function of the separation distance between the source and the shield. The reason the loss is distance-independent is because we assume that the shield is in the far field ($r > 2D^2/\lambda$) where the impedance of the wave (E/H) impinging on the shield is defined as $E = 120\pi H$ which is not a function of r.

<u>The multiple reflection correction term</u>, B, is defined by the equation:

$$B\ (\text{in dB}) = 20\log\left|1 - W^{-A/10}\ (\cos 0.23A - j\sin 0.23A)\right| \qquad (4\text{-}7)$$

where

$$W = 4\left[\frac{(1-m^2)^2 - 2m^2 - j2\ \ 2\ \ m\ (1-m^2)}{1 + (1 +\ \ 2\ \ m)^2}\right] \qquad (4\text{-}7a)$$

$$m = 0.766\ r\ \sqrt{fg/\mu} \qquad (4\text{-}7b)$$

$$B = 20\log\ (1-e^{-2t/\delta})\ \text{for thin shields} \qquad (4\text{-}7c)$$

where

t = thickness and δ = skin depth

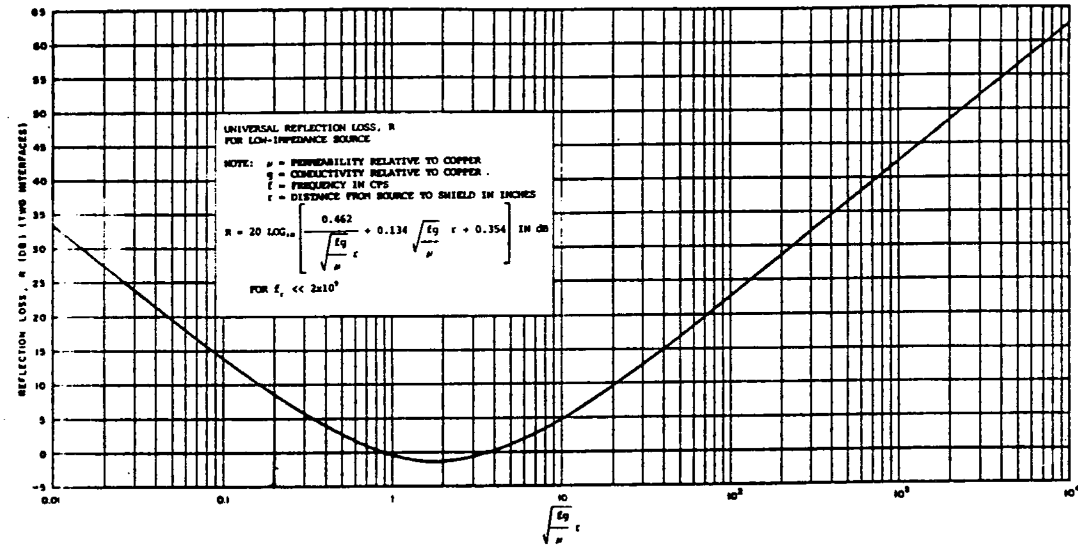


Figure 4-1 Universal Reflection Loss for Low-Impedance Source

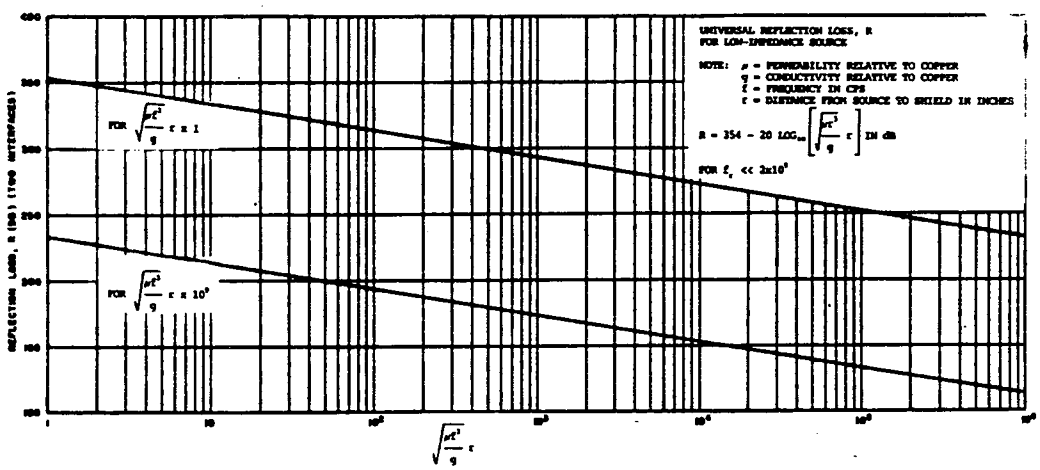


Figure 4-2 Universal Reflection Loss for High-Impedance Source

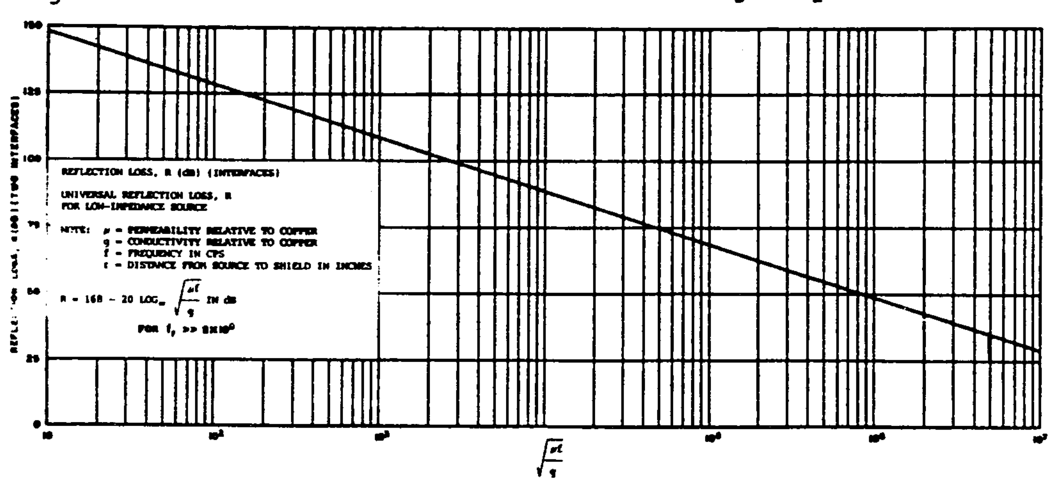


Figure 4-3 Universal Reflection Loss for Plane-Wave Impedance Source

$10^7 = \sqrt{.61f}$

As indicated previously, the multiple reflection correction term may be ignored if the absorption loss exceeds 15 dB. If the absorption loss is less than 15 dB, the correction must be determined.

The multiple reflection correction term is complicated to compute, and depends on material, dimensional, and frequency parameters. Fortunately, in almost all practical cases, W in equation (4-7a) equals unity. The only notable exception is the special case of extremely low-frequency shielding against low-impedance (chiefly magnetic) fields.

Equation (4-7) is combined with the absorption loss and is plotted in Figure 4-4 for W = 1. Note that the sign of B can be positive or negative, depending on the phase relationships between the reflections. When W is not equal to unity, its value may be obtained using Figure 4-5.

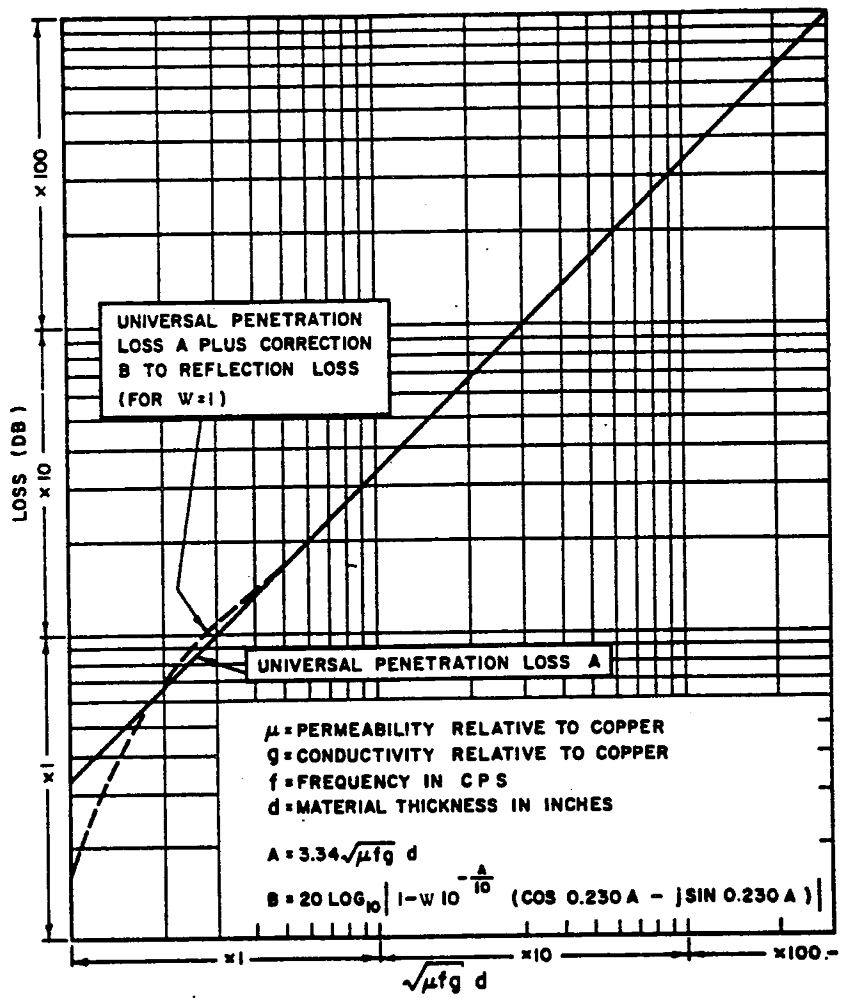


Figure 4-4 Penetration Loss and Multiple Reflection Correction Term When W = 1

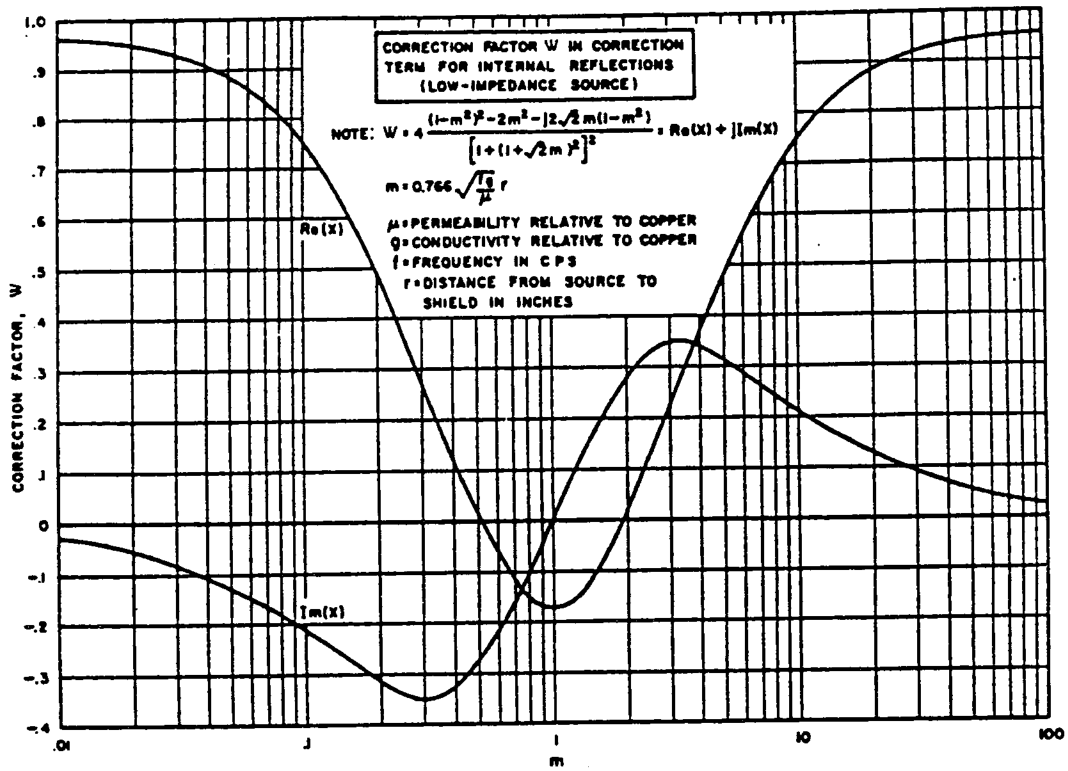


Figure 4-5 Correction Factor in Correction Term for Internal Reflections

Table 4-2 illustrates a typical set of shielding effectiveness calculations for solid sheets of copper and iron [2]. Note the relative magnitudes of the reflection and penetration components. Reflection loss in the electric field cases decreases inversely with frequency.

Nomographs for computing the shielding effectiveness of solid sheet materials may be found in Chapter 8 of this publication.

Representative solid sheet shielding effectiveness measurements are shown in Table 4-3. The data encompasses a variety of materials subjected to either low or plane-wave impedance energy at frequencies ranging from 100 Hz to 10 MHz. For the materials shown, the shielding effectiveness against high-impedance waves would exceed the plane-wave cases.

4.2.2 Additional Comments on Magnetic Shielding

At very low frequencies, it can be shown that the shielding effectiveness of solid materials to low impedance waves approaches

$$S \text{ (in dB)} = 20 \log (1 + \mu d/2r) \qquad (4\text{-}8)$$

Measurements confirm this relationship.

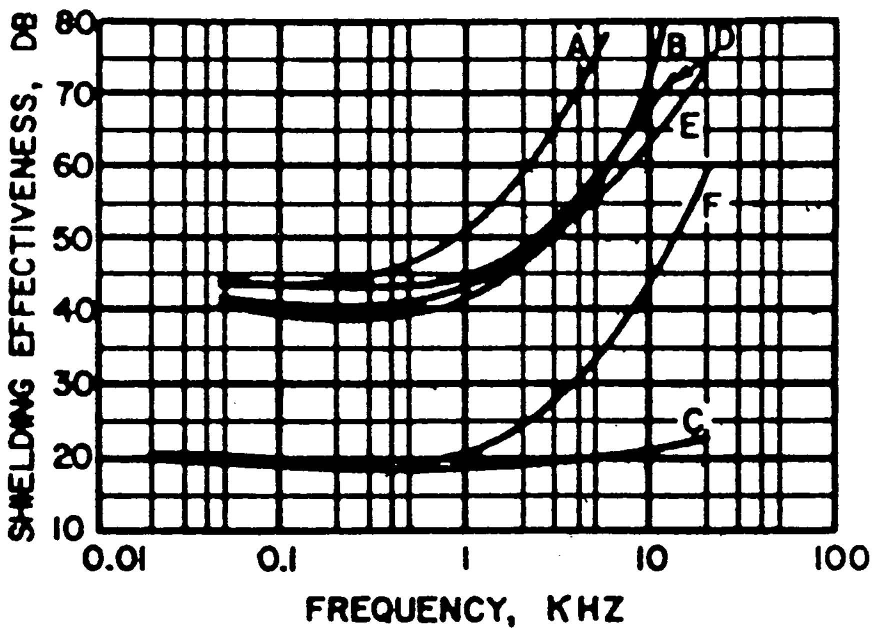
TABLE 4-2 TYPICAL CALCULATED VALUES OF SHIELDING EFFECTIVENESS OF SOLID SHEETS

Material	Frequency	Type of Field	Metal Thickness (mils)	R Reflection Loss (db)	A Penetration Loss (db)	B Correction Term (db)	Total Shielding Effectiveness S = R + A + B (db)
Copper	60 Hz	Magnetic	1	22.4	0.026	-22.2	0.23
	60 Hz	Magnetic	10	22.4	0.26	-19.2	3.46
	60 Hz	Magnetic	300	22.4	7.80	+0.32	30.52
	1 kHz	Magnetic	10	34.2	1.06	-10.37	24.89
	10 kHz	Magnetic	10	44.2	3.34	-2.62	44.92
	10 kHz	Electric	10	212.0	3.34	-2.61	212.73
	10 kHz	Plane Waves	10	128.0	3.34	-2.61	128.73
	10 kHz	Magnetic	30	44.2	10.02	+0.58	54.80
	150 kHz	Magnetic	10	56.0	12.9	+0.5	69.4
	150 kHz	Electric	10	176.8	12.9	+0.5	190.2
	150 kHz	Plane Waves	10	117.0	12.9	+0.5	130.4
	1 MHZ	Magnetic	10	64.2	33.4	0	97.6
	1 MHZ	Electric	10	152.0	33.4	0	185.4
Copper	1 MHZ	Plane Waves	10	108.2	33.4	0	141.6
Iron	60 Hz	Magnetic	1	-0.9	0.334	+0.95	0.38
	60 Hz	Magnetic	10	-0.9	3.34	+0.78	3.22
	60 Hz	Magnetic	300	-0.9	100.0	0	99.1
	1 kHz	Magnetic	10	0.9	13.70	+0.06	14.66
	10 kHz	Magnetic	10	8.0	43.5	0	51.5
	10 kHz	Electric	10	174.0	43.5	0	217.5
	10 kHz	Plane Waves	10	90.5	43.5	0	134.0
Iron	10 kHz	Magnetic	30	8.0	130.5	0	138.5

TABLE 4-3 MEASURED SHIELDING EFFECTIVENESS DATA FOR SOLID SHEET MATERIALS

Impinging Wave	Material	Thickness (mils)	Nominal Effectiveness							
			0.1 kHz	1 kHz	10 kHz	85 kHz	200 kHz	2 MHz	5 MHz	10 MHz
Plane	Cu	2.5					109	106	114	
Plane	Al	5					107	109	118	
Plane	Stainless Steel	18					97	95	99	
Plane	Steel (U_r=250)	4.5					105	99	101	
Plane	AA-Conetic Foil (U_r-10,000)	3.5					97	130		
Low Impedance	Cu	125	8	22	58					
		63				97				
		31	4	11	29	59		120		
		4.5				34		55		
Low Impedance	Al	125	5	18	50					
		63	1	16	35	78				
		31	1	10	24					
Low Impedance	Steel (u=242)	63	25	40	80					
		31	4	28	59	94		92		120
Low Impedance	Brass	31				42				
Low Impedance	Cu-Clad Steel Clad 2 Sides	31				107				
Low Impedance	Cu-Clad Steel Clad 1 Side	94				103				

Figure 4-6 illustrates representative shielding effectiveness data taken for a variety of high-permeability sheet materials. Loop sensors were located 1/8" from each sheet. The figure shows the typical leveling off in shielding effectiveness as frequency is decreased, with the breakpoint occurring in the 1 kHz range. Low frequency magnetic shielding is essentially achieved by establishing a low reluctance path in which the magnetic field is contained.



NOTE- The material designated as Mumetal is suspect of being ordinary steel.

A - HY-MU "80"
B - CO-NETIC "AA"
C - MUMETAL NO.1, 0.014" TK
D - AMPB-65, NO.1, 0.014" TK
E - AMPB-65, NO.2, 0.014" TK
F - GALVANNEALED STEEL FROM RFI ENCLOSURE, 0.028" TK

Figure 4-6 Shielding Effectiveness of High-Permeability Sheets

The variation of shielding effectiveness as a function of loop sensor separation is shown in Figure 4-7a for one of the materials plotted in the previous figure. A change in effectiveness of about 5 dB over the range of the test at a particular frequency is indicated.

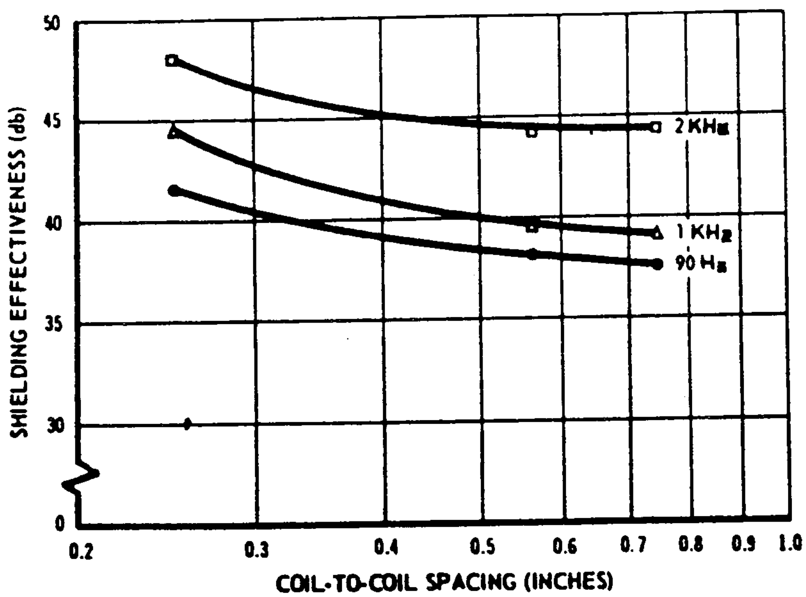


Figure 4-7a Shielding Effectiveness of AMPB-65 as a Function of Coil-to-Coil Spacing

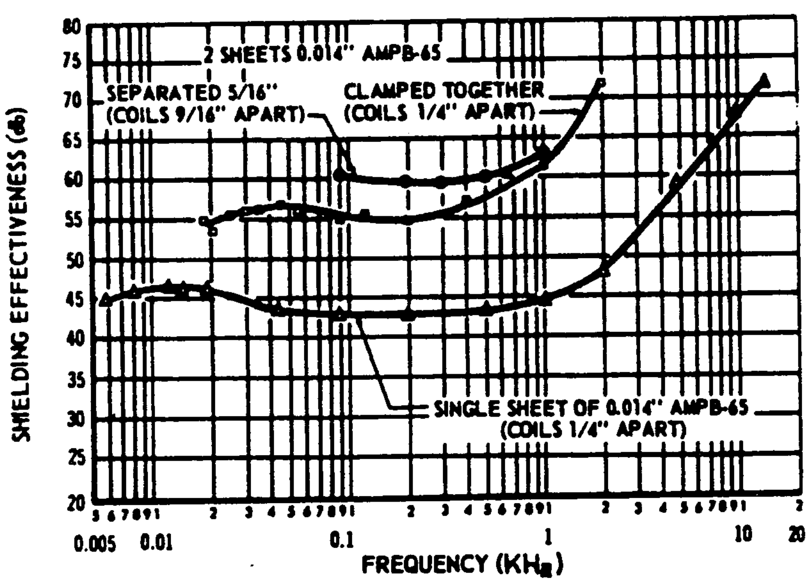


Figure 4-7b Shielding Effectiveness of Two Sheets of AMPB-65 Compared with Single Sheet

A difficulty with most magnetic shielding materials is their tendency to change permeability when formed, machined, subjected to rapid or extreme temperature changes, or dropped. These processes change the orientation of the magnetic domains in the material, and it is necessary to reorient the domains by annealing to restore the initial magnetic properties. A typical annealing process involves heating the material to about 2000° F (sometimes in an inert gas environment), holding it at that temperature for a couple of hours,and letting it slowly cool to room temperature.

4.2.3 Multiple Solid Shields

There are cases when it is appropriate to consider using two or even three layers of shielding material rather than a single sheet to obtain particular total shielding characteristics. The most frequently encountered circumstances are when good protection against both electric and magnetic fields is desired, although other situations also occur.

Although Mu-metal and similar types of high-permeability alloys provide good shielding for low-frequency weak magnetic fields, they tend to be less effective under the saturating effects of high-level fields. Where magnetic shielding in high level environments is necessary, it is often desirable to use a multiple shield, where the outer material has a lower permeability but a higher saturation level than the inner material. Such a structure might be created with materials having the following characteristics:

	INNER MATERIAL (CO-NETIC AA)	OUTER MATERIAL (NETIC S3-6)
Initial Permeability	20,000	300
Permeability at 200 Gauss	80,000	500
Saturation Inductance(Gauss)	7,500	22,000

The material thickness necessary would be dictated by the expected levels of external fields, and the desired suppression.

When much of the usefulness of shielding is due to reflection loss, 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. The separation of the two layers of metal is necessary to provide for the additional discontinuous surfaces.

Table 4-4 compares single and double shields of copper, bronze, and galvanized sheet steel at various frequencies. Table 4-4A compares single and double shielding, both nonisolated and isolated. Table 4-4B shows how the attenuation varies with frequency for various combinations. Curves for electric field shielding effectiveness often are not drawn because of the very high values encountered, often beyond measurement equipment capabilities. However, examples appear in Figure 4-13 and 4-14.

TABLE 4-4A COMPARISON OF SINGLE AND DOUBLE SHIELDS [26]: ATTENUATION AT 1 GHz

Type of Room	Electric Field and Plane Wave Field	
Single Shield Room	55-65	Copper Screen
Double Shield Room not Isolated	80	
Double Electrically Isolated Room	120	

TABLE 4-4B COMPARISON OF SINGLE AND DOUBLE SHIELDS [26]: ATTENUATION PROVIDED BY BARRIER TYPES IN DOUBLE ELECTRICALLY ISOLATED ROOMS

	Magnetic Field		Electric and Plane Wave Field		
Barrier Material	60 Hz	15 KHz	15 KHz	1 GHz	10 GHz
3 oz. (per sq. ft.) Copper, both layers	2-3	64	120	120	120
24 ga. Steel, both layers	15	80	120	120	90 to 106
Combination Copper and Steel	18-25	90	120	120	120
Copper Screen, both layers, 22 x 22 x .015	2-3	68	120	120	77
Bronze Screen, both layers, 18 x 20 x .011	0	40	120	110	57
Four shields, two 24 ga. Steel, two 24 oz. Copper	30-36	120	120	120	120

A similar advantage has been noted with magnetic sheet materials [5] (see Figure 4-7b).

4.2.4 Coatings and Thin-Film Shielding

Thin shielding has been employed in a variety of ways, ranging from metalized component packaging for protection against RF fields during shipping and storing, to conductive glass, to vacuum deposited shields for microelectronics applications[1]. Since future GSE developments might incorporate thin solid shields, some comments regarding such shields are considered appropriate.

1. Thin shielding is loosely defined as shield thickness less than one quarter wavelength at the propagation velocity dictated by the material.

Solid material shielding theory is applicable to thin-film shields. For shields much thinner than $\lambda/4$, the Absorption Loss is very small, but the Multiple Reflection Correction Term, B, is fairly large and negative, thus offsetting a portion of the Single Reflection Loss. The implication of the negative term is that the various reflections have additive phase relationships, and thus reduce the effectiveness of the shield. The shield effectiveness is essentially independent of frequency.

When the shield thickness exceeds $\lambda/4$, the Multiple Reflection Term becomes negligible, and there is no offsetting effect to the other losses. Thus the material shielding effectiveness increases, and is frequency dependent.

Table 4-5 provides representative calculations of the shielding effectiveness of thin-film copper for different thicknesses and frequencies, using equations 4-1, 4-2, 4-5 and 4-7. One-quarter wavelength in copper is approximately 32500 Angstroms (1 Angstrom = 3.94×10^{-9} in.) at 1 GHz, and it can be seen that shield effectiveness changes significantly above this thickness[2].

TABLE 4-5 CALCULATED VALUES OF COPPER THIN-FILM SHIELDING EFFECTIVENESS AGAINST PLANE-WAVE ENERGY

Shield Thickness	1050 Å		12500 Å		21960 Å		219600 Å	
Frequency	1MHz	1GHz	1MHz	1GHz	1MHz	1GHz	1MHz	1GHz
Absorption Loss, A	.014	.44	.16	5.2	.29	9.2	2.9	92
Single Reflection Loss, R	109	79	109	79	109	79	109	79
Multiple Reflection Correction Term, B	-47	-17	-26	-.6	-21	.6	-3.5	0
Shielding Effectiveness, S	62	62	83	84	88	90	108	171

Figures 4-8 and 4-9 contain measured shielding effectiveness data on copper, aluminum, silver, gold, and nickel materials [6]. A comparison between values computed in Table 4-5 and Figure 4-8b indicates differences between thin-film measurements and calculations of about 10 dB. One factor that may account for some of this difference is the difficulty in depositing a uniformly thick film. The dotted curves in Figures 4-8 and 4-9 are the experimenter's estimate of corrections necessary as a result of an error noted in his test setup that altered low frequency data.

2. One quarter wavelength = $.112 / (f\mu g)^{1/2}$ in meters.

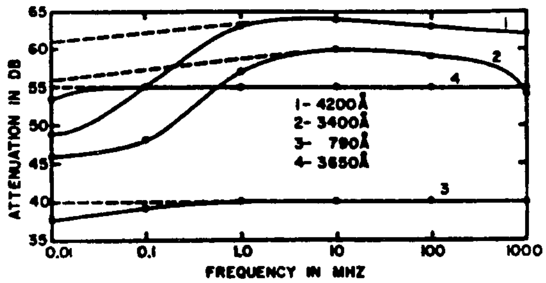


Figure 4-8a Shielding Effectiveness Curves, Copper

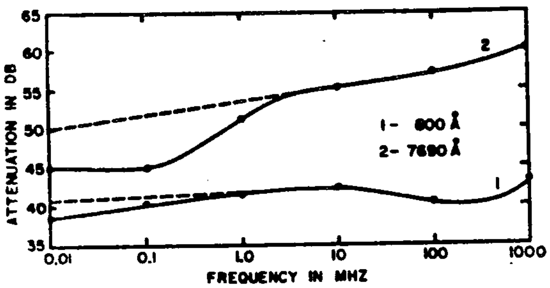


Figure 4-8b
Shielding Effectiveness
Curves, Copper

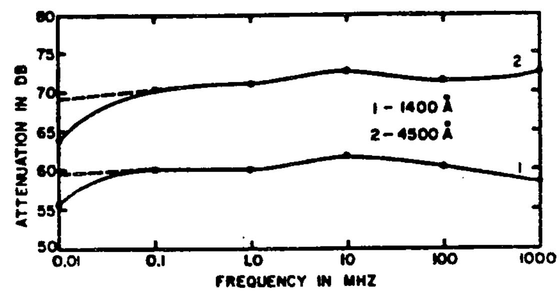


Figure 4-9a
Shielding Effectiveness
Curves, Silver

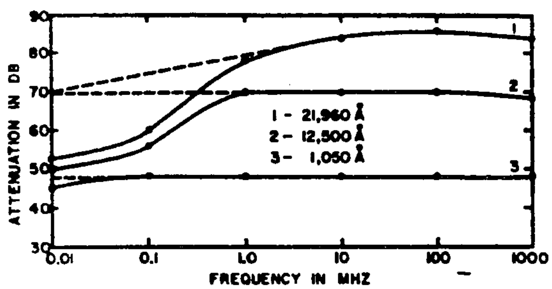


Figure 4-8c
Shielding Effectiveness
Curves, Aluminum

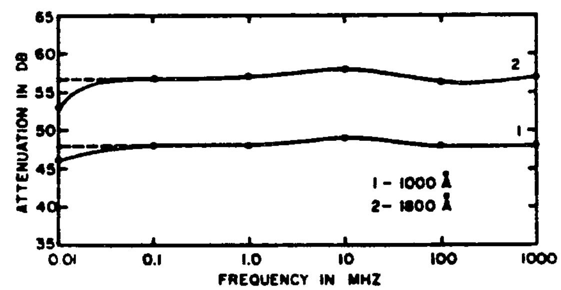


Figure 4-9b
Shielding Effectiveness
Curves, Gold

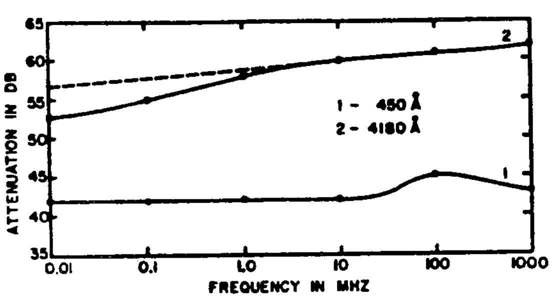


Figure 4-8d
Shielding Effectiveness
Curves, Aluminum

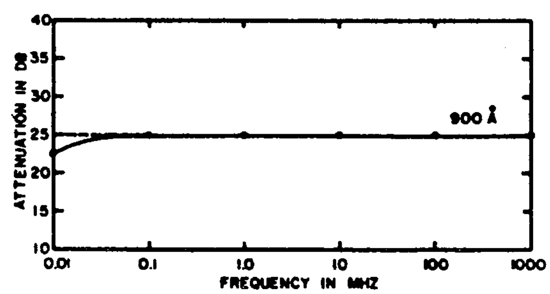


Figure 4-9c
Shielding Effectiveness
Curves, Nickel

Conductive coatings can be applied in a number of ways. Vacuum metalizing involves use of vacuum deposition equipment to apply a metallic film (usually aluminum) onto a surface; thicknesses of about 10^4 Angstroms are typical.

Flame spraying is a somewhat less expensive and time-consuming process than vacuum deposition and typically results in film thicknesses of 2.5×10^5 to 5×10^5 Angstroms for acceptable conductivity. One version consists of two consumable wires that are fed through electrode tips of a spray gun, where a 600 ampere electric current is transferred to them. As the wires emerge from the gun on intersecting paths, an arc is struck. The arc melts the consumable wires, and an air nozzle behind the wires atomizes the molten metal (see Figure 4-10a).

1. Electrode types
2. Consumable wire route
3. Cooling air circulation
4. Air nozzle

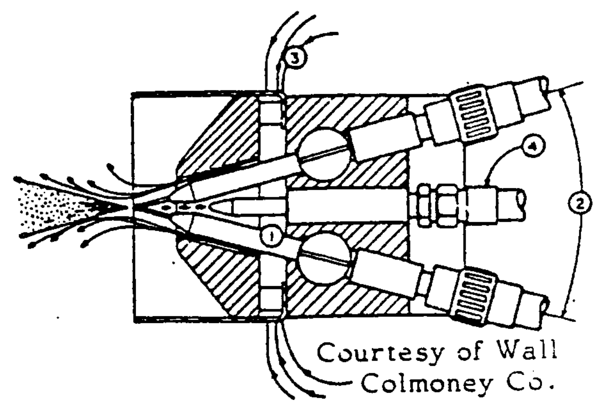


Figure 4-10a Details of Arc Flame Spray Gun

The flame-spray process indicated above, sometimes called arc or electrospray, operates at arc temperatures of 7000° F. This high temperature provides fast melting of the consumable wires, and a corresponding high spray deposition rate. For example, the flame spray process can deposit about 25 pounds of aluminum or 50 pounds of stainless steel per hour. Good bonds are achieved with less than ideal surface preparation, and relatively dense coatings are obtained.

Another form of flame spraying, sometimes called wire spray, involves a single consumable wire that is vaporized in a fuel gas and oxygen atmosphere. Compressed air impels the molten particles to the surface being sprayed. A cutaway view of a wire spray gun is shown in Figure 4-10b. Temperatures as high as 4500° F can be obtained, and a variety of spray rates can be achieved. This approach provides relatively fine-grain coatings at somewhat lower deposition rates than is available using the flame spray process.

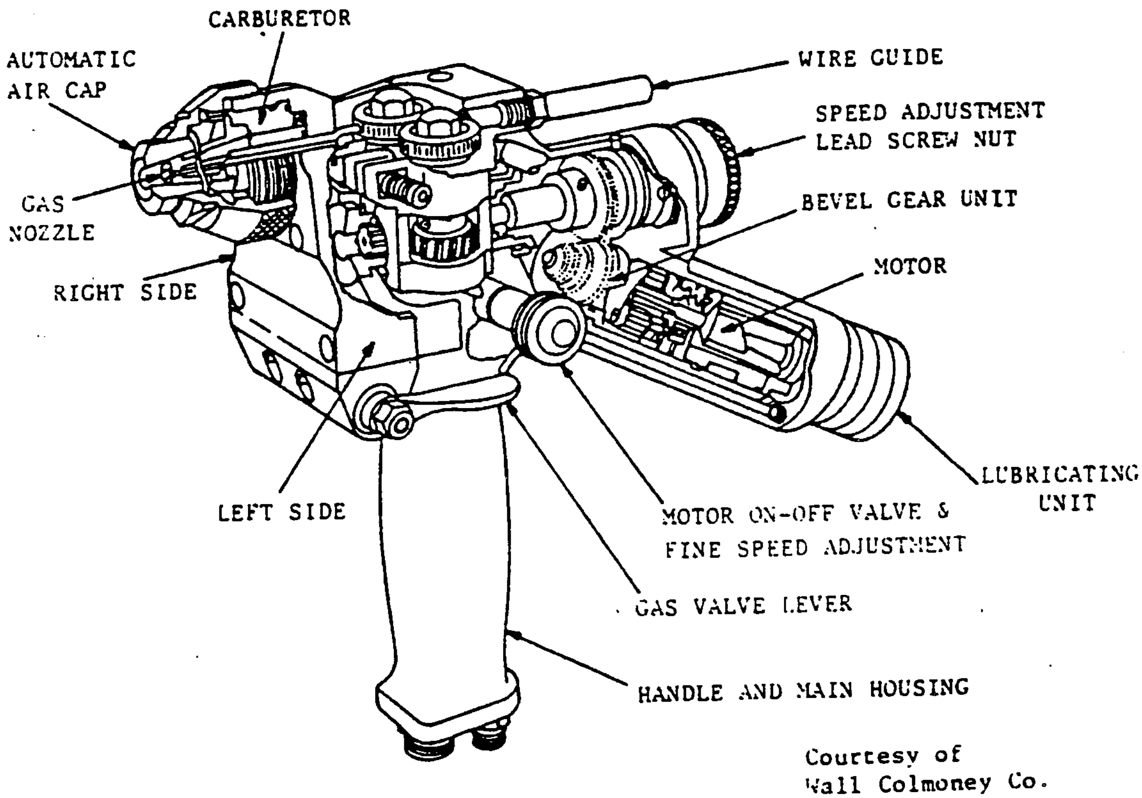
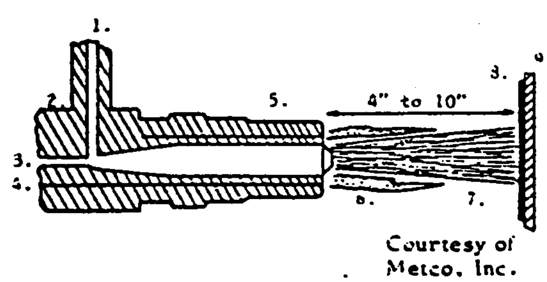


Figure 4-10b Cutaway of Wire spray Gun

A third form of flame spraying uses an oxy-gas flame in which powdered metal is melted. The powdered material is held in a hopper on top of the spray gun, is gravity-fed into the gun proper, and is carried to the gun nozzle by an oxygen-acetylene or hydrogen gas mixture. Then it is melted and projected onto the prepared surface. This process is illustrated in Figure 4-10c.



1.	Powder	2.	ThermoSpray Gun	3.	Powder Carrying Gas
4.	Fuel Gases	5.	Nozzle	6.	Burning Gases
7.	Spray Stream	8.	Sprayed Coating	9.	Prepared Base Material

Figure 4-10c Details of Powered Metal Spray Gun

The use of powdered metal offers a wide range of materials to be sprayed. The texture of the finish ranges from rough to smooth (depending on the powder used). Coating thickness can be typically 5000-8000 Angstroms, while coverage rates are as high as 500 square feet per hour at one mil thickness.

In addition to flame-spraying, conductive coatings can be applied in a number of other ways. These include:

Paint consisting of an acrylic-type carrier with silver-metallic or other conductive filler. The paint is sprayed on.

Plating of either rigid materials, or films of plastic.

Conductive tape or foil, with or without adhesive backing.

Conductive glass (a very thin conducting surface on a clear glass substrate) is often used in applications that require electromagnetic shielding properties but must maintain certain visual requirements. A typical shield thickness in this application is 10^4 Angstroms. A more detailed discussion of the applications of conductive glass will be found in Section 4.5.5.

4.3 NON-SOLID SHIELDING MATERIALS

4.3.1 Types of Discontinuities

An ideal shielded enclosure would be one of seamless construction with no openings or discontinuities. However, personnel, power lines, control cables, and ventilation ducts must have access to any practical enclosure. The design and construction of these discontinuities becomes very critical in order to incorporate them without appreciably reducing the shielding effectiveness of the enclosure. Some types of discontinuities commonly encountered include:

a. Seams between two metal surfaces, with the surfaces in intimate contact (such as two sheets of material that are riveted or screwed together).

b. Seams or openings between two metal surfaces that may be joined using a metallic gasket.

c. Holes for ventilation, or for exit or entry of wire, cable, light, film, water, meter faces, etc.

d. Screened openings.

Design construction considerations associated with such openings will be discussed in the next sections of this design guide.

4.3.2 Shielding Analysis [7],[8]

Leakage through openings in metal shields has been studied using transmission line theory. Based on these studies, the Shielding Effectiveness, S, of non-solid shielding materials has been defined as:

$$S = A_a + R_a + B_a + K_1 + K_2 + K_3 \qquad (4\text{-}9)$$

where

A_a = Absorption (Penetration) loss introduced by a particular discontinuity, in dB

R_a = Aperture Single Reflection Loss, in dB

B_a = Multiple Reflection Correction Term, in dB

K_1 = A correction term to take into account the number of like discontinuities

K_2 = A low-frequency correction term to take into account skin depth

K_3 = A correction term to take into account coupling between adjacent holes.

The first three terms in Equation (4-9) generally correspond to the three terms of Equation (4-1), while the last three terms encompass other factors that need not be considered for solid sheets. The analysis is most appropriate for single discontinuities, or for identical and uniformly spaced apertures (such as screening or perforated sheets), but can be extended to somewhat more complex configurations as well.

The absorption loss term, A_a, is obtained by assuming that the impinging wave is well below its cutoff frequency.[3] Under these circumstances:

$$A_a = 27.3\ d/w \text{ (for rectangular aperture)} \qquad (4\text{-}10a)$$

$$A_a = 32\ d/D \text{ (for circular aperture)} \qquad (4\text{-}10b)$$

where

d = depth of aperture, in inches

w = width of aperture perpendicular to the E-field vector, in inches

D = aperture diameter, in inches.

These equations are plotted in Figure 4-11a.

---

3. The cut-off frequency (F_c) is that frequency below which the propagation constant of the aperture is real. It is defined by $f_c = c/\lambda_c$, where c is the velocity of light and λ_c is the cut-off wavelength, in the same units. The cut-off wavelength is equal to twice the maximum dimension of a rectangular aperture, or ~~3.142~~ 3,412 times the radius of a circular aperture.

The multiplying factor to be applied to equation (4-10a and 4-10b) for frequencies f not well below cut-off is $[1 - (f/f_c)^2]^{1/2}$.

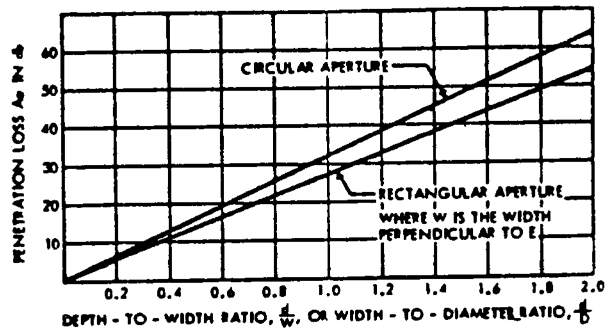


Figure 4-11a Aperture Shielding Absorption Loss

The aperture single reflection loss, R_a, is wave-impedance dependent and is also a function of aperture shape. It can be expressed as:

$$R_a = 20 \log \left[\frac{1+|k|^2}{4|k|} \right] \qquad (4\text{-}11)$$

where

K = $w/\pi r$, for low impedance fields and rectangular apertures

K = $D/3.682\ r$, for low impedance fields and circular apertures

K = $jfw(1.7\times10^{-4})$, for plane-wave fields and rectangular apertures

K = $jfd\ (1.4 \times 10^{-4})$, for plane-wave fields and circular apertures

f = frequency, in MHz

r = source to shield distance, in inches

$j = \sqrt{-1}$

Graphs of R_a for the low-impedance case are shown in Figure 4-11b.

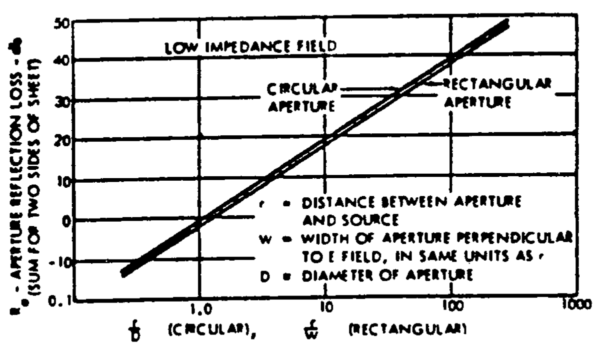


Figure 4-11b Aperture Shielding Reflection Loss

The multiple reflection correction term, B_a, is defined by the equation:

$$B_a = 20 \log \left[1 - \frac{(K-1)^2}{(K+1)^2} \; 10^{-Aa/10} \right] \tag{4-12}$$

and is applicable only when A_a is less than 15 dB. A curve for the cases when either $w/\pi r$ or $D/3.682\ r$ are much less than unity is plotted in Figure 4-11c.

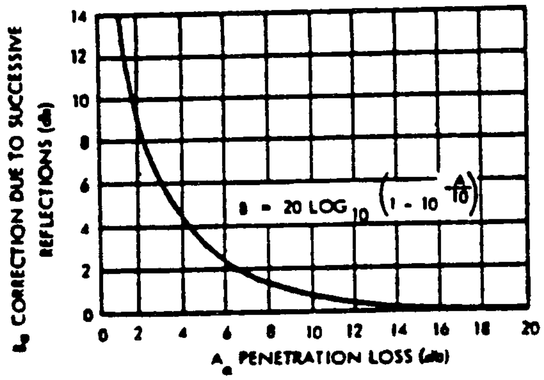


Figure 4-11c Aperture Shielding Correction Due to Re-reflection

The number of discontinuities correction term, K1, is employed when the source is located a large distance from the shield, in comparison with the aperture spacing in the shield. It is represented by:

$$K_1 = -10 \log (an) \tag{4-13}$$

where

a = area of each hole, in square inches

n = number of holes per square inch

The correction term can be ignored for sources close to the shield. Equation (4-13) is plotted in Figure 4-11d.

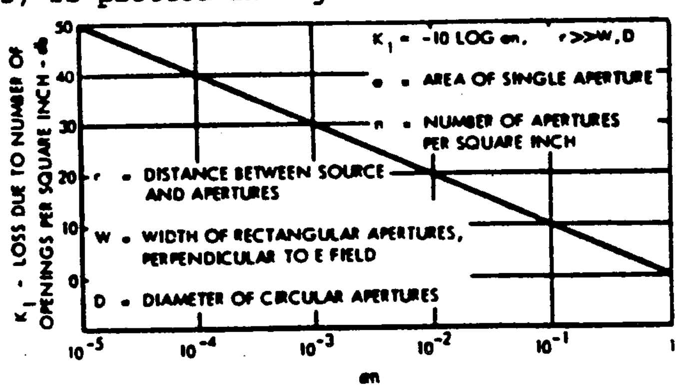


Figure 4-11d Aperture Shielding Loss Due to Density of Openings

The skin depth correction term, K_2, recognizes that, at low frequencies, when the skin depth becomes comparable to the screening wire diameter or dimension between apertures, a reduction in shielding effectiveness can occur. An empirical relationship for this effect was developed:

$$K_2 = -20 \log (1 + 35/p^{2.3}) \qquad (4\text{-}14)$$

where

p = the ratio of wire diameter to skin depth, for screening

p = the ratio of conductor width between holes to skin depth, for perforated sheets

For convenience, the equation for skin depth in copper, in inches, is

$$SD_{cu} = 2.6\text{x}10^{-3}/\sqrt{f} \qquad (4\text{-}15)$$

Equation (4-14) is presented in Figure 4-11e.

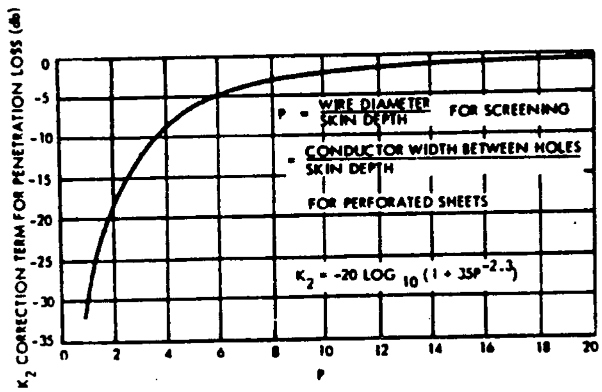


Figure 4-11e Aperture Shielding Correction for Skin Effect

The adjacent hole coupling correction term, K_3, is the result of noting that shielding efficiency is higher than expected when apertures in a shield are closely spaced and the depth of the openings are small compared to the aperture width. This is interpreted as the result of coupling between adjacent holes, and becomes important for small openings. The equation for computing K_3 is:

$$K_3 = -20 \log [\coth(A_a/8.686)] \qquad (4\text{-}16)$$

This equation is plotted in Figure 4-11f.

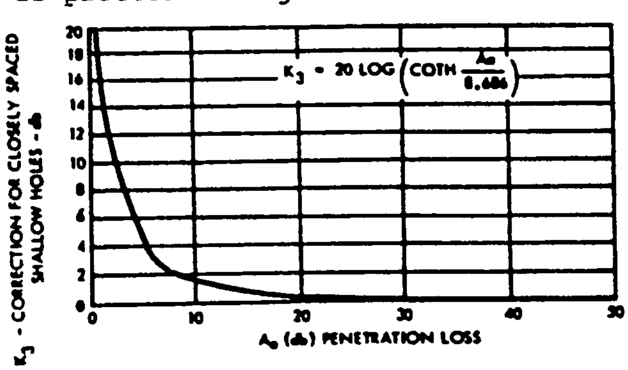


Figure 4-11f Aperture Shielding Correction for Closely Spaced Shallow Holes

The elements of the non-solid shielding effectiveness equation just discussed are summarized in Table 4-6 for convenience. Calculations in accordance with these equations have resulted in fairly good agreement with measurements, so long as the aperture pattern is uniform. Examples of this correlation are shown in Figure 4-12 for the low impedance shielding effectiveness of a variety of mesh copper screens. Both measurements and calculations used a loop-to-loop separation of 3.5 inches, and the measurements employed 3-inch diameter loops in coplanar orientation. Table 4-7 also provides similar comparative data [7].

TABLE 4-6 TERMS FOR APERTURE SHIELDING

Symbol	Item	Aperture		Comments
		Rectangular	Circular	
A_a	Penetration Loss	$27.3\frac{d}{w}$	$32\frac{d}{D}$	d=Depth of Aperture, inches w=Width of Rectangular Aperture, Perpendicular to E field, inches D=Dia. of Circular Aperture, inches
R_a	Aperture Reflection Loss (sum for 2 sides of sheet)	$20 \log_{10} 0.785\frac{r}{w}$	$20 \log_{10} 0.92\frac{r}{D}$	$\ll 377\,\Omega$
		$20 \log_{10}\frac{1+(1.7\times10^{-4}Fw)^2}{4(1.7\times10^{-4}fw)}$	$20 \log_{10}\frac{1+(1.47\times10^{-4}fw)^2}{4(1.47\times10^{-4}fw)}$	$377\,\Omega$
		----------------	-----------------	$\gg 377\,\Omega$
B_a	Correction Term Due to Successive Reflections When $A_a < 15$ db	$20 \log_{10} (1-10^{-\frac{A_a}{10}})$		$\frac{w}{\pi r'}\ \frac{D}{3.68r} \ll 1$
K_1	Loss Term Due to Number of Openings Per Unit Square	$-10 \log_{10} an,\ r \gg \omega, D$ zero, $r \ll \omega, D$		a Area of Single Aperture, inches^2 n=No. of Holes per sq. inch
K_2	Correction Term for Penetration of Conductor at Low Frequencies	$-20 \log_{10} (1+35p^{-2.3})$		P= Wire Dia./Skin Depth, Screening P= Conductor Width Between Holes/Skin Depth Perforated Sheets
K_3	Correction for Closely Spaced Shallow Holes	$20 \log_{10} (\coth \frac{Aa}{8.686})$		

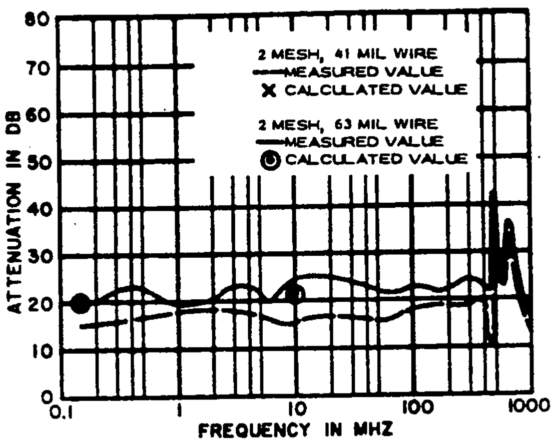


Figure 4.12a
Measured and Calculated
Insertion Loss of Copper
Screens

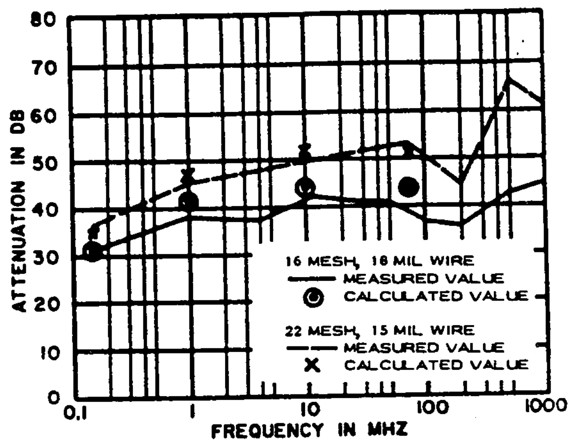


Figure 4-12b
Measured and Calculated
Insertion Loss of Copper
Screens (Cont'd)

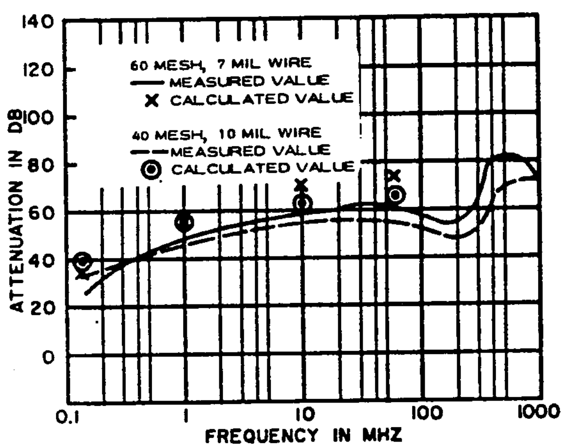


Figure 4-12c
Measured and Calculated
Insertion Loss of Copper
Screens (Cont'd)

TABLE 4-7
COMPARISON OF MEASUREMENTS
AND CALCULATIONS OF
SCREENING MATERIAL
SHIELDING EFFECTIVENESS

Screen Type*	Test Type	Freq (MHz)	Meas Values (DB)	Calc Values (DB)
No. 22 15 mil	Mag field	0.085 1.0 10.0	31 43 43	28 45 49
No. 22 15 mil	Plane Waves	0.2 1.0 5.0 100.0	118 106 100 80	124 110 95 70
No. 22 15 mil	Elec field	0.014 to 60	65	**65
No. 12 20 mil	Elec field	0.014 to 60	50	**53

* All screens made of copper.
** These values assume a wave impedance equal to that of a 30-inch square waveguide.

Figures 4-13 and 4-14 show the shielding effectiveness, respectively, of 3 Oz. (per square foot) copper screen and 24 oz. (per square foot) copper and silicon steel screen at frequencies up to 10 GHz [26]. The enclosure of Figure 4-13 is made up of two electrically insulated layers of 3 oz. copper sheet backed with saturated tar paper on both interior and exterior sides. The enclosure of Figure 4-14 is made up of two shields. Each shield consists of an outer layer of 24 oz. cold rolled copper laminated to an interior layer 24 gauge M-19 silicon steel, for a total of four layers of shielding material. The superiority of the shielding arrangement of Figure 4-14 is clearly evident.

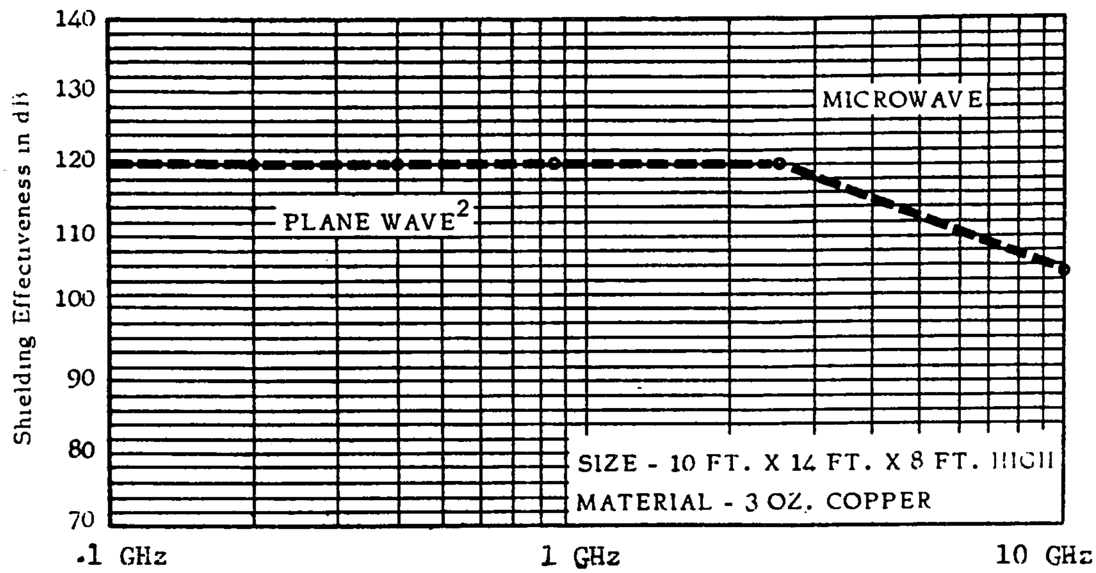


Figure 4-13 Shielding Effectiveness Tests on Double Electrically Shielded Enclosure

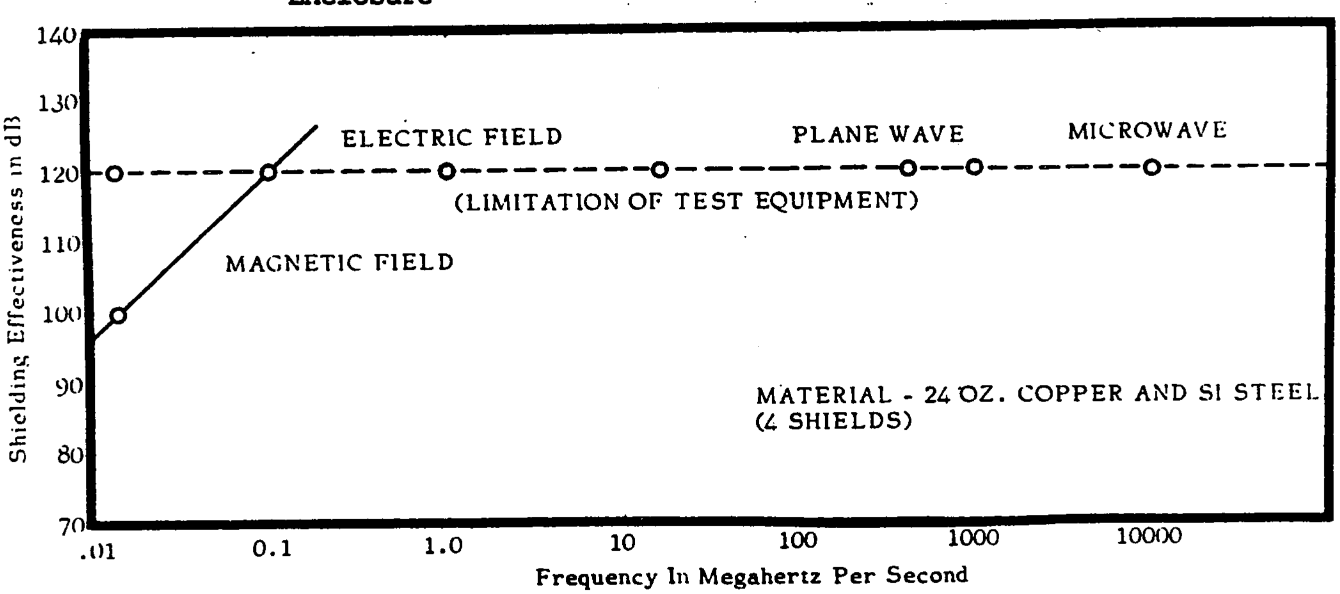


Figure 4-14 Shielding Effectiveness Test on a Double Electrically Insulated Shielded Cube 34" x 36" x 30"

Representative non-solid sheet shielding effectiveness measurements are shown in Tables 4-8 and 4-9, and in Figure 4-15 [2]. The two tables provide data on a variety of material forms, including meshes, perforated sheets, and cellular structures against low-impedance, high-impedance and plane waves. The curves of Figure 4-15 were obtained using source and pick-up coils on opposite sides of a flat sheet of shielding material. Experimental results strongly support the validity of plane-wave shielding theory, even at the extremely low frequencies used. Figure 4-15 illustrates how the low-impedance effectiveness of high permeability perforated sheet material changes with changes in hole size and hole separation [5].

TABLE 4-8
EFFECTIVENESS OF NON-SOLID SHIELDING MATERIALS AGAINST LOW-IMPEDANCE AND PLANE WAVES

<table>
<tr><th rowspan="3">Impinging Wave</th><th colspan="2" rowspan="2">Form</th><th rowspan="3">Material</th><th rowspan="3">Thick-ness (mils)</th><th colspan="6">Nominal Effectiveness (db)</th></tr>
<tr><th rowspan="2">0.1 kHz</th><th rowspan="2">1 kHz</th><th rowspan="2">10 kHz</th><th rowspan="2">85 kHz</th><th rowspan="2">1 MHz</th><th rowspan="2">10 MHz</th></tr>
<tr><th>General</th><th>Detail</th></tr>
<tr><td rowspan="4">Low Impe-dance</td><td rowspan="4">Mesh (Screening)</td><td>2 layers 1 inch apart</td><td>Cu (oxidized)</td><td></td><td>2</td><td>6</td><td>18</td><td></td><td></td><td></td></tr>
<tr><td>No. 22</td><td>Cu</td><td></td><td></td><td></td><td></td><td>31</td><td>43</td><td>43</td></tr>
<tr><td>No. 16</td><td>Bronze</td><td></td><td></td><td></td><td></td><td>18</td><td></td><td></td></tr>
<tr><td>No. 4</td><td>Galvan-ized Steel</td><td></td><td></td><td></td><td></td><td>10</td><td>17</td><td>21</td></tr>
<tr><td rowspan="6">Plane</td><td rowspan="2">Perfor-ated Sheet</td><td rowspan="2">45 mil dia., 225 sq. inch</td><td rowspan="2">Al</td><td rowspan="2">20</td><td colspan="3">3040 MHz</td><td colspan="3">9380 MHz</td></tr>
<tr><td colspan="3">60</td><td colspan="3">62</td></tr>
<tr><td rowspan="4">Mesh (Screening)</td><td rowspan="2">No. 16</td><td rowspan="2">Al</td><td rowspan="2">dia=13</td><td colspan="3" rowspan="2">34</td><td colspan="3" rowspan="2">36</td></tr>
<tr></tr>
<tr><td rowspan="2">No. 22</td><td rowspan="2">Cu -</td><td rowspan="2">dia=15</td><td colspan="2">200 kHz</td><td>1 MHz</td><td colspan="2">5 MHz</td><td>100 MHz</td></tr>
<tr><td colspan="2">118</td><td>106</td><td colspan="2">100</td><td>80</td></tr>
</table>

TABLE 4-9
EFFECTIVENESS OF NON-SOLID SHIELDING MATERIALS AGAINST HIGH-IMPEDANCE WAVES

FORM		MATERIAL	THICKNESS (mils)	NOMINAL EFFECTIVENESS (db)	OPEN AREA (%)	AIR-FLOW STATIC PRESSURE (inches of water)	
GENERAL	DETAILED			14 kc to 1000 mc		200 cu ft/min	400 cu ft/min
Hexcell	1/4-inch cell, 1 inch thick	Al	3	>90		0.06	.26
TV Shadow Masks (Photo-Etched)	9-mil holes, 28-mil centers	95% Cu 5% Ni	7	>90	12	>2	
					50	0.2	0.4
		100% Ni	3	>90	50	0.2	0.5
Lektromesh	40 count	Cu-Ni	7	>90	36	0.4	1.7
	25 count		5	78	49	0.2	0.5
	40 count	Cu	3	78	57	0.2	0.5
	25 count				56	0.2	0.4

FORM		MATERIAL	THICKNESS (mils)	NOMINAL EFFECTIVENESS (db)	OPEN AREA (%)	AIR-FLOW STATIC PRESSURE (inches of water)	
GENERAL	DETAILED			14 kc to 1000 mc		200 cu ft/min	400 cu ft/min
Perforated Sheet	1/8-inch dia., 3/16-inch centers	Steel	60	58		0.27	>0.6
	1/4-inch dia., 5/16-inch centers	Al	60	48	46		
	7/16-inch dia., 5/8-inch centers		37	35	45		
Mesh (Screening)	No. 16 16 x 16/sq.in.	Al	20 (dia)	55	36		
	No. 22	Cu		65(14 kc to 60 mc)			
	No. 12		20 (dia)	50	50		
	No. 16	Bronze		45(14 kc to 60 mc)			
	No. 10	Monel	18 (dia)	40			
	No. 4	Galvanized Steel	30 (dia)	35 28 (14 kc to 40 mc)	76		
	No. 2			24	88		

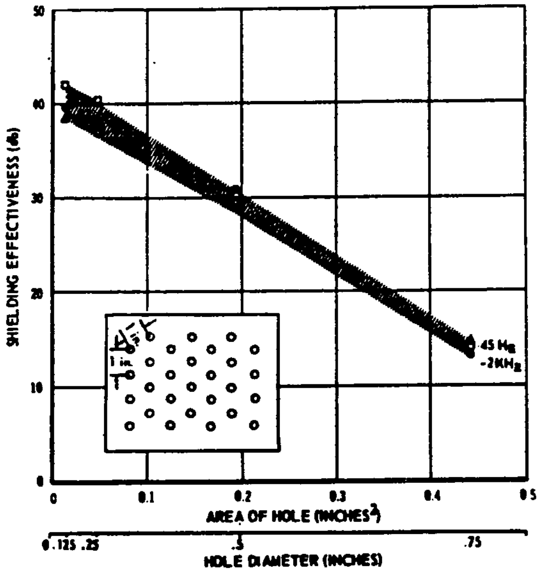


Figure 4-15a
Shielding Effectiveness
Perforated AMPB-65 as a
Function of Hole Size

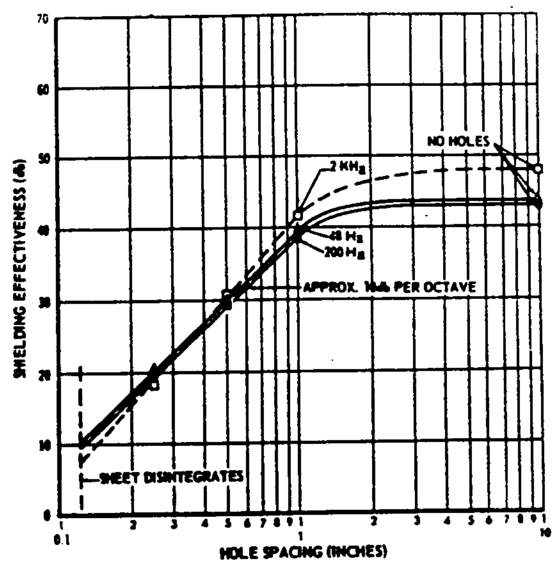


Figure 4-15b
Shielding Effectiveness of
Perforated AMPB-65 as a
Function of Hole spacing
(hole diameter = 0/125")

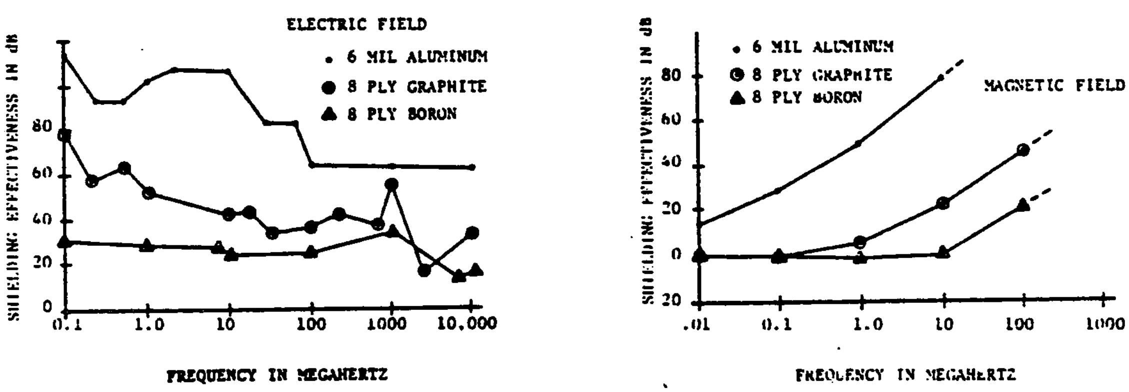


Figure 4-15c Typical Composite Shielding Effectiveness Characteristics

4.3.3 Composite Materials

In recent years, composite materials have been developed to the point that their strength and weight characteristics make them very attractive as substitutes for metals in various aircraft applications and even as cases for GSE. Composites are a family of polymeric or metal-reinforced materials having organic or inorganic fibers or filaments, woven fabric layering, and epoxy bonding. Boron, graphite, Kevlar 49 and glass fibers are employed, with boron and graphite most frequently used [1]. Recent approaches include nickel plating graphite fibers, aluminum coating glass or graphite fibers, and graphite coating stainless steel fibers. Applications of these composite materials include C-130 aircraft wing boxes and the AV-8B wing, as well as portions of F/A-18 and A-7 aircraft.

Boron filaments are produced by vapor deposition on a tungsten wire resulting in a very stiff, high strength, low density fiber. The boron composite is very limited in ability to transport electrical current. This fact makes the boron composite susceptible to high currents, such as lightning, and to charge-accumulated high potentials, such as static electricity.

Graphite filaments are made by pyrolysis of polyacrylonitrile, or decomposition of a material similar to rayon. This produces a strong, stiff low cost filament. The graphite material is more conductive than boron and, therefore somewhat less susceptible to electrical phenomena. In addition, graphite possesses basic electromagnetic characteristics which, although reduced from metals, are improvements over other composites.

The extent to which composites alter the electrical characteristics of the aircraft clearly depends on the amount and location of the material on the aircraft. The areas of major concern as far as composites are concerned are as follows:

Lightning -

The effects of lightning strikes to aircraft are known to be potentially severe even in aircraft of a typical metal variety. Investigations employing both destructive and non-destructive testing techniques have affirmed that lightning-produced currents may adversely affect boron and graphite-reinforced composites that are installed on aircraft. While composites offer significant structural advantages over conventional metals, they are much more easily damaged by the high currents associated with lightning [2]. Methods of improving the lightning protection of composites have included use of wire mesh, metalized paint, flame-sprayed aluminum and dielectric coatings.

Shielding Effectiveness -

The shielding effectiveness of composites is much lower than that of conductive metals. Essentially, the composite provides little protection from flux linkage of lightning energy to internal sub-systems and equipment. In addition, little attenuation is offered to other radiated environmental signals. Typical shielding effectiveness values of boron and graphite composites to electric and magnetic fields are compared against sheet aluminum in Figure 4-15c. The more limited shielding provided by the airframe transfers the shielding burden to internal sub-systems of the aircraft.

Electrical System Grounding -

The applications of composites in aircraft will undoubtedly have a significant impact on all aspects of electrical system grounding. Since composites are not capable of supporting significant current flow, they cannot serve as a system ground-plane. Alternative approaches, such as use of wired electrical power returns and ground buses become necessary.

Static Electricity -

Very little is known as to how composites will be affected when exposed to environments which cause static charging. Static electrical charging from various sources, such as rain, ice, dust, air, engine exhaust, etc. can cause equipment malfunction, composite damage, fuel charging, and personal shock. Traditional techniques for preventing static problems depend almost entirely on low resistance conductive paths for charge dissipation. Another approach is to apply anti-static coatings on composites. A new composite type, Varistor Composite Materials, may provide another approach [33]. Varistor composites are made of conductive powders and resins. The conductive particles are encased in a dielectric. For high voltage transients, like ESD, the conductivity increases. Thus, this material acts like a voltage clamp and absorbs short duration energy pulses capacitively, dissipating the charge resistively over a longer period of time.

In addition to the above items, composites may have other effects on aircraft operation and performance. These include changes of antenna performance (directivity, gain, VSWR, etc. when a metallic counterpoise is replaced by a composite), changes in aircraft radar cross-section, and others.

A unique problem associated with graphite composites is the contamination effect caused by the conductive filaments when the composite structure is altered. For example, in the event of an aircraft explosion or fire, it is possible for graphite fibers to disperse over a wide area. If some of these fibers come in contact with shipborne or shorebased electronic equipment, it is possible that they can cause circuit shorts, or otherwise degrade performance.

4.4 CABLES AND CONNECTORS

4.4.1 Cable Shielding

There are several methods for shielding cables. These include: (a) braid (b) flexible conduit, (c) rigid conduit, (d) spirally wound shields of high permeability materials, and (e) encasing the conductor in a ferrite material (Filter-line cable).

Braid, which constitutes woven or perforated material, is used for cable shielding in applications where the shield cannot be made of solid material. Advantages are ease of handling in cable make-up and lightness in weight. However, it must be remembered that for radiated fields, the shielding effectiveness of woven or braided materials decreases with increasing frequency and the shielding effectiveness increases with the density of the weave. The percentage coverage by a braided shield has been a critical parameter in the design of flight-deck operated GSE.

Conduit, either solid or flexible, may also be used to shield weapon system cables and wiring from the RF environment. The shielding effectiveness of solid conduit is the same, for RF purposes, as that of a solid sheet of the same thickness and material. Linked armor or flexible conduit may provide effective shielding at lower frequencies, but at higher frequencies the openings between individual links can take on slot-antenna characteristics, seriously degrading the shielding effectiveness. If linked armor conduit is required, all internal wiring should be individually shielded. Degradation of shielding conduit is usually not because of insufficient shielding properties of the conduit material but rather the result of discontinuities in the cable. These discontinuities usually result from splicing or improper termination of the shield [9].

Solenoids, or other devices associated with high in rush currents or incorporating switching devices that normally develop high-amplitude transients, might be located in the vicinity of shop-oriented GSE. For protection against this type of energy, shielding materials with high permeability are desirable. These materials cannot be drawn into tubing because they lose their shielding properties when cold worked; therefore, an adequate shield is often developed by wrapping a continuous layer of annealed metal tape around the cable. A protective rubber coating is recommended.

The principal types of shielded cables that are available include shielded single wire, shielded multiconductor, shielded twisted pair, and coaxial. Cables are also available in both single and multiple shields in many different forms and with a variety of physical characteristics.

One physical configuration is the filterline cable. Similar to annealed metal tape, the shielding material has a high permeability; in this case, however, the material is a ferrite. The ferrite encases the conductor and takes advantage of the loss versus frequency characteristics of this material (see Section 7.2.1.7 for information on Raychem filterline cable).

Data on the shielding effectiveness of cables is not easily measured. This is because a large number of parameters (some of them external to the cable itself) dictate the particular performance of a cable, including termination impedances, impinging signal direction and impedance, cable length frequency, the particular connectors employed, flexing requirements, and others. However, at frequencies where diffusion effects are negligible, we can determine the shielding effectiveness in a way that is invariant with changes in cable length or experimental conditions. We use the triaxial test method to obtain a dB versus frequency spectrum. We then determine the Surface Transfer Impedance (STI) and Geometric Transfer Ratio (GTR) that accompany a best fit between a calculated spectrum and the measured spectrum [31]. This enables us to predict the shield leakage of a cable at any frequency, for any length, and under any definable set of test conditions.

The general characteristics of four classes of shielded cables are identified in Table 4-10 [9]. The classes include rigid and flexible conduit, foil-wrapped cable, and braided shielded cable. As indicated previously, shielding effectiveness in most cabling applications is dependent on the percentage coverage of the cable provided by the shield.

Table 4-10
Comparison of Shielded Cables

	Copper Braid	Foil	Solid Conduit	Flexible Conduit
Shield Effectiveness (audio frequency)	Good	Exc.[1]	Exc.	Good
Shield Effectiveness (radio frequency)	Good	Exc.[1]	Exc.	Poor
Normal % of Coverage	60-95%	100%	100%	90-97%
Fatigue Life	Good	Fair	Poor	Fair
Tensile Strength	Exc.	Poor	Exc.	Fair

[1] may lose its effectiveness when flexed.

For example, MIL-C-7078/35 Notice 1 is one of several specifications sometimes referenced in GSE procurement. This specification calls for the percent shield coverage (the ratio of metal to total possible shield surface) to be at least 85%. It also imposes a requirement on the angle of the braid wires relative to a plane normal to the cable axis (a parameter that influences percentage shield coverage under cable flexing conditions) of between 10 and 40 degrees.

For GSE flight deck applications, tests made on particular systems indicate it is necessary to impose a shield coverage requirement of 94% on braided cable. The shield braid angle and percent braid coverage are determined in accordance with MIL-C-7078C, as follows:

$$\text{Tan } a = 2\pi\,(D + 2d)\,P/C \qquad (4\text{-}17)$$

$$K = 100\,(2F - F^2) \qquad (4\text{-}18)$$

where

a = Shield braid angle (angle of braid with axis of cable)

K = Percent coverage

d = Shield strand diameter in inches

C = Number of carriers

D = Diameter of cable under braid, in inches

F = NPd/Sin a

N = Number of strands per carrier

P = Picks per inch of cable length

Note: For 2-conductor cable only

$$D = \frac{(\pi+2)b}{\pi} \qquad (4\text{-}19)$$

where

b = diameter of basic wire

Some of these parameters are defined in Figure 4-16.

Figures 4-17 and 4-18 illustrate "bad" and "good" cable shielding for flight deck usage. Both pictures were taken with the cable flexed, and show the effects that may be expected from 85% and 94% coverage shields, respectively.

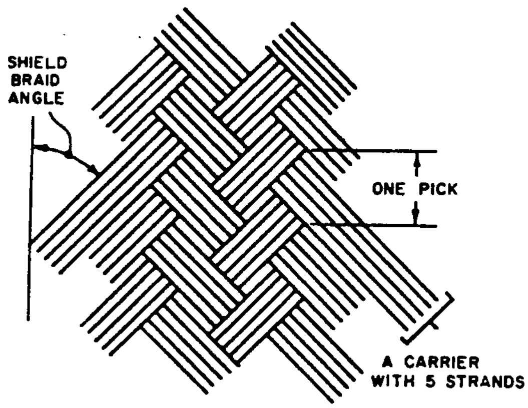


Figure 4-16
Definitions of Cable
Shield Parameters

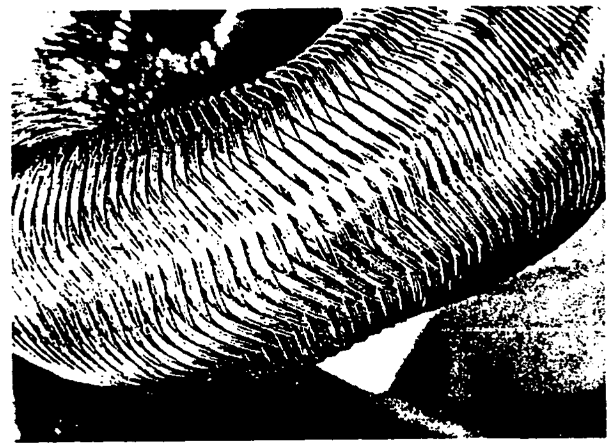


Figure 4-17
Braided Shield Cable with 85%
Coverage Under Flexing Condition

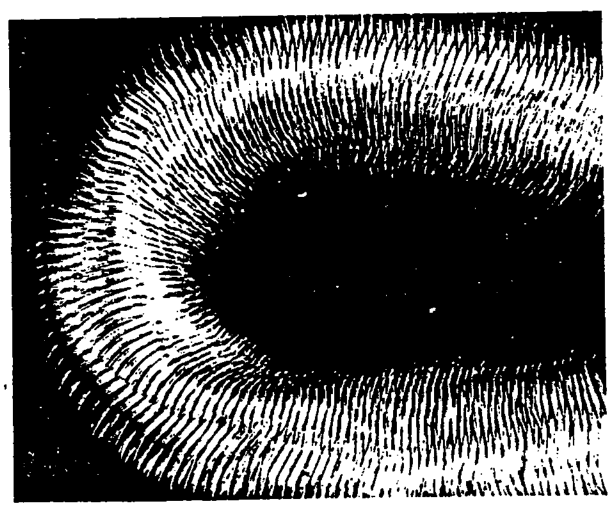


Figure 4-18 Braided Shield Cable with 94% Coverage Under Flexing Condition

Much of the previous discussion relates to the shielding of cables against plane wave or high impedance fields. For shielding against magnetic fields, the use of annealed high permeability metal strips wrapped around the cable has already been indicated. Multiple layers of counterspiral-wound nickel-iron or silicon iron alloys, or low carbon steel have proven effective under these circumstances. High permeability tape is available with or without adhesive backing. Also, combination high permeability, high conductivity tape is available to provide both electric and magnetic shielding.

4.4.2 Cable Shield Terminations and Connectors

If the effectiveness of a shield is to be maintained, the cable shield must be properly terminated. In otherwise adequately shielded systems, RF currents that are conducted along shields can be coupled to the system wiring from the point of an improper cable termination. This is a particularly important consideration in the case of cables exposed to high power RF fields. In fact, one school of thought is that EMC is affected mostly by system cable effects, rather than leakage through cracks, seams, etc. [32]. In the study conducted in [32], it was found that a shielded conductor carrying an RF current induces the negative of that current on the inside surface of the shield. When this return current is confined to the inside surface, the shielding is most effective. When this return current is not confined to the inside surface, some of it flows down the outside surface reducing shielding effectiveness.

Thus, in a properly terminated shield, the entire periphery of the shield is grounded to a low impedance reference, minimizing any RF potentials at the surface of the termination. Solder is undesirable in terminating RF coaxial cables because: (a) too much solder increases the center conductor diameter, thus increasing shunt capacitance and, (b) too little solder increases the current path, thus increasing series inductance. Specification MIL-E-45782B recommends against use of soldering to terminate shields because of the danger of damaging conductor insulation, and suggests a variety of termination methods, all involving crimping operations. A frequently used method of shield termination is illustrated in Figure 4-19. In this arrangement, the cable shield is flared so that it extends over the rear portion of the sleeve, and the crimp ring is slid into place over the sleeve. A crimping tool is then used to crimp the crimp ring onto the sleeve.

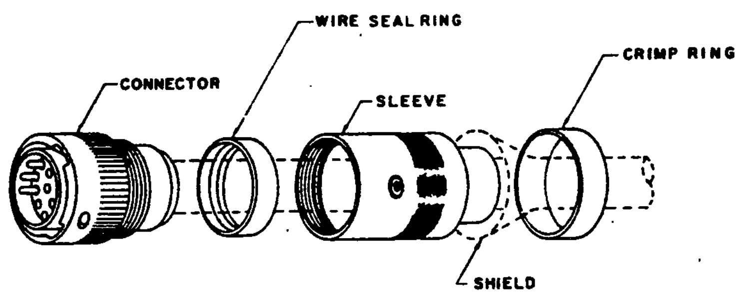
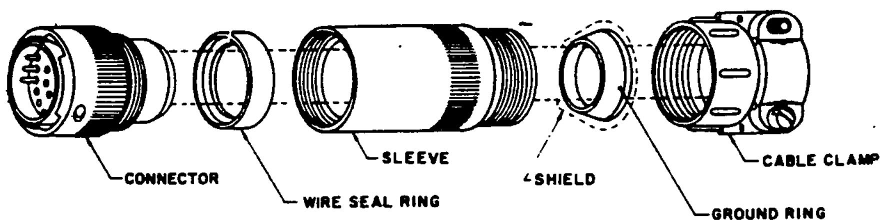
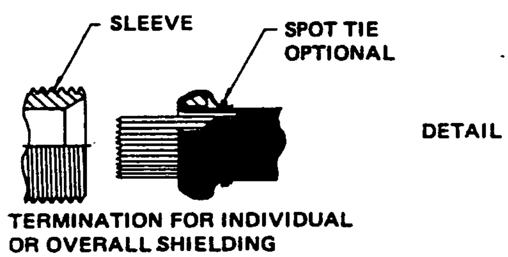


Figure 4-19 Shield Termination Using Crimping

An alternative to crimping is shown in Figure 4-20. The shield is placed through the ground ring and flared over and around the ring, and may be secured to the ring with a spot tie; see detail in Figure 4-20 a). The ground ring is then slid into the rear of the sleeve, which has a tapered base. Tightening the cable clamp onto the end of the sleeve assures positive 360 degrees grounding of the shield, and provides a strain relief for the cable. The use of silver epoxy or other synthetic conducting material has been found to be unacceptable for shield bonding because of lack of mechanical strength necessary for this application. A variation on this termination method includes environmental seals in the backshell construction; see Figure 4-20 b). This is of particular value for GSE which will operate in the corrosive marine environment of salt water/salt spray.



a)



b)

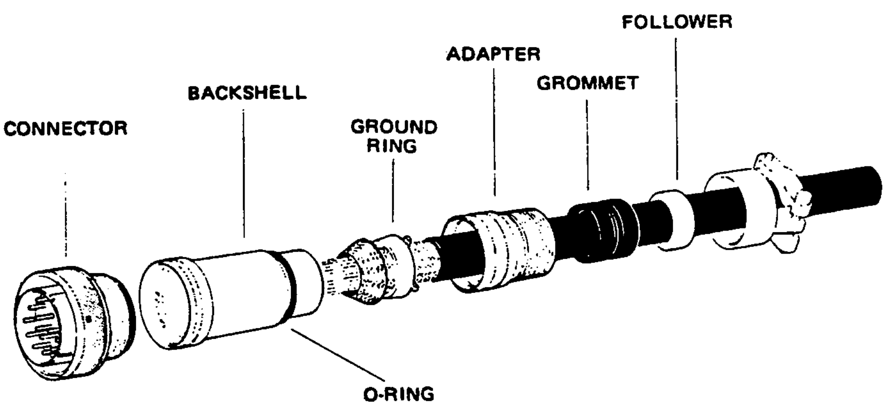


Figure 4-20 Shield Termination Using Threaded Assembly

For GSE that must operate in high field strength environments, both shields and solid cylindrical members have been terminated using the Magnaforming process. Magnaforming is a metal-forming technique that is used to shrink metal tubes and similar shapes around other forms such as collars, sleeves, rods, etc. The process uses a very intense, short duration magnetic field to induce an opposing current in the tubing or sleeve which is to be shrunk. The magnetic field produced by the induced current in the sleeve opposes the field produced by the current in the magnetic coil, and is of equal magnitude.

