HANDBOOK

ON

Radio Frequency Interference

VOLUME 3

METHODS OF ELECTROMAGNETIC INTERFERENCE-FREE DESIGN AND INTERFERENCE SUPPRESSION

PREFACE

This Volume on Methods of Electromagnetic Interference-Free Design and Interference Suppression is the third in our series of four Volumes on Radio Frequency Interference. In it we discuss specific interference reduction and suppression practices. System design is discussed in the first part of the Volume and in later parts of the Volume, attention is focussed on components and auxiliary devices. Some general aspects of the control of interference paths are discussed in the last Chapter. The Appendices present information on the details of filter design, and construction of bonds.

Since control of interference and the design of interference-free systems depend to a large extent upon the ability of instruments to measure electromagnetic energy precisely, it is necessary that their calibration be accurate. Calibration methods are, therefore, discussed in Appendix V.

Solutions of electronic compatibility and interference problems, though difficult, are not impossible. Good design practices are the key to interference suppression. Electromagnetic compatibility must be given as much consideration in initial design as other important considerations of reliability, size, weight, and similar parameters.

Designing for electromagnetic compatibility is not an inexpensive matter. It may be very costly in time and money. On the other hand, the alternatives -- poor performance of electronic equipment, or perhaps serious degradation of weapon systems — may not be acceptable. In fact, in some cases, the acceptance of these alternatives may result in complete abandonment of the entire system, or, as a minimum, a considerable reduction in system effectiveness.

But it is not alone the cost in time and money which is important - in this modern era, lives of thousands of individuals may depend on accurate and complete transmission of intelligence by electromagnetic devices. Only with thorough compatibility design can communication-electronic systems completely attain the effectiveness of operation necessary to meet the requirements of today.

I am grateful to Empire Devices, Inc.; Polarad Electronics Corporation; and Stoddart Aircraft Radio, Inc., for permission to use material concerning their interference measuring instruments.

I wish, also, to express my appreciation to A. H. Sullivan, Jr. as Editor of the Handbook; to J. A. Hopkins, as Editorial Assistant; to Robert Latorre for compilation of basic material; to L. Q. Fleenor T. H. Miller, J. L. Wibbe and H. D. Zink for considerable technical contributions in the preparation of the text; to R. E. Frickey, R. E. Hannah, D. Sabalos, H. Shaver, and R. E. Steinebach, Jr. for their careful and critical review of the final material; to H. M. Humbertson, R. F. Claxton, M. W. Tilly, and D. L. Barnes for production work; to R. Frederick who was responsible for the art work; and to Frances Zello, Marjorie Brown and Mary Schmidt for their efforts in typing the difficult copy.

Carl L. Frederick, Sr.
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HANDBOOK

ON

RADIO FREQUENCY INTERFERENCE

VOLUME 3

METHODS OF ELECTROMAGNETIC INTERFERENCE-FREE DESIGN AND INTERFERENCE SUPPRESSION

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DESIGN METHODS FOR INTERFERENCE-FREE SYSTEMS

1. INTERFERENCE CONSIDERATIONS IN COMMUNICATION-ELECTRONIC SYSTEMS

1.1 GENERAL

Individual components must be designed both for minimum generation of, and for minimum susceptibility to, radio interference. It is important to keep in mind that the component must perform satisfactorily not only alone but also in conjunction with other components as part of a system. Similarly, each system must be designed for interference-free operation not only when operating alone, but also when operating simultaneously with all other systems and their components.

In the design of equipment and in the layout of a system for interference-free operation, practical considerations will greatly influence the designer's decision as to which suppression technique will be applied to each interference problem as it arises. Factors other than technical which may influence the designer are as follows:

a. Weight

It is of the utmost importance to apply techniques which effectively provide the highest ratio of attenuation per unit of weight. It is the designer's responsibility, therefore, to study the interference problem in order to insure minimum weight consistent with interference levels which can be tolerated. The most effective suppression techniques are available during the initial design stages of components or systems. Once a design is completed, the designer is limited to techniques which are less effective, often resulting in corrective modifications and additions involving added weight. In many cases proper design may result in elimination, rather than suppression, of interference. In any event, it is obvious that the best control of radio interference will result if it is considered early in every stage of component or system design.

b. Space

The use of proper design techniques in designing components and systems from the standpoint of radio interference will minimize the size of components, the number of filter networks required, and the shields required in installations. This is particularly important in systems where high performance is essential. To achieve high performance all dimensions are critical and components and systems must be designed to fit the space available. This does not mean that radio interference can be ignored. Rather it requires better design practices in all phases of design.

c. Materials

Often the designer is influenced by availability of materials in the design of suppression devices to attenuate or eliminate radio interference. For example, inductances obtainable for filters in high current circuits are limited by the direct current resistance of the winding. Filter capacitors, especially those of high values, may change considerably at high temperatures. As a result of the use of materials of construction which will not "stay put" under operating conditions, the attenuation of radio interference may be permanently impaired, or if these changes vary rapidly with time, the change of impedance will actually result in additional sources of radio interference. These and similar factors must be borne in mind throughout the design of components and systems if the design is to be interference-free under all operating conditions.

Practical design must take into consideration all of the above, and the best compromise must be used. In addition to these factors the designer must choose the most efficient methods of eliminating or suppressing radio interference.

It may be shown that there are three places in which interference can be controlled: (1) at or in the source; (2) along the coupling path; and (3) at or in the receiver affected. Depending on the nature of the interference, the transmission path between source and receiver, and the receiver characteristics, the designer may find that there is a choice of several alternative solutions to control the interference. This permits the designer to choose the most effective method and to minimize the weight, size, and use of critical materials. The first and third are areas in which the designers of components and equipments have control. The designers of systems control the second.

Practically, a cooperative effort by all designers in all three areas is the only solution to the problem of radio interference. Since it is necessary to consider weight, size, and materials in the functional design of equipment as well as the problems of radio interference, the designer must choose a compromise which will permit satisfactory system operation. This implies radio noise suppression to a degree rather than complete elimination. This again is a practical compromise since there is no such thing as a realizable filter with infinite attentuation or a receiver with zero susceptibility to interference. Specifications have been written covering interference limits and receiver susceptibility. The limits were set as high as is believed to be tolerable in order to achieve what is considered interference-free operation.

These limits should never be exceeded and the designer should strive for as nearly interference-free operation as is practically possible. While the present limits seem to be satisfactory, they may at some future time be made more severe as the multiplicity of equipments increases. It is conceivable also that a particular installation of equipments, each of which is within specified limits, may interfere with each other because of some installation peculiarity.

In the design of any equipment, system, or installation, the designer should keep in mind the fact that it must operate satisfactorily in conjunction with all the other equipment which may be in the same system and which may operate at the same time. If this is considered at the inception of the design, it is possible to make most effective use of techniques for interference control. In many cases a study of various approaches to a problem yields one in which no appreciable weight need be added to achieve the required degree of suppression. In fact, proper design may even decrease the weight originally considered necessary, simply by careful routing of wires, correctly orienting parts or assemblies, shielding, or using correct design values in filter networks.

1.2 BASIC METHODS FOR SUPPRESSION OF RADIO INTERFERENCE

Basically there are three ways of combating an interfering signal to minimize: (1) its generation, (2) its transmission, and (3) its undesirable effect on the receiver.

To prevent the generation of interfering signals means to design all equipment in such a manner that no interference will be generated. This is obviously the ideal way of dealing with interference problems because, if no interference were generated, no further attention would have to be paid to the problem. In some cases, this can be achieved simply by a proper choice of components. For example, if there is a choice between two motors for a particular application, one with a commutator and one without, then all other things being equal, the one without a commutator should be chosen because this choice would eliminate all interference due to commutation. Or, in making a choice of a device to obtain alternating current from a direct current supply, vibrators should be avoided, because the arcing that accompanies their operation is a more serious source of interference than that associated with other types of inverters.

If the generation of interfering signals cannot be prevented entirely, as, for example, in those sources for which the generation of interference is inherent in their normal functions, good design may be able to minimize it. For example, in an ignition system the spark is essential and cannot be eliminated. But the exceedingly steep wave fronts which usually accompany the spark and are the primary cause of the strong interference fields may not always be essential to the proper operation of the ignition system. Here proper design may be able to prevent some of the interfering signals from being generated, though it can never eliminate all of them.

The prevention or minimizing of the generation of interfering signals is properly the task of the designer of components of electrical systems, though some prevention can also be achieved by the manufacturer in the assembly of systems.

Since it is never possible to prevent all interfering signals from being generated, the designer must try to keep them from reaching the receiver. This problem may be attacked at various points. The task of preventing the interfering signals from leaving the immediate vicinity of the source also belongs properly to the designer of components. For example, interference from a motor may be kept "bottled up" entirely within a metal housing, which is an integral part of the motor and, if carefully designed, may act as a complete shield. The task of preventing the transmission of interfering signals at points between components belongs to the designer of electrical systems and the engineer responsible for the installation of components and systems.

Finally, an effort must be made to prevent the interfering signals from affecting the receiver. Partly, this is a matter of preventing its transmission into the receiver or its reaching any sensitive circuits within the receiver. Partly, this is a matter of incorporating special circuits in the receiver which, in one way or another, reduce its nuisance value. The task of minimizing the effect of interference on the receiver belongs properly to the designer of receivers.

1.2.1 SUPPRESSION AT THE SOURCE

Suppression at the source has two basically different aspects: The prevention or minimizing of the generation of interfering signals, which is treated in this section, and the "bottling up" of the generated interfering signal within the source, which is a problem of transmission and will be treated in Section 1.2.3

Preventing or minimizing the generation of interference consists of preventing or minimizing all necessary variations of electromotive forces and impedances. This means that, for each piece of equipment, the designer must analyze the causes of variations of electromotive forces and impedances, eliminate those that are unessential to the proper operation, and reduce the essential ones to the absolute minimum. Further consideration of this aspect will be illustrated by the analysis of individual components in other sections of this Handbook. Two cases, however, involve considerations of a general type and will be taken up here: The prevention of radio interference by bonding and the suppression of arcs.

1.2.1.1 Bonding

Any two points of a metallic structure whether electrically connected or not, may develop a potential difference at some frequency. At those frequencies for which the dimensions of the structural member are of the order of magnitude of a wave length, such potential differences are impossible to avoid in the presence of an electric or magnetic field. At the lower frequencies, the circuit concept may be used to show that the potential difference between two points of the structure is proportional to the impedance between the same points. Reducing the impedance will therefore reduce the potential difference at all frequencies at which this impedance may be considered as one lumped element.

Between two points which are in an electromagnetic field and which are electrically insulated from each other, there will exist a comparatively strong electric, but weak magnetic field, the latter being caused by displacement currents only, which are negligible at frequencies below about 100 megacycles. When the two points are "bonded", i.e., connected through a path of low impedance, a conduction current will exist with which is associated a comparatively weak electric, but strong magnetic field. The conduction current with its magnetic field is much less important as an interference generator than the electric field between insulated points for the following reasons: (1) when the two points are insulated from each other, even a small amount of charge accumulated at the points may cause a large potential difference. When the points are connected permanently, no charge can accumulate and the resulting steady state current is usually negligible; (2) if the points are permanently bonded, the impedance between them is much more likely to be constant than if the points are separated by a distance which may vary with any mechanical shock or other random motion of the structure. Thus, bonding will eliminate the generation of interference caused by a varying impedance; (3) if the points are insulated from each other, the electric field between them may become large enough to cause an arc or spark discharge of the accumulated charge. Arcs and sparks are among the most serious sources of radio interference. This type of arcing is eliminated by proper bonding.

A truly low impedance path is possible only so long as the dimensions of the bonded members are small as compared to a wave length of the interfering signal. At high frequencies, the members must be considered as transmission lines whose impedance may be inductive or capacitive and have any magnitude whatever, depending on their geometrical shape and the frequency. Bonding in itself, therefore, does not assure the existence of a true "ground plane" in a system, i.e., the lack of an appreciable potential difference between any two points of the bonded members.

Bonding refers to the provision of a low impedance path not only between two points of the system, but also between one point of the system and a piece of equipment for which the system serves as ground. Poor bonding of this kind is a very frequent cause of interference though it is not ordinarily considered as a source. Rather, poor bonding keeps other measures of suppression, such as the insertion of filters and proper "grounding", from being effective.

1.2.1.2 Suppression of Arcs

Arcs occurring during switching and other processes, which do not perform any useful function, must be prevented entirely both because they are serious sources of radio interference and because they produce a rapid deterioration of the contacts. Devices which help to extinguish the arc once it is established, such as the mechanical insertion of a dielectric between the contacts or the placement of a strong magnet close to the gap, are not suitable for radio-interference reduction because shortening the duration of the arc does not reduce its effectiveness as a source of high-frequency disturbances. The only suitable methods are those which prevent the formation of the arc at the outset.

By far the most effective method of preventing arcing across the contacts of a switch is the insertion, in parallel with the switch, of a capacitance in series with a resistance. The purpose of the capacitor is to prevent the voltage across the contacts from building up sufficiently to produce arcing; the resistance helps to damp out any oscillations that may occur and provides a means of dissipating the energy stored in the magnetic field of the current that was flowing before the switch was opened. To understand these actions, consider the circuit of Figure 1-1. In this diagram, E is a direct-current voltage source. R_1 is the load resistor which is connected to the source when the switch, S, is closed. Since every circuit contains inductance, L cannot be zero, though it may be small. The quantity C_1 is the capacitance between the contacts of the switch when open, and R_2 and C_2 are the elements of the external network added for the purpose of the arc suppression. In general, C_1 is not constant since it depends on the position of the contacts. But if R_2 is not too large and C_2 is much larger than C_1, then C_1 may be neglected.

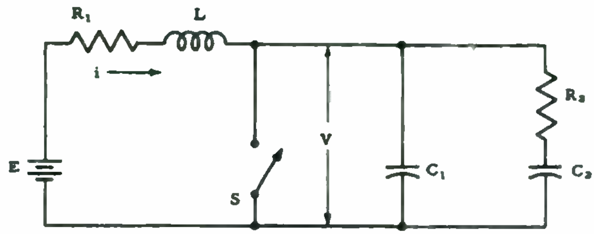


Figure 1-1. Switching Circuit with Arc-Suppressing Network

It is assumed that the switch, S, is initially closed so that a steady current $I = E/R_1$ flows through it and there is no voltage across either capacitor nor any current through R_2. If then, at time $t = 0$, the switch is suddenly opened, the question is: what happens to the voltage, V, across the switch?

Before answering this question, a simpler circuit will be considered: what happens to the voltage across the switch in the circuit of Figure 1-2, in which the external arc-suppressing network has been omitted? The integro-differential equation for this circuit, for all times subsequent to the opening of the switch, is as follows:

$$R_1 i + L \frac{di}{dt} + \frac{1}{C_1} \int_0^t i \, dt = E \tag{1-1}$$

where i is the instantaneous current, and the initial condition is that $i = I$ at $t = 0$. The solution of this equation may be written in the following form:

$$i = \left(I \cos \omega t + \frac{E}{2\omega L} \sin \omega t \right) e^{-\frac{R_1}{2L} t} \tag{1-2}$$

where: $\omega = \sqrt{(1/C_1 L) - (R_1/2L)^2}$

From this it follows that:

$$V = E + \left[\left(I\omega L - \frac{ER_1}{4\omega L} \right) \sin \omega t - E \cos \omega t \right] e^{-\frac{R_1}{2L} t} \tag{1-3}$$

Both i and V show damped sinusodial oscillations provided that ω is real. To estimate the maximum voltage that can build up across the switch, assume that R_1 is small so that there is little damping. Then the maximum value of V is approximately equal to $I \omega L = I \sqrt{L/C_1}$. Thus, if there were no capacitance across the contacts, the voltage would become infinite.

To return now to the circuit of Figure 1-1, let the capacitance C_1 be neglected since it is in parallel with C_2, assumed to be much larger than C_1. Then the circuit is similar to the simple one analyzed before with the capacitance C_2 substituted for C_1 and the total resistance $R_1 + R_2$ substituted for R_1. The major difference is that the voltage across the contacts is now the voltage across both R_2 and C_2 instead of being the voltage across the capacitance alone. The maxima of the voltages across R_2 and C_2 do not occur at the same time so that the maximum voltage across the switch is not equal to their sum. Nevertheless, in order to keep the contact voltage small, both these voltages must be small.

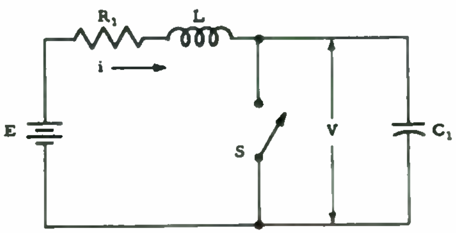


Figure 1-2. Simple Switching Circuit

Proceeding on the same basis as before, to keep the voltage across C_2 small for a given I and L, this capacitance should be as large as possible. In order to keep the voltage across R_2 small, this resistance should be as small as possible since the maximum value of this voltage is simply IR_2, using the same approximation as before. It may be noted here that for a value of R_2 equal to $\sqrt{L/C_2}$ the maximum voltage across C_2 and R_2 become equal.

On the other hand, to increase the damping, the total resistance of the circuit should be high. If weakly damped oscillations of large amplitudes are permitted to occur, they may produce as much radio interference as the arc that was suppressed. In fact, to prevent oscillations altogether, the total resistance should be sufficient for "critical damping", i. e., sufficient to reduce ω to zero or even make it imaginary. (A sine or cosine function with imaginary argument leads to an exponentially decaying function in this case.) The expression for ω shows that, for critical damping, the resistance should be equal to $2\sqrt{L/C_2}$, or just twice the value obtained before. If R_1 is very small, there could be no choice of R_2 that would satisfy both requirements, that of large damping as well as that of small contact voltage. Fortunately, in most practical cases, R_1 by itself is large enough to provide the damping, and a value of R_2 as little as 1/100 of the value required for critical damping may be a satisfactory choice.

The choice of C_2 is determined by the values of I and L, which must be known and can easily be measured, and by the values of the breakdown voltage between the contacts. The value of C_2 should be large enough to keep the maximum voltage, $I\sqrt{L/C_2}$, below that breakdown voltage by a wide margin of safety.

In determining the breakdown voltage between contacts, it must be remembered not only that tabulated values of dielectric strength (which is the ratio of breakdown voltage to the thickness of the dielectric between contacts) are very approximate and that the actual breakdown voltage is a function of many factors such as shape of the contact points, frequency and wave shape of the applied voltage, humidity, and temperature, but also that there is a definite and rapid decrease of the breakdown voltage with altitude due to the decrease of pressure and increase of ionization. The ratio of the breakdown voltage at any altitude to the breakdown voltage at sea level, other conditions being kept equal, is shown in Figure 1-3.

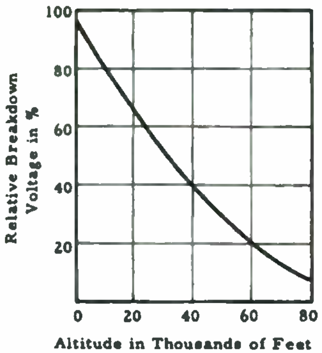


Figure 1-3. Effect of Altitude on Breakdown Voltage

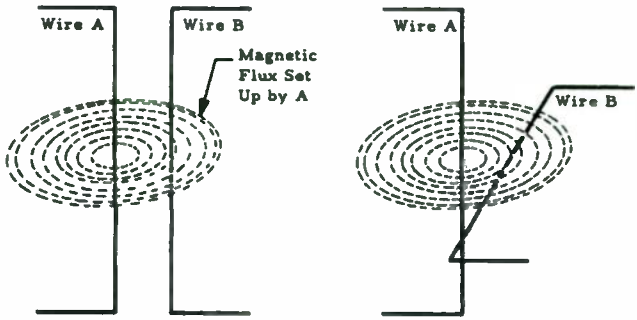
1.2.2 SUPPRESSION DURING TRANSMISSION

Once an interfering signal is generated, it must be prevented from reaching the receiver. Five methods are available for this: (1) location, (2) orientation, (3) grounding, (4) shielding, and (5) filtering.

The first two involve extremely simple principles, which can be stated in a few sentences. The other three require more detailed analysis and will be treated in later sections.

Prevention of the transmission of interfering signals by proper location simply means that equipment likely to generate interference or lead wires likely to carry interfering currents should be mounted or placed as far away as possible from all receivers and all power, control, input, or output leads connected to any receiver. It also means that all equipment should be placed so as to make the maximum use of natural metallic shielding. Because of the high sensitivity of modern receivers, because of the possibility of resonance excitation and because of the possibility of direct conduction, proper location cannot prevent all interfering signals from reaching and affecting the receiver, but it aids greatly in eliminating some and reducing most of the remaining interfering signals at the receiver. The proper location of equipment and wiring for minimum transmission of interfering signals is the responsibility of the systems design lay-out engineer.

Prevention of the transmission of interfering signals by proper orientation of wiring means the utilization of the fact that the inductive coupling between two circuits can always be reduced to zero by proper orientation of one circuit with respect to the other; an example is illustrated in Figure 1-4. Here a comparison is made between the inductive coupling of two circuits that contain a section of parallel wires and that of two circuits having wires crossing perpendicularly. The amount of inductive coupling between the two circuits depends on the amount of magnetic flux of one circuit linking with the other. Figure 1-4 shows that the number of flux linkages is a maximum when the two wires run parallel, and zero when they are perpendicular.



Parallel Conductors, Maximum Flux Linkages

Perpendicular Conductors, Zero Flux Linkages

Figure 1-4. Coupling of Long Straight Wires

Therefore, whenever conductors carrying interference currents and conductors sensitive to interference pickup must be close together, every effort must be made by the systems design and lay-out engineer to avoid paralleling them and to have them cross as nearly as possible at right angles.

1.2.2.1 Grounding

The word "grounding" has two different meanings in connection with electrical systems. Sometimes it is used synonymously with bonding to mean: connecting a point to the system through a low impedance path. At other times it is used to mean: connecting a point electrically in such a way that it becomes equipotential with all other "grounding" points in the system. It is this second meaning which gives rise to much confusion, and many radio interference problems can be directly traced to a faulty interpretation of this concept.

It is a common procedure in wiring and circuit diagrams to use the ground symbol to indicate that a point should be connected electrically to the system. This procedure is unambiguous for direct and low frequency power currents, but objectionable for radio frequency currents. As was pointed out and emphasized in Section 1.2.1.1, bonding to the structure does not, in itself, assure the existence of a true ground plane at most of the frequencies encountered in interference problems. And it must be remembered that the impedance between two points supposedly at the same "ground" potential need not be very large to cause interference to be transmitted to the receiver.

Consider, as an example, the circuit of Figure 1-5. There are four different "grounds", one for the receiver, one to which the antenna is coupled capacitively, one for the motor, which is a source of interference, and one to which the by-pass condenser of the motor is connected. These four points are labeled 1, 2, 3, and 4, respectively. Assume that points 1 and 4 are at the same potential and also points 2 and 3, but let there be a small impedance between points 2 and 4 as indicated by the dashed connection. This impedance, Z, may be caused by poor bonding or by the fact that the structural member between these points has a small effective resistance or inductive reactance at the frequency of interest. Inspection of the diagram shows that this impedance is common both to the interfering currents from the motor and the desired antenna currents in the receiver. Because of the high sensitivity of the antenna circuit of the receiver, even a very small interfering voltage developed across Z will cause an appreciable interfering signal at the output of the receiver. This example illustrates a very frequent cause of interference.

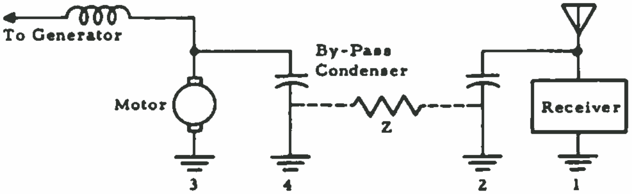


Figure 1-5. Ambiguity of Ground Points

The designer of the equipment and systems and the installation engineer must always be conscious of the possible ambiguity of ground points in the electrical system. He must keep clearly in mind the fact that proper "grounding" is not entirely a matter of good bonding, and that it may be impossible to have two points at the same potential over an appreciable range of frequencies unless the points are physically close together.

1.2.2.2 Shielding

A shield is a partition between two regions of space such that the electric and magnetic fields of interest are attenuated in passing from one of these regions to the other. All practical shields are made of metals of high conductivity.

The shielding action of metallic sheets may be explained in two ways depending on whether the field or circuit concept is used. According to the field concept, the shielding action of a metallic sheet is two-fold: An electromagnetic wave striking a metallic surface is partially reflected, and the transmitted part is attenuated in its passage through the sheet. According to the circuit concept, the currents flowing in circuits on one side of the sheet induce currents in the sheet. The induced currents produce fields on the other side which just cancel the fields due to the original currents. Mathematical analysis on either basis is very difficult, and is practical only for cases of simple geometry such as shields in the shape of infinite plane sheets or infinitely long circular cylinders. The second example finds application in the shielding of cables, but most shielding problems are not amenable to simple mathematical analysis. A compilation of the most important analytical expressions for the effectiveness of shielding in certain simple cases together with their derivations is given in Chapter 3.

One of the most significant results is that a plane electromagnetic wave is attenuated very rapidly in a metallic medium after entering it through a plane boundary surface. The fields decrease exponentially according to the law.

$$F = F_0 e^{-1.238\sqrt{\frac{\mu f}{\rho}}\,S} \tag{1-4}$$

where F is the electric or magnetic field intensity at a distance S inches away from the surface, F_0 the same field intensity at the surface in the same units as F, ρ the resistivity of the material in ohm-circular-mils per foot, μ the relative permeability (μ = 1 for all non-magnetic materials), and f the frequency in cycles per second. This variation is shown in Figure 1-6. The above equation is exact only when the metal extends to infinity in the direction of increasing S, but in case of a shield of finite thickness the equation is a good approximation when the field at the far end is sufficiently small. In practice, even a very thin metallic sheet will allow only a small fraction of the entering energy to pass through it. In addition, the reflection coefficients of most metallic surfaces to plane waves are large so that only a small fraction of the energy striking the surface will enter the metal. If a shield must support itself mechanically, the thickness required by mechanical considerations is usually in excess of that required for effective suppression of a plane wave except at low and very low frequencies. However, this does not apply to metallic coatings applied to mechanical supports of other materials. These coatings are often so thin that no effective shielding action is obtained. Of course, even a comparatively thick shield might not offer sufficient attenuation if the source is very strong and the receiver very sensitive.

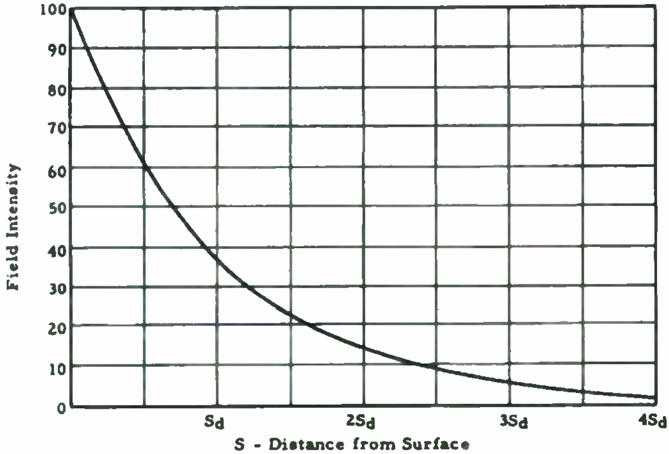


Figure 1-6. Variation of Electric or Magnetic Field Intensity Near the Surface of a Plane Conductor

It is often convenient to define the "depth of penetration" of a plane wave for a plane sheet. This is that value of S, called S_d, which makes the exponent in Equation (1-4) equal to -1. Thus

$$S_d = \frac{1}{1.238\sqrt{\frac{\mu f}{\rho}}} = 0.8079\sqrt{\frac{\rho}{\mu f}} \qquad (1-5)$$

Thus, the depth of penetration is the thickness through which the wave must travel to be attenuated to 1/e or about 37% of its original amplitude.

The depth of penetration of an electromagnetic wave is a concept closely related to that of the "skin effect". Just as the wave, at high frequencies, is very rapidly attenuated as it progresses into the metal, so the current decreases with the distance from the surface. This crowding of the current near the surface of a conductor is called the skin effect, and it gives rise to the increase of the effective resistance of a conductor with frequency. The shielding action of a metallic sheet may also be explained in terms of the skin effect. If the shield is thick enough so that the currents induced in one side of the sheet are practically reduced to zero on the other side, then no electromagnetic field can exist on the other side (provided that there is no source on that side) and the shielding action is complete. Since the current decreases exponentially, it actually never reaches the value of zero no matter how thick the shield is. This shows clearly that there cannot be a perfect shield theoretically. However, at higher frequencies the decay is so rapid that the shield need not be very thick to make the fields on the other side practically undetectable by the most sensitive receivers available.

Equation (1-4) shows that the effectiveness of shielding, as far as absorption within the metal is concerned, is directly proportional to the thickness of the shield, to the square root of its conductivity, and to the square root of its magnetic permeability. It also shows that the shielding effectiveness increases as the square root of the frequency so that the thickness, conductivity, and permeability of the shielding material should be chosen for the lowest frequency. However, when magnetic materials are used, it must be remembered that the permeability of most magnetic substances decreases with frequency. Therefore, an increase of shielding effectiveness with frequency is not always realized. It should also be noted that, as far as openings and joints in shields are concerned, their "leakiness" depends on their dimensions measured in wave lengths, hence their presence makes the shielding effectiveness decrease with frequency.

The discussion above, together with Equations (1-4) and (1-5), is based on plane waves. It is shown in Chapter 3 that within metals electromagnetic fields behave like plane waves under a variety of different conditions. For example, the remarks about the skin effect and Equation (1-5) are applicable to most metallic surfaces as a good approximation even for cylindrical waves unless the surface is curved with a radius of curvature much smaller than a wave length. On the other hand, the fields encountered in radio interference problems are often not associated with plane waves. The actual fields in the vicinity of sources of interference may be so complicated that the application of the results obtained on the basis of plane waves may lead to serious errors.

Practical shields are usually designed so as to enclose completely either the source in order to keep the interference in, or the receiver in order to keep the interference out. In either case, assuming that the shielding material has sufficient thickness, conductively, and permeability, the effectiveness of the shield depends on its completely enclosing the source. The inside (or outside) of the shield must form a continuous surface of lower impedance than any other possible current path leading to the outside (or inside) of the shield. This means that openings must be avoided, and any joints must be carefully designed so as to make sure that good electrical contact is made along a continuous line. In ignition systems, which depend entirely on shielding for interference-free operation, practically all interference troubles have been traced directly to faulty joint design. When openings are necessary, as for ventilating purposes, they must be specially designed for minimum transmission of interfering signals. They must interrupt the flow of induced currents in the shield as little as possible, and they must strongly attenuate any radiation through them. Protruding sleeves around the openings, acting as wave guides below their cut-off frequencies, have been found very helpful.

When long cables are shielded, the outside of the shield is another possible path for interfering currents from sources that may have no connection with the currents within the shield. To minimize this possibility as well as the effect of any faulty shield or shielding joint design, all cable shields must be properly grounded at least at each end, and for very long cables also at intermediate points. It might be argued that "floating" shields, or shields grounded at only one point, are also effective in preventing undesired currents through them. But it has been found in practice that grounding as suggested above is more effective despite the limitations pointed out in Section 1.2.2.1.

Certain types of special purpose shields are sometimes used to reduce one type of coupling without affecting another. The most important example of this is the "Faraday shield," which is used to prevent capacitive coupling between two coils without affecting the inductive coupling. A Faraday shield is a set of grounded metallic prongs, arranged somewhat like the teeth of a comb, placed between two coils. Since the prongs are not connected at one end, no induced currents can flow through them and the magnetic coupling is not affected. But the prongs are so close together that their plane is essentially equipotential. Thus, no electrostatic coupling can exist between the two coils. Another example of this is the electrostatic shield around a loop antenna, used to make the antenna responsive only to magnetic fields. This device is useful when the most important interfering signals have a much larger ratio of electric to magnetic field intensity than the desired signals.

1.2.2.3 Filtering

No matter how well a source is shielded, some electrical connections must be made to it that will break the shield because energy must be either supplied to it or carried away from it or both. In addition, control cables may have to be connected to it. These power and control leads will conduct interfering currents away from the source despite all efforts toward perfect shielding. The conduction of interfering currents may be impeded by the use of filters or other special networks.

A filter is a four terminal network designed to freely transmit currents or voltages of certain frequencies while attenuating all others. To do this, use is made mainly of reactive elements, i.e., inductances and capacitances. The presence of dissipative elements, that is, elements with effective resistance or pure resistances, prevents the free transmission of desired currents or voltages and is therefore usually avoided in filters.

Filters are classified as low-pass, high-pass, band-pass, and band elimination, depending on the bands of frequencies to be transmitted and attenuated.

The elements of the filter must be so chosen that the impedances looking into and out of the filter remain approximately the same as those of the transmission line into which it is to be inserted at the frequencies it is desired to transmit. This is necessary in order to insure that the filter does not impair the normal functioning of the equipment at both ends of the transmission line. In other words, the load impedance as seen by the generator should not be changed by the insertion of the filter so that the generator still delivers the current for which it was designed at the voltage for which it was designed. And the load should still be fed by a network of the same open circuit voltage and the same internal impedance so that it operates exactly in the way intended by the designer. On the other hand, at the frequencies the filter is designed to attenuate, the impedances as seen by the generator and the load are different from what they were before the insertion of the filter - usually either very

The design of filters is an art as well as a science since so much depends on the judgment and techniques used by the filter design engineer. Filters are discussed in detail in Chapter 4 of this Volume. Formulas, curves, and tables are given in Appendix I to assist the engineer in the design of low pass filters for the solution of interference suppression problems on a scientific basis, thus avoiding the pitfalls which are almost certain to be encountered in choosing values of elements on a trial and error basis.

1.2.3 SUPPRESSION IN RECEIVERS

In spite of all efforts to suppress the interfering signal at the source and to prevent its transmission, there will always be some such signals that reach the receiver. In the design of receivers which will not be susceptible to interference, the designer must strive (1) to prevent the interfering signals from entering the receiver and affecting any sensitive circuits, and (2) to minimize the effects of the interfering signals in case they have gained entrance and have affected a sensitive circuit. The first is a problem of transmission. The second is a matter of utilizing, in one way or another, any differences that may exist between the desired and the interfering signal.

1.2.3.1 Design for Minimum Transmission

The problems encountered in trying to prevent an interfering signal from entering the receiver have been discussed previously in Section 1.2.2. They are problems of shielding the entire receiver including all antenna lead-ins, filtering all power and control leads leading into the receiver, and designing the receiver in such a way that interfering signals entering through the output leads cannot affect any sensitive portion of the receiver. This last problem is a difficult one since filtering in the output leads is not usually practical, and once an interfering signal has entered the receiver, the problems of preventing its transmission to a sensitive circuit are greatly magnified. Extensive shielding and filtering within the receiver may, at times, be necessary.

Attention must also be paid to the location of switching devices related to receiver operation. Often an antenna relay is found within the receiver case, thus affording the interfering signals from the transmitter an easy entry. The number of leads entering a receiver must be kept at the absolute minimum, and any device, not an integral part of the receiver, should not be placed in the receiver case.

1.2.3.2 Bandwidth Considerations

One difference between the desired and interfering signals is their frequency content. The desired signal contains only frequencies within a well defined region of the frequency spectrum; for example, the carrier and two side bands for double side band transmission in amplitude modulation. On the other hand, the interfering signal is spread fairly evenly over a very large portion of the frequency spectrum. Of these frequencies, the receiver accepts only those that fall within its band of acceptance. Hence, a receiver should be designed so as to reduce the width of its acceptance band to the minimum required for the reception of the desired signal.

The last sentence must be interpreted in the light of the results of a statistical analysis of random noise. Such analysis shows that the larger the bandwidth of the receiver, the greater, potentially, could be the improvement of the signal-to-interference ratio provided the bandwidth is fully utilized for the improvement of this ratio. Hence, "minimum" means the smallest bandwidth required for making the fullest use of the potentialities of the system to improve the signal-to-interference ratio. Thus, for example, a receiver using frequency modulation will provide a greater signal-to-interference ratio at its output than one using amplitude modulation for the same signal-to-interference ratio at the input, even though it has by far the larger bandwidth. The reason for this is that the frequency modulated receiver makes more efficient use of its bandwidth.

On the other hand, for limiters to be effective, interfering pulses should undergo as little pulse lengthening as possible. Hence, the bandwidth should be as large as possible. Thus, the bandwidth requirements are contradictory, and the actual design must be a compromise. The designer must decide, in each case, whether it is better to allow more interference to enter the receiver and suppress it later by effective limiting action, or to allow as little interference as possible to enter, with the knowledge that any attempt to limit later will be quite ineffective.

1.2.3.3 Special Circuits

A large number of special circuits have been developed for the purpose of reducing radio interference in receivers. Section 2 of Chapter 4 of this Volume discusses five specific circuits presently in general use: (1) limiters, (2) wave traps, (3) blanking circuits, (4) phase cancelling circuits, and (5) audio filters.

Their general application to suspected receiver circuitry can be enumerated here. Limiters and blanking circuits are applicable when the interfering signal consists of large amplitude pulses whose duration is very short compared to their period. Wave traps are applicable if the interfering signal contains only one radio frequency or at most only a very narrow band of frequencies. Phase cancelling circuits are effective against interfering signals whose character and path of entry are precisely known. Audio filters are used when the interfering signal contains only a small number of fixed audio frequency components.

1.3 DESIGN CONSIDERATIONS FOR SUBASSEMBLIES AND CIRCUITS

1.3.1 GENERAL

Radio interference originates from the operation of the circuits and/or subassemblies of the system. Consideration of the system itself is essential only to establish techniques to prevent the interference which is generated by the circuits and/or subassemblies from being transmitted by radiation, conduction, or coupling to various other susceptible subsystems in the system. The ideal method of eliminating the effects of unwanted signals is to design all circuits and/or subassemblies in such a way that no unwanted signals are present.

Since source suppression can best be employed in the original circuit and subassembly design, the techniques and procedures described in the following paragraphs are of major importance. It is of utmost importance that all electrical circuits and subassemblies, regardless of their function or location within the system, be treated as potential sources of interference.

Electromagnetic interference generated by the operation of equipments generally originates in stationary resonant circuits. These include local oscillators, transmitters, and modulators, which perform functions essential to the operation of both radio and radar equipments. Consequently, each electronic component must be designed so that electrical oscillations are confined to the unit itself and not permitted to enter other electronic components either directly or indirectly, thus causing interference. The design and interference suppression methods relative to parasitic oscillations, transient oscillations, arcing, ignition sparks, and switching interference are discussed in the following paragraphs.

1.3.2 LOCAL OSCILLATORS

In superheterodyne receivers the local oscillator circuits serve a useful purpose in supplying an output whose frequency differs from a received signal frequency by a constant difference. The oscillator energy must at the same time be prevented from coupling through any path leading to the antenna or chassis and risking radiation which will interfere with the normal operation of adjacent electronic equipment. This is discussed under Section 1.4.3 of this chapter.

1.3.3 FEEDBACK AMPLIFIERS IN TRANSMITTERS

Radio transmitters are generators of radio frequency energy which is controlled by the intelligence to be transmitted. The very heart of such a generator is the oscillator circuit whose frequency of operation must be highly stable for the usual transmitter applications. To accomplish their purpose with a high quality of transmission, radio transmitters must be free from harmonic radiation, spurious sidebands, distortion, and hum. Radiation of harmonics is particularly troublesome in high powered transmitters. One percent of second harmonic radiation in a 500 kw transmitter corresponds to a 50 watt power signal level, and can readily produce an interfering signal over a considerable area.

Negative feedback is frequently used in transmitters. The block diagram shown in Figure 1-7 illustrates the operation of a feedback amplifier.

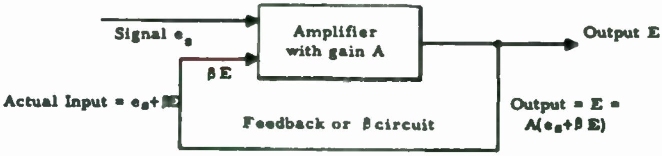


Figure 1-7. Feedback Amplifier Schematic

The quantity A represents the amplifier gain, and e_s the supplied signal. A fractional part, βE, of the output E is added to the external input, so that the actual input consists of the signal e_s plus this fractional part βE. The total input, multiplied by the gain, must equal the output. Thus, $E = A(e_s + \beta E)$. Representing the gain as the ratio of output to external input, it is simple to derive a gain formula.

$$\begin{aligned} &E/e_s = \text{actual gain} \\ &E = A(e_s + \beta E) \\ &E = Ae_s + A\beta E \qquad\qquad (1\text{-}6) \\ &Ae_s = E - A\beta E \\ \therefore\ &E/e_s = A/(1 - A\beta) = \text{gain with feedback.} \end{aligned}$$

If the feedback opposes the input signal, then β is considered negative. The quantity $A\beta$ is called the feedback factor, and if this quantity is much larger than unity, then the gain is reduced to $-1/\beta$. Thus when the feedback is large, the effective amplification depends only on the value of β, and is practically independent of the actual gain produced by the amplifier. If the feedback circuit employs a resistance network, the gain is almost independent of frequency but has phase shift of approximately 0° or 180°. If it is desired to have amplification vary with frequency, then the feedback circuit (β circuit) can be designed to have the desired transmission loss characteristic.

Negative feedback causes a reduction in amplitude distortion since some of the distortion is fed back to the input through the feedback circuit and reamplified in such phase as to cancel out most of the original distortion. If D indicates distortion in the output and d the distortion generated in the amplifier, then

$$D = d/(1 - A\beta) \qquad (1\text{-}7)$$

A large feedback factor, (recall that β is negative for opposing phase), will greatly reduce distortion in the output.

Feedback will modify the signal-to-interference ratio by the following relationship. Signal-to-interference ratio with feedback/signal-to-interference ratio without feedback = $A_f/A_o(1-A\beta)$. A_f represents the amplification taking place between the point where the interference is introduced and the output, with feedback. A_o represents the same amplification without feedback. The equation is based on the assumption that it is the same interference in each case and that the output voltages are the same. Analysis shows that feedback will reduce interference introduced in the high level stages of the amplifier, as for example, a poorly filtered power supply in the plate circuit of the final tube. It will not aid in reducing interference entering the low level stages, as microphonics or induced hum, since the feedback affects the interference and normal signal output to about the same extent.

The preceding paragraphs serve as background for appreciating some of the benefits of negative feedback. However, oscillations can result from such feedback because of accompanying phase shift. In the normal range of frequencies, circuit arrangements are such that feedback is negative. At the very low and very high frequencies, the amplifier stages produce phase shifts sufficient to cause the feedback factor, $A\beta$, to change from negative to positive. Oscillations are not usually encountered in two stage arrangements unless the feedback factor is made large. Where there are more than two stages, however, oscillations tend to take place even with a moderate amount of feedback.

Phase shift depends upon variation in the amplitude of transmission with frequency and its polarity on the sign of the slope of the transmission characteristics. Where the transmission is constant with frequency, that is a flat response, there is no phase shift. If the amplitude of transmission, "a", has a constant variation, then the phase shift is directly proportional to the slope of the amplitude characteristic. Expressed as an equation

$$\text{Phase shift (radians)} = \frac{\pi}{12} \times \frac{da}{du} \qquad (1\text{-}8)$$

where da/du represents variation in the amplitude of transmission expressed in decibels change in transmission for an octave change in frequency. In designing a feedback system to avoid oscillations it is only necessary to consider the way in which the amplitude of transmission varies with frequency. Thus if the feedback factor, $A\beta$, has shifted in phase by 180° and becomes positive, the magnitude of $A\beta$ must have decreased to less than unity to avoid oscillations.

In frequency modulated transmitters negative feedback will give the same benefits as in the amplitude modulated system. The problem of preventing oscillations will also be the same and, therefore, subject to the same design considerations previously discussed.

1.3.4 MODULATORS

Radar modulators are capable of producing large amounts of interference. The problem of reducing the interference to meet required specifications can and must be solved by proper design and internal shielding.

Measurements were made on a line type modulator to determine the best method for reducing interference conducted along the power line. A single copper sheet between primary and secondary windings served as an electrostatic shield for the power and filament transformers. Primary leads were placed as far from the thyratron as possible and at right angles to other wiring. The primary power leads were shielded with copper tubing from the transformers to the point where they left the modulator case, and this shielding was grounded. Partition shielding was used to separate the rectifier from the thyratron. These features are illustrated in Figure 1-8.

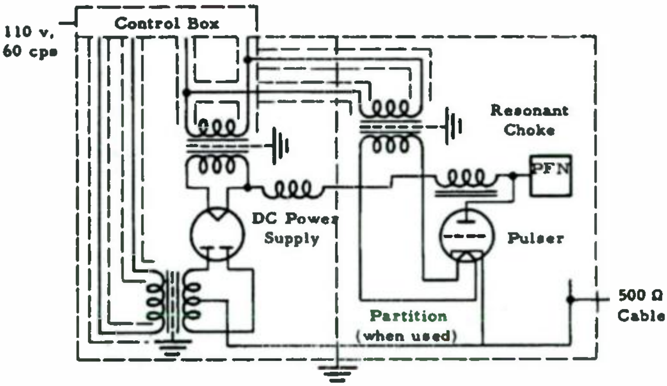


Figure 1-8. Radar Modulator and Pulse Lead Shielding

Interference was reduced to less than 50 microvolts over the entire spectrum, from 0.2 to 20 megacycles, without adding filters. When used, filters offered additional interference suppression from 0 to 5.5 mc, but the difference was minor above 5.5 mc, and therefore without justification in view of their additional weight and size. Individual shields were also used in place of the partition, and offered considerably greater reduction in the region of two megacycles. The relative merits of these design features are shown in the graph of Figure 1-9.

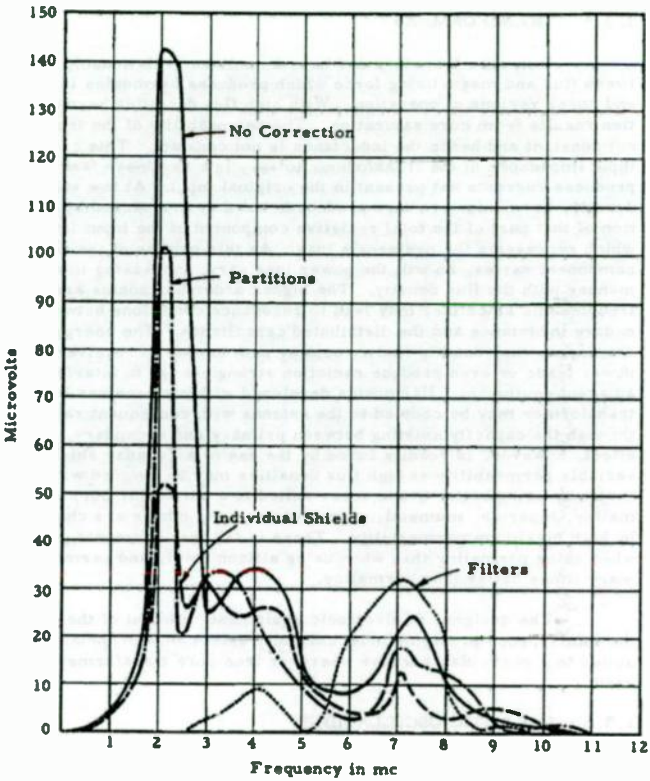


Figure 1-9. Interference Plotted Against Frequency

1.3.5 TRANSFORMERS

Iron core transformers have a nonlinear relationship existing between flux and magnetizing force which produces harmonics in the lower and upper regions of operation. With high flux densities harmonic generation results from core saturation. The permeability of the iron core is not constant and hence the inductance is not constant. This causes the input impedance of the transformer to vary in a nonlinear fashion which produces currents not present in the original input. At low values of flux density, harmonics are also produced, but they are caused by the variation of that part of the total resistive component of the input impedance which represents the hysteresis loss. As this portion of the resistive component varies, so will the power loss vary, increasing in a complex manner with the flux density. The higher order harmonics are particularly troublesome since they may lead to resonance conditions between the secondary inductance and the distributed capacitance. The energy, once developed, may readily find a coupling path to nearby receivers through power leads or even produce radiation strong enough to interfere with adjacent equipment. Harmonics developed within a receiver by the power transformer may be coupled to the antenna with consequent radiation, through the capacity existing between primary and secondary. This latter effect, however, is readily cured by the use of a Faraday shield. A variable permeability at high flux densities may be avoided within extended limits by using better grade cores suited to a particular purpose. Permalloy, hipernik, mumetal, perminvar, among others are characterized by high maximum permeability. There is far less harmonic generation when using permalloy than when using silicon steel, and perminvar is many times better than permalloy.

The designer of electronic equipment, mindful of the severity of the interference problem, must carefully select an appropriate transformer suited to a particular purpose wherever iron core transformers are to be used.

1.3.6 PARASITIC OSCILLATIONS

Oscillations which occur at other than a desired frequency, or outside a tank circuit, are called parasitic oscillations. They may take place in oscillators as well as in ordinary power amplifiers, and the energy they represent is capable of reducing the normal output at the operating frequency to a small fraction of its value. These spurious frequencies give rise to distortion in linear amplifiers and modulators, and may produce spurious sidebands, cause flashovers, and other undesirable effects.

Parasitics may be of higher or lower frequency than the normal operating frequency of the amplifier or oscillator. When a circuit possesses sufficient energy storage capabilities and enough feedback of the proper phase, it will oscillate, and the effect is normally superimposed on the output of the amplifier or oscillator.

High frequency parasitic oscillations, usually above 30 megacycles, may exist in tuned radio frequency electronic circuits. The circuits give rise to parasitic oscillations due to the lead inductance between tube and tank circuit and the interelectrode capacities of the tube. When large tubes are used, the long leads and rather large interelectrode capacities, as well as high transconductance of a given tube, increase the possibility of parasitic oscillations. Ordinary triodes have a feedback path through the grid plate capacitance. In pentodes there is coupling at high frequencies since the screen and suppressor grids are no longer at zero potential because of the ground lead inductance. The coupling capacities are from plate to suppressor, and from grid to cathode. Part of the lead inductance is due to the internal wiring of suppressor and cathode, as shown in Figure 1-10, and the remainder due to the lead wires of the by-pass capacitor to ground. The capacitor itself, at relatively high frequencies, is like a short circuit. This analysis will explain why tubes designed for ultra high frequency applications have a separate suppressor grid pin on the tube base. This should be directly connected to the chassis and not to the cathode.

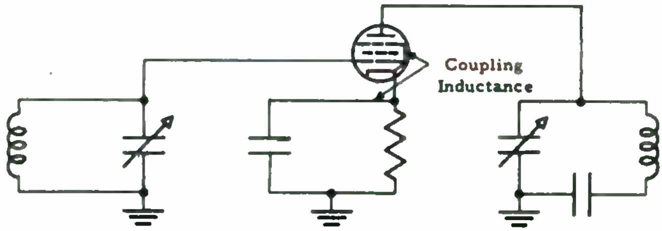


Figure 1-10. Schematic of Lead Inductance at High Frequencies

For a better understanding of parasitic oscillations and the means of preventing them at frequencies well above the normal range of operation, consider the Class C amplifier illustrated in Figure 1-11.

At high frequencies the capacitances C_1 and C_2 of the tank circuits are practically short circuits, while the tank inductances, L_1 and L_2

may be treated as open circuits due to the high impedances at the frequencies involved. Under these conditions, the circuit reduces to a tuned grid tuned plate oscillator type, as shown in Figure 1-12. The grid and plate tuning capacities are supplied by the interelectrode capacitances of the tube, and the inductances, L_g and L_p, by the lead inductances between electrodes. By comparing this figure with Figure 1-11, it is noted that the neutralizing capacity is not effective since it forms no part of the parasitic oscillatory circuit.

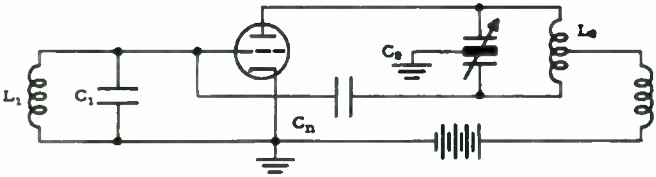


Figure 1-11. Conventional Class C Amplifier

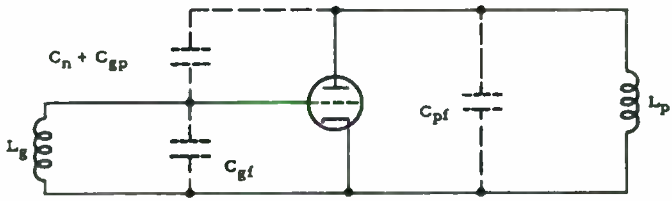


Figure 1-12. Class C Amplifier, Equivalent Circuit

The parasitic oscillations may be prevented by inserting a small resistance, about 1 to 25 ohms, in series with the grid or plate lead. It is preferable to insert it in the plate lead since it then affects the parasitic current directly, but is not in series with the main oscillating circuit. Detuning processes are effective in eliminating parasitics. The resonant frequency of the grid circuit may be increased, or that of the plate decreased, which causes the plate circuit to offer a capacitive reactance and thus introduces positive resistance into the grid circuit. The detuning process may also be accomplished by shortening the grid leads to decrease their inductance and lengthening the plate leads, or inserting a small choke in the plate lead next to the tube. The resonant frequency of the plate circuit being much lower than the grid circuit, oscillations cannot occur.

Low frequency parasitics occur where radio frequency chokes are used in series with the dc supply to both plate and grid, as shown in Figure 1-13. The equivalent circuit is shown in Figure 1-14. At frequencies

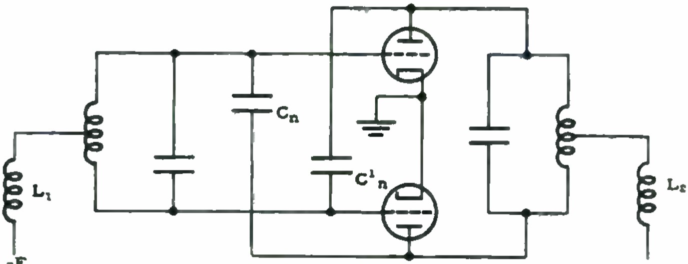


Figure 1-13. Amplifier Circuit

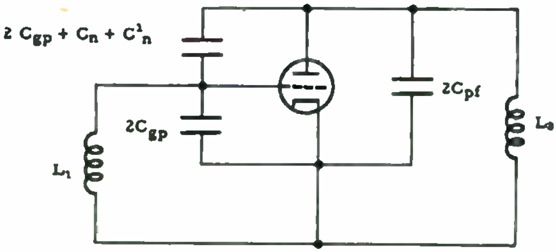


Figure 1-14. Equivalent Parasitic Circuit

well below the operating frequency the tank circuits' inductances are effectively short circuits, and the equivalent tuned grid tuned plate circuit employs the chokes, L_1 and L_2, as inductances and the interelectrode capacitances of the tubes as tuning capacitances. Again the neutralizing condensers are not effective in preventing coupling, since the tubes act in parallel for the parasitic action, rather than push-pull. Chokes are to be avoided, if possible. If, however, it is necessary to use them, oscillations will be prevented by a selection of chokes such that the resonant frequency of the grid circuit is higher than that of the plate.

Parasitic oscillations may occur in the grid circuit of a vacuum tube. Throughout a given voltage range, grid current may decrease as grid voltage increases because of secondary emission. This indicates a negative grid resistance over that range and may result in parasitic oscillations. Besides the possibility of underirable oscillations, severe distortion would be present in the output, and operation within the voltage range where this action occurs is to be entirely avoided by proper adjustment of the plate voltage and the grid bias.

Pentagrid and triode hexode converter tubes, which combine the functions of oscillator and mixer tubes, are characterized by a type of interaction which results from coupling between the signal grid and virtual cathode in the vicinity of the signal grid. The virtual cathode pulsates at the oscillator frequency and thus induces currents at the same frequency

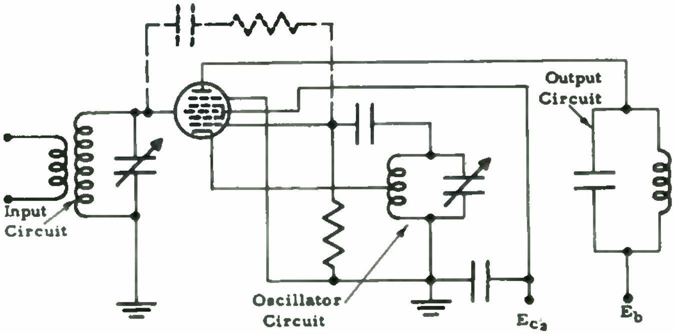


Figure 1-15. Neutralization of Space-Charge Coupling

in the signal grid circuit. This effect increases with frequency. At high frequencies, the tuned circuit of the signal input grid has a resonant frequency that may differ from the oscillator frequency by only a small amount. Thus, considerable impedance is offered to the induced current of the local oscillator frequency. The result is that the oscillator voltage developed on the signal grid causes the output to drop and to become relatively independent of the input tuned circuit. This space charge coupling can be effectively neutralized by inserting a capacitance in series with a small resistance between signal and oscillator grids. This arrangement is indicated, by dotted line, in Figure 1-15.

1.3.7 RESONANT CIRCUITS

The circuits previously mentioned in Paragraph 1.3.6, which were capable of setting up parasitic oscillations, may just as easily set up transient oscillations. The parasites represent the condition of sustained oscillations, while the transients are damped oscillations decaying with time. Transients may be produced in any circuit when it is disturbed by any sudden electrical change, such as a pulse of energy, or the discharge of a capacitor. Transient oscillations may produce interference in a receiver output such as reduction of the output and introduction of distortion and spurious frequencies. A group of damped oscillations, repeated periodically may produce interference in the output at the repetition frequency. If only one or two oscillations of appreciable intensity are present, or if single pulses of negligible width act on a resonant circuit, they may produce disturbances in the output due to their broad energy spectra.

Oscillations produced in coupled circuits are more complex than a simple damped train of oscillations in a single circuit. An analysis of transients in electrical systems results in a mathematical expression of the voltage time, current time, and charge time relationships of a system. The equations involved indicate the nature of the oscillations, and how they may be prevented. Applying Kirchoff's voltage law to the series LCR circuit shown in Figure 1-16 yields the equation

$$E - iR - L\frac{di}{dt} - \frac{1}{c}\int i\,dt = 0 \qquad (1\text{-}9)$$

An analysis of this equation was made in Section 1.2.1.2 and it was shown there that critical damping exists for the condition $R = 2\sqrt{L/C}$. If R is less than $2\sqrt{L/C}$, the circuit is underdamped, the dissipation is small, and the circuit will be subject to damped oscillations. If R is greater than $2\sqrt{L/C}$, it is overdamped and oscillations are prevented. The circuit shown is used chiefly for the purpose of analysis, and the oscillations could result from the actual closing of a switch, or from a pulse of energy inductively coupled to the circuit inductance.

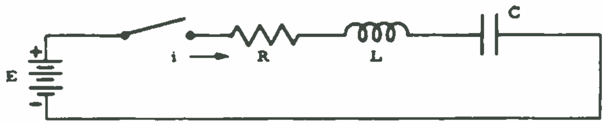


Figure 1-16. Series LCR Circuit

1.3.8 ARCING

Arcing as a source of radio interference is discussed in Volume I, Chapter I, Section 4. Most arcs that are troublesome in electrical or electronic equipment occur in connection with switching processes. Switching transients cause radio interference even in the absence of arcing, but usually the interference is increased by a large factor when arcs are present. Arcing is most severe in circuits having high inductance because it is the magnetic energy stored in the inductance that must be dissipated in the arc. Basic arc suppression techniques are discussed in Section 1.2.1.2.

1.3.8.1 Ignition Sparks

Present engine ignition systems are good examples of radio interference generators of the type discussed above because of the steep wave transients that ensue immediately after the firing of each spark plug. Figure 1-17 shows the wave shape of an ignition pulse when the out-put of a magneto is connected to a typical ignition system. This figure shows that the ignition pulse consists of a fundamental and a series of high frequency harmonics that are the cause of radio interference originating in the ignition system.

This interference can be prevented from being radiated or coupled into other wiring if the entire system is encased in a continuous metallic shield which is adequately bonded to the structure. The use of filters is usually not satisfactory because any filter sufficiently efficient to remove all high frequency components of an ignition pulse would also destroy the characteristic wave shape which is essential to the correct functioning of the ignition system. Spark plugs, which produce the ignition spark by using the power and voltage developed by the magneto, must also be shielded to prevent the radiation of radio interference energy. Figure 1-18 shows a typical spark plug, threaded into a recessed well in the cylinder head, as well as the finned metallic enclosure which is employed to shield and cool the plug.

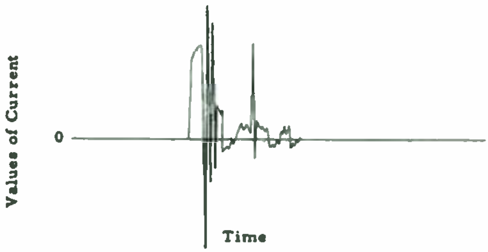


Figure 1-17. Wave Shape of an Ignition Pulse

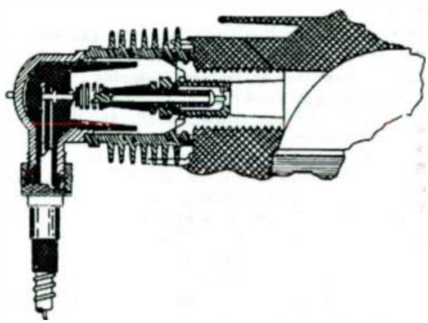


Figure 1-18. Spark-Plug Cooler and Radio Shield

Suppressors, i. e., impedance elements placed in series with the high tension lead of the distributor, are also employed to decrease the level of ignition interference. The purpose of suppressors is threefold:

a. To reduce the energy of the ignition impulse to such a value that spark plug erosion may be minimized.

b. To render the circuit nonoscillatory.

c. To reduce the steepness of the pulse wave front.

It is necessary to reduce the steepness of the wave front because even single periodic pulses are capable of setting up oscillations in the resonant circuits of adjacent receivers if the wave front is sufficiently steep. The steepness of the wave front is lowered by suppressors because the added resistance increases the time required for the current to rise from zero to its maximum value or to fall to zero from its maximum value.

1.3.9 T-R BOXES

The transmit-receive box is a cavity gas switch used in a microwave radar set employing a single antenna for transmission and reception. It prevents the transmitted pulse from entering the receiver and does not interfere with the reception of the reflected pulse. The box contains a tube, a resonant cavity which can be tuned, and provision for coupling the input and output circuits to the cavity. Two conical metallic electrodes separated by a short distance are enclosed in the tube, together with a slight amount of water vapor to improve the recovery time. The transmitter pulse causes a spark discharge in the tube which detunes the resonant cavity. This introduces a high degree of attenuation between transmitter and receiver circuits. The cavity in a discharging state is a means of rejecting the flow of radio frequency energy, whereas in the non-discharging state, there is a good match between input and output with very little reduction in delivered power. The various elements, with their connections, are illustrated in Figure 1-19.

The T-R box must be well shielded in order to prevent the interference due to the arc from affecting adjacent equipments. Energy entering the receiver must pass through the T-R box, and during the interval of a transmitted pulse, a small amount of transmitted energy will enter the receiver. This is referred to as leakage power. The leakage power consists of three components: (1) the "spike" of energy, (2) the flat power, and (3) direct coupling power. These terms are explained in Figure 1-20, which shows a typical leakage power pulse together with an idealized pulse for comparison.

The energy contained in the spike contributes to converter crystal failure, and to a far smaller extent so does the power in the flat section of the pulse. The time interval of the spike has been estimated at approximately 1/1000 of a microsecond, and the energy content is mostly dependent on the steepness of the transmitted pulse wave front and the repetition

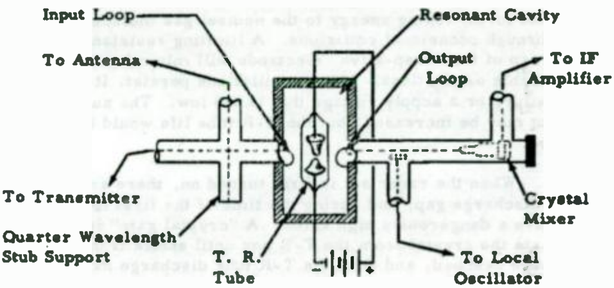


Figure 1-19. T-R Box and Crystal Mixer

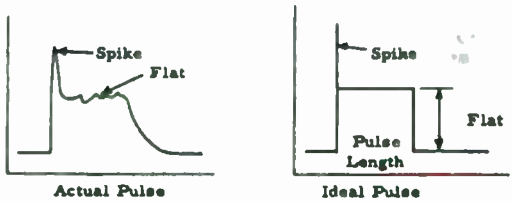


Figure 1-20. The Shape of the Leakage Power Pulse

rate. At low repetition rates, (less than 1000 pulses per second), the spike energy may be reduced by a direct current glow discharge near the radio frequency gap. A "keep-alive" electrode is supplied in all standard T-R tubes which provides a continuous supply of ions and free electrons to help establish the desired conditions in the radio frequency discharge path. However, oscillations may result due to the negative resistance characteristics of the low current discharge. This produces a cyclic variation in the number of free electrons and ions in the gap and causes the spike energy to fluctuate. The mean free path of an electron is, in general, of the same order of magnitude as the distance between electrodes, but very few electrons reach the electrodes because of the very rapid variations in the radio frequency field. Therefore, electrons oscillate back and forth, losing energy to the neutral gas molecules and to positive ions through occasional collisions. A limiting resistance mounted close to the cap of the "keep-alive" electrode will minimize the effects of these undesirable oscillations. If the oscillations persist, it is evidence of tube failure or a supply voltage that is too low. The auxiliary discharge current may be increased, but the T-R tube life would be correspondingly reduced.

When the radar set is first turned on, there are no residual ions in the discharge gap, and during the time of the first few pulses the spike may have a dangerously high value. A "crystal gate" is usually provided to isolate the crystal from the T-R box until stable transmitting conditions have been reached, and until the T-R tube discharge has been established. Another important function of the gate is to offer protection to the crystal, when the radar is idle, from damage by the radiated energy of other radars operating nearby.

The flat portion of leakage power indicated in Figure 1-20 is, in general, proportional to the spacing of the gap. Most crystals withstand flat power levels very well. Direct coupling, however, which is directly proportional to transmitter power and is a component of the flat portion, will result in damage to the crystal under improper operating conditions. This is so when the magnetron develops an appreciable amount of power at frequencies other than the frequency of normal operation. If they are in the vicinity of the resonant frequencies for these direct coupling modes in the T-R box, they may easily be transmitted with little attenuation. This is a threat to efficient operation in very high power systems.

1.3.10 RELAYS

High speed oscillographic measurements of radio interference produced by relays lead to the following conclusions:

a. The coil of the relay can be replaced by a capacitor having a capacitance equal to the distributed capacitance of the coil without altering the shape of the current transient wave during the first few microseconds after power is supplied to the circuit, as shown in Figures 1-21A and B. This indicates that the distributed capacitance effectively "shorts out" the coil during this short interval and is responsible for the generation of interference transients.

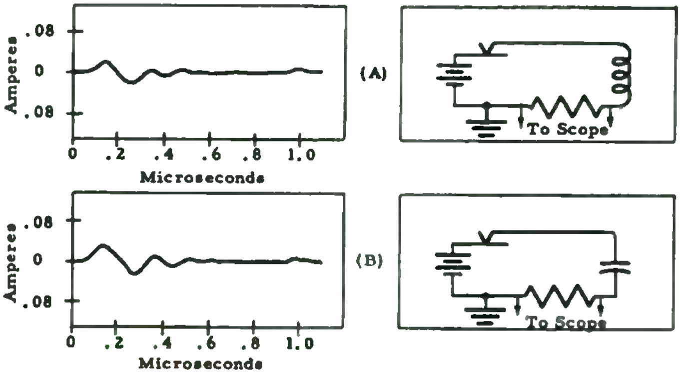


Figure 1-21. Equivalence of Relay Coil to Capacitance in Initial Closure Current Transient

b. Typical relay coils exhibit a high ratio of inductance to distributed capacitance, which results in high amplitude voltage surges with steep wave fronts caused by the collapse of the magnetic field about the relay coil when the current is interrupted. Figure 1-22A shows that the voltage across the coil rises quickly to the supply voltage of V volts dc when the circuit is closed, but on the break the potential rises to a value of approximately 100 times the supply voltage in about three microseconds and then decays to a zero value at a rate determined by the inductance, distributed capacitance, and resistance of the winding as shown in Figure 1-22B. It should be emphasized that this voltage surge possesses a steep wave front which is capable of producing violent shock excitations in receivers tunable over a wide range.

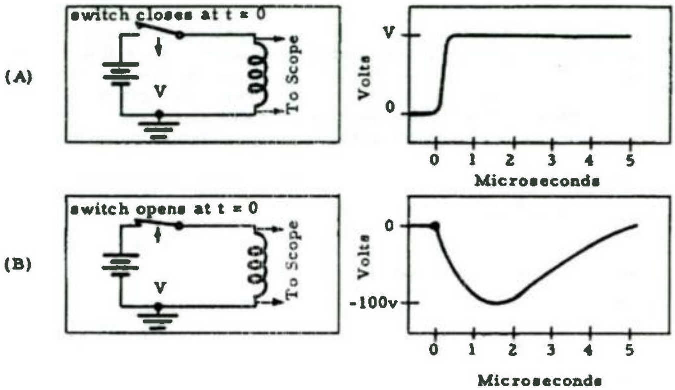


Figure 1-22. Make and Break Voltages Across a Relay Coil

c. Switching units, with the exception of mercury switches, display mechanical bounce or chatter which causes repetitive closures and interruptions of the current when the switch is closed. The long duration sweep shown in Figure 1-23 shows the effect of the bouncing switch contacts when the circuit is made. The high amplitude voltage surges shown in the figure are evidence that the points remain in contact sufficiently long to establish an appreciable current in the coil. These transients developed at the make of the circuit are of greater duration and severity than those developed at the break of the circuit.

d. In addition to the transients due to mechanical bouncing, there occur rapid closures and interruptions of the circuit. These are at a faster rate than those due to the mechanical bouncing of the relay contacts at the make of the circuit as described above. This is shown in Figure 1-24. As the contacts move outward, the contact area for the flow of current decreases resulting in local heating, which causes the contacts to pit and sputter until the circuit is broken. The amplitude of the resultant induced voltage is sufficiently high to imitate "cold" emission from the projecting area of the relay contacts. This is accompanied by local heating, which causes the contact material to melt and neck out until the circuit is again closed. This process repeats at an exceedingly rapid rate until the relay contacts are separated far enough to prevent the voltage gradient from rising to the value necessary for cold emission. These closures and interruptions of the circuit are also responsible for the generation of steep wave forms, which cause radio interference in adjacent electronic circuits.

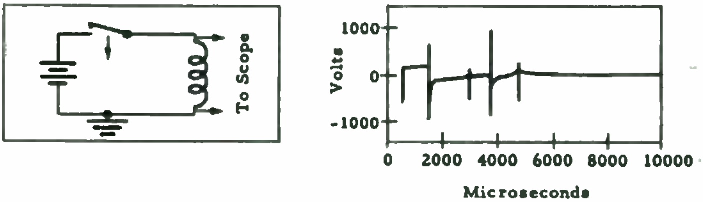


Figure 1-23. Bouncing Transients in Closing a Relay

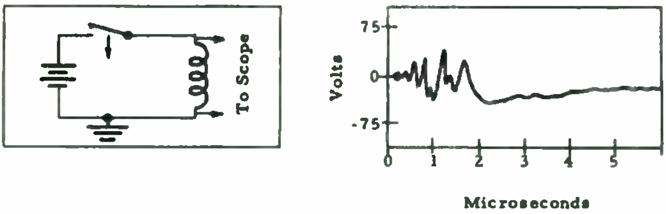


Figure 1-24. Pitting Transients in Opening a Relay

A reduction of the interference is obtained by enclosing the offending relay and its associated circuit within a metallic shield.

The use of low pass filters is probably the most practical means of suppressing the interference developed by relays or other devices which develop similar steep wave transients. Studies of low pass networks reveal that a series inductance, shunt capacitance network transmits the least high frequency energy. Filters of this type, however, fail to effect complete surge suppression because of the distributed capacitance of the inductance and the inductance that is inherently present in the capacitor leads. This suggests that the most effective low pass filter should consist of an inductance with the least possible distributed capacitance in conjunction with a feed-through capacitor. Furthermore, losses may be introduced into the filter to dampen the low frequency oscillations that may appear in this system.

A resistor in series with a capacitor connected across the contacts is another method used for the suppression of interference. A capacitor should never be connected across the contacts without including a series resistance because the discharge of the condenser when the contacts are closed can cause a heavy surge if not controlled by the resistor. Refer to Section 1.2.1.2 for a detailed discussion of this method of arc prevention.

1.3.11 BONDING CONTACTS

One of the purposes of bonding, as explained in Section 1.2.1.1, is to prevent arcing between adjoining metal parts. Bonding techniques are covered fully in Volume III, Chapter 3, and little radio interference trouble due to arcing between poorly bonded members may be expected if these techniques are strictly adhered to. In fact, since there are many other reasons for bonding, some of which impose much more stringent requirements, strict adherence to the bonding specifications will usually insure the elimination of interference producing arcs between members and parts covered by these specifications.

Arcing may occur between two metallic surfaces in the presence of a strong electric field, as near a transmitting antenna, if the bonding connection between them has opened up by corrosion. The resulting radio interference may well be the only indication of the fact that a poor bond exists. Another example is the arcing that may occur between the individual strands of copper mesh screening at the cross-over points if good electrical contact is not made. It has been observed that the copper mesh wall of a screened room can be the source of considerable radio interference when strong radio frequency fields are present. Good bonding of the strands at the cross-over points eliminates this interference entirely. Therefore, it is concluded that the same construction techniques must be employed for copper mesh screens used, for example, to shield ventilating louvres of electrical machines.

1.3.12 SWITCHES

Any switching device causes transients during opening and closing as explained in Volume I, Chapter 1, Section 7.1.2. Therefore, in the design of all switches used in electric systems, the radio interference problem must be considered from the outset.

As was pointed out in Section 1.2.1.2, arcing occurs during the normal operation of a switch when used to interrupt the flow of current. In fact, the interruption of a current in a circuit may be said to consist of substituting a highly ionized and therefore conducting gaseous medium, i.e., an arc, for a part of the metallic circuit, and then subjecting this arc to strong de-ionizing influences. The arc is extinguished when the energy stored in the inductance of the circuit is dissipated and the voltage drops below the value required to maintain the arc. To prevent the arc, the current, instead of being interrupted, is channeled into another branch containing a series capacitance and resistor. Thus, the energy is partly stored in the capacitor and partly dissipated in the resistor, which also serves to damp out any oscillations that may occur as a result of the added capacitance.

The design of these R-C arc-suppression networks is treated in Section 1.2.1.2. They must be used whenever switches or relays are used to interrupt currents large enough to cause radio interference. An alternative method is to completely shield and filter the unit, and when the currents to be interrupted are large, this may be the only effective method. But for small units, the use of an efficiently designed R-C network may make shielding and filtering unnecessary, and great savings in weight and space requirements may be effected.

1.3.3.6 Fluorescent Lamps

Fluorescent lamps contain mercury vapor at low pressure which under electron excitation, obtained by applying a difference of potential across the lamps, emits invisible ultra violet radiation. This in turn excites visible luminescence in the internal phosphor coating. A fluorescent lamp ballast - a coil of insulated copper wire wound on an iron core - is placed in series with the lamp to limit the current to its rated value.

Many lamps are also equipped with starters whose function is to complete a separate circuit so that a pre-heat current can flow through the cathode before the lamp operates.

Since basically the light source in a fluorescent lamp is an arc, considerable interference may be expected. This interference is both conducted away from the lamp through the power leads and also radiated directly from the lamp. The radiation takes place mostly at frequencies above 50 megacycles and is very difficult to control. Shielding with solid shielding material is obviously not feasible because there is no transparent conducting material. Shielding by means of copper mesh has been attempted, but such shielding also reduces considerably the amount of light available from the lamp. The exact frequencies and intensities of direct radiation are quite unpredictable and depend greatly on the age, condition, and type of lamp. For this reason, the use of fluorescent lamps in the vicinity of any electronic equipment or sensitive wiring should be avoided entirely.

As far as the conducted interference is concerned, the capacitor normally connected across the switch contacts to aid starting also aids greatly in by-passing the interfering current components and keeping them out of the power line. If the interference reduction with this one capacitor is not sufficient, a power-line filter, designed according to the procedures explained in Chapter 4, must be added. Thus, it is seen that the use of fluorescent lamps is undesirable also from the point of view of conducted interference. However, the interference originating in fluorescent tubes is sometimes put to use as the signal source of signal generators useful in measuring receiver response in the ultra high frequency range. The signal generators presently in use are adequate for general testing but because of their limited power output they are inadequate for testing wide range receivers.

1.4 DESIGN CONSIDERATIONS FOR RECEIVERS

To minimize the effects of interference, internal shielding of the individual radio frequency sections is essential. The first radio frequency section of any receiver is the most sensitive and any design feature causing a reduction in interference at this point has a healthy effect upon succeeding stages.

At least 90 percent of all interference enters a receiver through the input circuits of the first radio frequency amplifier stage. Interference conducted into the receiver by means of the power input cable may enter the first radio frequency stage through the filament circuit as well as through inductive coupling to the internal antenna circuits. The rest of the interference enters this stage through the external antenna circuits, (antenna, antenna lead-in) by means of inductive coupling with the interference carrying wires. The first radio frequency stage must be designed very carefully. Noted improvement is had by adding low-pass filters in series with the first radio frequency filament leads. Lower susceptibility will result from improved shielding design of the internal antenna input circuits. Shielding antenna leads inside the set, and shielding input coils and gang condensers will reduce the coupling to interference circuits, where the shielding is made as continuous as possible. Continuous shielding is obtained quite well by spacing the screws holding down the shield about one inch apart. However, the use of multiple contact serrated springs is much more effective for continuous shielding.

An effective way of constructing an internal shield to separate two stages is the use of a copper shield with a circular hole through which a tight fitting metal tube may be mounted. If the tube is of the type having a grid cap at one end and all other connections at the other end, its metal envelope effectively closes the hole in the shield, and there is complete separation of the grid or input and the plate or output circuits of that stage. Such an arrangement, used for the separation of the radio frequency amplifier from the mixer stage in a superheterodyne receiver, is shown schematically in Figure 1-25.

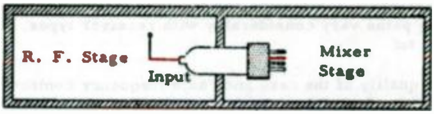


Figure 1-25. Mounting of Vacuum Tube Through Internal Shield

This method of tube mounting affords excellent isolation of stages to prevent stray coupling. The partition shield may serve as a common ground point for cathode and one side of the filament, and for plate and grid circuit returns. The tuning condensers may be insulated from the primary shield box, and copper straps used to connect the frame directly to the common ground on the partition shield. The control shaft of the condenser may be brought out of the primary shield box through a close fitting brass sleeve, which will be soldered to the walls of the box. Proper proportioning of the shaft hole and the length of the sleeve will make the brass sleeve act as an attenuating wave guide.

Interference problems in receivers have been solved in varying degrees by means of fixes. However, the best approach is to design, produce, and install the receivers to function with minimum susceptibility by means of appropriate routing of cables, shielding, filtering networks, and by adding circuits to the receiver to materially lower the amplitude of the offending interference voltages.

Susceptibility of a receiver to interference is a measure of undesirable response of the receiver to interfering voltages at all paths of entry. It may be expressed in terms of "coupling factor" or "susceptibility ratio." Coupling factor may be defined as the ratio of the antenna input voltage to the voltage input required at the various coupling paths to produce the same receiver output and is an index of the receiver's ability to reject conducted interference. Susceptibility ratio, which is sometimes used, is the inverse of this and may be measured as its reciprocal.

1.4.1 PATHS OF ENTRY

The major paths through which interference energy can enter a receiver are: (a) antenna lead-in, (b) power leads, (c) control leads, (d) output leads, and (e) receiver case.

Interference input paths and the attenuation of interference through these paths vary considerably with receiver types. These variations are due to:

a. quality of the case and radio frequency component shielding,
b. filtering of the input power circuits,
c. internal routing of interference carrying wires, and
d. filtering of radio frequency filament circuits.

Tight case shielding containing as few holes as is practical for cable outlets, and thorough internal shielding of the radio frequency and antenna circuits will lower the receiver interference susceptibility.

1.4.1.1 Antenna and Lead-In

Even though interference from all other possible paths of entry could be eliminated, the antenna would still provide a serious means for coupling interference into the receiver. Poor placement of the antenna or antenna lead-in with respect to the ignition system of an aircraft, the radiation field of a radar set, or of the components of its modulating pulse, may cause large induced voltages which severely affect the output of the receiver. Peak field strengths of radar transmitters, in the vicinity of a communications receiver antenna, may be of the order of hundreds of volts per meter. Under such severe conditions, interference may be introduced into the receiver case over the antenna lead-in or where poor shielding and bonding exist. The antenna and its lead-in provide an entry path capable of causing severe interference in the output despite the frequency discrimination of the receiver. The interference energy may be so intense as to saturate the first stage of the receiver. This type of interference may be caused by low frequency radar transmitters.

The use of a short shielded antenna lead-in wire is perhaps the best design practice for preventing interference from entering the first tuned circuit of the receiver due to antenna lead-in interference coupling. Amplitude limiting and suppression of the receiver during the radiation period of the radar pulse are other used methods. The method of approach will be determined in accordance with the established purpose of the receiver, although it is best to eliminate interference at the earliest possible point in a receiver. Appropriate location and orientation of the antenna is the only corrective measure available to reduce interference pick-up by the antenna itself.

High rejection should be designed into the receiver to provide maximum freedom from interference frequencies other than the reception frequency to which the receiver is tuned. When high level interference signals are present filter units or wave traps tunable to the radar frequencies may be designed and connected into the antenna lead-in wir[illegible] to suppress unwanted frequencies and at the same time offer a neglig[illegible] amount of attenuation to frequencies within the pass band of the re[illegible] Filtering is effective in p[illegible]nting the condition of free oscillati[illegible] caused by impulse type [illegible]rference, for it attenuates the radi[illegible] before they can shock [illegible]ite the first tuned circuit. An exam[illegible] type of filter used in [illegible] antenna circuit is illustrated in Fi[illegible] The attenuation freq[illegible]y bandwidth was sufficient to prev[illegible] ence due to sideban[illegible] quencies and carrier frequency v[illegible] e radar.

Wave tra[illegible] en used must be installed at th[illegible] [illegible]nna post or within the [illegible]ver case. Such wave traps or [illegible]nsist of parallel reson[illegible] [illegible]hoke coils or simply a quarte[illegible] [illegible]hoke coils are gen[illegible]ral[illegible] [illegible]erted in series with the rece[illegible] grid leads. Quarter wave stubs are generally connected between the receiver antenna post and receiver case with the far end open circuited. This provides satisfactory protection against very high or ultra high frequency interference below 600 mc. The stub is cut to quarter wave resonance at the frequency to be attenuated. Twisted wire pairs or parallel wires are convenient and practical for constructing the stubs. Wave traps made of a resonant stub using concentric transmission line will have a sharp cutoff characteristic and are not desirable for use against interfering energy coming in at a frequency of a few hundred megacycles.

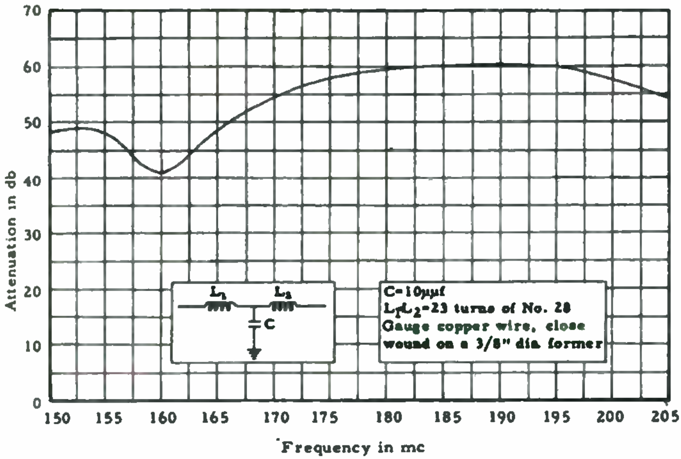


Figure 1-26. Antenna Filter Network for Suppressing Radar Interference in a Representative Communications Receiver

1.4.1.2 Power Leads

All primary power wiring in an electrical system is connected to a common bus bar. Interference sources always impress some portion of their output on the power wiring connected to them, unless completely filtered. Therefore, radio interference originating in the electrical and electronic equipments of the system, except the portion attenuated by the wiring, appears at the input to the receiver. Efficient removal requires knowledge of the frequency of the offending voltages. Protection against conducted interference over the power leads can best be controlled by suppression at the interference source and through good internal receiver design. When these means are improperly employed, some external controls are necessary. If the interfering frequencies fall within the bandpass of the receiver, a filter may be constructed for their removal, and inserted in the power line at the receiver plug.

Rotating equipment, radar, IFF, and similar equipments produce impulse type interference which may readily be conducted to the receiver by means of the power line. Network filters can be applied to the power line for attenuation purposes. The design procedure to be followed in constructing appropriate filters is found under Volume III, Chapter 4.

1.4.1.3 Control Cables

Electrical or mechanical control cables attached to a receiver, even though not connected to a source of interference, can readily act as an antenna, picking up interference by induction or radiation and transfering it to the receiver. Once they pass beyond the screening effect of the receiver case, there is nothing to prevent the interference from affecting the receiver output. Cables carrying pulse currents have little difficulty in injecting a portion of their energy into receiver lead wires, and consequently causing interference. For example, an overhead waterpipe or steel beam with radio frequency energy induced in it from a nearby transmitter can change from a normally good ground to act like a long wire antenna. At 28 mc there will be a high voltage point approximately every eight feet; that is, it will have standing waves on it and it will radiate. Any control cable greater than one-eighth wave length may, in a similar fashion, act as an antenna for its resonant frequency, and hence, there is some possible interference frequency at which it acts as an antenna.

This interference is eliminated by effective shielding of the control cable. Actually all leads carrying pulse currents should be shielded and routed separately. However, no interference path may be neglected.

The control leads should be grouped in a separate connector which incorporates an internal grounding ring for better grounding of the shields to the chassis. Feed-through capacitors located in the connector will help suppress high frequency interference.

Electrical control cables which terminate in the receiver and are run in a group of wires, may introduce interference signals into the receiver wiring through inductive or capacitive coupling. When feasible, isolation of cabling is effective; however, in some cases filters are required. The filter should be applied at the point of entry of the control cable to the receiver and may easily be incorporated in the original receiver design. Usually it is a low pass filter, and the frequencies to be filtered out may be ultra high, very high, or medium high frequencies depending on the frequency range of the receiver with which it is to be used. For this reason it is desirable to include it in the original receiver design. The low pass filter will not attenuate interference conducted to the receiver on the fundamental frequency of a radar transmitter. This will require a line filter of ultra high frequency design installed within the receiver.

1.4.1.4 Output Leads

Though designed to carry intelligence, output leads may also act as a pick-up for interference. Frequencies that gain access to the output circuit may be amplified sufficiently to become noticeable in the output. A receiver output circuit should incorporate design features that will minimize reverse coupling through the receiver stages and prevent amplification of unwanted frequencies. Output leads should be kept separate from any wiring carrying alternating current, and the compact construction of the receivers may make it advisable to shield the output system from the radio frequency section within the receiver case.

Once interference signals enter a receiver, they gain access to amplification stages in various ways and with varying degrees of attenuation. Tuned circuits may be susceptible to coupling with filament or relay wiring. Inadequate case and interstage shielding are coupling paths which also increase susceptibility. Production methods of cabling provide tight coupling between interference carrying wires and high impedance input and sensitivity control circuits. Receivers employing band switching appear to be more susceptible because of inadequate shielding between radio frequency coils.

The radio interference signals that gain access to the receiver case must eventually reach the last RF stage or second detector in order to adversely affect the receiver output. This internal interference path can generally be eliminated by proper design of the output stage. Reverse coupling either through the tube itself or through the associated circuitry, can be greatly reduced by careful selection of circuit components and internal stage shielding.

If the interference injected into the output leads is at radio frequencies, a simple by-pass capacitor is usually sufficient to prevent its entry into the receiver. Satisfactory operation is usually achieved by installing a capacitor between the output "hot" lead and the receiver case at the point where the lead enters the receiver.

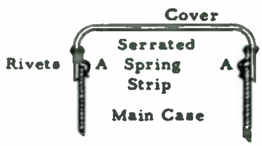
The use of filtering as well as shielding of the output leads is sometimes necessary to prevent radio frequency interference and any extraneous disturbances. Shielding is accomplished in the conventional manner.

1.4.1.5 Case Shielding

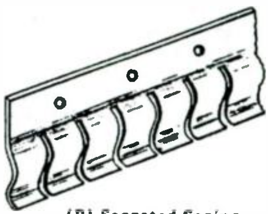
The primary purpose of the receiver case must be to shield the receiver from any external interference fields. The number of mechanical discontinuities must be kept to an absolute minimum, and those that are required must be electrically continuous across the interface. A multiple point, spring loaded contact is a very efficient method of obtaining electrical continuity.

Bonded screening of suitable conducting material must be used to cover all louvres and other apertures used for ventilation.

The multiple point, spring loaded contacts mentioned could be constructed in a number of ways. In general, a serrated shim inserted in the aperture of the discontinuity will be satisfactory. The serration gives enough spring pressure at its points of contact for electrical continuity, and no spring pressure is required at any other point on the shim. A sketch of a spring joint and a serrated spring is shown in Figure 1-27. The materials used in constructing the shims could be beryllium copper, German silver, phosphor-bronze, sheet steel, or tempered aluminum. The receiver case, however, should be constructed of the same material to prevent corrosion and resultant electrical discontinuity. If the materials are different, the shim must be protected by cadmium plating or alternative methods providing good electrical contact while preventing corrosion. This practice likewise conforms with service specifications.



(A) Serrated Spring Joint



(B) Serrated Spring

Figure 1-27. Multiple Point Spring Loaded Contact Joint

Receiver plugs must be free of paint or varnish between plug shell and receiver case, as well as between the plug and its shell. Such non-conducting materials at these points produce electrical discontinuities which could be a source of serious interference to normal receiver operation. The shield ground likewise must extend completely around the plug.

Shielding effectiveness also hinges on the thickness of the shield and the type of material used. The depth of penetration of interference currents in the metal wall is discussed in Volume III, Chapter 6. An excellent degree of attenuation is a definite possibility using thin walled ferrous metal shields of high effective permeability, and plated with non-ferrous metals of high conductivity. Plating the shield with a non-ferrous metal increases its reflection loss.

1.4.2 SPECIFIC CIRCUITS

Where specific circuits are devised for coping with a certain type of interference, they may be incorporated in any receiver if that receiver is susceptible to that interference type. For example, a gated amplifier is used to operate a radar receiver only during selected intervals of time. This same circuit may be used to block a counting circuit during an interfering pulse, as in the case of an altimeter. Such pulses may also be used as blanking pulses to turn off the beam of a cathode-ray tube, in order to eliminate a horizontal retrace line. A basic circuit performs a basic function. The circuit may be used wherever there is need for that function.

1.4.2.1 Radio Receivers

The tuned circuit of a radio frequency input stage is essentially a band pass filter. It usually has the characteristic of high selectivity, passing the frequencies within its acceptance band, while rejecting all others. When unwanted signals or interference, comparable in amplitude to the desired signals, appear in the frequency ranges adjacent to the pass band, they are effectively attenuated by the tuned circuit. Succeeding stages will further attenuate the minor amount of interference which does pass through the first stage. However, if the interference is very strong, of the order of volts as it may well be from a neighboring highpowered radar unit, the received impulse will produce ringing of the tuned circuit at its natural resonant frequency. This effect is passed along and amplified in conjunction with the normal signals, and the interference will appear in the output. The same is true of extremely strong radiation from a radio transmitter operating on an adjacent channel. Even though the transmitted energy is attenuated by the tuned circuit of the receiver, the response is still high within the pass band, and the familiar effect known as cross-talk occurs in the output. Therefore, it is necessary to provide for interference reduction by design techniques.

By using more than one radio frequency stage, additional selectivity may be provided between the antenna input and the converter unit. It is possible to increase both selectivity and fidelity by adding more stages, in cascade. This involves broader tuning of the individual stages to avoid a loss of fidelity by using compact coils wound with relatively small wire. The resultant reduced Q causes a loss in gain which is offset by the additional number of stages. A simple arrangement for increasing selectivity, by means of an extra tuned circuit, is illustrated in Figure 1-28.

Such circuits will also provide isolation of the oscillator stage and help prevent radio frequency energy at the oscillator frequency from reaching the antenna and causing undesirable radiation. However, they do not satisfactorily suppress interference consisting of strong pulses. A wave trap in the antenna circuit is required to properly eliminate strong interfering pulses.

Limiter circuits may be incorporated in receivers with the following limitations:

a. They are suitable only for use in receivers which have some radio frequency gain and rectify at a fairly high level, at least 0.1 volt.

b. Limiters are useful for suppressing low pulse recurrence frequencies below 100 times a second, but are less effective for suppressing those up to 500 times a second.

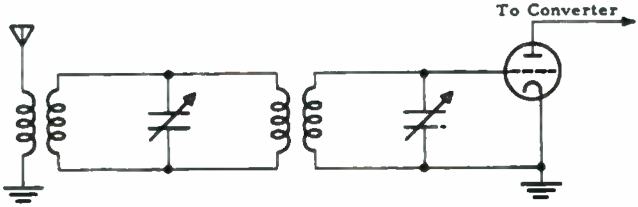


Figure 1-28. Additional Tuned Circuit Preceding Converter

The length of an interfering pulse may be considerably increased as it passes through successive stages of a receiver. This is a direct result of the tuned circuits "ringing" when subjected to pulse excitation. The time, t_o, in microseconds which it takes for the amplitude of the pulse to fall to approximately 4 percent of its initial value is given by the reciprocal of the bandwidth in megacycles of the tuned circuit:

$$t_o \text{ (microseconds)} = 1/(\text{Bandwidth mc})$$

The ringing waveform in combination with the local oscillator frequency will then produce a ring at the intermediate frequency of the receiver and appear in the output.

A circuit required to pass short pulses without distortion must be able to pass a wide range of frequencies. For example, a video pulse amplifier must have a reasonably flat frequency response up to high frequencies. If the bandwidth of a tuned circuit is large, the pulse will not be lengthened to any marked extent, and, because the pulse is of short duration, the limiting action will be much more effective. The pulse that is superimposed on the desired signal, is prevented from reaching the audio section by the action of the limiter. At the same time there is negligible distortion of the desired signal due to the short time interval of the pulse as compared to the period of the signal.

However, unavoidable interference power in the output of an amplifier increases in proportion to the bandwidth. Furthermore, the gain per stage in an amplifier is, in general, inversely related to bandwidth, so that for a given overall amplification a broadband amplifier requires more stages than one with a narrower band. It thus becomes important to judge the best bandwidth for a particular application.

Pulses with low repetition rates are adequately suppressed by limiters while higher recurrence rates would cause a proportionately larger amount of interference at the output terminals. Thus, low repetition rate and a large bandwidth represent ideal conditions under which to operate limiters.

The second detector is the first part of a receiver where an amplitude limiter can effectively be placed. The radio frequency voltages at the input to the receiver are too small to operate any known forms of limiter or rectifier. Usually the interfering pulse at the second detector has not become too long for effective limiting. Pulse lengthening is a function of initial energy and the bandwidth of intervening circuits. Limiters will require less operating time with low initial amplitudes of an interfering pulse and greater bandwidth in the tuned circuits. The duration of the pulse at the point of limiting is a very important factor because a portion of the desired signal is affected every time the limiter comes into action.

Limiters are primarily restrictive devices and distortion will result from their use, particularly when the input exceeds the limiting threshold. Limiters of the instantaneous interference peak type generally distort the output whenever the modulation of the incoming signal exceeds a definite percentage. The distortion effects can be intensified by the transient distortion characteristics of the audio amplifier. In general, it is desirable to use triode tubes in the audio amplifier or degenerative feedback sufficient to prevent oscillations because of insufficient damping of the output circuit.

When both a modulated carrier and pulses of interference are present at the output of the final intermediate amplifier, then the current through the audio frequency output resistor is of the form shown in Figure 1-29. If the cutoff point is adjusted to coincide with the negative peaks of the audio waveform then, as observed in the figure, maximum attenuation of the interfering pulse would occur with minimum distortion of the audio waveform. Under the ideal condition of wide bandwidth, as is commonly encountered in very high frequency receivers, a series limiter provides an average attenuation of 30 to 35 db of the unwanted pulse. These limiters also offer some degree of protection in receivers operating in the lower frequencies and at their widest possible band acceptance. Amplitude limiting gives a valuable degree of protection against atmospheric and ignition interference.

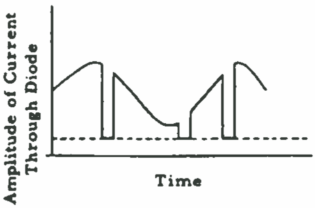


Figure 1-29. Current Through Limiting Diode Showing Limiting Action on Interfering Pulses

Parallel limiters are not as effective as series limiters, and under the same ideal conditions mentioned previously, provide an attenuation generally of about 20 db of the interfering pulse. Furthermore, they do produce distortion of the audio waveform since they reduce the output by shunting action, but are not a complete short circuit. The time, t_o, during which distortion occurs, varies inversely with the percent of modulation. In general, it is less than 0.1 millisecond above 50 percent modulation, as may be observed in the typical graph of Figure 1-30.

There are cases where interference pulses with frequencies of the order of 200 megacycles penetrate the receiver case or enter through external leads and are internally coupled to the audio frequency amplifiers at sufficient amplitude to result in grid circuit detection of the pulse. In such cases satisfactory suppression is obtained by the use of a resistance capacitance decoupling network. This combination has very little effect upon the audio frequency signals, but will greatly attenuate the interfering energy due to the low input impedance of the grid circuit at high frequencies. A typical circuit is shown in Figure 1-31.

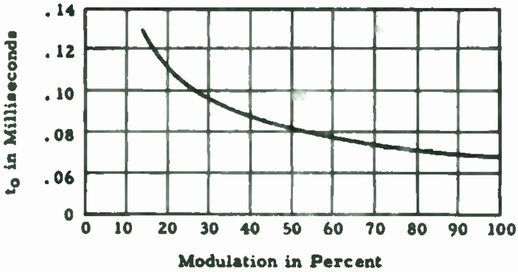


Figure 1-30. Distortion Interval of Audio Waveform Measured Against Percent Modulation

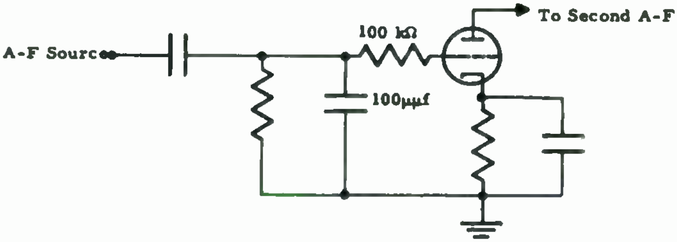


Figure 1-31. R-C Decoupling Network at Input to First AF Stage

It is quite possible for the interference entering through the amplifier case or external leads to the amplifier to be present in any part of the amplifier circuit. This necessitates precautions in each stage of the amplifier. Generally, the first stage is most important since greater gain results in subsequent stages. The blocking resistor of the resistance capacitance network is in series with the grid and located as close as possible to the tube terminal. The use of a short connecting lead wire will minimize interference injection between the series resistor and the tube. This combination is similar to an L-type filter with the inductance replaced by a resistance. The resistance should be designed to be much greater than the reactance of the capacitor for the interference frequencies. A ratio of 10:1 is useful in practice. This permits most of the interfering energy to be shunted to ground.

