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MILITARY HANDBOOK

MANAGEMENT AND DESIGN GUIDANCE
ELECTROMAGNETIC RADIATION HARDNESS
FOR AIR LAUNCHED ORDNANCE SYSTEMS

FSC - EMCS

NO INFORMATION REQUIREMENTS

FOREWORD

The increasing severity of military electromagnetic environments and the potentially adverse effects of these environments on the performance of air launched ordnance systems have reached the point where they pose a threat to the successful deployment of air launched ordnance systems. To counter this threat, the U.S. Air Force has developed an Electromagnetic Radiation (EMR) Hardness Program to ensure that adequate EMR hardening measures are incorporated into the design, development, and production of future systems to protect them from EMR operational environments.

This handbook provides guidance for establishing, implementing, and managing an effective EMR hardness program throughout the life cycle of an air launched ordnance system. EMR hardness is one of several disciplines concerned with the detrimental effects of electromagnetic environments. Other disciplines include EMC, EMP, ECCM, and HERO. While this handbook is directed specifically to EMR hardness, the program established should be coordinated and consolidated with all other electromagnetic effects disciplines invoked on the system to provide efficient and cost effective solutions to the electromagnetic effects problems.

Beneficial comments (recommendations, additions, deletions) and any pertinent data which may be of use in improving this document should be addressed to: Rome Air Development Center, RADC (RBE-2), Griffiss AFB, NY 13441, by using the self-addressed Standardization Document Improvement Proposal (DD Form 1426) appearing at the end of this document or by letter.

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SECTION 1 - INTRODUCTION

1.1 **PURPOSE.** The purpose of this document is to provide program managers and system designers with guidance for the design, development, and acquisition of air launched ordnance systems which are hardened against the detrimental effects of electromagnetic radiation (EMR).

1.2 **SCOPE.** The material in this handbook is intended to be applied during the development, design, production, and deployment of air launched ordnance systems. The material is designed to provide protection for air launched ordnance systems during that portion of the life cycle from the time the system is attached to the delivery aircraft until the system impacts a target. The EMR environments are considered to extend from 1 MHz to 100 GHz.

1.3 **FORMAT.** The material in the remainder of the handbook is divided into five sections and fifteen appendices.

Section 2 (Referenced Documents) lists military documents which may be tailored to invoke EMR hardness requirements and control into the acquisition process. While there are currently no government specifications or standards which specifically address EMR susceptibility or vulnerability of air launched ordnance systems, the more general electromagnetic compatibility (EMC) documents listed in this section can be tailored to address the EMR hardness problem in contractual documentation. The information in this section is intended for government management and procurement personnel who are responsible for ensuring that EMR hardness is adequately addressed in contract documents. (Information on tailoring specifications and standards is presented in Appendix J.)

Section 3 (The EMR Vulnerability Problem) describes the nature, causes, and effects of EMR vulnerability. The information in this section is intended for management, design, and engineering personnel. It seeks to convey an understanding of the overall EMR vulnerability problem and to provide general descriptions of the EMR environment, the environment-to-system coupling mechanisms, and the degradation effects of EMR vulnerability.

Section 4 (EMR Hardness Control and Management) describes an overall management approach for implementing an EMR hardness program over the entire life cycle of an air launched ordnance system. The material in this section is intended primarily for program managers and EMCAB personnel who are responsible for the development, implementation, and control of EMR hardness programs for air launched ordnance systems.

Section 5 (EMR Hardening Design) describes an overall approach and the specific procedures that contractor personnel may utilize to ensure that the EMR hardness requirements are satisfied during the development and fabrication of a system. The information in this section is intended primarily for contractors' management, design, and engineering personnel who are responsible for incorporating adequate EMR hardness into the design and fabrication of air launched ordnance systems. The material in this section will also be of interest to government program managers and EMCAB personnel who are responsible for monitoring and evaluating contractors' EMR hardness efforts.

Section 6 (EMR Hardness Measurement Program) describes an overall test and evaluation plan to verify the EMR hardness of a system. The information in this section is intended for both the program office personnel who are responsible for establishing and evaluating an overall measurement plan and the contractor personnel who are responsible for developing an EMR hardness test plan and performing EMR hardness tests.

The appendices describe in greater detail major areas which must be addressed in an EMR hardness program and provide additional guidance for addressing these areas. The appendices include the following:

A. EMR Environment

B. EMR Environment Forecasting Capabilities at ECAC

C. Analysis and Prediction

D. The Intrasystem Analysis Program (IAP)

E. Establishing Susceptibility Levels

F. Establishing EMR Hardness Criteria

G. EMR Hardness Design Practices

H. EMR Hardness Measurement Techniques

I. EMR Hardness Considerations in Program Documents

J. Tailoring of Specifications and Standards

K. Outline for EMR Hardness Program Plan

L. Outline for EMR Hardness Control Plan

M. Outline for EMR Hardness Test Plan

N. EMR Hardness Bibliography

O. Definitions and Acronyms

SECTION 2 - REFERENCED DOCUMENTS

2.1 **ISSUES OF DOCUMENTS.** The following documents of the issue in effect on date of invitation for bids or request for proposal, form a part of this handbook to the extent specified herein.

SPECIFICATIONS

MILITARY

MIL-E-6051	Electromagnetic Compatibility Requirements, Systems

STANDARDS

MILITARY

MIL-STD-461	Electromagnetic Interference Characteristics, Requirements For Equipment
MIL-STD-462	Electromagnetic Interference Characteristics, Measurement Of
MIL-STD-463	Definitions and System Of Units, Electromagnetic Interference Technology
MIL-STD-1377	Effectiveness of Cables, Connectors, Weapon Enclosure Shielding and Filters In Precluding Hazards of Electromagnetic Radiation to Ordnance; Measurement of
MIL-STD-1541	Electromagnetic Compatibility Requirements For Space Systems

HANDBOOKS

MILITARY

MIL-HDBK-237	Electromagnetic Compatibility/Interference Program Requirements

(Copies of specifications, standards, drawings, and publications required by contractors in connection with specific procurement functions should be obtained from the procuring activity or as directed by the contracting officer.)

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SECTION 3 - THE EMR VULNERABILITY PROBLEM

3.1 **CAUSES AND EFFECTS OF EMR VULNERABILITY.** The mission requirements of air launched ordnance systems dictate that they operate in intense and highly complex electromagnetic environments. From the time the system is attached to the delivery aircraft until it impacts a target, it is exposed to electromagnetic radiation from emitters aboard the delivery aircraft, from emitters aboard other aircraft, or from emitters located on the ground. The environment created by these emitters may contain signals which reach hundreds of volts per meter in level, employ complex modulations, and span a frequency range of 1 MHz to 100 GHz. If these signals are coupled into sensitive electronic circuits of the air launched ordnance system, then degradation of circuit performance may occur. Any degradation which is sufficient to compromise the system mission constitutes electromagnetic vulnerability.

3.2 **THE EMR ENVIRONMENT.** The increased use of high power electromagnetic emitters has brought about sources with effective radiated power (ERP) output levels ranging up to tens of megawatts. These sources, which are very often designed to intentionally radiate power within selected portions of the RF spectrum, can create very intense electromagnetic fields at large distances from the source location. Where tactical requirements or spatial limitations dictate the operation of sensitive electronic systems in close proximity to these sources, the systems may be exposed to electromagnetic field levels which far exceed the normal design requirements of the systems.

Any high power emitter may create EMR vulnerability problems, either through the unintentional or intentional radiation of EM energy. Intentional sources are those designed specifically to radiate EM energy, for example, radar, communication, EW, Navaids, or other type systems. The predominant sources are those with high power output levels and highly directive antennas. In particular, many pulsed radar systems radiate extremely high peak power levels. Systems which are illuminated by the main beams of pulsed radar antennas may be subjected to field intensity levels of hundreds of volts per meter.

In the performance of its intended missions, an air launched ordnance system will be exposed to several different electromagnetic environments created by different combinations of radiation from sources located on the delivery aircraft, on other aircraft, and on the ground. The composite of these environments will be characterized by intense electromagnetic fields, signal frequencies ranging from 1 MHz to 100 GHz, and complex signal modulations. Such an environment poses a major EMR threat to the operation of an air launched ordnance system. To insure that this threat is minimized or eliminated, it is mandatory that the operating environment of the system be well defined in order to permit the determination and incorporation of appropriate system hardening techniques.

The Electromagnetic Compatibility Analysis Center (ECAC) has developed a capability for defining composite environments for air launched ordnance systems. This capability provides a means of defining the EMR environment for a given system based on the system type and function, the type of delivery aircraft, other aircraft involved, the theater(s) of operation and the types of targets.

3.3 **ENVIRONMENT-TO-SYSTEM COUPLING.** Coupling is defined as the means by which a magnetic, electric, or electromagnetic field produced by one system induces a voltage or current in another system, and is broadly classified as conductive or free-space coupling. Conductive coupling occurs between two systems when the systems are physically connected with a conductor and share a common impedance. Free-space coupling is the transfer of electromagnetic energy between two or more systems not directly interconnected with a conductor. Depending upon the distance between the systems, free-space coupling is usually defined as either near-field or far-field. Near-field coupling can be subdivided into inductive or capacitive coupling, according to the nature of the electromagnetic field. In inductive coupling, the magnetic field set up by the source links the receptor. Capacitive coupling is produced by an electric field between the source and receptor.

Radiation of energy by electromagnetic waves is the principal coupling mechanism in far-field coupling. The term, radiated coupling, is sometimes used to describe both near-field and far-field coupling. However, radiated coupling is generally accepted as the transfer of energy from a source to a receptor by means of electromagnetic wave propagation through space according to the laws of wave propagation.

During the life cycle of an air launched ordnance system, undesired electromagnetic energy may be transferred to the system via radiated coupling from sources on the delivery aircraft, on other aircraft, or on the ground. The amount of energy which is coupled will be dependent upon the size and configuration of the system, the orientation of the system with respect to the energy source, and the frequency and polarization of the incident energy.

Energy which is coupled to a system from an incident electromagnetic field will cause current to flow on the surface of the system. If these currents are interrupted by a discontinuity in the form of a hole or seam in the surface, the field will penetrate into the interior of the system. The amount of penetration will depend upon the distribution of current on the system surface and the size and configuration of the hole. Once penetration has occurred, the undesired energy may be coupled to internal circuits and components through a combination of conducted/free-space coupling paths. If the system is unable to distinguish the coupled energy from legitimate signals, the system performance may be degraded.

3.4 **DEGRADATION EFFECTS.** Undesired RF energy which is coupled to a semiconductor device will be absorbed by the device. The amount of energy absorbed will depend on the level of energy coupled to the device via interconnecting wiring and cables, the frequency of the undesired signal, the type and operating conditions of the device involved, the device port into which the energy is coupled, and the impedance of the injection port. The effect of the coupled energy on the device performance can range from an alteration of the device operating characteristics to device failure. At low levels of power absorption the predominant effect will be device performance degradation whereas higher levels of power absorption will result in device failure.

The basic mechanism by which device performance is affected is rectification of the coupled RF energy at p-n junctions. The rectified current (voltage) will appear as a dc or video signal depending upon the modulation characteristics of the RF energy. For CW RF energy, the rectified current will produce a DC shift in the quiescent operating point of the device. For pulsed RF, the operating point effectively becomes a superposition of the original (no RF) value with a video signal which is a replica of the RF envelope.

The effect of device performance degradation or failure on system performance will depend upon the function of the circuit in which the device is employed and the function of interconnected circuits and subsystems. For example, CW RF energy which is rectified in an analog amplifier may cause a shift in the quiescent operating point of the amplifier or change the amplifier gain. Similarly, a change of state may be induced in a digital circuit. Where the undesired signal is modulated, the detected modulation waveform may appear as an output of the device. These effects may be propagated through interconnected circuits and subsystems and degrade the operation of the circuit or subsystem. The specific effects of coupled energy on circuit, subsystem, and system performance can only be determined by the system designer, beginning with a determination of the effects produced on specific devices and then analyzing the impact of device performance degradation or failure on circuit and subsystem behavior.

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SECTION 4 - EMR HARDNESS CONTROL AND MANAGEMENT.

4.1 **PROGRAM MANAGER'S RESPONSIBILITIES.** The program manager is responsible for ensuring that an effective EMR hardness program plan is developed early in the concept development phase of the program and that the EMR hardness program defined in this plan is implemented and evaluated throughout the life cycle of the system being developed. The program manager should ensure that:

a. EMR hardness requirements are defined and considered during trade-off studies of alternate concepts for satisfying the required operational capabilities.

b. Adequate schedules and budgets are established to accomplish the required EMR hardness actions.

c. The EMR hardness aspects of the system being developed are adequately addressed in the Decision Coordinating Paper (DCP) at the conclusion of each development phase.

d. The EMR hardness requirements for the system are adequately defined in the Request For Proposal (RFP).

e. The contractor is satisfying the EMR hardness requirements in the system design.

f. The contractor is satisfying the EMR hardness requirements in the fabrication of the system.

g. The pre-production prototype system is susceptibility tested and passes a vulnerability analysis.

h. The EMR hardness characteristics of the system are maintained during production.

i. The production model system is susceptibility tested and passes a vulnerability analysis.

To assist him in meeting these responsibilities, the program manager should establish an electromagnetic compatibility advisory board (EMCAB) early in the concept development phase of the program to serve as a major resource for the development, implementation, control, and review of an EMR hardness program.

4.2 **THE ELECTROMAGNETIC COMPATIBILITY ADVISORY BOARD (EMCAB).** THE purpose of the EMCAB is to assist the program manager in establishing, implementing, and controlling an electromagnetic effects program which shall include an EMR hardness program. The EMCAB should normally consist of 2 to 6 members (depending on the size and complexity of the program) with extensive experience and expertise in EMC engineering. In addition to the EMR hardness program, the EMCAB should have the responsibility for satisfying the requirements for any other electromagnetic effects specified in the Program Management Directive (PMD). These additional EM effects may include any combination of EMC, EMP, ECCM, HERO, RADHAZ, and other possible EM disciplines. The membership of the advisory board should be tailored to provide the expertise required to address the EM disciplines to be satisfied.

With respect to the EMR hardness program, the EMCAB should:

a. Obtain EMR environment forecasts as required during the development and acquisition phases of the program.

b. Generate an EMR hardness program plan.

c. Establish schedules and budget estimates for accomplishing the EMR hardness program.

d. Prepare EMR hardness inputs to Decision Coordinating Papers.

e. Prepare EMR hardness requirements for inclusion in the SOW and RFP.

f. Evaluate the EMR hardness aspects of proposals.

g. Evaluate contractor EMR hardness control plans and test plans.

h. Monitor contractor EMR hardness efforts and programs.

i. Identify and resolve EMR hardness problems which arise.

j. Evaluate contractor EMR hardness test data.

k. Assist in making arrangements for system susceptibility tests and vulnerability analyses.

The program manager should ensure that the EMCAB is organized sufficiently early in the concept development phase of the program so that the board can participate in trade-off studies of alternate concepts and can assist in establishing adequate schedule and budget estimates to implement a comprehensive EMR hardness program. In the organization of the advisory board, the program manager should ensure that the board's responsibilities are clearly defined and that sufficient authority is given to the board to accomplish its goals.

4.3 **THE EMR HARDNESS PROGRAM PLAN.** The EMR hardness program plan is the top level management document for the complete EMR hardness program to be conducted throughout the life cycle of the air launched ordnance system being developed. The EMR hardness program defined in the program plan should be tailored to the specific operational requirements and the anticipated EMR environment for the system being developed. The plan should be prepared in accordance with the outline presented in Appendix K.

The EMR hardness program plan should include the following types of information:

a. Description of the overall management, organizational and technical framework of the EMR hardness program.

b. Definition of tasks and milestones of the EMR hardness program.

c. Assignment of responsibilities for EMR hardness tasks.

d. Delegation of authority for EMR hardness actions.

e. Description of the implementation of the EMR hardness program.

f. Description of the EMR hardness tasks to be accomplished in each phase of the system life cycle.

The EMR hardness program plan should be reviewed and updated as the program progresses through the acquisition cycle. As a minimum, the program manager and the EMCAB should jointly review the EMR hardness program plan at the conclusion of each acquisition phase to ensure that the program plan accurately reflects the current requirements of the system being developed.

4.3.1 **ORGANIZATION AND MANAGEMENT.** The EMR hardness program plan should clearly describe the organizational and management structure of the EMR hardness program for the particular project. The program plan should establish which Air Force, other DoD, and contractor organizations will be required to participate in the EMR hardness program and identify the communication channels and contact points between the various organizations and the program office. The program plan should include schedule and budget estimates for each of the participating organizations.

4.3.2 **ASSIGNMENT OF RESPONSIBILITIES AND AUTHORITY.** The EMR hardness program plan should clearly define the responsibilities and goals of each participating organization in the overall EMR hardness program. The program plan should delegate adequate management responsibilities and authority to appropriate organizations and individuals to ensure that the entire EMR hardness program organization can function and accomplish its goals.

Under the direction and approval of the program manager, the EMCAB should have primary responsibility and authority for developing, implementing, and managing the EMR hardness program.

4.3.3 **INCORPORATING EMR HARDNESS IN PROGRAM LIFE CYCLE.** The EMR hardness program plan should clearly define the objectives, tasks, and milestones of the EMR hardness program. In addition to assigning responsibilities for the various EMR hardness tasks and actions, the program plan should describe how and when these tasks and actions are to be accomplished in relationship with the various phases of the system life cycle. The program plan should ensure that the EMR hardness actions are accomplished in a manner which will provide maximum benefit at minimum cost, and at the same time, will cause minimum delay in the development of the system.

4.3.3.1 **PROGRAM LIFE CYCLE PHASES.** The five life cycle phases of a typical air launched ordnance system are listed below:

a. Concept Development

b. Concept Validation

c. Full-scale Development

d. Production

e. Deployment

Certain EMR hardness activities must be accomplished during each of these life cycle phases. In addition, these activities must be accomplished in a certain sequence, both to assure the efficient accomplishment of the overall EMR hardness program and to maintain a smooth flow of the acquisition process.

4.3.3.2 **CONCEPT DEVELOPMENT PHASE.** The EMR hardness activities which should be accomplished in the concept development phase of the program are depicted in the flow diagram shown in Figure 4-1. The program manager should organize an EMCAB as early as possible in this phase of the program to assist him in establishing and implementing an EMR hardness program. The first actions of the EMCAB should be directed to obtaining an EMR environment forecast defining the EMR environment in which the system will be required to operate and to developing an EMR hardness program plan. The Electromagnetic Compatibility Analysis Center (ECAC) at Annapolis, Maryland has developed the capability for generating EMR environment forecasts for air launched ordnance systems. The EMCAB should establish contact with ECAC, alert them that a request for an EMR environment forecast is forthcoming, and obtain information as to what types of input information will be required to request the forecast. ECAC's request procedures are described in Appendix B. Operational deployment information (anticipated delivery platforms, mission scenarios, and anticipated target parameters) will be required to define the anticipated EMR environment. The accuracy and completeness of the EMR environment forecasts obtained in the early stages of a program will probably be limited by the degree to which the operational deployment information is defined. Hence, the EMCAB should make a concerted effort to supply ECAC with the best possible information in the request for an EMR environment forecast.

At the same time that the EMR environment forecast is being obtained from ECAC, the EMCAB should be generating an EMR hardness program plan tailored to the system to be developed. The program plan should completely describe the EMR hardness program to be conducted throughout the life cycle of the system to be developed.

After the EMR hardness program plan has been completed and approved and the EMR environment forecast has been received from ECAC, the EMCAB and the program manager should establish schedule and budget

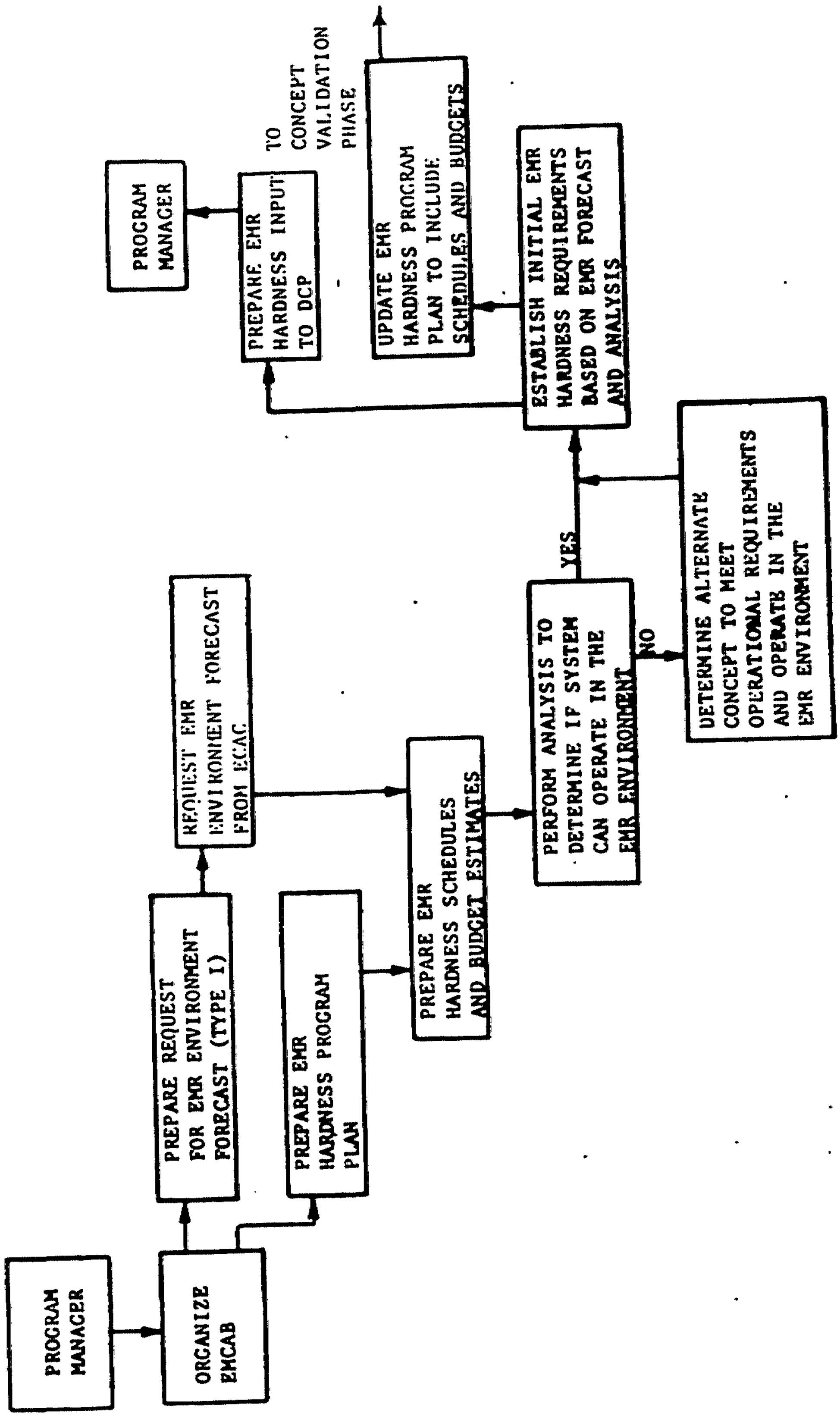

Figure 4-1. EMR Hardness Activities During Concept Development Phase of System Life Cycle.

estimates for accomplishing all of the tasks of the EMR hardness program. This effort will require the program manager and the EMCAB to interface with the organizations assigned responsibilities under the program plan to obtain their inputs to and concurrence with the schedule and budget estimates.

At this stage of the program, the EMCAB should perform an analysis to determine if the system, as proposed, can operate in the anticipated EMR environment. If more than one concept is being considered, this analysis can serve to establish the relative magnitudes of the EMR hardness requirements of the alternate concepts. If a single concept is being considered and the results from the analysis indicate that the system can not operate in the anticipated EMR environment, an alternate concept will have to be developed which satisfies the required operational capabilities, and, at the same time, is capable of operating in the anticipated EMR environment. The results from the analysis should be used to establish an initial estimate of the EMR hardness requirements for the remainder of the program.

Near the conclusion of the concept development phase, the EMCAB should prepare the EMR hardness documentation for inclusion in the Decision Coordinating Paper (DCP). This documentation should include a description of the tasks of the EMR hardness program that have been accomplished, the results from these tasks, an estimate of the tasks remaining to be accomplished, and an assessment of the risks involved in completing the EMR hardness program.

At the conclusion of the concept development phase, the EMCAB should update the EMR hardness program plan to reflect the current status of the EMR hardness program. The update of the EMR hardness program plan must be sufficiently detailed to assure continuity in the EMR hardness program in the transition from the concept development phase to the concept validation phase.

4.3.3.3 **CONCEPT VALIDATION PHASE.** The EMR hardness acivities which should be accomplished in the concept validation phase of the program are depicted in the flow diagram shown in Figure 4-2. The primary emphasis of the EMR hardness program during this phase should be directed to establishing the EMR hardness requirements for inclusion in the Request For Proposal (RFP) for the full-scale development system.

The EMCAB should contact ECAC, alert them that an updated EMR environment forecast will be required, and obtain information as to what types of input information will be required to request the updated forecast. The EMCAB should then initiate an effort to obtain updated performance, operational, and tactical information in accordance with ECAC's requirements and prepare a request for an updated EMR environment forecast. After the updated forecast is obtained, the EMCAB should perform an analysis to determine if the system, as being developed, can operate in the updated EMR environment. The results from the analysis should be used to establish the EMR hardness requirements for inclusion in the Statement Of Work (SOW) and the RFP for the full-scale development model of the system. After the EMR hardness requirements have been established, the EMR hardness test and analysis requirements

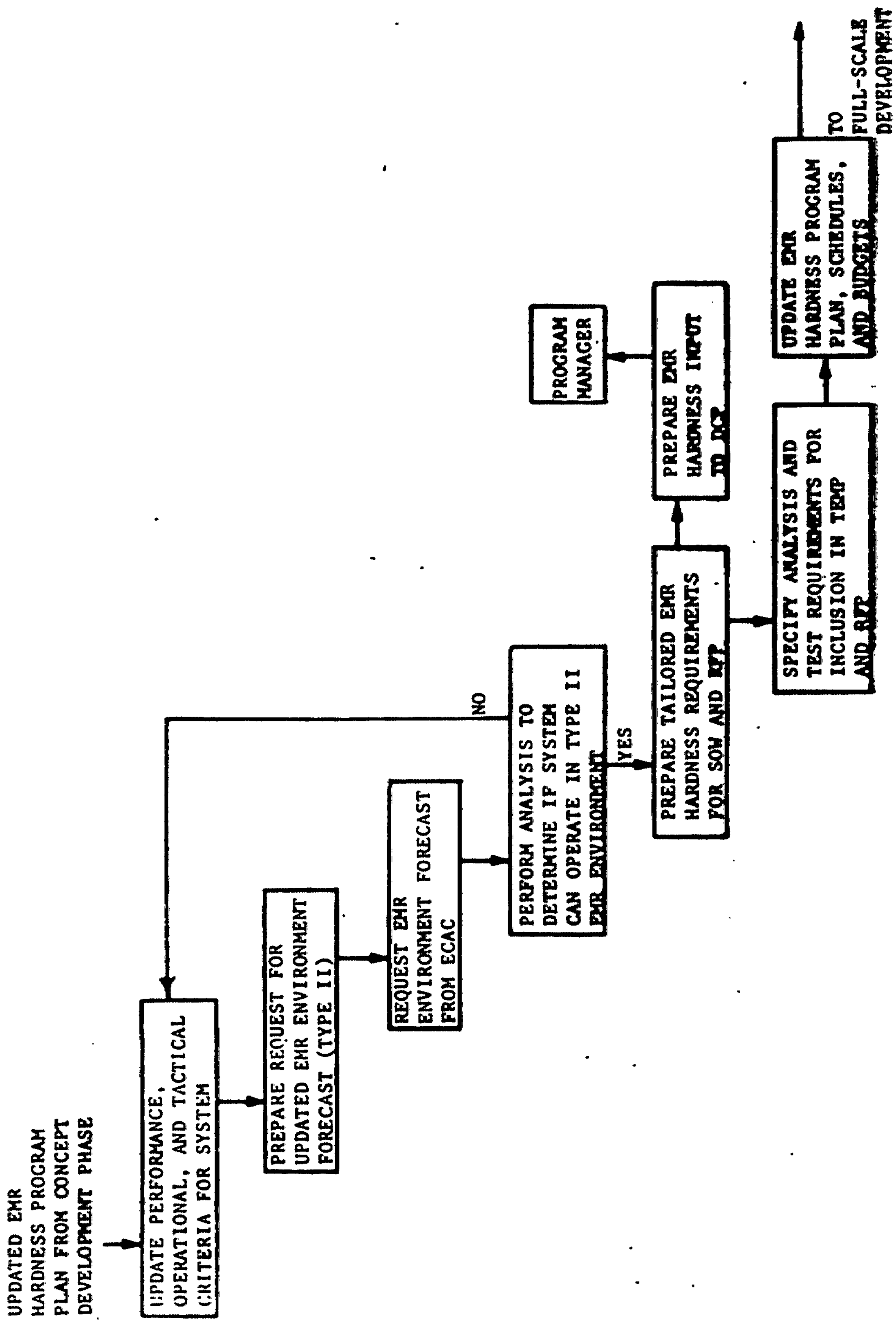

Figure 4-2. EMR Hardness Activities During Concept Validation Phase of System Life Cycle.

for inclusion in the RFP and the Test and Evaluation Master Plan (TEMP) should be developed.

Near the conclusion of the concept validation phase, the EMCAB should prepare the EMR hardness documentation for inclusion in the DCP. At the conclusion of the concept validation phase, the EMCAB should update the EMR hardness program plan to assure continuity in the EMR hardness program in the transition from the concept validation phase to the full-scale development phase.

4.3.3.4 **FULL-SCALE DEVELOPMENT PHASE.** The EMR hardness activities which should be accomplished in the full-scale development phase of the program are depicted in the flow diagram shown in Figure 4-3. The primary emphasis of the EMR hardness program during this phase should be directed to ensuring that the contractor incorporates sufficient EMR hardening in the design and fabrication of the full-scale development model of the system. The EMCAB should ensure that the RFP for the acquisition of a pre-production model of the system adequately defines the EMR hardness requirements for the system and requires the bidders to address how they propose to satisfy the EMR hardness requirements in their proposals. The contract should require the successful bidder to submit an EMR hardness control plan describing in detail how he will satisfy the EMR hardness requirements in the design and fabrication of the system and an EMR hardness test plan describing in detail how he will evaluate his EMR hardening efforts.

The EMCAB should participate in the evaluation of the proposals to determine if bidders adequately and realistically address the EMR hardness requirements in their proposals. The EMCAB should evaluate the control plan and test plan submitted by the successful bidder and require modifications if necessary to obtain satisfactory approaches. The EMCAB should participate in periodic design reviews to ensure that the EMR hardness requirements are being adequately addressed in the design of the system. The EMCAB should review and evaluate the contractor's EMR hardness test data to ensure that EMR hardening has been incorporated in the fabrication of the pre-production system. The program manager and the EMCAB should contact RADC/RBC and ECAC and alert them that a system susceptibility/vulnerability analysis of the pre-production model of the system is forthcoming. The EMCAB and RADC/RBC should jointly prepare a request to ECAC for an updated EMR environment forecast for the system susceptibility/vulnerability analysis at the RADC facility. The program manager and the EMCAB should make arrangements to provide a pre-production system to RADC/RBC for system susceptibility testing.

The EMCAB should establish the EMR hardness requirements and quality control verification tests for inclusion in the production contract. Near the conclusion of the full-scale development phase, the EMCAB should prepare the EMR hardness documentation for the DCP. At the conclusion of the full-scale development phase, the EMCAB should update the EMR hardness program plan to assure continuity in the EMR hardness program in the transition from the full-scale development phase to the production phase.

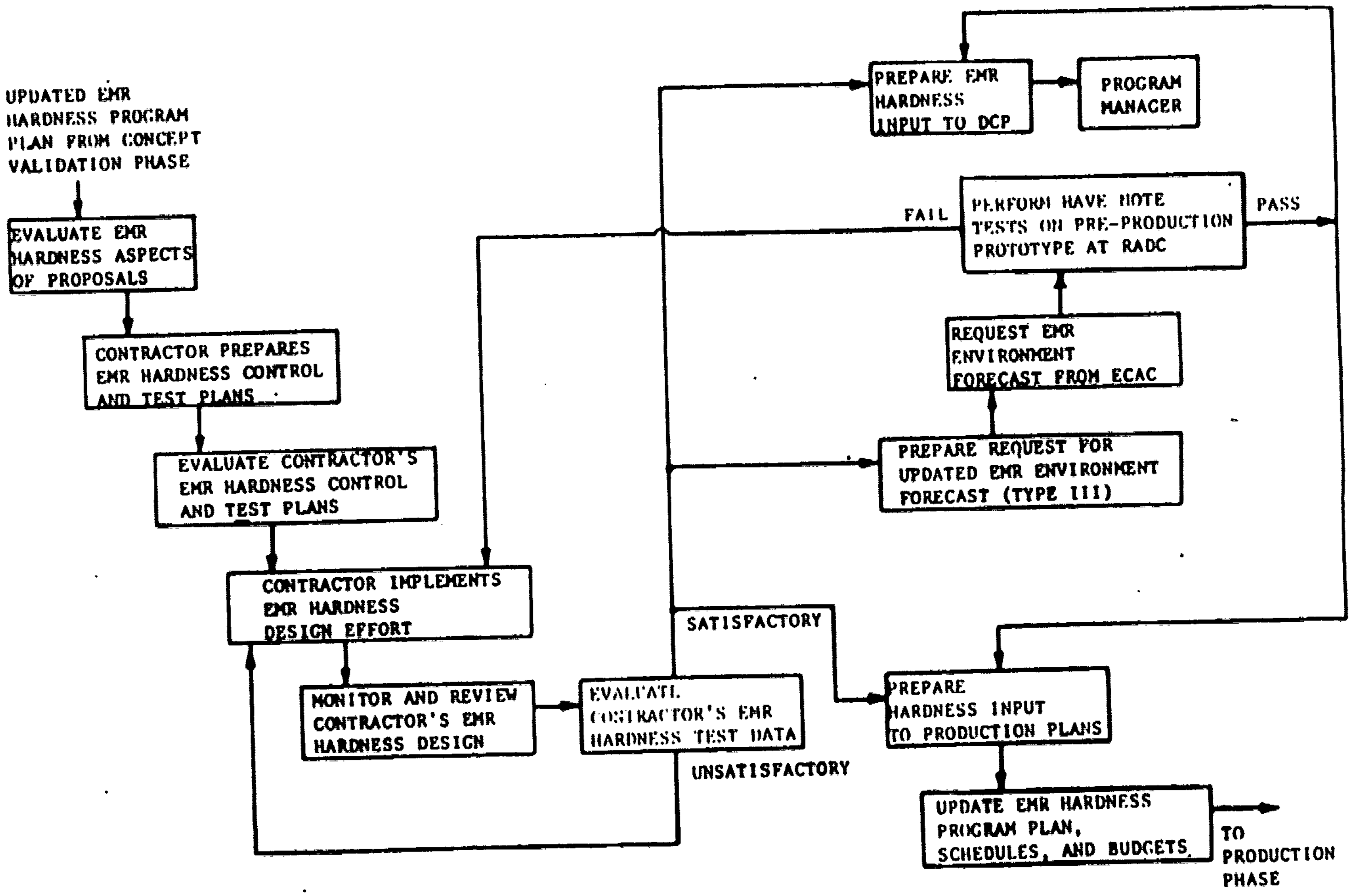

Figure 4-3. EMR Hardness Activities During Full-Scale Development Phase of System Life Cycle.

4.3.3.5 **PRODUCTION PHASE.** The EMR hardness activities which should be accomplished in the production phase of the program are depicted in the flow diagram shown in Figure 4-4. The primary emphasis of the EMR hardness program during this phase should be directed to ensuring that the EMR hardness characteristics of the pre-production model of the system are maintained in the production process. The EMCAB should verify that the production procedures and practices to be used in the production of the system will yield the required EMR hardness characteristics. The EMCAB should also ensure that adequate quality control tests are performed throughout the production cycle to assure that the required EMR hardness characteristics are being realized in the production systems. The EMCAB should ensure that the EMR hardness aspects of the system are adequately addressed in all operation, maintenance, and training documents for the system.

The program manager and the EMCAB should contact RADC/RBC and ECAC and alert them that a system susceptibility/vulnerability analysis of a production model of the system is forthcoming. The EMCAB and RADC/RBC should jointly prepare a request to ECAC for an updated EMR environment forecast for the system susceptibility/vulnerability analysis at the RADC facility. The program manager and the EMCAB should make arrangements to provide a production model system to RADC/RBC for system susceptibility testing.

If the production model of the system should fail to satisfy the system susceptibility/vulnerability analysis, the program manager and the EMCAB, with assistance as required from RADC/RBC, should develop modifications, engineering changes, or changes in the production practices to meet the EMR hardness requirements.

After the production model system satisfies the system EMR susceptibility/vulnerability analysis, the EMCAB should update the EMR hardness program plan to provide a complete record of the EMR hardness program to the logistics program manager.

4.3.3.6 **DEPLOYMENT PHASE.** The EMR activities which should be accomplished in the deployment phase of the program are depicted in the flow diagram shown in Figure 4-5. The deployment phase begins with the acceptance of the first operational system and extends until the last system is phased out of the inventory. There is usually a significant overlap between the production phase and the deployment phase. During this overlap period, a feed-back system should be established and maintained to ensure that any EMR deficiencies discovered during in-service performance are routed back to production for corrective actions. The EMR environment should be monitored and updated throughout the deployment phase. If the actual changes in the EMR environment differ significantly from the forecasted changes, the impact of the changes on the vulnerability of the system should be assessed and appropriate actions taken. The operational and maintenance procedures for the system should be monitored throughout the deployment phase, and their impact on the EMR hardness characteristics of the system should be assessed. Any proposed plans for modifications or engineering changes to the system should be reviewed to assess their impact on the EMR hardness characteristics of the system.

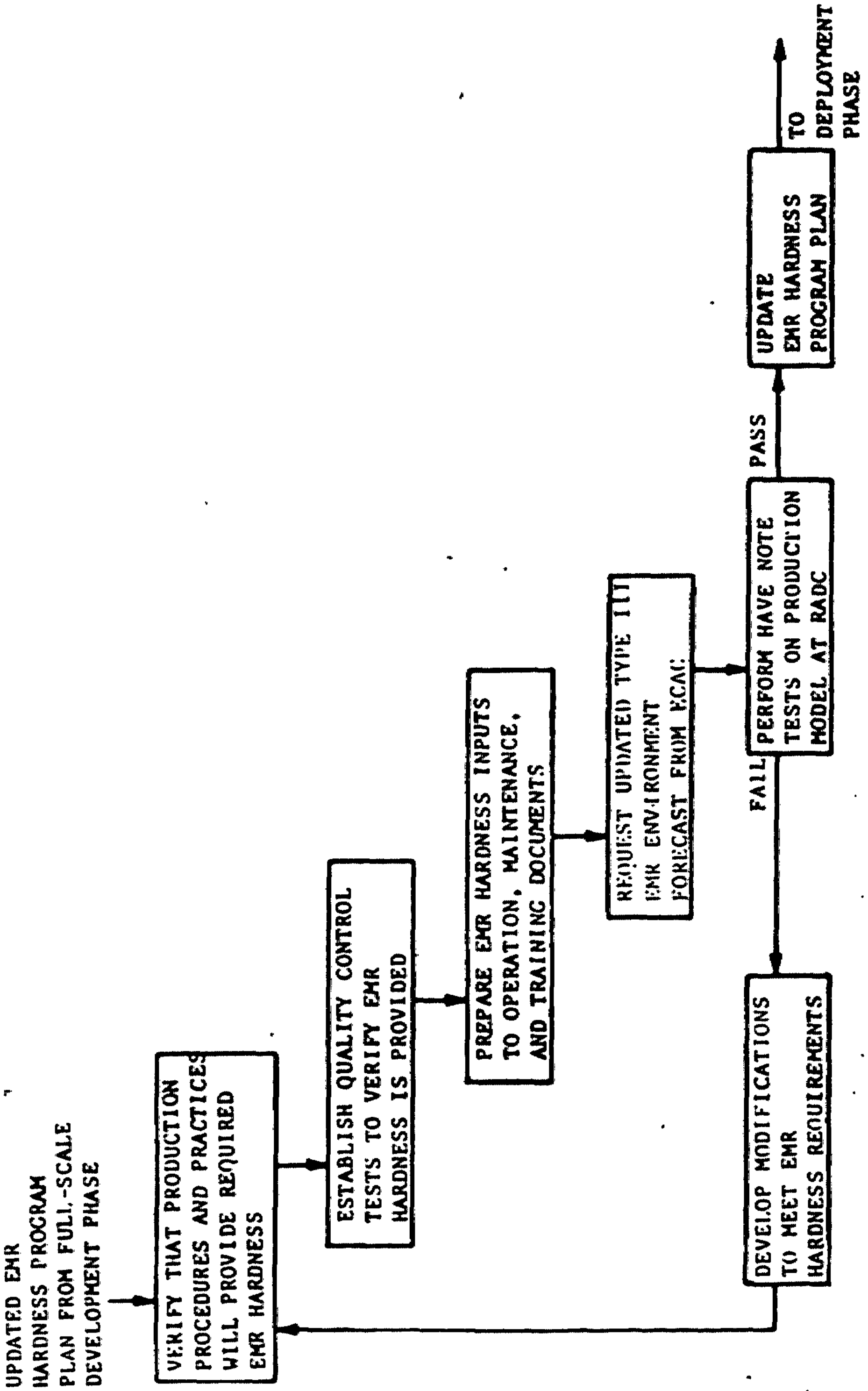

Figure 4-4. EMR Hardness Activities During Production Phase of System Life Cycle.

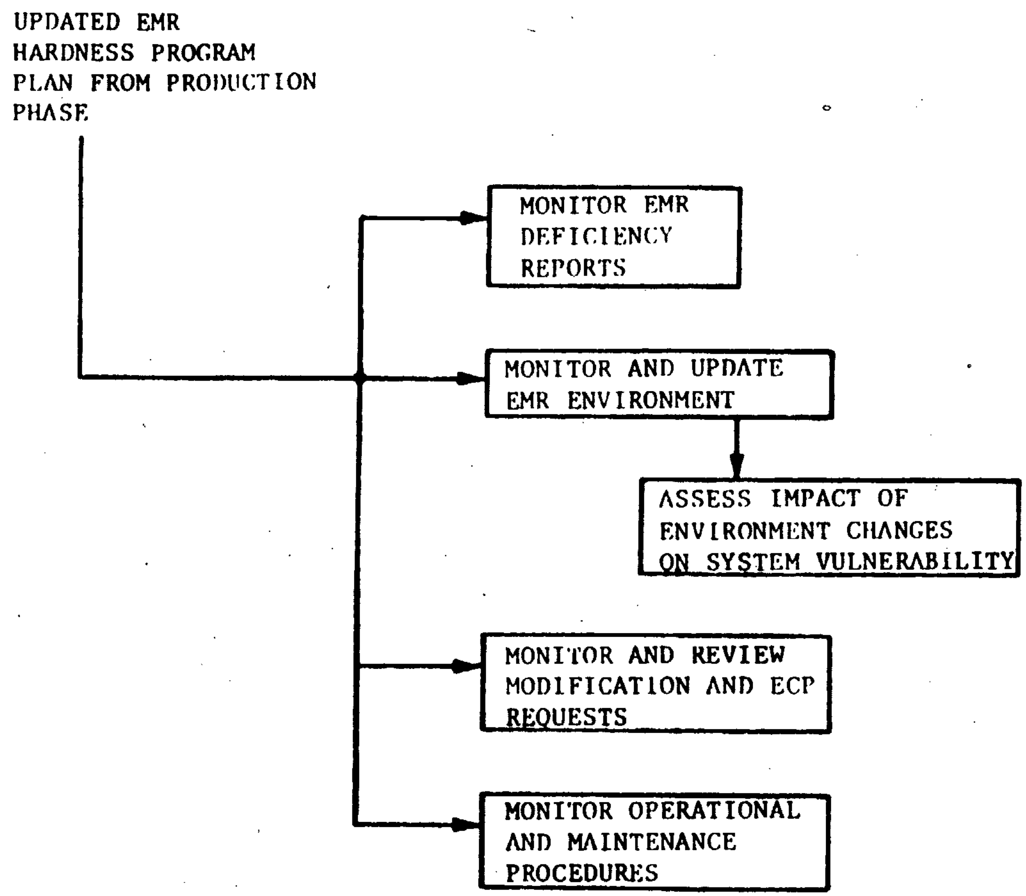

Figure 4-5. EMR Hardness Activities During Deployment Phase of System Life Cycle.

4.4 **EMR ENVIRONMENT FORECASTS.** An air launched ordnance system will be exposed to several different electromagnetic environments during its life cycle. The EMR hardness provided in the system must be sufficient to assure that the system can survive and accomplish its mission when exposed to all of these environments. Thus, an EMR environment profile depicting the maximum radiation levels for all the environments must be defined in order to establish the EMR hardness requirements for a system. In addition, since a system being developed may not be deployed for several years and may have an in-service life of several more years, the EMR environment profile used to establish the EMR hardness requirements must be extrapolated or forecasted to indicate the levels of radiation exposure anticipated at the end of the in-service life of the system.

The implementation of an Air Force EMR hardness program has created a continuous requirement for EMR environment forecasts as the developments of new systems are initiated. While the operational EMR environment of each system must be tailored for that specific system, the procedures for generating the EMR forecasts are essentially the same for all systems. In addition, the generation of an EMR forecast for a specific system will entail processing the same EMR data base to tailor the environment to the operational and tactical requirements of the specific system. Under these conditions, it is apparent that a permanent organization to provide EMR environment forecasts to all program offices is needed. The Electromagnetic Compatibility Analysis Center (ECAC) at Annapolis, Maryland has developed a capability for generating the required EMR environment forecasts.

A number of requirements should be considered in the generation of EMR environment forecasts. A minimum of three EMR environment forecasts is considered necessary during the acquisition cycle of a system. The points in the acquisition cycle at which these forecasts will be required are; (1) the Mission Element Need Statement (MENS) approval stage, (2) the Request For Proposal (RFP) preparation stage, and (3) the system EMR susceptibility/vulnerability evaluation stage.

The first EMR environment forecast (Type I) at the MENS approval stage, should be used in the feasibility analysis, trade-off studies of alternate approaches, and the definition of risks. This forecast should also be used to establish budgets and resources requirements for an EMR hardness program.

The second EMR environment forecast (Type II) at the RFP preparation stage, should be incorporated into the RFP to convey to the bidders the amount of EMR hardness that will be required. This forecast should also be used by the selected contractor to establish the EMR hardness criteria which will be used as the basis for the EMR hardness control plan and the EMR hardness test plan.

The third EMR environment forecast (Type III), at the system EMR susceptibility/vulnerability evaluation stage, should be used by the system test organization as guidance in conducting the system susceptibility tests and as the threat definition in vulnerability analyses.

Each EMR environment forecast must include all elements of all the electromagnetic environments to which the system will be exposed during all phases of the life cycle that the EMR hardness program is to address. The current concept is that the EMR hardness program will address the operational/threat environment present from the time the system is attached to the delivery aircraft until the system impacts the target.

While the basic objective of all three forecasts is the same (i.e., to define the anticipated operational/threat environment), the manner in which each forecast is to be used, the stage of development of the system when each forecast is generated, and the extended time periods between forecasts (possibly several years) dictate that the three EMR environment forecasts will be different.

The following specific features and information are considered necessary in the EMR forecasts in order for them to satisfy the requirements of an EMR hardness program:

TYPES

Type I - MENS Stage (25-year forecast)
Type II - RFP Stage (20-year forecast)
Type III - System EMR Evaluation Stage (15-year forecast)

FORMAT

Composite profiles including ground, on-board (Cosite), escort and intercept aircraft (Intersite), and approach-to-target environments.

Average and peak power profiles

FREQUENCY RANGE

1 MHz to 100 GHz

LIFE CYCLE PHASES

From attachment to delivery aircraft to impact on target

MODULATION CHARACTERISTICS

Probability distribution curves (or equivalent) of pulse width and pulse-repetition frequency over specified bands in the environmental profiles

ECAC has developed procedures to satisfy all of these requirements (including the integration of environments and a forecasting capability) in a capability for generating EMR environment forecasts. Hence, when an EMR environment forecast is needed, the program manager need only contact ECAC, define which type forecast he needs, and provide ECAC the operational and tactical information they require to generate the forecast. A more detailed description of EMR environment forecasts and examples of EMR environment profiles are presented in Appendix A. ECAC's capabilities and request procedures are described in Appendix B.

4.5 **ESTABLISHING EMR HARDNESS REQUIREMENTS.** The EMR hardness requirements need to be established early in the EMR hardness program and refined as the system develops. The establishment of the hardness requirements must be based on a systematic approach and commensurate with the needs at the various phases of the acquisition process. A systematic approach may flow as follows:

a. Based on the EMR environment forecasts, predict the internal EM fields within a system enclosure via aperture and external cable paths.

b. Predict coupling of the internal EM fields to cables and wires inside the system enclosure.

c. Determine susceptibility levels of the components, circuits and subsystems that have been identified as critical to system performance.

d. Assess the effect of the EM-induced response on system peformance, perform trade-offs, and establish adequate design margins in hardness requirements.

The approach for determining EMR hardness requirements varies greatly in complexity according to the available system configuration data, level of analysis, and accuracy requirements. As soon as possible in the system development, the analysis approach described should be balanced with testing and in some cases with computer model simulations. A well-balanced attack utilizing analysis, testing, and simulation is recommended since in many situations no one area provides all the necessary data or information desired. Establishing EMR hardness criteria is described in more detail in Appendix F.

4.5.1 **COUPLING ANALYSIS.** The level of coupling analysis will be different for the various phases of the acquisition process. During the conceptual phase, order-of-magnitude calculations may be sufficient. During these phases, little hardware information is available and system requirements are not well-defined. However, it is possible to use coupling analysis for trade-offs in alternative system designs and to study feasibility-type questions. This is true in part because of the wide range of quantities involved. For example, electromagnetic environment levels may be in the hundreds of volts/meter while component/circuit susceptibilities may exist at the millivolt or microvolt levels.

As the weapon system progresses through the acquisition cycle, the accuracy requirements for coupling analyses increase. In the validation and full-scale development phase, more system type information is available and coupling analyses may be approached with more detailed computer-aided analyses. Also in these phases, since hardware is available for testing, a balanced utilization of coupling analyses and measurements is possible. For example, analyses may be used to determine skin penetration levels and measurements based on these levels may be used to determine cable pickups.

The Air Force has available a family of EMC analysis computer programs to support both order-of-magnitude and detailed calculations. This collection of programs is referred to as the Intrasystem Analysis Program (IAP). RADC is the lead agency for development and use of the IAP. Recently, RADC has set up the EMC/IAP Support Center to provide a facility where government and industry may obtain support in executing and exchanging data and information on the IAP. This facility is located at RADC.

In the early stages of system development, the program manager may request system support through RADC/RBC which in turn will interface with the EMC/IAP Support Center. Once a contractor has been established, the program manager may request support for the contractor directly with the EMC/IAP Support Center. The support request procedure for government agencies is through a Memorandum of Agreement (MOA). For non-government organizations or contractors, the request for support must be on a specific tasking or an annual subscription basis.

Some of the major functions of the support center are:

a. Update and maintain computer programs.

b. Establish a configuration control system to maintain a record of all computer programs, data bases, center users, distribution schedules, etc.

c. Collect information, prepare documentation (such as newsletters, etc.) and disseminate information regarding activities, products and services of the center.

d. Prepare and present training courses for users of the computer programs and seminars for government and contractor personnel involved in the acquisition process.

e. Provide EMC liaison between the product divisions (SPO's) and their contractors.

f. Obtain, establish and maintain a library of data bases generated on system procurements.

g. Integrate all new software models into the computer programs and develop supporting documentation.

The IAP consists of two parts. The Intrasystem Electromagnetic Compatibility Analysis Program (IEMCAP), which is used for system level EMC analysis, is based on worst-case modeling techniques. IEMCAP is suitable for first level analyses and for order-of-magnitude calculations. The second part of IAP consists of off-line and supplemental computer models for higher levels of analysis and detail type calculations. The off-line and supplemental models are used for electromagnetic field analysis, wire and cable coupling, nonlinear circuit analysis, and lightning/precipitation static studies.

IEMCAP is the major component of IAP and was designed to provide an effective and cost beneficial method of analysis throughout the phases of a weapon system acquisition. IEMCAP may be used for such functions as coupling analysis, specification tailoring, and comparative analysis on which to base trade-off decisions. Some of the points in the acquisition where IEMCAP can be used to advantage are; (1) prior to and in the generation of the Statement of Work (SOW), (2) between contract award and the Preliminary Design Review (PDR) (perhaps as an aid in the generation of the EMR Hardness Control Plan), (3) at an intermediate time between the PDR and Critical Design Review (CDR), and (4) during system test planning.

Appendix D presents a more detailed description of IAP.

4.5.2 **COMPONENT, CIRCUIT, AND SUBSYSTEM SUSCEPTIBILITY LEVELS.** Electromagnetic susceptibility levels play an integral role in determining the EMR hardness requirements. These data and information, together with EM environment and coupling information, enable a designer to assess the system hardening requirements. However, effective hardness program planning often requires two different approaches to susceptibility evaluation.

A prediction of the susceptibility levels will be required early in the system life cycle so that realistic EMR hardness scheduling and budgeting estimates can be made. This requirement occurs typically during the concept development phase, when no actual hardware information is available. The lack of hardware information necessitates a worst-case analysis using the lowest interference levels of what might be termed "typical" circuits and devices. These levels may be obtained from the composite graphs in Appendix E.

A reassessment of the susceptibility levels will be required after information is available on the actual hardware to be employed in the system. It will now be possible to establish more accurate electromagnetic susceptibility levels using data on the particular components and circuits used in the design. When the scope of all possible devices, components, and circuits employed in modern ordnance systems is considered, the data in Appendix E appears quite limited. Unfortunately, a totally comprehensive data base is simply not available at the present time. There are three basic approaches to obtaining susceptibility levels for devices and circuits of concern on which no information is available. These include measurements, analytical modeling, and data extrapolation, all of which are addressed in more detail in Appendix E.

4.5.3 **SYSTEM EMR SUSCEPTIBILITY ANALYSIS AND PREDICTION.** The EMR environmental forecasts, the coupling analysis, and the susceptibility data form the basis for the system EMR susceptibility analysis. The EMR environment impinges on the outer skin of the weapon system, and through various modes of coupling, establishes an EM field in the interior of the weapon enclosure. The modes of coupling requiring consideration are direct field diffusion, aperture coupling, and coupling through external cables or other penetrations. The interior field itself couples to cables, wires, circuits, subsystems, etc., that comprise the weapon system electronics. Eventually, the RF currents that result from the interior fields are rectified in a nonlinear junction such as found in a discrete or integrated semiconductor device. For CW RF energy, the rectification results in a dc shift in the operating point of a circuit; for pulsed or modulated RF, the nonlinear junction is essentially an envelope detector. Once the RF currents are rectified, the rectified signal may propagate through the remainder of the circuit or subsystem as though it were a legitimate signal. If the circuit or subsystem is critical to system performance, the rectified signal may cause degradation of system performance and thus produce system EMR susceptibility.

In order to identify system susceptibilities before-the-fact, analysis and prediction methods must be relied on. Within the present state-of-the-art, coupling analysis must rely to a large extent on worst-case models. This approach is necessitated by the system complexities (e.g., electronic wiring), unavailability of physical configuration data, and the costs of detailed analyses. Guided weapon systems such as air-to-air and air-to-ground missiles are relatively small (as compared to an aircraft), are of simple geometrics (e.g., cylindrical shapes), and are self-contained. These properties make the weapons systems themselves reasonably manageable from a coupling analysis point-of-view. However, susceptibility analysis must be considered for these weapon systems in two configurations: inflight and onboard. For the inflight configuration, the weapon is in free flight and the EMR impinges most likely as uniform illumination. In the onboard configuration, however, the weapon is connected to an aircraft wing or fuselage and the coupling response becomes a function of the complete system: aircraft, weapon system and connecting cables. Also for the onboard configuration, the EMR field is most likely to be non-uniform or in the near-field of cosite emitters. Thus, it is seen that the onboard configuration is not as easily subjected to coupling analysis as the inflight configuration.

In the conceptual/validation phases, it may be necessary to estimate the proposed system sensitivity to EMR when:

a. System or component parameters are not well specified.

b. System geometry has not been determined.

c. Order-of-magnitude estimates of system sensitivity are sufficient.

For this situation, statistical analysis may be more appropriate. Statistical analysis techniques are under development, but are not yet sufficiently refined to be recommended. The alternative is to rely on simpler, deterministic, worst-case models and perform sensitivity-type analyses on the above cases.

IEMCAP has models, such as the field-to-wire and the wire-to-wire that are suitable for worst-case analysis. The field-to-wire model characterizes the coupling of the electromagnetic environment through apertures to interior wiring and cables. Exposed wires/cables are assumed to be adjacent to the aperture and the level of coupled EMR is a function of aperture size and location. A transmission-line model and a tuned-dipole model are used to compute the currents induced in the wires/cables. The transmission-line model is used for the lower frequencies (up to 100-800 MHz) and the tuned-dipole model is used for the higher frequencies (greater than 100-800 MHz). The frequency at which the two models crossover depends on the system geometry.

IEMCAP requires that the port susceptibilities (the nonrequired spectra) be user specified. In the conceptual/validation phase, the specification of these susceptibilities must rely on component data as given in Appendix E or on past experience with similar systems. Once circuits and equipments are developed or specified, these susceptibilities may be refined by tests or computer-aided circuit analyses.

Within IEMCAP, the user has the capability to specify interior field levels. Once the system external geometry is known, a more refined analysis may be appropriate to determine the interior fields. For example, body-of-revolution or finite-difference, time-domain codes are available to study apertures in cylindrical bodies. These codes may be used to investigate required apertures such as optical ports in a guided weapon system.

EMR analysis and prediction techniques are discussed in more detail in Appendix C, and the Intrasystem Analysis Program (IAP) is described in Appendix D.

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SECTION 5 - EMR HARDNESS DESIGN

5.1 **GENERAL.** The purpose of this section is to provide design guidance for hardening air launched ordnance systems against incident electromagnetic environments. The material presented is intended primarily for the system designer, although the various subsections contain information which should also be helpful to management, engineering, and test personnel. The order in which the material is presented generally corresponds to the flow of the hardening design process. Section 5.2 describes the nature and requirements of the EMR Hardness Control Plan, a contractually required document which establishes the contractor's management and engineering plan for achieving the required system hardness. The remaining sections, Sections 5.3 through 5.8, describe the major elements and tasks of the hardening design process. Section 5.3 presents an overall design methodology that the system designer may follow in the system hardening process, from program initiation through prototype completion and validation tests. Section 5.4 describes how environment definitions, coupling analyses, and system susceptibility data may be used to define system hardening requirements. Sections 5.5 and 5.6, respectively, describe an approach the designer may take in implementing the system hardening design, and identify hardness techniques and devices which may be employed. Section 5.7 discusses the use of design tradeoffs to ensure compatibility or to resolve conflicts between system functional and hardness requirements, and Section 5.8 describes an approach the designer may follow to verify system hardness prior to system test. EMR hardness design practices are described in Appendix G.

5.2 **EMR HARDNESS CONTROL PLAN.** The EMR Hardness Control Plan is the contractor prepared document which describes in detail his approach to ensuring system hardness. The plan is prepared, delivered, and updated as specifically required by the contract. The plan will be reviewed by, and must receive the approval of, the program manager and the EMCAB. The control plan should be prepared in accordance with the outline given in Appendix L.

The specific structure of the EMR Hardness Control Plan and the information to be documented will depend upon, and should be tailored to, the particular air launched ordnance being developed. Typically, the plan should address:

a. A definition of the applicable air launched ordnance system.

b. The EMR hardness program scope, objectives, and requirements.

c. The organization and management of the EMR hardness program.

d. The program tasks to be accomplished and the schedules and milestones to be met.

e. The documents (handbooks, standards, specifications, etc.) to be employed.

f. A definition of words, terms, or phrases used to describe the hardness program.

g. The approach to be followed in establishing system hardness requirements.

h. The EMR hardness requirements to be imposed on suppliers and subcontractors for vendor items and subsystems.

i. The design methodology, requirements, and techniques for achieving system hardness.

j. The analysis and measurement techniques to be used in defining or verifying system hardness.

k. The documentation to be provided to verify the hardness design.

The submission date of the initial EMR Hardness Control Plan and subsequent revisions or updates will be established by contractual requirements. Typically, the date of submission of the initial plan will be from 90 to 120 days after award of contract. The required dates of submission of updated plans will depend upon such factors as contract duration and system complexity. As with the initial plan, all revised plans will be subject to the review and approval of the program manager and the EMCAB.

5.3 **SYSTEM HARDENING METHODOLOGY.** A well organized EMR hardening design approach should be used by the system designer to ensure that the hardening of an air launched ordnance system is accomplished in a cost effective manner. Figure 5-1 illustrates a methodology that the system designer may follow in the system hardening process. The inputs required from the program manager to initiate the process are definitions of the operational environment and the functional and tactical requirements of the system. Given these inputs, the system designer should formulate a system design concept and employ data, analyses, and measurements to determine if the design concept is susceptible to the specified environment. The susceptibility assessment will require a determination of environment-to-system coupling and system susceptibility to the coupled signals. Methods for determining system susceptibility and environment-to-system coupling are discussed in Appendices C and E.

Based on the results of the susceptibility assessment, the system designer should define the system hardening requirement and proceed with a hardening design which will preclude system susceptibility. The designer should utilize documented design data and techniques, analyses, and measurements as necessary during the design process to achieve the required hardening level. Specific steps should be taken to thoroughly document all aspects of the design, including the design approach, the hardening techniques and devices employed, and the analyses and measurements performed to substantiate the degree of hardening incorporated at the device, circuit, or subsystems levels. Sufficient information and data should be included to verify that the design ap-

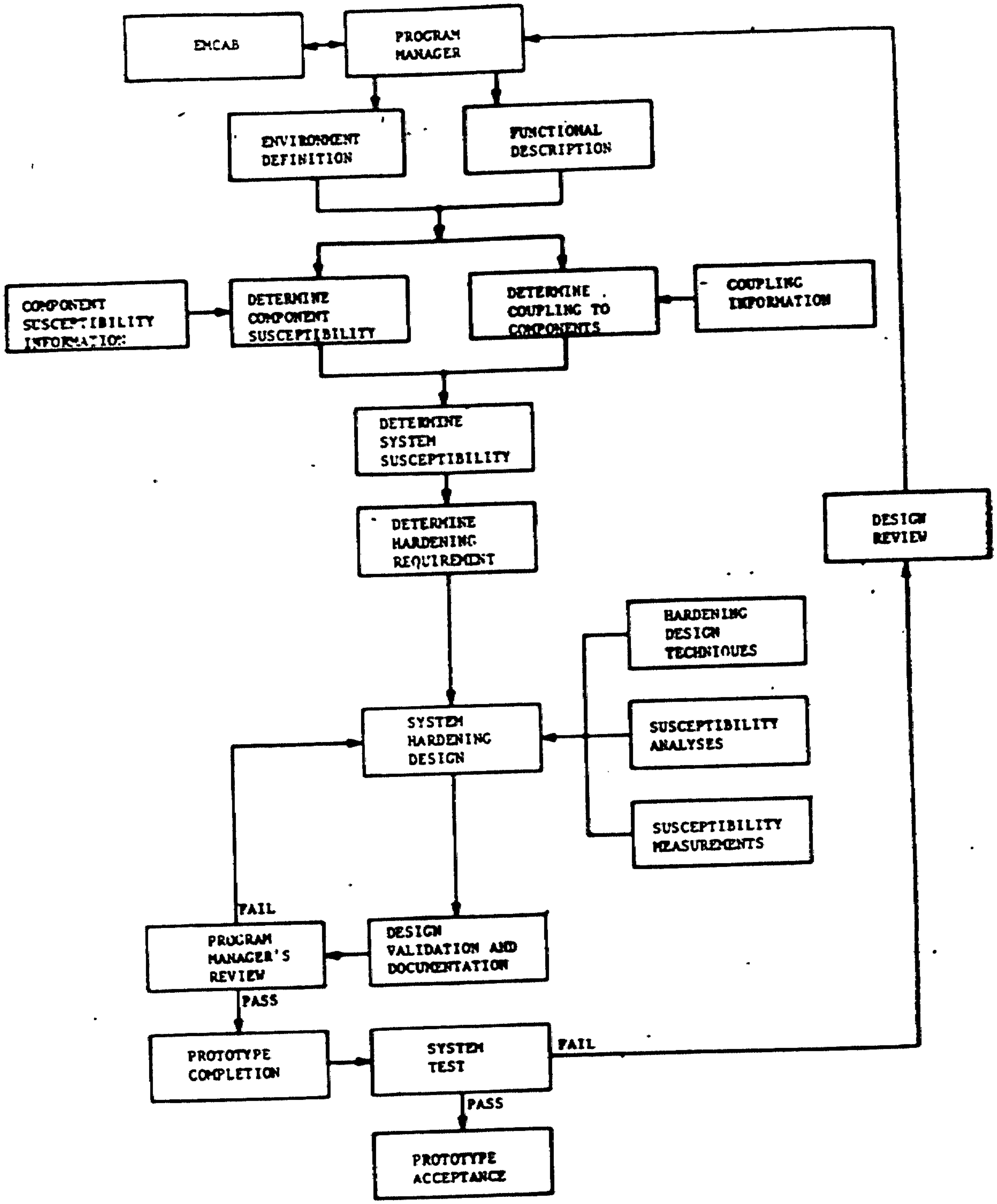

Figure 5-1. Methodology of System Hardening.

proach will satisfy the overall system hardening requirement and preclude the susceptibility of the system to the specified environment.

The program manager's review of the documented hardening design may lead to a decision requiring additional system hardening or additional verification of the design. Approval of the hardening design would lead to system tests of a prototype system and prototype acceptance if the test results proved satisfactory. Unsatisfactory results could lead to a design review by the program manager and EMCAB and a repeat of the hardening design cycle. This review could also include tradeoff analyses to assess the cost effectiveness of alternate approaches.

5.4 ESTABLISHING EMR HARDNESS REQUIREMENTS.

5.4.1 OVERALL APPROACH. The specific requirements for hardening an air launched ordnance system will vary from system to system, but the overall approach to establishing hardness requirements are the same for all systems. As shown in Figure 5-1, the system designer first formulates a system design concept based on the functional and environmental requirements provided under the contract. This preliminary design will form the basis for establishing environment-to-component coupling levels and component susceptibility thresholds. Using the system design concept and the environment definition, the environment-to-component coupling characteristics are then determined over the frequency range of concern. This step provides an estimate of the interference power level incident on system components and circuitry. Finally, the susceptibility thresholds of system components and circuitry are determined and compared with the incident interference power level to provide an assessment of system susceptibility. Hardening requirements are then established which will preclude system susceptibility in the specified environment.

It is important to recognize that the establishment of EMR hardness requirements is an iterative process. During the early stages of system design, detailed circuit and subsystem configurations will not have been defined. Thus hardness requirements must initially be derived from coupling and susceptibility analyses performed on a preliminary system design concept. From these preliminary analyses, an assessment of system susceptibility can be made and initial guidelines for hardening design can be established. As the system design progresses and hardware and circuitry becomes better defined, the coupling and susceptibility analyses should be continually refined and updated and the hardening requirements modified to reflect any changes. Also, as hardware becomes available, measurements should be performed as necessary to supplement the analyses or verify hardening techniques.

5.4.2 EMR ENVIRONMENT FORECASTS. The electromagnetic environment in which a system must operate must be defined before the system hardening design can be initiated. The environment definition, to be provided under the contract as a system design requirement, will characterize the environment as power level versus frequency profiles spanning the frequency range from 1 MHz to 100 GHz. Profiles for both peak power levels and average power levels will be provided to permit the designer to assess potential environmental effects on both a peak power and an average power basis. In addition, the modulation characteristics of pulsed signals will be provided to alert the designer to the most probable pulse repetition frequencies and pulse widths which are likely to be encountered.

5.4.3 **ENVIRONMENT-TO-COMPONENT COUPLING.** The term environment-to-component coupling is defined as the total process by which electromagnetic energy incident upon an air launched ordnance system is transferred to internal circuits and components. The process by which this energy transfer occurs is highly complex, involving several coupling mechanisms and numerous coupling paths. Several analytical techniques, ranging from simple approximations to complex computer codes, have been developed which permit the prediction of coupling levels. The more sophisticated techniques are generally applicable only to selected physical and electrical configurations.

At the early stages of system development, with no hardware defined, the system designer must resort to simple approximations to obtain an assessment of environment-to-component coupling. One approach is to assume that no shielding exists between the system components and the external environment, and that the effective aperture of all cables connected to internal components is that of a tuned half-wavelength dipole antenna which is matched to the impedance of the component. Using this effective-aperture model in conjunction with the defined interference environment, the power impinging on the components is the product of the incident field (in power density units) times the effective area of a half-wavelength dipole antenna:

$$P_c = P_d \times A_e \qquad (5\text{-}1)$$

where

P_c = power impinging on system component,

P_d = power density of the interference environment, and

A_e = effective aperture of 1/2 wavelength dipole antenna.

Since $A_e = 0.13\lambda^2$, Equation 5-1 may be written as

$$P_c = 0.13\lambda^2 P_d. \qquad (5\text{-}2)$$

Because the actual amount of energy which may be coupled to a cable is dependent upon such variables as aspect angle, termination impedance, and frequency, the predicted coupling using the half-wavelength dipole model has generally proven to be greater than measured values. Equation 5-2 thus represents the maximum power that will be coupled to a component from a specified environment, and provides a worst-case estimate of environment-to-component coupling. By comparing the results of Equation 5-2 with the susceptibility threshold of the most sensitive component envisioned for the system, the susceptibility of the system can be determined and the hardening requirements defined.

An expanded version of this coupling approach can be used to evaluate the effectiveness of hardening techniques applied during system development. This evaluation is accomplished by using Equation 5-3 to establish the relationship between the interference environment and the power delivered to the system components when equipment enclosure shielding and/or cable shielding are present.

$$P_c = P_d \times A_e \times S_{cable} \times S_{enclosure} \qquad (5\text{-}3)$$

where:

P_c = power delivered to equipment component,
P_d = power density of the interference environment,
A_e = effective aperture of unshielded cable,
S_{cable} = shielding effectiveness of cable shield, and
$S_{enclosure}$ = shielding effectiveness of enclosure.

The half-wavelength dipole antenna model is used for the effective aperture of the pick up cables. The shielding effectiveness values for the cables and the equipment enclosure are the values measured by the MIL-STD-1377 method. The calculated values of power impinging upon susceptible components are compared with the susceptibility levels of the components to determine if the realized shielding is adequate to protect the sensitive components.

As more detailed design information becomes available, the accuracy of environment-to-component coupling analyses may be improved through the use of more detailed computer-aided coupling models. The system designer may employ his own models or request analysis support through RADC. RADC has available the Intrasystem Analysis Program which provides both order-of-magnitude analyses as well as models for performing detailed calculations as described in Appendix D.

5.4.4 **COMPONENT SUSCEPTIBILITY THRESHOLDS.** A definition of component susceptibility thresholds is necessary to determine if the interference power coupled to a component will adversely affect its operation. There are three ways by which the component susceptibility information can be obtained. If the component is one of the types for which measured data are presented in Appendix E, the information can be obtained directly from the data presented, or if the component is similar to measured types, the information can be extrapolated from the measured results. If susceptibility data on a particular component type or similar components are not available, susceptibility measurements can be performed utilizing the measurement techniques described in Appendix E. The third possible approach to obtaining component susceptibility data is through the use of analytical models to predict component susceptibility characteristics. Analytical models based upon the physical mechanisms of interference are under development but at the present time are not complete. Thus, until validated analytical models become available, measured data will be required to define the susceptibility thresholds of discrete and integrated circuit components.

The designer should recognize that, in general, the susceptibility threshold of the system will be set by the susceptibility threshold of the most sensitive (to interference) system component. Thus, it usually will not be necessary to define susceptibility thresholds for all system components. It is also important to recognize that the susceptibility threshold for a particular component will depend upon such factors as the function of the component in the system, its bias conditions, its input signal level, its output signal level, and the modulation characteristics of the interference signal. Thus, the susceptibility threshold for a particular component may vary significantly within a system where the component is used for different functions or under different operating conditions.

5.4.5 **REQUIRED HARDENING LEVEL.** The level to which a system must be hardened is determined by first performing a system susceptibility assessment. This assessment is accomplished by comparing the interference power level coupled to system components with the susceptibility threshold of the most sensitive component to be utilized in the system. After the susceptibility level has been determined, the required hardening level is simply the ratio of the predicted maximum interference power level impinging on the most sensitive system component to the threshold susceptibility level of that component. If the interference power level is less than the threshold level of the most sensitive component, the hardening requirement is less than 0 dB (negative dB) and no hardening is required. If the interference power level is greater than the threshold level of the most sensitive component, the hardening requirement will be greater than 0 dB and the designer is required to develop a hardening approach. The selection of a hardening approach will be influenced by the magnitude of the hardening requirement.

5.5 **HARDENING APPROACH.** The requirements established for hardening an air launched ordnance system will lead to a definition of the level of hardening which must be achieved to prevent system vulnerability to incident EM energy. Once defined, the hardening level becomes a system design requirement, to be met through the application of appropriate hardening techniques and devices. The most effective approach to achieving this design requirement is that of layered hardening.

The layered hardening concept involves the layered application of hardening techniques and devices along the exterior-to-interior coupling paths of EM energy. For example, EM energy will first couple to the exterior of a system and set up skin currents and charge densities. These currents and charges excite penetrations such as antennas (real and virtual) and apertures, thus permitting the EM energy to penetrate to the system interior where it can couple to cables and wires leading to sensitive circuits and components. The layered hardening approach attempts to interrupt these coupling paths by first hardening the system exterior to reduce penetration. Next, the coupling to interior cables is reduced, and finally, the critical circuits and components are hardened. Various hardening techniques and devices are applied as necessary at each layer in the hardening process. Hardening techniques and devices are described in Appendix G.

Layered hardening represents a cost effective method of protecting a system for two reasons. One reason is that it takes advantage of hardening which is intrinsic to the system design, e.g., the shielding effectiveness intrinsic to the system exterior. The other reason is that it does not place the burden of achieving the total hardness on any one element or layer of the system. Hardening can be successively applied to various system elements until the required hardness level is realized.

Although the actual implementation of hardening techniques under the layered hardening approach typically follows an exterior-to-interior path, it is not to be construed that such a path is mandatory. The concept of layered hardening is concerned more with the application of hardening in layers rather than the order in which these layers are applied.

5.6 HARDENING TECHNIQUES AND DEVICES.

5.6.1 **GENERAL.** A number of techniques and devices are available for use in reducing the susceptibility of electronic systems to impinging electromagnetic fields. The principal techniques for EMR hardening are shielding, bonding, filtering, grounding, circuit design, and component selection. This section is intended to provide a general overview of these design approaches. A more detailed discussion of these hardening techniques is presented in Appendix G.

5.6.2 **SHIELDING.** Shielding is the establishment of an electromagnetic barrier between two regions. Shielding is the most direct method, and in many cases the most cost effective method, for protecting the circuits of a system from the EMR environment. Shielding has two main purposes: (a) to prevent radiated EM energy from entering a specific region; and (b) to keep radiated EM energy confined within a specific region. For the purposes of this handbook, the primary emphasis will be on shielding the interior of a system from the external EMR environment. The shield 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 shielding effectiveness levels.

5.6.3 **BONDING.** Bonding is the establishment of a low impedance path between two metal surfaces. This path may be between a ground reference and a component, circuit, shield, or structural element. The purpose of bonding is to establish an electrically homogeneous structure to prevent the development of electrical potentials between individual metal surfaces which can cause interference. Good bonding within a system is essential to minimizing interference.

5.6.4 **GROUNDING.** Grounding is the establishment of electrically conducting paths between selected points in a system and some common reference plane. The reference plane may be the system skin or a chassis or ground plane that may or may not be isolated from the system skin. An ideal grounding system would result in a system with a common potential reference point everywhere in the system so that no undesired potentials would exist between any two points in the system. However, because of the physical properties and electrical characteristics of grounding materials, no grounding system is ideal and some potential always exists between ground points within a system.

The extent to which potentials in the ground system are minimized and ground currents are reduced determines the effectiveness of the ground system. A poor ground system will make it possible for spurious voltages and currents to couple into circuits and subassemblies and can: (a) degrade the shielding effectiveness of well-shielded units; (b) bypass the advantages of filters; and (c) result in interference problems which are difficult to isolate and resolve.

5.6.5 **FILTERING.** Filters are devices which pass conducted energy over specified frequency ranges and reject or absorb conducted energy over other specified frequency ranges. Thus, a filter placed in-line with a wire or cable can be used to pass desired signals while reflecting or absorbing undesired signals outside the passband of the filter. Filters may be included in circuit designs and/or interconnecting wiring designs to prevent interfering signals from being conducted through the system circuits. In addition, filters may be inserted in wires and cables penetrating a shielded enclosure to maintain the integrity of the shielding effectiveness of the enclosure.

5.6.6 **COMPONENT SELECTION.** Minimizing the EMR susceptibility of a system should begin with the selection of the components and devices to be utilized in the system design. As wide a range as possible of components and devices with acceptable performance characteristics should be screened in order to obtain components with the highest possible interference susceptibility thresholds. A judicious choice of components can result in 10 to 30 dB of additional hardness in a system.

5.6.7 **CIRCUIT DESIGN.** In the selection of signal and impedance levels for the circuit designs, the designer should recognize that circuits operating with high signal levels and low impedance levels are less susceptible to interference. The susceptibility of circuits to radiated interference can be reduced by minimizing the length of interconnecting wiring between comonents and circuits and the use of shielded and twisted-pair wire for these interconnections. In general, digital circuits are less susceptible than linear circuits and low-speed digital circuits are less susceptible than high-speed circuits.

5.7 **DESIGN TRADEOFFS.** Design tradeoffs may be necessary to ensure compatibility or to resolve conflicts between the functional requirements and the hardening requirements of an air launched ordnance system. Also, tradeoffs may be necessary to achieve a cost effective hardening design. Such tradeoffs should be directed to design techniques which will permit compatibility of functional and hardening requirements to be realized. Examples of such techniques are:

a. Selection of system operational signal levels as high as practical commensurate with device and circuit requirements.

b. Selection of interconnecting wiring and cabling techniques which provide the best rejection of normal mode and common mode energy transfer.

c. Use of fiber optic guides versus conventional shielded cable.

d. Use of rigid or flexible solid shielding versus single or double insulated metallic braid.

e. Multiple utilization of load bearing structures such as air frames, cable raceways, and conduit to satisfy both functional and hardening requirements at relatively low cost.

f. Use of enclosure shielding and cable filtering versus internal cable and circuit hardening.

Any tradeoffs which involve a change in functional, hardening, or cost objectives are subject to the review and approval of the program manager. The course of action for the resolution of conflicts between hardening requirements and other system requirements will depend upon such factors as:

a. The impact of the tradeoff on system susceptibility and system functional performance.

b. The number of equipments, subsystems, and systems involved.

c. The impact on program cost and schedule.

5.8 **HARDENING VERIFICATION.** Testing of a complete prototype system is necessary to validate a hardening design which incorporates a combination of hardening techniques. However, prior to full system tests, the system designer can verify with reasonable assurance that his hardening design is sufficient to assure the satisfactory operation of the system in its operational environment. The hardening design can be verified through the use of data, analysis, measurements, or a combination of these three approaches. The most efficient approach to verifying the hardening design is to determine the effectiveness of the individual hardening techniques utilized in the overall design as the system design progresses. For example, the shielding effectiveness of the system or subsystem enclosures can be measured while the system circuitry is still under development, and any design changes required to realize the desired shielding can be incorporated into the enclosure with minimum effect on other design efforts. Also, component or circuit hardening can be accomplished independent of enclosure design efforts. Whatever approach is employed, a reliable estimate of the overall system hardness can be obtained simply by adding the levels of hardening achieved at each system layer. If the estimate of system hardness determined in this manner exceeds the design requirement, the system designer can be reasonably confident that the system tests will be satisfactory. An estimate which falls below the design requirement is an indication that additional hardening is required.

Once the hardening design is verified, a complete documentation of the design should be submitted for the review and approval of the program manager and the EMCAB. The documentation should describe in detail the hardening techniques and devices employed, the approach used in verifying the design, and supporting data which substantiate that the design requirements have been met.

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SECTION 6 - EMR HARDNESS MEASUREMENT PROGRAM

6.1 **GENERAL.** The program manager and the EMCAB should establish an overall EMR hardness measurement program. The measurement program plan should define all testing and evaluation efforts required to demonstrate compliance with the EMR hardness requirements and to ensure that the developed system is compatible with its anticipated operational EMR environment. The program manager and the EMCAB should also ensure that adequate budgets, time schedules, and resources (facilities) are allocated to accommodate the required testing and evaluation efforts.

6.2 **TEST AND EVALUATION MASTER PLAN (TEMP).** The TEMP is the controlling management document which defines all test and evaluation efforts to be accomplished in connection with a system acquisition. The program manager and EMCAB should ensure that all EMR hardness test and evaluation requirements are included in the TEMP which is prepared early in the program. These requirements should include the T&E efforts to be performed by other DoD organizations, as well as the T&E efforts to be performed by the contractor. The TEMP should be updated periodically to incorporate significant results achieved and any changes in plans and milestones.

6.3 **EMR HARDNESS TEST PLAN.** The contract should require the contractor to submit an EMR hardness test plan to the procuring activity for approval. The procuring activity may invoke specific tests on the contractor by tailoring existing EMC specifications and standards (such as MIL-E-6051, MIL-STD-461, MIL-STD-462, etc.) and making the tailored specifications a part of the contract. The test plan should describe, in the maximum detail possible, what tests the contractor plans to perform to demonstrate compliance with the EMR hardness requirements, how the tests will be conducted, and what types of data will be submitted as a result of these tests. The test plan should include a description of the tests that subcontractors will be required to perform on subsystems and components and how the subcontractor test results will be utilized in establishing system EMR hardness compliance. The test plan should be prepared in accordance with the outline presented in Appendix M.

6.4 **EMR HARDNESS TEST REPORT.** The contract should require the contractor to submit the results from all EMR hardness tests in an EMR hardness test report to the procuring activity for approval. The Data Item Description (DID) requiring the test report in the contract should ensure that the format of the test report is such that the data will be submitted in the most usable form for evaluation and subsequent analyses. Before approving the EMR hardness report, the program manager should be satisfied that the contractor has demonstrated his compliance with the EMR hardness requirements and that the system is ready for a system susceptibility/vulnerability analysis at RADC.

6.5 **SYSTEM SUSCEPTIBILITY AND VULNERABILITY ANALYSIS.** System susceptibility testing and vulnerability analyses of a pre-production prototype model and a production model of the system will be performed at the RADC Electromagnetic Compatibility Analysis Facility (EMCAF) by the Compatibility Branch personnel at the Rome Air Development Center. While neither program office personnel nor contractor personnel will be required to perform these tests and analyses, a brief description of the procedures and facilities involved are presented in this section to make the system developers aware of the degree of thoroughness with which a system will be evaluated before final acceptance.

The Electromagnetic Compatibility Analysis Facility is a dedicated, unique facility for testing Air Force systems in high power RF environments. The facility provides a capability to test weapon systems for susceptibility to radiated RF energy in a simulated free-space environment. There are three RF anechoic chambers within the facility. The characteristics of the three chambers are as follows:

CHAMBER NUMBER ONE

Size: 32 Ft. High x 40 Ft. Wide x 48 Ft. Long
Quiet Zone: 12 Ft. x 12 Ft. x 20 Ft. Long
Frequency Range: 50 MHz - 40 GHz
Maximum Sample System Size: 20 Ft. Long x 8 Ft. Diameter
Maximum Sample Weight: 4000 Pounds
Shielding: 100 dB
Shielded Instrumentation Room: 20 Ft. x 20 Ft. x 12 Ft.

CHAMBER NUMBER TWO

Size: 12 Ft. High x 12 Ft. Wide x 36 Ft. Long
Quiet Zone: 3 Ft. Diameter x 20 Ft. Long Cylinder
Frequency Range: 200 MHz - 40 GHz
Maximum Sample Size: 6 Ft. Long x 3 Ft. Diameter
Maximum Sample Weight: 1000 Pounds
Shielding: 100 dB

CHAMBER NUMBER THREE

Size: 18 Ft. High x 18 Ft. Wide x 55 Ft. Long (Tapered)
Quiet Zone: 6 Ft. Diameter x 20 Ft. Long Cylinder
Frequency Range: 200 MHz - 40 GHz
Maximum Sample Size: 14 Ft. Long x 6 Ft. Diameter
Maximum Sample Weight: 4000 Pounds
Unshielded

The facility has a group of wideband RF sources capable of generating high power RF signals in the 10 kHz to 40 GHz frequency range with a variety of modulation characteristics. In conjunction with these RF sources, the facility also maintains the necessary associated equipment such as antenna systems, transmission lines, and automated instrumentation and control systems to establish high intensity RF fields within the anechoic chambers. In addition to dedicated minicomputers, facility personnel have access to the RADC central computer and maintain extensive programs for use in test data processing and system vulnerability analysis.

When a system to be tested arrives at the facility, it is set up to operate in a typical operating mode in one of the anechoic chambers. Initial tests are performed with no EMR environment present to measure normal functional parameters, some of which are used to establish the criteria for determining system degradation. A "standard change" is established in each performance parameter used in the criteria. A change equal to or greater than a "standard change" in any one of these performance parameters is defined as constituting system degradation.

After the degradation criteria have been established, tests are performed with EMR fields incident on the system to determine the levels of fields necessary to cause degradation (susceptibility levels). In addition to frequency and power level, these tests take into account the effects of modulation, polarization, and aspect angle of the EMR field. The susceptibility levels established by the tests are compared with the corresponding levels in the EMR environment profile for the system. If all susceptibility levels are higher than the corresponding environment levels, the system is not vulnerable to the EMR environment. If any susceptibility levels are lower than the corresponding environment levels, the system is vulnerable, and an analysis is performed to determine the impact of the vulnerability on the system operation. If the results from the analysis indicate that the vulnerability will have a significant impact on the deployment of the system, a fix, modification, or redesign of the system may be necessary to eliminate the vulnerability.

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Preparing Activity
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APPENDIX A

EMR ENVIRONMENT

10. **INTRODUCTION.** The electromagnetic environment incident on an air launched ordnance system results from the electromagnetic radiation (EMR) from a number of radiating sources in the vicinity of the system. The radiating sources may include friendly and/or hostile radars, radio transmitters, jammers, and other relatively high power radiating systems. These radiating sources may be located on the ground, on ground vehicles, on ships, or on aircraft. The spectral distribution of the radiated energy in the electromagnetic environment (EME) is determined by the operating frequencies of the radiating sources. The levels of the EME are determined by the amount of power being radiated by the individual sources, the distance between the weapon system and the individual radiating sources, and the orientation of the weapon system relative to the radiation pattern of the individual sources.

While the system is mounted on a delivery aircraft, the radiation from electronic systems (such as radars, jammer pods, and communications transmitters) on board the delivery aircraft will be a primary source of the EME incident on the weapon system. If the delivery aircraft is flying in a formation, the radiation from escort aircraft may be a primary source of the EME incident on the weapon system. While the system is in free flight between the delivery aircraft and a target, the radiation from the ground environment may be a primary source of the EME. As the system approaches a target, the radiation from the target itself and emitters in the vicinity of the target will probably be the major source of the EME.

20. REQUIREMENTS FOR EMR ENVIRONMENT FORECASTS.

20.1 **BASIS FOR EMR HARDNESS REQUIREMENTS.** In order to design and develop a system which is hardened to survive and operate in its electromagnetic operational/threat environments, it is first necessary to define the electromagnetic environments the system will be exposed to during its deployment. Only after the EMR environments have been defined can the development of realistic system EMR hardness requirements begin to be addressed.

20.2 **FORECASTING.** In defining the EMR environments a system will be exposed to during deployment, consideration must be given to the fact that the system will probably be in development for several years, and in addition, the system will probably have an in-service life spanning several years. Hence, it is not sufficient to define the EMR environments that exist at the present time; it is necessary to predict the EMR environments the system will be exposed to during deployment until the end of its in-service life. For EMR environment definitions used in the early stages of a system development, the environment definitions will have to be projected by forecasting techniques to a time frame which is the sum of the acquisition cycle and the in-service life cycle of the system in the future. As the development of the system progresses the projection time for the updated EMR environment definitions will be reduced. If it is assumed that the acquisition time of a typical system is 10 years and the average in-service life is 15 years, the initial EMR environment definitions would need to be projected by a 25-year forecast. Environment definitions used midway in the acquisition cycle would need to be projected by 20 years, while the environments used at the end of the acquisition cycle would need to be projected 15 years.

20.3 **INTEGRATED ENVIRONMENT.** As mentioned previously, an air launched ordnance system will be exposed to several different EMR environments during its life cycle. Thus, to define the EMR environment a system is to be hardened to, it is necessary to define each of the independent environments the system will be exposed to. These individual environments include the ground environment, the cosite environment, the intersite environment, and the approach-to-target environment. The ground environment includes the radiation from all emitters (friendly and hostile) on the ground (and on water) over which the system will travel (both on the delivery aircraft and in free flight) in the performance of its missions. The cosite environment includes the radiation from emitters on-board the delivery aircraft. The intersite environment includes the radiation from emitters on escort aircraft, other friendly aircraft, and hostile aircraft. The approach-to-target environment includes the radiation from the target and emitters in the vicinity of the target. After the individual environments have been defined, each of the environments must be projected by forecasting techniques to be representative of the environments the system will be exposed to at the end of its service life.

In order to establish the overall EMR hardness requirements for a system, it is necessary to integrate the forecasted individual environments into a composite EMR environment profile which indicates the maximum radiation levels the system will be exposed to during its life cycle.