The resulting force from the magnetic coil field and the force from the field around the sleeve, combine to produce a uniform compression of the sleeve around the entire sleeve periphery. If the current applied to the magnetic coil is of sufficient amplitude, the combined resultant compressive force on the sleeve will deform the sleeve, reducing its diameter until it conforms to the shape of the underlying material. The basic process is shown in Figure 4-21.

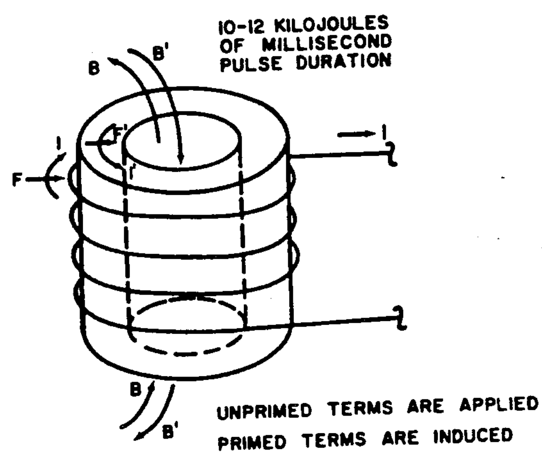


Figure 4-21 Relationship of F, B, and I in Magnaforming Process

Figure 4-22 shows a cross-section of a fixture in which a work piece is to be shrunk around a cable sleeve and connector. An insulating "former" is normally used between the magnetic coil assembly and the work piece, but the insulator does not directly enter into the forming operation.

Figure 4-23 is a cut-away photograph of two connectors terminating opposite ends of a short length of cabling in conduit. The Magnaforming process was used to make the terminations at the points identified in the figure.

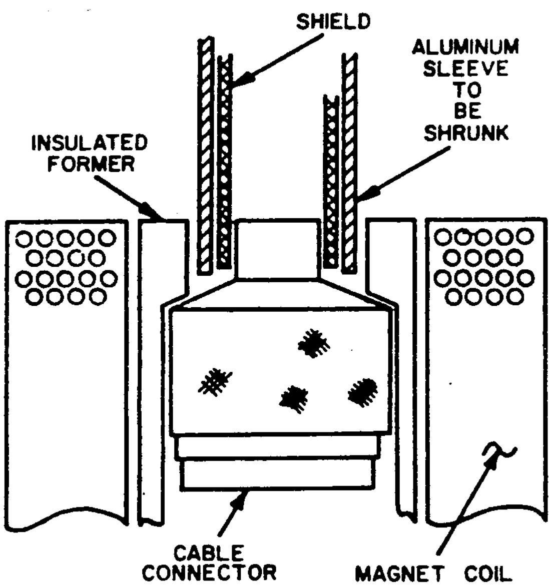


Figure 4-22
Typical Magnaforming Process

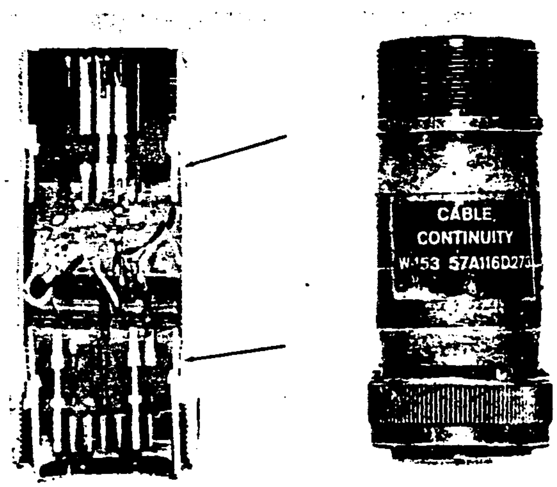


Figure 4-23
Cut- Away View of Magnaformed Cable Transition

When maintaining the shielding integrity of a connector pair (i.e., two interconnecting connectors), a good method to employ (see Figure 4-24) is to place spring contacts inside one portion of one connector so that positive contact is made along the circumference of the mating parts [9]. These contacts are extended so that the shell of the connector mates before the pins make contact on assembly of the connector and breaks after the pins on disassembly. A connector which meets these requirements is available under MIL-C-27599 and is the preferred type to be used in RF-proof designs.

The advantages gained using circumferential spring fingers over bayonet coupling is dramatically illustrated in Figure 4-25. In this case, the spring contacts were of silver-plated beryllium copper.

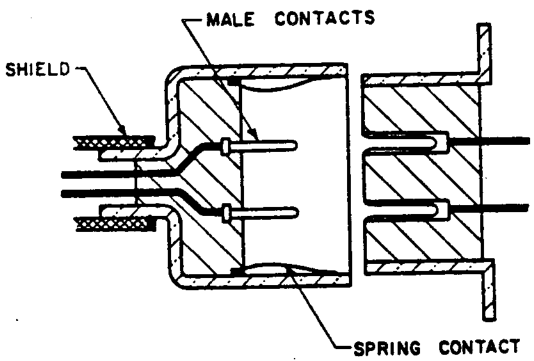
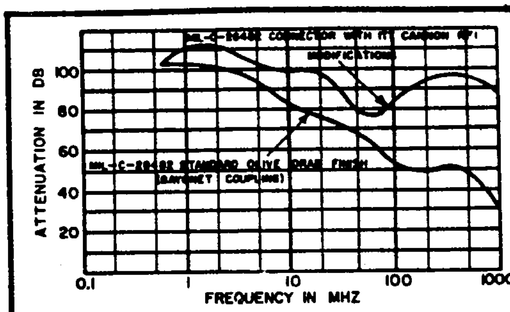


Figure 4-24
RF-Proof Connector



Comparison of the RFI shielding at the interface coupling of a MIL-C-26482 connector having standard olive drab finish with the same connector using peripheral spring contact fingers over cadmium plating (RFI type).

Figure 4-25
RFI Shielding at
Interface Coupling

Figure 4-26 illustrates the type of connector that should be used when a shielded cable assembly contains individual shielded wires[9]. The practice of pigtailing these shields and connecting them to one of the pins is not recommended. The individual shields should be connected to coaxial pins specifically adapted for this purpose, with the shields of the mating surfaces making contact before the pins.

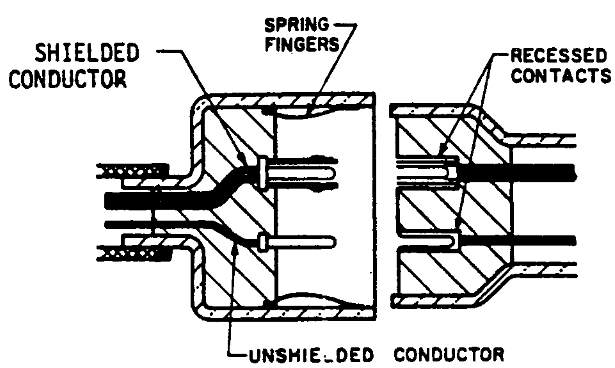


Figure 4-26 Connector for Shield Within a Shield

An alternative method of terminating individually shielded wires (if the shields can be tied together) is shown in Figure 4-27. A pick is used to strip the shield from each of the wires involved. The shields are then laid over a ground ring (Item 2 in Figure 4-27), and the entire assembly is secured by screwing items 1 and 4 together.

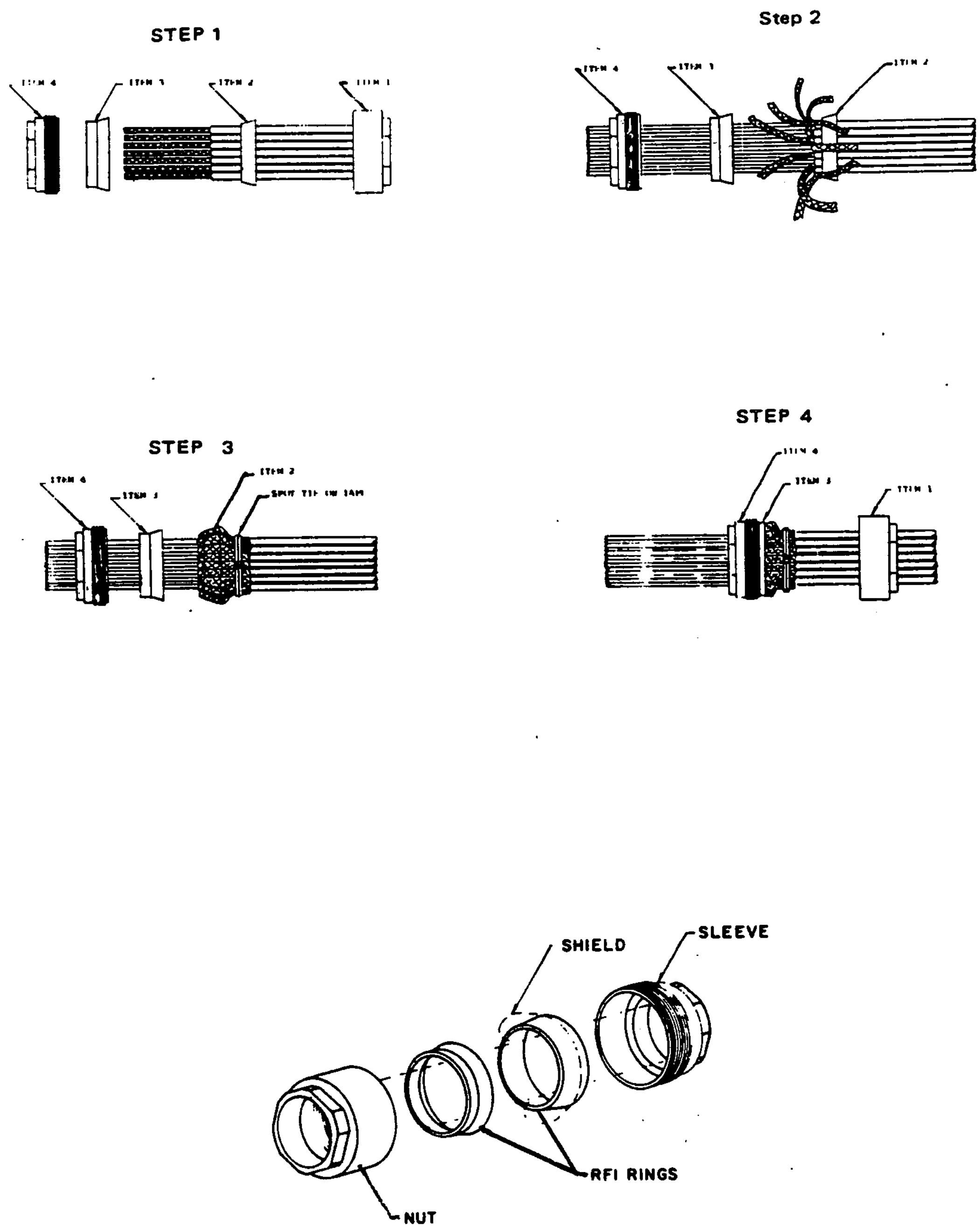


Figure 4-27 Shield Termination of Multiple Shielded Wires

RF arcing problems can occur in a shield that is adequate at audio frequencies. Induced RF currents can be conducted along cable shields and coupled to the system wiring at points of incorrect cable shield termination. RF potentials at the termination can be minimized by grounding the entire periphery to a low-impedance reference. Grounding may be required every 0.2λ based on the highest frequency at which currents may be present.

Two types of connectors with special features that improve shell-to-shell bonding in a high vibration environment are the Bendix "Tri-Start" series and the G & H Technology "Breech-Lok" series. Both are "scoop-proof" connectors that employ metric threads at the rear for attaching accessories. Scoop-proofing prevents shorts and pin damage, and metric threads provide more thread surface and thus a better bond between shells and accessories.

The Tri-Start[4] connectors are Series III connectors as defined in MIL-C-38999G and DoD-C-38999. They incorporate Series I high density inserts and contacts, and utilize a triple start Acme thread (one full turn to lock) with an anti-decoupling feature for plug-to-receptacle engagement. They also include metallic spring fingers for grounding. They were developed to meet rigid environmental and shock vibration requirements, but also provide improved EMI protection over the bayonette type Series I and II connectors in a high vibration environment.

The Breech-Lock[5] connectors are Series IV connectors as defined in MIL-C-38999G and DoD-C-38999. They incorporate Series I inserts and contacts, but utilize a solid-metal locking system (detent and lock in one-half turn) and internal drive threads for plug-to-receptacle engagement. They also include metallic spring fingers for grounding. They too were developed to meet rigid environmental and shock and vibration requirements, but also provide improved EMI protection over Series I and II connectors in a high vibration environment.

Two accessories that provide improved shield-to-connector bonding are the "IRIS" compression ring and "TAG" terminating and grounding ring. These devices provide additional options for terminating overall and individual shields respectively.

Figures 4-27a and 4-27b illustrate overall shield bonding by means of "IRIS" compression rings. The compression ring is part of the connector backshell or adapter and encircles the cable at the shield. The "IRIS" is a compressible metal coil spring, conductive elastomer, or other compressible conductive material that provides electrical contact between the overall shield and the backshell/adapter. The advantages of using "IRIS" compression rings are that the installation is simplified since the shield only has to be trimmed, and there is otherwise no disturbance to the shield. Also, the compression ring fits a range of cable sizes and makes a good bond to cables that are out of round.

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4. Tri-Start is a trademark of the Bendix Corporation.

5. Breech-Lock is a trademark of G&H Technology, Inc.

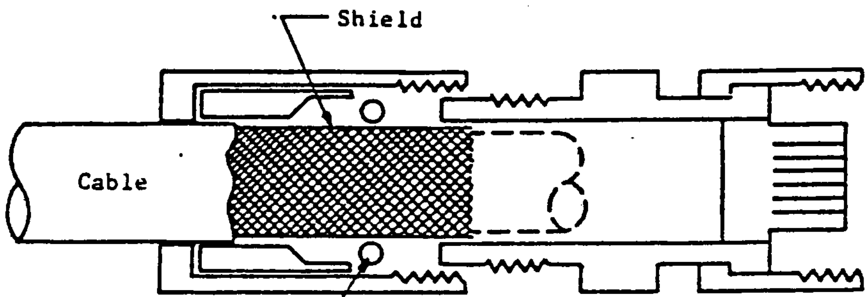

Figure 4-27a IRIS Coiled Spring Device

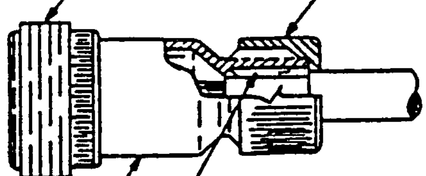

Figure 4-27b IRIS Conductive Elastomer Device

Figure 4-27c shows a TAG Ring[6] assembly that provides for terminating groups of individually shielded wires. The individual shields are placed in slots in the main body of the TAG Ring and are held there by a spring device built into the threaded clamp nut. The advantages of the TAG Ring are that installation can be performed without special tools, and that a range of cable sizes and shield thicknesses can be accommodated. Also, a TAG Ring and tapered cone can be installed in the same backshell/adapter to terminate individual shields and an overall shield when both are contained in the cable.

Another connector technology is the "filter connector". Although they were developed in the 1960's, filter connectors have only been available as production items since about 1980. Filter connectors are used to supplement shielded cable and shielding connectors when they cannot reduce incoming or outgoing interference to acceptable levels.

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6. TAG Ring is a registered trademark of Glenair, Inc.

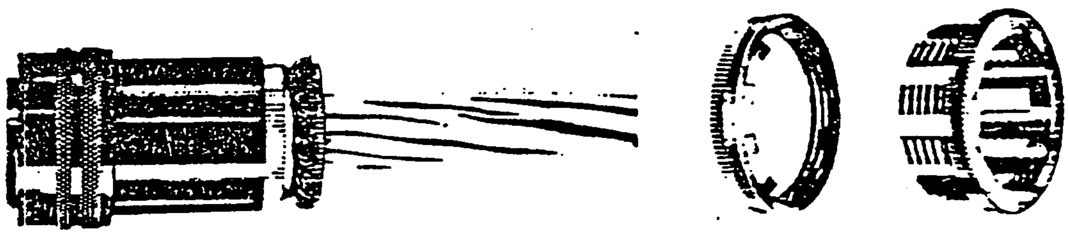


Figure 4-27c TAG Ring Assembly

A cross section of a filter connector is shown in Figure 4-27d. Individual filter elements are commonly capacitive (by-pass), L-Section, or π-section. Ferrites and capacitive ceramics provide the electrical properties. Filter element operation is shown in Figure 4-27e. Additional information on connectors having poor filter elements may be found in Section 7.2.1.7 of this manual.

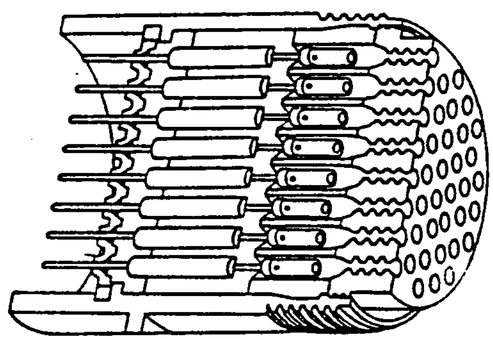


Figure 4-27d Filter Connector

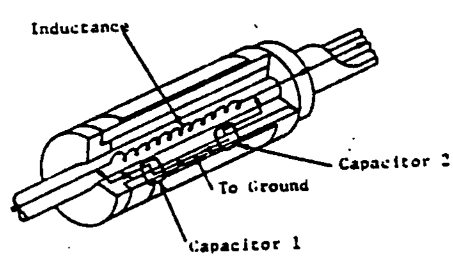


Figure 4-27e Filter Element

When repairing connectors, use should be made of the assembly instructions put out by the manufacturers. These instructions will also serve to disassemble the connector. During disassembly of the connector care should be taken not to damage the gaskets, rings, grommets, etc. and especially not the connector or backshell structure itself; this is because these are non-repairable items.

An example of such an assembly instruction is provided in Addendum 1 at the end of this chapter.

4.4.3 New Connector Technology

Some new connector technologies worth noting include conduit compatible backshells, EMI/RFI shrink boot adapters, and composite connectors.

The Breeze Illinois Inc. BI-SHELL 10/11 backshell can accommodate flexible metallic and non-metallic conduit to provide a hermetic/EMI seal. Both conduit and braided shield are inserted into the end of the backshell and terminated by brazing, or soldering the seam; the termination can also be accomplished by magna-forming. If environmental protection is required, an elastomer jacket can cover the underlying braided shield and flexible conduit slipping over the backshell collar. This jacket would then be terminated by bonding.

Glenair Inc. manufacturers EMI/RFI shrink boot adapters which feature a unique termination scheme. Shrink boots fit over the junction of the backshell and cable and "shrink" when heated to provide mechanical strain relief and some environmental protection. The unique termination scheme for these particular adapters is that the backshell collar and termination nut have sine wave threads which lock the braided shield between them when the shield is pulled over the collar threads and the termination nut is tightened (see Figure 4-28).

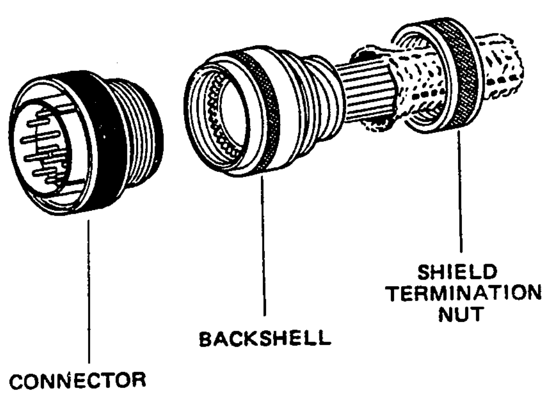


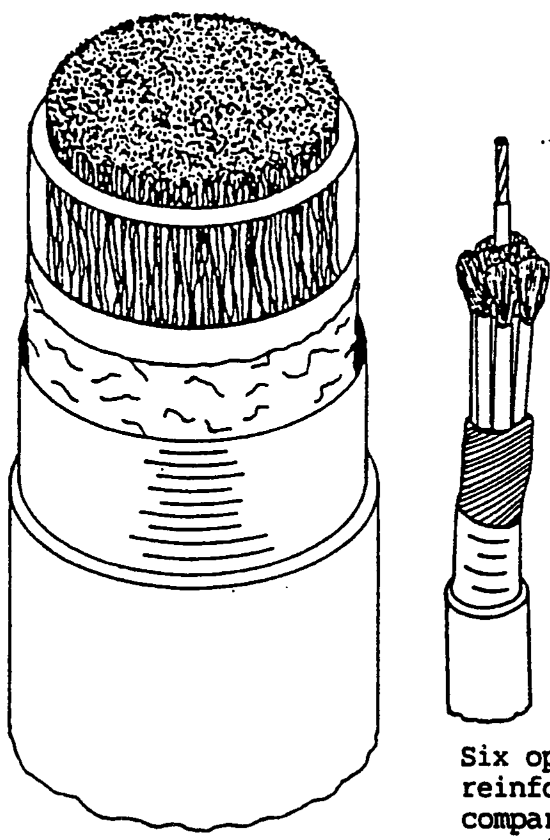
Figure 4-28 EMI/RFI Shrink Boot Adapter

Deutsch manufacturers the DG123 conductive coated composite connector. This corrosion resistant connector passes MIL-C-38999, class W, series III shielding effectiveness tests, and has a shell design based on the insert arrangement patterns of MIL-C-38999 and MIL-C-81511 connectors. This connector is capable of 1500 mating cycles without degradation and has features contributing to extended life cycle such as molded dielectric fingers in its inserts (for retention), and keys (for keyways) which are molded into the shell.

4.4.4 Fiber Optics

Fiber optics technology is gradually being incorporated into advanced weapons systems and in associated test equipment. It makes possible optical links between subsystems, by replacing hard-wired interfaces with glass fibers; and transmitting voice, video or data signals by light rather than by electricity. Infra-red lasers or light-emitting diodes serve as signal drivers.

The use of fiber optics offers weight and space advantages to the system designer. These benefits accrue largely because of the fiber's broadband transmission capabilities. Figure 4-29 illustrates these differences in a conventional and a fiber-optics cable having the same signal carrying capacity.



Six optical fibers in reinforced cable (right) compare in signal carrying capacity to the 900 pairs of conventional copper telephone wires (left).

Figure 4-29 Conventional and Fiber Optic Cables

From the standpoint of EMC, fiber optics cables are insensitive to electromagnetic field environments (although that is not necessarily true of the line termination devices) and therefore do not have to be shielded. They provide electrical isolation between source and load, eliminate potential problems of ground loop coupling, and allow real-time measurements over multiple channels. Additionally, they are totally safe for use in a fuel vapor or other explosive environment, and offer a secure communication means since they do not radiate detectable energy.

The advantages of broad bandwidth and better aerodynamic and EMC features must be weighed against other characteristics associated with fiber optics sub-system. These include the more complex and more costly line termination equipment, and the somewhat higher loss characteristics of the fibers themselves as compared with copper wire.

Typical fiber losses range from 3 to 15 dB/km when measured at 8000 Angstroms. Filament bandpass can be 25 - 400 MHz-km, depending on the type of fiber used, and the fiber diameter. The fibers can be clad and bundled in arrangements very similar to hard-wired cable.

4.5 OTHER DESIGN TECHNIQUES TO MAINTAIN SHIELDING EFFECTIVENESS

4.5.1 Seams Without Gaskets

Seams or openings in enclosure or compartment walls that are properly bonded will provide a low impedance to RF currents flowing across the seam. When good shielding characteristics are to be maintained, permanent mating surfaces of metallic members within an enclosure should be bonded together by welding, brazing, sweating, swaging, or other metal flow processes. To insure adequate and properly implemented bonding techniques, the following recommendations should be observed:

- All mating surfaces must be cleaned before bonding.

- All protective coatings having a conductivity less than that of the metals being bonded must be removed from the contact areas of the two mating surfaces before the bond connection is made.

- When protective coatings are necessary, they should be so designed that they can be easily removed from mating surfaces prior to bonding. Since the mating of bare metal to bare metal is essential for a satisfactory bond, a conflict may arise between the bonding and finish specifications. From the viewpoint of shielding effectiveness, it is preferable to remove the finish where compromising of the bonding effectiveness would occur.

- Certain protective metal platings such as cadmium, tin, or silver need not in general be removed. Similarly, low-impedance corrosion-resistant finishes suitable for aluminum alloys, such as alodine, iridite, oakite, turco and bonderrite, may be retained. Most other coatings, such as anodizing, are nonconductive and would destroy the concept of a bond offering a low impedance radio frequency path. See Figure 4-30 for shielding effectiveness data on selected surface finishes [11].

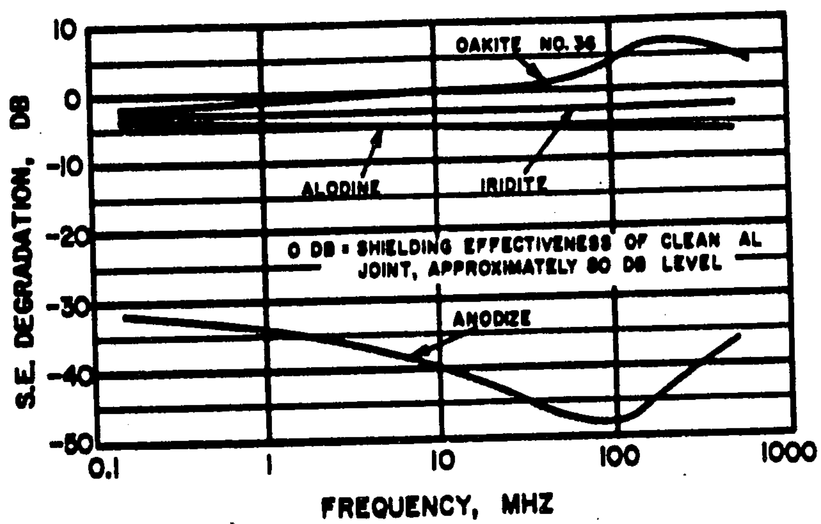


Figure 4-30 Shielding Effectiveness Degradation Caused by Surface Finishes on Aluminum

- o Mating surfaces should be bonded immediately after protective coatings are removed to avoid oxidation. Refinishing after bonding is acceptable from the standpoint of shielding effectiveness.

- o When two dissimilar metals must be bonded, metals that are close to one another in the electro-chemical series should be selected. Elements of the series are shown in Table 5-1 in the Bonding chapter of this design guide.

- o Soldering may be used to fill the resulting seam, but should not be employed to provide bond strength.

- o The most desirable bond is achieved through a continuous butt or lap weld. Spot-welding is less desirable because of the tendency for buckling, and the possibility of corrosion occurring between welds. Riveting or pinning is even less desirable because of the greater susceptibility of bond degradation with wear.

- o An overlap seam, accompanied by soldering or spot welding (see Figure 4-31) provides a relatively effective bond. Other types of crimped seams may be employed, so long as the crimping pressure is maintained.

There are often occasions when good temporary bonds must be obtained. Bolts, screws, or various types of clamp and slide fasteners have been used for this purpose.

Figure 4-31 Formation of Permanent Overlap Seam

The same general requirements of clean and intimate contact of mating surfaces, and minimized electrolytic (cathodic) effects apply to temporary bonds as well. Positive locking mechanisms that insure consistent contact pressure over an extended period of time should be used. Lockwashers should be employed that can "bite into" metal surfaces and fasteners and maintain a low bonding resistance. Bolts, nuts, screws, and washers that must be manufactured with material different from the surfaces to be bonded should be higher in the electromotive series than the surfaces themselves, so that any material migration erodes replaceable components.

A critical factor in temporary bonds (and in spot-welding permanent bonds as well) is the linear spacing of the fasteners or spot-welds [12]. Figure 4-32 provides an indication of the sensitivity of this parameter for a 1/2-inch aluminum lap joint. Data is taken at 200 MHz. The shielding effectiveness shown at 1-inch spacing is about 12 dB poorer than the identical configuration incorporating a 1/2-inch wide model mesh gasket; the effectiveness at 10-inch spacing is about 30 dB poorer than the same gasket. Use of conductive gaskets for this and other applications will be discussed shortly.

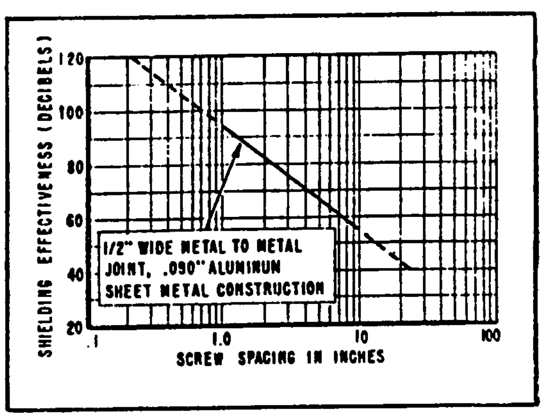


Figure 4-32 Influence of Screw Spacing

Similar techniques to those just described can be employed in connection with seams in magnetic materials [5]. Permanent seams can be butt or lap welded in an argon or helium atmosphere, recognizing that a final material heat treatment will be necessary. Temporary seams are usually screwed or bolted together. Figures 4-33 and 4-34 indicate the change in shielding effectiveness of an AMPB-65 seam at various frequencies as a function of screw spacing and lap joint width, respectively.

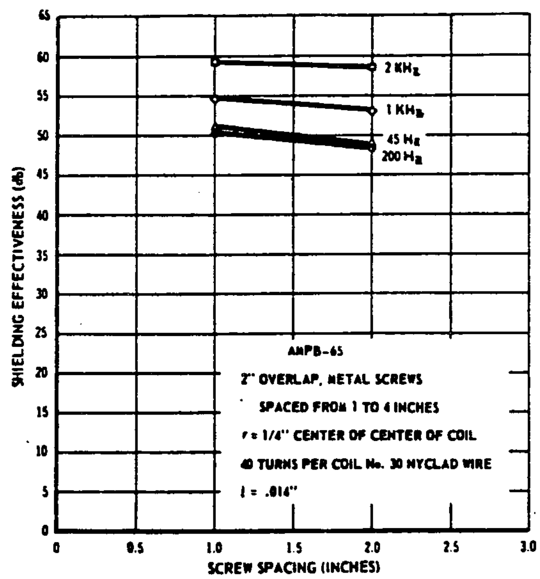


Figure 4-33 Shielding Effectiveness of AMPB-65 Overlap as a Function of Screw Spacing Along Two Rows, 1.5 Inches Apart

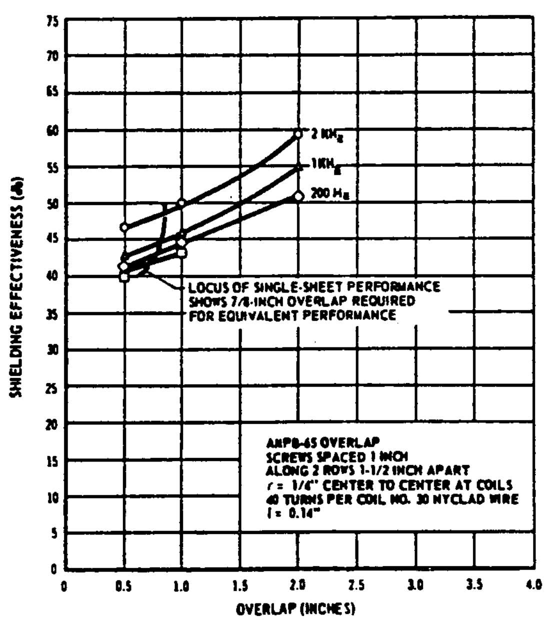


Figure 4-34 Shielding Effectiveness of AMPB-65 Joint as a Function of Overlap

Bonding techniques will be discussed in more detail in the next chapter of this design guide.

4.5.2 Seams and Gaskets

Considerable shielding improvement over direct metal-to-metal mating of shields used as temporary bonds can be obtained using flexible, resilient electrically conductive gaskets placed between shielding surfaces to be joined. Clean conductive mating surfaces and a good pressure contact are necessary.

The major material requirements for RF gaskets include compatibility with the mating surfaces, corrosion resistance, appropriate electrical properties (refer to Table 4-1), resilience (particularly when repeated compression and decompression of the gasket is expected), mechanical wear, and ability to form into the desired shape. On this basis, monel and silver-plated brass are generally the preferred materials, with aluminum surfaces. Beryllium-copper contact fingers are usually employed, with a variety of platings available, if desired. Mumetal and Permalloy have been used when magnetic shielding effectiveness is of concern.

For applications requiring moisture/pressure sealing as well as RF shielding, combination rubber-metal seals are available. These include metal mesh bonded to neoprene or silicone, aluminum screen impregnated with neoprene, oriented wires in silicone, conductive adhesives and sealants, and conductive rubber. The advantages and limitations of these, as well as non-sealing RF gaskets, are summarized in Table 4-11.

The problem in the marine environment of salt water/salt spray is that the fillers of conductive composite gaskets may corrode. For applications requiring corrosion resistance, special composite gaskets which avoid this may be used. One possible solution is to protect the conductive fillers in the matrix, with an organic coating [34]. If a metal mesh or knit is used, Raychem GELTEK (a silicon gel) may be used to impregnate the mesh or knit. This gel protects the conductive material from the corrosive elements (see Section 3.3.1 for more information on Raychem GELTEK and corrosion considerations).

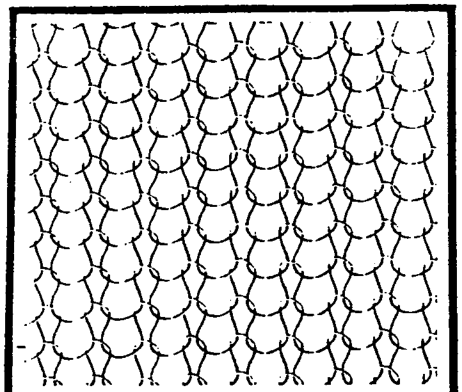
Similar to matching dissimilar metals to be bonded (see Section 4.5.1), the fillers of composite gaskets should be galvanically similar to the metal surfaces they join (i.e., close to each other in the electrochemical series). Elements of the series are shown in Table 5-1 in the next chapter (Bonding).

Knitted wire mesh gaskets can be manufactured in a variety of shapes and sizes. Figure 4-35 shows the structure of knitted wire mesh. Figures 4-36 and 4-37 illustrate representative cross sections and shapes, respectively. Gasket thickness is dependent on the unevenness of the joint to be sealed, the compressibility of the gasket, and the force available. Gasket shape depends on the particular application involved, as well as the space available, the manner in which the gasket is held in place, and the same parameters that influence gasket thickness [13].

Silver filled silicon rubber gaskets can be obtained in sheet, die-cut, molded or extruded form. The most popular, and most economical of these types, is the extrusion. Figure 4-38 shows typical extruded shapes and indicates recommended deflection limits for various shapes and sizes. Comments made above concerning thickness, shape and mounting methods for wire mesh gaskets also apply to conductive rubber gaskets.

TABLE 4-11 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, and in combination with neoprene and silicone.	Not available in sheet (Certain intricate shapes difficult to make). Must be 0.040 in. or thicker. Subject to compression set.
Brass or beryllium copper with punctured nail holes	Best break-thru of corrosion protection films.	Not truly resilient or generally reusable.
Oriented wires in rubber silicone	Combines fluid and rf seal Can be effective against corrosion films if ends of wires are sharp.	Might require wider or thicker size gasket for same effectiveness. Effectiveness reduces with mechanical use.
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 Rubber (carbon filled)	Combines fluid and conductive seal.	Provides moderate insertion loss.
Conductive Rubber (silver filled)	Combines fluid and RF seal. Excellent resilience with low compression set. Reuseable. Available in any shape or cross section.	Not as effective as metal in magnetic fields. May require salt spray environmental protection.
Contact fingers	Best suited for sliding contact.	Easily damaged. Few points of contact.



An enlarged view of knitted wire mesh. It is made up of many loop-shaped springs interlocked in a hinge-like manner; thus combining resilience, flexibility and cohesion.

Figure 4-35

Figure 4-36
Representative Gasket Cross Sections

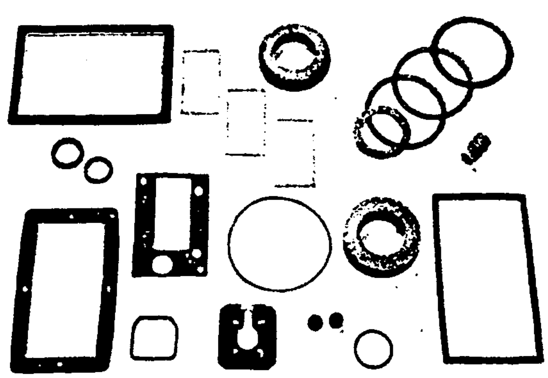
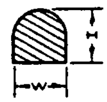
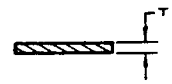


Figure 4-37 Representative Gasket Shapes

							
Deflection	W Dia.	Deflection	H	Deflection	T	Deflection	A
.007 – .018	.070	.006 – .012	.068	.001 – .002	.020	.025 – .080	.200
.010 – .026	.103	.008 – .016	.089	.001 – .003	.032	.030 – .125	.250
.013 – .031	.125	.012 – .024	.131	.003 – .006	.062	.075 – .250	.360
.014 – .035	.139	.014 – .029	.156	.003 – .009	.093		
		.016 – .032	.175				

from "EMI/RFI Gasket Design Manual," 1975, Chomerics, Inc.

Figure 4-38 Gasket Deflection Limits (in inches)

Shielding effectiveness of silver filled (or silver plated copper filled) silicones is especially high between 15 kHz and 10 GHz. Plane wave attenuation often improves with higher closure force, especially for die-cut gaskets. Best results are achieved with molded or extruded cross sections held in grooves.

Gaskets may be held in place by sidewall friction, by soldering, by adhesive, or by positioning in a slot or on a shoulder. Soldering must be controlled carefully to prevent its soaking into the gasket and destroying gasket resiliency. Adhesives (particularly non-conductive adhesives) should not be applied to gasket surfaces that mate for RF shielding purposes; auxiliary tabs should be used.

Typical gasket pressures for obtaining effective seals range from 5-100 psi. A usual pressure is 20 psi. Various ways in which wire mesh gaskets may be used are shown in Figure 4-39.

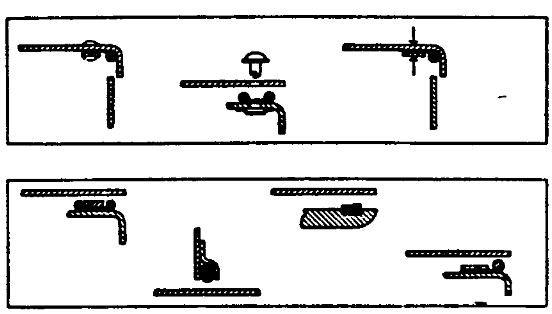


Figure 4-39 Representative Applications of Wire Mesh Gaskets

A number of examples of RF gaskets designed for GSE use can be cited. One such design is shown [14] in Figure 4-40. This gasket is used behind the front panel of the AN/AWM-54, Aircraft Firing Circuit Test Set. It consists of an outer Neoprene surface to waterproof the unit, plus an inner monel mesh.

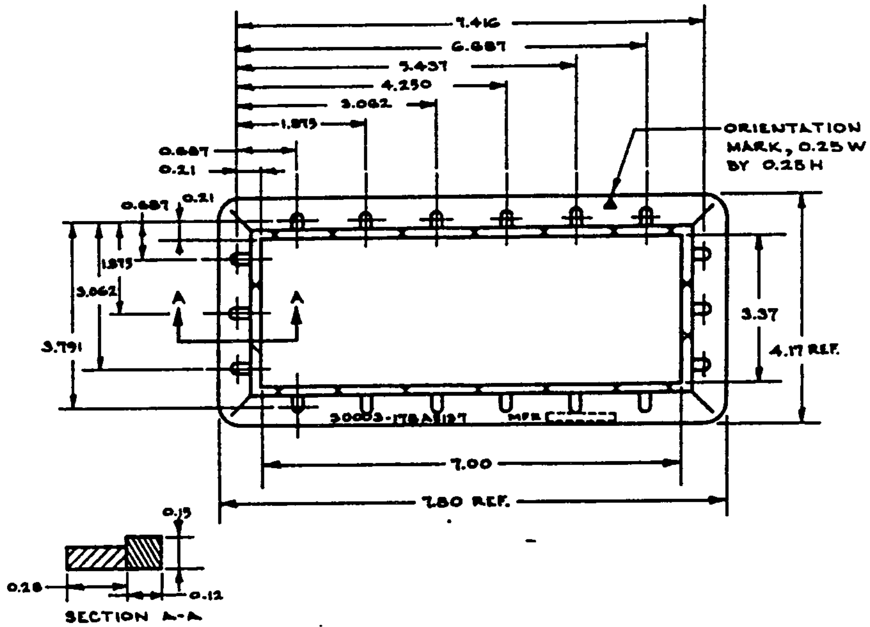


Figure 4-40 Front Panel Gasket for An/AWM-54

Five of the most often encountered RF gasket design problems are illustrated [15] in Figure 4-41, along with the means of overcoming the problems.

A practice presently in use is the dating of conductive and pressure gaskets, as well as other items that may suffer wear with continued use. The date is used during maintenance as an indicator of whether the gasket or other tagged item should be replaced.

4.5.3 Use of Waveguide Attenuators

The effects of seams, cracks, openings, holes and other breaks in a shield can often be studied by considering the opening as a waveguide. Alternatively, waveguide attenuators can be designed around control shafts, light receptacles and other openings to reduce interference leakage.

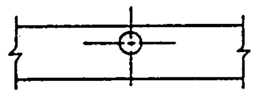
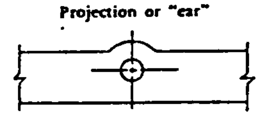
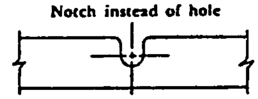
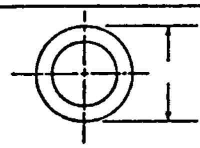
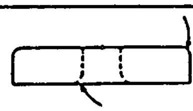
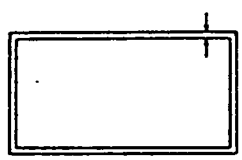
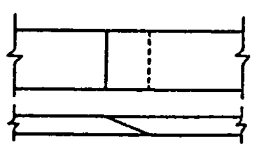
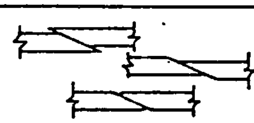
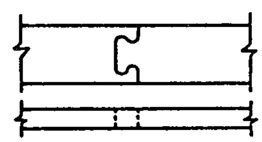
detail	why faulty	suggested remedy
 Bolt holes close to edge	Causes breakage in stripping and assembling	Projection or "ear"  Notch instead of hole 
 Metalworking tolerances applied to gasket thickness, diameters, length, width, etc.	Results in perfectly useable parts being rejected at incoming inspection. Requires time and correspondence to reach agreement on practical limits. Increases cost of parts and tooling. Delays deliveries	Most gasket materials are compressible. Many are affected by humidity changes. Try standard or commercial tolerances before concluding that special accuracy is required
 Transference of fillets, radii, etc., from mating metal parts to gasket	Unless part is molded, such features mean extra operations and higher cost	Most gasket stocks will conform to mating parts without preshaping. Be sure radii, chamfers, etc., are functional, not merely copied from metal members
 Thin walls, delicate cross section in relation to over-all size	High scrap loss: stretching or distortion in shipment or use. Restricts choice to high tensile strength materials	Have the gasket in mind during early design stages
 Large gaskets made in sections with beveled joints	 Extra operations to skive. Extra operations to glue. Difficult to obtain smooth, even joints without steps or transverse grooves	 Die-cut dovetail joint

Figure 4-41 Common Errors in Gasket Design

If the opening can be treated as a rectangular aperture, its cutoff wavelength, λ_c, is

$$\lambda_c = 2w \tag{4-20}$$

where

w = the long dimension of the opening.

Slot attenuation, in dB, can be computed from the relationship

$$A = 54.5\, d/\lambda_c \sqrt{\frac{1-\lambda_c}{\lambda}} \tag{4-21}$$

where

A = the attenuation per unit depth

d = slot depth

λ = interference wavelength

Similar relationships can be established for other opening configurations.

A useful rule to follow for circular holes is that, for 100 dB of attenuation, the length of the waveguide must be at least 3 times the diameter of the hole. The maximum permissible diameter (d_{max}) can be obtained by dividing the wavelength for the highest frequency under consideration (λ_{min}) by 3.4, where both diameter and wavelength are expressed in the same units [2].

$$d_{max} = \lambda_{min}/3.4 \qquad (4\text{-}22)$$

One example of the use of a waveguide over an opening to obtain attenuation may be found in the AN/AWM-54 Aircraft Firing Circuit Test Set. Figure 4-42 provides a side-view of the test set with its cover removed, and shows two circular guides above the GO/NO-GO indicator lights.

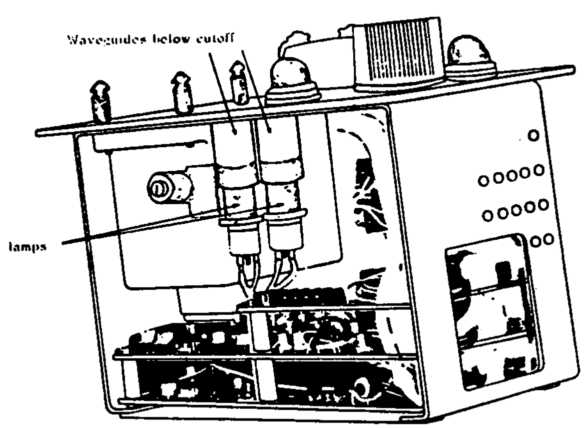


Figure 4-42 Application of Waveguide Below Cutoff Principle to AN/AWM-54 Aircraft Firing Circuit Test Set

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 (such as honeycomb) with openings that operate on the waveguide-below-cutoff principle. Honeycomb-type ventilation panels in place of screening:

a. Allow higher attenuation than can be obtained with mesh screening over a specified frequency range.

b. Allow more air to flow without pressure drop for the same diameter opening.

c. Cannot be damaged as easily as the mesh screen, and are therefore more reliable.

d. Are less subject to deterioration by oxidation and exposure.

All non-solid shielding materials, such as perforated metal, fine mesh copper screening, and metal honeycomb, present an impedance to air flow. Metal honeycomb is the best of these materials because it enables very high electric field attenuations to be obtained through the microwave band with negligible drops in air pressure. However, honeycomb has the disadvantages of occupying greater volume and costing more than screening or perforated metal. Also, it is often difficult to apply honeycomb paneling because flush mounting is required. Thus, screening and perforated sheet stock sometimes find application for purely physical design reasons, although honeycomb panels can achieve attenuations to 136 dB above 10 MHz [3].

4.5.4 Panel Openings

Panel openings that must accommodate control shafts may be shielded in one of several ways. A waveguide attenuator may be used around the panel opening as discussed under Section 4.5.3, so long as the shaft within the guide is non-conducting. Alternatively, the portion of the control that is behind the panel may be shielded to separate the control from the remainder of the equipment, and the control leads filtered.

A military requirement exists regarding RFI-shielded rotary-shaft seals (MIL-B-5423/16C). The requirement specifies a brass, nickel-plated nut with a knitted monel wire mesh insert to ground the control shaft. The nut is shown in Figure 4-43.

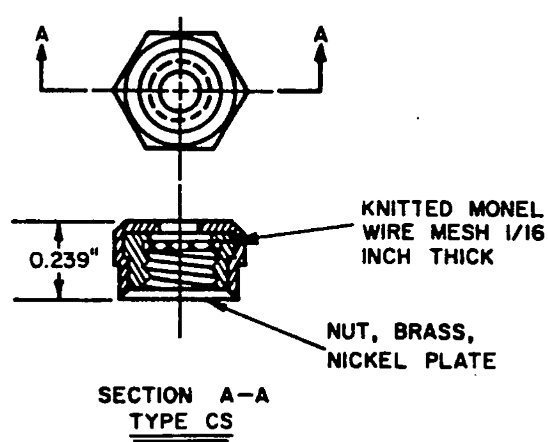


Figure 4-43 RFI-Shielded Rotary Shaft Seal

Fuseholders, phonejacks, panel connectors not in use, and other receptacles can be fitted with a metallic cap that provides an electrically continuous cover and maintains case integrity. An example of this type of cap for fuseholder applications is shown in Figure 4-44, as specified by MIL-I-49207/36A. The two conducting gaskets are knitted monel impregnated with silicone rubber.

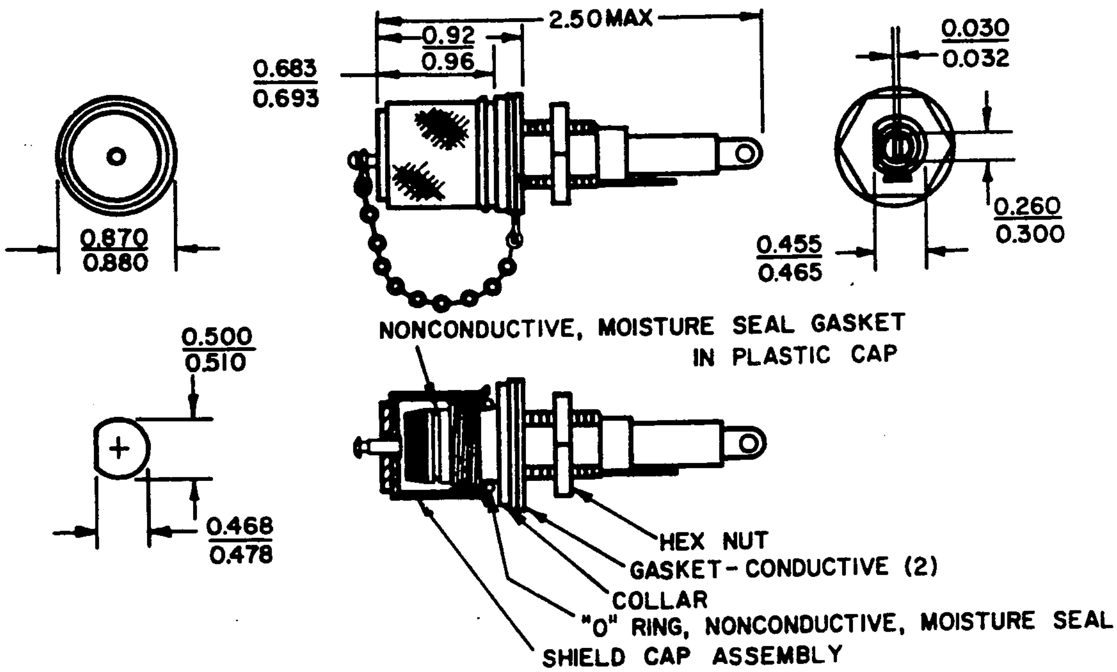


Figure 4-44 RFI-Shielded Extractor Post Type Fuseholder

Pushbutton switches are recommended to be shielded using boots having a mesh gasket insert (MIL-B-5423/7A). This approach (see Figure 4-45) appears to be more effective than the alternative of a conductive rubber boot.

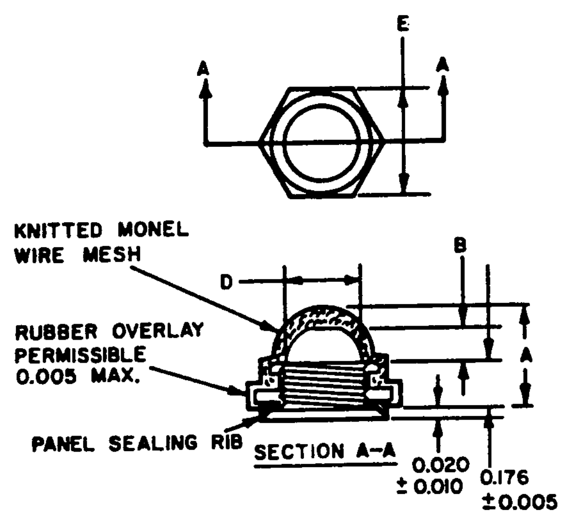


Figure 4-45 RFI-Shielded Pushbutton Switch

4.5.5 Required Visual Openings

Often it is necessary to provide RF shielding over pilot-light bulbs, digital display faces, meter faces, strip chart recorder outputs, oscilloscope faces, or similar devices that must be observed by the equipment user. The alternatives available include:

- o use of a waveguide attenuator
- o use of screening material
- o providing a shield behind the assembly of concern, and filtering all leads to the assembly
- o use of conducting glass

A waveguide attenuator is a practical approach for RF shielding of lamps, and an example has already been provided in Figure 4-38. This technique has the advantage of not introducing light transmission loss. However, it is not particularly suitable for most meter openings or larger apertures, because of the space requirements involved.

Use of screens over meter faces and other large apertures has often been employed for shielding purpose. A typical screen introduces a minimum of 15-20% optical loss, and can create difficulties in reading meters. If the device being shielded has a scale (such as an oscilloscope reticule), bothersome zoning patterns can result. However, these potential deficiencies are counterbalanced by good shielding efficiencies at a fairly low cost.

Figure 4-46 illustrates one method of mounting such screens, when they are not incorporated directly into the device to be shielded [1], and Figure 4-47 indicates the shielding effectiveness provided by commercially available screen assemblies [16]. In the latter case, data on either .0045 or .002 inch diameter monel, with either 12x12 or 6x4 openings per inch are presented. The screen may be embedded in the center of a single pane of acrylic, or incorporated into a glass sandwich. It should be tinned or otherwise bonded around its periphery to achieve good mating to the metallic plate. A variety of screen materials are available. Slight variations in the dimensions of the mesh openings are intentionally made to reduce the meter-reading and zoning problems.

A screened meter face on a prototype GSE test is shown in Figure 4-48.

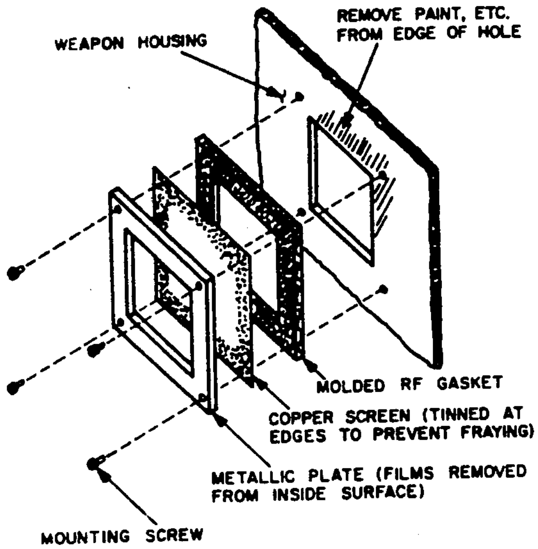


Figure 4-46
Method of Mounting Wire Screen Over A Large Aperture

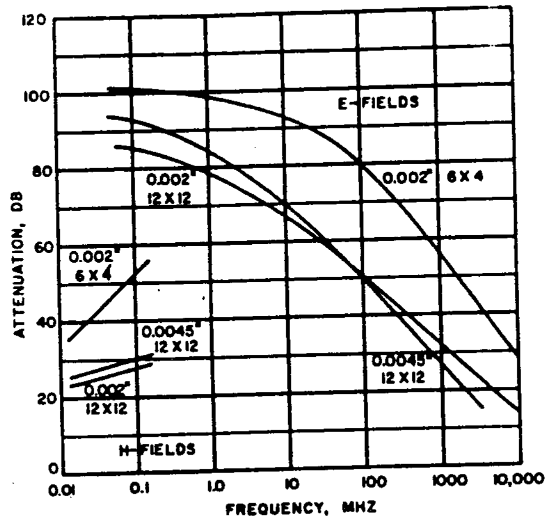


Figure 4-47
Shielding Effectiveness of Meter Screening Assembly

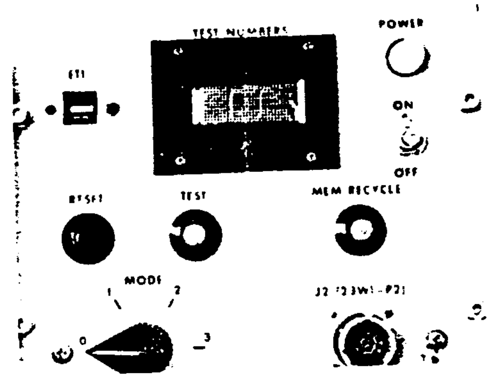


Figure 4-48 GSE With Shielded Digital Meter Face

Two approaches can be employed when shielding behind an assembly and then filtering all leads passing through the shield. These approaches are shown in Figure 4-49a and b, in the case of a meter. One method, when the meter involved is essentially an off-the-shelf item, is to build a supplementary enclosure and to pass the meter leads through feed-through capacitors or other appropriate filters to eliminate interference that may have been picked up through the meter face. The other method is to procure a meter whose back can be used in place of a supplementary shield, and one which incorporates the necessary lead filtering. In either case, it is assumed that external fields will not cause adverse effects to the operation of the meter itself.

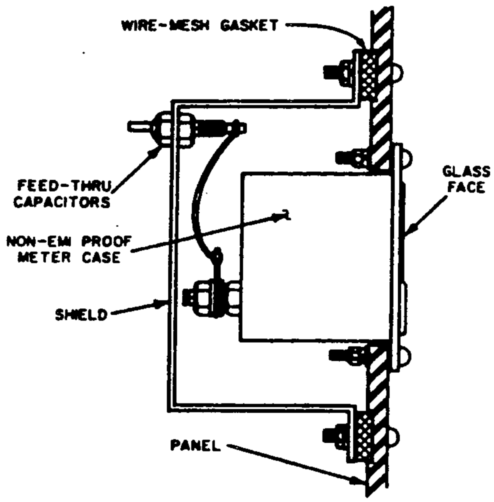


Figure 4-49a
Meter Shielding Techniques

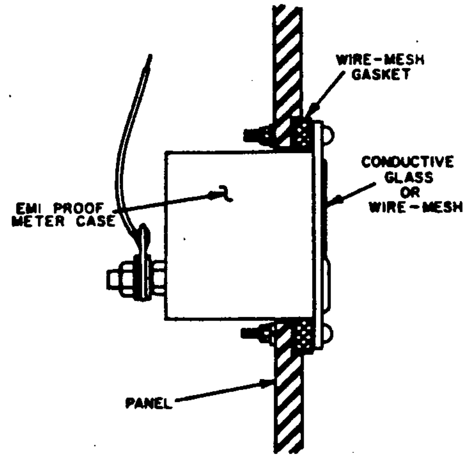


Figure 4-49b
Meter Shielding Techniques

Glass coated with conducting material such as silver can provide shielding across viewing surfaces with some loss in light transmission. Conductive glass is commercially available from a number of glass manufacturers.

Figure 4-50 provides shielding effective data on 50 and 200 ohms per square[7] silver-impregnated glass against electric arc discharges [17]. Figure 4-51 indicates similar characteristics against plane waves in the frequency range of 0.25 to 350 MHz [17]. The light transmission characteristics of this type of glass as a function of surface resistance [18] is presented in Figure 4-52. For effective shielding, good contact to the conducting surface of the glass must be maintained around its periphery.

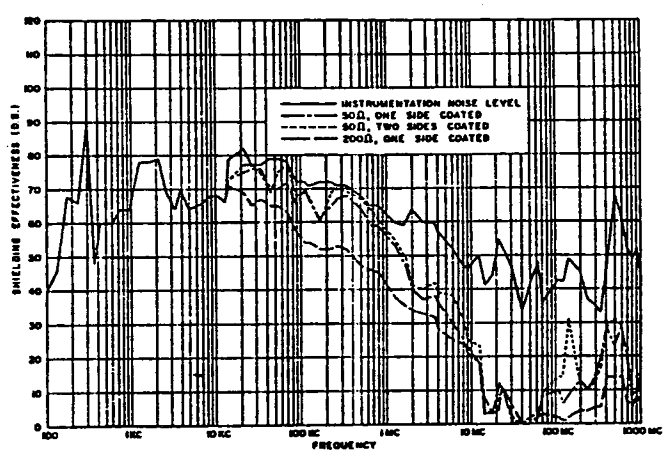


Figure 4-50 Shielding Effectiveness of Conductive Glass to High Impedance Waves

7. The resistance across a square section of uniformly conductive material is independent of the size of the section - hence the unusual dimension of "ohms per square".

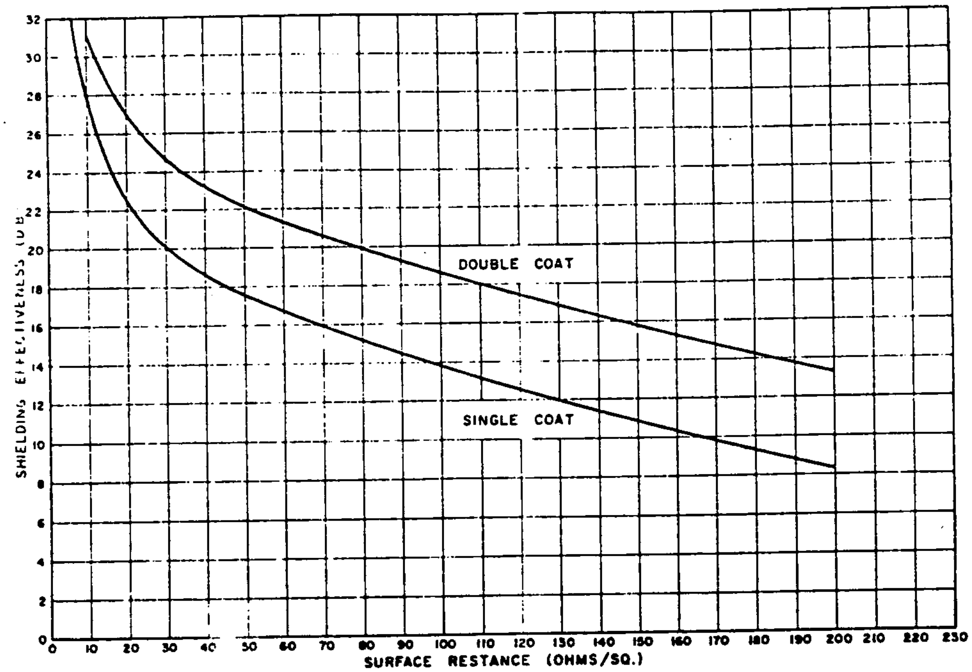


Figure 4-51 Shielding Effectiveness of Conductive Glass to Plane Wave Propagation

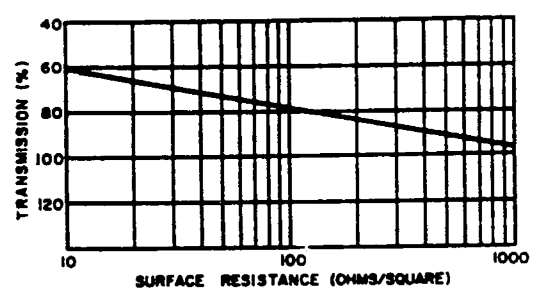


Figure 4-52 Light Transmission vs Surface Resistance for Transparent Conductive Glass

4.6 SHIELDING TESTS

4.6.1 General

The variety of test methods available for evaluating shielding effectiveness are due, at least in part, to the many different factors that can affect material shielding capabilities. These factors include the configuration of the shield (is it a sheet of material, or is it a box?), the frequency range of concern, whether or not the impinging wave is planar, the wave impedance, and others. This section will discuss the most frequently employed and generally applicable shielding effectiveness tests. While they are not all appropriate to evaluating a particular GSE design, they provide a set of established procedures from which designers can select applicable techniques. These tests include:

- Low Impedance Magnetic Field Testing Using Small Loops
- Low Impedance Magnetic Field Testing Using a Helmholtz Coil
- High Impedance Electric Field Testing Using Rod Antennas
- High Impedance Electric Field Testing Using a Parallel Line Radiator
- Plane Wave Testing Using Antennas
- Plane Wave Testing Using a Parallel Plate Transmission Line
- MIL-STD-1377 Testing

A number of the above tests are very similar to tests designed to measure equipment and system EMC in accordance with MIL-STD-462 [19]. They also are similar to tests performed to evaluate EM effectiveness of shielded enclosures used for testing purpose in accordance with MIL-STD-284 [20]. The GSE design engineer who is concerned with the measurement of shielding properties should familiarize himself with both of these standards.

The MIL-STD-1377 [21] tests to be discussed represent recommended procedures for evaluating the shielding (and filtering) effectiveness of Navy weapons systems. The specification contains a unique approach to shielding measurements, and its cable effectiveness evaluation methods are good illustrations of how cable and connector performance tests can be performed.

It should be pointed out that a high degree of measurement accuracy cannot generally be expected from shielding tests. Typically, wave impedances are not established when the tests are performed, antenna correction factors used for calibration purposes are based on plane-wave assumptions even though the test condition may not warrant this assumption, the degree of radiated field distortion by proximal structures is not known, and other factors limit the accuracy of the measurement. However, the tests can be expected to provide guidance to the GSE designer on the shielding design approaches he should take and the general effectiveness to be expected of those approaches.

Tabulated procedures are provided for each shielding test. These procedures are included for descriptive purposes only, and are not meant to be rigid rules for conducting a particular test. The reader is referred to the cited references for further details.

4.6.2 Low Impedance Magnetic Field Testing Using Small Loops[20],[22]

4.6.2.1 Objective

This test is designed to indicate the GSE enclosure or compartment effectiveness in reducing the intensity of predominantly magnetic field radiation impinging upon it. It employs two small loop antennas, and evaluates loop coupling with and without an intervening shield.

MIL-STD-285 incorporates a similar magnetic field small loop measurement procedure to evaluate the shielding effectiveness of shielded enclosures used for electronic testing purposes. The specific parameters of this procedure will be noted where applicable.

4.6.2.2 Test Setup

In this test, a pair of identical small loop antennas are used, one on one side of the shield and one on the other, spaced equidistant from the shield. If an enclosure is being tested, the usual practice is to have the test signal source within the enclosure and the receiving loop and detector outside the enclosure.

Figures 4-53 and 4-54 show the two basic loop orientations. In Figure 4-53 the loops are coaxial, that is, both loops are normal to a common loop axis. In Figure 4-54 the loops are coplanar, that is, the loop surfaces lie on the same plane. Tests using at least these two orientations should be employed, but other orientations that may result in a lower effectiveness figure should not be ignored.

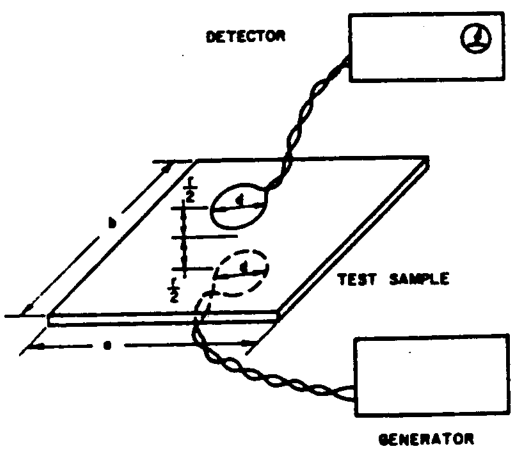


Figure 4-53
Small Loop to Loop Setup
(Coaxial Arrangement)
$r \ll a,b$

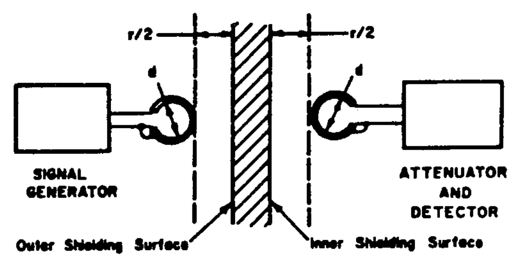


Figure 4-54
Small Loop to Loop Setup
(Coplanar Arrangement)

Both the loop diameters and the loop separations should be significantly less than the shortest dimension of the box, container, or enclosure being tested. Since this will result in only a small section of the shield being illuminated at one time, it will be necessary to move the loop over the entire surface of the shield to establish the effectiveness of the shield.

The frequency range over which this test can be performed is a function of the level of shielding effectiveness that must be measured (measurement system dynamic range), the sensitivity of the test equipment, the available power to drive the test transmitting loop, and the loop-to-shield separations. The limiting factors are usually the areas of the loops and the number of turns in the loops, since these establish the self-resonance frequency of the loop.

If the detection system is sensitive (0.1 μV input sensitivity) and the available test signal power is large (+10 dBm or greater), then small diameter loops with one, or at most a few, turns will be necessary. To compensate for low test signal level and detector insensitivity, or for small test sample size, smaller loops with more turns may have to be used. Loop-to-shield separation should not be closer than the loop diameter. Test personnel will have to establish loop parameters on a case by case basis, and wind test loops accordingly.

The signal source may be a signal generator or power oscillator of sufficient CW, MCW, or Pulsed CW output. When the radiating loop is connected to the output of a signal generator or power oscillator, a turning capacitor may be connected in series with the loop to obtain resonance at the test frequency used.

The detector can be any receiver or field strength meter having a low impedance input, or a high input impedance detector with an appropriate matching device. Tests should be performed in a shielded enclosure (screen room) to obtain maximum dynamic test range, and not be limited by the ambient environmental level at the test site.

If the signal generator has an accurate output attenuator, it may be used instead of the attenuator indicated in Figure 4-54.

The small loop-to-loop test setup specified in MIL-STD-285 is as shown in Figure 4-54, with the following parameter values employed:

Loop diameters (d) = 12 inches

Loop-to-shield separations (r/2) = 12 inches

Loops - One turn of No. 6 AWG Copper Wire

The test setup in this specification is intended to provide a minimum of 70 dB measurement range.

4.6.2.3 Test Procedure

To perform this test, set up the loops as described previously, and proceed as follows:

- o At each test frequency, increase the generator output level until the detector exhibits at least a 3 dB increase (and preferably a 10 dB increase) in output over noise.

- o Adjust the loop orientations to maximize the detector output.

- o Record the input signal amplitude (E_2).

- o Either remove the shield from between the loops, or move the loop antennas and set them up nearby in the same configuration that resulted in the previous maximum output condition.

- o Decrease the generator output level until the previous detector level is obtained. It may be necessary to add additional attenuation at the receiver input to sufficiently reduce the level.

- Record the input signal amplitude (E_1) and the amount of added attenuation.
- Compute the shielding effectiveness from the relationship

$$S(dB) = 20 \log [E_1 \ 10^{-A/10} / E_2] \qquad (4\text{-}23)$$

where A = added attenuation, in dB.

- Position the loops over various portions of the enclosure, particularly those incorporating seams, panels, and other discontinuities, and repeat the test.

4.6.2.4 Advantages/Disadvantages

- The test equipment is easily assembled and fabricated, but the test itself may be very time-consuming for large enclosures.
- The antenna sizes are convenient to use at low frequencies.
- A strong localized field is available.
- By controlling the distance between the loops and the shield, the effectiveness of shielding structures can be evaluated at different impedance levels of low-impedance fields.
- The method lends itself to measuring enclosures of widely varying size.
- Voltages induced in the small loops due to reflections from the enclosure are minimized.
- The method is applicable whether the signal source or the detector is inside the enclosure, thus providing considerable flexibility in the test setup.
- Since the radiating loop illuminates only a small section of the shield, a probing procedure must be employed.
- Dynamic range may be poor or, alternatively, the frequency range of the test may be limited.
- The test is fairly sensitive to small changes in loop spacing, and the test results are difficult to repeat.
- Loop diameter must be kept less than 0.01λ. Thus, a 12 inch receiving loop will provide accurate measurements at frequencies less than 10 MHz.

4.6.3 Low Impedance Magnetic Field Testing using a Helmholtz Coil [22],[23]

4.6.3.1 Objective

This test is also designed to measure enclosure magnetic shielding effectiveness. It uses a large radiating coil so that it bathes the entire enclosure in a magnetic field.

4.6.3.2 Test Setup

In this test, the test sample is placed inside a Helmholtz coil (large loop), with a small detection loop inside the test enclosure. The small loop is usually located near the center of the enclosure, but may also be used as a probe for leakage energy. The use of a Helmholtz coil enables a large portion of the enclosure to be illuminated at one time. Various orientations of the sample relative to the loop should be tried.

The frequency range is limited by the test sample size, which affects the size of the Helmholtz coil. Increasing coil size increases its inductance, reduces its self-resonant frequency, and decreases the frequency range over which its magnetic field intensity remains constant. The coil diameter should be at least two and preferably three times the longest test sample dimension. The upper frequency limit is typically 100-500 kHz.

The test arrangement is shown in Figure 4-55. On occasion, two coaxial coils instead of one have been used, with the test sample placed between the coils. The two coils are for the purpose of providing a more uniform field over the volume of the sample [23].

The leads from the detector loop should be twisted, shielded cable. When possible, the shield should be grounded circumferentially where it leaves the enclosure under test, and should be single-point grounded at the detector.

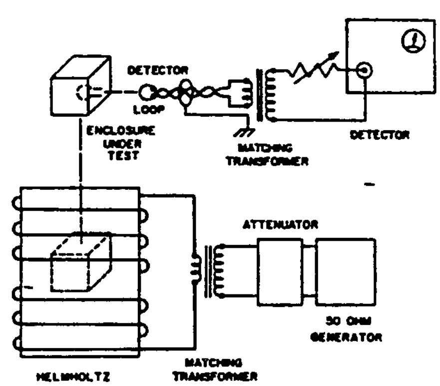


Figure 4-55 Helmholtz Coil Test Arrangement

4.6.3.3 Test Procedure

The test procedure using the Helmholtz coil is basically the same as the one employed for the two small loops method. However, instead of the two-loop probing operation, different orientations of the enclosure in the field of the Helmholtz coil are tested. Also, as indicated previously, some probing with the detector probe may be employed.

4.6.3.4 Advantages/Disadvantages

- The field generated by the Helmholtz coil can be made very uniform over a wide area or within a specified volume.
- The intensity of the generated field is reasonably strong; hence dynamic range is good.
- It provides a convenient way, to detect joint and seam leakage, however it does not locate the particular leakage source.
- Measurement repeatability is high.
- The test setup is quite complex and bulky.
- The maximum frequency at which this procedure can be used is considerably less than that for the two-loop method.
- Uniformity of the field as frequency increases can only be maintained by a reduction of the Helmholtz coil dimensions, thus limiting the sample size.
- Leaky joints of the enclosure that are parallel to the direction of the current flow in the large loop may not be detectable. Thus, to detect the effect of all possible seams and discontinuities, the enclosure must be measured for different orientations in the field.

4.6.4 <u>High Impedance Electric Field Testing Using Rod Antenna</u> [20],[22]

4.6.4.1 Objective

This test is intended to evaluate whether or not a GSE enclosure, or an internal compartment, affords adequate protection for the enclosed equipment against high impedance electric field energy. This is accomplished using rod antennas located on both sides of the shield. The frequency range covered by this test is typically 15 kHz to 30 MHz.

4.6.4.2 Test Setup

Figure 4-56 indicates a representative test setup for measuring high impedance wave affects. Rod antennas with passive impedance matching transformers are employed. The antennas must be matched to the respective terminating devices. Antennas such as the VR-105 and VA-105 rod antennas supplied with the NF-105 Field Intensity Meter can be used directly, since these antennas incorporate matching step-up auto-transformers.

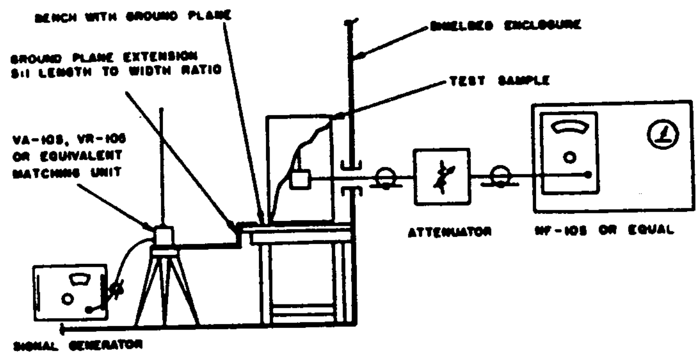


Figure 4-56 Test Equipment Setup for Performing High Impedance Electric Field Shielding Effectiveness Tests

Typically, the radiating rod antenna is 41 inches long. The detector antenna length is variable, depending on the particular test application. MIL-STD-285 specifies a detector antenna length of 41 inches, while other tests have used a 10 inch antenna. ANTENNA-TO-SHIELD SPACING IS ALSO dependent on the test involved; the MIL-STD-285 specification on spacing is 12 inches which limits the tests to frequencies below about 15 MHz.

The signal source should be capable of delivering at least +10 dBm into a 50 ohm load. Attenuators may be incorporated into the test setup ahead of the detector input as required.

The detector system is located external to the shielded enclosure containing the unit being tested, to prevent the radiated field from interfering with that system. Connection is made to the detector with coaxial cable that passes through a waveguide-below-cutoff filter in the shielded enclosure wall.

4.6.4.3 Test Procedure

- o Adjust the test signal generator frequency to the desired value, set the generator output level to its maximum value, and tune the detector for a received signal indication.

- o Adjust the detection receiver in gain and input attenuation to give mid-range output.

- o Orient the shield around the detection antenna until the highest output indication is obtained from the detection system. Also, position the source antenna anywhere around the enclosure under test, in any orientation so long as it remains parallel to the enclosure wall, for highest output indication. Continue to adjust the attenuator and gain control settings to maintain a mid-range output reading. Record the final attenuator and gain control settings.

- o Remove the shield under test from around the detection antenna, and place the test shield sample outside the test area.

- Decrease the signal generator output level until the detection system has the same output indication as with the shield in place around the detection antenna. It may also be necessary to increase the attenuator setting. The shielding effectiveness is then computed in accordance with Equation (4-23).

4.6.4.4 Advantages/Disadvantages

- The wave impedance generated by this procedure is high impedance so long as the general separation conditions of Figure 2-6 are maintained.

- The equipment used is readily available and the test setup is straightforward.

- The method is flexible, and allows measurement of different sample sizes.

- The test field can be generated inside or outside the enclosure under test, depending on sample size and screenroom availability.

- The field generated is non-uniform and thus has the same uncertain effects as a non-uniform field generated by a small loop.

- The reduction of measured data is usually based on plane wave theory. High-impedance field measurements may not be directly verifiable through application of plane wave considerations.

- Monopole antennas are susceptible to reflections from surrounding object, observers, etc., to the extent that the accuracy of the results becomes largely a function of the experience and capability of the testing personnel.

4.6.5 High Impedance Electric Field Testing Using a Parallel Line Radiator [24]

4.6.5.1 Objective

This test is also for the purpose of testing enclosure shielding effectiveness against high-impedance fields. It is identical to the procedure of Section 4.6.4 using rod antennas, except that the source antenna configuration is now a parallel line radiating device.

4.6.5.2 The Parallel Line Radiator

The Parallel Line Radiator is merely a parallel-wire line, with the high-impedance wave being the fringing field from this line. Such a radiator is shown end-on in Figure 4-57.

There are many ways in which the line itself can be constructed; one such configuration is also shown in Figure 4-57. The length of the line (if an end-fed line is used), should be $\lambda/8$ at the highest test frequency. The line separation is not critical, but a separation of 0.5 to 1 meter is suggested to get a fairly planar field of high impedance at a 1 meter distance to the test sample. Line impedance is computed from the relationship:

$$Z_1 \text{(ohms)} = 276 \log (2\ S/d) \text{ for } S/d >> 1 \qquad (4\text{-}24)$$

where

S = line separation

d = wire diameter, in same units as S

Using this equation, lines can be designed to work into other impedances. Lines of low impedance may not be able to provide adequate electric field levels, while lines of high impedance require a matching transformer having a higher turns ratio. The latter may reduce the maximum frequency capability of the line.

Courtesy of DON WHITE CONSULTANTS, INC.

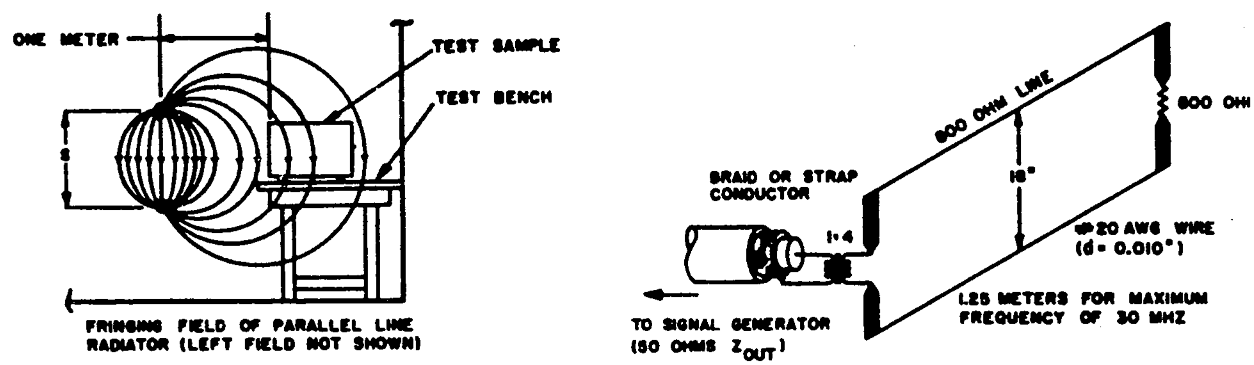


Figure 4-57 Parallel Line Radiator for Generating High Impedance Fields

4.6.5.3 Test Setup and Procedure

The test setup and procedure is the same for the parallel line radiator as for the rod antennas (Section 4.6.4).

4.6.5.4 Advantages/Disadvantages

- The field impedance using this method will be less than that achieved using the radiating rod antenna.
- The parallel line does not require a ground plane reference.
- The line is easily set up, but requires somewhat more space than the radiating rod antenna.

4.6.6 Plane Wave Testing Using Antennas [20],[22]

4.6.6.1 Objective

This test is designed to subject the enclosure under test to plane waves over the frequency range from 30 MHz to 35 GHz. The test is indicated as a "plane wave" test because, for the majority of the frequency range covered, the test distances are such that the wave characteristic approaches an impedance of 120 π ohms. The test uses a variety of radiating antennas, depending on frequency range. The 30-200 MHz range is generally covered with a broadband bi-conical antenna; while the range from 200 MHz to 12 GHz is covered with circular log-periodic conical antennas, (200-1000 MHz and 1-12 GHz). Horn antennas may also be used.

Because of test sample size limitations, it probably will not be practical to employ a detector antenna identical to that of the radiating antenna. A short dipole or monopole antenna is often used as the receiving antenna, located inside the test sample enclosure. In some cases, a horn antenna might also be used for receiving purposes.

4.6.6.2 Test Setup

The following equipment is required for this test:

- Signal generators covering the frequency range from 30 MHz to 35 GHz.
- A tuned voltmeter or detector/receiver covering 30 MHz to 35 GHz.
- Radiating antennas covering the frequency range of interest. The VSWR of these antennas should be less than 3:1 at all test frequencies.
- A shielded enclosure, preferably anechoic (non-reflective), in which to perform the test may be desirable.
- A short dipole or monopole having elements extending to a total distance less than the shortest inside dimension of the enclosure under test.

Test equipment should be arranged as shown in Figure 4-56. The source antenna should be located sufficiently far from the test sample so that far-field conditions apply. A similar test defined by MIL-STD-285 designates that the minimum separation shall be 6 feet.

4.6.6.3 Test Procedure

The procedure to be followed for this test is essentially the same as for the high-impedance rod antenna test (Section 4.6.4).

4.6.6.4 Advantages/Disadvantages

- The indicated antenna types are easily obtained or simple to construct, and generally have broadband characteristics.

- Conical spiral antennas, being circular polarized devices, are insensitive to GSE enclosure seam geometry.

- Log-periodic and horn antennas have relatively high gain, and therefore can provide a very high level of electric field and a subsequent wide dynamic range measurement capability.

4.6.7 Plane Wave Testing Using a Parallel Plate Transmission Line [24]

4.6.7.1 Objective

This test is designed to subject a GSE enclosure to a plane wave over a fairly wide frequency range, for the purpose of determining enclosure shielding effectiveness. The test uses a parallel plate transmission line for forming the field and controlling wave impedance. The field generated is planar. Parallel plate lines are normally used to 30 MHz [19], but can be designed to measure to 1 GHz if small test samples are involved.

4.6.7.2 Test Setup

The parallel Plate Transmission Line is essentially a strip line consisting of two rectangular sheets of conductive material, of equal length but one often wider than the other, and separated from each other. The sheets are driven and terminated at their respective narrow ends. A TEM mode field is created between the two sheets, as illustrated in the end-view of Figure 4-58. The test sample is placed between the sheets, where the electric field lines are essentially normal to the two surfaces.

Courtesy of DON WHITE CONSULTANTS, INC.