Resistance capacitance networks are usefully employed as decoupling networks in the plate circuits of an amplifier to prevent interstage coupling and possible oscillations. The voltage output obtained from a common B supply is not fixed but varies with the current demand. Also, any ripple appearing at the output of the power supply filter is impressed on all the amplifier grids, except the first. When using a decoupler the voltage across the condenser is very nearly constant and independent of any power supply variations. The networks act independently of one another and thus isolate the stages. A cascade arrangement, where the first stage possesses the largest amount of coupling, is illustrated in Figure 1-32.

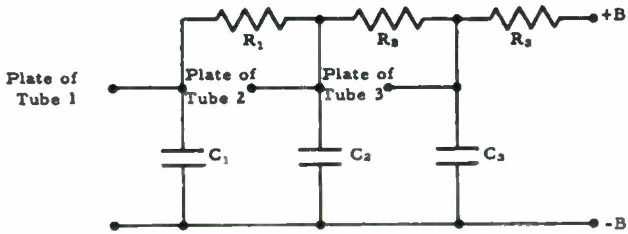


Figure 1-32. Decoupling Networks in Cascade

The problem of interference rejection in radar receivers is somewhat different from that encountered in communication receivers, since radar interference is nearly always the result of signals occurring so close to the radar frequency that very little can be done by improving receiver selectivity. The ability to deliver intelligible information to the radar indicators in the presence of on-frequency interference gives a better indication of the radar's quality than its selectivity in the usual sense of the word, i.e., the ability to reject disturbances at frequencies other than the desired signal frequency.

Most receivers have sufficient amplification to give appreciable interference outputs in the absence of a received signal, for which reason the sensitivity of a radar receiver is nearly always determined by its susceptibility to interference. Furthermore, amplitude modulated interference is of greater concern in radar receivers than frequency modulated interference because amplitude modulation envelopes approximate rectangular pulses in shape. Since adequate detection of rectangular envelopes depends more on receiver bandwidth and phase shift than on linearity of amplitude response, nonlinear amplitude response in the form of limiting is permissible in most radar receivers. The limiter suppresses the amplitude modulated interference and produces a visual output relatively free from interfering signals, even though the frequency modulated interference actually may be increased by the action of the limiter.

Three main types of receivers are used in radar applications which differ in regard to the frequency region within which most of the necessary signal amplification results: (1) superheterodyne receivers which convert the modulated radio frequency signals to an intermediate level, at about 30 megacycles, before amplification, (2) super-regenerative receivers, which use a regenerative radio frequency amplifier, with the oscillations quenched in a time interval about equal to a pulse width, and (3) crystal video receivers, which detect the modulation signals and amplify the resulting video signals. The majority of microwave radar receivers are of the superheterodyne type since this permits the largest amount of amplification to take place in a fixed tuned amplifier.

At frequencies below the microwave region, the first detector may be preceded by one or more stages of radio frequency amplification. The crystal mixer may be replaced by a vacuum tube mixer. The additional gain obtained from the radio frequency stages results in better image rejection, improved signal-to-interference ratio, and reduction of the radiation of local oscillator power.

Most radar systems today operate in the microwave region, and mixers for use at frequencies higher than 3000 mc are usually of the waveguide rather than coaxial type. At these frequencies the local oscillator contributes rather serious interference energy which can be sharply reduced by using a balanced mixer. One form of balanced mixer is the waveguide magic-tee, as illustrated in Figure 1-33, with the crystals installed in the arms of the waveguide, parallel to the electric field.

The arrows in the figure indicate the direction of the electric field in each arm. A waveguide type of mixer divides the local oscillator power equally between the two arms containing crystals A and B and prevents energy transmission out of the signal input arm. A simple and rather obvious explanation for such a characteristic follows directly from the geometrical arrangement of the four arms. The arms containing the crystal mixers and the local oscillator input arm form a shunt tee and provide for the continuity of the local oscillator electric field between the arms. Similarly, the crystal arms and the signal input arm form a series tee and provide for the continuity of the signal input magnetic field between the arms. Thus, the local oscillator and the signal input powers divide equally between the crystal arms when the impedances of these arms are matched. However, since there is no provision for continuity of either electric or magnetic components of the transverse electric wave between the signal input arm and the local oscillator arm, no energy can be transferred from one to the other. Therefore, local oscillator radiation is minimized through the use of a magic-tee balanced mixer.

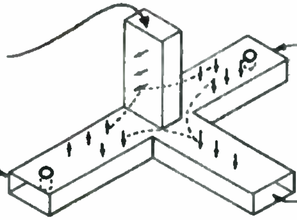


Figure 1-33. Magic-Tee Balanced Mixer

A crystal mixer has a gain of less than unity, making it mandatory to control the interference developed at the input stages of the intermediate frequency amplifier. Triodes are preferred to pentodes in these stages because of their lower interference characteristics, but they cannot always be used because of their large input capacitance which leads to extremely poor performance at frequencies as high as 30 megacycles. In the circuit shown in Figure 1-34 triodes are used in the intermediate frequency 30 megacycle amplifiers. The triode connected 6AK5 vacuum tube has a load impedance of about 200 ohms presented by the cathode of the second tube. This low impedance secures stability of the first tube, whereas in the second tube stability results from the grounded grid, shielding the input (on the cathode) from the output. The neutralizing coil between plate and grid of the first tube confers extra stability. It helps to prevent the output impedance of the first tube from falling off, and results in minimizing the interference from the second tube. This circuit yields an intermediate frequency interference figure about 2 db lower than a similar intermediate frequency input employing a pentode.

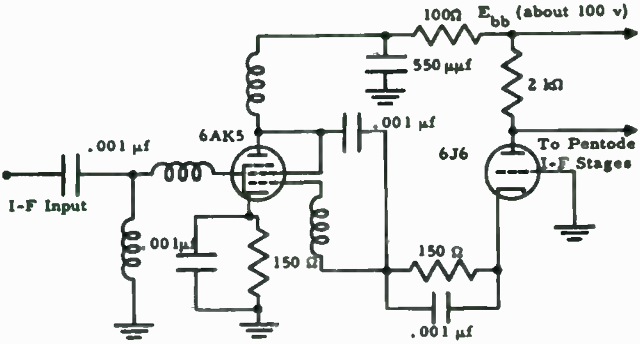


Figure 1-34. Diagram of Grounded-Cathode, Grounded-Grid Dual Triode Input for 30 MC IF Amplifier

Since interference in the output of an amplifier increases in direct proportion to the bandwidth, it is important to restrict the intermediate frequency bandwidth as much as possible. It must be sufficient, however, for adequate transmission of the signal. If a radar system is employed for search purposes, then the visibility of an echo in the presence of interference is the primary measure of performance. In that case the best circuit bandwidth has been found to be 1/T cps, where T is the pulse width. This results in an intermediate frequency amplifier bandwidth of 2/T cps in order to transmit the double sideband signal at this point. Thus, certain fire control radar equipments, making use of the leading or lagging edge of a received pulse for precise range measurements, require a broader bandwidth to assure a minimum rise time of the displayed pulse and consequently a more precise determination of the position of the pulse. For example, a radar pulse width of about 5 microseconds would require a bandwidth of $1/5 \times 10^6$ cps or 0.2 mc. A short pulse of 1/5 microsecond duration, however, requires a bandwidth of 5 mc.

Consideration must be given to the behavior of the radar receiver in the presence of excessive signal strength, as from neighboring radar or "jamming" signals. Intentional jamming may be of the "window" type, that is, it may be caused by strips of metallic foil which cause numerous fluctuating echoes and thus obscure the presence of aircraft or it may be caused by radio waves, either modulated or unmodulated from jamming transmitters. The most common form of modulation is that by long pulses, termed "railings" because of their appearance on a scope. Accidental jamming may be caused by strong echoes from land targets, rough water surfaces, or clouds. It may also be caused by other high frequency equipments, and if so will appear either as continuous wave or as "railings" jamming.

The main purpose of anti-jamming circuits is to prevent receiver saturation. Manual adjustments are impractical because it is quite impossible to follow the rapid fluctuations in jamming with a manual control. A signal of 80 db above normal receiver interference level coupled with 50 db of clutter would appear as a 30 db signal if the receiver gain were properly reduced. With normal gain, however, the signal would be invisible since the clutter would saturate the receiving system. This is particularly true in the case of an intensity modulated indicator, such as a plan position indicator. When the beam intensity is increased too much the focus is destroyed, and the spot is said to "bloom." The plan position indicators have a small dynamic range, around 10 to 20 db, and therefore require a limiter stage in the preceding video amplifier. This prevents strong signals from causing blooming.

The following are four types of circuits useful against jamming and clutter:

a. Sensitivity time control circuit. These circuits control the receiver gain as a function of time after the initial radar pulse. The gain is reduced when the radar pulse is first sent out and then gradually increased to normal value as determined by the time constants of the receiver. Since gain is made a function of distance, this circuit is useful only when a desired echo is greater in amplitude than the interference echoes at all ranges, and when their ratio is maintained for increasing ranges. This is possible only when the interference source is a target without strong directional characteristics, such as sea or land surfaces.

b. Automatic gain control circuits. An instantaneous automatic gain control rapidly decreases the gain of an intermediate frequency stage when the stage output increases beyond a value determined by the circuit constants. This action prevents stage saturation. It is advisable to protect the last two or three IF stages with this type of control. The circuit usually operates with a time constant of about 20 microseconds.

c. Short-time-constant networks. The coupling between the second detector and first video stage is provided with a very short-time-constant network. The network serves to remove or attenuate, by differentiator action, the dc and low frequency components encountered in continuous wave or low frequency modulated jamming. The time constant is usually made equal to the radar pulse width.

d. Bias-control circuits. These circuits automatically supply a bias to the second detector which prevents the high frequency components of interference modulated jamming or clutter from saturating the video section. The circuit can be designed with a short-time constant. For this reason a delay network is also necessary so that individual signals will not be reduced in amplitude, i. e., cut off too soon.

The short-time constant and the bias-control circuits are most effective when used in conjunction with the instantaneous automatic gain control circuit. At frequencies below the points where the short-time constant circuit cuts off, the fast time constant and instantaneous gain control are an effective combination against jamming by modulated or unmodulated continuous waves. Modulated jamming and most types of clutter, especially that caused by clouds, are best controlled by a combination of bias and instantaneous automatic gain control.

High power pulsed radar systems operating near each other can readily cause mutual interference, even though there is considerable separation of their operating frequencies. The radar receiver does not offer sufficient attenuation for extremely strong off-frequency signals. Interference may also be caused by pickup of the large video signals radiated from a nearby modulator. Blanking circuits have offered the most effective solution to this type of interference. A receiver gating pulse is developed and applied to one or more intermediate frequency grids and synchronized with the transmitted pulse of the interfering set.

Receiver gating can be accomplished in the intermediate frequency or video section. The stage may be cut off by reducing the plate or screen voltages, by making the suppressor or control grid voltages negative, or the cathode voltage positive. When applied to an intermediate frequency stage, the gating pulse does not produce any disturbance at the receiver output, since the amplifier cannot amplify the frequencies contained in the pulse. However, if a video amplifier is gated, the gating pulse produces a pedestal in the output on which the normal signals ride. This occurs because the video amplifier has an appreciable response in the range of frequencies contained in the pulse. When the pedestal is present in the output, it may be removed if necessary. For example, a pedestal is not permissible when signals are applied to an intensity modulated cathode ray tube since such a tube is not normally biased beyond visual cutoff.

A non-pedestaling cathode-gated video stage is illustrated in Figure 1-35. There is a gain of about unity and sufficient bandwidth to handle pulses of about one microsecond. The stage is capable of handling positive or negative signals of a few volts amplitude. A 6SN7 multivibrator (Eccles-Jordan circuit) supplies the gating pulse. V_{2A} and V_{2B} are cathode followers for driving the 6SL7 cathode, while V_{3B} carries the output plate current when V_{3A} is gated off. When V_{1A} is conducting, the positive potential on the cathode of V_{3B} rises sufficiently to cut the tube off. When V_{1B} is conducting, then V_{3A} is similarly cut off.

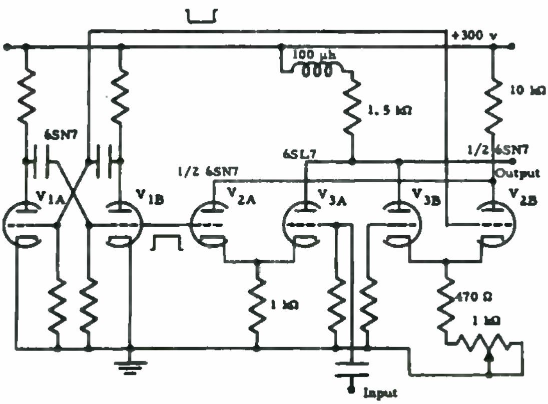


Figure 1-35. Cathode-Gated 6SL7 Video Stage, with no Pedestal in Output

1.4.3 DESIGN CONSIDERATIONS FOR MINIMUM OSCILLATOR RADIATION

Receivers must not be allowed to radiate signals whose frequency will adversely affect the basic function of other electronic equipments. There are several sources of incidental radiation in receivers, but by far the most serious is the high frequency local oscillator radiation of a receiving system. This is particularly true when the receivers operate in the very high frequency range or higher, since the local oscillator frequency is removed from the carrier by only a small percentage. Radiation interference from oscillators can be broken down to two types: antenna radiation and chassis radiation. Existing literature deals predominantly with antenna radiation with little mention made of the problems of chassis radiation. In the lower frequency ranges, where the chassis is small compared to a half wavelength, chassis radiation is no problem. However, when the chassis size is such as to approach half wave resonance at a particular operating frequency, it becomes an efficient radiator. This is true in the case of television receivers operating in the frequency range of 174 to 216 mc. The chassis usually is big enough to approach half wave resonance.

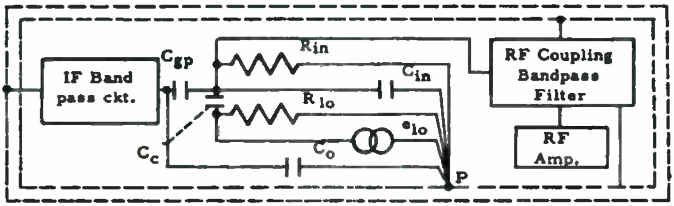
1.4.3.1 Chassis Radiation

Certain measures designed to minimize oscillator radiation are effective for both the antenna and chassis types of radiation. The appropriate technique of shielding will aid in confining the local oscillator energy so that reverse transmission by means of the antenna or lead-in will not occur. It will also prevent the excitation of the chassis, and thus prevent chassis radiation. Triggering of Airways Marker Beacon receivers by television local oscillator radiation has occurred and indicates that this form of undesirable signal radiation does take place. The local oscillator of a television receiver is about 21 mc above the station being tuned in and when channel 2 is used, 54-60 mc, the local oscillator frequency is within the region of 75 mc, which is the frequency of Airways Marker Beacons.

The problem of chassis radiation dictates the need to prevent the chassis from becoming excited by the local oscillator energy or to totally reflect the energy radiated if the chassis is excited. A metallic receiver case will aid in reflecting energy which is radiated by the chassis, but the case should be looked upon only as a secondary or outer shield, and emphasis placed on confining this energy to the region of the local oscillator itself. The metal case functions primarily to permit an interference free region within.

A high degree of shielding is required in the radio frequency section of receivers for the purpose of reducing their susceptibility to interference. It only remains to extend the principle of shielding to include the problem of oscillator radiation.

Figure 1-36 illustrates an equivalent circuit of shielded local oscillator. To restrict the electromagnetic fields to the immediate vicinity of the local oscillator, the oscillator and mixer circuits must be inclosed in a conducting shield which is as continuous as possible.



e_{lo} = Open circuit local Oscil. Voltage
R_{lo} = External damping on local oscil.
C_c = Coupling reactance to mixer grid
C_{in} = Input cap. of mixer
R_{in} = Input resistance of mixer circuit and tube
C_{gp} = Mixer grid plate capacitance
C_o = Mixer output capacitance

Figure 1-36. Shielded Local Oscillator Equivalent Circuit

The shield will have to be fastened to the next larger support member at points of equipotential to minimize excitation of the larger surfaces. All power supply leads entering the shielded area must be filtered for radio frequency disturbances. The oscillator must be designed such that its electromagnetic fields will produce a minimum of current in the shield material. This may be accomplished by (1) the use of a single point ground for the entire oscillator circuit, as illustrated in Figure 1-37, (2) orienting the oscillator coil so that its field induces minimum current in the surrounding metal, and (3) restricting the field due to the oscillator coil by its own individual shield or vestigial shielding such as a shorted turn surrounding the shield. Since the local oscillator drives the mixer or converter vacuum tube, it is necessary to design the band pass networks associated with the mixer tube for minimum transmission at the oscillator frequencies.

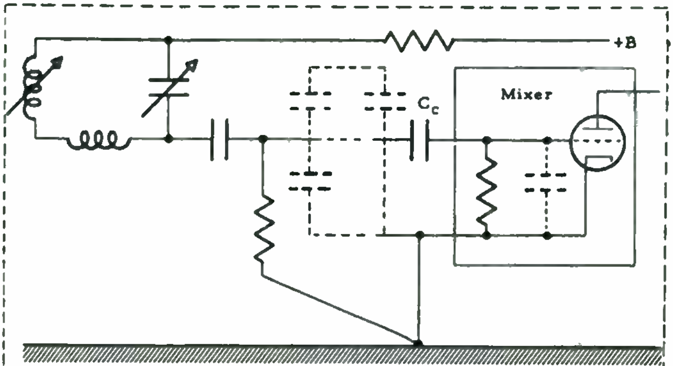
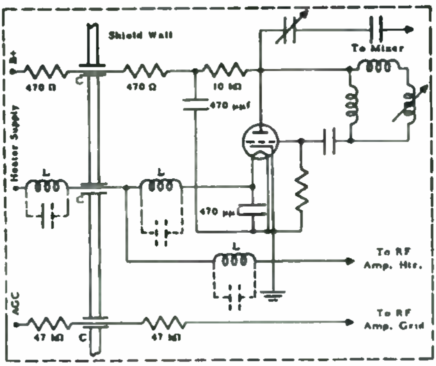


Figure 1-37. Good Oscillator Design Reduces Chassis Currents

The best approach in designing an oscillator shield is to enclose the oscillator circuit in a continuously shielded case, using soldered, "water-tight" joints, and even the use of double shielding, if necessary. Where there may be holes and slots in the shield, a difference of potential will exist across them. By proper orientation of the holes and slots, the electrostatic field may adequately be confined. However, this is not true for the electromagnetic field. Magnetic lines of force bulge through any opening and will excite the exterior of the oscillator shield. This, in turn, will excite the main chassis, which will act as a fairly efficient radiator. Holes and slots must be avoided. Where supply leads enter or leave the shielded compartment, they must be prevented from acting as a path of exit for the magnetic lines of force. There must be adequate low pass networks in the power supply leads. In the case of high voltage and automatic gain control leads, a series resistor shunt capacitor combination is used. The capacitor must not experience any anti-resonant effects within the tuning range of the oscillator. An excellent arrangement of a network designed to prevent spurious resonances within the oscillator tuning range of ordinary television receivers is shown in Figure 1-38. The use of a few hundred ohms in the output and input of this network helps to prevent resonance occurring in the supply leads exterior to the tuner. L-C networks are usually required in the filament leads to prevent excessive voltage drop. They are designed to prevent any spurious resonances in the tuning range of the oscillator. The capacitors shown in the figure are of the feed-through type.



Note: C = 500 μμf Low Inductance Type Feed-Through Capacitor First Anti-Resonance Above 300 mc
L = 1 μh with C distributed ≤ 0.5 μμf
R = All 1/2 Watt with Very Short Leads to Feed-Thru Capacitors

Figure 1-38. Network Arrangement to Prevent Spurious Resonances in Oscillator Range

Coupling between the magnetic field of the local oscillator coil and the shield or chassis must be held to a minimum. This can be accomplished by a high ratio of length to diameter, by a high permeability core which helps to confine the magnetic field, and proper spacing from the chassis. The spacing should not be less than two coil diameters. There is usually a rapid increase in chassis radiation at the high end of the tuning range when using permeability tuners, due to the field extending farther in space when the core is removed from the coil. To properly confine the magnetic field of the coil it is necessary to use both eddy current and permeability shielding. A combination of magnetic and non-ferrous conducting materials will serve the purpose. High conductivity metals are desirable since smaller thicknesses are necessary for a given attenuation of a confined field. The thinner shields also permit the formation of good joints in the shielding. A metal thickness of about ten times the depth of field penetration will produce an attenuation of approximately 86 db in the field intensity. At a minimum frequency of 80 mc the minimum amount of copper required would be 0.003 inches.

The oscillator shield joints must have as large an overlap as possible to prevent leakage. This may be accomplished by screws or spring pressure (multiple contact type) which assures continuity in the shielding. If overlays of copper are used, the shield should be formed with the copper on the inside. Cold rolled steel plated with copper offers adequate attenuation, but requires greater thickness of shield, and is subject to corrosion. If tolerances are of necessity loose, the greatest possible contact is secured in a practical way by the use of a metal textile gasket, as illustrated in Figure 1-39.

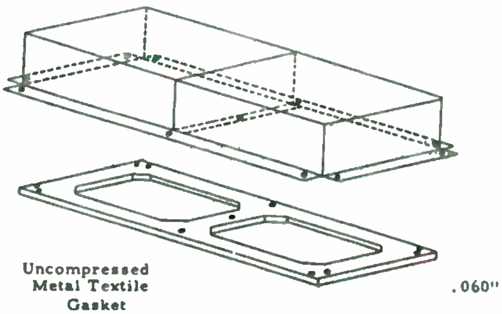


Figure 1-39. Typical Shield Construction

1.4.3.2 Antenna Radiation

Antenna radiation can adequately be suppressed by additional radio frequency stages, extra tuned link circuits, proper shielding and filtering, adequate RF by-passing, and the use of short leads and compact layout. A well screened radio frequency amplifier, together with proper shielding, prevents the energy of the oscillator coil from getting into the radio frequency amplifier grid circuit. The shielding processes employed by manufacturers of very high frequency standard signal generators represent some of the best present day techniques for the purpose of confining local oscillator energy.

However, a well shielded and properly filtered receiver is frequently a source of radio interference due to the local oscillator voltages appearing on the antenna. This generally occurs when all the possible paths for interference to reach the antenna are not taken into consideration during the design of the receiver.

This difficulty cannot be corrected simply by improving the shields or filters; since the local oscillator output is coupled to the antenna by the circuit elements themselves, rather than through improper shielding or lead filtering. An optimum design must be achieved to reduce the effective coupling between the oscillator and antenna terminal through the mixer and RF stages, and still not adversely affect the operating characteristic of the converter stage. In general, additional RF stages are the most effective means available to accomplish this. Refer to Section 1.4.4 for a detailed treatment of the use of additional RF stages in reducing local oscillator radiation.

A converter section of a typical receiver that was producing high level interference signals is considered herein to illustrate the problems and techniques involved in reducing radio interference caused by local oscillator antenna radiation. Figure 1-40 shows a simplified schematic diagram of the converter section of a typical radio receiver. The paths over which the local oscillator signals travel to reach the high and low band antenna terminals are represented by a series of dots.

The oscillator signal at the antenna terminal may be reduced in several ways. High "Q's" of the signal frequency tank circuit help to attenuate the oscillator signal appearing on the antenna. A limit is set by the increased difficulty of tracking and alignment of the various stages with increase in "Q." This limit was actually reached in the redesign of

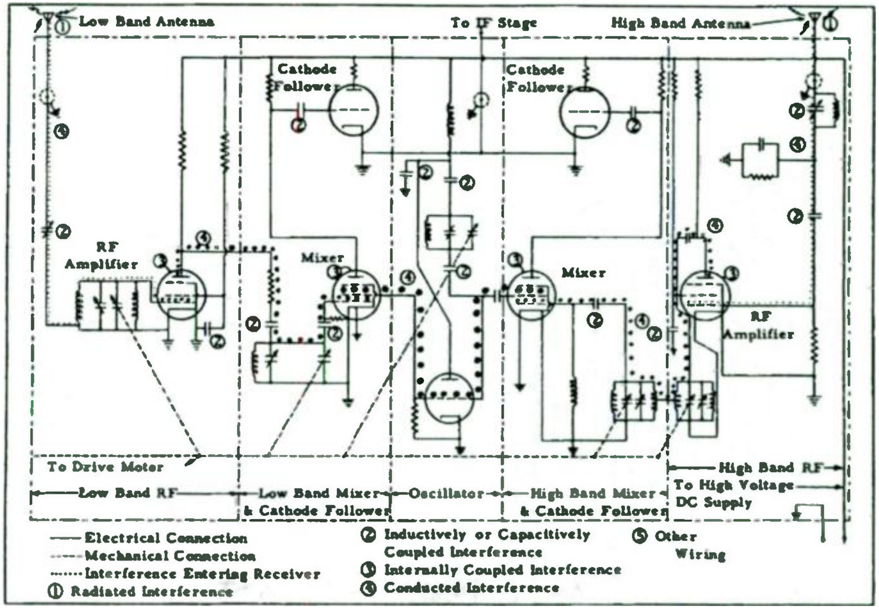


Figure 1-40. Path of Local Oscillator Interference Signals in the Converter Section of a Typical Scan Type Receiver

the preselector in the sample receiver under consideration and alignment difficulties due to the sharpness of the preselector were considerable. The "Q" of the mixer tank should also be increased, particularly in the frequency range where the largest oscillator signal on the antenna occurs. Test results showed that because of the above limitation only minor improvement could be obtained through increased "Q's" in the circuitry.

The oscillator is capacitively coupled to both the low and high band mixer circuits. Reduction of this coupling would certainly reduce the interference appearing at the antenna terminals. However, since this capacitor also couples the oscillator frequency into the mixer stages for superheterodyne receiver operation, excessive reduction of coupling adversely affects the receiver operation.

The following chart shows the effect of varying the coupling capacitor between the oscillator and mixer circuits on oscillator radiation and sensitivity in the typical converter section.

Coupling Capacitor	Oscillator Signal	Sensitivity
10.0 μμf	28,000 μv	10 μv
2.5 μμf	8,000 μv	14 μv
1.0 μμf	3,000 μv	25 μv

This clearly demonstrates that considerable oscillator attenuation by this method results in an appreciable loss of sensitivity. For this reason only minor improvement can be achieved through network coupling change.

Oscillation weakening by reduction of plate voltage, or by any other means, is in a class similar to the coupling problem. If the output of the local oscillator is reduced, the interference characteristics are improved at the sacrifice of performance. In fact, there is danger of the oscillator failing to oscillate at the low frequency end especially with a weak oscillator tube.

Reference to Figure 1-40 shows a coupling loop between the oscillator and high band RF stages. It is conceivable that sufficient capacity between the tank inductance and the coupling loop could affect the oscillator coupling. Tests performed on the typical receiver showed that this capacity was sufficient to produce a near-null in voltage at the center of the loop. However, further tests revealed that capacity coupling of the loop to circuit elements resulted in opposing voltages which tend to cancel, and reduce the overall coupling effect. The effect of the total capacity between loop and coil was also investigated by increasing the capacity of one end of the tank coil to ground, thus increasing the voltage between tank coil and loop, also destroying the cancellation mentioned above. Since no appreciable change in oscillator coupling was observed, it was concluded that the total coil to loop capacity, even without the benefit of the cancellation effect, produces negligible interstage coupling.

The above measures apparently must be considered as second order effects only. A device is required that will attenuate the oscillator voltage appearing on the antenna terminals without adversely affecting the sensitivity or other desired characteristics of the converter. Another RF amplifier stage should accomplish this since the sensitivity is improved by the additional stage, and the plate to control grid or screen grid to control grid capacitance coupling is relatively low. The choice of electron tube to be employed is quite important since the capacitance coupling varies with different tube types.

The following list shows the design considerations that produce the maximum reduction in the oscillator radiation without adversely affecting the overall operation:

a. pentode RF stage,

b. screen injection of oscillator signal into mixer,

c. connecting coupling lead near bottom end of the grounded tank plate circuit coil and near bottom of the grid tank coil of the high band mixer,

d. shielding between the input and output sides of the pentode socket.

These considerations serve to emphasize the fact that considerable time and expense can be saved by application of sound interference free design techniques in the original design rather than attempting to correct an interfering component or system once it has been installed.

INTERFERENCE CONSIDERATIONS IN THE DESIGN OF AEROSPACE SYSTEMS — CHAPTER 2

1. GENERAL ASPECTS OF AEROSPACE VEHICLE RFI

To solve the problem of the adverse effect caused by electromagnetic interference interacting with aerospace vehicle (aircraft, missile, and satellite) electronic circuits, it is necessary to classify the interference according to its source.

Different solutions are required depending on the source of the radio frequency (electromagnetic) interference. A primary source of RFI in an aerospace vehicle is that which originates from equipments inside the vehicle. This interference can degrade the performance of any susceptible circuit in the system. Problems associated with this type of interference will be discussed as system self-compatibility problems.

A second source of RFI for an aerospace vehicle is externally generated interference from other communications-electronics (C - E) equipments outside of the vehicle. Because the entry points of this type of interference are limited, only certain of the susceptible circuits can be degraded. Problems associated with this type of interference will be discussed as system-to-system compatibility problems.

Sources of RFI originating from outside the vehicle but not emanating from C-E systems will be discussed as other sources of electromagnetic interference. Signals from unfriendly C-E systems will not be considered directly since the philosophy for combating these signals is not the primary concern of RFI technology.

1.1 PROBLEMS OF SYSTEM SELF-COMPATIBILITY

At present, one of the most important unsolved problems in the field of systems engineering is the design of a complete aerospace and/or weapons system so that all the components and sub-systems are mutually compatible in every respect. For example, one area where compatibility is required concerns the interfaces between sub-systems. Because problems in this area directly concern the proper functioning of the system in an easily observable way, special engineering effort is now devoted to assuring compatibility between sub-systems as the design progresses rather than attempting remedial fixes after the sub-systems are joined together. Even with this effort, many compatibility problems still arise when an attempt is made to operate the whole system as a unit.

1.1.1 NECESSITY FOR EARLY RFI DESIGN COORDINATION

The coordinating efforts of the interface and general systems engineers deal largely with impedance and signal levels, information and control flow, and power problems. Very little attention is given to radio frequency interference problems created by radiating or susceptible subsystems and the detrimental effect that these phenomena may have on the system as a whole. As a result, system self-compatibility problems arise which are caused by RFI. Since many engineers do not consciously consider the possibilities of such interference, the system malfunction may not be attributed to RFI and may, in fact, be eliminated by cut and try methods without the engineers even being aware of the existence of an RFI problem or an RFI technology which could solve it. Fortunately, this situation is less likely to occur as more and more system engineers and designers become aware of the likelihood of RFI problems occurring.

1.1.2 CONSIDERATION OF RFI SPECIFICATIONS DURING DESIGN

The introduction of military specifications such as MIL-I-26600 and others dealing with acceptable levels of conducted interference, radiation, and susceptibility has done much to make systems and design engineers aware of the consequences of RFI. Unfortunately, in many instances, the system is designed and constructed without any thought being given to the applicable RFI specification until after the system is completed and its functional requirements have been met. When the finalized equipment is then tested, it usually fails to meet the RFI specification. Oftentimes, to meet the specification, a modification of the equipment is required which entails considerable time and expense.

1.1.3 NEED FOR TRAINING DESIGN ENGINEERS IN RFI

It becomes obvious that one of the first steps in the effective control of RFI in aerospace and/or weapons systems design is the education of the design engineer regarding the consequences of RFI and the preventive measures necessary to minimize it. At present, much more is known about the consequences of RFI than about its prevention and elimination, and so the objective of this chapter is to extend the scope of RFI technology in the areas of prevention and elimination.

1.1.4 INTERPRETATION OF SPECIFICATIONS REQUIRED TO AVOID DESIGN PITFALLS

Studies regarding RFI suppression techniques, show apparent deficiencies exist in specifications which deal with the measurement of radiation and susceptibility. As an example of the problem presented to a systems engineer who is seriously interested in preventing RFI by proper original design, rather than by remedial measures, consider two paragraphs from MIL-I-26600, and their implications discussed below as they affect the designer. (See Volume IV, for the complete specification.)

The first paragraph is (4. 3. 2) Radiated Interference; it states:

"Radiated interference fields in excess of the values given in figures 6, 7, 8 and 9 shall not radiate from any unit, cable (including control, pulse, IF, video, antenna transmission and power cables) or interconnecting wiring over the frequency range of 0. 15 to 10, 000 mc for CW and pulsed CW interference and 0. 15 to 400 mc for broadband impulsive interference. This requirement includes the transmitter fundamental, spurious radiation, oscillator radiation, other spurious emanations and broadband interference. This does not include radiation from antennas."

Paragraph (4. 3. 4. 2) on Radiated Susceptibility states:

"No change in indications, malfunctions or degradations of performance shall be produced when the equipment is subjected to a radio frequency field. This field shall be established with a signal generator driving the antenna listed below. Care shall be taken to use matching networks when required. The voltages specified are those calculated to exist across the antenna terminals. The test setup is shown in Figure 26 for the rod antenna and is similar to Figures 16 and 17 for the other antennas, with the signal source replacing the interference meter.

Frequency	Microvolts	Antenna
0. 10 to 25 mc	100, 000	41 inch rod
25 to 35 mc	100, 000	35 mc dipole
35 to 1000 mc	100, 000	tuned dipole
1000 to 10, 000 mc	100, 000	non-directive antennas"

First, consider the implications of the paragraph on radiated interference, (4. 3. 2) which, incidentally, uses the same antenna setup as required by paragraph (4. 3. 4. 2). It should be noted that the drawings which show how these radiated fields are to be measured indicate a standardized arrangement which is used regardless of the configuration being tested. No consideration is given to the fact that the RFI polarization might be different from the antenna polarization for some equipments but not for others. This means there could exist two equipments that both meet specification and yet one of them could radiate much more energy than the other. A careful examination of all the conditions for the measurement of radiated interference will show additional cases where similar situations could result. This is not to be construed as a condemnation of the specification, however. The problem of spurious and leakage radiations from all classes of equipments is a very complex one and to be solved exactly would require a different measurement setup for each different type of equipment, which of course, would be undesirable. The present specification has evolved as a means of standardizing all radiated interference measurements so that measurements made on an equipment at one facility can be compared with those made at another. Admittedly, there is a certain amount of arbitrariness about these setups, but until a better arrangement can be established, this method of measurement must be employed.

The same arguments as above apply to the susceptibility setups with the additional complication that no provision is made for testing susceptibility with broadband radiated energy. This fact is not an oversight but is simply the result of a limitation in the state of the art of acceptable signal generators. High-powered, broadband noise or impulse generators do not exist which have the precision required for repeatable, standardized measurements. The only possible way to test susceptibility at the present time is by applying the highest practical CW signal to the antennas specified and noting if any degradation of performance occurs.

From the above discussion, it should become fairly obvious that the relationship between the amount of radiated interference that one equipment generates and the susceptibility of another equipment to this same radiation is not defined by MIL-I-26600. Many factors beyond those discussed above enter into the reasons for this. Among them is: (1) a lack of knowledge of the incident field strength on the susceptible equipment since the specification only defines the voltage that should exist across the antenna terminals; (2) the lack of knowledge of the exact part of the susceptible equipment that responds to the incident energy; and (3) the lack of knowledge of the mechanism and location of the radiated interfering energy. For any specific.piece of equipment, all of these factors can be determined, but to do so required measurements which are considerably beyond the scope of MIL-I-26600. Because of these facts, it is necessary to make more specific tests than outlined in the specification when designing a system so that it will not interfere with itself. The nature of these tests can only be determined by consideration of the specific equipments involved.

Because of the nature of the specification, it is not valid to transfer the measured db level of radiated interference to the db level of susceptibility to interference, since any relationship would be pure chance. It is obvious, however, that in order to guarantee system self-compatibility, there must be a definable relationship between the energy radiated from an interfering source and the energy received by a susceptible source. Possible solutions to this problem must be investigated. Strict application to the present specifications for every system, however, would probably insure a large degree of system self-compatibility, because the requirements of MIL-I-26600 are as strict as possible within the ground rules set up by the specifications. There is a good chance that radiating and susceptible equipments which both pass the specification can operate side-by-side without interference, which is the primary objective. This cannot be guaranteed, however, and cases may easily come about where interference will be encountered.

1.2 PROBLEMS OF SYSTEM-TO-SYSTEM COMPATIBILITY

A full discussion of the question of system-to-system RFI compatibility covers an area that must be almost global in scope. Restricting attention to only those potential RFI sources within a comfortable geographic area or frequency is becoming increasingly unrealistic in view of the likelihood of interference of unanticipated origin. Problems of this nature are not unsolvable, however, and recent progress in RFI analysis and prediction, especially the spectrum signature approach, has removed much of the mystery surrounding system-to-system compatibility.

Potential interfering sources are not only the known transmitters, both ground-based and airborne, using the manifold modes of radio propagation discussed in this section, but also the freak variety whose use of these same modes elevates them to equal importance. Discussion here will be confined to the former type since the latter admits of little intelligent analysis with respect to its presence on the air. The next section, concerned with design solutions, will take it up more fully.

Ground sources of RFI can be cataloged according to their propagation modes: radars, sky wave systems, and scatter communications apply here. Aerospace RFI sources, although they add the parameter of mobility, are predictable and warrant discussion also. The above potential RFI contributors will be discussed in greater detail in appropriate sections when warranted.

1.2.1 RADARS

The narrow beam-width associated with most radars effectively limits the problem to those radars skin-tracking the aerospace vehicle itself. However, the very high power levels of these transmitters, not only at the principal frequency but also at their respective harmonics, can adversely affect the sensitive receivers on board, as well as injecting signals into and through the various parts on the skin of the craft.

1.2.2 RF COMMUNICATIONS

1.2.2.1 Sky-Wave Propagation

The RFI problem involved with sky-wave propagation communication is two-fold. The first is that the transmitted signal may not, for a number of reasons, follow its predicted path. Secondly, perfect operation of a multi-hop path may introduce an earth reflection in an undesirable area.

Unpredictable changes in the absorption on the ionoshpere and the interposition of extra refracting elements in the atmosphere and troposhpere account for most of the off-path problems. The "sporadic E-layer" phenomenon is a good example of this. The so-called E-layer, the lowest of the usable ionospheric layers, is also the least uniform; it is more of a stratum of discrete clouds, at times unreliable as a reflector or refractor of rf signal. Thus a signal intended for E-layer propagation may be passed, through a reduction in absorption uniformity to a higher layer destroying the intended path. Similarly, an increase in its absorption may cause attenuation, bending, or even full reflection of a signal traveling upward to a higher layer or one traveling downward. In addition, the composition of the ionosphere may be altered, sometimes drastically, by phenomena such as SID's sudden ionospheric distrubances due to solar flare, lighting, and the like. High level auroral activity is another contributor.

The other major cause of off-path transmission, non-ionospheric reflectors or refractors, can assume a number of forms. A dense cloud layer in the atmosphere can intercept a passing signal prematurely and, if enough power is involved and if the cloud layer is extensive, can create a tube-like channel of many hops extending for miles. Aircraft can also reflect signals, although for very short durations.

1.2.2.2 Scatter Propagation

Scatter propagation uses the troposphere and the ionosphere as instruments of wave reflection, refraction, and diffraction in order to effect trans-horizon propagation. Although higher, less infringing frequencies are involved, the showering of RF radiation over wide areas, other than that needed for the intended receiver, and at the power levels necessary for this technique, poses all the difficulties involved with skywave propagation, plus increasing the problem of undesirable spatial coverage.

1.2.2.3 Airborne Radiations

Signals from aircraft and satellite transmitters will be sharing frequency ranges and, as such, are worthy of discussion here. Aircraft communication, unlike ground-based, point-to-point communications, is multi-directional; signals aimed at no particular receiver are a matter of procedure. Satellites also transmit this way because pickup by a number of receivers throughout the world is desirable. Power levels involved here are much lower although spatial distribution per transmitter is vastly enlarged.

2. DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT COMPONENTS FOR MINIMUM GENERATION OF INTERFERENCE

Radio interference originates from the operation of the components of aircraft systems. Consideration of the aircraft system itself is essential only to establish techniques to prevent the interference which is generated by the components from being transmitted by radiation, conduction, or coupling to various other susceptible receivers in the aircraft. These system considerations are treated in detail under Section 2.1. The ideal method of eliminating the effects of unwanted signals is to design all components in such a way that no unwanted signals are generated.

Source suppression is by far the best method of controlling interference in most cases and should be applied whenever possible. Nevertheless, in some components the generation of signals is inherent to their normal function and source suppression cannot be employed. This is true for all transmitters where the signals appearing on the transmitter antennas are the desired result. Here the interference problem is primarily a system design consideration involving mounting and location of the antenna, and taking advantage of the shielding effects afforded by the aircraft structural members and metallic skin. However, source suppression techniques can be employed in the transmitter case design and in the elimination of harmonics appearing on the antenna to reduce the transmitter interference problem considerably by eliminating various unwanted signals associated with the generation of the desired output signal.

Source suppression is desirable from several standpoints other than that of interference-free design. Aircraft maintenance is one important reason for utilizing source suppression wherever practicable. The resulting decrease in required shielding greatly reduces the electrical maintenance problem. This is especially true when the aircraft is to be operated under combat conditions. Flak or gunfire damage at any point along a conduit would require replacement of an entire rigid conduit assembly and would considerably prolong the non-operational status of the aircraft while under repair. When flexible conduit is utilized for shielding purposes, as is the case in all late model aircraft, the maintenance problem is somewhat less severe. Nevertheless, damage to any required shielding increases the time and cost of repairs.

Since source suppression can best be employed in the original component design, the techniques described in the following sections are of major importance. Design engineers should be thoroughly acquainted with this material to insure good interference-free components. It is of utmost importance that all electrical components, regardless of their function or location within the aircraft, be treated as potential sources of interference.

2.1 DESIGN CONSIDERATIONS APPLIED TO AIRCRAFT SYSTEMS

Present day military aircraft are almost completely controlled by electrical systems. Even hydraulic systems depend upon electrical control circuits for their operation. The large increase in number and complexity of the electrical and electronic systems has greatly increased the radio interference problem. Close proximity of components, bundling of wiring into common cables, and high energy interference sources, have increased the number of paths over which interference signals may enter radio and radar receivers as well as the probability for such action. Radio interference levels that produced no adverse effect whatsoever in old model aircraft can no longer be tolerated in present day models. Since each crew member must operate one or more radio or radar sets, all indicators and controls involved must be located within the respective operator's reach. This requires bundling of power, indicator, and control wiring of several systems in the same cables. Figure 2-1 shows a typical radar compartment installation in a typical aircraft. Here the close equipment mounting, parallel wiring, and wire bundling can be seen clearly. The radar modulator, IFF unit, alternator, and inverter are located relatively close to one another and their interconnecting cables are bundled together. This offers the possibility for the high level radio interference emanating from the modulator unit to enter the IFF and AC power systems and gain access into many other electronic circuits throughout the aircraft. Also, the interconnecting cables passing through the compartment may pick up interference signals and conduct them into the various receivers. It should be stressed that electrical servo systems and control actuators may also introduce interfering signals into the electronic circuits.

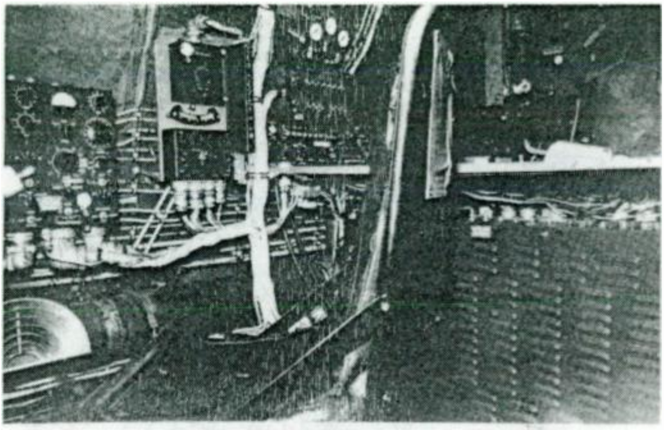


Figure 2-1. Typical Radar Compartment Installation

A typical aircraft electrical and electronic system installation is represented in Figure 2-2. This shows the necessity for bundling and paralleling of circuit wiring. The operational necessity for mounting control boxes and indicators close together is also illustrated in this figure. When it is fully realized that each system must operate in conjunction with the many other systems in close proximity, then the magnitude of the system installation problem can be appreciated.

Design engineers of electrical and electronic systems must be thoroughly familiar with the installation problems and techniques to insure interference-free operation in the original lay-out. The system must be so designed that interference signals cannot enter or leave the system due to conduction, radiation or inductive or capacitive coupling. Proper observance of good source suppression and systems design techniques will guarantee satisfactory functioning of all electrical and electronic systems regardless of their number or complexity.

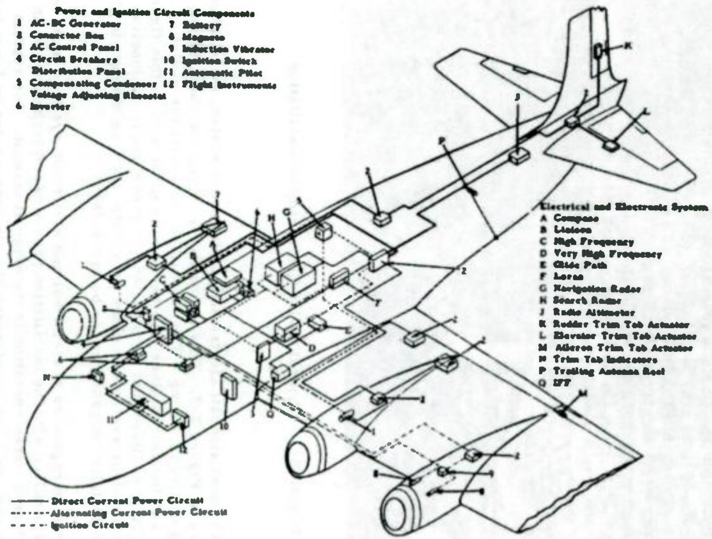


Figure 2-2. Typical Electrical and Electronic Power System

2.1.1 ELECTRONIC SYSTEMS

Electronic systems are generally susceptible to interference although any one of them may be capable of generating interference which can affect some other electronic system. The arrangement and interconnecting wires of the component parts of each of these systems within the aircraft presents an array by which interference may couple into the system. An analysis from the radio interference point of view of each of these systems can be made without any prior knowledge of how interference may actually couple into or leak out of the system. Various components of each system are either susceptible to picking up radio interference or generators of interference or harmonics. If the assumption is made that the source suppression of "noisy" components cannot be absolutely perfect, additional design considerations such as routing of wires, shielding, shading, arrangement and location of equipment are necessary to insure interference-free operation of the system and should be pointed out to the design engineer to increase his appreciation of the overall problem. Moreover consideration should be given to reducing the susceptibility of various parts of the system so as to tolerate the presence of interference fields without many adverse effects. This in no way advocates a "tailored" installation. It is merely intended to illustrate to any individual design engineer why and how he must broaden his viewpoint of the radio interference problem and where in the design of the general layout of a system attention must be given to certain special considerations. Many specific examples exist where interference entered or leaked out of a system as a result of poor design practice. The following sections describe typical installations which serve to point out some of the general considerations. Each of these has peculiar characteristics as to power supply, antenna location, receiver-transmitter combination, etc., which deserve enumeration and illustration to create a picture of the nature of the physical situation under consideration.

2.1.1.1 Radio Range and HF Systems

High frequency radio systems used in present day aircraft are designed to provide air-to-ground communication utilizing voice, code, and tone modulated CW transmission. Essentially, the system is a multichannel radio receiving and transmitting equipment used for HF command and radio range purposes consisting of (1) three receivers with frequency ranges of 190 - 550 kc, 3.0 - 6.0 mc and 6.0 - 9.1 mc; (2) two transmitters with 5.3 - 7.0 mc and 4.0 - 5.3 mc frequency ranges, (3) remote control boxes, and (4) a modulator, all located in the cabin on the top radio shelf in the navigator's compartment; (5) an antenna relay mounted in an inverted position on the cabin overhead, and (6) two wire antennas running between three masts located on the top of the fuselage, (one for the range receiver and the other for the transmitter and the command receivers). The HF equipment provides transmission on two preset channels and reception through the frequency range of 3.0 - 9.1 mc. The low frequency receiver receives radio range signals in the 190 - 550 kc frequency range. In a typical installation this system serves the following stations: (1) pilot, (2) co-pilot, (3) radio operator, (4) radar-navigator, (5) two observers, and (6) tail compartment stations.

The receivers and transmitters are potential radio interference sources and are also susceptible to radio interference. Since the components of this system are widely separated there is a strong possibility that interference signals can enter or leave the system by radiation or inductive (or capacitive) coupling from the system wiring. Conductive access to and from the HF system is also provided by the interphone and power systems. The paths over which interference can enter or leave the system as shown in Figure 2-3 are: (1) antenna leads, (2) power leads, (3) mechanical remote control cables, (4) interphone connections to the earphone and microphone, and (5) penetration of case. There is also the possibility of intersystem interference through the antenna relay.

The circuitry of the HF system should be arranged and components placed so that a minimum of shielding and filtering is required. Reference to Figure 2-3 shows that the HF system is directly connected to other electrical and electronic systems in the aircraft through connecting cables: (1) DC power supply to transmitter and receiver dynamotors, (2) receiver audio output to interphone systems, (3) transmitter microphone connection to interphone systems. These lines should be filtered to prevent interference from entering or leaving the system by conduction (refer to Chapter 4 of this Volume for details of filtering and filter design).

The long antennas, the relatively long antenna leads, and the long HF system wiring increase the susceptibility of the overall system to radiated interference as well as increase the interference-source potentialities of the system. This can be reduced by proper routing and shielding where necessary. In general, the antenna lead to the range receiver is adequately shielded and bonded to prevent interference difficulties over this path.

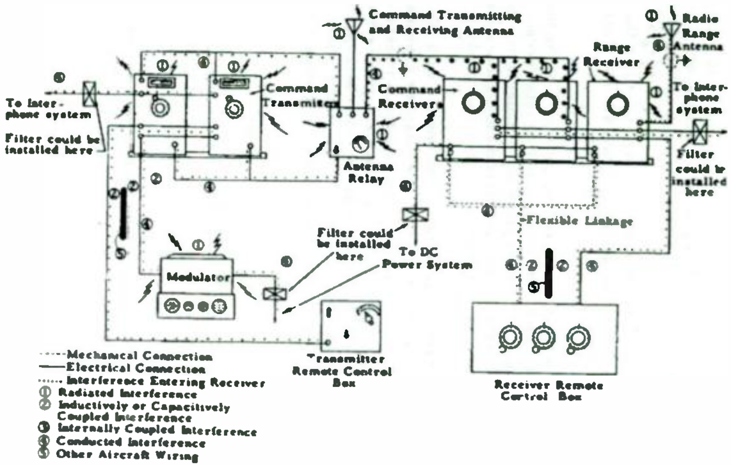


Figure 2-3. Paths of Interference Signals in a Typical Radio Range and HF System

However, if this antenna lead should be routed very close to a high energy interference source, additional shielding would be required. The command receiver leads are shielded between the receiver rack and the antenna relay so as to be relatively interference-free. The transmitting antenna lead between the transmitter rack and the antenna relay as well as the antenna lead from the antenna relay to the command antenna are not shielded and therefore introduce a serious interference problem.

The long HF system wiring extending the length of the fuselage in a typical installation provides a means of coupling interference signals into or out of the HF system. If these cables are bundled with other susceptible system wiring, or other probable interference-generating system wiring, they should be shielded, unless rerouting is practicable.

Satisfactory HF system operation in conjunction with the electrical and electronic systems in actual aircraft installations depends largely upon the care exercised by the designer in applying radio interference suppression techniques to the installations of other aircraft systems as well as the HF system itself. However, in any event, observation of design techniques outlined in this book together with specific attention to the particular problems discussed herein should result in a system free from objectionable radio interference.

2.1.1.2 Very High Frequency Radio Receiver Systems

Very high frequency radio systems used in present day aircraft are designed primarily to provide air-to-air or air-to-ground communication. The system is generally composed of a radio receiver, radio transmitter, power junction box, antenna, control units, and necessary interconnecting cords. In early models the transmitter, receiver, and power junction box were housed in separate cases while later more compact designs have incorporated all three components in one case.

Any one of eight channels within the VHF range may be selected for operation. Remote operation is provided by a remote control box and control cables. An audio output signal provided by the VHF receiver is available at any one of the interphone stations.

The installation of representative VHF Radio Set in a typical aircraft has been selected as an example. The discussion to follow applies specifically to this particular layout. However, since all such systems are functionally similar, generality is still maintained.

One antenna is utilized to radiate or receive radio-frequency energy by the VHF system. This antenna is generally mounted on the top or underside of the fuselage, near the pilot's or navigator's compartment. The pilot's remote control box is mounted within reach of the pilot when seated at the flight controls. In a typical installation one interphone control box each, located within reach of each crewmember, would be provided for the pilot, co-pilot, navigator, and radio operator. The receiver, transmitter, and power junction are mounted in the compartment behind the pilot. There are no components located in the tail or wing sections. Interconnecting cords are bundled with other electrical wiring passing through the compartment

The three principal components, receiver, transmitter, and power junction box, are potential radio interference sources. Of these, only the transmitter and receiver are susceptible to radio interference. Even though the functioning of the control and power junction boxes may be unaffected by interference signals, they may serve as paths of entry for undesired signals.

As shown in Figure 2-4, interfering signals may enter the transmitter case by way of the (1) antenna terminal, (2) antenna connection from the receiver, and (3) power and control cables.

Interfering signals may enter the receiver case over any of the following paths also shown in the figure. These are: (1) antenna terminal, (2) power and control cable, or (3) through penetration of the case. A detailed discussion of receiver design techniques for interference-free operation is given in Chapter 1, Section 1.4.

Interference entering the transmitter case may be radiated and manifest itself at receiving stations on the ground or in other aircraft, in the form of disturbing modulations producing "noise" in the respective audio output systems. Since this has not been particularly disturbing, the problem will not be discussed further except to point out the possibility and to recognize that these paths of entry do exist. Some future installation could alter the situation sufficiently to produce malfunctioning in the transmitter or in ground stations if proper design techniques are not carefully considered during the functional design of the transmitter system.

Transmitter case radiation and harmonic generation have caused considerable difficulty in aircraft installations. The design of the transmitter housing in the typical case selected, did not include screening of the ventilating louvres. The inspection plate on the base of the chassis made poor contact with the case and the electron tubes with glass envelopes

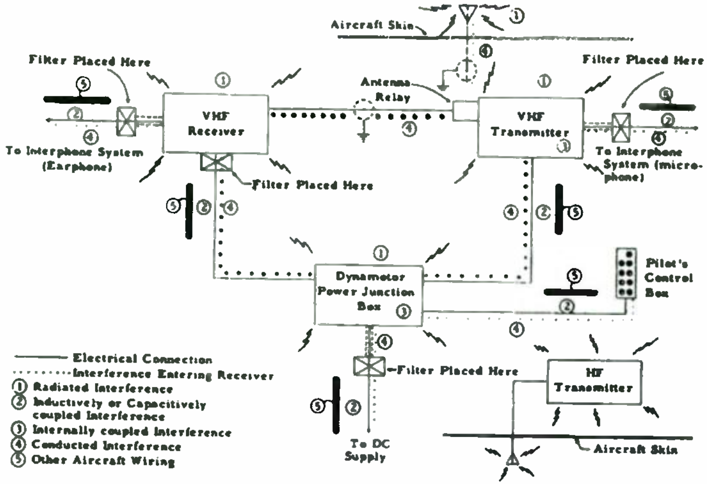


Figure 2-4. Paths of Entry of Interference Signals in a Typical VHF Communication System

were radiating appreciable energy. It was necessary to incorporate the following modifications in order to reduce the interference generated by the transmitter to within tolerable limits: (1) shield and bond first tripler and final stages, (2) rivet a metal screen to the inside of the cover to reduce radiation through the ventilating louvres, (3) bond the inspection plate on the base of the chassis with a copper gasket, (4) install a bonding strap around each shock-mount, and (5) replace troublesome glass tubes with metal tubes. These difficulties would not have appeared in the original installation if proper component design had been observed. A detailed discussion of interference-free transmitter feedback amplifier design is given in Chapter 1, Section 1. 3. 3.

The VHF receiver has caused interference with other electronic systems in the aircraft. Automatic selection of the various channels by an electrically operated, motor-driven, channel-selecting mechanism occurs when any one channel push button is depressed on any remote control box. This band-change motor caused excessive interference over a wide band of frequencies. The interfering signals were present on all interconnecting cables and also appeared at the supply terminals. Since the interference was present only during the "warm-up" period and during the band changing period, no corrective measures were attempted in the typical case selected for discussion. However, it should be emphasized that this type of interference need not appear in a system, even for short time intervals, if proper component design techniques were applied during the original design to suppress the interference at the source. A detailed discussion of interference-free design techniques for small motors and receivers is given in Chapter 1, Section 1. 4 of this Volume and Volume I, Chapter 1, Section 7. 1. 1.

Considerable interference was generated within the VHF system by the dynamotor power unit in the power junction box. Filters were installed in the power junction box to suppress the dynamotor commutator interference at the source. It was also discovered that the high voltage leads were exposed to radiation from the dynamotor after leaving the filter. Rerouting these leads reduced the interference to a permissible level.

In the original installation, components should be placed, and the circuitry arranged so that a minimum of filter components are necessary. Reference to Figure 2-4 shows that the VHF system is directly connected to other electrical systems in the aircraft through connecting cables: (1) receiver audio output to interphone system, (2) transmitter microphone connection to interphone system, (3) DC power supply to power junction box. These lines can be filtered to prevent interference from entering or leaving the system by conduction (refer to Chapter 4 of this Volume for details of filtering and filter design).

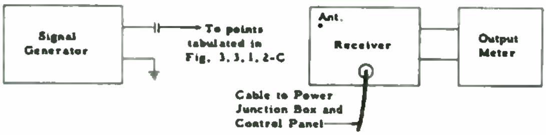


Figure 2-5a. Test for Source of Interference in Receiver

Lead Wire Function	Signal Generator Output In volts	Receiver Output In volts
Antenna Post	0.32	1.55
Channel Selector A	0.26	8.5
Channel Selector B	0.23	1.7 *
Channel Selector C	0.26	10.2
Channel Selector D	0.27	2.0 *
Channel Selector E	0.26	8.0
Channel Selector F	0.26	8.0
Channel Selector G	0.27	10.3
Channel Selector H	0.27	10.4
+200v	0.26	10.4
Audio Low	0.26	7.0
Time Delay	0.26	10.4
Ground	0.26	1.1 *
+28v	0.27	10.6

* Denotes wires which were not filtered. All other wires in this cable were filtered to prevent the 12 mc signal from entering the receiver.

Figure 2-5b. Susceptibility Tests of VHF Receiver

An interesting interference problem was encountered with this system which illustrates some poor design techniques. Interference was noted in this VHF receiver whenever the HF transmitter of another system was operated at 12 megacycles, the IF frequency of this receiver. Disconnecting the receiver antenna did not eliminate the interference. The test described below proved the interference was entering the receiver via the power and control leads. The signal generator was tuned to 12 megacycles and the modulation was 30 percent at 1000 cycles per second. The signal was applied to the antenna post and the individual wires in the cable shown in Figure 2-5a with the results tabulated in Figure 2-5b. The background interference level at the output was 0.65 volts with the antenna disconnected and no signal applied.

In this particular installation, the leads were filtered to eliminate the interference. Two faulty design practices created the need for ten filters; however, if these had not existed, the added weight, cost, etc., would not have been required. They are:

a. The HF transmitter antenna lead-in was unshielded and rather long. The 12 mc signal radiated by the lead-in was picked up on the control wires of the VHF receiver and interfered with its operation. The antenna lead-in had to remain unshielded because the transmitter output circuit was not designed for the load presented by a shielded lead-in.

b. The VHF receiver circuits were interference susceptible due to coupling between the IF wiring and the control and power wiring. It should be noted that the IF rejection of the antenna input circuit was far better than the equivalent rejection for the susceptible wires in the cable.

Overall design considerations for the system could be improved by reducing the number of separate components with the consequent elimination of external wiring. Some of the later model VHF systems have incorporated the receiver, transmitter and power unit in one case. This compact construction is advantageous from the standpoint of shielding, wiring, and overall interference-free design.

The VHF antenna system is provided with a feed-through insulator, coaxial cable to transmitter, a switching relay in the transmitter, and a coaxial cable to the receiver. This type of antenna design is in agreement with sound interference-suppression techniques and actual installation experience has proved this arrangement satisfactory.

This system serves as an excellent example to point out the necessity for thoroughness in considering the interference problems in the design of any system. Poor transmitter and dynamotor component design as well as poor system shielding and filtering have produced a system that is a relatively strong radiator of interference signals. If good interference-suppression design techniques had been employed in the overall system as well as to the components of the system, the need for the system fixes described above would have been eliminated and a satisfactory operation would have resulted in the original installation.

2.1.1.3 Search Radar Systems

Airborne search radar systems are designed to present a visual representation of a portion of the earth's surface or objects on a radar scope to supply accurate navigation and bombing information independent of weather and visibility conditions. This system is generally composed of (1) an antenna assembly, (2) transmitter-receiver unit, (3) modulator, (4) indicators, (5) synchronizer, (6) rectifier power unit, (7) junction boxes, (8) control units, (9) gyroscope, (10) directional coupler, (11) blowers, (12) servo amplifier, (13) camera attachment, and (14) various interconnecting cables. The antenna is operated by a servo motor so as to scan a portion of the area under and around the aircraft. Circuits, controls, etc., are arranged to radiate radio frequency pulses, synchronized with the indicator sweep circuits, into the region searched and to receive the returning reflected signals or echoes. These received echoes vary in strength because of the different reflecting properties of the objects in the area scanned. Each received signal is converted by the equipment into a light spot on a cathode-ray indicator. Since the intensity of the light spots are dependent upon the amplitude of the received signal, a light and dark map-like pattern appears on the radar scope. There is no audio output signal from the system. The antenna assembly is generally mounted on the underside or in the nose of the aircraft. All other components are mounted in the compartment behind the pilot where they will be accessible to the navigator or radar operator. Due to the large number of components, considerable interconnecting wiring is required and a large portion of the system wiring is bundled with other electrical and electronic system cables that pass through the compartment.

The installation of a representative model search radar set in a typical aircraft has been selected as an example for discussion of design techniques for interference-free operation. In this system, the receiver output is a visual pattern displayed on the radar indicator scope, and any unwanted signals which find their way into the radar receiver case must be capable of disturbing the indicator pattern in order toconstitute an interference problem. In Volume 1, Chapter II, Section 3.1 the nuisance value of interfering signals in various types of receivers is discussed in detail.

The components in a typical search radar installation that should be given special attention as potential interference generators are: (1) antenna assembly, (2) motor driven fans, (3) receiver-transmitter unit, (4) indicators, (5) synchronizer, (6) modulator, (7) power unit, (8) servo amplifier, and (9) the gyroscope unit. Of these units, those that can also be classed as "receivers" under the extended definition of receiver given in the introduction of Volume I are items (3), (4), (5), and (8). Even though the functioning of the other components of the system is unaffected by interference signals, they may serve as paths over which interference may enter or leave the system.

Figure 2-6 illustrates the paths of interference signals in a typical search radar system and shows the most vunerable points for interference signals to enter the system. Interfering signals have gained access to the various susceptible components over the paths shown and have eventually exercised sufficent influence on the indicator pattern to constitute an interference problem.

The location of the antenna assembly of a radar search system is particularly important. When the general design of the aircraft requires mounting the antenna in a location where the radar receiver-transmitter field to the sides or to the rear is blocked by reflecting surfaces, standing waves of high amplitude will generally be established in the antenna system. This results in faulty operation of the radar set. Considerable reduction of this type of radio interference can be obtained by coating the reflecting surfaces involved with material to absorb as much radio frequency energy as possible. Such measures are essentially the only means available for eliminating the disturbance when relocation of the antenna assembly is not feasible.

Interference in this typical installation entered the radar system also over the power cables. This disturbance was suppressed by inserting a filter at a convenient position between aircraft power supplies and the radar system junction box. Some difficulty was also encountered when the cable from the contol unit to the synchronizer and the cable from the synchronizer to the indicator were mounted too close to the power cables of either the radar system or other aircraft wiring. This was corrected by separating the power cables and the radar system indicator or synchronizer cables by at least 18 inches when installing the system wiring.

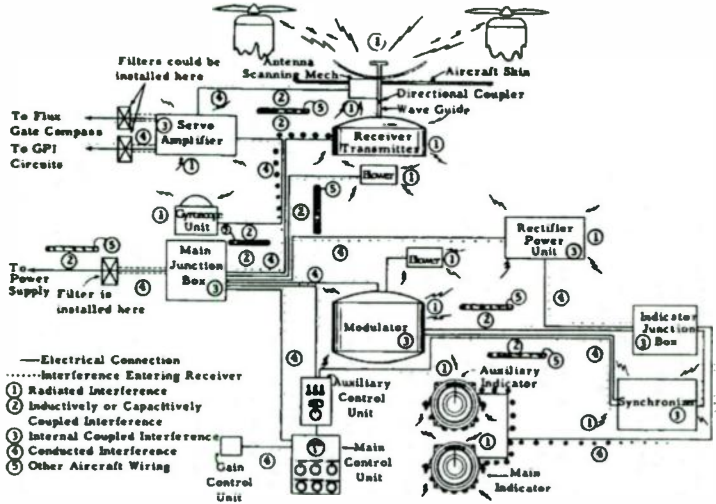


Figure 2-6. Paths of Interference Signals in a Typical Search Radar Set

Radio interference tests have demonstrated that search radar systems introduce a severe interference problem in other aircraft systems. In a typical installation the search radar pulse modulator circuits caused interference in the liaison and radio compass systems. Extended filtering and shielding was ineffective. Probing of the radar cabling indicated a high interference level over extensive lengths of various cables. This suggested that external filtering would be very difficult and that source suppression within the modulator together with the rerouting of certain portions of the cabling would be the most efficient solution. For a discussion of modulators refer to Chapter 1, Section 1.2.4.

Radio interference from the computer appeared in the audio output of the radio compass. Shielding of the radio compass sense antenna was required to prevent interfering signals from entering the radio compass receiver.

In the typical case under discussion, intermittent interference signals caused by the determination-switch operation and tracking-control operation appeared in other electronic systems on the aircraft. Attempts to filter all radar power cabling at the power connections proved useless. As shown in Figure 2-6, interfering signals from remote units must travel through long unshielded cables to reach the filter. This permits coupling of interference into other aircraft wiring as well as coupling into the search radar circuits. In addition, all power wiring had wire return circuits from the power unit to the main filter, and filters were provided for these wires just prior to grounding. This indiscriminate use of filters served no useful purpose, unnecessarily increased the length of the system's wiring, and required the use of unduly large connectors to handle the extra pins. Here again, source suppression would have more efficiently guaranteed an interference-free system in the original installation.

Reference to Figure 2-6 shows that the Search Radar System is directly connected to other electrical systems through the AC power supply cable into the main junction box. A filter is provided in this line to prevent interference from entering or leaving the system by conduction. However, the greatest interference problems in this typical search radar system were caused by coupling into other aircraft wiring. This coupling was traced to the facts that (1) the modulator unit was not adequately bonded to the airplane structure, (2) surplus pulse cable was coiled and clamped next to aircraft wiring, (3) the AC lead from the modulator to the external blower was shielded, but "grounds" on the shields were too long, (these leads were "hot" from the modulator pulse and it was necessary to shorten leads to one inch), and (4) the coaxial antenna cable for the Loran receiver passed through the interference field around the modulator, (although grounded at both ends, the cable provides an efficient path for conducting the interference to the liaison receiver because both receivers used the same antenna), and (5) coupling also occurred inside the liaison transmitter since the external power wires leading to the transmitter were exposed to the interference field and the interconnected internal wiring was in close proximity to the receiver antenna-grounding lead. Appreciable improvement in future designs and installations can be obtained by the application of the following procedures to obtain interference-free operation:

a. Ground all four pulse cable shields at the connectors. (It was found that in some cases only one of the shields was grounded because of greater ease of assembly, a practice which resulted in the loss of much of the shielding effectiveness.)

b. Install a suitable filter in series with an AC lead from modulator to external blowers.

c. Install the modulator and receiver transmitter as far as possible from other radio equipment, especially the liaison and radio compass receivers. These receivers should be in a separate compartment if possible.

d. The pulse and high voltage cables should not be bundled with, run parallel to, or placed less than one foot from other radio and aircraft wiring. At least eighteen inch separation should be maintained for cable lengths over twenty feet.

e. Keep all wiring associated with the radar set well isolated from other aircraft wiring.

f. Locate modulator and receiver-transmitter units so that pulse cable lengths are held to a minimum.

g. Since the pulse cables are prefabricated in fixed lengths, extra cable is sometime coiled to take up surplus. In case this is necessary, the coil should be placed in the bomb-bay or other isolated compartments and adequately spaced from other wiring.

h. All radar components should be properly bonded to the aircraft structure by application of the techniques discussed in Chapter 3 of this volume. The modulator and receiver-transmitter units should have at least two such bonds of shortest practicable length.

i. The ac lead from modulator to external blowers should be filtered with a portion of the lead between modulator and the filter shielded and grounded with short leads.

j. In case interference is encountered due to penetration or leakage of the pulse cable shield, the interference may be reduced by gounding the shield at intervals of approximately 5 feet.

Some of the later models of search radar equipment have incorporated improved design techniques resulting in a considerable reduction in the interference trouble described above. Much of the difficulty caused by long cable lengths and modulator radiation have been eliminated by placing the modulator, transmitter and receiver in one metal case. Fewer component parts, reduction of interconnecting wiring and better relative location of circuit elements have also improved the overall design from an interference standpoint. However, even with these improvements, considerable trouble is still caused in other electronic systems by search radar equipment. The high energy pulse output of the radar constitutes a prolific source of interference. In an improved version of a typical search radar system, shown in Figure 2-7 the paths for interfering signals to enter or leave the radar system are indicated. Some of the interference problems encountered in the improved version and the modifications made in an effort to attenuate interference are discussed below:

a. Excessive levels of interference were present in the high voltage, modulator pulse circuits which coupled interference to all interconnecting wiring and coaxial cable within the receiver-transmitter. This interference was decreased by the addition of filters and by-pass condensers on the interconnecting wiring, and by the substitution of double-shielded coaxial cable for the single-shielded cable. The leads to the external blowers were shielded to confine the interference to the receiver-transmitter unit. The grounding of the video and AN connectors to the junction box were improved by removing the paint from the connectors. Refer to Chapter 3 of this volume for a detailed discussion of direct bonding. Furthermore, the mating surface between the receiver-transmitter lid and case was cleaned in order to obtain a direct metal-to-metal contact.

b. Interference was generated within the synchronizer unit by the transients in the sweep and intensity-gate signals for the indicators as well as in the steep pulse for triggering the modulator. The level of interference was decreased by shielding the lines within the unit and the 120-volt leads to the antenna assembly and the two indicators.

c. A high level of video interference in the indicators was caused by thermal agitation and by the shot effect in the receivers and

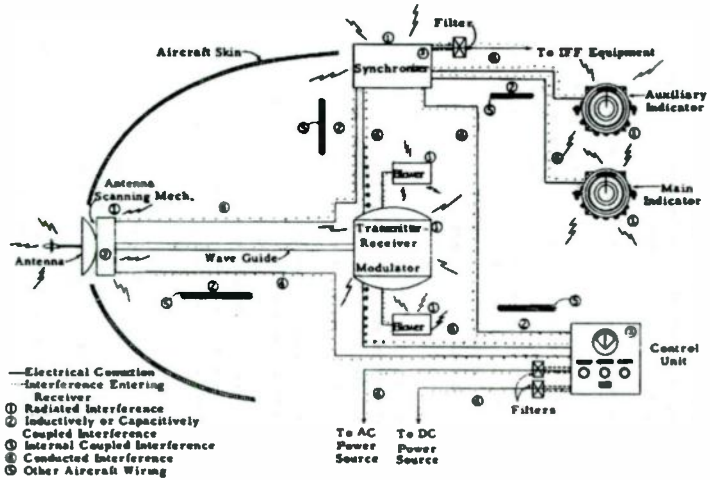


Figure 2-7. Paths of Interference Signals in an Improved Version of a Typical Search Radar Set

was coupled by induction into the interconnecting cables from the two B+ leads between the indicators and the synchronizer. This interference was attenuated by the insertion of an R-C decoupling circuit in the B+ leads. Interference emanating from the case of the indicator was attenuated by shielding and grounding the gain control potentiometer and grounding the case of the intensity-control potentiometer. Further attenuation of interference was obtained by grounding the indicator through a shock-mounted bonding jumper. Refer to Appendix IV for a detailed discussion of shock-mounted bonding jumpers.

d. Sparking at the commutators of the direct-current motors and the make and break action at the sector-scan switch points were the sources of interference within the antenna assembly. The motors were shielded to decrease the level of radiated energy and a line filter was used to attenuate conducted interference. Capacitor spark suppressors and filters were employed to attenuate interference generated by the sectoring switches.

From the above discussion it can be seen that this radar system continues to be a high level interference source in spite of the improvements. Although the corrective measures taken reduced the level of interference considerably, the radar set failed to meet the requirements of Specification No. 16E5 (Aer) and MIL-I-6181. In order to meet the specifications the system should be redesigned with the following interference-free design features incorporated:

a. The modulator unit should be totally enclosed within a mesh shield whose joints have been welded or soldered. Refer to Chapter 3, Section 1.3 for a detailed discussion of a mesh shield. To provide a low, radio-frequency impedance path to ground, the shield should be clamped or screwed to the chassis. Furthermore, all power and control leads should be filtered by the use of a feed-through capacitor at the point of entrance. The superior attenuation characteristics of feed-through capacitors are discussed in Chapter 4, Section 1.2.5. This method of shielding and filtering will confine all interference generated by the modular pulse circuits to the modulator unit. All other circuits in the receiver-transmitter unit should be relatively interference-free.

b. The use of inherently interference-free components in the radar system should be stressed. Alternating current induction motors, rather than fractional horsepower dc motors, should be used whenever possible. A detailed discussion of commutation, the worst offender of all sources of radio interference in rotating machines, is given in Chapter 5, Section 2.2. When a fractional horsepower dc motor must be used, it should be completely shielded and equipped with a line filter as explained in Chapter 5, Sections 2.3 and 2.4. Vacuum tube switching circuits are preferred to vibrating contacts such as relays or vibrators because these devices generate high levels of interference. If the use of relays or contacts cannot be avoided, the unit should be enclosed in a metal shield and all connections should be filtered.

2.1.1.4 Interphone Systems

Combat interphone systems are designed to provide interphone communication between the various stations of a multiple place aircraft Switching facilities are provided by means of jack boxes at each interphone station to enable the crewman to exert partial control over the radio system required for the discharge of his duties. Examples of radio systems which can be partially controlled at the various stations in addition to the interphone system are: VHF, Liaison, Command, and Radio Compass.

The major components in a typical interphone system include an interphone amplifier, dynamotor, jack boxes, headphones, microphones, microphone switches and control panels. The quantity and type of components used in a specific installation is dependent upon the tactical needs of the aircraft.

Paths of entry of interference are shown in Figure 2-8. Interference problems arising from typical aircraft installations are given below.

The dynamotor mounted on the chassis of the interphone amplifier causes interference in the output of many radio-frequency amplifiers incorporated in the various electronic systems due to the conduction of interference through the common power supply. This interference was satisfactorily attenuated by the use of a properly designed radio-frequency filter. Audio-frequency interference which was extremely apparent in the output of the interphone amplifier was decreased to a sufficiently low level by bonding the dynamotor to ground. Refer to Chapter 3 of this volume for a discussion on bonding. Dynamotors, small units designed to convert direct-current power from one voltage magnitude to another, are prolific sources of both audio and radio-frequency interference primarily because they contain two commutators. Refer to Chapter 5, Section 2.2 of this volume for a discussion on commutation.

Interference was found to enter the interphone amplifier by conduction through the direct-current bus where ripple voltages as high as 4.4 volts at a frequency of 4000 cycles per second were measured. This interference, originating in the surface-control hydraulic boost

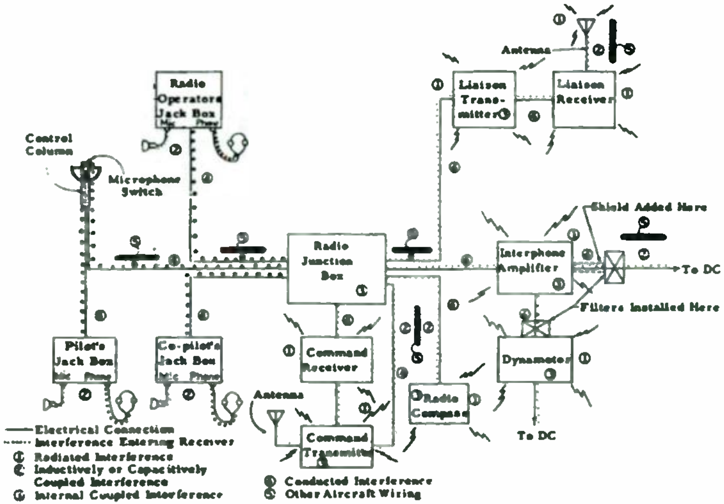


Figure 2-8. Paths of Interference Signals in a Typical Interphone System

pump motors, was effectively attenuated by the insertion of a filter illustrated schematically in Figure 2-9.

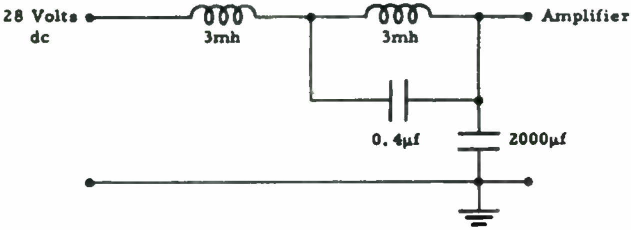


Figure 2-9. Line Filter for Surface Control Motors

The preceding problems were a result of interference being conducted into or out of the system. The following examples illustrate problems caused by interference coupling into or out of the system.

a. Interference, coupled by induction from the power wiring to the interphone amplifier, was prevented by maintaining a minimum spacing of 50 inches between the power wiring and the amplifier.

b. Inductive and capacitive coupling of interference from the power wiring into the interphone wiring of a specific aircraft installation was eliminated by replacing the original, single-wire interphone system with a two-wire system. The two wires were twisted, enclosed in a metal shield, and a minimum distance of 12 inches maintained between shield and the power wiring. In general, the results of interference tests indicated that for greater interference-free operation all single-wire systems should be replaced by two-wire systems. Variations of the two-wire systems have been satisfactorily used depending upon the function of the aircraft. For example, satisfactory operation of an interphone system in cargo planes, which are relatively free of interference-generating devices, has been accomplished by the use of two unshielded wires. In contrast, satisfactory operation of the same system installed in bombing aircraft could only be accomplished by the use of two wires enclosed within a shield. The wiring systems employed in aircraft installations in order to their effectiveness in suppressing interference are:

(1) Two twisted wires in a common shield.

(2) Two parallel wires in a common shield.

(3) One shielded wire with a common return path, or two unshielded wires.

(4) One unshielded wire using the structure as a return path.

c. Interference was present in the interphone system when the pilot's microphone switch was opened but disappeared when it was closed. This was caused by the coupling of interference by induction to the microphone switch wires which were run through the control column remote from the microphone as shown in Figure 2-10. Interference was attenuated to a certain degree by grounding the column, but much greater improvement was obtained by using a relay in the pilot's microphone wiring and by running relay control leads rather than the microphone switch wiring in the control column.

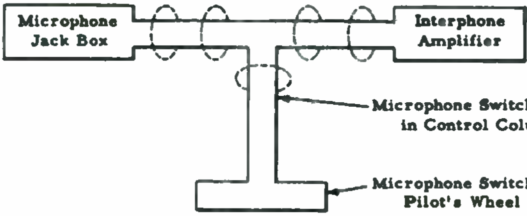


Figure 2-10. Routing of Microphone Switch Wires

d. The dynamotor mounted on the interphone amplifier chassis was the source of interference present in the liaison receiver in a certain installation. Interference was coupled from the power leads of the dynamotor into adjacent aircraft wiring and eventually coupled into the unshielded antenna leads of the liaison receiver. This interference was attenuated to a satisfactorily low level by the insertion of a filter in the power leads of the dynamotor.

e. In another typical installation, the liaison transmitter was mounted in close proximity to the interphone amplifier. Interference was inductively coupled to the audio input transformer of the amplifier from the armature windings of the dynamotor employed as the power source of the transmitter. The level of interference was decreased sufficiently by enclosing the transformer within a metallic shield.

f. The close proximity of the microphone and headset leads resulted in capacitive coupling between the input and output circuits of the interphone amplifier and caused audio oscillations in the interphone system. In order to prevent any oscillation from occurring, the amplifier components were connected so that the output and input circuits were as far out of phase as possible throughout the audio range of 100 to 20,000 cycles per second. This phasing was accomplished by connecting the secondary of the output transformer so that the signal voltage across the output lead and ground is out of phase with the signal voltage impressed across the input lead and ground.

The most recent design of an interphone system features an individual amplifier at each interphone station rather than a central amplifier. As a result, the audio signal is amplified at the source and conducted at higher levels and requires a minimum of amplification at the other interphone stations. This prevents the low "noise" levels in the lines from being amplified to an objectionable degree. Furthermore, the high level lines (high signal-to-interference ratio) permit the use of dynamic microphones which increases intelligibility by 65 to 95 percent. Boom, mask, and hand-held microphones are examples of the interference-cancelling types in use.

The design also features a two-wire system which has a central ground return path. It is possible that through the use of higher level lines some shielding previously essential in the original design could be eliminated. However, since interference tests have not as yet been conducted, sufficient shielding has been incorporated in the new design to insure adequate interference attenuation.

2.1.1.5 Liaison Systems

Airborne liaison systems are designed primarily to serve as an air-to-ground communication link utilizing either voice or code transmission.

In general, such systems consist of a transmitter, transmitter key, fixed antenna, radio receiver, antenna coupler, dynamotor, and interconnecting cables. The transmitter and receiver operate over a frequency range of from 200 to 500 kc and 1.5 to 18.0 mc. An audio output signal provided by the receiver is available at any one of the interphone stations. The installation of a liaison set in a typical aircraft has been selected as an example of an airborne liaison system. The discussion to follow applies specifically to this particular system. However, because of the similarity of all such systems, there will be no loss of generality.

A fixed antenna is mounted on the top or under the fuselage near the radioman's compartment. All tuning adjustments are made at the transmitter or receiver in the radioman's compartment. No remote controls are provided. A liaison receiver audio output and a liaison transmitter microphone input are provided at each interphone control box. In general, one interphone control box is located at each of the stations for the pilot, co-pilot, navigator, and radio operator. The power source bus, antenna coupler, and Loran receiver are mounted in the compartment behind the pilot. There are no components located in the tail or wing sections. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Navigation and search radar controls, radio altimeter indicator, Loran indicator, etc., are examples of types of equipment whose wiring is frequently found in a typical bundle. Exciter regulators and radars are examples of equipments that would be mounted in the same aircraft sections.

Since in this system the receiver output is an audio signal in the headset, any unwanted signals which find their way into the receiver case must be capable either of producing audible interference in the headset or of preventing functioning of the receiver in some other way to constitute an interference problem. In complete systems such as this one the paths of entry of unwanted signals may be many and devious. In this case of the particular receiver in this representative system, the many paths of entry are illustrated in Figure 2-11. This figure shows also that while the functioning of the other components of the system may be unaffected by interference signals, they may serve as a part of a path of entry for the undesirable signals. For example, interference was picked up from the radar modulator on the lead between the dynamotor and the liaison transmitter which would not have occurred had this lead been routed differently. This interference did not affect the operation of the transmitter, but within the transmitter itself the dynamotor and antenna leads were bundled together and the interference reached the liaison receiver via the antenna connection. Here it was amplified and appeared as serious audio interference in the receiver output.

A brief explanation of the interference problems as shown in Figure 2-11 follows, some solutions are shown in Figure 2-12 but for the sake of clarity only a few are included.

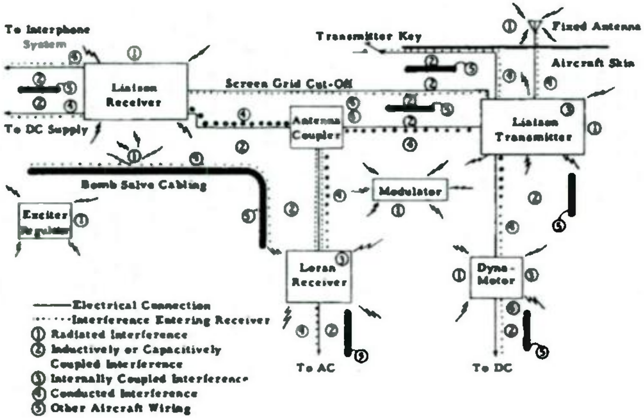


Figure 2-11. Paths of Interference Signals in a Liaison System

a. Interference was picked up directly on the open-wire antenna lead-in which was conducted to the liaison receiver causing unsatisfactory operation. It was determined that the antenna lead to the receiver could be shielded but that the transmitter antenna lead could not since the shielding capacitance would detune the transmitter output stage. The solution used is illustrated in Figure 2-12, and may be considered a satisfactory interim measure. Proper design, however, would permit the use of a shielded antenna lead-in for both the transmitter and receiver since harmonics as well as the carrier frequency of the transmitter are being radiated from the unshielded transmitter antenna lead-in and constitute a source of interference to other equipment.

b. Interference from the radar modulator was picked up on the lead between the dynamotor and the liaison transmitter and was conducted into the transmitter. Within the transmitter the antenna lead to one of the transmit-receive relays was bundled in with the dynamotor and other power wiring. The interference was thereby coupled into the antenna lead to the receiver and was conducted to the liaison receiver causing unsatisfactory operation. Removing the receiver antenna lead and rerouting it as in (a) above and as shown in Figure 2-12 is one possible solution. Another solution would be to isolate the antenna lead from the power wiring within the transmitter by shielding and/or rerouting. The latter would be preferable practice in initial design stages since it is poor practice in any event to bundle interference susceptible wiring with other wires likely to conduct radio interference. In other circumstances a filter in the lead from the dynamotor to the transmitter might be dictated, but only as a last resort since it would involve added equipment and weight.

c. Interference was conducted into the receiver on the interphone and DC power leads, which caused unsatisfactory operation. At this stage of design, filters were installed in the troublesome leads as shown in Figure 2-12 in order to reduce the interference to a tolerable level. This means added equipment, weight and cost. Preferably, the receiver should have been designed for less susceptibility to interference on these leads by improved shielding of the RF and/or IF stages and possibly better routing of these interphone and power leads within the receiver. Electrical cleanness (isolation) in the wiring of transmitters and receivers is preferable to mechanical neatness achieved by carefully bundling all the wires together into a cable and lacing them together.

d. In this particular installation the bomb salvo cabling (an entirely independent system) was routed very close to the exciter regulator and to the liaison receiver cover in the region of the cooling louvres. Interference was coupled from the exciter to the bomb salvo cabling and

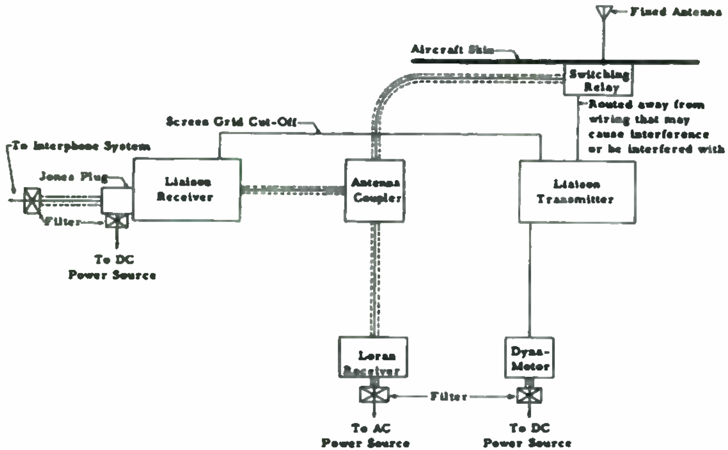


Figure 2-12. Modifications of the Liaison System to Minimize Interference

thence through the louvres in the receiver cover to the internal wiring of the receiver resulting in objectionable audio output in the headsets. Rerouting of the bomb salvo cabling in the neighborhood of the exciter regulator to a point not less than six inches away resulted in interference-free operation. Several faulty design practices are involved in creating this particular problem:

(1) The exciter regulator design was poor in that its generated interference was not suppressed properly.

(2) Cables should never be run close to known interference sources whenever practicable.

(3) The receiver case shielding properties were poor due to poor design. Wire mesh over the louvres and better bonding of the case to the frame are required.

e. Interference was coupled from the radar modulator to the AC power lead of the Loran receiver and was conducted into the case of this unit. Inside this receiver the antenna lead was bundled in with the power wiring and interference was therefore introduced into the liaison receiver on the antenna lead. One solution is the addition of a filter to suppress the interference as shown in Figure 2-12. Proper design of the internal wiring of the Loran receiver to properly isolate or shield the antenna lead would be preferable since the filter would then be unnecessary.

2.1.1.6 Loran Systems

The Loran system is a navigational aid that enables the operator in an aircraft to fix his position over land or sea by means of the reception on his receiver-indicator of special radio signals from the ground installations.

The ground installations consist of groups of transmitting stations operating on the same radio carrier frequency which emit a steady succession of pulses in all directions. The stations operate in pairs, a master station triggering a slave station by means of a radio link which synchronizes the pulses from the two stations.

The receiver-indicator receives and measures the timing of the pulsed signals and transcribes them to a visual indication on the Loran scope. With the aid of charts, tables, and tabulations the received signal can be interpreted and the location of the plane established.

The system operates in the high frequency range. The power supply is AC, 80 or 115 volts, 360 to 2460 cycles per second. In an aircraft the main units of the Loran system consist of an antenna, interconnecting cabling, either a passive coupler or a preamplifier coupler, and the receiver-indicator.

In a typical installation, the Loran system does not use its own antenna but couples onto the antenna of another installation. As a result unusually long runs of lead-in wire to the receiver-indicator are sometimes necessary. To compensate for the subsequent line-loss, a preamplifier is provided at the antenna coupler. When lead length is not excessive a passive coupler is used. The liaison system with its antenna approximates the needs of the Loran system and is the usual coupled system. In a heavy bomber installation, the Loran system uses the liaison antenna with an unshielded lead-in wire through a switching relay in the liaison transmitter to a preamplifier antenna coupler and a shielded lead-in wire from the antenna coupler to the receiver-indicator.

The Loran receiver, like any other receiver is subject to various sorts of interference. This interference may come from transmitters aboard the aircraft, aboard neighboring aircraft, from ground installations, or from deliberate enemy jamming. Typical interference patterns as seen on a scope would appear as shown in Figure 2-14.

The Loran receiver has not proven to be a source of interference itself. The local oscillator is adequately shielded, the case construction is apparently good. However, when operating a system, the performance has not been entirely satisfactory.

The unshielded antenna lead-in prior to the coupling stage has proven to be an efficient coupling path for radio interference currents. See Figure 2-13. The preamplifier at the antenna coupler amplifies the interference which is then conducted through the shielded lead-in into the receiver. In most cases of this interference, the Liaison receiver is also affected. By the nature of the reception of the two receivers, a higher level of interference can be tolerated in the Loran than in the Liaison so subsequent filtering, shielding, etc., of the interfering components that cleared up the interference on the Liaison also cleared it up on the Loran. (See Section 2.1.1.5 of this chapter.)

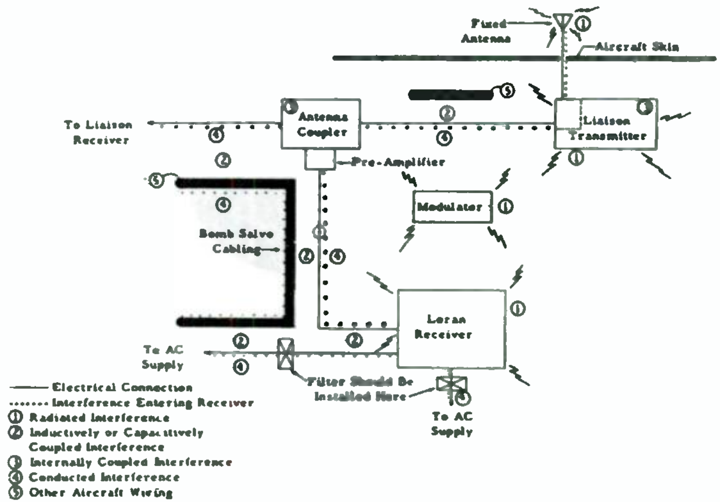
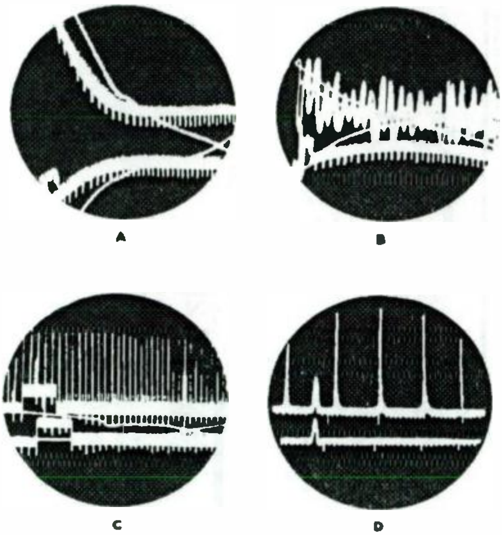


Figure 2-13. Paths of Interference Signals in a Typical Loran System



(A) CW or Radio Telegraph
(B) Radio Telephone
(C) Radar on Slow Sweep
(D) Radar on Fast Sweep

Figure 2-14. Typical Interference Patterns as Seen on Loran Indicator

Poor interference-free design is apparent when the AC power lead-in plug is a few inches from the antenna lead-in plug. This provides a coupling path onto the antenna lead-in cable for interference conducted on the power line. These interference currents ultimately affect the liaison receiver (see Section 2.1.1.5 on Liaison System) as well as appear as grass on the Loran scope. It was necessary to filter the AC line in order to eliminate this interference. Proper isolation in the original design would have eliminated the need for this filter.

The preamplifier produces interference problems particular to this installation. The non-linear elements in the preamplifier make the system susceptible to cross modulation. A simple illustration of this would be: The Loran receiver is tuned to 2 mc; a 3 mc and a 1 mc signal can mix in the preamplifier stage and the result be admitted as a 2 mc signal in the receiver.

Another problem presented by the preamplifier stage is the system's susceptibility to overloading because of the wide pass-band. An unwanted signal can be amplified to a proportion which overloads the front end of the Loran. In a heavy bomber installation, rendezvous equipment was producing interference in the Loran which was extremely serious because of its high magnitude and the dangerous fact that the interfering signals could be mistaken for an actual Loran station. The interference was coming from a strong, pulsed, radar signal at about 200 mc. It was admitted through the Loran antenna, amplified at the coupler stage, overloaded the front end of the receiver, and gave a visual signal on the indicator scope. Preliminary investigations and tests indicated that a low-pass filter inserted in series with the Loran antenna, between the preamplifier and the antenna, should eliminate most of the interference. Further tests are necessary to determine the feasibility of such a filter, and if so, the most efficient arrangement considering insertion loss and matching, and terminating impedances. Systems using the passive coupler have not been subject to this type of interference.

Improved original design of the preamplifier would have effectively reduced the susceptibility of the system to both cross modulation and overloading. Consideration should have been given to the rejection of frequencies of undesired signals. A tuning device on the front end of the amplifier would be one means of increasing rejection. Figure 2-15 shows such a tuning device. A band-pass filter with sharp cut-off characteristics covering the band of the Loran equipment as shown in Figure 2-16 would also give the necessary rejection characteristics.

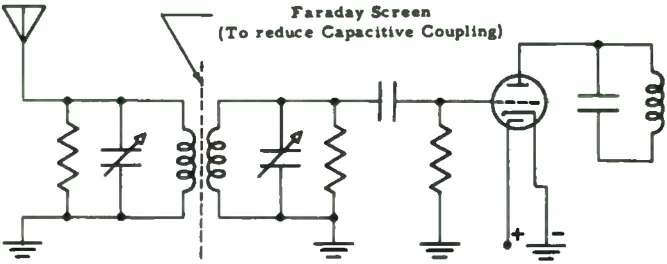


Figure 2-15. Tuning Device Installed
Ahead of Pre-Amplifier in Loran Receiver

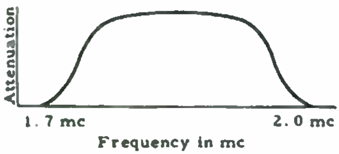


Figure 2-16. Attenuation Characteristic of Band-Pass
Filter for Use With Loran Receiver

The Loran system incorporates interference suppression techniques by providing for shielded lead-in cable, good case shielding against interference fields, and adequate shielding of the local oscillator to prevent interference with other equipment. Ideally, it should have its own antenna installation where it would be possible to minimize lead-in cable length; however, the increasing number of antennas in a bomber installation makes it necessary to double up with another system. Good installation practice can minimize this length, but it may still be long enough in some installations to make a preamplifier stage necessary. Rejection of undesirable frequencies should be provided for in the preamplifier to minimize malfunctioning. Poor design is exemplified by the proximity of the AC power plug and antenna lead-in plug. Coupling of interfering currents has occurred between the power lead and the antenna lead and filtering has been necessary. Proper isolation would have eliminated the need for this filter. The front-end rejection of undesirable frequencies in the receiver could be improved.

2.1.1.7 Shoran Systems

The Shoran Radio Set is a short range navigational system used in present day aircraft to determine the aircraft's position under instrument-flight conditions without visual reference to the surface of the earth or to celestial bodies. Two ground stations located a considerable distance apart are required in conjunction with the airborne installation in an operational system. During operation, pulses of amplitude-modulated radio waves of very high frequency are transmitted. Radiation from groundstations is moderately directional; radiation from the airplane is non-directional.

A typical Shoran system is shown in Figure 2-17. Two identical antennas and their bases (one for receiving and one for transmitting) are provided. The antennas must preferably be mounted on a large, relatively flat, metallic surface of the airplane. The particular location of each antenna depends on the type of airplane in which the equipment is installed and must be predetermined for each type by radiation field measurements. The antennas should preferably be shielded from each other by the fuselage or wings of the airplane. Also consideration should be given to the location in respect to the other receivers and transmitters in the airplane. In addition, attention is called to the fact that the transmission line (coaxial cable) which connects each antenna to the equipment enters the base of the antenna through a feed-through insulator and must be so laid out that the transmission lines make bends of no less than 4-inch radius to prevent damage to the lines.

The indicator is shock mounted in a frame which allows for a computer unit (when used). The receiver is installed within the indicator so that the controls of both units are accessible from the same position.