20.4 **POINTS IN ACQUISITION CYCLE WHERE EMR FORECASTS ARE REQUIRED.** A minimum of three EMR environment forecasts is considered necessary during the acquisition cycle of a system. The actual number of forecasts required will depend on the complexity of the system being developed and the time duration of the acquisition cycle. For a large complex system requiring an unusually long acquisition time frame, several EMR forecasts may be necessary.

The first EMR environment forecast (Type I) should be available for use during the evaluation of the required operational capabilities necessary to satisfy the mission need statement in the initial phase of the program. This forecast should be used in feasibility analyses, trade-off studies of alternate approaches and concepts, and the definition of risks.

The requirements of OMB Circular A-109 have increased the emphasis on investigating alternate concepts for satisfying mission needs. Under these conditions, the requirement to define the EMR environment early in the program to ensure that the concepts investigated are compatible with the operational EMR environment becomes more critical.

This initial EMR forecast should also be used by the program manager and the EMCAB to establish budgets, schedules, and resources requirements for inclusion in the EMR Hardness Program Plan and the Test and Evaluation Master Plan.

A second EMR environment forecast (Type II) should be available for use during the preparation of the RFP documentation. This EMR forecast should be incorporated into the RFP to convey to the bidders the amount of EMR hardness that will be required. The forecast should also be used to tailor the limits of any EMC specifications and standards (such as MIL-E-6051, MIL-STD-461, MIL-STD-462 and MIL-STD-463) which will be invoked by the contract. This forecast should also be used by the selected contractor to establish the EMR hardness criteria to be used as the basis for the EMR Hardness Control Plan and the EMR Hardness Test Plan.

A third EMR environment forecast (Type III) should be available for use during the system susceptibility/vulnerability evaluation phase. This forecast should be used by the system test organization as guidance in conducting the system susceptibility tests and as the threat definition in vulnerability analyses.

30. **OBTAINING EMR ENVIRONMENT FORECASTS.** From the previous discussions in this appendix, it is apparent that the generation of an EMR operational/threat environment forecast for a proposed air launched ordnance system is a complex and difficult process requiring extensive information describing the operational characteristics and geographical locations of friendly and hostile emitters world wide. The generation of EMR forecasts also requires extensive culling and processing of the emitter data and the application of validated forecasting techniques.

While the operational EMR environment forecasts for each system must be tailored for that specific system, the procedures and forecasting techniques for generating the EMR forecasts will be essentially the same for all systems. In addition, the generation of an EMR forecast for a specific system will entail processing the same EMR data base to tailor an environment to the operational and tactical requirements of the specific system.

Under these conditions, it is not feasible to require each program office to generate the EMR forecasts for the system for which it is responsible. This would require each program office to establish an extensive EMR data base and develop an organization, procedures, and forecasting techniques for generating EMR forecasts. A much more efficient approach is to assign a permanent organization the responsibility for satisfying the EMR environment forecast requirements for all program offices. With this approach, the basic EMR data base, the processing procedures, and the forecasting techniques are only developed one time, eliminating duplication of efforts.

30.1 **PREPARING ORGANIZATION.** The Electromagnetic Compatibility Analysis Center (ECAC) located at Annapolis, Maryland has been designated as the organization responsible for satisfying the EMR environment forecast requirements for Air Force air launched ordnance systems. This center is a joint-service Department of Defense facility, established to provide rapid analysis of electromagnetic compatibility problems of the military services. ECAC has an extensive electromagnetic environmental data base which includes a comprehensive listing of existing electromagnetic emitters throughout the world. In addition, ECAC has access to the required intelligence information and has extensive experience as to which organizations are the best sources of particular types of information. ECAC also has considerable experience in generating electromagnetic environmental profiles, similar to those of current interest, in support of RADC's HAVE NOTE test programs. The capabilities developed by ECAC for generating EMR environment forecasts are described in Appendix B.

30.2 **REQUESTING EMR ENVIRONMENT FORECASTS.** Contact should be established with ECAC at the initiation of the program, and they should be alerted that EMR environment forecasts will be required. If possible, the number of EMR forecasts that will be required and the approximate dates they will be required should be defined. Several types of operational and tactical information will be required for ECAC to generate an EMR forecast. To aid the requesting organization in providing this information, ECAC has prepared a data requirements questionnaire for requesting an EMR environment forecast. This questionnaire is described in Appendix B.

30.3 **TYPES OF EMR ENVIRONMENT FORECASTS.** There are three basic types of EMR environment forecasts. While the objective of all EMR forecasts is the same (i.e., to define the anticipated operational electromagnetic environment of the system), the manner in which each forecast will be used, the stage of development of the system when each forecast is generated, and the extended time periods between forecasts dictate that the EMR forecasts obtained at different points in the acquisition cycle of a system will be different.

Type I EMR Environment Forcast

The initial EMR forecast, obtained at the beginning of the program, will be based on a Type I analysis. It is anticipated that a great deal of the operational and tactical requirements, as well as the performance specifications, for the system will not have been defined at this stage of the program, and the environmental analysis will have to incorporate a number of assumptions. For example, if the theaters of operation have not been defined, then a worldwide environment will be considered in the analysis. If the type(s) of delivery aircraft has not been defined, then the worst-case, on-board environment for that class of aircraft will be considered in the analysis, etc.

A Type I analysis will provide baseline (current) environment profiles and forecasted profiles valid for up to 30 years in the future.

Type II EMR Environment Forecast

The EMR forecast obtained at the RFP preparation stage will be based on a Type II analysis. It is assumed that the operational/tactical requirements and performance specifications for the system have been defined so that adequate input information can be provided to closely approximate the actual EMR environment.

A Type II analysis will provide baseline profiles and forecasted profiles valid for up to 25 years in the future. Figures A-1 and A-2 show examples of Type II baseline profiles for a theater ground environment.

Type III EMR Environment Forecast

The EMR forecast obtained at the system EMR susceptibility/vulnerability evaluation stage will be based on a Type III analysis. It is not anticipated that the levels of the Type III forecasts will be significantly different from the levels of the Type II forecasts. However, the Type III analysis is specifically designed to provide forecasts which will provide maximum assistance in performing system susceptibility testing and vulnerability assessments. The environments for each theater will be provided in zonal increments, so that it is possible to generate environment profiles for specific mission scenarios. The Type III forecasted profiles will be valid for up to 15 years.

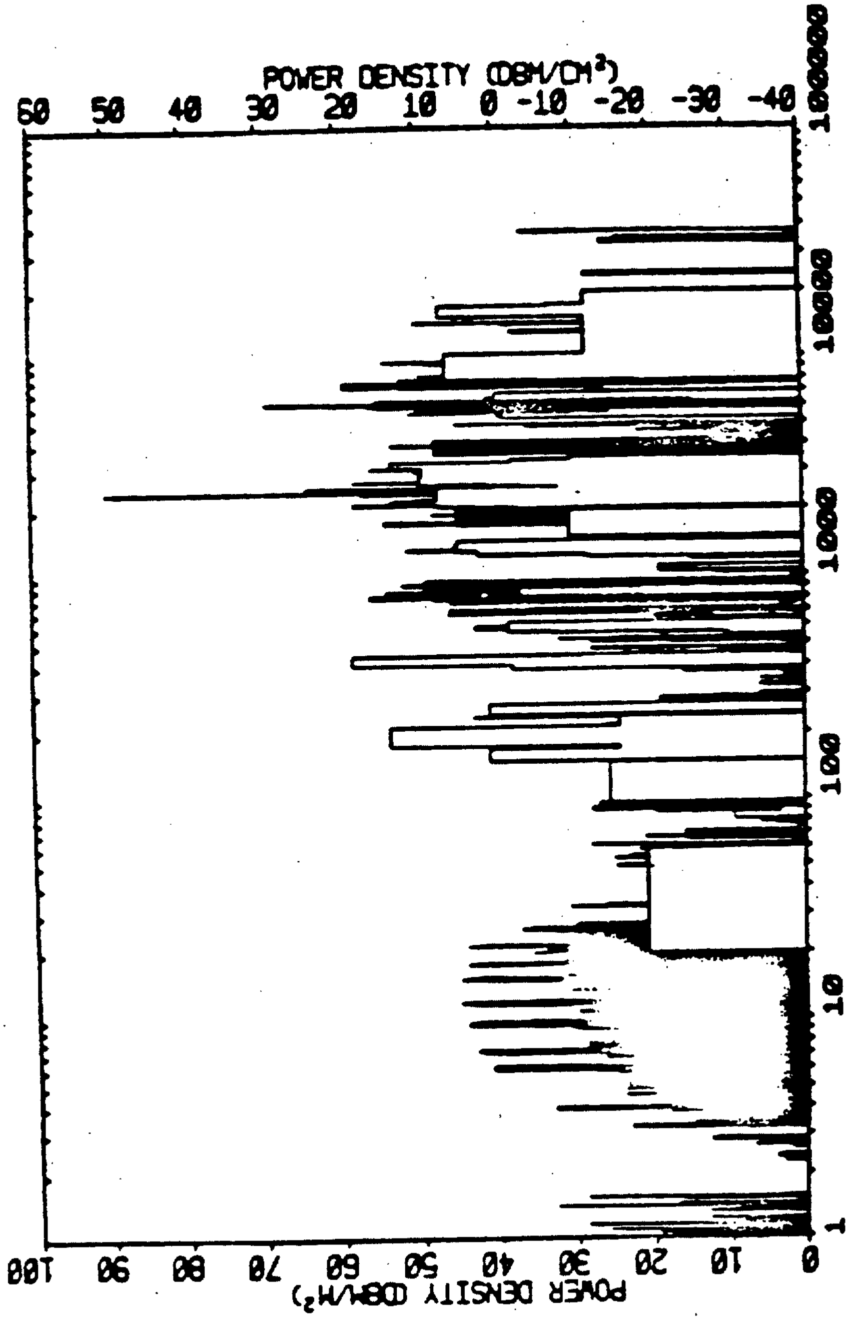

Figure A-1. Baseline, Theater, Average Power, Ground Environment Profile.

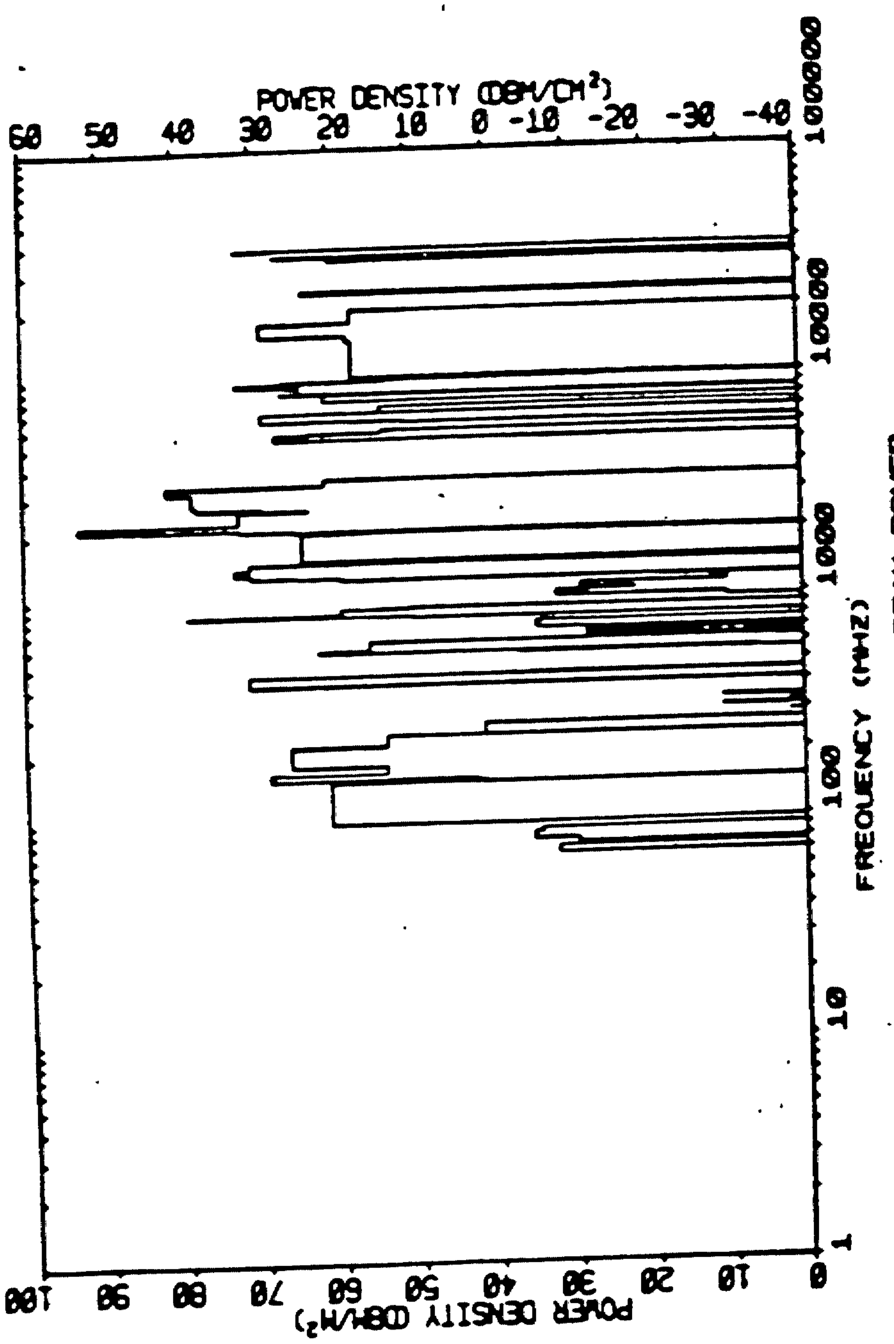

Figure A-2. Baseline, Theater, Peak Power, Ground Environment Profile.

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APPENDIX B

EMR ENVIRONMENT FORECASTING CAPABILITIES AT ECAC

10. **INTRODUCTION.** The U. S. Air Force Rome Air Development Center (RADC) has sponsored a program at the Department of Defense Electromagnetic Compatibility Analysis Center (ECAC) to develop a capability to rapidly produce current and future EMR environment "profiles" that are tailored to both the acquisition and deployment stages of the life cycle of specific weapon systems. This capability is intended to support the Air Force EMR Hardness Program.

20. **ENVIRONMENT DEFINITION SYSTEM.** The capability which ECAC has developed to accomplish this objective has been designated the Environmental Definition System (EDS). The EDS consists of two parts. The first part, called the Ground Environment Definition System (GEDS), was designed to produce current (baseline) and future (forecast) EMR ground environment profiles for the delivery phase of an air launched ordnance system. The second part, called the Aircraft Environment Definition System (AEDS), was designed to produce current (baseline) and future (forecast) EMR environment profiles for the aircraft cosite, aircraft intersite, immediately-after-launch, and approach-to-aircraft target phases for the delivery sequence of a weapon system.

ECAC is currently preparing a five-volume final report describing the Environmental Definition System. The titles and report numbers of the five volumes of the report are listed below.

- "Environmental Definition System (EDS) Volume 1: Ground Environment Definition System," ESD-TR-80-100 Vol. 1.
- "Environmental Definition System (EDS) Volume 2: Ground Environment Forecasting," ESD-TR-80-100 Vol. 2.
- "Environmental Definition System (EDS) Volume 3: Aircraft Environment Definition Sytem," ESD-TR-80-100 Vol. 3.
- "Environmental Definition System (EDS) Volume 4: Aircraft Environment Forecasting," ESD-TR-80-100 Vol. 4.
- "Environmental Definition System (EDS) Volume 5: Customer's Application Manual," ESD-TR-80-100 Vol. 5.

Volume 1 of the report describes the overall philosophy used to develop the Environmental Definition System and also describes the Ground Environment Definition System which is used to generate ground EMR environment baseline profiles. Volume 2 describes the forecasting techniques developed to predict the ground EMR environment conditions that will exist at specified times in the future. Volume 3 describes the Aircraft Environment Definition System which is used to generate aircraft EMR environment baseline profiles. Volume 4 describes the forecasting techniques developed to predict the aircraft environment conditions that will exist at specified times in the future. Volume 5 (Customer's Application Manual) describes in detail how an Air Force

Agency can request EMR environment forecasts for an existing or proposed weapon system. This volume contains specific information on input information requirements, available outputs, time and manpower estimates, and a glossary of special terms.

30. **OBTAINING ECAC's SERVICES.** To obtain ECAC's services for generating EMR environment forecasts, contact should be established with the Air Force Deputy Office at ECAC. The telephone numbers are autovon 281-2613 and commercial (301) 267-2613. The mailing address is the following:

Electromagnetic Compatibility Analysis Center
North Severn
Annapolis, Maryland 21402
Attn: CF/Air Force Deputy Office

Specific input information is required for ECAC to generate EMR environment forecasts. ECAC has developed DoD/ECAC Form EDS-1 to assist a requesting agency in providing the necessary information in the most usable format. To illustrate the type of information required, this form is shown in Table B-1.

TABLE B-1

EMR ENVIRONMENT FORECAST REQUEST FORM

EDS FORM-1

REQUEST FOR A TAILORED EMR ENVIRONMENT

TYPE I, II, IIIA

This form shall be considered unclassified until such time as it is filled out. The individual filling out this form shall determine the proper classification and down grading and shall mark this form accordingly. In addition, a copy of the current project security guide must be forwarded to ECAC. This form and the security guide shall be sent to:

Electromagnetic Compatibility Analysis Center
North Severn
Annapolis, MD 21402

In addition, the inside label shall have on it:

Attn: CF/Air Force Deputy Office

BACKGROUND INFORMATION

1. In order to obtain a tailored EMR environment from Dod/ECAC, several items of information are required as inputs. As an aid to the user of ECAC services the required information is being solicited by this questionnaire.

2. Section 3 of the Customer's Application Manual[a] contains information to aid in preparing this questionnaire. In addition, Section 2 contains the approximate cost and time required to do this analysis.

[a] ESD-TR-80-100-Vol 5.

(Continued)

TABLE B-1 (Continued)

EMR ENVIRONMENT FORECAST REQUEST FORM

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3. Unless special arrangements are made, the data base file selects for any analysis will be kept no longer than one year from the date of project completion.

B. Questionnaire

1. Administrative Information:

   Specify the actual engineering office, SOP, development lab or other agency responsible for this request. Include the office symbol(s) and points of contact with autovon phone numbers.

   ____

   ____

   ____

   CLAS( ), DOWNGRADING & AUTHORITY

2. Describe the class of weapon system for which a tailored EMR environment is required (i.e., air-to-air missile, air-to-surface missile, laser guided bomb, etc.).

   CLAS( ), DOWNGRADING & AUTHORITY

3. Unless requested otherwise, the EMR environment forecast will be based on a stockpile-to-target sequence (S-T-S) from the time the system is attached to the delivery aircraft

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(Continued)

TABLE B-1 (Continued)

EMR ENVIRONMENT FORECAST REQUEST FORM

until it impacts a target. If a different phase of the S-T-S is also desired, note that fact here.

CLAS( ), DOWNGRADING & AUTHORITY

4. Indicate the type of analysis required (check only one).

Type I ____ Type II ____ Type IIIA ____

CLAS( ), DOWNGRADING & AUTHORITY

5. Specify the anticipated delivery platforms (i.e., F-4J, A-7D, RF-4E, etc.). If unknown, so state.

CLAS( ), DOWNGRADING & AUTHORITY

6. Specify the anticipated aircraft that will be in formation with the aircraft carrying the weapon system. Also, if known, state the specific formations that will be used.

CLAS( ), DOWNGRADING & AUTHORITY

7. Specify the minimum altitude <u>above ground level</u> for the weapon delivery platform during the delivery phase. (Note: this altitude <u>must</u> be greater than or equal to 200 feet.)

CLAS( ), DOWNGRADING & AUTHORITY

(Continued)

TABLE B-1 (Continued)

EMR ENVIRONMENT FORECAST REQUEST FORM

8. Specify the lowest power density which, when incident on the weapon system, will cause harmful effects. The minimum value must be at least 0 dBm/m^2.

CLAS( ), DOWNGRADING & AUTHORITY

9. Specify the worst probable level for failure that you can tolerate for your system, assuming that all design goals are met and the system will fail only due to EMR.

CLAS( ), DOWNGRADING & AUTHORITY

10. Specify the probable radiating target classes for the weapon (i.e., SAM sites, radio relay stations, fighter aircraft, etc.), if any.

CLAS( ), DOWNGRADING & AUTHORITY

11. In general, the valid data of the EMR environment forecast is determined by the type of analysis requested in Question 4 above. Thus, Type I analysis normally includes a (current) baseline and a 20 to 30 year forecast (the exact interval depending on customer choice). Similarly, Type II analysis

(Continued)

TABLE B-1 (Concluded)

EMR ENVIRONMENT FORECAST REQUEST FORM

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includes a (current) baseline and a 15 to 25 year forecast. However, these intervals are primarily determined by customer requirements. (No forecast will be made for periods exceeding 30 years or for periods less than 5 years.) Specify the validation period of the forecast.

CLAS( ), DOWNGRADING & AUTHORITY

12. For Type II and IIIA analysis, specify the theaters for which a tailored EMR environment is required. The locations of these theaters are specified in Appendix C of the Customer's Application Manual.

1. Alaska - Kamchatka
2. Australia & New Zealand
3. Central America
4. European
5. Mediterranean
6. North Asian
7. Northeastern-North American
8. Pacific Islands
9. South African
10. South Central Asian
11. Southern Latin American
12. CONUS

CLAS( ), DOWNGRADING & AUTHORITY

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APPENDIX C

ANALYSIS AND PREDICTION

10. **INTRODUCTION.** Analysis is the process of separating something into its constituent parts. In EMR analysis, the goal is to break down the EMR response of a system into the constituent parts of the response. For example, the EMR response may be analyzed into the response of scattering structures, into the response of apertures, into cable response, and into electronic component and subsystem response. Mathematical models of the electromagnetic interaction with the system parts are used to predict each response. By considering the separate responses, the system response can be assessed.

Appendix D describes the Intersystem Analysis Program (IAP) which is essentially computer software programs of mathematical models used in analysis and prediction of responses to electromagnetic energy. The purpose of this appendix is to supplement Appendix D with pertinent information and to provide an insight as to analysis and prediction methods. The goal is to develop an overview of analysis and prediction with ample references for those desiring more depth. In some cases, the material presented in the following sections is a recapitulation of other source material. The material, plus additions, has been arranged and assimilated into a form more directly applicable to analysis and prediction of weapon system EMR response.

The role of analysis and prediction is worth discussing before proceeding. Within the present engineering state-of-the-art, neither experimental methods nor analytical methods should be relied on solely to assess a system. In most experimental situations, neither the time nor the capability exists to conduct sufficiently valid statistical or sensitivity tests to fully evaluate all the parameters of an electromagnetic environment. Usually only one weapon system is tested. The system that is tested may differ considerably from systems in the same class. These differences arise because of production differences such as in cable routing paths. It must be realized that all test configurations at best are only a simulation of a system in its actual environment. For example, instrumentation to detect or record responses can unintentionally modify the response being observed. Arguments can also be given why an EMR system assessment should not be based only on analysis and prediction. In general, analysis and prediction is presently at the stage of development that facilitates only "worst case" approaches. The worst case approach will usually result in an upper bound prediction and all potential susceptibilities are identified. However, non-susceptible situations may be predicted as susceptible because of the worst case approach. These situations lead to overhardening with unnecessary adverse effects on system performance. More accurate analysis and prediction can be implemented, but generally are more costly, and the required detailed geometric data is time consuming to obtain or may not be available.

The best overall approach is to utilize experimental and analytical methods in a complementary manner. Such an approach weighs the strong and weak points of both methods against each other. For example, the analysis and prediction method is the only approach available for use in the conceptual stage where little or no hardware is available for experimental assessment. In trade-off studies, two competing systems must be evaluated when only drawings and specifications are available. However, as a system development progresses and hardware becomes available, test results can be used as an input to the analysis and prediction effort to improve accuracy. In other situations, it may be appropriate to use analysis and prediction for test planning or to interpret test results. This complementary use of experimental and analysis/prediction methods assures the best possible methodology within the present engineering state-of-the-art.

20. **ELECTROMAGNETIC ENVIRONMENTS.** Appendix A discusses the EMR environment and defines the source of this environment as high power radiating emitters such as radar and communication transmitters, jammers, navigational equipments, etc. The EMR environment includes friendly and/or hostile emitters. Notice that the EMR environment results from emitters external to the weapon system and includes electromagnetic countermeasures (ECM) sources. There are many other environments, both natural and man-made, to which a weapon system may be susceptible [C-1]. For the analysis and prediction effort, it is advantageous to consider these different environments in a coordinated manner. Coordination of the environments is cost-effective and efficient, since they all require modeling of the same structures, apertures, cables, and circuits. Even though the environments are generated from different sources and their essential properties are divergent, they all give rise to the same generalized problem in the sense that unwanted and unintentional energy is coupled to electronic systems.

20.1 **ENVIRONMENT CHARACTERISTICS.** Other environments that should be considered in an analysis and prediction effort include electromagnetic pulse (EMP), electromagnetic interference (EMI), lightning, and precipitation static (P-static) electricity. Figure C-1 is a simplified spectrum comparison of these environments.

In general, there are various types of EMP, and the most important one is that resulting from a high altitude (exoatmospheric) burst. Here, the EMP is generated by the interaction of nuclear burst products such as gamma rays and x-rays with the upper region of the atmosphere and the earth's magnetic field. Electrons, produced by the Compton scattering of gamma rays and air molecules, spiral about the earth's magnetic field. The resultant effect is a substantial level of electromagnetic radiation below this source region with roughly uniform field intensity radiating outward in all directions. The intensity of the field and area coverage makes the high altitude burst EMP one of the most potent sources of unwanted electromagnetic energy. Other types of EMP include: (1) atmospheric burst EMP, (2) ground burst EMP, (3) dispersed EMP, (4) internal EMP, and (5) system-generated EMP. High altitude burst EMP has a very fast risetime (nanoseconds) and a 10 KHz to 100 MHz spectrum. Field levels can be 100 kv/m orders of magnitude.

EMI is electromagnetic noise in the sense it is unwanted; it is generated by electronic and electrical equipments, distribution networks, radiating subsystems, etc., of a system. These various sources can

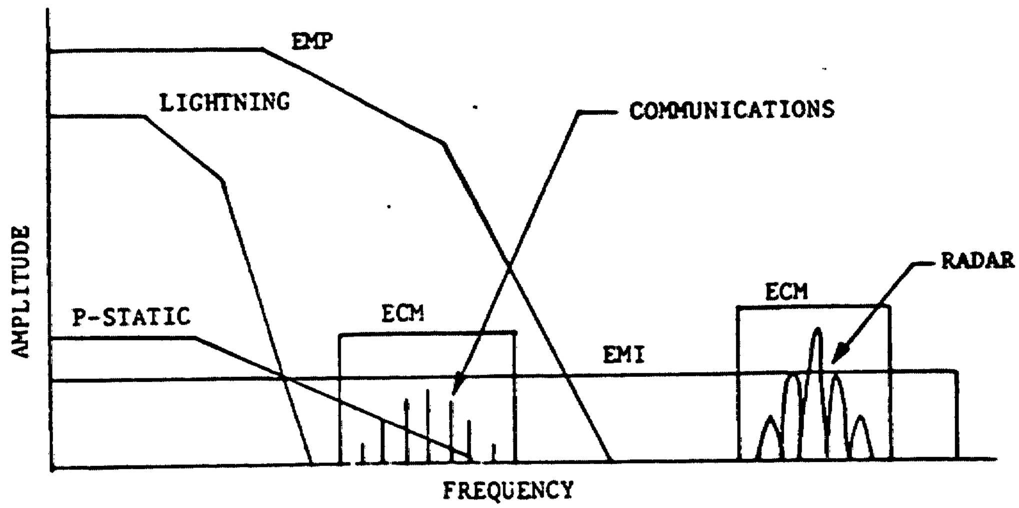


Figure C-1. Simplified Spectrum Comparison be broadly classified as radiated and conducted electromagnetic energy and may be both intersystem or intrasystem generated. The EMI spectrum is usually considered to extend from 400 Hz to 18.0 GHz for airborne systems. Radiated EMI may include levels in the order of 100 v/m.

Lightning strikes on airborne systems are considered to be direct hits for worst case considerations. The most common source of lightning is the electrical charge separated within thunder clouds. Charge separation and lightning can also be induced by thermonuclear detonation. An entry of an airborne system into an electrically charged region can also trigger lightning. A typical lightning stroke comprises multiple pulses, often superimposed on a relatively small continuing or "follow-on" current. The rise times of lightning pulses are on the order of microseconds and have a 1 KHz to 5 MHz spectrum. A direct strike can produce currents on the order of 10 KA.

Static electricity results from an electrical charge which is generated on the surface of an airborne system. When the static electricity leaks off the surface, an electrical interference called P-static results. Static electricity discharges have a 1 MHz to 100 MHz spectrum and voltage levels on the order of 1 Mv. Electrical discharges do not occur between objects when they are bonded together.

20.2 **EMR ENVIRONMENT PARAMETERS.** The primary concern of this handbook is the EMR environment, and parameters of this environment must be considered in an analysis and prediction effort. These parameters are as follows:

(1) Frequency. The EMR environment frequency range is from 1 MHz to 100 GHz. It is obvious that structures will range from small in terms of a wavelength to very large in terms of a wavelength.

(2) Power Density. Power densities may be greater than 60 dBm/m^2. Sources may be CW, AM/FM modulated or pulsed. Both peak and average powers must be considered.

(3) Modulation. Modulation is particularly important in assessing system susceptibilities. Induced signals with the proper modulation can be processed as legitimate signals and cause disruptions. Modulations which affect guidance may be different than those which affect fusing, etc., so each subsystem must be investigated independently.

(4) Polarization. Structures may respond better to some polarizations than they do to others. Horizontal and vertical polarizations are usually used in the analysis and prediction procedures.

(5) Aspect Angle. Since the system's structure and its electronic conductors behave as inadvertant antennas, the coupling between the external field and the structure is a function of the aspect angle.

(6) Dwell Time. Dwell time is a function of the environment and the weapon system dynamics. For example, it is the time that a weapon may remain in a radar mainbeam.

30. **SYSTEM CONSIDERATIONS.**

30.1 **ANALYSIS AND PREDICTION REQUIREMENTS.** EMR analysis and prediction techniques are based on external illumination of a weapon system either in the near-field or far-field of a source. The mathematical models utilized are in the frequency-domain. If the system requirements include EMP and lightning protection, mathematical models derived in the time-domain are usually more suitable. There are several good resources available for EMP and lightning time-domain calculations. The Air Force Weapons Laboratory has published an EMP handbook [C-2] for missiles and aircraft. The Air Force Flight Dynamics Laboratory has published a lightning handbook [C-3] for analysis and calculation of lightning interactions with aircraft electrical circuits.

30.2 **SYSTEM CONFIGURATION.** The system configuration influences the type and level of analysis and prediction techniques utilized. For example, an air-to-air missile in flight is relatively small and self-contained. Mathematical models based on smaller than a wavelength approximations may be appropriate for such a structure. However, when the same missile is on board an aircraft, such models may not be appropriate. Missiles in flight can usually be modeled as if in free-space with a plane electromagnetic wave illumination. For missiles on board an aircraft, near-field illumination is probable. When an aircraft with on-board missiles is on the ground, analysis and prediction calculations must consider the reflections from ground. In essence, the weapon system and its surrounding environment must be included in any analysis and prediction method.

30.3 **PLUME EFFECTS.** Plume effects are important considerations in the system configurations. In some cases, the ionized exhaust gases of the plume can alter a missile's skin current distribution, and in turn the coupling of EM energy into the interior of the missile. There have been several theoretical studies dealing with plume effects. Harrison's [C-4] study used a plume model on which the effective electrical conductivity was assumed to be an exponential function of the axial distance and the conductivities where taken to be high. Smith's [C-5] study was directed toward small missiles with solid propellant motors, and it is this study that is reported on here.

Both homogeneous and axially inhomogeneous plumes were considered. For the inhomogeneous case, a low altitude plume program was used to provide realistic values of the electrical properties along the length of the plume. The geometry of the thin-wire model used for the analysis is shown in Figure C-2. The missile and the plume are represented by cylinders with lengths ℓ_m and ℓ_p, respectively, and radius (a). An internal impedance per unit length z_p^i, which can be a function of the axial distance z, is used to describe the electrical properties of the plume. A uniform internal impedance z_m^i

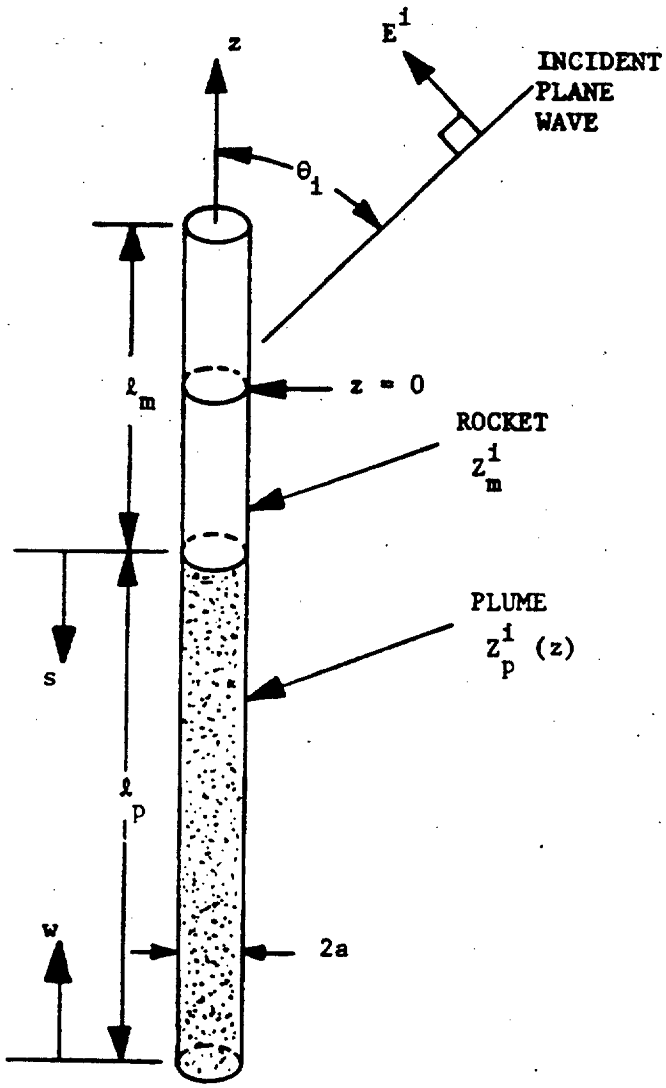


Figure C-2. Detail of Thin-Wire Model for a Missile with Plume. [C-5] is used for the highly conducting missile. The incident electromagnetic field is a linearly polarized plane wave with electric field E^i in the plane formed by the axis of the missile (z axis) and the propagation vector k. For an incident field in another direction, only the component of the field in this plane need be considered in the analysis since it is the only component that will excite the thin-wire structure. The electric field along the axis of the missile (z axis) is $E_z^i(\theta_i) = E_0 \sin \theta_i \exp(-j\beta_0 z \cos \theta_i)$ where $E_0 = E_i$, and the reference point for the phase of the incident field is at the center of the missile (z = 0).

The total axial current at any cross section of the rocket or plume is I(z). The integral equation for the current on this structure is the same as that used in the familiar analyses for linear antennas:

$$\int_{z'=-(\ell_p+\ell_m/2)}^{\ell_m/2} \int_{\phi'=0}^{\pi} \frac{I(Z')e^{-j\beta_o R}}{R}\, d\phi'\, dz'$$

$$= A \cos \beta_o z + B_o z$$

$$\cdot \frac{-j2\pi}{\zeta_o} \frac{(\lambda E_o)}{\sin\theta_i} e^{j\beta_o z \cos\theta_i}$$

$$\cdot \frac{+j4\pi^2}{\zeta_o} \int_{t=-(\ell_p+\ell_m/2)}^{z} z^i(t) I(t) \sin \beta_o (z - t)\, dt \qquad \text{(C-1)}$$

where:

$$z^i(z) = z_m^{\ i} \text{ for } -\ell_m/2 \leq z \leq \ell_m/2$$

$$= z_p^{\ i}(z) \text{ for } -(\ell_p + \ell_m/2) < z \leq -\ell_m/2 \qquad \text{(C-2)}$$

and:

$$R = [(z - z')^2 + 4a^2 \sin^2(\phi'/2)]^{1/2}. \qquad \text{(C-3)}$$

β_0 and ζ_0 are the propagation constant and the characteristic impedance for free space, respectively. The constants A and B in (C-1) are determined by imposing the boundary conditions at the ends of the structure, i.e., $I(\ell_m/2) = I(-\ell_p-\ell_m/2) = 0$.

The current was determined from the integral equation (C-1) using the method of moments technique which is described in a later section.

Figure C-3(a) shows results of the analysis for the missile model with a homogeneous plume. The distribution of the current on the missile without a plume is seen to have a form like $[\cos \beta_0 Z - \cos \beta_0(\ell_m/2)]$ which is characteristic of a thin-wire scatterer in an electric field parallel to its axis. The current builds up to maximum value at the center of the missile when the length of the structure is near that for resonance (ℓ_m/λ somewhat less than 0.5). With the plume present, the amplitude of the maximum current on the missile is reduced and is not as sharp a function of ℓ_m/λ as without the plume. The current along the plume is fairly uniform except near the ends. This is due to the high resistivity of the plume which attenuates axial currents that could produce a high standing wave on the structure. The effect of the plume on the current in the missile is quite different for missiles with different electrical lengths ℓ_m/λ. This is also illustrated in Figure C-3 where distributions of current are shown for a missile with a length near that for resonance and an electrically short missile. In Figure C-3(a), the maximum current on the missile with a plume is seen to be less than that on the resonant missile alone. The current near the tail of the missile is greatly increased when the plume is added. For an electrically short missile, $\ell_m/\lambda << 1$, the maximum current is much less than on the resonant missile, as shown in Figure C-3(b) (note the change of the current scale). With the plume present, the current on the electrically short missile is increased over the value without the plume at every point along its length. The relative changes in the current on a missile without a plume as compared with a missile with a plume are quite different. For the resonant missile, the distribution of the current on the missile is determined primarily by the length of the missile even when the plume is present. For the electrically short missile, however, the distribution of current on the missile is determined by the length of the missile and the plume, $\ell = \ell_m + \ell_p$. Note that the maximum current on the missile for all values of ℓ_m/λ used occurs on the missile without a plume near resonance.

In Figure C-4, the current distributions for a missile with an inhomogeneous plume model are shown. For the resonant missile, the currents are very similar for the homogeneous and inhomogeneous plume models. For the electrically short missile, Figure C-4(b), the situation is quite different. The maximum current on the missile with an inhomogeneous plume is about three times as large as that for the missile with a homogeneous plume. This difference is consistent with the idea that the distribution of current is determined by the electri-

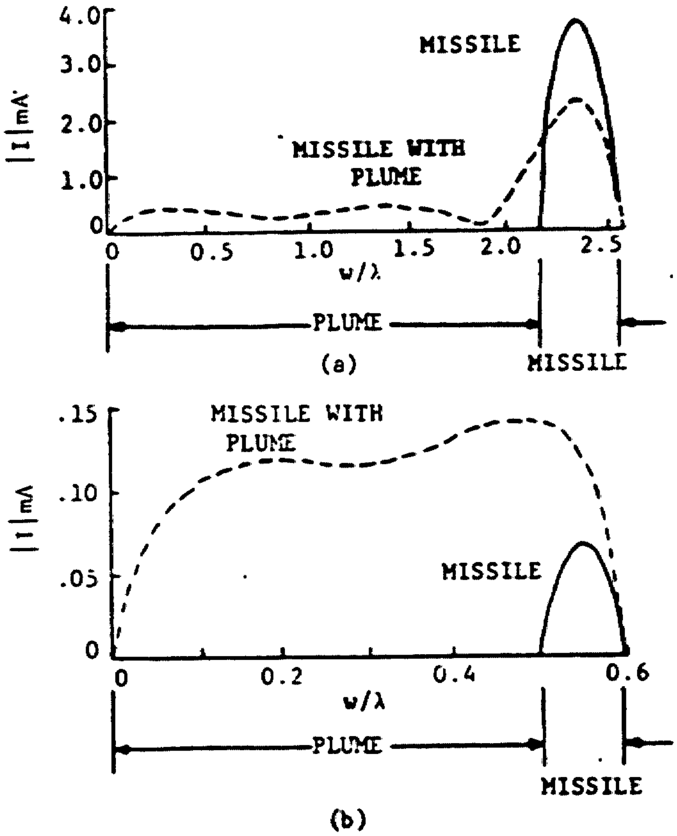


Figure C-3. Comparison of Magnitude of Current on Missile and Missile with Homogeneous Plume. (a) Length of Missile Near That for Resonance, $\ell m/\lambda = 0.39$. (b) Electrically Short Missile, $\ell m/\lambda = 0.09$. [C-5]

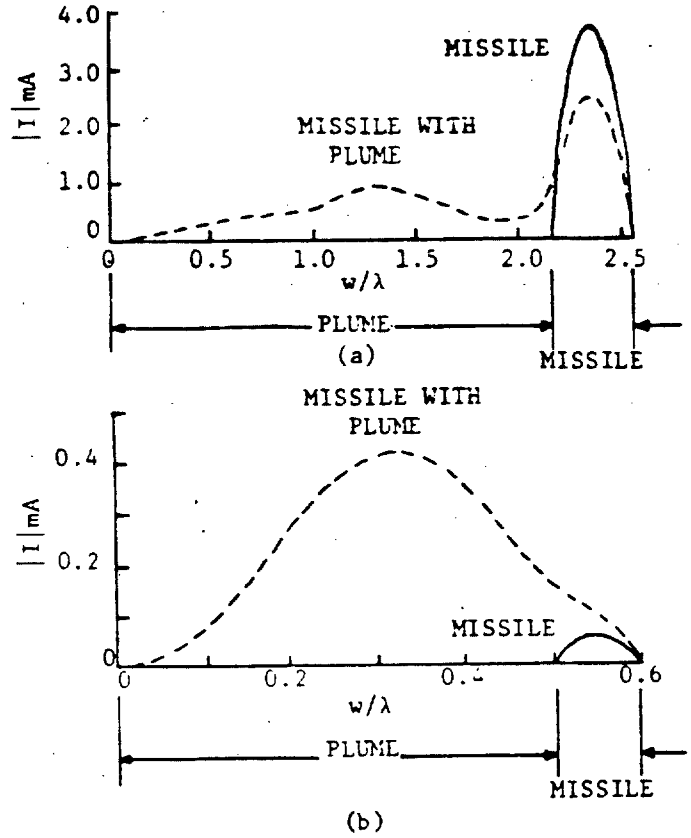


Figure C-4. Comparison of Magnitude of Current on Missile and Missile with Inhomogeneous Plume. (a) Length of Missile Near that for Resonance, $\ell m/\lambda = 0.39$. (b) Electrically Short Missile, $\ell m/\lambda = 0.09$. [C-5] cal length of the missile and the plume for the electrically short missile. A change in the internal impedance per unit length of the plume has a major effect on the current.

30.4 **AVAILABILITY OF DATA.** The data available for analyzing a weapon system is generally a function of the system's life cycle [C-6]. The various phases of the life cycle are discussed in Section 4. The availability of data impacts the selection of analysis and prediction procedures for each application. Some procedures require structure geometry specified to within fractions of a wavelength while other procedures require only approximate structural representations perhaps in terms of generic shapes. Also, the electrical characteristics needed for analysis and prediction can be detailed as time waveforms or as superficial as frequency assignments. Availability of data increases as a system progresses through each phase of the life cycle. The extent of the data at each phase is dependent upon the degree to which "off-the-shelf" equipment is utilized or the degree to which the system under development is similar to other existing systems. Table C-1 provides broad guidelines as to data availability as a function of the life cycle phases.

The performance criteria of subsystems is another type of data availability that impacts the choice of models and procedures. The performance criteria of a subsystem are its characteristics when viewed as a receptor, such as performance degradation curves or component susceptibility data. In some cases, the system performance criteria may be known and the subsystem criteria derived. In other situations, off-the-shelf equipment is utilized and performance criteria has been previously determined.

In summary, the structural, electrical, and performance criteria available at any point in the life cycle has a major impact on the selection of the analysis and prediction procedure selected. An analyst must judge and chose the best approximations and models based on the available data.

40. **COMPUTATIONAL TECHNIQUES.** There are a number of powerful computational techniques available for use in interaction and coupling analyses. A thorough discussion of any one of the computational techniques would be a lengthy treatise in itself. The scope here is limited to a survey to provide an introduction to the subject material. Computational techniques exist for both the frequency and time domains. Frequency domain models encompass those methods that either solve an electromagnetic problem at a single time-harmonic frequency (usually temporal variation assumed) or solve a quasi-static problem ($f \to 0$) or an asymptotic frequency problem ($f \to \infty$). Time domain models are used to find the impulse response of an electromagnetic system to a known stimulus. Of course, through Fourier transforms, frequency domain models can often be used to solve many time domain problems and vice versa.

TABLE C-1

DATA AVAILABLE VERSUS PHASES OF LIFE CYCLE

Conceptual

- o Subsystem Function Definitions
- o Organizations Responsible for Each Subsystem
- o Generic Subsystem
  - Power Circuits
  - Communication Circuits
  - Telemetry Circuits
  - Data Processing Circuits
- o Expected Geometry
  - Size
  - Weight
  - Generic Shape

Validation

- o Characteristics of Individual Susbystems
  - Power Requirements
  - Time Waveforms
  - Spectrum
  - Susceptibility
- o Prototype Specifications
  - Geometric Data
  - Schematics and Diagrams
  - Material Characteristics
- o Support Equipment Characteristics

(Continued)

TABLE C-1 (Concluded)

DATA AVAILABLE VERSUS PHASES OF LIFE CYCLE

Full Scale Development

- o Geometric Data
  - Shape and Surface Information
  - Wire Routing
  - Component Locations
- o Test Results

Production

- o Measured Data of Production Sample
- o Refined Data on Geometry and Electrical Characteristics
- o Description of Equipment Down to Brand Names

Deployment

- o Maintenance Statistics
- o Modification Specifications

40.1 **METHOD OF MOMENTS.** The method of moments (MOM) [C-7]-[C-9] is a numerical technique for obtaining approximate solutions to three dimensional integral equations of the form:

$$f(r) = \iiint_{v'} k(r, r')g(r')\, dv' \qquad \text{(C-4)}$$

where $f(r)$ is some known forcing function, $k(r, r')$ is a known integral kernel and $g(r')$ is the function to be determined. To illustrate the technique, consider the more simple one dimensional integral equation:

$$f(x) = \int_{P_1}^{P_2} k(x,x')g(x')\, dx' \qquad \text{(C-5)}$$

The method assumes an approximation to the unknown function $g(x')$ of the form:

$$g_a(x') = \sum_{i=1}^{N} g_i a_i(x') \qquad \text{(C-6)}$$

where the coefficients g_i are unknown and the $a_i(x')$ are linearly independent functions, usually called expansion or basis functions. The $a_i(x')$ are chosen in the hope that some combination such as (C-6) can accurately approximate the unknown. Next, another set of functions $w_k(x)$ ($k = 1, 2, ---, N$), usually called weighting functions, are chosen so that a similar linear combination can accurately approximate $f(x)$. Replacing $g(x')$ by $g_a(x)$ in (C-5) gives:

$$f(x) \approx \int_{P_1}^{P_2} k(x,x') \left[\sum_{i=1}^{N} g_i a_i(x')\right] dx'$$

$$\approx \sum_{i=1}^{N} \left[\int_{P_1}^{P_2} k(x,x')a_i(x')dx'\right] g_i \qquad \text{(C-7a)}$$

Multiplying both sides of (C-7a) by each of the N functions $w_k(x)$ and integrating with respect to x from p_1 to p_2 gives an inhomogeneous set of N linear equations in N unknowns:

$$\int_{P_1}^{P_2} w_k(x)f(x)dx = \int_{P_1}^{P_2} w_k(x)\left\{\sum_{i=1}^{N}\left[\int_{P_1}^{P_2} k(x,x')a_i(x')dx'\right]g_i\right\}dx \qquad \text{(C-7b)}$$

$$= \sum_{i=1}^{N}\left\{\int_{P_1}^{P_2}\int_{P_1}^{P_2} w_k(x)k(x,x')a_i(x')dx'dx\right\}g_i \qquad k = 1,2,\ldots,N$$

or

$$f_k = \sum_{i=1}^{N} \alpha_{ki}g_i \qquad \text{(C-7c)}$$

with

$$f_k = \int_{P_1}^{P_2} w_k(x)f(x)dx \qquad \alpha_{ki} = \int_{P_1}^{P_2}\int_{P_1}^{P_2} w_k(x)k(x,x')a_i(x')dx'dx$$

(C-7c) can be written in matrix form as:

$$\overline{F} = [\alpha]\,\overline{G} \qquad \text{(C-7d)}$$

which has the solution:

$$\overline{G} = [\alpha]^{-1}\,\overline{F} \qquad \text{(C-7e)}$$

The general function $g(r')$ is some unknown scalar or vector source distribution such as charge or current. The function $f(r)$ is a known field quantity such as potential, electric field intensity or magnetic field intensity. Thus the electromagnetic field problem in terms of an integral equation has been reduced to a matrix equation which is more readily solved using computers.

There are two integral equation formulations that can be used in the MOM; one is the electric-field integral equation (EFIE) and the other is the magnetic-field integral (MFIE). The MFIE is well suited for thin-wire structures of small or vanishing conductor volume while the MFIE, which fails for the thin-wire case, is more attractive for voluminous structures, especially those having large smooth surfaces.

When EFIE is applied to wire structures, certain assumptions [C-10] are made so that thin-wire approximations may be invoked. The assumptions applied are:

(a) Transverse currents can be neglected relative to axial currents on the wire.

(b) The circumferential variation in the axial current can be neglected.

(c) The current can be represented by a filament on the wire axis.

(d) The boundary condition on the electric field need be enforced in the axial direction only.

These widely used approximations are valid as long as the wire radius is much less than the wavelength and much less than the wire length.

As an example of the application of the MOM, consider the dipole antenna and its MOM model in Figure C-5. Thin-wire assumptions are used, and the dipole antenna is divided into n segments. For the illustration, assume the model is divided into 5 segments. The matrix equation for the MOM model becomes:

$$\begin{bmatrix} 0 \\ 0 \\ V_1 \\ 0 \\ 0 \end{bmatrix} = \begin{bmatrix} Z_{11} & Z_{12} & Z_{13} & Z_{14} & Z_{15} \\ Z_{21} & Z_{22} & Z_{23} & Z_{24} & Z_{25} \\ Z_{31} & Z_{32} & Z_{33} & Z_{34} & Z_{35} \\ Z_{41} & Z_{42} & Z_{43} & Z_{44} & Z_{45} \\ Z_{51} & Z_{52} & Z_{53} & Z_{54} & Z_{55} \end{bmatrix} \cdot \begin{bmatrix} I_1 \\ I_2 \\ I_3 \\ I_4 \\ I_5 \end{bmatrix} \qquad \text{(C-8)}$$

The voltage matrix is zero except at the gap of the antenna. The generalized impedance matrix represents all the interactions between the segments on the MOM model. The unknowns are the currents on each wire segment. The matrix equation is solved for the unknown currents. Once the currents are known, field quantities such as antenna patterns and impedances may be calculated.

As noted from the previous example, the impedance matrix size is n x n, where n is the number of segments. Thus the impedance matrix size increases in direct proportion to the size of the structure. The MOM is best suited to structures with dimensions up to several wavelengths. Although there is no theoretical size limit, the numerical solution requires a matrix equation of increasing order as the structure size is increased relative to wavelength. Hence, modeling very large structures may require more computer time and storage than is practical and still be within reasonable cost. The computer solution time is proportional to n^3 and the computer storage time is proportional to n^2. To scope the demands on the computer, consider a $\lambda/4$ monopole mounted on a $2\lambda \times 2\lambda$ ground plane. The ground plane can be approximated by a wire grid using ten segments per wavelength. Therefore, there will be a total of approximately 400 unknowns with the monopole. The 400 unknowns require storage of $(400)^2 = 160{,}000$ Z_{ij} impedance interactions which are all complex numbers.

Not only have thin-wire models been developed using MOM's, but surface integral equations have been used for the so-called surface patch models. Surface patch models have been based on the MFIE [C-11]

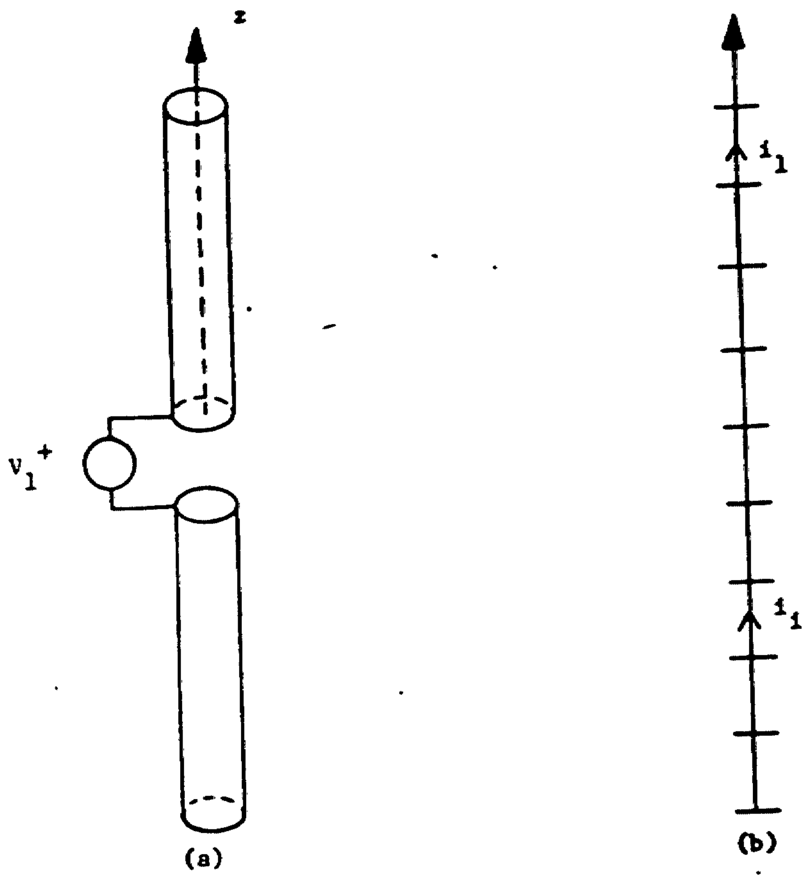


Figure C-5. (a) Dipole Antenna, (b) MOM Model [C-9].

and the EFIE [C-12]. The thin-wire models have been applied to surfaces by using wire-grid models. However, wire-grids have been shown to be poor models [C-13] of a closed surface for interaction calculations, and the surface patch models are preferred for closed surfaces.

40.2 **GEOMETRICAL THEORY OF DIFFRACTION.** As previously discussed, the MOM's solutions tend to be restricted to lower frequencies, based on the fundamental limitation on the size of matrices which computer can solve without excessive loss of accuracy or excessive costs. A computational technique more suited to high frequencies (the structure is large in terms of a wavelength) is the Geometrical Theory of Diffraction (GTD) [C-14]. Since GTD is essentially a high frequency solution, the lower frequency limit of this solution is dictated by the spacing between the various scattering centers (they should be at least a wavelength apart). Under this restriction, the low frequency limit is typically around 100 MHz. The upper frequency limit is dependent on how well the theoretical model simulates the important details of the actual structure.

GTD was introduced by Keller [C-15] as an extension of geometrical optics to include the diffracted field in the high frequency solution. The theory is based on the following postulates:

(1) The diffracted field propagates along rays which are determined by a generalization of Fermat's Principle to include points on edges, vertices, and smooth surfaces in the ray trajectory.

(2) Diffraction like reflection and transmission is a local phenomenon at high frequencies, i.e., it depends only on the nature of the boundary surface and the incident field in the immediate neighborhood of the point of diffraction.

(3) The diffracted wave propagates along its ray so that

   (a) power is conserved in a tube (or strip of rays),

   (b) the phase delay along the ray path equals the product of the wave number of the medium and the distance.

Using these postulates, one can express the diffracted field in the same form as a geometrical optics field with some coefficient of proportionality to the incident field at the point of diffraction. The coefficient is determined from a canonical problem and is referred to as a diffraction coefficient. For practical purposes, the GTD can be divided into two categories: (1) wedge diffraction theory - to treat diffraction by edges and (2) creeping wave theory - to treat diffraction by curved surfaces.

Complex structures such as a missile or an aircraft are modeled using a composite of wedges or curved surfaces to represent the important scattering surfaces. For example, wedges may be used to represent the scattering from aircraft wings and scattering from the fuselage may be represented by a cylinder.

To obtain some idea of the formulation of GTD solutions, consider diffraction from the wedge structure in Figure C-6 [C-16]. Consider a plane electromagnetic wave normally incident on the wedge of an angle $(z-n)$. The azimuthal angles ψ and ϕ_0 are associated with the angle of diffraction and the angle of incidence, respectively.

The field μ at point P is a solution to the scalar wave equation subject to the appropriate boundary conditions; the solution may be formulated as:

$$\mu(r,\psi) = \nu(r,\psi + \phi_o) \pm \nu(r, \psi - \phi_o) \tag{C-9}$$

Polarization determines the choice of sign such that, with the electric vector perpendicular (parallel) to the edge, the positive (negative) sign is chosen. It is convenient to represent the incident or reflected field in the form

$$\nu(r,\phi) = \nu(r,\psi \pm \psi_o) \tag{C-10}$$

such that the (-) sign yields the incident fields and the (+) sign yields the reflected fields. The component field is given by

$$\nu(r,\phi) = \nu^* + \nu_B \tag{C-11}$$

where ν^* is the geometrical optics field given by

$$\nu^* = \begin{cases} e^{j\rho \cos(\phi + 2\pi nN)} & \text{for } -\pi < (\phi + 2\pi nN) < \pi \\ 0 \text{ otherwise} & \end{cases}$$

$$N = 0, \pm 1, \pm 2 \; . \; . \; . \pm n \tag{C-12}$$

and ν_B is the diffracted field given by

$$\nu_B = \frac{1}{2\pi n} \int_c \frac{e^{j\rho \operatorname{Cos}\beta}}{1 - e^{-j(\beta+\phi)/n}} \, d\beta \tag{C-13}$$

where

$$\rho = kr \tag{C-14}$$

Here C is the appropriate path in the plane of the complex variable.

Asymptotic expressions for (C-13) have been derived by Sommerfield [C-17], Pauli [C-18] and Hutchins and Kouyoumjian [C-19]. It is the asymptotic expressions that are used in GTD computer computations.

40.3 **HYBRID TECHNIQUES.** Both the MOM and the GTD are useful computational methods within their class of problems. The characterization

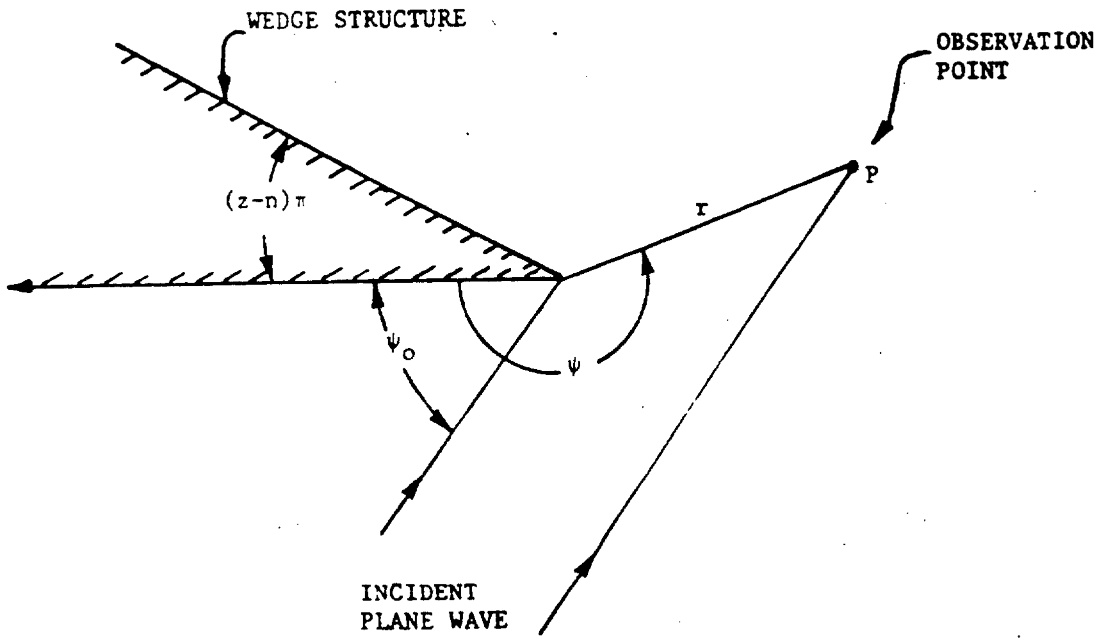


Figure C-6. Diffraction by Plane Wave by Wedge [C-16] of wires on or near a metallic surface by MOM is significantly limited by computer storage. This restriction applies to thin-wire or surface patch representations of a surface. Thus the MOM is a low-frequency technique since its practical use is generally limited to bodies that are not large in terms of a wavelength. However, the GTD is a high frequency technique applicable to bodies that are large in terms of a wavelength. The two computational methods can be combined to complement each other using the so-called hybrid techniques. Thiele [C-20] has extended the MOM via the use of GTD to include a wider class of problems where a large body (in terms of a wavelength) has a wire structure located on or near the body. The approach is to modify the MOM impedance matrix using GTD to account for the large body presence. Burnside [C-21] has developed a hybrid technique in which the GTD has been extended via the MOM. In this approach, the MOM is used to obtain the GTD diffraction coefficients, thus enabling GTD to be applied to additional structures that can not be handled with GTD alone.

40.4 **STATISTICAL METHODS.** EM coupling into electronic components and subsystems can not always be characterized in a deterministic manner. This is especially true when the EM coupling is via inadvertent paths such as seams, holes, cables, etc. While in principle an arbitrarily accurate analysis of the EM coupling can be derived by solving Maxwell equations in the context of a boundary valve problem, in reality even for a relatively simple system, such a classical deterministic approach often demands more effort and resources than are available. Not only does the EM coupling through inadvertent paths require statistical methods, but so does the incident EM energy, since it is a function of polarization, incident angle, etc. To circumvent this situation, EM coupling analysis may rely on worst-case models or on statistical models. The development of suitable statistical models are still in an early stage of development. Some of the recent efforts to formulate statistical models are reviewed here.

Graham [C-22] analyzed coupling to an electronic system in terms of random small dipole interactions for the frequency region where the system components, e.g. wire lengths, are small in terms of a wavelength. Both random coupling to the incident wave and random interactions among the dipoles were considered. The variables randomized were the incident direction and polarization, the sizes and orientations of the dipoles, the mutual coupling effects, and the lumped load impedances. The results of the statistical model showed that the EM coupling to large systems, when dominated by low frequency magnetic fields, is largely insensitive to the coupling detail and yields a distribution whose central part is nearly log-normal with a standard deviation of about 6 dB. However, it may be that the shapes of the extreme percentiles of the distribution may depend on and be sensitive to the detailed nature of the coupling.

Morgan [C-23], in an informal manner, developed basic statistical concepts concerning the statistical analysis of load impedance excitations induced on a random N-wire cable by an incident field. The N-wire unshielded and unbranched random cable is modeled by a stochastic random process such as the Monte Carlo random walk procedure. The reciprocity theorem is used to compute the load excitations.

Swink [C-24] considered the canonical problem of penetration of EM energy into a cylindrical enclosure through a small aperture as shown in Figure C-7. A parametric analysis of the model was performed by varying the incident angle of the incident field, the aperture location, and load impedances. The parametric data was then condensed and interpreted by utilizing basic statistical methods. Prehoda [C-25] extended the model investigation to consider the cylinder's antenna pattern effects and the internal transmission line structure for low frequencies. The antenna pattern effects were characterized using cumulative gain distributions. Some of the observations from the study were:

(1) The aperture admittance of a well-shielded enclosure presents a large mismatch to both the external structure equivalent antenna configuration, as well as to the internal transmission line configuration.

(2) Although the mean power available to widely different external structure equivalent antenna configurations is identical, the probability of receiving that power is highly dependent upon the gain of the antenna.

(3) A low-gain, external-structure, equivalent antenna, on the average, is likely to receive more power than a higher gain antenna for random angles of incidence.

(4) The 1% to 95% portion of the cumulative gain probability curves of the received power associated with a varying load at the end of a transmission line are not extremely sensitive to the distribution of the load impedance.

(5) The mean of the external-structure antenna response of a shielded electronic system is a constant. Similarly, the distribution of power received by an internal load as a function of internal variations is relatively stable. As a result, the coupling of electromagnetic energy into an electronic system is primarily dependent upon the charcteristics of the aperture itself.

40.5 **FINITE DIFFERENCE METHODS.** The finite-difference method is significantly different from integral equation approachs in that the method utilizes Maxwell's time dependent curl equations, and the equations are applied to a volume containing the structure of interest rather than a surface [C-26] [C-27]. The basic approach in using this method is to form a three-dimension lattice that surrounds the structure and solve the resulting finite-difference forms of Maxwell's equations in a time-stepping manner. By time-stepping, i.e., repeatedly solving the finite-difference analog of the curl equations at each point of a space lattice containing the structure of interest, an incident wave on the structure is tracked as it first propagates to the structure, and then interacts with it in some way (surface current excitation, diffusion, penetration, etc.). Wave tracking is completed for pulsed illumination when the desired early or late time behavior is observed; for sinusoidal illumination, the end point is the attainment of the sinusoidal steady state.

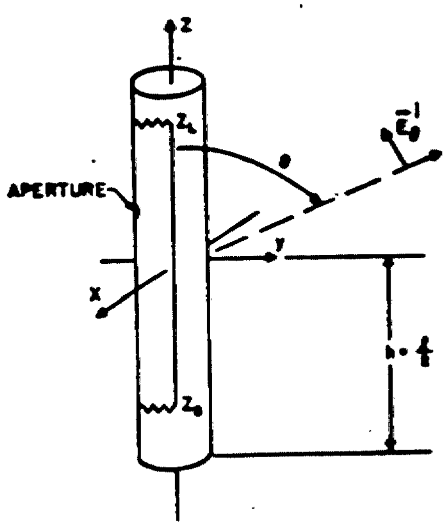


Figure C-7. Cylindrical Enclosure Model. [C-24]

Time-stepping for the finite-difference method is accomplished by what is termed an explicit finite-difference procedure. Here, the value of an electromagnetic field component at the latest time step is computed using only field quantities found during the previous time step, and stored in the computer memory. Thus, no simultaneous equations are needed to compute the fields at the latest time step. Further, computation can proceed one lattice point at a time, and the new field value at each point can be placed immediately in memory.

The finite-difference method formulation permits simple and straightforward modeling of arbitrary dielectric/conducting structures. The space containing the structure being modeled is divided into discrete volumes or unit cells. The simplest case is that of a cubic unit cell, which results in a cubic lattice approximation of the geometry. For this case, the structure of interest is mapped into the space lattice by first choosing the dimensions of the unit cell, and then assigning appropriate values of electrical permittivity and conductivity to each unit cell of the lattice. Thus, inhomogeneities or fine details of the structure can be modeled with a maximum resolution of one unit cell and then surfaces can be modeled as infinitely thin, stepped edge sheets.

Taflove C-26 recently extended the finite difference method to incorporate a satisfactory approximation to the free space condition at the lattice truncations, and the simulation of a long duration pulse or continuous wave incident on the structure of interest. Taflove also applied the method to map the field distribution within a small nose cone section of a missile which has an optical port and a seam type aperture.

50. **AVAILABLE COMPUTER CODES.** The availability of computer codes has put within the grasp of the EM analyst a technical data base to handle complex analysis and prediction procedures. The data base is not centralized, but is segmented through numerous organizations. Some organizations are prepared to transfer their codes to qualified users. Other organizations go a step farther and maintain output libraries which are available to users.

Although many computer codes are written in standard computer languages, such as FORTRAN IV, the codes are not truly machine independent. Invariably, subroutines must be modified to get a computer code up and running on a user's computer. For these reasons, when requesting computer codes, it is recommended that a two to three month lead time be used to allow ample time for code modification.

In the following sections, system-level, interaction/coupling, and circuit analysis codes are described. The description includes the code name, the developer organization, and the applicable geometry.

50.1 **SYSTEM LEVEL CODES.** System level codes are a collection of mathematical models used in solving for a system response. In EMR analysis, the system response to EM energy is determined by breaking down the response into constituent parts such as the response of scattering structures, apertures, cables, etc. Characteristically, system level codes also have provisions for calculating or inputting susceptibility data. The data is used to establish criteria to measure EM energy effects on system performance.

System level codes may be either intersystem or intrasystem. Intersystem concerns interactions between independent systems such as two aircraft. In the intrasystem case, interactions are between components or subsystems within a single system. EMR analysis and prediction is concerned with both intersystem and intrasystem configurations. For example, cosite analysis of two aircraft is an intersystem problem while an on-board analysis of a single aircraft is an intrasystem problem.

Several system level codes are described in this section. The codes have other available models that may provide additional capability or provide a better model than existing ones in IAP's IEMCAP system level code.

50.1.1 **SPECIFICATION AND ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM (SEMCAP).** SEMCAP is a large scale computer program for intrasystem analysis, and has been in use since 1968, when it was developed as a spacecraft oriented intrasystem analysis program for NASA [C-28] [C-29].

Basically, this program performs an analysis between modeled interference generators and modeled interference receptors for various interference transfer functions. The generators and receptors are modeled in terms of their electrical and physical parameters. The system's physical characteristics are also modeled so that transfer functions can be computed. The computer calculates the spectrum of the generator circuits and transfers the energy via the transfer function to the receptor terminals. The received spectrum is limited by the receptor bandwidth and integrated over the complete frequency range from 10 Hz to 10 GHz (or higher, at the user's option). The integral then represents the voltage available at the receptor terminals. This received voltage is compared to the threshold of the particular circuits to determine compatibility status. In addition, the computer stores the voltage received from a particular generator and proceeds through the complete generator list until the contribution of each generator is determined. These contributions are summed to determine if the receptor is compatible with the sum of all generating sources modeled.

SEMCAP was developed to perform analyses of spacecraft, where the major problems are in the area of wire-to-wire coupling in complex cable harnesses. Antenna-to-antenna radiation on most satellites does not represent a serious analysis problem because, unlike an aircraft, there are few antennas on most satellites, and these usually operate in the GHz frequency range. Therefore, the emphasis was placed on what happens inside the spacecraft. Consequently, SEMCAP models are virtually unconstrained by the complexity of cable harnesses regardless of the number of different wires or types of wires used, the number of harness segments, unequal spacing of wires in the harness, different heights above ground, different pigtail lengths, etc. A unique feature of SEMCAP is that the model may simultaneously define a large number of different separation distances between wires in the same harness,or between wires in different harnesses. In addition to handling the effects of shielded wires, SEMCAP simultaneously can deal with the effects of group shields, bulkhead shielding, various values of ground return resistances, various values of common return paths, etc.

However, because it is recognized that not all problems deal with wire-to-wire coupling, a flexible method was developed for modeling antenna-to-antenna, antenna-to-wire, and field-to-wire coupling. This method requires definition of the field rather than definition of typical antenna parameters. In fact, the antenna coupling capability of SEMCAP has been used very little because of the nature of the analytical problems to which it has been addressed. On the other hand, coupling from external fields originating from arc discharges and EMP has been analyzed using the E and H-field models. A feature of this flexibility is that any number of various internal and external fields can be modeled simultaneously, in either the time or frequency domain, and a large number of different structural shielding characteristics can be modeled simultaneously, with the shielding factors being either constant or variable as a function of frequency.

Another unique feature of SEMCAP is that it generates its own frequency base using as many points as necessary to define the spectrums, and has a standard frequency base of 1801 points which are logarithmically spaced to give 1% resolution. On the other hand, if the user wishes to select frequencies as in frequency amplitude pairs, he is virtually unlimited as to the number he may use.

SEMCAP models for generators are described in either the time or frequency domain. In the time domain, it accepts models for sinewaves, single pulses, pulse trains, and ramp steps. In the frequency domain, there is no practical limit to the number of frequency amplitude pairs that may be used. Voltage sources and current sources are defined independently. E and H-field sources are derived from the voltage and current sources, or may be entered independently, at the user's option. Filters may be used for each source and may be defined in either the frequency domain or by standard parameters such as cutoff frequency, slope, inband insertion loss, etc. Two filters may be cascaded for each source. The use of generator filters is optional, and in practice is used infrequently.

SEMCAP models for receptors are modeled as voltage thresholds combined with a frequency response curve. This allows a reasonably accurate representation of both analog and digital (or bi-level) circuits. The frequency response curve can be defined in the frequency domain or by standard filter parameters. If desired, two filters can be cascaded for each receptor.

The development of SEMCAP Version 8 provides an enchanced version of a proven system level analysis program which has been successfully used for many years.

50.1.2 **INTERFERENCE PREDICTION PROCESS - VERSION 1 (IPP-1).** The system level computer code IPP-1 [C-30] [C-31] is a versatile computer program designed to assess transmitter-to-receiver interference and to provide useful parameters and data for optimizing compatibility in EM environments.

Development of the operating program was sponsored by RADC and contractually carried out by the Atlantic Research Corporation, Alexandria, Virginia.

IPP-1 may be used to assess interaction between equipments over a broad frequency span ranging from VLF through microwave systems. While pulse and non-pulse systems are both within the capability of IPP-1, many of the submodels were generated to handle the special interaction mechanisms of non-pulse systems.

IPP-1, operating under the control of an "executive routine," can, through user options, "order" several basic types of analyses to be performed. The analysis options are:

- o EMC analysis,
- o data base management,
- o power density/field strength analysis,
- o frequency/distance analysis,
- o intermodulation analysis,
- o adjacent signal analysis.

The EMC analysis is performed in three basic phases termed:

- o Rapid Cull Phase,
- o Frequency Cull Phase,
- o Detailed Analysis Phase.

In the Rapid Cull Phase, simplified, conservative estimates are made to eliminate obvious, non-interfering situations which exist in the environment. In certain applications, the electromagnetic environment may be quite large and rapid means must be used to eliminate the low likelihood interference cases. The Rapid Cull Phase performs this function.

The Frequency Cull Phase, using more refined analysis techniques, operates on the reduced environment (cases not eliminated by Rapid Culling) to effect further reduction of the environment, i.e., cases which have a higher likelihood of occurring are examined in this phase.

Once the rapid and frequency culls have been applied, the environment (cases to be examined) should be significantly reduced from the original level. The detailed analysis, a fairly time-consuming operation, can now be used to "fine-grain" analyze the cases which appear to have an appreciable chance of interference.

Output from each analysis phase, described above, is used as input to the next phase. This type of approach (sequential culling) attempts to optimize computer time (minimize computer run-time).

50.1.3 **INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS (IECA).** The IECA [C-30], [C-31] was developed under Air Force sponsorship to analyze and predict interference between avionic subsystems on aerospace vehicles. IECA includes four interrelated programs:

a. Antenna-to-Antenna Compatibility Analysis Program (ATACAP) coupling -- Analyzes interference from transmitters to receivers when the path is between their antennas.

b. Wire-to-Wire Compatiblity Analysis Program (WTWCAP) -- Analyzes interference resulting from cross coupling within a wire bundle.

c. Field-to-Wire Compatiblity Analysis Program (FTWCAP) -- Analyzes interference induced in the loads of an aircraft wire bundle from exposure to on-board antenna radiation through dielectric apertures in the vehicle skin.

d. Box-to-Box Compatiblity Analysis Program (BTBCAP) -- Analyzes interference resulting from low frequency magnetic fields coupling into sensitive transformers and electron beam devices within equipment boxes.

Each program exists as a separate deck of punched cards, and input data formats are compatible between them. All four programs can be run for a given vehicle to obtain a complete analysis or they can be run independently, as desired. The programs are in FORTRAN IV language and were written for the CDC 6600 computer.

Since analyses and prediction are most valuable early in the conceptual and design phases of vehicle development, the programs may be used before many of the basic equipment parameters are known. Therefore, the program has many built-in default parameters which can be used for the unknown parameters. The values are based on the applicable military specifications or on mathematical expressions. An analysis can be performed initially using the default values so that the major problem areas can be determined and corrective measures taken. Later, when the actual data and specifications become available, new analyses can be made to update the previous ones. Each program prints a summary of all data, including any default values that were inserted. Thus, a record of the data on which the analysis was based is provided. All default values are identified in the printout.

50.2 **ANTENNA-TO-ANTENNA ANALYSIS CODES.** Several codes designed specifically for aircraft antenna-to-antenna interference analyses are available. ATACAP as described in the previous section is one of those codes. This section describes other antenna-to-antenna codes.

50.2.1 **AVIONICS INTERFERENCE PREDICTION MODEL (AVPAK3).** Under contractual tasks with the Federal Aviation Administration (FAA), ECAC developed AVPAK3 [C-32] to determine the mutual effects of introducing new avionics equipment to an existing airframe.

The analysis of the mutual effects of the operation of equipment on an airframe is accomplished by predicting the expected level of interference relative to the degradation threshold of each receiver. Antennas are assumed to be isotropic and may be located on the aircraft or on a neighboring aircraft. Equipment examined for possible interference includes those with overlapping or immediately adjacent operating frequencies, and also, those with harmonically related operating fre-uencies for which inadequate transmitter harmonic attenuation exists. Nonlinear effects are not included in the analysis and must be dealt with manually. To ensure that only far-field coupling conditions are considered, only those equipments operating above 30 MHz are treated.

The model offers the option of either a purely deterministic calculation or a probabilistic calculation that estimates the probability of interference. The body of the airframe is modeled as a cylinder of finite length to which appropriate conical sections may be added. The calculation takes into account the factors of airframe curvature, airfoil obstructions, and bulkhead obstruction.

In addition to interference analysis, the model calculates the power density at user-specified points, resulting from the operation of transmitters located on an aircraft. These points may be located anywhere on the airframe, including wing pods, or they may be "raised" from the airframe (i.e., not lying on the fuselage skin), including locations on neighboring aircraft. The model also has the capability of calculating a cumulative power density due to the effects of more than one transmitter.

An empirical method to compute airfoil obstruction loss was developed for AVPAK. The obstruction loss is determined by calculating the free-space loss around the airfoil and adding a curvature correction factor.

50.2.2 **ADVANCED ATACAP PLUS GRAPHICS (AAPG).** During antenna-to-antenna analyses, the geometry calculations required to define paths between antennas are difficult to verify as to correctness. The ATACAP and IEMCAP provide the user with large quantities of data which requires careful and time consuming examination. Puthermore, the design of compatible systems involves several iterations of: (1) program execution, (2) output evalution, and (3) system redesign. Each iteration recomputes the majority of parameters which remain unchanged from one run to the next. This is not efficient use of computer time or of the analyst's time.

The AAPG C-33 has been developed to make the user-computer interaction more natural, rapid, and productive. A user now communicates with the computer program via a collection of graphical input/output modules whereby he receives almost instantaneously plots, illustrations and tables of coupling path information. This approach allows the user considerable flexibility in redesign and in determination of interference margins.

50.3 **INTERACTION AND COUPLING CODES.** The interaction and coupling codes supplement the system level codes. If there is a requirement to perform a more accurate or detailed analysis of a separate response, the interaction and coupling codes can be utilized. Numerous interaction and coupling codes exist. Bevensee, et. al., [C-34] has done an extensive classification and cataloging of these codes. Other information on coupling codes may be found in the DNA EMP Handbook, Volume 6 -Computer Codes [C-35]. Some of the interaction and coupling codes are highlighted here using Bevensee's classification for code names.

50.3.1 **THIN-WIRE CODES.** Thin-wire codes are based on the method of moments computational technique. These codes are most applicable to wire antennas in free space or to wire antennas protruding from flat surfaces. These codes are also used to compute induced currents on solid bodies through wire-grid approximations. Two well-known thin wire codes are the NEC and the WF-OSU.

- NEC -- The Numerical Electromagnetic Code (NEC) [C-36] [C-37] developed at the Lawrence Livermore Laboratory under the sponsorship of the Naval Ocean Systems Center and the Air Force Weapons Laboratory. NEC is a user-oriented computer program for the analysis of interactions of EM waves with conducting structures. The program is based on the numerical solution of integral equations for the currents induced on a structure by an existing field.

  NEC combines an integral equation for smooth surfaces with one for wires to provide convenient and accurate modeling of a wide range of structures. A model may include nonradiating networks and transmission lines, perfect and imperfect conductors, lumped element loading, and ground planes which may be either perfectly or imperfectly conducting.

  The excitation in NEC may be either voltage sources, plane waves of linear or elliptic polarization, or fields due to a Hertzian source. The output may include induced current and charge densities, near- or far-zone electric or magnetic fields, and impedance or admittance. In addition, many of the commonly used quantities such as gain, directivity, and power budget are also available.

- WF-OSU -- The WF-OSU [C-38] [C-39] code was developed at the Electro Science Laboratory of the Ohio State University. The code is used to perform a frequency domain analysis of thin-wire antennas and scatters of arbitrary geometry. The computer model is a piecewise linear representation of the actual geometry so that the structure is approximated by straight-wire segments. The segments may have finite conductivity, lumped loads, and/or lossy insulating sleeves. The homogeneous, isotropic, ambient medium may be a lossy dielectric. Antenna computations include current distribution, input impedance, radiation efficiency, gain, far-field patterns, and near-zone fields.

50.3.2 **SURFACE CODES.** Surface codes are used to solve boundary value problems over solid surfaces. Depending on frequency, the models have been developed around solid surface integral equations, thin wire integral equations, and quasi-optical techniques. Two surface codes are the S3F-SYR and the G3F-TUD1.

- S3F-SYR -- This code was developed by Syracuse University [C-40] and is used to solve in the frequency domain radiation and scattering from bodies of revolution. The bodies may be solid, open with zero-thickness shells (such as open cylinders) and may have points (cones) and edges (discs). The program is written for bodies in free space only. The S3F-SYR/LLL1 [C-41] is an extension of the S3F-SYR code that includes an option for computing electric near fields at selected test points.

- G3F-TUD1 -- This surface code was developed by the Technical University of Denmark [C-42] and uses GTD to obtain induced currents on simple models. The code has been used to compute radiation from an antenna near a closed, convex polyhedral satellite including direct, reflected, simply-diffracted and doubly-diffracted rays. The code computes the radiation created by a magnetic dipole at the observation point and invokes the reciprocity principle to find the surface currents caused by an incident wave. This is the usual technique for finding surface currents with GTD.

50.3.3 **APERTURE CODES.** The classes of apertures that are of interest can be categorized by: (1) apertures in planes, (2) apertures in two or three dimensional bodies, and (3) apertures with wires behind them. Two aperture codes are the DASC and the BOR3.

- DASC (Diffraction by an axially slotted cylinder). This code was developed at the Harry Diamond Labs [C-43] and was written to yield basic information concerning the scattering of monochromatic plane waves by a missile-like body with an axial slot. The missile body is simulated with a cylinder of infinite length.

- BOR3 -- Syracuse University [C-44] developed this aperture code. The code incorporates a method for predicting the field penetrating a circumferential opening in a body of revolution and is based on the method of moments and an aperture equivalence theorem. The code has been applied to missile-like cavities illuminated by an obliquely incident plane wave.

50.3.4 **CABLE CODES.** The cable or transmission-line codes treat cables, wires and many antennas by using circuit-type equations to express the currents on the wires and potentials between the wires. Some codes operate in the time domain while others use Laplace or Fourier transforms. Two principal types of transmission line configurations are treated: (1) injected signals propagating down the line in the transverse electromagnetic mode and (2) longitudinal electric-field coupling into the lines. The codes all assume that the wavelengths of interest are large compared to transverse dimensions of the line.

A number of cable codes have been developed for transient analysis intended primarily for EMP analysis. In many cases the codes are for special configurations such as underground cables or cables above lossy earths and the codes are not adaptable to a generalized EMR analysis. However, it is conceivable that transient analysis may be used in EMR analysis in special cases, such as for short pulse radar environments. For this reason, several cable codes that utilize transient analysis are included here.

- NLINE (N-Conductor Transmission Line) -- NLINE was developed by the Harry Diamond Laboratories [C-45] and is used to compute the currents and voltages induced by an incident electromagnetic field at the terminations of a multiconductor transmission line.

- TART -- This code was developed by BDM [C-46] to calculate the transient voltage or current response at the terminals of a dipole, monopole, or loop antenna, or a two-wire transmission line excited by a general transient electromagnetic field.

50.3.5 **CIRCUIT ANALYSIS CODES.** Circuit analysis codes incorporate models of resistors, capacitors, inductors, transistors, tubes, etc., and are used to perform a detailed analysis of coupled EM energy into the circuit level. Such an analysis may use the component data and equivalent circuits found in Appendix E. Generally circuit analysis programs can treat ac and dc circuits and can be used for frequency sensitivity calculations. Codes are available for the frequency and time domains. Two circuit analysis codes are SCEPTRE and TRAFFIC.

- SCEPTRE (System for Circuit Evaluation and Prediction of Transient Radiation Effects) -- The code was developed by IBM under the sponsorship of the Air Force Weapons Laboratory and the Defense Nuclear Agency [C-47]. SCEPTRE analyzes the transient and frequency response of electronic circuits. The code uses a free format, problem oriented language to describe circuit topology.

- TRAFFIC (Transfer Function for Internal Coupling) -- TRAFFIC [C48] is a component of the larger PRESTO program. Both were developed by the Boeing Aerospace Company under Defense Nuclear Agency sponsorship. TRAFFIC uses a nodal admittance matrix approach to solve linear circuits at user-specified frequencies. The calculations are performed in the frequency domain. Sparse-matrix techniques are used for efficient solution of large networks.

60. **INTERACTION AND COUPLING APPLICATIONS.** While rigorous solution of typical EMR interaction and coupling problems is largely impractical, results of engineering accuracy are obtainable using canonical models. A canonical model [C-49] is a structure having a simple configuration that is used to represent the more complex real world structure. For example, the straight-cylinder canonical model may be used to represent a missile from an electromagnetic standpoint. An aircraft may be represented by a combination of canonical models such as a cylinder and flat plates. It is the task of the analyst to select canonical models for the real world problem he is attempting

to solve and to determine how "good" the model actually represents the real world problem. The goodness evaluation may be done by data comparison with measured data from other systems or by data comparison with other computer models. In some cases data can be obtained with scale models.

In selecting canonical models, the analyst must consider the size of the structure in terms of the wavelength. For example, stick models (connected thin-wires) may be adequate to study responses on an aircraft when the aircraft is small in terms of a wavelength. However when the aircraft is large in terms of a wavelength, the stick model is not appropriate since circumferential currents and diffraction from wing edges must be included. As another example, consider the aperture problem. The fields and currents around a small aperture may be treated as localized when the aperture is in a structure that is large in terms of a wavelength. When the structure is small in terms of a wavelength, localized treatment is not appropriate since resonances on the structure may affect the current distribution in the vicinity of the aperture.

Another approach in lieu of the canonical model technique is the use of worst-case models that establish an upper bound for a response. The approach is particularly useful when little information is available about the functions and geometry of a system, as in the early stages of development. The tendency with this approach is to over harden or over protect a system thus causing unnecessary adverse effects on system performance. The analyses must be refined through more detailed analysis or testing as a system progresses in its development.

An important consideration in interaction and coupling applications is the degree of accuracy in the geometric data describing a system. Many times, the coupling paths into a system are inadvertent and not truly definable. Consider a small inadvertent aperture with a diameter (d) located in a structure. The coupling through the aperture is proportional to d^3. A small inaccuracy in the dimensions of the aperture can have a significant effect on the coupling calculation accuracy. In situations like this, the best approach may be to initially assume an external structure provides no shielding protection, and when hardware becomes available, refine coupling estimates based on shielding effectiveness tests.

60.1 **BASIC MODEL RESPONSE.** This section treats techniques and basic model responses that may be used in interaction and coupling applications. The basic models may be used to obtain coupling estimates (order of magnitude type calculations) early in the system development or may be used to cross check computer code calculation results.

60.1.1 **TRANSFER FUNCTION TECHNIQUE** [C-50] [C-51]. The kind of problem to be solved is illustrated in Figure C-8. A survey of the initial system design will identify certain holes, domes, access doors, etc., in the missile skin which are possible severe ports of entry (POE). Thus, it is desired to determine the response of a component in a system to an electromagnetic field, E_i, incident on the system. The only POE illustrated is a side opening. To describe the coupling of high

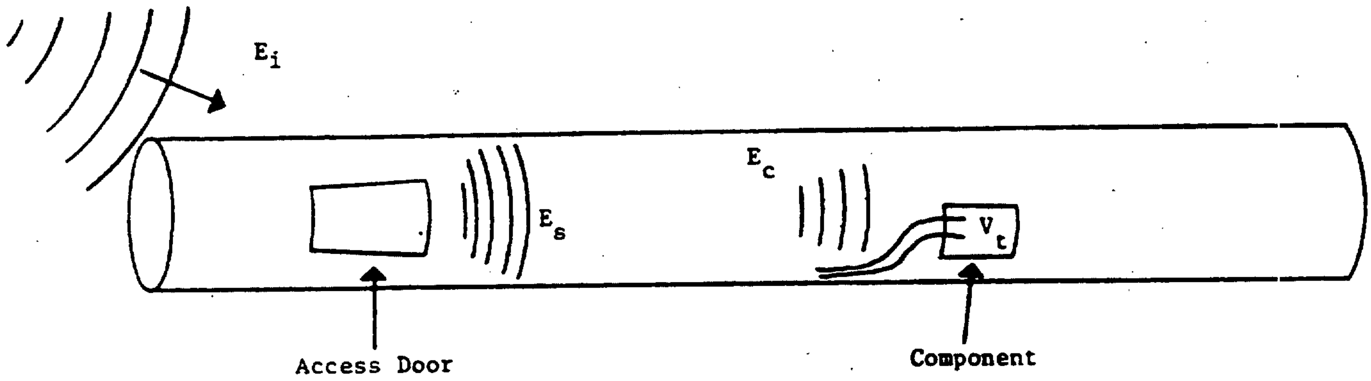


Figure C-8. Aperture Coupling Approximation [C-28] frequency electromagnetic energy to the component of interest, the amount of coupling through the aperture in the system skin must first be determined. Next, the way in which the energy is distributed through the system, the degree to which the energy couples to system cables, and the amount of loss encountered in conducting the energy along cables are determined.