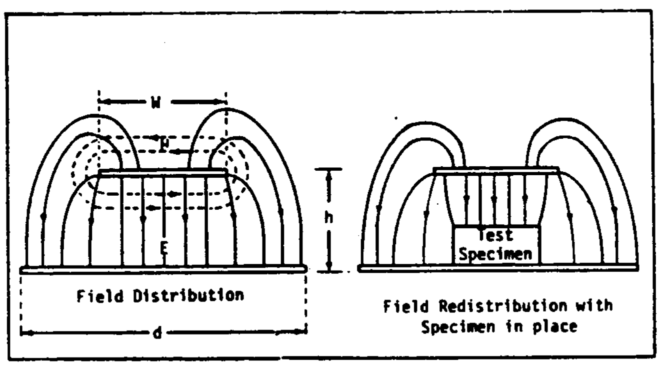


Figure 4-58 TEM Field Inside of Strip Line

The major advantage of the parallel plate line is that it is capable of establishing a reasonably high planar electric field strength. Field strength (in volts/meter) can be computed from the relationship:

$$E\ (v/m) = 39.4\ (V_t\ 10^{-A/10}/h) \tag{4-25}$$

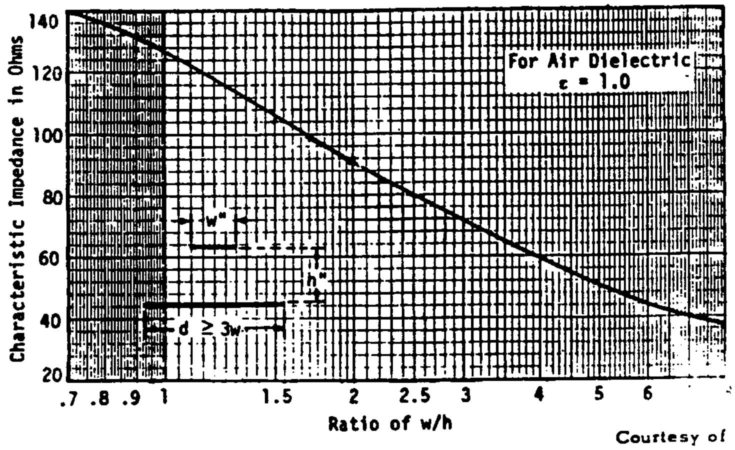
where

V_t = input voltage to the line, in volts

A = matching network loss, in dB

h = separation between transmission line plates, in inches

The design parameters for this type of line are shown in Figure 4-59. The figure points out that the lower plate should be at least three times wider than the upper plate, and that the characteristic impedance of the line is a function of upper plate width and plate separation [24].



Courtesy of DON WHITE CONSULTANTS, INC.

Figure 4-59 Characteristic Impedance of Strip Line vs. w/h Ratio

Impedance matching is relatively unimportant so long as the line length is less than 0.1λ. However, at frequencies when this relationship does not apply, the line acts like a transmission line, and standing waves will exist unless careful matching is accomplished.

When a test sample is inserted into the line, it will perturb the field, and two effects will result. First, the VSWR of the line will change. This change may be approximated from the following equations:

$$VSWR = R_L/Z_D \quad (4\text{-}26)$$

$$Z_D = [(-jX_c)R_L] / [R_L - jX_c] \quad (4\text{-}27)$$

$$X_c = 1/(2\pi f C_d) \quad (4\text{-}28)$$

$$C_d = 0.22\,A\,[(1/h_D) - (1/h)] \quad (4\text{-}29)$$

where

R_L = characteristic impedance of the line, in ohms

A = area in square inches of the test specimen face parallel to the plates

h_D = separation between transmission line plates, in inches

Second, the electric field strength in the region above the test specimen will increase because of the effective increase in capacitance in that portion of the structure. As an approximation, the increase in field strength would be inversely proportional to the separation between the upper plate and the top face of the test sample. Thus, if the separation is reduced by one-third (to 2/3 of its normal spacing), the field strength will increase to 1.5 times its normal value.

An example of a 90 ohm parallel plate transmission line capable of plane wave measurements to 200 MHz is shown in Figure 4-60. The line can test samples of cigar-box size while maintaining a VSWR of less than 1.2:1. The upper plate tapers to an input connector and a 50-to-90 ohm resistive matching network. The termination resistors in this arrangement are three 270 ohm distributed resistance cards connected in parallel.

Courtesy of DON WHITE CONSULTANTS, INC.

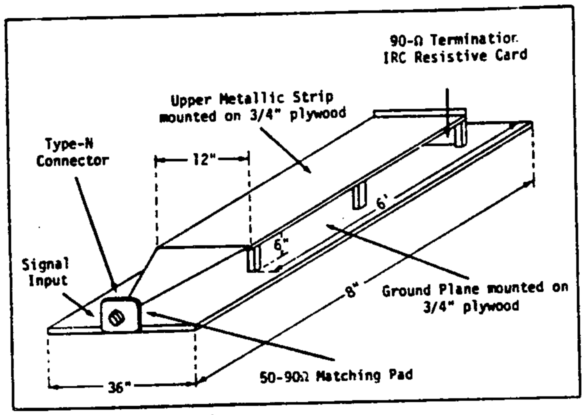


Figure 4-60 90 ohm Strip Transmission Line

The lower plate of the parallel plate transmission line should be bonded to the shielded enclosure wall. The equipment under test should be placed as near to the center of the parallel plate line as possible. No other grounding scheme should be employed.

A small loop antenna or short stub monopole antenna should be placed inside the enclosure under test. The output of the antenna inside the enclosure should be routed through a double shielded cable to a detector.

4.6.7.3 Test Procedure

The following steps should be used in performing this test:

- o Connect a signal generator to the input of the parallel plate line. Be sure that the line is properly terminated for the generator and that the generator output level is sufficient to produce the desired field intensities.
- o Place the small antenna inside the enclosure under test and connect it to the detector.
- o Adjust the signal generator frequency and detector frequency appropriately.
- o Increase the signal generator output level until an indication of received signal is given by the detector. Continue to increase the generator output level until the detector output is at least 3 dB (and preferably 10 dB) above the system noise level. Note the output indication of the detector and record the generator input level (E_2).

- Remove the test enclosure from the test fixture, keeping the receiving antenna in the same position in the transmission line as when the test enclosure was in place.
- Reduce the signal generator output level until the detector indication is the same as that recorded with the test enclosure in place.
- Record the output signal generator level (E_1).
- Calculate the shielding effectiveness measured from the relationship:

$$SE(dB) = 20 \log (E_1/E_2) \tag{4-30}$$

4.6.7.4 Advantages/Disadvantages

- The parallel plate line is easily obtained or simple to construct, and has broadband characteristics.
- The field generated is uniform and of known impedance level.
- High field intensities may be established, with nearly all energy concentrated within the line.
- Design parameters limit the capabilities of this device to the testing of medium to small-size enclosures.
- The strip line concept is adaptable to generating planar fields having a wide range of wave impedance, by controlling line termination impedance.

4.6.8 MIL-STD-1377 Testing [21],[25]

This standard is employed by weapon system developers to evaluate shielding and filtering effectiveness. It defines applicable shielding effectiveness measurement techniques covering the frequency ranges of 100 kHz to 30 MHz, and 1000 MHz to 10 GHz; and filtering effectiveness measurement techniques covering the frequency range of 100 kHz to 10 GHz. Although not originally intended for this purpose, it is being brought to the attention of the GSE designer, since he may want to consider an approach taken in MIL-STD-1377 to evaluate the equipment or system he is developing.

At frequencies below 30 MHz, shielding in MIL-STD-1377 is evaluated on the basis of Surface Transfer Impedance (STI). STI is the ratio of the longitudinal voltage drop on the outer surface of the shield to the current creating that voltage drop. If a cable or cable/connector is under test, it is terminated in a short, and driven by a signal source. The STI is measured with an RF voltmeter, as shown in Figure 4-61. Shield enclosure discontinuities are measured in a similar way. An STI of about 10 milliohms has been correlated to a level of 15% of Maximum No Fire Current for typical electroexplosive devices [25].

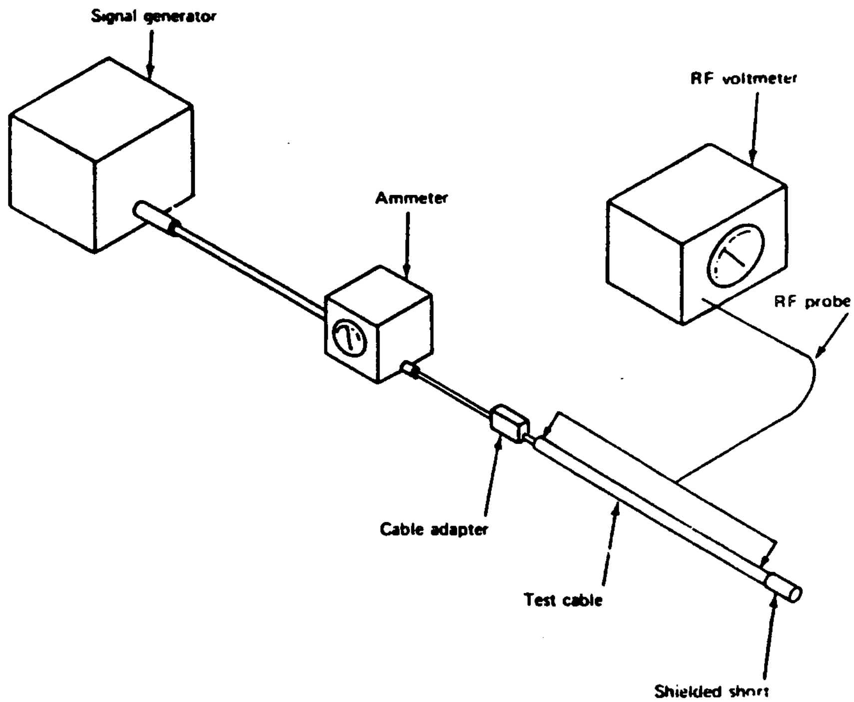


Figure 4-61 STI Measurement Configuration for Cables

At frequencies above 1 GHz, a special test enclosure is employed. The enclosure is a conductive container into which the test sample is placed. If a cable/connector shielding effectiveness test is being performed, the cable assembly is mounted in the enclosure. One end is terminated in a short, and the other end is brought out of the test enclosure to a power meter. A wire strung diagonally within the enclosure is used to generate a test field. Another wire strung diagonally within the enclosure is used for calibration purposes. A paddle wheel tuner is employed as a mode converter, so that a maximum level environment can be established.

Figures 4-62 and 4-63 illustrate the test set up and enclosure for this test. A similar procedure is employed for determining the shielding effectiveness of equipment and system enclosures. An illustration of a test set up for shielding effectiveness testing using a mode-stirred chamber is presented in Figure 4-64.

4.7 SUMMARY OF GOOD SHIELDING PRACTICES

The following represent what might be considered the more salient points on GSE shielding design considerations covered or implied in the previous discussion:

a. Good conductors such as copper, aluminum, and magnesium should be used for high-frequency electric-field shields to obtain the highest reflection loss.

b. Magnetic materials such as iron and Mumetal should be used for low-frequency, magnetic-field shields to obtain the highest penetration loss.

c. Any shielding material strong enough to support itself will usually be thick enough for shielding electric fields at any frequency.

d. In the case of thin-film shields, the effectiveness of the shield is fairly constant for material thicknesses below $\lambda/4$, and increases markedly above that thickness.

e. Multiple shields (for both enclosures and cables) can provide both higher shield effectiveness and extended shielding frequency range. Cost considerations will probably be the deciding factor between use of multiple shields and using other means of achieving EMC, although factors such as reduced cable flexibility with double braids may also come into play.

f. All openings or discontinuities should be treated in the design process, to assure minimum reduction in shield effectiveness. Particular attention should be paid to selection of materials that are not only suitable from the shielding standpoint, but from the electro-chemical corrosion viewpoint as well.

g. When other aspects of system design will permit, continuous butt or lap weld seams are most desirable. The important consideration is to get intimate contact between mating surfaces over as much as the seam surfaces as possible.

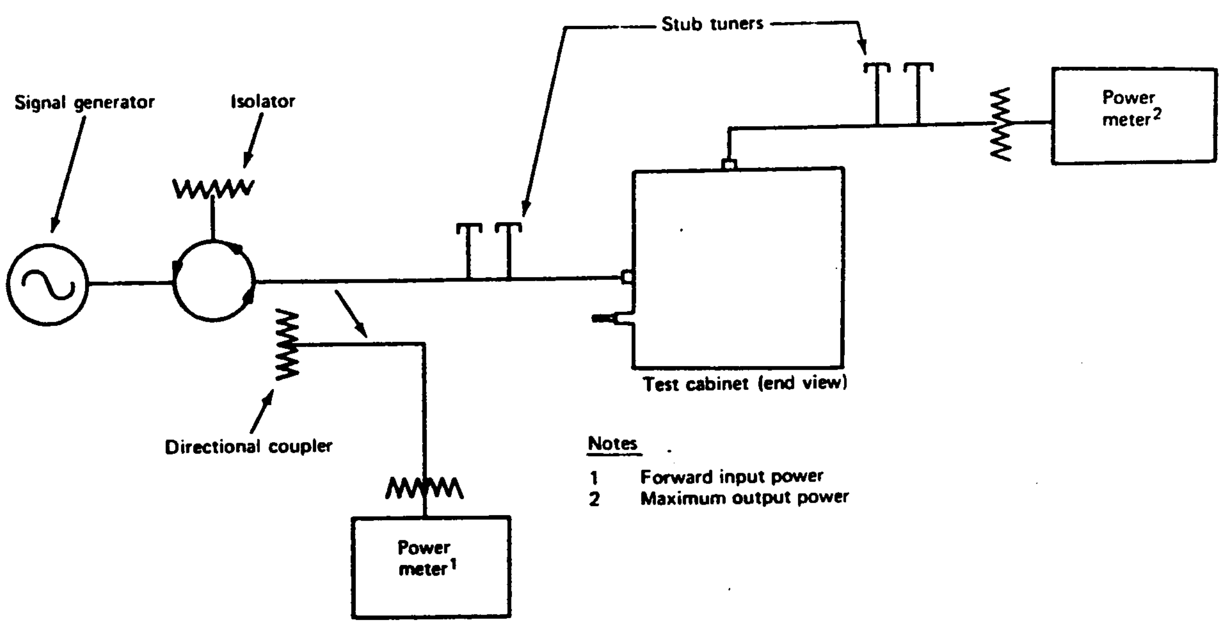


Figure 4-62 Arrangement of Test Apparatus for MIL-STD-13771 1-10 GHZ Tests

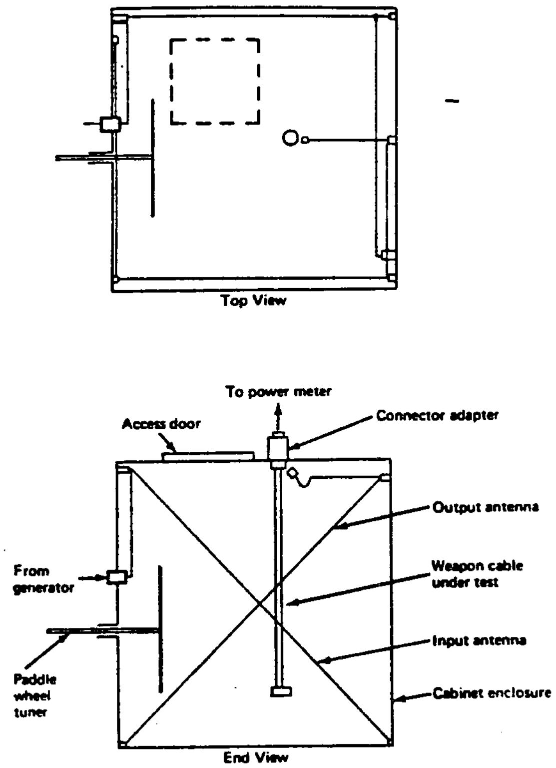


Figure 4-63 Test Cabinet (With Cable Under Test)

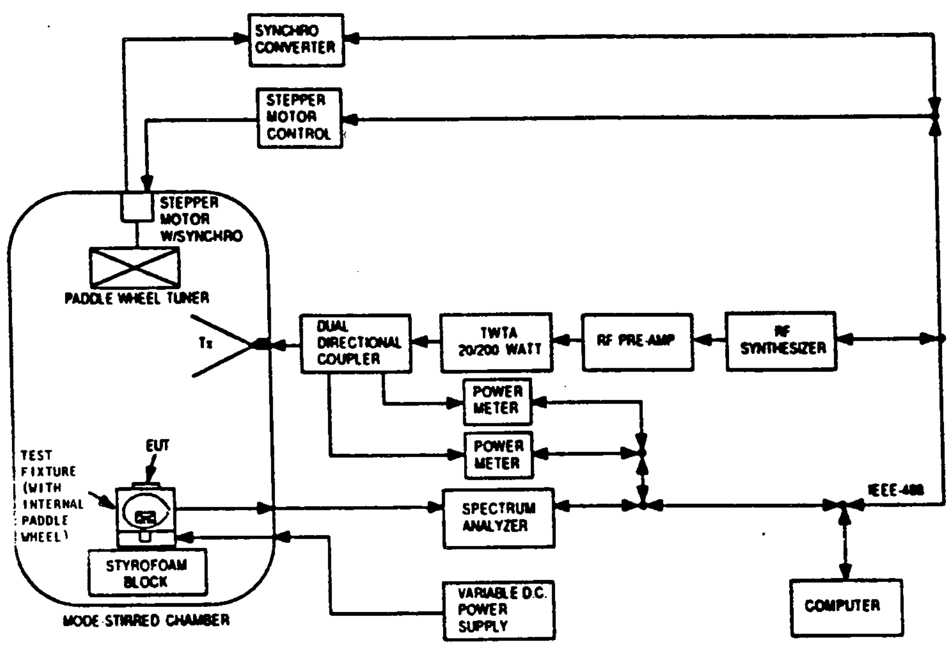


Figure 4-64 Shielding Effectiveness Test Set Up Using Mode-Stirred Chamber

h. Surfaces to be mated must be clean and free from nonconducting finishes unless the bonding process positively and effectively cuts through the finish. When EMC and finishing specifications conflict, it is important that the finishing requirement be modified.

i. The critical factors in flight-deck cable shielding are shield coverage under operational cable flexing conditions, and cable shield termination at the connector. A minimum of 94% shield coverage is recommended for these applications. Shields should be peripherally bonded to connector back shells to maintain shielding effectiveness at mating surfaces.

j. Conductive gaskets and spring fingers, waveguide attenuators, screens and louvers, and conducting glass are the major devices and mechanisms available for maintaining enclosure shield effectiveness. Many factors in addition to shielding capabilities per se, ranging from space availability to cost, and from air circulation requirements to visibility factors, will affect particular methods employed in particular situations.

k. Shielding represents only one method of reducing equipment EM interactions, and should not be considered without also considering tradeoffs of filtering, grounding, and bonding techniques that simplify or eliminate requirements for shields.

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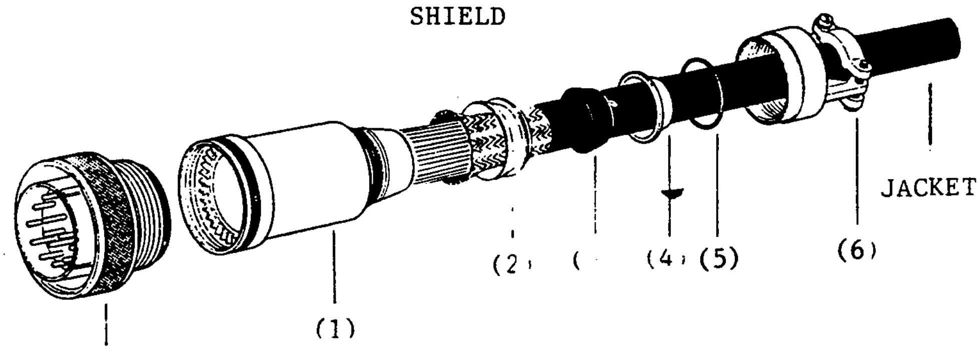
ASSEMBLY INSTRUCTIONS FOR SHIELDED AND JACKETED CABLES



(1) Adapter	(3) Grommet	(5) Teflon Washer
(2) RFI Ferrule	(4) Grommet Ferrule	(6) Clamp

The following suggested procedure serves as a guide for proper assembly and installation of Glenair Lightweight RFI Environmental Adapters. It is recommended that trial samples of appropriate cables be used to determine proper trim dimensions of the cable jacket, shield and individual conductors. (Numbers in parenthese refer to the exploded view shown.)

NOTE: As with any electrical connector assembly procedure be sure to use the proper tools. For convenient, reliable assembly of the connector and adapter, it is suggested that Glenair's connector wrenches, pliers, etc., be used.

(a) Temporarily assemble adapter (1) to connector.

(b) Place remaining adapter assembly components (2 thru 6) on cable in sequence shown. Keep these components at a convenient distance from end of cable, so they will not interfere with subsequent assembly steps.

(c) Insert cable into adapter (1) and bottom against connector. Hold cable in position and mark cable jacket at rear end of adapter (1).

CAUTION: If cable conductors are to have service loops, or if conductors will have cross-overs, etc. allow sufficient added length to cable to compensate for these factors.

(d) Remove adapter (1) from connector and place on cable with components in step (b) above.

(e) Trim cable jacket and shield at mark made in step (c) above (allowing for service loops and cross-overs).

(f) Strip jacket 1/4 inch back from trim point in step (e) to expose shield.

ASSEMBLY INSTRUCTIONS FOR SHIELDED AND JACKETED CABLES CONT'D

(g) Prepare and terminate cable conductors in accordance with established practices.

(h) Assemble adapter (1) to connector and tighten securely.

(i) Flare shield over tapered end of adapter (1) and slide RFI ferrule (2) into place over shield. Hold ferrule in position and trim any exposed shield strands adjacent to rear threads on adapter.

(j) Slide grommet (3), grommet ferrule (4) and washer (5) against RFI ferrule (2).

(k) Engage clamp (6) with adapter and tighten securely. Tighten clamp saddles on cable jacket.

5.0 BONDING

5.1 GENERAL

Bonding is the establishment of a low impedance path between two metal surfaces. This path may be between two points on a system ground plane, or between ground reference and a component, a circuit or a structural element. The purpose of the bond is to make the structure homogeneous with respect to the flow of RF currents, thus avoiding the development of electric potentials between metallic parts which can produce interference.

Generally there are two types of bonding: direct bonding, where there is metal-to-metal contact between the members to be bonded; and indirect bonding through the use of conductive jumpers.

Permanent joints of metallic parts made by welding, brazing, sweating, swaging and soldering are the best direct bonds. Semi-permanent joints of machined metallic surfaces rigidly held together provide excellent direct bonds, as long as the contact areas are clean and all non-conductive coatings are removed prior to assembly. Joints that are press-fitted or jointed by self-tapping or sheet metal screws cannot be relied upon to provide a low-impedance bond at high frequencies. Riveted joints on 3/4" centers are acceptable if the rivet holes are bare. Direct bond must always be made through continuous contact to base or conductively finished metals.

An indirect bond (or bonding jumper or strap) is an intermediate electrical conductor used to connect two isolated items. Because jumpers often have significant impedance at frequencies above 10 MHz, their use is less preferable than direct bonds. However, there are obviously circumstances when direct bonds cannot be used (such as when equipment must be removable, or when mechanical shock-mounting is necessary), and it is these situations that warrant indirect bonds.

Good bonding between equipment and a ground reference plane is essential to minimizing interference.

- Good bonding enables the design objectives of other methods of EMI suppression, such as shields and filters, to be more nearly achieved.
- Good bonding minimizes the build-up of RF voltage differences and ground current loops.
- An adequate bond deters the build-up of static charges in equipment operation.

In addition to reducing interference, good bonding minimizes damage which might be caused by lightning strikes, and protects personnel from the shock hazard that could result if primary power were inadvertently shorted to an enclosure.

An example of the effects of a poor bond is shown in Figure 5-1. If the impedance of the bond is significant, the interference signal, which is supposed to be bypassed by the filter capacitors (Path 1), will be coupled into the susceptible equipment (Path 2).

L_B = INDUCTANCE OF A POOR BOND
R_B = CONTACT RESISTANCE OF A POOR BOND

NOTE – WHEN $(1/j\omega C + R_L) < (R_B + j\omega L_B)$ INTERFERENCE CURRENTS WILL FOLLOW PATH (2) TO SUSCEPTIBLE EQUIPMENT.

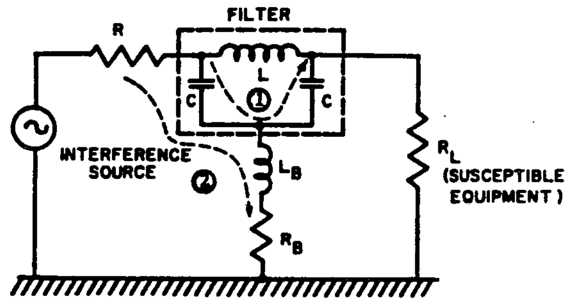


Figure 5-1 Circuit Representing Poor Bonding Between a Filter and Ground

5.2 SURFACE TREATMENT

Both direct and indirect bonding connections require metal-to-metal contact of bare surfaces. It is frequently necessary to remove protective coatings from metals to provide a satisfactory bond. The area cleaned for bonding should be slightly larger than the area to be bonded. Ridges of paint around the periphery of the bonding area can prevent good metal-to-metal contact. Washers or fittings must fit inside the cleaned area. Immediately prior to bonding, all chips, paint, grease, or other foreign matter must be removed with a proper cleaning solution[1].

After bonding, the exposed areas should be refinished as soon as possible with the original finish. However, if the paint used is too thin, refinishing paint may seep under the edges of bonded components and impair the quality of the bond.

A suitable conductive coating may be used when removable components must be provided with a protective finish. Where aluminum or its alloys are used, corrosion resistant finishes that offer low electrical resistance are available. Refer to Figure 4-28 of Chapter 4 for data showing the degradation of shielding effectiveness of metal-to-metal joints caused by anodizing aluminum, and the characteristics of various conductive aluminum finishes.

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1. See also Paragraph 5.5.1 for additional information in connection with the bonding of enclosure seams.

5.3 CORROSION

Corrosion occurs between two dissimilar metals in solution, since they form an electro-chemical cell. The extent of corrosion depends on the metals comprising the electro-chemical cell and the conditions under which the dissimilar metals come into contact with each other. By properly modifying these two factors, the extent of corrosion in the vicinity of a bond can be reduced.

No appreciable corrosive action occurs between two metals in the same group in the electro-chemical series shown in Table 5-1. If the two metals are in different groups, the metal coming first in the list of Table 5-1 will form the anode and be relatively heavily corroded, whereas the metal coming later will form the cathode and be relatively free from corrosion and will be protected.

The greatest degree of corrosion occurs when dissimilar metals are openly exposed to salt water, salt spray, rain, gasoline, jet fuel, or other liquids that may act as an electrolyte. Minimum corrosion occurs when metals are kept dry and completely free from exposure to moisture. The following three exposure conditions are defined:

Exposed:

Metal has an open, unprotected exposure to weather.

Sheltered:

Milder exposure than above, and metal surfaces receive limited protection from direct action of weather.

Housed:

Metal surfaces of equipment are housed in weatherproof buildings.

For a given pair of dissimilar metals in contact under the three exposure conditions enumerated above, the extent of corrosion can be minimized. Table 5-2 may be used for this purpose [1]. The table indicates the protection to be applied to a bond for the expected environmental conditions in which the particular type of bond is to be used. The table must be used in conjunction with Table 5-1 to identify the group associated with each bond metal.

Table 5-1

Electrochemical Series Ordered by Decreasing Sensitivity to Corrosion

Anodic End (most easily corroded)

Group I	Magnesium
Group II	Aluminum, aluminum alloys, zinc, cadmium
Group III	Carbon steel, iron, lead, tin, tin-lead solder
Group IV	Nickel, chromium, stainless steel
Group V	Copper, silver, gold, platnum, titanium

Cathodic End (least corroded)

Table 5-2

Groups of Materials Recommended for Providing Protective Bond Between Two Dissimilar Metals Used as Anode and Cathode

Condition of Exposure	Anode				Cathode
	I	II	III	IV	
Exposed	A	A			II
Sheltered	A	A			
Housed	A	A			
Exposed	C	A	B		III
Sheltered	A	B	B		
Housed	A	B	B		
Exposed	C	A	B	B	IV
Sheltered	A	A	B	B	
Housed	A	B	B	B	
Exposed	C	C	C	A	V
Sheltered	A	A	A	B	
Housed	A	A	B	B	

Notes: Bond Protection Code

A. The couple must have a protective finish after metal-to-metal contact has been established so that no liquid film can bridge the two elements of the couple.

B. The two metals may be joined with bare metal exposed at junction surfaces. The remainder must be given an appropriate protective finish.

C. This combination cannot be used except where short life expectancy can be tolerated or when the equipment is normally stored and exposed for only short intervals. Protective coatings are mandatory.

Consideration should be given to the electro-chemical series to assure that, where possible, corrosion occurs only to replaceable elements of a bond. For example, if an aluminum (Group II) equipment case is to be bonded to a stainless steel (Group IV) frame, it is good practice to interpose a tin- (Group III) or cadmium - (Group II) plated washer between the two metal surfaces. If the protective coating is then chipped, the washer instead of the aluminum case will be attacked by corrosion.

5.4 BONDING EFFECTIVENESS CHARACTERISTICS

5.4.1 Bond Jumper Equivalent Circuit

The use of bonding jumpers in indirect bonding is equivalent to the problem of maintaining low impedance paths. At low frequencies, bonding jumpers do not present any special problems except for resistance, and any reasonable length jumper can be used. At higher frequencies, however, the RF impedance of the bond becomes a critical design consideration.

The equivalent circuit of a bond strap at frequencies where the length of the strap is short compared to a wavelength is a simple parallel-tuned circuit, as illustrated in Figure (5-2a) [2]. As such, the bonding jumper has the usual electrical parameters of resistance, inductance, and capacitance. Of these parameters, its resistance is an inherent property of jumper resistivity, depending on the material selected; its capacitance is dependent upon the physical configuration and separation from the bonded units; and its inductance is dependent upon the physical dimensions of the bonding jumper.

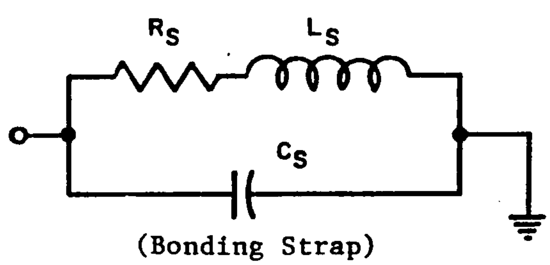


Figure 5-2a
Bonding Equivalent Circuits

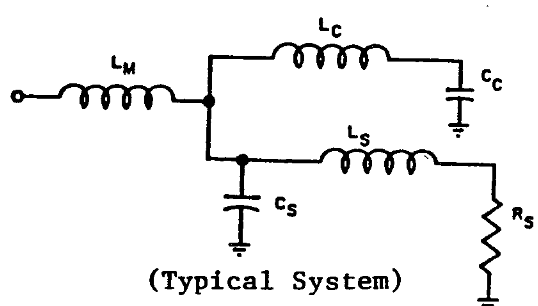


Figure 5-2b
Bonding Equivalent Circuits

The dc resistance of a bonding strap per unit length of the bond can be obtained from the relationship:

$$R_{dc} \text{ (ohm/cm)} = \rho/A \qquad (5\text{-}1)$$

where

ρ = the specific resistivity of the material, and

A = the cross-sectional area of the bond.

If ρ is in units of ohm-centimeters (for copper, $\rho = 1.724 \times 10^{-6}$ ohm cm) and A is in square-centimeters, then R_{dc} is in ohms-per-centimeter of bond length.

At frequencies where skin effect becomes significant, the ac resistance of the bond can differ significantly from its dc value. Skin effect is defined as the depth at which the current density is 1/e (about 37%) of its surface current density. This depth, in centimeters, is defined as[3]:

$$\delta = 5033 \sqrt{\rho/\mu f} \qquad (5\text{-}2)$$

where

f = the frequency, in Hertz

μ = material permeability, relative to copper

The assumption that all of the current in a conductor is within the skin depth defined by equation (5-2) results, in the case of circular conductors, in the following equation for ac resistance:

$$R_{ac} \text{ (ohms/cm)} = \frac{\rho}{2\pi r\delta} \qquad (5\text{-}3)$$

where

r = the conductor radius, in centimeters.

Figure 5-3 indicates the ratio of ac-to-dc resistance for several sizes of copper wire over the frequency range of 1-1000 MHz.

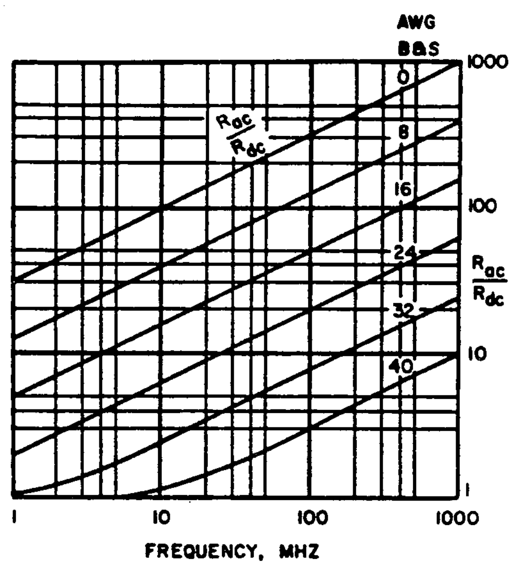


Figure 5-3 Ratio of ac to dc Resistance for Various Wire Sizes

For a straight bonding strap of non-magnetic metal, bonding inductance, L, is defined by [3] as:

$$L = 0.00508\ a[2.303\ LOG_{10}(2a/b+c) + 0.5 + 0.2205(b+c/a)] \qquad (5\text{-}4)$$

If the bond is made with a straight piece of wire of circular cross section, L is given by:

$$L = 0.00508\ a[2.303\ LOG_{10}(4a/d) - 0.75] \qquad (5\text{-}5)$$

In the last two equations,

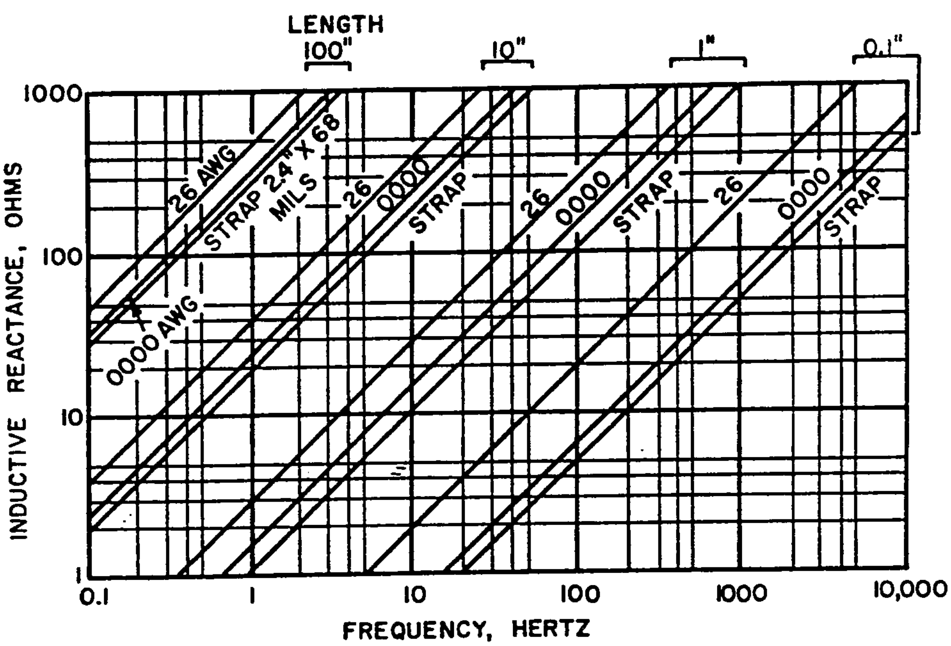
a = the length of the strap in inches

b = the width of the strap in inches

c = the thickness of the strap in inches, and

d = the wire diameter in inches

Figure 5-4 is a plot of the calculated inductive reactance of three specific bonds as a function of frequency. Two of the bonds are copper wire, while the third is a copper strap.



CALCULATED INDUCTIVE REACTANCE, IN UH, FROM,

$$L = 0.00508\ a\left[2.303\ LOG_{10}(4a/d) - 0.75\right],\ \text{OR}$$

$$L = 0.00508\ a\left[2.303\ LOG_{10}(2a/b+c) + 0.5 + 0.2205(b+c/a)\right]$$

WHERE a = Length, (Inches)
b = Width, (Inches)
c = Thickness, (Inches)
d = Diameter, (Inches)

Figure 5-4 Inductive Reactance of Wire and Strap Bond Jumpers

At frequencies where the length of the bond approaches or exceeds the wavelength of concern, the bond can act as a transmission line. The result is the development of standing-waves on the bond, and a deterioration in bond effectiveness.

Typical bond straps may be manufactured as flat, solid straps; as braided straps; or a solid wire. The material used is generally copper or aluminum, but in the case of flat straps can also be phosphor bronze.

5.4.2 Equipment Effects on Indirect Bonds

The representation of the equivalent circuit of a bond as a parallel-tuned circuit ignores the effects of the equipment enclosure or other item being bonded, and the latter can significantly affect bond strap performance. Figure 5-2b shows an equivalent circuit when taking the system being bonded into account. In Figure 5-2b, the inherent inductance of an equipment case is represented by L_c, its capacitance to other equipments and the reference plane is indicated by C_c, and the impedance contributed by the insertion loss measuring circuit is represented by L_m. Measurements confirm this representation, but additionally denote that, under certain circumstances, the dip in shielding effectiveness caused by resonant effects can go below 0 dB(-10 to -20 dB have been noted), indicating that the induced voltage on the unit being bonded can be increased by attachment of a bondstrap.

Figures 5-5 and 5-6 contain plots of measured bonding effectiveness of two bond straps of different lengths that are connected between an equipment case and a ground plane [2]. The proximity of the bond strap to the ground plane is also a parameter in the plots. The bonding effectiveness is the difference (expressed in dB) between the induced voltages on the equipment case with and without the bond strap. A negative value of bonding effectiveness in the plots indicates that the bond strap increases the amount of voltage developed on an equipment case at the indicated frequency.

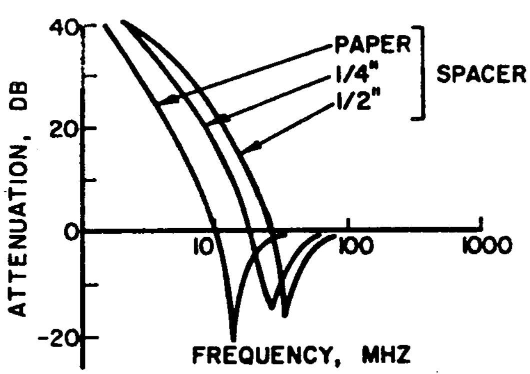


Figure 5-5
Bonding Effectiveness of 9 1/2 inch Bonding Strap, Measured Using Shunt-T Insertion Loss Technique

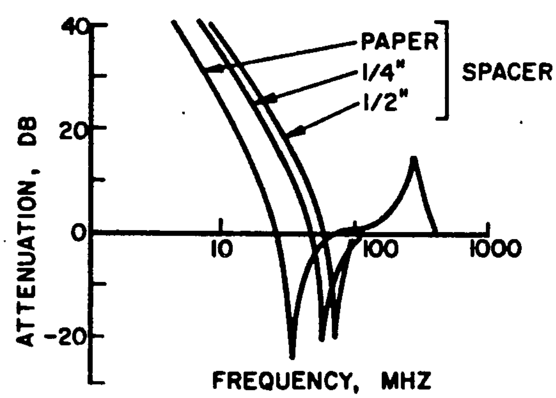


Figure 5-6
Bonding Effectiveness of 2 3/8 inch Bonding Strap, Measured Using Shunt-T Insertion Loss Technique

It should be noted in Figures 5-5 and 5-6 that above about 100 MHz for 9-1/2" straps and above 400 MHz for the 2-3/8" straps (but below the frequencies where the length of the bond is a significant part of a wavelength), a bond will generally not affect the voltage developed on the equipment case. Also, the bonding effectiveness of the 2-3/8" straps are typically 20 dB better at 10 MHz than with the longer straps.

The data in Figures 5-5 and 5-6 show a significant frequency range for each bond where the bond creates the negative bonding effectiveness condition cited earlier. This range can be increased in frequency (and reduced somewhat in frequency range) by employing shorter bonds located away from the ground plane.

In selecting a bond strap, the resonant frequency of the strap must be well above the highest interfering frequency which is expected to be encountered by the GSE/Avionics. This resonant frequency corresponds closely to the frequency at which the worst bonding effectiveness value is obtained.

To reduce the RF impedance of the bond, and thus increase its bonding effectiveness, a high C to L ratio should be achieved, by minimizing the case to ground spacing and the length to width ratio of the bondstrap. The inductance of the bonding strap can be minimized by selecting a strap whose length is less than 5 times its width [4]. A good bond should have an inductance of less than 0.025 microhenry [5].

5.4.3 Bonding Resistance

Although measurement of the dc resistance of a bond is obviously not an indication of the bond's ac characteristics, it is often used as a guide to the anticipated performance of the bond. Depending on the purpose of the bond, military specifications designate the maximum dc resistance allowable for a good bond.

For example, bonds that are installed to prevent shock hazards are required by both MIL-B-5087B(2) INT AMD 3 and MIL-STD-1310E (SHIPS) to have a resistance of less than 0.1 ohms. Radio-Frequency bond requirements in MIL-B-5087B specify a resistance of less than 2.5 milliohm. Additionally, in areas prone to explosion or fire hazards, maximum values of bond resistance are designated; these values are functions of anticipated maximum fault current in the event a power line to ground short occurs. A guideline as far as a good RF bond is concerned, is a dc resistance value of between 0.25 and 2.5 milliohms [4].

Measurements of various types of bond straps indicate the flat, solid strap provides less inherent ac resistance than other types [6]. However, this advantage is often a minor one, and may not offset the advantages of flexibility of braid and lower cost of solid wire.

5.5 BONDING TESTS

5.5.1 Introduction

The bonding tests described below are intended to identify practical test techniques for measuring the parameters of an electrical bond. The specific tests to be discussed include:

- o DC Resistance Measurement
- o Swept Frequency/Shunt-T Insertion Loss Measurement

Tabulated procedures are provided for each of these tests. They are included for descriptive purposes only, and may have to be altered depending on specific test conditions.

5.5.2 DC Resistance Measurement

5.5.2.1 Objective

This test is intended to give design personnel a general indication of bond adequacy after installation, based on the dc resistance of the bond. The test is often performed to establish whether or not the bond resistance is within pre-established design limits. However, as pointed out previously, measurement of bond resistance is not a complete indicator of the electrical characteristics of the bond.

5.5.2.2 Instrumentation and Test Setup

Equipment required for this test is as follows:

- o A dc resistance bridge capable of measuring to about 0.001 ohm. The bridge should be portable and easily moved, and should not be position-sensitive. Connection of the test sample to the bridge terminals should be easily performed without cumbersome adapters or special tools.
- o A pair of heavy-duty spring clip leads for connection between the bridge and the bonded junction. Clip leads may be connected to braided straps and plugs to make connection to bridge. Total resistance of external connectors and leads should not be greater than 0.001 ohms.

Equipment set-up is as follows:

- o Using the heavy-duty spring clips and braid, connect the leads to the bridge.
- o Place the bridge in operation according to the manufacturer's operation manual.
- o Zero the bridge, including leads, and connect the clip-leads across the bonded junction.

5.5.2.3 Test Procedure

- Adjust the bridge balance until a null is obtained.
- Compare the indicated resistance to the tolerance or design limit.
- If the measured resistance is above the required value, check to see that the bond has been properly installed. Tighten to design limit any bolts, clamps, etc., and repeat the test.
- If bond resistance is still not within limits, redesign of the bond may be necessary.

5.5.2.4 Special Precautions

- Be sure that the clip-leads and connections are not contributing significantly to the total resistance measured.
- Be sure that the bridge is zero balanced to take into account the effects of the external connectors and straps.

5.5.2.5 Advantages/Disadvantages

- A DC resistance measurement of a bond will give an indication of bond performance under dc or low-frequency high current flow, but does not provide RF impedance levels.
- Variations in bond RF impedance can severely affect system performance and safety. Since the dc resistance test does not give such data, its usefulness is restricted to an indicator of good mechanical union between metallic surfaces. Physical form factors of the bond, external to the bond junction, affect bond RF performance. Additional tests may be necessary to insure adequate RF performance.

5.5.3 Swept Frequency/Shunt-T Insertion Loss Measurement

5.5.3.1 Objective

The shunt-T insertion loss test, using a swept frequency source and wideband detector, is designed as a rapid method of characterizing bonding impedance over the frequency range where bonding performance degradation might ordinarily occur. The test, with careful test fixture design, can measure bond impedance up to about 400 MHz [2], [7].

5.5.3.2 Discussion

The insertion loss-shunt-T test is based on the fact that if the shunt arm impedance of a T-attenuator is low in comparison to the series arm impedances, the current through the shunt arm will be essentially constant for varying values of shunt arm impedance. For a fixed voltage input to the T-attenuator, variations in the T-attenuator output voltage will be proportional to changes in the shunt arm impedance.

Figure 5-7 is an equivalent circuit of an insertion loss impedance measurement device. Two isolation resistors, R_s, are placed in series with the connections to the unknown impedance to form a T-network. The isolation resistors are chosen to be large with respect to the unknown impedance. The symbols R_1, R_2, L_1 and L_2 denote the connection impedances.

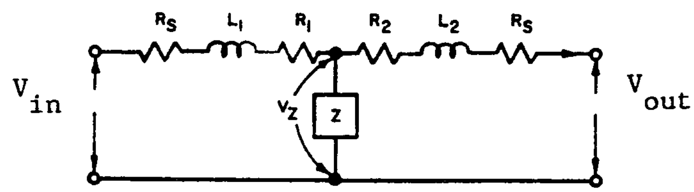


Figure 5-7 Equivalent Circuit of the Impedance Measuring Device

The relationship between the unknown impedance and the insertion loss through the test fixture is given as:

$$Z = \frac{R_s(R_s + 50)}{50} \left| \frac{V_{out}}{V_{in}} \right|, \text{ ohms} \qquad (5\text{-}6)$$

This equation assumes a 50 ohm measurement system. It also applies so long as the absolute value of the bond impedance is less than 0.1 R_s. In addition, with the measurement terminals open circuited, a minimum attenuation value will exist; the equation applies only for values that are well above this minimum.

The overall construction of a device for use in performing shunt-T insertion loss measurements is shown in Figure 5-8. The unit is intended to hold all of the test components and connectors, and maintain short lead lengths in the process. The bolt on the test fixture is connected to the bond strap under test. The case of the test fixture must be connected to the ground plane using threaded holes on the bottom of the fixture.

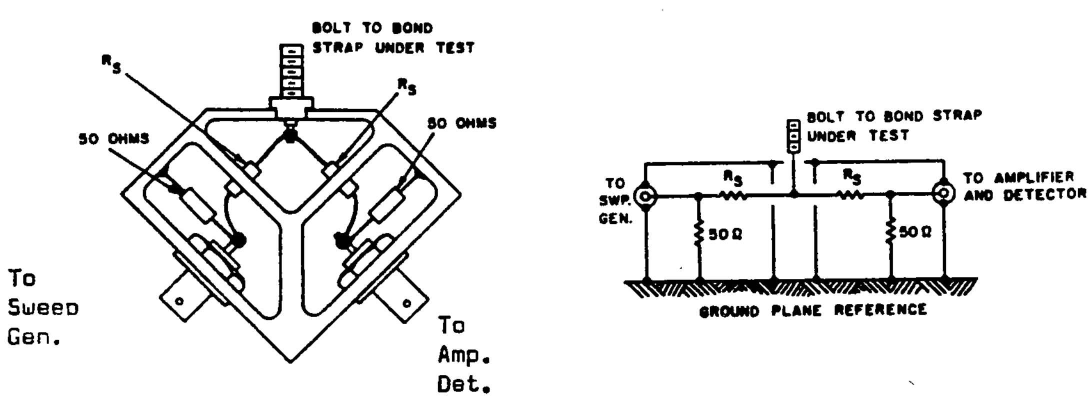


Figure 5-8 Schematic Diagram and Mechanical Details of Bonding Impedance Measurement Test Fixture

The two 50 ohm resistors in the test fixture are not shown in Figure 5-7. They are used for properly terminating the test equipment, and are incorporated in the fixture only when the value chosen for R_s is large compared to 50 ohms. In that case, the bond impedance becomes four times the value computed using equation (5-6).

5.5.3.3 Instrumentation and Test Setup

The test equipment is arranged as shown in the block diagram of Figure 5-9. The following equipment is required for the test:

- o A swept-frequency source covering the desired frequency range. The output signal level of the source should be constant over frequency, and the X-axis sweep should be a known function of frequency, preferably either linear or logarithmic. Its output signal level should be between 0 dBm and +10 dBm for achieving a reasonable dynamic range.

- o A wideband amplifier, with gain according to the following scale:

Anticipated Minimum Bond Impedance, Ohms	0.01	0.1	1.0	10	100
Amplifier Gain Required, dB	80	60	40	20	Not required

The amplifier should be impedance-matched to the sweep generator and to the detection system. The test fixture design has assumed both the sweep generator output and amplifier input impedances are 50 ohms.

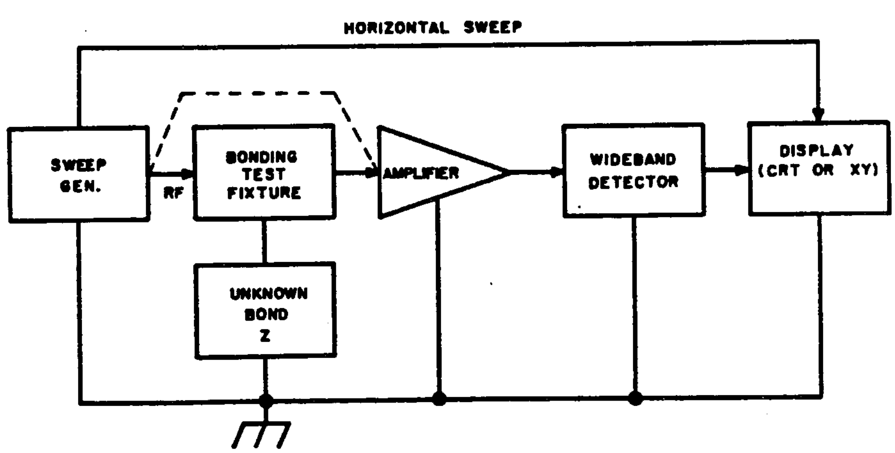


Figure 5-9 Swept Frequency System for Measuring Bonding Impedance

- A wideband detector, which may be a diode/low-pass filter combination, or a wideband logarithmic amplifier with dc output that is a function of log (E_{in}). The latter type is preferred.

- Display and readout may be either an oscilloscope or X-Y plotter. The input to the display device should be matched to the output of the wideband detector.

- A bonding test fixture, as shown in Figure 5-8. The test can be performed using commercially available 50 ohm resistive splitters without the shunt-matching 50 ohms, or a test fixture can be fabricated from commercially available components.

5.5.3.4 Test Procedure

- Attach the bonding test fixture to the bond to be tested, using bolts or clamps to insure good connection to the ground plane.

- Connect the test equipment as shown in Figure 5-9. Initially by-pass the test fixture, and connect the sweep source directly to the amplifier.

- Adjust the sweep generator for maximum desired deflection on the display/readout device and for the desired frequency range to be swept. Adjustment of wideband amplifier gain and display/readout deflection (X and Y) may be necessary to optimize the display.

- Break the connection between the sweep generator and the wideband amplifier, and connect the sweep generator output to the input of the test fixture. Connect the output of the test fixture to the input of the wideband amplifier.

- A display similar to those shown in Figure 5-10 should be obtained. The attenuator on the sweep generator output, or the input attenuator on the display input can be used to calibrate the vertical display in terms of insertion loss.

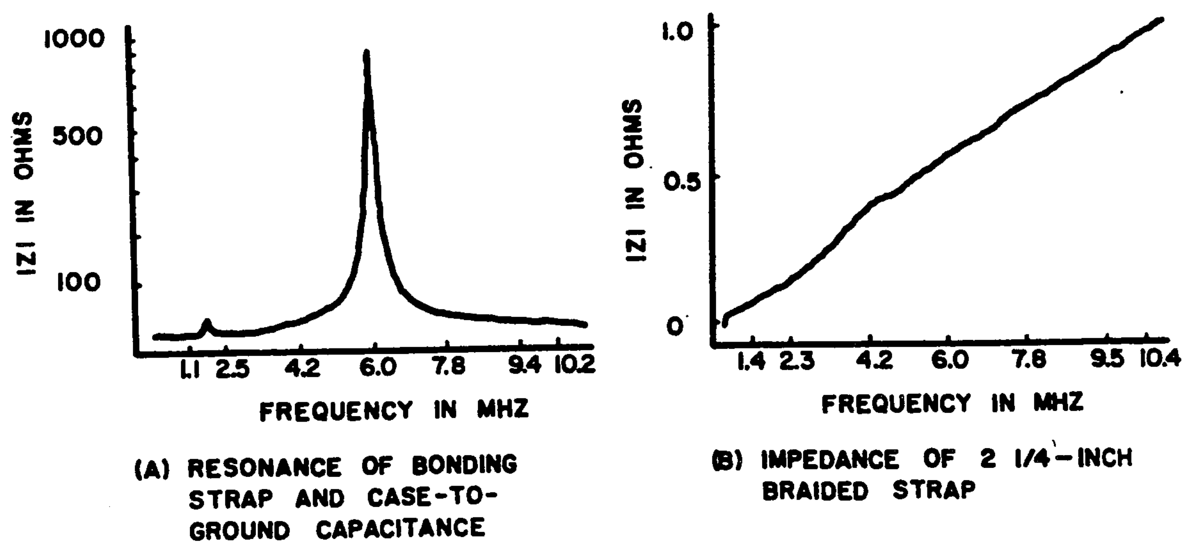


Figure 5-10 Typical Bond Impedance Vs. Frequency Measurements

- From equation (5-6), determine the bond impedance at a frequency of interest. Don't forget the 12 dB (4X) correction factor to be used if the test fixture contains 50 ohm shunt resistors. The impedance corresponding to a particular value of insertion loss should be compared with both the value of R_s, in the test fixture and with the minimum value of impedance obtained with the measurement circuit open. The measured value should be less than 0.1 R_s and well above the minimum impedance value.

- If the above conditions are not met, a test fixture with a different value of R_s must be used.

5.5.3.5 Advantages/Disadvantages

The shunt-T insertion loss test technique is convenient and easy to use, once the test fixture is available. With a suitable, leveled output sweep generator and a wideband detector, and a carefully constructed test fixture, the method can be accurate to 400 MHz, and down to absolute values of bond impedance of 0.01 ohm, although the low limit of impedance is frequency-dependent. The frequency range of the test is limited by the series reactance of the test fixture and by the amount of insertion loss introduced as a result.

5.6 SINGLE VS. MULTI-POINT BONDING

At frequencies sufficiently low that the entire structure is less than $\lambda/20$ in size, a single point bond or ground generally is most desirable since it avoids the problems of common-mode impedance coupling. In fact, efforts are made to assure that contact between two ground planes can take place at only one point to avoid noise that may result when irregular multiple point contacts are permitted, as when two large metallic surfaces are in surface contact under only moderate pressure.

Problems arise in implementing single point ground contacts when parasitic capacitance is significant between equipment housings or between different subsystem grounds. Problems also arise when interconnecting cables are used, especially those having cable shields longer than $\lambda/20$ between the source and the load. When cable shields interconnect subsystems, multiple grounding paths from a given subsystem to the system ground (reference) point exist. Furthermore, at frequencies high enough that the lengths in the system exceed $\lambda/20$, parasitic capacitive reactance can constitute a low impedance path while bond inductance can provide a comparable, but inductive, reactance. Anti-resonance can even occur, thus effectively isolating a circuit from its intended ground point at the anti-resonant frequency of the bond and the parasitic capacitance.

Thus, while efforts should be made to preserve a single-point bond for audio frequency applications, a multi-point bond is desirable at high frequencies [8]. The purpose of the multi-point bond is to obtain a low enough bonding inductance (reactance) that no significant impedance can appear between the two points that are connected by the bond.

5.7 BONDING DESIGN GUIDELINES

The effectiveness of a bond depends on its application, frequency range, magnitude of current, and environmental conditions such as vibration, temperature, humidity, fungus, and salt content in the ambient. Figures 5-11 through 5-15 provide examples of direct and indirect bonds, to illustrate the variety of techniques that have been employed to obtain low impedance connections. Many other examples are available in the literature, and particularly in MIL-STD-1310E.

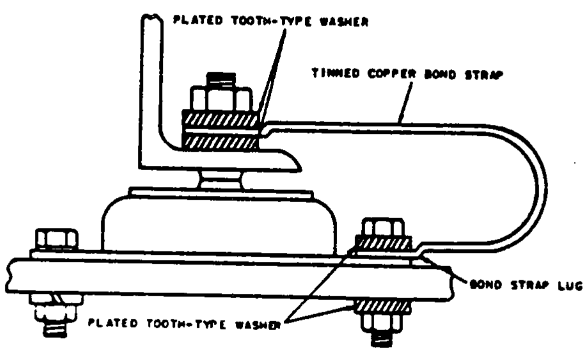


Figure 5-11
Typical Shock Mount Bond

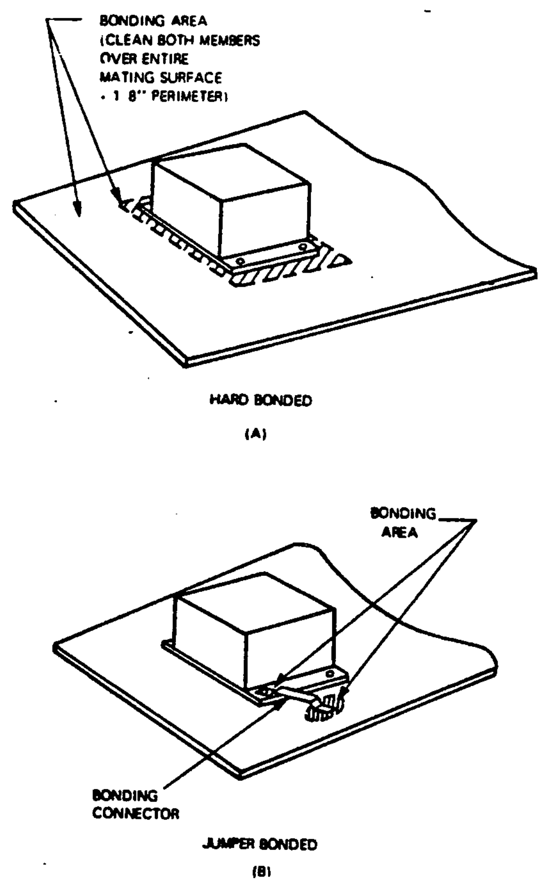


Figure 5-12
Base Mount Components

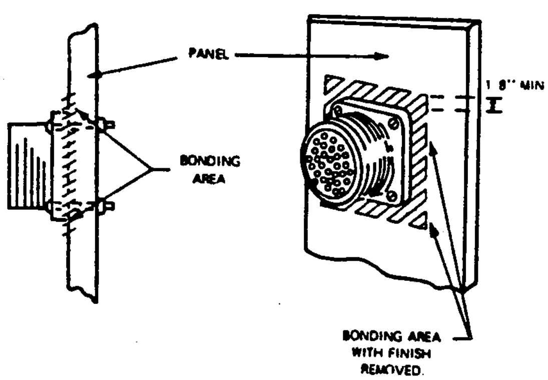


Figure 5-13
Bonding of a Connector

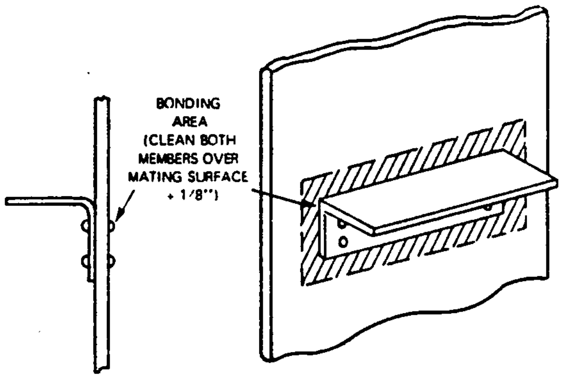


Figure 5-14
Bolted Members

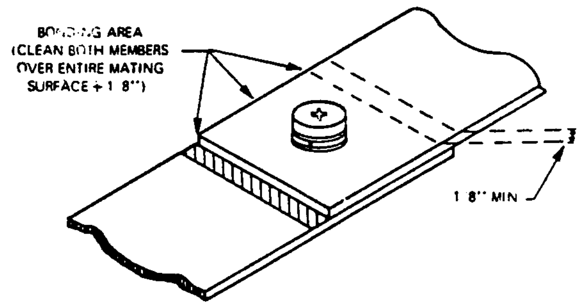


Figure 5-15
Bracket Installation
(Rivet or Weld)

Some general guidelines to obtaining good bonds are provided below.

The secret to good bonding is intimate contact between metal surfaces. Surfaces must be smooth, clean, and free of non-conductive finishes. The fastening method must exert sufficient pressure to hold the surfaces in contact in the presence of deforming stresses, shock, and vibrations associated with the equipment and its environments.

Bonds are always best made by joining similar metals. If this is not possible, special attention must be paid to the possibility of bond corrosion, through the choice of the materials to be bonded, the selection of supplementary components (such as washers) that will assure any corrosion will affect replaceable elements only, and the use of protective finishes.

Solder should not be used to provide mechanical strength to a bond.

Protection of the bond from moisture and other corrosion effects must be provided where necessary.

Bonding jumpers are only a substitute for direct bonds. The jumpers should be kept as short as possible, have a low resistance and low L/C, and not be lower in the electro-chemical series than the bonded members. A good rule to use is that the jumper should have a length-to-width ratio of less than 5.

Jumpers should be bonded directly to the basic structure, rather than through an adjacent part. They should not be connected with self-tapping screws, or by any other means where screw threads are the primary means of bonding.

It is always important in the broadest types of bonding application that the bonding jumper or direct bond is sufficient to be able to carry the currents that may be required to flow through it.

Use single point bonds at low frequencies (circuit dimensions $< \lambda/20$) and multi-point bonds at high frequencies.

BONDING BIBLIOGRAPHY

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6.0 GROUNDING

6.1 GENERAL

Grounding involves the establishment of an electrically conductive path between two points, with one point generally being an electrical/electronic element of a system and the other being a reference point. When the system element of concern is an equipment enclosure (such as one rack of the VAST AN/USM-247(V) System), then the reference point may be a ship structural member. When the system element of concern is a circuit within Avionics/GSE, then the reference point can be the equipment case or a ground plane that may or may not be isolated from the case.

A good, basic ground plane or reference is the foundation for obtaining reliable, interference-free equipment operation. An ideal ground plane would be a zero-potential, zero-impedance body that can be used as a reference for all signals in the associated circuitry, and to which any undesirable signal can be transferred for its elimination. An ideal ground plane would provide equipment with a common potential reference point anywhere in the system, so that no voltage would exist between any two points. However, because of the physical properties and characteristics of grounding materials, no ground plane is ideal, and some potential always exists between ground points in a system.

The extent to which potentials in the ground system can be minimized and ground currents can be reduced will determine the effectiveness of the ground system. A poor system, by enabling these spurious voltages and currents to couple into a circuit, subassembly, or equipment, can degrade the shielding effectiveness of well-shielded units, can essentially bypass the advantages of good filters, and can result in EMI problems that may be rather difficult to resolve after-the-fact.

Various ground systems must meet requirements for personnel safety, as related to the electrical power system, for lightning protection of personnel and property, and for providing a signal ground bus as a common electronic circuit return. Emphasis in this chapter will be placed on the establishment of good signal grounds, since this aspect of grounding will be of most concern to Avionics/GSE design engineers.

It is important to note that the GSE designer, even if he is developing a small, portable tester, must consider grounding from a system point of view. The system involved includes not only the test device, but the equipment or system that the device will be testing. The ground systems must be compatible under the interfaced condition.

6.2 GROUNDING TECHNIQUES

There are three fundamental grounding concepts that can be employed, as illustrated in Figure 6-1. The approaches can be used separately or in combination in any given equipment or system. They are: a) Floating

Ground System, b) Single-Point Grounding System, and c) Multiple-Point Grounding System.

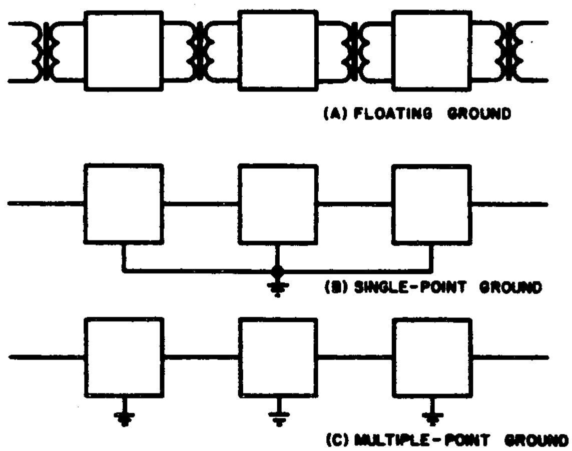

Figure 6-1 Grounding Methods

The Floating Ground System is a method to electrically isolate circuits or equipments from a common ground plane, or from any common wiring that might introduce circulating currents. Floating grounds depend, for their effectiveness, on truly "floating".

For many situations, this complete isolation may be very difficult to achieve. This is particularly true of a piece of GSE that will be hard-wired to an aircraft or weapon in order to perform its test function. Also, certain hazards exist in the use of floating systems, in that static charges or lightning potentials may accumulate between the floating grounds and other accessible grounds such as the flight-/weather-deck or other portions of the ship structure, power line neutrals, the skin of the weapon system under test, etc..

Examples of isolation techniques are shown in Figures 6-2 and 6-3. Figure 6-2 employs transformer isolation, while Figure 6-3 obtains isolation by optical means.

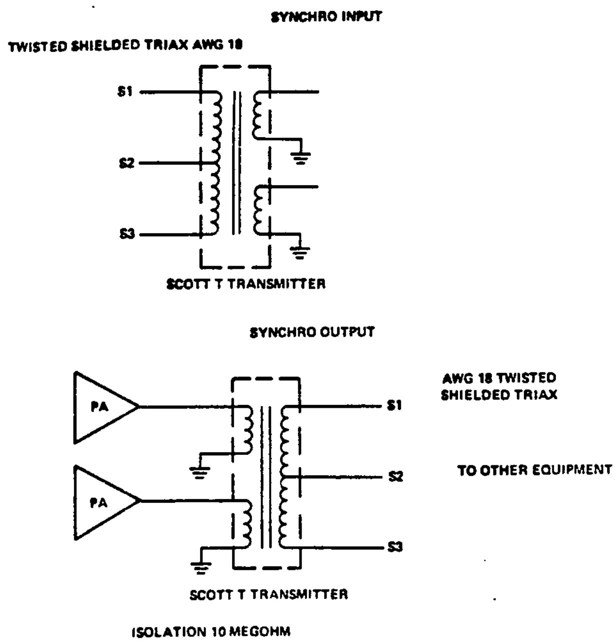

Figure 6-2
Typical Synchro Input, and Synchro Output Isolation Circuitry

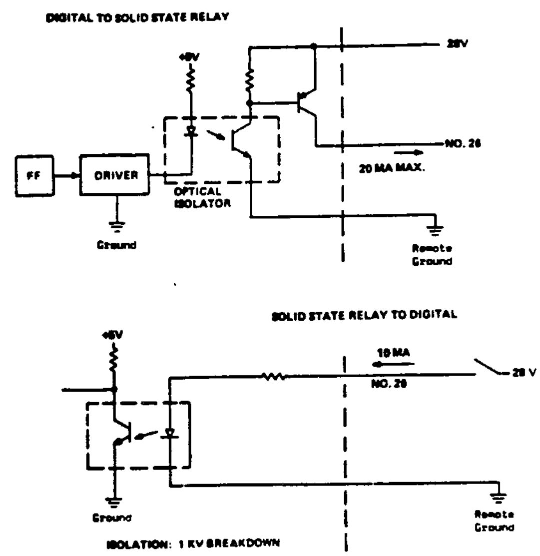

Figure 6-3
Digital to Solid-State Relay, and Solid-State Relay to Digital Isolation Circuitry

In a Single-Point Grounding System, a single physical point in the circuitry is defined as a ground reference point. All ground connections are tied to this point.

For multiple cabinet configurations, the cabinet and electronic circuit grounds are often kept separate, with the single-point grounding concept being used independently for each ground system. The interconnection between ground systems is then made only at the reference point. This isolates the rack or cabinets, and prevents circulating currents in one ground system from affecting the other [1].

Again, because of the appendage nature of most GSE, it would be extremely difficult to maintain a complete single-point grounding system. However, under certain conditions, it may be advantageous to use single-point grounding for the GSE and combination grounding for the GSE/weapon system.

At high frequencies (frequencies whose wavelengths approach equipment ground plane dimensions or cable lengths), single-point grounding systems are no longer practical. Therefore, it is important for the equipment/system designer to have an understanding of the frequency-susceptibility of his device, as well as the internal and environmental frequencies that are around to create EMI, before he finalizes his grounding system design.

The Multiple-Point Grounding System is one in which a ground plane is used instead of individual return wires for each circuit. The ground plane might be an equipment chassis, or a ground wire that is carried throughout a system, or (in the case of a very large system) building structural members.

The advantages of a multiple-point grounding approach are that circuit construction is easier, and it is the only way to avoid standing-wave effects in the ground system at high frequencies. Also, a large conductive mass can be selected for the ground plane. However, since multiple-point grounding creates many ground loops, the quality of the ground system becomes very important. Also, the ground plane must be carefully maintained, particularly with reference to corrosion, vibration, and mechanical damage, which can introduce high impedances into the ground system.[1]

6.3 CIRCUIT GROUNDING CONSIDERATIONS[1],[2],[3]

If the ground plane is fairly large, significant potential differences may exist between two points on the ground plane. These voltages must be considered when defining the permissible ambient interference level in the system, and when determining the expected signal-to-interference ratio of the signal transmission circuits.

The simplest and most direct approach to keeping potential differences introduced by the circuit ground plane to a minimum is to arrange circuit components physically so that ground return paths are short and direct, and have the fewest possible crossings of these paths. In this way, the inter-circuit coupling of the ground currents will be low and isolated.

The effect of ground potential can be cancelled by electrically isolating the circuits using a floating ground system. This method is especially effective at audio and low radio frequencies. Above these frequencies, however, its effectiveness progressively diminishes because, as the frequency of equipment operation increases, coupling paths appear that bypass the isolation transformers.

Band pass filters are another method of obtaining circuit isolation. These filters are only effective when the frequency of the interference energy is outside of the frequency band of the desired signal. Then the band pass filter will reject the interference signal across the load, causing the voltage drop to occur elsewhere in the circuit and not at the load. When using band pass filters, care must be taken to insure that "ringing" is not initiated by the undesired signal when it is processed by the filter.

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[1] See Sections 3.3 and 4.5.2 for information on corrosion control.

Differential or balanced circuitry can help reduce the effects of ground circuit interference. Since a differential circuit responds only to the potential difference between its input leads, the noise voltage at the source may be above ground potential by a considerable amount without degrading circuit performance. Figure 6-4 illustrates such a differential circuit. The input voltage, V_g, is the voltage to which the device responds. V_n, the interference voltage, is simultaneously impressed on both input leads but is balanced out in the input to the device because each input lead has the same impedance to ground. Thus, the device does not respond to the ground circuit signal. In theory, the ambient noise voltage is cancelled out, assuming the impedance of V_g, is zero ohms. In practice, there is always some unbalance in the differential device or associated circuitry, and some part of V_n will appear as a difference voltage across an equivalent resistance R. The noise voltage differential results in a reduced output signal-to-noise ratio. Figure 6-5 shows how the unbalance causes a portion of V_n, ΔV_n, to appear across the input terminals of the device.

Figure 6-6 summarizes the above discussion. It shows the ground circuit voltage, V_n, introducing an incremental voltage, ΔV_n, at the input to the differential circuit.

In transistorized digital circuits, the input and output impedance of the circuits are generally relatively low. This makes the circuits more susceptible to the effects of low impedance (magnetic) fields than to the effects of high impedance (electric) fields. One of the important parameters controlling interference due to a low impedance field is the loop area of the circuits causing and picking up the interference. By minimizing the loop area of these circuits, much of the interference problem can be eliminated.

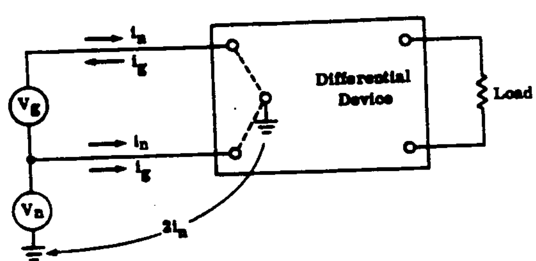


Figure 6-4 Schematic Diagram of a Differential or Balanced Circuit

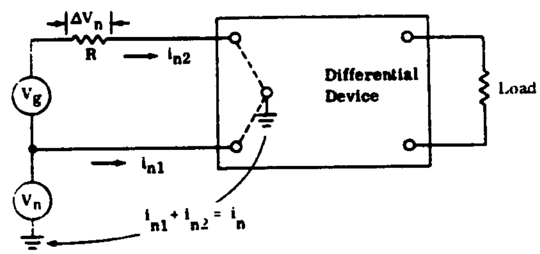


Figure 6-5 Effect of Unbalance in a Differential Circuit

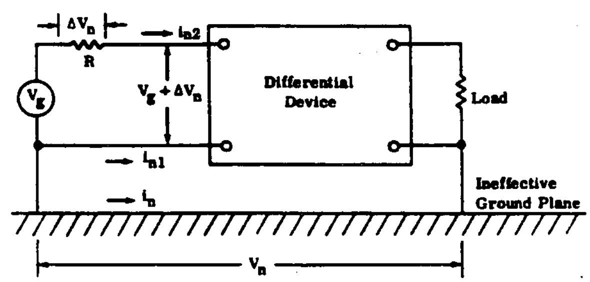


Figure 6-6 Common-mode Voltage Generated by Current Flowing in a Finitely Conducting Ground Plane

Where modular type construction is used, one method of minimizing the loop area of grounded circuits is to mount modules on a sheet of good conducting material, such as copper or aluminum, that is connected to the circuit ground as directly as possible. All intra-module wiring should be run as close to this sheet as possible. This technique can reduce the loop area of these circuits to an extremely small value, but the capacitive coupling of the circuits is increased. The effect of this increase in capacitive coupling on circuit operation should be considered if this technique is used.

The use of high impedance input or output circuit impedances when signals must be transmitted over even a few inches should, in general, be scrupulously avoided. Where the use of such signals cannot be avoided, the interconnecting lead must be shielded and the shield grounded at each end [4].

6.4 POWER SUPPLY CONDITIONS[1]

The ideal method of utilizing ac power supplies in GSE would be to have an individual power supply for each electronic circuit. This would give complete decoupling of circuits viewed from the power supply, and since this common conducting path would be eliminated, interference from this source would also be eliminated. However, this is obviously a very impractical solution in most cases, both from the standpoints of cost and space requirements. Other alternatives must be considered.

It is generally good practice to isolate power and signal grounds from each other. This will go a long way to minimizing the possible coupling of signals between these two major types of paths.

Groups of electronic circuits can be connected to one power supply based on a pre-arranged logical pattern. Critical circuits must be kept separated (i.e., critical with respect to interference susceptibility), and each given their own power supply when possible. If it is decided that these critical circuits cannot be kept separate, a voltage regulator can be introduced to act as decoupling component; Zener diodes can usually be used quite effectively in such applications.

Use of a large capacitor or an RC network at the load end of power supply leads can offer some degree of decoupling. This is particularly helpful in digital circuits to reduce the inductive effects of long power supply leads, but can introduce "ringing" at frequencies where the line and capacitor resonate. This condition can be circumvented by shunting the decoupling capacitor with a small RF bypass capacitor.

When deciding on which circuits to connect to which power supplies, the physical location of the electronic circuits must be considered. If no other factors are involved, all circuits fed by a given power supply should be grouped in the same area. However, in cases where various circuits will have to operate simultaneously (as in a synchronous digital system) some spatial separation may be advantageous to prevent coupling of radiated energy.

Alternating currents of high amperage are potential sources of interference coupling to adjacent lines. Transposed or twisted lines should be used for all ac power circuits. Routing of these lines should be away from susceptible lines. AC power circuits in which switching transients are expected may beneficially use shielding to contain the higher frequency components of the transient. Such shielding should be over the entire power wiring bundle and should be grounded at both ends.

A power system that drives electro-mechanical relays should not generally be used as a supply voltage for any electronic circuits because of the large amount of noise in this system due to relay operations. In addition to providing power relays, this power source may also furnish power for indicator lights, small motors, electro-magnetic clutches and brakes, and these devices may also be sources of noise. The power supply itself generally consists of a transformer-rectifier combination, or a motor generator set. In either case, the source is usually regulated to insure a constant voltage over a wide range of loads.

The primary cause of relay interference is due to the de-energizing of relays. When a relay circuit is opened, the collapsing magnetic field of the relay coil induces a large voltage of the opposite polarity across the coil. This induced voltage initiates an arc across the contacts that are breaking the circuit. The arc contains energy, in a wide band of frequencies, that may be transmitted as interference both by conduction and radiation. There are many standard arc suppression techniques, but even when these are used, a certain amount of noise will be conducted or radiated from the source.

Another cause of conducted interference can be the changes of current in the relay power supply busses due to the continually changing demands when the equipment is in normal operation. The conducted interference can be reduced by isolation of the relay circuits from the rest of the equipment.

To maintain isolation of the relay circuits and to prevent the system from floating and possibly reaching a high potential above ground, the relay system should preferably be grounded at a single point. Because of the unusually wide distribution of relay circuits, and because it is the point of maximum current, the ground connection should be made at the source.

In designing the power distribution system for relay systems where large currents are expected, it is important to consider the IR drops of the conductors in addition to their current-carrying capacities, although there is relatively low voltage involved. Also, the generation of varying magnetic fields due to the variation of the relay power supply currents must be considered. The effects of these variations can be minimized by running the supply and return leads as close together as possible throughout the system, thereby reducing the loop areas of the various relay circuits.

To minimize the possibility of dc supply lines coupling interference to other circuits, and to protect them from receiving interference from other lines, the best procedure is to use a twisted-pair for the supply line and its return. Shielding offers little advantage on lines of this low impedance. The routing should be away from ac power and control lines. When possible, run separate supply lines from the various circuits using the supply to the supply output. Do not share supply return lines with other circuit returns.

6.5 PRIME POWER CONSIDERATIONS[5]

There are two primary concepts regarding power returns. These are ground or structure common return and wired return. Although the GSE designer will probably have no control over the prime power configuration into which his equipment must work, he should nevertheless be familiar with these two concepts.

In the ground or structure common concept, one side of the power system is grounded at the power source, and all loads use the vehicle frame or structure as the return conductor. In three-phase connected ac systems, the neutral is grounded and all single phase loads use the structure for the return circuit. The principal advantage of this system is the reduced weight resulting from the elimination of a great many heavy power return wires.

The disadvantage of the common return system is that it does not distribute the power efficiently. The flow of currents through structure produces voltage drops in the structure. These voltages are normally small compared to the operating level of the power system, but are large compared to the operating level of electronic systems. This can create potential interference problems in any electronic system using the structure as a power return. Even systems using structure as a ground plane for shield grounds are subject to induced voltages in susceptible circuits.

In the wired return concept, all systems are grounded at one point only and have wires for all return circuits. The wired return system eliminates the vehicle structure as an impedance common to all systems and therefore eliminates the complex ground loops which exist with the common return system.

6.6 CABLING CONSIDERATIONS[5]

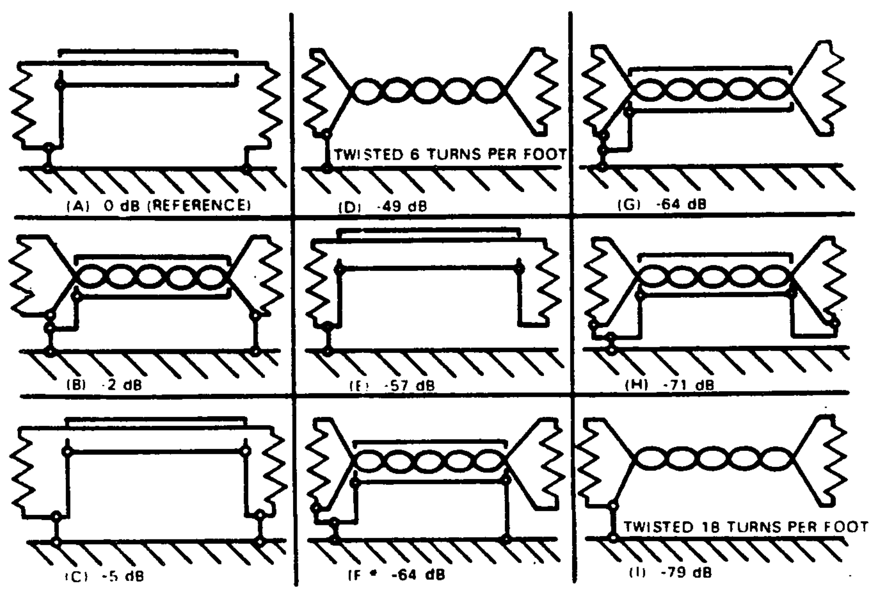
The problem of EMC in a complex electrical or electronic system is in many cases dependent on the treatment of the shielding and the grounding of the shields of interconnecting leads. Injudicious application of a grounded shield to a wire may cause coupling problems that otherwise would not exist.

Grounding of the shields may be accomplished as single-point or multiple-point grounding. Factors that influence the selection of single-point or multiple-point grounding include the interference signal frequencies involved, the length of the transmission line, and the relative sensitivity of the circuit to high- or low-impedance fields.

a. Single-point shield grounding. For multilead systems, each shield may be grounded at a different physical point as long as individual shields are insulated from each other. Single-point shield grounding is more effective than multiple-point shield grounding only for short shield length. 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 [4].

b. <u>Multiple-point shield grounding</u>. For L/λ ratios greater than 0.15, multiple-point grounding at intervals of 0.15λ is recommended, since the shield can act as an antenna that is relatively efficient at L = λ/4 when one end is grounded. When such grounding of the shield at intervals of 0.15λ is impracticable, shields should at least be grounded at each end. Multiple-point shield grounding is effective in reducing all types of electro-static coupling, so long as large ground currents do not exist. In general, multiple-point shield grounding solves most problems, but in audio circuits, single-point shield grounding may be more effective because of the ground current problem.

An interesting comparison of the magnetic interference susceptibility of cable-connected circuitry is shown in Figure 6-7. The evaluation was performed at 100 kHz, with the measurement parameters having to do with the type of cable used, whether or not the load is grounded, and how the cable shield is grounded. Among other results, the comparison shows (for the cases indicated) the disadvantage of returning the load to the ground plane, and the advantages gained using tightly twisted leads.



*PREFERRED CIRCUIT FOR HIGH FREQUENCIES

VALUES GIVEN ARE FOR CIRCUITS 1 INCH ABOVE GROUND PLANE BUT ARE ABOUT THE SAME FOR OTHER DISTANCES FROM GROUND PLANE.

Figure 6-7 Relative Susceptibility of Circuits to Magnetic Interference

The impact of cable shield and termination grounding on crosstalk has also been explored. Figure 6-8 shows a coaxial cable test setup for investigating single- versus multiple-point grounding, with single-point grounding always applied at the input ends of two 17-foot long cables[6]. The induced voltage in the receiving cable was measured for four conditions:

a. Sending and receiving cable resistor termination floating at the receiving end.

b. Sending and receiving cable resistor termination grounded at the receiving end.

c. Sending cable resistor termination grounded and receiving cable resistor termination floating at the receiving end.

d. Sending cable resistor termination floating and receiving cable resistor termination grounded at the receiving end.

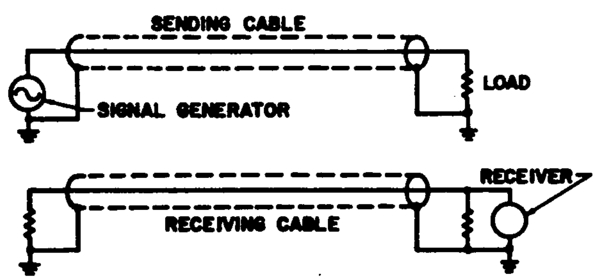


Figure 6-8 Test Setup for Coaxial Cable Crosstalk Measurements

Figure 6-9 shows some of the results of this test. It points out that specific rules on how to ground cannot be defined, and that the best grounding technique to employ is the one that works best for the particular circuitry involved. In this case, lowest crosstalk is obtained when the sending cable is single-point grounded and the receiving cable is multiple-point grounded.

A comparison between two of the coaxial curves of Figure 6-9 and a shielded twisted pair configuration is shown in Figure 6-10. The sending and receiving cables of the twisted pair were grounded as shown in Figure 6-7 (F), that is, the cable shields were grounded at both ends, while the signal return leads were grounded only at the input ends. The advantages of the latter configuration can be seen in Figure 6-10.

The performance of tests of this type indicate that, for low-level signals and low-impedance circuits where the distance between input connector and circuit input is small, i.e., an inch or two, the use of a twisted-pair alone may prove adequate. For long runs, the use of twisted-pair shielded wire becomes mandatory; this is true with unbalanced as well as with balanced output devices. In the balanced case, there should be no grounding except for the shield; in the unbalanced case, it is often necessary to ground one of the conductors as well as the shield. The shield and conductor should be grounded at the same point. Single-point grounding of the shield should be used for short runs and multiple-point grounding for long runs. If a shielded twisted-pair is part of a cable bundle and the bundle must go through a connector, three pins should be made available on the connector to provide insulated passage of the twisted-pair leads and the shield when it is not desirable to ground the shield at the connector.

