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ELECTRONICS INSTALLATION AND MAINTENANCE BOOK

ELECTROMAGNETIC INTERFERENCE REDUCTION

DEPARTMENT OF THE NAVY
NAVAL SHIP ENGINEERING CENTER

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PREFACE

POLICY AND PURPOSE

The Electronics Installation and Maintenance Book (EIMB) was established as the medium for collecting, publishing, and distributing, in one convenient source document, those subordinate maintenance and repair policies, installation practices, and overall electronic equipment and material-handling procedures required to implement the major policies set forth in Chapter 9670 of the Naval Ships Technical Manual. All data contained within the EIMB derive their authority from Chapter 9670 of the Naval Ships Technical Manual, as established in accordance with Article 1201, U. S. Navy Regulations.

Since its inception the EIMB has been expanded to include selected information of general interest to electronic installation and maintenance personnel. These items are such as would generally be contained in textbooks, periodicals, or technical papers, and form (along with the information cited above) a comprehensive reference document. In application, the EIMB is to be used for information and guidance by all military and civilian personnel involved in the installation, maintenance, and repair of electronic equipment under cognizance, or technical control, of the Naval Ship Systems Command (NAVSHIPS). The information, instructions, and procedures, in the EIMB supplement instructions and data supplied in equipment technical manuals and other approved maintenance publications.

ORGANIZATION

The EIMB is organized into a series of handbooks to afford maximum flexibility and ease in handling. The handbooks are stocked and issued as separate items so that individual handbooks may be obtained as needed.

The handbooks fall within two categories: general information handbooks, and equipment-oriented handbooks. The general information handbooks contain data which are of interest to all personnel involved in installation and maintenance, regardless of their equipment specialty. The titles of the various general information handbooks give an overall idea of their data content; the General Handbook includes more complete descriptions of each handbook.

The equipment handbooks are devoted to information about particular classes of equipment. They include general test procedures, adjustments, general servicing information, and field change identification data.

All handbooks of the series are listed below with their NAVSHIPS numbers. The NAVSHIPS numbers serve also as the stock numbers to be used on any requisitions submitted.

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EIMB Section Dividers	0967-000-0009

PREFACE

INFORMATION SOURCES

Periodic revisions are made to provide the best current data in the EIMB and keep abreast of new developments. In doing this, many source documents are researched to obtain pertinent information. Some of these sources include the Electronics Information Bulletin (EIB), the Naval Ship Systems Command Technical News, electronics and other textbooks, industry magazines and periodicals, and various military installation and maintenance-related publications. In certain cases, NAVSHIPS publications have been incorporated into the EIMB in their entirety and, as a result, have been cancelled. A list of the documents which have been superseded by the EIMB and are no longer available is given in Section 1 of the General Handbook.

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NOTES

SECTION 1 – PURPOSE AND SCOPE

1.1 PURPOSE

This book is a source of information for persons involved in efforts to achieve electromagnetic compatibility among shipboard electronic systems.

1.2 SCOPE

Significant electromagnetic interference (EMI) sources, methods of recognizing them, and methods of reducing interference are discussed in this book. Hull-generated intermodulation interference (rusty bolt effect), a relative newcomer to the list of interference offenders, is treated in some detail.

Discussions throughout this manual are directed more toward field techniques than initial equipment design. Conceptional design to meet operational requirements is a very broad and complex field that is not within the scope of this book. Ultimately, shipboard interference problems will require a total approach embracing all aspects of electronics design and installation – from initial design concepts through equipment manufacture to the equipment operation.

SECTION 2 – ELECTROMAGNETIC INTERFERENCE

2.1 BACKGROUND

The multiplicity of electronic systems developed during World War II brought to light a problem heretofore given little consideration, that is, electromagnetic interference (EMI). Since that time, rapid advances have been made in all phases of design and implementation of sophisticated electronic systems. Ships and aircraft have become mere platforms to house the multitude of highly sensitive and complex electronic weapons and communication systems.

Although acceptable limits of undesired radiation from each piece of equipment have been established, too little effort has been directed toward achieving compatibility among systems. The consequences were inevitable and, to an extent, predictable – mutual interference between the prolific radiators and susceptible equipments. Such interference threatens the very existence of man's ability to communicate via electronic means.

2.2 DEFINITIONS

EMI – An electromagnetic or electrostatic disturbance that causes malfunctioning or an undesirable response in one or more electronic receivers, or a condition which does not meet the requirements of interference tests.

Electronic equipment – Any item serving functionally by electromagnetically generating, transmitting, conveying, acquiring, receiving, storing, processing, or utilizing information in the broadest sense. Communications, radar, sonar, countermeasures, navigation, control, armament, computers, and test equipment are examples.

Electrical equipment – Equipment that, by design, is not intended to generate radio frequency energy. Examples are: electric motors, office equipment, laundry and repair shop equipment, power supplies, ultrasonic equipment, medical equipment and fluorescent lights.

Electronic space – A space used primarily to house installed units of electronic equipment. Spaces include radar rooms; sonar equipment and control rooms; ECM rooms; radio central; radio transmitter rooms (VHF-UHF); crypto rooms; missile director control rooms; chart rooms; WDE, CIC, ASW control rooms; television equipment spaces; CATCC, and computer equipment spaces.

Electrical space – A space used primarily to house installed units of electrical equipment. Spaces include 60-and 400-cycle motor-generator (MG) rooms, power switchboard distribution rooms, gyrocompass rooms, and IC spaces. Spaces occupied by both electrical and electronic equipments are classified as electronic spaces.

Grounding (ground) – The process of physically providing a metallic surface with a low resistance or impedance path to ground potential. (On metallic hull ships, the ship hull is considered to be ground potential.)

Bonding (bond) – The process of physically connecting or bridging two metallic surfaces to provide a low impedance path for RF current.

Direct bond – The joining of two metallic surfaces by welding, brazing, or soldering.

Indirect bond – All bonds achieved by use of interconnecting straps, cables, etc.

Intermodulation – The production, in a nonlinear element, of frequencies corresponding to the sums and differences of the fundamental and harmonics of two or more frequencies.

Topside areas – On ships having flight or antenna decks, such as carriers and some communication ships, those areas on and above the galley deck that are exposed to weather. On all other ships, those areas on and above the main deck that are exposed to weather.

Below decks and interior areas – The inner or inside area of a ship. It consists of all areas in the interior of a ship from the lowest to the highest level. These areas include pilot house, flag bridge, and air control areas.

Conduit – A metallic structure containing one or more ducts used to protect and support wires and cables of electrical and electronic systems and to provide sufficient attenuation to radio frequencies for reduction of RF interference.

Transmission lines – Structures forming a continuous path from one place to another, for directing the transmission of electric or electromagnetic energy along this path. The term transmission lines includes such items as telephone lines, power cables, waveguides, and coaxial cables.

SECTION 3 – GENERAL SOURCES OF EMI

There are three sources of Electromagnetic Interference: (1) man-made, (2) natural, and (3) inherent. The first of these is the major source of concern and, unless it is given constant attention, it can completely destroy man's ability to transmit information via electronic means.

3.1 MAN-MADE INTERFERENCE

The most common sources of man-made interference are:

a. Transmitters – communications, radar, active ECM, TACAN, IFF sonar, and other types
b. Electrical controllers
c. Motors and generators
d. Engines and igniters
e. Lighting
f. Power lines
g. Welders
h. Local oscillators (in heterodyne receivers)

3.1.1 Transmitters – Communications, Radar, and Other Types

In theory, a communications, navigation, radar, or other type of transmitter is required to radiate energy only over the band of frequencies necessary to convey the intelligence that it is processing. In practice, however, a transmitter may emit energy at a great number of spurious frequencies, and thus create a potential interference problem.

3.1.1.1 Communications Transmitters. These transmitters can produce spurious emissions from:

a. Overdriven amplifiers.
b. Frequency multiplier stages.
c. Sideband splatter caused by over-modulation, excessive modulator bandwidths, or modulator non-linearity.
d. Modulator noise.
e. Transmitter intermodulation and cross modulation caused by interaction between two or more transmitters.

3.1.1.2 Radar Transmitters. These transmitters can produce spurious emissions by:

a. Modulator pulsing – the extremely short pulses used in radar result in a frequency spectrum which can extend into the radio frequency region (see figure 9-1).
b. Magnetron moding.
c. Magnetron drift.
d. Sideband power caused by pulse modulation.
e. Harmonic outputs that may be only 40 dB below the fundamental signal level.
f. Induced arcing between the armor of cables illuminated by the radar antenna.
g. Arcing in PA stage, waveguide rotary joint, or antenna.

3.1.2 Electrical Controllers

The controller equipment associated with electrical installations has as its function the control of the voltage and/or current applied to (or from) an electrical system. The control system may be a refined device with continuous, precise settings, or it may be a simple on-off switch. Since continuous-type control results in slow variations of current waveform, there is relatively little interference. The make-break contact results in sharp changes in current waveform and causes severe interference over a wide frequency band.

3.1.3 Motors and Generators

Electrical motors and generators constitute the most common sources of electromagnetic interference. Rotating electrical machinery interference is generated by the following:

a. Arcing from brushes to commutator segments in DC motors or generators, or from brushes to slip rings in AC motors or generators.
b. Induction of interference into nearby electronic equipment from the high energy magnetic fields associated with rotating machinery.
c. Ripple which is present to some degree in the output of all DC generators.
d. Slot harmonics appearing as harmonic frequencies in the output of generators. These harmonics are the results of lack of uniformity in the magnetic field caused by the effect of the armature slots on the distribution of the magnetic flux.

3.1.4 Engines and Igniters

Ignition systems of gasoline engines can be sources of interference with an energy distribution up to 500 MHz. Igniters for jet engines, using high voltage and high current, can cause severe interference.

3.1.5 Lighting

The steep waveforms associated with the operation of fluorescent, mercury, sodium arc, and vapor lamps can cause broadband interference.

3.1.6 Power Lines

This type of interference is caused by electromagnetic coupling between wires or the degraded condition of conductors, insulators, and associated hardware.

3.2 NATURAL INTERFERENCE

Interference caused by natural phenomena, such as electrical storms, snowstorms, rain particles, and interstellar radiation, is designated as atmospheric noise or static. This interference is characterized by the following types of noise in a receiver output:

a. Impulses of high intensity, occurring intermittently, caused by local thunderstorms.

b. A steady rattling or cracking produced by distant thunderstorms.

c. A continuous noise caused by the impact of charged particles against an antenna. This is known as precipitation static.

d. A steady hiss type of static observed at high frequencies, apparently having an interstellar origin.

3.3 INHERENT INTERFERENCE

A certain amount of interference inherent in receiving equipment is caused by thermal agitation of electrons in the circuit resistances. Random motion of free electrons develops voltages in the conductors and components, and these voltages contribute to receiver background noise.

In vacuum tubes, there is some fluctuation in the large number of electrons traveling from the cathode to the plate. This randomness, commonly called the "shot" effect, is responsible for additional background noise in receivers.

SECTION 4 – TYPES OF EMI

EMI can be classified according to spectrum distribution as either narrowband or broadband.

4.1 NARROWBAND INTERFERENCE

Narrowband interference consists of a single frequency or a narrow band of frequencies that occupy little space in the receiver passband. Another way of describing narrowband signals is that they are unaffected by the bandwidth of the receiving device; i.e., they are narrow with respect to the receiver bandwidth. Sometimes an interfering narrowband signal can be prevented from causing an undesired response by tuning the susceptible receiver away from it; however, even out-of-band signals may interfere if they are strong enough to force their way past the attenuation of the preselector or RF stages.

4.2 BROADBAND INTERFERENCE

Broadband interference is not of a discrete frequency; i.e., it occupies a relatively large part of the radio frequency spectrum. It is usually not possible to tune away from it since it is much wider than either the assigned frequency channel or the receiver bandwidth.

Broadband emissions have a spectral distribution sufficiently broad, uniform, and continuous so that the response of the measuring receiver in use does not vary significantly when it is tuned over a wide band.

This type interference is usually caused by arcing or corona and it is misleading in that it is sometimes misinterpreted by inexperienced operators as a high background noise level.

SECTION 5 – METHODS OF COUPLING EMI

Interference is transferred from an interfering source to the affected equipment by two general methods: (1) conduction and (2) radiation.

5.1 CONDUCTED EMI

Conducted EMI involves the transfer of undesired energy through conductors between a source of interference and a susceptible device. The complete circuit required for flow of interference conduction currents can be made up entirely of metallic conductors, or the return path may be through earth ground. The most common, and most important, paths for conduction currents are:

a. Power supply cables
b. Control and accessory cables
c. Grounding systems
d. Transmission lines

5.1.1 Factors Affecting the Degree of Conducted EMI

The type of circuit, frequencies involved, power level, and amount of capacitive and inductive coupling between parts of the circuit all have a bearing on the generation of interference that can be conducted from one equipment to another. Different types of circuits (such as pulse, CW, audio) are prone to generation of interference in different ways and to different degrees. The care taken with lead dress, filtering, and shielding affects the amount of interference conducted from an equipment and also the degree of susceptibility of an equipment to this interference. In addition to the equipment features that may be responsible for conducted interference, energy radiated from high-power transmitters can be coupled into interconnecting cables and conducted into equipment.

5.1.2 Containment of Conducted EMI

The most effective approach to avoid conducted interference is through equipment design and installation practices that prevent it. Among the methods used are:

a. Filtering to shunt the interference to ground.
b. Shielding to confine interfering signals.
c. Proper grounding of circuits, shields, and interconnecting cabling.
d. Use of isolation transformers for circuit (and ground) isolation.
e. Separation of low-level signal cables from other cables.

5.2 RADIATED EMI

Radiated EMI is any interfering signal transferred through space by an electromagnetic field. The radiated field represents energy that escapes from a source and spreads out in free space according to the laws of wave propagation.

5.2.1 Factors Affecting Strength of Radiated Field

An electromagnetic field is generated whenever current flows in a conductor and some energy in this field is radiated.

Several factors determine the strength of the radiated signal.

a. Amount of current flow in the conductor from which the field radiates.
b. Efficiency of the conductor as an antenna, e.g., physical configuration and orientation, and amount of shielding offered by nearby metal items.
c. Frequency of the current waveform causing the field. Very little energy is propagated at frequencies below about 15 kHz; progressively stronger fields are radiated as the frequency of the current causing the field increases.

5.2.2 Containment of Radiated EMI

Suppression methods for radiated EMI include confinement of the interference to the source to prevent its being radiated. Metal shielding is used around the EMI source to reduce the strength of any stray radiation and prevent it from reaching susceptible equipments. It is usually less expensive and less difficult to suppress interference at the source than to eliminate it once it has been radiated.

SECTION 6 – INTERFERENCE FROM ELECTRICAL DEVICES

6.1 GENERATION OF INTERFERENCE BY ROTATING MACHINES

Motors and mechanically-driven generators are included in the large category of rotating machinery. In this category are the rotary inverter (a DC-to-AC converter), the dynamotor (a DC motor and generator, operated from a single magnetic field, that functions either to step up or to step down DC voltages), and the alternator (an AC generator). These machines operate on the principle of converting mechanical energy to electrical energy, or electrical energy to mechanical energy. Because of the nature of their operation, these equipments create various forms of radio interference caused by the following.

6.1.1 Arcing

Arcing occurs when brushes sweep over commutator segments in DC motors and generators. Although microscopic arcing results from brush action on the slip rings of an AC motor or generator, it is much less intense than that which occurs in the DC type and is normally not a problem.

6.1.2 Induction

Interference voltages may be induced in communications equipment by action of the magnetic fields associated with rotating electrical machines.

6.1.3 Ripple

The characteristic ripple found in the output of all DC generators can result in interference in communications equipment if it is not adequately filtered.

6.1.4 Slot Harmonics

The presence of harmonics in the output of ship service generators is usually caused by a non-uniform field brought about primarily by the effect of the armature slots on the distribution of the magnetic flux. This causes harmonics to appear in the output. Harmonics also are produced by nonlinearities existing in the electrical system. Therefore, loads as well as generators can be the source of harmonic frequencies. The amplitude of these harmonics is dependent upon the extent of flux variations produced by the armature slots. Existing specifications for generators state that the RMS value of any one harmonic shall not exceed 6 percent of the nominal output voltage, and the RMS value of the sum of the harmonics shall not exceed 10 percent of the nominal output voltage. Severe interference may be encountered in facsimile, or similar equipment, if the ship service generators produce a harmonic within the audio frequency band used in the transmission process.

6.2 METHODS FOR SUPPRESSING RADIO INTERFERENCE GENERATED BY ROTATING MACHINES

The following methods of suppressing interference must be adapted to the requirements of the particular equipment to which they are being applied.

6.2.1 DC Motors and Generators

Interference from DC motors and generators may be suppressed by one or more of the following methods:

a. Capacitors installed at the brushes or other accessible terminals. (See figure 6-1 for illustration of suppression capacitors connected from brushes to ground.)

b. Shielding provided by the housing. The housing should have screened louvers. If there is an inspection plate, it should be fitted tightly to the housing.

6.2.2 Alternators

Interference from alternators may be suppressed by one or more of the following:

a. Capacitors installed at the slip ring brushes.

b. Capacitors installed at the exciter brushes.

c. Feedthrough capacitors installed in the output leads (preferred) or bypass capacitors connected to the brush terminal outlet inside the shield.

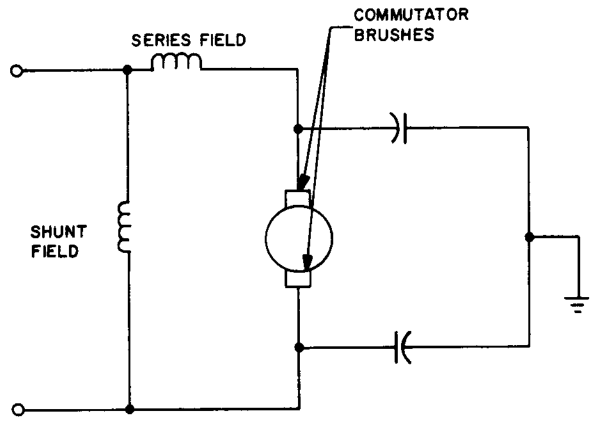


Figure 6-1. Suppression Capacitors Connected from Brushes to Ground

6.2.3 Synchronous Motors

Interference from large synchronous motors with slip rings and brushes similar to those of alternators are suppressed by the same techniques as those used for alternators. Since small synchronous motors do not have slip rings, they are not normally a source of interference.

6.2.4 Dynamotors

Interference from dynamotors may be suppressed by the same techniques as those used for generators and motors. Feedthrough capacitors should be mounted through the shield on both the input and output leads.

Semiconductor rectifiers and varistors are used as voltage surge suppressors also, especially in circuits containing relay coils. Semiconductors or varistors, when placed across the relay coil, will help prevent the voltage across the coil from building up appreciably as a result of the collapsing magnetic field when the circuit is broken. A capacitor or choke also can be used to suppress interference-causing transients across the relay coil. The suppression components can be connected across the relay coil, the switch, or the power line. See figure 6-2 for preferred location of filters in make-and-break circuits.

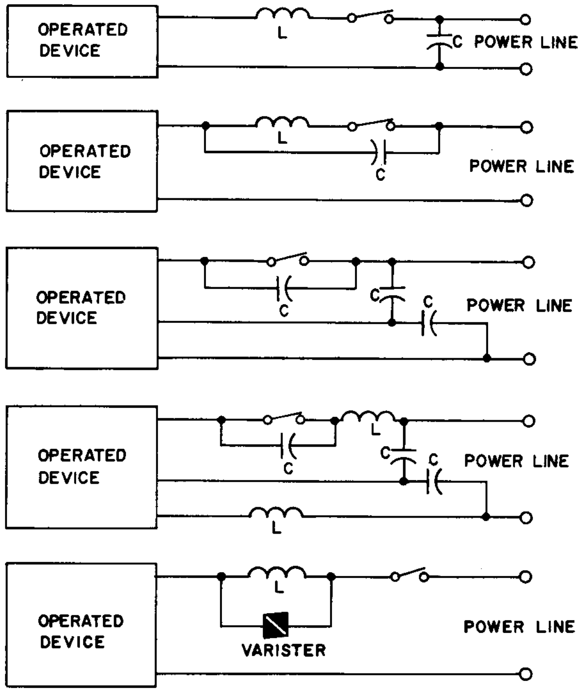


Figure 6-2. Filters Used With Make-and-Break Circuits

SECTION 7 – COMMUNICATIONS EQUIPMENT INTERFERENCE

7.1 INTERFERENCE PRODUCED BY TRANSMITTERS

7.1.1 General

Transmitters usually are not susceptible to interference, but they may cause considerable interference to other equipments. The equipment designer provides for adequate interference suppression in the manufacture of the equipment; however, there is much the ultimate user can do to ensure that design capabilities are realized. Too often, the designer's best efforts are negated by improper installation or operating practices. The following discussion of interference from transmitters is slanted toward installation maintenance personnel rather than equipment designers.

7.1.2 Spurious Emissions

Any emission from the generating source other than the one meant to transmit intelligence is a spurious emission that can cause interference. Spurious emissions are usually categorized as:

a. Broadband emissions caused by arcing.
b. Spurious sidebands (splatter).
c. Harmonic radiation.
d. Spurious outputs caused by parasitic oscillations.
e. Spurious outputs from oscillator, frequency multipliers, and frequency synthesizer stages.
f. Transmitter noise.
g. Cross modulation and intermodulation.

7.1.2.1 Broadband Emissions. These emissions are recognized by their distribution across a wide portion of the tuning range of the affected receiver. They have a characteristic irregular, ragged hash sound in the receiver output and are usually caused by arcing either in the transmission line, tuner/coupler, transmitter, or in the rigging and structural appendages of the ship.

High-power shipboard transmitters induce very high RF currents and voltages in surrounding structures. When discontinuities are present in these structures, there is a strong tendency toward arcing which generates broad-band interference. Cable armor can arc to a ship hull even with the armor grounded at regular intervals. Arcing or corona can often be detected visually at night when the transmitters are radiating.

One solution to this problem is to eliminate discontinuities such as corroded junctions, stray metal objects, long lengths of guy wires, unused cables and antennas, and the like. The ultimate goal is to make the entire topside area of the ship a single conducting structure by welding, brazing, bonding, and other methods. Interference from arcing is especially misleading in that it is sometimes misinterpreted as a high ambient noise level and no remedial actions are taken.

7.1.2.2 Spurious Sidebands (Splatter). When a carrier is modulated, either in frequency or amplitude by an intelligence-bearing signal, a group of new frequencies is generated around the carrier frequency. These frequencies are caused by, and are a function of, the modulating or intelligence signal. The usable and important range of frequencies necessary to transmit intelligence determines the bandwidth of a signal. Sidebands occurring outside this necessary bandwidth are undesirable and should be eliminated. Spurious sidebands may arise from improperly tuned circuits, from over-modulation, or from faulty equipments.

This form of interference is prevalent to such a degree in naval communication complexes that it is almost accepted as normal. In addition to creating wasteful and undesirable sidebands, splatter caused by overmodulation distorts the desired transmission. The received signal has a characteristic "mush" sound instead of the crispness associated with properly operated transmitters.

In the majority of cases, the splatter is caused by overdriving the intermediate or final output stage. Many operators simply do not follow published procedures for tuning, loading, and operating the present generation of transmitters. For example, when the output power meter on the AN/URC-32 transmitter deflects to about 175 watts on voice peaks (in the SSB mode), the transmitter is putting out its rated 500 watts peak. Some operators still insist on "hitting the 500 mark" on the output power meter by advancing the exciter RF gain control. This practice creates intolerable distortion and many interference signals. The only solution, aside from completely automated operation, is for supervisory personnel to insist on a demonstration of correct tuning, loading, and operating procedures by equipment operators.

Another reason for over-modulation stems from different speech characteristics of individual operators. One operator might speak softly into a microphone located several inches away; another operator may practically shout into the microphone located almost at his lips. This leads to widely varying speech levels throughout the transmitter and, consequently, varying average power output levels. If the transmitter were properly adjusted for the softer speaking operator, then overmodulation would result when a louder speaking operator used the transmitter. Voice compression circuits and transmitter gain control circuits are standard accessories on present generation transmitters and full advantage should be taken of these aids to reduce radiated output power fluctuations caused by individual operator characteristics. Usually, these circuits are automatic in operation once they are switched into the system, but they must be properly adjusted and maintained.

7.1.2.3 Harmonic Radiation. Harmonic radiation is present to some degree in the output of all transmitters because of nonlinearity of the power output stage. Harmonics, integral multiples of the fundamental frequency, are undesirable for two reasons: They are sources of interference, and they waste power. Power transmitted in harmonics contributes nothing to desired communications.

Good design and correct tuning and loading usually ensure sufficient harmonic attenuation. Military specifications now require that the second harmonic level be at least 60 dB below the fundamental; third and higher order harmonics at least 80 dB below the fundamental. The use of multicouplers will attenuate further transmitter-generated harmonics, thereby reducing their importance as sources of interference.

7.1.2.4 Spurious Outputs Caused by Parasitic Oscillations. Parasitic oscillations cause emission at radio frequencies which are neither harmonics of the fundamental nor intermodulation product frequencies. Parasitic emissions occur when a circuit is self-excited and goes into oscillation at a frequency other than the desired one. These emissions include shock excitation due to internal transient phenomena and unintential excitation of circuit components by the carrier signal.

The usual cause of parasitic oscillation, in a well designed circuit, is improper lead dress following component replacement. The technician should ensure that component placement and wire positions are not haphazardly disturbed during troubleshooting or alignment. Parasitic oscillations also occur in the output stage of a transmitter because of improper neutralization following tube replacement in that stage.

7.1.2.5 Spurious Outputs from Oscillator, Frequency Multiplier, and Frequency Synthesizer Stages. In some transmitter systems, a mixer is used to produce the system fundamental output frequency or a submultiple thereof. When this is the case, spurious signals are generated because of nonlinearity of the mixer; hence, tuned circuits are required following the mixer stage to attenuate undesired frequencies. Usually four or five tuned circuits must be provided.

A second type of signal generation, utilizing frequency multiplier stages and a single master oscillator, is used in many transmitting systems. Since frequency multipliers are nonlinear, harmonics of the master oscillator frequency are generated. Unless extremely high Q circuits are used in the multiplier stages of the transmitter, high levels of harmonics of the master oscillator frequency will be emitted along with the fundamental frequency of the transmitter.

7.1.2.6 Transmitter Noise. Transmitter noise is generated in the various RF stages, together with noise from the audio system and power supply. It is inherent in the radiation process, evenly distributed in the spectrum near the carrier frequency, and its effect is similar to atmospheric noise.

In general, this noise does not degrade the desired transmitter signal appreciably because the depth of modulation or deviation is small compared to the desired signal modulation. It is usually ignored as a cause of interference because adjacent receivers are not tuned near enough to the transmitter frequency for its effects to be noticed.

7.1.2.7 Cross-Modulation and Intermodulation. Although cross-modulation and intermodulation occur in both receivers and transmitters, only transmitters are considered in the following discussion. In both types of modulation, the mechanism responsible involves two or more signals present simultaneously in a nonlinear element. Both intermodulation and cross-modulation can occur at the same time.

Cross-modulation involves the transfer of modulation from one carrier to another. Intermodulation involves generation of numerous new frequencies from two or more original signals.

When transmitting antennas are closely spaced, a large degree of coupling exists between them. Powerful off-frequency signals from one transmitter feed back into the final stage of a second via the closely coupled antenna systems. The output tuned coupling network of the second transmitter may not offer sufficient attenuation to the offending signal to prevent it from reaching the nonlinear final stage and interfering with the desired signal.

One solution is to physically separate the transmitting antennas to decrease the transfer of RF energy between them. This approach is limited aboard ship since space is at a premium and other factors must be considered in antenna placement.

The most common solution at present is to use tunable filters (multicouplers) between transmitters and antennas. The additional attenuation offered by such filters to off-frequency signals is usually adequate to prevent them from coupling into a susceptible output stage. Another major benefit derived from multicouplers is that several transmitters at different frequencies can be coupled to one broadband antenna, thereby reducing antenna requirements aboard ship.

7.2 RECEIVER SUSCEPTIBILITY

If a receiver would not respond to any electromagnetic radiation other than the desired signal, there could be no interference to that receiver. All present receiving devices, however, are less than ideal and all are susceptible in varying degrees to undesired signals.

Three general mechanisms by which undesired signals intrude on a selected or desired signal are (1) linear intrusion via normal input terminals, (2) non-linear intrusion via normal input terminals, and (3) intrusion through ports not intended as signal inputs.

7.2.1 Linear Intrusions

The block diagram of figure 7-1 shows the essential elements of a receiver. The superheterodyne circuit, usually employed in current receiver applications, is inherently susceptible to certain frequencies other than the frequency to which it is tuned. Linear intrusion is possible because the receiver acts as a normal bandpass filter which accepts any input containing frequency components in the receiver passband. Unwanted inputs with a spectrum centered at or near the tuned frequency of the RF filter are the usual interference sources. These inputs arise from a variety of sources:

a. Broadband noise arising from natural or man-made sources. The spectrum of such noise is essentially flat over the bandpass of a typical receiver.

b. Signals from communications sources assigned to a frequency at or near the receiver center frequency. When these signals are separated in frequency by an amount less than the receiver bandwidth, the signals are co-channel.

c. Signals from communications sources whose frequencies are separated from the receiver center frequency by more than the receiver bandwidth (adjacent-channel interference).

d. Signals from communications sources assigned to frequencies within one of the internal pass frequencies (IF interference).

e. Discrete (narrow-band) signals generated in nonlinear elements in the shipboard electromagnetic environment. These signals are both intermodulation products and harmonics.

7.2.1.1 Broadband Noise. Typical natural broadband noise sources are thermal noise, shot noise, galactic noise, solar noise, and atmospheric noise. Typical man-made sources include discharges on high-voltage lines and devices, noise from automobile ignition systems, commutator noise, noise in complex switching systems, noise generated by fluorescent lamps, arcing across structural discontinuities located in high-energy radar beams, and arcing of the HPA or waveguide in a radar system. In some of these sources a certain amount of regularity exists. Atmospheric noise bursts sometimes have some coherence because of multiple propagation paths. Furthermore, there are long term fluctuations, depending on the time of day, season, and sunspot cycle, which are roughly predictable. Corona noise on high-voltage lines and fluorescent lamp noise are usually modulated by the power-line frequency. A single, stationary ignition noise source is more or less periodic; noise from one or more randomly passing vehicles or from many stationary vehicles is not periodic.

Remedies for broadband noise that overlaps the receiver bandpass must take advantage of differences between signal and noise characteristics. The most common methods of dealing with noise of a discrete impulse nature are limiting and blanking, both done

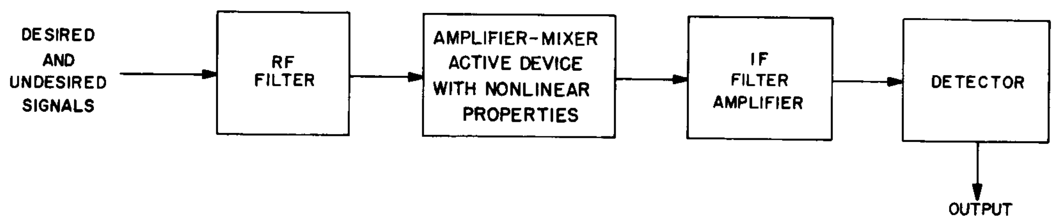


Figure 7-1. Block Diagram of Basic Receiver Elements

before the broadband pulses have been filtered in the IF amplifier. This type of noise, having short duration and large peak value, can be limited to the level of the desired signal or it can be totally blanked out for its brief duration. In either case, the signal is eliminated for the duration of the impulse, with no significant loss of information. A filter preceding the limiter or blanker would widen the interference pulse and make these processors less effective as noise reducers. For example, a typical impulse type noise burst, which may be only one or two microseconds wide at the receiver input, can be lengthened to 100 microseconds or more by the selectivity of the receiver IF amplifier. This wider pulse is more difficult to remove than the narrow pulse. A gap of 100 microseconds in the detector output signal is much more noticeable than a gap of one or two microseconds.

Blanking systems require information concerning the interfering impulses. For example, systems for eliminating the periodic pulses of a nearby radar system may use direct synchronization from the radar source. When there is no access to the source, the receiver itself must sense the pulse in one branch to eliminate it in a second branch.

Broadband interference, such as white noise, that extends throughout the tuning range of a receiver can be suppressed by methods similar to those used against impulse noise, but the emphasis is on signal enhancement rather than noise rejection. Advantage can be taken of known signal characteristics to design modulator and demodulator processes that increase the signal-to-noise ratio. For example, a matched filter detector uses information about the shape of the signal pulses to discriminate between the signal and noise. The operator can increase the signal-to-noise ratio by selecting the minimum receiver bandwidth necessary for satisfactory signal reception.

7.2.1.2 Co-channel Interference. Co-channel interference involves communications systems that are assigned the same, or nearly the same, carrier frequencies. Co-channel interference occurs when the center frequency or carrier of an undesired signal falls in any portion of the passband of another receiver (see figure 7-2). The interfering signal may be a normal authorized frequency; however, since its carrier frequency and some sidebands are within the bandpass of the victim receiver; they will follow a normal signal path and appear as interference. The susceptibility of a receiver to co-channel interference is greater than other types of interference since a co-channel signal is received, amplified, heterodyned, and detected in the same manner as a desired signal.

Co-channel frequencies normally are not assigned except when the possibility of conflict between two systems is remote. For example, co-channel assignments among satellite transmitters, and widely separated transmitter sites can be allowed. The result is that, although the interference susceptibility is greatest in the co-channel mode, frequency management makes the probability of this type of interference remote.

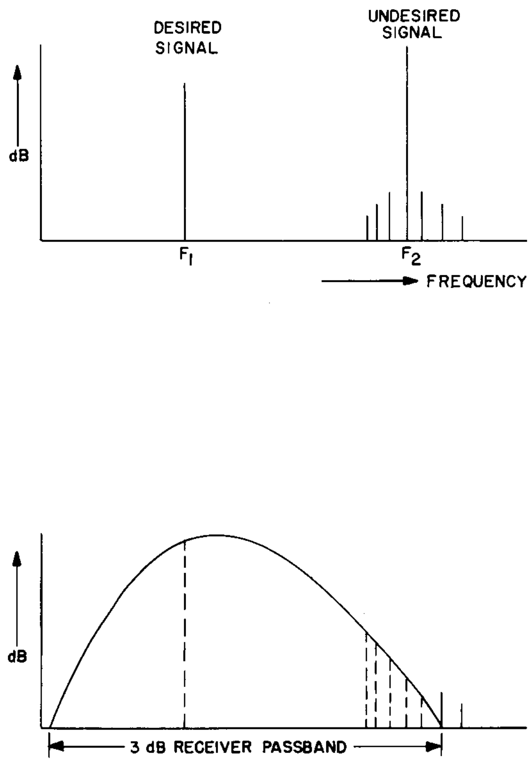


Figure 7-2. Co-channel Interference

7.2.1.3 Adjacent Channel Interference. This type of interference occurs when some sidebands, but not the center frequency or carrier of an undesired signal, are within the receiver bandpass. Adjacent channel interference occurs between communication systems that have been assigned neighboring channels. This is shown in figure 7-3, where energy on the skirt of the adjacent channel signal spectrum overlaps the receiver bandpass. Usually, the sensitivity in this mode is low compared to the in-band sensitivity, but receivers located close to an adjacent-channel transmitter can be exposed to very high levels of unwanted signals.

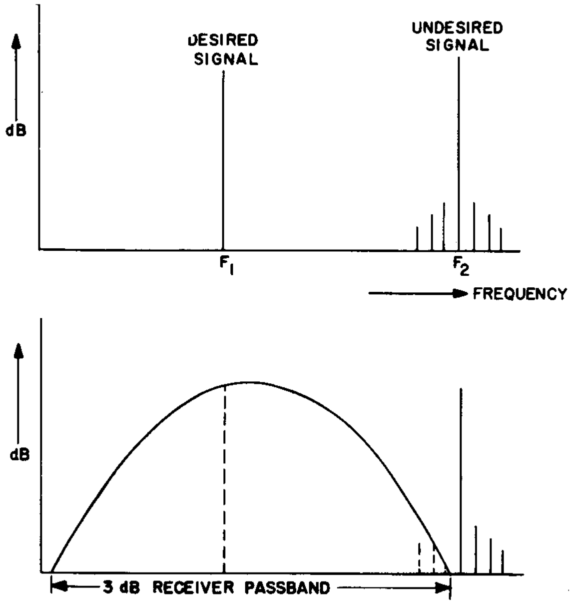


Figure 7-3. Adjacent Channel Interference

7 2.1.4 Image Frequency Interference. In a heterodyne-type receiver, there are two signal frequencies that can combine (heterodyne) with the local oscillator signal to produce the receiver intermediate frequency (IF). One of these signals is the desired, or tuned, signal and the other is called the image frequency.

The image frequency is separated from the desired signal by twice the IF and is located on the side of the local oscillator signal opposite the desired frequency.

To keep an undesired signal at the image frequency from interfering, it is necessary to prevent it from reaching the mixer stage where the unwanted combining takes place. Some receivers employ one or more resonant circuits, amplifier stages, or frequency selective filters between the antenna and the mixer stage. These resonant circuits or filters pass the desired signal and reject or attenuate the image frequency.

Little can be done to a receiver with poor inherent image rejection, but the technician can ensure that the receiver is carefully aligned, especially the front end. When additional image rejection is mandatory, an external preselector can be used with the receiver.

7.2.1.5 Intermediate Frequency Interference. One final mechanism involving linear phenomena is the penetration of unwanted signals that are centered at any pass frequency within the receiver. For instance, a large amplitude signal centered at one of the IF amplifier frequencies may manage to get by the input selective circuits to the IF amplifier. Once there, the unwanted signal proceeds through the rest of the receiver in a normal manner. To correct this problem, the input circuit selectivity and/or stray paths to the sensitive circuits must be controlled. As a rule, the first IF amplifier frequency is the most vulnerable, but consideration should be given to any other internal passbands used.

A practical method of reducing this type of interference is to insert a series or parallel resonant circuit, tuned to the susceptible frequency, in the RF stage. This will attenuate or reject any undesired signal at the intermediate frequency.

7.2.2 Nonlinear Intrusion

Interference of the nonlinear intrusion type occurs when a strong unwanted signal penetrates the input filter circuits and encounters a nonlinear element such as the mixer or an overloaded RF amplifier stage. The unwanted signal mixes the desired signal to produce many new signals. If any of these new frequencies fall within the IF passband, they will be processed along with and interfere with the desired signal. The unwanted mixing can also take place in nonlinear elements external to the receiver. Any in-channel new signals, radiated from the nonlinearity, can cause interference once they are processed in the receiver. This latter phenomenon, hull-generated interference, is discussed in Section 8.

Because a strong signal is required to cause interference by nonlinear intrusion, nearby transmitters are usually the only significant sources. Sources not designed to radiate, spurious outputs of transmitters, and broadband noise sources are rarely found to be major contributors to nonlinear intrusion.

7.2.2.1 Spurious Responses. The spurious responses of a receiver are the result of (1) nonlinearity in an early stage which gives rise to harmonics and intermodulation products of incoming signals, (2) nonlinearity in the mixer which results in oscillator and signal harmonics, and (3) frequency multiplication in the local oscillator and its related circuits.

A different set of possible spurious response frequencies exists at each receiver frequency. Whenever one of these spurious frequencies coincides with the intermediate frequency, it becomes a potential source of interference. Maximum conversion transconductance at the fundamental frequency is obtained with a large oscillator input, but this results in more than a proportionate increase in the conversion transconductance to harmonics.

The relative levels of interference and signal magnitude are not easily calculated since, up to the mixer, the gains of all possible interfering signals at all frequencies of tuning are not ordinarily known.

Furthermore, the oscillator level, plus harmonic and subharmonic content as a function of tuned frequency, are not ordinarily known. It is more common to measure the intensity of spurious responses than to calculate them. The usual procedure is to set the tuning control of a receiver to three points in each band and to tune an input signal generator through the regions of potential response at each point. The points are the band center and the band edges. The ratio of signal-to-interference carrier-voltage levels at the input required to give equal outputs is the observed quantity. The ratio may depend on the input level.

7.2.2.2 Intermodulation and Cross-Modulation. Intermodulation and cross-modulation, discussed under transmitter mechanisms (paragraph 7.1.2.7), are important in receivers also. The mechanisms are essentially the same. Intermodulation or cross-modulation in receivers, however, occurs when two or more signals are present simultaneously at the input. Intermodulation results in the generation of many new frequencies; cross-modulation involves the transfer of information from an undesired to a desired carrier. In either case, it is caused by nonlinearity in a circuit near the receiver input.

Intermodulation is the most important of these mechanisms. It becomes especially important when a range of frequencies is subdivided into separate communication channels and when a number of these closely spaced channels are used simultaneously. When these signals meet in a nonlinear element, new signals are generated that sometimes are spaced closely in frequency to the original signals and fall within the tuned passband of the receiver.

With three channels at frequencies f_1, f_2, f_3, intermodulation products (that is, those products near in frequency to the original generating frequencies but not coincident with them), are:

a. $f_1 + f_2 - f_3$	e. $2f_2 - f_1$
b. $f_1 - f_2 + f_3$	f. $2f_2 - f_3$
c. $2f_1 - f_2$	g. $2f_3 - f_1$
d. $2f_1 - f_3$	h. $2f_3 - f_2$

The above examples cite only a few of the possible combinations – there are many more.

7.2.2.3 Desensitization. Desensitization refers to a reduction in overall receiver gain or sensitivity, or both, when a large unwanted signal enters the receiver. The interfering signal alone may not even be heard if it is either unmodulated or is modulated in a way to which the receiver is not receptive. Typical mechanisms of desensitization are:

a. A large-amplitude, unwanted carrier passing through a receiver with automatic gain control will sometimes depress the receiver gain with regard to a desired signal. The gain control voltage is determined by the carrier level at the detector input and any signals there will affect the gain. In envelope detector systems, a large undesired signal will tend to "capture" the detector.

b. Desensitization also occurs when unwanted strong signals overload one of the early receiver stages. This can occur even with unwanted signals at frequencies relatively far from the tuned frequency of the receiver because of the large bandwidth of the early stages. The mechanism varies according to the circuit. The unwanted signal may overload the first active device, causing periods of saturation and cutoff during which the signal is suppressed. In systems having R-C networks for bias generation, or automatic gain control in the early stages, overloading will cause a change in bias and reduction in gain. The changed bias is sustained for a time, dependent upon the R-C time constant and the peak value of the undesired signal.

c. Low duty-cycle pulsed signals, such as radar emissions, can be especially troublesome because of their large peak amplitudes. Microwave radar interference to lower frequency UHF and VHF communications receivers is common, mostly because the input tuned circuits of the receivers are virtually useless as filters for microwave energy. Large-amplitude pulsed signals can cause input circuit overloads at the first amplifier stage. As a result, the amplifier may be driven into nonlinear operation, or, in the extreme, it may be driven between saturation and cut off. The desired signal is altered at the pulse rate, or, in effect, modulated by the pulse signal, and the modulation can produce numerous frequencies that lie in the passband of following receiver stages.

d. An unwanted signal that is normally rejected by the receiver will sometimes cause interference by transferring its information sidebands to the carrier of the desired signal. Overdriven amplifiers increase the likelyhood of cross-modulation, but any nonlinearity of the input-output characteristic of an active device may be a cause of cross-modulation even though unwanted signals are not large enough to cause overload directly at the input.

e. Diode mixers act naturally as harmonic mixers and thus create spurious frequencies. They are also subject to desensitization effects, particularly in microwave receivers, e.g., radar receivers, where the mixer is the first electronic device following the input terminals. The conversion transconductance is altered by the presence of a large unwanted signal, and the effective impedance of the mixer IF output is altered by the unwanted signal. The IF input and the mixer impedances become mismatched when the unwanted signal appears. Tests show a drop in conversion efficiency of 3 dB for an unwanted sinusoid equal in amplitude to the local oscillator signal. The larger the local oscillator input power to the mixer, the larger the unwanted signal that can be tolerated; however, harmonic conversion transconductance becomes significant with larger local oscillator inputs. A compromise is needed, therefore, between high local oscillator power to minimize desensitization potential and low local oscillator power to minimize spurious response potential.

7.2.2.4 Interference Following the Detector. Intrusion of unwanted signals by the various mechanisms just described can have quite variable effects on the signal output of a receiver. The effect of desensitization is the most straightforward since it simply reduces the desired signal level. Other effects of unwanted signal intrusion are more difficult to predict because they depend upon the type of intruding signal and the receiver circuit design. The transfer characteristics of the detector and processing devices following the detector may nullify the effects of an intruding unwanted signal. That is, an intruding signal is not necessarily an interfering signal. If the intruding signal alters only the amplitude of the desired signal, a good phase or frequency modulation detector will not be affected (if the envelope exceeds the threshold level). If the phase alone is altered, an envelope detector will not respond.

Both desired and undesired signals are present simultaneously at the detector input when the unwanted signal is admitted as a separate entity; e.g., in the various spurious entry modes. The signals mix in the detector, generally not in a simple additive manner. For instance, envelope detectors for AM signals and frequency modulation detectors both have capture characteristics. Their performance is dominated by the larger of two signals. These detectors are used frequently for suppressed-carrier double and single sideband, and for pulse code modulation (PCM) signals using phase- or frequency-shift keying. Coherent detector oscillators may lock on the undesired carrier frequencies, and this usually destroys the desired information. If locking to the wrong signal can be avoided, coherent detectors will effectively suppress an undesired signal.

7.2.3 Intrusion Through Ports not Intended as Signal Inputs

Numerous leads penetrate the case of a typical receiver. Each lead is potentially a vehicle for the transfer of electromagnetic energy into or out of the receiver. A method of dealing with this kind of problem is to install bypass capacitors and/or RF chokes to strip the wires of RF before they enter susceptible receiver circuits. This also prevents conduction and subsequent radiation of interference energy from the receiver to other susceptible equipments. The local oscillator signal is the most significant energy source in the receiver, and radiation of this signal is a source of interference to nearby receivers. An RF stage ahead of the mixer serves to attenuate and thus reduce the strength of the local oscillator signal reaching the antenna.

Holes put in equipment cabinets to provide air cooling provide a route by which interference can be conveyed to susceptible equipment. Two techniques used to reduce the intrusion of interference are (1) use of metal mesh screens to effectively close the air passage to RF, and (2) selection of the hole size to be equivalent to dimensions of a waveguide that is below cutoff for the interfering frequency.

SECTION 8 – HULL-GENERATED INTERMODULATION INTERFERENCE

The hull-generated interference phenomenon has become a problem of major proportions aboard ship because of:

a. the increase in number and power level of transmitters.

b. the increase in sensitivity and number of receivers.

c. the requirement for increased use of the spectrum.

d. the increase in number of antennas.

e. the same, or in some cases, less mounting space for antennas.

Interference signals arising from hull nonlinearities are related directly to the number and power level of simultaneous radiations from shipboard emitters. The probability of interference increases with the number of simultaneous radiations.