The purpose of the coupling analysis is to determine the relationship between fields (E_i) incident on the system and the response (V_t) in various components of the system. It is convenient for analysis purposes to break the transfer function V_t/E_i into smaller parts. One possible way to subdivide the overall transfer function is shown in Equation C-15:

$$\frac{V_t}{E_i} = \frac{E_s}{E_i} \times \frac{E_c}{E_s} \times \frac{V_t}{E_c} \qquad \text{(C-15)}$$

There is an implicit assumption in this relationship that each response is unaffected by other responses. In other words, the reaction of the interior region does not affect the current distribution about a POE. The first term on the right side of Equation C-15 represents the relationship between the field incident on the system (E_i) and the field present inside the system (E_s). To determine the ratio of fields inside the missile to incident fields, techniques to analyze the coupling of electromagnetic energy through the various kinds of apertures are required.

The second term of the transfer function relates the field (E_s) inside the system near some POE to the field (E_c) incident on the cable connected to the component of interest. This part of the coupling analysis requires that the fields coupled into cavities with complex internal geometries be determined. Theory to rigorously define such fields is under development [C-26]. The method to be used in this approximate analysis will make use of the transmission characteristics of apertures in infinite conducting planes to develop a first order approximation for this term of the transfer function.

The final term relates the field (E_c) incident on the component wiring to the response (V_t) of the component. This term will depend on the receiving characteristics of the component and its wiring. Ideally, the component wiring can be represented by antenna or transmission line models and the component can be replaced by an equivalent load.

In analyzing specific systems, it may be necessary to expand the terms of the transfer function to account for different internal structural arrangements. For example, if a component were located inside a metal package, it might be desirable to add a term to the transfer function to account for coupling of energy through the package POE's. Similarly, if the point of exposure of the component cable is a significant distance from the component terminals, the cable attenuation must be accounted for by an additional term.

Once the transfer function relating component response to incident field has been established, it can be used with degradation data for the component under study (V_f) to determine the field (E_f) at which failure may be expected as indicated in the following equation:

$$V_f \times \frac{E_i}{V_t} = E_f \qquad \text{(C-16)}$$

By using the transfer function approach, the analysis may be made as general as is desired. The magnitude of each term of Equation C-15 can be determined in the same manner regardless of the type of component under study. When studies of the susceptibility of other components are conducted, the only quantities that change are the final term of Equation C-15 and V_f of Equation C-16.

If some simple, approximate models are assumed to describe the interaction of electromagnetic energy with systems, the maximum worst-case response of the components can thus be computed. These models permit the rapid, conservative estimation of the effects of particular electromagnetic environments on systems. The models to be discussed are for the high-frequency range (above 100 MHz) where coupling is primarily controlled by openings in the system skin.

Consider the response characteristics of apertures in conducting surfaces. Diffraction theory predicts that the field intensity on the shadow side of an aperture in an infinite conducting screen increases (approximately) directly with frequency up to the frequency at which the aperture becomes resonant. (The resonant frequency for an aperture is approximately that frequency at which its characteristic dimension is one-half the wavelength of the incident radiation.) At frequencies above resonance, the field intensity on the shadow side of the aperture is essentially equal to that of the incident wave.

These response characteristics are illustrated in Figure C-9. In this figure, the transmission coefficient (defined as the ratio of total power radiated by the aperture to the power of the incident wave) of an elliptical aperture is plotted. At frequencies above resonance, the value of the transmission coefficient is approximately unity; that is, all the energy striking the aperture area passes through. The point at which the value of unity is first reached is at $2a/\lambda \approx 0.5$. Below resonance, the transmission coefficient of the aperture varies (approximately) directly with frequency.

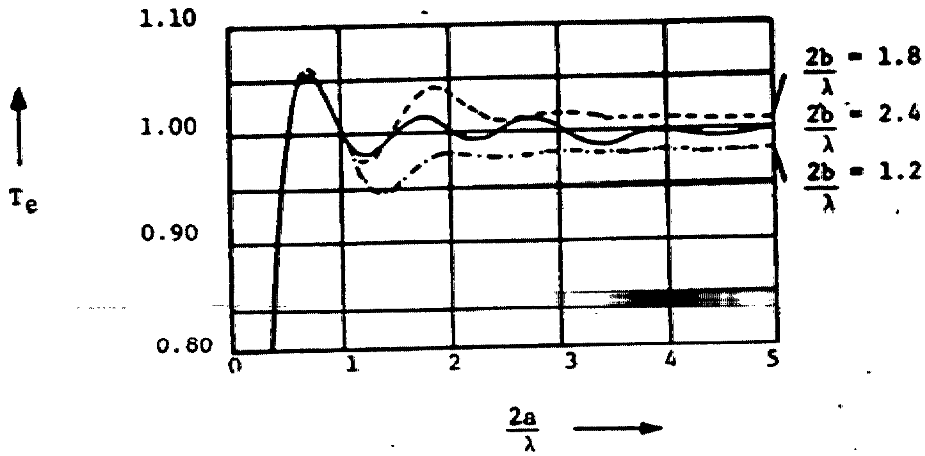


Figure C-9. Transmission Coefficient (T_e) of an Elliptical Aperture in a Conducting Screen. (a = ellipse major axis, b = ellipse minor axis) [C-51]

The model used for the coupling through an aperture in a conducting plane is expressed in terms of the transfer function Equation C-15 as:

$$\frac{E_s}{E_i} = \begin{cases} \frac{f}{f_o} & \text{for } f \le f_o \\ 1 & \text{for } f > f_o \end{cases} \qquad \text{(C-17)}$$

where f is the frequency and f_o is the resonant frequency of the aperture. The approximate transfer function is then represented by the curve of Figure C-10. This transfer function can then be used to estimate the magnitude of fields inside a missile skin in the near vicinity of an aperture. These interior fields then couple to cables and wires which are connected to the critical circuits in question. The next step in the approximate transfer function definition is that of describing the interior coupling.

The two remaining terms of Equation C-15 (viz. E_c/E_s and V_t/E_c) are discussed next. Two frequency regions will be discussed for the second of these two terms; a low frequency region where wavelengths are longer than the circuit wire and cable dimensions and a high frequency region where these dimensions are comparable to wavelength. First, consider the term E_c/E_s.

The maximum response should occur when cables connected to the component of interest are uniformly exposed to the field just behind the aperture. Therefore, for the model, the second term of Equation C-15 is taken to be unity; that is:

$$\frac{E_c}{E_s} = 1 \qquad \text{(C-18)}$$

The final term (V_t/E_c) of the transfer function represents the response of the component wiring to the local field inside the system. It is necessary to determine the internal coupling characteristics for cables which are, in general, long in terms of wavelength. As might be expected, the actual response of internal wiring is extremely complicated. However, approximations may be made which should yield reasonable estimates of the maximum coupling.

In the <u>low frequency region</u>, where circuit dimensions are small compared to a wavelength, transmission line models are used. Paul [C-52] has given an extensive discussion of a field-to-wire coupling model which includes the transmission line model.

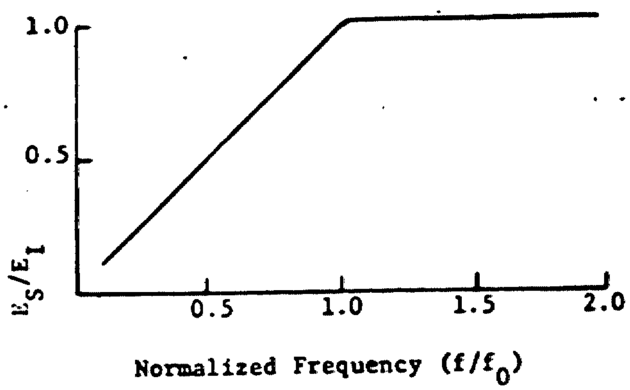


Figure C-10. Approximate Aperture Transfer Function. [C-51]

In the high frequency region, the estimation is made by assuming that the maximum expected coupling is no greater than that of a tuned half-wave dipole for each frequency in question. In this case, the Thevenin voltage (V_t) is given as:

$$\frac{V_t}{E_c} = \frac{\lambda}{\pi} = \left(\frac{c}{\pi}\right)\frac{1}{f} \tag{C-19}$$

Then combining Equations C-19, C-18, and C-17 gives a total transfer function for the high frequency response:

$$\frac{V_t}{E_i} = \begin{cases} \dfrac{c/\pi}{f_o} & \text{for } f < f_o \\ \dfrac{c/\pi}{f} & \text{for } f > f_o \end{cases} \tag{C-20}$$

where f_o is the resonant frequency of the aperture. This total transfer function is shown in Figure C-11.

An alternative approach for the high frequency coupling problem is based on directly estimating the power coupled to a resonant, matched dipole. This method uses the above relation in Equation C-19 to determine the effective aperture of the tuned dipole to be:

$$A_{eff} = 0.13\lambda^2 \tag{C-21}$$

The terminating impedance is assumed to be the conjugate of the antenna impedance and antenna losses are neglected. Then the maximum power an unshielded wire will pick up is given by:

$$P = A_{eff} P_i$$

or

$$P = 0.13\lambda^2 P_i \text{ (watts)} \tag{C-22}$$

which is the well known result for a half-wavelength dipole. Again, P_i is the power density incident on the interior subsystem of the missile. Above the resonance of the largest aperture, this is simply the power in the incident field exterior to the missile.

A second assumption for very high frequencies ($\lambda \leq$ circuit dimensions) would be to assume that the effective area of the complete cir-

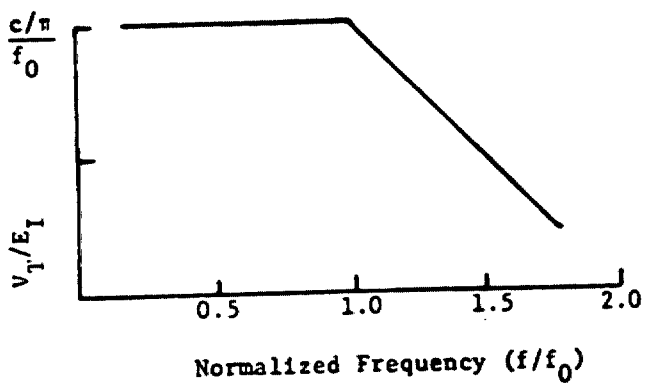


Figure C-11. Over-all Transfer Function. [C-51] cuit was equal to its actual area and that all power incident on the circuit was absorbed. For a given circuit, the larger of the two effective areas should be used to assure a conservative prediction.

60.1.2 **DIPOLE MODELS.** The electric field dipole model was discussed in the previous section. This model is often used in worst-case type calculations to bound the coupled power. For this reason, futher insight into the dipole model is warranted.

Figure C-12 shows an equivalent circuit [C-53] for the dipole model. Z_A is the dipole impedance which is a function of frequency and Z_T is the terminal or load impedance connected to the dipole. The equation $V = hE$ is the relationship between the voltage induced in the dipole and the incident electric field on the antenna. The effective height is the quantity h. The maximum effective aperture of a matched dipole is given by the equation:

$$A_{em} = \frac{V^2}{4PR_r} \tag{C-23}$$

where V = induced voltage (volts)
P = incident power density (watts)
R_r = radiation resistance (ohms).

For a lossless, resonant (half-wavelength) dipole in free space with a matched load, R_r is 72 ohms. The effective height (h) of a half-wavelength dipole is equal to λ/π . The incident power density (P) is equal to $E^2/120\pi$. Substituting these quantities into Equation C-23 gives:

$$A_{em} = \frac{V^2}{4PR_r} = \frac{120\pi}{4\pi^2}\,\frac{E^2\lambda^2}{E^2 72}$$

$$= \frac{30\lambda^2}{72\pi} = 0.13\lambda^2 \tag{C-24}$$

This is the relationship given by Equation C-21 in the previous section. Notice this is the maximum effective aperture and to obtain this value the dipole must be matched with 72 ohms.

Next it is instructive to consider a small dipole ($\ell << \lambda$). The effective height of the small dipole is approximately equal to its length. The maximum effective aperture of a short dipole is given by:

$$A_{em} = 0.119\lambda^2 \tag{C-25}$$

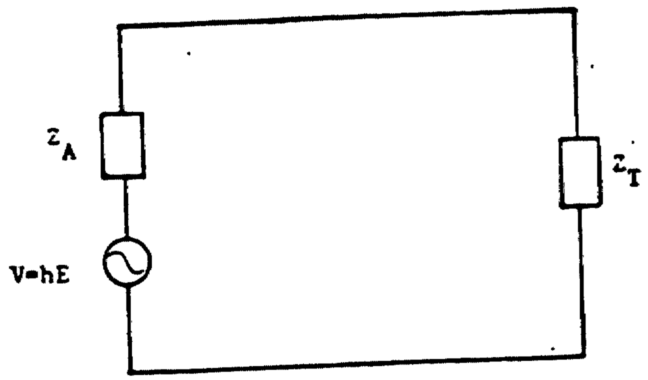


Figure C-12. Equivalent Circuit for Dipole Model.

The maximum effective aperture of the half-wavelength dipole is about 10 percent greater than that of the short dipole. The impedance of the short dipole is given approximately by the equation:

$$Z_a = \frac{80\pi^2 \ell^2}{\lambda^2} - j\,\frac{60}{\pi}\,\frac{\lambda}{\ell}\left(\ln \frac{2\ell}{a} - 1\right) \qquad \text{(C-26)}$$

where ℓ = length (meters)
λ = wavelength (meters)
a = radius (meters)

The real term (radiation resistance) rapidly becomes smaller as the wavelength increases and the capacitive reactance term becomes larger as the wavelength increases. For maximum power transfer, the small dipole must be conjugately matched which means a large inductance (high-Q) must be added to the load.

Notice that the maximum effective aperture of a matched dipole and a matched isotropic antenna are not equal. The maximum effective aperture of an isotropic antenna is:

$$A_{em} = \frac{\lambda^2}{4\pi} = 0.079\lambda^2 \qquad \text{(C-27)}$$

The use of the matched half-wavelength dipole should be used with caution for low frequencies where the structure becomes small in terms of a wavelength. Swink [C-54] has done an extensive study of the straight-cylinder with aperture canonical model. His results show that the power delivered at low frequencies to a transmission line in a cylinder is much less than the power delivered by a matched half-wavelength dipole.

It is informative to derive the maximum available power from a fixed length dipole for frequencies from well below to well above the resonant frequency of the dipole. Figure C-13 is a curve of the power delivered to the matched load of a 4.57 meter long dipole with a one-volt per meter incident electric field.

60.1.3 **CYLINDRICAL MODELS** [C-2]. As previously discussed, the straight-cylinder structure can be utilized as a canonical model of a missile. To develop further insight of this model, the external current distribution and interior fields of a cylindrical model are considered.

The model considered is a perfectly conducting tubular cylinder of length (2h) and radius (a) shown in Figure C-14. It is assumed that the incident electric field is directed along the axis of the cylinder and that the radius of the antenna is small at the highest frequency of interest.

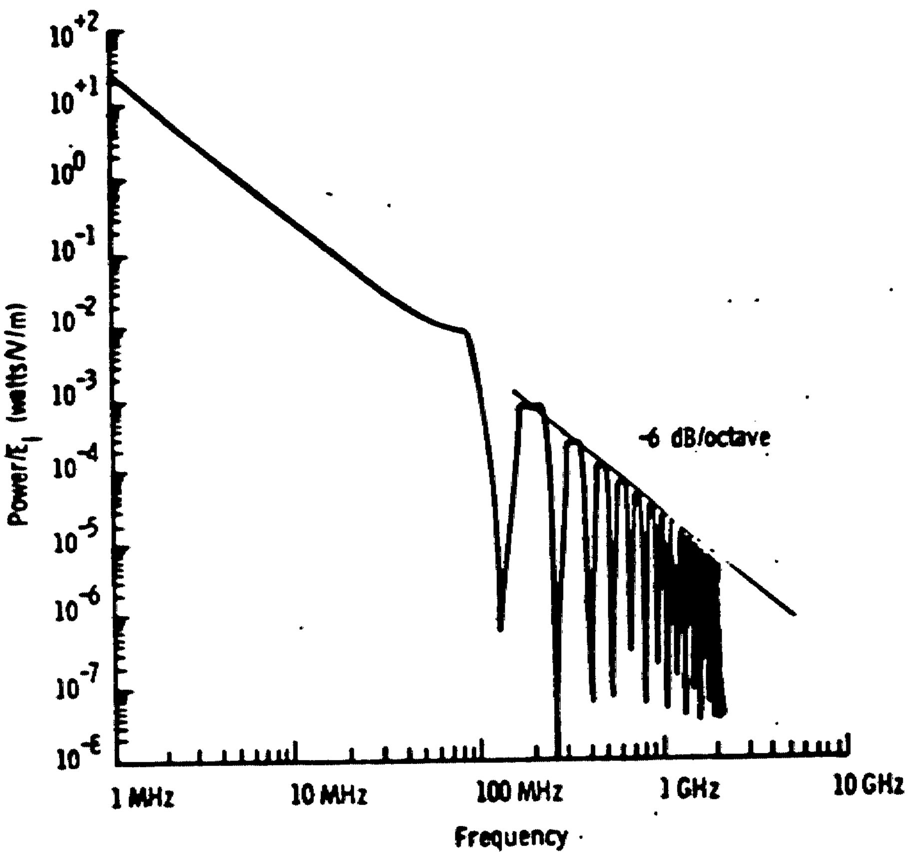


Figure C-13. Dipole Maximum Power Transfer. [C-51]

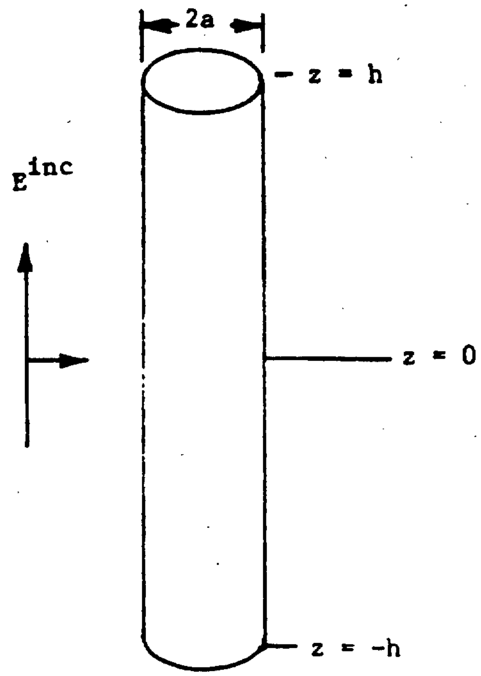


Figure C-14. Cylindrical Model [C-2]

It can be shown that the peak current at any point on a cylindrical body is proportional to the length of the structure for CW excitation. This allows normalization to be used in response functions. A variable used in the normalization is the cylinder fatness, traditionally specified as Ω where:

$$\Omega = 2 \ln \frac{2h}{a} \qquad \text{(C-28)}$$

A missile which might be modeled by a cylindrical body usually has a relatively small total length to radius ratio. The appropriate fatness factor Ω is about 5 or 6. Figure C-15 illustrates the dependence of the induced currents on Ω. Generally, the fatter cylinder carries more current.

The currents observed at five points along the cylindrical body with end-caps were computed using the finite-difference approach. The normalized CW transfer functions (magnitude) along the structure are displayed as a function of normalized frequency in Figure C-16.

To illustrate how the scattered field components on the outside of a cylindrical body affect the fields penetrating a small aperture, several examples are given, showing how the fields behind a circular aperture in the cylinder depend upon the frequency and the location of the hole in the cylinder. The geometry of the example is given in Figure C-17.

Fields near a small aperture ($k_o a \ll 1$) inside a finite body have been shown to be approximated closely by the fields leaking through the same size hole in an infinite sheet. This result allows classical diffraction theory to be extended to predict the fields near the hole inside a cylindrical body.

Formulas are provided here for selected scattered field components near an elliptical aperture in an infinite plate (see Figure C-18). The scattered field components are normalized to the electric field normal to the plate or the magnetic field component tangent to the plate <u>before the aperture was present</u>. More general formulas are available. We note the static $(1/r^3)$ dependence of the principle field components (E_x/E_{nor}, H_y/H_{tan}) near the aperture, and the frequency independence of the principle components. It should also be noted that internal fields are proportional to l_1^3 so that doubling the size of an aperture would result in an eight-fold increase in the interior fields.

The data for the CW fields behind a circular hole in the side of a finite cylinder are shown in Figures C-19 through C-21. These data reveal that for holes near the center of the structure, the relative importance of penetration from the H_{tan} component is greater than for holes near the ends of the cylinder. Near the end of the structure, cables that run away from the hole (but not past it) may be most strongly

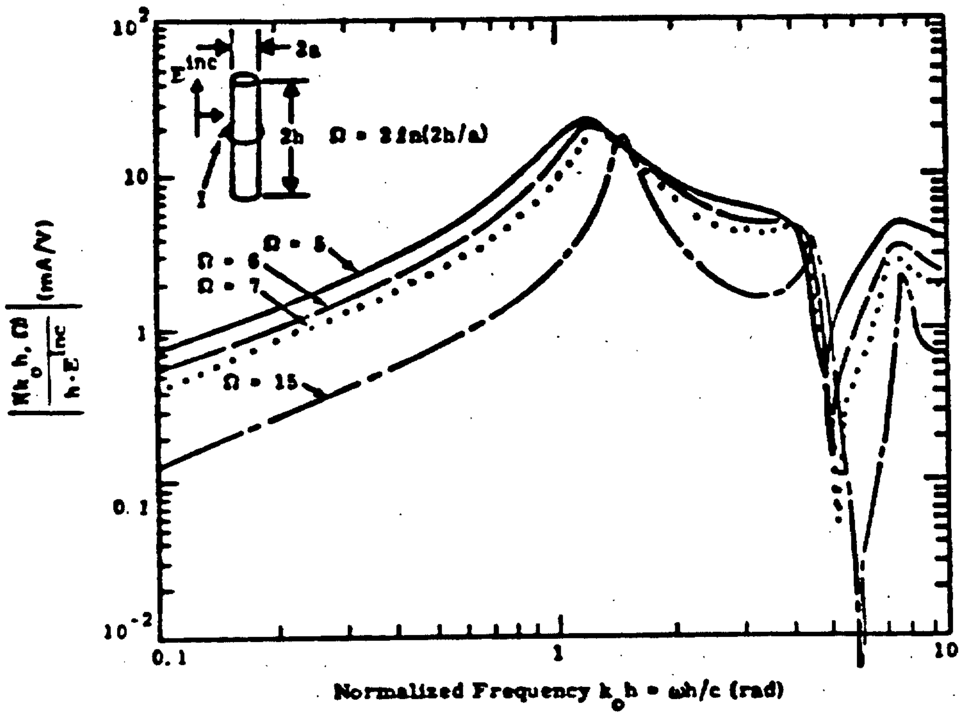


Figure C-15. Dependence of Center Current on Fatness. [C-2]

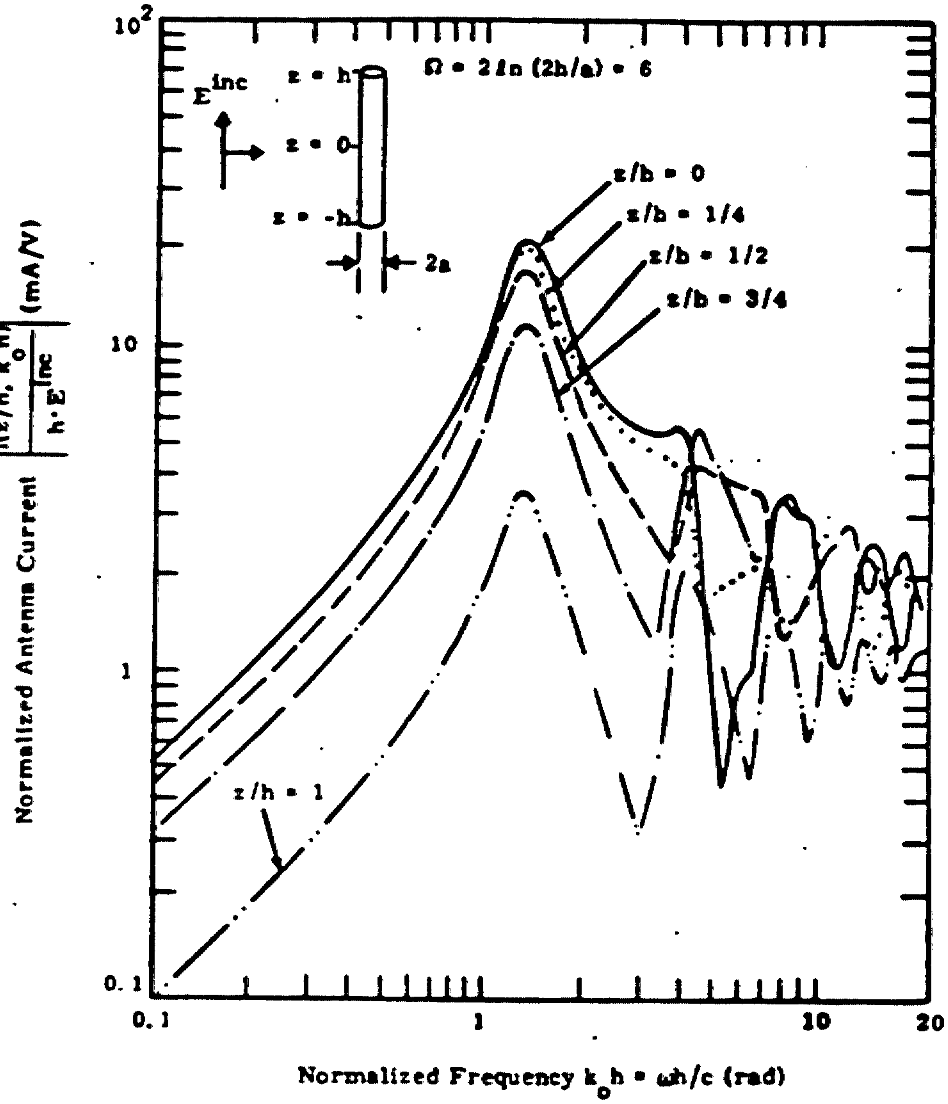


Figure C-16. Transfer Function for Total Axial Current on a Cylinder with Endcaps. [C-2]

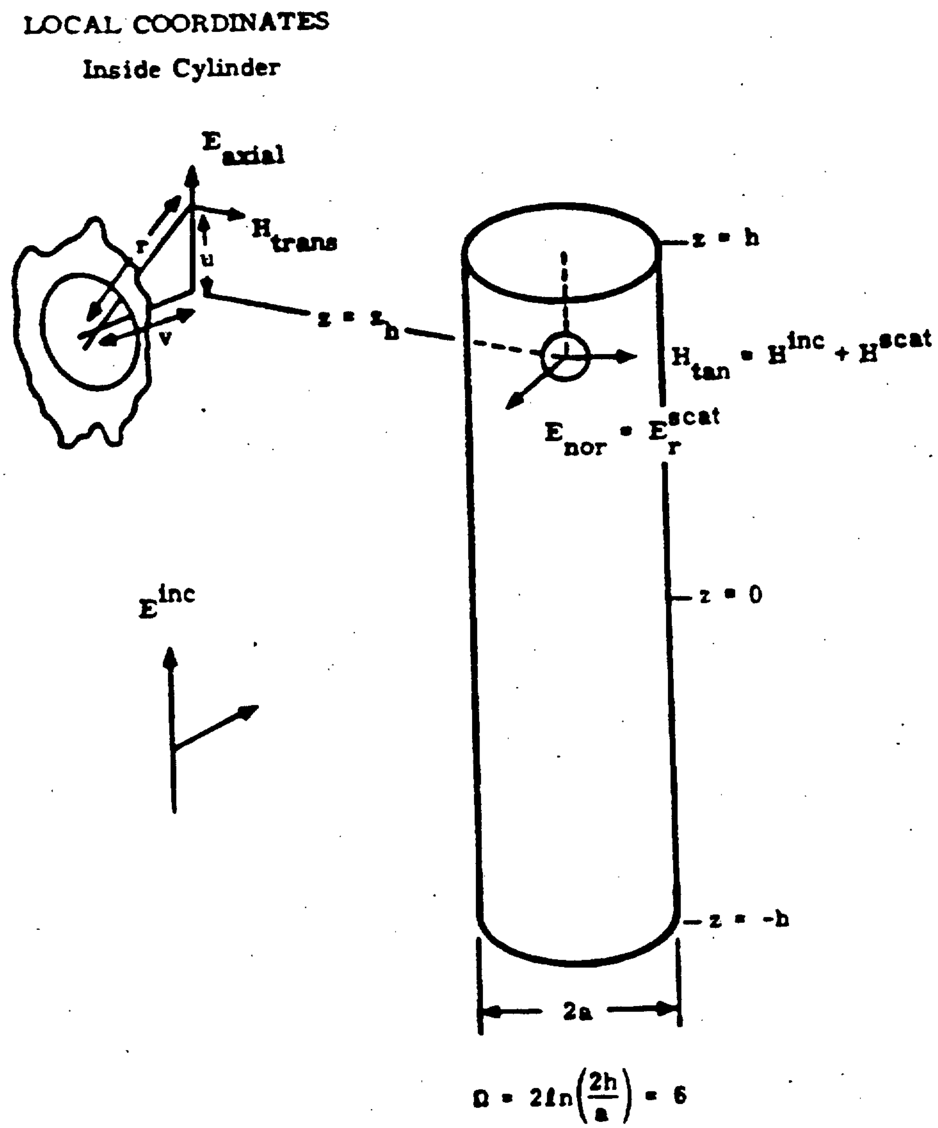
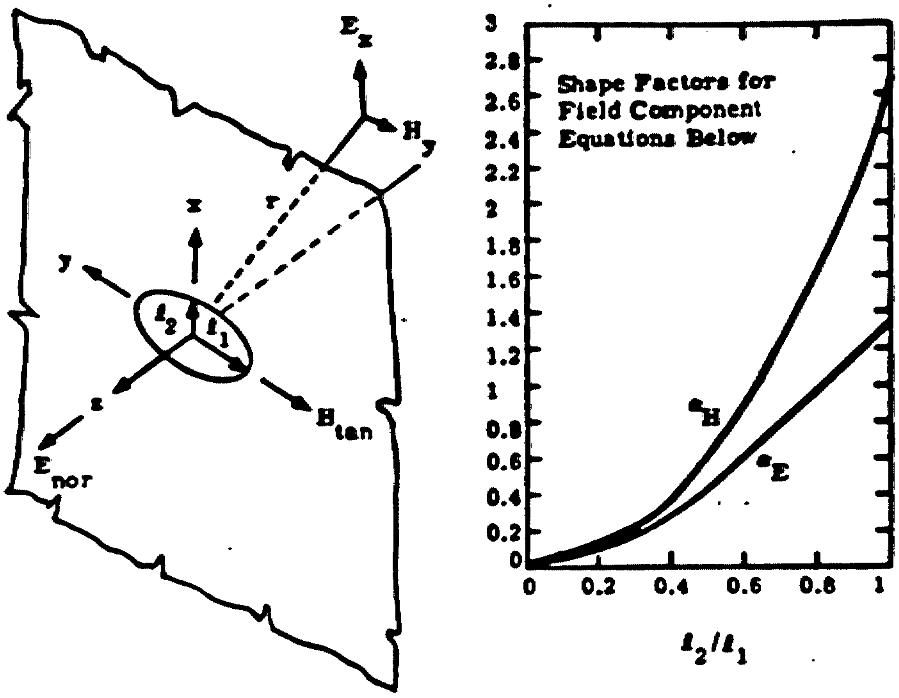


Figure C-17. Cylindrical Model for Calculating the Fields Penetrating a Circular Aperture in a Finite Cylinder. [C-2]

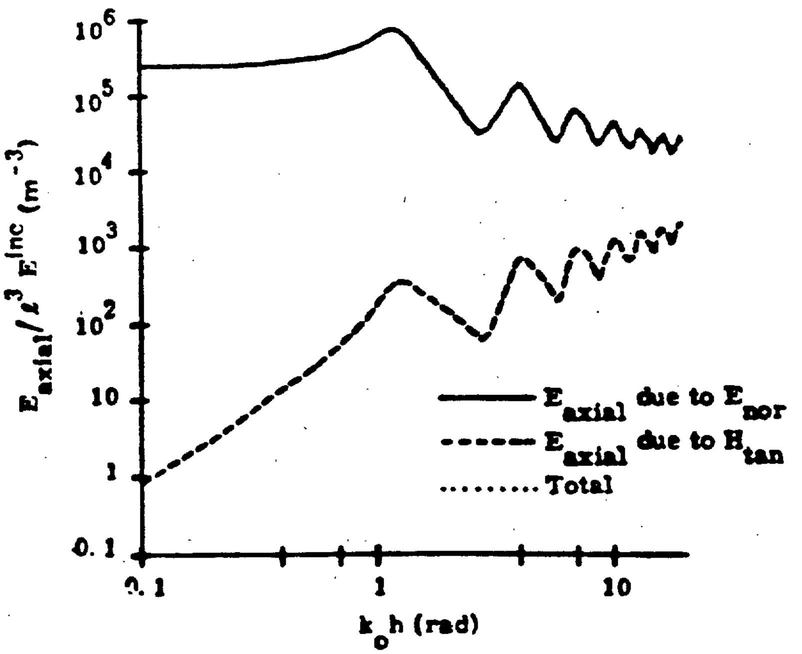


$$\frac{E_x(x,o,z)}{E_{nor}} = -\frac{\alpha_E\,\ell_1^3\,xz}{r^5}, \qquad \frac{H_y(x,o,z)}{E_{nor}} = -\frac{\alpha_E\,\ell_1^3\,jk_o z}{480\pi^2 r^3}$$

$$\frac{E_x(x,o,z)}{H_{tan}} = \frac{30\,\alpha_H\,\ell_1^3\,jk_o z}{r^3}, \qquad \frac{H_y(x,o,z)}{H_{tan}} = \frac{\alpha_H\,\ell_1^3}{4\pi r^3}$$

$$r = \sqrt{x^2 + z^2}$$

Figure C-18. Selected Scattered Field Component for Diffraction Through an Elliptical Aperture in an Infinite Perfectly Conducting Sheet [C-2]



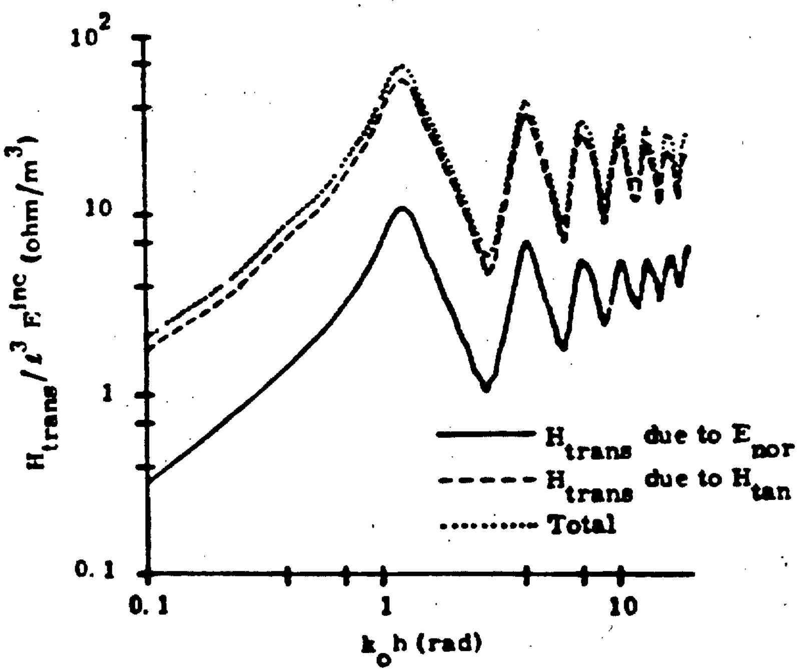
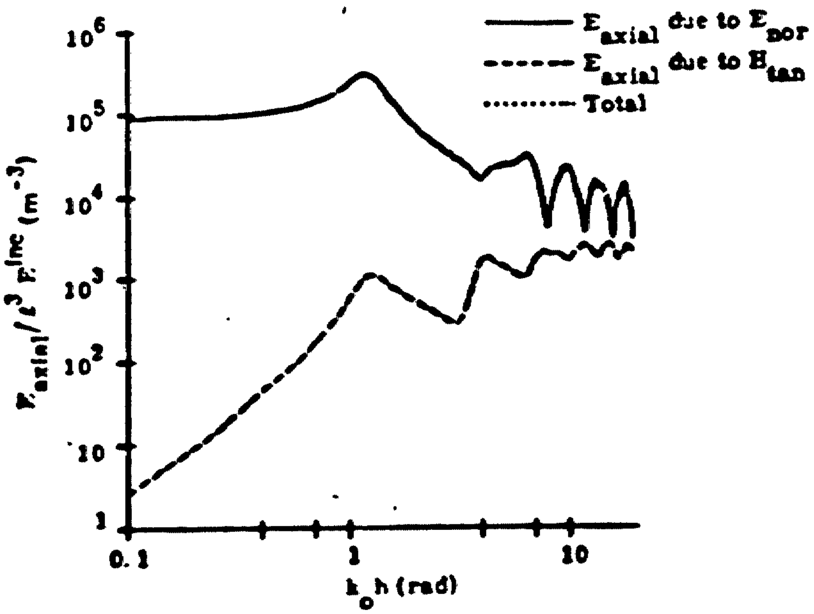


Figure C-19. Normalized CW Fields Behind a Circular Hole in the Side of a Finite Cylinder, $\Omega = 6$, $u = a/8$, $v = a/4$. Hole Located at $Z \sim h$. [C-2]



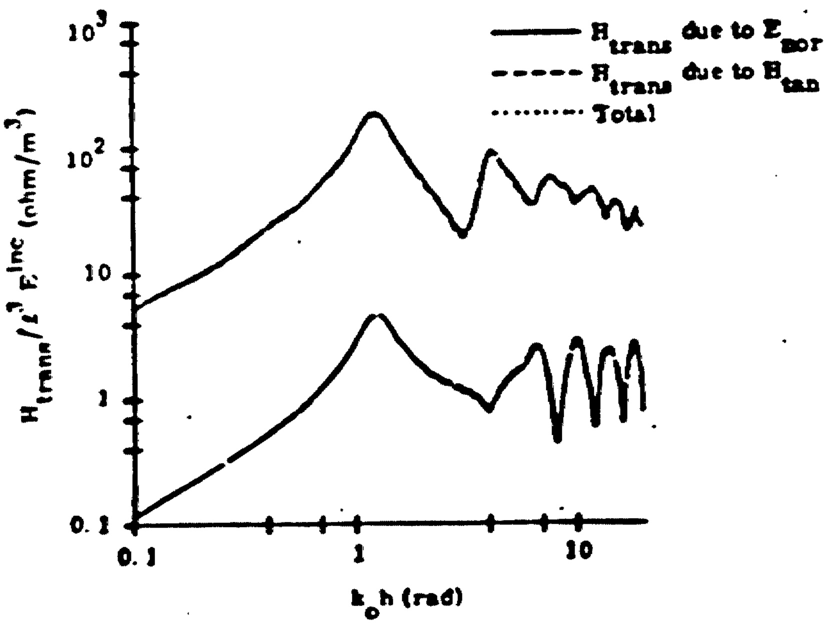


Figure C-20. Normalized CW Fields Behind a Circular Hole in the Side of a Finite Cylinder, $\Omega = 6$, $u = a/8$, $v = a/4$. Hole Located at $z = 3\ h/4$. [C-2]

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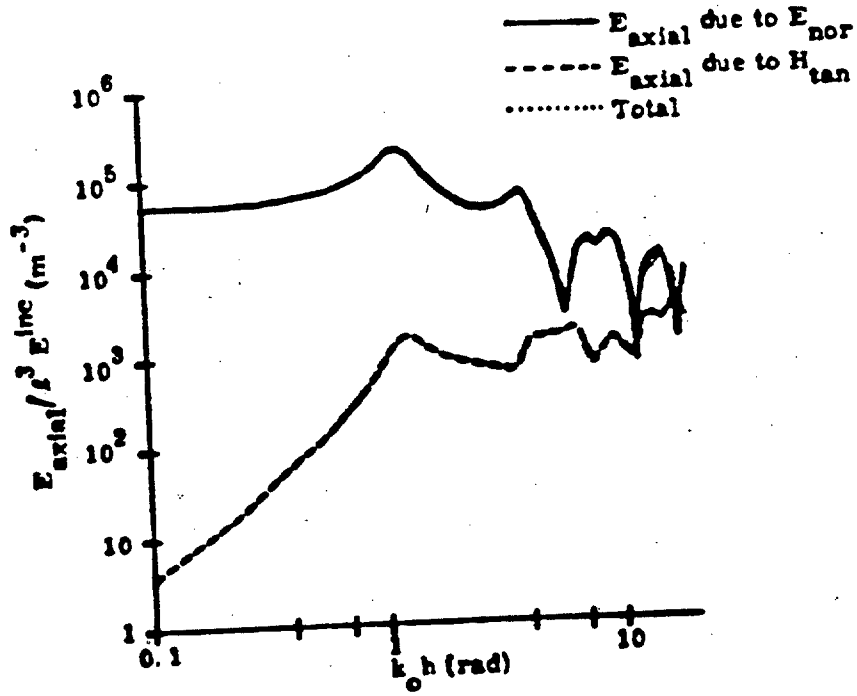

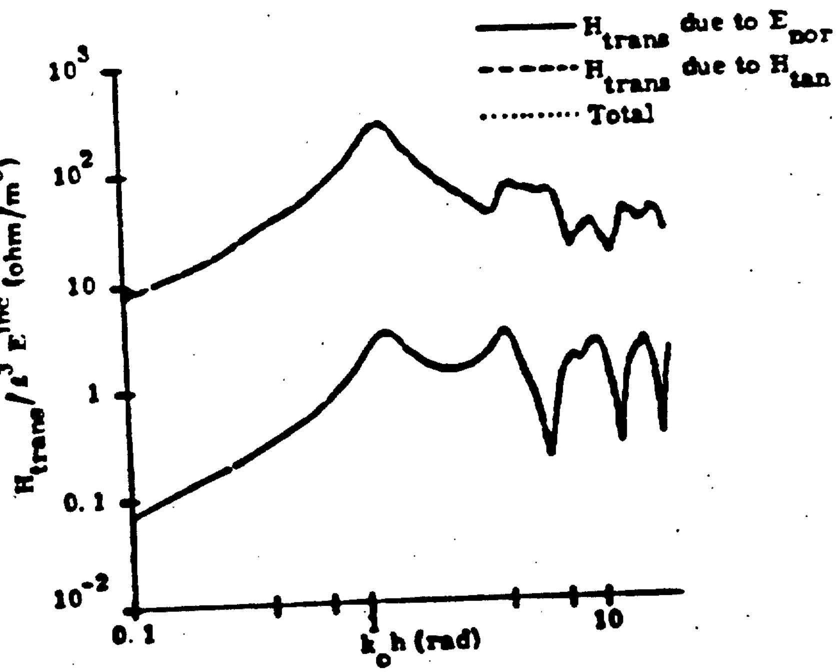

Figure C-21. Normalized CW Fields Behind a Circular Hole in the Side of a Finite Cylinder, Ω = 6, u = a/8, v = a/4. Hole Located at Z = h/2. [C-2] excited by the penetrating electric field. Lengthening the structure increases this effect, since the scattered electric field outside the cylinder is directly proportional to the length of the structure while the scattered magnetic field is not (if the fatness factor, Ω , is unchanged).

These predicted fields can be used to estimate the response of a small dipole or loop inside an aircraft or missile; however, extension of the available theory is required to obtain estimates of current in more complex cables. Holes near the end of the cylinder can be used to approximate penetration through detector domes in the nose of a missile. Holes at other locations can be used for penetrations such as canard drive openings, access ports, etc.

60.2 **SYSTEM RESPONSE.** The system response is determined by measuring or calculating a voltage or current at a terminal in the missile electronics. The voltage or current is induced by EM energy which illuminates the missile externally. The external environment may be characterized in terms of either peak or average values and the corresponding peak or average current defined at an electronics terminal. Generally, the system response is characterized through a "standard" response. A standard response is that level of external EM energy, as a function of frequency, required to produce a specified voltage or current at a circuit's terminal.

In order to present a clearer understanding of system level calculations, consider the system response of the straight cylinder model with a side aperture and a single transmission line located in the interior of the cylinder. The general shape of the cylindrical model coupling is shown in Figure C-22. The cylindrical model response is shown in comparison to a matched isotropic antenna. Next assume it is desired to calculate the rectified current in a component located at the end of the interior transmission line. Let the maximum rectification efficiency of the component be as shown in Figure C-23 (see Appendix E for this type of data). Next multiply the coupling curve of Figure C-22 and the component response curve of Figure C-23 to obtain the composite curve shown in Figure C-24. The units of the composite curve are $A/(w/m^2)$. Next determine the standard response necessary to produce 1.0 mA of rectified current at the output of the component. The standard response curve of Figure C-25 is obtained by dividing the 1.0 mA by the composite curve of Figure C-24 and converting the power density to dBm/cm^2. It is this standard response curve that can be compared to the EM environment profiles, or to measured data, to identify susceptibilities and hardening requirements.

70. **ANALYSIS METHODOLOGY AND PROCEDURES.** Now that the system considerations, computational methods, computer codes, etc., have been discussed, a fundamental analysis methodology/procedures will be outlined to better define the steps in an analysis and prediction effort.

70.1 **FUNDAMENTAL METHODOLOGY.** The fundamental methodology for an analysis and prediction effort may be represented by the basic blocks shown in Figure C-26. The procedures follow the general flow discussed

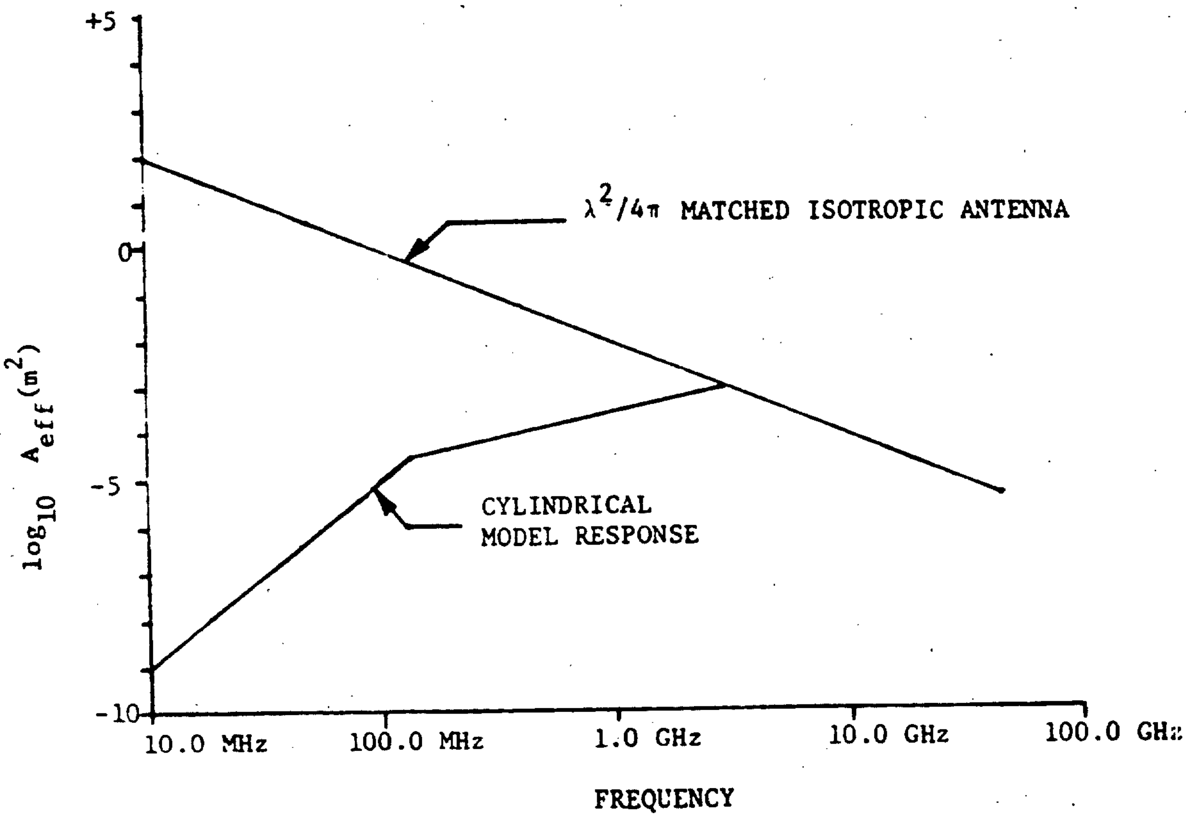

Figure C-22. Asymptotic Coupling Response of Cylindrical Model with Aperture. [C-24]

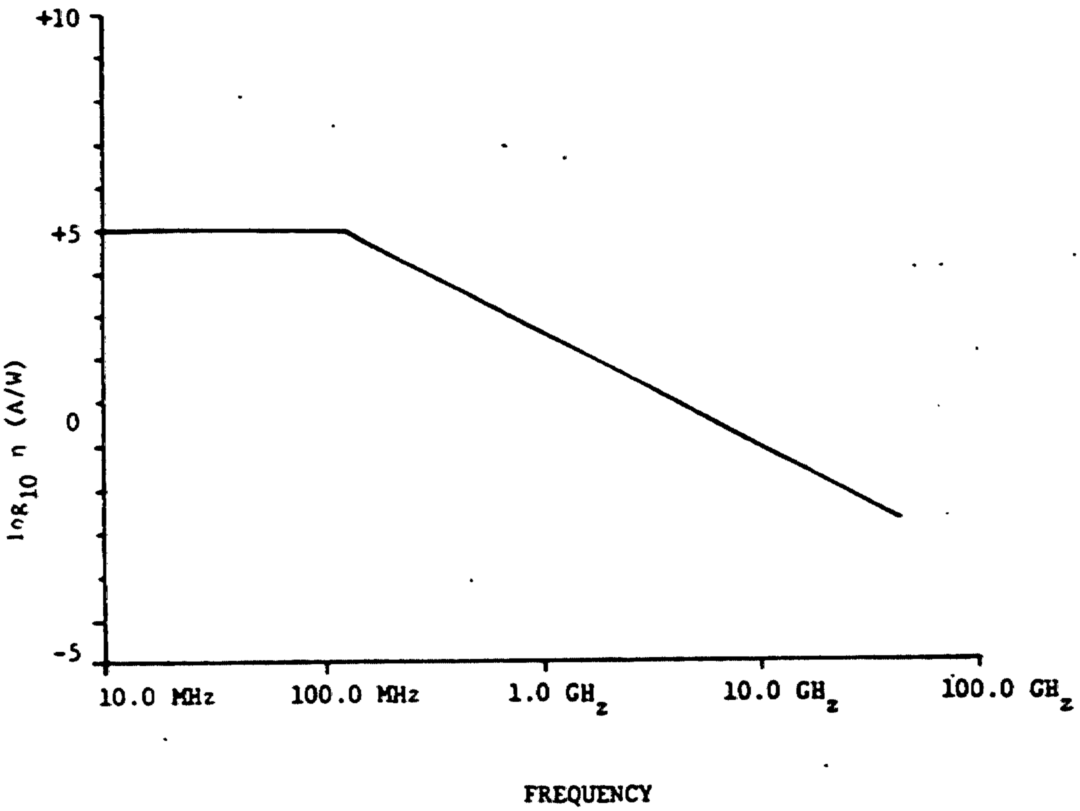

Figure C-23. Maximum Rectification Efficiency of a Component. [C-24]

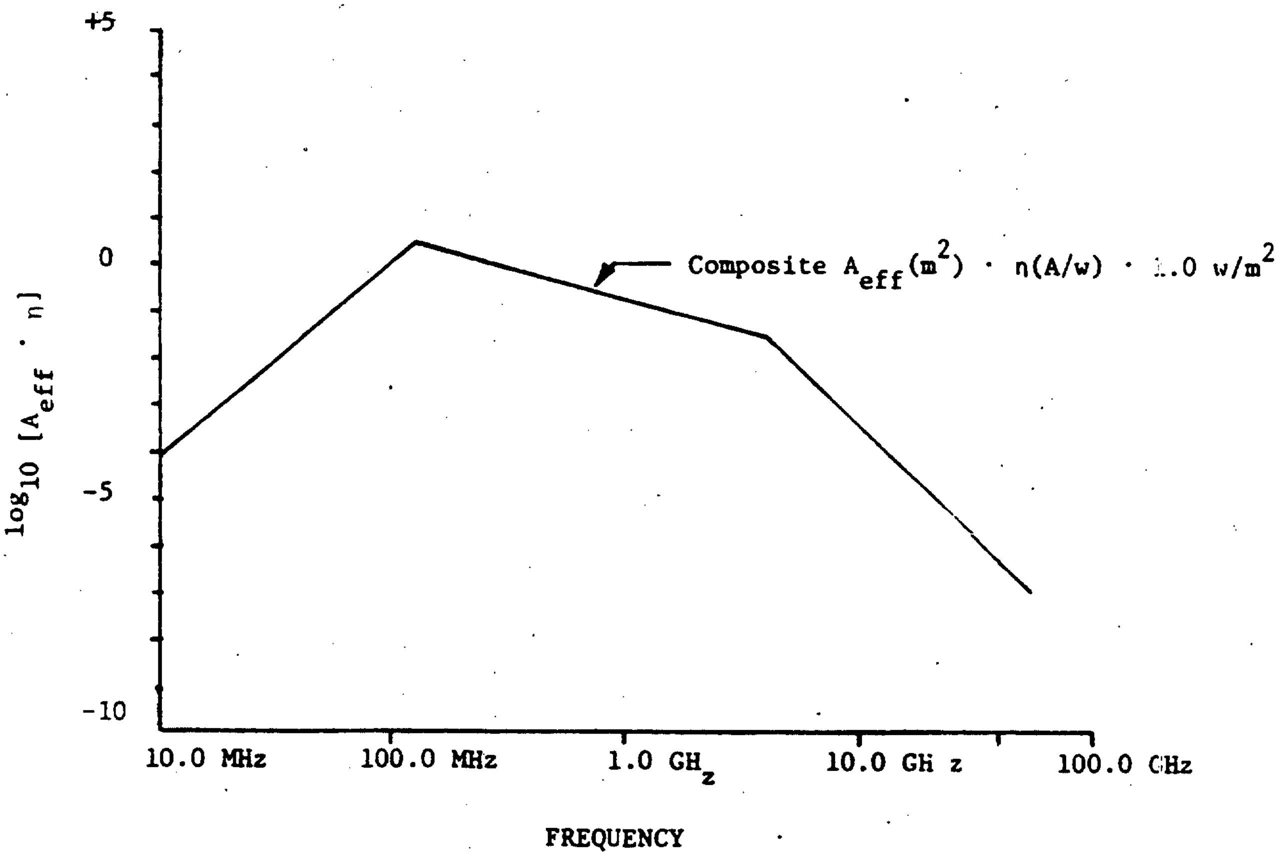

Figure C-24. Composite curve of Coupled Power and Component Response. [C-24]

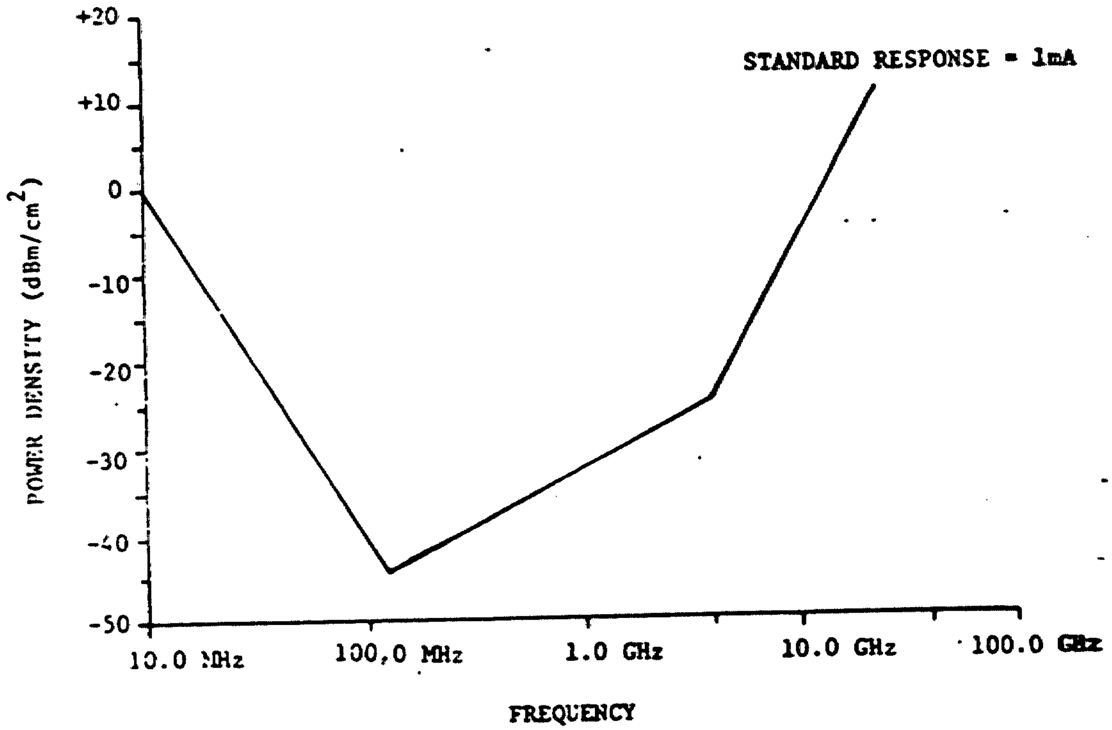

Figure C-25. System Standard Response Curve.

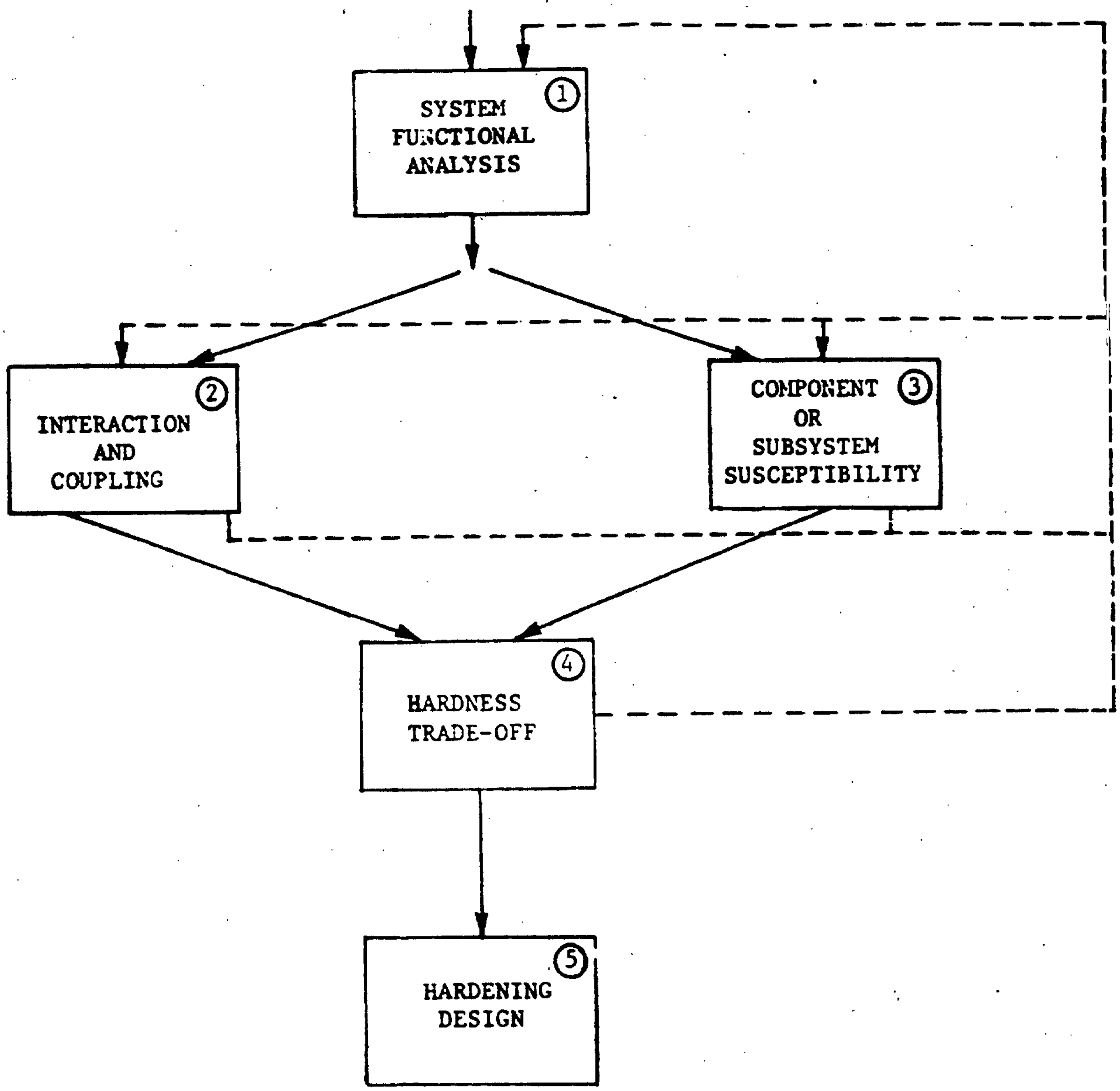

Figure C-26. Fundamental Methodology. [C-55] in Section 5 - EMR Hardness Design, but the emphasis here is on the steps dealing with analysis and prediction. The methodology set forth here was adapted from work by Latorre [C-55]. Each block of Figure C-26 consists of a decision structure with associated engineering operations. The decision structure takes into account constraints on time, money, and other resources. The cost-effective use of resources is provided through efficient program management.

The first block -- System Functional Analysis -- provides operations generally involving the entire system. In this block, the system is broken into manageable subsystems, and performance specifications are set for each subsystem. Here a criterion is considered to be some measure of system performance, while a specification is a predetermined value for that criterion. Typical system performance criteria may be the tracking rate or the circular error probability (CEP).

Following these operations, the effect of the local electromagnetic environment on the subsystems must be determined. This is accomplished in two steps. The first step is to determine the relationship of performance to parameters of the local environment. This relationship is defined as Component/Subsystem Susceptibility; operations required to obtain this relationship are performed in the corresponding block of Figure C-26. Estimates of parameters of the local environment combined with subsystem susceptibility predict the performance of the subsystem. Making these estimates constitutes the second step, and is provided through analysis and/or testing performed in the Interaction and Coupling block. Note that the analyst does not care what the precise susceptibility curve is or what the exact values of the environmental parameters are. He wants to know if this system is hard. By predicting upper bounds of performance and comparing them to lower-bound susceptibility curves (curves implying a greater susceptibility than actually exists), the analyst is assured that when he predicts that the performance of a subsystem will lie within its specification, it will.

In the block entitled Hardness Trade-Off, the analyst must decide whether to protect a system based on worst-case predictions of performance for all subsystems. If all subsystems appear susceptible, he may elect to protect, or he may choose to reassess the system. The dashed lines in Figure C-26 show the iteration path implied by the latter option.

Operations within the final block (Hardening Design) identify, develop, and enforce procedures that will assure and maintain the hardness integrity of the system.

70.2 **SYSTEM ANALYSIS.** The Systems Functional Analysis block of Figure C-27 has two purposes. First, it must provide some environmental description of the system from which estimates of the environmental parameters of subsystems may be obtained. Second, this block must set the subsystem performance specifications that identify the acceptable performance ranges for subsystems. A supplementary purpose is to identify system factors influencing the hardness of subsystems. For a specific hard-

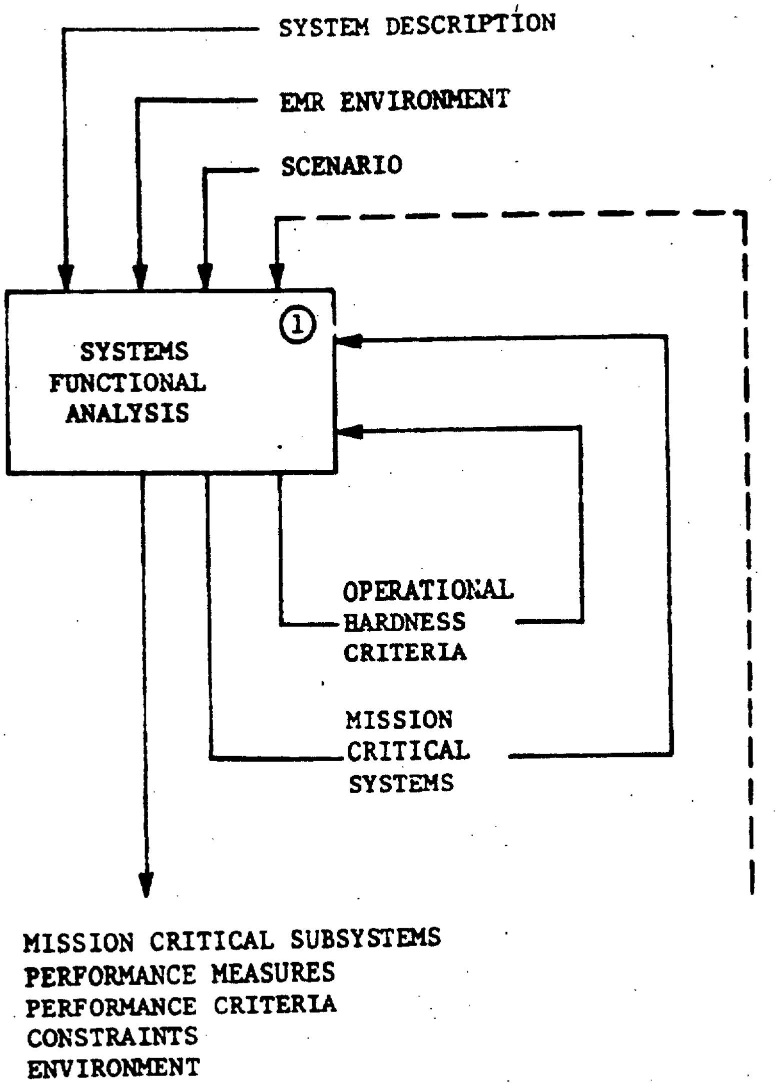

Figure C-27. Expanded System Analysis Block. [C-55] ness problem, the ingredients necessary to fulfill this block's tasks are:

- o Input data.
- o A logic structure that allows selection of necessary operations commensurate with problem definition and the quality and quantity of input data.
- o The mechanics by which required operations may be implemented.

Consider the hardness problem in which the analyst is given subsystem specifications and an environmental description that allows prediction of environmental parameters. In this case, the role of the System Functional Analysis block is minimal, and the analyst proceeds to the other blocks of Figure C-26.

Now consider the case when subsystem performance specifications are not given. The relevant specifications are generated from an overall system specification. Subsystem performances are obtained from values of subsystem performance descriptors. Thus, to obtain performance specifications of subsystems, it is necessary to know:

- o The overall system specification.
- o Quantities that describe subsystem performance (performance descriptors).
- o A description of the system by subsystems critical to the mission and amenable to hardness assessment.

If these factors are known, the analyst can then determine subsystem performance specifications. So far, the procedures have assumed that system specifications, subsystem performance descriptors, and mission critical subsystems were given. Generally, system specifications or specifications of a portion of the system are provided and the analyst is confronted in the System Functional Analysis block with the problem of first decomposing the system into subsystems and then identifying subsystem performance descriptors that adequately measure performance of each subsystem. Operations using this information and the system specifications will then yield subsystem specifications. If, however, the specifications of a portion of the system are not given, the analyst must rely on system's analysis to provide these specifications from the overall system specification. This involves the identification of performance descriptors of the system elements and the determination of the ranges these descriptors are allowed to assume, based on the overall system specification. If the overall system specification is not provided, the analyst must first identify how the system should perform in its environment (operational hardness criteria), then establish system performance descriptors, and finally determine system specifications.

An analyst may feel it is not his responsibility to determine specifications on any portion of the system. He may feel that this is the user's job. First, however, consider the case of subsystem performance specifications. Generally, the functional breakdown of a system and an assessment breakdown will not be the same. Thus the analyst is left with the task of determining how the subsystems resulting from his decomposition must behave in the EMR environment. Acceptable tolerances in this behavior are subsystem specifications.

Regardless of their origin, system and function specifications are required to provide subsystem specifications. As stated previously, these specifications are ranges of performance descriptors, such that values within these ranges assure tolerable system performance. Obviously, before the analyst can identify subsystem performance descriptors, he must first decompose the system into subsystems. Subsystem decomposition must satisfy two requirements. First, the subsystems must be critical to the mission (i.e., degradation in their performance would deleteriously influence the outcome of missions). Second, subsystems must be amenable (best compromise) to both subsystem susceptibility analyses and interaction and coupling analyses. To satisfy this last requirement, the analyst must weigh the accuracies of subsystem susceptibility analyses for rough system decomposition against the difficulties and accuracies of interaction and coupling analyses for a finer system decomposition. Decomposition will generally be made at metallic interfaces. Inputs required for the decomposition include system description, system configuration, modes of operation, and constraints in operation.

In the diagram of Figure C-26, the feedback paths imply that the analyst must generate information and then perform additional analyses on that information to obtain his required output. The dashed line covers the situation in which the analyst asks if the performance specifications on certain subsystems cannot be relaxed. Consider the case where, through analysis and testing, a set of subsystems are found to be hard, but an additional set soft. It may be found that the hard subsystems would remain hard if their specifications were tightened, thereby relaxing those of the soft subsystems. The end result may be a hard system.

The role of the System Functional Analysis block ceases when subsystem performance specifications and environmental descriptions have been generated and the allocation of resources for subsequent analyses has been provided. The analyst must now execute operations performed in the Subsystem Susceptibility and Interaction and Coupling blocks.

70.3 **SUBSYSTEM SUSCEPTIBILITY.** Performance specifications provide the allowable tolerances of performances. Changes in the values of performances result from the existence of an EMR environment local to that subsystem. Such a local environment includes incident fields, currents on penetrating conductors, currents on cable sheaths, etc. The variation in performances may be estimated from values of governing parameters contained within environmental parameters. These environmental parameters include amplitude and modulation.

In the Subsystem Susceptibility block of Figure C-28, the analyst determines the relationship of performances to environmental parameters. These relationships will be referred to as susceptibility curves (Appendix E). From such curves, allowable limits of environmental parameters are derived. These limits, when compared to estimated values of environmental parameters derived through interaction and coupling analyses, indicate the possible need for subsystem hardening.

Because of the complexity of most subsystems, exact determinations of susceptibility curves are impossible. Consequently, estimation is required. Estimates of susceptibility curves should satisfy the criterion that if environmental parameters fall within environmental limits, the system is hard, whereas, if they fall outside, the system is considered soft but may, in reality, be hard.

The specific susceptibilities of guided weapon systems and their subsytems is classified information, but pertinent discussion may be found in References[C-56] through [C-60].

70.4 **INTERACTION AND COUPLING.** The purpose of the Interaction and Coupling block is to provide estimates of the values of those parameters of the EM environment of a subsystem that are required in estimating the performance of that subsystem. The parameters requiring estimation are identified in the Subsystem Susceptibility block and are an input to the Interaction and Coupling block. Parameters that may be of interest are amplitude and modulation within a given frequency band, and amplitude samples of the frequency spectrum.

The estimates used in subsystem susceptibility are similar to interaction and coupling estimates in that they should be made on a worst-case basis so that a non-susceptible system really is, and a susceptible one may actually be non-susceptible. Once again, it is desirable to have the capability to provide varying degrees of estimation accuracy.

Estimates of interaction and coupling can be made in various ways. Figure C-29 illustrates the general coupling mechanisms involved in the estimation problem. Estimates of coupling to subsystems can be made based on knowledge of fields and currents within the system structure or from external surface currents, or made directly from the free field. When internal fields and currents are used to estimate the environmental parameters, it is necessary that external-to-internal coupling modes be identified and the appropriate estimates provided. When the external-to-internal coupling results from surface currents are used, it is necessary to estimate the external coupling modes that result from the direct interaction of the incident field with the external system configuration.

The analyst, in order to obtain estimates of environmental parameters, must first identify his subsystem coupling problem. Then he determines which canonical form will allow him to make the estimates. This canonical form permits him to identify not only physical parameters associated with the subsystem problem but also required independent variables. These variables are electromagnetic

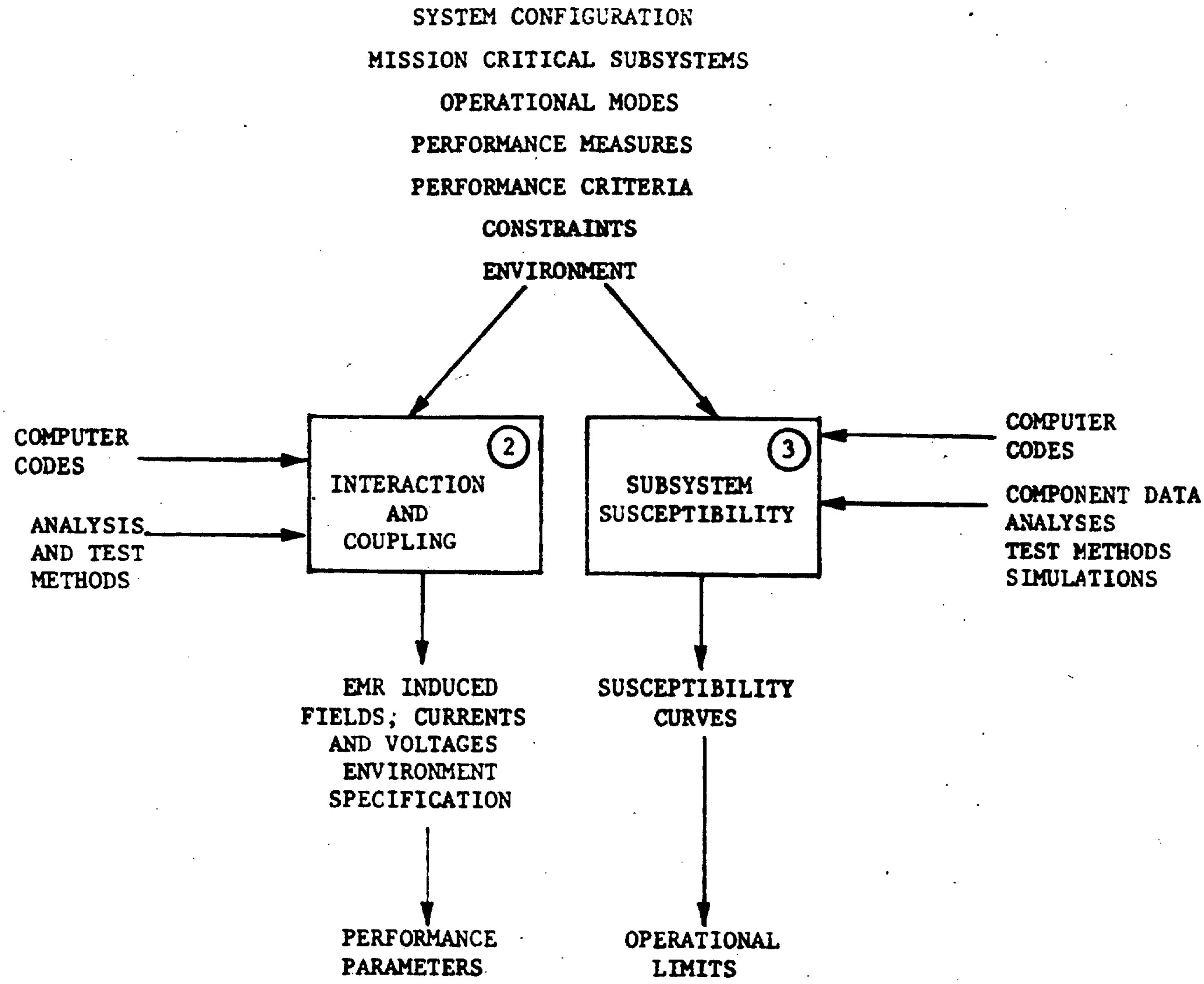

Figure C-28. Expanded Interaction and Coupling and Subsystem Susceptibility blocks. [C-55]

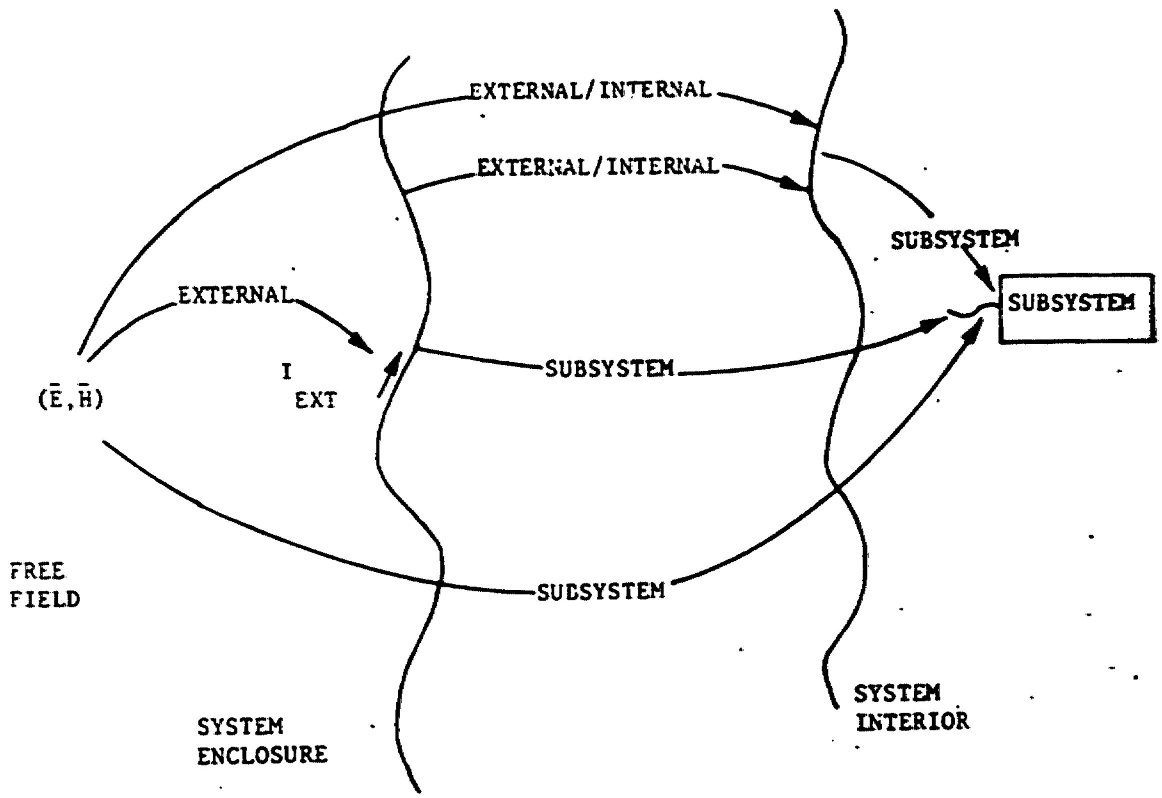

Figure C-29. General Coupling Modes. [C-55] quantities obtained either from external-to-internal coupling estimates or directly from the incident field. To determine values of the independent variables for external coupling modes, it is necessary to identify the coupling problem that exists and the canonical form associated with that problem. This new canonical form identifies physical parameters associated with the solution of that problem, and new independent variables. These new independent variables are electromagnetic quantities representing the free field or currents induced on the external system configuration. To obtain estimates of these new independent variables when they represent external currents, we have to consider an external coupling problem. Again, the analyst must identify the physical external coupling problem and associate a canonical form with it. With this form, he determines the physical parameters associated with the problem, and obtains estimates employing independent variables derived from the free field. These estimates are values to be used in estimating the exernal-to-internal coupling, which in turn are employed in subsystem coupling problems to provide estimates of the environmental parameters. A similar procedure is used when the other coupling modes are appropriate.

As in the Subsystem Susceptibility block, the first ingredient is appropriate logic to assist the analyst in the selection of the proper interaction and coupling method consistent with the input information, available resources, and problem requirements. A complete description of the available methods and the necessary tools is also required. The tools include computer programs, analytical methods, and descriptions of and results of experimental testing to obtain the basic coupling data for canonical problem sets. From these data, the analyst can estimate the coupling in a portion of the system.

The data needed to solve internal coupling problems include direct conductor penetration to subsystems, coupling to cables, and field diffusion to subsystem shields. For the external-to-internal coupling problem, data are required for three major mechanisms: coupling through apertures, field diffusion through structure shields, and direct conductor penetration of structural shields. External coupling data encompasses information on enclosures, long wires, deliberate antennas, and appendages on enclosures.

Data obtained through interaction and coupling analysis and that provided by subsystem susceptibility investigations are used to determine hardness requirements. This step is performed in the Hardness Trade-off block.

70.5 **HARDNESS TRADE-OFF.** In the Hardness Trade-off block, the analyst arrives at hardness requirements and, if protection is required, selects the most cost-effective scheme.

To accomplish the hardness trade-off, the analyst must make estimates of both subsystem susceptibility and interaction and coupling. In addition, he must have access to hardness methods and devices, device and method data, costs (monetary, performance, maintenance, and reliability), analysis and test methods, and costs of analysis and test. The above items, along with a logical trade-off structure, are the essential ingredients of the Trade-off block.

In the Trade-off block, the need for hardness has been determined, hardness requirements have been identified, and the protection method has been selected. The next facet of engineering is the actual design, fabrication, and maintenance of the system.

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APPENDIX D

THE INTRASYSTEM ANALYSIS PROGRAM

10. **INTRODUCTION.** This appendix describes the Air Force Intrasystem Analysis Program (IAP) including its components and related IAP studies. The information in this appendix is intended to give a user an overview of IAP so that parts may be adapted for an analysis and prediction effort or so that a user can make knowledgeable requests from RADC/RBCT. The IAP's capability is periodically upgraded as new developments become available. To determine the latest IAP information or to inquire as to IAP support and services, users should contact RADC/RBCT or RADC/RBCTI. The telephone numbers for the EMC/IAP Support Center (RADC/RBCTI) are commercial 315/339-3830 and autovon 587-2780/81/88.

During the 1971-72 time frame, the Air Force, along with industry and the Rand Corporation, performed an extensive evaluation on the effectiveness of system modeling programs used in the acquisition process. From the evaluation, it was concluded that an analysis and prediction capability would improve the effectiveness and reduce the costs in achieving system compatibility. As a result of these studies, HQ AFSC assigned RADC the task of developing an analysis and prediction capability applicable to the acquisition of ground and aerospace systems. As RADC's work progressed, the Air Force realized a need to expand the concept, and in 1974 introduced the Intrasystem Analysis Program. The IAP provides a computer analysis capability for:

- o Assessing system compatibility in a systematic approach.
- o Tailoring equipment specifications.
- o Performing waiver analyses.
- o Analyzing design tradeoffs.
- o Reducing the number of required tests.
- o Predicting the effectiveness of hardness controls prior to system design.

Since 1974, IAP has been, and is presently being, upgraded and improved to increase its capability. Improved models, data base management techniques, and computer code updates have been, or will soon be, implemented within IAP. In 1978, RADC initiated the EMC/IAP Support Center to provide a central facility for IAP support and service to both government and industry.

20. **OVERVIEW OF IAP.** The IAP consists of several parts: the Intrasystem Electromagnetic Compatibility Analysis Program (IEMCAP) and several off-line models. The IEMCAP provides the basic systems level analysis and is the building block from which the total program evolves.

IEMCAP is written in USA Standard FORTRAN IV, which makes it readily adaptable to most computers. Included within IEMCAP are EM emitter models, coupling models, and receptor models. The off-line models that comprise the second part of IAP extend the analysis capability beyond that of IEMCAP. These capabilities include wire/cable coupling, computer-aided circuit analysis, and electromagnetic field analysis. Additional supplemental programs and models are available, or are currently under development, and will be included as a part of IAP in the future. Supplemental programs are available for precipitation static analysis and TEMPEST.

The wire/cable coupling package consists of the programs XTALK, XTALK2, FLATPAK, FLATPAK2, GETCAP, and WIRE. The computer-aided circuit analysis is contained in the computer code NCAP (Nonlinear Circuit Analysis Program). The General Electromagnetic Model for the Analysis of Complex Systems (GEMACS) is a method-of-moments code for use in electromagnetic field analysis. Each of these off-line programs is described further in the following sections.

30. **IEMCAP.** IEMCAP is a link between equipment and subsystem EMC performance and total system performance in an electromagnetic environment. IEMCAP was developed for the Air Force by McDonnell Aircraft Company (MCAIR) in 1974 to facilitate computerized analysis in the engineering of cost-effective EMC. Since its development, IEMCAP has undergone several revisions. The newest released version is IEMCAP-05. The documents describing IEMCAP and its use are:

- "Volume I - INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM - User's Manual Engineering Section," December 1974, RADC-TR-74-342, AD-A008-526.

- "Volume II - INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM - User's Manual Usage Section," December 1974, RADC-TR-74-342, AD-A008-527.

- "Volume III - INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM - Computer Program Documentation," December 1974, RADC-TR-74-342, AD-A008-528.

Changes, additions, and deletions have been made to these volumes. These updates may be obtained by contacting the EMC/IAP Support Center at RADC. Also, the FORTRAN computer listing for IEMCAP may be obtained through the Support Center.

Since the release of IEMCAP, additional studies and investigations have been performed in support of IEMCAP. One study was performed to determine how IAP and IEMCAP can best be implemented in the weapons system procurement process. These reports are:

- "Volume I - IEMCAP IMPLEMENTATION STUDY," RADC-TR-77-376, December 1977.

- "Volume II - IEMCAP IMPLEMENTATION STUDY, Annex: Electromagnetic Compatibility Handbook for System Development and Procurement," RADC-TR-77-376, December 1977, AD-A094-738.

Another study was performed to summarize, in a concise manner, the required input parameters for the emitter models within IEMCAP. Along with this information, the form of the power spectral density and suggested frequency table input values to adequately represent the spectrum are presented. This study is reported on in:

- "A SUMMARY OF REQUIRED INPUT PARAMETERS FOR EMITTER MODELS IN IEMCAP," RADC-TR-78-140, June 1978, AD-A056-805.

The validity and usefulness of IEMCAP was assessed in a study effort. IEMCAP was used to predict the EMC performance characteristics of the F-15 air superiority fighter aircraft. The aircraft was simulated using a combination of known, measured, and approximated data. In this validation, the IEMCAP predicted the overall system compatibility, some isolated cases of interference, and the compatibility effectiveness of the subsequent fixes. This study is documented in reports:

- "Part I - INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM (IEMCAP) F-15 VALIDATION - Validation and Sensitivity Study," RADC-TR-77-290, September 1977, AD-A045-034.

- "Part II - INTRASYSTEM ELECTROMAGNETIC COMPATIBILITY ANALYSIS PROGRAM (IEMCAP) F-15 VALIDATION - Interpretation of the Integrated Margin," RADC-TR-77-290, September 1977, AD-A045-035.

Another study, by the Aeronautical Systems Division (ASD), assessed the effectiveness of the IEMCAP in predicting antenna-coupled interference. The B-52 aircraft was used in the case study and the IEMCAP predictions were compared with actual measurements. The study concluded that IEMCAP does correctly predict a high percentage of actual interference problems.

When using IAP for systems (e.g., aircraft, satellite, missiles, etc.), data pertaining to the physical and electrical characteristics must be collected. Much of these data are fixed quantities which are used repeatedly with each analysis of a system. For example, knowledge of the geometrical shape of the exterior surface of an aircraft is usually needed for an antenna-to-antenna coupling analysis of the system. Also, the locations and electrical characteristics of all electrical equipment and intrasystem wire cabling are required for major analyses. In order to aid users in performing analyses, it is intended that a data base of physical and electrical characteristics of a large number of systems be collected. The data for each specific system are stored on a System Data File (SDF) for that system. Consequently, it is intended that a separate SDF exists for each physical system. The primary application of the SDF is to provide a source of input data for any of the various IAP computer programs as needed for a particular analysis effort. The SDF is documented in reports:

o "System Data File (SDF) for the Intrasystem Analysis Program (IAP)," Volume I - Description, Volume II - Surface Geometry, RADC-TR-79-213, December 1979, AD-A080-584 and AD-A080-585.

The IEMCAP documentation that has been described provides all the necessary information for one to become familiar with IEMCAP. However, it is an extensive task to start with those documents and proceed to implement IEMCAP. The alternative and recommended approach is to work through the EMC/IAP Support Center, which teaches training courses and sponsors user forums.

40. **WIRE-COUPLING MODELS.** Crosstalk or electromagnetic coupling between wires (cylindrical conductors) in densely packed cable bundles can be a major contributor to the performance degradation in weapon systems. IEMCAP provides a general analysis capability for wire-to-wire coupling. The IEMCAP subroutines do not consider the simultaneous interactions between all wires in a cable bundle when computing the coupling between a generator-receptor circuit pair. Each generator-receptor circuit pair is considered individually, and the effects of other "parasitic" wire circuits in a bundle on the coupling between a generator-receptor circuit pair is not considered. This approach was used for two reasons. First of all, such a model tends to give an upper bound which is in keeping with the worst-case approach used in IEMCAP. Secondly, if the interactions of all the wire circuits in the cable bundle were considered, an NxN complex matrix must be solved at each frequency. For large cable bundles, much of the IEMCAP execution time would be consumed by the wire-to-wire coupling calculations.

To supplement IEMCAP's capability, a number of stand-alone computer programs have been developed. Once IEMCAP pinpoints a marginal wire-coupling problem, these programs may be used to perform a more fine-grain analysis to determine if, in fact, a problem exists. These stand-alone programs are referred to as XTALK, XTALK2, FLATPAK, FLATPAK2, GETCAP, and WIRE. XTALK analyzes three configurations of transmission lines: (1) (n + 1) bare wires; (2) n bare wires above an infinite ground plane; and (3) n wires within a cylindrical shield which is filled with a homogeneous dielectric. All conductors are considered to be perfect conductors. XTALK2 analyzes the same three structural configurations as XTALK, except that the conductors are considered to be imperfect conductors. FLATPAK analyzes (n + 1) wire ribbon cables. All wires are assumed to be perfect conductors. FLATPAK2 analyzes the same configuration as FLATPAK, except that the wires are considered to be imperfect conductors. GETCAP (an acronym for GEneralized and Transmission line CAPacitance matrices) is utilized to calculate the per-unit-length generalized and transmission line capacitance matrices for ribbon cables. WIRE is a computer program which is designed to calculate the sinusoidal, steady state, terminal currents induced at the ends of a uniform, multiconductor transmission line which is illuminated by an incident electromagnetic field. The incident field can be either a uniform plane wave or a general nonuniform field. Three types of transmission line structures are considered. Type 1 structures consist of (n + 1) parallel wires. Type 2 structures consist of n wires above an infinite ground plane. Type 3 structures consist of n wires within an overall, cylindrical shield.

At present, these stand-alone wire coupling programs do not directly consider branched cables. Also, they do not consider individually shielded wires or twisted pairs. Current work is being directed towards obtaining a single program which considers all these factors directly and incorporates the transmission line model. The IEMCAP has models for all these situations. However, the models in the IEMCAP are simple, lumped approximations to the coupling phenomena, whereas the stand-alone routines consist of the more exact multiconductor transmission line parameter models.

An additional effort is directed to the field-to-wire coupling model in the sense that the WIRE program considers interactions between all wires in the bundle, whereas the field-to-wire subroutine in the IEMCAP does not. Current efforts are being directed toward verifying, with experimental data, the predictions of the WIRE programs and updating the programs to directly handle branched bundles, individually shielded wires, and twisted pairs.

The stand-alone wire coupling programs, which were developed by the University of Kentucky, are described in a seven-volume series entitled, "Applications of Multiconductor Transmission Line Theory to the Prediction of Cable Coupling," RADC-TR-76-101.

- "Volume I - Multiconductor Transmission Line Theory," April 1976, AD-A025-028.
- "Volume II - Computation of the Capacitance Matrices for Ribbon Cables." April 1976, AD-A025-029.
- "Volume III - Prediction of Crosstalk in Random Cable Bundles," February 1977, AD-A038-316.
- "Volume IV - Prediction of Crosstalk in Ribbon Cables," February 1978, AD-A053-548.
- "Volume V - Prediction of Crosstalk Involving Twisted Wire Pairs," February 1978, AD-A053-559.
- "Volume VI - A Digital Computer Program for Determining Terminal Currents Induced in a Multiconductor Transmission Line by an Incident Electromagnetic Field," February 1978, AD-A053-560.
- "Volume VII - Digital Computer Programs for the Analysis of Multiconductor Transmission Lines," July 1977, AD-A046-662.

50. NCAP. IEMCAP determines and analyzes system susceptibilties using equipment emitter-receptor ports as characterized by the user. For finer analysis at the circuit level, it is necessary to utilize a computer-aided circuit analysis program. Such a code is available

in IAP and is referred to as the Nonlinear Circuit Analysis Program (NCAP). Both linear and nonlinear component and source models are available in NCAP. Included are models for semiconductor diodes, bipolar junction transistors, and field-effect transistors. Diode, triode and pentode models are available for vacuum tubes. Nonlinear models for resistors, capacitors, and inductors are also available. NCAP uses a circuit-oriented procedure, based upon frequency domain analysis, for predicting many nonlinear effects in electronic circuits.

NCAP can be used to analyze the nonlinear effects such as gain expansion/compression, desensitization, cross-modulation, intermodulation, and demodulation. For low frequencies, these nonlinear effects are relatively easy to analyze; at higher frequencies, the task becomes more difficult due to the presence of parasitic components. Thus at higher frequencies, NCAP is limited to the 100-500 MHz frequency range.

NCAP may also be used to study the effects of modulations induced into a circuit via the rectification mechanism for out-of-band RF or microwave signals. A computer-aided analysis procedure, based upon a modified Ebers-Moll transistor model, may be applied to predict the induced modulation signal levels. Once the modified Ebers-Moll transistor model is characterized, NCAP may be used to determine the effect of the induced modulation. Information on the use of the modified Ebers-Moll model in other computer-aided circuit analysis programs may be found in the McDonnell Douglas report MDC E1929, entitled "Integrated Circuit Electromagnetic Susceptibility Handbook, Integrated Circuit Electromagnetic Susceptibility Investigation - Phase III." Further details on component susceptibility to electromagnetic fields may be found in Appendix E. The documentation that describes user's information has been released and is:

- "Nonlinear Circuit Analysis Program (NCAP) Documentation," RADC-TR-79-245, Volume I - Engineering Manual, Volume II - User's Manual, Volume III - Programmer's Manual, September 1979, AD-A076-384, AD-A076-596 and AD-A076-317.

60. **GEMACS.** IEMCAP calculates radiation coupling paths using transmission line or dipole models that are simplistic, but which nevertheless are an upper bound for the level of coupling. For more accurate analysis than is possible with IEMCAP, GEMACS must be used to determine radiation coupling paths. GEMACS uses the method-of-moments (MOM) technique with the expansion function (sine + cosine + constant) and the collocation scheme. A user may obtain the electrical currents, far-field and near-field radiation patterns, antenna input impedance, and antenna coupling parameters for wire antennas on structures represented by wire grid models. Thus calculation may be made to determine: the coupling between a pair of antennas or between an antenna and a conductor; the coupling of an external field to a conductor; the modification of an antenna far-field beam pattern by the presence of nearby obstacles causing radiation in undesired directions and; the level of potential field hazards.

Basically, the MOM formulation takes the integral electromagnetic field equation and transforms it into a matrix equation. The foundation of this transformation is the subdivision of the structure being analyzed into a number of subsections, each of which is small compared to the wavelength. The elements within the matrix represent the interactions between the subsections into which the geometry is divided.

In the past, the MOM matrix has been limited to small systems that can be represented by 300 subsections or less. Electrically, this roughly corresponds to a size of 30 wavelengths of wires or a surface with an area of one square wavelength. This has not been due to a limitation of the theory or the technique, but is a limitation of computer resources needed to perform a MOM analysis. The computer core storage goes up as N^2, and the solution time increases as N^3, where N is the number of subsections. GEMACS has circumvented the computer resource limitations by introducing the banded matrix technique and out-of-core manipulation capability. GEMACS can handle matrices much greater than 300. However, matrix sizes greater than 1000 become very expensive, and use of such large matrices should be weighed carefully in terms of the cost-benefit.

The MOM wire-grid model can be used only to solve external type problems. For example, coupling through apertures in the skin of a structure and coupling between antennas located on opposite sides of a structure cannot be treated with confidence using a wire-grid model. The reason for this is that the wire grid "leaks" through the mesh in the model causing undesired coupling. If a "surface-patch" model is utilized, these limitations will not be present. Efforts are currently underway to implement such a capability within GEMACS.

GEMACS has been very well documented and the GEMACS reports are:

- "Volume I - GENERAL ELECTROMAGNETIC MODEL FOR THE ANALYSIS OF COMPLEX SYSTEMS - User's Manual," April 1977, AD-A040-026.
- "Volume II - GENERAL ELECTROMAGNETIC MODEL FOR THE ANALYSIS OF COMPLEX SYSTEMS - Engineering Manual," April 1977, AD-A040-027.
- "AN INTRODUCTION TO THE GENERAL ELECTROMAGNETIC MODEL FOR THE ANALYSIS OF COMPLEX SYSTEMS (GEMACS)," RADC-TR-78-181, September 1978, AD-A060-319.

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APPENDIX E

ESTABLISHING SUSCEPTIBILITY LEVELS

10. **INTRODUCTION.** The objective of this appendix is to provide data and information on the susceptibility levels of a variety of solid-state circuits and devices commonly employed in modern air launched ordnance systems. It must be emphasized that a totally comprehensive set of electromagnetic susceptibility (EMS) data is not available at the present time; indeed, it appears to be an impossible task as the development of new devices and integrated circuits outdistances any attempt at a quantitative analysis of their susceptibility levels. However, a wide variety of the available susceptibility information and data is documented in this appendix. These data are representative of components used in system hardware design applications.

The domain of the EMS data presented herein is confined to the frequency range from 220 megahertz to 9.1 gigahertz. The upper frequency limit is considered sufficient, owing to two major factors: (1) the frequency response of semiconductor devices and integrated circuits (IC's) roll off quite rapidly due to parasitic and distributed capacitances, and (2) the coupling of RF energy onto system wiring falls off as the square of the wavelength. Below 220 MHz, the component responses level off as frequency decreases, as do the pickup responses which are limited by mismatch effects. It is therefore believed that the worst-case susceptibility levels are adequately covered by this frequency domain.

The material which follows is divided into four major sections. Section 20 is a qualitative overview of some general susceptibility characteristics. This should assist the system designer in understanding the cause and nature of device susceptibility and direct him in the selection of design guidelines for limiting potential system susceptibility problems. Section 30 documents the available susceptibility data on a variety of semiconductor circuits and devices. Section 40 discusses three possible approaches to extending the data base provided in Section 30. Section 50 lists reference material.

20. **GENERAL SUSCEPTIBILITY CHARACTERISTICS.** Electromagnetic susceptibility is defined as the characteristic of electronic equipment that permits undesirable responses when the equipment is subjected to electromagnetic energy. There are three separate classifications pertaining to the effects of coupled energy on electronic device performance [E-1]. These classifications serve to categorize the disruptive effects into varying degrees of severity and these are: (1) interference, (2) degradation, and (3) catastrophic failure. Interference is the least severe effect and is defined as a reduction in the operational capability of one or more semiconductor device parameters in the presence of coupled RF energy, with a return to normal operation when the energy

is removed. A device which has undergone degradation will still function after removal of the RF energy, but with an alteration in one or more parameters and/or a decrease in device lifetime. Catastrophic failure is the most severe effect and is defined as permanent physical or electrical damage which renders a device nonfunctional with respect to its intended use.

The effect of coupled energy on a circuit or subsystem level can only be determined by the system designer. The analysis begins with a determination of the effects on specific devices. By pinpointing the susceptible devices, the designer may then analyze the impact of interference, degradation, or catastrophic failure on circuit and subsystem performance. This will, of course, depend on numerous factors including the characteristics of the coupled energy, device parameters and operating conditions, and the particular function of each circuit or subsystem.

20.1 **RF POWER ABSORPTION.** One of the most influential factors concerning disruptive effects on semiconductor devices is the amount of RF power absorbed. Holding all other parameters fixed, an increase in absorbed power causes an increase in the likelihood of a device malfunctioning. It has been demonstrated that the ratio of incident power to absorbed power may vary from approximately 2 dB to more than 20 dB [E-2]. It is thus the absorbed power that proves to be the more reliable susceptibility parameter. The amount of power absorbed for a given incident power is influenced by the frequency of the RF signal, the operating conditions of the device, the entry port characteristics, as well as the impedance of the injection port.

The basic mechanism in the disruptive effects of RF energy is rectification of the coupled energy at p-n junctions. Rectification is effected through different physical phenomena depending on the device type and fabrication technology. Studies done on bipolar transistors, for example, have indicated that rectification is caused by both the nonlinear characteristics of device p-n junctions and RF-induced current crowding resulting from the redistribution of junction current [E-3] However, for the purposes of this appendix, attention will be focused on the general nature of the mechanism and its relationship to device susceptibility levels.

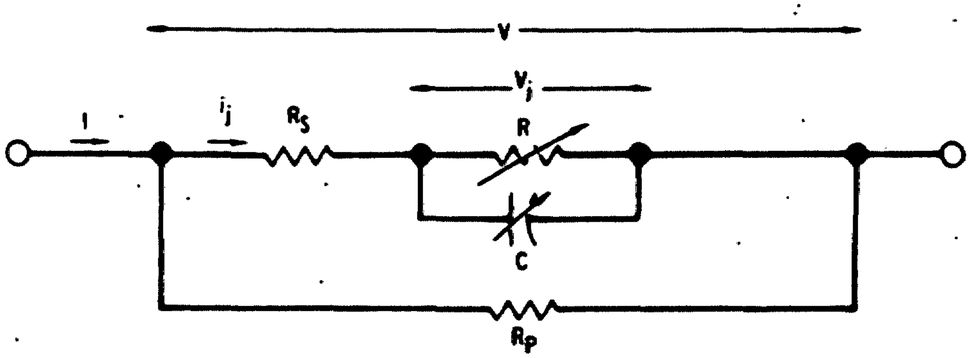
The initial effect of RF power absorption on semiconductor device performance is interference. This typically occurs at absorbed power levels of 10 mW or less. As the amount of absorbed power increases, device degradation and eventually catastrophic failure will occur. Device failure occurs when the RF power level is sufficient to generate enough heat to cause permanent physical damage to the device.

This appendix uses a worst-case approach to susceptibility analysis and therefore will only include the interference levels for the devices and circuits. It has been demonstrated that the upper interference level and the lower failure threshold nearly coincide for digital devices [E-4]. On the other hand, the interference level in linear devices can only be defined in relation to the operating conditions and circuit function (causing a very large spread in the possible levels for a single device). A logical approach to this problem is to perform a worst-case analysis using some predetermined maximum tolerance as the interference level. The susceptibility level is obtained by reference to a graph which includes a curve for several different fixed parametric values (all of which may represent possible interference levels). Once the maximum tolerable change in a particular parameter is determined, the suceptibility is then read off of the absorbed power axis.

20.2 **FREQUENCY AND MODULATION.** The susceptibility levels of nearly all devices and IC's are influenced by the frequency and modulation characteristics of the RF signal. Most components become less susceptible as the frequency of the signal increases. This is due, at least in part, to the fact that most system components are relatively slow reacting. This trend can be expressed theoretically in the form of a rectification efficiency, which is shown to be frequency dependent. Figure E-1 shows a model which may be used to describe the nonlinear characteristics of a p-n junction [E-5]. The nonlinear resistance of the junction is described in terms of the voltage-current relationship given by:

$$i_j = I_o(e^{QVj}-1) \qquad \text{(E-1)}$$

where I_o and Q are treated as parameters to be determined as necessary



NOTE:

$i_j = i_o(e^{QV_j} - 1)$

$C = C_o\left(\frac{\phi}{\phi - V_j}\right)^n$

C_o = JUNCTION CAPACITANCE AT $V_j = 0$

ϕ = WORK FUNCTION OF JUNCTION

n IS DEPENDENT UPON DOPING GRADIENTS AND USUALLY LIES IN THE RANGE OF 0.1 TO 0.5

Figure E-1. Model for p-n Junction. [E-5] to give the best representation of the junction. The variable capacitance represents the junction capacitance which varies with V_j. Nominally, R_s represents the bulk resistance and R_p the various leakage paths. The rectification efficiency (η) is derived using a Taylor series expansion about the dc bias voltage, thereby arriving at a series expression for the average current through the junction. The assumption of a small RF signal amplitude leads to an expression involving the rectification efficiency, viz,

$$I_g = \eta P_a \tag{E-2}$$

where:

I_g = rectification current produced by the RF signal

P_a = RF power absorbed

η = rectification efficiency

$$\eta = \frac{Q}{2(1 + R_s/R)^2} \cdot \frac{1}{1 + \frac{\omega^2 C^2 R_s R^2}{R + R_s}} \tag{E-3}$$

R = dV/di for the junction characteristics.

Figure E-2 is a graph of Equation (E-3) showing the predicted frequency dependence of the rectification efficiency. Clearly, the decrease in η for frequencies greater than "cutoff" (f_o) implies a decrease in rectification current for a fixed absorbed power level. Two examples of measured rectification current in IC's (measured while the devices were unpowered) are shown in Figure E-3 [E-4]. The 7400 NAND gate (bipolar) example uses the rectification current in the collector isolation junction of the output transistor. The 4011 CMOS NAND gate example monitors the rectification current in the drain-substrate junction of the output transistor. Both demonstrate rectification efficiencies varying inversely with frequency, as predicted. This trend will also be quite apparent when viewing the composite worst-case susceptibility curves of Section 30. In general, it can be seen that the susceptibility levels increase with frequency (indicating a decrease in susceptibility).

While some variation in susceptibility can be measured over broad frequency ranges, the primary response to an RF signal is not to the carrier frequency, but rather to the envelope of the modulation. Tests have been conducted where a video pulse shaped like the modulation envelope was substituted for the modulated microwave signal. The failure results were the same when either the video or the modulated microwave signals were applied [E-6].

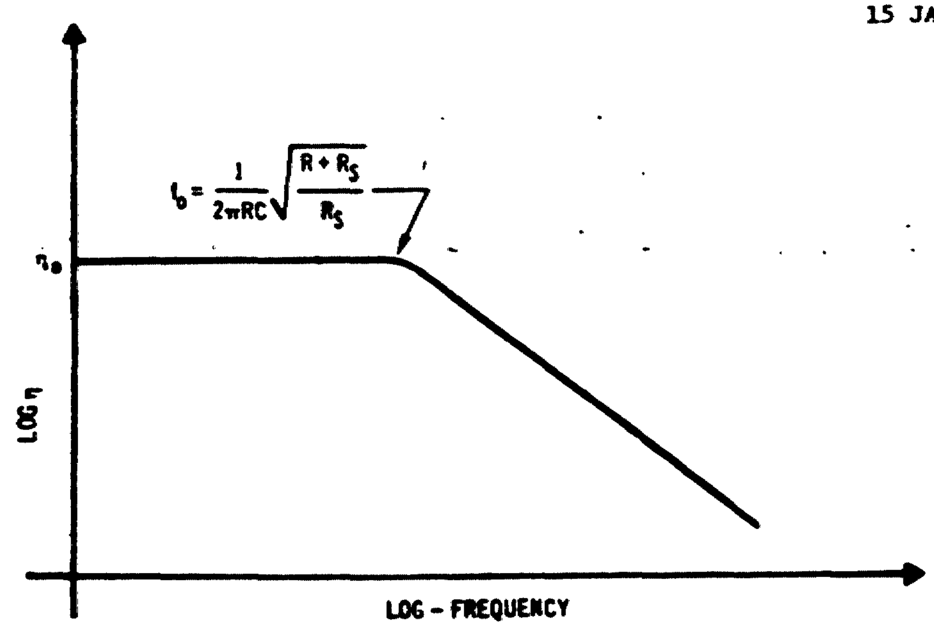


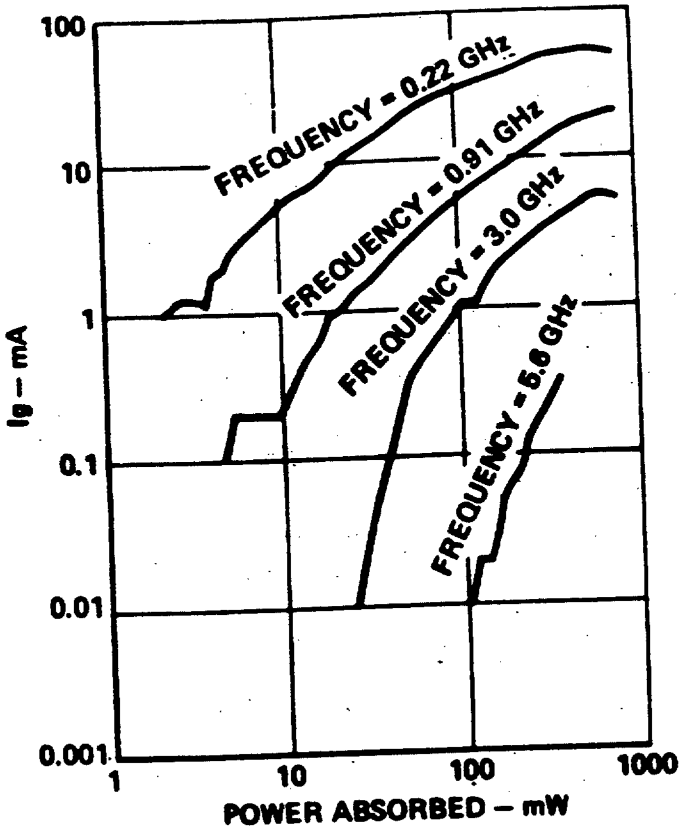
Figure E-2. Predicted Form of the Maximum Rectification Efficiency Plotted vs. Frequency. [E-5]

The rectified RF signal will appear as a dc or video signal depending on the modulation characteristics of the RF energy. For a continuous wave (CW) RF signal, the rectified current (voltage) will produce a dc shift in the quiescent operating point of the device. A pulsed RF signal causes a superposition of a video signal, which is a replica of the RF envelope, onto the original operating point. The resulting dc or video signal may now propagate through the circuit as though it were a legitimate signal and may thus affect circuit and system performance.

20.3 **GENERAL CONSIDERATIONS CONCERNING SUSCEPTIBILITY.** The previous two sections addressed source-related factors (RF power, frequency, and modulation) which affect the susceptibility of solid-state devices. This section is directed toward system-related factors which may influence device susceptibility.

20.3.1 **DIGITAL VS. LINEAR.** Semiconductor components function as digital, linear, or hybrid units. In digital units, the output is intended to vary discretely between two states, interpreted as being either high or low. In linear units, the output varies continuously and the information is derived from the exact output level. A hybrid circuit utilizes a combination of both digital and linear units.

Digital device susceptibility is relatively easy to define. One clear-cut way is to consider the absorbed power required to change the output state of a device for a particular set of conditions. However, this creates some degree of obscurity in that a range of voltages



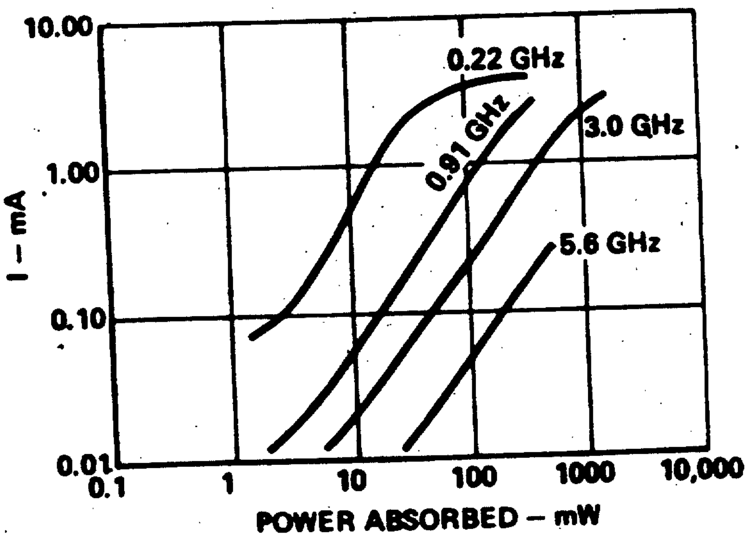


Figure E-3. Illustration of Rectification Currents in 7400 NAND Gate (top) and 4011 NAND Gate (bottom). [E-4] exist which may be interpreted by the circuit as high or low. One solution to this apparent difficulty is to define separate signal levels which, for the purpose of an example, shall be designated as voltages A, B, and C. Level A pertains to the manufacturer's specification of the guaranteed voltage limit for proper interpretation of the logic state. Levels B and C correspond to increasing degrees of uncertainty (to be defined in Section 30) and increasing noise level. Operation below Level A guarantees correct interpretation of state while operation above Level A may be somewhat risky. Therefore, level A may be used for an initial worst-case analysis of a system.

Linear device susceptibility, in contrast to that of digital devices, can only be defined in terms of circuit or subsystem function and operating characteristics. Each linear unit has a specific function to perform and the accuracy or stability with which a signal must be maintained depends on the particular system requirements. For example, Figure E-4 illustrates the effect on output voltage of injecting 220 MHz energy into the non-inverting input of a 741 operational amplifier [E-7]. Clearly, the level of absorbed power required to cause interference is dependent upon the circuit function. The circuit designer

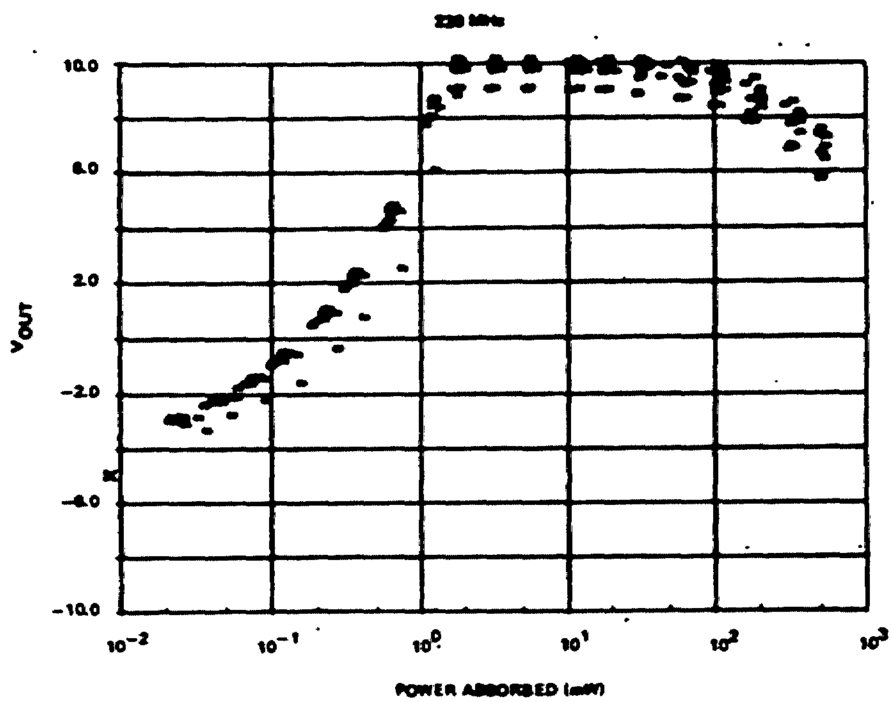


Figure E-4. 741 Operational Amplifier Susceptibility Data-- RF Injected Into Non-Inverting Input. [E-7] must determine what constitutes interference for the circuit in question based upon a knowledge of the susceptibility of representative linear devices, their interrelationship within the circuit or subsystem, and the system requirements.

Linear circuits are generally used to perform more sensitive functions than digital circuits. It might therefore be expected that digital circuits would be less susceptible than linear circuits to coupled RF energy. Studies conducted to compare the relative susceptibility levels of representative digital and linear devices have demonstrated this to be true [E-8]. Twenty different digital and linear devices were tested using reasonable criteria for the minimum and maximum susceptibility threshold levels. The particular devices used are listed in Tables E-1 and E-2, and a summary of the results is shown in Figure E-5. The designer is usually concerned with the worst-case conditions and therefore is more interested in the minimum threshold levels. The minimum interference levels of the linear devices are below those for the digital devices at all four test frequencies, by the following amounts: 30.8 dB at 220 MHz, 12.3 dB at 0.91 GHz, 19.7 dB at 3.0 GHz, and 31.0 dB at 5.6 GHz. For linear devices, the most susceptible port was generally found to be an input (usually the inverting), while for the digital devices it was most often the output port operating in the low state.

20.3.2 **FABRICATION TECHNOLOGY.** There are three major technologies available for device selection by system designers: bipolar, MOS, and hybrid. Investigations have been conducted concerning the possible differences in EMS between similar devices from each technology [E-9] The devices selected were a 7400 bipolar NAND gate, a 4011 CMOS NAND gate, and a 2002 DTL high power driver. Interference and failure data were recorded at test frequencies of 220 MHz, 910 MHz, 3.0 GHz, and 5.6 GHz.

Interference levels were selected from the individual device specification sheets. Figure E-6 shows the results of the interference measurements. Although the CMOS devices appeared to be slightly less susceptible to RF interference than the bipolar and hybrid devices, the variation was limited to a 10-dB band and may be considered insignificant. A comparison of the peak pulse power level required to cause failure for the three types also resulted in a 10-dB maximum spread and at a level approximately 30 dB above the minimum interference threshold level band.

It appears that the particular fabrication technology, in itself, does not significantly influence the RF susceptibility of semiconductor devices.

20.3.3 **PACKAGE EFFECTS.** Considering that semiconductor device packages are not designed for RF transmission, it would be reasonable to assume that reflective and absorptive losses would vary with package style. Reflective losses arise from the microwave mismatch provided by the packaged IC to the nonideal transmission line represented by the system cabling. Absorptive losses result when energy is dissipated in the package as it is delivered to the chip. If these losses did

TABLE E-1

DIGITAL DEVICES TESTED. [E-8]

Device Number	Family Type	Logical Function
7432	TTL	Quad 2-input OR gate
7402	TTL	Quad 2-input NOR gate
7404	TTL	Hex inverter
7405	TTL	Hex inverter (open collector)
7450	TTL	Expandable dual 2-wide 2-input AND-OR invert gate
7473	TTL	Dual type J-K flip-flop
7479	TTL	Dual type D flip-flop
3021	TTL	Quad exclusive OR gate (high speed)
4011	CMOS	Quad 2-input NAND gate
2002	Hybrid	High power driver

TABLE E-2

LINEAR DEVICES TESTED. [E-8]

Device Number	Device Type
201	Operational amplifier
307	Operational amplifier
310	Voltage follower
316	Operational amplifier
324	Quad operational amplifier
339	Quad comparator
725	Instrumentation operational amplifier
747	Dual operational amplifier
309	Positive voltage regulator
320	Negative voltage regulator

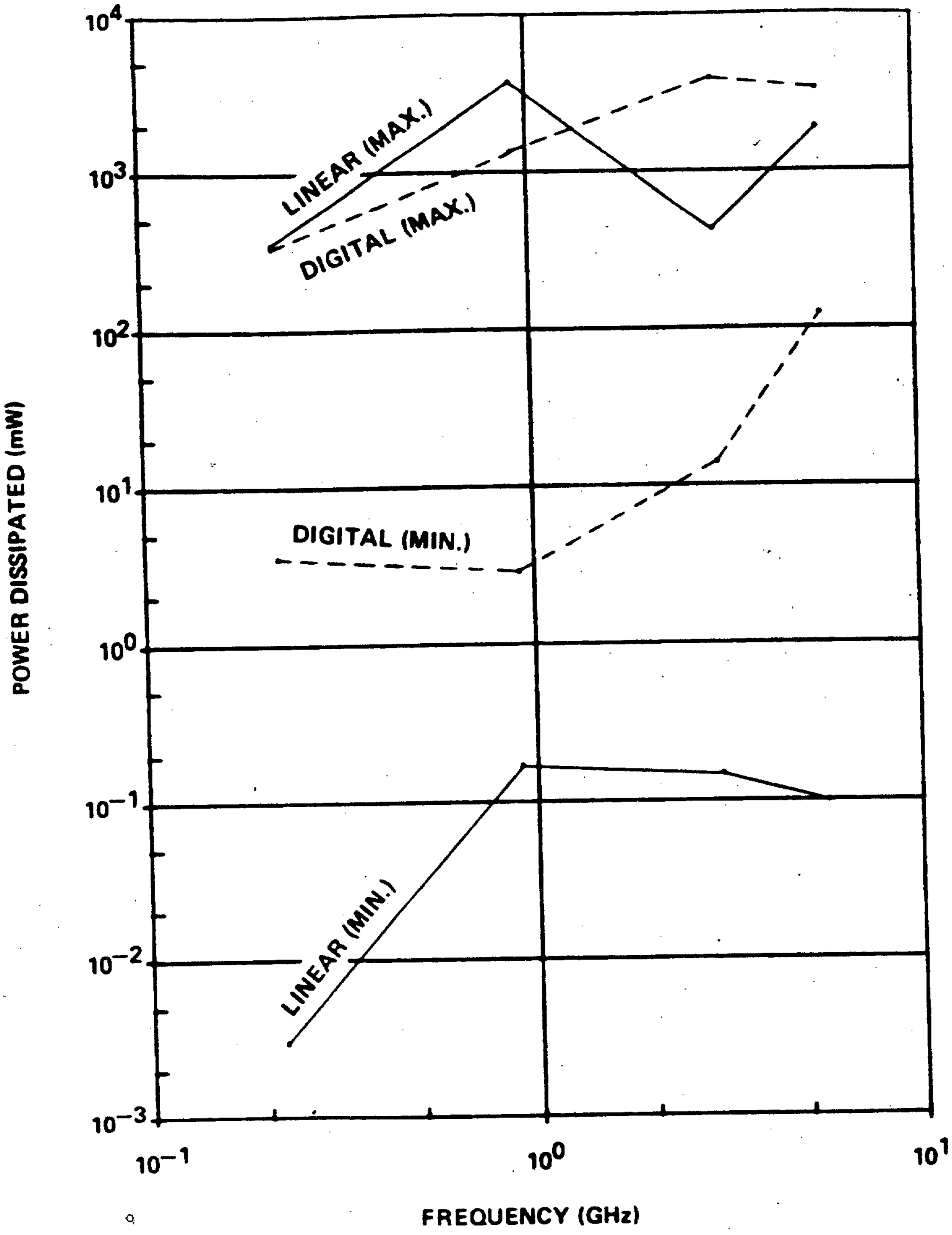


Figure E-5. Maximum and Minimum Susceptibility Threshold Levels for Linear and Digital Devices. [E-8]

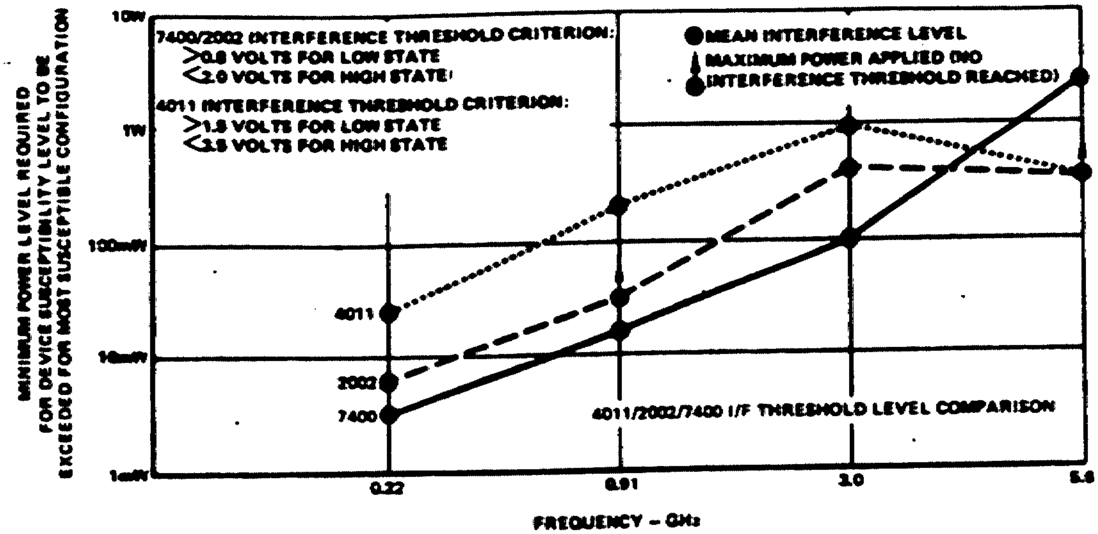


Figure E-6. 4011/2002/7400 Minimum Interference Level Comparisons at Four Frequencies. [E-9]

indeed vary between package styles, the designer would have a means of reducing circuit susceptibility by selecting the least susceptible package type.

Studies to determine whether or not potentially significant differences in reflective and absorptive losses exist have been performed using three representative IC package types [E-2]. Figure E-7 contains outlined illustrations of the packages used in this study. Absorptive loss measurements were performed on specially fabricated TO-5 and DIP packages, and reflective loss measurements were performed on 7400 NAND gates in DIP and flat packages and 741 operational amplifiers in TO-5 and flat packages. Measurements did not indicate that reflective or absorptive losses were significantly influenced by package style. These results suggest that the choice of package style is insignificant in terms of affecting system susceptibility.

20.3.4 **GAIN-BANDWIDTH PRODUCT.** An important transistor parameter indicative of high-frequency response is the gain-bandwidth product (F_T). It equates to the frequency at which the current gain falls to unity magnitude. Devices with high values of F_T would be expected to be more susceptible to RF energy than those with low F_T values since, by definition, they are more responsive to high frequency signals.

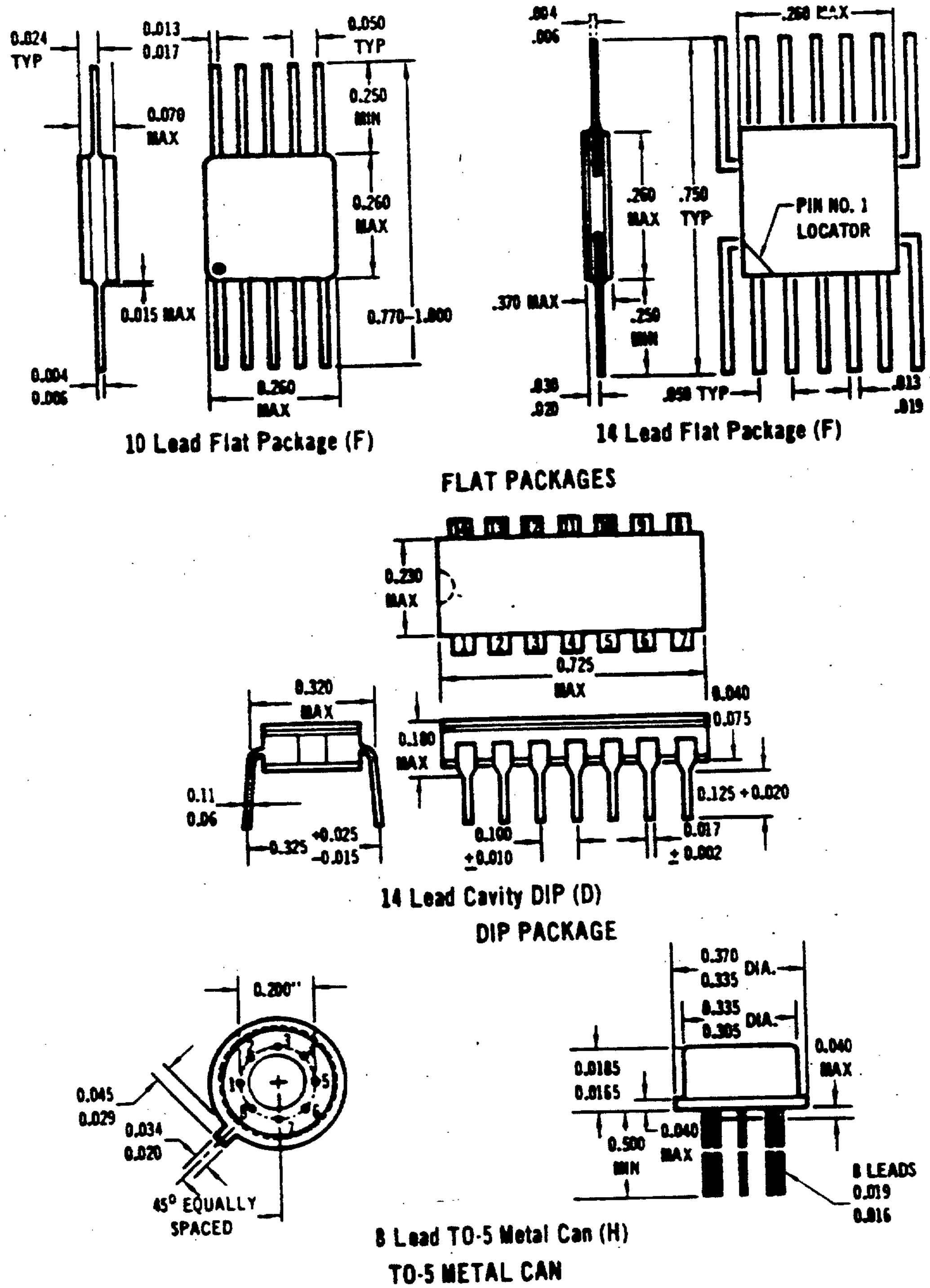


Figure E-7. Illustrations of Representative Integrated Circuit Package Types. [E-2]

There exists a significant relationship between F_T and the rectification efficiency. This is illustrated in Figure E-8, where the measured rectification efficiencies for various transistor types are plotted as a function of their manufacturer's published F_T values [E-3]. This figure represents a diverse sampling including several device types and manufacturing processes with each vertical bar representing the range of η measured for 10 transistors of each type. Despite the diversity in the devices, transistor types with high F_T's generally have larger rectification efficiencies than those with low values of F_T and are therefore more susceptible to RF energy.

20.3.5 **OPERATING CONDITIONS.** The susceptibility of a circuit to RF interference is influenced to some extent by the operating conditions of the semiconductor devices. The EMS of digital devices may be influenced by the output states and the supply voltages (i.e. signal levels), while bias levels, offset null settings, gain, and input levels may influence the susceptibility of linear circuits.

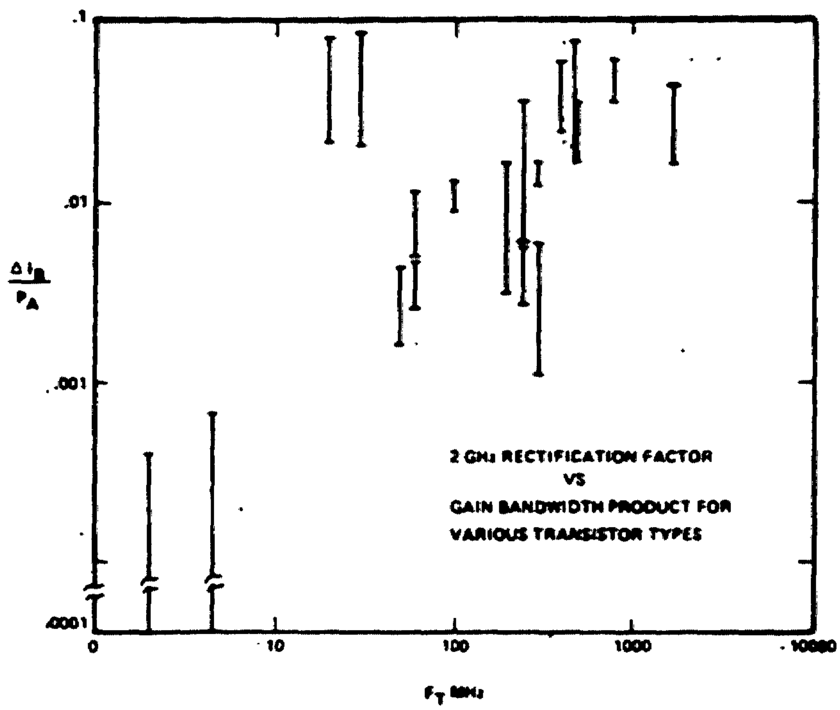


Figure E-8. Rectification Factor (η) of Various Transistor Types vs. Manufacturer Specified Gain-Bandwidth Product. [E-3]

Investigations of transistor susceptibility to conducted RF energy have shown that the base and collector bias networks can strongly influence transistor susceptibility [E-10]. Through proper biasing, excessively large collector currents caused by RF absorption can be prevented. It is usually advantageous to the hardening problem to operate potentially susceptible devices at high collector current levels and low gain, if this proves compatible with other requirements. Large signal levels also reduce susceptibility in that, for a given RF induced offset, digital circuits are less apt to change states and analog circuits will have a lower percentage change in the desired signal.

20.3.6 **ENTRY POINTS.** The sensitivity of the various device ports to coupled RF energy can be a critical factor in relation to system susceptibility. For example, signals coupled into a sensitive port such as the base terminal of a transistor may cause interference at low absorbed power levels. On the other hand, a ground lead may be able to tolerate relatively large RF power levels before interference occurs. Figure E-9 illustrates the effect of injection port on device susceptibility for a 7400 NAND gate operated in an output high state [E-11].

The actual injection port(s) will depend upon the interconnection of device leads to the system cabling, which act as antennas coupling the RF signal to the internal circuitry. One technique which may be adopted is the use of common mode rejection. The desired signal is fed into a differential amplifier through an RF coupling device which insures that interfering signals appear in equal magnitude and in phase on each amplifier input. In general, the sensitive, highly susceptible circuitry should be isolated from probable points of entry of RF energy.

20.4 **PULSE INTERFERENCE EFFECTS.** Most of the severe electromagnetic environments to be encountered by air launched ordnance systems will be due to pulsed radar transmitters which radiate high peak power in short pulses. A replica of the RF pulse envelope is created through the rectification mechanism. The effect on circuit performance of the resultant video signal can be predicted by considering basic device limitations such as switching speed and propagation delay time in digital devices and output slew rate in linear devices. Digital devices exhibit bit errors under RF stimulation which can be related to peak RF environment levels. Linear devices often respond to the average level of the environment. The following two sections summarize potential effects of pulse signals on digital and linear circuits, including the results from measurements on representative digital and linear devices.

20.4.1 **DIGITAL CIRCUITS.** RF pulse measurements made on a 7400 NAND gate have demonstrated that the peak interference effect corresponds to the peak RF power level according to CW response predictions [E-12]. These measurements were performed using pulse widths as low as one microsecond, at PRF's of up to 10 kHz, and at test frequencies of 0.22, 0.91, 3.0, and 5.6 GHz. Figure E-10 displays typical observations of induced pulses on the device output. The device response due to

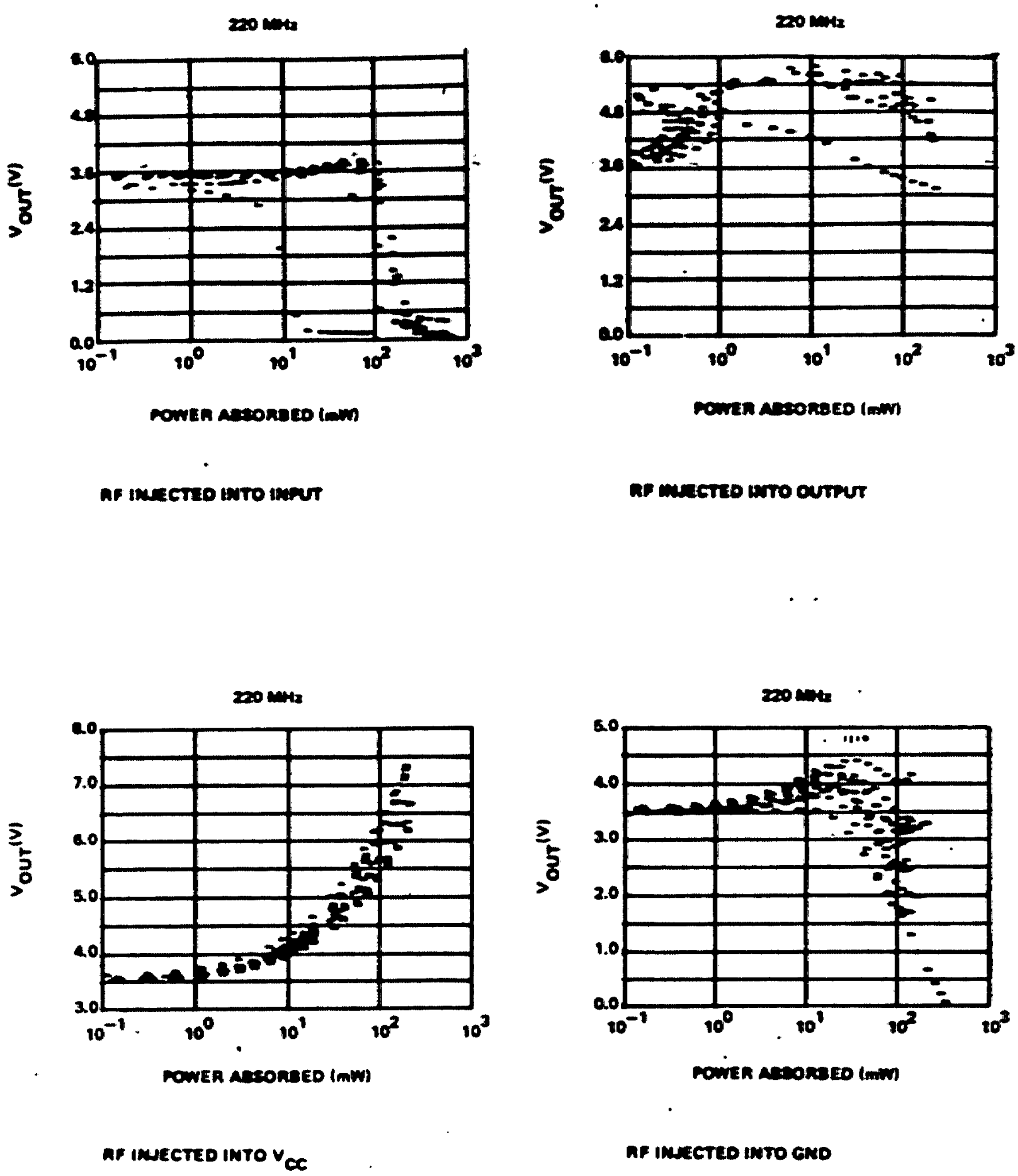
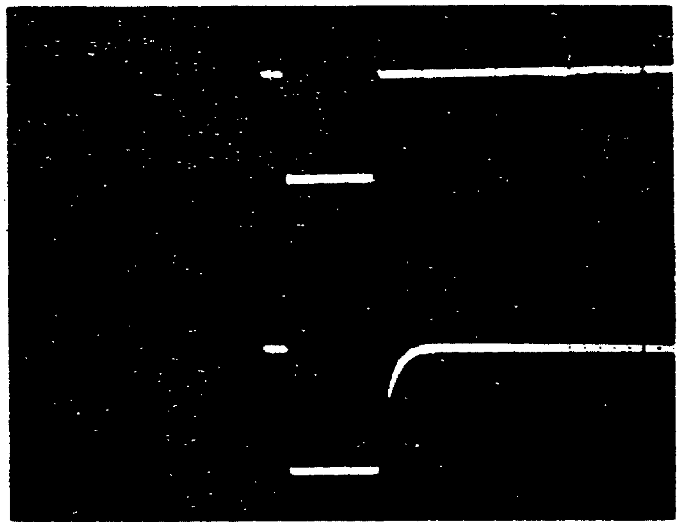


Figure E-9. Illustration of Susceptibility vs. Injection Port for 7400 NAND Gate - Output High. [E-11]

RF INTO INPUT PORT, OUTPUT HIGH

.91 GHz



$V_{out}\left(\frac{2V}{DIV}\right)$

RF PULSE ENVELOPE

10 μsec/DIV.

Figure E-10. Typical Pulse Interference Effect on 7400 NAND Gate. [E-12]

RF pulses was essentially the same as the video pulse response in all cases tested. Also, the peak output during the time the pulse was present was identical to the peak output observed under CW stimulation.

Interference due to RF pulse signals may manifest itself in a bit error rate or, more subtly, as an increase in the probability of bit errors due to a rise in the overall noise level. For a repeated interfering stimulus (e.g., a pulsed radar environment), the overall effect as measured by the bit error rate depends upon the information stream being processed by the circuit, the clock rate of the information, the pulse width and pulse interval of the interfering signal and, to some extent, the relative phasing of the two pulse streams.

Figure E-11 illustrates a few of these concepts. A data stream consisting of alternating highs and lows was supplied to a 7400 NAND gate [E-12]. An RF pulse injected into the input inhibited the output from going high, which causes a bit error every other bit for the duration of the pulse. For the case of an RF pulse injected into the output, the output remained high, once again giving rise to a bit error on

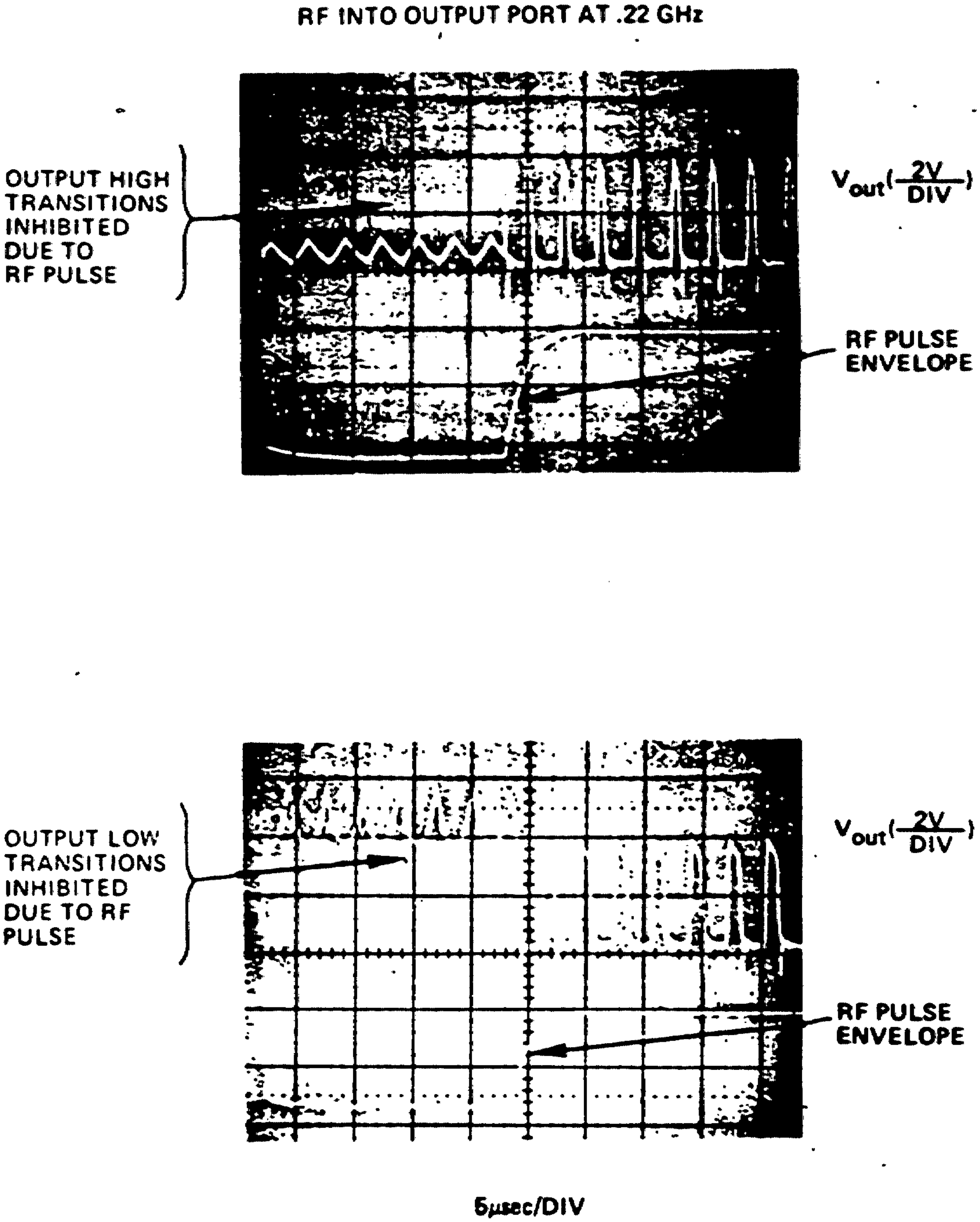


Figure E-11. RF Pulse Response of 7400 NAND Gate Under Dynamic Conditions. [E-12] every other bit. The maximum number of bit errors per RF pulse is given by:

$$\text{Bit errors (max)} = \text{(data rate)} \times \text{(pulse width)} \qquad \text{(E-4)}$$

Therefore, the maximum bit error rate is given by:

$$\text{Bit error rate (max)} = \text{(data rate)} \times \text{(pulse width)} \times \text{(PRF)} \qquad \text{(E-5)}$$

Equation (E-5) can be normalized by dividing through by the data rate term to give:

$$\text{Functional bit error rate (max)} = \text{(pulse width)} \times \text{(PRF)} = \text{duty cycle} \qquad \text{(E-6)}$$

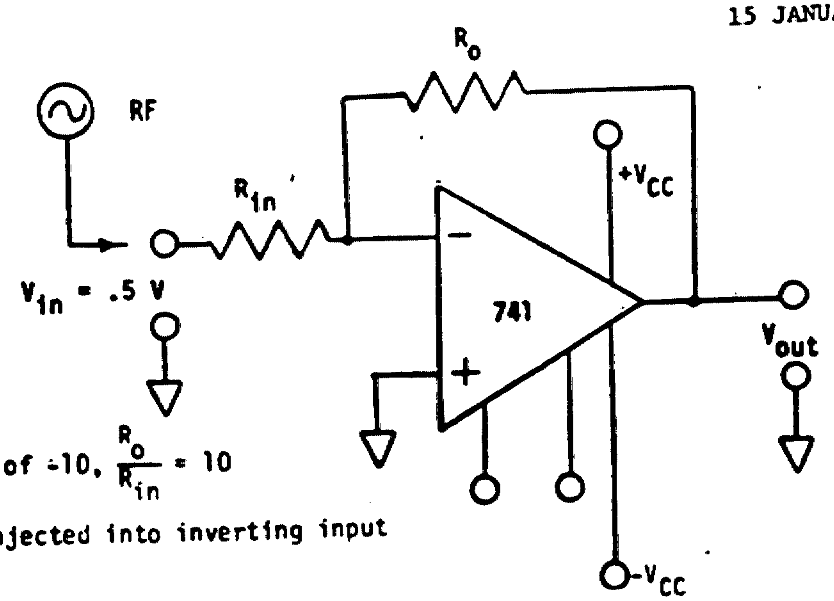
The duty cycle of a radar gives the ratio of average output power to peak output power. Consider a simplistic example which ignores simultaneous interference on different devices and the possibility of multiple emitters. Using a duty cycle of .001 for a pulsed radar emitter powerful enough to cause interference, a system designer could expect a bit error rate equal to 0.1% of his clock rate.

20.4.2 **LINEAR CIRCUITS.** RF pulse tests performed on a 741 operational amplifier have demonstrated that the interference effect in a linear circuit is a simple superposition of the interference signal and the normal device signal [E-12]. The test configuration is illustrated in Figure E-12. The interference model which explains the observed interference effects is shown in Figure E-13 and consists simply of an offset voltage generator (V_{os}) in the inverting input arm. The dependence of V_{os} upon the RF drive level is plotted in the lower half of the figure. For the particular feedback network used, the influence of V_{os} on the output voltage (V_{out}) can be shown to be given by:

$$V_{out} = \frac{-R_o}{R_{in}} V_{in} + \frac{R_{in} + R_o}{R_{in}} V_{os} \qquad \text{(E-7)}$$

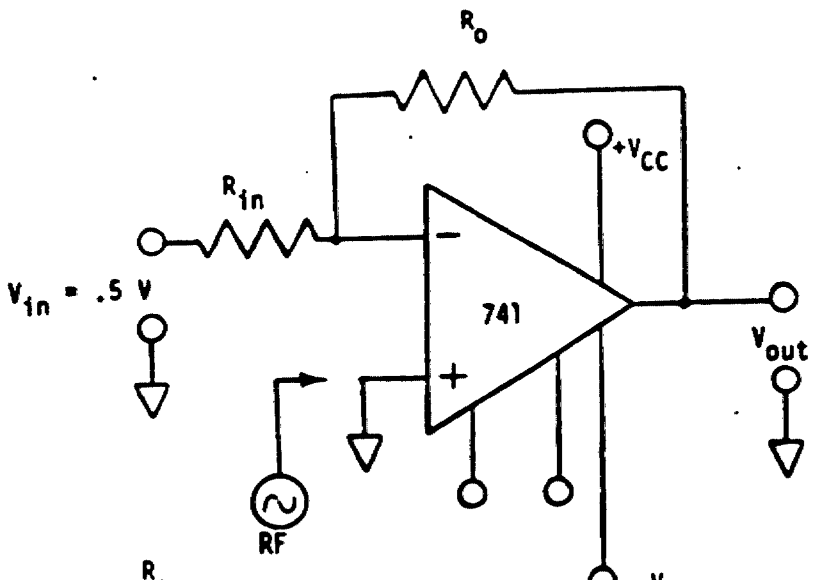
The value of V_{os} can be positive or negative depending on the port of injection and the type of input transistors [E-13]. If NPN transistors are used, V_{os} is negative for injection into the inverting input and positive for injection into the non-inverting input. The polarities are reversed for PNP input transistors. The other ports (such as output, $+V_{cc}$, $-V_{cc}$, offset null, etc.) show a similar dichotomy and are referred to as inverting input-like and non-inverting input-like according to the sign of V_{os}. Figure E-14 demonstrates the excellent agreement of this model with the observed phenomena.

The output saturation limits, which are determined by the positive and negative supply voltages, set absolute limits beyond which the output voltage cannot go. Thus, the maximum RF interference effect depends upon the sign of V_{os} and the value of V_{out} (with no RF present). The output slew rate also sets a fundamental limit on the pulse response



for gain of -10, $\frac{R_o}{R_{in}} = 10$

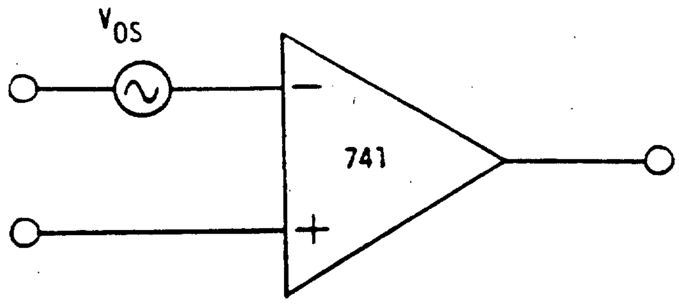
a) RF injected into inverting input



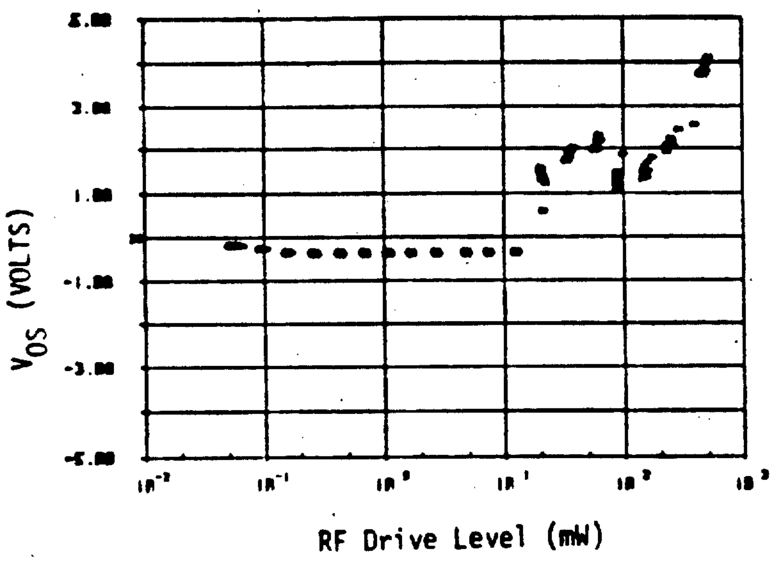
for gain of -10, $\frac{R_o}{R_{in}} = 10$

b) RF injected into non-inverting input

Figure E-12. Test Configurations for 741 Pulsed Interference Tests. [E-'2]



(a) Interference Model



(b) Functional Dependence of V_{OS} on RF Drive Level

Figure E-13. Interference Model for 741 Operational Amplifier. [E-12]

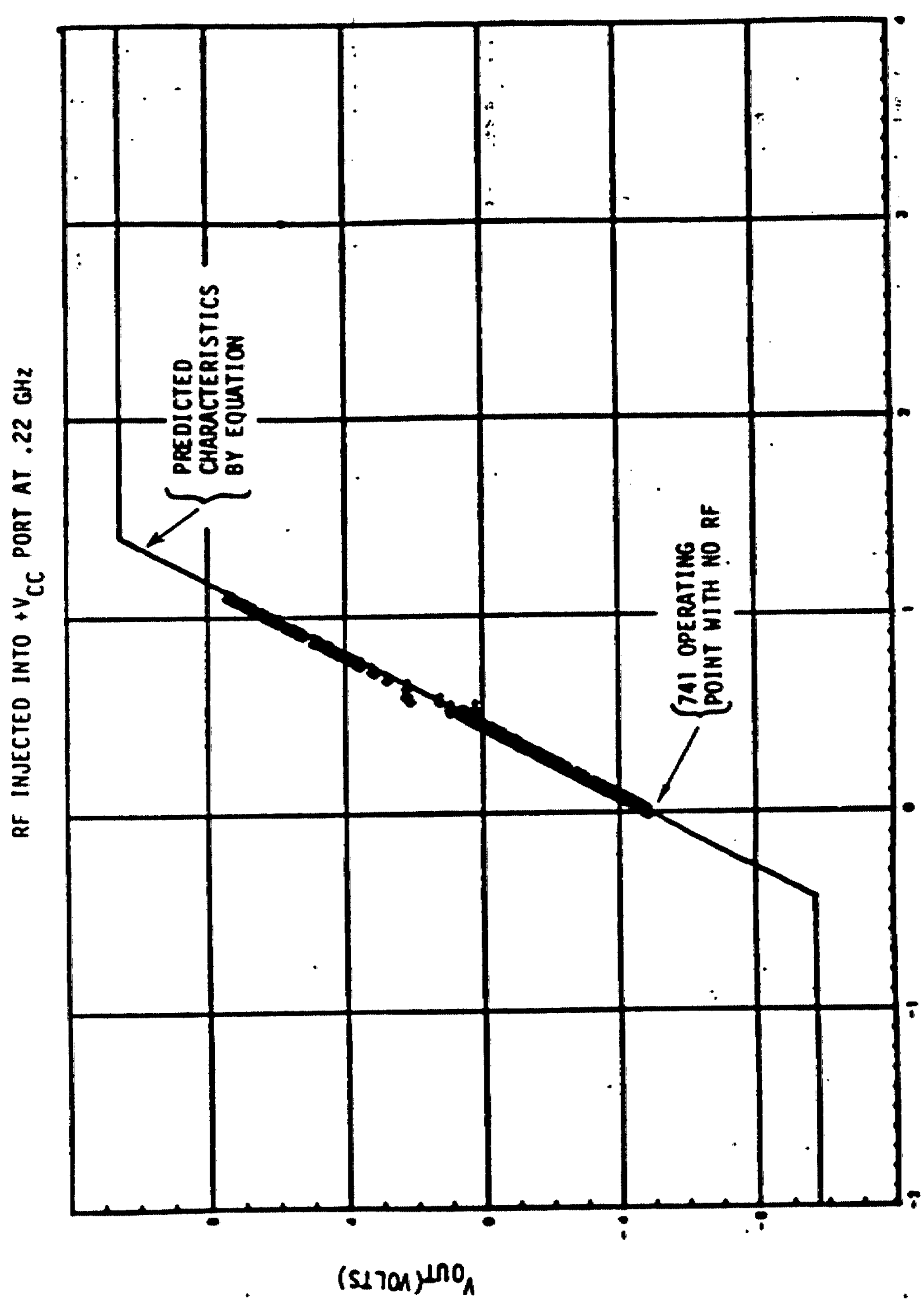


Figure E-18. 741 Input-Output Characteristics with RF Response Superimposed. [E-12] capability of the amplifier. For an input pulse (square wave) with a given pulse width, PRF, and amplitude, the output pulse will be either trapezoidal or triangular with rise and fall times determined by the slew rate. A triangular output would imply, neglecting the limiting case, that the peak value was not reached and the inteference would therefore be greater for longer pulse widths. As with the 7400 digital device, no PRF difficulties arise up to a maximum of 10 kHz [E-12].

Interference in linear circuits due to RF pulses depends on many parameters such as the level of both the intended and interfering signals, the pulse width and PRF of the interfering signal, response times, and slew rates. In many circuits, the interference may be treated as noise and its effect on the signal-to-noise (S/N) ratio can be determined. For relatively simple interference pulse trains, Fourier analysis will quickly identify the spectral components of the interference. The zero frequency (dc) term is related to the average interference level and is easily calculted from:

Average interference level = (peak value) x (duty cycle) (E-8)

As an example, an interference pulse train with a peak amplitude of one volt and a duty cycle of .001 will have an average interference level of one millivolt. Such an interference level could be quite serious in a low level pre-amplifier, for instance, but would have little effect in a power stage. Also, many linear circuits interface with transducers, electromechanical devices, etc., which have slow response times compared to typical radar pulse widths. This class of circuits would respond to the average value only.

On the other hand, tuned circuits can be expected to "ring" when driven by short pulses, and hence, could respond to the peak amplitude and PRF of the interfering signal. Comparator circuits would also be expected to respond to the peak levels of the interfering pulse; however, their finite slew rates may limit the response.

30. **SUSCEPTIBILITY DATA.** The objective of this section is to document representative data on the susceptibility levels of many commonly used devices and IC's. As was already mentioned, a totally comprehensive set of EMS data is not available, and yet some means is necessary for evaluating a component on which no data is accessible. Meaningfulness, practicality, and manageability considerations tend to dictate the use of composite worst-case curves.

Information obtained from worst-case curves prove to be more meaningful than data on individual devices. Susceptibility level consistency is not designed into the manufacturing of present-day devices and IC's. As a result, a sampling of a single device or IC type often demonstrates a widespread variation in susceptibility thresholds. As an example, consider Figure E-15, which was derived from a group of ten "identical" NAND gates operating in the output high state (input low) [E-11]. The effect of injecting a 220 MHz signal of sufficient strength into the input port was to change the output state from high to low. Two of the gates changed state at approximately 10 milliwatts while most of the others required over 100 milliwatts of absorbed power

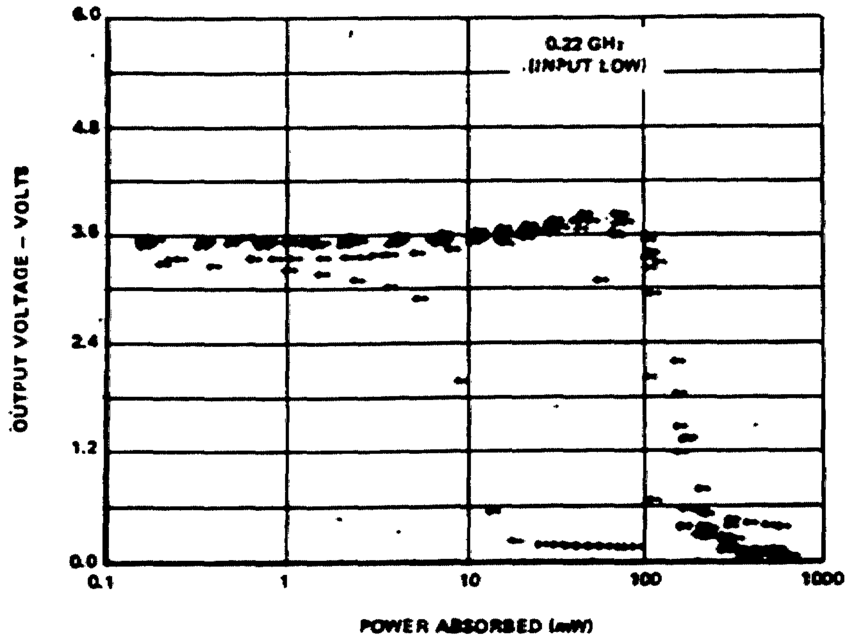


Figure E-15. Output Voltage vs. RF Power Injected into Input of NAND Gate. [E-11]

to induce a similar transition. This represents approximately a 10 dB variation in susceptibility levels. A worst-case value of 10 mW must be used unless the capability exists for measuring each device individually (unlikely in a system with a few thousand components).

The use of composite curves aids in eliminating the need for an unmanageably large volume of data. Figure E-15 pertains to only one particular type of NAND gate. In all likelihood, a designer would choose a NAND gate different from this particular one. Obtaining data of sufficient scope using this approach would require a virtually limitless number of graphs (consider the number of graphs similar to Figure E-15 required to cover but a small portion of the device and IC types available to a system designer). A set of composite worst-case curves tend to be more practical. These curves can be updated quite easily as more data becomes available. Figure E-15 would represent but one point on a composite worst-case curve and this point would appear only if the threshold (10 mW) were lower than that of all other NAND gate types measured at 220 MHz. If a particular type of component is not included in the graph, the designer can be more confident in using a value that is assumed to be representative (in the worst-case) of the component under consideration. This worst-case approach affords some degree of safety but it does not necessarily create unrealistic hardening requirements. If the electromagnetic environment (EME) defines a wide frequency band exposure, then there is a good possibility that at some point the worst-case assumptions will come close to being satisfied.

This section includes data from thousands of tests performed on semiconductor devices and integrated circuits. Continuous wave signals at frequencies of 220 MHz, 910 MHz, 3.0 GHz, 5.6 GHz, and 9.1 GHz were used in the testing of integrated circuits. EMS information is documented on linear circuits including operational amplifiers, voltage regulators and comparators, digital circuits of the TTL and CMOS families, and interface circuits of the line driver and receiver type. Also included are interference data for bipolar transistors and failure data for microwave point-contact diodes.

30.1 **OPERATIONAL AMPLIFIERS.** Operational amplifiers are often used as functional blocks in more complex integrated circuits, as they are probably the most common type of linear integrated circuit. The results of numerous susceptibility measurements performed on several representative types are presented in this section. Table E-3 lists the types of op amps that were tested [E-13].

TABLE E-3

OP AMPS TESTED. [E-13]

741
108A
201A
207
0042C
531

Op amps were found to be most susceptible to RF energy conducted into either of the input terminals. The interference effect, in this case, is the generation of an offset voltage at the particular input terminal that the RF entered. Figure E-16 illustrates the offset voltage generator, represented by v_{II}, which occurs due to rectification of the RF signal for the case of an op amp with NPN input transistors. The polarities of the offset generator, v_{II}, would be reversed for PNP type input transistors.

The susceptibility criterion used for these measurements was the magnitude of v_{II}. Figure E-17 shows the minimum power levels required to cause offsets of magnitude 0.05, 0.10, 0.15, and 0.20 volt. Other effects such as power supply current increases were observed, but these were either linked to the input offset through circuit interactions or they occurred at higher power levels.

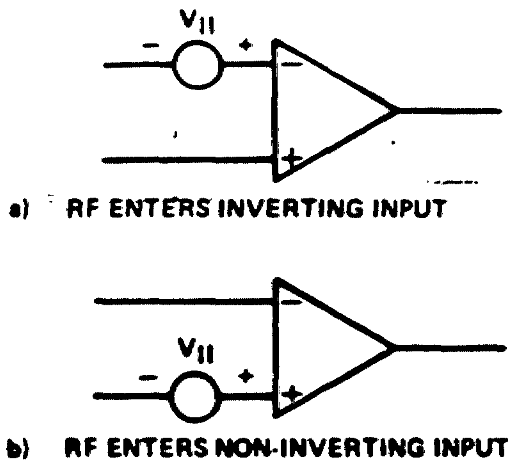


Figure E-16. Location of Offset Voltage Generator due to Rectification of RF Signal at Op Amp Inputs. [E-13]

Figure E-18 shows an op amp operating in the inverting mode with RF entering the inverting input. This figure is used to illustrate the technique for deriving the relationship between v_{II} and v_{OUT}. A simple analysis using the virtual ground concept allows one to write three equations which are sufficient for solution:

$$-v_{IN} + i_{IN} R_{IN} - v_{II} = 0 \qquad \text{(E-9)}$$

$$v_{II} - i_F R_F + v_{OUT} = 0 \qquad \text{(E-10)}$$

$$i_F = -i_{IN} = -(v_{IN} + v_{II})/R_{IN} \qquad \text{(E-11)}$$

Solving Equation (E-10) for v_{OUT} in terms of v_{IN}, v_{II}, R_F, and R_{IN} yields:

$$v_{OUT} = -\left(\frac{v_{IN} + v_{II}}{R_{IN}}\right)R_F - v_{II} = -v_{IN}\frac{R_F}{R_{IN}} - v_{II}\left(\frac{R_F}{R_{IN}} + 1\right) \qquad \text{(E-12)}$$

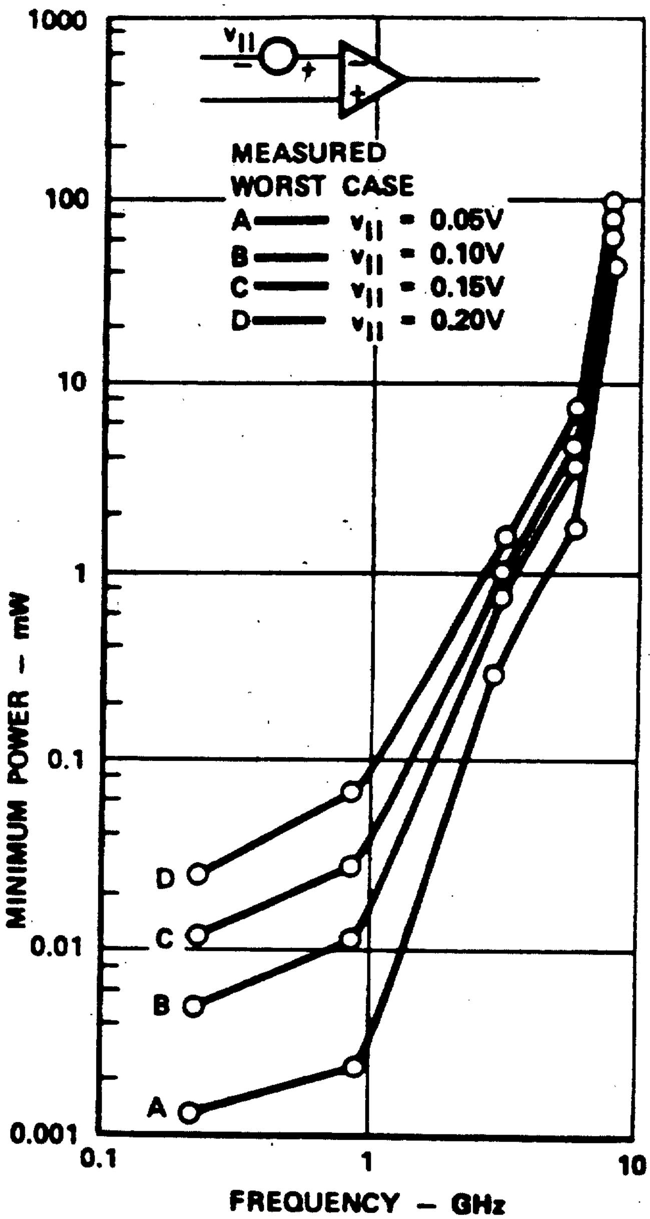


Figure E-17. Composite Worst-Case Susceptibility Values for Op Amps. [E-13]

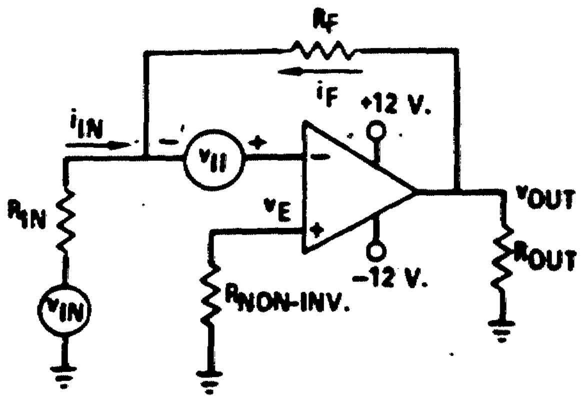


Figure E-18. Inverting Amplifier Circuit with Offset Generator Shown at Op Amp Inverting-Input Terminal. [E-13]

The interference effect seen at the output is a voltage offset which depends on v_{II} and the ratio of R_F to R_{IN}. Clearly, an op amp operated with high gain (R_F/R_{IN}) and low input signal level will be highly susceptible to RF injected into the input terminal. Other op amp circuits can be analyzed for interference effects in a similar manner.

The minimum susceptibility levels for offset voltages other than those shown in Figure E-17 can be estimated from the available data. For offset magnitudes of less than 0.05 volt, the offset voltage is approximately proportional to the minimum RF power level, $P(f,v_{II})$. $P(f,v_{II})$ indicates that the minimum power level is a function of frequency and offset voltage. For offsets greater than 0.20 volt, the offset voltage is approximately proportional to the square root of the RF power level. Thus, a reasonable procedure for estimating the minimum RF power required to cause offsets not shown in Figure E-17 is:

$$P(f,v_{II}) = \begin{cases} (v_{II}/0.05V) \cdot P(f,0.05V) & \text{for } v_{II} < 0.05V \\ \text{use Figure E-17} & \text{for } 0.05V \leq v_{II} \leq 0.20V \\ (v_{II}/0.20V)^2 \cdot P(f,0.20V) & \text{for } v_{II} > 0.20V, \end{cases} \qquad \text{(E-13)}$$

where P(f,0.05V) and P(f,0.20V) are determined from Figure E-17 at the desired frequency.

30.2 **VOLTAGE REGULATORS.** Voltage regulators are common linear IC's, of which many varieties are available. These devices may be divided into two groups consisting of 3-pin regulators and multi-pin regulators. The 3-pin devices simply have input, output, and ground terminals. They require no external components, which is the major reason for their widespread use. Multi-pin regulators do require external components, usually resistive dividers and compensation capacitors, but they are more versatile than their 3-pin counterparts. In general, multi-pin regulators have either 4, 8, or 10 terminals.

Measurements have been made to determine the RF susceptibility of both 3-pin and multi-pin regulators [E-13]. The tests were performed on 3-pin regulators, each having a nominal output voltage of 5 volts, and on 8-pin regulators designed to yield a nominal output of 12 volts. Table E-4 lists the types tested.

TABLE E-4

VOLTAGE REGULATORS TESTED. [E-13]

3-PIN (5 VOLT)
309
320
78M05
8-PIN
300
305

The 3-pin regulators were tested using a 7-volt input and six different load conditions: output currents of 1mA, 20mA, 50mA, 100mA, 150mA, and 200mA. The 8-pin regulators were tested in the configuration shown in Figure E-19. The susceptibility criterion for all devices was a 0.25 volt change in the output voltage from the "no-RF" condition. RF was conducted into each possible port; the output terminal was the most susceptible in the 3-pin regulators, while the reference-bypass and feedback terminals were most susceptible in the 8-pin devices.

The composite worst-case curves for voltage regulators are shown in Figure E-20. Clearly, the 8-pin regulators are more susceptible than the 3-pin regulators (by approximately 12 dB). An analysis of the regulator circuit reveals why this is so. A functional diagram of the basic regulator circuit is illustrated in Figure E-21. Two of the pins in the 8-pin devices correspond directly to the op amp inputs. When RF is conducted into these pins, rectification occurs and an offset voltage appears at the amplifier input terminals. This interferes with the op amp's ability to compare the voltage across R_2 to the reference voltage, which results in a deviation in the output voltage.

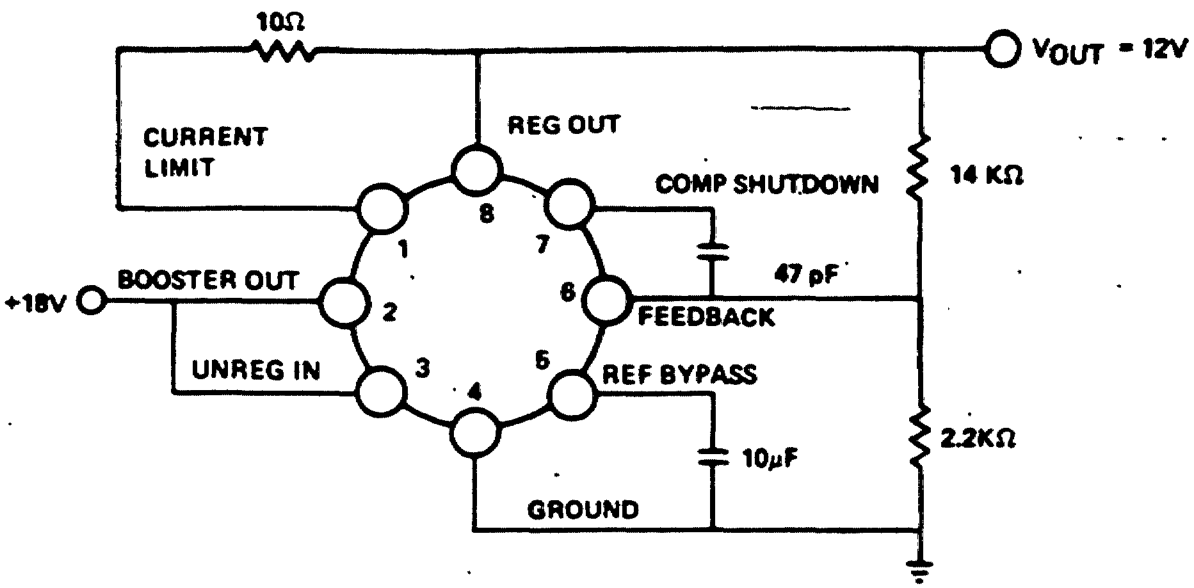


Figure E-19. Circuit for 8-Pin Voltage Regulator Susceptibility Tests. [E-13]

In 3-pin regulators, however, the resistive divider is manufactured directly on the chip. Therefore, the op amp inputs are inaccessible at the regulator terminals, which accounts for the lower susceptibility of these devices. While the compensation and bypass capacitors may offer some degree of protection by shunting the RF energy away from the amplifier inputs, the difference in the measured susceptibilities appears to be significant [E-13].

30.3 **COMPARATORS.** Comparators are common linear IC's used to detect voltage levels in electronic equipment. Measurements have been made on the RF susceptibility of several types of comparators. Table E-5 lists the types tested [E-13].

TABLE E-5

COMPARATORS TESTED. [E-13]

306
311
339
360
710
760

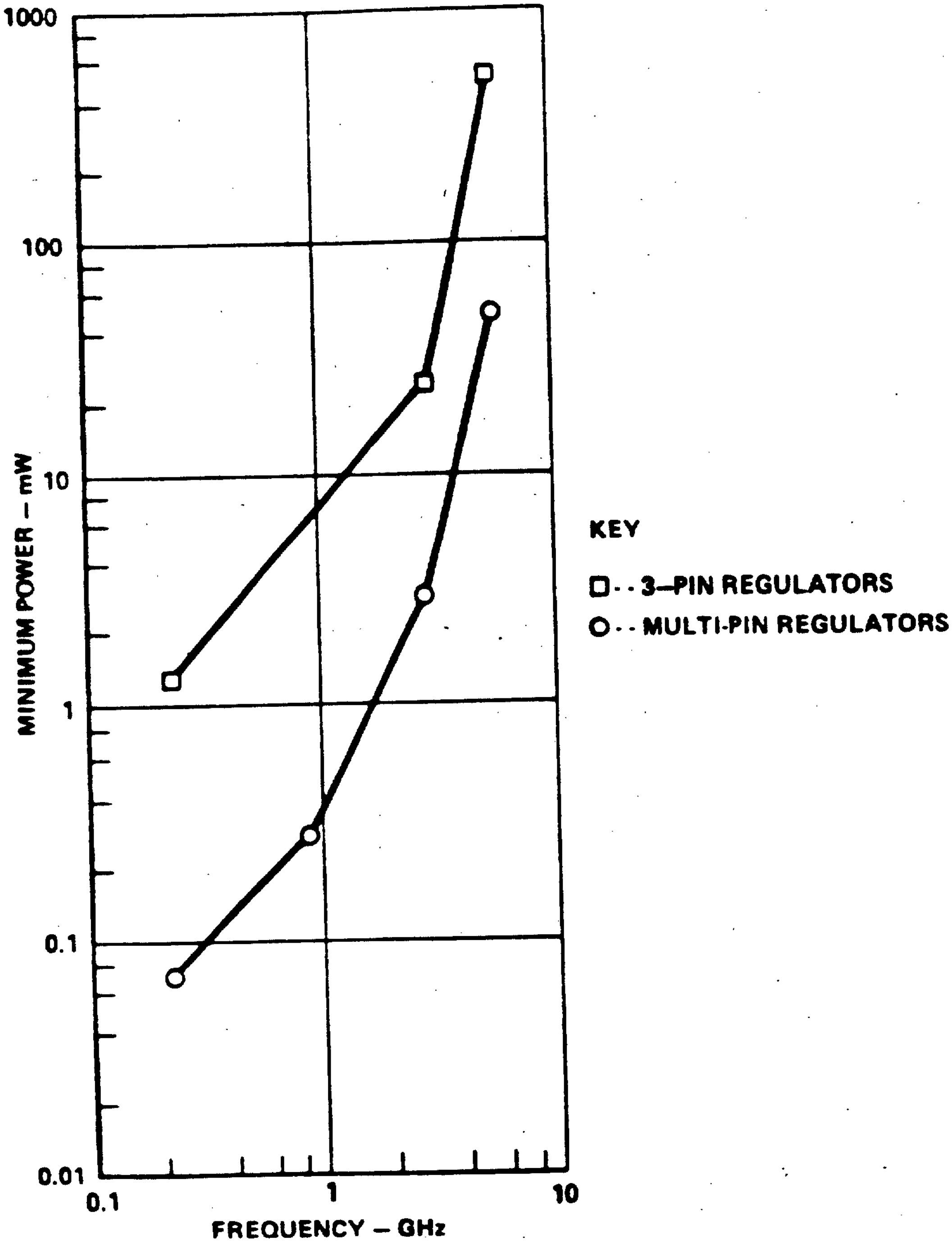


Figure E-20. Composite Worst-Case Susceptibility Values for Voltage Regulators. Output Voltge Change of 0.25 Volt is Susceptibility Criterion. [E-13]

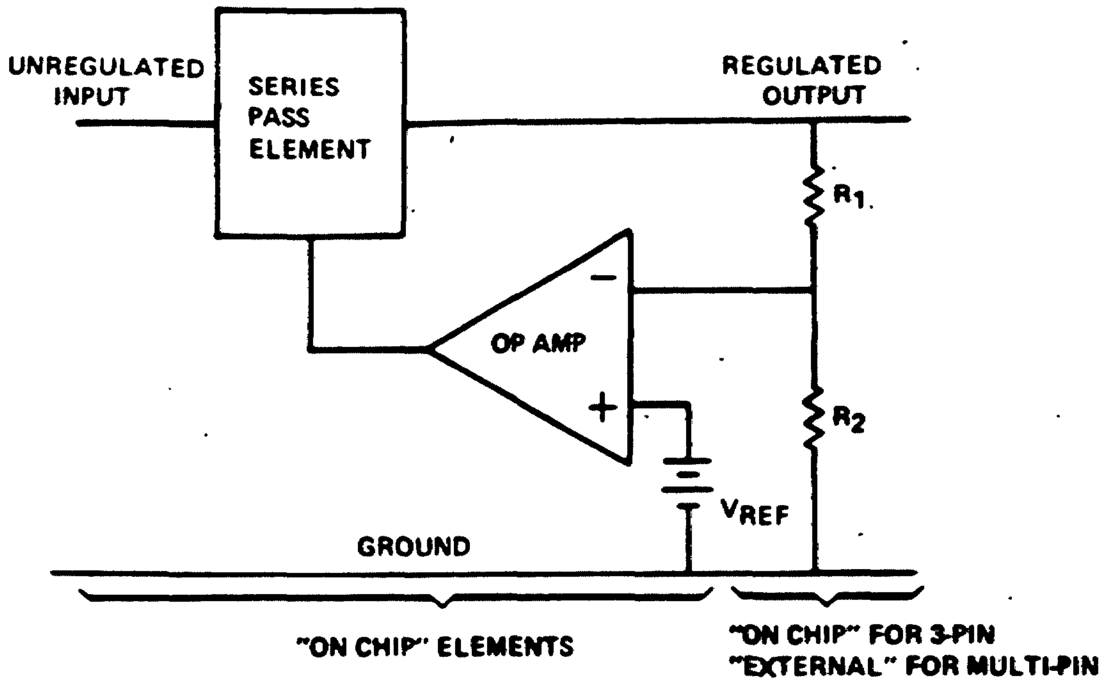


Figure E-21. Basic Series Regulator Circuit. [E-13]

Susceptibility was defined in terms of changes in the comparator switchpoints. For example, Figure E-22 shows a typical transfer curve for a 710 type comparator. The output voltage switches between high and low values for input voltages between -1.0 mV and +1.0 mV. Manufacturer specifications guarantee that switching will occur at input voltages between -3.0 mV and +3.0 mV. However, coupled RF energy can cause offsets in this switchpoint, thereby diminishing the accuracy with which the comparator detects voltage levels.

The testing revealed that comparators are most susceptible to RF energy conducted into the input terminals. This result was to be expected, since comparators contain a differential pair input stage similar to that in op amps, which are very sensitive to RF conducted into their input terminals. Rectification of the RF signal gives rise to an offset voltage at the input terminal into which the RF is conducted, causing a similar offset to occur in the comparator switchpoint.

Switchpoint offsets of +0.05, +0.10, +0.20, and +0.50 volt, representing varying degrees of interference effect, were sought in the testing. Figure E-23 illustrates the minimum powers observed to cause interference as defined by these four susceptibility criteria. The devices were susceptible to a minimum of 0.025 mW at 20 MHz, using the 0.05 volt offset criterion as the definition of susceptibility.

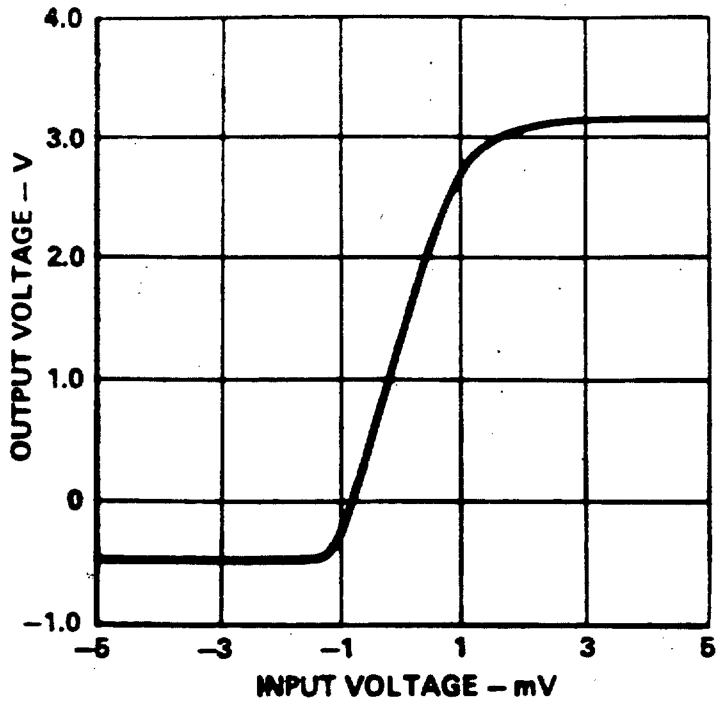


Figure E-22. Typical Voltage-Transfer Curve for Type 710 Comparators. [E-13]

30.4 **TTL DEVICES.** The TTL line is the most widely used family of digital IC's. Several types of TTL devices have been measured to determine their minimum susceptibility levels to conducted RF energy [E-13]. Table E-6 lists the specific devices tested.

TABLE E-6

TTL DEVICES TESTED. [E-13]

DEVICE NO.	DEVICE TYPE
7400	QUAD 2 INPUT NAND GATE
7402	QUAD 2 INPUT NOR GATE
7404	HEX INVERTER
7405	HEX INVERTER (OPEN COLLECTOR)
7408	QUAD 2 INPUT AND GATE
7432	QUAD 2 INPUT OR GATE
7450	EXPANDABLE DUAL 2 WIDE, 2 INPUT AND–OR–INVERT GATE
7473	DUAL J-K FLIP-FLOP
7479	DUAL D FLIP-FLOP

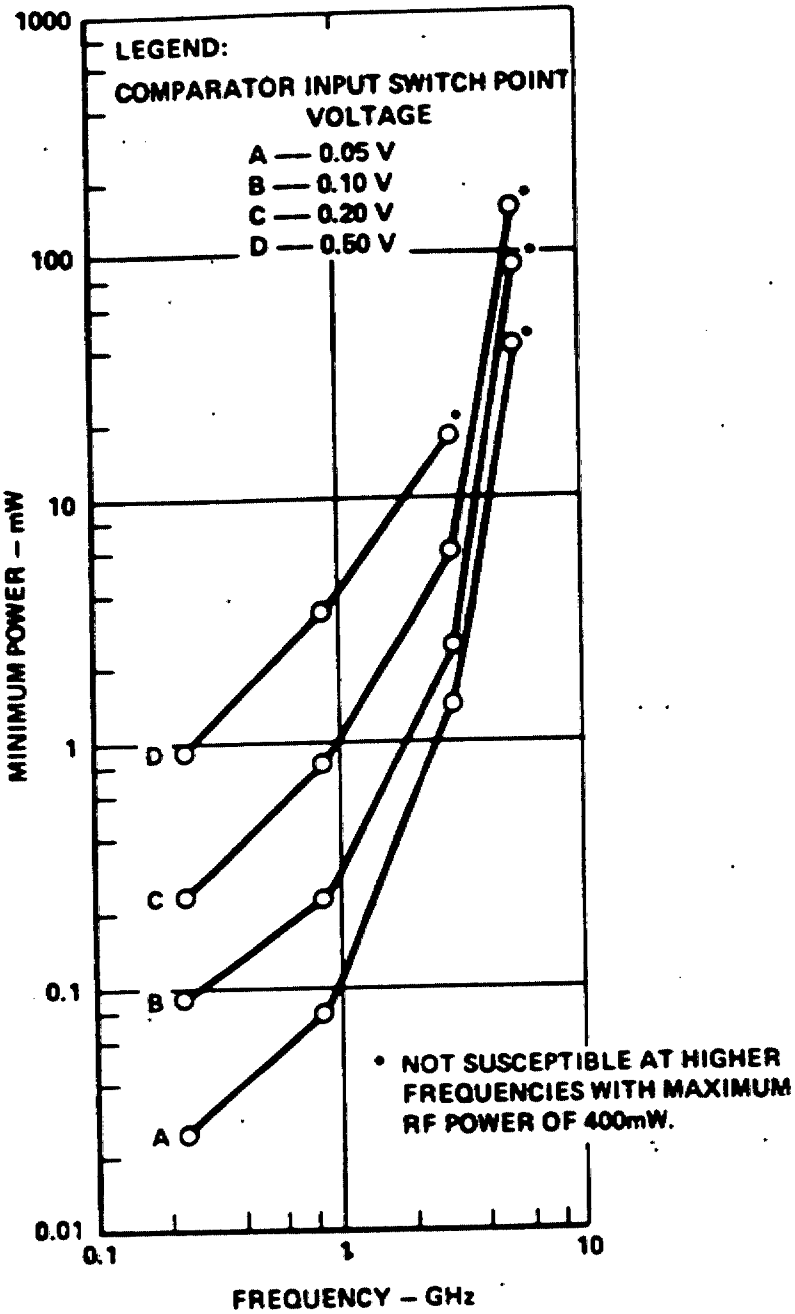


Figure E-23. Composite Worst-Case Susceptibility Values for Comparators. [E-13]

Three susceptibility criteria were used to define varying degrees of interference. These criteria were based on the manufacturers' specifications for voltage levels in TTL circuits. The least severe interference effect, designated by criterion A, is the manufacturer's guaranteed specification limit. This criterion characterizes interference as a low state output voltage which exceeds 0.4 volt or a high state output voltage below 2.4 volts. RF powers greater than the susceptibility values given by criterion A do not necessarily cause malfunction of the device, but the usual 0.4 volt noise margin is reduced, which increases the risk of operation. At RF powers below the susceptibility values given by criterion A, no interference effect will occur.

Criterion B is the outer edge of the noise margin; it is exceeded when the device low state output voltage is greater than 0.8 volt or when a high state output voltage is less than 2.0 volts. Beyond these thresholds, succeeding stages may misinterpret the logic state, resulting in a bit error. Operation with RF powers above the susceptibility limits for criterion B is not recommended due to the high likelihood of logic state errors.

Criterion C defines the most severe interference effect. The output voltage limits for this case, low output voltage greater than 2.0 volts or high output voltage less than 0.8 volt, are the voltages at which state changes are certain. Bit errors and incorrect system outputs would be expected for absorbed RF powers greater than the thresholds specified by criterion C.

Figure E-24 displays the composite worst-case susceptibility levels as defined by each of the three interference criteria. Measurements were also made using changes in the package supply current as the susceptibility criteria [E-13]. However, it was found that significant increases did not occur until the RF power level was far above the levels sufficient to induce state changes in the output voltage. It was also found that the differences in the susceptibilities of standard TTL series (54/74), the low power (54L/74L), and the high speed (54H/74H) TTL circuits was probably not great enough to be significant [E-13]

30.5 **CMOS DEVICES.** CMOS IC's are widely used in logic applications requiring low power consumption. The RF susceptibility of several types of CMOS devices have been tested, including types with and without protective input diodes [E-13]. Table E-7 contains a list of the device types tested.

TABLE E-7

CMOS DEVICES TESTED. [E-13]

DEVICE NO.	DEVICE TYPE
4011A	QUAD 2 INPUT NAND GATE
4011B	QUAD 2 INPUT NAND GATE
4007A	DUAL COMPLEMENTARY PAIR PLUS INVERTER
4007B	DUAL COMPLEMENTARY PAIR PLUS INVERTER
4001A	QUAD 2 INPUT NOR GATE
4013A	DUAL "D" – TYPE FLIP-FLOP

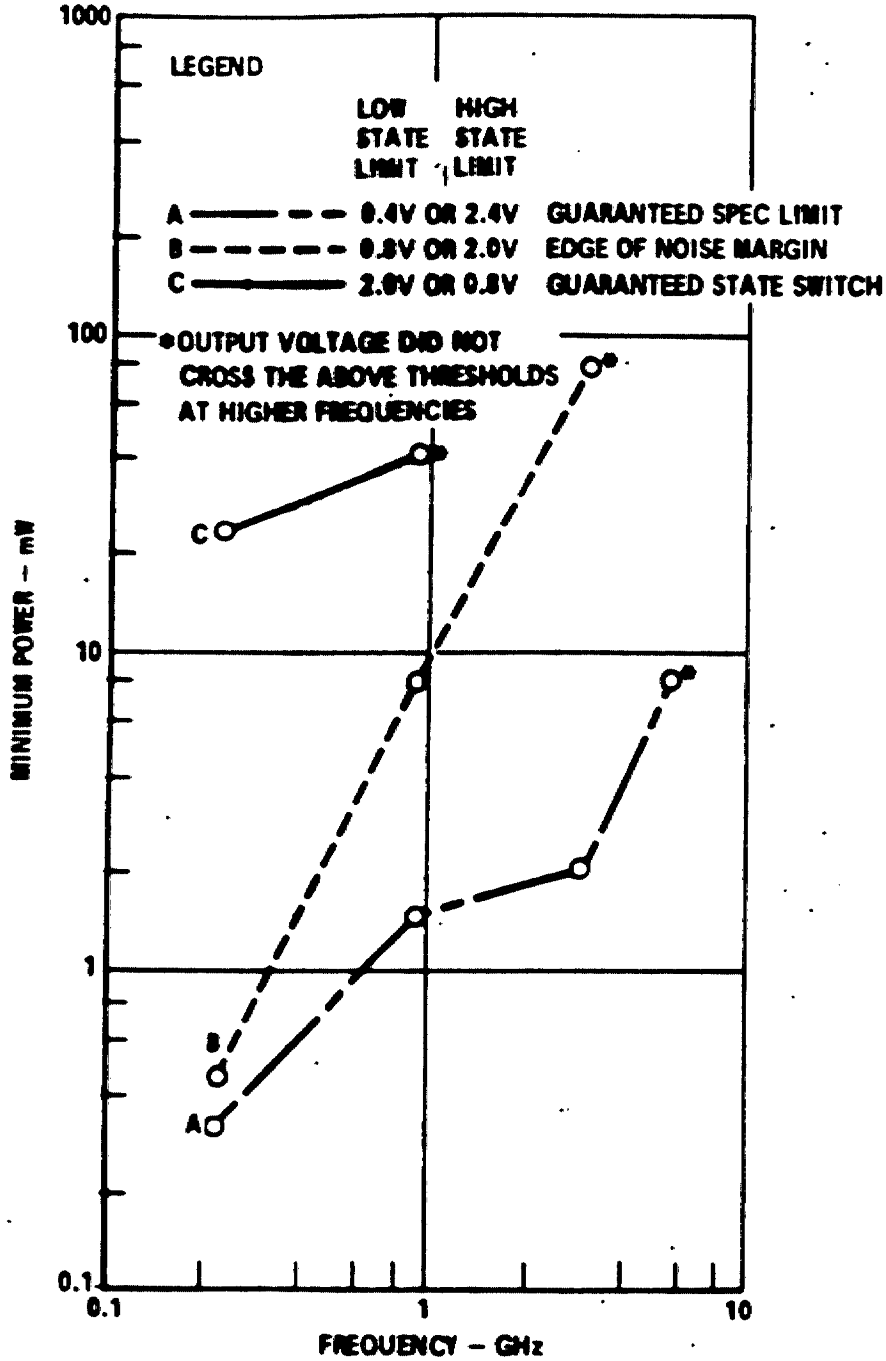


Figure E-24. Composite Worst-Case Susceptibility Values for TTL Devices. [E-13]

As with the TTL susceptibility data, the thresholds for high and low output voltages were combined and three distinct interference criteria were defined. The first criterion indicates when the output voltage is no longer within the guaranteed specification limits. Manufacturers guarantee that the maximum low state output voltage is 0.05 volt and the minimum high state output voltage is 4.95 volts for a supply voltage of 5 volts. The devices continue to operate beyond these thresholds, but with a reduced noise margin. The second criteria represents the edge of a 1-volt noise margin, meaning the maximum low output voltage is 1.05 volts and the minimum high output voltage is 3.95 volts. These values are guaranteed by the manufacturer to be correctly recognized by succeeding CMOS devices. Operation outside this range could result in logic state errors and is not recommended. The third criterion was arbitrarily defined as a 2 volt offset from the ideal output voltages. The thresholds are a maximum 2 volts for the low output state and a minimum 3 volts for the high output state. Operation outside this range has a higher probability of logic state errors than the second criterion, and should be avoided.

Figure E-25 shows the composite worst-case susceptibility values for the device types tested. The results indicate a minimum susceptibility of 1 mW of RF power at 220 MHz. A comparison with the minimum susceptibility for TTL devices at 220 MHz shows CMOS to be approximately 5 dB less susceptible. If the wider noise margin of CMOS is taken into consideration, these devices appear to be approximately 10 dB less susceptible than TTL devices.

30.6 **LINE DRIVERS AND RECEIVERS.** Line drivers and receivers are often used to transmit digital data over long system interconnect cables. The amount of RF energy conducted into these devices may be greater than that of most other system components in that long interconnect cables may be relatively efficient receptors of RF energy. Therefore, special care should be taken to prevent interference from occurring in these devices. Adequate shielding and a reduction of the data transmission rate will ensure signals of acceptable quality. Measurements have been made on the susceptibility of several representative line drivers and receivers [E-13]. The data presented in this section should enable designers to estimate the susceptibility of line driver and receiver pairs and the reduction in data rate required for quality transmission.

Table E-8 lists the line drivers and receivers used in the testing. Tests of drivers and receivers were conducted independently. The susceptibility criteria for line drivers were based on changes in the output voltage from the nominal value. Each output terminal was considered separately, and the device was considered susceptible if either output crossed the appropriate interference threshold. The 8830 and 9614 line drivers were tested with resistors across the output terminals simulating normal terminations. The type 55109 and 55110 line drivers have open collector (current type) outputs which were connected to pull-up resistors and a +5-volt supply to give a 0-5 volt range for the output voltage. When the drivers were in a nominal low state, output voltage thresholds of 0.4, 0.8, and 2.0 volts defined increasing degrees of interference. When the output voltage was in a nominal

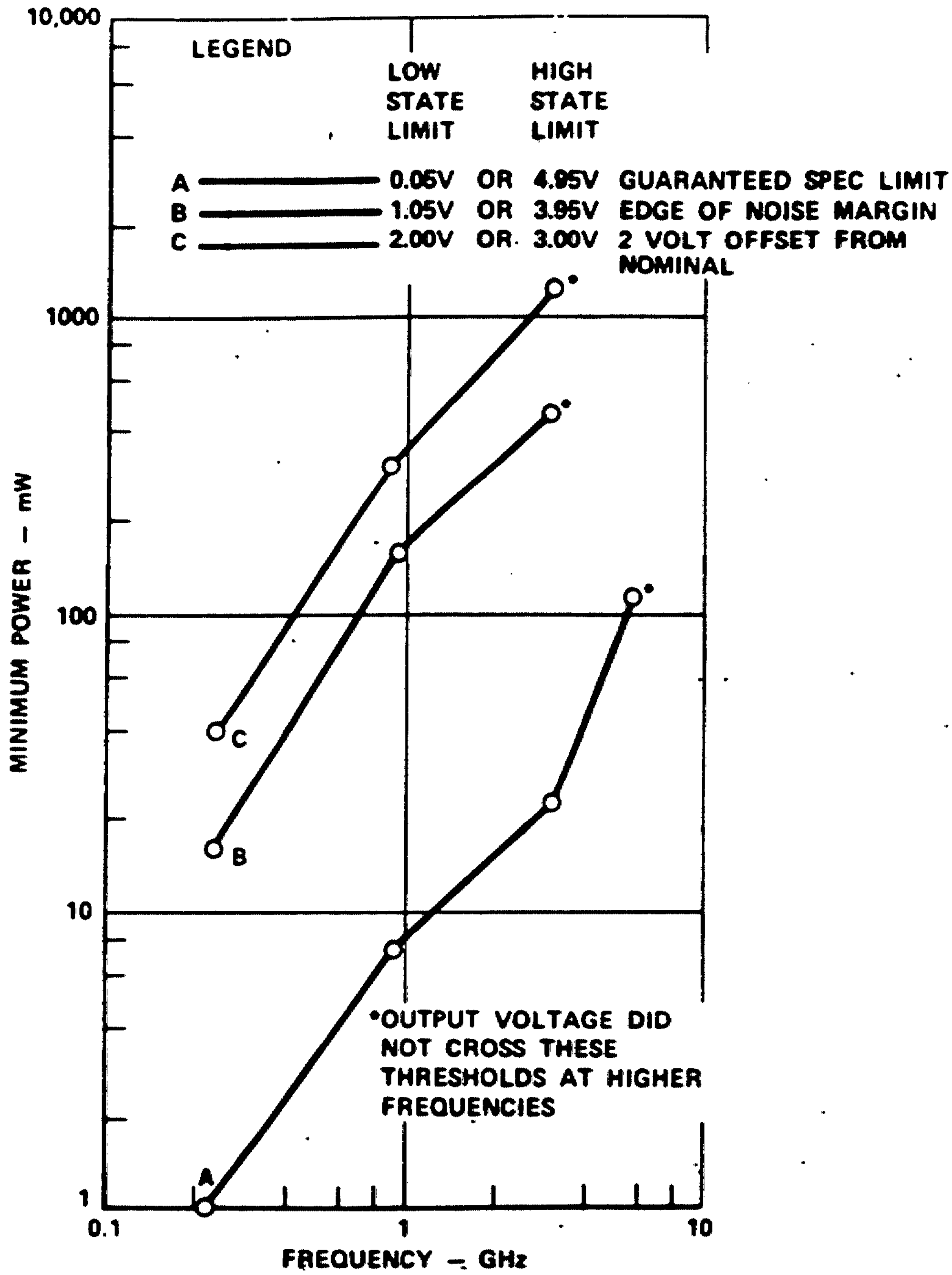


Figure E-25. Composite Worst-Case Susceptibility Values for CMOS Devices. [E-13]

TABLE E-8

LINE DRIVERS AND RECEIVERS TESTED. [E-13]

LINE DRIVERS	LINE RECEIVERS
8830	8820
9614	9615
55109	55107A
55110	

high state, increasing interference was defined by 2.4, 2.0, and 0.8 volt thresholds.

Susceptibility for line receivers was defined in terms of changes in the input threshold voltage which determined the receiver switchpoint. As an example, Figure E-26 shows the input-output transfer curve for

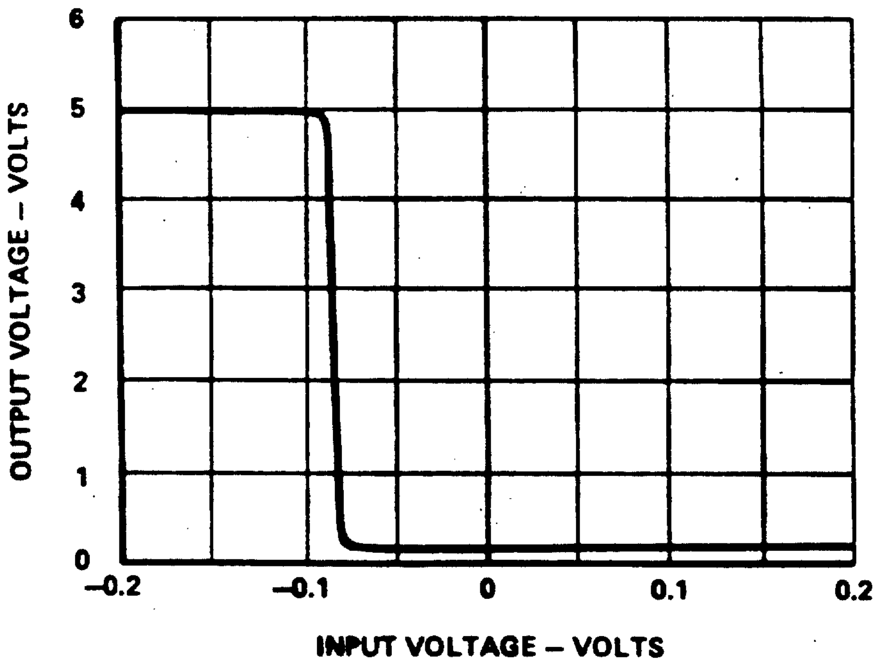


Figure E-26. Typical Input-Output Transfer Characteristic for 9615 Line Receiver. [E-13] a 9615 type receiver. At input voltages (differential input voltage between the two input terminals) below -0.08 volt, the receiver input voltage is 5.0 volts, which is a high state output. When the input voltage is greater than -0.08 volt, the output voltage is 0.2 volt, a low state output. Thus, -0.08 volt is the input threshold voltage (V_{th}). Manufacturer specifications guarantee this threshold will be between -0.5 and +0.5 volt for this device. A threshold voltage outside this range reduces the noise margin of the device and may cause bit errors in noisy environments. Input threshold voltage changes of 0.5, 1.0, 2.0, and 5.0 volts were used for the susceptibility criteria during the testing. Threshold changes of 0.5, 1.0, and 2.0 volts represent decreasing system noise margins. A 5.0 volt threshold change denotes zero noise margin, and probable malfunction of the device.

Figure E-27 shows the minimum susceptibilities measured for line driver and receiver pairs. Line receivers were found to be significantly more susceptible than line drivers, so Figure E-27 is actually a plot of the susceptibility levels of line receivers. Since line receivers are the "weak link" of a line driver and receiver system, the susceptibility of the pair is adequately described by the susceptibility of the receivers alone. Line receivers were found to be approximately 7 dB more susceptible than line drivers. (However, line driver susceptibility lies within 0.5 dB of receiver susceptibility at 910 MHz.)

The strobe and response-control terminals were found to be the most susceptible line receiver terminals. However, the strobe and response-control terminals, unlike the inputs, are rarely connected to system interconnect lines, which may be the major receptors of RF energy. Thus, the susceptibility of the input terminals may be more important to the system designer than the susceptibilities of the other terminals. The inputs were found to be approximately 4 dB less susceptible than is indicated in Figure E-27 (all points of which occurred with RF conducted into the strobe and response-control terminals).

Offsets in the threshold voltage have other effects on a signal besides reducing the noise margin. Where long lines are used, threshold offsets can cause time variations in the received signals from those sent by the driver. As a result, pulses may appear shifted in time in the received signal, or have longer or shorter durations than in the original signal. The quality of a received signal can be expressed in terms of what is called "percent jitter" [E-15]. This is a ratio of the maximum relative time variations in the original and received signals to the minimum pulse period. For example, Figure E-28 shows two pulse trains. The upper trace is the pulse train entering the driver, and which is to be sent by the system. The lower trace represents the pulse train which emerges from the receiver after transmission via the long signal line. In addition to a propagation delay, the second pulse in the received train is shifted in time with respect to its position in the original train. The percent jitter is:

$$\text{Percent Jitter} = \frac{t_3 - t_2}{t_1 - t_0} \times 100\% \qquad \text{(E-14)}$$

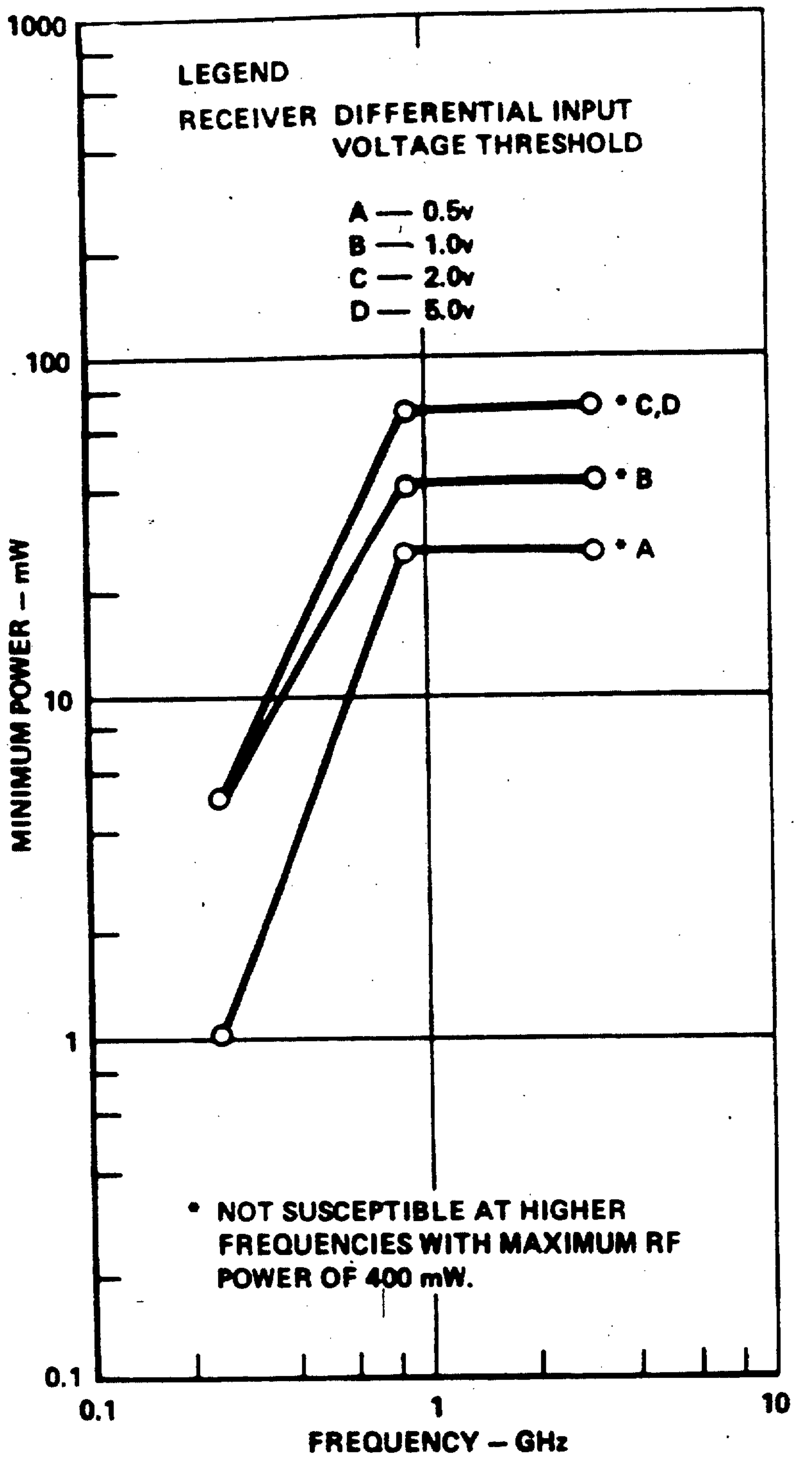


Figure E-27. Composite Worst-Case Susceptibility Values for Line Drivers and Receivers. [E-13]

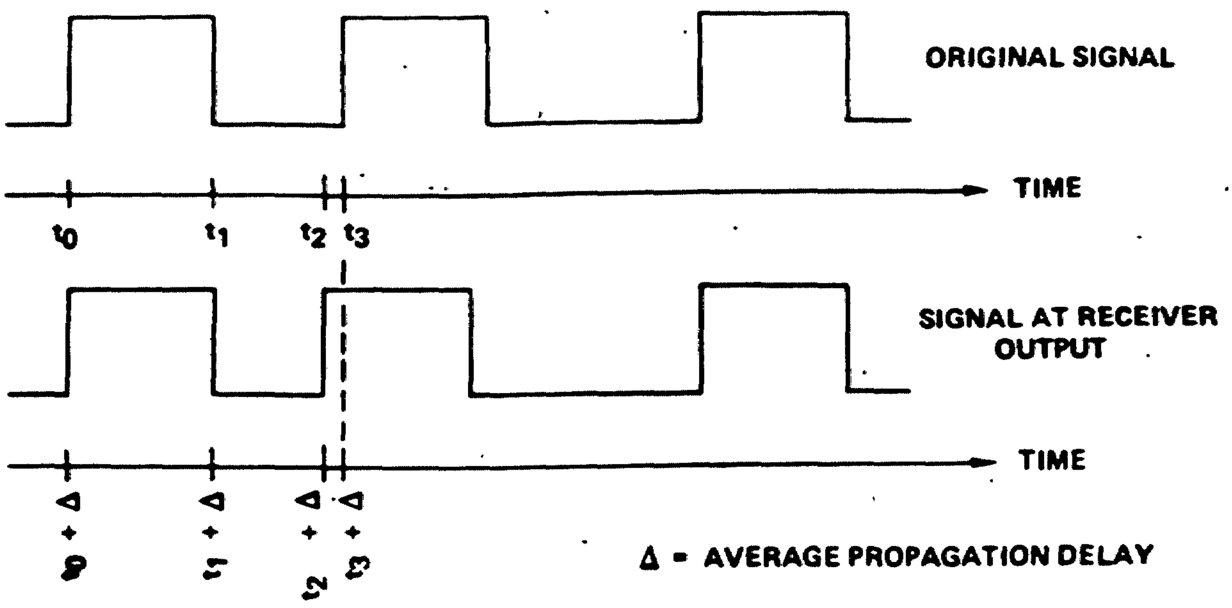


Figure E-28. Illustration of Jitter in Signal After Transmission via Long Line. [E-13]

The jitter is related to the data rate (or minimum pulse width) and line length as shown in Figure E-29. This graph was made using the following assumptions:

a. The driver "1" and "0" levels are matched exactly.

b. The receiver threshold is exactly the mean of the "1" and "0" levels produced by the driver.

c. Time delays through both driver and receiver, for both logic states, are symmetrical and have zero skew.

d. The line is perfectly terminated.

e. The line charges at an exponential rate.

The line was assumed to have a time delay of 1.7 nsec/ft, which is a typical value for a twisted pair line. Reference E-15 recommends that systems be operated with a minimum pulse width (t_{ui}) greater than 4 times the rise time of the line (t_r), which yields a jitter of less than 0.002% under these conditions. Data with jitter greater than 100% are probably not recoverable.

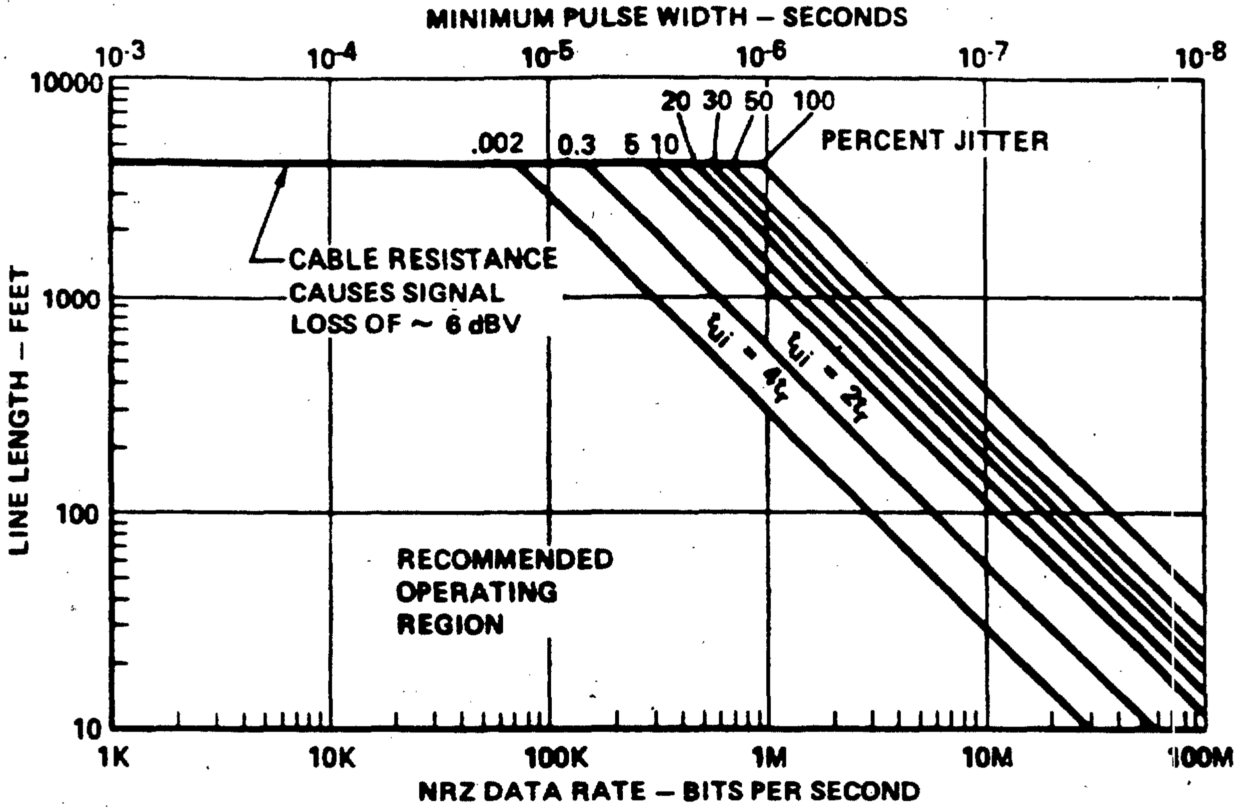
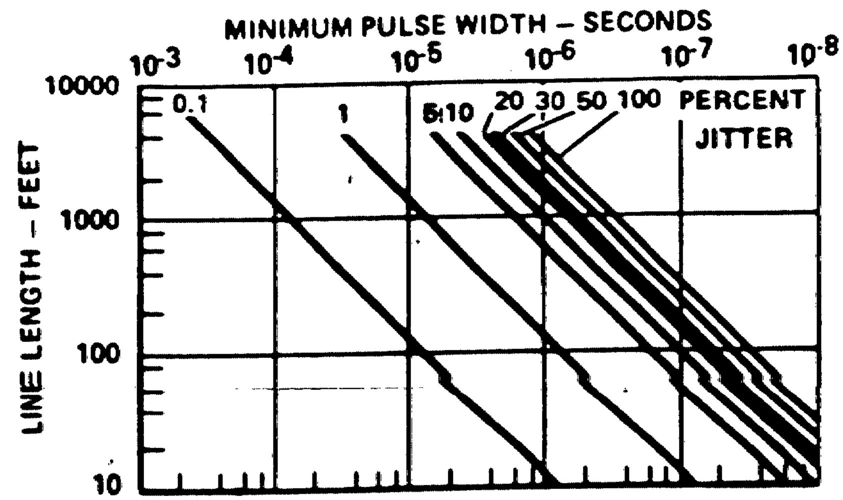


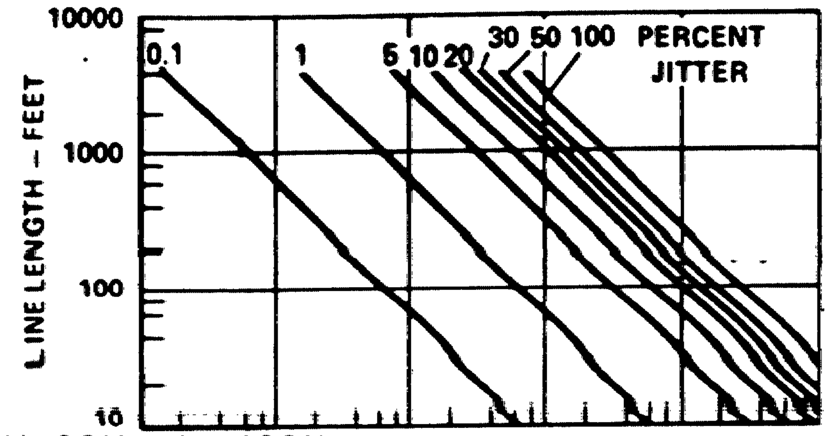
Figure E-29. Signal Quality as as Function of Line Length and Data Rate. [E-13]

If the effects of threshold variations are included in Figure E-29, the graphs shown in Figure E-30 result. The differential line voltage is assumed to be driven by voltages of $\pm V_{cc}$. Three conditions are shown: Figure E-30(a) shows the jitter which results when the threshold voltage (V_{th}) is given by $-0.1V_{cc} \leq V_{th} \leq 0.1V_{cc}$, Figure E-30(b) shows the jitter when $-0.2V_{cc} \leq V_{th} \leq 0.2V_{cc}$, and Figure E-30(c) shows the jitter when $-0.4V_{cc} \leq V_{th} \leq 0.4V_{cc}$. Figures E-30(a) through (c) correspond to 0.5, 1.0, and 2.0 volt threshold changes for the receivers tested. Comparison with Figure E-29 shows that the jitter increases due to threshold voltage variations.

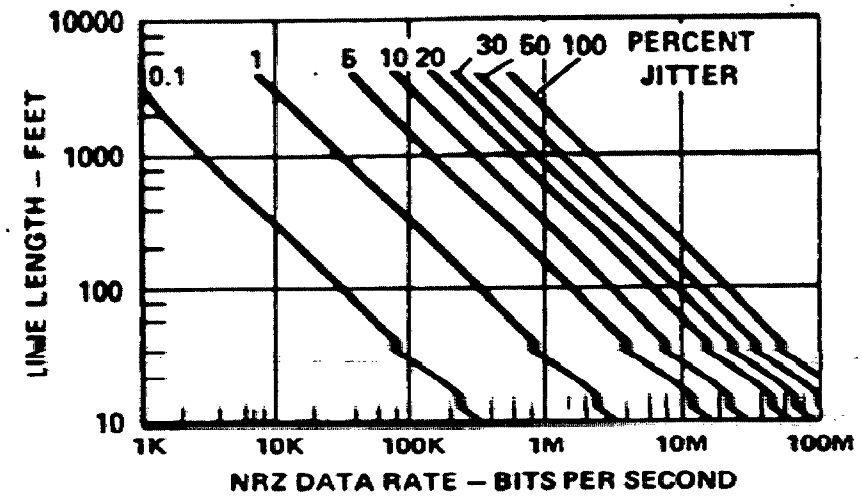
As an example of the use of these graphs, suppose that a designer must drive a 100-foot line, and desires a jitter of less than 5%. The maximum data rate is determined from Figure E-29 to be 12 MHz. If the maximum interfering signal expected to enter the system is 1 mW at 220 MHz, Figure E-27 shows that threshold variations of 0.5 volt may occur. Figure E-30(a), which applies to the 0.5 volt threshold case, shows that a data rate of 12 MHz will result in approximately 15% jitter, substantially higher than the desired 5%. However, by reducing the data rate to 6 MHz, the 5% jitter requirement can be satisfied even with the interfering signal present. This example clearly



a) $-0.1\ V_{cc} \leq V_{th} \leq 0.1\ V_{cc}$



b) $-0.2\ V_{cc} \leq V_{th} \leq 0.2\ V_{cc}$



c) $-0.4\ V_{cc} \leq V_{th} \leq 0.4\ V_{cc}$

Figure E-30. Signal Quality as a Function of Line Length and Data Rate Including Effects of Threshold Voltage Offsets. [E-13] illustrates that in high intensity EM environments, it may be necessary to reduce the data transmission rate of a system.

30.7 **BIPOLAR TRANSISTORS.** For relatively low absorbed power levels (typically below 10 mW), the response of an actively-operated bipolar transistor can be explained, in part, by rectification of the RF signal at a p-n junction. A complete explanation of the interference mechanism must include the effects of RF-induced current crowding [E-16]. Exposure to the RF signal causes the current distribution to be more heavily weighted toward the emitter edge (a low-gain region of the device). Thus, the DC gain falls off quite rapidly with increasing RF power. At higher values of absorbed power, the AC gain is also reduced, and some stages of the system may saturate or malfunction as the rectified current levels become comparable in magnitude to bias current levels. The rectified signal manifests itself as either a DC change in bias level or a video signal, each of which corresponds to the envelope of the RF signal. The sensitivity of a particular transistor to interference is influenced by the circuit configuration and operating conditions, the transistor structure, the speed of the device (gain-bandwidth product), and the frequency of the interfering signal.

RF injection into the base terminal, with rectification at the emitter-base junction, produces the greatest interference effect in bipolar transistors [E-3]. This particular interference configuration may be used for a worst-case analysis since the effects of injecting RF power into the emitter and collector terminals are similar, though less drastic. When an RF signal is applied at the base terminal, additional current must be supplied by the base bias source to maintain a constant collector current. It has been found that the base current which must be added is proportional to the absorbed RF power (square-law rectification) up to power levels on the order of 10 to 100 mW. This linear relationship has been found to hold at very low power levels in measurements extending down to a few nanowatts. At higher power levels, other effects become operative and the linearity is lost. Under these conditions, the emitter-base junction may become completely reverse-biased, seriously degrading both the DC and AC current gain [E-17]. Degradation or device failure typically occur at absorbed power levels on the order of 100 mW.

The observed response, for small absorbed power levels, may be expressed by the following equations:

$$\Delta I_B = I_B(P_a) - I_{BO} = \eta P_a \qquad \text{(E-15)}$$

$$I_c = \text{constant}$$

where:

η = rectification factor for microwave energy

$I_B(P_a)$ = base current of transistor when microwave energy is applied

I_{BO} = base current of transistor in absence of microwave signal

P_a = absorbed microwave power

I_c = collector current.

If the base bias current is held constant and the collector current allowed to vary,

$$\Delta I_c = \beta \eta P_a \tag{E-16}$$

where β is the low frequency AC current gain of the device. It has been suggested that bias conditions might possibly be adjusted to minimize interference effects (decrease η), but more work is required in this area [E-1].

As the frequency of the interfering signal increases, the magnitude of η generally decreases. This is illustrated in Figure E-31 where η (here shown as a function of incident rather than absorbed RF power) was measured for several stripline mounted 2N708's with the aid of a Hewlett-Packard model 11608A transistor fixture [E-18]. The rectification factor typically decreases at the rate of about 6 dB per octave (i.e., an f^{-1} relationship).

The sensitivity of several device types is listed in Table E-9 and represented graphically in Figure E-32, where the measured rectification factor for 2 GHz microwave energy is plotted versus the manufacturer's specified gain-bandwidth product (F_T). Each vertical bar represents the range of η values measured on a sample of 10 transistors of each 2N-type, and it is seen that there is often a large variation in η between otherwise "identical" devices. Despite the diversity in the devices, transistor types with high F_T values have, in general, a relatively large rectification factor and therefore are highly susceptible to RF interference.

The data also indicates that the structure of the device has an appreciable influence on η. As a class, low frequency germanium alloy transistors are the least susceptible devices, with some types showing no directly measurable ($\eta < 10^{-5}$ mA/mW) response to 2 GHz energy (although they do respond to VHF energy). Silicon planar transistors show a general trend of increasing sensitivity with increasing F_T, although when F_T approaches 2 GHz, the measurement frequency, it begins to decrease. Grown junction devices show a marked response to 2 GHz energy even though they have a rather low gain-bandwidth product. These devices have a large emitter-base overlap diode where the base wire contact is made to the base region of the transistor. This results in a very low β for that region of the device which the RF energy excites.

The use of the rectification factor in determining susceptibility can be demonstrated through a simple example. Suppose that the susceptibility is to be determined at 6 GHz (CW) for a preamplifier stage utilizing a transistor with a maximum η of .005 amps/watt at 2 GHz.

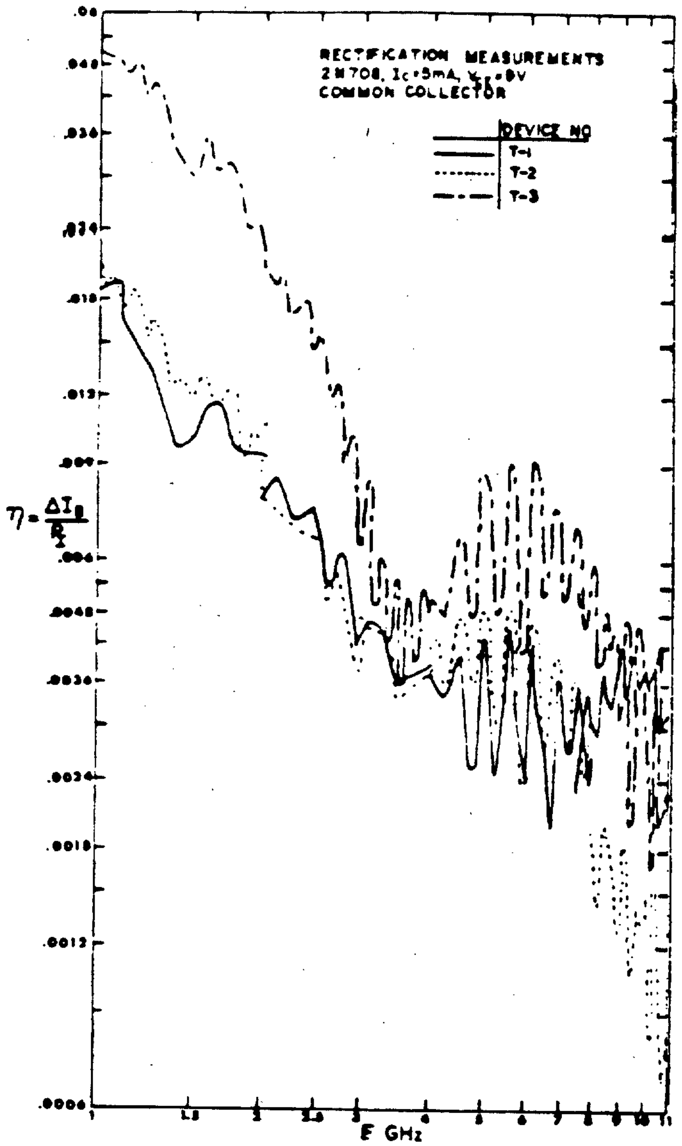


Figure E-31. Rectification factor (2 GHz) vs Manufacturers' Specified Gain Bandwidth Product. The Device Numbers Refer to the Device Types Listed in Table E-9. [E-18]

TABLE E-9

MEASURED RECTIFICATION FACTOR, F_T AND STRUCTURE OF VARIOUS TRANSISTOR TYPES [E-19]

$$\eta = \frac{\text{amps}}{\text{watt}}$$

#	MFG	DEVICE	MIN	AVG	MAX	F_T*	STRUCTURE
1	GE	2N337	.0217	.0460	.0817	20M	N-GD
2	GE	2N338	.0198	.0464	.0877	30M	N-GD
3	TI	2N388	0	0	0	15M	N-A
4	TI	2N395	0	.00025	.00067	4.5M	P-A
5	RCA	2N404A	0	0	0	4.0M	P-AΔ
6	TI	2N697	.0090	.0100	.0134	100M	N-DMEΔ
7	TI	2N705	.0011	.0034	.0059	300M	P-DME
8	M	2N706	.0247	.0389	.0588	400M	N-DME
9	NAT	2N708	.0200	.0500	.0800	480M	N-PL
10	M	2N834	.0169	.0267	.0372	500M	N-DME
11	F	2N914	.0157	.0240	.0443	480M	N
12	F	2N1132	.0025	.0048	.0073	60M	P-D
13	TI	2N1303	0	0	0	-	P∅
14	TI	JAN2N1304	0	0	0	5.0M	N-A
15	F	2N1613	.0050	.0082	.0118	60M	N-PLΔ
16	TI	2N1605	0	0	0	14M	N-A
17	TI	2N2000	0	.00019	.00040	2.0M	P-AΔ
18	TI	2N2192	.0016	.0027	.0043	50M	N
19	F	2N2219	.0059	.0210	.0361	250M	N
20	TI	2N2222	.0027	.0040	.0059	250M	N
21	TI	2N2369A	.0350	.0431	.0618	800M	N
22	F	2N2894	.0257	.0301	.0414	400M	P
23	F	2N2907	.0031	.0068	.0167	200M	P
24	M	2N3251A	.0124	.0145	.0163	300M	P-ϕ
25	F	2N4888	0	0	0	160M	P
26	M	2N5837	.0158	.326	.0443	1700MΔ	N-Δ

MFG

GE - General Electric
TI - Texas Instruments
RCA - RCA
M - Motorola
F- Fairchild
NAT - National

STRUCTURE

N - NPN
P - PNP
A - Alloy
D - Diffused
DM - Diffuses mesa
E - Epitaxial
G- Grown
PL - Planar
∅ - radiation resistant device
ϕ - noise figure 8 dB or below
Δ - switching, other uses

*Data taken from Transistor D.A.T.A. Book 32nd Edition.

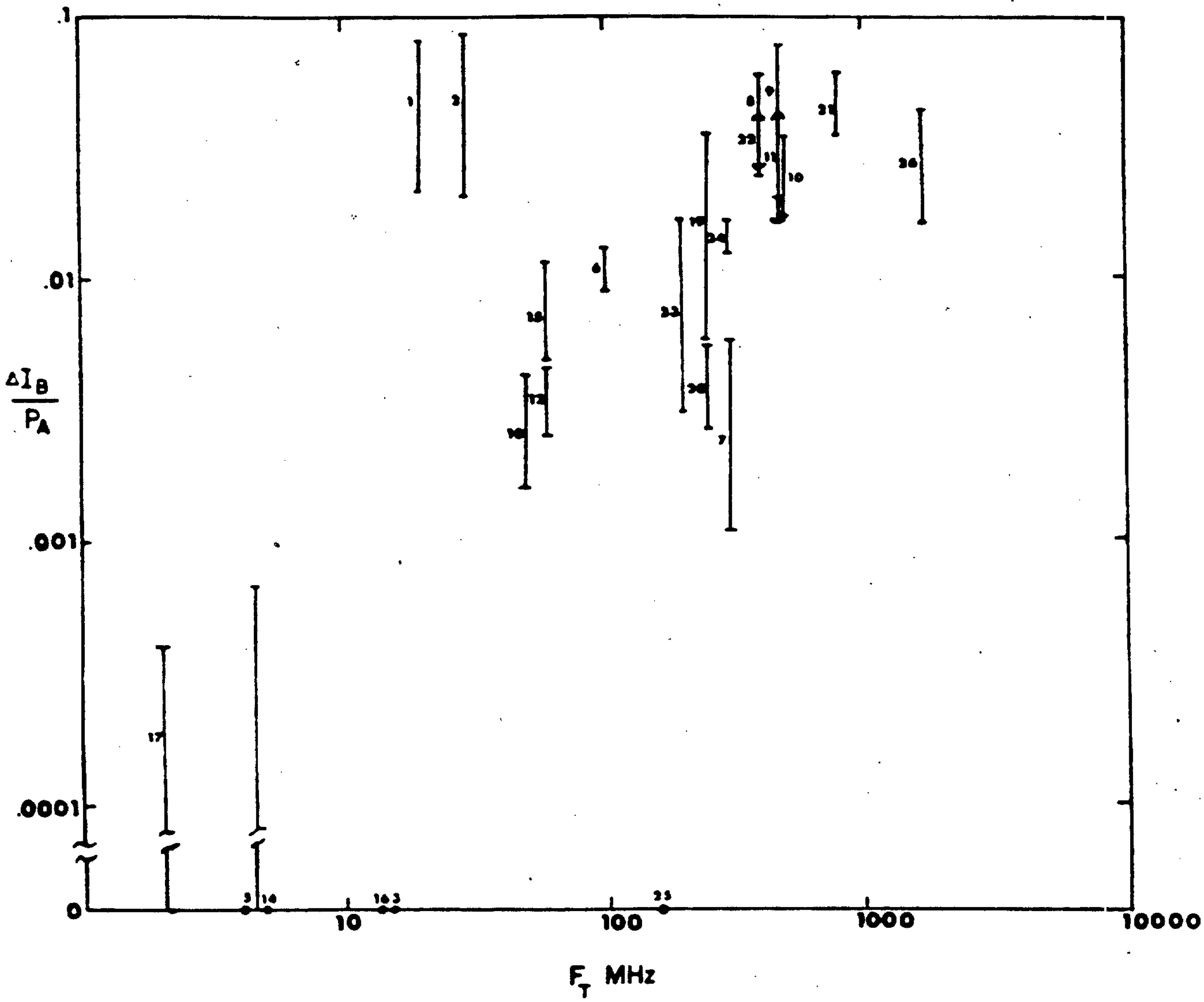


Figure E-32. Rectification Factor at 2 GHz vs Gain Bandwidth Product for Stripline-Packages 2N708 Transistors. [E-18]

(Additional factors such as the system passband capability for the modulation envelope may also be important, but are not considered in this simple example.) It is determined that a 0 dB signal-to-noise ratio (SNR) is the susceptibility threshold and a one microampere base (signal) current flows with no RF stimulus present. The rectification factor at f = 6 GHz is readily obtained (assuming a 6 dB/octave rolloff):

$$\eta \bigg|_{6\ \mathrm{GHz}} \approx .005(2/6) \approx .00167\ \mathrm{Amp/Watt}$$

The microwave power required to yield a 0 dB SNR would then be:

$$P_a = \frac{1 \times 10^{-6}\ \mathrm{Amp}}{.00167\ \mathrm{Amp/Watt}} = 6 \times 10^{-4}\ \mathrm{Watts} = 0.6\ \mathrm{mW}$$

The value obtained by such a procedure yields a worst-case susceptibility level. Suitable termination impedances (RF and video) could cause the current variations to be somewhat less. The "right" combination of impedances is quite fortuitous, however, and the result should not be viewed as overly pessimistic.

30.8 **MICROWAVE POINT-CONTACT DIODES.** Extensive studies have been conducted on microwave point-contact diodes. Within electronic systems, these devices are generally employed to intentionally receive electromagnetic signals coupled through an antenna system. Thus, these components are prime receptors of extraneous high power radiation.

Considerable testing has been performed on X-band diodes at 9.375 GHz to determine the percentage of diodes failing as a function of absorbed power for various pulse widths, pulse repetition frequencies, and numbers of applied pulses [E-17][E-19]. A change in noise figure greater than 10 dB was used for the failure criterion. Over 750 diodes were tested with each diode stressed only once to prevent cumulative effects from distorting the statistics. A total of eleven different parameter combinations were tested and these are shown in Table E-10. A minimum of 25 test samples were required to plot percent failed versus power absorbed. Typical data are given in Figures E-33 through E-42.

Tests were conducted using a repetition rate of 1 Hz, 450 Hz, 1 kHz, and 10 kHz for 10 pulse exposures at a 1 μsec pulse width. It can be seen from Figure E-43 that the failure level is essentially independent of the pulse repetition rate. This data would indicate a thermal relaxation time for the diode faster than 100 μsec, which is reasonable. In addition, the assumption may reasonably be made that the failure level remains essentially independent of the repetition rate for pulses of any width having 100 μsec or more between pulses.

Figures E-44 and E-45 are a compilation of the data for the 50% failure level. Figure E-44 is a plot of the failure level as a function of the number of pulses applied, for various pulse widths.

TABLE E-10

LIST OF RF PARAMETERS USED IN TESTS [E-19]

Pulse Width (μs)	Repetition Rate (Hz)	No. of Pulses
1	1K	10
1	1K	100
1	1K	2K
1	1K	10K
1	400	10
1	1	10
1	10K	10
3	1K	10
3	1K	100
10	1K	10
10	1K	100
100	1K	10
0.3	1K	10
0.3	1K	100
0.3	1K	3.5K

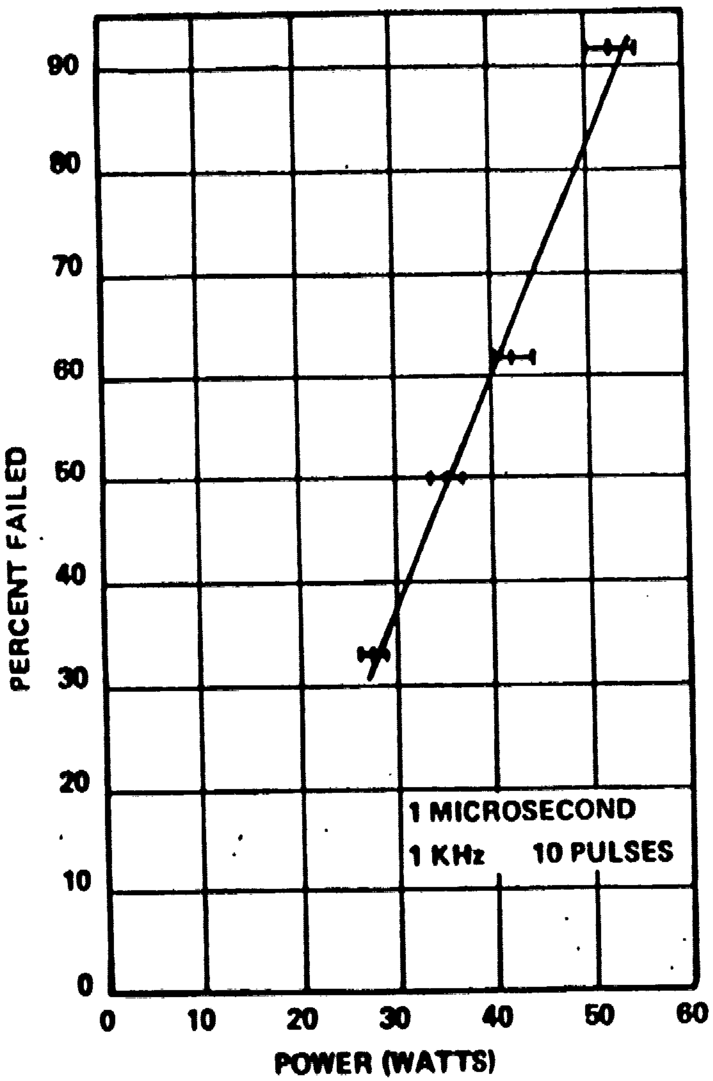
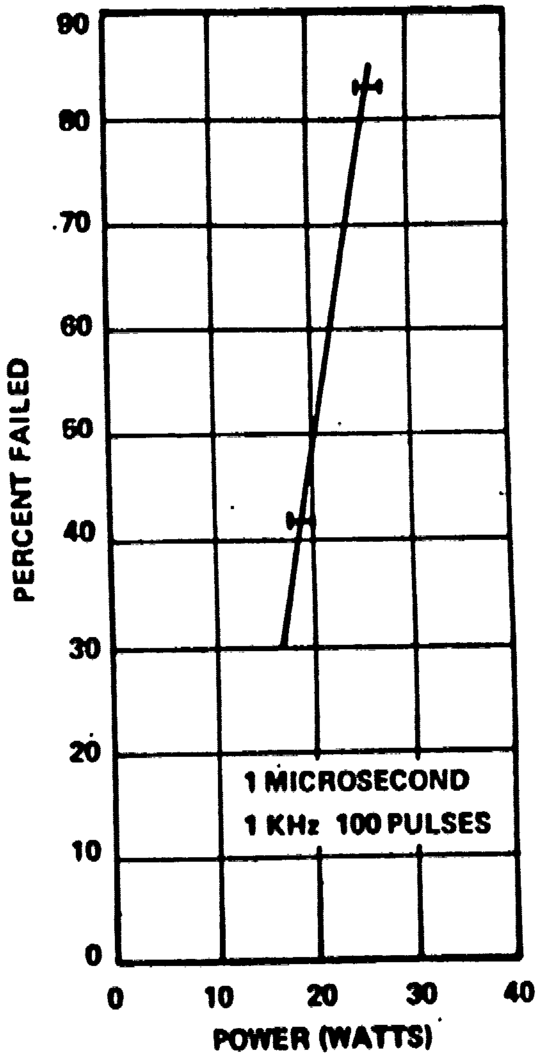


Figure E-33. Percent Failure Levels of 1N23 Diodes as a Function of Peak absorbed Power for Exposure Conditions: 1-μS Pulse Width, 1-kHz PRF, 10 Pulses. [E-19]



E-34. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 1 - μS Pulse Width, 1 - kHz PRF, 100 Pulses. [E-19]

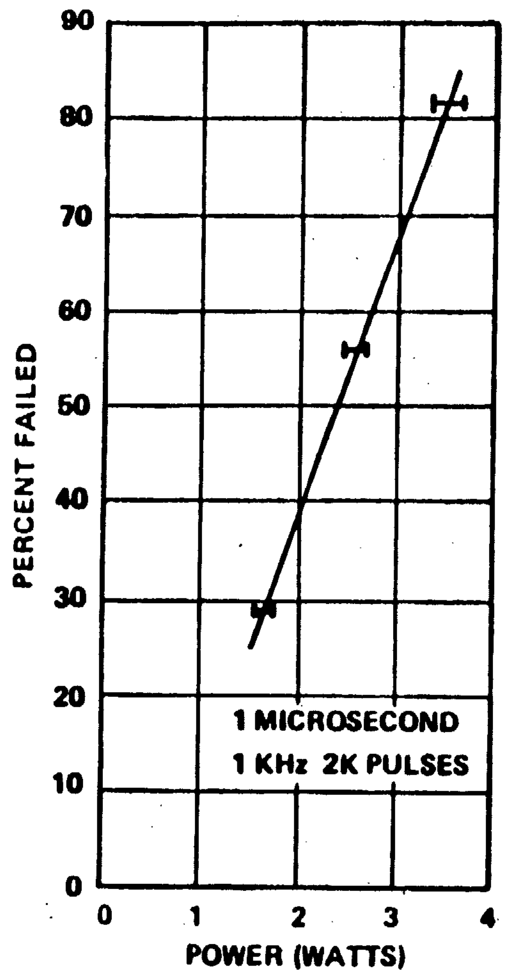


Figure E-35. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions:
1 - μS Pulse Width,
1 - kHz PRF, 2 K Pulses.
[E-19]

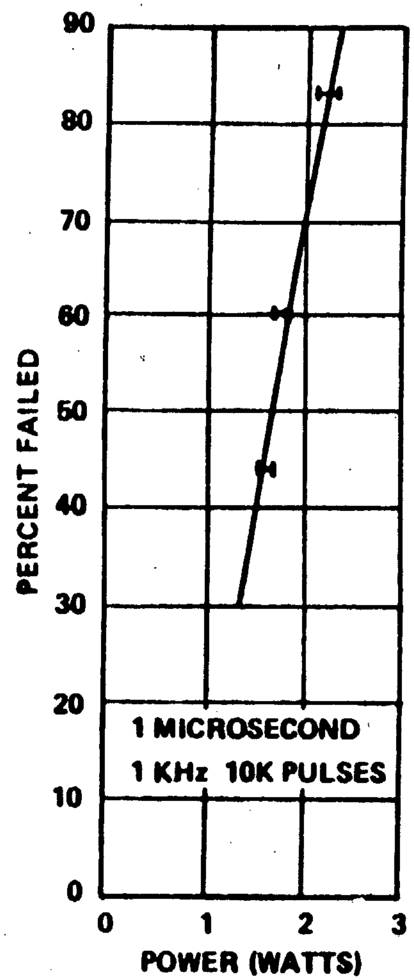


Figure E-36. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions:
1 - μS Pulse Width,
1 - kHz PRF, 10 K Pulses.
[E-19]

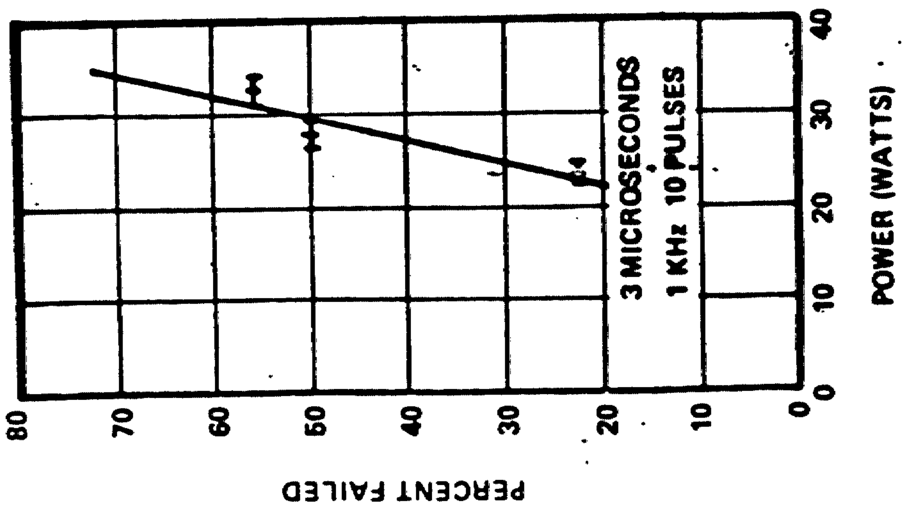


Figure E-37. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 3-μS Pulse Width, 1-kHz PRF, 10 Pulses. [E-19]

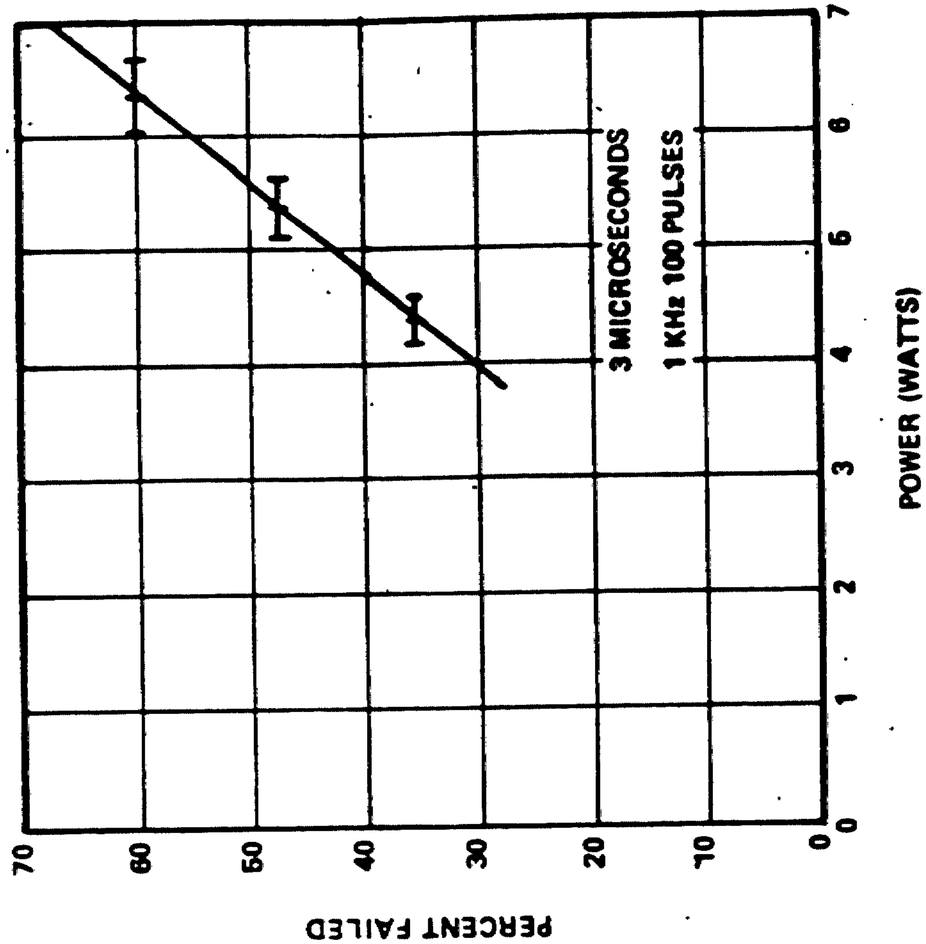


Figure E-38. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 3-μS Pulse Width, 1-kHz PRF, 100 Pulses. [E-19]

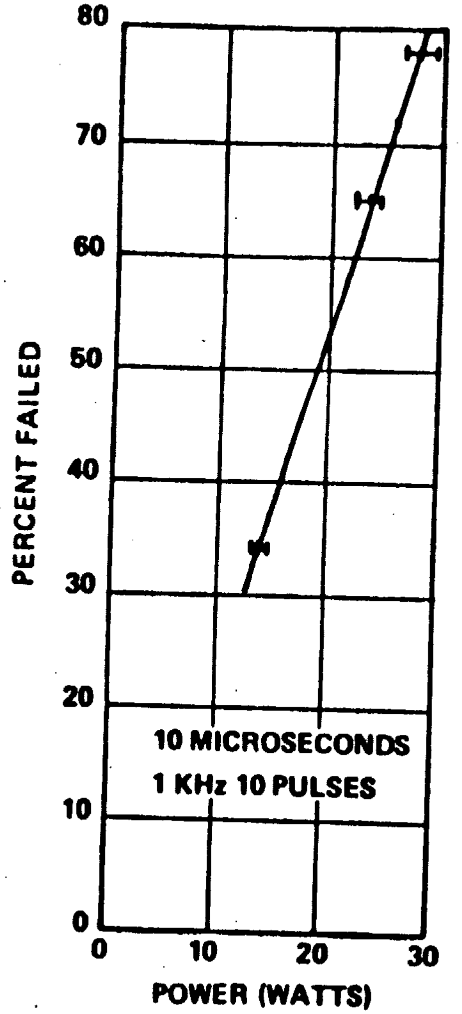


Figure E-39. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 10-μS Pulse Width, 1-kHz PRF, 10 Pulses. [E-19]

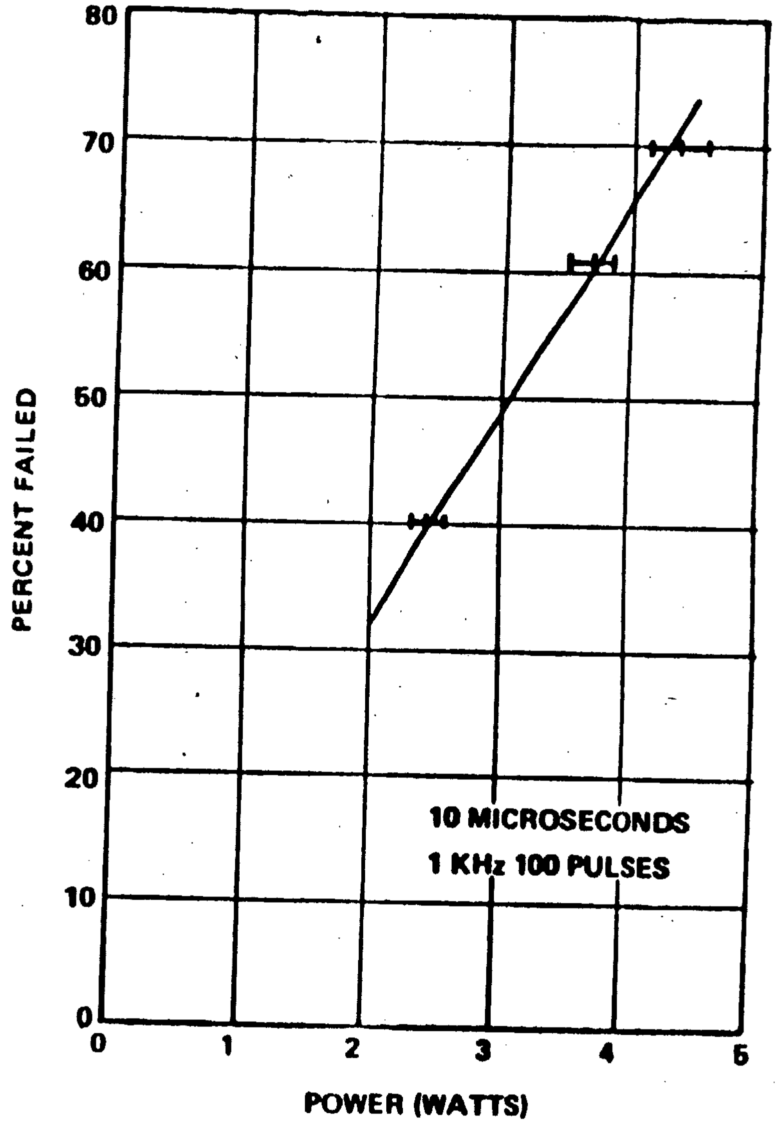


Figure E-40. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 10-μS Pulse Width, 1-kHz PRF, 100 Pulses. [E-19]

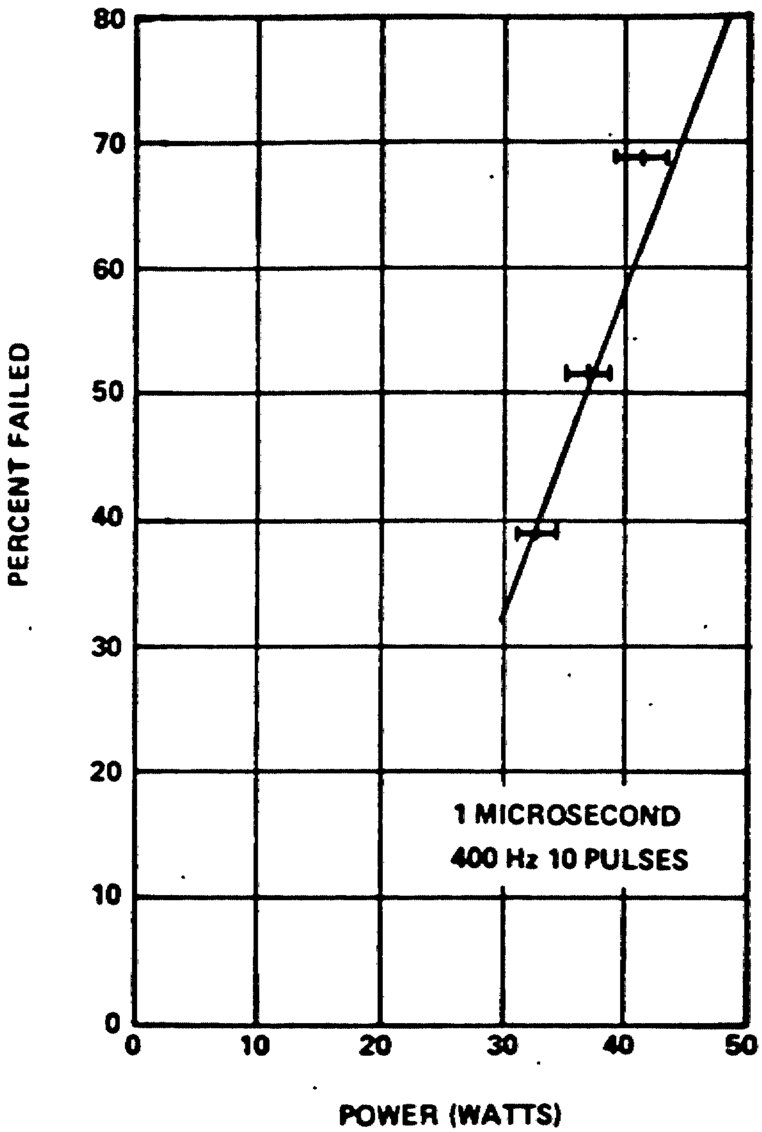


Figure E-41. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 1-μS Pulse Width, 400-Hz PRF, 10 Pulses. [E-19]

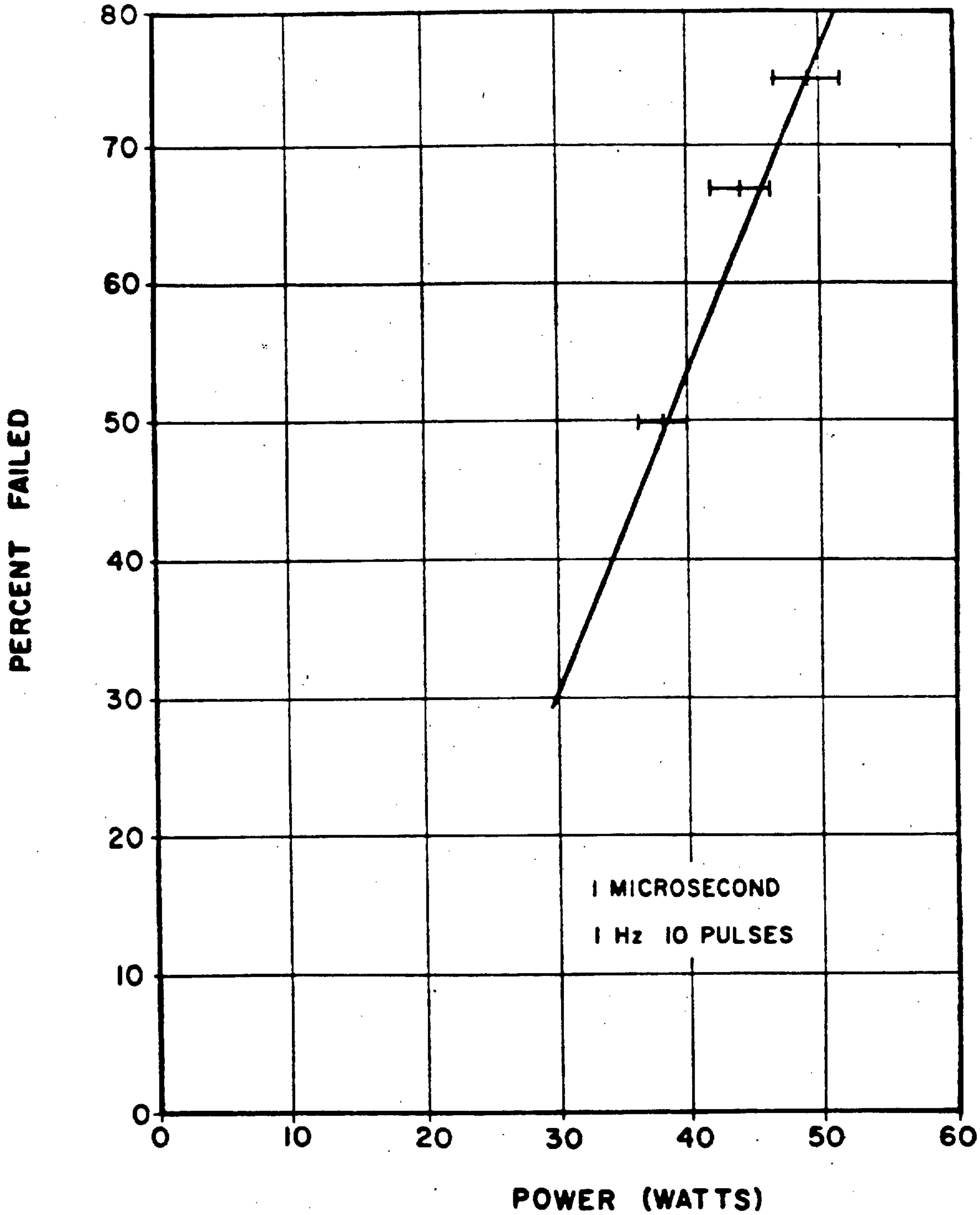


Figure E-42. Percent Failure Levels of 1N23 Diodes as a Function of Peak Absorbed Power for Exposure Conditions: 1-μS Pulse Width, 1-Hz, 10 Pulses. [E-19]

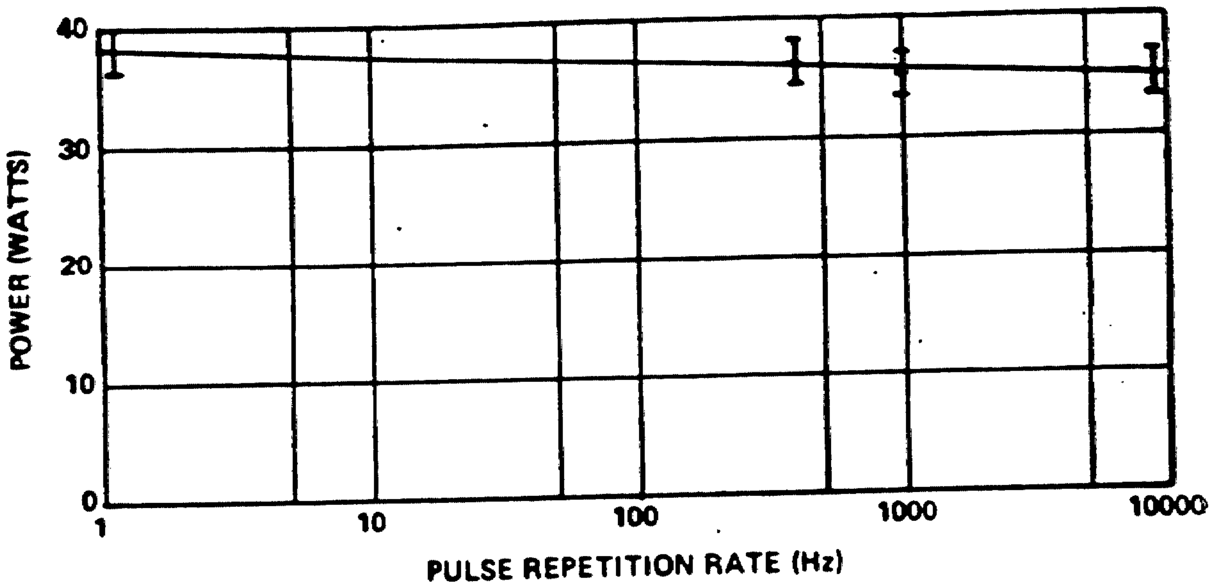


Figure E-43. Peak Absorbed Power Required for 1N23 Diode 50-Percent Failure for 1μS Pulse Width. [E-19]

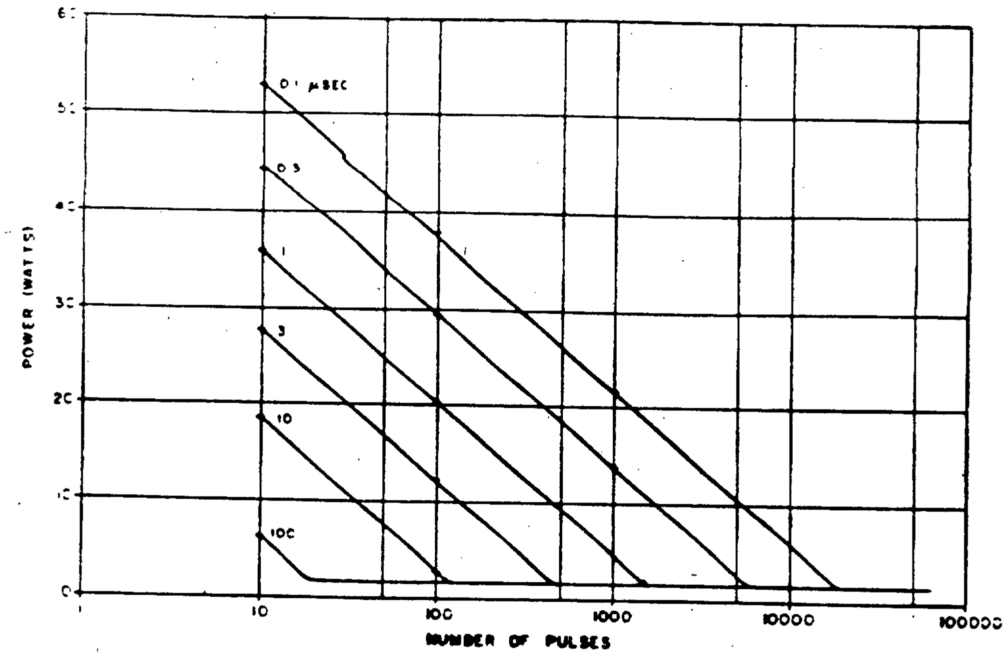


Figure E-44. Peak Absorbed Power for 1N23 Diode 50-Percent Failure Level vs Number of Applied Pulses for Various Pulse Widths. [E-19]

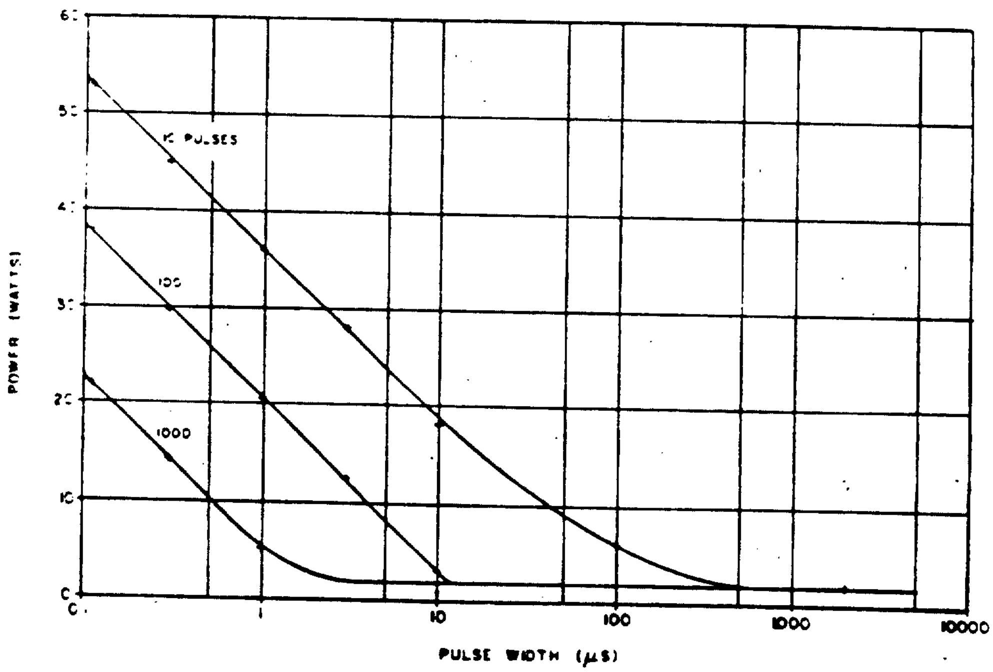


Figure E-45. Peak Absorbed Power for 1N23 Diode 50-Percent Failure Level vs Pulse Width for Various Numbers of Applied Pulses. [E-19]

All the curves on this graph approach the CW failure level for long exposures, as is expected. The number of pulses can be translated into exposure time through the repetition period for a given repetition rate. Figure E-45 is a plot of failure as a function of pulse width for various numbers of pulses (i.e.. exposure times). These also approach the CW level for long pulse widths. These two graphs give sufficient information to determine susceptibility, once the amount of absorbed power is known. The absorbed power can be determined from the incident power using Figure E-46. This graph shows the percent power reflected as a function of incident power. At 1 mW, the diodes are well matched, and will absorb essentially all the incident power. As the incident power is increased, the diodes become more and more mismatched and reflect more power. This mismatch saturates at about 50 watts where approximately 47% of the power is reflected.

An empirical expression was derived which predicts the 50% failure levels for various pulse widths and number of applied pulses. This was made possible by the equally sloped lines of Figure E-44 which. when plotted as a function of the product of pulse width and number of applied pulses, resulted in Figure E-47. A straight line drawn through the data points was determined as a least square fit to the data and is given by the empirical formula on the graph. For any given pulse repetition rate up to 10 kHz and pulse width greater than 0.1 μsec, the 50% failure level can be determined by the empirical expressions given in Figure E-47.

As a demonstration of the use of this data. assume one is concerned with the possible effects on the mixer diodes in a receiving system from a particular radar threat. The characteristics of the receiver are as follows:

Antenna gain: 19 dB
Operating frequency: X-band
Diode holder: Balanced mixer

As a possible radar threat, assume the following operational characteristics:

Power output: 800 kW
Pulse width: 0.8 sec
Pulse repetition rate: 1200 Hz
Frequency: 9600 MHz
Antenna gain: 39 dB

It is possible for the receiver to be exposed to the radar at a distance of 200 m for 85 msec. The equations needed to analyze this problem are as follows:

$$P_R = \frac{P_T G_T A_R}{4\pi R^2}(1-L)M \qquad \text{(E-17)}$$

$$P_A = P_R(1-F) \qquad \text{(E-18)}$$

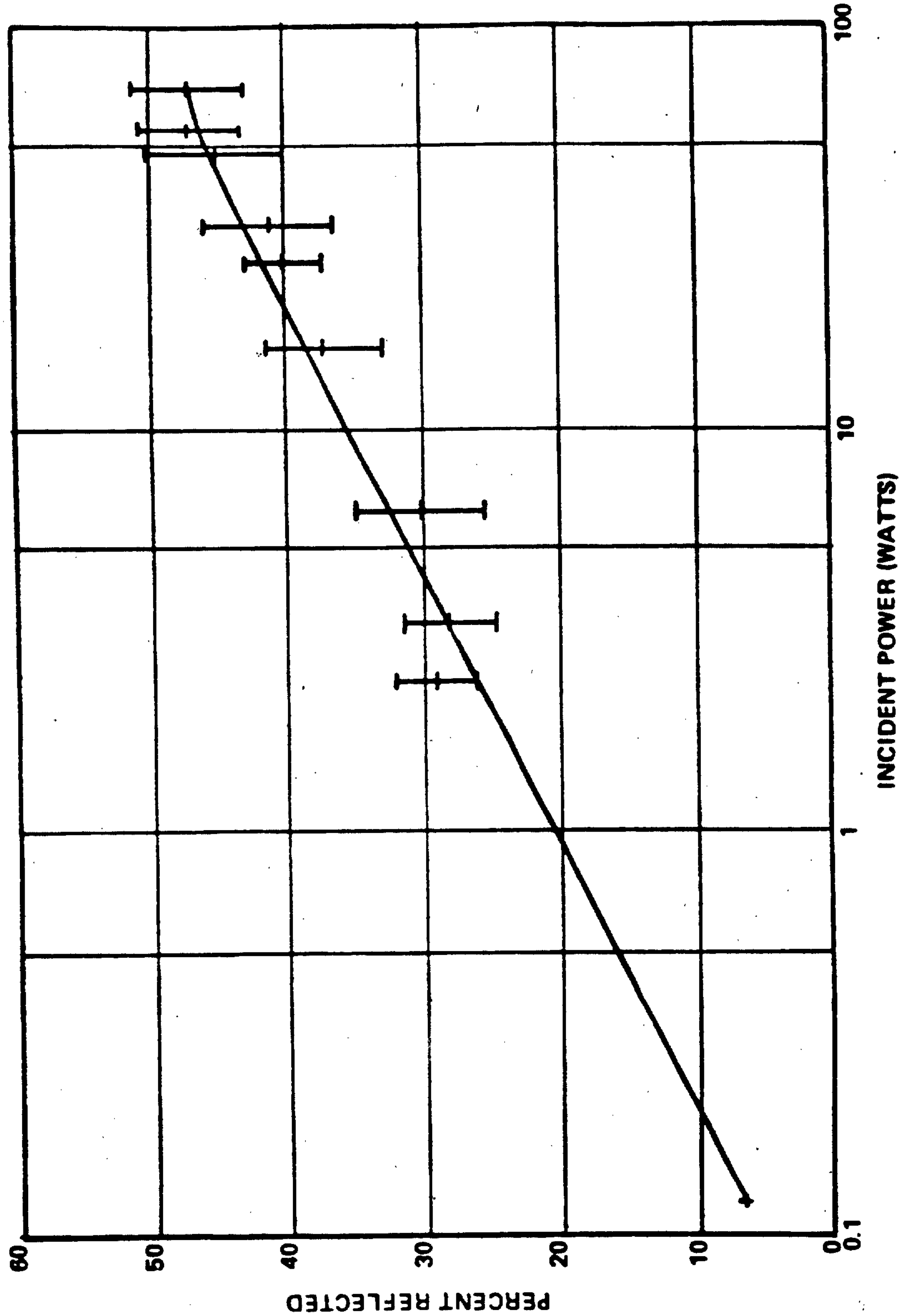


Figure E-46. Percent of Power Reflected vs Incident Power for 1N23 Diodes. [E-19]

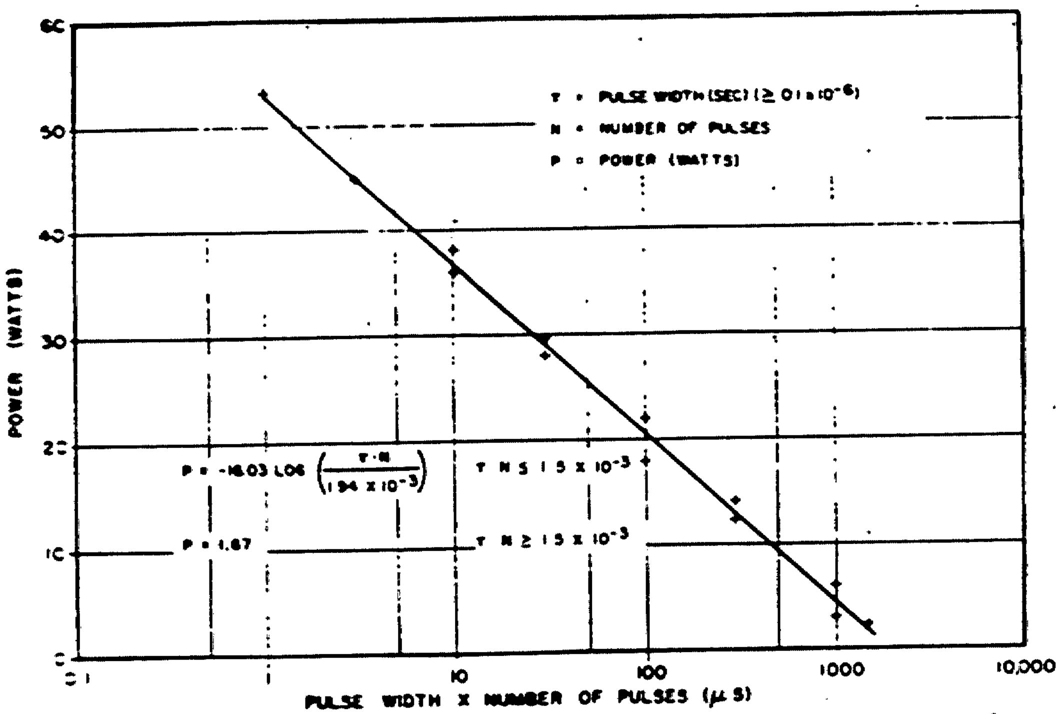


Figure E-47. Experimental 50-Percent Failure Levels of 1N23 Diodes as a Function of the Product of the Pulse Width and Number of Applied Pulses. (The line drawn is a best fit to the data points as given by the above empirical equation.) [E-19]

$$P_F = -16.03 \text{ Log} \left[\frac{\tau \cdot N}{(1.94 \times 10^{-3})} \right] \qquad \text{(E-19)}$$

where:

P_R = Power (peak) reaching diode, in watts
P_T = Power (peak) transmitted by threat radar, in watts
G_T = Threat radar antenna gain
A_R = Receiver antenna effective area = $\lambda^2/4 \pi G_R$, in square meters
G_R = Receiver antenna gain
L = Fractional losses between receiver antenna and mixer
M = A number determined by the mixer characteristics
R = Distance between threat radar and receiving antenna, in meters
P_A = Power (peak) absorbed by diode, in watts
F = Fractional power reflected by diode
P_F = 50-percent power failure level, in watts
= Pulse width, in seconds
N = Number of pulses.

As a worst-case, assume there are no losses between the receiving antenna and the mixer. Since the mixer is balanced, with each of the diodes receiving half of the power incident upon the mixer, M is 0.5. The power reaching the diode at a distance of 200 meters is calculated from Equation E-17 to be about 39 watts. Using Figure E-46 and Equation E-18, the power absorbed by the diode would be about 22 watts.

Under the assumed condition of an 85 msec exposure, 102 pulses would be received. Using Equation E-19, the failure level for these conditions is about 22 watts. The receiver vulnerability is marginal in this case, and some protection should be built into the receiver to ensure survivability in the anticipated environment.

40. **EXTENDING THE DATA BASE.** Due to the limited amount of available susceptibility data, alternate approaches must be taken in those instances where susceptibility information on a particular device or circuit of concern is not documented. The three major options are, in order of increasing costs and accuracy, data extrapolation, analytical modeling, and actual measurements. In some cases, the extrapolation of susceptibility data from one device type to another particular device is an appealing approach. However, a judgement must be made in which the reliability of the extrapolated value and the criticality of possible error relative to system hardening is considered. If, for example, this particular device is influential on the overall system susceptibility, extrapolation error could lead to exorbitant costs as a result of inadequate or excessive hardening. Analytical models are being developed which could prove useful by providing a more accurate estimate of susceptibility levels for circuits and devices. Susceptibility measurements represent the most accurate approach; however, cost considerations may outweigh this benefit in many situations. These three approaches are discussed in the following paragraphs.

40.1 **EXTRAPOLATION.** The extrapolation of susceptibility data is based upon the similarities between device design and susceptibility characteristics. Discretion is advised in that significant differences in susceptibility may exist even between devices of the same general type or class. However, there are two situations where data extrapolation appears to be a reasonable approach to susceptibility evaluation. These cases arise when data is required on: (1) a particular component for which susceptibility data is already documented on other components of the same general class, or (2) a component which, though of a different class, contains circuitry that parallels that of the components of known susceptibility. The first situation is strengthened by using the worst-case susceptibility curves of Section 30. The more devices that have been measured for each class, the "safer" will be the extrapolation. As an example of the second situation, consider the differences in the susceptibility levels found between 3-pin and multi-pin voltage regulators (Section 30.2). The explanation rested upon the fact that the output of the multi-pin regulator directly exposed the input terminals of an op amp, which was known to be very susceptible. In this manner, estimates of the susceptibility thresholds can be made for those devices on which no data is available.

40.2 **MODELING.** Although still in the relatively early stages of development, modeling is demonstrating promise as an effective means of predicting susceptibility levels. The data in Section 30 was obtained under well-controlled laboratory conditions, although worst-case conditions were estimated for conservatism [E-13]. The objective of modeling is to gain a greater understanding of the phenomena involved and to extend the results to devices and configurations not actually tested. Also, it can be used as an aid in obtaining estimates of susceptibility levels for a device operated under less than ideal conditions [E-20]. A complete listing of all the available modeling techniques will not be attempted at this time; rather, exemplary diode and transistor models will be presented along with a brief description of their role in circuit susceptibility prediction.

The approach taken for the modeling of interference effects in integrated circuits has been to develop models that account for interference in individual p-n junctions and transistors, using these as building blocks to construct models of complete circuits. Figure E-48 illustrates a diode model used to account for rectification effects [E-13]. Diode D1 models the diode with no RF stimulation and obeys the standard diode equation:

$$i_{D1} = I_{DS}\left(e^{\frac{qv_D}{kT}} - 1\right). \qquad \text{(E-20)}$$

where: i_{D1} is the current through diode D1.

v_D is the voltage across D1,

I_{DS} is the diode reverse saturation current,

q is the electron charge,

k is Boltzmann's constant, and

T is the junction temperature in degrees Kelvin.

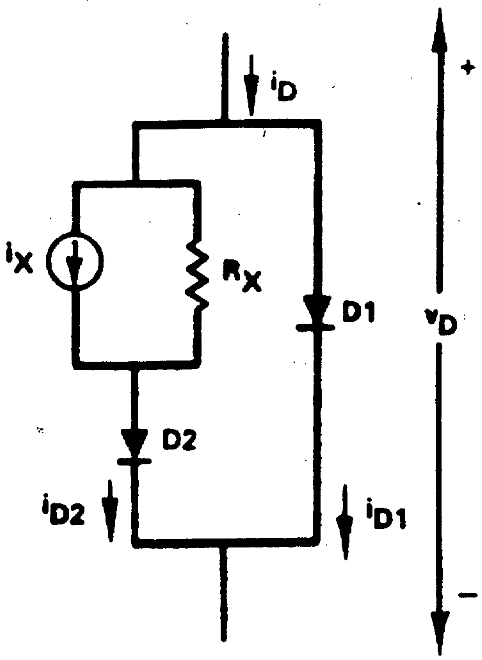


Figure E-48. **Circuit Model of Diode Under RF Influence.** [E-13]

The Norton equivalent comprised of i_x and R_x, and diode D2 model the video current and voltage offsets due to RF. For simplicity, diode D2 is assumed to have the same characteristics as D1. The value of current source i_x depends on the RF power level, frequency, and RF source impedance. For large RF signals (i.e., RF voltage comparable to, or greater than kT/q), i_x is proportional to the square root of the RF power level. That is,

$$i_x = K\sqrt{P_{RF}} \tag{E-21}$$

where K is a constant dependent on the frequency and source impedance of the interfering RF signal. The value of R_x also depends on the frequency and source impedance, but is independent of power level. In general, R_x increases with increasing frequency or increasing source impedance, while K decreases.

For modulated RF signals, the value of current source i_x varies with the envelope of the signal. The RF power level, P_{RF}, follows the instantaneous envelope of the RF signal, and the instantaneous value of i_x is given by Equation (E-21).

An analysis involving ideal diodes yields an estimate of the expected ranges of the parameters K and R_x. The RF source is represented by a Thevenin equivalent consisting of a voltage source $V_s \sin \omega t$ in series with an impedance $R_s + jY_s$. The diode junction is modeled by an ideal diode with a constant shunt capacitance C and series resistance, r_s. Computer-aided studies of the effects of the parameters R_s, r_s, and Y_s on the video model parameters K and R_x together with an analysis of max-min conditions show that the values of K and R_x (which always occur as ordered pairs) occupy a definite region in the K-R_x plane. For the absolute worst case of no extrinsic loss (i.e., $r_s = 0$), the region of the K_x-R_x plane in which possible values of K and R_x lie is shown in Figure E-49. The upper boundary is described by the relationship:

$$K_{max} = (8/R_x)^{1/2} \tag{E-22}$$

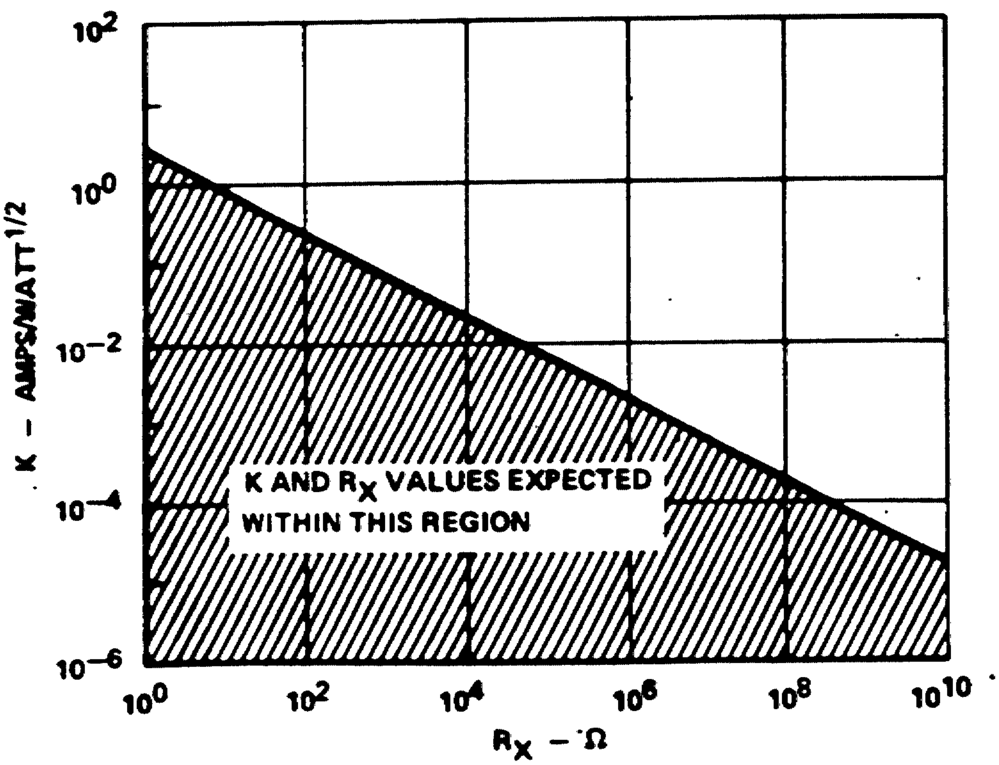


Figure E-49. Range of Parameters for Junction Model. [E-13]

The lossy element r_s provides a degrading effect on the rectification, and limits the maximum value of K which can occur. For a given value of r_s, the maximum K obtainable is:

$$K_{max} = (2/r_s)^{1/2} \qquad \text{(E-23)}$$

$$r_s \neq 0.$$

Likewise, there exists a minimum value of R_x (r_s) and a maximum value of R_x $[1/r_s(\omega C)^2]$.

Figure E-50 illustrates the implications that the various diode model parameter possibilities have on circuit modeling. Three cases are selected in Figures E-50(a) as possible values of K and R_x depending upon the values of the RF driving impedance. The value of K is chosen constant in this example, while the value of R_x differs in each of

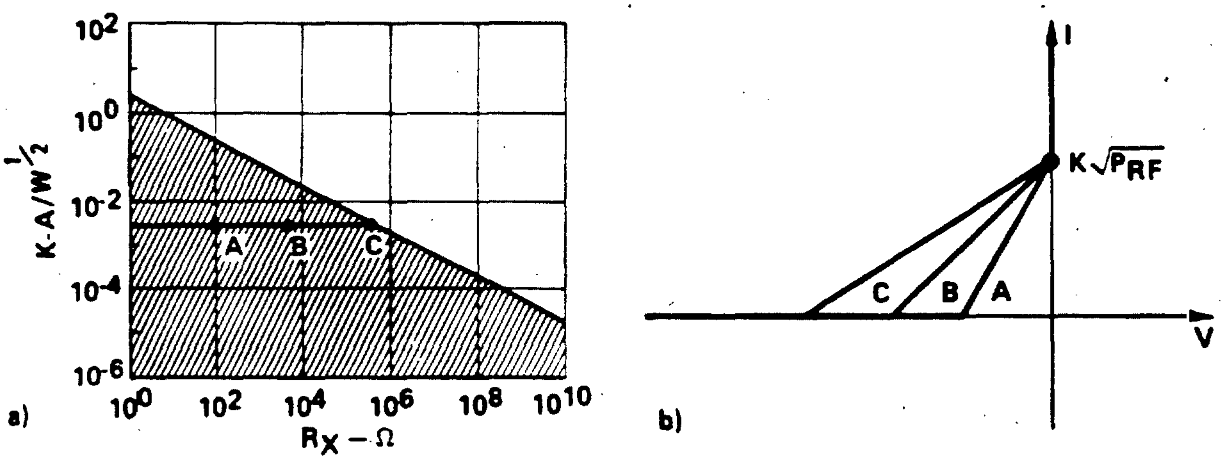


Figure E-50. **Illustration of Relationship Between Possible Choices of Rectification Parameters and Ideal Diode IV Characteristics** [E-13]

the three cases. Figure E-50(b) shows the piece-wise linear IV characteristics for the three values of K and R_x chosen. For worst-case circuit analysis, an iterative procedure of selecting possible K and R_x values followed by evaluation of circuit effects may be required.

RF effects in transistors may be accounted for by a modified version of the standard Ebers-Moll representation. The Ebers-Moll model is a widely used large-signal transistor model that includes the nonlinear effect of the transistor junctions. As such, it is accurate in all regions of operation: saturation, cutoff, forward active, and reverse active regions. The standard Ebers-Moll model was modified to include rectification effects by substituting the junction rectification model (Figure E-48) for each of the transistor junctions. The modified Ebers-Moll model is shown in Figure E-51.

It has been demonstrated that electronic circuit analysis programs such as SPICE (Simulation Program with Integrated Circuit Emphasis) can be used to predict the effects of RFI upon bipolar digital IC's such as 7400 NAND gates [E-21]. The modified Ebers-Moll model is substituted for the transistor in which RF is injected. The procedures for using this as an external model in the simulation program SPICE is described in Reference [E-21]. No change in an existing SPICE computer code is required. These procedures were applied to determine the EM susceptibility of a 7400 NAND gate with both inputs high when RF power was injected into its output (the most susceptible case). The simulations used standard SPICE models for all components in the 7400 NAND gate, except for the output transistor. The output transistor, into which RF was injected, was modeled using the modified Ebers-Moll model.

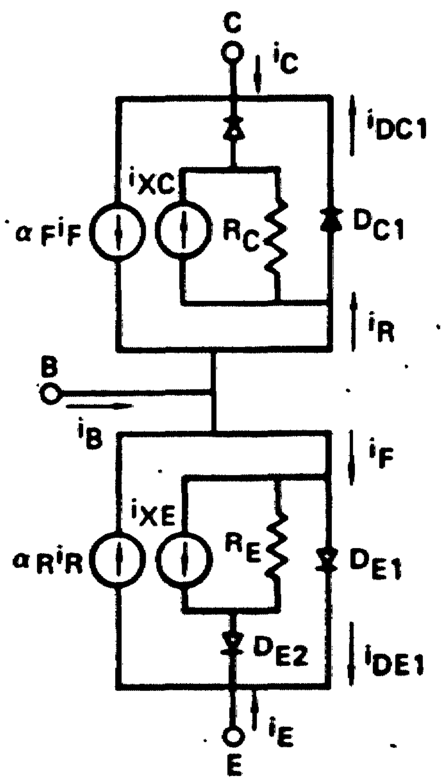


Figure E-51. Modified Ebers-Moll for a Transistor Under RF Influence. [E-13]

The EM susceptibility of three types of NAND gates was investigated. Experimentally determined values were used for the RF-induced parameters in the modified Ebers-Moll model. The SPICE simulations indicated that the low-power 74L00 series NAND gates are the most susceptible and that the high-speed 74H00 series are the least susceptible. Variations in fanout cause less than a 2 dB variation in the incident RF power required to cause various threshold levels to be exceeded. All three types of 7400 NAND gates will malfunction in most circuit applications at injected RF power levels in the 6 to 16-dBm range.

A worst-case analysis was also undertaken in which the impedance, RGEN, of the Thevenin equivalent RF source was varied in a systematic manner [E-21]. This procedure is especially useful when the RGEN value is not known apriori. Values for the RF power, PINC, in the range -4 to +4 dBm caused the three selected EM susceptibility threshold levels to be exceeded. For the special case RGEN = 50 , the simulation results and experiental results agreed within 4 dB. The predicted results are more conservative than the experimental results, which is desirable in a worst-case analysis. The close agreement for the 50 case increases the confidence with which these procedures can be extended to the more general case where RGEN is not 50 .

Other computer codes, such as NCAP (Nonlinear Circuit Analysis Program), have been developed and may prove useful in interference modeling. Examples of NCAP applications include an accurate prediction of RFI effects in an AF JFET preamplifier and a prediction of RFI effects in a bipolar linear IC broadband amplifier [E-22], although the accuracy appears to fall off at frequencies greater than $f_T/2$ [E-23]. For a more detailed discussion on these procedures, see the referenced material.

40.3 **MEASUREMENTS.** Susceptibility measurements are made by conductively coupling the RF energy to the device under test. Provisions are made for establishing realistic bias levels as well as monitoring the device's operating parameters. Device response is measured as a function of the absorbed RF power using predetermined changes in device operating parameters as susceptibility criteria. The validity of the technique rests upon the principle of the conservation of power.

40.3.1 **DISCRETE COMPONENT SUSCEPTIBILITY TESTS.** Susceptibility data on discrete components (diodes and transistors) is extremely limited and independent measurements can be a cost-effective means of obtaining the necessary data. The basic measurement technique for discrete components is shown schematicaly in Figure E-52. The arrangement shown is typical, but many variations are possible. The device being tested is mounted in a commercial 50-ohm test fixture which matches the characteristic impedance of the transmission line. Directional couplers are used to sample the incident, reflected, and transmitted microwave powers which are then measured with suitable instrumentation (e.g., power meters or crystal detectors). The couplers and the power measuring instruments are carefully calibrated. The device is biased through networks designed to decouple the RF from the bias sources and monitoring instrumentation. For non-critical continuous wave measurements, the bias network may consist simply of a resistor or RF choke. Commercial bias networks are available with limited passbands for biasing signals. Bias networks can be designed to combine video and microwave signals

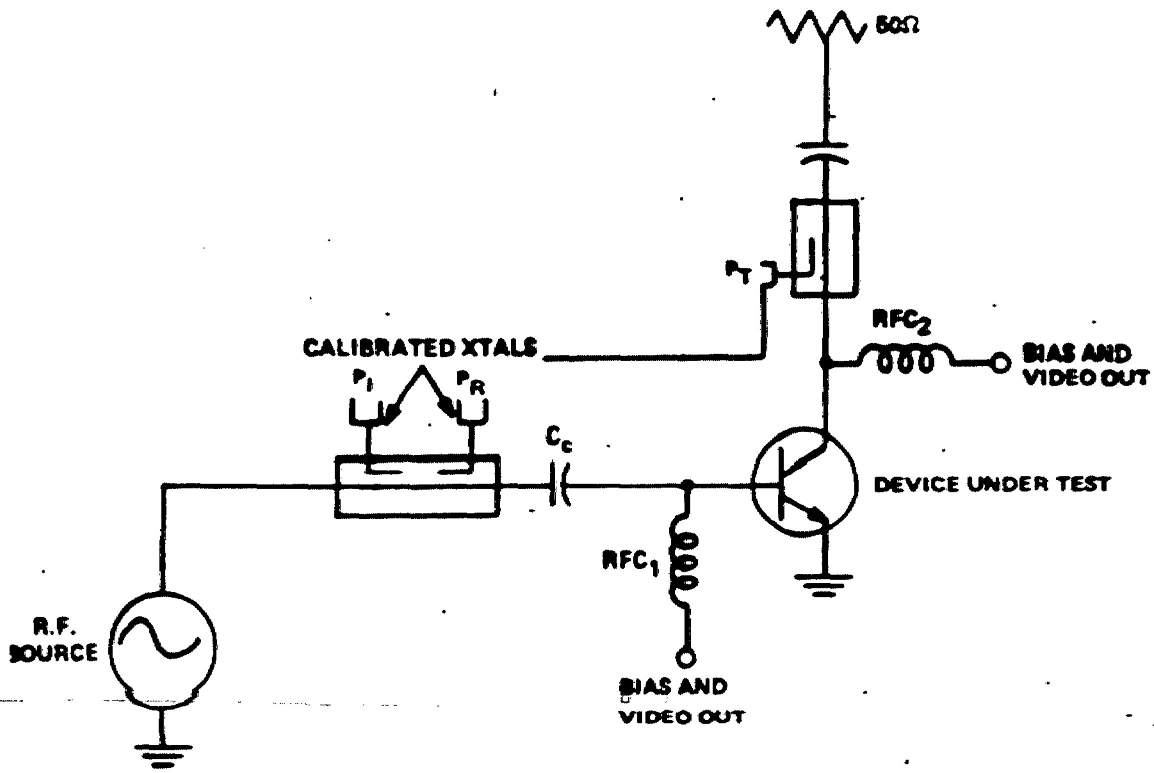


Figure E-52. Schematic Diagram of Basic Susceptibility Measurement Setup for Discrete Devices. [E-3]

while allowing independent adjustment with excellent isolation (Figure E-53). The microwave arm is essentially a high pass filter with a cutoff frequency near 100 MHz. The video arm offers excellent isolation from RF (low pass characterisic) while having a fast risetime capability [E-24].

The operation of the device under test is monitored through the low-frequency arm of the bias network while the interference stimulus is injected. Taking samples of the incident, reflected, and transmitted power, and using the calibration factors for couplers, bias networks, and detectors permit a determination of the power absorbed by the device.

40.3.2 **INTEGRATED CIRCUIT SUSCEPTIBILITY TESTS.** A relatively large amount of information and data on the susceptibility of electronic components have been devoted towards integrated circuits. The equipment required to obtain this data was very complex and extremely expensive.

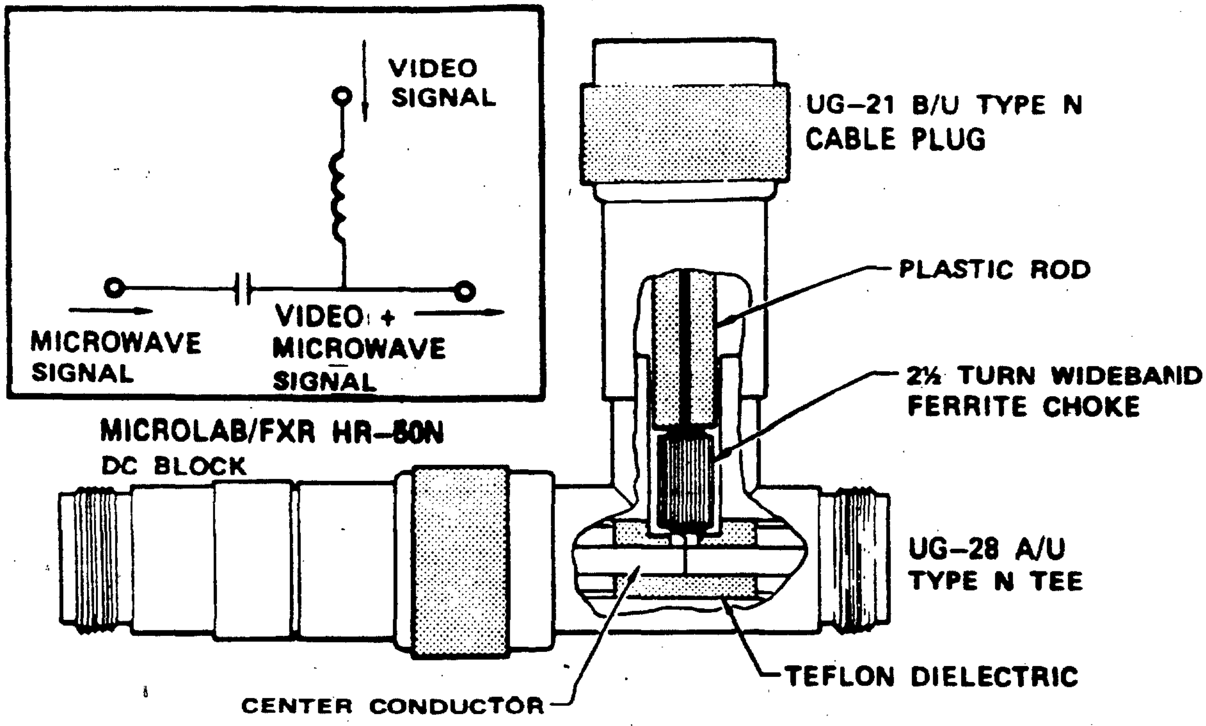


Figure E-53. Microwave Bias Unit. [E-13]

For example, the construction of several different test fixtures was required to accommodate each package style [E-13]. Also, for a typical IC under test, the number of pins involved necessitates a large quantity of instrumentation and biasing circuitry while simultaneously increasing the demand for test and measurement accuracy. These factors all contribute to substantial expense with the possibility of unreliable test results. Thus, it does not appear at the present time that such measurement activity is a cost-effective means of obtaining susceptibility data for integrated circuits.

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APPENDIX F

ESTABLISHING EMR HARDNESS CRITERIA

10. **INTRODUCTION.** The EMR hardness criteria for a system are established by the differences between the maximum EMR environment levels the system will be exposed to in the performance of its intended missions and the susceptibility levels of the system's most sensitive subsystems and circuitry. A comparison of the maximum EMR environment level and the susceptibility level of the most sensitive subsystem at any given frequency defines the minimum system EMR hardness required at that frequency. The accuracy with which the system EMR hardness requirements can be defined at any point in the life cycle of a system is dependent on how accurately the EMR environment levels and the subsystem susceptibility levels can be defined at that point in time.

20. **CONCEPTUAL PHASES.** In most cases, the EMR hardness criteria established in the early conceptual phases of a system will be rough approximations due to the limited input information available to accurately define the EMR environment levels and subsystem susceptibility levels. In particular, the definition of subsystem susceptibility levels will be difficult during these early stages of system development since neither the subsystem hardware nor the specifications will be available. It will probably be necessary to assume some typical susceptibility levels based on the information provided in Appendix E.

Even though the EMR hardness criteria are rough approximations, this information will be extremely useful in the early phases of a system development in defining the feasibility of approaches, estimating the magnitude of the EMR hardness program, identifying the risks and devlopment involved in satisfying the EMR hardness requirements, and establishing schedules and budgets for implementing an EMR hardness program.

30. **FULL-SCALE DEVELOPMENT PHASE.** It is extremely important that a complete, accurate EMR environment forecast be included in the RFP for the full-scale development phase. This forecast must convey to the bidders the EMR environment in which the system must operate. This forecast should also be used by the selected contractor to establish the system EMR hardness criteria. After the system designers have selected the subsystems and developed the specifications, the susceptibility levels of the most sensitive and most critical subsystems should be defined. A comparison of these susceptibility levels with the EMR environment forecast will yield the minimum system EMR hardness requirements. If alternate subsystem approaches are feasible, trade-off studies should be performed to determine the impact of the alternate approaches on system EMR hardness criteria. The hardness criteria should be updated as hardware is developed and susceptibility test data become available.

40. **SYSTEM SUSCEPTIBILITY/VULNERABILITY ASSESSMENT.** The system test organization has the responsibility for establishing the EMR hardness testing and evaluation criteria. A Type III EMR environment forecast should be used to define the frequencies, power levels, and modulation characteristics for the system susceptibility tests. This forecast should also be used to define the mission profiles for use in the vulnerability assessments.

APPENDIX G

EMR HARDNESS DESIGN PRACTICES

10. **INTRODUCTION.** To assure that a system is not adversely affected by its operational environment, it is imperative that EMR hardening considerations be an integral part of the system design cycle. Given a definition of the EMR environment, the system designer must determine the EM energy coupled to the system and the susceptibility of the system to this energy. If the EMR environment is a threat to system operation, the designer must harden the system by reducing the coupling of energy, reducing the system response to the coupled energy, or both.

This appendix documents design practices a designer may employ to harden a system. Section 20 presents a hardening methodology which defines the major tasks of a hardening design program, and Section 30 identifies hardening techniques and devices for minimizing EMR effects. Emphasis is given to hardening practices for systems containing solid state devices, for EMR effects which are caused by out-of-band frequencies, and for EMR environments covering a frequency range from 1 MHz to 100 GHz.

20. **HARDENING METHODOLOGY.**

20.1 **OVERALL HARDENING PROCESS.** A well organized EMR hardening design approach should be used by the system designer to ensure that the hardening of an air launched ordnance system is accomplished in a cost effective manner. Figure G-1 illustrates a methodology that the system designer may follow in the system hardening process. The inputs required from the program manager to initiate the process are definitions of the operational environment and the functional and tactical requirements of the system. Given these inputs, the system designer should formulate a system design concept and employ data, analyses, and measurements to determine if the design concept is susceptible to the specified environment. The susceptibility assessment will require a determination of environment-to-system coupling and system susceptibility to the coupled signals. Methods for determining system susceptibility and environment-to-system coupling are discussed in Appendices C and E.

Based on the results of the susceptibility assessment, the system designer should define the system hardening requirement and proceed with a hardening design which will preclude system susceptibility. The designer should utilize documented design data and techniques, analyses, and measurements as necessary during the design process to achieve the required hardening level. Specific steps should be taken to thoroughly document all aspects of the design, including the design approach, the hardening techniques and devices employed, and the analyses and measurements performed to substantiate the degree of hardening incorporated at the device, circuit, or subsystems levels. Sufficient information and data should be included to verify that the design approach will satisfy the overall system hardening requirement and preclude the susceptibility of the system to the specified environment.

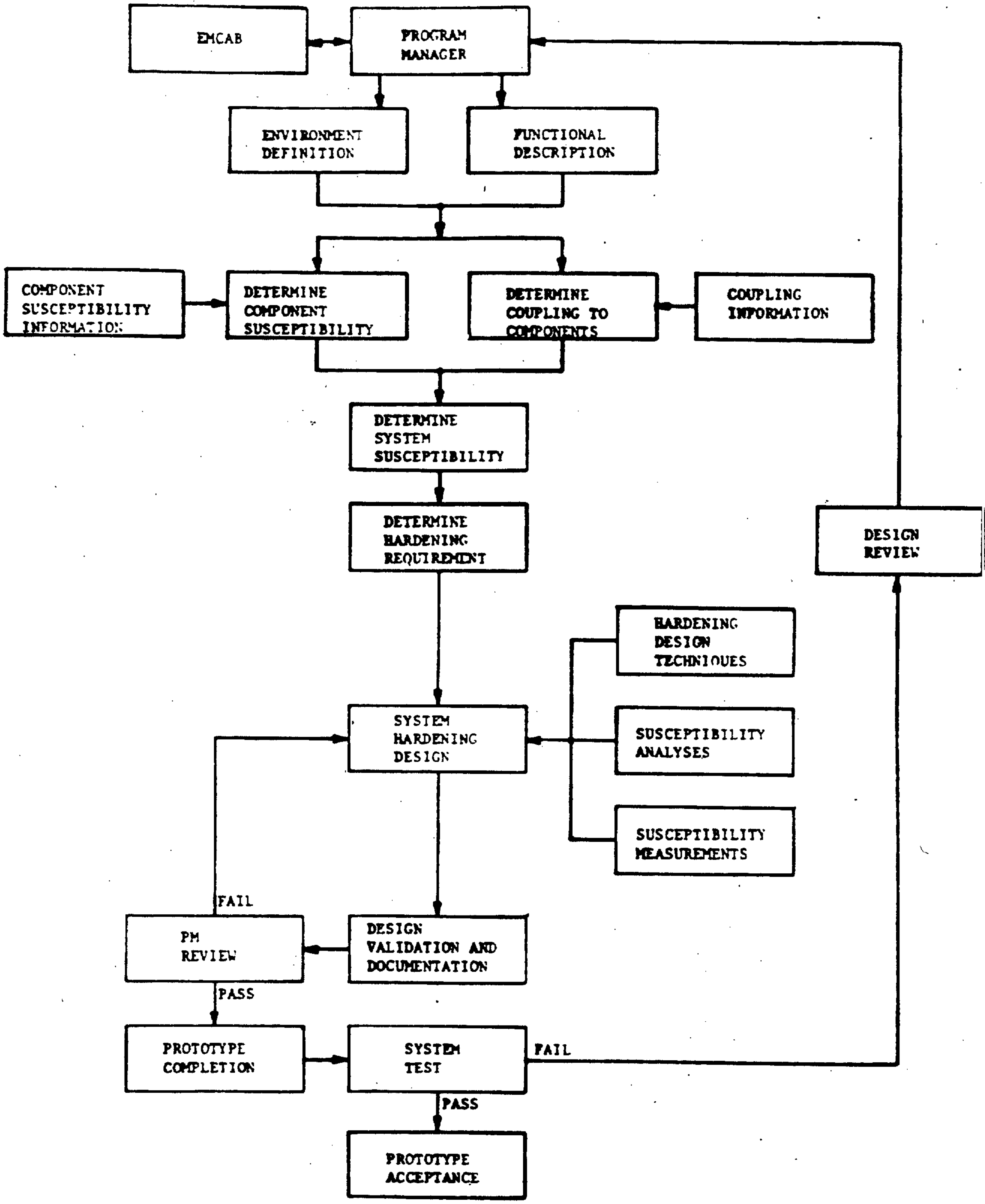


Figure G-1. Methodology of System Hardening.

The program manager's review of the documented hardening design may lead to a decision requiring additional system hardening or additional verification of the design. Approval of the hardening design would lead to system tests of a prototype system and prototype acceptance if the test results proved satisfactory. Unsatisfactory results could lead to a design review by the program manager and EMCAB and a repeat of the hardening design cycle. This review could also include tradeoff analyses to assess the cost effectiveness of alternate approaches.

20.2 SELECTING THE HARDENING APPROACH.

20.2.1 **LAYERED HARDENING CONCEPT.** When a system is exposed to an EMR environment, EM energy is coupled to internal system components by a series of paths beginning at the outer enclosure of the system. The EM energy first couples to the outer enclosure and sets up skin currents and charge densities. These currents and charge densities excite enclosure apertures, allowing the energy to penetrate to the interior of the enclosure where it can couple to internal wiring and cables which are connected to sensitive circuits or components. The concept of layered hardening is to interrupt these coupling paths via successive layers of hardening. The outer enclosure is first hardened to reduce energy penetration and thus reduce internal fields. Next, cables and wiring are hardened to reduce internal field-to-cable coupling. Finally, sensitive subsystems, circuits, and components are hardened.

Layered hardening is a cost effective approach to system hardening because it takes advantage of the intrinsic shielding of the system and subsystem enclosures. Also, the application of hardening in layers reduces the possibility of overly stringent hardening requirements on individual subsystems or circuits. As the isolation of each layer is improved, the level of hardening required for the remaining layers is reduced.

20.2.2 **ENCLOSURE HARDENING.** As stated above, the EM environment first couples to the system exterior to produce skin currents and charge densities. These currents and charge densities in turn excite points-of-entry (POE) to provide leakage of EM energy into the system. A first step in system hardening is to reduce the coupling through the POE's. The POE's are basically of two types: (a) antennas (deliberate and inadvertent); and (b) apertures. Apertures are inadvertent POE's in the sense that they are not intentional collectors of EM energy. Antennas can either be deliberate (real) or inadvertent (virtual). As an example, the detector in a missile is a deliberate antenna. A canard on a missile is an inadvertent blade antenna. One effective means of closing apertures in a system exterior is shielding. Exterior shielding can best be accomplished in the initial design phases of the system where required apertures can be kept to a minimum. For those apertures which are required for system operation and cannot be removed, measures must be taken to "close" points-of-entry of incident energy.

20.2.2.1 **SEEKER DOMES AND OTHER REQUIRED OPENINGS.** Often it is necessary to provide shielding over required openings such as seeker domes. The alternatives available are the use of screening material or conductive glass; the use of conductive gaskets around canard drive entry ports; and minimizing the size of openings.

(1) Use of screens over apertures has been employed for shielding purposes. A typical screen introduces a minimum of 15-20% optical loss. In some cases, screens can give good shielding at fairly low cost. Typical values may approach 20 dB at 10 GHz.

(2) Conductive gaskets, spring fingers, waveguide attenuators, screens and louvers are major devices and mechanisms available for maintaining enclosure shield effectiveness.

(3) Glass coatings using conducting materials such as gold or silver can provide shielding across viewing surfaces with some loss in light transmission. Conductive glass is commercially available from a number of glass manufacturers.

(4) 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.

20.2.2.2 **REDUCING COUPLING FROM ANTENNAS.** As stated above, system antennas from a protection point of view are of two types: deliberate and inadvertent. These two generic classes must be treated in different manners. A deliberate antenna is required for system operation and therefore cannot be removed. Filtering techniques must be used here to prevent undesired EM energy from being transmitted to the system interior. The coupling path from an inadvertent antenna, on the other hand, can be totally interrupted by the use of isolation techniques. For example, a missile canard could be connected to its mechanical drive mechanism by non-conductive devices instead of good conducting metallic rods which carry the EM energy to the system interior. As mentioned previously, the associated aperture for allowing connection to the canard must be kept as small as possible. Also, conductive gaskets are available for closing the annular apertures around the necessary penetration for the canard drive mechanism.

20.2.3 **CABLE HARDENING.** After the POE's in the system exterior are closed to the extent possible, the next layer of hardening is applied to interior cables and wiring. The coupling of energy to cables and wiring can be reduced by shielding and by the selection of cable types (fiber optics, twisted pair, etc.) which minimize pickup. Cable hardening can also be accomplished through the appropriate use of filtering to prevent coupled energy from being conducted into sensitive subsystems and components. Detailed design techniques for cable hardening are presented in Section 30.

20.2.4 **SUBSYSTEM HARDENING.** The final layer of hardening is applied at the subsystem level. Subsystem hardening can be accomplished through the application of numerous hardening techniques and devices. A properly filtered subsystem enclosure can be employed to prevent the coupling of energy to subsystem circuits and components. Circuits and components can be hardened through the selection of components or circuit designs which reduce the effects of injected EM energy. Subsystem designs can be employed which discriminate against arbitrary inputs or circumvent the effects of injected energy through appropriate logic circuitry. Detailed information on subsystem hardening techniques is given in Section 30.

20.3 **APPLICATION OF HARDENING DESIGN TECHNIQUES.** After the hardening approach has been defined, the designer should proceed to the detailed hardening design task. This task includes the selection of specific design techniques and applicable components (circuits, filters, gaskets, cables, connectors, etc.). In addition, the designer must specify how the design techniques and components are to be incorporated into the system to assure the integrity of the system.

20.4 **DESIGN TRADEOFFS.** Design tradeoffs may be necessary to ensure compatibility or resolve conflicts between the functional requirements and the hardening requirements of an air launched ordnance system. Also, tradeoffs may be necessary to achieve a cost-effective hardening design. Such tradeoffs should be directed to design techniques which will permit compatibility of functional and hardening requirements to be realized. Examples of such techniques are:

a. Selection of system operational signal levels as high as practical commensurate with device and circuit requirements.

b. Selection of interconnecting wiring and cabling techniques which provide the best rejection of normal mode and common mode energy transfer.

c. Use of fiber optic guides versus conventional shielded cable.

d. Use of rigid or flexible solid shielding versus single or double insulated metallic braid.

e. Multiple utilization of load bearing structures such as air frames, cable raceways, and conduit to satisfy both functional and hardening requirements at relatively low cost.

f. Use of enclosure shielding and cable filtering versus internal cable and circuit hardening.

Any tradeoffs which involve a change in functional, hardening, or cost objectives are subject to the review and approval of the program manager. The course of action for the resolution of conflicts between hardening requirements and other system requirements will depend upon such factors as:

a. The impact of the tradeoff on system susceptibility and system functional performance.

b. The number of equipments, subsystems, and systems involved.

c. The impact on program cost and schedule.

20.5 **HARDENING VERIFICATION.** Testing of a complete prototype system is necessary to validate a hardening design which incorporates a combination of hardening techniques. However, prior to full system tests, the system designer can verify, with reasonable assurance, that his hardening design is sufficient to assure the satisfactory operation of the system in its operational environment. The hardening design can be verified through the use of data, analysis, or measurements, or a combination of these three approaches. The most efficient approach to verifying the hardening design is to determine the effectiveness of the individual hardening techniques utilized in the overall design as the system design progresses. For example, the shielding effectiveness of the system or subsystem enclosures can be measured while the system circuitry is still under development, and any design changes required to realize the desired shielding can be incorporated into the enclosure with minimum effect on other design efforts. Also, component or circuit hardening can be accomplished independent of enclosure design efforts. Whatever approach is employed, a reliable estimate of the overall system hardness can be obtained simply by adding the level of hardening achieved at each system layer. If the estimate of system hardness determined in this manner exceeds the design requirement, the system designer can be reasonably confident that the system tests will be satisfactory. An estimate which falls below the design requirement is an indication that additional hardening is required.

Once the hardening design is verified, a complete documentation of the design should be submitted for the review and approval of the program manager and the EMCAB. The documentation should describe in detail the hardening techniques and devices employed, the approach used in verifying the design, and supporting data which substantiate that the design requirements have been met.

30. HARDENING TECHNIQUES AND DEVICES.

30.1 SHIELDING.

30.1.1 **INTRODUCTION.** Shielding is the most obvious method, and in most cases the most cost effective method, for protecting the circuits of a system from the EM enviroment. Shielding has two main purposes: (1) to prevent radiated EM energy from entering a specific region; and (2) to keep radiated EM energy confined within a specific region. Thus, shielding is essentially a decoupling mechanism used to reduce the radiated interactions beween systems or between portions of a given system. Here, the primary emphasis is on shielding the interior of a system from the external environment. This can be accomplished by minimizing the apertures in the system and subsystem enclosures to approximate a continuous metallic enclosure. The greatest shielding effectiveness is provided by a completely closed enclosure with no apertures. However, in practice this is seldom achieved since some apertures are usually necessary to meet system functional requirements. Thus, compromises in shielding effectiveness must be accepted and measures taken to minimize environmental coupling through those POE's which are required for system operation and cannot be removed.

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, the intrinsic characteristics of the shielding materials, and the number and shapes of shield discontinuities. The bibliography at the end of this section identifies sources for the theoretical aspects of shielding, as well as detailed engineering design techniques. The material which follows is primarily an overview of good design practices rather than detailed design information.

30.1.2 **SHIELDING MATERIALS.**

30.1.2.1 **SOLID MATERIALS.** Interference signal attenuation by a solid shield is due to two distinct effects: (1) reflection 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 (2) 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 Penetration or Absorption Loss. The combined loss due to these two effects is 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 where the absorption loss is greater than about 15 dB, the multiple reflection term can be ignored.

Using transmission line theory, the Shielding Effectiveness, S, of solid shielding materials is defined as:

$$S = A + R + B, \qquad \text{(G-1)}$$

where:

A = Penetration or Absorption Loss by the material, in dB,

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

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. The penetration (absorption) loss, which is essentially independent of wave impedance, is the same for electric and magnetic fields.

The aborption loss, single reflection loss, and multiple reflection correction terms can be described by relationships involving the shield thickness, material conductivity (g_r) and permeability (μ_r) relative to copper, and frequency. 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.

The single reflection loss term depends upon the degree of mismatch between the impedance of the air medium and the impedance of the shield. The impedance of the impinging wave is given by the ratio of its electric to magnetic field strength in space in the vicinity of the shield. The impedance of the shield is a complex function of the shield electrical parameters, shield thickness, and frequency of interest; in general, it is low for good conductivity shields and high for shields that have high permeability.

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

The multiple reflection correction term is a complex function of material, dimensional, and frequency parameters. As indicated previously, the multiple reflection correction term may be neglected if the absorption loss exceeds 15 dB. If the absorption loss is less than 15 dB, the correction must be determined.

30.1.2.2 **NON-SOLID MATERIALS.** Leakage through the 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:

$$S = A_a + R_a + B_a + K_1 + K_2 + K_3, \qquad \text{(G-2)}$$

where:

A_a = The Attenuation introduced by a particular discontinuity, in dB,

R_a = The Aperture Single Reflection Loss, in dB,

B_a = The 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, and

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

The first three terms in Equation G-2 generally correspond to the three terms of Equation G-1, while the last three terms encompass other factors that need not be considered for solid sheets. The number of discontinuities correction term, K_1, is employed when the source is located a large distance from the shield relative to the aperture spacing in the shield. This term can be ignored for sources close to the shield. The skin depth correction term, K_2, serves to reduce shielding effectiveness at low frequencies, when the skin depth becomes comparable to the screening wire diameter or dimension between apertures.

The adjacent hole coupling correction term, K_3, accounts for the fact 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. Non-solid material shielding analysis is most appropriate for single discontinuities or for identical and uniformly spaced apertures (such as screens or perforated sheets), but can be extended to somewhat more complex configurations as well.

30.1.3 SHIELDED ENCLOSURE DESIGN.

30.1.3.1 **GENERAL CONSIDERATIONS.** In the design of a shielded enclosure to provide a specified shielding effectiveness over a particular frequency range, the designer has a number of options available to him, and he is required to make a multitude of decisions concerned with the selection of shielding materials, fabrication techniques, cable and connector types, gasket types, etc. The information in the following sections is intended to aid the designer in making these decisions.

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.

Thin shielding utilizing coatings and thin-film shields have been employed in a variety of ways, ranging from metallized component packaging for protection against RF fields during shipping and storing, to conductive glass, to vacuum deposited shields for microelectronics applications. Since future integrated circuit packages may include thin solid shields, some comments regarding such shields are considered appropriate.

Thin shielding is loosely defined as shields whose thickness is less than 1/4 wavelength at the propagation velocity of the shield material. Solid 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 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 G-1 provides representative calculations of the shielding effectiveness of thin-film copper for different thicknesses and frequencies. One-quarter wavelength in copper is approximately 32500 A (3.25 x 10^{-4} centimeters) at 1 GHz, and it can be seen that shield effectiveness changes significantly above this thickness.

There are many applications in which the shield cannot be made of a solid material but must be made of a transparent or perforated material. Examples of these are covers for meters and gauges, which must be read through the shield, and apertures for ventilation or circuit adjustment. Woven materials such as wire mesh can be used over instruments and perforated materials or honeycomb panels can be used for ventilation or where circuit adjustment capabilities are required.

For perforated sheets, the fewer and smaller the holes the better the shielding effectiveness. For woven wire mesh, the larger the wire size and the greater the density, or wires per square inch, the better the shielding effectiveness. Table G-2 shows the measured shielding effectiveness of two common types of woven wire mesh for radiated fields.

TABLE G-1

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

Shield Thickness	1050 A		12500 A		21960 A		219600 A	
Frequency	1 MHz	1 GHz	1 MHz	1 GHz	1 MHz	1 GHz	1 MHz	1 GHz
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

TABLE G-2

WIRE MESH: RADIATED FIELD ATTENUATION VS. FREQUENCY

Frequency (MHz)	Copper		Galvanized Steel	
	18 x 18 (Wires/in^2)	22 x 22 (Wires/in^2)	22 x 22 (Wires/in^2)	26 x 26 (Wires/in^2)
	Attenuation (dB)		Attenuation (dB)	
0.01	103.6	109.1	137.7	143.9
0.03	104.7	110.2	135.4	141.6
0.06	105.4	110.2	132.1	138.3
0.1	105.4	113.6	129.1	135.3
0.3	105.0	110.5	120.8	127.0
0.6	103.4	108.9	115.1	121.3
1	101.3	106.8	110.8	117.0
3	94.5	100.0	101.4	107.6
6	89.3	94.8	95.4	101.6
10	85.1	90.6	91.0	97.2
30	75.8	81.3	81.4	87.6
60	69.9	75.4	75.4	81.6
100	65.6	71.0	71.0	77.2
300	55.9	61.4	61.4	67.6
600	49.9	55.4	55.4	61.6
1,000	45.5	51.0	51.0	57.2
3,000	35.9	41.4	41.4	47.6
6,000	29.9	35.4	35.4	41.6
10,000	25.5	31.0	31.0	37.2

The calculated shielding effectiveness of wire grids as a function of wire-diameter-to-wire-spacing ratio is shown in Figure G-2. The family of curves covers the wavelength-to-wire spacing ratio range from 10 to 10,000. The percentage transmission of IR and visible light as a function of the wire-diameter-to-wire-spacing is also shown in the figure.

Honeycomb panels are formed as a series of cylindrical, rectangular, or hexagonal tubular openings. Each individual opening acts as a waveguide-below-cutoff attenuator. The depth of the aperture determines the amount of attenuation realized, and the diameter of the individual openings determines the cutoff frequency. The shielding effectiveness of a hexagonal honeycomb with 0.317-cm openings and 1.27-cm long is shown in Table G-3.

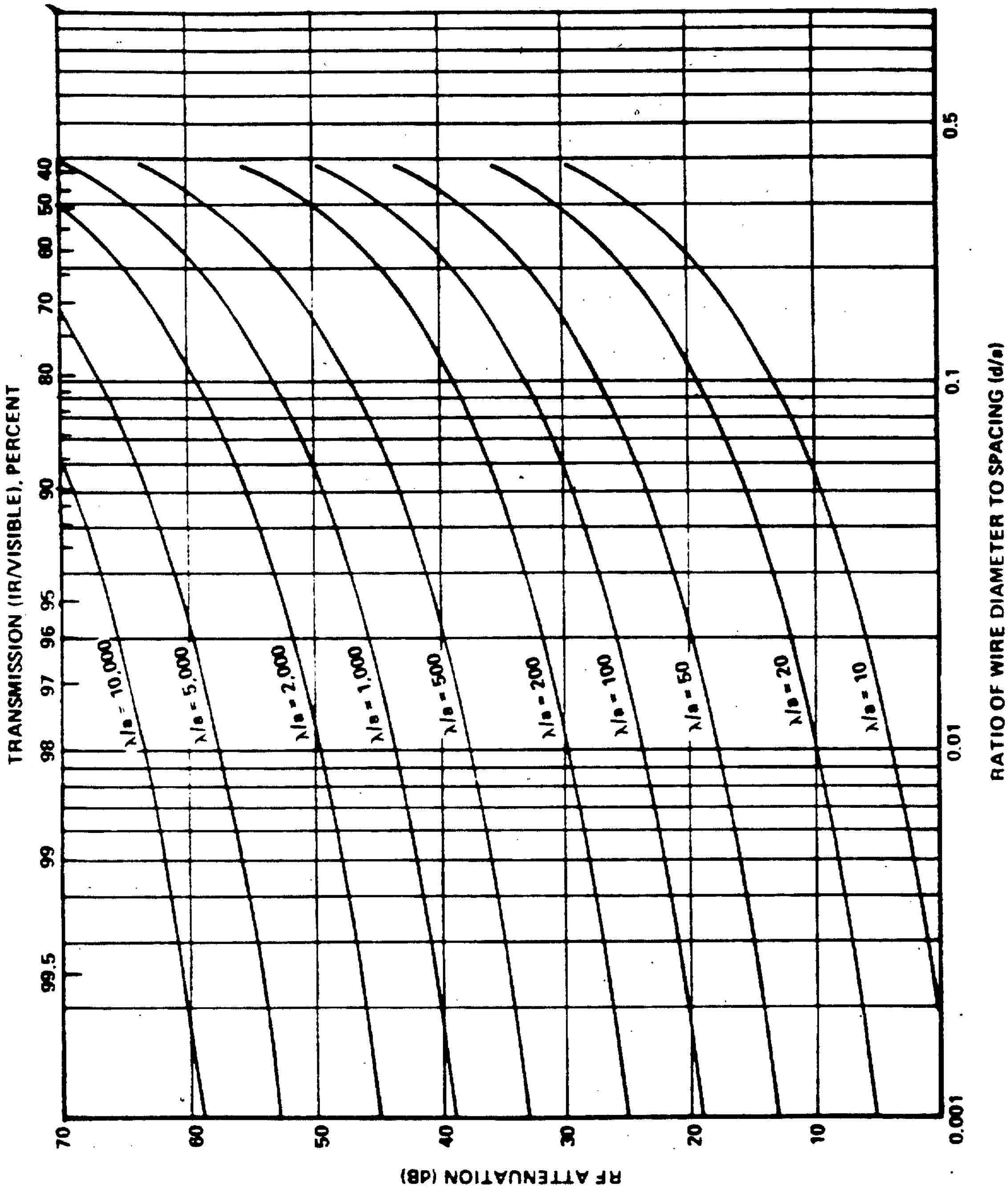


Figure G-[illegible]. Attenuation of Wire Grids as a Function of Wire Diameter-to-Wire Spacing Ratio.

TABLE G-3

SHIELDING EFFECTIVENESS OF HEXAGONAL HONEYCOMB MADE OF STEEL, WITH 0.317-CM OPENINGS, 1.27-CM LONG

Frequency	Shielding Effectiveness
100 kHz	45 dB
50 MHz	51 dB
100 MHz	57 dB
400 MHz	56 dB
2200 MHz	47 dB

30.1.3.2 **DESIGN PRACTICES.** An ideal shielded enclosure would be one of seamless construction with no openings or discontinuities. However, any practical enclosure will have seams to facilitate installing the system circuitry in the enclosure and will have apertures and penetrations for ventilation and entry of wire, cable, controls, meter faces, etc. Since each of these discontinuities represents a potential degradation of the shielding effectiveness of the enclosure, the selection and implementation of techniques to provide shielding integrity at these interfaces are important factors in the design of the shielded enclosure. Figure G-3 illustrates some of these interfaces and proper and improper techniques to maintain the shielding integrity at the apertures and penetrations.

In the design and fabrication of the basic shielded enclosure shell, the number of seams and openings should be kept to a minimum. Where possible, all permanent seams should be welded, brazed, or soldered such that the joint is continuous. However, satisfacory results may be obtained utilizing closely spaced rivets or spot welding if sufficient care is taken in the preparation of the mating surfaces and the installation of the fastening devices. To ensure satisfactory bonding at the seams of the enclosure, the bonding design practices recommended in Section 30.3 should be observed.

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 beween shielding surfaces to be joined. Clean conductive mating surfaces and a good pressure contact are necessary.

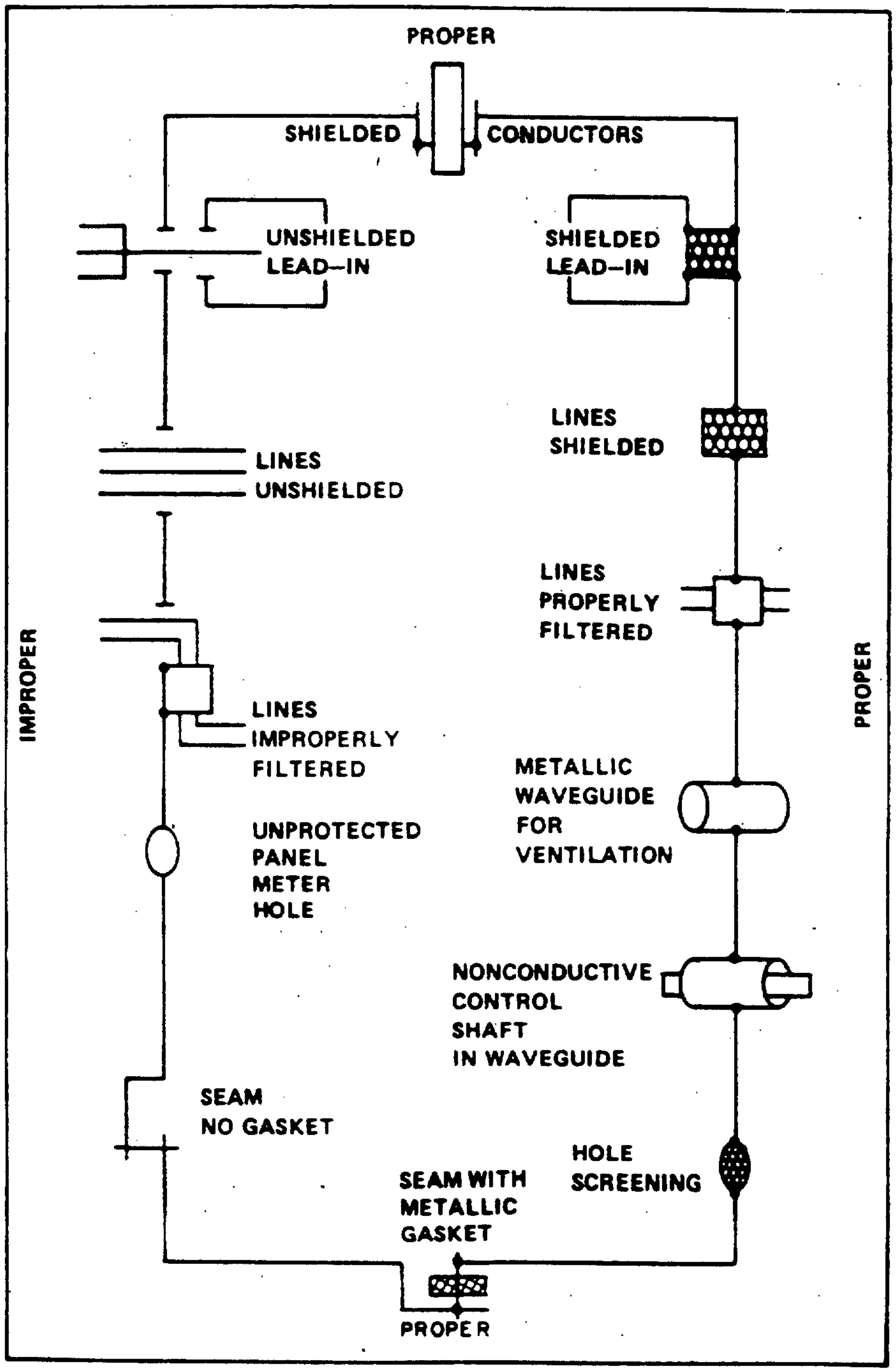


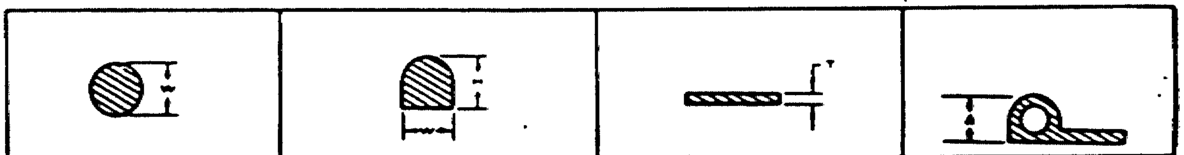
Figure G-3. Typical Shielding Enclosure Discontinuities with Proper and Improper Controls. [G-6]

The major material requirements for RF gaskets include compatibility with the mating surfaces, corrosion resistance, appropriate electrical propeties, 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 used only for gasketing between two aluminum surfaces. Beryllium-copper contact fingers are usually employed, with a variety of platings available. Mumetal and Permalloy are commonly 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 advantage and limitations of these, as well as non-sealing RF gaskets, are summarized in Table G-4.

Silver-filled silicone 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 G-4 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.

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.



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 – .008	.093		
		.016 – .032	.175				

Figure G-4. Gasket Deflection Limits (in inches). [G-1]

TABLE G-4

CHARACTERISTICS OF CONDUCTIVE GASKETING MATERIALS

Material	Chief Advantages	Chief Limitations
Compressed knitted wire	Most resilient all-metal gask-ket (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-through 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. Reusable. 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.

Metal mesh 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 nonconductive 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 G-5.

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.

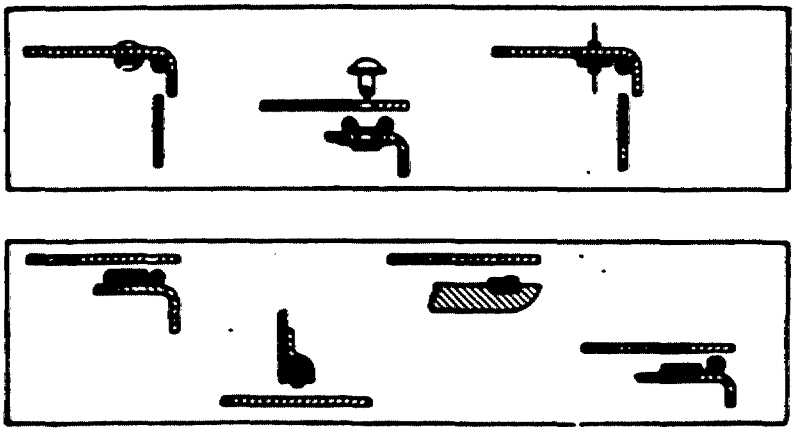


Figure G-5. Representative Applications of Wire Mesh Gaskets. [G-1]

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.

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 long as the shaft within the guide is nonconducting. 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.

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

(a) Use of a waveguide attenuator,

(b) Use of screening material,

(c) Providing a shield behind the assembly of concern, and filtering all leads to the assembly, and

(d) Use of conducting glass.

A waveguide attenuator is a practical approach for RF shielding of lamps. 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 purposes. 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, bothersome zoning patterns can result. However, these potential deficiencies are counterbalanced by good shielding efficiencies at a fairly low cost.

Figure G-6 illustrates one method of mounting such screens when they are not incorporated directly into the device to be shielded. The screen may be imbedded 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 made to reduce the meter-reading and zoning problems.

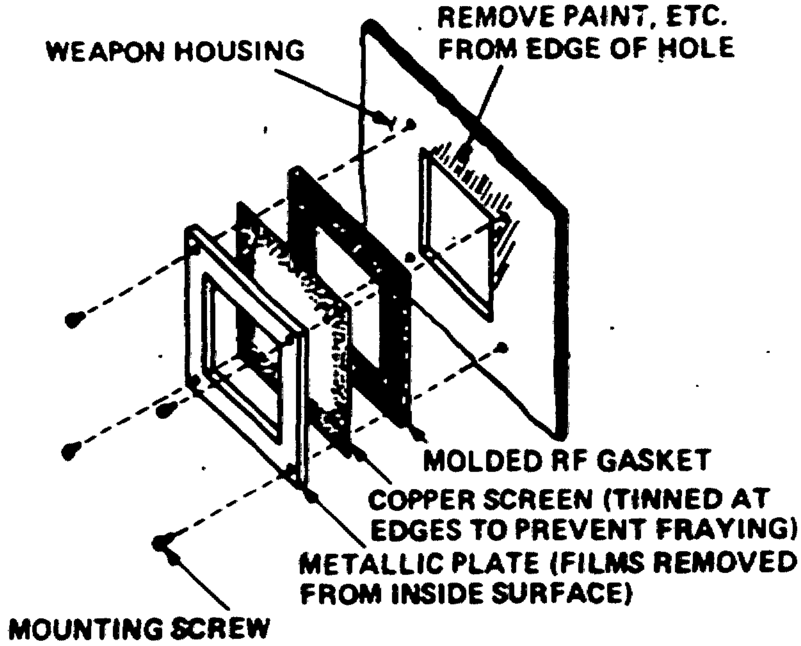


Figure G-6. Method of Mounting Wire Screen Over a Large Aperture. [G-1]

Two approaches can be employed when shielding a meter assembly. These approaches are shown in Figure G-7. 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. Glass coating 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.

30.1.4 **CABLE AND CONNECTOR SHIELDING.**

30.1.4.1 **TYPES OF SHIELDED CABLES.** Unshielded and unfiltered conductors which are exposed to the interference environment and penetrate the shielded enclosure may completely negate the shielding effectiveness of the enclosure. Therefore, it is necessary to shield and/or filter all conductors which penetrate the shielded enclosure.

There are several methods for shielding cables. These include: (1) braid; (2) flexible conduit; (3) rigid conduit; and (4) spirally wound shields of high permeability materials.

Braid is used for cable shielding in applications where the shield cannot be made of solid material. Advantages are ease of handling

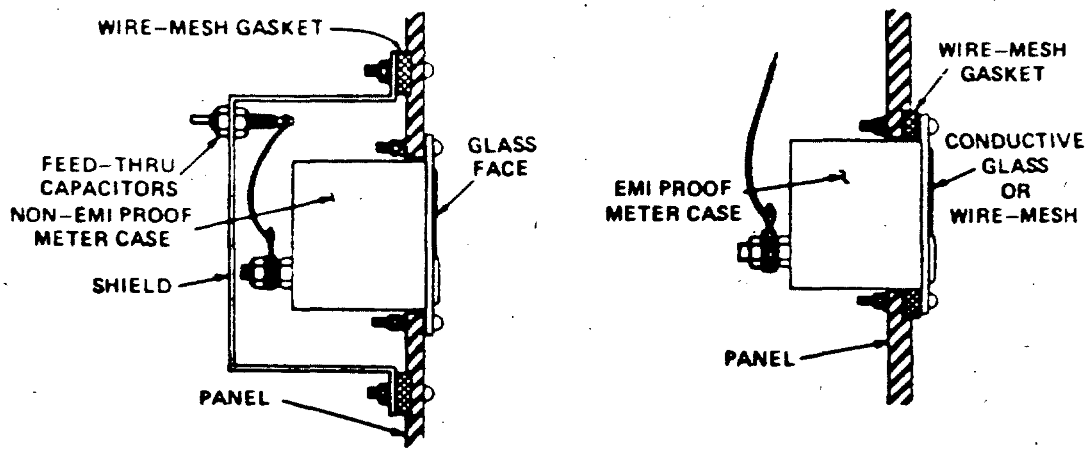


Figure G-7. Meter Shielding Techniques. [G-6]

in cable makeup 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.

Conduit, either solid or flexible, may also be used to shield 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.

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.

Data on the shielding effectiveness of cables is not readily available. The lack of data is primarily due to the fact that there is no standardized procedure for collecting such information, and because a large number of parameters (some of them external to the cable itself) influence the particular performance of a cable. These parameters include termination impedances, impinging signal direction and impedance, cable length relative to the interference signal frequency, the particular connectors employed. flexing requirement, and others.

The general characteristics of four classes of shielded cables are identified in Table G-5. 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.

30.1.4.2 **SHIELD TERMINATIONS.** If the effectiveness of a shield is to be maintained, the cable shield must be properly terminated. In an otherwise adequately shielded system, 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 cable exposed to high power RF fields.

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: (1) too much solder increases the center conductor diameter, thus increasing shunt capacitance; and (2) too little solder decreases 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 G-8. 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.

TABLE G-5

COMPARISON OF SHIELDED CABLES

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

[1]may lose its effectiveness when flexed.

An alternative to crimping is shown in Figure G-9. 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 G-9). 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° 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.

Figure G-10 illustrates the type of connector that should be used when a shielded cable assembly contains individual shielded wires. 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.

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.

30.1.4.3 **FIBER OPTICS.** In recent years, there has been a significant increase in the use of fiber optics (optical waveguides) as a replacement for conventional cables. Fiber optics offer significant performance and operational advantages over conventional cables such as wide bandwidths, low cost, and freedom from conventional transmission line problems such as standing waves, and the dependence of transmission loss upon frequency.

Fiber optics also offer significant advantages over conventional cables in terms of their EMC/EMI properties, particularly in terms of their immunity to the coupling of undesired signals. When exposed to a radiated environment, a fiber optics link will be esentially transparent to signals in the environment. Hence, fiber optics offer an attractive alternative to the use of shielded cables for the prevention of signal coupling.

Since the link is transparent to the incident EMR environment, the coupling of undesired signals to the link (which might occur with conventional cables, even if well-shielded) will not be a problem. Even though the fiber optics link itself will not serve as a source of EMR signals, the entry of the link into the subsystems must be appropriately treated to eliminate apertures through which undesired signals might enter. In most cases, a wave-guide-below-cutoff-treatment of the aperture should be sufficient to eliminate signal entry.

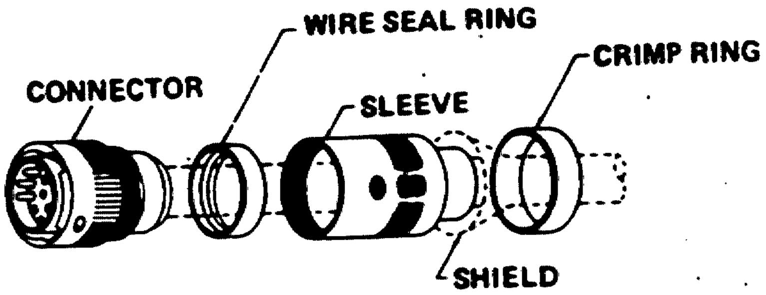


Figure G-8. Shield Termination Using Crimping. [G-1]

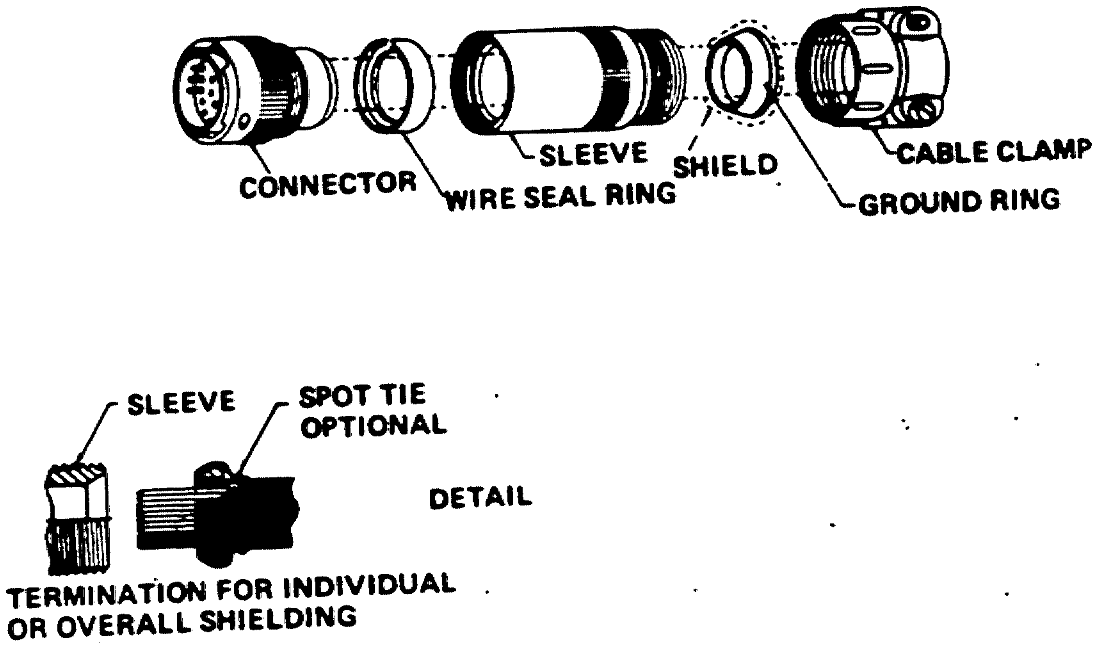


Figure G-9. Shield Termination Using Threaded Assembly. [G-1]

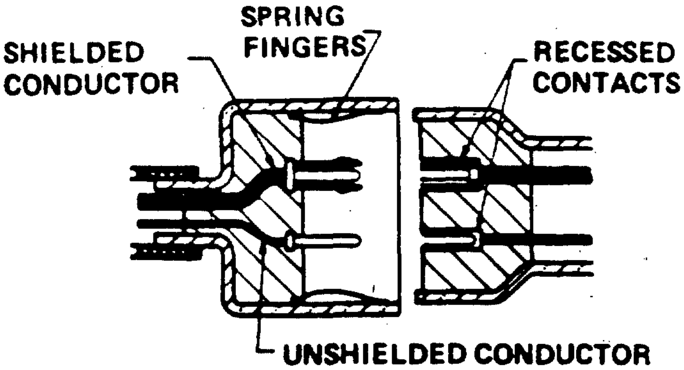


Figure G-10. Connector for Shield Within a Shield. [G-1]

30.1.5 **COMPOSITE MATERIALS.** Fiber reinforced laminates or composites are increasingly being used for the design of system enclosures. Composite materials offer the mechanical advantage of high strength and rigidity coupled with significant weight savings. Thus, the use of composite materials in place of metal for structure designs can improve system performance in terms of such factors as mission range, speed, and payload.

One area of concern which must be addressed when composite materials are used for the structure or framework of a system is the shielding effectiveness of the materials. Generally, composite materials will not provide the degree of shielding that can be obtained with metallic structures. Thus, if system enclosures (outer shell, inner enclosures, etc.) are to be constructed using composite materials, care must be taken to ensure that the required system hardening is not compromised.

Depending upon the system hardening requirement, the use of composite rather than metallic material for structure or enclosure construction may impose more stringent hardening requirements on components, circuitry, or subsystems which are internal to the enclosure. If the external enclosure does not provide sufficient shielding to protect internal components and circuits from the EMR environment, then internal metallic shields around these components and circuits and/or other internal hardening techniques will be required. The capability for achieving the required hardening under this approach as well as possible cost penalties involved must be weighed against the performance advantages gained through the use of composites.

The state-of-the-art of the EM properties of composite materials is quite limited. Although investigations of these properties are underway by various military and industrial organizations, well-documented design information is not yet available for use in evaluating the application of composite materials from an EMR hardness viewpoint. Until design data on composites becomes available and is reduced to practice, any system design which employs composite materials must be accompanied by a program to assess the EM impact of the design. References G-51 through G-66 of Section 40 identify material which will be helpful in such a program.

30.1.6 **SUMMARY OF SHIELDING GUIDELINES.** The following represent the major considerations in the design of an equipment shielded enclosure:

(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 EM compatibility, although factors such as reduced cable flexibility with double braids may be decisive.

(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 the selection of materials that are not only suitable from the shielding standpoint, but from the electrochemical 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 of the seam surfaces as possible.

(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 shielding and finishing specifications conflict, it is important that the finishing requirement be modified.

(i) The critical factors in 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 the particular methods employed in particular situations.

30.2 **FILTERING.**

30.2.1 **INTRODUCTION.** In the design of an equipment shielded enclosure which is to protect a system's circuits from an interference environment, it is important that any wire or cable which will be exposed to the interference environment and penetrates the shielded enclosure be filtered to maintain the integrity of the shielding effectiveness of the enclosure. In addition, the designer should consider including filters in circuit designs and/or interconnecting wiring designs to prevent interfering signals from being conducted through the system circuits.

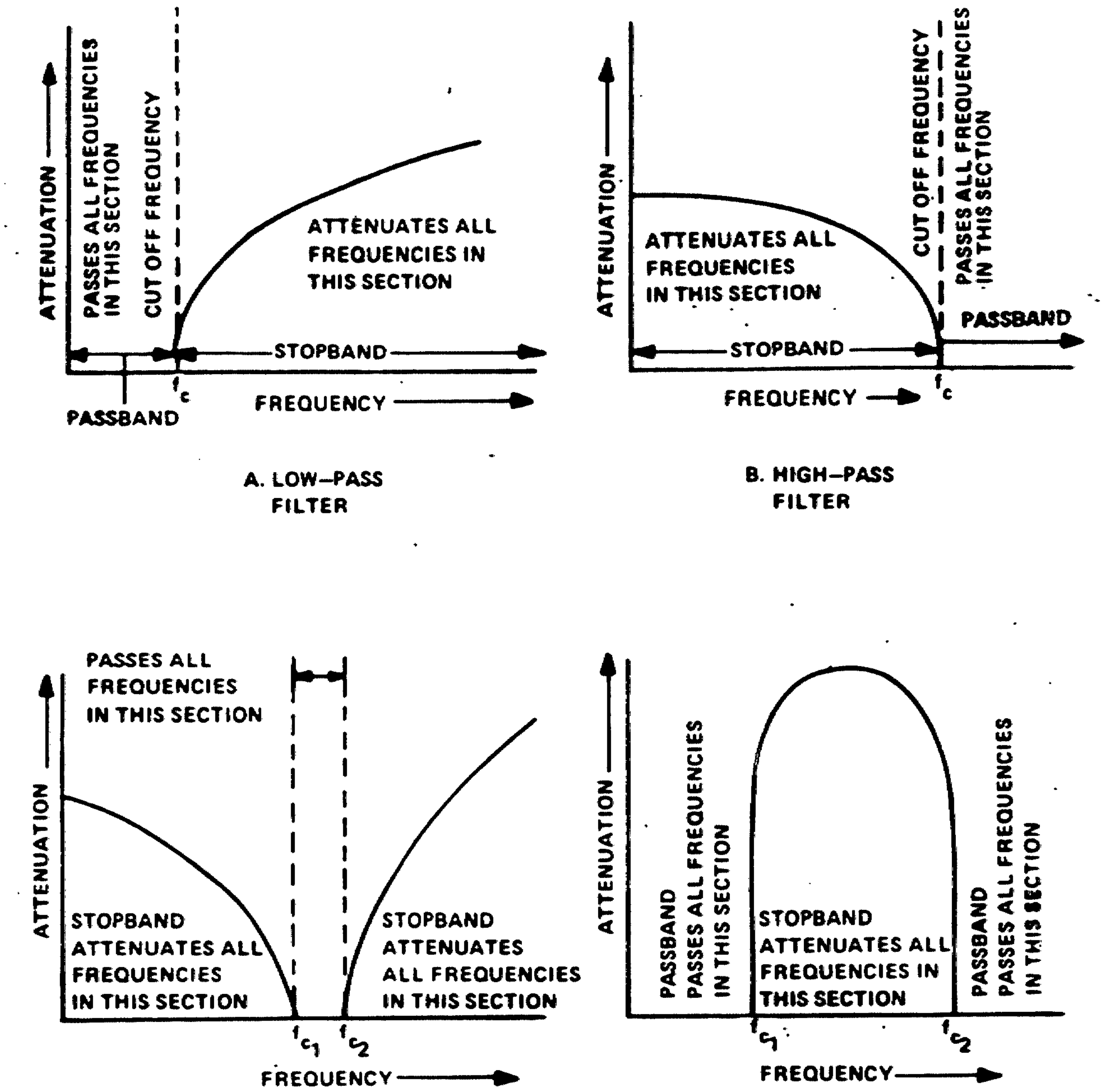
The purpose of this section is to provide the designer with general filter information which will be useful in the selection and application of filters for EMI hardening purposes. A comprehensive discussion of filter design techniques is considered beyond the scope of this handbook. If this type information is required, it is available in a number of documents listed in the bibliography at the end of this section. The emphasis in this section is to make the designer aware of the types of filter available, how these filters may be applied to EMI problems, and the parameters which must be considered in specifying a filter for a particular application.

30.2.2 **FILTER TYPES AND APPLICATIONS.** An electrical filter can be defined as a network of lumped or distributed constant elements (capacitors, inductors, and resistors, or their equivalent) that permits the transmission of signals at some frequencies and impedes the transmission of signals at other frequencies. The passband of a filter is the frequency range in which there is little or no attenuation. The stopband is the frequency range in which attenuation is desired.

Filters are divided into four basic classes according to the relative positions of the passbands and stopbands in the frequency spectrum. The four basic classes of filters are low-pass, high-pass, bandpass, and band-reject. The attenuation as a function of frequency for each of the classes is illustrated in Figure G-11.

A low-pass filter (illustrated in Figure G-11A) passes all frequencies below its cutoff frequency (f_c) and theoretically attenuates all frequencies above the cutoff frequency. This type filter is used extensively in EMI control applications. Power line filters are low-pass type filters which pass dc or ac power frequencies without significant power loss while attenuating signals above these frequencies. In addition, low-pass filters are used on control and signal lines where all interference frequencies are above the desired signal frequencies.

A high-pass filter (Figure G-11B) passes all frequencies above its cut-off frequency and attenuates all frequencies below the cutoff



A. LOW–PASS FILTER

B. HIGH–PASS FILTER

C. BAND–PASS FILTER

D. BAND–REJECT FILTER

Figure G-11. The Four Basic Classes of Filters.

frequency. High-pass filters are used on lines where all of the interference frequencies are lower than the desired signal frequencies. In particular, such filters are used to remove ac power line frequencies from signal channels.

A bandpass filter (Figure G-11C) passes all frequencies between a lower cutoff frequency (f_{c1}) and an upper cutoff frequency (f_{c2}) and attenuates all frequencies below f_{c1} and above f_{c2}. This type filter is used in cases where interference frequencies are both lower and higher than the desired signal frequencies.

A band-reject filter (Figure G-11D) attenuates all frequencies between a lower cutoff frequency (f_{c1}) and an upper cutoff frequency (f_{c2}) and passes all frequencies below f_{c1} and above f_{c2}. This type filter is used where the interference or undesired signals are within a restricted frequency range and the desired signal frequencies may be located over a considerable frequency range both above and below the undesired signal band.

Filters are also classified according to the manner by which attenuation is achieved. Reactive, or lossless, filters provide attenuation of unwanted signals by reflecting energy back to the source. Absorptive, or lossy, filters attenuate unwanted signals by converting them to heat in a lossy dielecric or in a thin layer of resistance material.

There are two factors which significantly affect the effectiveness of reactive, or reflective, type filters. These factors become extremely important in applications where the filters are required to exhibit either passband or stopband characteristics over extremely wide frequency ranges (for example, a low-pass interference suppression filter which is required to attenuate interference frequencies over the frequency range from 1 MHz to 18 GHz). In order for a reflective type filter to exhibit the design or specified bandpass and stopband characteristics, it is necessary that both the input and output terminals of the filter be terminated with the design impedances of the filter. It is important to note that these matched impedance conditions must be satisfied over the entire stopband region as well as the passband region if the specified attenuation is to be realized. In cases where the desired stopband (or passband in the case of a high-pass or band-reject filter) covers several octaves or decades of frequency range and extends into the microwave frequency range, it is extremely difficult (if not impossible) to maintain the matched impedance conditions (even if they are known). In addition, for some appliations such as power line filtering, the source, or input, impedance is probably unknown and may vary drastically with frequency. Under these conditions, it should be realized that the performance of the filter will differ from the design specifications. A second factor to be considered with reflective type filters is the fact that they will exhibit spurious resonances which will degrade the stopband or passband characteristics where the bands extend over the large frequency ranges of several octaves. These spurious resonances result from the stray, or parasitic, reactance associated with lumped element filters and from the inherent periodicity in transmission line filters. The effects of these spurious responses on the attenuation characteristics of a lumped element filter and a transmission line filter are illustrated in Figures G-12 and G-13 respectively.

It is apparent from the discussions above that the use of reflective, or low loss, filters in applications requiring either stopband or passband characteristics over very broad frequency ranges extending into the microwave frequency range may result in the performance of the filters being significantly different from the design specifications. These deviations of performance from design specifications result from an inability to maintain an impedance match at the input and output terminals of the filters and/or spurious resonances of the filters within the frequency range of interest. In most cases where reactive filters are used in these applications, it will be necessary to perform measurements to determine the characteristics of the filters after they are installed.

The deficiencies of reflective type filters in EMI applications led to the development of lossy, or dissipative, filters that take advantage of the loss-versus-frequency characteristics of materials such as ferrite compounds and carbonyl iron mixes. These materials have the unique charactristic of low dc attenuation and good high frequency attenuation over broad continuous frequency ranges. The attenuation of the lossy filter is directly proportional to the distance that the signal travels through the lossy material and is specified in terms of dB per Megahertz per unit length. A significant feaure of dissipative filters is that they do not exhibit spurious passbands in the stopband region. In addition, since the undesired energy is absorbed in the lossy material of the filter, an impedance mismatch at the input and/or the output terminals of the filter has no significant effect on the attenuation characteristics of the filter. Typical attenuation characteristics of a lossy filter are illustrated in Figure G-14. The filter becomes extremely lossy in the frequency range where either electric or magnetic losses, or both, become large and increase rapidly with frequency. Dissipative filters of this type are necessarily low-pass; and a major application is general purpose power line filtering.

In cases where more rapid attenuation slopes are required, a hybrid dissipative-reflective filter can be utilized. With proper design, the sharp cutoff characteristics of the reflective filter can be realized while the dissipative features of the filter will eliminate the spurious passbands in the stopband region and reduce the impedance matching requirements for the filter. Typical attenuation characteristics of a hybrid dissipative-reflective filter are shown in Figure G-15.

Another concept of the lossy filter is the use of ferrite beads. A ferrite bead slipped over a wire produces a single-turn RF choke and exhibits low impedance at low frequencies and moderately high impedance over a wide high frequency range. Adding more or longer beads provides additional series inductance and resistance in direct proportion.

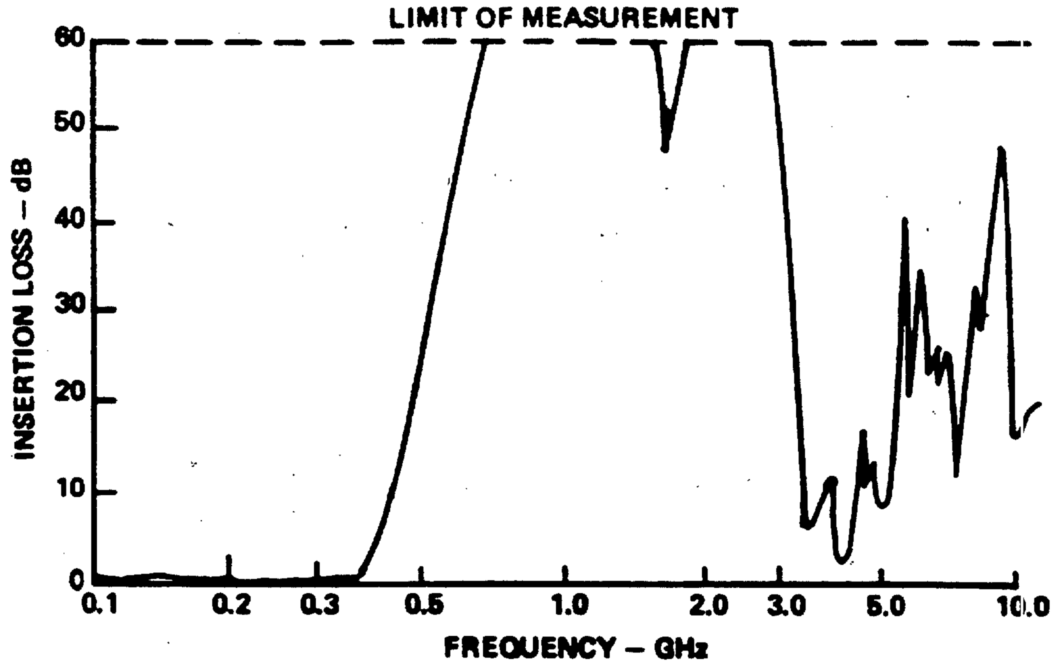


Figure G-12. Typical Low-Pass Attenuation Characteristics of a Lumped Element Filter.

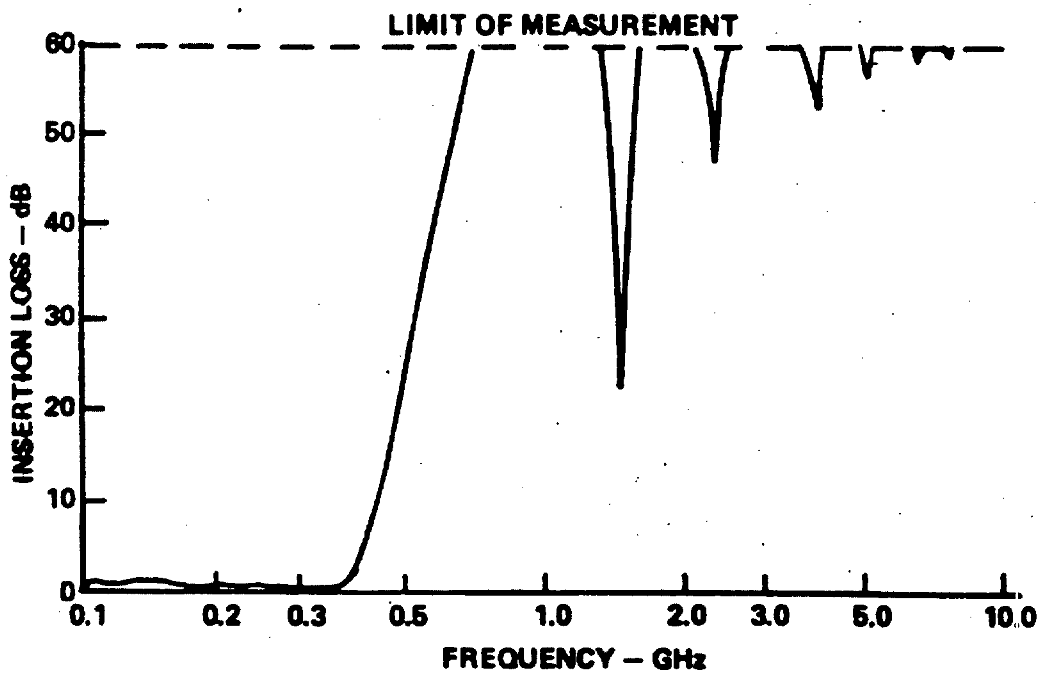


Figure G-13. Typical Low-Pass Attenuation Characteristics of a Transmission Line Filter.

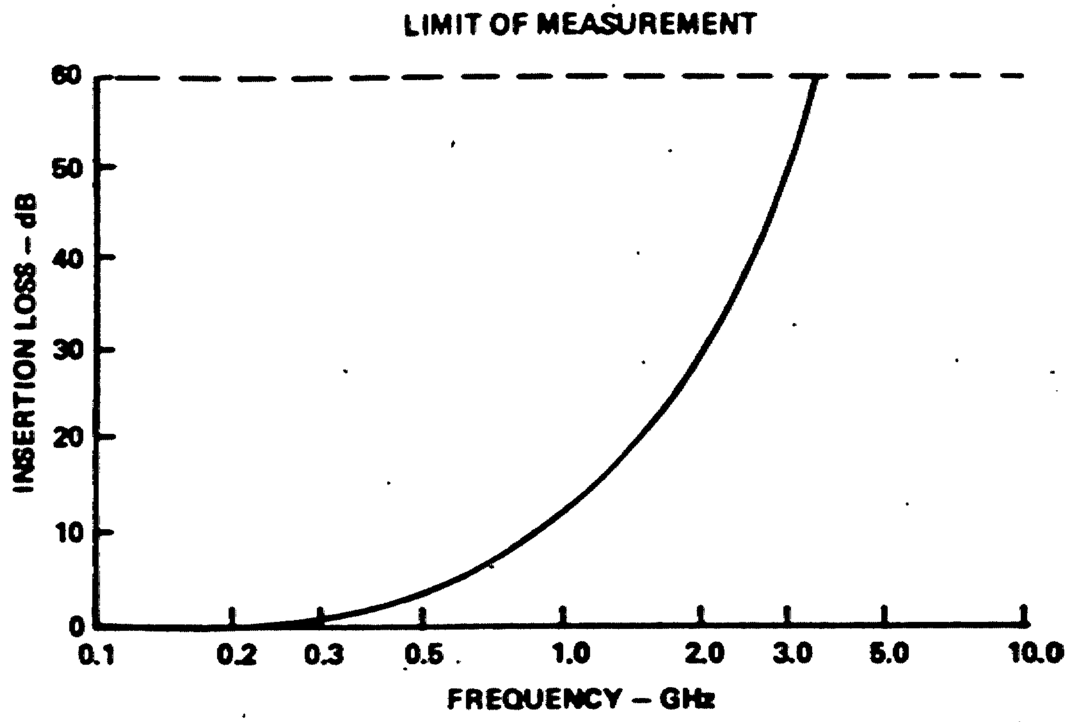


Figure G-14. Attenuation Characteristics of a Dissipative Filter.

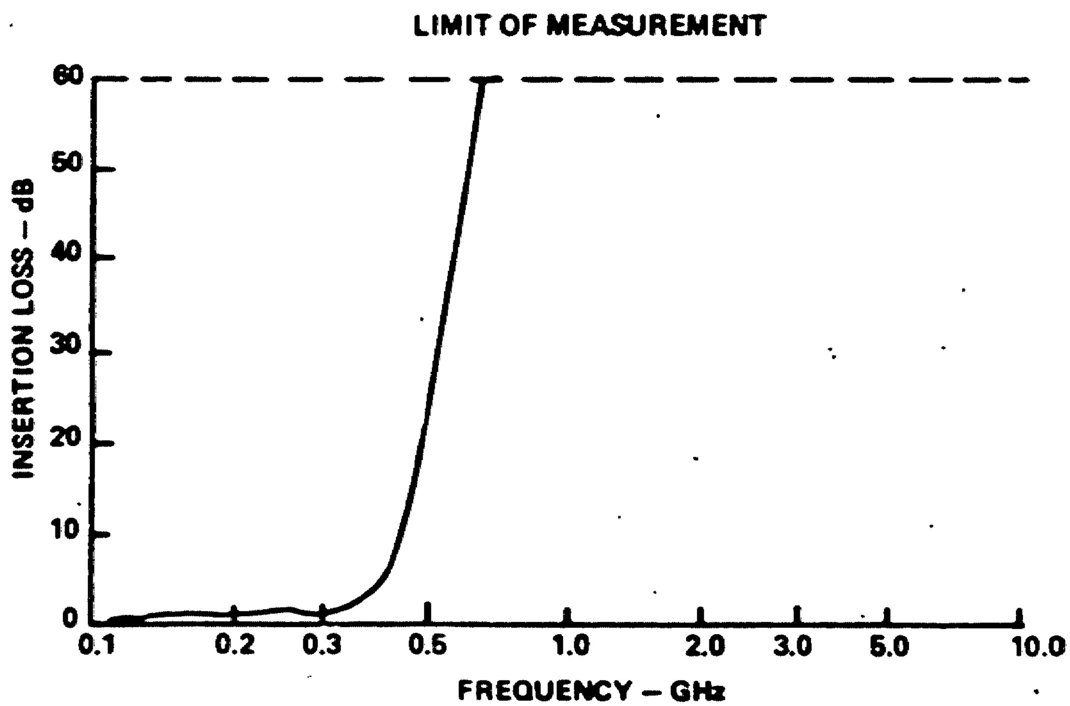


Figure G-15. Attenuation Characteristics of a Hybrid Dissipative - Reflective Filter.

Still another concept of lossy filtering is available in filter-pin connectors. In these type connectors, the filter is built into the cable-pin assembly. Each filter-pin is configured as a π-type connector through the use of lossy material surrounding the pin, and shunt capacitors between the pin and the connector shell. Filter-pins have been miniaturized to such small size that filter-pin connectors are now available with as many as 128 pins. However, because the shunt capacitance and series inductance that can be constructed in the pin are limited, filters of this small type offer little attenuation below about 1 MHz. In a 50 ohm system, the typical attenuation offered by filter-pins is approximately 20 dB at 10 MHz and up to 80 dB at 100 MHz.

30.2.3 **FILTER INSTALLATION AND MOUNTING.** In order to achieve the desired results with filters, it is absolutely necessary to adhere to certain guidelines with respect to the installation and mounting of the filters. The RF impedance between the filter case and ground must be made as low as possible. Otherwise, the filter insertion loss may be seriously degraded at the higher frquencies. The preferred contact between the filter case and ground is accomplished by a metal-to-metal bond between the filter case and the shielded enclosure wall or equipment chassis. In addition, effective separation between the input and output wiring of the filter is mandatory to prevent radiation from the input wiring to the output wiring from circumventing or degrading the effects of the filter. If complete isolation is effected between input and output, a filter insertion loss approaching the design specification can be realized.

Where possible, the use of bulkhead mounted feed-through filters is recommended since this configuration is optimum for establishing a good RF bond between the filter case and ground and provides complete isolation between the input and output terminals of the filter. In cases where feedthrough filters are not used, it will probably be necessary to provide additional compartmental shielding to isolate the input and output terminals of the filters.

30.2.4 **SPECIFYING FILTERS.** In selecting or designing a filter for a particular application, many parameters must be taken into account if the filter is to be effective. The attenuation versus frequency characteristic is obviously the primary factor that determines the suitability of a filter for a particular application. However, other electrical and mechanical requirements must be specified. These include the following:

(a) Impedance Matching - The input and output impedances must be specified to match the impedance of the line into which the filter will be inserted. The impedance matching is particularly critical for transmission lines, so that the filter does not impair the normal operation of the equipment on both ends of the line. In addition, care must be taken that the filters to be used do not degrade the desired performance of circuits within the system. This includes waveform distortion as well

(b) Voltage Rating - The voltage rating of the filters must be specified to assure that each filter is adequate for it particular application. The filter voltage ratings must be sufficient to provide reliable operation under the extreme conditions expected. However, specifying a rating higher than required will result in penalties in size, weight, and additional cost.

(c) Current Rating - The current rating of the filter should be specified for the maximum allowable continuous operation of the circuit in which it is installed and should be consistent with the current rating of the wire, components, circuit breakers and fuses with which it will be used. A current rating higher than required will add size, weight, and cost penalties.

(d) Voltage Drop - The maximum allowable voltage drop through the filter should be specified. With the maximum current specified, the voltage drop requirement specifies the maximum passband insertion loss of the filter.

(e) Frequency - The relative frquencies and magnitudes of the desired and undesired signals must be considered when specifying the frequency characteristics of a filter. In general, the size, weight, and cost of a filter rise rapidly as the attenuation slope increases.

(f) Temperature - Filters must be able to withstand the environmental operating ranges of the equipment in which they are used. The specified temperature range for the filters must include both the extreme low and extreme high temperatures in which the equipment will be required to operate.

(g) Size and Weight - In most cases, size and weight will be important considerations in the selection of filters. Filter manufacturers are fairly flexible in being able to provide a wide choice in the shape of a filter case, the method of mounting, and the types of terminals and connectors.

30.2.5 **SUMMARY OF FILTERING GUIDELINES.** The following represent the major considerations in the selection and installation of filters:

(a) In the selection, procurement, or design of filters for various applications in a system, the parameters of the filters should be specified in accordance with Section 30.2.4.

(b) The RF impedance between the filter case and ground must be made as low as possible. Otherwise, the filter insertion loss may be seriously degraded at the higher frequencies. The preferred contact between filter case and ground is accomplished by a metal-to-metal bond between the filter case and the shielded enclosure wall or the equipment chassis.

(c) Complete isolation between the input and output wiring is mandatory to prevent radiation from the input wiring to the output wiring from degrading the effects of the filter. The use of bulkhead mounted feed-through filters is recommended since this configuration is optimum for establishing a good RF bond between the filter case and ground and provides complete isolation between the input and output terminals of the filter.

30.3 **BONDING.**

30.3.1 **INTRODUCTION.** Bonding is the establishment of an electrical union between two metal surfaces to provide a low impedance path between them. This path may be between a ground reference and a component, circuit, shield, or structural element. The purpose of the bond is to establish an electrically homogeneous structure to prevent the development of electrical potentials between individual metal surfaces which can produce interference.

Good bonding within a system is essential for minimizing interference. Some of the desirable characteristics of good bonding are the following:

(a) Good bonding enables the design objectives of other methods of interference suppression, such as shields and filters, to be achieved.

(b) Good bonding minimizes the buildup of RF voltage differences.

(c) An adequate bond deters the buildup of static charges during equipment operation.

There are two types of bonds: direct bonds where there is metal-to-metal contact between the surfaces to be bonded, and indirect bonds where the surfaces to be bonded cannot be placed in direct contact.

30.3.2 **DIRECT BONDS.** Direct bonds include both permanent and semi-permanent metal-to-metal joints. Permanent joints made by welding, brazing, sweating, and soldering are the best direct bonds. Semi-permanent joints of machined metal surfaces rigidly held together provide excellent direct bonds, as long as the contact areas are clean prior to assembly and the clamping pressure is applied in a manner to assure that a constant pressure is maintained under stress and vibration conditions. Joints that are press-fitted or joined by self-tapping or sheet metal screws cannot be relied upon to provide a low impedance bond at high frequencies. Direct bonds must always be made through continuous contact to bare or conductively finished metals.

30.3.3 **INDIRECT BONDS.** An indirect bond is an intermediate electrical conductor used to connect two isolated metal surfaces. It should be noted that indirect bonds are only substitutes for direct bonds and should be used only when direct bonds are not possible (such as when motion is required between surfaces to be joined). Indirect bonds are accomplished by means of bonding jumpers or bonding straps. Bonding jumpers are short woven braid conductors for use in applications where the interfering signal frequency is below a few megahertz. They are generally used in low-frequency devices, and where the development of static charges must be prevented. Bonding straps are preferably

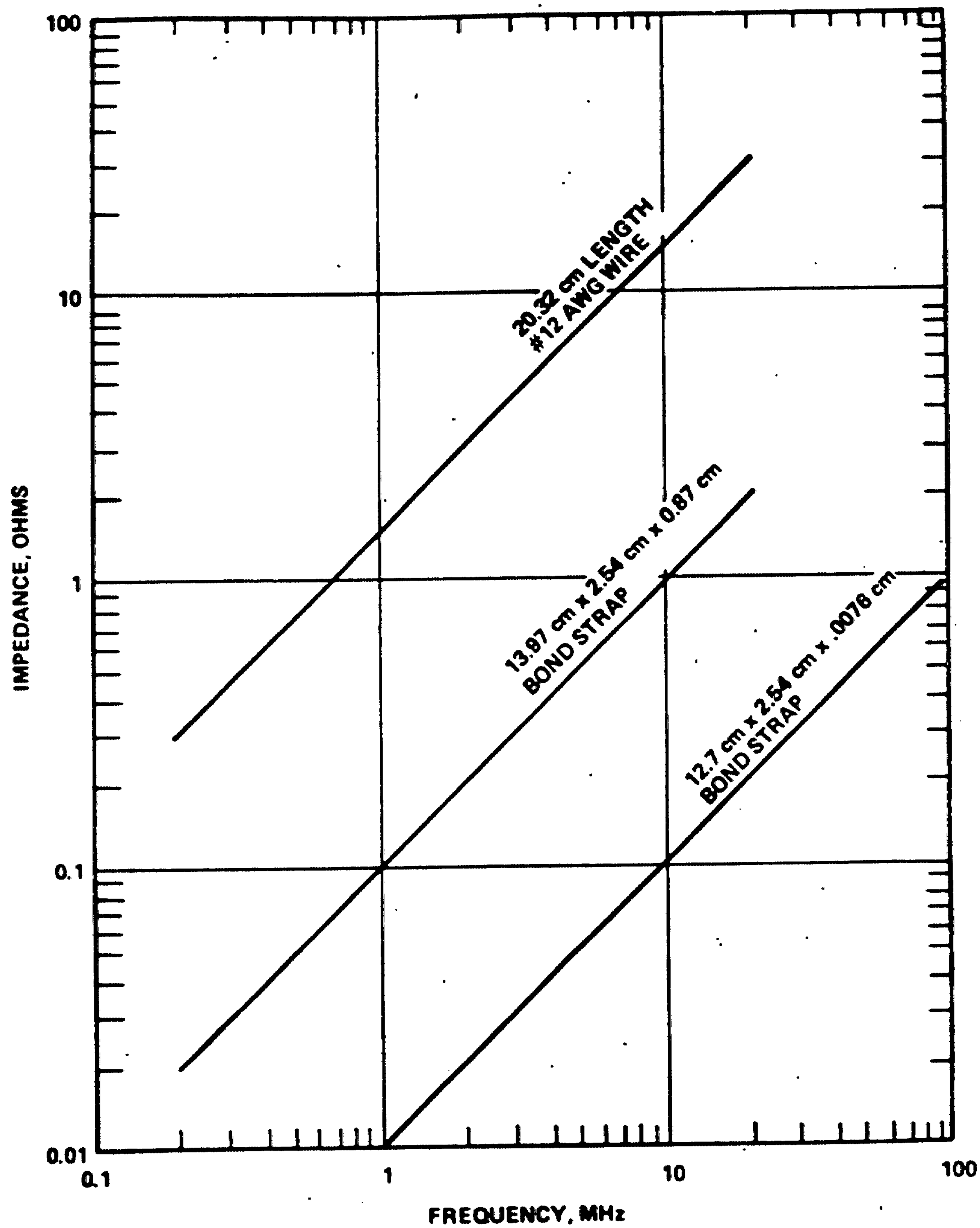


Figure G-16. Impedance of Bond Straps and No. 12 AWG Wire. [G-6]

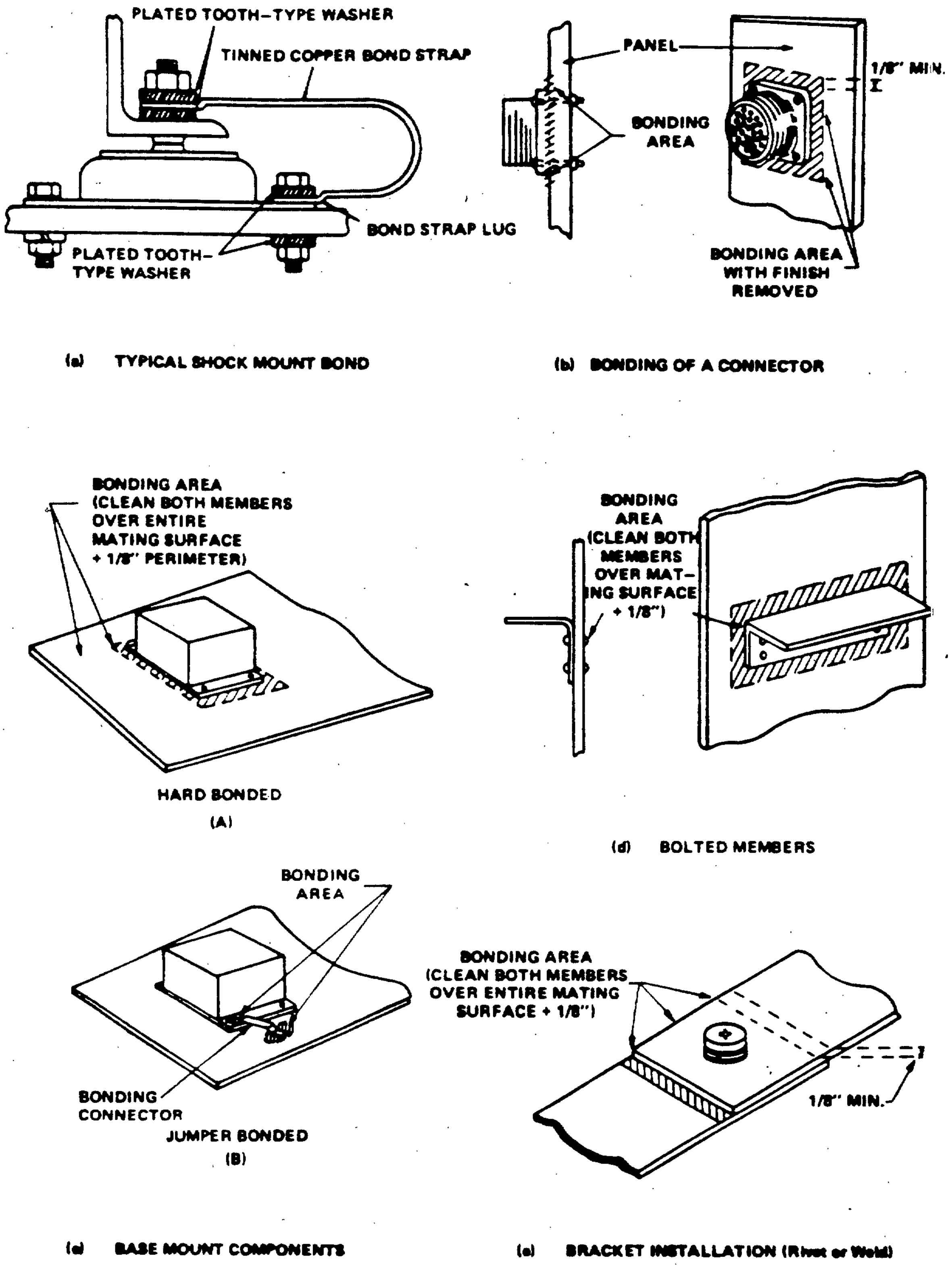


Figure G-17. Typical Examples of Direct and Indirect Bonds. [G-1] wide, flat, thin, and short straps of the same type of metal as the two surfaces to be bonded. The bonding strap should provide large contact areas and low RF impedance between the two surfaces to be bonded over the frequency range of interest. The influence of the physical characteristic of the bond strap is illustrated in Figure G-16 where the impedances of two bonding straps and a 20.32 cm length of No. 12 wire are shown plotted as a function of frequency. The relatively high impedance at high frequencies illustrates that there is no adequate substitute for direct metal-to-metal contact. A rule of thumb for achieving minimum bond strap impedance is that the length-to-width ratio of the strap should be 5:1 or less. When an indirect bond is absolutely necessary, the length should be made as short as possible.

Typical examples of direct and indirect bonds are shown in Figure G-17 to illustrate the variety of techniques that are employed to obtain low-impedance connections. Figure G-18 illustrates how an overlap seam, accompanied by soldering or spot welding, can provide a relatively effective bond. Many other examples are available in the literature, particularly in MIL-STD-1310B.

30.3.4 **TEMPORARY BONDS.** 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.

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 ensure consistent contact pressure over an extended period of time sould 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-welded permanent bonds as well) is the linear spacing of the fasteners or spotwelds. Figure G-19 provides an indication of the sensitivity of this parameter for a 1.27-cm aluminum lap joint. Data is taken at 200 MHz. The shielding effectiveness shown at 2.54-cm spacing is about 12 dB poorer than the identical configuration incorporating a 1.27-cm wide monel mesh gasket; the effectiveness of 25.4-cm spacing is about 30 dB poorer with the same gasket.

30.3.5 **SURFACE PREPARATION.** Both direct and indirect bonding connections require metal-to-metal contact of bare surfaces. It is frequently necessary to remove protective coatings to provide a satisfactory bond. The area cleaned for bonding should be slightly larger than the area to be bonded. Immdiately prior to bonding, all chips, grease, or other foreign matter should be removed with a cleaning solution.

After bonding, the exposed areas should be refinished as soon as possible. A suitable conductive coating should be used when removable components must be provided with a protective finish.

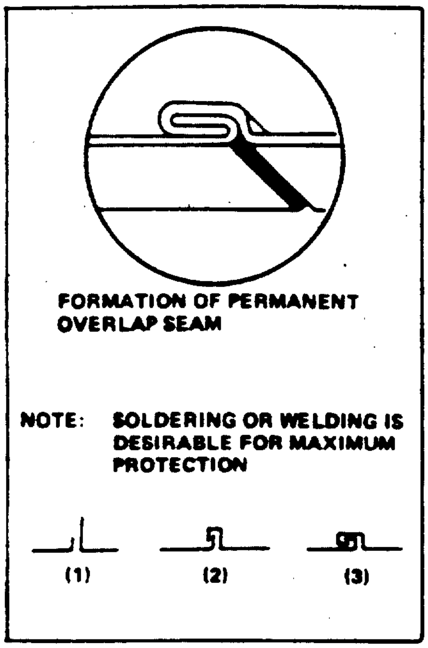


Figure G-18. Formation of Permanent Overlap Seam. [G-1]

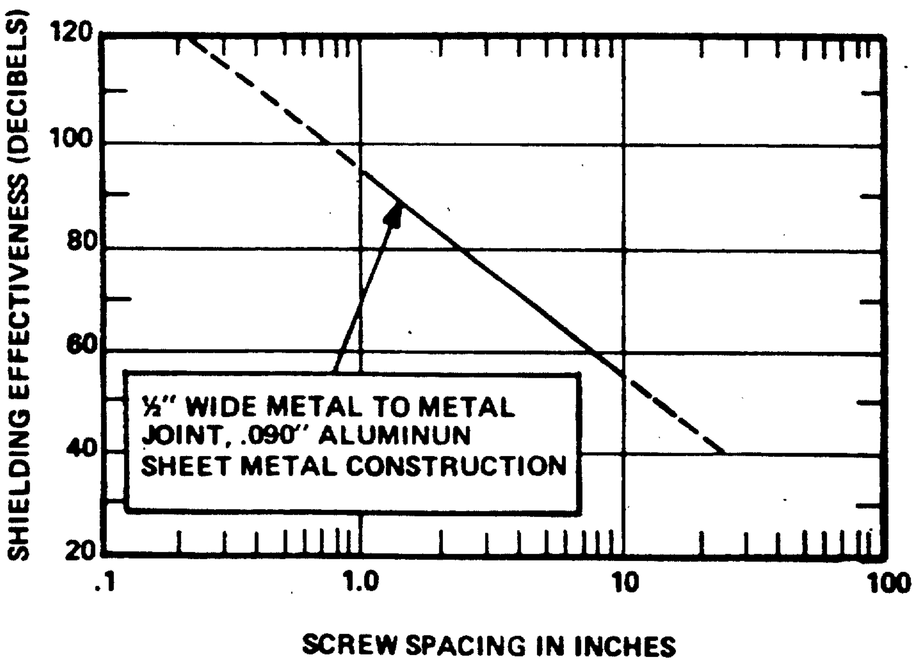


Figure G-19. Influence of Screw Spacing. [G-1]

In bonding, the necessity for joining dissimilar metals is frequently unavoidable. In such cases, galvanic corrosion becomes an important consideration. Factors contributing to galvanic corrosion are the relative closeness of the metals in the electromotive series and the amount of moisture present.

Several methods can be employed for minimizing or preventing corrosion and its adverse effects on bonding. One method is to use metals low on the activity table, such as copper, lead, or tin. Where the two metals to be joined are widely separated on the acivity table, it is sometimes practical to use a plating such as cadmium or zinc. Thin, bimetal plates, formed by mechanically bonding dissimilar metals cold flowed together under high pressures, are sometimes used to interconnect two dissimilar metals. The possibility of galvanic and/or electrolytic action necessitates extreme care in assembling joints that serve as bonds. Surfaces should be absolutely dry before mating, and should be held together under high pressure to minimize the possibility of moisture entering the joints.

30.3.6 **SUMMARY OF BONDING GUIDELINES.** Some general guidelines to obtaining good bonds are listed below:

(a) All mating surfaces must be cleaned before bonding.

(b) Direct metal-to-metal joints between similar metals formed by welding, sweating, or brazing represent the best bonds.

(c) 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.

(d) 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.

(e) 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 bonderite, 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.

(f) 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.

(g) When two dissimilar metals must be bonded, metals that are close to one another in the electrochemical series should be selected.

(h) Indirect bonds are only a substitute for direct bonds. Indirect bonding conductors should be in strap form; broad in width, thin, and as short as possible to provide low-impedance paths at higher frequencies. The length-to-width of the bonding straps should be less than 5:1.

(i) Bonds should provide good metal-to-metal contact over the entire mating surfaces. The fastening method should exert sufficient pressure to hold the surfaces in contact in the presence of deforming stresses, shock, and vibrations.

(j) Where bonding requires the joining of two dissimilar metals, special attention must be paid to the possibility of bond corrosion. Protection of the bond from moisture and other corrosive effects must be provided where necessary.

(k) Soldering may be used to fill the resulting seam, but should not be employed to provide bond strength.

(l) 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 occuring between welds. Riveting or pinning is even less desirable because of the greater susceptibility of bond degradation with wear.

(m) Any overlap seam, accompanied by soldering or spot welding provides a relatively effective bond. Other types of crimped seams may be employed, so long as the crimping pressure is maintained.

30.4 GROUNDING.

30.4.1 **INTRODUCTION.** Grounding is the establishment of electrically conductive paths between selected points in a system and some common reference plane. The reference plane may be a chassis, a shielded enclosure, or a structure with which the system is to operate. An ideal grounding system would result in a system with a common potential reference point anywhere in the system so that no undesired voltage potentials would exist between any two points in the system. However, because of the physical properties and electrical characteristics of grounding materials, no grounding system is ideal and some potential always exists between ground points within a system.

A good ground plane or reference is the foundation for obtaining reliable, interference-free system operation. An ideal ground plane would be a zero-potential, zero-impedance body that could be used as a reference for all signals and power sources in the associated circuitry, and to which any undesired signals could be transferred for their elimination.

The extent to which potentials in the ground system are minimized and ground currents are reduced determines the effectiveness of the ground system. A poor ground system will make it possible for spurious voltages and currents to couple into circuits and subassemblies and can: (1) degrade the shielding effectiveness of well-shielded units; (2) bypass the advantages of filters; and (3) result in EMI problems that are difficult to isolate and resolve.

30.4.2 **GROUNDING TECHNIQUES.** There are three basic signal grounding concepts that are frequently employed in electronic systems, as illustrated in Figure G-20. The approaches can be used separately or in combination in a given system. They are:

(a) Floating ground system,

(b) Single-point grounding system, and

(c) Multiple-point grounding system.

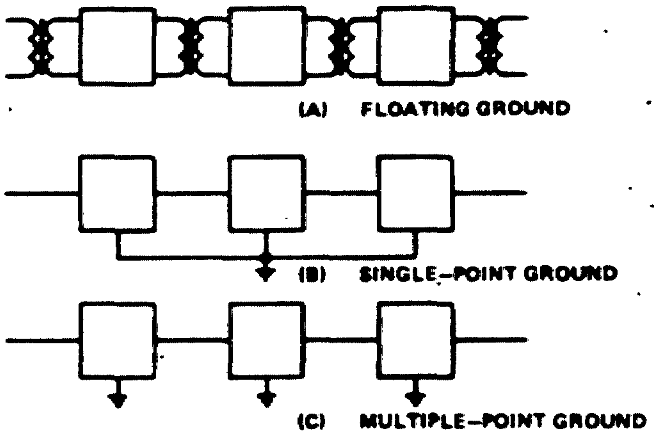


Figure G-20. Grounding Methods. [G-1]

30.4.2.1 **FLOATING GROUNDING SYSTEM.** The floating ground system is a method of electrically isolating circuits or subassemblies from a common ground plane or from any common wiring that might introduce circulating currents. The effectiveness of a floating ground system depends on the degree of isolation that can be established and maintained between the circuits of interest and other references.

In many cases, complete isolation may be very difficult to achieve, particularly at high frequencies. Also, certain hazards exist in the use of floating ground systems, in that static charges or lightning potentials may accumulate between the floating grounds and other references such as the equipment enclosure or power line neutrals of nearby structures.

Isolation transformers, optical isolators, and bandpass filters are techniques commonly utilized to obtain the isolation required for a floating ground system. This method is especially effective at audio and low radio frequencies. However, for higher operating frequencies or interference frequencies, the effectiveness of the technique diminishes due to coupling paths that bypass the isolation devices.

30.4.2.2 **SINGLE-POINT GROUNDING SYSTEM.** A single-point grounding system is one in which a single physical point in the circuitry is defined as the ground reference point. All ground connections are tied to this single point as illustrated in Figure G-20B. Under the single-point grounding concept, the potentials existing at different points on the ground plane and the currents circulating in the ground plane do not affect the grounding of the individual circuits.

At high frequencies where the wavelengths involved approach the length of the ground wires going to the single-point reference, single-point ground systems are no longer practical. At these frequencies, the ground wires become efficient antennas and the stray inductance and capacitance associated with the ground wires prevent the single-point reference from being realized.

30.4.2.3 **MULTIPLE-POINT GROUNDING SYSTEM.** A multiple-point grounding system is a concept in which a ground plane is used instead of individual return wires for each circuit as illustrated in Figure G-20C. The ground plane may be a chassis, enclosure, or a ground bus wire carried throughout a system.

The advantages of a multiple-point grounding concept are that circuit construction is easier, the ground system wiring is significantly reduced, and it is the only way to avoid standing-wave effects in the ground system at high frequencies. However, since multiple-point grounding creates many ground loops, the design of the ground system must include considerations for minimizing the areas of the ground loops. In addition, precautions to prevent stress, vibrations, and corrosion from introducing high impedances into the ground system must be included in the design.

30.4.3 **CIRCUIT GROUNDING CONSIDERATIONS.** A significant potential difference may exist between points on a 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 ground plane to a minimum is to arrange circuit components physically so that the ground return paths are short and direct, and have the fewest possible crossings of paths. In this way, the intercircuit coupling of the ground currents will be low and isolated.

Differential or balanced circuits can help reduce the effects of ground circuit interference. Since a differential circuit responds only to the potential difference between its input leads, considerable common-mode interference voltage due to the ground circuit may be simultaneously impressed on both input leads without degrading the circuit performance.

In transistorized digital circuits, the input and output impedances 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 or picking up the interference. By reducing the loop area of these circuits to a minimum, much of the interference problem can be eliminated.

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 avoided. Where the use of such circuits cannot be avoided, the interconnnecting lead must be shielded and the shield grounded at each end.

It is generally good practice to isolate power and signal grounds from each other. This will minimize the possible coupling of signals between these two major types of paths.

30.4.5 **GROUNDING CABLE SHIELDS.** The coupling of interference into circuits of a system can be significantly affected by the grounding of the shields of the interconnecting leads.

Grounding of the shields may be accomplished as single-point or multiple-point grounding. Factors which 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 multi-lead systems, each shield may be grounded at a different physical point as long as individual shields are isolated from each other. Single-point grounding is more effective than multiple-point shield grounding only for short shield lengths. Single-point grounding begins to be ineffective when conductor-length-to-wavelength (L/λ) ratios are greater than 0.1, where the wavelength is that of the highest frequency to be used or the highest frequency interference to be expected on the wire or in the system.

(b) Multiple-Point Shield Grounding - For L/λ ratios greater than 0.1, multiple-point grounding at intervals of 0.1λ is recommended, since the shield can act as an antenna that is relatively efficient at $\lambda/4$ when one end is grounded. When such grounding of the shield at intervals of 0.1λ is impractical, shields should at least be grounded at each end. Multiple-point shield grounding is effective in reducing all types of plane wave coupling, so long as large ground currents do not exist.

When cable shields are grounded, good electrical contact to the shield must 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.

There may be situations when multiple-shielded cable usage is considered necessary. 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. It is considered good practice to use multiple-shielded cable for unbalanced input circuits with sensitivity thresholds better than 50 millivolts and operating frequencies beween 5 kHz and 1 MHz. When biaxial or triaxial cables are employed, single-point grounding of the inner shield should be used. An illustration of how to ground double-shielded cable is shown in Figure G-21.

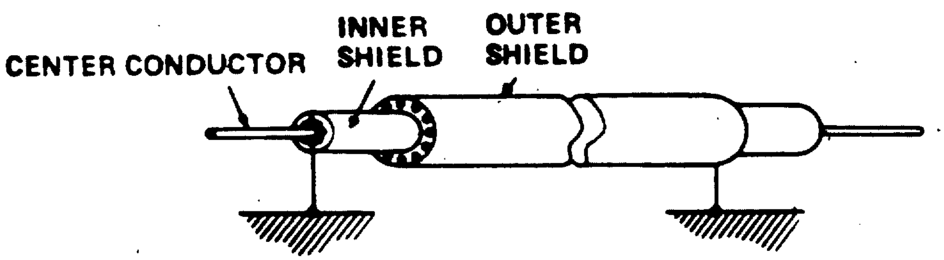


Figure G-21. Examples of Grounding A Double-Shielded Coaxial Cable. [G-1]

30.4.5 **SUMMARY OF GROUNDING GUIDELINES.** The specific grounding philosophy employed on a system will be influenced by the physical configuration, the operational requirements, and the detailed design objectives of the particular system. Several general grounding guidelines are stated below to aid in the selection of grounding approaches and to aid in the implementation of the grounding techniques selected.

(a) Use single-point grounding when the dimensions of the circuit or component under consideration are small compared to the wavelength of concern (typically less than 0.1λ). When possible, ground large circuits or components at several locations, so that the separation between grounds is never greater than 0.1λ

(b) There are occasions when transformer isolation and other isolation techniques can be used to alleviate common-mode noise problems arising from ground network pickup.

(c) Keep all ground leads as short and direct as possible. Avoid pigtails when terminating cables.

(d) 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.

(e) Ground reference planes should be designed so that they have high electrical conductivity, and so that this conductivity is maintained under the stress and vibration conditions encountered in system operation.

(f) Grounds for low-level signals should be isolated from all other grounds.

(g) 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.

(h) Use of differential or balanced circuitry can significantly reduce the effects of ground circuit interference.

(i) 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 reduced equipment susceptibility.

(j) For coaxial or shielded cables, single-point grounding at the source end offers certain advantages at low frequencies. At high frequencies, multiple-point grounding is required.

(k) Low-level sensitive transmission lines may require multiple shields. Single-point grounding of each shield is recommended.

30.5 **SUBSYSTEM HARDENING TECHNIQUES.**

30.5.1 **INTRODUCTION.** The final level of hardening in the layered approach is that of subsystem hardening. Here the critical subsystems with their associated circuits are hardened. If the first two layers (system exterior and cables) have been appropriately hardened, then this level becomes significantly easier to harden. Otherwise, it is a formidable task to harden subsystems when no attention has been given to interrupt the coupling paths leading to the subsystem; this makes it very difficult to determine what are the sources of deleterious energy. Nevertheless, some subsystem hardening is inevitable. From a system operation standpoint, it is usually impossible to close all POE's and adequately shield and filter all cable leads. Various techniques are available to provide the adequate hardening at the subsystem/-circuit level. These are discussed in the following subsections.

30.5.2 **COMPARTMENTALIZATION.** Place all sensitive electronic circuitry in one compartment. This can greatly reduce the amount of subsystem shielding required. It will also make the filtering problem more tractable. In addition, place all noise generators, if possible, in a separate compartment. A high-impedance static ground (one-point ground) should be in the electronic circuitry compartment. This would, for example, prevent charge buildup when this sensitive compartment is separate from the remainder of the system, which may occur when packages are recycled for checkout.

In conjunction with compartmentalization, the separated compartments must be shielded. This, in effect, is building the entire sensitive electrical and electronics subsystems in solid-metal enclosures. The practices necessary for good shielding described in Section 30.1 of this appendix apply here also. Compartmentalization and associated shielding of subsystems is a powerful EMR hardening concept. Clearly, this concept must be violated to some extent to accommodate required penetrations in a practical system.

30.5.3 **COMPONENT/CIRCUIT HARDENING.** Consideration should be given in the initial stages of the design of a system to minimizing the susceptibility of the system components and circuits to interference. The effort should begin with the selection of discrete and IC solid-state devices to be utilized in the system design. The available devices should be screened in order to obtain devices with the highest possible interference susceptibility thresholds. In some cases, this screening/selection process may become difficult, since it may involve some compromise between the desired operating characteristics and the interference susceptibility threshold levels of the devices.

In the selection of signal levels and impedance levels for the circuit designs, the designer should recognize that circuits operating with high signal levels and low impedance levels are less susceptible to interference. Therefore, the circuits should be designed to operate with the highest signal levels and lowest impedance levels that are compatible with the devices involved and consistent with the achievement of the intended functions of the circuits.

The susceptibility of the circuits to radiated interference can be reduced by minimizing the length of interconnecting wiring between components and IC devices and the use of shielded and twisted-pair wire for these interconnections. Since the interference usually has a different spectral content from that of the desired signals, a significant reduction in the susceptibility of circuits to interference can frequently be realized by incorporating filters, chokes, ferrite beads, and lossy ferrite sockets into the interconnecting wiring and power leads of the circuits.

A technique especially applicable to the reduction of interference signals up through the VHF region is the use of common mode cancellation in a differential amplifier. The desired signal source may be either differential or single-ended. The desired signal is fed into a differential amplifier through an RF coupling device which ensures that RF energy appears in equal amounts on each amplifier input. The resulting rectified signal is then rejected through common mode cancellation. Such a technique may provide 10 to 20 dB of system RF hardness and may be used in addition to other shielding and filtering techniques.

In summary, the designer should keep the following considerations in mind:

(a) In general, low-speed devices are less susceptible than high-speed devices. Circuit overdesign with respect to speed and frequency response should be avoided.

(b) If compatible with other requirements, critical or potentially susceptible devices should be operated at low gain and high collector current levels.

(c) Some interference immunity may be provided by employing common-mode cancellation features of differential amplifiers.

30.5.4 **FUNCTIONAL HARDENING.** Functional hardening can be defined as a logical design that insures protection against arbitrary transients through coding and/or timing discrimination, although with a price paid in complexity. This approach is usually used to prevent one-shot failure, such as EED detonation due to a single stray transient. It also has possible application for missiles which must fly in and out of scanning radar beams. For example, a sequential system can be designed so that a particular sequence of pulses, instead of a single relatively short duration signal, is necessary for a state change.

Time discrimination can be used such that a signal must be present for a particular length of time that is longer than that likely due to the deleterious environment.

Under some conditions, parity techniques, such as requiring a signal to appear on two differently routed lines, can provide effective discrimination against arbitrary transients.

30.5.5 **CIRCUMVENTION.** Circumvention involves sensing an EM field that could cause upset and mission failure, and discontinuing signal processing while there is danger of transient malfunction. There are two basic circumvention approaches that should be considered in radiated susceptibility problems of air launched ordnance systems. In the first of these, if EM fields high enough to cause logic changes are detected, the system circumvents by blocking all inputs and recycling the sequence of operations to the last safe reference point. After a designated time delay, operation resumes by resetting where needed and updating to account for the elapsed time. The time delay must be short for missiles. A second type of circumvention involves an extremely hard but slow digital processor (that does not respond to transients) working in conjunction with a soft, fast processor. Ordinarily, the system is under control of the fast processor. However, if severe EM fields are detected, the operation is temporarily turned over to the hard processor, which subsequently resets the fast processor. The first of these two types will probably have the greatest applications.

40. **REFERENCES.** The references in this section are organized into five categories: (1) general references that are applicable to various areas of EMR hardening design; (2) shielding references; (3) filtering references; (4) bonding references; (5) grounding references; and (6) composite material references. Also referenced are articles and documents which, although not used directly as source material, will provide the handbook user with additional information related to EMR hardness design.

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APPENDIX H

EMR HARDNESS MEASUREMENTS

10. **INTRODUCTION.** This appendix describes the elements of an EMR hardness measurement program for air launched ordnance systems. A well planned and implemented measurement program is critical to the design of a system which will perform satisfactorily in its intended operational environment. Measurement data are required at various stages of system design and development to (1) establish hardening design requirements and criteria, (2) verify the hardening design approach, and (3) ensure that the hardening design criteria are being satisfied. The performance of EMR susceptibility tests early in the design and procurement cycle is a very cost-effective approach for achieving overall system hardness goals since the need for extensive system level and field testing in the latter phases of the procurement cycle will be reduced.

EMR hardness measurement programs and conventional EMC test programs are similar in nature in that the goal of both types of programs is to provide measurement data for use in assuring the electromagnetic compatibility between a system and its intended operating environment. However, there are significant differences between the measurement requirements and techniques of EMR and EMC test programs, partially due to the differences in the objectives of the two types of test programs and partially due to the differences in the magnitudes and mechanisms of the interference involved. EMR hardness measurements are designed to define, and aid in tailoring, the susceptibility characteristics of a system to a specified EMR environment, whereas EMC tests are not tailored to a specific environment, do not provide susceptibility data for intense RF environments, and involve both susceptibility and emission measurements for both radiated and conducted interference. For these reasons, the several specifications and standards which establish EMC test plans and define measurement techniques, data requirements, and specification limits for electromagnetic compatibility are not directly applicable to the EMR interference problem and no equivalent documents are available to cover this area. Hence, the system designer/developer must generate his own test program or tailor existing EMC specifications and standards to assure the compatibility of the system with a given EMR environment.

20. **TEST PROGRAM.**

20.1 **GENERAL.** The objective of an EMR hardness test program is to ensure that a system and its integral components will function in a specified manner in its intended operational environment. To meet this objective, measurement data will be required at various stages in the system development cycle, from the initiation of the system design to system delivery. For example, interference susceptibility data on the components to be used in the system are required at the initiation of the system design in order to establish the hardening requirements for the system. As the design progresses, measurement data are needed to establish EMR hardening approaches and techniques and to verify that the hardening design is satisfactory. The following paragraphs summarize the major elements to be considered in an EMR hardness test program.

20.2 **TEST FACILITIES.** EMR hardness measurements generally require extensive test facilities for performing the required measurements. These facilities may include shielded enclosures for isolating the test environment, exposure chambers (anechoic chamber, TEM cell, etc.) for establishing a radiated test environment, and various instrumentation (signal sources, field intensity meters, antennas, and other ancillary test equipment). A detailed description of test facilities for performing EMR hardness measurements is given in Section 30.

20.3 **TYPES OF TESTS.** Three types of tests will usually be required in an EMR hardness measurement program - injection tests, radiated susceptibility tests, and shielding effectiveness tests. Injection tests are simply conducted interference measurements where the EMR signal is injected directly into a port of the unit under test. Injection tests are required for measuring the susceptibility of discrete components/circuits but may also be useful for higher level testing (i.e., testing of subsystems). Radiated tests are used for subsystems where it is necessary to determine the overall susceptibility of a unit comprised of interconnected components, circuits and enclosures. Radiated tests are not applicable to the testing of components because of the special component test fixtures required. Shielding effectiveness measurements are employed to determine the shielding provided by subsystem enclosures or cabling. A detailed description of these three types of tests and the required test facilities is given in Sections 30 and 40.

20.4 **LEVELS OF SUSCEPTIBILITY TESTS.** For the purposes of this handbook, two levels of EMR hardness susceptibility tests are defined -- component/circuit tests and subsystem tests. This definition categorizes units to be tested according to the type of susceptibility test employed. For discrete components/circuits, injection testing is the only feasible means of obtaining susceptibility data, whereas all other units can usually be tested using either injection testing, radiation testing, or both. Thus all system units other than discrete components/circuits are defined as subsystems.

The above definition should not be misconstrued to mean that only one subsystem test is required above the component/circuit level. As the system design and development progresses, numerous subsystem tests may be required to ensure that the system hardening design is proceeding as planned. For example, subsystem testing of circuit boards may be followed by the testing of a configuration of circuit boards, the testing of interconnected equipments. etc.

20.5 **TEST CRITERIA.** The ultimate criteria for EMR testing is derived from the overall system hardening requirement. Given the characteristics of the intended operational environment, the system designer must determine the overall system hardening requirement by comparing the environment characteristics to the system susceptibility characteristics. Since the system susceptibility characteristics are ultimately dependent upon the susceptibility threshold of the lowest level system components, i.e., discrete components/circuits, the difference between the environmental level and component susceptibility threshold level defines the overall system hardening requirement.

Once the overall system hardening requirement is established, hardening design requirements are assigned to the various "layers" or subsystems of the system in a manner such that the sum of the hardening requirements of individual layers equals or exceeds the overall system hardening requirement. Test criteria for a particular layer are thus defined by the system hardening requirement for that layer. As a simple illustrative example, assume that (1) a system is comprised of a single circuit board within a metallic enclosure, (2) an overall system hardening requirement of 30 dB has been established, and (3) it has been determined from the data in Appendix E and other sources that the optimum hardening approach is to assign a 15 dB hardening requirement to both the enclosure and the circuit board. Test criteria for the enclosure would thus involve shielding effectiveness tests to ensure that 15 dB of isolation was achieved. Test criteria for the circuit board would involve radiated susceptibility tests to establish that the applied design techniques (component selection, lossy sockets, filtering, lead shielding, etc.) provided the required 15 dB of hardening.

Inherent in any test criteria is the specification of test parameters and parameter tolerances to be employed for defining interference susceptibility. Generally, the parameters and parameter tolerances used for susceptibility testing are the same as those which are critical to functional design or operation of the unit under test. The system designer must define the test parameters and tolerances, based on a knowledge of how changes in the parameters would affect the functional operation of the system.

20.6 **TEST PLAN.** The EMR test plan is a contractor prepared document which describes in detail the contractor's EMR test program for assuring compliance with the contractual EMR hardness requirements. In general, the test plan describes the test program organization and objectives, the test facilities to be employed, the types of tests to be performed, the test configurations, techniques, and procedures to be employed, the type and format of measurement data to be recorded, and how the measurement data will be used in the EMR hardness program. The specific contents and details of the plan will depend upon the complexity of the system involved. Generally, however, the contents of the test plan should be as described in Appendix M, and the plan should contain sufficient detail to enable the procuring agency to ascertain that the test program is valid. Approval of the test plan by the procuring agency is normally required prior to proceeding with the system design.

30. **TEST FACILITIES.** The performance of EMR hardness measurements requires the use of specialized test facilities. These facilities include test instrumentation (signal sources, antennas, receivers, etc.), anechoic chambers, various forms of shielded enclosures, and other equipments necessary to establish and measure the effects of a given EMR environment. The purpose of this section is to describe the types of facilities which can be employed in EMR hardness measure-

ments and their basic characteristics and limitations. The manner in which the facilities are employed to perform EMR hardness measurements is further described in Section 40, Test Approach.

The facilities described below are oriented toward those facilities which provide isolation between the EMR test environment and the ambient environment. It is possible to perform either conducted or radiated type EMR hardness measurements in the laboratory or other open-field site where the measurement setup is not isolated from the ambient EM environment. However, the measurement results may be influenced by extraneous signals, and signals radiated from the test setup may interfere with external operations. Furthermore, outdoor open-field measurements which require geographically large sites are expensive and are subject to weather conditions. For these reasons, open-site test facilities are not recommended for performing EMR hardness measurements and consequently will not be addressed in this handbook.

30.1 SHIELDED ENCLOSURES

Shielded enclosures are closed metallic structures which provide a high degree of isolation between the electromagnetic environment within the enclosure and the external environment. Signals present in the environment external to the enclosure will have little effect on measurements made in the enclosure, and signals radiated in the enclosure during testing will not interfere with external operations. Because of the isolation provided, shielded enclosures in various configurations can be used for conducted or radiated EMR hardness measurements. The basic shielded enclosure configurations discussed below are the conventional shielded enclosure, the TEM cell, and the tuned-mode shielded enclosure.

30.1.1 CONVENTIONAL SHIELDED ENCLOSURE

The conventional shielded enclosure is a shielded room whose dimensions are usually multiples of 4 feet (i.e., 8' x 8', 8' x 12', 12' x 20', etc.). These dimensions result from the fact that the enclosures are assembled in 4-foot panel sections. The practical requirements of power access, personnel access, light, heat, and air conditioning result in discontinuities in the enclosure which must be considered a part of the enclosure design.

There are three basic types of shielded enclosures. They include:

- o Single wall,
- o Double wall, electrically isolated, and
- o Double wall, not electrically isolated.

Conventional enclosures are made of either screen or solid metal. Typical screen materials are copper (22 x 22-.015) or bronze (18 x 20-.010). Typical solid sheet materials include copper and 24 gauge galvanized steel, or a combination of copper and steel sheets in double wall enclosures.

The enclosures are assembled in panel sections. The panels are normally 4' x 8'; however, 4' x 10', 4' x 12', and 4' x 14' panels are sometimes used where ceiling heights greater than 8 feet are required. The sections may be bolted, clamped, or welded together to create the room. In the case of non-isolated double-wall enclosures. the inner and outer walls are connected together around the periphery of each panel. In the case of electrically isolated double-wall enclosures, the inner and outer walls are maintained electrically isolated as the panels are assembled.

All power and signal line penetrations into the shielded enclosure are accomplished by filtered wires or by means of coaxial penetrations through the enclosure walls. Each phase conductor and the neutral of power lines are filtered at the entrance to the enclosure. Adequate filters and coaxial penetrations are commercially available from a number of manufacturers. Openings for heating and air conditioning are accomplished with honeycomb barrier panels. The most critical opening in the enclosure is the access door(s). A variety of door sealing techniques are used by the different enclosure manufacturers. The lighting in the enclosure in all cases should be provided by incandescent sources, because other types of lighting sources are prone to produce RF noise.

A summary of typical shielded enclosure performance characteristics is presented in Table H-1. The 120 dB limit usually represents a testing limit and not the limit of the enclosure itself.

TABLE H-1

SUMMARY OF TYPICAL SHIELDED ENCLOSURE PERFORMANCE

Enclosure Type	Wall Type	Material	Magnetic Fields		Electric and Plane Waves	
			60 Hz	15 KHz	1 GHz	10 GHz
Double	Screen	Copper	2 dB	68 dB	120 dB	77 dB
Electrically		Bronze	0 dB	40 dB	110 dB	57 dB
Isolated		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-	Screen	Copper		52 dB	90 dB	
Electrically		Bronze				
Isolated	Solid	24 Ga. Steel		68 dB	90 dB	
		Copper			90 dB	
Single	Screen	Copper	6 dB	42 dB	60 dB	
		Bronze			75 dB	
Shield	Solid	Copper		48 dB	80 dB	

The primary use of the conventional shielded enclosure is to provide isolation when performing conducted susceptibility measurements. In performing conducted measurments, the device, equipment, or system under test and all test instrumentation is located within the shielded enclosure, thus isolating the test environment from the external environment.

Historically, the conventional shielded enclosure has been used to perform radiated susceptibility measurements where the unit under test is illuminated by a source antenna located within the enclosure. However, this type of measurement will not yield valid results since reflections from the enclosure walls prohibit the accurate calibration of the exposure field. Results from experimental investigations have shown that the exposure field level is extremely sensitive to the size and shape of the enclosure, the location of test equipment and personnel within the enclosure, and the spacing between the equipment under test and the radiating antenna, and that field level calibration errors as great as ± 40 dB are possible. Thus the conventional shielded enclosure should not be used as a radiated susceptibility measurements facility, but only for isolation purposes.

30.1.2 **TEM CELL**

The TEM cell is a shielded enclosure in the form of a section of TEM transmission line which can be used to establish a uniform TEM field inside the shielded cell for susceptibility testing. The cell configuration, illustrated in Figure H-1, consists of a section of rectangular TEM mode transmission line inside a shielded enclosure.

The line is tapered at each end to a transition which mates with a standard coaxial cable. A commercially available TEM cell is shown in Figure H-2. The TEM cell provides a means of generating accurately calibrated high intensity RF fields with moderate power RF sources. For example, with an input to the cell of 10 watts, field intensities as high as 200 volts/meter can be generated. Errors in determining the absolute level of the exposure field are within ± 1 dB.

The TEM cell offers several advantages in measuring the susceptibility characteristics of small equipments and devices. It is portable, simple to build, provides isolation of the test environment, and can be used from dc to the lowest order TE mode (TE_{10}) cutoff frequency (f_{c10}) to provide fields from 10 V/m to 500 V/m. The construction cost of TEM cells is lower than the cost for conventional anechoic chambers and shielded enclosures, and it is a relatively accurate (1 to 2 dB) susceptibility test chamber. Because of its TEM mode of operation, the cell has a linear phase response from dc to near f_{c10} and thus can be used for swept frequency measurements.

The major limitation of the TEM cell is the small test volume available at high test frequencies. The cell must be used at frequencies at which only the fundamental TEM modes exist. For example, a cell with 1.2 m x 1.2 m cross section that can be used at frequencies from dc to approximately 150 MHz can accommodate a test specimen approximately 1.2 m x 0.4 m x 0.4 m in size. Smaller cells with comparably reduced test volumes may be used at higher frequencies.

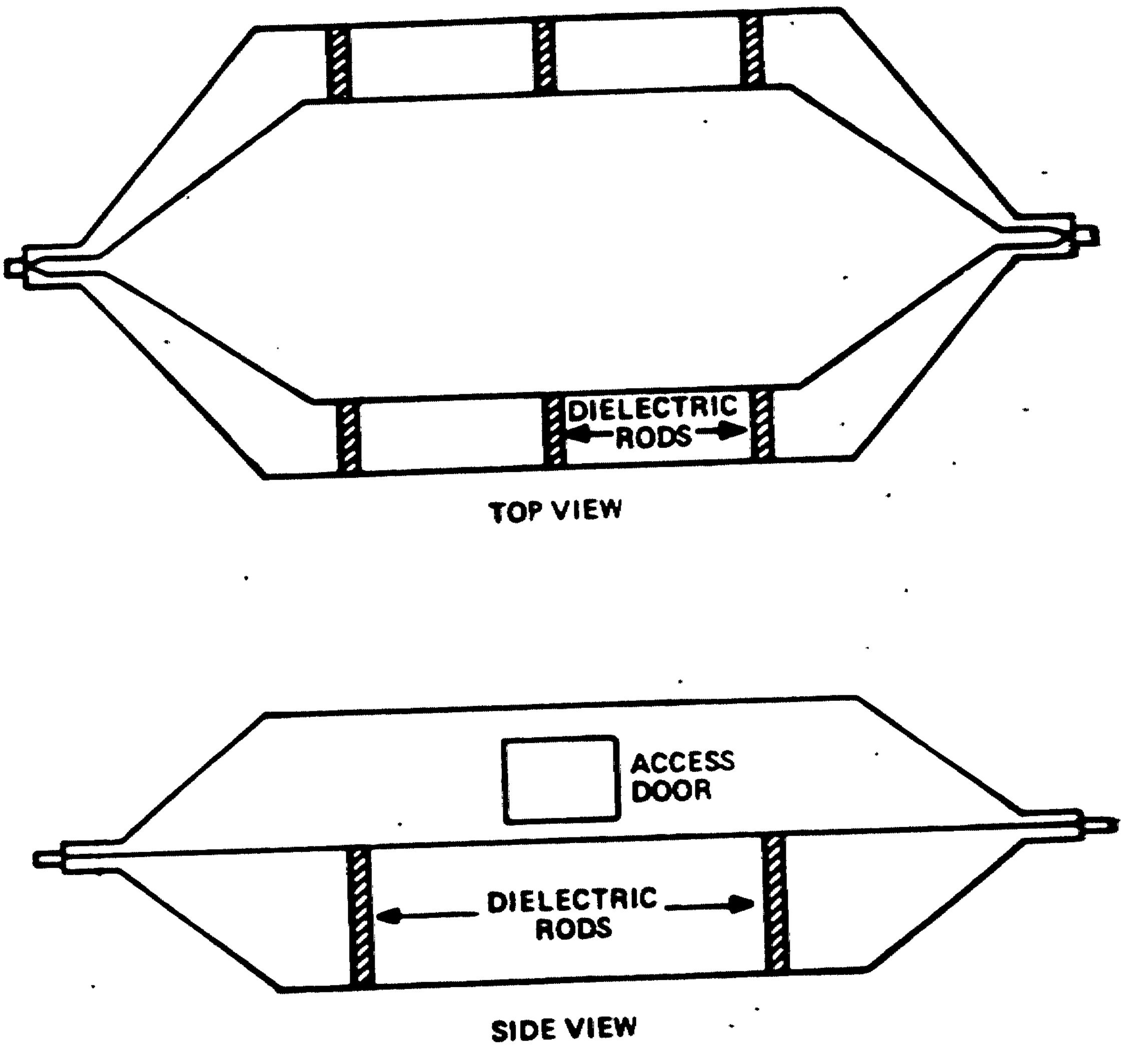


Figure H-1. TEM Cell.

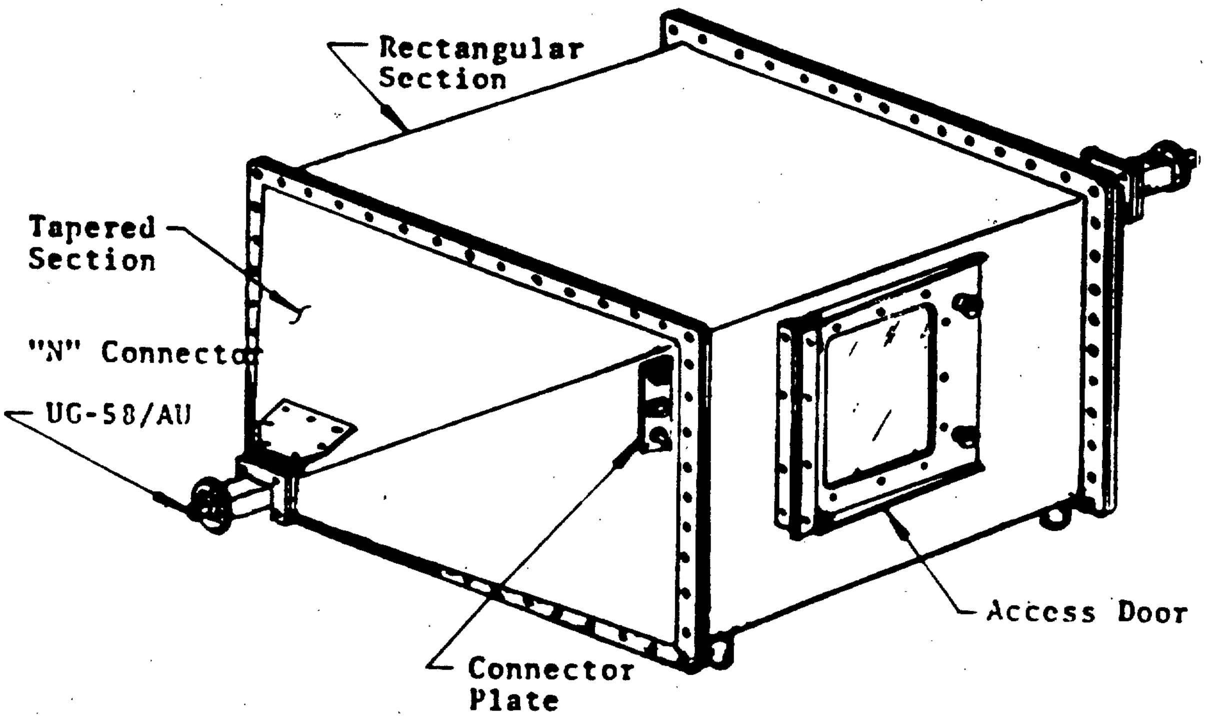


Figure H-2. Configuration of TEM Cell.

30.1.3 **TUNED-MODE SHIELDING ENCLOSURES.** Tuned-mode shielded enclosures which utilize the reflective nature of the enclosure walls can be used to measure the shielding effectiveness of equipments and components. A shielded-enclosure measurement technique which involves mode tuning has been incorporated into MIL-STD-1377 (Navy) as a test method for measuring the shielding effectiveness of cables, connectors, enclosures, and filters over the frequency range of 1 to 10 GHz [H-3]. This technique involves placing the component or equipment to be tested inside a multimoded, tuned shield enclosure in which paddle wheel tuners are used to redistribute the energy within the enclosure.

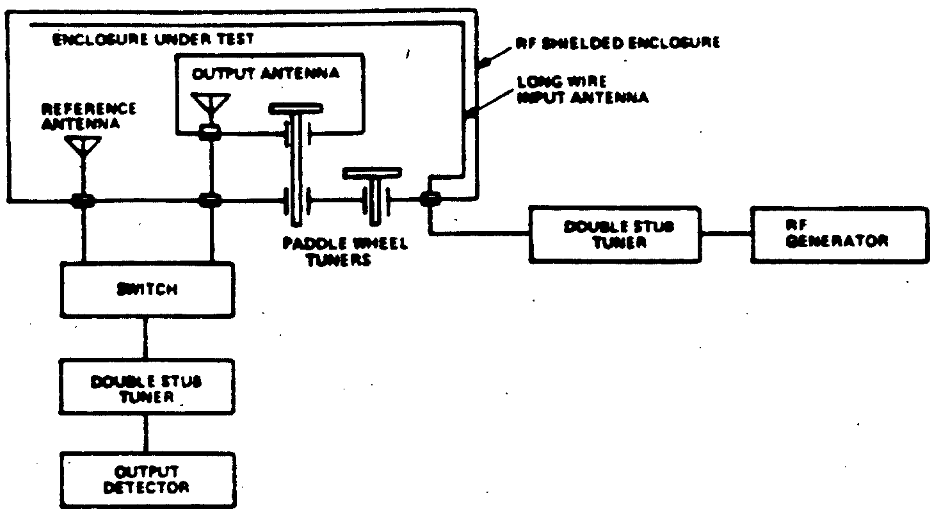
Typical tuned-mode shielded enclosure configurations for performing shielding effectiveness measurements are illustrated in Figure H-3. To determine shielding effectiveness, the equipment to be tested is placed within the enclosure and the enclosure is energized at specific frequencies of interest via the long-wire input antenna. The paddle wheel tuners are then adjusted to maximize the power coupled through the test specimen equipment to the external receiver. Because the equipment under test is subjected to a composite field pattern, both in position and polarity, many local maxima are possible as the tuners are moved. The largest local maxima is indicative of the intrinsic shielding effectiveness of the test specimen equipment. Comparison of this value with the maximum power coupled to the receiver via the unshielded reference antenna yields the shielding effectiveness of the test specimen.

The MIL-STD-1377 type test method has several advantages as a shielding effectiveness test technique. Elaborate test equipment is not necessary, and testing can be performed with an enclosure only slightly larger than the equipment under test. Test techniques are relatively simple, and experience to date indicates that measurements are repeatable within a few decibels. A basic limitation of the technique is that its use is restricted to frequencies above a few hundred megahertz in order to obtain a sufficient number of modes.

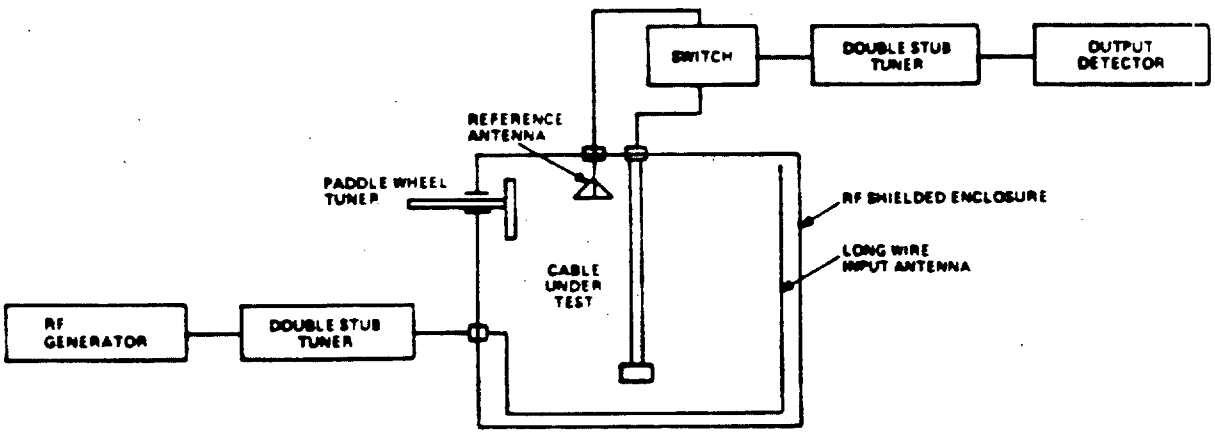
30.2 **PARALLEL-PLATE TRANSMISSION LINES**

Parallel-plate transmission lines can be employed for susceptibility measurements on small equipment items at frequencies ranging from dc up to a few hundred megahertz. The line is configured with two parallel plates, separated by a distance h, which serve as a conductor pair to connect a load to a signal source. The ends of the plates are terminated with matched loads so that the structure operates in the TEM mode. Figure H-4 illustrates a typical configuration for a parallel-plate transmission line.

Major considerations in the use of parallel-plate transmission lines for susceptibility measurements are (1) the field intensity requirements, (2) the size of the equipment to be tested, and (3) the upper frequency range of concern. The maximum field intensity which can be established is dependent upon the spacing between the plate, the output power of the signal source, and the power rating of the matching network and load. Field intensities up to several hundred volts per meter are commonly achieved with parallel-plate structures.



ENCLOSURE SHIELDING EFFECTIVENESS TEST



CABLE SHIELDING EFFECTIVENESS TEST

Figure H-3. Typical Test Setups Using Tuned-Mode Shielded Enclosures.

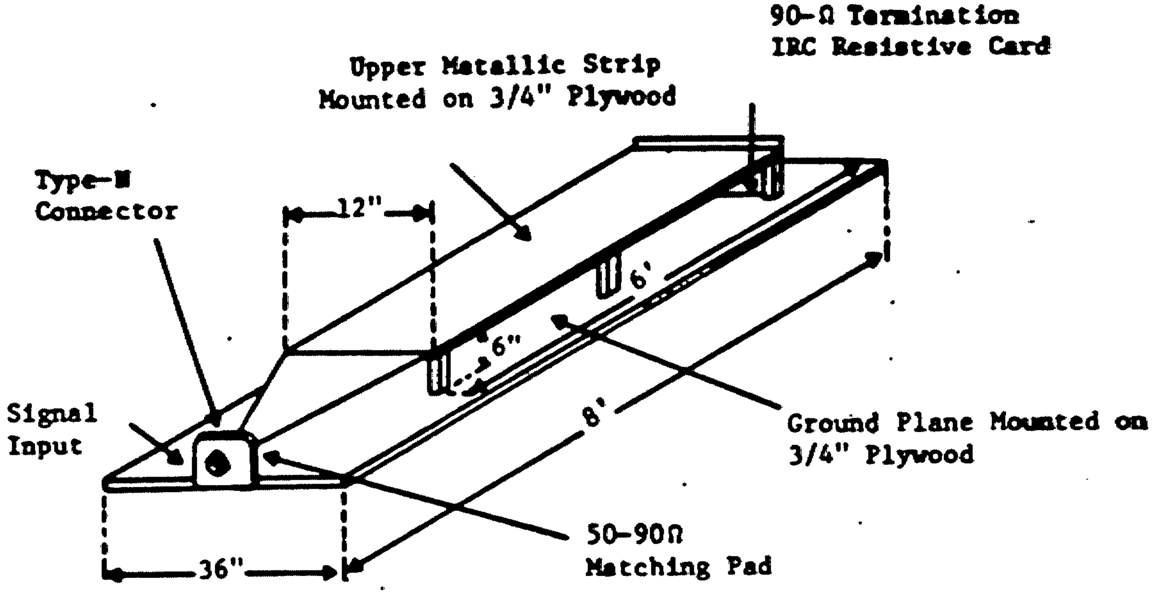


Figure H-4. Parallel-Plate Transmission Line.

The size of the equipment under test is limited by the plate separation distance. The height of the equipment under test should be less than one-third the plate separation to prevent perturbations of the field between the plates. Equipments which exceed this dimension can cause field perturbations sufficiently large to prevent accurate determinations of the field intensity.

The highest frequency at which a parallel-plate structure can be employed is dependent upon the plate separation distance. Theoretically, the structure can be used up to a frequency for which the plate separation distance is equal to one-half wavelength. When the separation distance exceeds one-half wavelength, transmission modes other than the TEM mode can exist and the field will lose its uniformity. In practice, it is difficult to construct a parallel-plate structure which will operate satisfactorily up to the theoretical upper frequency limit. Most structures are thus operated at frequencies considerably below the theoretical limit (typically, less than one-third the upper frequency limit).

Another consideration in the use of a parallel-plate structure is that the structure does not perform as a shielded enclosure. Thus, the structure is exposed to the ambient electromagnetic environment, and the structure itself radiates electromagnetic fields. However, the structure can be located within a conventional shielded enclosure to achieve isolation from the ambient environment.

30.3 SHIELDED ANECHOIC CHAMBER

A shielded anechoic chamber is a shielded enclosure with anechoic material mounted on the inside walls, floor, and ceiling of the enclosure. The anechoic material absorbs the energy which is radiated within the enclosure, and hence, minimizes reflections from the enclosure walls. In good quality anechoic chambers, the levels of reflected energy are 30 to 60 dB below test radiation levels. The shielded anechoic chamber thus provides a nearly free-space test volume which is isolated from the outside environment.

A region inside the anechoic chamber in the vicinity of where the equipment to be tested will be located is designated as the quiet-zone of the chamber. The quiet zone should be located so that the equipment to be tested will be located in the center of the quiet-zone volume; the size of the quiet-zone volume should be sufficiently large to completely encompass the largest piece of equipment to be tested. In the design of the chamber, the size and shape of the chamber and the absorbing materials to be used are selected to satisfy the quiet-zone volume requirements.

RF absorbing materials [H-4] function basically by converting the incident electromagnetic energy which impinges on them into heat. The absorbing materials normally used in anechoic chambers are lossy dielectric materials in the form of an open-cell polyurethane foam impregnated with a carbon-latex solution. (An exception is ferrite absorbing material; however, the cost of ferrite material is approximately $100 per square foot which does not make it economically feasible for anechoic chamber designs.) The material is normally shaped into pyramids, cones, or wedges to provide a gradual transition from free-space into the absorbing material. In order to provide a good air-to-dielectric transition, and hence low reflecivity characteristics, the thickness of the material (height of the pyramids, cones, or wedges) is required to be a minimum of a quarter-wavelength to the lowest frequency of interest. Absorbing materials with thicknesses from 1 to 180 inches are available commercially.

Since wavelength is inversely proportional to frequency, the thickness of the anechoic material must be increased as the low frequency limit of chamber operation is decreased. This requirement limits the low frequency threshold of anechoic chambers because as the thickness of the material increases, (1) the cost of the material increases, (2) the size of the shielded enclosure must increase to retain the same test volume, hence increasing its cost, and (3) it becomes more difficult to mount the material and prevent sagging of the pyramids. For these reasons most existing anechoic chambers have low frequency limits of 200 MHz or higher. A few large chambers have 100 MHz low-frequency thresholds.

The power handling capability of absorbing material is also an important consideration in chambers where tests involving high intensity radiated fields are to be conducted. Absorbing materials having power handling capabilities from 0.15 watt/sq. cm to 31 watts/sq. cm are commercially available.

Generally, the accuracy of measurements made in a shielded anechoic enclosure will be within ± 1 dB. However, the measurement accuracy is dependent upon several test parameters, including:

1) location of the test setup in the chamber;

2) directivity of transmitting and receiving antennas;

3) separation between source and receptor;

4) frequency;

5) reflectivity of the anechoic material; and

6) the magnitude of the energy being measured relative to the maximum energy being radiated in the chamber.

30.4 TEST INSTRUMENTATION

The test instrumentation required for a particular EMR measurement will depend upon such factors as the characteristics of the unit under test, the type of test to be performed, the test technique to be employed, and the frequency range of concern. Generally, a wide array of test instrumentation will be required under an EMR measurement program: thus it is not feasible to identify and describe every test instrument and device which may be needed. The purpose of this section is to identify the more common test instrumentation employed in EMR hardness measurements.

30.4.1 SIGNAL SOURCES

Signal sources are usually required in all EMR hardness measurements to provide test signals which simulate those of the defined EMR environment. In selecting a signal source for a particular measurment, the major source characteristics of concern are frequency, modulation capability, output impedance, and output power level. The source frequency range and modulation capability are chosen to correspond to the characteristics of the defined EMR environment. The output impedance of most test instrumentation, including signal sources is normally 50 ohms. If the source is to be connected to a load which is not 50 ohms, an appropriate matching network must be employed.

The power output requirements of a signal source are dependent upon the level of the EMR environment to be simulated and the type of measurement being performed. For conducted susceptibility measurements, where the signal is injected directly into the unit under test, the source power output necessary to reach a susceptibility threshold will be relatively low. Generally, conducted susceptibility measurements can be performed at power levels not exceeding a few watts.

For radiated susceptibility measurements, the source power output requirements can increase significantly, depending upon such factors as the level of the radiated test environment to be simulated, the characteristics of the transmitting antenna employed, spacing requirements between the transmitting antenna and the unit under test to realize a far-field test condition, and power losses in ancillary test equipment. To illustrate the source power qutput requirements for a typical radiated measurement, assume that the unit under test is to be exposed to a 200 V/m field intensity, and that a three-meter spacing between the transmitting antenna and the unit under test is required to satisfy far-field testing conditions. The RF power required to establish a desired field intensity can be determined by means of Equation H-1:

$$P_{RF} = \frac{4\pi R^2 P_D}{G_A} \tag{H-1}$$

where:

P_{RF} = RF power required,

R = distance between radiation antenna and field intensity reference point (meters),

P_D = field intensity in power density terms (W/m^2), and

G_A = gain of radiating antenna as a power ratio (i.e., 3 dB = 2, 6 dB = 4, etc.).

The 200 V/m field intensity requirement can be converted to the power density form of Equation H-1 by means of Equation H-2.

$$P_D = \frac{(FI)^2}{120\pi} \tag{H-2}$$

$$P_D = \frac{(200\ V/m)^2}{120\pi} = 106.1\ W/m^2$$

Substituting 106.1 W/m^2 for P_D and 3 meters for R in Equation H-1 yields:

$$P_{RF} = \frac{4\pi(3)^2(106.1)}{G_A}$$

$$= \frac{12000}{G_A} \text{ watts}$$

Assuming an antenna gain of 6 dB (G_A = 6 dB = 4),

$$P_{RF} = \frac{12000}{4} = 3000 \text{ watts}$$

or assuming an antenna gain of 15 dB (G_A = 15 dB = 31.62).

$$P_{RF} = \frac{12000}{31.62} = 380 \text{ watts}$$

These results indicate that to establish a 200 V/m field intensity at three meters from a radiating antenna having a 6 dB gain requires an RF source capable of providing approximately 3 kilowtts of RF power, while to establish the same level field with a radiating antenna having a 15 dB gain requires an RF source capable of providing approximately 380 watts of RF power. It is apparent from these results that the power output requirements for the RF signal sources can be significant, and are heavily influenced by the characteristics of the radiating antennas, as well as the spacing between the radiating antenna and the test specimen. Conventional signal generators and sweepers will not provide sufficient RF power to perform radiated high power interference susceptibility tests.

The RF signal source requirements may be satisfied with either a series of tunable power oscillators or lower-power signal generators and/or sweepers used in conjunction with broadband RF power amplifiers to provide RF power output levels sufficient to perform the required susceptibility tests.

30.4.2 **ANTENNAS**

The selection of antennas and exposure chambers to be used to establish high intensity fields for radiated susceptibility tests is an important consideration in the development of the testing facilities. As illustrated in the previous section on signal sources, the characteristics of the radiating antenna have a significant influence on the amount of RF power required to perform the tests. For each 3-dB increase in antenna gain, the RF power requirement is reduced by 1/2. On the other hand, the radiating antenna gain must be limited to the extent that the width of the main beam is sufficient to adequately illuminate the test specimen. Further, it is desirable that the main beam illumination be such that the amplitude and phase distibutions across the test specimen are essentially planar. Thus, the selection of a radiating antenna becomes a compromise between obtaining the maximum gain possible to minimize the RF power requirements, and at the same time, obtaining sufficiently broad beamwidth characteristics (sufficiently low gain) to properly illuminate the test specimen.

Since susceptibility tests are currently required over the 50 MHz to 18 GHz frequency range (a spectrum width of over eight octaves), a number of antennas representing several antenna types are required for the tests. The low frequency end of the spectrum presents the most difficult problem as far as selecting a suitable radiating antenna is concerned. In the 50 to 200 MHz frequency range, antennas having sufficient gain characteristics are physically too large to fit inside normal size shielded enclosures, or if they will fit inside, the walls are so close to the extremes of the antenna structure that they significantly affect the characteristics of the antenna. This situation is true for tuned dipole. log-periodic, log-conical. and tuned vertical antennas. Hence, in this frequency range, non-resonant short dipole, biconical, or short vertical whip antennas are normally used as radiators. These antennas have very low gain characteristics, and hence, relatively large RF power levels are required to generate moderate field intensity levels. For example, a typical biconical dipole antenna used as a standard in EMC measurements has a gain of approximately 0.15 dB relative to an isotropic source at 50 MHz. To establish a 200 V/m field one meter from the antenna would require approximately 1.285 watts of RF power. At a frequency above 200 MHz where a log-periodic antenna with a 6 dB gain could be used, only 350 watts of RF power would be required. At a still higher frequency where a standard gain horn antenna with a gain of 15 dB could be used. only 50 watts of RF power would be required. On the other hand, over the 50 to 200 MHz frequency range, high power broadband RF amplifiers are readily available and large RF power requirements are not as much of a problem or as costly as they are in the higher frequency ranges.

Over the 200 to 1000 MHz frequency range, resonant antennas such as the tuned half-wave dipole, tuned vertical whip, log-periodic, and log-conical antennas are sufficiently small to fit inside normal size shielded enclosures. The log-periodic and log-conical are the preferred antenna types for this frequency range because a single antenna can cover the entire range without any tuning required. In addition, these antenna types exhibit gain characteristics in the range of 5 to 6 dB which significantly reduce the RF power requirements compared to the 50-200 MHz frequency range.

Over the 1 GHz to 18 GHz frequency range, the size of the radiating antenna is no longer a problem and several types of antennas can be used. These include log-periodic, log-conical, cavity-backed spiral, standard-gain horn, and ridged-guide horn antennas. The preferred radiating antennas for this frequency range are waveguide horns due to their good unidirectional, high gain, and VSWR characteristics. The entire frequency range (1-18 GHz) can be covered with two ridged waveguide horns with gains from 6 to 12 dB. The frequency range can also be covered with seven standard gain horn antennas with gains from 14 to 24 dB.

30.4.3 FIELD INTENSITY METERS

Sensitive RF detectors with provisions for accurate calibration in terms of standard EMI measurement units are required to perform the tests described in this chapter. For measurements of the EM environment, calibrated detectors are required to define the field intensity levels over the frequency spectrum of interest. For susceptibility measurements, accurate detectors are required to calibrate the fields to which the test specimen is to be exposed and to detect the response of the test specimen when exposed to the test fields.

These detector requirements for the tests can be satisfied by two types of instruments, EMI receivers and spectrum analyzers. Both of these instruments are well-shielded, sensitive receivers with wide dynamic ranges and provisions for calibration in standard EM measurement units. Instruments in both categories are available to cover the frequency spectrum from 20 Hz to 40 GHz.

30.4.3.1 EMI RECEIVERS

EMI receivers are basically superheterodyne receivers with special circuits and parameters to enhance the accurate measurement of EMI emission characteristics of equipment and systems. Specific design parameters and some operational techniques may vary depending upon the portion of the frequency range being covered. However, the basic principles of field intensity meters are the same over the frequency spectrum presently covered (20 Hz to 40 GHz).

Most EMI receivers include the following characteristics:

a) Have an extremely wide frequency tuning range so that a single instrument can be used over as much of the frequency spectrum as possible. The frequency tuning accuracy is normally 1 percent or better.

b) Accept a wide range of input signal levels. The actual dynamic range of the receiver is normally 20 to 60 dB. Additional input signal range is obtained with a set of wideband, calibrated attenuators at the input of the meter. Most meters will accept input signals in the range from -100 dBm to +20 dBm.

c) Present a constant real impedance (normally 50 ohms) at the input terminal so that the meter properly terminates the transmission line and also so that voltage levels at the input can be specified in terms of power.

d) Have selectable, calibrated IF bandwidths. The narrow IF bandwidths make it possible to obtain maximum sensitivity for narrowband signals and the wider IF bandwidths make it possible to encompass more of the spectrum of wider band signals. The different bandwidths make it possible to distinguish between narrowband and broadband signals.

e) Have internal sources for calibrating the gain and IF bandwidth of the meter at the frequency of interest. The impulse generator is the most common type calibrator in current EMI receivers.

f) Have an output meter and different detector and metering circuits which make it possible to measure signals having different types of modulation.

A number of EMI receivers are available from several manufacturers which cover the frequency range from 20 Hz to 18 GHz. Generally, three instruments are required to cover this frequency range. In some cases, several plug-in RF heads are also required. Converter units are available which make it possible to perform measurements up to 40 GHz. Table H-2 shows some typical characteristics of currently available EMI receivers.

TABLE H-2

TYPICAL CHARACTERISTICS OF CURRENT EMI RECEIVERS

Frequency Range	Number of Bands	Sensitivity	Measurement Range
20 Hz to 15 kHz	1	-127 to -150 dBm	126 dB
14 kHz to 1 GHz	15	-85 to -120 dBm	120 dB
1 GHz to 18 GHz	5	-79 to -90 dBm	120 dB

30.4.3.2 **SPECTRUM ANALYZERS**

Modern spectrum analyzers can also be used as field intensity meters. When compared to EMI receivers, spectrum analyzers are characterized by untuned front ends, built-in automatic frequency sweeping and dispersion capabilities, a larger choice of IF bandwidths, and CRT output displays with variable persistence. Accurate amplitude calibrations of spectrum analyzers are more difficult than calibrations of EMI receivers, particularly for peak amplitudes of pulsed signals. The advantage of spectrum analyzers over EMI receivers are their flexibility in frequency sweeping, the availability of a spectrum output display, and lower cost for equivalent frequency coverage.

Within the past 5 to 10 years, significant progress has been made in the spectrum analyzer technology. This progress has resulted in the addition of several important functions to the current generation of spectrum analyzers. The functions include variable persistence/storage CRT displays, internal calibration capabilites, wide spectrum scans (several GHz), flat frequency response, and significantly increased dynamic ranges.

Several spectrum analyzers which cover the frequency range from 10 MHz to 18 GHz are availble from a number of manufacturers. In most cases, a single instrument is capable of covering the entire frequency range. External tracking preselectors are available to provide RF preselection where required. In addition, external mixers and converter units are available which make it possible to perform measurements up to 40 or even 90 GHz. There is such a broad range of characteristics available and such rapid advances in the state-of-the-art, it is recommended that suppliers or manufacturers of spectrum analyzers be contacted to establish the present state-of-the-art in these instruments.

30.4.4 AUXILIARY DEVICES

A large number of auxiliary devices are required to interconnect the various test instruments into the required test setups, and, at the same time, assure that the test specimen and each of the test instruments will function in the desired manner. A listing of typical auxiliary devices is shown in Table H-3.

The types and specific properties of the auxiliary devices required for the tests are dependent to a large extent on the characteristics of the tests, the specific types of test instruments to be used, and the frequency range over which the tests are to be performed. Thus, it is not feasible to attempt to describe the characteristics of every device which might be needed under all these conditions. In selecting auxiliary devices, the frequency range (bandwidth), impedance, and power rating of the devices must be compatible with the test specimen, test instruments, and other components in the test setup. The principal requirement is that the incorporation of auxiliary devices into a measurement setup not degrade the accuracy or validity of the measurement results.

TABLE H-3

TYPICAL AUXILIARY DEVICES FOR HIGH POWER SUSCEPTIBILITY TESTS

Filters	Switches
Directional Couplers	Coaxial Cables
Signal Samplers	Coaxial Connectors
Attenuators	Coaxial Adapters
Isolators	Waveguide
Hybrid Junctions	Waveguide Hardware
Dummy Loads	Isolation Transformers

As a general rule, double-shielded, 50-ohm, coaxial cable should be used for interconnecting cables. The double shield minimizes the effects of interconnecting cable leakage on the measurement results, and the 50-ohm impedance is compatible with the majority of test instruments and auxiliary devices.

40. **TEST APPROACH.**

40.1 **COMPONENT/CIRCUIT SUSCEPTIBILITY MEASUREMENTS.**

40.1.1 **GENERAL CONSIDERATIONS.** Ultimately, it is some disturbance of the operation of a circuit or component that will cause system degradation or failure. Thus susceptibility information on the solid-state components used in a system is essential to the design and development of a system which is compatible with a specified EMR environment. The susceptibility data presently available on both discrete and integrated circuit components are presented in Appendix E. However, measurements may be necessary to obtain the required data on many existing devices, as well as all future devices.

40.1.2 **IDENTIFICATION OF CRITICAL COMPONENTS/CIRCUITS.** Many of the components and circuits used in an air launched ordnance system may be relatively immune to the coupled EMR environment. Attempting to define the susceptibility levels of all system components would thus not be a cost effective approach to determining system hardening requirements. The system designer must identify those components and circuits which are critical in terms of their potential susceptibility to the EMR environment, and where susceptibility data are not available, implement a measurement program to determine component/circuit susceptibility.

Only the system designer can identify the specific critical circuits of a system, through a knowledge of the circuit function and an assessment of how this function might be affected by coupled energy. However, some general guidelines are presented below which should aid the designer in this identification process.

When undesired energy is coupled to a system circuit, an upset in the circuit operation may occur which is reflected throughout the system. An upset is normally considered as any response other than the prescribed response. These upsets are usually viewed as an undesired change in signal characteristics. These include voltage and current levels, transmission or reception levels, or improper control of other related systems. Upsets may be categorized as those in digital (nonlinear) circuits and those in analog (linear) circuits. Table H-4 gives examples of the most frequently observed system effects and some of the circuit types in which upsets occur.

A systematic categorization of circuits in terms of parameters which influence circuit susceptibility can also be helpful in the identification of critical circuits. One method of performing this categorization is shown in the histograms of Figure H-5. Note from this figure that the first step of the circuit categorization is to order the circuits in accordance with the coupled RF environment as in Figure H-5a. If the coupling analyses show that a servo circuit will be exposed

TABLE H-4

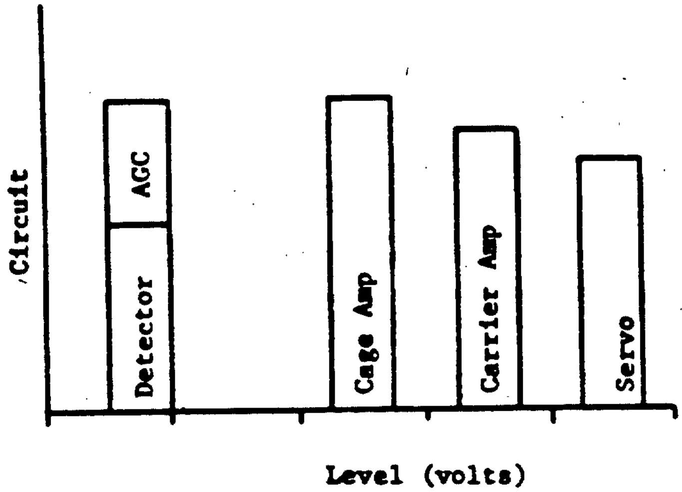
CIRCUIT UPSETS

I. DIGITAL (NONLINEAR) CIRCUIT UPSETS

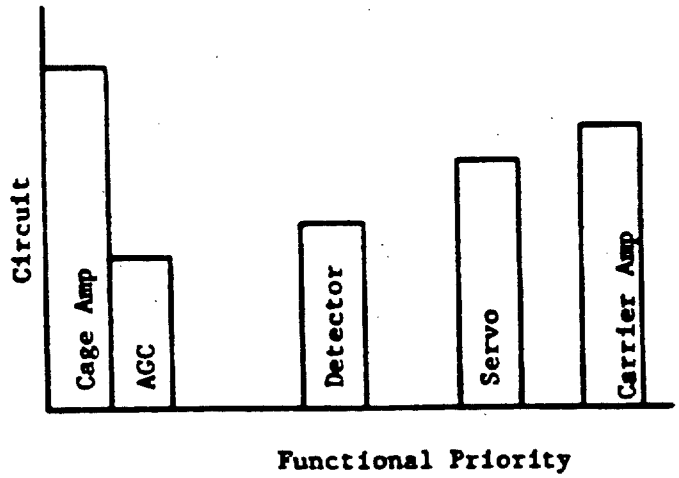
Functional Effect	Produced by upset of:
1. Changes in command functions	Flip-flops, discrete circuits
2. Alteration of stored information	Registers, memory elements
3. Changes in system timing	Clocks, counters, oscillators
4. Changes in drive states	Logic buffers, interface circuits
5. Changes in processing states registers	Non-synchronous logic

II. ANALOG (LINEAR) CIRCUIT UPSETS

Functional Effect	Produced by upset of:
1. Changes in signal level	Amplifiers - ac/dc
2. Loss of regulation	Voltage and current sources
3. Loss of synchronous states	Phase-lock circuits
4 Control instability	Servomechanisms
5. Loss of information in process	Demodulators, choppers, amplifiers
6. Premature activation/ function loss of protective circuits	Power supply "crowbars" and current limiting circuits
7. Loss of signal generation	Sinusoidal oscillators, dc-dc inverters, dc-ac converters
8. Improper wave shapes	Active, passive filter networks



(a) Coupled RF Environment Levels.



(b) Circuit Functional Priority.

Figure H-5. Circuit Categorization.

to the highest RF environment. the servo circuit is represented at the highest position on the histogram. To provide a firmer basis for judgement as to the effort that should be expended in testing a circuit. a different ordering of the circuits is used. In this case, as shown in Figure H-5b. a functional priority from the standpoint of mission success should be assigned to each and plotted. Determining the functional priority is a much more difficult task since an appreciation of the overall. detailed, integrated-system operation must be available. With a systematic approach. the critical circuits can be identified and intelligent cost trade-offs as to the necessary testing can be greatly facilitated. from both the suceptibility and functional priority points of view.

The circuits that should be considered in the histogram analysis can be identified according to the following criteria:

- o The circuits that are at terminations of external cables.
- o The circuits that are sensitive devices.
- o The circuits that have large gain-bandwidth products.
- o The circuits that perform critical system functions.
- o The circuits that are connected to cable routes that might have great sensitivity to pickup.

Coupling to low level analog circuits is especially critical. Such coupling is illustrated in Figure H-6. RF energy is coupled through three ports-of-entry (POE). Coupling through POE 1 to the control circuitry would not be critical since the control signal amplitude would be large compared to any induced RF voltages. However. coupling to the "front-end" amplifier through POE 2 is critical since rectified waveforms would appear as legitimate signals and be amplified and thus create disturbances throughout the system electronics. Likewise, the coupling from POE 3 is critical since the connecting cables can couple the RF energy to the "front-end" amplifier. These types of observations should be made in the design stage so that the critical circuits can be identified and tested.

40.1.3 **MEASUREMENT TECHNIQUES.**

40.1.3.1 **INJECTION TESTING.** Injected susceptibility measurements are made by conductively coupling the RF energy to the device under test. Provisions are made for biasing the devices as well as for monitoring the device's operating parameters such as voltages and currents. This approach removes the variables associated with coupling from a free-field and permits concentration on device behavior under the influence of the RF energy. This experimental arrangement is realistic because individual circuits are usually fairly well shielded in systems (i.e., protected from radiated RF energy) but have cable or wires running between various circuits or subsystems. These wires or cables can act as inadvertent antennas and couple RF energy from field to device.

Device response is measured as a function of the absorbed RF power. The reasons for treating power as the independent variable are as follows:

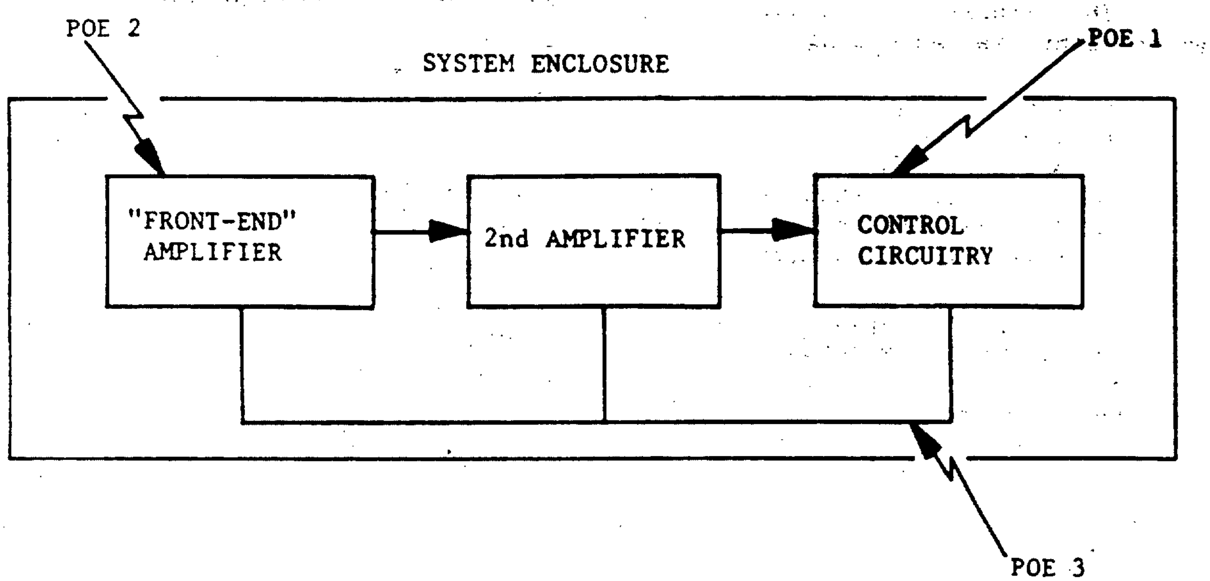


Figure H-6. Coupling to Low-Level Analog Circuits.

(a) At microwave frequencies, power and impedance are the fundamental, observable quantities; voltages and current are derived from power and impedance. At low frequencies, the converse is true.

(b) The semiconductor junction acts as a square-law detector (induced-current change is proportional to absorbed power). This square-law dependence holds for absorbed power levels up to the order of 100 mW or so, above which the dependence gradually diminishes to a region where the interference effect is proportional to the square root of P_A. For RF-induced catastrophic failures, the failure level is a function of absorbed power and time.

Placement of the device under test in a test fixture which can be well defined with respect to fixture losses is essential to making an accurate measurement of absorbed power. The validity of the technique rests upon the principle of conservation of energy (or, more precisely, upon its time derivative, i.e., power). It should be noted that the technique which is described here does not yield a measurement of device complex impedance, since no phase information is given. The complex impedance must be measured separately by using conventional devices, i.e., slotted line, impedance bridge, network analyzer, etc. However, the power measurement technique does yield a measurement of the reflection coefficient, which is related to the standing wave ratio and impedance magnitude.

The basic measurement technique for components/circuits is illustrated in Figure H-7. A high power RF source is used to generate the necessary RF energy. The RF source might consist of a signal generator, a low power amplifier, a variable attenuator, and a high power amplifier, if needed. The RF source is decoupled through use of an isolator. Following the RF source is a dual directional coupler for monitoring the incident power (P_I) and the reflected power (P_R). The circuit input signal (DC, video, or audio) is combined with the RF energy in a bias injection unit. The purpose of this unit is to decouple the RF from the bias sources and monitoring instrumentation. A bias injection unit which has been designed for use in performing component susceptibility measurements is illustrated in Figure H-8 [H-23]. The unit is usable over the frequency range of 100 MHz to 12 GHz and will pass video pulses having a rise time of 80 nanoseconds. At lower frequencies, a bias insertion unit could be designed using discrete circuit elements (capacitors and inductors).

For discrete devices (i.e., transistors) the device under test is mounted in a commercial 50 ohm test fixture. The bias injection unit must be coupled to this fixture through an appropriate matching network. At microwave frequencies this network must be in the form of a stripline transition or launcher. The specific configuration of this transisition will depend upon the circuit under test.

The operation of the device under test is monitored through the low-frequency arm of the bias network while RF energy is injected.

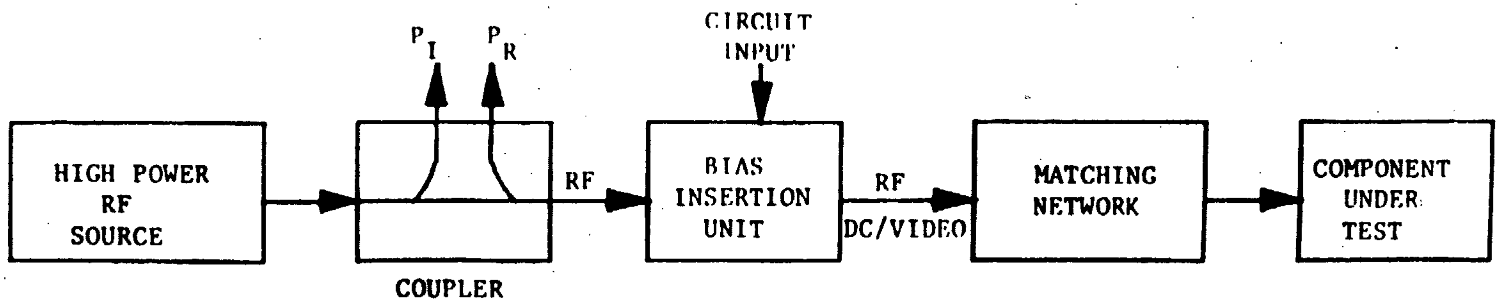


Figure H-7. Basic Test Configuration for Component Susceptibility Measurements.

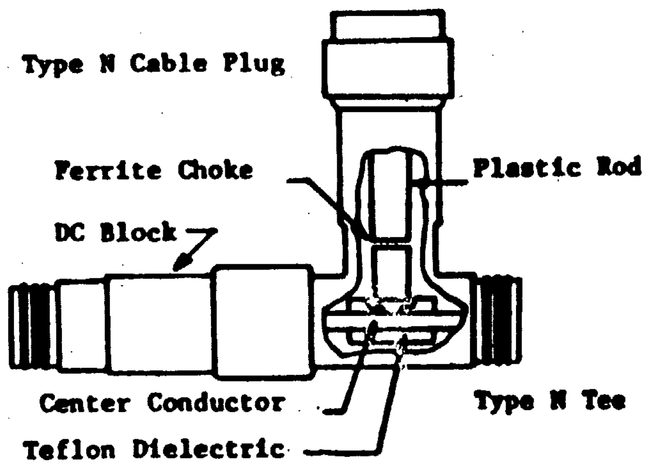


Figure H-8. Bias Insertion Unit. [H-23]

Taking samples of the incident, reflected, and transmitted power, and using the calibration factors for coupler, bias networks, and detectors permits a determination of the device susceptibility (i.e., a change in output voltage, etc.) as a function of the power absorbed in the device under test.

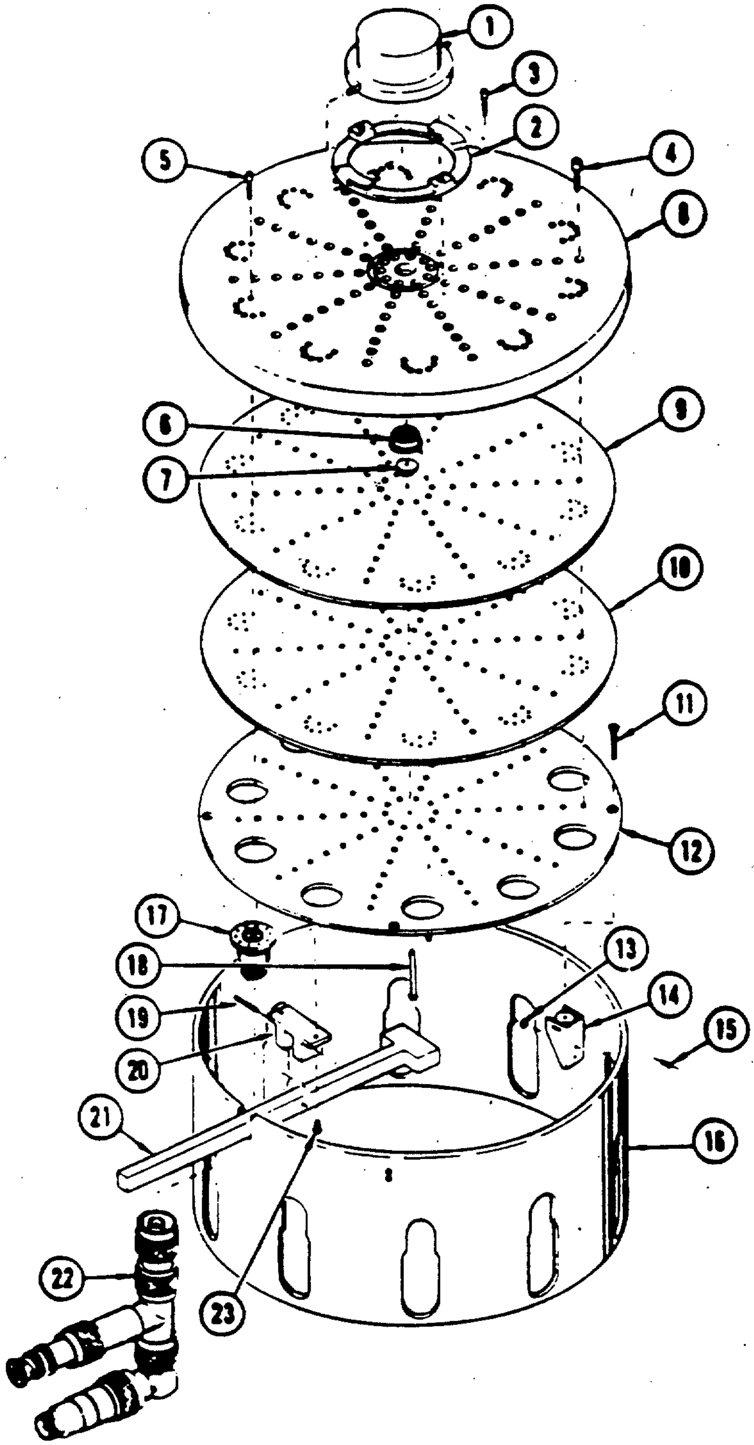
Susceptibility measurments on integrated circuits (IC's) involve the same basic measurement technique used for transistors. The transistor is treated as a 2-port device for susceptibility measurement (one terminal of the transistor is at an RF ground). However, a small-scale integrated circuit may have from 8 to 16 terminals, and commercial RF test fixtures are not available. In addition, different IC package styles will require different fixtures. An exploded view of a test fixture developed for IC susceptibility testing is shown in Figure H-9. In a manner similar to that for discrete devices, a relationship is established between the absorbed power and: (1) incident power; (2) reflected (at entry terminal) power; and (3) transmitted power, at each of the terminals other than the entry terminal. The relationship takes into account losses arising from the test fixture, bias networks (one required per IC pin), and calibration factors for the crystal detector.

A large number of parameters must be measured and recorded. For example, a single series in a 16-pin IC will consist of 17 power measurements in addition to a voltage and current measurement at each pin, for a total of 49 parameter values. When measurements must be made at a number of frequencies, power levels, and operating conditions, manual data acquisition becomes burdensome. Therefore, an automated data acquisition system (Figure H-10) has been developed for use with the IC measurements. A minicomputer is used to control test parameters, process and record data, and perform "on-line" analysis of results. Figure H-11 shows data taken with and plotted by the automated data acquisition system for one device at a single RF frequency. The data shows the device going from an "output low" state (0.4 volts) to an "output high" state (2.4 volts) under the influence of RF energy. Since the data are also stored on magnetic tape, other parameters may be plotted as a function of power (or any other variable) at a later time and statistical analyses performed without the necessity of retaking the data.

40.1.3.2 **RADIATED TESTING.** Measurements of the susceptibility of discrete components/circuits require the use of special test fixtures for mounting the components/circuits under tests, for injecting test signals (desired and interference signals), and for monitoring susceptibility characteristics. Radiated testing would involve the exposure of these fixtures as well as the components/circuits under test; thus it would be extremely difficult to differentiate between the effects of the test fixture and the susceptibility characteristics of the components/circuits under test. For this reason, radiated testing of discrete components/circuits is not a viable test approach.

40.2 **SUBSYSTEM SUSCEPTIBILITY TESTS.**

40.2.1 **GENERAL CONSIDERATIONS.** For the purpose of this handbook, a subsystem is defined as a configuration of components and circuits



PARTS LIST		
ITEM NO.	NAME	NUMBER REQUIRED
1	COVER PLUG	1
2	SPRING LOADED HOLD DOWN	1
3	SCREW 4-40 X 3/4	2
4	SCREW 6-32 X 1/2	80
5	SCREW 2-56 X 1/2	70
6	TO-5 TEST SOCKET	1
7	TO-5 SOCKET	1
8	TOP COVER PLATE (TO-5)	1
9	TOP STRIPLINE BOARD (TO-5)	1
10	BOTTOM STRIPLINE BOARD (TO-5)	1
11	SCREW 6-32 X 1/2	4
12	BOTTOM COVER PLATE (TO-5)	1
13	NUT 2-56	8
14	FASTENER	4
15	SCREW 2-56 X 1/4	8
16	TEST FIXTURE SKIRT	1
17	STRIPLINE LAUNCHER	10
18	EJECTOR PIN	1
19	EJECTOR PIVOT PIN	1
20	EJECTOR PIVOT	1
21	EJECTOR ARM	1
22	BIAS UNIT	10
23	SCREW 4-40 X 1/8	2

Figure H-9. Exploded View of IC Suscptibility Test Fixture. [H-23]

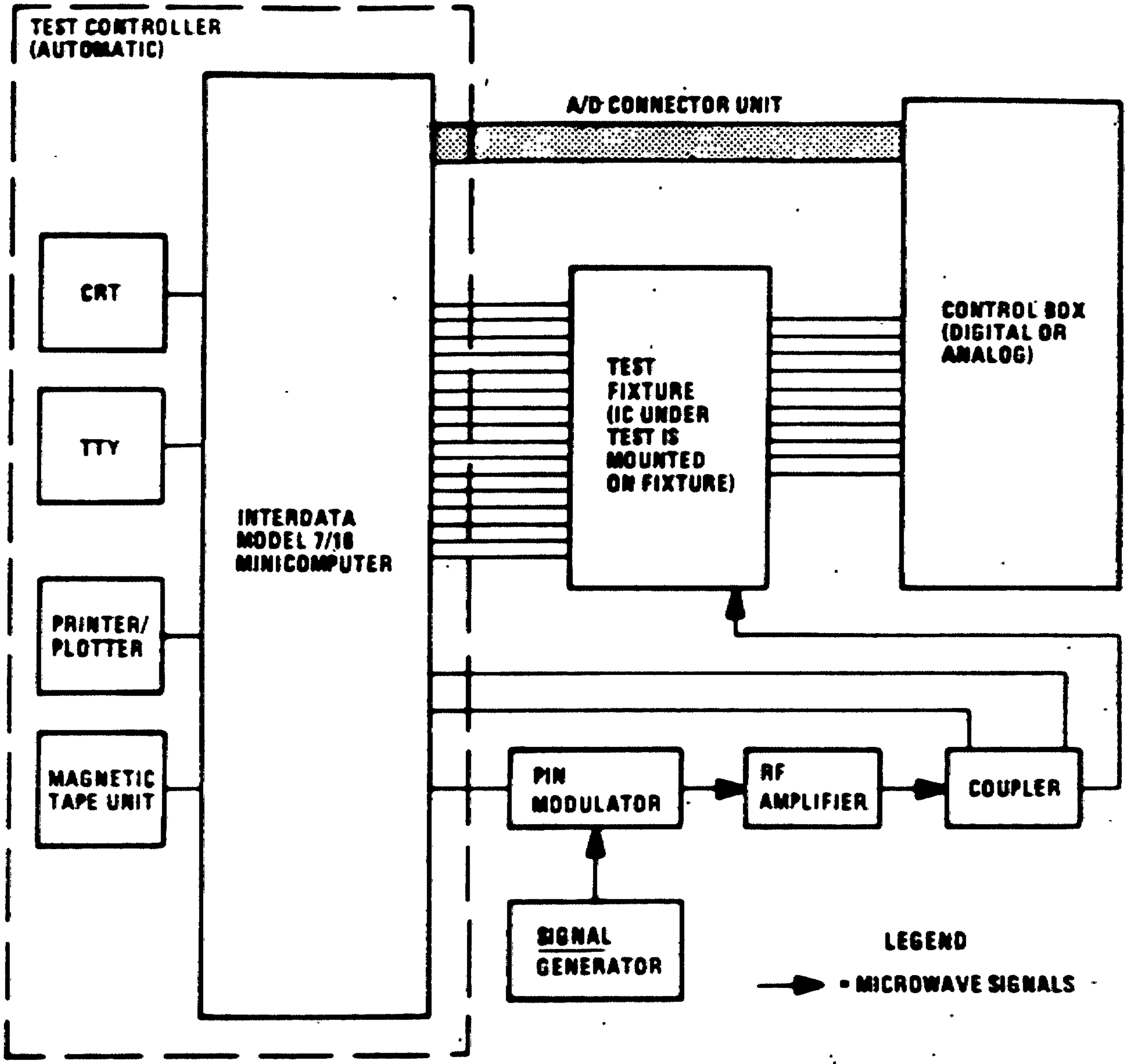


Figure H-10. Schematic of Automated Data Acquisition System. [H-23]

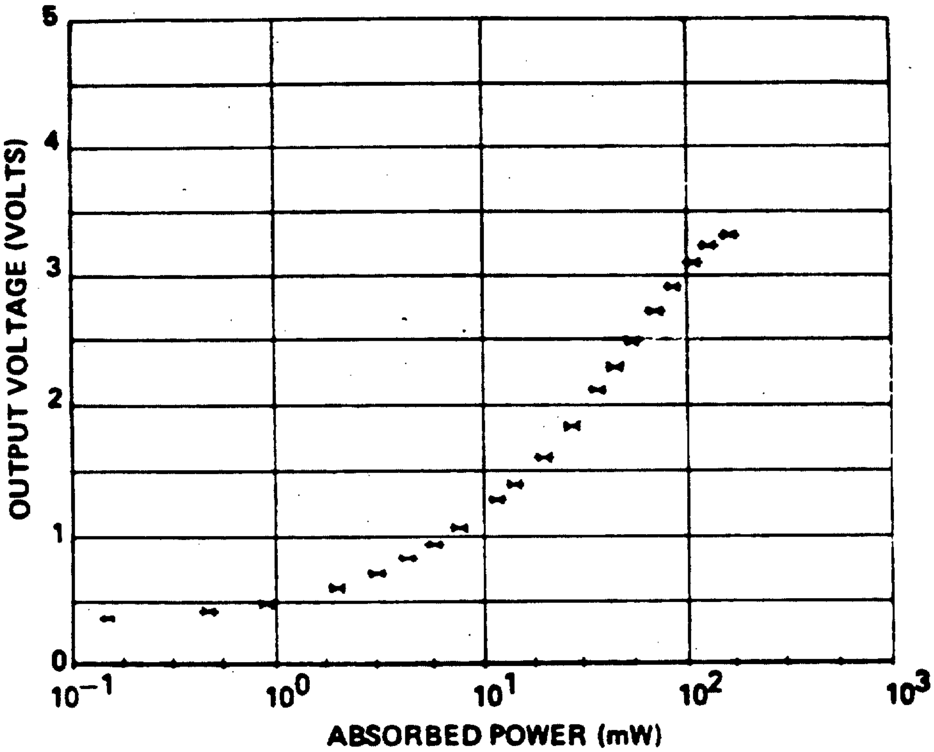


Figure H-11. Interference Effect (Change of Output Voltage) on a 7400 NAND Gate as a Function of RF Power Level. [H-23] which are connected together in a subassembly, performs a specific function with given input/output signal requirements, and does not require a special test fixture for susceptibility tests. The prime reason for this definition is to broadly categorize units to be tested as those which require specialized test fixtures and techniques (discrete components/circuits) and those which do not. Under this definition, individual or interconnected configurations of circuit boards, modules, or equipments could comprise a subassembly.

As subsystems are developed, it is highly desirable to perform tests to determine the susceptibility characteristics of the systems in order to evaluate wiring, bonding, and grounding techniques as well as to evaluate the composite susceptibility characteristics of all the components of a particular subsystem when operating in their normal configuration and environment.

In contrast to discrete component/circuit susceptibility measurements, EMR susceptibility measurements on subsystems may be performed either by injection testing or radiation testing. Generally, however, injection testing is better suited to testing of lower level (i.e., smaller, simpler, less complex) subsystems whereas radiated testing should be used for higher level subsystems. One reason for this approach is that low level subsystems (e.g., modules, circuit boards, etc.) tend to be located at points within a system which are "remote" from the external EMR environment (e.g., a circuit board may be separated from the external environment by two or more shielded enclosures), and EMR signals which reach the circuit board are more likely to be coupled (conducted) via interconnecting wiring than by radiation. Thus, injection tests should provide a valid description of EMR effects. A second reason involves the complexity of injection versus radiated tests. Generally, injection testing is simpler than radiated testing in terms of facility requirements and should be used as a test method whenever feasible.

40.2.2 **IDENTIFICATION OF CRITICAL SUBSYSTEMS.** The techniques used to identify the critical subsystems are the same as used at the circuit level. In fact, a critical circuit in a subsystem causes the subsystem to be critical. However, because of the increased complexity, a subsystem may have more than one coupling penetration and a number of critical parameters which cause degradation. Subsystems usually have design parameter specifications and susceptibility degradations can be defined in terms of such specifications.

40.2.3 **MEASUREMENT TECHNIQUES.**

40.2.3.1 **INJECTION TESTING.** Instrumentation for injecting RF energy into a subsystem uses the same basic techniques as for circuit injection testing. However, the test fixtures for subsystems will differ considerably from the fixtures used in component/circuit injection. As for component/circuit injection, a problem arises in measuring accurately the injected RF energy. This is particularly true when the injection must be on a lead or wire in a cable bundle. When the subsystem has a coaxial input, the injection is relatively simple. When the subsystems are connected together through a cable bundle a technique for injecting microwave RF energy into the appropriate cables will have to be developed.

If the subsystems are shielded, coupling through the connecting cables will be the dominant coupling mode. For unshielded subsystems, coupling may occur over more than one path and determination of the best injection technique may be difficult.

40.2.3.2 **RADIATION TESTING.** Radiation testing of subsystems is cost effective in the early design stage because shielding weaknesses can be readily identified. Radiation testing also evaluates cable shields and compartmentalization techniques. Establishing good ground connections with RF gaskets, filters, etc. can be a problem at microwave frequencies and radiation testing is necessary to prove the RF hardness. At high frequencies where no dominant coupling lead or cable exists, the absorbed power is more related to the effective area of a circuit board. For this case meaningful measurements can only be made using a radiation testing technique.

If the threat environment is well defined, the levels at which to peform the subsystem tests are based on the interior environment as determined from the interaction and coupling analyses of Chapter 5. Such an approach tailors the hardening specifications to the particular subsystem. The approach used in electromagnetic compatibility (EMC) testing is to perform susceptibility testing according to MIL-STD-461, "Electromagnetic Interference Characteristics, Requirements for Equipment" and to MIL-STD-462, "Electromagnetic Interference Characteristics, Measurement Of." These standards are general purpose specifications applicable to subsystems, and they specify constant susceptibility levels in a given frequency range. For example, MIL-STD-461, Notice 4, which applies to all Army procurements, requires susceptibility levels (depending on frequency range) of 10, 20, and 50 volts/meter for sheltered equipment and 5 or 10 volts/meter for non-sheltered equipment. However, the fixed specification approach as opposed to the tailored specification approach has certain limitations.

Some of the limitations of the MIL-STD-460 series are summarized in the following paragraphs:

- The MIL-STD-460 series is a general purpose document which can result in an overdesign or underdesign of equipment at the subsystem level and thus cause increased cost to eventually develop a system.

- Compliance with the untailored MIL-STD-460 series does not guarantee system hardening since the coupling of subsystems to the real-world EMR environment is not considered.

- Subsystem data cannot be related to system performance.

- SPO management cannot assess the impact of the contractor's failure to harden a subsystem.

Tailored specifications along with a well-organized approach to hardening will result in a cost-effective and well engineered design.

The measurement techniques to be employed in the radiation testing of subsystems are described in the following paragraphs. The specific technique to be used for a given subsystem will depend on the physical dimensions of the subsystem. Radiation testing of small subsystems can be performed using parallel-plate, TEM cell, or anechoic chamber test facilities. For larger subsystems, anechoic chamber testing will be required.

40.2.3.2.1 **TEM CELL TESTING.** As discussed in Section 30.1.2, the TEM cell provides a shielded exposure chamber for RF susceptibility tests [H-27]. A typical test configuration for performing subsystem susceptibility tests with the cell is shown in Figure H-12. In operation, a high power oscillator or pulse generator is connected to the input port of the cell and a matched termination is attached to the output port. With this configuration, the TEM cell provides a means of generating accurately calibrated high intensity RF fields with moderate power RF sources. The subsystem to be tested is placed within the cell and its performance is monitored as a function of the exposure field level. The main limitation of the TEM cell, as for the parallel-plate transmission line, is the small test volume available at high test frequencies, i.e., the cell must be used at frequencies for which only the fundamental TEM mode exists.

40.2.3.2.2 **PARALLEL-PLATE TRANSMISSION LINE TESTING.** The susceptibility of a circuit to radiated fields can be measured using a parallel-plate transmission line (see Section 30.2). Transmission lines can be constructed to operate from DC up to a few hundred megahertz. The upper frequency range is limited by the line spacing. When the line spacing becomes greater than one-half wavelength, the field within the line begins to lose its uniformity.

In operation, the line is driven by a signal source connected to one end, and a receiver or RF voltmeter is connected to the other end to measure the voltage (V) between the plates. The magnitude of the field intensity (E) between the plates is given by $E = V/h$, where h is the separation distance between the plates. Susceptibility measurements are performed by placing the equipment to be tested between the plates and monitoring its operation for malfunction or degradation of performance.

The major limitation of the parallel-plate transmission line is the small test volume available at high test frequencies. Also, the line does not provide a shielded test volume.

40.2.3.2.3 **ANECHOIC CHAMBER TESTING.** It is recommended that subsystem radiation testing above 200 MHz be accomplished in an anechoic chamber as shown in Figure H-13. Ideally, the anechoic chamber should have a 3-foot diameter (spherical) quiet zone with 20 dB of quieting at 200 MHz. The chamber facility should have the capability of operating to 18 GHz. The chamber walls should have a power rating of at least 1 watt/sq. in. For test automation, the chamber facility should have high power swept sources capable of generating CW or pulsed field environments. Test antennas should have polarization diversity (horizontal, vertical, and circular) in order to test all orientations.

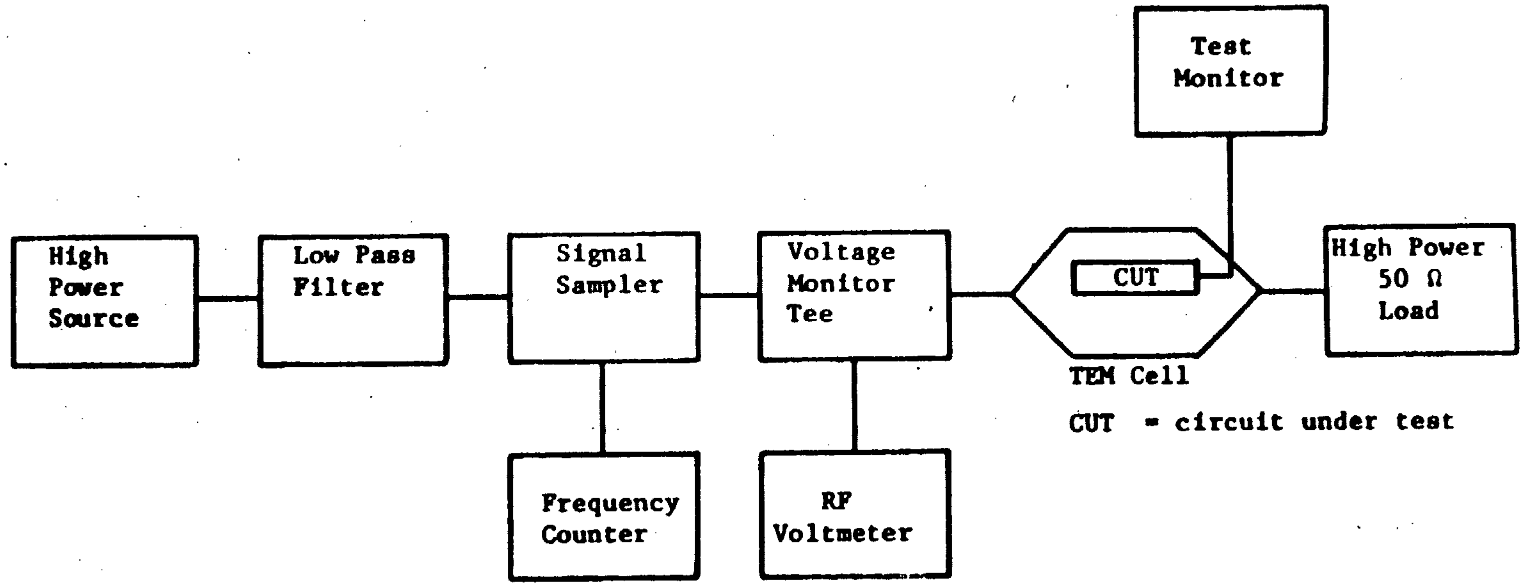


Figure H-12. TEM-Cell TEST Configuration

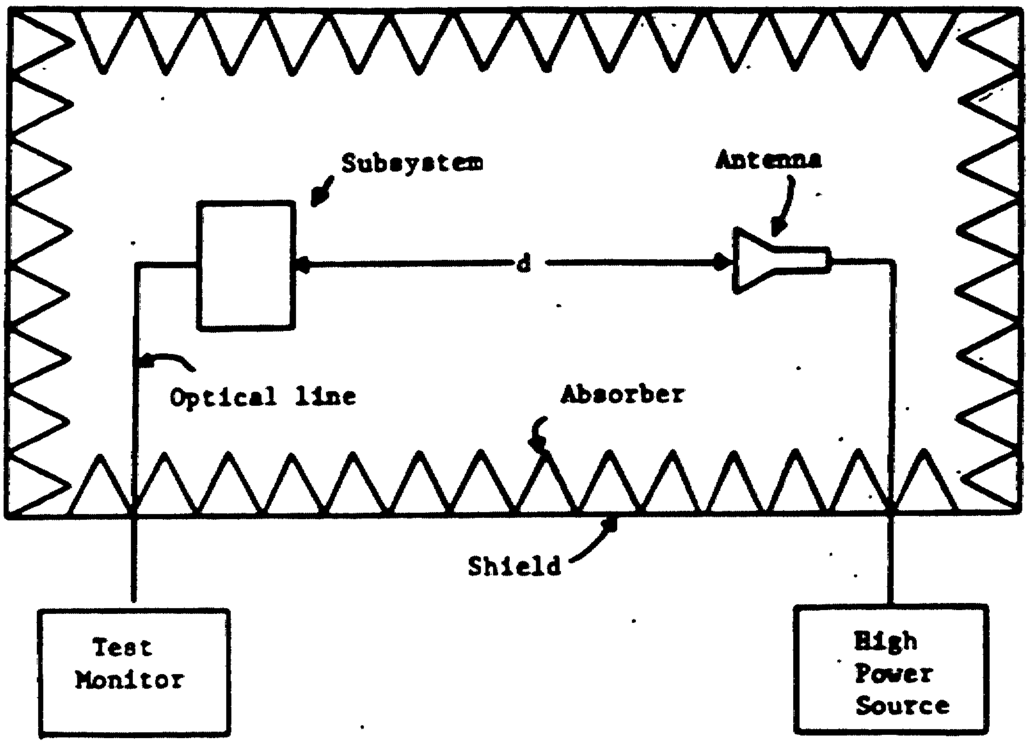


Figure H-13. Radiation Testing in a Chamber Facility.

The impact of these requirements on a weapon contractor may be severe, depending on the facilities presently available to the contractor. However, the use of the parallel-plate transmission line up to 200 MHz for subsystem testing, reduces the requirements for the anechoic chamber considerably. A 50 MHz anechoic chamber is very expensive because of the thickness requirement of the absorbing material. In a 200-MHz chamber, the test antenna size is smaller by a factor of four, thus reducing the required test volume. The drive power for a 200-MHz chamber is less than for a 50-MHz chamber, because of the increased antenna gain for comparable size antennas.

Quad-ridged horns are recommended for polarization diversity. Through a polarization switching and phasing network, vertical, horizontal, and circular polarizations can be selected. The gain of a quad horn at 200 MHz is typically 10 dB. To generate a 200 volts/meter field at a spacing of 3 meters with this antenna requires an input power of 1.2 KW. The beamwidth of the quad antenna is typically 30 degrees at the upper frequency limit. Figure H-14 shows the maximum width of a subsystem that can be uniformly illuminated. For the 30-degree beamwidth, the maximum width is approximately 5 feet. Thus, from this example it is evident that the antenna beamwidth should not be narrower than the angle subtended by the subsystem.

The simplest means of radiation testing above 1 GHz is with horn antennas. A single ridged-guide horn antenna covers the frequency range of 1 to 12 GHz. Figure H-15 shows a test configuration using horn antennas. The radiated field is calibrated as shown in Figure H-15a. A high power source drives the transmit horn, and the field at distance d is calibrated using a receive horn and a calibrated receiver. If standard gain horn antennas are used, the field can be calculated at a distance d within ±1 dB and the receive horn and calibrated receiver are not required. The test separation distance (d) should be greater than $2D^2/\lambda$ where D is the maximum dimension of the horn opening or the maximum dimension of the subsystem, whichever is the larger.

The drive power required to generate a field intensity E at the test separation distance (d) can be calculated using the following equation:

$$P_T = \frac{4\pi d^2 E^2}{Z_o G} \qquad \text{(H-3)}$$

where: P_T = power required (watts),
E = field intensity (volts/meter),
Z_o = free space impedance (ohms),
d = test separation distance (meters), and
G = antenna gain in power ratio.

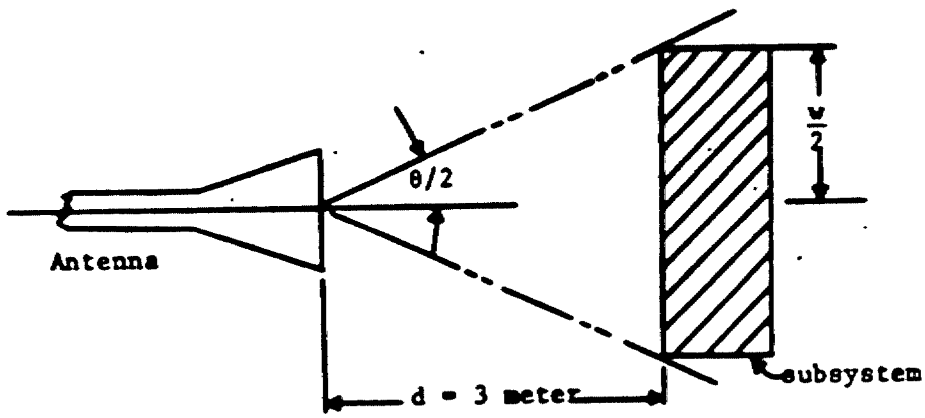
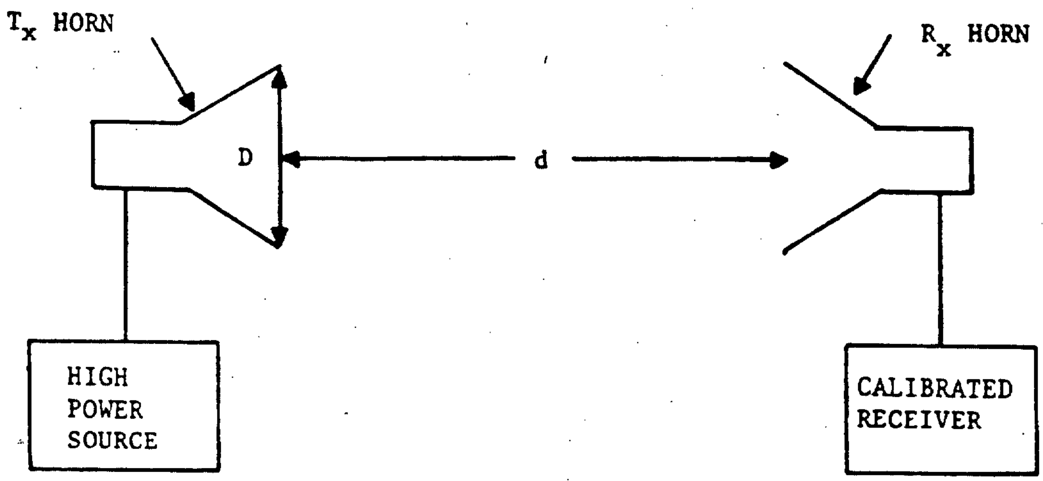
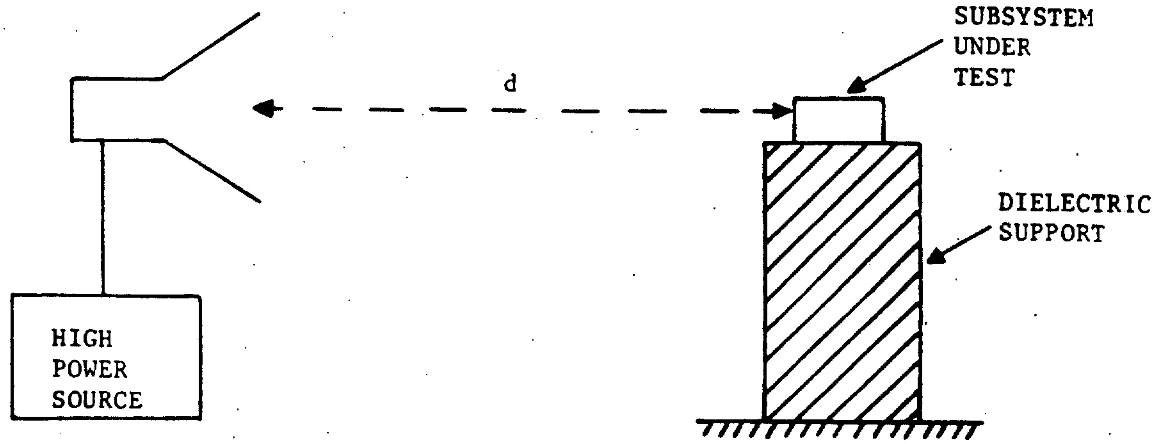


Figure H-14. Relationship Between Antenna Beamwidth and Subsystem Size.



(a) Calibration of Radiated Field.



(b) Radiation of Subsystem

Figure H-15. Subsystem Radiation from 1-18 GHz.

A typical ridged-guide horn antenna has a gain of 7 dB at 1 GHz. A drive power of 666 mW (+28.2 dBm) is required to generate a field intensity of 10 volts/meter at a test separation distance of 1 meter. At 12 GHz, the ridged-guide horn antenna gain is typically 13 dB, and a drive power of 167 mW (+22.2 dBm) is required.

Once the radiation field at the test separation distance has been calibrated, the subsystem is placed in the field as shown in Figure H-15b. When the subsystem is introduced, field perturbations should be minimized. For example, signal and power line lengths should be short. Figure H-16 illustrates a test configuration that will minimize field perturbations. The battery power pack and a test instrumentation package are placed in a small shielded enclosure. The outside of the enclosure is covered with an absorbing material (6 inches thick). The subsystem is placed above the shielded enclosure on dielectric supports. The critical test parameters are monitored through a fiber-optic link. The shielded enclosure is supported on a dielectric support column.

40.2.3.2.4 **SHIELDED ENCLOSURE TESTING.** The conventional shielded enclosure is not recommended for radiated susceptibility measurements because of errors that can result (due to reflections and multipath) in the measurement data. Shielded enclosures should be used only for isolating the test environment from the external environment during injection tests, parallel plate transmission line tests, or other type tests which do not require the establishment of a radiated environment within the enclosure.

40.2.3.3 **SHIELDING EFFECTIVENESS MEASUREMENT TECHNIQUE.** The basic idea of the MIL-STD-1377 shielding effectiveness measurement is to place the test specimen inside a specially constructed test chamber with provisions for radiating an electromagnetic signal while monitoring the amount of power picked up by the test specimen. Efforts are then made to maximize the power coupled to the test specimen by adjusting various tuning devices. The underlying concept is that if one finds the maximum amount of power that can be transferred to the test antenna or cable, one can infer that the difference between it and a reference antenna is intrinsic to the test specimen and its shielding effectiveness. Figure H-3 of Section 30.1.3 shows schematically the MIL-STD-1377 test chamber and tuning devices. Standard coaxial lines are used for input and output with stub tuners to match the impedance at the two ports. Inside the chamber, paddle wheel tuners, consisting of dipoles which can be rotated and moved laterally, provide for adjustment of the internal fields. When cable shielding effectiveness tests are being run, only one dipole tuner is needed; enclosure tests require a second dipole tuner located inside the enclosure under test. To minimize the losses and other undesirable effects, the dipoles are mounted on dielectric rods which pass through the chamber walls via waveguide below-cutoff tubes. In practice, the cabinet losses (as measured between the injection antenna and a reference pickup antenna) can be quite small over a usable frequency range of 1 to 10 GHz.

In operation, the test specimen is placed inside the chamber, a microwave signal of the desired frequency is injected into the chamber,

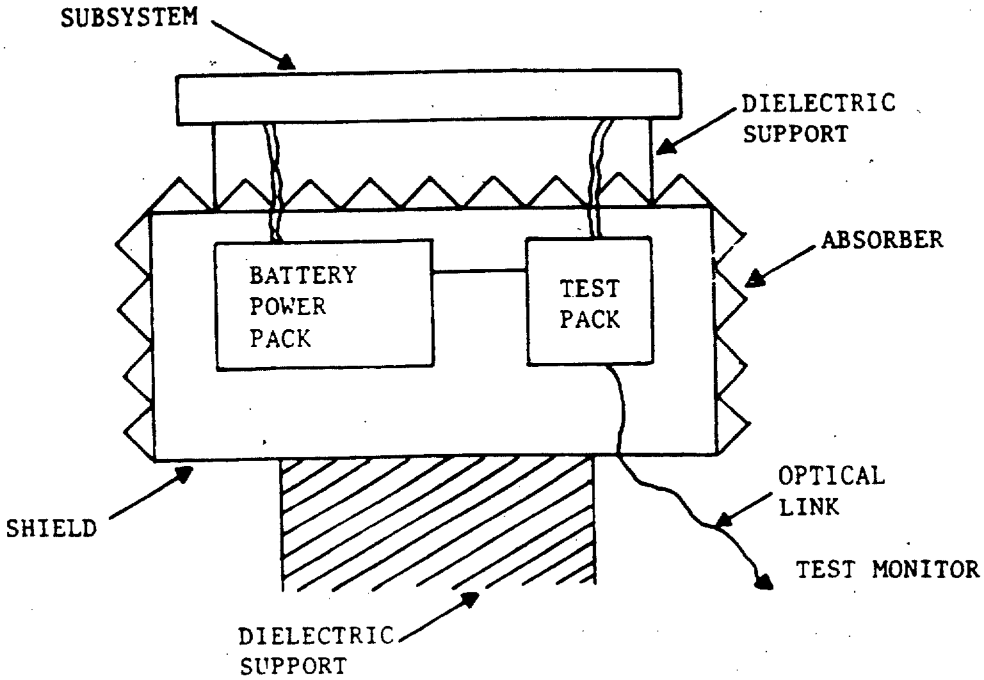


Figure H-16. Subsystem Test Configuration.

and the operator manually adjusts the various tuning devices in a somewhat iterative manner to find the maximum power that can be coupled through the test specimen. In the process, many local maxima are observed in which a tuning position is found for what any further tuning reduces the power transfer. The largest of these local maxima is indicative of the intrinsic shielding effectiveness of the specimen. Comparison of this result with that obtained by looking at the unshielded reference antenna (calibration measurement) inside the chamber indicates the shielding effectiveness value for the test specimen. The principal feature of this procedure is that it is highly repeatable with different operators and reconstructions of the experiments. The relative ease with which the absolute maximum tuning condition can be found depends on the test chamber parameters, tuning system design, and operator skill.

50. **REFERENCES.** The references in this section are organized into various categories related to EMI test instrumentation and techniques. In addition to the referenced source material used in the preparation of this appendix, other selected articles and documents are also listed to identify additional sources of information on EMI test instrumentation and technique.

SHIELDED ENCLOSURES AND ANECHOIC CHAMBERS

H-1. "Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," NAVAIR AD 1115, December 1975.

H-2. D. R. J. White, "A Handbook Series of Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-3. MIL-STD-1377, "Effectiveness of Cable, Connector, and Weapon Enclosure Shielding and Filters in Precluding Hazards of Electromagnetic Radiation to Ordnance; Measurement of," Department of the Navy, 20 August 1971.

H-4. Galagan, Steven, "Understanding Microwave Absorbing Materials and Anechoic Chambers," Microwaves, Part 1, pp. 38-41, December 1969; Part 2, pp. 44-49, January 1970; Part 3, pp. 47-50, April 1970; Part 4, pp. 69-73, May 1970.

H-5. S. L. O'Young et al., "Survey of Techniques for Measuring RF Shielding Enclosures," IEEE Transactions on Electromagnetic Compatibility, Vol. EMC-10, No. 1, March 1968, pp. 72-81.

SIGNAL SOURCES

H-6. D. R. J. White, "A Handbook Series of Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-7. "Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," NAVAIR AD 1115, December 1975.

ANTENNAS

H-8. D. R. J. White, "A Handbook Series on Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-9. MIL-STD-461A, "Electromagnetic Interference Characteristics, Requirements for Equipment," Department of Defense, Washington, DC, 1 August 1968.

H-10. Hollis, J. S., Lyon, T. J., and Clayton, L., "Microwave Antenna Measurements," Scientific-Atlanta, Inc. Publications, pp. 14-16 to 14-41, July 1970.

FIELD INTENSITY METERS

H-11. D. R. J. White, "A Handbook Series on Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-12. MIL-STD-461A, "Electromagnetic Interference Characteristics, Requirements for Equipment," Department of Defense, Washington, DC, 1 August 1968.

AUXILIARY DEVICES

H-13. D. R. J. White, "A Handbook Series on Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-14. MIL-STD-461A, "Electromagnetic Interference Characteristics, Requirements for Equipment," Department of Defense, Washington, DC, 1 August 1968.

H-15. "Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," NAVAIR AD 1115, December 1975.

ELECTROMAGNETIC ENVIRONMENT MEASUREMENTS

H-16. D. R. J. White, "A Handbook Series on Electromagnetic Interference and Compatibility, Vol. 4, EMI Test Instrumentation and Systems," 1971.

H-17. "Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," NAVAIR AD 1115, December 1975.

COMPONENT AND CIRCUIT TESTS

H-18. R. A. Amadori, W. J. Peters, and V. G. Puglielli, "NWL Microwave Susceptibility Measurement System for Solid State Components," NWL Technical Note No. TN-F/97-71, May 1971.

H-19. R. E. Richardson, V. G. Puglielli, and R. A. Amadori, "Microwave Interference Effect in Bipolar Transistors," IEEE Transactions on Electromagnetic Compatibility, Vol. EMC-17, No. 4, 216, 1975.

H-20. R. A. Amadori, V. G. Puglielli, and R. E. Richardson, "Prediction Methods for the Susceptibility of Solid-State Devices to Interference and Degradation of Microwave Energy," Proceedings of the International Electromagnetic Compatibility Symposium, 1973, pp. 184-200.

H-21. Roe, J. M., "Microwave Interference Effects in Integrated Circuits," IEEE EMC Symposium Record, San Antonio, Texas, 7-9 October 1975, pp. 4AIb3-6.

H-22. "Integrated Circuit Electromagnetic Susceptibility Investigation - Phase II: Test and Measurement Systems," Report No. MDC E1099, Contract No. N00178-73-C-0362, McDonnell-Douglas Astronautics Co., St. Louis, MO, July 1974.

H-23. "Integrated Circuit Electromagnetic Susceptibility Investigation - Phase III, IC Susceptibility Handbook, Draft I," Report MDC E1515, Contract No. N60921-76-C-A030, McDonnell-Douglas Astronautics Company, St. Louis, MO, 4 June 1976.

SUBSYSTEM TESTS

H-24. AFSC Design Handbook, "Electromagnetic Compatibility," DH 1-4, Chapter 6, Tests and Procedures.

H-25. EMP (Electromagnetic Pulse) Handbook (U), Vol. 2, Analysis and Testing, DNA2114H-2 (CONFIDENTIAL) Section 16.2.2.

H-26. Sandia Laboratories, "Electromagnetic Pulse Handbook for Missiles and Aircraft in Flight," Air Force Weapons Laboratory Publication AFWL-TR-37-68 (AD 919-395L), September 1972.

H-27. Crawford, M. L., "Generation of Standard EM Fields Using TEM Transmission Cells," IEEE Transactions on EMC, Vol. EMC-16, No. 4, pp. 189-195, November 1974.

H-28. Riley, Leon H., "Infrared Data Link for Electromagnetic Effects Testing," U.S. Army Missile Command Technical Report RT-75-8, Redstone Arsenal, Alabama, December 1974.

H-30. Heverly, J. R., and Doeppner, T. W., "Integrated Electromagnetic (EM) Analysis and Testing," IEEE Electromagnetic Compatibility Symposium Record, San Antonio, Texas, pp. 6AIf1 to 6AIf6, October 1975.

H-31. Grinoch, Abraham, "Electromagnetic Test and Hardening of Weapon Systems," IEEE International Electromagnetic Compatibility Symposium Record, New York, pp. 132-143, June 1973.

H-32. Sachs, H. M., "Developing Tailored EMI Systems Specifications," IEEE Electromagnetic Compatibility Symposium Record, San Antonio, Texas, pp. 105-109, October 1975.

H-33. Baldwin, Thomas, E., Jr., "Generation of Tailored EMC Specifications," IEEE Electromagnetic Compatibility Symposium Record, Washington, DC, pp. 246-249, July 1975.

SHIELDING EFFECTIVENESS TESTS

H-34. W. Jarva, "Shielding Tests for Cables and Small Enclosures in the 1 to 10 GHz Range," IEEE Transactions on Electromagnetic Compatibility, Vol. EMC-12, No. 1, February 1970.

H-35. MIL-STD-1377, "Effectiveness of Cable, Connector, and Weapon Enclosure Shielding and Filters in Precluding Hazards of Electromagnetic Radiation to Ordnance; Measurement of," Department of the Navy, 20 August 1971.

H-36. "Characterization of Shielding Effectiveness Measurement Technique," Report MDC E1516, Contract No. N60921-76-C-A076, McDonnell-Douglas Astronautics Company, St. Louis, MO, 19 July 1976.

APPENDIX I

EMR HARDNESS CONSIDERATIONS IN PROGRAM DOCUMENTS

10. **INTRODUCTION.** The major planning and procurement documents form a logical flow of activity representing an orderly progression from concept formulation to a final operational deployment, and, hence, they can provide a vehicle for implementing and documenting an EMR hardness program throughout the life cycle of a system. To accomplish this implementation and documentation, the EMR hardness requirements must be addressed in the following program documents.

SON	Statement of Operational Need
MENS	Mission Element Need Statement
PMD	Program Management Directive
FORM 56	AFSC Program Direction
PMP	Program Management Plan
DCP I	Decision Coordinating Paper (Milestone I)
APP	Advanced Procurement Plan
TEMP	Test and Evaluation Master Plan
RFP	Request for Proposal (Full-Scale Development)
DCP II	Decision Coordinating Paper (Milestone II)
RFP	Request for Proposal (Production)
OMTD	Operation, Maintenance, and Training Documents
DCP III	Decision Coordinating Paper (Milestone III)
PMRT	Program Management Responsibility Transfer

20. **SON AND MENS.** The Statement of Operational Need (SON) document is the mechanism for identifying operational needs within the Air Force. A prerequisite to the identification of operational needs is a Mission Area Analysis (MAA). The MAA examines the mission responsibilities of a command and assesses the command's ability to perform each task in terms of current capabilities, the present and projected threat, the operational environment, and any other constraints which may limit solutions to accomplishing the stated task. In analyses involving air launched ordnance systems, the EMR environment and its potential affects on the performance of the system should be included.

When the MAA identifies an operational need, the need is documented in a SON. Upon publication, the SON is submitted to HQ USAF for validation. Upon validation of the SON, a Mission Element Need Statement (MENS) is prepared and submitted to the Secretary of Defense (Secretary of Air Force for non-major programs) for approval. Upon approval of the MENS, Milestone Zero is achieved and the Conceptual Phase of the program begins.

30. **PMD AND FORM 56.** Upon approval of the MENS, HQ USAF issues a Program Management Directive (PMD) to initiate action in the Air Force. For air launched ordnance systems, the PMD should require that an EMR hardness program be implemented to ensure that the system developed will be compatible with its EMR operational environment. After receipt of the PMD from HQ USAF, HQ AFSC establishes the program priority and issues guidance and direction to the AFSC organizations via AFSC Form 56 (AFSC Program Direction). The Form 56 for air launched ordnance sys-

tems should require the program office to implement an EMR hardness program to ensure that the system developed is compatible with its EMR operational environment.

40. **DCP.** The Decision Coordinating Paper (DCP) is the documentation requesting approval to enter the next phase of system acquisition. DCP I requests approval to proceed to the Concept Validation Phase (Milestone I). DCP II requests approval to proceed to the Full-Scale Development Phase (Milestone II). DCP III requests approval to proceed to the Production Phase (Milestone III). The EMR hardness considerations for inclusion in the Decision Coordinating Papers should include the following:

- o Describe the EMR hardness plan for achieving the EMR hardness goals.
- o State tests planned to demonstrate EMR hardness.
- o Describe EMR hardness tasks and tests accomplished and those remaining to be accomplished.
- o State anticipated problems and risks in accomplishing EMR hardness goals.
- o Specify anticipated resolution dates for identified problems and risks.

50. **PMP AND TEMP.** The Program Management Plan (PMP) is prepared by the Program Manager early in the conceptual phase of the program and describes the overall program management approach, the master program schedule, and the estimated program cost. A description of the development, management, and implementation of an EMR Hardness Program should be included in this document.

The Test and Evaluation Master Plan (TEMP) is the controlling management document which defines all test and evaluation (T&E) efforts to be accomplished in connection with a system acquisition. Descriptions of the T&E efforts required to implement and validate the EMR Hardness Program should be included in this document. These descriptions should include the roles and responsibilities of other DoD T&E organizations, as well as the T&E efforts to be performed by the contractor.

60. **RFP.** The Request for Proposal (RFP) document package completely describes the proposal requirements and the tasks to be accomplished during each acquisition phase. The RFP's for all acquisition phases of air launched ordnance systems should address the requirement that the system must be capable of performing its mission in the EMR operational environment. However, special emphasis should be placed on defining the EMR Hardness requirements in the RFP for the Full-Scale Development Phase and the RFP for the Production Phase.

The RFP for the Full-Scale Development Phase should include a complete description of the forecasted EMR operational environment.

This EMR environment forecast should be the most accurate possible at this stage of the program, since it will serve as a basis for the system designers to determine the amount of EMR hardness required for the system.

The RFP for the Production Phase should emphasize that the production techniques utilized shall not compromise the EMR hardness incorporated in the full-scale development model and should require that adequate quality control tests are performed to ensure that the required EMR hardness characteristics are realized in the production systems.

70. **OMTD.** The Operation, Maintenance, and Training Documents (OMTD) for air launched ordnance systems should emphasize the importance of maintaining the EMR hardness characteristics of these systems. Sufficient instructions should be included in these documents to ensure that EMR hardness is not compromised during maintenance, check-out, and deployment.

80. **PMRT.** Program Management Responsibility Transfer (PMRT) is the formal act of transferring management responsibility for a system to the Air Force Logistics Command (AFLC). The transfer documentation for air launched ordnance systems should emphasize the importance of maintaining the EMR hardness characteristics of the systems. This documentation should also stress the importance of evaluating the impact of system modifications and engineering change proposal (ECP) requests on the EMR hardness of the systems.

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APPENDIX J

TAILORING OF SPECIFICATIONS AND STANDARDS

10. **INTRODUCTION.** In many cases, it may be desirable to contractually obligate the system contractor to provide adequate EMR hardness and testing by invoking specifications and standards. Unfortunately, there are currently no specifications or standards which address the EMR hardness problem as defined in this handbook. However, since the EMR hardness problem is a special case in a subset (radiated susceptibility) of electromagnetic compatibility (EMC), several of the current EMC specifications and standards can be tailored to achieve the goals of the EMR hardness program.

20. **TAILORING.** The tailoring of specifications and standards is the process by which the requirements of a specification document are adapted to the peculiarities, operational requirements, or characteristics of a system being developed and the electromagnetic environment in which the system must operate. The tailoring process should accomplish two purposes. The first objective is to modify existing requirements and test procedures and, if necessary, add additional requirements and test procedures so that the document assures that all required functions are performed and validated. The second objective is to delete all requirements and test procedures which are not essential to satisfying the specific requirements of the system. Thus, a properly tailored specification or standard should ensure that the essential requirements of a system will be satisfied, and at the same time, will not require the performance of any nonessential functions or testing.

It is recommended that the tailoring of specifications and standards to achieve the goals of the EMR hardness program be accomplished by the EMCAB under the direction of the program manager. The tailoring effort will constitute a part of the preparation of the solicitation documentation. Some allowance should be provided to permit the contractor to recommend subsequent tailoring of the specifications subject to approval by the EMCAB and the program manager.

30. **APPLICABLE SPECIFICATION AND STANDARDS.** The specifications and standards most appropriate for application to the EMR hardness program are listed below and discussed in the following paragraphs.

MIL-E-6051	Electromagnetic Compatibility Requirements, Systems
MIL-HDBK-237	Electromagnetic Compatibility/Interference Program Requirements
MIL-STD-461	Electromagnetic Interference Characteristics, Requirements for Equipment

MIL-STD-462	Electromagnetic Interference Characteristics, Measurement of
MIL-STD-463	Definitions and Systems of Units, Electro-magnetic Interference Technology
MIL-STD-1512	Electroexplosive Subsystems, Electrically Initiated, Design Requirements and Test Methods
AMCP 706-235	Hardening Weapon Systems Against RF Energy
MIL-STD-1377	Effectiveness of Cable, Connector, and Weapon Shielding and Filters In Precluding Hazards of Electromagnetic Radiation to Ordnance, Measurement of
MIL-B-5087	Bonding, Electrical, and Lightning Protection For Aerospace Systems

30.1 **MIL-E-6051.** This specification establishes the overall electromagnetic compatibility requirements for electronic systems, including control of the system electronic environment, lightning protection, static electricity, bonding, and grounding. It is applicable to complete systems, including all associated subsystems and equipments. The use of this specification is officially mandatory for all Departments and Agencies of the DoD in procurements of electronic systems, implying that the EMC requirements outlined will be imposed on all systems. Under these circumstances, the tailoring of this specification to include the EMR hardness requirements is relatively simple. The tailoring process consists of: (1) ensuring that the external electromagnetic environment discussed in paragraph 3.2.13 includes the EMR environment forecast, and (2) deleting the phrase "--except deliberately generated interference--" from the definition of electromagnetic interference in paragraph 6.2.3. In addition, it may be desirable to modify the limits and requirements of MIL-STD-461 and 462 as discussed in paragraph 3.2.4.1, depending on how these standards are to be used in controlling the subsystems and equipments (see discussion in 30.3 below).

In the event that a system procurement does not invoke EMC requirements, tailoring MIL-E-6051 to address only the requirements of the EMR hardness program is a much more difficult task. The tailoring process requires the deletion of all the EMC requirements which do not directly address the radiated susceptibility characteristics of the system.

30.2 **MIL-HDBK-237.** This handbook provides additional guidance and more detailed procedures for adapting the MIL-E-6051 specification to a particular program. As a handbook, the document is intended to be referenced in procurement specifications for information purposes only, and is not intended to supersede any specification requirement. However, the use of the information and guidelines provided in this handbook will assist the program office and contractor personnel in establishing, managing, implementing, and validating an effective EMC/EMR hardness program. As long as EMC requirements are invoked on the system procurement, no tailoring of this handbook to include an EMR hardness program in the overall EMC program is required other than to ensure that the EMR environment forecast is included in the electromagnetic environment.

30.3 **MIL-STD-461, 462, AND 463.** This series of standards establishes the electromagnetic compatibility requirements for electronic, electrical, and electromechanical equipments and subsystems. MIL-STD-461 defines the general requirements, measuring equipment, and test limits for the EMC measurements required. MIL-STD-462 establishes the measurement techniques and setups to be used in measuring the EMC characteristics. MIL-STD-463 presents definitions of the terms, abbreviations, acronyms, and system of units used in the other two standards. Unless otherwise specified in the contract, invoking MIL-E-6051 automatically invokes MIL-STD-461 and MIL-STD-462.

MIL-STD-461 and MIL-STD-462 can be used to control the EMR hardness of subsystems, components, and interwiring harness of the system being developed. Either one of two approaches can be taken to accomplish this control. One approach is to require all subsystems, components, and interwiring to meet the test limits specified in MIL-STD-461. This establishes a susceptibility level in the interior of the system at the level specified by the radiated susceptibility test limits of MIL-STD-461. With this approach, it is necessary to provide sufficient shielding effectiveness in the outer skin or shield of the system to ensure that the EM environment in the interior of the system does not exceed the susceptibility test limits of MIL-STD-461 when the system is exposed to the maximum exterior EMR environment.

If it is not possible, or desirable, to provide sufficient shielding in the outer skin of the system to reduce the penetration of the exterior EMR environment to a level below the susceptibility test limits, the susceptibility test limits in MIL-STD-461 and the test procedures in MIL-STD-462 may be modified to require the subsystems, components and interwiring to be immune to a higher level interior EM environment. This approach places more of the hardening burden on the design and fabrication of the subsystems of the system and constitutes a layered hardening approach. In the modification of the susceptibility test limits, care must be exercised to ensure that the subsystem hardening requirements are realistic and obtainable.

30.4 **MIL-STD-1512.** This standard establishes the general requirements and test methods for the design and development of electroexplosive subsystems and associated items to preclude hazards from unintentional initiation. This standard should be imposed on all subsystems utilizing electrically initiated explosive, or pyrotechnic, components to be included in the system being developed. The electromagnetic environment used to establish the hardness criteria for such subsystems should also include the forecasted EMR environment.

30.5 **AMC PAMPHLET 706-235.** This handbook provides detailed information on hardening weapon systems against RF energy. The information is directed specifically to hardening systems and subsystems containing electroexplosive devices (EED's) against electromagnetic environments and is primarily intended to assist the design engineer in meeting the requirements of specification documents such as MIL-STD-1512. However, a great deal of the information is applicable to all aspects of the EMR hardness program and should be useful in the development of program and test plans, as well as in system design. This handbook may be referenced for information purposes, particularly in support of MIL-STD-1512, without any tailoring.

30.6 **MIL-STD-1377.** This standard is intended to provide system designers and developers with test methods for determining if particular portions of a system are providing the anticipated hardness. It is not intended to be a substitute for full-scale EMR hardness tests of a system, but rather an aid in developing a system with a high probability of successfully passing such environmental tests. The standard describes methods of evaluating the shielding effectiveness of shielding enclosures, cables, and cable connectors over the frequency range from 1 GHz to 10 GHz. The standard also describes methods of measuring the filtering effectiveness of interference suppression filters over the frequency range from 100 kHz to 10 GHz. This standard may be invoked to support MIL-E-6051, MIL-STD-461, and MIL-STD-462 without any tailoring.

30.7 **MIL-B-5087.** This specification establishes the requirements, procedures, and test methods for electrical bonding and grounding in aerospace systems. The application of this specification ensures that the bonding and grounding practices utilized in the system development comply with accepted EMC practices. This specification may be invoked to support the EMR hardness program without any tailoring. This specification is invoked by reference by MIL-E-6051 and MIL-STD-1512.

APPENDIX K

OUTLINE FOR EMR HARDNESS PROGRAM PLAN

1. Introduction
   1.1 Purpose of Plan
   1.2 Scope of Program
   1.3 Description of System
      1.3.1 Performance Functions
      1.3.2 Intended Missions
      1.3.3 Delivery Platforms
      1.3.4 Anticipated Targets

2. Management Concept, Organization, and Responsibilities
   2.1 Program Manager's Responsibilities
   2.2 Key Program Office Personnel
   2.3 EMCAB's Functions and Responsibilities
   2.4 Prime Contractor's Responsibilities
   2.5 Interface Between Program Office and Contractor
   2.6 Tasks and Responsibilities of Other Organizations
      2.6.1 HQ AFSC
      2.6.2 AFTEC
      2.6.3 RADC
      2.6.4 ECAC
      2.6.5 Others

3. EMR Hardness Activities Scheduling and Milestones
   3.1 Concept Development Phase
   3.2 Concept Validation Phase
   3.3 Full Scale Development Phase
   3.4 Production Phase
   3.5 Deployment Phase

4. Applicable Documents and Specifications
   4.1 Military Standards, Specifications, and Handbooks
   4.2 Other Government Publications
   4.3 Industry Publications

5. Prediction and Analysis
   5.1 Description of Prediction and Analysis Techniques to be Employed
   5.2 Milestones for Analyses
   5.3 Definition of Guidelines for EMR Environment Forecasts
   5.4 Identification of Potential EMR Hardness Problems
   5.5 Definition of Risks

6. EMR Hardness Testing
   6.1 Engineering Development
   6.2 Prototype System
   6.3 Production System

7. Documentation and Schedule
   7.1 Charter for EMCAB
   7.2 Test and Evaluation Master Plan (TEMP)
   7.3 Request for Proposal (RFP)
   7.4 Contractor's Control Plan
   7.5 Contractor's Test Plan
   7.6 Contractor's Test Report
   7.7 Prototype System Test Plan
   7.8 Production System Test Plan

APPENDIX L

OUTLINE FOR EMR HARDNESS CONTROL PLAN

10. **INTRODUCTION.** This section should describe in detail the system to be developed and the EMR hardness requirements which must be satisfied by the EMR hardness control plan.

20. **MANAGEMENT.** This section should describe the management organization structure to be utilized by the contractor to implement and direct the EMR hardness program. The description should define how the hardness control management structure fits within the contractor's organization management structure and the overall program management structure. The individual who will implement and direct the EMR hardness program should be identified and his responsibility and authority should be defined. The proposed method for interface between the contractor's EMR hardness management and the program office should be described.

30. **MECHANICAL DESIGN.** This section should describe the EMR hardness considerations which will be included in the mechanical design of the system. As a minimum, this section should include specific data in the following areas:

(a) Types of materials, finishes and hardware to be employed in the mechanical design.
(b) Shielding design practices to be employed to realize the required degree of shielding effectiveness.
(c) The types and characteristics of RF gaskets to be used on all internal and external mating surfaces.
(d) The types and characteristics of shielding devices to be used on required openings such as ventilation ports and control shaft apertures.
(e) Design practices for isolation bulkheads and compartmentalization.
(f) Corrosion control practices.

40. **WIRING DESIGN.** This section should describe the contractor's proposed electrical and electronic wiring design. As a minimum, this section should include detailed descriptions of the following areas:

(a) Grounding philosophy and practices
(b) Bonding philosophy and practices
(c) Cable shielding practices
(d) Wiring hardness design practices

50. **CIRCUIT DESIGN.** This section should describe the EMR hardness considerations which will be included in the design and development of the electronic circuitry of the system. As a minimum, this section should include detailed descriptions of the contractor's approaches and procedures in the following areas:

(a) Criteria and procedures for the selection of components and circuit configurations.
(b) Procedure for identifying critical circuits.

(c) Practices for shielding and isolating critical circuits.
(d) Descriptions of suppression techniques and devices to be utilized and their applications.
(e) Bonding and grounding criteria.

60. **SUBSYSTEM DESIGN AND CONTROL.** This section should describe in detail how the contractor plans to ensure that all subsystems, equipments, and components supplied by subcontractors and vendors meet the specified EMR hardness requirements.

70. **PREDICTION AND ANALYSIS.** This section should describe in detail the prediction and analysis techniques the contractor plans to use in his EMR hardness program. In addition, this section should describe how the results from the analyses will be utilized in the development of the system and provide a proposed schedule for the planned analyses.

80. **ANTICIPATED PROBLEM AREAS AND PROPOSED SOLUTIONS.** This section should identify the areas in which the contractor anticipates the major EMR hardening problems will be encountered and describe the contractor's proposed solutions to these problems.

90. **UPDATING.** This section should describe the contractor's plan to update the EMR hardness control plan to ensure that the plan accurately reflects the current EMR hardness requirements throughout the development program.

APPENDIX M

OUTLINE FOR EMR HARDNESS TEST PLAN

10. **INTRODUCTION.** This section should describe in detail the EMR hardness tests to be performed in conjunction with the system design and development. The following information should be included:

(a) A description of the system to be developed.

(b) The objective of the test plan and its relationship to the overall EMR hardness program for the system.

(c) The scope of the test program in terms of the types of tests to be performed, the test facilities and procedures to be employed, and the utilization of the test data.

(d) A general description of the scheduling of the EMR hardness tests with respect to the system design/development cycle.

20. **APPLICABLE DOCUMENTS.** This section should identify those documents which are applicable to the test program, and describe how the documents will be utilized.

30. **TEST FACILITIES AND TEST INSTRUMENTATION.** This section will describe the EMR test facilities and test instrumentation to be employed for each type of test to be performed, and will include:

(a) The location of the test facilities.

(b) A detailed description of the technical characteristics and specifications for each test facility (anechoic chamber, TEM cell, etc.).

(c) A detailed description of the technical characteristics and specifications of the test instrumentation (signal sources, receivers, antennas, etc.) used in conjunction with each test facility.

(d) A detailed description of the test configuration to be employed for EMR hardness tests at each facility, including the physical layout and electrical interconnections of the facility, test instrumentation, and unit under test.

40. **UNIT UNDER TEST OPERATION.** This section will describe in detail the operating status of the unit under test during each EMR hardness test. The rationale for the operating status will be defined, and all operating parameters and control settings will be documented.

50. **TEST PROCEDURE.** This section will fully document the test procedures to be employed in the EMR hardness tests including:

(a) Block diagram depicting the test setup for each test method.

(b) Detailed step-by-step procedures enumerating the performance of each test.

(c) A definition of the test criteria for each test and the relationship of these criteria to the EMR hardness of the unit under test.

60. **TEST DATA.** This section will describe the data to be recorded during each test, and will include sample data sheets, test logs, and graphs.

APPENDIX N

BIBLIOGRAPHY

10. **INTRODUCTION.** The following is a list of handbooks and design guides which may be of benefit to program managers, contractors, and design engineers in the area of EMR hardness.

10.1 **HANDBOOKS AND DESIGN GUIDES.**

AD-619666	"Interference Reduction Guide for Design Engineers," Vol. 1, U.S. Army Electronics Laboratories, Ft. Monmouth, N. J., Aug. 1964.
AD-619667	"Interference Reduction Guide for Design Engineers," Vol. 2, U.S. Army Electronics Laboratories, Ft. Monmouth, N. J., Aug. 1964.
AFSC DH 1-4	"Air Force Systems Command Design Handbook, DH 1-4, Electromagnetic Compatibility," ASD/ENYEH, Wright-Patterson, AFB, OH, 5 January 1975.
AFSC DH 2-5	"Air Force Systems Command Design Handbook, DH 2-5, Armament," ASD/ENYHM, Wright Patterson AFB, OH, 15 May 1975.
AFWL TR 73-68	"Electromagnetic Pulse Handbook for Missile and Aircraft in Flight," Air Force Weapons Laboratory, Kirtland AFB, NM, September 1972.
AMC Pamphlet 706-235	"Engineering Design Handbook, Hardening Weapon Systems Against RF Energy," Headquarters, U.S. Army Materiel Command, Feb. 1972.
DARCOM Pamphlet 706-410	"Engineering Design Handbook, Electromagnetic Compatibility," Headquarters, U.S. Army Development and Readiness Command, March 1977.
DNA 2114H-1	"DNA EMP Handbook, Vol. 1 -- Design Principles," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.
DNA 2114H-2	"DNA EMP Handbook, Vol. 2 -- Coupling Analysis," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.
DNA 2114H-3	"DNA EMP Handbook, Vol. 3 -- Component Response and Test Methods," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.
DNA 2114H-4	"DNA EMP Handbook, Vol. 4 -- Environment and Applications," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.

DNA 2114H-5	"DNA EMP Handbook, Vol. 5 -- Resources," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.
DNA 2114H-6	"DNA EMP Handbook, Vol. 6 -- Computer Codes," (CONFIDENTIAL), Defense Nuclear Agency, Washington, D.C., 5 July 1979.
FAA-RD-75-215	"Grounding, Bonding and Shielding Practices for Electronic Equipments and Facilities," Vols. 1-3, Federal Aviation Administration, Systems Research & Development Service, Washington, D.C., Dec. 1975.
MDC E1929	"Integrated Circuit Electromagnetic Susceptibility Handbook," McDonnell-Douglas Astronautics Co., St. Louis, MO, 1 Aug. 1978.
MIL-HDBK-235-1	"Electromagnetic (Radiated) Environment Considerations for Design and Procurement of Electrical and Electronic Equipment (General Information)," 23 June 1972.
MIL-HDBK-235-2	"Electromagnetic Radiation Environment from Friendly or Own Force Emitters," (CONFIDENTIAL), 23 June 1972.
MIL-HDBK-235-3	"Main Beam Electromagnetic Radiation Environment from Hostile Force Emitters," (SECRET), 23 June 1972.
MIL-HDBK-237A	"EMC Management Guide for Platforms, Systems, and Equipment," 20 April 1973.
MIL-HDBK-253	"Guidance for the Design and Test of Systems Protected Against the Effects of Electromagnetic Energy," 28 July 1978.
MIL-HDBK-760	"U.S. Army SEMI Handbook," (SECRET), Office of Missile Electronic Warfare, White Sands Missile Range, NM., 11 Sept. 1978.
NASA NHB 5320.3	"Electromagnetic Compatibility Principles and Practices," National Aeronautics and Space Administration, Washington, D.C., Oct. 1965.
NASA SP-3067	"Radio Frequency Interference Handbook," National Aeronautics and Space Administration, Washington, D.C., 1972.
NAVAIR AD-1115	"Electromagnetic Compatibility Design Guide for Avionics and Related Ground Support Equipment," Dec. 1975.

NAVAIR 5335	"Electromagnetic Compatibility Manual," April 1972.
NAVELEX 0101, 106	"Naval Shore Electronics Criteria Electromagnetic Compatibility and Electromagnetic Radiation Hazards," Aug. 1971.
NAVWEPS OD 30393	"Design Principles and Practices for Controlling Hazards of Electromagnetic Radiation to Ordnance (HERO Design Guide)," 6 June 1965.
NSWC	"Design Handbook for Control of RF/Microwave Electromagnetic Effects," Naval Surface Weapons Center, Dahlgren, VA, 30 Sept. 1977.
NAVMAT	"TESSAC Electromagnetic Environment Effects RDT&E Program Plan," Naval Material Command, Washington, D.C., 24 May 1978.
NSWC	"TESSAC Final Report for EM Vulnerability," Naval Surface Weapons Center, Dahlgren, VA, 1 June 1977.
NAVSEC	"TESSAC Electromagnetic Safety Study Group," Naval Ship Engineering Center, Washington, D.C., 31 May 1977.
NOSC	"TESSAC EM-Power Task Technology Capabilities Specifications," Naval Ocean Systems Center, San Diego, CA, 12 July 1977.
NRL	"TESSAC Final Report on Electronic Counter-Countermeasures," (SECRET), Naval Research Laboratory, Washington, D.C., 15 July 1977.
NAVMAT	"TESSAC Electromagnetic Environment Effects -- Summary Report to Chief of Naval Material," Naval Material Command, Washington, D.C., 30 Sept. 1977.
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NPGS	"TESSAC Electromagnetic Compatibility Survey," Naval Post Graduate School, Monterey, CA, September 1977.
NSWC	"TESSAC EMP Protection Engineering Study," Naval Surface Weapons Center, White Oak, MD, 1 July 1977.
NAVMAT	"TESSAC In-Service Support Plan for Electromagnetic Environment Effects," Naval Material Command, Washington, D.C., 5 May 1978.

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APPENDIX O

DEFINITIONS AND ACRONYMS

10. DEFINITIONS.

Analog (Linear) Device - A device that operates with variables represented by continuously varying physical quantities.

Aperture - An opening in a shield through which electromagnetic energy may pass.

Composite Environment - A single environmental profile which represents an integration of the various electromagnetic environments to which the system is exposed.

Coupling - The transfer of electromagnetic energy between fields, wires, circuits, equipments, or systems.

Degradation - An undesired change in the operational performance of a device, circuit, subsystem, or system as the result of interference.

Digital Device - A device that operates on the basis of discrete numerical techniques in which the variables are represented by coded pulses or states.

Electromagnetic Compatibility - The capability of equipments or systems to be operated in their intended operational environment at designed levels of efficiency without causing or experiencing degradation owing to unintentional electromagnetic interference.

Electromagnetic Environment - The electromagnetic levels that impinge on the system at a particular time and location in space. The ground environment includes all ground-based and ship-based emitters in a specified geographical area. The cosite environment includes all emitters on-board the delivery aircraft. The intersite environment consists of all emitters from aircraft other than the delivery aircraft. The approach-to-target environment will include any radiation from the target itself and the ground environment adjusted for the critical range of the system.

EMR Hardness - The degree or extent to which a system is immune to EMR susceptibility.

Electromagnetic Interference - The phenomenon resulting when undesired electromagnetic energy causes unacceptable responses in the intended operation of electronic equipment, subsystems, or systems.

EMR Susceptibility - The characteristic of a component or system which permits undesired responses when exposed to a radiated electromagnetic enviroment. The system susceptibility levels are those determined to cause degradation in the intended performance of the system.

EMR Vulnerability - The characteristic of a system that permits degradation sufficient to compromise the system mission when exposed to its anticipated EMR environment. EMR vulnerability exists when an EMR environment level for a system equals or exceeds the corresponding EMR susceptibility level.

Far-Field - Two major criteria exist for defining the far-field region of an emitter. One criterion is based on the maximum allowable phase deviation across an aperture; in this case a value of $2D^2/\lambda$ is commonly used, where D is the maximum aperture dimension and λ is the longest wavelength under consideration. The other criterion concerns the distance required for the radiation fields to predominate over the inductive and electrostatic fields, and this distance is commonly taken to be three to five wavelengths from the radiating antenna.

Field Strength - The magnitude of the electric or magnetic field vector, measured in V/m or A/m, respectively. At frequencies above about 100 MHz, field strength in the far field is conventionally specified as a power density (see Power Density).

Hardening - The use of techniques and devices designed to improve the susceptibility characteristics of a device, subsystem, or system.

Near-Field - The region between an emitter and its far-field region, the value of which depends on the particular criterion being considered (see Far-Field).

Plane Wave - An electromagnetic wave whose equiphase surface approproximates a planar wavefront.

Platform - As used in this handbook, the platform is considered to be synonymous with the delivery aircraft.

Power Density - The rate of energy flow per unit area (may be expressed as W/m^2, mW/cm^2, dBm/cm^2, etc...). Power density may be used to define the intensity of an electromagnetic environment.

Radiation - The emission of energy in the form of electromagnetic waves.

Rectification - As used in this handbook, rectification is the basic interference mechanism whereby the modulation of an RF signal is detected (rectified) in a nonlinear element.

Subsystem - A collection of devices or equipments designed and integrated to function as a major part of a system and to perform an operational function or functions.

System Acquisition Cycle - The time period beginning with the concept development phase and extending to the completion of full-scale production of a system.

System Life Cycle - The time period beginning with the concept development phase of a system and extending until the system has been phased out of inventory.

System - A composite of subsystems and equipments capable of performing an operational role.

Survivability - The ability of a system to perform its required mission during or after its exposure to an EMR environment.

Tailoring - The tailoring of specifications and standards is the process by which the requirements of a specification document are adapted to the peculiarities, operational requirements, or characteristics of a system being developed and the electromagnetic environment in which the system must operate.

Theatre - A geographical area of operation (i.e., European, Mediterranean, Pacific, etc.).

Waiver Analysis - Determination of the impact of relaxing particular specifications or requirements on overall system performance.

Wave Impedance - The ratio of the electric to magnetic field vectors at a particular point in space.

20. ACRONYMS.

CDR	Critical Design Review
CW	Continuous Wave
DCP	Decision Coordinating Paper
DID	Data Item Description
DoD	Department of Defense
E^3	Electromagnetic Environment Effects
ECAC	Electromagnetic Compatibility Analysis Center
ECCM	Electronic Counter Counter Measures
ECM	Electronic Counter Measures
ECP	Engineering Change Proposal
EED	Electro-Explosive Device
EMC	Electromagnetic Compatibility
EMCAB	Electromagnetic Compatibility Advisory Board
EME	Electromagnetic Environment
EMI	Electromagnetic Interference
EMP	Electromagnetic Pulse
EMR	Electromagnetic Radiation
EMS	Electromagnetic Susceptibility

EMV	Electromagnetic Vulnerability
ERP	Effective Radiated Power
FEBA	Forward Edge of the Battle Area
GEMACS	General Electromagntic Model for the Analysis of Complex Systems
HERO	Hazards of Electromagnetic Radiation to Ordnance
IAP	Intrasystem Analysis Program
IEMCAP	Intrasystem Electromagnetic Compatibility Analysis Program
MENS	Mission Element Need Statement
MOA	Memorandum of Agreement
MOM	Method of Moments
NCAP	Nonlinear Circuit Analysis Program
PDR	Preliminary Design Review
PM	Program Manager
PMD	Program Management Directive
POE	Points of Entry
PRF	Pulse Repetition Frequency
PW	Pulse Width
RADC	Rome Air Development Center
RF	Radio Frequency
RFP	Request for Proposal
SON	Statement of Need
SOW	Statement of Work
SPICE	Simulation Program with Integrated Circuit Emphasis
SPO	System Program Office
STS	Stockpile-to-Target Sequence
TEMP	Test and Evaluation Master Plan

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