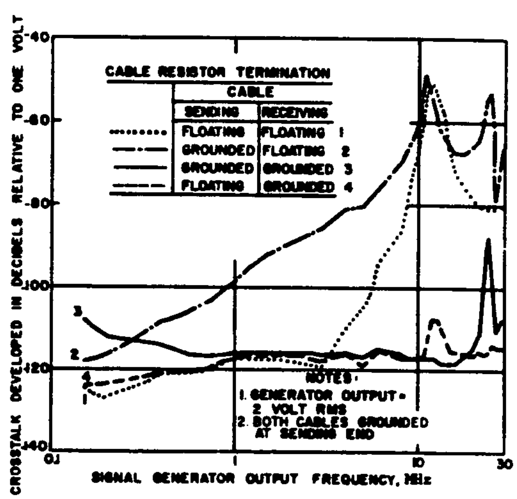


Figure 6-9 Measured RF Coupling Between Coaxial Cables

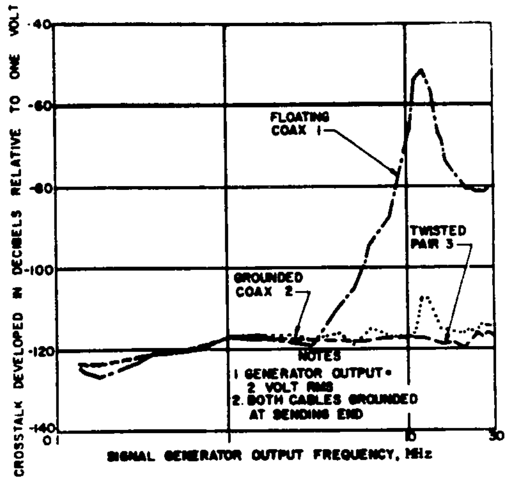


Figure 6-10 Measured RF Coupling in Shielded Cables

Signals of high level will, in general, not be bothered by susceptibility of circuits. Rather, it may be a source of interference to lower level signal lines. For this reason, and dependent on other characteristics of the signal, either a twisted-pair or a shielded lead should be used. Multiple shielding may be required if the signal is of sufficient level. Grounding should be applied at both ends to prevent electric field radiation from the cable.

When cable shields are grounded, good electrical contact to the shield mus be established. If possible, the shield should be grounded completely around the periphery of the connector shell. The use of pigtail grounding should be avoided on all cables, particularly those carrying signals above 1 MHz. When it becomes necessary to use a pigtail, the length should be minimized.

Figure 6-11 provides an indication of how the length of these ground leads can affect the coupling between two shielded cables [7].

There may be situations when multiple-shielded cable usage is considered necessary. For example, one suggestion has been made to use multiple-shielded cable for unbalanced input circuits with sensitivity thresholds better than 50 millivolts and operating frequencies between 5 kHz and 1 MHz [3]. When biaxial or triaxial cables are employed, single-point grounding should be used. An illustration of how to ground double-shielded cable is provided in Figure 6-12. Double or triple-shielded cable may be necessary for feeding high input or output impedance circuitry, particularly if the circuit is in a high electric field environment.

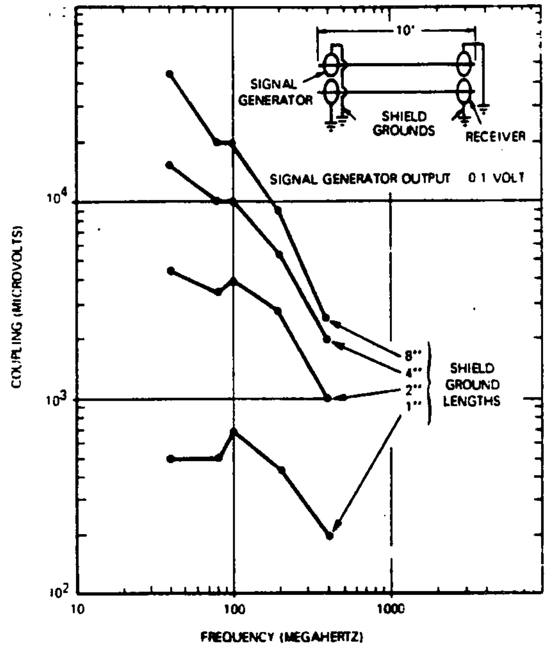


Figure 6-11 Cable Coupling Vs. Shield Ground Length

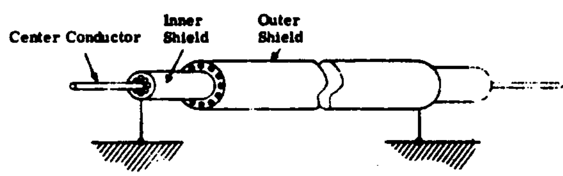


Figure 6-12 Example of Grounding a Double-Shielded Coaxial Cable

6.7 GROUNDING DESIGN GUIDELINES

Although the specific grounding philosophy to be employed in Avionics/GSE is very dependent on some of the detailed design objectives of that equipment, it is nevertheless possible to establish general grounding guidelines to follow. These are stated below, with the understanding that they should not be applied rigidly, and that alternate grounding methods should be tested before selecting a final design.

- o Remember that the ground system of concern includes the combined effects of both the GSE and the equipment under test.

- o Use single-point grounding when the dimensions of the circuit or component under consideration is small compared to the wavelength of concern (typically less than 0.15λ). Use multiple-point grounding when circuit or component dimensions exceed 0.15λ. When possible, ground large circuits or components at several locations, so that the separation between grounds is never greater than 0.15λ.

- There are occasions when transformer isolation and other isolation techniques can be used to prevent ground loops from occurring.

- Keep all ground leads as short and direct as possible. Avoid pigtails when terminating cables.

- It is advisable to maintain separate ground systems for signal returns, signal shield returns, power system returns, and chassis or case ground. They can be tied together at a single ground reference point.

- Ground reference planes should be designed so that they have high electrical conductivity, and so that they can be easily maintained to retain good conductivity.

- Circuits that produce large, abrupt current variations should have a separate grounding system, or should be provided with a separate return lead to the single-point ground. This will reduce transient pickup in other circuits.

- Grounds for low-level signals should be isolated from all other grounds.

- Never run supply and return leads separately, or in separate shields. A twisted pair is the best configuration for the supply bus and its return. Also, avoid carrying signal and power leads in the same bundle or in close proximity to one another. When signal and power leads must cross, make the crossing so that the wires are at right angles to each other.

- Use of differential or balanced circuitry can significantly reduce the effects of ground circuit interference.

- For circuits that operate below 1 MHz, tightly twisted pairs of wires (either shielded or unshielded, depending on application) that are single-point grounded offer the best approach to reduce equipment susceptibility.

- When coaxial cable is necessary for signal transmission, it appears that signal return through the shield and single-point grounding at the generator end (see Figure 6-7 (E)) offers certain advantages at the lower frequencies. However, other grounding arrangements should be considered. At high frequencies, multiple-point shield grounding is required.

- Low-level sensitive transmission lines may require multiple shields. Single-point grounding of each shield is recommended.

GROUNDING BIBLIOGRAPHY

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7.0 FILTERING

7.1 GENERAL

Even when a system has been well designed and incorporates proper shielding and grounding considerations, undesired energy can still be conducted through the system to degrade performance or cause malfunction. Filters can reduce this unwanted conducted energy to levels at which the system can function satisfactorily. Because of this role, filters are important in contributing to EMC.

An electrical filter can be defined as a network of lumped or distributed constant resistors, inductors, and capacitors or their equivalent, or any combination thereof, that offers comparatively little opposition to certain frequencies, or to direct current, while blocking the passage of other frequencies [1]. The design of filters is an art as well as a science since much depends on the judgment and techniques used by the filter design engineer.

The purpose of this chapter is to provide the Avionics/GSE design engineer with general filter design information that can be useful in the development of compatible avionics support units. However, some overall comments on EMI filtering are first considered appropriate.

First of all, it is important to point out that filters are often used only as stop-gap measures to problems that might have been resolved in a different way earlier in system design. For example, improved circuit linearity requirements might have obviated the need for a harmonic filter. Similarly, improved isolation of a relay circuit could have resolved a problem that later had to be attacked with transient suppression filters. An initial engineering effort should be made to design circuits that are inherently free of EMI. As part of this engineering effort, filters should be considered to limit the magnitude of EMI currents and to confine the currents to the smallest practical physical area, but the decision to use a filter should be made at a stage of development that provides maximum choice in alternative EMC approaches.

The impetus to the establishment of equipment filtering requirements (or shielding or grounding requirements for that matter) are the formal and informal specifications imposed on the designer. Thus, formal interference specifications such as MIL-STD-461 limit the amount of conducted interference that may be introduced on a powerline. Tolerable interference levels on critical internal equipment leads must be defined during an early stage of Avionics/GSE design so that circuit designers know the conditions their subassemblies must meet. The ability to comply with these specification limits can then be continuously assessed in the breadboard stage, etc.. Discrepancies between design objectives and actual EMI levels can often be most easily resolved by adding or modifying filters, particularly as portions of the design are frozen.

While filters are necessary and should be placed where needed, care should be taken to avoid redundant filtering caused by uncoordinated efforts of separate design groups. Redundancy usually occurs when each "black box" is required to meet an interference control specification regardless of final location. Although trade-offs must be made, there is no substitute for a well thought-out system EMC control plan. If formulated well ahead of the system design, filter duplication can be avoided.

The effectiveness of any EMI filter is greatly influenced by the impedance of its source and load terminations. Manufacturers of EMI suppression filters normally specify the filter characteristics with fixed source and load impedances (usually 50 ohms). The actual characteristics may be different when used in a circuit that requires other terminations, or when used in circuits whose impedances are not fixed. This aspect must be taken into consideration when designing, specifying or using EMI filters.

The uncertainty in knowing the impedances into which a filter will work when it is installed in an equipment makes it difficult to evaluate what its effectiveness will be in a given situation. This is a particular problem with regard to power line filtering. Initial steps have been made to establish the statistics of filtering effectiveness based on measured values of line and load impedance [20].

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 containing more series filter elements is indicated (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. The series element faces the low-impedance side of the system.

The basic characteristic used to describe filter performance is its insertion loss. Insertion loss is defined as the ratio of voltages appearing across the system terminals immediately beyond the point of insertion of a filter, before and after insertion [12], [14]. This ratio is expressed in decibels (dB) as follows:

$$\text{Insertion loss} = 20 \log [E_1/E_2] \qquad (7\text{-}1)$$

where:

E_1 = load voltage without filter.

E_2 = load voltage with the filter inserted.

A number of other filter characteristics are important. The input and output impedance requirements have already been cited. Others include the attenuation in the pass-band, the skirt falloff characteristic (rate at which the filter insertion loss changes as a function of frequency), steady-state and transient voltage ratings, etc.. These and other parameters will be discussed in more detail in Section 7.8 of this chapter.

7.2 FILTER DESIGN[1]

As indicated previously, filters are designed to attenuate at certain frequencies while permitting energy at other frequencies to pass unchanged.

Reflective filters do this by using combinations of capacitances and inductances to set up a high series impedance or a low shunt impedance for the interfering currents. Lossy filters do this by absorbing the interference energy.

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. The attenuation may vary in the stopband and is usually least near the cutoff frequency (the frequency at which a 3 dB insertion loss is obtained), rising to high values of attenuation at frequencies considerably removed from the cutoff frequency.

Filters can be grossly classified 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 will deal with these classes. Attenuation as a function of frequency for each of the classes is shown in Figure 7-1.

Design considerations for these four filter classes will next be presented. It should be pointed out that the Tables and Nomographs chapter of this design guide (Chapter 8) contains several normalized graphs and nomographs for convenience in making insertion loss calculations or defining filter parameters for selected filter types.

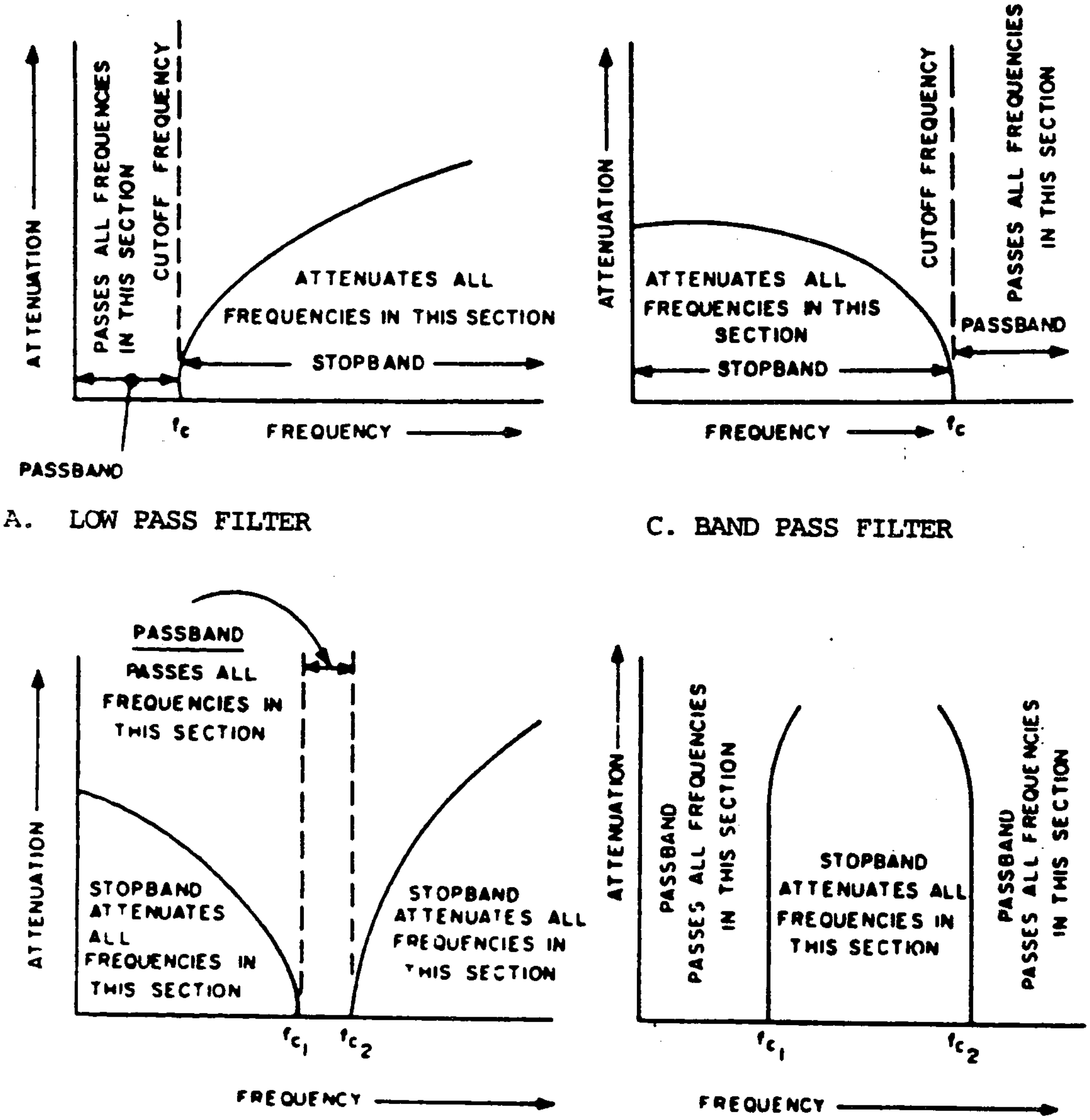
7.2.1 Low-Pass Filters

EMI control usually requires filters of the low-pass type. Power line filters are low-pass filters that pass dc or power frequency currents without significant power loss, while attenuating signals above these frequencies. Filters incorporated in amplifier circuits and transmitter output circuits are usually of the low-pass type so that the fundamental signal frequency can be passed while harmonics and the spurious signals are attenuated.

7.2.1.1 Shunt Capacitive Filters, and General Capacitor Characteristics

The simplest low-pass EMI filter is a shunt capacitor connected from the interference-carrying conductor to ground. It serves to bypass HF energy, as indicated in the ideal representation of Figure 7-2.

1. Except for ferrite filters, only lumped constant filters are considered in this design guide.



A. LOW PASS FILTER

C. BAND PASS FILTER

B. HIGH PASS FILTER.

D. BAND REJECT FILTER

Figure 7-1 The Four Basic Classes of Filters

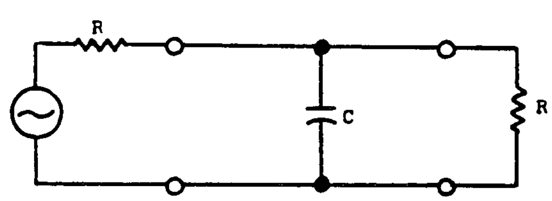


Figure 7-2 Capacitor Low-Pass Filter

Under these circumstances, the insertion loss of a shunt capacitor filter is defined [1], [3] by the relationship:

$$IL(dB) = 10 \log (1 + F^2) \qquad (7\text{-}2)$$

where:

$F = \pi fRC$

f = frequency, in Hertz

R = driving or termination resistance, in ohms

C = filter capacitance, in farads

An actual capacitor incorporates both resistance and inductance. These effects are due to such factors as the foil inductance of the capacitor plates, lead inductance, foil resistance, and lead-to-foil contact resistance.

The variations in these inductive and resistive effects depend upon the type of capacitor. 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 also a source of radio noise as the dielectric punctures and self-heals by burning away the metal film. This effect is indicated by the switch in the equivalent circuit shown in Figure 7-3. The standard wound aluminum foil capacitor may be employed as a radio frequency bypass in the frequency range up to 20 MHz. Its useful frequency range of operation is a function of capacitance and lead length. Its equivalent circuit is the same as that of the Metalized paper capacitor, but R_s and S of Figure 7-3 are not in the circuit.

R = Lead-to-Foil Contact Resistance
R^1 = Resistance of Metallized Foil
L = Lead Inductance
C = Capacitance
L^1 = Foil Inductance
R_s = Short Circuit Resistance
S = Short Circuit due to Voltage Puncture

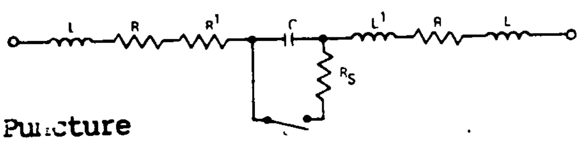


Figure 7-3 Metalized Capacitor Equivalent Circuit

As a result of these inductive effects, a capacitor will exhibit a capacitive reactance at low frequencies, and this situation will be maintained until its self-resonant frequency is reached. Above this frequency, the capacitor behaves like an inductive reactance. This effect is illustrated in Figure 7-4. Also note the effect of changing capacitor lead length on this self-resonant frequency.

Mica and ceramic capacitors of small values are useful up to about 200 MHz. A capacitor of flat construction, if the capacitor plates are round as in a ceramic disc capacitor, will remain effective to higher frequencies than one of square or rectangular construction.

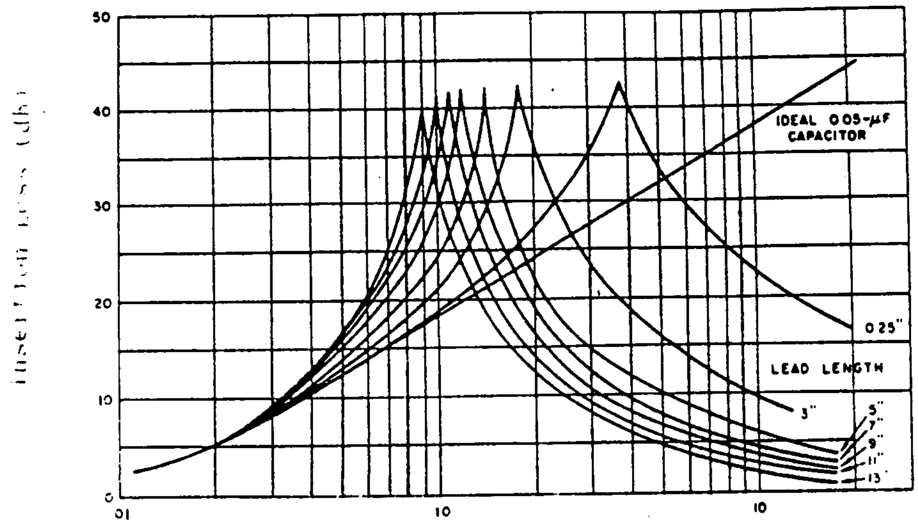


Figure 7-4 Insertion Loss of an 0.05-μf Aluminum Foil Shunt Capacitor

Other factors must be considered in selecting ceramic filter capacitors. 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. The dielectric composition can be adjusted to obtain a desired characteristic such as negative temperature coefficient, or minimum size. 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 negative to the extent that 50 percent capacity exists at full operating voltage, and full ambient temperature may cause an additional sizable reduction in capacity. Also, from the time of firing of the ceramic, the dielectric constant of the materials used may decrease; after 1000 hours, the capacitance may be as low as 75 percent of the original value. The designer should make ceramic capacitor selection based on required capacity under the most adverse operating conditions, and taking into account aging effects.

Capacitors of short-lead construction, and feed-through capacitors, are three-terminal capacitors designed to reduce inherent end lead inductances. Figure 7-5 shows the construction of these three-terminal types [4]. In each case, the inductance of the lead is not included in the shunt circuit. The wound foil short-lead capacitor is made with an extended foil type construction so that each plate of the capacitor can be soldered to a washer-shaped terminal. One washer is, in turn, soldered to the center lead, while the other is soldered to the case that is the ground terminal.

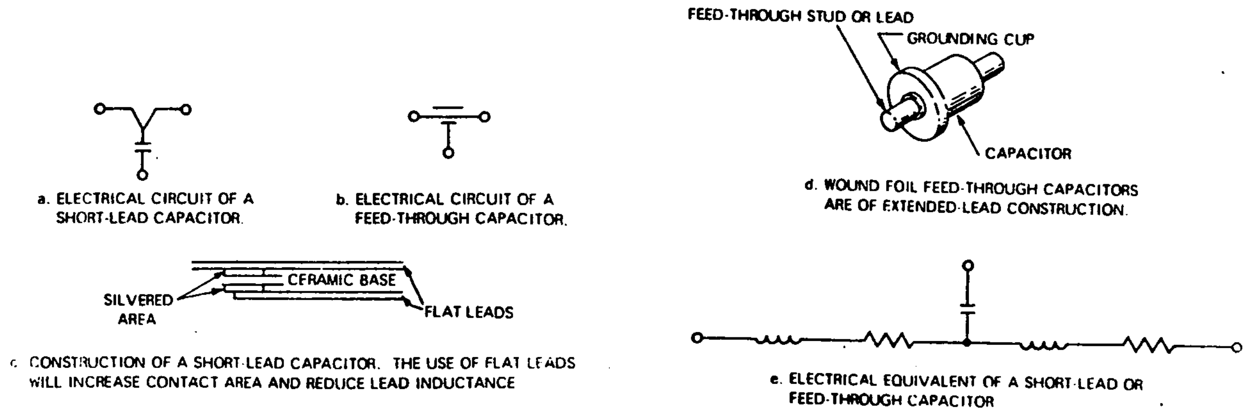


Figure 7-5 Three-Terminal Capacitor Construction

Theoretical insertion loss of three-terminal capacitors is the same as for an ideal two-terminal capacitor. However, the insertion loss of a practical three-terminal capacitor follows the ideal curve much more closely than does a two-terminal capacitor. The useful frequency range of a feed-through capacitor is improved further by its case construction, enabling a bulkhead or shield to isolate the input and output terminals from each other.

While the short-lead construction capacitor is ideally suited for EMI suppression in the frequency range of 1 to 1000 MHz, feed-through capacitors are available with a resonant frequency well above 1 GHz. The feed-through current rating is determined by the stud diameter. Figure 7-6 shows the insertion loss characteristics of typical feed-through capacitors, while Figure 7-7 indicates the construction details of a feed-through unit.

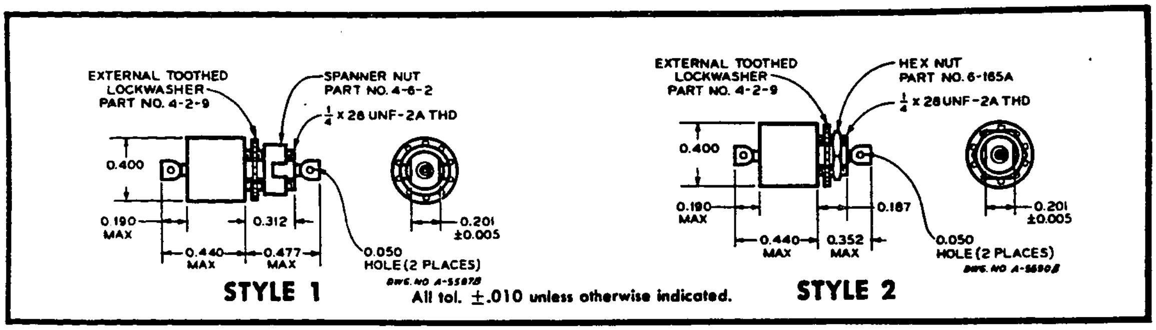
Capacitor selection for shunt capacitive filters, or any other filter application, is determined in part by the voltage, temperature, and frequency range in which the filter must operate. For 28 V dc applications, capacitors rated at 100 working volts dc (WVDC) are adequate. Metalized mylar capacitors offer the most compact design and good reliability. Their dissipation factor is very low, and lead length can generally be kept short to improve high frequency performance.

Wet-type electrolytic capacitors are used for dc filtering and sometimes in EMI filters. They are single polarity devices, and their high dissipation factor or series resistance 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.

If a large value of capacitance is required in a small space, tantalum capacitors may be considered. Because tantalum capacitors are electrolytics, they are more sensitive to over-voltages, and are damaged by reverse polarity. The dissipation factor is considerably higher than for mylar or paper capacitors, and high frequency characteristics are poor. A large tantalum capacitor reaches its minimum impedance at 2 to 5 MHz or less, depending upon construction and capacitance value.

Capacitors for 120 V ac applications should be rated at 400 WVDC and be suitable for ac use. A unit of mylar and foil or of paper-mylar and foil is recommended. Dissipation factor is low and high frequency performance is good. For 240 V ac applications, an oil-impregnated paper and foil unit is recommended.

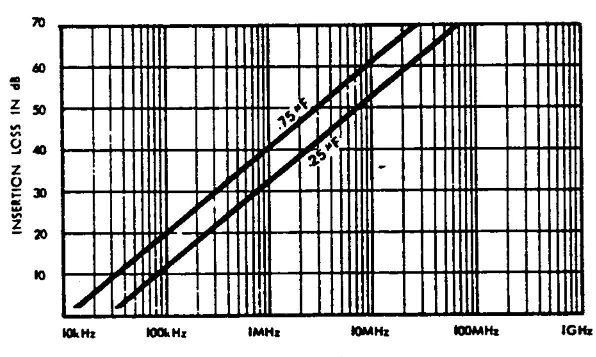
If good capacitor performance is to be expected above about 50 MHz, it is necessary to use designs incorporating feed-through techniques. As noted previously, lead inductance in a feed-through capacitor is not part of the shunt circuit, so that, compared to capacitors with leads, its insertion loss is not degraded as rapidly with increase in frequency.



NOTE: Style 1 units are supplied with spanner nut and external toothed lockwasher, as shown above. If hex nut is preferred, add the letter "X" to the basic catalog number when ordering; i.e., 7JX2103X.

SPECIFICATIONS

Catalog No. Style 1	Catalog No. Style 2	Capacitance GMV (μF)	Max. D-C Ohms	Current (Amperes)	Voltage Rating
7JX2103	7JX2503	0.75	0.01	10.0	100VDC @ 85°C 50VDC @ 125°C
7JX2105	7JX2505	0.25	0.01	10.0	250VDC @ 85°C 150VDC @ 125°C
5JX3102	5JX3502	0.25	0.01	5.0 7.0	125VAC, 400Hz @ 125°C 125VAC, 60Hz @ 125°C



Courtesy of Sprague Electric

Figure 7-6 Typical Feed-Through Capacitor Characteristics

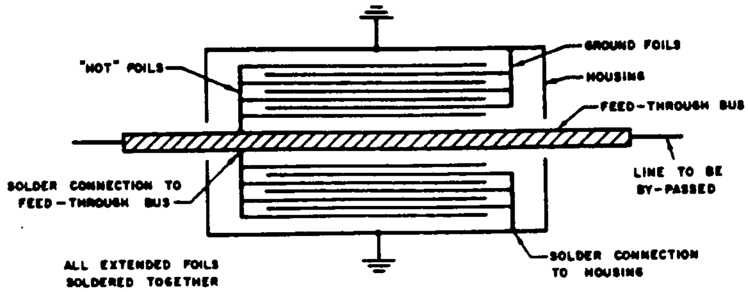


Figure 7-7 Construction of Feed-Through Capacitor

7.2.1.2 Series Inductive Filters, and General Inductor Characteristics

Another simple form of low-pass filter is an inductor connected in series with the interference carrying conductor. It is ideally represented in Figure 7-8. In practice, its insertion loss can be defined [1],[3] by the relationship:

$$IL(dB) = 10 \log (1 + F^2) \tag{7-3}$$

where:

$$F = \pi f \frac{L}{R}$$

f = frequency, in Hertz

R = driving or termination resistance, in ohms.

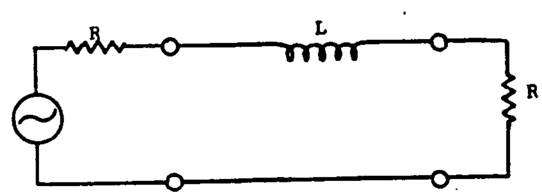


Figure 7-8 Inductor Low-Pass Filter

In practice, an inductor exhibits inductive reactance only until its self-resonant frequency is reached. Above self-resonance, it appears as a capacitive reactance, with the interwinding capacitance becoming 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 (number of turns multiplied by the peak current) should not drive the core to more than 50 percent of magnetic saturation.

The choice of core materials is determined by operating frequency and current rating. Powdered iron cores can be used for all dc applications and for most 60 Hz applications. For high current 60 Hz devices, and for all 400 Hz applications, molybdenum permalloy cores should be used. Ferrite materials can be considered when the current that will flow through the inductor will not saturate the core.

Stray or distributed capacitance in a filter inductor has two detrimental effects: EMI may be coupled from input to output of the filter via the capacitance, and the capacitance may cause the filter to become self-resonant at one or more critical frequencies. Windings should be placed on the coil so that input and output turns are separated as much as possible to keep stray capacitance low. 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.

Coil loss resistance is a measure of all power losses, hysteresis losses, and frequency-dependent absorption losses in the core. Loss resistance increases with frequency because of skin effect in the conductor, and due to changes in core loss with frequency. An increase in loss resistance represents an increase in attenuation in the filter passband. Losses in the core are not particularly detrimental, except that the insertion loss in the passband must be kept low.

7.2.1.3 Low-Pass "L" Section Filters

A primary disadvantage of single element filters is that their out-of-band falloff rate is only 6 dB per frequency octave (20 dB per decade). By combining both a shunt capacitor and a series inductance single element filter into an "L" configuration, a falloff rate of 12 dB per frequency octave can be obtained.

The two possible representations of low-pass "L" section filters are shown in Figure 7-9. In 7-9a, the capacitor shunts the source impedance, while in 7-9b, the capacitor shunts the load impedance. The insertion loss for the "L" section filter is independent of the direction of inserting the "L" section into the line, if source and load impedances are equal. When source and load impedance are not equal, the greatest insertion loss will usually be achieved when the capacitor shunts the higher impedance.

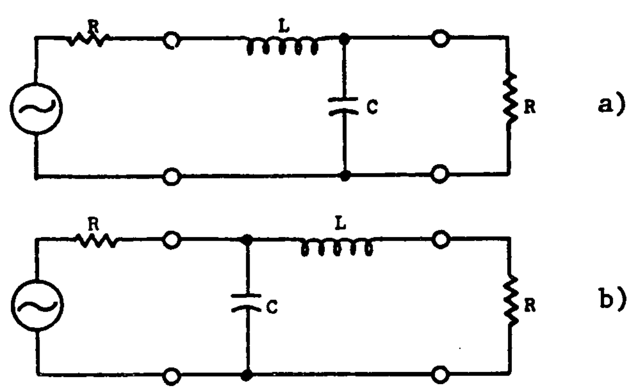


Figure 7-9 Low-Pass "L" Section Filter

The insertion loss of an "L" section lumped-constant network into 50 ohm resistance source and load impedances is [3]:

$$IL(dB) = 10 \log \left(1 + \frac{F^2 D^2}{2} + F^4\right) \tag{7-4}$$

where:

$$D = \frac{1 - d}{\sqrt{d}}$$

$d = L/CR^2$ = damping ratio

$$\omega_o = \frac{\sqrt{2}\,(R)}{L} = \frac{\sqrt{2}}{RC} \quad (\text{if } d = 1)$$

$$\omega_o = \sqrt{2/LC} \quad (\text{if } d \neq 1)$$

$F = \omega/\omega_o = f/f_o$ = normalized frequency

$\omega = 2\pi f$ = angular frequency (radians/second)

The "damping ratio", d, relates the magnitudes of the filter elements to the magnitudes of the source and load impedances. It is defined so that setting d equal to one (ideal damping) results in the elimination of the squared frequency term from the insertion loss equation and produces an abrupt transition from the pass-band to the stop region. The equations for Butterworth filter designs are obtained when d is set equal to one.

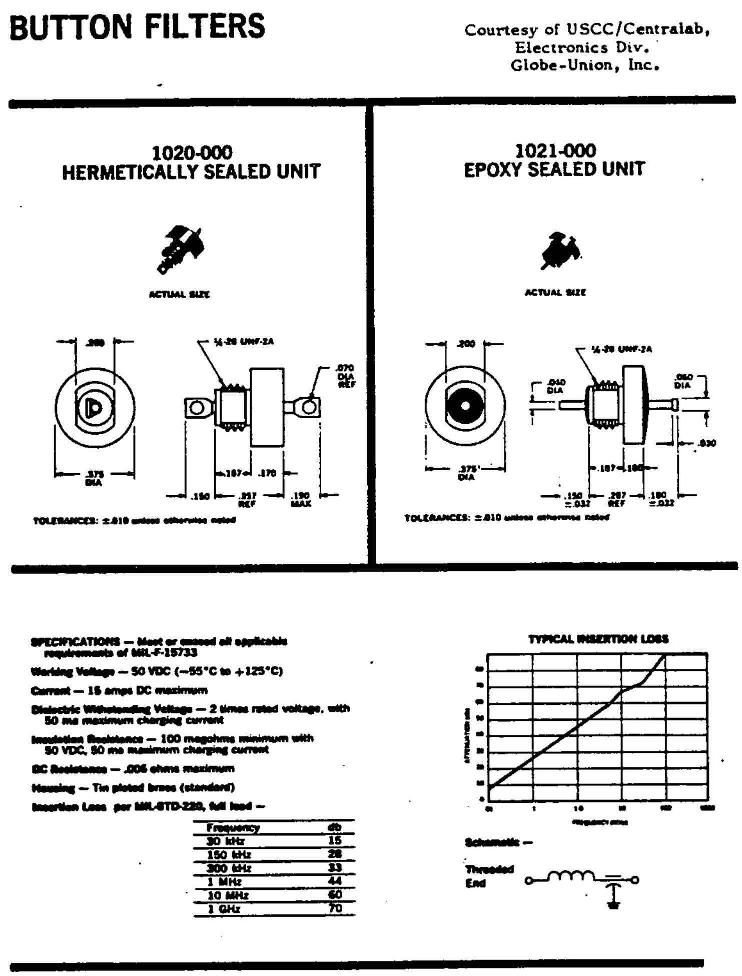
Values of d less than one result in insertion loss curves identical to those obtained when d is greater than one. That is, for two element filters:

$$IL\ (\text{for } d = n) = IL\ (\text{for } d = 1/n), \text{ for } n = \text{any real number} \tag{7-5}$$

The insertion loss of a two-element filter is not changed when it is "turned around" so that the source and load terminals are transposed, so long as the source and load impedances are equal.

The physical size of an "L" section filter depends upon insertion loss requirement, current rating, and voltage rating, with the first two usually predominant. The "L" section type of filter may give poor high frequency attenuation because of stray inter-turn capacitance. In some cases, the "L" type may resonate and oscillate when excited by transients.

Figure 7-10 provides an example of a commercially available "L"-section low-pass filter. This type of configuration enables a unit to be manufactured that can maintain an adequate rejection level to 1 GHz.



BUTTON FILTERS

Courtesy of USCC/Centralab,
Electronics Div.
Globe-Union, Inc.

SPECIFICATIONS — Meet or exceed all applicable requirements of MIL-F-15733

Working Voltage — 50 VDC (−55°C to +125°C)

Current — 15 amps DC maximum

Dielectric Withstanding Voltage — 2 times rated voltage, with 50 ma maximum charging current

Insulation Resistance — 100 megohms minimum with 50 VDC, 50 ma maximum charging current

DC Resistance — .006 ohms maximum

Housing — Tin plated brass (standard)

Insertion Loss per MIL-STD-220, full load —

Frequency	db
30 kHz	15
150 kHz	28
300 kHz	33
1 MHz	44
10 MHz	60
1 GHz	70

Figure 7-10 Example of Commercial "L" Section Filter

7.2.1.4 π-Section Filters

The "π-" section filter is the most common type of radio frequency interference suppression network. Figure 7-11 shows the circuit of the π-section filter. Advantages are ease of manufacture, high insertion loss over a wide frequency range, and moderate space requirements. Although voltage rating must be considered, current rating and attenuation are the most important factors in determining the size of the filter.

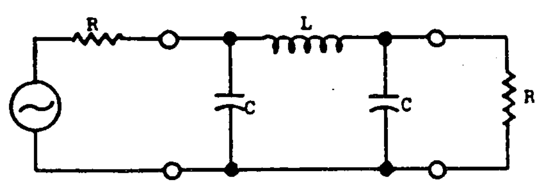


Figure 7-11 Low-Pass π Filter

The insertion loss of a lossless π-section network operating with equal source and load impedance is [3]:

$$IL\ (dB) = 10 \log (1 + F^2 D^2 - 2F^4 D + F^6) \qquad (7\text{-}6)$$

where:

$$D = \frac{1 - d}{\sqrt[3]{d}}$$

$d = L/2CR^2$ = damping factor

$\omega_o = \sqrt{2/LC} = 2R/L = 1/RC$ (if $d=1$)

$\omega_o = \sqrt[3]{2/RLC^2}$ (if $d \neq 1$)

$F = \omega/\omega_o$

$\omega = 2\pi f$ = angular frequency

Unlike the "L" section filter case, overdamping or underdamping of π-section (and T-section) filters result in entirely different effects. This is discussed further in Section 7.2.1.6 of this design guide.

A typical attenuation curve of a π-section filter has a slope of approximately 18 dB per octave; 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. Representative data on a series of commercial low-pass "π" filters are shown in Figure 7-12.

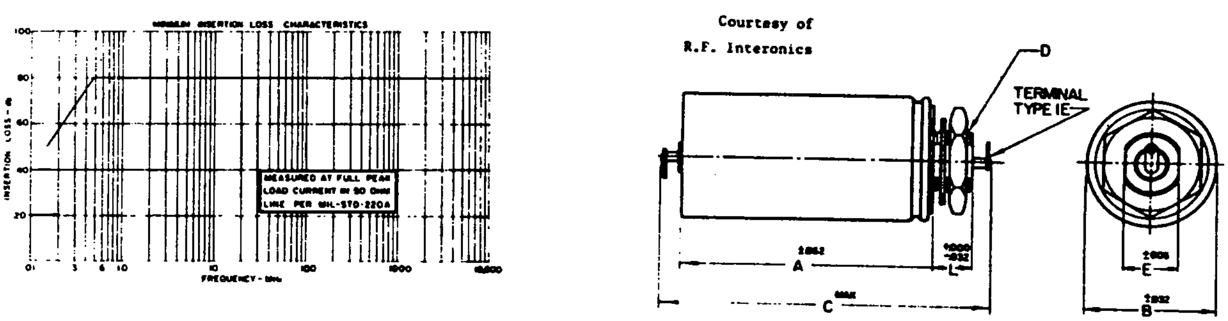


Figure 7-12 Representative π-Section Filter Characteristics

The multiple π-section filter (cascaded π-sections) has characteristics identical to those of the multiple L-section filter. The attenuation curve of the theoretical multiple π-section filter rises at a rate of 20 dB more per decade of frequency than does a multiple "L" filter of the same number of sections. Although this may not be a significant factor when three or more sections are used, it does provide a capacitive input at both ends of the filter that is sometimes advantageous. An extensive use for this type of network is as a power-line filter in large installations, and for shielded enclosures where high attenuation is needed at very low frequencies.

7.2.1.5 "T" Section Filters

The "L" type low-pass filter can also be improved by the introduction of another series inductor. This addition forms a "T" section filter, which consists of two inductors in series with a shunt capacitor connected from the junction of the two inductors to ground (see Figure 7-13). Insertion loss is given by:

$$IL\ (dB) = 10 \log (1 + F^2D^2 - 2F^4D + F^6) \qquad (7\text{-}7)$$

where:

$$D = \frac{1 - d}{\sqrt[3]{d}}$$

$$d = R^2C/2L = \text{damping factor}$$

$$\omega_o = \sqrt{2/LC} = R/L = 2/RC \text{ (if } d = 1)$$

$$\omega_o = \sqrt[3]{2R/L^2C} \text{ (if } d \neq 1)$$

$$F = \omega/\omega_o$$

$$\omega = 2\pi f$$

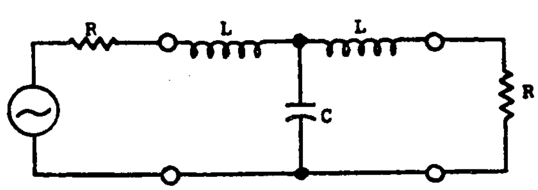


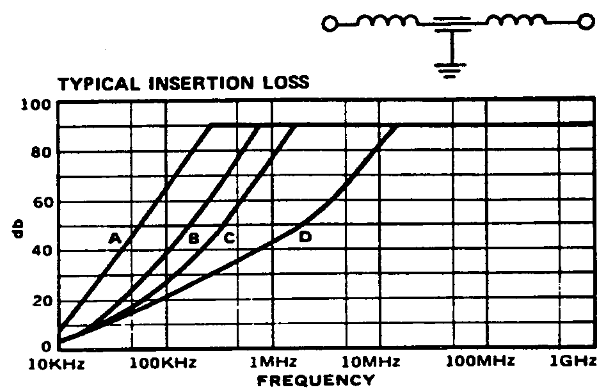
Figure 7-13 Low-Pass "T" Filter

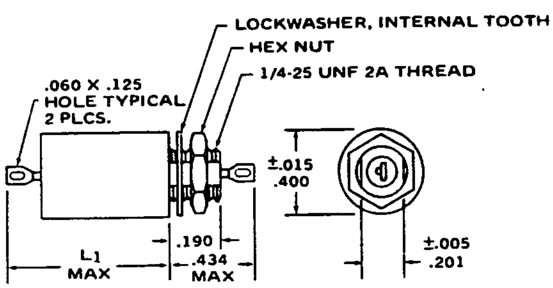
The "T" type of filter is a very effective form of the lumped-constant type of filter for reducing transient interference. Its major disadvantage is the requirement for two inductors, which under some circumstances may present a size penalty. It provides the same out-of-band falloff rate as the π-section filter, that is, 18 dB/octave (60 dB/decade) for a single section.

Figure 7-14 provides representative information on commercially available T-section low-pass filters.

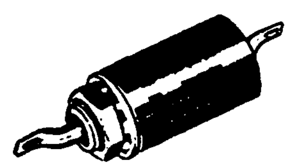
RATINGS

PART NO.	CURRENT AMPS	DCR AT 25°C OHMS	INSERTION LOSS CURVE	L1 MAX INCHES	L2 MAX INCHES
GF51F3A	0.10	3.20	A	.99	.94
GF51F3B	0.50	.66	B	.99	.94
GF51F3C	1	.36	C	.99	.94
GF51F3D	5	.022	D	.99	.94





1. Tolerance: ± 0.010 unless otherwise specified.
2. Case material and finish: steel, gold plated (optional).
3. Case marked with Genisco G and part number.
4. Typical weight: 5 to 12 grams.
5. Recommended torque during installation: 40 in. oz.
6. Each filter is supplied complete with lock washer and nut.



Courtesy of Genisco Technology Corp.

Figure 7-14 Representative Commercial Low-Pass T-Section Filter Characteristics

7.2.1.6 Insertion Loss Calculations for "π" and "T" Section Filters

The equations for the insertion loss of a T-circuit and a π-circuit as given by Equations (7-6) and (7-7) are of the same form, but differ with respect to the definition of the equation parameters. The equation has three modes of response. When d equals one, the response is optimally damped and is the ideal (Butterworth) response curve. When d is greater than one, the response is in an overdamped mode. When d is less than one, the response is in an underdamped mode. In the underdamped case, the curve has a maximum in band loss of:

$$IL = 10 \log (1 + 4D^3/27) \qquad (7\text{-}8)$$

at the frequency where F = D/3. A minimum loss point will also occur at the frequency where F = D.

7.2.1.7 Lossy Line Filters

While the input and output impedances of some filters can be expected to match their intended source and load impedances over a fairly broad frequency range, it is more often the case that such matches will not occur. For example, the input impedance of a powerline filter almost never achieves a match with the impedance of its associated powerline. As another example, a transmitter harmonic filter is generally designed to match the transmitter output stage over the fundamental frequency range, but not necessarily at its harmonic frequencies.

Because of such mismatch situations, there have been many cases when the insertion of a filter into a line carrying interference has actually resulted in more, rather than less, interference voltage appearing on the line beyond the point of its application [3]. This deficiency in all filters composed of low loss elements has led to the development of dissipative filters that take advantage of the loss-versus-frequency characteristics of magnetic materials such as ferrites.

Ferrites are inert ceramics containing granulated iron compounds. They are free of any organic substances, and are not degraded by most environments. Ferrites have both a large relative permeability (μ_r) and a large relative permittivity (ε_r). When $\mu_r = \varepsilon_r$ for a given ferrite the wave impedance of the material is equal to the wave impedance of free space, and an impinging EM wave is absorbed without reflection [39]. This occurs over a small relative bandwidth and is the reason for the loss-versus-frequency characteristics of ferrites.

One form of dissipative filter uses a short length of ferrite tube with conductive silver coatings deposited on the inner and outer surfaces to form the conductors of a coaxial transmission line. The line becomes extremely lossy at radio frequencies, that is, it has high attenuation per unit length in the frequency range where either electric or magnetic losses, or both, become large. An example of the performance of a lossy line filter of this type is shown in Figure 7-15.

Dissipative filters of this type are necessarily low-pass. One of the large uses of such filters is in general-purpose power-line filtering, in which the dissipative filter is combined with conventional low loss elements to obtain the necessary low cutoff frequency.

Another method of achieving a dissipative filter is by use of lossy beads. Tubular ferrite toroids offer a simple, economical method for attenuating unwanted high frequency noise or oscillations. One bead slipped over a wire produces a single-turn RF choke that possesses low impedance at low frequencies and moderately high impedance over a wide high frequency band.

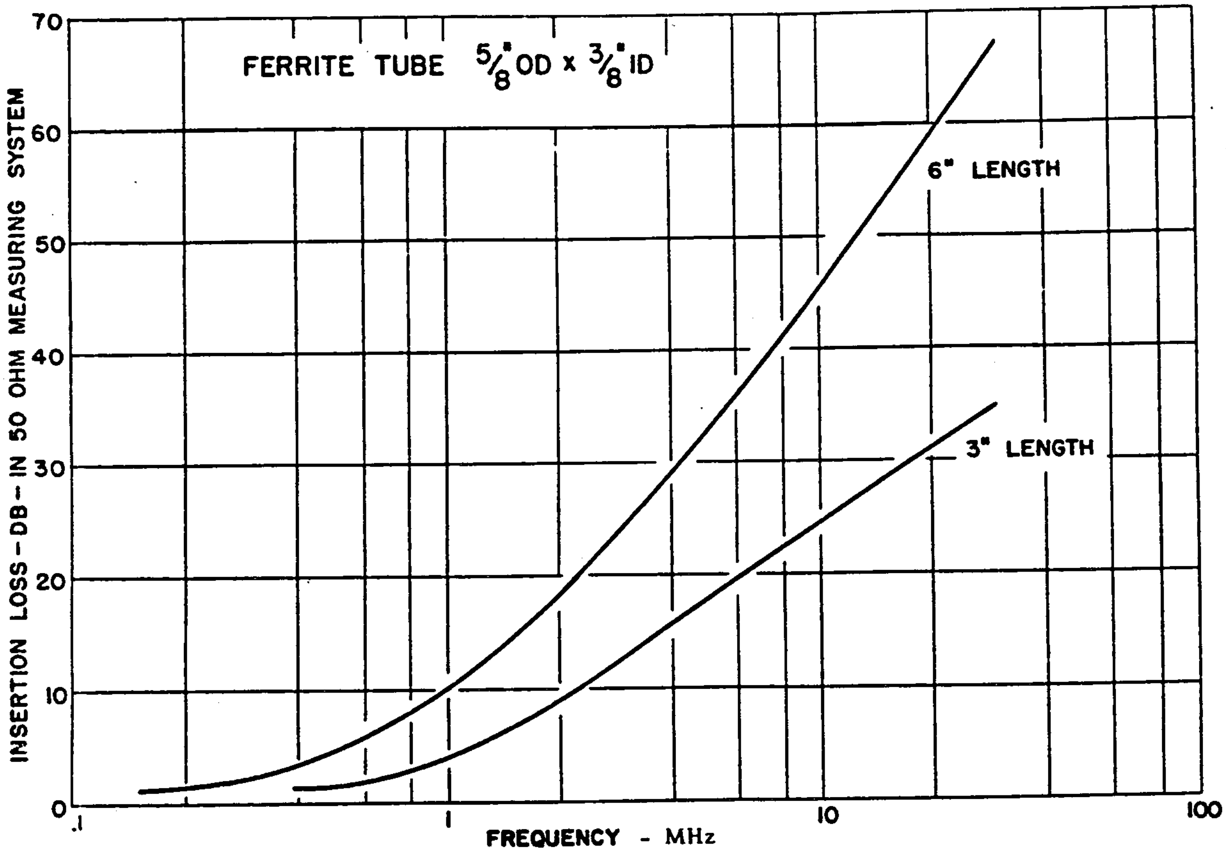
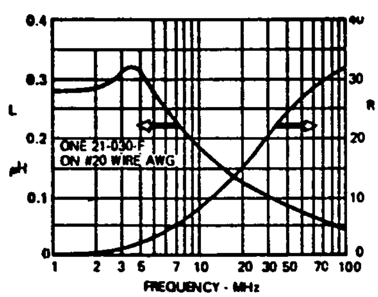


Figure 7-15 Insertion Loss of a Ferrite Tube Low-Pass EMI Filter

The presence of a ferrite bead on the wire causes a local increase of series impedance (largely resistive) presented to currents in the wire. Figure 7-16 illustrates the effects of one ferrite bead on a length of wire [4]. Adding more or longer beads provides additional units of series inductance and resistance in direct proportion. Extra turns of wire can be passed through the bead, increasing both resistance and inductance in proportion to the square of the number of turns. Because of distributed winding capacitance, this technique is most effective at the lower frequencies. Because of the high resistivity of ferrite beads, they may be considered insulators for most applications.



Typical values of equivalent series resistance and inductance attributable to ferrite beads.

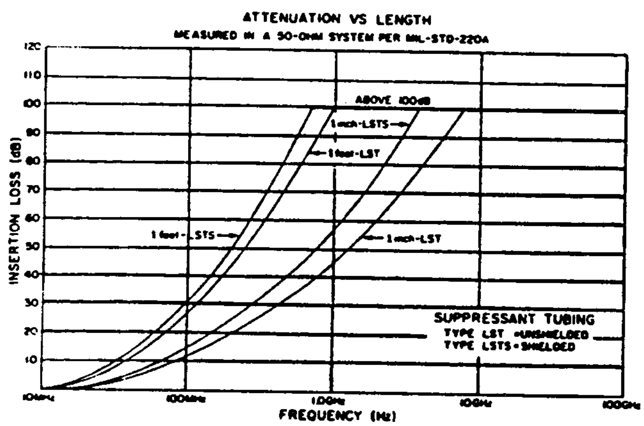
TYPICAL PROPERTIES	
Flux Density (B) at 5 O_e	2400 G
Coercive Force (Hc)	0.56 O_e
Hysteresis Factor (h/μ^2)	22×10^{-6}
Initial Permeability (μo)	450
Permeability (μ) at 250G	900
Resistivity - Ohm - cm	$\geq 10^7$
Curie Temperature	≥ 156 °C

Figure 7-16 Filter Characteristics of Ferrite Beads.

High amplitude signals below 50 MHz may cause some reduction in the suppression effect due to ferrite saturation. However, as long as only one turn links the core, fairly high currents can be tolerated using representative materials before saturation is approached. At saturation, inductance and resistance will be low, but will return to normal values upon removal of the high field.

Lossy line suppressant tubing also provides an efficient means for suppressing undesired EMI and other spurious signals. The tubing can be slipped over standard wire and cable and suppresses both radiated and conducted energy. It can be used in environments from -55 degrees to +250 degrees C without electrical or mechanical degradation. The tubing will provide shielding from low frequency electro-static interference and magnetic fields, and will not cause dc or low frequency ac losses. RF power handling capability is in excess of 10 watts per inch of tubing. Representative data on this type of tubing is shown in Figure 7-17.

ATTENUATION CURVES



SIZES AVAILABLE

Lundy Part No.	Inside Diameter (Inches)	Outside Diameter (Inches)	Tolerance on I.D. & O.D. (Inches)
LST-060	.060	150	±.004
LST-080	.080	.170	±.005
LST-100	.100	.200	±.006
LST-125	.125	.225	±.006
LST-150	.150	.250	±.007
LST-175	.175	.300	±.008
LST-200	200	.325	±.008
LST-225	.225	.350	±.010
LST-250	.250	375	±.010
LST-275	.275	.400	±.010
LST-300	.300	.450	±.010
LST-325	.325	475	±.010
LST-350	.350	.500	±.010
LST-375	.375	525	±.010
LST-400	.400	.550	±.010
LST-425	.425	575	±.010
LST-450	450	600	± 010
LST-475	.470	650	±.010
LST-500	.500	700	±.010

Figure 7-17 Typical Characteristics of Lossy Line Suppressant Tubing

Lossy line coaxial cable can also serve as a dissipative low pass filter. As an example, data on Raychem Electro Loss filterline cables is provided in Figure 7-18.

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, and offer low-pass filter performance as shown in Figure 7-19.

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

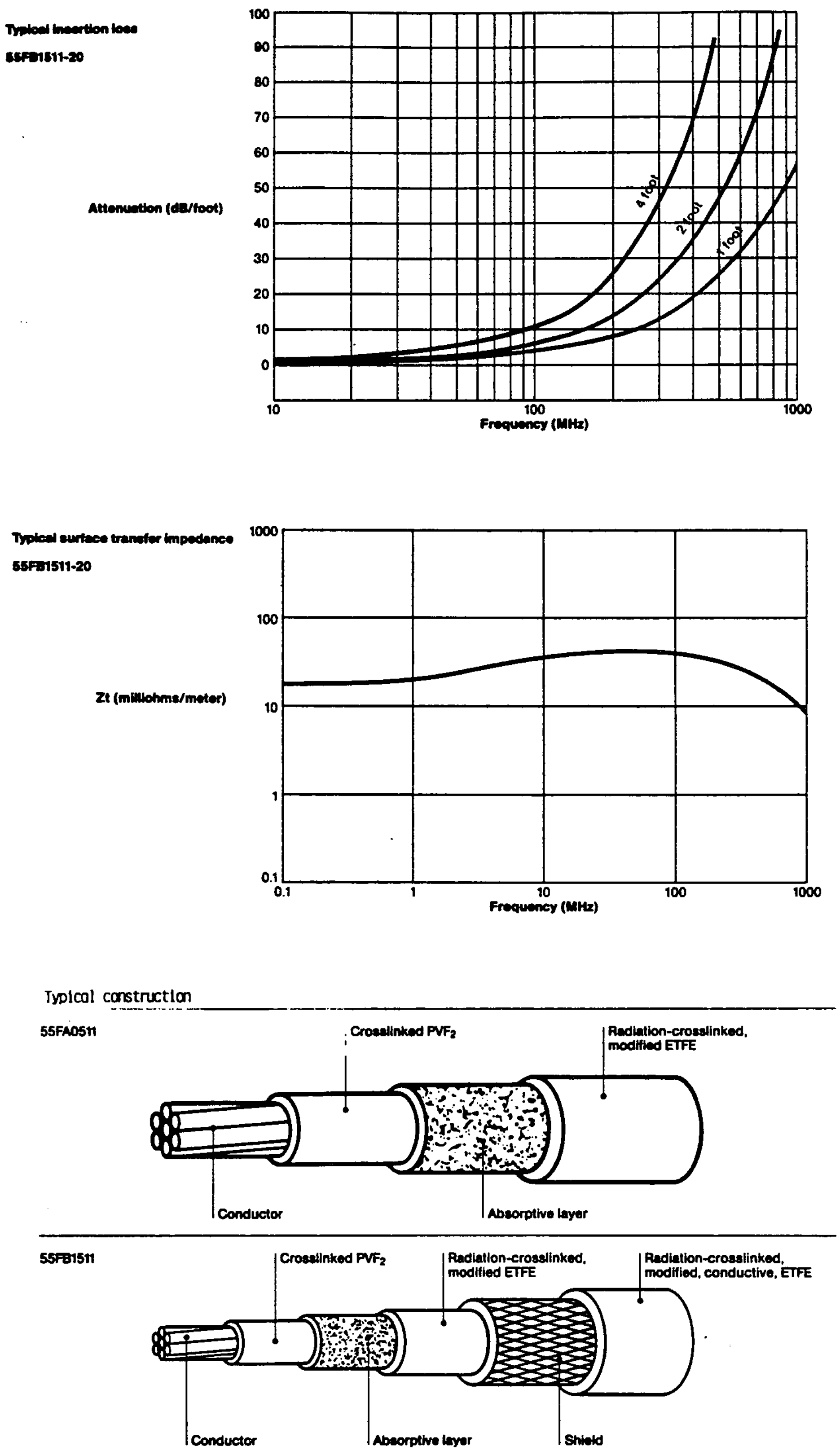
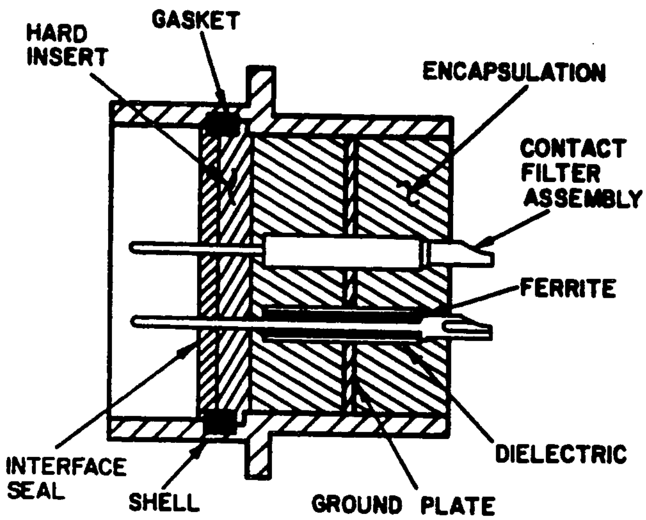
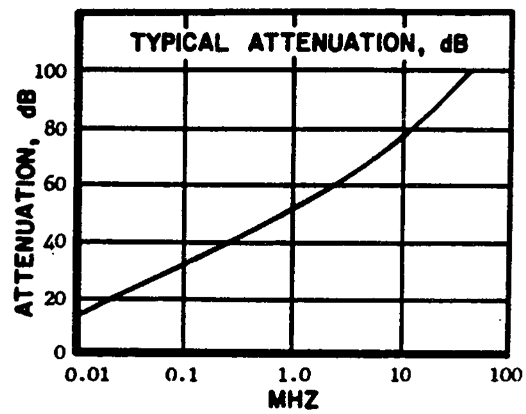


Figure 7-18 Typical Characteristics of Lossy Coaxial Cable





Minimum Attenuation from −55°C to +125°C and 100 MHz to 10 GHz .... 80 dB
D.C. Working Voltage (includes summation of the D.C. and low level A.C. superimposed peak voltages) .... 50 VDC
Dielectric Strength (for 5 sec. with charging current of 50 milliamperes maximum) .... 100 VDC
Feed Through Current (Nominal) D.C. and/or audio frequency RMS .... 7.5 Amperes
Insulation Resistance .... 250 Megohms
R.F. Current .... 0.25 Amperes
Operating Temperature Range .... -55 C to + 125 C
Capacitance (μf) .... 1μf Nominal

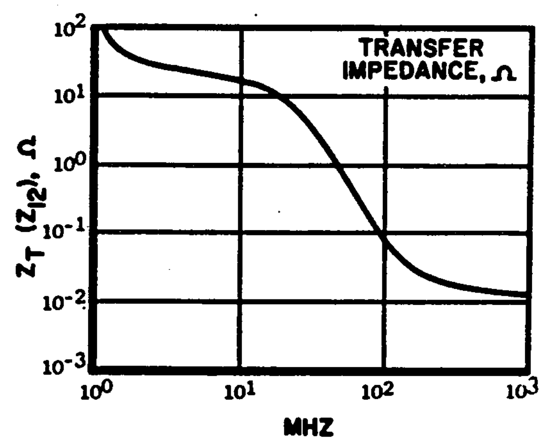


Figure 7-19 Typical Characteristics of Lossy Connector

band attenuation. An example of the improvement in stop-band attenuation can be gained by preceding a reactive filter with a lossy line section is illustrated in Figure 7-20.

Figure 7-20a 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 coaxial line whose dielectric space is filled with a 6:1 ratio of iron-to-epoxy dielectric material, the attenuation characteristic is altered to that shown in Figure 7-20b. The addition of the lossy section has increased the passband attenuation only slightly, but the stop band attenuation has been increased to greater than 60 dB.

When a lossy line section is used in cascade with a conventional low-pass filter, the passband insertion loss can be minimized by the proper choice of the dielectric material. However, there is always some passband loss introduced by the lossy dielectric. Such passband losses can be reduced by designing the reactive filter to have as wide a region as possible between the low-pass cutoff frequency and the first spurious passband, so that a minimum of lossy material is needed to provide the required stopband attenuation.

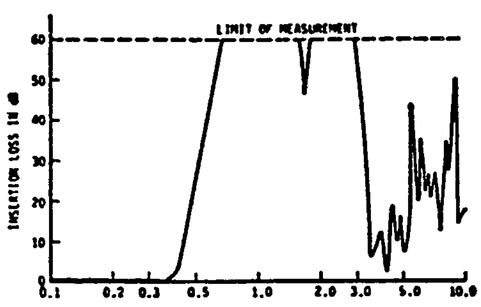


Figure 7-20a
Typical Low-Pass Filter Loss Characteristics, Low-Pass Filter Only

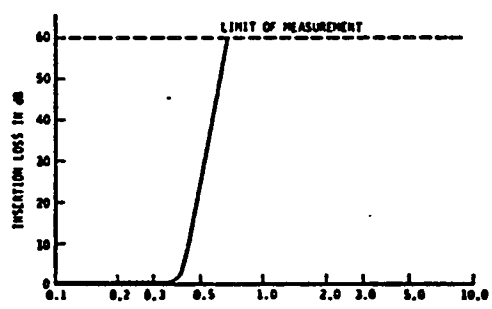
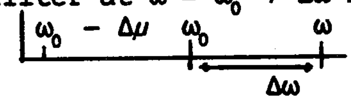


Figure 7-20b
Typical Low-Pass Filter Loss Characteristics, Plus Lossy Filter Section

7.2.2 High-Pass Filters

Although not as common as the low-pass type, high-pass filters also have an application in EMI reduction. In particular, such filters have been used to remove ac power line frequencies from signal channels and to reject particular lower 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 [5]. 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 with the same impedance terminations and cutoff frequencies by replacing each coil with a capacitor, and vice versa, and by replacing the element values by their reciprocals. Thus, 2 Henries become 0.5 Farad, 10 Farads become 0.1 Henry, etc. The attenuation given by the low-pass filter at $\omega = \omega_0 + \Delta\omega$ is now given by the high-pass filter at $\omega = \omega_0 - \Delta\omega$, 

, where ω_0 is the cutoff frequency.

For example, a Butterworth low-pass, π-section has the element values shown in Figure 7-21. The cutoff frequency is 10 kHz. The filter is shown transformed into a high-pass filter with the same input and output impedances, and the same cutoff frequency. A similar transform relative to a T-section filter is also provided.

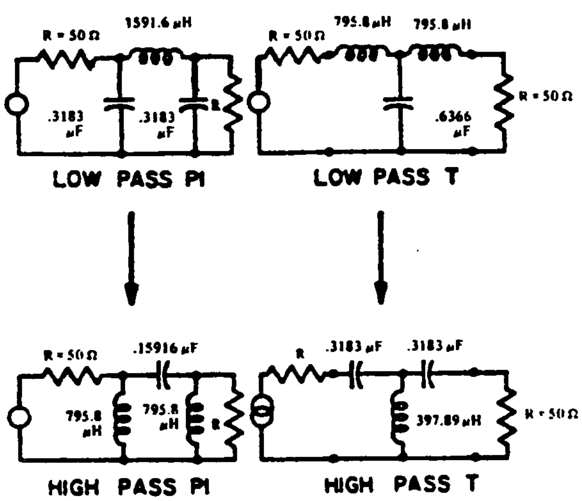


Figure 7-21 Low-pass to High-pass Transforms

A transformation can be made from a low-pass filter to a high-pass filter having different impedance terminations and/or cutoff frequencies by applying scaling factors during the process of making the transformation (see detailed transformation procedure in Chapter 7 Addendum #1). The scaling operations, if used before taking the reciprocals, are as follows:

$$L(\text{scaled}) = L(\text{original}) \cdot \frac{R(\text{scaled})}{R(\text{original})} \cdot \frac{W(\text{original})}{W(\text{scaled})}$$

$$C(\text{scaled}) = C(\text{original}) \cdot \frac{R(\text{original})}{R(\text{scaled})} \cdot \frac{W(\text{original})}{W(\text{scaled})}$$

7.2.3 Bandpass Filters

Each low-pass filter can also 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 an arbitrary frequency band and reject signals outside that band) can be readily established 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 3dB bandwidth and insertion loss as the bandpass filter.

b. In the case of single section L, T and π filters having 50 ohm input and output impedances, select a low-pass filter with the required attenuation using the two and three element filter design equations discussed in this 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 at 1.0 MHz, and a skirt roll-off rate of at least 15 dB/octave. The required impedance level is 50 ohms input and output. There is to be no ripple in the pass-band; that is, response should at all points be monotonic. 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. From Equation (7-6), it is found that the L and C values for such a filter are 160 μH AND 0.03 μF, using a cutoff frequency of 100 kHz and a damping factor of unity.

Each of the above components are next resonated at 1.0 MHz using the relationship:

$$f = \frac{1}{2\pi\sqrt{LC}} \tag{7-9}$$

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 Figure 7-22.

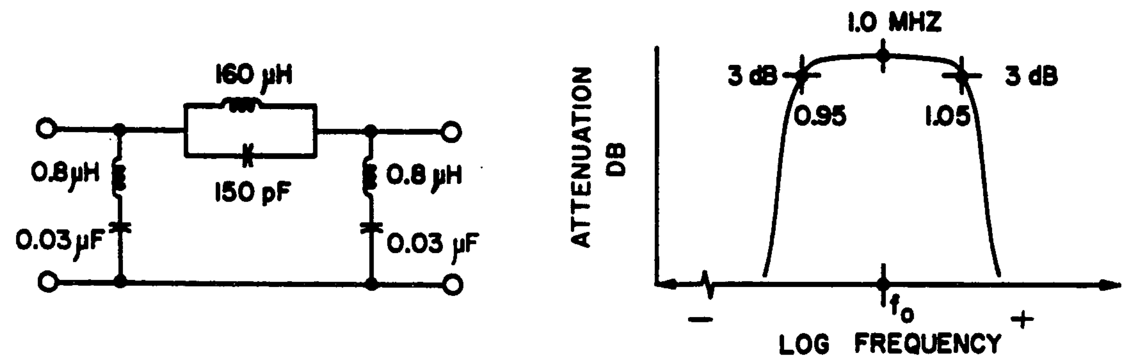


Figure 7-22 Example of Band-Pass Filter Design

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_o is a mirror image of the equivalent displacement below f_o. Alternatively if the attenuation at a frequency xf_o, is N dB, then the attenuation will be the same at f_o/x. For the example above, the band-pass filter cutoff frequencies are shown in Figure 7-22 as 0.95 and 1.05 MHz, since the logarithmic effect is not evident at frequencies close to the tuned frequency of the filter.

Transformation is thus the essential principle in the design of bandpass filters. It reduces the design to a procedure of specifying the element values of a low-pass filter section and transforming this section into a bandpass filter. In this transformation, a capacitor is added across each coil of a size to resonate the coil at f_o. A coil is added in series with each capacitor of the low-pass filter of a size to resonate the capacitor at f_o.

The above 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 pass-band can be tolerated, then a steeper descent into the attenuation band can be obtained. Chebyshev filters have a greater roll-off rate than do Butterworth filters for the same number of components, and are generally used in bandpass designs where bandpass ripple can be tolerated. Tabular approaches to the design of Chebyshev bandpass filters are also available [5],[6].

7.2.4 Band-Rejection Filters

Band-rejection and notch filters are networks that, from an EMC standpoint, are designed to attenuate a specific narrow band of frequencies that may be causing interference problems. This type of device is normally used as a series rejection device between the interference source and the load. An alternative is to use a bandpass configuration that shunts the interference to ground.

Typical applications and locations of band-rejection notch filters include the following:

- At input terminals to reject strong out-of-band interference that would otherwise cause overload.
- At receiver input terminals to reject troublesome image frequencies.
- At receiver input terminals to reject IF-feedthrough signals.
- At transmitter output or interstage terminals to reject harmonics.
- In ac or dc power distribution leads to reject radar PRF, computer-clock surges, or rectifier ripple.
- At audio amplifier input or interstage terminals to reject IF or BFO feedthrough, unwanted heterodynes, signal tones, radar PRF.

A notch filter or wavetrap 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 is often found to be an acceptable configuration.

The simplest types of wavetrap are a parallel or series resonant circuit such as those shown in (A) and (B) of Figure 7-23. The configuration of Figure 7-23(A) will give a high impedance at the resonant frequency, while the configuration of Figure 7-23(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, as also illustrated in Figure 7-23.

The details on the design of the above 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.

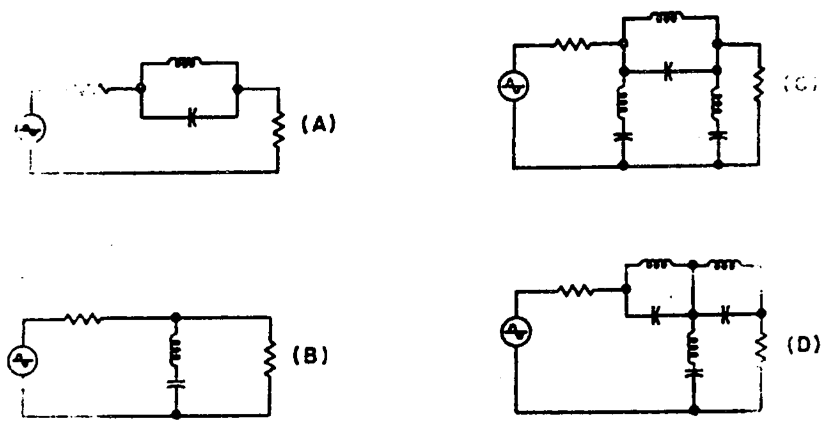


Figure 7-23 Band-Reject Filter Configurations

The twin-T notch filter shown in Figure 7-24 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 is determined by [7]:

$$f_n = \frac{0.1592\sqrt{K}}{\sqrt[2]{R_1 \ R_2 \ C_1 \ C_2}} \qquad (7\text{-}10a)$$

where

$$K = \frac{C_1 + C_2}{C_3} = \frac{R_1 \ R_2}{R_3 (R_1 + R_2)}$$

$$R_3 = \frac{R_1}{2K}$$

$$C_3 = \frac{C_1 + C_2}{K}$$

$$\omega = \sqrt{\frac{K}{R_1 R_2 C_1 C_2}}$$

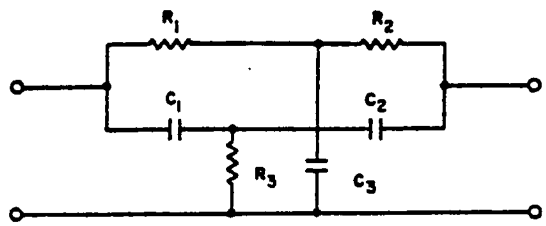


Figure 7-24 Twin-T Notch Filter

Three special cases are of interest. The case when K = 1 gives the symmetrical form of Figure 7-25. With K = 1/2, a circuit with three equal resistances as shown in Figure 7-26 is obtained. In Figure 7-27, with K = 2, three equal capacitances result.

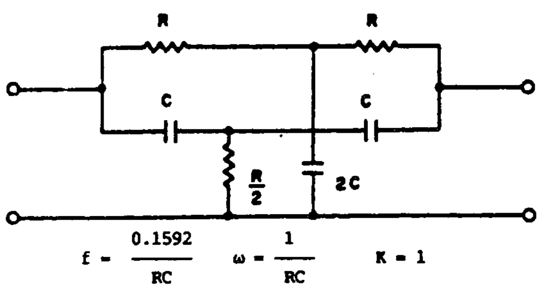


Figure 7-25
Twin-T Network with K = 1

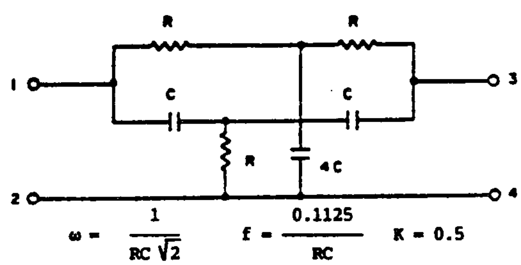


Figure 7-26
Twin-T Network with K = 0.5

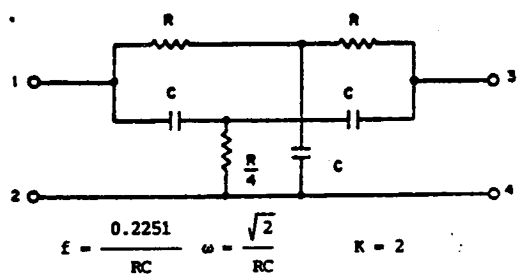


Figure 7-27
Twin-T Network with K = 2

Various transfer characteristics can be obtained with twin-T filters by varying the parameter K. For example, for the unloaded network and $R_1 = R_2$ and $C_1 = C_2$, the transfer function at $\omega = \omega_o$ can be shown to be:

$$\frac{E_o}{E_i} = \frac{K(K-1)}{2+K(K+1)} \qquad (7\text{-}10b)$$

Thus for $K = 1$, a complete null is obtained, while for $K = 2$, $E_o/E_i = 0.25$. For $K = 0.5$, a 180 degrees phase reversal is obtained, since $E_o/E_i = -0.091$.

A wide variety of characteristics can be obtained by removing the restrictions $R_1 = R_2$ and $C_1 = C_2$, as may be required to handle differing source and load impedances.

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.

7.3 TRANSIENT SUPPRESSION

The making and breaking of current, either mechanically or electronically, can introduce transient radiated and conducted effects. The transient may be generated within a GSE as a result of a switching function, or external to the test set, such as from another equipment tied to the same power line, or from the weapon system being tested.

If the source of the transient is external to the GSE and cannot be reduced at its point of origin, then the shielding, bonding, grounding and filtering techniques already discussed in this publication must be applied by the test equipment designer to prevent the transient energy from affecting GSE performance. If the source of the transient is within the GSE, the designer should be aware of particular switch transient filter techniques he can apply.

7.3.1 Inductive Loads

When an inductive load is opened by a switch contact, such as when a relay coil is opened, a reverse voltage is produced by the collapsing magnetic field. This reverse voltage increases until an arc occurs across the contact. The arc produces a wide frequency band of interference which is conducted and radiated away from the switch. Shielding and filtering will be necessary to reduce noise conduction and re-radiation from the switch contacts and wiring.

Figure 7-28 shows a number of suppression techniques that can be used to minimize transients in circuits that switch inductive loads [8]. These techniques are as follows:

a. RESISTANCE DAMPING - Use of a non-inductive resistor across the load is the least expensive approach but it has drawbacks because the resistor will continuously dissipate power, and because the suppressor will increase relay dropout time. Maximum protection is obtained when the value of the suppression resistor, R, is equal to the coil resistance, R_L, but this will generally result in severe steady state power consumption requirements. Practically, R is kept as low as possible consistent with power capabilities and dropout time. A typical resistance value chosen is 10 times R_L, but this size may cause the dropout time of the relay to increase by as much as a factor of ten.

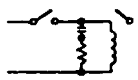
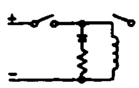
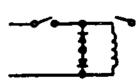
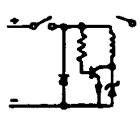
TYPES OF INDUCTIVE SUPPRESSION	VOLTAGE INPUT	RELAY CONTACTS		REMARKS
		CLOSING	DROPOUT	
A  RESISTANCE DAMPING	AC or DC	NO EFFECT	FUNCTION OF RESISTANCE	Increase in power consumption. Resistance should be as low as practicable. Observe power rating, E^2/R, and heat dissipation.
B  CAPACITANCE SUPPRESSION	AC or DC	SLIGHT EFFECT	SLIGHT EFFECT	Need series resistance of a few ohms. Capacitance value around .01 to 1 uF. Capacitance rated 10 times the input voltage.
C  SINGLE DIODE SUPPRESSION	DC ONLY	NO EFFECT	SLIGHT EFFECT	Polarity is critical; diode is connected in backward or nonconductive direction. PIV should be higher than any transient voltage plus safety factor. Series resistance of a few ohms might be needed to increase inductance life.
D  BACK-TO-BACK DIODE SUPPRESSION	AC	NO EFFECT	NO EFFECT	Avalanche voltage should be above input voltage. Power dissipation should be sufficient for transient current. Cost of device becoming a significant factor.
E  DIODE-TRANSISTOR SUPPRESSION	DC	NO EFFECT	NO EFFECT	Most effective in transient suppression. Voltage transient and dropout time negligible. Most expensive technique.

Figure 7-28 Comparison of Various Suppression Devices Across an Inductor

b. CAPACITANCE SUPPRESSION - A very popular relay suppression circuit consists of a series resistor capacitor combination placed across the coil. Care must be taken in this arrangement to avoid circuit resonance effects that will cause contact chatter, and to avoid overdamping so that dropout time does not become excessive. A typical arrangement is for R to be 1/4 to 1/2 of R_L. The value of the capacitance, C, can then be obtained from the relationship:

$$C = \frac{L}{RR_L} \qquad (7\text{-}11)$$

where L is the inductance of the load.

If the computed capacitance is very large, use the closest practical smaller capacitor. Capacitance between .01 to 1 uf is usually sufficient to minimize most transient effects.

The RC suppressor can create EMI problems when the relay is actuated, since there is an increase in current flow until the capacitor is charged. It is often preferable to place the suppressor across the relay contacts rather than across the coil to reduce this problem.

c. SINGLE DIODE SUPPRESSION - A series diode-resistor combination across the coil eliminates the power dissipation problem of the Resistor Suppressor. When the relay is energized from a dc source, the diode is back-biased so that the suppression circuit has no effect on relay operation. When current to the relay coil is opened, a current flows through the suppression circuit; this current is generally a function of the value of the suppression resistor R, (since the value of R is usually significantly larger than the forward resistance of the diode). The resistor value should be kept as small as possible consistent with relay dropout requirements.

   Silicon diodes are used for this application because of their low cost, adequate current ratings, and ability to meet peak inverse voltage (PIV) requirements. Zener diodes have also been employed for this purpose.

d. DUAL DIODE BACK-TO-BACK SUPPRESSION - An extension to the single diode suppression approach is the use of back-to-back diodes across the relay coil. This arrangement is insensitive to the polarity of the relay control signal and can therefore be employed with ac relays. The component that conducts when power to the relay coil is removed may be a Zener diode, to clamp the transient at a specific level. Use of a Zener diode can help the dropout time problem as well.

e. DIODE-TRANSISTOR SUPPRESSION - A very effective transient suppression approach is to use two diodes, a resistor, and a transistor in an arrangement as shown in Figure 7-28E. When the switch is opened, the transistor is cut off, forcing the transient energy to dissipate through the Zener diode Z and the conventional diode D. While this suppression circuit results in good transient control, it is a relatively expensive alternative, and should only be considered when the previously discussed approaches are not suitable.

7.3.2 Mechanical Switches

As indicated previously, any switching device causes transients during opening and closing. The arc generated across the switch contacts is extinguished when the energy stored in the inductance of the circuit is dissipated and the voltage drops below the value required to maintain the arc. To prevent or reduce the arc, the current, instead of being interrupted, can be channeled into another branch containing a series capacitance and resistor of a few ohms, as was suggested for reducing relay coil transients. In this way, the energy is partly stored in the capacitor and partly dissipated in the resistor, which also serves to damp any oscillations that may occur as a result of the added capacitance.

RC arc-suppression networks should be used whenever switches or relays are employed and equipment transient susceptibility is of concern. A capacitor should never be connected across contacts without including a series resistance, because the capacitor discharge through the switch contacts when the contacts are closed can cause a heavy current surge if not controlled by the resistor.

In addition to arcs being generated, relays can also chatter. This occurs when the polarity of the voltage across the relay alternates, and can be caused by stray RF being coupled directly into the switch (inductive coupling) or onto the powerline and then into the switch (conductive coupling). Conductive coupling, in fact, was the scenario sited in Section 3.2.1 with the Airborne/Ground power switch for the A-7E.

To determine the mode of coupling, causing the relay to chatter, one can perform EMI tests to MIL-STD-461 requirements for conducted and radiated susceptibility using the counterpart test methods from MIL-STD-462. Such tests consist of determining the frequency and threshold of susceptibility.

7.3.3 Transformer Switching

Transformer cores for GSE are often made from grain-oriented silicon steel. The high permeability of this material results in low exciting volt-amperes and low core loss. Because the cores operate at very high flux densities, the high inrush currents are of particular concern. The maximum inrush current occurs when the switch circuit is closed at the instant the voltage is zero.

The butt-joint air gap of the core can be increased to control the inrush magnetizing current. In typical large cores, a spacer 0.005 inch thicker than the gap, will reduce inrush magnetizing currents by a factor of two or three; however, the exciting current is increased by about 2%. Another solution is to control the timing of the energizing switch or relay so that the transformer is connected when the voltage is at a maximum. The switch used in this application must be able to close the circuit at or near voltage maximum, when the flux change is approximately zero.

Both electrostatic and electromagnetic transformer shielding can be effective in reducing transformer-generated interference. The purpose of electrostatic shielding is to significantly lower the capacitance between primary and secondary transformer windings, and thus stop the direct flow of high-frequency currents through the transformer. This can be accomplished by enclosing one or both windings in a thin metal conducting shield, usually of copper or aluminum, with the ends of each shield insulated from one another. Either the shields are internally grounded to the core and case of the transformer, or a connection is provided so that external grounding can be accomplished.

EM shielding is necessary to reduce stray magnetic fields outside the transformer. One method of doing this is to wrap and solder a copper band of .005 to .025 inch thick material around the winding and core (see Figure 7-28A). This band, called a shading ring, is a shorted turn of low resistance, and has the effect of creating a magnetic field that cancels much of the original transformer field. Another more expensive approach is to enclose the transformer in a high permeability case, or to line its regular steel case with high permeability material such as silicon steel.

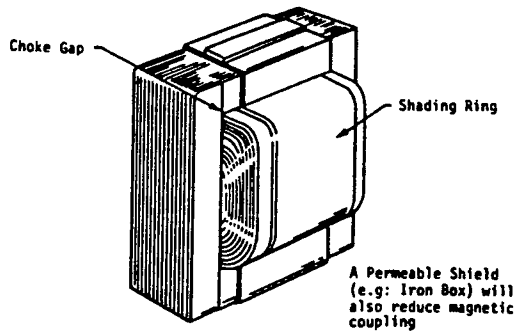


Figure 7-28A Shading Ring in a Transformer or Choke

7.3.4 Semiconductor Transients

Semiconductors that are used in power conversion or certain digital processing circuits operate as switches. The rapid current changes occurring in these devices when they are either turned on or turned off can give rise to EMI containing significant high-frequency spectral components. The spectral effects generally associated with pulsed signals are discussed in Section 2.2.2 of this design guide. However, some of the unique characteristics of transistors and other semiconductors that further contribute to spurious signal emissions are noted here [17].

7.3.4.1 Diode Recovery Time

Reverse recovery 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 ("sweep out") the minority carriers from the n and p regions and reunite their covalent bonds. T_{rr} for switching diodes ranges from 200 nsec to 12 nsec and is related to the magnitude of peak reverse current (I_{pr}). It has been shown that I_{pr} can be five times greater than the forward current for a 200 nsec diode and decreases to a value less than the forward current for a 12 nsec diode. In power supply applications, the reverse current pulses, once produced, are 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 taking advantage of parasitic circuit capacitance.

7.3.4.2 Forward Recovery

Forward recovery transients have a similar interference generating effect as 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 minimum tolerable rise time, (2) keep the junction slightly forward biased, (3) reduce the drive pulse amplitude to the lowest level to maintain reliable switching action, and (4) select a device with fast recovery characteristics.