The indicator unit includes the circuits for generating the various sweep and blanking voltages necessary for operation of the indicator tube. It also contains timing and phasing circuits, pulse-selector circuits, and keying-pulse-generating circuits for controlling the operation of the airplane transmitter. The indicator is cooled by air circulation which is provided by a blower motor within the unit with unscreened louvres over the air intake. The cathode ray indicator tube operates with a circular sweep and the signals, applied to an axial electrode within the tube, cause

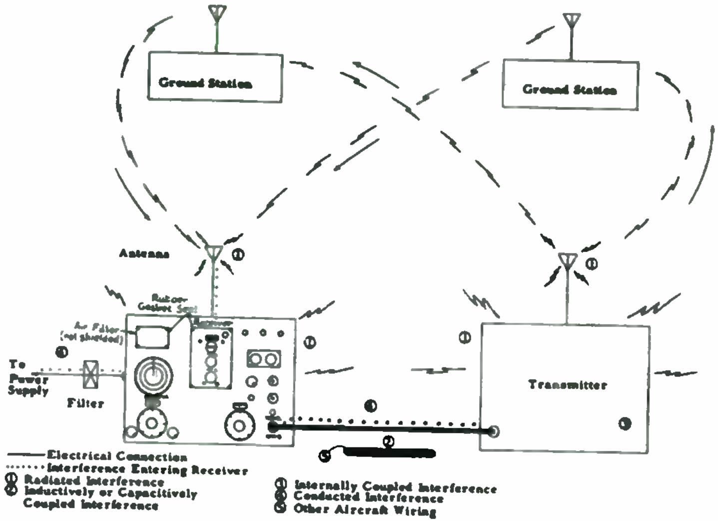


Figure 2-17. Paths of Interference Signals in a Typical Shoran System

radial deflections of the electron beam. The signals thus appear as inner or outer pulses protruding from a circle on the screen of the indicator tube, and their relative positions serve to determine the position of the airplane.

The receiver utilizes a superheterodyne circuit through which signals from the ground stations are amplified and converted into a video signal. This signal is than applied to the axial electrode of the indicator tube. The oscillator is the main source of interference. It radiates from the case, external wiring, and antenna.

The case is described as a dust cover and that is the only function it serves. There are rubber gaskets between the air filter and indicator case, the receiver and indicator case, and the connected sides of the indicator case. There is no screening on the front panel to prevent radiation both in and out at the various controls and other breaks in the case. The case is poorly constructed with loose tolerances and of non-rigid construction so that sides buckle in with little applied pressure. Tests in a shielded room using a probe showed high levels of oscillator radiation from the rubber-gasketed joints. Paint was scraped off the contact surfaces of these joints and shielding gasket braid was substituted for rubber gaskets. Shielding was applied to all apertures in the front panel so that contact with the control rods was made. Screening was placed over the air filter for shielding purposes. These measures effectively "bottled up" the oscillator radiation from the case.

All these paths of leakage described above also serve as paths of entry into the receiver case for radio interference. Difficulties have been encountered in installations where it was necessary to shut off other transmitting equipment in order to operate Shoran systems. Information as to coupling paths is not available and particular examples are classified material. It is assumed that the corrective measures taken to contain the oscillator radiation will also serve to keep out radio interference; however, tests were only conducted in a shielded room and it has not been proved in an actual installation.

Installation difficulties have been encountered with the Shoran transmitter in interfering with other receivers by both shock excitation and the generation of harmonics. The liaison receiver has been particularly susceptible, although other receivers have been affected by the Shoran transmitter. In a typical aircraft installation, the following modifications were necessary to produce a satisfactory system.

a. The mounting of the liaison receiver was modified by the addition of grounding straps mounted inside of each shock absorber housing and by the removal of paint around the mounting bolts.

b. The back end of the liaison receiver was filtered to prevent radio interference from Shoran transmitters and a wave trap was connected to the antenna binding post to prevent shock excitation.

c. On the radio compass a choke coil was installed at the antenna binding post which acts as a low-pass filter to prevent shock excitation from Shoran transmitters.

d. A capacitor was installed on the radio compass "look" antenna (reason not known).

Paths of entry into receivers have been both from the back and the front ends. Filters have been installed in the back end and wave traps in the antenna leads. No information is available as to the coupling paths in any particular installation. Operating at a very high frequency and being pulse modulated, the Shoran transmitter has caused difficulties by overloading the receivers in the airplane and causing shock excitation. Most of the difficulty has been through antenna radiation although interference has been radiating from the case and from external wiring.

These difficulties can be overcome by incorporating the following features. Screening should be put across the air filter and conductive mesh gasket material installed instead of rubber gasket. All external wiring should be shielded to prevent radiation and possible coupling. The transmitter should be installed in the plane so as to take advantage of shielding by bulkheads, etc. Wiring should be routed to prevent coupling. Rejection of receivers can be improved by adding wave traps to antenna lead-ins and by filtering the back end when feasible.

In a shielded room with Shoran transmitter and liaison receiver antennas set 6 feet apart, a high level of interference was recorded in the VHF and UHF frequency range. The liaison receiver was picking up harmonics of the fundamental with spurious radiations, harmonics of the pulse repetition rate with spurious radiations, and shock excitations from the fundamental. Better internal circuit design is necessary to prevent the generation of harmonics (see Section 2.1.1.2 on VHF System). Specific examples of bad practice in circuit design of Shoran transmitters are not available.

2.1.1.8 Navigational Radar Set

A navigational radar set is a dual-purpose equipment designed to operate by switch selection either as a radar beacon or an interrogator-responder. When used as an interrogator-responder, its function is to guide the airplane within 200 yards of the beacon location. As a beacon, it responds to other interrogator-responder radars within its frequency range. As shown in Figure 2-18, a representative set would consist of: (a) receiver-transmitter, (b) indicator, (c) control box (power), (d) control box (frequency), (e) video gate, (f) transmitting antenna, (g) receiving antennas, (h) reflector antennas, and (i) mountings, cables, interconnecting plugs, etc.

The functional operation of a typical radar set when used as an interrogator-responder includes the following. The pulse-generating circuits furnish the modulator with pulses at various rates which may or may not be synchronized with associated radar equipment. The pulse generating circuits also furnish a synchronizing pulse for the indicator and a suppression pulse for the IFF equipment. The pulses from the modulator cause the transmitting oscillator to oscillate for four to six micro-seconds at the pulse repetition rate. This radio-frequency energy is fed to the transmitting antenna and the transmitted energy is received by an associated transponder, which transmits a radio-frequency reply. The reply is picked up by the receiving antennas, the outputs of which are alternately fed into the receiver section by the antenna switch located in the indicator. The video output of the receiver is synchronously switched with the receiving antenna in the indicator and applied to the horizontal plates of the cathode-ray tube in the indicator.

When the radar set is used as a transponder (beacon), the video output is connected to the pulse-generating circuit. Thus the transmitter is triggered each time a pulse is received from the associated interrogator-responder.

The set operates in the VHF range on any one of eight preset frequencies. The receiver-transmitter unit is designed to maintain a high degree of frequency stability; the frequency drift of both components is less than one megacycle between ambient temperatures, -55° C. and +71° C. Power is obtained from a 115-volt, 400-1600 cps supply. In addition, a 24-28 volts, 2-ampere, DC source is required.

The transmitting antenna is usually installed on the underside of the nose of the airplane and, when possible, in the center of a 36-inch diameter, flat surface. Best results can be achieved if the antenna is within ten degrees of vertical to the airplane's line of level flight.

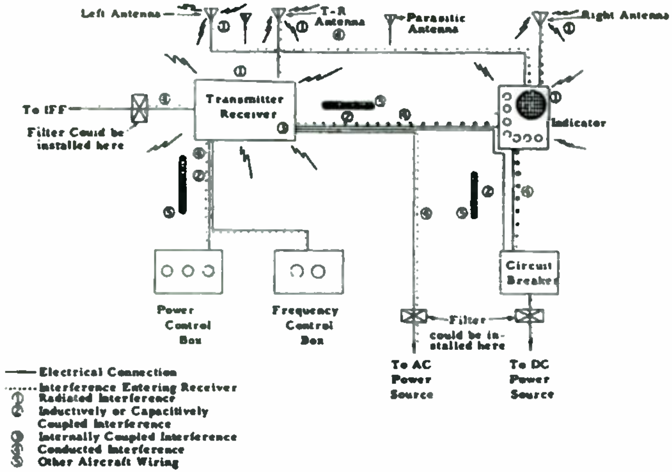


Figure 2-18. Paths of Interference Signals in a Typical Radar Navigation System

Receiving antennas are installed on each side of the airplane, symmetrical about the line of flight. The exact location for best results would have to be determined experimentally because it would differ with each model of plane but generally the two antennas are installed on the underside of the wing, outboard from a vertical reflecting surface. Installation of the indicator and receiver-transmitter in the plane should be convenient to the operator while sitting in a comfortable position to view the display tube. A minimum of wiring length and isolation from sensitive receivers are also installation considerations.

Coaxial cable is used from the receiver-transmitter to the antennas and from the receiver-transmitter to the indicator. Power leads and control wiring are unshielded. As installed in a typical airplane, an analysis revealed that the internal circuitry of the receiver-transmitter was not originally designed to suppress modulator interference. Wires carrying interference were coupled with "clean" leads and the cooling blower motor was generating undesirable voltages which interfered with the receiver-transmitter as well as with other equipment. Harmonics and overloading from pulses coming from antenna radiation affected other receivers. Interference fields were leaking out of the case and getting to the power and control wiring which then proveded a coupling path to any adjacent wiring. The video and antenna leads were shielded according to good design practice; however, neither the AC nor DC power leads were filtered and hence constituted a source of interference. Also the lobe-switch drive motor was found to be an interference generator. While not a source of interference in itself, the interconnecting wiring provided a coupling path for interference generated by the receiver-transmitter.

In order to suppress interference from this typical navigational radar set numerous modifications were necessary. The following list of modifications were recommended by the manufacturer:

a. The blower motor was changed from DC to AC to eliminate triggering of the beacon function by brush-spark interference and to prevent interference in other equipment.

b. The receptacles for power and control wiring were changed to a new type having ceramic by-pass condensers integral with the contact-pin construction to reduce leakage from the receiver oscillator and other interference sources in the unit.

c. A shield was designed and installed on the radio-frequency tuner chassis of the receiver to enclose and confine the field of the line-type oscillator. Since it was not considered in the original design, the shield had to be modified to provide clearance holes for turret-wheel contacts and the openings used for access to trimmer condensers had to be capped. This added shielding materially reduced oscillator leakage and also suppressed the oscillations radiating from the radio-frequency stage which had occurred in the original design.

d. Additional bonding was necessary for the RF tuner chassis to reduce chassis potential and leakage. It was also necessary to install chokes, by-pass condensers, and filter resistors both in and outside of the RF tuner to reduce leakage and radiated interference. A series resistor was installed in the mixer-plate output lead to reduce the oscillator leakage through the IF stages into the circuit wiring and radiation from the receiver.

e. A small amount of delay bias on the video detector diode reduced the interference due to triggering of the beacon.

f. The screen resistor of the RF amplifier tube was lowered from 33 to 10 kilohms to increase stage gain and improve the signal to interference ratio of the receiver.

g. The tuning-slug screw in the oscillator-vernier mounting block was found to move with normal vibration and caused both amplitude and frequency modulation of the oscillator. A detent device was incorporated to prevent this vibration and correct the difficulty.

h. Decoupling resistors were provided in the suppressor input and output circuits so that interference leakage along the interconnecting cables would not cause faulty operation of paired units in beacon and interrogator-responder service.

i. The receiver-transmitter plate-supply voltage was lowered to approximately 265 volts, which is 30 volts less than that used in the original design. The loss of receiver gain which accompanied this change was made up by lowering the cathode resistor of the fifth IF stage from 470 to 270 ohms. The chief advantage of this change was to lower the plate dissipation in all the tubes, especially in the 6AK5 of the RF stage.

Lowered plate voltage for the transmitter also was intended to eliminate voltage breakdown and reduce corona on high voltage leads and terminals at high altitudes, a condition which had made the triggering due to interference so severe that the beacon function was not usable. Transmitter-power output indication was satisfactory at the lowered plate voltage provided a tube with ample filament emission was used in the power-output indicator device. In addition, several detailed changes were made to increase air-gap length and decrease corona on the high-voltage, power supply layout. A recent investigation on a heavy bomber installation revealed that the navigational radar was producing interference in the command receiver, the radio compass, the liaison receiver, the VHF receiver, and the Loran receiver. The interference produced in the Loran was particularly serious because of its high magnitude and the dangerous fact that it can easily be mistaken for an actual Loran station.

Preliminary tests indicated that the interference was being radiated from the radar transmitting antenna and being picked up by the antennas of the above receivers. Overloading was occurring and the interfering signal was being admitted. A low-pass filter was inserted in series with the Loran antenna and it appeared that most of the interfering currents were eliminated. Further investigations have to be made to determine the feasibility of this type of filter as well as to consider the insertion loss and the matching and terminating impedances.

The above discussion should serve as an excellent example of what is likely to happen if proper design practices are not made an integral part of all phases of design. While it may be admitted that this system was made operable by making certain modifications, it must be understood that fundamentally "fixes" are inefficient and costly in terms of time, weight, and space.

2.1.1.9 Radio Altimeter System

Airborne radio altimeter systems are designed to indicate the altitude of the aircraft above the terrain by accurately measuring the time required for a radio signal to travel to the earth's surface and return, and interpreting this time interval in terms of distance in feet above the reflecting surface directly under the aircraft.

Representative equipment is designed to emit a frequency-modulated radio wave in a downward direction from the transmitter antenna. For both the low and high ranges, 0 - 400 feet and 400 - 4000 feet respectively, the carrier signal is in the UHF range. The earth's surface reflects a portion of this radiated energy, and this is received on a separate receiver antenna. During the time interval required for the signal to travel to earth and return, the frequency of the transmitter will have changed. A small fraction of the transmitter signal is transmitted directly to the receiver and mixed with the reflected signal from the earth. Since the transmitter is continuously changing its frequency, there will be a difference in frequency between the direct and reflected signal at any instant of time and this difference will be proportional to the distance of the aircraft above ground. The detector circuit converts this difference in frequency to a direct current which is indicated by a meter calibrated directly in feet of altitude above the ground. The detected current also operates a pair of relays which are preset to energize altitude-limit indicators.

In general, radio altimeter systems are composed of (1) transmitter antenna and receiver antenna, (2) a transmitter-receiver, (3) indicators, (4) limit-switch assembly, (5) limit-light indicator, and (6) inter-connecting cables. The installation of a representative radio altimeter set in a typical aircraft has been selected as an example for discussion of design techniques for interference-free operation.

The transmitter and receiver antennas used on the radio altimeter are identical and interchangeable. Each antenna consists of a one-half wavelength dipole on a suitable mounting structure with provisions for connecting a 50-ohm transmission line through a coaxial elbow adapter. The antennas are mounted on a suitable metallic surface which acts as a reflector. The antenna system is designed to radiate and receive maximum signal strength in a generally downward direction with a minimum transfer of energy from the transmitting antenna to the receiving antenna, and to minimize variations in the received reflected signal while executing a reasonable dive, climb, or banking maneuver. The pilot's indicator, limit and marker lights are mounted within reach of the pilot when seated at the flight controls. There is no audio output for the radio altimeter; all indications are visual.

The transmitter-receiver unit is usually located as close to the antenna installations as practicable and consequently appears in the compartment nearest the antennas. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Navigational radar receiver-transmitter controls, search radar synchronizer, navigational radar indicator, Loran indicator, and radio altimeter indicator, are examples of types of wiring that are frequently found in a typical bundle.

Radio interference signals have gained access to the receiver circuits over paths of entry, as shown in Figure 2-19, and caused unsatisfactory operation of the radio altimeter system. Interference signals generated by the system may be introduced into other aircraft wiring systems over the same routes. In the radio altimeter system only the transmitter-receiver combination need be considered as a potential

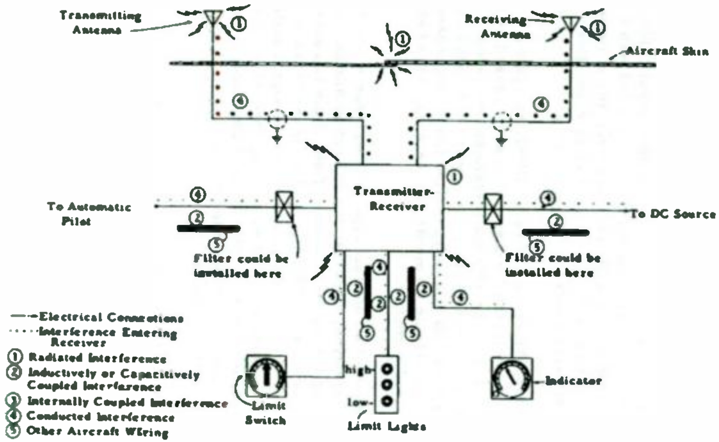


Figure 2-19. Paths of Interference Signals in a Typical Radio Altimeter System

source of interference. It is also the only component of the system that is susceptible to interference.

In general, interfering signals can enter the transmitter-receiver case through the following paths: (a) transmitting-antenna lead, (b) receiving antenna lead, (c) automatic-pilot altitude-control cable, (d) limit-switch cable, (e) indicator cable, and (f) the power-input cable. Interfering signals can be introduced into the receiver over any of these paths by conduction along the cables or through inductive or capacitive coupling between other aircraft wiring and the system wiring.

Interference problems in the radio altimeter system of the typical aircraft installation selected for discussion were caused by the following: (a) loose connections in the antenna transmission system due to the fact that plugs, adapters, receptacles, etc., were not tight or properly meshed and (b) field modulation which resulted from "secondary feed-through" paths having variable elements and were capable of effectively modulating the received signal.

As an example of an unusual type of interference which may affect radio altimeter systems, consider the reflecting surface of a propeller which is moving through the common field of the two antennas. The intermittent reflected signal thus picked up by the receiving antenna combines with and effectively modulates the desired signal reflected by the earth. If this field modulation, after passing through the detector and audio-frequency amplifier stages is of sufficient amplitude to operate the square-wave limiter, counts will be added to the limiter output signal and an excessive altitude will be indicated.

A similar but more erratic effect is produced if there is a loose metallic part of the aircraft within the fields of the antennas which can shift or vibrate so as to cause an intermittent or rubbing contact with another metal part or surface. In this case, induced currents, set up in one of the surfaces by the transmitted signal, are modified both in phase and in amplitude by intermittent contact with the other surface similarly charged. Thus, transient currents, covering a wide band of frequencies, are produced in the field of the receiver antenna. Unusual sources of such trouble are bomb bay doors, loose access plates, loose wheel fairing, trailing wire antennas, etc. When a source of field modulation is found, an attempt should be made to correct the condition either by electrically bonding or by completely insulating the two surfaces which are causing the trouble. If such correction is not feasible, it may be necessary to relocate the antennas so as to remove the source from the common field of the antennas.

Difficulties have been encountered in a typical aircraft installation where audio ripple originating in a motor has been conducted by the power leads into the transmitter and receiver affecting the altitude indicator by as much as 300 feet on the high range scale. The altitude indicator is in a particularly susceptible position because of the circuit arrangement. In a controlled situation, an audio oscillator could vary readings by 300 feet with little difficulty. In a typical installation, a 250 μfd capacitor was installed at an aileron booster motor in order to suppress the audio ripple that was getting in through power leads and affecting the indicator. Proper transmitter-receiver design should have made the unit insusceptible to interference on these leads.

Interference in radio altimeters often manifests itself by (1) erratic altitude indications observed as jumps of 200 - 400 feet when flying over land at altitudes from 400 - 3000 feet, and (2) marked increase in altitude indication when switching over the high range while in flight at altitudes near 400 feet. It has been found that these effects are usually caused by electrical interference which effectively adds counts to the audio-frequency signal output of the detector. Such conditions are most likely to be encountered when the reflected signal is relatively low in amplitude as when flying over dry land.

The difficulties encountered with this system as explained above and as shown in Figure 2-19 could have been avoided had proper design and installation techniques been adhered to from the start. Two types of problems can be shown to exist by inspection of the figure. One group concerns the equipment designer and may be outlined as follows:

a. The antenna lead-ins may be considered potential sources of interference to the system itself. Proper design will permit both to be completely shielded from the antennas to the transmitter-receiver, including correctly matched input and output circuits.

b. Cabling between the transmitter-receiver and other units to which it must be connected may be considered similarly. Proper design of the internal circuits of the transmitter-receiver will insure that they are not susceptible to interference picked up on the cables. Additionally, the design should be such that interference will not be conducted out of the transmitter-receiver on these cables. If a signal essential to the system operation is to be conducted between units of the system and which might cause interference in other equipment, the signal should be contained within the cable by shielding.

The second group of problems concerns the installation designer and may be outlined as follows:

a. The antennas should be installed outside interference fields of other equipment and antennas.

b. Install the antennas so that the receiving antenna is shielded from the transmitting antenna.

c. Install the other components of the system in as close proximity as possible using proper interconnecting cabling, and where practicable, away from strong interference sources and fields.

2.1.1.10 Radio Compass System

Airborne radio compass systems in use today are basically similar in function and design. The system consists of a radio receiver using a superheterodyne circuit and certain additional circuits necessary for radio compass operation. Two remote control boxes and two indicators permit operation of the compass from either of two separate positions in the aircraft. A relay unit including an autotransformer is provided to permit switching from one control box to the other. A vertical rod, non-directional antenna and a center-tapped, loop antenna together with the necessary interconnecting cords and power cable complete the system.

In general, radio compass systems operate over a frequency range of approximately 100 to 2000 kilocycles and are capable of providing:

a. an automatic visual bearing indication of the direction of arrival of radio frequency signals,

b. aural reception of modulated radio-frequency energy using either a non-directional or loop antenna, and

c. aural-null directional indications of the arrival of modulated radio-frequency signals using a loop antenna.

The installation of a representative radio compass set in a typical aircraft has been selected as an example. The discussion to follow applies specifically to this particular system. However, because of the similarity of all such systems, no sacrifice of generality is suffered.

Loop and sense antennas are mounted outside the aircraft on the top or under the fuselage near the pilot and navigator compartments. The pilot's control panel and indicator are mounted within reach of the pilot when seated at the flight controls. There is another remote control panel and indicator mounted in the navigator's compartment. A radio-compass audio output is provided at each interphone control box. In general, there would be a pilot, co-pilot, navigator, and radio operator interphone control box, each located within easy reach of the crew member. The relay, power shield, and receiver are mounted in the compartment behind the pilot. An inverter in the same compartment provides a 400 cycle per second, ac supply to the radio compass system. There are no components located in the tail section or wing sections. Interconnecting cords and cables are generally bundled with other electrical wiring passing through the compartment. Radar sets, computers, and auto-pilot trays are frequently mounted in the same compartment with the radio compass receiver.

Radio interference signals have caused unsatisfactory operation of the radio compass system. Visual bearing indications are particularly susceptible to interfering signals due to the fact that unwanted voltages appearing on the grids of the thyratrons, which position the loop antenna, can override the desired signal and cause erroneous or unstable visual indications. Aural reception and aural-null indications may become unintelligible when interfering signals of sufficient nuisance value are allowed to produce audible sounds in the headset.

There are two components in the radio compass system that must be considered as potential interference generators: (1) the drive motor in the loop antenna, and (2) the receiver itself. Design considerations for these components are discussed in Volume I, Chapter 1, Section 7.1.1, and Chapter 1, Section 1.4 of this Volume, respectively. The receiver requires particular attention because of its local oscillator and thyratron output.

An analysis of each component in the compass system reveals that only the receiver is susceptible to radio interference. Even though the functioning of the other components of the system is unaffected by interference signals, they may serve as paths of entry for undesired signals.

In general, interfering signals can enter the receiver case itself, through any one of the following paths, as shown in Figure 2-20:

a. Separate inverter ac power supply cable,
b. Dc power supply cable,
c. Loop and non-directional antenna cables,
d. Flexible tuning shafts,

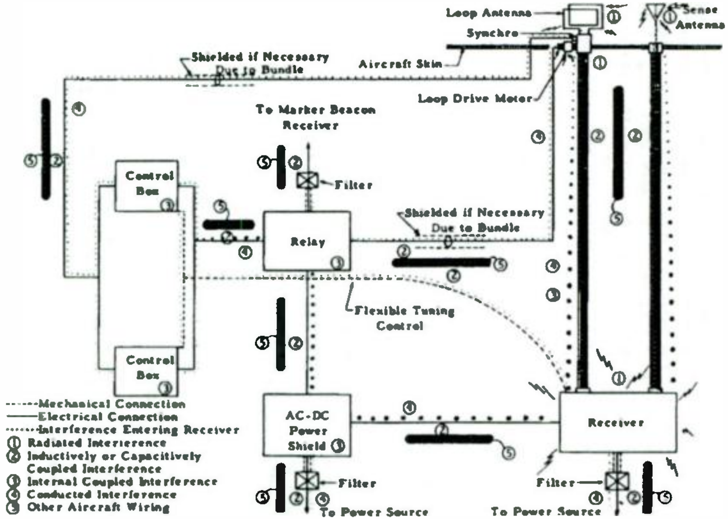


Figure 2-20. Paths of Interference Signals in a Typical Radio Compass System

e. Marker beacon receiving equipment, 14 to 28 and 220 volt dc supply cable,
f. Remote control connecting cables,
g. Headphone output cord, and
h. Penetration of the receiver case or shield.

Interfering signals may enter the radio compass receiver by conduction along the cables as a result of inductive or capacitive coupling links between any of the system cables and other aircraft wiring. Such cable connections are those between the control boxes and the relay, relay and marker beacon receiver, relay and ac-dc power shield, relay and loop antenna, and indirectly, those from the loop-antenna motor control circuits.

Unwanted signals can also be introduced into the receiver through any one or more of the five openings in the metal receiver case. These openings are provided for the loop antenna and sense antenna leads, the ac-dc power cable, the headset or interphone audio-output cord, and the flexible tuning linkages. Interfering signals can be inductively or capacitively coupled into these cables directly and gain access to the receiver case. In a typical radio compass installation, an auto-pilot tray was located close to the receiver and the power cables of the auto-pilot tray paralleled the antenna leads and power cable of the receiver. This difficulty was increased by the use of an unshielded sense-antenna wire to the radio compass receiver. Interference was reduced by substituting a shielded coaxial cable for the unshielded wire, rerouting the loop cable away from the auto-pilot cabling, and braid shielding the entire cable bundle. These steps could have been taken in the original installation; however, the offending source, namely, the auto-pilot, should have been rendered interference-free by the manufacturer in the original design by application of techniques outlines in Section 1. Observance of good design techniques would have obviated the necessity for the preceding fixes.

Figure 2-20 illustrates the most vulnerable points for interference signals to enter the radio compass system. Unwanted signals picked up by the marker beacon antenna can reach the receiver through the marker beacon receiver and relay to the ac-dc power shield. From the power shield, a direct path is provided over the power cables to the receiver. Disturbing signals picked up by the loop or sense antennas have a direct path into the receiver case over the antenna wiring.

The degree of freedom from interference attained in any radio compass system will depend upon the care of the designer in applying radio interference elimination techniques, as discussed in other paragraphs of this book. The major interference in radio compass systems is due to conducted signals over the system wiring and also to inductively or capacitively coupled signals into the system wiring. In radio compass systems, particular attention must be paid to the filtering, shielding, and bonding necessary to avoid conducted and radiated interference from neighboring systems operating from the same power supplies and in close proximity. An effort should be made to avoid bundling the system cables with other cables that are likely to couple unwanted signals into the compass system. Also, the components of the system, as well as the wiring, should be located and routed so as to avoid proximity to obvious or probable interference generators.

Components should be placed and circuitry arranged in such order as to result in a minimum susceptibility to interference. Reference to Figure 2-20 shows that the compass system wiring is directly connected to other electrical systems in the aircraft through three connecting cables, i.e., (1) receiver audio output to interphone system, (2) relay power supply to marker beacon receiver, (3) ac-dc power supply to power shield. These lines can be filtered to prevent interference from entering or leaving the system by conduction. Details of filtering and filter design are discussed in Chapter 4 of this volume. Proper design techniques, as outlined in other sections of the book, however, should have been applied in the initial design so as to make the equipment insusceptible to interference on these leads. There is also the possibility that the components of the radio compass system may not be at the same ground potential for a certain band of interference frequencies. Such a condition frequently occurs, due to varying impedance-to-ground characteristics for the circuit components, and consequently provides a path for circulatory interfering currents. Proper bonding will prevent these conditions, as pointed out in Chapter 3 of this volume. Each component should be adequately bonded, and each shielded cable should be bonded, not only at its terminal points, but also at each supporting strap, in the case of long cables.

Loop antenna and sense antenna leads should be electrostatically shielded. The synchro connections from the loop antenna to the pilot and navigator indicator are generally bundled together with other aircraft wiring and provide a means of coupling interference from the loop drive motor into other aircraft electrical and electronic systems. Consideration must be given, also, to the possibility of coupling interference from the other systems into the receiver through the loop antenna lead-in. If these cables are bundled with other susceptible-system wiring, or other probable interference-generating system wiring, they should be shielded, unless rerouting is practicable. The example, discussed above, where the autopilot caused considerable interference in the compass system, is a good illustration of the difficulties that can result when proper shielding techniques are not observed. Basically, however, the auto-pilot is the offender and source suppression of interference was not properly considered in its design.

2.1.1.11 Radar Fire-Control System

Radar fire-control and gun-laying radar systems (see Paragraph 2. 1. 1. 12) have essentially the same components and serve basically the same function. These two representative systems were chosen to contrast design techniques pertinent to radio interference.

The radar fire-control system chosen is an all-weather system designed for high altitude interceptors that employ fixed armament. It has two primary functions, search and attack. In two place aircraft a transitional function, hand-tracking, is provided for the radar operator. In the search function, the fire-control system detects the enemy aircraft. Once the target has been located, the operator directs the antenna toward the target by hand-tracking and places the equipment on automatic track. At this time, the pilot's indicator provides the attack display in terms of range, azimuth, and elevation of the target, and an artificial horizon display of his own altitude.

A fire-control system has many components installed in separate housings. The recommended practice of incorporating as many units as practicable in one housing to minimize interconnecting wiring is not applicable in this system because of the space limitations in fighter aircraft. Many smaller units facilitate installation engineering. However, the resulting maze of interconnecting wiring magnifies the radio interference problem.

Components of such a system would include: antenna, radar modulator, receiver-transmitter, servo amplifier, computer, range servo, dimmer control, rocket-setting control, sight head, range indicator, power supply control, power supply, manual range control, gyroscope, electronic control amplifier, gun-laying radar central, two azimuth-elevation indicators, antenna (hand control), blowers, wave guide, and cabling. The main sources of interference in this typical fire-control system would include the antenna, receiver-transmitter, and the modulator.

The antenna unit which includes potential interference generators such as relay switches, servo motors, and drive motors, necessitates careful consideration in the design stage. In this system's original design, relays were suppressed by the addition of capacitors in parallel with the relay switch. Feed-through capacitors (0.01 μf) were also included in series with the relay input leads. However, circuit arrangement was such as to couple interference back into the filtered lead nullifying the suppression techniques. More effective suppression could have been had with the addition of a resistance-capacitance circuit in parallel with the switch rather than a capacitor; proper routing of leads would have eliminated coupling of interference back into the "clean" leads. Filters were installed on the output leads of the servo motors. However, again, the "clean" leads out of the filter were exposed to interference coupling.

The receiver-transmitter (RF Unit) proved to be an indirect source of interference currents by providing a coupling path for interference from the modulator pulse. The pulse entered the receiver-transmitter through a shielded cable and the interference coupled through the circuits to appear finally on the external wiring enabling the interference to couple into other airplane wiring. Improved shielding or filtering techniques within the receiver-transmitter could have eliminated this coupling path. Case construction of the receiver-transmitter was poor because it utilized a rubber gasket for its pressurized seal. Radiated interference from currents originating in the modulator was leaking from the case; conductive gasket material substituted for the rubber gasket would have eliminated this source of leakage.

The pulse cable between the modulator and the receiver-transmitter proved to be a severe source of radiated interference. A loop probe measured 12,000 microvolts of radiated interference. The cable used was of approved coaxial type with four layers of shielding; however, the connectors or plugs provided poor shielding continuity with the cable, and leakage caused interference to appear on all parts of the coaxial line and to radiate or couple into other airplane wiring. A different type of coaxial line and plugs was used and materially reduced the interference. The coaxial cable replacement had inferior shielding properties (2 layers of shielding); however, good continuity was achieved with the new plugs. Better results could have been obtained if the original cable had been used with an improved type of plug.

The modulator in this system, a source of interference by the nature of the function it has to perform (see Chapter 1, Section 1.3.4 on Modulators), incorporated widespread shielding of internal wiring. The effectiveness of this shielding was considerably reduced by leaving about four inches of pulse cable unshielded within the case from the plug to the capacitor. This enabled interference to radiate to all susceptible leads and circuits within the modulator and also to be radiated and conducted from the case. The most efficient pick-up of this interference was by the coils of relays and blowers. Subsequently, the interference could be conducted out of the case by the relay and blower, power and control leads. Shielding effectiveness was further reduced by not adequately bonding the shielding conduit and by using throughout the modulator what amounted to "floating" shields. As in the receiver-transmitters, the modulator employed a rubber gasket to maintain its pressurized seal which permitted radiation of interference from the case. Shielding braid substituted for the rubber gasket again would reduce this case radiation.

Poor design practices observed throughout the set would include:

a. "Noisy" and "clean" leads were bundled together, a malpractice which permitted coupling of interference into "clean" leads.

b. Filter and by-pass capacitors were made ineffective by routing the filtered lead back through the interference field.

c. Filters were not mounted close to the interference source, and the portion of the lead between the source and the filter was left unshielded and free to radiate.

d. Shielding braid was not grounded.

e. Ground wires were unnecessarily long, a poor design practice which creates high impedance paths and permits radiation of radio interference from these ground leads.

To illustrate the importance of proper routing, grounding, and shielding techniques, two tables have been prepared which show conducted and radiated interference before and after the following remedial actions were taken:

a. "Noisy" leads were shielded, or when shielding was impractical, rerouted or filtered.

b. Rerouting of leads coming out of interference suppression units was accomplished so as to avoid coupling of the original interference back into the "clean" lead.

c. Filters were mounted as close to the interference source as possible.

d. All "floating" shields were effectively grounded at both ends and at intermediate points when necessary with leads whose length was no longer than two inches.

(e) Other ground leads, some as long as 10 feet, were eliminated and short leads (2 inches) substituted.

(f) The pulse cable and plugs were improved and properly grounded.

(g) Shielding braid was placed between the mating surfaces of the modulator and the receiver-transmitter.

(h) Leads conducting interference out of the modulator and receiver-transmitter were filtered at a point as close to the pin connector as possible.

The results of the tests demonstrated in Figure 2-21 and Figure 2-22 were conducted in accordance with Specification MIL-I-6181D. The corrective action taken on this fire-control system was not intended to be complete, but merely to demonstrate the appreciable improvements that can be had by proper design considerations.

The interaction of this equipment with other aircraft systems can be demonstrated by the following installation difficulties which also point out other important design considerations.

In a fighter installation, two troublesome sources of radiated interference were the antenna spin motor and the thyratrons in the antenna drive-motor circuits. The spin-motor incorporated a current-interrupting governor, the contacts of which proved to be an efficient interference-generating device - affecting the liaison and intercommunication equipment. Another motor was tried and while commutor interference was higher than in the original model, the interference pick-up, when installed, was negligible. Apparently, the type of interference from the second motor (commutator) was more easily suppressed than in the original model. Proper original motor design would have eliminated the need for modification (see Volume I, Chapter 1, Section 7.1.1 on Rotating Machinery).

The thyratrons fed into a changing load and standing waves were unavoidable on the leads from the thyratrons to the antenna drive motors. Thyratrons, a non-linear device, also produce steep wave fronts and, consequently, are rich in harmonics. External fixes to suppress interference originating from thyratrons would include shielding, bonding, and filtering of leads carrying the interference. Internally suppressor circuits such as resistance-capacitance and inductance-capacitance combinations should be installed as close to the thyratrons as possible. When

RADIATED INTERFERENCE IN MICROVOLTS BEFORE SUPPRESSION TECHNIQUES AND AFTER				
Freq. in MC	Before		After	
	Search	Hand Control	Search	Hand Control
0.16	31	80	8	7
0.18	37	100	12	12
0.24	60	145	12	12
0.28	85	160	13	13
0.32	100	180	11	10
0.6	85	140	10	10
1.0	54	100	9	8
1.4	26	58	3	3
1.8	7	34	3	2
2.8	16	16	2.5	2
3.8	11	12	2	1
5	10	11	4	4
7	11	15	2	2
9	5	5	2	1.5
11	14	14	1.5	1.0
14	14	14	2	2
16	12	15	2	1
18	65	15	2.5	1
20	4	5	3	2
30	100	100	80	30
32	250	380	90	40
36	50	60	30	18
38	450	750	50	18
40	180	250	22	12
45	50	60	45	8
50	100	115	50	22
65	110	120	70	70
75	220	160	30	30
80	60	50	50	50
90	55	45	50	45
100	300	86	18	10
110	55	38	5	5
120	60	58	22	22
140	100	110	24	20
150	80	90	15	10

Figure 2-21. Radiated Interference in Microvolts Before Suppression Techniques and After

CONDUCTED INTERFERENCE IN MICROVOLTS BEFORE SUPPRESSION TECHNIQUES AND AFTER

Freq. in MC	Before AC Line		After AC Line		Before DC Line		After DC Line	
	Search	Hand Control	Search	Hand Control	Search	Hand Control	Search	Hand Control
0.16	40	40	25	25	55	55	2.5	1
0.18	60	60	50	50	70	70	3	3
0.24	55	55	45	45	140	140	1	1
0.28	57	50	45	40	60	58	1	1
0.32	90	92	45	45	75	70	1	1
0.6	28	28	20	18	9	6	1	1
1.0	28	28	7	7	9	9	1	1
1.4	6	6	6	5	10	5.5	1	1
1.8	4	4	3	2	14	25	1	1
2.8	11	11	1.5	1	18	5	1	1
3.8	20	18	1	1	11	8	1	1
5	40	37	2	2	30	22	2	2
7	20	18	4	4	80	60	1	1
9	23	25	2	2	27	28	2	2
11	65	60	2	2	75	72	2	2
14	100	75	4	3	150	120	20	18
16	65	55	6	5	90	60	5	4
18	50	50	7	3	70	55	2	1
20	20	18	2	1	30	30	3	2

Figure 2-22. Conducted Interference in Microvolts Before Suppression Techniques and After

equipment is designed that includes the use of thyratrons, consideration should be given to their interference characteristics. Maximum isolation, most efficient shielding, suppressor circuits, careful routing of leads, should all be included in original equipment design.

In another typical installation, "noise" clicks occurred in the intercommunication amplifier during radar scanning. The clicks were a result of the switching action of the relays controlling the antenna-drive motors. Shielding of the drive-motor circuit leads from the antenna to the radar control and from the antenna to the radar control and from the antenna to the operator's indicator was necessary. The general effect of the shielding was to reduce the maximum interference values about 50 to 90 percent, at which level the clicks became barely perceptible. The most efficient way would be to incorporate a resistance-capacitance circuit across the relay suppressing the clicks at the source, and eliminating the need for heavy and expensive shielding of external wiring.

2.1.1.12 Gun-Laying Radar System

A gun-laying radar system is a light-weight equipment used for tail protection of bombers. It operates in the microwave region and can perform the function of "searching" and "tracking" at the selection of the operator. It is similar to the fire control system, also discussed in this section, except that gun laying works with a turret while a fire control system operates in conjunction with fixed guns. Space consideration would not be as much of a factor in a bomber installation and would make possible grouping of units and elimination of interconnecting wiring. The principles of good design incorporated in this system would be applicable to the system discussed previously. The system's essentials are the same. Gun-laying radar sets include six major assemblies: The radar central (with a number of sub-assemblies), the RF unit, the antenna, the indicator, the radar junction box and the indicator junction box.

This system is unusual in that interference suppression was included in the original design as contrasted to the equipment discussed previously. On the first developmental model, without consideration for interference suppression, high levels of both conducted and radiated interference were measured. As discussed in other radar systems, the modulator was the main source of interference. In this installation, it generated high-voltage, short pulses supplied through a pulse cable to the RF unit. High levels of conducted noise were measured on the thyratron and rectifier transformer, ac power leads, the plate power lead, and the common (filament and plate) ac power leads.

A systematic study of the developmental model was conducted with the objective of building a second model which would include good design practices pertinent to interference suppression.

The modulator, a potential source of interference in any radar system, was completely redesigned with the U-shape layout changed to a rectangular design with the tubes and transformer in separate sealed compartments filled with insulating oil. This insulating-oil technique did not prove a practical one because of maintenance difficulties. In order to replace the oil after any repair work, it was necessary to prevent any air bubbles or moisture from getting into the compartment, which proved to be a difficult task. Solid copper shielding was used on all wiring coming from the transformer compartment. Ceramic capacitors (0.001 µf) were used from each terminal to ground. The filament and plate transformer were electrostatically shielded. Filters were installed in the filament supply, plate supply, 115-volt ac lines. Tin foil gaskets were placed under the covers in the transformer and tube compartments. Components and leads were rearranged within the modulator to isolate high pulse currents from power leads and power supply components. As a result of this redesign, the weight of the modulator was increased from 18 to 24 lbs; however, by the use of more compact design, the dimensions remained the same.

The pulse cable from the modulator to the RF unit is a frequent source of interference due to leakage. This leakage can be caused by inadequate shielding properties of the coaxial cable, or, more commonly, from poor shielding continuity of the plug-coaxial connection. In this equipment, special plugs were designed to minimize this leakage. It was found that by using a cable with its shield composed of four layers of braid, a reduction of the residual interference from peaks of 50 microvolts to less than 10 microvolts could be obtained.

The RF unit incorporated design features to minimize interference generation as follows: The pulse transformer and the magnetron were built into one unit. The magnetron section of this unit was pressure-sealed while the pulse transformer section was oil-filled and sealed. Special leakage-proof connectors were used for the triggering pulse. Filters were added to the blower motor, relay bus, and two in the secondary of the filament transformer.

In the synchronizer, all pulse cables, e.g., gating, timing, ranging, etc., over three inches long were shielded or rerouted to eliminate possible coupling into "clean" leads.

Two capacitor filters were added to the primary power sources in the radar central, one in the dc supply and one in the 400-cycle supply. A 7-ampere filter was added to the 27-volt lead in the radar junction box. One filter, in each of the three regulated power supply leads, 315, 150, 300 volts, is located in the radar central mounting rack.

In the radar junction box, the wires on the terminal board were rearranged to avoid pick-up from adjacent terminals. Harnessing was rearranged and some shielding provided for high-impedance lines. Resistance-capacitance filters were placed across microswitches (see Figure 2-23a) with shielded leads to the filters. Two of the switches were redesigned to provide low-impedance paths to ground from the switch terminals when they are not in active use.

Suppression measures were taken on all "noise" motors in the system. Constant armature-current dc motors were used for scanning and spinning; with the use of filters and the grounding of the negative lead, interference was reduced to well within specification limits. Three 0.25 µf filter capacitors were used in each pair of servo motors. See Figure 2-23b.

To suppress interference above 15 mc, two small coils of 1 to 2 µh were placed in series with the armature and two 0.001 µf capacitors were placed across the brushes. To save space, each coil was wound around a capacitor and the assembled filter placed within the motor casing. Blower motors were all filtered as close to the motor as possible with adequate grounding provided.

A filtering arrangement was considered for the motor-control relays. See Figure 2-23c. While this arrangement was satisfactory in its suppression characteristics, space limitations prevented its use. (See Chapter 1, Section 1.3.10 on Relays.)

All leads two inches long or longer showing 70 microvolts or more of conducted interference were shielded when possible. Pulse-carrying leads merited special attention to avoid affecting the pulse shape; low-capacity shielded leads were used when necessary. All servo motor leads were also shielded. Non-shielded high impedance leads were bypassed to ground by capacitors.

Careful consideration was given to adequate grounding throughout the system. It was found that anodizing, usually specified for aluminum parts, produces an insulating film which interferes with proper grounding techniques. The following protective coatings are recommended as substitutes for anodizing: chrom-aluminum, zincate process, caustic dip and zinc plate.

The total weight increase, emphasizing internal modifications, amounted to approximately 10 lbs. It is estimated that if these internal preventive measures had not been taken, the increase in weight for shielding of interconnecting wiring alone would have been in excess of 50 lbs.

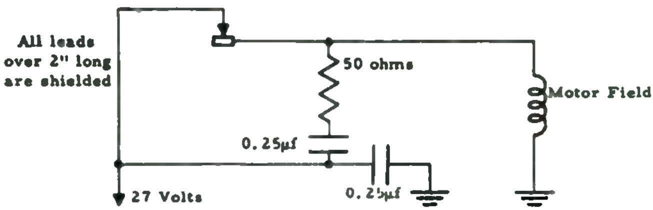


Figure 2-23a. Filtering Arrangement for Microswitches

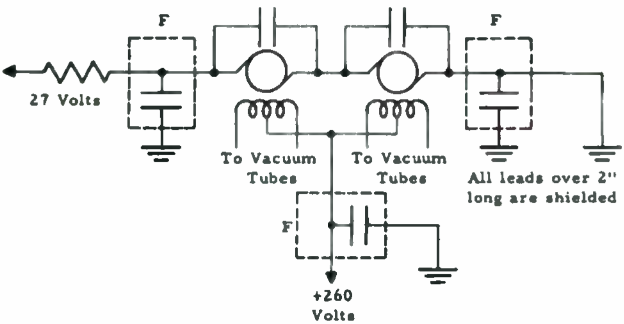


Figure 2-23b. Filter for Servo Motors Above 15 mc

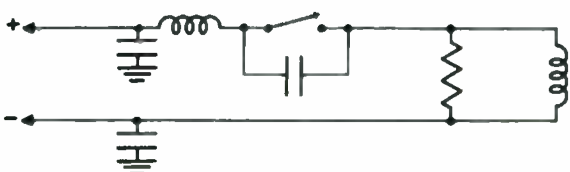


Figure 2-23c. Filter for Motor-Control Relay

Furthermore, the interference suppression measures incorporated in this system do not take advantage of all present day approved techniques; e.g., the miniaturization of suppression devices which would have eliminated some of the discussed difficulties encountered in the design.

2.1.2 ELECTRICAL SYSTEMS

Electrical systems are primarily generators of interference transients and fields as a result of the operation of the electrical components of each system. Without the application of any suppression techniques, interference would be conducted and inductively or capacitively coupled to electronic systems throughout the aircraft. In general, an attempt should be made to suppress the interference generated by components at the source. This involves improved mechanical design of the component, adequate shielding, filtering and bonding. Since it is not always feasible or practical to adequately filter all "noisy" components, some interference is conducted through the interconnecting wiring and couples into the electronic systems. Filtering or shielding at points, such as junction boxes, may be sufficient to eliminate these coupling paths. Particular attention should be paid to the routing of interconnecting wires of each system so that "clean" leads are not bundled with "noisy" leads or routed near strong pulse or harmonic generators. Analysis of these electrical systems generally serves to illustrate the type and location of interference-generating components to permit logical references to those paragraphs wherein the components are considered in detail for source suppression. Thus these electrical systems serve to incorporate information available from interference checks on military aircraft as to the types of equipment generally responsible for the production of interference. The following paragraphs describe typical installations to point out some of the general considerations.

2.1.2.1 AC and DC Power Systems

Modern military aircraft are controlled and operated largely through the use of the many electrical and electronic systems installed in the aircraft. A typical medium or heavy bomber would contain the following electronic systems: (1) HF Command, (2) Homing, (3) IFF, (4) Interphone, (5) Liaison, (6) Localizer and Glide Path, (7) Loran, (8) Shoran, (9) Marker Beacon, (10) Navigational Radar, (11) Search Radar, (12) Radio Altimeter, (13) Radio Compass, (14) VHF Command, and (15) Jamming Equipment; in addition, the following electrical systems would generally be installed: (1) Auto-Pilot, (2) Flight Instruments, (3) Cabin Heating and Ventilation, (4) Interior and Exterior Lights, (5) Fire Detector, (6) Fuel Booster Pumps, (7) Fuel Gage, (8) Tachometer, (9) Operational Indicators and Engine Instruments, (10) Propeller Feathering, Unfeathering, Reversing, and RPM Control, (11) Propeller De-Icer, (12) Trim Tab Controls, (13) Windshield Anti-Icer, (14) Wing and Tail Surface De-Icer, and (15) Aileron, Rudder and Elevator Controls. The electrical energy required to power these systems is supplied by common ac and dc power circuits. This provides a means for introducing interference signals from either power system into any electronic system as well as a common path between any of the electrical or electronic systems in the aircraft.

The aircraft ignition system is not interconnected with other electrical systems. An isolated ignition circuit complete with power source is provided for each engine. This precludes the possibility of conductive coupling of interfering signals out of the ignition system during operation. A conductive path for interfering signals into the dc power system exists when induction boosters are used for starting. This power is usually broken by the starting relay after the engines are started. However, there is a strong possibility for inductive or capacitive coupling and radiation from the system wiring. Interference problems peculiar to ignition systems are treated in Section 2.1.2.4 of this chapter.

A typical ignition system is shown in Figure 2-2 by dot-dash lines to illustrate the compactness of the ignition system and at the same time point out the possible paths for interference signals to gain access to the other aircraft systems.

In general, a typical ac power circuit would consist of the following components: (1) Bus Bars, (2) Voltage and Power Selector Switches, (3) Alternators, (4) AC Voltmeters, (5) Compensating Condenser, (6) Voltage-Adjusting Rheostats, (7) AC Voltage Regulators, (8) Circuit Breakers, (9) Inverters, (10) Junction Boxes, (11) Control Panels, (12) Various Plugs, Receptacles and Interconnecting Cables, and (13) Fluorescent Lights. A typical ac power circuit installation is illustrated in Figure 2-2. This diagram shows the approximate location of the various electronic systems and how these systems are interconnected through their power leads in the aircraft. The ac power cables are represented by broken lines (dashed).

A typical dc power circuit would consist of the following components: (1) DC Generators, (2) Batteries, (3) Bus Bars, (4) Generator Ammeters, (5) Battery Ammeters, (6) Voltmeters, (7) Voltage Selector Switch, (8) Circuit Breakers, (9) Internal and External Lights, (10) Voltage Regulators, (11) Relays, (12) Condenser, (13) Connector Boxes, (14) Control Panels, and (15) Various Plugs, Receptacles and Interconnecting

Cables. The majority of electrical actuators and servos are energized by the dc source. A number of such devices are shown in Figure 2-2. This diagram illustrates the relative locations of the various electrical systems and how these systems are conductively linked to the electronic systems through their respective power leads. Direct conductive links between the various dc electrical systems generally occur in the junction boxes, circuit breakers, and control panels. Since in many typical installations the aircraft generators are ac-dc, a possible conductive path also exists between the electrical and electronic systems. The ac-dc generators frequently utilize the same coil windings for each external system. A commutator is provided to rectify the induced alternating voltage for the dc system and slip rings are provided for the ac system. Some isolation effect is obtained through the use of separate power supplies for the various systems, i.e., dynamotors, inverters, etc. Bundling of the power cables with other system wiring introduces the possibility of inductive or capacitive coupling of interfering signals from one system to another. The dc power circuits are represented by solid lines.

The functioning of the power systems is not impaired by the presence of interfering signals along the power cables or in the system components. However, a great number of interference difficulties have been traced to the location, routing, shielding, etc. of the power cables. These troubles are primarily caused by other systems coupling interfering signals into the power lines and eventually introducing these signals into the various receivers in the aircraft. The generators, regulators, and relays used in the power systems are possible sources of interference. The design considerations applied to aircraft components for minimum generation of interference are discussed in Section 2 of this book.

2.1.2.2 Propeller Systems

Propeller systems are potential and actual sources of radio interference because they contain electrical and electronic equipment and circuits. This is true regardless of whether the propeller is of the electric or the hydraulic type since electrical means are employed to accomplish the following diverse functions in present-day designs of propeller systems:

a. Propeller Control

 (1) Speed control

  (a) Reciprocating engines

  (b) Turbo-prop engines

(2) Synchronization

(a) Speed synchronization

(b) Phase synchronization

(3) "Beta" operation (primarily in turbo-prop applications)*

(4) Reverse-pitch operation (primarily in reciprocating engine applications)

(5) Feathering

(a) Manual initiated

(b) Automatic type

(6) Remote control of speed settings of governors and synchronizers.

b. Propeller De-icing

c. Propeller blade-angle (beta-angle) indication.

d. Electrical lamp indication of propeller operations.

Basically the problem of radio interference prevention on such equipment is the same as for other aircraft equipment and the same principles and practices might be expected to apply. Such equipment, incidental to its basic operation, generates rapidly changing currents and voltages. A certain amount of coupling exists in the aircraft due to the proximity of equipment in the necessarily close confines. This coupling comes from two principal sources:

a. Use of common power system and battery for both the propeller and the radio and electronic equipment, and

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*In "beta" operation, the pilot has direct control, through manual operation of the lower lever, of the propeller blade angle. This type of operation is used for taxiing and other ground handling as well as reverse-pitch operation for ground braking.

b. Mutual inductance between propeller wiring and antenna lead-ins of radio equipment.

Two factors seem to distinguish propeller systems from other sources of radio interference in aircraft:

a. The current values used are very high, particularly on 28-volt dc systems, and normal operation involves continual switching on and off of such currents. From a weight penalty standpoint, the margin of error in the direction of supplying a generous amount of suppression filtering is narrow.

b. Many important sources of interference in propeller systems produce interference of the click- or pulse-type of low and very low repetition rate. This situation is illustrated in Figure 2-24. The practical result is to create a difficult problem in measurement technique.

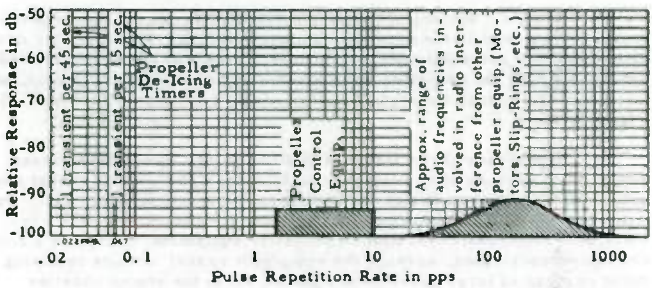


Figure 2-24. Approximate Repetition Rate of Important Sources of Radio Interference of the Click or Pulse Type

The present specified measuring set-up prescribes two standard service receivers. It also involves the use of a cathode-ray oscilloscope as an indicating instrument rather than a D'Arsonval meter used in conventional noise meters. The reason for this approach lies in the wide range of so-called repetition rates which are involved in the types of interference that are produced by propeller circuits as shown in Figure 2-24. Particular note should be paid to the low values of repetition rates involved in the pulse-type "noise" produced by many propeller control systems and the very low values for certain types of de-icing timers. This interference is primarily a series of clicks or pops with "clear" places between.

Sources of interference in propeller systems may be found in (a) pitch change motors, (b) pitch change solenoids, (c) slip rings, (d) governors, (e) synchronizers, (f) de-icing timers, (g) de-icing relays, (h) inverters, and (i) the various switches and contactors. Many of these items can readily be seen illustrated in Figure 2-25. Such equipment, incidental to its basic operation, generates rapidly changing currents and voltages.

A certain amount of coupling exists in aircraft due to the proximity of equipment in the necessarily close confines. This coupling can come from two principal sources: (a) use of common power system and battery for both the propeller and the radio and electronic equipment, and (b) mutual inductance between propeller wiring and antenna lead-ins. See Figure 2-26.

Experience shows that it is difficult to add radio-interference suppression devices after propeller designs are complete. For that reason, it is far better to design the equipment correctly to begin with. Practical methods of prevention include: (a) use of non-electrical and non-electronic methods of operation in propeller equipment, (b) where electrical equipment is used, arrange the equipment so that circuits involving rapid changes of large currents are placed out in the engine nacelles rather than in the fuselage and particularly not near the radio equipment, (c) in cases where propeller equipment is to be powered from a common electrical system with the radio gear, provide effective filtering in the power leads, and (d) in cases where propeller equipment containing interference sources is to be installed near radio equipment, provide adequate shielding containers for the propeller equipment and filter all leads from the equipment.

The filtering of leads from interference sources in propeller equipment requires the use of the following components: (a) capacitors, (b) inductors, and (c) transient suppressors of the following types: (1) dry-plate rectifiers (magnesium copper sulfide), (2) point rectifiers (germanium), (3) non-linear resistors of the "Globar" or "Thyrite" type,

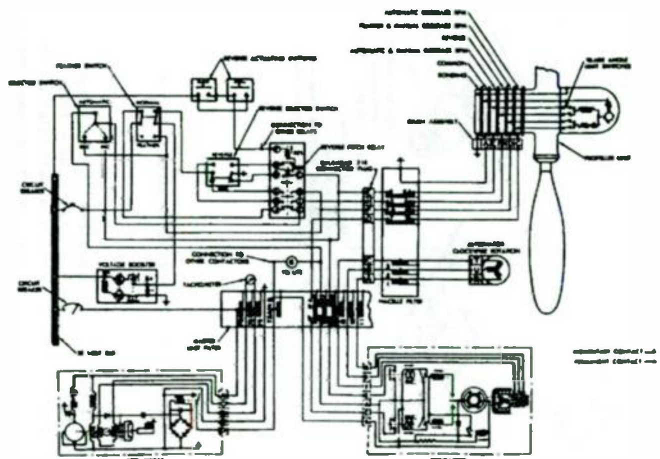


Figure 2-25. Schematic Wiring Diagram of a Representative Propeller System Installed in a Typical Airplane

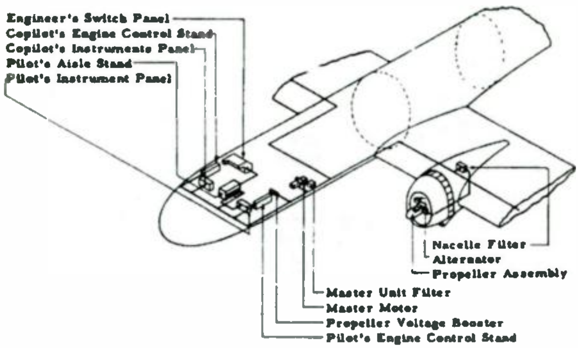


Figure 2-26. Propeller Control Diagram

(4) gaseous discharge tubes, (5) vacuum tubes, and (6) resistor-capacitor networks. Capacitors find universal application across lines from interference sources as suppressors; they are particularly effective on dc motors and generators. Addition of an inductor between the capacitor and relay contact makes an effective combination in suppressing relay and contactor interference. Additional inductors in series and capacitors in shunt can be used in the form of radio-interference filters for further suppression. For filtering to be adequate it must be supplemented by the use of shielding containers. Proper shielding involves the use of metallic (not necessarily magnetic) boxes and containers that are free of openings and non-conducting joints. Size and shape should be as required for

the propeller equipment. The design techniques for shields and joints are discussed in Chapter 3 in this volume.

The designer must be guided by experience in the amount, type, and extent of filtering and shielding to use. The final criterion is that the propeller equipment cause no interference in the radio and electronic gear in the actual aircraft installation.

There are, however, obvious practical disadvantages to waiting until equipment is installed in the aircraft before being able to determine radio-interference performance. The answer to this difficulty has been to set up test specifications setting forth the procedures, measuring equipment and maximum allowable limits to be used in determining the radio-interference performance of electrical equipment for aircraft. Propeller equipment has received its share of attention in this effort.

The thought in setting up a radio-interference specification for propeller equipment has been to provide the manufacturer with a guide as to how good to make his equipment. The expectation is not that equipment meeting such a test will never cause interference when installed; the range of variables involved is too great for that. Rather, equipment, when passed, can be expected to cause interference in less than about 10 percent of actual cases. These latter problems could then be handled individually. Working that close to the margin requires good test equipment and procedures together with maximum allowable limits founded on a sound basis.

In this connection, it may be helpful to review the general basis upon which future plans are founded. This envisages that a radio-interference survey will be made as part of the type test of each new propeller design. This approach would be analogous to the use of vibratory-stress surveys on new designs. Indeed, it is believed that the radio-interference survey can be conducted just before or just after the vibration study. The type test approach was chosen rather than a production inspection method because of conditions which exist in the propeller industry. Production methods in the industry are such that while operational checks are continually made of components of propeller systems, very seldom, if ever, are complete setups made during production of systems to check operation prior to installation on the airplane. In many cases new designs of propeller systems consist of only minor changes from previous designs or consist of changes of a mechanical or similar nature not usually affecting the radio-interference performance of the system. It is believed that if a system has been once tested for radio interference and found satisfactory, approval can be granted on various modifications of the system without costly retesting of each modification. A similar procedure is currently used in handling vibratory stress approval on propellers differing in minor detail from a previously tested and approved design. In this connection, attention of contractors is directed toward Specification MIL-P-5449, Amendment 1, requiring the furnishing with the propeller installation model specification of detailed data on the radio-interference suppression provisions.

Two alternative forms for the type test are prescribed: (a) engine tests, and (b) bench tests.

The engine test procedure is to be used for testing all propeller systems and equipment the radio interference from which can reasonably be expected to be influenced by engine rotation, vibration and propeller power loading. Such effects can be expected in systems involving such items as: (a) electrical current-carrying slip rings in the hub or on the blades, (b) engine-mounted propeller governors, and (c) pitch-changing mechanisms of an electrical nature.

The bench test procedure is to be used for testing all propeller systems and equipment the ratio interference from which can be expected to be unchanged in magnitude by the presence or absence of engine vibration or propeller power loading. Typical examples of such equipment are: (a) blade-angle indicators involving use of ac selsyns powered by 400 cps inverters, and (b) de-icing system using hub-mounted generators together with a timer.

Manufacturers of propellers and propeller components are cautioned that it is the practice of the services to follow up all complaints of interference and to request modifications, where required, of propeller equipment. Responsibility for performing tests to check compliance of such equipment with applicable radio-interference specification lies with the group having jurisdiction over the equipment. All questions of compliance, test methods, and equipment design should be referred to the group, laboratory, etc., having jurisdiction over the equipment.

2.1.2.3 Automatic Pilot

An Automatic Pilot is generally an electromechanical-electronic system designed to provide automatic control of airplane surfaces to maintain a predetermined course of flight. The plane is stabilized on longitudinal, lateral, and vertical axes with minimum angular displacement and accelerations in any axis. Manual control of the Automatic Pilot to accomplish dives, climbs, and coordinated banking turns through wide limits is provided.

The components of a typical system include: (1) Flight Control Rate Gyros, (2) Directional Panel, (3) Turn and Pitch Controller, (4) Turn Control Transfer Switch, (5) Three Servo Motors, (6) Multiple Channel Amplifier, (7) Calibrator Unit, (8) Mounting Chassis, (9) Amplifier, (10) Turn Controller, (11) Formation Sticks, (12) Interconnecting Cabling, (13) Flight Control Vertical Gyro, and (14) Gyro Directional Stabilizer.

In a typical installation, in a bombardment type aircraft as shown in block form in Figure 2-27, the control mechanisms which include Turn Control Transfer Switch, the Turn and Pitch Controller, and the Formation Sticks would be readily accessible to both pilot and co-pilot. The Directional Panel is installed in the bombardier's compartment to be used in conjunction with the bomb-sight for flight control during bombing runs. The chassis equipment assembly, which includes vertical and rate gyros, calibrator, and amplifier mounted in a tray, is installed for ease of maintenance. The three servo motors are installed close to the control surfaces and are connected to the amplifier with lengths of electrical cable.

This Automatic Pilot operates on the principle of the balanced bridge. A balanced bridge is provided for each of the three axes: aileron, rudder, and elevator. When any one of these bridges is unbalanced by a signal from any one of the gyros, formation sticks, remote control, or turn and pitch controller, a relay switch in the amplifier activates the appropriate servo to position the control surface to a predetermined or new setting. The process is one of continuous correction with subsequent continuous operation of the relays signalling the servos. The rate of the corrective response is regulated by the calibrator.

The Automatic Pilot, by the nature of the components necessary for its operation, can be a prolific source of radio interference to other systems in an aircraft. The continuous operation of relays and small motors generates interference over a wide range of frequencies, and requires consideration in the original design of such equipment. In the original design of the Automatic Pilot under discussion, interference suppression was not stressed. Examples of poor interference suppression design include:

a. Mating surfaces were anodized preventing good ground and bond connections.

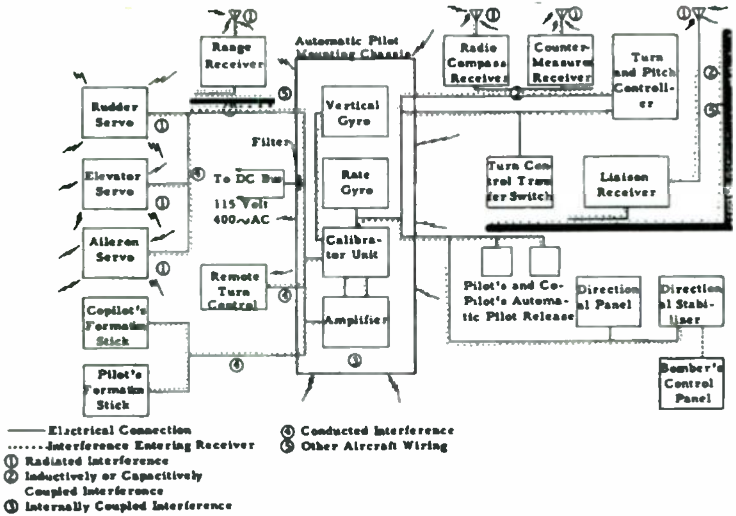


Figure 2-27. Paths of Interference Signals in a Typical Automatic Pilot System

b. Bolts, nuts and washers were treated with a corrosion resistant compound which was non-conductive and contributed high impedance paths to ground.

c. Poor case and internal shielding permitted radiation and coupling of interference into other airplane wiring.

d. Filters were used with long unshielded leads thus permitting radiation and coupling of noise into other wiring.

e. "Noisy" leads were bundled together with "clean" leads resulting in interference appearing on most external wiring and making it difficult to apply suppressive measures. Leads coming out of filters were exposed to interference fields which nullified the effect of the filter.

The principle components of the system from which the interference would originate includes the Calibrator, Amplifier, Directional Panel, and Turn and Pitch and Controller.

The calibrator with its numerous relays, e.g., aileron engage, rudder engage, elevator engage, auto recovery, etc., proved to be a prolific source of interference. The action of the relays caused currents with steep wave fronts which resulted in radiation and coupling into other airplane wiring. In the original design, suppression networks were not included in the relay circuits. External filters were inadequately grounded and improperly located to suppress interference adequately. Shielding cups prevented radiation from above; however, interference was free to radiate from beneath the relay. The calibrator case had poor shielding continuity, bonding, and grounding, and permitted radiation to emanate from the case. It was necessary to modify the calibrator internally to prevent it from interfering with other aircraft equipment.

In the amplifier the six servo relay switches were the principle sources of radio interference. It was particularly serious in this component because of the long lengths of cabling to the three servo motors which could cause interference in a considerable amount of airplane wiring in the original design. The only recognition of the interference problem caused by relays and their associated wiring took the form of inadequately shielding each relay internally. Interference radiated from the poorly grounded shield, and transients in the servo wiring affected other airplane wiring. Poor shielding continuity and grounding of the amplifier case also permitted radiated interference to affect other circuits.

In the Directional Panel, interference from the dc Arm Lock Motor Commutator appeared on external interconnecting wiring and was free to radiate or couple into other airplane wiring. In the original design, the motor was shielded but the leads were not shielded or filtered and conducted interference appeared on external wiring.

The Turn and Pitch Controller has three direct current motors used for rudder, aileron and elevator centering. In the original installation, these motors were individually shielded but no internal suppression measures were taken to prevent the commutator interference from being conducted through the shield by the motor leads. These unshielded leads were bundled in with other wiring and decreased the effectiveness of external filters of dc power leads. Coated surfaces used in mounting the motors did not provide adequate bonding and reduced the shielding effectiveness of the cases.

As a result of lack of adequate suppressive measures used in the design stages, installation difficulties were encountered with the operation of the Automatic Pilot in a bomber installation because interference was found on all bands of the Liaison and one of the Countermeasures Receivers. The relays caused a rapid clicking noise in both receivers with peaks measured in the range from 3 to 21 megacycles. Since an internal modification of the Automatic Pilot System was the only practicable way to eliminate the interference, no installation "fix" was attempted. In a later model of the same type of aircraft, the interference from the relays was reduced to a low level by modifications in the Automatic Pilot; however, commutator noise, from the Gyro Torque motor in the Directional Stabilizer and the Arm Lock Motor located in the Directional Stabilizer appeared in the same receivers. A "fix" was attempted by mounting a filter and a 4 microfarad capacitor near the directional stabilizer. The filter was connected in series with the 28 volt dc power lead to the gyro torque motor. The capacitor was connected from the torque motor side of the filter to ground. This "fix" proved to be successful; however, a similar configuration tried on the Arm Lock Motor did not remove the interference from the Countermeasure Receiver. The same receivers being affected in both examples of Automatic Pilot interference suggests that improvement may have been possible in the installation mock-up to eliminate the coupling paths for the interference to the receivers. However, this does not remove the premise that inadequate source suppression measures were taken in the original design of the Automatic Pilot.

In another bomber installation, the Automatic Pilot interfered with the Radio Compass, Range and Liaison receivers. The interference was of three types:

a. The continual or repeated pecking noise from the contacts of the six servo relays. This same type of interference was evident also when the Automatic Pilot was disengaged, but the centering motors in the control panel were being actuated. A "fix" was attempted by using a filter on the load side of each of six servo relays but this only reduced the noise and did not eliminate it. Complete suppression would necessitate modification of the Automatic Pilot.

b. "Clicks" and "pops" appeared on all three of the affected receivers when the Turn Control was moved in and out of the detent position. Part of the interference was due to the make and break action of the switch contacts at the detent position during the make and break of the Arm Lock Relay coil circuit. This interference was particularly prominent on all except the low frequency band of the Liaison receiver with peaks in the range of 6 to 9 megacycles. A "fix" was attempted by using a 0.5 microfarad metallized paper condenser across the switch contacts. This action effectively suppressed this part of the interference. The remainder of the interference was caused by the Arm Lock Relay Contacts actuating the Arm Lock Motor. These clicks and pops were the sharpest and most annoying of the Automatic Pilot interference and most serious on the Radio Compass and Range Receivers. To suppress this part of the interference, a filter was placed on the load side of each of the three contacts that actuate the Arm Lock Motor. A filter was also placed in the 28 volt power lead to the relay.

c. The occasional or intermittent "clicks" and "pops" due to the Engaging, Anti-engaging, Transfer, and Rudder Engage Relays in the Calibrator Unit appeared on all three receivers. Suppression measures were not taken on the Automatic Pilot but an attempt was made to eliminate coupling paths and increase the rejection characteristics of the receivers. The radio compass antenna lead-in was double shielded and a 0.5 microfarad condenser, 400 volts, dc rating was connected on the 400 cycle power lead of the Radio Compass to ground. Bonding of the Liaison and Range receivers was improved and leads rerouted when possible. This attenuated the interference, but did not eliminate it.

External filtering in this installation was not satisfactory because of the lack of circuit isolation. Internal cabling was laced in large bundles with interference coupling throughout the equipment. To be effective, filters would have to be applied virtually to each individual wire of the many wires coming out of the units.

Extensive modification of the Automatic Pilot was necessary in order to insure interference-free installations in aircraft. The following interference-free design techniques were applied:

a. Good bonding and grounding surfaces were insured by changing the anodizing protective process to cadmium plate over copper plate over zinc plate on the aluminum.

b. Non-conductive protective coating was removed from bolts and nuts to lower the impedance paths to ground as discussed in Chapter 3, Section 1.3.7 of this book.

c. Both internal and case shielding were improved by new design as well as by the previous improvement of contact surfaces as discussed in Chapter 3, Section 1.3.

d. Filter ground leads were shortened and the filters were relocated close to the "noise" source with a shielded lead into the filter as discussed in Chapter 4.

e. The internal circuitry was improved which resulted in better isolation of interference sources and interference conducting leads.

f. Calibrator relay contact surfaces were improved and the leads into the filters were shielded thus preventing the relay interference from coupling into other wiring.

g. Case shielding continuity, bonding, and grounds were improved to prevent case radiation.

h. Suppressor networks were installed inside the amplifier and connected to the servo relays. These networks were connected to the signal leads of each of six relays as close to the relay shielded case as possible and a capacitor was connected across the relay contact points to reduce the effect of transients as shown schematically in Figure 2-28. The relay shield was improved as well as overall case design.

i. The Arm Lock Motor in the Directional Panel was internally filtered with a shielded lead into the filter. This effectively eliminated commutator interference by confining it to its internal shield.

j. In the Turn Pitch Controller, filters for the three dc motors were more advantageously located, provided with better contacts, and connected to ground by shorter leads. Leads which conducted interference were shielded prior to entering the filters. Rerouting was accomplished to protect the filtered lead from exposure to an interference field. Shielding effectiveness of the case was improved by changing the anodizing to a plating process.

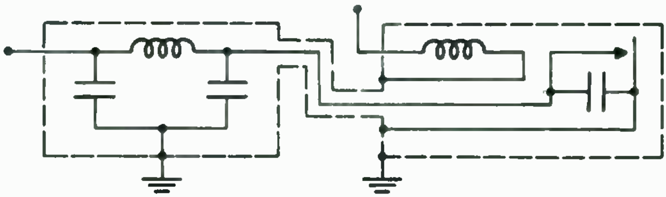


Figure 2-28. Schematic Diagram of Suppressor Network

Up to this point, the interference generating characteristics of the Automatic Pilot have been stressed; now the Automatic Pilot's susceptibility to interference from other systems in the aircraft will be considered.

Interference which entered the Automatic Pilot System in one particular installation caused erratic operation of the control surfaces, known as "jitter". Investigation showed that the signal leads to the servo units picked up interference and conducted it to sensitive circuits in the amplifier at which point the circuits were unable to distinguish the interference signals from the desired ones. By installing a filter in each of the six signal leads, as close as possible to the amplifier, this interference was effectively eliminated. Routing the servo motor signal leads away from interference fields may possibly eliminate the need for these filters; however, owing to the long lengths of cable necessary, filters may be the simplest way to eliminate this as an installation problem. Interference from other systems has not been a major problem in the operation of the Automatic Pilot in this installation.

However, interference from other systems in the aircraft has been a major problem in another type of Automatic Pilot System and a brief discussion of this is included because it still constitutes a problem in recent Automatic Pilot installations.

The principle of operation of the two Automatic Pilots is basically the same. However, instead of a potentiometer type of signal pick-up, rate gyros for the rate signals, and relay pulsing type servo, this equipment uses a synchro type signal pick-up, resistance-capacitance type of rate circuit, and an amplidyne servo system. This system is sensitive to interference on both the ac and dc power systems. In a typical medium bomber installation, oscillation or "jitter" of the control surfaces was caused by commutator ripple in the dc power supply and low amplitude modulation of the 400 cycle wave in the ac supply.

Two control surface hydraulic booster pump motors produced commutator ripple on the dc system. When these motors operate simultaneously their ripple voltages combine to form a beat effect which caused trouble. This interference eventually coupled into the sensitive circuits of the Automatic Pilot. With low batteries, or with batteries disconnected from the system, the "jitter" is exaggerated but when the batteries are in good condition they tend to minimize the effect of the commutator ripple. The "fix" recommended for this installation was to install a 500 microfarad, or larger, capacitor from the dc bus to ground near the booster pump motor terminals. As much isolation as was possible was provided between sensitive Automatic Pilot wiring and dc power wiring.

In planning the design of future systems and installations of this kind, early consideration should be given to the isolation of sensitive circuits and wiring from the effects of ripple on the dc power supply because this has been an important source of interference to electronic systems in various aircraft. Ideally, the interference characteristics of motors, such as the booster pump motor, should be suppressed in the design stages; however, the rejection characteristics of a receiver to this type of interference should be equally emphasized in its design stages.

The inverter, in this installation, was putting out a low frequency, low amplitude modulation of the 400 cycle wave in the ac power supply. This modulation was aggravated, by the ac ignition system used, to the point where the Automatic Pilot could not distinguish it from the normal sensing signals. The recommended corrective action in this instance was to provide separate sources of ac supply for the ignition system and the Automatic Pilot. To illustrate the importance of a "clean" ac power supply for this equipment, an Air Force specification required that an ac source be provided with no modulation between frequency limits 0.1 to 60 cycles per second and a total harmonic content of not more than 5% of the fundamental. The original design of this equipment should include considerations that make it compatible with the operation of the other systems in the aircraft. With the increasing use of ac power supply in aircraft, this Automatic Pilot system should be able to operate as satisfactorily as other equipment on the same ac supply.

In the design stages of any equipment, recognized interference sources should be properly isolated by means of shielding, filtering, bonding, or other effective means as necessary considering weight, space and materials as discussed in Chapter 1, Section 1.1. Study of the original design of this Automatic Pilot equipment revealed that the main reliance was on shielding. The techniques used, however, were not effective in this case. While it is recognized that shielding is frequently necessary to suppress interference from ignition systems, modulators, etc., it is one of the more expensive methods and may result in greater weight than if a combination of techniques is used. For example, a simple R-C network, properly designed and connected across the contacts of relays may have reduced the transients sufficiently to eliminate the need for shielding of each individual relay. (See Chapter 1, Section 1.3.10.) Proper routing of leads that may conduct interference would still further reduce the need for this extensive shielding. Small motors with interference suppressed by means of techniques discussed in Chapter 5 and proper isolation would obviate the need for extra heavy shielding.

To overcome the lack of foresight in the original design of this equipment with regard to interference-free operation, good engineering practice was employed in locating interference suppression devices as close to the source of the interference as was possible. The "fixes" used showed a good understanding of and an application of the principles which must be used in combating interference. The result was a considerable improvement in the overall design which reduced the generation, conduction, and radiation of the interference.

2.1.2.4 Spark-Type Ignition System

Ignition systems used in present day reciprocating aircraft engines are designed to produce an electric-spark to ignite the compressed fuel inside the cylinder. The ignition system must deliver a spark at the proper instant in each cylinder to give smooth operation. Reliability and efficiency are prime design considerations and extreme environmental conditions must be provided for in the design of aircraft ignition systems. The system must operate at temperatures of -40° to -60° F. as a lower limit. At the other extreme, however, the magneto is exposed to 250° F., the cable to 350° F., and the plugs to 500° F. Since the system is mounted on the engine, all parts are subject to severe vibration at frequencies below 200 cps, with forces as high as 75 g's.

The reduction of dielectric strength and ionization potential which accompanies increase in altitude is very damaging. Corona discharge forms ozone and nitric oxides. The oxides combine with moisture to make nitric acid, which corrodes all metal parts. Both corona leakage and weakened dielectric combat attempts to increase ignition voltage and thereby gain more effective ignition. To avoid the detrimental effects of altitude, the ignition system has been pressurized in some designs and in others it has been filled with a sealing compound. These aging, breaking, or other damaging effects due to environmental conditions generally tend to increase the interference problem created by the ignition system.

The aircraft ignition system includes the following component parts: (1) magnetos and distributors; in some systems these are integral (two per engine), (2) harness assembly and spark plug leads, (3) external transformer coils, used only with low tension ignition, (4) spark plugs (two per cylinder), (5) starting booster coil or induction vibrator, (6) ignition and magneto grounding switch, and (7) flexible metal conduit used to cover the wiring between the engine ignition system, the ignition switch, and the starting assemblies. A front view of a typical ignition system installed on a radial aircraft engine is shown in Figure 2-29.

While radio interference signals entering the ignition system will not adversely affect the system itself, the components of the ignition system are extremely strong interference sources and could cause serious interference in the aircraft electronic systems if proper suppression techniques are not observed.

In general, spark-type ignition systems, if allowed to radiate, can be listed high among the major sources of radio interference. This is true because of the deliberate generation of high frequency transient currents in the ignition circuit as a necessary function in aircraft engine operation.

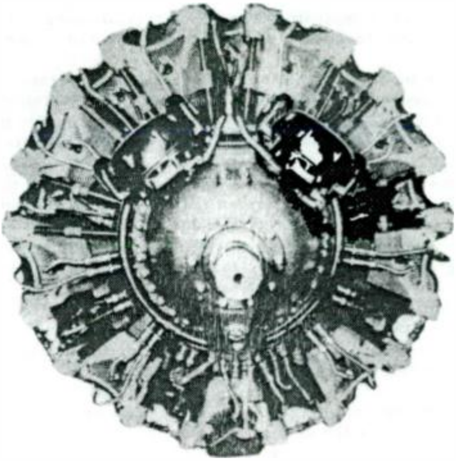


Figure 2-29. Typical Spark-Type Aircraft Ignition System

Whenever an electrical circuit is opened or closed, such as occurs with a distributor, there is a transient or variable-current state immediately following the making or breaking of the circuit, during which current is either rising or falling. These "steep wavefront transients" are not of the simple sinusoidal type, but have been shown to consist of a fundamental wave of relatively low frequency, upon which are superimposed a multiplicity of higher frequency components or transients.

The components of an ignition system must, therefore, be regarded as generators of periodic short duration waves containing not only low frequency components, but high frequency components extending across the useful radio-frequency spectrum.

These high frequency oscillations, if not suppressed through shielding, are radiated by the various ignition system components which act as antennas. The radiation results in uncontrolled frequency waves in the radio frequency spectrum above 10 mc and is especially high in annoyance value depending upon the characteristics of the transient and the sensitivity, frequency response, and location of the receiver.

Although, as pointed out in earlier paragraphs, noise suppression may generally be accomplished by filtering or shielding at the source or the receiver, or both; such is not generally the case with respect to interfering waves generated by an engine ignition system. With the various aircraft antennas subject to direct uncontrolled frequency radiations of the engine ignition system, any effort to filter or shield at the radio receiver will result not only in the suppression of interfering waves but also of the desired intelligence signal radiated by a controlled frequency radio transmitting station. The possibility of suppression by filtering at the source is equally remote. Any filter capable of eliminating all of the high frequency components of an ignition transient would also be capable of removing a large part of the ignition wave itself, thereby impairing the efficiency of the entire engine ignition system. Some recent success, however, has been found through use of resistor-type spark plugs.

It follows, therefore, that complete shielding of the aircraft ignition system is the only practicable method to prevent radio frequency energy of appreciable magnitude from radiating into space and resulting in aircraft radio interference.

The existence of this condition is fairly well accepted, although not necessarily appreciated by the industry. As a result radio interference emanating from the aircraft ignition system still is troublesome. The detectable interference now generally results from (a) poor shielding joints, (b) poor flexible shielding conduit, and (c) insufficient shield wall thickness and/or improper material.

In the transmission of high frequency transient currents in a coaxial system, such as is the case in the modern ignition system, the current tends to follow the path of least impedance. As a result, the current will flow on the outside surface of an ignition cable conductor and on the inside surface of the shielding. Any poor junction in the shielding will provide an opening to the shielding exterior thereby permitting the current to flow on the outside surface of the shielding where it can radiate. No bonds are required with a perfect shield.

The problem of ignition noise largely reduces itself to one of joint or flange design and frequent bonding of the shielding to the airframe. Major emphasis should be placed upon proper initial joint design for little can be done in the field to make satisfactory corrections of a permanent nature. Obviously, the number of joints should be kept to a minimum. The most satisfactory joining of mating surfaces is by welding, brazing, or soldering. Even a good solder joint will exhibit an appreciable contact resistance and is never as good a conductor as a brazed or welded joint. However, welding and brazing cannot generally be used because of the difficulties of field service and because the heat required causes damage to insulated conductors.

Radio interference signals may enter or leave the ignition systems over any of the paths shown in Figure 2-30.

Radio interference caused by engine ignition systems can be summed up as follows:

a. Interference due to radiation leaking past poorly mated joints located in any portion of the ignition system.

b. Interference from damaged portions of the ignition harness assembly, and

c. Interference from loose nuts or other fastening devices which are part of the ignition assembly.

The magneto is a form of high-frequency generator and consequently all joints and covers in the magneto are potential interference sources. Figure 2-31 illustrates several types of joints ordinarily found to be a source of interference in magnetos. In order to keep the radio interference energy within the magneto shield it is necessary to make all joints low impedance paths, as indicated above, and discussed in Chapter 3, Section 1.3.

The distributor comprises a distributor rotor and terminals connecting individual spark plugs. It may be considered as a rotating mechanical switch which transfers the electrical energy from the high tension coil of the magneto to the spark plugs. The distributor finger or rotor does not touch the terminals or electrodes but passes over them with close clearance. Since the high voltage produced by the magneto must jump the air gap between the rotor finger and the terminal of the distributor in addition to the spark plug gap, the distributor can be

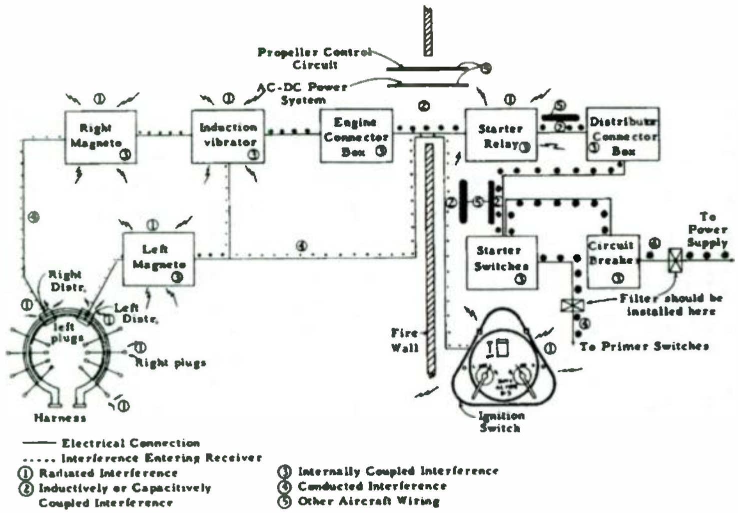


Figure 2-30. Paths of Interference Signals in a Typical Aircraft Ignition System

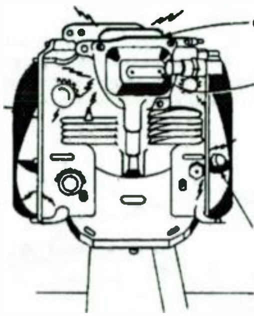


Figure 2-31. Typical Sources of Interference in Magnetos

compared to a spark transmitter. Volume I, Chapter 1, Section 4 and Chapter 1, Section 1.3.8.1 of this Volume give a detailed discussion of spark gap interference. Again it is a problem of keeping the joints electrically tight at high frequencies in much the same manner as described for the magneto. Types of troublesome distributor joints are shown in Figure 2-32a.

The harness assembly is again a problem of keeping radio interference energy inside the harness or shield. If any portion of the harness assembly is cracked or broken and any connections not properly tightened interference signals will radiate from the harness assembly. Typical potential sources of noise leakage from the harness section are shown in Figure 2-32b.

Spark plugs are normally well shielded and seldom present an interference problem, unless they are damaged or not operating properly.

It has been demonstrated by design experience that proper radio interference control of ignition systems is obtainable by using carefully designed shielding assemblies and interference filters. A well-shielded ignition system is well protected against fire hazard, shock hazard, heat, lightning, weather, and vibration. However, shielding adds to the cost and weight of an ignition system and increases spark plug electrode erosion, thus requiring careful application of suppression techniques in the original design. Shielding and filtering techniques are treated in Chapter 3, Section 1.3 and Chapter 4.

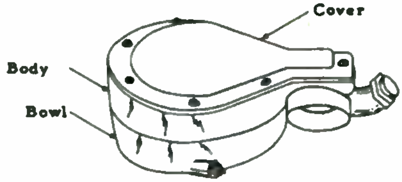


Figure 2-32a. Typical Sources of Interference in Distributors

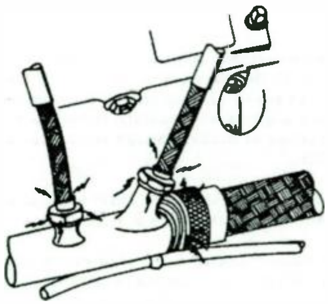


Figure 2-32b. Interference Leaks in Harness Assembly

3. RADIO FREQUENCY INTERFERENCE FROM PRECIPITATION STATIC

Increased speed and higher altitude flying, the result of improved design and the availability of more powerful engines, both dictated by military necessity, brought ever-increasing losses due to the precipitation-static type of radio interference. The Military Services, aircraft manufacturers, and commercial flight operators have made attempts to remove this hazard to aerial navigation. From numerous investigations, in flight and in the laboratory, there resulted a series of programs for "cleaning-up" airplane structure and design. These included the elimination of sharp-edged or pointed protuberances, the effect of various coatings on the metal surface, and the bonding of separated parts of the aircraft structure. In addition, much research and development has been done to make available dielectrically insulated antenna wire and fittings as well as devices for discharging static charges harmlessly from the aircraft while in flight.

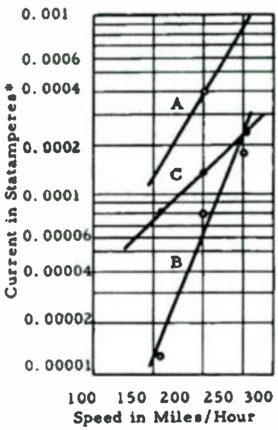
3.1 OPERATION OF AIRCRAFT

Pilots have frequently observed that the severity of precipitation interference is a function of the speed. Using as a working equation, charging current equals a coefficient times the speed to the n^{th} power, it was found that the average value of n is 3. Thus the radio interference goes up approximately as the cube of the speed. Reduction of airspeed from 400 miles per hour to 200 miles per hour would, therefore, reduce the charging current, and hence the corona discharge, by a factor of eight as shown in Figure 2-33.

The variation of the coefficient with temperature is not linear. It is a maximum between -7° C and -9° C, while the exponent, n, reaches its peak value at about -7° C, as is shown by the curve in Figure 2-34. The severity of precipitation interference may therefore be reduced both by decreasing speed and by descending to lower levels where the temperature is higher. The second alternative is, of course, always accompanied by the dangers caused by icing.

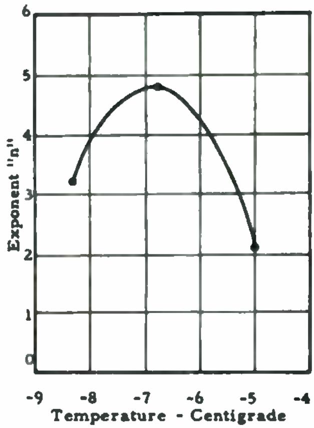
3.2 EFFECTS OF COATINGS ON METAL SURFACES

Both as a means to protect a shiny aluminum surface against the elements and to reduce visibility to anti-aircraft and other enemy fire, airplanes have customarily been covered with various varnishes and camouflage paints. It has been shown that the charging rate of painted surfaces is several times as great as that for clean aluminum surfaces. Surfaces of clean aluminum and those coated with aircraft



Curve A - Temp. -83 C Slope n=3.3
Curve B - Temp. -67 C Slope n=4.7
Curve C - Temp. -50 C Slope n=2.1

Figure 2-33. Charging Current as a Function of Speed & Temperature



Charging Current = Coefficient (Speed)^n. (Nose-piece coated with aluminum.)

Figure 2-34. Temperature Effect on Exponent in the Equation

wax produce negative charging, but coatings of TiO_2 and colloidal silica give positive charges when bombarded with driving snow. The results of these and other tests show that the paints in common use on aircraft surfaces consistently give negative charges to the airplane surface under precipitation charging conditions.

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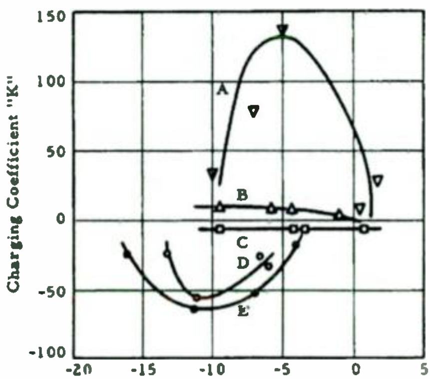
*One Statampere (electrostatic cgs units) is equivalent to 3.33560 x 10^{10} amperes (absolute).

The fact that coatings with positive charging coefficients are available has led to attempts to produce a chargeless airplane by using equivalent amounts of positive and negative materials on different sections of the plane. Experience proved, however, that contamination of positive surfaces, caused by handling of the plane by the ground crew, resulted in a reversal of sign for the charging coefficient. The best possible surface for minimum precipitation static is clean, smooth aluminum. The charging coefficient, K, for various coatings as a function of temperature is shown in Figure 2-35. The coefficient, K, is defined by the equation, K = I/W, where I is the current, in electrostatic units per square meter of surface exposed to a blast of snow, and W the weight of snow which strikes this unit surface in unit time.



A--TiO_2(Anatase Form), Thin Film; Sol No. 155
B--Colloidal Silica No. 155 in Cellulose Nitrate
C--Bare 52-S Alumini
D--Aircraft Wax on 52-S Aluminum
E--Amphibious Transport Paint

Figure 2-35. Charging Coefficient as a Function of Temperature

3.3 DESIGN OF AIRCRAFT STRUCTURES

From the standpoint of freedom from interference from precipitation static, the ideal shape for an airplane would be a smooth sphere. This being functionally impossible, there remain, as a means of a practical solution to the problem, attempts to reduce the curvature of sharp-edged and pointed structures without interfering with their mechanical operation. This includes rivet heads, projecting edges of sheet-metal, exposed surfaces of Pitot tubes and thermometers, all of which have received attention in recent aircraft design. Since corona discharge begins at the point of sharpest curvature, it is necessary to reduce the curvature of all critical areas to the same value, so far as possible, in order to keep the potential of the whole aircraft uniformly high and to prevent breakdown below the operating voltage of static dischargers. In some cases, covering the sharp edges with plastic insulation is effective, unless it interferes with normal air-stream flow.

The quantity of electricity, Q, on any metallic body is equal to the product of its electrical capacity, C, multiplied by the potential, V, applied to it, thus $Q = CV$. The electrical capacity of an insulated sphere in electrostatic units is equal to its radius, r, in centimeters, that is, $C = r$. Electrical charges distribute themselves equally over the entire surface so that $V = Q/r$, but if the sphere is surrounded by a medium of dielectric constant, k, then $V = Q/kr$. In the consideration of aircraft, k is approximately one for air.

The surface density of charge, s, equals the total charge divided by the area of the sphere; therefore,

$$s = \frac{Q}{4\pi r^2} = \frac{V\ r}{4\pi r^2} = \frac{V}{4\pi\ r} \qquad (2\text{-}1)$$

For any given potential, therefore, the charge density varies inversely as the curvature, $1/r$. It approaches zero for a plane surface and becomes infinite at the tip of a sharp-pointed needle.

The atmosphere always contains considerable numbers of ions produced by ultra-violet light, cosmic rays, radio-active substances, engine exhausts, etc. Such ions are accelerated towards or away from intense fields, according to their signs, whether positive or negative, and may gain sufficiently high velocities to ionize more molecules by collision and thus initiate a corona discharge with its resulting static interference. Corona discharge generally begins at about 1/2 to 2/3 the potential required for a disruptive spark, although, depending on the geometry involved, it has been observed at 1/10 this potential.

Laboratory experiments have shown that the break-down potential between polished spheres, 1 inch in diameter, is approximately three times as great as for needle-points. One hundred kilovolts can bridge a six-inch gap between needle-points in dry air. Exact break-down voltages have been shown to depend on materials used, temperature, and pressure.

The break-down potential from a metal surface to the surrounding air depends on the density of the air, being a function of the mean free path of the molecules. For aircraft at high altitude the pressure decreases with the altitude while the temperature falls at the same time, increasing the air density. While these two factors affect the air density in opposite directions, the relations are not linear and the pressure effect predominates. At 5000-feet elevation, the break-down potential is about eight-tenths (0.8) and, at 10,000 feet, sixty-seven hundredths (0.67) of that at sea level.

Because of the camber, or curvature, of the upper surface of an airplane wing, the air going over the top of the wing surface must travel farther, and hence have greater relative velocity, than the air passing the comparatively straight under surface. According to the well-known theorem of Bernouilli, the greater the speed of the air over a surface, the less the pressure. Thus is produced the important lifting effect which keeps the plane in the air but incidentally increases the tendency to go into corona over these areas of low pressure. Because of the pitch and relatively high speed of the propeller tips, they are likewise surrounded by a region of reduced pressure as they plough through the ambient snow and ice particles. Consequently, the propeller tips tend to burst into corona almost as soon as bare wire antennas and other sharp metal points. Obviously, both of these Bernouilli effects are inherent and essential to the functioning of the plane. The change in pressure on the wing surface due to camber is not serious because of its large radius of curvature. The only remedy for the propellers is to bond them effectively to the fuselage to keep their potential relative to the airframe as low as possible, and to keep their surfaces free from offending paints and oils.

3.4 IMPROVED ANTENNA WIRE AND INSULATORS

Before the recent advent of antenna wire and fittings of improved design most of the interference to radio reception originated in the antenna and its auxiliary parts. The antenna is of necessity located in an exposed position of high potential. The fine wire of which it was made, to reduce wind-drag, had a very small radius of curvature and the connections to inexpensive strain insulators were crudely fashioned and had sharp-ended, projecting points.

Precipitation static on aircraft employing external wire antennas is usually a result of corona discharge from the antennas. Charging of the whole aircraft is responsible for the corona. This situation has been effectively dealt with by raising the corona threshold of the antennas and providing a discharge path for charges to leave the airframe without creating radio interference. The corona threshold of the antenna may be raised by increasing its radius of curvature or by coating it with insulating material of very high dielectric strength. The Air Force and the Navy now use an anti-corona antenna wire, consisting of a 50-mil diameter, #16 copper-weld, conductor coated to an outer diameter of 183 mils with polyethylene, and the so-called "wick" dischargers on reciprocating-engine transports and bombers.

Antenna fittings currently used with the wire employ smooth metal surfaces having large radii of curvature for increased corona threshold. Figure 2-36 shows the components of antenna assembly AS-315/A.

New fittings are now in production development which will provide complete external insulation for wire antennas. Figure 2-37 shows the components of these new type fittings and antenna masts. Figure 2-38 shows some of the fittings assembled in greater detail. Mechanical limitations make it impractical to apply this type of equipment to high-speed aircraft.

The transmitting and receiving antennas on aircraft are necessarily mounted in such a manner that they are badly exposed to the sources of precipitation static. The relation between the speed of an aircraft and the severity of precipitation static was discussed in Section 3.1. The great advances in aircraft speed accomplished in recent years have necessitated a multitude of improvements in aircraft design, one of the most important of which is the adoption of greatly improved antennas, antenna insulators, and masts.

Field and laboratory tests have proved beyond doubt that the corona can be suppressed by using a wire covered with a polyethylene insulation and supplied with fittings which protect the ends of the wires

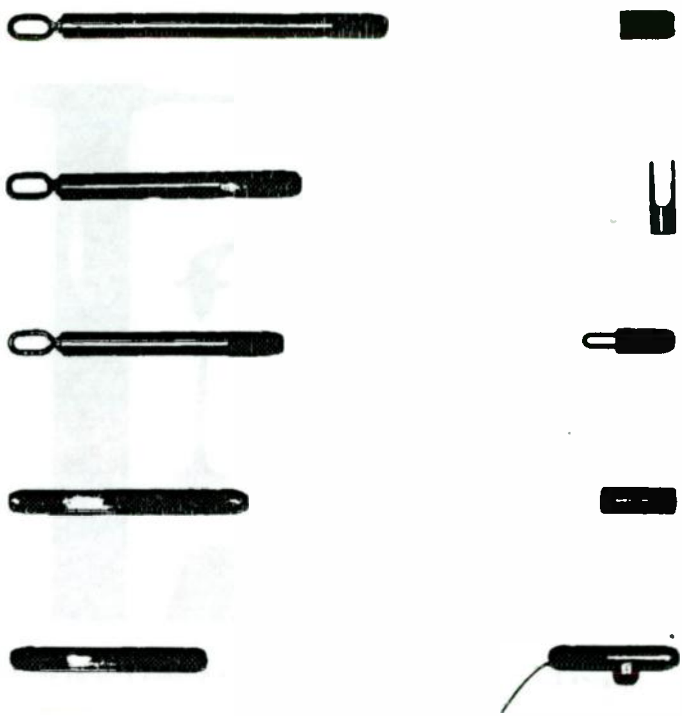


Figure 2-36. Components of Antenna Assembly AS-315/A

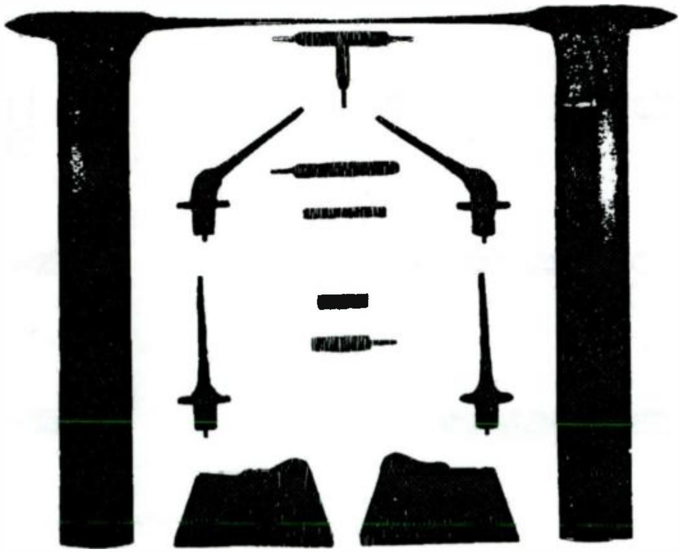


Figure 2-37. New Type Completely Insulated Antenna Fittings and Masts

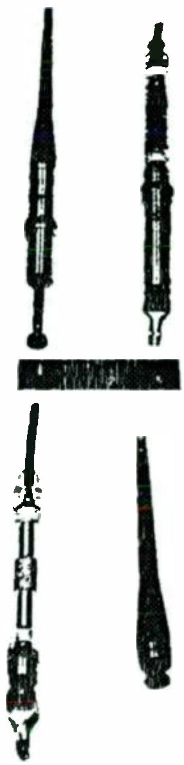


Figure 2-38. Anti-Precipitation Static Antenna Hardware Assembled

from exposure by layers of insulating material. It was found that a #16 copper-weld conductor with a diameter of 50 mils, insulated by a sheath of polyethylene to bring the total outside diameter to 183 mils was a great improvement over the bare wires previously used. Insulators with rounded corners that could be covered with thick layers of tape were used for connections but it was still found that these fittings were the most likely place for the incidence of corona. Difficulty was also encountered from scratches and nicks in the ends of the wires. The new type of antenna hardware includes silicone rubber sleeves, polyethylene thrust washers and molded lucite caps. Instead of coiling the wire around an insulator and twisting its end around a lead-in, a new fitting has been devised into which the end of the wire is forced and clamped between jaws which hold it firmly without twisting.

The new type of strain insulator shown in Figure 2-37 and 2-38 contains two wire-holding chucks separated by an RF insulator and encased in a molded sheath which contains a sealing cavity and is closed at each end with a threaded cap. The Tee-Splice also shown contains three wire-holding chucks fastened to a metal tee spacer for mechanical and electrical connection. The insulation and sealing arrangements at the ends are similar to that of the strain insulator. The Dead-End Mast shown in Figure 2-37 consists of a streamlined plastic (Fibreglass) mast 24 inches long, fitted with an adjustable insert at the upper end for protecting and shielding the outer end of the antenna. The insert contains a wire-holding chuck encased in a molded covering of insulating material. The inside of the masthead is threaded to fit the plastic nut on each end of the insert, thus providing a small amount of "take up".

The Lead-Through Mast illustrated in Figure 2-37 has the same general construction as the Dead-End Mast. No insert is used, but a 1/4-inch hole is molded into the center of the mast to permit the antenna wire to pass through so that it may terminate on the inside of the fuselage of the plane. The Mast Socket also shown consists of a pair of molded brackets designed to mount and support either type of mast on the skin of the aircraft.

3.5 LOOP TYPE ANTENNAS

The advantages of loop antennas are that they can be electrostatically shielded against high impedance fields, that is, fields for which the ratio of electric to magnetic components is large, they can be brought closer to the surface of the aircraft and can be designed to avoid sharp points or small radii of curvature.

The regulation flat-top antennas are, of necessity, mounted as high as possible in order to intercept the required amount of energy from the desired signal. This is where the electrical fields caused by precipitation static discharges produce strong surges into the receiving equipment. A loop antenna, on the other hand, by making use of a large number of turns in the loop, acquires a larger induced voltage at a lower elevation above the aircraft.

For the condition that the circumference or the perimeter of the loop is small as compared to the wavelength, the induced voltage is given by the equation

$$V = \frac{2\pi}{\lambda} ANE \cos\theta \qquad (2\text{-}2)$$

where V is the induced voltage, A the area of the loop in square meters, λ the wavelength in meters, E the field strength in volts/meter, N the number of turns, and θ is the angle between the incoming wave and the plane of the loop. The maximum value for V is obtained when $\cos\theta = 1$. The effective length of the antenna, L_e, in meters, is defined as V/E and has the value

$$L_e = \frac{2\pi AN}{\lambda} \qquad (2\text{-}3)$$

for the condition $\cos\theta = 1$. The current which flows through a loop of impedance Z is

$$I_c = \frac{V}{Z} \qquad (2\text{-}4)$$

The reactive part, X_L, of the impedance Z is always inductive for a small loop. If the loop itself is designed so that the inductive reactance is equal to a capacitive reactance which is either designed into the loop or supplied externally, then the current flowing in the circuit would be a maximum and limited only by the effective resistance. In this case,

$$I_c = \frac{V}{R_e} = \frac{V}{X_L} Q \qquad (2\text{-}5)$$

where Q is defined as the ratio of the reactance, X_L, to the effective resistance, R_e. It is seen, therefore, that the induced current is proportional to the "Q" of the loop.

In using the loop as a radio compass, for example Radio Compass AN/ARN-7 used as a homing device, the width of the loop null is a function only of the radio interference present. As the null point is reached, the automatic volume control in the receiving equipment increases the amplification, and amplifies any interference in the circuit together with the signal. If the receiver is entirely free from interference, the null point may be obtained very accurately, at least to within one or two degrees. One of the disadvantages of the high impedance loop lies in the necessity for physically locating it at such a distance from the receiver that the capacitance of the connecting cable necessitates other changes in the circuit which consequently decrease the effective height of the loop and also reduce its induced voltage. Present practice employs a low impedance loop which is connected to the receiving set through a special transformer and coaxial cable. Either type of loop is enclosed by a metalic shield connected to the airplane structure. When the shielded loop is put in a position parallel to the line of flight, in which case the azimuth angle of the compass needle is at 90°, the loop antenna gives readable signals long after signals from open antennas are lost in crashing noises. In this 90° azimuth position, the loop intercepts the maximum signal and minimum interference. With the local disturbances held to a minimum, the signal voltages can usually be amplified to the point where they are readable above the irreducible interference background.

3.6 TRAILING WIRE DISCHARGERS

Trailing wire dischargers were employed several years before Static Dischargers AN/ASA-3 became available. This older type consisted of a fine steel wire connected through a resistance of 100,000 ohms or more to a high potential point at the rear end of the fuselage. Such a wire in actual use was 15 to 30 feet long. Because of the small diameter of the bare wire, there was a tendency for it to go into corona before any other part of the airplane. Corona discharge at the end of the trailing antenna would often couple back to the receiving antenna and appear as another source of interference. A large resistance in the circuit was used to prevent oscillatory discharges; however, it was found that it was of great importance to have the resistance distributed rather than lumped in one place. In its most useful form, such a trailing antenna is put into a small shield from which it can be released by the pilot through an electromagnetic device operated from the cockpit. A cup attached to the outer end of the wire holds it in the windstream. The use of this device has frequently enabled a pilot to reduce the precipitation interference enough to get range signals through; however, it does add additional hazards so far as lightning strokes are concerned and is falling into disuse. Both commercial airlines and the military have experienced a high incidence of lightning strikes to trailing dischargers.

3.7 FLUSH ANTENNAS, CANOPIES, AND RADOMES

Integral antennas may be divided into two classes for consideration of precipitation-static effects. One class is comprised of antennas which utilize portions of the external structure, such as wing and tailcap antennas; the second is comprised of antennas housed within insulating material.

One problem which is important for integral antennas of the first class is their corona threshold. Flight tests have shown that although jet engines are in themselves fairly efficient dischargers they do not completely compensate for higher rates of impact charging produced by higher speeds. Net charge accumulated on the airframe raises the potential of the aircraft to a point where corona discharge from extremities may occur. Since integral antennas are often located at such points, corona discharge is still a problem. There are three methods of dealing with this situation. The first is to obtain more efficient dischargers, so that the airplane potential remains low. The second is to shape and locate the antennas, if possible, in such a way that they have a high corona threshold. The third method is to control antenna discharges to greatly reduce the magnitude of radio-frequency interference accompanying the corona.

The most fruitful approach at present is that of designing the antenna for high corona thresholds so that large amounts of charge can accumulate on the airframe without causing interference. It seems doubtful that much consideration has been given to this factor in the past, and investigation of some of the aspects of the problem would be very worthwhile.

In general, some of the same factors which provide good efficiencies on low-frequency antennas tend to lower the corona threshold; that is, the antenna may function best if it is located at points where electrostatic field intensification is high. The problem of computing electrostatic field intensities at various points on an aircraft is tedious and difficult. Some measurements of electric field at different points on a typical bomber have been made in flight, and these measurements may serve to indicate the orders of magnitude of some of the ratios. Figure 2-39a shows the points at which measurements were taken, and the ratio of the field at these points to a point near the center of gravity.

Some flight data obtained from corona threshold measurements on dischargers showed that the field intensity at the tips of the horizontal and vertical stabilizers is comparable to the field intensity at the wing tips. Any designs utilizing aircraft extremities as antennas should be carefully considered from the standpoint of probable corona threshold, because high values of electric field are obtained at these points with comparatively small amounts of charge residing on the airframe. The best solution would be to avoid placing antennas on the extremities, if that can be done without affecting performance. If performance requirements dictate placement on wing or tail extremities, every effort should be made to maintain large radii of curvature of the conducting surfaces.

A comparatively large number of integral antenna designs utilize some form of housing constructed of insulating material, with antenna conductors close to the inner surface of the housing. If there is no conducting path provided across the outer surface of the housing, it is possible for bound charges to build up on the insulating material as a result of impact charging. When the field gradient between charges or between charge and airframe becomes high, disruptive discharges occur. This "streamering" can produce interference with characteristics similar to those of antenna corona, with the exception that the number of bursts per second may not be as high. This type of interference may be very intense with small charging currents, which results in disruption of reception even in very light precipitation. Radio-compass sense antennas installed underneath plexiglass canopies have proven very susceptible to this type of interference. Plexiglass has excellent insulating properties so that very small charging currents to the plexiglass surface develop high voltage and consequent streamering. Some flight measurements made on the nose canopy of a typical bomber showed 10,000 microvolts of interference developed on a test antenna mounted on the inner surface when the charging current to the outer surface was considerably less than one microampere per square foot. The interference measurements were taken at 300 kilocycles with an effective bandwidth of about eight kilocycles. At the same time the interference was measured on the test antennas, the radio-compass, which uses a sense antenna mounted on the underside of the pilot's canopy, was inoperative on range signals. On this airplane it has also been observed that the interference from the compass loop, which is mounted at the rear of the pilot's canopy, becomes very large at slightly higher charging currents.

An additional undesirable effect can occur from charge accumulation on insulating surfaces. Electric fields are produced on the inside of the housing which may reach intensities sufficient to cause corona from the conducting structure of the antenna. Some measurements of corona current of this type have been made on a typical bomber antenna. The corona is usually intermittent, but in conditions where high values of charging current are present it can contribute significantly to the interference.

Interference-producing effects on insulating structures can be greatly diminished by providing suitable conduction paths over the outer surface. Since the currents involved are quite small, a high order of conduction is not required. If the surface is coated with a thin layer of conducting material, the lower limit of conductivity is set by the maximum voltage difference which is permitted between the airframe and the surface of the housing and the upper limit is set by shielding effects on radio signals. A limited investigation has shown that, in general, higher conductivity is permissible at higher frequencies, and that surface resistivities ranging from three to ten megohms per square are satisfactory for most applications. Calculated and measured attentuation effects at low and medium frequencies are shown in Figure 2-39b. It can be noted that increased losses occur at lower frequencies for a given coating resistivity.

The problem of interference coupled from corona or streamering regions to antennas at some distance away has not been given much consideration up to the present time. Coupled interference is more serious for the case of weak-signal reception; scanty observations seem to show that it usually is not a severe problem with moderate signal strengths. Capacitive antennas are apparently more susceptible to interference pick-up from points in corona than loop antennas. For example, corona points on the wingtip of a typical bomber coupled interference to the top-wire antennas. Flight and laboratory tests made with compass loop antennas showed that corona points more than four feet away from the loops did not couple in much interference unless the loops were oriented for maximum pick-up in the direction of the point. Interference coupled to wire antennas was greatly reduced or eliminated if the corona point was screened from the wire by portions

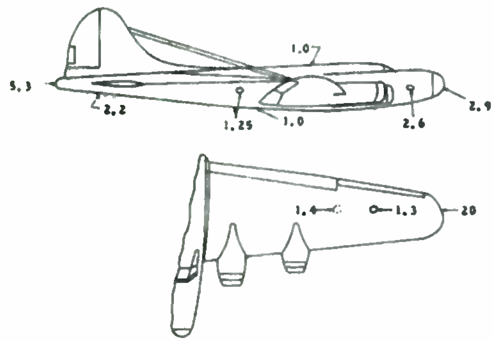


Figure 2-39a. Relative Measurements of the Electric Field as Measured on the Surface of a Typical Bomber in Flight

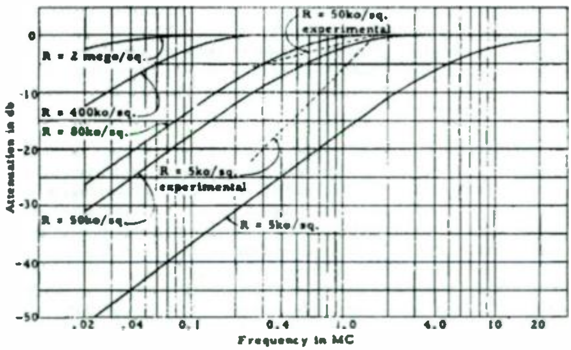


Figure 2-39b. Attenuating Effect of Shield Coatings of Various Surface Resistivities

of the fuselage. A point worth stressing is that coupled interference may prove troublesome if the more usual types of interference are eliminated.

3.8 STATIC DISCHARGER AN/ASA-3

Experiments have proved that charges can escape from sharp points on an airplane without producing radio interference. This helps to keep the aircraft at a low potential with respect to the surrounding atmosphere and thus minimizes the possibility of corona discharges. Steel needle-points have a tendency to become corroded so that they are no longer sharp points. Fine cotton fibers have been found to be effective and practical when formed into a so-called wick, as in the AN/ASA-3 Static Discharger. Each fiber serves as a separate point from which charges can escape. The cotton is treated with a compound which reduces its resistance but keeps it high enough so that the current escaping from each fiber is very small and cannot become oscillatory because of the large resistance in the discharge circuit. The dischargers are mounted at points of maximum electric intensity and removed as far as possible from antenna leads to reduce coupling effects. They are not completely effective but are capable of very greatly reducing the radio interference. When properly installed and serviced they will discharge, without appreciable interference, up to 250 microamperes per wick.

After exhaustive tests by both civilian and military personnel the "Static Discharger AN/ASA-3" has been officially adopted for use on military aircraft. This discharger consists of a conducting cotton wick 13 inches long and enclosed in a plastic sheath. At one end is an aluminum tube which fits tightly around the wick. The outer end of this tube is flattened to provide for two mounting holes. In operation, the plastic sheath is removed from 1-1/2 inches of the free end of the wick, as shown in Figure 2-40a.

As shown in Figure 2-40b, the dischargers are mounted on the trailing edges of wings, rudders, and stabilizers where the potential produced by precipitation static is highest, but in far enough to escape the extreme turbulence at the wingtips. They should not be placed where oil spraying from an engine exhaust may produce matting of the fibers. At least sixteen dischargers are required per airplane, depending on the size, speed, and type of airplane. The only servicing required is to keep 1-1/2 inches of clean cotton wick exposed

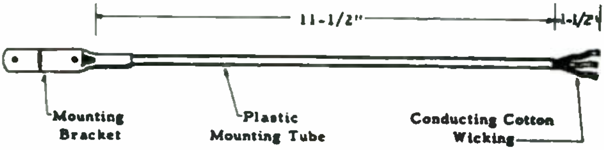


Figure 2-40a. Static Discharger AN/ASA-3

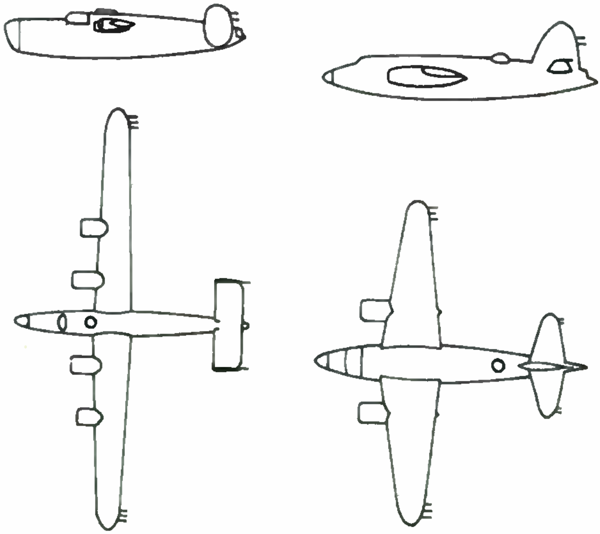


Figure 2-40b. Suggested Locations of Static Discharger AN/ASA-3

outside the plastic sheath at the free end. Replacement should be made when the wicks become shortened by wear to a length of 6 inches.

3.9 BLOCK AND SQUIRTER DISCHARGERS

It has been repeatedly observed that precipitation static charges on an airplane can be reduced by use of the radio transmitters in the plane so that radio signals can be received for a short time immediately after transmitting, even through the most serious interference. It has also been noted that following a lightning discharge in the vicinity of a plane under thunderstorm conditions there are a few seconds of comparative quiet during which signals come through clearly. The explanation is that the additional potential applied to the plane by the transmitter adds a sufficient impulse to discharge the accumulated negative electricity, which is responsible for the precipitation interference. This is because the plane acts as ground for the radio transmitter while its antenna is highly insulated from the remainder of the aircraft. The additional potential given the plane and antenna during the transmission helps to discharge the accumulated static down to a level low enough for the signals to get through. In like manner, the presence of a thunder cloud produces increased negative potentials on one wing and increased positive on the other, both of which may be above the corona point. When the lightning discharge occurs, the space charges within the cloud are reduced and the induced charges on the airplane neutralize each other to such an extent that signals can get through.

Making use of this principle, attempts have been made by several experimenters to produce a system within the plane by which the receiving circuits are momentarily blocked while additional potentials are applied to the plane, or to special dischargers attached to the plane. The time interval during which the receivers are blocked may be three or four times as great as the listening time and is usually adjustable, but there still is a sufficient interval of listening time to obtain range signals clearly. Tests have proved that such a system is expensive to operate. It adds considerable weight and a fire hazard due to the high potentials, and is the equivalent of only two of the AN/ASA-3 dischargers in reducing the interference.

4. INTERFERENCE PARAMETERS FOR PASSIVE SATELLITE COMMUNICATIONS

The following discussion provides a summary of probable propagation phenomena that will provide interference to a passive satellite global communications complex. These phenomena are discussed, and parameters are described which are necessary for obtaining a good estimation of interference. In the discussion on interfering signals, important considerations are presented relative to the sources of propagation and spectrum signature. The effect of the interfering signals on the receiver is dependent upon receiver spectral signatures, detector considerations, antenna noise temperature, and other factors discussed in the text. The evaluation of interference is made on some characteristic of the output signal. A discussion of the necessity of determining subsets, i.e., a set of signals containing samples that have essentially the same effect on the measure of interference is contained in this section.

The interference prediction technique involves the computer simulation of the entire electromagnetic environment of the system. This portion discusses necessary simulation for obtaining estimates of fields emanating from the variety of interference sources. The subset classification of sources is discussed as part of the technique. The necessary parameters involved may not be obtained through empirically derived models, and for this reason brief mention is made of instrumentation.

4.1 INTERFERENCE SOURCES AND PARAMETERS

Interference that appears at a receiving site for satellite communications may be categorized in three major groups. They are:

a. Interference which results from undesired radio frequency reflections from the satellite.

b. Ambient radio frequency energy level that exists at the geographic location of the site; this interference is present in the absence of the satellite and the receiving system.

c. Interference resulting within the system itself; this type is strictly a system function and excludes groups (a) and (b) above.

The description of these interference sources indicates parameters necessary for the successful estimations of expected interference. It is the purpose of this section to provide the sources that cause interference in passive satellite relaying and the parameters that are necessary for a reliable estimate of interference level.

4.1.1 INTERFERENCE RESULTING FROM UNDESIRED REFLECTIONS FROM THE PASSIVE SATELLITE

Because of the attenuation that exists in the various modes of propagation, interference resulting from reflections from the satellite will be caused from sources within radio line-of-sight of the satellite. Assuming that the passive relay stations will orbit at an altitude of 2500 miles, the solid angle in which interference sources may exist at a given time is described by a cone with its apex at the satellite having an apex angle of approximately 76 degrees. This implies that the satellite is in view of approximately one-fifth the surface area of the earth at all times. Electromagnetic radiations within this region are possible sources of radio frequency interference. The probability of a particular emanation of radio frequency energy being a contribution is a function of a great number of parameters. These parameters are discussed in the following subsections.

4.1.1.1 Reflection From a Passive Satellite

The received signal level from the passive repeater may be determined by exactly the same means by which the return to a radar is calculated. In satellite communication, the received signal characteristics are altered by Doppler shifts, Faraday rotation, scintillation and atmospheric losses.

a. General Loss Expression

In determining the propagation loss in a communication circuit using passive repeaters, it is generally assumed that the signal strength is attenuated by free-space loss to the repeater. The energy is then collected by the effective area of the passive repeater, and reradiated isotropically to the receiver site. Additional free space loss also results in the transmission from the passive repeater location to the receiver. The received signal level becomes

$$S_R = \left(\frac{P_T G_T}{4\pi R_1^2}\right) \left(\frac{A_R}{4\pi R_2^2}\right) \left(\sigma(\theta, \varphi)\right) \qquad (2\text{-}6)$$

where: P_T = transmitter power

G_T = transmitter antenna gain

R_1 = distance from transmitter to passive repeater

R_2 = distance from passive repeater to receiver

A_R = effective area of receiver antenna

$\sigma(\theta, \varphi)$ = scattering cross section as a function of direction to receiver

The passive repeater in this discussion is a spherical satellite. Assuming that the dimensions of this satellite are considerably greater than a wavelength, the scattering cross section is

$$\sigma(\theta, \varphi) = \pi a^2 \tag{2-7}$$

where a is the radius of the spherical satellite. With this assumption, the expression for the received power becomes

$$S_R = \frac{G_T A_R a^2 P_T}{16\pi R_1^2 R_2^2} \tag{2-8}$$

Expressed in decibels relative to one watt (dbw), Equation (2-8) becomes

$$(S_R)_{dbw} = (P_T)_{dbw} + 20 \log_{10} (a)_{ft} - 20 \log_{10} (R_1)_{mi} - 20 \log$$

$$(R_2)_{mi} - 20 \log (f)_{mc} + (G_T)_{db} + (G_R)_{db} - 117 \tag{2-9}$$

When the passive relay satellite is a 100-foot radius balloon, Equation (2-9) becomes

$$(S_R)dbw = (P_T)_{dbw} + (G_T)_{db} + (G_R)_{db} - 313 \text{ db} \tag{2-10}$$

assuming an overhead satellite at an orbiting altitudeof 2500 miles and a frequency of 1 kmc. The noise level in an ideal receiver with an ambient noise temperature of 300° K is -174 dbw per kilocycle of bandwidth. Therefore, to provide a detectable signal it becomes necessary to provide sufficient transmitted power and/or antenna gains and/or reduced receiver noise temperature. In the 1 kmc region, the maximum CW power easily attainable is approximately 10 kw (40 dbw). There are also limitations on the size of movable antennas that can be constructed. It, therefore, becomes necessary to incorporate in the system design of a satellite relay system extremely low temperature receivers (~30° K). Although these receivers satisfy the system gain requirements, they provide a severe limitation on the maximum acceptable interference level.

The gain requirements indicated by the received level expression must be altered to include the additional free space loss that occurs since the free space distance to a satellite will be a minimum of 2500 miles. The maximum additional loss due to this factor is approximately 6 db. In addition, the loss must also be altered to account for scintillation, atmospheric absorption, and Faraday rotation in linearly polarized transmissions. Although Doppler shift may not increase gain requirements, this transmission characteristic does affect system performance and, therefore, requires consideration.

b. Scintillation

Scintillation is the fluctuation of the receiver signal level caused by irregular refraction in the atmosphere. These refractive index variations are the result of irregularities existing in the electron density in the ionosphere and turbulence in the troposphere. The mean square fractional deviation of amplitude is proportional to the 4th power wavelength and to the cube of the secant of the zenith angle. Scintillation experiments indicate that the secant law is obeyed except when the zenith angle approaches 90 degrees. At frequencies of 9375 mc and 32, 320 mc, measurements indicate that fluctuations are maximum up to about 3 degrees elevation (87 degrees zenith angle) and decrease to very small perturbations after about 5 degrees. The period of the scintillation cycle averages about 15 to 22 seconds at 9375 mc and 35 to 55 seconds at 32, 320 mc. The depth of fade at any wavelength will vary from a fraction of 1% to as much as 20%. The average fluctuation at the most disturbed elevation is between 5 and 10% regardless of the frequency band. The reduction of scintillation can be greatly reduced in the 1 to 10 kmc frequency band by operating at an elevation greater than 5 degrees.

c. Atmospheric Absorption

Microwave energy is absorbed by constituents of the atmosphere, primarily oxygen and water vapor. In the 1 to 10 kmc regions, the expected attenuation resulting from the absorption is less than 0.01 db per kilometer. Transmission toward the horizon will pass through approximately 200 km of atmosphere. Therefore, at this zero degree horizon angle, less than 2 db of attenuation is expected. The expected loss decreases as the elevation angle is increased since less atmosphere is encountered in the transmission. Although the loss appears almost insignificant, serious absorption may be encountered during heavy rains.

d. Faraday Rotation

Faraday rotation is a propagation phenomenon that results in a rotation of the polarization of an electromagnetic wave that passes through the ionosphere. A linearly polarized system will exhibit a fading characteristic due to the variation in polarization. The depth of a fade due to Faraday rotation is related to the degree of polarization shift and the polarization isolation at the receiver site. Faraday rotation, like scintillation, is related to the frequency and the secant of the zenith angle. In general, in the frequency region of interest, the rotation will be only a few degrees. However, at the horizon (zenith angle = 90 degrees), it may be necessary to consider the effects of Faraday rotation.

e. Doppler Shift

The existence of a relative velocity between the transmitter and the satellite, and the satellite and the receiver, provides a shift in the received frequency. At an orbiting altitude of 2500 miles, a maximum shift of approximately 20 kc will occur in 1 kmc transmissions and a shift of 200 kc in 10 kmc transmissions. The doppler shift in frequency will provide an increase in frequency as the satellite is approaching and a decrease as the satellite is departing. The absolute value of doppler shift is a function of the spatial position and velocity of the satellite with respect to the transmitting and receiving site locations and velocities. In addition to the frequency variation, the transmitted spectrum will be slightly varied due to the frequency dependence of doppler shift.

4.1.1.2 Possible Sources of Interference Caused by Undesired Reflections

Sources that provide interference due to the undesired reflection of their transmissions must, because of the severe path loss encountered, be within the radio line-of-sight of the passive satellite. This does not greatly localize sources since the field of view of the satellite subtends approximately one-fifth of the earth's surface. Also in order to significantly contribute to the interference, the sources must be high power radiators. It is the purpose of the following discussion to indicate possible sources and characteristics which are pertinent to the determination of interference.

a. Radars

Perhaps the greatest contributor to this particular type of interference will be both stationary and movable (i.e., airborne, shipboard installations, etc.) radars. These radars are of sufficient power to provide significant reflections from the satellite. To determine the interference resulting from a radar, it is necessary to know the general mode of operation, type of transmission, and the spectrum signature associated with its transmission.

The mode of operation explains the function of the radar and therefore, indicates the programmed scan. This is essential since it provides the probability of satellite intercept as a function of the relative position of satellite and radar. As an example, a horizon search radar will concentrate most of its transmitted power in the direction of the horizon, thus maximizing its contribution to interference at the time when the satellite is near the radar's radio horizon.

To determine the total reflected energy from the satellite, it is also necessary to know the type of transmission that is being employed. Of major concern is the pulse repetition rate and the duty cycle. The interference effects on a desired transmission is definitely a function of these parameters.

Spectrum signature is often defined to include both mode of operation and type of transmission. In this discussion, spectrum signature is defined to include all radiation (fundamental, spurious, and harmonics) in terms of the spatial distribution of emanated energy as a function of frequency. Spectrum signature in general does not include terrain effects on the free space patterns and this effect must also be considered. A detailed spectrum signature for this type of interference may not be necessary since the extreme propagation loss implies negligible contributions to the greatly suppressed spurious and harmonic transmissions.

An indication of the necessity of spectrum signature may be supplied by consideration of a horizon search radar and a height finding radar. The scans employed may be quite similar, but the lobing that may occur on the height finding radar presents an entirely different interference effect at a particular satellite position than does the horizon search radar.

The interference from airborne and shipboard radars must be determined in essentially the same manner as a fixed installation radar, but with the added complexity of the probabilistic model determined from the estimated missions of the carrier.

b. Tropospheric Scatter Terminals

Communication circuits employing the tropospheric scatter mode of propagation are also characterized by severe propagation losses. Therefore, it is expected that a long range communication circuit utilizing tropospheric scatter will employ high power transmitters and high gain antennas and, therefore, may provide a significant interference level at the received site.

The antenna position of a transmitting terminal of a tropospheric scatter site is fixed. Since propagation loss increases rapidly with respect to elevation angle, the antenna will be beamed at the horizon. Antenna beamwidth will, in general, be less than 5 degrees. The optimum frequency of operation is 1 to 2 kmc. The most probable type of transmission is CW with narrow band FM modulation.

Interference resulting from the reflection of energy radiated from a tropospheric scatter terminal will be most significant when the satellite is at the radio horizon of the transmitting terminal. In this case, it is also necessary that the spectrum signature of the transmitting site be known before a successful estimate of interference can be made.

c. Other Main Trunk Sites

Another source of high power radiations within the frequency band of interest is the transmitter of other sites that are also in view of the satellite. To realize the fields created at a receiver site, it is necessary to know the spectrum signature of the transmitting sites. Also of interest, is the backscatter that will result from transmissions of the transmitter located in the vicinity of the receiver. In general, this interference may not be too significant since these factors may be carefully minimized by antenna design, receiver design and proper frequency allocation. The probability of significant interference resulting from the reflection of energy from a satellite to a satellite and then to the receiver is essentially zero because of the additional attenuation created by the extra "bounce."

d. Other Sources

Several specialized systems exist that may appear to be sources. In general, these systems must be approached in a manner similar to the approaches indicated in the preceding sections. Reflection from the sun may be thought to be a possible source. However, even with an albedo of unity, this radiation is insignificant because of the very small solid angle subtended by the satellite as viewed from the receiving site. A truly significant source would be the inclusion of the sun in the main beam of the receiving antenna or even perhaps in a minor lobe. A detailed discussion of this effect is given in Section 4.1.2.

4.1.1.3 Received Field Strength from Undesired Reflections from the Satellite

Section 4.1.1.1 has provided an expression for the expected signal strength of a circuit consisting of transmission to the satellite, the reflection from that satellite, and the transmission to a receiver site. If the aperture of the receive antenna, A_R, is omitted from Equation (2-8) the equation becomes the power density at the receive antenna. This power density, however, must be altered by multiplying the loss factors discussed in 4.1.1.1.b, 4.1.1.1.c, and 4.1.1.1.d to provide the expected power densities. Fading about the expected value must be determined from the characteristics of the individual factors and incorporated in power density computations if appreciable. The product of P_T and G_T in the expression is the spectrum signature determined at the frequency of interest. In equation form, this product may be written as

$$P_T G_T = H(f, \theta, \varphi) \qquad (2\text{-}11)$$

where θ and φ are the orthogonal angles necessary to provide the unique angular location of the satellite with respect to the source of the undesired transmission. R_1 and R_2 have been defined on page 2-118, and are obtained from the spherical geometry of the system.

4.1.2 AMBIENT RF ENERGY LEVEL

The ambient radio frequency energy level referred to in this section is the energy level present at the receiving site excluding the received energy due to the presence of the system. In general, this ambient is composed of contributions from:

a. Equipments designed to emit radio frequency energy
b. Galactic and ground noise
c. Man-made noise

A discussion of possible sources and required information concerning these sources is contained in the following subsections.

4.1.2.1 Equipments Designed to Emit RF Energy

The low noise temperature of the receiver dictates the consideration not only of radio line-of-sight transmissions to the receiver site, but also of radiations received from beyond the horizon sources.

a. Energy Received from Radio Line-of-Sight Sources

Energy which may be sufficient to cause interference may emanate from essentially all radio frequency energy emitters in the line-of-sight region of the receiver. Sources will include microwave links, radars (particularly airborne), television transmitters, tropospheric scatter terminals, etc.

The basic equation for the determination of the energy density received from a particular source is

$$P_R = \frac{KH(f, \theta, \varphi)}{4\pi R^2} \qquad (2\text{-}12)$$

where: $H(f, \theta, \varphi)$ = spectrum signature of the transmitter

f = frequency region of interest

θ = azimuth angle to receiver site

φ = elevation angle to receiver site

R = distance to the receiver site

K = factor added to include terrain effects in the receive direction and reflections from obstacles.

Since the distances involved are small and the distance dependence is less than in the case of the satellite reflection, it is essential that an extremely detailed spectrum signature be available for the calculation of the expected power density.

In the line-of-sight transmission of microwaves, fading occurs due to variations in the refraction index between transmit and receive sites. A detailed interference study must include these statistical variations.

As stated in preceding sections, the interference realized is not only due to the received field strength, but also to the type of transmission.

b. Energy Received from Beyond the Horizon Sources

The possible sources of interference from beyond the horizon radiations is somewhat limited since the modes of propagation providing transmission do incur losses in addition to the free space loss. Modes of propagation that may exist are reflection, diffraction, tropospheric scatter and ducting.

(1) Reflection

A high power radio frequency source illuminating an obstacle which is in line-of-sight of both transmit and receive locations may yield a signal of strength sufficient to produce interference. Unfortunately, methods of estimating the scattering cross section of irregular terrain or experimentally developed estimates of the scattering cross section has not been obtained. Some experimental information is available, however, the available information is not sufficient to formulate necessary functional relationships.

The necessary modification of Equation (2-12) to provide an expression for the determination of power density due to reflection is

$$P_{RR} = \frac{KH(f, \theta, \varphi)}{4\pi R_1^2} \cdot \gamma \qquad (2\text{-}13)$$

where: R_1 is the distance from transmitter to obstacle and

$$\gamma = \frac{\sigma_0}{4\pi R_2^2} \qquad (2\text{-}14)$$

σ_0 is the scattering cross section of obstacle and R_2 is the distance from the obstacle to the receiving site. The scattering cross section must be a function of reflectivity of the obstacle, the angles of arrival and departure, degree of roughness and, therefore, frequency and orientation with respect to propagation path. The reflectivity will be long term time variant and, therefore, a long term fading characteristic is associated with γ.

(2) Ducting

The propagation phenomenon known as ducting results from a slope inversion in the refractive index profile. In tropospheric scatter systems, the occurrence of a duct can increase the received signal levels by as much as 40 to 50 db. Ducting is more prevalent in overwater circuits.

The signal received by means of ducting does not exhibit fast fading characteristics. The occurrence of a duct is not predictable and the creation of a ducting condition can occur rapidly and disappear just as quickly. Because of the tremendous enhancement of signal strength due to ducting, it must be considered in an interference study. The only logical method of approach for the inclusion of ducting is statistical and based on the known meteorological conditions between transmitter and receiver.

The expression necessary for computation of a received field strength due to ducting is as shown in Equation (2-13). γ must now include meteorological information and contain the proper statistical variations as indicated by intervening geographical considerations. Spectrum signature information is necessary only for small horizon angles.

(3) Tropospheric Scatter Propagation

Tropospheric scatter propagation, unlike ducting, does provide a persistent signal. This mode of propagation is the result of refractive index in homogeneities caused by turbulence. The received signal contains a rapid fade characteristic (period of approximately 0.1 cps at 1 kmc and linearly increasing with frequency). This fading has a Rayleigh probability density function. In addition to this fading, long term fading (period greater than an hour) and seasonal variations in the medium level do exist.

Methods of estimating the expected transmission loss with reasonable accuracy do exist. The estimation of this loss may still be provided by Equation (2-13). However, now γ becomes a function of:

(a) Great circle path length
(b) Frequency
(c) Horizon angles

The short term fading characteristic is well defined by the Rayleigh probability density function. The long term characteristic (diurnal and seasonal) is described by a Gaussian probability density function with a standard deviation that is dependent on the path length. Spectrum signature and near terrain effects must be provided at elevation angles not greatly exceeding the site elevation angles.

(4) Diffraction

Beyond the horizon transmission can be obtained by diffraction. Diffraction occurs because of the disturbance in Fresnel zones caused by the presence of an obstacle which is within the radio horizon of both transmitter and receiver sites. An example of an obstacle which will provide diffraction is a ridge that forms the horizon, along the great circle path of the transmitting and receiving site. A diffracted signal is predictable with a fair degree of accuracy and does not exhibit a fading characteristic. γ, of Equation (2-13), when applied to diffraction becomes a function of:

(a) Curvature of edge
(b) Path length
(c) Frequency
(d) Horizon angle, along great circle path, of both sites.

As in the case of tropospheric scatter propagation and ducting, the spectrum signature information and near terrain effects must be obtained for angles not greatly exceeding the horizon angles and in the direction of the great circle path.

4.1.2.2 Galactic, Solar, and Ground Noise

The advent of low noise microwave amplifiers has shifted, at least partially, the responsibility for the lower limit on the detectable signal from the amplifier to the antenna and the radio frequency hardware. Additionally, the term noise temperature is frequently used rather than noise figure. Galactic, solar, and ground noise all contribute to the antenna noise temperature. The following section provides a short discussion of antenna temperature, and is followed by a discussion of the sources and examples of antenna temperature under various conditions.

a. Antenna Temperature

The intensity or brightness of an object such as the sun, a star or a satellite, or a field of view such as the sky may be represented by a temperature distribution $T(\theta, \varphi)$. This is the effective temperature presented by the source in a differential element of solid angle whose position is specified by θ and φ. The origin of the coordinate system is located at the receiving antenna. The temperature required by a black thermal emitter in order to produce the same flux observed at the antenna and having the same location and geometrical configuration as the actual source is defined as the effective temperature of the source. The actual source may be either thermal or nonthermal.

Using the temperature distribution, $T(\theta, \varphi)$, of a source and a differential element of solid angle, $d\omega$, the antenna temperature presented by the source is T_a.

$$T_a = \frac{1}{4\pi} \int T(\theta, \varphi) G(\theta, \varphi)\, d\omega \qquad (2\text{-}15)$$

$G(\theta, \varphi)$ is the gain function of the antenna relative to an isotropic radiator. Using the definition

$$G_o \Omega_a = 4\pi \qquad (2\text{-}16)$$

where: G_o is the antenna gain in the direction of maximum gain and Ω_a is solid angle representing the effective antenna beam, Equation (2-15) becomes

$$T_a = \frac{1}{G_o \Omega_a} \int T(\theta, \varphi) G(\theta, \varphi) d\omega \qquad (2\text{-}17)$$

The integration may be approximated by a summation of the temperatures associated with each lobe of the antenna as

$$T_a \doteq \frac{1}{G_o \Omega_a} \sum_{n=o}^{N} G_n T_{sn} \Omega_n \qquad (2\text{-}18)$$

Where: G_n is the maximum gain in the nth lobe, T_{sn} is the effective temperature of a source emitting energy to the nth lobe and Ω_n is smaller of the two solid angles, (1) the effective solid angle of the nth source or (2) the effective solid angle of the nth lobe. An important special case is one in which the sidelobes are all relatively unimportant and a source of temperature T_s, high relative to the background, fills a solid angle Ω_s within the main beam of the antenna. For this case the antenna temperature is approximated by

$$T_a \doteq \frac{T_s \Omega_s}{\Omega_a} \qquad (2\text{-}19)$$

b. Antenna Noise Sources

The sources of noise to be discussed in this section will include normal sky background temperature, solar temperature, and ground temperature. Man-made noise will be discussed briefly in the next section.

The effective noise temperature of the sky excluding the sun varies quite radically over the frequency range 1 kmc to 10 kmc. For the noisiest portion of the sky excluding the sun, the temperature is roughly 60°K at 1 kmc, 6°K at 2.5 kmc, and 30°K at 10 kmc. From 20 mc to 2.5 kmc, the major contributing source to the temperature is cosmic noise. From 2.5 kmc to 100 kmc, the major contribution is from atmospheric absorption noise. There is an apparent minimum in the noise temperature at approximately 2.5 kmc. At this frequency the temperature is due mainly to the reradiation of thermal energy by oxygen molecules.

The effective noise temperature of the sun also varies greatly as a function of frequency. Measured average temperatures at 218 mc, 9.375 kmc, and 32.32 kmc are 1,000,000°K, 16,000°K, and 5280°K, respectively. The angular diameter of the sun is approximately one-half of a degree. Since the solid angle subtended by the sun is small, and the antennas used have narrow beamwidths, the probability of the sun being in the antenna beam is small.

The effective noise temperature of the earth is approximately 300°K. The ground temperature becomes important in satellite communication systems because of the minor lobes, backlobes, and spillover.

c. Examples of Antenna Temperature

An approximate expression for the determination of the effective antenna temperature has been developed in Section 4.1.2.2.a and is repeated here.

$$T_a \doteq \frac{1}{G_o \Omega_a} \sum_{n=o}^{N} G_n T_{sn} \Omega_n \qquad (2\text{-}20)$$

G_o is the maximum gain the main antenna lobe, G_n is the maximum gain in the nth antenna lobe, Ω_a is the solid angle representing the effective main antenna beam, T_{sn} is the effective temperature of a source emitting energy to the nth lobe and Ω_n is the smaller of the two solid angles: (1) the effective solid angle of the nth source or (2) the effective solid angle of the nth lobe.

The simplest example is one in which the main beam is the only one that provides an appreciable contribution and that beam is entirely filled with a source such as the sky with an effective noise temperature T_{so}. In this situation, Equation (2-20) becomes

$$T_a \doteq \frac{1}{G_o \Omega_a} (G_o T_{so} \Omega_a) = T_{so} \qquad (2\text{-}21)$$

At a frequency of 10 kmc, the maximum T_{so} in the entire surveillance of the sky will be approximately 30°K.

Suppose now that while the main lobe is illuminated by the sky that a backlobe of gain of 0.001 relative to the maximum gain G_o and an effective solid angle Ω_1 = 0.1 steradians is illuminated by the ground (T_{s_1} = 300°K). Assume also that the effective solid angle Ω_a of the main beam is 0.0025 steradians. This will correspond to a beamwidth of approximately 3 degrees in both the horizontal and vertical planes. From Equation (2-20), the effective antenna temperature will be

$$T_a = \frac{G_o T_{so} \Omega_a}{G_o \Omega_a} + \frac{G_1 T_{s_1} \Omega_1}{G_o \Omega_a}$$

$$= 30°K + \frac{(0.001)(300°K)(0.1)}{(0.0025)}$$

$$= 30°K + 12°K = 42°K \tag{2-22}$$

Again a frequency of 10 kmc was used.

As a final example, suppose that in addition to the conditions existing in the previous example the sun appears in a minor lobe. At a frequency of 10 kmc, the sun temperature T_{s_2} will be approximately 16,000°K and the effective solid angle Ω_2 is approximately 0.000076 steradians. The gain G_2 of this minor lobe relative to G_o will be 0.001. From Equation (2-20), the antenna temperature will be

$$T_a = \frac{G_o T_{so} \Omega_a}{G_o \Omega_a} + \frac{G_1 T_{s_1} \Omega_1}{G_o \Omega_a} + \frac{G_2 T_{s2} \Omega_2}{G_o \Omega_a}$$

$$= 30°K + 12°K + \frac{(0.001)(16,000°K)(0.000076)}{(0.0025)} = 42°K + 0.5°K$$

$$= 42.5°K \tag{2-23}$$

The examples given in this section simply illustrate the calculation of the antenna temperature and do not necessarily represent any real physical antenna configuration or interference situation.

4.1.2.3 Man-Made Noise

Man-made noise will include interference produced by sources such as high-tension lines, electric motors, electric switching gear, engine ignition systems, diathermy and other industrial equipment. The interference produced by these sources will be greatest in densely populated and industrial areas and will be greatly effected by the terrain near the site. Certain man-made noise will be predictable with a reasonable degree of accuracy. However, the greatest portion of this noise is not predictable and, therefore, measurements are necessary. In many cases, man-made noise may be negligible because of site location and the fact that the main antenna beam is pointing skyward.

4.1.3 UNDESIRED RF ENERGY CREATED BY COMMUNICATION TRANSMITTER AND ASSOCIATED SYSTEM

The preceding sections have provided a discussion of RF energy at the receiver site produced by sources external to the system. Consideration must also be given to the mutual interference effects that exist between the receiver and the major components of the system. For instance, the receiver site will probably be in near proximity of a tracking radar necessary for receive antenna tracking information and in addition, the transmitter that provides the relaying function. These sources of interference are contained under a separate heading since the effects are somewhat controllable though geographic placement of major components and original major component design. Essentially the parameters already discussed in previous sections will provide interfering signal levels. However, it is possible that near field spectrum signatures will have to be obtained for particular arrangements.

4.1.4 COMMUNICATION RECEIVER PARAMETERS

In addition to the desired signal reflected from a passive satellite, a wide variety of potentially interfering electromagnetic radiation is present at the communication receiver antenna. The basic purpose of the receiver is to recover the original information input to the transmitter such that a minimum amount of error is caused by interference. This section discusses the factors at the receiver which determine interference level.

A general block diagram of the receiver, with special emphasis on interference analysis, is given in Figure 2-41. Each type of radiation

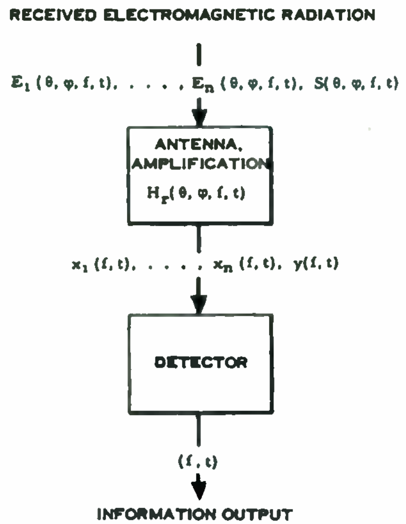


Figure 2-41. Communication Receiver Block Diagram Showing Parameters for Interference Analysis

input $E_j(\theta, \varphi, f, t)$ to the receiver antenna, as discussed in the previous sections, is described by a direction of arrival, field strength, frequency spectrum, and time variation.

The antenna radio frequency (RF) preamplifier, and intermediate frequency (IF) amplifier pickup and amplify the radiation signal at the receiving terminal. The transfer function of this receiver section $H_r(\theta, \varphi, f)$ is a function of antenna direction (θ, φ) and signal frequency (f), and is usually described by antenna spatial characteristics, and receiver sensitivity and selectivity.

The amplified signals from the receiver "front end" are then processed in a decision or detection section. The characteristics of this section are primarily determined by both known parameters of the message source and expected or average parameters of interfering signals.

Finally, the processed signal and interference is displayed or presented in some form to the message destination, such as digital computer input or a human operator. A measure of interference is the effect of interfering signals upon recognition of the desired signal at the "message" destination.

The receiver parameters required for an analysis of interference effects are discussed in the following sections.

4.1.4.1 Antenna and Radio Frequency Amplication

At the low signal levels expected in a passive satellite communications system, the receiving antenna characteristics are very important. Maximum antenna gain with minimal side and backlobes should be realized in terms of state-of-the-art, economics, and tracking requirements.

Since the surface of the earth has an effective noise temperature of 300° K, the noise which will be picked up by the antenna side and backlobes depends on both the antenna spatial characteristics and antenna elevation angle, thus determining a minimum elevation angle. Even with restricted elevation angle, antenna noise equivalent temperature could possibly be larger than receiver preamplifier temperature.

Signal power in the receiver from interfering sources on the surface of the earth, especially line-of-sight, is also very much a function of antenna characteristics. To minimize interference from these sources, sidelobes should be minimized. In addition, minimal sidelobes will minimize interference from extraterrestial sources including the sun. A discussion of antenna temperature and examples are contained in Section 4.1.2.2.

At the low signal levels at a satellite communications receiver, it is desirable to use maser or parametric radio frequency preamplifiers. However, this increases the susceptibility of the system to interference, thus requiring careful receiver design with regard to spurious response and selectivity.

The receiver bandwidth should be selected such that maximum channel capacity is realized, giving maximum signal-to-noise ratio. Actual bandwidth will thus be determined by the modulation technique and information bandwidth utilized in communication. In addition, the design must assure that spurious responses be negligible.

4.1.4.2 Detector

Incoming signals from all sources at the receiver antenna, after being picked up and amplified, are processed (demodulation and detection) to provide a message or signal format desired at the destination. This processing is designed to enhance the desired signal and reduce interference effects, and may include sophisticated techniques such as matched filters or correlation detectors. For simplicity in this discussion, detection will be defined as the signal processing from signal output of the intermediate frequency amplifier (or similar linear radio frequency amplifier) to final user of desired signal information. This detection process is performed by a detector; the term is used in this broad sense, and should not be interpreted as radio frequency detection such as associated with the function of a crystal diode.

Referring again to Figure 2-41, the transfer function relating electromagnetic radiation to the output signal of the radio frequency amplifier is defined as $H_r(\theta, \varphi, f, t)$ where θ is azimuth angle at the antenna site, φ is elevation angle, and f is signal frequency. Thus, letting $E_j(\theta, \varphi, f, t)$ be one particular interfering signal arriving at the receiver, and $x_j(f, t)$ be the corresponding output of the RF amplifier section.

$$x_j = x_j(f, t) = E_j(\theta, \varphi, f, t)H_r(\theta, \varphi, f, t) \qquad (2\text{-}24)$$

Also, defining $S(\theta, \varphi, f, t)$ as the desired signal,

$$y = Y(f, t) = S(\theta, \varphi, f, t)H_r(\theta, \varphi, f, t) \qquad)2\text{-}25)$$

Note that the transfer function $H_r(\theta, \varphi, f, t)$ includes antenna pattern. Thus, at some particular time t and frequency f, the antenna pattern is defined in terms of θ and φ. For this reason, $H_r(\theta, \varphi, f, t)$ is nonstationary, or dependent on time reference.

In this section, $H_r(\theta, \varphi, f, t)$ includes all linear signal amplification prior to detection. The superposition principle applies to this transfer function, such that each incoming signal from each source is picked up and amplified as though that signal alone were present. Thus, no interaction between the input signals $E_j(\theta, \varphi, f, t)$ will take place.

The actual method of detection is largely dependent upon the format of the transmitted signal. For this reason, the interference analysis at the receiver can only be described in general terms.

All signals at the input to the detection system may be described by a set

$$(\tilde{x}, y) = (x_1, x_2, \ldots, x_n, y) \tag{2-26}$$

as in Figure 2-42, where each x_j represents the interfering signal from one particular source, such as a radar transmitter, and is a function of frequency spectrum and time. The desired signal is represented by y, also a function of frequency and time.

The detection system operates on both desired and undesired signals to produce an output ζ. The amount of usable information conveyed by ζ, $I\left\{\zeta\right\}$, is a measure of interference. That is, information output with no interfering signal, $\left[I\left\{\zeta\right\} \middle| \tilde{x} = o\right]$, is a measure of maximum possible information output; interference may be expressed as

$$\text{Interference} = \frac{I\left\{\zeta\right\}}{I\left\{\zeta\right\}_{\tilde{x} = o}} \tag{2-27}$$

which is a maximum of 1.0 for no loss of information due to interference.

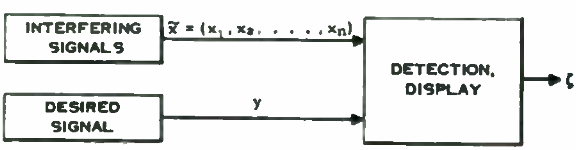


Figure 2-42. Input-Output Relationship of Detector

The set of received interfering signals $\tilde{x} = (x_1, x_2, \ldots, x_n)$ may be arranged in N subsets, such that the members of each subset x_j have the same interfering effects at the detector. For example, one such subset of interfering signals is almost all CW signals in the radio frequency and intermediate frequency pass band. Even though power, time variation, and direction of arrival is somewhat different for each individual member of the subset, the total or subset average may be treated as a form of non-Gaussian noise if the number N is large. The effect of this noise-like signal on desired signal may then be computed, rather than attempting to individually compute the effects of each signal.

Furthermore, the characteristic parameters of each x_j signal, such as power, θ, ϕ, f, and time dependence, may be expected to have a statistical variation, with a given mean (μ_i) and variance (σ^2).

For convenience, the mean value of the signal input to the detector corresponding to a particular source "j" has been designated as $x_j(f, t)$. This is, then, the average signal "j" as a function of frequency and time which would be obtained from a large number of measurements. The corresponding variance $\sigma^2(x_j)$ corresponds to the "mean" $x_j(f, t)$.

The subset $x_i(f, t)$ is chosen to represent the total of the subset of signals with identical interference effect on the desired signal (y). First, assume that the frequency components of each $x_j(f, t)$ are statistically independent; this condition is easily realized, since signals with identical components in the receiver pass band may be added directly. The mean value of the subset signal is then

$$x_i(f, t) = \frac{1}{N} \sum_{j=1}^{N} x_j(f, t) \qquad (2\text{-}28)$$

with N members of the subset. Variance of this mean is defined by

$$\sigma^2(x_i) = \frac{1}{N} \sum_{j=1}^{N} \sigma^2(x_j) \qquad (2\text{-}29)$$

The set of signals at the detector may thus be classified into subsets for interference analysis with particularly distinctive or high power signals, such as interference from the tracking radar, being treated individually. A statistical analysis, giving variance or confidence limits on the interference, is also possible.

4.1.5 DATA REQUIRED FOR DETERMINATION OF SATELLITE COMMUNICATION INTERFERENCE

This section is intended to draw together considerations that must be made in the successful analysis of interference. It should be realized that each consideration is a multivariant problem. The summary of these considerations is as follows:

a. Satellite Information - physical information about the satellite such as its size and scattering cross section and orbital information such as velocity, altitude, and statistical data about its position.

b. Source System Information - position of source, use-time, function, e.g., communications, navigational aid, air search radar, sector operation. This will include information on intentional and unintentional man-made emissions and sources such as the sun, sky, water, and ground.

c. Source Spectrum Signature - spectral power densities over the entire significant frequency range, spatial patterns at significant frequencies, polarization, nature of signal, e.g., pulse, CW, SSB, AM, etc.

d. Geographical Environment - topographical and meterological data having any possible effect on the propagation of an interference signal.

e. Mode of Propagation - reflection, diffraction, line-of-sight, tropospheric scatter, etc.

f. Communication Complex Parameters - positions of sites, receiving system characteristics including antenna, tracking sector of antenna, possible mutual effects from tracking system and other equipments in the system.

It should be stressed that this action provides only a summary of considerations necessary. A detailed analysis of each of the many items included must be performed before an estimation of interference with reasonable confidence limits can be obtained.

4.2 PREDICTION TECHNIQUE FOR SATELLITE COMMUNICATION INTERFERENCE

RFI prediction is a complex multivariant problem. The problem is a formidable one just for radio line-of-sight propagation between earth stations. By adding the constellation of orbiting passive satellites at such great distances above the earth's surface the interference region has become greatly enlarged. Even this problem of increasing the interference producing area is belittled by the fact that the receivers used in in the satellite communication system are generally the low noise type. The total interference effect at one receiver input is the sum of all emitter and noise characteristics from the enlarged region. Because of the multivariant aspects and the associated calculations the evaluation of the interference can be accomplished with a general purpose, digital computer.

4.2.1 GENERAL TECHNIQUE

The interference analysis technique is referenced to a measure of the message information at the receiver output in the absence of interfering signals, $I\{\zeta\}\,|\,\tilde{x} = 0$, with actual receiver output to the information user designated as $I\{\zeta\}$. Interference effects are defined as

$$\text{Interference} = \frac{I\{\zeta\}}{I\{\zeta\}\,|\,\tilde{x} = 0} \qquad (2\text{-}30)$$

The set $\tilde{x} = (x_1, x_2, \ldots, x_n)$ has been defined as all separable interfering signals in the receiver after pickup and linear noninteracting amplification of the electromagnetic radiation present at antenna $(E_1, E_2, \ldots, E_n)$. Also, the set $\tilde{x}$ may be categorized into subsets $(x_1, x_2, \ldots, x_n)$ such that each member of the subset has identical interfering effects. The criteria for assigning a particular signal x_j to a subset x_i is that the inclusion of this signal will not change the character of the interference due to subset x_i, but only increase the magnitude. Rules for assigning signal to a subset will, of course, be a function of the particular receiver to be used. For example, it may be assumed in most cases that Gaussian noise from all sources will form one subset.

Each incoming signal at the receiver $E_j(\theta, \varphi, f, t)$ is related to its source by a transmission path and propagation losses. Although the nature of the emitter will primarily determine classification into subset at the detector input, the transmission path from emitter to receiver is a convenient criteria for classification of input radiation. Two distinct transmission paths have been discussed in previous sections; the first involves reflection from the satellite, and is thus primarily a function of satellite position relative to both emitter and receiver. The second main transmission path is directly line-of-sight from source to receiver, and may be a function of antenna position. Since the communication antenna will track the satellite, line-of-sight interference spatial radiation characteristic changes in a somewhat random time function, and it is necessary to apply statistical analysis methods. However, the statistical approximation is very much improved if a number of emitters may be grouped together for analysis. Two conditions must be satisfied for an averaging of this nature. First, the radiated signal characteristics of these emitters must all be members of one subset x_i at the detector. Second, it must be possible to describe an average density of emitters with an average directional radiation pattern from the surface of the earth. This second condition is essentially the replacement of many emitters in a surface area by an average emitter with an average spatial radiation characteristic, such that the probable analysis error is not excessively large.

The computer simulation interference analysis follows the factors discussed above. In general, radiation at the receiver is determined from (1) emitter characteristics, either individual or average, (2) transmission path and losses, and (3) receiver susceptibility to the interfering signals.

4.2.2 COORDINATE SYSTEMS

The reference coordinate system to be used in spherical where the fixed reference point, called the origin, is the center of the earth. The set of space coordinate is given as ρ, α, and β. ρ is defined as the positive straight line distance from the origin to the point in question. α is defined as the longitudinal angle in the equatorial plane, to the point in question, representing meridians going from the north to the south pole through the point in question and measured from the prime meridian (0°) that passes through Greenwich, England. β is defined as the latitudinal angle measured on the meridian arcs from the equatorial plane to the point in question.

The satellite orbital position, as well as the emitter and receptor sites, are reported in this coordinate system. Since this is a cumbersome system for reporting antenna radiation characteristics each emitter or receptor antenna is considered as an origin for a subspherical coordinate system.

The common system employed at antenna sites defines the azimuthal angle θ as the angle in the tangential plane to the earth's surface at the antenna and measured from the north-south meridian of the antenna site, and the elevation angle φ as the angle measured from the tangential plane to the point in question.

4.2.3 SPECIFIC COMPUTATION FOR SATELLITE COMMUNICATION INTERFERENCE

Section 4.1 has defined interference sources and parameters. The collection of these parameters in flow diagram form is contained in the following subsections. Interference caused by system components is not separately modeled since this type of interference is essentially the same as that resulting near local sources not related to the system. Detailed modeling of antenna noise and man-made noise is not provided since these inputs are direct receive antenna inputs and do not require the detailed simulation demanded by the other types of interference.

4.2.3.1 Satellite Interference

Figure 2-43 shows a general flow diagram where each block represents the parameters of the propagation of energy from one emitter to one satellite to one receiver's detector. Block (1) depicts the transmission of energy from the emitter in the direction of the satellite as a function of the emitter site, frequency spectrum, and time plus satellite position and time function. Block (2) is the inclusion of the independent variables, time, and frequency. Blocks (4) and (6) are the fixed position parameters of the emitter and receiver. Block (5) is the orbiting repeater position as a function of time. Block (3) represents the translation from the reference coordinate system to the site coordinate system. The possible propagation losses due to beaming toward the horizon as a function of elevation angle is simulated in Block (9). The signal at the receiver is given by Equation (2-8) in Section 4.1.1.1.a:

$$S_R = \frac{P_T G_T A_R a^2}{16 \pi R_1^2 R_2^2} \qquad (2\text{-}31)$$

where: P_T = transmitter power

G_T = transmitter antenna gain

A_R = effective area of receiver

a = radius of repeater

R_1 = transmitter to repeater distance

R_2 = repeater to receiver distance

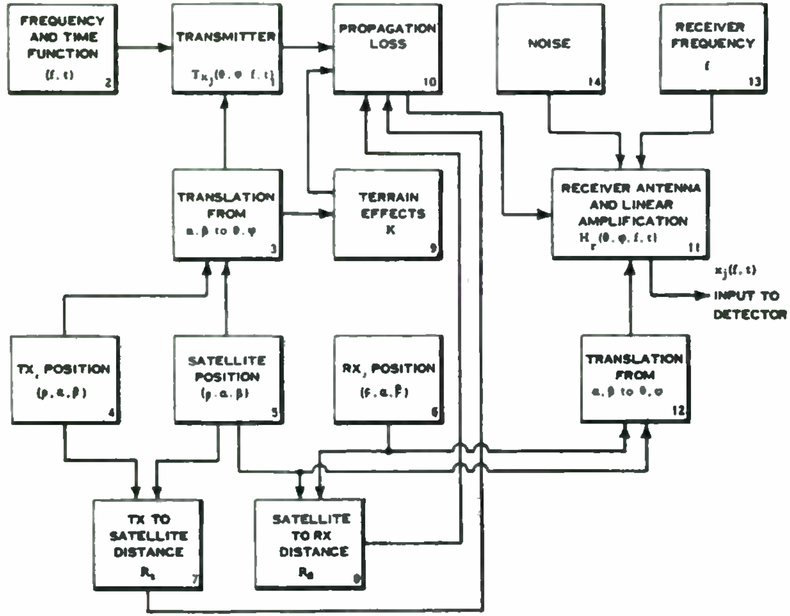


Figure 2-43. Satellite Interference Flow Diagram

Propagation loss is a function of transmitter to repeater to receiver distance; therefore, Blocks (7) and (8) simulate these distances at an instant in time. Block (10) depicts the propagation losses as functions of the above parameter. Block (9) supplies the variation in propagation loss due to scintillation, Faraday rotation, and atmospheric absorption.

The linear portions of the receiver site are represented by Block (11). Antenna position information provided by Block (12) is obtained in turn by the correct operation of positions contained in Blocks (5) and (6). Other required inputs to Block (11) are Blocks (13) and (14) which provide the noise input to the receiver and the frequency at which interference is being evaluated.

The output of Block (11) is the signal strength at the input to the nonlinear detection device. The flow diagram, therefore, represents the total signal characteristic through all linear components of transmission propagation and receptors.

4.2.3.2 Interference From Electromagnetic Sources Having a Direct Propagation to the Receiver

Figure 2-44 shows a general flow diagram where again each block represents the parameters of the propagation of energy from one emitter to the receiver by direct propagation. Block (1) depicts the transmission of energy from the emitter in the direction of the receiver (or obstacle in the case of a reflection circuit) as a function of the emitter site, frequency spectrum, and time.

Block (2) is the inclusion of the independent variable time and frequency. Blocks (4) and (5) are the fixed position parameters of the emitter and receiver (or obstacle). Block (3) represents the translation from the reference coordinate system to the site coordinate system. The possible propagation losses due to terrain effects as a function of the direction of propagation and description is simulated in Block (6). Propagation losses are a function of transmitter to receiver (or obstacle) distance; therefore, Block (7) simulates this distance. The simulation thus far depicts Equation (2-13) previously discussed in Section 4.1.2.1.b.(1). This equation, although used to define the propagation characteristics of a reflection circuit, is applicable to all other modes of propagation. Equation (2-13) is given as follows:

$$P_{RR} = K \frac{H(f, \theta, \varphi)}{4 \pi R_1^2} \cdot \gamma \qquad (2\text{-}13)$$

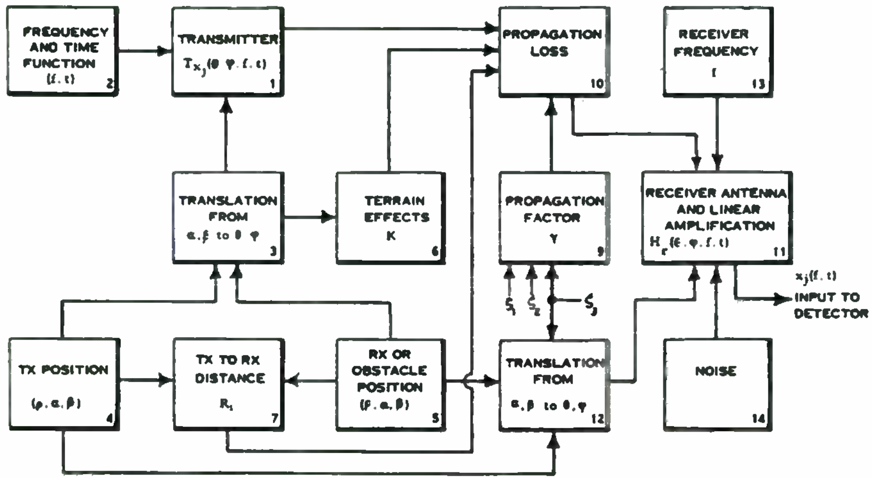


Figure 2-44. Environmental Interference Flow Diagram

where: R = transmitter to receiver distance or obstacle in the case of a reflection circuit

γ = the propagation factor that provides loss in excess of the free space loss.

All other parameters have been previously defined. This equation does not contain the receiver function since it is the formulation of power density at the receive antenna. The receiver characteristics, i.e., Blocks (11), (13), and (14) have been adequately explained in the preceding section and, therefore, will not be discussed further except as required by the inputs from Block (12)

In reflection circuits, Block (5) is the obstacle position. The R_1 calculation of Block (7), therefore, becomes the distance to the obstacle. The distance from the obstacle to the receiver is necessary and, therefore, ζ inputs to Block (9) must contain obstacle and receiver position. The receiver position in turn must also be entered in Block (12) along with obstacle position for computation of necessary angles. In addition to the position, coordinate γ also requires ζ inputs of reflectivity, degree of roughness, frequency, and obstacle orientation.

In the case of ducting, Block (5) is receiver position and R_1 actual distance. The ζ inputs now include refractive index conditions (meteorological) and probably receiver and transmitter positions for the increased propagation loss due to distance.

Tropospheric scatter propagation requires ζ input from Block (7). In addition, horizon angles in the direction of propagation must be provided by the ζ inputs. Frequency is another input that is required.

Diffraction requires the ζ inputs to be curvature of edge, path length, frequency, and horizon angles along the great circle path from both sites.

In general, with the exception of ducting, the mode of propagation that provides the greatest interfering level can be determined simply by an investigation of path topography. The levels of signal provided by diffraction and tropospheric scatter modes of propagation are predictable with a fair degree of confidence.

4.2.3.3 Detection and Decision Simulation

In the preceding sections, the computer simulation method is given for determining the set of signals $(\widetilde{x}, y) = (x_1, x_2, \ldots, x_n, y)$ at the input to the detector. Again, detector is defined as that section of the receiver following linear antenna and amplification stages. Since interaction between signals will very probably occur in the detector, it is necessary to include the complete set of signals in analysis.

The analysis and simulation procedure is considerably simplified by classifying the signals into subsets, as discussed in Section 4.1.4.2. Criteria for classification will definitely depend upon actual detector characteristics for a particular receiver. Classification criteria should be established by analytical procedure, using communications theory methods, whenever possible. However, this method may not be feasible for certain complex interfering signals, and an experimental determination may be necessary.

The criteria for subset classification will undoubtedly be a complex function of spectral characteristics of the detector input signals, best handled by computer sorting techniques. At the same time, mean and variance of the subset characteristics may be computed.

A measure of the interference effect of each subset is required to complete interference analysis. This requires an extension of the criteria classification discussed above, since a quantitative measure is required. Unless the number of subset parameters to be considered is quite small, digital simulation is the most convenient method of calculation.

4.2.3.4 Summary of Interference Analysis

The interference situation in the postulated passive satellite communication system includes a very large number of potential interference source emitters; in addition, the receiver will very probably be high gain with sophisticated detection circuitry. The large number of parameters involved virtually necessitates a computer analysis in addition to statistical averaging of emitters and signals.

Briefly reviewing, a set of signals $(\widetilde{x}, y) = (x_1, x_2, \ldots, x_n, y)$ at the input to the detector of the communications receiver is formulated from interfering transmitter, propagation, and receiver input characteristics. Each member x_j of the set corresponds to a particular transmitter (individual, average, or noise), and is computed from the transmitter signature, propagation, and receiver antenna and radio frequency section parameters. The set is classified into subsets $(x_1 x_2, \ldots, x_n)$ with each subset having the same interference effects at the output.

Final calculation of interference, as a function of applicable parameters of interfering signals, will largely depend upon the form of transmitted data. Interference has been defined as the ratio of information received in the presence of interfering signals to maximum received information if no interference were present. This definition separates actual interference effects from information loss due to other factors in the receiver.

A speech communication system may be used as an example. Suppose speech output with no interference yields 100% word intelligibility, and that word intelligibility is a satisfactory measure of information output. Interference signals, in this example, may produce noise at the output, with a noise spectrum which may be calculated from known interference. Interference may then be expressed by word intelligibility as a function of noise spectrum, and thus as a function of interfering signals. Actual interference as a function of time or position of satellite can then be predicted. However, it may be necessary to experimentally measure word intelligibility as a function of particular noise spectrums.

This example should illustrate the type of results to be expected from an interference analysis. The inclusion of such an analysis in design and development will lead to a successful and useful satellite communication system.

1. SHIELDING

The use of shielding is an important method of reducing or eliminating radio frequency interference. In fact, without the use of shielding, it is frequently impossible to reduce radio frequency interference to acceptable levels. Closely allied with shielding are bonding and grounding methods. Careful application of shielding, bonding and grounding techniques is essential regardless of any other RFI suppression measures. In this section, the theory, practice and measurement of shielding is discussed in detail. Sections 2 and 3, which follow, contain information pertaining to bonding and grounding. It is important to note that most design and installation considerations overlap in the three areas. Thus, equipment and system engineering for minimum interference must treat shielding, bonding and grounding from an overall view if optimum results are to be obtained.

1.1 FUNDAMENTALS OF SHIELDING

The interference caused by the radiation of electromagnetic energy to or from an equipment can be reduced by the use of a shield which reflects and/or absorbs this energy. An electromagnetic shield may be defined as a thin metallic sheet or mesh interposed in a transmission path. All practical shields are made of metals of high conductivity and are usually designed to enclose a source of potential interference completely or to exclude existing interference. Three types of shields commonly used are the solid metal covering, the flexible metal conduit, and the shielded cage. The attenuation of an electromagnetic wave in a mesh is considerably less than that of a solid cover. In some applications, double shielding may become necessary in order to prevent the pickup of undesirable signals.

A satisfactory shielded enclosure should have a shielding effectiveness of 50 to 100 db depending upon the intensity of the undesired signals present and the type of electromagnetic fields. The designer should strive for maximum shielding effectiveness within the limitations of weight, size, mechanical rigidity, and cost.

Certain types of special-purpose shields are used to reduce the coupling of electric or magnetic fields to or from an equipment. This may be done to reject an undesired field, or to reduce one type coupling without affecting another.

The shielding definition admits of division into three specific aspects: electromagnetic, magnetic, and electrostatic shielding.

1.1.1 ELECTROMAGNETIC SHIELDING

Since electromagnetic radiation fields have both electric and magnetic components, shielding requirements are severe because appreciable currents can flow across gaps which may act as radiating antennas. At high frequencies it is necessary to attain almost complete watertightness of a shield before the leakage is brought to an adequate level. A more efficient approach to the problem of reduction of radiation coupling is to attempt to contain the radiation within the equipment by use of multiple shielding and by providing suitable filters for all leads and shafts entering the enclosure.

1.1.2 MAGNETIC SHIELDING

The shielding requirements necessary to prevent the coupling of alternating magnetic fields become more difficult to meet at low frequencies than at frequencies above several megacycles. Such shields must have either high conductivity or high permeability. In the normal power-frequency range, for example, copper must be very thick to be a practical magnetic shield. Mu-metal and similar-type high-permeability alloys will provide good shielding for weak fields. Because of this, multiple magnetic shielding is recommended for strong fields where reduction of coupling is required. Power transformers and audio transformers mounted near each other may require multiple shielding to prevent magnetic coupling and minimize interference.

1.1.3 ELECTROSTATIC SHIELDING

Electrostatic shielding is used to prevent stray capacitance coupling between two circuits. A good ground is essential for effective operation of this type shielding. Electrostatic shielding should be used in high-impedance circuits where static fields may exist due to high potential. Power transformers, audio tranformers, and input transformers of receivers require such treatment. An example of this type of shielding, which might be used in a transformer circuit, is shown in Figure 3-1. A Faraday shield is placed between the transformer windings and is used to prevent capacitance coupling without affecting the inductance coupling. This type of shield is a set of grounded magnetic prongs, arranged somewhat like the teeth of a comb. Since the prongs are not connected at one end, no induced currents can flow through them and the

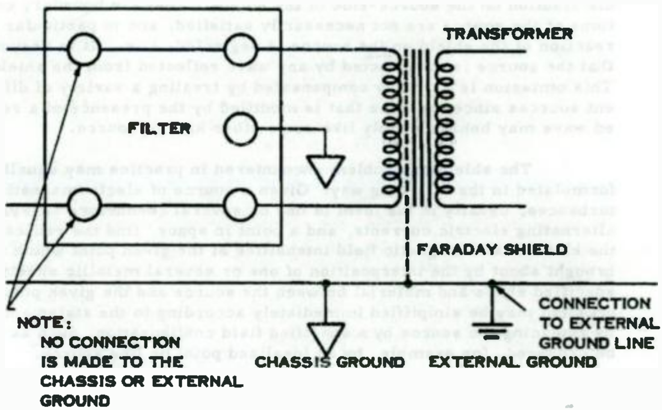


Figure 3-1. Power Transformer with Electrostatic Shield and Filter

magnetic coupling is not affected. However, the prongs are so close together that their plane is essentially equipotential and, therefore, no electrostatic coupling can exist between the windings.

1.2 THEORY OF SHIELDING

An exact solution to any shielding problem can be obtained only by solving a so-called boundary-value problem. This means that a solution to Maxwell's equations, which govern the behavior of all electromagnetic fields, must be found which satisfies the boundary conditions imposed by the metallic surfaces both of the source of the electromagnetic field and of the conductors used as shields. Boundary-value problems belong to the most difficult problems of electrical engineering, and usually an effort is made to find an approximate solution to a practical shielding problem by other means. A very fruitful approach is that of treating the problem as a boundary-value problem only for the conductors used as shields and making some kind of reasonable assumption about the field distribution on the source-side of the shield. Then the boundary conditions at the source are not necessarily satisfied, and in particular the reaction of the shield on the source is neglected, i. e., it is assumed that the source is not affected by any wave reflected from the shield. This omission is partially compensated by treating a variety of different sources since a source that is modified by the presence of a reflected wave may behave simply like some other kind of source.

The shielding problem encountered in practice may usually be formulated in the following way: Given a source of electromagnetic disturbances, usually in the form of one or several conductors carrying alternating electric currents, and a point in space, find the reduction in the electric and magnetic field intensities at the given point which is brought about by the interposition of one or several metallic sheets of specified shape and material between the source and the given point. The problem may be simplified immediately according to the statement above by replacing the source by a specified field configuration, such as would be produced, for example, by an idealized point or line source.

Even after this first simplification, the general problem is still much too complicated to allow a simple solution. If the shape of the shield is at all irregular, the analytical approach becomes all but hopeless. Only the three simplest regular shapes will be treated here, and in each case it will be assumed that the fields exhibit the same kinds of symmetries as the shields themselves. The cases to be treated are that of a plane shield of infinite extent in the presence of plane waves (Section 1. 2. 1), that of a shield in the form of a circular cylinder in the presence of cylindrical waves concentric with the shield (Section 1. 2. 2), and that of a spherical shield in the presence of spherical waves concentric with the shield (Section 1. 2. 3). While none of these three cases is exactly duplicated in practice, almost every practical case can be closely approximated by one or a combination of several of these. Therefore, the results that will be obtained here are of considerable practical importance despite their restrictions.

In any shielding problem, a valuable aid is offered by the analogy between the propagation of electromagnetic waves through a medium and the transmission of electrical energy through a transmission line. This analogy has been extensively exploited in the literature, and it is especially valuable to the engineer who is much more familiar with transmission-line theory than with propagation phenomena. It is found that such familiar concepts as characteristic impedance, propagation constant, reflection and transmission factors, attenuation, and phase shift, all have their counterparts in the propagation of electromagnetic waves. Some of these require re-definition with a slight change in meaning, but once the new definitions have been introduced properly, all the laws of transmission-line theory can be applied directly. For example, just as the characteristic impedance of a transmission line is defined as the ratio of voltage to current along a line on which no reflected wave is present (i.e., along a line that is either infinitely long or terminated in its characteristic impedance), so the "intrinsic impedance" of a medium is defined as the ratio of electric to magnetic field intensity in a medium in which no reflected wave is present. In general, the analogy requires that the electric field intensity in the medium be substituted for the voltage along the transmission line, and that the magnetic field intensity be substituted for the current. It should be noted that the units for electric field intensity are volts per meter or volts per inch and those for magnetic field intensity are amperes per meter or amperes per inch, so that their ratio, the intrinsic impedance, is measured in ohms just as is the characteristic impedance of a transmission line. The units of length drop out when the ratio is taken.

There is one complication in the extension of transmission-line concepts to the propagation of electromagnetic waves: The impedance is a function not only of the properties of the medium, but also of the type of wave that is being considered. It is true that the same complication arises also in transmission-line theory at high frequencies when modes of transmission other than the fundamental may be present. Then the impedances associated with the various modes will, in general, be different. But the presence of higher modes in transmission lines is a comparatively rare phenomenon in practice, and usually the term "transmission-line theory" refers to the behavior of the fundamental mode only. In the propagation of electromagnetic waves, on the other hand, it must be remembered that, at all frequencies, the impedance is a function not only of the mode, when several different modes exist, but also of the type of wave, i.e., the impedance is different for plane, cylindrical, and spherical waves.

1.2.1 PLANE WAVES

It is assumed that the field intensities are functions of one space coordinate only. The electric field intensity is assumed to have an x-component only and the magnetic field intensity to have only a z-component. The variation then takes place along the y-axis, which is also the direction of propagation. With these assumptions, Maxwell's equations reduce to the following form:

$$\frac{\partial E_x}{\partial y} = \mu \frac{\partial H_z}{\partial t} \qquad (3\text{-}1)$$

$$\frac{\partial H_z}{\partial y} = \sigma E_x + \frac{\epsilon \partial E_x}{\partial t} \qquad (3\text{-}2)$$

Where E_x and H_z are the electric and magnetic field intensities, respectively, μ is the permeability of the medium, σ its conductivity, ϵ its permitivity, x, y, and z are rectangular coordinates forming a right-handed system, and t is time.

Assuming sinusoidal time variations of all field intensities at the angular frequency $\omega = 2\pi f$, where f is the frequency, the solution to Equations (3-1) and (3-2) may be written in the following form:

$$E = E_1 e^{\gamma y} + E_2 e^{-\gamma y} \qquad (3\text{-}3)$$

$$H = \frac{E_1}{Z_0} e^{\gamma y} - \frac{E_2}{Z_0} e^{-\gamma y} \qquad (3\text{-}4)$$

where $E = E_x e^{j\omega t}$, $H = H_z e^{j\omega t}$, $j^2 = -1$, E_1 and E_2 are arbitrary constants which must be determined from the boundary conditions, and γ and Z_0 are defined as follows:

$$\gamma = \text{propagation constant of the medium} = \sqrt{j\omega\mu(\sigma + j\omega\epsilon)} \qquad (3\text{-}5)$$

$$Z_0 = \text{intrinsic impedance of the medium} = \sqrt{\frac{j\omega\mu}{\sigma + j\omega\epsilon}} \qquad (3\text{-}6)$$

Both σ and Z_0 are complex quantities, in general, so that one can write:

$$\gamma = \alpha + j\beta \qquad (3\text{-}7)$$

$$Z_0 = R_0 + jX_0 \qquad (3\text{-}8)$$

where α is called the attenuation constant and β the phase constant of the medium, and R_0 and X_0 are the resistive and reactive parts, respectively, of the intrinsic impedance.

For free space or air, $\sigma = 0$, $\mu = 4\pi \times 10^{-7}$ henries per meter, and $\epsilon = 8.85 \times 10^{-12}$ farads per meter. With these values substituted, the intrinsic impedance of free space or air is 376.6 ohms and its propagation constant is $j\omega/c = 2\pi/\lambda$, where $c = 3 \times 10^{8}$ meters per second is the velocity of light in free space and λ is the wavelength.

For all metals, the conductivity σ is much larger than the product $\omega\epsilon$ at all frequencies now used or likely to be used for radio communication purposes. Therefore, the term $j\omega\epsilon$ may be neglected and one obtains for metals

$$\gamma = \sqrt{\frac{\omega\mu\sigma}{2}}\,(1 + j) \tag{3-9}$$

$$Z_0 = \sqrt{\frac{\omega\mu}{2\sigma}}\,(1 + j) \tag{3-10}$$

$$\alpha = \beta = \sqrt{\frac{\omega\mu\sigma}{2}} \tag{3-11}$$

$$R_0 = X_0 = \sqrt{\frac{\omega\mu}{2\sigma}} \tag{3-12}$$

It is seen that the intrinsic impedance of metals is extremely small in comparison with that of air and most other dielectric media. On the other hand, both the real and the imaginary parts of the propagation constant are very large in comparison with those of dielectrics, which indicates a large attenuation and a small wavelength in metals.

In order to apply these results to shielding problems, the behavior of the wave at the boundary between two media must be investigated. Let it be assumed that a plane wave is propagated in air and impinges on a plane metal surface. For simplicity, choose a set of rectangular coordinates so that the boundary surface coincides with the plane $y = 0$. Let the region $y > 0$ be air and the region $y < 0$ be the metal. Let the incident wave be propagated in the direction of decreasing y, as shown in Figure 3-2. The problem is to find a set of solutions of the type given by Equations (3-3) and (3-4) which satisfy the boundary conditions, viz., the conditions that both E and H must be continuous everywhere, including the surface $y = 0$. The transmission-line analogy of this situation is shown in Figure 3-3: Two semi-infinite lines of different characteristic impedances are joined so that each line is terminated at one end in the characteristic impedance of the other.

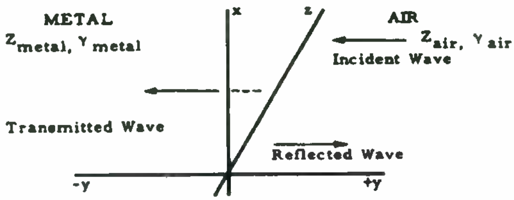


Figure 3-2. Plane Wave Striking Plane Boundary Surface Between Air and Metal

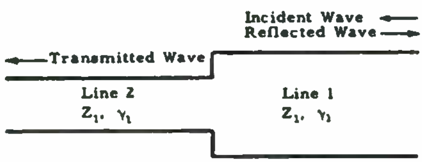


Figure 3-3. Transmission-Line Analogy of Plane-Wave Boundary Problem

The problem is solved by postulating the existence of a transmitted wave in the metal and a reflected wave in air traveling in the direction of increasing y, in addition to the incident wave. One must write:

$$E_{air} = E_1 e^{\gamma y} + E_2 e^{-\gamma y} \qquad (3\text{-}13)$$

$$E_{metal} = E_1' e^{\gamma y} \qquad (3\text{-}14)$$

$$H_{air} = \frac{E_1}{Z_{air}} e^{\gamma y} - \frac{E_2}{Z_{air}} e^{-\gamma y} \qquad (3\text{-}15)$$

$$H_{metal} = \frac{E_1'}{Z_{metal}} e^{\gamma y} \tag{3-16}$$

Here E_1 and E_1/Z_{air} are the amplitudes of the incident wave, E_2 and E_2/Z_{air} the amplitudes of the reflected wave, and E_1' and E_1'/Z_{metal} the amplitudes of the transmitted wave. E_1, E_2, and E_1' are three constants, two of which may be evaluated from the conditions that, at $y = 0$, E_{air} must equal E_{metal} and H_{air} must equal H_{metal}. The third, which is best taken as the amplitude of the incident wave, remains arbitrary, of course. The ratio of the amplitude of the electric field intensity of the reflected wave to that of the incident wave is called the reflection factor, F_r, and the ratio of the electric field intensity of the transmitted wave to that of the incident wave is called the transmission factor, F_t. Using the boundary conditions stated, one obtains after a short computation

$$F_r = \frac{E_2}{E_1} = \frac{Z_{metal} - Z_{air}}{Z_{metal} + Z_{air}} \tag{3-17}$$

$$F_t = \frac{E_1'}{E_1} = \frac{2\,Z_{metal}}{Z_{metal} + Z_{air}} \tag{3-18}$$

These expressions are identical with the defining equations for the reflection and transmission factors of a transmission line.

The quantity of interest here is $20 \log F_t$, which gives, in decibels, the attenuation experienced by the electric field intensity in entering the metal. It represents a reflection loss, due to the fact that the incident wave is only partially transmitted, the rest being reflected. It must be distinguished from the absorption loss within the metal, which will be discussed below.

A wave, in passing through an actual shield, experiences reflections at two boundary surfaces: once when it enters the shield, and again when it leaves the shield. The transmission factor for the second surface may be obtained from Equation (3-18) by replacing Z_{metal} with Z_{air} and vice versa. The reflection loss in this case is not equal to $20 \log F_t$ because the first medium does not extend to infinity. There will exist multiple reflections between the two surfaces of the shield, which will affect the transmission loss at the second surface. However, as will be seen presently, the absorption loss in the metal is so large that, for all practical shields, the effect of multiple reflections may be neglected.

The reflection factor was defined above as the ratio of the amplitudes of the electric field intensities. A similar factor might be defined as the ratio of the amplitudes of the magnetic field intensities. A simple calculation shows that the reflection loss for the magnetic field at the first surface is just equal to the reflection loss for the electric field at the second surface (neglecting multiple reflections), and vice versa. Thus, while the attenuation experienced by the magnetic field at any one surface is quite different from that experienced by the electric field (much smaller at the first and much larger at the second surface), the combined effect of the two surfaces is the same for both fields, as it must be in accordance with the assumption that the character of the wave does not change in passing through the shield and, therefore, that the impedances on both sides of the shield are equal.

As was shown before, in all practical cases Z_{metal} is much smaller than Z_{air} and may be neglected in the denominator of Equation (3-18) Using this approximation for both surfaces, one obtains for the combined reflection loss in decibels for either the electric or the magnetic field:

$$\text{Total reflection loss} = 20 \log (Z_{air}/2\,Z_{metal})$$

$$+ 20 \log (Z_{air}/2\,Z_{air}) = 20 \log (Z_{air}/4\,Z_{metal}) \qquad (3\text{-}19)$$

Here the ratio is inverted as compared to Equation (3-18) so that the reflection loss is positive when the wave is attenuated.

In addition to the reflection loss, there is an absorption loss within the metal. Equations (3-14) and (3-16) show that both the electric and the magnetic fields within the metal contain the propagation factor $\exp \gamma y = (\exp \alpha y)(\exp j\beta y)$. The second of these factors is a phase factor that does not affect the amplitude of the fields. The first factor shows that the amplitudes decrease exponentially within the metal. If the thickness of the shield is S, the amplitudes at the second surface are smaller than those at the first surface within the metal by a factor $\exp(-\alpha S)$. The negative sign arises because within the metal y is negative. Hence, the absorption loss is

$$\text{Absorption loss} = \alpha_{metal}\,S \qquad (3\text{-}20)$$

Equations (3-19) and (3-20) may be combined. After substituting numerical values from Equations (3-10) and (3-11) and converting to practical units, one obtains for the total loss in decibels:

$$\text{Total loss} = 3.34\sqrt{f_m \mu_r \sigma_r}\,S + 108.2$$

$$+ 10 \log (\sigma_r / f_m \mu_r) \qquad (3\text{-}21)$$

where f_m is the frequency in megacycles per second, μ_r the relative magnetic permeability of the shielding material ($\mu_r = 1$ for all non-magnetic materials), σ_r the relative conductivity of the shielding material ($\sigma_r = 1$ for copper), and S the thickness of the shield in mils. For example, the total loss experienced by a plane wave in passing through a plane copper sheet 5 mils thick at 1 mc would be 124.9 decibels. In Figure 3-4, the absorption loss, the reflection loss, and the total loss are plotted as functions of frequency for a plane copper sheet 5 mils thick. It is seen that at low frequencies the reflection loss is by far the larger, but at high frequencies the absorption loss becomes the larger. For a copper sheet 5 mils thick, the two losses are equal at about 30 mc, but for a thicker sheet, that point would occur at a lower frequency. The total loss has a minimum at about 0.3 mc, and the minimum loss is about 122 decibels.

Decreasing the conductivity decreases both the reflection and the absorption losses, so that the shielding material should always have as high a conductivity as possible. Increasing the relative permeability, i.e., using magnetic materials, increases the absorption loss but decreases the reflection loss. To obtain some idea of the effects of using magnetic materials for shielding purposes, Figure 3-5 shows the absorption loss, the reflection loss, and the total loss plotted as a function of frequency for a material having a constant relative permeability of 1000. Since all known magnetic materials have conductivities considerably lower than copper, the relative conductivity was given a representative value of 0.1. It is seen that the total loss now reaches its minimum at about 3 kc, and the minimum loss itself is about 103 db. Thus it appears that, for the example chosen, the magnetic material is preferable at the higher frequencies, but poorer at the lower frequencies. It must be remembered, however, that the material in the example chosen is quite thin (5 mils thick). For a thicker sample, the reflection loss remains the same, but the absorption loss is increased proportionally. Since the magnetic material has the higher absorption loss, it follows that, the thicker the shield, the more advantageous does the use of magnetic materials become.

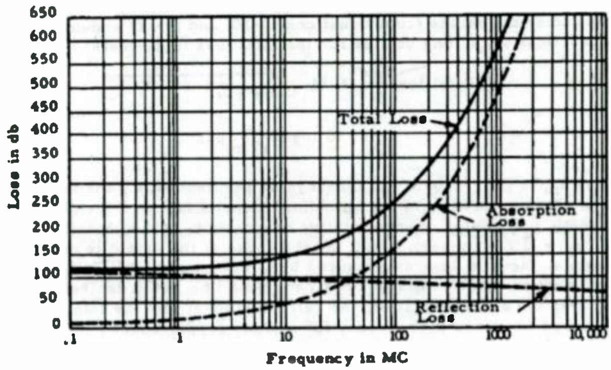


Figure 3-4. Absorption Reflection and Total Losses in 5 Mil Sheet of Copper

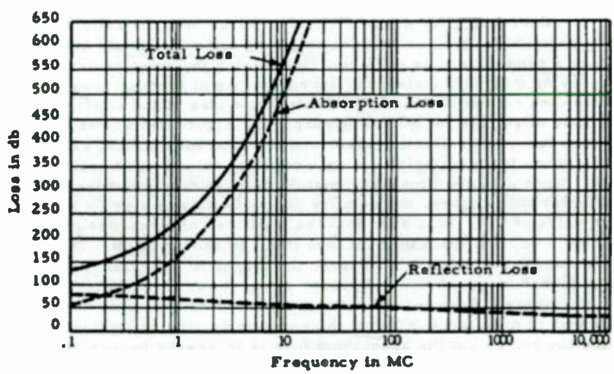


Figure 3-5. Absorption, Reflection, and Total Losses in 5 Mil Sheet of Magnetic Material. $\sigma_r = 0.1$, $\mu_r = 1000$

Another fact to remember is that the permeability of all magnetic materials decreases with frequency. For a typical sample having a permeability of 1000 at frequencies up to 10 mc, the permeability might decrease to 100 at 100 mc, to 10 at 1000 mc, and down to unity at 10,000 mc. Thus, the extremely high losses at the high frequencies shown in Figure 3-5 cannot be realized in practice, but they also are not usually required.

Equation (3-21) was derived on the basis of a plane wave striking a plane surface normally. If the direction of propagation of the wave makes an angle with the normal to the surface other than zero degrees, the situation is more complicated. It is then necessary to treat two cases separately: that of the electric field intensity being parallel to the surface, and that of the magnetic field intensity being parallel to the surface. In the first case one speaks of a transverse electric wave; in the second case of a transverse magnetic wave. In the case of normal incidence the two cases merge into one. The analysis is simplified by the fact that, no matter at what angle a plane wave strikes a plane metal surface, the refracted wave, which is transmitted into the metal, always travels in the direction normal to the surface. Therefore, the absorption loss within the metal and the reflection loss at the second surface are independent of the angle of incidence. The detailed analysis is not carried out here, but it can be shown that the reflection loss at the first surface for oblique incidence is never less than for normal incidence, so that Equation (3-21) gives a total loss which is at worst too low.

Finally, it must be pointed out that Equation (3-21) is not valid in practical cases for either very small values of S or very small values of f. This is evident from the fact that, if one sets $S = 0$, Equation (3-21) still indicates a reflection loss. Actually, $S = 0$ means that there is no shield, and, therefore, there can be no loss. Also, setting $f = 0$, one finds an infinite reflection loss, while it is known that shields are very ineffective for static magnetic fields. The reasons for these discrepancies are not immediately obvious.

Since multiple reflections within the shield were neglected in the derivation of Equation (3-21), one might think that this omission causes the equation to cease being valid for small values of S. This, however, is not the case. Analysis shows that Equation (3-21) is theoretically correct, even when multiple reflections are considered, for values of S down to the order of 10^{-6} mils. This is a distance of the order of magnitude of an atomic diameter. Hence, Equation (3-21) is theoretically valid even for a shield consisting of a monatomic or monomolecular layer of metal. Yet, it is known that very thin shields, such as coatings of metallic paints, are sometimes very ineffective. The fallacy lies in the fact that Maxwell's equations themselves, with their macroscopic concepts of conductivity, permeability, and dielectric constant, cease to be valid in the realm of atomic or molecular phenomena. In other words, Equation (3-21) is valid only as long as one deals with shields that are thick enough to allow the application of concepts, such as conductivity, that require the presence of a very large number of atoms or molecules. While no definite limit can be set, it should be expected that Equation (3-21) will begin to break down for thicknesses less than about 0.01 or 0.001 mils.

As to the application of Equation (3-21) to very low frequencies, there is nothing in its derivation that would indicate that it might cease to be valid for any communication frequency. And, indeed, it does remain valid even for very low frequencies for the conditions under which it was derived: the presence of plane waves. In practice, any wave becomes approximately plane at a distance of several wavelengths from the source. At very low frequencies the wavelength may be several miles, and it is usually quite impossible to even approximate a plane wave. Hence, at power frequencies and, in the limit, at zero frequency, plane waves never exist, in practice, and Equation (3-21) is not applicable. In addition, Maxwell's equations remain satisfied if any arbitrary constant static field is added to the solution given by Equations (3-3) and (3-4). This constant field was neglected entirely in the treatment given above. Hence, the results are not applicable to a static field except in the special case of a static field obtained from a plane wave in the limit as the frequency goes to zero, which case is never met in practice.

1.2.2 CYLINDRICAL WAVES

Practical cases involving cylindrical symmetry are encountered frequently. Out of the many possible cases, only a very small number will be investigated here. The most important examples are: (1) electromagnetic waves traveling axially in a cylindrical enclosure either with an internal conductor (coaxial transmission line) or without (circular waveguide); (2) electromagnetic waves traveling radially and originating from a centrally located line current source; and (3) electromagnetic waves traveling radially and originating from a centrally located line loop source. These cases do not normally occur separately. Usually there exists a superposition of several occurring simultaneously. The first case is best attacked from the point of view of surface transfer impedance and will not be treated here; this case is discussed in Section 1.4.

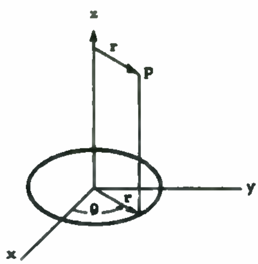


Figure 3-6. Cylindrical Coordinates

For the second and third cases, it is assumed that the fields have circular symmetry and do not vary along the axis of the cylinder. Consider the cylindrical coordinate system shown in Figure 3-6. It is assumed, then, that the fields are functions of r only, independent of z and θ. With these assumptions, Maxwell's equations read:

$$\frac{\partial E_z}{\partial r} = \mu \frac{\partial H_\theta}{\partial t} \tag{3-22}$$

$$\frac{\partial}{\partial r}(r H_\theta) = r(\sigma E_z + \epsilon \frac{\partial E_z}{\partial t}) \tag{3-23}$$

$$\frac{\partial H_z}{\partial r} = -(\sigma E_\theta + \epsilon \frac{\partial E_\theta}{\partial t}) \tag{3-24}$$

$$\frac{\partial}{\partial r}(r E_\theta) = -\mu r \frac{\partial H_z}{\partial t} \tag{3-25}$$

It is seen that this set naturally splits up into two pairs: the pair involving E_z and H_θ, Equations (3-22) and (3-23), and the pair involving E_θ and H_z, Equations (3-24) and (3-25). Each pair may be treated separately.

Setting $E_\theta = H_z = 0$, one is led to case (1) above: the fields act as if they were produced by a source in the form of an infinitely long and infinitely thin conductor at the center carrying a current that is everywhere in phase. It will be seen that such a field, in the vicinity of the source, is associated with a very low impedance, and therefore, this case will be called the "low-impedance case." If one sets $E_z = H_\theta = 0$, case (2) results: the fields act as if they were produced by an infinite number of very small, closely spaced coaxial loops carrying equal and uniform currents that are everywhere in phase. This field is associated with a high impedance near the source and will be called the "high-impedance case."

Just as in the case of plane waves, a static field may be superimposed on the dynamic fields. Such a field might be produced by a static line charge at the center and would vary as $1/r$. The possible presence of this field will be neglected here.

The solution to the pair of Equations (3-22) and (3-23) may be written as follows, assuming sinusoidal time variations as before:

$$E_z = E_1 H_0(j\gamma r) + E_2 H_0'(j\gamma r) \tag{3-26}$$

$$H_\theta = -\frac{j}{Z_0}\left[E_1 H_1(j\gamma r) + E_2 H_1'(j\gamma r)\right] \tag{3-27}$$

These equations are very similar to Equations (3-3) and (3-4) for plane waves except that the exponential functions are replaced by the functions H_0, H_0', H_1, and H_1'. These functions are the Hankel functions of first and second order and of the first and second kind. The subscripts refer to the order of the Hankel function; the unprimed symbols are Hankel functions of the first kind, and the primed symbols Hankel functions of the second kind. The Hankel functions of the first kind represent waves traveling radially inward, while the Hankel functions of the second kind represent waves traveling radially outward.

The impedance is again found as the ratio of the electric to magnetic field intensity, but a complication arises due to the fact that the impedances for the incoming and outgoing waves are not equal. One obtains from Equations (3-26) and (3-27) treating the incoming and outgoing parts separately:

$$Z_1 = j Z_0 H_0(j\gamma r) / H_1(j\gamma r) \tag{3-28}$$

$$Z_1' = jZ_0H_0'(j\gamma r) / H_1'(j\gamma r) \tag{3-29}$$

where Z_1 and Z_1' are the radial impedances for the incoming and outgoing waves, respectively, in the low-impedance case. It is seen that these impedances are functions of r.

If the same analysis is carried through for the high-impedance case, starting with Equations (3-24) and (3-25), the following set of impedances is obtained:

$$Z_2 = jZ_0H_1(j\gamma r) / H_0(j\gamma r) \tag{3-30}$$

$$Z_2' = jZ_0H_1'(j\gamma r) / H_0'(j\gamma r) \tag{3-31}$$

where Z_2 and Z_2' are the radial impedances for the incoming and outgoing waves, respectively, in the high-impedance case.

An extensive analysis of these impedances requires a detailed consideration of the behavior of the Hankel functions. This will not be carried out here. Great simplifications result if the absolute value of the argument $j\gamma r$ is either very large or very small. If it is large, i.e., at large distances from the center of the cylinder, all impedances approach the value of Z_0 for plane waves. Hence, as might have been expected, cylindrical waves behave like plane waves at large distances from the center. When the absolute value of $j\gamma r$ is very small, the following approximations may be used:

$$\left.\begin{aligned} H_0(x) &\to 1 + (2j/\pi)\ln|x| \\ H_0'(x) &\to 1 - (2j/\pi)\ln|x| \\ H_1(x) &\to -2j/(\pi x) \\ H_1'(x) &\to 2j/(\pi x) \end{aligned}\right\} \text{ as } x \to 0 \tag{3-32}$$

With these approximations, one obtains for the four impedances in the case of small values of $j\gamma r$:

$$Z_1 = -(1/2)j\gamma r Z_0(\pi + 2j\ln|j\gamma r|) \tag{3-33}$$

$$Z_1' = -(1/2)j\gamma r Z_0(\pi - 2j\ln|j\gamma r|) \tag{3-34}$$

$$Z_3 = -\frac{2 Z_0}{j\gamma r(\pi^2 + 4 \ln^2 |j\gamma r|)} (\pi - 2j \ln |j\gamma r|) \qquad (3\text{-}35)$$

$$Z_3 = -\frac{2 Z_0}{j\gamma r (\pi^2 + 4 \ln^2 |j\gamma r|)} (\pi + 2j \ln |j\gamma r|) \qquad (3\text{-}36)$$

It is seen that the impedances for the incoming and outgoing waves are negative complex conjugates of each other in each case. It is also seen that, for small $j\gamma r$, the first two impedances are much smaller than Z_0 while the second set of two impedances are much larger than Z_0, thus justifying the nomenclature used.

Now let there be a metallic medium filling all space outside a cylindrical region such that for $r < a$ the medium is air and for $r > a$ the medium is a good conductor as shown in Figure 3-7. The surface $r = a$ is the boundary surface between the two media, and the task at hand is to determine how much of the incident wave, originating at the center, is transmitted into the metal.

As in the case of plane waves, the presence of an incident and a reflected wave must be postulated on the inside together with a transmitted wave within the metal. To obtain a general expression for the transmission factor, taking into consideration the fact that the impedances for outgoing and incoming waves need not be equal, the following equations may be written based on the continuity of the electric and magnetic fields across the boundary surface:

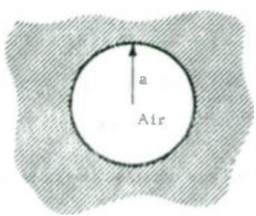


Figure 3-7. Cylindrical Boundary Surface Between Air and Metal

$$E'_{metal} = E'_{air} + E_{air} \tag{3-37}$$

$$\frac{E'_{metal}}{Z'_{metal}} = \frac{E'_{air}}{Z'_{air}} - \frac{E_{air}}{Z_{air}} \tag{3-38}$$

where the primed quantities refer to the outgoing waves and the unprimed ones to the incoming waves. In addition, the subscripts 1 or 2 must be added depending on whether the low-impedance case or the high-impedance case is dealt with.

The above equations may be solved for the transmission factor, E'_{metal}/E'_{air}:

$$F_t = \frac{Z'_{metal}}{Z'_{metal} + Z_{air}} \left(1 + \frac{Z_{air}}{Z'_{air}} \right) \tag{3-39}$$

It is seen that this reduces to Equation (3-18) when the primed and unprimed impedances are equal.

When the quantity $|\gamma|$ is computed for a metal from Equation (3-9), one finds $|\gamma| = 545 \sqrt{f_m \mu_r \sigma_r}$ per inch. Hence the absolute value of the quantity $j\gamma r$ will be much larger than unity in most practical applications. Therefore, within the metal cylindrical waves behave like plane waves. In particular, the absorption loss within the metal is the same as that obtained for plane waves, Equation (3-20), and the impedance of the metal to cylindrical waves is the same as for plane waves. An exception to this might occur only when the quantity $r \sqrt{f_m \mu_r \sigma_r}$ (r in inches) is less than, say, 0.1, and even then the assumption of plane waves will be a fair approximation down to $r \sqrt{f_m \mu_r \sigma_r} = 0.01$, approximately. The exceptional case in which this might not be true will not be treated here.

In evaluating the four impedances for air from Equations (3-33) to (3-36), it is found that Z_1 and Z'_1 (low-impedance case) are very small while Z_2 and Z'_2 (high-impedance case) are very large as compared with Z_0 in the region of validity of these equations. A more detailed computation of these quantities is hardly justified since the physical conditions on which the derivation of these equations was based are highly idealized.

In practice, the presence of an inner conductor imposes additional boundary conditions, which were neglected here. However, the following qualitative conclusion drawn from the form of Equation (3-39) remains valid: In the high-impedance case the reflection loss is at least as great as with plane waves; but in the low-impedance case the reflection loss may be very small and cannot be relied upon for effective shielding action.

The practical conclusions for cylindrical shields may be summarized as follows:

a. Within the shielding metal, the cylindrical waves behave like plane waves. The absorption loss may be computed using Equation (3-20) as for plane waves.

b. If the electric field near the source at the center is predominantly axial, the radial impedance is low, and very little shielding action may be expected due to reflection.

c. If the electric field near the source at the center is predominantly concentric, the radial impedance is high, and the shielding action due to reflection is at least as good as that for plane waves.

d. In the absence of detailed information about the fields near the source, it is best to assume that a substantial low-impedance component is present, and to proceed on the assumption that shielding action is due to absorption only.

1.2.3 SPHERICAL WAVES

Practical cases involving spherical symmetry are very rare in shielding problems. If the same approach is taken in this case as was taken for cylindrical waves, similar results are obtained except that the resulting functions are spherical Hankel functions, or Hankel functions of half-integral order, instead of the cylindrical Hankel functions H_0, H_1, H_0', and H_1'. This analysis, however, is not carried out here.

If the radius of the sphere is very much larger than a wavelength, the waves behave exactly like plane waves. If the radius of the sphere is very much smaller than a wavelength, the conclusions a to d at the end of the section on cylindrical waves remain essentially valid. The intermediate case, where resonant excitation of the spherical enclosure may occur, is again too complicated to be treated here. The main reason why spherical waves are mentioned here is that the conclusions just mentioned remain valid qualitatively when a point source is shielded by an enclosure of other than spherical shape, as for example when a spark gap, acting like a point source, is enclosed by a rectangular shielding box. Here, an exact solution of the boundary problem involved is practically impossible. Yet, it may be said that the shielding effectiveness of the box is at least as great as that computed on the basis of the absorption loss for plane waves, and may be considerably increased by the reflection loss if the source is of the high-impedance type.

1.3 SHIELDING DESIGN CONSIDERATIONS

The purpose of a shield is to keep all radio interference energy "bottled up" within a specified region, or to prevent all radio interference energy from entering a specified region. The first type is used for ignition systems, motors, and other sources of radio interference. The second type is used for receivers or leads leading to receivers. Because power for control energy must always be supplied or removed from the region within the shield, and because the techniques of construction as well as the necessity for accessibility and serviceability demand that shields be made of more than one part, openings, seams, joints, or other discontinuities must always be present. The problem of constructing an effective shield has therefore two separate phases. One is the prevention of the penetration of electromagnetic energy through the shielding wall itself, and the other is the prevention of leakage through the discontinuities in the shield. The second of these two problems - the proper design of the necessary discontinuities so that effectiveness of the entire shield is not impaired - requires the greater consideration and attention. Just as a chain is no stronger than its weakest link, a shield is no more effective than its poorest joint. The major portion of the following paragraphs is devoted to the principles and techniques employed in the design of discontinuities in shields for minimum leakage.

1.3.1 SHIELDING MATERIALS

The problem of preventing penetration through the shielding wall itself is comparatively simple. As was pointed out in Chapter 1 under certain simplified conditions the ratio of the electromagnetic energy that has penetrated a shield to that entering it, expressed in decibels, varies inversely as the thickness and the square root of the magnetic permeability and directly as the square root of the resistivity. While shielding effectiveness depends on other factors, such as the impedance of the wave and the geometrical shape of the shield, in a very complicated way, this dependence can be entirely neglected provided that the three factors mentioned above are chosen large (or small) enough. This leads to walls that may be much thicker than necessary for the desired degree of shielding, but usually not thicker than necessary for mechanical reasons whenever the shield must support itself mechanically. In the absence of more detailed information about the effect of the other factors, it is suggested that the minimum thickness of shielding material be based on an absorption loss of about 33 db at 1 mc. The absorption loss in decibels is given by Equation (3-21):

$$\text{Absorption Loss} = 3.34\,S\,\sqrt{f_m \mu_r \sigma_r} \tag{3-40}$$

where S is the thickness in mils, f_m is the frequency in megacycles per second, μ_r is the relative magnetic permeability ($\mu_r = 1$ for all non-magnetic materials) and σ_r is the conductivity relative to copper (see Figure 3-8). Hence, the thickness for about 33 db at 1 mc is given by:

$$S = \frac{10}{\sqrt{\mu_r \sigma_r}} \tag{3-41}$$

This leads to a minimum thickness of 10 mils for copper - a choice that has proved satisfactory in practice.

When magnetic materials are used, care must be exercised in the evaluation of μ_r because it varies with saturation and frequency. If the frequency is high enough, the permeability, even of the best magnetic materials, decreases to unity. Therefore, the shielding effectiveness of

Metal	Relative Conductivity, σ_r	Minimum Thickness in Mils
Aluminum	0.61	13
Brass	0.25	20
Copper	1.00	10
Magnesium	0.37	16.5
Silver	1.05	10
Steel	0.035 - 0.16	25 - 55
Tin	0.15	26
Zinc	0.29	18.5

Figure 3-8. Minimum Thickness Recommended for Shielding for Metals Having Conductivities Shown

magnetic materials, such as steel, should be carefully tested before using Equation (3-40) or (3-41) with a value of μ_r obtained from low-frequency measurements. Figure 3-8 shows the minimum thickness of shielding recommended for several common shielding materials together with their relative conductivities. The value for steel is conservatively based on a permeability of unity. In most cases this thickness is insufficient from a mechanical point of view, but it is the suggested minimum when there are no mechanical considerations.

The shielding effectiveness of a solid metallic wall increases with frequency (except possibly for magnetic materials). Therefore, measurements of this effectiveness need to be made only at the lower frequencies. In fact, it has been found in practice that, if a particular material and thickness are satisfactory below 20 mc, they will be satisfactory above that frequency. This condition may be vitiated, however, by the effect of seams, joints, or other discontinuities. For such discontinuities the opposite is true: Their shielding effectiveness decreases with frequency so that joints which are entirely satisfactory at low and medium frequencies may be quite "leaky" at high, very high, or ultra-high frequencies.

Instead of solid metal walls, meshes of metallic wires are sometimes used for shielding purposes. The attenuation of an electromagnetic wave in a mesh is considerably less than that in a solid screen. Therefore, the principal shielding action of a mesh is due to reflection. Tests have shown that mesh with 50 percent open area and sixty or more strands per wavelength introduces a reflection loss very nearly equal to that of a solid sheet of the same material. For this to be true, it is necessary that the mesh be so constructed that the individual strands are permanently joined at their points of intersection by some kind of fusing process so that good permanent electrical contact is made. Figure 3-4 shows that reflection loss of a solid shield decreases with frequency. In addition, the reflection loss in a mesh depends on the number of strands per wavelength, and since the wavelength decreases with frequency, the shielding effectiveness of meshes decreases with frequency faster than indicated by Figure 3-4. Since the reflection loss is also that portion which is most affected by the impedance of the wave, and hence the configuration of the source, it is best to make careful tests whenever meshes are to be used for shielding rather than to rely on theoretical considerations.

1.3.2 OPENINGS IN SHIELDS

A case designed to completely enclose a unit such as a receiver or motor must have openings and other discontinuities for the following purposes: to pass power, control, and output leads; to allow access for maintenance and servicing; to aid the ease of manufacture and assembly; and to permit proper ventilation, drainage, and heat transfer. Of these only the considerations of ventilation, i.e., the circulation of air, and drainage of any condensed moisture require actual openings. All the other requirements can be satisfied with temporary, semi-permanent, or permanent seals.

Leakage of electromagnetic energy through actual openings may be minimized by controlling either the size or the shape of the holes. The amount of electromagnetic energy that can escape through a hole in a shield is roughly proportional to the size of the hole provided that its dimensions are small compared to a wavelength. This means that the leakage can be kept negligible simply by making the holes sufficiently small. When the only purpose of the hole is the drainage of condensed moisture, a small number of very small holes of no more than 1/8-inch diameter is usually sufficient, and the leakage through these holes is negligible except in the case of extremely powerful interference sources such as ignition systems or radar modulators. For proper ventilation, larger openings are required. Such openings must then be covered with fine-mesh copper screen which must be soldered or welded along a continuous line around the edge of the opening. The type of mesh must be chosen in accordance with the principles explained in the previous paragraph. Since a mesh for effective shielding action has rarely more than 50 percent open area, and frequently less than that, the size of the opening must be correspondingly increased for effective ventilation. If the mesh must be easily removable, it must be attached with screws or bolts in sufficient number to insure a high pressure contact along a continuous line completely around the edge. Here, as in other joints, maintenance constitutes the largest problem. The contact surfaces must be thoroughly cleaned each time the mesh is screwed back into place.

An alternative to reducing the size of the openings, either by making the holes themselves sufficiently small or by covering them with metallic meshes effectively making many small holes out of one big one, is to design the shape of the openings in such a way that the escape of electromagnetic energy through them is prevented. The most effective way of doing this is to surround the openings by protruding sleeves which, effectively, convert the openings into waveguides. To be considered a waveguide, the length of the sleeve should be at least three times its longest cross-sectional dimension, or three times its diameter if of circular cross-section.

Waveguides act like high-pass filters. They pass without attenuation (neglecting losses in the walls and dielectric) all frequencies above the cut-off frequency and attenuate all frequencies below the cut-off frequency. Since a particular waveguide permits many modes of transmission of electromagnetic waves through it, there are many cut-off frequencies associated with each guide, one for each mode of transmission. The lowest cut-off frequencies for rectangular and circular waveguides are given by the following expressions:

$$f_c = \frac{5900}{b} \tag{3-42}$$

$$f_c = \frac{6920}{d} \tag{3-43}$$

where f_c is the lowest cut-off frequency in megacycles per second, b is the longer inside dimension of the rectangular guide in inches, and d is the inside diameter of the circular guide in inches.

In a waveguide, the attenuation per unit length for a wave below the cut-off frequency of the waveguide, neglecting dissipation, is given by the following expression:

$$\alpha = 0.00463 \; f \; \sqrt{(f_c/f) \;\; - 1} \tag{3-44}$$

where the attenuation, α, is in decibels per inch, and f is the frequency in megacycles per second. When the frequency is less than one-tenth of cut-off frequency, a good approximation, independent of frequency, may be obtained in the following form:

$$\alpha = \frac{27.3}{b} \tag{3-45}$$

$$\alpha = \frac{32}{d} \tag{3-46}$$

for rectangular and circular guides, respectively. Figure 3-9 shows a plot of Equations (3-45) and (3-46). From these curves the attenuation per unit of length (decibels/inch) may be read directly for any size of rectangular or circular waveguide. The same figure also shows the cut-off frequency (Equations (3-42) and (3-43)) in each case, so that the limit of validity can be determined from the same figure. For example, if 70 db are required up to at least 600 mc in a circular waveguide, the inside diameter must be no more than 1.15 inches based on a cut-off frequency of 6000 mc. For that size, the attenuation is 27.8 db per inch, so that about 2.5 inches of length would apparently give 70 db of attenuation. But remembering that the minimum ratio of length to diameter is three, a length of three inches should be chosen. It is seen that, since the minimum length is 3b or 3d, the minimum attenuation is always at least 82 or 96 db for all frequencies up to about one-tenth of the cut-off frequency. A picture of an end-cap for a direct current motor whose openings for ventilation were designed in accordance with these considerations, which proved very effective in actual tests, is shown in Figure 5-8 in Chapter 5 of this volume.

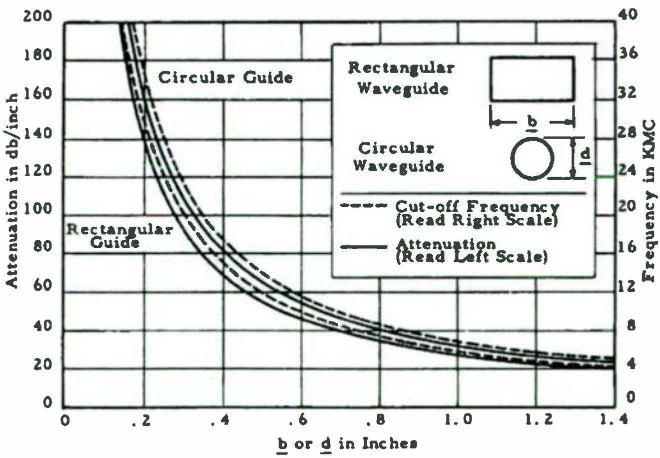


Figure 3-9. Cut-off Frequency and Attenuation of Waveguides

1.3.3 JOINTS

When it becomes necessary to join together several parts of a complete shield, the first consideration must be to keep the number of such joints down to the absolute minimum. In practically all cases where joints are necessary to allow ease of manufacture and access for maintenance, the shield should be constructed of no more than two parts having only one joint. In some cases it is most desirable to make a permanent joint by welding or soldering and sacrifice accessibility. This is done, for example, with ignition coils which are sealed permanently in a metallic shield. When the coil fails, the shield and coil are considered expendable and are replaced as one unit. This extreme solution of the problem of joint design cannot, of course, apply to more complicated pieces of equipment that may require frequent servicing.

When joints are made, the most important requirement is that a continuous metal-to-metal contact be maintained along a continuous line. When the pressure is maintained by means of screws or bolts, a sufficient number must be used to assure high unit pressure even at the points most distant, i.e., farthest away, from any screw or bolt. Lack of stiffness of the mating members is an important factor in producing distortion of the mating surfaces, a condition which results in bulging and insufficient pressure for good electrical contact. Flange and cover-plate joints should be made circular wherever possible because of the ease with which the surface can be machined either plane, grooved, or tapered. An example of a good shield design with a circular joint is shown in Figures 3-10 and 3-11 which show a photograph of a generator-regulator shield assembly open, and a similar assembly closed. A retaining band is used to maintain high unit pressure all around the joint.

A modification of the taper or wedge cover-plate design is shown in Figure 3-12. Here the shape of the mating members assures positive contact along two continuous lines. For thinner shielding materials, when screws cannot be used because the members are not stiff enough to maintain high pressure between screws, the "paint can" cover illustrated by Figure 3-13 gives good results. It can be used only for shields that need not be designed for strength and rigidity.

In the event of a gap or break in the shielding, such as might be encountered at poor joints in equipment racks or component housings, the opposition to current flow is increased. It is at these points that electrical interference may emanate. Some of the current may travel along the

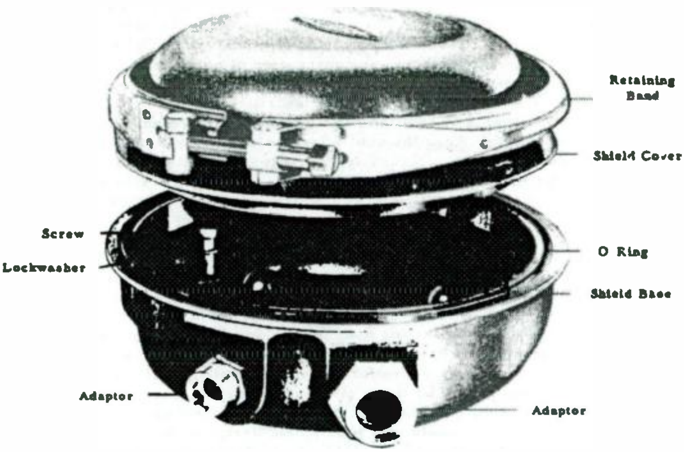


Figure 3-10. Generator Regulator Shield Assembly Open

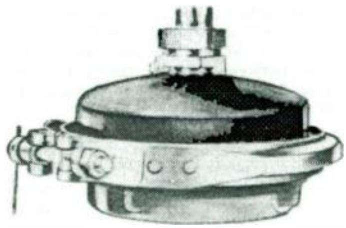


Figure 3-11. Tachometer Shield Assembly Closed

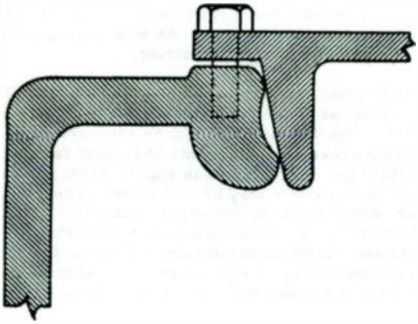


Figure 3-12. Design of a Tapered Cover Plate

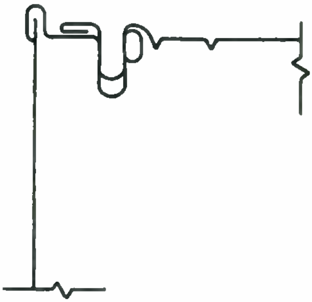


Figure 3-13. "Paint Can" Cover Plate

outside surface of the shield where it can establish an electromagnetic field and cause electrical interference. As with joining separate shield parts, electrical continuity is very important.

A multiple point, spring-loaded contact is a very efficient method of obtaining electrical continuity. In general, a serrated shim inserted in the aperture of the discontinuity will be satisfactory. The serration gives enough spring pressure at its points of contact for electrical continuity. A suggested type is illustrated in Figure 3-14. The materials used in constructing the shims may be beryllium copper, German silver, phosphor bronze, sheet steel, or tempered aluminum. To prevent corrosion and resultant electrical discontinuity, the case should be constructed of the same material. If the materials are different, the shim must be protected by cadmium plating or equally effective alternative methods to provide good electrical contact while at the same time preventing corrosion.

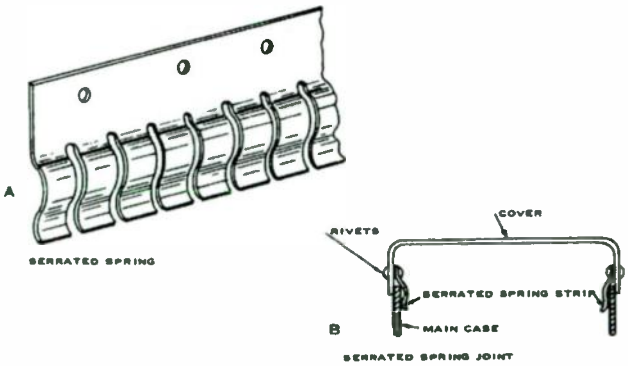


Figure 3-14. Multiple Point Spring Loaded Contact

1.3.4 CONDUCTIVE GASKETS

Failure of the flange and cover-plate designs to provide continuous line contact has necessitated the use of various types of conductive gaskets at the interfaces of mating surfaces. Two examples of designs of cover plates using conductive gaskets are shown in Figure 3-15.

The use of conductive gaskets in any shielding assembly is an admission of the inadequacy of the joint design because if continuous line contact at all mating surfaces were attained, there would be no need for gaskets even for providing moisture and gas seals at the joint interfaces. Yet, conducting gaskets are used in many types of shielding of present design and consequently must be discussed in some detail.

Any gasket must have a degree of compressibility if it is to conform to the mating surfaces. The degree of conformance is largely dependent on the available pressure and the compressibility of the material of which the gasket is made. It can safely be said that there is no one ideal gasket material for all purposes, but the successful employment of any type of gasket can be greatly improved if its use is properly taken into design consideration at the outset. The introduction of a gasket at a later date may not be as highly successful.

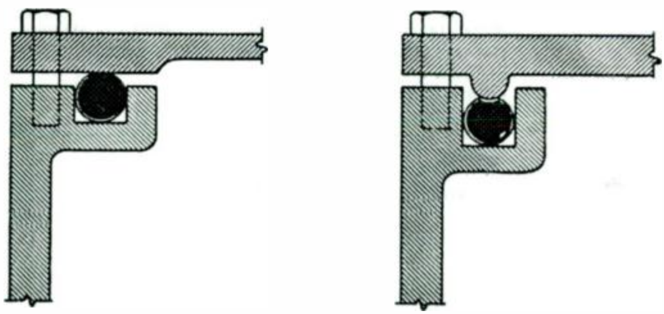


Figure 3-15. Cover Plates with Conductive Gaskets

Seals for radio interference reduction generally fall into one of two classes, the multiple-contact type or the continuous-contact type. A typical example of the former is a Neoprene-impregnated screen, and of the latter, the foil-wrapped gasket. Since gaskets are essentially interposed in the shielding system, they must satisfy the requisite degree of attenuation. For most metals, this can be calculated from depth of penetration formulas. In gaskets this is difficult except for the foil-wrapped type, where it is obvious that the foil thickness must be that required for effective penetration loss. In the multiple-contact type, the greater the number of contacts, the more effective the gasket becomes.

Conductive gaskets ordinarily are used only on the flange and cover-plate type of joints. The various types of gaskets either now in use or proposed for use are:

a. Metal screen impregnated with Neoprene (excess Neoprene removed from wire-mesh surface by use of abrasives)

b. Metal foil over Coroprene core

c. Wire sleeving over Neoprene core (for use in slotted flange and cover-plate design)

d. Sprayed-metal Neoprene gaskets

e. Aluminum-Magnesium alloy crystals suspended in Dow Corning No. 4 Compound

f. Serrated washer-type metal gasket

g. Metal-brush type (for use in slotted flange and cover-plate design)

h. Aluminum tubing filled with a Neoprene core (for use in slotted flange and cover-plate design

i. Compressed knitted wire mesh of copper, monel, or other metals.

A brief discussion of the above types of conductive gaskets, based on the experiences of the Military Services and on performance tests made by various equipment manufacturers follows. The letters refer to the gasket types described above.

The wire sleeving over a Neoprene core, a, the aluminum tubing filled with a Neoprene core, h, and the knitted wire mesh, i, for use both in flange and cover plates of slotted design, are the most promising of the several suggested types for providing interference-free service.

All gaskets that depend on multipoint contact for shielding, namely, the wire mesh, Neoprene-impregnated, a, the serrated metal washer, f, the metal-brush type, g, the Al-Mg alloy crystals, e, and the knitted wire mesh, i, are effective in proportion to the density of the points of contact, or in other words, to the degree to which line contact is approached.

The use of finely divided Al-Mg alloy crystals, while effective in providing multipoint contact, which closely approaches line contact provided the space between mating surfaces is not greater than the crystal size, is not recommended by the Military Services because of the danger that might be encountered should the metal particles inadvertently come in contact with the surfaces of bearings.

The metal-foil covered gaskets, b, while suitable for use on flat surfaces, have caused trouble on curved surfaces due to buckling or breaking of the foil.

Sprayed-metal Coroprene gaskets, d, were found by test to be inferior to other gaskets because of the porosity of the coating.

The knitted wire mesh, i, is far superior to any gasket made of woven mesh. It can be made from any metal or alloy that can be drawn into wire. It can also be combined with a sealing medium such as Neoprene when hermetic sealing, as well as shielding, is required.

Conductive gaskets can be made in practically any size or shape to fit the equipment with which it is to be used. The resiliency and density of the gasket material can also be varied according to the requirements of each application. A well-designed gasket insures good all-around contact even with appreciable unevenness of the mating surfaces or warping. Special machining for a close fit is not required.

Shielding gaskets for use in the covers of sheet-metal chasses and cabinets are available in round strips, strips with one or two fins, and double strips with a connecting web to meet varied installation requirements. Typical examples of the use of these types are shown in Figure 3-16.

When conductive gaskets are used with flange joints, the gasket should be thicker than the depth of the groove that holds it to insure complete contact, as shown in Figure 3-17. It is also important that the gasket be located inside the flange bolts to prevent leakage through the bolt holes. This is illustrated in Figure 3-18.

In choosing the material for the conductive gasket, the electrical conductivity is only one of many factors to be considered. For example, gaskets of monel metal have been found to shield as effectively, if not better than, gaskets made of silver-plated copper. In addition, monel metal gaskets have greater resiliency for the same volume of metal and are

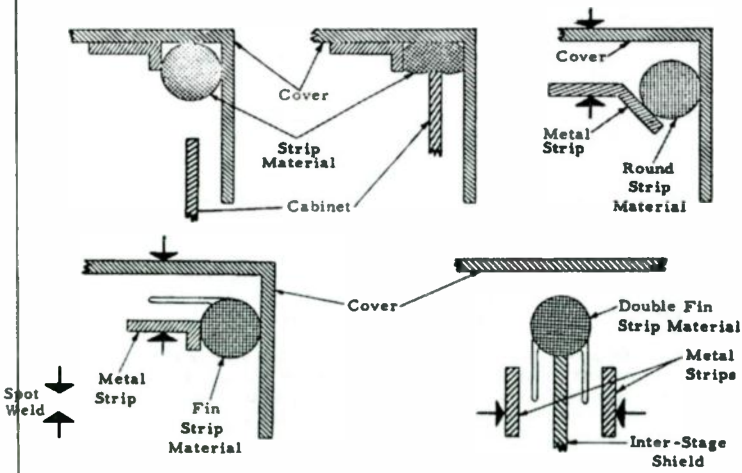


Figure 3-16. Use of Round, Single-Fin and Double-Fin Strips of Conductive Shielding Gaskets

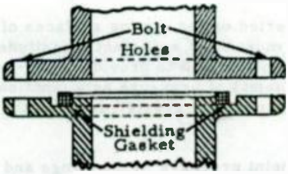


Figure 3-17. Use of Conductive Gasket with Flange Joint

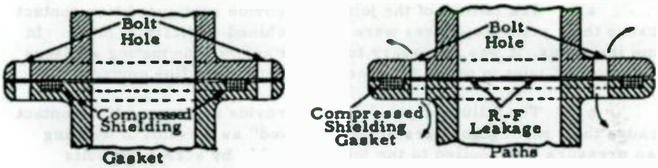


Figure 3-18. Proper Location of Shielding Gasket

much more resistant to corrosion. In fact, the ability to resist corrosive action, atmospheric or galvanic, is often as important as the electrical conductivity.

In spite of the many advantages of conductive gaskets, it must be remembered that a joint with a gasket can never produce as good a shield as a good joint with direct metal-to-metal contact. Tests have shown that a gasketed joint may have eight times the impedance of a similar metal-to-metal joint. This is due to the fact that a gasketed joint comprises two joints in series and may have a useful contact area of roughly one quarter of its apparent contact area. But good direct joints are very difficult to obtain, and still more difficult to maintain, and in general gaskets may improve the performance of average-to-poor joints.

1.3.5 JOINT PRESSURE

The pressure exerted on the mating surfaces of any joint in a radio interference shield must be of sufficient magnitude to maintain line contact of the mating metals and to provide a low-impedance path at their interface. The subject of pressure as a function of the quality of electrical contact is discussed in greater detail in the next section (1.3.6).

The problem of joint pressure in the flange and cover-plate type of shielding joint is of great importance because of the comparatively large and irregular surface area that must be joined together. Many joints of this type, as designed in the past, have not been satisfactory, with regard to prevention of radio interference, due to one of the following causes:

a. The failure of the joints to provide continuous line contact because their mating surfaces were not machined sufficiently plane. In some instances, it was necessary to scrape and lap the mating surfaces of this type of joint in order to achieve the necessary line contact.

b. The failure of the joints to provide continuous line contact because their mating surfaces became "waved" as a result of warping when pressure was applied to the joint assembly by screws or bolts spaced at intervals along the periphery; also, similar trouble has been encountered as a result of thermal changes in the joint assembly. The causes for this trouble can be traced to a lack of stiffness of the flange and plate edges and to excessive distances between the points of application of pressure.

c. The failure of the joint design to provide suitable means for maintaining constant pressure at the mating surfaces, especially during operation under service conditions when vibration and heat tend to loosen the fastening mechanism and reduce the joint pressure.

1.3.6 CONTACT IMPEDANCE

In the design of joints in radio interference shields, there are conflicts between mechanical, electrical, and chemical requirements, all of which need careful consideration in view of their close interrelation. A good design, as always, is one that achieves the best possible compromise among the various factors involved.

The impedance between two conductors that are pressed together is in most cases largely resistive; however, it can be inductive or capacitive depending on the nature of the joint and the wavelength involved. Therefore, the general term impedance is used to include the cases of the highest frequencies and dimensions. In any case, the total impedance across two conductors in contact must be carefully considered by the designer. It consists of the following components:

a. The impedance of a major part of the first conductor, in which the current density is almost constant, owing to the comparative remoteness of the interface.

b. The impedance of the remainder of the first conductor in the proximity of the contact area or areas where the current density is quite variable.

c. The impedance of the film between the conductors at their contact area or areas. The film may consist of one, two, or three films in series. Each contact surface, in general, has a certain amount of "bound" film due to corrosion or surface treatment; also there is a certain amount of "floating" film, such as grease, not securely attached to either contact surface. However, bound films are not always tenacious, and floating films sometimes adhere firmly.

d. The impedance of the second conductor just beyond the interface, where the current density is quite variable.

e. The impedance of the second conductor for some distance beyond the interface, a region in which the current density is almost constant.

In most practical cases, contact impedance consists mainly of "interfacial impedance," c, and the combined "constrictional impedance," b, plus d. Contact impedances are, therefore, sensitive to a number of factors, many of which do not prevail in ordinary solid conduction. The factors to be considered especially in designing shielding joints are (1) the conductivity, hardness, density, roughness, cleanliness, and corrosion-resistance of the materials, (2) the size and geometrical shape of both mating surfaces, and (3) the pressure, that is, the force per unit area binding the surfaces together.

Since in most cases the reactive portion of the impedance is very small compared to the resistive component, it can be neglected and the effective contact resistance, or real part of the impedance, can be approximately determined by the equation:

$$R = KP^{-N} \tag{3-47}$$

where R is in ohms, K is constant for a system, P is the pressure, and N is an exponent usually lying between 0.5 and 2.0, depending upon the nature of the contact surfaces. The determination of K and N is somewhat difficult because of their interdependency on the materials and geometry. The test data on R as a function of P may be plotted to great advantage on log-log graph paper. The resulting points approximate a straight line of negative slope N.

Logarithmic curves of R plotted as a function of P usually show a marked linearity - with erratic departure from a straight line taking place only at very low and extremely high pressures. Extremely rough or warped surfaces show a very high contact resistance at low pressure, but a rapid decrease in resistance with added pressure. Such surfaces at first may touch at only two or three points, but due to high unit pressure, these points are eventually depressed, permitting other points to make contact. Extremely smooth or lapped surfaces also show very high contact resistance at low pressure, but their decrease in resistance with added pressure is less marked. Such surfaces may, at first, engage films of corrosion, oil, air, and water, all of which are difficult to eliminate with low unit pressure.

Theory and experience indicate that the best contact surface is a slightly rough, though clean, surface. It has been noted that highly polished platinum contact points, often used in breaker assemblies, operate best after a brief run-in period. The preferred range of roughness must be determined in various cases. It should not be so rough as to cause severe air leakage around an ungasketed joint. A surface may not actually appear rough except under a magnifying glass. Instruments are available for the measurement of roughness, but it might be unwise and unnecessary to attempt to use them in production testing because once a satisfactory range of roughness has been determined, it is not difficult to specify a control standard.

In the design of shielding joints, an important question arises as to whether any particular flange width or area results in a minimum direct current resistance when the clamping force is constant. This question may be approached analytically. The measured resistance R of a metallic conductor is directly proportional to the length L of the conductor and inversely proportional to its cross-sectional area A. Introducing a constant of proportionality, ρ, called the resistivity of the metal, permits expressing the measured resistance as:

$$R = \rho L/A \tag{3-48}$$

For the case in question, ρ refers to the joint itself rather than to the metal at either side. The length L refers to the thickness of the interstice between the mating surfaces, and A refers to the actual cross-sectional area, even though all of it does not effectively conduct current. Obviously, the values of ρ and L depend primarily on the nature of the mating surfaces and are independent of the area A. All three factors on the right side of Equation (3-48) are actually functions of the pressure but A, as defined for this case, changes a negligible amount with pressure. Therefore Equation (3-48) can be written as:

$$R = f(P)/A \tag{3-49}$$

where the function, f(P), accounts for the change in ρ and L with pressure. Equation (3-47), derived empirically, accounts for the change of these two factors with pressure by the value of the exponent N.

The problem is to determine the effect of a change in the actual cross-sectional area A on the measured resistance R when the force F holding two mating surfaces together is kept constant, all surface conditions remaining the same. Analogy to Equation (3-49) shows that K of Equation (3-47) is, in general, a function of the area A. Since the resistance is inversely proportional to A, as seen from Equation (3-49), the function K must be equal to K'/A, where K' is a true constant independent of both the area and the pressure. If F is the constant force, the pressure P is F/A. Substitution of these expressions into Equation (3-47) gives:

$$R = \frac{K'}{A}\left(\frac{F}{A}\right)^{-N} = K'F^{-N}A^{N-1} \tag{3-50}$$

It follows then, for a constant F, that R is independent of A when $N = 1$, increases as A increases when $N > 1$, and increases as A decreases when $N < 1$.

This means that the flange width has no effect on the direct current resistance when the total clamping force is constant, providing the log-log curve (abscissa and ordinate having same scale) of R plotted as a function of P has a 45° slope. If the slope exceeds 45°, as in the case of very sharp or rough surfaces, such as the perforated steel-core gasket, the flange width should be small. If the slope is less than 45° (N less than 1), as in the case of all other gaskets and all surfaces not covered with metal paint, the flange width should be large. As N approaches zero, an increase in clamping force is less effective than a corresponding increase in area in reducing resistance. When N varies from 0.5 to 0, the effect of area changes from an inverse square root to an inverse proportion (as in wire conductors).

As a general design consideration for interference-free operation, it is well to keep in mind that the direct current resistance of practically all contacts is reduced by the use of wider contact surfaces; however, radio frequency impedance is not proportionately reduced, and gaskets, when used, may be less satisfactory. A narrow contact is generally a good contact because it has high unit pressure. It thus promotes good shielding if it is wide enough to pass substantially all of the radio frequency currents.

1.3.7 CORROSION

Corrosion is a very important factor in the choice of design of shielding joints for permanent interference-free operation. Corrosion is a chemical action or effect whereby metals are gradually disintegrated and converted to high-resistance compounds. Exposure to ordinary air results in surface corrosion on all but a few metals. Corrosion may be avoided by the extensive use of protective coatings, such as paint, varnish, lacquer, or grease, and chemical treatment.

One of the best protections for magnesium is a chemical treatment, known as Dow No. 7 or sodium dichromate treatment. This treatment results in very high resistance at electrical joints, and thus is objectionable in radio shielding applications. Moderate protection of magnesium, together with reasonable contact resistance, is afforded by another chemical treatment known as the Dow No. 1 chrome-pickle treatment. This is a brief-dip process, long used as standard practice for the protection of unfinished parts and as a base for subsequent paint coats.

Until the advent of VHF radio equipment, Dow No. 7 treatments were not particularly troublesome. However, as a result of recent radio interference troubles, these coatings were removed at the radio shielding joints by abrasion, and light coatings of white petrolatum were substituted for whatever amount of corrosion protection they might afford. Experience with magnesium joints thus treated has shown that, with reasonable care, the sanding of joints did little harm to bearings, gears, and dielectric parts and that the petrolatum offered fair protection against corrosion without insulating the joints. Unless subjected to salt spray, these almost unprotected magnesium surfaces withstand rugged field service with no more corrosion in evidence than that usually seen on solder or lead.

Investigations of improved methods of surface treatment to solve the combined problems of corrosion and contact resistance lead to the following observations:

a. Dow No. 1 treatment is a good temporary compromise.

b. Various metal paints, such as "Metal-X," "Alumilastic," "Alkyd Graphite Varnish," and powdered-metal lacquers have been investigated, but none have been found to possess high enough conductivity for use in radio shielding applications. On the other hand, greases containing metal powders (such as DC-52) are good conductors, but have a serious disadvantage in most practical applications when used between rotary parts and a dielectric material.

c. The plating of nonconductors, such as gasket material, and of metals, such as magnesium and stainless steel, which have poor adhesion to platings, has shown that the plated surfaces are never quite impervious to moisture, and thus, when two dissimilar metals are in contact with an electrolyte, serious corrosion may eventually occur. Corrosion would take place rapidly in the case of nickel-plated magnesium exposed to the atmosphere. After a slight amount of corrosion sets in, most plated surfaces tend to peel off, especially when severe mechanical conditions are imposed.

d. Metal sprays are subject to some of the same limitations as electroplating, but they may be very adherent and highly conductive.

e. Iridite No. 14, a chromate process, provides an effective corrosion resistant finish, acceptable under many Armed Services specifications, for aluminum and its alloys. It can be applied to shields, mounting brackets, waveguides, connecting plugs, etc., without appreciably interfering with electrical contact, because films of Iridite No. 14 offer low resistance to direct and to low or high frequency alternating currents. In addition, Iridite No. 14 provides an extremely tight paint-bond for either baked or air-dried paints and prevents the penetration of moisture to the base surface through the pores in the paint. Because of the above properties and its ease of application, Iridite No. 14 is replacing electrolytic anodizing in many applications.

Galvanic corrosion may occur when two or more kinds of metal are united in one assembly in the presence of moisture. In such galvanic couples, the corrosion of one metal is accelerated while the other metal corrodes less or not at all.

It is often necessary to combine unlike metals because of shortages and of various design requirements. Combinations such as silver and platinum, copper and monel, cadmium and steel, are known by experience to be quite compatible. New combinations of metals may be considered, in which cases little corrosion data may be available. Fortunately, it is still possible to anticipate, and thus avoid serious galvanic corrosion, without resorting to lengthy field and laboratory tests.

The first principle to observe is that joined metals should lie close together in the electromotive force series. (See Section 2.1.) Thus, magnesium should not be joined to brass or nickel as these combinations cause excessive corrosion of the magnesium. Certain aluminum alloys combine harmlessly with magnesium.

If the first principle cannot be strictly followed, then the second principle to observe is that the joined metals should be of such relative sizes that the attacked metal is the more abundant. For example, iron bolts on magnesium castings are fairly satisfactory, whereas aluminum rivets on brass plates are quite unsatisfactory. In the first instance, it is wise to cadmium-plate the iron and to chrome-pickle the magnesium. A third principle to observe is that joints should be kept tight and well coated in order to bar the entrance or exit of liquids and gases. A galvanic cell is powerless without moisture. It is also enfeebled (polarized) when the electrodes are coated with gas that cannot escape.

A fourth principle, often very useful, is generally inapplicable in the case of radio shielding joints. This principle requires the insulation of the joints wherever possible. Although solid insulation may be out of the question, there is no real objection to the use of semisolids such as petrolatum, which can insulate only those portions of the joints that cannot make contact anyway. Moreover, any protruding metal particles that meet are incapable of producing galvanic currents so long as moisture is kept from them by the semisolid. Naturally, in preventing corrosion, the contact impedance is kept low for a long period of time.

While the problems of corrosion and contact impedance are not yet fully solved, and until further research provides other materials with which to raise corrosion resistance while lowering contact resistance, the designer may well consider the use of bare metal surfaces for shielding joints or chrome-pickled surfaces upon which is applied an inhibiting grease, having low surface tension and suitable chemical properties as discussed above.

1.3.8 SOLDERED JOINTS

The corrosion problem is ever present in soldering. Corroded surfaces are difficult to solder, and soldering usually results in corroded surfaces, unless resin is used as a soldering flux. Three types of fluxes, namely, the chloride, organic acid (waxes), and organic base types, are all corrosive, differing in their rate of attack rather than in the end effects. The grease-paste emulsion of salt or acid content is little better than its fluxing ingredient, although it may be more conveniently applied. The cooling grease may limit the travel of the liquid flux, but it does not prevent internal corrosion of the affected parts, an action which may proceed even in the absence of air or external moisture. Resin, on the other hand, is noncorrosive because it is a solid that is quite impervious to liquids and gases. For this reason, electrical joints between small clean metal parts are usually made with resin flux which may be applied as a core within solder wire.

Even a good soldered joint is likely to exhibit an appreciable contact resistance. Grade A solder has a conductivity of about 12.2 percent. Thus an ideal soldered joint is never as good a conductor as a brazed or welded joint.

1.3.9 MAINTENANCE

The matter of cleanliness is perhaps the most important, yet least understood, factor in a good joint. An article may appear physically clean when wiped with a "not-too-dirty" cloth. It may be chemically clean when subjected to alkaline cleansers, solvents, or chromic acid. However, metal surfaces are not electrically clean unless they make good contact, a condition exceedingly difficult to prescribe or detect.

The wiping blades of a switch may be electrically clean although coated with grease. They may be electrically dirty when bearing an invisible film of oil. Magnesium may be electrically clean when chrome-pickled (Dow No. 1 treatment) and electrically dirty when polished with a fine abrasive cloth a few hours previously. Thus, the use of certain greases between mating surfaces may provide excellent enduring contacts because they inhibit corrosion, exclude foreign matter, and have low surface tension, permitting the intimate association of adjacent metal surface particles.

The great difficulties in maintaining a joint electrically clean necessitate a continued search for types of joints that will maintain their electrical properties unchanged with time and repeated use. Improper or insufficient maintenance of shielding joints is one of the largest single causes of radio interference, observed in the field, in equipment that was found satisfactory in the laboratory. Many times the cleaning of a mating surface or an additional turn of a screw that seemed tight, but was not, is all that is required to eliminate a major source of interference.

1.3.10 RIGID CONDUIT

Rigid conduit is rarely used for shielding purposes only, since damage to it requires too time-consuming repairs or entire replacement. When the use of rigid conduit is necessitated for other reasons, its incidental shielding properties may, of course, be utilized. Rigid conduit is usually made of aluminum or aluminum alloy having a wall thickness of at least 22 mils. A solid aluminum wall introduces an attenuation due to absorption of 2.6 db per mil to a plane electromagnetic wave at a frequency of 1 megacycle. Thus, 22 mils of aluminum gives an attenuation of about 57 db and this value may also be used, approximately, for a cylindrical shield. To this must be added the losses due to reflection, which depend on many factors, such as the number and position of the wires within and the shape of the conduit. The attenuation due to absorption increases as the square root of the frequency, so that at 100 megacycles the attenuation is at least 570 db. Thus, it is seen that any rigid conduit that satisfies all mechanical requirements usually is also adequate for shielding, except possibly for shielding very strong interfering currents of low frequencies. In this last case, it may be necessary to require conduit walls somewhat thicker than those demanded by purely mechanical considerations.

Bends in rigid conduit do not affect its shielding properties provided that the metal is not damaged in any way during the process of bending. Fittings at the end of the conduit require special considerations if shielding effectiveness is not to be impaired. These will be discussed later in Section 1.3.12.

1.3.11 FLEXIBLE CONDUIT

Since flexible conduit is relatively heavy and expensive, its use should be held to a minimum. Yet, there are conditions that require the use of flexible conduit, and these conditions exist quite frequently in connection with radio interference shielding. Flexible shielding conduit must be used whenever:

a. The ends of the conduit have an appreciable relative movement, such as the conduit leading to shock-mounted equipment

b. Shielding conduit is required at removable plugs

c. Conduit is required at equipment subject to frequency removal

d. Conduit is required in the engine-section wiring subject to large vibrations.

The most important examples of the use of flexible conduit for radio interference shielding are the antenna lead-in of receivers, the wiring between different units of radar systems, and the leads to the spark plugs in ignition systems.

Flexible shielding conduit is made of strip metal formed either into spiral bellows or into some other kind of spiral that allows interlocking of adjacent strips. It may be either soldered at the seams or allowed to provide sliding action between turns. For more effective shielding, it may be covered with one or more layers of woven metal braid. One or more layers of metal braid may also be used alone without any metallic tubing inside. If so, stiffness is provided entirely by the wires carried inside the braid. It is obvious from the method of construction that flexible shielding is relatively "porous" as compared to metal tubing. It is found, as would be expected, that for the same weight and kind of metal a seamless metallic tube has greater shielding effectiveness than a flexible one.

Leakage of electromagnetic energy from flexible conduit is of two distinct types: the penetration through the metal, called "penetration-leakage," and the escape through breaks, joints, or openings, called "opening-leakage." Penetration-leakage decreases with frequency while opening-leakage usually increases with frequency. Both are present simultaneously, but the first is negligible at higher frequencies while the second is usually negligible at lower frequencies. Total leakage plotted as a function of frequency usually follows a curve as shown in Figure 3-19. The slope at low frequencies is determined by the thickness of the metal, the slope at high frequencies by the size and shape of the openings, and the position of the minimum, which is characteristic of practically all flexible conduits, is determined by the design details.

The bellow construction (usually of soft brass) with soldered seams really has no openings and could, therefore, be considered "electrically tight." However, since the conductivity of solder is much less than that of brass, the shielding effectiveness is greatly reduced. Interlocked flexible metal hose, made by winding a suitably formed metal strip on an arbor and folding the edges in, as shown in Figure 3-20, has

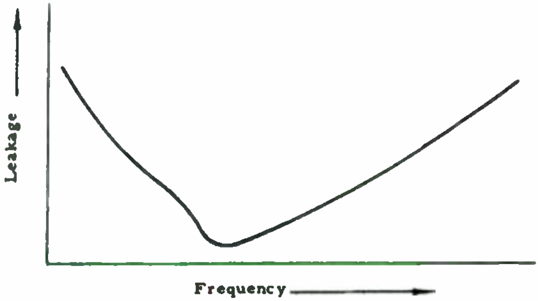


Figure 3-19. Leakage from Typical Flexible Conduit

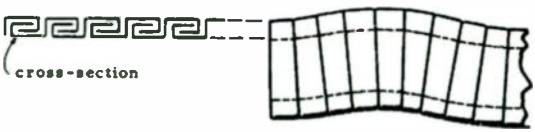


Figure 3-20. Construction of Interlocked Flexible Metal Hose

been widely used commercially in the manufacture of flexible tubing for shielding conduits. This construction provides three parallel metal-to-metal sliding joints between adjacent convolutions. Properly made, this type of hose provides excellent flexibility, long life under vibration, considerable ruggedness, and a very substantial degree of shielding. Compared with soldered-convoluted and seamless corrugated types of flexible metal hose, it has definite advantages with respect to shielding properties at the lower frequencies. This follows because, with the interlocked construction, it is possible to employ a much heavier gauge metal and still retain flexibility. Furthermore, the strip is folded over in such a manner that the wall of the hose is composed of four thicknesses of metal, except between convolutions.

The interlocked construction, however, depends for its effectiveness on the attainment of good electrical contact between adjacent convolutions. If the contact is poor, relatively large opening-leakage results, which impairs the shielding effectiveness at the higher frequencies. A seamless hose or a properly made soldered-convoluted hose, on the other hand, does not suffer from this drawback. Because of its excellent mechanical characteristics and low frequency shielding superiority, interlocked hose is attractive to the designer of flexible shielding conduit. The only real problem is that of obtaining and maintaining (throughout the useful life of the conduit) good electrical contact between convolutions.

In manufacturing interlocked hose, aluminum strip is sometimes employed. Aluminum is desirable from a weight standpoint, but is a notoriously poor contact metal. Insulating films form on aluminum surfaces almost immediately upon exposure to the atmosphere. This is what makes soldering of aluminum so difficult and also why it is poorly suited for fabrication of interlocked hose for shielding purposes. Bronze and copper are found better in this respect. But even these tarnish rather quickly and the hose loses its effectiveness. Stainless steel of the so-called "magnetic" variety is found to be usable and is adopted commercially in a few cases. It has definite mechanical advantages over copper and bronze and is highly resistant to corrosion. But it is not a particularly good contact metal. Therefore, it is found difficult to manufacture satisfactory shielding hose from this metal.

Various coatings may be applied on the strip with a view to improving convolution contact. Silver is effective but tends to be costly. It is possible to apply tin to stainless steel strip and produce a very much improved hose. This development results in the type of ignition conduit known as HTCD, which has been widely used. Recent investigations have shown that further improvements may be achieved by substituting tin-coated cold-rolled steel for the tin-coated stainless steel commonly used. The improvement observed is apparently due to the much higher magnetic permeability of the cold-rolled steel. However, these observations are based on tests performed at frequencies from 50 kilocycles to 12 megacycles, and little is known about the behavior at frequencies above 12 megacycles.

The designer must be warned against the use of interlocked metal hose with an insulating cord packing which has been widely used for various applications. Because the insulating packing, wound into the convolutions, prevents electrical contact between adjacent strips, large effective openings are present and the shielding properties at high frequencies are very poor.

The use of tightly woven metal braid to cover the flexible hose is to be recommended when the hose alone does not give sufficient attenuation. At low frequencies (below 1 mc), about 45 db additional attenuation may be expected from one layer of braid, but only about 25 db more (a total of 70 db) from a second layer. At higher frequencies, each layer contributes approximately the same amount of attenuation, and a double layer may be expected to give as much as 90 db of additional attenuation.

The shielding effectiveness of various types of flexible conduit for frequencies from 50 kilocycles to 12 megacycles is plotted in Figure 3-21. Here shielding effectiveness is measured in terms of the surface transfer impedance (see Section 1.4). For proper interpretation, it must be remembered that a low impedance means good shielding effectiveness and a high impedance means poor shielding effectiveness. The construction of these types is described in Figure 3-22.

While a search for improved types of flexible shielding conduit with minimum weight continues, it may be said that, for those applications where a large degree of shielding is required (as in ignition and radar systems), a convoluted type of hose with two braids should be used with special attention given to good electrical contact between adjacent convolutions. When a lesser degree of shielding suffices, a single braid may be used. If effective shielding at the lower frequencies is especially important, the use of magnetic material such as cold-rolled steel is recommended. In all cases, good electrical contact between adjacent convolutions is more important than low resistivity of the shielding material. As always, the maintenance problem is ever present. It

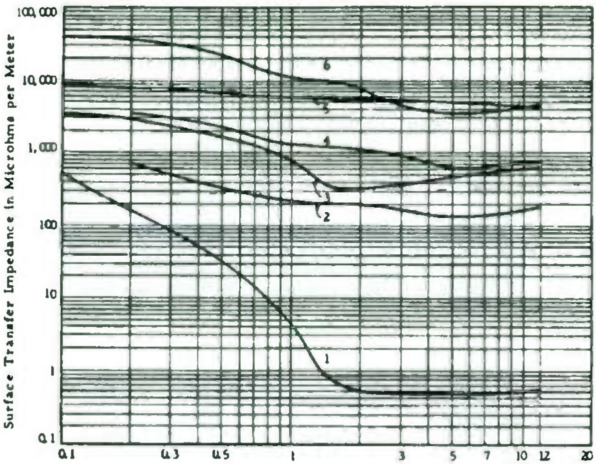


Figure 3-21. Shielding Effectiveness of Various Types of Flexible Conduit Described in Figure 3-22

Specimen No.	Construction Details
1	Interlocked flexible metal hose, tinned stainless steel, plus two tinned-copper wire braids.
2	Interlocked flexible metal hose, tinned copper plus one tinned-copper wire braid.
3	Interlocked flexible metal hose, silver-laminated bronze, no braid.
4	Soldered convoluted brass or bronze hose plus one copper-clad steel wire braid.
5	Soldered convoluted brass or bronze hose plus one bronze-wire braid.
6	Square locked hose, aluminum plus aluminum braid.

Figure 3-22. Construction Details for Various Types of Flexible Conduit (See Figure 3-21)

is not sufficient that good electrical contact exist initially; it must be maintained under all service conditions.

As with rigid conduit, fittings and connectors pose special problems. These will be discussed in the next section.

1.3.12 FITTINGS AND CONNECTORS

In fittings and connectors used to terminate either rigid or flexible conduit, the most important consideration, from a radio interference point of view, is that of obtaining and maintaining continuous line contact between the mating members. The problems are the same as in the design of joints. Maintenance is usually the more difficult aspect of the two.

In joining the connector to the conduit, the metal tube or hose, as well as the covering braid or braids, if used, must be welded or brazed to the metal housing of the connector in order to insure good permanent electrical contact. Great attention must be given to mechanical strength, particularly with flexible conduits, because the points of maximum flexure are usually adjacent to the connectors.

The mating members of the connectors must be designed to provide a high pressure contact along a continuous line even under a slight misalignment of the shielding components. Tapered or wedge-shaped designs have been found satisfactory, and the conical, or spherical type shown in Figure 3-23, gives excellent results even under adverse service conditions. It should be emphasized that practically all connectors which rely for electrical contact entirely on the high pressure between the threads have been found unsatisfactory in service.

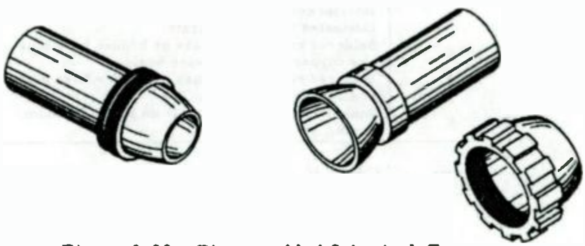


Figure 3-23. Disassembled Spherical-Type Connector for Flexible Shielding Conduit

1.4 METHODS OF MEASURING THE EFFECTIVENESS OF SHIELDS

All practical shields consist of metallic walls, which may be solid or may consist of mesh or braid, separating two regions of space. The effectiveness of a shield is due to its ability to isolate electromagnetic phenomena in one of these two regions. This effectiveness depends not only on the shield itself, i.e., its shape, material and physical dimensions, but also on the type of electromagnetic waves used during the measurements, i.e., their impedance (ratio of electric to magnetic field intensity), frequency, and polarization. Therefore, there is no absolute measure of the effectiveness of a shield. Measurements can determine only the relative shielding effectiveness under a given set of test conditions.

No standard methods are available for measuring the effectiveness of shields used to enclose completely interference-generating units such as motors, or interference-susceptible units such as receivers. Here effectiveness must be determined by actual operation. The motor is run in the way that resembles as closely as possible actual operating conditions and the region outside the shield is explored with suitable pickup devices at all frequencies of interest. The shielding is considered effective if no signal can be detected. A receiver may be tested similarly with a strong interference source placed directly outside and precautions taken that the signal cannot enter any other way than through the shield.

The testing of cables and conduit for shielding effectiveness is particularly important in connection with the suppression of radio interference. Special standard test methods have been developed, two of which will be described in detail here.

Briefly, the first method consists of measuring the voltage drop on the outside of the shield when a specified current flows through the conductor or conductors inside the shield. This method is based on the fact that, in order for electric and magnetic fields to exist outside a region completely enclosed by the metallic shield and containing no other sources, currents and charges must be present on the surface of the shield on the side facing that region. A perfect shield would restrict all currents and charges to its inside; hence, no fields could be present in the outside region. The voltage drop along the outside of the shield, which is a measure of the integrated effect of the currents and charges on the outside, is, therefore, a measure of the effectiveness of the shield. This method is particularly suited for coaxial cables consisting of one inner conductor and a concentric outer sheath serving both as shield and as return path for the current. It is also applicable to conduit carrying more than one conductor, but a difficulty arises because it is not immediately clear what is meant by "the current inside" if there are several conductors carrying different currents, possibly in opposite directions. It is, however, the only possible method for conduit filled with a solid dielectric, in which the conductors are embedded, since the second method requires the replacement of the inner conductors by a radiating coil.

The second method consists of measuring the field strength at a point in the vicinity of the specified source both with and without the source being enclosed by the shield to be evaluated. This may be called a "direct" method of measurement since the ratio of the two field strengths is a direct measure of the attenuation introduced by the shield. It is a measure, however, only for the particular type of field produced by the source, which may be quite different from the actual fields encountered in practice. This method is particularly suited to shielding conduit with more than one inner conductor in which the shield carries little or none of the line current. It is not directly applicable to shields carrying the return current because, when the shield is used as a return path, removal of the shield destroys the circuit, and, hence, the conditions could not be kept similar for measurements with and without the shield. Also, this method can obviously not be used when the conduit is filled with a solid dielectric.

A detailed description of the two methods follows.

1.4.1 METHOD I

1.4.1.1 Principle of Operation

This method utilizes the concept of surface transfer impedance, which is defined as the longitudinal voltage drop along the outside of the shield per ampere of current carried by the shield. The impedances of primary interest are small; hence, units are given in microhms. Comparative tests are referred to a standard. A tube as described in Figure 3-24 is recommended. For best results, the test specimen and the standard should have the same size. Relative leakage of zero db indicates leakage equal to that from the standard.

The lower the transfer impedance per unit length, the better the shield. Thus its reciprocal, the transfer admittance, would more appropriately represent a unit expressing shielding effectiveness.

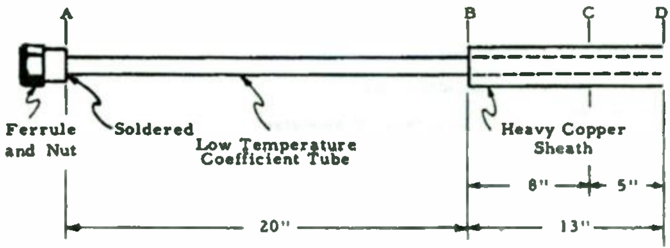


Figure 3-24. Tubular Shielding Standard

Longitudinal external voltage drops, E_x and E_s, across specimen and standard respectively, are compared with the same currents and at the same frequency. Leakage is expressed in decibels and computed from the equation

Leakage of specimen relative to a standard =

$$20 \log_{10} \frac{E_x}{E_s} \tag{3-51}$$

For shielding effectiveness, the same equation is used with the voltage ratio inverted.

A calibrated radio frequency microvoltmeter can be used to find the external voltage drop across the specimen. However, it is preferred to evaluate this voltage by comparison with a signal of equal intensity from a standard signal generator. By means of a switch, a suitable receiving device receives the signal from specimen and generator, alternately, and the generator is adjusted until equal receiver outputs are obtained.

To measure transfer impedance directly, it is necessary to measure the current in the center lead. The basic circuit is indicated

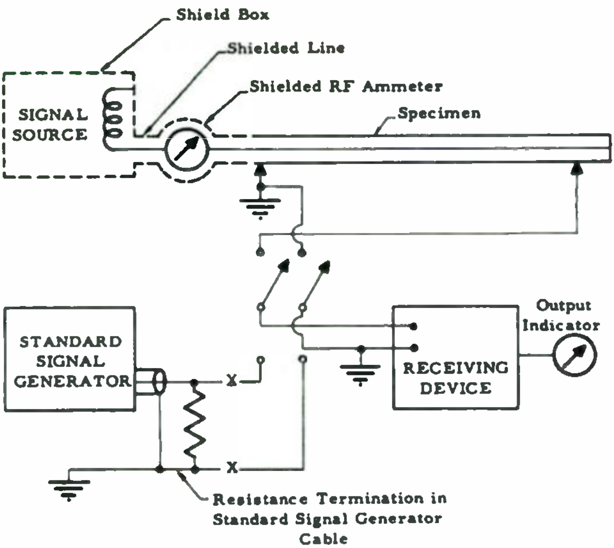


Figure 3-25. Basic Circuit Used for Measuring Transfer Impedance

in Figure 3-25. A current indicating device, such as a shielded radio frequency ammeter, is inserted in series with the inner conductor of the coaxial line of which the specimen forms the outer conductor. The voltage, read from the signal generator setting in the same way as before, is divided by the current indicated on the ammeter, yielding the desired transfer impedance.

1.4.1.2 Instructions for Test

Using Figure 3-24 as a guide, the following itemized procedure should be followed:

a. Tube made of 70-30 Cupro-nickel, a non-magnetic alloy, which has an attenuation of about 21 percent and a thermal coefficient of attenuation of about 1.2 percent of that of copper.

b. Change in attenuation is absolutely negligible for ordinary temperature variations.

c. A tube of this alloy is usable as a standard (to provide a given amount of attenuation) at frequencies more than 20 times as high as is a copper tube of the same wall thickness.

d. Wall thickness should be made as uniform as possible. A 0.125-inch-thick tube is suitable for radio frequencies up to about 1 mc; a thickness of 0.050 inches is satisfactory from 0.1 to about 8 mc.

e. A straight stiff center conductor (preferably 1/16 to 1/8-inch diameter hard copper) is held centrally by means of 4 or 5 equally-spaced thin polystyrene wafers.

f. Connection from the center conductor to the inside of the tube is made at point "C." The sheath extension beyond the point of contact C (indicated by the section CB) permits an equal current distribution around the periphery; without this extension lack of perfectly symmetrical contact around the entire circumference would result in a non-symmetrical current distribution.

g. Contacts for measuring leakage are applied across section AB; about 20 inches is a convenient length.

h. A heavy copper sheat BD (at least 1/16-inch thick) is used to reduce leakage from section BD to a point where it is negligible compared with the leakage from test section AB. The sheath fits snugly over the cupro-nickel tube and is sweated on with soft solder.

i. Section CD acts as a cut-off tube preventing leakage from the open end D. Its length should be at least five times the inside diameter of the inner tube.

j. The standard has a spherical-faced brass ferrule soldered to end A. The assembly fits into a conically beveled seat and is secured by means of the nut shown in the figure.

1.4.1.3 Equipment Required and Suggested Types

Figure 3-25 and the following list suggest the test equipment to carry out this test.

a. Oscillator-Amplifier Signal Source. Hewlett-Packard 606A and 608C, or equivalent. Care must be exercised in maintaining a good sinusoidal waveform, especially at the lower frequencies.

b. RF Ammeter, Thermocouple Type. Weston, Model 640 or equivalent.

c. Standard Signal Generator. Measurements Corporation Model 65-B or equivalent.

d. A suitable receiving device. Stoddard RFI Meter or equivalent.

e. A frequency standard, used to maintain signal source and generator within close tolerances. Millen Type 90501 or equivalent.

This method was used to investigate shielding effectiveness for frequencies up to 12 megacycles, and the range can probably be extended to several hundred megacycles.

1.4.2 METHOD II

1.4.2.1 Principle of Operation

The radio frequency output of a signal generator is applied across the terminals of a radiating coil located within a completely shielded test cabinet, as indicated in Figure 3-26. The coil is magnetically coupled to a probe, which is connected directly to the output of a radio receiver. The output of the generator and the gain of the receiver are then adjusted to give a 10-milliwatt reading on an output power meter connected to the receiver output.

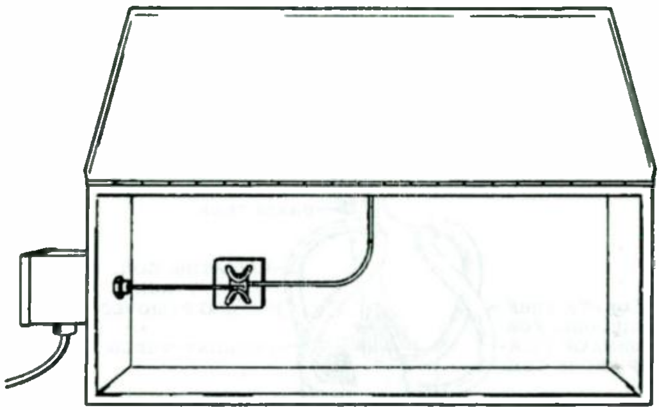


Figure 3-26. Test Cabinet Illustrating Complete Shielding of Conduit Tester

A specimen of shield is then inserted between the radiating coil and probe. The signal generator only is adjusted to obtain the same output as before.

The ratio of generator voltage required with the shield for a given output to the voltage required without the shield is a measure of the shielding effectiveness.

3.4.2.2 Basic Test Setup

The equipment specified is suitable for measuring shielding effectiveness of electromagnetic field strength attenuation less than 100 db

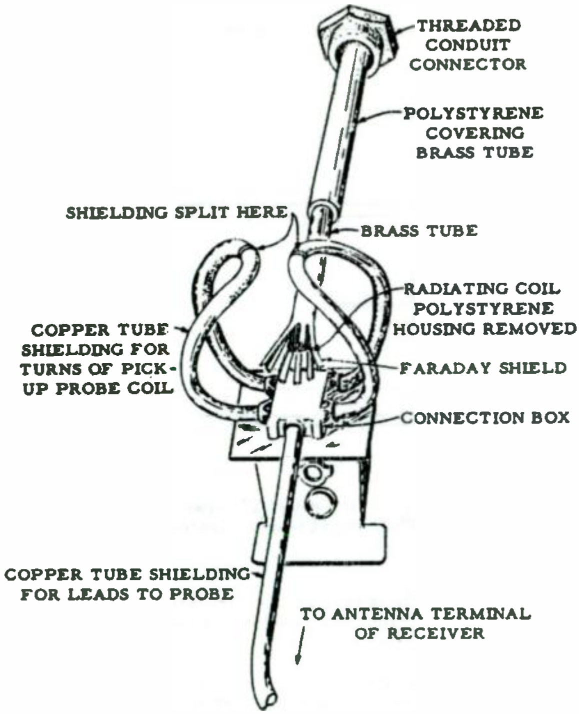


Figure 3-27. Detail of Pickup Probe and Radiating Coil

in the range 0.15 to 50 mc, and less than 45 db in the range of 50 to 156 mc. These limits may be extended by using signal generators of greater output or receivers of greater sensitivity.

In this frequency range, a typical radiating coil consists of seven turns of copper wire wound in the form of a solenoid, located within a polystyrene sheath, and surrounded by a Faraday shield as shown in Figure 3-27. The Faraday shield insures repeatable measurements in the frequency range above 100 mc.

The pickup probe consists of two loops of copper tube shielding, within each of which is an insulated conductor. The loops are placed one on each side of the radiating coil in that position which permits linking the maximum number of flux lines in the field of the radiating coil. The loops are soldered into a small copper connection box at the base as shown in Figure 3-27. The copper tube shields the insulated conductor against electrostatic coupling, but in order to permit unimpaired magnetic coupling to the insulated conductor, the shielding is split at the top of each turn and the two cut ends are separated by a short air gap.

The pair of insulated conductors is continuously shielded to the rear wall of the test cabinet where one of them is connected to the inner surface of the tube that contains it, which is grounded to the wall of the cabinet. The other conductor is connected to the antenna post of the test receiver.

1.4.2.3 Equipment Required and Suggested Types

The following list of equipment is suggested for carrying out this test.

a. Standard Signal Generator. General Radio Model 805-C or equivalent. (Range, 16 kc - 50 mc; maximum voltage output, 2 volts.)

b. Standard Signal Generator. General Radio Model 1021 AV or equivalent. (Range, 40 - 250 mc; maximum voltage output, 1.0 volt.)

c. Commercially available receivers (range to match signal generator).

d. Output Power Meter. General Radio Type 583-A or equivalent.

1.4.2.4 Instructions for Test

The following itemized procedure should be employed, guided by Figure 3-28.

a. Place the test cabinet on a grounded metal test bench. Solder a strip of bonding braid (as short as possible) from each of the four corners of the test cabinet to the metal test bench.

b. Connect the output of the signal generator to the coaxial cable connector mounted on the cabinet connection box.

c. Connect a coaxial cable from the probe terminating connector on the rear of the test cabinet to the antenna post of the receiver. The cable should be bonded to the test bench at 18-inch intervals. (The antenna post should be shielded to prevent stray pickup.) Connect the receiver ground post (or chassis) to the metal test bench with a short strip of bonding braid.

d. Connect the output power meter to the receiver and adjust its impedance to correspond with the receiver output impedance.

e. Connect the instruments to a 110-volt ac regulated power source. Allow a 1/2 hour warm-up period.

f. With no conduit installed, but with the cabinet lid tightly closed, set the signal generator to the desired test frequency. Increase the output to enable detection of the signal.

g. Tune the radio receiver to the same frequency, observing the output power meter for maximum indication. (Be sure that the receiver is not tuned to the "image" frequency.)

h. Reduce the signal generator output to zero and adjust the receiver gain to obtain not more than 2 mw of residual noise.

i. Increase the signal generator output until a 10 mw reading is obtained on the output power meter. Record the signal generator output as E_2.

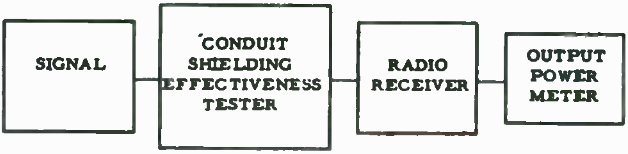


Figure 3-28. Block Diagram for Shielding Effectiveness Test

j. Install the conduit sample.

k. Tightly close the cabinet lid.

l. Increase the signal generator output until a 10 mw reading is again obtained on the output power meter.' Record the signal generator output as E_1.

m. Determine the ratio of the two signal generator output signals. Convert the ratio into decibels using the expression 20 $\log_{10}$ (E_1/E_2). This figure represents the shielding effectiveness of the conduit.

2. BONDING

The purposes of bonding have been enumerated in Chapter 1. This section deals with the techniques of bonding, with special emphasis on those applications of bonding which have a direct or indirect bearing on the suppression of radio interference. As was pointed out before, poor bonding may contribute to radio interference in three distinct ways: it may allow electric charges to build up which can produce a spark, it may allow a varying electromagnetic field to exist when the poorly bonded members are subject to shock or vibrations, and it may produce coupling paths due to the impedance of the poor bond being common to two circuits. The first and second conditions are usually found when structural parts are unbonded or poorly bonded. The third condition may exist in structural members also, but more often, it is found when electronic equipment, filters, or shields are inadequately grounded because of poor bonding.

Any one of the causes of trouble mentioned above is completely eliminated when a low impedance path is produced between the two conductors to be bonded. The impedance that is important in this connection is the impedance at radio frequencies. Actual measurements have shown that there is almost no correlation between the direct current resistance of a bond and its radio frequency impedance. Therefore, the direct current resistance cannot be used as a measure of the effectiveness of bonding. A method of measuring the radio frequency impedance of bonds is described in Appendix II of Volume II.

Moreover, even the measured radio frequency impedance is not a perfect indication of the effectiveness of the bond in an actual installation. Since in the actual installation the artificial bond, a jumper, rivet, or screw connection, is always in parallel with some kind of natural bond, such as a direct metallic contact or the capacitance between two surfaces not in direct contact, the total impedance between the members to be bonded must be evaluated by considering the various parallel paths through which a radio frequency current may flow. Thus, it may happen that a bonding jumper which was measured to have a satisfactorily low inductance combines with the capacitance of the system to form an antiresonant circuit of extremely high impedance. This shows that bonding techniques must take into consideration the actual installation in which the bond is to be used.

In general, two main types of bonding may be distinguished: direct bonding and bonding by means of jumpers. Direct bonding is achieved by insuring permanent metal-to-metal contact between the members to be bonded. If this method is practical, it is always preferable. But when the contacts are subject to frequent separation, such as the edges of doors, or if clearance between the bonded members must be maintained for mechanical reasons, such as with control surfaces, or if the equipment to be bonded is shockmounted, then direct bonding is not feasible and bonding jumpers must be substituted. It must be borne in mind that a jumper is never more than a substitute for a direct bond. At best, its impedance may be only slightly larger than that of a direct bond. At worst, it may actually increase the impedance.

2.1 DIRECT BONDING

Direct bonding is accomplished by direct metal-to-metal contact between two surfaces under high and uniform unit pressure. If properly constructed, a bond of this type has a low ohmic resistance as well as low radio frequency impedance. Permanent joints of metallic parts made by welding, brazing, sweating, or swaging; semipermanent joints of machined metallic surfaces held together by lock-threaded devices, rivets, tie rods, or structural wires under heavy tension; pinned fittings driven tight and not subjected to wear; and clamped fittings normally permanent and immovable: All these are considered as meeting the bonding requirements inherently if all protective coatings are removed from the contact areas before assembly.

Bonds formed by direct metal-to-metal contact through mating surfaces held together by clamping devices may deteriorate with time. This is brought about by corrosive action which in time makes the bond ineffective by causing the contact resistance to increase beyond tolerable limits. Corrosive action may be either the galvanic or the electrolytic type or both depending on the nature of the metals in contact and on whether or not 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.

In the galvanic type of corrosion, the two mating surfaces in contact with moisture act like a chemical cell of two metal electrodes immersed in a solution. In general, when an electrode is immersed in a solution, a potential difference develops across the junction of the electrode and the solution. The reason for this is that when a metal is placed in water, or any other ionizing solvent, some of the metal in contact with the water passes into solution as positively charged ions. This process leaves an equivalent amount of negative charge on the metal electrode. There is a tendency for such a process to occur at each of the two electrodes of the chemical cell which the moist mating surfaces of the bond resemble. However, if the electrodes are of the same metal and the solution in contact with them is homogeneous, no net potential difference across the two electrodes can be detected because the conditions at each are the same, exactly balancing one another. This would be the case for mating surfaces of the same metal exposed to moisture, and no such galvanic action would occur.

On the other hand, if the bond is formed by direct contact of dissimilar metals, in the presence of moisture, a chemical cell (resembling a voltaic pile or galvanic couple) is formed across which a net potential difference is developed. The reason for this is that each of the metals has a different tendency to go into solution as ions. Hence the potential difference (electrode potential) across the junction of one metal electrode and the solution is greater than that which exists across the other. Construction of a series of units, each made from two sheets of dissimilar metals separated by a wet cloth, shows that metals can be arranged in an "electromotive series" so that each is positive when placed in contact with the one next below it in the series. This series, for selected metals, is given in Figure 3-29. The chemical action* accompanying the establishment of these electrode potentials in a galvanic couple is such that the more positive electrode (higher in the series) corrodes by loss of metal while the other electrode does not. Hence, the nonreplaceable part of a joint formed by dissimilar metals should be a metal lower in the series than its mate. Moreover, the further apart any two metals are in the series means that a greater potential difference across the pair could be established and the chemical action (corrosion) would be more severe. Therefore, it is essential to select metals close to one another in the electromotive-force series when contact between dissimilar metals cannot be avoided. For example, the contact of a copper fitting and a magnesium casting would lead to excessive corrosive action because these two metals are too far apart in the electromotive-force series. This corrosive action could be minimized by plating the copper fitting with zinc or cadmium.

In addition to the galvanic action described above, there is another phenomenon, called electrolysis, which also produces corrosion due to chemical action. Electrolytic action takes place when a direct current flows between two metal surfaces in contact with a conducting solvent. The occurrence of electrolytic-type corrosion is independent of the nature of the metals in contact. It may occur along with galvanic action in joints of dissimilar metals, but, by itself, can account for the corrosion at joints formed by surfaces of the same metal in contact. Since structures and casings of equipment are often used for the ground return path of direct current power, there is the possibility of large dc currents through joints and connections serving as bonds to ground. If the joint contains moisture with dissolved salts or other impurities, the mechanism of electrolysis alone (by chemical reactions at the metal surfaces and the flow of ions) permits the passage of current, if no other path of much lower resistance exists. Depending upon the type of metal and the magnitude and direction of the voltage drop across the moist joint, dissolution and deposition of metal can occur at the metal surfaces of the

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*The above explanation of galvanic action is purposely simplified to give merely a qualitative picture of the phenomenon and to establish general rules for design practice. An exact description of the process, especially as to the actual seat of the EMF's developed, still lacks verification. Details concerning ion concentration and temperature dependence have been omitted from the discussion in this book.

Magnesium
Magnesium Alloys
Zinc
Aluminum 2S
Cadmium
Aluminum 17ST
Steel or Iron
Cast Iron
Chromium Iron (Active)
Ni-Resist
18-8 Stainless (Active)
18-8-3 Stainless (Inactive)
Lead-Tin Solders
Lead
Tin
Nickel (Active)
Inconel (Active)
Brass
Copper
Bronze
Copper-Nickel Alloys
Monel
Silver Solder
Nickel (Passive)
Inconel (Passive)
Chromium-Iron (Passive)
18-8 Stainless (Passive)
18-8-3 Stainless (Passive)
Silver
Graphite
Gold
Platinum

Figure 3-29. Electromotive-Force Series for Selected Metals (Listed in Decreasing Order of Tendency to go into Solution as Ions)

joint. Moreover, chemical action of the dissolved impurities at the metal surfaces causes rapid contamination, corrosion, and destruction of the joint.

As explained above, the possibility of galvanic or electrolytic action necessitates the use of extreme care in assembling joints which serve as bonds for the ground return path. The surfaces should be absolutely dry before mating and held together under high pressure to minimize the chance of moisture creeping into the joint. After a joint is assembled with no moisture occluded, it is good practice to seal the periphery of the exposed edge with a suitable protective compound.

Direct bonding may be improved, and its use may be extended to surfaces in relative motion provided that the clearance between them remains very small at all times, by the use of non-hardening conductive silver pastes, for which the following claims are made by the manufacturer: They are rubbery, adhesive solids, resistant to oxidation and corrosion caused by moisture, heat or fumes. They exhibit the same electrical characteristics as solder, and can be used to replace it where physical contact between the parts is maintained by mechanical means. Furthermore, if relative motion exists between the parts, a conductive paste which has the consistency of chewing gum has been designed to insure good electrical conductivity. Iron and other base metal contacts can be used in circuit-breaking equipment when treated with these pastes. Pastes are also effective in maintaining the electrical continuity of shields at seams, junctions, and joints.

2.2 BONDING JUMPERS

For direct or low frequency alternating currents, bonding of equipment is easily accomplished. A wire or a length of tinned-copper braid suffices. However, at radio frequencies the same jumpers present considerable impedance. In order to understand the factors which determine the magnitude of the impedance, the equivalent circuit shown in Figure 3-30 must be analyzed. In this diagram, R is the ohmic resistance including the increase due to the skin effect, L is the total series inductance of the jumper, and C is the combined capacitance due to the distributed capacitance of the jumper and the capacitance of the bonded members.

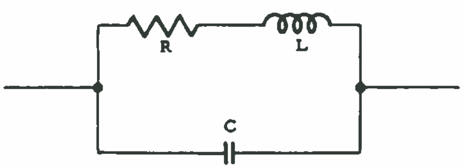


Figure 3-30. Schematic Diagram of a Bonding Jumper

Experiments have shown that the effective value of the resistance despite its rise due to skin effect at higher frequencies is negligible except near the point of anti-resonance. If the resistance is neglected, formulas for the impedance Z of the equivalent circuit are given as:

$$|Z| = \frac{\omega L}{1 - \omega^2 LC} = \frac{1}{\omega C}\left[\frac{1}{1 - \frac{1}{\omega^2 LC}}\right] \qquad (3\text{-}52)$$

where ω is the angular frequency.

If $\omega^2 LC$ is less than one, the circuit operates below its anti-resonant frequency and acts as an inductance. An increase of the values of capacitance or inductance, up to the point where $\omega^2 LC$ equals one, results in a decrease of the impedance. If the circuit operates above the anti-resonant point, $\omega^2 LC$ is greater than one, and the circuit acts as a capacitance. With an increase of the values of capacitance or inductance the impedance decreases. Therefore, to keep the radio frequency impedance low at frequencies below the anti-resonant frequency, the values of capacitance and inductance must be comparatively low, but for frequencies above the anti-resonant frequency, the values of capacitance and inductance must be comparatively high to obtain low values of impedance as shown in Figure 3-31A, B and C.

The region of frequencies of interest is almost always such that the anti-resonant frequency of the jumper occurs near the upper end of this region. In order to obtain values of radio frequency impedance as low as possible, it is necessary to have a comparatively low LC product. Lowering the LC product raises the anti-resonant frequency, and, as may be seen from Figure 3-32, lowers the impedance in the range of the frequencies of interest.

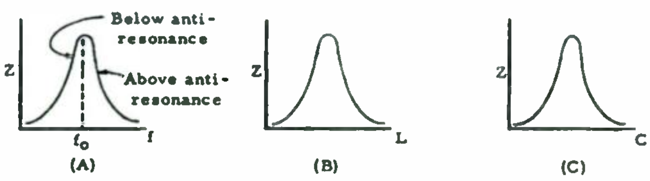


Figure 3-31. Magnitude of Impedance of a Parallel Circuit as a Function of: (B) Frequency, (C) Inductance, and (D) Capacitance

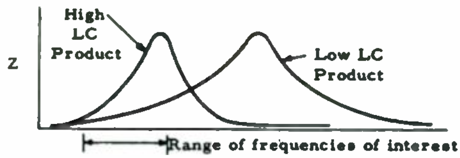


Figure 3-32. Magnitude of Impedance of a Parallel Circuit as a Function of Frequency for Two Different Values of the LC Product

Experiments have shown that the physical characteristics of a jumper have a marked effect on their radio frequency impedance. An increase in the length of a jumper causes its impedance to increase proportionally, but an increase in either its cross-sectional or its surface area causes a non-proportional decrease in its impedance. However, the change in the surface area exerts a greater effect on the impedance of a jumper than a corresponding change in its cross-sectional area. In the design of a bonding jumper, these results must be considered along with the requirements noted in Appendix IV of this volume.

For certain radio frequency suppression bonding applications, "round" bonding jumpers consisting of strands of wire arranged in a rope twist can be substituted for the more expensive "flat-braid" bonding jumpers consisting of woven strands of wire. Radio frequency measurements made in the range of 0.15 to 30 megacycles show that the impedance of the round jumper is only slightly higher than that of the flat braid of comparable size at all frequencies within this range. The curves of Figure 3-33 indicate this clearly. They also show the direct, approximately linear, relationship between the impedance and the frequency. This is evidence that the resistive component of the radio frequency impedance is comparatively low, and that its reactive component is inductive. Despite the slightly higher radio frequency impedance values of the round bonding jumpers, their use is justified in view of the greater dependence of impedance upon other conditions and characteristics such as the length of the bond and its orientation with respect to the ground plane.

In Chapter 1 several applications of bonding other than the suppression of radio interference are given. Among these is the minimization of lightning damage. To provide adequate bonding for lightning protection, bonding jumpers of tinned-copper stranded cable should have a

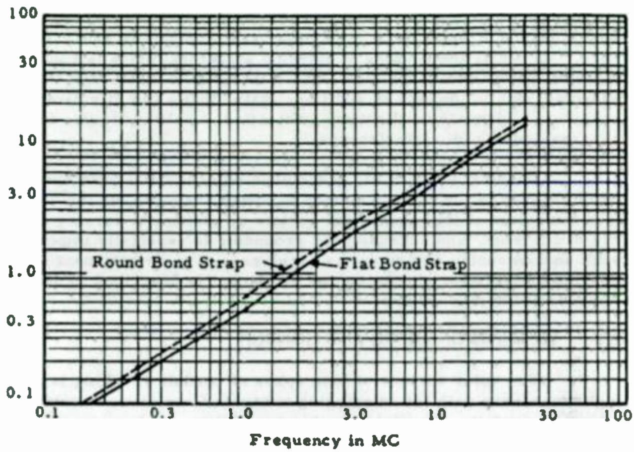


Figure 3-33. Impedance Characteristics of Flat and Round Bonding Jumper

minimum cross-sectional area of 6475 circular mils to withstand a maximum current surge of 100,000 amperes built up in 10 microseconds and damped to one-half its maximum value in 20 microseconds. If stranded aluminum cable is used, its minimum cross-sectional area should be 10,000 circular mils. Twisted cables are superior to braided cables of the same cross-sectional area because the latter crystallize and break more readily than the former due to the magnetic stresses accompanying the surge currents. Due to the oscillatory nature of the discharge, stranded wire must be used to minimize the skin effect.

Although the effective resistance provides a fairly accurate evaluation of the effectiveness of bonding when applied to lightning protection, it does not have, as previously mentioned, any significance in indicating the effectiveness of bonding used to suppress radio interference. As far as the bonding of external structural parts is concerned, lightning protection is the most important consideration, and it must be assumed, due to the lack of additional data, that a low effective-resistance bond, adequate for lightning protection, is also adequate for the suppression of radio interference.

Bonding jumpers for shockmounted equipment pose special problems because of the limited amount of space available for them, because they form part of the mechanical system in addition to serving as carriers of currents, and because good grounding is especially important for such shockmounted equipment as receivers. The development of special bonding jumpers for shockmounted equipment is described in detail in Appendix IV of this volume.

2.3 BONDING OF TUBING AND CONDUIT

The outer surface of long spans of conduit or tubing is a possible high-impedance path for interfering currents from sources outside the tubing or conduit. To minimize this possibility, such spans should be properly bonded to ground at both ends and several intermediate points.

Ordinary clamps cannot be used to bond flexible conduit because the pressure exerted on a comparatively small surface area of the conduit is sufficiently high to compress it or force it to give way. To overcome this a flared split-sleeve is fitted around the conduit, as shown in Figure 3-34, which distributes the high pressure delivered by the bonding clamp over a larger area, thereby resulting in a low unit pressure on the conduit. Contact is further improved by soldering the sleeve to the conduit, when the materials permit, through several holes in the sleeve provided for this purpose.

Figure 3-35 illustrates a method for bonding rigid conduit or tubing to a structure through supporting attachments. The number of mechanical supports required is generally adequate to provide an efficient bond even when the conduit is carrying interference signals.

The conduit or tubing to which bonding clamps are attached should be cleansed of paint and foreign material over the entire area covered by the clamp. All insulating finishes should be removed from the contact area before assembly, and anodized screws, nuts, or washers should not be used for attaching parts in making bonding contact. If, in

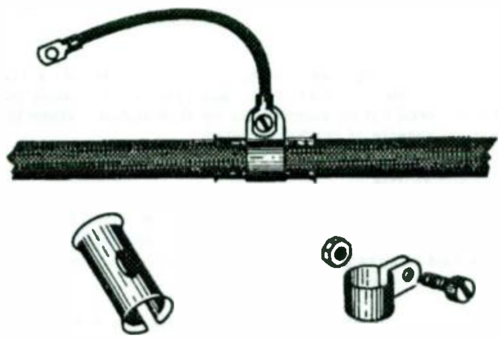


Figure 3-34. An Acceptable Method of Bonding Flexible Conduit

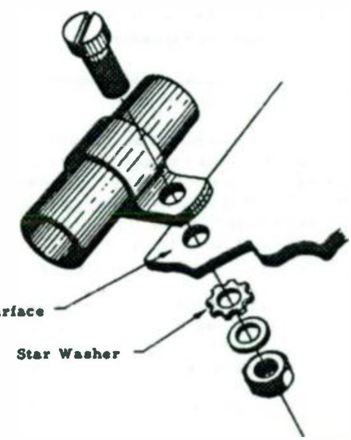


Figure 3-35. Tubing Clamp Bonded to Painted Surface

bolting the bonding clamp to the structural surface, a star washer is used, as shown in Figure 3-35, any protective coating (unless very thick or tough) need not be removed from this surface since the points of the washer penetrate to the bare metal.

3. GROUNDING

Ground systems have always been considered an important part of electrical circuitry; they have basic functions in dc, 60-cycle ac, telephone and telegraph circuits. Generally speaking, they provide the following three benefits:

a. Increased operational safety in electrical devices through establishing a zero potential for all metallic parts and structures with which persons might come in contact

b. A power "sink" for carrying off lightning strokes and dissipating excessive power from circuit malfunctions

c. An equipotential element to reduce possible interference-causing currents.

Grounding is not to be confused with bonding. Grounding refers solely to the electrical connection of points or equipment to insure that all related points in the same system operate at the same potential as the earth. Two methods of grounding in common usage are the single grounded system and the multiple grounded system. Ungrounded networks should be avoided except where unusual mounting and/or circuit conditions preclude the use of grounding methods.

A common ground for both signal and power circuits should be avoided as this can cause increased interference pickup in the signal circuit. Ground currents, which flow through the common ground return impedance, will cause voltages to appear in series with the signal voltage in the signal loop. A separate, low impedance ground return should be used in all signal circuits, avoiding chassis or structure for this ground return.

Consideration should be given to the mutual and self-impedance of the ground circuits (see Figures 3-36 through 3-39). The use of long common ground lines, with consequent high mutual impedance, should also be avoided in signal circuits which operate at widely different levels. In such specific cases, separate ground lines should be provided for each

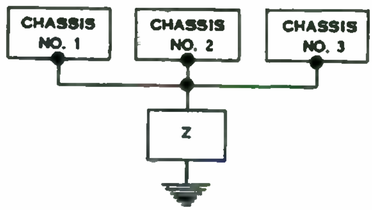


Figure 3-36. Ground System for Strap-Connected Chassis Returned to External Ground by a Single Line

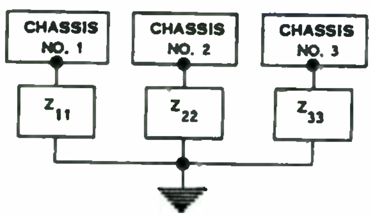


Figure 3-37. Ground Return Circuit with Insulated Chassis and Separate Lines to Ground

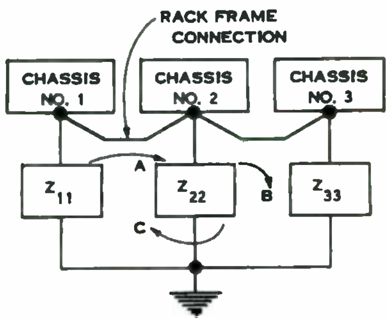


Figure 3-38. Multiple Ground Return Circuit with Rack Frame Connections

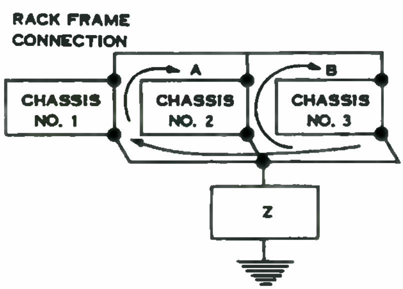


Figure 3-39. Ground System for Strap-Connected Chassis Returned to External Ground by a Single Line and Having Additional Interconnections Through the Supporting Rack

circuit to chassis frame. A short bonding strap should be used to ground chassis and circuits directly through chassis-to-frame ground, since long ground lines may be a source of induced interference. Under most operating conditions, a single ground line may still be used between the ground and chassis frame. However, in sensitive low level circuits, it may be necessary to provide a separate, isolated, shielded ground line for each circuit to remove the possibility of interference through ground loop currents. In these cases, each shield should be connected to the ground at only one point near the connection to the earth ground.

In circuitry design, the use of separate ground lines for sensitive circuits does not automatically preclude interference from mutual ground couplings. In addition, the mutual resistance and impedance of common ground lines, and the possibility of mutual inductive and capacitative effects remain in lines which are cabled together. Figure 3-40 and 3-41 show two methods of designing equipment input circuits which stress the mutual impedance coupling of the ground circuits. In these circuits, shunt capacitors are used at the input. Figure 3-40 shows the condition of these circuits when the input is connected directly to the chassis ground. In this case, mutual coupling, Z_{12}, is present due to stray input capacitances to ground or due to the filter capacitance at the input.

In Figure 3-41, the input circuitry is provided with high impedance dc ground return paths (R_3 and R_4). C_3 and C_4 are the stray capacitances of the input circuit (transformer case) to the chassis ground. The combination of C_3 and R_3 provides a high impedance path compared to that of C_1. As a consequence, C_3R_3 limit the flow of mutually coupled currents in the mutual impedance Z_{12}. The equivalent circuit is also included for comparison and clarification.

When the shield around a conductor is used to reduce interference pickup, it should be grounded at one end only to prevent current flow through the shield. When used to prevent radiation from a wire in a nonsensitive circuit, multiple grounds on the shield may be used. A general rule to follow is, "If the current flow through the shield will produce interference in the shielded circuit, ground the shield at one point only." When using single-point grounding on the shield, the ground point should be located near the low level or input side of the circuit.

Ground straps should not be connected from one unit of test equipment to another because circulating ground currents may flow through

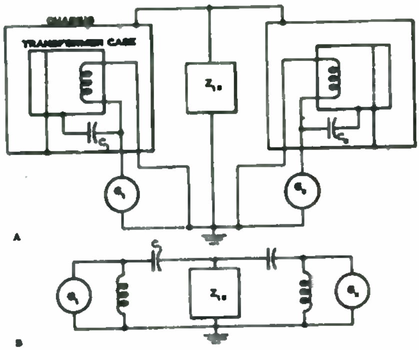


Figure 3-40. Normal Input Circuit Connection and Equivalent Circuit

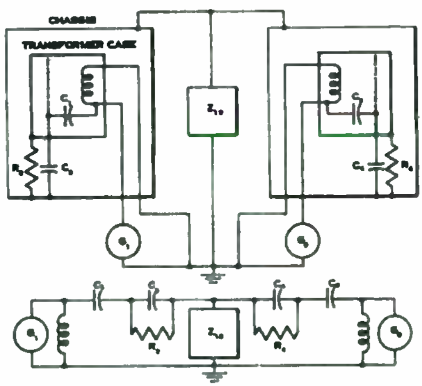


Figure 3-41. Modified Low Level Signal Circuit Connection and Equivalent Circuit

one of the units causing interference to appear on other equipment. When using ground straps, they should be connected between each equipment and ground. Decorative metal trim strips, cabinetry, and structures that are poor conductors should not be used as grounding points.

The use of the earth as a circuit element provides a conductor of high power capacity and low impedance; this is, in many cases, the most economical way to provide the desired functions.

The earth presents some undesirable traits however. Its impedance, although low in most cases, may appear high in others. Such factors as moisture content, degree of acidity and alkalinity may cause variations in the impedance. Also the fact that ground contact is made with a metallic part of the circuit makes corrosion an important consideration, since this may increase the ground impedance to the point of vitiating the ground system entirely.

An important consideration of RF grounding is the high impedance that can develop between the RF circuit and ground due to the connection itself. Conductors and transmission lines are effective solutions here; grounds such as ordinary wire or ground bus should be avoided.

Attention should be paid to the effect the distance above ground has on the impedance problem. Since communications/electronic equipment is normally in a building or tower which is physically higher than ground in terms of wavelength, the ground system used will be at least as long as that distance and, most likely, longer.

Finally, the frequency dependence of ground system impedance must be considered at radio frequencies. If the ground system represents an odd multiple of a quarter wavelength, its impedance will be very high; at even multiples, low impedance may be achieved, providing ideal transmission line characteristics exist in the ground system. The extent to which this ideal is approached determines the possibility of the low impedance condition.

3.1 SINGLE GROUND CIRCUITS

The single ground circuit provides a path to a single ground point from a number of equipment's and circuits. Two methods commonly employed in this type of circuit are shown in Figures 3-36 and 3-37. In Figure 3-36, each of the circuits or chassis is provided with short jumpers or straps connecting each other and with a single line connecting the equipment to the earth ground. In Figure 3-37, each circuit or chassis is returned to the earth ground through separate lines. With only a single path to ground in each circuit, no ground loop currents can flow.

3.2 MULTIPLE GROUND CIRCUITS

When additional ground circuits are provided as shown in Figures 3-38 and 3-39, ground loop currents may flow. Such loop currents may cause induced interference in low level circuits. Circulating ground currents are shown by arrows on both illustrations. Ground currents of this type should be avoided both in high level and low level circuits. This type of grounding should be used only if the circulating currents are known to be of a low level and the physical location makes this type of grounding unavoidable.

FILTERS AND SPECIAL CIRCUITS — CHAPTER 4

1. FILTERS AND OTHER SUPPRESSION NETWORKS

In this section, filters and other suppression networks will be considered from the design standpoint (Section 1.1), and the application of these design principles will be enumerated through a number of specific examples (Section 1.2).

The purpose of suppression networks, in general, is to attenuate interfering signals conducted from interference sources or into interference-susceptible equipment. Their importance is underlined by the fact that, no matter how well an interference source is shielded, a problem remains in that access must be provided through the shield for the transfer of energy, which involves leads capable of conducting currents away from the source. It is here, in fact, that filters and other like networks find their most frequent application: suppression of the conduction of all signals except the one desired.

The filter, itself, is a four-terminal network designed to freely transmit currents or voltages of certain frequencies while attenuating all others. To accomplish this, use is made mainly of reactive elements. Dissipative elements are usually avoided in practical filter networks since they act to prevent the free transmission of the desired signal.

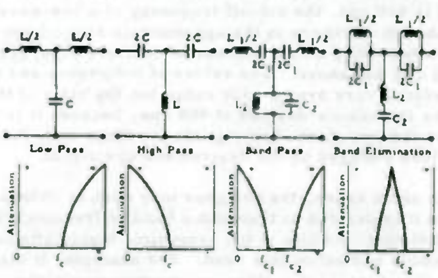


Figure 4-1. Typical Filter Sections Showing Attenuation As a Function of Frequency

Filters admit of a general classification according to the band of frequencies to be transmitted or attenuated: low-pass, high-pass, band-pass, and band-elimination. Typical examples of the simplest basic structures for each of these types, together with typical attenuation curves for each, are shown in Figure 4-1. The frequencies that separate transmission and attenuation regions are called the cut-off frequencies, f_c, of the filter as noted in the diagrams.

1.1 DESIGN CONSIDERATIONS

As was noted in Chapter I, the design of filters is an art as well as a science since so much depends on the judgment and techniques used by the design engineer. He must determine his requirements exactly, and then make a prudent selection of the suppression device. Much practical material is available to assist him, including the very useful tools of passive network synthesis.

In most practical interference-suppression problems, the main consideration is to pass freely a power or control current comprising a single frequency or at most a very narrow band of frequencies, while at the same time greatly attenuating all other frequencies. In many cases, this can be accomplished by designing a simple low-pass filter in accordance with the data given for element values in Appendix 1, Volume III. The designer of interference suppression filters usually will have a wide choice of parameters at his disposal, depending upon the ratio of the lowest frequency to be suppressed and the frequency to be transmitted, and also the attenuation requirement to be met. For example, if the interfering frequencies to be suppressed are above 100 kc while the frequency to be passed is 400 cps, the cut-off frequency of a low-pass type filter could be chosen anywhere in the approximate range from 500 cps to 90,000 cps depending on the number of decibels suppression required at 100,000 cps and above. The values of inductance and capacitance can, therefore, vary over a wide range but the ratio of their values is fixed by the impedance desired at 400 cps, because it is this ratio which determines the impedance looking into the filter and, therefore, the attenuation loss suffered by the desired 400 cps signal.

In some cases, the designer may wish to choose a band-pass filter when it is desired to transmit a band of frequencies without attenuation and distortion while at the same time highly attenuating all frequencies above and below this band. For example, it may be desired to transmit voice frequencies along an intercommunication system in aircraft, or the input circuit to a receiver may require a broad band filter.

Also, circuits within a receiver itself, such as audio output or intermediate frequency circuits, often require faithful transmission over a fairly wide band of frequencies while suppressing all other frequencies. In these cases, the designer first of all determines the approximate impedance level of the line into which the filter is to be inserted, the upper and lower cut-off frequencies, and the attenuation required for all other frequencies. By use of the formulas and charts given in Appendix 1 of this book, the correct values of elements can then be readily computed for maximum performance.

Unless the ratio of the upper to the lower cut-off frequencies is less than 2 to 1, it will usually be found advantageous to employ high-pass and low-pass filter sections in series. This advantage will be apparent if the size of the inductors and capacitors are computed both for band-pass and for low or high-pass sections.

It should be pointed out that there may be other good reasons for choosing a band-pass type filter, as, for example, to make use of impedance transformations to increase or decrease the current or voltage in the line (to avoid large magnetic or electric fields), or to obtain more easily realizable values of elements.

Sometimes, radio interference filters may require unique characteristics due to the complexity of present-day installations of electronic equipment. For example, a filter may be required with large attenuation from 150 kc to 1000 mc, a range that cannot usually be covered with a single-section filter. In these cases, where complete information may not be available over the entire frequency range that must be considered, a fair approximation of the impedance/frequency characteristic of the circuit at the frequencies to be passed suffices to work out the remaining network parameters.

For normal suppression conditions, however, the following enumeration of design goals is an effective guide in solving filter problems.

1.1.1 IMPEDANCE CONSIDERATIONS

In order to preserve the normal functioning of the equipment at both ends of a transmission line selected for insertion of a filter, the elements of the filter must be chosen so that the impedances looking into and out of the filter match those of the line.

The input impedance of any filter, as seen by the source, is either very high or very low over most of the attenuation band as compared to that required for a good match in the transmission band. This fact has an important consequence for the behavior of the circuit before the point at which the network is applied. If the source of interfering signals has a low internal impedance at those frequencies, a low impedance input network increases the amplitude of the interfering currents between it and the source. If there is an appreciable amount of coupling between these interfering currents and a receiver through, for example, magnetic induction, the suppressing network may increase the interference rather than decrease it. If, on the other hand, the source has a high internal impedance, a high impedance network increases the voltage across its input. As before, this voltage may affect a receiver through, say, capacitive coupling, so as to increase the interference rather than decrease it. In practice, the source may, at times, have a low internal impedance and at other times have a high internal impedance, depending on the frequency or mode of operation. Therefore, great care must be taken by the designer in the choice of suppressing networks to insure a reduction of all types of interference at the receiver and to avoid the pitfalls of reducing the effect of conducted interference at the expense of interference coupled to the receiver in other ways.

It must be pointed out that there are many cases in which the impedance of the filter at the frequency to be transmitted is of a secondary importance. If the current to be passed is direct current, then a low-pass filter does not change the impedance level at all, looking either way, except for the effects of possible coil resistance. Also, neither a series inductor, whose impedance is zero at dc, nor a shunt capacitor, whose impedance is infinite at dc, in any way affects any direct currents or voltages in the system. It can be shown that this statement is substantially true also for low frequency alternating currents, provided only that the frequency to be transmitted is sufficiently removed from the cutoff of the low-pass filter. To show this, the effect of the insertion of a symmetrical filter into a circuit is considered. Let a generator of voltage E having an internal impedance Z_s be connected directly to a load of impedance Z_R. Then the current is $E/(Z_s + Z_R)$ at all frequencies. After inserting a filter to suppress the radio interference frequencies, it is essential that the current at the desired frequency or frequencies be the same as before, but currents at all other frequencies be much less. If this can be accomplished, then the functional operation of the generator and the load is the same as before the insertion of the filter. Now let a filter, with image impedances Z_{I_1} and Z_{I_2} and with image transfer constant θ, be inserted as shown in Figure 4-2. Since the filter is assumed to be symmetrical, $Z_{I_1} = Z_{I_2} = Z_I$.

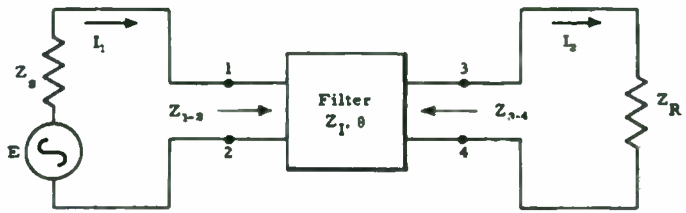


Figure 4-2. Insertion of Filter Between Generator and Load

The input current I_1 is given by the following equation:

$$I_1 \simeq \frac{E_o}{Z_o} \sin \omega t \left(1 - A_1 \frac{E_o \sin \omega t}{Z_o^2} - A_2 \frac{E_o^2 \sin^2 \omega t}{Z_o^3} \right) \qquad (4\text{-}1)$$

From this equation, the impedance Z_{1-2} looking into the filter terminated in Z_R may be obtained. It is given by

$$Z_{1-2} = Z_1 \left(\frac{1 + F_r' e^{-2\theta}}{1 - F_r' e^{-2\theta}} \right) \qquad (4\text{-}2)$$

where $F_r' = (Z_R - Z_I)/(Z_R + Z_I)$. Since it is required that the impedance as seen by the generator should still be Z_R at the desired frequency, the important relationship is that between Z_{1-2} and Z_R. The expression for the input impedance may be changed to the following form:

$$Z_{1-2} = Z_R \left[\frac{(1 + e^{-2\theta}) + \frac{Z_I}{Z_R}(1 - e^{-2\theta})}{(1 + e^{-2\theta}) + \frac{Z_R}{Z_I}(1 - e^{-2\theta})} \right] \qquad (4\text{-}3)$$

To obtain a clearer picture of the effect on the input impedance, assume that θ is small as compared to unity, so that $e^{-2\theta} \simeq 1 - 2\theta$. Then,

$$Z_{1-2} \simeq Z_R \frac{1 + \left(\frac{Z_I}{Z_R} - 1\right)\theta}{1 + \left(\frac{Z_R}{Z_I} - 1\right)\theta} \simeq Z_R \left[1 + \theta\left(\frac{Z_I}{Z_R} - \frac{Z_R}{Z_I}\right)\right] \qquad (4\text{-}4)$$

It is seen that $Z_{1-2} = Z_R$; i. e., the impedance as seen by the generator remains unchanged, when either $\theta = 0$ or $Z_R = Z_I$.

In a low-pass filter, both θ and Z_I vary with frequency. In the pass band, θ is imaginary, $\theta = j\beta$. The phase shift, β, is given by:

$$\beta = \tan^{-1}\left[2\frac{f}{f_c}\,\frac{1 - \left(\frac{f}{f_c}\right)^2}{2\left(\frac{f}{f_c}\right)^2 - 1}\right] \qquad (4\text{-}5)$$

where f is the frequency and f_c is the cut-off frequency of the low-pass filter. If f/f_c is small, β is also small, and one may write, approximately,

$$\beta \simeq -2\frac{f}{f_c} \qquad (4\text{-}6)$$

The image impedance of a constant -k low-pass filter varies with frequency as follows:

$$Z_I = R\sqrt{1 - \left(\frac{f}{f_c}\right)^2} \simeq R \qquad (4\text{-}7)$$

where R is the design resistance of the filter. Combining these equations, one obtains:

$$Z_{1-2} = Z_R\left[1 + 2\frac{f}{f_c}\left(\frac{R}{Z_R} - \frac{Z_R}{R}\right)\right] \qquad (4\text{-}8)$$

This shows that, in order to keep Z_{1-2} substantially equal to Z_R, the term

$$2 \frac{f}{f_c} \left(\frac{R}{Z_R} - \frac{Z_R}{R} \right) \tag{4-9}$$

must be kept small. To do this, neither f/f_c nor R/Z_R (or Z_R/R) can be large.

In the Figure 4-3, the quantity Z_{1-2}/Z_R is plotted as a function of the mismatch-ratio, R/Z_R, for different values of f/f_c. For example, it is seen that, when f = 400 cps and f_c = 40 kc, so that f/f_c = 0.01, the ratio R/Z_R may be as small as minus one and as large as ten without changing the impedance level more than 20 per cent . If a change of no more than 10 per cent is desired, either the cut-off frequency must be raised or the ratio R/Z_R must be made closer to unity.

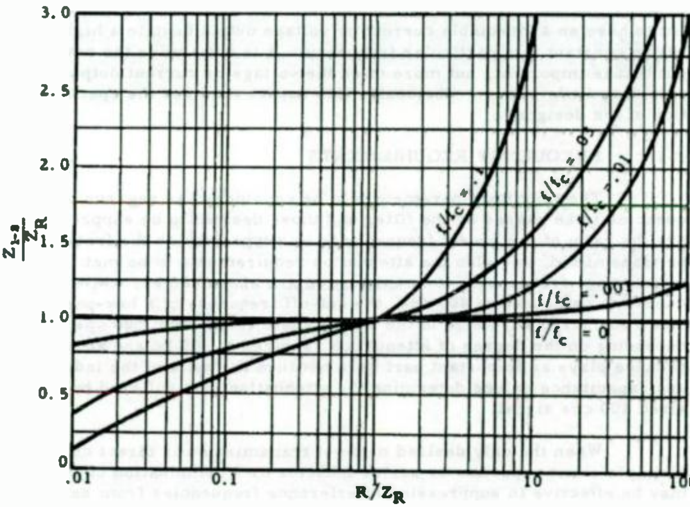


Figure 4-3. The Effect of Mismatch on Input Impedance of Low-Pass Filter

The equations derived hold equally well for both sides of the filter. From the relationships given, it follows that, if the ratio of Z_s to Z_R is very high or very low, it is most desirable to choose R near the geometric mean, $\sqrt{Z_s Z_R}$, in order to keep both R/Z_R and R/Z_s as close to unity as possible.

Terminating impedances are not constant with frequency and are usually not even known quantities, which leads to considerable complications in filter design. Most filter design considerations, as found in standard text books, are based on constant resistance terminations. In this case, no distinction need be made between transmission of power on the one hand, and that of currents or voltages on the other because, for a resistive termination, there exists the simple relationship $P = I^2R = E^2/R$, so that, when the power is zero, both current and voltage must also be zero. This is not true when the terminating impedances can be reactive and can even assume values of zero or infinity; i. e., become resonant or anti-resonant, at certain frequencies. Hence, it is possible for a filter to have an appreciable current or voltage output despite a high attenuation constant at a particular frequency. It is not always the power output that is important, but more often the voltage or current output, even with very little power. Obviously, this aspect requires the special attention of the designer.

1.1.2 FREQUENCY REQUIREMENTS

The important parameters to be considered as regards the frequencies to be passed by the filter and those desired to be suppressed are the ratio of the lowest frequency to be suppressed to the frequency to be transmitted, and also the attenuation requirements to be met. If the interfering frequencies to be attenuated are above 100 kc, while the frequency to be passed is 400 cps, the cut-off frequency of a low-pass filter could be chosen anywhere in the approximate range from 500 cps to 90 kc, depending on the degree of attenuation required at 100 kc and above. Impedance plays an important part here because the ratio of the inductance and capacitance values determine the attenuation loss suffered by the desired 400 cps signal.

When the only desired mode of transmission is direct current, a simple shunt capacitor or series inductor or a combination of the two may be effective in suppressing interference frequencies from zero to infinity. This type of network is often called a "brute-force" filter (Section 1. 2. 1) because it does not satisfy the previously mentioned requirements for filters. Since these networks are often used to suppress interference in circuits where only direct currents are to be passed, they are usually designed without regard to specific frequency or impedance requirements.

Typical arrangements of elements for "brute-force" filter networks are shown in Figure 4-4.

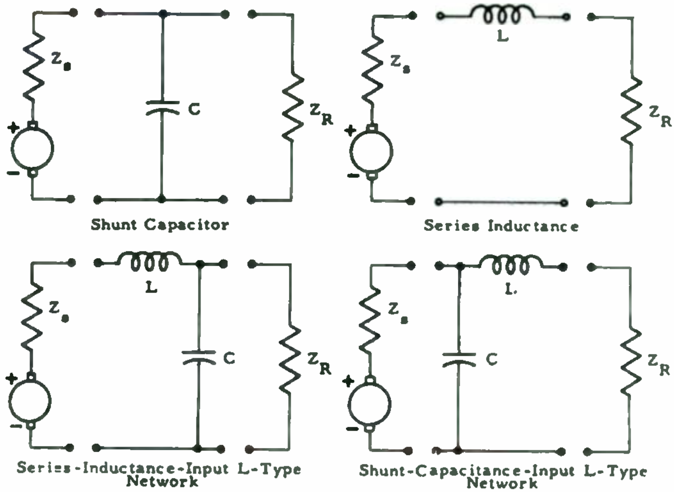


Figure 4-4. "Brute-Force" Filters

If filters are to be constructed for frequencies that are too high for lumped elements, sections of transmission lines must be used instead. Because of the difficulty in constructing filters that are effective over very wide ranges of frequencies, it may be necessary at times to use two or more filters in series to cover the entire frequency band.

1.1.3 FILTER LOCATION

The location of the filter or other suppressing network is another item that must receive careful consideration. It is evident that one filter at each of the possible input paths to the receiver will take the place and do the work of many filters at the outputs of all the interference sources. On the other hand, the effect of increasing interference currents or voltages before the point of application of the filter obviously increases as the filter is moved further from the source and is minimized by placing the filter at the source. Care must also be taken that filter coils and leads are not themselves placed in interfering fields. For example, by bunching the output leads of a filter together with the input leads, or by placing the coil of a filter close to a coil carrying interfering currents, the effect of the filter may be completely nullified. To avoid this last possibility, it may be necessary to shield the filter coil or perhaps use special core designs. All these factors must be considered in arriving at a decision as to the best location of the filters. Figure 4-5 shows possible filter locations.

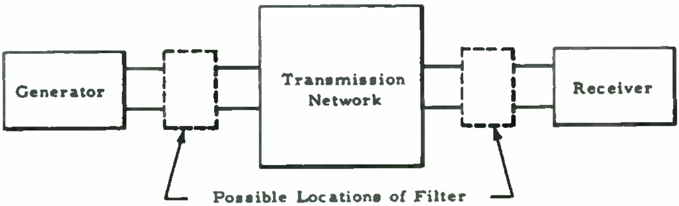


Figure 4-5. Shows Possible Filter Locations

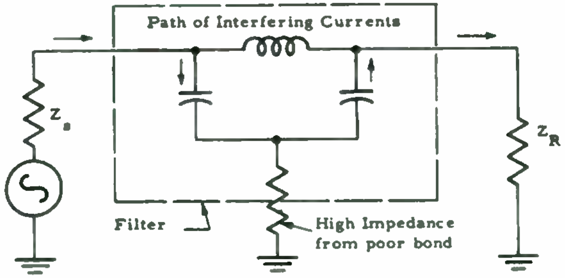


Figure 4-6. Effect of Poor Grounding on Filter Effectiveness

1.1.4 GROUNDING

The extreme importance of good grounding of filters must be emphasized. A filter can be made completely ineffective by improper grounding techniques. Figure 4-6 shows how a high impedance bond to ground offers the interfering currents an effective by-pass path, thus vitiating the purpose of the filter

1.1.5 EFFECTS OF REACTIVE ELEMENTS

The fact that filters are composed of reactive elements which cannot dissipate any energy leads to a considerable increase in the energy that must be dissipated in the network before the filter. This is not harmful if the filter is placed at the output of an interference source which is well shielded and has no other possible outlets for the interfering energy. In this case, the energy is completely bottled-up, the filter acting as an effective stopper, and is dissipated inside of the shield, where it can do no harm. When shielding is not complete, the reflection of the energy back into the source may be harmful and lead to increased leakage; but, it must be remembered that one function of a properly designed filter is to prevent the generation of real power at the frequencies to be attenuated, by offering a reactive impedance at these frequencies. If this function is properly performed, there is no need for increased dissipation of real energy. There exists apparent, or reactive, power which fluctuates back and forth between the filter and the source and need not be dissipated.

1.1.6 ADDITIONAL LIMITATIONS

As has been noted, very adequate treatments of filter design can be found in a variety of textbooks and handbooks. The design formulas do, however, have limitations which should be clearly recognized.

The first of these limitations, the resistance of coils, is especially troublesome for direct currents since it may cause an excessive voltage drop across the coil and a corresponding reduction of useful output. It is also of great importance in the design of wave traps where a very narrow band of frequencies is to be attenuated. It is also of importance in obtaining sharp cutoffs in filters; but, as stated elsewhere, most filters used for interference reduction do not require sharp cutoff characteristics. The resistance of a coil is usually measured in terms of a figure of merit, called the Q of the coil, and is defined as the ratio of the inductive reactance to the resistance. Usually, over a limited range of radio frequencies, it is found that the Q of a coil is more nearly constant than its resistance. If the coil is used in combination with a capacitance, either in series or in parallel, then the Q specified usually refers to the resonant or antiresonant frequency of the combination.

A second limitation applies particularly at higher frequencies. The distributed capacitance of an inductance coil may cause it to have a negative reactance at frequencies as low as 10 kc , but this frequency may be greatly increased by proper design. Also, the lead inductance of a capacitor may cause its reactance to become inductive, and for capacitors with external leads this usually happens between 100 kc and 10 mc, approximately. In the feed-through capacitors which have no external leads in series with the capacitance, this frequency may be increased to 1000 mc and higher. Obviously, filters will lose their effectiveness when their elements undergo such radical changes.

A third limitation, that of stray coupling between coils and leads, is troublesome at all frequencies. It may be inductive or capactive. It occurs most frequently between the input and output leads of a filter, between the series arms of a T network, and between the shunt arms of a pi network.

1.2 TYPES OF FILTERS AND SUPPRESSION NETWORKS

Three types of suppression networks are most important for the suppression of conducted radio interference. The "brute-force" filters, consisting of a single capacitor, a single inductor, or a simple L-section, are used mostly in direct current circuits; power-line filters are used in 60, 400, or 800-cycle alternating current circuits either at the output of a source of interference or at the input of a receiver; and harmonic-suppression filters are used in the output of transmitters. These types are treated in the next three sections. Following them, the design of individual elements for suppression networks is discussed in detail.

1.2.1 "BRUTE-FORCE" FILTERS

The simplest type of suppression network is a single capacitor connected from a conductor carrying interfering currents to ground. Such a capacitor offers a low impedance to ground to the interfering currents while offering infinite impedance to direct currents. It is very effective as long as the capacitor behaves like a capacitance, i.e., as long as the inductance of the capacitor leads is negligible. The problems associated with this network are those of capacitor design, to be discussed in Section 1.2.4 and 1.2.5, and those of proper installation, to be discussed here.

The capacitor must be installed as close to the actual point of interference generation as is physically possible. For equipments enclosed in a case or shield, this always means that the capacitor must be located within the case or shield. In the case of direct current motors or generators, the capacitor should be connected directly across the brushes. Grounding of the capacitor directly to the negative brush is preferable to grounding it to the frame or case. An exception to this occurs in the case of feed-through capacitors used when the interference source is completely shielded (see Section 1.2. 5).

It is extremely important to install the capacitor in such a way that the connecting leads are as short as possible. Any lead wire has inductance, and at a certain frequency this inductance resonates with the capacitance. Above the resonant frequency, the effectiveness of the capacitor for the reduction of radio interference decreases very rapidly. This is shown by Figure 4-7, which gives the insertion loss as a function of frequency for a 4μf condenser, installed in a 28-volt direct current line operating into a 10-ohm resistive load. Curve A is plotted for a condenser connected by means of a lead wire 4 inches long; Curve B is plotted for a condenser with connections made directly at the terminals. In general, the insertion loss in decibels of an ideal capacitor, connected across a load Z_R which is fed by a generator of internal impedance Z_s, is given by the expression:

$$\text{Insertion Loss} = 20 \log \left| 1 + j\,2\pi f C \frac{Z_s Z_R}{Z_s + Z_R} \right| \qquad (4\text{-}10)$$

where f is the frequency in megacycles per second, C the capacitance in microfarads, and the impedances are in ohms. If both Z_s and Z_R are pure resistances, i.e., $Z_s = R_s$ and $Z_R = R_R$, then this reduces to:

$$\text{Insertion Loss} = 10 \log \left[1 + \left(2\pi f C \frac{R_s R_R}{R_s + R_R} \right)^2 \right] \qquad (4\text{-}11)$$

Figure 4-8 gives the resonant frequencies of capacitors of various sizes plotted versus the total length of both leads. These curves must be considered approximate since the geometrical arrangements of the leads and the internal construction of the capacitor itself affects the resonant frequency to a considerable extent. The curves shown are based on experimental data, but similar curves (actually straight lines) would be obtained by assuming that the leads have an inductance of 25milli-microhenries per inch. Such an inductance would exist if the leads are run parallel to a ground plane and three wire diameters away from it.

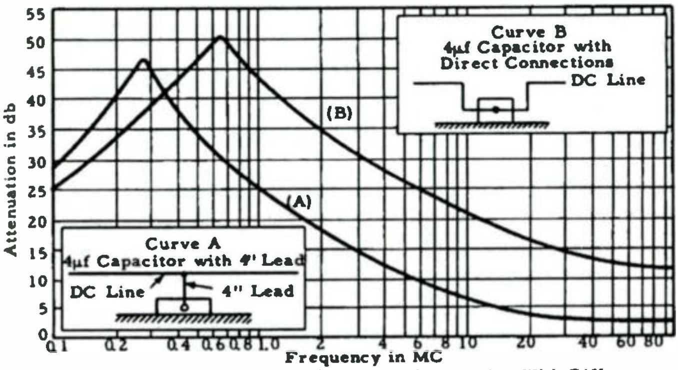


Figure 4-7. Difference in Capacitor Attenuation With Different Arrangement of Leads

When a capacitor is used with an alternating current generator, additional requirements must be met as follows:

a. The resonance of the capacitor with the internal inductance of the generator at the fundamental or the lower harmonic frequencies must be avoided to prevent the build-up of dangerously high currents and voltages.

b. The capacitance must not be so large as to increase the generator output sufficiently to cause overheating.

c. The excitation of a generator with automatic voltage regulation must not be increased by the capacitor current more than the amount for which the regulator can compensate.

To meet these requirements, the maximum value of capacitance that can usually be used on alternating current generators is 0.5 µf with a value of 0.05 µf preferred for most applications. Since alternating current generators have no commutators, the interference generated by them is not as severe as that generated by direct current generators, and the reduction of radio interference afforded by a 0.05 µf capacitor is usually sufficient.

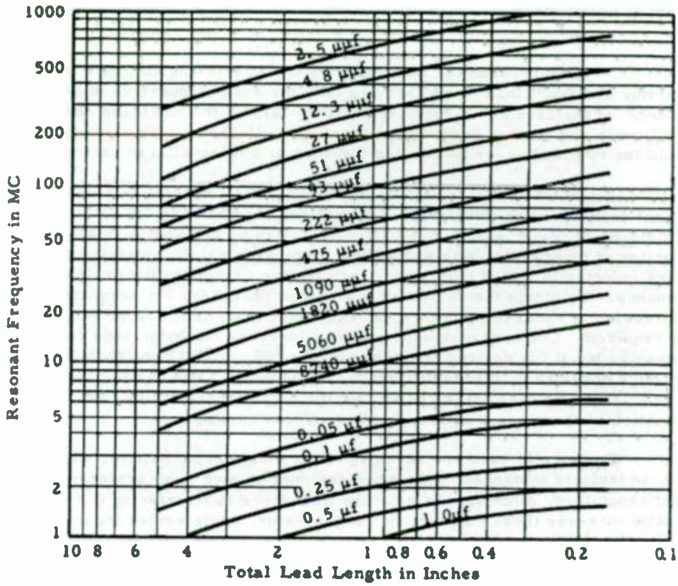


Figure 4-8. Series Resonance of Small Capacitors as a Function of Lead Length

Another important factor in the installation of capacitors is provision for a good ground connection. Since the purpose of the capacitor is to provide a low impedance path, and since the grounding connection is in series with the capacitor itself, it is clear that the connection must be a low impedance bond at all frequencies at which the capacitor should be effective. A single-terminal capacitor in a grounded metal case is most effective. The grounding connection should be a bare metal-to-metal contact with high pressure maintained mechanically. An exception to this occurs when the negative lead of the capacitor must be connected directly to the negative side of the interference generator. In this case, a capacitor with two terminals should be selected, and both leads should be kept as short as possible.

Coupling between the lead carrying the interfering currents to the capacitor and the "clean" lead carrying only the desired currents must be avoided. In the case of a single-terminal capacitor, the two leads should be collinear and pointed in opposite directions. Under no circumstances should the two leads ever be bundled together or run parallel to each other.

Instead of using a single capacitor, a single inductor in series with the lead carrying interfering currents can also be used. This is very rarely done because usually more attenuation can be obtained with a shunt capacitor of comparable size and weight. The greatest disadvantage of the series inductor is that it must carry the full line current at the desired frequency. To obtain the amount of inductance necessary for adequate suppression of radio interference, an excessive amount of copper is usually required. Cores of magnetic material increase the inductance considerably, but if the desired frequency is 60, 400, or 800 cps, such cores introduce undesirable losses at the power frequency. For use in direct current leads, no such losses exist, but the direct current may produce saturation of the core, thus lowering the effective inductance.

Single coils have one advantage over single capacitances, which may, in isolated instances, induce the designer to give them preference: Shunt capacitors, since they offer a low impedance to interfering currents, actually increase these currents in the generator, while series inductors, which offer a high impedance to the interfering currents, decrease these currents everywhere.

The design of inductors will be treated in detail in Section 1.2.6. As far as the installation of inductance coils is concerned, the most important consideration is to keep the coils out of any interfering fields. If no interference-field-free location can be found, the coils must be shielded. If they are allowed to pick up any interference through inductive or capacitive coupling, their effectiveness is obviously lost.

The upper frequency limit of an inductance coil is determined by its distributed capacitance. At some frequency this capacitance will resonate with the inductance, and at frequencies above that, the effectiveness of the coil in suppressing radio interference decreases rapidly. The distributed capacitance of a 100-millihenry iron-core coil is about 3 to 15 μμf, depending on the kind of winding. The resonant frequency of this coil could be no higher than 250 kc, and it would not be suitable for radio interference suppression. An inductance of 100 μh is usually sufficient, and with proper care in design (see Section 1.2.6) the resonant frequency of such a coil may be as high as 20 mc. For the suppression of frequencies above 20 mc, a single coil can be used only if inductance values below 100 μh are sufficient.

Finally, "brute-force" filters may be constructed by combining shunt capacitors and series inductances into L- or Pi-sections as shown in Figure 4-9. These sections give higher attenuation than single elements and should be used when the attenuation afforded by single inductors or capacitors is insufficient.

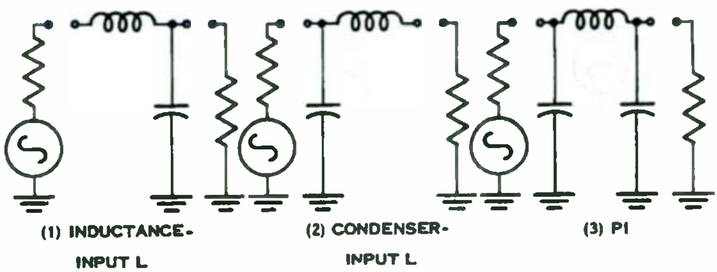


Figure 4-9. "Brute-Force" Filter Sections

The individual requirements for the installation of single capacitors or single inductors apply equally when such elements are used in combination. The requirement for good grounding connections acquires special significance for the Pi-section, as is evident from Figure 4-10. Here a poor bond actually allows the interfering currents to by-pass the inductance coil and return to the line thus vitiating the purpose of the network.

The choice between a condenser-input and an inductance-input L-section is determined by the following considerations: The inductance-input L-section (Figure 4-9) offers the higher input impedance to the interfering currents; therefore, it should be chosen whenever there is a reason for making it desirable to reduce the interfering currents before the point of application of the suppressing network. Such reasons might be the prevention of overloading of a generator, or the prevention of inductive coupling of these currents to other circuits. In the absence of such reasons, the condenser-input L-section (Figure 4-9) should be chosen because it results in greater attenuation. This is proved for a somewhat special case in Appendix II.

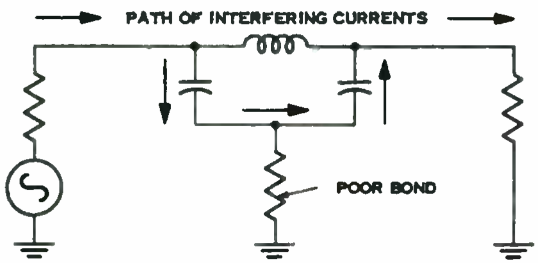


Figure 4-10. Effect of Poor Bond in Pi-Section

In a Pi-section, great care must be taken that there be no coupling, inductive or capacitive, between the two capacitors. Figure 4-11 shows how such coupling allows the interfering currents to return to the line. Such coupling may be prevented by extreme care in the installation or, in severe cases, by complete shielding of at least one of the capacitors. More details about these procedures will be found in the following sections.

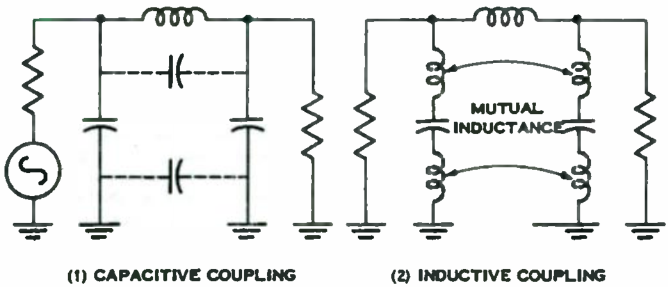
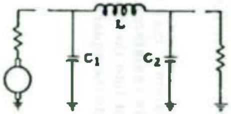


Figure 4-11. Coupling Between Condenser Leads in Pi-Section



Remarks

Determine wire size for coil from table on right.

Voltage rating of capacitors should be about twice generator or load voltage. (50 volts recommended for 28 volt systems).

	L	C_1	C_2
150 kc to 20 mc	12 Turns single cotton enameled magnet wire on 3/4" diameter form, close wound. Laminated iron core 3/8" x 5/8" with sufficient length to accommodate winding. Laminations 0.025" thick.	4 µf Single terminal (grounded metal case)	1.3 µf Single terminal (grounded metal case)
20 mc to 100 mc	10 Turns enameled magnet wire on 1/2" diameter form, space wound (equivalent to diameter of conductor).	0.5 µf Single terminal (grounded metal case)	0.01 µf mica (leads as short as possible)
100 mc to 150 mc	8 Turns enameled magnet wire on 1/4" diameter form, space wound (equivalent to diameter of conductor).	500 µµf mica (leads as short as possible)	500 µµf mica (leads as short as possible)

Max. Current in Amps	Wire Size	Diameter in Mils	k* No. of turns per inch for close windings: EN & SSC	k* No. of turns per inch for close windings: DSC & SCC
1.5	22	25.3	37	34
2	21	28.5	32.9	29.8
4	18	40.3	23.7	22.0
6	16	50.8	18.9	17.9
10	14	64.1	15	14.2
15	12	80.8	12	11.5
25	10	102	9.6	9.3
40	8	128	7.6	7.4
70	6	162	----	----
100	4	204	----	----
175	2	258	----	----

* Approximate only since thickness of insulation depends on type of insulation and manufacturer.

EN - Enamel
SSC - Single Silk Covered
DSC - Double Silk Covered
SCC - Single Cotton Covered

Figure 4-12. Design Instructions for Brute-Force Pi-Section in Direct Current Systems

ERC

Figure 4-12 gives three designs of practical brute-force Pi-sections. For complete suppression of all frequencies from 150 kc to 150 mc, the three filters must be used in series. These design instructions are approximate and should be used only as a guide.

1.2.2 POWER LINE FILTERS

Power line filters are used at the output of interference sources such as motors, generators, or inverters, and at the input power lines of receivers. Filters of this type are also required for the leads carrying power into shielded rooms. In this last case, the power frequency is usually 60 cps, but in all other cases the power frequency will be assumed to be 400 cps.

Power line filters for the suppression of radio interference are low-pass filters. If they are to operate efficiently, i.e., if they are to transmit the power frequency with little or no losses, a knowledge of the impedances between which they are to operate is essential for their design. An exact knowledge of these impedances is practically never available, but an approximate value can usually be obtained at the power frequency. Fortunately, as was explained in Section 1.1.1, the effect of an impedance mismatch in the pass band can be kept negligibly small by choosing the cut-off frequency high enough. As was pointed out there, to reduce the effect of the filter on the desired power currents to a minimum, the image impedance of the filter should be chosen near the geometric mean of the impedances between which it is to operate. In the attenuation band, an impedance mismatch usually introduces additional losses, a condition which is desirable, or at worst it introduces a reflection gain, which is never more than a small fraction of the total attenuation. Therefore, in the design of power line filters, it is only necessary to consider the impedances at the power frequency.

Most loads, such as receivers and lighting systems, may be assumed to have a purely resistive input impedance. This resistance may be computed either by dividing the normal input current into the nominal input voltage, or by dividing the normal input power into the square of the nominal input voltage:

$$R = \frac{V}{I} = \frac{V^2}{P} \qquad (4\text{-}12)$$

depending on whether the load is rated in terms of current or power. For example, a 110-volt load rated at 500 ma would have a resistance of 110/.5

or 220 ohms. The same load might be rated at 55 watts, and its resistance would be computed as $110^2/55$ or 220 ohms.

Loads such as motors or relays usually have inductance as well as resistance. For practical purposes it is best to ignore the inductance and design the filter on basis of the resistance only. For this purpose, the exact value of the resistance may be obtained by dividing the square of the current into the power, $R = P/I^2$, but the error in using Equation (4-12) is small. Since an approximate value of resistance is sufficient, the errors caused by assuming unity power factor are not significant.

In conventional filter design, much attention is given to maintaining a good impedance match over the entire pass-band. The development of m-derived terminating half-sections was motivated entirely by this necessity. In power-line filters only one frequency is of importance in the pass-band. Therefore, wide-band matching is not a problem. However, it may still be desirable to use m-derived sections in order to meet specific attenuation requirements.

The image impedance of a conventional low-pass filter of the constant-k or m-derived type is exactly equal to the design resistance R only at zero frequency. For other frequencies in the pass-band, for a Pi-section, it is given by:

$$Z = \frac{R}{\sqrt{1 - (f/f_c)^2}} \tag{4-13}$$

and for a T-section, it is

$$Z = R\sqrt{1 - (f/f_c)^2} \tag{4-14}$$

where Z is the input impedance, f the frequency, and f_c the cut-off frequency in the same units as f. For a radio interference filter, f_c may not be lower than 100 kc. With a power frequency f of 0.4 kc, it is obvious that for all practical purposes Z may be taken equal to R.

Once the cut-off frequency and the value of R have been decided upon, the design may proceed according to the procedure outlined in Appendix 1. For the installation of the filter, all considerations outlined in Section 1.1 for the installation of capacitors, inductance coils, and "brute-force" filter sections apply. When the filter consists of more than one section, say two constant-k and one m-derived sections, coupling of any sort between sections must be prevented. Where large attenuation is required, each section should be enclosed separately in a complete shield.

Examples of the design of typical filters for use as radio interference filters are in Appendix 1.

1.2.3 HARMONIC SUPPRESSION FILTERS

Harmonic suppression filters are used at the output of transmitters to prevent any harmonic of the desired transmitter frequency from reaching the antenna. They are band-pass filters, or possibly low-pass filters, since the frequencies below the desired transmitter frequency are not usually of interest. The cut-off frequency of such a filter should be between the desired fundamental, f_d, and its second harmonic, $2f_d$. A value of $1.5f_d$ is usually a good choice.

Harmonic suppression filters are inserted between the output of the transmitter and the transmission line feeding the antenna, or between the transmission line and the antenna. Assuming that the system is designed so that there would be a good impedance match without the filter at both the transmitter and the antenna ends, the filter should be designed to operate between impedances equal to the characteristic impedance of the transmission line. For coaxial lines, the commonly used impedances are 52 and 75 ohms.

Conventional filters using a combination of constant-k and m-derived sections are effective even at ultra-high frequencies. The only special consideration at the higher frequencies is that short sections of transmission lines must be used as filter elements.

The design objectives for a harmonic-suppression filter are the following:

a. In the pass band the filter must perform electrically as though it were simply another section of conventional transmission-line cable (52 ohm or 75 ohm, or any other impedance which may be required).

b. The attenuation characteristic should be such that a certain minimum must be exceeded over a very wide band. A good choice would be to require an attenuation of not less than 60 db between 10% above and four times the cut-off frequency (to attenuate the strong lower harmonics), and of not less than 30 db between four and ten times the cut-off frequency.

c. In the pass band, the insertion loss should be less than 0.5 db and the voltage standing-wave ratio caused by the filter when properly terminated should be less than 1.10 over the complete pass-band range.

d. The physical size and weight of the filter should be held to a minimum consistent with good engineering practice.

e. The average and peak power capacity must be sufficient for the transmitter considered.

f. The filter should be hermetically sealed to insure satisfactory operation under all atmospheric conditions encountered in flight.

If attenuations greater than those stipulated under Objective (b) are required it is better to use two or more complete filters in series than to attempt the design of a filter meeting more severe requirements.

The individual filter elements are extremely critical and may be constructed so that they can be tuned after the complete filter has been assembled. One satisfactory method of accomplishing this consists of varying the capacity of the condensers by rotating the dielectric spacers between parallel plates. The spacers are shaped like cams so that their rotation effectively controls the dielectric constant and thus changes the capacitance. After tuning, all of the adjustments must be permanently locked.

It is found that in a filter of this type, the insertion loss in the pass band is due primarily to the reflection coefficient and not the losses due to resistive components. Therefore, the voltage standing-wave ratio in the pass band with the filter properly terminated is an excellent indication of the filter's pass-band performance and efficiency. To keep the reflection coefficient small over the entire pass band, the input impedance of the filter should be as constant as possible. Terminating m-derived half-sections can be constructed specially for this purpose. A value of m near 0.6 gives the best results. But if the image impedance of such a half-section is plotted as a function of frequency, it is seen that there is a rapid decrease in its magnitude in the region from about 90 to 100 percent of the mathematical cut-off frequency. A portion of this region must necessarily be considered as belonging in the attenuation band, if low values of voltage standing-wave ratio are to be obtained in the pass band. Therefore, the design cut-off frequency f_c' should be taken as a little lower than the mathematical cut-off frequency f_c.

A good rule to follow is

$$f_c' = 0.965\, f_c \quad \text{or} \quad f_c = 1.035\, f_c' \tag{4-15}$$

To increase the sharpness of cut-off, a value of m a little below 0.6 for the terminating half-sections is desirable. If a value of 0.5 is chosen, it must be remembered that the image impedance of such a section is a little above the nominal image impedance R_o (i.e., the image impedance used in the design of the filter) for most of the pass band. A good choice for the relation between the actual load impedance R_L and R_o is

$$R_o = 0.925\, R_L \quad \text{or} \quad R_L = 1.081\, R_o \tag{4-16}$$

At frequencies above 70 or 80 mc, it becomes convenient to use short sections of coaxial transmission lines as filter elements. These lines are used to simulate the required values of lumped inductance and capacitance. The dimensions of the lines required may be computed in the following way:

Any uniform lossless transmission line of characteristic impedance Z_o and electrical length, θ radians, can be exactly represented by the equivalent circuit shown in Figure 4-13.

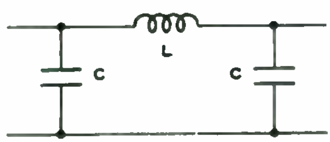


Figure 4-13. Equivalent Circuit of Transmission Line

The elements have the following values:

$$L = \frac{Z_o \sin\theta}{2\pi f} \tag{4-17}$$

$$C = \frac{\tan(\theta/2)}{Z_o 2\pi f} \tag{4-18}$$

where Z_o is in ohms and f, the frequency, in cycles per second. When θ is small (less than 15°), these values can be replaced to a good approximation by:

$$L \text{ (in } \mu\mu h) = 84.7\, Z_o\, d \tag{4-19}$$

$$C \text{ (in } \mu\mu f) = \frac{42.3}{Z_o}\, d \tag{4-20}$$

where d is the actual length of the line in inches. It is seen that for an electrically short line, the equivalent lumped parameters are independent of frequency, which is an important requirement if the line is to simulate constant elements. It is necessary, therefore, that the line be less than 15-electrical degrees long at all frequencies for which the filter is to maintain its essential properties.

The characteristic impedance Z_o of a coaxial transmission line using a solid dielectric is plotted in Figure 4-14 as a function of the ratio r of the inner diameter of the outer conductor to the outer diameter of the inner conductor for several values of the dielectric constant e of the medium between the two conductors. The analytical expression for the characteristic impedance is

$$Z_o = \frac{138}{\sqrt{e}} \log_{10} r \tag{4-21}$$

An abrupt change in either the inner or outer conductor diameter of a coaxial transmission line causes a distortion of the field which produces an admittance that can be represented electrically as a shunt capacitance across the line at the place of discontinuity. The value of this discontinuity capacitance is appreciable and must be considered in high frequency filter design where short sections of transmission line are employed as lumped elements. Two such discontinuities are pictured in

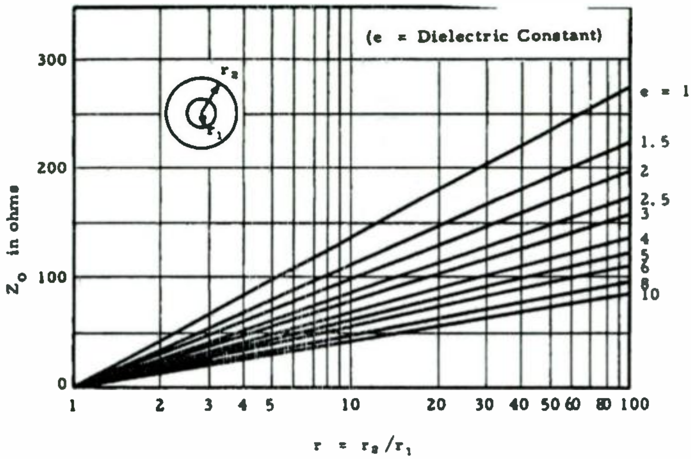


Figure 4-14. Characteristic Impedance of Coaxial Transmission Lines

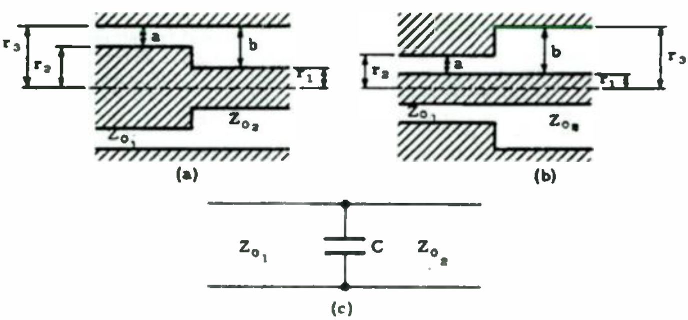


Figure 4-15. Discontinuities in Coaxial Lines and Equivalent Circuit

Figure 4-15 together with their equivalent circuit. The values of the shunt capacitances may be found from the curves on Figure 4-16. In each case, to obtain the actual shunt capacitance in micromicrofarads, the value read from the curve must be multiplied by the circumference of the unchanged conductor measured in centimeters. If there is a simultaneous discontinuity in both the outer and the inner conductor, an approximate but very good value of the equivalent shunt capacitance is obtainable by breaking it into two parts and using the same curves as before. That is, the capacitance is the sum of the two capacitances obtained by assuming first that the inner conductor remains continuous while the outer one changes, and then that the outer conductor remains continuous while the inner one changes.

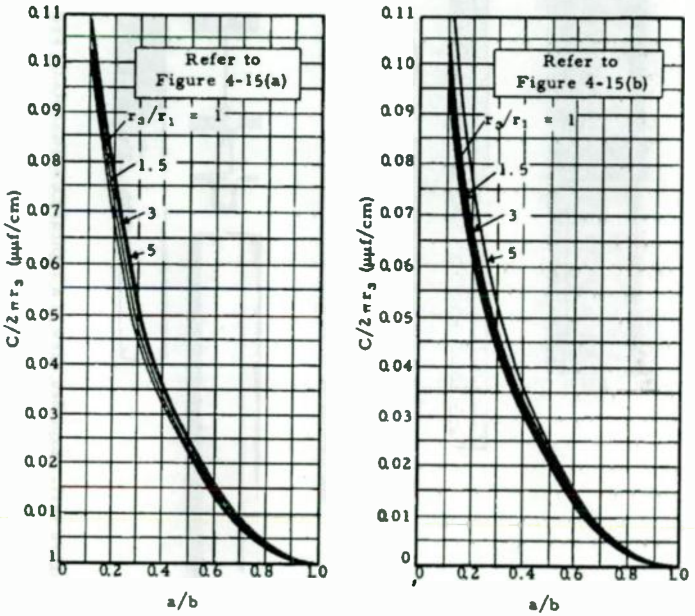


Figure 4-16. Curves for Computation of Discontinuity Shunt Capacitance in Coaxial Lines

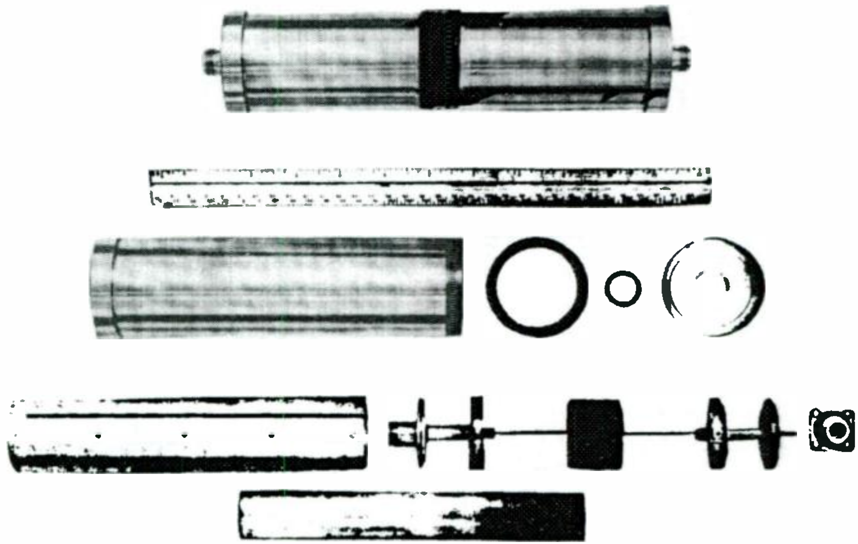


Figure 4-17. Exploded View of a Typical Low Pass Radio Frequency Filter
(52 ohm image impedance, 500 mc cutoff frequency,
50 db attenuation at frequencies over 550 mc)

An exploded view of a filter designed and constructed in accordance with the procedure outlined in this Section is shown in Figure 4-17.

Low-pass filters of the type described here are also sometimes useful in the antenna input circuit of receivers. While in a well-designed receiver the antenna input circuit acts as a band-pass filter attenuating all frequencies above and below its pass band, it may happen that a very strong signal outside the pass band overrides the desired signal despite this attenuation, so that an additional external filter becomes necessary.

1.2.4 FILTER CAPACITORS

Capacitors for use in radio interference suppression networks must meet the following mechanical requirements:

a. Extreme ruggedness. Must be unaffected by severe shock.

b. High resistance to extreme altitudes, temperatures, and humidities.

c. Durability. Must retain their mechanical and electrical properties without maintenance indefinitely.

d. Ability to resist corrosion.

The electrical requirements are that they have the specified capacitance with minimum weight and space, that they be capable of withstanding the operating voltages without danger of flash-over, and that they neither produce nor be affected by external electric or magnetic fields. They must also preserve their electrical properties over as wide a frequency range as possible.

These requirements make it mandatory that the capacitors be hermetically sealed, if at all possible, unless they are part of a unit which is hermetically sealed as a whole. In general, the design should follow the requirements of the joint Army-Navy specification of date of issue at the time the design is formulated.

If hermetical sealing is impossible, some sealing compound must be used to protect the capacitors against the effects of high temperatures and humidities. Dow Corning 4 compound has been widely used for this purpose. Recent tests by the U. S. Navy have shown that, due to its "flow" properties and loss of insulation resistance with absorption of moisture, this compound cannot be relied upon as a means of sealing an open-type, oil-dielectric capacitor. Breakdown and failure occurred frequently in these capacitors because the Dow Corning 4 compound mixed with the oil dielectric resulting in deterioration of the capacitor. Dow Corning 4 compound is satisfactory when used with synthetic rubber, neoprene, or phenolic insulating materials.

Capacitors to be used in suppression networks should always be of the non-inductive type. For example, a rolled impregnated-paper capacitor should make use of the extended foil construction as shown in Figure 4-18.

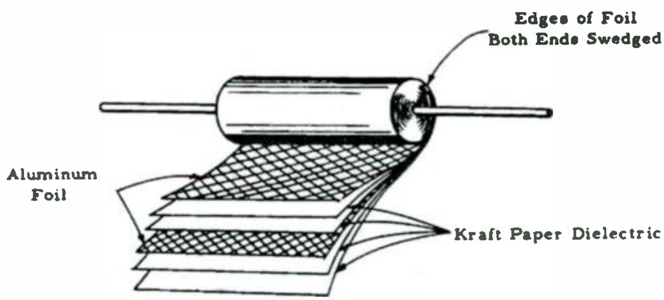


Figure 4-18. Construction of Non-Inductive Capacitor

Capacitors used as shunts to ground should be of the single-terminal type with the metal case grounded. An exception to this occurs for low-capacitance mica condensers completely embedded in a plastic mold. Since the mold is not conducting, it cannot be grounded. Provision should be made to mount the capacitor at the point where the lead leaves the mold.

Capacitors to be used as series elements should be constructed so that the two leads leave the case or mold at opposite sides and so that connections can be made with a minimum length of leads. The internal lead length must also be kept at a minimum in order to keep the series inductance as small as possible.

1.2.5 FEED-THROUGH CAPACITORS

The "feed-through" capacitor differs from the conventional, or "lead-type", capacitor in that all ground leads have been completely eliminated. As shown in Figure 4-19 it consists of a feed-through bus that passes through the center of the capacitor section which is rolled in the extended foil manner. Alternate foils on each side of the feed-through bus are soldered together; one set is soldered to the housing and the other set is soldered to the feed-through bus. This type of a configuration limits the inherent inductance from line to ground to an almost negligible value thus raising the resonant frequency of the capacitor and extending its useful frequency range of effective by-passing.

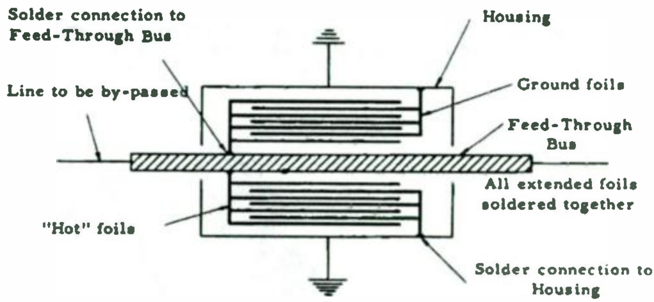


Figure 4-19. Schematic Diagram of a Feed-Through Capacitor

The feed-through capacitor is essentially a three-terminal element since the line from which the radio frequency currents are to be by-passed must be broken and the capacitor inserted between the separated ends. The feed-through capacitor functions most effectively when mounted through a shield wall so that contact to ground is afforded along a continuous symmetrical line around the circumference of the housing. This minimizes the inductance and resistance from the housing to ground. Furthermore, the shield isolates the input and output leads of the capacitor from each other. Because of this, feed-through capacitors find their greatest usefulness at the exit points of leads from shielded enclosures containing interference sources. In this application they are far superior to any other single element though their insertion loss is not as great as that of a well-designed filter.

One method of mounting, as shown in Figure 4-20 (1), makes use of a flange, an integral part of the capacitor wall, which is screwed to the shield. Better results are obtained, however, by using "press-in" or "cone" capacitors. The press-in capacitor is cylindrical in shape having an excess diameter of about a thousandth of an inch, while the cone capacitor, as the name implies, is conical. Both of these capacitors are forced into a bored opening under great pressure. Another method of mounting, consistent with good results, employs a condenser cylindrical in shape whose surface has been threaded as shown in Figure 4-20 (2). This capacitor is inserted into an opening in the shield and is held securely to both sides of the shield by a washer and nut arrangement. A similar type of mounting, shown in Figure 4-20 (3), employs a smooth cylindrical condenser with a threaded mounting stud which is inserted into an opening in the shield and held securely to the shield by a lock washer and nut.

(1) Mounting Flange (2) Threaded Surface (3) Threaded Stud

Figure 4-20. Methods of Mounting Feed-Through Capacitors

An analysis of the insertion loss - frequency curves of the feed-through, lead-type, and the ideal capacitor (i. e., one without any resistance or inductance), as shown in Figure 4-21, clearly shows the superiority of the feed-through capacitor. The dip below the ideal observed in the curve of the feed-through capacitor is typical of all capacitors of this type and usually occurs in the 50-600 megacycle range. The cause of the dip is not precisely known, but it is probably caused by a complex interaction of currents and fields in and surrounding the different layers of foil. Despite the dip, the feed-through capacitor is superior to the lead-type capacitor since no resonant effects are apparent and the insertion loss continues rising after the dip.

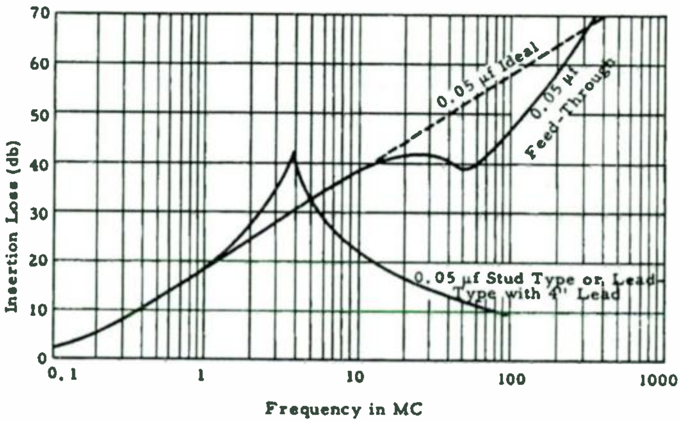


Figure 4-21. Curves of Insertion Loss as a Function of Frequency Comparing Feed-Through and Lead-Type Capacitors with Ideal Capacitors

Production samples of feed-through capacitors show wide variations in their insertion loss - frequency curves in the region approaching 1000 megacycles. This variation is probably caused by the radio-frequency resistance which at these frequencies is equal to or greater than the capacitive reactance of the capacitor thereby exerting the greater influence on the shape of the curve.

Another type of a feed-through capacitor, shown in Figure 4-22 consists essentially of a feed-through bus passing through the center of the capacitive section which consists of discs rather than rolled foil. Alternate discs on both sides of the feed-through bus are connected to the housing while the remaining discs are connected to the feed-through bus. This type of configuration is best adapted for use with ceramic dielectric discs whose surfaces have been coated with silver to form the conducting plates.

Figure 4-22. Schematic Diagram of a Stacked-Discs Type of Feed-Through Capacitor

1.2.6 FILTER INDUCTORS

Inductors used in radio interference suppression networks must meet the same mechanical requirements as capacitors, which are enumerated in Section 1.2.4. The electrical requirements are that they have the specified inductance with a minimum weight and space, that they be capable of passing the operating current without excessive heating, and that they neither produce nor be affected by external electric or magnetic fields. They must also preserve their electrical properties over as wide a frequency range as possible.

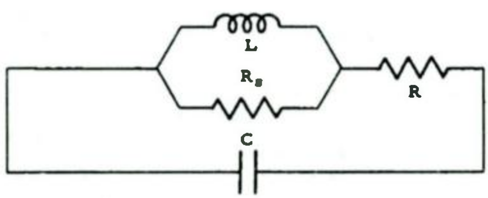


Figure 4-23. Equivalent Circuit of Inductance Coil

Figure 4-23 shows the equivalent circuit of an inductance coil. The quantity L is the inductance, R_s represents the shunting effect of the losses in the surrounding medium, R is the actual winding resistance taking into account the skin effect, and C is the distributed capacitance. The distributed capacitance may have losses, but these are usually negligible compared to those already mentioned. For direct and low frequency alternating currents, the core losses are very small and R approaches the direct current value of the winding. For use in an effective suppression network, C should be as small as possible so that the anti-resonant frequency, above which the inductor behaves like a capacitance and will lose its effectiveness, is as high as possible. The inductance L should be large for the frequencies of the interfering currents, but low for the desired power frequencies. The value of R is not important as far as the undesired frequencies are concerned, but a low value of R is desirable in order to keep the voltage drop across R as small as possible for the desired currents.

At the radio frequencies which are to be suppressed, practically all the losses associated with the inductance coil occur in R_s. Thus the "Q" of the coil is determined primarily by R_s. It is known from filter analysis that the effect of dissipative elements is negligible in all cases except in the vicinity of the critical frequencies (the cut-off frequencies and the frequencies of infinite attenuation). Since radio interference suppressing filters or networks usually operate well beyond cut-off, relatively low values of R_s may be tolerated. Values of "Q" as low as three reduce the effective impedance of a coil by only 30 percent as compared to its value for the ideal "Q" of infinity.

It is very important to reduce the capacitance C as much as possible, and this may be done by connecting several coils in series so that the various capacitances of the individual coils are in series. Since this also lessens the inductance of the coil, because it eliminates some of the mutual inductance between the turns of different coils, a compromise must be found.

Multiple-layer coils should be avoided since their capacitance is appreciably greater than that of single-layer coils. The size of wire to be used depends on the current to be carried. The wire size should be chosen so that 100 percent overload can safely be carried except in speial cases where circumstances may require a larger factor of safety. In determining the current, the radio interference currents may be neglected in comparison with the power currents. The correct wire size may be found in the table of Figure 4-12.

For inductances below about 1000 microhenries, usually no magnetic material need be used unless special space requirements exist. The inductance of a coil is a very complicated function of its physical dimensions and the number and arrangement of the turns. However, the following equation may be relied on to give correct results for a single-layer air-core coil within about 2 percent provided that the ratio of diameter to length, a/b, is neither very large nor very small as compared to unity.

$$L = \frac{0.5\, a^2 b^2 k^2}{9a + 20b} \tag{4-22}$$

The inductance, L, is in microhenries, the average coil diameter a is in inches, (see Figure 4-24), the length of winding b is in inches (see Figure 4-24), and k is the number of turns per inch. Figure 4-25 is a nomograph based on Equation (4-22) for the rapid computation of either the inductance of a coil of given dimensions, or the dimensions of a coil for a given inductance. The use of this nomograph is as follows:

To find the inductance of a coil whose dimensions and number of turns are known, find the reference point on the "turning" scale by drawing a straight line through the approximate points on the "b/a" and the "a" scales. Then draw a straight line through this reference point and the approximate point on the "k" scale and extend this line until it intersects the "L" scale.

The following example illustrates the use of the nomograph in solving the above type of problem. Refer to Figure 4-25. Find the inductance of a coil close wound with 13 turns of No. 22 enamel wire (37 turns per inch), if its diameter and length are 3.5 and 0.35 inches, respectively. Connect b/a = 0.1 and a = 3.5 with a straight line and rotate this line about the reference point on the turning scale to k = 37, and read L = 30 microhenries.

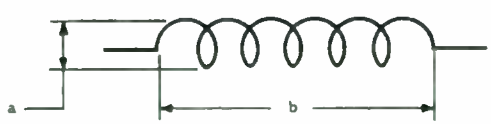


Figure 4-24. Dimensions of Single-Layer Air-Core Coil

The length and the number of turns of a coil of known diameter, inductance, and current carrying capacity can be found by the use of the nomograph. by finding first the number of turns per inch from the table of Figure 4-12. Then, a straight line is drawn between the appropriate points on the "k" and "L" scales to locate the reference point on the "turning" scale. This point, in turn, is connected to the appropriate point on the "A" scale by a straight line which is extended until it intersects the "b/a" scale. Since the value of "A" is known, the value of "b" can be readily calculated and used to obtain the total number of turns, N, by means of the relationship, N = kb.

The following examples illustrate the use of the nomograph for the above conditions. Refer to Figure 4-25. Find the length and the number of turns of a 30-microhenry coil, designed to carry a maximum current of 1.5 amperes, and whose diameter is 3.5 inches. Figure 4-12 shows that 37 turns per inch of No. 22 enamel wire are required. Connect k = 37 and L = 30 with a straight line and rotate this line about the reference point on the turning scale to a = 3.5 and read b/a = 0.1. Since a = 3.5 inches, b = 0.35 inches. Therefore, the number of turns required is 0.35 x 37 = 13 turns.

If space is ample, it is usually desirable to make "k" smaller than its maximum value read from the table of Figure 4-12 because a decrease of "k" will reduce the distributed capacitance of the coil. This reduction of the value of "k" is obtained by space winding rather than close winding the coil.. It should be noted, however, that an error in the value of the inductance is introduced because the value of "k" used in the nomograph of Figure 4-25 is based on a close-wound coil.

The distributed capacity C_o of a single-layer coil, shown in cross-section in Figure 4-26, can be readily found by the use of the nomograph of Figure 4-26, if the diameter of the coil "a", the diameter of the wire used "d", and the distance between the centers of adjacent turns "s" are known. The procedure for the use of the nomograph is as follows:

Draw a straight line connecting the appropriate points on the "s/d" and "a" scales. The point of intersection of this line and the "C_o" scale gives the value of the distributed capacity. If the diameter of the coil is "m" times larger or smaller than the values on the "a" scale, the value of the distributed capacity is also "m" times larger or smaller than the value read on the "C_o" scale. In this case, "m" is any constant factor.

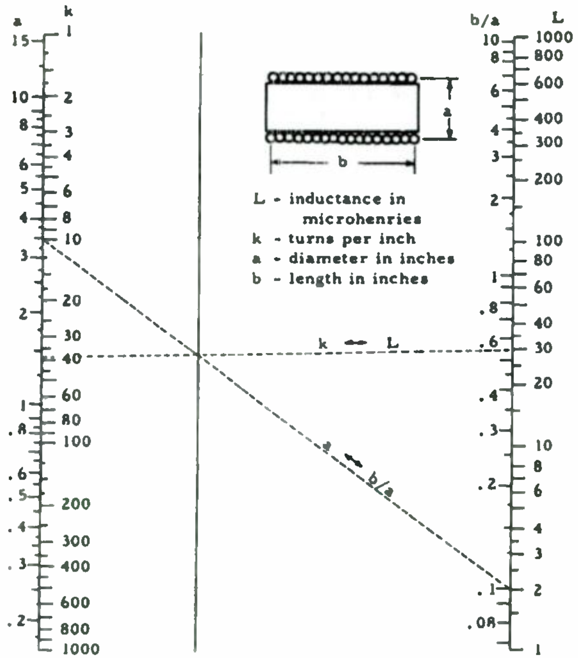


Figure 4-25. Nomograph for Single-Layer Coil Design

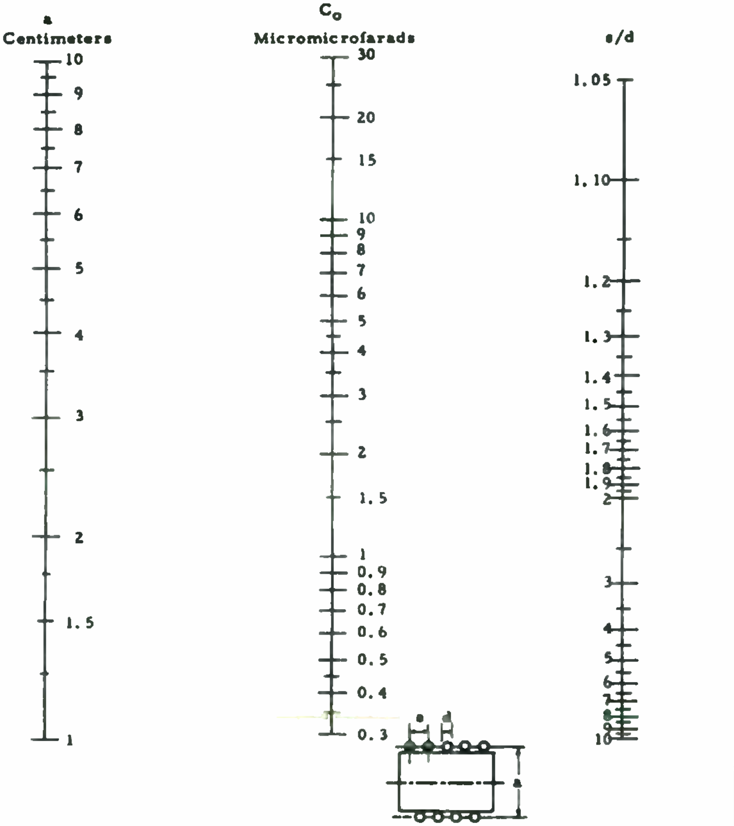


Figure 4-26. Nomograph for Determining Distributed Capacitance of Single-Layer Coils

If more inductance is needed in a given space than can be obtained in an air-core coil, a core of magnetic material must be used. Suitable coils using magnetic cores may take many forms. Several important arrangements are shown in Figure 4-27. Arrangement (1) in this figure shows a simple straight solenoid whose greatest merit is ease of manufacture. Arrangement (2) shows a toroid, which provides a satisfactory filter inductor as the external flux is almost nil. The distributed capacitance may be satisfactorily low; and, for a given space requirement, it provides a good value of inductance. Finally, arrangement (3) is an improved design which makes excellent use of available space, allows a very close-fitting shield to be used, and has a relatively low value of distributed capacitance.

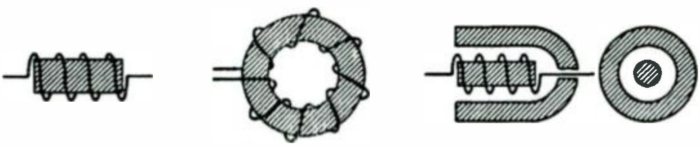


Figure 4-27. Designs of Magnetic Core Inductors

If the power current is direct current, the saturation effect of the load currents must be carefully checked. In arrangement (1), the large airgap, plus the intrinsic airgaps due to the interstices between iron particles in the case of powdered cores, generally prevents saturation. A toroidal core of laminated strips must have an airgap to prevent saturation. A toroidal core of powdered iron may be free from saturation effects for moderate load currents, but may still require an airgap for very large load currents. In arrangement (3), special care must be taken, since saturation may easily occur. This is particularly important should end-caps of magnetic material be added.

A magnetic slug in a straight solenoid will increase its inductance by a factor of about two to four. The inductance of a toroid of rectangular cross-section with or without intrinsic airgap, but without external airgap, may be found from the equation,

$$L = 0.00508\, N^2\, b\, \mu_d \ln (r_2/r_1) \qquad (4\text{-}23)$$

where L is the inductance in microhenries, b is the core width in inches, μ_d is the average incremental permeability, r_1 and r_2 are the inside and outside radii, respectively, and N is the total number of turns. See Figure 4-28. The incremental permeability must be used because the quantity of interests is the incremental inductance for small alternating currents superimposed on a large direct current. Curves of μ_d as a function of the magnetization H in ampere-turns per inch are plotted in Figure 4-29. The quantity H may be computed from the equation,

$$H = \frac{N\,I}{2\pi r_{av}} = \frac{N\,I}{\pi(r_1 + r_2)} \tag{4-24}$$

where I is the direct current in amperes.

For toroids with round cross-section, the inductance is

$$L = 0.0319\, N^2 \mu_d \left(r_m - \sqrt{r_m^{\,2} - \frac{b^2}{4}} \right) \tag{4-25}$$

where r_m is the mean radius of the toroid and b is the radius of the cross-section, both in inches, as shown in Figure 4-30.

Another design, which is particularly useful for frequencies above about 50 mc, is shown in Figure 4-31. It consists of a straight conductor surrounded by an annular core. It is essentially equivalent to a one-turn toroid, and Equations (4-23) and (4-24) may be used with N taken as unity. The package-ability, low voltage drop for direct current, and extremely low capacitance combine to make this an ideal design when a very large inductance is not required. The optimum inductance is obtained by using

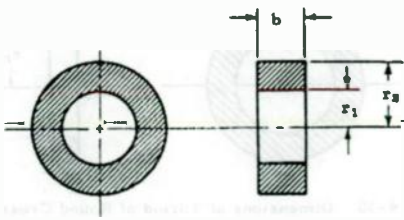


Figure 4-28. Dimensions of Toroid of Rectangular Cross-Section

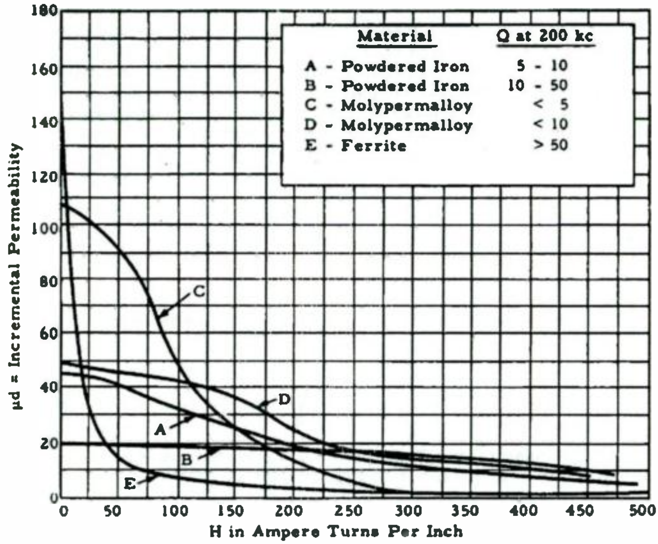


Figure 4-29. Incremental Permeability as a Function of Magnetic Field Intensity for Different Materials

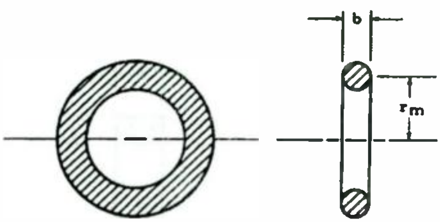


Figure 4-30. Dimensions of Toroid of Round Cross-Section

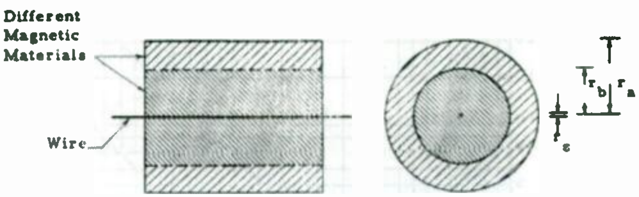


Figure 4-31. High Frequency Inductor

several layers of different material in the annular core. The design of such an inductor may be carried out with the aid of the curves of Figure 4-32.

It is assumed that the outside diameter of the core is known from a consideration of the available space. The direct current determines the wire size and thus the inside diameter of the core. The range of values of H may then be determined from the curves of Figure 4-32, which give the magnetic field intensity at any arbitrary distance r from the center. The materials having the largest μ_d may be chosen from the curves of Figure 4-29. Note that the intersection of two curves on this graph marks the value of H for which one material becomes better than another. Thus, Figure 4-32 may be used to determine the radius or radii at which one material should be replaced by another. Finally, the inductance per unit length is found by adding the inductance values for the individual cores:

$$L' = 0.00508 \left(\mu_{d_1} \ln\frac{r_a}{r_b} + \mu_{d_2} \ln\frac{r_b}{r_c} + \ldots \right) \qquad (4\text{-}26)$$

where L' is in microhenries per inch and r_a, r_b, and r_c are explained in Figure 4-31; ln stands for $\log_e$.

As an example, let it be required to determine the optimum inductance per unit length of a conductor to carry 10 amperes direct current, with a maximum diameter of 0.6 inches available for the core. From the table in Figure 4-12, a wire size No. 14 is chosen. Allowing a hole of 0.08 inches in diameter, the radius of the core varies from 0.04 to 0.3

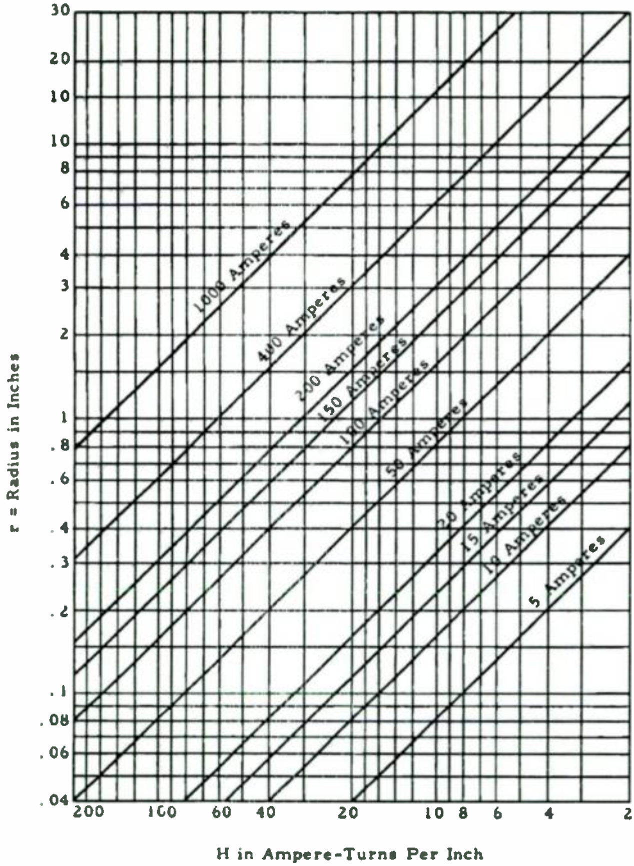


Figure 4-32. Relation Between Radius from Center of Cylinder and Magnetic Field Intensity for Different Currents

inches. From Figure 4-32, this corresponds to a variation in the field intensity H from 40 to 5.5 ampere-turns per inch. Figure 4-29 shows that, for H having values below 12 ampere-turns per inch, material E (ferrite) has the highest incremental permeability, while, for H between 12 and 40, material C (molypermalloy) should be chosen. The value H = 12, from Figure 4-32, corresponds to a radius of 0.14 inches for a current of ten amperes. Hence, the core should consist of two concentric annular rings. The inner ring, with an inner diameter of 0.08 inches and an outer diameter of 0.28 inches, is made of molypermalloy, and the outer ring, with an inner diameter of 0.28 inches and an outer diameter of 0.6 inches, is made of ferrite. The resultant inductance may be computed as follows:

$$L' = 0.00508 \left(\mu_{d_E} \ln \frac{0.3}{0.14} + \mu_{d_C} \ln \frac{0.14}{0.04} \right) \qquad (4\text{-}27)$$

The quantity μ_{d_E} varies from 135 to 105 in the region considered. An average of 120 may be used. The quantity μ_{d_C} varies from 105 to 97. An average of 101 may be used. With these values the inductance, L', is 1.11 microhenries per inch.

In choosing a magnetic core for an inductor, three factors must be considered: permeability, saturation, and losses. When inductors are used in networks that must pass only direct or very low frequency currents, the losses need not usually be considered. The losses always increase with frequency, and when all high frequencies are to be suppressed, losses are not objectionable. They are just additional aids in attenuating the interfering currents. But in harmonic-suppression filters or other networks that must pass radio frequencies, low-loss materials must be chosen. Figure 4-29 shows that the ferrites have the highest initial permeability and the lowest losses at 200 kc. However, they reach saturation very quickly, and their losses increase very rapidly at frequencies above 5 mc. Cores of powdered iron or special alloys such as molypermalloy have very low losses even at very high frequencies. They are not saturated easily, but their initial permeabilities are not very high. They are still the best compromise available when fairly high inductances and low losses at high frequencies are required.

It is often desirable to embed coreless inductance coils in plastic material in order to protect them from shock. When doing this, the heat-conducting properties of the material become important. Powdered quartz has been found particularly useful since its large heat conductivity allows an increase of the current rating, thus permitting larger values of inductance without increase in volume.

2. SPECIAL CIRCUITS

In Chapter 1; Section 1.2.3.3 the basic types of circuits used for interference reduction in receivers were enumerated. The specific circuits, with examples reflecting extensive practical application, are presented here.

Only those circuits are treated here which attempt to reduce the interference after it has entered the antenna or any other point along the normal route of the desired signal. In other words the circuits to be discussed act after the interfering and desired signals have mixed, and therefore they must separate the two on the basis of intrinsic difference between them. This difference may lie in the wave form, in the frequency distribution, or in some definite phase relations that apply to one, but not the other.

2.1 LIMITERS

The action of an amplitude limiter is based on the fact that, in amplitude modulation, the amplitude of the desired signal varies from zero to, at most, twice the carrier amplitude, reaching that amplitude only during the rare modulation peaks. If an interfering signal contains pulses of short duration whose amplitude rises above that value, its effect will be greatly reduced if all amplitudes are limited to that same value. This limiting action does not affect the desired signal. Noise limiters of this type are quite effective and are now a standard part in many receivers.

Limiting action in these circuits is usually provided either by the switching action of a diode (in so-called "diode limiters") or by saturation of some tube. A diode limiter may be either the series or shunt type. In the series type, the diode is in series with the normal plate current flow so that it is conducting as long as the signal does not exceed twice the carrier level, but becomes non-conducting for a short-time interval when the signal exceeds this value. In the shunt type, the diode is in parallel with the normal plate current flow so that it is normally non-conducting, but becomes conducting for excessive signal amplitudes. In either case, the signal becomes greatly attenuated during the time that the limiting action of the diode takes place. Clearly, this time must be small enough so that the desired signal remains unaffected. Making use of the saturation effect in a vacuum tube has the advantage that no special tube is needed to provide noise limiting action, but that an existing tube may be utilized for this purpose. Practically, however, it has been found that this method is less effective and introduces greater distortion in the desired signal than a separate limiter stage.

Because of the pulse-lengthening action of the various stages of the receiver, the most desirable location of the noise limiter circuit is before the radio frequency amplifier, or at any rate, as close to it as possible. Diode limiters, however, need both direct current pulses and considerable voltage ampiltudes for their operation, which normally are not available in the early stages of the receiver. The generally accepted location of diode limiters is therefore the second detector stage, where the necessary voltages and currents are available.

In many cases limiting action can be made even more effective by allowing it to take place below the 100 percent modulation level. This will, of course, introduce distortion into the desired signal during modulation peaks; but, remembering that, in practice, an average modulation level of from 30 to 40 percent is rarely exceeded, the resulting distortion is quite small as compared to the improvement obtained due to the suppression of interference peaks.

When designing limiters as a means of interference suppression, it must be borne in mind that their effectiveness is highest when the frequency selectivity in the circuits after the limiter is greater than the selectivity preceding the limiting.

The limiter shown in Figure 4-33 operates satisfactorily on both modulated and unmodulated reception. It is a simple, convenient type requiring only a fixed capacitor, two fixed resistors, and an independent diode besides the normal components of a diode second detector.

When the switch is in the "OFF" position, C_3 connects to junction point B of the detector diode load resistors R_1 and R_2 and the limiter diode has no effect on circuit performance. When the switch is "ON", C_3 is connected directly to the cathode of D_2, putting the diode switch in the circuit.

Assuming a potential of 10 volts across R_1 and R_2 by a constant carrier, the cathode of the limiter would be 10 volts negative with respect to ground if the diode were not conducting. The plate, connected to point B, is 5 volts negative with respect to ground. Hence the plate is 5 volts positive with respect to the cathode, and the limiter diode conducts, having

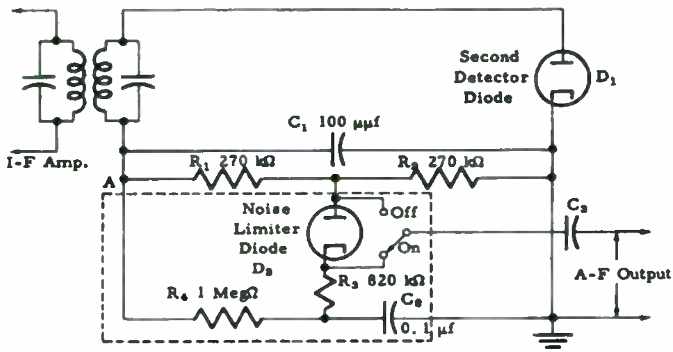


Figure 4-33. Series Limiter

fairly low resistance compared to other circuit resistances. The output capacitor C_3 is then connected to point B through the limiter diode, providing audio frequency output. This output is reduced to 45 percent of what it would be without the limiter, but generally the reduction is of little significance.

The difference in time constants between plate and cathode circuits allows the diode resistance to become very high when a large interference voltage appears, effectively preventing conduction through the interference limiting diode D_2 and cutting off C_3 from point B. The amplifier will have no appreciable input for the duration of the interfering signal. By the time the cathode of the limiter diode goes negative with respect to its plate, the interfering signal will have decayed, restoring audio frequency input to the amplifier.

Distortion in this limiter is noticeable on an oscilloscope only above 40 percent modulation. Speech and coded transmissions maintaining an average modulation level of 30 to 40 percent are commonly encountered.

The modified shunt type of noise-peak limiter circuit, Figure 4-34, is similar to the series type except that the plate of the limiter diode and the low end of the cathode resistor are interchanged. When an interference peak makes the diode D_2 conduct, it acts to reduce the output voltage. Grounding the low end of the intermediate frequency secondary increases the stability of the system.

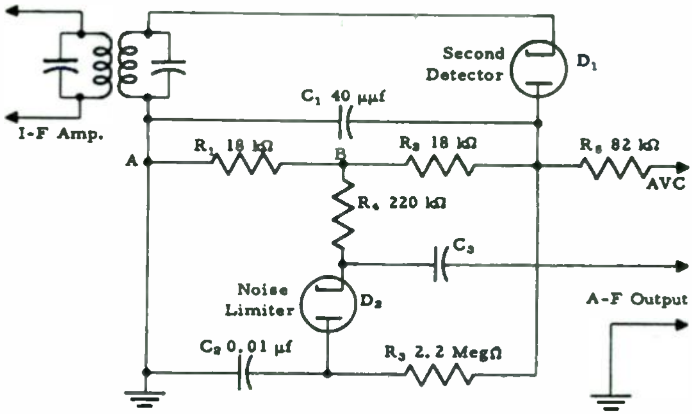


Figure 4-34. Shunt Limiter

In this circuit normally the cathode is positive with respect to the plate, and D_2 is not conducting. The time constant in the plate circuit is more than 10,000 times longer than that of the cathode circuit, so that any interference pulse in excess of normal bias drives the cathode negative and the diode conducts, shunting the input of the following audio frequency stage. Shunting action is more complete with the addition of R_4 which acts as part of a voltage divider when D_2 is conducting and helps attenuate detector-load voltage peaks.

The limiting action ceases when the interference pulse decays or C_2 charges, since then point C becomes positive with respect to point D. Audio distortion begins at about 100 percent modulation for values shown.

If the ratio of R_2 to R_1 is 0.4, distortion begins at approximately 40 per-cent modulation. This limiter is not as good as the simple series type at lower carrier frequencies.

The system given in Figure 4-35 for interference suppression silences the receiver momentarily when an interfering pulse stronger than the desired signal is received. Here the blanking and limiting principles are combined.

The interference amplifier (6J7) has its grid connected in parallel with the grid of the final intermediate frequency amplifier, and delivers its output to an auxiliary rectifier. The direct-current bias thus obtained is applied to the third grid of the 6L7 final intermediate-frequency amplifier. When properly adjusted any interference voltage whose peak amplitude exceeds the signal being received will develop enough bias to make the final intermediate frequency amplifier tube inoperative, thus silencing the receiver for the duration of the pulse. This is accompanied by some distortion, but reception is much improved over that obtained without the interference suppressor.

The same effective principle is applied in the counter-modulation type of interference-reducing circuit shown in Figure 4-36.

In this circuit, intermediate frequency signals and interference voltages are injected into a push-pull controlled amplifier stage, paralleled by a push-pull interference amplifier whose plate circuit is tuned lower than the intermediate frequency value. This allows more interference and less signal voltage at the interference-rectifier input than at the second detector input. Broad tuning of the intermediate frequency stages permits sufficient detuning of the interference amplifier to virtually eliminate signal voltage in this circuit. Any signal voltage still present is prevented from affecting the circuit operation by the automatic threshold control tube. The threshold control tube regulates a variable-delay bias for the interference rectifier, allowing rectification only above the level of the signal component in the voltage being rectified. The automatic volume control voltage fed to the grids of the threshold tube regulates the delay level for a change of signal strength, and the interference component is taken from the interference rectifier load resistor. Then it is fed through a blocking condenser to the 6L7 grids. The level of zero axis for the countermodulating voltage is set by a direct current bias supplied to these grids. Once the delay and bias adjustments are made, they need not be changed.

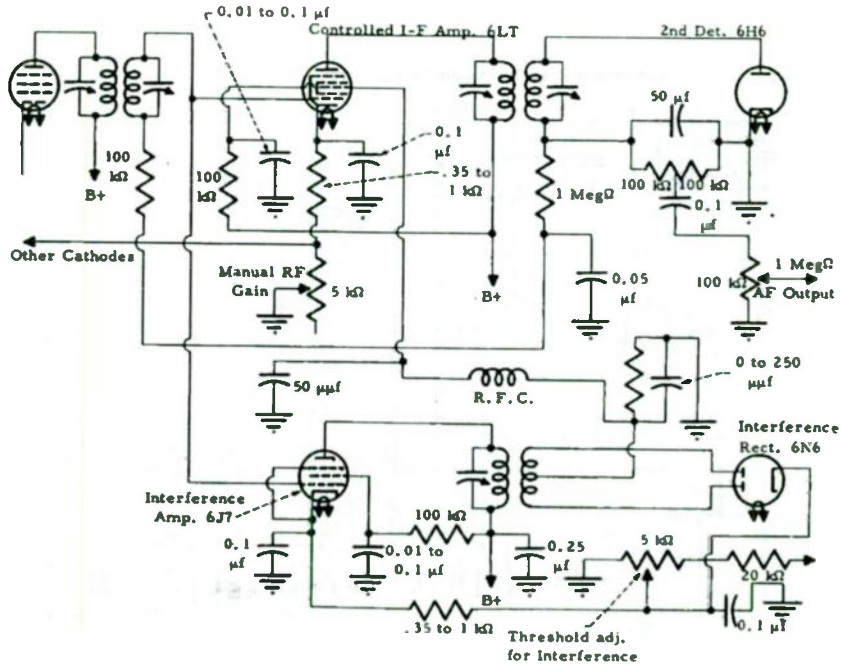


Figure 4-35. Combined Limiter-Blanking Circuit

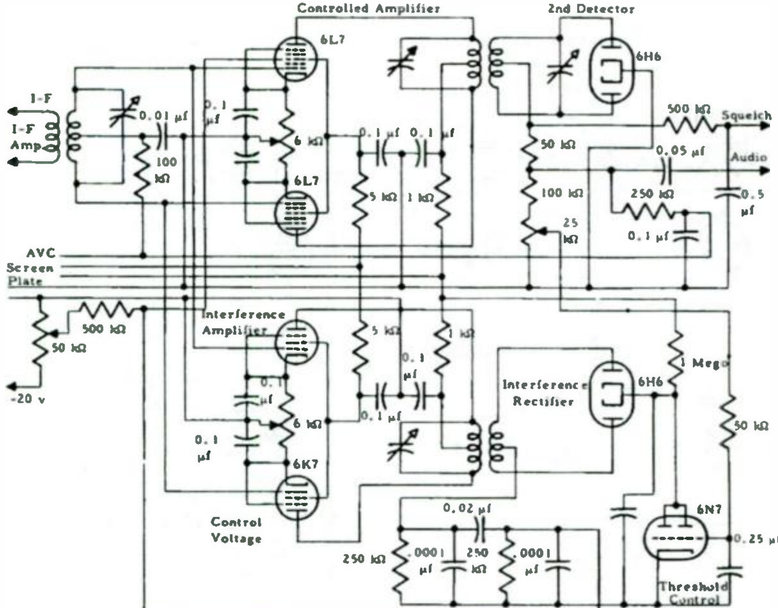


Figure 4-36. Limiter Employing Counter-Modulation

Though not especially simple, the circuit is effective and fully automatic. It reduces the interference voltage before recitification and does not increase the automatic volume control voltages. Thus receiver gain, interference rectifier delay, and interference amplifier gain are all controlled by the strength of the desired signal.

The limiter circuit shown in Figure 4-37 employs degeneration as a means of obtaining limiting action. The input stage of the amplifier uses one triode section of the 6F8G tube, the other section being used for an auxiliary amplifier for the 6H6 control rectifier. The input stage is transformer coupled to a push-pull output stage using another 6F8G tube. Degeneration is used on this stage, and the feedback factor is determined by R_1 R_2, and the plate resistances of two 6L7 tubes. The feedback factor can be controlled by varying the plate resistances of the 6L7 tubes since they are effectively in parallel with the resistors R_1. An increasing signal causes the grid bias on the 6L7 tubes to increase, which increases the plate resistance and the feedback factor and results in decreased gain, thus producing compression. Negative feedback has the added advantage of holding distortion to a low value.

By putting an initial positive bias on the 6H6 full wave rectifier, the control of the auxiliary gain will delay compression till any desired output is reached, within the limitations of the amplifier. An amplifier plate resistance at 7700 ohms, a transformer ratio from primary to secondary of 2 to 1, and a resistance of 1000 ohms for the 6H6, in conjunction with 0.5 microfarads for C_2, give an acting time of 1.5 milliseconds. Releasing time, with R_3 equal to 2 megohms, is 1 second.

A push-pull compressor stage with balanced feedback network is necessary to eliminate transient distortion due to compressor action, and is also desirable from the standpoint of low inherent distortion. Noise and hum level is 75 db below 6 milliwatts output, largely because of a regulated power supply.

In broadcast service, this amplifier is able to handle all ordinary peaks. There are no thumps when compression takes hold, and the general operation is quite smooth. Feedback does not help reduce interference at low power levels, such as thermal agitation, induced hum voltages, and microphonics.

Where automatic volume control is employed in a receiver of considerable sensitivity, a disagreeable amount of interference will be heard in the output when no carrier is present. An arrangement for

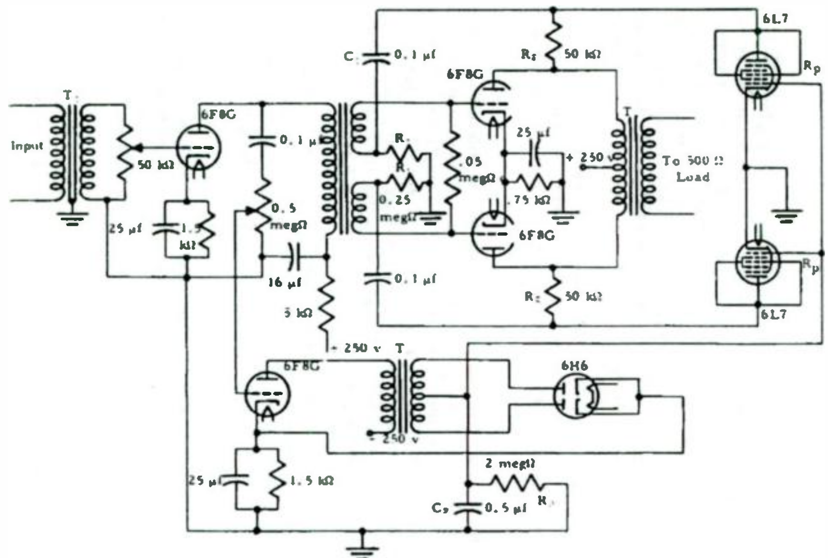


Figure 4-37. Limiter Employing Feedback

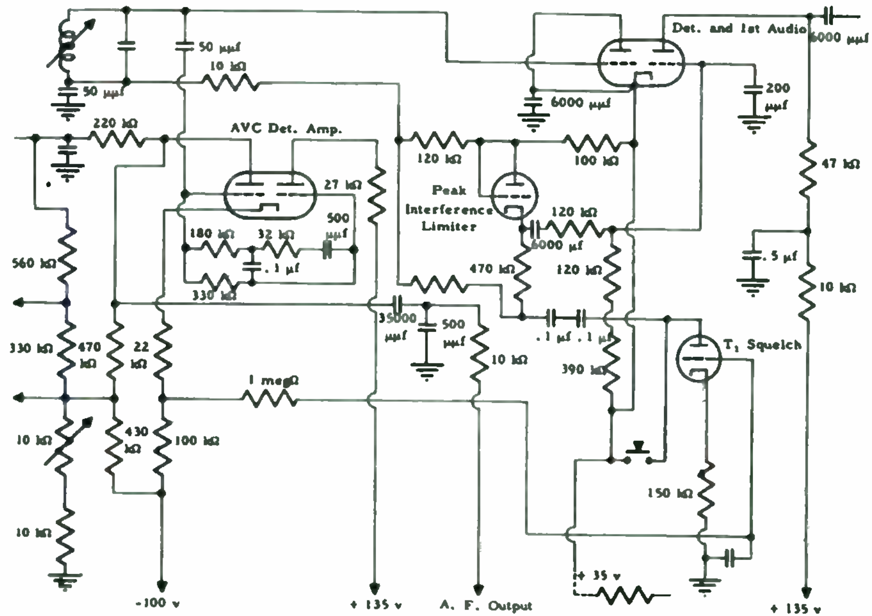


Figure 4-38. Squelch Circuit

blanking the receiver during the tuning process, often referred to as a squelch system, will suppress this form of interference and is indicated in a typical schematic form in Figure 4-38.

The first audio tube is biased beyond cut-off by the action of Tube T_1 unless the grid bias of Tube T_1 approaches or exceeds cut-off. By using the automatic volume control system to bias T_1, the receiver can be made inoperative until a carrier of pre-determined amplitude is present.

A modification of this arrangement is to operate Tube T_1 from a separate branch of the intermediate-frequency amplifier that delivers its output to a second diode. By making this auxiliary intermediate-frequency branch very selective, the signal-to-interference ratio will be much higher for the interference-suppressor diode than for the detector diode. The threshold level of the system may be set such that the signal is so low with respect to the interference as to be barely usable. The receiver will then deliver no output until tuned exactly to the desired carrier frequency.

2.2 WAVE TRAPS

A wave trap is a circuit designed to attenuate greatly one frequency, or a very narrow band of frequencies, while passing without appreciable attenuation all other frequencies. Wave traps are usually inserted into the antenna circuit of a receiver to eliminate one particular frequency which happens to be particularly disturbing. Since the input circuit of a receiver itself acts like a band-pass filter, frequencies far removed from the acceptance band of the receiver will usually be sufficiently attenuated by the receiver itself unless they are extremely strong. Wave traps are, therefore, most frequently employed when the frequency to be suppressed lies just above or just below the frequencies passed by the receiver. In the case of very strong interfering signals, however, the attenuation produced by the receiver may be insufficient, and a wave trap may have to be used even though the interfering frequency is several octaves above or below the frequencies passed by the receiver.

The simplest type of wave trap is an anti-resonant circuit, i.e., a parallel combination of inductance and capacitance as shown in Figure 4-39(A). This circuit has a very high impedance at the anti-resonant frequency and, therefore, attenuates the currents at that frequency. How rapidly the impedance decreases on either side of the anti-resonant frequency depends on the ratio of L/C and on the "Q" of the circuit. This is shown by writing the expression for the impedance of the anti-resonant circuit:

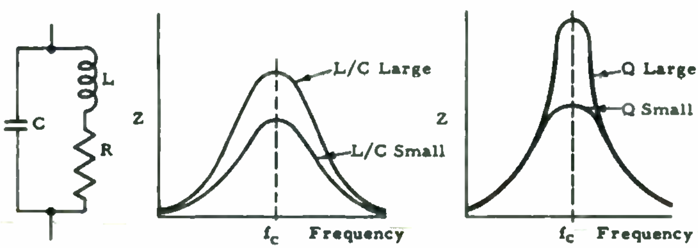
$$|Z| = \left| \frac{R + j\omega L}{j\omega C \left[R + j\left(\omega L - \frac{1}{\omega C} \right) \right]} \right| \qquad (4\text{-}28)$$

$$= \frac{f_c^2}{f} \, 2\pi L \sqrt{\frac{Q^2 + 1}{Q^2 \left[\left(\frac{f_c}{f} \right)^2 - 1 \right]^2 + 1}} \qquad (4\text{-}29)$$

where f is the frequency and $f_c = 1/(2\pi\sqrt{LC})$ is the approximate anti-resonant frequency. This expression shows that the impedance is decreased at all frequencies if L is decreased with a constant Q and f_c. Since f_c is determined by the product LC, decreasing L with constant f_c means increasing C.

On the other hand, increasing Q with a constant L and f_c has no effect except when f/f_c is close to unity. For it is seen that, if f/f_c is either much larger or much smaller than unity, the impedance is practically independent of Q, provided that Q is larger than about 10 - a condition usually satisfied in practice. But when f/f_c is very close to unity, the impedance becomes practically proportional to Q.

The effects on Z of variations of L/C and Q are illustrated in Figures 4-39(B) and 4-39(C). It is seen that, for a wave trap of the type



(A) Circuit

(B) Effect of the Ratio L/C on the Impedance

(C) Effect of Q on the Impedance

Figure 4-39. Characteristics of Simple Wave Trap under consideration to have the minimum effect on the transmission of the desired band and to provide the maximum attenuation of the interfering signal, the ratio L/C should be as small as possible and Q should be as large as possible.

Simple wave traps can consist of either a parallel-tuned circuit connected in series with the receiver antenna (Figure 4-40) or a series-tuned circuit in parallel (Figure 4-41). As has been noted, proper attenuation of the interference and low insertion loss depend upon a low L/C ratio in a parallel circuit (Figure 4-40); similarly a high L/C ratio is required by the series circuit (Figure 4-41).

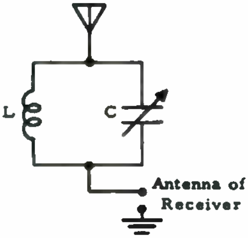


Figure 4-40. Parallel-Resonant Wave Trap

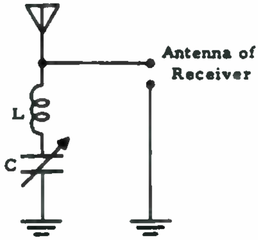


Figure 4-41. Series-Resonant Wave Trap

Suitable values for the elements of the circuit of Figure 4-40, for selected frequencies, are given in Figure 4-42. Wave traps consisting of lumped elements of this type are rarely used for frequencies above 30 to 40 megacycles.

Frequency in Megacycles	Capacitance in $\mu\mu f$	Inductance in μh	Coil Design Data
3.5	140	16	32 turns No. 22, 1" dia., 1" long
7	100	6	19 turns No. 22, 1" dia., 1" long
14	50	3.5	14 turns No. 18, 1" dia., 1" long
21	35	2.2	12 turns No. 18, 1" dia., 1" long
28	25	1.5	9 turns No. 18, 1" dia., 1" long

Figure 4-42. Representative Values for Basic Wave Trap

The following types of wave traps have been successfully applied in reducing interference from IFF and similar systems operating at frequencies about and above 50 megacycles.

a. A radio frequency choke coil, designed to resonate with its distributed capacity, is installed in the antenna circuit of the communication receiver. Such a coil consists of three series windings of fifteen to thirty turns each, depending on the interfering frequency, wound on a bakelite form roughly 3/8 inches in diameter and 4 inches long.

b. A quarter-wave length open-circuited stub is connected between the antenna post and ground of the communication receiver. This stub consists of No. 18 solid copper, insulated, push-back type wires, about one quarter of a wavelength long and twisted about 9/10 of their length, the remainder serving as leads, as shown in Figure 4-43. The stub should be cut back experimentally until maximum interference reduction is obtained. This point is critical, and the wires should not be cut back more than 1/8 inch at a time in order to make sure that this point is not missed. A short length of coaxial cable may be used instead of the twisted wires. Its length must be adjusted by trial and error in the same way as described above for the twisted wires.

c. A three-section parallel-resonant series wave trap constructed from coaxial cable, similar to the radio frequency choke coil, may be inserted between the antenna relay and the receiver antenna post. Each section is approximately a quarter wave length long and short-circuited, as shown in Figure 4-44. The exact dimensions are determined by trial and error.

The choke coil described in "a" is suitable in the range from about 40 to 100 magacycles. Types "b" and "c" are suitable above 100 megacycles.

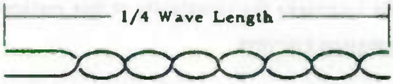


Figure 4-43. Twisted Quarter-Wave Series-Resonant Parallel Wave Trap

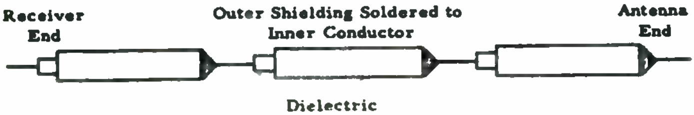


Figure 4-44. Three Section Parallel-Resonant Series Wave Trap

2.3 BLANKING CIRCUITS

In a blanking circuit, the entire receiver is rendered inoperative for the duration of an interfering pulse. Such action can take place before the signal enters the first stage of the receiver, and therefore is not subject to the restrictions imposed by the pulse-lengthening effect of the receiver circuits. Blanking action may be triggered by the interfering pulse itself, in which case a delay line or circuit must be provided for the desired signal so that the receiver is blocked before the undesired pulse can enter it. It may also be triggered by an independent signal in those cases where the arrival of an interfering pulse is known in advance, as for example when the interfering pulse comes from a radar transmitter in the same system. In this case, the triggering pulse may be provided by a signal taken directly from that radar transmitter. Blanking action is usually provided by a simple amplifier stage that can be biased to cut-off by the suitably amplified trigger pulse.

Blanking circuits are often used when no other methods of protecting the receiver are available, but they lack the simplicity of limiters and wave traps. They are complete units in themselves including amplifiers, trigger pulse amplifiers, and delay circuits. In addition, their action itself may be a source of interference inasmuch as cutting off the carrier periodically is a form of modulation, which will appear in the audio output as noise. This can be overcome by supplying the carrier locally when the blanking circuit punches "holes" into the carrier arriving from the outside, but this increases the complexity of this method still further.

2.4 PHASE-CANCELLING CIRCUITS

A phase-cancelling circuit is a circuit that allows the transmission of the interfering signal by two different paths. When the two portions of the interfering signal recombine, they are exactly 180° out of phase and cancel. Because the phase shift of any network is usually a function of frequency, phase-cancelling networks are not designed for signals containing more than one or at most a few frequencies. They can only be used when the path of entry as well as the nature of the interfering signal is fully known. Therefore, their use is practically restricted to interfering signals from radar transmitters or modulators in the same system, or from similar sources of definite frequency.

The block diagram of a typical phase-cancelling circuit is given in Figure 4-45.

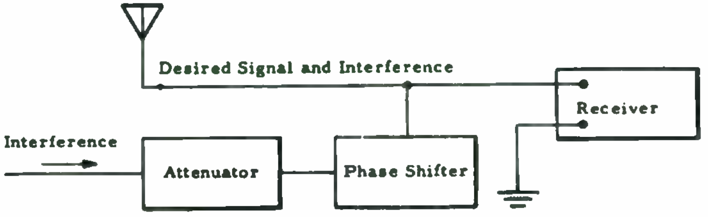


Figure 4-45. Block Diagram of Phase-Cancelling Network

2.5 AUDIO FILTERS

In those cases where the interfering signal contains only a small number of fixed audio frequencies, it is possible to design a filter that eliminates these frequencies and to place it in one of the audio stages of the receiver. For example, when a local radar transmitter causes severe interference at the frequency of its pulse repetition rate, the audio frequencies present in the output of the receiver will be only that repetition frequency and its harmonics. An audio filter can be designed that attenuates the very narrow bands of frequencies about each of these frequencies.

Since the ear is normally quite discriminating against audio signals of different pitch, the desired signals will usually remain intelligible in the presence of even fairly loud interference of definite pitch. An audio filter, in eliminating this interference, will decrease the annoyance caused by the interference, but will not usually appreciably add to the intelligibility of the desired signal. Therefore, because of their inherent complexity, these filters are rarely used in existing receivers.

1. GENERAL ASPECTS OF POWER SYSTEM INTERFERENCE

Without the application of proper design and suppression practices, interference from power systems becomes a serious problem. Modern complex communication-electronic systems incorporate a large number of components and devices which use power in common; each of these is at the same time susceptible to interference and also a potential generator through power input connections.

A preventive and corrective approach then to interference-free functioning of communication-electronic systems is most effectively done at the component unit level. Every electric motional source must be designed (or later corrected) so that its inherent electrical transients are confined to the unit itself and prevented from entering other units. The following paragraphs discuss interference problems and suppression techniques in power systems.

2. ROTATING MACHINERY

2.1 BRUSHES

In all types of electrical machines (except certain types of induction motors) electrical contact must be made between two conducting surfaces that are in relative motion. Such contact is normally made by brushes sliding on a metallic surface, and this is always accompanied by the generation of interference. In dc machines having good commutation, most of the interference is directly attributable to the sliding brush contact. This so-called brush interference may be reduced by careful consideration of the following factors in the design or choice of brushes and of the metal surface in contact with the brush face.

2.1.1 BRUSH PRESSURE

Interference generated decreases with increasing brush pressure at all frequencies as shown in Figure 5-1. As the brush pressure is increased, the unit pressure over the entire brush face in contact with the metal surface is maintained more constant and uniform, thereby reducing the variation in contact resistance across the sliding surfaces which produces what is known as surface contact transients. In addition, the possibility of brush bounce which causes severe arcing and transients is reduced. Since the amount of brush vibration and chatter increases with the

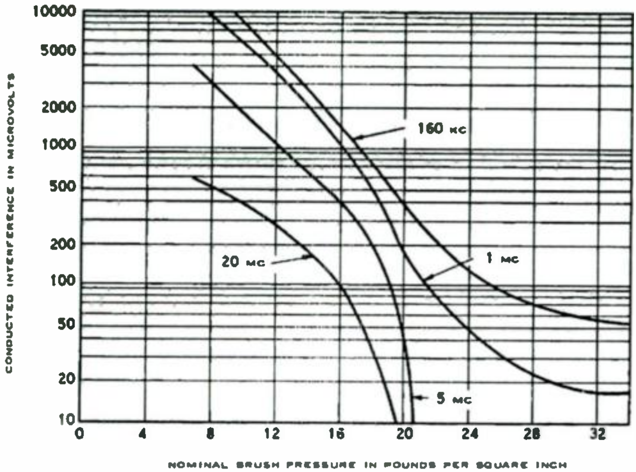


Figure 5-1. Effect of Brush Pressure on Generation of Interference at Various Frequencies

peripheral speed of the sliding metal surface in contact with the brush, the brush pressure selected for any particular application should at least be adequate to damp-out the vibration expected at the designed peripheral speed. Increased brush pressure increases the rate of wear, but the necessity of more frequent replacement should be considered a reasonable compromise for the sake of decreased interference.

2.1.2 CURRENT DENSITY

Interference generated increases with increased current density as shown in Figure 5-2. As the current density is increased, more heat is generated in the contact resistance and the temperature of the brush surface in contact with the commutator increases. This increase in temperature hastens the formation of an oxide film of considerable thickness on the sliding metal surface. Rapid variations in sliding contact resistance due to irregularities in the oxide film modulate the direct current and may give rise to radio interference. The magnitude of interference so produced is increased as the current density, hence voltage drop, across the brush face increases. Therefore, somewhat larger brush surface area should be provided than is demanded only by consideration of the dissipation of heat and losses due to mechanical friction. However, if too low a current density is used, non-uniform grooves or threading develop on the metal surface or commutator, and frequently a high friction coefficient occurs which sets the brushes into a noisy chatter. General design practice calls for a contact current density of 55-65 amperes per square inch at full-load rating for electrographitic carbon brushes, and 65-90 amperes per square inch for metal-graphite brushes.

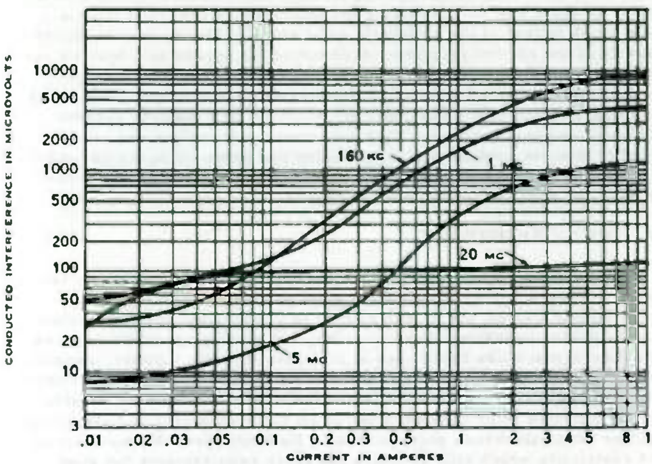


Figure 5-2. Effect of Brush Current on Generated Interference at Various Frequencies

2.1.3 SURFACE TREATMENT

Interference generated is materially decreased by treating the brush surface with a colloidal graphite material. This treatment also decreases the mechanical friction and the voltage drop across the brushes. The metal surface upon which the brushes bear deserves special consideration because of the prolonged effects of wear and high temperatures to which it is subjected. A newly finished metal surface develops an oxide film within several hours. For instance, in the case of a copper commutator in contact with a carbon or graphite brush, a layer of copper oxide, mixed with carbon particles from brush wear, forms on the commutator. The presence of this copper oxide film introduces unidirectional electrical properties (polarity effects) consistent with the well-known copper-oxide rectification. The oxide layer displays a non-linear resistance of higher value to a brush used as a cathode than to one used as an anode. The cathode brush passes current in discontinuous high current density surges. Approximately ten times as much radio interference may come from the cathode brush as from the anode brush. Promising experimental results have been obtained by plating the copper commutator with chromium to a thickness of about one mil, reducing the interference observed from a cathode brush to that of the relatively quiet anode. The somewhat higher resistivity of the chromium plating does not cause appreciable loss as regards heat dissipation at the contact inasmuch as the commutator-film power loss is relatively large. The thinness of the chromium oxide layer and the fine division of the finished chromium surface seem to account for its excellent performance. This performance is maintained because the hard chromium surface avoids threading and grooving of the commutator. Wear rate and sliding friction of many brush materials on chromium is of the same order as that for copper.

2.1.4 BRUSH RESISTIVITY

Interference generated is less for brush materials of lower resistivity. General design practice for good performance is to use an electrographitic carbon brush with 0.0015 to 0.0025 ohm specific resistance in machines operating above 50 volts, and a metal-graphite mixture for brushes in machines being used at less than 50 volts. Silver, copper, or cadmium impregnated graphite is available in the form of low-resistance metal-graphite brushes. The final selection of a brush material is actually a compromise after consideration of all the mechanical and electrical properties which the brush must possess. Nevertheless, the material of lowest resistivity which still satisfies the other requirements for good functional performance should be the preferred choice. The resistivity of brushes used for commutation should also be in accord with the requirements for good commutation as noted in Section 2.2. On the other hand, considerable leeway is permitted in the choice of design and brush material for slip-ring applications since no switching action is involved. The use of low-resistance brushes can be applied to good advantage. For instance, the use of a low-resistance brush composed of many flexible metal contacts sliding on a chromium surfaced slip-ring gives radio interference reduction substantially greater than 10 to 1 over normal brush contacts.

2.1.5 ALTITUDE TREATMENT

Brush and slip-ring or commutator devices used in aircraft rotating electrical devices must be designed for high altitude operation. Brush wear at high altitudes is much more severe than at ground level. This has been attributed to the rarified atmosphere and lack of moisture at high altitudes. Under such conditions a layer of oxide does not readily form on the sliding metal surface. While the presence of this film is undesirable from a radio interference standpoint because of the resulting variations in contact resistance, its presence does reduce wear by serving as a lubricant. Severe wear, especially if it is not uniform, is just as bad from a radio interference standpoint as the presence of a film layer. Consequently, special brushes are used which contain the ingredients necessary for a film layer to form. Impregnated brushes have been developed for this purpose with built-in lubrication and/or oxygen supply. The most successful of such impregnated brushes are those utilizing barium compounds in varying percentages. It should be noted that hermetical sealing of units produces conditions within the container which approximate those at high altitudes. Hence special brushes should be used in such units also.

2.2 COMMUTATION

Commutation is essentially a switching action, and as such is normally accompanied by interference producing transients called break transients. This commutation interference is apart from the brush interference, or surface contact transients, explained in Section 2.1. However, commutation is always achieved by brushes bearing on a commutator so that both contribute to the interference generated by commutator devices. This combined interference is commonly called "motor hash" - a poor term to use in design practice since no distinction is made between brush and commutator interference. For design purposes, such a distinction should be made since the techniques applicable to suppressing each are different. The reduction of brush interference necessitates providing a low, non-varying contact resistance between brush and sliding metal, while reducing commutation interference necessitates the use of special features designed to provide as smooth a transition as possible from one value of current to another. Consideration of the requirements for good commutation may in some cases conflict with the choice of technique used to suppress brush interference. In general, good commutation deserves first consideration.

For this reason, machines requiring commutation are doubly troublesome from a design standpoint. The design engineer should first determine whether a machine requiring commutation is absolutely necessary for a particular application. If, for example, an induction motor can be substituted for a direct current motor, it should be done even at some sacrifice of cost, ease of wiring, or ease of control. Unfortunately, such a substitution is not always possible, mainly because of the high starting torque of certain direct current motors, which cannot be duplicated by alternating current machines except by the addition of bulky and complicated devices. Therefore, where commutation is judged to be essential, special attention must be given to incorporating design features which will minimize the interference generated.

Commutation interference itself, aside from any consideration of brush interference, may be reduced by the use of (1) interpoles, (2) compensating windings, (3) laminated brushes, and (4) careful machining techniques to insure clean, symmetrical mechanical design. All these techniques are intended to smooth out the commutation break transient as much as possible. Figure 5-3 shows a typical oscillographic voltage and current trace of an armature coil undergoing commutation in a machine which employs none of these special design techniques. The diagram actually illustrates an example of poor commutation where strong interference is generated.

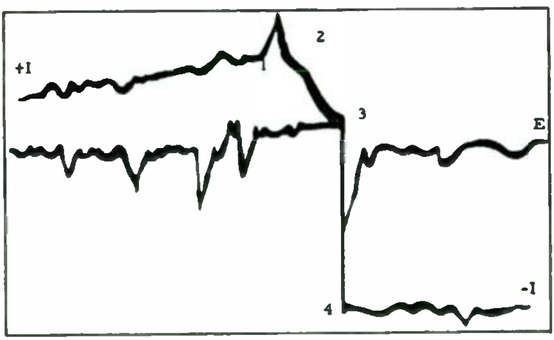


Figure 5-3. Voltage and Current Oscillogram During Commutation Period for an Armature Coil

At point 2 in this figure, the current reversal in the coil nominally begins with initial contact of the leading edge of the brush with the commutator bar. At point 4, the reversal of current I from plus to minus is complete although the trailing edge of the brush broke contact with the bar at point 3 where a break transient is visible on the voltage wave E. Therefore, the commutation break transients are seen to originate at the trailing edges of the brushes; i.e., at the end of the commutation period for each armature coil. The rise in current evident in Figure 5-3 between points 1 and 2 on the current wave I is the result of a break transient produced in the coil immediately adjacent to and in advance of the coil undergoing commutation in the figure. It is evident that transition in current from point 1 to 4 is far removed from the ideal commutation discussed in Volume I, Chapter 1, Section 7.1.1.1, and the generation of severe interference is to be expected. In order to reduce tne generation of commutation interference the steepness of the break transient should be reduced and the initial rate of change of current in the armature coil should be greater. The following techniques are employed to achieve this goal.

2.2.1 INTERPOLES

The best way of improving commutation is by the addition of interpoles. The main function of interpoles in dc machines is to (1) counterbalance the self-induction of the armature coils during the commutation period, and (2) reduce the induced voltage in the armature coils resulting from the coils cutting fringing flux from the pole pieces during the commutation period. Hence, the use of properly designed interpoles would produce a more rapid change in the armature coil current at the beginning of the commutation period and thereby reduce the steepness of the break transient at the end of the commutation period. Interpoles are usually not practical on small machines because of the added weight and lack of space. Yet, careful attention should be given to the possibility of using interpoles even though contrary to normal practice.

2.2.2 COMPENSATING WINDINGS

When interpoles cannot be used, compensating windings in the pole pieces will produce the same effect as interpoles. Compensating windings are rarely used because of the expense involved in cutting the slots into the pole pieces and inserting the windings. In many cases, it will be found that the additional expense is justified in view of the decrease in the generation of interference.

2.2.3 LAMINATED BRUSHES

If practical limits prevent the use of either interpoles or compensating windings, resistance commutation may be used. This scheme should reduce the current in the receding bar to zero at the time the bar leaves the trailing edge of the brush. Actually, as Figure 5-3 indicates, the current does not reach zero at the appropriate time and a steep break transient is produced when the bar leaves the brush.

Instantaneous values of coil current during commutation should depend only upon the ratio of the contact drops of the brush to the commutator bars connected to each end of the armature coil being commutated. Since the self-inductance of the armature coil tends to oppose the change in coil current, the armature current does not divide in the same ratio as the contact drops of the brush to the commutator bars. Circulating currents flow in the coil undergoing commutation by way of the commutator bars and brush, and the rate of change of coil current is not constant, being very slow initially in the commutation period as shown near point 2 in Figure 5-3. If the brush resistivity is large enough to reduce circulating currents in the beginning of the commutation period, the initial current reversal would be at a faster rate. Hence, low brush resistivity, while desirable from the standpoint of reducing brush interference as mentioned in Section 2.1, is undesirable from the standpoint of good commutation.

Lowering the resistance presented to the approaching commutator bar and raising the resistance presented to the receding bar would favor the flow of load current in the approaching bar and also would introduce higher resistance to the circulating current. The use of laminated brushes with brush materials of different resistivity produces such control and hence gives a more linear current reversal. Promising experimental results have been achieved with a brush consisting of segments of different resistivities, the trailing segment having the highest and the leading segment having the lowest. The segments are well insulated from each other.

The ideal operation of laminated brushes is indicated in Figure 5-4. In this diagram, a three-lamination brush is bridging the commutator bars attached to each end of the coil undergoing commutation. The brush resistance increases as the commutator bar progresses from the leading edge to the trailing edge of the laminated brush. The segments are insulated from one another by some suitable glue and electrically connected only at the end in contact with the brush spring. Hence, circulating

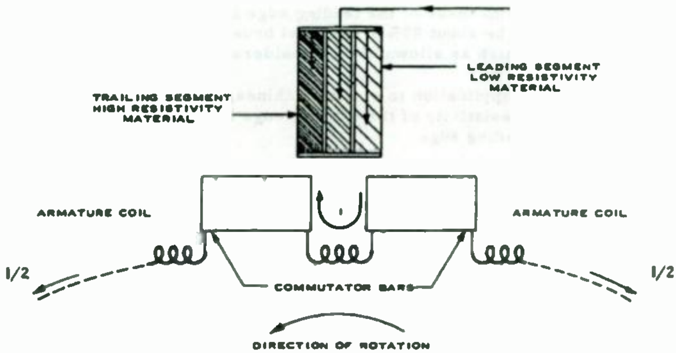


Figure 5-4. Commutation of an Armature Coil by Using Laminated Brushes

currents resulting from the self-inductance of the coil under commutation and from the coil cutting fringe flux from the pole pieces must flow through the entire length of two brush laminations whose total resistance is much greater than that presented by a direct path across the face of the brush as would occur with a nonlaminated brush. Consequently, circulating currents are reduced early in the commutation period and a more desirable division of current through the two adjacent commutator bars is achieved. A more linear coil current reversal is produced and break transients are considerably reduced.

Another consequential advantage to the use of laminated brushes is that good commutation can be achieved over a fairly wide range of brush positions relative to the magnetic neutral, so that this position becomes less critical and less dependent on the armature current.

The details of laminated brush designs are discussed in Appendix III. The lamination design may be summarized as follows:

a. Two, or at most three, laminations give best performance.

b. The thickness of the leading edge lamination of a two lamination brush should be about 90% of the total brush thickness and its resistivity should be as high as allowable by consideration of heat dissipation.

c. For application to small machines, good performance is obtained when the resistivity of the trailing edge lamination is about 15 times that of the leading edge.

d. The trailing lamination should be thick enough to avoid mechanical weakness.

e. The glue thickness should be sufficient to provide electrical insulation and to avoid the formation of a smear of conducting particles from brush wear on the rubbing edge which would short-circuit the laminations. (A thermosetting glue of six-mil thickness has been found satisfactory.)

Reference to Figure 5-3 shows a series of radio interference transients before point 1 where commutation begins. These are the surface contact transients mentioned in Section 2.1. For the case illustrated in this figure the break transient at point 3 is relatively large compared to the surface contact transients. This is usually the situation observed for poor commutation and is accompanied by trailing edge burning of the bars or brushes. Radio interference reduction by the use of laminated brushes is generally limited to such cases where the break transients are relatively large. In other cases, interference reduction can be better achieved by employing one or more of the techniques noted for the reduction of brush interference.

2.2.4 MACHINING TECHNIQUES

Any other design feature that improves commutation in general will also reduce the generation of interference. Considerations such as uniform distribution of armature coils with respect to the commutator bars should be given careful attention in the light of the requirements for good commutation. Careful machining techniques are necessary to insure good electrical and mechanical symmetry. For instance, commutator interference can be reduced by grinding the commutator precisely about its true rotating axis. In a typical case of a very troublesome dc generator, the armature was chucked on the shaft centers, and the commutator was grounded. The interference reduction resulting from this operation was in the ratio of approximately 2:1. The same commutator was then grounded, with the armature chucked in the generator bearings. This resulted in an interference reduction by a factor of about 25:1. The first operation represented the expected rotation axis, while the second operation represented the true rotation axis.

2.3 USE OF CAPACITORS AND FILTERS

Even with the best design, some interference will still be generated at the brushes and during commutation. To prevent this from being conducted to other equipment, capacitors or filters should be incorporated in the original design. In some cases, a simple capacitor of from 0.05 to 1 microfarad, depending on the size of the machine and the amount of interference generated, connected directly across the brushes, is sufficient. In other cases, a complete filter in the output leads may be required.

In present day practice, machines are often constructed without filters, and a filter is added later if excessive interference is produced. These filters are often found ineffective because, once the machine is completed, the brush terminals are not accessible, so that leads of considerable length may be required to connect the filter, or the filter is grounded to the frame of the machine which serves as ground for one of the terminals. Extremely short and direct connections are very important for the effectiveness of a filter. The great advantage of incorporating filters or capacitors into the original design is that connections can be made at the point where the suppressing action is most effective. Much is to be gained, for example, as far as avoiding radiation and capacitive coupling is concerned if the interfering currents can be kept entirely out of the frame of the machine. In addition, a capacitor connected directly across the brushes will be much more effective than one connected from the output lead to the frame, but located several inches away from the brushes.

In the installation of filters or capacitors, the consideration of good bonding is extremely important. In many actual cases, recent experiences have shown that a filter or capacitor was ineffective only because some protective coating was not removed so that no electrical contact was made between the filter and its base. Where installation directly at the brushes is impossible, great care should be exercised in shielding the input leads from the brushes to the filter to the point of entry. Obviously, care should be given to the prevention of any sort of coupling between the "clean" output leads and the "noisy" input leads to a filter.

Filters for 400 cycle alternating current machines require special attention. An ordinary capacitor may draw too much current at the fundamental frequency to be practical. In that case, it is necessary to design a low-pass filter at the proper impedance level with a cut-off frequency of about 600 cycles per second according to the design equations of Appendix I.

2.4 SHIELDING

The brushes and brush leads are the most likely regions from which interference may be radiated or coupled with other circuits. Therefore, unless the entire machine is completely enclosed in such a way as to be adequately shielded, the brushes, brush holders, and brush leads should be shielded as completely as is possible without disturbing their normal functioning. Complete shielding of the entire machine should be used, when practical, in the case of small direct-current machines such as dynamotors, which generate a considerable amount of interference even with the best design. However, the shaft provides a path for interference since it must penetrate the shielding. The shielding of the shaft is done in any of the following ways: using a nonconducting fiber coupling in the shaft, or using a conductive packing for the bearings where the shaft penetrates the shield, or using an additional set of brushes for the shaft near the place where it passes through the shielding.

2.5 SERIES DIRECT-CURRENT MOTORS

Series direct-current motors are frequently used because of their high starting torque. Many of them have split field windings so that they can be reversed quickly without changing more than one connection.

The considerations of Sections 2.1 and 2.2 are applicable to series direct-current motors. They may require filters though first consideration should be given to a good, clean design, which makes filters unnecessary. If filters are used, the series winding may be utilized as their series element so that only one or two additional capacitors are required. The capacitor should be placed as shown in Figure 5-5a because first only one instead of two capacitors is required in case of a split field winding, and, secondly, this position leads to greater attenuation as proved in Appendix II. If this is not practical because of lack of space, two separate capacitors must be used as shown in Figure 5-5b. In several experimental cases, the attenuation was increased by splitting the series field and adding the capacitors as shown in Figure 5-5c. In extreme cases, a pi-section should be constructed by using capacitors simultaneously in all three places. Consideration should be given to

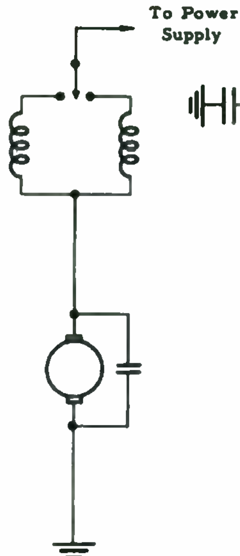


Figure 5-5 a. Location of Capacitor in Series Motor

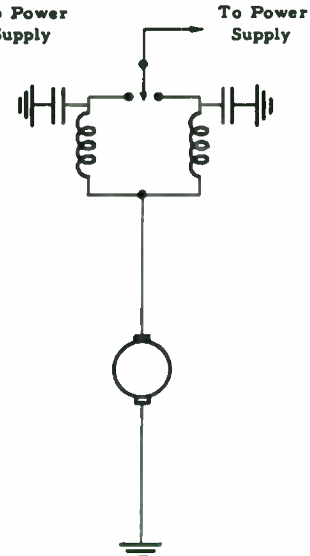


Figure 5-5 b. Alternate Location of Capacitors in Series Motor

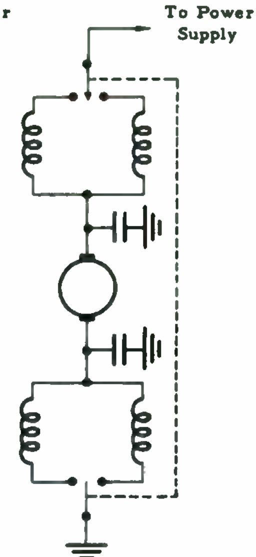


Figure 5-5 c. Location of Capacitors in Split Series Field Motor

isolation of field leads when utilized as a part of a pi-section filter. These leads should be shielded between the armature and field connection. The power input lead should be routed to be as remote as possible from all other leads. The size of the capacitors is determined by the size of the machine and the amount of interference generated, as explained in Appendix I and will normally be about 0.05 to 0.5 microfarads.

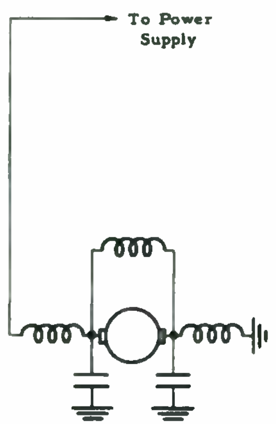


Figure 5-6a. Installation of Capacitors in Direct Current Split-Field Compound Motor

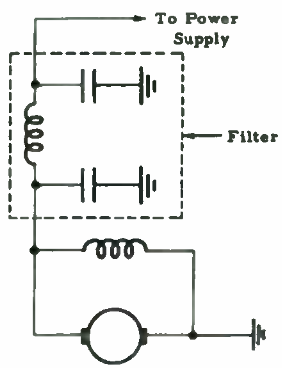


Figure 5-6b. Installation of Filter in Direct Current Shunt Motor

2.6 OTHER DIRECT-CURRENT MOTORS

The same considerations also apply to all other direct-current motors. In a compound motor, the series field may be used as part of the filter as in the series motor as shown in Figure 5-6a. But in a shunt motor, a complete filter must be installed as shown in Figure 5-6b because the shunt field cannot be utilized for that purpose. In most applications, dynamotors are small units placed directly at the location where their output is used. Since they have two commutators, they are fairly effective interference generators, but filters, other than comparatively small capacitors across the brushes, are not practical because of their weight and size. The most practical solution for these small units is complete shielding.

2.7 DIRECT-CURRENT GENERATORS

Direct-current generators are basically very similar to direct-current motors and the same design considerations apply. Since the field is normally fed from the output of the generator itself, this would be classified as a shunt machine, and the field cannot be used as part of the filter. Adequate filters should be incorporated in the positive output lead as close to the brushes as practical. Even here, direct connection of the filter ground terminal to the negative brush is preferable to grounding it to the frame only. Careful attention must be given to proper bonding as pointed out in Section 2.3. If the negative lead is not grounded, a filter must be used in each output lead and a capacitor should be connected directly across the brushes. It should be remembered, of course, that filters are used only as a last resort, and that a good design may make all filters unnecessary.

2.8 ALTERNATORS

The considerations of Sections 2.1 through 2.4 apply to all alternating current generators. In addition, careful attention must be given to the prevention of the generation of harmonics. Production of as pure a sine wave as possible is one of the important "normal" considerations in the design of alternators. But this requirement acquires a new importance and is put to a much more severe test when the generation of radio interference is considered. A comparatively minute harmonic content might be quite tolerable from all points of view except that of radio interference.

Alternators usually use direct-current exciters to provide the necessary magnetic field. These should be designed in accordance with the recommendations of Section 2.7 except that their size usually does not warrant the use of a complete filter, and a single capacitor connected across the brushes will normally provide sufficient filtering action.

To reduce the generation of harmonics, special attention should be given to the following items:

a. Flux Distribution

The most important factor determining the wave form of the generated voltage is the distribution of the magnetic flux around the periphery of the armature. Sinusoidal distribution may be achieved by chamfering the pole tips or skewing the pole faces.

b. Symmetry

For a perfectly symmetrical machine, all even harmonics automatically disappear. Therefore, special care must be exercised in constructing identical pole pieces, making the yoke and armature perfectly symmetrical, producing a perfectly uniform winding on the armature, and avoiding all other irregularities.

c. External Connections

In a three-phase alternator, the third harmonic and its multiples disappear at the terminals except when the machine is star connected and has its neutral grounded, in which case third harmonics are present in the voltage from any phase to neutral. Hence, this connection should be avoided, or else special attention must be given to the prevention of the third harmonic and its multiples.

d. Chord Factor

The harmonics generated may be considerably reduced by the choice of a suitable chord factor. If the difference between the pole pitch and the coil pitch is θ electrical degrees, the chord factor for the nth harmonic is $\cos(n\theta/2)$: A value of 30° is often recommended for θ because this greatly reduces the chord factors for the 5th and 7th harmonics while affecting the fundamental very little.

e. Distribution Factor

If the winding is distributed over several slots per pole per phase and there are m slots per pole per phase, the distribution factor for the nth harmonic is $\sin(mn\theta/2) \;/\; m \sin(n\theta/2)$, where θ is the slot pitch in electrical degrees. This decreases with increasing m and should be chosen so as to eliminate the lowest harminic not eliminated by any of the devices mentioned in b, c, and d.

f. Tooth Ripples

The generation of tooth ripples can be greatly decreased by skewing through one slot pitch either the pole shoes or the armature slots. Also, tooth ripples may be eliminated altogether by making the number of armature slots per pole-pair an odd number. For, in this case, the chord factors for the harmonics that are contained in the tooth ripples reduce to zero.

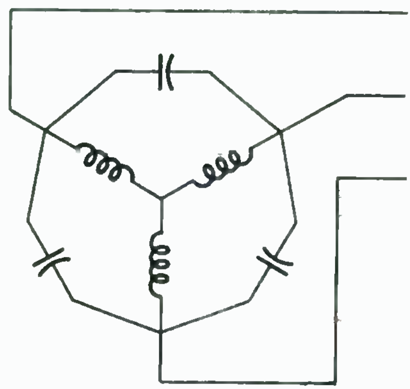


Figure 5-7. Installation of Condensers in Three-Phase Alternator with Ungrounded Neutral

Incorporating the preceeding considerations into the original design of an alternator produces a machine which generates a minimum of interference. Such practice is highly desirable since less demands have to be made on filtering techniques to bring the remaining conducted interference within tolerable limits.

Complete filters are not usually required in the output of well-designed alternators. In a three-phase machine, capacitors should be connected directly across each set of brushes as shown in Figure 5-7, but their value must be somewhat smaller than that recommended for direct-current machines because the current at the fundamental power frequency must be kept low. If filters should be required, they must be designed as low-pass filters with a cut-off frequency of about 3/2 times the fundamental power frequency, and one is required in each lead except the ground lead.

Even a well-designed machine will radiate some interference which must be prevented from leaking out of the alternator housing. This necessitates a casing which acts as a perfect shield and is designed to permit a good bond to ground. The additional requirement of adequate ventilation, especially for a machine in continuous operation, makes it difficult to design a perfect shield. This difficulty can be overcome by providing ventilation through tubular air vents which act as waveguide attenuators to the interference.

Consideration of the problems encountered in a typical motor-alternator unit will illustrate some design procedures that are useful in bringing the generated interference within tolerable limits. The example chosen and discussed below is actually a case of a motor alternator that was well-designed from a functional standpoint and contained a filter as an integral part of the unit. However, from the radio interference standpoint, the original design did not adequately provide for good shielding and bonding. The modifications made to the unit could easily have been incorporated into the original design.

The motor-alternator was subjected to extensive tests which indicated that, after obtaining satisfactory reduction of conducted interference by the use of a filter in the dc power line, excessive radiated interference was present in the medium high and the ultra high frequency ranges. A unique technique for the attenuation of radiated interference in motor-alternators was developed by the Navy. A circular section containing the ventilating louvres was cut from the commutator end-bell of the housing and replaced by a similar section having tubes of small diameter, which act as waveguide attenuators, and at the same time provide the necessary ventilation. This section of the housing before and after modification is illustrated in Figure 5-8. The inside diameter of each tube used was

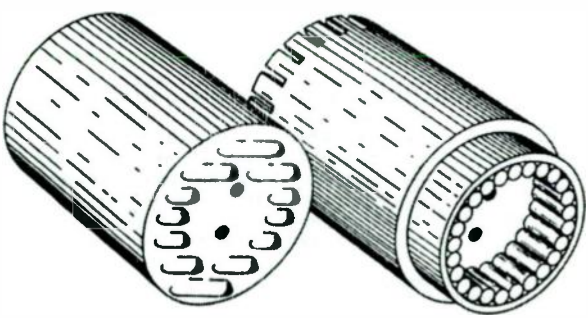


Figure 5-8. Section of Motor-Alternator Shield Before and After Modification

19/64 inch, and its length was in excess of 57/64 inch. The number of tubes used was determined by the requirement that the total cross-sectional area of the openings should not be less than 1.6 square inches to assure sufficient ventilation. Copper was used for this modification but any metal having high conductivity, such as aluminum or brass, could be used provided that all joints are capable of being soldered or welded so that no opening, other than that of the tubes, exists. The necessity for avoiding openings in the shield must be emphasized. In addition, a metal collar no less than 1/32 inch in thickness was placed around these tubes to provide rigidity and protection. The rim of the commutator end-bell was slotted, and the resulting "fingers" were cleaned and polished as well as sprung slightly inward to assure continuous and positive metal-to-metal contact of the friction fitting that joins the commutator end-bell section to the rest of the alternator housing. Such modification to the casing of the alternator provides for adequate attenuation of the radiated interference. In general, the use of standard ventilating louvres should be avoided since the leakage of high frequency interference through such openings can be severe.

Since the example under consideration is a motor-alternator unit, the motor itself must be designed for interference-free operation. Attenuation of the observed motor interference was accomplished by bonding the negative lead from the motor directly to the case as shown by the dashed connection of Figure 5-9. Formerly this lead was bonded to an external

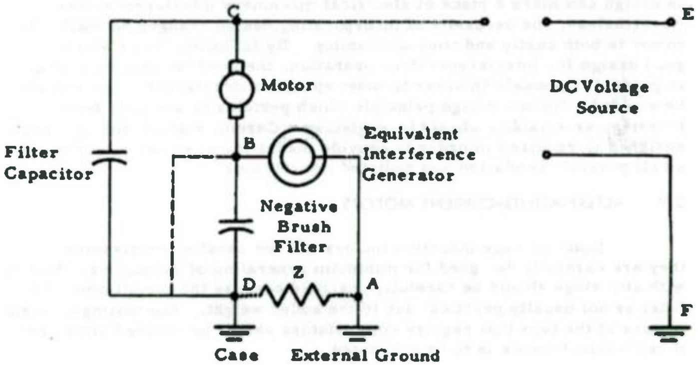


Figure 5-9. Ambiguity of Ground Points

structure which is not at the same potential as the case. Bonding to a structure does not, in itself, assure the existence of a true ground plane at most of the frequencies encountered in interference problems. As a result an impedance, Z, existed between these points as shown in Figure 5-9. This impedance is a part of the closed loop, A-B-C-D-A, through which interference currents generated by the motor flowed. The portion of the same loop, A-B-D-A, which was formed by connecting the negative lead from the motor terminal to the external ground, passed through strong interference fields. Interference voltages were induced in this loop as indicated by the equivalent interference generator in the figure. Since this impedance is also a portion of the loop, E-C-D-A-F, which contains the external power source, any interference voltage that appears across it causes interference currents to flow in the receivers which employ the same power source.

As a result of incorporating these modifications, it was found that in the region of 5 to 20 megacycles the radiated interference was reduced from a maximum value of 42 microvolts and an average value of approximately 20 microvolts down to a maximum value of 2 microvolts and an average value of approximately 1/3 microvolt. In the region of 1000 megacycles, no interference was measurable at any distance from the commutator end-bell after modification.

This example illustrates how oftentimes relatively minor changes in design can make a piece of electrical machinery interference-free. Nevertheless, the necessity of incorporating design changes no matter how minor is both costly and time-consuming. By following the principles of good design for interference-free operation, the need for making changes in production models in order to meet specified interference limits could be avoided. No one design principle which pertains to adequate bonding, filtering, or shielding should be neglected. Careful study of the unit being designed is required in order to provide means for attenuating or bottling up all possible conducted and radiated interference.

2.9 ALTERNATING-CURRENT MOTORS

Squirrel cage induction motors are not usually troublesome if they are carefully designed for minimum generation of harmonics. Motors with slip rings should be carefully designed because the installation of a filter is not usually practical due to the added weight. Alternating-current motors of the type that require commutators should be avoided altogether if radio interference is to be prevented.

2.10 INVERTERS

Inverters are needed to supply alternating current in aircraft whose primary power system consists of a set of direct-current generators. Rotary inverters are inherently sources of interference because they have commutators as well as slip rings, and, in addition, they produce a large number of harmonics of considerable amplitude both on the input and on the output sides.

Since basically a rotary inverter is a direct-current machine with added taps on the armature winding and slip rings connected to these taps, the design considerations for direct-current generators apply here in full force. But in addition to this, filters must be provided in all output and input leads. This imposes a particularly stringent requirement on the alternating current side because the required filters are heavy and bulky; but, for inverters of large size such installation is usually necessary and justified.

3. VIBRATORS

Vibrators are used to convert direct current into alternating current by means of vibrating contacts which alternately make and break the direct-current line. The wave forms of the resulting currents and voltages are more nearly rectangular than sinusoidal. They are therefore very rich in high harmonics and capable of producing a large amount of radio interference. If at all possible, the use of vibrators should be avoided and other means of converting direct into alternating current should be employed in aircraft. When it is necessary to use vibrators, complete shielding and extensive filtering must be employed to keep the interference from causing damage.

There is one particular application of vibrators where the radio interference is not objectionable and where they are being used extensively: Engine-starting vibrators operate only during the brief interval of starting the aircraft engine while the engine speed is not sufficient to allow the magneto to develop a high enough voltage. Once the motor has reached a predetermined speed, the starter switch is disengaged, which also cuts off the vibrator. Such an arrangement would not be suitable, for example, in aircraft operating on an aircraft carrier because radio interference even during the short starting period is not permissible in the vicinity of the many receivers on the carrier itself. But in land-based aircraft, starting vibrators are commonly used.

This starting vibrator, while not in itself a source of radio interference once the engines are running, causes radio interference problems of a different kind: The vibrator unit offers coupling paths through which the ignition interference from the magneto is introduced into the direct-current power system of the aircraft. The interference could be prevented from leaving the vibrator housing by filtering the power lead connecting the unit.

However, filters add weight and do not represent the most desirable approach for radio interference suppression in this case. It was found that appropriate design techniques applied to the vibrator unit itself accomplished the same purpose without the use of filters.

A sketch of a typical vibrator unit is given in Figure 5-10. There are two main component parts, an off-on relay and a vibrator, housed in a metal case.

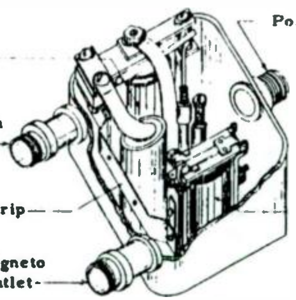


Figure 5-10. Aircraft Vibrator Unit, Original

The moving contact arms of relay and vibrator are attached to their metal frame and are electrically insulated from ground. A shorting strip connects the magneto and ignition switch outlets. A starting mesh-switch, not part of this unit, allows the relay contacts to close, which energizes the vibrator and supplies high surge currents to the magneto primary. This produces a high voltage in the secondary coil of the magneto, which supplies normal output to the spark plugs during the interval required by the engine to reach a minimum running speed. The primary winding of the magneto is in series with a set of breaker points, where the opening and closing of the points create low tension pulses. In this way, steep-wave-front transients are produced, and the resulting radio-frequency energy is conducted back through the primary lead of the magneto to the vibrator.

The interference voltages gain admittance to the direct current aircraft wiring by the following coupling paths:

a. Capacitive coupling across the open contacts of the off-on relay.

b. Capacitive coupling between the shorting bar of the two outlets and the coil windings of the relay and vibrator.

c. Capacitive coupling between the relay frame, which is directly connected to the magneto primary circuit and the shorting bar, and the winding.

d. Electromagnetic radiation from the relay frame.

e. Electromagnetic radiation from the shielding case, since the shield is broken by insulating gaskets on the base plate and cover.

A typical unit was subjected to five modifications, which resulted in a considerable reduction of the radio interference coupled to the direct-current power system. The modifications are the following:

a. The movable contact arm of the relay was tied to the vibrator instead of the magneto lead, as shown in Figure 5-11b.

b. The magneto lead, including the stationary contact point of the relay, was surrounded by a shielding bushing having only one small opening to allow the movable arm to make contact.

c. The shorting bar was removed entirely. The connections to the ignition switch were no longer made through the vibrator unit.

d. The vibrator coil was rewired to minimize capacitive coupling.

e. A conductive gasket was installed to insure good electrical contact between cover and base plate.

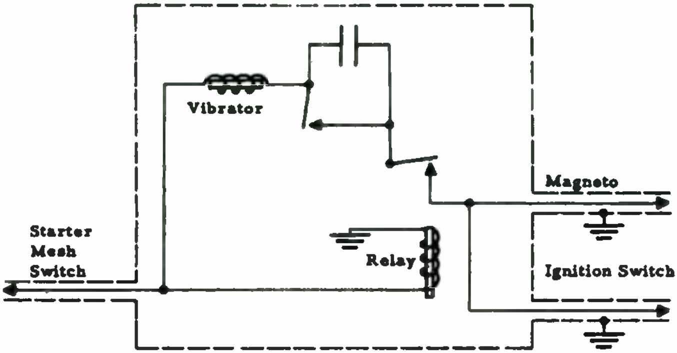


Figure 5-11a. Schematic Diagram, Original Circuit

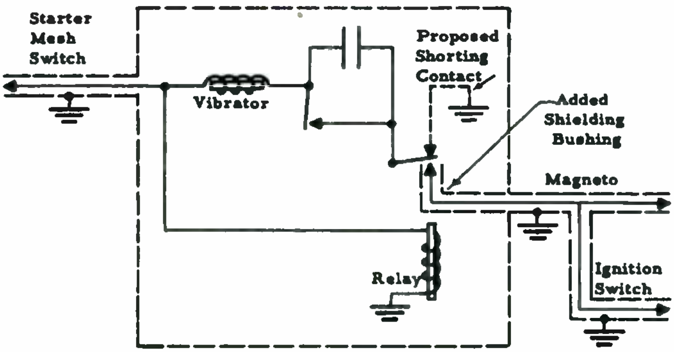
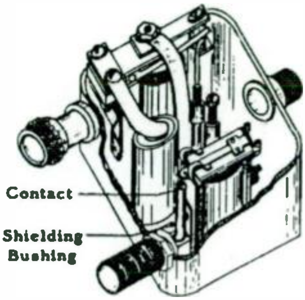


Figure 5-11b. Schematic Diagram, Modified Circuit



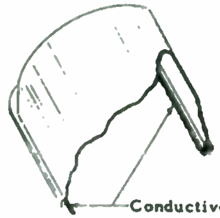


Figure 5-12. Modified Vibrator Unit

The vibrator unit featuring these modifications is shown in Figure 5-12. An additional modification is suggested (but was not carried through in the example shown): The capacitive coupling across the open relay contacts may be eliminated by a grounding contact on the relay instead of the existing open position.

The schematic diagrams of the original and modified type are illustrated in Figures 5-11a and b for comparison purposes. Tests on the modified unit indicate an average reduction of 60 db in the frequency range of 0.25 to 18 megacycles, dropping to 40 db between 30 to 144 megacycles. Grounding the movable contact arm of the relay results in a reduction greater than 100 db between input and output circuits of the vibrator throughout the frequency range of 0.20 to 144 megacycles. The results of such modifications, without the application of filters, indicate that interference in aircraft equipment can be reduced to an acceptable level by the application of appropriate design techniques.

1. GENERAL CONSIDERATIONS

Previous Chapters in this Handbook have presented discussions on methods of suppressing and controlling interference through the use of grounds, bonds, shields, filters, and appropriate interference reducing circuitry. In this Chapter we will discuss some of the overall considerations in controlling interference paths.

In general, interference may be controlled in four ways:

a. By preventing the generation of electromagnetic energy at the source.

b. By confining electromagnetic energy in a predetermined space or circuit.

c. By preventing electromagnetic energy from entering a space or circuit.

d. By modifying the waveform of the energy so that resulting voltages and currents change at a low rate.

It must be particularly emphasized that in the reference to "spaces" and "circuits," we are making reference to all "spaces" and "circuits" which contain electric current or electromagnetic fields, or which contain elements which can be affected by electric current or electromagnetic fields. Although easy to define in this manner, these "spaces" and "circuits" are frequently neglected or overlooked with the result that unwanted electromagnetic energy emanates from or enters a "space" or "circuit" and causes interference with normal circuit functioning.

By controlling entry to and exit from "spaces" and "circuits," we can effectively control the paths by which electromagnetic energy can travel and thus we can control interference. From an engineering management point of view, a program to control interference paths must have its start in basic concepts of design, development, purchasing, maintenance, and operation. By application of electrical engineering fundamentals pertaining to currents, voltages, impedances, and fields, control of interference paths can be exercised throughout a system or equipment.

The degree to which any one or all of the electrical principles can be applied must necessarily depend upon the circumstances of design and functional use. It is important that the interference reduction measures to be taken be decided at an early stage in the design. This decision must, of course, be based largely on two factors:

a. Degree of system degradation that can be tolerated.

b. Cost of incorporating necessary interference-free design features.

It is obvious that a greater degree of interference can be tolerated in some systems such as television broadcasting, for example, than in other systems such as military command and control systems. In many military and missile systems, even a low level of interference can cause catastrophic system degradation. In such cases, the cost of eliminating interference will necessarily be a secondary consideration.

2. EFFECT OF CHANGING FIELDS AND CURRENTS

As pointed out in Section 6, Chapter 1, Volume I, of this Handbook, the basic condition for no interference is that:

$$\frac{di}{dt} = 0 \qquad (6\text{-}1)$$

where: i is the current

$\frac{d}{dt}$ stands for differentiation with respect to time

According to Faraday's law if a magnetic flux, varying with time, is surrounded by a closed linear path, then the electromotive force induced around the path is proportional to the rate of change of flux:

$$e \propto \frac{d\phi}{dt} \qquad (6\text{-}2)$$

Thus, the condition for no interference due to flux variation is:

$$\frac{d\phi}{dt} = 0 \qquad (6\text{-}3)$$

However, again referring to Section 6, Chapter 1, Volume I, variations in a field are found to be due to variations in a current. Thus, if electric circuitry and components can be so arranged that variations in current can be prevented from causing flux variations, or can be prevented from entering interference susceptible electrical elements, then interference can be prevented.

The basic approach to interference prevention discussed above is frequently lost sight of in electrical design. Resulting interference problems may be of such size and complexity that considerable effort and cost is involved in interference elimination, even though the interference source, or transmission media is of a relatively simple nature. In a later part of this Chapter we will discuss some examples of interference sources and transmission media frequently overlooked.

3. SOURCES AND PATHS OF INTERFERENCE

If the ideal condition for no interference is zero change of current with time, then we now have a criterion by which we can examine all parts of an electric circuit to determine the interference control measures to be applied. It may be easier in some cases to directly control current variations while in other cases it may be more appropriate to indirectly control the current variations by controlling field variations. To control current variations, we may use filters, special circuits, grounds, and bonds. To control the effect of field variations, we rely upon shields or upon the placement of susceptible units in positions where field strength and field variation are low. Application of these various methods of interference control is discussed in detail in Chapters 3 and 4 of this Volume.

Since rapid variations of currents or fields are the basis of interference, then circuits and systems which are susceptible to interference must be carefully examined to determine whether unwanted currents and fields are present; and, if present, their rate of variation and their magnitude must be analyzed. Although this would appear to be a simple procedure, it is sometimes, in practice, difficult and complex.

In the first place, some interference sources are frequently overlooked, either because they are not normally considered to be generators of current or fields, or because it may be thought that their rate of change and magnitude is so small as to be negligible in its effect. In the second place, measurement of the rate of change and the magnitude of the current or field may be difficult due to inherent limitations of test instruments and measurement techniques. And in the third place, determination of the location of the source may not be easy for both technical and physical reasons.

Possible interference sources are so numerous that a complete listing is almost impossible. Some general categories of such sources are listed and discussed in Chapters 1, 2, and 3 of Volume I of this Handbook. These sources are listed again below for easy reference.

- Radio and Radar Equipment
- Rotating Machinery
- Mechanical Switches
- Transmission Lines and Connectors
- Electronic Devices
- Telephone and Telegraph Lines, Circuits, Switches, and Dials
- Incandescent Lamps
- Fluorescent Lamps
- Mercury and Sodium Lamps
- Electric Welders
- Engine Ignition Systems
- Industrial, Scientific, and Medical Equipment
- Household Appliances
- Business Machines
- Electric Power Controllers
- Atmospheric Disturbances
- Star and Solar Activity
- Precipitation Static
- Non-electric Sources

In brief, our list shows that, not only can almost any electrical or electronic device be a source of interference, but, in addition, interference may arise from sources not normally considered to be part of an electric or electronic circuit. Thus, it becomes even more apparent that interference analysis of circuitry and equipment must be made from the standpoint of fundamental principles.

Once the source of interference has been determined, it is then possible to analyze the possible paths by which the interference may travel to a receiver. The definition of "receiver" here is the same as that presented in the introduction to Volume I - "any type of electrical and electronic equipment in which radio interference may cause an undesirable response or malfunctioning." The interference paths are arbitrarily categorized into two types - space paths and physical circuit paths. This means of categorizing the two types of paths is selected only because it allows ready mental visualization of the media by which interference can travel to a receiver. The space paths can be considered to be general paths through air, gas, water, or vacuum, whereas the physical circuit paths are considered to be those through wires, coaxial cables, vacuum tubes, semi-conductors, and similar solids. The theory of propagation in these various media is well covered in the literature, and it will suffice to say here that there is as wide a variety of paths as there is of sources.

The specific control measure to be taken for a particular interference situation will depend upon the circumstances of the situation and the characteristics of the interference. For example, if it is desired to eliminate (or reduce) spurious emissions from a very high power radar, it may be determined from the technical and military circumstances of the radar operation that the only feasible method of suppression is the installation of a filter in the antenna feed system. Such a filter is very costly and is difficult to design and construct because of the large RF power handling capabilities required. It would only be used if no other interference control method was feasible. Thus, in this case the interference path would be controlled by using methods c. and d. of Section 1 of this Chapter (preventing electromagnetic energy from entering a space, and modification of waveform).

On the other hand, an interference problem in a miniaturized equipment might necessitate the use of a special shield because of the proximity of RF sensitive circuit elements and the difficulty of using other control measures in the available space. Thus, in this case, the path of the interference is controlled by using method b of Section 1 of this Chapter (confinement of electromagnetic energy in a predetermined space or circuit).

4. APPLICATION OF CONTROL METHODS

As has been pointed out, the decision as to the interference control methods to be used must be based on both technical and cost factors. From a purely theoretical standpoint, it would always be desirable to use method a of Section 1 for interference control (prevention of generation of electromagnetic energy). Unfortunately, in many electronic and electrical systems this method of control cannot be exercised as, for example, in most dc motors and generators, or other electrical machines using commutation devices. In these cases, there is no alternative to the application of one of the other three control methods.

However, in general, the equipment and system designer should seek to prevent interference from being generated. Although this may require some additional engineering effort in the design stage, it may prevent large-scale, costly problems from arising at a later date.

Prevention of interference generation in original design has an additional advantage in that it allows more freedom in other aspects of system planning, since there will be a reduction in problems pertaining to control of interference paths.

If interference generation cannot be prevented, then control of interference paths becomes mandatory by one of the other methods shown in Section 1. In general, it is desirable to use filters only if other methods are either too costly or technically too difficult.

4.1 PRINCIPLES OF APPLICATION

Certain basic interference control principles can be applied in the design and use of communications and electronics equipments specifically, and electrical equipments generally. It must be remembered that any changing electrical current, whether in a radio transmitter or in an arc welder, is capable of causing interference if its frequency and power can affect a receiver. The following paragraphs discuss briefly the control principles which should be applied as necessary.

4.1.1 EXAMINATION FOR INTERFERENCE SOURCES AND PATHS

The circuit and the equipment installation should be examined carefully to determine whether changing currents or fields exist which can cause interference. Be sure that all parts of the circuit and equipment are studied, including those parts not normally considered as interference sources and paths such as, for example, audio circuitry, ac power input, low voltage ac in filament circuits, openings where RF energy may enter or exit, and lighting circuits.

4.1.2 SEPARATION OF CIRCUITS

Separate as far as possible wires, cables, and leads containing RF and ac from each other and from dc and audio wires.

4.1.3 AVOIDANCE OF GROUND LOOPS AND HIGH GROUND IMPEDANCES

Use appropriate grounding methods in the equipment so as to avoid ground loops or spurious emissions, and at the same time keep all ground impedances as low as possible.

4.1.4 APPROPRIATE CONTROL METHODS FOR ALL FREQUENCIES INVOLVED

Use appropriate shielding, grounding, bonding, and filtering methods for all frequencies which may cause interference. For example, the pulses of radar transmitters generally have low fundamental frequencies with many harmonics due to sharpness of wavefront. These harmonics may extend over many megacycles of the radio spectrum from LF to UHF. Control of interference from these pulses is required in addition to control of interference from the fundamental RF output.

4.1.5 INTERFERENCE CONTROL IN DESIGN DETAILS

Use care in details of design, construction, and installation. Make sure that the integrity of shielding is not violated by small openings (such as those around covers or doors) or inconspicuous openings (such as those around power lead or plumbing entries). See that all separate parts of the shield are in very low resistance electrical contact, with no discontinuities, using sufficient screws and screw pressure (or other methods of fastening such as soldering or welding). At all metal-to-metal contacts, be sure that the metals are compatible and that corrosion will not cause a rise in contact resistance. Use lock washers or other devices where necessary on ground, shield, and bond connections to ascertain that there is direct pressure contact to the metal to penetrate surface corrosion. Likewise, be sure that metal-to-metal contacts are chemically and electrically clean so that surface preservatives or other treatments do not cause high contact resistance. Keep RF emitters and RF sensitive devices as far as possible from shields and from shield discontinuities.

4.1.6 USE OF RF GENERATORS

Use vacuum tubes and other RF generators in such a way as to cause minimum interference. This implies the use of:

a. Minimum power consistent with operational requirements.

b. Tubes of low transconductance so as to reduce spurious frequency generation.

c. Tubes of minimum HF performance in order to reduce emissions at frequencies higher than the fundamental.

d. As little frequency multiplication as possible in order to reduce overall RF emissions.

f. RF generator loads which do not cause broadband emissions.

4.1.7 GROUNDING AT SHIELDING EXITS AND ENTRANCES

Ground cable shields, coaxial feed-throughs, and filters at shields at point of exit or point of entrance.

4.1.8 RELATION OF LENGTH OF GROUND LEADS TO FREQUENCY

Check length and electrical characteristics of ground leads against frequency. At the higher frequencies, a ground lead, which has a length which is an odd multiple of a quarter wavelength, may have a very high impedance to ground. Likewise, high impedances in ground leads may be caused by sudden changes in configuration as when leaving a metal duct or otherwise changing the electrical relationship to metallic bodies. Such changes produce serious reactive mismatches which vary with frequency.

4.1.9 SWITCHES AND SWITCHING CIRCUITS

Suspect all switches, switching circuits, relays, controllers, vibrators, and other similar devices as RFI sources. Except in cases where special attention has been given to interference-free design, these components are generally large-scale generators of both conducted and radiated interference. Semi-conductors, when used as switches or rectifiers, will almost always cause interference.

5. CONTROL TECHNIQUES

Control techniques based on the application of control methods as discussed in Section 4 of this Chapter involve the use of any method or combination of methods which will provide control of interference paths. In this Section we will discuss control techniques for some general interference situations.

5.1 STRUCTURES AND ENCLOSURES

When sensitive electronic equipment is used in a building, or near a building containing devices which emit electromagnetic energy, it becomes desirable to construct the building so that the building enclosure prevents passage of the energy. The RF characteristics of the ultimate building enclosure from an interference standpoint would approach those of a completely shielded room as discussed in Chapter 5 of Volume II and Chapter 3 of this Volume. For large buildings the materials become very expensive using presently available techniques. Difficult problems are encountered in providing proper control of openings for doors, stairways, elevators, ventilators, air conditioners, utility entries, heating systems, and handling of building supplies.

Some work has been done in the study of interference shielding in buildings by the use of materials other than metal. For example, the Naval Research Laboratory has investigated the use of coke-aggregate concrete as a shield to electromagnetic radiation.[1] Insertion loss varied from about 30 db at frequencies lower than 30 mc to as high as 80 to 100 db at higher frequencies. (These figures were obtained with test samples in the laboratory and therefore would not necessarily be representative of attenuation possibilities in an actual building.)

Use of electromagnetic shielding is necessary or desirable in many types of buildings. As interference manifestations become more critical in modern civilization, it becomes increasingly necessary to add some degree of shielding to old buildings and to have architects design RF suppression into new construction in order to prevent degradation of electronic device operation in such establishments as hospitals, scientific laboratories, electronic manufacturing facilities, and communications centers.

The type of shielding installation and the extent of shielding in a building will obviously depend on the building function, and the amount and placement of electronic and electrical equipment. Probably the most widespread and most critical requirements are those of the military services where there is increasing use of very sensitive receivers and high power transmitters in communications, radar, navigation, and control systems. The wide variety of environmental conditions encountered in military operations further amplifies the problems of obtaining proper installation and maintenance of shielded structures. Construction and maintenance supervisors and workers are frequently unfamiliar with and, sometimes, completely unaware of the importance of providing shielding integrity throughout the structure. Loose bolts, unsatisfactory welds, poorly installed doors, and similar conditions frequently reduce shielding attenuation below the minimum required.

To provide proper interference shielding in a structure, the shield must be looked upon as an electronic device or an integral part of an electronic system. Appropriate realistic design criteria should be established, and construction specifications and plans should reflect these criteria. Construction must not only be sound mechanically but must be carefully done so as to meet the interference specifications. Great care must be used in shipping and handling the shielding materials. In fact the same care should be used as with any other fragile electronic equipment. Torque wrenches should be used on bolts, welds should be very carefully accomplished, and, in particular, auxiliary devices used in doors, penetrations, and other openings should be handled with care.

Waveguide filters used in ventilating openings, door frames, finger stock for door stripping, gaskets, and door locking mechanisms are particularly susceptible to damage in shipping, handling, and installation. Damage or improper installation of these items may result in reduction of attenuation by as much as 30 to 40 db.

Development is now in progress on construction materials which may reduce the cost and technical complexities of constructing RFI-proof buildings. Conductive building cement, wall boards, paint, and similar materials may eventually become available. Copper foil in large sheets has already been used in laboratories, hospitals, and industrial buildings as an RFI shield (Figure 6-1).

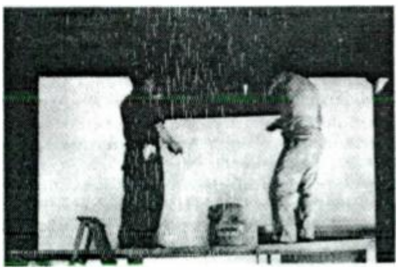


Figure 6-1. Dielectric Welding Department in an Industrial Plant Electrostatically Shielded with Two-ounce Anaconda "Electro-Sheet" Foil Bonded to Building Paper
(The shielding was applied with adhesive and the seams clinch locked.)

(COURTESY AMERICAN BRASS COMPANY)

In general, wherever possible electrical work in a communications-electronics installation should comply with the National Electrical Code.

This code is sponsored by the National Fire Protection Association and the general purpose of the code is to reduce to a minimum the hazards of electric circuitry and wiring. The NFPA and other groups are also concerned with the engineering and installation of ground systems. The interest of each group is, somewhat different but, in general, safety and electric system stability have been prime objectives of past studies. In recent years the importance of grounds in interference control has been recognized and the military services, in particular, have devoted considerable attention to design of ground systems for communication-electronic installations. In the case of receiver and transmitter stations, typical grounding specifications might be similar to the following:

a. All metal pieces that are greater in any dimension than 9 feet and are not energized parts of the electrical distribution system, or are not specifically covered by the drawings or notes thereon, shall be connected to the nearest ground by copper conductors.

b. All grounding conductors or bonding jumpers, except where otherwise noted, are to be No. 10 soft or medium-hard-drawn copper wire.

c. All connections of copper to copper shall be brazed. Other connections can be welded.

d. Sheet copper ground strap shall be made continuous if possible with folds for change of direction. Where connections are necessary, they shall be lapped 6 inches and brazed. Copper strap, 20 gage, 1-inch wide, can be used in place of the No. 10 wire where greater flexibility is necessary.

e. An operations building ground system comprised of a 6 inch by 1/32 inch copper strap perimeter ring with ground rods and buried radial wires should be installed. Ground straps from trenches, structural metal, etc., should be connected to this ground system.

f. Electrical conduit, duct, raceways, etc., and their fittings, such as outlet boxes, fuse panels, etc., are to be spot welded at joints and connected to building ground at one point.

g. All piping and fittings for plumbing and heating (such as valves) shall be spot welded at each joint; for cast iron pipe, a copper jumper shall be used at each joint.

h. All ducts, roof ventilators, cooling equipment, plaster rings, etc., shall be connected at each joint with jumpers and at one point connected to nearest ground.

i. All joints in roof flashing, counterflashing, scuppers, leader heads, leaders, etc., are to be spot connected by soldering, brazing, or welding, or bonded with short jumpers. All parts are to be connected to nearest grounded metal. Flashing and counterflashing are to be bonded to each of the vertical ground straps in the walls approximately 10 feet from center to center. Leaders should be connected to the perimeter ring at lower ends.

j. All steel framing members shall be connected by copper ground strap. All metal-to-metal contact, including bridging contact, shall have at least one spot weld.

k. Steel lintels shall have one spot welded to steel sash and be connected to the tier with copper jumpers. Louvers are to be connected to the tier and to ground with copper jumpers.

l. All metal partition work shall be spot welded at all joints and connected to the building ground at one point with a 2 inch by 1/32 inch copper strap.

m. Machinery, motors, pumps, fans, etc., shall be connected to nearest ground with copper jumpers.

n. Where reinforcing rods are used in foundations or walls, all vertical rods shall be welded to the bottom horizontal rod; the bottom horizontal rod shall be connected to the perimeter ground ring with 2 inch by 1/32 inch (minimum) copper straps at intervals not greater than 10 feet. All horizontal rods shall be connected to vertical rods at intervals not greater than 10 feet.

o. Heat insulation batts with vapor barrier metal foil will not be used. Where required, heat insulation material shall be limited to electrically nonconducting type.

Obtaining a low ground resistance is of importance in all ground systems with the ultimate objective of obtaining zero resistance to earth. Actual resistance to earth for a well-designed ground system may be in the neighborhood of one ohm. Earth resistances at small substations and at industrial plants may be five ohms or lower. Resistivity of soils varies considerably in accordance with moisture content, temperature, type of soil, chemical composition, and vegetation. Soil resistivity may be reduced from 15 to 90 percent by treatment with such chemicals as common salt (sodium chloride), magnesium sulphate, copper sulphate, and calcium chloride. Chemicals are generally placed in a trench around each ground electrode but not in direct contact with the electrode. The effects of the treatment may not become apparent immediately. They can be hastened by saturating the area with water. The treatment is not permanent and must be repeated from time to time.

5.2 COMPATIBILITY PACKAGING

Closely allied with and technically similar to RFI problems in structures are RFI problems in packaging of equipment. Mechanically, these problems are somewhat easier to solve because of the relatively small size of the packages. Nevertheless, the same electrical problems exist and the same general principles are followed in obtaining RFI-tight packages.

The compatibility packaging design for electronic equipment is based on the use of shielded enclosures. Filters for critical circuits must be provided and openings must be protected.

Development work in recent years has resulted in greatly improved methods of sealing enclosure openings such as filter feed-throughs, small access doors, cover plates, and hinged lids. Joints in these openings are made mechanically sound and RFI-tight by the appropriate application of gaskets of various types. A few examples of these gaskets are shown in Figures 6-2 through 6-5. Details of gasket application are given in Chapter 3 of this Volume. For permanent joint closure soldering, welding, or brazing may be used. New conductive adhesives and plastic materials are becoming available which may lower the cost of joint closures.

In electronic compatibility packaging, the same features of design are critical as they are in structures. Care is required in design, in assembly, and in use. Each package must be made exactly in accordance with the plans and specifications with the same relation of parts maintained in each equipment. Bonding, grounding, shielding, and filtering practices for electronically compatible packages are discussed in Chapters 3 and 4 of this Volume.

5.3 WIRES, CABLES, AND TRANSMISSION LINES

An important interference path is the metallic electric current conductor as represented by a wire, cable, or transmission line. Since

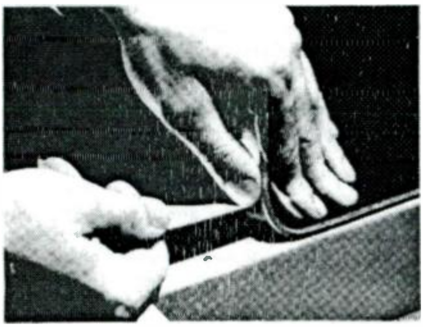


Figure 6-2. Pressure Sensitive Adhesive-Backed RFI Gasket and Elastomeric Neoprene Fluid Seal
(COURTESY TECHNICAL WIRE PRODUCTS, INC.)

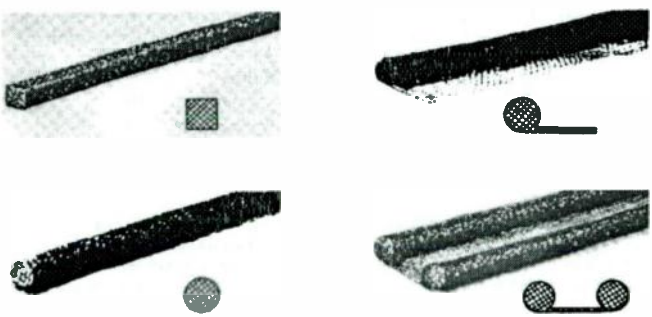


Figure 6-3. Several Configurations of RFI Gaskets
(COURTESY TECHNICAL WIRE PRODUCTS, INC.)

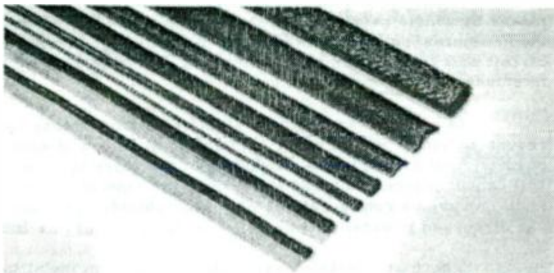


Figure 6-4. Special RFI Gaskets
(COURTESY METEX ELECTRONICS CORPORATION)

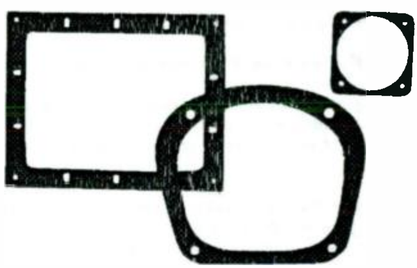


Figure 6-5. Combination RF Gaskets for Use in Pressure Tight and RF Tight Joints
(COURTESY METEX ELECTRONICS CORPORATION)

the conductor provides for a flow of current from one point to another in a system, it presents an ideal path for concentrations of unwanted interference currents. It is well to recognize that this is one of the most important means by which interference can travel between two points. In addition, the conductor may act as a radiator to scatter interference throughout the circuit or system. It may also act as a pickup antenna by which outside interference is brought into the system.

In general, appropriate shielding of conductors can materially reduce radiation and pickup. Application of filters at sources can prevent conduction of interference. Twisting of the pairs in a two-wire circuit will cancel induced voltages and reduce the amount of pickup. Interference tests on various configurations of wires, shields, grounds, and sources are discussed in detail in Chapter 7 of Volume II of this Handbook.

There are certain basic rules to follow in installing wires and cables for minimum interference:

a. Lines carrying dc keying signals should be shielded and separated from lines carrying RF and AF currents. Building cables should be installed in ducts, preferably with power and dc keying cables in the floor ducts and AF and RF cables in overhead ducts.

b. AF lines (including tone signal lines) should be shielded pairs with the shields well grounded.

c. RF lines should be coaxial cables.

d. Telephone and power cables entering communications-electronics installations should use buried approaches to the communications building, and particularly should be buried when passing near or under antennas.

e. Care should be used in grounding wire and cable shields in order that the ground circuit does not provide a path for interference and hum. Particular attention must be paid to those circuits containing very low level signals as in instrument circuits.

5.4 POWER TRANSMISSION LINES

Overhead transmission lines are a common source and path of interference. The widespread use of ac power systems in modern civilization has established a network capable of generating, conducting, and radiating interference over large areas. Coupling through transformers and ground circuits can carry this interference directly into electronics equipment through distribution system low voltage building power supplies.

A study made by the Canadian Department of Transport, Telecommunication Division, showed that over a period of a year, power and transmission line sources were 37 percent of all RFI sources reported.

Approximately half of the interference from power lines has been found to originate at loose tie wires and defective insulators. Proper insulator design, proper application of tie wires, and cleanliness of the insulator-tie wire assembly will materially reduce this type of interference.

Good design practices, care in construction, and considerable maintenance are necessary if power transmission lines are to be prevented from being interference sources. Important considerations from an RFI standpoint are:

a. Insulators should be post type of proper voltage rating. If suspension insulators are used, they should be weight loaded by 340 lbs.

b. Tie wires should be installed with ends that are rounded or square to avoid sharp points. They should make good contact with the conductor and the ends should be as far from the conductor as possible (see Figure 6-6).

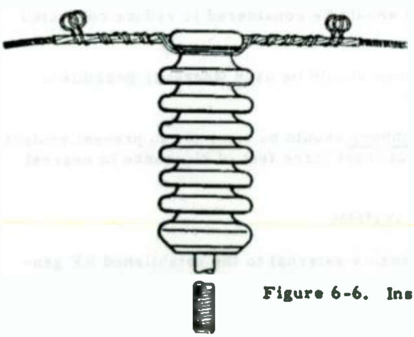


Figure 6-6. Insulator Line Tie

c. Conductors should be as large as economically feasible. Use pressure sleeves or automatic splices for splicing .

d. Poles and crossarms should be of wood construction if at all feasible. This will prevent corrosion leaks and rubbing metal surfaces.

e. Hardware should be static-proof with spring washers or lock nuts. Hardware should be spaced at least two inches apart.

f. Guy wires should be clear of other metal structures along the line and pole and should be kept taut.

g. Lightning arresters should be interference-free types.

h. Ground wires should be clear of all other metal parts. Resistance to earth should be less than 10 ohms.

i. Pole top switches should be at least two inches away from all other unbonded metal on the pole. Switch contact surfaces should be clean and should make positive connections.

j. Insulator contamination should be checked carefully, particularly in industrial areas. Insulator washing may be necessary at regular intervals.

k. Substation low voltage lines should leave the station at 90° to the high voltage lines. Bushings should be of an interference-free type.

l. RF choke coils should be considered to reduce conducted interference.

m. Underground lines should be used wherever possible to obtain additional attenuation.

n. Trees and shrubbery should be trimmed to prevent contact with wires. There should be at least three feet of clearance to nearest branches.

5.5 EXTERNAL NON-LINEAR SYSTEMS

Equipment and components external to the established RF generators of a communication-electronic system may contain or develop non-linear elements which act as detectors or generators of RF energy at frequencies other than the fundamental of the system. These non-linear devices may be vacuum tubes, corroded metals, semi-conductors in electronics circuits, and any other device in which the output is not linearly proportional to the input.

Many natural products and manufactured devices are rectifiers. Oxides and other corrosion products, lead sulfide, silicon, and germanium diodes fall in this category. If the material or device will pass current better in one direction than in the other, harmonics will be produced of the alternating currents through it. The strength of the harmonics depends upon the amount of current, the resonant frequency of the circuit of which the device is part, and the efficiency of the device as a rectifier.

A house or building, if observed as a metallic skeleton, could be seen to have a maze of conductors, joints, and fixtures which might act as sources and radiators of RFI. Any point where there is a rusty or corroded joint, or where there is intermittent contact between two metallic parts, may act as a generator of RF energy and associated pipes and other metal parts may act as transmitting antennas. Likewise, such rectifiers may be affected by incoming RF energy and reradiate or conduct energy at harmonic frequencies.

Non-linear devices and materials may be found in pipe joints, furnaces, hot water installations, guy wires, power and telephone installations, metal roofs and gutters, structural steel, house wiring for bells, intercoms, lights, and household appliances, and fences, to cite a few examples. In addition, non-linear devices may be found in any electrical or electronic apparatus.

External non-linear devices have frequently been found to be the cause of interference in situations where considerable difficulty was found in locating sources. Unfortunately, since such sources may be buried in the walls, floors, or ceilings of a building, the removal or correction of the problem, although simple in itself, may require a major operation and repair on the building. Some troubles with unaccessible non-linear devices may be solved by bonding to the pipe or metallic structure on each side to ground, effectively grounding the system.

6. TYPICAL INTERFERENCE CONTROL PROBLEMS

Interference may originate in many ways and no circuit, material, device, or equipment should automatically be eliminated from consideration as a possible RFI source. The examples presented in this

Section have been selected at random from the literature and are representative of the large variety of practical interference situations which may arise:

a. Flashing neon sign caused television interference due to arcing between high voltage lead and metal body of sign. Arcing due to broken porcelain bushing.

b. Intermittent interference buzz occurred 20 to 40 times per minute received in 40 to 120 mc band. Found after considerable investigative difficulty to be a fault in the rewind motor of an electric clock.

c. Oil burner furnace produced interference when ignition was on. High voltage leads were shortened and suppressor filter was applied.

d. Interference was caused from 4400-volt line due to conductor being off of insulator and in contact with steel insulator pin. Interference existed over a wide area along the line. The situation was found by hitting pole with a hammer which caused obvious changes in interference intensity. (The hammer test on electric power poles is frequently used when a transmission line or pole-mounted equipment is suspect.)

e. Interference was caused by a cash register. Cured by a resistor across speed control contacts and a filter in motor cable.

f. An industrial heater caused interference on a marine distress frequency at 100 pulses a second due to its use in the manufacture of ball bearings at that rate.

The above interference situations are but a few examples of the type of sources that must be considered in interference studies. Many others could be presented. Since any changing electric current may be an interference source, the possibilities are unlimited.

7. FUTURE INTERFERENCE CONTROL PROBLEMS

The future holds many new developments in communications and electronics. The vastly increased use of electronics for military command and control, the new uses of the radio spectrum by individual citizens in communicating with each other, and the advent of space exploration, and the use of space relays all indicate an enlargement of interference problems for the future. Receivers will be more sensitive, transmitters more powerful, greater numbers of equipments will be in use, and redundancy factors will tend to decrease. Every sign points to the importance of increasing efforts to control interference sources and paths.

One source[2] estimates that the message rate will about double in the next ten years; increase in message demand due to increasing traffic congestion will also double; and that ionospheric and scatter phenomena, natural and artificial, will lead to an additional doubling. Thus, there will be in ten years about an eight-fold increase in traffic density.

Whatever the actual increase in use of the radio spectrum, our requirements for prevention of interference by improved technical use of the spectrum and by better spectrum management will continually become more severe, perhaps by an order of magnitude or more. Revision and tightening of equipment and system specifications and standards, operating rules and regulations, and communications laws will be necessary. New developments will be required in techniques for locating sources and paths of all types of electromagnetic energy and for making measurements. Indeed, it is probably in RFI measurement instrumentation that we may expect to have the next large improvement in test equipment.

In light of our forecasts, we can readily see that control of interference paths will require intensified engineering attention in all communications-electronics systems. From design through production to operation, interference control may become a prime objective in the development of manufacturing methods, prediction and measurement techniques, and the application of interference suppression measures.

References

1. P. F. Nicholson: Study of Coke-Aggregate Concrete as a Shield to Electromagnetic Radiation, NRL Report 5473, U. S. Naval Research Laboratory, Washington, D. C., 6 May 1960.

2. J. E. Hacke, Jr.: The Future Natural, Technological, and Social Environment of Global Communications, Second Annual National Symposium on Global Communications, St. Petersburg, Florida, 3 December 1958.

3. C. C. Borden: Shielded Rooms for Electronic Equipment, Architectural Record, September 1961.

4. J. J. O'Neil: R. F. Shielded Structures, Ground Support Equipment Magazine, October/November 1961.

5. R. Daniels: Compatibility Packaging, Electronic Packaging and Production, July/August 1961.

6. Grounding of Industrial Power Systems, AIEE No. 953, A Report of the Industrial Power Systems Committee, American Institute of Electrical Engineers, October 1956.

7. Television Interference, Third Edition, Remington Rand Laboratory of Advanced Research, South Norwalk, Connecticut.

8. M. Seybold: Harmonic Radiation from External Nonlinear Systems, QST, January 1953.

9. Circulars of the Telecommunications Division, Canadian Department of Transport.

10. J. Darr: Eliminating Man-Made Interference, a booklet published by Howard W. Sams and Co., Inc., November 1960.

11. Policy Planning for Space Telecommunications, Staff Report prepared for the Committee on Aeronautical and Space Sciences, U. S. Senate, 4 December 1960.

Interference filters may be constructed from reactances in various possible configurations; however, the information presented in this appendix will be confined to ladder filters of the low pass type shown in Figure I-1.

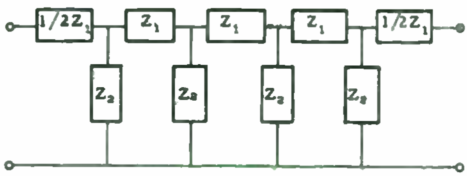


Figure I-1. Ladder Network Filter

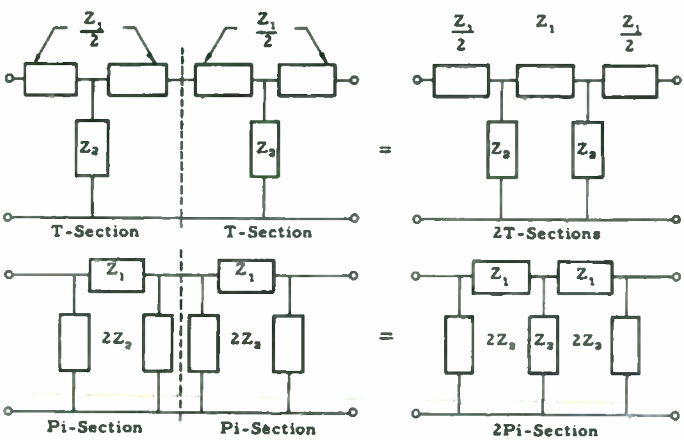


Figure I-2. Basic Configurations Included in Ladder Filters

The series impedance Z_1, and the shunt impedance Z_2 consist of inductive or capacitive reactances or a combination of both.

Both the T- and Pi-sections, as shown in Figure I-2, are grouped under the common heading of ladder networks. Each of the two series arms of the T-section is equal to $Z_1/2$ resulting in a full series impedance of Z_1, while each of the two shunt arms of the pi-section is equal to $2Z_2$ resulting in a full shunt impedance of Z_2.

Figure I-3 shows a T network with three arms Z_a, Z_b, and Z_c, connected to a source of voltage E whose internal impedance is Z_{I_1}, and terminated by an impedance Z_{I_2}. If the impedance Z_{I_1} is equal to the impedance looking in from the terminals 1-2, and similarly, if the impedance Z_{I_2} is equal to the impedance looking in from terminals 3-4, then the impedances Z_{I_1}, and Z_{I_2} are called the image impedances of the T network. A similar relationship can also exist for a Pi network.

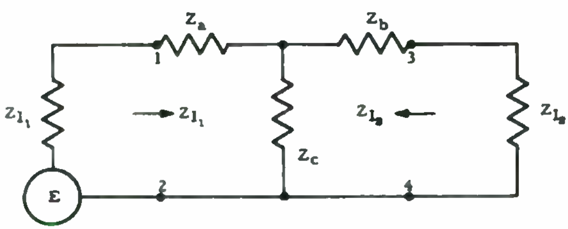


Figure I-3. A T Network Terminated by Its Image Impedances

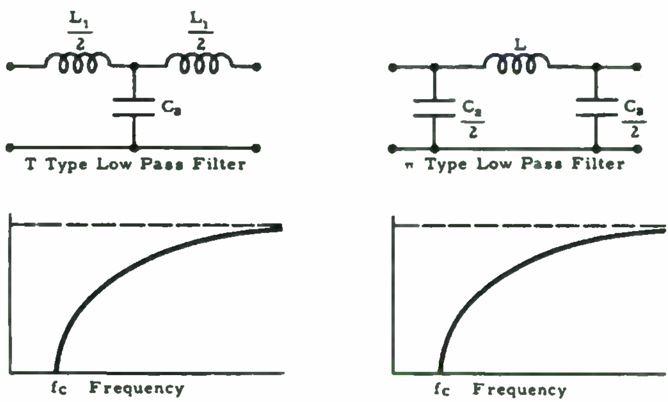
In terms of open-circuit and short-circuit measurement the image impedances Z_{I_1} and Z_{I_2} for both T and Pi networks are expressed as

$$Z_{I_1} = Z_{OC}\, Z_{SC} \qquad (1)$$

$$Z_{I_2} = Z'_{OC}\, Z'_{SC} \qquad (2)$$

where Z_{OC} and Z_{SC} are the impedances looking into terminals 1-2 with terminals 3-4 open-circuited and short-circuited respectively, and Z'_{OC} and Z'_{SC} are the impedance looking into terminals 3-4 with terminals 1-2 open-circuited and short-circuited, respectively. When the two image impedances are equal, $Z_a = Z_b$, the filter is symmetrical, and the impedances are equal to Z_o, the characteristic impedance of the network.

Figure I-4 gives the schematic diagram of two constant-k filters, the equation for their elements, and curves of attenuation plotted as a function of frequency. The filters are so-called because the product $Z_1 Z_2$



$$f_c = \frac{1}{\pi \sqrt{L_1 C_2}} \qquad f_c = \frac{1}{\pi \sqrt{L_1 C_2}}$$

$$R = \sqrt{\frac{L_1}{C_2}} \qquad R = \sqrt{\frac{L_1}{C_2}}$$

$$L_1 = \frac{R}{\pi f_c} \qquad L_1 = \frac{R}{\pi f_c}$$

$$C_2 = \frac{1}{\pi f_c R} \qquad C_2 = \frac{1}{\pi f_c R}$$

Figure I-4. Conventional Constant-k Filters is a constant for all frequencies and is equal to k^2. The value k is equal to the value R used in the table of Figure I-5.

f_c (cps)	L_1 (mh)	C_2 (μf)
30	5.31×10^2	2.12×10^2
100	1.59×10^2	6.37×10
150	1.06×10^2	4.24×10
200	7.96×10	3.18×10
250	6.37×10	2.55×10
300	5.31×10	2.12×10
350	4.55×10	1.82×10
400	3.98×10	1.59×10
450	3.54×10	1.41×10
500	3.18×10	1.27×10
550	2.89×10	1.16×10
600	2.65×10	1.06×10
650	2.45×10	9.79
700	2.27×10	9.09
750	2.12×10	8.49
800	1.99×10	7.96
850	1.87×10	7.49
900	1.77×10	7.07
950	1.68×10	6.70
1×10^3	1.59×10	6.37
3×10^3	5.31	2.12
10×10^3	1.59	6.37×10^{-1}
30×10^3	5.31×10^{-1}	2.12×10^{-1}
100×10^3	1.59×10^{-1}	6.37×10^{-2}
300×10^3	5.31×10^{-2}	2.12×10^{-2}
1×10^6	1.59×10^{-2}	6.37×10^{-3}
3×10^6	5.31×10^{-3}	2.12×10^{-3}
10×10^6	1.59×10^{-3}	6.37×10^{-4}
30×10^6	5.31×10^{-4}	2.12×10^{-4}

Figure I-5. Constant-k Low Pass Filter (R = 50Ω) - Table-1

The table given in Figure I-5 lists the element values for the full series and shunt arms of a low pass constant-k filter whose image impedances are equal to ohms. The following example illustrates the method by which values of L_1 and C_2 for cut-off frequencies not listed in the table may be obtained. A low pass filter working out of and into a 50-ohm line having a cut-off frequency of 150 cycles requires an inductance of 0.106 henries and a capacitance of 42.4 μf. For a cut-off frequency of 1.5 mc both of these values must be divided by 10^4, the ratio of the desired frequency to the listed frequency, because the product $L_1 C_2$ varies inversely with the cut-off frequency.

The expression $R = \sqrt{L_1/C_2}$ suggests a method by which data for a filter whose characteristic impedance is R, different from the value given in the table, can be obtained. First find the required data for a 50-ohm filter then multiply the inductance by R/50 and divide the capacitance by the same value.

Constant-k type filters act as a resistive load throughout the pass band if they are properly terminated in their image impedance. However, at the cut-off frequency the load becomes zero for the T network and infinite for the Pi network, as shown in Figure I-6. At frequencies beyond cut-off the load becomes imaginary--that is, in the attenuation band the filter acts as a reactive load, does not take energy from the interference source, and, therefore, does not transmit energy to the terminal impedance. However, infinite attenuation of the interference frequencies is obtained only by a filter with purely reactive arms which, of course, exists only in theory.

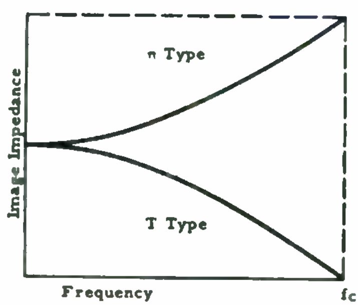


Figure I-6. Variation of Image Impedance with Frequency in Low Pass Constant-k T and Pi Type Filters

If a sharper cut-off (higher attenuation in the region beyond the cut-off frequency) than that exhibited by a constant-k type section is desired, it can be obtained by adding additional impedances to the prototype, the constant-k section. When the values of the added impedances are derived from those of the prototype, the resultant section is called an m-derived filter. These impedances sharpen the cut-off of the section by providing an attenuation which approaches infinity at a frequency beyond cut-off as shown in Figure I-7.

The position of the added impedance elements in the filter network determines the specific nomenclature of the section. If the addi-

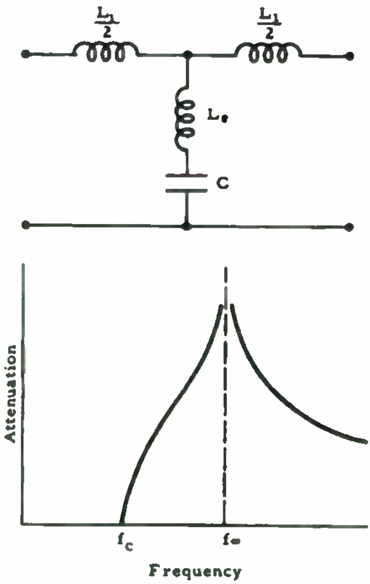
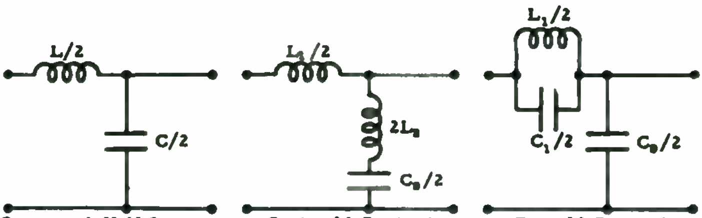
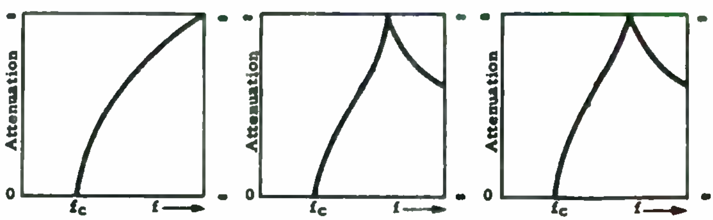


Figure I-7. Variation of Attenuation with Frequency in a Low Pass M-Derived T Type Filter



Constant-k Half Section	Series M-Derived Half Section	Shunt M-Derived Half Section
$L = \frac{R}{\pi f_c}$	$L_1 = mL$	$L_1 = mL$
$C = \frac{1}{\pi f_c R}$	$L_2 = \frac{1-m^2}{4m} L$	$C_1 = \frac{1-m^2}{4m} C$
$m = 1 - \frac{f_c}{f_\infty}^2$	$C_2 = mC$	$C_2 = mC$



Note: L in Henrys
C in Farads

Figure I-8. Constant-k and M-Derived Filter Sections tional impedances are added to the series arm of the section, the section is shunt derived. The section is series derived if additional impedances are added to the shunt arm. Schematic diagrams of series derived m-type low pass filter sections, the expressions used to obtain their component values from the basic data, their basic formulas, the expressions used to obtain their component values from k-values and their curve of attenuation plotted as a function of frequency are shown in Figure I-8. Constant-k values are designed by the subscript (k) in the expressions given in Figure I-8.

Sharpness of cut-off in the m-derived filter section is a function of m. To obtain sharp cut-off, the filter section should have a frequency of infinite attenuation, f_∞, close to the cut-off frequency, f_c. The expression

$$m = 1 - \frac{f_c^2}{f_\infty} \qquad (3)$$

$$f_\infty = \frac{f_c}{1 - m^2} \qquad (4)$$

which are valid for a low pass m-derived filter shows that as the ratio, f_c/f_∞ approaches unity the value of m approaches zero, or as m approaches unity the values f_∞ and f_c become more nearly equal.

Figure I-9 illustrates the variation of attenuation with the ratio of the cut-off frequency f_c to the interference frequency, f. The attenuation offered by m-derived filter sections for all values of m less than unity becomes infinitely high at some finite frequency, f_∞, and then decreases and approaches zero at higher frequencies. Although the sharpness of cut-off is more pronounced for the lower values of m, it is accompanied by correspondingly lower values of attenuation which approach zero at frequencies beyond the frequency of infinite attenuation. To compensate for this undesirable characteristic, a constant-k section, m = 1, is often used in conjunction with m-derived sections, but its attenuation increases constantly and approaches infinity as the frequency increases.

In order to join two or more filter sections for the purpose of obtaining higher attenuation throughout the attenuating band, their cut-off frequencies as well as their image impedances must be identical, but their frequencies of infinite attenuation may be unequal. M-derived T sections, each with a different value of m, make this possible because the image impedances of a T section at any frequency is the same regardless of the

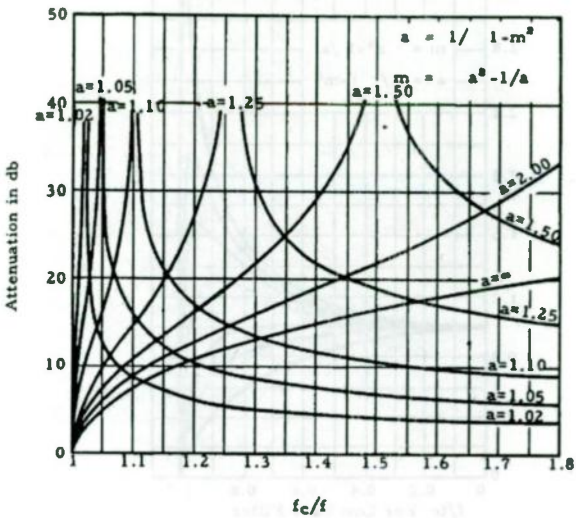


Figure I-9. Attenuation of M-Derived Low Pass Filters

value of m as shown in Figure I-10. Therefore, an impedance match is obtained between the T sections and reflections will not, of course, take place. However, it is difficult to terminate these sections properly due to the variation of image impedance with frequency, but by the use of terminating half Pi-sections it is possible to keep the image impedance of the filter constant at all frequencies up to approximately 90 percent of cut-off if the value of m selected is 0.6 (a = 1.25) as illustrated in Figure I-10. The proper termination can be accomplished by designing each section as a T network and then rearranging to form a Pi-section as shown in Figure I-11. Although the entire filter looks like 3 Pi-sections, the sections between the dotted lines have a T configuration. It is thus possible to alter the attenuating characteristics of a filter without varying its image impedances which are equal to the terminating impedance of the half Pi-section.

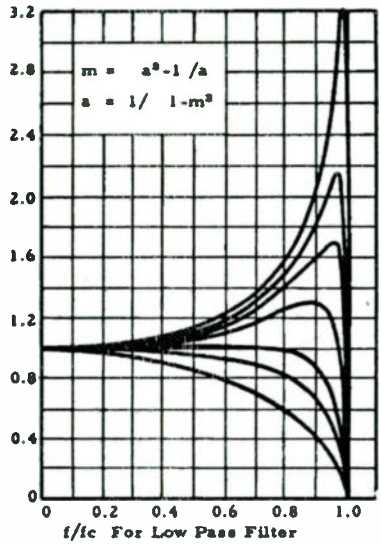


Figure I-10. Variation of Mid-Series and Mid-Shunt Image Impedance for Low Pass M-Derived Filter Sections

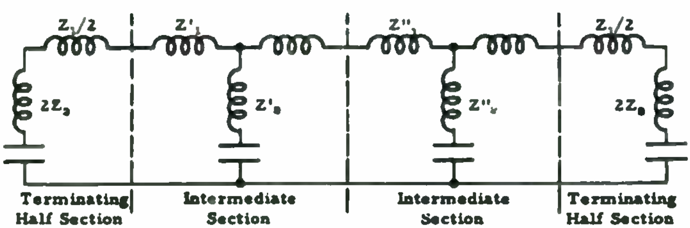


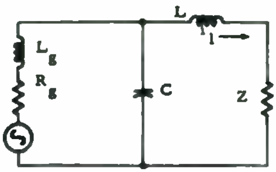
Figure I-11. Development of a Filter To Have a Desirable Impedance Variation

Figures I-12 through I-15 give in tabular form values of inductance and capacitance for m-derived low pass filters having m values of 0.1, 0.2, 0.4, and 0.6, respectively. The image impedance, R, of each filter is 50 ohms. If it is necessary to employ a value of R other than the value for which the filter components are listed, the component values given in the table must first be found. The listed inductance value or values must be multiplied by R/50, while the listed capacitance value or values must be divided by R/50 in order to obtain the desired values if the filter is to be designed at some other nominal impedance.

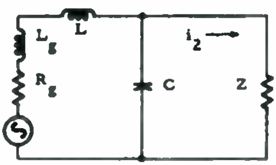
COMPARISON OF CONDENSER INPUT AND INDUCTANCE INPUT L-TYPE NETWORKS FOR USE IN DC MOTORS

APPENDIX II

It is required to determine the proper position of a condenser to be installed in a direct current motor with a series field. Assume that the interference source acts as a voltage generator with an inductive internal impedance. Consider the following two networks and their matrices.



$$\begin{bmatrix} R_g + j\left(\omega L_g - \frac{1}{\omega C}\right) & \frac{j}{\omega C} \\ \frac{j}{\omega C} & Z + j\left(\omega L - \frac{1}{\omega C}\right) \end{bmatrix}$$



$$\begin{bmatrix} R_g + j\left(\omega(L_g + L) - \frac{1}{\omega C}\right) & \frac{j}{\omega C} \\ \frac{j}{\omega C} & Z - \frac{j}{\omega C} \end{bmatrix}$$

R_g is the internal resistance of the generator, L_g is the inductive internal impedance of the generator, L is the inductance of the series field, C is the condenser to be installed in the motor, Z is the load impedance seen by the motor, which may be resistive, inductive, or capacitive. Hence, the ratio of the current is:

$$\frac{i_2}{i_1} = \frac{\left\{R_g + j\left(\omega L_g - \frac{1}{\omega C}\right)\right\}\left\{Z + j\left(\omega L - \frac{1}{\omega C}\right)\right\} + \left(\frac{1}{\omega C}\right)^2}{\left\{R_g + j\left(\omega(L_g + L) - \frac{1}{\omega C}\right)\right\}\left\{Z - \frac{j}{\omega C}\right\} + \left(\frac{1}{\omega C}\right)^2} \qquad (1)$$

Let $Z = R + jX$. Then the ratio reduces to:

$$\frac{i_2}{i_1} = \frac{A - \omega L_g + j(B + R_g)}{A - X + j(B + R)} \qquad (2)$$

where: $$A = \frac{R_g R}{\omega L} + \frac{1}{\omega C} + \frac{X}{\omega^2 LC} + \frac{L_g}{\omega LC} - \frac{L_g X}{L}$$

and $$B = \frac{R_g X}{\omega L} + \frac{L_g R}{L} - \frac{R_g + R}{\omega^2 LC}$$

Now the question is whether the ratio $\left| i_2/i_1 \right|$ is larger or smaller than unity. Consider the following cases:

(a) The load is resistive; $X = 0$. Then the absolute value of the numerator is larger than that of the denominator for high frequencies, say $\omega > 10^5$, since A decreases rapidly with frequency and the imaginary parts do not differ from one another by much for normal values of R_g and R. Hence, for the frequencies of interest,

$$\left| \frac{i_2}{i_1} \right| > 1$$

(b) The load is inductive; $X = \omega L_L > 0$. Then, for small A, the denominator increases as ωL_L and the numerator as ωL_g. L_g, the armature inductance of the motor, is normally about 1 to 5 henries. No load inductance as large as this is likely to be encountered. Hence it may be assumed that $L_L < L_g$, and again

$$\left| \frac{i_2}{i_1} \right| > 1$$

(c) The load is capacitive; $X = -(1/\omega C_L) < 0$. Then, for small A, the numerator increases as ωL_L and the denominator as $1/\omega C_L$. For normal values of C_L (10^{-11} to 10^{-7} farads) and the frequencies of interest, the numerator will again be larger than the denominator. Hence, again

$$\left| \frac{i_2}{i_1} \right| > 1$$

It is concluded that the condenser arrangement of the first circuit shown is always preferable except for those rare cases where the above assumptions do not hold.

DETAILS OF LAMINATE BRUSH DESIGN

APPENDIX III

The use of laminated brushes for reducing communication interference is explained in Volume I, Paragraph 7.1.1.1. The details for constructing laminated brushes are given in this appendix.

Either a mixture of copper and bakelite or a mixture of graphite and bakelite is satisfactory for preparing brush stocks of different resistivity. The materials selected are ground to powder consistency, passed through a 297 micron sieve to assure particles of uniform size, and mixed in a ball mill in various proportions to obtain various degrees of resistivity. Figures III-1 and 2 are curves giving the required mixing data necessary to predetermine resistivity with respect to percentage of mixture. The mixture is then placed in molds in a hydraulic press and baked for one hour at a temperature of 190° C. under a pressure of 2500 pounds per square inch. Laminations of different thicknesses are cut from the brush stock after removal from the mold. The laminations selected to fabricate a brush are coated with six mil leaf glue (No. R612 Minnesota Mining and Manufacturing Company is suitable) and are baked for one hour at a temperature of 180° under a pressure of 500 pounds per square inch. To maintain a constant temperature during the baking process, the oven is covered with a glass cloth and a variac is used to control the heat lamp employed. The brush thus formed is sanded to its exact dimensions, painted on both sides with G. E. 7031 adhesive, and baked at a moderate temperature to dry the adhesive. Cables are now tamped into the brush top, and to insure electrical contact across one end of the laminations, the brush top is subjected to a copper spray.

The techniques described above have been found satisfactory for constructing laminated brushes. Previously it was impossible to construct brushes having satisfactory radio interference characteristics because of the intermingling of the conducting and insulating materials. This was caused by the formation of minute pockets in the insulation between laminations which become filled with the conducting dust of the laminations due to brush wear. The gluing process described above and the use of a superior glue which can be subjected to relatively high pressures with a minimum of extrusion overcame this difficulty. Furthermore, brushes constructed as described possess satisfactory mechanical characteristics.

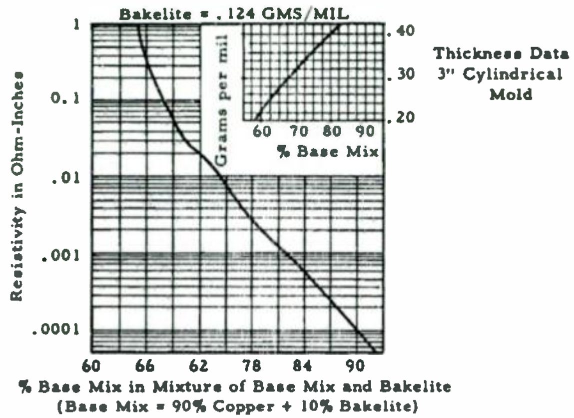


Figure III-1. Mixing Data for Copper Base Material

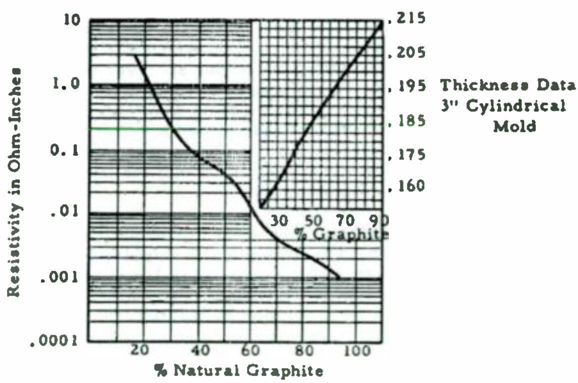


Figure III-2. Mixing Data for Graphite Bakelite

THE CONSTRUCTION OF BONDING JUMPERS FOR SHOCK MOUNTED EQUIPMENT

APPENDIX IV

An investigation into the best type of bonding jumper for shock-mounted equipment has led to the development of a new design, the essential features of which are described in this appendix.

The requirements for a good bonding jumper are the following:

(a) Low direct current resistance

(b) Low radio frequency impedance at all frequencies.

(c) Good mechanical properties as follows:

1. Ability to withstand ambient conditions and endurance requirements

2. Minimum height so that the overall mounting height of the shockmount is not increased

3. Absence of, or adequate guards against, all sharp edges for personnel protection

4. Minimum weight

5. Minimum volume

6. Minimum stiffness so that the vibration-isolation characteristics of the shockmount are not adversely affected.

Consideration of all these requirements has led to the development of the types illustrated in Figure IV-1. The most important feature is the employment of two metallic strips whose width is large compared to their thickness.

These jumpers consist of three major parts: the two bonding strips, the base plate, and the spacer washers. The strips are displaced 90° in position and connected externally between the top and bottom of the shockmount. Four strips could be used, but the addition of two more strips decreases the impedance by only ten percent and adds considerably to the damping. Therefore, it is not recommended.

Figure IV-1. Capacitive Types Incorporated in Shockmount

A possible alloy to be used in the construction of the strips is Ni-Span "C" whose inherent properties are those of high modulus of elasticity, high tensile strength, and a high degree of hardness.

To limit the stress on the bonding strip its length should be made as great as possible while its thickness should be kept to a minimum. However, the electrical requirements of the bonding connector limit the length of the strips while ease in manufacture and fragility of relatively thin strips limit their thickness. For shockmounts whose load capacities are one pound or less, a thickness of 0.002 of an inch is recommended; for mounts whose load capacities exceed one pound, a thickness of 0.003 of an inch should be used. A good compromise for the other dimensions is a length of from 1 to 3 inches and a width of from 3/4 inches to 1 inch.

The sharp edges of the thin leaves are safety-shielded with glass fibre sleeving, a material that will withstand the extreme ambient conditions expected, for the protection of personnel. The sleeving is coated with Dow-Corning No. 801 cement, and partially cured in an oven at approximately 300° F. Following this the strip is cut into sections of desired length, slipped over the bonding leaves, re-cemented, and cured completely.

The baseplate to which the bonding strips are welded is made of cadmium-plated SAE 1030 sheet steel. Where possible, the base is designed for top mounting because this design does not add appreciably to the overall mounting height of the shockmount.

Both spacer washers are made of cadmium-plated cold-rolled steel. The thickness of the upper spacer washer is sufficient to prevent contact between the bonding leaves and the plate which is to be shockmounted. Because of the possibility of excessive vibration the bottom spacer washer is made sufficiently large to function as a snubber.

Torsional stresses resulting from sideway motion of the top of the mount with respect to the base will be lessened by connecting the tops of the bonding leaves to the spacer washers by a loosely stacked joint. This joint lowers the calculated stresses (developed on the basis of a doubly clamped beam), and furthermore allows the rotation of the leaves plus slight lateral movements.

The pin type joint at the top of the mount, designed with as thin sections as practical to reduce overall height, also reduces torsional stresses in the leaves. This joint is so designed that a positive electrical contact is assured, yet allows rotational movement without binding.

The direct current resistance of bonds may vary from 0.012 to 0.059 ohms depending upon the length of the jumper. The resistance is not seriously affected by the mutual movements of the strips at the top of the shockmount. The impedance, at 1000 megacycles, may vary from 13.5 to 14.8 ohms and decreased with compression of the mount. The resonant frequency of all these jumpers was in the vicinity of 3500 mc.

TEST EQUIPMENT CALIBRATION PROCEDURES

APPENDIX V

1. INTRODUCTION

The Military Agencies require that RFI data taken by measurement equipment be accurate to very close tolerances. The basic measurement quantities are frequency, power, voltage, and impedance. The most stringent tolerance required for each of these quantities is:

a. Frequency accuracy of one part in 1000 or better.

b. Power output accuracy of signal generators within 2db at any attenuator setting.

c. Frequency-selective voltmeter accuracy within 2 db of the indicated value.

d. Impedance of signal generator output shall be unbalanced 50 ohms, resistive, with a VSWR not to exceed 1.3:1. The calibration of attenuators shall be known within 1 db at each measuring frequency.

These values are to be maintained throughout the measurements program by reference to laboratory standards.

To meet these requirements for measurement accuracy, it will be necessary to establish two calibration programs, one for maintenance of the accuracy of the laboratory standards and the other for the maintenance of the accuracy of the measurement equipment. The laboratory standards must be periodically compared with standards which have calibration accuracies traceable to the National Bureau of Standards. These then become the local laboratory standards for maintenance of the accuracy of the measurement equipment. Before any measurement task is undertaken, it will be required to establish the accuracy of what will be considered the laboratory standards of frequency, power, voltage, and impedance. A schedule of equipment maintenance and calibration must be established to prepare a set of measurement equipments for each measurement task assigned.

A definitive equipment calibration schedule will have to be established in order to keep all instrumentation up to peak performance. Calibration checks of laboratory standards and measurement equipments should be scheduled in accordance with the following outline:

Schedule of Calibration Checks of Laboratory Standards.

Frequency Meter	every 6 months
Power Meter	every 3 months
Voltage Meter	every 3 months
Attenuators; Impedance Measurement Equipment	every 6 months

Schedule of Calibration Checks of Measurement Equipment.

AC Voltmeter	every 3 months
DC Voltmeter	every 3 months
Frequency Meters	every 6 months
Oscilloscopes	every 3 months
Spectrum Analyzers	every 3 months
Signal Generators	every 6 months
Frequency Selective Voltmeters (Field Intensity Meters)	every 3 months
Multifunction Test and Support Equipment	every 6 months

It should be understood that these schedules will have to be modified to accommodate intermediate checks of individual equipments which, by their performance on the job, show that maintenance and calibration are necessary before their scheduled check.

A file must be maintained to account for all maintenance performed, calibration schedules, records of calibration data, and the results of calibration checks. A copy of the latest calibration record in effect should always be kept with the equipment.

2. CALIBRATION PROCEDURES SIGNAL GENERATOR MODEL HP-608C

FRC

2. CALIBRATION PROCEDURES FOR RF SIGNAL GENERATOR HP-608C

2.1 CALIBRATION CHECKS REQUIRED

a. RF OUTPUT accuracy at zero dbm throughout the frequency range.

b. ATTENUATOR dial linearity at selected frequencies.

c. FREQUENCY dial tracking throughout the frequency range.

d. PERCENT MODULATION meter accuracy from 0 to 100%.

2.2 EQUIPMENT REQUIRED OR EQUIVALENT

Power Meter, HP-430C

Thermistor Mount, HP-477B

Frequency Meter, Lavoie LA-70A

Attenuators, Weinschel 50 Series

Receiver, Stoddart NM-30A

Cables as required

2.3 PRELIMINARY OPERATION

Connect all equipment to a suitable a-c power source, energize equipment and allow a warm-up time of 30 minutes.

2.4 OUTPUT POWER MEASUREMENTS

a. Setup of operating controls for HP-430C

COEF switch	NEG
RES switch	200
POWER RANGE switch	+ 5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

b. Connect equipment as shown in Figure 2-1.

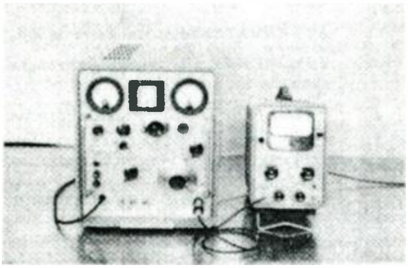


Figure 2-1. Power Measurement Test Configuration

CAUTION

The power meter BIAS CURRENT switch must be set to OFF before connecting or disconnecting the thermistor mount.

c. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

d. Adjust the HP-430C ZERO SET controls for a zero indication on the top meter scale.

e. Set the HP-608C controls as follows:

MOD SELECTOR switch	CW
MOD LEVEL control	Maximum counter-clockwise
OUTPUT LEVEL control	SET LEVEL indication on the OUTPUT VOLTS meter
RF OUTPUT control	for 0 dbm output

f. Set the HP-608C MOD SELECTOR switch to PULSE and re-zero HP-430C.

g. Verify the SET LEVEL indication on the HP-608C OUTPUT VOLTS meter and a zero indication on the HP-430C by switching the MOD SELECTOR switch from CW to PULSE.

h. Switch the MOD SELECTOR switch to CW and record the variation from zero dbm as indicated on the HP-430C.

i. Repeat Steps (e) through (h) for each frequency to be calibrated.

j. Verify that the indications recorded in Step (h) are within tolerance.

k. Plot a graph of output deviation from zero dbm as a function of frequency.

l. Turn HP-430C ZERO SET controls maximum counter-clockwise. Set the HP-430C BIAS CURRENT switch to OFF and disconnect the HP-477B from HP-608C. Turn HP-430C LINE POWER switch to OFF.

2.5 OUTPUT ATTENUATOR DIAL LINEARITY

a. Set HP-608C controls as follows:

FREQUENCY RANGE switch	D
FREQUENCY tuning control	200 mc
MOD SELECTOR switch	CW
RF OUTPUT control	0 dbm
OUTPUT LEVEL control	SET LEVEL indication on the OUTPUT VOLTS meter
MOD LEVEL control	Maximum counter-clockwise

b. Set NM-30A controls as follows:

BAND SWITCH	BAND 5
Frequency TUNING control	200 mc
Function selector switch	FIELD INTENSITY
CALIBRATE control D adjust for	2 db meter indication
Attenuator	X10

c. Connect a 50-5 attenuator between the RF OUTPUT of the HP-608C and the RF INPUT of the NM-30A as shown in Figure 2-2.

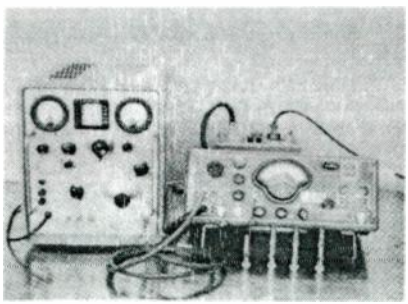


Figure 2-2. Tracking Error Measurement Configuration

d. Tune NM-30A for maximum indication of signal output of HP-608C.

e. Adjust HP-608C OUTPUT LEVEL control for a SET LEVEL indication on OUTPUT VOLTS meter.

f. Set NM-30A meter to read 12 db by means of CALIBRATE control.

g. Remove 50-5 attenuator from RF OUTPUT of HP-608C and connect RF OUTPUT of HP-608C to NM-30 A RF INPUT.

h. Reduce HP-608C output until the NM-30A indicates the reference level of 12 db.

i. Record HP-608C output ATTEN indication.

j. Repeat Steps (c) through (i) substituting appropriate standard attenuators until the HP-608C ATTEN has been calibrated from 5 to 90 db in 5 db steps.

k. Disconnect attenuator and NM-30A from RF OUTPUT of HP-608C.

2.6 FREQUENCY DIAL TRACKING ERROR

a. Set HP-608C panel controls as follows:

FREQUENCY RANGE switch	A
Frequency tuning control	10 mc
ATTEN dial	0 dbm
MOD SELECTOR switch	CW
MOD LEVEL control	Maximum counter-clockwise
AMP TRIMMER control	Maximum output
OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter

b. Set controls on front panel of LA-70A as follows:

SEARCH-MC LOCK-LOCK switch	SEARCH
MC RANGE switch	0.01 to 30
DIAL C	0
DIAL B	0

c. Connect a 50-5 attenuator between the RF OUTPUT of the HP-608C and the 10 kc to 1000 mc Input of the LA-70A, as shown in Figure 2-3.

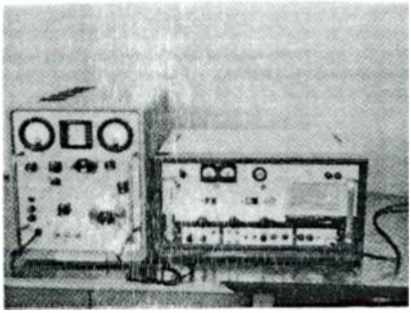


Figure 2-3. Frequency Dial Tracking Error
Measurement Configuration

d. Tune in the output signal of the HP-608C on the LA-70A by searching for a beat while tuning DIAL A in the vicinity of 10 mc.

e. On locating the beat note, turn function selector switch to MC LOCK and turn DIAL B until meter A deflects and the beat note disappears.

f. Regain beat note by rocking DIAL B back and forth and verify that the MC stage is still locked.

g. Turn function selector switch to LOCK and adjust DIAL B to the first lock point lower in frequency than the beat note frequency and regain the beat note by tuning DIAL C.

h. Read and record the frequency as indicated by DIALS A, B, and C. Verify that the frequency is within a tolerance of 1%. If error is greater than 1%, refer to instruction manual for alignment. Alternate solution is to make a correction chart for the errors found.

i. For frequencies above the direct reading frequency of the LA-70, the indications of DIALS A, B, and C are multiplied by the harmonic used.

j. Repeat Steps (d) through (h) at selected frequencies across FREQUENCY RANGE bands A, B, C, D, and E of HP-608C.

k. Disconnect equipment from RF OUTPUT of HP-608C.

2.7 MODULATION METER ACCURACY

a. Set HP-608C controls as follows:

FREQUENCY RANGE switch	C
Frequency tuning control	80 mc
MOD SELECTOR switch	1000
AMP TRIMMER control	Adjust for maximum output
OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter
MOD LEVEL control	0%
ATTEN dial	0 dbm

b. Connect HP-608C RF OUTPUT to NM-30A as shown in Figure 2-4 and tune in 80 mc signal.

c. Set NM-30A function selector switch to QUASI PEAK and set a reference level of 20 μvolts by varying CALIBRATE control.

d. Increase MOD LEVEL control for a 100% indication on the MODULATION meter.

e. Readjust OUTPUT VOLTS meter for a SET LEVEL indication.

f. The NM-30A should indicate a level of 40 μvolts.

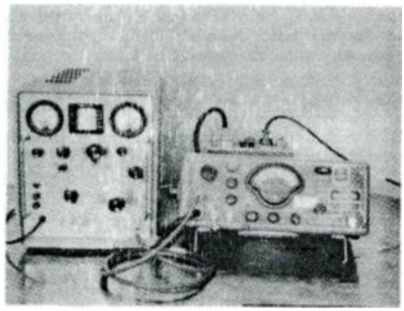


Figure 2-4. Modulation Meter Accuracy Test Configuration

g. Check PERCENT MODULATION meter calibration for 30%, 50%, 75%, and 100% modulation. A 30% modulation of the HP-608C output should give a reading on the NM-30A of 1.3 times the 0% modulation reference level established in Step (c); a 50% modulation should give 1.5 times the reference level of Step (c); and so forth. Modulation accuracy shall be within 5%. If error is greater than 5%, refer to instruction manual.

3. CALIBRATION PROCEDURES SIGNAL GENERATOR MODEL HP-612A

FRC

3. CALIBRATION PROCEDURES FOR RF SIGNAL GENERATOR HP-612A

3.1 CALIBRATION CHECKS REQUIRED

a. Power output accuracy at zero dbm throughout the frequency range.

b. Attenuator dial linearity at selected frequencies.

c. Frequency dial tracking throughout the frequency range.

d. Percent modulation meter accuracy from 0 to 100%.

3.2 EQUIPMENT REQUIRED OR EQUIVALENT

Power Meter, HP-430C

Thermistor Mount, HP-477B

Frequency Meter, Lavoie LA-70A

Attenuators, Weinschel 50 Series

Receiver, Stoddart NM-50A

Cables as required

3.3 PRELIMINARY OPERATION

Connect all equipment to a suitable a-c power source, energize equipment and allow a warm-up time of 30 minutes.

3.4 OUTPUT POWER MEASUREMENT

a. Setup of operating controls for HP-430C

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

b. Connect equipment as shown in Figure 3-1.

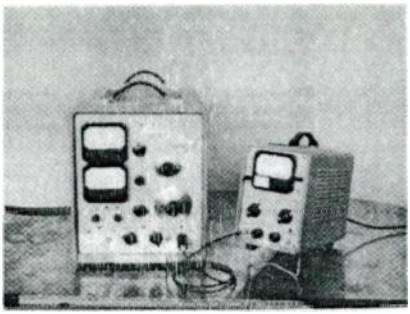


Figure 3-1. Power Measurement Test Configuration

CAUTION

The power meter BIAS CURRENT switch must be set to OFF before connecting or disconnecting the thermistor mount.

c. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

d. Adjust the HP-430C ZERO SET controls for a zero indication on the top meter scale.

e. Set the HP-612A controls as follows:

Modulation selector switch	CW
MOD LEVEL control	Maximum counter-clockwise

OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter
ATTEN control	0 dbm

f. Set HP-612A modulation selector switch to PULSE and re-zero HP-430C.

g. Verify a SET LEVEL indication on HP-612A and a zero indication on HP-430C by switching modulation selector switch from CW to PULSE.

h. Switch the modulation selector switch to CW and record the variation from zero dbm as indicated on HP-430C.

i. Repeat Steps (e) through (h) for each frequency to be calibrated.

j. Verify that the indications recorded in Step (h) are within tolerance.

k. Plot a graph of output deviation from zero dbm as a function of frequency.

l. Turn HP-430C ZERO SET control maximum counter-clockwise. Set the HP-430C BIAS CURRENT switch to OFF and disconnect the HP-477B from HP-612A. Turn HP-430C LINE POWER switch to OFF.

3.5 ATTENUATOR DIAL LINEARITY

a. Set HP-612A front panel controls as follows:

MEGACYCLES tuning control	1000
Function selector switch	CW
Modulation selector switch	NORMAL
ATTEN control	0 dbm
OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter

b. Set NM-50A front panel controls as follows:

TUNING AID-PULSE STRETCHER switch	OFF
Function switch	FI

TUNE control	1000
CAL control	2 db meter indication
Attenuator	$X10^4$

c. Connect a 50-5 attenuator between the OUTPUT of the HP-612A and the RF INPUT of the NM-50A as shown in Figure 3-2.

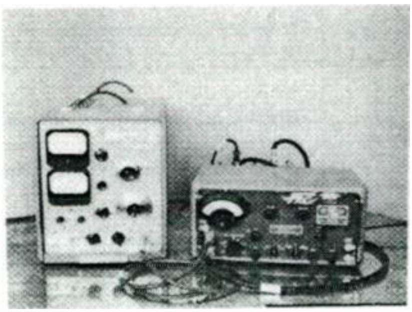


Figure 3-2. Attenuator Tracking Error Measurement Configuration

d. Tune in output signal of HP-612A on NM-50A. Peak RF TRIM and MIXER TRIM controls.

e. Set the NM-50A meter needle for a reference of 25 db by means of CAL control.

f. Verify a SET LEVEL indication on HP-612A OUTPUT VOLTS meter.

g. Remove 50-5 attenuator from OUTPUT of HP-612A and connect OUTPUT of HP-612A directly to NM-50A RF INPUT.

h. Decrease HP-612A ATTEN until the reference level of 25 originally set in Step (e) is indicated on NM-50A meter.

i. Record HP-612A ATTEN indication.

j. Adding an appropriate standard attenuator in 5 db steps, repeat Steps (e) through (i) until the HP-612A ATTEN control has been calibrated from 5 to 90 db in 5 db steps.

k. Disconnect attenuator and receiver from OUTPUT of the HP-612A.

3.6 FREQUENCY DIAL TRACKING ERROR

a. Set HP-612A front panel controls as follows:

Frequency tuning control	450 mc
ATTEN control	0 dbm
Modulation selector switch	CW
OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter

b. Set controls on front panel of LA-70A as follows:

SEARCH - MC LOCK - LOCK switch	SEARCH
MC RANGE switch	42 - 60
DIAL C	0
DIAL B	0

c. Connect a 50-5 attenuator between the OUTPUT of HP-612A and the 10 kc - 1000 mc INPUT of the LA-70A as shown in Figure 3-3.

d. Tune in the output signal of the HP-612A on the LA-70A by searching for a beat while tuning DIAL A in the vicinity of 10 mc.

e. On locating the beat note, turn function selector switch of LA-70A to MC LOCK and turn DIAL B until meter A deflects and the beat note disappears.

f. Regain beat note by rocking DIAL B back and forth and verify that the MC stage is still locked.

g. Turn function selector switch to LOCK and adjust DIAL B to the first lock point lower in frequency than the beat note frequency and regain the beat note by tuning DIAL C.

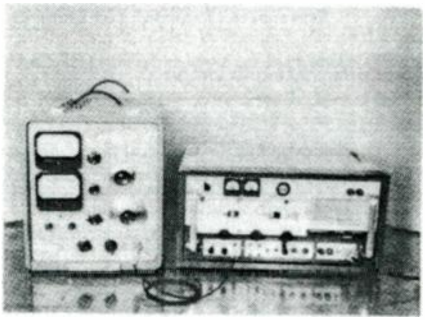


Figure 3-3. Dial Tracking Error
Measurement Configuration

h. Read and record the frequency as indicated by DIALS A, B, and C. Verify that the frequency is within a tolerance of 1%. If error is greater than 1%, refer to instruction manual for alignment. Alternate solution is to make a correction chart for the errors found.

i. For frequencies above the direct reading frequency of the LA-70A, the indications of DIALS A, B, and C are multiplied by the harmonic used.

j. Repeat above Steps (d) through (h) at selected frequencies across tuning bands A, B, C, D, and E of HP-612A.

k. Disconnect equipment from OUTPUT of HP-612A.

3.7 MODULATION METER ACCURACY

a. Adjust front panel controls of HP-612A as follows:

Frequency tuning control	850 mc
MOD LEVEL control	Maximum counter-clockwise
OUTPUT LEVEL control	SET LEVEL indication on OUTPUT VOLTS meter
ATTEN control	-100 dbm
Function selector switch	1000 cps

b. Adjust front panel controls of HP-430C as follows:

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

c. Connect the equipment as shown in Figure 3-4.

Figure 3-4. Modulation Meter Accuracy Test Configuration

d. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

e. Adjust the HP-430C ZERO SET controls for a zero indication on the top meter scale.

f. Verify a SET LEVEL indication on OUTPUT VOLTS meter of HP-612A.

g. Rotate the ATTEN control of HP-612A until HP-430C reads 1 mw on the 3 mw range.

h. Rotate MOD LEVEL control of HP-612A CW until HP-430C reads 1.5 mw. This represents 100% modulation. Repeat for 40% and 80% modulation (1.2 mw and 1.4 mw).

i. Turn BIAS CURRENT selector switch of HP-430C to OFF.

j. Disconnect calibration setup.

4. CALIBRATION PROCEDURES

RF SIGNAL GENERATOR MODEL GR-1001A

ERC

FRC

4. CALIBRATION PROCEDURES FOR RF SIGNAL GENERATOR GR-1001A

4.1 CALIBRATION CHECKS REQUIRED

a. Voltage output calibrated at selected levels and frequencies.

b. Frequency dial tracking throughout the frequency range.

c. Percent modulation meter tracking from 0 to 100%.

4.2 EQUIPMENT REQUIRED OR EQUIVALENT

Frequency Meter, Lavoie LA-70A

Voltmeter, HP-400H

Oscilloscope, Tektronix 545A

Cables and connectors as required

4.3 PRELIMINARY OPERATION

Connect all equipment to a suitable a-c power source, energize equipment and allow a warm-up time of 30 minutes.

4.4 OUTPUT ATTENUATION AND MULTIPLIER TRACKING MEASUREMENTS

a. Set the controls of the GR-1001A as follows:

FREQ RANGE switch	150-500 kc
Tuning control	250 kc
METER switch	CARRIER
OUTPUT switch	ATTEN <400 kc
MULTIPLIER switch	100 millivolts

b. Connect the HP-400H INPUT terminals to the GR-1001A ATTEN output terminal as shown in Figure 4-1. Ensure that circuit ground terminals are interconnected.

c. Measure the output of the GR-1001A in accordance with the following chart; record indications.

MULTIPLIER	2.0	1.8	1.6	1.4	1.2	1.0	0.8	0.6	0.4	0.2	MAX Error (db)
100 mv											
10 mv											
1 mv											
100 µv											

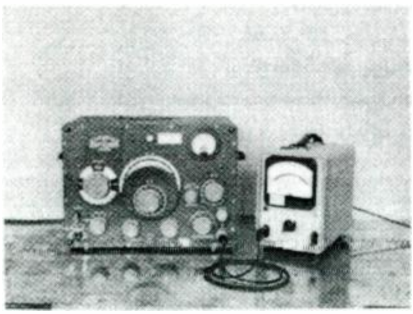


Figure 4-1. Output Attenuation and Multiplier Tracking Measurement Configuration

d. Verify that each voltage indication is within tolerance of 2 db. Refer to instruction manual for repair, alignment, or adjustment if required to achieve 2 db tolerance.

e. Disconnect calibration setup.

4.5 FREQUENCY DIAL TRACKING ERROR

a. Set the controls of GR-1001A as follows:

FREQ RANGE switch	5-15 mc
OUTPUT selector switch	ATTEN

METER selector switch	CARRIER
CARRIER control	SET CARRIER
MODULATION switch	OFF
Tuning control	10 mc

b. Set the controls of LA-70A as follows:

SEARCH - MC LOCK - LOCK switch	SEARCH
MC RANGE switch	0.01 - 30
DIAL C	0
DIAL B	0

c. Connect a 50-5 attenuator to the ATTEN output of the GR-1001A. Connect the 10 kc - 1000 mc INPUT/OUTPUT terminal of the the LA-70A to the 50-5 attenuator as shown in Figure 4-2.

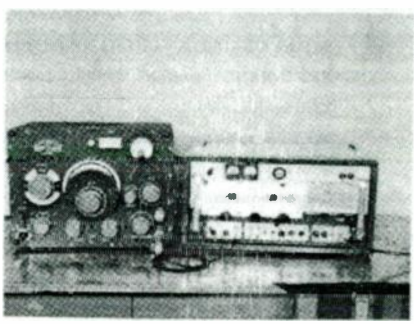


Figure 4-2. Dial Tracking Error Measurement Configuration

d. Tune in the output signal of the GR-1001A on the LA-70A by searching for a beat while tuning DIAL A in the vicinity of 10 mc.

e. On locating the beat note, turn function selector switch of LA-70A to MC LOCK and turn DIAL B until meter A deflects and the beat note disappears.

f. Regain beat note by rocking DIAL B back and forth and verify that the MC stage is still locked.

g. Turn function selector switch to LOCK and adjust DIAL B to the first lock point lower in frequency than the beat note frequency and regain the beat note by tuning DIAL C.

h. Read and record the frequency as indicated by DIALS A, B, and C. Verify that the frequency is within a tolerance of 1%. If error is greater than 1%, refer to instruction manual for alignment. Alternate solution is to make a correction chart for the errors found.

i. Repeat Steps (d) through (h) at selected frequencies over the frequency range of GR-1001A.

j. Disconnect equipment from output of GR-1001A.

4.6 MODULATION METER ACCURACY

a. Set the GR-1001A controls as follows:

FREQ RANGE switch	150 - 500 kc
Tuning control	200 kc
METER selector switch	CARRIER
OUTPUT selector switch	ATTEN
Output control	2

b. Set the 545A Oscilloscope controls as follows:

Vertical sensitivity switch	0. 1v/cm
Vertical sensitivity vernier control	CALIBRATED
VARIABLE TIME/CM switch	1 millisecond
VARIABLE TIME/CM vernier control	CALIBRATED
TRIGGERING MODE TRIGGER SLOPE switch	INT +
Sync switch	AUTO

HORIZONTAL DISPLAY switch	A
POLARITY switch	NORMAL
MAGNIFIER switch	OFF

c. Connect the GR-1001A ATTEN output terminal to the 545A CHANNEL A input terminal as shown in Figure 4-3.

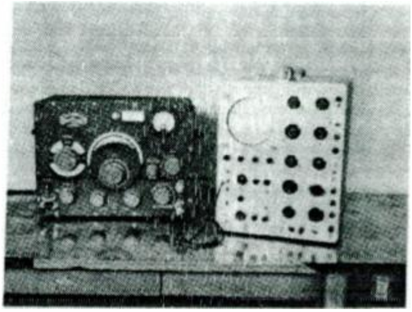


Figure 4-3. Modulation Meter Accuracy Measurement Configuration

d. Adjust the GR-1001A CARRIER control for a SET LEVEL indication on the % MODULATION meter.

e. Adjust the GR-1001A output dial to obtain 4 cm of presentation on the oscilloscope.

f. Set the GR-1001A METER switch to MODULATION and the MODULATION switch to 400 cps.

g. Adjust the GR-1001A MODULATION control for an indication of 70 on the % MODULATION meter.

h. Set the GR-1001A METER switch to CARRIER and verify that the % MODULATION meter is indicating SET LEVEL. Set the METER switch to MODULATION.

i. Adjust the 545A STABILITY TRIGGER LEVEL control for a stable condition.

j. Adjust the 545A VERTICAL POSITION and HORIZONTAL POSITION controls to center the presentation.

k. Record the vertical deflection from center of the troughs and peaks of the presentation as in Figure 4-4.

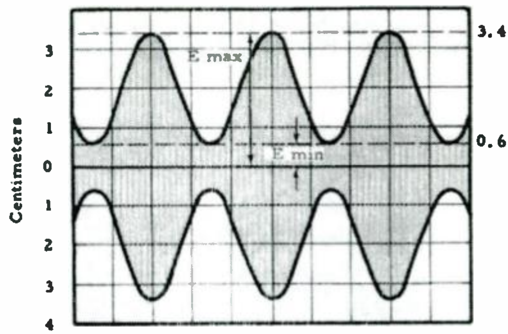


Figure 4-4. Percent Modulation Presentation on Oscilloscope

l. Record the percent modulation as derived from the following formula:

$$\% \text{ MOD} = \frac{E_{max} - E_{min}}{E_{max} + E_{min}} \text{ X } 100$$

at 70% MOD: $E_{max} = 3.4$ cm

$E_{min} = 0.6$ cm

$$\% \text{ MOD M} = \frac{3.4 - 0.6}{3.4 + 0.6} \text{ X } 100 = 70\%$$

m. Measure and record as in Steps (h) through (l) at each of the remaining GR-1001A % MODULATION meter major scale divisions.

5. CALIBRATION PROCEDURES
MICROWAVE SIGNAL GENERATOR
POLARAD MODEL MSG-1

5. MICROWAVE SIGNAL GENERATOR POLARAD MODEL MSG-1

5.1 CALIBRATION CHECKS REQUIRED

a. The calibration of the FREQUENCY dial is verified at selected frequencies by using LA-70A as a standard.

b. The power output is checked at zero dbm by monitoring the power with a power meter and measuring any variation from this level resulting from a change in frequency.

c. The OUTPUT ATTENUATOR dial is calibrated to determine its linearity at selected output frequencies. With zero dbm a known attenuator is inserted in the line between the signal generator and receiver. With the attenuator removed, the output of the MSG-1 OUTPUT ATTENUATOR dial is varied for a comparable receiver indication.

5.2 EQUIPMENT REQUIRED OR EQUIVALENT

Power Meter, HP-430C

Thermistor Mount, HP-477B

Frequency Meter, Lavoie LA-70A

Transfer Oscillator, HP-540B

Attenuators, Weinschel 50 Series

Receiver, Stoddart NM-50A

Interconnecting cables as required

5.3 PRELIMINARY OPERATION

a. Verify that all POWER switches are OFF.

b. Connect the power cable of MSG-1 to a source of 115 volts ac, 50-60 cycle.

c. Turn the POWER ON-OFF switch of MSG-1 to ON.

d. Set the HP-430C BIAS CURRENT switch to OFF.

e. Connect power cable of HP-430C to a source of 115/230 volts ± 10%, 50/1000 cycles.

f. Turn LINE POWER switch of HP-430C to ON.

g. Connect power cable of LA-70A to a source of 115 volts ac, 50-420 cps.

h. Turn the POWER ON-OFF switch of LA-70A to ON.

i. Connect the NM-50A power cable (91947-1) between J-301 on power supply 90399-4 and POWER receptacle on NM-50A.

j. Connect the NM-50A a-c power cable (91258-1) to J-302 on power supply 90399-4.

k. Connect power cable (91258-1) to a source of 105-125 volts ac, 50-1600 cps, single phase.

l. Turn power switch on NM-50A power supply to ON

m. Allow equipment a 30-minute warm-up time.

5.4 FREQUENCY DIAL TRACKING ERROR

a. Set MSG-1 front panel controls as follows:

RF OSC	1.0 kmc
MODULATION SELECTOR switch	CW
OUTPUT ATTENUATOR control	0 dbm
FM AMPLITUDE control	Maximum counter-clockwise

b. Connect an attenuator 50-5 to RF OUTPUT of MSG-1.

c. Connect RF OUTPUT of MSG-1 to OSCILLATOR INPUT of HP-540B by means of a 4-foot length of RG-9/U as shown in Figure 5-1.

d. Set LA-70A front panel controls as follows:

SEARCH-MC LOCK-LOCK switch	SEARCH
MC RANGE switch	42-60
DIAL C	0
DIAL B	0

e. Connect (by means of a 3-foot length of RG-9/U) between 10 kc to 1000 mc input terminal of LA-70A to FREQUENCY METER connector on HP-540B.

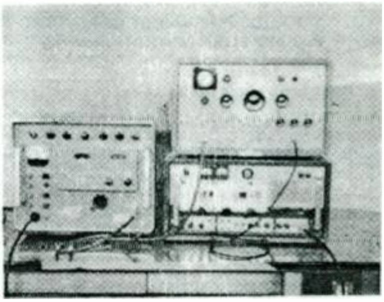


Figure 5-1. Frequency Dial Tracking Test Configuration

f. Turn MODULATION SELECTOR switch on MSG-1 to ZERO SET and turn ZERO SET control for a ZERO indication on the output meter.

g. Turn MODULATION SELECTOR switch on MSG-1 to CW, adjust POWER SET control for red line indication on output meter.

h. Tune the COARSE VERNIER and then FINE VERNIER of HP-540B until the highest harmonic zero beat of MSG-1 is indicated on the scope of HP-540B. Repeat the above step for the next lowest zero beat. The harmonic can be found by use of the following formula:

$$N = \frac{F_2}{(F_2 - F_1)}$$

where:

N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lowest zero beat

i. Search for the highest fundamental output of HP-540B by rotating DIAL A of LA-70A.

j. On locating beat with DIAL A of LA-70A, turn function selector switch to MC LOCK.

k. Turn DIAL B on LA-70A until meter A deflects (beat disappears).

l. Regain zero beat by tuning DIAL B of LA-70A back and forth. Verify that MC stage is still locked.

m. Turn function switch of LA-70A to LOCK and adjust DIAL B to first lock point lower in frequency from beat note.

n. Regain zero beat with DIAL C of LA-70A.

o. Read frequency as indicated by DIALS A, B, and C.

p. Multiply the indicated frequency by the harmonics used.

q. Harmonic number for LA-70A can be found by using the following formula:

$$N = \frac{F_2}{(F_2 - F_1)}$$

where: N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lowest zero beat

r. The output frequency of MSG-1 is a product of the harmonic of the HP-540B times the harmonic of LA-70A times the frequency as indicated by DIALS A, B, and C of LA-70A.

s. Verify that frequency is within tolerance (1%).

t. Repeat above steps at selected output frequencies over the frequency range of MSG-1.

u. Disconnect equipment from RF OUTPUT of MSG-1.

5.5 OUTPUT TRACKING ERROR AT ZERO DBM

a. Connect the power cable of MSG-1 to a source of 115 volts ac, 50 - 60 cps.

b. Connect the power cable of HP-430C to a source of 115/230 volts ± 1%, 50/1000 cycles.

c. Connect HP-477B to RF OUTPUT of MSG-1. Connect a 3-foot length of RG-58/U from HP-477B to BOLOMETER input of HP-430C as shown in Figure 5-2.

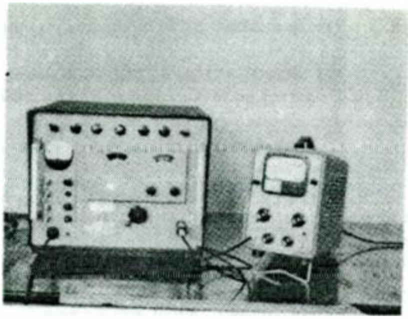


Figure 5-2. Output Tracking Error Test Configuration

d. Set controls on front panel of MSG-1 as follows:

RF OSC tuning control	1 kmc
MODULATION SELECTOR switch	ZERO SET
OUTPUT ATTENUATOR control	0 dbm
ZERO SET control	POWER SET indication

e. Set controls on front panel of HP-430C as follows:

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

f. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

g. Adjust the HP-430C ZERO SET controls for a zero indication on the top meter scale.

h. Verify a ZERO SET indication on output meter of MSG-1.

i. Set MODULATION SELECTOR switch on MSG-1 to CW. Set POWER SET control on MSG-1 for a POWER SET indication on output meter.

j. Set OUTPUT ATTENUATOR control on MSG-1 to zero dbm.

k. Switch MODULATION SELECTOR switch on MSG-1 from ZERO SET to CW verifying a ZERO SET and a POWER SET indication on output meter.

l. Switch MODULATION SELECTOR switch on MSG-1 to ZERO SET and ZERO SET HP-430C.

m. Switch MODULATION SELECTOR switch of MSG-1 to CW and record the variation from zero dbm as indicated on HP-430C.

n. Repeat Steps (h) through (m) for each frequency of calibration.

o. Verify that the indications recorded in Step (m) are within tolerance.

p. Turn HP-430C ZERO SET control maximum counterclockwise. Set the HP-430C BIAS CURRENT switch to OFF and disconnect HP-477B from RF OUTPUT of MSG-1. Turn LINE POWER switch of HP-430C to OFF.

q. Plot a graph of output deviation from zero dbm as a function of frequency.

5.6 OUTPUT DIAL ATTENUATION TRACKING ERROR

a. Connect the power cable of MSG-1 to a source of 115 volts ac, 50 - 60 cps.

b. Connect the power cable of NM-50A, to a source of 105-125 or 210-250 volts ac, single phase, 50-1600 cps.

c. Set MSG-1 front panel controls as follows:

RF OSC tuning control	1 kmc
MODULATION SELECTOR switch	CW
OUTPUT ATTENUATOR control	0 dbm
POWER SET control	POWER SET indication

d. Set NM-50A front panel controls as follows:

TUNING AID - PULSE STRETCHER switch	OFF
Function switch	FI
PULL-TURN-PUSH Attenuator	X10
TUNE control	1000
CAL control	2 db on microvolt meter

e. Insert a standard attenuator 50-5 in RF OUTPUT of MSG-1.

f. Set NM-50A PULL-TURN-PUSH attenuator to $X10^4$.

g. Connect one end of a 6-foot length of RG-9/U to open end of attenuator 50-5 and the other end to RF INPUT of NM-50A as shown in Figure 5-3.

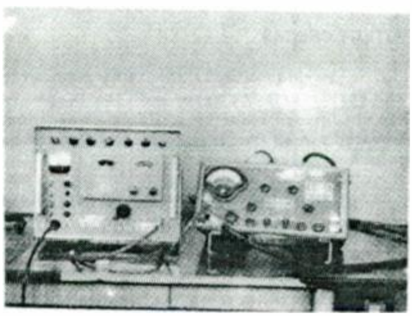


Figure 5-3. Output Dial Attenuation Tracking Test Configuration

h. Tune in output signal of MSG-1 on NM-50A. Peak RF TRIM and MIXER TRIM controls.

i. Switch MODULATION SELECTOR switch on MSG-1 from ZERO SET to CW, verify a ZERO SET and a POWER SET indication on output meter and a zero dbm output indication on attenuation dial.

j. Set NM-50A meter needle for a reference level by means of CAL control.

k. Remove attenuator 50-5 from RF OUTPUT of MSG-1 and connect the RF OUTPUT to RF INPUT of NM-50A by means of a 6-foot length of RG-9/U.

l. Decrease MSG-1 OUTPUT ATTENUATOR dial until the reference level originally set in Step (j) is indicated on NM-50A.

m. Record MSG-1 attenuation indication.

n. Repeat Steps (e) through (m) until the OUTPUT ATTENUATOR dial on MSG-1 has been tracked from 5 to 90 db in 5 db steps.

o. Disconnect attenuator 50-5 and NM-50A from RF OUTPUT of MSG-1.

5.7 PULSED OUTPUT CHECKS

a. Periodic checks should be made of the pulsed output to verify that the characteristics agree with the respective control dial settings. These checks are made by utilizing an oscilloscope with an accurately calibrated sweep frequency and a calibrated base line. The characteristics to be checked are:

(1) Pulse Rate - at 40 and 400 pulses per second

(2) Pulse Delay - at 300 microseconds delay

(3) Pulse Width - at 10 microseconds pulse output

b. Verify that the pulse amplitude output equals the CW peak output by measuring both with the appropriate FIM equipment of suitable frequency range.

6. CALIBRATION PROCEDURES

RF SIGNAL GENERATOR MSG-2A

6. RF SIGNAL GENERATOR MSG-2A

6.1 CALIBRATION CHECKS REQUIRED

a. Power output at zero dbm throughout the frequency range.

b. Attenuator dial linearity at selected frequencies.

c. Frequency dial tracking throughout the frequency range.

d. Percent modulation meter from 0 to 100%.

6.2 EQUIPMENT REQUIRED OR EQUIVALENT

Power Meter, HP-430C

Thermistor Mount, HP-477B

Frequency Meter, Lavoie LA-70A

Attenuators, Weinschel 50 Series

Receiver, Polarad FIM

Cables as required

6.3 PRELIMINARY OPERATION

Connect all equipment to a suitable a-c power source, energize equipment and allow a warm-up time of 30 minutes.

6.4 OUTPUT POWER MEASUREMENT

a. Setup of operating controls for HP-430C

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

b. Connect equipment as shown in Figure 6-1.

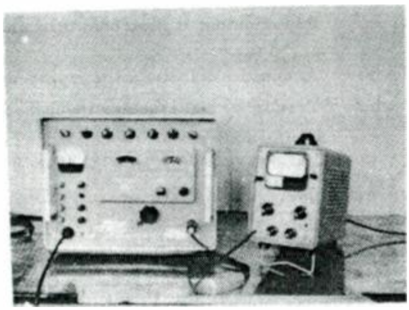


Figure 6-1. Power Measurement Test Configuration

c. Set controls on front panel of MSG-2A as follows:

RF OSC tuning control	2 kmc
MODULATION SELECTOR switch	ZERO SET
OUTPUT ATTENUATOR	0 dbm
POWER SET control	POWER SET red line indication

d. Set HP-430C front panel controls as follows:

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	Maximum clockwise
ZERO SET FINE control	Maximum clockwise

CAUTION

The power meter BIAS CURRENT switch must be set to OFF before connecting or disconnecting the thermistor mount.

e. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

f. Adjust HP-430C ZERO SET controls for a zero indication on the top meter scale.

g. Verify a ZERO SET condition on output meter of MSG-2A.

h. Set MODULATION SELECTOR on MSG-2A to CW. Adjust POWER SET control on MSG-2A for a POWER SET (red line) indication on output meter.

i. Set OUTPUT ATTENUATOR on MSG-2A to zero dbm.

j. Switch MODULATION SELECTOR switch on MSG-2A from ZERO SET to CW. Verify a ZERO and POWER SET indication on output meter.

k. Switch MODULATION SELECTOR switch on MSG-2A to ZERO SET. Zero HP-430C.

l. Switch MODULATION SELECTOR switch on MSG-2A to CW and record the variation from zero dbm as indicated on HP-430C.

m. Repeat Steps (g) through (l) for each frequency of calibration.

n. Verify that the indications recorded in Step (l) are within tolerance.

o. Turn HP-430C ZERO SET controls maximum counter-clockwise. Set the HP-430C BIAS CURRENT switch to OFF and disconnect HP-477B from RF OUTPUT of MSG-2A. Turn HP-430C LINE POWER switch OFF.

p. Plot a graph of output deviation from zero dbm as a function of frequency.

6.5 OUTPUT DIAL ATTENUATION

a. Set MSG-2A front panel controls as follows:

RF OSC tuning control	3 kmc
MODULATION SELECTOR switch	CW
OUTPUT ATTENUATOR	0 dbm
POWER SET control	POWER SET indication on output meter

b. Set FIM front panel controls as follows:

Monitor Unit

Function selector switch	AVERAGE
INPUT ATTENUATOR	0 db
AFC switch	OUT
IF GAIN control	10
METER SELECTOR switch	POWER MON
TUNING dial	3 kmc
POWER MONITOR meter function selector switch	FIM

Power Unit

Voltmeter selector switch	BIAS
KLYSTRON CATHODE CURRENT selector switch	LOCAL OSCILLATOR

c. Switch the FIM Power Unit VOLTMETER selector switch from BIAS to LOW VOLTAGE to REPELLER to BEAM. Verify a red line indication on VOLTMETER for all switch positions.

d. Verify that the KLYSTRON CATHODE CURRENT meter on the FIM Power Unit is indicating between 20 and 30 mcs.

e. Connect a 50-5 attenuator between the OUTPUT of the MSG-2A and INPUT of the FIM Monitor Unit as shown in Figure 6-2.

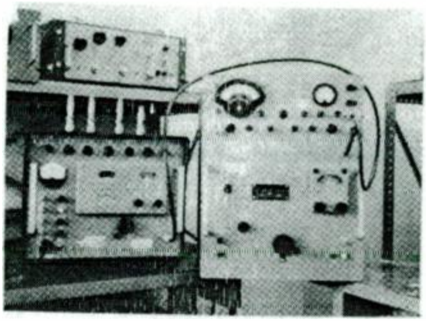


Figure 6-2. Output Attenuator Linearity Test Configuration

f. Tune in output signal of Signal Generator MSG-2A on Polarad FIM.

g. Vary INPUT ATTENUATOR on Polarad FIM Monitor Unit to cause an on-scale meter indication.

h. Switch MODULATION SELECTOR switch on Signal Generator MSG-2A from ZERO SET to CW, verify a ZERO and POWER SET indication on output meter and a zero dbm output indication on attenuator dial.

i. Set FIM Monitor Unit meter needle for a reference level by means of Monitor Unit IF GAIN control.

j. Remove 50-5 attenuator from RF OUTPUT of MSG-2A and connect the RF OUTPUT of MSG-2A to RF INPUT of FIM Monitor Unit by means of a 6-foot length of RF-9/U.

k. Decrease the OUTPUT ATTENUATOR dial of MSG-2A until the reference level originally set in Step (i) is indicated on the Polarad FIM Monitor Unit Meter.

l. Record MSG-2A attenuation indication.

m. Repeat Steps (e) through (l) until the output attenuator dial of MSG-2A has been tracked from 5 to 90 db in 5 db steps.

6.6 FREQUENCY DIAL TRACKING ERROR

a. Set MSG-2A front panel controls as follows:

RF OSC tuning control	2 kmc
MODULATION SELECTOR switch	CW
OUTPUT ATTENUATOR	0 dbm
FM AMPLITUDE control	Maximum counter-clockwise

b. Set LA-70A front panel controls as follows:

SEARCH-MC LOCK-LOCK switch	SEARCH
MC RANGE switch	42-60
DIAL C	0
DIAL B	0

c. Connect a 50-5 attenuator between the RF OUTPUT of MSG-2A and the MIXER input of HP-540B as shown in Figure 6. 3.

d. Connect 10 kc to 1000 mc input of LA-70A to FREQUENCY METER INPUT of HP-540B as shown in Figure 6. 3.

e. Turn MODULATION SELECTOR switch on MSG-2A to ZERO SET and turn ZERO SET control for a zero indication on the output meter.

f. Turn MODULATION SELECTOR switch on MSG-2A to CW, adjust POWER SET control for red line indication on output meter.

g. Tune the Course Vernier and then Fine Vernier of HP-540B until the highest harmonic zero beat of MSG-2A is indicated on the scope of the HP-540B. Repeat the above step for the next lowest zero beat. The harmonic can be found by use of the following formula:

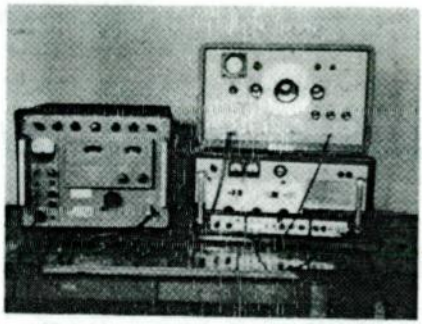


Figure 6-3. Frequency Tracking Test Configuration

$$N = \frac{F_2}{(F_2 - F_1)}$$

where: N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lowest zero beat

h. Search for the highest fundamental output of HP-540B by rotating DIAL A of LA-70A.

i. On locating beat with DIAL A of LA-70A, turn function selector switch to MC LOCK.

j. Turn DIAL B on LA-70A until meter A deflects (beat disappears).

k. Regain zero beat by tuning DIAL B of LA-70A back and forth. Verify that MC stage is still locked.

l. Turn function selector switch of LA-70A to LOCK and adjust DIAL B to first lock point lower in frequency from beat note.

m. Regain zero beat with DIAL C of LA-70A.

n. Read frequency as indicated by DIALS A, B, and C.

o. Multiply the indicated frequency by the harmonics used.

p. Harmonic number for LA-70A can be found by using the following formula:

$$N = \frac{F_2}{(F_2 - F_1)}$$

where: N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lower zero beat

q. The output frequency of MSG-2A is a product of the harmonic of the HP-540B times the harmonic of LA-70A times the frequency as indicated by DIALS A, B, and C of LA-70A.

r. Verify that frequency is within tolerance (1%).

s. Repeat above steps at selected output frequencies.

t. Disconnect equipment from output of MSG-2A.

6.7 PULSED OUTPUT CHECKS

a. Periodic checks should be made of the pulsed output to verify that the characteristics agree with the respective control dial settings. These checks are made by utilizing an oscilloscope with an accurately calibrated sweep frequency and a calibrated base line. The characteristics to be checked are:

(1) Pulse Rate - at 40 and 400 pulses per second

(2) Pulse Delay - at 300 microseconds delay

(3) Pulse Width - at 10 microseconds pulse output

b. Verify that the pulse amplitude output equals the CW peak output by measuring both with the appropriate FIM equipment of suitable frequency range.

7. CALIBRATION PROCEDURES

RADIO INTERFERENCE - FIELD INTENSITY METER

STODDART MODEL NM-10A

FRC

7. CALIBRATION PROCEDURE FOR RADIO INTERFERENCE - FIELD INTENSITY METER STODDART MODEL NM-10A

7.1 CALIBRATION TESTS

a. Utilizing 3 frequencies in each band, equipment is calibrated for standard gain as prescribed by the instruction manual.

b. The meter is accurately calibrated in the field intensity position against the internal calibrator, which provides a known source of random noise. Direct meter readings of random interference are then available and are converted to "microvolts-per-kilocycle bandwidth" by the use of the calibration charts. If calibration charts 90152-4 (8 each) are more than 10% inaccurate, new curves will be drawn.

7.2 EQUIPMENT REQUIRED

a. Standard Signal Generator, General Radio Type 1001A.

b. Six foot length of RG-9/U coaxial cable properly terminated.

7.3 PRELIMINARY OPERATION

a. Connect equipment to a suitable a-c power source, energize equipment and allow a warm-up time of one hour.

7.4 FINAL ADJUSTMENT OF THE X10 ATTENUATOR

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X1
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply an unmodulated 125 kc signal and adjust its output to exactly 100 microvolts.

d. Carefully tune in the signal on the NM-10A and adjust the CAL control for a meter reading of exactly 10 microvolts.

e. Increase the GR-1001A output to 1000 microvolts and operate the NM-10A ATTENUATOR to X10. Verify a 10 microvolts indication on the NM-10A meter.

7.5 FINAL ADJUSTMENT OF THE X10^2 ATTENUATOR

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X10
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply and unmodulated 125 kc signal and adjust its output to exactly 100 microvolts.

d. Carefully tune in the signal on the NM-10A and adjust the CAL control for a meter reading of exactly 10 microvolts.

e. Increase the GR-1001A output to 1000 microvolts and operate the ATTENUATOR to X10^2.

7.6 FINAL ADJUSTMENT OF THE X10^3 ATTENUATOR

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X10^3
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply an unmodulated 125 kc signal and adjust its output to exactly 1000 microvolts.

d. Carefully tune in the signal of the NM-10A and adjust the CAL control for a meter reading of exactly 10 microvolts.

e. Increase the GR-1001A output to 10,000 microvolts and operate the ATTENUATOR to X10^3. Adjust the trimmer C-112 (X10^3 TRIM), located on the bottom of the antenna can, for maximum meter indication.

f. After adjusting C-112 (X10^3 TRIM) for maximum meter indication, note meter reading. If it is not exactly 10 microvolts, it will be necessary to readjust C-111 (X10^3 ATTN). Moving C-111 (X10^3 ATTN) in either direction will cause meter indication to decrease, and the proper direction of adjustment of C-111 (X10^3 ATTN) must be determined as follows: Adjust C-111 (X10^3 ATTN) in either direction to decrease meter reading about 1 db and note carefully in which direction adjustment is made. Readjust C-112 (X10^3 TRIM) for maximum meter indication. If reading is now closer to 10 microvolts than it was at the beginning of this step, adjustment of C-111 (X10^3 ATTN) was in the proper direction. If reading is farther from 10 microvolts than it was at the beginning of this step, C-111 (X10^3 ATTN) must be adjusted in the opposite direction.

g. Note distance of meter indication from 10 microvolts. Adjust C-111 (X10^3 ATTN) in the proper direction as determined in Step (f), to decrease indication by an amount about equal to that distance. Readjust C-112 (X10^3 TRIM) for maximum indication.

h. Repeat Step (g) as required to bring the meter indication to exactly 10 microvolts with a final adjustment of C-112 (X10^3 TRIM).

7.7 FINAL ADJUSTMENT OF THE X10^4 ATTENUATOR

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X10^3
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set GR-1001A controls to supply an unmodulated 125 kc signal and adjust its output to exactly 10,000 microvolts.

d. Carefully tune in the signal of the NM-10A and adjust the CAL control for a meter reading of exactly 10 microvolts.

e. Increase the GR-1001A output to 100,000 microvolts and operate the ATTENUATOR to X10^4. Verify a meter indication of exactly 10 microvolts.

7.8 FINAL ADJUSTMENT OF THE ADJ 10 AND ADJ 100 CONTROLS

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X10
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply an unmodulated 125 kc signal and adjust its output to 10 microvolts.

d. Carefully tune in the signal of the NM-10A and adjust the CAL control for exactly 1 microvolt meter indication.

e. Set the GR-1001A output to 1000 microvolts. Verify a 100 microvolts meter indication.

f. Set the GR-1001A output to 100 microvolts. Verify a meter indication of exactly 10 microvolts.

g. Set the GR-1001A output to 10 microvolts. Verify a 1 microvolt meter indication.

7.9 ADJUSTMENT OF THE PEAK FULL SCALE CONTROL

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X1
BFO switch	OFF
Band switch	60 - 125 kc

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply an unmodulated 125 kc signal and adjust its output to 100 microvolts.

d. Carefully tune in the signal on the NM-10A and adjust the CAL control for exactly 100 microvolts meter indication.

e. Operate FUNCTION switch to QUASI-PEAK. Verify a 100 microvolts meter indication.

f. After the equipment has been completely aligned, obtain data for Chart No. 1 at each of the following frequencies:

BAND 1 14 - 30 kc	BAND 2 30 - 60 kc	BAND 3 60 - 125 kc	BAND 4 125 - 250 kc
14	30	60	125
18	37	75	156
22	45	90	187
26	52	105	218
30	60	125	250

7.10 CALIBRATION PROCEDURE

a. Setup of operating controls:

Input selector switch	ANTENNA OPERATE
FUNCTION switch	FIELD INTENSITY
ATTENUATOR	X1

BFO switch	OFF
Band switch	As called for

b. Use the GR-1001A output cable terminating in a coaxial connector type UG-21/U. Connect Impedance Matching Network 90081-4 into the ANTENNA receptacle and connect the GR-1001A cable to this network.

c. Set the GR-1001A controls to supply an unmodulated 14 kc signal with 100 microvolts output.

d. Set the band switch to 14 - 30 kc and carefully tune in the signal on the NM-10A. Adjust the CAL control for exactly 100 microvolts meter indication.

e. Operate the input switch to ANTENNA CAL and the FUNCTION switch to CAL.

f. Record the meter indication in microvolts.

g. Repeat Steps (c), (d), (e), and (f) for each frequency.

h. Using the right-hand vertical scales of Chart No. 1, plot the data at each frequency. Draw a smooth curve through the five plotted points in each frequency band.

8. CALIBRATION PROCEDURES

RADIO INTERFERENCE - FIELD INTENSITY METER

STODDART MODEL NM-20A

FRC

8. CALIBRATION PROCEDURES FOR RADIO INTERFERENCE-FIELD INTENSITY METER, STODDART MODEL NM-20A

8.1 CALIBRATION TESTS

a. Utilizing 3 frequencies in each band, equipment is calibrated for standard gain as prescribed by the instruction manual.

b. The NM-20A includes a highly sensitive radio receiver covering the 0.15 to 25 mc portion of the radio frequency spectrum. It also contains internal means for calibrating or standardizing its gain, thus permitting direct readings in indicated microvolts or microvolts-per-meter. The internal calibrator supplies a shot noise calibrating voltage that follows the same signal path from RF stage to indicating meter. With the calibrator on, the overall gain of the IF section is adjusted, by means of the CAL control, to the standardized signal level obtained from the calibration chart for the frequency of operation. The built-in step attenuator provides the following step ratios: 1, 10, 100, 1000, and 10,000. The meter is accurately standardized in the field intensity position against the shot noise source to obtain a previously calibrated standard meter reading. Direct meter readings on random interference are then available, and are converted to "microvolts-per-kilocycle-of-bandwidth" by the use of the calibration chart. If the calibration charts are more than 10% inaccurate, new curves will be drawn.

8.2 PRELIMINARY OPERATION

a. Connect the power cable to a suitable a-c source. The clip on the extra lead at the power plug end should be securely connected to a good ground.

b. Allow equipment one hour warm-up time prior to start of calibration.

8.3 METER TRACKING

a. Setup of operating controls:

Input control	Any
Tuning dial frequency	Any
Attenuator control	$X10^2$
BAND selector	Band 2
Power switch	AC
Function switch	FI

'A' ADJ and 'B' ADJ	Meter indication at red mark.
CAL	As required
BFO	OFF

b. Connect the GR-1001A to the ANT input connector of NM-20A.

c. Set CAL control approximately two-thirds clockwise. Set the BAND selector to Band 2 and the GR-1001A controls to supply an unmodulated 535 kc signal. Set the function switch to the FI position. Set GR-1001A output for 1000 microvolts and adjust GR-1001A frequency for maximum deflection of M-101. Adjust CAL control for a meter reading of one microvolt.

d. Increase the GR-1001A output exactly 100 times. Verify a meter reading of 100.

e. Reduce the GR-1001A output exactly 10 times. Verify a meter reading of 10.

f. Reduce the GR-1001A output exactly 10 times. Verify a meter reading of 1.

8.4 ADJUSTMENT OF THE QP AND PEAK FULL SCALE CONTROLS

a. Setup of operating controls:

Input control	As called for
Tuning dial frequency	As called for
Attenuator control	$X10^0$
BAND selector	As called for
Power switch	AC
Function switch	As called for
'A' ADJ and 'B' ADJ	Meter indication at red mark
CAL control	As called for
BFO	OFF

b. Connect the GR-1001A to the ANT input connector of NM-20A.

c. Set the BAND selector to Band 2 and the GR-1001A controls to supply an umodulated 535 kc signal. Set the function switch to the FI position. Set GR-1001A output for 1000 microvolts and adjust GR-1001A frequency for maximum deflection of M-101.

d. Adjust CAL control for a meter reading of one microvolt.

e. Set the function switch to the QP position. Verify a meter reading of one microvolt.

f. Increase GR-1001A output 100 times. Verify a meter reading of 100 microvolts.

g. Decrease GR-1001A output 100 times and check for a meter reading of one microvolt. Set function switch to QP position and check that a meter reading of one microvolt is obtained. Repeat Steps (e) and (f) if necessary.

8.5 CALIBRATION PROCEDURE

a. Setup of operating controls:

Input control	As called for
Tuning dial frequency	As called for
Attenuator control	X10
BAND selector	As called for
Power switch	AC
Function switch	As called for
'A' ADJ and 'B' ADJ	Meter indication at red mark
CAL control	As called for
BFO	OFF

NOTE

THE ABSOLUTE VALUE OF CALIBRATION DEPENDS UPON THE ACCURACY OF THE CALIBRATING SOURCE. BE CERTAIN THAT THE SIGNAL GENERATOR IS PROPERLY TERMINATED, PARTICULARLY AT FREQUENCIES ABOVE TEN MEGACYCLES.

b. Connect the signal generator to the NM-20A.

c. Check the meter zero and diode current by setting the function switch to these positions and observe that the meter does read at zero and within the diode current range. Check the A and B voltage adjustments.

d. Set the signal generator at the desired frequency and adjust the signal generator output to provide 1000 microvolts at the input of the NM-20A. Observe any correction factor necessary because of termination requirements.

e. Set the input control to ANT or LOOP, according to the input device in use. Set the function switch to the FI position.

f. Set the BAND selector to the desired band and carefully tune to the test signal.

g. Adjust the CAL control to give a meter reading of 100. Quickly check and adjust the A and B voltages as necessary. For greatest accuracy, the B voltage must be adjusted again after the signal has been introduced, especially for signals reading in the upper portion of the scale. This will be most evident with operation on fresh B batteries.

h. Set the attenuator control to the X1 position.

i. Set the input control to CAL and the function switch to CAL.

j. Record the meter reading on the db scale of M-101 as the calibrating figure to be plotted on the appropriate frequency band of charts Nos. 1 through 6, according to the input device used.

NOTE

FOR GREATEST ACCURACY, CORRECT THE B VOLTAGE ADJUSTMENT JUST BEFORE READING AND RECORDING THE METER READING.

9. CALIBRATION PROCEDURES

RADIO INTERFERENCE - FIELD INTENSITY METER

STODDART MODEL NM-30A

FRC

9. CALIBRATION PROCEDURES FOR RADIO INTERFERENCE-FIELD INTENSITY METER, STODDART MODEL NM-30A

9.1 CALIBRATION TESTS

a. Utilizing 5 frequencies in each band, equipment is calibrated for standard gain as prescribed by the instruction manual.

b. The NM-30A includes a highly sensitive radio receiver covering the range of 20 to 400 mc. The meter is accurately standardized in the field intensity position against the shot noise source to obtain a previously calibrated standard meter reading. Direct meter readings on random interference are then available, and are converted to microvolts-per-kilocycle-of-bandwidth by the use of calibration charts. If the calibration charts are more than 10% inaccurate, new curves will be drawn.

9.2 PRELIMINARY OPERATION

a. Connect the power cable to a suitable a-c source. The clip on the extra lead at the power plug end should be securely connected to a good ground.

b. Allow equipment one hour warm-up time prior to start of calibration.

9.3 CHARTS NO. 6 AND NO. 7 (RANDOM NOISE BANDWIDTH)

The overall bandwidth of the NM-30A is influenced by the selectivity of the RF tuned circuits; therefore, bandwidth data must be taken at radio frequencies listed immediately preceding.

A signal generator is ordinarily used for measuring bandwidth. This method is suitable for the low end of the frequency range but becomes too critical at the higher frequencies. At the higher frequencies the internal impulse noise calibrator will be used in determining bandwidth. Advantage will be taken of its constant output per unit bandwidth and that the peak value of an impulse signal is directly proportional to bandwidth. By dividing the calibrator output by the bandwidth at a low frequency, a factor is derived which can be used at other frequencies. The accuracy of measuring bandwidth by the following procedure depends on the accuracy of the signal generator output.

a. Setup of operating controls:

Attenuator	X1
Function switch	FIELD INTENSITY
CALIBRATE	As called for

b. Connect the output of the signal generator to J101 using the 90933-1 RF transmission line.

c. Adjust the signal generator frequency to 20 megacycles. Adjust output to 100 microvolts.

d. Tune the NM-30A for maximum response to the test signal. Adjust the CALIBRATE control for M101 indication of 50 microvolts.

e. Increase the signal generator output by exactly 6 db (2 times).

f. Decrease the signal generator frequency to about 18 megacycles, then slowly increase the frequency beyond 20 megacycles.

Observe the two signal generator frequencies at which M101 indicates 50 microvolts. Record the difference in frequency between the two observed points; this is the overall 6 db bandwidth at this test frequency. In order to counteract backlash in the signal generator tuning mechanism, always sweep the signal generator frequency continuously in one direction from a point below the calibrating frequency upward to a point above the calibrating frequency. The two frequencies will be quite close together. These frequencies must be measured with the LA-70A frequency meter for maximum accuracy.

g. Standardize the gain as per instructions on first page of chart, set at test frequency of 20 mc; but instead of using the calibrator setting provided on Chart 1, use 80 microvolts.

h. Turn attenuator to X1 and function switch to FIELD INTENSITY position.

i. Adjust signal generator to 20 megacycles. Tune NM-30A for maximum response. Adjust generator output for an NM-30A output meter indication of 80 microvolts, same as the calibrator setting in Step (g). Note the generator output in microvolts.

j. Divide the signal generator output in microvolts by the 6 db bandwidth in kilocycles determined in Step (f). This factor is the peak output of the calibrator in microvolts per 1 kc bandwidth.

k. To use the factor derived in Step (j) in determining bandwidth at other frequencies, repeat Steps (g) through (i) at each test frequency. The generator output in microvolts from Step (i), divided by the factor in Step (j), will be the 6 db bandwidth in kilocycles. The random noise bandwidth can be determined by dividing the 6 db bandwidth by 1.3.

l. Using the vertical scale of Charts No. 6 and No. 7, plot the points of random noise bandwidth determined in Step (k). Draw a smooth curve through the plotted points.

9.4 CHARTS NOS. 1 AND 2 CALIBRATION SETTINGS

a. Setup of operating controls:

Attenuator	As called for
Function switch	As called for
PEAK	As called for
PEAK SENS.	As called for
TUNING dial frequency	As called for

b. Using the RF transmission line (90933-1), connect the NM-30A to the output of the signal generator.

NOTE

THE ABSOLUTE VALUE OF CALIBRATION DEPENDS ON THE ACCURACY OF THE CALIBRATING SOURCE.

c. Adjust the signal generator frequency to 20 megacycles. Adjust the output to 100 microvolts.

d. Set the NM-30A attenuator to X1 position, function switch to FIELD INTENSITY position. Tune the NM-30A for maximum response to the test signal. Adjust the CALIBRATE control for M101 indication of 100 microvolts.

e. Turn the attenuator to CAL position. Set the function switch to CAL. Turn the PEAK control maximum clockwise.

f. Rotate the PEAK SENS. control clockwise until visual PEAK indicator lamp glows. Then rotate PEAK SENS. control counter-clockwise until lamp just ceases to light. (The visual null indicator circuit is now adjusted to its most sensitive operating point for the detection of the calibrator pulse.)

g. Turn the PEAK control counter-clockwise until the visual PEAK indicator lamp blinks in unison with pulses emitted by calibrator (headphones may be used for aural identification). Then rotate the PEAK control clockwise until lamp just ceases to blink. Note M101 indication on db scale and record.

NOTE

THE AURAL NULL METHOD MAY BE USED TO VERIFY VISUAL NULL MEASUREMENT OF CALIBRATOR PULSE AMPLITUDE.

h. Repeat Steps (c) through (g) for each of the test frequencies in Bands 1 through 4.

i. Using the vertical scale of Charts No. 1 and No. 2, plot the values of calibrator pulse amplitude in db which were determined in Step (g). Draw a smooth curve through the plotted points.

j. Repeat Steps (c) and (d) at test frequency of 145 megacycles in Band 5. Then reduce signal generator output to zero. If M101 indication is less than one microvolt, use 100 microvolt signal generator output as suggested in Step (c). If M101 indication is greater than one microvolt, use a signal generator output of 200 microvolts in Step (c).

k. Repeat Steps (c) through (g) for each of the test frequencies in Band 5 but using signal generator output indicated by Step (j). If 100 microvolts is used, the correction factor for Band 5 will be 1. If 200 microvolts is used, the correction factor for Band 5 will be 2.

l. Using the vertical scale of Charts No. 1 and No. 2, plot the values of calibrator amplitude in db which were determined in Step (g). Draw a smooth curve through the plotted points.

m. For test frequencies in Band 6, a different procedure will be used as the sensitivity will not reach the two microvolt level at some frequencies.

To achieve maximum sensitivity capabilities of this band, operate with a signal-to-noise ratio close to unity. This means the noise level will be maintained at or just below the 1 microvolt level. Take data as follows:

(1) Set up the following operating controls:

Attenuator	X1
Function switch	FIELD INTENSITY

(2) Tune the NM-30A for maximum response to a generator test signal at a frequency of 240 megacycles.

(3) Decrease signal generator output to zero. Adjust CAL control for M101 indication of one microvolt. Adjust signal generator output for an M101 indication of 100 microvolts. Note signal generator output in microvolts. This divided by 100 will be the correction factor.

(4) Repeat Steps (1) through (3) at a test frequency of 400 megacycles.

(5) The chart for Band 6 can now be started. The ordinate will be a linear scale marked off in frequencies (240 to 400 megacycles). There will be two abscissas. The one on the left will be calibration settings in db. The one on the right will be correction factors in units (1 to 6). Plot the two correction factors taken in Step (3). Draw a straight line between these two points.

(6) From this correction factor line, we will derive the corresponding calibrator settings. Tune signal generator to 240 megacycles. Adjust the generator output to 100X microvolts where X = correction factor at the test frequency of 290 megacycles.

(7) Tune the NM-30A for maximum response to the generator signal. Adjust the CAL control for an M101 indication of 100 microvolts.

(8) Change the attenuator to CAL position. Set the function switch to CAL. Turn the PEAK control maximum clockwise. Rotate the PEAK SENS. control clockwise until the PEAK indicator lamp glows, then rotate counter-clockwise until lamp just ceases to light. (The visual null indicator circuit is now adjusted to its most sensitive operating point for the detection of the calibrator pulse.)

(9) Rotate PEAK control counter-clockwise until PEAK indicator lamp blinks in step with calibrator pulses, then adjust clockwise until lamp just ceases to blink. Note M101 indication in db. This is the calibrator setting at this test frequency.

(10) Repeat Steps (6) through (9) for each remaining test frequency in Band 6. Plot calibrator settings on Band 6 chart. Draw a smooth curve through the plotted points.

Should there be insufficient range in the CAL control for Step (7) at test frequencies in the middle portion of band, RF realignment of Band 6 is indicated.

9.5 OTHER CHART SETTINGS

Chart No. 3, an antenna adjustment table, Charts No. 4 and No. 5, antenna pickup device factors, and Chart No. 8, signal plus noise chart, are not influenced by realignment of the NM-30A. No corrections, therefore, need be made.

10. CALIBRATION PROCEDURES
RADIO INTERFERENCE-FIELD INTENSITY METER
STODDART MODEL NM-50A

FRC

10. CALIBRATION PROCEDURE FOR RADIO INTERFERENCE - FIELD INTENSITY METER, STODDART MODEL NM-50A

10.1 CALIBRATION CHECKS REQUIRED

a. Utilizing 18 frequencies, the equipment is calibrated for standard gain as prescribed by the instruction manual.

b. The NM-50A is a sensitive radio receiver operating as a selective radio frequency voltmeter over the 375 to 1000 mc radio spectrum. It contains an internal means for calibrating or standardizing its gain. Direct meter readings on random interference are then available, and are converted to microvolts-per-megacycle-of-bandwidth. If the calibration charts are more than 10% inaccurate, new curves will be drawn.

10.2 PRELIMINARY OPERATION

a. Connect the power cable to a suitable a-c source. The clip on the extra lead at the power plug end should be securely connected to a good ground.

b. Allow equipment one hour warm-up time prior to start of calibration.

10.3 CALIBRATING PROCEDURE—STANDARD GAIN

a. Setup of operating controls:

Attenuator	CAL
Function switch	As called for
TUNE control	As called for
TUNING AID - PULSE STRETCHER switch	OFF
ADJ control	As called for
CAL control	As called for

b. Set function switch to ADJ position. Adjust ADJ control for a reading of 10 microvolts on M-101.

c. Set function switch to CAL position. Tune in internal calibrating signal (approximately 800 megacycles) and adjust RF TRIM and MIXER TRIM controls for maximum meter indication.

d. Adjust CAL control for a meter reading of 100 microvolts. This setting of the CAL control should not be changed until the equipment gain is again standardized.

10.4 CHART NO. 2, CORRECTION FACTORS FOR USE WITH LINE PROBE 90338-2 AND MATCHING IMPEDANCE 90339-2

NOTE

BECAUSE THE EQUIPMENT SETUP IS THE SAME, DATA FOR CHART NO. 3 CAN BE TAKEN ALONG WITH DATA FOR CHART NO. 2. SEE PARAGRAPH 10. 3d.

a. Setup of operating controls:

Attenuator	As called for
Function switch	FIELD INTENSITY
TUNE control	As called for
TUNING AID - PULSE STRETCHER switch	OFF

b. Using a transmission line (90343-1), connect the NM-50A to the appropriate signal generator.

NOTE

THE ABSOLUTE VALUE OF CALIBRATION DEPENDS ON THE ACCURACY OF THE CALIBRATING SOURCE.

c. Set the NM-50A attenuator to X10 position. Adjust NM-50A to 375 megacycles on the tuning dial and adjust RF TRIM and MIXER TRIM controls for maximum meter indication on signal from signal generator.

d. Adjust the signal generator output to provide a 100 microvolt deflection at M101.

e. Divide the NM-50A indicated output (100 microvolts times attenuator factor X10 equals 1000 microvolts) by the output from the signal generator. This product is the correction factor at 370 megacycles for attenuator positions X10 and X10^2. This factor applies to both attenuator positions since the RF termination is the same.

f. Repeat Steps (c) and (d), using the X10^3 attenuator position. Compute correction factor as in Step (e) for attenuator positions X10^3, X10^4, and X10^5. This factor applies to these attenuator positions since the RF termination remains the same.

g. Repeat Steps (a) through (f) for each calibration frequency.

h. Using the vertical scales of Chart No. 2, plot the data at each frequency. Draw a smooth curve through the plotted points of each group.

10.5 CHART NO. 1, CORRECTION FACTORS FOR USE WITH ANTENNA 90330-2

This correction factor, K_1, is computed from the formula:

$$K_1 = K_s \cdot K_d$$

where: K_s is the factor obtained from Chart No. 2,
K_d is the dipole factor obtained from Figure 10-1.

The dipole factors are derived from measurements taken of impedance between the dipole elements and from measurements of physical dimensions. These factors are plotted in Figure 10-1 and apply to all antenna (90330-2) units. Data shown in Figure 10-1 is required for plotting data in Chart No. 1 only.

Compute the corresponding K_1 factor from each of the K_s factors determined in paragraph 10. 3b. Using vertical scales of Chart No. 1, plot K_1 factor data at each frequency. Draw a smooth curve through the plotted points of each group.

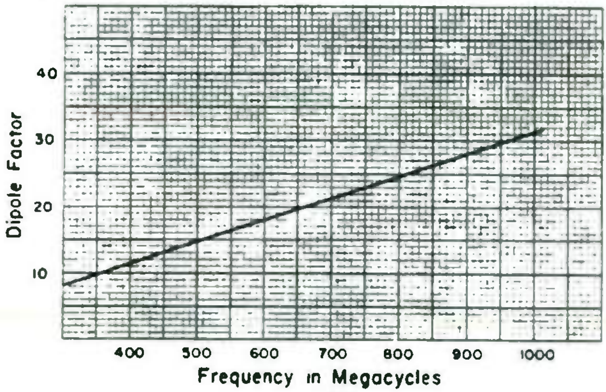


Figure 10-1. Antenna Dipole Factors for Plotting Chart No. 1

10.6 CHART NO. 3, EFFECTIVE BANDWIDTH

The overall bandwidth of the NM-50A is influenced by the selectivity of the RF tuned circuits Z-102 and Z-103 (RF and mixer butterfly circuits). Bandwidth data must therefore be taken at radio frequencies.

a. Setup of operating controls:

Attenuator	X10
Function switch	QP
Tune control	As called for
TUNING AID - PULSE STRETCHER switch	OFF
CAL control	Adjust for Standard Gain
ADJ control	Adjust for Standard Gain

b. Using transmission line CG-92D/U (20'0"), connect the NM-50A to the appropriate signal generator.

c. Adjust the signal generator to 375 megacycles and tune the NM-50A to the signal generator, adjusting RF TRIM and MIXER TRIM controls for maximum meter indication.

d. Adjust the signal generator output to provide some convenient up-scale deflection on the NM-50A, such as 60 microvolts.

e. Increase signal generator output by 3.5 db.

f. Decrease the signal generator frequency to about 370 megacycles, then slowly increase the frequency beyond 375 megacycles. Measure the two signal generator frequencies at which the indicating meter of the NM-50A reads 60 microvolts. Record the difference in frequency between the two observed points; this difference is the effective bandwidth at that calibrating frequency. In order to counteract backlash in the signal generator tuning mechanism, always sweep the signal generator frequency continuously in one direction from a point below the calibrating frequency upward to a point above the calibrating frequency.

g. Repeat Steps (c) through (f) for each of the following frequencies: 550, 750, and 1000 megacycles.

h. Using the vertical scale of Chart No. 3, plot the points of impulse bandwidth determined in Step (f). Draw a smooth curve through the plotted points. Typical effective bandwidth values are as follows:

Frequency (Megacycles)	Effective Bandwidth (Megacycles)
375	0.6
550	0.75
750	0.9
1000	1.05

11. CALIBRATION PROCEDURES

RF SIGNAL GENERATOR POLARAD FIM-L, S, M, AND X

11. CALIBRATION PROCEDURES FOR RF SIGNAL GENERATOR POLARAD FIM-L, S, M, and X

11.1 CALIBRATION CHECKS REQUIRED

a. The calibration of the frequency dial is verified at selected frequencies by using Transfer Oscillator HP-540B together with Frequency Meter, Lavoie LA-70A, as a secondary standard.

b. The power output is checked at zero dbm by monitoring the power output with a power meter and measuring any variation from this level resulting from a change in frequency.

c. The OUTPUT ATTENUATOR dial is checked to determine its linearity at selected output frequencies. With zero dbm output, a known attenuator is inserted in the line between the TUNING UNIT and the receiver. With the attenuator removed, the TUNING UNIT OUTPUT ATTENUATOR is varied for a comparable receiver indication.

11.2 EQUIPMENT REQUIRED

a. Power Meter, HP-430C

b. Thermistor Mount, HP-477B

c. Frequency Meter, Lavoie LA-70A

d. Transfer Oscillator, HP-540B

e. Attenuator, Weinschel 50 Series

f. Interconnecting cables as required

11.3 PRELIMINARY OPERATIONS

a. Connect equipment as shown in Figure 11-1.

b. Verify that all power switches are OFF.

c. Set the HP-430C BIAS CURRENT switch to OFF.

d. Plug power cable of HP-430C into a suitable power source.

e. Turn HP-430C POWER switch to ON.

f. Connect the power cable of LA-70A to an a-c source of 115 v, 50-420 cps.

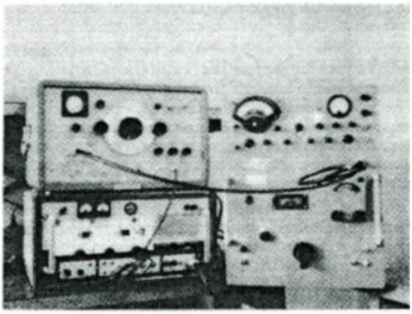


Figure 11-1. Frequency Dial Tracking Error Test Configuration

g. Turn Lavoie LA-70A power switch to ON.

h. Plug power cable of HP-540B into a suitable a-c power source.

i. Turn HP-540B POWER switch to ON.

j. Connect FIM interconnecting cable W903 (8 ft.) to receptacle J201 at the rear of FIM POWER UNIT and connect the other end of cable to receptacle J18 at the rear of FIM MONITOR UNIT.

k. Connect FIM power cable to power input connector J203 on FIM POWER UNIT and plug in to a suitable voltage source.

l. Remove TUNING UNIT FIM-L, S, M, or X from its dust cover by unfastening the four cam locks located in each of the four corners of the front panel.

m. Extend the sliding rails of the FIM MONITOR UNIT fully and insert tuning unit. Push TUNING UNIT into MONITOR UNIT and secure the four cam locks.

n. Open the 6 x 8 inch door located on the left side of the FIM MONITOR UNIT dust cover and connect the following receptacles:

TUNING UNIT FIM-L

(1) J411 to P118
(2) J427 to P427
(3) J438 to P138
(4) J439 to P139

TUNING UNIT FIM-S

(1) J511 to P118
(2) J527 to P427
(3) J538 to P138
(4) J539 to P139

TUNING UNIT FIM-M

(1) J611 to P118
(2) J627 to P427
(3) J638 to P138
(4) J639 to P139

TUNING UNIT FIM-X

(1) J711 to P118
(2) J727 to P427
(3) J738 to P138
(4) J739 to P139

o. Close side door on MONITOR UNIT and turn power switch to ON.

11.4 FREQUENCY DIAL TRACKING ERROR

a. Set the controls on the front panel of the TUNING UNIT under calibration as follows:

MONITOR UNIT

METER SELECTOR switch	POWER MON.
Function selector switch	ZERO SET
ZERO SET control	ZERO POWER

TUNING UNIT FIM-L

TUNING control	1 kmc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

TUNING UNIT FIM-S

TUNING control	2.14 kmc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

TUNING UNIT FIM-M

TUNING control	4.20 kmc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

TUNING UNIT FIM-X

TUNING control	7.36 kmc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

b. Connect a 5 db attenuator, 50-5, to tuning unit SIG GEN OUT connector.

c. Connect the SIGNAL INPUT jack of HP-540B, by means of a 6-foot length of RG-9/U, to the attenuator in (b) above.

d. Set LA-70A front panel controls as follows:

SEARCH-MC LOCK switch	SEARCH
MC RANGE switch	42 - 60
DIAL C	0
DIAL B	0

e. Connect 10 kc - 1000 mc input of LA-70A to FREQUENCY METER connector on HP-540B by means of a 3-foot length of RG-58/U.

f. Turn POWER MONITOR meter function selector switch to CAL SIG GEN. Set POWER SET control on tuning unit for a CAL-line indication on POWER MONITOR meter.

g. Set TUNING UNIT OUTPUT ATTENUATOR control for zero dbm.

h. Tune the Course Vernier of HP-540B until the highest harmonic zero beat of the tuning unit output signal is indicated on the scope of the HP-540B. Repeat the above step for the next lower zero beat. The correct harmonic can be determined by use of the following formula:

$$N = \frac{F_2}{(F_2 - F_1)}$$

where: N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lower zero beat

i. Search for the highest fundamental output of HP-540B by rotating DIAL A on LA-70A.

j. On locating the highest zero beat with DIAL A of LA-70A, turn function selector switch to MC LOCK.

k. Turn DIAL B on LA-70A until meter A deflects; zero beat fades out.

l. Regain zero beat by tuning DIAL B of LA-70A back and forth. Verify that MC stage is still locked.

m. Turn function selector switch of LA-70A to LOCK and adjust DIAL B to first lock point lower in frequency from beat note.

n. Regain zero beat by tuning DIAL C of LA-70A.

o. Record frequency as indicated by DIALS A, B, and C.

p. Repeat Steps (i) through (n) for the next lower zero beat of LA-70A.

q. The harmonic number for the frequency meter can be determined from the following formula:

$$N = \frac{F_2}{(F_2 - F_1)}$$

where: N = harmonic number of F_1
F_2 = highest zero beat
F_1 = next lower zero beat

r. The output frequency of the tuning unit is a product of the harmonic of the HP-540B times the harmonic of LA-70A, times the frequency as indicated by DIALS A, B, and C of LA-70A.

s. Verify that the frequency is within the 1% tolerance.

t. Repeat the above procedures at selected output frequencies of TUNING UNITS FIM-L, S, M, and X.

11.5 OUTPUT TRACKING ERROR AT ZERO DBM

a. Set the controls on the front panel of the FIM TUNING UNIT under calibration as follows:

MONITOR UNIT

METER SELECTOR switch	POWER MON.
POWER MONITOR meter function switch	ZERO SET
ZERO SET control	ZERO POWER

TUNING UNIT FIM-L

TUNING control	1 kmc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

TUNING UNIT FIM-S

TUNING control	2.14 mc
SIGNAL CALIBRATOR OUTPUT ATTENUATOR	0 dbm

TUNING UNIT FIM-M

TUNING control	4.2 kmc

SIGNAL CALIBRATOR OUTPUT ATTEN-UATOR	0 dbm

TUNING UNIT FIM-X

TUNING control	7.36 kmc
SIGNAL CALIBRATOR OUTPUT ATTEN-UATOR	0 dbm

b. Set Power Meter HP-430C front panel controls as follows:

COEF switch	NEG
RES switch	200
POWER RANGE switch	+5 dbm
ZERO SET COARSE control	fully CW
ZERO SET FINE control	fully CW

c. Connect Thermistor mount HP-477B to SIG GEN OUT of tuning unit. Connect a 3 foot length of RG-58/U cable from the Thermistor Mount HP-477B to BOLOMETER INPUT of Power Meter HP-430C as shown in Figure 11-2.

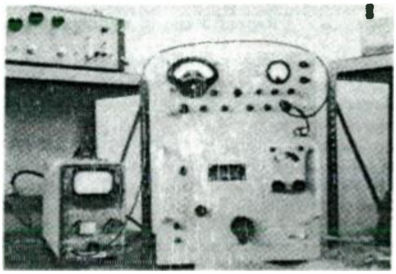


Figure 11-2. Output Tracking Error at Zero dbm Configuration

d. Set the HP-430C BIAS CURRENT switch to the first position which produces a CW off-scale meter indication.

e. Adjust HP-430C ZERO SET controls for a zero indication on the top meter scale.

f. Verify a ZERO SET condition on FIM MONITOR UNIT POWER MONITOR meter.

g. Switch the MONITOR UNIT function switch to CAL SIG GEN. Adjust POWER SET control on TUNING UNIT for a CAL-line POWER MONITOR meter indication.

h. Set TUNING UNIT SIGNAL CALIBRATOR OUTPUT ATTENUATOR for zero dbm.

i. Switch POWER MONITOR meter function selector switch from ZERO SET to CAL SIG GEN. Verify a ZERO SET and red-line POWER SET indication on POWER MONITOR meter.

j. Switch POWER MONITOR meter function selector switch to ZERO SET and ZERO SET power meter HP-430C.

k. Switch POWER MONITOR meter function selector switch to CAL SIG GEN and record the variation from zero dbm as indicated on HP-430C.

l. Repeat Steps (f) through (k) for each frequency of calibration.

m. Verify that the indications recorded in Step (k) are within tolerance.

n. Turn HP-430C ZERO SET control maximum counterclockwise. Set the HP-430C BIAS CURRENT switch to OFF and disconnect HP-477B from the tuning unit SIG GEN OUT. Turn HP-430C LINE POWER switch to OFF.

o. Plot a graph of output deviation from zero dbm as a function of frequency.

11.6 OUTPUT DIAL ATTENUATION TRACKING ERROR

a. Set controls on the front panel of the TUNING UNIT under calibration as follows:

MONITOR UNIT

Function Selector switch	AVERAGE
AFC switch	OUT
IF Gain control	10
METER SELECTOR switch	POWER MON
POWER MONITOR meter function selector switch	ZERO SET

TUNING UNIT FIM-L

TUNING control	1 kmc
SIGNAL CALIBRATOR OUTPUT ATTEN-UATOR	0 dbm

TUNING UNIT FIM-S

TUNING control	3 kmc
SIGNAL CALIBRATOR OUTPUT ATTEN-UATOR	0 dbm

TUNING UNIT FIM-M

TUNING control	5 kmc
SIGNAL GENERATOR OUTPUT ATTEN-UATOR	0 dbm

TUNING UNIT FIM-X

TUNING control	10 kmc
SIGNAL GENERATOR OUTPUT ATTEN-UATOR	0 dbm

POWER UNIT

VOLTMETER selector switch	BIAS
Klystron cathode current selector switch	Local Oscillator

b. Switch the FIM Power Unit VOLTMETER selector switch from BIAS to LOW VOLTAGE to REPELLER to BEAM. Verify a CAL-line indication on VOLTMETER for all switch positions.

c. Verify that the KLYSTRON CATHODE CURRENT meter on the FIM Power Unit is indicating between 20 and 30 ma.

d. Insert a 50-5 attenuator between the SIG GEN OUT connector of the tuning unit and the CAL IN connector, as shown in Figure 11-3.

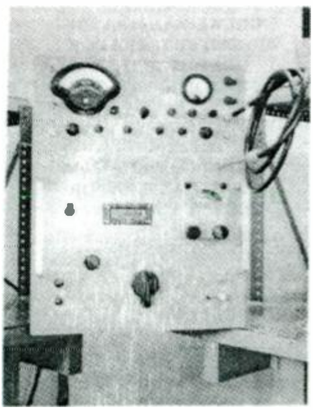


Figure 11-3. Output Dial Attenuation Tracking Error Configuration

e. Switch the POWER MONITOR meter function selector switch to CAL SIG GEN and tune in output signal by varying SIGNAL CALIBRATOR COURSE TUNING and SIGNAL CALIBRATOR FINE TUNING controls. Increase input attenuator to maintain an on-scale meter indication.

f. Switch POWER MONITOR meter function selector switch from ZERO SET to CAL SIG GEN, verify a ZERO SET and redline indication on POWER MONITOR and a SIGNAL CALIBRATOR OUTPUT ATTENUATOR indication of zero dbm.

g. Set the FIM MONITOR OUTPUT METER for a reference level by varying MONITOR UNIT IF GAIN control.

h. Remove the 50-5 attenuator from SIG GEN OUT connector of TUNING UNIT and connect the CAL IN connector of the MONITOR UNIT to the SIG GEN OUT connector of the TUNING UNIT by means of the 4 foot length of RG-9/U.

i. Decrease the SIGNAL CALIBRATOR OUTPUT ATTENUATOR control until the reference level originally set in Step (g) is indicated on the MONITOR UNIT OUTPUT METER.

j. Repeat Steps (c) through (i) until the OUTPUT ATTENUATOR dials of all TUNING UNITS have been calibrated from 5 to 90 db in steps of 5 db.

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