The shipboard environment possesses all the necessary elements to produce a harmonic generating and radiating system. The elements are present in the complex ship structures, appendages, and other objects immersed in high intensity rf fields. RF energy from shipboard transmitters induces current flow in these structures. If a corroded joint or oxidized fastening is in the path of current flow, rectification occurs. The distorted waveforms resulting from this process contain new signals. These signals, created by nonlinear junctions in the ship structures, are commonly called the "rusty bolt" effect.

8.1 FORMATION OF NONLINEARITY IN SHIP SUPERSTRUCTURE

There are thousands of nonlinear elements present in the topside areas of steel ships including steel hulls with aluminum superstructures. Steel is inherently nonlinear even when free from oxidation. The nonlinearities that cause the majority of intermodulation problems, however, are metallic junctions exposed to weather.

Two pieces of clean metal held tightly together have near zero impedance at the junction which has little restriction to current flow. Once the joint is exposed to weather, however, deterioration of the metal starts, due to corrosion. The oxides formed, since they are a semiconductor, destroy the metal-to-metal contact and the low impedance of the joint. The impedance varies under the influence of induced rf current from shipboard transmitters, i.e., the junction becomes nonlinear. Corrosion is the culprit.

The deterioration of metal in a corrosive environment is a familiar process. When metal objects are joined, and moisture is present, the junction will corrode. Corrosion is a complex process. Corrosive action may be either galvanic or electrolytic, or both, depending on the nature of the metals in contact and on whether the metal-to-metal contact is part of a direct current circuit. But, both types of corrosion take place only when moisture is in contact with the mating surfaces. The following is a partial list of the major causes of corrosion.

8.1.1 Galvanic Corrosion

(Refer to Table 8-1, Galvanic Series of Metals.) When a more noble metal is joined to a less noble metal by the same corroding medium, electrochemical corrosion will occur. This simple battery cell action produces a high corrosion rate on the less noble metal while the more noble metal remains unharmed. One of the attendant problems here involves the coating of the more noble metal with its less noble counterpart, resulting not only in mechanical weakness, but also creating a high impedance joint. In fact, a rectifier of sorts is formed with all the attendant problems relative to intermodulation.

Adjacent metals, listed in Table 8-1 are considered safe from galvanic action generation and may be classified as reasonably safe for electrical contact with each other.

Metals listed very far apart should not be mated because galvanic action may attack the metal at the top of the list. Metals that are listed as closely together as possible should be chosen; the further apart on the list, the greater the corrosion tendency through the galvanic process.

To have galvanic action, current, or course, must flow; in some cases, the current uses the metals as the prime conductor; in others, such as structures in a high rf field, the metals will intercept rf energy and become unexpected conductors. Reduction of galvanic corrosion requires a knowledgeable choice of construction materials and good construction techniques.

8.1.2 Fatigue Corrosion

The dangers inherent from fatigue corrosion are those from breakdown of a protective film on the metals caused by bending or vibration. Some metals have "self-repair" characteristics, and the corrosion process will be slowed, while others, not having a "repair' characteristic will corrode rapidly. The use

Table 8-1
Galvanic Series of Metals

Corroded End (anodic or less noble)

- Magnesium
- Magnesium Alloys
- Zinc
- Aluminum 1100
- Cadmium
- Aluminum 2017
- Steel or Iron
- Cast Iron
- Chromium Iron (active)
- Ni-Resist. Irons
- 18-8 Chromium-nickel-iron (active)
- 18-8-3 Cr-Ni-Mo-Fe (active)
- Lead-Tin Solders
- Lead
- Tin
- Nickel (active)
- Inconel (active)
- Hastelloy C (active)
- Brasses
- Copper
- Bronzes
- Copper Nickel Alloys
- Monel
- Silver Solder
- Nickel (passive)
- Inconel (passive)
- Chromium Iron (passive)
- Titanium
- 18-8 Chromium-nickel-iron (passive)
- 18-8-3 Cr-Ni-Mo-Fe (passive)
- Hastelloy C (passive)
- Silver
- Graphite
- Gold
- Platinum

Protected End (cathodic, or more noble)

of suitable materials and stringent mechanical design will largely obviate this problem.

8.1.3 Crevice Corrosion

Corrosion is likely to form in crevices because crevices have a propensity for retaining corrosive solutions. Crevices can lead to differences in metal ion concentration at different locations. Corrosion will take place at the edges of a crevice area having a higher metal ion concentration. Elimination of this problem can be accomplished by smoothing surfaces, filling crevices, and/or accomplishment of both through structure design (e.g., rounded corners). Crevices can be minimized by using welds instead of mechanical fasteners. Regardless of the torque applied to a bolt, it is practically impossible to eliminate crevices into which fluids can penetrate and cause corrosion.

8.1.4 Stress Corrosion

Stress corrosion occurs when internally or externally stressed metals are exposed to a corrosive environment. The loss of good bonds or structural integrity will be dependent upon (1) the magnitude of stress, (2) corrosion medium present, and (3) the structural configuration of the base metal. Stress corrosion is one of the most important and common types, but it is virtually impossible to predict since the same conditions that cause cracking in one metal will not influence another in the same general category. In general, high strength aluminum alloys are susceptible to cracking and should be avoided if at all possible.

8.1.5 Welding Corrosion

Corrosion can occur in areas where variations in grain size are produced by the heat from welding. Corrosion rates vary according to the heat input of the welding method and the geometry of the joint. Here the grain structure is changed by high temperatures caused by the inability of the heat to dissipate. The possibility of corrosion in these areas can be avoided by increasing the surface area so that heat is dissipated more rapidly.

8.2 MECHANISM OF NEW SIGNAL GENERATION

The phenomenon of mixing two signals in a nonlinear element to produce a new frequency is familiar to persons acquainted with the operation of a simple heterodyne receiver. A locally generated signal is applied along with the broadcast signal to a mixer or first detector stage. This stage is operated intentionally so as to distort and produce harmonics of the two input signals.

The output of this nonlinear stage consists of the original two frequencies plus the sums and differences of the two original frequencies and all of their harmonics. In a typical mixer stage the difference frequency of the two original signals is normally selected for further processing. In some frequency translating schemes, however, any number of frequencies can be selected by inserting appropriately tuned circuits.

The point to keep in mind is that when a single frequency sine wave is distorted in any manner, harmonics of that frequency are generated. When two or more single frequency sine waves are mixed in a distortion-producing (nonlinear) element, then many new frequencies are generated.

This process is called heterodyning. It requires but little additional study to understand how a similar process occurs when corroded metallic objects are immersed in high intensity rf fields.

8.2.1 Frequency Content of Complex Waveforms

Spectrum analysis of a sine wave reveals only one frequency. Any other waveform, when examined in the same manner, would yield a fundamental frequency plus harmonics. The more the deviation from a sine function, the higher the harmonic content of any particular waveform. Figure 8-1 illustrates the formation of a $\sin^2 \theta$ waveform from the addition of two pure sine waves.

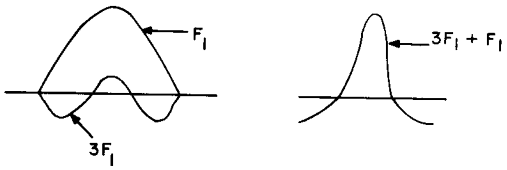


Figure 8-1. Addition of Two Sine Waves

A perfect square wave, for example, would contain a fundamental and an infinite number of odd harmonics. See figure 8-2.

All periodic wave functions, no matter how complex, are made up of a sum of harmonically related components having various amplitudes and phase relations to one another.

A mathematical or physical analysis of a periodic wave function determines its harmonic content. This process is called Fourier analysis.

8.2.2 Distortion of Sine Waves by Nonlinearity

A linear device has a linear transfer characteristic. An induced current through such a device is at all times directly proportional to the voltage causing the current. The output would be an exact replica of the input (see Figure 8-3a). When a pure sine wave (single frequency) current flows through a linear device there is no distortion and no harmonics of the single frequency current are generated.

A nonlinear element is simply one in which the current through the element is not at all times proportional to the *voltage across* the element. Ohm's law has been violated. The current is not a replica of the voltage that produced it; therefore the impedance must have changed under the influence of the driving voltage, i.e., the element has a nonlinear transfer characteristic (see Figure 8-3b). This characteristic produces distortion with an attendant generation of harmonics.

8.2.3 Mixing of Harmonics to Produce Intermodulation Products

Two or more frequencies present simultaneously in a nonlinear impedance will give rise to new frequencies that are the sums and differences of the fundamentals and harmonics. Hundreds of combinations are possible when higher order harmonics are considered.

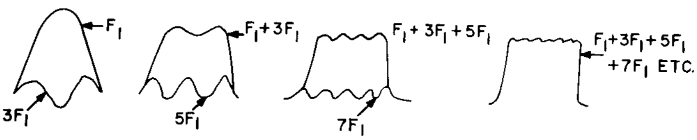
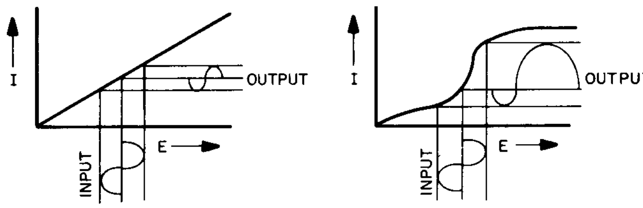


Figure 8-2. Square Wave Formation from Sine Waves



a. LINEAR TRANSFER CHARACTERISTICS b. NONLINEAR TRANSFER CHARACTERISTICS

Figure 8-3. Characteristics of Linear and Nonlinear Transfer Mechanisms

8.2.4 Product Order

When the second harmonics of two signals mix to produce a third, the new signal is a 4th order product, e.g., $2f_1 \pm 2f2$ = 4th order.

Product order is the sum of the harmonics contained in any discrete output frequency from a nonlinear element. For example, when two signals f_1 and f_2 mix in a nonlinearity the following frequencies would appear:

$f_1 + f_2 = f_3$
$f_1 - f_2$ (assuming f_1 is the higher frequency) $= f_4$
$2f_1 + f_2 = f_5$
$2f_1 - f_2 = f_6$
$f_1 + 2f_2 = f_7$
$f_1 - 2f_2 = f_8$, and so forth

The first two frequencies (f_3 and f_4) are 2nd order products; the next four (f_5, f_6, f_7, f_8) are 3rd order products. Significant intermodulation products aboard ship extend through about the 9th order. Higher order products are usually too low in amplitude to interfere with reception. Figure 8-4 shows the number of frequencies generated as a function of product order and the number of transmitters radiating simultaneously. Figure 8-5 illustrates the large increase in intermods generated when an additional transmitter is added to an existing radiation environment.

8.2.5 Reradiation of Signals from Linear Junction

To illustrate the hull-generated interference phenomenon, imagine that a welded lifeline stanchion (linear junction) aboard ship receives energy at 4 MHz from a shipboard transmitter. Induced rf current, undistorted by the welded junction, causes reradiation of the driving signal and no others (see Figure 8-6).

The receiver can detect the fundamental at 4 MHz but does not detect any signal at 8 MHz, 12 MHz, etc., since they are not generated by the welded (linear) joint.

8.2.6 Radiation of Harmonics from Hull Nonlinearity

Now assume that the stanchion was bolted or riveted instead of welded to the hull. Assume further that the bolted (or riveted) junction has corroded and a nonlinearity has been formed due to metallic oxides between the two structures. The induced rf current flowing through the nonlinear impedance of the junction would not be an exact replica of the rf driving voltage, i.e., a harmonic content has been added by the nonlinearity. These new signals are radiated and the receiver can detect signals at 4 MHz, 8 MHz, 12 MHz, etc. See Figure 8-7.

8.2.7 Intermodulation Product Radiation from Hull Nonlinearity

Assume now that a second transmitter is radiated at 3 MHz in addition to the original transmitter radiating at 4 MHz. Both signals illuminate the nonlinear junction and harmonics of each fundamental mix with each other and the two fundamentals. See Figure 8-8.

Many new signals are generated in this process. A few of these signals are:

$f_1 + f_2$ = 7 MHZ
$f_1 - f_2$ = 1 MHz
$f_1 + 2f_2$ = 10 MHz
$2f_1 + f_2$ = 11 MHz
$2f_1 - f_2$ = 5 MHz
$3f_1 + f_2$ = 15 MHz
$3f_1 - f_2$ = 9 MHz, and so forth.

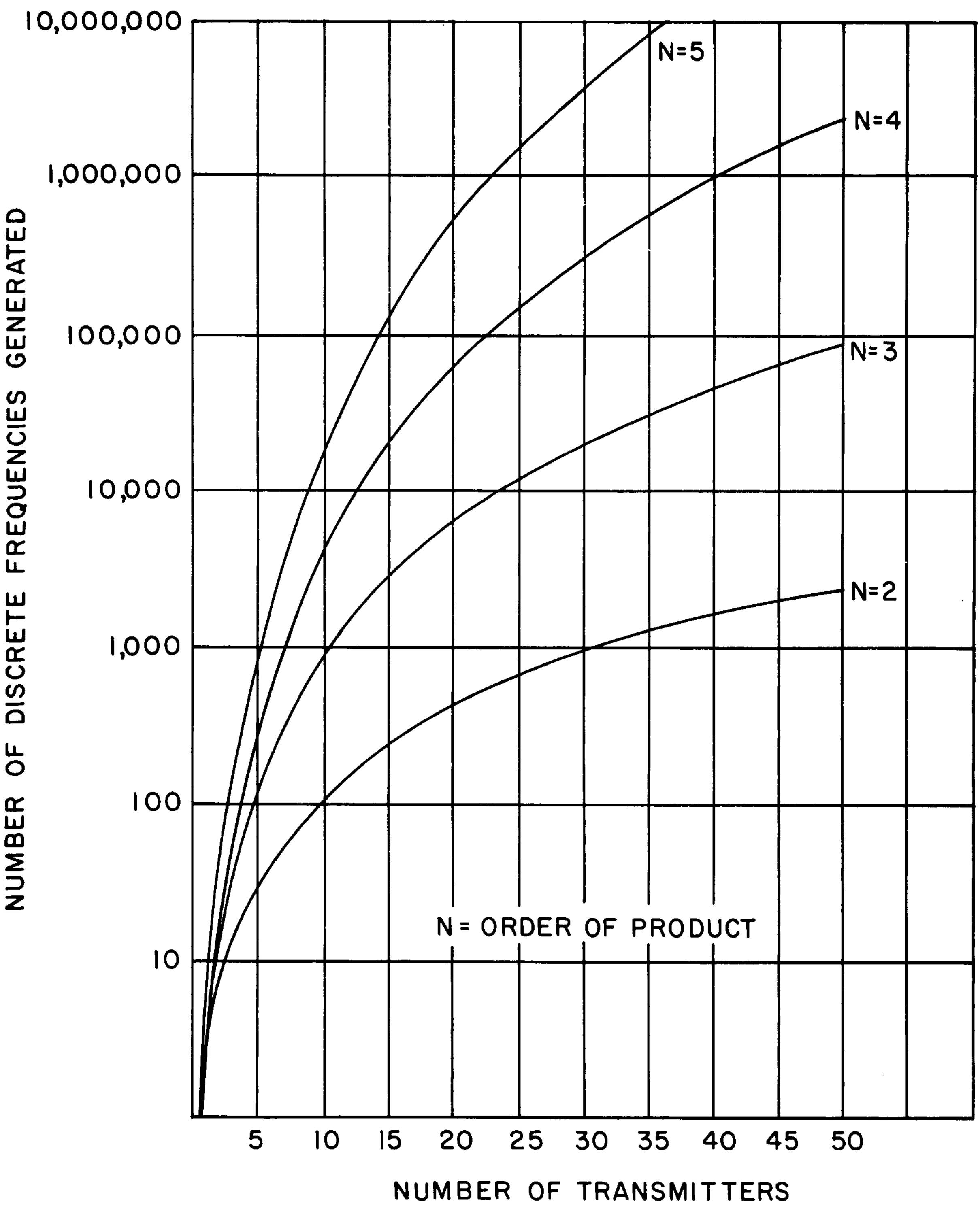


Figure 8-4. Effect of Product Order on Number of Frequencies Generated

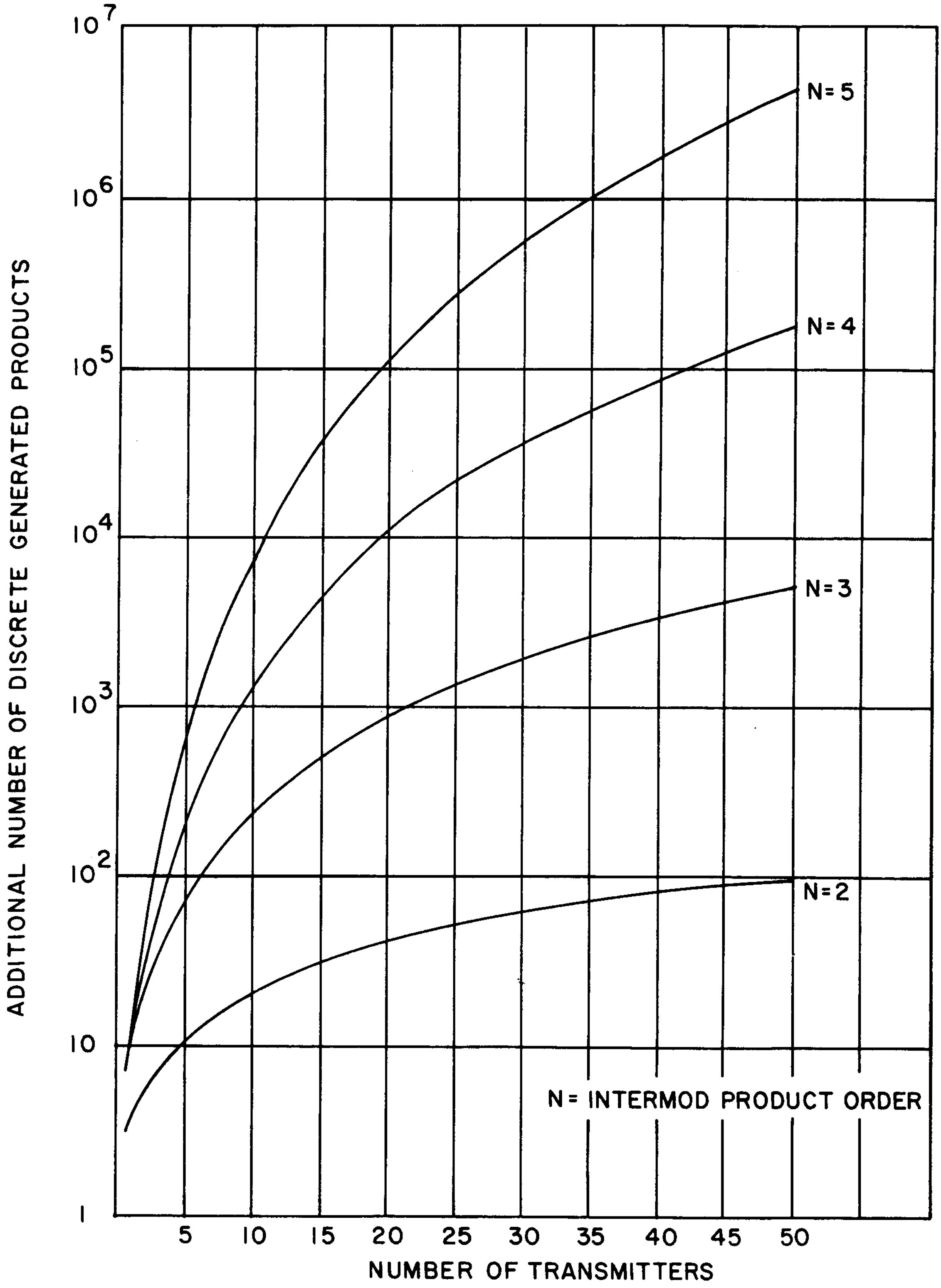


Figure 8-5. Effect of Adding One More Transmitter

The receiver can now detect signals at each of the frequencies listed above and if one of these frequencies is close to a desired frequency, then interference will result. Typical hull-generated interference sources are shown in Figures 8-9 through 8-16.

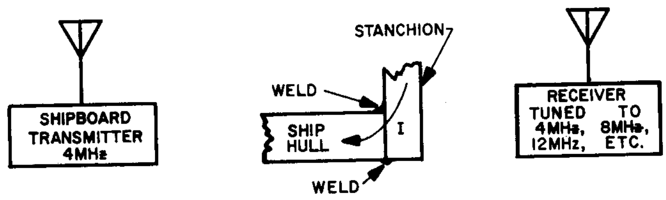


Figure 8-6. Signal Reradiation from Welded Junction

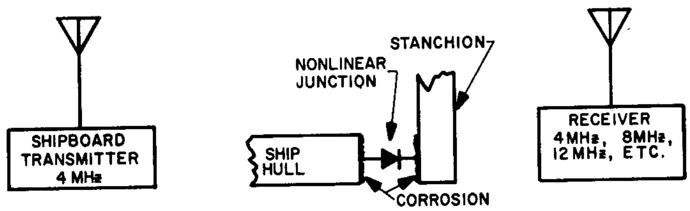


Figure 8-7. Harmonic Generation from Hull Nonlinearity

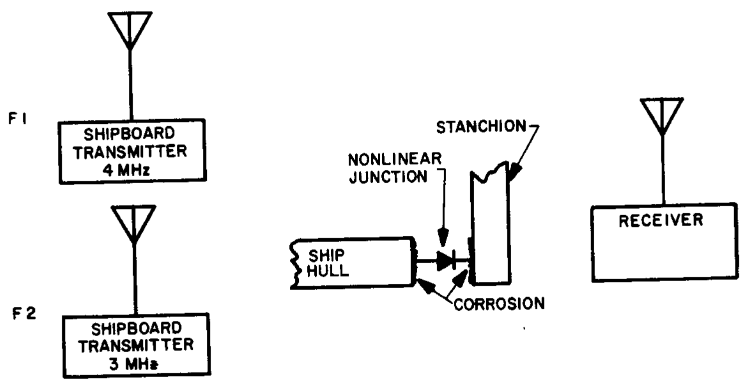
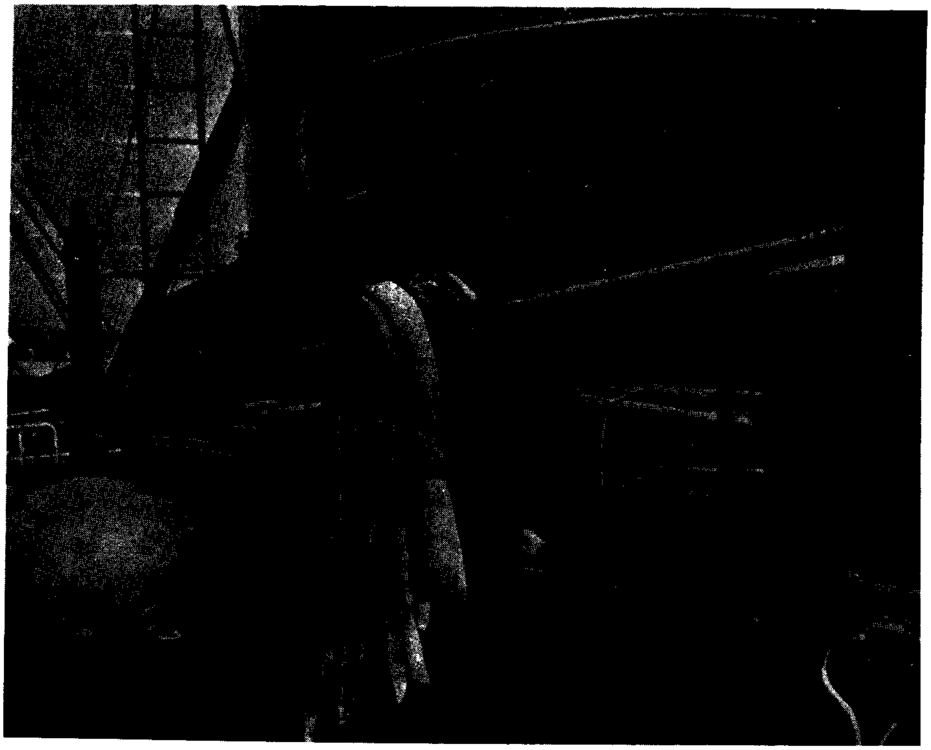
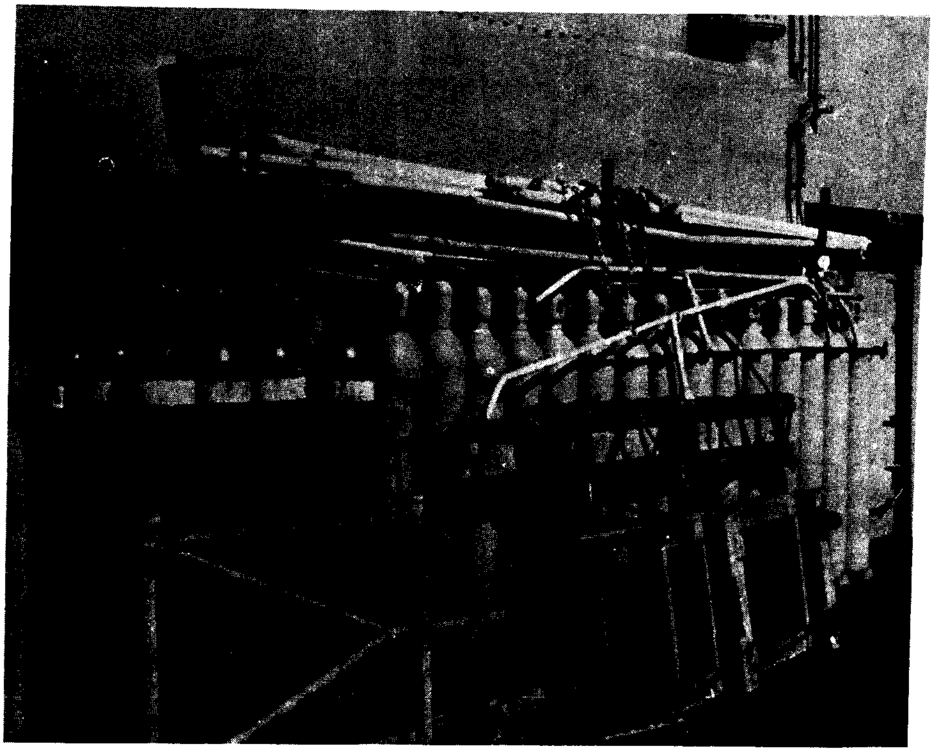


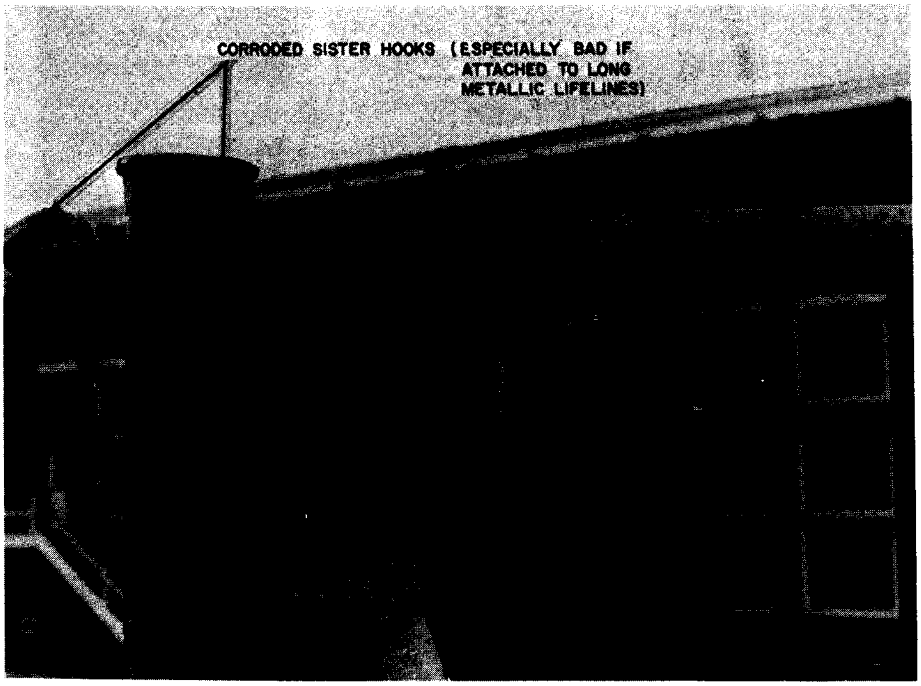
Figure 8-8. Intermodulation Product Radiation from Hull Nonlinearity



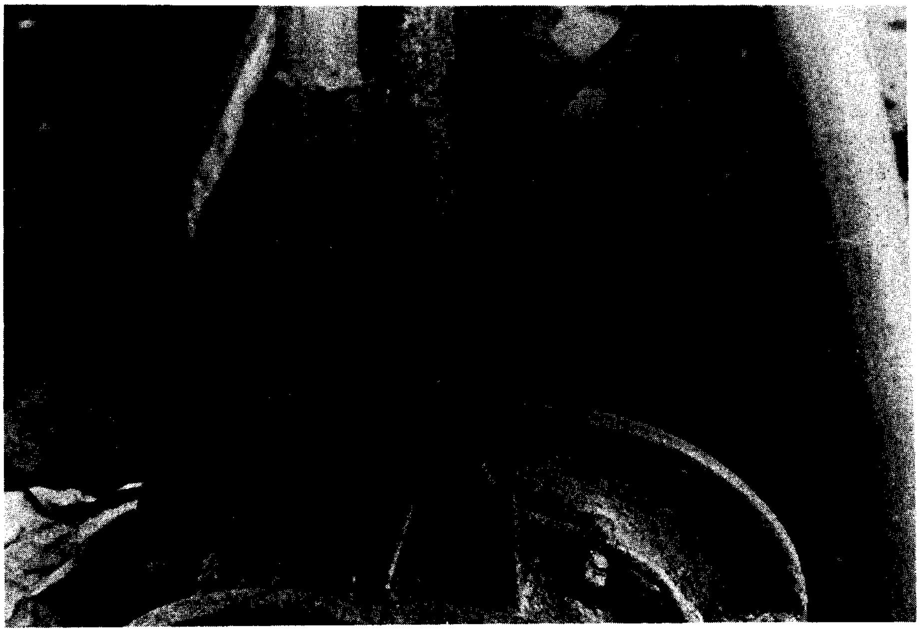
a. Loose Metallic Items in Topside Area
Figure 8-9. Typical Intermodulation Product Sources



b. Loose Metallic Items in Topside Area
Figure 8-9. Typical Intermodulation Product Sources



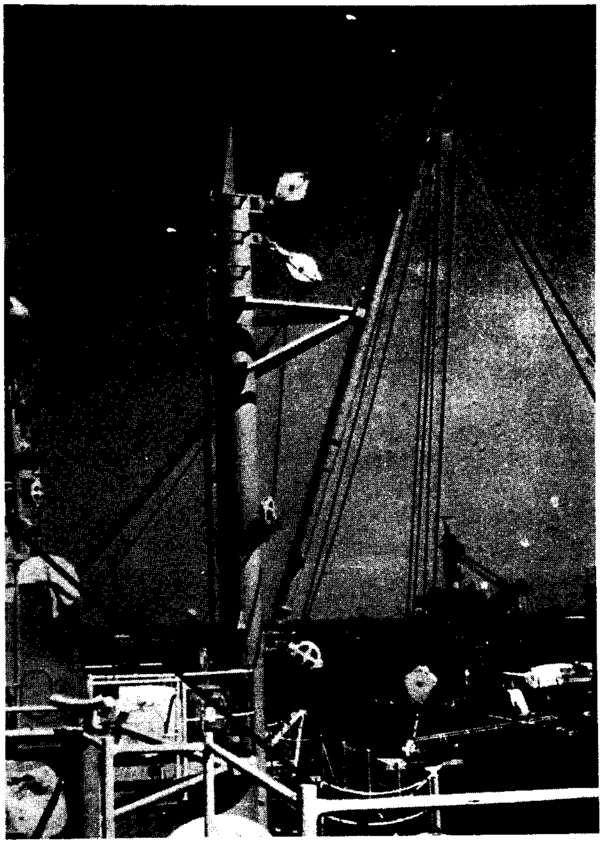
a. Corroded Stanchion and Metallic Lifelines
Figure 8-10. Typical Intermodulation Product Sources



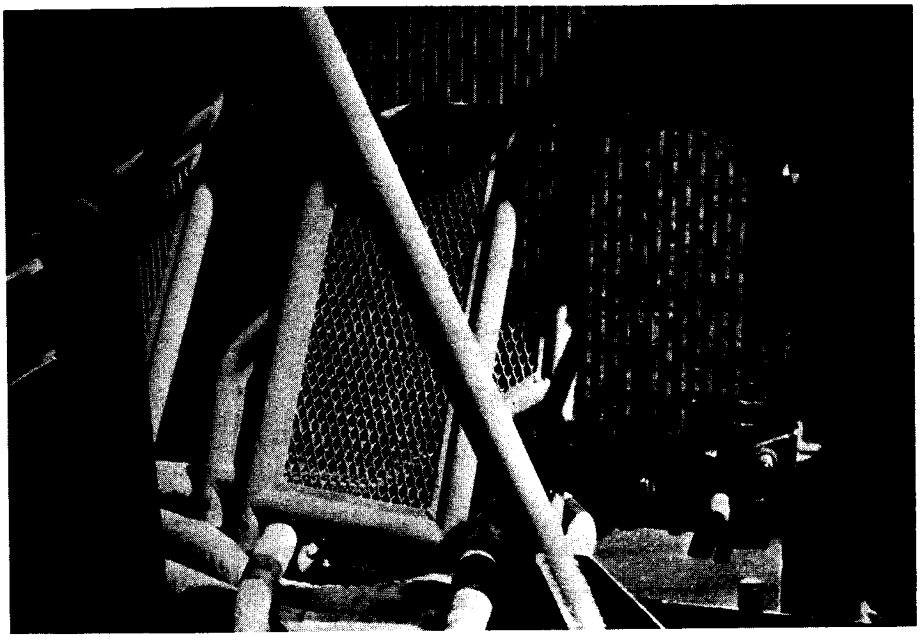
b. Corroded Stanchion and Metallic Lifelines
Figure 8-10. Typical Intermodulation Product Sources

a. Corroded Door Hinge
Figure 8-11. Typical Intermodulation Product Sources

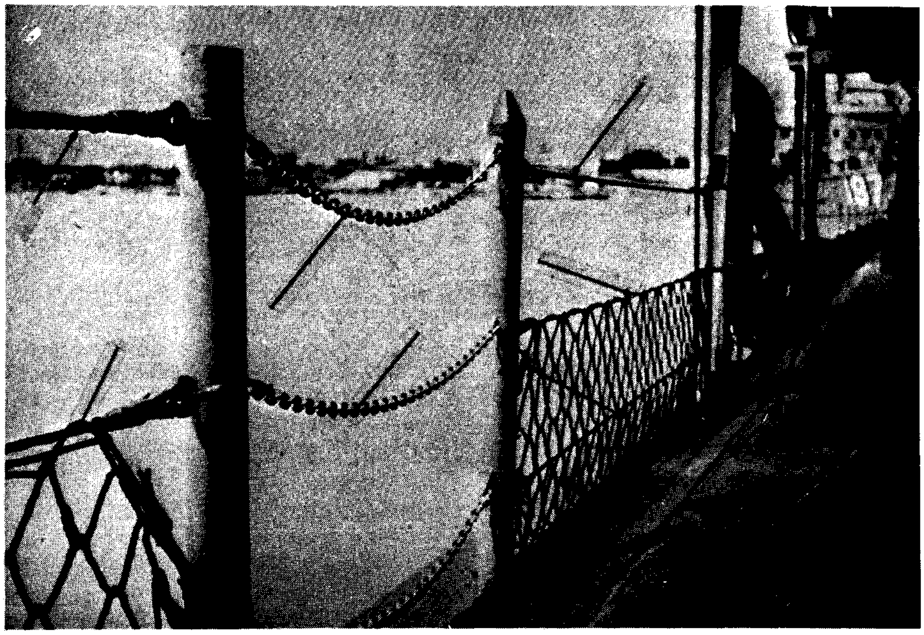
b. Rusty Anchor Chain and Metallic Cables
Figure 8-11. Typical Intermodulation Product Sources



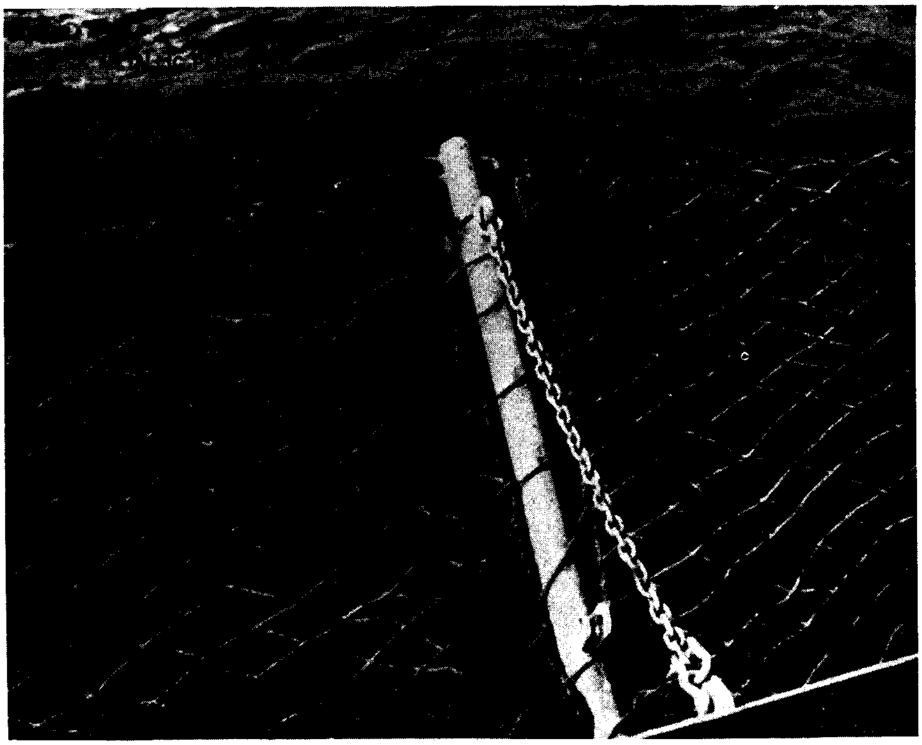
a. Booms and Associated Rigging
Figure 8-12. Typical Intermodulation Product Sources



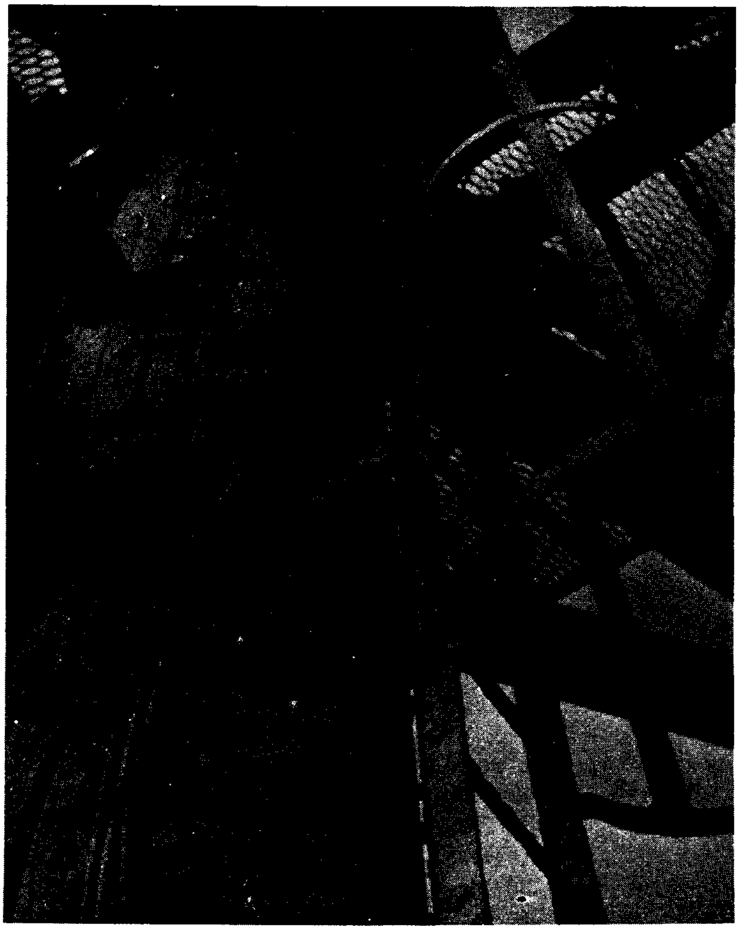
b. Accommodation Ladder and Accessories-Stowed
Figure 8-12. Typical Intermodulation Product Sources



a. Safety Chains and Metallic Lifelines
Figure 8-13. Typical Intermodulation Product Sources

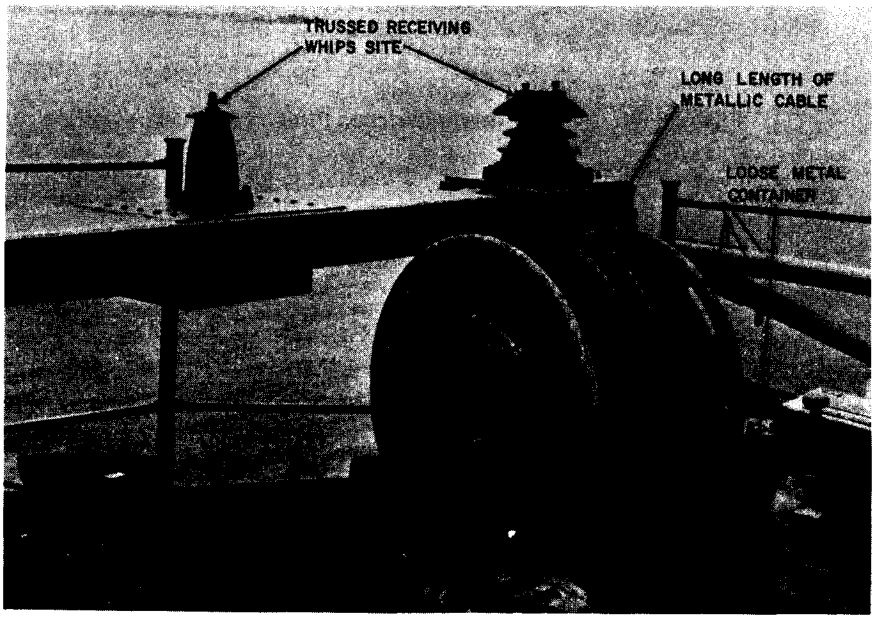


b. Metallic Safety Nets, Chain and Cable
Figure 8-13. Typical Intermodulation Product Sources

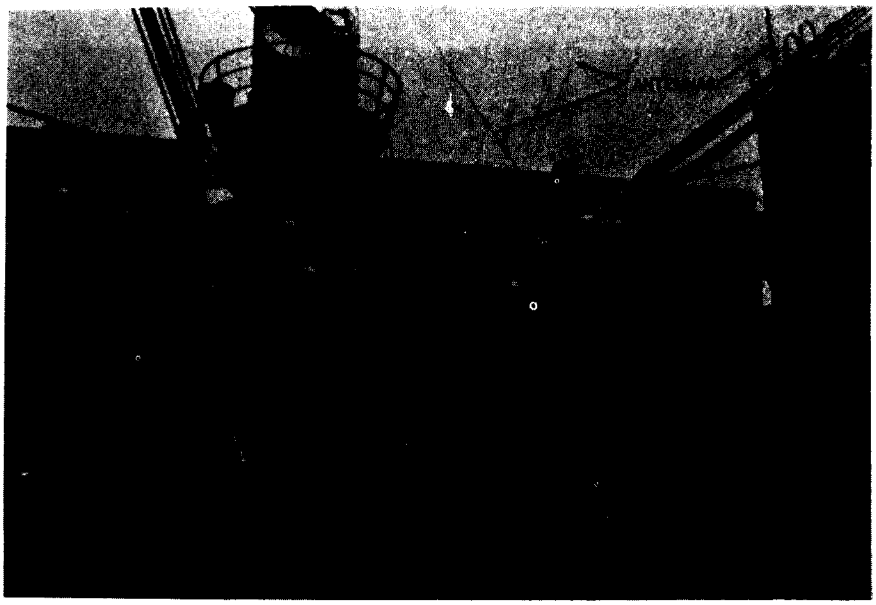


a. Mast Items
Figure 8-14. Typical Intermodulation Product Sources

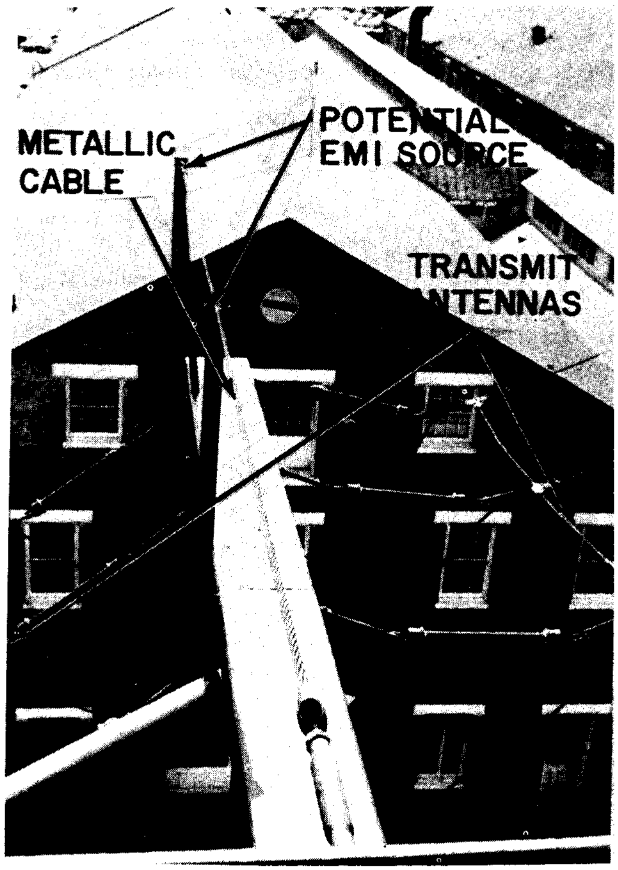
b. Corroded Pinned Joint
Figure 8-14. Typical Intermodulation Product Sources



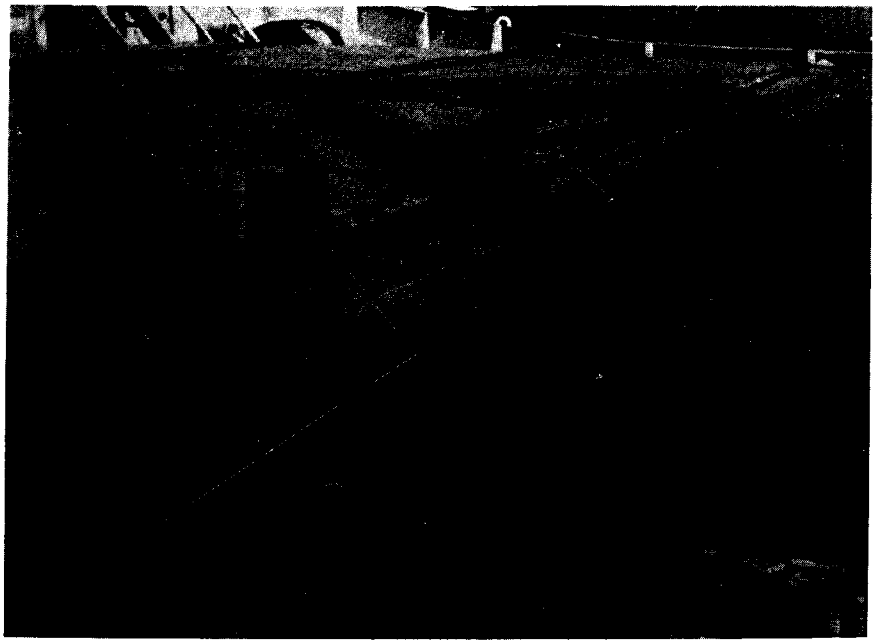
a. Loose Metal Items Close to Antenna
Figure 8-15. Typical Intermodulation Product Sources



b. Booms, Cables, Shackles and Pulleys
Figures 8-15. Typical Intermodulation Product Sources



a. Metallic Yardarm Cable
Figure 8-16. Typical Intermodulation Product Sources



b. Cargo Hatch Hold-Down Straps
Figure 8-16. Typical Intermodulation Product Sources

8.2.8 Levels of Hull-Generated Interference

The level of hull-generated interference signals is determined by (1) output power of the transmitters illuminating the nonlinearity; (2) efficiency of the nonlinearity as a rectifier, i.e., the degree of nonlinearity; (3) coupling and amount of shielding between the transmitter and the nonlinear element; (4) the physical size, configuration and orientation of the metallic objects comprising the junction i.e., the larger the objects are with respect to a given wavelength the more energy they will intercept and reradiate; and (5) physical properties, i e., gain of the transmit antennas.

8.2.9 Additional Nonlinear Sources

IITRI studies show that steel is inherently nonlinear and will produce intermodulation products. However, the level of these products is well below those levels produced by corroded junctions and become significant only after all other nonlinear sources have been eliminated.

Intermodulation occurs also in receiver front ends when two or more strong signals penetrate the preselector attenuation, overdrive the rf amplifier stage and cause it to become nonlinear. The same mixing occurs if the signals are strong enough to penetrate to the mixer stage which is normally operated nonlinearly. The cure here is to insert additional attenuation to off-frequency signals ahead of the receiver. A tunable bandpass filter such as the Collins F-871/U, or equivalent, that offers about 70 dB attenuation to off-frequency signals is usually sufficient.

Another nonlinear element across which two or more signals may intermodulate is the final or output stage of a shipboard transmitter. Signals from one transmitter may couple, via closely spaced antenna systems, into the output stage of a second transmitter and intermodulate with the functional signal energy present in that stage. Intermodulation products from the mixing process are then dumped into the environment and become interference sources. A solution to this problem is to insert a multicoupler between the transmitter and its antenna. The multicoupler serves to attenuate off-frequency signals regardless of whether they are entering or leaving the transmitter. If strong off-frequency signals are prevented from intruding in the final amplifier stage then intermodulation products originating in transmitters is eliminated.

8.3 SUPPRESSION TECHNIQUES

The understanding of specific interference sources without the application of such knowledge to the reduction or elimination of interference is wasted effort. For most effective results, suppression techniques should be applied during each phase of ship construction. A retrofit normally costs more and accomplishes less.

At least in the foreseeable future, ships will continue to be constructed of metal – both steel and aluminum. A desirable goal (which probably will never be fully achieved) to reduce or eliminate hull-generated interference is to make the entire topside structure of such ships a single homogeneous conducting structure. This structure should present a low linear impedance to the induced rf current from shipboard transmitters and be completely devoid of loose metal items. If there were no discontinuities in the topside area of a ship, there would be no hull-generated interference.

Refer to Section 10, EMI Reduction Methods, for specific reduction techniques.

SECTION 9 – RADAR SYSTEMS INTERFERENCE

9.1 INTERFERENCE PRODUCED BY RADAR SYSTEMS

9.1.1 Modulators

In a pulse radar system the output of the modulator is a high-voltage rectangular pulse used to key the transmitter. The basic circuit for most modulators is shown in Figure 9-1. This circuit is applicable to both the hard-tube and the line-type modulator.

Figure 9-2 shows the output circuit of a hard-tube modulator where capacitor C_0 is the energy storage element. If a positive pulse is applied to the control grid of T_1, electrons flow out of C_0 and around the load circuit (to the right of A-A') until the driving pulse on the grid of T_1 is removed.

The basic circuit for a line-type modulator, shown in Figure 9-3, consists of a power supply, charging impedance, switch, pulse forming network, and pulse cable leading to the pulse transformer and load. The pulse forming network is an artificial transmission line. The switch may be a rotary gap, a "trigatron," a series gap, a hydrogen thyratron, or a mercury thyratron. Thyratrons are most commonly used.

The main source of interference in a modulator is its own output waveform.

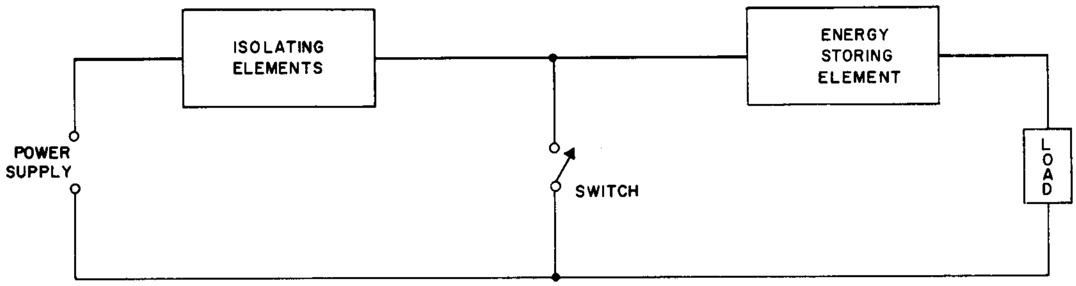


Figure 9-1. Basic Circuit for Pulse Modulator

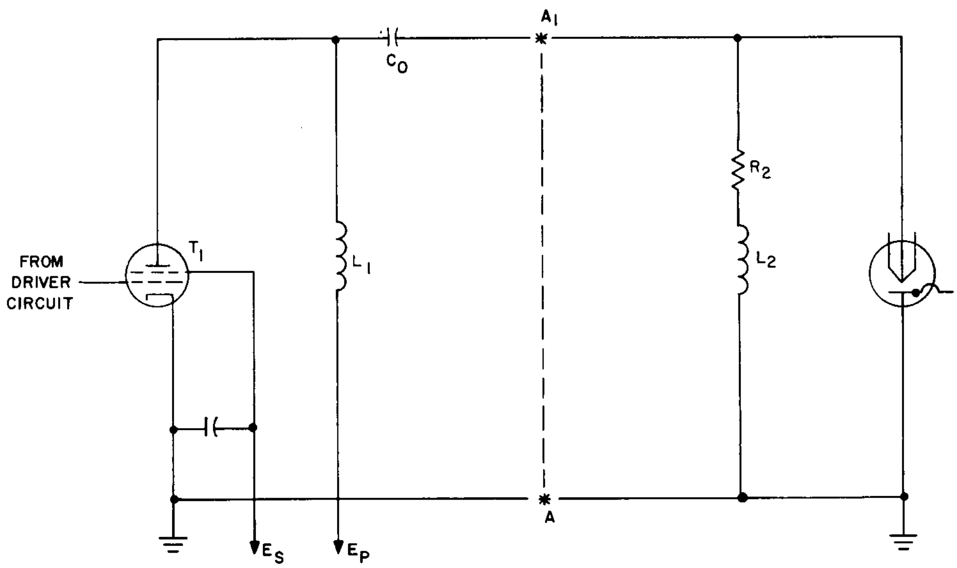


Figure 9-2. Output Stage of Hard-Tube Modulator

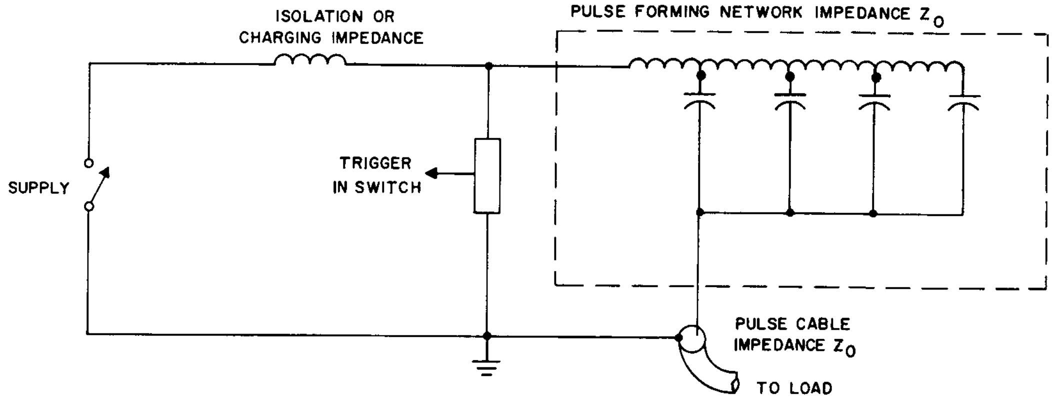


Figure 9-3. Line-Type Modulator

This waveform is a high-voltage rectangular pulse which causes the generation of harmonics over a broad portion of the frequency spectrum. Any extraneous voltage spikes on the waveform will necessarily increase the related interference level. Since most radar transmitters require a modulation pulse with fast rise and fall times, very little can be done to prevent the generation of harmonics. However, care should be taken to see that all unnecessary spikes are eliminated.

Since little can be done to prevent the generation of interference produced from the modulator pulse, the interference produced therein must be isolated by shielding from other electronic equipment.

9.1.2 Magnetron Spurious Outputs

The most commonly used microwave power source is the plate-modulated magnetron, and it is also the most troublesome from an interference standpoint. Magnetrons emit energy at harmonics of the fundamental frequency and also at spurious frequencies that bear no relation to the fundamental. The harmonic and spurious output level is unpredictable and will vary between tubes of the same design.

Since the causes of harmonic and spurious emissions are not fully understood nor their effects easily predicted, a filter should always be placed between the transmitter and the antenna to attenuate harmonic and spurious outputs.

9.1.3 Magnetron Frequency Pulling

Magnetrons are self-excited oscillators whose frequency of operation depends on the loading. When the radar antenna is scanning, it is inevitable that variations in magnetron loading will occur as a result of changing reflections from rotary joints, antenna housing, and nearby large reflecting objects. Such changes in loading, produce changes in frequency which may be large. The effect of loading on the frequency of a magnetron is usually expressed by what is called the "pulling figure." It is defined as the total frequency excursion experienced when a voltage with a standing wave ratio of 1:5 to 1 is presented to the magnetron and varied in phase over at least 180 degrees.

The problem of pulling of the magnetron frequency due to variation in load is attacked in two ways: (1) the magnetron is designed so that its frequency shift with changing rf loading is small; (2) variations in rf loading are reduced by careful design and construction of the rf components.

It is possible in some cases to combine a high-Q cavity with the magnetron coupled to it in such a way as to reduce the pulling figure by a large amount with little or no loss in efficiency. The addition of this stabilizing cavity has the disadvantage of reducing the tuning range of the magnetron.

9.1.4 Klystron Harmonic Outputs

Klystrons do not generate spurious outputs, but do generate harmonics. As with magnetrons, the harmonic output level from klystrons is unpredictable.

9.1.5 Pulse Transformer Leakage

Energy can be radiated from a pulse transformer due to the high peak currents in the windings. Usually, a pulse transformer is enclosed in a shield with good rf integrity. In fact, all components involved in forming the high amplitude transmitter pulse should be enclosed in shield compartments to prevent interference energy from being radiated.

9.1.6 RF Leakage from Waveguides

Leakage of rf power can occur at any waveguide or coaxial cable joint. Leakage from waveguide joints constitutes an interference source and also a loss in transmission efficiency.

To minimize leakage the waveguide should be firmly connected with an rf shielding gasket between sections. The bolts holding the waveguide joints together should be external to the high intensity rf area. The gasket should be located between the bolt holes and the rf area. Routing the waveguide through a hole in the transmitter cabinet allows interference to be conducted on the outside of the waveguide and to be radiated after leaving the cabinet. A short section of flexible waveguide may be used in connecting the waveguide to the transmitter to adjust for slight misalignment of the waveguide and transmitter.

9.1.7 Duplexer Leakage

Duplexers may use gas TR (transmit-receive) and ATR (anti-TR) tubes or ferrite devices. The TR and ATR tubes function as switches through ionization of their internal gases during the transmit time. The ionization process produces interference which may be conducted or radiated to other equipments. If the tubes required a "keep-alive" voltage, the "keep-alive" voltage lead should be filtered and shielded.

If a CW transmitter is operating in the vicinity of a pulsed radar, the CW power level may be sufficient to continuously fire the TR tube. The continuous firing of the TR tube isolates the receiver from the antenna insofar as target signals are concerned. This interference to the radar operation is relatively independent of frequency providing the CW energy can propagate down the waveguide of the radar. The insertion of filters between the antenna and duplexer may reduce CW power to levels that will not affect the TR tube.

9.1.8 Arcing in Waveguides, Rotary Joints and Mast Cables

Another kind of interference, often difficult to detect, is the broadband noise caused by arcing or corona. Arcing may result when a rotary joint is dirty, worn or misaligned. This is a cumulative process since arcing causes further deterioration of the joint which results in more arcing. Sustained arcing normally causes a complete breakdown of the rotary joint. Arcing can occur also in waveguides due to foreign objects (including moisture) in the waveguide, corroded or flaked-off coating, a high voltage standing-wave ratio (VSWR), or a sharp bend in the waveguide run.

Arcing can result when high energy radar beams illuminate mast discontinuities such as corroded cables, cable hanger straps, service platform safety lines (or chains), and the like. This kind of arcing can be best detected at night when the radars are operating. Normally, the arc can be seen and heard. Good housekeeping practices, such as the replacement of corroded cables, the cleaning and tightening of loose strap brackets, and the replacement of all metal safety lines with nonmetallic rope, will eliminate interference from this source. Current practice is to run mast cables inside the mast or on the opposite side of any radar antenna, or to install rf shielding over mast cables which are located in the main beam of radars. No matter what the source, arcing results in a very intense, broad distribution of interference that may interfere at frequencies far removed from the radar fundamental frequency.

9.1.9 PRF Overload

Receivers that are sited close to a radar antenna will experience interference due to "brute force" penetration of the rf pulse energy. The susceptible device does not have to be tuned to the frequency of the radar for this action to occur. Intense rf energy simply forces its way pass the front end selectivity and enters the active amplifier stage. The amplifier is driven alternately between cutoff and saturation by the interfering pulse which gives the characteristic radar PRF buzz at the receiver output. Usually, the interference is heard only when the receiving antenna is illuminated by the radar beam.

The most effective method for alleviating this kind of interference, is to locate all receiving antennas out of the main beam of nearby high power radars. In some cases, it is necessary to install additional shielding of rf transmission cables that pass through the main radar beam.

It is also possible to run antenna cables inside the mast or on the side of the mast opposite the one being illuminated by the radar beam. The mast then serves as a shield to reduce the energy pickup on the antenna cables.

9.1.10 Sideband Splatter and Pulse Shaping

Present radars with a pulsed transmitter strive for a fast rise time of the pulse. Analysis of such a pulse shows a broad spectrum–the faster the rise time, the broader the spectrum. A spectrum analysis of a rectangular pulse of 0.5 microsecond width shows that the peak power of the first transmitted sideband is down only 13 dB, and for several megahertz away from the carrier frequency, each successive sideband contains about one-half as much peak power as the preceding one. Beyond a certain frequency, which is an inverse function of the pulse width, the peaks fall off more gradually. The net result is that the theoretical spectrum of a narrow pulse may be reduced only 60 dB at 100 megahertz from the center frequency. Other radars operating in the same frequency band may be affected by this sideband splatter. Today's crowded frequency spectrum demands conservative use and a reduction in required channel widths.

Studies now in progress indicate that for many applications, even in high-resolution fire-control radars, a very fast rise time not only may be undesirable, but unnecessary.

It is possible to reduce the power transmitted in the sidebands by controlling the rise and decay times of the radiated pulse. The Gaussian pulse has the narrowest spectrum of any of the pulses (see Figure 9-4). Pulse shaping techniques require a modulation pulse that does not have fast rise and fall times; hence, interference levels are automatically lower.

9.1.11 Local Oscillator Radiation

The local oscillator input to the mixer, in the superheterodyne receiver, is at least ten times the magnitude of the desired signal input. This ratio is desirable for good conversion efficiency to the intermediate frequency. In most radar receivers, the high frequencies used preclude an rf amplifier stage ahead of the mixer. The result is that the strong local oscillator signal is fed back to the antenna system and is radiated as an interfering signal. A major disadvantage of L.O. radiation is its capability of detection by enemy ECM equipment even when the RADAR is not transmitting. A secondary consideration, at least in the radar frequencies, is the possibility of this signal interfering with nearby receiving equipment.

Usual fixes for this kind of interference include:

a. Use of a balanced mixer which greatly attenuates the L.O. signal leaking through to the antenna.

b. Use of rf amplifier or preselection ahead of the mixer stage to attenuate the L.O. signal from getting to the antenna.

c. Proper shielding and filtering of the local oscillator stage including components and wiring.

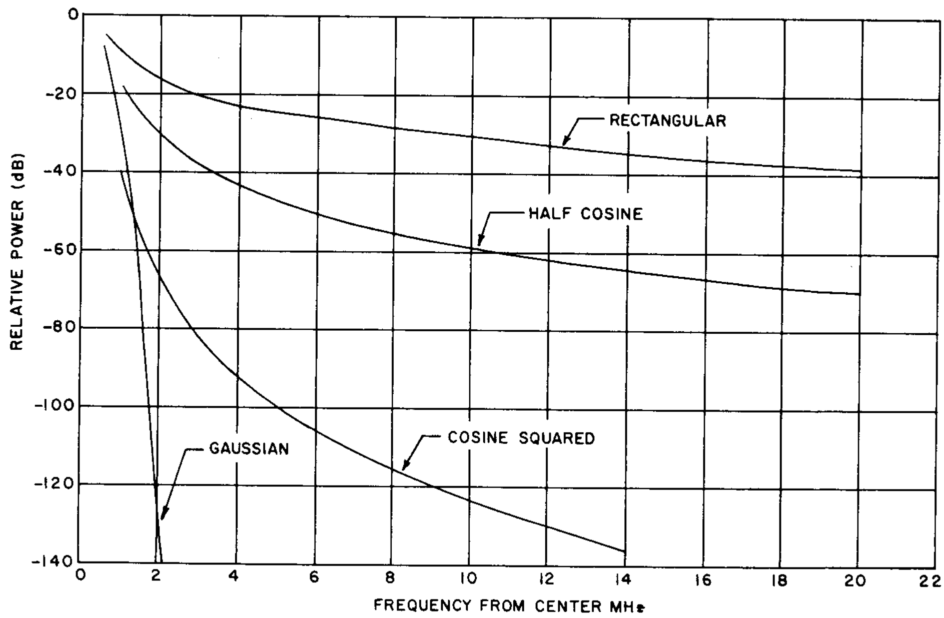


Figure 9-4. Distribution of Spectral Energy for a One-Microsecond Pulse

9.2 RADAR SUSCEPTIBILITY TO INTERFERENCE

9.2.1 Receiver Spurious Responses

The heterodyning process in the superheterodyne receiver gives rise to spurious responses. A spurious response occurs when a signal other than the desired signal mixes with the local oscillator frequency or a harmonic of the local oscillator frequency in such a manner as to cause an IF signal. An illustration of spurious responses would be a radio receiver tuned to 1000 MHz having a local oscillator frequency of 1030 MHz.

These two signals are combined in the mixer and the resulting difference frequency of 30 MHz is amplified in the IF amplifier. If an interference signal of 1060 MHz enters the mixer, it beats with the local oscillator signal and the difference frequency is also 30 MHz, which will be amplified by the IF amplifier. This IF signal then is a spurious response since the 1060 MHz signal is an unwanted frequency. Spurious responses can also be caused by signals that beat with harmonics of the local oscillator frequency to give a sum or difference frequency which is equal to the IF or a harmonic of the IF.

In order to prevent the generation of spurious responses in the heterodyning process, the interfering signal must be prevented from reaching the mixer stage. Waveguide filters are effective in attenuating any interfering signal before it reaches the mixer. Recent advances in microwave tube development have made available traveling wave tubes and backward wave tubes which can be used as rf amplifiers in radar receivers. Moreover, narrow-band rf amplifiers provide preselection as well as rf amplication, and therefore help reduce spurious responses.

When spurious responses are caused by a signal mixing with a harmonic of the local oscillator frequency, a bandpass filter placed in the line between the local oscillator and the mixer will help attenuate these harmonics.

9.2.2 Front End Selectivity

Radar receivers do not normally have as high a front end selectivity as do lower frequency communications receivers. It is difficult to design selective rf amplifiers for the microwave frequencies used in radar systems. Consequently, in radar receivers it is easier for strong off-frequency signals to penetrate to the mixer stage and interfere with desired signal reception.

The problem is not as severe as might first appear, however, since (1) the radar portion of the spectrum is not as crowded as the communications portion and (2) radars of the same, or very nearly the same, frequency are normally never co-located.

9.3 RADIATED INTERFERENCE MECHANISMS

In addition to the more or less standard sources of interference, there are others that are not normally considered. Some of these are discussed in this section.

9.3.1 Passive (Non-Intentional) Interference

There are other forms of interference to radar systems that come under the heading of natural phenomena. Radar signals can be reflected from metallic items such as masts, stacks, etc., located in the main beam. These reflections interfere with the normal propagation and cause blank sectors on the radar indicators, i.e., a full 360° azimuth coverage is not obtained. Targets within the affected sector either will not appear on the radar scopes or will not return as strong an echo as do targets in unaffected areas. Much closer scrutiny is required of the operator to detect targets in the blanked sector.

The radar operator should be aware of these effects even though there is little or no corrective action possible on his part. Antenna and ship design personnel are striving constantly to obtain clear coverage by all radars.

It is also possible, especially in L-Band radars, to experience a loss of air targets due to a phenomena called "ducting." A rapid change in air temperature at a fairly low altitude creates, in effect, a huge waveguide that tends to propagate the radar energy along the curvature of the earth. Aircraft flying above this "duct" are not illuminated by the radar beam and do not return an echo to the radar receiver. An experienced operator can recognize the presence of the effect very simply: Land targets are observed at much greater distances than are normal and practically no air targets are seen. This indicates that the radar energy is not propagated at slant angles for air target illumination but is being confined and forced to follow the earth's curvature.

9.3.2 Active (Intentional) Interference (Jamming)

Electronic warfare (EW) is a very real and complex science in today's technology. It has become a deadly serious "cat and mouse" game of threat and counter threat. Specific details are classified and are not within the scope of this book. Radar operators should strive to recognize the effects of intentional, deliberate jamming and learn to meet such threats by proper use of any anti-jam features built into his equipment.

9.4 RADAR SPECTRUM UTILIZATION

Military uses of radar have increased causing a subsequent increase in the demands on the electromagnetic spectrum. These demands have created a serious concern about the formulation and implementation of better frequency management standards for radar systems. The adoption of minimum engineering design requirements is a logical first step toward the achievement of electromagnetic compatibility and improvement of the accommodation of expanding radar requirements in the limited spectrum space.

A document, MIL-STD-469, was published as an aid in achieving the above goal. Engineering design requirements in MIL-STD-469 were established to control the spectral characteristics of all new radar systems in the frequency range of 100 to 40,000 MHz.

Spectrum users are urged to avail themselves of the information in MIL-STD-469 in order to make more intelligent use of the rf spectrum.

9.5 METHODS OF REDUCING INTERFERENCE IN RADAR SYSTEMS

Less target information is available in an interference environment than during normal operation. Interference, either deliberate or unintentional, presents a formidable challenge to the weapons system. The employment of certain ECCM techniques, however, minimizes this loss of information, and the system may still be operated effectively, though perhaps at a somewhat reduced capability. This section describes the various ECCM techniques available and the means by which they counter various types of interference.

9.5.1 Linear-Logarithmic Receivers

IF amplifier stages that have an output proportional to the logarithim of the input may be used to prevent saturation of the video amplifier and to handle a wide range of input signal strengths. This type receiver has the greatest dynamic range of any receiving mode. A typical log IF strip processes strong signals 80 dB greater than low level signals without saturation, i.e., it has a dynamic range of 80 dB

It can also process weak signals in the presence of extremely strong ones. A change in signal strength of a given ratio at low level will produce the same difference in output as the same ratio of signal strength at high level. Typically, if a tenfold increase in signal strength of a weak target (i.e., from 10 to 100μv) increases the output from 2 v to 3 v (i.e., difference = 1 v), a ten-fold increase in signal strength of a strong target (i.e., from 1000 to 10,000 μv) will increase the output from 4 v to 5 v (i.e., a difference of 1 v). Thus, in an interference environment, either deliberate or unintentional, the video output of a log receiver is approximately constant and independent of the interference level.

The response is essentially linear for low-level inputs and shifts to a logarithmic response for stronger signals. A detector is connected to each IF stage, and the outputs of the detectors are added linearly. When a small signal is present, the video output will consist almost entirely of the output of the last IF stage; this condition will exist as the signal becomes larger until the last IF stage reaches saturation. After saturation of the last stage, the major portion of the video will be contributed by the second-to-last stage until it reaches saturation, etc. The response is linear as signal strength increases until saturation of the last IF stage occurs and then the response becomes logarithmic. Therefore these amplifiers are often called linear-logarithmic or "lin-log" amplifiers.

9.5.2 Constant False Alarm Rate (CFAR) Receivers

CFAR receivers have the general property of maintaining a practically constant noise level or constant number of target-like noise pulses (false alarms) at the radar receiver output. A characteristic of a CFAR receiver is that the receiver gain is automatically reduced as the radar scans through a sector containing noise interference. Noise is thereby restrained from crossing the threshold so that the operator does not see on his display an increase in noise level or an increase in the number of apparent targets due to interfering signals.

One advantage of CFAR is that the operator is not distracted by noise, and, if automatic detection circuitry is used, this circuitry is not overloaded with false targets. Also, the sensitivity of the receiver is controlled to make it possible to detect targets in the immediate vicinity of strong interfering signals. An extremely important disadvantage is that with a CFAR receiver an operator may not know that his radar is being interfered with, since noise may merely reduce the radar receiver sensitivity and prevent the detection of interference and weaker targets. Since many modern search radars employ CFAR receivers, the operator must realize the interference is difficult to detect and be alert for subtle indications of its presence.

9.5.3 Instantaneous Automatic Gain Control (IAGC)

IAGC is a signal-derived radar receiver gain control circuit designed to automatically adjust the gain of the receiver for each signal to obtain a substantially constant video peak amplitude output with a varying signal peak amplitude input. To be effective, IAGC circuits must respond to sudden changes in clutter amplitude. Typical response time is a microsecond or so; sufficiently fast to operate during the time a signal is passing through the receiver.

In a typical IAGC circuit the average dc component of the rectified second detector voltage is applied to the control grid of the next-to-last IF amplifier stage. When a strong video signal appears at the second detector, the IF amplifier acquires a bias voltage that will decrease the IF gain until the video voltage at the second detector is reduced to a value determined by the design of the circuit. In order to prevent saturation in earlier IF stages, another IAGC ring can be added there. This IAGC ring will be complete with a detector operating from an earlier stage.

The response time of IAGC can be improved by using a single degenerative ring around each IF stage. The first ring should be added to the last IF stage then to each preceding stage until adequate overall dynamic range is obtained.

9.5.4 Detector Balanced Bias Circuits

To reduce clutter fluctuation and saturation of the video amplifiers, a circuit similar to IAGC is used to bias the second detector. This circuit is called the detector balanced bias circuit or DBB and is shown in Figure 9-5(a). Another diode is added in parallel with the second detector, and output voltage from this diode is applied to bias the second detector. The amount of bias can be controlled by adjustment of the gain of the bootstrap cathode follower V_{2a}.

Unless a slow action or a delay in time is provided, desired echoes as well as clutter will be removed. To eliminate the removal of desired signals, a delay line is inserted between the diode and cathode follower as shown in Figure 9-5(b). If the delay is on the order of the pulse length there will be little distortion to the desired signals.

9.5.5 Sensitivity Time Control (STC)

Clutter, resulting from sea returns is fairly constant at all azimuth angles and is a steadily decreasing function of range. It is possible to devise a circuit that lowers the receiver gain immediately following the transmission pulse and then increase it as the range increases. Receiver gain should arrive at its maximum value just when clutter disappears.

Generally, the slope and depth of the STC curve can be varied by the operator so that sea clutter returns and strong signals from large nearby targets do not cause receiver saturation from antenna sidelobe returns. In most radars, STC influences receiver gain only on short ranges, but does not affect radar operation at longer ranges. When using STC, the operator should be careful that adequate receiver gain is maintained at short ranges if he wants to detect targets at these ranges.

9.5.6 Moving Target Indicator (MTI)

When only moving targets need be displayed in a radar system, a method is available to eliminate echoes from stationary or slow moving targets. Land masses, clouds, sea returns, chaff, etc., could then be prevented from cluttering up the PPI scopes and obscuring desired returns from fast moving aircraft or ships.

The basic principle for all MTI systems is the doppler effect. RF energy striking a moving target undergoes a frequency shift which is proportional to target speed. Thus reflected energy from a moving target is returned to the radar at a different frequency from the transmitted signal. This frequency shift is called the doppler frequency; and it is this difference in reflected signals from moving and stationary targets that permits selective discrimination and subsequent processing.

An A-scope showing stationary targets and a moving target is illustrated in Figure 9-6. Stationary targets exhibit constant phase and a fixed amplitude from pulse to pulse. The moving target gives a butterfly-like appearance as a result of varying phase from pulse to pulse.

The frequency of the reflected signal is given by

$$f' = \frac{c+v}{c-v}(f)$$

where

f' = reflected signal frequency in hertz
f = transmitted signal frequency in hertz
c = speed of light in miles per hour
v = radial velocity of target in miles per hour.

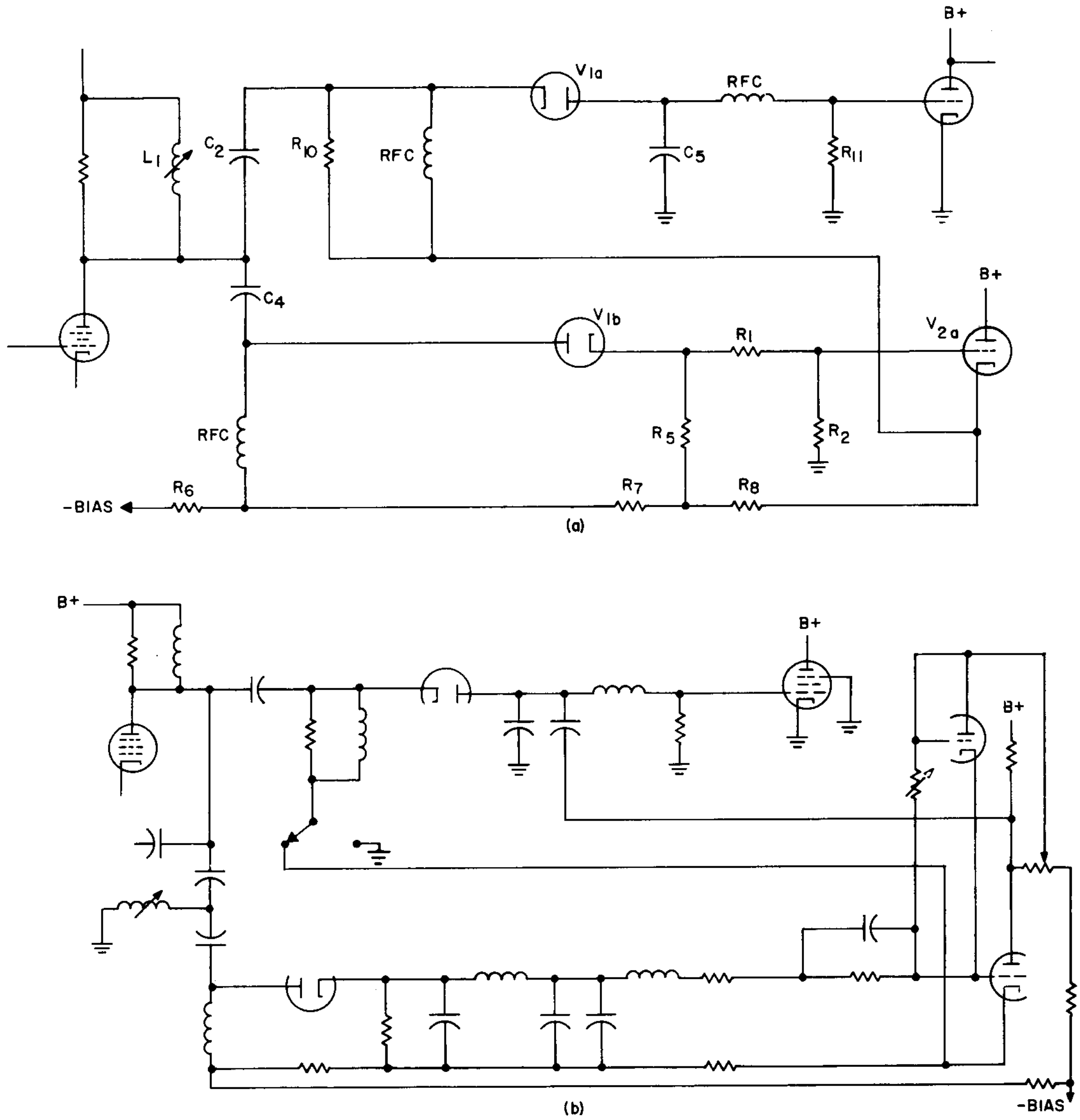


Figure 9-5. Detector Balanced Bias Circuits

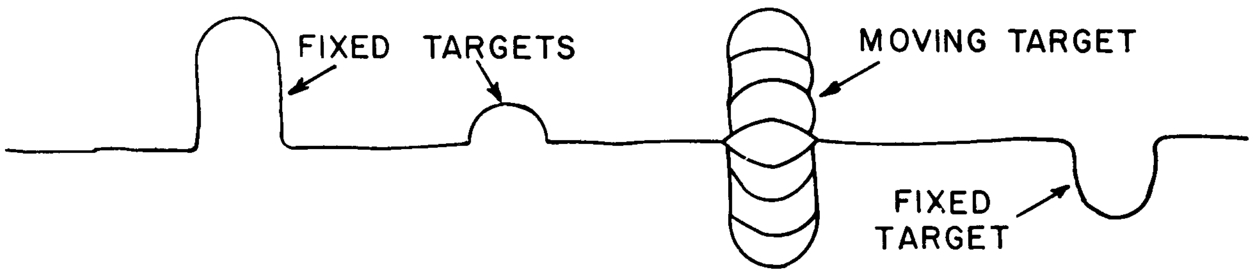


Figure 9-6. A-Scope of a Moving Target and Stationary Targets

If the transmitted signal is mixed with the reflected signal, a doppler beat frequency f_d, results:

$$f_d = (89.4)\frac{v}{\lambda}$$

where v = radial velocity of target in miles per hour
λ = wavelength in centimeters

Several methods exist for recovery of the doppler frequency and each method has its advantages and disadvantages. Two of the more important ones are (1) coherent MTI systems, and (2) noncoherent MTI systems.

9.5.6.1 Coherent MTI. In the coherent detection process, target return signals are mixed in a phase detector with a reference signal from a coherent oscillator that is locked in phase and frequency to the transmitted pulse. See Figure 9-7. A reference oscillator provides a CW signal to beat with the echo signal. Since the transmitter starts with random phase from pulse to pulse, the reference oscillator must be matched in phase to the transmitter for each transmitted pulse. This can be accomplished by allowing a sufficient amount of power from the transmitter to enter the resonant cavity of the reference oscillator, which then forces the oscillator into step with the transmitter.

Echoes from stationary targets produce pulses which do not vary in amplitude or phase on successive returns. A delay line, with a delay time exactly equal to the pulse period, stores the target returns from one period so a comparison can be made between echoes on successive returns.

When delayed and undelayed video returns are compared in a subtractive process, the fixed returns cancel; but fluctuating returns from moving targets leave a residue. This residue represents a moving target on the display.

The principle of pulse cancellation is illustrated in Figure 9-8. Sweep 1 is delayed for a repetition period and then subtracted from Sweep 2, etc. As the stationary targets have constant amplitude from pulse to pulse, the result of cancellation is evidenced in no stationary targets.

A practical MTI system is shown in Figure 9-9. Instead of beating the signals at radio frequencies, the beating is done at the IF by applying the superheterodyne principle to both the echo signal and the locking pulse. A coherent locking pulse (coho lock pulse) from the first mixer which is in phase with the transmitted signal is applied to a coherent oscillator. The frequency of the coho oscillator is the receiver IF.

9.5.6.2 Noncoherent MTI. If the radar platform is moving, as on a ship, there will be a doppler frequency even on stationary targets. Any relative motion between the target and radar antenna will cause the reflected signal to return at a frequency different from the transmitted signal. The effect is one of incomplete cancellation of stationary targets.

This disadvantage can be overcome if the signal from a stationary target is used as the reference signal. The reference oscillator is forced into exact frequency and phase with the clutter and it is this signal which is compared against other signals. Fast moving targets when compared with the slower moving clutter still give a fluctuating output. Returning signals from other clutter, which have little or no motion relative to the reference clutter, will cancel. In effect then, the use of clutter as a reference, instead of a sample of the transmitted pulse, will eliminate the effects of target platform movement and will give good clutter cancellation. Moving targets will still be displayed.

9.5.6.3 Automatic Clutter Gating. Some MTI systems have provisions for detecting clutter and gating on MTI circuits only when clutter is present. This can be done with coherent and noncoherent systems. In the noncoherent MTI with this automatic gating feature, the first bit of returning clutter does two things: (1) in turns on the MTI circuits, and (2) it becomes the sync or locking signal for the reference oscillator.

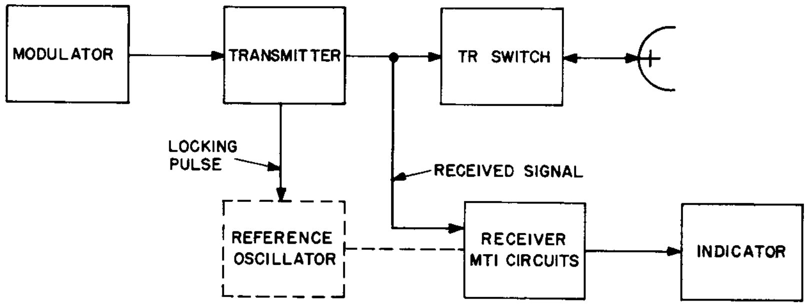


Figure 9-7. Doppler System for Microwave

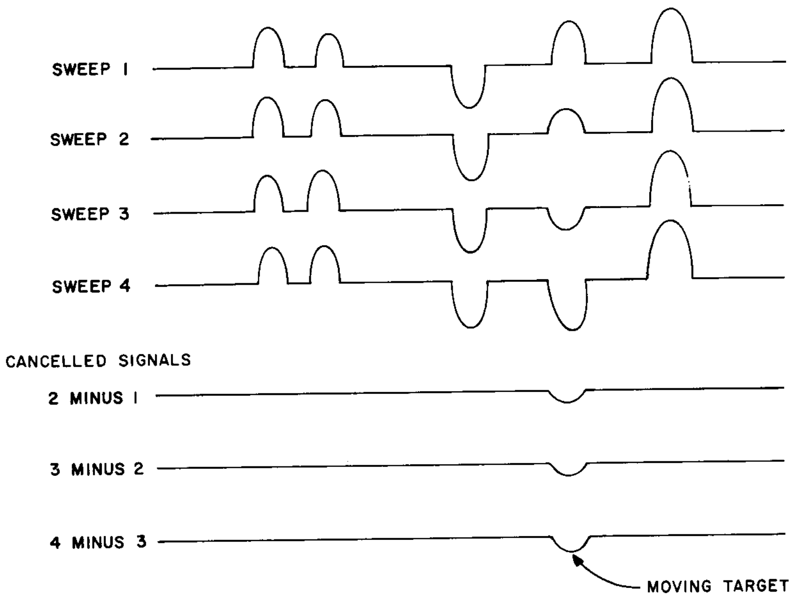


Figure 9-8. Pulse Cancellation

9.5.6.4 Sector Gated MTI. This mode of MTI operation requires a manual adjustment by the operator of "front panel" controls to gate on the MTI circuits in a chosen sector. The areas selected normally contain clutter that the operator wants to eliminate from the display.

9.5.7 Video Integration

Video integration in search radars is particularly effective against nonsynchronous interference because this technique integrates signals that repeat at the same range, and discriminates against pulses that do not repeat at the same range. It thus enhances the detection of weak signals in noise or in continuous interference and is particularly effective against other radar interference. The most common form of video integration is delay line storage in which the signals from one pulse repetition period are delayed exactly on repetition period in passage through the delay line, and are added to those in the next repetition period. The sum signal is recirculated through the delay line and added to the returns from the succeeding pulse. This summation process lasts effectively through a large number of repetition periods. The result is that a number of video pulses occurring at the same range (characteristic of a true target) are added together (integrated) and thereby built up. Noise pulses, on the other hand, do not recur repeatedly at the same time interval and do not build up. The video integrator builds up synchronous signals and performs its integration and discrimination automatically, and its output can be fed directly to automatic signal processing equipment.

9.5.8 Fast Time Constant (FTC)

This circuit is used to break up low frequency or dc returns from clutter, or jamming pulses longer than the transmitted radar pulse. It is essentially a differentiating or high-pass video circuit using a short time constant network. Short echo pulses from targets will be mostly unaffected. Use of FTC gives an apparent three-dimensional view on the PPI display.

9.5.9 Side-Lobe Blanking (SLB)

The direction or azimuth from which the interference reaches the radar system can be used to discriminate between desired and undesired signals. A device utilizing the azimuth of the arriving pulse is the side-lobe blanker, shown in block diagram form in Figure 9-10. To the normal radar system is added another receiver with an omnidirectional antenna. The receiver is fed by the same local oscillator as is the normal radar receiver, and they both have

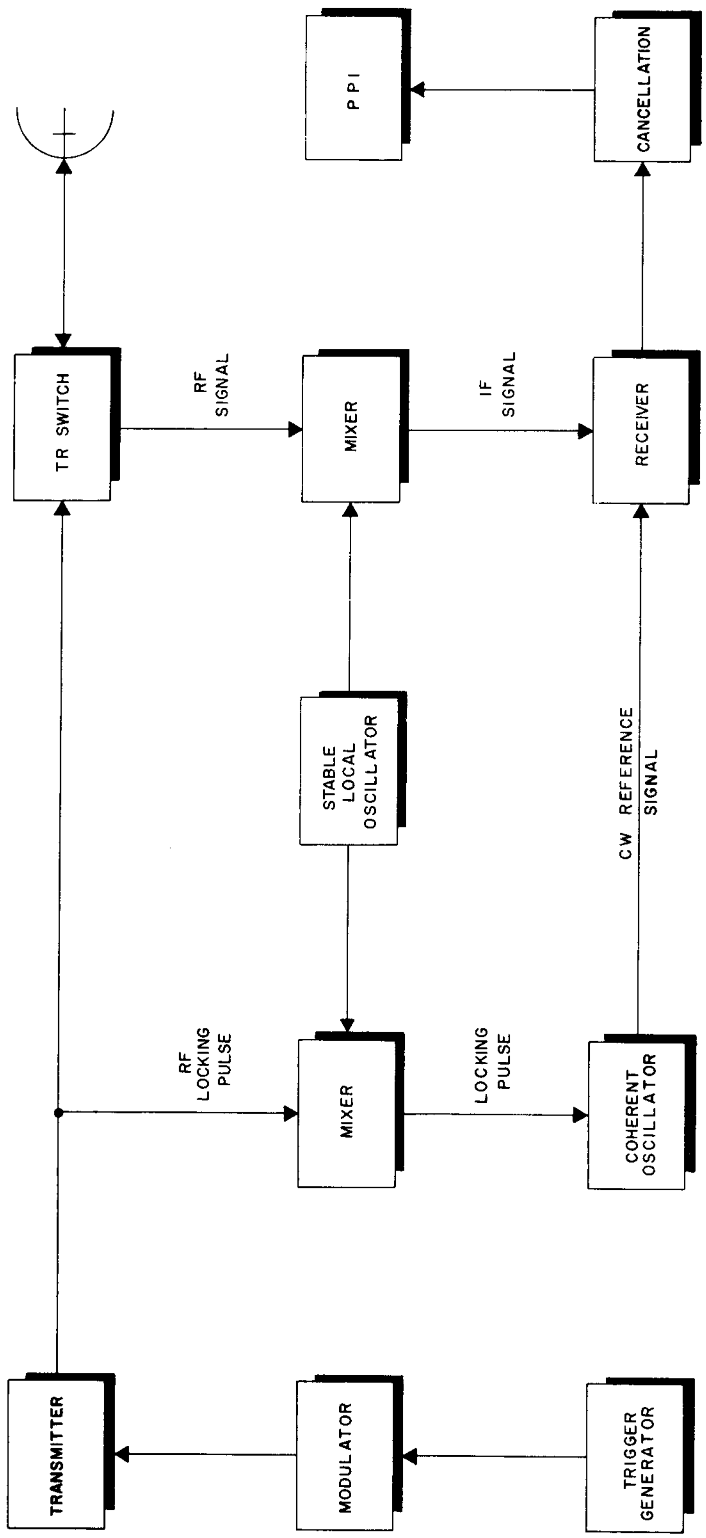


Figure 9-9. Practical MTI System

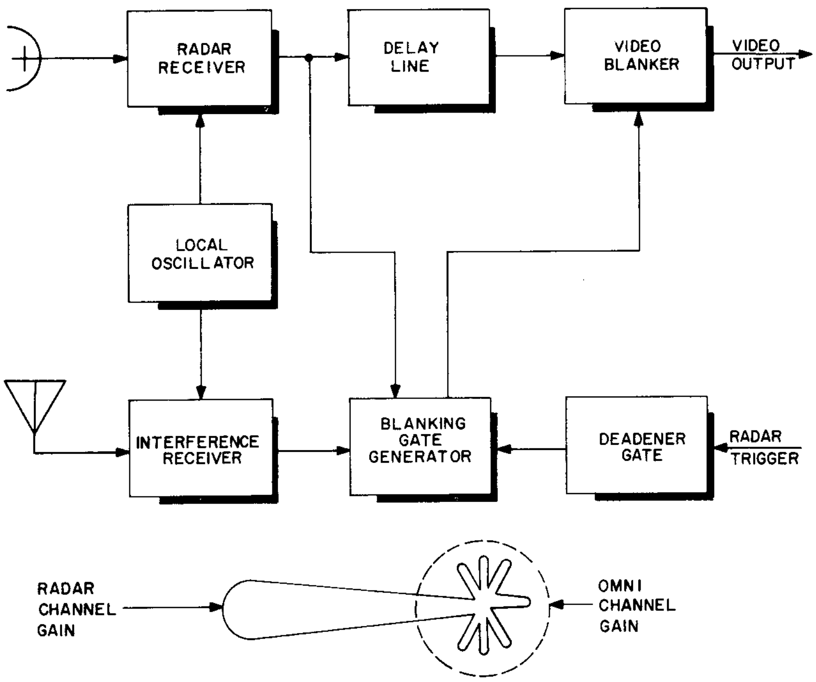


Figure 9-10. Side-Lobe Blanker Block Diagram

identical IAGC circuits so that the gain of the normal radar channel and the auxiliary or interference channel bear a constant relationship.

The normal radar antenna and the auxiliary antenna have identical patterns in the vertical plane. The gain of the interference channel is adjusted to be larger than any of the side lobes of the normal radar channel in the horizontal plane. The main-lobe gain of the normal channel, however, is much larger than that of the interference channel. See Figure 9-11. If a signal enters on the side lobes of the normal channel, the output from the radar receiver will be smaller than the output from the auxiliary receiver; however, if the signal arrives at the main lobe, the radar receiver output is larger. Since desired signals enter only through the main lobe, any signal entering by the side lobes can be assumed to be an interference signal.

The outputs of both receivers are fed to the blanking gate generator and compared. If a signal from the auxiliary channel is the larger, a blanking gate is generated. The blanking gate prevents the signal in the radar channel from reaching the display. A delay is necessary in the radar video to compensate for the time required to generate the blanking gate. A deadener gate, operated by the system trigger, is used to keep the side-lobe blanker from operating immediately following the transmitted pulse.

9.5.10 Dicke Fix (WBL)

The bandwidth of a normal radar receiver is matched to the bandwidth of the radar pulses to optimize the overall radar sensitivity; the required bandwidth being an inverse function of the pulse length. When a strong noise spike enters such a receiver it causes the narrow band IF and video coupling circuits to ring or oscillate at their resonate frequency. Such ringing results in a video output pulse that is stretched to about the normal radar pulse length. This video pulse resembles a target signal and constitutes a false alarm. Strong short pulses can saturate the receiver and cause it to ring longer than one pulse length. If the short pulses have a high repetition rate, as in a case of deliberate interference, they can keep the receiver ringing constantly and thereby prevent reception of normal signals.

Dicke Fix (sometimes called wideband limiting) is a method to combat these "False Alarms." It has a three-step operation for processing received signals: (1) wideband amplification to avoid "ringing," (2) amplitude limiting of all signals at a CFAR threshold, and (3) bandwidth limiting to separate normal radar signals from noise or interference spikes.

A Dicke Fix receiver consists of a wideband amplifier with "hard" amplitude limiting, followed

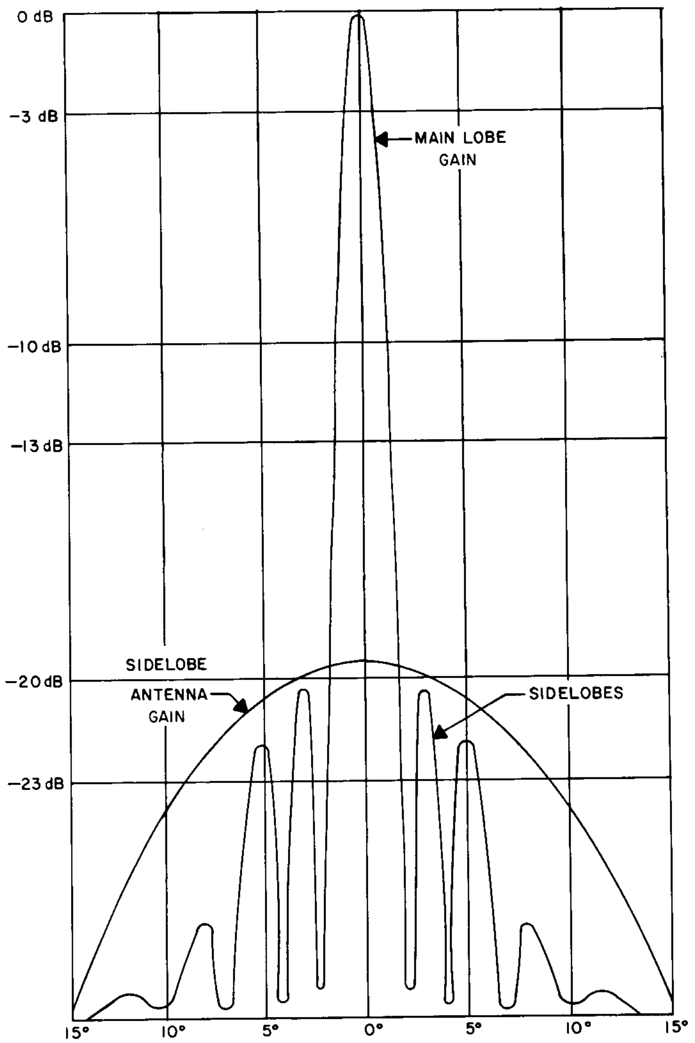


Figure 9-11. Typical Side-Lobe Antenna and Main Antenna Gain Patterns

by a normal receiver with bandwidth limiting. The wideband limiting amplifier causes all signals - normal radar signals, short and long pulse jamming signals, and noise jamming signals - to be amplified without ringing, then limited to the same maximum amplitude, generally near or below the radar noise level. Because the bandwidth is wide, the amplifer stages are capable of responding to a fast-rising narrow pulse input without ringing. Amplitude limiters prevent signals strong enough to cause significant ringing from being delivered to subsequent stages.

The bandwidth of amplifier stages following WBL is optimized for best recovery of desired echo pulses. Since most of the spectral energy of narrow pulses lie outside this bandwidth, they are not recovered at the detector at the same amplitude as normal radar signals.

9.5.11 Lamb Silencer

Another circuit designed to discriminate against off-frequency interference is the Lamb noise-silencing circuit. This circuit places a wideband IF amplifier and limiter ahead of the normal bandwidth IF stages, as shown in Figure 9-12 (a). Figure 9-12 (b) shows how the sidebands of an off-frequency interference signal can cause interference in the normal bandwidth receiver. When a wideband amplifier and limiter are used ahead of the normal channel with bandwidths shown in Figure 9-12 (c), the sidebands of the interference are reduced to a tolerable level as shown by Figure 9-12 (d).

The limiting level is adjusted so that no information will be lost on desired signals.

9.5.12 PRF Synchronization and Blanking

When interference is present from other nearby radars, it can be eliminated by pulsing all the radars simultaneously from a common trigger source. Since the radar receivers are disabled when the transmitters pulse, the high pulse energy is prevented from intruding and causing interference on associated radar indicators. PRF synchronization can be used only between radars having the same basic repetition rate.

When PRF synchronization is not possible, another method is used to blank or disable the offended equipment during radar pulse time. A pre-trigger from the offending radar is used to generate a blanking gate of sufficient width to disable susceptible receivers. A typical shipboard blanker would have the following features: (1) Provisions for multiple inputs and outputs, (2) Both negative and positive outputs, (3) Adjustable blanking gate width and amplitude, and (4) Variable delay between pre-trigger time and start of blanking gate.

Blankers with the above features would be completely versatile and could be used to blank several offended equipments during the pulse time of any offending radar.

9.5.13 Spectrum-Centered Receiver

9.5.13.1 Description. The spectrum-centered receiver is designed to be added to existing radars to effectively eliminate off-frequency pulse interference. The unique bandpass of the receiver and characteristics of the interfering pulse are utilized to cause cancellation of the pulse. In a dense radar environment, the spectrum-centered receiver provides immunity to mutual interference between radars operating in the same frequency band.

The spectrum-centered receiver is a completely transistorized unit capable of operating with most radars having a 30-MHz intermediate frequency. It consists of a low noise preamplifier, several stages of IF amplification and video gating circuitry. The receiver can operate on either positive or negative

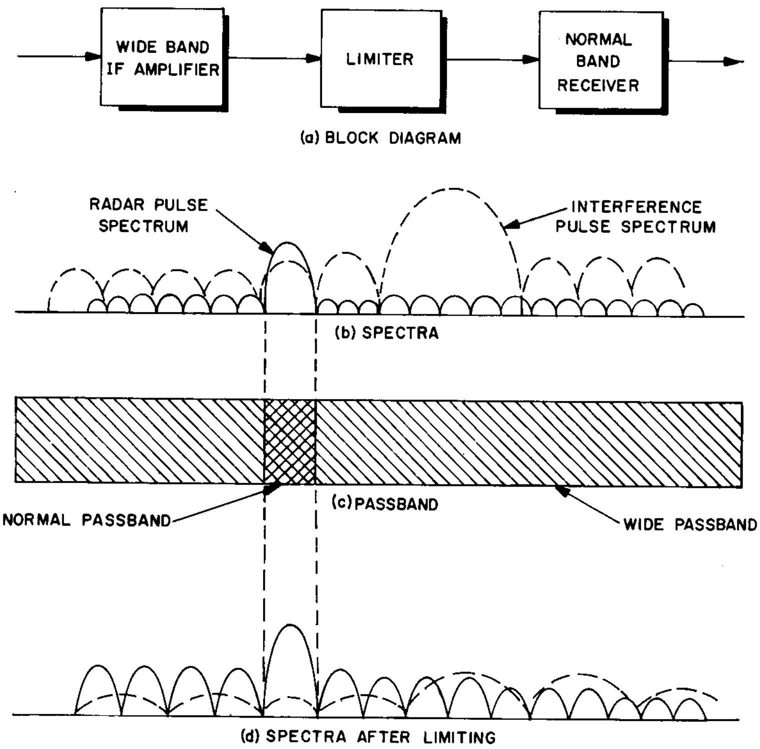


Figure 9-12. Lamb Noise-Silencing Circuit

video from the radar. The spectrum-centered receiver model 6620 employs single channel video; the model 6619 has dual channel capability for MTI operation. Operation is from 24 vdc which can be obtained either from the radar or an optional power supply available for converting 115 vac 60 to 400 Hz, into 24 vdc. Manufacture of the spectrum-centered receiver is based on Military Specification MIL-E-16400.

9.5.13.2 Off-Frequency Interference. Mutual interference among radars operating in the same general area is usually minimized by operating the equipments at different frequencies. However, the spectrum of the transmitted pulse from a radar set can extend well beyond the bandwidth of its own receiver. See Figure 9-13 (a). Consequently, extensive interference energy is received by other radars, although their operating frequencies may be far removed from the frequency of the interfering radar. See Figure 9-13 (b).

The usual radar receiver is tuned to f_1, Figure 9-13 (b), the center frequency of the magnetron output, and the receiver passband is sufficiently wide to encompass all the centerlobe energy returned from a target. Note that sideband energy from an undesired signal f_2 can fall within the receiver passband f_1.

Since this undesired signal travels directly from the offending radar to the victim receiver, it could be much stronger than the desired echo from a target and can cause considerable interference on the radar display.

9.5.13.3 Theory of Operation. The principle of operation of the spectrum-centered receiver is that the spectrum of the desired radar return is centered about the receiver passband, whereas the spectrum of an undesired signal is offset from the receiver passband. The spectrum-centered receiver, which has a wide passband, detects off-frequency energy and excludes energy in the radar receiver passband. This is depicted in the frequency response curves of Figure 9-14.

9.5.13.4 Functional Operation. The functional operation of the spectrum-centered receiver is described by means of the system interconnection diagram (see Figure 9-15).

Radar energy at the intermediate frequency of 30 MHz is fed from the radar mixer diode directly to the spectrum-centered receiver. A low-noise preamplifier in the spectrum-centered receiver amplifies and then couples a portion of the signal out of connector J2 to the input of the normal radar IF amplifier. This is done to compensate for the 3-dB loss due to division of the signal. The radar receiver therefore does not suffer a loss of sensitivity, as it would if a passive power divider were employed.

The spectrum-centered receiver functions as a wideband amplifier with a deep attenuation notch at

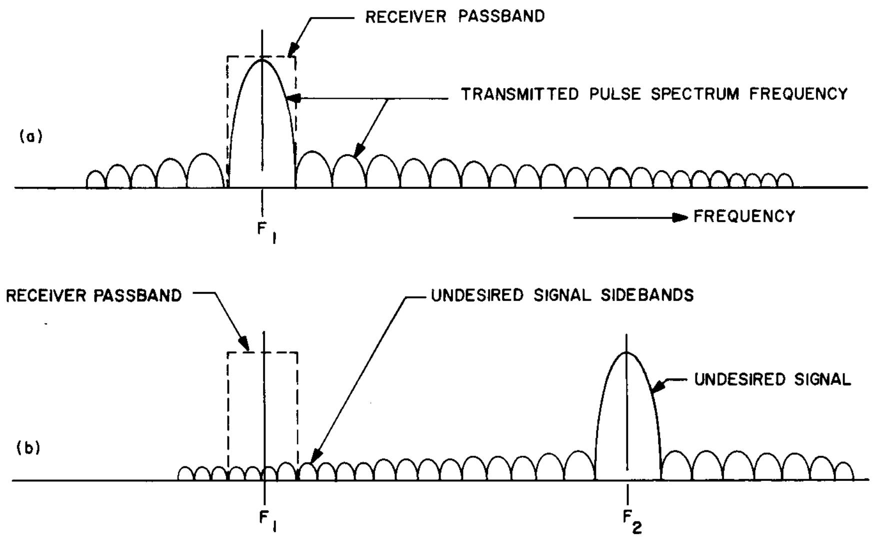


Figure 9-13. Spectrum of Desired and Undesired Signals with Respect to Receiver Passband

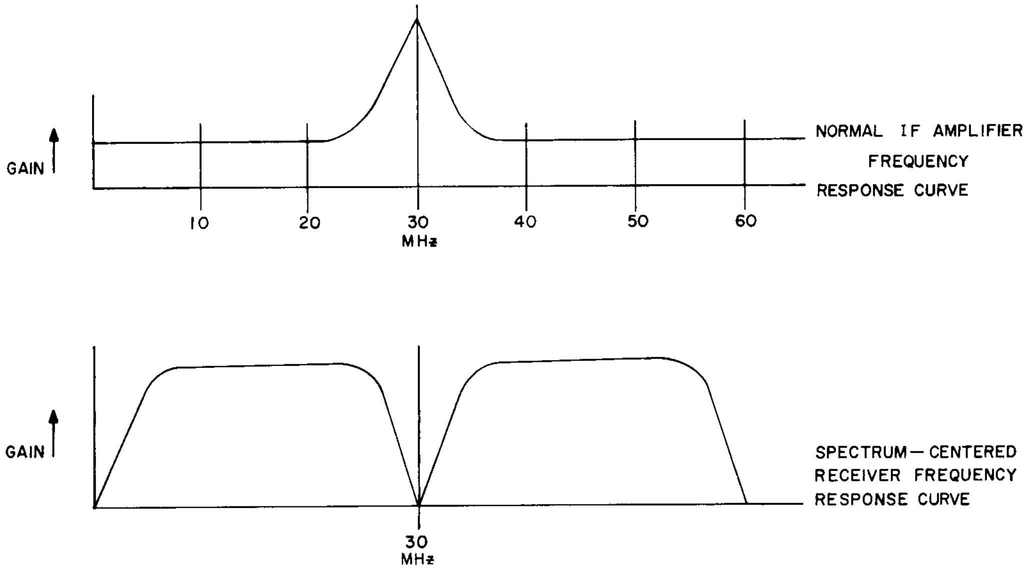


Figure 9-14. Idealized Frequency Response Curves

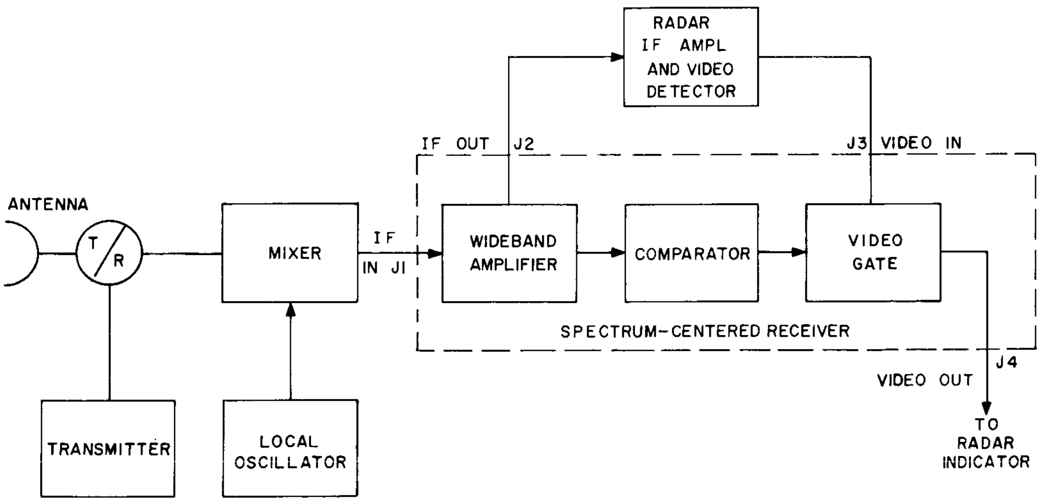


Figure 9-15. Spectrum-Centered Receiver-System Interconnection Diagram

the normal intermediate frequency of 30 MHz. In the event an off-frequency signal is detected by the spectrum-centered receiver, and is of sufficient amplitude to overcome the threshold level of the comparator circuit, a gate signal is generated. This gate signal cuts off the video gate diode thereby inhibiting the flow of video information to the radar indicator.

Note that the radar receiver operates as it normally does in the system. The spectrum-centered receiver merely senses undesirable off-frequency signals and causes the video signal to the radar display to be cut off during this time.

The video gate is designed to accommodate video of varying characteristics (including positive, negative, and bipolar), and amplitudes as high as 12 volts, peak-to-peak. The bipolar video gate is most useful in those radar systems employing moving target indicator (MTI) capability. The nature of the video requires careful design of gating circuitry to ensure minimum transient generation.

The video gate is buffered from the input and the output by means of an emitter follower; the output impedance, nominally less than 50 ohms, may be modified to meet a particular requirement.

9.5.14 Pulse Compression-Expansion

A technique for raising the subclutter visibility of an echo in a noisy environment is that of frequency-coding the transmitted pulse, then separating wanted echoes from noise through recognition of the coding.

Coding is applied to the transmitted signal by routing a short trigger spike through a frequency-sensitive "Expansion" filter. The high frequency components of the pulse transit the filter faster than do the low frequency components (for a highpass expansion filter) so that it becomes a "stretched" pulse of long duration whose carrier frequency is swept, or "chirped" at a known rate.

The advantage of a long transmitted pulse for increasing detection range has already been demonstrated. The increased average power of the longer transmitted pulses increases the detection range of the radar compared with a short-pulse radar having the same peak power. A loss of radar resolution can be expected from a long pulse, but if the frequency of the pulse is swept during the pulse width, the advantages of a long transmitted pulse can be realized, and then the lost resolution recovered by "compression" in the receiver.

The received echo signal is passed through a compression filter which matches the pulse expansion of the transmitter. The frequency of the leading edge of the pulse is delayed most and that of the trailing edge least. Thus, successive parts of the pulse are delayed progressively less so that they are compressed or "stacked" to form a short pulse of higher amplitude. The disadvantage of PC radars is that the compressed pulses are accompanied by "residues" or filter transients before and after the compressed pulse. These residues are typically one-tenth to one-fifteenth the amplitude of the compressed pulse (i.e., 20 to 34 dB down), and they constitute unwanted "noise" extending before and after the compressed pulse. A signal limiter is frequently used ahead of a PC network to suppress the PC residues, but it also suppresses weak targets that overlap strong ones. This suppression sometimes produces a "swiss cheese" effect on the PPI when holes in a clutter background result from the suppression of weak clutter by strong overlapping individual clutter or target signals.

Frequencies from the expansion filter have to be translated to the final transmit frequency by frequency converters. A technique known as "chirp" is utilized to shift the frequency distribution in the pulse. See Figure 9-16.

9.5.15 Frequency Agility (FA)

The rapid change of operating frequency, generally on a pulse-to-pulse basis, is called frequency agility. The three-dimensional frequency scanning radars presently in the fleet possess inherently a large degree of frequency agility. Since the operating frequency is constantly changing, any discrete interference frequency in the scanned band will interfere at one frequency only, which does not prevent the radar from obtaining targets on the clear frequencies.

9.6 RADAR-TO-RADAR INTERFERENCE

9.6.1 Effects of Radar-to-Radar Interference

Interference between radars usually appears as a series of moving dots, "running rabbits," on the PPI scopes. It occurs when several ships with the same type radars are operating within a few miles of each other. It can also occur between radars of different types aboard one ship. Since the offending radar is not at exactly the same PRF, the dots are moving radially along the indicator sweep creating spiral interference patterns. A trained operator can usually read through this interference; however, if the pulses are sufficiently high, they can obscure desired target pips.

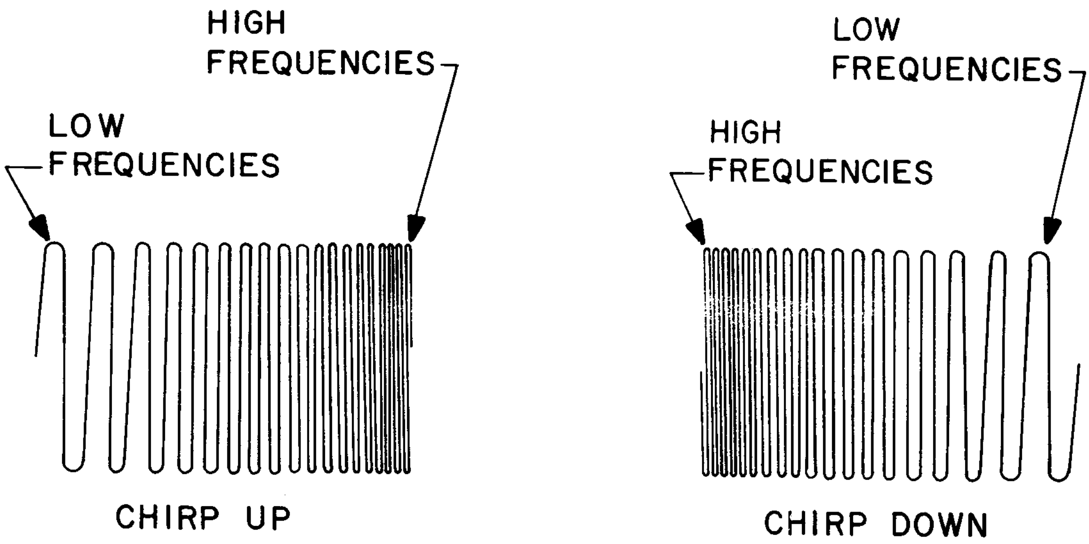


Figure 9-16. Typical "Chirp" Frequency Distribution

9.6.2 Reduction Techniques

9.6.2.1 Frequency Separation. Sometimes, the problem can be overcome or minimized by shifting the offended radar to another operating frequency. This technique is feasible when the spectrum distribution of the individual radars extend only slightly beyond the center frequency of the pulse. However, when rectangular pulses are involved and the radars operate in the same frequency band, there may not be sufficient space for adequate frequency separation to avoid mutual interference. In that case, other methods of reducing interference must be utilized.

9.6.2.2 Video Integration. The use of one or more ECCM features such as log receiver, video integration, etc., will reduce the effects of interference. Video integration is the most effective method of eliminating pulse type interference from an offending radar. Ships having this feature should ensure it is kept fully operational for just such a need. See paragraph 9.5.7.

9.6.2.3 Bandpass Filters. Bandpass filters are incorporated in some high power radar transmitters to suppress out-of-band harmonics or other spurious outputs which are potential interference sources. These filters are connected between the final stage and the antenna, usually inserted directly in the transmission line. Research is continuing on development of filters with higher attenuation and power handling capability for use with higher powered radars. These filters also reduce exposure of the associated radar receiver to out-of-band rf energy, which helps reduce the number of spurious signals that get into the receiver.

9.6.2.4 Antenna Placement. Ship designers should be aware that radar antenna placements affect the interference picture as well as the radiation patterns of the associated radars. It is never good practice to have one radar antenna "looking directly at" a second radar antenna even if the radars are widely separated in frequency. High energy out-of-band rf can force its way past the pre-selection of a radar receiver, causing undesired responses or injuring sensitive components such as crystal mixers, rf amplifier tubes or transistors.

Because of these problems, radar antennas should be separated vertically as well as horizontally. This will reduce the cross coupling of rf from one system to the other.

Another facet of antenna placement is one of blind spots or dead sectors in the radiation pattern caused by large metal objects in the main radar beam. The problem is more acute at microwave frequencies since bending of radar beams around the obstacle is less.

Since it is obviously impossible to have all radar antennas completely in the clear, certain trade-offs have to be accepted. However, ship designers should be aware of the problem areas in order to arrive at intelligent compromises with conflicting requirements.

SECTION 10 – EMI REDUCTION METHODS

Ideally, there would be no need for EMI reduction methods once the electronic systems are installed aboard ship. Ship design and equipment design personnel would have designed, manufactured and installed the equipments in such a manner as to have anticipated and eliminated all potential interference situations. Although this may be an ideal goal, it is not yet fully realizable. Therefore field EMI reduction techniques are still necessary.

This section of the book is devoted to a discussion of the most practical methods for alleviating shipboard EMI. Among the many methods available are (1) bonding/grounding to reduce interference currents; (2) shielding of the interference source or of the susceptible equipment; (3) antenna placement to reduce mutual interaction between antennas; (4) filtering to reject unwanted signals and pass desired signals; (5) use of nonmetallic materials that are electromagnetically transparent, hence do not affect radio wave propagation; (6) use of computer prediction techniques for frequency assignments which will avoid potential interference situations; (7) use of one carrier frequency to transmit many information channels (Multiplex); (8) use of blanking devices to disable a susceptible equipment during the time an interference pulse is present; and (9) adequate training and motivation of equipment operators and maintenance personnel in order that designed-in EMI reduction features of electronic systems are fully utilized and maintained.

This last item is all too easily overlooked and sometimes causes devastating consequences. Loss of information, caused by (1) equipment malfunction due to lack of preventative maintenance or (2) sloppy operating techniques due to lack of motivation or training, can seriously affect the outcome of a tactical situation.

10.1 BONDING - GROUNDING

Bonding is the process of providing a low impedance union between two metallic conductors. Grounding implies an extension of this definition to include the establishment of a common reference point with regard to electric potential. Bonding, or the formation of a fixed union, can be accomplished by a wide variety of means, each of which has the same objective: to obtain minimum resistance between the two surfaces to be bonded. For reduction of electromagnetic interference it is generally the case that the best method is the one in which the joint becomes a homogeneous mass of the joined metals. This is accomplished by welding or brazing. Of lesser value are (1) methods that make extremely good physical contact such as pressure connections obtained by bolting or riveting and (2) installation of a low rf impedance bonding strap that acts as a low linear impedance shunt around a deteriorated junction.

The importance of equipotential ground planes cannot be over-emphasized for proper equipment operation, EMI suppression, and personnel safety. An equipotential ground plane implies a mass, or masses, of conducting media which, when bonded together, offers a negligible impedance to current flow. A single connection that offers a significant impedance to current flow can place an entire grounding system at a high potential with respect to ground. Shielding connected to the system would then be completely ineffective. When such a high impedance connection is subject to large amounts of current, resultant potentials can be extremely hazardous to personnel and equipment. Oxides may form at mating surfaces of metallic media which will greatly increase the impedance of the bond and form nonlinear systems capable of generating and radiating various harmonic signals.

Effective bonding techniques must be utilized to prevent degenerative and hazardous effects of high impedance bonds. Compatible test procedures and limitations must be understood clearly to ensure that adequate bonding techniques are implemented properly.

10.1.1 General Bonding Techniques

There are two general techniques for bonding: direct and indirect. Direct bonding is attained by direct metal-to-metal contact between two metal items through the process of welding or brazing. Indirect bonding is attained by the use of a bonding jumper or bolts between two metal items. Direct bonding is always the preferred method, since the microscopic path length involved in direct bonding has a lower rf impedance than that of any practical indirect bonding method.

Indirect bonding is at best only a substitute where physical constraints make it impractical to use direct bonding. When bonding by the indirect method, the length of the bond path should be as short as possible. In addition, when using bond straps, the welded-at-both-ends bond strap is preferred over the removable-end strap.

10.1.2 EMI Contributors

EMI, in the form of arcing or intemodulation, can be generated by many shipboard topside items. A few of the major EMI contributors aboard ship are (1) metallic lifelines and chains; (2) boom cables, winch cables and other long cables in the vicinity of transmit or receive antennas; (3) corroded armored cables in the illumination path of high-powered radars; and (4) any loose, corroded metallic junction between metal items in topside areas when they have sufficient physical size to intercept appreciable rf energy from shipboard transmitters.

10.1.3 Bonding Specifications

MIL-B-5087A states that the bonding impedance must be less than 80 milliohms at all frequencies below 20 MHz. To meet this requirement, calculations show that the diameter of a simple round bond strap one inch long must be 1.3 inches. This is an impractical size for most applications. Consequently, it is almost impossible to meet the idealized requirement for bond strap impedance given in MIL-B-5087A.

MIL-STD-001310B (SHIPS) is the present authority for all bonding/grounding procedures aboard Navy ships. Items to be bonded, bond strap fabrication, installation details and attachment methods are all described in the above mentioned publication.

This publication (MIL-STD-001310B (SHIPS) covers three main areas of interest:

a. Bonding/grounding
   1. items to be bonded
   2. bond strap fabrication
   3. attachment methods
b. Use of nonmetallic materials
   1. replacement of metal lifelines with a glass base line.
   2. replacement of metal handrails, stanchions, flagstaffs, lockers, and other topside items with fiberglass or epoxy equivalents.
c. Isolation (insulation) of portable items from contact with ship hull
   1. gas bottles
   2. portable tools
   3. foam nozzles

10.1.4 Bond Strap Types

The most commonly used bond strap for the reduction of hull generated interference is made from extra flexible welding cable TRXF-84 as specified in MIL-STD-001310B (SHIPS). This publication describes the strap length, attachment methods, and fabrication materials, including end hardware. The strap, which is waterproof and extremely flexible, is ideally suited for use in a rugged shipboard environment. For instance, if the strap is fastened across a door hinge, or the like, it will withstand the rigors of usage. In those cases where both ends of the strap can be welded, then permanency is assured with little or no maintenance required.

Sometimes it is necessary for a bonded item to be removable; thus, the bond strap is welded at one end only. The other end is punched to allow for attachment to a threaded stud. This end of the bond strap can then be removed quickly to allow removal of the bonded item. A bond strap installed in this manner must be given more attention than a bond strap welded at both ends. Otherwise, the bolted end contact surface may deteriorate because of corrosion, thereby causing a high resistance connection.

The third type of bond strap is a cadmium-plated copper strap used to bond electronic equipments to ground potential internal to the ship.

The fourth type of bond strap is a zinc-plated hose clamp strap. This strap is used for bonding of rf round transmission lines and conduit, both internal and external, on a ship.

10.1.5 Bonding Applications

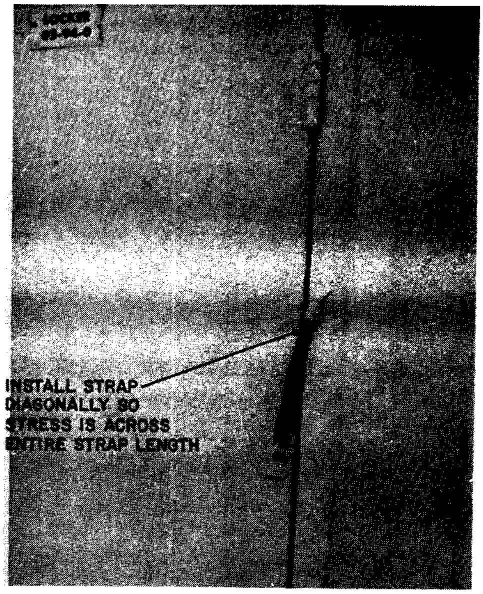
It is difficult to illustrate and discuss all of the items aboard ship that could become interference sources and that should be bonded. Rigid bonding procedures whereby every metallic junction is bonded without regard to its location, configuration, or physical size, are prohibitively expensive and time consuming. Personnel who work in accordance with the requirements of MIL-STD-001310B (SHIPS) should assign priorities and install bond straps selectively so as to obtain maximum effectiveness with available funds. In difficult cases, it sometimes becomes a matter of interpretation based on experience in deciding whether or not an item should be bonded. Investigations are continuing to develop a simple, fast method of locating and assessing the impact of potential EMI contributors in topside structures.

Figures 10-1 through 10-5 illustrate some of the items aboard ship that require bonding in addition to illustrating acceptable techniques for strap installation.

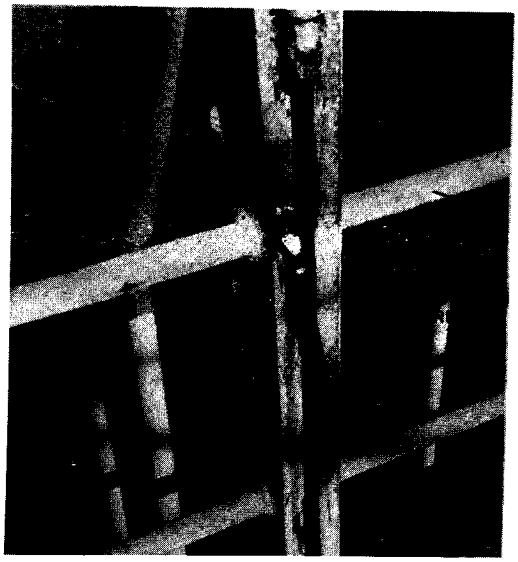
Following are some general guidelines to follow in the use of bond straps.

a. Mast Guy Turnbuckle and Shackles
Figure 10-1. Typical Bonding Applications

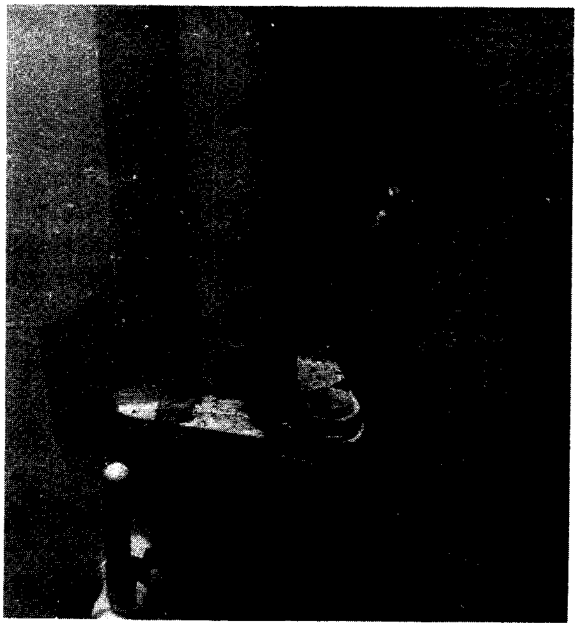
b. Loading Davit
Figure 10-1. Typical Bonding Applications



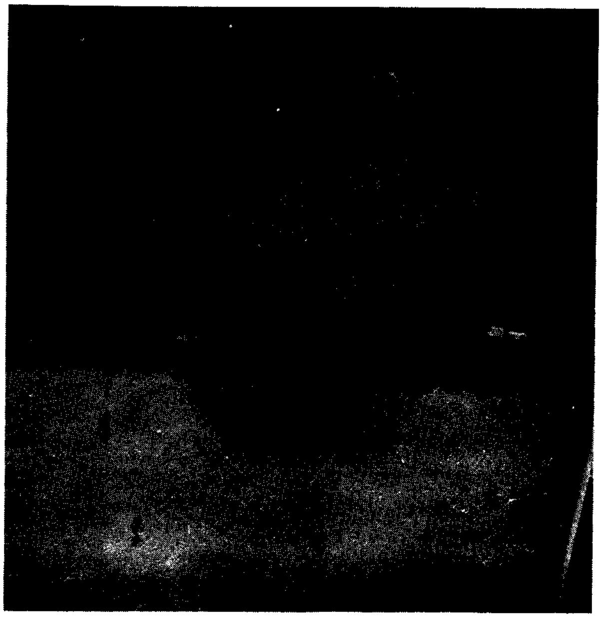
a. Flush-Mounted Locker Door
Figure 10-2. Typical Bonding Applications



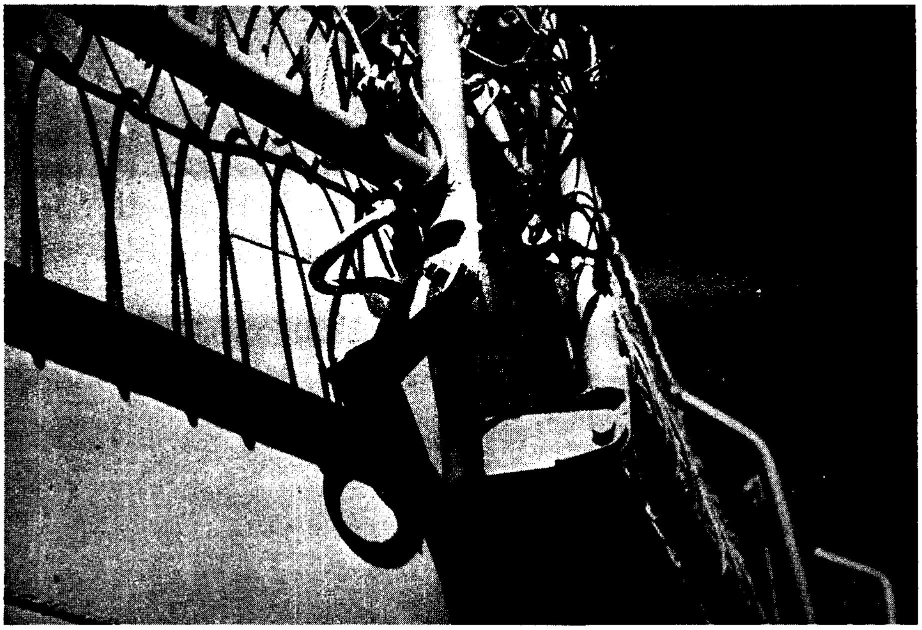
b. Lifeboat Access Safety Gate
Figure 10-2. Typical Bonding Applications



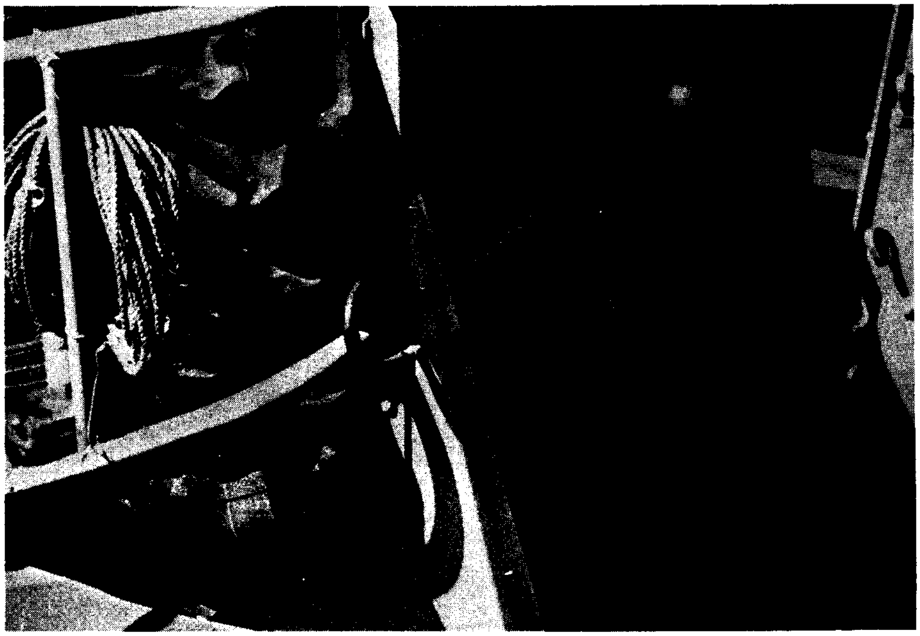
a. Watertight Door Hinge
Figure 10-3. Typical Bonding Applications



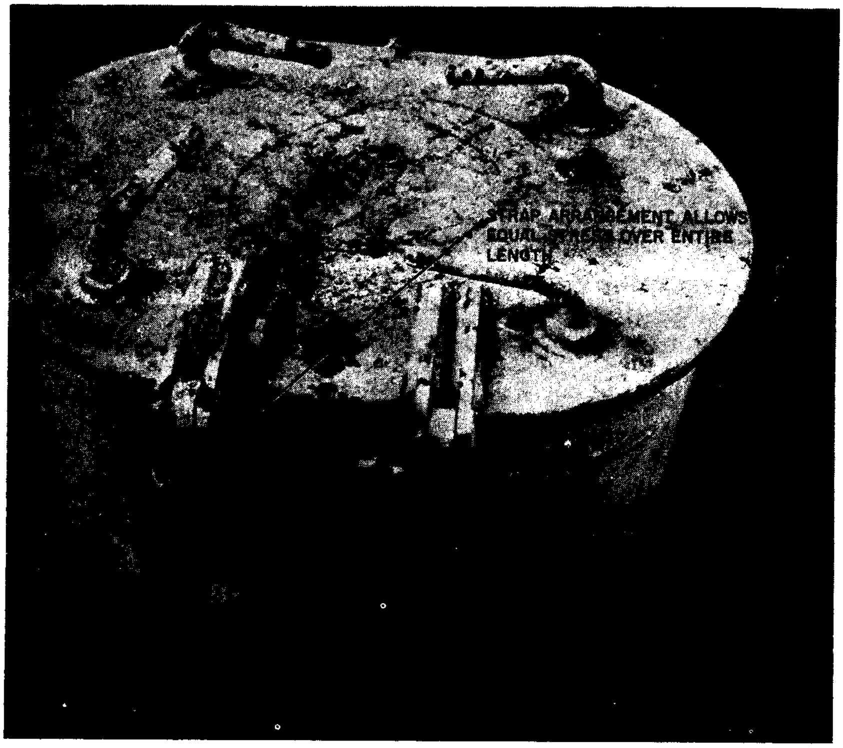
b. Bulkhead Mounted Locker Door
Figure 10-3. Typical Bonding Applications



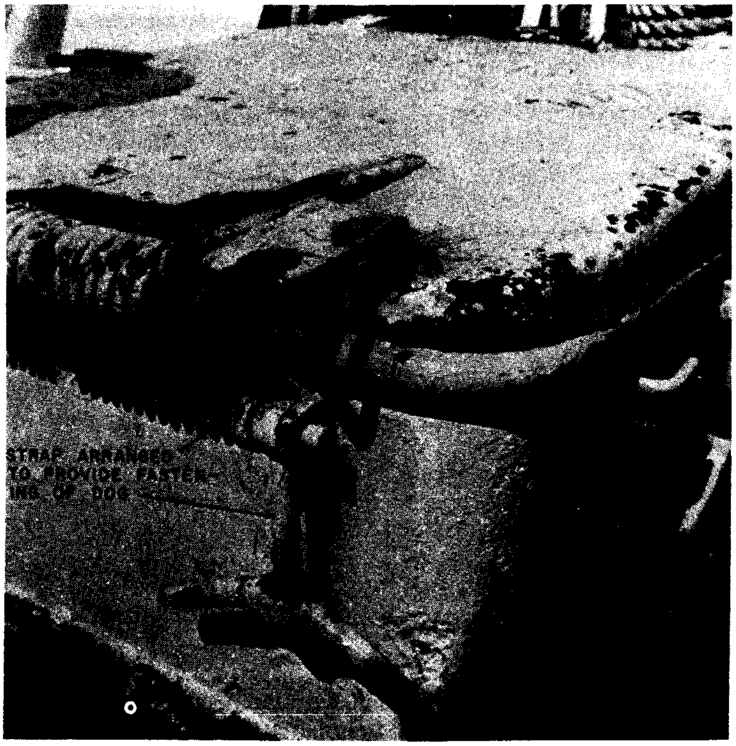
a. Safety Net Bonding
Figure 10-4. Typical Bonding Applications



b. Liferaft Holder — Dissimilar Metals
Figure 10-4. Typical Bonding Applications



a. Scuttle Cover
Figure 10-5. Typical Bonding Applications



b. Hatch Cover
Figure 10-5. Typical Bonding Applications

Keep the bond strap as short as physically possible in all applications. In the case of bonding across pivoted joints such as davits, doors, and the like, the strap must be long enough to allow normal unrestricted movement of the item being bonded. The strap should be tack welded or hand-held in place, when installing it across door hinges, scuttle hinges, etc., to check for freedom of movement and to ensure that the strap does not kink or flex unduly when the door is opened.

Use a welded-at-both-ends strap whenever possible. It is practical to use this type strap even when bonding across mast turnbuckles and the like even though there is a requirement to remove or adjust those items. The welded strap can be very quickly knocked loose and rewelded when adjustments are complete. The only time a one-end bolted type strap is acceptable is when the bonded item must be removed frequently. Such items would be davits (when stowed), portable servicing stands, and portable stanchions.

If a priority is necessary, due to lack of funds for a complete bonding program, the physically large items should be tackled first; especially if the item is mounted loosely. A loose, corroded junction between two physically large items near a transmit or receive antenna should receive the highest priority.

The bond strap should be installed so that it is not likely to have undue stress applied to the strap. Some ships have had bond straps broken due to improper installations.

10.1.6 Problems Associated With Bonding/Grounding

To accomplish interference suppression through an effective grounding system, all of the system electrical and structural components must be maintained at the same reference potential. This is generally accomplished by setting up separate grounding systems for the structural and the various electrical parts of the system and combining these separate grounds only at one common reference point or plane. The object is to prevent any electromagnetic interference generated by one unit of the system from being transferred through a common ground impedance to the other units. If potential differences are not allowed to exist, interference currents cannot flow and spurious signals can neither be radiated nor conducted to the susceptible parts of the system. Obviously, the larger the system, the less this ideal can be achieved. No conductor has zero impedance, hence zero potential difference can only be approached, never completely achieved.

Even in an ideal situation, such as the availability of a tremendously large mass of a highly conductive type of material, problems would still arise. These problems are caused by the necessity to make connections to this large ground mass. In order for these connections to be effective, they should not present any discontinuity to the electric current when it passes through the connection.

10.1.6.1 Development of RF Potentials. An equivalent circuit which illustrates how potentials develop is shown in Figure 10-6.

The media through which the source of rf potential is connected to the equipment or structure in question might be the impedance of a cable, the capacity between two objects, or the intrinsic impedance of free space.

In Figure 10-7(a) a source located in one equipment is driving a load in another. When the cable length is short, the resistance (R) of the cable shield is negligible compared with X_C, the capacitive reactance of the distributed capacity between box 2 and the ground plane. Since the potential (V) will then also be negligible (see equivalent circuit 10-7(b)), bonding of box 2 is not necessary. Longer lengths of cable, however, may raise V to a value such that a bond (Z_B) would be required.

Figure 10-8(a) illustrates a situation where rf energy is coupled from internal electronics or electrical equipment to equipment enclosures only through the stray capacities C_1 and C_2. If these capacities can be evaluated, Figure 10-8(b) can be used to determine the necessity and magnitude of a bond impedance Z_B. It is evident here also that bonding is required only when long cables are involved (large R_1 and R_2). Since there are two shunting paths, R_1 and R_2 longer lengths of shielded twin lead can be used, rather than a single lead coaxial cable, before bonding is required to reduce potentials.

If the source is external to the equipment or structure to be bonded and within a wavelength or two of it, the situation in Figure 10-9 exists. The magnitude of potential V before and after bonding with Z_B can be evaluated if the capacities C and C_1 can be determined.

10.1.6.2 Impedance of Jumpers. Bonding problems result when the bonding impedance becomes significant. They can most often be traced to a very large reactance of a jumper, a result which is comparatively independent of the jumper dc resistance. Direct bonding (welding or brazing) will always assure low impedance at any frequency, since the microscopic path lengths cannot contribute significant reactance.

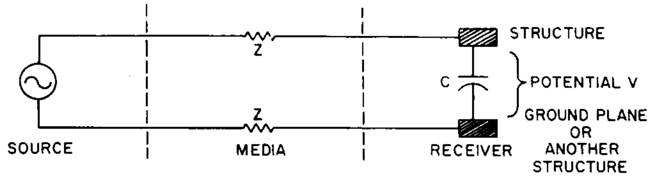


Figure 10-6. Generalized Equivalent Circuit Showing How Potentials Develop

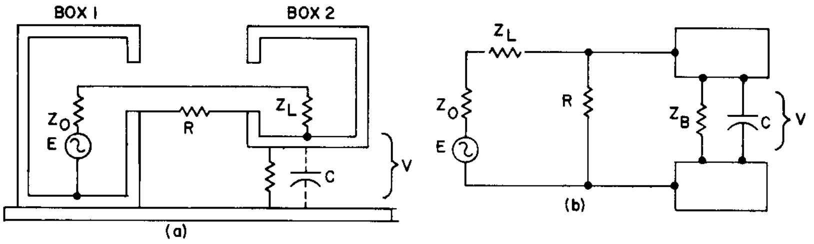


Figure 10-7. Internal Source - Single Wire

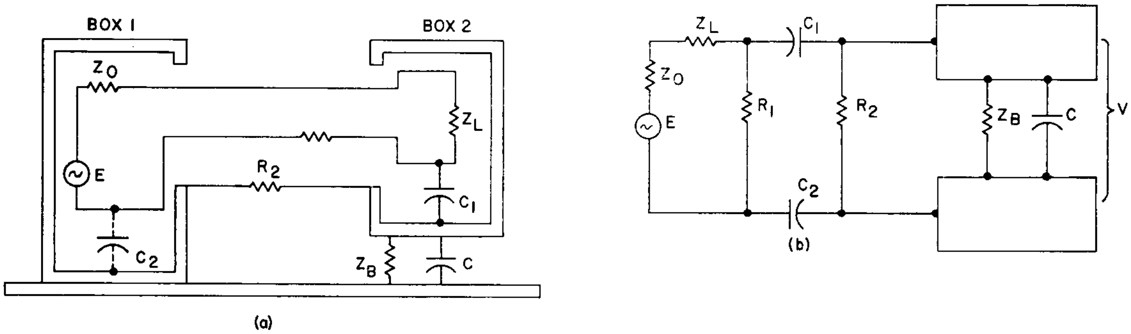


Figure 10-8. Internal Source - Double Wire

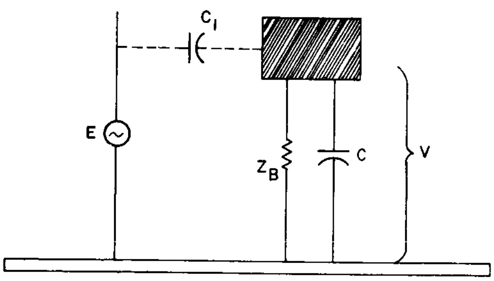


Figure 10-9. Near External Source

Figure 10-10 represents bonding of a shock mounted enclosure, equipment case, or chassis to a ground plane.

The impedance between chassis and ground, ignoring any electrical contribution by the shock mount, consists of stray capacity in parallel with the resistance and inductance of the jumper. The resistive component is in the milliohm range at power frequencies, and is negligible compared with X_L and X_C at radio frequencies. The variation of impedance with frequency for such a jumper will then assume a curve similar to that shown in Figure 10-11.

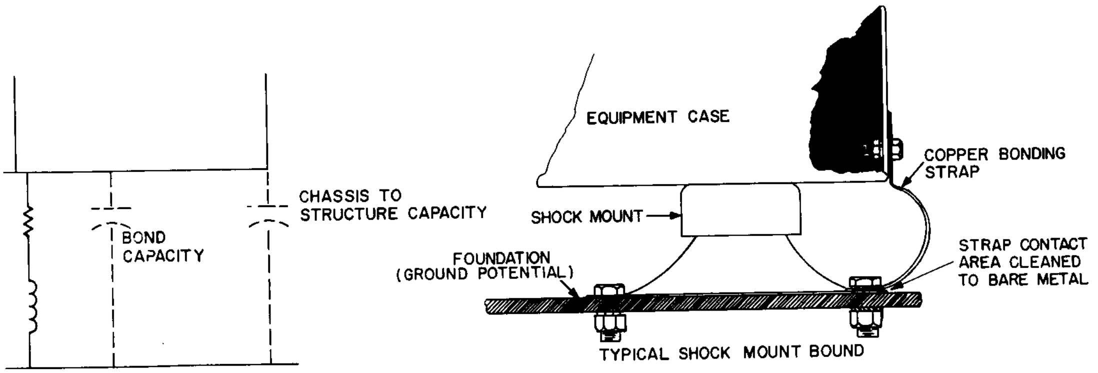


Figure 10-10. Bonding by Means of a Jumper

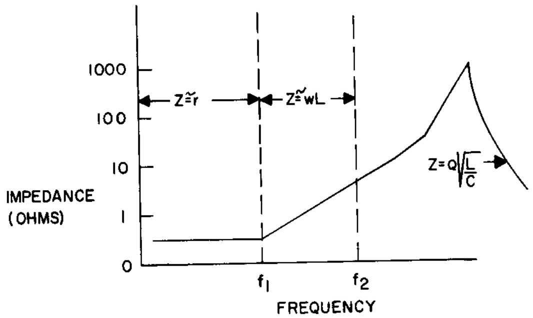


Figure 10-11. Variation of Jumper Impedance with Frequency

Three distinct frequency ranges can be discerned in Figure 10-11. From DC to f_1, the impedance is primarily resistive, being composed of jumper resistance and contact resistance. From f_1 to f_2, the inductive reactance of the jumper predominates since R is less than either X_L or X_C. Above f_2, parallel resonance with the stray capacity occurs. With the inductance being at most a one turn coil, and the capacity having an air dielectric, the resultant Q is very high (in the order of 200 or 300) which results in an impedance of many thousands of ohms. Figure 10-12 illustrates a possible condition where the attached wiring, shielding, and mechanical linkages result in composite shunting impedances. This combination of shunting impedances reduces the resonant "Q" and correspondingly lowers the impedance at resonance. This condition can be compared to a "loaded" parallel resonant circuit.

A bonding jumper has the usual electrical parameters of R, L, and C; of these parameters, R is an inherent property of the jumper (depending on the material selected), C is dependent upon the physical configuration and separation between the bonded members, and L is the inductance in microhenrys of the bonding jumper that is dependent on its physical dimensions. For a straight bonding strap, L is given by:

$$L = 0.00508a \left[2.303 \log\left(\frac{2a}{b+c}\right) + 0.5 + 0.2235 \left(\frac{b+c}{a}\right)\right]$$

where a = length of strap in inches

b = width

c = thickness (variations of thickness does not affect inductance once the skin depth is exceeded)

If the bonding strap is bent into a "U" shape, the strap takes on the characteristics of a parallel LC circuit. Not considering R (resistance of the strap), the equation for the impedance, Z, is given by:

$$Z = \frac{\omega L}{1 - \omega^2 LC}$$

where ω is the angular frequency

If $\omega^2 LC$ is less than one, the bonding strap appears principally as an inductance. As $\omega^2 LC$ approaches 1, the magnitude of the impedance increases and as $\omega^2 LC$ goes above 1 and continues to increase, the bonding strap appears to be principally capacitive. (Figure 10-13).

Consequently, to keep the impedance of the bonding strap low, the values of L and C must be such that the value of ω^2 LC remains as far as possible from 1, i.e., the resonant frequency of the bonding strap should be kept as far away as possible from the system frequencies. In theory, this resonant point of the bond could be located above or below the system frequencies encountered as indicated in Figure 10-14.

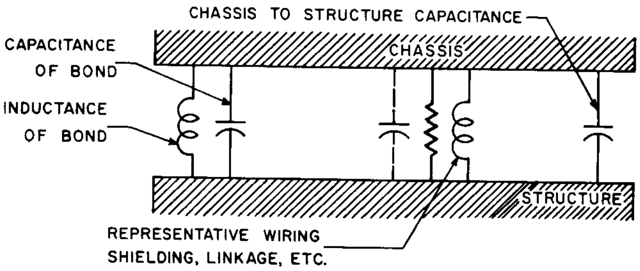


Figure 10-12. Electrical Configuration
(with Wiring, Shielding, Mechanical Linkages, Etc.)

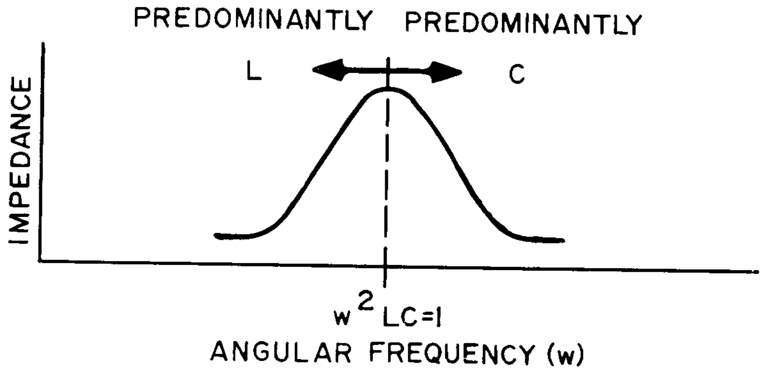


Figure 10-13. Bond Strap Impedance

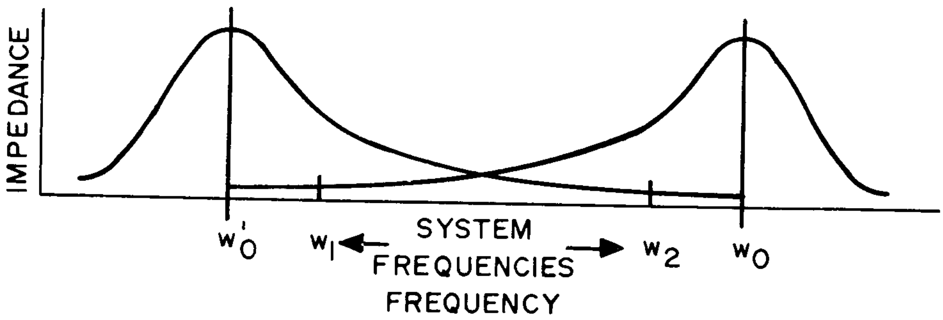


Figure 10-14. Bond Strap Resonant Frequencies

From the expression for resonance: $\omega_o^2 = \frac{1}{LC}$,

the LC product must be held low. As shown in Figure 10-14, the lower that L and C can be held the higher the resonant frequency will become; which, in turn, will lower the bond strap impedance. In the case of the resonant frequency below the system frequency range, the LC product must be large (L = 1 henry and C = 1 μf). Obviously, these large components are out of the question. Hence, for practical purposes, only the high resonant frequency case can be considered.

10.1.6.3 Skin Effect. Skin effect is a physical phenomenon which is somewhat complicated to describe but which has been measured and tabulated in numerous engineering handbooks for most practical conductor configurations. The effect is present at all frequencies but becomes more noticeable as frequency increases, and therefore must be considered in the design when higher frequencies are used. Briefly, the rapidly changing current of an ac wave produces a proportionately moving magnetic field. As this magnetic field moves across the conductor, it induces an electric field (back emf) in such a direction as to oppose the original current flow in the conductor. Since more of the magnetic flux lines of the moving field link the center of the conductor instead of the area near the surface, the opposing electric field (or inductance) will be the strongest in the center. At radio frequencies, the reactance of this extra inductance is sufficiently great to seriously affect the flow of current, most of which flows near the surface where the impedance is low.

The skin effect in a current-carrying conductor forces the current to the outer surfaces of the conductor. Figure 10-15 illustrates cross-sectional views of different shaped conductors.

For the resistance calculation, assume that all the current is flowing uniformly distributed in a cylindrical shell of wall thickness δ. The cross-sectional area of this shell is then δ times $2\pi a$, the perimeter of the conductor, where a is the radius of

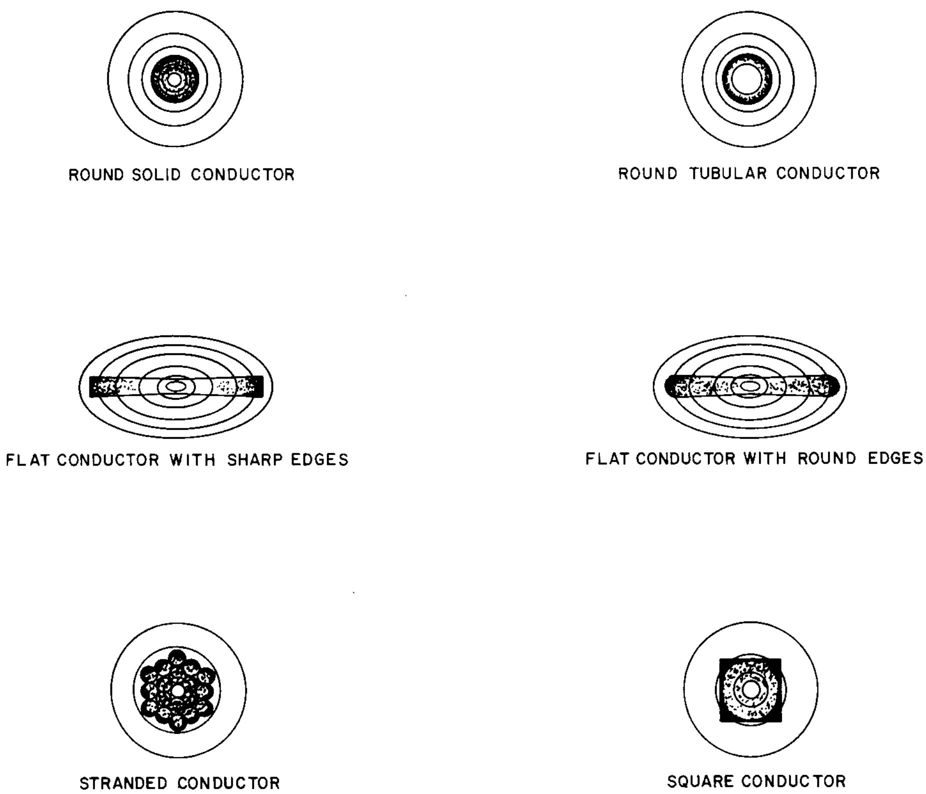


Figure 10-15. Cross Sectional View of Current Distribution in Various Conductors Due to Skin Effect

the conductor. The ac resistance of the conductor is then the dc resistance of the cylindrical shell:

$$R = \frac{\rho}{2\pi a} \text{ ohms per unit length}$$

Note that this calculation is valid for a conductor of any shape provided its thickness is at least 10 δ and it has no sharp corners.

When two or more wires are in proximity, the ac resistance is increased further by the proximity effect. Actually, this effect is but an extension of the skin effect phenomenon.

Figure 10-16 illustrates proximity effect in two conductors with current in the same direction and with current in opposing directions. As expected, when the current is in the same direction in the conductors, current density increases away from the other wire; but when the currents are flowing in opposite directions, the current density tends to concentrate at the near surface.

The ratio of effective ac resistance at a specific frequency to the dc resistance of a conductor is termed the resistance ratio. The resistance ratio increases with frequency, conductivity, and size of the conductor. This is because a higher frequency causes the extra inductance at the center of the conductor to have a higher reactance. Similarly, greater conductivity makes the reactance of the extra inductance important in determining the distribution of current, while a greater cross-section provides a larger central region. It is to be noted, however, that a larger conductor always has less radio frequency resistance than a smaller one because, although the ac-to-dc resistance ratio is less favorable, this is more than made up by the greater amount of conductor cross-section present.

10.2 SHIELDING THEORY

10.2.1 Purpose

Shielding has two main purposes: (1) to keep EMI confined within a specific region, and (2) to prevent EMI from entering a specific region. Ideally, only a transmitter would radiate energy and only a receiver would receive energy. This hardly ever happens; electronic equipment is constantly producing energy that may interfere with adjacent equipment. Shielding is a means of decoupling and thus, reduces interaction between equipments.

This does not mean that shielding should be thought of as the all purpose interference eliminator. On the contrary, the most efficient method of reduction would be one that uses the lowest number of strategically located components for a given interference source. This means that primary sources of interference must be recognized in the design stages. Isolation of each source must be effected by decoupling, by component orientation and location from an interference viewpoint. The circuit design should also be reevaluated to determine if the circuit parameters may be changed to alleviate the interference condition. Shielding is considered after the accomplishment of these isolation methods.

Only those components that actually require shielding should be included within the shield. Where physically possible, all supplementary components should be external to the shield. These include relays, motors, power transformers, chokes, filters, resistors, capacitors, metering devices, and wiring which is routed through the shield, but which does not terminate at the interference source. Shielding integrity can then be effected by confining the interference to this shield. It should be remembered that the amount of internal shielding required to merely permit a receiver to operate without oscillation or instability is not normally adequate to meet EMI requirements. The number of shields necessary to reduce interference energy transfer is minimized by exposing the minimum number of components to the interference source.

10.2.2 Shield Attenuation

Interference attenuation by a shield is due to two distinct effects: (1) Reflection of the interference

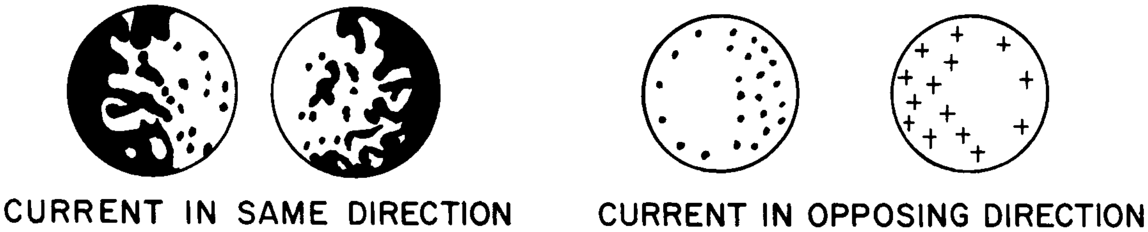


Figure 10-16. Proximity Effect of Current in Two Parallel Wires

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 "reflection loss" and the second is "penetration loss." The combined loss due to these two effects is sometimes called "attenuation of the shield" or "effectiveness of the shield."

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.

10.2.2.1 Reflection Loss. Reflection loss will be considered first since this is the first loss which the oncoming wave suffers. For simplicity, the interference wave considered will be a plane wave. This is a wave originating at a great distance from the shield so that in the vicinity of the shield the electric and magnetic components of the wave are equal or related by a constant coefficient. Also, the shield will be assumed to be an infinite flat plane. In the case of shielding an interference source, neither of these assumptions is correct, but this does not materially affect discussion of the problem.

Reflection of the wave at the air-metal boundary is precisely the analogue of reflection at an impedance discontinuity in a transmission line. The line impedance and lumped load impedance, in the latter case, are identically replaceable, respectively, by the impedance of the medium and the surface impedance of the metal shield in the former case. As in the transmission line, the amplitude of the reflected wave depends upon the degree of mismatch between the impedance of the air medium and the impedance of the metal sheet. The impedance of free space is given by the ratio of the electric to magnetic field strength in space and is 377 ohms. In order that the reflected wave be as large as possible, or that the reflection loss be great, the shielding sheet should have an impedance that is either very much greater than 377 ohms or very much less. Because a high or infinite-impedance sheet is impractical, the opposite extreme, a sheet of very low impedance, is used. The impedance of the metal sheet facing the oncoming wave is the ratio of the electric to magnetic field strength at the surface of the shield.

Table 10-1 gives the relative conductivity of metals. Copper is high in conductivity, second only to

Table 10-1. Conductivity of Metal

Metal	Relative Conductivity
Silver	1.05
Copper, annealed	1.00
Copper, hard drawn	0.97
Gold	0.70
Aluminum	0.61
Magnesium	0.38
Zinc	0.29
Brass	0.26
Cadmium	0.23
Nickel	0.20
Phosphor-bronze	0.18
Iron	0.17
Tin	0.15
Steel, SAW 1045	0.10
Beryllium	0.10
Lead	0.08
Hypernick	0.06
Monel	0.04
Mumetal	0.03
Permalloy	0.03
Stainless Steel	0.02

silver, which is the reason copper is used in screen room construction. Because of this high conductivity, copper has the highest reflection loss for an electric field, as shown in Table 10-2. A ferromagnetic substance, such as iron or steel, would give an inferior reflection loss compared to copper (see Table 10-3).

Reflection loss of the low impedance magnetic field is poor since the impedance discontinuity of the air-to-metal boundary is not as great for the magnetic field (see Table 10-4).

10.2.2.2 Penetration Loss. The intrinsic propagation constant for any medium describes the propagation of an electromagnetic disturbance through that medium. Penetration loss of a metal shield is essentially independent of wave impedance and it is the same for electric and magnetic fields. It depends primarily on shield material and frequency; being worse at low frequencies. High permeability material is desirable for high penetration loss. This is obvious from Table 10-5.

10.2.2.3 Shielding Effectiveness. The shielding effectiveness can be designed so that part of the burden is carried by the reflection loss and the rest by the absorption loss. Reflection loss in the electric field decreases inversely with frequency while reflection loss in the magnetic field increases inversely with

Table 10-2. Reflection Loss in Electric Field (Wave Impedance Much Greater Than 377 Ohms) of Solid Copper, Aluminum, and Iron Shields for Signal Source 12 Inches From Shield

Frequency	dB Loss		
	Copper	Aluminum	Iron
60 Hz	279	- - -	241
1000 Hz	242	- - -	204
10 kHz	212	- - -	174
150 kHz	177	175	- - -
1 MHz	152	150	116
15 MHz	117	115	83
100 MHz	92	90	64
1500 MHz	*	- - -	*
10,000 MHz	*	- - -	*

*At these frequencies, the fields approach plane waves with an impedance of 377 ohms; see Table 10-3 for plane waves.

Table 10-3. Reflection Loss in Plane Wave Field (Wave Impedance Equal to 377 Ohms) of Solid Copper and Iron Shields for Signal Source Greater Than 2 λ From the Shield

Frequency	(Loss in dB)	
	Copper	Iron
60 Hz	150	113
1,000 Hz	138	100
10 kHz	128	90
150 kHz	117	79
1 MHz	108	72
15 MHz	96	63
100 MHz	88	60
1,500 MHz	76	57
10,000 MHz	68	60

frequency. Penetration loss of iron at 10 MHz is approximately 70 dB greater than the magnetic field reflection loss, but at 100 kHz, it is a few dB less than the reflection loss. These factors should be fully considered by those who design metal shields. For example, if shielding against electric fields at frequencies of about 10 kHz is necessary, the tremendous reflection loss should be utilized rather than the penetration loss alone.

Table 10-4. Reflection Loss in Magnetic Field (Wave Impedance Much Smaller Than 377 OHms) of Solid Copper, Aluminum, and Iron Shields for Signal Source 12 Inches From the Shield

Frequency	dB Loss		
	Copper	Aluminum	Iron
60 Hz	22	- - -	-1
1,000 Hz	34	- - -	10
10 kHz	44	- - -	8
150 kHz	56	54	19
1 MHz	64	62	28
15 MHz	76	74	42
100 MHz	84	82	56
1,500 MHz	*	- - -	*
10,000 MHz	*	- - -	*

*At these frequencies, the fields approach 377 ohms in impedance and become plane waves; see Table 10-3 for plane waves.

To summarize, shielding effectiveness is the result of reflection losses and penetration losses together, not any single part. At frequencies as low as 60 Hertz, penetration loss and reflection loss become negligible for magnetic fields so that, for lower frequencies, very thick metallic barriers may be necessary to shield against magnetic fields. For example, a metallic barrier of iron with a thickness of 300 mils is required to obtain a shielding effectiveness of 100 dB at 60 Hertz for magnetic fields. For copper and iron, reflection and penetration losses are small for magnetic fields at low frequencies. Since magnetic materials, such as Mumetal, have high permeability at low frequencies, and therefore high penetration loss, they are more effective as shields for low frequencies. The resulting increase in penetration loss is obtained at the expense of a reduction in reflection loss. The following are general rules for selection of shielding materials:

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

b. Magnetic materials such as iron and Mumetal should be used for low-frequency 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. To provide a given degree of shielding, reference to the curves of penetration loss permit quick estimates of the required metal and thickness. In most applications, it is necessary to reduce an interference field to some specified level. This shielding requirement can be determined by measuring the interference fields without the shield and comparing them with the specification limits.

When the penetration loss is less than 10 dB the shield is said to be electrically thin. In this case, a correction factor (B) which accounts for the effect of successive re-reflections must be added. The total shielding effectiveness (S) becomes:

$$S = A + R + B$$

where

A = penetration or absorption loss in dB of the shield.

R = total reflection loss in dB from both surfaces of the shield.

B = positive or negative correction factor when A is less than 10.

In most practical cases, the correction factor can be ignored. However, at extremely low frequencies when shielding is required against magnetic fields, it must be considered.

Table 10-5. Penetration Loss of Metals

Metal	Penetration Loss (at 150 kHz, dB/mil)
Silver	1.32
Copper, annealed	1.29
Copper, hard drawn	1.26
Gold	1.08
Aluminum	1.01
Magnesium	0.79
Zinc	0.70
Brass	0.66
Cadmium	0.62
Nickel	0.58
Phosphor-bronze	0.55
Iron	16.9
Tin	0.50
Steel, SAW 1045	12.9
Beryllium	0.41
Lead	0.36
Hypernick	88.5*
Monel	0.26
Mumetal	63.2*
Permalloy	63.2*
Stainless Steel	5.7

*Obtainable only if the incident field does not saturate the metal.

10.2.3 Multiple Shielding

The shielding requirements necessary to protect against magnetic field transference are more difficult to achieve at low frequencies than at frequencies above several megahertz. Low-frequency shields must have either high conductivity or high permeability. In the normal power-frequency range, for example, copper must be very thick to serve as a practical magnetic shield. Although Mumetal and similar type high-permeability alloys provide good shielding for low-frequency weak fields, multiple magnetic shielding is recommended for low-frequency strong fields. Good shielding effectiveness for electrical fields, obtained with shields of high conductivity such as copper or aluminum, is infinite at zero frequency and decreases with increasing frequency. On the other hand, magnetic (low-impedance) fields are difficult to shield at low frequencies because reflection losses may approach zero for certain combinations of material and frequency. Reflection and penetration losses decrease with decreasing frequency for non-magnetic materials. At high frequencies, the shielding effectiveness is good because of reflections from the surface and rapid dissipation of the field by penetration losses. At low frequencies, it is also possible to obtain good reflection losses in magnetic materials that may be used to provide a safety factor in design. 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. Copper, Mumetal, iron, Co-netic and Netic type materials, and other metals, some with excellent electric field reflection loss and some with excellent magnetic field penetration loss properties, can be used effectively in combination. The results are composite antimagnetic and anti-electric field shields of usable physical proportions. In many applications, it is possible to reduce shielding effectiveness requirements for the overall equipment housing by employing suppression techniques within the equipment. The recommended techniques make use of component shields, filters at the source of the undesired signal interference, partial shields, circuit isolation by decoupling, short leads, and use of the ground plane as the ground return lead.

10.2.4 Shielding Effectiveness for Perforated Metal Shields

An ideal shielded enclosure would be one of seamless construction with no openings or discontinuities. However, personnel, power lines, control cables and ventilation ducts must have access to any practical enclosure. The particular design that will emerge to accommodate these varying configurations is a critical factor. Every design must incorporate the discontinuities in the shield without substantially diminishing the overall shielding effectiveness. Typical shield enclosure discontinuities are illustrated in Figure 10-17.

Usual methods for coping with these difficulties are:

a. Metal Seams – No Gasket – It is necessary to have clean metal-to-metal mating surfaces. Good pressure contact between the metal surfaces should be assured by using set screws or rivets. If the metal seams become corroded or anodized, a serious interference source will develop.

b. Metal Seams – Metallic Gasket – Metal seams can be improved by using metallic gaskets placed between the surfaces. These gaskets, often made of knitted wire mesh, are resilient and are good conductors. The metal-to-metal contacts must still be clean, maintained by a pressure of approximately 20 psi. Another type of gasket that is also effective is a 10-mil-thick strip of beryllium copper with jagged points on both surfaces. It requires very little pressure to give metal-to-metal contact and it is even effective on metal-to-metal mating surfaces that are not clean.

c. Holes and Screening – Any holes or openings in shielding enclosures cause a decrease in shielding effectiveness, but they usually cannot be avoided. Large holes, particularly in higher-frequency applications, are bad offenders. Therefore, holes should be as small as possible, but if large holes are present, they must be covered with copper mesh screening.

d. Waveguides – In handling large holes that cannot be screened over, the use of waveguide attenuators is recommended. Waveguides are tubes of conducting material that will pass high-frequency energy with little attenuation above some cutoff frequency and will greatly attenuate energy below this cutoff frequency. This latter property makes them very useful in shield designs.

e. Control Shaft (Grounded) – Openings must be provided in a shielded enclosure for control shafts to be available outside the case for operating or maintenance purposes. If the control shaft is of non-conducting material, it can be set inside of a waveguide attenuator.

f. Fuse Receptacles, Phone and Meter Jacks – A fuse receptacle must be covered with a metallic cap that makes the shielded enclosure electrically continuous at that opening. The same solution can be used for phone jacks, meter jacks, and pilot lamps.

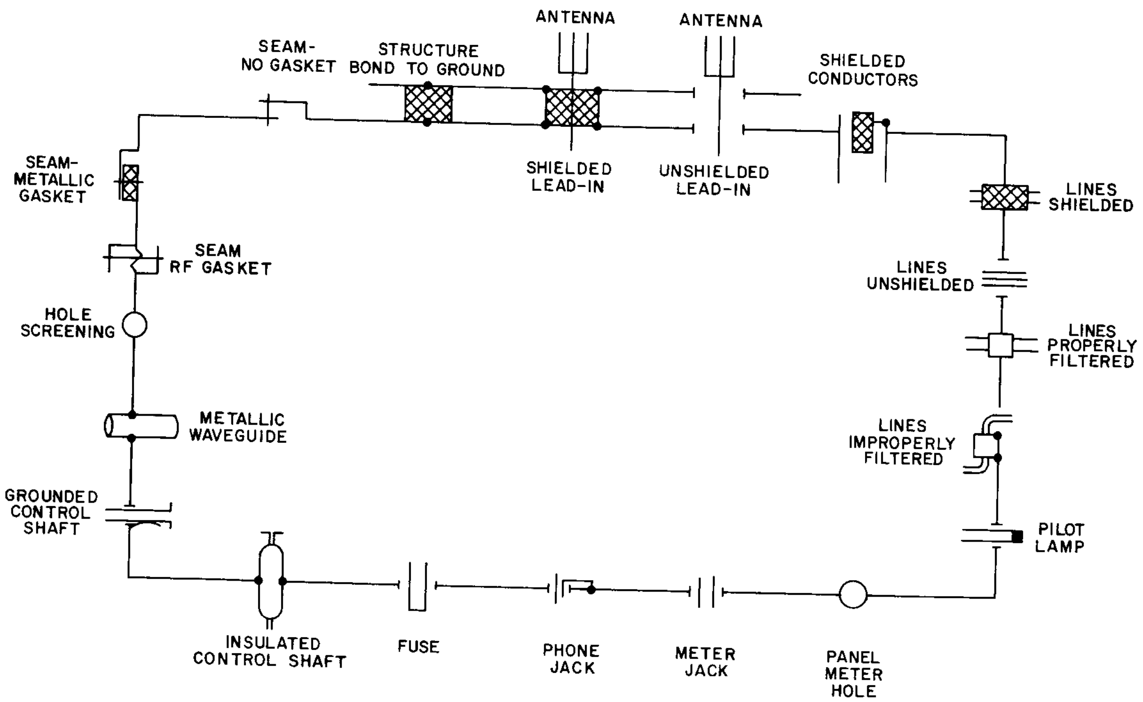


Figure 10-17. Typical Shielded Enclosure Discontinuities

10.2.4.1 Screening. Reflection loss of a screen, which accounts for the major loss, is less than that of a solid metal sheet. Even so, wire screen with 50 percent open area and 60 or more strands per wave length has a reflection loss very close to that of a solid sheet of the same material.

At low frequencies, when the skin depth becomes comparable to the radius of the wire, a considerable loss occurs in shielding effectiveness.

Screen construction should have individual strands joined permanently at points of intersection by a fusing process so that permanent electrical contact is made and oxidation does not reduce shielding against electric (high impedance) fields at low frequencies, because the losses will be primarily caused by reflection.

Louvered openings are good for ventilation purposes but are extremely poor for rf integrity because of their long narrow gaps. Screening is sometimes used on louvered panels to improve rf integrity.

10.2.4.2 Waveguide-Below-Cutoff Devices. Another method of preventing rf leakage through a ventilation duct or other necessary opening is the use of devices acting as a waveguide-below-cutoff. Over 130 dB of attenuation is available by this method. The waveguide attenuator is also of considerable value when control shafts must pass through an enclosure. When an insulated control shaft passes through a waveguide attenuator, the control function can be accomplished with almost no interference leakage.

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 easily as can the mesh screen, and are therefore, more reliable.

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

All nonsolid 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. Honeycomb, however, has the disadvantages of occupying greater volume and costing more than screening or perforated metal. Also, it is often difficult to apply honeycomb paneling because flush mounting is required. Thus, screening and perforated sheet stock sometimes find application for purely physical design reasons, although honeycomb panels can achieve attenuations to 136 dB, above 10 MHz.

Screened openings usually must be large to permit sufficient air to flow. When frequencies above 1000 MHz are to be attenuated to a high degree, ventilation openings must be designed as waveguide attenuators operating below cutoff at their lowest propagating frequency. In this manner, shielding effectiveness of over 100 dB can be obtained at frequencies of 10,000 MHz. A 1/4 inch diameter tube, 1 inch in length, would have 102 dB of shielding effectiveness at 10,000 MHz; a 1/2-inch diameter tube, 2-1/4 inches long, would have 100 dB of shielding effectiveness at 10,000 MHz. Openings of 1 inch or more in diameter would have little or no attenuation at 10,000 MHz. To obtain an opening of sufficient size to admit the required volume of ventilating air, tubes should be placed side by side until sufficient air flow is achieved.

10.2.5 Special Shielding Techniques

10.2.5.1 Carbon Arc Lamps. Carbon arc lamps, widely used in projectors and reproduction devices, produce a continuous electrical arc between two carbon electrodes. They are very severe interference generators. Because carbon arc lamps are usually used in small enclosures, and light is beamed in a single direction, partial shielding is feasible and effective. In addition, feed-through or bypass capacitors may be installed in power leads to prevent conduction of interference out of the enclosure.

10.2.5.2 Gas Discharge Lamps. Gas discharge lamps, such as fluorescent lights, ultraviolet, and neon lamps are intrinsic sources of interference, control of which is largely a matter of isolation and containment. This interference arises from the rapid ionization that takes place with each cycle of the line frequency that generates impulsive bursts of wide-band interference. The ionic discharge often generates a single frequency, usually in the low megahertz range, which appears at the terminals of the lamp and is radiated from the glass bulb and interconnecting leads. Fluorescent lamps contain mercury vapor at low pressure which is ionized by the flow of electrons in the tube. The subsequent deionization causes ultraviolet radiation which excites the phosphor coating on the inside of the tube, causing it to emit light in the visible region. Since this process employs a continuous arc, it gives rise to radio interference.

There are three ways in which these lamps can transmit interference to receivers: (1) by radiation directly from the lamp, (2) by radiation from power leads, or (3) by transmittal by conduction through a common power supply system. Suppression of direct lamp radiation is a function of lamp shielding. This approach is not often very effective, however, because shielding can interfere with the normal lighting function. Elimination of most power-line conducted or radiated fluorescent lamp interference is accomplished with the aid of feedthrough or bypass capacitors. For systems that employ starters, such capacitors may be placed across starter terminals. For starterless systems, the capacitors are mounted in the ballast. Most lighting fixtures have these capacitors built in, but this may not be the case when manual starting is employed. For manual starting a capacitor with a value between 0.006 μf to 0.01 μf should be installed across lamp leads. In cases where it is desirable to reduce interference further in power leads, filters should be used.

Lamps of this type cannot be entirely shielded with sheet metal without eliminating all of the useful light output. However, screen wire can be used, but reduction in light transmission by suitable screening may be as great as 50 percent. Progress has been made in the development of coated glass which transmits over 90 percent of the visible light while providing adequate interference shielding. The function of the coating on the glass is to intercept and ground out radiated interference. Typical construction consists of a heat-resistant, borosilicate glass panel having a permanently bonded transparent, electrically-conducting film applied to its smooth side. A 0.25-inch-wide metal grounding strip is fired onto the film around the periphery of the glass panel. A conductive silver paint is applied to make good contact between the glass panel and the frame, and the frame is bonded to the metal fixture ground plane. The glass coatings usually exhibit resistances ranging from 50 to 200 ohms per square inch and substantially reduce radiated interference in the frequency range from 0.014 to 25 MHz.

The standard slimline fluorescent lighting fixtures represent a potential source of conducted and radiated interference. Tests performed on an 8-foot slimline fixture indicated conducted interference of 700 microvolts per kHz bandwidth at 0.5 megahertz, and 18 microvolts per kHz at 10 MHz. Radiated interference exceeded the limits of MIL-I-6181B up through 1 MHz. As an example, 700 microvolts per kHz broadband interference energy in a circuit having a bandwidth of 20 kHz could produce the equivalent of a 14,000 microvolt signal. Further, a large number of such fixtures could generate even higher interference voltage amplitudes, depending on inherent phase relationships. To deal effectively with these levels of broadband interference, waveguide type filter shields of honeycomb material can be successfully employed. Low rf impedance bonding of the shield assembly to the fixture frame and, in turn, to ground through conduits, etc., becomes of utmost importance. The use of capacitor networks (bypass or feedthroughs) or power line filters (and conduits) are recommended for attenuation of conducted interference phenomena.

10.2.6 Magnetic Shielding Techniques

10.2.6.1 General. Well engineered circuits can be seriously affected by interference from spurious magnetic fields set up by neighboring components. These effects can be minimized by judicious orientation of components on the chassis. There are many instances, however, where this cannot be accomplished because of space limitations. In addition, many low-level applications can be upset by the earth's magnetic field. This problem can be critical in amplifier circuits where the extraneous signal may be amplified thousands of times. For example, the field generated by the driving motor of a tape recorder is of sufficient magnitude to dominate the signal completely unless the record and playback heads are well shielded. The electron beam of a cathode-ray tube can be bent by the field of a transformer used in the circuit. A magnetic compass is of such strength as to affect a radar scope six feet away. Because of such effects, input transformers handling very low-level signals must be protected from any spurious fields. In general, the solution of most of these problems lies in the use of a magnetic shield around the affected component.

A magnetic shield is actually a low-reluctance path in which the magnetic field is contained. For this reason, shields are generally made of high permeability nickel-iron alloy such as Mumetal. There are many low-level applications where a single shield will not reduce the field to a low enough value. In such cases, a nest of shields is necessary. Additional shields are placed over the first one to achieve the required degree of isolation.

10.2.6.2 Cable Shielding. Both high and low-impedance waves exist in typical shipboard and non-shipboard environments. For the purpose of shielding, high and low wave impedance is referred to the impedance of free space, whereas wave impedance greater than 377 ohms is termed high impedance and wave impednace less than 377 ohms is termed low impedance. The impedance of any wave within a good conductor at ELF and VLF is much less than 377 ohms.

The amount of electromagnetic shielding provided by any metallic material is a direct result of the reflection and penetration losses. The amount of reflection loss in db is directly proportional to the impedance mismatch between an electromagnetic wave and the shielding material. When the impedance of the wave approximates the impedance of the shielding material, little or no reflection loss occurs and most of the wave energy penetrates the shield. When the wave impedance is higher than the shield impedance most of the wave energy is reflected by the shielding and very little penetration is realized. In either case the wave energy that penetrates the shielding suffers a penetration loss which is related to the type and thickness of the shielding material. Most shielding losses occur by reflection which indicates wave impedances higher than that of the shielding material. Reflection losses are independent of shielding thickness; therefore, a thin layer of metal of high conductivity is quite effective as an electromagnetic shield. This reflection loss cannot be relied upon for low impedance electromagnetic waves and the only alternative in designing a shield for low impedance waves is to provide for the maximum penetration loss.

$$A = 3.34(10^{-3}t\sqrt{f\mu_r\sigma_r})$$

Where

t = shield thickness in mils,
f = frequency in Hz,
μ_r = permeability relative to μ_0, and
σ_r = conductivity relative to copper.

The relative permeability for existing "magnetic shielding" materials ranges from 10^2 to 10^6, while the range of σ_r for these same materials varies from 10^2 to 10^{-1}. Hence, the absorption function is most readily maximized at a given frequency by maximizing permeability; i.e., in ELF and VLF ranges the best shielding material is obtained by using a nickel-iron alloy (roughly 80 percent nickel and 20 percent iron), such as Mumetal, Hypernom, and Co-Netic. The permeability of these materials is maximized by annealing. Generally, the annealing process consists of heating the material in a dry hydrogen atmosphere to around 2100° C, holding it at this temperature for several hours, and then cooling it at a slow, controlled rate. Cold working reduces the permeability; therefore, the material should not be annealed until after it has been formed into its final shape. Because of furnace size limitations, it is not always practicable to anneal long lengths of stock; therefore, a compromise in the anneal process is not unusual. At least one supplier passes his conduit through a hot flame (at a controlled rate) primarily for the purpose of brazing the copper (strip) packing to the "magnetic" material. Obviously, the conduit is not annealed properly in this case.

Since the majority of conduits made from magnetic materials are strip wound and since the exact permeability is not usually known, analytical prediction of the shielding effectiveness of these conduits is prohibitive. Therefore, a more practical way to obtain these data is to measure the shielding effectiveness in a configuration that most closely approximates the intended application. Most often, the philosophy of shielding has been to apply shielding to low-level signal cables. The test fixture shown in Figures 10-18 and 10-19 was designed to simulate most shipboard cable shielding installations. A McIntosh power amplifier drives the transmitting loop, which impresses equal fields at the reference and shielded loop probes. The induced field at the shielded probe adds vectorially with the impressed field at the probe, and, therefore, the insertion loss (shielding effectiveness) is the ratio of the impressed field at the reference probe to the total field at the shielded probe. The impressed field level is 10 dB/ /1 gauss.

Figure 10-20 is a nomograph designed as an aid for estimating the amount of shielding required for a shipboard installation. Notice that, for a given cable and system sensitivity level, the expected level of induced noise voltage and, hence, the shielding requirements decrease with increasing frequency. Thus, the ideal shielding characteristic in the ELF and VLF range is a high value at the lowest frequencies and monotonically decreases from this high value as the frequency increases.

INSTRUCTIONS FOR USE OF NOMOGRAPH

1. This nomograph is based on the following: (a) the low-frequency flux density vs. frequency limit line set forth in Mil-Std-461, (b) Faraday's Law: $V = \partial \psi / \partial t$, and (c) the readily obtained equivalent areas of twisted and coaxial cables.

2. The expected induced voltage on any coaxial or twisted signal cable is easily calculated by connecting the desired frequency and cable (or equivalent loop area if the cable in question is not listed) with a straight line and noting the intersect on the induced voltage axis. This intersect on the induced

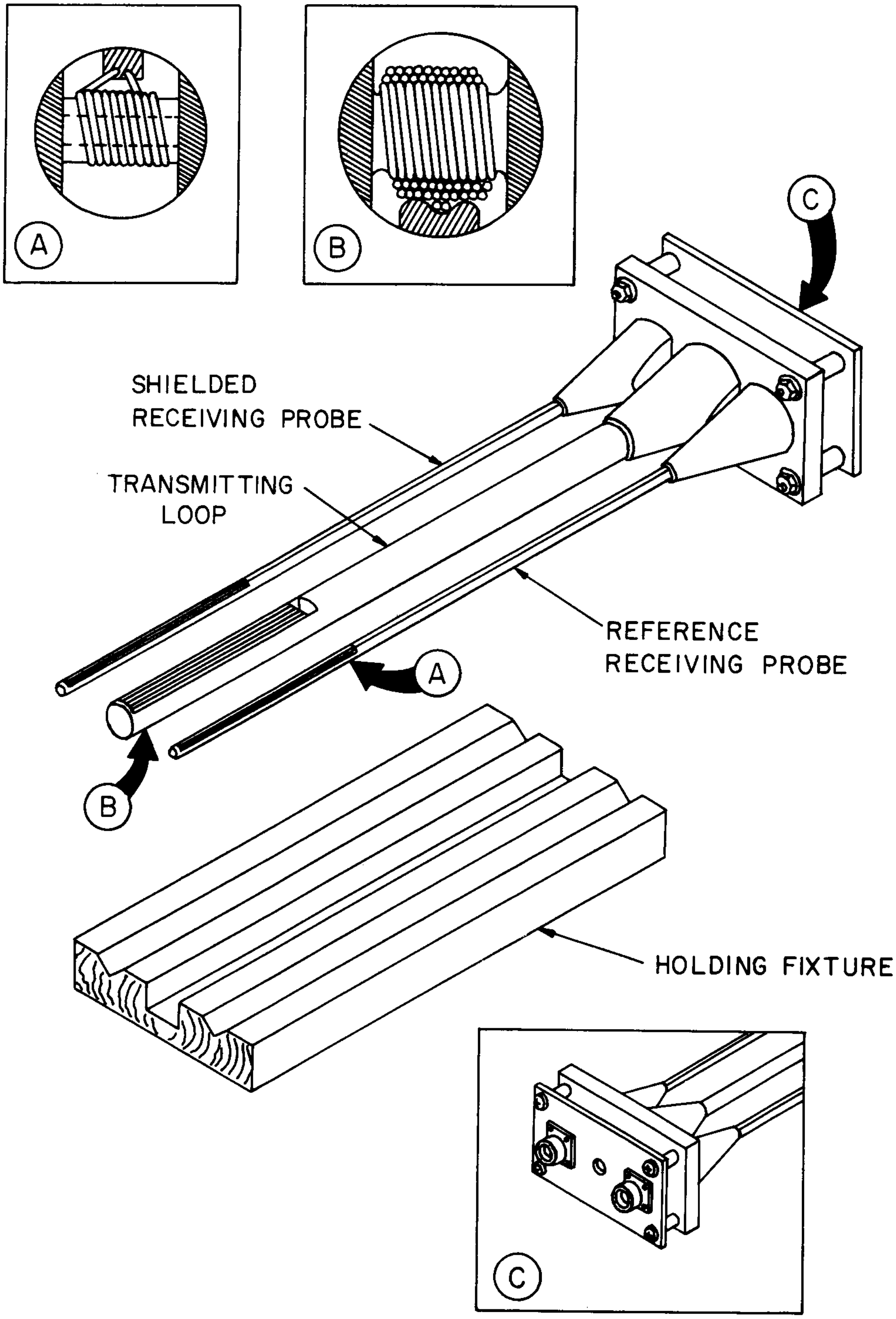


Figure 10-18. EM Shielding Test Fixture for Conduits

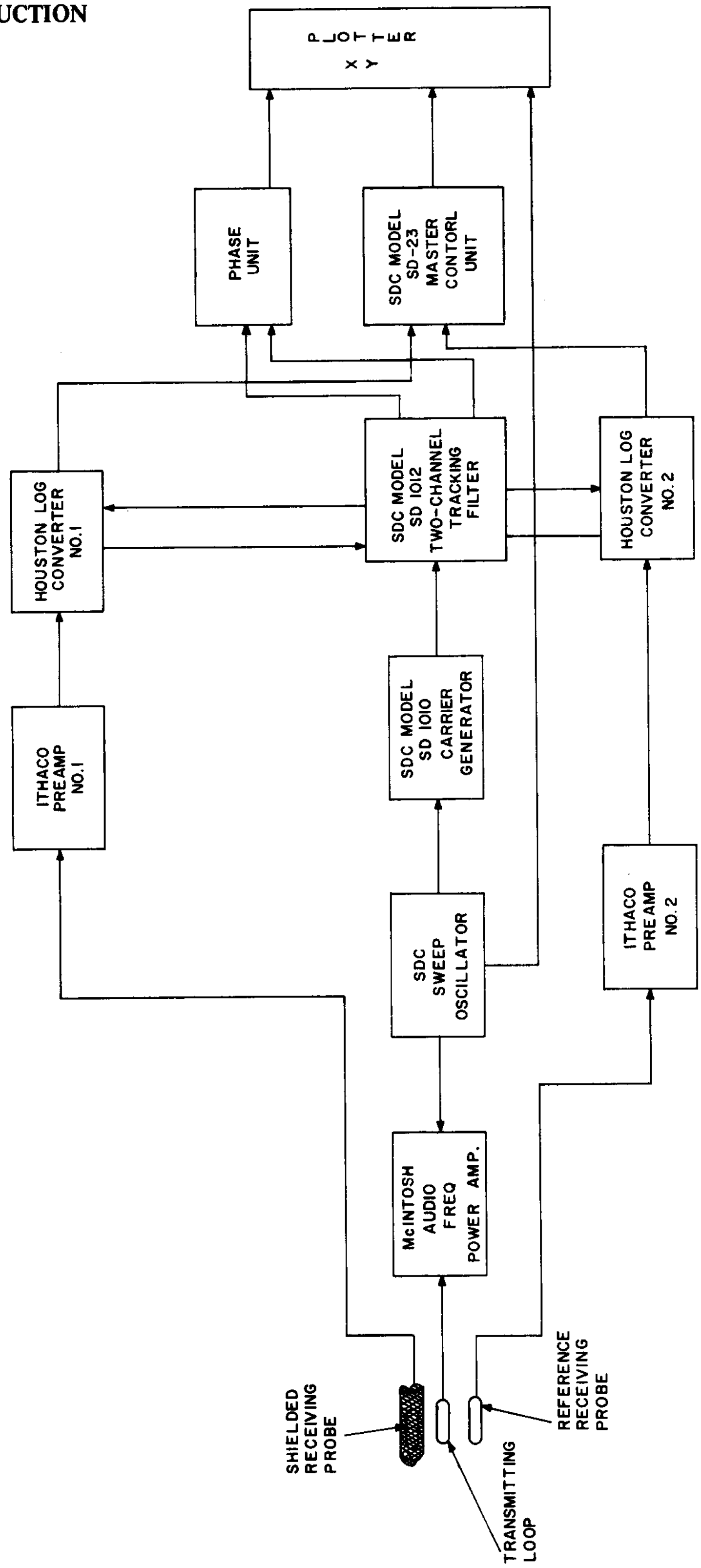


Figure 10-19. Automated System for Obtaining Shielding Effectiveness (Insertion Loss) Versus Frequency Data

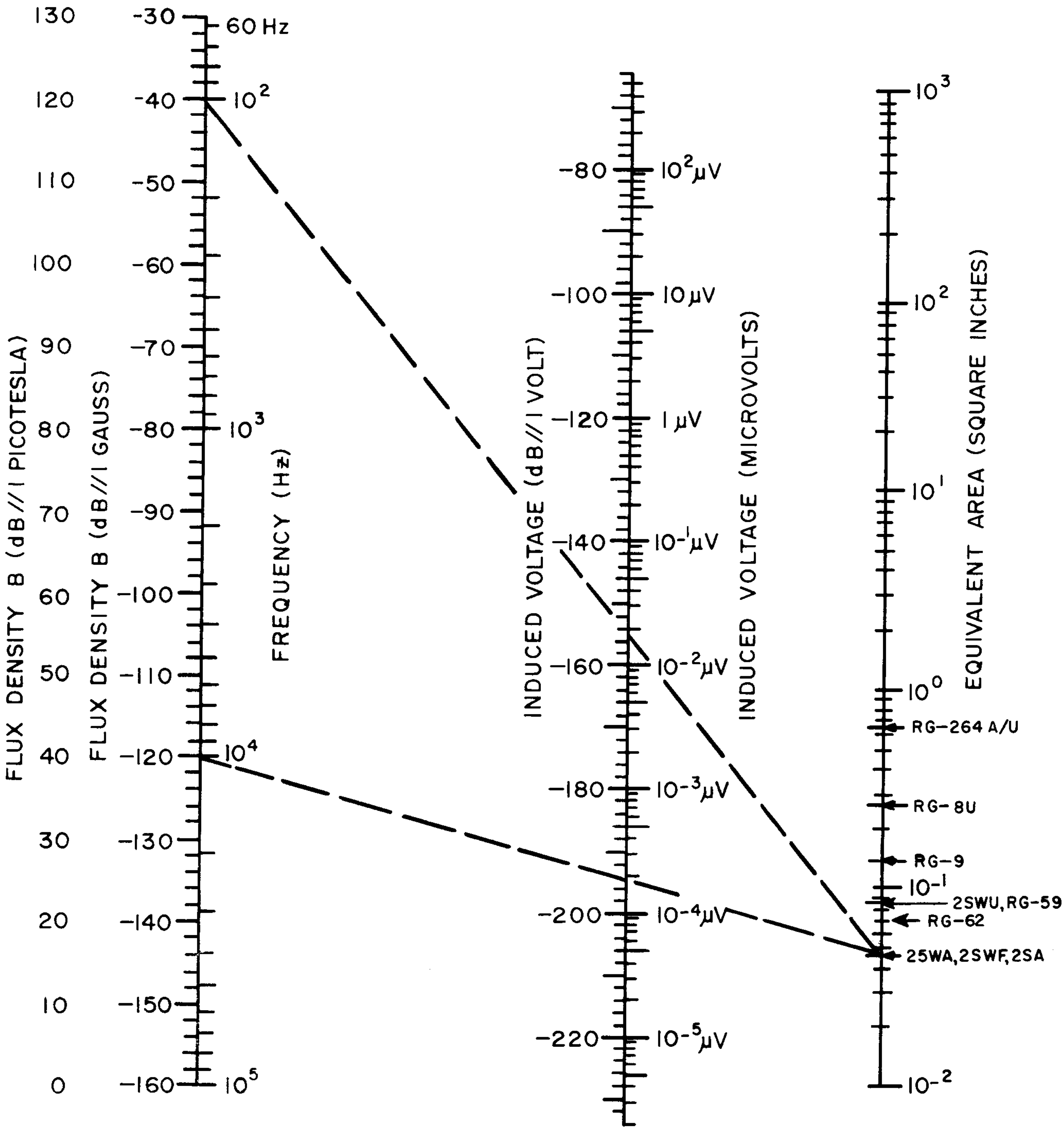


Figure 10-20. Nomograph for Estimating the Amount of Shielding Required for Shipboard Conduit Installation

voltage in (a) μ volts or (b) dB/ /1 volt. Since Mil-Std-461 is an upper decile presentation of observed shipboard fields, the following quantities are required to insure safety when computing submarine shielding requirements:

	400 Hz and below	400-4000 Hz	4 kHz - 100 kHz
Add to induced voltage level	40 dB	30 dB	20 dB

EXAMPLE

Given a sonar cable whose signal and return are in the same magnetically unshielded 2SWA cable. The passband of the sonar set is 10^2 to 10^4 Hz. If the noise voltage must be kept below -140 dB/ /1V Hz within the passband to effect electromagnetic compatibility, what is the minimum amount of shielding required?

Straight lines are drawn from the 2SWA point (4.5×10^{-2} in.^{2}) on the equivalent area axis to the frequency extremes of interest. One sees that the expected induced voltage at 100 Hz is - 155 dB/ /1V and at 10 kHz the expected voltage level is -195 dB/ /1V. Adding the recommended correction quantities, we obtain a worse-case voltage of -115 dB//1V at 100 Hz and -175 dB//1V at 10 kHz. Clearly, 25 dB of magnetic field shielding is required to ensure electromagnetic compatibility at 100 Hz. Conversely, there is a 35-dB safety margin at 10 kHz, and no shielding would therefore be required at this frequency. Because of the linear nature of the Faraday relationship, $/e_{ind}/ = .406fBA$, shielding requirements between the two frequency extremes would be intermediate to these calculated. Therefore the minimum amount of shielding effectiveness required is 25 dB.

Note 1: The cables along the equivalent area axis are correctly placed along that axis only if the signal and its return are in the same cable.

Note 2: Careful routing of signal cables a minimum of six inches away from interference sources would eliminate the need for the safety margins. In the above example, no shielding of the 2SWA cable would be required anywhere throughout the band of interest if isolation of signal and power cables were strictly maintained at six inches.

Note 3: The limit line from Mil-Std-461 is based on submarine shipboard measurements of ambient magnetic fields. A spectrum plot of these field levels versus frequency invariably yields "spikes" at the power line frequencies (60 and 400 Hz) and their harmonics. The limit line represents the locus of the tonal levels (upper decile) measured.

The shielding characteristics versus frequency data for this one-inch conduit are presented in Figure 10-21. The curious "resonance" at 2.7 kHz and the negative shielding slope above this frequency were quite unexpected. Several attempts were made to establish that the data above 2.7 kHz were contaminated by end effects. The conduit was moved two feet off center, and the end caps were welded. Neither act caused any change in the shielding function. (A 3/8-inch-diameter hole was drilled in one end cap to accept the probe.) Since the anomaly cannot be attributed to geometrical factors, the conclusion must be that the $\mu\ \sigma$ product decreases rapidly above 2.7 kHz. Further study of this conduit should not be discouraged.

Figures 10-22, 10-23 and 10-24 show the shielding functions obtained for 1-, 3/4-, and 1/2-inch conduits. These conduits do not provide much shielding effectiveness in the sonar frequency range; the characteristics are far from the ideal shielding requirements outlined in the previous section. Data obtained from annealed material are included in Figures 10-22 and 10-24. The anneal improved the shielding effectiveness by less than 10 dB.

Figure 10-25 presents the shielding characteristics of one-inch material. This conduit is comprised of 4-mil-thick strips of nickel-iron alloy. After the conduit is formed, the wall thickness is effectively tripled. As shown in Figure 10-25, there is little difference between the shielding characteristics of the Mumetal and bronze wire braids. The shielded probe was centered within the conduit.

A special conduit was made from 8-mil-thick nickel-iron stock (the effective wall thickness was 24 mils). Figure 10-26 shows the improved shielding effectiveness versus frequency characteristics for this conduit.

Figure 10-27 presents the shielding effectiveness versus frequency data for 3/8-inch hard and soft drawn copper conduits. The shielding is inadequate in the sonar frequency range and is significant only above 15 kHz.

Conduits of high initial permeability are best suited for shipboard low-impedance, low-frequency EM shielding. These light-weight nickel-iron conduits will ultimately replace the less flexible, present-day steel conduits, although the transition will be delayed by mechanical factors, such as bonding techniques, end fittings, and reliability.

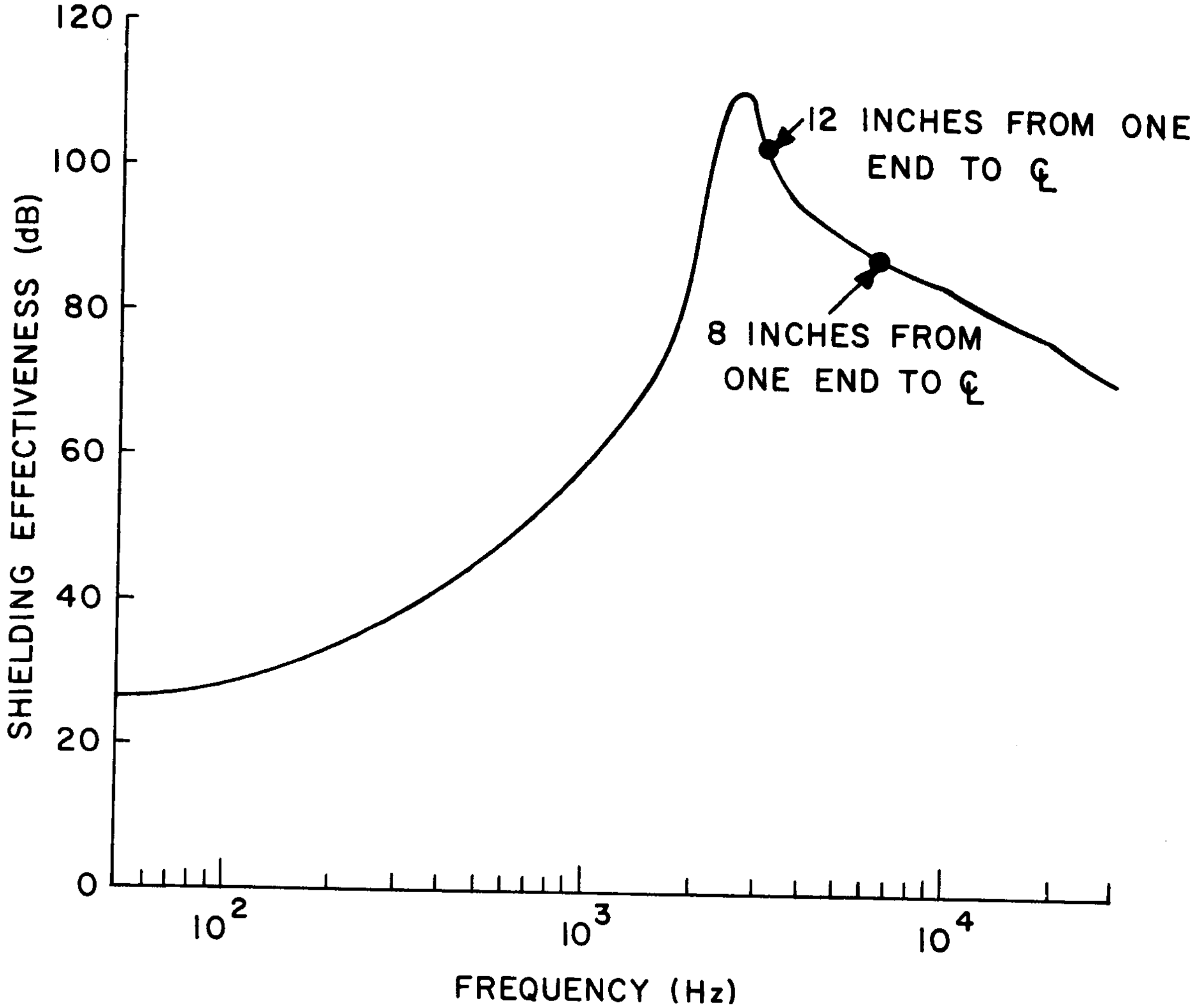


Figure 10-21. Shielding Effectiveness Versus Frequency for 1-inch Seamless Steel J Pipe (MIL-T-0020 157C Type E) Conduit

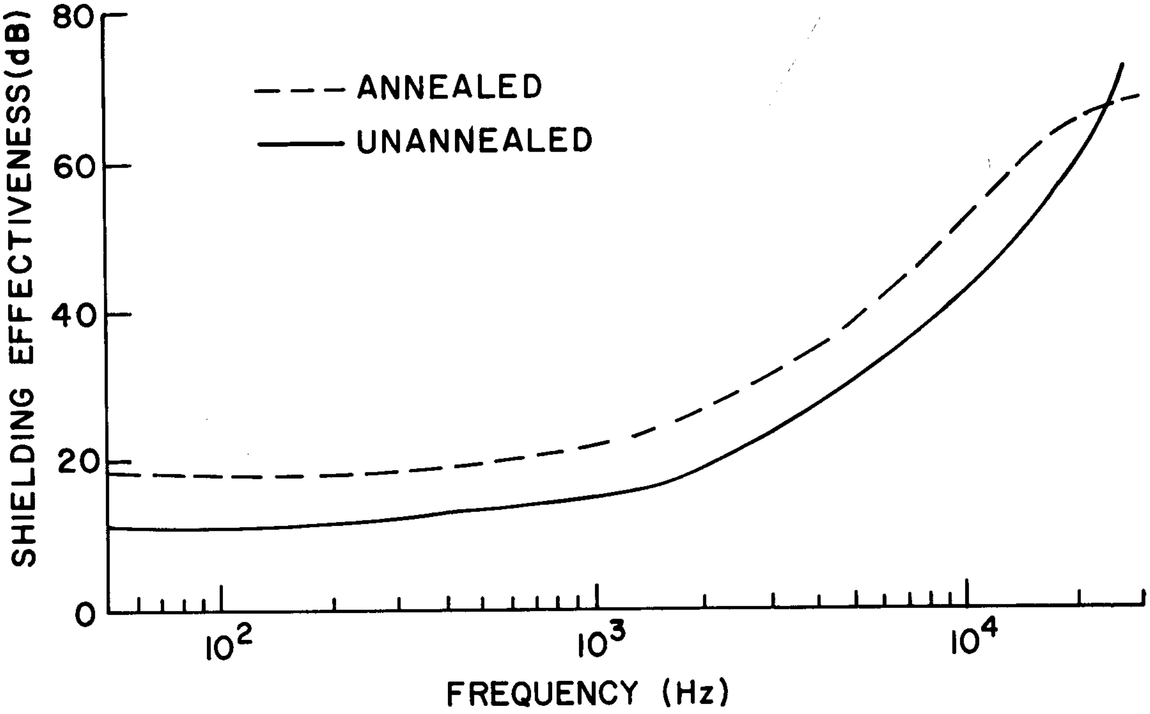


Figure 10-22. Shielding Effectiveness Versus Frequency for 1-Inch Corrugated Steel Inner-Core (Annealed and Unannealed) Conduits

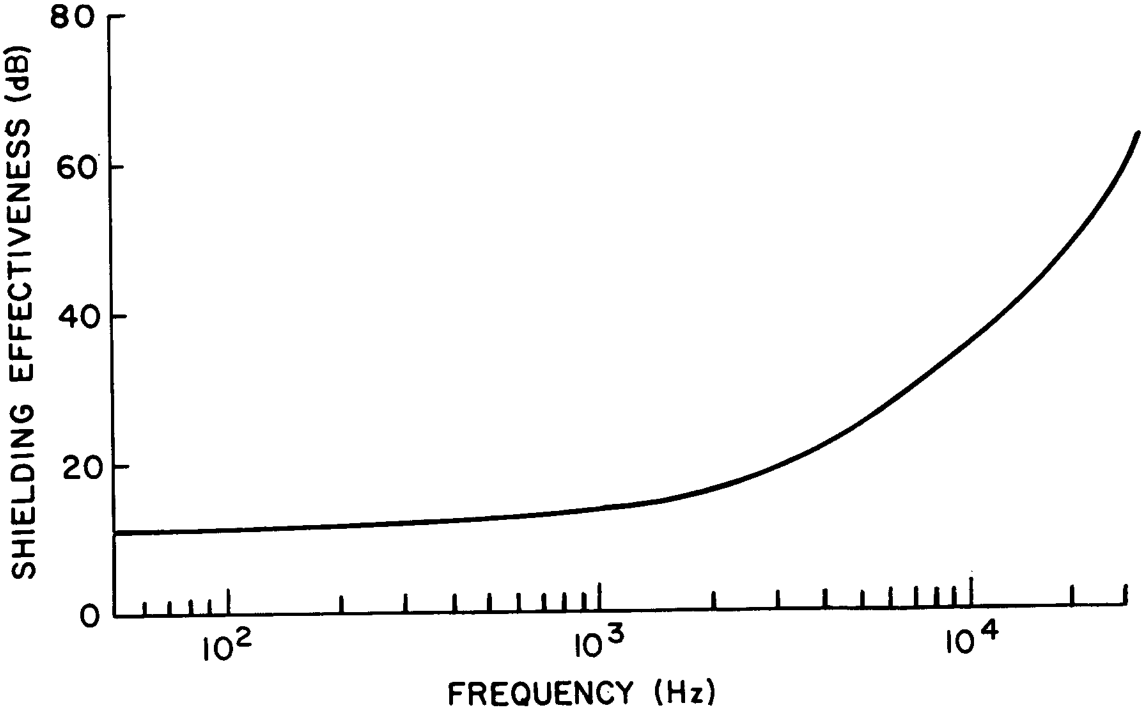


Figure 10-23. Shielding Effectiveness Versus Frequency for 3/4-Inch Corrugated Steel Inner-Core (Unannealed) Conduit

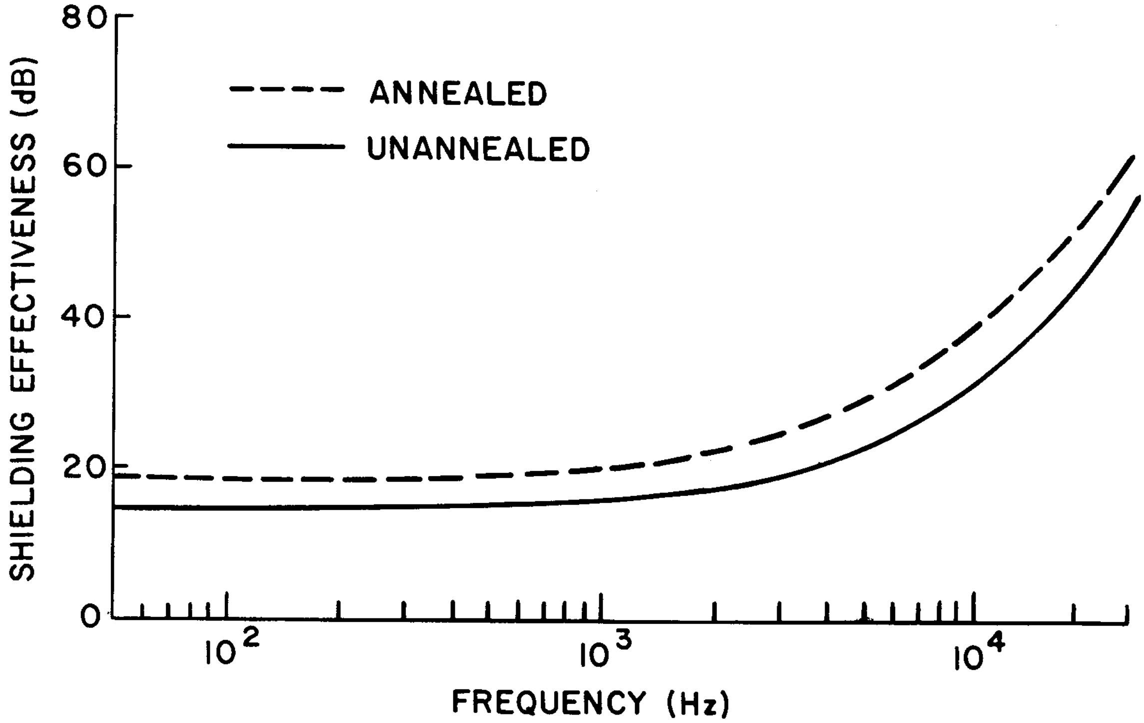


Figure 10-24. Shielding Effectiveness Versus Frequency for 1/2-Inch Corrugated Steel Inner-Core (Annealed and Unannealed) Conduits

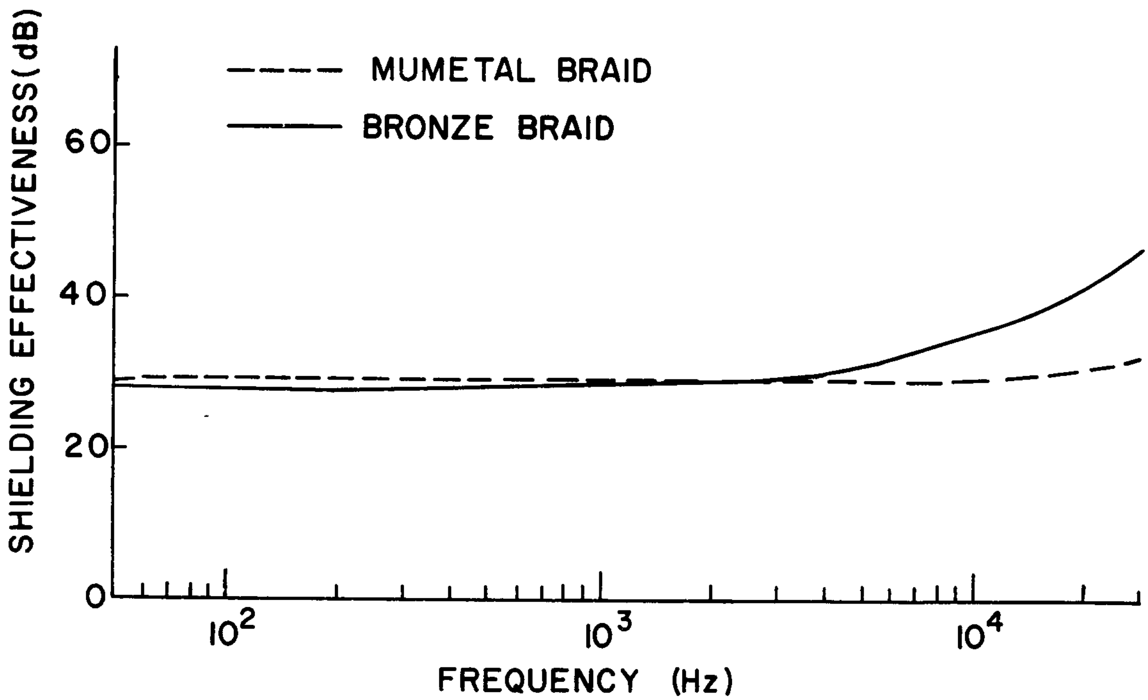


Figure 10-25. Shielding Effectiveness Versus Frequency for 1-inch, 4-mil (Mumetal and Bronze Braids) Conduits

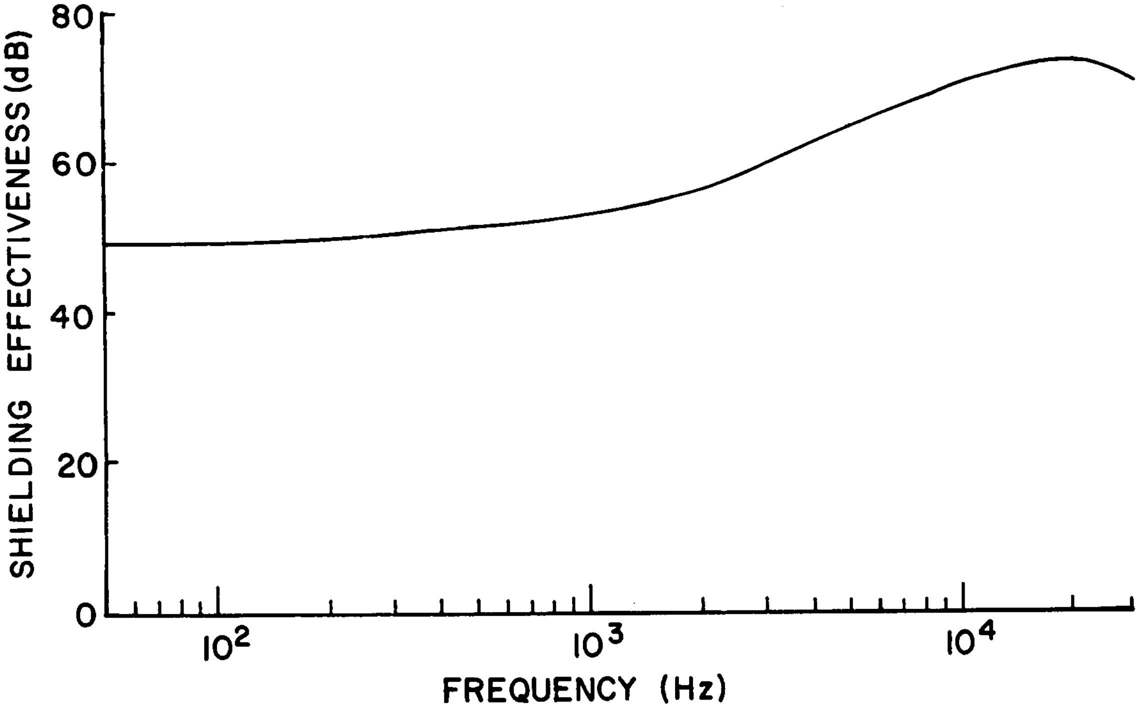


Figure 10-26. Shielding Effectiveness Versus
Frequency for 1-inch, 8-mil (Mumetal Braid) Conduits

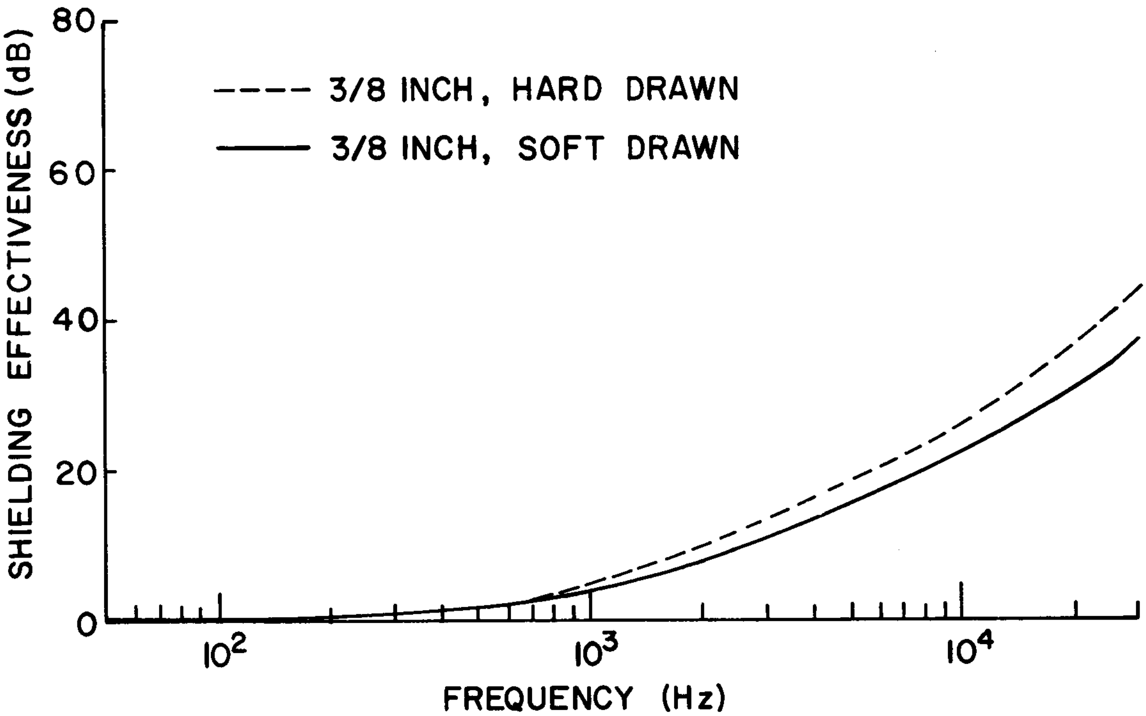


Figure 10-27. Shielding Effectiveness Versus
Frequency for 3/8-Inch Copper Tubing (Hard and Soft Drawn) Conduits

The nomograph in Figure 10-20 will be useful for estimating shipboard low-frequency shielding requirements for various low-level twisted and coaxial transmission lines.

10.2.7 Data Output Openings

Data output openings, such as those for direct view storage tubes, cathode-ray tubes, and meters, represent a large discontinuity in an equipment case. Meters and other visual readout devices often present difficult interference protection problems, particularly at high frequencies, because, at present, materials that are both optically transparent and conductive are ineffectual shields. If fine knitted mesh or conductive glass is used for display shielding, it may not furnish the degree of attenuation required. In most cases, this problem is overcome by installing a shield around the rear of the readout device and filtering or bypassing all leads entering and leaving it. A number of semitransparent shielding materials are available for application to data output and display devices. Included are copper mesh screening, perforated metal, conductively coated glass, and conductively coated plastic. The optical and electrical transmission properties of some of these materials are summarized in Table 10-6. A comparison of relative levels at three frequencies indicates that, for virtually all of the conductively coated materials, the ratio of electrical transmittance to optical transmittance is relatively low. For instance, a 30-micron-thick gold film on plastic yields an electrical transmittance of .16 percent

Table 10-6. Microwave and Optical Properties of Semitransparent Shielding Materials

Material	Microwave Transmittance (Percent)			Optical Transmittance (Percent)
	5.9 GHz	9.7 GHz	18.8 GHz	
Gold film about 11 μm thick on plastic (300 ohms/square)	23	10	0.8	49
Gold film about 30 μm thick on plastic (12 ohms/square)	0.16	0.1	0.01	24
Gold film about 75μm thick on glass (1.5 ohms/square)	0.04	0.01	0.004	3.2
Copper mesh (20 per inch)	0.1	0.2	0.2	50
Copper mesh (8 per inch)	1.0	1.3	2.5	60
Lead glass (X-ray protective, 1/4-inch thick)	30	25	16	85
Lucite (3/16 inch thick)	80	50	25	92
Libby-Owens-Ford Electrapane glass, with conductive coating about 150 μm thick (120 ohms/square)	16	16	16	85
Libby-Owens-Ford Electrapane glass, with conductive coating about 300 μm thick (70 ohms/square)	9	10	8	80
Corning heating panel glass, with conductive coating about 1.5 μm thick (15 ohms/square)	1.6	1.2	0.08	45

(at 5.9 GHz) and an optical transmittance of 24 percent. The electrical transmittance decreases with frequency; at higher frequencies considerably higher electrical transmittance (poorer shielding qualities) can be expected. The twenty-mesh copper screen, while showing the same order of electrical transmittance as the 30μm gold film at 5.9 GHz, has approximately twice the optical transmittance. In addition, the shielding effectiveness of copper mesh improves with decreasing frequency; although at higher frequencies, comparison of the two materials would favor the 30μm gold film. Economically, copper mesh is more practical because deposition of thin metal films is an expensive process compared with utilization of copper mesh.

10.2.7.1 Cathode-Ray Tubes. Cathode-ray tube openings represent one of the largest discontinuities and are of a type most difficult to treat. The disruption of the shield is such that, if not considered in the initial design stages, it may become extremely difficult to achieve an acceptable final product. Treatment of the opening is made additionally difficult by the requirement for unrestricted transparency. For good shielding construction, it is necessary for all items that penetrate the shielding, such as pipes and conduits, to be electrically bonded to the shielding at the point of entrance by soldering, brazing, or welding. Handles, latches, screw heads, nails and other metal projections that pierce the shield should be brazed or soldered to the shield; all breaks should be bonded in continuous seams. These precautions prevent the antenna effect: e.g., a metal element that projects through the shielding can act as a receiving antenna on one side, picking up radiated energy and reradiating signals in the opposite direction on the other side. At high frequencies, such isolated hardware is comparable to, and can radiate as, a waveguide probe.

10.2.7.2 Indicating and Elapsed Time Meters. Meter movements when shielded, are not affected by ac or dc magnetic fields. The most common sources of interference to unshielded meters are transformers, motors, generators, current-carrying cables or buses, solenoids, and other components producing magnetic fields. Shielding against interference from such items permits unrestricted use of meters in otherwise difficult environments. Figure 10-28 illustrates typical shielding of a panel meter.

10.2.7.3 Fuse Holder and Indicator Lamp Openings. In electronic equipment, both active and spare fuses may be sources of radiation because they act as antennas. The lack of shielding in most fuseholders permits internal high frequency interference

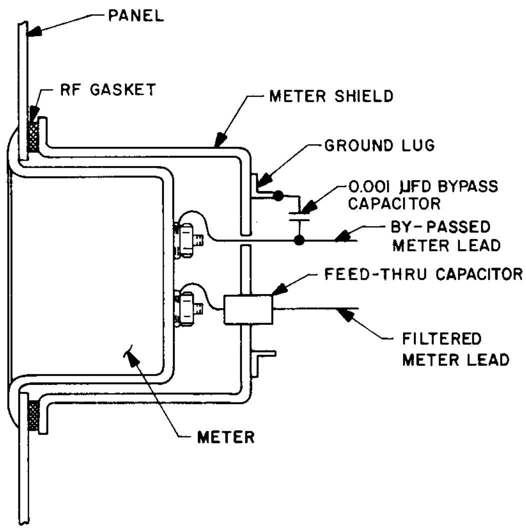


Figure 10-28. Meter Shielding and Isolation

to leak through the opening in an otherwise well shielded panel. One solution is to group all fuseholders together so that a solid metal shield with wire mesh gasketing may then be used to surround the fuse cluster. This method can be used for indicating lamps, provided screening or special conducting glass is substituted for the solid metal fuse holder shield.

10.2.8 Spiral-Wrap Shielding

Switching devices such as relays, switches, etc., sometimes require high surge currents in their operation. The inrush of current through interconnecting leads creates high energy fields which are radiated as interference transients. A reduction method for this type of interference involves wrapping the offending wires with Mumetal tape. Successive spiral wraps of tape ensure a minimum of gaps and permit flexibility. Such spiral-wound shielded cables are commercially available. Zipper tubing is not recommended for cable shielding, but it can be used to hold the foil wrap in place.

10.2.9 Conductive Surface Coatings

Metal-to-metal contact can be improved significantly by applying conductive surface coating specifically formulated for shielding applications. It is composed of fine, silver-based lacquers that adhere excellently to metal, plastic, ceramic, wood and concrete. When this coating is applied to a nonconductor, the surface resistivity becomes substantially less than one ohm/square inch and successive coats further reduce the surface resistivity. The coating can be either brushed or sprayed on and it is sufficiently fluid so that it readily flows into cracks.

For gross voids, a putty-like conductive coating is available which can be applied by hand or with an air-activated gun.

10.2.10 Conduit for Exterior Cable Runs

10.2.10.1 Introduction Runs of armored cable ascending the mast structure are exposed to an extremely corrosive environment that rapidly deteriorates the metal armor. When several cables are in one bundle, the number of potential nonlinear joints due to galvanic corrosion is endless. These nonlinearities are a strong contributor to the hull-generated intermodulation problem.

In addition, intense illumination from radio and radar transmitters induces high rf currents in the metallic cable armor with subsequent arcing to adjacent metal structures. Broadband interference results.

10.2.10.2 Basic Premise of Mast Conduit. A fairly recent addition to shielding techniques is the use of rectangular conduit to completely enclose all armored cables ascending the mast structures. The conduit is locally fabricated to fit the specific requirements of a ship. The intent is to shield weather-exposed armored cable from the weather as well as from illumination from shipboard emitters. Such conduit is made from materials compatible with the ship's mast structure and welded to the mast structure. Access should be provided in order to install or remove cables within the conduit. Use of conduit precludes the requirement for bonding of the cable armor inside the conduit.

10.2.10.3 Typical Conduit Installations. There are no rigid guidelines at present for installation of conduit, as each ship arrangement is different. However, Figure 10-29, 10-30 and 10-31 illustrate some general methods used.

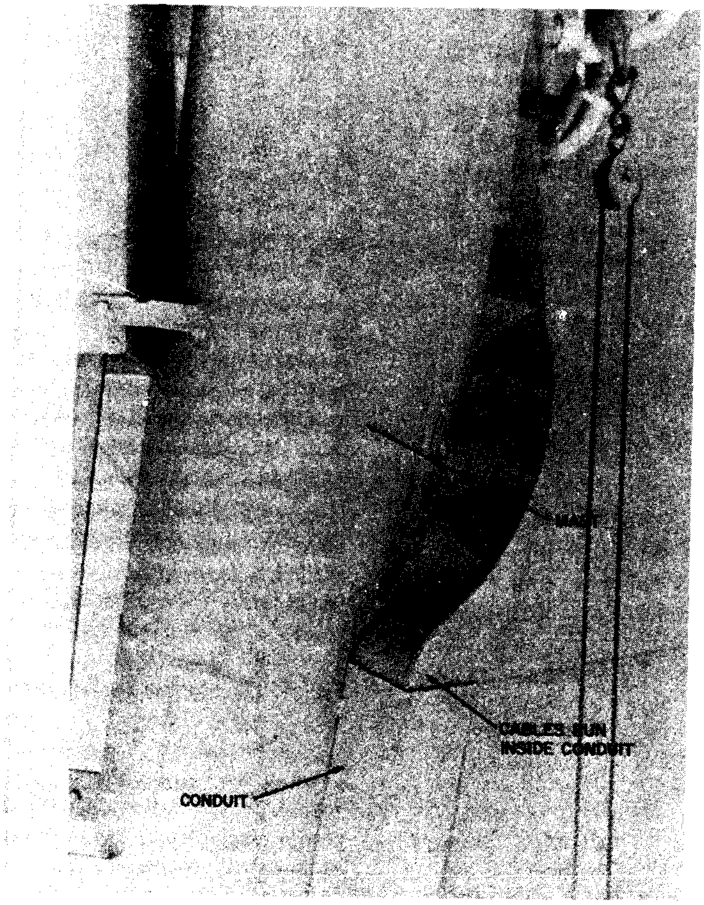


Figure 10-29. Typical Usage of Mast Conduit

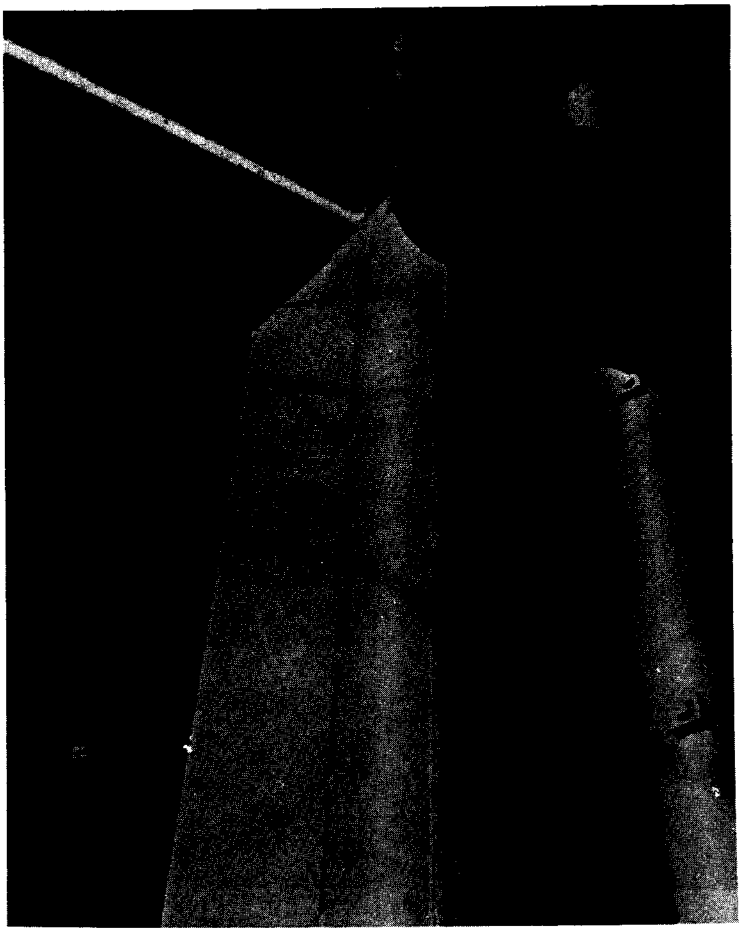


Figure 10-30. Typical Usage of Mast Conduit

10.3 FILTERING

10.3.1 Introduction

Filters are combinations of circuit components designed to pass currents at certain frequencies or to attenuate currents at other frequencies. They utilize the resonance characteristics of series and parallel combinations of inductance and capacitance.

These reactances reduce interference by introducing a high impedance in series with the interference currents and/or shunting interference currents to ground through a low impedance. Figure 10-32 shows the attenuation versus frequency curves for four common filters.

10.3.2 Definitions

For a better understanding of filters, several terms need defining:

a. **Insertion loss** is the attenuation of the desired signal due to the insertion of the filter when properly terminated. Ideally, this should be zero; practically, it is kept as small as possible.

b. **Mismatch loss,** as opposed to insertion loss, is the attenuation suffered because of impedance mismatch in the system. Sometimes mismatch loss and insertion loss are inseparable; in other cases, they are distinctly different.

Figure 10-31. Typical Usage of Mast Conduit

c. **Cutoff frequency** is that frequency or those frequencies toward the edge of the response at which the attenuation is 3 dB greater than the insertion loss value. The reason for specifying the edges of the response is that some practical bandpass filters might tolerate a 3 dB (or greater ripple) in the band pass.

d. **Bandwidth** is the frequency separation between the 3 dB cutoff frequencies. If the term is applied to a low-pass filter, it means the band from zero frequency up to the cutoff frequency.

e. **Q-factor** (applies to bandpass) is the ratio between the center frequency (defined as f_1 f_2) to the 3 dB bandwidth (defined as $f_2 - f_1$).

f. **Shape factor** (for bandpass) is the ratio of the 60 dB bandwidth to the 6 dB bandwidth.

g. **Impedance level** is the value in ohms from which and into which the filter is to work. The input and output impedances may be the same or different, though the former is often the case in communications work.

h. **Power handling** capacity is the average power that the filter will safely pass without degradation or failure of any component.

10.3.3 Classes of Filters

Filters can be classified according to their method of construction and operation.

10.3.3.1 Resistance-Capacitance (RC) Filters. For a "brute force" filtering technique, the RC filter (Figure 10-33) provides a cheap way of accomplishing some frequency selective properties. It is commonly used for interstage coupling of audio amplifiers. The "skirt selectivity" of these filters leaves much to be desired. For bandpass applications two RC filters may be placed in tandem or if high selectivity is required, a parallel T may be employed.

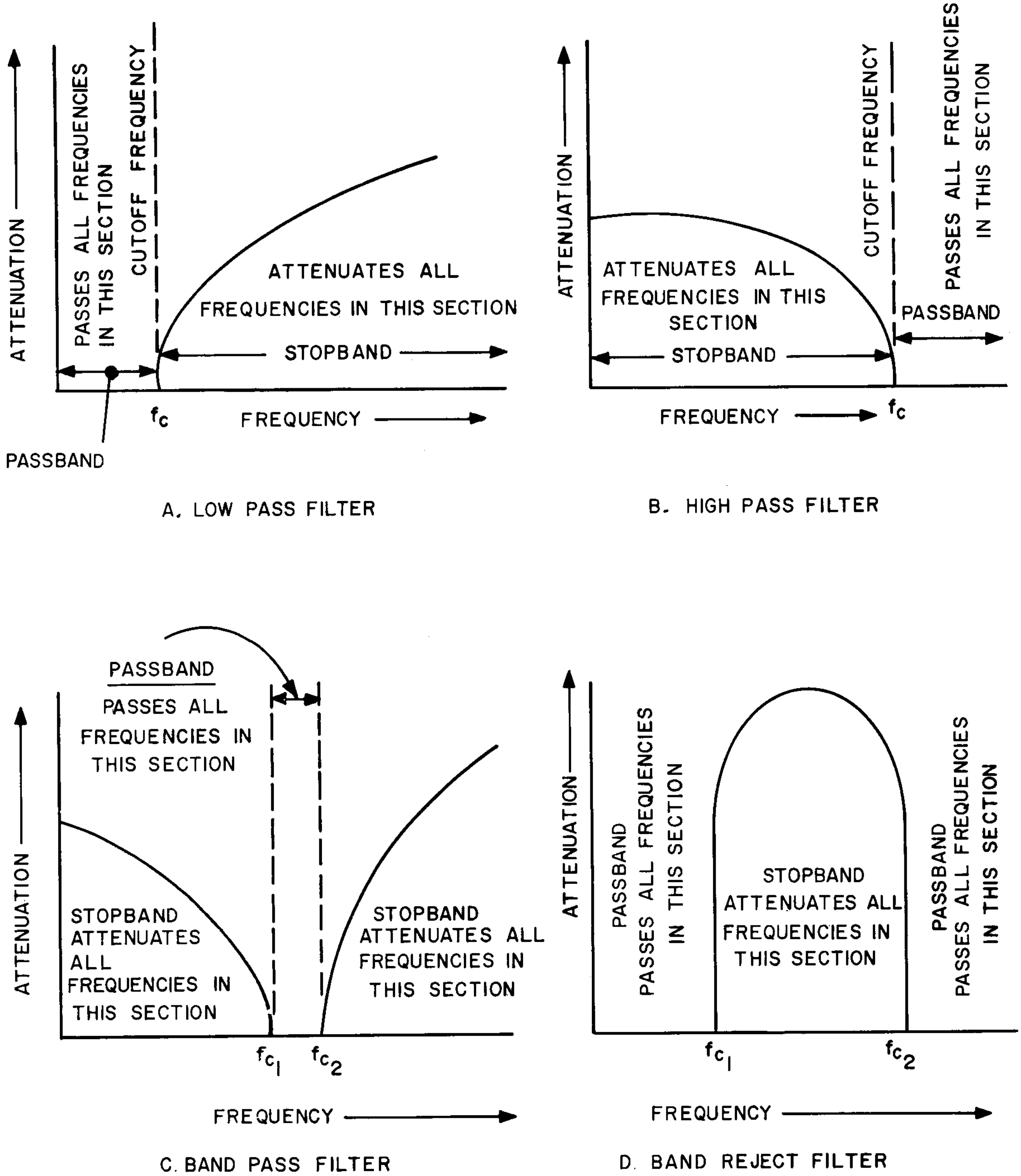


Figure 10-32. Filter Attenuation – Frequency

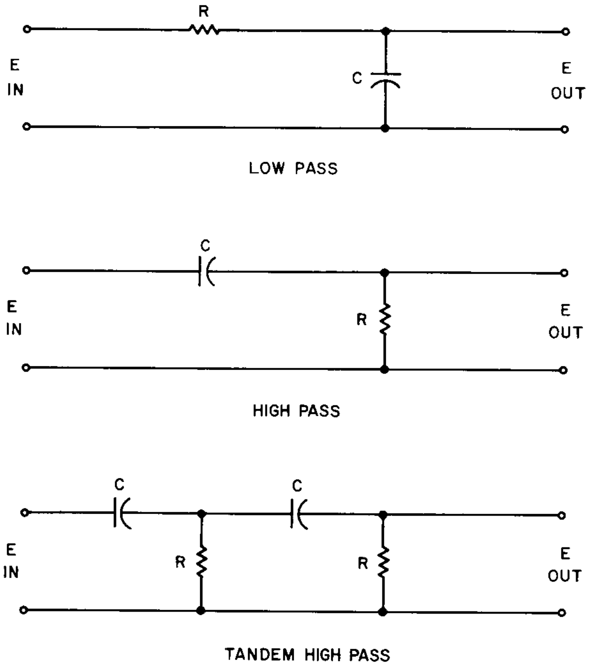


Figure 10-33. RC Filters

10.3.3.2 Constant-k Filters. Constant-k filters are designed by what is known as the image parameter method. Image impedances of a network are those impedances, not necessarily equal, one of which is used to terminate the network at one end resulting in a value equal to the other at the opposite end of the network, and vice versa. Image impedances are equal for symmetrical networks, sometimes called characteristic impedance, and unequal for unsymmetrical networks.

Constant-k filters may be represented either as "T" or "π" sections. The impedance characteristics of these are different.

10.3.3.3 M-Derived Filter. M-derived filters are arrived at by modifying the elements of the constant-k filter prototype by another factor m and adding one or more elements. Two things are accomplished: (1) sharper cutoff; (2) a more favorable impedance characteristic.

10.3.3.4 Butterworth Filter. The Butterworth filter, sometimes termed the "maximally-flat" filter, yields a passband which is "flat" on the top, and a skirt response which is much sharper than the constant-k. Although the constant-k, m-derived composite filter, may, at some point beyond cutoff, give superior performance, the overall response curve of the Butterworth is more desirable.

10.3.3.5 Tchebycheff Filter. The Tchebycheff filter is used when a still sharper skirt response is desired, but "flatness" in the response region can be degraded to allow "ripple." Sometimes this type of filter is called an "equal ripple" filter. It is exemplified by some overcoupled tuned IF transformers. (See Figure 10-34 for a rough comparison of the responses of the types mentioned.)

10.3.3.6 Mechanical Filter. The mechanical filter is a relatively new concept in the field of selectivity. It contains an input transducer, a resonant mechanical section comprised of a number of metal discs, and an output transducer.

The frequency characteristics of the resonant mechanical section provide the almost rectangular selectivity curves shown in Figure 10-35. The input and output transducers serve only as electrical-to-mechanical coupling devices and do not affect the selectivity characteristics which are determined by the metal discs. An electrical signal applied to the input terminals is converted into a mechanical vibration at the input transducer by means of magnetostriction. This mechanical vibration travels through the resonant mechanical section to the output transducer, where it is converted by magnetostriction to an electrical signal which appears at the output terminals.

In order to provide the most efficient electromechanical coupling, a small magnet in the mounting above each transducer applies a magnetic bias to the nickel transducer core. The electrical impulses then add to or subtract from this magnetic bias, causing vibration of the filter elements that corresponds to the driving signal. There is no mechanical motion except for the imperceptible vibration of the metal discs.

Magnetostrictively-driven mechanical filters have several advantages over electrical equivalents. In the range of 100 MHz to 500 MHz, the mechanical elements are extremely small, and a mechanical filter having better selectivity than the best of conventional IF systems may be enclosed in a package smaller than one IF transformer. Mechanical elements with Qs of 5000 or more are readily obtainable. The frequency characteristics of the mechanical filter are permanent, and no adjustment is required or is possible. The filter is enclosed in a hermetically sealed case.

In order to realize full benefit from the mechanical filter's selectivity characteristics, it is necessary to provide shielding between the external input and output circuits capable of reducing transfer

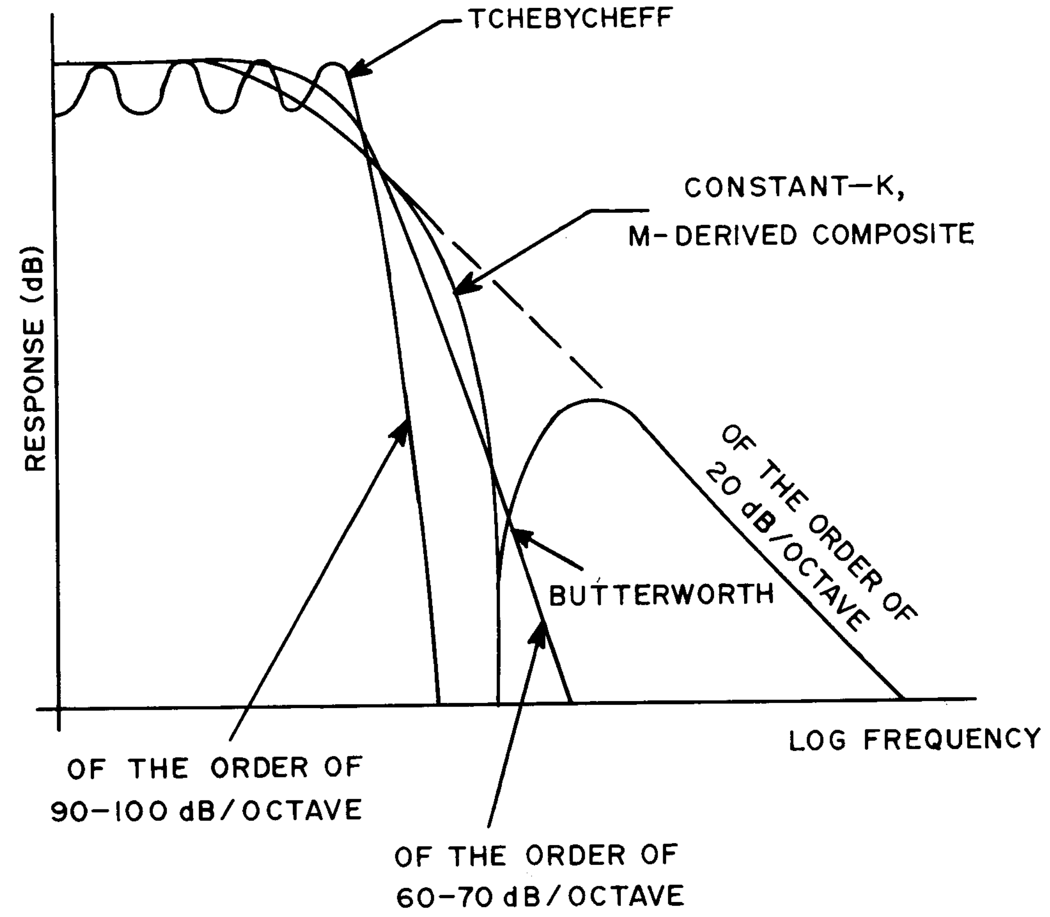


Figure 10-34. Rough Comparison of Filter Types Having Equal Number of Sections

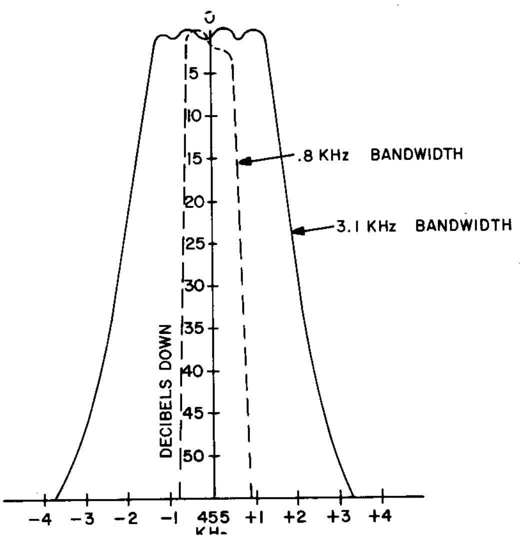


Figure 10-35. Selectivity Curves of 455 kHz Mechanical Filters

of energy external to the filter by a minimum value of 100 dB. If the input circuit is allowed to couple energy into the output circuit external to the filter, the excellent skirt selectivity will deteriorate and the passband characteristics will be distorted.

As with almost any mechanically resonant circuit, elements of the mechanical filter have multiple resonances. These result in spurious modes of transmission through the filter and produce minor passbands at frequencies on both sides of the primary passband. Design of the filter reduces these sub-bands to a low level and removes them from the immediate area of the major passband. Two conventional IF transformers supply increased attenuation to these spurious responses and are sufficient to reduce them to an insignificant level.

10.3.3.7 Crystal Filter. The passband of an intermediate frequency amplifier can be made very narrow by use of a piezoelectric filter crystal employed as a series resonant circuit in a bridge arrangement known as a crystal filter. The shape factor is quite poor, but the very narrow passband obtainable as a result of the extremely high Q of the crystal makes this type filter useful for c-w telegraphy reception. The passband of a typical 455 kHz crystal filter can be made as narrow as 50 hertz; the narrowest passband obtainable with a 455 kHz tuned circuit is about 5 kHz.

10.3.4 Filter Characteristics

10.3.4.1 Ratings. Filters are usually inserted in a circuit so that all circuit energy passes through them; they must, therefore, perform their functions without impairing normal circuit operation. Filters are generally rated in terms of the voltage and current parameters of the circuit in which they operate.

10.3.4.2 Attenuation and Insertion Loss. Attenuation and the frequency range of attenuation are the primary characteristics that determine filter suitability for interference reduction. If a selected filter does not provide the minimum attenuation required in the stop-band, it is not satisfactory, regardless of its other characteristics. Figure 10-36 illustrates a typical attenuation-versus-frequency curve for a power-line filter. The attenuation of the filter is expressed as the ratio of the filter input voltage to the filter output voltage, measured under normal circuit conditions.

The attenuation figure, however, does not take into consideration the source and load impedances and, therefore, does not represent a true indication of the suppression effectiveness of a filter. The use of the insertion loss criteria is a far more realistic measure of the effectiveness of a filter, as it is a function of both source and load impedances, as well as a function of the filter network itself.

10.3.4.3 Filter Considerations. The following characteristics are common to all filter installations and should be carefully considered in filter selection.

a. Voltage rating of the circuit in which the filter is to be inserted and the maximum current that will pass through the filter. Voltage and current ratings required of a filter, unless otherwise specified, are the maximum allowable for continuous operation. Any filter will perform satisfactorily when operated below its nameplate rating. The breakdown voltage of capacitors used in filters should also be considered. A safety factor of approximately 100 percent should be used. For a given application, the working voltage of a standard filter capacitor should be twice the voltage of the circuit in which it is used. In general, filter test voltage should be twice the filter's nameplate rating.

b. Duty cycle of the filter; this applies to the decreased load current of intermittent operation.

c. Operating frequency of the circuit and the frequencies to be filtered. Power-frequency specifications are primarily applicable to low-pass line filters. Filters should not be operated at power frequencies above those specified by the manufacturer. They will operate satisfactorily at frequencies below those marked on the nameplates.

d. Voltage drop that can be tolerated at the operating frequency. All filters using series inductors

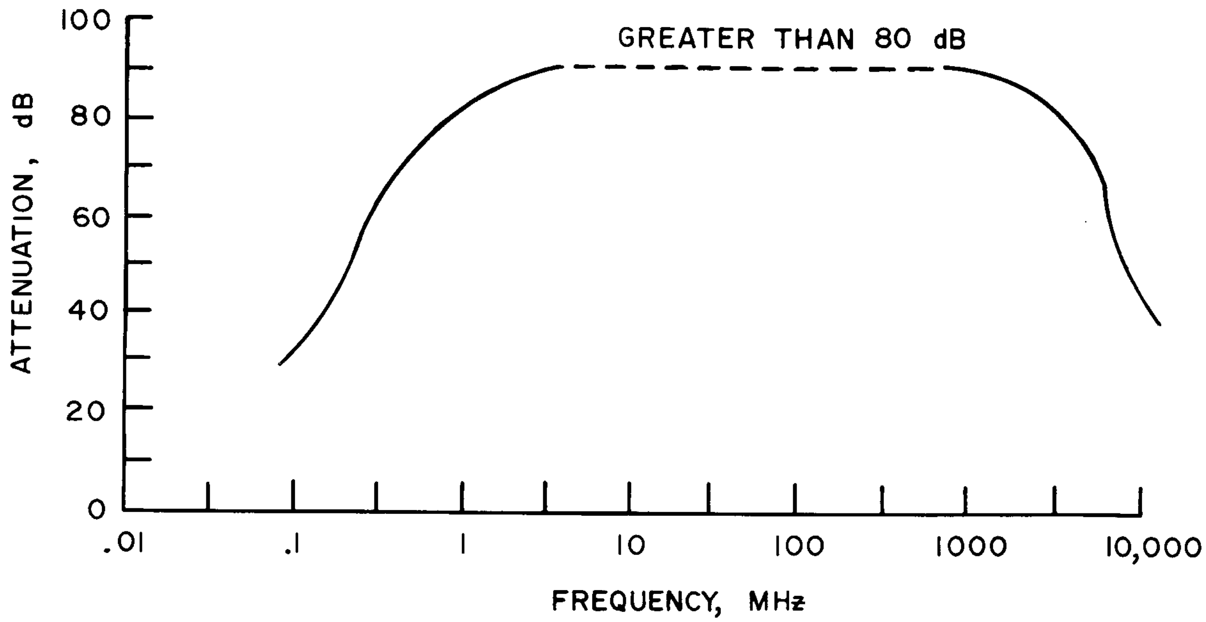


Figure 10-36. Attenuation Versus Frequency for a Power Line Filter

cause some voltage drop. The magnitude of this drop is determined by the series resistance of the filter and may be expressed, for example, as: maximum voltage drop at rated current = 0.1 volt. In some cases, only the series resistance is given. When this occurs, the voltage drop for dc filters, or ac filters when the cut-off frequency is far removed from the operating frequency, can be calculated by the application of Ohm's law.

e. Maximum ambient temperature at which the filter must operate. If the ambient temperature deviates from the range specified for the filter, failure or shortened service life may result.

f. Attenuation required of the filter for adequate interference reduction.

g. Minimum filter life – the number of hours a unit will operate satisfactorily under rated conditions and at maximum ambient temperature.

h. Circuit requirements such as minimum (or maximum) capacitance or insulation resistance. The capacitance to ground of a filter is often a determining factor in the filter's application. Some circuits may limit the maximum capacitance to ground to prevent long time-constant currents from charging the capacitors through a resistor. Another instance in which capacitive limitations are important is when capacitors might cause danger to personnel because of charging currents.

The insulation resistance of a filter decreases continually during the life of a filter. The resistance of a new filter usually measures several hundred megohms, and after several years of operation may measure 50 megohms. Because most power cables can be used with an insulation resistance of one megohm, the insulation resistance of a filter has little effect on such cables.

10.3.5 Lossy Line Filter

The usual π or T filters are generally composed of lossless or very nearly lossless inductive and capacitive lumped elements. Such filters do not dissipate energy; they merely reflect it, reroute it, or transform it. Under certain conditions, the energy may reappear elsewhere as an undesirable signal or interference.

In the lossy type filter, the loss versus frequency characteristics of a ferrite material is utilized to adsorb and dissipate rf power. It can be constructed of a short length of ferrite tube with conducting silver coatings which form the conductors of a coaxial transmission line. The line is extremely lossy; that is, it has high attenuation which increases with frequency. It is essentially a low-pass filter used mostly for power line filtering.

10.3.6 Transmission Line Filters

This type filter differs from lumped filter elements in two respects: (1) extremely high Q, and (2) very different frequency characteristics.

Both of these qualities can be used to advantage in some applications. For example, "wave traps" can be effectively realized with resonant transmission line elements. Suppose, for example, one wished to short-circuit the second harmonic component of a signal in a transmission line. A shunt short-circuited stub a quarter-wave long at the fundamental does this nicely. At the fundamental frequency the stub presents an infinite impedance to the line, and thus has no effect. At the second harmonic, however, it appears as a half-wave long short-circuited line and presents zero impedance across the line, thus effectively short-circuiting the second harmonic component. Several of these in succession across the line will reduce the undesired signal to very low values. In a similar way, other harmonics can be blocked.

In addition to the behavior discussed above, there is the possibility of utilizing the below-cutoff characteristic of waveguides for filtering since a waveguide is a natural highpass filter.

10.3.7 Resonant Cavity Filter

Another form of ultrahigh frequency and microwave filter having very high Q is one that utilizes resonant cavities. These are sometimes used in receiver preselector stages and provide selective response while giving greater bandwidth than would be obtained with a single resonant cavity.

10.3.8 Filter Applications

The object of a filter is to discriminate against some frequencies while not affecting others. Typical applications are:

a. Power-line filtering: prevents HF signals from being conveyed by the power line to vulnerable equipment.

b. Bypassing: allows the passage of low frequencies or dc, while providing a low impedance path to "ground" for high frequencies.

c. Feedthrough: allows LF signals to be passed through a panel or bulkhead while bypassing or blocking HF interference.

d. Wave trapping: use of high Q resonant or anti-resonant circuits to reject certain frequencies.

e. Harmonic suppression: prevents harmonics from reaching the antenna or radiating in any other way.

f. Duct and shaft filtering: prevents unwanted signals from leaving an equipment via air ducts or control shafts.

10.3.8.1 Power-Line Filters. Power-line filters are used to prevent unwanted signals from being coupled between equipments by a common power connection. A common technique is to utilize a "brute force" filter, i.e., several ladder sections in which the choice of components is not critical. Sometimes it is necessary for such filters to have extremely broadband performance, e.g., filters for screened rooms. In these cases, multiple sections are used, each designed to reject different portions of the spectrum.

Although many precautions must be taken in the construction of all filters, some particular ones must be followed for this type of filter. Since substantial amounts of power-line current may be required, the filter inductors may require rather large sized wire. This places severe limitations on how much inductive reactance can be secured.

If the power-line filter is for a transmitter of any type then the circulating currents in the inductors and capacitors may be very large. Care must be taken in these cases to choose components that can withstand such currents. Transmitting mica capacitors are usually rated in capacity, frequency of operation, and maximum current.

10.3.8.2. Bypassing. Bypassing may be a part of normal circuit design or it may be an EMC problem. The objective in bypassing is to provide a low impedance shunt path to effectively "short circuit" unwanted signals. Typically, a capacitor is used where its impedance, at the frequency in question, is less than a tenth of that of the circuit being bypassed. It is to be noted that a 10:1 impedance ratio may not be enough in some cases. Unfortunately, it does not always suffice to merely use a larger capacitor, since these elements have internal and lead inductance which provide a frequency limit above which their capacitive properties cannot be realized. The frequency limit is imposed by the self resonant properties of the capacitor due to internal and lead inductance. The self-resonant property is sometimes utilized to obtain nearly perfect bypassing in a narrow band of frequencies; the resonance is a high Q series resonance which provides, in the region of resonance, a very low impedance. This technique has been used in high frequency amplifier design. Capacitors with minimal leads are available which attempt to place the series resonant frequency above the range of interest and provide a large amount of capacity in a small package. Standoff and button capacitors are representative of this class.

10.3.8.3 Feedthrough. The bypass problem is virtually duplicated in feedthrough applications. The solution lies generally in the use of a feedthrough capacitor which provides a conductive path from one terminal to the other, but gives at the same time a high capacity (up to 2300 pf) to a bulkhead or mounting panel. In special cases, transmission line section may be used instead, making use of the impedance transforming property of a quarterwave line.

10.3.8.4 Wave Traps. The nomenclature of "wave trap" evolved in the early days of radio when interference problems were solved mostly by the provision of resonant or anti-resonant circuits in appropriate places to "trap out" the undesired signal.

The inclusion of shunt series resonant circuits in the category of wave traps actually brings us back to the series resonance bypass application already discussed, and no more need be said.

10.3.8.5 Harmonic Suppression. A generator normally produces harmonics of its fundamental. Some types of generators are more notorious than others in this regard. At low power levels the existence of some harmonics may not be a serious problem. With high power systems such as radar, however, even if the harmonic is 60 dB down from the fundamental, it may still represent an interfering power measured in watts.

One way of suppressing harmonics is by utilization of the wave trap or bypass principles already mentioned. In the case of a transmission line system, the filter elements may take the form of shunt or series stubs.

The difficulty with this technique is that the harmonic power is reflected back to the generator which may misbehave because of the presence of these extra reactive volt-amperes. One way of avoiding this difficulty in theory would be to use an isolator between the generator and filter. Unfortunately, most isolators are designed to work as isolators at the fundamental frequency, and their performance at harmonic frequencies is unspecified. This becomes increasingly true at the higher microwave frequencies.

10.3.8.6 Duct and Shaft Filters. When holes are put in equipment cabinets for the purpose of providing air cooling, these same holes potentially provide a route by which interference can be conveyed to vulnerable equipment. Two techniques are used to reduce the tendency to interference. The first of these, a shielding technique, uses effectively metal mesh filters to close the air passages electrically. In actuality, this is similar to the second technique which deliberately leaves openings but chooses their size to be below cutoff when considered as waveguides. More often, the holes are connected with definite air tubes. The size and number of tubing must also be consistent with air flow requirements.

The use of tubes makes the attenuation much more readily calculable than in the case of just the holes alone in a panel because, in the latter case, there are "edge" effects which obscure the behavior.

The wire mesh mentioned first is, in actuality, a proliferation of small irises, all of which are below cutoff.

10.3.9 Departure of Filter From Ideal

Generally speaking, circuits never behave quite as predicted. In the case of filters with lumped elements, the presence of dissipation and stray reactive effects causes departure from ideal characteristics. At very low frequencies, dissipation in reactors becomes a limiting factor. In the radio frequency range, dissipation and stray effects are important. At higher frequencies, stray effects are likely to be more important than losses.

In filters that utilize transmission lines and waveguides, good performance can be obtained in the VHF, UHF, and lower microwave frequencies, but at higher frequencies, losses and mechanical tolerances begin to play a role in degrading performance. Even so, fairly close performance to design specifications can be obtained by careful attention to details.

10.3.10 Filter Installation and Mounting Techniques

10.3.10.1 General. Electromagnetic interference can emanate from equipment both by radiation through space and by conduction through power lines and control circuits. No matter how well shielded a source may be, the shielding effectiveness can be nullified by conduction interfering currents in power, control, and instrumentation leads. Capacitors and filters prevent interference from reaching other circuits by introducing high-impedance paths for the interfering currents, or shunting them from the load through a lower impedance to ground, or providing a reflective mismatch to the interference.

When filters are used, proper installation is absolutely necessary to achieve good results. Effective separation of input and output wiring is mandatory, particularly for good high-frequency performance, because the radiation from wires carrying interference signals can couple directly to output wiring, thus circumventing and nullifying the effects of shielding and filtering. Input and output terminal isolation is most easily accomplished by the use of a filter which mounts through a bulkhead or chassis. In all cases where this type isolation is not feasible, isolation by shielded wiring is mandatory. It is highly desirable to locate suppression components in or on the device generating the interference. The rf impedance between filter-case and ground must be as low as possible. The methods of mounting a filter become a very critical at high frequencies If complete isolation is effected between input and output, filter insertion loss will approach the design figure.

The impedance to ground of an improperly mounted filter, from the standpoint of rf grounding, can become sufficiently large in value to cause interference voltages to develop across the impedance at radio frequencies. These voltages reduce the effectiveness of the filter. An important factor in filter performance is the bonding of the filter case to the ground plane structure of the interference source. This requirement is of utmost importance if the filter is to achieve its design performance capability.

It is imperative that the surface on which a filter is mounted, as well as the mounting surface of the filter itself, be clean and unpainted. The rf impedance of the filter case to ground should be as close to zero as possible. If the surfaces are aluminum, a good bond is required; the surfaces should be iridited, never anodized. The mounting ears, or studs, must ensure firm and positive contact over the entire area of the mounting surface. Although the location of the filter depends on the individual application, in general, it should be installed as close as possible to the interference source. In cases where there is no control over the interference source, the filter should be installed at the point of susceptibility. Grounding plays an important role in the application of filters and capacitors, whether the two-terminal bypass type or the three-terminal feedthrough type is used. Grounding is an important factor in the case of feedthrough capacitors and filters because these devices are inherently more effective in the high-frequency ranges than two-terminal capacitors, and there is therefore, more to lose by excessive impedance in the ground circuit. When bypass capacitors are used for interference reduction purposes, only metal-cased bypass capacitors should be employed. The cases should always be grounded directly to the chassis, either by suitable clamps or by threaded-neck type construction. Grounding of bypass capacitors by pigtail leads simply adds additional series inductance; thus the resonant frequency is lowered, and the usefulness of the bypass capacitor is diminished.

10.3.10.2 Chassis Mounting. In general, any of the following five methods can be employed to mount filters on a chassis: (1) tabs on the filter body, (2) screws or bolts on the filter body, (3) a flange on the filter body, (4) a clamp on the filter body, or (5) a feedthrough stud for bulkhead mounting. Figure 10-37 is representative of filter mounting techniques. A typical filter installation is shown in Figure 10-38. The top Figure 10-38 illustrates the preferred method of filter installation; one that is integral with the interference source–in this case, a dc motor. Figure 10-38 also illustrates the difference between proper and improper filter installation. In Figure 10-38A the bulkhead mounting principle is used, and the filter's input and output circuits are completely isolated. Figure 10-38B shows how direct input-output coupling can reduce the effectiveness of the filter, particularly when extremely high magnetic fields exist within the shielded area. This results in interference being coupled to all leads which enter or leave the area. Often, filters fail to perform because of improper installation. Poor installation can result from improper lead routing (Figure 10-39). In this figure, two incorrect methods of mounting a filter are illustrated; both are ineffective at high frequencies. In Figure 10-39, the input and output leads are physically crossed, completely nullifying the effectiveness of the filter. In Figure 10-39, isolation between input and output circuits is not complete due to lack of shielding on the leads–although there is the advantage of ease of assembly and some isolation up to about 5 MHz. The insertion loss drops rapidly because of the coupling of energy across the filter, regardless of the insertion loss originally designed into the filter.

Proper installation of power-line filters is shown in Figure 10-40. One method of achieving designed insertion loss is to shield either the input or output lead, or both. To be really effective, shielded wire must be continuous from the interference source to the filter and/or from the filter to the point of exit from the radiation area. The best way to effect a satisfactory filter installation is to specify feedthrough or bulkhead mounting wherever feasible and to employ circumferential grounding of the filter-case to the bulkhead. Feedthrough capacitors, because of their superior characteristics, are recommended even when mounting in a shield is not feasible. Spot facing of areas around filter mounting bolts is one reliable technique for obtaining an adequate bond. A far more permanent and better bond can be achieved by making the filter housing integral with the housing of the interference generator. The same bonding requirements apply when filters must be mounted on the susceptible unit rather than on the source.

10.4 CABLE SEGREGATION

One means of reducing or eliminating electromagnetic interference onboard ship is by segregation of cables and equipments. Through the use of separation and structures such as false decks, bulkheads, doors, and masts, energy in high level circuits may be prevented from coupling to low level circuits. High energy circuits may interfere with medium and low level circuits either by direct coupling or inductive coupling. Whenever practicable, interference reduction segregation measures should be considered before equipment is installed and should prove more effective than shielding and grounding at a later date.

10.4.1 Definition of Cables by Function

All cables are grouped into three major types: active, passive, and susceptible. Active cables (high level) are radar modulator pulse cables, radio transmitter antenna and transmission lines, and sonar transducer leads when transmitting. Passive cables (medium level) are power and lighting cables, control cables, intercommunication and fire control cables, and cables other than those which are active or susceptible. Susceptible cables (low level) are radio receiving antennas and transmission lines, hydrophone cables, IFF cables, and sonar transducer leads when receiving. All cables functionally categorized with fall into one of the preceding types.

10.4.2 Radar Modulator Cables

Radar modulator pulse cables should be segregated an absolute minimum of 18 inches from all other cables - active, passive, and susceptible - except at the point of entrance to the modulator and transmitter. These cables carry extremely high power over a frequency spectrum from the pulse repetition rate upwards.

10.4.3 Sonar Transducer Cables

Sonar transducer cables should be separated from power cables by 18 inches.

10.4.4 Direct Current Cables

Cables carrying high levels of direct current should be separated from each other to reduce undesired external magnetic fields to a minimum.

10.4.5 AC Power Cables

Cables carrying varying current or voltage should be routed separately from cables carrying dc. Where it

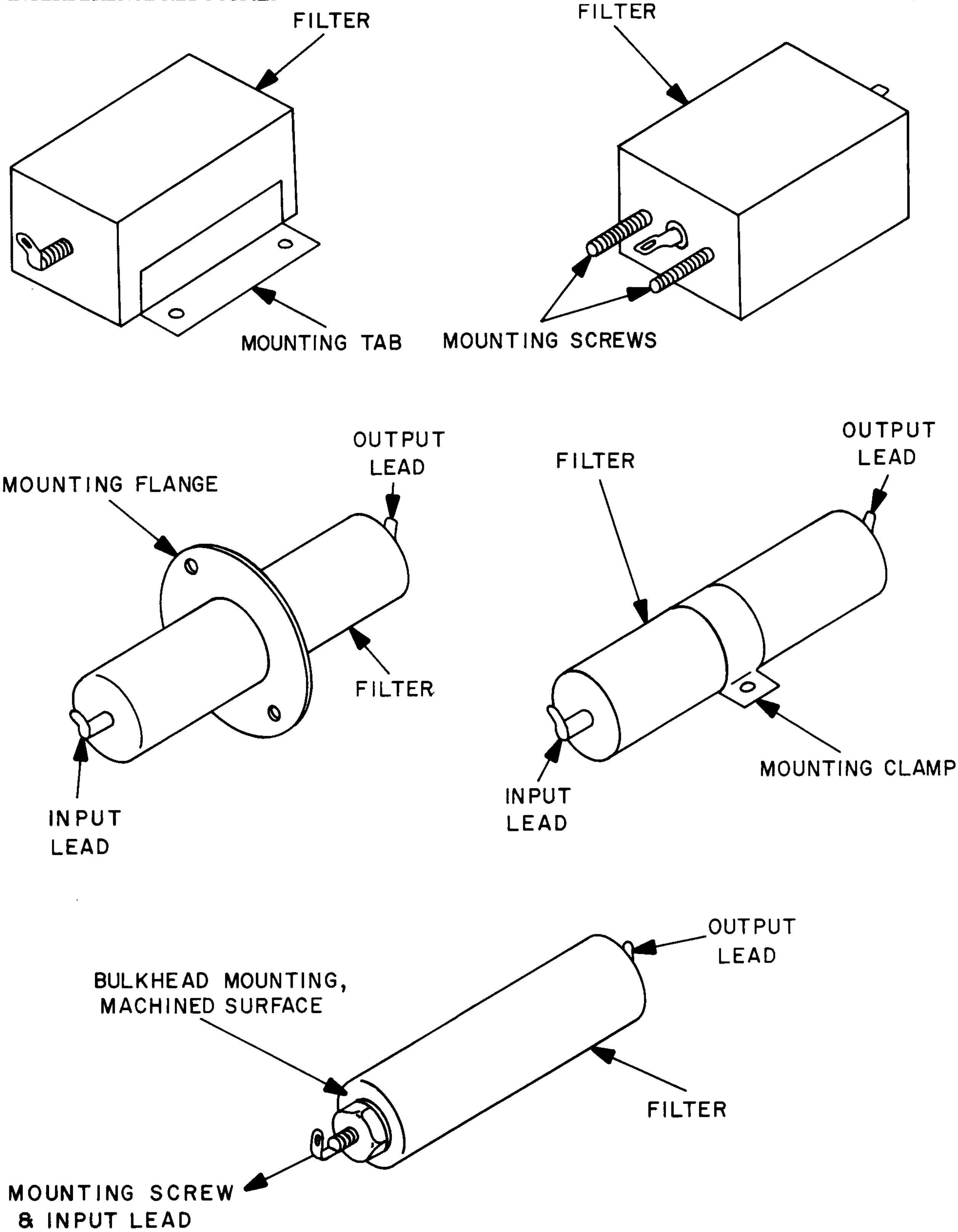
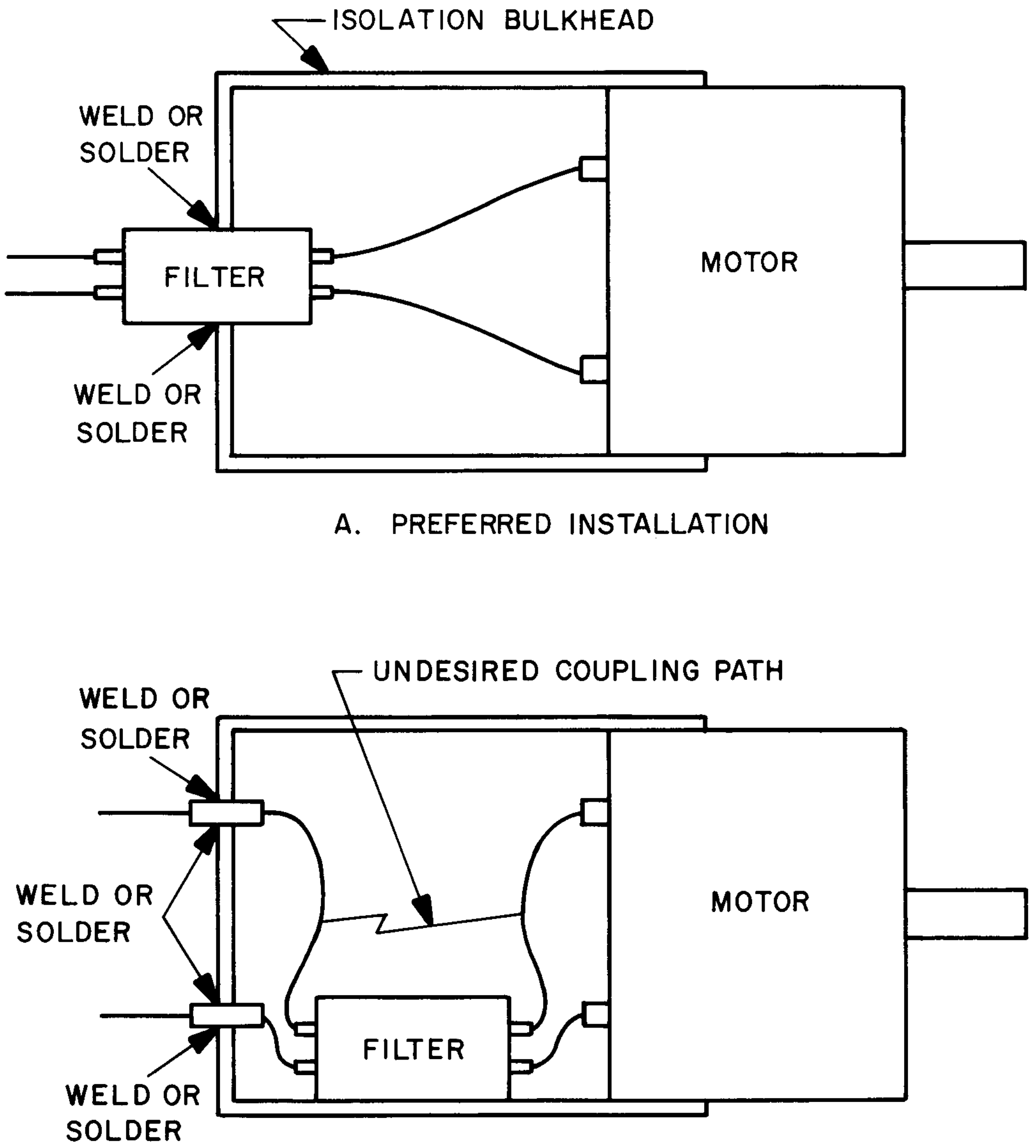


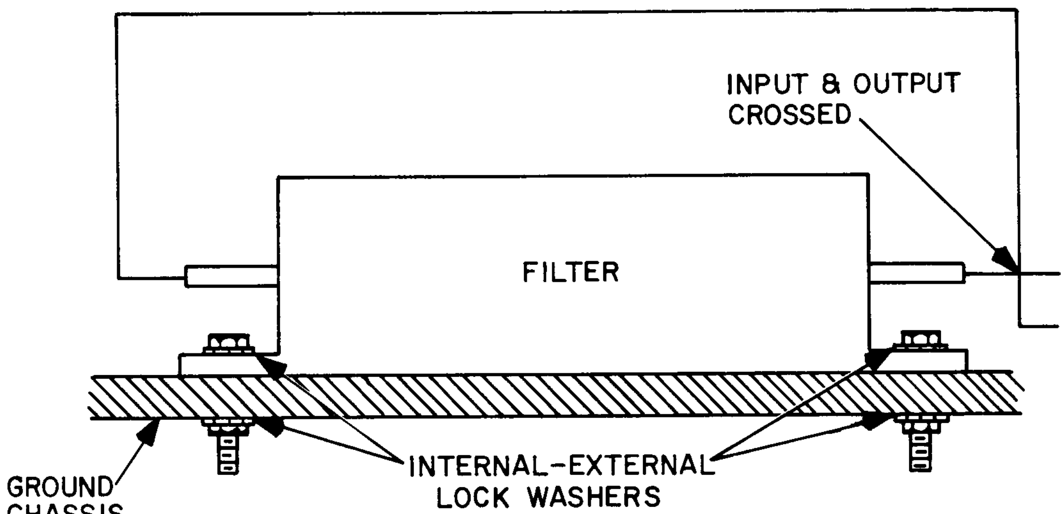
Figure 10-37. Typical Filter Mounting Techniques



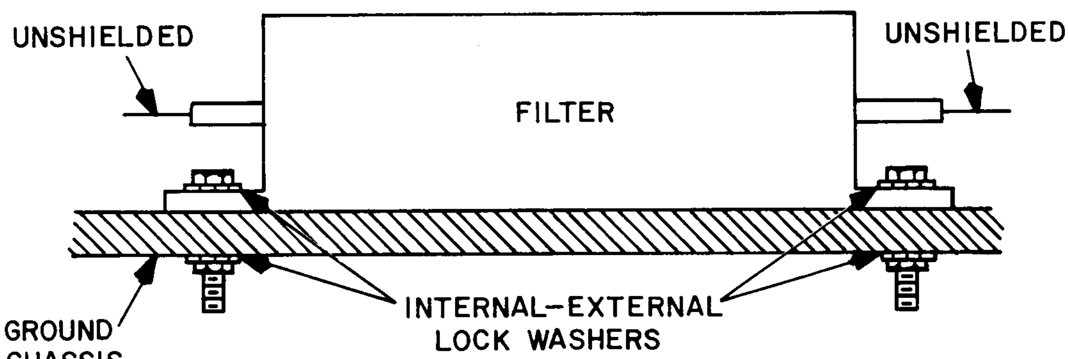
A. PREFERRED INSTALLATION

B. NOT RECOMMENDED

Figure 10-38. Filter Installation



A. CROSSED INPUT AND OUTPUT LEADS



B. UNSHIELDED LEADS

Figure 10-39. Incorrect Filter Mounting Methods

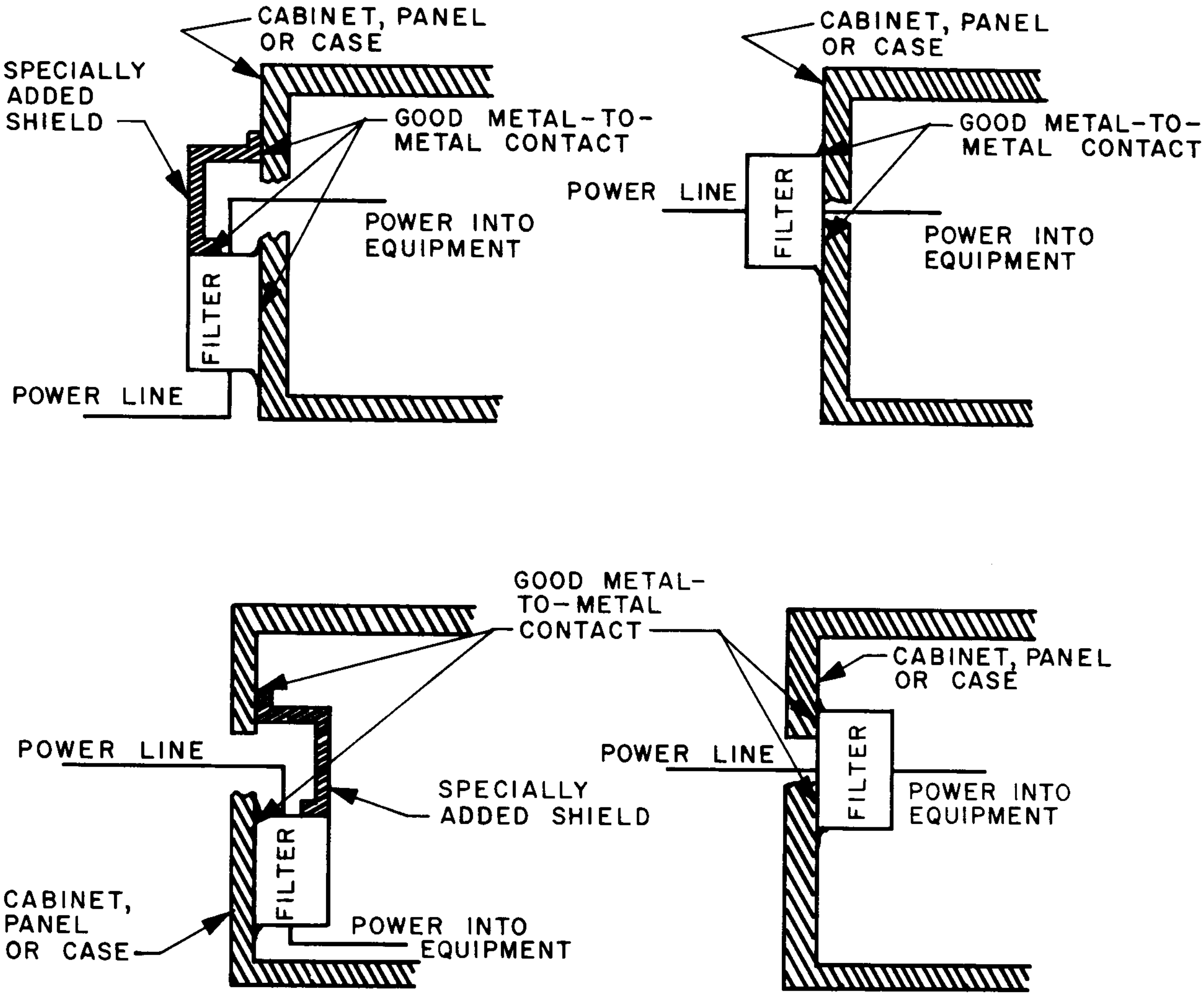


Figure 10-40. Proper Installation of Power-Line Filters

is necessary to run separate phase conductors for alternating current cables, the individual conauctors should be as close as practicable throughout their entire length.

10.4.6 Audio Frequency Cables

Cables carrying audio frequency should be routed separately from each other. In installations where adequate separation is not practicable, audio frequency lines should be twisted pairs with an outer shield.

10.4.7 Negative Power Leads

One-wire system negative power leads should be routed as directly as possible to basic ground with the leads as short as possible. Loops made for neatness of appearance can be very detrimental to reduction of interference.

10.4.8 Interconnecting Cables

Cables interconnecting the units of individual equipments (radar equipment particularly) should be grouped and routed separately from all other cables.

10.4.9 Coaxial and Unshielded Antenna Cables

Coaxial antenna cables should be routed separately from all other antenna cables, such as control cables.

Unshielded antenna transmission lines should be routed in the shortest, most direct route practicable and as free from other antenna cabling as possible throughout their length.

10.4.10 Other Cables

Active cables, other than radar modulator pulse cables, should be separated 18 inches from all passive and susceptible cables; susceptible cables should be segregated at least two inches from passive cables. If tests indicate that interference is present and if segregation is physically impossible, as when cables are terminated at an equipment, shielding is then necessary.

10.4.11 Cable Runs

The shortest cable run between equipments is not necessarily the best run, if the cable is active and passes through a susceptible area compartment. Running the active cable around the susceptible compartment may prevent many problems at a later date.

When mast cables are routed externally, they should be installed so that the cables are not in the main beam of high power radars. If conditions permit, the cables should be run inside the mast. Figure 10-41 illustrates a typical cable run inside a mast.

10.5 USE OF NONMETALLIC MATERIALS

The use of nonmetallic materials in topside areas of ships will reduce hull generated interference as well as minimize antenna pattern perturbations caused by metal structures near the antennas. Nonmetallic material is essentially transparent to rf energy and thus does not intercept or reradiate this energy. Another problem with metallic items – material deterioration caused by corrosion – is eliminated by the use of nonmetallic material. Applications and techniques are constantly improving.

a. Lifelines – Metallic lifelines are among the worst interference offenders aboard ships. Prestretched, double-braided nylon rope and Mylar rope were used for a time as replacements for metalllic lifelines, but their use was discontinued, primarily because they stretched and sagged. To overcome this problem, a glass fiber cable was tested and proven adequate on several ships. This glass-based line is resistant to stretch and has high tensile strength. End fittings can be epoxyed in place, making fabrication easier and faster than that of metallic lifelines. The glass-based line currently approved for use is NUPLAGLAS. Figure 10-42 illustrates nonmetallic lifelines aboard ship.

b. Guys – Ropes made of Mylar, and other synthetic fibers, are being used for antenna guys, mast guys, and similar applications. Although nylon rope is no longer used for lifelines, it remains satisfactory for flag halyards and antenna downhauls.

c. Mast Items – Epoxy glass material is used on some ships to fabricate vertical mast ladders as replacements for existing welded metal ladders. Nonmetallic equivalents also are replacing most service platform metal handrails, and yardarm foot and hand safety rails.

d. Flag Masts – Wood masts, in a few cases, have replaced the metal jackstaffs aboard ships. No bond straps or support posts are necessary for the wood mast, and it meets the requirement for portability. Figure 10-43 and Figure 10-44 illustrate the use of wood masts aboard ship.

e. Whip Antennas – Fiberglass is used to construct whip antennas that are very nearly maintenance free. These antennas have an inner conducting core that serves as the radiating element.

f. Nylon life nets are replacing metallic life nets in all applications except those areas subjected to missile blasts or jet exhausts. Figure 10-45 shows a nylon web safety net aboard ship.

g. Flagboxes, stanchions, and utility boxes are being constructed of nonmetallic material.

10.6 TREATMENT OF MISCELLANEOUS METAL OBJECTS IN TOPSIDE AREAS

Loose metal items in intermittent contact, when exposed to strong rf fields, are prolific sources of interference to nearby receivers. The interference can be broadband, when arcing occurs at the metallic junctions, or it can be discrete, frequency-type interference if the junction is nonlinear and is illuminated by two or more rf signals.

Whenever possible, all of these junctions should be eliminated by (1) installation of a bond strap, (2) isolation (insulation) of the item from contact with the ship hull, or (3) stowage of the item in below-deck areas.

10.7 INSULATION OF SMALL ITEMS

Many of the smaller metallic items such as dog wrenches, fire extinguishers, fog and fire nozzles, do

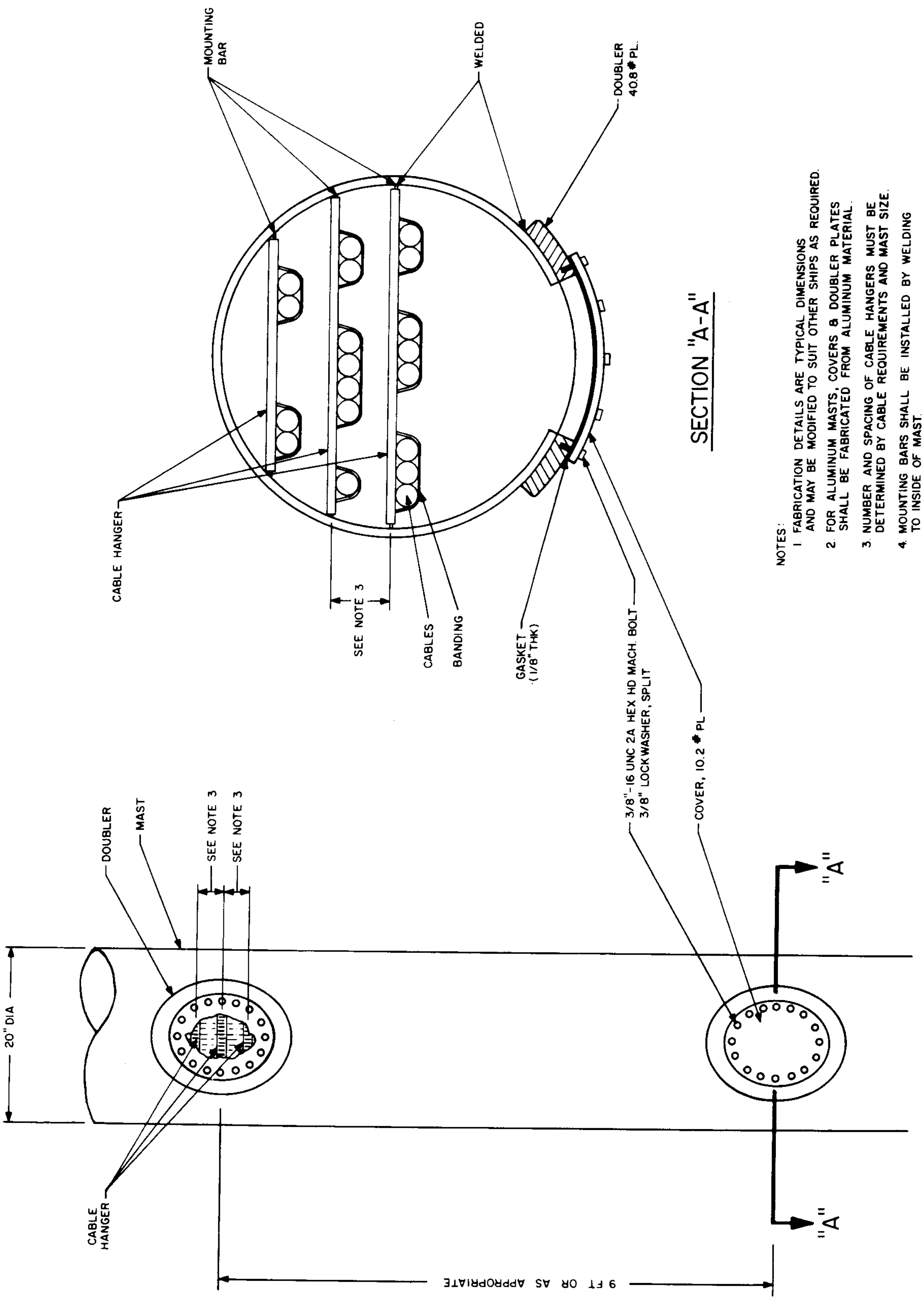


Figure 10-41. Mast Cables Located Within Mast, Typical

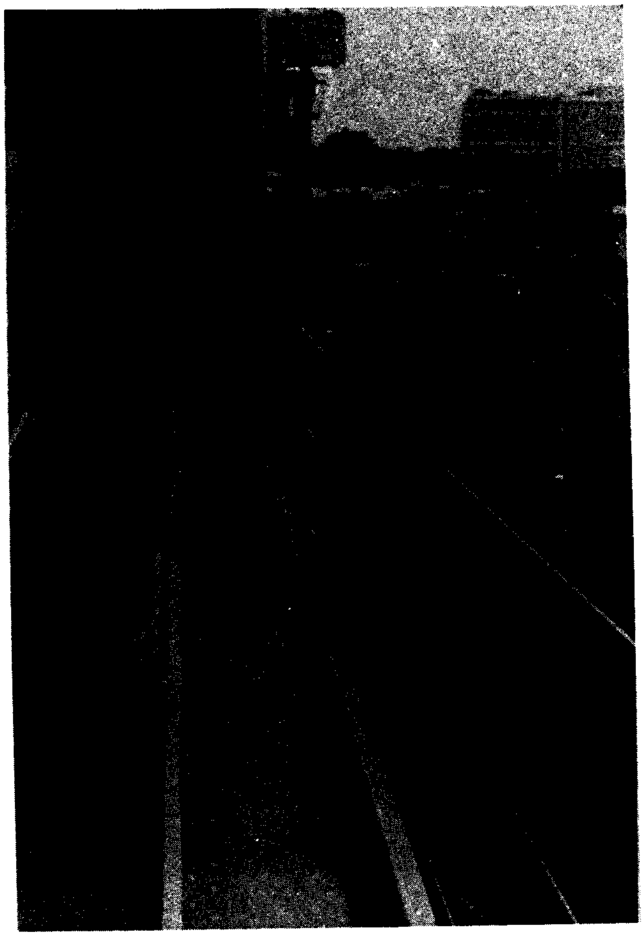


Figure 10-42. Nonmetallic Lifelines in Use Aboard Ship

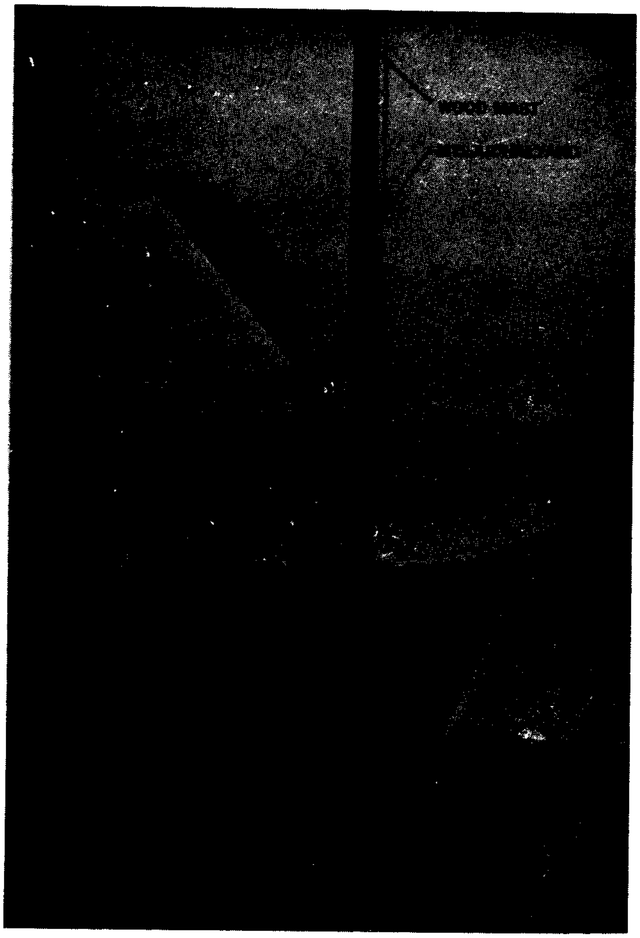


Figure 10-43. Wood Flag Mast in Use Aboard Ship

Figure 10-44. Wood Flag Mast in Use Aboard Ship

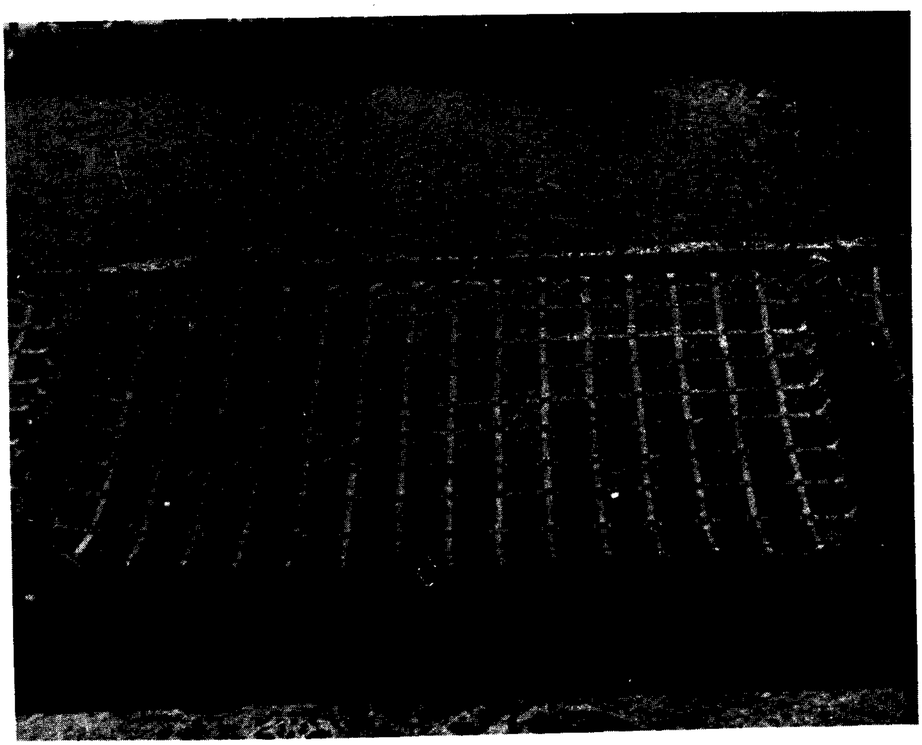


Figure 10-45. Nylon Web Safety Net

not lend themselves to standard bonding techniques. In some instances, the bond strap would be almost as large as the item to be bonded.

The preferred method is to isolate these objects from the ship hull to prevent corrosion and nonlinear junction formation. The isolation can be accomplished by the use of heat-shrinkable tubing on the mounting brackets of such items. The point to keep in mind is that the item has to be electrically isolated from the ship hull by an insulating material.

Figure 10-46 illustrates the use of rubber hose and pads to isolate gas bottles from the ship hull.

10.8 COMPUTER PREDICTION FOR FREQUENCY ASSIGNMENTS

A new technique, used most effectively on the AGMR class communication ships, is the use of a computer to assist in making frequency assignments. The computer determines the best transmit frequency for time of day, distance to be covered, geographical location, and weather conditions. This information is updated continuously by inputs from the National Bureau of Standards.

In addition, when the desired transmit frequencies are inserted, the computer will predict all possible interference products from harmonics and intermodulation mixes. Any desired width "clear" segment in the HF band can be predicted, and "receive" frequencies quickly assigned in these segments.

10.9 USE OF MULTIPLEX TO AVOID MULTIPLE FREQUENCY ASSIGNMENTS

Each time a new transmit frequency is added to an existing transmitter complex, the number of possible intermodulation interference products increase disproportionally to the number of transmitters energized. Since it is almost impossible to eliminate the nonlinearities causing the intermodulation products, the alternate solution is to reduce the number of transmitters radiating simultaneously.

A method is available whereby many channels (commonly 16 channels in shipboard HF circuits) of information can be transmitted via one carrier frequency. This method is called multiplexing or "MUX." Multiplexing is the art of combining two or more information channels into one composite channel or trunk for the simultaneous transmission of two or more messages over a single transmitter. The receiving station utilizes terminal equipment to "DEMUX" the signals into individual channels.

Multichannel broadcasts and ship/shore terminations utilize frequency multiplexing whereby each channel of the composite tone package of the broadcast or termination is assigned an audio frequency.

10.10 BLANKING DEVICES

Another method for reducing the effects of interference is by cutting off or blanking the susceptible equipment during the time of the interference. This is very effective for pulse type interference but is not applicable for CW or continuous interference sources.

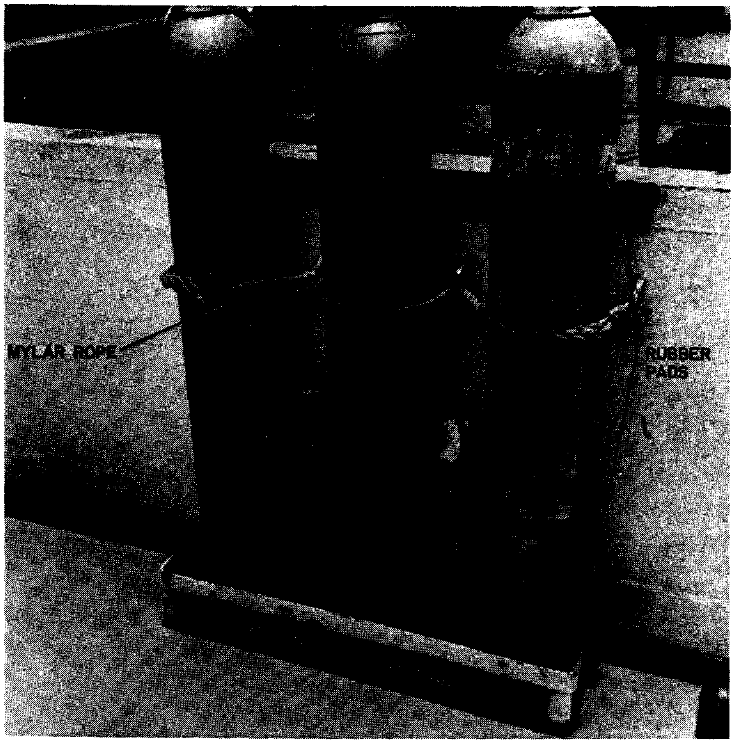
10.10.1 Shipboard ECM Blankers

The interfering pulse can be used to generate a blanking pulse or, if the interference is periodic, as in radar pulses, a pretrigger can be taken from the interfering equipment and used to develop a blanking pulse. An example of pulse blanking is the method used to disable shipboard ECM equipment when own ship radars are pulsed. A pretrigger from each radar, fed into a central pulse blanking device, generates blanking pulses which are used to turn off the ECM display. These "holes" in the display caused by blanking are not nearly as objectionable as the interference would be. If the blanking time is short, no significant detail is lost due to blanking.

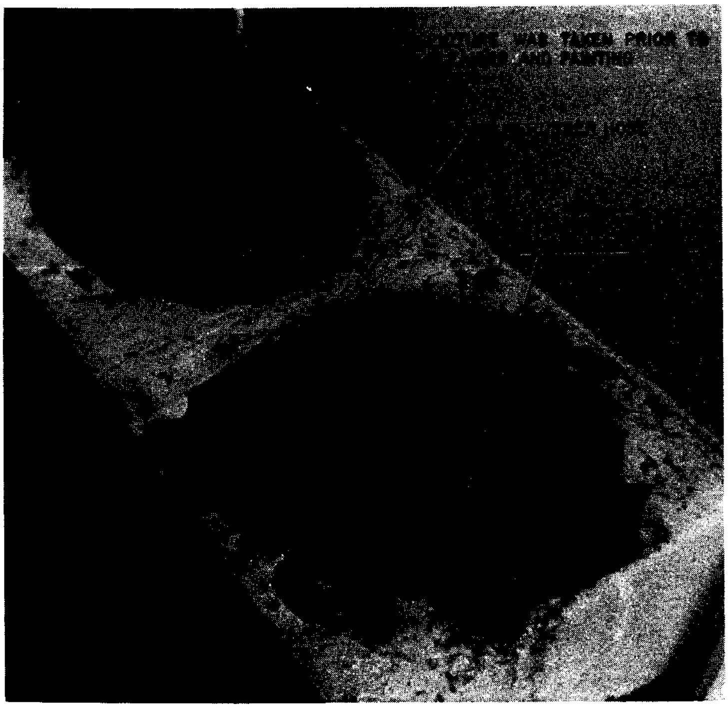
10.10.2 Mobile Noise Blankers

Another example of blanking is the pulse blanker used with mobile radios to eliminate ignition pulses and other high amplitude transients. Ignition interference is broadband in nature and peaks at about 40 MHz. A fixed tuned receiver tuned to this frequency is used to generate blanking waveforms from ignition pulses and to blank the companion receiver during this interval.

Blankers are available for almost any application. They are usually auxiliary devices that can be adapted to almost any susceptible equipment.



a. Gas Bottle Isolation
Figure 10-46. Isolation Techniques



b. Gas Bottle Isolation
Figure 10-46. Isolation Techniques

SECTION 11 – SHIPBOARD ELECTROMAGNETIC INTERFERENCE TESTS

11.1 GENERAL REQUIREMENTS

An electromagnetic interference survey should be conducted, upon completion of a regular overhaul, or other work of similar or greater extent, to assure that performance of the ship's electronic systems is degraded as little as possible by electromagnetic interference. Some electromagnetic interference is certain to be present simply because of the large number of high-powered transmitters and sensitive receivers, but this interference must be reduced to an acceptable level. Careful planning is essential for conducting a comprehensive EMI survey to avoid lost time due to conflicts between survey requirements and other shipboard activities. An EMI survey can be needlessly time consuming and may be inconclusive if the activities of all participants are not carefully scheduled and coordinated.

Electromagnetic interference is difficult to define quantitatively because there are so many variables involved. A signal of a certain frequency and intensity may be a serious problem in some receivers on one ship, while the same signal could be just a minor nuisance on another ship. Each case must be evaluated from the results of quantitative measurements. On the other hand, qualitative definitions are of some help in discussions and data taking as a way of broadly indicating the degree of interference. Definitions for mild, medium, and severe interference are included in NAVSHIPS Instruction 9671.25A.

11.2 DETAILED INSTRUCTIONS

NAVSHIPS Instruction 9671.25A was issued to specify standard procedures for conducting and reporting the results of surface ship EMI surveys. Detailed test procedures and reporting requirements are given along with directions for selection of monitoring receivers, preparation of the ship for testing, and other factors that lead to a comprehensive, economical test. NAVSHIPS Instruction 9671.25A is an essential reference document for anyone conducting a shipboard EMI survey.

11.3 TEST PHILOSOPHY

The test philosophy of NAVSHIPS Instruction 9671.25A is based on techniques used and results obtained during years of testing for electromagnetic interference aboard ships. This valuable experience has been applied to improving test procedures and to reducing testing time by specifying tests only between categories of equipment that are likely to interfere with each other. The instruction contains a table listing categories of potential sources of interference versus the equipments that should be monitored for interference.

The overall test procedure is divided into two phases. During Phase I, which is normally conducted in port, individual electrical devices are tested to determine, if they are likely to cause interference to an operating electronic system. Many devices are eliminated as potential interference sources during Phase I and therefore require no testing during Phase II, which is conducted at sea. Consequently, the at-sea time needed for testing electrical devices is reduced to a minimum.

During Phase II, the effects of self-generated electromagnetic energy upon the electrical and electronic systems of the ship are determined. The ship must be underway and in its normal operating configuration with all boats and normal deck equipment secured in place. The basic requirement for the test area is an ambient noise level no greater than 25 microvolts/meter/5kHz from 2 to 25 MHz; however, certain kinds of noise exceeding this limit can be accepted. NAVSHIPS Instruction 9671.25A specifies certain allowable deviations from the desired ambient noise level. Above the HF band, the background noise from man-made devices and natural sources falls off rapidly and ambient noise measurements are not required.

Past EMI survey experience also has highlighted the value of a standard report format. Such a standard reporting format is specified in NAVSHIPS Instruction 9671.25A. A uniform reporting procedure, with all reports in the same format, simplifies evaluation of the interference problems reported. The instruction emphasizes the principal objective of presenting sufficient detailed data and descriptive information to completely define each interference condition detected. The report must contain all the information needed to make decisions concerning corrective action.

SECTION 12 – OPERATING PRACTICES FOR EMI REDUCTION

The best efforts of equipment designers can be negated by unindoctrinated operators when the equipment is installed aboard ship. Present trends are to automate more fully and thus eliminate the human error factor from equipment operation. Yet, it is extremely difficult to remove all operator influence since present technology still depends on skilled operators to manipulate the sophisticated electronic systems aboard today's ships. Conscientious operators can do much to enhance communications by fully utilizing radar and weapons systems capabilities. Following are some pertinent comments directed towards improvement.

12.1 TUNING/LOADING PROCEDURES FOR COMMUNICATIONS TRANSMITTERS

An ideal transmitter would radiate only the signals necessary for the desired communication. This implies that the emitted spectrum and output amplitude would be the minimum necessary to convey the intelligence over the distance to be covered.

A significant portion of spurious emissions often result from improper transmitter tuning procedures. Close adherence to transmitter tuning instructions enhances shipboard communications operations by ensuring the radiation of maximum power at the transmitter fundamental. This in turn ensures that minimum power will be transmitted at harmonic and spurious frequencies.

A very good method for observing the effects of tuning/loading is the use of a spectrum analyzer to sample transmitter output. Effects of tuning on harmonic levels and other spurious outputs can be easily determined, and transmitter intermodulation products can be detected. Along with correct tuning/loading procedures, it is essential to utilize only enough radiated power to communicate satisfactorily over the distance to be covered. It is very poor operating technique to utilize full transmitter power for short range communications.

Different transmitters have different tuning procedures and the best way to ensure correct tuning is to become thoroughly familiar with these procedures by reference to appropriate equipment manuals and tuning procedures posted on each transmitter. The following paragraphs include the proper tuning and adjustment procedures for certain transmitters that are often found to be tuned and adjusted incorrectly.

12.1.1 Single Sideband and Independent Sideband Operational Considerations

The introduction of single sideband transmission to the fleet resulted in revolutionary new types of transmitter and transceiver equipments. Examples of the new type units are: AN/URC-32, -35, -58; AN/URT-23, -24; AN/WRC-1 and AN/WRT-2.

Single sideband transmitters use linear amplifiers for developing power rather than the Class C amplifiers which were common to the older units. The linear amplifier places a special requirement on SSB transmitters. Linear power generating devices are limited by their peak-power-handling capability. However, power meters on most equipments are calibrated to read average power. The AN/URT-23 is an exception. Its meter is calibrated to read peak-envelope-power. The average undistorted power output of a single sideband transmitter is limited by the type of signal being amplified and by the characteristics of the antenna with which it is being used. That is, it is limited by the peak-to-average power ratio of the modulation envelope and by the voltage-standing-wave-ratio (VSWR) of the antenna. In the case of independent sideband (ISB) transmitters, the combined modulation in both sidebands determines the ratio of peak-to-average power.

Figure 12-1 shows the average power that corresponds to 1000 watts PEP for various numbers of tones. Two curves are obtained by changing the percentage of the rated PEP. When the PEP exceeds the 1000-watt limit for a percentage of the time (one percent), the average power that can be transmitted, increases, especially at the higher numbers of modulating tones. However, this increase in average power is primarily an increase in noise, rather than in useful tone power as will be seen in the next paragraph.

12.1.2. Transmitter Overmodulation

Consider what happens to the transmitter output when modulated by a multichannel RATT signal. If a single-sideband transmitter is modulated to the limit of its peak power capability, the signal will be at least 35 dB above the noise that the transmitter is also generating. This condition corresponds to the transmitter putting out about 120 watts of average power. If the transmitter output is doubled by increasing the audio level of the multiplexed tones, the signal spectrum power is increased by 3 dB and the noise power is increased by 25 dB or by a power ratio

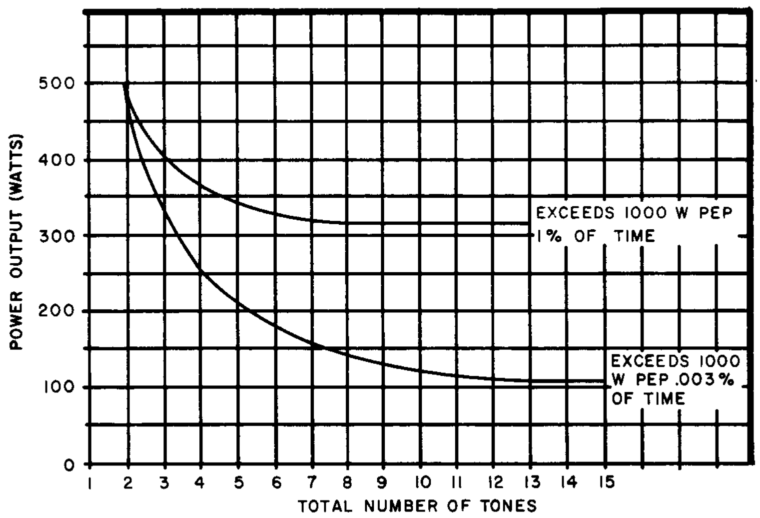


Figure 12-1. Average Power Output When Peak Power Levels Are Exceeded

of 320. When the transmitter power output is increased to its maximum capability of 320 watts, the noise power is again increased by a factor of about 320 *with no noticeable change in the power of the signal.*

This increase of noise output means that the transmitter is no longer radiating a narrow signal of 3 kHz with only some small additional noise within 10 kHz of the assigned frequency. The transmitter will now emit significant noise power hundreds of kHz from the assigned frequency and will radiate very large harmonic signals. If several transmitters are keyed simultaneously, intermodulation products will be generated throughout the spectrum.

The same considerations that apply to multi-channel RATT circuits also apply to voice circuits because the transmitter is required to handle the same peak-to-average-power ratios with voice signals. Table 12-1 is a list of commonly used HF equipments in the fleet.

Regardless of the problems associated with single-sideband equipment, the performance of single-sideband circuits is much better than that of standard AM circuits. Single-sideband advantages include:

a. Definite power advantages.

b. Elimination of heterodyning between carrier frequencies.

c. The availability of receivers with superior noise-eliminating features.

(Single-sideband voice must lie within narrow frequency tolerance to be usable.)

When an AN/WRT-2 transmitter is used, the power levels listed below are considered to be the maximum for the transmitter as indicated by the power output meter:

CW:	500 watts
FSK:	400 watts when shifting, 500 watts for steady/mark or space.
SSB:Voice:	0 watts unmodulated, 20 to 200 modulated.
AM Phone:	300 watts unmodulated, 300 to 400 modulated.
16-Channel Multiplex:	64 to 100 watts.
8-Channel Multiplex:	125 to 175 watts.

In virtually all cases, the tuning procedure of the technical manual should be followed and the above limits used as a check. Frequently the VSWR of an antenna and aging of transmitter output tubes will cause the power output to be substantially less than the maximum values given above. Such lower values may be the consequence even though proper tuning procedures were followed.

Tests have proven that voice transmissions are distorted to various degrees by overmodulation. This

Table 12-1. High Frequency SSB Equipments Most Commonly Used in the Fleet

	NOMENCLATURE	FUNCTION	FREQ. RANGE (MHz)	MODES OF OPERATION	P. E. P. Watts	AVG. PWR. Watts	TYPE TUNING INCREMENTS (kHZ)	SYN. ?	EXT. FREQ. STD. (MHz)	AFTS TONE (Hz)*	COMMENTS
TRANSMITTERS	AN/WRT-2	XMIT	2 - 30	CW, SSB, ISB, AFTS, AM, MULTI-CHAN. RATT	1000	500	1.0	YES	3.0	0	425 HZ tone in USB & LSB for AFTS.
	AN/URT-23 Series	XMIT	2 - 30	CW, SSB, ISB, AFTS, AM, MULTI-CHAN. RATT	1000	1000	0.5/or 0.1 with latest model	YES	5.0	2000/ 2550*	Companion Antenna Coupler is AN/URA-38.
	AN/FRT-39B	XMIT	2 - 28	CW*, SSB, ISB, AFTS, AM, MULTI-CHAN. RATT	10,000	5000	0.1	YES	1.0	2000/ 2550*	
	AN/URT-24	XMIT	2 - 30	CW, SSB, ISB, AFTS, AM	100	50	0.1	YES	5.0	2000/ 2550*	
RECEIVERS	R-390A & CV-591A	RCVE	0.5- 32	CW, SSB, AM, AFTS			CONTINUOUS	NO	NONE		Geneal Purpose Rcvr. Not used for MULTI-CHANNEL RATT
	AN/WRR-2 (FRR-59) Series	RCVE	2 - 32	CW, SSB, ISB, AM, AFTS MULTI-CHAN. RATT			1.0/or 0.5 with FLD CHANGE + CONTIN-UOUS	YES	1.0		
	R-1051/ URR . Series	RCVE	2 - 30	CW, SSB, ISB, AM, AFTS MULTI-CHAN. RATT			0.5/ or 0.1 with latest model + CONTINUOUS	YES	5.0		
TRANSCEIVERS	AN/URC-32 Series	TRANS-CVE	2 - 30	CW**, SSB, ISB, AM, AFTS, MULTI-CHAN RATT.	500	500	1.0/ or 0.1 with FLD CHANGE	YES	0.1	2000	
	AN/URC-35	TRANS-CVE	2 - 30	CW, SSB, AM	100	50	0.1 + CONTIN-UOUS	YES	5.0	NONE	
	AN/URC-58	TRANS-CVE	2 - 15	CW*, AM, SSB	100	100	1.0 + CONTIN-UOUS	YES	NONE	NONE	
TRANSMITTER/ RECEIVER	AN/WRC-1 Series	XMIT & RCVE	2 - 30	CW, SSB, ISB, AM, AFTS	100	50	0.5/ or 0.1 with latest model + CONTINUOUS ON RECEIVE	YES	5.0	2000/ 2550	

* CW tone is 1000 Hz in the upper sideband
** CW tone is 1000 Hz or 1500 cps selectable

* = Center Freq. — Tones are ± 425 Hz.

was more prevalent when single-sideband transmitters were used in the sideband-plus-carrier mode than when used in the single-sideband suppressed carrier mode. Thirty percent of the stations on one amplitude modulation net participating in these tests were overmodulating to the point at which their transmission quality was seriously degraded. Nearly all the transmitters on this net were identified as SSB type by the characteristic absence of one sideband on the spectrum analyzer display. Degradation resulted from the improper adjustment of the carrier reinsert.

Usually when a transmitter voice circuit sounds as if the operator's speech were impaired, the sound distortion is due to overmodulation. Overmodulation can result from causes other than the improper setup of the transmitter. A major cause is the inability of the transmitter to compensate for different voice (modulation) input levels applied to the transmitters. This inability may be due, in turn, to differences in voice handset output levels or speech levels used by operators.

12.1.3 Transmitting Antennas

A rigorous antenna system maintenance schedule is essential in preserving reliable communications systems. Insulators should be cleaned weekly. Leakage tests and component inspections should be made monthly. Transmitter maintenance procedures, inspections, and test schedules are very nearly the same as those used for receiving antennas. It is therefore essential that applicable tests, inspections, cleaning schedules, and recommendations specified for receiving antennas also be performed for transmitting antenna systems. Complete inspection and testing details for specific transmitter antenna types are provided in NAVSHIPS 0967-177-3020, Shipboard Antenna Details, Volume II, and should be closely followed to maintain reliable communications. This NAVSHIPS publication also specified proper corrosion inhibitors, preservatives, and waterproofing materials required for a sound maintenance program.

Dried salt deposits act as insulators; moist salt deposits become conductive. For this reason, the antennas should be checked with a Megger during humid or damp conditions.

12.2 USE OF MULTICOUPLERS

The use of multicouplers is fast becoming almost mandatory with today's crowded frequency spectrum. Multicouplers are basically designed to permit several transmitters, or receivers, or combinations of both to operate simultaneously from a common antenna. A secondary benefit realized from use of these devices, almost as important as that for which they were designed, is their EMI reduction characteristics. Some of the advantages gained by use of multicouplers are:

a. Reduced exposure of transmitter output stages to off-frequency signals with an attendant reduction of transmitter-generated intermodulation products.

b. One transmit antenna can be used with several transmitters, which reduces the number of transmit antennas required. Better radiation patterns can then be realized from the smaller number of antennas.

c. Spurious outputs separated in frequency from the fundamental are attenuated which helps reduce the clutter of undesired signals radiated.

d. Less frequency separation between transmit-to-receive, transmit-to-transmit or receive-to-receive frequency assignments. Without multicouplers, it is necessary for receivers to maintain a minimum frequency separation of at least 15 percent from same-ship transmitter fundamental frequencies and a minimum of five percent from same-ship transmitter harmonics. When tunable receiver multicouplers, such as the AN/SRA-38 (2 to 6 MHz), AN/SRA-39 (4 to 12 MHz), and AN/SRA-40 (10 to 30 MHz) are used, receivers need be separated only five percent from same-ship transmitter fundamental frequencies.

There is a power loss in multicouplers, significant in some cases, and the additional tuning requirements add to overall complexity but these are acceptable trade-offs to obtain the advantages listed above.

Table 12-2 shows the frequency separation that should be maintained between same-ship transmitters and between same-ship receivers and transmitters when various multicouplers are employed.

Figure 12-2 illustrates the attenuation in dB versus frequency characteristics of the four tunable couplers in HF Antenna Coupler Group AN/SRA-34(V).

The multicoupler must be considered as an integral component of any antenna system in which it has been installed and should never be bypassed except in emergencies such as multicoupler malfunction. Multicouplers must be periodically inspected as a matter of routine maintenance. Due to the power levels involved, transmitter multicouplers are subject to arcing and electrical breakdowns. Coupler units should be frequently checked for arc over, insulator and circuit component conditions, and to ensure that the plating on coils is complete and unblemished.

Improper tuning of multicouplers, especially those associated with transmitting systems, results in

Table 12-2. Frequency Separations Specified for Various Multicouplers

Multicoupler System	Multicoupler	Pair	Frequency Separation
Receiving	AN/SRA-12 -12A	TxRx	*
Transmitting	AN/SRA-13 -14 -15 -16	TxTx	10%
Transmitting	AN/SRA-23	TxTx	10%
Transmitting/ receiving	AN/SRA-34 AN/SRA-34	TxTx TxRx	5% 15%
Transmitting	AN/SRA-35 -36 -37	TxTx	5%
Receiving	AN/SRA-38 -39 -40	TxRx	5%
Receiving	AN/SRA-44 -45 -46	TxRx	10%

*Receiver must be out-of-filter band occupied by transmitter.

intolerable increases in intermodulation products and harmonic levels. Precise tuning is a "must." Technical manual instructions should be closely followed to ensure proper tuning is accomplished each time multicoupler retuning or adjustments are required.

12.3 RECEIVER OPERATING TECHNIQUES FOR EMI REDUCTION

Present day receivers are very sophisticated devices compared to receivers of just a few years ago. Nevertheless, they are still subject to some of the same limitations as older receivers. Skilled operators can do much to enhance operation of either type.

12.3.1 Overload in Receiver Front-Ends

A hard-to-recognize type of interference occurs when a sensitive receiver is exposed to a crowded, high energy spectrum. There is a limit to the amount of signal amplitude an amplifying stage can linearly process; signals stronger than this create unwanted harmonics and intermodulation products due to nonlinearity of the overloaded stage. The main difficulty from this kind of interference is the inability of operators to detect its presence, since it is almost impossible, from an operator standpoint, to determine the source of an interfering signal.

When an intense off-frequency rf signal (normally from an intraship transmitter) penetrates to the first active stage in a typical receiver, intermodulation with the desired signal may occur. However, presence of the strong signal will not be indicated on the carrier level meter since the meter normally monitors AGC voltage. Note that penetration of a strong signal is independent of, and bears no relationship to, receiver tuned frequency. The 40 to 60 dB off-frequency signal attenuation of present preselectors is not sufficient to exclude intense off-frequency signals from the rf amplifier stage.

Even when it can be determined that an interfering signal is generated in the receiver front end, the question arises as to how to eliminate it. An acceptable method at present is to insert a tunable bandpass filter immediately ahead of the receiver. This filter gives an additional 70 dB, or so, attenuation to out-of-band signals. The combined attenuation of the bandpass filter and receiver preselector usually serve to eliminate interference caused by fundamental overload from own ship transmitters.

An alternate practice is to physically separate receive and transmit antennas to reduce the cross-coupling between them. Additional space attenuation provided by increased separation will serve to attenuate strong fundamental signals with the same results as the filter method. Physical separation is increasingly difficult to achieve due to space limitations imposed by priority demands on topside surface area aboard ships.

A simple, and in most cases reliable, method to determine if a receiver is overloaded is presently in use by field personnel. Known dB amounts of attenuation are inserted ahead of the receiver and the

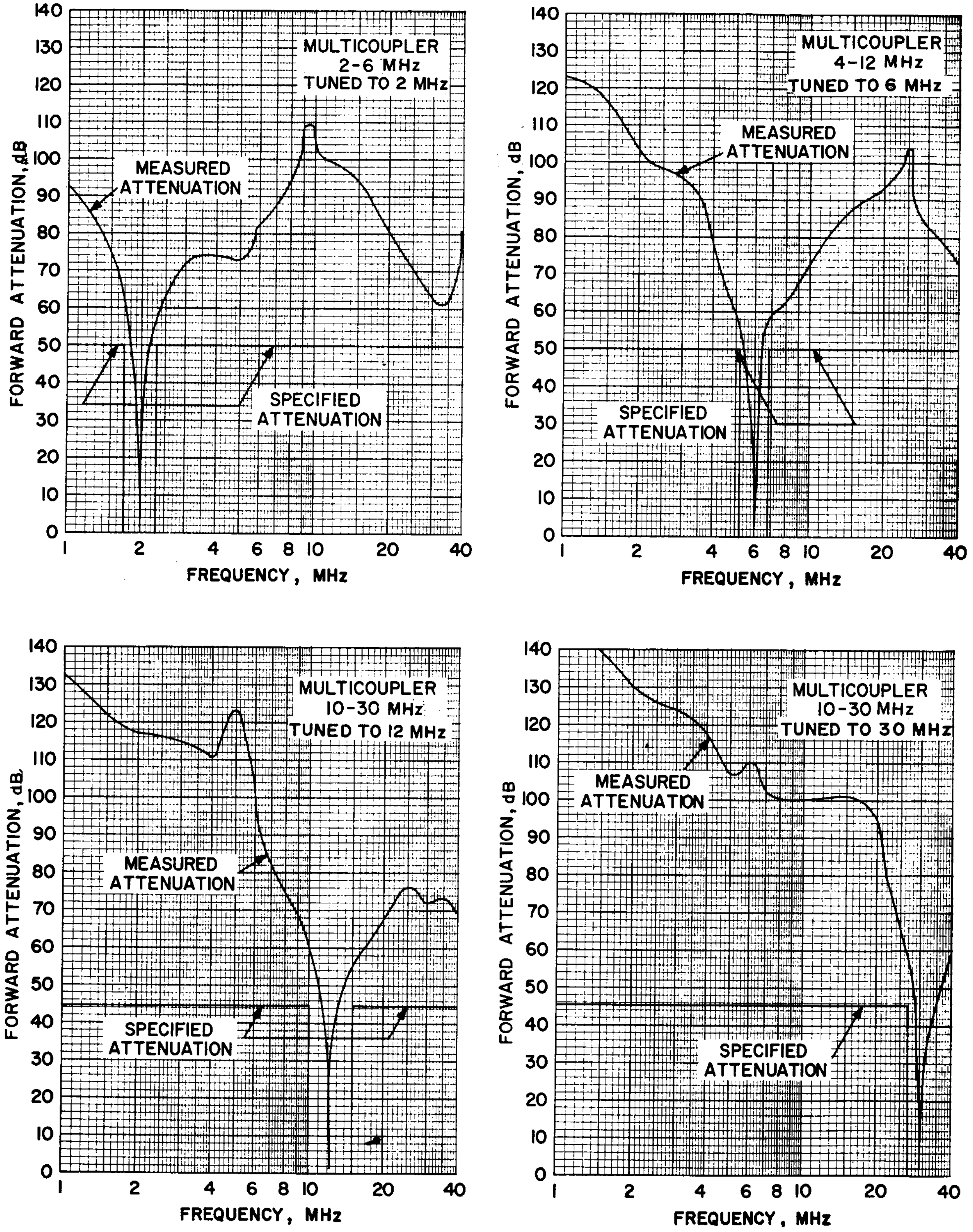


Figure 12-2. AN/SRA-34(V) HF Antenna Group Selectivity Curves

same dB decrease is observed in carrier level meter reading. The procedure follows:

a. Adjust the receiver for normal reception of the desired signal and note the indicated dB level on the CARRIER meter.

b. Insert an attenuator of known value in the receiver transmission line and note the drop in signal level.

c. If the meter reading is reduced precisely by the value of the attenuator, the receiver is experiencing no overload. If the level is reduced by an amount either less or greater than the value of the attenuator, the receiver is operating in an overloaded condition.

d. In the event an overload condition is detected, repeat the test with reduced RF gain settings until a linear relationship exists between the value of the attenuator inserted and the drop in the indicated meter reading. For example, insertion of a 10-dB attenuator should provide a 10-dB reduction in signal level.

Lacking the necessary time or attenuator for the above test, an operator should maintain the carrier level meter indication between 20 and 60 dB by use of the rf gain control. This will help to minimize the chance of overloading the receiver.

12.3.2 Choice of Receiver Bandwidth

Another, mostly overlooked, consideration in operating techniques to reduce the effects of EMI is judicious selection of bandwidth on receivers provided with this feature. The narrowest bandwidth that will accommodate the intelligence content in the desired signal should be selected. Narrow bandwidths minimize the chance of degradation of receiver operation by closely spaced undesired signals.

12.3.3 Maintenance of Receiver to Meet POMSEE Standards

It should be self-evident that a properly aligned and maintained receiver will perform better, especially in a high density environment, than a poorly aligned receiver of reduced sensitivity. The designed-in sensitivity and undesired signal rejection characteristics of receivers can be realized only when the equipments are satisfactorily maintained by competent technicians.

This presupposes that operators learn to interpret receiver performance in order to recognize and report a malfunctioning receiver.

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