7.3.4.3 SCR Recovery

Silicon controlled rectifiers (SCRs) can produce reverse recovery transient currents and induce EMI similar to p-n junction diodes. The reverse recovery transients in SCR applications are intensified by increased forward current, predominately inductive loads, and the large junction areas associated with these rectifiers. A T_{rr} for a typical SCR is on the order of microseconds. In full-wave SCR applications, the reverse recovery current has to be carried by the complementary SCRs. The reverse conducting SCR can produce an excessive forward current rise time in the conducting SCR. When SCRs 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 SCRs. Consideration of load current and surge current sharing is necessary when parallel SCR operation is required.

7.4 ACTIVE POWER LINE 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 more easily accommodated with active devices.

Active filters for a dc line may contain capacitors as storage elements, modified series regulators to create a high impedance path, and modified shunt regulators in combination with high gain feedback systems for cancellation of interference. In the case of ac lines, cancellation is 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 Figure 7-29 uses phase cancellation and operates as follows: The input signal is fed into an ac-coupled amplifier through a notch filter, which is tuned to the fundamental of 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 power line frequency and correct any change in the filter tuning itself. The voltage attenuation curve for a 220 volt, 20 ampere unit is also shown in Figure 7-29.

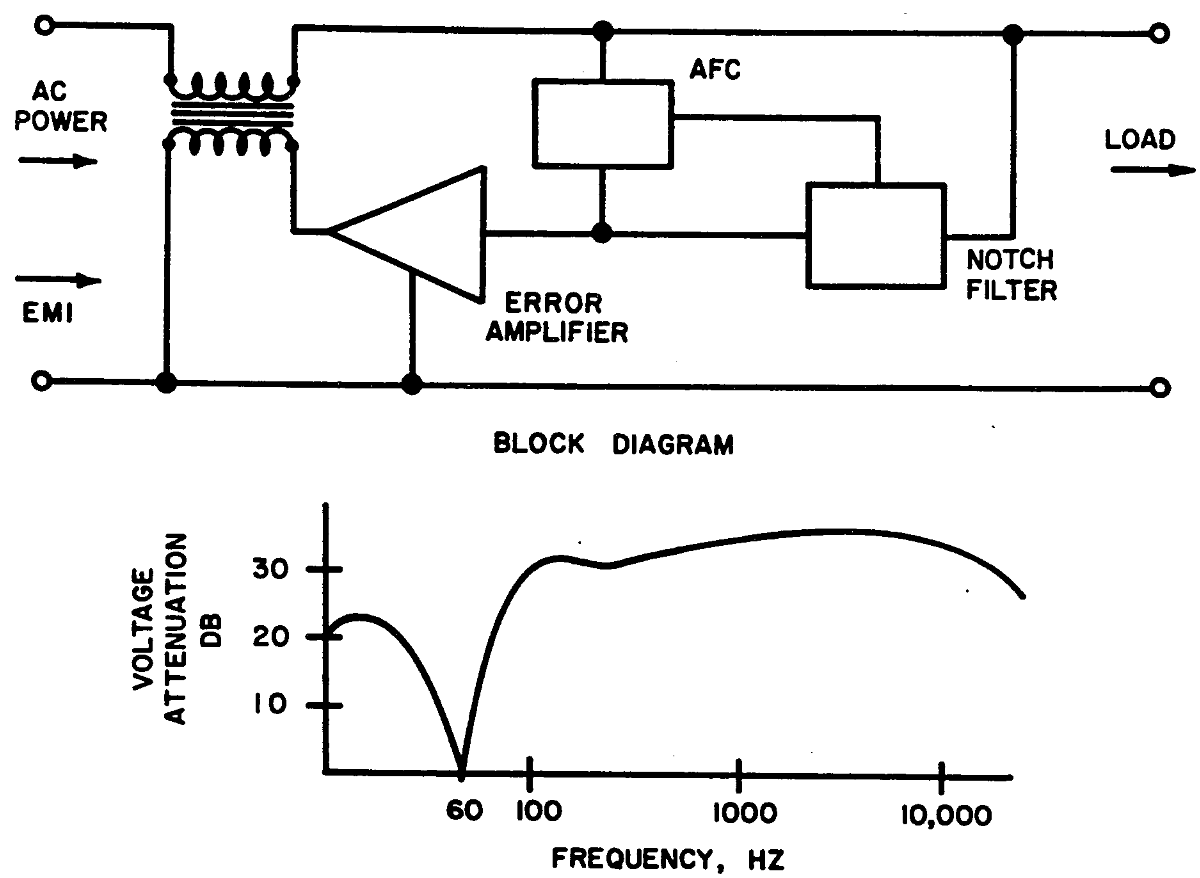


Figure 7-29 Active Filter for Powerline Interference

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 volt for a load power factor approaching unity and up to 6 volts 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 percent have been realized for 200 volt, 20 ampere filters.

7.5 NOISE BLANKING, CANCELING, AND LIMITING CIRCUITS

7.5.1 Noise Blanking

If the interfering signal is a pulse, and it is not possible to prevent the pulse from getting into the system being developed, a blanking circuit might be used. The function of a blanker in the EMC sense is to render the system inoperative for the duration of the interference.

The blanking circuit can be triggered by the interfering pulse, or by an independent signal arriving before the interference pulse. When the interference pulse triggers the blanking circuit, delay lines must be provided to delay the desired signal long enough to allow the blanking circuit to turn off the system before the interference pulse reaches it. An independent pulse can be used if the arrival of the interference pulse is known beforehand by way of a trigger pulse, and the trigger can be hardwired to the GSE. Blanking action is usually provided by a simple gate circuit that can be actuated by the trigger pulse, and a timing circuit to control the duration of the gate.

Blanking circuits lack the simplicity of wavetraps and limiters, and can also cause interference. The periodic gating of the signal is a form of modulation, which will appear in the output as noise. The major disadvantage of blankers is that they introduce dead-time in a signal channel. For some applications, this dead-time cannot be tolerated.

7.5.2 Cancellation

An interference canceling circuit suppresses interference by allowing the interference signal to travel along two paths. One of the paths carries only the interference signal. The interference signal is shifted 180 degrees in phase, adjusted in amplitude, and added to the channel containing the signal and interference. The result is the cancellation of the interference while leaving the desired signal.

This method can only be used when the path of entry and the nature of the interfering signal are known, and when the source of interference is accessible. Interference canceling circuits are useful in suppressing interference from equipment when the interference signal has unique properties relative to a desired signal, such as radar transmitter interference to an analog device.

Figure 7-30 illustrates one arrangement for eliminating interference by cancellation. Two directional couplers are shown. One, indicated by DC-1, samples the interfering signal at the offending transmitter. The signal is routed through delay lines to introduce the required phase shift, attenuated to set the power level, and then coupled to the receiver via another directional coupler, indicated as DC-2.

For canceling interference sources not a part of the same system being developed, a different pickup arrangement will be needed, and the phase and attenuation control methods will be more complex. One pickup arrangement that has been employed is a separate receiver to pick up the interfering signal. For this configuration, it is necessary to include automatic phase (or frequency) and gain control circuitry to control the characteristics of the signal in the cancellation path. This type of EMI reduction technique should be considered only when the type and source of interference is known, and when system operation and outside influences can be tightly controlled.

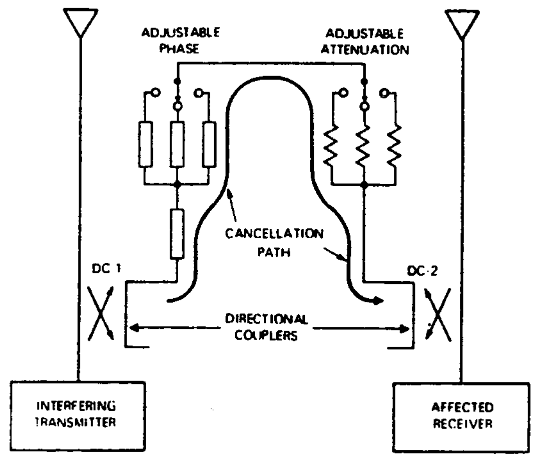


Figure 7-30 Block Diagram of an Interference-Canceling Network

As an example of the expected capabilities of cancellation circuitry, consider a prototype canceller designed to suppress analog signals in the 225-400 MHz range [9]. Such a device typically provided at least 35-40 dB of rejection to the interference signal at desired-to-interference signal frequency differences of as small as 0.025 MHz, so long as the interference-to-signal ratio in the interference channel was 20 dB or more.

7.5.3 Limiting

For interfering signals consisting of short impulses of large amplitude, noise limiters can prevent the peak amplitude of the interference from rising above the desired signal and overloading the system. Noise limiters are not effective against thermal noise, or steady-state EMI. There are two types of noise limiters: peak amplitude limiters and gated noise limiters.

The peak amplitude limiter reduces the effects of interference by clipping all impulses above a pre-established threshold level. To be most effective, the clipping level should be as close as possible to the peak value of the desired signal, but should not be adjusted to clip so heavily as to distort the desired signal. Diode devices are commonly used as limiters, although tube or transistor amplifiers biased to swing into a nonlinear region on strong signals may also be employed. The disadvantage of peak amplitude limiters is that they introduce nonlinear effects, with the resultant possibility of cross-modulation on strong signals.

The gated noise limiter is a low frequency device that might also be classified as a blanker. Basically, the circuit consists of a diode or combination of diodes having a fast time constant network at the input, and a slower time constant network at the output. As long as the charge-discharge period at the output can follow the desired input signal, the diode gate continues to conduct and the input signal is delivered to the output. However, an impulse will possess a rise-decay rate greater than that provided for at the noise gate output. This will cause the gate to be back-biased and cut off momentarily until the slower time constant at the output can catch up with the input. Noise impulse steep wavefronts are thus gated out. The gated noise limiter is generally effective only when used with near-sinusoid waveforms.

The usefulness of both types of noise limiters depends on the width of the system bandpass. The bandpass must be broad enough to avoid "ringing" on noise impulses. Also, bandwidth-limiting can alter the noise impulse envelope so that subsequent limiter action is ineffective.

7.6 FILTER TESTS

7.6.1 General

There exist several techniques for measuring filter performance, based on the type of filter and its intended application. The tests to be described are those which appear to provide the most meaningful results of such performance. They include:

- Insertion Loss Test in accordance with MIL-STD-220A, Notice 1
- Filter Admittance Transfer Test
- Parallel Signal Injection Test
- Series Signal Injection Test

These test methods are applicable to both signal line and power line filters. In all cases, the test procedures are designed to measure the filter insertion loss while the filter is carrying rated load current.

The MIL-STD-220A, Notice 1, Insertion Loss Test [2] and the Filter Admittance Transfer Test are very similar. They differ mainly in that the former test is restricted to the situation of 50 ohm input and output impedance terminations. This case may not be representative of the actual filter operating condition. The Filter Admittance Transfer Test provides the flexibility of varying the output load impedance.

In determining power-line filter attenuation characteristics, it is also important to use measurement techniques which give results that demonstrate installed performance. For example, Figure 7-31 provides one illustration of the variation of ship power line impedance. Not shown in the figure are the very low values of expected line impedance (well below one ohm) below 15 kHz. Variations of impedance as a function of location on a power-line also occur.

A meaningful measurement should indicate minimum attenuation that the installed filter will exhibit under such impedance variations. A worst-case filter test method is therefore a desirable technique for evaluation of filters, when the filters will be used in a variety of installations where impedance conditions can vary widely. That is the objective of the Parallel and Series Signal Injection Tests.

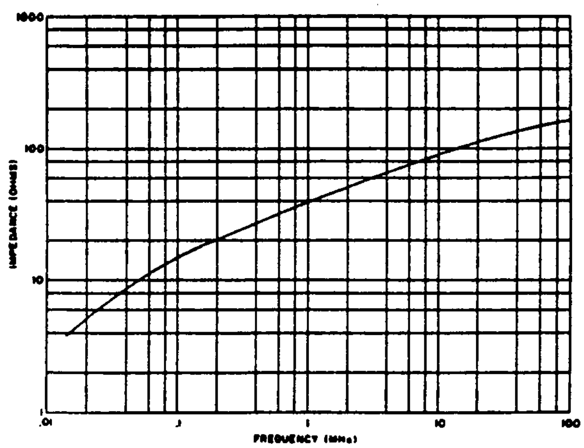


Figure 7-31 Sample of Power Line Impedance

For simulated installation testing, the current injection probe technique used in the Series Signal Injection Tests will furnish meaningful results, provided that the current injection probe design does not cause changes in loading conditions by inserting large values of impedance into the filter circuit, and provided that it does not cause coupling between filter input and output. However, the limited frequency range offered by this test necessitates use of a different test method at high frequencies, usually that defined by MIL-STD-220A, Notice 1.

Before discussing the individual filter test procedures, it is important that a few unique terms to be used in this section of the design guide be defined:

- <u>Buffer Network</u>: an isolation network, consisting of (a) an inductor designed to have a very high self-resonant frequency and capable of carrying the full rated load current of the filter under test, and (b) a source-end shunt by-pass capacitor.
- <u>Current probe</u>: A calibrated toroidal winding capable of being placed around a current-carrying conductor. This arrangement then becomes a 1:N transformer, where N is number of turns on the toroidal core. Probes of this type are used for both current detection and current injection. In the latter case, the probe windings must be heavy enough to carry the required current for specified test power levels.
- <u>Power bias</u>: Direct or low frequency current passed through a filter under test. The presence of such current enables filter performance to be tested under load.
- <u>Substitution measurements</u>: A technique for finding the amplitude of an unknown voltage or current by using a detector to indicate an arbitrary value at some combination of detector device control settings and, without changing the detector system, replacing the unknown source with a known calibrated voltage or current that can be adjusted to give the same detector indication.

- <u>Transfer admittance</u>: The ratio of network signal output current divided by the network input signal voltage.

- <u>Transfer impedance</u>: The ratio of network signal output voltage divided by the network input signal current.

The test procedures provided in this chapter are for guidance purposes only; refer to the cited references for further test details.

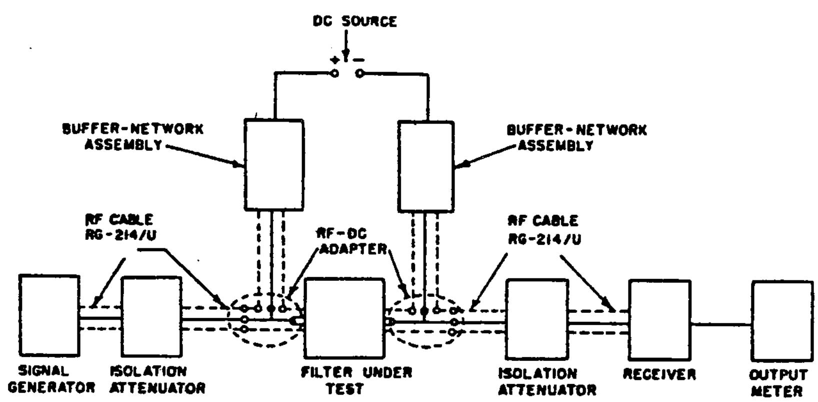
7.6.2 <u>MIL-STD-220A, Notice 1 Insertion Loss Test</u>

7.6.2.1 Objective

MIL-STD-220A, Notice 1 covers a method of measuring, in a 50 ohm system, the insertion loss of feed-through suppression capacitors, and of single-/multiple-circuit RF filters. MIL-STD-220A, Notice 1 defines insertion loss in accordance with Equation (7-1).

7.6.2.2 Test Set-Up

Figure 7-32 is a general test set-up for filter insertion loss measurements as given in MIL-STD-220A, Notice 1. The isolation attenuators in this set-up have input and output impedances of 50-ohms to provide matching for the signal generator and the receiver. These networks also present 50-ohm loading at both the input and output of the filter. Using the insertion-loss measuring equipment specified, the test set-up is capable of measuring insertion loss within ± 1.0 dB over the required frequency range.



7-32 Test Arrangement for MIL-STD-220A, Notice 1 Filter Test

All test equipment must be shielded and filtered so that leakage from the signal generator or any portion of the signal-source circuitry does not affect the output of the receiver when the generator and receiver are operating at the output level and sensitivity needed to make the required maximum insertion-loss measurement. The dc source used in making these measurements must be a floating dc source and not connected to ground.

It is essential that cable with the general characteristics of type RG-214/U be used to connect the isolation attenuators together for the filter-out condition, and to connect the filter under test to the isolation attenuators for the filter-in condition. The length of cable connecting the isolation attenuators in the filter-out condition should be within 6 inches of the combined length of the two cables connecting the filter under test to the isolation attenuators in the filter-in condition. Type N 50-ohm coaxial connectors conforming to specification MIL-C-71 should be used where applicable. When coaxial switches are used, they should have a 50-ohm characteristic impedance, and a maximum voltage standing wave ratio (VSWR) of 1.1 to 1 at the frequency of measurement.

The buffer-network assembly and RF-dc adapter assemblies shown in Figure 7-33 and in Figure 7-34 are specified by MIL-STD-220A, Notice 1 when performing insertion-loss measurements with rated current applied. The buffer-network assembly provides isolation between the RF test signal and the dc power supply. The complete measurements system can be used for filter tests over the frequency range of 100 kHz to 20 MHz. The indicated buffer-network is suitable for continuous use with currents up to 100 amperes.

7.6.2.3 Procedure

The test equipment is set up as shown in Figure 7-32 except that a cable of the type and length specified for the filter-out condition may be inserted between the attenuators when applicable.

- Adjust the signal generator to the desired test frequency, with the attenuator set for the lowest convenient value of output voltage.
- In the filter-out condition, tune the receiver to resonance at the frequency of the generator, and set the gain controls so that receiver sensitivity will be high enough and the level of circuit noise low enough to allow clear reception of the signal required for the measurements.
- Adjust the output of the generator to give the lowest possible stable and readable indication on the output meter, care being taken not to saturate or overload the receiver. Record the generator output level (E_2).
- Remove the cable used for the filter-out condition, and insert the filter under test and its connecting cables between the isolation attenuators.
- Readjust the receiver to resonance, and adjust the output of the generator until the output meter gives the same indication as that obtained for the filter-out condition. Record the generator level (E_1).
- The insertion loss of the filter being tested under the specified conditions and at the frequency of measurement may then be expressed in dB as:

$$\text{Insertion Loss} = 20 \log \left[E_1/E_2\right] \qquad (7\text{-}12)$$

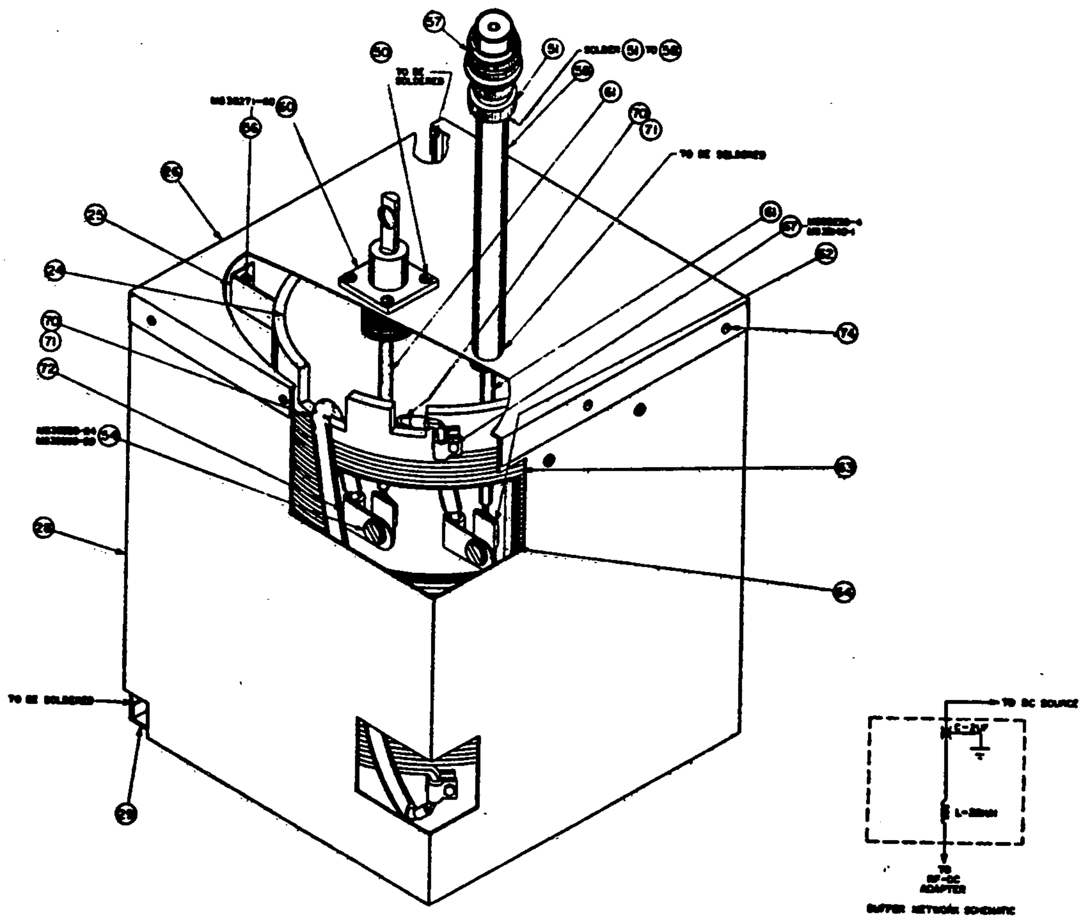


Figure 7-33 MIL-STD-220A Buffer Network

- o If a multiple-circuit filter is under test, measure the insertion loss of each circuit of that filter with each of the other circuits open-circuited, and also with each of the other circuits short-circuited. By definition, the insertion loss of the filter at the frequency of the test is considered to be equal to the lesser of these two measurements.

- o If full-load insertion-loss measurements are being performed, the nominal dc rated current of dc components, or the dc equivalent of the peak ac rated current of ac components should be applied during the above test.

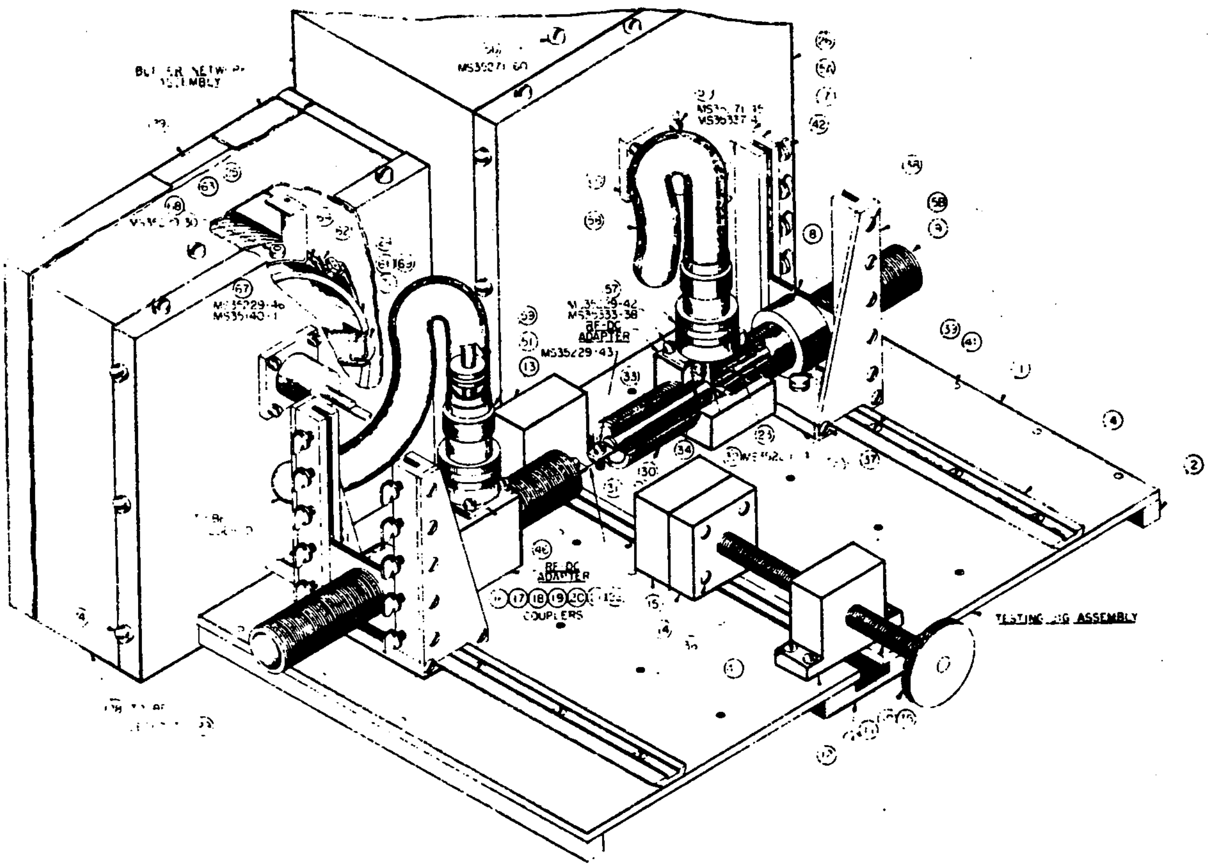


Figure 7-34 MIL-STD-220A Filter Test Fixture

7.6.2.4 Limitations and Precautions

- Measurements using MIL-STD-220A, Notice 1 can be extended to 1000 MHz, but caution should be exercised in judging the accuracy of the test results at frequencies above 30 MHz. The MIL-STD-220A, Notice 1 buffer-network limits measurement accuracy below 100 kHz and above 20-30 MHz for rated load tests.
- Be sure the dc power supply is an ungrounded supply.
- As a precautionary measure, apply the power bias starting with zero voltage from the power supply.

7.6.2.5 Advantages/Disadvantages

- The 50-ohm system used allows the performance of different filters to be measured against a common standard.
- Power bias load currents of up to 100 amperes may be used in the filter tests.

o The buffer-networks, as defined by MIL-STD-220A, Notice 1, limit low frequency measurement down to 100 kHz and high frequency measurements up to approximately 30 MHz, for full power load tests. For no load tests, the test range may be extended up to 1000 MHz. Figure 7-35 provides a comparison of load and no-load tests taken in accordance with this procedure at frequencies below 100 kHz [10].

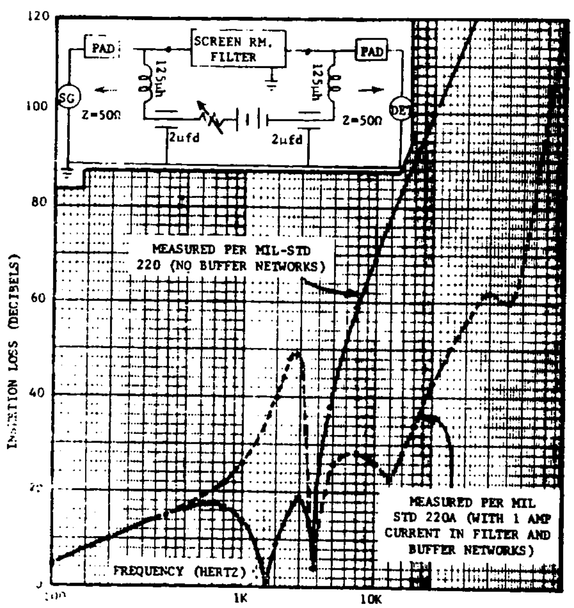


Figure 7-35a
Insertion Loss of 100 Ampere
Screen Room Filter

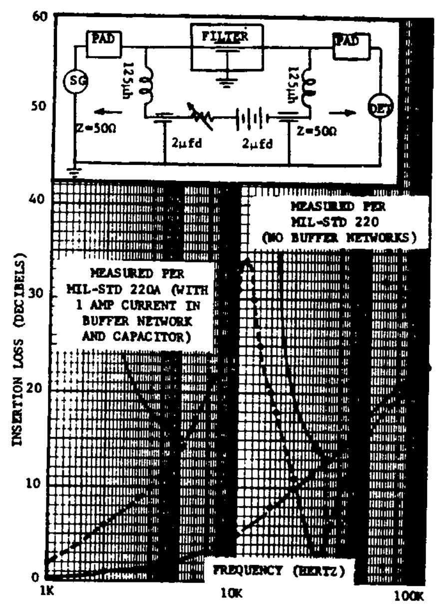


Figure 7-35b
Insertion Loss of 1.0 μF
Feed Through Capacitor

o This test is based on a 50-ohm matched impedance condition at both input and output of the filter under test. Impedance conditions of other than 50-ohms will result in different insertion loss values. It should be recognized that GSE terminations are not necessarily 50 ohms.

7.6.3 <u>Filter Admittance Transfer Test</u> [11]

7.6.3.1 Objective

The objective of this test is to measure filter performance under extremes of load impedance with readily available instruments.

The measurement set-up for the admittance transfer test is shown in Figure 7-36. The method uses a current probe to measure the filter network output current, and a direct measurement of input test voltage to the filter. The insertion loss of the network is calculated from:

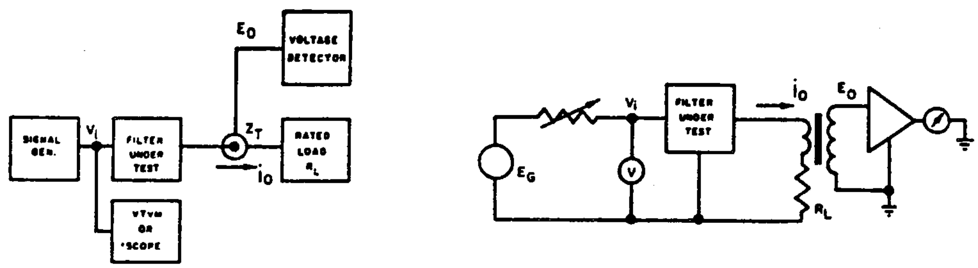


Figure 7-36 Transfer Admittance Test Block Diagram and Schematic Diagram

$$\text{Insertion Loss} = 20 \log_{10} \left[\frac{V_i + i_o R_L}{i_o R_L} \right], \text{ in db} \qquad (7\text{-}13)$$

where:

V_i = Input test signal voltage

R_L = Filter Load Impedance

i_o = Filter output current into load = E_o/Z_T

Z_T = Current probe transfer impedance

E_o = Current probe output voltage

This relation is obtained by assuming that the filter is a two-terminal device, and is applicable for absorptive-type filters with any load R_L. However, for lumped parameter L-C type filters, which are essentially four-terminal devices, the formula is applicable only when R_L is much smaller than the series transfer resistance, R_T, of the filter, i.e., V_i/i_o.

In many respects, this test is similar to MIL-STD-220A and results can be correlated when R_L = 50 ohms, if R_T is much greater than 50 ohms. The test has the advantages over the MIL-STD-220A, Notice 1 procedure in that measurements can be made with different filter output load impedance terminations. In this respect, the transfer admittance test can more closely approximate actual usage conditions, since the output load resistance can be made very low. However, it should be pointed out that at frequencies below filter cut-off, the series resistance of a lumped parameter low-pass L-C filter will be small and it will be difficult to satisfy the condition that R_L be much smaller than R_T.

7.6.3.2 Test Set-Up

The test block diagram is shown in Figure 7-36. The signal generator should have a power output level, in dB, at least equal to the detector sensitivity in dB, plus the attenuation of the filter, plus the probe gain. For example, assume the generator has a +10 dBm maximum output (600,000 μV, or 117 dBμV), the detector is a receiver with 0.1 μV sensitivity (-20 dBμV), and the anticipated filter specification at the test frequency is 100 dB. The probe transfer impedance is 0.31 ohms, or -10 dB re 1 ohm. The calculation is:

$$\text{Generator Output} - \text{Filter Loss} + \text{Probe Gain} - \text{Detector Noise} = \text{Measurement Margin} \qquad (7\text{-}14)$$

$$(117) - (100) + (-10) - (-20) = 27 \text{ dB}$$

Since the measurement margin is positive, the signal generator is adequate.

The output voltage should be measured with a detector having a high input impedance over the entire test frequency range. The current probe should have a transfer impedance between 0.1 and 10 ohms, and should be placed around the filter output conductor as close to the filter output as possible. Do not clamp the probe around the outside shield of a coaxial cable; clamp it around only the "high" side or center conductor of the coax. Use a ground plane for current return. Be sure a cable of correct characteristic impedance connects the probe to the detector.

7.6.3.3 Test Procedure

At each test frequency, perform the following steps and record the indicated data:

- o Adjust the signal generator frequency and output level.
- o Record the output frequency and generator output voltage level.
- o Adjust the detector until the reading on the output indicator is greater than one-third full scale.
- o Measure the voltage delivered by probe to the detector input, using either substitution or direct-reading techniques.
- o Using the probe transfer impedance, the voltage delivered from the probe, and the input voltage to the filter, calculate the filter insertion loss for the test frequency and filter termination used.

7.6.3.4 Limitations and Precautions

- o The series impedance of the test filter must be much greater than the load impedance.
- o Insertion loss can be accurately measured by this method only when the frequencies are well into the attenuation band, e.g., when the filter insertion loss is greater than 20 dB. Thus, neither the shape of the pass-band nor the 3 dB point can be accurately determined in this way.

- The current probe turns ratio must be high enough not to reflect significant series impedance levels compared to the value of test load impedance. At the same time, the probe must not have too many turns, or the value of transfer impedance will be too low and the probe frequency response and sensitivity will be limited.

- For high values of insertion loss, care must be taken that the input and output terminals of the filter are electrically isolated, and that generator and detector are not coupled in anyway to by-pass the filter under test. While this requirement applies to all tests, it is particularly important when using current probes.

7.6.3.5 Advantages/Disadvantages

- Insertion loss values for varying load conditions can readily be found using this procedure, subject to the constraints indicated above.

- High levels of attenuation can be measured, with proper isolation of filter terminals.

- A wide dynamic measurement range is possible.

- Test results can be correlated with MIL-STD-220A when R_L is 50 ohms and the filter series impedance is much greater than 50 ohms.

- Low values of insertion loss cannot be accurately measured.

- The filter 3 dB point cannot be accurately measured.

- The test arrangement is not designed to measure filter attenuation under load, although the method could be adapted to incorporate this feature.

7.6.4 Parallel Signal Injection Test [11],[12]

7.6.4.1 Objective

This test is intended as a "worst-case" filter test technique, where the test filter is operated under rated power conditions, and the input and output capacitors of the filter are resonated at each test frequency. The test is designed for low pass pi-filters, having as the last element at each end a shunt capacitor. Under these conditions, the filter insertion loss is found, and will be the minimum value for any service load or source impedance to been countered. This test uses a method of parallel test signal injection.

7.6.4.2 Test Set-Up

The test arrangement is shown in Figure 7-37.

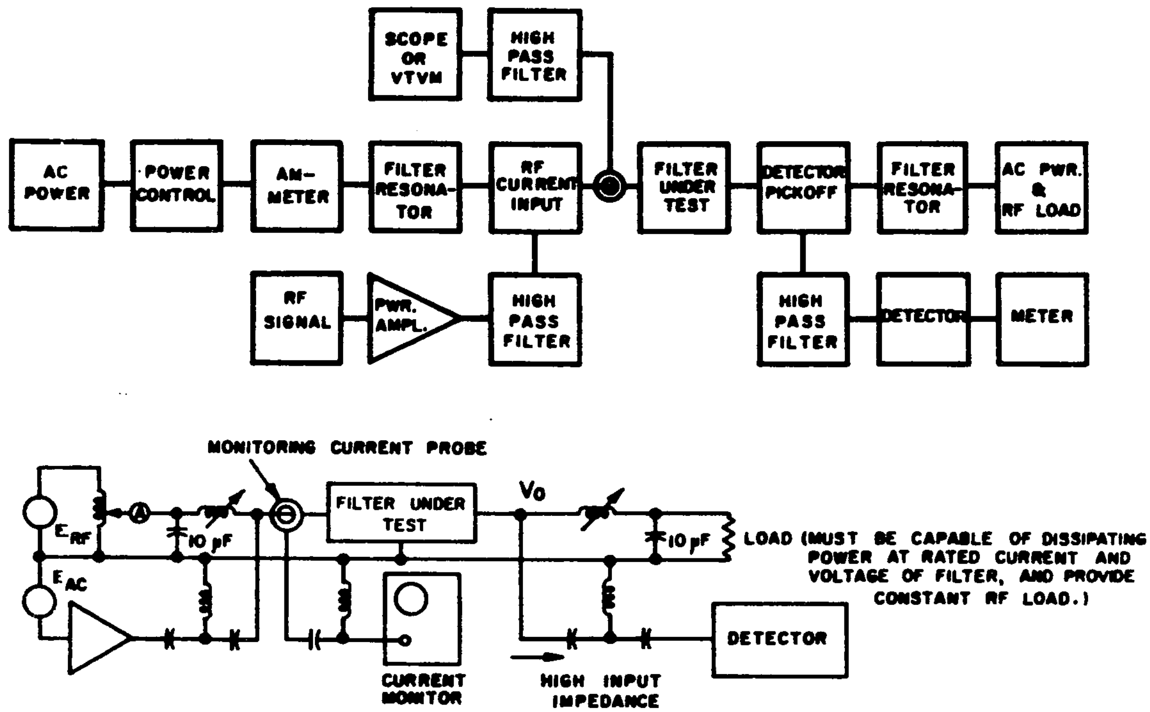


Figure 7-37 Parallel Signal Injection Block Diagram and Schematic Diagram

A test signal source is inserted in parallel with the filter input and a detector is inserted in parallel with the filter output. The input current level is monitored with a probe; the probe-inserted impedance is not critical since the impedance level of the current source is large in comparison to the signal source impedance of about 50 ohms.

The high-pass filters shown in Figure 7-37 are needed to keep the low-frequency power-biasing voltage from reaching the signal source and detector. To avoid errors resulting from radiation, the oscillator, amplifier, and detector should be placed outside the shielded room used in performing the test, or the filter should be mounted within a separate enclosure in the shielded room.

The power bias source for testing the filter under rated load can be either the standard ac line voltage or a dc power supply of variable output voltage and required maximum test current output. The set-up shown in Figure 7-37 assumes a 115 V, 60 Hz source, with amplitude control using an autotransformer and an ammeter to measure power load current. The variable frequency test signal current source is a standard signal generator and, if the generator output level is insufficient, a booster amplifier.

The variable inductors shown in Figure 7-37 are used to resonate the filter shunt capacitors at each test frequency.

7.6.4.3 Test Procedure

With the equipment arranged as shown in Figure 7-37, the following steps can be performed to find the lowest possible insertion loss of a particular pi-section filter:

- o Determine the value of the π-filter shunt capacitances, both input and output.
- o Calculate the range of resonating series inductor values required for the range of test frequencies.
- o Place the correct series inductors in the test circuit and adjust for resonance at the test frequency.
- o Adjust the test signal generator at the test frequency, and adjust the detector for indication of the test signal.
- o With the filter input and output resonating, and with power bias of the desired level applied, record the detector indication of received test signal.
- o Also record the injected test current level into the filter (I_2), and the power bias level from the ac or dc power supply.
- o Record the frequency used in the test.
- o Remove the filter under test, and connect the test signal source directly to the detector and load (filter-out condition).
- o Decrease the test signal current level until the same detector indication is obtained as the test filter in the circuit.
- o Record the input test (I_1).
- o Calculate the insertion loss of the filter from:

$$\text{Insertion Loss} = 20 \log_{10} (I_1/I_2) \tag{7-15}$$

7.6.4.4 Limitations and Precautions

- o The upper frequency limit of the Parallel Signal Injection approach is dictated by inductance introduced into the test circuit by the circuit configuration, test leads, and components used. In general, an upper limit of 5 to 10 MHz can be expected.
- o The lower frequency limit of this approach is affected by the size of the inductor needed for resonance and the feed through capacitor needed for isolation. The capacitor is shown as having a value of 10 μf in Figure 7-37. The lower frequency value may be approximated using Table 7-1.

Table 7-1

Theoretical Low-Frequency Limit
for Parallel Injection Testing
(Q = 10; R = 50 ohms)

Filter Capacitance (μF)	Lower Frequency Limit (kHz)	Value of Resonating Inductor (μH)
1	32	25
10	3.2	250
20	1.6	500
30	1.1	750
40	0.8	1000
50	0.64	1250

- Be sure that filter input and output terminals are electrically isolated.
- Apply power bias gradually, starting with zero input power voltage to the test set-up.
- Be sure that the resonating series inductors can carry the power bias current.
- Protect the power bias circuitry from test filter short circuit failure with fuses or circuit breakers.
- Values of source and load impedance will depend in part on the shunting effects of the test signal current source and the test signal detector.
- Be sure the ammeter used to measure power bias current has full scale values at least equal to or preferably greater than rated filter load current. Some π-filters draw appreciable amounts of reactive current, and some allowance should be made for a meter safety factor.

7.6.4.5 Advantages/Disadvantages

- The method of resonating the input and output filter capacitors, gives "worst-case" performance data for the filter.
- The filter is tested under rated load current conditions.
- Good low frequency performance data can be obtained, limited only by the required value of the series resonating inductors and shunt by-pass capacitors.

- Accurate measurements of worst-case insertion loss can be made up to about 10 MHz.
- Load impedance may be varied to simulate actual operating conditions.
- Measurements at VHF and UHF frequencies cannot be accurately performed.
- A wide range of series resonating inductor values is required for general purpose testing.
- For a wide dynamic range (large values of filter attenuation), power booster amplification may be needed for the injected test current signal.

7.6.5 Series Signal Injection Test [11],[12]

7.6.5.1 Objective

The objective of this test is to perform a "worst-case" evaluation of a pi-section filter (resonating the filter input and output) without hard wire connection to the filter under test. As indicated previously, hard wired connections limit the range of values of source and load impedance under which the filter can be tested. In this test a signal injection current probe acts as a series current source for the test filter input. The test filter can be operated under rated power conditions during the test.

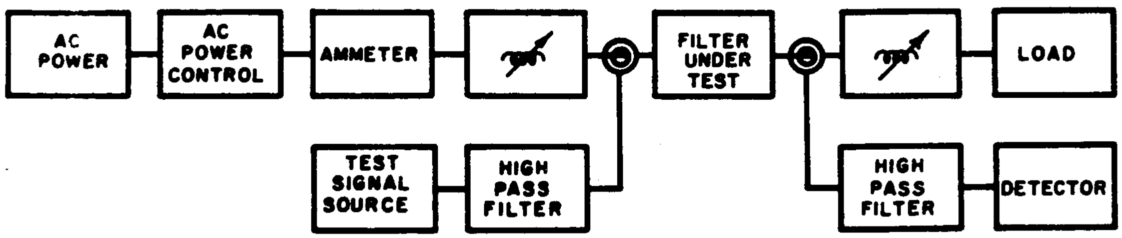
7.6.5.2 Test Set-Up

The instrumentation arrangement is given in Figure 7-38. Two current probes are used for injecting the test signal and for detecting the filter output.

The input test signal is connected through a high-pass filter to a current injection probe; the high-pass filter prevents the power bias signal picked up by the probe from reaching the test signal generator during power bias tests. A second probe couples current on the output side of the filter to a signal detector.

The large feedthrough capacitors are needed during power bias tests to bypass any external power-line impedance, forming essentially a short circuit to ground for the test signal. These capacitors should be at least five times as large as the test filter input capacitors, and should have very low self-inductance.

The power bias source for testing the filter under rated load can be either the standard ac line voltage or a dc power supply of variable output voltage and required maximum test current output. The equipment set-up shown in Figure 7-38 assumes a 115 V., 60 Hz power bias source, with power bias level control using a variable autotransformer. An ammeter is used to measure power bias current level. The test signal current source is a standard signal generator; a power booster amplifier may be needed if the filter under test has high insertion loss and the detector has low sensitivity.



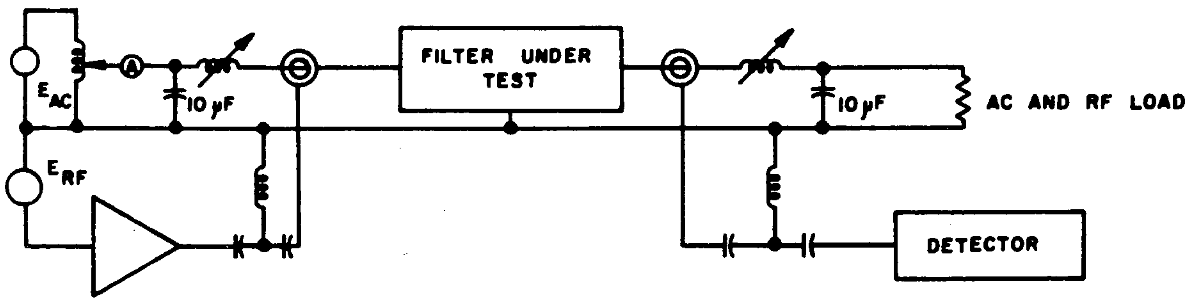


Figure 7-38 Series Signal Injection Block Diagram and Schematic Diagram

The input current probe should be designed to minimize the resistance inserted into the test circuit, and should be capable of carrying the maximum current to be injected as a test signal. Neither the input nor the output current probe should inject appreciable series impedance at the filter input or output, respectively. The variable inductors shown in Figure 7-38 are used to resonate the input and the output capacitors in the pi-filter under test.

7.6.5.3 Test Procedure

With the test equipment arranged as shown in Figure 7-38, the following steps can be performed to find the lowest insertion loss of a π-section filter:

- o Determine the value of the input and output shunt capacitance in the pi-filter to be tested.
- o Calculate the range of resonating series inductor values required for the range of test frequencies.
- o Place the correct series inductors in the test circuit and adjust for resonance at the test frequency.
- o Adjust the test signal generator at the test frequency, and adjust the detector for indication of the test signal.
- o With the filter input and output resonating, and with power bias applied, record the detector indication of the received test signal.

- Record the power bias level from the ac or dc power supply, and the injected test voltage level (E_2).
- Record the frequency used in the test.
- Remove the filter under test, and connect the test signal source directly to the load (filter-out condition).
- Adjust input signal voltage level to obtain the same indication on the detector as was obtained with the test filter installed.
- Record the test signal (E_1).
- Calculate the insertion loss of the filter using Equation (7-1).

7.6.5.4 Limitations and Precautions

- The lower frequency limit of the Series Signal Injection Method is dictated by the high value of inductance needed to resonate the filter capacitance, the feedthrough capacitor impedance needed to provide isolation, and the roll off of the transfer impedance curve of the current probes.
- The high frequency limit of this method is determined by the amount of inductive reactance inserted into the test circuit by the current probes. The maximum frequency value, based on input and output filter capacitances and probe-inserted reactance, may be approximated using Table 7-2.

Table 7-2

Upper Frequency Limit for
Worst-Case Testing for Series Injection

Filter Capacitance (µF)	Upper Frequency Limit (MHz)	Inserted Inductive Reactance (mΩ)
52	0.100	30
25	0.150	43
2.3	0.500	140
0.52	1.000	300

- Care must be taken to insure that the input and output sections of the filter under test are electronically isolated.

- Injected series impedance from the current probe transformation must be kept as low as possible, consistent with adequate probe transfer impedance. Refer to Section 7.6.6 of this publication for further details on the design parameters and characteristics of current injection probes.

- Be sure that the test load power bias dissipation capability is not exceeded.

- Apply power bias gradually starting with zero input power voltage to the test set-up.

- Be sure the series resonating inductors can carry the power bias current.

- Protect the power bias circuitry from test filter short circuit failure by fuses or circuit breakers.

- Be sure the ammeter used to measure power bias current has full scale values at least equal to or preferably greater than rated filter load current.

7.6.5.5 Advantages/Disadvantages

- Series signal injection using current probes eliminates hard-wired connections between the signal generator and the filter under test, and between the filter and the detector.

- Filter load impedances can be varied over a wider range.

- Possible damage to generation and detection equipment from the power bias source is practically eliminated.

- The series injection technique has a lower maximum frequency limit for measuring π-filters than does the parallel injection method. The current probes inject a series impedance into the test circuit, limiting the high frequency measurement range.

- Low frequency measurement limits are established by the large values of series inductors needed to resonate the filter capacitors.

- For wide dynamic range, current probes able to handle appreciable amounts of power must be available.

- Large by-pass capacitors will be necessary if testing to very low frequencies is required. Additionally, probe low frequency transfer impedance performance may also establish the low frequency test limit.

7.6.6 Current Injection Probes

7.6.6.1 General

The Series Signal Injection Test method just discussed (Section 7.6.5) for performing filter evaluations under load requires that a test signal be coupled to the filter input through a current probe device. Since commercially available probes are not normally employed in this way, some discussion is needed of the design characteristics of a current probe that can be used for signal injection applications.

There are five major design objectives for a current injection probe. These objectives are as follows:

- <u>Low insertion resistance</u> - The effective resistance the probe couples to the filter input circuit must be kept as low as possible to assure that filter resonances are not damped out, and that other measurement errors are not introduced.

- <u>Low insertion reactance</u> - The reactance introduced into the filter input circuit by the probe must also be kept to a minimum, since the higher this inductance, the lower will be the maximum frequency at which the Series Signal Injection test can be performed.

- <u>High transfer impedance</u> - The current injection probe is actually being used to inject a voltage in series with the filter under test and this voltage level is proportional to the probe transfer impedance.

- <u>High current handling capability</u> - The probe must be designed to handle the rated prime power load current (100 amperes or more in some cases). Also, the primary winding of the probe must be able to carry a current that will result in adequate dynamic range for the insertion loss test.

- <u>Good external shielding</u> - Sufficient shielding must be provided to assure that the probe is not the cause of the test signal by-passing the filter.

As in most designs, the above requirements are not mutually exclusive. For example, the higher the number of primary turns on the probe, the lower the insertion resistance. However, the transfer impedance will decrease with increased turns. Probe design must therefore be a compromise in performance objectives.

7.6.6.2 Sample Probe Design

One example of a current probe that has been used successfully in making Series Signal Injection filter tests is shown in Figure 7-38A [18]. The probe is constructed for a U core and ribbon conductors. The core is 12.9 sq. cm, in area, and consists of 2 mil grain-oriented silicon iron weighing 5.8 pounds. The upper half of the core is removable in order to allow the placement of a conductor within the probe aperture.

An airgap of 80 mils total is provided by 40 mil spacers in each side of the core.

The primary windings of the probe consist of two turns of flat ribbon conductors that are as wide as the vertical sides of the core. The two-turn windings on each core side are connected together by a one-inch wide strip that also forms a fifth turn. The ends of the conductor windings are brought out for connection to the driving source.

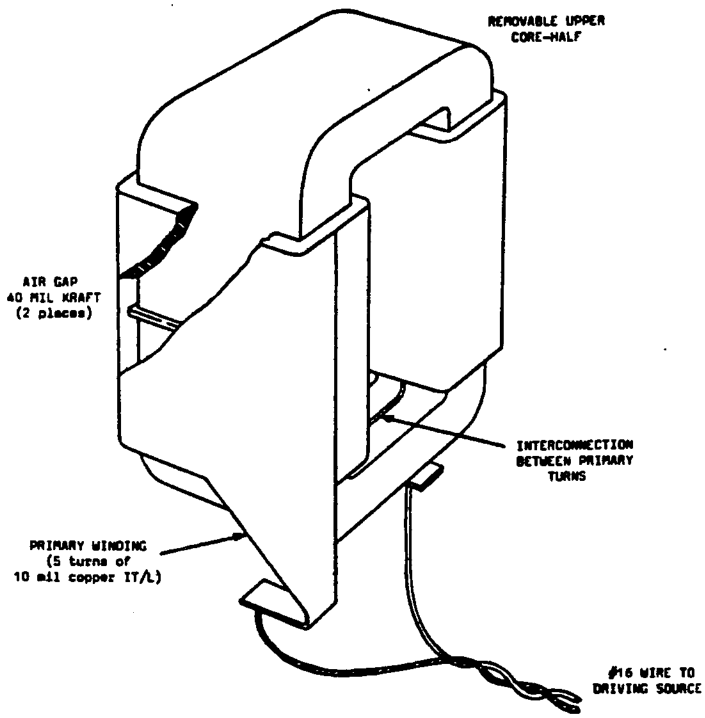


Figure 7-38A Sample Current Injection Probe

The unit is encapsulated in a high temperature epoxy compound and mounted in a brass electrostatic shield case. Heavy copper terminals are provided in order to facilitate high current excitation at low frequencies. All materials used in the construction are rated at least for +200 degrees C.

One test arrangement employed an audio oscillator and amplifier combination driving the probe. The amplifier was rated at 75 watts and had 4 and 8 ohm output terminals. The impedance reflected into the input circuit under test was found to be as shown in Figure 7-38B. This impedance level can be reduced by using a separate step-down transformer at the amplifier output, or by designing a probe having more primary turns[2].

The level of current that can be injected into the filter input at frequencies between 4 and 300 kHz can be raised significantly by capacitively tuning the probe. This increases the impedance of the capacitor-probe combination so that it more closely matches that of the driving source. Approximately 100 uf are required at 4 kHz, decreasing to 0.1 uf at 300 kHz. The capacitor is adjusted for a voltage peak across the probe windings.

2. See Reference 12 for an example of a current injection probe with 30:1 turns ratio.

Figure 7-38C shows the representative dynamic range capability of the probe when it is tuned as indicated above. The advantage of this technique at frequencies higher than 4 kHz is evident from the figure.

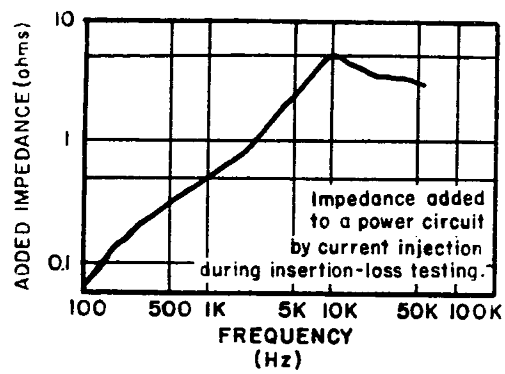


Figure 7-38B
Impedance Added by Sample Probe

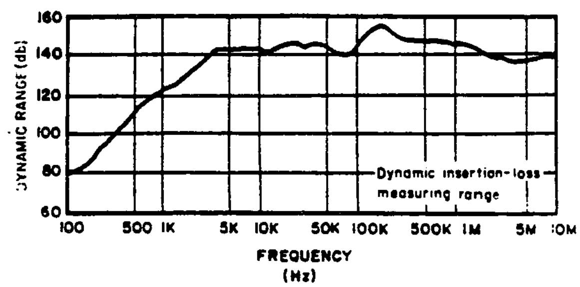


Figure 7-38C
Sample Probe Dynamic Range

A typical test setup using the current injection probe is illustrated in Figure 7-38D [19]. Note the shielding precautions taken to prevent the input signal energy from by-passing the filter under test.

It should be noted that the use of an input current probe is not restricted to the Series Signal Injection test, where the filter input and output are resonated to obtain a "worst-case" test result. Figure 7-38E shows the results of using an input signal probe to test a filter under different line conditions, for fixed terminating resistances. The partial saturation of the filter inductor cores under loaded conditions causes the reduced insertion loss above 10 kHz.

7.6.7 Leakage Current Test

7.6.7.1 Objective

The objective of this test is to determine the leakage current per phase of a power line filter when driven by any ac power source. This may be done at 60 Hz or 400 Hz as appropriate, with the filter separate from, or installed in, a GSE. The leakage current test may be performed in conjunction with the insertion loss test (Para. 7.6.2) or EMI tests (per MIL-STD-462). See paragraph 7.8.j for safety considerations.

7.6.7.2 Test Set-Up

The test configuration is shown in Figure 7-38F. Three alternatives for measuring the leakage current are shown. One is to use a millivolt-meter (with negligible internal impedance) inserted in the ground return to make the measurement directly. An alternative is to insert a known resistance (250 ohms or less) in the ground return, use a millivoltmeter to measure voltage drop across the resistor, and calculate the leakage current.

1. Signal Generator, such as HP-200-AB, AN/URM-25F
2. Power Amplifier, McIntosh 75 watts
3. RF Voltmeter, HP-400-E
4. Variable Capacitor, 0 to 110 uF
5. Current Injection Probe
6. Variable resistive load bank
7. Power filter being measured
8. Clamp-on power current probe meter
9. RF current probe, 10 Hz to 15 MHz, Genistron GCP-5120
10. Isolation pad, 10 dB
11. RF Field Strength meter

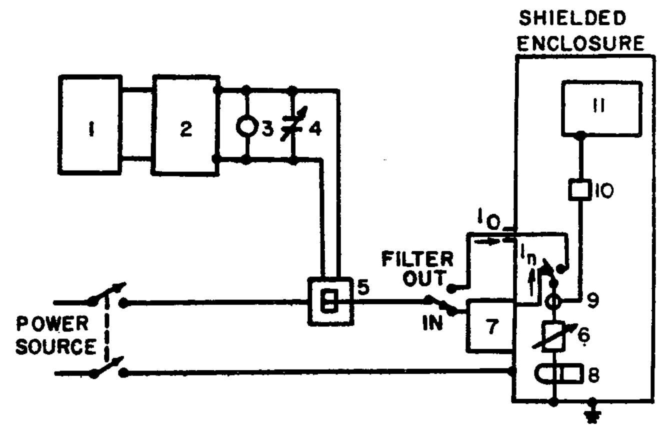


Figure 7-38D
Filter Test Block Diagram

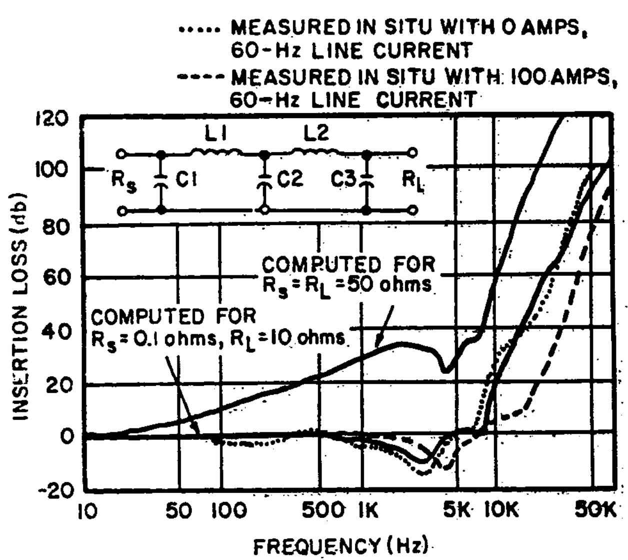


Figure 38E
Insertion Loss Data Using
Current Injection Probe

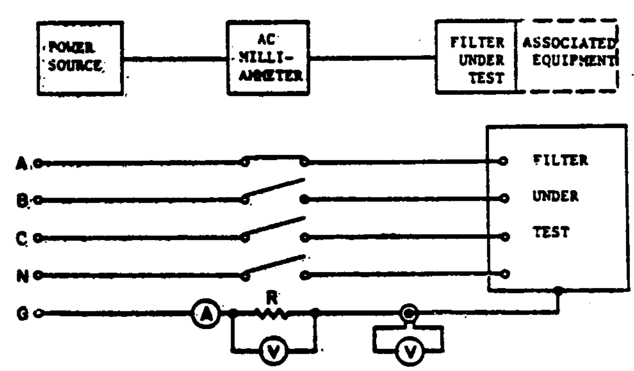


Figure 7-38F
Leakage Current Test
Block Diagram and Schematic Diagram

Another alternative is to put a current probe clamp around the ground return, measure voltage out of the probe with a millivolt-meter, and calculate the leakage current from the probe calibration data. One phase of the power input is energized while the other two phases and the neutral are open circuited.

If the filter is tested separately, the power source should provide a waveform that has the characteristics specified in Table 7-3 (beginning on Page 7-62. The filter may be terminated in the GSE or may be short circuited during the test. True RMS voltmeters should be used for voltage measurements, or individual frequency components can be recorded separately and combined on an RMS basis.

7.6.7.3 Test Procedure

With the equipment arranged as shown in Figure 7-38F, perform the following steps:

Energize one input to the filter and determine the appropriate ammeter or voltmeter reading.

Convert voltmeter reading to current value using resistance value or clamp calibration factor.

Repeat above steps for other inputs to the filter and add the results vectorially.

7.7 FILTER INSTALLATION AND MOUNTING

When filters are used, it is absolutely necessary to follow certain installation guidelines if good results are to be obtained. The RF impedance between case and ground must be kept as low as possible. Otherwise, the filter insertion loss will be seriously degraded at the higher frequencies. Effective separation of input and output wiring is mandatory because the radiation from wires carrying interference signals can couple directly in output wiring, thus circumventing and nullifying the effects of shielding and filtering. If complete isolation is affected between input and output, filter insertion loss will approach the design figure.

Both filter mounting guidelines given above can be readily satisfied by the use of bulkhead mounted feedthrough filters such as shown in Figure 7-39. A wide variety of these filters is available as off the shelf items from a number of manufacturers that can satisfy most filtering requirements.

Figure 7-40 shows the preferred method of mounting feedthrough filters in equipment that must be used in a high level electromagnetic environment such as exists on an aircraft carrier flight deck. Here the filters are mounted in a metal enclosure behind the front panel known as a "doghouse" that provides excellent isolation of filter input and output terminals.

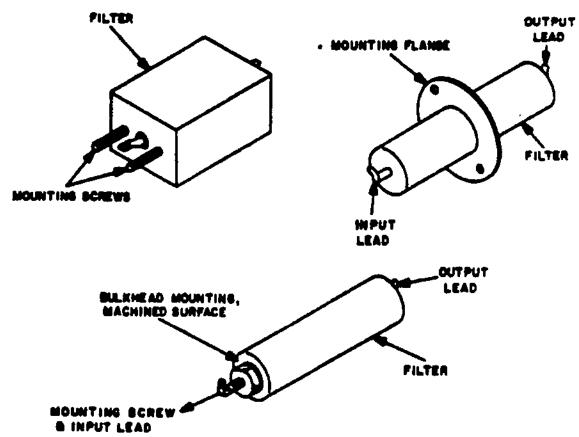


Figure 7-39 Typical Feed Through Filters for Bulkhead Mounting

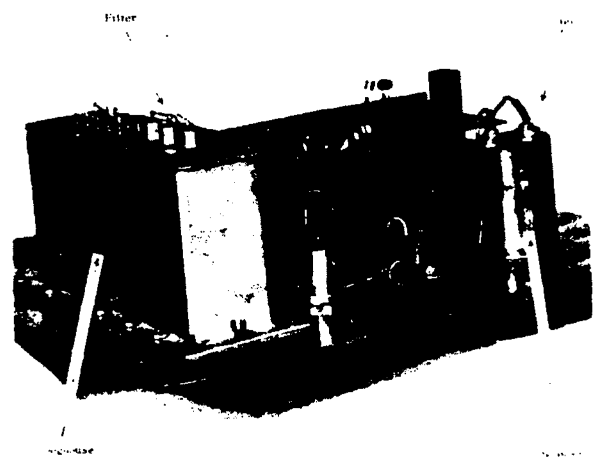


Figure 7-40 Example of Filter "Doghouse"

This enclosure can be constructed with an access panel on one of the sides. However, the panel should be attached to the doghouse using an RF gasket following the manufacturers recommendations for obtaining an effective seal (see Section 4.5.2).

Another effective filtering method uses a filtered pin connector. This arrangement, called a lossy connector, was shown in Figure 7-19. In this case the EMI filter is an integral part of the connector. The filtered pin connector will generally require less equipment space for mounting than will a connector/filter/doghouse assembly.

7.8 SPECIFYING FILTERS

In designing or selecting a filter for a particular application, many parameters must be taken into account if the filter is to be effective. Its insertion loss versus frequency curve is obviously the primary characteristic that determines the suitability of the filter for a particular EMI application. However, other electrical and mechanical requirements must also be designated. These include the following:

a. IMPEDANCE MATCHING - The elements of the filter must be chosen so that the impedance network matches the line into which it is inserted. This is especially true of transmission lines, so that the filter does not impair the normal function of the equipment at both ends of this line. When a filter must be installed in a circuit where its source impedance and/or load impedance is either not known or may vary over a relatively wide range, it may be desired to terminate the filter into fixed impedances to stabilize its performance.

b. VOLTAGE RATING - Consider the voltage rating on the filter, particularly if used on power lines. Under some conditions, the voltage may deviate by a large amount from its normal value. In addition, short duration pulses whose amplitudes are well above rated line voltage may be on the power circuits, both AC and DC. The filter voltage ratings must be sufficient to provide reliable operation under the extreme considerations expected.

c. VOLTAGE DROP - Determine the maximum allowable voltage drop through the filter and design accordingly.

d. CURRENT RATING - Current rating should be for a maximum allowable continuous operation of the filter. Calculate the current rating for filter elements, such as capacitors, inductors, and resistors. Whenever possible, the current rating of filters should be consistent with the current rating of the wire, circuit breakers, or fuse with which the filter will be used. A filter with a higher current rating than the circuit in which it is installed will often add a weight and space penalty. A filter with a lower rating will have poor reliability and maybe a safety hazard. The safety used in rating filters should also be consistent with those used for other circuit components.

e. FREQUENCY - Consider both the operating frequencies of the circuit and the frequencies to be attenuated. In general, the cost of a filter rises rapidly as the required rate of skirt falloff goes up, so be careful in identifying insertion loss versus frequency needs. Also, be sure to select the appropriate type of filter for your needs - you might be better off with a band-reject filter to reduce the level of a single close-in narrowband source, than to use a sharp falloff low-pass filter.

f. INSULATION RESISTANCE - The insulation resistance of the filter may vary during the life of the filter. Determine the maximum allowable variation of this resistance for proper filter operation, and design accordingly.

g. SIZE AND WEIGHT - Size and weight may be important in some GSE filter applications. When space is at a premium, adding or subtracting various filter elements may be traded against reduced size and weight of filter. Filter manufacturers are fairly flexible in being able to provide a wide choice in shape of the filter unit, its method of mounting, and the methods of making connections.

h. TEMPERATURE - The filter must be able to withstand the environmental operating ranges of the equipment in which it will be used. In most cases, the GSE requirement will probably necessitate a -65 to +85 degree C temperature range.

i. RELIABILITY - Filter component reliability must be commensurate with the equipment reliability requirement, and should be high relative to other equipment components. This is primarily dictated by the fact that faults in EMI filters may be somewhat more difficult to locate than faults in other components.

As an aid to filter procurement, Table 7-3 is provided. This type of form is similar to that contained in the Appendix of MIL-F-18327C, but is applicable both MIL-F-18327C [14] and MIL-F-15733E [15] requirements. It can be useful to the GSE designer in establishing his filter requirements, whether they are to be met in-house or by an outside source. The designer who must specify EMI filter requirements should familiarize himself with MIL-F-18327C and MIL-F-15733E (as well as MIL-STD-202D [16], not only as far as Table 7-3 is concerned, but to obtain a clear understanding of the requirements that must be imposed on filters. To make filter design easier, an article entitled "7-Element Chebyshev Filters Using Standard Value Capacitors," by E.E. Wetherhold, from the 1981 ITEM appears as Addendum #2 at the end of Section 7.

j. LEAKAGE CURRENT (POWER LINE FILTERS) - The capacitance between individual inputs to a power line filter and the filter case should be minimized to control leakage currents on the ground return. Low leakage will prevent shock hazards to personnel if the filter case should become ungrounded, and will also reduce common-mode interference (hum) as well as prevent inadvertent tripping of ground fault indicating devices. Maximum filter leakage current in a single equipment should not exceed 30mA per phase (5mA per phase in test equipment) in order to comply with existing specifications. This translates to 0.6 μf and 0.1 μf (0.1 μf and 0.016 μf for test equipment) of input capacitance per phase at 60 Hz and 400 Hz, respectively. Where filtering requirements cannot be met without exceeding the input capacitance limits, an isolation transformer should be used to isolate the equipment from the power source. Other means such as grounding the powerline neutral at the filter case should be explored. Paragraph 7.6.7 outlines a test procedure to test for leakage current.

Table 7-3
Filter Information Sheet (Cont'd)

Duration of test voltage (if other than 60 seconds) _____ seconds.
Points of application of test voltage _____.
At reduced barometric pressure: Method 105, MIL-STD-2020.
Test condition B _____. Other (specify) _____.
Insulation resistance: Method 302, MIL-STD-2020.
Test condition A _____ Other (specify) _____.

Insulation resistance: Method 302, MIL-STD-2020.
Test condition A _____ Other (specify) _____.
Points of measurement: Between terminal and mounting _____, between terminal and case _____, Other (specify) _____.

Overload: Paragraph 4.6.10, MIL-F-15733E _____.

Stability at temperature extremes: Paragraph 4.7.6, MIL-F-18327C.

Life: Method 108, MIL-STD-2020.
Test condition B _____, Test condition D _____, Test condition F _____. Duration of test if other than test condition B,D, & F _____ hrs.

Operating conditions: Test potential _____, Duty cycle _____, Load _____, Test Temp. _____, Tolerance _____.

Temperature rise: Paragraph 4.7.9, MIL-F-18327C _____. Paragraph 4.5.4, MIL-F-15733E _____.

Vibration: Low frequency _____, Method 201, MIL-STD-2020. Load condition _____.

High frequency _____, Method 204, MIL-STD-2020. Load condition _____.

Shock: Method 205, MIL-STD-2020, Paragraph 4.7.12, MIL-F-18327C.

Temperature cycling: Method 104, Condition A, MIL-STD-2020, and Paragraph 4.5.15.2, MIL-F-15733E _____.

Immersion: Method 104, MIL-STD-2020, Test condition A _____, B _____, C _____.

Moisture resistance: Method 106, MIL-STD-2020, Paragraph 4.7.14, MIL-F18327C _____.

Flammability (grades 5 and 7): Method 111, MIL-STD-2020.

Salt Spray: Method 101 of MIL-STD-2020.

Electrical characteristics

Impedance: Input _____ ohms. Output _____ ohms. Transfer _____ + _____ ohms (if applicable). Load _____ ohms.

Table 7-3
Filter Information Sheet (Cont'd)

Resonating capacity _____ pf + _____ pf (if applicable).

Capacitance to ground: Method 305, MIL-STD-2020 and 4.6.3, MIL-F-15733E:
_____ pf at _____ kHz, _____ pf at _____ kHz,

Signal input voltage _____ volts.

Current rating: AC, _____ amps DC, _____ amps

Voltage rating: AC, _____ volts DC, _____ volts.

Duty Cycle (if intermittent): Time on _____ Time off _____

Maximum allowable voltage drop: _____ volts.

Reference frequency _____ (Hz) _____ (kHz) _____ (MHz) _____.

Insertion loss (at reference frequency) _____ dB (min.).

Discrimination Frequency Range _____ (Hz) _____ (kHz) _____ (MHz)	Min (dB)	Max (dB)

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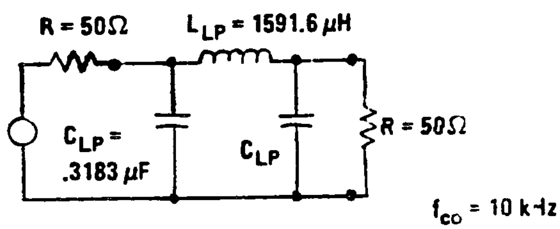
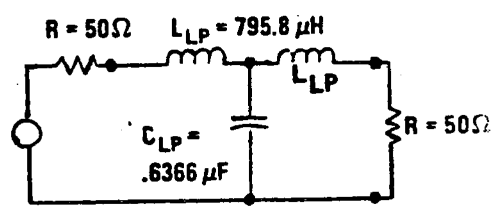
CHAPTER 7 ADDENDUM #1

PROCEDURE FOR TRANSFORMING A LOWPASS FILTER INTO ITS EQUIVALENT HIGHPASS FILTER



LOWPASS π FILTER



LOWPASS T FILTER
(BUTTERWORTH DESIGN)

The following equations are used when transforming lowpass filter component values into equivalent highpass filter component values:

$$C_{HP} = 1/[L_{LP} \cdot \omega^2] \text{ and } L_{HP} = 1/[C_{LP} \cdot \omega^2]$$

where

C_{HP} = HP filter capacitance in Farads,

L_{HP} = Equivalent LP filter inductance in Henries,

L_{HP} = HP filter inductance in Henries,

C_{LP} = Equivalent LP filter capacitance in Farads, and

ω = $2\pi f$ where f = cutoff frequency in Hz.

For example, for a cutoff frequency of 10 kHz, the lowpass π and T filters shown above are transformed into equivalent highpass π and T filters as follows:

$$\omega = 2\pi f = 2\pi \cdot 10 \text{ kHz} = 2\pi \cdot 10^4 \text{ and } \omega^2 = 4\pi^2 \cdot 10^8$$

CHAPTER 7 ADDENDUM #1 (Cont'd)

FILTER TRANSFORMATION CALCULATIONS

LOWPASS π TO HIGHPASS π

$C_{HP} = 1/[L_{LP} \cdot \omega^2] = 1/[(5000/\pi) \cdot 10^{-6} \cdot 4\pi^2 \cdot 10^8]F$

$= 1/(.5\pi \cdot 4 \cdot 10^6) = 1/(2\pi \cdot 10^6) = (1/2\pi)\ \mu F$

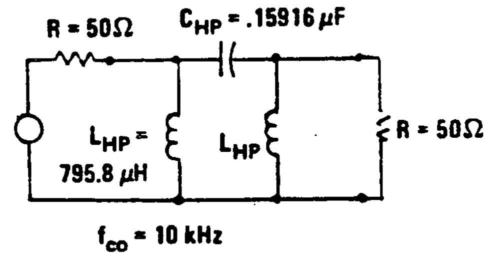
$= 0.15916\ \mu F$

$L_{HP} = 1/[C_{LP} \cdot \omega^2] = 1/[(10^{-6}/\pi) \cdot 4\pi^2 \cdot 10^8]H$

$= 1/(400\pi) = (2500/\pi)\ \mu H$

$= 795.8\ \mu H$

EQUIVALENT HIGHPASS π FILTER



LOWPASS T TO HIGHPASS T

$C_{HP} = 1/[L_{LP} \cdot \omega^2] = 1/[(2500 \cdot 10^{-6}/\pi) \cdot 4\pi^2 \cdot 10^8]\ F$

$= 1/(\pi \cdot 10^6) = (1/\pi)\ \mu F$

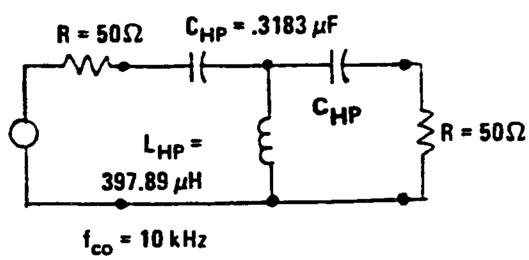
$= 0.3183\ \mu F$

$L_{HP} = 1/[C_{LP} \cdot \omega^2] = 1/[(2 \cdot 10^{-6}/\pi) \cdot 4\pi^2 \cdot 10^8]\ H$

$= 1/(800\pi) = (1250/\pi)\ \mu H$

$= 397.89\ \mu H$

EQUIVALENT HIGHPASS T FILTER



CHAPTER 7 ADDENDUM #2

7-ELEMENT CHEBYSHEV FILTERS USING STANDARD VALUE CAPACITOR[1]

INTRODUCTION

An article in [1980's] ITEM (1) discussed the design of passive LC 7-element Chebyshev filters .... The design and construction of these filters were greatly simplified by the use of tables of pre-calculated designs requiring only standard value capacitors. Thirty lowpass and thirty highpass 50 ohm designs were tabulated with the filter cutoff frequencies covering the 3-50 kHz range in small increments so virtually any desired cutoff frequency could be approximated. After this article was published, the computer program used to generate the filter tables was improved and expanded, and these improvements were incorporated in several recently published articles (2,3,4). In this year's ITEM article, the improvements developed over the past year are included in the two comprehensive filter tables presented in this article. One table lists lowpass filters and the other lists highpass filters, and each consists of 178 designs over the 1 to 10 MHz range.

Because the computer program used to generate last year's filter tables was not completely developed in time to meet the publication deadline, all of the possible combinations of standard-capacitor values were not included. This omission has been corrected in the new tables presented in this article. All possible capacitor values up to a ratio of 2-to-1 [equivalent to a filter reflection coefficient (RC) of 8.1%] are included for all feasible combinations of the twenty-four standard-capacitor values of the 5% series [10, 11, 12, 13, 15, 16, 18, 20 etc.] This permits a total of 178 lowpass and 178 highpass designs to be realized between reflection coefficients of zero (equivalent to a Butterworth) and 8.1%.

Because the maximum permissible impedance variation in a 50-ohm detection system is ± 10 ohms (corresponding to a maximum VSWR of 1:2), the 8.1% RC value was selected as the upper limit because it corresponds to a filter VSWR of 1:17. Also, for an RC = 8.1%, one of the two capacitor values of the lowpass and highpass filters is exactly double that of the other. In some cases this may facilitate the filter construction if six capacitors having the same value are available.

Although the tabulated designs are based on a 50-ohm impedance system over the 1 to 10 MHz range, filters with standard-value capacitors for any impedance level and cut off frequency are easily determined from the filter tables by using simplified scaling procedures. Because of the comprehensiveness of the lowpass and highpass tables, and because of the convenience of needing only standard-value capacitors, these pre-calculated designs should be used whenever equally-terminated 7-element C-input/output Chebyshev filters are required.

Use of the Lowpass tabulation (Table 1) will be discussed, and these comments, where applicable, will also apply to the Highpass tabulation (Table 2).

1. Adapted from 1981 ITEM.

Table 1. Lowpass Filters

Figures 1(A) and 1(B) are the schematic diagram and attenuation response corresponding to the design parameters and component values listed in Table 1. Note that the C1, C3, C5, and C7 designations in Figure 1(A) indicate the physical placement within the filter of the capacitors whose values are tabulate under the headings of C1,7 and C3,5. Because C1 and C7 are identical, the values of C1 and C7 are listed in a single column. The same is true for capacitors C3 and C5. The inductor circuit placement and tabulation are presented in a similar manner.

For easy identification, each filter design has a corresponding filter number, and these numbers are listed in the first and last columns of both tables. In the following discussion, a particular design will be referred to by an "LP" or "HP" designation (signifying lowpass or highpass) followed by the filter number.

The frequencies in the attenuation response curve [Figure 1(B)] correspond to the attenuation levels at A_p, 1,3, 6, 30 and 50-dB and they are listed (in MHz) under the appropriate headings of Table 1. The A_p-dB frequency is listed because this is the value used to calculate all the other parameters for a particular design. The 1-dB frequency is listed because one may wish to know at what frequency the attenuation begins to become significant. The 3 and 6-dB frequencies are listed because these frequencies are commonly used as a cutoff frequency specification. The 30 and 50-dB frequencies give the user an indication of the filter stopband performance.

The parameter that defines a particular Chebyshev design is the reflection coefficient (RC), and this parameter is listed in percent under the "R.C. (%)" heading of the tables 1 and 2. An exponential format is used to express this parameter because of its extremely wide variation. For example, the smallest value of RC is 0.554 x 10^{-6}% (see LP#44), and this is virtually identical to a Butterworth design. The maximum RC is 8.1%. By selecting particular values of RC and A_p-cutoff frequency, it is possible to generate a filter design that uses only standard-value capacitors. This accounts for the apparently odd values of RC and A_p-cutoff frequency that are listed in the tables.

Because of the small increments in the 3-dB cutoff frequency from one design to the next, virtually any required cutoff frequency in the 1-10 MHz decade can be obtained.

Although the cutoff frequencies in the A_p, 1, 3 and 6-dB columns generally increase with increasing filter number, there is an occasional regression in the cutoff frequency as the RC values change from a high value in one group to a low value in the next group. Thus, when searching for a particular cutoff frequency, be sure to examine at least two consecutive groups to find a design having the desired cutoff frequency. Also note that the same cutoff frequency may be obtained with different designs. For example, designs LP #14 and LP #17 both have a 3-dB cutoff frequency of 1.26 MHz, but filter #14 has a 5.78% RC while #17 has a 0.233% RC.

Although filter #14 has a slightly steeper attenuation response than #17, it is not significantly better than #17. If a choice had to be made between these two designs, it might be based on the fact that design #17 is more convenient to assemble because its capacitor values (1500 and 4700 pF) are more commonly available than the values of design #14 (2400 and 5100 pF).

Table 2. Highpass Filters

Most of the explanation of the lowpass filters in Table 1 is applicable to the highpass filters in Table 2. In both cases, the 3-dB cutoff frequencies of the tables start and end at approximately one and ten megahertz. Another similarity is that if the same capacitor values are used in the lowpass and highpass designs the RC value of both designs will be identical. For example, filter design LP #154 (RC = 1.0%) has capacitor values of 300 and 820 pF for C1,7 and C3,5, while design HP #138 has the same capacitor and RC values, except the values of C1,7 and C3,5 are reversed. Other differences are that for each group of highpass designs, the RC values start high (with a maximum of 8.1%) and gradually decrease to a minimum value, whereas the reverse is true in the lowpass table. Also, as the highpass filter attenuation increases, the corresponding frequencies decrease, whereas in the lowpass table the attenuation and frequency both increase with increasing attenuation. Other minor differences and similarities can be found between the two tables, and these differences and similarities should be recognized as being normal.

Verification Of Filter Tables

Before the reader uses tabulated data, such as presented in Tables 1 and 2, some attempt should be made to first check the correctness of the data. If the data of one design can be demonstrated to be correct, then it is reasonable to expect that all the designs will be correct because they were generated by the same computer program.

The attenuation versus frequency data can be quickly checked by estimating the filter attenuation slope. It is common knowledge that a Butterworth attenuation response has a slope of 6-dB/octave for each element, and since each filter in the tables has seven elements, the attenuation slope should be near 7 x 6-dB/octave = 42-dB/octave for those designs approximating a Butterworth design (RC values less than 0.1 %). For higher RC values, the slope will be a few dB greater than 42-dB/octave. For example, referring to design LP #44 (RC = $.554 \times 10^{-6}$% if the 30 and 50-dB frequencies are plotted on semi-log paper, the slope between 2.84 and 3.95 MHz can be verified to be 42.5-dB/octave, and this is a satisfactory check on the attenuation versus frequency parameters.

In a similar manner, the correctness of the capacitor and inductor values can be independently verified. This is done by using previously published normalized C and L values. By scaling the normalized values to the impedance and A_p-cutoff frequency used in a particular design, it can be demonstrated that the independently calculated values will be identical to the tabulated values in Tables 1 and 2. Previously published normalized data for the Chebyshev filter are available from several authoritative sources (5, 6), and either of the references provide normalized C and L data for reflection coefficients of 1, 2, 3, 4, 5, 8 and 10 percent and higher. However, for a meaningful comparison to be made, it is necessary to use a tabulated design that has an RC value that closely approximates one of the published RC values. Design LP #154 has an RC value of 1.00%, and this matches one of the published RC values. Because of this, it will be used to demonstrate the validity of the tabulated computer-calculated C and L values. The A_p-cutoff frequency of LP #154 is 5.68 MHz, and this parameter will be used in the scaling process required to transform the normalized filter component values (normalized for an A_p-cutoff frequency of one radian per second and an impedance of one ohm) to the A_p-cutoff frequency and impedance level (5.68 MHz and 50 ohms) of filter design LP #154. From the normalized tables of either Saal or Zverev, the normalized component values for C1 and C3 are 0.5354 and 1.464 farads. The normalized values for L2 and L4 are 1.179 and 1.500 henries. Based on a new impedance and cutoff frequency of 50 ohms and 5.68 MHz, the C and L scaling factors are:

$$C_s = 1/R\omega, \text{ where } R = 50 \text{ ohms and } \omega = 2\pi f_{Ap},$$

$$\text{with } f_{Ap} = 5.68 \times 10^6 \text{ and } \omega = 35.6885 \cdot 10^6.$$

$$C_s = 1/1(1784.425\text{x}10^6) = 560.4 \times 10^{-12}.$$

$$L_s = R/\omega = 50/(35.6885 \times 10^6) = 1.401 \times 10^6.$$

The normalized C and L values are scaled to the desired impedance and cutoff frequency by multiplying them by their corresponding scaling factors:

$$C1 = 0.5354\text{F} \times (560.4) \times 10^{-12} = 300 \text{ pF}$$
$$C3 = 1.464\text{F} \ (560.4) \times 10^{-12} = 820 \text{ pF}$$

$$L2 = 1.179\text{H} \times (1.401) \times 10^{-6} = 1.65 \ \mu\text{H}$$
$$L4 = 1.500\text{H} \ (1.401) \times 10^{-6} = 2.10 \ \mu\text{H}$$

These independently calculated values of C and L are identical with the tabulated values of LP #154; consequently, the correctness of the tabulated values is verified. Since the same computer program was used to calculate all the designs, the correctness of all the designs is cofirmed.

HOW TO USE THE TABLES

For 50-ohm Filters

A suitable lowpass or high pass filter is selected by entering the appropriate table under the 3 or 6-dB column heading, and by searching for the desired cutoff frequency. Because the TEMPEST specification is relatively lenient regarding the permissible range of the 6-dB cutoff frequency, several different designs will usually be suitable. Examine the component values, and , if possible, select the design which has convenient capacitor values and a low value of RC.

If the desired cutoff frequency is outside the tabulated 1-10 MHz frequency range, then a simple procedure is used to scale the tabulated data to the desired frequency decade. This involves shifting the decimal points of the frequency and component values to the right or left. To find lowpass or highpass filter designs for the 10 - 100 or 100 - 1000 MHz range, multiply all tabulated frequencies by 10 or 100, respectively, and divide all C and L values by the same number. The RC values remain unchanged. For example, to find a lowpass filter having a 10.4 MHz 3-dB cutoff frequency, multiply all the tabulated frequencies by ten and divide the component values by ten. For this particular application, design LP #1 is suitable, and the corresponding C and L component values are 150 and 560 pH and 1.011 and 1.526 μH.

To scale the filter tables to cover the 1 - 10 kHz, 10 - 100 kHz or .1-1 MHz decade, divide the tabulated frequencies by 1000, 100 or 10 respectively, and multiply the component values by the same number. For example, to find a lowpass filter having a 1.79 kHz 3-dB cutoff frequency, change Table 1 to a 1-10 kHz decade by dividing all tabulated frequencies by 1000 and multiplying all component values by the same number. In this particular case, design LP #46 would be suitable, and the C and L values are 1.0 and 3.3 μF and 6.30 and 8.80 mH. Regardless of the scaling factor used, the capacitor values will always be standard as long as the capacitance does not exceed one microfarad.

For Filter Impedance Levels Other Than 50 ohms

Although filters are most frequently used with 50-ohm receivers, there may be occasions when the output signals of transmitters must be filtered to attenuate harmonics. In this case, impedances other than 50 ohms may be encountered. The ability to quickly and accurately design filters using standard-value capacitors for any impedance level is very useful, and a simple scaling procedure using Tables 1 and 2 will be explained as follows:

(1) For any desired filter impedance in ohms Z_x, select the desired 3-dB cutoff frequency in Hz, F_3^{x}

(2) Calculate $F_3^{50} = F_3^{x} (Z_x/50)$

(3) From Tables 1 or 2, select the design (lowpass or highpass) that has a 3-dB cutoff frequency closest to the F_3^{50} value calculated in (2). Record the tabulated capacitor values of the selected design as they will be the C1,7 and C3,5 values of the new filter. Also note the inductor values.

(4) Calculate the inductor values of the new filter by multiplying the inductor values noted in (3) by the square of $Z_x/50$.

(5) Calculate the frequencies of the new filter at A_p, 1, 3, 6, 30 and 50-dB by multiplying the tabulated frequencies of the selected filter by $50/Z_x$. This concludes the scaling procedure. An example will demonstrate the impedance scaling procedure.

Assume it is desired to find a 75-ohm lowpass filter having a 3.0 MHz 3-dB cutoff frequency. The calculation steps previously explained follow:

(1) Z_x = 75 ohms and F_3^x = 3.0 MHz.

(2) F_3^{50} = 3(75/50) MHz = 4.5 MHz

(3) From Table 1, the lowpass design having F_3^{50} closest to 4.5 MHz is LP #115, and this design is selected for scaling. The C1,7 and C3,5 capacitor values of the new 75-ohm filter will be 360 and 1300 pF, respectively. The L2, 6 and L4 inductor values of 2.39 and 3.52 μH are noted.

(4) The 75-ohm filter inductor values are calculated:
L2,6 = 2.39 $(75/50)^2$ μH = 5.38 μH,
L4 = 3.52 $(75/50)^2$ μH = 7.92 μH.

(5) The A_p, 1, 3, 6, 30 and 50-dB frequencies for the new filter are calculated by scaling the corresponding frequencies of design LP #115 to the new 75-ohm impedance level. The scaling factor = $50/Z_x$ = 50/75 = 0.6667. Thus, for F_{Ap} of LP #115 = 2.17 MHz, the corresponding F_{Ap} of the new 75-ohm 3.0 MHz lowpass filter is 2.17(0.6667) = 1.45 MHz. In a similar way, the F1, 3, 6, 30 and 50-dB frequencies of the 75-ohm filter are 2.75, 2.99, 3.21, 4.71 and 6.47 MHz. Note that the desired 3-dB cutoff frequency was 3.0 MHz but the closest that could be obtained from Table 1 was a design with a 2.99 MHz cutoff frequency.

In a similar manner, highpass filters having impedance levels other than 50 ohms may be calculated. For example, if a 75-ohm 3-dB 3.0 MHz highpass filter is desired, design HP #19(F_3^{50} = 4.49 MHz) would be used, and the C1, 7 and C3,5 capacitor values would be 1500 and 390 pF, respectively. The L2,6 and L4 inductor values would be 3.08 and 2.00 μH for the 75-ohm filter. The F_{Ap} frequency of design HP #119 is 13.35 MHz, and the corresponding F_{Ap}, frequency of the new 75-ohm 3.0 MHz highpass filter is 13.35(.6667) MHz = 8.90 MHz. In a similar way, the F1, 3, 6, 30 and 50-dB frequencies of the new 75-ohm filter are 3.27, 2.99, 2.78, 1.86 and 1.35 MHz.

Capacitor And Inductor Recommendations For Filter Construction

Using Tables 1 and 2 eliminates most of the problems associated with the selection of suitable filter designs; however, the problem of component selection still remains. The capacitor type most suitable for general purpose filtering applications above 100 kHz is the polystyrene type, and it is recommended for constructing the filter designs listed in Tables 1 and 2. Fortunately, the 2.5% tolerance of the Mallory Type SXM polystyrene capacitor makes it possible to use this part as purchased. Because of its tight tolerance there is no need to individually measure and select capacitors, as each capacitor is close enough to its nominal value so it can be used directly in any design requiring its nominal value. The type SXM capacitor is available in a capacity rating up to .056 μF; consequently, 50-ohm filters can be constructed using this capacitor and having a 3-dB cutoff frequency as low as 100 kHz. For filters having a cutoff frequency below 100 kHz and requiring capacitors larger than .056 μF, the Mylar[2] capacitor is more attractive than polystyrene because of its smaller size and lower cost. However, the Mylar capacitor usually has a tolerance of ten percent, and each one should be measured and selected for its particular application.

The inductors required to build filters in the frequency range above 10 kHz are usually hand-wound for expediency. Up to about 100 kHz, molybdenum permalloy cores are used, and above 100 kHz, powdered iron cores are used. A detailed discussion of inductor construction suggestions is beyond the scope of this article, and the interested reader is referred to references (1), (7) and (8) for further information.

Lowpass and Highpass Filters Used to Get a Bandpass Response

In addition to the usual lowpass and highpass filtering application, most of the filter designs of tables 1 and 2 may be cascaded to achieve a bandpass response. This procedure is especially useful when the passband is required to be extremely wide (more than one decade of frequency). In this case, it is often more convenient to treat the design separately as individual lowpass and highpass filtering requirements instead of treating it as a bandpass filter design. The use of separate lowpass and highpass filters to achieve a bandpass response will seldom give any trouble as long as the passband is greater than an octave wide. Below this width, however, the design becomes increasingly critical because of interaction between the two filters.

Each filter will operate as expected if it is correctly terminated, but neither filter will see this condition unless the other one has a very constant terminal impedance. For this reason, any filters intended for connection in cascade should have low values of reflection coefficient and VSWR. An RC value of less than 5% is suggested [See reference (9), p. 51].

If a standard bandpass filter design is required to obtain a passband less than an octave wide (but not less than about 10% of the center frequency), the standard lowpass-to-bandpass transformation procedure should be used. See reference (10) for an example of this procedure. Of course, the lowpass designs of Table 1 could be used to select a lowpass design suitable for transformation to a bandpass response, but the number of components (7 capacitors and 7 inductors) would be excessive for TEMPEST

2. DuPont Trademark.

applications. A more practical bandpass design can be obtained by transforming a four or five element lowpass filter, and in this case, the number of elements will be either eight or ten, and therefore more reasonable.

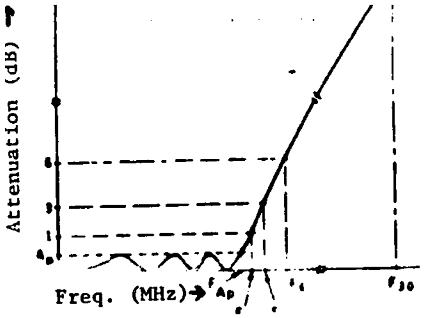
CONCLUSIONS

It is important that the test engineer be able to quickly and conveniently design and assemble low and high-pass filters over a wide range of cut-off frequencies. The precalculated 50-ohm filter designs presented in this article provide a wide selection of designs for virtually any cut-off frequency between 1 and 10 MHz.

By using these designs the likelihood of calculation error is eliminated, and the filter construction is simplified because each design requires only standard-value capacitors. Although the designs are based on equal 50-ohm terminations in the 1 to 10 MHz range, designs for other termination resistances and other frequency ranges are easily calculated with simple scaling procedures while the advantage of standard-value capacitors is included in the new design.

Because most of the pre-calculated designs have a reflection coeffient less than four percent, the filter terminal impedance is relatively constant across the passband. This means that the lowpass and highpass filters can be cascaded to achieve a bandpass response with no difficulty as long as the passband is not less than one octave wide.

The design procedures discussed in this article should be useful to those responsible for performing TEMPEST testing, or for designing and constructing filters. Any comments, suggestions, or criticisms will be appreciated and they should be addressed to the author, Ed Wetherhold, Honeywell Inc., POB 391, Annapolis, MD 21404.

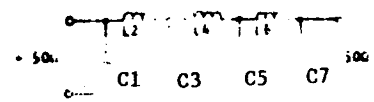


Attenuation Response Curve

(B) Typical Attenuation Response

CHAPTER 7 ADDENDUM #2

Lowpass Filter Schematic Diagram



(A) Schematic Diagram

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
1	.430	.951	1.04	1.12	1.65	2.28	.455E-02	1500	5600	10.11	15.26	1
2	.559	.966	1.05	1.12	1.63	2.23	.339E-01	1600	5600	10.43	15.08	2
3	.714	.991	1.06	1.13	1.60	2.17	.255E-00	1800	5600	10.92	14.77	3
4	.812	1.01	1.08	1.14	1.58	2.11	.806E-00	2000	5600	11.23	14.44	4
5	.886	1.04	1.10	1.15	1.56	2.08	.178E-01	2200	5600	11.36	14.04	5
6	.949	1.07	1.12	1.17	1.56	2.06	.327E-01	2400	5600	11.31	13.54	6
7	1.04	1.12	1.16	1.21	1.58	2.06	.663E-01	2700	5600	10.90	12.57	7
8	.302	1.03	1.13	1.22	1.83	2.54	.173E-03	1300	5100	8.98	14.05	8
9	.663	1.07	1.15	1.23	1.78	2.43	.630E-01	1500	5100	9.61	13.66	9
10	.755	1.08	1.16	1.24	1.76	2.39	.184E-00	1600	5100	9.86	13.51	10
11	.881	1.11	1.18	1.25	1.73	2.33	.720E-00	1800	5100	10.20	13.19	11
12	.972	1.14	1.20	1.26	1.72	2.28	.176E-01	2000	5100	10.34	12.79	12
13	1.05	1.17	1.23	1.28	1.71	2.26	.341E-01	2200	5100	10.29	12.29	13
14	1.12	1.21	1.26	1.31	1.72	2.26	.578E-01	2400	5100	10.04	11.66	14
15	.336	1.12	1.23	1.32	1.99	2.75	.206E-03	1200	4700	8.28	12.94	15
16	.596	1.14	1.24	1.33	1.96	2.69	.143E-01	1300	4700	8.62	12.73	16
17	.842	1.18	1.26	1.34	1.91	2.58	.233E-00	1500	4700	9.14	12.41	17
18	.918	1.20	1.28	1.35	1.98	2.54	.500E-00	1600	4700	9.32	12.25	18
19	1.03	1.23	1.30	1.36	1.86	2.49	.147E-01	1800	4700	9.52	11.88	19
20	1.12	1.27	1.33	1.39	1.86	2.45	.312E-01	2000	4700	9.50	11.40	20
21	1.21	1.31	1.37	1.42	1.87	2.45	.560E-01	2200	4700	9.27	10.78	21
22	.378	1.22	1.34	1.45	2.17	3.00	.251E-03	1100	4300	7.58	11.83	22
23	.674	1.25	1.36	1.46	2.13	2.93	.185E-01	1200	4300	7.92	11.63	23
24	.836	1.27	1.37	1.46	2.10	2.87	.104E-00	1300	4300	8.20	11.47	24
25	1.03	1.31	1.40	1.48	2.06	2.77	.643E-00	1500	4300	8.58	11.15	25
26	1.10	1.33	1.41	1.49	2.04	2.73	.116E-01	1600	4300	8.68	10.97	26
27	1.21	1.38	1.45	1.51	2.03	2.69	.280E-01	1800	4300	8.71	10.51	27
28	1.31	1.43	1.49	1.55	2.04	2.68	.540E-01	2000	4300	8.50	9.91	28
29	.429	1.35	1.48	1.60	2.39	3.31	.314E-03	1000	3900	6.89	10.73	29

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
30	.771	1.38	1.50	1.61	2.35	3.22	.247E-01	1100	3900	7.22	10.53	30
31	.954	1.41	1.52	1.61	2.31	3.14	.138E-00	1200	3900	7.49	10.37	31
32	1.08	1.43	1.53	1.62	2.28	3.08	.397E-00	1300	3900	7.69	10.21	32
33	1.25	1.48	1.57	1.65	2.25	2.99	.151E-01	1500	3900	7.90	9.85	33
34	1.32	1.51	1.59	1.66	2.24	2.97	.244E-01	1600	3900	7.91	9.62	34
35	1.44	1.57	1.64	1.71	2.25	2.95	.516E-01	1800	3900	7.73	9.04	35
36	.369	1.46	1.60	1.73	2.60	3.60	.593E-04	910	3600	6.31	9.93	36
37	.791	1.49	1.62	1.74	2.55	3.50	.162E-01	1000	3600	6.62	9.74	37
38	1.02	1.53	1.64	1.75	2.51	3.41	.123E-00	1100	3600	6.89	9.58	38
39	1.17	1.55	1.66	1.76	2.47	3.34	.397E-00	1200	3600	7.10	9.43	39
40	1.28	1.58	1.68	1.77	2.45	3.28	.892E-00	1300	3600	7.23	9.26	40
41	1.44	1.64	1.73	1.81	2.43	3.21	.271E-01	1500	3600	7.29	8.82	41
42	1.52	1.68	1.76	1.83	2.43	3.20	.411E-01	1600	3600	7.22	8.54	42
43	1.66	1.77	1.84	1.90	2.47	3.22	.810E-01	1800	3600	6.86	7.83	43
44	.208	1.58	1.74	1.88	2.84	3.95	.554E-06	820	3300	5.73	9.14	44
45	.839	1.63	1.77	1.90	2.79	3.83	.130E-01	910	3300	6.04	8.94	45
46	1.09	1.66	1.79	1.91	2.74	3.73	.108E-00	1000	3300	6.30	8.80	46
47	1.27	1.70	1.81	1.92	2.70	3.64	.397E-00	1100	3300	6.51	8.64	47
48	1.40	1.73	1.83	1.93	2.67	3.57	.949E-00	1200	3300	6.64	8.47	48
49	1.51	1.76	1.86	1.95	2.65	3.53	.182E-01	1300	3300	6.69	8.26	49
50	1.68	1.85	1.93	2.01	2.65	3.49	.471E-01	1500	3300	6.58	7.72	50
51	1.77	1.90	1.98	2.05	2.68	3.50	.684E-01	1600	3300	6.40	7.73	51
52	.322	1.74	1.92	2.07	3.12	4.34	.623E-05	750	3000	5.23	8.30	52
53	.889	1.78	1.94	2.08	3.07	4.22	.978E-02	820	3000	5.47	8.15	53
54	1.21	1.83	1.97	2.10	3.01	4.10	.110E-00	910	3000	5.73	7.99	54
55	1.40	1.87	1.99	2.11	2.97	4.01	.397E-00	1000	3000	5.91	7.85	55
56	1.56	1.91	2.02	2.13	2.93	3.93	.102E-01	1100	3000	6.04	7.68	56
57	1.68	1.95	2.05	2.15	2.91	3.87	.204E-01	1200	3000	6.09	7.47	57
58	1.79	2.00	2.09	2.18	2.91	3.84	.351E-01	1300	3000	6.05	7.21	58
59	1.99	2.12	2.20	2.29	2.96	3.86	.810E-01	1500	3000	5.72	6.52	59
60	.452	1.94	2.13	2.30	3.47	4.81	.326E-04	680	2700	4.72	7.46	60

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
61	1.06	1.99	2.16	2.32	3.40	4.67	.162E-01	750	2700	4.96	7.31	61
62	1.34	2.03	2.19	2.33	3.35	4.56	.112E-00	820	2700	5.16	7.19	62
63	1.58	2.08	2.22	2.35	3.29	4.44	.449E-00	910	2700	5.34	7.05	63
64	1.75	2.12	2.25	2.37	3.26	4.35	.111E-01	1000	2700	5.45	6.89	64
65	1.89	2.18	2.29	2.40	3.24	4.29	.232E-01	1100	2700	5.48	6.68	65
66	2.02	2.24	2.34	2.44	3.24	4.26	.411E-01	1200	2700	5.41	6.40	66
67	2.15	2.31	2.41	2.50	3.27	4.27	.658E-01	1300	2700	5.26	6.06	67
68	.761	2.20	2.41	2.60	3.88	5.37	.590E-03	620	2400	4.25	6.59	68
69	1.27	2.25	2.44	2.61	3.81	5.23	.277E-01	680	2400	4.45	6.47	69
70	1.59	2.30	2.47	2.63	3.75	5.09	.174E-00	750	2400	4.63	6.36	70
71	1.81	2.34	2.50	2.64	3.70	4.98	.520E-00	820	2400	4.76	6.25	71
72	2.01	2.40	2.54	2.67	3.65	4.88	.135E-01	910	2400	4.86	6.09	72
73	2.17	2.46	2.59	2.71	3.64	4.82	.271E-01	1000	2400	4.86	5.88	73
74	2.33	2.55	2.66	2.77	3.65	4.79	.495E-01	1100	2400	4.77	5.59	74
75	2.49	2.65	2.76	2.86	3.70	4.82	.810E-01	1200	2400	4.57	5.22	75
76	.685	2.39	2.62	2.83	4.25	5.88	.148E-03	560	2200	3.87	6.06	76
77	1.36	2.45	2.66	2.85	4.16	5.71	.242E-01	620	2200	4.07	5.94	77
78	1.70	2.50	2.69	2.86	4.10	5.57	.148E-00	680	2200	4.23	5.84	78
79	1.96	2.56	2.73	2.88	4.03	5.43	.508E-00	750	2200	4.36	5.73	79
80	2.16	2.61	2.76	2.91	3.99	5.34	.117E-01	820	2200	4.44	5.61	80
81	2.35	2.68	2.82	2.95	3.97	5.26	.258E-01	910	2200	4.46	5.41	81
82	2.52	2.77	2.89	3.01	3.98	5.23	.471E-01	1000	2200	4.38	5.15	82
83	2.72	2.90	3.01	3.12	4.04	5.26	.810E-01	1100	2200	4.19	4.78	83
84	.775	2.63	2.88	3.11	4.67	6.47	.181E-03	510	2000	3.52	5.51	84
85	1.47	2.69	2.92	3.13	4.59	6.29	.203E-01	560	2000	3.69	5.41	85
86	1.88	2.75	2.96	3.15	4.50	6.12	.154E-00	620	2000	3.85	5.31	86
87	2.15	2.81	3.00	3.17	4.44	5.98	.494E-00	680	2000	3.96	5.22	87
88	2.38	2.87	3.04	3.20	4.39	5.86	.124E-01	750	2000	4.04	5.09	88
89	2.57	2.94	3.09	3.24	4.37	5.79	.243E-01	820	2000	4.06	4.93	89
90	2.78	3.05	3.18	3.32	4.38	5.75	.474E-01	910	2000	3.99	4.68	90
91	2.99	3.18	3.31	3.43	4.45	5.79	.810E-01	1000	2000	3.81	4.35	91

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

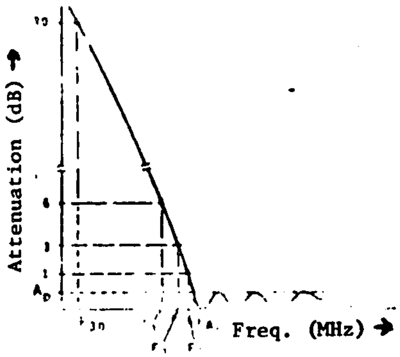
FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB: 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
92	1.12	2.94	3.22	3.46	5.16	7.13	.124E-02	470	1800	3.21	4.93	92
93	1.69	3.00	3.25	3.48	5.08	6.97	.277E-01	510	1800	3.34	4.85	93
94	2.11	3.06	3.29	3.50	5.00	6.79	.163E-00	560	1800	3.47	4.78	94
95	2.43	3.13	3.34	3.53	4.92	6.63	.566E-00	620	1800	3.58	4.68	95
96	2.67	3.20	3.38	3.56	4.87	6.51	.131E-01	680	1800	3.64	4.57	96
97	2.89	3.29	3.45	3.61	4.85	6.42	.271E-01	750	1800	3.65	4.41	97
98	3.09	3.39	3.54	3.69	4.87	6.39	.478E-01	820	1800	3.59	4.21	98
99	1.53	3.33	3.63	3.90	5.78	7.96	.522E-02	430	1600	2.89	4.36	99
100	2.11	3.40	3.68	3.93	5.68	7.76	.614E-01	470	1600	3.01	4.29	100
101	2.47	3.46	3.72	3.94	5.60	7.59	.229E-00	510	1600	3.11	4.23	101
102	2.78	3.53	3.76	3.97	5.53	7.43	.663E-00	560	1600	3.19	4.15	102
103	3.07	3.62	3.82	4.02	5.47	7.29	.160E-01	620	1600	3.24	4.03	103
104	3.30	3.72	3.90	4.08	5.46	7.21	.309E-01	680	1600	3.24	3.88	104
105	3.55	3.86	4.02	4.18	5.49	7.19	.565E-01	750	1600	3.15	3.67	105
106	1.30	3.52	3.86	4.15	6.20	8.57	.938E-03	390	1500	2.67	4.11	106
107	2.11	3.61	3.91	4.18	6.09	8.33	.366E-01	430	1500	2.80	4.04	107
108	2.56	3.68	3.95	4.20	5.99	8.14	.181E-00	470	1500	2.90	3.97	108
109	2.87	3.75	4.00	4.23	5.92	7.98	.4943-00	510	1500	2.97	3.91	109
110	3.17	3.83	4.05	4.26	5.86	7.83	.119E-01	560	1500	3.03	3.82	110
111	3.45	3.93	4.13	4.33	5.82	7.71	.257E-01	620	1500	3.04	3.69	111
112	3.69	4.06	4.24	4.42	5.84	7.67	.464E-01	680	1500	2.99	3.52	112
113	3.99	4.25	4.41	4.57	5.93	7.72	.810E-01	750	1500	2.86	3.26	113
114	1.11	4.04	4.43	4.78	7.19	9.96	.106E-03	330	1300	2.28	3.58	114
115	2.17	4.13	4.49	4.81	7.07	9.71	.148E-01	360	1300	2.39	3.52	115
116	2.72	4.21	4.54	4.84	6.97	9.50	.908E-01	390	1300	2.47	3.47	116
117	3.20	4.30	4.59	4.87	6.86	9.26	.363E-00	430	1300	2.56	3.41	117
118	3.54	4.38	4.65	4.90	6.78	9.08	.901E-00	470	1300	2.61	3.34	118
119	3.81	4.47	4.72	4.95	6.73	8.96	.176E-01	510	1300	2.64	3.26	119
120	4.10	4.60	4.82	5.03	6.72	8.87	.338E-01	560	1300	2.62	3.14	120
121	4.43	4.79	4.98	5.18	6.78	8.86	.623E-01	620	1300	2.54	2.94	121
122	.805	4.36	4.79	5.18	7.81	10.84	.623E-05	300	1200	2.09	3.32	122

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
123	2.28	4.47	4.86	5.21	7.67	10.5	.119E-01	330	1200	2.19	3.25	123
124	2.95	4.56	4.92	5.24	7.55	10.3	.908E-01	360	1200	2.28	3.20	124
125	3.39	4.64	4.97	5.27	7.45	10.1	.294E-00	390	1200	2.35	3.16	125
126	3.80	4.74	5.03	5.31	7.35	9.86	.831E-00	430	1200	2.41	3.09	126
127	4.13	4.84	5.11	5.36	7.29	9.70	.174E-01	470	1200	2.43	3.01	127
128	4.40	4.96	5.20	5.44	7.28	9.61	.309E-01	510	1200	2.43	2.91	128
129	4.72	5.13	5.35	5.57	7.32	9.59	.550E-01	560	1200	2.37	2.76	129
130	2.40	4.87	5.30	5.68	8.38	11.5	.907E-02	300	1100	2.00	2.99	130
131	3.22	4.97	5.36	5.72	8.23	11.2	.908E-01	330	1100	2.09	2.94	131
132	3.73	5.07	5.42	5.75	8.12	11.0	.320E-00	360	1100	2.16	2.89	132
133	4.10	5.16	5.48	5.78	8.03	10.8	.752E-00	390	1100	2.20	2.84	133
134	4.49	5.28	5.57	5.85	7.96	10.6	.172E-01	430	1100	2.23	2.76	134
135	4.82	5.42	5.68	5.94	7.94	10.5	.320E-01	470	1100	2.22	2.66	135
136	5.12	5.59	5.83	6.06	7.98	10.5	.530E-01	510	1100	2.18	2.54	136
137	2.52	5.34	5.82	6.25	9.24	12.7	.628E-02	270	1000	1.81	2.72	137
138	3.54	5.47	5.90	6.29	9.06	12.3	.908E-01	300	1000	1.90	2.67	138
139	4.15	5.58	5.97	6.33	8.92	12.0	.353E-00	330	1000	1.97	2.62	139
140	4.59	5.69	6.04	6.37	8.81	11.8	.867E-00	360	1000	2.01	2.57	140
141	4.93	5.80	6.13	6.43	8.75	11.7	.169E-01	390	1000	2.03	2.51	141
142	5.33	5.97	6.26	6.54	8.73	11.5	.334E-01	430	1000	2.02	2.41	142
143	5.69	6.18	6.44	6.70	8.79	11.5	.573E-01	470	1000	1.97	2.29	143
144	2.40	5.83	6.37	6.86	10.2	14.1	.219E-02	240	910	1.63	2.49	144
145	3.79	6.00	6.47	6.91	9.97	13.6	.744E-01	270	910	1.72	2.43	145
146	4.55	6.13	6.56	6.95	9.80	13.2	.349E-00	300	910	1.79	2.39	146
147	5.08	6.26	6.65	7.01	9.68	13.0	.926E-00	330	910	1.83	2.34	147
148	5.49	6.40	6.75	7.08	9.61	12.8	.188E-01	360	910	1.85	2.28	148
149	5.84	6.56	6.88	7.18	9.60	12.7	.327E-01	390	910	1.84	2.20	149
150	6.28	6.81	7.09	7.37	9.37	12.7	.591E-01	410	910	1.79	2.07	150
151	2.97	6.50	7.09	7.62	11.3	15.5	.487E-02	220	820	1.48	2.23	151
152	4.08	6.64	7.17	7.66	11.1	15.2	.572E-01	240	820	1.54	2.20	152
153	5.05	6.81	7.28	7.71	10.9	14.7	.344E-00	270	820	1.61	2.15	153

TABLE 1. LOWPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONCLUDED)

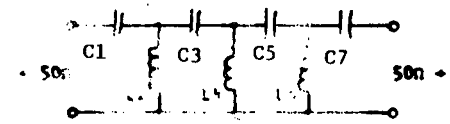
FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
154	5.68	6.97	7.39	7.78	10.7	14.4	.100E-01	300	820	1.65	2.10	154
155	6.17	7.14	7.52	7.88	10.7	14.1	.213E-01	330	820	1.66	2.04	155
156	6.60	7.34	7.68	8.01	10.7	14.0	.381E-01	360	820	1.65	1.96	156
157	7.01	7.58	7.89	8.20	10.7	14.0	.614E-01	390	820	1.61	1.86	157
158	3.13	7.10	7.74	8.32	12.3	17.0	.374E-02	200	750	1.35	2.05	158
159	4.48	7.26	7.84	8.37	12.1	16.6	.598E-01	220	750	1.41	2.01	159
160	5.30	7.40	7.93	8.42	11.9	16.2	.241E-00	240	750	1.46	1.98	160
161	6.12	7.59	8.06	8.49	11.8	15.8	.867E-00	270	750	1.51	1.93	161
162	6.72	7.79	8.21	8.61	11.7	15.5	.204E-01	300	750	1.52	1.87	162
163	7.23	8.03	8.40	8.77	11.7	15.4	.386E-01	330	750	1.51	1.79	163
164	7.72	8.32	8.66	9.00	11.8	15.4	.646E-01	360	750	1.46	1.69	164
165	3.29	7.81	8.53	9.18	13.6	18.8	.264E-02	180	680	1.22	1.86	165
166	4.97	8.01	8.65	9.24	13.4	18.3	.630E-01	200	680	1.28	1.82	166
167	5.94	8.18	8.76	9.29	13.2	17.8	.278E-00	220	680	1.33	1.79	167
168	6.61	8.33	8.86	9.35	13.0	17.4	.720E-00	240	680	1.36	1.76	168
169	7.36	8.58	9.04	9.48	12.9	17.1	.193E-01	270	680	1.38	1.70	169
170	7.98	8.86	9.28	9.68	12.9	16.9	.393E-01	300	680	1.37	1.62	170
171	8.58	9.22	9.60	9.96	13.0	17.0	.687E-01	330	680	1.32	1.52	171
172	2.91	8.51	9.32	10.0	15.0	20.8	.544E-03	160	620	1.10	1.70	172
173	5.28	8.76	9.48	10.1	14.7	20.1	.484E-01	180	620	1.16	1.67	173
174	6.48	8.96	9.60	10.2	14.4	19.5	.267E-00	200	620	1.21	1.63	174
175	7.29	9.15	9.73	10.3	14.2	19.1	.758E-00	220	620	1.24	1.60	175
176	7.91	9.34	9.87	10.4	14.1	18.8	.159E-01	240	620	1.26	1.56	176
177	8.67	9.68	10.1	10.6	14.1	18.6	.362E-01	270	620	1.25	1.49	177
178	9.39	10.10	10.5	10.9	14.2	18.6	.676E-01	300	620	1.20	1.39	178



Attenuation Response Curve

(B) Typical Attenuation Response

Highpass Filter Schematic Diagram



(A) Schematic Diagram

CHAPTER 7 ADDENDUM #2 CONT'D

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
1	1.09	1.01	.971	.935	.717	.549	.684E-01	3300	1600	5.15	4.48	1
2	1.16	1.05	1.00	.961	.724	.550	.411E-01	3600	1600	4.99	4.22	2
3	1.23	1.08	1.02	.977	.725	.547	.244E-01	3900	1600	4.93	4.06	3
4	1.34	1.10	1.04	.990	.721	.539	.115E-01	4300	1600	4.95	3.92	4
5	1.47	1.13	1.06	.999	.713	.529	.500E-00	4700	1600	5.04	3.84	5
6	1.65	1.15	1.07	1.00	.705	.519	.184E-00	5100	1600	5.17	3.78	6
7	2.02	1.17	1.08	1.01	.693	.506	.339E-01	5600	1600	5.37	3.71	7
8	3.51	1.20	1.10	1.02	.680	.491	.544E-03	6200	1600	5.64	3.64	8
9	1.13	1.06	1.02	.985	.760	.583	.810E-01	3000	1500	4.92	4.31	9
10	1.22	1.11	1.06	1.02	.771	.587	.471E-01	3300	1500	4.70	4.01	10
11	1.30	1.14	1.09	1.04	.774	.584	.271E-01	3600	1500	4.63	3.83	11
12	1.39	1.17	1.11	1.05	.771	.578	.151E-01	3900	1500	4.63	3.71	12
13	1.52	1.20	1.12	1.06	.763	.568	.643E-00	4300	1500	4.70	3.62	13
14	1.72	1.22	1.14	1.07	.754	.556	.233E-00	4700	1500	4.82	3.55	14
15	2.00	1.24	1.15	1.08	.743	.544	.630E-01	5100	1500	4.97	3.50	15
16	2.80	1.27	1.16	1.08	.730	.530	.455E-02	5600	1500	5.19	3.44	16
17	1.34	1.25	1.20	1.15	.883	.676	.658E-01	2700	1300	4.17	3.62	17
18	1.45	1.30	1.24	1.19	.892	.676	.351E-01	3000	1300	4.03	3.38	18
19	1.57	1.34	1.27	1.21	.891	.670	.183E-01	3300	1300	4.00	3.24	19
20	1.69	1.37	1.29	1.22	.885	.660	.892E-00	3600	1300	4.04	3.16	20
21	1.85	1.39	1.30	1.23	.875	.648	.397E-00	3900	1300	4.12	3.10	21
22	2.17	1.42	1.32	1.24	.862	.633	.104E-00	4300	1300	4.26	3.05	22
23	2.78	1.45	1.33	1.25	.848	.617	.143E-01	4700	1300	4.43	3.00	23
24	5.06	1.48	1.35	1.25	.835	.603	.173E-03	5100	1300	4.62	2.95	24
25	1.41	1.33	1.28	1.23	.950	.729	.810E-01	2400	1200	3.94	3.45	25
26	1.55	1.40	1.34	1.28	.966	.734	.411E-01	2700	1200	3.74	3.16	26
27	1.68	1.44	1.37	1.31	.966	.727	.204E-01	3000	1200	3.70	3.01	27
28	1.82	1.48	1.40	1.32	.959	.716	.949E-00	3300	1200	3.73	2.92	28
29	2.01	1.51	1.41	1.33	.948	.703	.397E-00	3600	1200	3.80	2.86	29
[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	[illegible]	30

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
31	2.91	1.57	1.44	1.35	.920	.670	.185E-01	4300	1200	4.07	2.77	31
32	5.35	1.61	1.46	1.36	.905	.653	.206E-03	4700	1200	4.26	2.72	32
33	1.54	1.45	1.39	1.34	1.04	.795	.810E-01	2200	1100	3.61	3.16	33
34	1.65	1.51	1.44	1.39	1.05	.801	.495E-01	2400	1100	3.46	2.95	34
35	1.80	1.57	1.49	1.42	1.05	.795	.232E-01	2700	1100	3.39	2.78	35
36	1.97	1.61	1.52	1.44	1.05	.782	.102E-01	3000	1100	3.41	2.68	36
37	2.19	1.65	1.54	1.46	1.03	.766	.397E-00	3300	1100	3.49	2.63	37
38	2.51	1.68	1.56	1.46	1.02	.750	.123E-00	3600	1100	3.59	2.58	38
39	3.06	1.71	1.57	1.47	1.01	.733	.247E-01	3900	1100	3.71	2.55	39
40	5.67	1.75	1.60	1.48	.987	.713	.251E-03	4300	1100	3.90	2.50	40
41	1.70	1.59	1.53	1.48	1.14	.875	.810E-01	2000	1000	3.28	2.87	41
42	1.82	1.66	1.59	1.53	1.16	.881	.471E-01	2200	1000	3.14	2.67	42
43	1.95	1.71	1.63	1.56	1.16	.877	.271E-01	2400	1000	3.08	2.55	43
44	2.15	1.77	1.67	1.59	1.15	.862	.111E-01	2700	1000	3.10	2.45	44
45	2.41	1.81	1.69	1.60	1.14	.843	.397E-00	3000	1000	3.17	2.39	45
46	2.81	1.85	1.72	1.61	1.12	.823	.108E-00	3300	1000	3.28	2.34	46
47	3.56	1.88	1.73	1.62	1.10	.803	.162E-01	3600	1000	3.40	2.31	47
48	6.05	1.92	1.76	1.63	1.09	.785	.341E-03	3900	1000	3.54	2.27	48
49	2.00	1.83	1.75	1.68	1.27	.968	.474E-01	2000	910	2.85	2.43	49
50	2.15	1.89	1.80	1.72	1.28	.963	.258E-01	2200	910	2.81	2.31	50
51	2.31	1.93	1.83	1.74	1.27	.951	.135E-01	2400	910	2.81	2.24	51
52	2.61	1.99	1.86	1.76	1.25	.929	.449E-00	2700	910	2.88	2.18	52
53	3.08	2.03	1.88	1.77	1.23	.905	.110E-00	3000	910	2.98	2.13	53
54	4.02	2.07	1.91	1.78	1.21	.881	.130E-01	3300	910	3.11	2.10	54
55	8.39	2.12	1.93	1.79	1.19	.859	.593E-04	3600	910	3.25	2.06	55
56	2.22	2.03	1.94	1.86	1.41	1.07	.478E-01	1800	820	2.57	2.19	56
57	2.41	2.10	2.00	1.91	1.41	1.07	.243E-01	2000	820	2.53	2.08	57
58	2.61	2.15	2.03	1.93	1.41	1.05	.117E-01	2200	820	2.54	2.01	58
59	2.85	2.20	2.06	1.95	1.39	1.03	.520E-00	2400	820	2.58	1.97	59
60	3.41	2.25	2.09	1.96	1.37	1.00	.112E-00	2700	820	2.68	1.92	60
61	4.63	2.31	2.12	1.98	1.34	.975	.978E-02	3000	820	2.81	1.89	61

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	Ap-DB	FREQ. (MHZ) AT Ap, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
62	18.02	2.37	2.15	1.99	1.32	.948	.554E-06	3300	820	2.95	1.85	62
63	2.26	2.12	2.04	1.97	1.52	1.17	.810E-01	1500	750	2.46	2.16	63
64	2.38	2.19	2.10	2.02	1.54	1.17	.565E-01	1600	750	2.38	2.05	64
65	2.60	2.28	2.17	2.08	1.55	1.17	.271E-01	1800	750	2.31	1.91	65
66	2.83	2.35	2.22	2.11	1.54	1.15	.124E-01	2000	750	2.32	1.84	66
67	3.13	2.40	2.25	2.13	1.52	1.13	.508E-00	2200	750	2.36	1.80	67
68	3.53	2.45	2.28	2.14	1.50	1.11	.174E-00	2400	750	2.43	1.77	68
69	4.74	2.51	2.31	2.16	1.47	1.07	.162E-01	2700	750	2.55	1.73	69
70	13.98	2.58	2.35	2.18	1.44	1.04	.623E-05	3000	750	2.69	1.69	70
71	2.69	2.45	2.34	2.25	1.70	1.30	.464E-01	1500	680	2.13	1.81	71
72	2.82	2.50	2.39	2.28	1.71	1.29	.309E-01	1600	680	2.10	1.75	72
73	3.10	2.59	2.45	2.33	1.70	1.27	.131E-01	1800	680	2.10	1.67	73
74	3.46	2.65	2.49	2.35	1.68	1.25	.494E-00	2000	680	2.14	1.63	74
75	3.97	2.71	2.52	2.37	1.65	1.22	.148E-00	2200	680	2.21	1.60	75
76	4.88	2.76	2.54	2.38	1.63	1.19	.277E-01	2400	680	2.29	1.58	76
77	12.21	2.84	2.59	2.40	1.59	1.15	.326E-04	2700	680	2.43	1.54	77
78	2.84	2.63	2.52	2.43	1.86	1.42	.623E-01	1300	620	1.98	1.71	78
79	3.16	2.77	2.64	2.52	1.87	1.41	.257E-01	1500	620	1.91	1.58	79
80	3.33	2.82	2.67	2.54	1.87	1.40	.160E-01	1600	620	1.91	1.54	80
81	3.74	2.90	2.72	2.58	1.84	1.37	.566E-00	1800	620	1.95	1.49	81
82	4.34	2.97	2.76	2.59	1.81	1.34	.154E-00	2000	620	2.01	1.46	82
83	5.45	3.03	2.79	2.61	1.78	1.30	.242E-01	2200	620	2.09	1.44	83
84	8.95	3.10	2.83	2.62	1.75	1.27	.590E-03	2400	620	2.19	1.41	84
85	3.19	2.94	2.82	2.71	2.06	1.57	.550E-01	1200	560	1.77	1.52	85
86	3.39	3.03	2.89	2.77	2.07	1.57	.338E-01	1300	560	1.73	1.45	86
87	3.81	3.15	2.98	2.83	2.06	1.54	.119E-01	1500	560	1.73	1.37	87
88	4.06	3.20	3.01	2.85	2.05	1.52	.664E-00	1600	560	1.75	1.35	88
89	4.77	3.28	3.05	2.87	2.01	1.48	.163E-00	1800	560	1.82	1.32	89
90	6.17	3.36	3.09	2.89	1.97	1.44	.203E-01	2000	560	1.90	1.29	90
91	12.00	3.44	3.14	2.91	1.94	1.40	.148E-03	2200	560	1.99	1.27	91
92	3.53	3.23	3.10	2.98	2.26	1.73	.530E-01	1100	510	1.61	1.38	92

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
93	3.76	3.34	3.18	3.04	2.28	1.72	.309E-01	1200	510	1.58	1.31	93
94	4.01	3.42	3.24	3.09	2.27	1.71	.176E-01	1300	510	1.57	1.27	94
95	4.61	3.54	3.31	3.13	2.24	1.66	.494E-00	1500	510	1.61	1.22	95
96	5.03	3.58	3.34	3.15	2.22	1.64	.229E-00	1600	510	1.64	1.21	96
97	6.51	3.68	3.39	3.17	2.17	1.58	.277E-01	1800	510	1.72	1.18	97
98	12.81	3.78	3.44	3.19	2.13	1.54	.181E-03	2000	510	1.81	1.16	98
99	3.79	3.49	3.35	3.22	2.45	1.87	.573E-01	1000	470	1.49	1.29	99
100	4.07	3.62	3.45	3.30	2.47	1.87	.320E-01	1100	470	1.45	1.21	100
101	4.35	3.71	3.52	3.35	2.46	1.85	.174E-01	1200	470	1.45	1.17	101
102	4.68	3.78	3.57	3.38	2.45	1.83	.901E-00	1300	470	1.46	1.14	102
103	5.61	3.90	3.63	3.42	2.40	1.77	.181E-00	1500	470	1.52	1.11	103
104	6.39	3.96	3.66	3.43	2.37	1.74	.614E-01	1600	470	1.56	1.10	104
105	10.66	4.07	3.72	3.46	2.32	1.68	.124E-02	1800	470	1.65	1.07	105
106	4.12	3.80	3.65	3.51	2.68	2.05	.591E-01	910	430	1.37	1.18	106
107	4.42	3.94	3.76	3.60	2.70	2.04	.334E-01	1000	430	1.33	1.11	107
108	4.77	4.06	3.84	3.66	2.69	2.02	.172E-01	1100	430	1.33	1.07	108
109	5.16	4.14	3.90	3.70	2.67	1.99	.831E-00	1200	430	1.34	1.04	109
110	5.66	4.22	3.95	3.72	2.64	1.96	.363E-00	1300	430	1.37	1.02	110
111	7.45	4.35	4.02	3.76	2.58	1.88	.366E-01	1500	430	1.44	.999	111
112	9.60	4.42	4.05	3.77	2.55	1.85	.522E-02	1600	430	1.49	.987	112
113	4.52	4.18	4.01	3.86	2.95	2.26	.614E-01	820	390	1.24	1.07	113
114	4.89	4.35	4.15	3.97	2.98	2.25	.327E-01	910	390	1.21	1.01	114
115	5.27	4.48	4.24	4.04	2.97	2.23	.169E-01	1000	390	1.20	.970	115
116	5.75	4.58	4.31	4.08	2.94	2.19	.752E-00	1100	390	1.22	.944	116
117	6.39	4.67	4.36	4.11	2.91	2.15	.294E-00	1200	390	1.25	.926	117
118	7.34	4.75	4.40	4.13	2.87	2.10	.908E-01	1300	390	1.28	.913	118
119	13.35	4.91	4.49	4.17	2.79	2.02	.938E-03	1500	390	1.37	.889	119
120	4.86	4.51	4.33	4.17	3.19	2.44	.646E-01	750	360	1.15	.999	120
121	5.20	4.68	4.47	4.28	3.22	2.44	.381E-01	820	360	1.12	.942	121
122	5.64	4.83	4.58	4.37	3.22	2.42	.188E-01	910	360	1.11	.900	122
123	6.14	4.95	4.66	4.42	3.19	2.28	.867E-00	1000	360	1.12	.874	123

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONTINUED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
124	6.86	5.05	4.72	4.45	3.15	2.33	.320E-00	1100	360	1.15	.856	124
125	7.96	5.14	4.77	4.48	3.11	2.28	.908E-01	1200	360	1.18	.843	125
126	10.00	5.24	4.82	4.50	3.06	2.23	.148E-01	1300	360	1.23	.831	126
127	5.26	4.90	4.71	4.53	3.48	2.66	.687E-01	680	330	1.06	.924	127
128	5.67	5.10	4.87	4.67	3.51	2.67	.386E-01	750	330	1.03	.865	128
129	6.07	5.25	4.98	4.75	3.51	2.65	.213E-01	820	330	1.02	.830	129
130	6.65	5.39	5.08	4.82	3.49	2.60	.926E-00	910	330	1.03	.803	130
131	7.40	5.50	5.14	4.85	3.44	2.55	.353E-00	1000	330	1.05	.786	131
132	8.68	5.61	5.20	4.88	3.39	2.49	.908E-01	1100	330	1.09	.772	132
133	11.21	5.73	5.26	4.91	3.34	2.43	.119E-01	1200	330	1.13	.761	133
134	21.33	5.85	5.33	4.94	3.29	2.37	.106E-03	1300	330	1.17	.748	134
135	5.80	5.39	5.18	4.99	3.83	2.93	.676E-01	620	300	.965	.838	135
136	6.22	5.60	5.35	5.13	3.86	2.93	.393E-01	680	300	.933	.787	136
137	6.71	5.78	5.49	5.23	3.86	2.91	.204E-01	750	300	.924	.753	137
138	7.25	5.91	5.58	5.29	3.84	2.87	.100E-01	820	300	.931	.732	138
139	8.15	6.05	5.66	5.34	3.79	2.80	.349E-00	910	300	.954	.715	139
140	9.55	6.17	5.72	5.37	3.73	2.74	.908E-01	1000	300	.986	.702	140
141	12.79	6.31	5.80	5.40	3.66	2.66	.907E-02	1100	300	1.03	.690	141
142	34.95	6.46	5.88	5.44	3.60	2.60	.623E-05	1200	300	1.08	.678	142
143	6.46	6.00	5.77	5.55	4.25	3.25	.663E-01	560	270	.867	.752	143
144	6.98	6.26	5.97	5.72	4.30	3.26	.362E-01	620	270	.838	.704	144
145	7.50	6.44	6.10	5.82	4.29	3.23	.193E-01	680	270	.832	.676	145
146	8.18	6.59	6.21	5.89	4.26	3.18	.867E-00	750	270	.840	.656	146
147	9.07	6.72	6.29	5.93	4.21	3.11	.344E-00	820	270	.859	.643	147
148	10.87	6.88	6.37	5.97	4.13	3.03	.744E-01	910	270	.892	.631	148
149	14.92	7.03	6.45	6.01	4.06	2.95	.628E-02	1000	270	.931	.620	149
150	7.41	6.83	6.55	6.30	4.80	3.67	.578E-01	510	240	.762	.656	150
151	7.95	7.08	6.75	6.46	4.83	3.66	.327E-01	560	240	.743	.620	151
152	8.61	7.29	6.90	6.57	4.82	3.62	.159E-01	620	240	.740	.595	152
153	9.39	7.45	7.01	6.64	4.78	3.56	.720E-00	680	240	.750	.580	153
154	10.63	7.61	7.10	6.69	4.71	3.48	.241E-00	750	240	.771	.568	154

TABLE 2. HIGHPASS 50-OHM 7-ELEMENT CHEBYSHEV FILTERS (CONCLUDED)

FLTR NO.	A_p-DB	FREQ. (MHZ) AT A_p, 1,3,6,30 & 50 DB 1-DB	3-DB	6-DB	30-DB	50-DB	R.C. (%)	C1.7 (pF)	C3.5 (pF)	L2.6 (uH)	L4 (uH)	FLTR NO.
155	12.63	7.76	7.18	6.72	4.64	3.40	.572E-01	820	240	.797	.559	155
156	19.33	7.96	7.28	6.77	4.55	3.30	.219E-02	910	240	.837	.549	156
157	8.11	7.47	7.16	6.88	5.24	4.00	.560E-01	470	220	.697	.599	157
158	8.63	7.70	7.35	7.04	5.27	4.00	.341E-01	510	220	.681	.571	158
159	9.28	7.92	7.51	7.15	5.26	3.96	.178E-01	560	220	.678	.548	159
160	10.19	8.12	7.64	7.24	5.22	3.89	.758E-00	620	220	.687	.532	160
161	11.41	8.28	7.73	7.29	5.15	3.80	.278E-00	680	220	.704	.522	161
162	13.71	8.46	7.83	7.33	5.06	3.71	.598E-01	750	220	.730	.513	162
163	18.94	8.64	7.92	7.37	4.98	3.61	.487E-02	820	220	.761	.505	163
164	8.97	8.23	7.89	7.59	5.77	4.40	.540E-01	430	200	.632	.543	164
165	9.59	8.51	8.11	7.76	5.80	4.39	.312E-01	470	200	.618	.515	165
166	10.22	8.72	8.26	7.87	5.79	4.35	.176E-01	510	200	.616	.498	166
167	11.14	8.92	8.39	7.96	5.74	4.28	.806E-00	560	200	.623	.485	167
168	12.60	9.12	8.51	8.02	5.66	4.18	.267E-00	620	200	.640	.474	168
169	14.98	9.30	8.61	8.06	5.57	4.08	.630E-02	680	200	.663	.467	169
170	21.57	9.52	8.72	8.11	5.47	3.97	.374E-02	750	200	.694	.458	170
171	9.42	8.84	8.51	8.21	6.33	4.86	.810E-01	360	180	.590	.517	171
172	10.02	9.18	8.80	8.45	6.42	4.89	.516E-01	390	180	.567	.486	172
173	10.79	9.51	9.05	8.65	6.45	4.87	.280E-01	430	180	.555	.460	173
174	11.59	9.74	9.22	8.77	6.42	4.82	.147E-01	470	180	.555	.445	174
175	12.53	9.94	9.34	8.85	6.37	4.74	.720E-00	510	180	.562	.435	175
176	14.08	10.14	9.46	8.91	6.29	4.64	.255E-00	560	180	.577	.427	176
177	17.20	10.36	9.58	8.97	6.18	4.52	.484E-01	620	180	.600	.419	177
178	25.12	10.60	9.70	9.02	6.07	4.40	.264E-02	680	180	.627	.412	178

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(9) P. Geffe, Simplified Modern Filter Design, John F. Rider, New York, 1964.

(10) E. Wetherhold, "A Narrowband Non-tunable Detection System for TEMPEST Testing," pp. 68-71, ITEM 1975.

This article was prepared for ITEM by Edward E. Wetherhold, Honeywell, Inc., Signal Analysis Center, P.O. Box 391, Annapolis, MD 21404.

8.0 TABLES AND NOMOGRAPHS

8.1 INTRODUCTION

This chapter provides a convenient reference for miscellaneous data useful in the design of electromagnetically compatible Avionics/GSE. The chapter contains tables for various conversions of ratio units to decibels, as well as tables, equations and nomographs for convenient calculation of EM wave shielding parameters, field strength and power density, cable separation distance, transmission path loss, and various useful characteristics.

8.2 ORDER OF CONTENTS

PAGE

ORDER OF CONTENTS (CONT'D)

SHIELDING-RELATED NOMOGRAPHS (CONT'D)

ORDER OF CONTENTS (CONT'D)

USEFUL CONVERSION EQUATIONS (CONT'D)

DBM	u/volts 50 ohm	u/volts 72 ohm	u/volts 600 ohm	uu/watts
0	223,607.0	268,328.0	774,596.7	1,000,000,000.0
-3	158,314.0	189,976.0	548,379.4	501,200,000.0
-6	112,094.0	134,513.0	388,265.4	251,250,000.0
-9	79,358.0	95,230.0	274,845.4	125,900,000.0
-12	56,192.0	67,431.0	194,576.5	63,100,000.0
-15	39,780.0	47,736.0	137,738.9	31,620,000.0
-18	28,174.0	33,809.0	97,519.2	15,850,000.0
-21	19,932.0	23,919.0	69,034.8	7,943,000.0
-24	14,112.0	16,934.0	48,873.3	3,981,000.0
-27	9,990.0	11,988.0	34,597.7	1,995,000.0
-30	7,073.0	8,487.0	24,494.9	1,000,000.0
-33	5,009.0	6,011.0	17,341.3	501,200.0
-36	3,546.0	4,256.0	12,276.8	251,200.0
-39	2,511.0	3,013.0	8,691.4	125,900.0
-42	1,776.0	2,132.0	6,153.0	63,100.0
-45	1,258.0	1,509.0	4,355.7	31,620.0
-48	890.0	1,068.0	3,083.8	15,850.0
-51	630.0	756.0	2,183.1	7,943.0
-54	446.0	536.0	1,545.5	3,981.0
-57	316.0	379.0	1,094.0	1,995.0
-60	223.607	268.328	774.597	1,000.0
-63	158.314	189.976	548.379	501.2
-66	112.094	134.513	388.265	251.25
-69	79.358	95.230	274.845	125.9
-72	56.192	67.431	194.576	63.1
-75	39.780	47.736	137.739	31.62
-78	28.174	33.809	97.519	15.85
-81	19.932	23.919	69.035	7.943
-84	14.112	16.934	48.873	3.981
-87	9.990	11.988	34.598	1.995
-90	7.073	8.487	24.495	1.0
-93	5.009	6.011	17.341	0.5012
-96	3.546	4.256	12.277	0.2512
-99	2.511	3.013	8.691	0.1259
-102	1.776	2.132	6.153	0.0631
-105	1.257	1.509	4.356	0.03162
-107	0.999	1.199	3.260	0.01995

Figure 8-1
Voltage Into Different Impedances and Power to dBm

dBm	dBµV	Volts
+53	160	100
+48	155	56
+43	150	32
+38	145	18
+33	140	10
+28	135	5.6
+23	130	3.2
+18	125	1.8
+13	120	1.0
+8	115	.56
+3	110	.32
-2	105	.18
-7	100	.10
-8	99	89 mV
-9	98	79 mV
-10	97	71 mV
-11	96	63 mV
-12	95	56 mV
-13	94	50 mV
-14	93	45 mV
-15	92	40 mV
-16	91	35 mV
-17	90	32 mV
-18	89	28 mV
-19	88	25 mV
-20	87	22 mV
-21	86	20 mV
-22	85	18 mV
-23	84	16 mV
-24	83	14 mV
-25	82	13 mV
-26	81	11 mV
-27	80	10 mV
-28	79	8.9 mV
-29	78	7.9 mV
-30	77	7.1 mV
-31	76	6.3 mV
-32	75	5.6 mV
-33	74	5.0 mV
-34	73	4.5 mV
-35	72	4.0 mV
-36	71	3.5 mV
-37	70	3.2 mV
-38	69	2.8 mV
-39	68	2.5 mV
-40	67	2.2 mV
-41	66	2.0 mV
-42	65	1.8 mV
-43	64	1.6 mV
-44	63	1.4 mV
-45	62	1.3 mV
-46	61	1.1 mV
-47	60	1.0 mV
-48	59	.89 mV
-49	58	.79 mV
-50	57	.71 mV
-51	56	.63 mV
-52	55	.56 mV
-53	54	.50 mV
-54	53	.45 mV
-55	52	.40 mV
-56	51	.35 mV
-57	50	.32 mV
-58	49	.28 mV
-59	48	.25 mV
-60	47	.22 mV
-61	46	.20 mV
-62	45	.18 mV
-63	44	.16 mV
-64	43	.14 mV
-65	42	.13 mV
-66	41	.11 mV
-67	40	.10 mV

dBm	dBµV	Microvolts
-68	39	89
-69	38	79
-70	37	71
-71	36	63
-72	35	56
-73	34	50
-74	33	45
-75	32	40
-76	31	35
-77	30	32
-78	29	28
-79	28	25
-80	27	22
-81	26	20
-82	25	18
-83	24	16
-84	23	14
-85	22	13
-86	21	11
-87	20	10
-88	19	8.9
-89	18	7.9
-90	17	7.1
-91	16	6.3
-92	15	5.6
-93	14	5.0
-94	13	4.5
-95	12	4.0
-96	11	3.5
-97	10	3.2
-98	9	2.8
-99	8	2.5
-100	7	2.2
-101	6	2.0
-102	5	1.8
-103	4	1.6
-104	3	1.4
-105	2	1.3
-106	1	1.1
-107	0	1.0
-108	-1	.89
-109	-2	.79
-110	-3	.71
-111	-4	.63
-112	-5	.56
-113	-6	.50
-114	-7	.45
-115	-8	.40
-116	-9	.35
-117	-10	.32
-118	-11	.28
-119	-12	.25
-120	-13	.22
-121	-14	.20
-122	-15	.18
-123	-16	.16
-124	-17	.14
-125	-18	.13
-126	-19	.11
-127	-20	.10
-132	-25	.056
-137	-30	.032
-142	-35	.018
-147	-40	.010
-152	-45	.006

Figure 8-2
dB Above 1 Microvolt and Volts Into 50 Ohms to dBm

dB	Voltage or Current Ratio	Power Ratio
0	1.000	1.000
1.0	1.122	1.259
2.0	1.259	1.585
3.0	1.413	1.995
4.0	1.585	2.512
5.0	1.778	3.162
6.0	1.995	3.931
7.0	2.239	5.012
8.0	2.512	6.310
9.0	2.818	7.943
10.0	3.162	10.000
10.5	3.350	11.22
11.0	3.548	12.59
11.5	3.758	14.13
12.0	3.981	15.85
12.5	4.217	17.78
13.0	4.467	19.95
13.5	4.732	22.39
14.0	5.012	25.12
14.5	5.309	28.18
15.0	5.623	31.62
16.0	6.310	39.81
17.0	7.079	50.12
18.0	7.943	63.10
19.0	8.913	79.43
20.0	10.000	10^2
25.0	17.8	3.16×10^2
30.0	31.62	10^3
35.0	56.2	3.16×10^3
40.0	100.0	10^4
45.0	177.8	3.16×10^4
50.0	316.2	10^5
60.0	1000	10^6
70.0	3162.0	10^7
80.0	10,000	10^8
90.0	31,620	10^9
100.0	100,000	10^{10}

Figure 8-3
Conversion Ratios, dB to Voltage, Current and Power

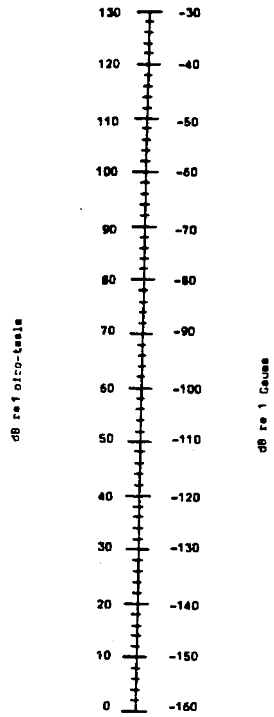


Figure 8-4
Converting Flux Density
From Pico-Tesla to Gauss

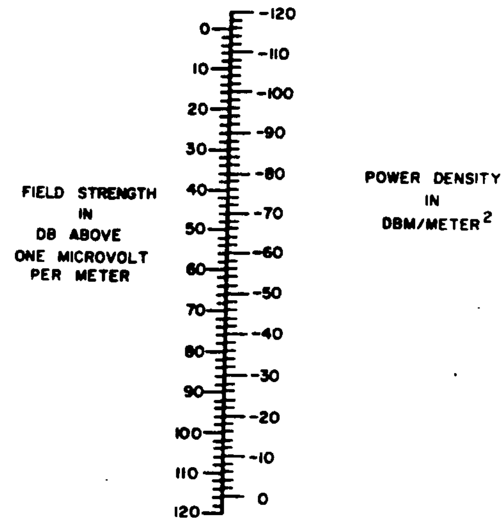


Figure 8-6
Converting dB Above 1 Microvolt-per-Meter to dBm-per-Square-Meter

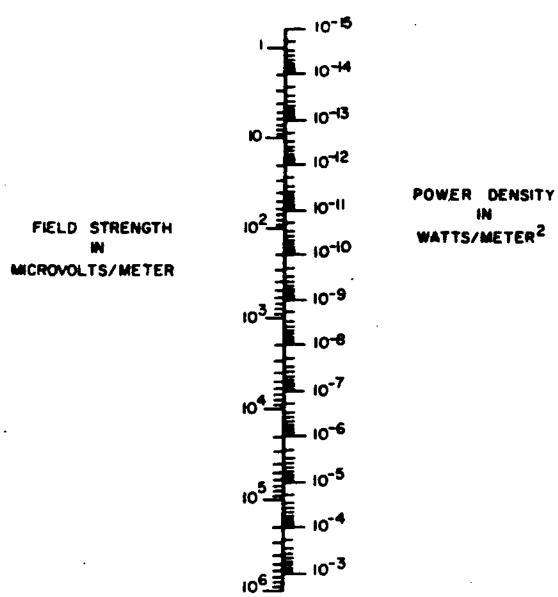


Figure 8-5
Converting Microvolts-per-Meter to Watts-per-Square-Meter

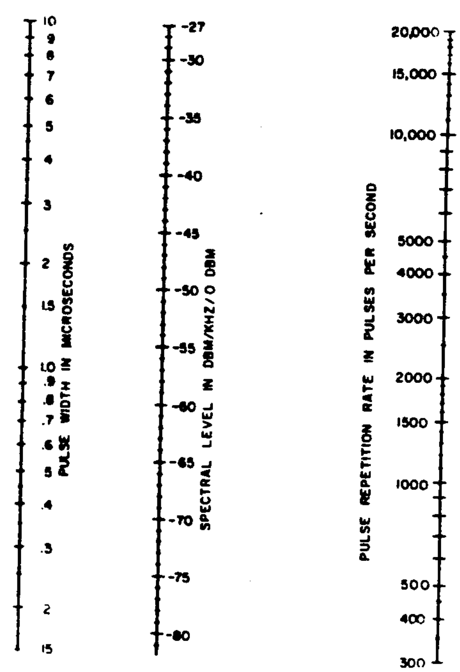


Figure 8-7
Determining Spectral Level from Pulse Width and Pulse Repetition Rate

NOMOGRAPHS

Nomograph for the Relationship Between Pulse Width, Pulse Repetition Rate, and Duty Cycle (Figure 8-8)

The duty cycle, expressed in decibels, is:

$$dB = 10 \log_{10} \frac{T_{on}}{T_{on} + T_{off}}$$

where T_{on} is the on-time, and T_{off} the off-time.

As an example of the use of this nomograph, the dashed line shows that pulses of one microsecond width and a 30 dB (0.001) duty cycle correspond to a pulse repetition frequency of 1000.

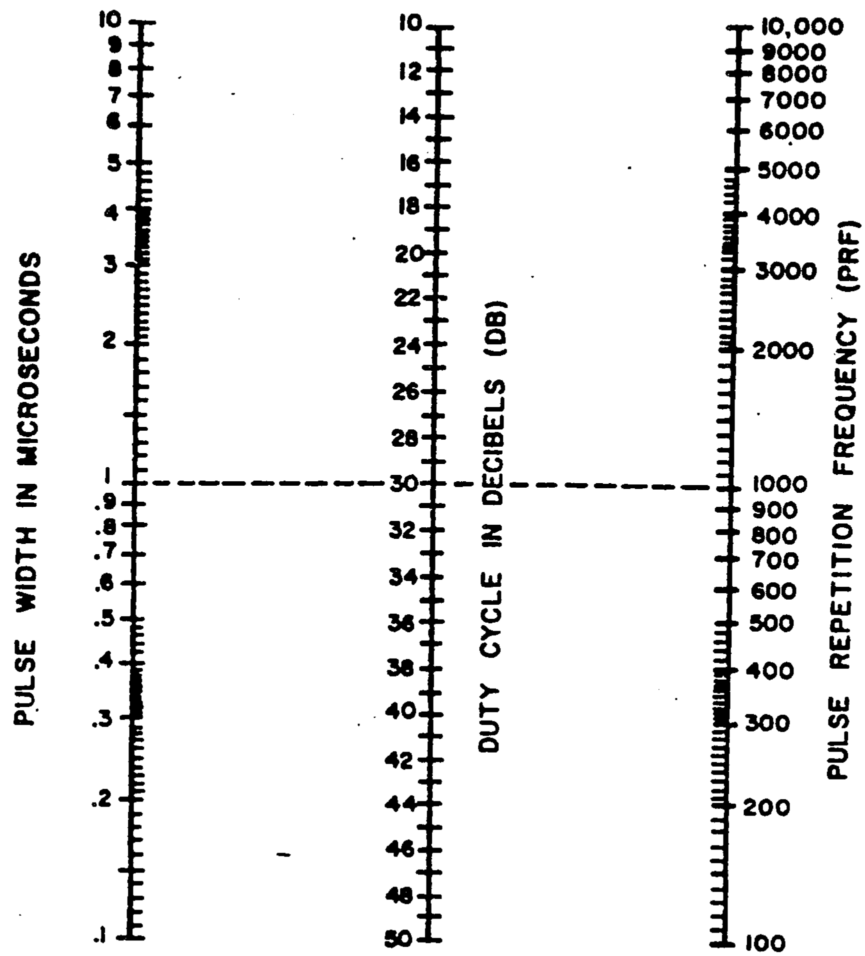


Figure 8-8
Relationship Between Pulse Width, Pulse Repetition Rate, and Duty Cycle

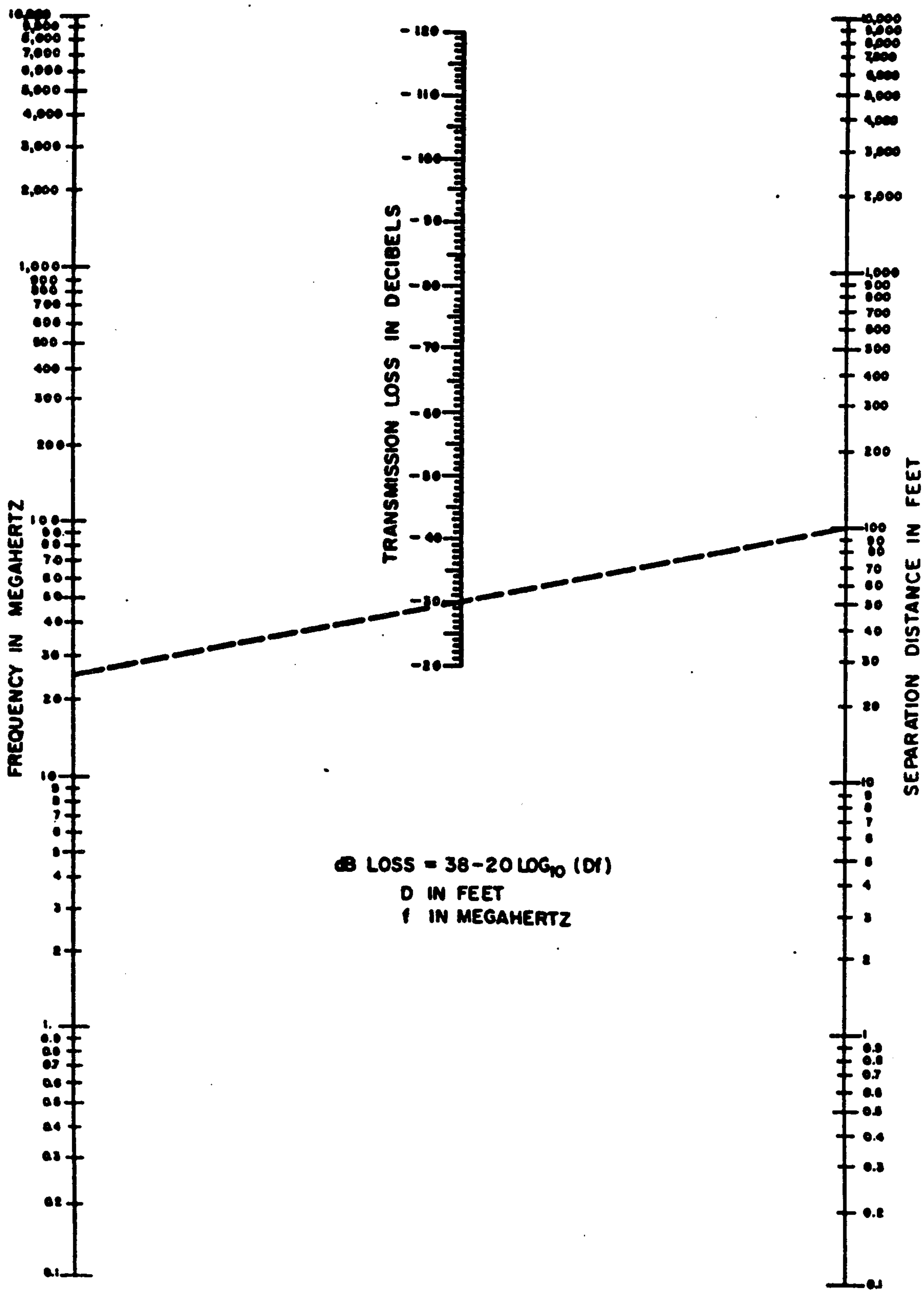


Figure 8-9
Free Space Nomograph (Feet and Megahertz)

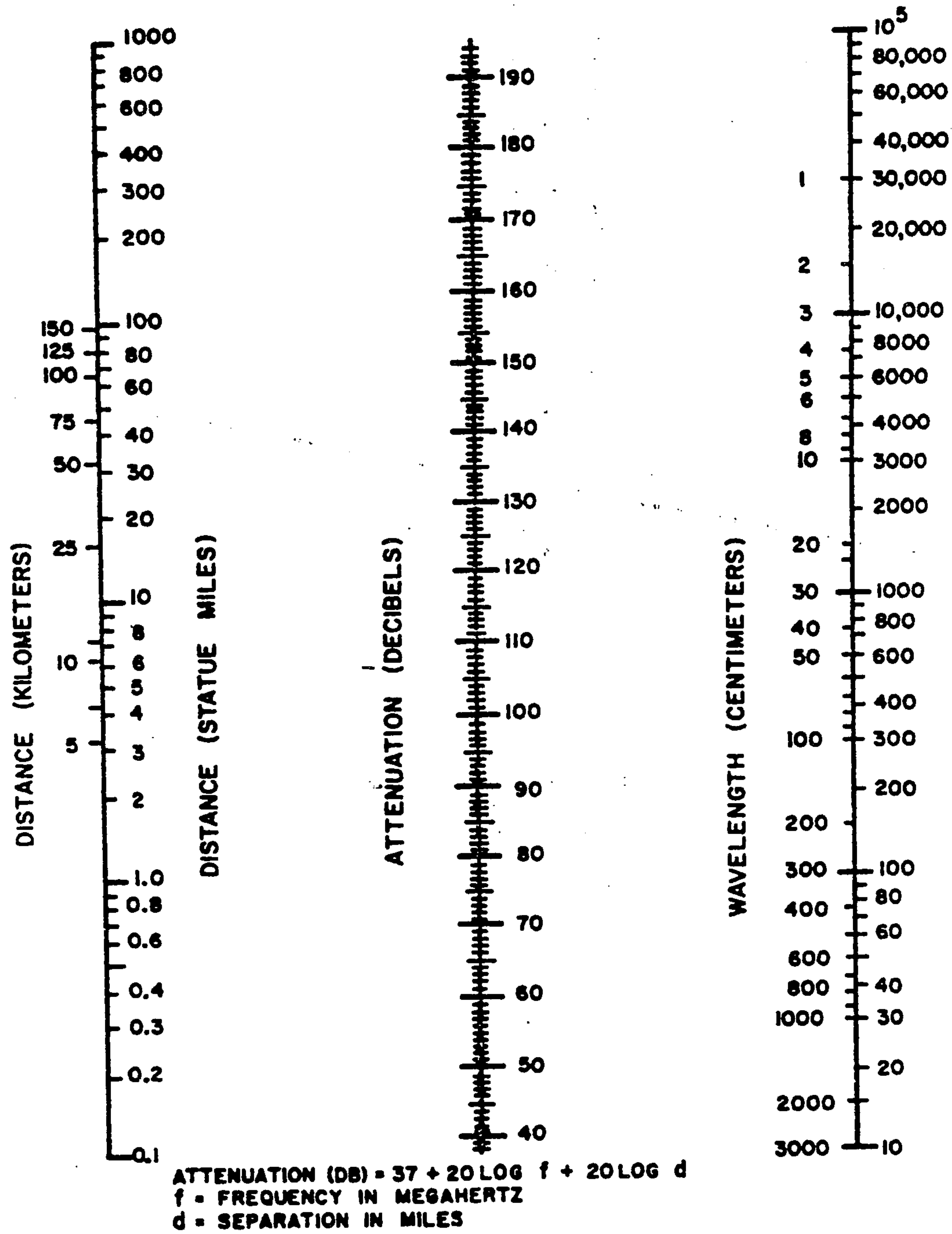


Figure 8-10
Free Space Nomograph (Miles and Megahertz)

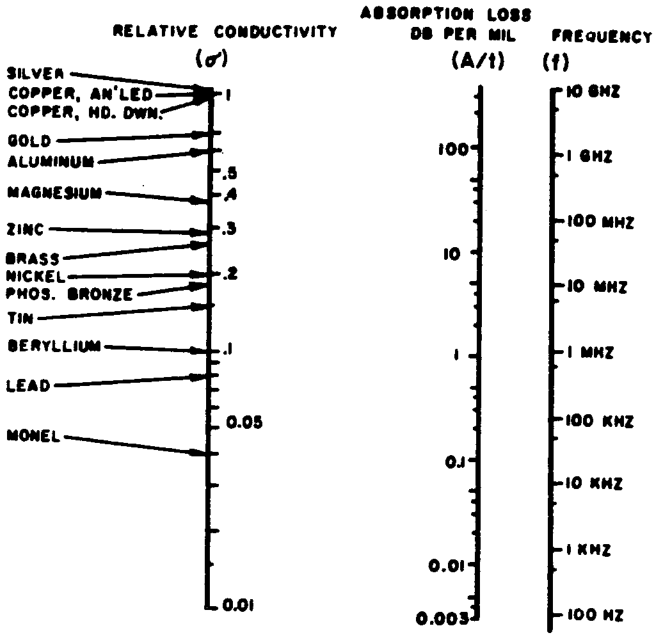


Figure 8-11
Absorption Loss for
Nonmagnetic Materials

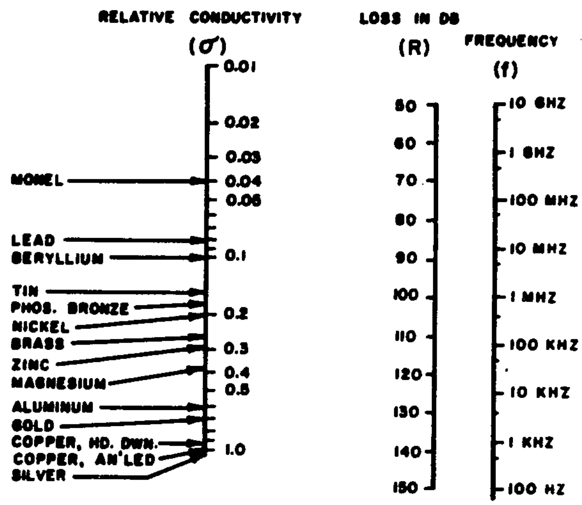


Figure 8-13
Plane Wave Reflection
Loss for Nonmagnetic Materials

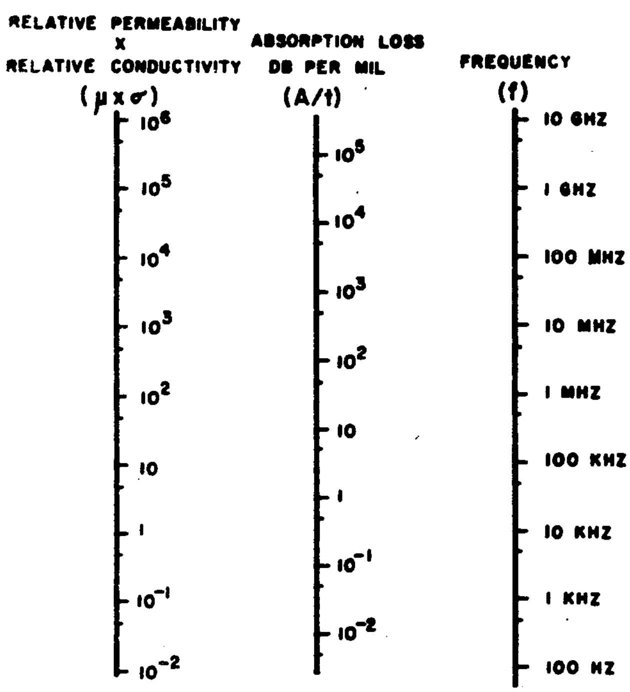


Figure 8-12
Absorption Loss for
Magnetic Materials

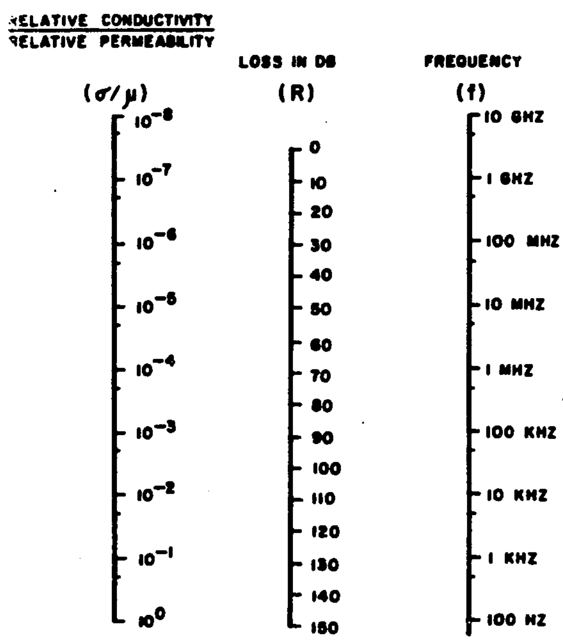


Figure 8-14
Plane Wave Reflection
Loss for Magnetic Materials

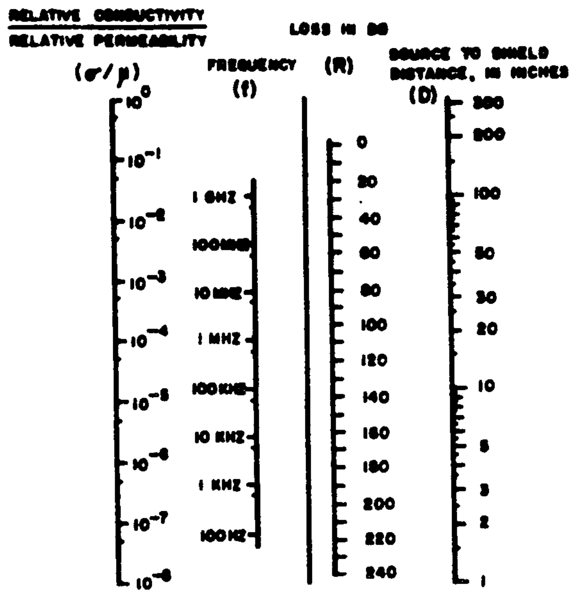


Figure 8-15
Electric Field Reflection Loss for Magnetic Materials

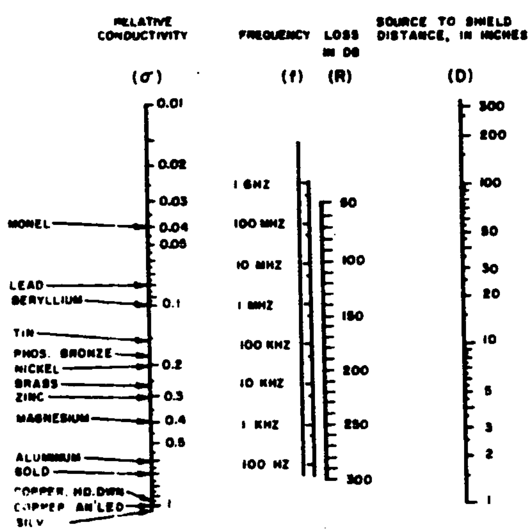


Figure 8-16
Magnetic Field Reflection
Loss for Nonmagnetic Materials

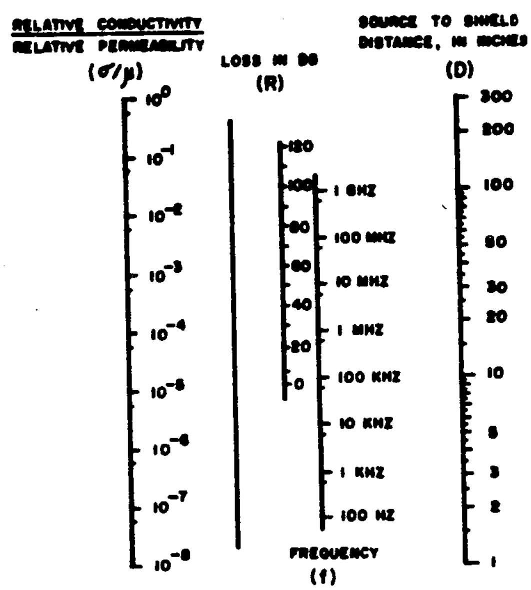


Figure 8-17
Magnetic Field Reflection Loss
for Magnetic Materials

Nomograph for Estimating Allowable Distance Between Parallel Conductors (Figure 8-18)

At frequencies up to where a cable run is approximately one-sixteenth of a wavelength, the nomograph in Figure 8-18 can be used to estimate the amount of distance allowable between a radiating line and a susceptible line.

Example:

A = level in radiating cable = -25 dBm W

B = susceptible level of receiving cable = 30 dBm W

C = cable run length factor at 20 ft.

D = frequency factor at 1000 Hz

E = elevation or distance from ground plane or shield

X = compensated difference between emitting cable and susceptible cable = (A)-(B)+(C)+(D), in dB

R = required cable separation, in inches

Procedure:

1. For C, Length Factor from Figure 8-18
2. For D, Frequency Factor from Figure 8-18
3. Solve for X = (-25)-(-30)+4.5 + 8.0 = +17.5 dB
4. Find R from nomograph, entering at X = 17.5 and E = 2 inches. Find R = 4 inch required separation.

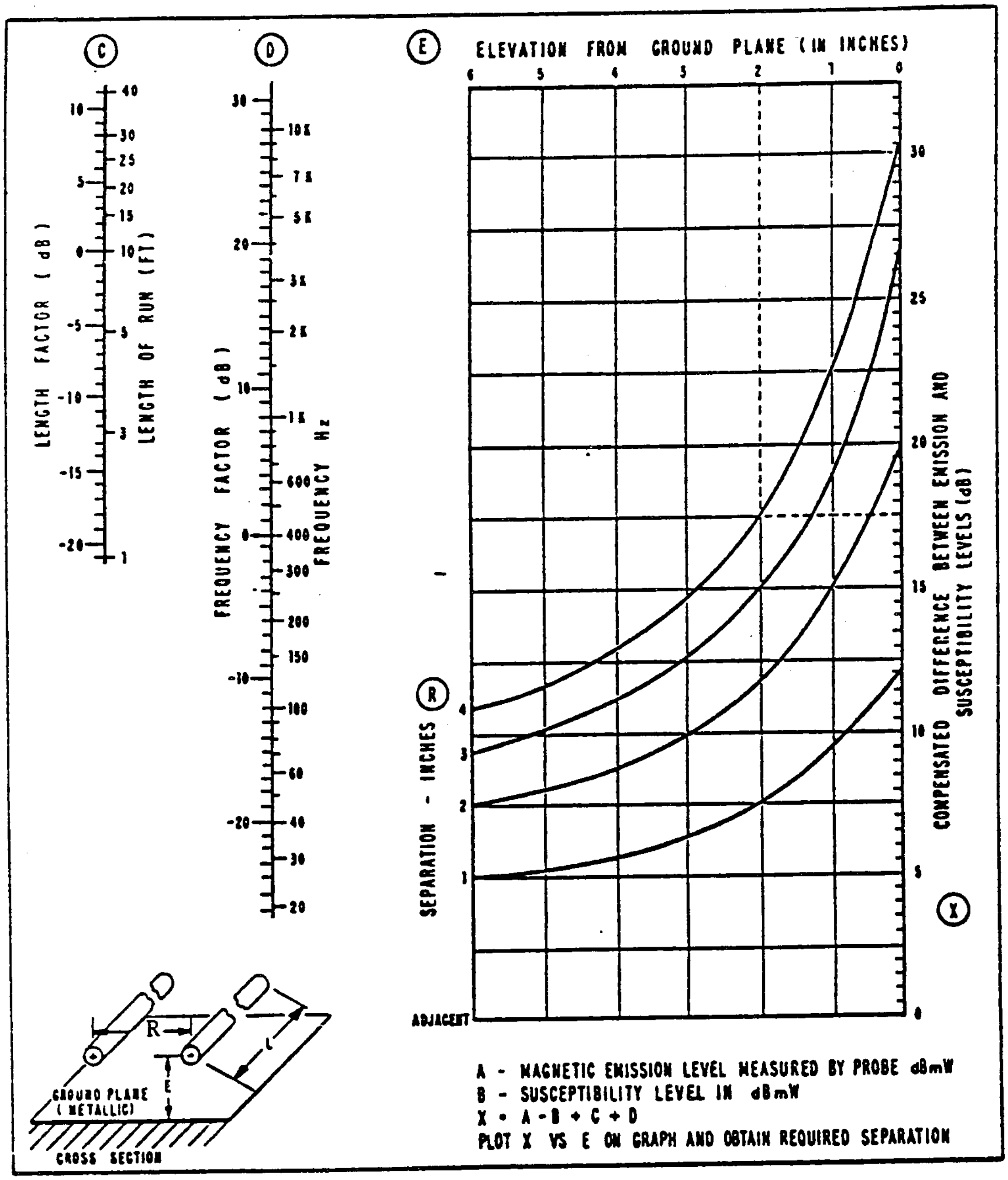


Figure 8-18
Nomograph for Estimating Allowable Distance Between Parallel Conductors

Estimation of Capacitance in Unshielded Cable Harnesses and Runs

The capacitance between unshielded wiring is given in Figure 8-19. The chart may be used to estimate total capacity for a given length of cable harness.

The capacitance between wires in a long cable run can exceed circuit requirements; standard coaxial and shielded cable can be used to control capacitance. The table in Figure 8-20, "Capacitance of Various Shielded AN Wires", is given to assist the designer in selecting the proper cable configuration for his needs.

Capacitance vs. Wire Size for Various Grouping of Wires

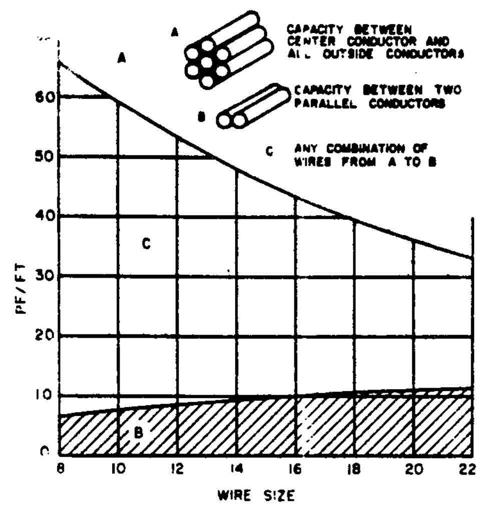


Figure 8-19 Estimation of Capacitance in Unshielded Cable Harnesses and Runs

CAPACITANCE IN PICOFARAD PER FOOT												
TYPE	ONE WIRE SHIELDED				TWO WIRE SHIELDED				THREE WIRE SHIELDED		ONE WIRE DOUBLE SHIELDED	
CONFIGURATION												
WIRE SIZE	16	18	20	22	16	18	20	22	20	22	22	22
CONDUCTOR TO SHIELD pf/ft	89	91	74	98	66	65	64	62.5	60	52	98 (CONDUCTOR TO INNER SHIELD)	340 (INNER SHIELD TO OUTER SHIELD)
					(EITHER CONDUCTOR TO SHIELD)				(ANY CONDUCTOR TO SHIELD)			
CONDUCTOR TO CONDUCTOR pf/ft					42	39.5	38	36.5	36	30		

Figure 8-20 Capacitance of Various Shielded AN Wires

Insertion Loss and Parameter Value Calculations for Single-Element Filters (Figure 8-21)

See Equations (7-2) and (7-3) of this publication for the single element filter design equations.

To find the insertion loss at any frequency for a given shunt capacitance, using the chart in Figure 8-21, proceed as follows:

a. Place a straight edge so that it lies parallel to the sloping guide lines and intersects the given value of capacitance on the C scale. The straight edge now lies along the insertion loss characteristic for that ideal element.

b. At any frequency along the abscissa, note where the frequency intersects the straight edge. Read insertion loss in dB along the ordinate corresponding to that intersection.

To find insertion loss at any frequency for a given series inductance, using the chart, proceed as above, but use the L scale instead of the C scale.

To find the amount of shunt capacitance or series inductance required to produce a desired amount of insertion loss at a given frequency, using the chart, proceed as follows:

a. Place a straight edge so that it lies parallel to the guide lines and passes through the desired amount of insertion loss at the given frequency.

b. Read the required capacitance on the C scale or the required inductance on the L scale.

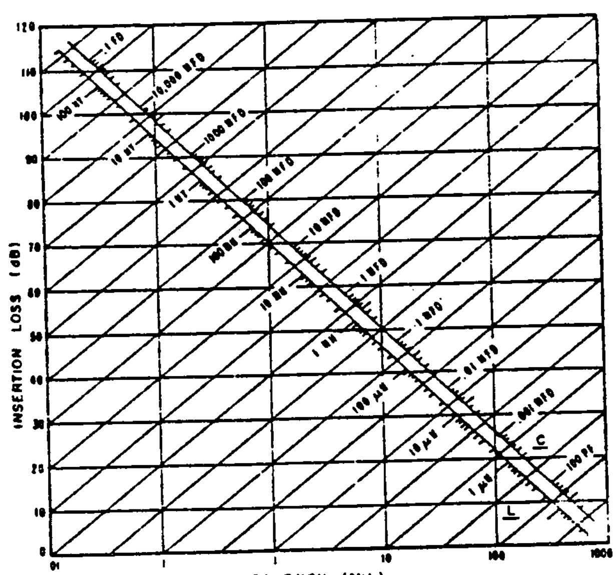


Figure 8-21 Chart For Computing Single-Element Filters

Insertion Loss and Parameter Value Calculations for Ideally-Damped Two-Element Filters (Figure 8-22)

See Equation (7-4) of this publication for the two-element filter design equation.

To find the insertion loss of an ideally-damped "L" section filter (d = 1) at any frequency when f_o is known, using the chart in Figure 8-22, proceed as follows:

a. Place a straight edge parallel to the guide lines and pass it through the known cutoff frequency. The straight edge now lies along the insertion loss characteristic for the ideally-damped L section having that cutoff frequency.

b. At any frequency along the abscissa, read the insertion loss in dB along the ordinate.

To find the cutoff frequency (f_o) required to produce a desired amount of insertion loss at a given frequency,

a. Place a straight edge parallel to the guide lines and pass it through the desired insertion loss at the given frequency.

b. Read the cutoff frequency on the f_o scale.

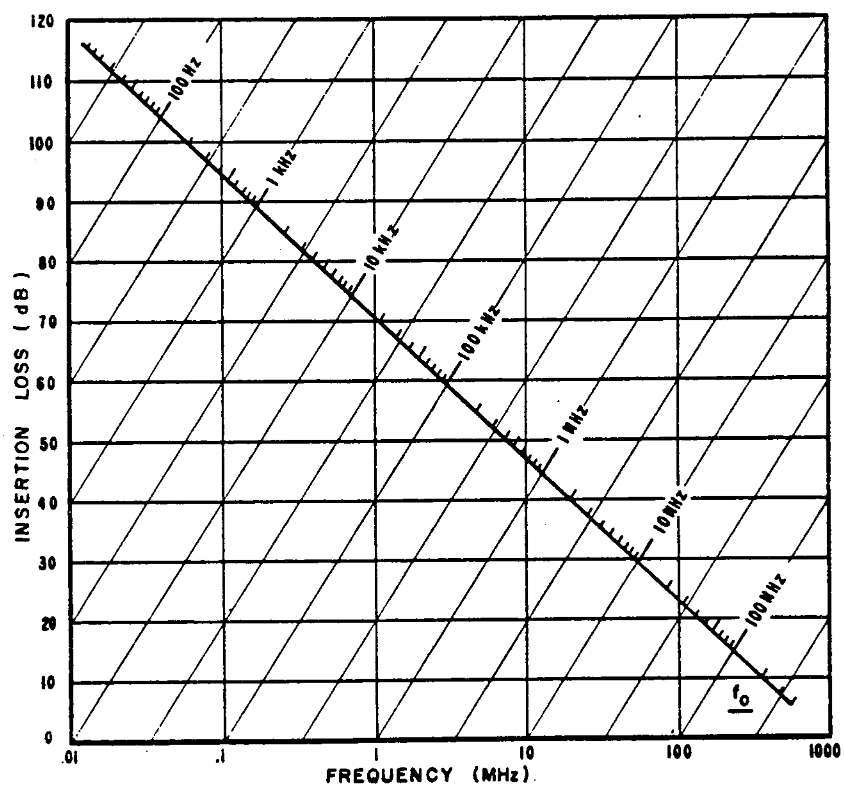


Figure 8-22 Chart for Computing Ideally-Damped Two-Element Filters

Insertion Loss and Parameter Value Calculations for Non-Ideally-Damped Two-Element Filters (Figures 8-23 and 8-24)

See Equation (7-4) of this publication for the two-element filter design equation.

To find the insertion loss of a non-ideally-damped L section filter at any frequency when the cutoff frequency (f_o) and the damping ratio (d) are known, proceed as follows, using Chart A in Figure 8-23.

a. This chart is normalized for $f_o = 1$; thus, divide the frequency of interest by the known f_o to determine the normalized frequency of interest.

b. Locate the characteristic for which d is equal to the known damping ratio. Read the approximate value of d.

c. At the normalized frequency of interest, read the insertion loss in dB.

To establish the relationship between inductance, capacitance, cutoff frequency, and damping factor of two-element filters, use Chart B in Figure 8-24. To find the cutoff frequency and the damping ratio for a given inductance and capacitance:

a. Place a straight edge so that it passes through the known L and C values.

b. Read the cutoff frequency (f_o) on the center scale.

c. The damping ratio slope line which lies parallel to the straight edge indicates the damping ratio. Note that the point at which the straight edge intersects the damping ratio scale is irrelevant; the "d" sloped line parallel to the straight edge determines the damping ratio.

To find the circuit elements required to produce a desired cutoff frequency at a given damping ratio, use Chart B and:

a. Place a straight edge so that it passes through the desired cutoff frequency and lies parallel to the given damping ratio line.

b. Read the inductance and capacitance on the L and C scales.

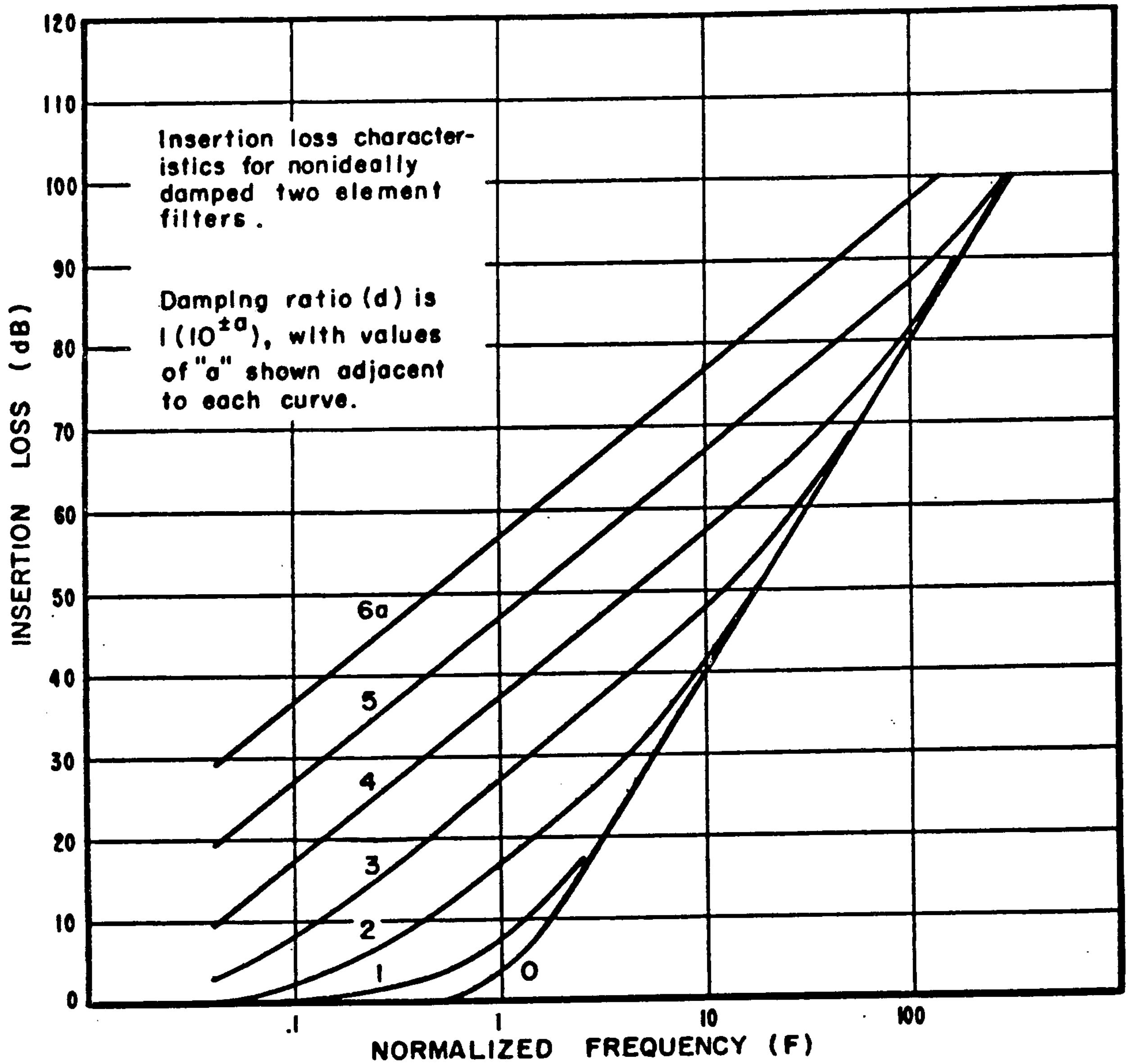


Figure 8-23 Chart A for Computing Non-Ideally-Damped Two-Element Filters

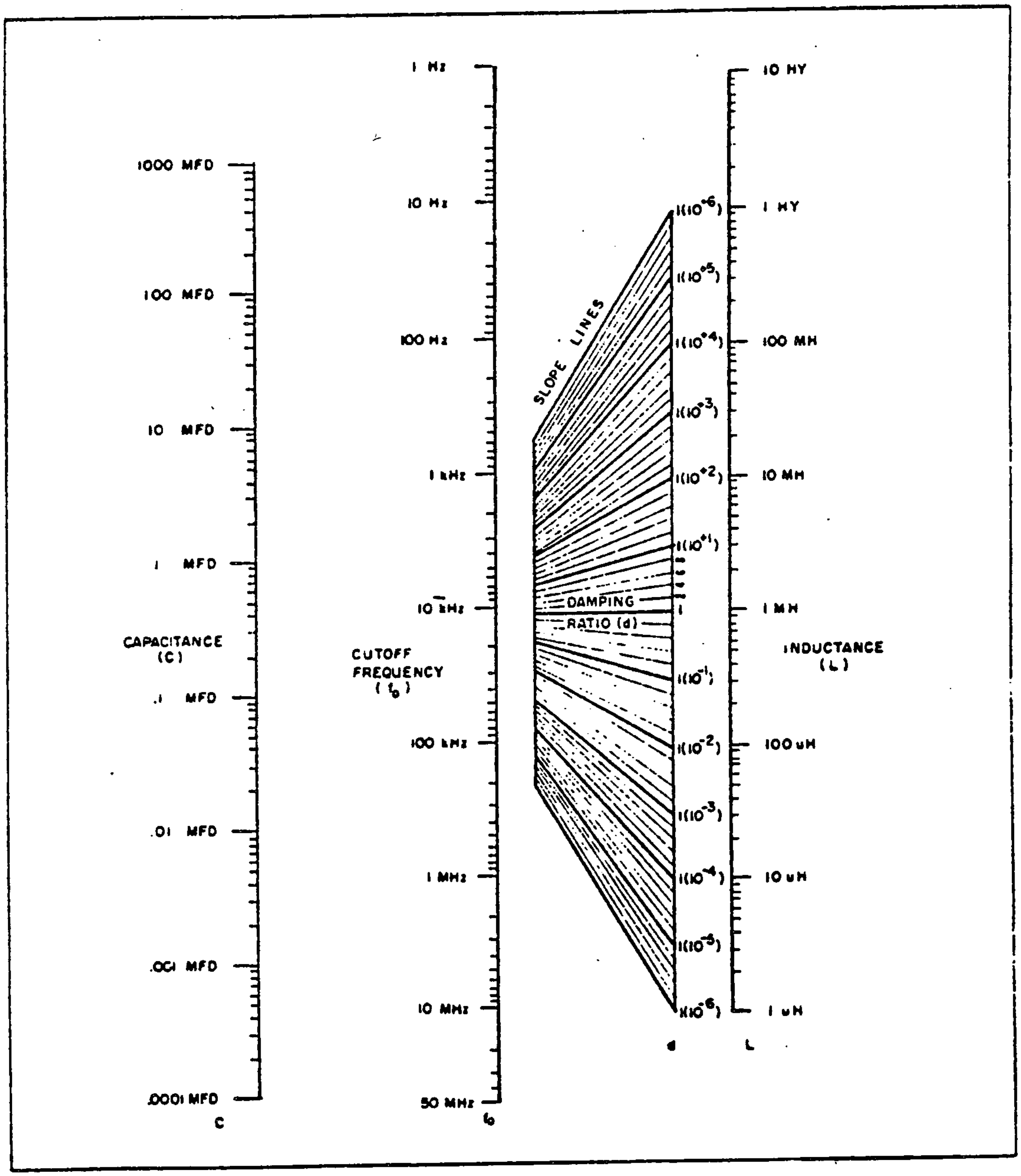


Figure 8-24 Chart B for Computing Two-Element Filters For Any Damping Ratio

Insertion Loss and Parameter Value Calculations for Ideally-Damped Three-Element Filters (Figure 8-25)

See Equations (7-6) and (7-7) of this publication for the three-element filter design equations.

To find the insertion loss of an ideally-damped "T" or "π" section filter (d = 1) at any frequency when f_o is known, using the chart in Figure 8-25, proceed as follows:

a. Place a straight edge parallel to the guide lines and pass it through the known cutoff frequency. The straight edge now lies along the insertion loss characteristic for the ideally-damped L section having that cutoff frequency.

b. At any frequency along the abscissa, read the insertion loss in dB along the ordinate.

To find the cutoff frequency (f_o) required to produce a desired amount of insertion loss at a given frequency,

a. Place a straight edge parallel to the guide lines and pass it through the desired insertion loss at the given frequency.

b. Read the cutoff frequency on the f_o scale.

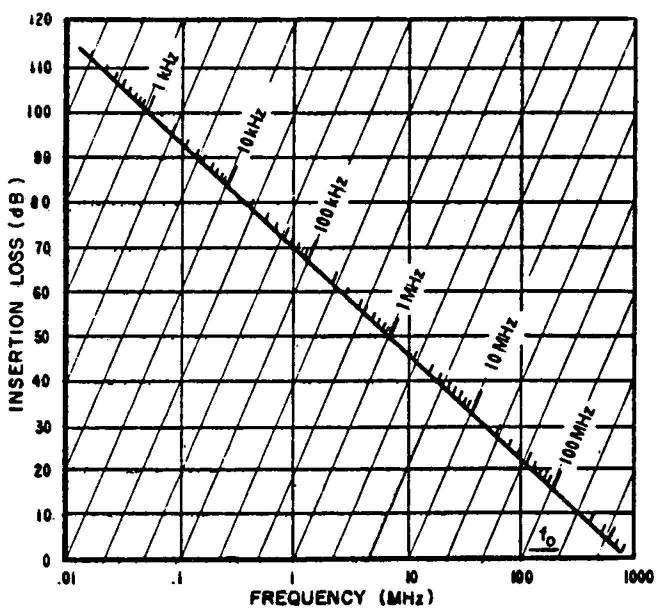


Figure 8-25 Chart for Computing Ideally-Damped Three-Element Filters

Insertion Loss and Parameter Value Calculations for Non-Ideally-Damped Three-Element Filters (Figures 8-26, 8-27, 8-28, and 8-29)

See Equations (7-6) and (7-7) of this publication for the three-element filter design equations.

To find the insertion loss of a non-ideally-damped "T" or "π" section filter at any frequency when the cutoff frequency (f_o) and the damping ratio (d) are known, use Chart A in Figure 8-26 for the overdamped case (d > 1) or Chart B in Figure 8-27 for the underdamped case (d < 1). Proceed as follows:

a. These charts are normalized for $f_o = 1$; thus, divide the frequency of interest by the known f_o to determine the normalized frequency of interest.

b. Locate the characteristic for which d is equal to the known damping ratio. Read the approximate value of d.

c. At the normalized frequency of interest, read the insertion loss in dB.

To establish the relationship between inductance, capacitance, cutoff frequency, and damping factor of three-element filters, use Charts C or D in Figures 8-28 or 8-29 respectively. Chart C applies to π section filters, while Chart D applies to "T" section filters. To find the cutoff frequency and the damping ratio for a given inductance and capacitance:

a. Place a straight edge so that it passes through the known L and C values.

b. Read the cutoff frequency (f_o) on the center scale.

c. The damping ratio slope line which lies parallel to the straight edge indicates the damping ratio. Note that the point at which the straight edge intersects the damping ratio scale is irrelevant; the "d" sloped line parallel to the straight edge determines the damping ratio.

To find the circuit elements required to produce a desired cutoff frequency at a given damping ratio, use Chart C or D and:

a. Place a straight edge so that it passes through the desired cutoff frequency and lies parallel to the given damping ratio line.

b. Read the inductance and capacitance on the L and C scales.

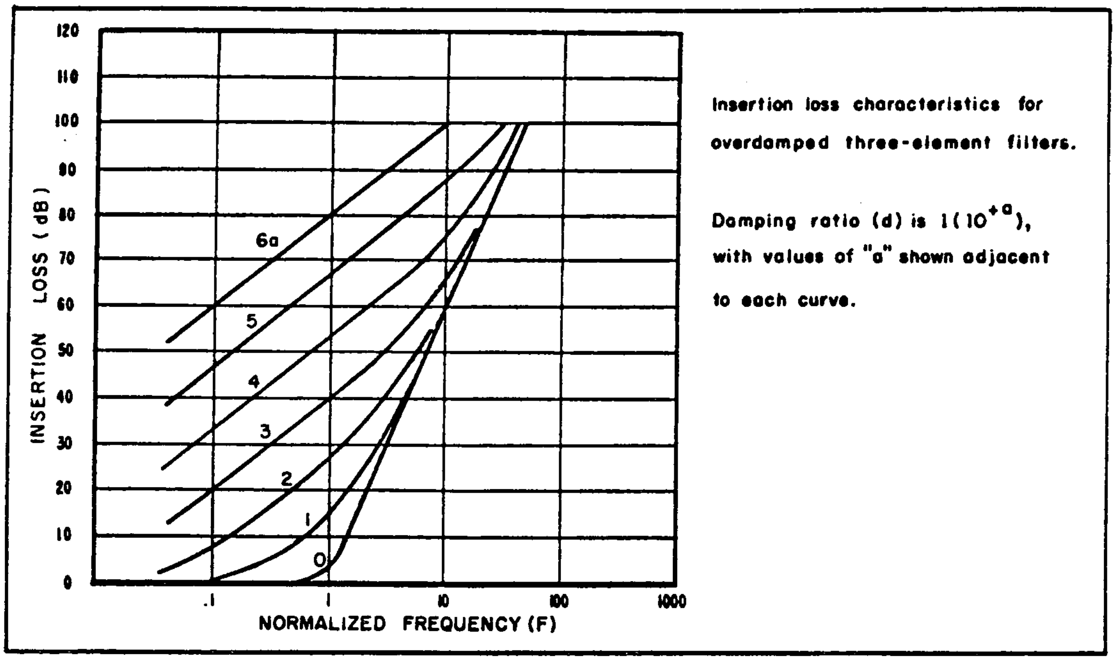


Figure 8-26 Chart A For Computing Overdamped Three-Element Filters

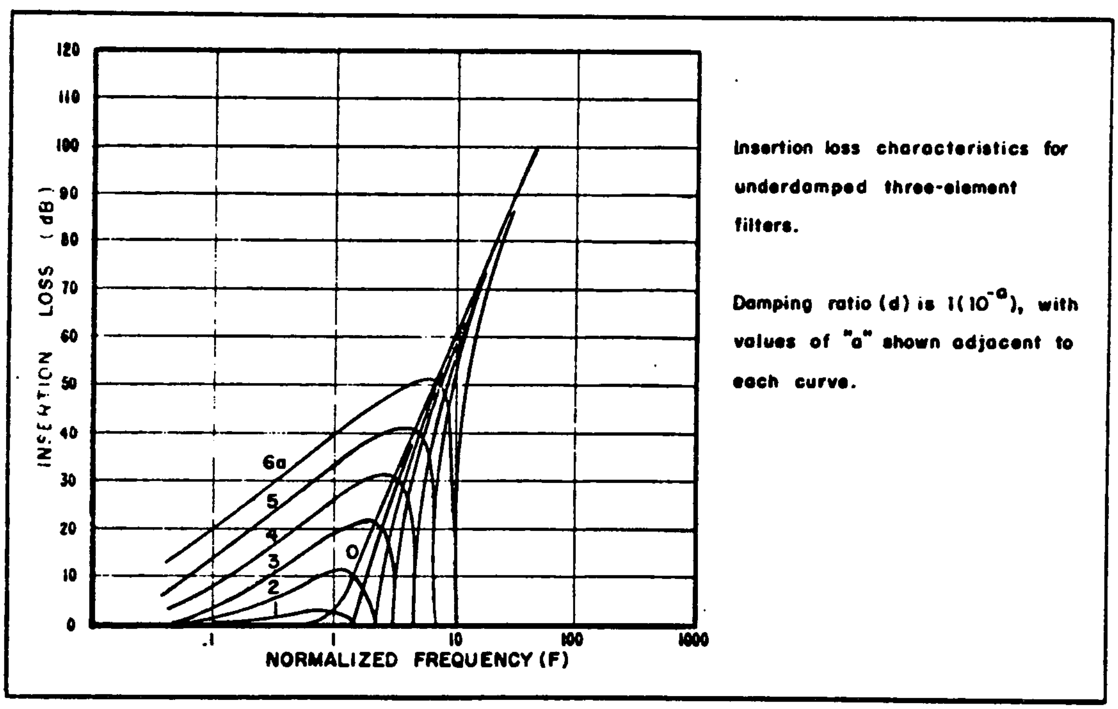


Figure 8-27 Chart B For Computing Underdamped Three-Element Filters

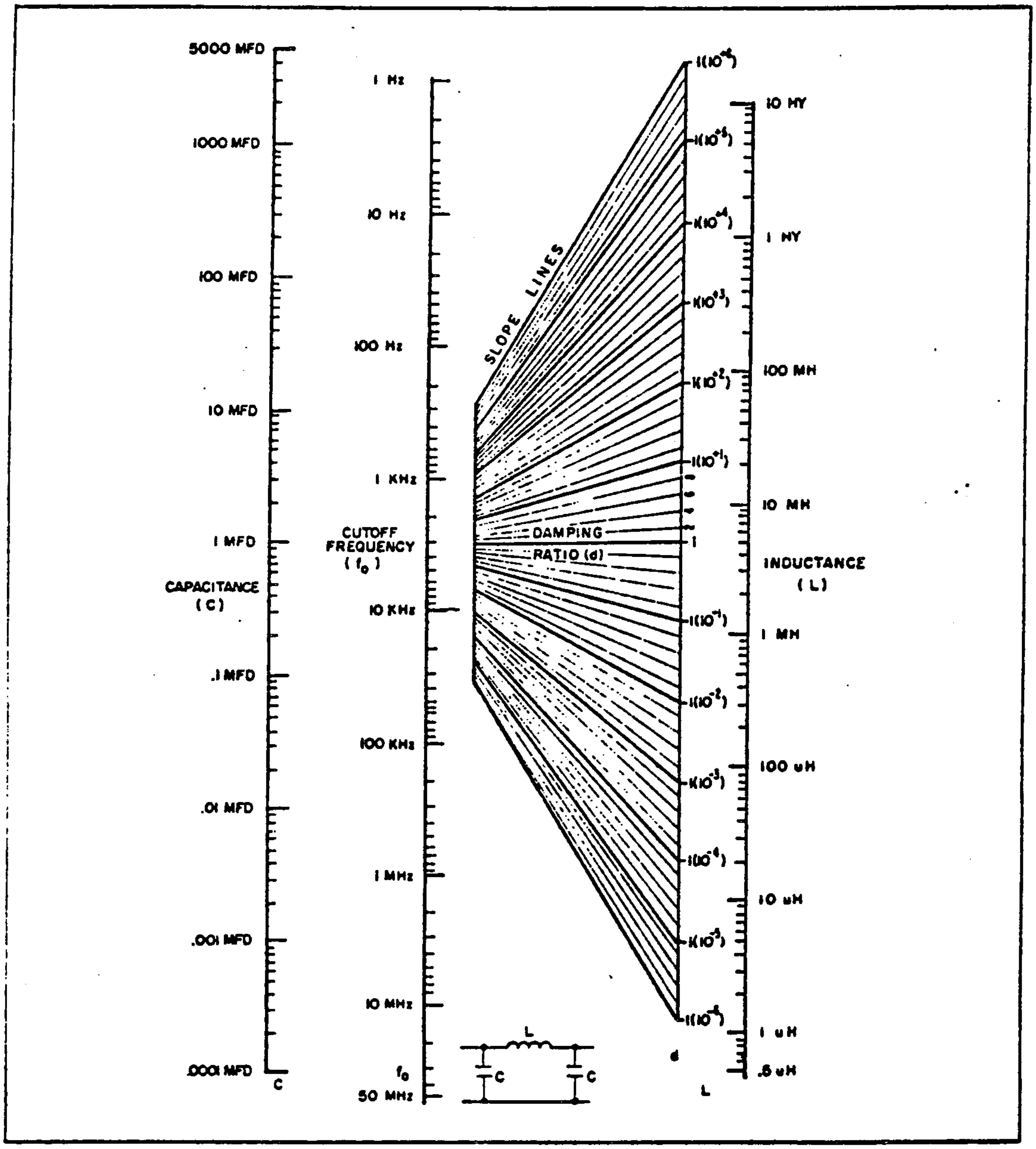


Figure 8-28 Chart C For Computing π Filters For Any Damping Ratio

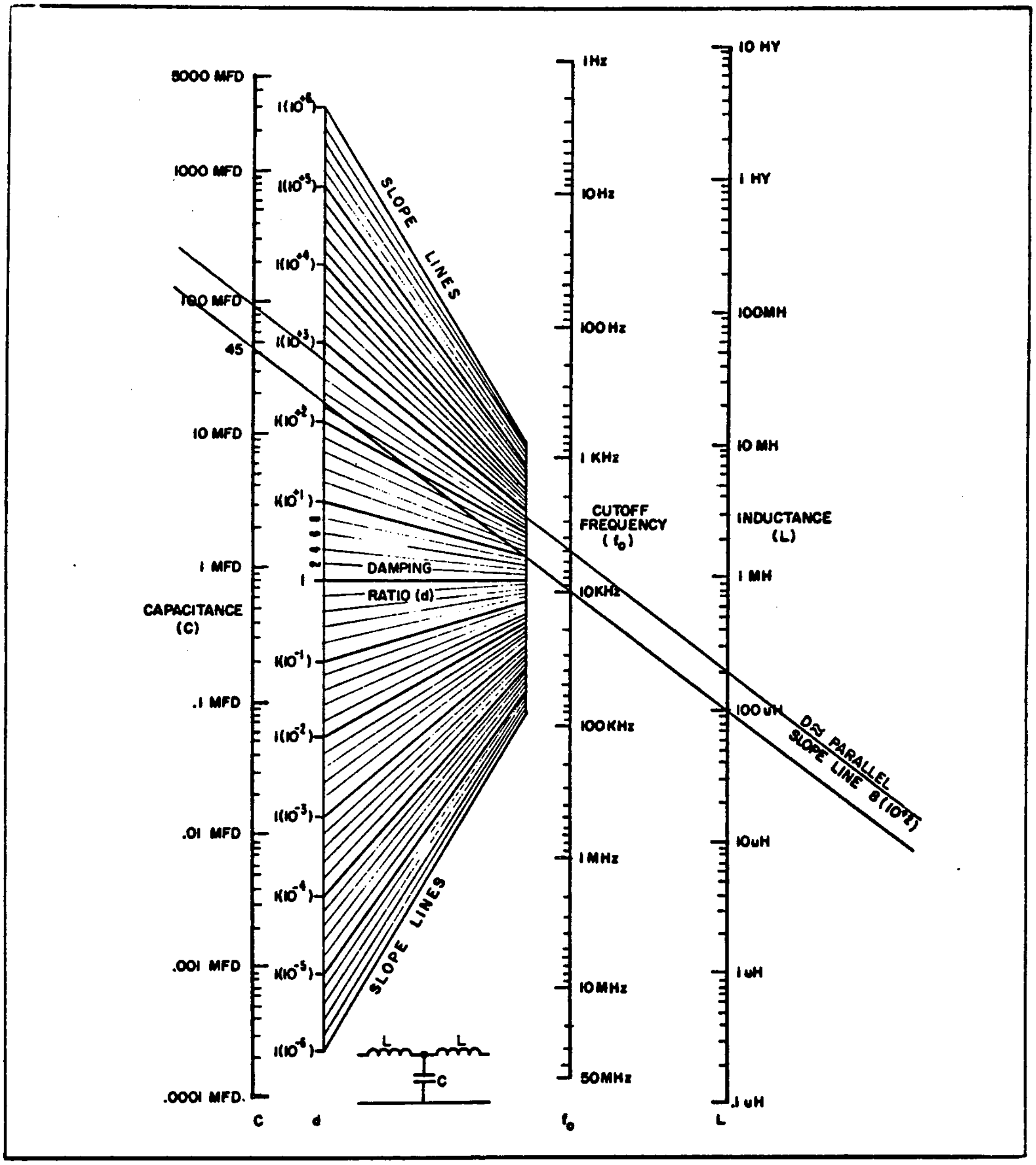


Figure 8-29 Chart D For Computing T Filters For Any Damping Ratio

Parameter Calculations for Twin-T Notch Filters for K = 1 or 2 (Figures 8-30 and (8-31)

See Equation (7-10) of this publication for the twin-T notch filter design equation.

The following nomographs are provided for rapidly computing the parameters of twin-T notch filters when K = 1 or 2. Nomograph A (Figure 8-30) is used to relate the source resistance, load resistance and R, while Nomograph B, (Figure 8-31) identifies the relationship between f_o, R and C [1].

The example shown assumes a source resistance of 300 ohms, a load resistance of 100,000 ohms, a notch frequency of 60 Hertz, and K= 1. Under these circumstances, R from Nomograph A is 7500 ohms, and C from Nomograph B is 0.4 μf.

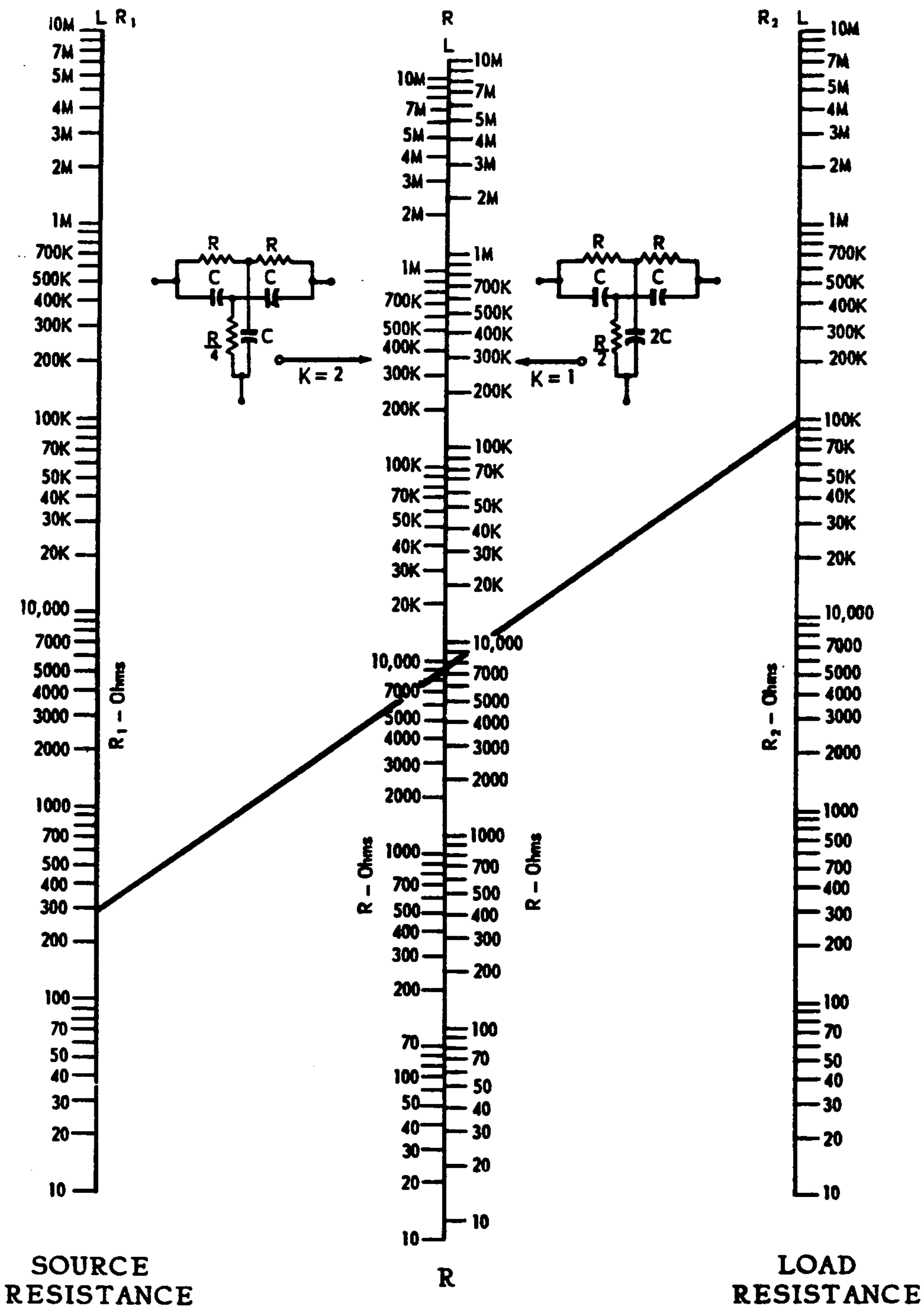


Figure 8-30 Nomograph A for Determining Resistance Values for Twin-T Notch, K = 1 and K = 2

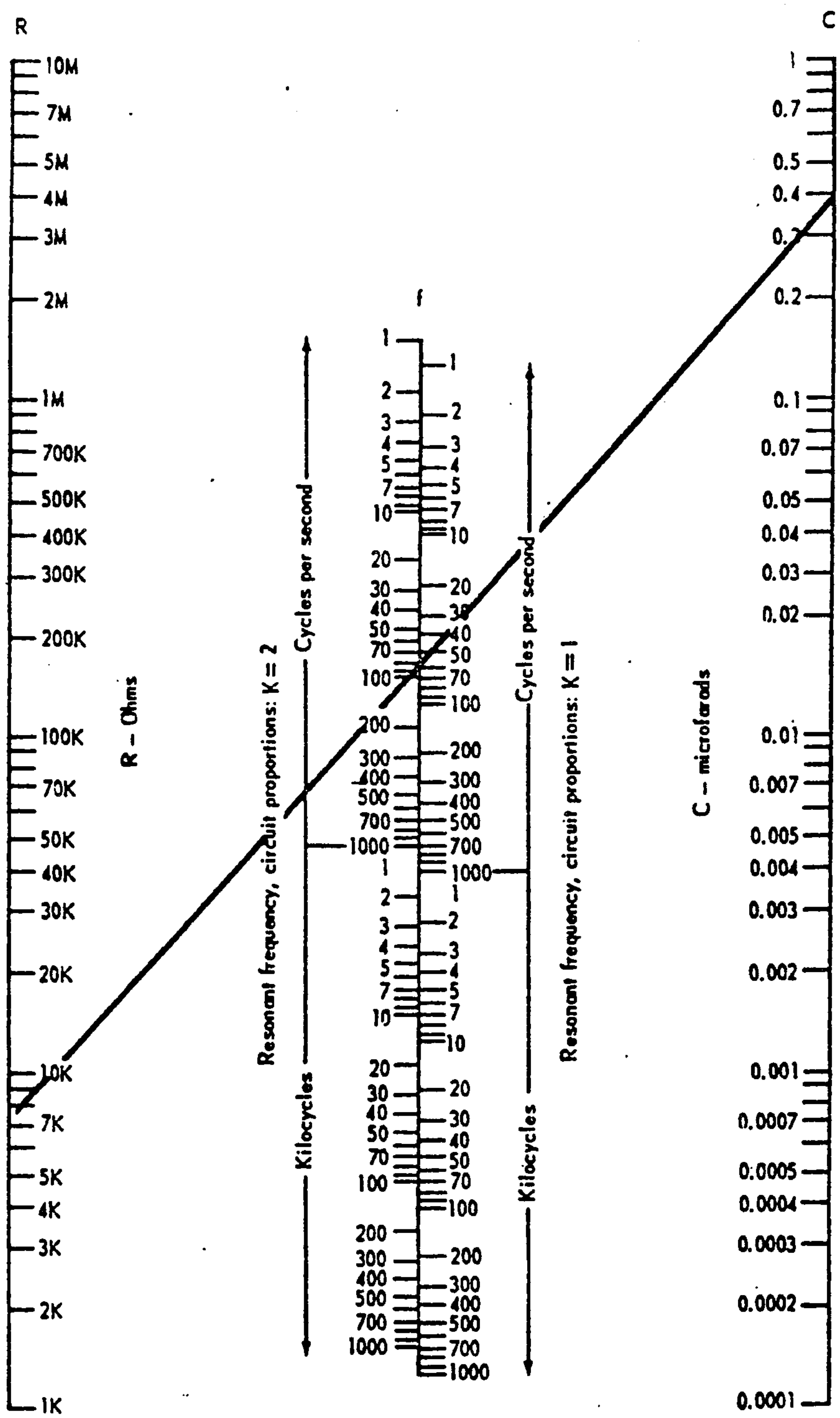


Figure 8-31 Nomograph B for Determining Capacitor Values for Twin-T Notch, K = 1 and K = 2

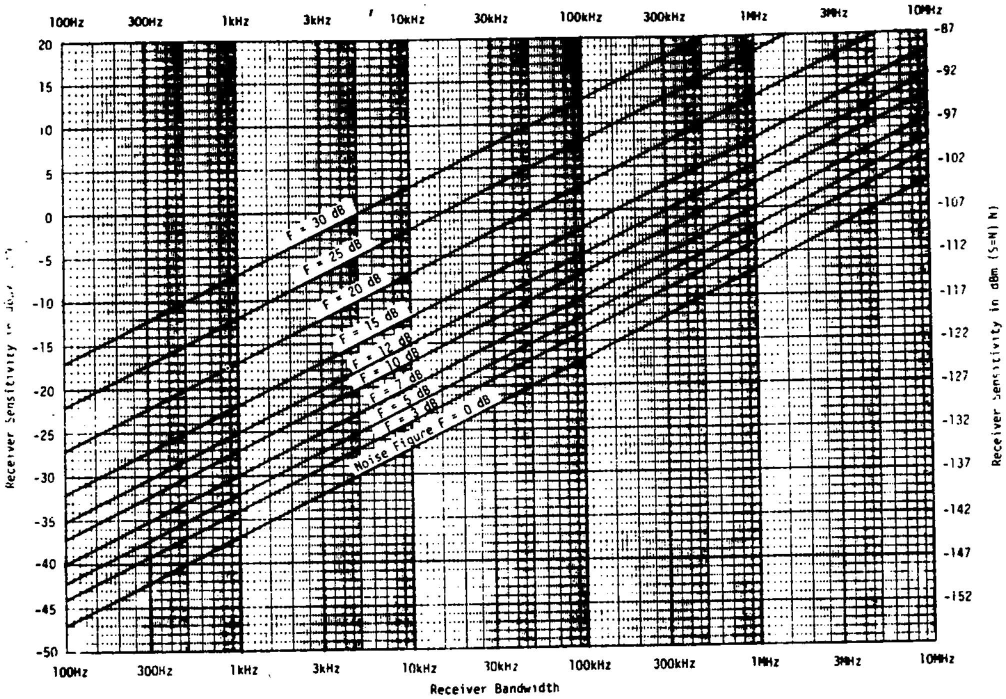


Figure 8-32 Receiver Narrowband Sensitivity as a Function of Noise Figure and Bandwidth

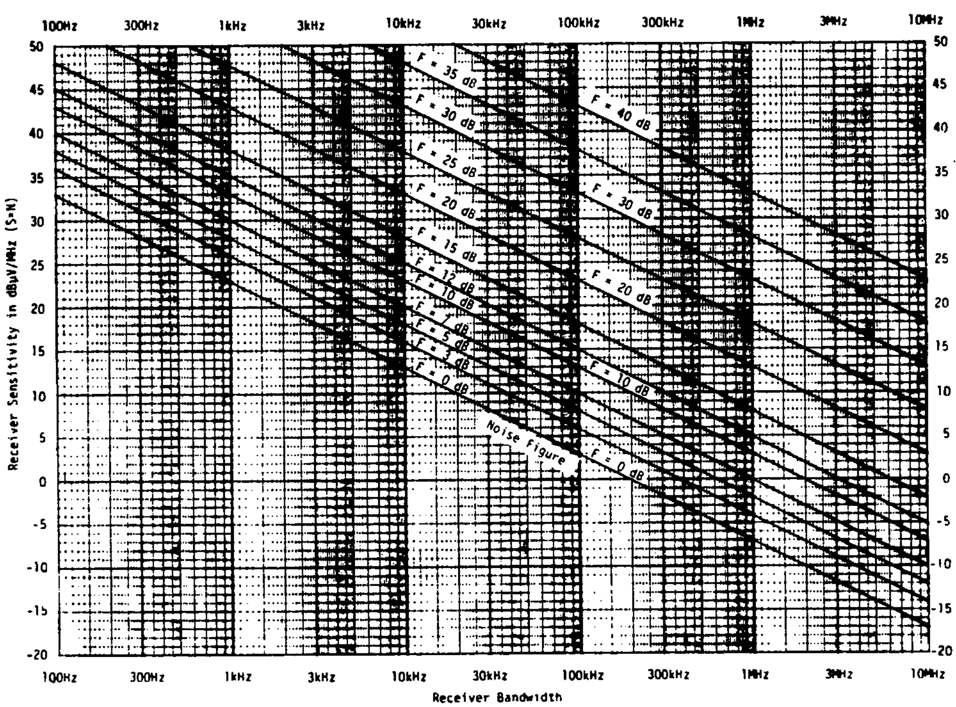


Figure 8-33 Receiver Broadband Impulse Noise Sensitivity as a Function of Noise Figure and Bandwidth

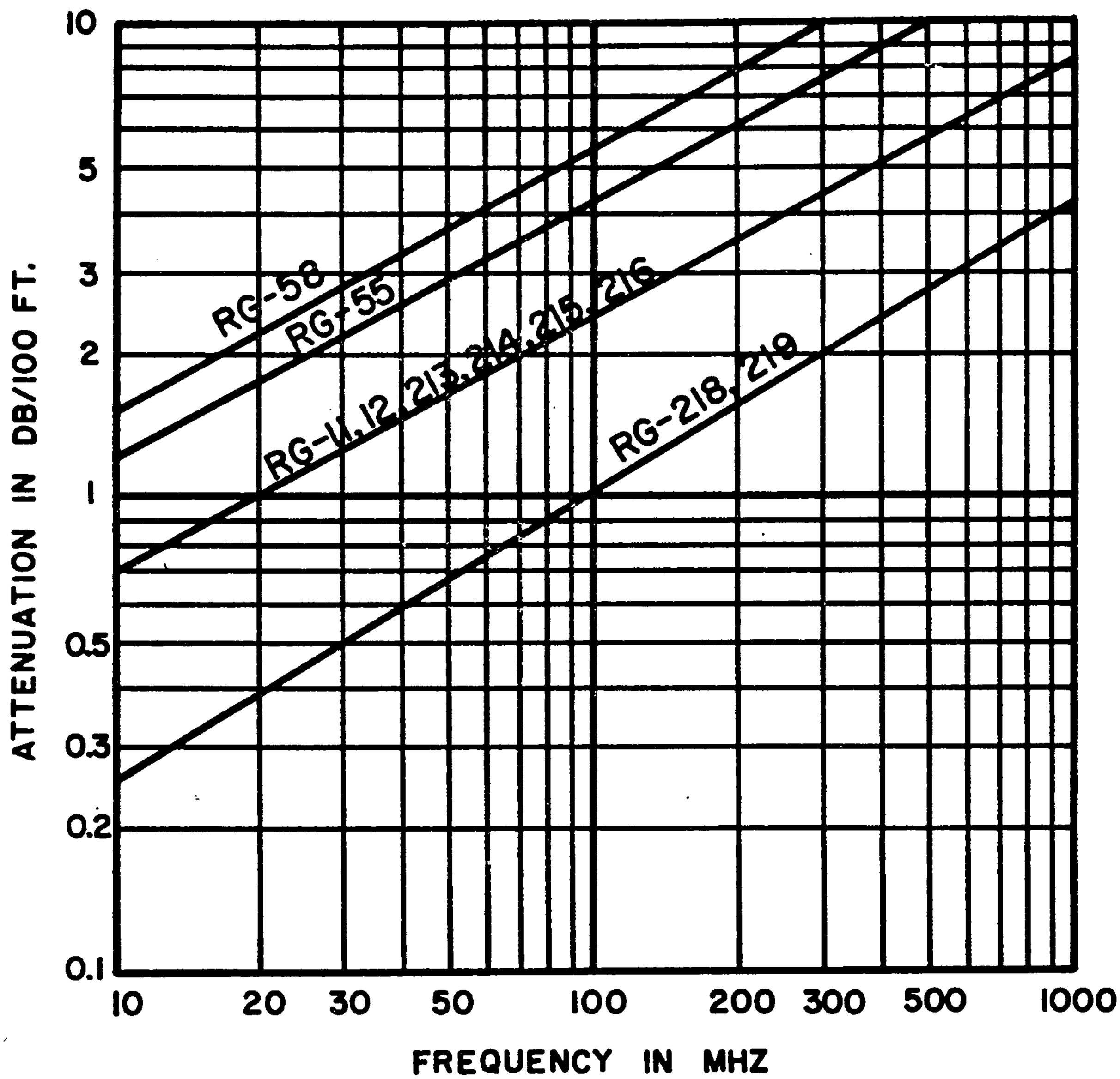


Figure 8-34 Attenuation of Coaxial Cable

MISCELLANEOUS CONVERSIONS

λ (meters) = $300/f_{(MHz)}$

λ (feet) = $984/f_{(MHz)}$

1 Meter = 3.28 feet = 6.214×10^{-4} statute miles

1 Foot = .3048 meters

1 Mile (statute) = 5280 feet = 1609 meters

Statute Miles = Nautical Miles x 1.1508

Free Space Impedance, Z = 377 ohms

Diameter of the Earth = 7917 nautical miles
= 9112 statute miles

Maximum line of sight distance over smooth earth in statute miles,

$$D_{(st.\ miles)} = (2h_1)^{1/2} + (2h_2)^{1/2}$$

h_1 is the height of the first antenna, in feet

h_2 is the height of the second antenna, in feet

$$P_p = \frac{P_A}{\text{duty cycle}}$$

BASIC NOISE POWER IN A SYSTEM

$$P_N = KT\beta$$

$K = 1.380 \times 10^{-23}$ joule/degree Kelvin

degrees Kelvin = °C + 273

room temperature is taken as 68° F = 20° C = 293° K

$P_N = 1.380 \times 10^{-23} \times 273 \times 1$ for $\beta = 1$ Hertz and P_N in watts

$P_N = 4.05 \times 10^{-21}$ watts/Hertz or 4.05×10^{-18} milliwatts/Hertz

P_N = -173 dBm/Hertz

To convert to wider bandwidths, add 10 log β in Hertz

BASIC NOISE POWER IN A SYSTEM (Continued)

Example: for β = 1 MHz, 10 log 10^6 = 60

$$P_N = -173 + 60 = -113 \text{ dBm/MHz}$$

To compute real noise level in a receiver, add the noise figure in dB to the computed noise power.

$$P_N = P_N + NF \text{ (dB)}$$

For a typical 1 MHz receiver, NF = + 9dB

$$\text{Sensitivity} = -113 + 9 = -104 \text{ dBm}$$

POWER DENSITY RELATIONSHIPS

$$\text{dBm/m}^2 = \text{dBm/ft}^2 + 10.3$$

$$\text{dBm/m}^2 = \text{dBm/cm}^2 + 40$$

$$\text{dBm/m}^2 = 10 \log (\text{mW/cm}^2) -10$$

$$61.4 \sqrt{\text{mW/cm}^2} = \text{V/m}$$

POWER-VOLTAGE CONVERSIONS

For a 50 ohm load: $P_{dBm} = E_{dB\mu V/m} - 107$

For a 72 ohm load: $P_{dBm} = E_{dB\mu V/m} - 108.6$

For a 300 ohm load: $P_{dBm} = E_{db\mu V/m} - 114.8$

For a 377 ohm load (free space): $\xi_{(dB\mu/m)} = P_{d(dBm/m^2)} + 116$

PATH LOSS AND PROPAGATION LOSS

Path Loss in dB (Free space propagation between isotropic antennas)

(1) for R in feet $L_i = 20 \log f_{MHz} + 20 \log R_{(feet)} - 37.9$

(2) for R in yards $L_i = 20 \log f_{MHz} + 20 \log R_{(yards)} - 27.5$

(3) for R in meters $L_i = 20 \log f_{MHz} + 20 \log R_{(meters)} - 26.8$

(4) for R in K yards $L_i = 20 \log f_{MHz} + 20 \log R_{(K\ yards)} + 32.5$

(5) for R in miles $L_i = 20 \log f_{MHz} + 20 \log R_{(miles)} + 36.5$

(6) for R in Nautical miles $L_i = 20 \log f_{MHz} + 20 \log R_{(N\ miles)} + 37.8$

Propagation Loss in dB (Free space propagation)

Gain-Coupling Loss

(1) for R in feet $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(feet)} - G_t - G_r - 37.9$

(2) for R in yards $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(yards)} - G_t - G_r - 27.5$

(3) for R in meters $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(meters)} - G_t - G_r - 26.8$

(4) for R in K yards $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(K\ yards)} - G_t - G_r + 32.5$

(5) for R in miles $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(miles)} - G_t - G_r + 36.5$

(6) for R in Nautical miles $L_{gc} = 20 \log f_{MHz} + 20 \log R_{(N\ miles)} - G_t - G_r + 37.8$

REDUCTION OF POWER DENSITY TO TRANSMITTER POWER OUTPUT

(1) $P_{t(dBm)} = P_{d(dBm/m^2)} - G + 20 \log R + 11$ (R in meters)

(2) $P_{t(dBm)} = P_{d(dBm/m^2)} - G + 20 \log R + 0.7$ (R in feet)

(3) $P_{t(dBm)} = P_{d(dBm/m^2)} - G + 20 \log R + 75$ (R in miles)

(1, 2, 3, hold for free space propagation with P_d measured in the far field.)

REDUCTION OF FIELD STRENGTH TO EFFECTIVE RADIATED POWER (P_{erp}, dBm)

(1) $P_{erp\ (dBm)} = \xi_{(db\mu v/m)} + 20 \log R - 105$ (R in meters)

(2) $P_{erp\ (dBm)} = \xi_{(dB\mu v/m)} + 20 \log R - 115$ (R in feet)

(3) $P_{erp\ (dBm)} = \xi_{(dB\mu v/m)} + 20 \log R - 41$ (R in miles)

(1, 2, 3, hold for free space propagation only.)

POWER, VOLTAGE, FIELD STRENGTH AND POWER DENSITY RELATIONSHIPS

1. Effective area A_e, and effective length L_e, for the same antenna in a constant field:

$A_e = L_e^{\ 2} \dfrac{377}{4R_L}$ where R_L is the antenna matched load

2. Field Strength and Received Voltage:

$$\xi_{(dB\mu V/m)} = E\ (_{dB\mu V/m}) - 10 \log G_r + 20 \log f_{MHz} - 29.8$$

3. Power and Volatage in a 50 ohm load:

$$P_{(dBm)} = E_{(dB\mu V/m)} - 107$$

4. Power Density and Field Strength (Free Space Z = 377):

$$P_d(dBm/m^2) = \xi(_{dB\mu V/m}) - 116$$

5. Power Density and Received Voltage (for P_d at the Rx antenna):

$$P_d(dBm/m^2) = E_R(_{dB\mu V/m}) - G_R(dB) + 20 \log f_{MHz} - 146$$

6. Power Density and Received Power (for P_d at the Rx antenna):

$$P_d(dBm/m^2) = P_R(dBm) - G_R(dB) + 20 \log f_{MHz} - 38.5$$

FREE SPACE PROPAGATION VELOCITY

186,280 Statute Miles/Second

161,300 Nautical Miles/Second

984×10^6 Feet/Second

328×10^6 Yards/Second

299.8×10^6 Meters/Second

Coupling of EM Fields Into Wire Loops

The nomographs in Figures 8-35 and 8-36 allow an EMC engineer to determine the electromagnetic field induced open circuit voltage in a rectangular loop (hxL) such as is found in common mode coupling.

Analysis of the magnitude of these signals can be performed by determining the equivalent loop aperture of the wire. An isolated single wire can be analyzed as a loop by considering the end-to-end capacitance as the return path. The wire would therefore act as an unintentional antenna. A pair of wires in a cable could similarly be considered as a loop when the capacitive and inductive coupling between them is taken into account. These effects can be differential mode and common mode. The loop for differential mode signals between two wires would be defined by the common run length and spacing between the wires. The signal would be 180 degrees out-of-phase between the wires, therefore, "differential-mode". The loop for common mode signals would be defined by the wires and the closest ground plane. The induced signal would be in-phase on both wires, hence, "common-mode". The open circuit voltage induced in a loop by either an electric or magnetic field is a function of the area of the loop and its orientation in the field.

For an electric field;

$$V_i/E = 2L \cos\Theta \sin\left(\frac{\pi h}{\lambda}\cos\alpha\right), \text{ for h and L} \ll \lambda$$

where V_i = open circuit induced voltage (volts)
E = electric field (V/m)
h = height of loop (meters)
L = length of loop (meters)
λ = wavelength (meters)
α = angle between plane of loop and direction of propagation of field
Θ = angle between the wire and the E-field polarization

The maximum value would occur when the loop is orthogonal to the direction of propagation and the wire and field polarization are in alignment. In such a condition, $\cos\Theta = \cos\alpha = 1$.

Therefore,

$$(Vi/E)_{max} = 2L \sin\left(\frac{\pi h}{\lambda}\right), \text{ for h and L} \ll \lambda$$

$$\sim \frac{2\pi\, Lh}{\lambda}$$

$$\sim 2.1\ (10)\ LhF_{Hz}$$

Similarly, magnetic coupling,

$$Vi/B = 2Lc\ (\sin\Theta)\ \left(\sin\left(\frac{\pi h}{\lambda}\cos\alpha\right)\right), \text{ for h and L} \ll \lambda$$

where

Vi = open circuit induced voltage (Volts)
B = magnetic flux density (Gauss)
λ = wavelength (meters)
c = velocity of propagation = 3×10^8 m/s
h = height of loop (meters)
L = length of loop (meters)
α = angle between plane of the loop and the direction of propagation of the field
Θ = angle between the wire and plane orthogonal to the field

The maximum, therefore, is

$$(Vi/B)_{max} = 2Lc \sin\left(\frac{\pi h}{\lambda}\right) \text{ radians, for h and L} \ll \lambda$$

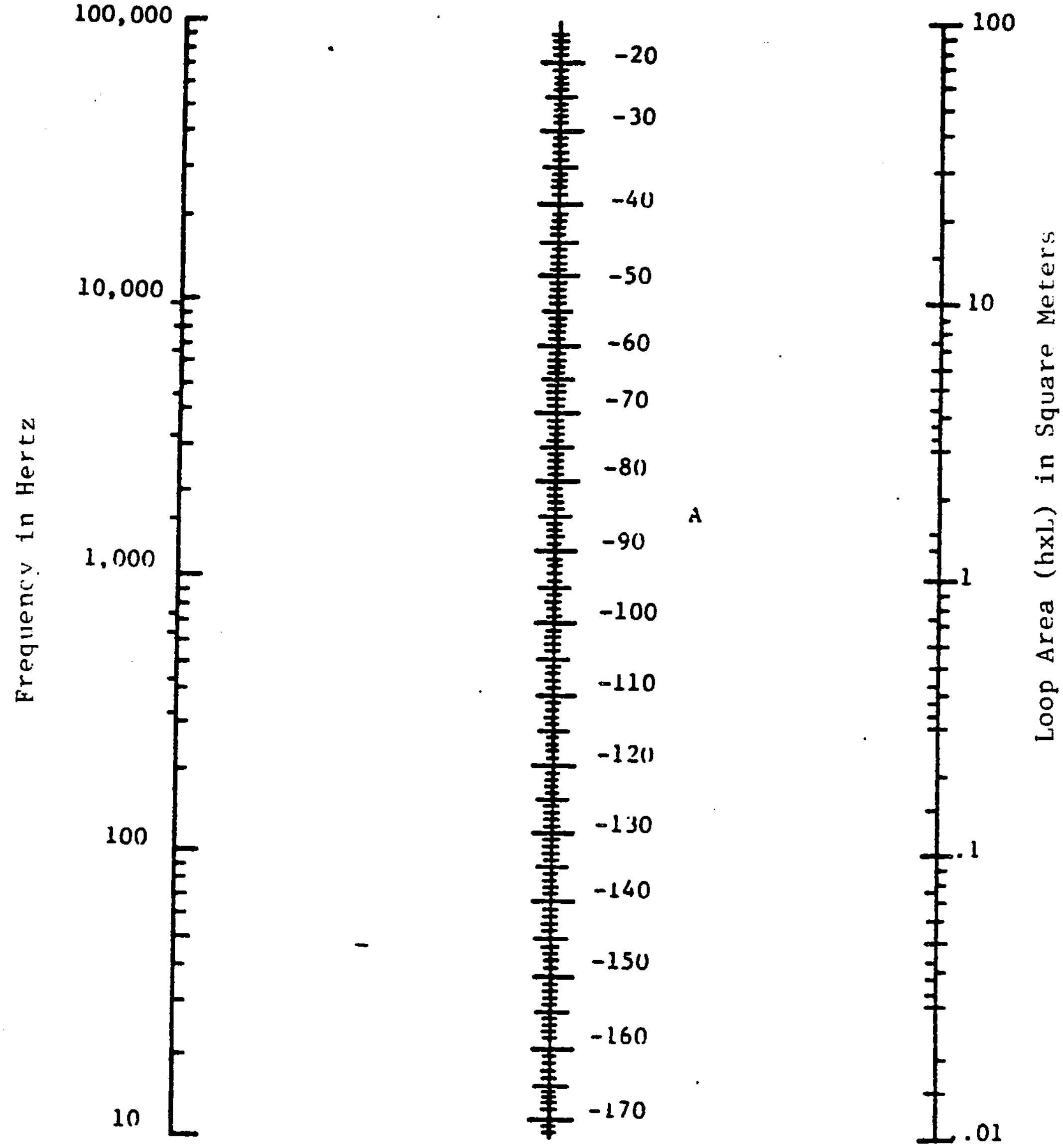
$$\sim \frac{2\pi LhC}{\lambda}, \text{ for h and L} \ll \lambda$$

$$\sim 6.28 \times 10^{-4}\, hLF_{Hz}$$

The maximum coupling values in dB,

$20 \log (Vi/E)_{max}$, dB (meter) and $20 \log (Vi/B)_{max}$, dB(VT)

as a function of loop area (hL) can be determined using the nomographs.



A(dB Meter) = 153.56 + 20 log (hL) + 20 log F_{Hz}, h and L

V (dBV) = A + E (dBV/M)

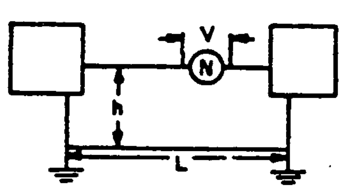


Figure 8-35 Maximum Open Circuit Loop Voltage Due to Electric Field

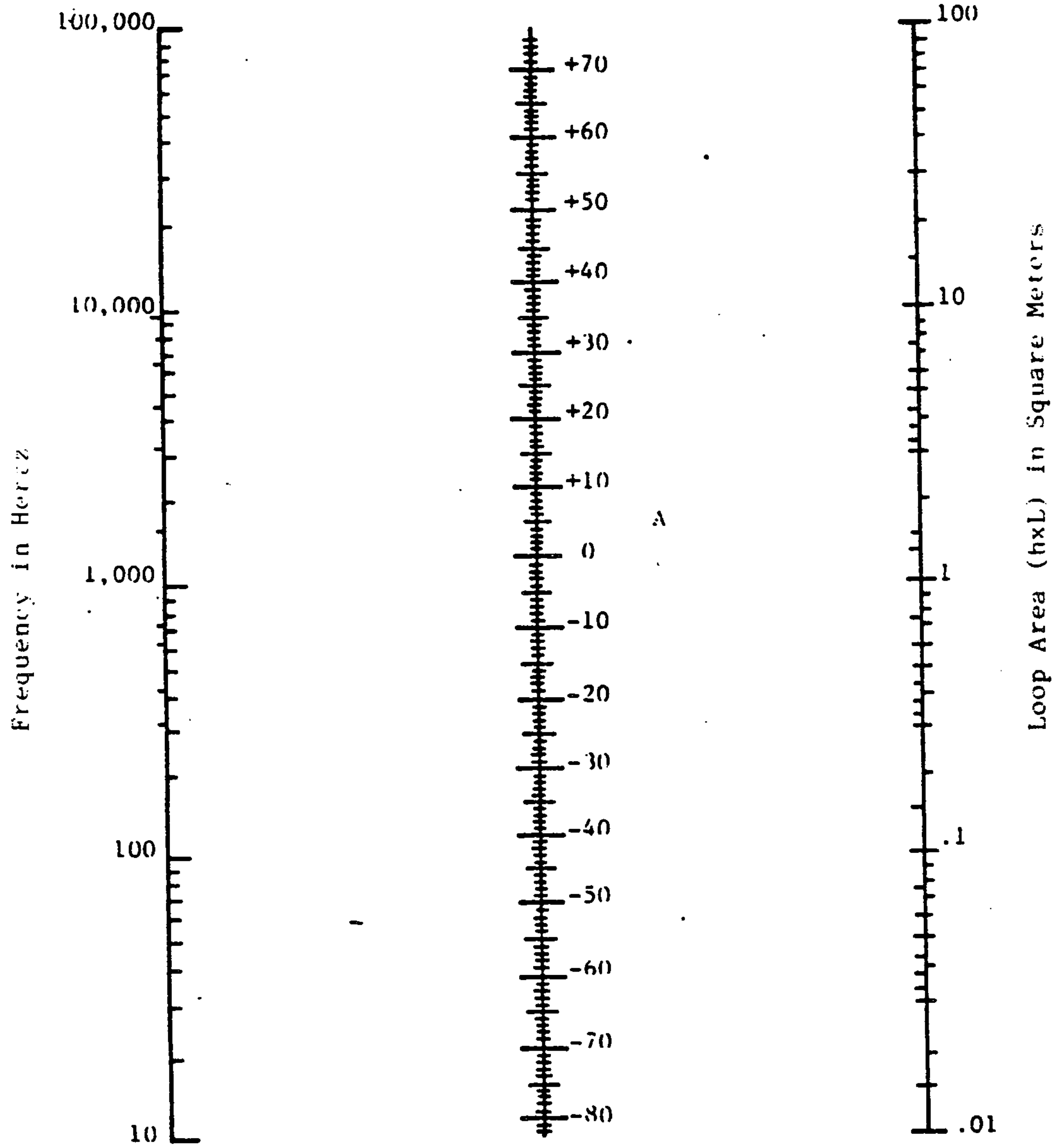


Figure 8-36 Maximum Open Circuit Loop Voltage Due to Magnetic Field

TO CONVERT A POWER DENSITY (mW/cm^2) TO FIELD INTENSITY (V/m) OR VISA VERSA

$$\text{Field Intensity in (V/m)} = 61.4 \sqrt{\text{Power Density in } (mW/cm^2)}$$

Ex. To convert a Power Density of 9 mW/cm^2 to a Field Intensity

$$V/m = 61.4 \sqrt{9\ mW/cm^2}$$

$$V/m = 61.4\ (3)$$

$$V/m = 184.2$$

so 9 mW/cm^2 = 184.2 V/m

TO CONVERT A FIELD INTENSITY (V/m) TO POWER DENSITY (mW/cm^2)

$$\left(\frac{\text{Field Intensity (V/m)}}{61.4}\right)^2 = \text{Power Density } (mW/cm^2)$$

Ex. To convert 200 V/m to a Power Density (mW/cm^2)

$$\left(\frac{200\ V/m}{61.4}\right)^2 = mW/cm^2$$

$$(3.26)^2 = mW/cm^2$$

$$10.61 = mW/cm^2$$

So 200 V/M = 10.61 mW/cm^2

Another Ex. Convert a Power Density of 4 mW/cm^2 to a Field Intensity (V/m)

$$V/m = 61.4 \sqrt{4\ mW/cm^2}$$

$$V/m = 61.4\ (2)$$

$$V/m = 122.8$$

So 4 mW/cm^2 = 122.8 V/m

Another Ex. Convert a Field Intensity of 61.4 V/m to a Power Density (mW/cm^2)

$$\left(\frac{61.4\ V/m}{61.4}\right)^2 = mW/cm^2$$

$$(1)^2 = mW/cm^2$$

$$1 = mW/cm^2$$

So 61.4 V/m = 1 mW/cm^2

DUTY CYCLE

Duty Cycle (DC) = Pulse Repetition Frequency (PRF) in Hz X Pulse Width (PW) in μsec.

Ex. A Pulsed Radar operating at 1000 Hz and 2 μsec or 2 X 10^{-6} sec has a Duty Cycle of

$$D.C. = 1000\ Hz\ (2 \times 10^{-6}\ sec)$$

$$D.C. = 2000 \times 10^{-6}\ Hz\ sec \qquad Hz = \frac{1}{sec}$$

$$D.C. = .002\ \ 1/sec\ (sec)$$

$$D.C. = .002$$

AVERAGE POWER DENSITY TO PEAK POWER DENSITY

To Derive a Peak Power Density (P_P) from the Average Power Density (P_A) of a Pulsed Radar

Average Power Density (P_A) = Peak Power Density (P_P) X Duty Cycle (D.C.)

$$P_A = P_P \times D.C.$$

$$\frac{P_A}{D.C.} = P_P$$

Ex. So a Pulsed Radar generating an Average Power Density of 9 mW/cm^2 with a Pulse Width of 2 μsec or 2 X 10^{-6} sec and a Pulse Repetition Frequency of 1000 Hz is generating a Peak Power Density of

$$P_P = \frac{P_A}{D.C.} \qquad D.C. = PRF \times PW$$

$$P_P = \frac{9}{1000 \text{ Hz} \times 2 \times 10^{-6} \text{ sec}}$$

$$P_P = \frac{9}{.002 \text{ Hz per sec}}$$

$$P_P = 4500 \text{ mW/cm}^2$$

Ex. A Pulsed Radar generating a Peak Power Density of 9000 mW/cm^2 with a Pulse Width of 1 μsec or 1 X 10^{-6} sec and a PRF of 1000 Hz is generating an Average Power Density of

$$P_A = P_P \times D.C.$$

$$P_A = (9000 \text{ mW/cm}^2)\ (1 \times 10^{-6} \text{ sec})\ (1000 \text{ Hz})$$

$$P_A = 9000 \text{ mW/cm}^2\ (.001 \text{ sec Hz})$$

$$P_A = 9 \text{ mW/cm}^2\ (\text{sec})\ (1/\text{sec})$$

$$P_A = 9 \text{ mW/cm}^2$$

<u>dBm to mW/cm^2</u>

$$\text{Power (in dBm)} = 10 \log \text{Power (in mW/cm)}^2$$

$$\text{Power (in mW/cm}^2) = 10^x$$

$$\text{where } X = \frac{\text{Power (in dBm)}}{10}$$

AVERAGE FIELD INTENSITY TO PEAK FIELD INTENSITY

$$\frac{V/m_{ave}}{D.C.} = V/m_{Peak} \qquad \text{for Pulse Modulation}$$

where D.C. = Duty Cycle

$$V/m_{ave} = V/m_{Peak} \left(\sqrt{D.C.}\right) \qquad \text{for Pulse Modulation}$$

$$V/m_{max} = A_o (1+m) \qquad \text{for AM}$$

where A_o = carrier voltage

m = % of Modulation/100

$$V/m_{max} = A_o \qquad \text{for FM}$$

where A_o = carrier voltage

REFERENCES

[1] From J.F. Sorado, "Twin-T Network Design" 1958 Electrical Design Manual, Rogers Printing Co., Englewood, CA, 1958, pp. 27-30.

9.0 EMC MANAGEMENT PROGRAM

9.1 GENERAL

The success in controlling EMI effects during Avionics/GSE design is very often dependent on the extent to which an EMC management program is established and implemented. This program is necessary whether the equipment under development is a sophisticated aircraft radar, or a relatively simple and self-contained HI-POT and continuity tester, although the type of management effort required in these two instances will obviously be distinctly different.

This chapter of the design guide provides guidance to the Avionics/GSE developer in organizing and implementing efficient interference regulation. Specifically, it discusses one of the major documents that will be needed to achieve a high level of EMI control. This document, which is normally prepared by the design activity, is the EMC Control Plan.

9.2 THE EMC CONTROL PLAN

9.2.1 The Role of the Control Plan

The Interference Control Plan is the formal document that will place into effect the control procedures and activities to be employed during an equipment design/development cycle. In the case of a hardware contractor, the plan will respond directly to commitments of a bid proposal or contract. For government in-house developers, the plan will reply to directives that define or interpret the design and operational objectives of the Avionics/GSE.

The Interference Control Plan is the document that should communicate to each engineer, each department head, each subcontractor, and each representative of the procuring agency the EMC work effort, emphasis and guidance to be imposed during the project. It is the mechanism for establishing EMC engineering and management milestones, including points in time at which the effectiveness of the program can be judged, and is the device in which individual and group responsibilities are defined.

The importance of the EMC Control Plan cannot be emphasized enough. Since it will outline the entire program by which subsequent EMC documentation, requirements, specifications, test methods, and criteria are controlled, the manner in which the plan is developed can be seen to have an important bearing on successful achievement of EMC.

Even when not called for separately, an EMC Control Plan is automatically imposed on a development program when compliance with certain interference specifications is required. For example, both MIL-STD-461 and MIL-E-6051 identify the need for such a plan and provide outlines of general plan content. The initial version of the plan is usually prepared by the equipment developer and distributed within 90 days after program initiation or contract award. However, periodic review and update is encouraged (typically on a six-month basis) to be sure that the document continually reflects the current objectives and needs of the Avionics/GSE development effort.

9.2.2 The Contents of the Control Plan

The basic outline of a representative EMC Control Plan is as follows:

1. Introduction
2. Applicable Documents
3. Specification Interpretation
4. Program Management Requirements
5. Design Requirements
6. Test Requirements
7. Quality Assurance Tests
8. Definition of Terms

A discussion of the possible contents of the various sections of this outline should provide a better indication of the role the plan can play in support of Avionics/GSE systems engineering.

9.2.2.1 Introduction

This section presents the scope and objectives of the control plan. It should introduce the system, to which the plan is applied, in enough detail to acquaint the reader with the functional characteristics of the equipment involved and the EM conditions it is required to function under. Plan review and up-dating procedures should be specified; as indicated previously, a review cycle of no more than six months is recommended.

9.2.2.2 Applicable Documents

All pertinent standards, specifications, drawings, plans and reports related to EMI control are listed in this section. The EMC documents developed for the system by the military, the contractor, the subcontractors and vendors should be included. The effective dates of all documents and any restrictions or limitations on their use should be stated.

9.2.2.3 Specification Interpretation

Many EMC specifications and standards are directed toward establishing limits on individual components or units and, by themselves, do not guarantee any specific degree of compatibility when the units are assembled as part of a larger system. A particular limit may be set too low, thereby causing devices to be very susceptible to their EME, or be established too high, resulting in an unnecessary expenditure of suppression dollars. For this reason, and because two specifications and standards may conflict in some areas, it is necessary to analyze and modify specification requirements so that they more realistically represent the limits of expected operating conditions.

It should be noted that the need to tailor specifications to an anticipated EME apply to small and large GSE alike. The broad scope of a specification such as MIL-STD-461 may require modification even in the case of a battery powered, hand-held test set, when that test set must operate in a flight-/weather-deck environment, or in direct contact with a Sidewinder Missile Launcher.

This section of the EMC Control Plan should outline the approach the developer will take to interpret designated specifications, establish their applicability, and propose modified or additional requirements. Many of the analysis techniques to do this evaluation are contained in this design guide.

9.2.2.4 Program Management Requirements

One of the most important functions of the EMC Control Plan is that of documenting the responsibility and authority of the individuals who will direct and implement the EMC Program of the equipment developer. The success of the EMC Program will rest upon its effective administration.

The management portion of the plan should establish the organization and responsibilities of the EMC group, stating its functions and objectives. In general, this group participates in periodic design reviews, acts in an advisory capacity on EMC matters, performs EMC testing and generates recommendations for the solution or further definition of EMC problems uncovered. Management of EMC extends to the standards and specifications imposed contractually on vendors and suppliers of equipments and components.

The management portion of the EMC Control Plan should also specify all EMC documents and reports to be submitted during the course of equipment development. Target dates and milestones for the EMC effort should be included. A typical EMC program milestone chart is shown in Figure 9-1. Minimally, the milestones should include:

a. Preparation and submission of the EMC Control Plan

b. Qualification testing of equipment to interference and susceptibility requirements

c. Preparation of the system EMC test plan

d. Preparation of the system EMC test procedures

e. Initiation of the EMC test effort

f. Initiation of the general acceptance test

g. Any other milestones or target dates that are considered pertinent.

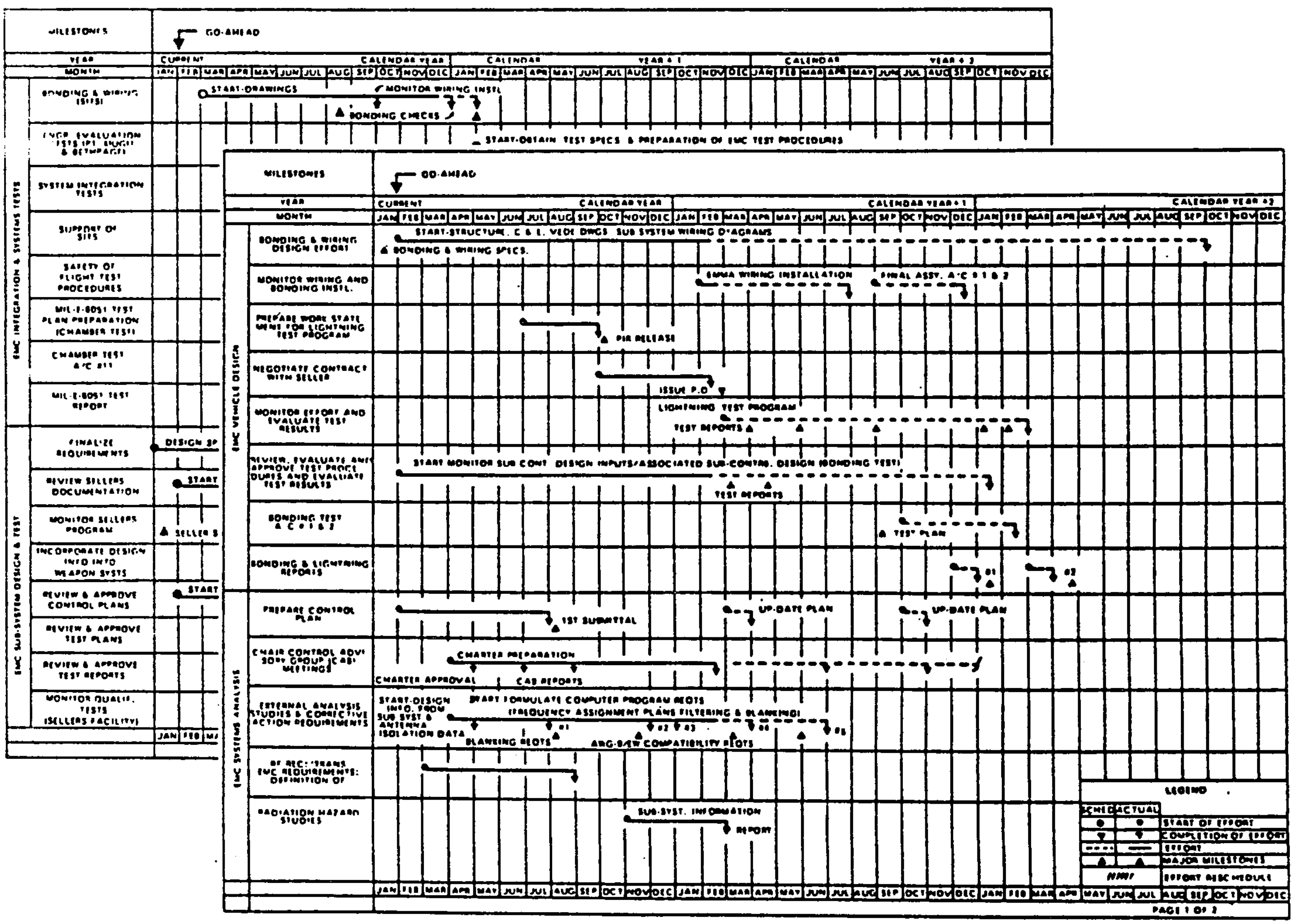


Figure 9-1 Typical EMC Program Milestones

Another key management action for Avionics/GSE of even moderate size and complexity should be the establishment of an EMC Advisory Board, Group, or Committee (EMCAB, EMCAG, EMCAC). This activity would act in an advisory capacity to the EMC program, to assist in resolving problems that may arise, expediting solutions, and establishing channels of coordination. NAVAIR AR-43, "Electromagnetic Compatibility Advisory Board, Requirements for," provides guidance for the formation and operation of an EMCAB and is written for use on EMC programs organized for NAVAIR contracts. EMCAB activities include EMC program review, discussion and evaluation of potential problem areas, definition of specific problems, determination of possible problem solutions, recommendation of best solutions to the responsible activity, and review of results of the implementation of the solutions.

Members of the board are typically representatives of the developer/ contractor, subcontractors, vendors, and the procuring activity. The charter for the board is usually included in the EMC Control Plan. Typically, the charter details membership, defines the normal interval between regular meetings and the meeting locations, and identifies the schedule for EMC design reviews.

9.2.2.5 Design Requirements

The means of assuring satisfactory EMI/EMC design control should be presented in the EMC Control Plan. This includes a statement of the design guidelines that are to be used, how the design guidelines are to be presented to the design organizations, the controls to assure that the design guidelines are being followed, and the means by which the design organizations are to be informed of the EMI/EMC philosophies of the EMC Control Plan. The guidelines should be written to apply to management as well as to the individual designers. Design organizations should be told of these guidelines in formal training sessions, through publications prepared and distributed by the responsible organization, and in conferences with representatives from all affected organizations.

As a basis for establishing Avionics/GSE design guidelines, it is necessary to identify degradation or performance threshold criteria that must be met at various locations within a system, and the consequence of failure to meet these criteria. One way in which the impact of interference may be categorized is as follows:

Category I: Subsystems or equipments where interference could result in loss of life or vehicle, aborted mission, costly delays, and unacceptable reduction in systems effectiveness.

Category II: Subsystems or equipments where interference could result in injury, damage to vehicle, or a reduction of systems effectiveness.

Category III: Subsystems or equipments where interference will cause only annoyance, minor discomfort, or loss of performance that does not decrease desired system performance.

For equipment in Categories I or II it may be advisable to develop additional degradation criteria beyond the limits specified in standards.

In addition, it may also be desirable to specify interference safety margins as a function of these categories, to assure that more critical operations receive a high level of protection (e.g., 6 dB margins for Categories II and III, 20 dB margin for Category I). This problem should be addressed in the control plan.

The purpose of this design requirements task is to assure that every essential EMI/EMC control measure is incorporated as early as possible into equipment design. Otherwise, costly time delays, major redesigns, or serious compromise of equipment operation will result.

In most cases, a subsystem from one organization will ultimately be joined with a subsystem from another organization. The design portion of EMC Control Plans submitted by the two different organizations must present EMC design policy in enough detail to assure the procuring activity that the design of one system will not be violated by the design of an associated system.

Some of the areas for which EMC design specifications must be prepared are:

a. Electrical bonding and grounding: Equipment-to-ground reference plane, electrical interface with other equipments or systems and subsystems, electrical power returns, and conductor shields.

b. Shielding: Equipment and subsystem case shielding, shielding provisions of the system structure, shielding and twisting of conductors, and anticipated EM and electrostatic environments.

c. Transient control: Suppression of transients from inductive sources, suppression of contactor transients, and surge limiting within the system power profile.

d. Radiated signal control: Spurious signals from intentional radiation sources, unintentional radiation sources, control of response to radiated signals, and projected intentional radiation environment created by the system.

e. Interference and susceptibility prediction: Frequency allocations, antenna locations, antenna patterns, and signal levels, when applicable.

f. Cable and/or conductor routine consideration: Conductor separation and isolation, and location of equipments and subsystems.

g. Electromagnetic radiation hazard (RADHAZ) and hazards of electromagnetic radiation to ordnance (HERO): Design policy and test procedures to prevent electromagnetic radiation hazard to personnel or harmful operation of weapon systems and electro-explosive devices.

h. Any special EMC considerations: Conditions peculiar to the type of Avionics/GSE involved, or to the nature of its intended mission may call for more stringent EMC control or variations to the cited specifications.

9.2.2.6 Test Requirements

The test requirements of the EMC Control Plan should provide for:

a. Preparation of an EMC Test Plan that defines equipment, subsystem, and system level testing to be performed, and that, in the case of a contracted program, is a requirement over which the procuring activity has approval authority. Section 11.2 of this design guide will discuss test plan preparation in some detail.

b. A statement of criteria used in determining which equipments and subsystems shall be tested.

c. Identification, by nomenclature and serial numbers if possible, of equipments and subsystems to be tested.

d. A statement of criteria used in determining the extent of EMC tests to be performed.

e. A design review to assure the equipments and subsystems have complied with the applicable design requirements of the EMC Control Plan.

In large Avionics/GSE development programs, many different organizations may be involved in testing equipments and subsystems to EMC specifications and standards. When this is the case, the EMC Control Plan should indicate the particular test to be performed by each of the organizations and specify how the results of these tests are to be coordinated from a systems point of view. Active liaison is especially important between testing organizations performing subsystem and equipment tests to MIL-STD-461 and MIL-STD-462 standards, and those performing tests to assure overall system performance.

This is true because the results of the subsystems and equipment level tests are used both to indicate points to be monitored and to establish tolerable limits for the systems level testing.

9.2.2.7 Other Information

Additional information is normally provided in the EMC Control Plan, including definitions of terms used; methods to be used in the education of design, management, and test personnel to EMC awareness; system quality assurance and production control; EMC maintenance procedures for operational equipment; or any studies that will be initiated for improvement of the EMC control program.

9.2.3 Control Plan Checklist

The checklist contained in Table 9-1 [2], in addition to the design guidelines sections of several chapters of this publication, provides good initial guidance on the features to incorporate in an EMC Control Plan. This must be supplemented with any requirements unique to the equipment or system under development.

TABLE 9-1 EMC CONTROL PLAN CHECKLIST

1. Introduction

1) Is the purpose of the document specified?

2) Is the document applicable to all subsystems and components?

3) Is the document applicable to all subcontractors and to all off-the-shelf purchased parts and components?

2. Management Section

1) Is there a positive statement concerning program management's intent to support fully an effective EMC control program?

2) Is there a sufficient organizational presentation to assure ready implementation of the EMC control program?

3) Have all critical organizational elements, including program management, EMC engineering, testing, quality assurance, materials, publications, and system design been identified in the organization?

4) Does the control plan specify all applicable EMC documentation to be provided?

5) Has the responsibility for preparation of critical EMC documentation including control plans, procedures, test plans, and other documentation been assigned?

6) Has responsibility for testing at subsystem and component levels as well as system levels been assigned?

7) Have appropriate distinctions been made between qualification, acceptance and compliance testing?

8) Is an individual or group of individuals designated to provide representation on behalf of EMC during critical design reviews and preliminary design reviews?

9) If the representation provided above is made by a team of individuals, is at least one of the individuals an EMI engineer?

TABLE 9-1 EMC CONTROL PLAN CHECKLIST

10) Has the Organization responsible for preparing, approving and submitting deviation requests and/or waivers against EMC requirements, limits, or specifications been identified?

11) Does the document provide for related status of EMC efforts in relation to critical milestones of the entire program?

12) Does the document show a program milestone for EMC?

13) Does the document provide for a duly constituted EMC board that will merge management and EMC engineering organizations?

14) Is it clear how design testing and management organizations interface together?

15) Does descriptive information on the EMC organization make clear how liaison among the elements is to be carried out?

16) Is there provision for EMC presentations to be made concurrent with preliminary design reviews and critical design reviews?

3. Applicable Documents

1) Have all applicable system documents for systems and subsystems been listed?

2) Have the latest or applicable version of EMC specifications been listed?

3) Have all applicable documents relating to reliability and quality assurance been listed?

4) Have all applicable specifications regarding materials and processes been indicated including those for metal treatment, finishes, chemical films, etc.?

5) Have all ancillary drawings and specifications regarding grounding planes, harnessing schemes, cable routing, etc., been listed?

6) Where more than one specification applies, is the order of precedence indicated?

7) Is it clear which way the control plan interfaces with applicable directives and all other specifications regarding EMC?

8) If separate individual specifications have been developed for system level testing and component and subsystem level testing, have these been listed and in the correct order of precedence?

9) Have all special provisions for deviations from specification limits, frequency ranges, test procedures, etc., been listed?

TABLE 9-1 EMC CONTROL PLAN CHECKLIST

4. Analysis and Design

1) Have key operational parameters of the overall system been adequately and concisely described?

2) Have the operating frequency and power output levels of all transmitters and susceptible devices been delineated?

3) Has a general presentation been provided on potential problem areas?

4) Have critical potential victim elements been defined?

5) Has an analysis been made of the power system utilization and potential susceptibilities via the power system, including grounds?

6) Have harmonics been considered when providing a table for predicting interference problems?

7) Have potential nonlinear elements in the system been identified which could generate spurious responses?

8) Have basic and specific guidance been provided toward the following areas?

   a) Selection of interference-free parts and components?

   b) Bonding and grounding.

   c) Wiring and cabling, including connectors, grouping, bundling, and routing.

   d) Shielding and filtering of sensitive circuits.

   e) Printed circuit coupling.

   f) Diode and transistor logic levels.

   g) Filters.

BIBLIOGRAPHY

[1] NAVAIR 5335, Naval Air Systems Command, Electromagnetic Compatibility Manual

[2] P.J. Mondin, "Control and Test Plans in EMC Programs," Seminar on Electromagnetic Compatibility, May 1973.

10.0 SPECIFICATIONS & STANDARDS

10.1 GENERAL

The previous chapters of this handbook have dealt with the objectives of EMC design, and possible methods for achieving the desired end result of EM compatibility. This chapter summarizes the appropriate government design and test specifications that can be employed to quantitatively control or influence GSE/Avionics EMC design and performance. These specifications identify performance requirements to be met, specific design practices to be employed, and other guidance pertinent to meeting desired EMC design goals.

The specifications selected for inclusion in this design guide have been chosen for appropriateness to Navy avionics/GSE. The list is not intended to be all-inclusive as regards the entire EMI/EMC discipline. Specifications have been omitted from the list relating to more specialized EMC problem areas or to types of situations not generally expected to be encountered during GSE/Avionics design. The basis for specification selection were as follows:

- o Direct relationship to EMC design of Navy Avionics and related GSE
- o General EMC design guidance for GSE/Avionics
- o Checklist for general design to prevent inadvertent oversight or omission of specific types of problems
- o General information regarding system engineering and performance requirements

Information is also provided in this chapter on how to obtain copies of required government specifications.

10.2 APPLICABLE SPECIFICATIONS

The specifications considered appropriate to EMC design of GSE/Avionics are contained in Table 10-1. The table lists the title and number of the specification, its scope of application and coverage, and comments regarding usage, content, and any unique features.[1]

The data in Table 10-1 reflect three general categories of specifications:

- o Criteria and tests for measuring system performance.
- o Sub-system test and design techniques and requirements.
- o Component and hardware selection, design and application.

1. Of these applicable specifications, MIL-STD-461 is given special attention. Because of the considerable usage of this standard, its entry in Table 10-1 includes the differences between versions A, B, and C of this document.

For those special cases or situations where the data of Table 10-1 are not sufficient, reference should be made to the Department of Defense Index of Specifications and Standards for all current related documents.

10.3 OBTAINING SPECIFICATIONS

The Department of Defense single stock point (DOD-SSP) for unclassified specifications and standards is:

Naval Publications and Forms Center
5801 Tabor Avenue
Philadelphia, Pennsylvania 19120

Phone: 215/697-3321

The DOD-SSP has the responsibility to index, issue, and control the current specifications and standards listed in the Department of Defense Index of Specifications and Standards (DOD-ISS).

Types of documents stocked and distributed are:

- Military Specifications
- Military Standards
- Federal Specifications
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- Qualified Products Lists
- Industry Documents
- Military Handbooks
- Air Force-Navy Aeronautical Standards
- Air Force-Navy Aeronautical Design Standards
- Air Force-Navy Aeronautical Specifications
- U.S. Air Force Specifications
- Other Departmental Documents
- Air Force-Navy Aeronautical Bulletins
- US Air Force Specification Bulletins

The Department of Defense Index of Specifications and Standards (DODISS) is prepared at NPFC and is published in three parts annually with bi-monthly supplements. Parts I and II of the DODISS and the Federal Supply Classification Listing are available to private industry on request from the Superintendent of Documents, Government Printing Office, Washington, DC 20402. Both include the basic index and the cumulative bi-monthly supplements for each part as they are published.

A sample of Form DD 1425, Specifications and Standards Requisition, to be used when requesting documents from NPFC, is shown in Figure 10-1.

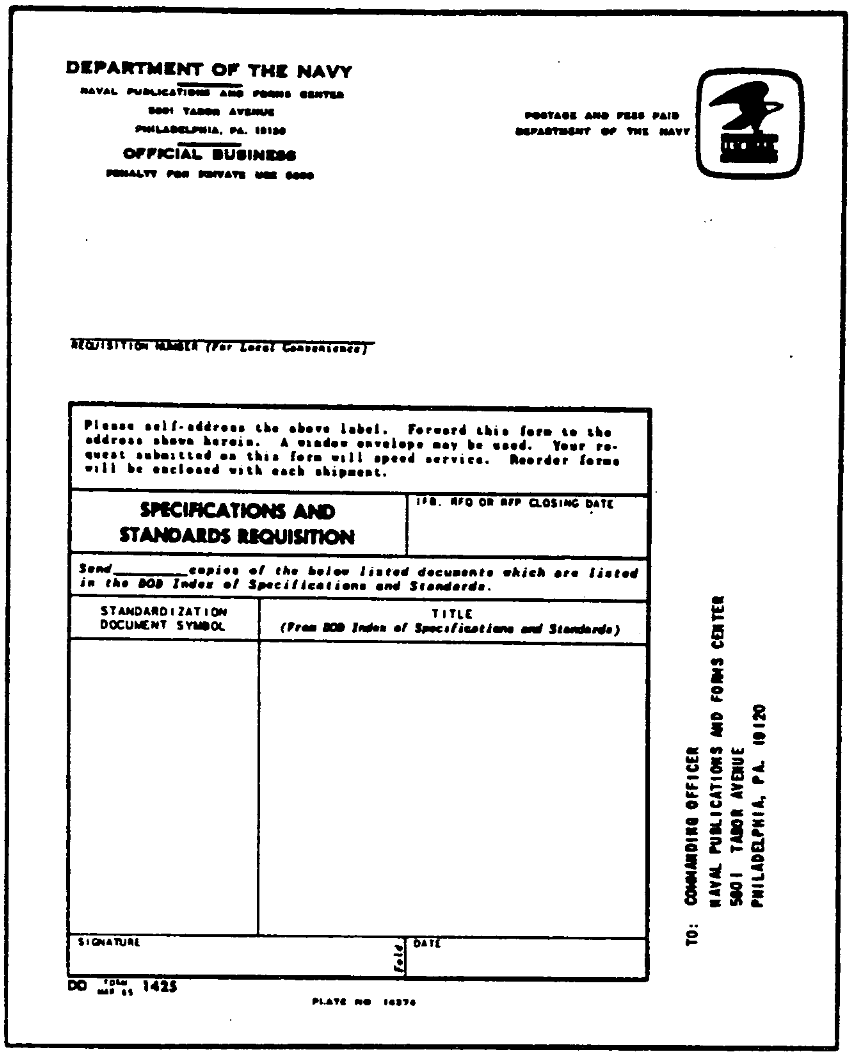

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Figure 10-1 Standard Order Form, DD 1425, Specifications and Standards Requisition

Table 10-1 Specifications for EMC Design

(1) **Specification:** MIL-B-5087B (2), INT AMD 3, 24 October 1984

Title: MILITARY SPECIFICATION BONDING, ELECTRICAL AND LIGHTNING PROTECTION, FOR AEROSPACE SYSTEMS. This specification has been approved by the Department of the Air Force and by the Bureau of Naval Weapons.

Scope: This specification covers the characteristics, application, and testing of electrical bonding for aerospace systems, as well as bonding for the installation and interconnection of electrical and electronic equipment therein, and lighting protection.

(2) **Specification:** MIL-B-5423/16C, 18 February 1987

Title: MILITARY SPECIFICATION BOOTS, DUST AND WATER SEAL (FOR TOGGLE AND PUSH BUTTON SWITCHES AND ROTARY - ACTUATED PARTS), GENERAL SPECIFICATION FOR. This specification has been approved by the Department of Defense and is mandatory for use by the Departments of the Army, the Navy and the Air Force.

Scope: This specification covers the general requirements for molded silicone-rubber RF-proof boots which can be used on toggle and push button switches and rotary-actuated parts such as rotary switches, variable resistors, 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.

(3) **Specification:** MIL-C-5541C, 14 April 1981

Title: MILITARY SPECIFICATION CHEMICAL CONVERSION COATINGS ON ALUMINUM AND ALUMINUM ALLOYS. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.

Scope: This specification covers the requirements for chemical conversion coatings formed by the reaction of chemical conversion materials and the surfaces of aluminum and aluminum alloys. Chemical coatings resulting from treatment of aluminum and aluminum alloy surfaces with chemical conversion coating materials shall be of the following classes, as specified:

- Class 1A - For maximum protection against corrosion, painted or unpainted.
- Class 3 - For protection against corrosion where low electrical resistance is required.

Comments: Aluminum Alloy 2024, Plate and Sheet. Aluminum Alloy 6061, Plate and Sheet. Anodic Coatings, For Aluminum and Aluminum Alloys.

(4) **Specification:** MIL-C-7078/35 Notice 1, 8 September 1987

Title: MILITARY SPECIFICATION CABLE, ELECTRIC, AEROSPACE VEHICLE, GENERAL SPECIFICATION FOR: This specification is mandatory for use by all Departments and Agencies of the Department of Defense.

Scope: This specification covers electric cable for use in aerospace vehicles and other applications for which its performance characteristics are suitable.

Table 10-1 Specifications for EMC Design (Cont'd)

Comments: The cable will be as described in the applicable military specification sheets and will be of the following types:
Unshielded-unjacketed: Two or more spirally laid coded wires with no overall jacket or shield.
Jacketed: Two or more spirally laid coded wires with no shield over the spirally laid wires but with an overall jacket.
Shielded: A single wire or two or more spirally laid coded wires with an overall shield but with no jacket over the shield.
Shielded and Jacketed: A single wire or two or more spirally laid coded wires with an overall shield and jacket over the shield.

(5) Specification: MIL-C-11693/8C, 25 July 1980
Title: CAPACITORS, FEED THROUGH, RADIO-INTERFERENCE REDUCTION AC AND DC (HERMETICALLY SEALED IN METAL CASES), ESTABLISHED AND NON-ESTABLISHED RELIABILITY GENERAL SPECIFICATION FOR
Scope: This specification covers the general requirements for established reliability (ER) and non-ER capacitors designed for operation with alternating current (ac) and direct current (dc), feed-through capacitors, hermetically sealed in metal cases for use primarily in broadband, radio interference suppression applications.
Comments: Capacitors meeting the established reliability requirements specified have a maximum failure rate of 1%/1000 hours. This failure rate is established with 90% confidence limit and maintained at a 10-percent producer's risk. An acceleration factor of 5:1 has been used to relate the life test data obtained at 140 percent of rated voltage at the applicable high test temperature. Styles which are of metallized construction 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.

(6) Specification: MIL-C-12889/2B, 3 June 1980
Title: CAPACITORS, BY-PASS, RADIO-INTERFERENCE REDUCTION, PAPER DIELECTRIC, AC AND DC, (HERMETICALLY SEALED IN METALLIC CASES), GENERAL SPECIFICATION FOR
Scope: This specification covers the 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 applications. 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.

Table 10-1 Specifications for EMC Design (Cont'd)

(7) Specification: MIL-C-13909C Notice 1, 20 May 1987
Title: MILITARY SPECIFICATION CONDUIT, METAL, FLEXIBLE; ELECTRICAL, SHIELDED. This specification has been approved by the Department of Defense and is mandatory for use by the Departments of the Army, the Navy, and the Air Force.
Scope: This specification covers shielded, electrical, flexible metal conduit. The conduit consists of a core of flexible metal tubing with a covering of wire braid.
Comments: Conduit shall be furnished in the following types, as specified: Type I Waterproof, Type II Non-waterproof.
Conduit shall be furnished in the following grades, as specified:
Grade A - Double wire braid. Grade B - Single wire braid.

(8) Specification: MIL-C-25200B Notice 1, 12 February 1986
Title: MILITARY SPECIFICATION CABLE ASSEMBLIES, SPECIAL WEAPONS, ELECTRICAL, GENERAL REQUIREMENTS FOR
Scope: This specification establishes the detailed requirements for the design, manufacture and testing of Air Force special weapons cable assemblies, including interconnecting cables.

(9) Specification: MIL-C-38999H Supp 1 21 September 1981
Title: CONNECTORS, ELECTRICAL, CIRCULAR, MINIATURE, HIGH DENSITY, QUICK DISCONNECT (BAYONET, THREADED, AND BREECH COUPLING), ENVIRONMENT RESISTANT, REMOVABLE CRIMP AND HERMETIC SOLDER CONTACTS, GENERAL SPECIFICATION FOR.

(10) Specification: MIL-C-81511/56A, 7 January 1977
Title: MILITARY SPECIFICATION CONNECTORS, ELECTRICAL, CIRCULAR, HIGH DENSITY, QUICK DISCONNECT, ENVIRONMENT RESISTING: AND ACCESSORIES GENERAL SPECIFICATION FOR. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.
Scope: This specification covers two series of electrical connectors (plugs and receptacles) with removable crimp and non-removable solder contacts and accessories. Electrical, mechanical, and environmental features of these connectors include: (1) Resistance from environment at sea level and high altitude. (2) Quick disconnect. (3) RFI/EMI protection. (4) High density insert arrangements. (5) Low level circuit capabilities. (6) Scoop proof. (7) Fluid resistant class should be provided. (8) High temperature class should be provided. (9) Availability of several voltage service ratings.

Comments: The connectors are identified as series 1 or series 2. Both series are designed to prevent inadvertent electrical contact and to provide contact protection during mating. They can be mounted in the same size panel cutout but are not intermateable. They have the following features: Series 1 - Long Shell. Scoop-proof when pins are installed in either the plug or the receptacle (formerly known as 100 percent no-scoop).

Table 10-1 Specifications for EMC Design (Cont'd)

Series 2 - Light-weight, short shell. Scoop-proof when pins are installed in the receptacle. Not scoop-proof when pins are installed in the plug (formerly known as 50 percent no-scoop). These connectors are rated to operate within a temperature range specified for each class. The upper temperature limit is the maximum internal hot-spot temperature resulting from any combination of electrical load and ambient temperature. These connectors are rated for continuous operation, 1000 hours at the specified maximum internal hot-spot temperature.

(11) Specification: MIL-C-85485A Supp 1 10 May 1983
Title: CABLE, ELECTRIC, FILTER LINE, RADIO FREQUENCY ABSORPTIVE
Scope: This specification covers the requirements for radio frequency absorptive component wires and finished cables which function electrically as distributed low-pass filters. Materials and construction details are specified in the applicable specification sheet.

(12) Specification: MIL-E-6051D (1), 5 July 1968
Title: MILITARY SPECIFICATION ELECTROMAGNETIC COMPATIBILITY REQUIREMENTS, SYSTEMS. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.
Scope: This specification outlines the overall requirements for systems electromagnetic compatibility, 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.

(13) Specification: MIL-E-16400H, 13 July 1987
Title: MILITARY SPECIFICATION ELECTRONIC EQUIPMENT, NAVAL SHIP AND SHORE: GENERAL SPECIFICATION. All interested bureaus of the Navy Department and the Marine Corps have concurred in the use of this specification.
Scope: This specification covers the general requirements applicable to the design and construction of electronic equipment and associated and auxiliary electronic apparatus furnished as part of a complete system intended for Naval ship or shore applications. The intent of this specification is to set forth the ambient conditions within which equipment must operate satisfactorily and reliably; the general material, the process for selection and application of parts, and to detail the means by which equipment as a whole will be tested to determine whether it will so operate. Throughout the design and manufacture of the equipment, maximum effort shall be made to attain the basic design objectives in that the equipment will meet the needs of the Naval service. Requirements applicable to individual equipment specification.
Comments:
- o Unitized modular construction,
- o Plug-in techniques,
- o Light weight Modular maintenance testing,
- o Commercially acceptable metals,
- o Planned-for maintenance.

Table 10-1 Specifications for EMC Design (Cont'd)

(14) Specification: MIL-E-45782B (1), 15 December 1980

Title: MILITARY SPECIFICATION ELECTRICAL WIRING, PROCEDURES FOR. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.

Scope: This specification covers the fabrication and termination of shielded electrical harness and cable assemblies and the wiring of electrical and electronic circuits of subassemblies used in missile systems.

Comments:
- o Shield Termination,
- o Securing of harnesses,
- o Use of ferrules.

(15) Specification: MIL-F-15733/75 (1), 18 January 1984

Title: MILITARY SPECIFICATION FILTERS, RADIO INTERFERENCE, GENERAL SPECIFICATION FOR. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.

Scope: This specification covers the general requirements for current-carrying filters, alternating-current (ac) and direct-current (dc), for use primarily in the reduction of broadband radio interference.

Comments: The type designation shall be in the following form, and as specified:
- o Style,
- o Current rating,
- o Insertion-loss characteristic,
- o Terminal identification,
- o Operating temperature range,
- o Grade.

(16) Specification: MIL-F-18327/77, 22 October 1984

Title: MILITARY SPECIFICATION FILTERS; HIGH PASS, LOW PASS, BAND PASS; BAND SUPPRESSION, AND DUAL FUNCTIONING, GENERAL SPECIFICATION FOR. This specification is mandatory for use by all Departments and Agencies of the Department of Defense.

Scope: This specification covers the general requirements for passive frequency-selective networks, such as dual functioning, band suppression, band pass, low pass and high pass (or any combination thereof) electric-wave filters, including those employing electromechanical and piezoelectric elements, for use over the frequency range of 0 to 150 megahertz. Filters covered by this specification are intended for use where operation under various environmental conditions is required. This specification covers filters weighing not more than 50 pounds and requiring root-mean-square test voltage ratings not greater than 5,000 volts.

Comments: The type designation shall be in the form appearing below, and as specified. The type designation does not describe a discrete item. For complete identification, the part number and type designation should be referenced.
- o Component,
- o Grade,

Table 10-1 Specifications for EMC Design (Cont'd)

- o Class,
- o Life Expectancy,
- o Family,
- o Case and Mounting Dimensions,
- o Composition.

(17) Specification: MIL-F-19207/39A (1), 22 July 1986

Title: MILITARY SPECIFICATION FUSEHOLDERS, EXTRACTOR POST TYPE, BLOWN FUSE INDICATING AND NONINDICATING GENERAL SPECIFICATION FOR

Scope: This specification covers enclosed, panel mounted, extractor post type electrical fuseholders, blown fuse nonindicating, EMI shielded.

Comments:
- o Gaskets and o-rings, conductive Knitted monel impregnated with silicone rubber, ZZ-R-765.
- o Shield cap, collar & associated hardware: Brass, w/lusterless black finish (collar & chain retainer rings etc.).
- o Rating: 30 amp., 250 volts max.
- o Nonindicating

(18) Specification: MIL-HDBK-235-1A, 5 February 1979, Superseding, MIL-HDBK-235-1 (NAVY), 23 June 1972.

Title: ELECTRO (RADIATED) ENVIRONMENT CONSIDERATIONS FOR DESIGN AND PROCUREMENT OF ELECTRICAL AND ELECTRONIC EQUIPMENT, SUBSYSTEMS AND SYSTEMS

Scope: The intent of this handbook is to provide guidance and establish a uniform approach for the protection of military electronics from the adverse effects of the electromagnetic environment. The handbook is applicalbe to any electrical and electronic equipment, subsystem or system which may be exposed to an electromagnetic environment during its life cycle, including the following:

(a) Aerospace and weapons systems and associated subsystems and equipments.

(b) Ordnance.

(c) Support and checkout equipment and instruments for (a) and above.

(19) Specification: MIL-HDBK-237A, Interim Notice 1 (NAVY), 6 June 1986.

Title: ELECTROMAGNETIC COMPATIBILITY MANAGEMENT GUIDE FOR PLATFORMS, SYSTEMS AND EQUIPMENT

Scope: This document is intended to provice managers responsible for the design, development and acquisition of DoD platforms, systems and equipments with the guidance necessary to establish an effective program for achieving the desired degree of EMC. The handbook describes the steps which must be taken to ensure that EMC considerations are incorporated during the life cycle of the platform, system, or equipment.

Table 10-1 Specifications for EMC Design (Cont'd)

(20) Specification: MIL-HDBK-241, 2 August 1974.
Title: DESIGN GUIDE FOR EMI REDUCTION IN POWER SUPPLIES
Scope: The intent of this handbook is to provide information relating to methods and techniques that an equipment engineer may use to reduce electromagnetic interference (EMI). This handbook is directed toward basic techniques for reducing EMI, particularly in power supplies. Use of the techniques set forth in this handbook should enable design engineers to develop comparisons between EMI electromagnetic compatibility (EMC), weight, size, cost, reliability, maintainability, temperature, humidy, human engineering and performance.

(21) Specification: MIL-HDBK-248 AS, 1 April 1977.
Title: TAILORING GUIDE FOR APPLICATION OF SPECIFICATIONS AND STANDARDS IN NAVAL WEAPONS SYSTEMS ACQUISITIONS
Scope: The guidelines contained herein are directed primarily to the application of military specifications and standards in major acquisition contracts for complex military weapons systems, such as aircraft and missile systems, and major subsystems, such as propulsion systems, avionics and weapons, requiring multiphased development and procurement.

(22) Specification: MIL-HDBK-253, 28 July 1978.
Title: GUIDANCE FOR THE DESIGN AND TEST OF SYSTEMS PROTECTED AGAINST THE EFFECTS OF ELECTROMAGNETIC ENERGY
Scope: The purpose of this document is to provide program managers with guidance for the design and test of electronic systems which are to be immume to the detrimental effects of electromagnetic energy. This handbook is applicable to any electronic system or equipment which may be exposed to electromagnetic energy during its life cycle, including the following:

(a) Aerospace and weapons systems and associated subsystems
(b) Ordnance
(c) Support and checkout equipments for a and b

(23) Specification: MIL-I-6181D Notice 7, 18 May 1987.
Title: MILITARY SPECIFICATION INTERFERENCE CONTROL REQUIREMENTS, AIRCRAFT EQUIPMENT. This specification has been approved by the Department of Defense and is mandatory for use by the Departments of the Army, the Navy, and the Air Force.
Scope: This specification covers design requirements, interference test procedures and limits for electrical and electronic aeronautical equipment to be installed in or closely associated with aircraft.

The test procedures which are specified cover the following types of tests:

(a) Interference test: Conducted and radiated tests which measure the magnitude of the interference signals emanating from the equipment under test.

(b) Susceptibility tests: Conducted and radiated, intermodulation and front-end rejection tests which determine whether an equipment will operate satisfactorily when exposed to external interference signals.

Table 10-1 Specifications for EMC Design (Cont'd)

Comments: Electrical and electronic equipment shall operate satisfactorily, not only independently but also in conjunction with other equipment which may be installed nearby. This requires that the operation of such equipment shall not be diversely affected by interference voltages and fields reaching it from external sources, and also requires that such equipment shall not, in itself, be a source of interference which might adversely affect the operation of other equipments. The limits specified herein are established to insure that the air vehicles will meet the requirements of Specification MIL-I-6051 or other applicable system specification.

(24) Specification: MIL-I-16165E, 14 October 1984
Title: MILITARY SPECIFICATION INTERFERENCE SHIELDING, ENGINE ELECTRICAL SYSTEMS.
Scope: This specification covers requirements for interference shielding items and shielded harnesses for engine electrical systems aboard Naval ships, at advance bases, and in the vicinity of electronic installations. It includes the allowable interference limits for such items and the permissible limits for auxiliary devices normally installed on electrical wiring systems associated with these engines.
Comments: The interference shielding electrical systems furnished under this specification shall be a product which has been tested and passed the qualification tests specified herein and has been listed or approved for listing on the applicable qualified products list. Reliability of operation shall be stressed throughout the design and manufacture of the shielding.

(25) Specification: MIL-P-24014 Notice 1, 10 November 1972
Title: MILITARY SPECIFICATION PRECLUSION OF HAZARDS FROM ELECTROMAGNETIC RADIATION TO ORDNANCE, GENERAL REQUIREMENTS FOR
Scope: This specification establishes general requirements for weapon systems to preclude hazards from environmental electromagnetic fields in the frequency range of 1 kHz to 40 GHz. These requirements apply to all Navy weapon systems, including safety and emergency devices and other ancillary equipment which contain electrically initiated, explosive or pyrotechnic components.

(26) Specification: MIL-STD-202F, 11 April 1986
Title: MILITARY STANDARD TEST METHODS FOR ELECTRONICS AND ELECTRICAL COMPONENT PARTS.
Scope: The standard specifies test conditions obtainable in the laboratory which give results equivalent to actual service conditions existing in the fields. All of the test methods of a similar character which appeared in the various joint or single-service electronic and electrical component-parts

specifications; those newly-developed test methods which are feasible for use in several specifications; and the recognized extreme environments, particularly temperatures, barometric pressures, etc., at which component parts will be tested under some of the presently-standardized testing procedures have been consolidated.

Comments: This standard establishes uniform methods for testing electronic and electrical component parts, including basic environmental tests to determine resistance to natural elements and conditions surrounding military operations, and physical and electrical tests. This standard is intended to apply only to small parts weighing up to 300 pounds or having a root-mean-square test voltage up to 50,000 volts unless otherwise specifically invoked

(27) Specification: MIL-STD-220A Notice 1, 8 March 1978
Title: MILITARY STANDARD METHOD OF INSERTION-LOSS MEASUREMENT
Scope: This standard covers a method of measuring, in a 50-ohm system, the insertion loss of feed-through suppression capacitors, and of single- and multiple-circuit, radio-frequency (RF) filters at frequencies up to 1000 megahertz (MHz).

(28) Specification: MIL-STD-285, 25 June 1956
Title: ATTENUATION MEASUREMENTS FOR ENCLOSURES, ELECTROMAGNETIC SHIELDING, FOR ELECTRONIC TEST PURPOSES, METHOD OF
Scope: This standard covers a method of measuring the attenuation characteristics of electromagnetic shielding and enclosures used for electronic test purposes over the frequency range 100 kilohertz to 10,000 megahertz.
Comments: Attenuation is the ratio, expressed in decibels (db), of the received powers on opposite sides of a shield when the shield is illuminated by electromagnetic radiation; and as used in this standard, is the figure of merit to designate the shielding effectiveness of electromagnetic enclosures.

(29) Specification: MIL-STD-449D Notice 1, 18 May 1976
Title: MILITARY STANDARD RADIO FREQUENCY SPECTRUM CHARACTERISTICS, MEASUREMENT OF
Scope: This technical standard establishes uniform measurement techniques that are applicable to the determination of the spectral characteristics of radio-frequency transmitters and receivers. The ultimate goal is to ensure the compatibility of present and future systems.
Comments: The data obtained from the measurements described in this standard will comprise one of the principal aids for (a) predicting the performance of equipment and systems in an operational electromagnetic environment, and (b) predicting the effect of a particular equipment or system on the electromagnetic environment of other equipments or systems. This radio-frequency-measurement standard will apply to all equipments and systems that are designed to emit or will respond to electromagnetic energy in the radio-frequency range of from 0.014 to 12,000 megahertz (MHz). This standard does not necessarily apply to all parameters of all systems.

Table 10-1 Specifications for EMC Design (Cont'd)

(30) Specification: MIL-STD-454K Notice 2, 26 February 1987
Title: STANDARD GENERAL REQUIREMENTS FOR ELECTRONIC EQUIPMENT
Scope: Requirements applicable to electronic equipment. This standard covers some of the common requirements to be used in military specifications for electronics equipment.
Comments: The requirements contained herein are intended to provide uniform requirements applicable to electronic equipment and shall be incorporated by reference in general equipment specifications.

(31) Specification: MIL-STD-461C Notice 1, 1 April 1987
Title: ELECTROMAGNETIC INTERFERENCE CHARACTERISTICS REQUIREMENTS FOR EQUIPMENT
Scope: This standard covers the requirements and test limits for the measurement and determination of the electromagnetic interference characteristics (emission and susceptibility) of electronic, electrical and electromechanical equipment. The requirements shall be applied for general or multi-service procurements and single service procurements, as specified in the individual equipment specification, or the contract or order.
The requirements specified in this standard are established to:
(a) Insure that interference control is considered and incorporated into the design of equipment.
(b) Enable compatible operation of the equipment in a complex electromagnetic environment.
Comments: This standard shall be used in conjunction with MIL-STD-463 and MIL-STD-462.

Differences Between 461A, 461B and 461C

A. Diferences between 461A and 461B

The differences between MIL-STD-461A and 461B are mainly additions and deletions of requirements in MIL-STD-461A; these changes are as follows:

1) Additions

CE07: A time domain transient limit applicable to the power lines for Navy Air Force procurements. It limits conducted switching transients to + 50% of nominal rms voltage on ac leads, and + 50%, - 150% of nominal line voltage on dc leads. No time duration is given.

CS09: A requirement to inject noise currents between the equipment case and structure. It applies to Navy procurements only and is for equipment and subsystems that have an operating frequency range of 100 kHz or less, and an operating sensitivity of 1 microvolt or less. The injected current starts at 1 ampere at 60 hertz and decays in a varied fashion until it reaches 100 ma at 100 kHz.

Table 10-1 Specifications for EMC Design (Cont'd)

UM03: Limits E-field broadband emissions for tactical, special purpose vehicles, and engine driven equipments, including the electrical equipments and parts installed thereon.

UM04: Limits broadband EM conducted and radiated emissions. Also restricts exhibition of malfunction, degradation of performance, or deviation from tolerances of equipment specifications by engine generators subjected to E-field radiations in frequency range of 2 MHz to 10 GHz. UM04 applies to engine generators and associated components, UPS and MEP equipment supplying power to, or used in, critical areas.

UM05: Limits broadband electromagnetic conducted and radiated emissions for Group I commercial electrical and electromechanical equipments procured for use in critical areas.

2) Deletions:

There are some significant deletions from MIL-STD-461A and earlier versions of MIL-STD-461B in the latest issue. The inverse filter method of measuring conducted emissions (CE05) has been deleted, as has the radiated requirement for overhead power lines (RE06). RE05 covering vehicles and engine driven equipment has been replaced by UM03 whereas CE02 and CE04 have been combined with CE01 and CE03. The requirement to measure magnetic fields over the 0.02 to 50 kHz frequency range has been dropped, as has the one signal generator method of measuring a receiver's rejection of undesired signals from 30 Hz to 10 GHz (CS08).

B. Differences between 461B and 461C

The major differences between MIL-STD-461C and its predecessor are the addition of test methods RS05, CS10, and CS11. These tests have been added to parts 2, 4, 5 and 6 of the requirements. The test procedures are as follows:

CS10: Determines susceptibility to damped sinusoidal transients injected into various pins and terminals of the equipment or subsystems under test.

CS11: Determines equipment susceptibility to damped sinusoidal transients contained on control, signal, and power cables.

RS05: Determines susceptibility of equipment to a high voltage transient pulse field.

Table 10-1 Specifications for EMC Design (Cont'd)

In addition to new test methods, MIL-STD-461C incorporates changes to the previous MIL-STD-461B issue. A major difference between the two versions is the testing with addition of categories A1h and A2d. These categories are to be used for the testing of class A3 equipment (equipment and subsystems installed in ground facilities - fixed and mobile, including tracked and wheeled vehicles) procured for Air Force use.

Other major changes can be found in the test method RS03. With the increased use of synthetic composite materials in areas where only metal was previously found, a new category has been added to Part 5 for testing in areas of non-metallic hulls. In addition, RS03 allows for the use of antennas which will generate circularly polarized fields instead of the previous "linear only" requirement.

(32) Specification: MIL-STD-462 Notice 6, 15 October 1987
Title: ELECTROMAGNETIC INTERFERENCE CHARACTERISTICS, MEASUREMENT OF
Scope: This standard establishes techniques to be used for the measurement and determination of the electromagnetic interference characteristics (emission and susceptibility) of electrical, electronic, and electromechanical equipment, as required by MIL-STD-461.
Comments: Notice 3 of this specification was released 9 February 1971; it is applicable to Army procurements and established techniques to be used for the measurement and determination of the electromagnetic interference characteristics (emission and susceptibility) of electrical, electronic and electromechanical equipments, sub-systems and systems as required by MIL-STD-461A, Notice 4.

(33) Specification: MIL-STD-463A, 1 June 1977
Title: DEFINITIONS AND SYSTEM OF UNITS, ELECTROMAGNETIC INTERFERENCE TECHNOLOGY
Scope: This standard contains general interference definitions, abbreviations, and acronyms used in MIL-STD-461 and MIL-STD-462. Definitions of abbreviations and terms are limited to statements of meaning as related to this and referenced standards, rather than encyclopedia or textbook discussions.

(34) Specification: MIL-STD-469 Notice 1, 30 March 1967
Title: RADAR ENGINEERING DESIGN REQUIREMENTS, ELECTROMAGNETIC COMPATIBILITY
Scope: The engineering design requirements are established to control the spectral characteristics of all new radar systems operating between 100 and 40,000 megahertz (MHz) in an effort to achieve electromagnetic compatibility and to conserve the frequency spectrum available to Military radar systems.

Table 10-1 Specifications for EMC Design (Con'd)

(35) Specification: MIL-STD-704D, 30 September 1980
Title: ELECTRIC POWER, AIRCRAFT, CHARACTERISTICS AND UTILIZATION OF
Scope: This standard delineates the characteristics of electric power supplied to airborne equipment at the equipment terminals and the requirements for the utilization of such electric power by the airborne equipment.
Comments: The purpose of this standard is to foster compatibility between aircraft electric systems or ground support electric systems and airborne utilization equipment to the extent of confining the aircraft and ground support electric power characteristics within definitive limits and restricting the requirements imposed on the electric power by the airborne utilization equipment.

(36) Specification: MIL-STD-1310E, 18 August 1987
Title: SHIPBOARD BONDING, GROUNDING AND OTHER TECHNIQUES FOR ELECTROMAGNETIC COMPATIBILITY AND SAFETY
Scope: This standard provides shipboard bonding, grounding and other techniques for electromagnetic compatibility and safety.
Comments: The requirements of this standard apply to all new shipboard installations and to that part of existing installations receiving modifications. It is not the intent of this standard to retrofit existing installations not programmed for modification or to retrofit work accomplished according to previous requirements. The procedures and methods specified in this standard shall be utilized where ever it is required to bond, ground, insulate or use nonmetallic materials for the following applications, as specified:
(a) To provide electromagnetic compatibility.
(b) To provide personnel safety from electrical shock hazards.
(c) To safeguard electrical transmission of classified information.
(d) To provide a dc reference ground.

(37) Specification: MIL-STD-1377, 20 August 1971
Title: MILITARY STANDARD EFFECTIVENESS OF CABLE, CONNECTOR, AND WEAPON ENCLOSURE SHIELDING AND FILTERS IN PRECLUDING HAZARDS OF ELECTROMAGNETIC RADIATION TO ORDNANCE; MEASUREMENT OF
Scope: This standard is intended to provide a weapon developer or designer with shielding and filter effectiveness test methods for determining 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 test.
Comments: This standard covers the methods of shielding effectiveness of weapon enclosures, cables, and cable connectors over the frequency ranges of from 100 kilohertz (kHz) to 30 megahertz (MHz) and from 1000 MHz to 10 gigahertz (GHz). The shielding effectiveness test methods are conducted at two separate frequency ranges because of hardware limitations. This standard

Table 10-1 Specifications for EMC Design (Cont'd)

also covers the methods of measuring the filtering effectiveness of radio frequency (RF) suppression devices (filters) for weapon firing circuits over the frequency range of from 100 kHz to 10 GHz.

(38) Specification: MIL-STD-1385B, 1 August 1986
Title: MILITARY STANDARD PRECLUSION OF ORDNANCE HAZARDS IN ELECTROMAGNETIC FIELDS; GENERAL REQUIREMENTS FOR
Scope: This standard establishes the general requirements to preclude hazards resulting from ordnance having electro-explosive devices when exposed to electromagnetic fields. The nominal frequency range covered by this standard is from 10 kilohertz (10^4 Hertz) by 40 gigahertz (4 x 10^{10} Hertz
Comments: These requirements apply to all Naval Weapon Systems, safety and emergency devices and other ancillary equipment containing electrically initiated explosives, propellant or pyrotechnic components. The trend in radar and communications equipment towards greater radiated power has resulted in growing concern with electromagnetic radiation hazards to ordnance. These hazards stem from the use of electro-explosive devices which can be initiated spuriously by means of electromagnetic energy. In addition to the hazards that could occur, reliability factors are also considered.

(39) Specification: MIL-STD-2757A, 20 July 1983, Superseding, MIL-STD-1757, 17 June 1980.
Title: LIGHTNING QUALIFICATION TEST TECHNIQUES FOR AEROSPACE VEHICLES AND HARDWARE
Scope: This document presents a set of standard test waveforms and techniques for lightning qualification testing of aerospace vehicles and hardware. The test waveforms presented in this document are intended to reproduce the significant effects of the natural environment and are therefore independent of vehicle type or configuration. The tests include high voltage and high current physical damage tests of fuel, structural and electrical hardware, as well as indirect effects associated with lightning strikes to externally mounted electrical hardware.

(40) Specification: MIL-T-28800D, 30 September 1986
Title: TEST EQUIPMENT FOR USE WITH ELECTRICAL AND ELECTRONIC EQUIPMENT, GENERAL SPECIFICATION FOR.

(41) Specification: MIL-W-6858D Notice 1, 30 July 1987
Title: MILITARY SPECIFICATION WELDING, RESISTANCE: ALUMINUM, MAGNESIUM, NONHARDENING STEELS OR ALLOYS, NICKEL ALLOYS, HEAT-RESISTING ALLOYS, AND TITANIUM ALLOYS; SPOT AND SEAM. This specification is mandatory for use by all Departments of Defense.
Scope: This specification covers 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.

Table 10-1 Specification for EMC Design (Cont'd)

Comments: Classification shall be based on function of the spot and seam welded joint, as follows:
Class A - Used in joints, the single failure of which during any operating condition would cause loss of the weapons system or one of its major components, loss of control, unintentional release of, or inability to release any armament store, failure of gun installation components, or which may cause significant injury to occupants of the manned weapons system.
Class B - Used in joints, the failure of which would reduce the overall strength of the weapons system or preclude the intended functioning or use of equipment.
Class C - A weld which is considered non-critical and for which no stress analysis is required.

11.0 TEST REQUIREMENTS, PLANS AND TECHNIQUES

11.1 INTRODUCTION

Specialized equipment testing is necessary to insure that GSE/Avionics will ultimately operate in their anticipated EM environments without performance degradation, and that they will not degrade other equipment in these environments. EMC testing of GSE/Avionics is necessary at both the "component" or sub-system level, as well as at the equipment and system level, and specifically in the following areas:

- Design testing of breadboard or engineering models, to obtain component and circuit engineering performance estimates that will assist in meeting intra-equipment and intra-system needs.
- Performance testing of equipment or system prototypes to evaluate operation in intended environments, or to evaluate compliance to applicable specifications.
- Acceptance testing of production units, to formally demonstrate or provide proof of operation as specified or required on the first article produced.

Planning to support these test areas should be initiated early in GSE/Avionics design and development stages, so that sufficient consideration can be given to the EMI implications of the particular design employed, and meaningful tests can be defined and implemented. Such tests should not only provide input data from which basic EMC design decisions can be made, but should be able to offer essentially continuous information that can lead to design updates, reduce redundant EMC design, and avoid EMC retrofits.

Previous chapters of this design guide have already provided considerable inputs to the EMI testing area. Test techniques to evaluate shielding effectiveness (Section 4.6), bonding effectiveness (Section 5.5), and filtering effectiveness (Section 7.6) have been discussed. The purpose of this chapter is to provide additional technical information the GSE/Avionics designer may need in planning and conducting EMI tests. This includes:

- The EMC Test Plan
- Shielded enclosure requirements
- Testing guidelines
- Radiated and conducted equipment testing
- MIL-STD-1377 Testing

11.2 THE EMC TEST PLAN

The purpose of an EMC test plan is to provide the details necessary to implement the collection of equipment or systems performance characteristics pertinent to EMI evaluation. It should precede any major testing effort as a matter of good engineering practice, even if it does not represent a requirement imposed by the Navy.

Very often, the requirement to prepare the EMC test plan is levied by a specification or standard. For example, MIL-STD-461 requires such a plan be submitted and approved prior to formal compliance testing. A 90-day review of a plan by the cognizant organization is typical.

A suitable GSE/Avionics test plan should contain but not be limited to the following:

- Justification for running each test, including how the test will support any GSE/Avionics performance evaluation, and the application of the test results in identifying and resolving real or potential problem areas.

- A description of each test to be performed, including the test instrumentation to be used, the test configurations to be employed, the status of all controls and switches that can affect the test, the test modulations to be employed, the specific locations (by terminal or pin identification) of test points, the performance criteria to be used, and the output format for the results of each test.

- Detailed step-by-step test procedures.

- A description of how the required characteristics of any simulated signal will be established, and how these signals will be obtained and synchronized.

- A detailed description (including characteristics, operating procedures, calibration procedures, and circuit diagrams) of any specialized or unique test equipment or test component not normally employed for the particular test in question.

- A Description of how the uncontrolled ambient electromagnetic environment will be measured, and the action to be taken to prevent the uncontrolled ambient level from introducing errors into the test measurements.

- A description of procedures to be employed to assure proper operation of the systems or equipments under test prior to and during the testing.

- A description of test instrumentation calibration procedures to be employed.

An outline of a typical test plan is shown in Table 11-1.

When preparing the test plan, care must be taken to be sure that conditions that result in maximum interference impact are tested. For example, emitter/repeater pairs should be studied to be sure that possible tuning arrangements in which fundamental, harmonic, sub-harmonic, or spurious frequency overlaps can occur are investigated. In a similar vein, extensive probing should be specified for radiated tests to be sure that the most serious case leakage paths are located.

The requirement for testing in a low ambient environment usually dictates use of a shielded enclosure (See next section). The test plan must be particularly detailed regarding precautions to be taken when using an enclosure, to be sure that the test results obtained are meaningful. The plan should define the physical and electrical requirements of the enclosure and ancillary equipment within the unit (such as its inside dimensions, the size and type of ground plane, the method to use to ground or bond the ground plane to the enclosure and to the equipment under test, the room attenuation characteristics,etc.). It should also identify the locations of the test equipment and the GSE within the enclosure, and the interconnecting lead dress, to minimize reflection and undesired coupling effects.

The test procedures themselves should be sufficiently detailed to enable repeating the same test at a later date if desired. The procedures should include the output criteria to be employed or monitored, the specific test points at which signals are to be injected or removed, the matching techniques to be used at these points (including detailed descriptions of couplers, matching networks, etc.), test modulations, position of all adjustable controls, test setup block diagrams, and any other factors that have to be defined before a test can be unambiguously performed.

A tabulation of planned test equipment should accompany the test plan. This tabulation should at least include general calibration and accuracy data (See Table 11-2), but should be expanded considerably for unique or non-standard devices. In many cases (such as when testing in accordance with MIL-STD-461), approval must be obtained on instrumentation before it can be used.

It must be appreciated that an EMC Test Plan is an evolving document. The initial draft of the plan can be expected to emphasize the types of tests that should be performed and the general philosophy to be implemented in the testing program. As the GSE design proceeds, the plan should be updated and expanded so that, when detailed test procedures are needed, they will be available. The checklist of Table 11-3 offers a good start in identifying items to be considered in test plan preparation.

TABLE 11-1 OUTLINE OF TYPICAL TEST PROCEDURE

<u>Contents</u>
Includes major paragraphs and appendices

<u>Administrative Data</u>
Include the following headings:
Purpose of Test
Applicable Documents - Specifications
Standards, Drawings
Test Sample Identification
Manufacturer
Security Classification
Test to be Performed by
Test Location
Date of Test
Disposition of Test Sample

<u>Scope</u>
Define the scope of the EMC tests to be performed, and identify the applicable EMC specification

<u>Description of Test Sample</u>
Functional and Physical Description
Connectors Associated with the Test Sample
External Controls

<u>EMI Test To Be Performed</u>
Interference Tests
Susceptibility Tests
Tests Not Applicable
Exceptions to EMI Control Specification

<u>Anticipated Interference Test Equipment</u>
Test Equipment Required
List all ground support equipment (GSE), EMC test equipment, hardware, special test interfaces, include part no., S/N and calibration dates
Test Equipment Calibration
Line Impedance Stabilization Networks (LISN) or Feed-Through Filters
Test Equipment Bonding

<u>Test Conditions</u>
Test Location
Test Arrangement
Test Sample Bonding
External Loads
Describe electrical and/or mechanical loads to be used
Lead Lengths and Position with respect to Ground Plane
1. Primary Power Leads
2. Interconnecting Leads
3. Load Leads
Shielded Leads

TABLE 11-1 OUTLINE OF TYPICAL TEST PROCEDURE (Cont'd)

Coaxial Cables
Test Sample Operation
1. Modes of Operation
2. Steady State Interference Test Mode
3. Transient Interference Test Mode
4. Susceptibility Test Mode
5. Functional Test
6. Control Settings

Accuracy of Test Equipment
Define recalibration conditions, etc.

Detailed Test Procedure
Operation of Interference Measuring Instruments
(State frequency range, detector function and calibration)
Selection of Test Frequencies
Ambient Interference Measurements
Test Limits
Describe design parameters and specify acceptable variations
Future Criteria
List conditions that constitute failure due to susceptibility, i.e., where changes from operational test values beyond allowable limits or degradation of performance occurs
Specific EMI test procedures for each EMI test to be performed
Test #1, Conducted emissions from GSE power leads (30 Hz to 20 KHz)
Test #2, Conducted emissions form GSE output leads(30 Hz to 20 KHz)
(Continue list as required)

Test Report
Acknowledge that a test report will be prepared

Test Personnel
Acknowledge that personnel performing test will be identified in test report

Appendix I
Includes test arrangement diagram, electrical loads, test equipment list

Appendix II
Sample data sheets and graphs

TABLE 11-2 SAMPLE TEST EQUIPMENT LIST

SPECIFICATION: MIL-STD-461 TEST METHOD: RE02

Frequency Range	Item	Manufacturer	Model	S/M	Cal. Period	Date of Last Cal.	Accuracy
.014-1000 MHz	RIFI Meter	Singer Metrics	NF-105	1824	6 months	12-15-87	freq ± 3% Amp ± 2dB
.014-.14 MHz	Tuning Unit	Singer Metrics	TX/WF-105	3105	6 months	1-12-88	freq ± 3% Amp ± 2dB
.15-30 MHz	Tuning Unit	Singer Metrics	TA/WF-106	3057	6 months	12-20-87	freq ± 3% Amp ± 2dB
20-200 MHz	Tuning Unit	Singer Metrics	T1/WF-105	1817	6 months	10-13-87	freq ± 3% Amp ± 2dB
200-400 MHz	Tuning Unit	Singer Metrics	T2/WF-105	1817	6 months	1-15-88	freq ± 3% Amp ± 2dB
400-1000 MHz	Tuning Unit	Singer Metrics	T3/NF-105	1817	6 months	2-23-88	freq ± 3% Amp ± 2dB
.014-.15 MHz	41" Rod Antenna	Singer Metrics	VR1-105	L-428	12 months	10-12-87	± 2dB
.15-30 MHz	41" Rod Antenna	Singer Metrics	VA-105	1817	12 months	10-12-87	± 2dB
20-200 MHz	Bi Conical Antenna	NSL					
200-1000 MHz	Log Sprial Antenna	EMCO					

TABLE 11-3 CHECKLIST FOR PREPARATION OF TEST PLAN

1. Introduction

1) Is the purpose of the test indicated?

2) Are applicable drawings, documents, specifications, and standards listed? Is the listing complete?

3) Is the test sample or system completely identified?

4) Is it clear who would perform the test and at what location?

2. Description of Test Sample

1) Is the test sample or system completely identified?

2) Is it clear whether the system under test is a production sample, prototype, or engineering model?

3) Are external controls, ground support equipment, or ancillary systems required for testing identified?

4) Is the operational configuration of the system or sample during test clearly specified?

3. Test Conditions

1) Is the test facility and test location completely identified and described?

2) Are the facilities for test including shielded enclosures, power line filtering, power line load distribution completely specified?

3) Are the ambient conditions clearly identified?

4) Is the arrangement of the test system during test clearly specified?

5) Do the test setup drawings and specifications completely identify the power distribution to the system under test, location and orientation of exercising equipment and placement location of breakout boxes monitoring circuitry and equipment locations?

6) Are external loads clearly labeled?

7) Is the test sample bonding in accordance with MIL-STD-461 and 462?

8) Are test and monitoring points appropriately listed?

9) Are the system operating modes clearly delineated?

10) Are separate tests to be performed in terms of steady state interference and susceptibility and transient performance?

TABLE 11-3 CHECKLIST FOR PREPARATION OF TEST PLAN (Cont'd)

11) If so are these completely delineated?

12) Are susceptibility test modes responsive to stimuli requirements of the applicable specifications?

13) Are all possible susceptible parameters listed and specified?

14) Are the initial system or sample operating conditions and control settings listed for susceptibility and interference testing?

15) Are critical cables between support equipment and bench test equipment shielded in order to eliminate interaction with the test setup?

16) Has care been taken to ensure that the test monitoring equipment themselves are not susceptible or interfering sources?

4. <u>EMC Test to be Performed</u>

1) Are all tests in consonance with applicable specifications?

2) Are deviations and special exceptions clearly documented?

3) Is the listing of each test to be performed complete?

4) Are all legitimate operating modes of the systems considered for these tests?

5) Is the test listing compatible with performance requirements

6) Are there any tests that are not required by reason of prior qualification? If so, has the prior qualification been evaluated with regard to the applicable specifications?

7) Are exceptions being made to the EMC control plan?

8) Are the EMC tests to be performed in complete agreement with the EMC control plan?

9) Are there any deviations from the EMC control plan test requirements?

10) Are test limits thoroughly described for interference?

11) Have special tests been identified in relation to unique system requirements?

12) Have susceptibility threshold limits been carefully interpreted?

5. <u>Test Equipment</u>

1) Has all of the required EMC test instrumentation been listed?

TABLE 11-3 CHECKLIST FOR PREPARATION OF TEST PLAN (Cont'd)

2) Are the support equipments for monitoring and exercising the system or sample been made complete?

3) Have special test interfaces of breakout boxes been identified?

4) Is the equipment list complete?

5) Does the equipment list include all applicable calibration dates?

6) Does the calibration conform to MIL-C-45662?

7) Have all special-purpose devices been identified and described, including feed-through filters, LISN's and voltage probes?

8) Does the list of equipment include all applicable operational parameters including frequency range sensitivity, etc., as applicable to the parameters to be investigated?

9) Is the accuracy specified?

6. Detailed Test Procedures

1) Do the test procedures conform to MIL-STD-461, MIL-STD-462, MIL-E-6051D, or specific specification?

2) Have test frequencies been selected on the basis of real system operational requirements?

3) Have deviations necessitated by system operational parameters of from commonly used specifications been clearly delineated?

4) Have the test limits been accurately identified?

5) Have the failure criteria been interpreted into meaningful test procedures?

6) Is there a specific EMI test procedure and detail including the operation of each monitoring and EMC test instrument for each test procedure?

7. General Requirements

1) Does the test plan acknowledge that the test report will be prepared?

2) Is there a sample test data sheet, including provisions for all critical system measurement equipment parameters, antenna correction factors, cable losses, etc.?

3) Are the test personnel to be performing the tests identified?

TABLE 11-3 CHECKLIST FOR PREPARATION OF TEST PLAN (cont'd)

4) Are all important test arrangement diagrams, electrical loads, test equipment lists, etc., included?

5) Does the documentation methods proposed conform to MIL-STD-831 or the applicable documentation standard?

11.3 SHIELDED ENCLOSURE REQUIREMENTS

11.3.1 Enclosure Limitations

A shielded enclosure for equipment test purposes is nothing more than a conductive "box" for enclosing all or part of a test setup. The practical requirements of power access, personnel access, light, heat, and air conditioning, result in discontinuities to this enclosure that must be considered a part of enclosure design.

The main advantage of the shielded enclosure for interference measurements is the RF isolation it provides to and from the outside world. Its use allows meaningful emission and susceptibility measurements to be made, both conducted and radiated, in locations where such testing would not ordinarily be possible because of the level of the ambient environment. It also prevents high level signals that must be generated during the course of testing from causing interference to other devices that may or may not be part of the test.

There are several difficulties in making measurements inside of a shielded enclosure. One is that many radiated tests employ antennas whose dimensions may be large compared to the size of the available enclosure. The result of antenna elements coming close to the conducting walls of the room is to load the antenna and cause changes in its calibration and radiation characteristics.

A second difficulty is that the multiple reflections of emitted signals that will occur in a shielded enclosure will result in standing waves in the room and potential measurement errors. This effect becomes particularly significant above the lowest resonant frequency of the room (see Figure 11-1)[1]. It can be aggravated by personnel within the enclosure, who can cause the standing-wave pattern in the room to change as they move about. There is no way to avoid this problem, but it may be alleviated to an extent in some measurements by moving the test antenna over a small volume to determine the worst-case susceptibility/interference radiation, by using microwave absorbing materials to reduce reflections, and by keeping the number of people and volume of extraneous equipment in the room to a minimum. A third difficulty is that practical room sizes result in many radiated tests being performed at distances for which $r \ll D^2/2\lambda$ (see Section 2.3.2), where levels of field strength are very difficult to establish and measure.

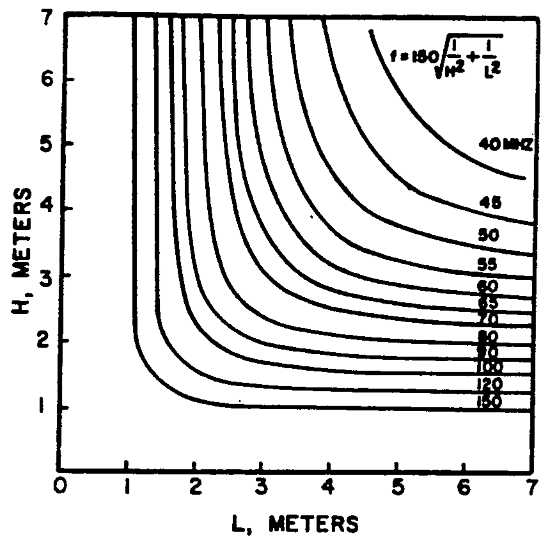


Figure 11-1 Resonant Frequency Chart

The alternative to testing in a shielded enclosure is open-site testing[2]. This requires a location several miles from any concentration of machinery or electronics and also some distance from overhead power lines. Such a site assures a reasonable ambient level, usually limited only by atmospheric noise and signals in known crowded frequency bands. The advantage of such a site is offset by inconvenience in having to go to a remote site for testing, and in the special precautions necessary to get power to the site. Also, use of an open-site does not eliminate the need for a shielded enclosure where it is necessary to isolate one part of a particular test employing a radiating signal source from another part of the test that uses a sensitive detection device.

11.3.2 Enclosure Design Considerations

There are three basic types of shielded enclosures. They include:

- Single wall
- Double wall, with the walls connected together at many points
- Double wall, with the walls connected together at only one point

The enclosures are assembled in panel sections. The sections may be bolted, clamped, or welded together to create the room. In the case of the multiply-connected double-wall enclosure, the inner and outer walls are connected around the periphery of each panel member. In the case of the single-point connected double-wall enclosure, the inner and outer walls are obviously kept separate electrically when the panel members are assembled.

Enclosures are made of either screen or solid metal. Typical screen materials are copper (22 x 22 - .015, measuring 22 x 22 wires per inch, each wire being 0.015" in diameter), bronze (18 x 20 - .011), and hot galvanized steel (.011 -162). Typical solid sheet materials used include 24.02 copper and 24 gage galvanized steel, or a combination of copper and steel sheets in double wall enclosures [3].

Room cost increases as one goes from a single-wall to a double-wall multiply-connected, to a double-wall single-point connected room. It also is higher for a solid wall room than for a screened room, and for a welded room versus a bolted or clamped room. Since room shielding effectiveness also goes in these same directions, it is important to carefully define the requirements needed in an enclosure before taking steps to purchase a room.

As far as power and signal feed-through requirements are concerned, a number of adequate connectors are available commercially that can provide the necessary continuity through the enclosure wall, and the isolation of signal leads that must be obtained for effective testing. The power lines are filtered for each phase separately and for the neutral at the entrance to the enclosure. Adequate filters are available commercially from a number of manufacturers [3].

The heating and air-conditioning ducts for both fresh air and return air are generally built into a single barrier panel. Their form is that of a honeycomb whose element dimensions act as waveguides operating below cutoff. The cutoff frequency of the honeycomb is many times that of the highest frequency used in the enclosure in order to protect the enclosure characteristics. The correct penetration of the barrier panel is generally specified by the manufacturer [3].

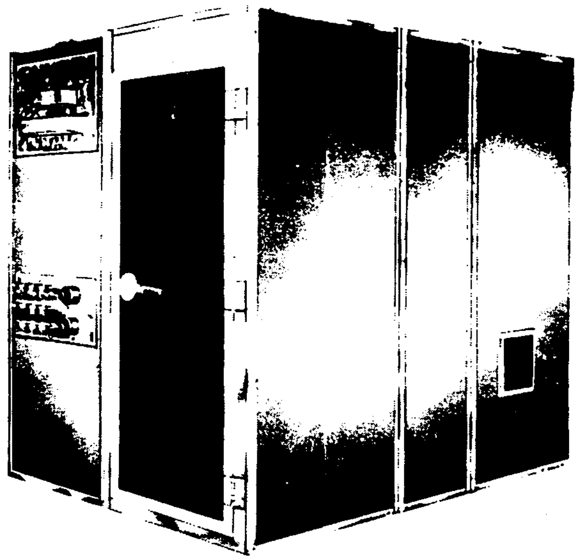
The opening of the enclosure that is the most critical and the one that can cause the most EMI problems is the entrance/exit door. There are many methods that will preserve the continuity of the room, and a wide variation is found amongst manufacturers.

The most common method of providing the barrier continuity is through the use of finger stock around the perimeter of the door. The finger stock can either be mounted around the door itself, or around the door frame. When the door is closed, the fingers are compressed, providing a good electrical contact throughout the door perimeter. One manufacturer employs a rather complex system that makes use of a set of pressurized air bags to force the door edges against a mating flange. This provides a good intimate contact so that the required uniformity of attenuation characteristics is protected.

There must be lighting in the enclosure. This should in all cases be incandescent, because other types of lighting usually involve ionization processes and subsequently produce RF noise.

Figure 11-2 provides an illustration of a commercially available solid-wall shielded enclosure. In Table 11-4, some of the general electrical characteristics of the three major room types are summarized. The 120 dB limit usually represents a testing limit, and not the limit of the room itself.

The characteristics of rooms designed to meet special performance requirements are not included in Table 11-4. For example, rooms have been designed and built that provide a magnetic shielding effectiveness of about 60 dB at 60 Hz, and reach the limit of typical measurements (120 dB) at 200 Hz.



Courtesy of Erik A.
Lindgren & Assoc., Inc.

Figure 11-2 Representative Solid Wall Shielded Enclosure

TABLE 11-4 Summary of Typical Shielded Enclosure Performance

Enclosure Type	Wall Type	Material	Magnetic Fields		Electric & Plane Waves	
			60 Hz	15 kHz	1 GHz	10 GHz
Double Electrically Isolated	Screen	Copper	2 dB	68 dB	120 dB	77 dB
		Bronze	0 dB	40 dB	110 dB	57 dB
		Galvanized		50 dB	50 dB	
	Solid	24 Ga. Steel	15 dB	84 dB	120 dB	90 dB
		Cu & Steel	18 dB	86 dB	120 dB	106 dB
Double Non-Electrically Isolated	Screen	Copper		52 dB	90 dB	
		Bronze				
	Solid	24 Ga. Steel		68 dB	90 dB	
		Copper			90 dB	
Single Shield	Screen	Copper	6 dB	42 dB	60 dB	
		Bronze				
	Solid	Copper			75 dB	
		24 Ga. Steel		48 dB	80 dB	

Note: Measurements made in accordance with MIL-STD-285.

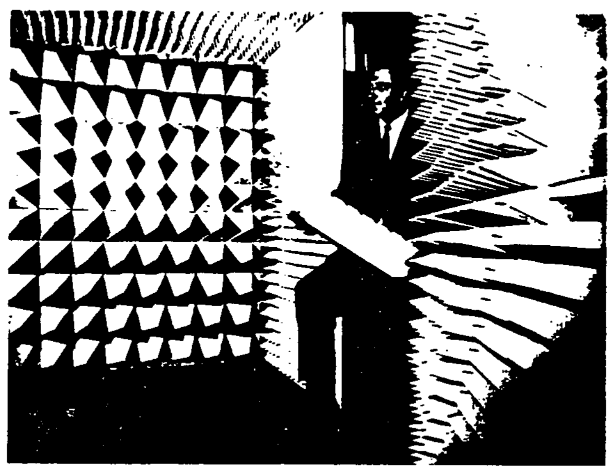
11.3.3 Microwave Absorbers

A well-designed RF anechoic (echoless) chamber must first be a good shielded enclosure, and then be further modified or finished into an anechoic enclosure. The basic requirements of such a chamber are high attenuation through the walls and high absorption on the inside walls.

Microwave-absorbent materials such as polyurethane foam, vinyl, and epoxies are used in anechoic chambers. These materials, usually pyramidal in shape, are placed on the surfaces of the chamber to absorb microwave energy. Signals from the test equipment or equipment being tested that are not absorbed by the equipment can be absorbed by the walls so that reflections are prevented or minimized. Cone-shaped sections of the absorbent materials are also available for incorporation in standard shielded enclosures.

The size of the cones or wedges determines, to a great extent, the lowest frequency at which the absorbent material is effective. The lowest practical limit at the present time is about 200 MHz, and requires using very large wedges. This rather high frequency, from an interference test standpoint, reduces the usefulness of the RF anechoic chamber for general purpose radiated energy measurements.

Figure 11-3 indicates a representative application of microwave-absorbent material. The material shown in the figure is effective about 300 MHz, and provides a reflection loss of about 40 dB at 1 GHz.



Courtesy of Emerson & Cuming, Inc.

Figure 11-3 Representative Anechoic Chamber

11.3.4 Enclosure Test Requirements

Shielded enclosures should be tested to MIL-STD-285 "Attenuation Measurements For Enclosures, Electromagnetic Shielding For Electronic Test Purposes, Method Of". This standard covers measuring the attenuation of the EM shielding over the frequency range of 100 kHz to 10 GHz and using this figure to designate the shielding effectiveness. The attenuation is the ratio (in dB) of received powers, on opposite sides of the shield, from the RF test source. Measuring these received powers yields data in the form of Attenuation vs Frequency. For the purposes of shielded enclosures used to test Avionics/GSE, the test points should be in increments of three frequencies per decade.

11.4 TESTING GUIDELINES

11.4.1 General

In performing EMI testing of GSE, both standard and specialized test equipment may have to be employed. In either case, general guidelines can be established regarding the required characteristics of this equipment, and the test setups in which they might be used. The following paragraphs are intended to document some of these guidelines.

11.4.2 Signal Sources

General laboratory signal sources and generators are usually suitable for EMC applications as well. The generator should have good frequency and amplitude stability, with an output level calibrated to within at least 2 dB. Its output VSWR should not exceed 1.3:1 (<0.1 dB mismatch loss) when terminated in the design output impedance.

Many signal generators exhibit some case or power line leakage that becomes particularly noticeable when the generator attenuator is near its maximum setting. This leakage may create a problem under low-level testing conditions. Particular attention should be given to keeping this leakage level down, through addition of a power-line filter to the generator, by improving the generator case shielding or enclosing the generator in a small shielded enclosure, or by using a lower leakage generator.

Care should be taken that signal generator harmonics are attenuated sufficiently to prevent false GSE responses from these signals. Low-pass or bandpass filters of known insertion loss at the desired frequency may be used for this purpose.

Many EMI tests necessitate a large dynamic measurement range. This would be the case in order to evaluate shielding or filtering effectiveness. Under these circumstances, high generator or source output levels are desirable. The particular level required depends on the type of test being performed, the sensitivity of the detection device being used, and other factors. Signal generator/power amplifier combinations can provide an adequate power level. Pulse generators can often be used to advantage when high peak power levels are needed.

Some pulse signal generators of the master-oscillator, keyed-amplifier type generate a cw output during the interval between pulses. This residual cw energy should be well below the peak pulse level (perhaps 30 dB down), and data collected using this type of generator should be carefully monitored to be sure a false response is not being obtained from the cw component of the signal.

11.4.3 Sweep Generators and Oscillators

Much EMI testing must cover measurements over a wide frequency range. Under these circumstances, it is convenient to use a frequency sweeping technique, and automatically record the detected output signal through a broadband receiver.

Commercial sweep oscillators are available that can be used for this purpose. They typically sweep an octave range at microwave frequencies (and wider at lower frequencies), have a wide range of sweep rates, have a fairly flat output level over the sweep range (within 0.1-0.2 dB), and put out ten to 100 milliwatts of power [3].

11.4.4 Attenuators

Fixed and variable attenuators are used extensively in EMI testing. The units should be accurate to within 0.5 dB when terminated in their characteristic impedance. As noted elsewhere in this design guide, proper termination of attenuators, as well as other test equipment, is important in maintaining test accuracy.

Variable or stepped attenuator assemblies are also frequently built into signal sources and detection devices. When attenuators are incorporated into a detector, it is important to know where they are located relative to the detector input. For example, it is often assumed that the attenuator in a frequency-selective voltmeter is at the input terminals of the voltmeter,and insertion of attenuation will prevent voltmeter front-end saturation; this is not always the case for low attenuation values. The initial attenuation may be added at IF, and therefore have no effect on the front-end situation.

When attenuators are clustered in a switch arrangement, particular care should be taken to be sure that energy does not get by passed through the switch assembly and result in erroneous readings. About 60 dB of isolation appears to be a practical level of switch decoupling to be expected[4].

11.4.5 Detectors

A wide variety of detection systems are employed for EMC testing. These devices might include power meters, crystal-video detectors, wide-band RF voltmeters, oscilloscopes, spectrum analyzers, and various other standard indicators. However, there is one detection system that is unique to interference analysis; that is an Electromagnetic Interference (EMI) meter.

The EMI meter is a two-terminal tunable voltmeter. It is essentially a calibrated super-heterodyne receiver having a wide dynamic range. Calibration is usually built into the EMI meter, using cw, random noise, and impulse noise sources. An impulse noise source operates by generating a very narrow (less than a nanosecond) repetitive pulse by alternately charging and discharging a short length of transmission line.

For performing radiated field measurements, a series of calibrated antennas are used. EMI meters are available that cover the frequency range of 30 Hz to 40 GHz. Peak and average detection modes are available.

Although current EMI meters are fairly well shielded, the dynamic range requirements of some tests may necessitate additional meter shielding. One way of doing this is to enclose the meter (or other type of detector) in a small shielded enclosure.

11.5 RADIATED AND CONDUCTED EQUIPMENT TESTING

11.5.1 General

In addition to the shielding, bonding, and filtering tests already discussed in this design guide, a number of other test procedures can be identified that are concerned with the EMI performance of a total equipment or system being employed. These procedures are for the purpose of exploring the radiated and conducted emission and susceptibility characteristics of an electrical/electronics device over a wide frequency range.

The generally accepted standard for these types of tests is MIL-STD-462 [5]. The tests contained in MIL-STD-462 are listed in Table 11-5. The reader who will be concerned with equipment-oriented EMC measurements, either to obtain an indication of overall GSE performance in the course of equipment design, or to evaluate a contract compliance requirement, is urged to familiarize himself with this specification.

11.5.2 Emission Tests

Most GSE do not constitute EMI emission hazards. However, equipment that employ digital clocks, or that generate switching transients, or that incorporate coherent or incoherent signal sources, may be potential interferences over either radiated or conducted paths.

Conducted emission tests at one time were performed by direct electrical coupling to the power or control lead under test through an impedance transformation device. Today, these tests are almost exclusively executed using a magnetically coupled probe device, calibrated in terms of the current flowing in the line under test. If the test is on a power line, a 10 micro-farad capacitor is used across each line to establish a low impedance between the line and ground.

Radiated emission tests use an EMI meter or equivalent detector coupled to either a magnetic loop or an electric field sensor. The antennas probe the field around the unit under test and measure the maximum level of radiation. The MIL-STD-462 test distances are 7 centimeters for the magnetic field sensor and 1 meter for the electric field sensor. Sensor calibration is in terms of far-field equivalent levels, even though it is recognized that a significant measurement error can result in using this calibration procedure.

11.5.3 Susceptibility Tests

Equipment-oriented EMI susceptibility tests are very similar in most respects to tests already discussed in this publication. The procedures employed to make radiated susceptibility measurements are the same as the shielding effectiveness tests, except that the indicator is no longer a

Table 11-5

INDEX OF MIL-STD-462 EMI MEASUREMENT TESTS

Test	Title
	CONDUCTED EMISSION (CE)
CE01	30 Hz to 20 kHz, Power Leads
CE02	30 Hz to 20 kHz, Control and Signal Leads
CE03	20 kHz to 50 MHz, Power Leads
CE04	20 kHz to 50 MHz, Control and Signal Leads
CE05	30 Hz to 50 MHz, Inverse Filter Method
CE06	10 kHz to 12.4 GHz, Antenna Terminal
	CONDUCTED SUSCEPTIBILITY (CS)
CS01	30 Hz to 50 kHz, Power Lead
CS02	50 kHz to 400 MHz, Power Lead
CS03	30 Hz to 10 GHz, Intermodulation, Two Signal
CS04	30 Hz to 10 GHz, Rejection of Undesired Signals At Input Terminals (2-Signal Generator Method)
CS05	30 Hz to 10 GHz, Cross-Modulation
CS06	Spike, Power Leads
CS07	Squelch Circuits
CS08	30 Hz to 10 GHz, Rejection of Undesired Signals At Input Terminals (1-Signal Generator Method)
CS09	60 Hz to 100 kHz, Structure Current (Common Mode Current)
CS10	10 kHz to 100 MHz, Damped Sinusoidal Transients, Pins and Terminals
CS11	10 kHz to 100 MHz, Damped Sinusoidal Transients, Cables
	RADIATED EMISSION (RE)
RE01	30 Hz to 30 kHz, Magnetic Field
RE02	14 kHz to 10 GHz, Electric Field
RE03	Spurious and Harmonic Emissions to 10 kHz to 40 GHz
RE04	20 Hz to 50 kHz, Magnetic Field
RE05	150 kHz to 1 GHz, Vehicles and Engine-Driven Equipment
RE06	14 Khz to 16 Hz, Overhead Power Lines
	RADIATED SUSCEPTIBILITY (RS)
RS01	30 Hz to 30 Khz, Magnetic Field
RS02	Magnetic Induction Fields
RS03	14 kHz to 10 GHz, Electric Field
RS04	14 kHz to 30 MHz, Electric Field
RS05	Electromagnetic Pulse Field, Transient

detection system whose sensor is located inside of a box, but is the system under evaluation. Similarly, conducted susceptibility tests are very much like the filter effectiveness tests, but with device performance being the susceptibility indicator. The MIL-STD-462 conducted procedures couple the test signal onto the line under test through a special impedance matching transformer for tests below 50 kHz, and through a capacitor whose reactance is less than 5 ohms for tests above 50 kHz.

The devices employed for radiated susceptibility measurements in MIL-STD-462 include the following:

- o Magnetic loop, with current through the loop being the indicator of input level, and the loop located 5 cm from the equipment under test.
- o Antennas, located 1 meter from the equipment under test.
- o Parallel Plate Transmission Line, of different construction than the one discussed in Section 4.6.7, since the specification requires testing using this device to only 30 MHz.

The direct application of MIL-STD-462 susceptibility testing under some circumstances may not provide a complete picture of GSE operational capabilities and limitations. Differences in test procedures that the GSE test engineer should consider relative to MIL-STD-462 susceptibility testing procedures include (a) the use of modulated input signals for the tests, and (b) the use of levels more representative of the expected GSE operational environment. Both of these modifications can result in more realistic indications of GSE performance.

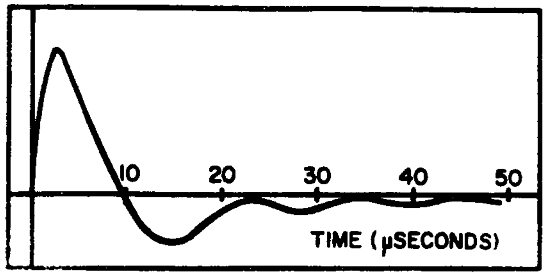
11.5.4 Transient Testing

Both conducted and radiated transient response tests are designated in MIL-STD-462. The unique nature of this type of test is worthy of comment here.

A standardized pulse is employed to test equipment transient response. The characteristics of this pulse are shown in Figure 11-4. Transient or spike generators are commercially available that generate this pulse, with peak amplitudes as high as 600 volts [3].

The pulse may not be representative of many signals that might be encountered in a GSE environment, and the test engineer should consider modifying the pulse shape to more realistically represent his needs. In particular, the pulse rise time of the test signal may be significantly longer than pulses created by fluorescent lighting, ignition systems, and rotating machinery[3].

The conducted transient susceptibility test is performed on AC or DC power lines. Figures 11-5 and 11-6 show alternative series and parallel test arrangements. Tests should be performed at different spike repetition rates (including those coincident with any clock rates or gating operations of the equipment under test), at different phase positions relative to the normal power-line signal, and at both positive and negative polarities.



CHARACTERISTICS

(1) Pulse Width = 10 microseconds to first zero crossing, 5.5 microseconds at half-voltage point

(2) Pulse Repetition Rate = 3 to 10 p.p.s.

(3) Pulse Rise Time = approximately 1.5 microseconds, 10 to 90% voltage levels

(4) Voltage Output = Not less than 200 V. peak

(5) Output Control = Adjustable from 0 to 200 V. peak

(6) Output Spectrum = 150 dB uv/MHz at 25 kHz decreasing to 115 dB v/MHz at 30 MHz

(7) Phase Positioning = 0 to 360 degrees

(8) Source Impedance (with injection transformer) = 0.06 ohms

(9) Transformer (current capacity) = 30 amperes

(10) External Synch = 50 to 800 Hz
External Trigger = 0 to 20 p.p.s.

Figure 11-4 MIL-STD-462 Transient Pulse Characteristics

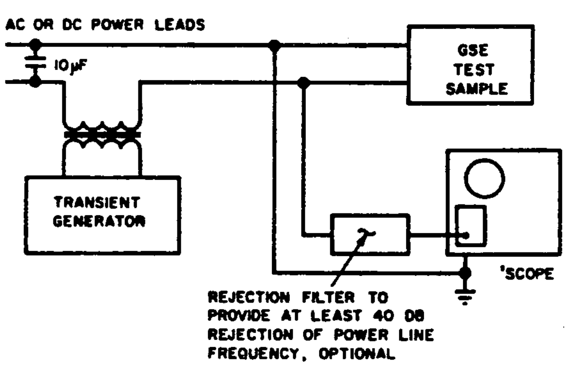


Figure 11-5
Conducted Transient
Susceptibility, Series Injection

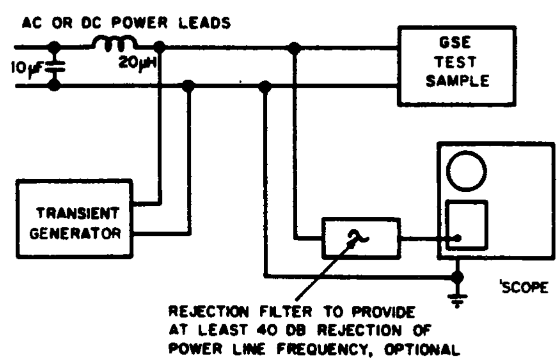


Figure 11-6
Conducted Transient
Susceptibility, Parallel
Injection

The radiated transient susceptibility test is an induction field test. It can be performed on input cable bundles or on equipment cases. To measure equipment transient susceptibility via cable bundles, the output of the transient generator is fed to a coil that is wrapped around the cable (see Figure 11-7). To measure transient susceptibility through the case, three turns of wire are wrapped around the case and excited by the generator (see Figure 11-8). The generator parameters are then varied as for the conducted transient susceptibility case, to establish the susceptibility threshold of the unit under test.

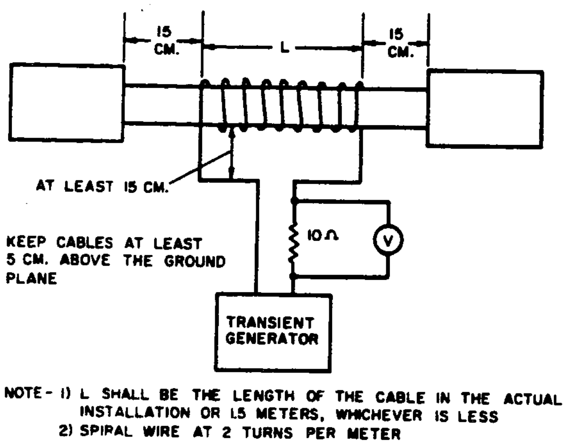


Figure 11-7
Radiated Transient
Susceptibility, Cable Test

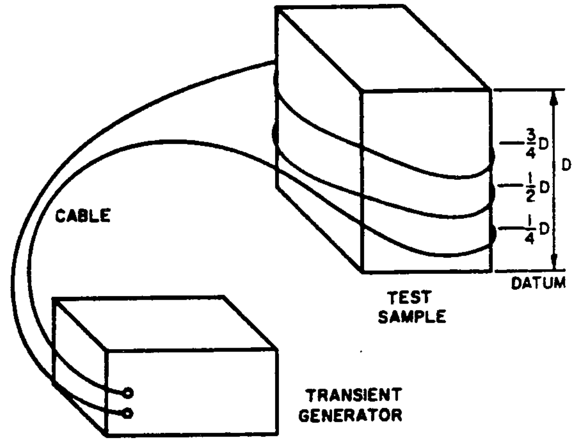


Figure 11-8
Radiated Transient
Susceptibility, Case Test

11.6 MIL-STD-1377 TESTING

This testing has been described and discussed in Section 4.6.8. It cannot be overemphasized that it is not enough for Avionics/GSE to meet MIL-STD-461 design criteria and MIL-STD-462 testing alone when the equipment being designed is to operate aboard Naval vessels. The most comprehensive test is when the unit is exposed to EMEs and experimental conditions that simulate onboard conditions in the fleet. This type of testing is performed by NSWC at NSWC Dahlgren, Virginia and NATC Patuxent River, Maryland. The facilities available consist of a mode-stirred chamber and a Ground Plane Testing platform to simulate the metal decks of Navy vessels.

A mode-stirred chamber is a large shielded enclosure with internal reflective surfaces. When the chamber is excited by EM energy, complex standing wave patterns are created which are then stirred by a field perturbing device. This facility can be used to measure shielding effectiveness (see Figure 11-9).

The ground plane testing platform is made of painted steel plates to simulate the deck of a Navy ship. The unit under test is set on this platform, and HF whips and pulsed radars are energized to provide the EME.

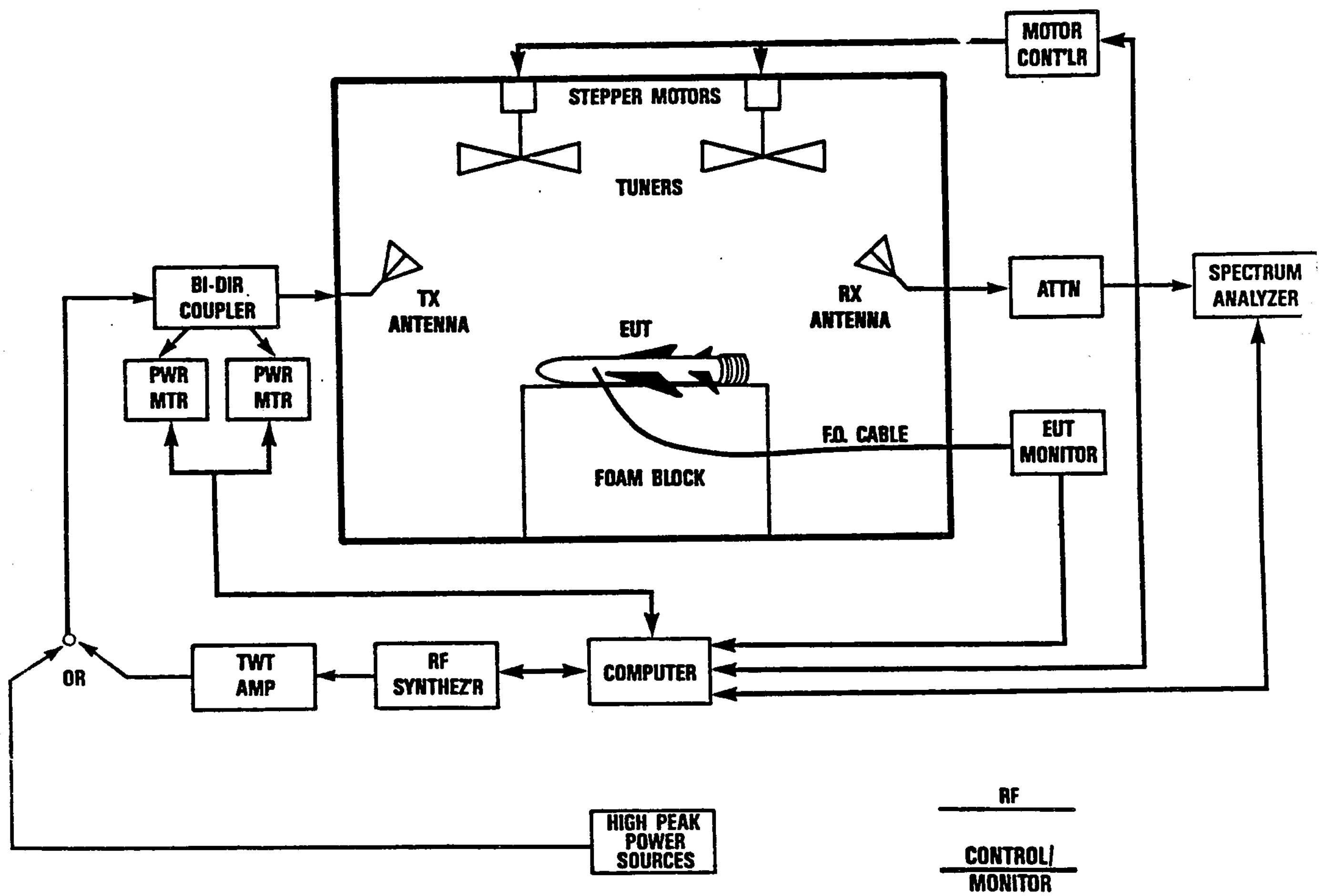


Figure 11-9 Mode-Stirred Chamber Test Set-Up

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[5] Department of Defense, Washington, DC, MIL-STD-462, "Electromagnetic Interference Characteristics, Measurement of," 31 July 1967.

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ERRORS OR DEFICIENCIES IN THE GUIDE:

RECOMMENDATIONS FOR CORRECTING THE DEFICIENCIES:

ADDITIONAL COMMENTS:

SUBMITTED BY: (Printed or typed name & activity)	DATE: