# SECTION III. DESIGN APPLICATION

## 3.1 Introduction

3.1.1 Purpose. The purpose of this section is to provide the necessary information for the consideration and implementation of the FCC system for electrical network interconnections.

The advantages, limitations, various design considerations, methods of implementation, special considerations, and mechanized design are included.

This section provides pertinent information for personnel responsible for selection and design of the interconnecting harness system to be used on a given program. The many advantages to be gained by FCC must be weighed against its limitations or restrictions early in the program, and an intelligent decision made on the type and extent of FCC system to be employed.

3.1.2 Comparison of FCC to RWC Systems. RWC interconnecting network systems have been used successfully for many years. Although higher-density connectors, thinner and better cable insulations, and high-strength alloy conductors are being used, the basic RWC harnesses generally still require individual conductor handling, identification, and termination. The individual conductors and insulation are subjected to potential concentrated strain and abrasion. The RWC system for interconnecting networks is generally limited to the size of conductors that can be used by mechanical and handling requirements.

The FCC systems, with conductors laminated between high-performance lamination sheets, can provide very small gages for minimum weight and space with maximum reliability and minimum cost. These are important factors which will influence the extensive use of FCC systems in the future.

3.1.2.1 Advantages of FCC System. Previous studies and system testing have indicated that up to 75 percent of the existing shielded cable on RWC systems can be replaced by nonshielded FCC by controlling the conductor registration in the harness runs. This is a major advantage of the FCC systems; however, in the comparisons that follow, this factor has not been included. Therefore, the actual advantages will be even greater than indicated. Also, all comparisons are made on equivalent conductor cross-section areas. In practice, the large surface areas of the FCC permit higher current densities (See Paragraph 3.2.3.1.3).

3.1.2.1.1 Weight Comparison. From previous studies, the weights for FCC supports and clamps are considerably less than those used with RWC. This saving results primarily from cable stacking and from simplification of clamping and supports for FCC.

The weight comparison for electrical connectors is dependent upon the connector system selected. In general, the FCC conductor-contact connector system, developed by NASA/MSFC, provides appreciable weight savings over the current, miniature, round connectors. These savings are realized primarily from the conductor-contact plug; however, it can be generalized that the weight of an FCC connector system will be equal to or less than the weight of an equivalent RWC connector system.

The cable, itself, provides the major weight savings for FCC system, especially in the smaller conductor cross-sections.

Figure 3-1 shows actual weight comparisons of three types of RWC with varying wall thicknesses and types of insulation, and FCC H/FEP standard and high-density cable per MIL-C-55543. The conductor area in square mils is plotted against weight in pounds per 1000 conductor feet to provide actual weight comparisons.

It can be concluded that appreciable weight savings can be achieved by the use of the FCC system and that the savings are greatest when the conductor cross-sections are minimized. Savings for shielded FCC are even greater than those for nonshielded FCC.

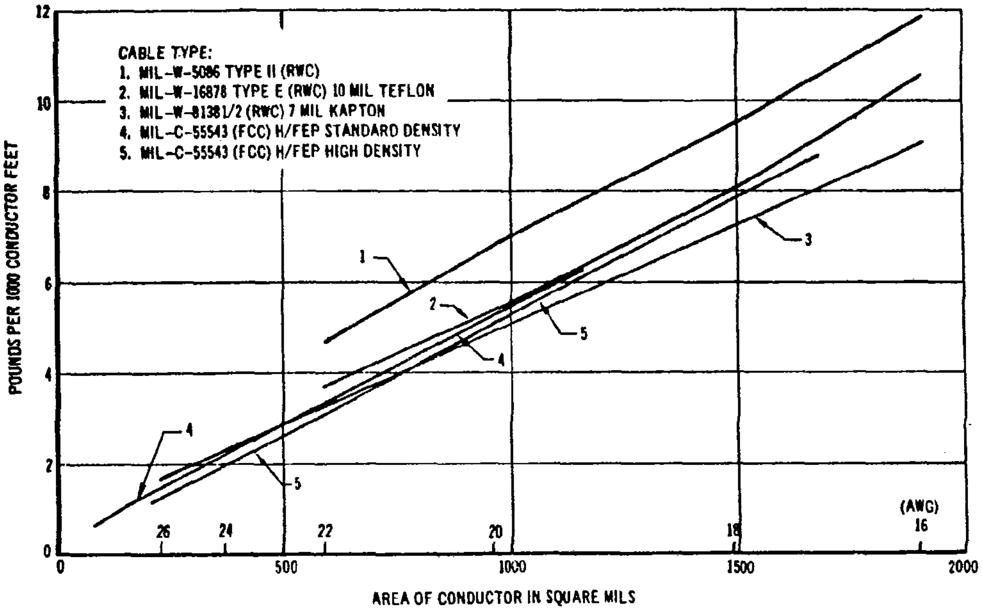



FIGURE 3-1. Nonshielded cable weight comparison.

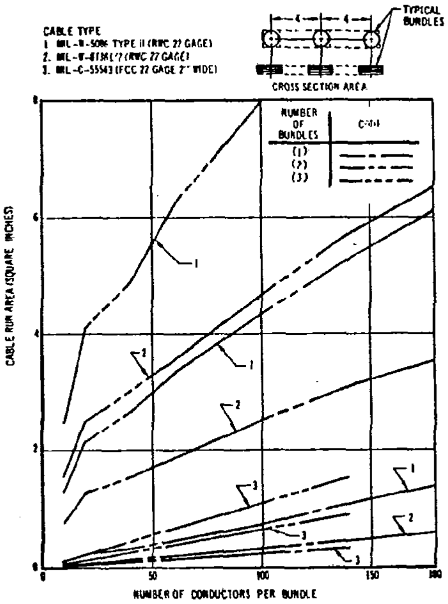



FIGURE 3-2. Nonshielded cable space comparison.

3.1.2.1.2 Space Comparison. Space comparisons have been made for various bundle run configurations and for the panel space required on electronic units for the installation of connectors.

Space saving for the FCC cable is inherent in its geometric cross-section and in the positive control of the locations of each conductor in the bundle runs which permits the use of fewer bundle runs.

Figure 3-2 shows actual comparisons of the bundle run areas required for one, two, and three bundles for two types of RWC and H/FEP FCC. It can be seen that the space saving is minimum for a single bundle and increases rapidly with the number of bundles. The space savings shown are particularly advantageous in tunnel and other congested areas where a minimum height cross-sectioned area is available for harness runs.

Appreciable connector mounting and handling area savings are achieved by the FCC connector system (Figs. 1-5 and 1-6).

It can be concluded that the FCC system offers major space savings in bundle runs and in panel mounting and handling areas.

3.1.2.1.3 Cost. To arrive at realistic cost comparisons, the cost of materials (including cable, connectors, supports, etc.) design, development, harness fabrication, and installation must be considered. Table 3-1 lists the cost comparison between FCC and RWC systems. The basis for many of the cost figures were taken from a study which included the development, fabrication, and installation of over 100 FCC harnesses of the MSFC conductor-contact system. Because of the many variables involved, certain general assumptions and conclusions have been made as follows:

a. To realize the cost savings indicated, FCC must be applied early in the program. Studies pointed out the additional costs of redesign, including those for design, development, and requalification.

b. The major cost saving for the FCC system is the recurring harness fabrication cost which is reduced 80 percent. This saving is realized by handling, identifying, and terminating all conductors in each plug layer, simultaneously.

c. The cost of the connectors will be approximately equal if the pin-and-socket-type FCC connector is used. The MSFC conductor-contact FCC connector will provide a cost saving of approximately 35 percent for a mated pair.

d. The cost saving of FCC increases as the conductor cross-sections decreases.

e. The cost of supports and clamps for the FCC system will generally be substantially less because of the simplicity of installations and reduction of parts.

Figure 1-7 gives cost comparisons for a 2-inch-wide H/FEP FCC cable and Kapton-insulated alloy conductor RWC, per MIL-W-81381/2. An FCC predicted cost of $1.50 per linear foot for nonshielded and $3.00 for shielded cable was used. The cost of the RWC is based upon CY 1968 prices quoted for 25,000 linear foot lengths. The advantage trend for FCC in the smaller conductor cross section is maintained.

In general, it can be stated that substantial cost savings can be realized with the FCC system. The actual savings will depend upon the program requirements and the application techniques employed.

TABLE 3-1. COST COMPARISON — FCC VERSUS RWC SYSTEMS

| Item | Subitem of Major Item (%) | FCC Cost Saving (%) | |
|---|---|---|---|
| | | Subitem | Major Item |
| Engineering | | | -5 |
| System | 25 | -10 | |
| Harness Layout | 25 | -10 | |
| Production Drawings | 25 | 0 | |
| Schematics, etc. | 25 | 0 | |
| Development | | | 20 |
| Materials | | | 28 |
| Cable | 40 | 0 | |
| Connectors | 40 | 35 | |
| Clamps | 5 | 50 | |
| Supports | 15 | 75 | |
| Harness Fabrication | | | 80 |
| Harness Installation | | | 40 |

Notes:
1. Percentages listed above are average and can vary from program to program.
2. Subitem percentages have been listed from typical space system programs.
3. Cost savings for all major items except engineering have been influenced heavily by the Reference 1-1 study; however, cost comparison reports by other major contractors verify these percentages.
4. To establish actual program percent or dollar savings would require the establishment of dollar costs for each of the major items listed above.
5. The recurring harness fabrication cost saving of 80 percent is realistic for all programs properly utilizing FCC systems, and will contribute heavily to its future use.

3.1.2.1.4 System Performance. System performance, through the use of FCC interconnecting harnesses, can be greatly improved. The FCC configurations can be selected to provide the electrical characteristics actually required by the system. Past routine selection of twisted, shielded, and other special RWC configurations has resulted in expensive, overly designed, and overweight systems requiring excessive space for installation. The positive placement control of each conductor circuit in its cable and in the bundle runs assures predictable performance plus repeatability from unit to unit.

3.1.2.1.5 Others. Other advantages of FCC include:

a. Increased mechanical strength provided by high-performance, laminated insulation layers or sheets.

b. Permits identification, termination, and handling of cable by layers, rather than individual conductors.

c. Provides many special application possibilities as explained in Paragraph 3.2.8.

3.1.2.2 Limitations in Use of FCC System. It is very important that the personnel responsible for the program application of FCC understand the current status of the system being considered, and the inherent characteristics that differ from the RWC system that has been used routinely for such a long time.

3.1.2.2.1 Hardware Availability. In the past, FCC and FCC connectors, qualified to military specifications acceptable to government and prime contractor agencies, were not available. Now, triservice coordinated military specifications, MIL-C-55543 for nonshielded FCC and MIL-C-55544 for environmental shielded and nonshielded FCC connectors, are available. It is planned to release a military specification for shielded FCC cable in the near future. These specifications will provide the prime contractors and vendors with the guiding requirements for the development and qualification of FCC systems which can be used for general application on future programs.

Many existing programs have used FCC systems utilizing cable to NASA/MSFC, NAS, or company specifications with existing connectors used as is, or modified as required. The applications, which have been used so successfully, should be understood and considered for all future programs prior to the availability of a complete line of cable and connectors qualified to the military specifications.

Extensive evaluation and qualification tests to the requirements of MSFC-SPEC-220A have been successfully performed by and for NASA/MSFC on various FCC configurations. Qualification tests per MSFC-SPEC-219 have been successfully performed by two independent agencies on the NASA/MSFC conductor-contact connector system. Extensive FCC evaluation testing has been performed per NAS729 by USAEC of Ft. Monmouth, N.J. Since the performance requirements of MIL-C-55543 and MIL-C-55544 are essentially in accordance with the above listed specifications, it is anticipated that little difficulty will be encountered in qualification to the new specifications.

A complete line of military specification NASA/MSFC conductor-contact connectors, utilizing molded and premolded plugs, has been tooled and is currently available, with contacts on 75-mil centers. Additional effort is required for other centerline spacings, final production tooling, and formal qualification. One size single-layer pin-and-socket, shielded-type, military specification connector with 100-mil spacing has been developed and is being evaluation tested. Additional connector sizes and center spacings are required with qualification.

Studies have indicated all other hardware required for general FCC application is available, with the exception of a suitable production, high-reliability, distribution-unit system. The requirements for these units will be discussed in Paragraph 3.2.5 and design concepts will be covered in Section IV.

3.1.2.2.2 Harness Electrical Flexibility. The current RWC system permits random registration of conductors to connectors. Pin-assignment changes on connectors with removable crimp contacts can be readily made after initial connector assembly. Soldered and potted connectors are readily adaptable to pin-assignment design changes accomplished on new harness fabrication. The flexibility of the RWC system has resulted in the licensing, by all phases of engineering, of their right to avoid final harness pin assignments until very late in the development of the prototype or first article unit. The black-box designer was permitted to make wiring and pin-assignment changes, the system designer could change the systems' interconnections, and the harness designer made the final pin-assignment changes late in the development stages. This flexibility was provided at hidden costs which were more or less taken for granted. First, all major harness-run drawings required many engineering changes; 50 or more is common for todays complicated space systems. These engineering changes required costly planning and manufacturing changes. Second, this system provides random conductor registration with no assurance of electrical repeatability from unit to unit.

The FCC systems using the military specification connectors cannot make individual pin-assignment changes at the connectors on the FCC side. Although cable segments can be reversed and relocated, and pin-assignment-change devices can be incorporated into initial design, or added later as required, the harness pin-assignment flexibility is limited without distributors. The FCC system with distributors assures minimum harness drawing changes, permits registration control for all conductors, and provides the maximum interconnecting wiring network flexibility.

3.1.2.2.3 Terminating with Round Connectors. The termination of FCC to round connectors will be required when mating to existing or black boxes having round connectors, or when round connectors are used to provide pin-assignment flexibility. Several systems have been developed for this termination. The first shown in Figure 3-3 provides pin assignment changes, and the second shown in Figure 3-4 essentially has predetermined pin assignments. One method of manufacturing the Figure 3-3 transition is covered in Section VI. All methods add additional circuit joints and reduce the major harness cost saving afforded by layer termination to the connectors. The Figure 3-4 method should be used only for very special applications.

3.1.2.2.4 Multiple-Plane Bending. FCC permits very small bend radius and efficient multiplane bending and routing as described in Paragraph 3.2.6. However, the planar nature of FCC could present problems in some special routing requirements. An example would be of routing through conduit with bends in multiple planes.

3.1.2.2.5 Familiarization and Training

3.1.2.2.5.1 Design. New design philosophy, application techniques, and drawing preparation are required for the successful application of the FCC system. It is the primary purpose of Section III to present information for the design and application of the FCC system to electrical interconnecting networks.

3.1.2.2.5.2 Manufacturing. Manufacturing facilities producing and installing FCC systems must establish the required capability (including equipment), facility space, procedures, and trained personnel. It is the primary purpose of Section VI to detail these requirements for a typical FCC system.

3.1.2.2.5.3 Quality Control. It is important to establish the proper quality control on new production technologies. Section VII gives an extensive quality control treatise on receiving inspection for incoming cable and connectors, and for subsequent manufacturing processes for cable assembly and installation of FCC systems.

3.1.3 Conclusions. It can be concluded that the FCC system offers many advantages for future application of electrical interconnecting networks. A thorough understanding of the advantages which can be achieved, and the effort which must be expended, is necessary before FCC application can be intelligently considered. The information contained in Section III should be thoroughly reviewed, prior to proceeding with FCC application.

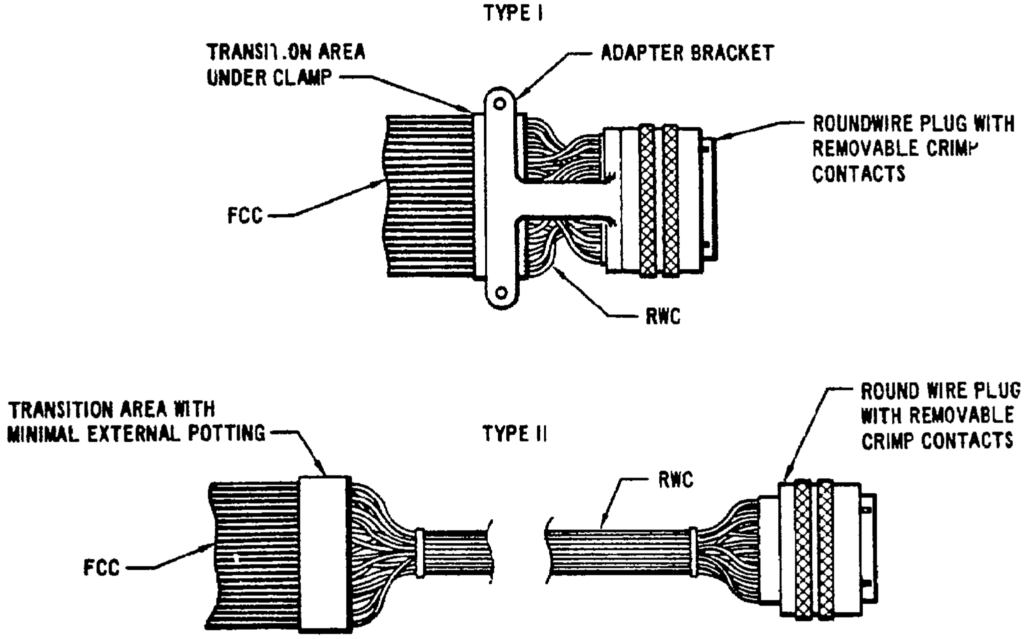



FIGURE 3-3. FCC to round connectors - with pin-assignment changes.

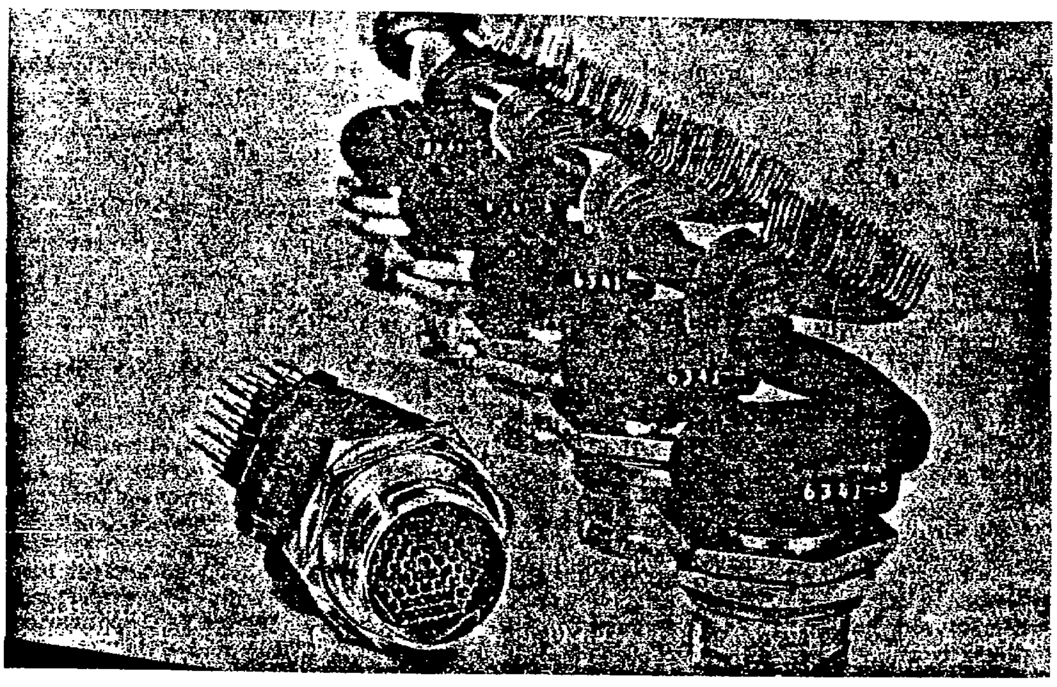


FIGURE 3-4. Flat to round connectors - without pin-assignment changes (Ansley West).

## 3.2 FCC Application

3.2.1 Program Considerations. Certain questions must be asked, and considerations given early in the program to determine the feasibility of utilizing the FCC system for electrical interconnecting networks.

3.2.1.1 Program Requirements. How important are weight and space savings? What are the schedule requirements? Will the program electronic units be newly designed, or will existing units or those common with other programs be utilized? What quantity of program end items will be required? What are the conductivity and special wiring characteristic requirements?

3.2.1.2 Approval for Use. Does the customer procuring document require, encourage, or permit the use of the FCC system? With the FCC and FCC connector triservice military specification issuances, MIL-C-55543 and MIL-C-55544, this customer approval is easier to obtain. If the program can benefit substantially by using FCC, then a request for approval, listing the program advantages, should be acceptable.

3.2.1.3 Company Capability. What is the company current FCC capability in design, manufacturing, and quality control? Does the magnitude of the current program or future proposed programs warrant establishing the capability if it does not exist? If a major contractor expects to remain active and successful in future programs, he would do well to review Figure 1-7 and Table 1-1 before answering this question.

3.2.1.4 Preliminary Selection of FCC system. After proper consideration is given to those items listed above, an early program decision should be made on the areas for FCC application consideration.

3.2.2 Network System Requirements. With the decision made to use FCC on the program, the various system requirements should be analyzed and defined to establish a basis for the design requirements.

3.2.2.1 System Definition. All electronic system requirements should be established that include a definition of the number and type of subsystems, their performance requirements, and their mutual interactions.

3.2.2.2 Electrical Interface Requirements. Electrical interfaces between end items and associated equipment (umbilicals, flight disconnects, stage joining, etc.) must be established for the number and type of circuits, the interfacing connectors, and for their physical location. To obtain the maximum advantages from FCC interconnecting networks, it is highly desirable to utilize FCC interface connectors in locations best suited for FCC routing and support.

3.2.2.3 Environmental Requirements. The system environmental requirements should be compared with those of the proposed FCC system procurement specifications. In general, these specifications cover very rigid environmental requirements. The geometry and inherent characteristics of FCC make it adaptable to meeting high-vibration, flexing cleanliness, and other requirements. However, special requirements, such as hermetically sealed connectors, cryogenic temperatures, extremely high temperatures, and exposure to nuclear radiation, should be considered.

3.2.3 Design Considerations. The design of FCC interconnecting harnesses requires careful consideration of the electrical characteristics of FCC plus the design and implementation of the selected system.

3.2.3.1 FCC Electrical Characteristics

3.2.3.1.1 Introduction. The FCC electrical characteristics useful to the designer are discussed in the following paragraphs. Some characteristics, such as resistance and current-carrying capability, can be used directly by the designer to select the required conductor cross-sections. Other characteristics can be used for more complicated considerations in determining the FCC component compatibility in meeting the overall requirements for proper system functions.

## TABLE 3-2. FCC CABLE SELECTION CHART-CONDUCTOR SIZE AND RESISTANCE

| Centerline Spacing (in.) | Conductor Configuration | Conductor Size (in) | | Equiv AWG Size[a] | Square Mils | Nominal Resistance Ohms per 1000 ft at 20°C | Resistance Tolerance in Ohms (±) |
|---|---|---|---|---|---|---|---|
| | | Width (in.) | Thickness (in.) | | | | |
| 0.050 | Standard | 0.025 | 0.003 | 30 | 75 | 140 | 15.0 |
| | | | 0.004 | 29 | 100 | 98 | 7.8 |
| | | | 0.005 | 28 | 125 | 76 | 4.9 |
| | High Density | 0.040 | 0.003 | 28 | 120 | 87 | 5.8 |
| | | | 0.004 | 27 | 160 | 61 | 3.1 |
| | | | 0.005 | 26 | 200 | 47 | 1.9 |
| 0.075 | Standard | 0.040 | 0.003 | 28 | 120 | 87 | 5.8 |
| | | | 0.004 | 27 | 160 | 61 | 3.1 |
| | | | 0.005 | 26 | 200 | 47 | 1.9 |
| | High Density | 0.065 | 0.003 | 26 | 195 | 53 | 2.2 |
| | | | 0.004 | 25 | 260 | 38 | 1.2 |
| | | | 0.005 | 24 | 325 | 29 | 0.72 |
| 0.100 | Standard | 0.065 | 0.004 | 25 | 260 | 38 | 1.2 |
| | | | 0.005 | 24 | 325 | 29 | 0.84 |
| | | | 0.006 | 23 | 390 | 24 | 0.50 |
| | High Density | 0.090 | 0.004 | 24 | 360 | 27 | 0.59 |
| | | | 0.005 | 23 | 450 | 21 | 0.38 |
| | | | 0.006 | 22 | 540 | 17 | 0.25 |
| 0.150 | Standard | 0.115 | 0.004 | 22 | 460 | 21 | 0.36 |
| | | | 0.005 | 22 | 575 | 16 | 0.22 |
| | | | 0.006 | 21 | 690 | 13 | 0.16 |
| | High Density | 0.140 | 0.004 | 22 | 560 | 18 | 0.25 |
| | | | 0.005 | 21 | 700 | 14 | 0.15 |
| | | | 0.006 | 20 | 840 | 11 | 0.11 |

Note: Only the cable configurations shown in MIL-C-55543 are shown.
For resistance tolerance explanation see Paragraph 3.2.3.1.2.

a. Shown graphically on Figure 3-5.

3.2.3.1.2 Resistance. The resistance of an electrical conductor is a function of its material, cross-section, temperature, and length.

The nominal resistances in ohms per 1000 feet, for copper conductors for the various MIL-C-55543 conductor cross-sections, are given in Table 3-2. Figure 3-5 shows various FCC cross sections for equivalent AWG sizes 20 through 30.

The tolerances for the conductor width and the thickness of conductors provide a ± tolerance to the nominal cross-section. Using the dimensional tolerances listed in MIL-C-55543, the last column of Table 3-2 lists the resistance tolerances for the various conductor configurations.

The resistance change with temperature change is a function of the temperature coefficient of the conductor material which is, in this case, annealed copper. Conversions must be made in determining resistance for conductors at temperature other than 20°C. This may be done using the formula

$$R_T = R_{20°C} \left[1 + \alpha_T (T - 20°)\right]$$

where

$R_T$ = resistance at a revised temperature T

$R_{20°}$ = resistance at a temperature of 20°C

T = revised temperature (°C)

$\alpha_T$ = temperature coefficient of resistivity as a function of T, as given by Knowlton.

The resistance temperature correction factors $[1 + \alpha_T (T - 20°)]$ are given in Figure 3-6. Multiplying the resistance values listed in Table 3-2 by the correction factors given in Figure 3-6 will give the resistance at temperatures other than 20°C.

3.2.3.1.3 Current-Carrying Capacity. The current-carrying capacity of conductors in a cable is primarily a function of the temperature rise that can be tolerated in the cable. This, in turn, is dependent on the density of power dissipation from ohmic losses in the conductor, which result from current levels and conductor resistance. Temperature rise is also dependent on the medium for conducting heat away from the cable, and on the cross-sectional shape of the cable. This shape determines the amount of surface area on the outside of the cable for heat dissipation.

The geometry of FCC lends itself uniquely to effective surface-heat dissipation. The cross-sectional surface of an FCC is significantly greater than the surface of an RWC bundle, when both have the same aggregate of cross-sectional conductor area (Fig. 3-7).

The net effect of the surface difference allows a greater current-carrying capacity for FCC. This fact is demonstrated by a NASA study in comparative data of a 2-inch-wide cable of 25 0.004- by 0.040-inch conductors (27-gage equivalent), and a 25-wire bundle of 26- or 28-gage round wires. See Figures 3-8 and 3-9 for the data for both air and vacuum surroundings.

Additional data are available for FCC, for both single cables, and for 3- and 10-stacked cables of 0.004- by 0.040-inch conductors with 0.075-inch centerline spacing, both vacuum and air. These data are illustrated in Figures 3-10 and 3-11, as maximum allowable current for a stated maximum temperature rise in the hottest conductor of the cable(s).

To apply these data to FCC of other conductor configurations, use the following correction formula:

$$I \text{ (new cable)} = K_C \cdot I \text{ (0.004 x 0.040 x 0.075 centerline spacing)},$$

where

I (new cable) = allowable current for each conductor of the new cable

I (0.004 x 0.040 x 0.075 centerline spacing) = allowable current as given in Figures 3-10 or 3-11 for air or vacuum as appropriate.

$K_C$ = configuration factor as tabulated in Table 3-3.

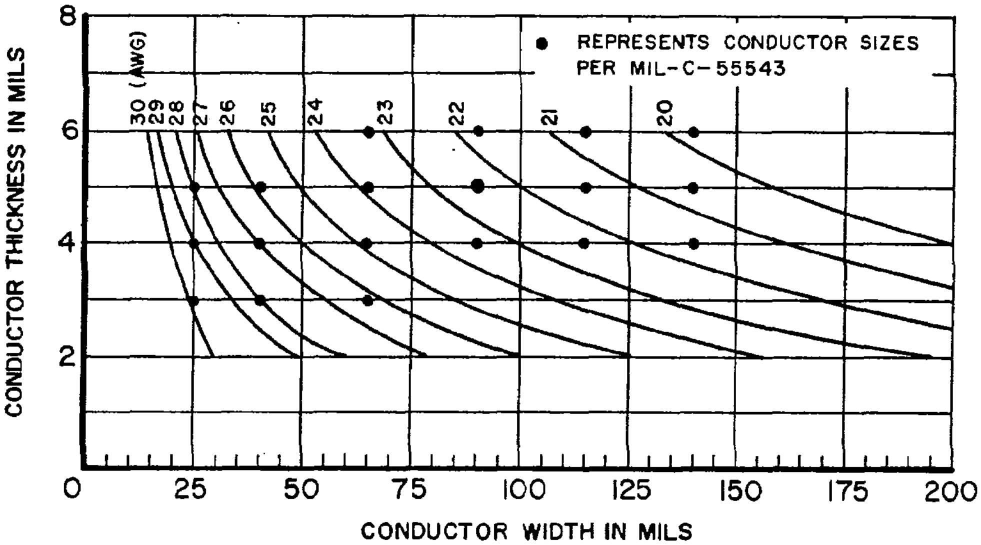



FIGURE 3-5. FCC cross-sections - equivalent AWG 20 through 30.

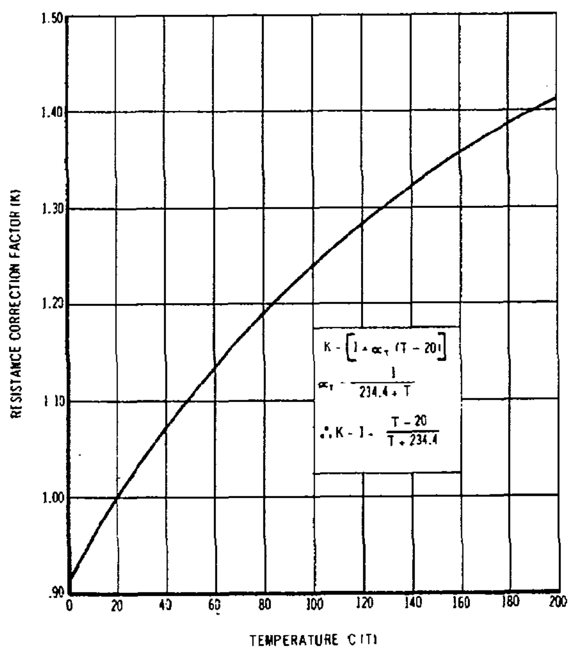



FIGURE 3-6. Resistance correction factor.

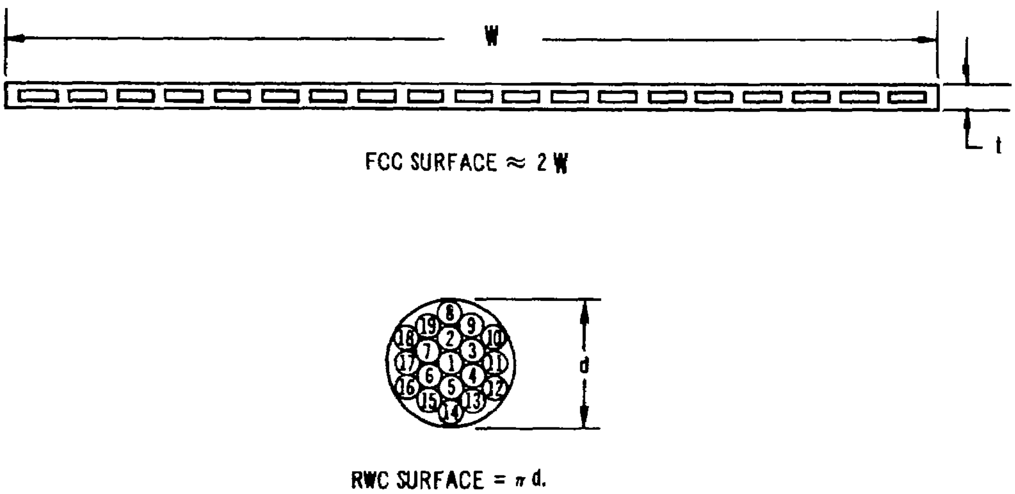



FIGURE 3-7. Surface cross-section comparison - FCC to RWC.

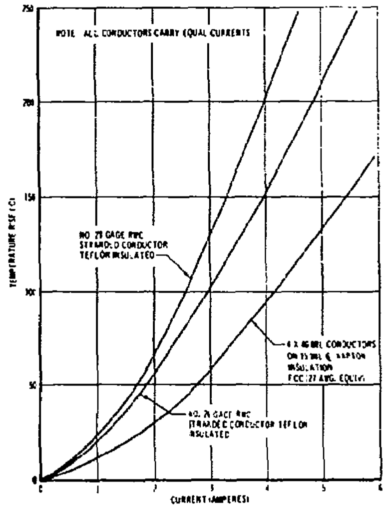



FIGURE 3-8. Current versus temperature rise in hottest conductor in air (25 conductors).

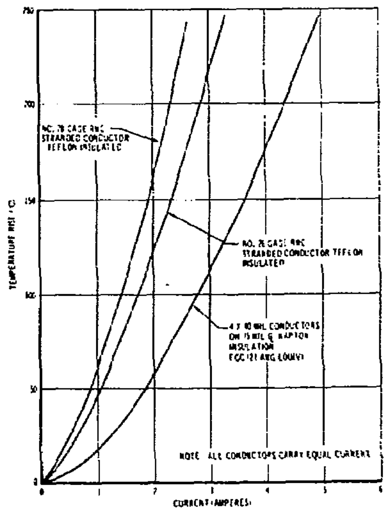



FIGURE 3-9. Current versus temperature rise in hottest conductor in vacuum (25 conductors).

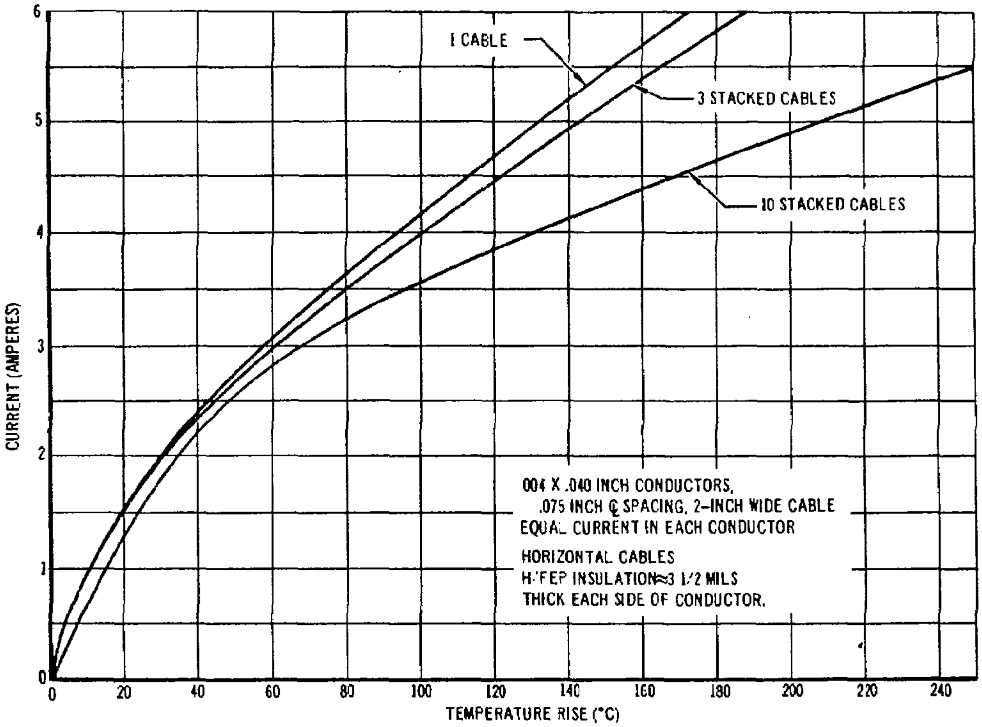



FIGURE 3-10. Maximum current versus temperature rise in hottest conductor in air.

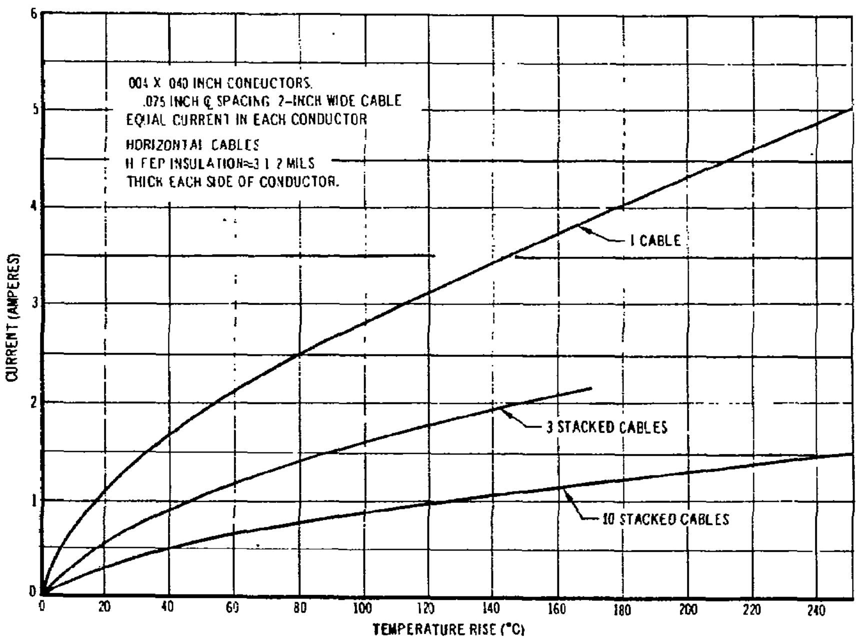



FIGURE 3-11. Maximum current versus temperature rise in hottest conductor in vacuum.

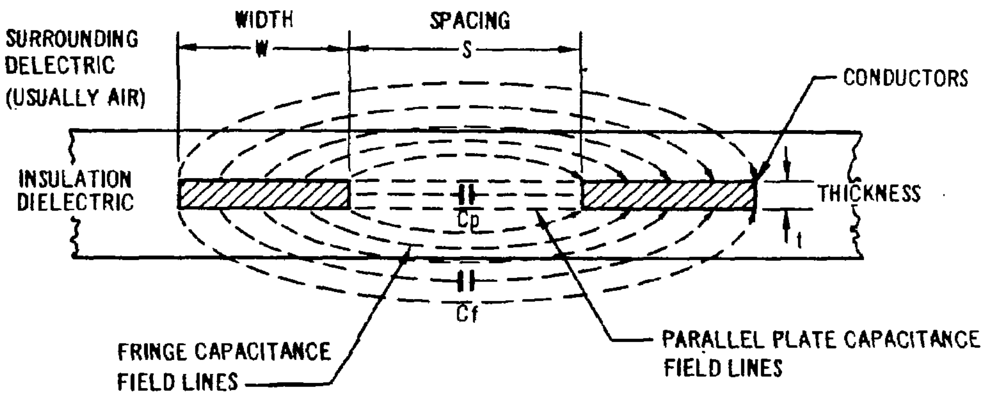



FIGURE 3-12. Distributed capacitance between conductors.

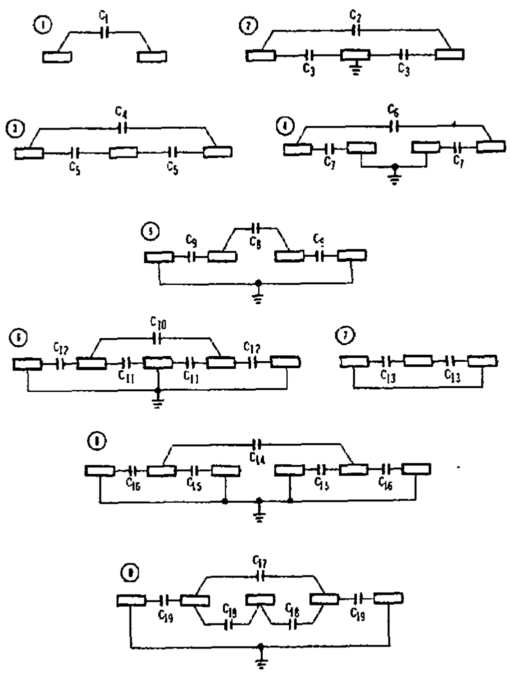


FIGURE 3-13. Inner conductor capacitance configurations for unshielded flat cable.

This formula and Table 3-3 correction factors are based on the assumption that the maximum conductor temperature for the same type and thickness of FCC insulation is dependent on the $I^2R$ losses of the conductors per unit of cable width and length; therefore, FCC cables with equal $I^2R$ losses per unit width and length will have equal maximum conductor temperatures.

3.2.3.1.4 Capacitance Data. Capacitance per unit length between conductors is a function of the cross-sectional geometry of the cable, the dielectric material, the shielding configuration, and the mounting technique. Capacitance between adjacent conductors consists of parallel plate capacitance and fringing capacitance. Parallel plate capacitance is usually small relative to the total capacitance, except when the conductors are very closely spaced.

For an unshielded cable, the fringing field extends into the air outside the insulation (Fig. 3-12). Calculation of exact capacitance values is difficult, since the capacitance depends on thickness and dielectric constant of the insulation, dielectric constant of the surrounding medium (usually air), widths and thicknesses of the conductors, spacing of the conductors, presence of nearby grounds, and operating frequency of the circuit. One approach which can be used to approximate distributed capacitance is to establish a "composite dielectric" value which falls between that of the insulation material and the air.

Tables 3-4 and 3-5 provide standard flat-cable capacitance values for those configurations shown in Figure 3-13. These tables were compiled by using calculated values and all available experimental data. It should be pointed out to the designer that manufacturing tolerances for insulation thickness, conductor spacing, and dielectric material play an important part in the capacitance values. The range of values found in Tables 3-4 and 3-5 are the values which should be anticipated by the designer because of manufacturing tolerances and variations in measuring techniques. These tables are for isolated conductors not influenced by ground planes or close conductors. Table 3-6 shows calculated capacitance values for stacked FCC conductors.

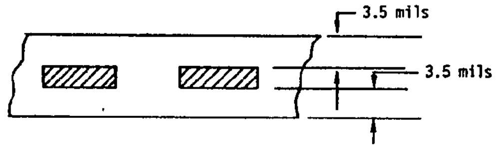
The use of electrical shields permits reduction of capacitive cross-coupling and minimization of conducted and radiated electro-magnetic interference. The flat-cable geometry lends itself to effective shielding. Use of a grounded shield, bonded to one or two sides of a cable, lowers the direct conductor-to-conductor capacitance. Suitable shielding and grounding arrangements can be used to reduce the direct capacitance in various configurations of FCC. For example, the capacitance of approximately 6.5 pf/ft for $C_1$ of Figure 3-13 for an unshielded cable with conductors 4 by 40 mils spaced on 75 mil centers and $\epsilon$ equal to 2.9 (Table 3-5) can be reduced to approximately 0.051 pf/ft by adding a double shield as shown in Configuration 1 of Figure 3-14. Shielding material may be foil, wire mesh, or, in some cases, vacuum-deposited aluminum, copper, or silver. The foil used may be solid, perforated, wrinkled, or expanded. For the double-shielded cable, the field is contained within the shields, with a minimal field, if any, ever extending beyond the shield. Table 3-7 provides shielded flat-cable capacitance values for those configurations shown in Figure 3-14. These tables were compiled by using calculated values and all available experimental data. The range of values found in Table 3-7 are values which the designer should anticipate because of manufacturing tolerances and variations in measuring techniques. The tolerances used for compiling Table 3-7 were: ±0.5 mil on the distance between the surface of conductor and the shield (this variation assumed to occur on either side of conductor simultaneously for analysis); ±2.0 mils on the conductor width; ±0.4 mil on conductor thickness; and ±2.0 mils on conductor separation for high-density cable; and ±5.0 mils on conductor separation for standard cable. From a magnitude standpoint, the distance between the conductor and shield had the greatest impact. Since the conductor and shield are relatively large area plates closely spaced, small changes in distance result in sizeable variations of capacitance; for example, it would be expected that the capacitance $C_2$ (Fig. 3-14) for a 40- by 4-mil cable would go from 186 to 286 pF/ft to 163 to 360 pF/ft if the conductor-to-shield tolerance were increased to ±1.0 mil with all other tolerances remaining the same.

For cables other than the standard cables listed in Table 3-7, the conductor-to-shield capacitance can be determined by using the curves in Figures 3-15 through 3-19 [3-1]. If the geometry of the cable is known, $C_{pp}$ is obtained from Figures 3-15 through 3-18. The total fringe capacitance, which is four times $C_f$, is obtained from Figure 3-19. The sum of $C_{pp}$ and $4C'_F$ is equal to the total conductor-to-shield capacitance. The capacitance between adjacent conductors can be expected to decrease to approximately 1/45 to 1/150 of the equivalent value for an unshielded cable.

TABLE 3-3. CONFIGURATION CORRECTION FACTORS FOR CURRENT RATINGS

| Cable Configuration (in.) | | | Configuration Correction Factor $K_C$ |
|---|---|---|---|
| Centerline Spacing | Width | Thickness | |
| 0.050 | 0.025 | 0.003 | 0.56 |
| 0.050 | 0.025 | 0.004 | 0.64 |
| 0.050 | 0.040 | 0.003 | 0.71 |
| 0.050 | 0.040 | 0.004 | 0.82 |
| 0.050 | 0.040 | 0.005 | 0.91 |
| 0.075 | 0.040 | 0.003 | 0.87 |
| 0.075 | 0.040 | 0.004 | 1.00 |
| 0.075 | 0.040 | 0.005 | 1.12 |
| 0.075 | 0.065 | 0.003 | 1.10 |
| 0.075 | 0.065 | 0.004 | 1.27 |
| 0.075 | 0.065 | 0.005 | 1.42 |
| 0.100 | 0.065 | 0.004 | 1.47 |
| 0.100 | 0.065 | 0.005 | 1.65 |
| 0.100 | 0.065 | 0.006 | 1.80 |
| 0.100 | 0.090 | 0.004 | 1.73 |
| 0.100 | 0.090 | 0.005 | 1.94 |
| 0.100 | 0.090 | 0.006 | 2.12 |
| 0.150 | 0.115 | 0.004 | 2.36 |
| 0.150 | 0.115 | 0.005 | 2.68 |
| 0.150 | 0.115 | 0.006 | 2.72 |
| 0.150 | 0.140 | 0.004 | 2.65 |
| 0.150 | 0.140 | 0.005 | 2.96 |
| 0.150 | 0.140 | 0.006 | 3.24 |

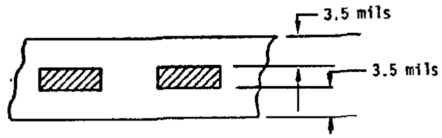
# TABLE 3-4. CAPACITANCE VALUES IN pF/ft FOR MYLAR INSULATED FLAT CABLE ($\epsilon = 2.5$) (SEE FIGURE 3-13 FOR CAPACITANCE SYMBOL DESIGNATION)




Note: 4 X 40 mil conductors, spaced on 75 mil center-line spacings. See Figure 3-13 for $C_1$ - $C_{19}$ measurement.

| Centerline Spacing (in.) | Conductor Separation (in.) | Conductor Width (in.) | Conductor Thickness (in.) | $C_1, C_3, C_5$ | $C_2$ | $C_4$ | $C_6$ | $C_7$ | $C_8$ | $C_9$ | $C_{10}$ | $C_{11}, C_{12}, C_{13}, C_{15}, C_{16}$ | $C_{14}$ | $C_{17}$ | $C_{18}$ | $C_{19}$ |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0.050 | 0.025 | 0.025 (Std) | 0.003 | 5.3- 7.6 | 1.2-1.7 | 3.3-4.8 | 0.7-1.1 | 5.7- 8.2 | 4.3- 6.1 | 7.4-10.7 | 0.5-0.8 | 10.6-15.2 | 0.2-0.4 | 2.7-4.0 | 5.1- 7.6 | 6.7-10.0 |
| | | | 0.004 | 5.4- 7.7 | 1.2-1.8 | 3.4-4.9 | 0.8-1.1 | 5.8- 8.3 | 4.3- 6.2 | 7.6-10.8 | 0.5-0.8 | 10.8-15.4 | 0.2-0.4 | 2.8-4.0 | 5.3- 7.6 | 7.0-10.0 |
| 0.050 | 0.010 | 0.040 (Hi Den) | 0.003 | 9.5-12.3 | 1.4-1.8 | 5.5-7.1 | 0.9-1.1 | 10.2-13.2 | 7.6- 9.8 | 13.3-17.2 | 0.6-0.8 | 19.0-24.5 | 0.3-0.5 | 3.9-5.0 | 7.4- 9.5 | 11.3-14.5 |
| | | | 0.004 | 9.8-12.6 | 1.5-1.9 | 5.7-7.3 | 0.9-1.1 | 10.5-13.5 | 7.8-10.1 | 13.7-17.6 | 0.6-0.8 | 19.8-25.2 | 0.3-0.5 | 4.0-5.2 | 7.6- 9.9 | 11.6-15.0 |
| | | | 0.005 | 10.1-12.9 | 1.5-1.9 | 5.8-7.5 | 0.9-1.2 | 10.8-13.8 | 8.1-10.3 | 14.1-18.0 | 0.6-0.8 | 20.2-25.8 | 0.3-0.5 | 4.1-5.3 | 7.8-10.0 | 11.9-15.4 |
| 0.075 | 0.035 | 0.040 (Std) | 0.003 | 5.3- 7.0 | 1.3-1.8 | 3.5-4.7 | 0.8-1.1 | 5.7- 7.5 | 4.3- 5.6 | 7.4- 9.8 | 0.5-0.7 | 10.6-14.0 | 0.2-0.4 | 3.2-4.2 | 6.1- 8.0 | 8.0-10.5 |
| | | | 0.004 | 5.4- 7.1 | 1.3-1.8 | 3.6-4.8 | 0.8-1.1 | 5.8- 7.6 | 4.3- 5.7 | 7.5-10.0 | 0.5-0.7 | 10.8-14.2 | 0.2-0.4 | 3.3-4.3 | 6.3- 8.2 | 8.3-10.8 |
| | | | 0.005 | 5.4- 7.1 | 1.3-1.8 | 3.6-4.8 | 0.8-1.1 | 5.8- 7.6 | 4.3- 5.7 | 7.5-10.0 | 0.5-0.7 | 10.8-14.2 | 0.2-0.4 | 3.3-4.3 | 6.3- 8.2 | 8.3-10.8 |
| 0.075 | 0.010 | 0.065 (Hi Den) | 0.003 | 10.0-12.7 | 1.5-1.9 | 5.8-7.4 | 0.9-1.1 | 10.7-13.6 | 8.0-10.2 | 14.0-17.8 | 0.6-0.8 | 20.0-25.3 | 0.3-0.5 | 4.1-5.2 | 7.8- 9.7 | 11.9-15.2 |
| | | | 0.004 | 10.3-13.0 | 1.5-2.0 | 6.0-7.5 | 0.9-1.2 | 11.1-14.0 | 8.2-10.4 | 14.4-18.2 | 0.6-0.8 | 20.6-26.0 | 0.3-0.5 | 4.2-5.3 | 8.0- 9.8 | 12.2-15.4 |
| | | | 0.005 | 10.6-13.3 | 1.6-2.0 | 6.1-7.7 | 0.9-1.2 | 11.4-14.2 | 8.5-10.6 | 14.8-18.6 | 0.6-0.8 | 21.2-26.5 | 0.3-0.5 | 4.3-5.5 | 8.2-10.4 | 12.5-16.0 |
| 0.100 | 0.035 | 0.065 (Std) | 0.004 | 6.0- 7.7 | 1.5-1.9 | 4.0-5.2 | 0.9-1.2 | 6.4- 8.3 | 4.8- 6.2 | 8.4-10.8 | 0.6-0.8 | 12.0-15.4 | 0.3-0.5 | 3.6-4.6 | 6.8- 8.7 | 9.0-11.5 |
| | | | 0.005 | 6.0- 7.7 | 1.5-1.9 | 4.0-5.2 | 0.9-1.2 | 6.4- 8.3 | 4.8- 6.2 | 8.4-10.8 | 0.6-0.8 | 12.0-15.4 | 0.3-0.5 | 3.6-4.6 | 6.8- 8.7 | 9.0-11.5 |
| | | | 0.006 | 6.2- 7.9 | 1.6-2.0 | 4.2-5.3 | 0.9-1.2 | 6.6- 8.5 | 5.0- 6.3 | 8.7-11.0 | 0.6-0.8 | 12.4-15.8 | 0.3-0.5 | 3.7-4.8 | 7.0- 9.1 | 9.2-12.0 |
| 0.100 | 0.010 | 0.090 (Hi Den) | 0.004 | 10.6-13.3 | 1.6-2.0 | 6.1-7.7 | 1.0-1.2 | 11.4-14.2 | 8.5-10.6 | 14.8-18.6 | 0.6-0.8 | 21.2-27.2 | 0.3-0.5 | 4.4-5.4 | 8.3-10.2 | 12.8-15.6 |
| | | | 0.005 | 11.0-13.7 | 1.7-2.1 | 6.4-7.9 | 1.0-1.2 | 11.8-14.7 | 8.8-11.0 | 15.4-19.2 | 0.7-0.9 | 22.0-27.4 | 0.3-0.6 | 4.5-5.6 | 8.5-10.6 | 13.0-16.2 |
| | | | 0.006 | 11.3-14.0 | 1.7-2.1 | 6.6-8.1 | 1.0-1.3 | 12.0-15.0 | 9.0-11.2 | 15.8-19.6 | 0.7-0.9 | 22.5-28.0 | 0.3-0.6 | 4.6-5.7 | 8.7-10.9 | 13.4-16.8 |
| 0.150 | 0.035 | 0.115 (Std) | 0.004 | 6.6- 8.1 | 1.7-2.0 | 4.4-5.4 | 1.0-1.2 | 7.1- 8.7 | 5.3- 6.5 | 9.2-11.3 | 0.6-0.8 | 13.2-16.2 | 0.3-0.5 | 4.0-4.9 | 7.6- 9.3 | 10.0-12.2 |
| | | | 0.005 | 6.8- 8.3 | 1.7-2.1 | 4.6-5.6 | 1.0-1.2 | 7.3- 8.9 | 5.4- 6.6 | 9.5-11.6 | 0.6-0.8 | 13.6-16.3 | 0.3-0.5 | 4.1-5.0 | 7.8- 9.5 | 10.2-12.5 |
| | | | 0.006 | 6.9- 8.4 | 1.7-2.1 | 4.6-5.6 | 1.0-1.3 | 7.4- 9.0 | 5.5- 6.7 | 9.7-11.8 | 0.6-0.8 | 13.8-16.8 | 0.3-0.5 | 4.2-5.1 | 8.0- 9.7 | 10.5-12.8 |
| 0.150 | 0.010 | 0.140 (Hi Den) | 0.004 | 11.2-13.7 | 1.7-2.1 | 6.5-7.9 | 1.0-1.2 | 12.0-14.7 | 9.0-11.0 | 15.6-19.2 | 0.7-0.9 | 22.4-27.4 | 0.3-0.5 | 4.6-5.6 | 8.7-10.6 | 13.4-16.2 |
| | | | 0.005 | 11.5-14.0 | 1.7-2.1 | 6.6-8.1 | 1.1-1.3 | 12.3-15.0 | 9.2-11.2 | 16.0-19.6 | 0.7-0.9 | 23.0-28.0 | 0.3-0.6 | 4.7-5.7 | 8.9-10.8 | 13.6-16.6 |
| | | | 0.006 | 11.8-14.3 | 1.8-2.1 | 6.9-8.3 | 1.1-1.3 | 12.6-15.4 | 9.5-11.4 | 16.5-20.0 | 0.7-0.9 | 23.5-28.5 | 0.3-0.6 | 4.8-5.9 | 9.1-11.2 | 14.0-17.1 |

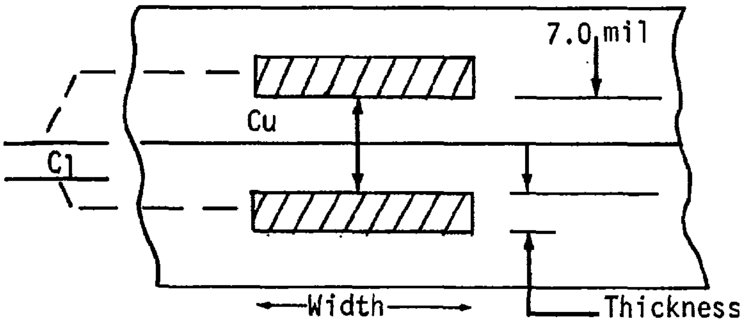
# TABLE 3-5. CAPACITANCE VALUES IN pF/ft FOR H-FILM/FEP INSULATED FLAT CABLE ($\epsilon = 2.9$-$3.1$) (SEE FIGURE 3-13 FOR CAPACITANCE SYMBOL DESIGNATION)




Note: 4 X 40 mil conductors, spaced on 75 mil center-line spacings. See Figure 3-13 for $C_1$ - $C_{19}$ measurement locations.

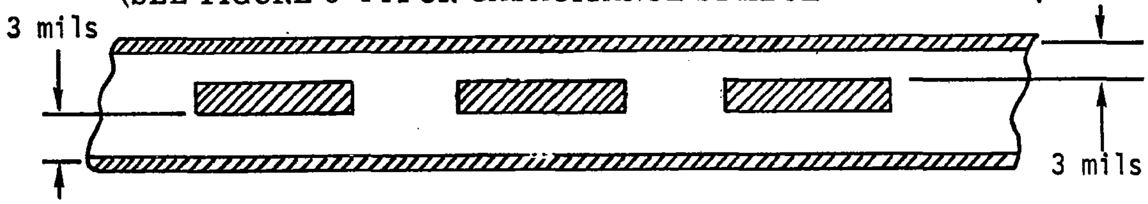
| Centerline Spacing (in.) | Conductor Separation (in.) | Conductor Width (in.) | Conductor Thickness (in.) | $C_1, C_3, C_5$ | $C_2$ | $C_4$ | $C_6$ | $C_7$ | $C_8$ | $C_9$ | $C_{10}$ | $C_{11}, C_{12}, C_{13}, C_{15}, C_{16}$ | $C_{14}$ | $C_{17}$ | $C_{18}$ | $C_{19}$ |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| 0.050 | 0.025 | 0.025 (Std) | 0.003 | 5.7- 8.1 | 1.3-1.9 | 3.6-5.1 | 0.8-1.1 | 6.1- 8.7 | 4.6- 6.5 | 8.0-11.4 | 0.5-0.8 | 11.4-16.1 | 0.2-0.4 | 3.0-4.2 | 5.7- 8.0 | 7.5-10.5 |
| | | | 0.004 | 5.8- 8.2 | 1.3-1.9 | 3.7-5.2 | 0.8-1.1 | 6.2- 8.8 | 4.7- 6.6 | 8.1-11.4 | 0.5-0.8 | 11.6-16.2 | 0.2-0.4 | 3.0-4.3 | 5.7- 8.2 | 7.5-10.8 |
| 0.050 | 0.010 | 0.040 (Hi Den) | 0.003 | 10.6-13.7 | 1.6-2.0 | 6.1-7.9 | 1.0-1.2 | 11.4-14.7 | 8.5-11.0 | 14.8-19.2 | 0.6-0.8 | 21.2-27.4 | 0.3-0.5 | 4.4-5.6 | 8.3-10.6 | 12.8-16.2 |
| | | | 0.004 | 10.9-14.0 | 1.6-2.1 | 6.3-8.1 | 1.0-1.3 | 11.7-15.0 | 8.8-11.2 | 15.1-19.6 | 0.7-0.9 | 21.7-28.0 | 0.3-0.6 | 4.5-5.7 | 8.5-10.8 | 13.0-16.8 |
| | | | 0.005 | 11.2-14.3 | 1.6-2.1 | 6.5-8.4 | 1.0-1.3 | 12.0-15.3 | 9.0-11.4 | 15.4-20.0 | 0.7-0.9 | 22.4-28.5 | 0.3-0.6 | 4.6-5.9 | 8.7-11.2 | 13.4-17.1 |
| 0.075 | 0.035 | 0.040 (Std) | 0.003 | 5.7- 7.5 | 1.4-1.9 | 3.8-5.0 | 0.8-1.1 | 6.1- 8.1 | 4.6- 6.0 | 8.0-10.5 | 0.5-0.8 | 11.4-15.0 | 0.2-0.4 | 3.4-4.5 | 6.4- 8.5 | 8.5-11.2 |
| | | | 0.004 | 5.8- 7.6 | 1.5-1.9 | 3.9-5.1 | 0.8-1.1 | 6.2- 8.2 | 4.7- 6.1 | 8.1-10.6 | 0.5-0.8 | 11.6-15.2 | 0.2-0.4 | 3.5-4.6 | 6.6- 8.7 | 8.7-11.5 |
| | | | 0.005 | 5.8- 7.6 | 1.5-1.9 | 4.0-5.1 | 0.8-1.1 | 6.2- 8.2 | 4.7- 6.1 | 8.1-10.6 | 0.5-0.8 | 11.6-15.2 | 0.2-0.4 | 3.5-4.6 | 6.6- 8.7 | 8.7-11.5 |
| 0.075 | 0.010 | 0.065 (Hi Den) | 0.003 | 11.1-14.1 | 1.6-2.1 | 6.4-8.2 | 1.0-1.3 | 11.9-15.2 | 8.9-11.2 | 15.5-19.6 | 0.6-0.8 | 22.2-28.1 | 0.3-0.5 | 4.6-5.8 | 8.7-11.0 | 13.3-16.8 |
| | | | 0.004 | 11.1-14.1 | 1.6-2.1 | 6.4-8.2 | 1.0-1.3 | 11.9-15.2 | 8.9-11.2 | 15.5-19.8 | 0.6-0.8 | 22.2-28.1 | 0.3-0.5 | 4.6-5.8 | 8.7-11.0 | 13.3-16.8 |
| | | | 0.005 | 11.7-14.7 | 1.7-2.2 | 6.8-8.5 | 1.1-1.3 | 12.5-15.8 | 9.4-11.8 | 16.4-20.5 | 0.7-0.9 | 23.4-29.3 | 0.3-0.6 | 4.8-6.0 | 9.1-11.2 | 14.0-17.4 |
| 0.100 | 0.035 | 0.065 (Std) | 0.004 | 6.4- 8.1 | 1.6-2.0 | 4.3-5.4 | 1.0-1.2 | 6.9- 8.7 | 5.1- 6.5 | 9.0-11.3 | 0.6-0.8 | 12.8-16.2 | 0.3-0.5 | 3.9-4.9 | 7.4 9.3 | 9.8-12.2 |
| | | | 0.005 | 6.5- 8.2 | 1.6-2.1 | 4.3-5.5 | 1.0-1.2 | 7.0- 8.8 | 5.2- 6.6 | 9.1-11.5 | 0.6-0.8 | 13.0-16.4 | 0.3-0.5 | 3.9-4.9 | 7.4- 9.3 | 9.8-12.2 |
| | | | 0.006 | 6.6- 8.3 | 1.7-2.1 | 4.4-5.6 | 1.0-1.2 | 7.1- 8.9 | 5.3- 6.6 | 9.2-11.6 | 0.6-0.8 | 13.2-16.6 | 0.3-0.5 | 4.0-5.0 | 7.6- 9.5 | 10.0-12.5 |
| 0.100 | 0.010 | 0.090 (Hi Den) | 0.004 | 11.6-14.5 | 1.7-2.2 | 6.7-8.4 | 1.0-1.3 | 12.4-15.6 | 9.3-11.6 | 16.2-20.3 | 0.7-0.9 | 23.2-29.0 | 0.4-0.6 | 4.7-5.9 | 8.9-11.2 | 13.6-17.1 |
| | | | 0.005 | 12.0-14.8 | 1.8-2.2 | 7.0-8.6 | 1.1-1.3 | 12.8-15.8 | 9.6-11.8 | 16.8-20.8 | 0.7-0.9 | 24.0-29.5 | 0.4-0.6 | 4.9-6.1 | 9.3-11.6 | 14.2-17.7 |
| | | | 0.006 | 12.2-15.1 | 1.8-2.3 | 7.2-8.8 | 1.1-1.4 | 13.1-16.2 | 9.7-12.1 | 17.0-21.1 | 0.7-0.9 | 24.4-30.2 | 0.4-0.6 | 5.0-6.2 | 9.5-11.8 | 14.5-18.0 |
| 0.150 | 0.035 | 0.115 (Std) | 0.004 | 7.1- 9.2 | 1.8-2.3 | 4.8-6.2 | 1.1-1.4 | 7.6- 9.9 | 5.7- 7.3 | 9.9-12.8 | 0.7-0.9 | 14.2-18.4 | 0.3-0.5 | 4.2-5.5 | 8.0-10.4 | 10.5-13.8 |
| | | | 0.005 | 7.2- 9.3 | 1.8-2.3 | 4.8-6.2 | 1.1-1.4 | 7.7-10.0 | 5.8- 7.4 | 10.1-13.0 | 0.7-0.9 | 14.4-18.6 | 0.3-0.5 | 4.3-5.6 | 8.2-10.6 | 10.7-14.0 |
| | | | 0.006 | 7.3- 9.4 | 1.8-2.4 | 4.9-6.3 | 1.1-1.4 | 7.8-10.1 | 5.8- 7.5 | 10.2-13.2 | 0.7-0.9 | 14.6-18.8 | 0.3-0.5 | 4.4-5.7 | 8.4-10.8 | 11.0-14.2 |
| 0.150 | 0.010 | 0.140 (Hi Den) | 0.004 | 12.1-14.9 | 1.8-2.2 | 7.0-8.6 | 1.1-1.3 | 13.0-16.0 | 9.7-11.9 | 17.0-20.8 | 0.7-0.9 | 24.2-29.8 | 0.4-0.6 | 5.0-6.1 | 9.5-11.6 | 14.5-17.6 |
| | | | 0.005 | 12.4-15.2 | 1.9-2.3 | 7.2-8.8 | 1.1-1.4 | 13.3-16.3 | 9.9-12.2 | 17.4-21.2 | 0.7-0.9 | 24.8-30.4 | 0.4-0.6 | 5.1-6.2 | 9.7-11.8 | 14.8-18.0 |
| | | | 0.006 | 12.7-15.5 | 1.0-2.3 | 7.4-9.0 | 1.1-1.4 | 13.6-16.6 | 10.2-12.4 | 17.8-21.6 | 0.8-0.9 | 25.3-30.4 | 0.4-0.6 | 5.2-6.4 | 9.9-12.2 | 15.0-18.6 |

TABLE 3-6. CAPACITANCE VALUES FOR MYLAR ($\epsilon = 2.5$) AND H-FILM/FEP ($\epsilon = 2.9 - 3.1$) INSULATED FLAT CABLE




| Conductor Width (in.) | Conductor Thickness (in.) | Mylar $C_1$ (pf/ft) | H-Film/FEP $C_1$ (pf/ft) |
|---|---|---|---|
| 0.025 | 0.003 | 23.5 - 44.4 | 27.2 - 51.9 |
| 0.025 | 0.004 | 24.1 - 45.1 | 27.8 - 52.7 |
| 0.040 | 0.003 | 35.8 - 64.7 | 40.2 - 75.5 |
| 0.040 | 0.004 | 35.3 - 65.3 | 40.8 - 76.3 |
| 0.040 | 0.005 | 35.8 - 65.9 | 41.4 - 77.0 |
| 0.065 | 0.003 | 53.5 - 98.4 | 61.8 - 114.9 |
| 0.065 | 0.004 | 54.1 - 99.1 | 62.5 - 115.7 |
| 0.065 | 0.005 | 54.6 - 99.7 | 63.1 - 116.4 |
| 0.065 | 0.006 | 55.0 - 100.2 | 63.6 - 117.0 |
| 0.090 | 0.004 | 72.8 - 132.8 | 84.2 - 115.1 |
| 0.090 | 0.005 | 73.3 - 133.4 | 84.8 - 155.8 |
| 0.090 | 0.006 | 73.8 - 133.9 | 85.3 - 156.4 |
| 0.115 | 0.004 | 91.6 - 166.6 | 105.8 - 194.5 |
| 0.115 | 0.005 | 92.1 - 167.1 | 106.4 - 195.2 |
| 0.115 | 0.006 | 92.5 - 167.7 | 106.9 - 195.8 |
| 0.140 | 0.004 | 110.3 - 200.3 | 127.5 - 233.9 |
| 0.140 | 0.005 | 110.8 - 200.9 | 128.1 - 234.6 |
| 0.140 | 0.006 | 111.3 - 201.4 | 128.6 - 235.2 |

TABLE 3-7. CAPACITANCE VALUES IN pF FOR A DOUBLE-SHIELDED H-FILM/FEP CABLE ( = 2.9-3.1)
(SEE FIGURE 3-14 FOR CAPACITANCE SYMBOL DESIGNATION)




| Centerline Spacing (in.) | Conductor Separation (in.) | Conductor Width (in.) | Conductor Thickness (in.) | $C_1$, $C_3$, $C_8$ | $C_2$, $C_4$, $C_7$, $C_6$, $C_9$ | $C_5$, $C_{10}$ |
|---|---|---|---|---|---|---|
| 0.050 | 0.025 (Std) | 0.025 | 0.003 | 0.04 - 0.07 | 116 - 192 | 0.03 - 0.06 |
| | | | 0.004 | 0.05 - 0.09 | 118 - 195 | 0.04 - 0.08 |
| 0.050 | 0.010 | 0.040 (Hi Den) | 0.003 | 0.09 - 0.18 | 184 - 284 | 0.08 - 0.15 |
| | | | 0.004 | 0.13 - 0.23 | 186 - 286 | 0.11 - 0.20 |
| | | | 0.005 | 0.16 - 0.28 | 188 - 290 | 0.14 - 0.24 |
| 0.075 | 0.035 | 0.040 (Std) | 0.003 | 0.03 - 0.05 | 184 - 284 | 0.02 - 0.04 |
| | | | 0.004 | 0.04 - 0.06 | 186 - 286 | 0.03 - 0.05 |
| | | | 0.005 | 0.05 - 0.08 | 188 - 290 | 0.04 - 0.06 |
| 0.075 | 0.010 | 0.065 (Hi Den) | 0.003 | 0.09 - 0.18 | 282 - 420 | 0.08 - 0.15 |
| | | | 0.004 | 0.13 - 0.23 | 285 - 425 | 0.10 - 0.20 |
| | | | 0.005 | 0.16 - 0.28 | 286 - 426 | 0.14 - 0.24 |
| 0.100 | 0.035 | 0.065 (Std) | 0.004 | 0.04 - 0.06 | 285 - 425 | 0.03 - 0.05 |
| | | | 0.005 | 0.05 - 0.08 | 286 - 426 | 0.04 - 0.07 |
| | | | 0.006 | 0.06 - 0.09 | 287 - 427 | 0.05 - 0.08 |
| 0.100 | 0.010 | 0.090 (Hi Den) | 0.004 | 0.13 - 0.23 | 437 - 638 | 0.11 - 0.20 |
| | | | 0.005 | 0.16 - 0.28 | 438 - 642 | 0.14 - 0.24 |
| | | | 0.006 | 0.19 - 0.33 | 440 - 645 | 0.16 - 0.28 |
| 0.150 | 0.035 | 0.115 (Std) | 0.004 | 0.04 - 0.06 | 632 - 920 | 0.03 - 0.05 |
| | | | 0.005 | 0.05 - 0.08 | 634 - 922 | 0.04 - 0.07 |
| | | | 0.006 | 0.06 - 0.09 | 635 - 823 | 0.05 - 0.08 |
| 0.150 | 0.010 | 0.140 | 0.004 | 0.13 - 0.23 | 780 - 1170 | 0.11 - 0.20 |
| | | | 0.005 | 0.16 - 0.28 | 782 - 1172 | 0.16 - 0.28 |
| | | | 0.006 | 0.19 - 0.33 | 782 - 1172 | 0.16 - 0.28 |

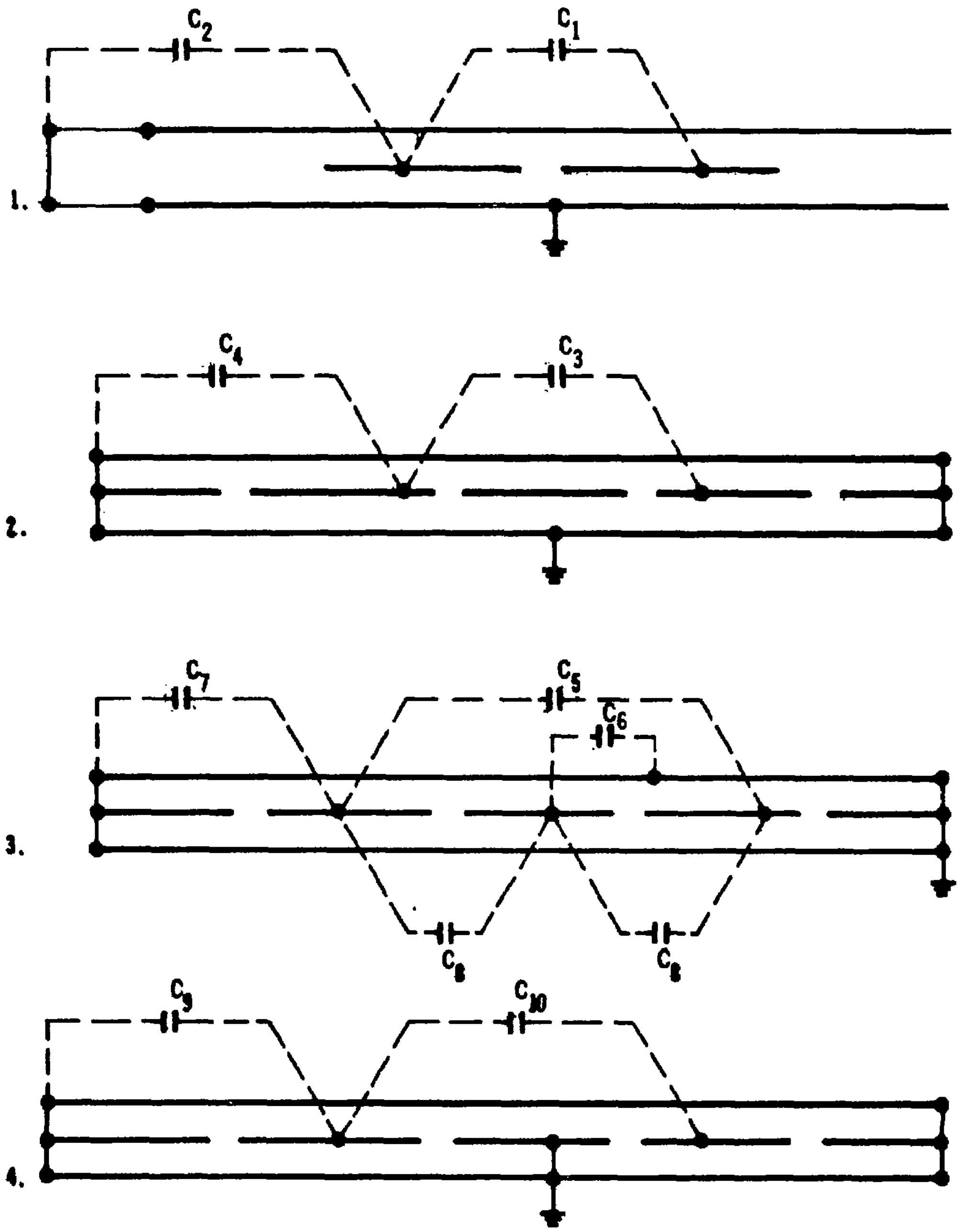



FIGURE 3-14. Double-shielded cable capacitance configurations.

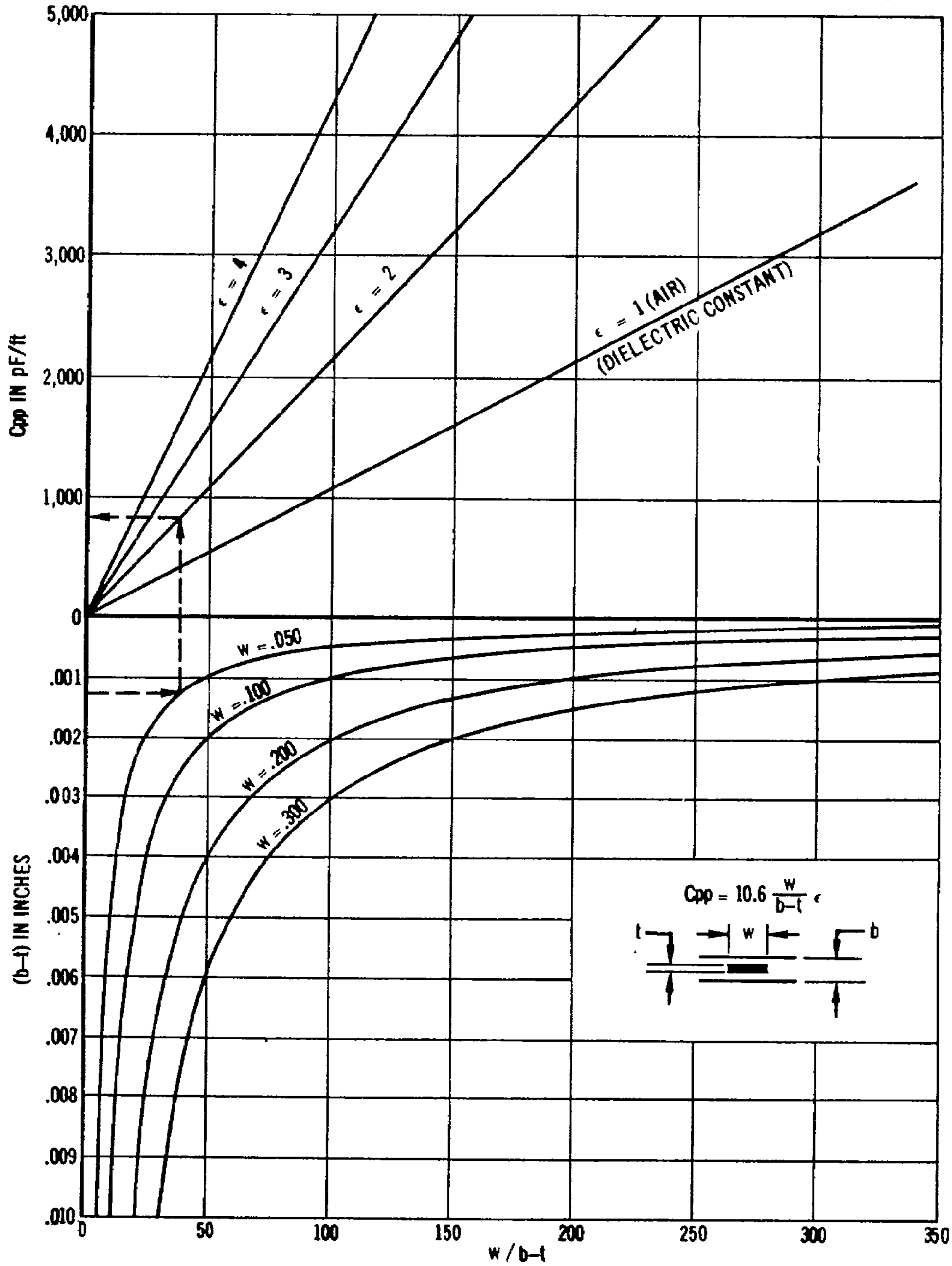



FIGURE 3-15. Conductor-to-shield capacitance for double-shielded FCC.

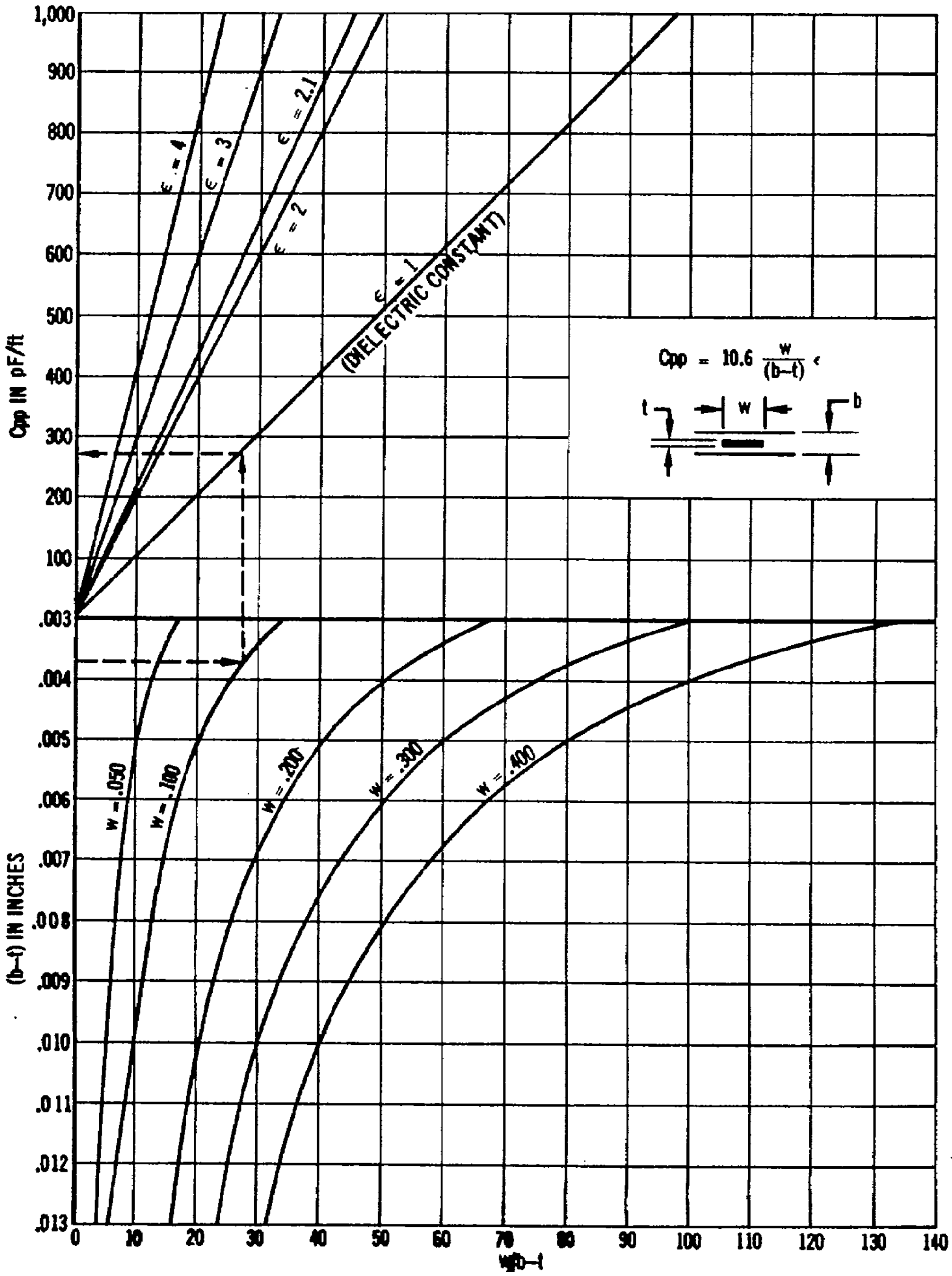



FIGURE 3-16. Conductor-to-shield capacitance for double-shielded FCC.

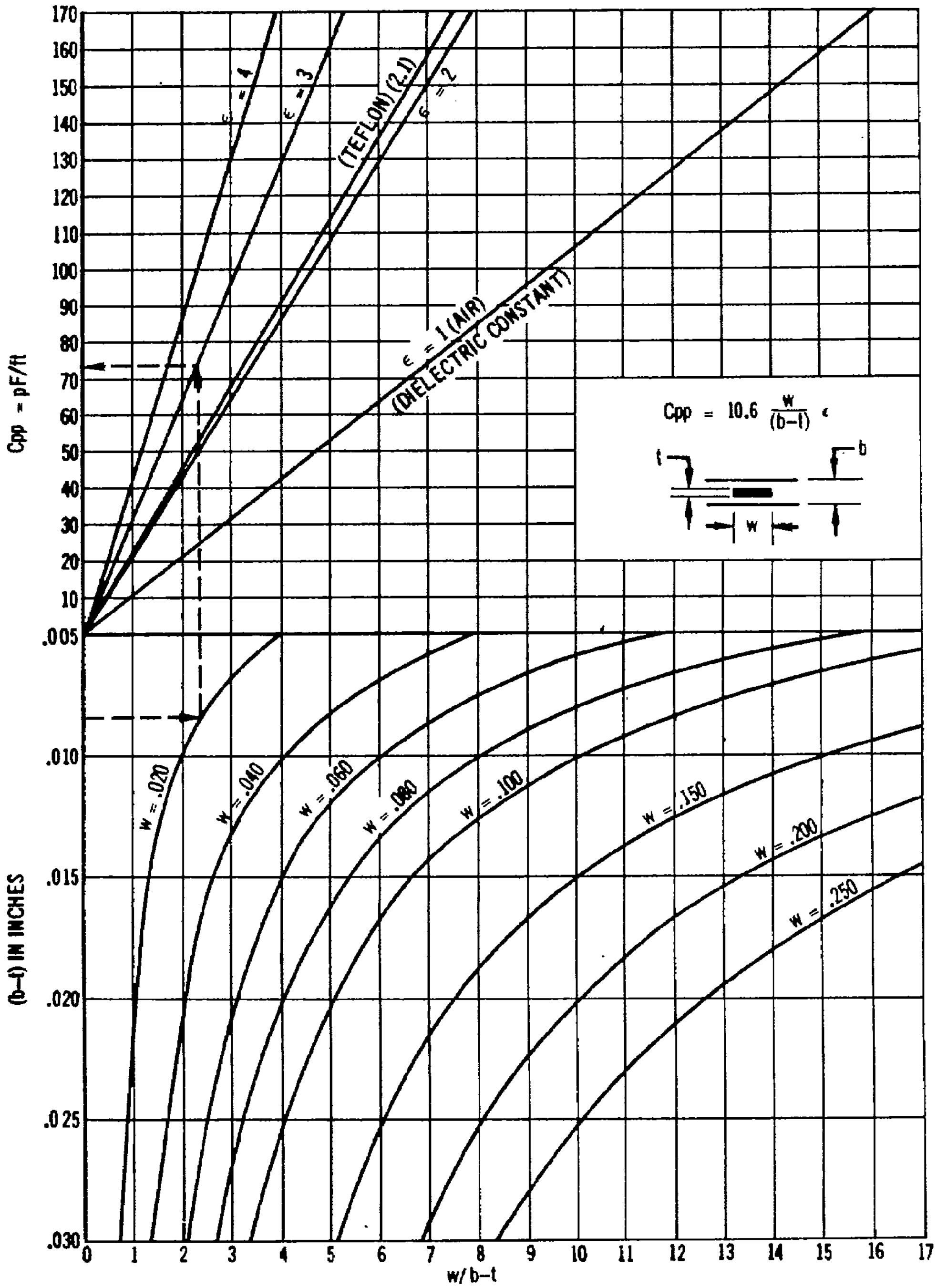



FIGURE 3-17. Conductor-to-shield capacitance for double-shielded FCC.

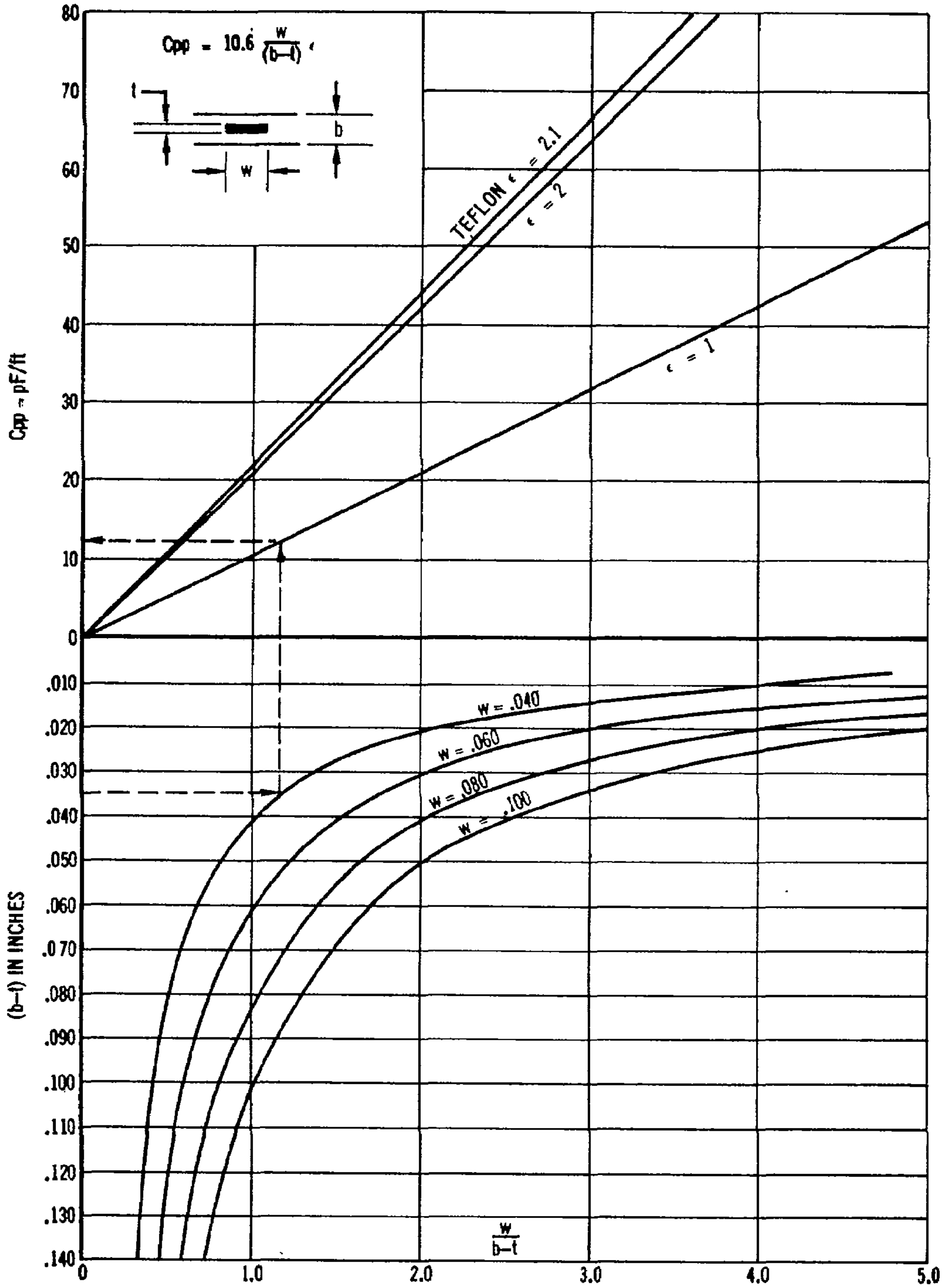



FIGURE 3-18. Conductor-to-shield capacitance for double-shielded FCC.

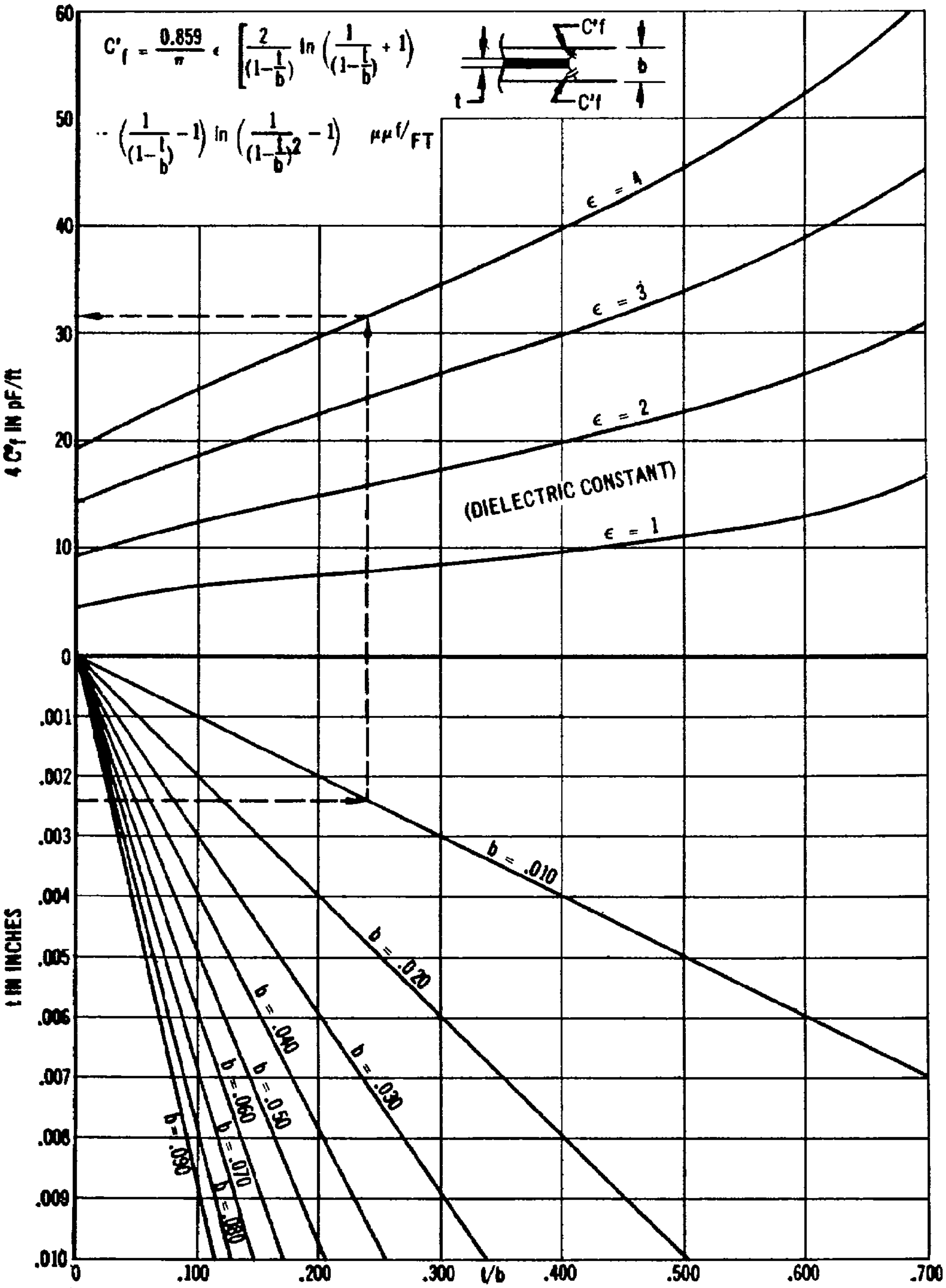



FIGURE 3-19. Conductor-to-shield fringe capacitance for double-shielded FCC.

3.2.3.1.5 Inductance Data. Any selected pair of conductors within moderately close proximity exhibits some amount of inductive coupling between the two conductors whenever current flows through one of the conductors. This can be measured as mutual inductance, M. In addition, any selected pair of differential elements (in the differential calculus sense) of a single conductor interacts inductively when a current flows through these elements. When this effect is considered over the entire conductor, it can be measured as self-inductance, $L_{self}$. If no magnetic materials are in the vicinity of the conductors, inductance of conductors can be calculated analytically for simple geometrical configurations. Dr. Frederich W. Grover provides formulas and tables that are applicable to FCC[1]. In the case of a straight conductor of rectangular cross-section, self-inductance is given by

$$L_{self} = 0.00200\,A \left[\ln \frac{2A}{B+C} + \frac{1}{2} - \ln E + \frac{0.2235(B+C)}{A} - \frac{0.04995(B+C)^2}{4\,L^2}\right]$$

where

$L_{self}$ = microhenries

A = length of conductor in centimeters

B and C = width and thickness, respectively, of the conductor in centimeters

E = a correction term computed and tabulated by Grover[2]

Mutual inductance of a pair of straight, parallel conductors is given by

$$M = 0.00200\,A \left[\ln \frac{2A}{R} - 1 + \frac{R}{A} - \frac{R^2}{4A^2}\right]$$

where

M = microhenries

A = length of the conductors in centimeters

R = geometric mean distance in centimeters between all elements of the cross-section of one conductor and all elements of the cross-section of the other conductor.

Data for calculating geometric mean distances for adjacent conductors of rectangular cross sections are tabulated by Grover[3].

Calculations of self and mutual inductance have been made for adjacent parallel, straight conductors in 23 standard configurations of FCC, both on an edge-to-edge basis within a cable and on an over-and-under basis for conductors in stacked cables, with no magnetic material in close proximity. These calculations were made for cable lengths of 0.1 through 100 feet. From these calculated data, the total inductance of these pairs of conductors was computed for connection in a current-opposing circuit, as in transmission lines. This transmission line (*tl*) value of inductance is given by

$$L_{tl} = 2L_{self} - 2M$$

with all dimensions in consistent units (microhenries in our case).

The inductance, in terms of microhenries per foot, is essentially independent of length for each cable configuration for the transmission line circuit, while the ratio increases with length toward an asymptotic limit for both self and mutual inductance. This effect is shown graphically in Figure 3-20 for 0.004- by 0.040-inch conductors on 0.075-inch centerline spacing.

1. Grover, Frederich W.: Inductance Calculations: Working Formulas and Tables. Dover Publications, Inc., New York, N.Y., 1962.
2. Ibid, p. 14.
3. Ibid, p. 14.

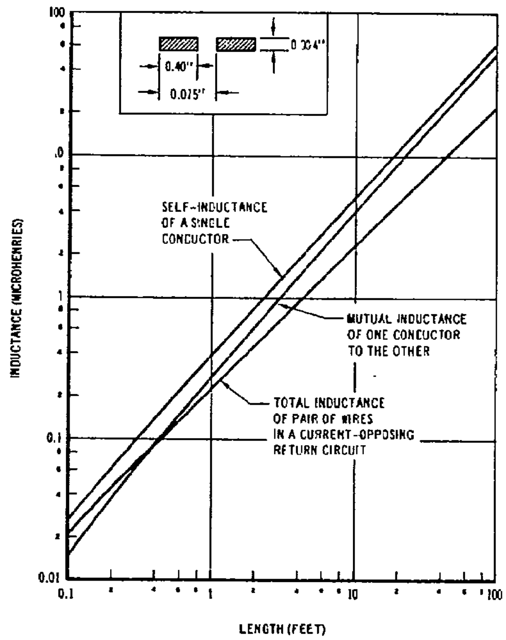



FIGURE 3-20. Components of inductance in a flat cable running in a straight line.

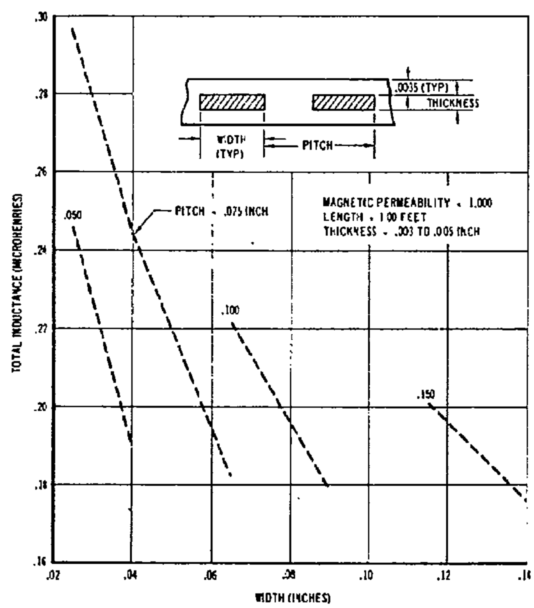



FIGURE 3-21. Total inductance of edge-to-edge conductors in a return circuit.

Because of the close coupling between parallel pairs of adjacent conductors, these pairs, when connected in a transmission-line circuit, exhibit a total inductance that is almost entirely dependent on only this coupling, and essentially independent of whether the pairs of wires travel a straight or a circuitous path between their sources and destinations. As a result, calculations of transmission-line inductance that are made for straight conductors are also valid for other-than-straight routing that can be expected in the wiring of electronic equipment. These values of inductance for 23 standard cable configurations (edge-to-edge and over-and-under) are shown in Figures 3-21 and 3-22, respectively.

The total inductance tolerance of a transmission-line pair of conductors in edge-to-edge configuration will be primarily a function of pitch or centerline spacing between the conductor and of conductor width. This can be seen from examination of Figure 3-21. As an example, for conductors of 0.040-inch width, an increase from 0.050 to 0.075 inch in pitch (an increment of 0.025 inch) results in an increase in inductance from 0.190 to 0.244 microhenry per foot (an increment of 0.054). As a result, inductance increases by approximately 0.6 percent per 1.0 percent pitch change. For a constant pitch of 0.075 inch, inductance will decrease by 0.9 percent per 1.0 percent width change. Other variation ratios may be derived from the figure.

By similar analysis, the total inductance tolerance of a transmission-line pair of conductors in over-and-under configurations is primarily a function of conductor width and of pitch between the stacked conductors. For constant width, the variation ratio for stacked cable with 0.004- by 0.065-inch conductors is approximately 1.5 percent inductance change per 1.0 percent pitch change. For constant pitch, the ratio will be approximately 1.0 percent inductance change per 1.0 percent width change.

3.2.3.1.6 Characteristic Impedance. Characteristic impedance of transmission-line pairs can be determined from capacitance, inductance, series resistance, and shunt conductance, according to

$$Z_o = \left( \frac{\sqrt{R^2 + 4\pi^2 f^2 L_{t\ell}^2}}{\sqrt{G^2 + 4\pi^2 f^2 C^2}} \right)^{1/2}$$

where

$Z_O$ = characteristic impedance in ohms

R = series resistance in ohms per unit length

$L_{t\ell}$ = inductance of the transmission line in henries per unit length

G = shunt conductance in ohms per unit length

C = shunt capacitance in farads per unit length

f = frequency of the applied signal.

The unit of length must obviously be consistent for all of the electrical parameters; e.g., a "per foot" basis.

It is appropriate for lower frequencies to simplify this formula by assuming that resistance and conductance components are insignificant. This is valid for FCC that are described in this report, for frequencies of a few megahertz or less. The simplified formula is

$$Z_o = \sqrt{\frac{L_{t\ell}}{C}}$$

The characteristic impedance of adjacent pairs of unshielded conductors for edge-to-edge conductors is plotted in Figures 3-23 and 3-24, and for over-and-under (stacked) conductors in Figures 3-25 and 3-26, using the simplified formula.

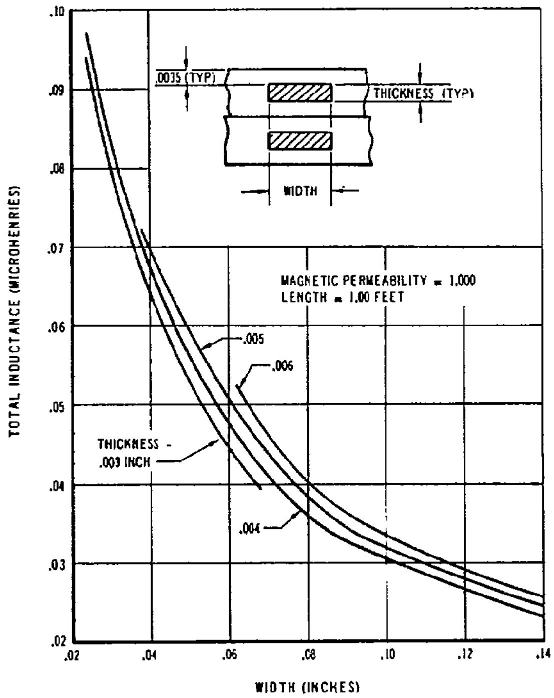



FIGURE 3-22. Total inductance of over-and-under conductors in a return circuit.

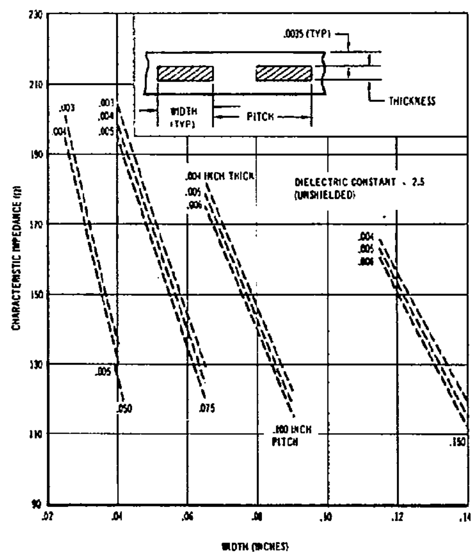



FIGURE 3-23. Characteristic impedance of edge-to-edge conductors in a transmission-line circuit, dielectric constant = 2.5.

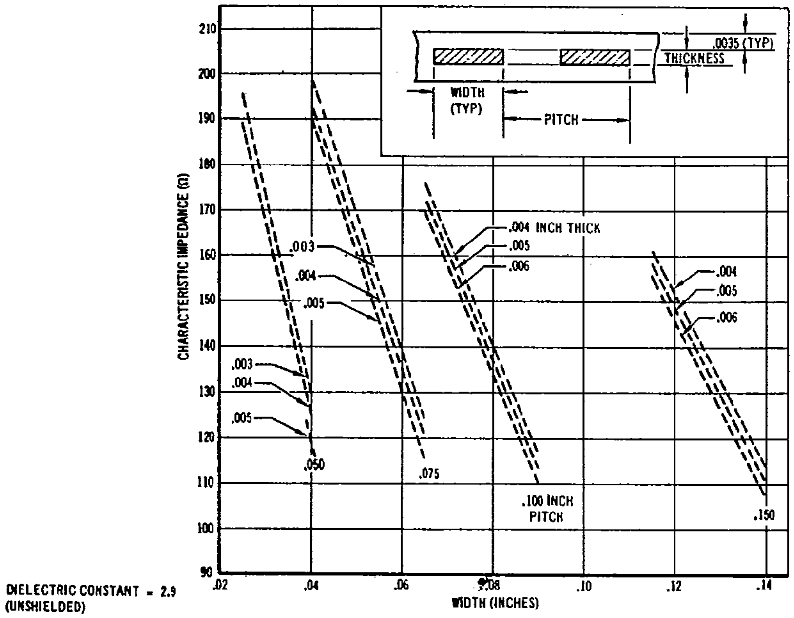



FIGURE 3-24. Characteristic impedance of edge-to-edge conductors in a transmission-line circuit, dielectric constant = 2.9.

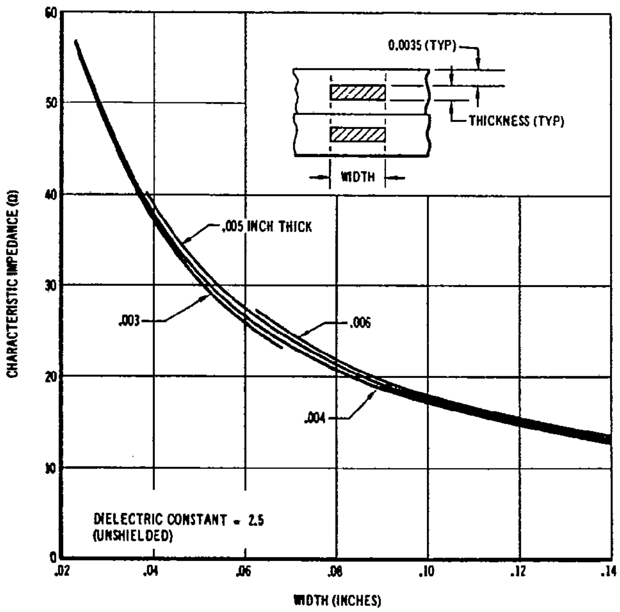



FIGURE 3-25. Characteristic impedance of over-and-under conductors in a transmission-line circuit, dielectric constant = 2.5.

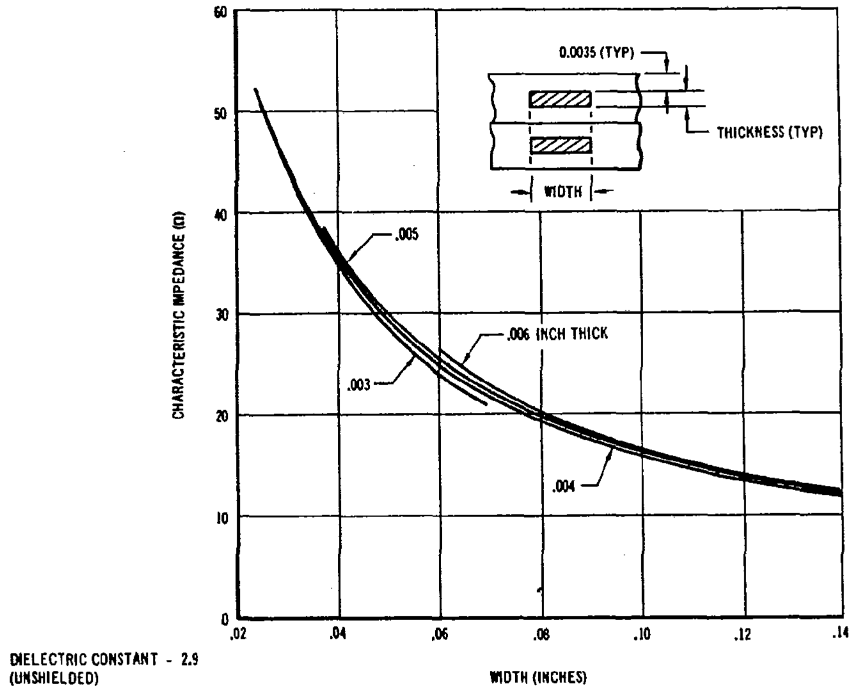



FIGURE 3-26. Characteristic impedance of over-and-under conductors in a transmission-line circuit, dielectric constant = 2. 9.

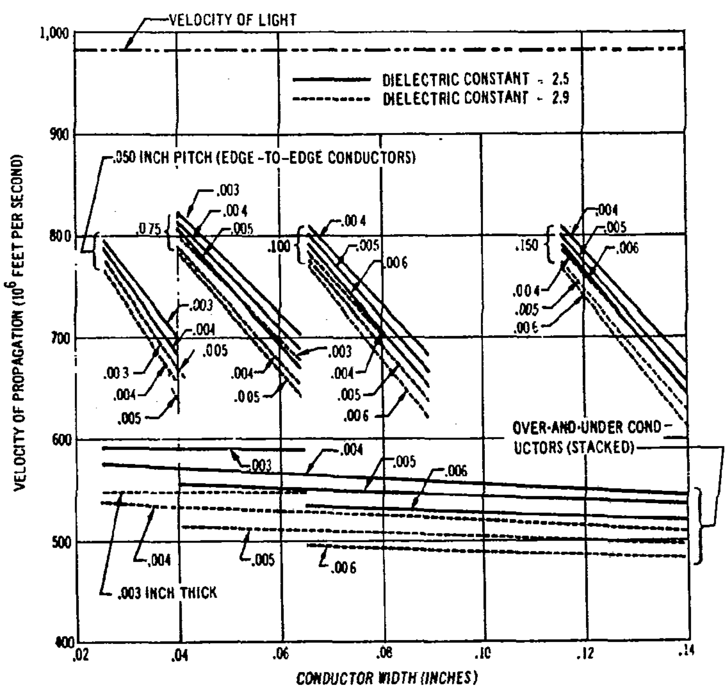



FIGURE 3-27. Velocity of propagation for conductors in a transmission-line circuit (unshielded).

3.2.3.1.7 Velocity of Propagation. Velocity of propagation of a wavefront in a transmission line is given by the formula

$$V = \frac{2\pi f}{\beta}$$

where $\beta$ is the imaginary part of the complex quantity

$$\sqrt{\underline{Z} \cdot \underline{Y}}$$

where

$$\underline{Z} = R + j\,2\pi f\,L_{t\ell}$$

$$\underline{Y} = G + j\,2\pi f C$$

and where

$R$ = series resistance in ohms per unit length

$L_{t\ell}$ = inductance of the transmission line in henries per unit length

$G$ = shunt conductance in mhos per unit length

$C$ = shunt capacitance in farads per unit length

$f$ = frequency of the applied signal

The dimensions of velocity are units of length per second where the unit of length is the same as used in the electrical parameters; e.g., foot-per-second, when the electrical parameters are given in ohms per foot, farads per foot, etc.

When the resistive and conductive components of $\beta$ are insignificant, say for frequencies less than a few megahertz, the formula for velocity of propagation simplifies to

$$V = \frac{1}{\sqrt{L_{t\ell} C}}$$

Velocity of propagation is plotted, using the simplified formula, in Figure 3-27 for adjacent pairs of conductors for the standard unshielded FCC configurations of this report. In this case of edge-to-edge conductors, the actual values are shown as points with straight lines drawn between the points for identification. In the case of over-and-under (stacked) conductors, the velocity values for the various thicknesses of conductor fall very close to straight lines within the range of conductor widths shown.

3.2.3.2 Electromagnetic Compatibility Considerations in Cable Design. Most of us are familiar with the slapstick comedy in which a handyman makes plumbing repairs, and water spurts from the lighting fixture when the switch is turned on. Complex electrical/electronic systems have exhibited similar tendencies under far less humorous conditions. The electrical/electronic components of any system must be mutually compatible within the system, with components in other systems of the electrical/electronic complex, and with the electromagnetic environment of the electrical/electronic installation.

A major source of electromagnetic incompatibilities, in any electrical/electronic system, is the system interconnection cabling. Most electronic units receive sufficient testing during the breadboard and prototype phases of design, so that internal spurious coupling is identified and corrected. Large systems are interconnected with cable assemblies too massive for convenient laboratory testing; therefore, most spurious system coupling problems are identified and corrected during system testing, acceptance testing, or operational usage of the hardware. The consequences of uncorrected spurious coupling occurring during a critical mission, either manned or unmanned, involving complex assemblies of nonreusable hardware, can be serious.

FCC has unique features which can be taken advantage of, in many instances, to obtain a wiring system that is superior to any cable of conventional round wires. On the other hand, the close proximity of adjacent conductors in FCC can cause more severe problems than with round-wire cable if proper and detailed attention is not paid to EMC considerations.

The ensuing paragraphs in this section discuss three major aspects of EMC design; (1) a simplified approach to optimization of conductor placement, (2) crosstalk between conductors, and (3) electric and magnetic shielding of FCC's.

The user who has not developed a sound background of experience in EMC design is strongly advised to study fundamentals of EMC, as discussed in Paragraph 3. 2. 10, and to consult with an EMC specialist to resolve remaining difficulties. To further emphasize this matter, an additional warning notice is given here.

WARNING

Experience has shown that certain interference and susceptibility characteristics are typical of the average electrical/electronic system, and that these characteristics can be extrapolated for use in more specialized types of systems. Because of the extremely broad spectrum of hardware involved, these characteristics are subject to wide variations and should be utilized only as a preliminary design goal until more specific hardware design and test information becomes available. As confidence in system performance increases, arbitrarily specified, initial characteristics should be superseded or refined with analyses of actual system performance.

3.2.3.2.1 Simplified Optimization of Conductor Placement. An examination of the factors influencing the actual spurious-coupling loss reveals that these factors (Tables 3-8 through 3-12), when properly chosen, can be multiplied by each other, thus providing a criticality figure ranging from an extremely severe source of interference, through a neutral condition, to an extremely sensitive susceptible circuit. By using a logarithmic scale, it becomes possible to add the transformed log values instead of engaging in a multiplication process. Avoiding multiplication provides smaller and more convenient numbers.

When the numeric values are summed, and the two alphabetic prefixes are carried through; an alphanumeric criticality figure for electromagnetic compatibility is generated for the conductor. This criticality figure identifies the function as being susceptible (S) to or a source of interference (I) that is dominantly magnetic (M) or electric-field (E) coupled, and has a magnitude removed between 20 times the numeric value (in decibels) and 20 times the difference of the numeric value less 1 (in decibels) from a neutral level.

After assigning an alphanumeric criticality figure to each function, list the functions in criticality sequence with the most-critical susceptible circuit at one end of the list and the most-critical interference source at the other end of the list. Keep conductor pairs or multiples in the same circuit adjacent to each other in the criticality sequence list. Low-impedance return or neutral conductors are partially effective as electric-field shields at low frequencies.

Within the constraints imposed on the cable segment being designed, make function assignments to conductors within the cable-segment cross-section on the basis of the function location in the criticality sequence list. Try to alternate magnetic and electric field dominant functions within the same grouping in the criticality sequence to minimize crosstalk between adjacent conductors with similar coupling characteristics.

Complete an electromagnetic compatibility analysis on several of the most-critical conductor combinations to establish a reference point for later differential analysis of the bulk of conductor combinations, and to provide a basis for the modification of the weighing factors supplied, it this becomes necessary.

Determine the distributed-parameter values of typical lengths of typical conductor combinations. Determine the ratios that exist between the various mutual capacitances, shunt capacitances, mutual inductances, series self-inductances, and other variables as the cable and cable conductor configurations are changed. Convert these ratios to equivalent decibels of change in coupling, for use as constraints for comparison with the difference in criticality figures (in decibels) of closely spaced functions assigned to conductors in accordance with the criticality sequence list.

# TABLE 3-8. TERMINATION IMPEDANCE COUPLING FACTOR

| Units Magnetic Coupling | Resistance/ Reactance | Reactive Component Value | | | |
|---|---|---|---|---|---|
| | | 60 Hz | 400 Hz | 4 kHz | 50 kHz |
| M4 | Under 10 mΩ | Above 0.3 F<br>Under 50μH | Above 0.03 F<br>Under 5μH | Above 3000μF<br>Under 0.5μH | Above 300μF<br>Under 0.05μH |
| M3 | 10 to 100 mΩ | 0.03 to 0.3 F<br>50μH to 500μH | 3000μF to 0.03 F<br>5μH to 50μH | 300μF to 3000μF<br>0.5μH to 5μH | 30μF to 300μF<br>0.05μH to 0.5μH |
| M2 | 100 mΩ to 1Ω | 3000μF to 0.03 F<br>500μH to 5mH | 300μF to 3000μF<br>50μH to 500μH | 30μF to 300μF<br>5μH to 50μH | 3μF to 30μF<br>0.5μH to 5μH |
| M1 | 1 to 10Ω | 300μF to 3000μF<br>5mH to 50mH | 30μF to 300μF<br>500μH to 5mH | 3μF to 30μF<br>50μH to 500μH | 0.3 F to 3μF<br>5μH to 50μH |
| M0 | 10 to 100Ω | 30μF to 300μF<br>50mH to 500mH | 3μF to 30μF<br>5mH to 50mH | 0.3μF to 3μF<br>500μH to 5mH | 0.03μF to 0.3μF<br>50μH to 500μH |
| Units Electric Coupling | | | | | |
| E0 | 100Ω to 1 kΩ | 3μF to 30μF<br>500mH to 5H | 0.3μF to 3μF<br>50mH to 500mH | 0.03μF to 0.3μF<br>5mH to 50mH | 0.003μF to 0.03μF<br>500μH to 5mH |
| E1 | 1 to 10 kΩ | 0.3μF to 3μF<br>5H to 50H | 0.03μF to 0.3μF<br>500mH to 5H | 0.003μF to 0.03μF<br>50mH to 500mH | 300pF to 0.003μF<br>5mH to 50mH |
| E2 | 10 to 100 kΩ | 0.03μF to 0.3μF<br>50H to 500H | 0.003μF to 0.03μF<br>5H to 50H | 300pF to 0.003μF<br>500mH to 5H | 30pF to 300pF<br>50mH to 500mH |
| E3 | 100 kΩ to 1 MΩ | 0.003μF to 0.03μF<br>500H to 5000H | 300pF to 0.003μF<br>50H to 500H | 30pF to 300pF<br>5H to 50H | 3pF to 30pF<br>500mH to 5H |
| E4 | Above 1 MΩ | Under 0.003μF<br>Above 5000H | Under 300pF<br>Above 500H | Under 30pF<br>Above 50H | Under 3pF<br>Above 5H |

TABLE 3-9. FREQUENCY/TIME COUPLING FACTOR

| Units of Coupling | Frequency | Equivalent Pulse Width |
|---|---|---|
| 0 | Under 3 Hz | Over 100 ms |
| 1 | 3 to 30 Hz | 10 to 100 ms |
| 2 | 30 to 300 Hz | 1 to 10 ms |
| 3 | 300 Hz to 3 kHz | 100 μs to 1 ms |
| 4 | 3 to 30 kHz | 10 to 100 μs |
| 5 | 30 to 300 kHz | 1 to 10 μs |
| 6 | 300 kHz to 3 MHz | 100 ns to 1 μs |
| 7 | 3 to 30 MHz | 10 to 100 ns |
| 8 | 30 to 300 MHz | 1 to 10 ns |
| 9 | Above 300 MHz | Under 1 ns |

TABLE 3-10. LENGTH COUPLING FACTOR

| Units of Spurious Coupling | Unshield Length of Conductors |
|---|---|
| 0 | Under 1 mm |
| 1 | 1 to 10 mm |
| 2 | 10 to 100 mm |
| 3 | 100 mm to 1 m |
| 4 | 1 to 10 m |
| 5 | 10 to 100 m |
| 6 | 100 m to 1 km |
| 7 | 1 to 10 km |
| 8 | 10 to 100 km |
| 9 | Over 100 km |

TABLE 3-11. AMPLITUDE SUSCEPTIBILITY/
INTERFERENCE FACTOR

| Units of Susceptibility | Voltage | Current |
|---|---|---|
| S4 | Under 10 $\mu$V | Under 0.1 $\mu$A |
| S3 | 10 to 100 $\mu$V | 0.1 to 1 $\mu$A |
| S2 | 100 $\mu$V to 1 mV | 1 to 10 $\mu$A |
| S1 | 1 to 10 mV | 10 to 100 $\mu$A |
| S0 | 10 to 100 mV | 100 $\mu$A to 1 mA |
| Units of Interference | | |
| I0 | 100 mV to 1 V | 1 to 10 mA |
| I1 | 1 to 10 V | 10 to 100 mA |
| I2 | 10 to 100 V | 100 mA to 1 A |
| I3 | 100 V to 1 kV | 1 to 10 A |
| I4 | Above 1 kV | Above 10 A |

TABLE 3-12. RESOLUTION/ACCURACY
SUSCEPTIBILITY FACTOR

| Units of Susceptibility | Resolution or Accuracy Required |
|---|---|
| S4 | Better than 0.1% |
| S3 | 0.1 to 1% |
| S2 | 1 to 10% |
| S1 | 10 to 100% |
| S0 | On-off and other bilevel functions |
| Units of Interference | |
| I2 | Average value for interference |

Crosstalk and shielding-effectiveness values, available in decibel units relative to unshielded cable, may also be used to evaluate the decrease in coupling achieved by the use of devices such as shielding.

Detailed design procedures are highly dependent on the specific types of cables selected for the system and on the specific assembly geometry of the mounted cabling; therefore, these procedures must be developed, refined, and expanded by the user in accordance with the general guidelines that have been previously defined.

3.2.3.2.2 Crosstalk. Crosstalk is the spurious coupling of energy from one conductor to another, both of which are located in a common electromagnetic environment, either unshielded or within a common shield. At termination impedances below the characteristic impedance of the conductor, the spurious coupling will be predominantly magnetic. At termination impedances above the characteristic impedance of the conductor, the spurious coupling will be predominantly electric. The magnitude of coupling is critically dependent on the terminating impedances, but generally becomes significant above midaudio frequencies and approaches unity coupling in the supersonic and radio-frequency spectrums.

A more detailed discussion relating to crosstalk may be found in Paragraph 3.2.10 of this handbook. Since crosstalk or spurious coupling is so highly dependent on the conductor termination characteristics and other function properties, generalizations are often misleading and individual analyses are mandatory.

Measured crosstalks between adjacent, centrally located conductors in a variety of cable configurations is shown in Figures 3-28 through 3-39.

Crosstalk can be divided into the two categories of electric and magnetic coupling.

3.2.3.2.2.1 Electric-Field Crosstalk. A classic case of electric-coupled crosstalk would exist if two adjacent, vertically stacked, unshielded conductors, 40 meters long, resulted in the assignment of the following values to Figure 3-40(A):

$R_{S,\ INT} = 1\ 000\Omega R_{L,\ INT} = 10\ 000\Omega$

$R_{S,\ SUS} = 10\ 000\Omega R_{L,\ SUS} = 100\ 000\Omega$

$C_C = 10\ 000$ PF $\qquad C_S = 1000$ pF

$V_{INT} = 10$ Vrms

$F_{INT} =$ 100-kHz sine wave

$$V_{SUS} = V_{INT} \times \frac{Z_{SUS}}{XC_C + Z_{SUS}}$$

$$V_{SUS} = 10V \times \frac{1361}{1521} = 8.948 \text{ V (Fig. 3-40)}$$

In this case, a signal of 8.9 volts was generated in the susceptible circuit. Reactance values can be read from a reactance chart or calculated. Phase shift was not considered because the coupling impedance and susceptible circuit impedance were widely different values. Reasonable approximations were used for convenience because a high degree of precision was not required.

3.2.3.2.2.2 Magnetic-Field Crosstalk. A classic case of magnetic-coupled crosstalk would exist if two adjacent, vertically stacked, unshielded conductors, 40 meters long, resulted in the assignment of the following values to Figure 3-41:

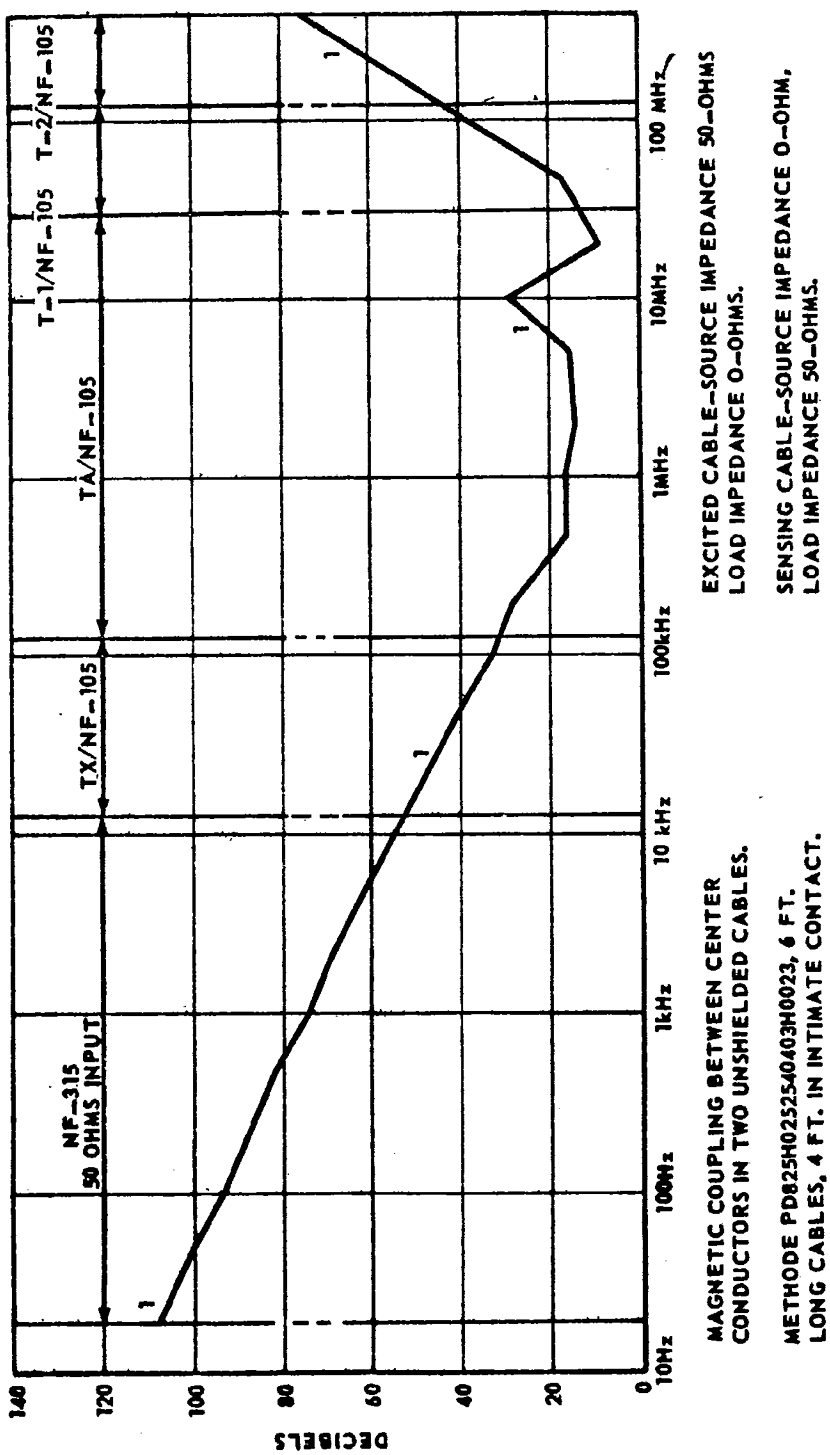
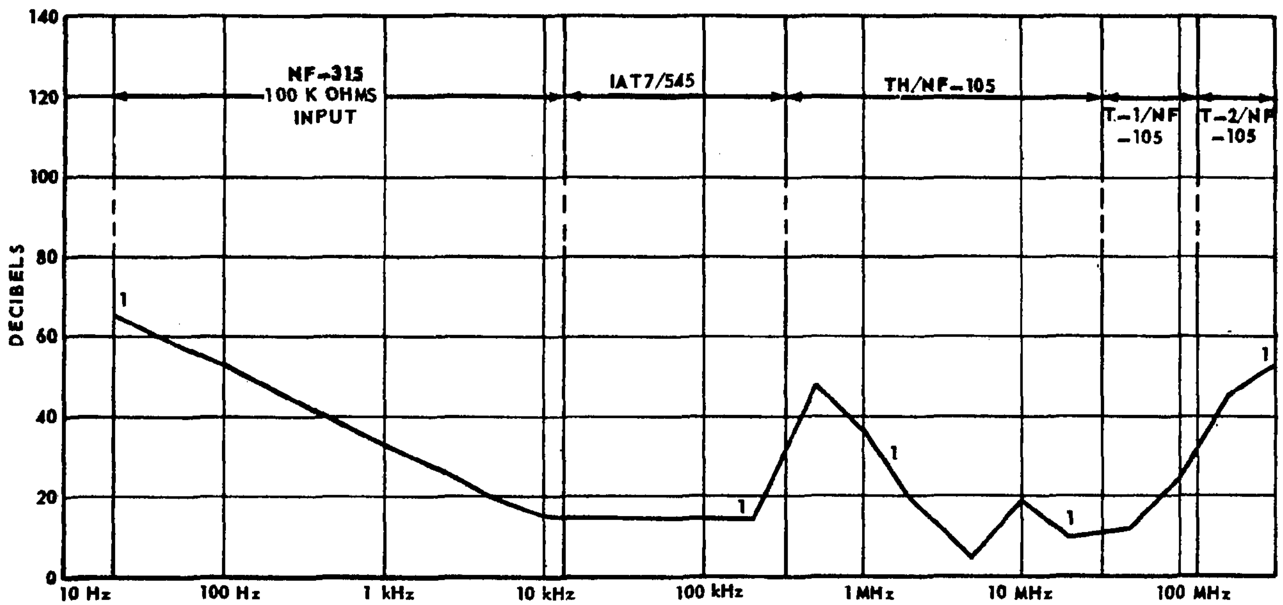



FIGURE 3-28. Magnetic crosstalk - adjacent center conductors - stacked unshielded cables.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTORS IN TWO UNSHIELDED CABLES.

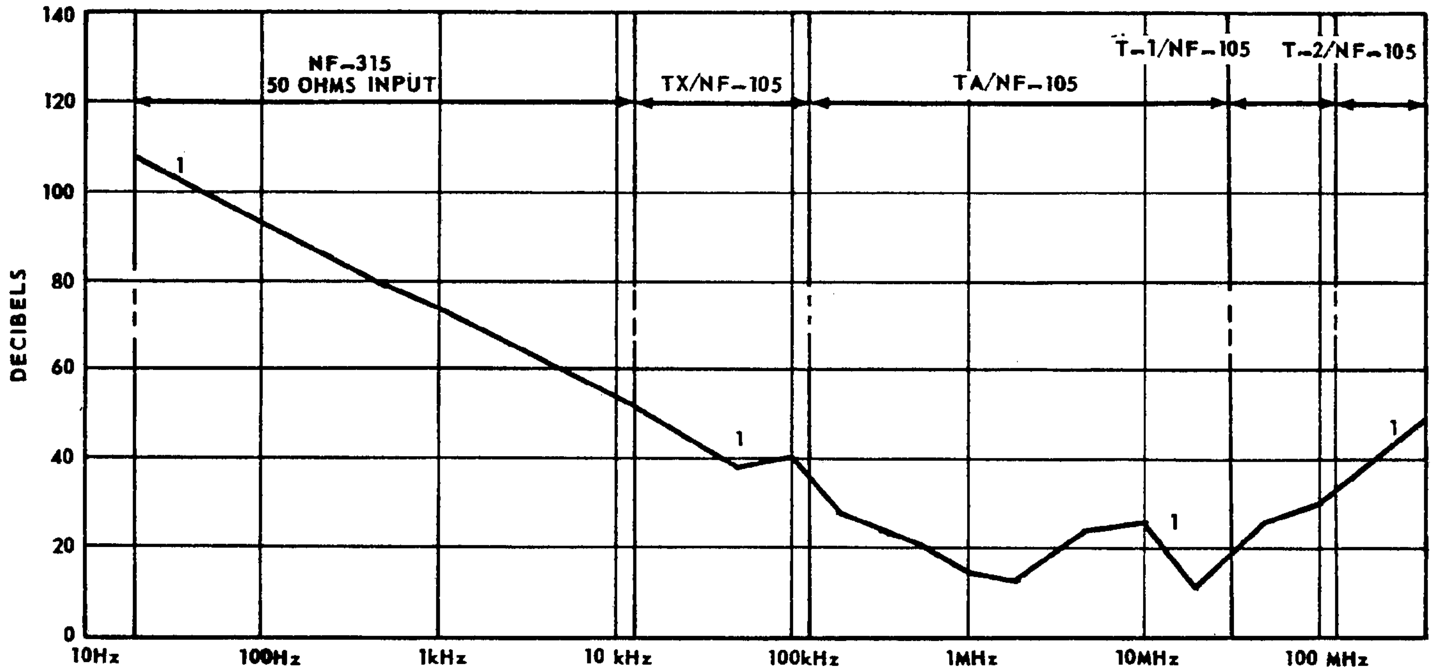
METHODE PD825HC252540403H0023, 6 FT. LONG CABLES, 4 FT. IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS LOAD IMPEDANCE ∞-OHMS.

SENSING CABLE-SOURCE IMPEDANCE ∞-OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 ——— UNSHIELDED CABLE MEASUREMENT

FIGURE 3-29. Electrostatic crosstalk - adjacent center conductors - stacked unshielded cables.



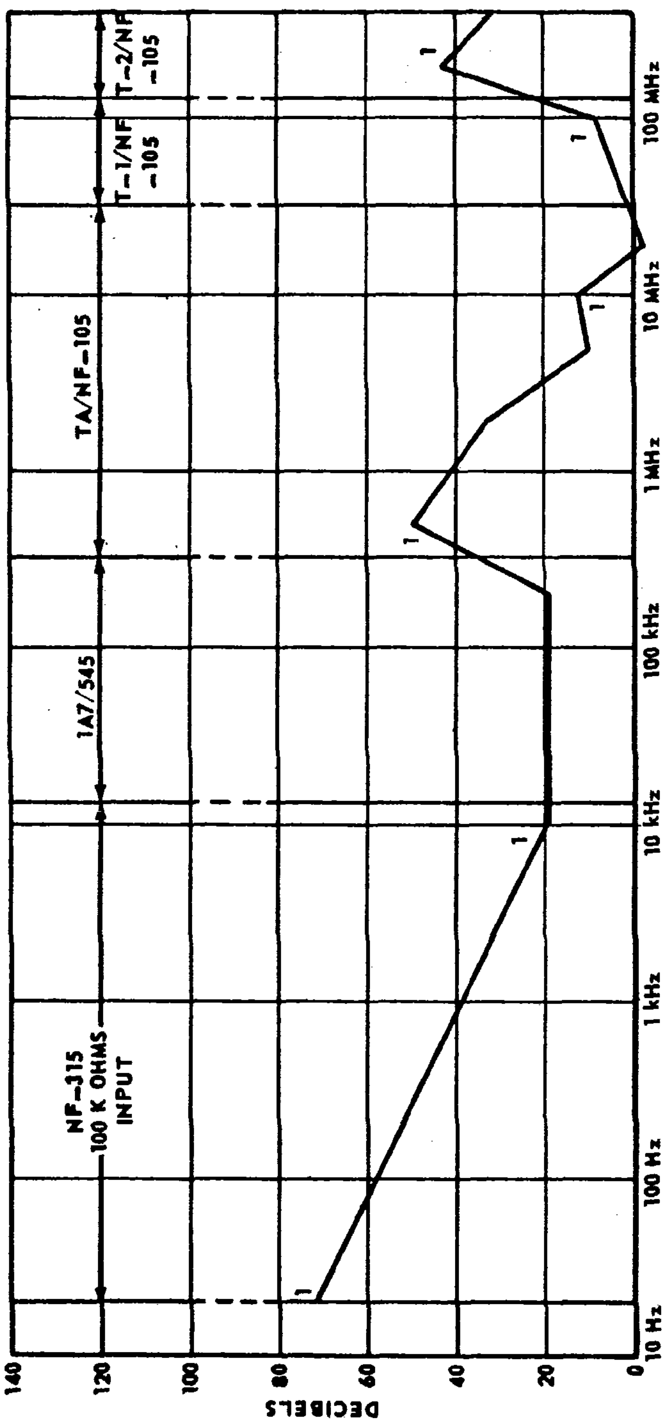

MAGNETIC COUPLING BETWEEN CENTER CONDUCTOR AND ADJACENT CONDUCTOR IN THE SAME UNSHIELDED CABLE.

METHODE PD825HC252540403H0023 6 FT. LONG CABLES.

EXCITED CONDUCTOR, SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHM .

SENSING CONDUCTOR, SOURCE IMPEDANCE 0-OHM , LOAD IMPEDANCE 50-OHMS.

FIGURE 3-30. Magnetic crosstalk - adjacent center conductors - same unshielded cables.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTOR AND ADJACENT CONDUCTOR IN THE SAME UNSHIELDED CABLE.

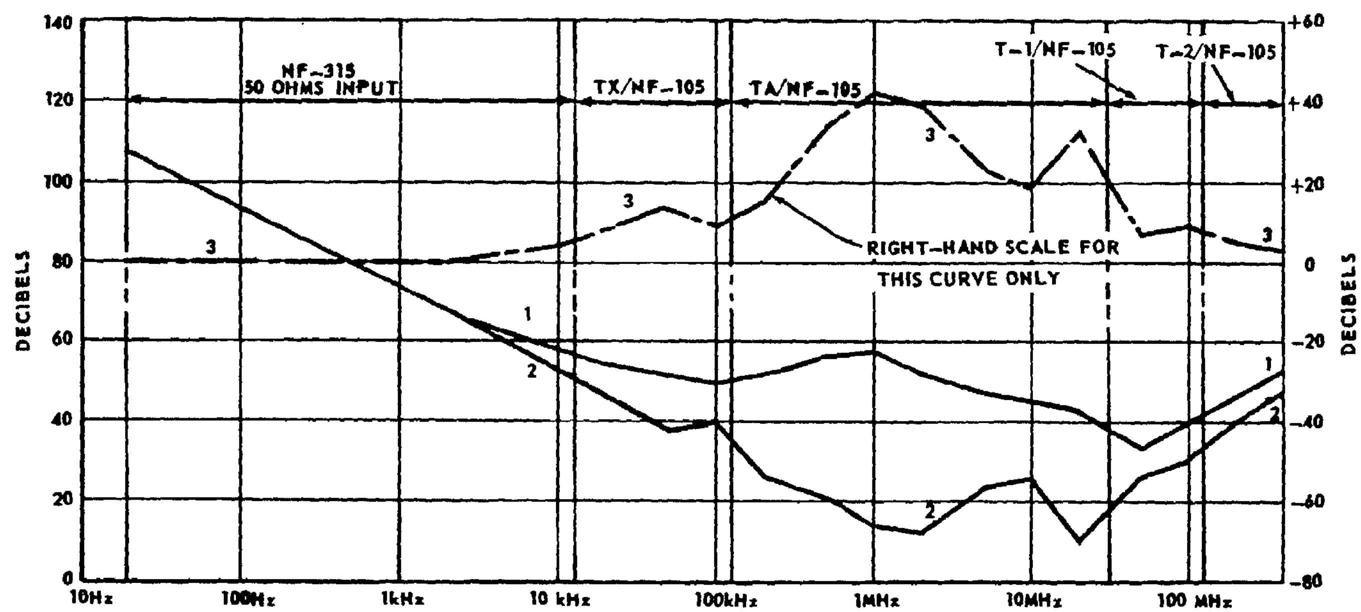
METHODE PD825MC2525404O3H0023 6 FT. LONG CABLES.

EXCITED CONDUCTOR, SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE ∞ OHMS.

SENSING CONDUCTOR, SOURCE IMPEDANCE ∞ OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 —— UNSHIELDED CABLE MEASUREMENT

FIGURE 3-31. Electrostatic crosstalk - adjacent center conductors - same unshielded cables.




MAGNETIC COUPLING BETWEEN CENTER CONDUCTOR AND ADJACENT CONDUCTOR IN THIS SAME NASA/METHODE SPEC 220 COPPER SHIELDED CABLE WITH OPEN EDGES. CABLE LENGTH 6 FEET.

EXCITED CABLE- SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHM

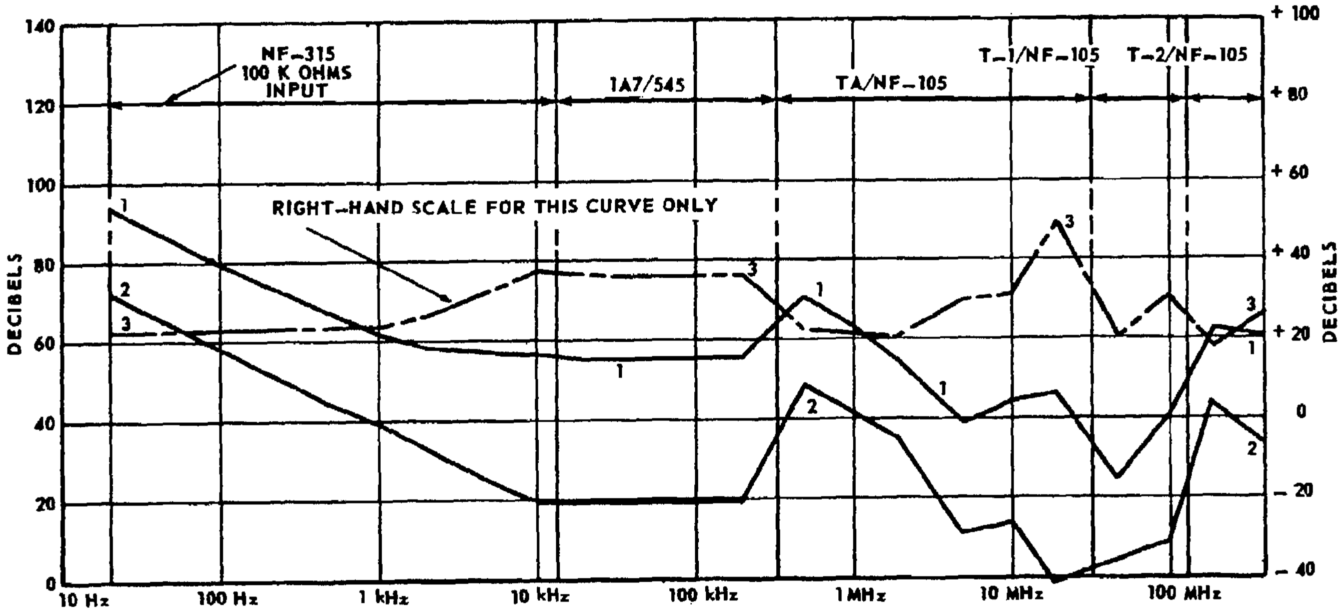
SENSING CABLE-SOURCE IMPEDANCE 0-OHM, LOAD IMPEDANCE 50-OHMS.

1 ——— SHIELDED CABLE MEASUREMENT

2 ——— UNSHIELDED CABLE (FIG. 3-30)

3 - - - SHIELDING EFFECTIVENESS

FIGURE 3-32. Magnetic crosstalk - adjacent center conductors - same solid copper shielded cable - open edges.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTOR AND ADJACENT CONDUCTOR IN THE SAME NASA/METHODE SPEC 220 SHIELDED CABLE WITH OPEN EDGES. CABLE LENGTH 6 FEET.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE ∞ OHMS.

SENSING CABLE - SOURCE IMPEDANCE ∞ OHMS, LOAD IMPEDANCE AS NOTED.

1 ———— SHIELDED CABLE MEASUREMENT

2 ———— UNSHIELDED CABLE (FIG. 3-31)

3 — — — SHIELDING EFFECTIVENESS

FIGURE 3-33. Electrostatic crosstalk - adjacent center conductors - same solid copper shielded cable - open edges.

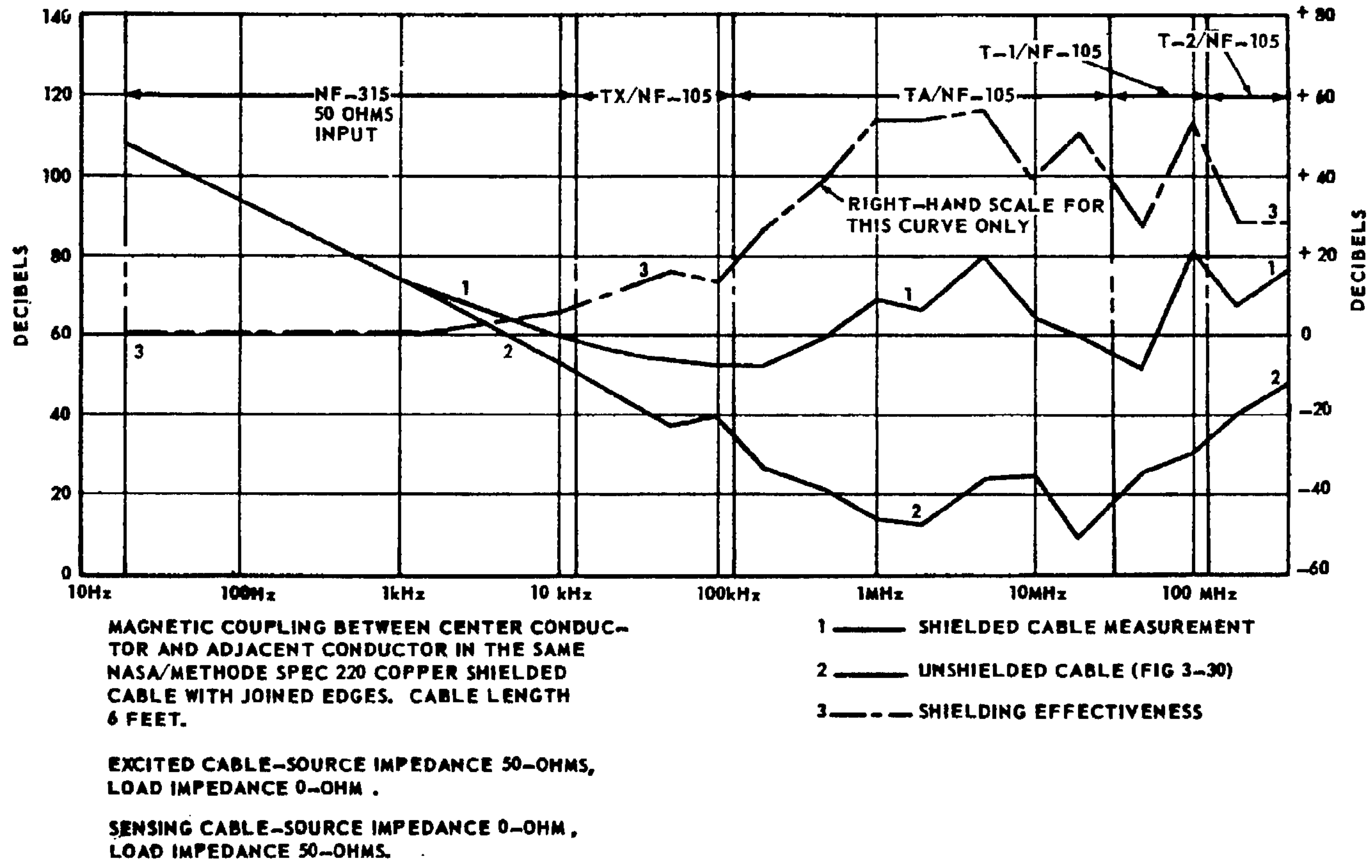



FIGURE 3-34. Magnetic crosstalk - adjacent center conductors - same solid copper shielded cable - joined edges.

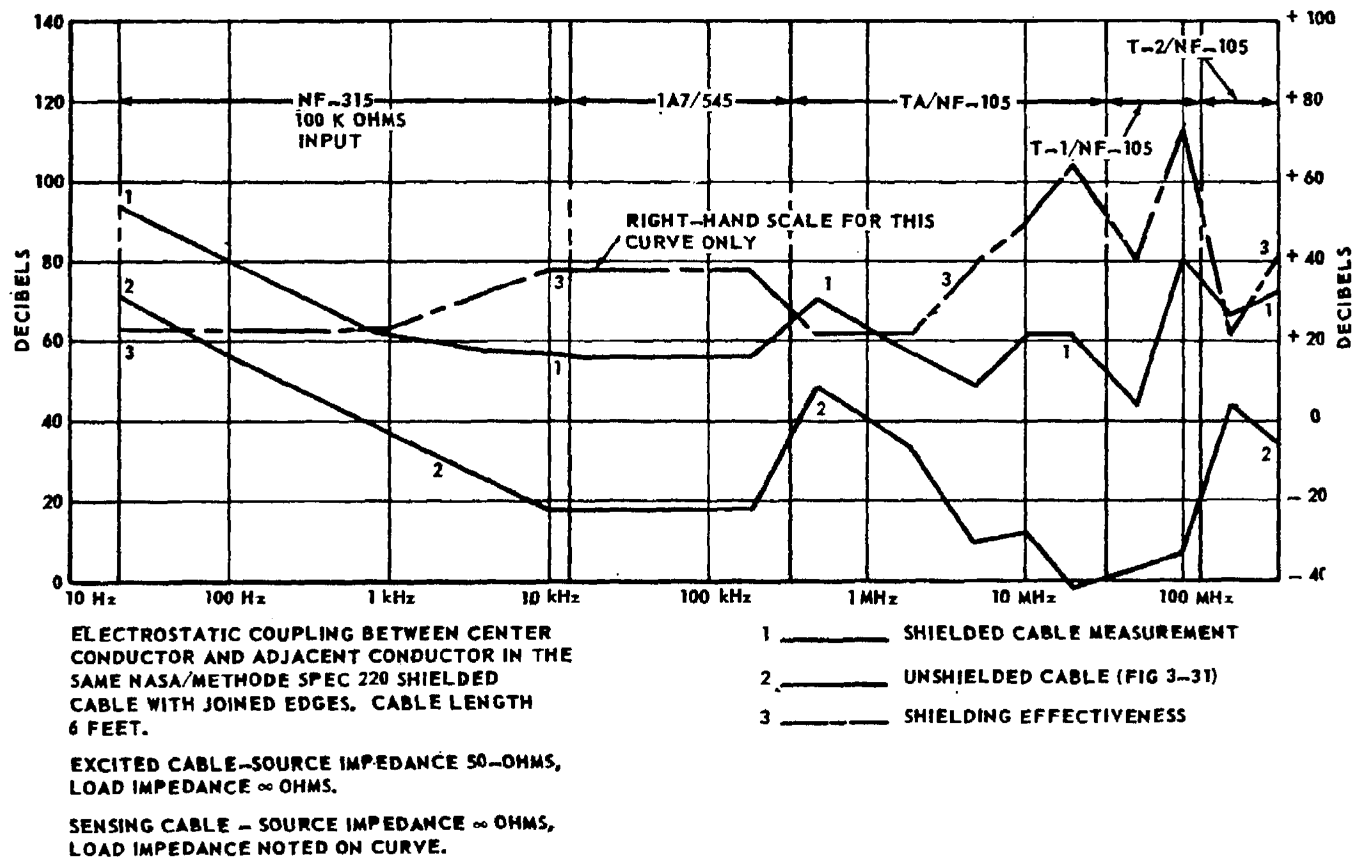
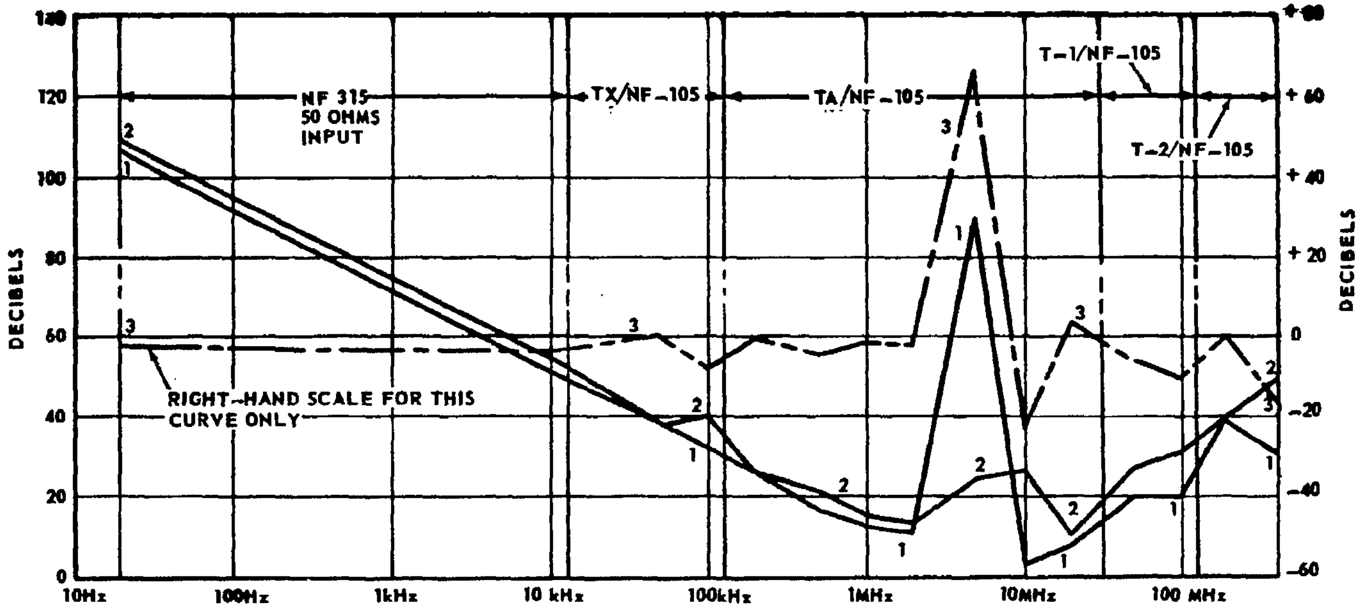



FIGURE 3-35. Electrostatic crosstalk - adjacent center conductors - same solid copper shielded cable - joined edges.



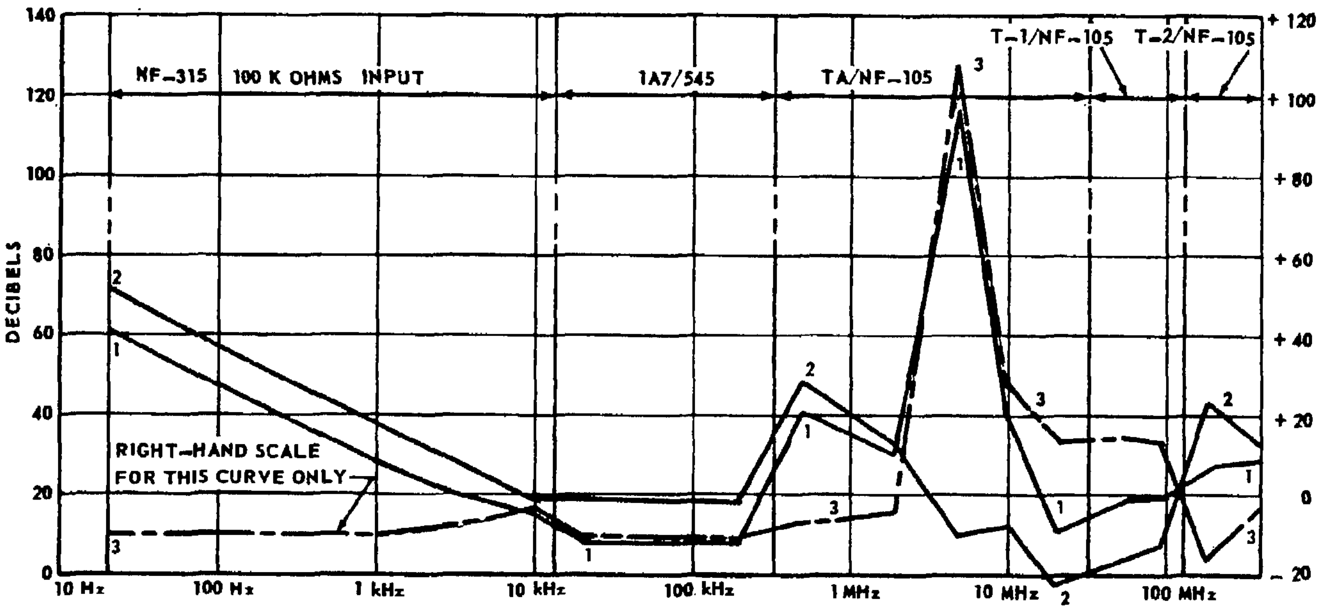

MAGNETIC COUPLING BETWEEN ONE CONDUCTOR OF A SUPRENANT 25-R04538 (300) JA 10 (15) SHIELDED TWISTED PAIR ROUND WIRE CABLE AND THE SECOND CONDUCTOR IN THE SAME CABLE. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CONDUCTOR - SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHM.

SENSING CONDUCTOR - SOURCE IMPEDANCE 0-OHM, LOAD IMPEDANCE 50-OHMS.

1 ———— SHIELDED CABLE MEASUREMENT
2 ———— UNSHIELDED CABLE (FIG 3-30)
3 — — — SHIELDING EFFECTIVENESS

FIGURE 3-36. Magnetic crosstalk - adjacent conductors - same woven copper shielded RWC.




ELECTROSTATIC COUPLING BETWEEN ONE CONDUCTOR OF A SUPRENANT 2S–HC4538 (300) JA 10 (15) SHIELDED TWISTED PAIR ROUND WIRE CABLE AND THE SECOND CONDUCTOR IN THE SAME CABLE. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CONDUCTOR – SOURCE IMPEDANCE 50–OHMS, LOAD IMPEDANCE OHMS.

SENSING CONDUCTOR – SOURCE IMPEDANCE OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 —— SHIELDED CABLE MEASUREMENT
2 —— UNSHIELDED CABLE (FIG 3–31)
3 – – – SHIELDING EFFECTIVENESS

FIGURE 3-37. Electrostatic crosstalk - adjacent conductors - same woven copper shielded RWC.

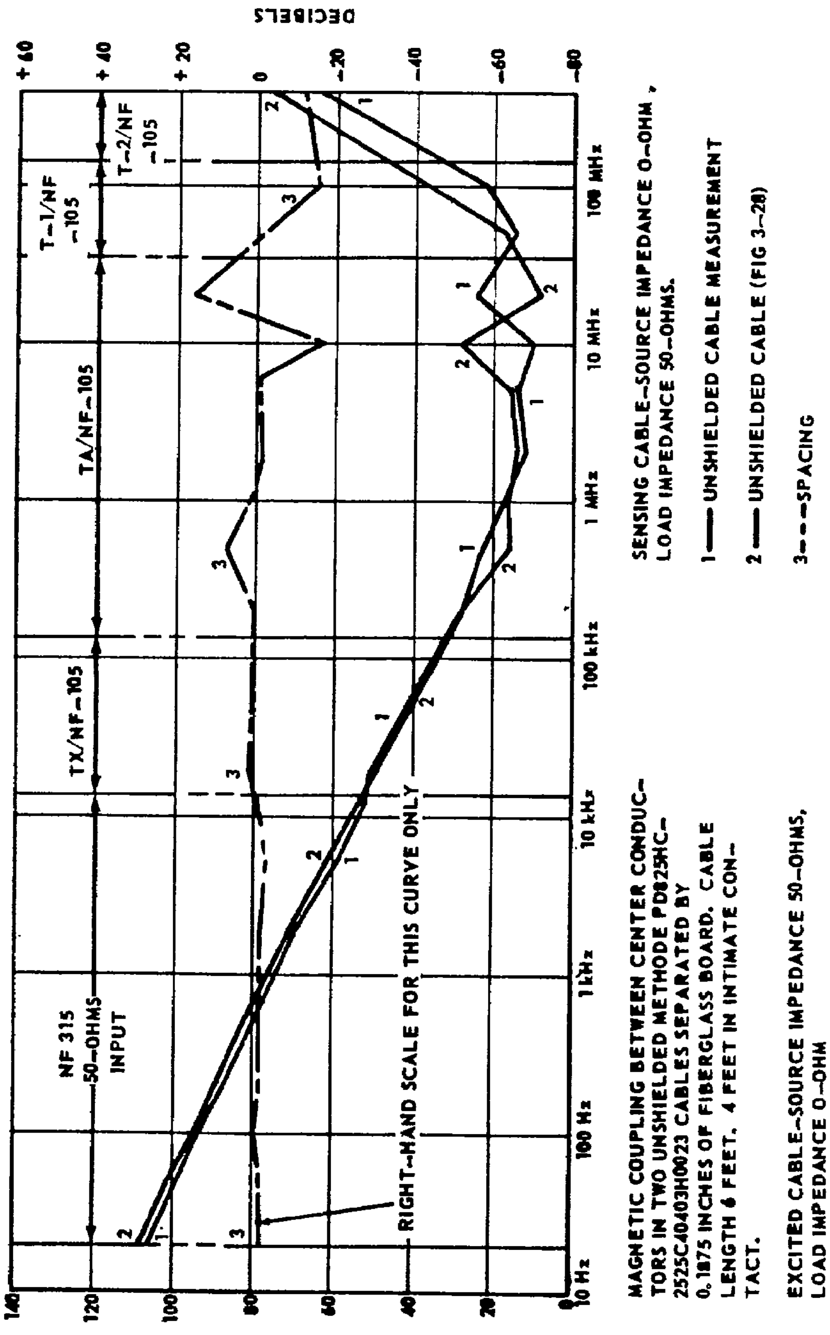
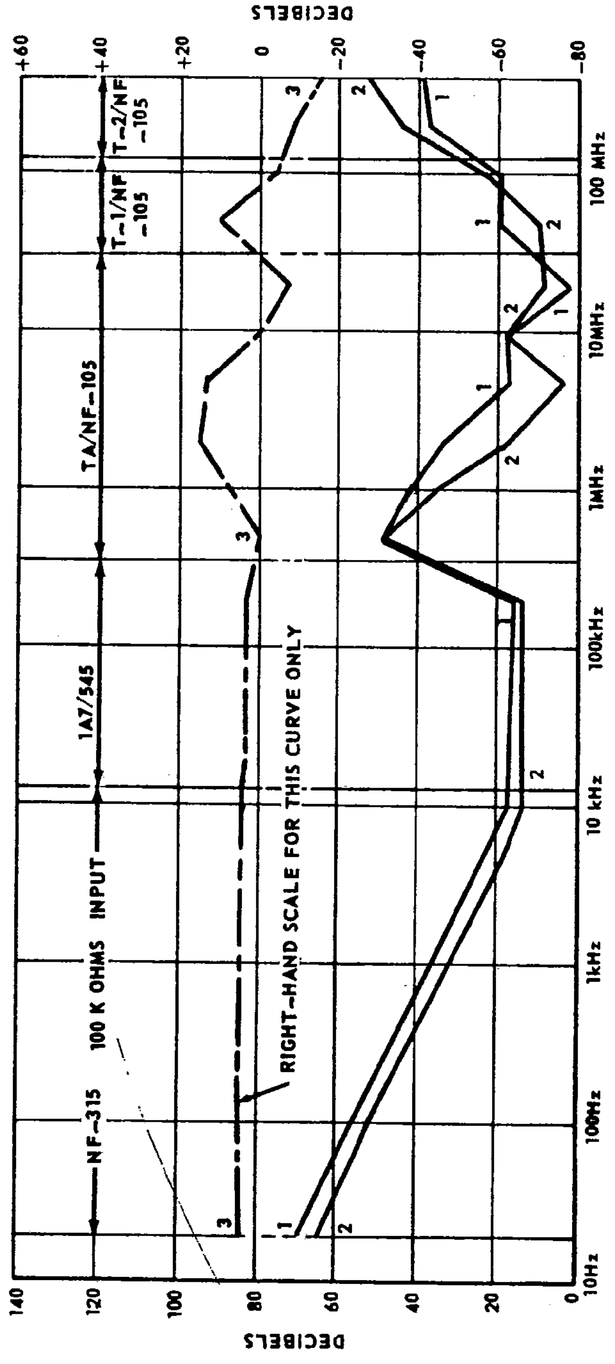



FIGURE 3-38. Magnetic crosstalk reduction - stacked unshielded cables - 0.1875-inch dielectric spacer.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTORS IN TWO UNSHIELDED METHODE PD-825HC2525C4040 3H0023 CABLES SEPARATED BY 0.1875 INCH OF FIBERGLASS BOARD. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE ∞OHMS.

SENSING CABLE-SOURCE IMPEDANCE ∞ OHMS, LOAD IMPEDANCE NOTED ON CURVE.

SEPARATING

1 —— UNSHIELDED CABLE MEASUREMENT

2 —— UNSHIELDED CABLE (FIG 3-29)

3 - - - SPACING

FIGURE 3-39. Electrostatic crosstalk reduction - stacked unshielded cables - 0.1875-inch dielectric spacer.

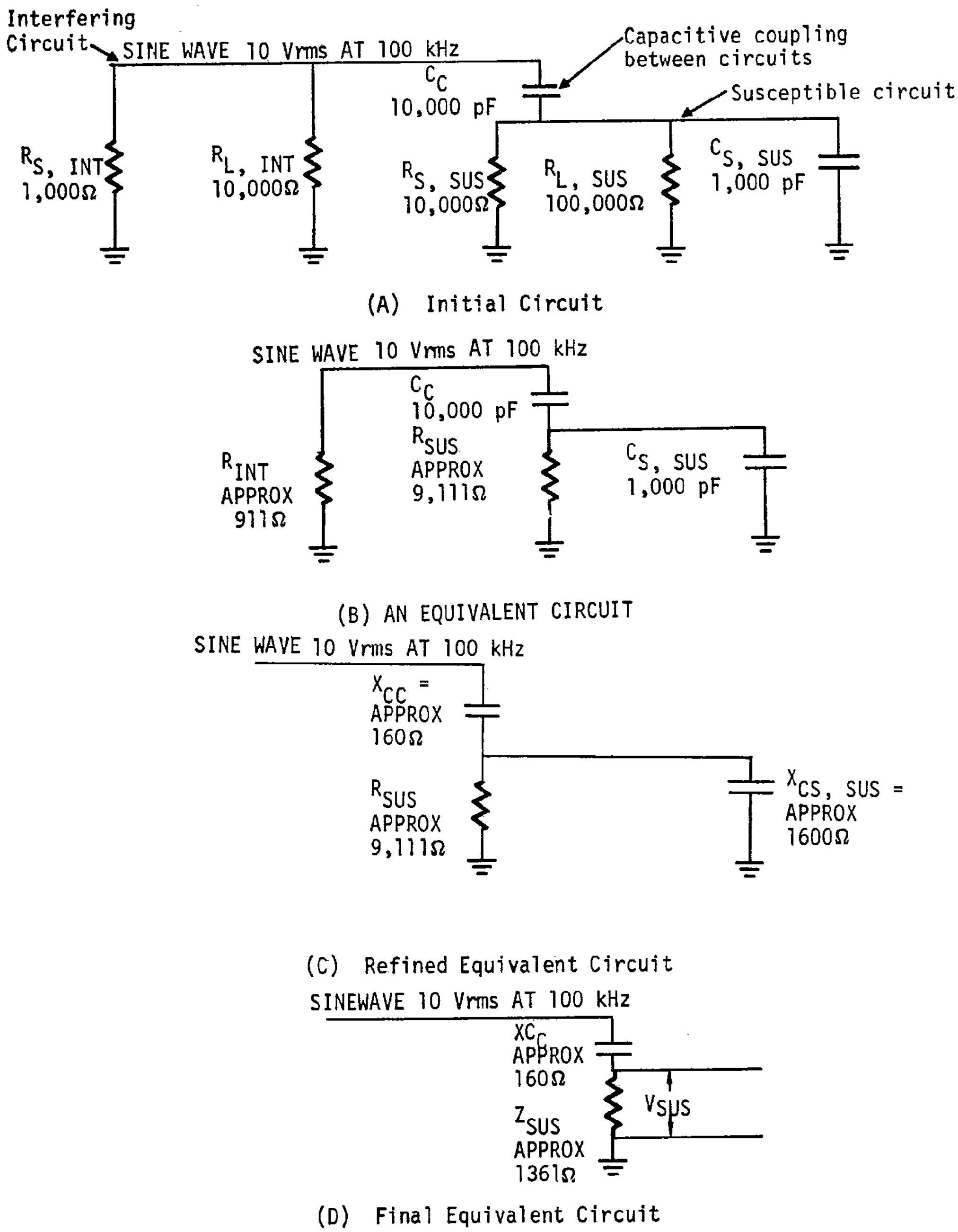
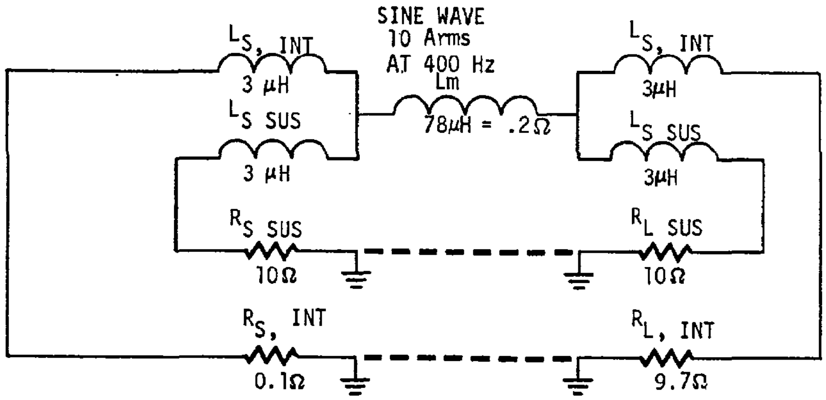
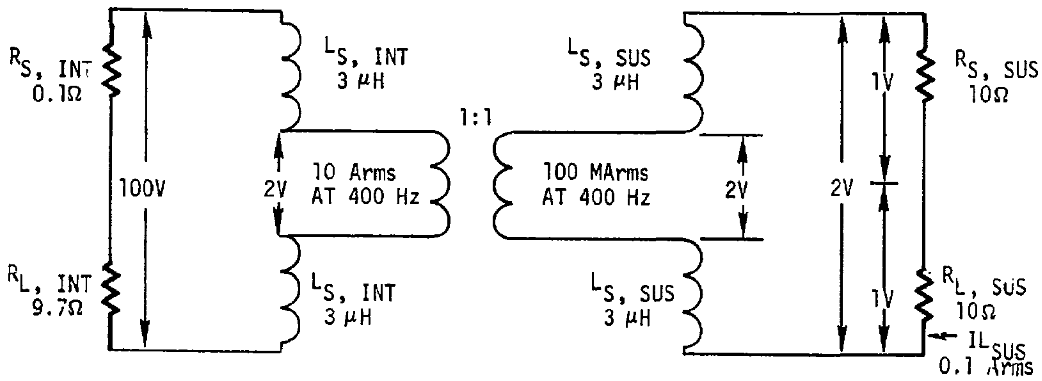



FIGURE 3-40. Electric-field crosstalk.




(A) Initial Circuit




(B) Equivalent Circuit

$$V_{Lm} = I_{INT} \times XL_m$$

$$V_{Lm} = 10 \times 0.2\Omega$$

$$V_{Lm} = 2V$$

$$I_{SUS} = \frac{2V}{R_{S,SUS} + R_{L,SUS}}$$

$$I_{SUS} = \frac{2V}{10\Omega + 10\Omega}$$

$$I_{SUS} = \frac{2V}{20\Omega} = 0.10 \text{ Arms}$$

**FIGURE 3-41.** Magnetic-field crosstalk.

$R_{L,\ INT} = 9.7\Omega$

$R_{L,\ SUS} = 10\Omega$

$R_{S,\ INT} = 0.1\Omega$

$R_{S,\ SUS} = 10\Omega$

$L_{S,\ INT} = 3\ \mu H + 3\ \mu H = 6\ \mu H$

$L_{S,\ SUS} = 3\ \mu H + 3\ \mu H = 6\ \mu H$

$X_{Lm} = 0.2\Omega$

$L_m = 78\ \mu H$

$V_{INT} = 100V$

$F_{INT} =$ 400-Hz sine wave

In this case, a signal of 0.10 ampere was generated in the susceptible circuit. Reactance values can be read from a reactance chart or calculated. Phase shift was not considered because the coupling impedance and susceptible circuit impedance were widely different values. The loop or series inductance $L_S$ was negligible at 400 Hz and was not considered. Reasonable approximations were used for convenience because a high degree of precision was not required.

3.2.3.2.3 Shielding of Cables. Shielded conductors can reduce the spurious coupling of electric and/or magnetic energy from one conductor to another, compared to equivalent unshielded conductors. This reduction in coupling occurs when the electromagnetic environments of the two conductors are isolated by insertion of a shielding medium between the conductors.

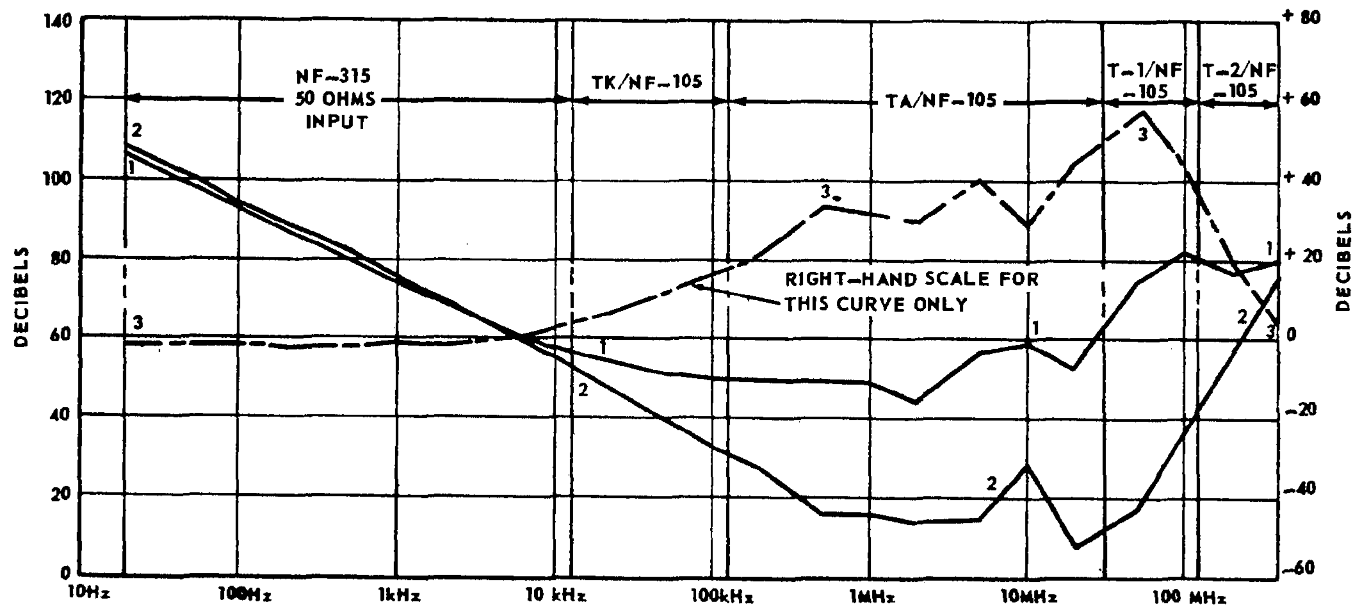
The figure of merit used to measure shielding is "shielding effectiveness." This is given in terms of decibels reduction in received energy at a conductor when a shielded conductor is substituted for the equivalent unshielded conductor.

Measured shielding effectiveness between centrally located conductors in stacked cables for a variety of cable configurations is shown in Figures 3-42 through 3-51.

3.2.3.2.3.1 Detailed Shield Design. The shielding effectiveness of any given solid-shield configuration versus frequency for magnetic, electric, or plane-wave fields can be plotted quickly with the aid of the nomographs of Figures 3-52 through 3-55. Similar nomographs with reduced ranges were originally prepared by Robert B. Cowdell of the Genisco Technology Corporation. The following paragraphs explain the use of Figures 3-52 through 3-55.

3.2.3.2.3.2 Practical Shield Design. Shielding effectiveness, or attenuation, is the sum of the reflection and absorption losses of the shield material selected. Reflection losses are not critically dependent on material thickness, but are dependent on a low-reluctance, circumferential, magnetic path through the shield. Absorption losses are proportional to material thickness. After the reflection loss versus frequency is plotted for the type of field under consideration, it is possible to determine whether additional losses are required at some frequencies. If the reflection losses are inadequate, absorption losses must be utilized to achieve the desired level of shield attenuation and the thickness of the material becomes significant.

3.2.3.2.3.3 Plane-Wave Reflection Losses. Plane-wave reflection losses are dependent only on the type of shield material selected, and on frequency. A straight edge, connecting a point on the G/u (material) scale to a point on the frequency scale, will also intersect the dB (attenuation) scale to provide the plane-wave reflection loss (Fig. 3-52). Under most conditions, this reflection loss will provide adequate plane-wave attenuation; therefore, absorption losses are not significant, and the thinnest material capable of meeting nonelectrical requirements is satisfactory. Thickness-dependent absorption losses are rarely required.




MAGNETIC COUPLING BETWEEN CENTER CONDUCTORS IN ONE UNSHIELDED METHODE PD825CH-252540403H0023 CABLE AND ONE NASA/METHODE SPEC 220 COPPER SHIELDED CABLE WITH OPEN EDGES. EXCITED CABLE WAS SHIELDED.

CABLES WERE 6 FEET LONG CABLES. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHM .

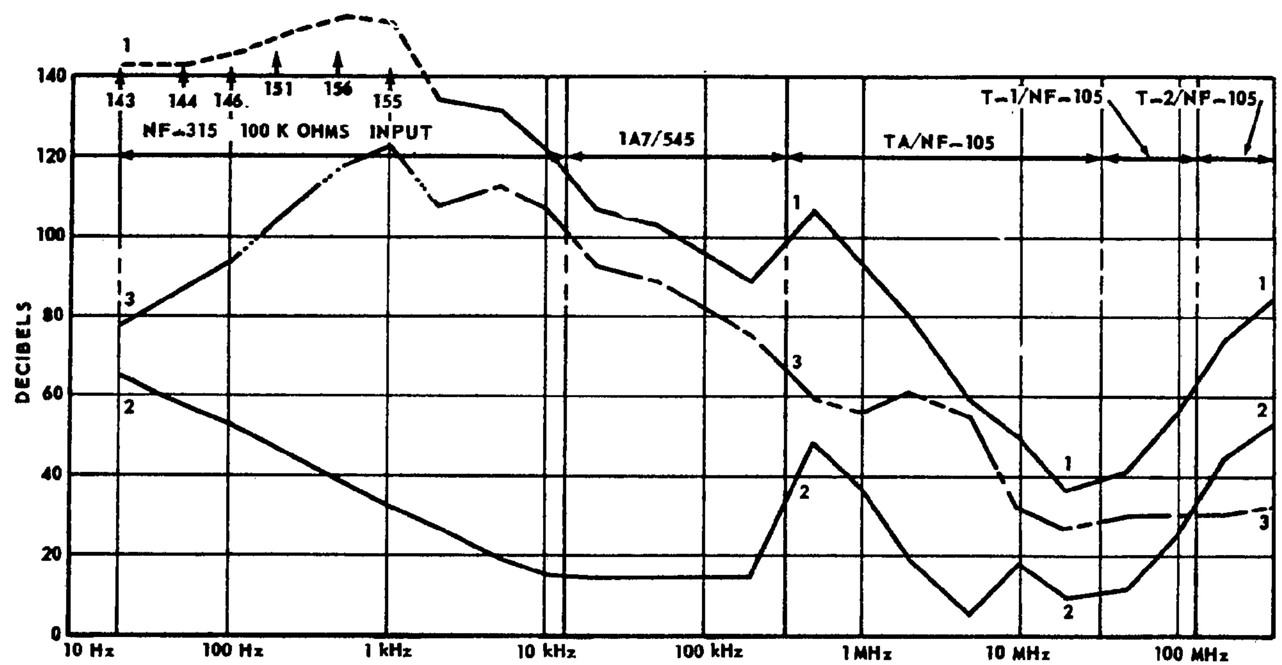
SENSING CABLE-SOURCE IMPEDANCE 0-OHM LOAD IMPEDANCE 50-OHMS.

1 UNSHIELDED CABLE MEASUREMENT

2 UNSHIELDED CABLE (FIG 3-28)

3 SHIELDING EFFECTIVENESS

FIGURE 3-42. Magnetic shielding effectiveness - solid copper shielded cable - open edges.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTORS IN ONE UNSHIELDED METHODE PD825CH252540403H0023 CABLE AND ONE NASA/METHODE SPEC 220 COPPER SHIELDED CABLE WITH OPEN EDGES. EXCITED CABLE WAS SHIELDED.

CABLES WERE 6 FEET LONG CABLES. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE – SOURCE IMPEDANCE 50 OHMS, LOAD IMPEDANCE ∞ OHMS.

SENSING CABLE – SOURCE IMPEDANCE ∞OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 ——— UNSHIELDED CABLE MEASUREMENT

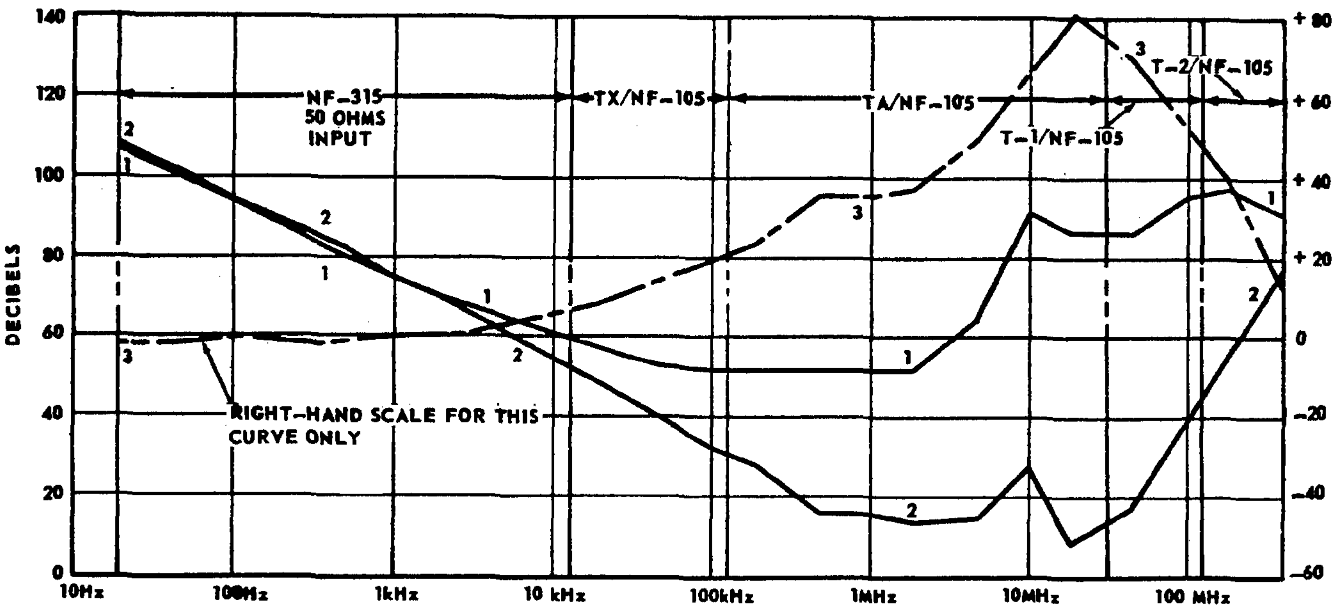
1 – – – – DYNAMIC RANGE OF INSTRUMENTATION

2 ——— UNSHIELDED CABLE (FIG 3–29)

3 — — — SHIELDING EFFECTIVENESS

3 —·-·-· SHIELDING EFFECTIVENESS IN EXCESS OF VALUES SHOWN

FIGURE 3-43. Electrostatic shielding effectiveness - solid copper shielded cable - open edges.




MAGNETIC COUPLING BETWEEN CENTER CONDUCTORS IN ONE UNSHIELDED METHODE PD825CH-252540403H0023 CABLE AND ONE NASA/METHODE SPEC 220 COPPER SHIELDED CABLE WITH JOINED EDGES. EXCITED CABLE WAS SHIELDED.

CABLES WERE 6 FEET LONG CABLES, 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHMS.

SENSING CABLE-SOURCE IMPEDANCE 0-OHMS, LOAD IMPEDANCE 50-OHMS.

1 ———— UNSHIELDED CABLE MEASUREMENT

2 ———— UNSHIELDED CABLE (FIG 3-28)

3 — — — SHIELDING EFFECTIVENESS

FIGURE 3-44. Magnetic shielding effectiveness - solid copper shielded cable - joined edges.

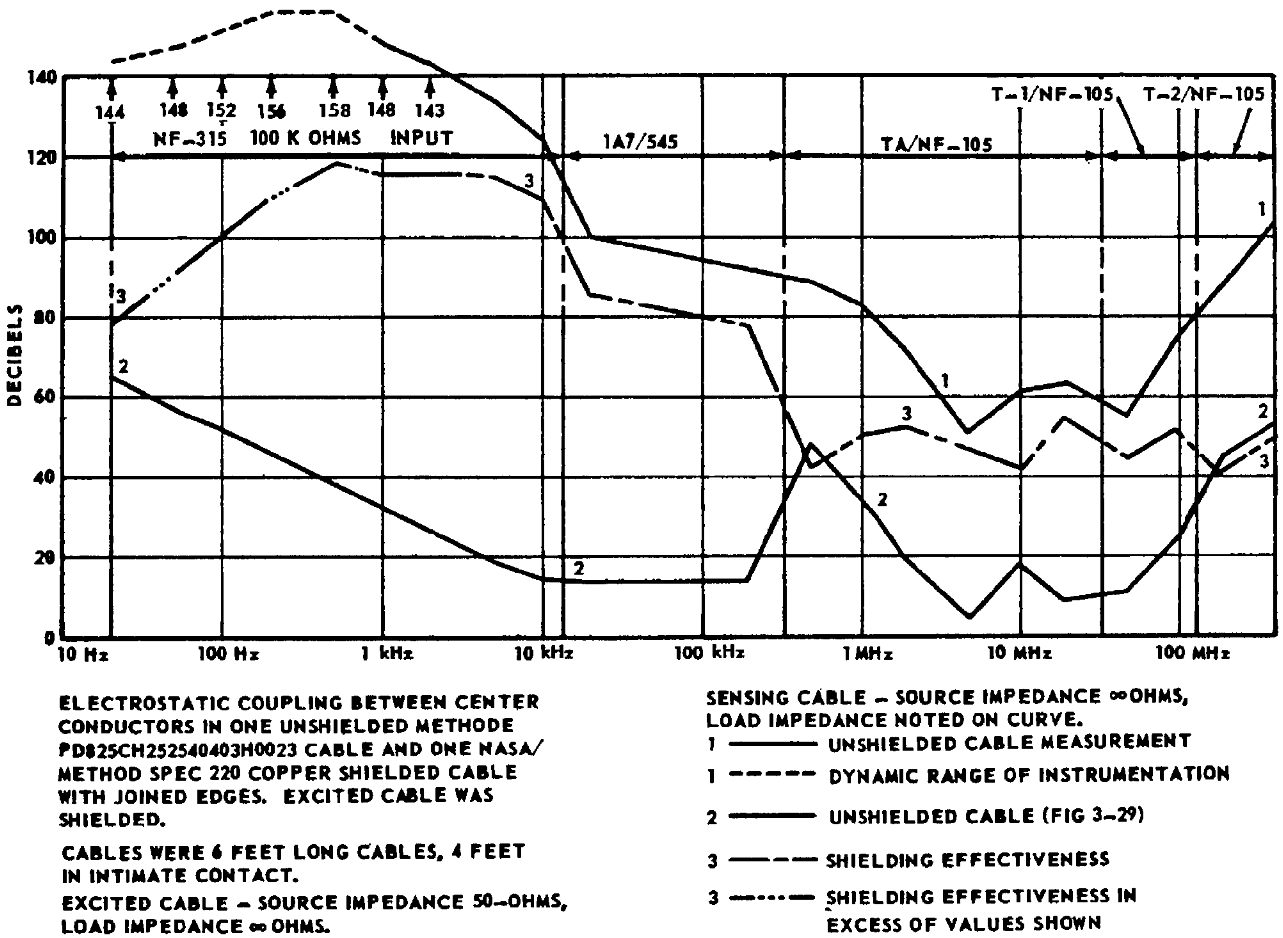
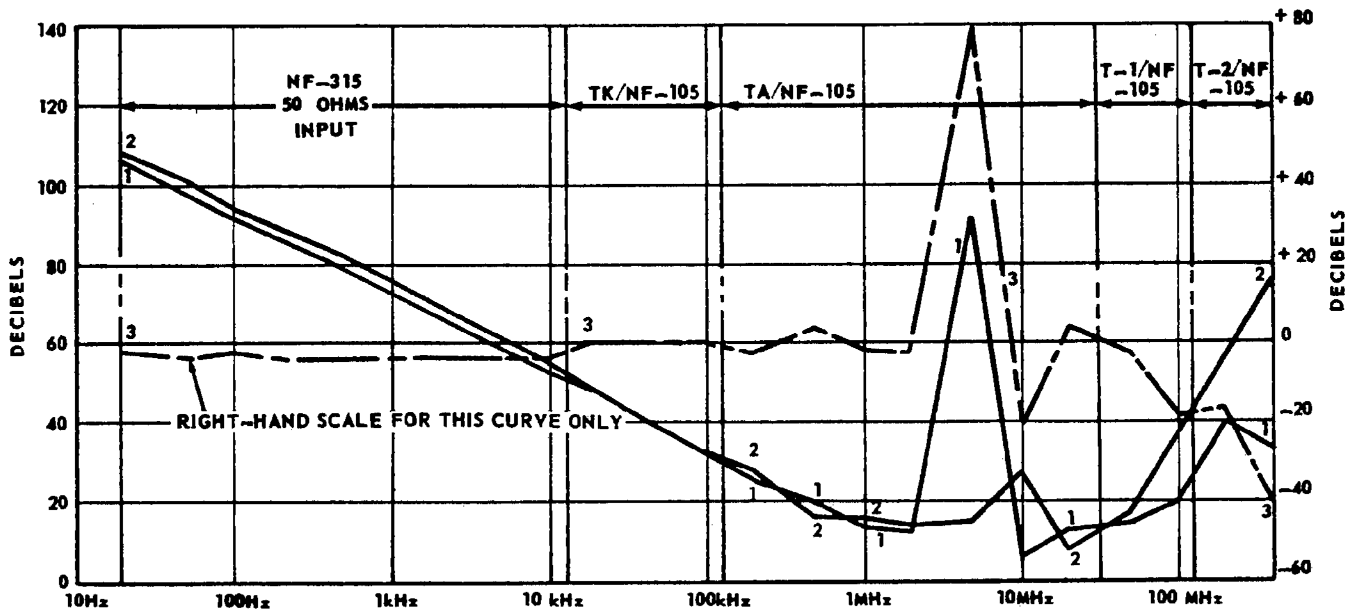



FIGURE 3-45. Electrostatic shielding effectiveness - solid copper shielded cable - joined edges.




MAGNETIC COUPLING BETWEEN ONE CONDUCTOR OF A SUPRENANT 2S-RC4538N (300) JA 10 (15) SHIELDED TWISTED PAIR BOUND WIRE CABLE AND AN ADJACENT 20-GUAGE SINGLE UNSHIELDED ROUND CONDUCTOR. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0 OHM .

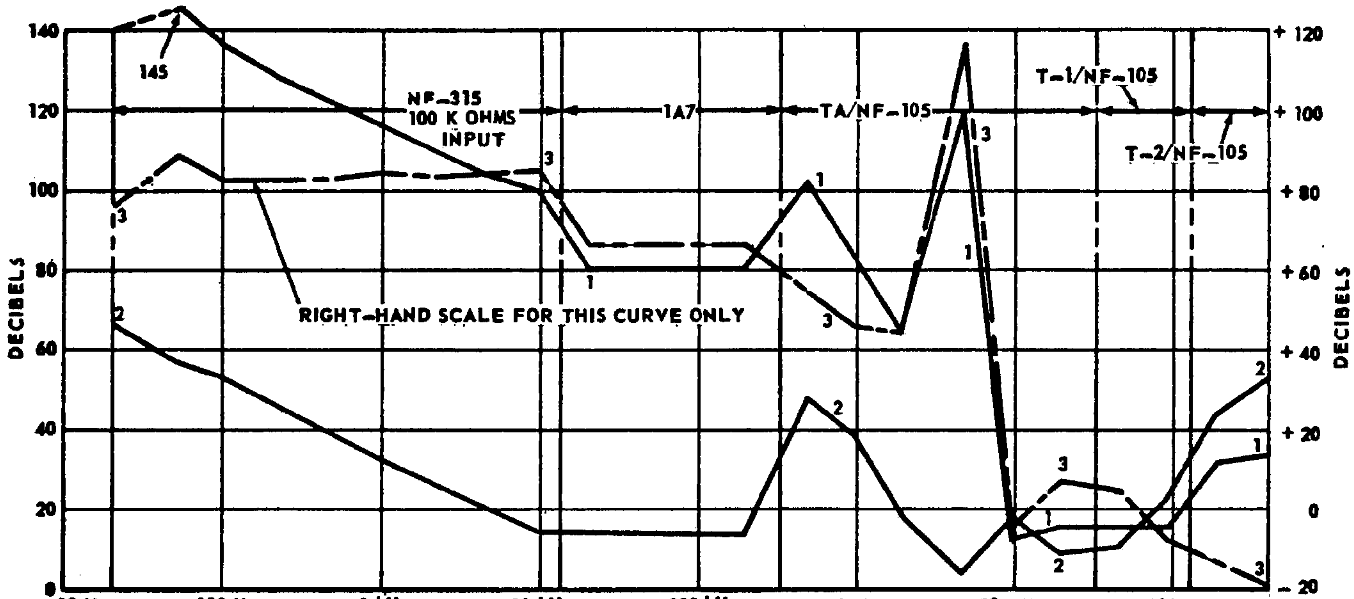
SENSING CABLE-SOURCE IMPEDANCE 0 OHM LOAD IMPEDANCE 50-OHMS.

1 —— UNSHIELDED CABLE MEASUREMENT

2 —— UNSHIELDED CABLE (FIG 3-28)

3 - - - SHIELDING EFFECTIVENESS

FIGURE 3-46. Magnetic shielding effectiveness - woven copper shielded cable - RWC.



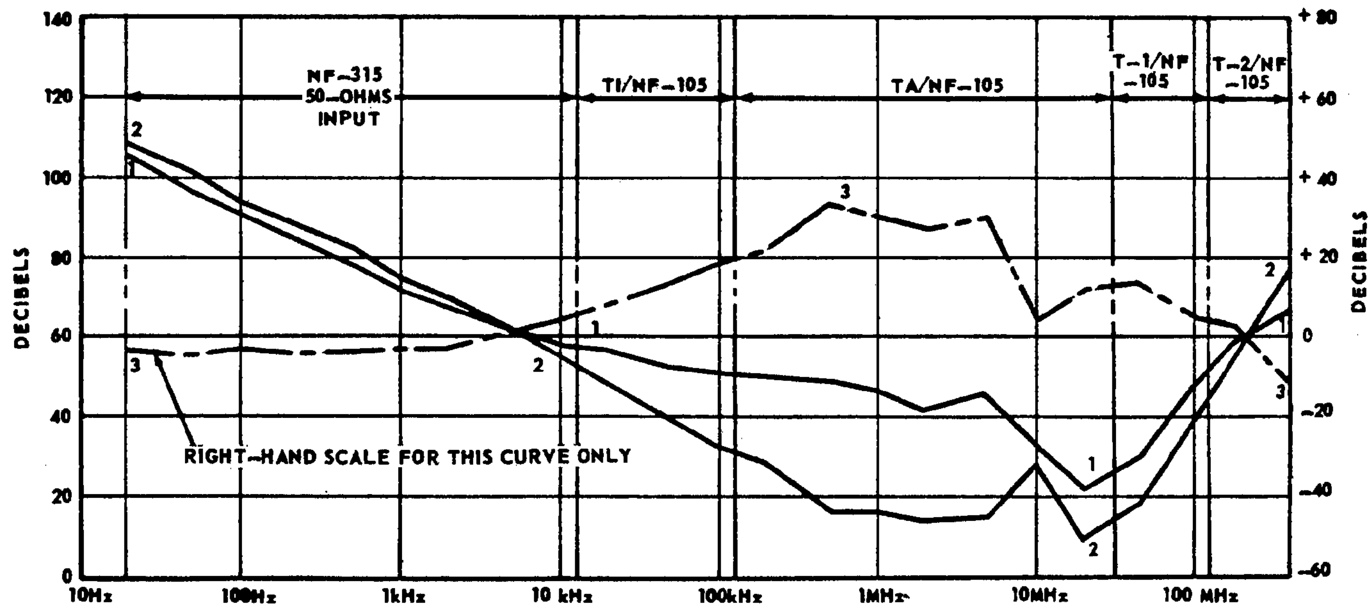

ELECTROSTATIC COUPLING BETWEEN ONE CONDUCTOR OF A SUPRENANT 2S-RC4538N (300)' JA 10 (15) SHIELDED TWISTED PAIR ROUND WIRE CABLE AND AN ADJACENT 20-GUAGE SINGLE UNSHIELDED ROUND CONDUCTOR. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50 OHMS, LOAD IMPEDANCE ∞ OHMS.

SENSING CABLE-SOURCE IMPEDANCE ∞ OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 ——— SHIELDED CABLE MEASUREMENT
1 - - - - DYNAMIC RANGE OF INSTRUMENTATION
2 ——— UNSHIELDED CABLE (FIG 3-29)
3 — — — SHIELDING EFFECTIVENESS
3 — · · — SHIELDING EFFECTIVENESS IN EXCESS OF VALUES SHOWN DUE TO TO INSTRUMENTATION LIMITATION

FIGURE 3-47. Electrostatic shielding effectiveness - woven copper shielded cable - RWC.




MAGNETIC COUPLING BETWEEN CENTER CONDUCTOR IN AN UNSHIELDED METHODE PD825HC25-25C40403H0023 CABLE AND A SECOND CENTER CONDUCTOR IN A SECOND UNSHIELDED METHODE PD825HC2525C40403H0023 CABLE. CABLES ISOLATED BY A 2-INCH WIDE 1 MIL THICK COPPER FOIL. CABLE LENGTH 6 FEET. 4 FEET IN INITIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHMS.

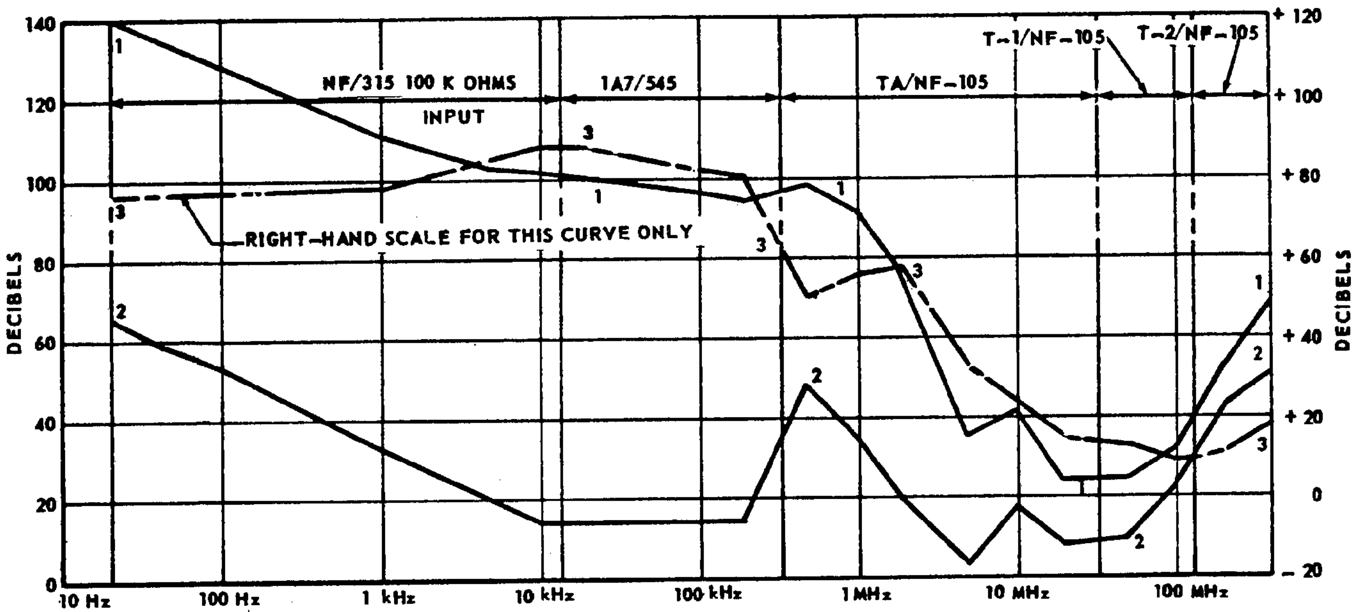
SENSING CABLE-SOURCE IMPEDANCE 0-OHMS, LOAD IMPEDANCE 50-OHMS.

1 —— UNSHIELDED CABLE MEASUREMENT

2 —— UNSHIELDED CABLE (FIG 3-28)

3 – – – SHIELDING EFFECTIVENESS

FIGURE 3-48. Magnetic shielding effectiveness - stacked unshielded cables - metallic spacer - one side.



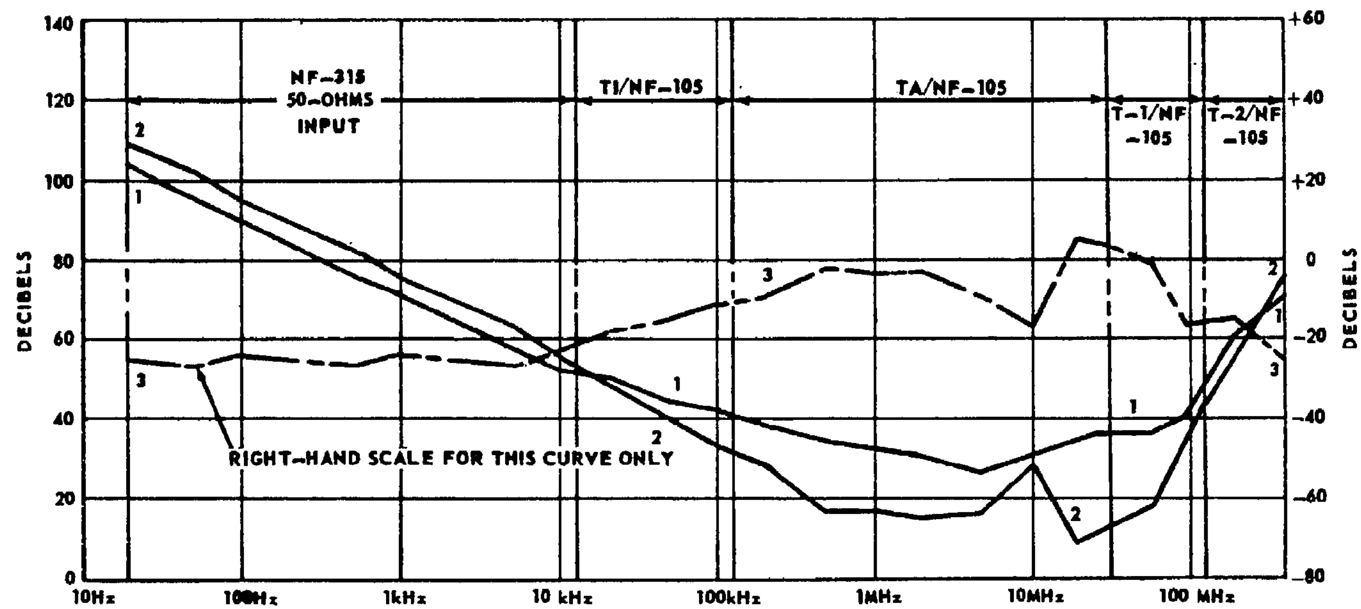

MAGNETIC COUPLING BETWEEN CENTER CONDUCTOR IN AN UNSHIELDED METHODE PD825HC25-25C40403H0023 CABLE AND A SECOND CENTER CONDUCTOR IN A SECOND UNSHIELDED METHODE PD825HC2525C40403H0023 CABLE. CABLES ISOLATED BY A 2-INCH WIDE 1-MIL THICK COPPER FOIL. CABLE LENGTH 6 FEET. 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE -SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0-OHM.

SENSING CABLE-SOURCE IMPEDANCE 0-OHMS, LOAD IMPEDANCE NOTED ON CURVE

1 ——— UNSHIELDED CABLE MEASUREMENT
1 – – – DYNAMIC RANGE OF INSTRUMENTATION
2 ——— UNSHIELDED CABLE (FIG 3-29)
3 — — — SHIELDING EFFECTIVENESS
3 —·— — — SHIELDING EFFECTIVENESS IN EXCESS OF VALUES SHOWN DUE TO INSTRUMENTATION LIMITATIONS

FIGURE 3-49. Electrostatic shielding effectiveness - stacked unshielded cables - metallic spacer - one side.




MAGNETIC COUPLING BETWEEN CENTER CONDUCTOR IN A SOLID 4-79 PERMALLOY SHIELDED METHODE C252C040403H2723 CABLE WITH OPEN EDGES AND ALUMINUM NASA/DAC CONNECTORS, 5-FEET LONG WITH 3-INCH SHIELD GROUNDS AT BOTH ENDS, AND CENTER CONDUCTOR IN AN UNSHIELDED METHODE PD825HC2525C40403-H0023 CABLE 6 FEET LONG WITH 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE-SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE 0 OHM .

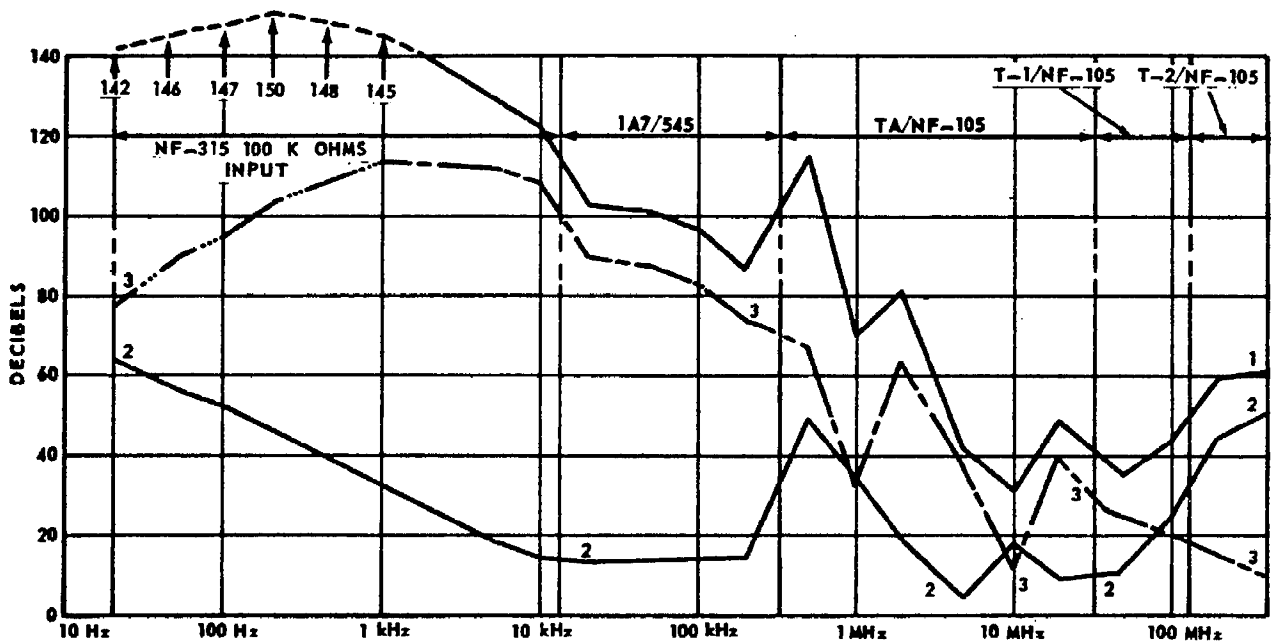
SENSING CABLE-SOURCE IMPEDANCE 0 OHM LOAD IMPEDANCE NOTED ON CURVE.

1 —— UNSHIELDED CABLE MEASUREMENT

2 —— UNSHIELDED CABLE

3 — — SHIELDING EFFECTIVENESS

FIGURE 3-50. Magnetic shielding effectiveness - solid 4-79 Permalloy shielded cable - NASA/DAC aluminum connectors.




ELECTROSTATIC COUPLING BETWEEN CENTER CONDUCTOR IN A SOLID 4-79 PERMALLOY SHIELDED METHODE C2520C40403H2723 CABLE WITH OPEN EDGES AND ALUMINUM NASA/DAC CONNECTOR, 5 FEET LONG WITH 3-INCH SHIELD GROUNDS AT BOTH ENDS, AND CENTER CONDUCTOR IN AN UNSHIELDED METHODE PD-825HC2525C40403H0023 CABLE 6 FEET LONG WITH 4 FEET IN INTIMATE CONTACT.

EXCITED CABLE - SOURCE IMPEDANCE 50-OHMS, LOAD IMPEDANCE ∞ OHMS.

SENSING CABLE - SOURCE IMPEDANCE ∞ OHMS, LOAD IMPEDANCE NOTED ON CURVE.

1 ——— UNSHIELDED CABLE MEASUREMENT

2 ——— UNSHIELDED CABLE (REF)

3 —-·— SHIELDING EFFECTIVENESS

FIGURE 3-51. Electrostatic shielding effectiveness - solid 4-79 Permalloy shielded cable - NASA/DAC aluminum connectors.

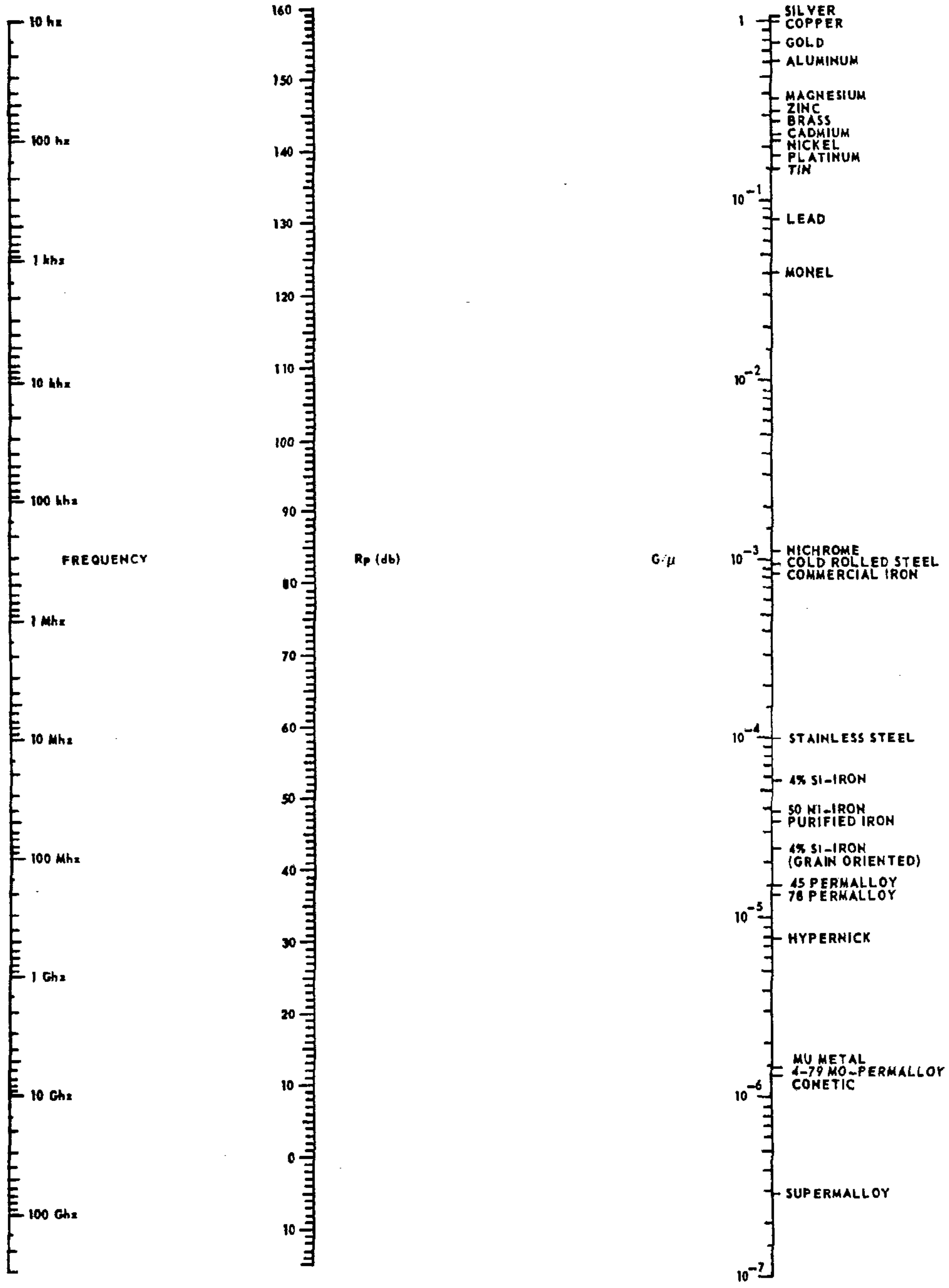



FIGURE 3-52. Plane-wave reflection losses.

3.2.3.2.3.4 Electric-Field Reflection Losses. Electric-field reflection losses are dependent only on the spacing between the source of interference and the shield, the type of shield material, and frequency. A straight edge, connecting a point on the inches (separation distance) scale to a point on the G/u (material) scale, will also intersect a point on the uncalibrated vertical scale (Fig. 3-53). A straight edge, connecting this point on the uncalibrated vertical scale to a point on the frequency scale, will also intersect the dB (attenuation) scale to provide the electric-field reflection loss. Under most conditions, this reflection loss will provide adequate electric-field attenuation; therefore, absorption losses are not significant, and the thinnest material capable of meeting nonelectrical requirements is satisfactory. Thickness-dependent absorption losses are rarely required.

3.2.3.2.3.5 Magnetic-Field Reflection Losses. Magnetic-field reflection losses are dependent only on the spacing between the source of interference and the shield, the type of shield material, and frequency. Shield thickness is not critical, but a low-reluctance circumferential path through the shield is required. A straight edge, connecting a point on the inches (separation distance) scale to a point on the G/u (material) scale (Fig. 3-54), will also intersect a point on the uncalibrated vertical scale. A straight edge, connecting this point on the uncalibrated vertical scale to a point on the frequency scale, will also intersect the dB (attenuation) scale, providing the magnetic-field reflection losses. Under most conditions, this reflection loss will not provide adequate magnetic-field attenuation; therefore, absorption losses are required and material thickness is significant, unless special high-permeability alloys are utilized.

The magnetic-field reflection loss, when plotted against frequency, has a peculiar shape when compared with the electric-field and plane-wave reflection-loss curves. Since the shield impedance is low compared with the high-impedance electric and plane-wave fields, and exhibits a rising characteristic but does not achieve an impedance match, these reflection losses have a magnitude inversely proportional to frequency. The magnetic field has a relatively low impedance that matches the shield impedance at some frequency in the useful spectrum, producing a magnetic reflection-loss null. At this null, the shield is magnetically transparent and unfortunately, most shields lack useful absorption-loss characteristics.

At low frequencies, the shield impedance is lower than the magnetic-field impedance but exhibits a rising characteristic; therefore, the impedance match between the shield and magnetic field improves as frequency increases, and a more efficient transfer of energy occurs. At these lower frequencies, the magnetic reflection-loss curve resembles the electric field and plane-wave reflection loss curves. In the midfrequency range, the shield impedance matches the magnetic-field impedance, an efficient transfer of power occurs, and the shield magnetic reflection loss is negligible. Because of secondary reflections within the shield, a slight transmission gain through the shield may occur at the tip of the null. At the high frequencies, the shield impedance is higher than the magnetic-field impedance, and exhibits a rising characteristic; therefore, the impedance match between the shield and magnetic field becomes worse as frequency increases, and a less-efficient transfer of energy occurs. At these frequencies, the magnetic reflection-loss curve becomes the inverse of the electric-field and plane-wave reflection-loss curve.

The magnetic reflection-loss null usually occurs in the audio or lower video frequency spectrum when conventional shielding materials are utilized; therefore, adequate shield absorption losses are essential for the achievement of useful degrees of shield attenuation. High-permeability materials have the triple advantage of raising the frequency of the magnetic-field reflection-loss null beyond the upper response limit of much system hardware, increasing the absorption loss significantly, and increasing significantly the inductive losses exhibited by conductors adjacent to the high-permeability shield.

3.2.3.2.3.6 Absorption Losses. Absorption losses are dependent only on the type of shield material, frequency, and shield thickness. A straight edge, connecting a point on the G/u (material) scale to a point on the thickness scale, will also intersect a point on the uncalibrated vertical scale (Fig. 3-55). A straight edge, connecting this point on the uncalibrated vertical scale to a point on the frequency scale, will also intersect the dB (attenuation) scale, providing the absorption loss through the shield material.

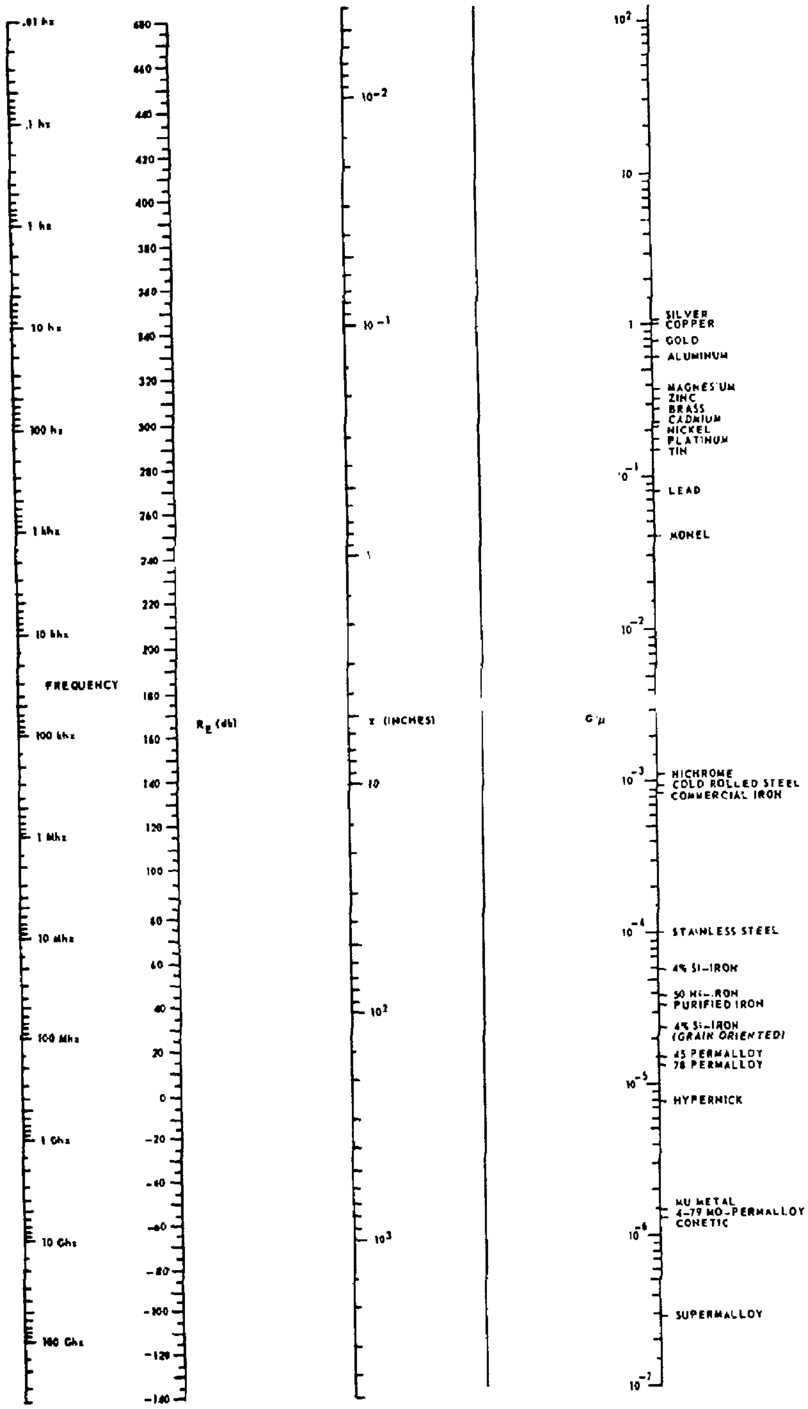



FIGURE 3-53. Electric-field reflection losses.

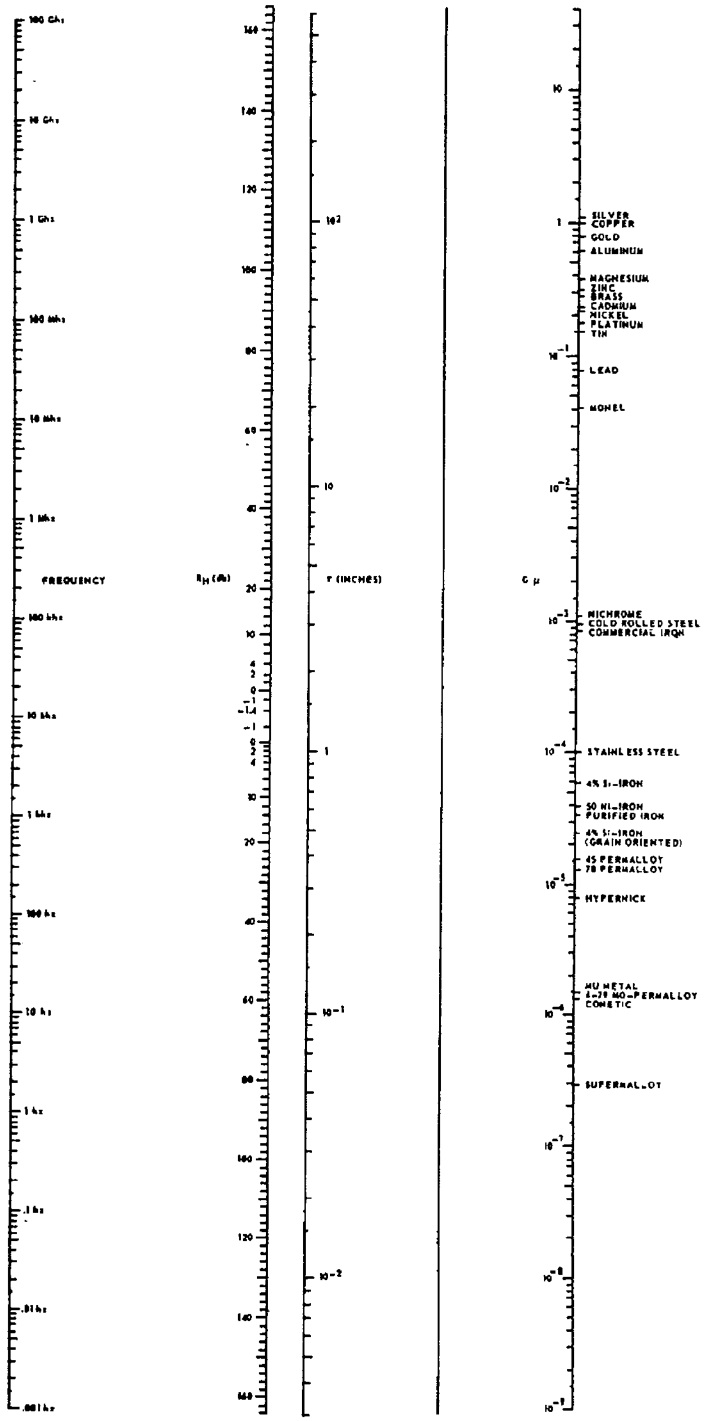



FIGURE 3-54. Magnetic-field reflection losses.

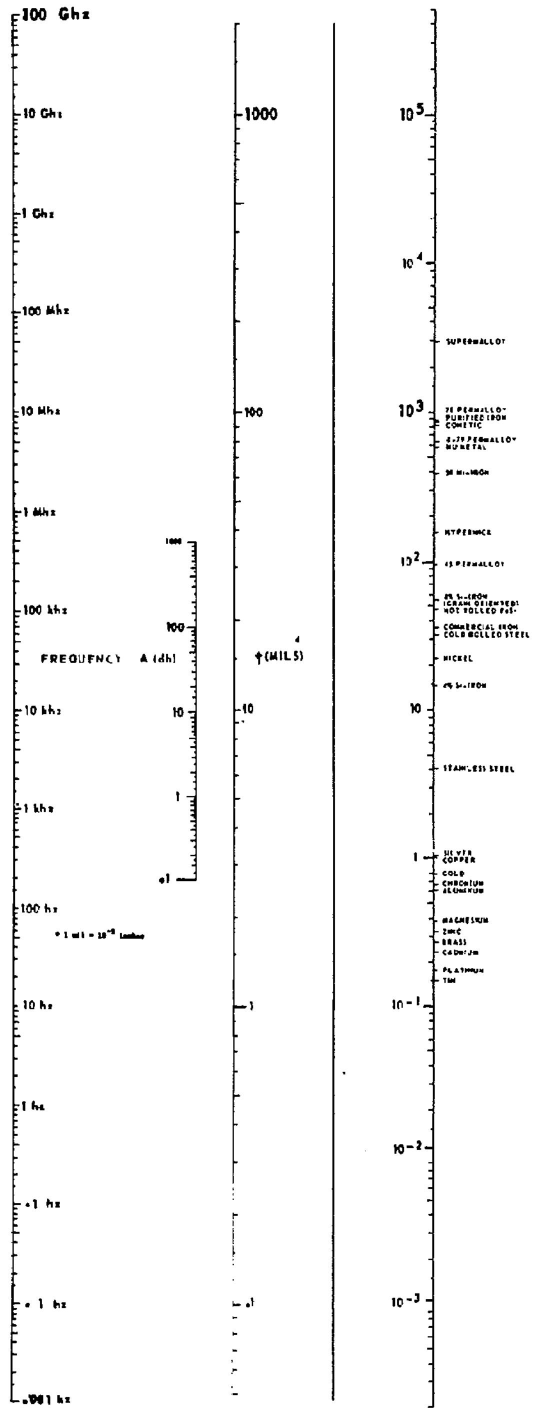



FIGURE 3-55. Absorption losses.

Absorption losses are a function of the material selected; they are proportional to thickness and increase rapidly with frequency, as noted previously. The use of ferrous metals, having high permeabilities, produces the greatest absorption losses for a given thickness of shielding material at a given frequency. Increasing the shield thickness also increases absorption losses, but prohibitive shield thickness may be required to provide adequate magnetic-field attenuation at audio frequencies, unless the material is selected carefully. Since absorption losses rise rapidly as frequency increases, and since high-permeability materials have high magnetic-reflection-loss null frequencies, the use of high-permeability materials produces a shield with relatively high absorption losses at the higher frequencies, where the magnetic-reflection-loss null of these materials tends to occur.

Ordinarily, absorption losses need to be considered only where magnetic-field attenuation is required. Electric-field and plane-wave reflection losses are usually adequate, even when absorption losses are negligible.

3.2.3.2.3.7 Multiple-Layer Shields. A single layer of shield material has reflection losses from both sides and an absorption loss proportional to thickness. Doubling the shield thickness does not increase the number of reflection losses or magnitude of total reflection losses, but merely doubles the attenuation because of absorption losses. Dividing the same double thickness of material between two independent layers of shield doubles the number of reflection losses and the magnitude of total reflection losses, in addition to doubling the attenuation because of absorption losses.

The use of multiple layers provides the opportunity to select electrically complementary materials, each of which supplements the characteristics of the other. This is also necessary to prevent fusing adjacent layers of similar material into a single, thicker layer. Nonferrous materials are analogous to a dielectric when inserted between ferrous materials. Since reflection losses are a dominant factor in shield attenuation, this increased number of interfaces is highly desirable.

Each layer of material is treated independently during calculations, with the adjacent layer considered as the source of interference. Attenuation versus frequency is plotted for each layer. The sums of the attenuations of the layers are plotted to provide a curve of the total shielding effectiveness versus frequency for the type of field (magnetic, electric, or plane wave) being considered.

3.2.3.2.3.8 Shield Perforations. The shielding effectiveness of an unperforated shield is degraded seriously when perforations are introduced. Magnetic-shielding effectiveness is extremely dependent on shield integrity, because of the flow of leakage currents through the perforations; these couple the induced currents on one shield surface to the opposite shield surface, from which magnetic radiation and magnetic coupling occur.

Electric-shielding effectiveness is also dependent on shield integrity; however, the presence of a surrounding shield structure provides a partially effective shadowing effect for the open areas analogous to the cone of protection provided by a lightning rod. The degree of protection supplied is dependent on the ratio of perforation size to shield thickness.

A thick shield with small perforations is analogous to looking through a long tube with a restricted field of view. The fringe capacitance is virtually eliminated and only the effective plate-to-plate capacitance, limited by the aperture size, is available to couple the electric field between objects isolated by the shield.

Plane-wave shielding, that eliminates predominantly capacitive coupling, at the higher frequencies where plane-wave geometry may exist, is affected by the same basic factors controlling electric-field shielding, with the addition of increased high-frequency coupling through the perforations. Above the frequency where the largest perforated, cross-sectional dimension exceeds a half wavelength, the perforation becomes a low-loss coupling path between the shield-isolated objects.

Shield perforations are equivalent to a waveguide attenuator with the same cross-sectional dimensions as the perforations, and the same length as the shield thickness. The large variety of potential perforated shield designs places detailed analyses of perforation effects beyond the scope of this publication. Numerous proceedings for the many conferences on radio interference reduction and electromagnetic compatibility, conducted by the Armour Research Foundation of the

Illinois Institute of Technology under triservice sponsorship, contain much valuable information related to shielding effectiveness.

3.2.3.3 Network Interconnection System. The wiring interconnection systems vary from complex multibranched wire-harness assemblies (Fig. 3-56) to simple two-ended wire-harness assemblies (Fig. 3-57).

In the past, the multibranched harnesses have been used extensively for airborne and other mobile vehicles, while two-ended harnesses have been commonly used for ground support, with provisions for interconnections provided by wire-wrap panels, patch panels, etc. Of course, many programs have used a compromise system with fixed branched harnesses in critical circuits less likely to change, and circuit-change devices in instrumentation and other systems more likely to change.

When considering FCC, it is essential to understand how the various harness-system requirements can be satisfied. First, the two-ended harnesses are ideally adaptable to the FCC system. Lengths of the required cable widths can be cut and terminated to plugs to make the simplest, lightest, most reliable, and most economical harnesses possible. FCC can also be applied to the large multibranched harnesses. Multiple cable widths can be terminated in each FCC plug to provide the required multiplicity of interconnections; however, the FCC harness complexity would be increased by the many cable segments required.

The early selection of the minimum number of multibranched harnesses is often made with the purpose that more important matters can then be persued; furthermore, this system has always worked in the past. One object of this handbook is to discourage such thinking and action and to promote careful and early consideration of the optimum interconnecting network system.

3.2.3.4 Connector Selections. This selection will be influenced by: the program requirements; the availability of connector hardware; and the company's manufacturing capability. The conductivity and voltage requirements will dictate the contact size and spacing. The handling and service requirements will dictate the ruggedness and configuration of the connector housings. Minimum space and weight requirements will dictate the degree of miniaturization required. The FCC connector system selected should accommodate the following requirements:

a. The preparation and termination of FCC conductors by layer.

b. The capability of accommodating various cable segments, both shielded and nonshielded or a combination, in the same cable layer (Fig. 3-58).

c. The capability of accommodating FCC with conductor centerlines a multiple of the connector contact centerlines. This may be accomplished in the conductor-contact connector by properly preparing the cable end to form the required number of conductors as shown in Figure 3-59. The pin-and-socket connector system can meet this requirement by making multiple contact terminations through the insulation or to the properly prepared conductor ends.

3.2.3.5 Cable Selection. The proper FCC must be selected to meet the program environmental, handling conductivity, and special requirements. The conductor material will usually be bare or plated copper. The cable construction must be compatible with the cable-end termination preparation requirements. The shielding material and conductor cross-section must meet the electrical requirements.

The decision on whether to use FCC for the larger conductor sizes (greater than AWG 20) must be made. If the conductivity is required to limit voltage drop rather than for high currents, FCC power cables defined in Section II can be used with existing FCC connector designs by using parallel contacts. FCC can be used for high-current applications with proper termination techniques.

3.2.3.6 Termination Systems. The termination system to be utilized with FCC is dictated by the type of connectors selected. With the NASA/MSFC conductor-contact system, the conductors are stripped and plated prior to plug assembly, and act as the plug contact. In other molded-on FCC plug concepts, such as that used by Rogers Corporation, the nonstripped cable is molded to the plug; the cable is then stripped and plated in the contact area. Pin-and-socket connectors

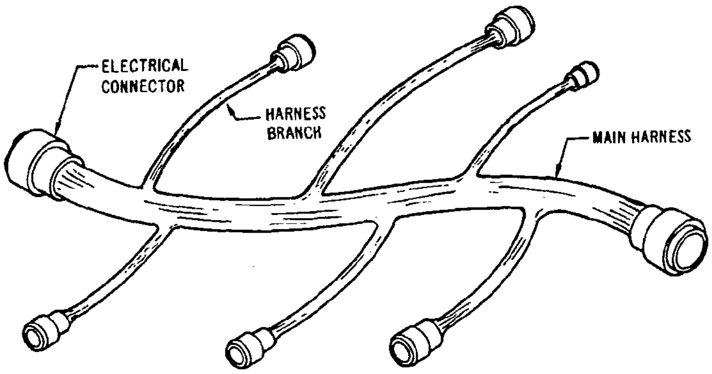



FIGURE 3-56. Multibranched wire harness assembly.

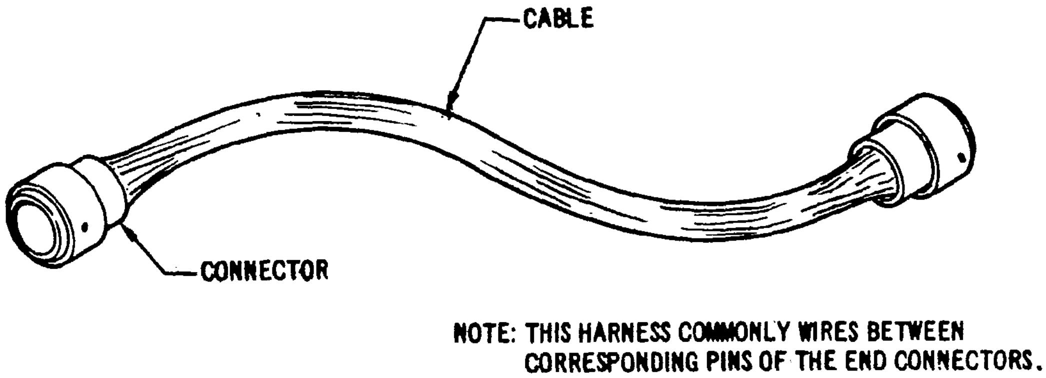
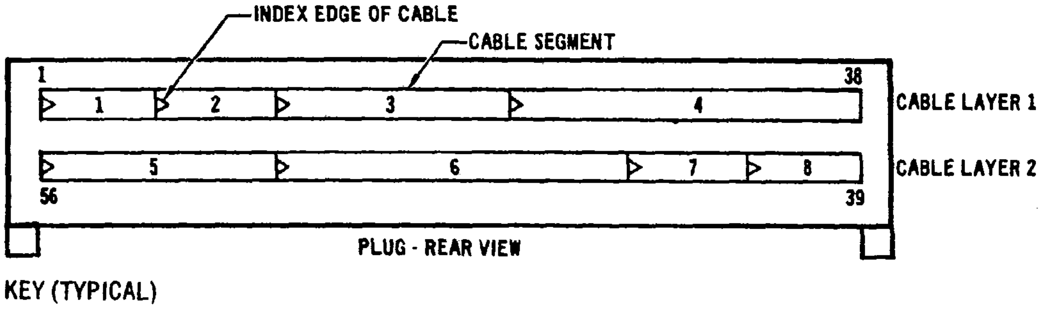



FIGURE 3-57. Two-ended wire-harness assembly.




| CABLE SEGMENT | WIDTH | SHIELD TYPE | CONDUCTOR CENTER LINE | DENSITY | CONDUCTOR THICKNESS |
|---|---|---|---|---|---|
| 1 | 1/2 | 2S | .075 | STD | .003 |
| 2 | 1/2 | 2S | .075 | STD | .003 |
| 3 | 1 | NON-S | .075 | STD | .004 |
| 4 | 1-1/2 | NON-S | .150 | HIGH | .006 |
| 5 | 1 | NON-S | .075 | STD | .004 |
| 6 | 1-1/2 | NON-S | .150 | HIGH | .006 |
| 7 | 1/2 | 2S | .075 | STD | .003 |
| 8 | 1/2 | 2S | .075 | STD | .003 |

FIGURE 3-58. FCC connector versatility.

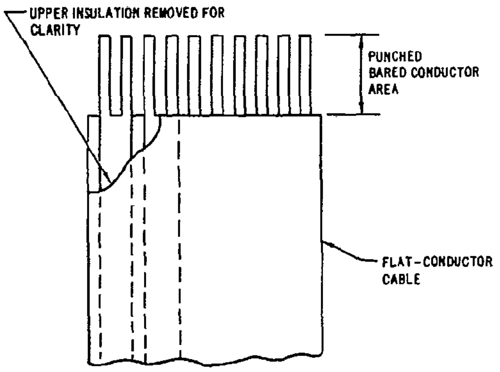



FIGURE 3-59. FCC reduced center cable preparation.

utilize various types of termination systems including crimping, soldering, ultrasonic welding, and and insulation penetrations. The termination system to be used will strongly influence the selection of the FCC and connectors.

3.2.4 FCC Power Cables. Various FCC power cables have been designed by NASA/MSFC and prototypes manufactured by Methode Electronics. These cables are 1, 2, and 3 inches wide with two or three power conductors in each cable with equivalent AWG conductor sizes ranging from Nos. 8 through 15 (Fig. 2-6).

Figure 3-60 is a graph for selecting various FCC cross-sections for equivalent AWG sizes 8 through 18. The centerline spacing and distance between adjacent conductors would be based on the termination system hardware and on the system electrical requirements.

Table 3-13 lists various FCC cross-sections for equivalent AWG wire sizes from Nos. 8 through 23. Conductor thickness is 10 mils. For installations not requiring frequent flexing, the conductor thicknesses could be more than 10 mils.

There are numerous installation advantages to be realized in using FCC power cables. Bonding the cable directly to the primary structure provides the following advantages:

a. A minimum of space is required.

b. Very little, or no, support on attaching structure is required.

c. The structure will provide an efficient heat sink for conductor heat dissipation.

Using FCC power cables with the equivalent 20-gage and smaller MIL-C-55543, FCC simplifies harness fabrication, routing, and support (see Paragraph 3.2.6).

Electrical advantages of FCC power cables are:

a. Electrical characteristics can be accurately predicted.

b. For dc and relatively low-power frequencies, the capacitance between conductors and between conductors and ground automatically provides a very efficient noise-suppression filter.

Preliminary calculations for typical cable requirements for dc power frequencies up to 400 Hz indicate acceptable line losses.

For characteristic impedances most acceptable for system faults and general system performance, it is recommended that the configurations shown in Figure 3-61 be used.

For power frequencies above 400 Hz, special considerations must be given to skin effect and capacitive high-frequency losses. These can be calculated from the FCC harness cross-section and length by using formulas in published electrical handbooks.

For FCC power cable termination in relatively low current applications, MIL-C-55544 FCC connectors can be used with multiple, parallel contacts for each power conductor. This is illustrated in Figure 2-8 for the MSFC conductor-contact plug assembly. For high-current applications, the FCC conductors can be terminated to bus-bar type terminals for bolting directly to the utilizing equipment, or transition can be made to RWC by welding, crimping, etc., for routine RWC terminations. All termination areas should be properly sealed for corrosion and strain relief.

The use of FCC power cables with aluminum conductors offers additional advantages in weight, cost, and availability. By maintaining an efficient seal over the conductors and the termination areas, the FCC power cable system can overcome previous corrosion problems associated with the use of aluminum for electrical conductors.

In conclusion, the use of FCC power cables presents many potential advantages over RWC. The mechanical, electrical, and hardware requirements should be detailed early in the program to determine those areas in which FCC power cables can be efficiently used.

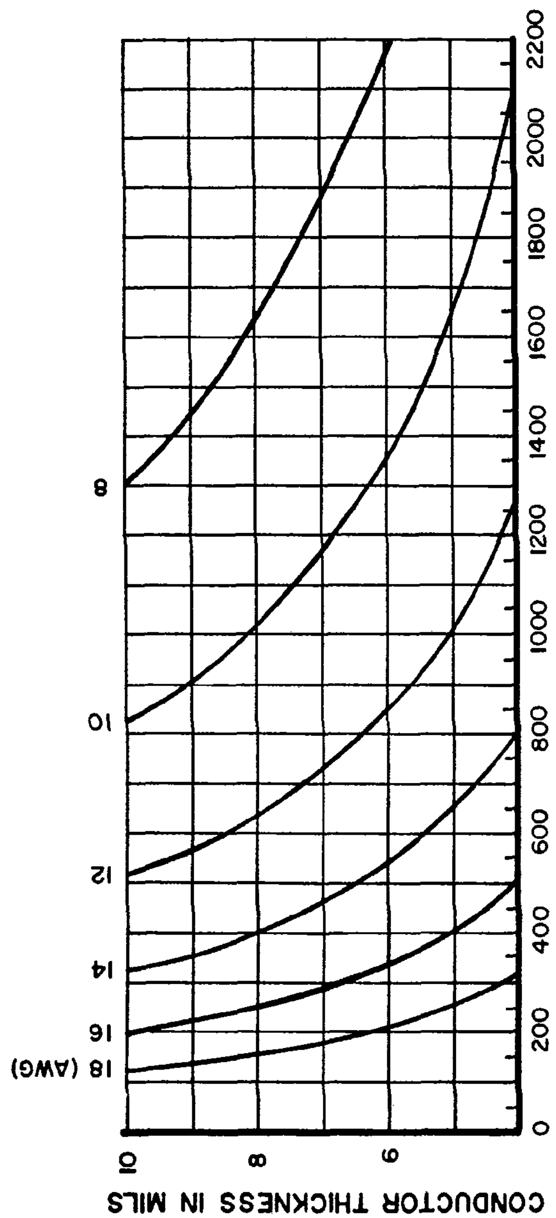
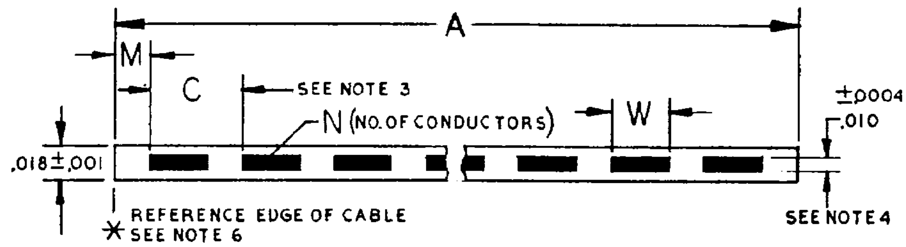



FIGURE 3-60. FCC cross-sections-equivalent AWG 8 through 18.

## TABLE 3-13. FCC POWER CABLE CONDUCTOR CROSS SECTIONS




NOTES:
1. Material: Polyimide/Fluorinated Ethylene Propylene Insulated, Nickel Coated Strip-Copper Conductor
2. Dimensions are in inches.
3. Typical, non cumulative.
4. Thickness of insulation on each side of the conductors shall be uniform within 0.001 inch.
5. Based on conductor cross-section.
6. Cable marking shall be placed along reference edge of cable.
7. Per requirement of MIL-C-55543 except "Flexing."
8. Presently, no military specification items.

| Cable Width "A" ±0.005 | No. of Conductors "N" | Conductor Width "W" ±0.002 | Basic Spacing "C" ±0.005 | Cable Margin "M" ±0.005 | Nearest Awg. Wire Size (Note 4) | Max. Cond. DC Resistance Ohms/1000 ft. at 20°C | Max. Cable Weight Lbs/1000 ft. |
|---|---|---|---|---|---|---|---|
| 0.5 | 2 | 0.190 | 0.225 | 0.0425 | 16 | 4.29 | 19.2 |
| 0.5 | 3 | 0.115 | 0.150 | 0.0425 | 18 | 7.08 | 18.2 |
| 0.5 | 6 | 0.040 | 0.075 | 0.0425 | 23 | 20.40 | 15.2 |
| 1.0 | 2 | 0.415 | 0.450 | 0.0675 | 13 | 1.96 | 40.3 |
| 1.0 | 3 | 0.265 | 0.300 | 0.0675 | 15 | 3.07 | 39.3 |
| 1.0 | 4 | 0.190 | 0.225 | 0.0675 | 16 | 4.29 | 38.3 |
| 1.0 | 6 | 0.115 | 0.150 | 0.0675 | 18 | 7.08 | 36.3 |
| 1.0 | 12 | 0.040 | 0.075 | 0.0675 | 23 | 20.40 | 30.3 |
| 1.5 | 2 | 0.640 | 0.675 | 0.0925 | 11 | 1.27 | 61.5 |
| 1.5 | 3 | 0.415 | 0.450 | 0.0925 | 13 | 1.96 | 60.5 |
| 1.3 | 4 | 0.265 | 0.300 | 0.0675 | 15 | 3.07 | 51.9 |
| 1.5 | 6 | 0.190 | 0.225 | 0.0925 | 16 | 4.29 | 57.5 |
| 1.5 | 9 | 0.115 | 0.150 | 0.0925 | 18 | 7.08 | 54.5 |
| 1.5 | 18 | 0.040 | 0.075 | 0.0925 | 23 | 20.40 | 45.5 |
| 1.9 | 2 | 0.865 | 0.900 | 0.0675 | 10 | 0.94 | 81.1 |
| 1.9 | 3 | 0.565 | 0.600 | 0.0675 | 12 | 1.44 | 80.1 |
| 1.9 | 4 | 0.415 | 0.450 | 0.0675 | 13 | 1.96 | 79.1 |
| 2.0 | 5 | 0.340 | 0.375 | 0.080 | 14 | 2.40 | 81.9 |
| 1.9 | 6 | 0.265 | 0.300 | 0.0675 | 15 | 3.07 | 77.1 |
| 1.9 | 8 | 0.190 | 0.225 | 0.0675 | 16 | 4.29 | 75.1 |
| 1.9 | 12 | 0.115 | 0.150 | 0.0675 | 18 | 7.08 | 71.1 |
| 2.0 | 25 | 0.040 | 0.075 | 0.080 | 23 | 20.40 | 61.9 |
| 2.5 | 2 | 1.165 | 1.200 | 0.0675 | 9 | 0.70 | 108.3 |
| 2.4 | 3 | 0.715 | 0.750 | 0.0925 | 11 | 1.14 | 101.2 |
| 2.5 | 4 | 0.565 | 0.600 | 0.0675 | 12 | 1.44 | 106.3 |
| 2.4 | 5 | 0.415 | 0.450 | 0.0925 | 13 | 1.96 | 99.2 |
| 2.4 | 6 | 0.340 | 0.375 | 0.0925 | 14 | 2.40 | 98.2 |
| 2.5 | 8 | 0.265 | 0.300 | 0.0675 | 15 | 3.07 | 102.3 |
| 2.4 | 10 | 0.190 | 0.225 | 0.0925 | 16 | 4.29 | 94.2 |
| 2.5 | 16 | 0.115 | 0.150 | 0.0675 | 18 | 7.08 | 94.3 |
| 2.5 | 32 | 0.040 | 0.075 | 0.0675 | 23 | 20.40 | 78.3 |
| 3.0 | 2 | 1.390 | 1.425 | 0.0925 | 8 | 0.59 | 129.4 |
| 2.8 | 3 | 0.865 | 0.900 | 0.0675 | 10 | 0.94 | 120.8 |
| 2.8 | 4 | 0.640 | 0.675 | 0.0675 | 11 | 1.27 | 119.8 |
| 2.8 | 6 | 0.415 | 0.450 | 0.0675 | 13 | 1.96 | 118.0 |
| 2.8 | 9 | 0.265 | 0.300 | 0.0675 | 15 | 3.07 | 114.8 |
| 2.8 | 12 | 0.190 | 0.225 | 0.0675 | 16 | 4.29 | 111.8 |
| 3.0 | 19 | 0.115 | 0.150 | 0.0925 | 18 | 7.08 | 111.7 |
| 3.0 | 38 | 0.040 | 0.075 | 0.0925 | 23 | 20.40 | 93.4 |

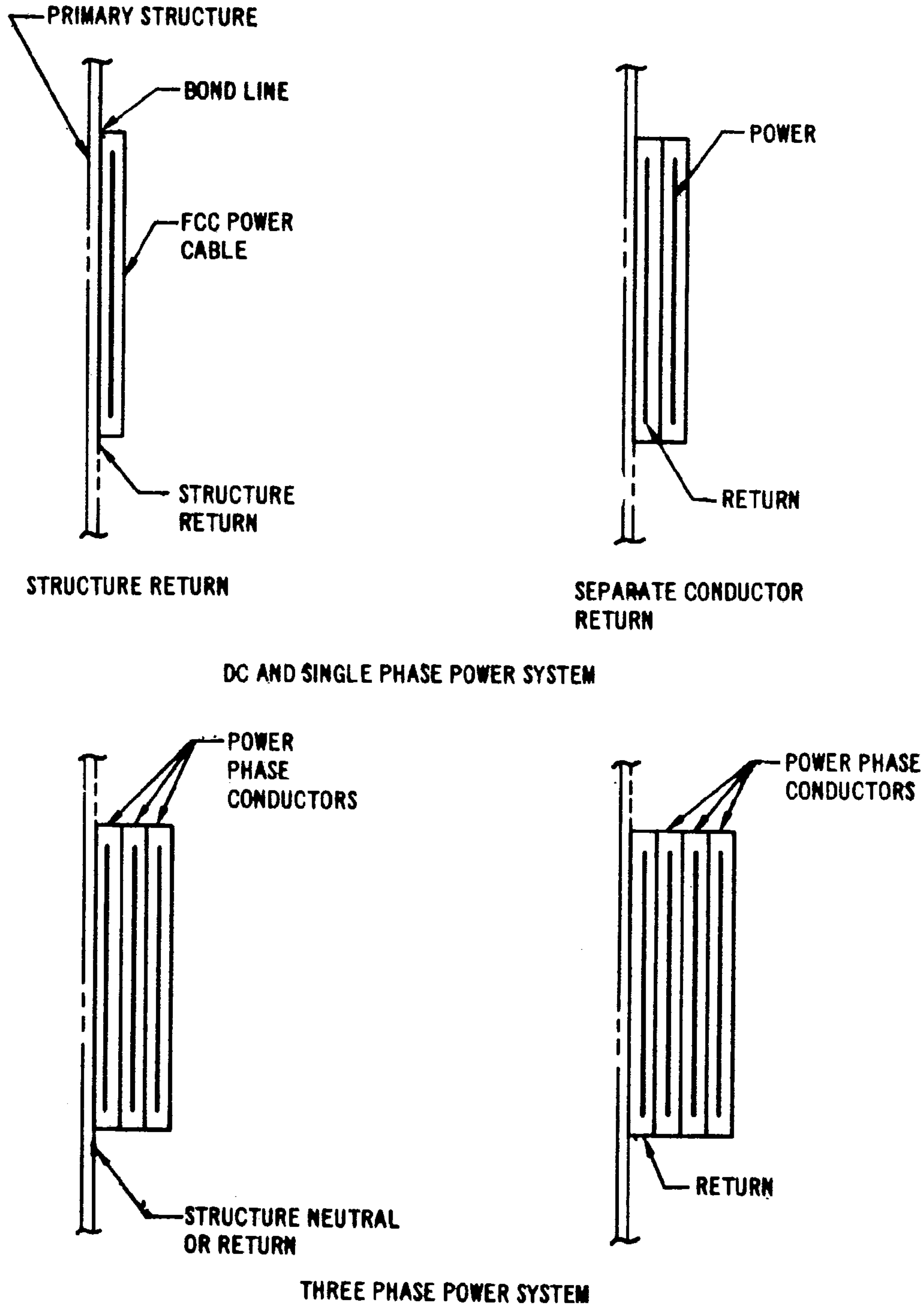



FIGURE 3-61. FCC power cable installations configurations.

3.2.5 Harness Design. Prior to the interconnecting harness detail design, the overall system design must be accomplished so that all components comprising the functional system are grouped or packaged together, located relative to each other, and interconnections established to provide a logical and simple flow of power, control, sensing, etc. Requirements for heat dissipation, weight distribution, and accessibility for installation and service must be considered. The importance of the initial system design cannot be overemphasized, and its efforts will be rewarded manifold in subsequent development and detail design.

Since this document deals primarily with FCC application, it will be assumed that an efficient system design has been accomplished to define (1) the packaged electronic units and components in each system, (2) the interconnecting requirements of each system, (3) the interconnecting requirements between systems, and (4) the interface requirements to units or systems mating with the end item under design. The design of the interconnecting FCC harness network can then proceed as defined in the following subparagraphs.

3.2.5.1 Interconnecting Wire Requirements. The system design should define all system components (units requiring cable network interconnecting) and the interconnecting requirements within and between systems, and to electrical interfaces. This definition should provide a designation number for each system component, together with the number of circuits, circuit functions, classification for grouping, conductivity, and any other special consideration requirement for the interconnecting circuits. See Table 3-14 for typical interconnecting requirement tabulation.

The zoning classification philosophy for grouping is explained in Paragraphs 3.2, 3.2.1 and 3.2.10.2. Each system has its own peculiarities which must be thoroughly analyzed to establish the circuit classifications and zonings required to provide adequate system performance. In the past, the random conductor registration in conventional round-wire bundles has resulted in overdesign.

Excessive shielded cable was specified and, where space permitted, an excessive number of bundles was routed and supported separately. Testing accomplished by MDC on existing missile and aircraft systems indicated that up to 75 percent of the existing round-wire shielded cable could be eliminated with FCC interconnecting harnesses. The FCC system, with its positive registration control of each conductor, permits the interconnecting harness to be considered as a predictable, repeatable system component.

The complete definition of the program systems interconnecting requirements (Table 3-14) and the establishment of the proper zoning classification for each circuit conductor is a major task. Once this is accomplished and recorded, work can begin on the wiring layout.

3.2.5.2 Wiring Layout. With the end-item interconnecting electrical wiring requirements defined in a series of tabulated charts typified by Table 3-14, work can begin on the wiring layout. A simple example will be given to illustrate a logical method of procedure.

An initial wiring layout is made as shown in Figure 3-62. This figure contains the basic elements required for the analysis and design of the interconnecting network. Each electronic unit is shown in its area, the required interconnect paths are indicated, and the number and zone classification of each circuit with conductivity requirements are defined.

In most modern systems, there would be many more units to be interconnected and more zoning classifications than the three as follows:

| Class | Definition |
|---|---|
| A | Interference circuits |
| B | Noncritical circuits |
| C | Susceptible circuits |

Figure 3-62 also assumes that only two conductor conductivities are required; 200 square mils (AWG 26) for symbol 1 and 840 square mils (AWG 20) for symbol 2. In most instances, there would be additional conductivity requirements, including those for power, which may be many times the average required for general use. Power cables are given in Paragraph 3.2.4 and other special configurations are given in Paragraph 3.2.8.

# TABLE 3-14. TYPICAL INTERCONNECTING REQUIREMENTS

| System (Note 1) | Component Unit (Note 2) | Interconnecting Requirements: Circuit No. | Circuit Function | Minimum Square Mil Conductor Cross-Section | Zoning Classification (Note 3) | Special Requirements (Note 4) | Termination (Note 5) |
|---|---|---|---|---|---|---|---|
| 1 | 1 | 1 | 5 Vdc TM | 150 | A2 | | 6-1-3 |
| | | 2 | -20 Vdc test | 150 | A4 | | 8-3-8 |
| | | 3 | 2000 Hz TM, continued | 150 | A7 | | 10-2-15 |
| | 2 | 1 | 20 Vdc output | 1500 | A5 | | 5-3-6 |
| | | 2 | 20 Vdc return | 1500 | A5 | | 5-3-7 |
| | | 3 | RAGS #1 continued | 800 | A3 | | 4-2-9 |
| | 3 | 1 | Trans input | 800 | B3 | | 6-3-10 |
| | | 2 | Cal Sig Rtn | 100 | C8 | | 4-2-7 |
| | | 3 | Cal Sig Continued | 100 | C8 | | 4-2-8 |
| | Additional Units | | | | | | |
| 2 | 1 | 1 | Meas hr | 75 | C4 | | 4-2-7 |
| | | 2 | Pwr Xfer Com | 190 | A4 | | 3-3-9 |
| | | 3 | Mstr Rtn continued | 800 | C3 | | 4-2-15 |
| | 2 | 1 | Ball test | 150 | A2 | | 7-2-3 |
| | | 2 | Bolt test | 150 | A2 | | 7-2-7 |
| | | 3 | Meas rms continued | 75 | C4 | | 7-2-10 |
| | 3 | 1 | Coax Input | NA | D | RG-59 | 3-2-1 |
| | | 2 | Out to 1 | | D | RG-59 | 3-5-2 |
| | | 3 | Out to 2 Continued | | D | RG-59 | 3-6-2 |
| | Additional Units | | | | | | |
| Additional Systems | | | | | | | |

Notes:
1. Electronic system identification.
2. Particular electronic unit of a system which requires network interconnections.
3. Zoning classification. (Ground rules for grouping and isolation must be established for each program, see Paragraph 3.2.3.2.1.)
4. Special requirements (coded or defined).
5. Termination designation indicates the system, unit, and unit circuit number to which the circuit terminates.

The callouts listed in the table do not represent actual systems but are given as examples of typical callout.

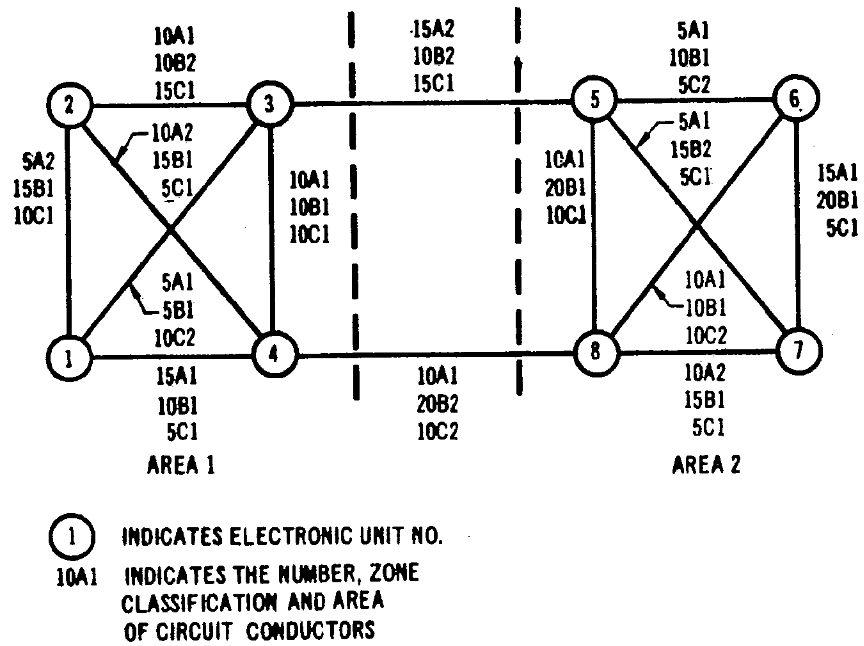



FIGURE 3-62. Initial wiring layout.

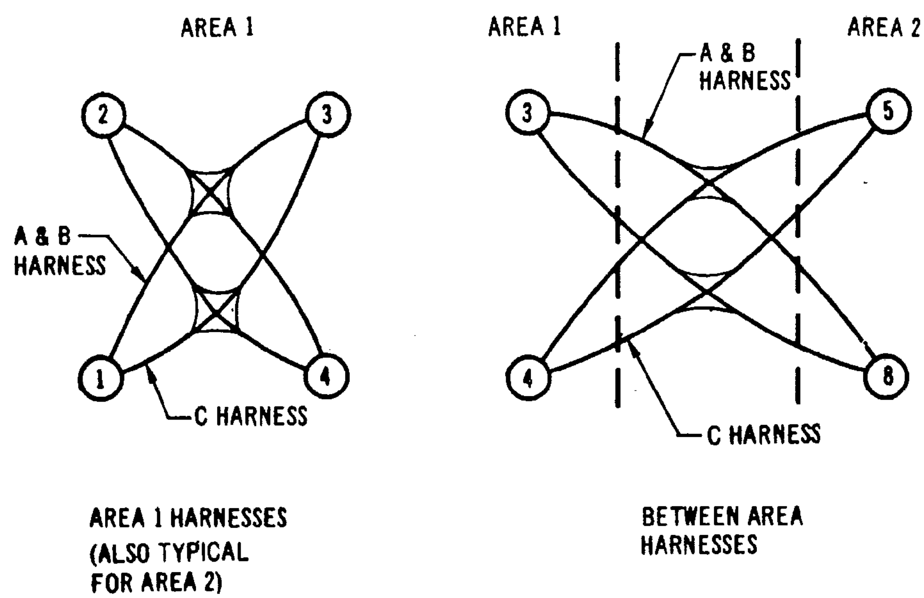



FIGURE 3-63. RWC wiring layout.

3.2.5.2.1 Multibranched Harnesses. If the system shown in Figure 3-62 were to be interconnected with the conventional RWC system, it would utilize multibranched harnesses in each area and between areas as shown in Figure 3-63. No. 22 AWG (500 square mils) wire could be selected for the No. 1, and No. 20 AWG (804 square mils) wire for the No. 2 conductivity requirements. The number and sizes of connectors would be selected to accommodate the number of conductors and the zoning requirements.

There would be a minimum of two RWC harnesses each in areas 1 and 2 and two harnesses to interconnect areas 1 and 2, a total of six harnesses with a minimum of 24 connectors. The use of two harnesses in each area would provide the required circuit isolation.

With multibranched FCC harnesses utilizing multiple cable segments, as required in each FCC connector contact layer, the number of branched harnesses could be reduced to three; one each in areas 1 and 2, and one to interconnect areas 1 and 2. No attempt will be made to design this harness, but Figure 3-64 shows the wiring layout, and Figure 3-65 shows the concept. The circuit conductivity and zoning requirements per Figure 3-62 are shown in Figure 3-65 at electronic unit No. 1 and through the harness run between units 2 and 3. The circuit isolation in the multibranched FCC harnesses is provided by conductor registration.

By using the method described in subsequent paragraphs, the cable and connectors can be selected from Tables 3-2, 3-15, 3-16, and Figure 3-5, and the harness design completed. For example, the FCC connectors at unit 1 (Fig. 3-62) would require a minimum total of 95 contacts on 0.075 centerline (2 for each circuit with a conductivity requirement of 2). Table 3-16 shows that this requirement might be satisfied with a 3-inch wide, 3-layer connector or two 2-inch-wide, 2-layer connectors. The cable layer segments would be selected from Tables 3-2 and 3-15. Many narrow-width cable segments may be required to provide the cable branching.

So we see that the FCC-branched harness can reduce by 50 percent the number of harness runs and the number of connectors over that required by the RWC branched harnesses. If transitions with pin-change capabilities are made to round connectors, or if the electronic units can readily be rewired internally, then the system flexibility can be retained with FCC-branched harnesses. All the advantages of weight, space, and reliability can be realized with the branched FCC harnesses. Although the manufacturing cost will increase slightly due to the handling of the multiple-cable segments, the major cost saving will still be realized.

With all these advantages, it is important to recognize the system requirements' imposed by the use of the FCC-branched harness described above. To efficiently utilize this system, the pin assignment and connector selection at each electronic unit would be controlled by the FCC-branched harness design; the pin assignments at each electronic unit could not be made until after the harness design was completed and, once defined, the pin-assignment changes would be limited to those which could be accomplished by reversing or relocating cable segments (not individual conductors) in the FCC plugs. Section IV defines FCC wire-change methods that can be used to accomplish isolated changes; however, in general, the multibranched FCC harnesses described should be used on those programs requiring many identical end items, and those which can permit pin-assignment control by the interconnecting harness networks.

3.2.5.2.2 Branched Harnesses with Distributors. Many existing airborne designs utilize branched interconnecting harness networks with distributors or other circuit-changing capabilities incorporated into those systems most likely to change during the program. Typical systems, which change frequently, are instrumentation and control. Figure 3-66 shows such a branched harness system with partial system distributors located in separate units. Electronic unit 2 and distributor unit I, and electronic unit 4 and distributor unit II can be located within the same box structure.

Figure 3-66 is a compromise design that has evolved over many years experience with RWC interconnecting systems. This system is flexible in permitting pin-assignment changes in the primary of "fixed" systems during design, and in the secondary or "changeable" systems after the design is completed and essentially frozen.

The FCC system can be readily adapted to the "branched harness with distributor" design. Adjacent to the distributors, the two-ended harnesses will provide all the advantages of FCC.

TABLE 3-15. FCC CABLE SELECTOR CHART

| Cable Width (in.) | Number of Conductors: 50-Mil $C_L$ Spacing | 75-Mil $C_L$ Spacing | 100-Mil $C_L$ Spacing | 150-Mil $C_L$ Spacing |
|---|---|---|---|---|
| 0.25 | | 3 | | |
| 0.50 | 7 | 6 | 4 | 3 |
| 1.00 | 17 | 12 | 9 | 6 |
| 1.50 | 27 | 18 | 14 | 9 |
| 2.00 | 37 | 25 | 19 | 12 |
| 2.50 | 47 | 32 | 24 | 16 |
| 3.00 | 57 | 38 | 29 | 19 |

TABLE 3-16. FCC CONNECTOR SELECTOR CHART

| Number of Contacts | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|
| 50-Mil $C_L$ Contact Spacing | | | | 75-Mil $C_L$ Contact Spacing | | | | 100-Mil $C_L$ Contact Spacing | | | |
| Cable Width (in.) | No. of Layers 1 | 2 | 3 | Cable Width (in.) | No. of Layers 1 | 2 | 3 | Cable Width (in.) | No. of Layers 1 | 2 | 3 |
| 0.25 | | | | 0.25 | 3 | 6 | 9 | 0.25 | | | |
| 0.50 | 7 | 14 | 21 | 0.50 | 6 | 12 | 18 | 0.50 | 4 | 8 | 12 |
| 1.00 | 17 | 34 | 51 | 1.00 | 12 | 24 | 36 | 1.00 | 9 | 18 | 27 |
| 1.50 | 27 | 54 | 81 | 1.50 | 18 | 36 | 54 | 1.50 | 14 | 28 | 42 |
| 2.00 | 37 | 74 | 111 | 2.00 | 25 | 50 | 75 | 2.00 | 19 | 38 | 57 |
| 2.50 | 47 | 94 | 141 | 2.50 | 32 | 64 | 96 | 2.50 | 24 | 48 | 72 |
| 3.00 | 57 | 114 | 171 | 3.00 | 38 | 76 | 114 | 3.00 | 29 | 58 | 87 |

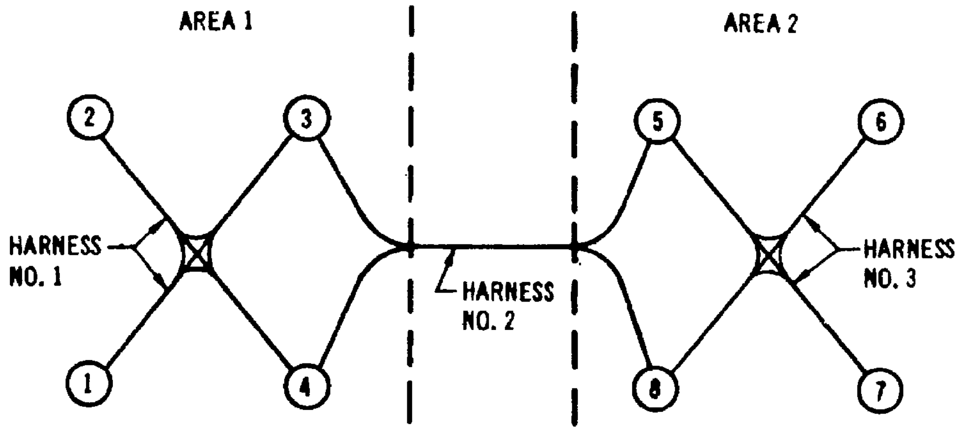



FIGURE 3-64. Multibranched FCC wiring layout.

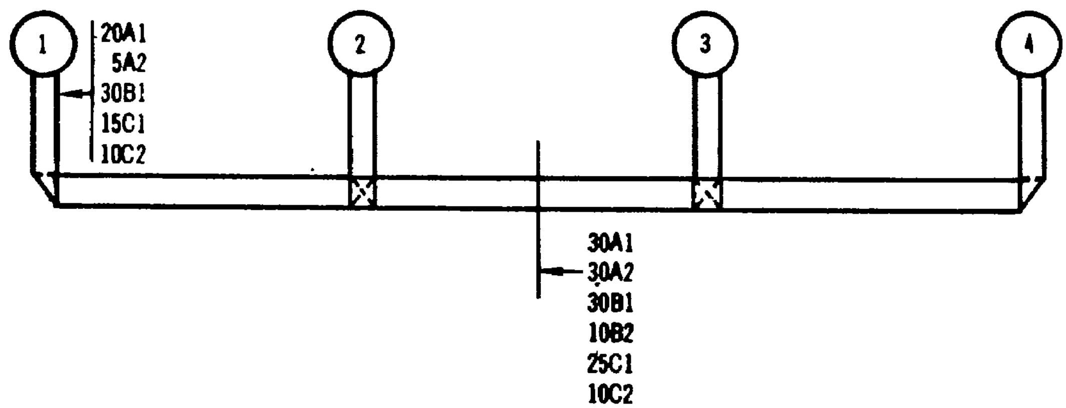



FIGURE 3-65. FCC branched harness.

3.2.5.2.3 Two-Ended Harnesses with Distributors. Returning to the example initial wiring layout (Fig. 3-62), a third design approach will be taken using distributors for all circuits.

For definition purposes, a distributor is a device used in the electrical interconnecting network that simplifies the wiring harnesses and facilitates wiring changes. A thorough description of the objectives and proposed design for distributor systems is given in Section IV.

Figure 3-67 shows the distribution unit, equivalent wiring layout of Figure 3-2. An examination of Figure 3-67 shows that 4 each FCC harnesses are required in areas 1 and 2, and one between the two areas. The most important advantages of distributors are:

a. Harnesses are two-ended and can be terminated in a predetermined configuration for all program connectors.

b. The electronic unit pin assignments can be selected for optimum circuit zoning that will automatically provide optimum zoning in the FCC harnesses.

c. All circuit or pin-assignment changes can be made in the distributors without affecting the FCC harnesses or electronic units.

d. The FCC harnesses between the distributors can be designed and installed prior to the final circuit designs of the electrical systems. The optimum cable widths and connector sizes can be used on these harnesses.

e. The minimum number and smallest sizes of connectors can be used on each electronic unit.

f. Electrical characteristics will be practically identified on subsequent installation of the same configuration.

The advantages listed above are very important and make a major contricution to the wiring breakthrough that is so badly needed. Many agents, both government and prime contractor, responsible for overall electrical systems in space, defense, and commercial programs, are convinced that for the complicated systems of tomorrow, the distributor design system is needed, whether the RWC or FCC interconnecting systems are used.

3.2.5.3 Selection of Cables and Connectors. The FCC harness design will now be made for the wiring layout with distributors shown in Figure 3-67.

The selection of cable and connectors requires close coordination. First, the centerline spacing of the cable must be the same as, or a multiple of, that of the connector and, second, the cable construction and insulation system must be compatible with the termination system used by the connector.

For this program, let us assume that the NASA/MSFC conductor-contact connector system will be used.

Figure 3-67 shows that two conductivity sizes are required; No. 1 requires 200-square-mil cross-section, and No. 2 requires 840-square mils as previously stated. From Table 3-2, these requirements can be met by:

a. 200 square mils - 50-mil centerline, 5- by 40-mil high density
- 75-mil centerline, 5- by 40-mil standard density

b. 840 square mils - 150-mil centerline, 6- by 140-mil high density.

The 75-mil centerline NASA/MSFC conductor-contact connector system is selected with the 5- by 40-mil standard density and the 6- by 140-mil high-density cables. Two conductor contacts will be used in parallel for the 150-mil centerline cable (see Paragraph 3.2.3.4.c). The cable type will be selected to meet the program environmental requirements. For this example, the type selected is polyimide/FEP insulated, laminated, nickel-plated copper conductor cable.

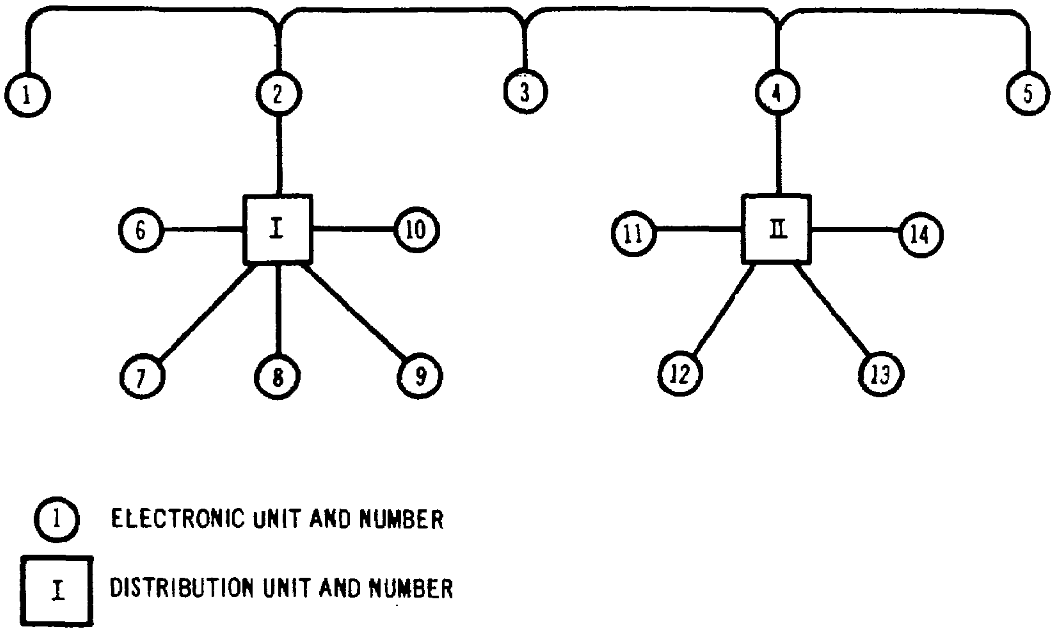



FIGURE 3-66. Branched harness with distributors.

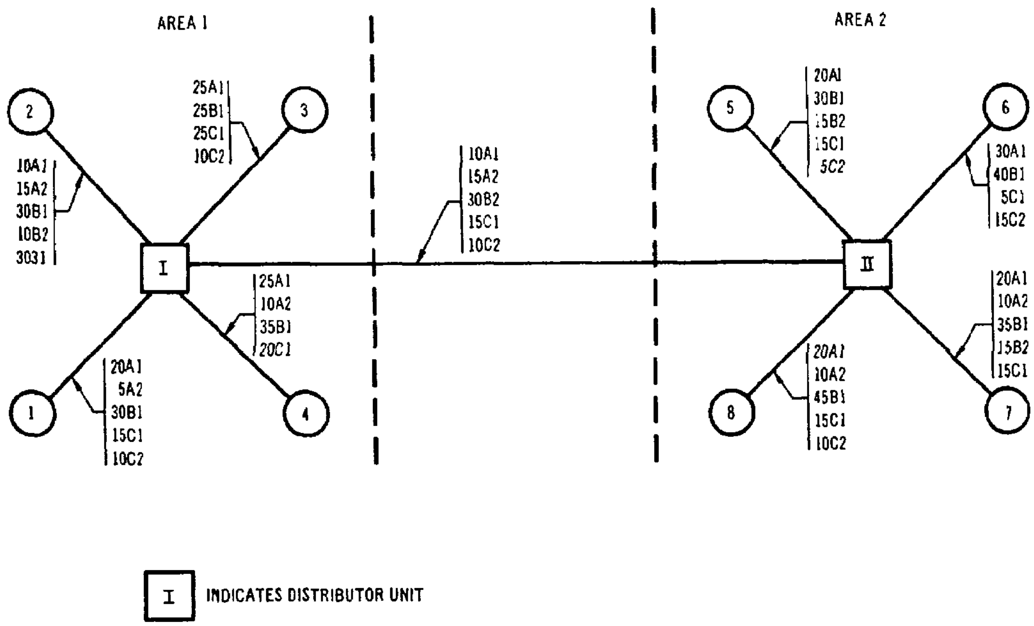



FIGURE 3-67. Wiring layout with distributors.

Now the connector sizes and cable segments can be selected for each FCC harness. First, consider the harness (1-I) between electronic unit 1 and distributor I. The circuit requirements are:

| | | |
|---|---|---|
| 20A1 | 30B1 | 15C1 |
| 5A2 | | 10C2 |

Since the number 2 conductivity will require two 0.075-centerline spacing contacts, the total number of contacts is 95. From Table 3-16, two 2.50-inch two-layer connectors are selected. Next, by using the cable information from Table 3-15, Table 3-17 can be prepared to provide the cable segments to meet the conductivity and zoning requirements. The cable segments selected are shown together with their registration to the plug. The selection shown provides the conductivity, zoning isolations, and spare conductors in each conductivity size and in each zoning category.

The connectors, cable segments, and zone grouping can be selected in a similar manner for the remaining eight harnesses.

3.2.5.4 Pin Assignment and Zoning. During the process of the selection of the cable segments, the zoning requirements were already considered. After the connector sizes and cable segments have been selected and registered, the circuit pin assignments can be added to the interconnecting wiring requirement data of Table 3-14. The pin assignment should be made, where possible, to obtain even greater isolation. If the most susceptible circuits are placed at one extreme edge, and the worst interference circuits at the other extreme edge of the interconnecting harness, then the maximum isolation will have been achieved. In cases where this is not possible, shielded FCC may be used. It should be noted that the connector and zoning selection provides the optimum design, both inside and outside the electronic unit.

It remains only to carry the established zoning through the harness routing with other cables.

3.2.5.5 Wire Harness Drawings. The FCC wire harness drawings are similar to RWC harness drawings. This drawing can be a computer printout drawing, showing a tabulation for point-to-point interconnection between all terminating items such as connectors, ground lugs, etc.; parts list; picture presentation of wire harness; and production test requirements for the completed harness. Figure 3-68 shows the essential elements of an FCC harness drawing. The harness requirements were taken from the upper plug of Table 3-17. These are 2.5-inch plugs at each end with one cable segment in layer 1 and two in layer 2, as shown.

The wire list shown in Figure 3-68 defines the point-to-point interconnections for each cable to control the exact cable-segment registration in the connectors; to provide a means of circuit tracing; and for production testing requirements. Two connector pins are required for each conductor of cables 1 and 3 that have conductors on 150-mil centers. The cable conductor number is positively identified in each cable by counting 1, 2, 3, etc., from the part number identification edge of the cable. This identification is required on all FCC during manufacture. Each cable segment is also identified with the harness and cable numbers so that each circuit can be positively identified without stamping the wire numbers on individual conductors. If the program requires a wire number identification for each conductor segment, it can be added as shown in Figure 3-68.

3.2.6 FCC Harness Installation. Concurrent with the electrical design of the FCC harnesses, consideration should be given to harness development, routing, and installation. The largest effort to date for development and installation of FCC harnesses has been on studies performed by MDC for NASA/MSFC. In one study over 100 FCC harnesses were developed, manufactured, and installed in a 180-degree mockup section of the aft skirt of the S-IV-B (Figs. 1-1 and 1-2). In another study, FCC was considered for all harnesses on a proposed S-IV-B baseline vehicle to be used with special kits to meet the mission requirements. The following paragraphs include information from these two programs, together with additional information obtained from other FCC programs and studies.

TABLE 3-17. CABLE SELECTION AND ZONING — HARNESS 1-I

| Plug No. | Layer | Zone A | Zone B | Zone C | Cable Segments |
|---|---|---|---|---|---|
| 1 | 1 | 5A2 | (S2-6) | 5C2 | (1) 2.5 - 0.150 |
| | 2 | 10A1 | 8B1 \| (S2-1) | 5C2 | (1) 1.5 - 0.075, (1) 1.0 - 0.150 |
| 2 | 3 | 10A1 | 11B1 \| (S1-1) | 10C1 | (1) 2.5 - 0.075 |
| | 4 | (S1-3) | 11B1 \| (S1-6) | 5C1 | (1) 2.0 - 0.075 |

2.500

Plug Key

Vertical line indicates assignment division within cable.

NOTES: Layers 1 and 2 are used for Plug No. 1

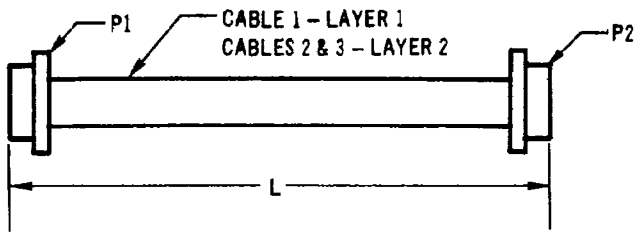
Layers 3 and 4 are used for Plug No. 2

S2-6 indicates 6 spare conductors on 0.150 centerlines.

S1-3 indicates 3 spare conductors on 0.075 centerlines.

▶ Indicates identifications edge of cable.

CABLE ASSY




WIRE LIST

| REF DES - PIN | REF DES - PIN | CABLE CONDUCTOR | WIRE NO. |
|---|---|---|---|
| P1-1 & 2 | P2 - 32 & 31 | 1-1 | 110A |
| P1 - 3 & 4 | P2 - 30 & 29 | 1-2 | 111A |
| P1 - 29 & 30 | P2 - 4 & 3 | 1-15 | 323C |
| P1 - 31 & 32 | P2 - 2 & 1 | 1-16 | 324C |
| P1 - 64 | P2 - 33 | 2-1 | 37B |
| P1 - 63 | P2 - 34 | 2-2 | 38B |
| P1 - 48 | P2 - 47 | 2-17 | 16C |
| P1 - 47 | P2 - 48 | 2-18 | 17C |
| P1 - 43& 44 | P2 -54 & 53 | 3-1 | 42A |
| P1 - 41 & 42 | P2 -56 & 55 | 3-2 | 43A |
| P1 -39& 38 | P2-60 & 59 | 3-4 | 75D |
| P1 -35& 36 | P2-62 & 61 | 3-5 | 76D |

REFERENCE DESIGNATION

| | | |
|---|---|---|
| P1 | PART NO. | 2-1/2" PREMOLDED PLUG |
| P2 | PART NO. | 2-1/2" PREMOLDED PLUG |
| CABLE 1 | PART NO. | 2-1/2" 150 MIL CENTER HIGH DENSITY (16 CONDUCTORS) |
| CABLE 2 | PART NO. | 1-1/2" 75 MIL CENTER STD DENSITY (13 CONDUCTORS) |
| CABLE 3 | PART NO. | 1" 150 MIL CENTER HIGH DENSITY ( 6 CONDUCTORS) |

PLUG ASSY (CABLE SEGMENTS)

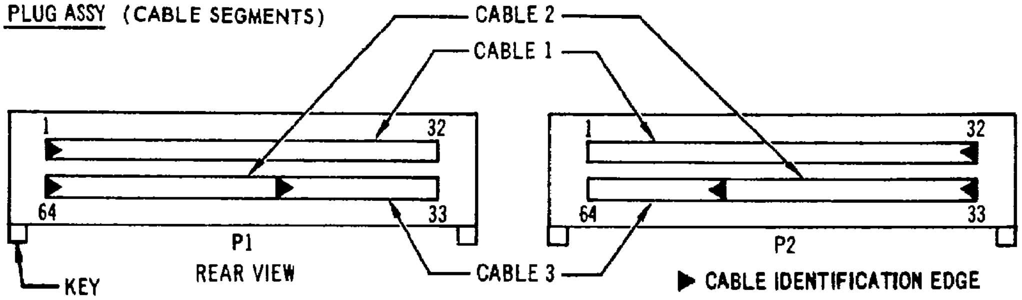



FIGURE 3-68. FCC wire harness drawing.

3.2.6.1 Electromechanical Layout. The installation engineer is primarily concerned with the outline, installation, and interconnecting requirements of the electronic units. The inboard profile drawing, included with the program detail specification, is usually the first scaled drawing for the general arrangement of the end item.

When the program is approved and preliminary design is begun, a scaled electromechanical layout is prepared to show the basic structure, other technology installations, and all electronic units and electrical interfaces. All external electrical connectors, defined by the harness drawings, are shown. By making transparent overlays over the basic layouts, various wire-harness routing and support schemes can be considered. Figure 3-69 shows a typical layout section of FCC harness routing and support. All units would be identified with their external connectors shown and defined.

3.2.6.2 Objectives of FCC Harness Development. Three separate development schemes were tried on the S-IV-B mockup before the final configuration was selected. The general FCC harness development objectives resulting from this program are listed as follows:

a. Keep each harness assembly as simple as possible with none or a minimum number of branches.

b. Route all harnesses inboard of structural cutouts, etc., so that cable-end threading through openings is eliminated or reduced to a minimum.

c. Do not route harnesses behind or under installations that would require removal of equipment or supports for FCC harness installation or removal.

d. Join or group individual harness assemblies in main bundle runs to permit installation, removal, and replacement of separate harnesses with the minimum of disturbance to the other harnesses in the bundle run.

e. Route FCC harnesses close to structure for bundle ground-plane effect.

f. Have wide axis of FCC connectors on electronic units parallel to main bundle runs to permit routing with simple folds.

g. Utilize existing structure sections where possible for FCC support. Add simple lightweight angles, "Z" sections, or hat sections for additional supports. Modular clamp mounting holes should be provided to accommodate clamp-size changes without making new holes in mounting brackets.

3.2.6.3 Full-Scale Development Fixture or Mockup. Although much preliminary design can be accomplished on drawing layouts and transparent overlays, it is highly desirable to use a full-scale, three-dimensional development fixture or mockup for the actual routing, support, and harness definitions. The mockup development should be coordinated with manufacturing and inspection personnel.

A mockup is usually made early in the program from wood and other materials to simulate the structure support and all equipment installations. Very close dimensional control is not required; however, approximate dimensions must be maintained. A development fixture utilizes structure and support essentially in accord with the production drawings with dimensional control so that tubing and wire harnesses developed on the fixture can be installed directly on the production units.

3.2.6.3.1 Initial Installation. By using the manufacturing development techniques defined in Section VI, all FCC harnesses are prepared from the wire-harness drawings and installed per the schemes developed on the engineering layout drawings. The general development objectives of Paragraph 3.2.6.2 and any special program requirements are used in this development.

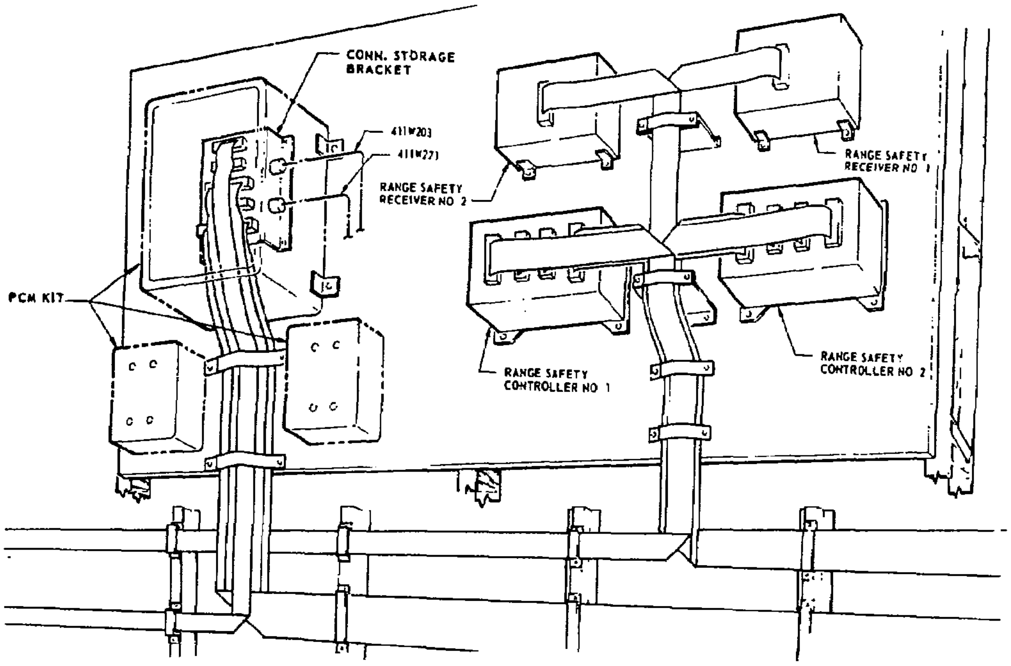



FIGURE 3-69. Typical FCC installation.

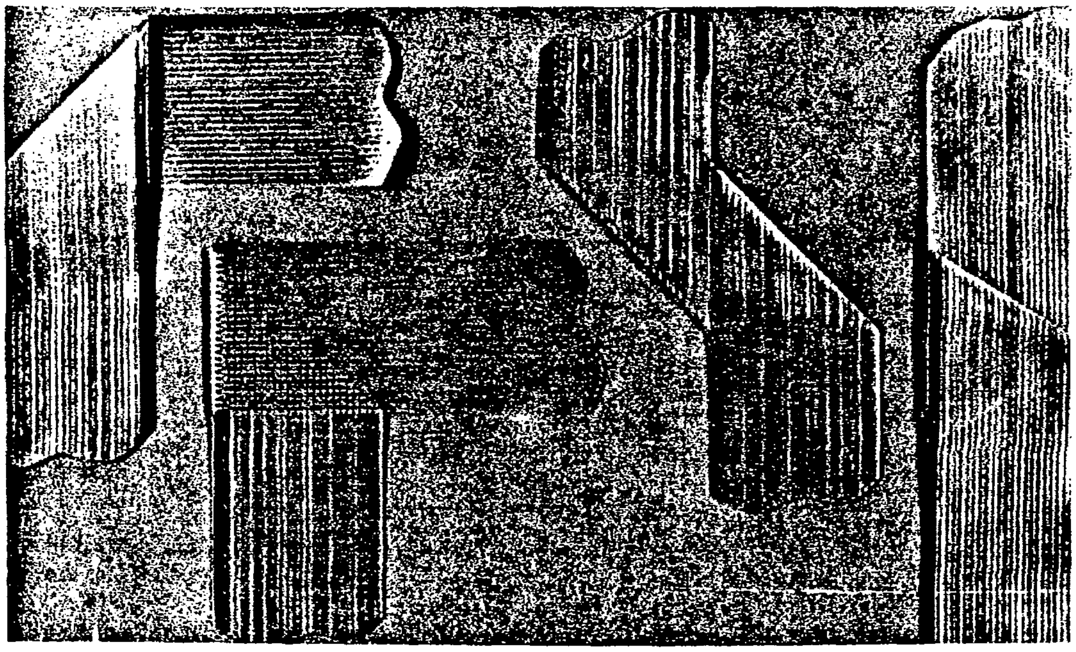


FIGURE 3-70. Cable folding techniques.

3.2.6.3.2 Folding Techniques. Various cable-folding techniques are used, as shown in Figure 3-70, to provide the direction changes and cable registration required. Nonshielded FCC can be folded flat on itself with no bend radius required. Shielded FCC should always be folded with a minimum bend radius, usually 1/8 inch. A permanently installed filler or other device should be used to assure that this radius will be maintained. In those areas where cables branch out of major runs, or where the major run changes its direction by folding, there are two methods to be considered, as shown in Figure 3-71. Folding by group provides neater bundles and additional support, with fewer exposed edges. Folding by cable makes it much easier to install, remove, and replace individual cable assemblies.

3.2.6.3.3 Clamping and Supports. Maximum utilization should be made of existing primary and secondary structure and existing brackets, etc., for FCC harness support. Generally, all that is required is a 5/8-inch-wide flat surface with the rigidity or strength provided by angles, Z's, or other cross-sections. Simple lightweight support sections can be added as required for additional support (Fig. 3-72). By providing modular-spaced mounting holes in these sections, the required clamp sizes and main-bundle-run spacing can be revised without changing the support brackets.

Various type clamps are shown on Figure 3-73 for general FCC support. They are: lap clamps, both single and multiple, for low temperature (100°C max) application; and metal clamps, both cushioned and noncushioned, for high-temperature (200°C) application. Additional information on these clamps and others is contained in Section II. All clamps shown in Figure 3-73 have captivated hardware and are suitable for tabulated widths to accommodate the standard cable widths.

To assist in the proper registration of the cable segments in the FCC harness runs, adhesive tape can be used under all clamps and between clamps. This will assure that cable segments will remain in the proper harness layer and will help maintain each segment in its proper layer position.

3.2.6.3.4 Cable-Segment Registration. It is very important to maintain the cable-segment registration in the harness run as required by the engineering design (Fig. 3-74). It is mandatory that this registration be accomplished on the final mockup or development fixture harnesses since the production units will be patterned from them. This will require that each cable segment be identified together with its index edge. The use of adhesive tape will aid in maintaining this registration during development and installations.

3.2.6.3.5 Final Approval. After the final development has been accomplished, engineering approval is required prior to removal of developed FCC harnesses, and subsequent to their use for patterns in making the production harnesses. At this time, the cable segment registration should be checked in each bundle-run section and at each connector, as well as the general routing and support. Figure 3-75 shows a section of a completed final mockup development.

3.2.7 Other Drawings.

3.2.7.1 Cable Network. The cable network drawings required by many programs show all electronic units listed by title and reference designation number in each area or zone of the end item, each interconnecting harness, and the reference designation for each external electrical connector. For FCC application, it is necessary only to establish a code so that the FCC and RWC harnesses can be easily differentiated.

3.2.7.2 Schematics. Unit schematics define the components and interconnections inside electronic units with no reference to external harnesses; therefore, these drawings are not affected by FCC application to interconnecting harness networks.

Advanced functional schematics define the electrical interconnections between and through electronic units on a system or subsystem basis. It is necessary only to establish a code for FCC so that it can be distinguished from RWC.

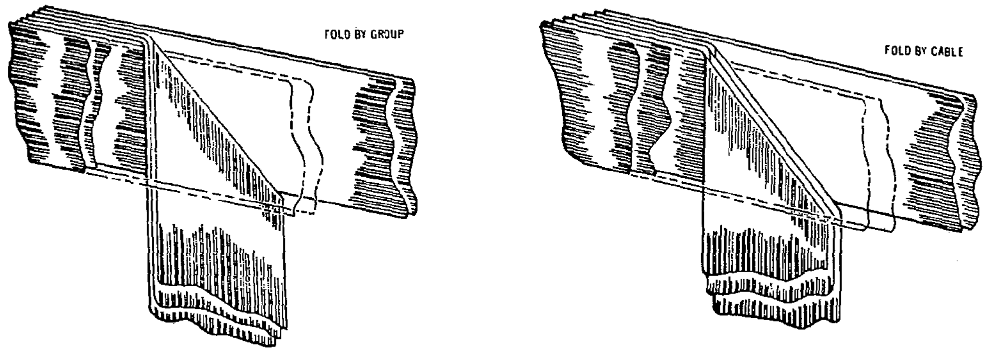



FIGURE 3-71. FCC branchout from bundle run.

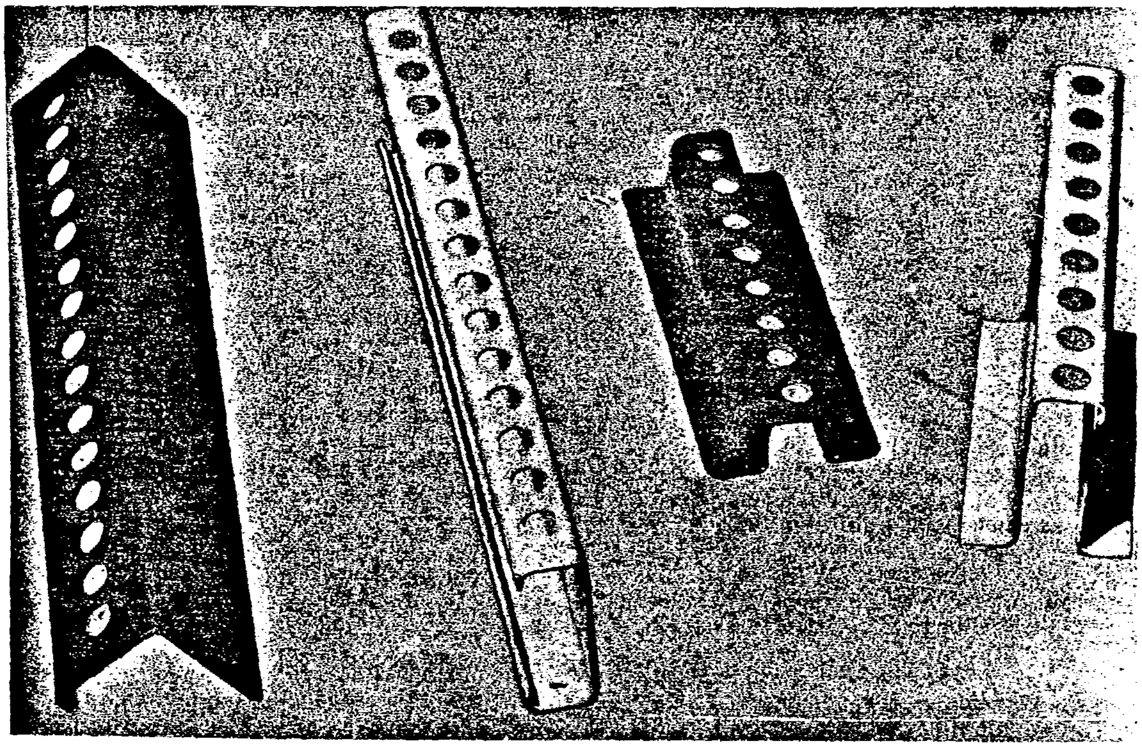


FIGURE 3-72. Sections for FCC support.

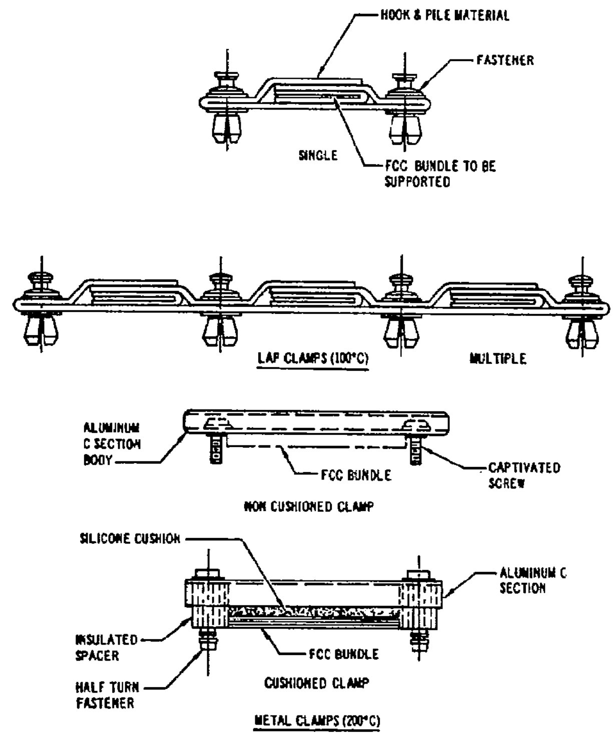



FIGURE 3-73. FCC clamp types.

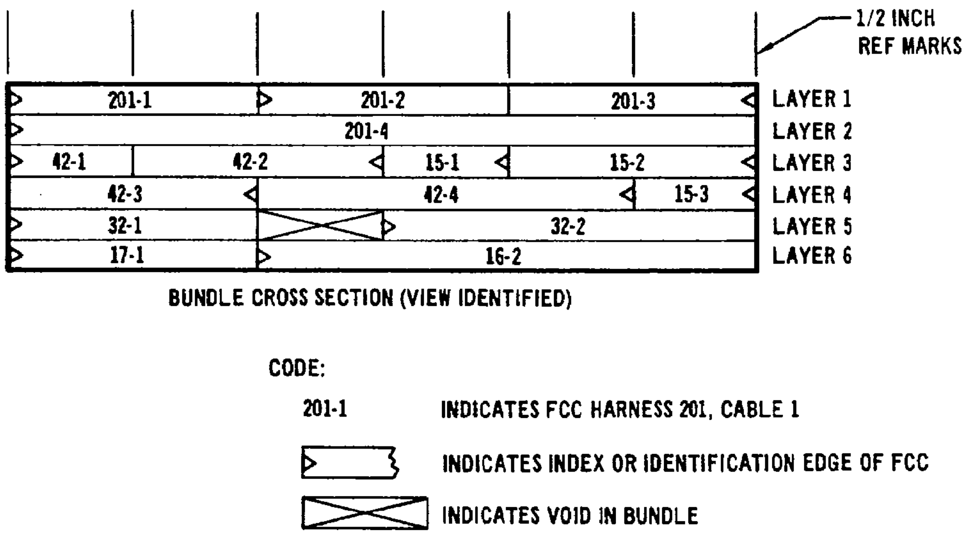



FIGURE 3-74. FCC cable registration.

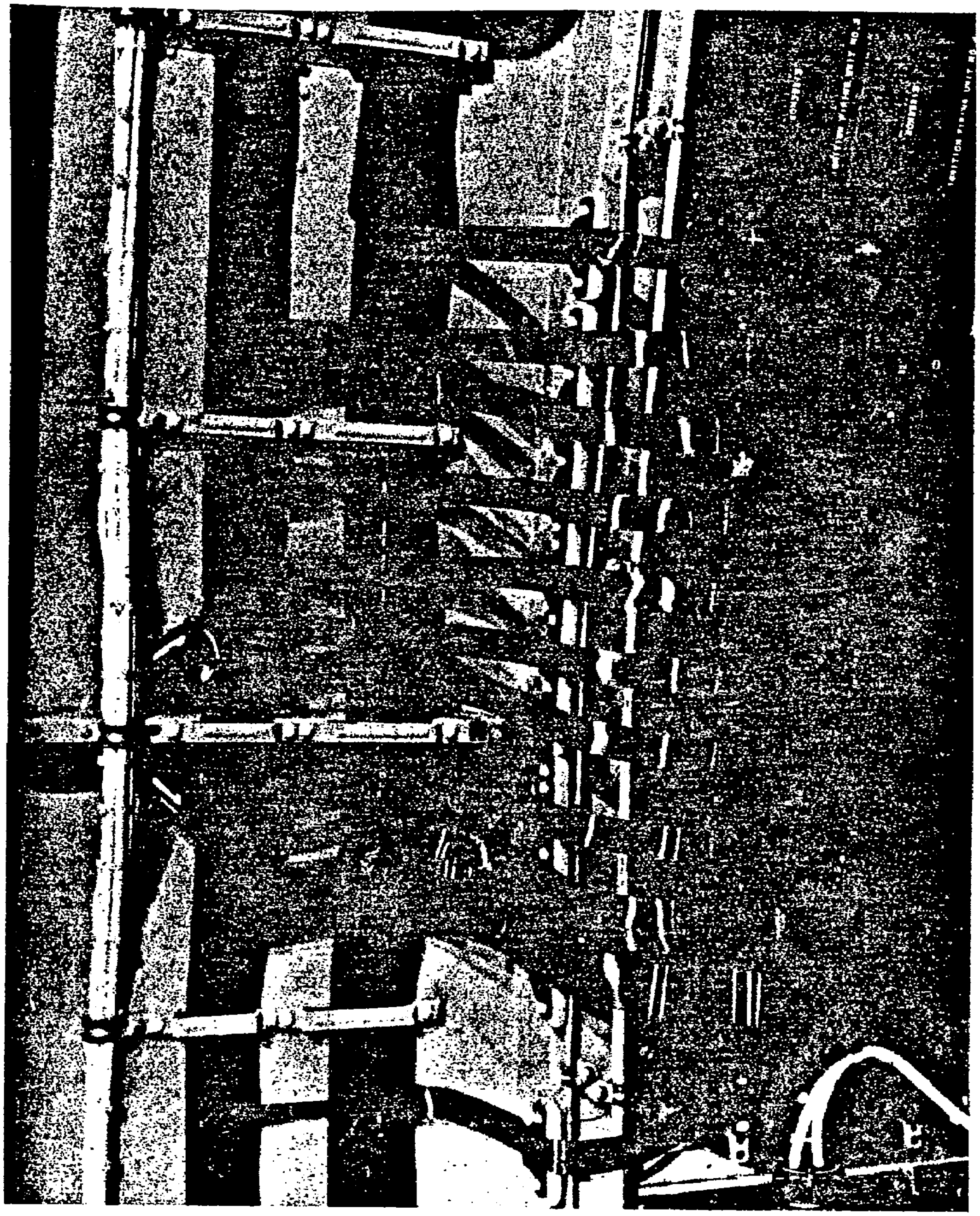


FIGURE 3-75. Final development mockup.

3.2.7.3 Electrical Interface. The electrical interface drawings define the pin-assignment functions and electrical connectors that interconnect with other items such as umbilicals, flight disconnects, or assembly disconnects that mate on joining. These drawings are essentially the same for FCC or RWC systems. If both type interfaces are used, the proper differentiation should be made.

3.2.8 Special Applications. The electrical and mechanical properties of FCC, and the many possible configurations in which it can be manufactured, make it adaptable to numerous special applications.

3.2.8.1 GSE Cabinet Cables. FCC has been used in many GSE cabinet applications to provide electrical conductors between fixed structure and hinged panels or electrical drawers that are usually mounted on slides.

Figure 3-76 shows a current application by the Librascope Group of General Precision for the Mark 48 Fire Control System. The shielded FCC assemblies were fabricated by Ansley West with continuous, very small corrugations at right angle to the cable length to provide the flexibility required. Both panel rotation and drawer translation, plus rotation movements, have been accommodated.

By using insulation materials, such as Teflon and Mylar which can be heat-formed to provide a built-in memory, both corrugated and convoluted coils can be fabricated from FCC to provide retractable cable assemblies for drawer application. Figures 3-77 and 3-78 show examples of the memory-type retractable cable assemblies.

A standard FCC cable can be used in a loop or U-shaped configuration to accommodate drawer extension as shown in Figure 3-79.

3.2.8.2 Storage and Deployment. Requirements for stowage and deployment of electrical interconnecting cables must be accommodated on many programs. The systems to be described are especially simple and efficient. Generally, they are not self-retractable.

Bendix designed an FCC cable-and-stowage reel to meet these requirements on a lunar application (ASLEP) to interconnect lunar experiments to a central data package. Figure 3-80 shows the stowage-reel concept used. Deployments up to 60 feet with FCC, having 32-AWG conductors, were accommodated with the advantages of simplicity, lightweight, small space, and high reliability. These advantages will warrant consideration of this system for all future stowage and deployment requirements.

A sample corrugated FCC configuration (Fig. 3-81) offers unique advantages. The sample 2-inch-wide cable with 50-mil centerline conductors can be compressed to 1.25 inches for stowage. The free length is 10 inches; the extendible without exceeding the elastic limit of the corrugated cable is 15 feet; and the total cable length is 30 feet. It is difficult to envision how any other system could stow, for reliable deployment, the number and length of conductors in a 1.25- by 2.0- by 2.0-inch stowage volume.

3.2.8.3 Hinge Applications. The geometry and physical characteristics of FCC make it ideally suited for use as a hinge medium, providing both the hinge and the means of transferring many electrical signals across the hinge line.

Two basic configurations have been developed by NASA/MSFC and are shown in Figure 3-82. The FCC can be bonded directly to the adjacent structure to absorb the required sheer load and to provide simple cable support. The Type A hinge is suitable for low torque, small angular movement. The Type B hinge can be rotated through larger angles with greater torque required.

A little imagination by the designer will result in many practical FCC hinge-line applications for interwiring of electronic units and for external interconnecting harnesses.

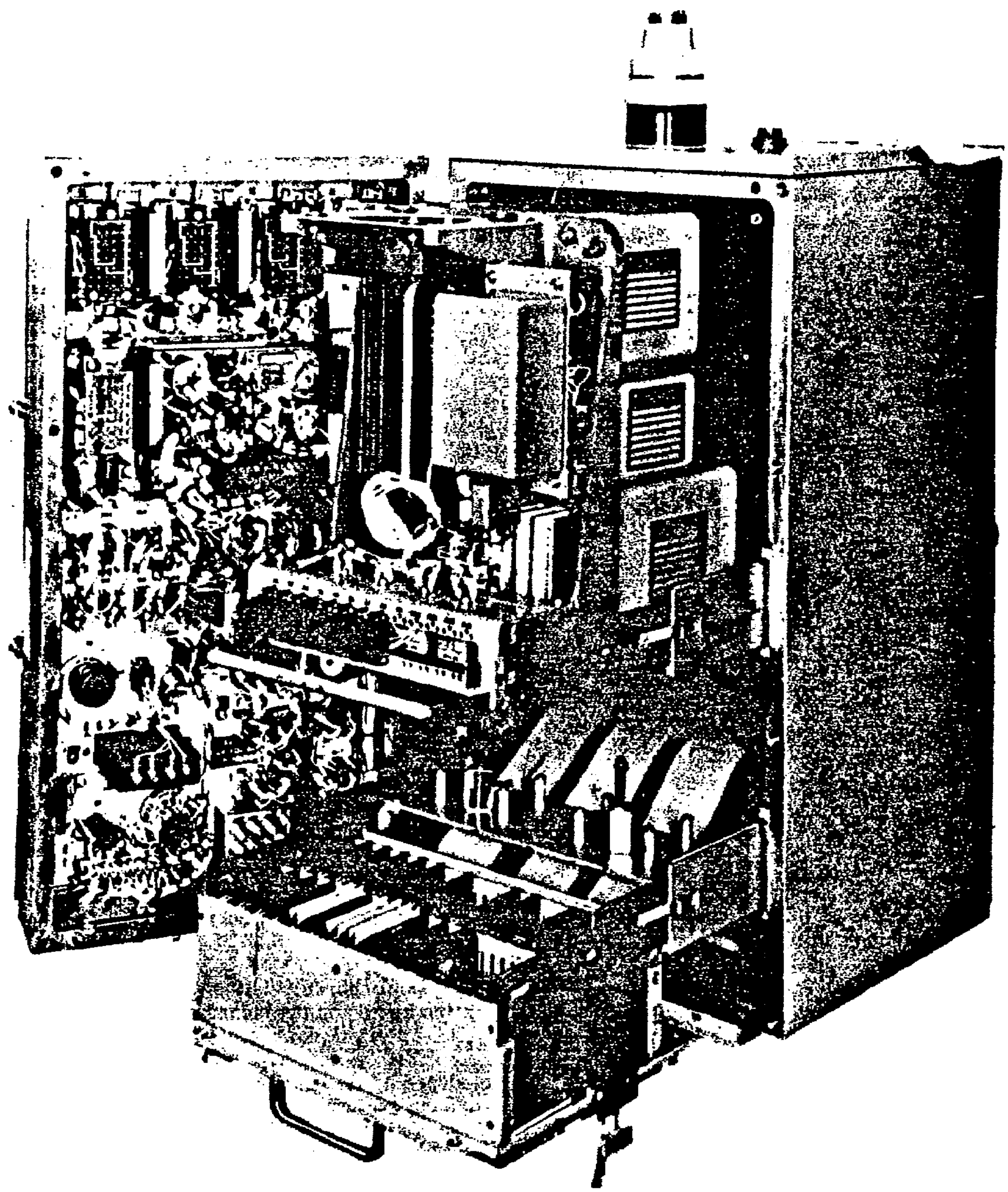


FIGURE 3-76. Mark 48 Fire Control - Libroscope Group of General Precision.

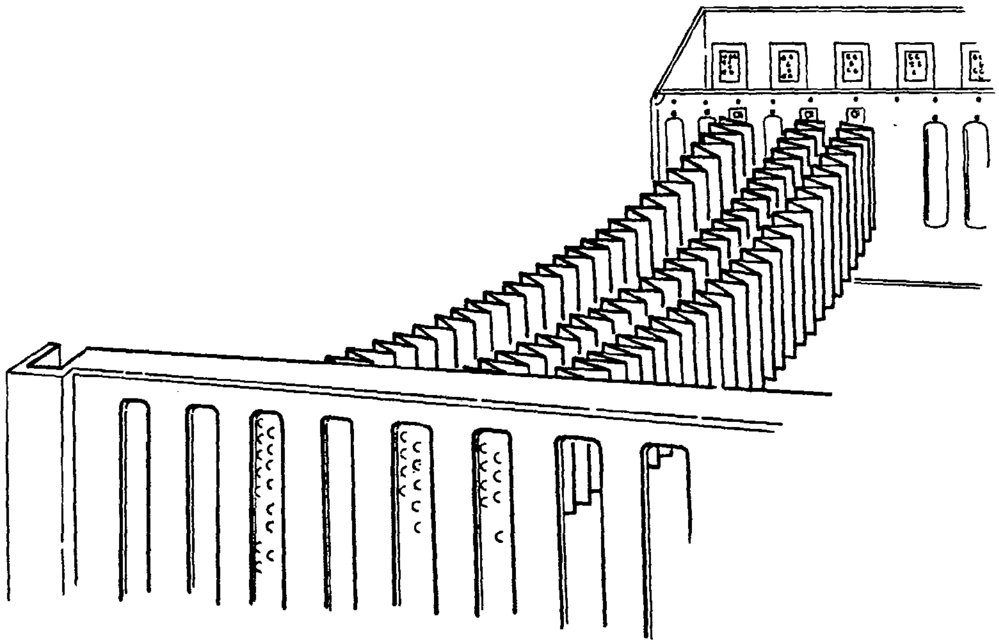


FIGURE 3-77. Retractable FCC rack-mounted drawer assembly.

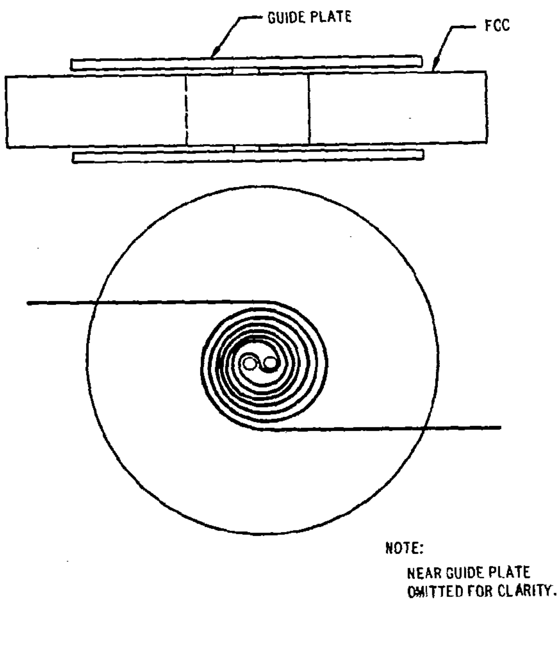



FIGURE 3-78. Convoluted coil.

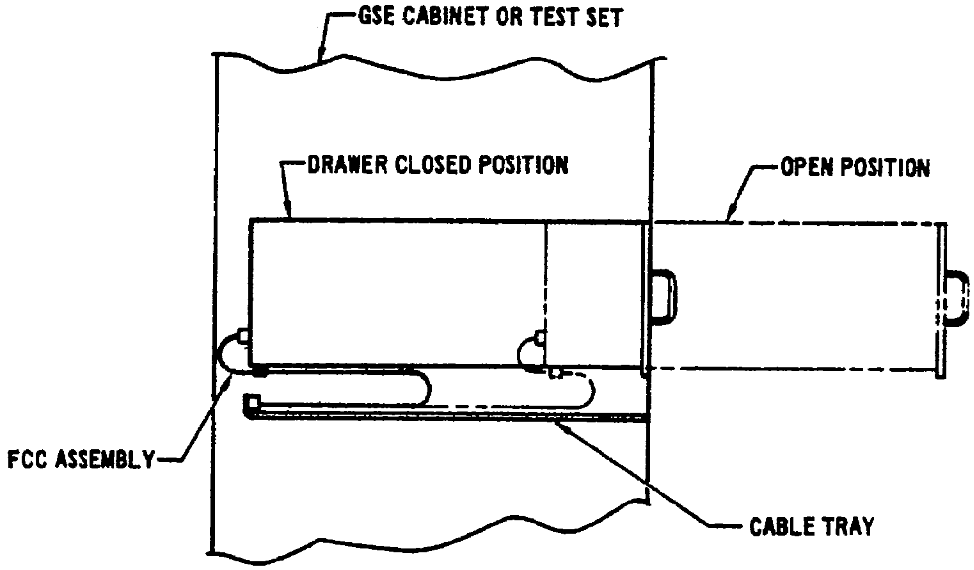



FIGURE 3-79. U-shape trailing cable.

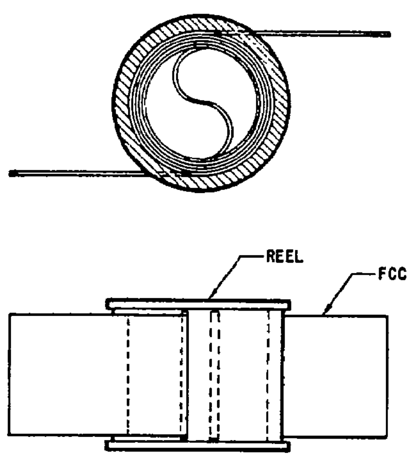



FIGURE 3-80. FCC stowage reel.

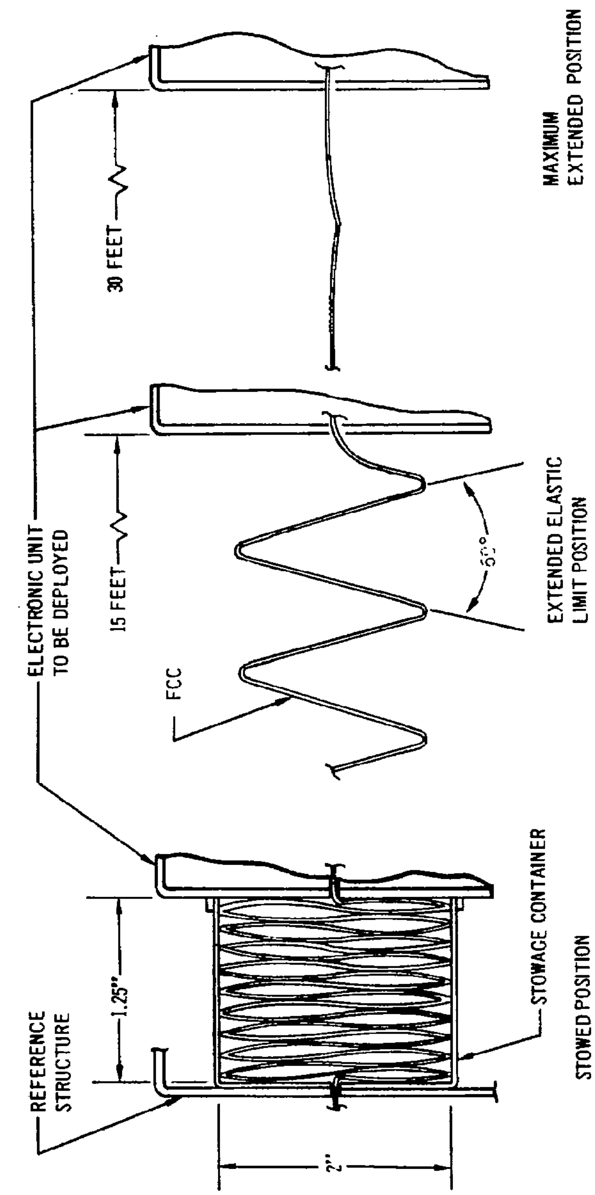
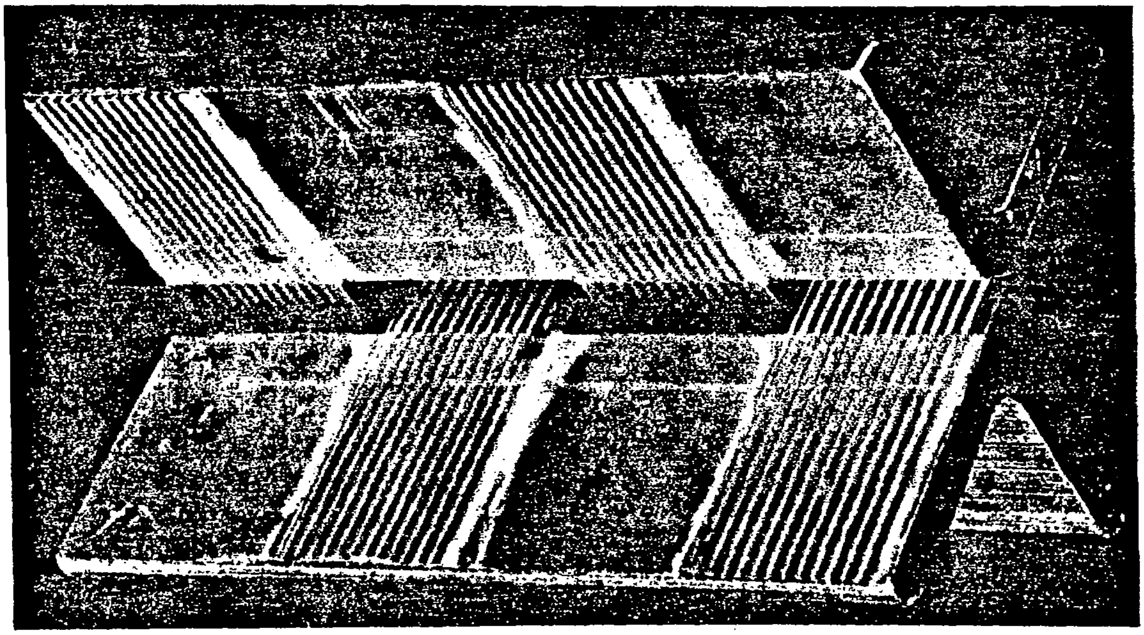
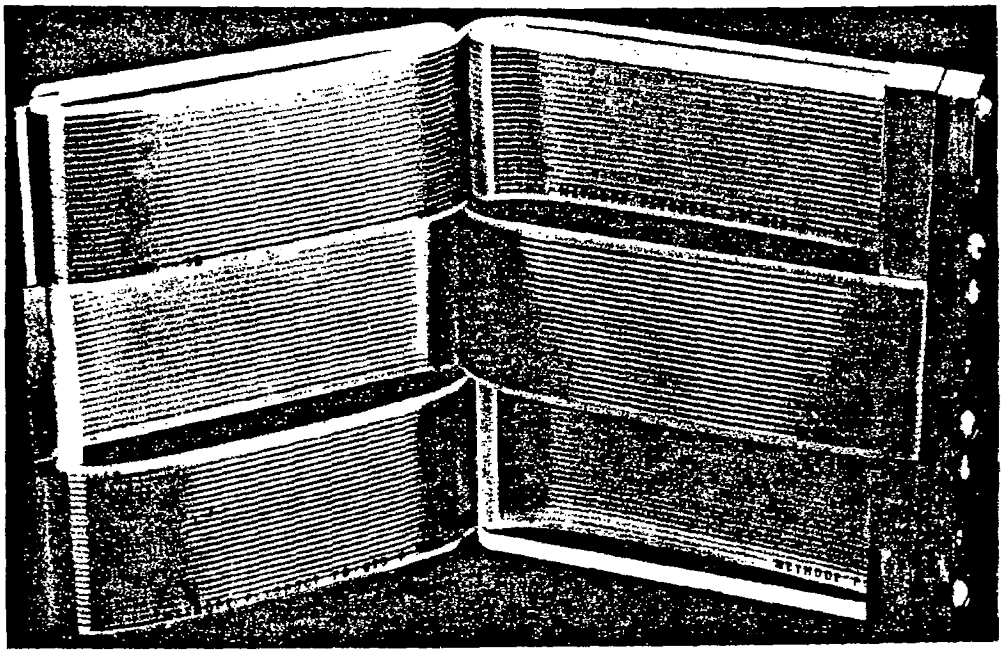



FIGURE 3-81. Corrugated FCC cable stowing.



TYPE A - LOW-TORQUE SMALL ANGLE OF ROTATION



TYPE B - HIGHER-TORQUE LARGE ANGLE OF ROTATION

FIGURE 3-82. FCC hinge applications.

3.2.8.4 Low-Torque Application. FCC is to be used on the Apollo Telescopic Mount (ATM) program by MSFC to transmit 2500 conductors across two sensitive gimbal systems. It was a requirement that the crossing electrical harness require a minimum of torque for gimbal rotation. Extensive studies indicated that the FCC system required approximately one-tenth the torque required for the conventional RWC system. Figure 3-83 shows the harness configuration which can be used for gimbal crossings. Figures 3-84 and 3-85 show a method used for determining and the values required for rotation and bending of the FCC harness.

3.2.8.5 Angular Rotation. The physical characteristics of FCC make it especially suitable for applications requiring the transfer of electrical signals across a rotating joint. Continuous oscillation of 180 degrees is used frequently for gun turrets, radar antenna arrays, etc. One shot or limited operations can be made for spin-up and other requirements of many rotations for missile, satellite, and other applications. Figure 3-86 shows the basic configuration used for accommodating limited rotation with FCC.

For the limited rotation case of 180 degrees, the design is essentially noncritical, and a sample experiment with a mockup FCC harness will determine the drum and arbor diameters and the number of FCC turns required.

For maximum angular rotation in one shot or limited operation, studies have indicated that up to 20 complete revolutions can be made with a drum diameter of 1.7 inches and an arbor diameter of 1 inch. Up to 50 complete revolutions can be made if the drum diameter is increased to 2 inches. For maximum rotation and reliability, the cable thickness should be kept to a minimum, and the insulation material should have as small a coefficient of friction as possible.

3.2.8.6 Special Electrical Configurations.

3.2.8.6.1 Minimum-Spurious-Coupling Configurations. Minimum-spurious-coupling configurations are discussed as follows:

a. Capacitive coupling - When two conductors, or other objects, are in close proximity to each other, they form a small but finite capacitor. The conductors are capacitor plates and the conductor insulation and other nonconductive material form the capacitor dielectric. Any varying voltage on one of the conductors is capacitively coupled into the impedances connected to the other conductor.

b. Magnetic coupling - The mere existence of a longitudinal conductor creates a small but finite inductor. When two conductors are in close proximity to each other, they form a transformer because of their coupling or mutual inductance. Any varying current flow in one conductor is inductively coupled into the other conductor as an induced voltage because of the transformer action.

c. Single-ended circuitry with a common return - Conventional circuitry frequently utilizes a structural ground plane as a reference point and circuitry return path.

   1. Capacitive coupling - Since any two objects have a mutual capacitance (sometimes referred to as series capacitance), a conductor has mutual capacitance to other conductors or objects that are a source of interference. The same conductor also has a mutual capacitance to the return path conductor and surrounding objects, other than the interference source, that is commonly known as shunt capacitance.

      Interference is capacitively coupled from the interfering conductor, through the series capacitance, to the susceptible conductor which has a parallel shunt capacitance between it and the circuit return path. The shunt capacitance is also normally paralleled by additional terminating impedances at each end of the conductor. The capacitive coupling circuit is basically a frequency-sensitive, capacitive voltage divider with an output loaded by the paralleled source and load termination impedances. Voltages appearing between the function conductor and the structural ground plane, or other common return path conductor, are known as common-mode voltages.

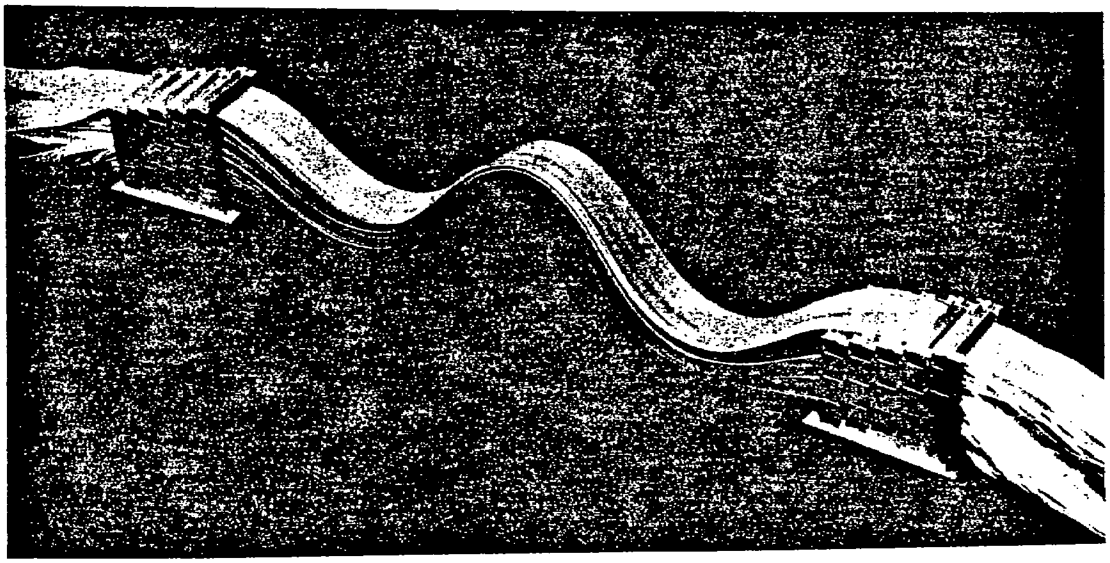


FIGURE 3-83. Low-torque FCC harness.

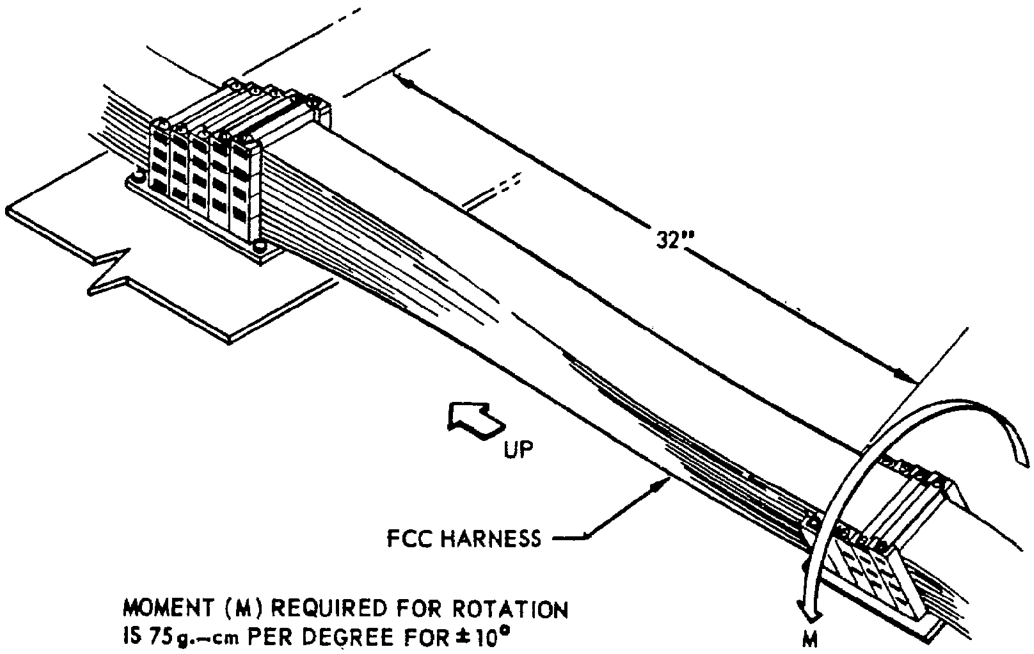
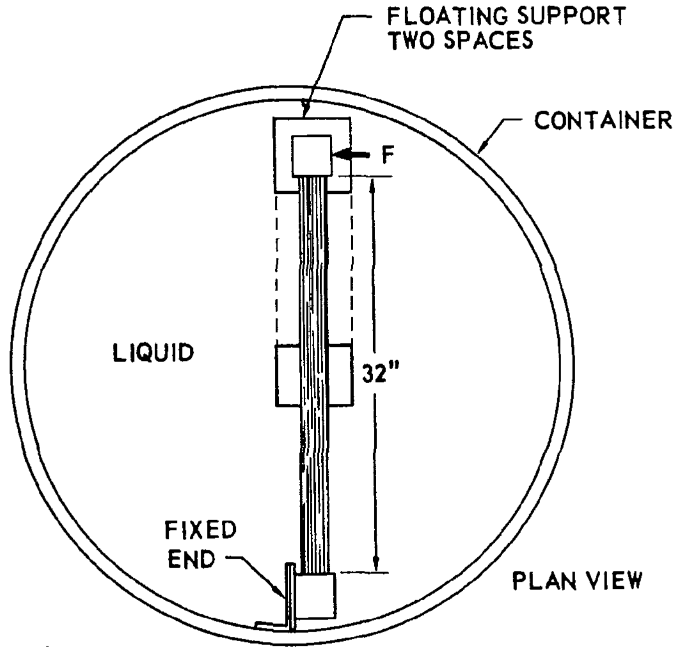



FIGURE 3-84. Torque required for rotation.




FORCE (F) REQUIRED IS 1.5 GRAMS PER DEGREE OF BENDING FOR MOVEMENTS LESS THAN 8 INCHES.

FIGURE 3-85. Force required for bending.

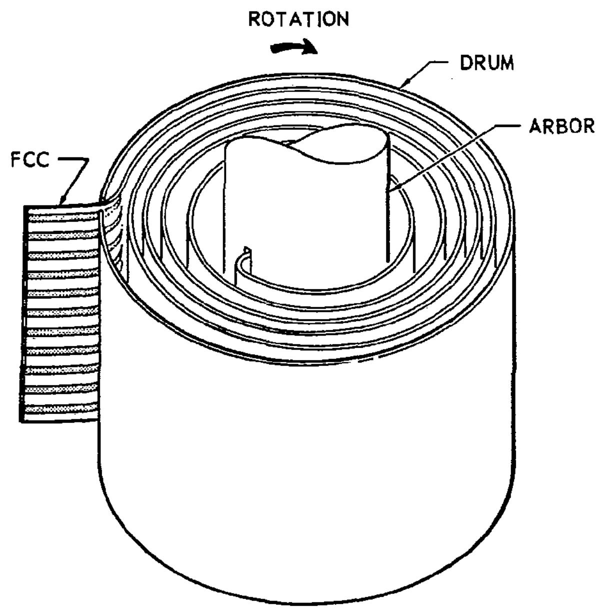



FIGURE 3-86. FCC application for rotational devices.

2. Magnetic coupling - Since any two objects also have a mutual inductance, a conductor has mutual inductance common to other conductors or objects that are a source of interference. The same conductor also has a loop-inductance component that appears in series with the conductor, but is not mutual or common to other conductors. The sum of the mutual inductance and loop inductance forms the self, or total, inductance of the conductor that interacts with circuitry in the termination hardware.

   The interference-carrying conductor is equivalent to a transformer primary winding. The susceptible conductor is equivalent to a transformer secondary winding with the same number of turns as the primary winding. If this transformer had unity coupling between windings, which would be the case if an iron core were available, virtually identical voltages and currents would appear in both windings if similar source and load impedances were connected to the transformer windings. Cable conductors do not have a ferrous core; therefore, the resultant air-core transformer has less-than-unity coupling.

   Currents flowing through the primary winding generate a magnetic field, but only a portion of this magnetic field intercepts the secondary winding because of the absence of an iron core capable of confining the magnetic field to a path that intercepts the secondary winding. Since only a portion of the energy in the magnetic field intercepts turns in the secondary winding, the resultant-induced voltages and currents are smaller than those present in the primary winding. This is the equivalent effect that would be produced by a transformer in which only part of the primary-winding turns and part of the secondary-winding turns were wound on a common iron core, while the remaining primary-winding turns and secondary-winding turns were magnetically isolated from each other.

   The resultant inductive Tee network has a common inductive reactance or mutual inductance that is equivalent to a unity-coupled transformer connected between the source and load terminations. A series, or loop, inductance appears between the primary winding of the transformer and the source termination. A second series, or loop, inductance appears between the secondary winding of the transformer and the load termination. The conductor equivalent to the transformer primary winding has a total, or self, inductance equal to the sum of the source circuit-loop inductance and the primary inductance of the mutual-inductance virtual transformer. The conductor equivalent to the transformer secondary winding has a total, or self, inductance equal to the sum of the load-circuit loop inductance and the secondary inductance of the mutual-inductance virtual transformer.

   Interference is inductively coupled from the interfering conductor, through the transformer action of the mutual inductance to the susceptible conductor, which has a termination impedance connected from each end to the common return path, forming a complete current loop. The inductive-coupling circuit is basically a frequency-sensitive, two-section, inductive voltage divider with an output loaded by the susceptible circuit source and load termination impedances.

Balanced Transmission Systems

Since the voltage induced in each conductor is proportional to the field intensity at the conductor, it is possible to induce identical voltages in two conductors if they can be made to occupy points of identical intensity within the field. If identical voltages are induced in both conductors, no difference exists between them; the functional circuit has no interference voltage impressed across it, and the induced interference voltages have been cancelled.

The FCC configuration that comes closest to causing both conductors to occupy the same point in space is an extremely low-profile cable with two layers of conductors, in registration, arranged to provide over-and-under pairs of conductors with a minimum dielectric thickness separating the paired conductors. If the conductor layers are in perfect registration, the field intensity generated by an interference source located at either edge of the cable will be essentially identical at both conductors. If the interference source were located above or below the cable, the difference in field intensity appearing at the two over-and-under conductors would be equal to the ratio of the vertical conductor center-to-center spacing to the geometric mean separation distance of the interference source. Since the vertical conductor separation is approximately 0.10 millimeter, and the interference source separation distance is ordinarily orders of magnitude greater than this dimension, virtually perfect cancellation of the interference is achieved. Even when the source of interference is in the same cable stack, a useful reduction in the effective line-to-line mode, interference level will result from the use of this configuration.

The FCC configuration that provides the next closest approach to causing both conductors to occupy the same point in space is side-by-side conductors in the same layer of cable. If the conductors are in the same vertical plane, the field intensity generated by an interference source centered above or below the conductors will be identical at both conductors. If the interference source were located off either edge of the cable, the difference in field intensity appearing in the two side-by-side conductors would be equal to the ratio of the horizontal conductor center-to-center spacing to the geometric mean separation distance of the interference source. Since the horizontal-conductor separation is approximately 2 millimeters, and the interference source separation distance is ordinarily orders of magnitude greater than this dimension, virtually perfect cancellation of the interference is achieved. Even when the source of interference is in another edge adjacent layer of cable in the same cable assembly, a useful reduction in the effective line-to-line mode interference level will result from the use of this configuration.

Balanced or Differential Circuitry

Differential devices respond to differences in the absolute levels existing at the differential inputs and are extremely insensitive to absolute levels simultaneously present at the differential inputs within the dynamic range of the available circuitry. Until recently, the inherent advantages of balanced and/or differential devices could be used only infrequently because of increased circuit and hardware complexity.

Each balanced circuit is essentially two single-ended circuits connected face-to-face electrically. A differential circuit is basically a balanced circuit with additional circuitry for the optimization and maintenance of balance between halves of the basic balanced circuit. Since individual integrated circuits, medium-scale integrated circuit arrays, and large-scale integrated circuit arrays are becoming commercially available, the limitations previously imposed by increased circuit complexity are no longer critical factors.

Since balanced or differential circuitry is sensitive to line-to-line mode rather than common mode interference, the effective interference level at the susceptible circuit imputs is equal to the common mode interference amplitude divided by the reciprocal of the fractional circuit unbalance existing at the susceptible circuit inputs.

3.2.8.6.2 Transmission Lines. Transmission lines are conductors with controlled electrical characteristics used for the transmission of high-frequency or narrow-pulse type signals. The impedance of a transmission line is a function of the distributed series inductance and distributed shunt capacitance of either the balanced conductor pair or the single-ended conductor-and-shield ground plane as a pair. The characteristic impedances of typical conductor configurations, and the formula for deriving the characteristic impedance of a transmission line from the basic electrical properties of the conductors, are given in Paragraph 3.2.3.1. Many standard conductor configurations, both side-by-side and over-and-under, are suitable for transmission line use.

The distributed inductance and capacitance of a transmission line form a high-frequency resonant circuit. When the termination impedances at both ends of the conductor equal the characteristic impedance of the transmission line, a matched-impedance situation exists and the distributed resonant circuit is critically damped.

If the source and/or load terminations do not match the transmission line characteristic impedance, the distributed resonant circuit is not critically damped, and any rapid change in applied voltage or current causes the resonant circuit to ring and generate a damped wavetrain at the frequency of the resonant circuit. Typical cables have relatively high ringing frequencies; therefore, the damped wavetrain appears as a positive or negative spike, or fine fuzz, wherever rapid transitions in waveform amplitude occur. The resultant pulse distortion cannot be tolerated in many systems.

When high-frequency sinusoidal waveforms are applied to a mismatched transmission line, standing waves are distributed along the length of the conductor because of the reactive current flows and voltages that exist in a resonant circuit, and the function amplitude becomes dependent on cable length and frequency.

Impedance-mismatched transmission lines are highly reactive at most frequencies. At high frequencies, relative to cable length, the sign and magnitude of this reactance changes rapidly, causing the transmission losses and received function amplitude to fluctuate with frequency. Except at a few isolated frequencies where a coincidental impedance match occurs, the magnitude of these losses is extremely high and relatively difficult to predict.

Electromagnetic energy travels through free space at 300, 000, 000 meters per second. Most practical transmission lines have a velocity of propagation about two-thirds that of air, or approximately 200, 000, 000 meters per second. The transit time of digital pulses through even short lengths of cable can be significant in current high-speec digital systems. Variations in cable lengths can also cause digital pulses to arrive in the wrong time sequence where parallel pulse transmission is involved.

Nonferrous-shielded conductors have limited bandwidth, compared with unshielded conductors, because of the large shunt capacitances between the conductors and shields, and are therefore suitable only on low-speed digital systems. Ferrous-shielded conductors have a very restricted bandwidth due to both the large series conductor inductance and shunt conductor-to-shield capacitance; therefore, ferrous-shielded conductors are generally unsuitable for digital transmission systems. The high-resultant characteristic impedance of this system is generally incompatible with digital hardware.

Measured transmission loss for energy conducted by a centrally located conductor in a variety of cable configurations is shown in Figures 3-87 through 3-90.

3.2.9 Mechanized Design. A large portion of the design of FCC harnesses can be mechanized with the use of a digital computer and appropriate programs. This can aid designers of a complex electronic system by providing relief from much of the repetitive, tedious, and time-consuming phases of harness design processes.

3.2.9.1 Capability of Mechanized Design. A mechanized harness-design system can accomplish much of the routine assignment of identification information, bookkeeping, and drawing effort that would ordinarily be done by designers' aides.

A description of tasks that can be performed effectively by the computer system is offered. This system can:

a. Read, compile, and file (e.g., magnetic tape) information on approved parts, with appropriate performance characteristics and specifications from which harness assemblies may be fabricated.

b. Read, compile, and file point-to-point wiring requirements.

c. Process and analyze wiring requirements, and sort out those which cannot be met by available approved parts. The system can then analyze the remaining requirements to group conductors into cables, and within each cable, according to the location of each conductor termination and to EMC zoning criteria (see Paragraphs 3.2.3.2.1 and 3.2.10.2).

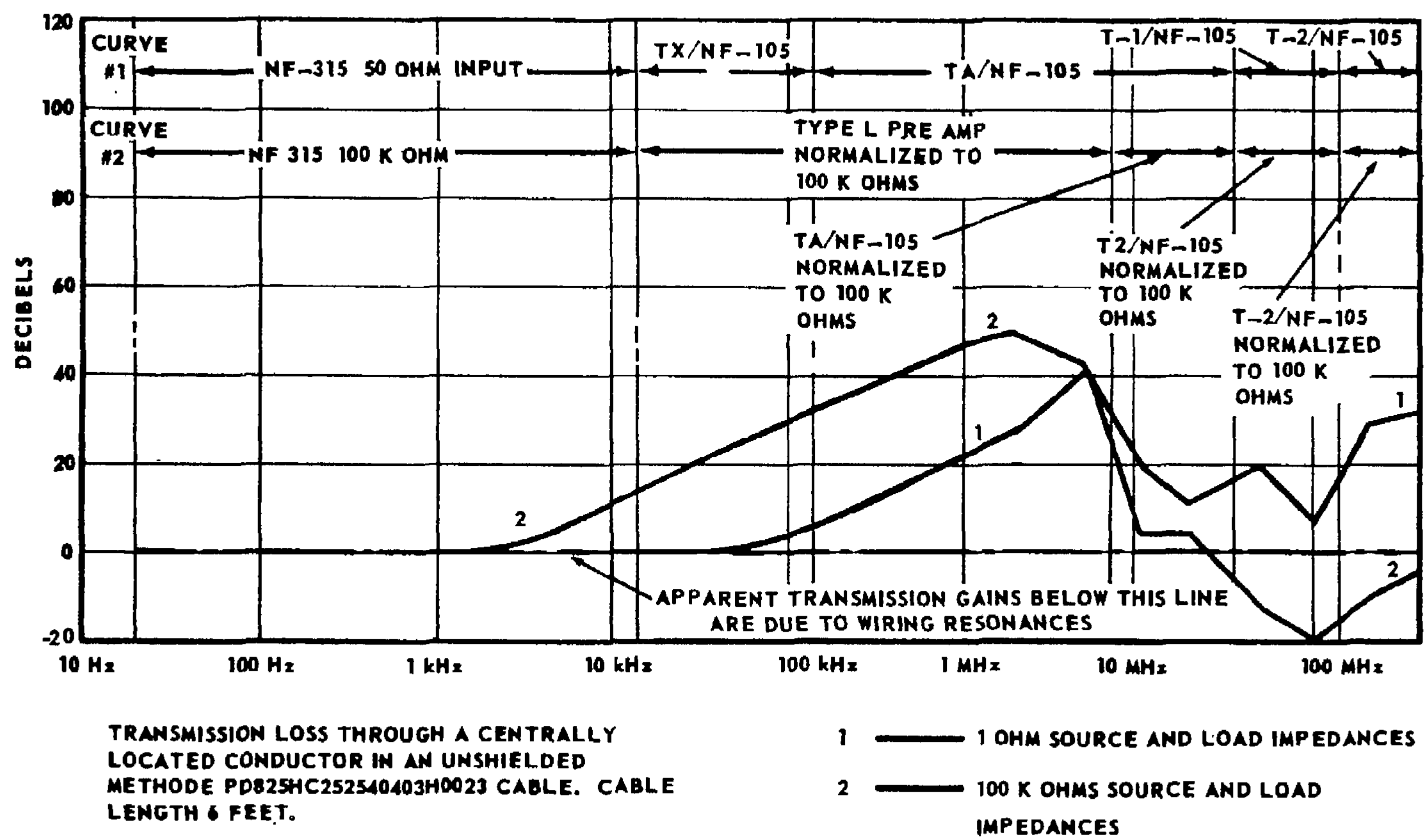
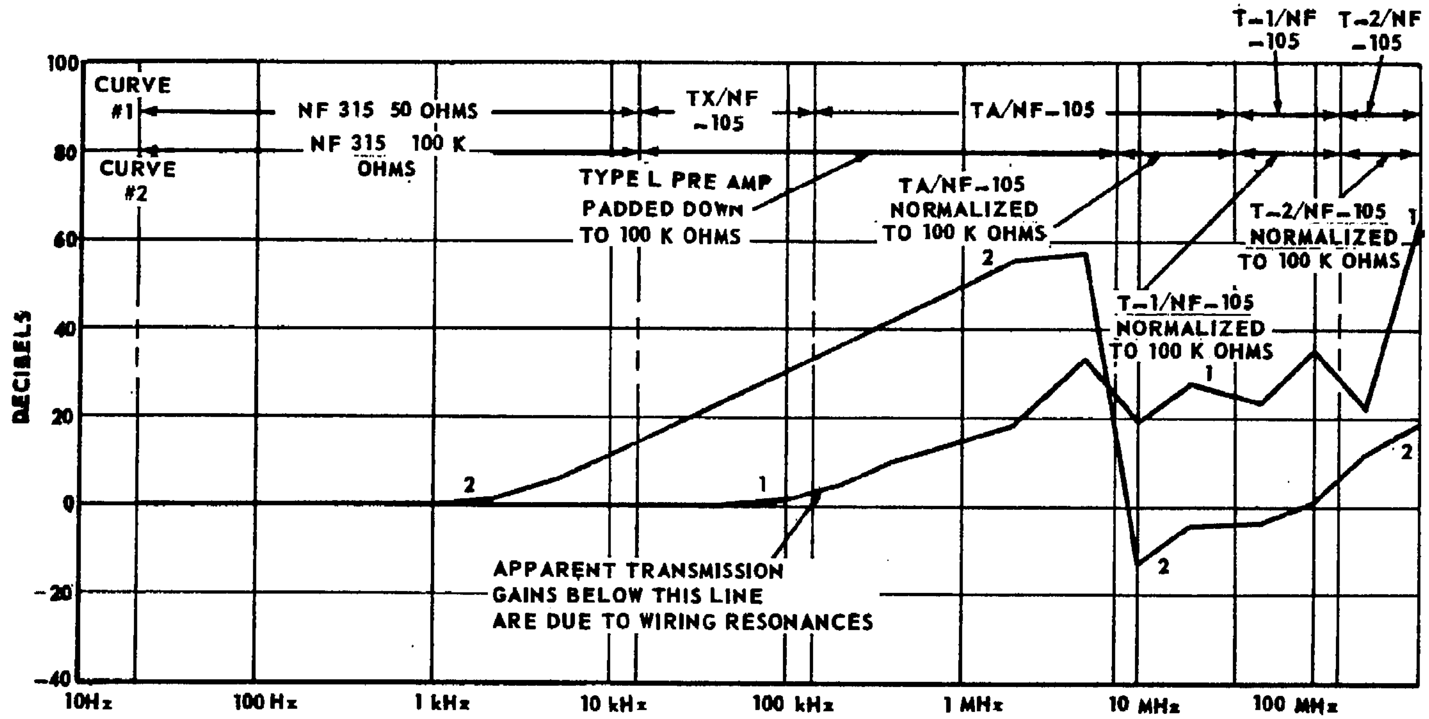



FIGURE 3-87. Transmission loss – unshielded cable.




TRANSMISSION LOSS THROUGH CENTER CONDUCTOR OF METHODE PD818H821T0292509 MESH SHIELDED CABLE. CABLE LENGTH 6 FEET.

1 —— 1 OHM SOURCE AND LOAD IMPEDANCES

2 —— 100 K OHM SOURCE AND LOAD IMPEDANCES

FIGURE 3-88. Transmission loss - mesh shielded cable.

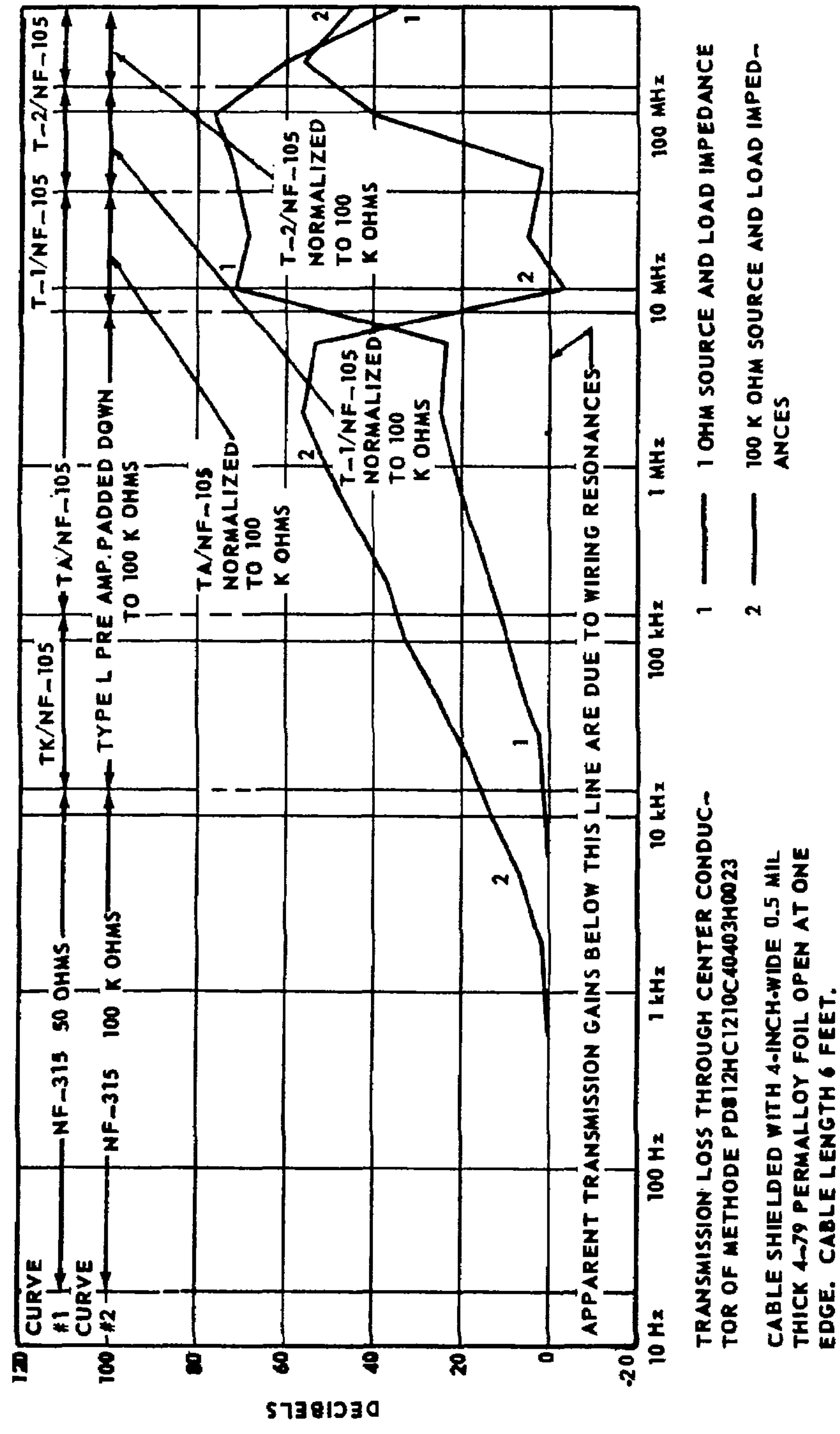
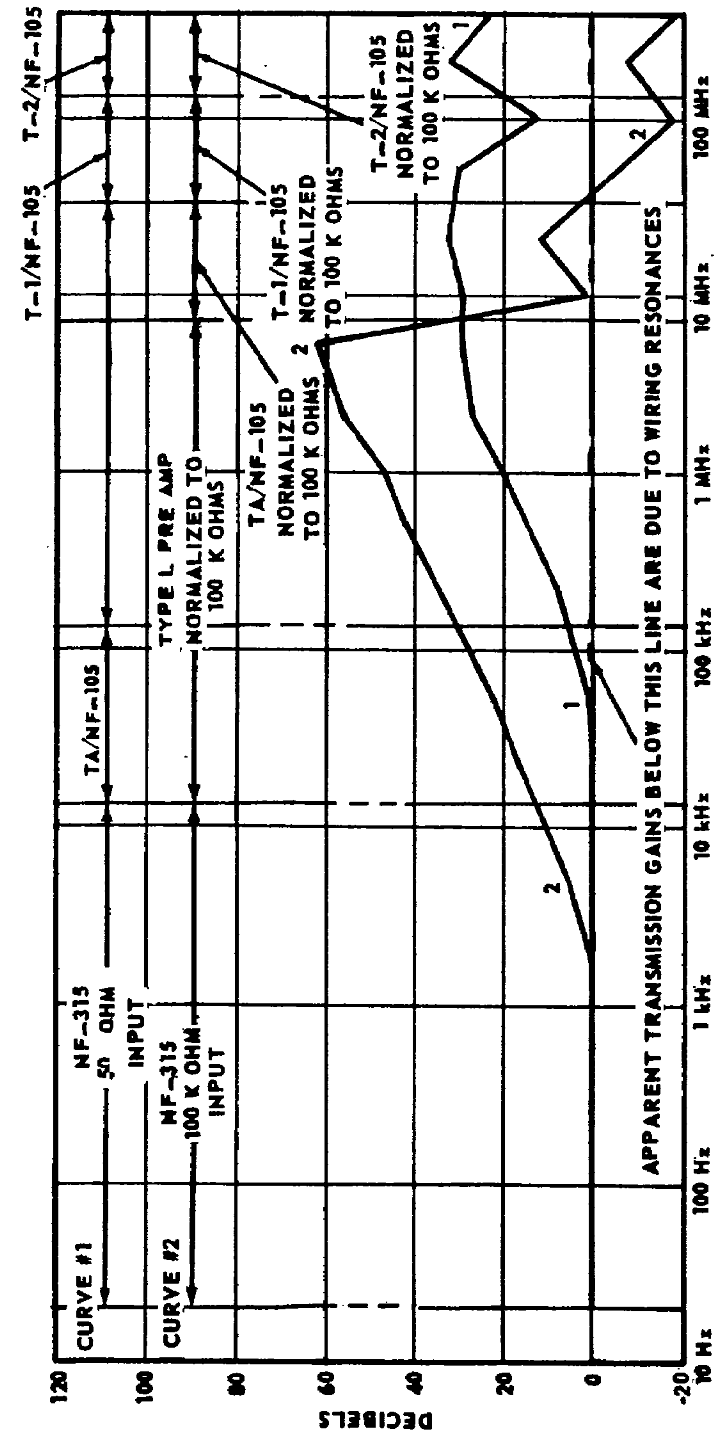



FIGURE 3-89. Transmission loss - solid 4-79 Permalloy shielded cable.




TRANSMISSION LOSS THROUGH ONE CONDUCTOR
PF A SUPRENANT RC-4538N (300) JA 10 (15)
SHIELDED TWISTED PAIR ROUND WIRE CABLE.
CABLE LENGTH 6 FEET.

1 —— 1 OHM SOURCE AND LOAD IMPEDANCE

2 —— 100 K OHMS SOURCE AND LOAD IMPEDANCE

FIGURE 3-90. Transmission loss - woven copper shielded cable - RWC.

d. Assign cable network path routing from information generated in item c above including a specified quantity of spares for each network path. It can then assign, from the approved parts list, the type and quantity of cable.

e. The computer system can select the type and required quantity of approved connectors to meet the cable-termination requirements.

f. The computer system can sequentially assign identification numbers for each cable segment.

g. The computer system can prepare wiring and fabrication information for automatic placement and pictorialization on manufacturing drawings.

An optional, but desirable, subprogram within the mechanized design system would be a routine to optimize the cable routing and layout, in conjunction with equipment layout, according to some objective function (e.g., weight and total cost of materials and labor) within limits of specified constraints (e.g., location of mechanical obstructions, specified minimum percentage of spares in each path of the cable network, and EMC isolation requirements). This is compared with a simpler assignment of cable routing according to preestablished ground rules, which may give reasonably good, but less than optimal, results. Because of the complex nature of such a subprogram, it would be appropriate to add it to the system at a later date after the simpler routines are adequately checked out and operate satisfactorily.

3.2.9.2 Computer Program Design Requirements. The computer program and subroutines written to accomplish automated flat-cable design should be written and documented to meet the following requirements:

a. The program should provide the capability of allowing the user to extend the contents of data tables to add items previously undefined. This extended capability should be allowed with little or no program modification and without causing obsolescence of existing data files.

b. The program should be modularly constructed to facilitate addition of extended capabilities and to facilitate modifications. Each discrete function should become a subroutine. All subroutines should be under control of a supervisory routine or executive.

c. Programming should be done in a higher-level machine-independent language such as FORTRAN IV. This facilitates implementation of the program on more than one computer hardware configuration.

d. The program should be documented sufficiently to facilitate interpretation and modification. Comment cards should be associated with each decisive statement to reflect what the program is doing. Each subroutine should be documented in the following manner:

1. The purpose of the subroutine.

2. Input arguments.

3. Output arguments.

4. Error returns.

5. Internal variables used.

6. Restrictions to using the routine.

7. Names of all subroutines used by this subroutine.

3.2.9.3 Computer Description. The system described in this handbook, if written in a machine-independent language such as FORTRAN IV, could be run on almost any medium- or large-scale computer. The extent to which this system is implemented is the only major variable in the extent of machine configuration. Examples of machine configurations are the UNIVAC 1108, IBM SYSTEM 360 (MODEL 65), and the SDS 9300.

A basic system hardware configuration should include the following:

a. Console and mainframe - Used by the computer operator to control operation of the program.

b. 32K words of core memory - Holds system software and controls execution of program.

c. Card reader - Input options and control information.

d. Typewriter - Messages to operator for mounting of tapes, system errors.

e. Off-line printer - Generate printed output.

f. Magnetic tape units - Means for inputting standard files for program access.

g. Random access bulk storage unit - Store large amounts of data to be operated on by the program.

3.2.10 EMC Theory

3.2.10.1 EMC Fundamentals. As stated previously in Paragraph 3.2.3.2, a major source of electromagnetic incompatibility in a large electrical/electronic system is interconnection cabling within the system. This cabling network can be too massive for convenient laboratory testing, and so many spurious coupling problems may not be identified until late in the program; e.g., during acceptance testing or during early operational usage of the system hardware.

Analytical methods are available for use during design stages to help reduce the likelihood of serious system compatibility problems. These methods when judiciously applied can result in sizeable net savings of funds through the reduction of costly redesign and modification efforts. Moreover, their application can help avoid the compromise of system effectiveness or even personnel safety.

3.2.10.1.1 Spurious Coupling in Flat-Cable Systems. Spurious coupling in cabling is due to the sharing of a common conductor impedance, capacitive coupling between conductors, and magnetic coupling between conductors. It is also possible for the conductor to react to the electric and magnetic fields producing capacitive coupling and magnetic coupling to objects other than adjacent conductors as follows:

a. Common impedance coupling - When a conductor is shared by several circuits, the flow of current from one circuit through this common impedance produces a voltage drop that may affect the operation of other circuits sharing the same conductor. If the common impedance is a signal return for several similar channels, interchannel crosstalk will occur.

b. Electric coupling - When two conductors are in close proximity to each other, a small but definite capacitor is formed. The conductors are capacitor electrodes, and the cable insulation and any other nonconductive materials form the capacitor dielectric. Any varying voltage or one of the conductors is capacitively coupled into the impedances connected to the other conductor.

c. Magnetic coupling - The presence of a longitudinal conductor produces a small but definite inductance. When two conductors are in close proximity to each other, a transformer is produced because of the mutual inductance that exists between conductors. Any varying current flow in one conductor is magnetically coupled into the other conductor as induced voltage because of this transformer action.

d. Radiated susceptibility - A conductor immersed in an electric field acts like the output electrode of a capacitor, supplying energy to the terminating impedances at each end of the conductor. A conductor immersed in a magnetic field acts like the secondary winding of a transformer, supplying energy to the terminating impedances at each end of the conductor.

e. Radiated interference - When a potential is applied to a conductor, the conductor acts like the input electrode of a capacitor, creating an electric field that links the conductor to surrounding objects. When a current flows through a conductor, the conductor acts like the primary winding of a transformer, creating a magnetic field that links the conductor to surrounding objects.

3.2.10.1.2 Allowable Signal-to-Noise Ratio. The amplitude of the desired signal being transmitted by the conductor is a function of the terminating hardware design, and may vary between the extreme limits of a few hundreds of nanovolts to a few hundreds of volts, but will generally be between a few millivolts and a few tens of volts. Higher voltages require special cable and connector dielectric materials and configurations, while lower voltages require special cable and connector-shielding materials and configurations.

The transmitted signal resolution, accuracy, or signal-to-noise ratio determine what fraction of the total signal amplitude may be noise without significantly affecting system performance. The amplitude of the desired signal, and the fractional part of this amplitude which may be noise, determines the maximum allowable amplitude of noise in the susceptible circuit.

The total allowable amplitude of noise must be divided between the interference under discussion and the other sources of noise in the signal channel. If the sources of interference within the system and system environment have frequency spectrums outside the normal operational bandwidth of the susceptible circuit, the potentially desirable effects of susceptible circuit frequency discrimination in rejecting the undesired interference should be determined.

3.2.10.1.3 Required Decoupling Ratio Between Interference Source and Susceptible Circuit. The amplitude of the interfering signals being transmitted by adjacent conductors is a function of the terminating hardware design, and may vary between the extreme limits of a few milliamperes or millivolts and a few amperes or hundreds of volts, but will generally be between a few hundreds of milliamperes or millivolts and a few amperes or volts. Higher amplitudes often require special harness assemblies that would normally be routed separately, while lower amplitudes would not represent a significant hazard to other circuits and would often be classified as susceptible circuits. The amplitude of radiated interference is a function of the specific system environment, but generally has an equivalent conducted interference amplitude lower than the other conducted interference amplitudes found in the same system, except where high-power transmitters are involved. The amplitude which may appear as noise in the susceptible circuit without significantly affecting system performance determines the required decoupling ratio between the interference source and the susceptible circuit.

3.2.10.1.4 Spurious-Coupling Ratio of Interconnection Wiring. Any coupling circuit consists of a source, one or more transmission paths, and a load. Spurious coupling paths consist of common conductor impedances, mutual conductor capacitances, mutual conductor inductances, radiated electric fields, and radiated magnetic fields. Spurious signals from the sources of interference are coupled through these distributed transmission paths into the susceptible circuits. The amplitude of interference coupled into the susceptible circuit is determined by the amplitude of the source of interference and the ratio of interference division between the transmission path and the susceptible circuit terminating impedances.

3.2.10.1.4.1 Cable Capacitance. Cable capacitances for typical (0.075 C/L, 4-mil-thick Mylar) unshielded standard-density and high-density flat cable are given below:

| Cable Type (Typical) | Cable Configuration | Mutual Capacitance PF/m |
|---|---|---|
| Standard-Density Unshielded | Side-by-side, same layer | 21.5 |
| | Over-and-under, tightly stacked multiple layer | 165.0 |
| High-Density Unshielded | Side-by-side, same layer | 40.0 |
| | Over-and-under, tightly stacked multiple layers | 255.0 |

Specific values of mutual capacitance for many standard flat cable configurations, including several common conductor wiring connections for each cable configuration, are given in Paragraph 3.2.3.1.4.

Edge-to-edge capacitance is primarily a function of the dielectric thickness between conductors and the dielectric constant. Since the dielectric constant for commonly used insulating materials does not change drastically, and the dielectric thickness is determined primarily by the selection of either standard-density or high-density cable, only two values of edge-to-edge capacitance are required for most preliminary designs.

3.2.10.1.4.2 Cable Inductance. Cable inductance for typical unshielded standard and high-density flat cables running in a straight line is tabulated below:

| Cable Type (Typical) | Cable Configuration | Mutual Inductance ($\mu$H/m) | Self Inductance ($\mu$H/m) | Coupling Coefficient |
|---|---|---|---|---|
| Standard-Density Unshielded | Side-by-side (normal cable lengths) | 1.65 | 2.1 | 0.79 |
| | Over-and-under (normal cable lengths) | 1.95 | 2.1 | 0.93 |
| High-Density Unshielded | Side-by-side (normal cable lengths) | 1.66 | 2.0 | 0.83 |
| | Over-and-under (normal cable lengths) | 1.90 | 2.0 | 0.95 |

Typical copper and other nonferrous shielded cables should have inductance characteristics identical to unshielded cables, if the shield material is actually nonferrous. Inductance measurements on certain copper-shielded cables have indicated effective inductances approximately eight times greater than expected. The addition of copper foils, of known purity, to unshielded flat cable confirmed that the increased inductance was because of ferrous impurities in the copper shield foils and/or conductors.

Typical ferrous-shielded flat cables should have inductance values equal to the inductance values of unshielded flat cables, multiplied by the effective permeability of the ferrous shield. Since the ferrous shield has an extremely high permeability and an extremely thin-shield foil with a relatively small cross-sectional area, magnetic saturation, with a consequent reduction of shielding effectiveness and inductance at combined conductor currents in the order of tens of milliamperes, is a distinct possibility. Space isolation from nearby high-current conductors is necessary.

The high self-inductance of ferrous-shielded flat cable should reduce the bandwidth of the flat-cable transmission system significantly. The high mutual-inductance of conductors within ferrous-shielded flat cable should significantly reduce the frequency above which crosstalk becomes objectionable.

3.2.10.1.4.3 Cable-Shield Discontinuities at Terminations. When the shield is removed from a section of shielded flat cable to permit the installation of an unshielded connector or provide access to the conductors for other termination hardware, the stripped section is, in effect, unshielded flat cable and exhibits the electrical characteristics of the equivalent unshielded flat cable. Since the mutual capacitance and inductance (ferrous shielded only at low frequencies) of shielded flat cable are significantly lower than the equivalent unshielded flat cable, the insertion of even a short length of unshielded flat cable in a shielded transmission system reduced significantly the transmission loss through the spurious coupling path between conductors.

3.2.10.1.5 Terminating Impedances. Transmission lines are terminated in a source impedance, supplying the transmitted function, at one end and a load impedance, utilizing the transmitted function, at the other end. Complex transmission systems may interconnect multiple sources and loads. The magnitudes of complex termination impedances are frequency dépendent.

3.2.10.1.5.1 Function-Frequency and Time-Variable Characteristics. When the amplitude of the function exhibits a variation with time, this characteristic is defined in terms of either frequency or rate of change. Analog waveforms, exhibiting relatively smooth and continuous cyclic variations, are defined in terms of frequency and waveform. Digital and pulse waveforms, exhibiting rapid and discontinuous variations in instantaneous amplitude that are either cyclic, randomly repetitive, or nonrepetitive, are defined in terms of rates of change, waveform shape, and repetition frequency. These nonsinusoidal waveforms contain energy distributed over a broad frequency spectrum. The exact frequency versus amplitude distribution is dependent on the specific digital or pulse waveform and requires appropriate analysis for conversion from the time domain to the frequency domain. Where hardware is available, appropriate spectrum analyzers may be utilized to measure these functions directly in terms of frequency versus amplitude.

3.2.10.1.5.2 Frequency Dependence of Reactive Components. The resistive component of a termination is not frequency dependent; however, virtually all terminations contain reactive components, either lumped constant or distributed, that have a significant magnitude at some frequency of interest. The values of lumped-constant capacitances are readily obtained or measured. The reactance of a capacitance can be calculated or obtained from a reactance chart for the frequency of interest, and will exhibit a magnitude inversely proportional to frequency. The values of lumped-constant inductances are readily obtained or measured. The reactance of an inductance can be calculated, or obtained from a reactance chart for the frequency of interest, and will exhibit a magnitude proportional to frequency.

3.2.10.1.5.3 Transformers. Transformers are merely multiple inductors wound on a common core. If the inductance values are not specified, it is possible to measure the inductances of the windings if hardware is available, or to estimate the inductances if the transformer design details are available. With a knowledge of the transformer-winding inductances and interfacing circuit parameters, it becomes possible to calculate the reflected impedance presented by the connected transformer winding. The values of connected load resistances, frequency-dependent reactances, winding transformation ratios, and copper and core losses must be considered.

3.2.10.1.5.4 Distributed Parameters. Distributed capacitances and inductances frequently have significant reactances at surprisingly low frequencies when associated with either extremely high or low circuit impedances, or lumped constants having large reactive values of the opposite sign. Unfortunately, no comprehensive compilation of distributed parameter values for typical hardware configurations is in existence at this time. Therefore, the most practical approach to determining distributed parameter values is an estimate by personnel experienced in high-frequency design, radio-frequency interference, or electromagnetic compatibility.

3.2.10.1.5.4.1 Capacitors. The distributed series inductance of capacitors can be measured by placing a noninductive resistor with a value several times greater than the magnitude of the capacitive reactance at the lowest frequency of interest, in series with the capacitor, to form an essentially constant-current circuit. The output of a variable-frequency signal source is applied across this series circuit and maintained at a constant amplitude. The output voltage of the signal source will divide between the resistor and capacitor in proportion to the ratio of capacitive reactance to resistance. Since the noninductive resistor is not frequency-dependent, the voltage measured across the capacitor, as the frequency of the signal source is varied, provides an indication of capacitor impedance versus frequency.

At relatively low frequencies, where the magnitude of the series inductive reactance is small compared with the magnitude of the series capacitive reactance, the voltage variation across the capacitor will be inversely proportional to frequency. At relatively high frequencies, where the magnitude of the series inductive reactance becomes large compared with the magnitude of the series capacitive reactance, the voltage variation across the capacitor will be proportional to frequency. At some intermediate frequency, the two curves with opposing slopes cross, indicating a series resonance in the form of a broad negative peak.

The value of series impedance measured at and above this fundamental series resonant frequency cannot be reduced by increasing the value of capacitance used, because the series inductive reactance tends to increase with capacitance in a given capacitor configuration, lowering the fundamental series resonance, and dominating an even larger portion of the frequency spectrum. As the signal-source frequency is increased above the fundamental series resonant frequency, the capacitor impedance will alternate between a series of low-impedance series resonances and high-impedance parallel resonances that are a function of the capacitor configuration.

The capacitor fails to function effectively in this portion of the frequency spectrum and is ordinarily replaced or supplemented with an improved capacitor configuration or supplemented with external inductance to form a multiple-element filter. Some capacitor manufacturers have curves and other information describing the impedance of specific capacitor values and configurations versus frequency. Since the distributed parameters of capacitors are not ordinarily specified, large, uncontrolled variations in these values may be expected.

Conventional foil-wound capacitors, having terminations of the inserted tab type, have the lowest fundamental resonant frequencies. A very high-value, foil-wound, tantalytic capacitor might be resonant at audio frequencies. The same basic capacitors, having terminations of the extended foil type, have significantly higher fundamental resonant frequencies, but high-frequency performance is still severely limited by associated wiring. This same capacitor in a feedthru configuration has a much higher fundamental resonant frequency, and performance is essentially independent of associated wiring but imposes mounting limitations. The feedthru capacitor is a three-terminal device, similar to a filter, that is oridnarily mounted directly on the ground-plane structure to function properly, usually in a bulkhead type mount through a metallic panel to provide the improved input-output isolation necessary to take full advantage of the superior performance available from this configuration.

3.2.10.1.5.4.2 Inductors. The distributed shunt capacitance of inductors can be measured using the same techniques described for capacitors if the resistor has a value several times greater than the magnitude of the inductive reactant at the highest frequency of interest at relatively low frequency where the magnitude of the shunt capacitive reactance is relatively high compared with the magnitude of inductive reactance, the shunt capacitive current flow will be negligible and the voltage across the inductor will be proportional to frequency. At relatively high frequencies, where the magnitude of the shunt capacitive reactance becomes low compared with the magnitude of the inductive reactance, the shunt capacitive current flow will dominate the inductive current flow and the voltage variation across the inductor will be inversely proportional to frequency. At some intermediate frequency, the two curves with opposing slopes cross, indicating parallel resonance in the form of a broad peak.

Above this fundamental parallel resonance, the inductor impedance will alternate between a series of high-impedance parallel resonances and low-impedance series resonances that are a function of inductor configuration. The inductor fails to function effectively in this portion of the frequency spectrum and is ordinarily replaced or supplemented with an improved inductor configuration or supplemented with external capacitance to form a multiple element filter. Some inductor manufacturers supply shunt capacitance values or impedance/reactance curves and other information describing the impedance of specific inductor values and configurations versus frequency. Since the distributed parameters of inductors are not ordinarily specified, large uncontrolled variations in these values may be expected.

3.2.10.1.5.5 Impedance. The vector sum of the resistance, inductive reactance, and capacitive reactance provides the net low-frequency complex impedance of the terminating circuit but does not adequately describe the high-frequency response of the termination near and above the fundamental resonances produced by distributed parameters. The equivalent high-frequency circuit configuration, including all distributed parameters, is required for a meaningful circuit analysis.

3.2.10.1.5.6 Nonlinear Devices. The presence of nonlinear devices may result in several unusual effects. The impedance, both resistive and reactive, of circuitry associated with nonlinear devices becomes amplitude sensitive. Nonlinear devices produce distortion products not present in the original function. When severe nonlinearity occurs, alternating currents are rectified producing direct-current offset voltages and low-frequency modulation products that alter the characteristics of the original function.

3.2.10.1.5.7 Frequency Discrimination in Terminating Circuitry. The amplitude of interference coupled through a terminating circuit is determined by the amplitude of the source of interference at the input of the terminating device and the ratio of interference division between components of the function transmission path internal to the terminating device. When these components are reactive, the transmission path internal to the terminating device becomes frequency selective, often providing useful discrimination that supplements the shielding and transmission characteristics of the external interconnection cabling and intentional filtering.

3.2.10.2 Conductor Zoning Fundamentals

3.2.10.2.1 Introduction to Electrical Zoning of System Wiring. A major source of electromagnetic incompatibilities in any electrical/electronic system is the system interconnection wiring. Paragraph 3.2.10.1 explains spurious-coupling mechanisms, describes a method of determining the magnitude of extraneous energy tolerated by a circuit without impairment of function, and the degree of isolation required between circuits. A method is described for determining the actual degree of isolation provided by specific cable and conductor configurations transmitting specific interfering and susceptible functions, and terminated in a specific manner.

Comparing the degree of isolation required with the degree of isolation actually existing provides an indication of the safety margin either available or required for satisfactory system operation. This section describes a method of design that reduces the effort required to achieve an initial wiring configuration with optimum conductor-to-conductor isolation, and a high probability of obtaining the required electromagnetic safety margin. The application of this method of wiring design also ensures that differences between the characteristics of FCC and conventional RWC are adequately considered by personnel with minimal FCC experience.

3.2.10.2.2 Conductor Relationships. Every conductor is related electromagnetically to every other conductor in the wiring system. Some functions are natural sources of interference. Other functions are generally susceptible to interference. It is quite possible for one group of functions to be susceptible to interference from another group of functions, and simultaneously be a source of interference to a third group of functions.

Experience has shown that certain levels of interference and susceptibility are typical of the average electrical/electronic system, and that these levels can be extrapolated for use in more specialized types of systems. Because of the extremely broad spectrum of hardware involved, these levels are subject to wide variations and are utilized only as a preliminary design goal until more specific hardware design and test information become available. As confidence in system performance increases, arbitrarily specified, initial levels should be superseded or refined with analyses of actual system performance.

This section provides gross methods for the initial evaluation of wiring performance and a general approach for the refinement of this information as more specific information becomes available. The user must modify the assigned units of coupling and generalized approaches for improved compatibility with the particular system involved, as experience is accumulated, so that useful degrees of performance-prediction accuracy will be achieved.

3.2.10.2.3 Spurious-Coupling Circuit Configuration. A schematic of the interference circuit, the susceptible circuit, and the spurious-coupling circuit must be generated. The interference circuit should show the electrical configuration of the interference-carrying conductor and the associated return conductor, or other path, including any shield terminations. Conductor-termination circuitry at both ends should be shown in detail up to the input or output element of the first isolating device, such as a transistor. The susceptible circuit should be shown in the same detail. When large, reactive terminations and/or radio frequencies are involved, the distributed parameters of the terminating circuits must be considered.

The lengths of conductors, conductor breakouts, and shield terminations must be detailed. When conductors follow separate routes for part of the total path length, the common and separate path lengths must be dimensioned. Reasonable approximations are usually satisfactory. The cable assembly cross-sectional geometry and composite dielectric constant or generic materials description will be required so that the cable-distributed parameters can be either taken from the prepared tables or calculated. When interaction between conductors in separate strips of the cable assembly is being evaluated, the cable-stack geometry and the effect of air gaps, dielectric space strips, metallic spacer strips, and intervening shielded cables over all or part of the transmission path length must be considered.

The resultant spurious-coupling circuit schematic will resemble one of the basic sketches shown in Figures 3-91 through 3-96. The evolutionary process, from distributed parameters to equivalent lumped constant values, is described in Paragraph 3.2.3.2.2. This selection of electric- or magnetic-field configurations is determined primarily by the relationship existing between the conductor characteristic impedance and the termination impedances of the interference-generating and susceptible circuits. The conductor characteristic impedance can be obtained from Paragraph 3.2.3.1.

3.2.10.2.4 Termination Impedance Dependence of Spurious-Coupling Modes. Electric (capacitive) coupling is a voltage-related phenomenon that is independent of current flow. When the terminating impedances of a transmission line are high compared with the characteristic impedance of that particular conductor configuration, a predominantly electric field is generated or sensed. Magnetic (inductive) coupling is a current-related phenomenon that is independent of voltage level. When the terminating impedance of a transmission line is low compared with the characteristic impedance of that particular conductor configuration, a predominantly magnetic field is generated or sensed.

Both electric- and magnetic-field coupling occur when the terminating impedances of a transmission line are equal to the characteristic impedance of that particular conductor configuration; however, the magnitude of the total electromagnetic coupling is minimized. Additional advantages of a matched-impedance transmission system are highly efficient power transfer, minimum waveform distortion, insensitivity to frequency, insensitivity to cable length, resistive reflected impedances. minimum circuit loading, and a high degree of performance predictability.

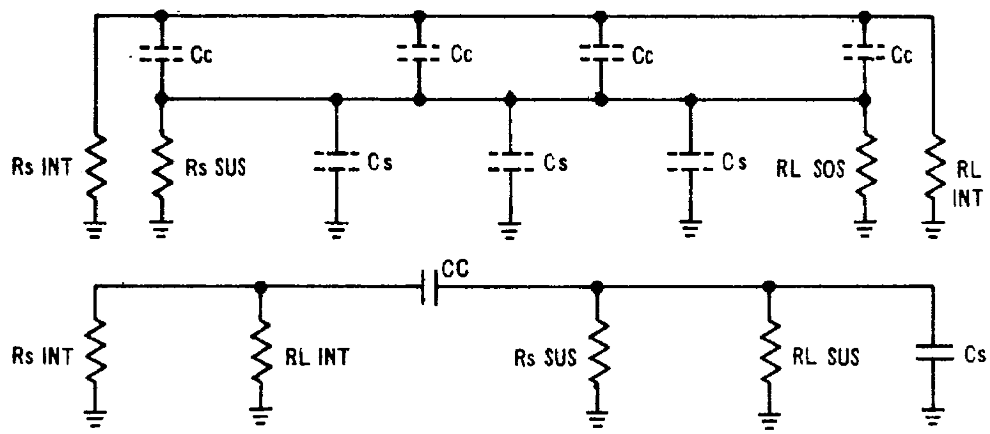



FIGURE 3-91. Electrical coupling - unshielded.

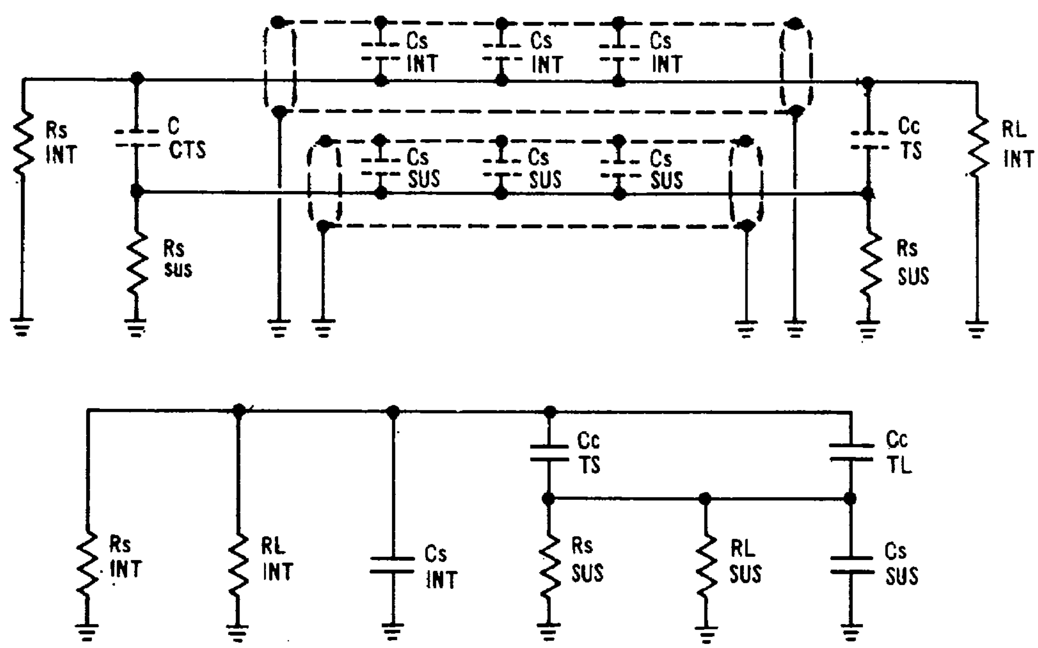



FIGURE 3-92. Electrical coupling - shielded.

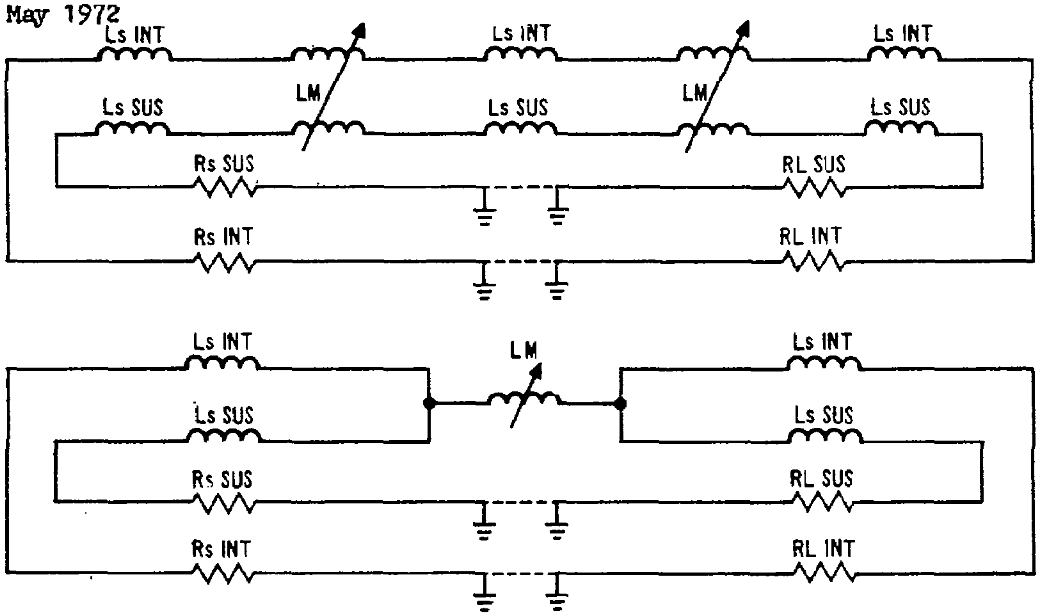



FIGURE 3-93. Magnetic coupling - unshielded.

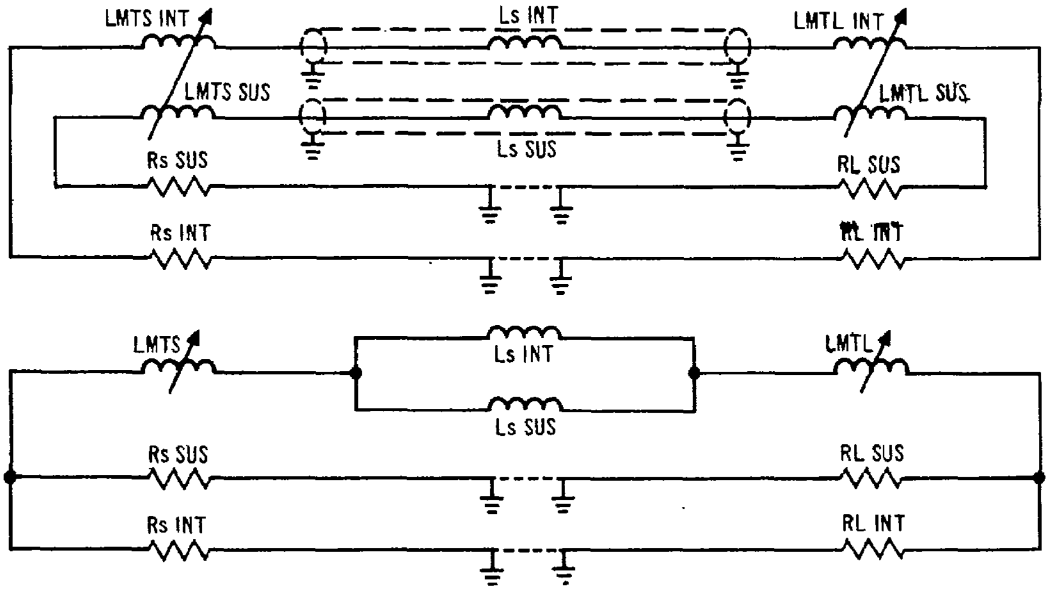



FIGURE 3-94. Magnetic coupling - shielded.

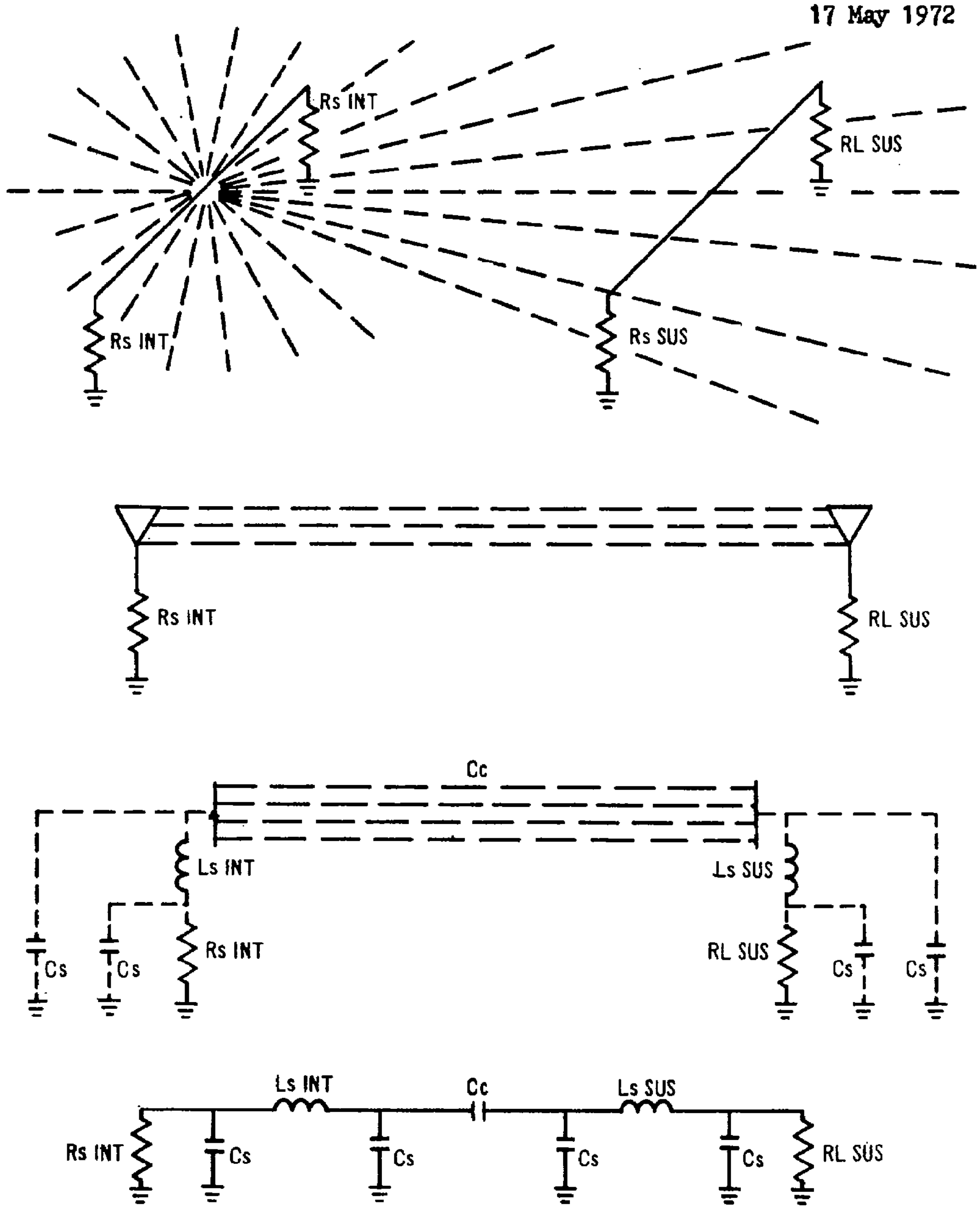



FIGURE 3-95. Radiated field coupling - electric field.

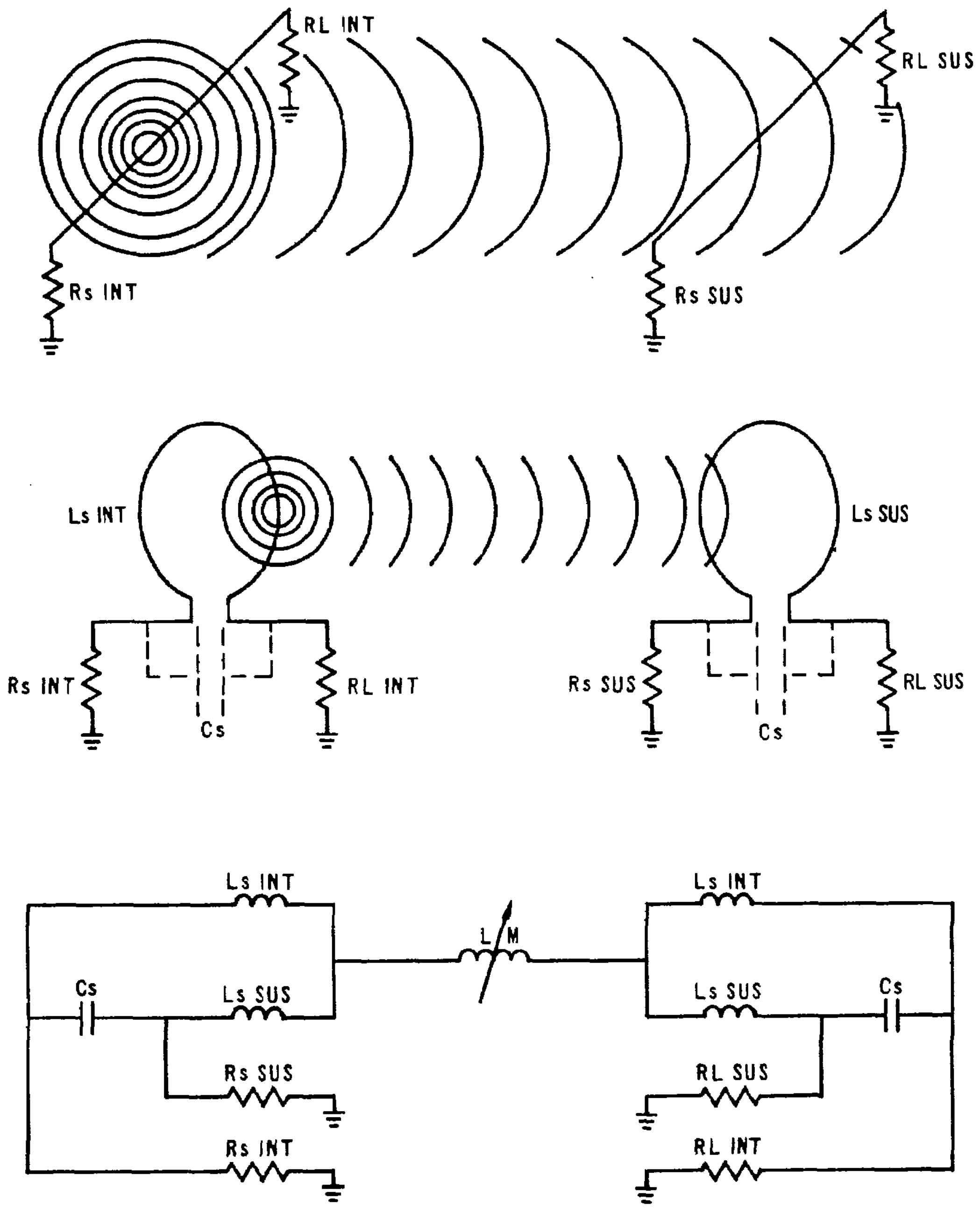



FIGURE 3-96. Radiated field coupling — magnetic field.

Both electric- and magnetic-field coupling also occur when the source and load terminating impedances have opposite magnitudes and differ greatly from the characteristic impedance of that particular conductor configuration. The magnitude of the total electromagnetic coupling is minimized when the geometric mean of the terminating impedances is equal to the characteristic impedance of that particular conductor configuration and when the terminating impedances differ from the characteristic impedance by the greatest possible magnitude. The low-impedance termination reduces the capacitively coupled voltage available while the high impedance termination limits the inductively coupled current available.

When the use of an impedance-mismatched transmission system is inevitable, this configuration minimizes spurious coupling but has the serious disadvantages of being the ultimate in inefficient power transfer, maximizing waveform distortion, extreme sensitivity to frequency, extreme sensitivity to cable length, reactive reflected impedances, reactive circuit loading, and poor performance predictability at medium and high frequencies. Low-frequency, low-level signal-type functions are not ordinarily affected by these limitations.

Minimum coupling will also occur between conductors when one conductor is terminated in low impedances and the other conductor is terminated in high impedances because the dominant generated field and dominant sensed field are incompatible and an inefficient transfer of spuriously coupled power occurs.

3.2.10.2.5 Determining Termination Impedances and Sensitivity to Spurious Coupling. Typical FCC configurations have characteristic impedances of a few tens of ohms for unshielded over-and-under or single conductors with nonferrous shields, to a few hundreds of ohms for unshielded, side-by-side or single conductors with structural returns. Ferrous shields increase these characteristic impedances from 10 times for low-grade steels to 100 times for high-permeability alloys that are not magnetically saturated.

When the termination is reactive instead of resistive, the equivalent reactance at the interference frequency is substituted for the resistance value. The values of capacitance and inductance, having similar reactive magnitudes at a variety of common power frequencies, is shown for each range of resistance/reactance levels.

If the shunt reactance is higher than the terminating resistance, use the resistance value. If the shunt reactance is lower than the terminating resistance, use the reactance value. If the series reactance is lower than the terminating resistance, use the resistance value. If the series reactance is higher than the terminating resistance, use the reactance value.

A high-impedance parallel resonance will occur at a frequency where the reactances of parallel-connected capacitances and inductances are equal. A low-impedance series resonance will occur at a frequency where the reactances of series-connected capacitances and inductances are equal. The increase or degrease in impedance, relative to the termination resistance, is a function of the resonant circuit "Q." Since the capacitor Q is normally high compared with the inductor Q, the circuit Q is limited by the inductor Q. Low-frequency inductors normally have a Q of less than 10. High-frequency inductors normally have a Q of between 10 and 100.

When the two ends of the conductor are terminated in widely differing values of impedance, the electric coupling will be determined by the two terminating impedances in parallel, and the magnetic coupling will be determined by the two terminating impedances in series. The geometric mean of the two terminating impedances will indicate which coupling mode is dominant and the extent to which that mode is dominant.

3.2.10.2.6 Frequency Dependence Spurious Coupling. The following tabulation indicates the degree of magnetic or electric coupling (on a logarithmic scale) that occurs at various frequencies for a given coupling circuit configuration:

| Units of Coupling | Frequency | Equivalent Pulse Width |
|---|---|---|
| 0 | Under 3 Hz | Over 100 ms |
| 1 | 3 to 30 Hz | 10 to 100 ms |
| 2 | 30 to 300 Hz | 1 to 10 ms |
| 3 | 300 Hz to 3 kHz | 100 $\mu$s to 1 ms |
| 4 | 3 to 30 kHz | 10 to 100 $\mu$s |
| 5 | 30 to 300 kHz | 1 to 10 $\mu$s |
| 6 | 300 kHz to 3 MHz | 100 ns to 1 $\mu$s |
| 7 | 3 to 30 MHz | 10 to 100 ns |
| 8 | 30 to 300 MHz | 1 to 10 ns |
| 9 | Above 300 MHz | Under 1 ns |

Basically, the frequency factor is a multiplier for the other variables involved. Frequencies and pulse widths in category zero a negligible spurious-coupling problem. Circuits with one unit or less of coupling rarely present a problem. Two units of coupling provide a problem only when extremely sensitive hardware is involved.

Three units of coupling may be a problem where sensitive hardware is used. Four units present a serious, spurious-coupling problem and may require special wiring practices. Five units will produce near-unity coupling in typical unshielded conductors, special wiring practices are required, and wiring resonances will occur where moderately long cable lengths of the order of 300 feet are involved. For six units of coupling, the same conditions apply except that resonance will occur in shorter cable lengths of the order of 30 feet, and for seven units of coupling, resonance occurs with cables of the order of 3 feet in length. Circuits with eight or nine units produce near-unity coupling in unshielded conductors, require special wiring practices, and will exhibit wiring resonances for all cable lengths.

It is essential that the transmission losses of shielded cable, regardless of shielding material and any unshielded conductors containing or coated with ferrous materials, be evaluated because both the undesired, spuriously coupled energy and the desired function may be attenuated significantly, even in the audio- and power-frequency spectrums. If unshielded cable is in intimate contact with a metallic structure, adjacent-shielded cables or adjacent-metallic spacer strips, a pseudosingle- or double-shielded flat cable is created, and the resulting transmission losses must be evaluated.

3.2.10.2.7 Length Dependence of Spurious Coupling. The degree of magnetic or electric coupling (on a logarithmic scale) that occurs at various lengths of unshielded conductors for a given coupling circuit configuration is tabulated below:

| Units of Spurious Coupling | Unshielded Length of Conductors |
|---|---|
| 0 | Under 1 mm |
| 1 | 1 to 10 mm |
| 2 | 10 to 100 mm |
| 3 | 100 mm to 1 m |
| 4 | 1 to 10 m |
| 5 | 10 to 100 m |
| 6 | 100 m to 1 km |
| 7 | 1 to 10 km |
| 8 | 10 to 100 km |
| 9 | Over 100 km |

Basically, the factor for the unshielded length of conductors is a multiplier for the other variables involved. Conductors that are shielded, or otherwise isolated from each other, have some length of unshielded conductors in the terminating hardware at distributors, junction boxes, and connector interfaces. The accumulated total of these shield breakouts is equivalent to an unshielded cable of the same length. When this accumulated length is less than the total circuit length divided by the shielding effectiveness ratio, relative to unshielded cable, the spurious coupling becomes dependent on the shielding effectiveness ratio instead of the accumulated total unshielded lengths of conductors. This condition will normally occur only in the extremely long cables associated with large land-based installations and large-wired communications systems unless sophisticated shielding techniques are utilized.

3.2.10.2.8 Amplitude Dependence of Electromagnetic Compatibility (Time-Varying Component). The degree of susceptibility or interference (on a logarithmic scale) associated with a conductor carrying various voltages and currents is tabulated below:

| Units of Susceptibility | Voltage | Current |
|---|---|---|
| S4 | Under 10 $\mu$V | Under 0.1 $\mu$A |
| S3 | 10 to 100 $\mu$V | 0.1 to 1 $\mu$A |
| S2 | 100 $\mu$V to 1 mV | 1 to 10 $\mu$A |
| S1 | 1 to 10 mV | 10 to 100 $\mu$A |
| S0 | 10 to 100 mV | 100 $\mu$A to 1 mA |

| Units of Interference | Voltage | Current |
|---|---|---|
| I0 | 100 mV to 1 V | 1 to 10 mA |
| I1 | 1 to 10 V | 10 to 100 mA |
| I2 | 10 to 100 V | 100 mA to 1 A |
| I3 | 100 V to 1 kV | 1 to 10 A |
| I4 | Above 1 kV | Above 10 A |

Experience has shown that certain levels of interference and susceptibility are typical of the average electrical/electronic system, and that these levels can be extrapolated for use in more specialized types of system. Because of the extremely broad spectrum of hardware involved, these levels are subject to wide variations and should only be utilized as a preliminary design goal until more specific hardware design and test information become available. As confidence in system performance increases, arbitrarily specified initial levels should be refined with the aid of measurement and analyses of actual system performance.

3.2.10.2.9 Resolution/Accuracy Dependence of Electromagnetic Compatibility. The degree of susceptibility or interference (on a logarithmic scale) associated with a conductor function having various resolution and/or accuracy requirements is tabulated below:

| Units of Susceptibility | Resolution or Accuracy Required (%) |
|---|---|
| S4 | Better than 0.1 |
| S3 | 0.1 to 1 |
| S2 | 1 to 10 |
| S1 | 10 to 100 |
| S0 | On-off or other bilevel functions |
| **Units of Interference** | |
| I2 | Average value for interference |

Basically, the resolution/accuracy factor is a divider for the susceptible-conductor-function amplitude to provide a threshold amplitude of undesired signal above which the system will malfunction or exhibit an unacceptable error. The selection of a resolution/accuracy value for a component or subsystem must provide for both the division of the allowable tolerance between the components of the system chain and a division of tolerances between spurious coupling and other sources of error.

3.2.10.2.10 Functional Dependence of Electromagnetic Compatibility. Wiring-design personnel should consider the secondary characteristics of the function being analyzed when electrical wire zoning is in process. These secondary characteristics often dominate the primary characteristics of the function during an evaluation of system electromagnetic compatibility.

Direct-current power wiring would not be a source of interference if a truly pure source of direct current were available. Experience shows that many practical direct-current power distribution systems used to supply electrical/electronic hardware are modulated by a complex noise waveform having a peak-to-peak amplitude equal to 3 percent of the power supply voltage, and may have long-duration transient peaks equal to 30 percent of the power supply voltage when large electrical loads are applied or removed intermittently.

The complex noise waveform consists of both random noise and repetitive waveform created by the flow of instantaneously variable current demands through the common impedance of the power distribution wiring in response to functions generated in the connected hardware. The long-duration transients occur when high current demands on the power distribution system are made by large electrical devices and secondary power conversion hardware for these devices. The durations of these instantaneous current demands exceed the time constants of energy storage elements in practical passive-interference filters.

Based on the primary functional characteristics of a direct-current power distribution system, significant interference levels would not be anticipated. A close look at the secondary characteristics of this same direct-current power distribution system reveals the existence of a very significant interference source. Many other electrical/electronic functions have similarly significant secondary characteristics.

A repetitive pulse might appear harmless if the repetition rate is evaluated in sine-wave terms. An examination of the pulse rate-of-rise and rate-of-fall, or of pulse width, would reveal a much higher equivalent sine-wave frequency with a proportionate increase in coupling.

A nonrepetitive pulse, because of the actuation of a switching function, might appear significant on the basis of waveform or rate-of-rise and rate-of-fall when evaluated in repetitive pulse terms. An examination of the energy content of the single pulse and the time of occurrence relative to other functions in the system might well indicate that this same nonrepetitive pulse is harmless.

A pure sine-wave is normally considered the least offensive waveform, because the energy content is restricted to a single frequency and harmonics are not present. At low to mid-audio frequencies, spurious coupling of sine-wave energy is normally not a problem in typical interconnection wiring. Above the midaudio frequency range, even sine-wave energy may produce serious spurious-coupling problems in typical cabling. In the supersonic and radio-frequency spectrums, near-unity coupling often occurs. Even at low frequencies, a small amount of sine-wave energy, offset slightly from the frequency of an alternating current power or reference source, might appear as a false error signal in a servo system.

Because of the extremely broad spectrum of hardware involved, specific recommendations for the evaluation of conductor function cannot be made. The users must become thoroughly familiar with the system theory of operation and with hardware design details so that they will be aware of secondary considerations that may affect the functional separation of conductors. The previous examples are intended to point out the inconsistencies that may occur if the primary function is taken at face value without an adequate evaluation of secondary considerations.

3.2.10.2.11 Frequency Dependence of Conductor Transmission Losses. The series inductance and shunt capacitance of a conductor form a distributed low-pass filter that limits the bandwidth of the conductor. Reactive terminations will interact with the distributed parameters of the conductor and alter the inherent cable bandwidth. A match-impedance transmission system, with terminating impedances equal to the conductor characteristic impedance, will have a -3-decibel (50-percent power loss or 29-percent voltage or current loss) upper frequency limit approximately equal to the frequency at which the reactance of the conductor series inductance becomes equal to the reactance of the conductor shunt capacitance. Above this frequency, the transmitted function will exhibit a transmission loss of 12 decibels per octave (94-percent power loss or 75-percent voltage or current loss as frequency is doubled) or 40 decibels per decade (10,000 times power loss or 100 times voltage or current loss as frequency is increased 10 times).

A typical, matched-impedance, nonferrous, shielded transmission line would have a -3-decibel bandwidth of approximately 5 megahertz for a 1-meter length, approximately 500 kilohertz for a 10-meter length, and approximately 50 kilohertz for a 100-meter length. The -3-decibel frequency is inversely proportional to cable length.

A ferrous-shield material would lower these -3-decibel frequencies by an additional factor of approximately 100 for high-permeability alloys such as 4-79 Permalloy or a factor of approximately 10 for commercial steels if the shield is magnetically unsaturated.

3.2.10.2.12 Termination Impedance Dependence of Conductor Transmission Losses. The series inductance and shunt capacitance of a conductor form a distributed low-pass filter that limits the bandwidth of the conductor. When a matched impedance transmission line condition exists, with the terminating impedances equal to the characteristic impedance of the conductor, the reactances of the conductor series inductance and shunt capacitance vary symmetrically about the terminating impedance value, as frequency changes, and make equal contributions to the transmission losses experienced by the transmitted functions.

When the transmission line is mismatched, one of the reactive elements in this distributed low-pass filter becomes relatively ineffective as a frequency discriminating device; therefore, the -3-decibel (50-percent power loss or 29-percent voltage or current loss) upper frequency limit is approximately equal to the frequency at which the reactance of the effective reactive element becomes equal to the composite terminating impedance. Above this frequency, the transmitted function will exhibit a transmission loss of 6 decibels per octave (75-percent power loss or 50-percent current or voltage loss as frequency is doubled) or 20 decibels per decade (100 times power loss or 10 times voltage or current loss as frequency is increased 10 times).

In the electrostatic-coupling mode, where the paralleled terminating impedances are high compared with the characteristic impedance of the conductor, the frequency at which the series inductive reactance exceeds the paralleled terminating impedances is considerably higher than the frequency at which the shunt-capacitive reactance falls below the paralleled terminating impedances. Therefore, the shunt capacitance dominates the high-frequency transmission losses. The conductor is relatively useless as a transmission line at the still higher frequencies where the series inductive reactance becomes significant.

In the magnetic-coupling mode, where the seriesed terminating impedances are low compared with the characteristic impedance of the conductor, the frequency at which the shunt-capacitive reactance falls below the terminating impedances is considerably higher than the frequency at which the series-inductive reactance exceeds the seriesed terminating impedances. Therefore, the series inductance dominates the high-frequency transmission losses. The conductor is relatively useless as a transmission line at the frequencies where the shunt-capacitive reactance becomes significant.

A deliberately impedance-mismatched transmission system, in which one terminating impedance is selected lower than the conductor characteristic impedance, and the other terminating impedance is selected higher than the conductor characteristic impedance by similar ratios, has the conductor characteristic impedance at the geometric mean of the terminating impedances. The bandwidth characteristics will be determined by the series-inductive reactance and shunt-capacitive reactance of the conductor as elements of a distributed low-pass filter.

The bandwidth of this transmission system will be similar to that of a matched-impedance transmission system; however, this system has all the disadvantages of mismatched transmission systems, including inefficient power transfer, waveform distortion, sensitivity to frequency, sensitivity to cable length, reactive reflected impedances, reactive circuit loading, and poor performance predictability at medium and high frequencies.

3.2.10.2.13 Analyzing the Electromagnetic Compatibility of Two Conductors. The following steps define the method for analyzing the electronic compatibility of two conductors:

a. Determine the allowable level of undesired energy that may be spuriously coupled into the susceptible circuit without causing the system to malfunction or produce an unacceptable error. Error analyses made by the cognizant design engineer during mandatory design analyses performed in accordance with the requirements of good engineering practice and the system specification will provide the necessary basic information for the derivation of the allowable susceptibility level. The allowable susceptibility level will be some fractional part of the total allowable malfunction or unacceptable error level, because spuriously coupled, undesired energy is only one of several noise sources that may be present simultaneously in the system.

   The allowable susceptibility level may vary with frequency if the susceptible circuit provides frequency discrimination in the spectrum of interest. The effects of direct-current shift and modulation detection should be considered. Beat-frequency generation, in carrier-type amplifiers, including two-phase servos, in the presence of harmonically related, spuriously coupled energy, should be considered. An adequate safety margin should be included to allow for performance-prediction uncertainties. Good engineering practive and most specifications require a minimum safety margin of 6 decibels, or one-half the voltage or current threshold, and one-fourth of the power threshold. Safety margins in excess of 6 decibels should be value engineered because of the extra effort that may be required to achieve this increased confidence level.

b. Determine the ratio by which the interference source amplitude must be reduced to achieve the acceptable level of spuriously coupled, undesired energy at the susceptible circuit. This is the required isolation loss ratio. If the source of interference has an amplitude that varies with frequency and/or the susceptible circuit has a threshold level that varies with frequency, the required isolation ratio may vary with frequency.

c. Determine the ratio by which the interference source amplitude is actually reduced at the susceptible circuit. This is the actual spurious-coupling loss. The actual spurious-coupling loss may be expected to vary with frequency.

   1. Generate a schematic of the interference source circuit, the susceptible circuit, and the spurious-coupling circuit, including all source and load termination circuitry up to the input or output element of an isolation device such as a transistor.

   2. Determine which coupling modes exist and become dominant in various parts of the frequency spectrum. Both coupling modes may be significant simultaneously, if the matched and the deliberately mismatched-impedance transmission-line conditions exist.

   3. Calculate the losses that occur when energy, at a variety of frequencies, passes through the spurious-coupling network. Be sure to include all critical frequencies related to both the interference function and the susceptible function. This spurious-coupling network is merely a frequency-sensitive voltage, current, or power divider.

      If the network appears to be too complex for convenient manipulation, break the circuit down into several subsections that are easier to analyze. When analyzing the network, look for any interactions that may occur between isolated subsections of the circuitry when these subsections are combined to form a complete circuit. The reactance chart is a convenient tool that simplifies this analytical process considerably.

d. Determine the magnitude of the existing safety margin or the degree of additional isolation required to produce an acceptable safety margin.

   1. If the actual spurious-coupling loss has a greater magnitude than the required isolation ratio, divide the actual spurious-coupling loss ratio by the required isolation ratio to determine the existing safety-margin ratio.

   2. If the required isolation ratio has a greater magnitude than the actual spurious-coupling loss ratio, divide the required isolation ratio by the actual spurious-coupling loss ratio to determine how much the spurious-coupling loss ratio must be increased.

e. Determine what method will be used to increase the actual spurious-coupling loss ratio.

   1. Increasing the separation distance between two conductors will reduce the mutual capacitance and mutual inductance of the conductors and increase the spurious-coupling loss.

   2. Grounding guard conductors between the two conductors will reduce the mutual capacitance and mutual inductance by the creation of a pseudoshield and increase the spurious-coupling loss. Guard conductors, grounded at one end only, provide electric-field isolation only. Guard conductors, grounded at both ends, provide both electric- and magnetic-field isolation.

   3. Dielectric isolation-spacer strips can be used to increase the separation distance between layers of stacked cables when moderately susceptible and moderately interfering conductors are not available to create a buffer zone, a relatively small improvement is required, and schedule limitations preclude the rearrangement of cabling.

4. Metallic isolation-spacer strips increase the isolation between layers of stacked cables, with performance similar to electrostatically shielded cables, but have the disadvantage of incomplete isolation in the connector region; therefore, the potential shielding effectiveness of the cable shield is degraded significantly. Moderate increases in electric-field isolation will be achieved with long cables, and very little improvement will occur when used with short cables. Magnetic-field isolation will be negligible except at radio frequencies where high transmission losses through the cable tend to limit the usefulness of shielded cable. A shielded cable placed between unshielded cables is equivalent to a metallic-isolation spacer strip.

5. Nonferrous-shielded cable greatly increases the electric-field isolation between stacked layers of cable but produces a negligible magnetic-field improvement except at radio frequencies where high transmission losses through the cable tend to limit the usefulness of shielded cable. Shielded cable also serves as a pseudometallic, isolation spacer strip when placed between layers of unshielded cable.

6. Ferrous-shielded cable greatly increases both the electric- and magnetic-field isolation between stacked layers throughout the frequency spectrum. The ferrous-shielded cable has a limited audiofrequency bandwidth that is inversely proportional to length and the square root of the effective shield-permeability. A ferrous-shielded cable serves as a pseudometallic, isolation spacer strip when placed between layers of unshielded cable but does not provide magnetic-field isolation throughout the frequency spectrum in this application.

7. Adjacent conductors in a cable used as "hot" and return conductors, a conductor pair, either floating or balanced, or conductor sets in a multiple-phase circuit, sense or radiate almost equal induced-interference levels that tend to cancel each other at the susceptible circuit.

   This is virtually the same effect produced by twisting conductors. The resultant pseudotwist greatly increases the spurious-coupling loss. This conductor configuration supplements and complements, but is not a complete substitute for the shielded cable.

   The increase in the spurious-coupling loss, when compared with widely separated conductors, is equal to the average amplitude of the induced interference in the individual conductors, divided by the difference in amplitude between conductors in the circuit. Over-and-under conductors, separated by the thinnest practical layer of dielectric, are superior to side-by-side conductors on conventional centerline spacings.

8. Unshielded conductors that contain any ferrous materials, either alloyed or deposited, exhibit the high-frequency transmission-loss characteristics of the equivalent ferrous-shielded conductors. Ferrous or ferrous-coated, unshielded conductors form a distributed low-pass filter capable of rejecting radio-frequency energy, without the complications introduced by shielding, if the ferrous material is maintained in a magnetically unsaturated state.

   The ferrous-coated configuration would saturate at relatively low levels. Since the transmission loss of this conductor configuration is amplitude-sensitive, it would be feasible to produce a cable that had limited bandwidth when used to carry low-amplitude functions, and a relatively unlimited bandwidth when used to carry high-amplitude functions, provided that the surrounding environment was free of high-current conductors and other sources of strong magnetic fields.

f. Using the new cable configuration, recalculate the actual spurious-coupling loss ratio and redetermine the existence of an adequate safety margin.

Repeat the foregoing process, as required, until an optimal wiring design for all the circuit combinations involved is achieved for the system.

3.2.10.3 Shielding Fundamentals. Shielding effectiveness is the reduction in spurious coupling of energy from one conductor to another, compared to equivalent unshielded conductors, that occurs when the electromagnetic environments of the two conductors are isolated by the insertion of a shielding medium between the conductors. Shielding effectiveness is usually measured in decibels.

3.2.10.3.1 Theory of Shields (Types of Fields Encountered). Three interrelated types of fields must be considered during any shield design. The plane wave or far field exists when the shield intercepts energy at a point distant from the source of energy in terms of wavelengths, and the resultant wavefront has a negligible curvature. The plane wave contains both electric- and magnetic-field components in a known relationship. The plane wave is not normally considered at low frequencies because the large separation distances required to produce low-frequency plane waves generally attenuate the received energy to acceptable low levels.

The near field, or induction field, exists when the shield intercepts energy at a point near the source of energy in terms of wavelengths, and the resultant wavefront has a significant curvature. The near field contains both electric- and magnetic-field components, but these components lack the relatively fixed relationship that exists between them in the far field; therefore, they are considered separately. The near field and its components, the electric and magnetic fields, are not normally considered at relatively high radio frequencies because the extremely small separation distances required to produce relatively high frequency near fields rarely exist in practical systems and hardware.

The electric field is created by the existence of a potential between two conductive objects isolated by a dielectric and is the mechanism by which a capacitor functions. Transducers such as the monopole and dipole antenna, used to generate and sense electric fields, are basically capacitor plates separated by the propagation distance as a dielectric. Since the mutual capacitance of these capacitor plates is extremely small at any reasonable separation distance and the capacitively coupled current in the circuit loop is negligible, a virtual open circuit exists. Therefore, the electric field is dependent on the existence of a voltage between objects, is not significantly dependent on the flow of current in either the transmitting or receiving circuits, and, as a result, is a high-impedance phenomenon.

The magnetic field is created by the existence of a current flow through a conductive object, which in turn generates a reciprocal current flow in conductive objects and is the mechanism by which a transformer functions. Transducers, such as the loop antenna and current probe used to generate and sense magnetic fields, are basically transformer windings separated by the propagation distance as a transformer core. Since the reactance of the mutual capacitance between windings is extremely high compared with the inductive reactance of the transformer windings, the capacitively coupled voltage across the secondary circuit and the capacitively coupled current through the secondary circuit are limited to relatively low values at any reasonable separation distance, and a near short circuit exists at the secondary winding. Therefore, the magnetic field is dependent on the flow of current through conductive objects and produces reciprocal current flows in conductive objects, is not significantly dependent on the existence of a voltage between objects, and, as a result, is a low-impedance phenomenon.

3.2.10.3.2 Type of Losses Contributing to Shielding Effectiveness. Shielding effectiveness for electric, magnetic, and plane-wave fields is the sum of absorption and reflection losses.

3.2.10.3.2.1 Absorption Loss. When a conductive object is immersed in a magnetic field, a current flow is produced in the object which, in turn, generates a mirror image of the original magnetic field and, in effect, cancels the original magnetic field. Since the object is not a perfect conductor, the secondary field generated is not a perfect replica of the original field, and the existence of this unbalance produces magnetic-field attenuation, rather than perfect magnetic field cancellation, in the practical case. When an object having a high-magnetic permeability is immersed in a magnetic field, the magnetic field tends to detour through, or be concentrated in, the high-permeability material, since high-permeability materials have a low-magnetic resistance or reluctance. If the high-permeability material surrounds another object, the magnetic field detours through the low reluctance of the high-permeability material in preference to flowing through the high-reluctance path of a nonferrous material. Diverting the magnetic field from the interior of the shielded object is equivalent to eliminating the magnetic field. Since the degree of diversion is proportional to the permeability of the ferrous material, and infinitely high permeabilities are not available in practice, a small, residual magnetic field continues to exist, and the magnetic field is attenuated rather than eliminated.

Both of these effects contribute to the absorption loss through the shielding material to a degree dependent on the conductivity and permeability of the materials used.

3.2.10.3.2.2 Reflection Loss. A basic requirement for the efficient transfer of power from one circuit or medium to another is that the source and load impedances be equal or matched. When the impedance of the shield, considered as an inadvertent antenna, does not match the impedance of the electric, magnetic, or plane-wave field, the energy is reflected rather than transferred from one circuit or medium to the other, and the efficiency of the antenna as a transducer is decreased significantly. This decrease of energy transfer or conversion efficiency is known as the reflection loss. The reflection loss is actually the sum of the interface mismatch losses or both side of the shield. The shield is a receiving antenna on one side and a transmitting antenna on the other side.

3.2.10.3.2.3 Total Shielding Effectiveness. The total shielding effectiveness of a shielded layer is the sum of the absorption and reflection losses for the type of field under consideration. When the shield comprises multiple layers, either individual laminated, or coated sheets, the total shielding effectiveness of the composite assembly is equal to the sums of the individual shielding layers.

The shielding effectiveness of any given shield configuration is a function of the shielding material used, the thickness of shielding material, frequency, the type of field being shielded against, the geometry of the interference source, shield assembly and susceptible item, and the configuration of any perforations or discontinuities in the shield.

3.2.10.3.3 Loss Characteristics of Shielding Materials. The two basic types of materials used for shielding are the nonferrous materials such as copper and aluminum that cannot be magnetized and the ferrous materials such as iron, steel, and nickel-iron alloys that are magnetic. The ferrites have electrical properties similar to the ferrous metals, but generally lack the mechanical properties required of a shield, and are limited to applications such as cup cores for self-shielding, high-frequency transformers, and other usages not related to shielding.

3.2.10.3.3.1 Reflection Losses. Any metal of any thickness will provide reflection losses that are a function of the material used, frequency, the geometry of the source of interference, shield assembly and susceptible item, and the type of field being shielded against. Electric-field reflection losses vary inversely with frequency at the rate of 30 decibels per decade or 9 decibels per octave. Plane-wave reflection losses vary inversely with frequency at the rate of 10 decibels per decade, or 3 decibels per octave and are not dependent on geometry. Magnetic-field reflection losses vary at the rate of 10 decibels per decade or 3 decibels per octave at the extremes of the frequency range but have a nonlinear characteristic in the midportion of the frequency spectrum with the losses varying proportionately with frequency at the higher frequencies and inversely with frequency at the lower frequencies. The transition points are dependent on geometry and shield material.

High-conductivity, nonferrous metals provide the highest reflection losses in an electric field because the shield has a low impedance that is mismatched to the high-impedance electric field and relatively little power is transferred between the electric field and the shield. High-resistivity materials provide a better impedance match to the electric field; therefore, the reflection losses are lower. As frequency increases, the inductive reactance of the shield increases providing a better impedance match to the electric field; therefore, the reflection losses vary inversely with frequency. Ferrous materials have an inductance that is considerably higher than nonferrous materials because of the inductance multiplication provided by the magnetic permeability of these materials. Since the inductive reactance of a ferrous shield is considerably higher than that of a nonferrous shield, the ferrous shield provides a better impedance match to the electric field; therefore, the reflection losses are lower.

Plane-wave reflection losses have a different rate of change with frequency and are not geometry dependent but behave like electric-field reflection losses in most other respects.

Magnetic-field reflection losses have a complex relationship to frequency and geometry that is material dependent. High-conductivity, nonferrous metals do not exhibit useful magnetic reflection losses in the power and audiofrequency spectrum with the normally encountered interference source-to-shield separation distances. At frequencies above the audio spectrum, useful reflection losses are obtained. High-permeability ferrous metals do not exhibit useful magnetic reflection losses in the audio and low video frequency spectrum with the normally encountered interference source-to-shield separation distances. At subsonic frequencies and frequencies above the medium video spectrum, useful reflection losses are obtained. At frequencies where useful losses are achieved, the magnetic reflection losses vary proportionately with frequency at the rate of 10 decibels per decade or 3 decibels per octave, except for the case of subsonic frequencies. At subsonic frequencies, where useful losses are achieved, the magnetic reflection losses vary inversely with frequency at the rate of 10 decibels per decade or 3 decibels per octave.

3.2.10.3.3.2 Absorption Losses. Any metal will provide absorption losses that are a function of the material used, material thickness, and frequency. Absorption losses are proportional to the thickness of the material used. Absorption losses, in decibels, increase with the square root of frequency.

Nonferrous metals, in the commonly used thicknesses, do not have useful absorption losses in the power and audiofrequency spectrum. Absorption losses become significant at approximately 10 megahertz and increase rapidly to extremely high values beyond that point.

Ferrous metals of modest permeability, such as cold-rolled steel, in the commonly used thicknesses, do not have useful absorption losses in the power and low-to-medium audiofrequency spectrum. Absorption losses become significant at approximately 1 megahertz and increase rapidly to extremely high values beyond that point.

Ferrous metals of high permeability, such as 4-79 Permalloy in the commonly used thickness, do not have useful absorption losses in the power and audiofrequency spectrum. Absorption losses become significant at supersonic frequencies and increase rapidly to extremely high values beyond that point.

3.2.10.3.3.3 Secondary Reflections. Secondary reflections occurring within the metallic-shielding structure introduce minor errors when neglected if the shield has a low attenuation factor. Most shields are designed for medium-to-high attenuation factors that make the calculation of secondary reflections unnecessary.

3.2.10.4 Electromagnetic Compatibility Tests, Analyses, and Curves. Electrostatic Crosstalk, Magnetic Crosstalk, and Transmission Loss tests were made on various types of unshielded and shielded FCC.

3.2.10.4.1 Electrostatic Crosstalk. The electrostatic crosstalk tests were made on unshielded and shielded cables with the electrical test conditions as follows:

a. Excited conductor

   1. Source impedance - 50 ohms (all cases)
   2. Load impedance - open circuit

b. Sensing conductor

   1. Source impedance - open circuit
   2. Load impedance - 20 Hz to 10 kHz, nominal 100 K ohm receiver input

      - 20 to 200 kHz, Tektronix 1A7 plug-in preamplifier and Tektronix 545 oscilloscope (nominal 1-megohm input shunted by 47 plus 50 pF of cable capacity).

      - 50 kHz to 400 MHz, nominal 50 ohm receiver input.

When two conductors are in close proximity to each other, a small but finite capacitance exists. The conductors are capacitor electrodes and the conductor insulation and other nonconductive materials form the capacitor dielectric. Any varying voltage on one conductor is capacitively coupled onto the other conductor as an induced voltage. The induced voltage appears across the parallel-connected combination of terminating impedances connected to the ends of the second conductor.

The capacitance between the second conductor and objects other than the first conductor also appears in parallel with the terminating impedances unless balanced circuitry is used. This shunt capacitance is a function of the total circuit length while the series coupling capacitance between the two conductors is a function of the shared path length. The series and shunt capacitances form a capacitive voltage divider. The shunt capacitance in the lower leg of the voltage divider is in parallel with the terminating impedances connected across the ends of the conductor.

At low frequencies, the induced voltage is inversely proportional to the reactance of the conductor-to-conductor capacitance. The conductor-to-conductor capacitive reactance is extremely high compared with the value of the parallel-connected terminating impedances connected across the ends of the conductor and the shunt capacitance between the conductor and structural ground plane. The voltage at the interfering conductor is divided between the two legs of this voltage divider in proportion to the impedances in the upper and lower legs of the divider.

As frequency increases, the reactance of the conductor-to-conductor capacitance decreases while the terminating impedances remain relatively constant because of the resistive component usually present. Therefore, the induced voltage that appears between the susceptible conductor and the structural ground plane increases with frequency at the rate of 6 decibels per octave or 20 decibels per decade at low frequencies.

At high frequencies, the induced voltage at the susceptible conductor remains relatively constant as frequency changes. The reactance of the conductor-to-conductor capacitance becomes insignificantly low compared to the relatively constant, resistive component of the conductor terminations, and essentially the entire interfering voltage appears on the susceptible conductor.

If the resistive component of the terminating impedances is dominated by the lower reactance of the parallel capacitance in the terminating hardware and/or the distributed shunt capacitance of the conductor, the resultant capacitive voltage divider has a division ratio that is relatively independent of frequency.

The reactance of this capacitive divider is inversely proportional to frequency, and will load the interference generator at relatively high frequencies if the reactance of the capacitive voltage divider is significantly lower than the output impedance of the interference generator. Under these conditions, the induced voltage will decrease at the rate of 6 decibels per octave or 20 decibels per decade.

If the terminating hardware is dominantly inductive, the reduction in loading on the interference generator would increase the induced voltage at the rate of 6 decibels per octave or 20 decibels per decade.

The measured data are significantly affected by the instrumentation and by the interconnection wiring and test configuration used. Physical limitations reduced the coupled length of the conductors to 4 feet out of a total conductor length of 6 feet. A short-shielded conductor, used to connect the susceptible conductor to the input of the instrumentation, had significant capacitance to the structural ground plane. This shunt reactance formed a capacitive voltage divider in conjunction with the conductor-to-conductor capacitance, preventing the full interference voltage from appearing at the input of the instrumentation.

Because of this pseudoreduction in induced voltage, the coupling loss, even at the high frequencies, does not approach zero decibels. In actual system wiring, this reduction in induced voltage would be considerably smaller except where large capacitive terminations were encountered.

The data break appearing between 200 and 500 kilohertz occurred because available high-input impedance instrumentation was limited to frequencies below 200 kilohertz. Above 500 kilohertz, commercially available instrumentation with the necessary sensitivity was limited to an input impedance of 50 ohms, causing an artifical change in induced voltage.

The distributed electrical parameters of the 6-foot-long test specimens produced conductor resonance at approximately 5 megahertz and a series of harmonically related frequencies above this point. Data above this fundamental resonance in the 2- to 10-megahertz region are dominated by the conductor impedance characteristics and do not reflect the conductor-coupling characteristics.

3.2.10.4.1.1 Electrostatic Crosstalk, Unshielded FCC. Electrostatic crosstalk tests were run on four configurations of unshielded cable setups. Analyses of the resulting information and curves are as given in the following paragraphs.

Figure 3-29 (Curve 1) shows the magnitude of induced voltage appearing across the high-impedance susceptible load referenced to the magnitude of the disturbing voltage on the interfering conductor. Note the perturbation because of resonance near 10 megahertz.

Figure 3-31 (Curve 1) shows the magnitude of induced voltage appearing across the susceptible high-impedance load referenced to the magnitude of the distrubing voltage on the interfering conductor. Note the perturbation because of resonance near 10 megahertz. A lower-frequency perturbation between 200 and 500 kilohertz was because of a reduction of instrumentation input impedance.

Figure 3-39 (Curve 3, referenced to the right-hand vertical scale) shows the change in induced voltage appearing across the susceptible high-impedance load that occurs when various thicknesses of dielectric spacer strips are inserted between stacked unshielded cables.

The dielectric spacer strips produce a relatively small reduction in electric crosstalk when inserted between stacked unshielded cables. Excessive dielectric thicknesses would be required to achieve generally useful degrees of isolation for typical electronic systems. The use of dielectric spacer strips appears to be limited to applications, such as digital systems, where a marginally unsatisfactory condition can be made marginally satisfactory with an extremely small reduction in crosstalk. The dielectric spacer has the advantage of not increasing the capacitance between the conductor and the structural ground plane.

Figure 3-49 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced voltage appearing across the susceptible high-impedance load when one grounded, nonferrous, metallic-shielding foil is inserted between unshielded cables. The nonferrous foil provided an excellent reduction in electric coupling between adjacent stacked cables below 2 megahertz, provided a generally useful reduction in electric coupling between 2 and 5 megahertz, and produced some reduction in electric coupling between adjacent stacked cables between 5 and 400 megahertz.

The electric-shielding effectiveness exceeded 76 decibels between 20 and 50 hertz, was approximately a measured 77 decibels between 50 hertz and 1 kilohertz, gradually increasing to 88 decibels between 10 and 20 kilohertz, gradually decreasing to 80 decibels at 200 kilohertz, decreasing sharply to 55 decibels between 500 kilohertz and 2 megahertz, decreasing to approximately 10 decibels between 100 and 200 megahertz, and increasing to 18 decibels at 400 megahertz. The measured data are of questionable validity above 5 megahertz because of conductor resonances. The low-frequency electric-shielding effectiveness greatly exceeded the values shown but could not be measured because of the state-of-the-art limitations of commercially available monitoring instrumentation.

3.2.10.4.1.2 Electrostatic Crosstalk, Shielded FCC. Electrostatic crosstalk tests were run on seven configurations of shielded cable setups. Analyses of the resulting information and curves are as given in the following paragraphs.

Figure 3-33 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced voltage appearing across the susceptible high-impedance load that occurs when a nonferrous solid-shielded cable with open edges is substituted for an unshielded cable. The nonferrous solid shield with open edges produces some reduction in electric-field crosstalk throughout the measured frequency spectrum, but does not achieve really useful degrees of electric crosstalk reduction. Electric crosstalk reduction is 22 decibels at 1 kilohertz, rising slowly to 36 decibels between 10 and 200 kilohertz, dropping to 22 decibels between 500 kilohertz and 2 megahertz, and gradually rising to 29 decibels at 5 megahertz. A rise to a peak of 48 decibels at 20 megahertz and decrease to 25 decibels at 400 megahertz is a questionable validity because of conductor resonances.

Figure 3-35 (Curve 3, referenced to the right-hand vertical scale) shows that reduction in induced voltage appearing across the susceptible high-impedance load that occurs when a nonferrous solid-shielded cable with joined edges is substituted for an unshielded cable. The nonferrous solid shield with joined edges produces some reduction in electric-field crosstalk throughout the measured frequency spectrum, but does not achieve really useful degrees of electric crosstalk reduction. Electric crosstalk reduction is 23 decibels, rising slowly to 38 decibels between 10 and 200 kilohertz, dropping to 22 decibels at 500 kilohertz, and rising irregularly to 40 decibels at 5 megahertz. A rise to a peak of 64 decibels at 20 megahertz and 72 decibels at 100 megahertz and decrease to 40 decibels at 400 megahertz is of questionable validity because of conductor resonances.

Figure 3-37 (Curve 3, referenced to the right-hand vertical scale) shows the change in induced voltage appearing across the susceptible, high-impedance load that occurs when a woven, nonferrous-shielded, twisted-pair conductor is substituted for unshielded cable. The woven, nonferrous shield produces a moderate increase in electric crosstalk below 200 kilohertz, a small reduction in electric crosstalk between 200 kilohertz and 2 megahertz, indifferent performance above 2 megahertz, and does not achieve really useful degrees of electric crosstalk reduction. A spurious peak occurs at 5 megahertz because of conductor resonance. Electric crosstalk is approximately 10 decibels worse than unshielded flat cable below 200 kilohertz and approximately 6 decibels worse between 200 kilohertz and 2 megahertz. A long rise to a large peak of 108 decibels of electric crosstalk reduction at 5 megahertz, decreasing to zero decibel reduction at 70 megahertz, and a degradation rising to 5 decibels at 400 megahertz is of questionable validity because of conductor resonances. The large peak in Curve 3, near 5 megahertz, is characteristic of the higher "Q" resonance exhibited by RWC. The existence of these higher "Q" resonances was discussed in DAC-56440, "Flat Cable Applications Study," final report.

The increase in electric crosstalk below 2 megahertz is because of the higher conductor-to-conductor capacitance of round conductors. The smaller edge-to-edge surface area of flat conductors in the same cable-layer results in a lower coupling capacitance between flat conductors. Even though the round conductor-to-shield capacitance, plus the instrumentation capacitance, is approximately twice the instrumentation capacitance associated with the unshielded flat conductor and tends to reduce the electric crosstalk, and the much higher conductor-to-conductor capacitance of the shielded round conductor pair more than offsets this trend and increases electric crosstalk by a factor of three times (9 dB) compared to unshielded flat cable below 2 megahertz.

Above 2 megahertz, the higher capacitance to the structural ground plane of the shielded round conductors produces a higher degree of unintentional magnetic coupling, not electric coupling, than the unshielded flat cable test configuration. Capacitance between the conductor and ground plate completes a capacitively coupled, magnetically efficient loop at both the interfering and susceptible conductors. This problem is made even more difficult by the use of monitoring instrumentation with an input impedance of 50 ohms at the higher frequencies. Sensitive high-frequency monitoring instrumentation is universally low-impedance because of physical limitations imposed by high-frequency transmissions systems. The existence of this spuriously coupled, high-frequency magnetic component must be recognized and properly interpreted since it cannot be eliminated at the present state-of-the-measurement art.

Figure 3-43 (Curve 3, referenced to the left-hand vertical scale) shows the reduction in the induced voltage appearing across the susceptible high-impedance load when a nonferrous, solid-shielded cable with open edges is substituted for unshielded cable. The nonferrous, solid shield with open edges provided an excellent reduction in electric coupling between adjacent stacked cables below 5 megahertz and provided a generally useful reduction in electric coupling between adjacent-stacked flat cables between 5 and 400 megahertz.

The electric-shielding effectiveness exceeded 78 decibels at 20 hertz, rising to a value exceeding 122 decibels at 1 kilohertz, decreased to a value exceeding 108 decibels at 2 kilohertz, and decreasing gradually to approximately 40 decibels between 10 and 400 megahertz. The measured data were of questionable validity above 5 megahertz because of conductor resonances. The low-frequency electric-shielding effectiveness greatly exceeded the values shown, but could not be measured because of state-of-the-art limitations of commercially available monitoring instrumentation.

Figure 3-45 (Curve 3, referenced to the left-hand vertical scale) shows the reduction in the induced voltage appearing across the susceptible high-impedance load when a nonferrous, solid-shielded cable with joined edges is substituted for unshielded cable. The nonferrous shield with joined edges provided an excellent reduction in electric coupling between adjacent-stacked cables below 500 kilohertz and a generally useful reduction in electric coupling between adjacent-stacked cables between 500 kilohertz and 400 megahertz. The electric-shielding effectiveness exceeded 78 decibels at 20 hertz, rising to a value exceeding 118 decibels at 500 hertz, decreased to a value exceeding 116 decibels at 1 kilohertz, was measured as approximately 116 decibels between 1 and 5 kilohertz, decreased sharply to 85 decibels at 20 kilohertz, decreased gradually to 78 decibels at 200 kilohertz, and decreased sharply to approximately 46 decibels between 500 kilohertz and 400 megahertz. The measured data were of questionable validity above 5 megahertz because of conductor resonances. The low-frequency electric-shielding effectiveness greatly exceeded the values shown, but could not be measured because of state-of-the-art limitations of commercially available monitoring instrumentation.

Figure 3-47 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced voltage appearing across the susceptible high-impedance load when a woven, nonferrous-shielded cable is substituted for unshielded cable. The woven, nonferrous shield provided an excellent reduction in electric coupling between adjacent-stacked cables below 7 megahertz, decreasing to a negligible effect on the electric coupling between 10 and 100 megahertz, and produced some increase in electric coupling between adjacent-stacked cables between 100 and 400 megahertz.

The electric-shielding effectiveness exceeded 76 decibels at 20 hertz, increasing to a measured 89 decibels at 50 hertz, decreasing to 82 decibels between 100 hertz and 10 kilohertz, decreasing to approximately 66 decibels between 20 and 200 kilohertz, decreasing to approximately 45 decibels between 1 and 2 megahertz, rising to a large peak of 117 decibels at 5 megahertz, decreasing to approximately zero decibels between 10 and 100 megahertz, and increased the electric coupling between adjacent-stacked cables by 19 decibels at 400 megahertz. The measured data were of questionable validity above 2 megahertz because of conductor resonances. The low-frequency electric-shielding effectiveness greatly exceeded the values shown, but could not be measured because of state-of-the-art limitations of commercially available monitoring instrumentation.

Figure 3-51 (Curve 3, referenced to the left-hand vertical scale) shows the reduction in induced voltage appearing across the susceptible high-impedance load when a high-permeability-ferrous solid shield with open edges and NASA/DAC aluminum connectors is substituted for unshielded cable. The Chromerics shielding gaskets in the aluminum connector were removed and short, copper jumper strips were substituted between the shield foils and the structural ground plane, while the aluminum connector shell was grounded through existing mounting clips. The high-permeability-ferrous solid shield and aluminum connectors provided an excellent reduction in electric coupling between adjacent stacked cables below 3 megahertz, provided a generally useful reduction in electric coupling between 3 and 6 megahertz, and produced some reduction in electric coupling between adjacent stacked cables between 6 and 400 megahertz.

The electric-shielding effectiveness exceeded 78 decibels at 20 hertz, increasing to a value exceeding approximately 112 decibels between 50 hertz and 10 kilohertz, decreasing to 12 decibels at 10 megahertz, increasing to a peak of 40 decibels at 20 megahertz, and decreasing to 10 decibels at 400 megahertz. The measured data are of questionable validity above 5 megahertz because of conductor resonances. The low-frequency electric-shielding effectiveness greatly exceeded the values shown, but could not be measured because of state-of-the-art limitations of commercially available monitoring instrumentation.

3.2.10.4.2 Magnetic Crosstalk. The magnetic crosstalk tests were made on unshielded and shielded cables with the electrical test conditions as follows:

a. Excited conductor

   1. Source impedance - 50 ohms (all cases)
   2. Load impedance - Short to ground-plane return

b. Sensing conductor

   1. Source impedance - Short to ground-plane return
   2. Load impedance - Nominal 50-ohm receiver input

The mere existence of a longitudinal conductor creates a small but finite inductance. When two conductors are in close proximity to each other, a transformer is created because of the mutual inductance of, or magnetic coupling between, the inductances. Any varying current flow in one conductor is inductively coupled into the other conductor as an induced voltage because of the transformer action. The induced voltage appears across the winding-load impedance and causes a current flow through the load impedance that is a function of the varying current amplitude in the first conductor, the coupling coefficient of the conductors, the output winding impedance, and the impedance of the load.

Turns ratio is ordinarly fixed at 1 to 1 by the geometry of the procured cable. The coupling coefficient and mutual inductance are based on a uniform geometry between conductors over the entire lengths of the conductors. Extensions of the conductor lengths beyond the points of common geometry increase the total series inductance of the conductors without increasing the total mutual inductance of the conductors and, therefore, reduces the effective mutual inductance and effective coupling coefficient of the conductors.

At low frequencies, the induced voltage is proportional to the impedance of the susceptible conductor. The conductor-source impedance is extremely low compared with the value of the series-connected terminating impedances at the ends of the conductor. The unloaded induced-output voltage appears across the series-connected terminating impedances.

The current through the terminating impedances is a function of the unloaded induced-output voltage of the susceptible conductor and the value of the series-connected terminating impedances at the ends of the conductor. As frequency increases, the source impedance of the susceptible conductor increases proportionately; therefore, the impedance match between the conductor and the series-connected load impedances improves as frequency increases, and the induced current through the series-connected load impedances increases at the rate of 6 decibels per octave or 20 decibels per decade at low frequencies.

At high frequencies, the induced voltage is inversely proportional to the impedance of the susceptible conductor. The conductor-source impedance is extremely high compared with the value of the series connected at the ends of the conductor. The induced-output voltage is heavily loaded by the series-connected terminating impedances.

If the two ends of the conductor were connected together, a short-circuit current would flow. This short-circuit current has twice the amplitude of a current flowing from the conductor through a load impedance that matches the conductor-source impedance.

The short-circuit current is a function of the unloaded induced voltage and the conductor-source impedance. When the relatively low, series-connected terminating impedances are inserted in series with the source impedance of the conductor, the induced current through the terminating impedances is reduced proportionately.

As frequency increases, the source impedance of the susceptible conductor increases proportionately; therefore, the impedance match between the conductor and the series-connected load impedances gets worse as frequency increases, and the induced current through the series-connected load impedances decreases at the rate of 6 decibels per octave or 20 decibels per decade at high frequencies.

At some intermediate frequency, where the two curves with opposing slopes meet, optimum coupling between the conductors occurs. At and near this frequency, the current flowing through the series-connected terminating impedances is equal to the current flowing through the interfering conductor multiplied by the coupling coefficient of the conductors. Since the coupling coefficient of most adjacent conductors approaches unity, the induced current through the terminating impedances almost equals the disturbing current through the interfering conductor near this optimum-coupling frequency.

The measured data are affected by the characteristics of the instrumentation and interconnection wiring and the test configuration used. The effective coupling coefficient of the two conductors was reduced significantly because of the relatively short length of the available test specimens. Physical limitations reduced the coupled length of the conductors to 4 feet out of a total conductor length of 6 feet. Since this additional uncoupled inductance is in series with both the interfering and susceptible conductors, the coupling coefficient is reduced in proportion to the square of this ratio. Interconnection wiring at the ends of the test specimen also contributed to the reduction in effective-coupling coefficient.

Because of this pseudoreduction in the coupling coefficient of the test specimens, the coupling loss, even at the optimum coupling frequency, does not approach zero decibel. In actual system wiring, this reduction in effective-coupling coefficient would not occur except for isolated cases.

3.2.10.4.2.1 Magnetic Crosstalk, Unshielded FCC. Magnetic crosstalk tests were run on four configurations of unshielded cable setups. Analyses of the resulting information and curves are as given in the following paragraphs.

Figure 3-28 (Curve 1) shows the magnitude of induced current flowing through the susceptible 50-ohm load referenced to the magnitude of the disturbing current flowing in the interfering conductor. Note the perturbations because of resonances near 10 megahertz and 20 megahertz.

Figure 3-30 (Curve 1) shows the magnitude of induced current flowing through the susceptible 50-ohm load referenced to the magnitude of the disturbing current flowing in the interfering conductor. Note the perturbations because of resonances between 5 and 10 megahertz and near 20 megahertz. A lower-frequency perturbation of approximately 6 decibels near 100 kilohertz is probably nonexistent. Speculation indicates that personnel operating the signal sources may have supplied incorrect disturbing-current amplitude information during measurements made at this frequency.

Figure 3-38 (Curve 3, referenced to the right-hand vertical scale) shows the change in induced current flowing through the susceptible 50-ohm load that occurs when various thicknesses of dielectric spacer strips are inserted between stacked unshielded cables. The dielectric spacer strips produce a relatively small reduction in magnetic crosstalk when inserted between stacked unshielded cables. Excessive dielectric thicknesses would be required to achieve generally useful degrees of isolation for typical electronic systems. The use of dielectric spacer strips appears to be limited to applications such as digital systems where a marginally unsatisfactory condition can be made marginally satisfactory with an extremely small reduction in crosstalk. The dielectric spacer has the advantage of not increasing the capacitance between the conductor and the structural ground plane.

Figure 3-48 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load when one grounded, nonferrous, metallic-shielding foil is inserted between unshielded cables. The nonferrous foil had a negligible effect on the magnetic coupling between adjacent stacked conductors below 5 kilohertz and produced some reduction in magnetic coupling between 5 kilohertz and produced some reduction in magnetic coupling between 5 kilohertz and 400 megahertz but did not provide a generally useful reduction in magnetic coupling at any point in the entire measured frequency spectrum. Above approximately 100 kilohertz, over most of the frequency range where some reduction in magnetic coupling occurs, the transmission losses of this cable configuration become so high that the cable becomes useless as a medium for the transmission of information in electronic systems.

The magnetic-shielding effectiveness was actually zero decibel below 5 kilohertz. The low negative values shown are because of small, long-term drifts in the monitoring instrumentation. The magnetic-shielding effectiveness gradually increased above 5 kilohertz to approximately 32 decibels between 500 kilohertz and 5 megahertz, dipped to 4 decibels at 10 megahertz, went through zero decibels at 250 megahertz, and dropped to a negative value of 10 decibels at 400 megahertz. The measured data are of questionable validity above 5 megahertz because of conductor resonances.

3.2.10.4.2.2 Magnetic Crosstalk, Shielded FCC. Magnetic crosstalk tests were run on seven configurations of shielded cable setups. Analyses of the resulting information and curves are as given in the following paragraphs.

Figure 3-32 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load that occurs when a nonferrous solid-shielded cable with open edges is substituted for an unshielded cable. The nonferrous solid shield with open edges has no effect on adjacent conductor magnetic crosstalk within the shield below 10 kilohertz, produces some magnetic crosstalk reduction above 10 kilohertz, and only provides a useful magnetic crosstalk reduction at one point near 1 megahertz, at which frequencies high transmission losses tend to make the cable relatively useless as a medium for the transmission of system information. Note the perturbations because of the reference cable resonance between 5 and 10 megahertz and near 20 megahertz, and the shielded cable resonance near 50 megahertz.

Figure 3-34 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load that occurs when a nonferrous solid-shielded cable with joined edges is substituted for an unshielded cable. The nonferrous solid shield with joined edges has no effect on adjacent-conductor magnetic crosstalk within the shield below 5 kilohertz, produces some magnetic crosstalk reduction between 5 and 200 kilohertz, and provides a useful magnetic crosstalk reduction above 200 kilohertz, at which frequencies high transmission losses tend to make the cable relatively useless as a medium for the transmission of system information. Note the perturbations because of shielded-cable resonance near 10, 20, 100, and 200 megahertz.

The distributed electrical parameters of the 6-foot-long test specimens produced conductor resonance at approximately 5 megahertz and a series of harmonically related frequencies above this point. Data above this fundamental resonance in the 2- to 10-megahertz region are dominated by the conductor impedance characteristics and do not accurately reflect the conductor coupling characteristics.

Figure 3-36 (Curve 3, referenced to the right-hand vertical scale) shown the change in induced current flowing through the susceptible 50-ohm load that occurs when a woven, nonferrous-shielded, twisted-pair conductor is substituted for unshielded cable. The woven nonferrous shield has no effect on adjacent conductor magnetic crosstalk within the shield over the entire frequency spectrum measured. The low-frequency end of Curve 3 is actually on the zero-decibel line, but is shown as approximately -3 decibels because of small long-term drifts in the instrumentation. The large peak in curve 3 near 5 megahertz is characteristic of the higher "Q" resonances exhibited by RWC.

Figure 3-42 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load when a nonferrous solid-shielded cable with open edges is substituted for unshielded cable. The nonferrous solid shield had a negligible effect on the magnetic coupling between adjacent stacked conductors below 5 kilohertz, produced some reduction in magnetic coupling between 5 and 500 kilohertz, provided a generally useful reduction in magnetic coupling between 500 kilohertz and 150 megahertz, and produced some reduction in magnetic coupling between 150 and 400 megahertz. Above approximately 100 kilohertz, at the frequencies where a useful reduction in magnetic coupling occurs, the transmission losses through the cable become so high that the cable is useless as a medium for the transmission of information in electronic systems.

The magnetic-shielding effectiveness was actually zero decibel below 5 kilohertz. The low-negative value shown is the result of small, long-term drifts in the monitoring instrumentation. The magnetic-shielding effectiveness gradually increased from zero decibel at 5 kilohertz to an irregular 34 decibels between 500 kilohertz and 10 megahertz, increased to a peak of 58 decibels at 50 megahertz, and dropped to 4 decibels at 400 kilohertz. The measured data have questionable validity above approximately 5 megahertz because of conductor resonances.

Figure 3-44 (Curve 3, referenced to right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load when a nonferrous solid-shielded cable with joined edges is substituted for unshielded cable. The nonferrous solid shield had a negligible effect on the magnetic coupling between adjacent stacked conductors below 2 kilohertz, produced some reduction in magnetic coupling between 2 and 500 kilohertz, provided a generally useful reduction in magnetic coupling between 500 kilohertz and 250 megahertz, and produced some reduction in magnetic coupling between 250 and 400 megahertz. Above approximately 100 kilohertz, at the frequencies where a useful reduction in magnetic coupling occurs, the transmission losses through the cable become so high that the cable is useless as a medium for the transmission of information in electronic systems.

The magnetic-shielding effectiveness was actually zero decibels below 2 kilohertz. The low-negative value shown is the result of small, long-term drifts in the monitoring instrumentation. The magnetic-shielding effectiveness gradually increased from zero decibel at 2 kilohertz to 36 decibels between 500 kilohertz and 2 megahertz, increased to a peak of 80 decibels at 20 megahertz, and dropped to 14 decibels at 400 megahertz. The measured data have questionable validity above approximately 5 megahertz because of conductor resonances.

Figure 3-46 (Curve 3, referenced to the right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load when a woven nonferrous-shielded round-conductor cable is substituted for an unshielded cable. The woven nonferrous shield had a negligible effect on the magnetic coupling between adjacent round conductors throughout the entire measured frequency spectrum. The magnetic-shielding effectiveness was actually zero decibel below 2 megahertz. The low values shown are because of small long-term drifts in the monitoring instrumentation.

The peaks at 5 megahertz, between 20 and 50 megahertz, at 200 megahertz, and the dips at 10, 100, and 400 megahertz are because of conductor resonances. The measured data have questionable validity above approximately 2 megahertz because of conductor resonances. The large peaks and dips in curve 3, above 2 megahertz, are characteristic of the high "Q" resonances exhibited by round conductors.

Figure 3-50 (Curve 3, referenced to right-hand vertical scale) shows the reduction in induced current flowing through the susceptible 50-ohm load when a high-permeability-ferrous, solid shield with open edges and NASA/DAC aluminum connectors is substituted for unshielded cable. The Chromeric shielding gaskets in the aluminum connector were removed, and short, copper jumper strips were substituted between the shield foils and the structural ground plane, while the aluminum connector shell was grounded through existing mounting clips. The high-permeability-ferrous solid shield and aluminum connectors had a negligible effect on the magnetic coupling between adjacent stacked cables below 20 kilohertz, produced some reduction in magnetic coupling between adjacent-stacked cables between 20 kilohertz and 200 megahertz, and produced some increase in magnetic coupling between adjacent stacked cables between 200 and 400 megahertz, at which frequencies the transmission losses through the cable become so high that the cable is useless as a medium for the transmission of information in electronic systems.

The magnetic-shielding effectiveness was actually zero decibel below 20 kilohertz. The low-negative value shown is the result of small, long-term drifts in the monitoring instrumentation. The magnetic-shielding effectiveness gradually increased to approximately 16 decibels between 500 kilohertz and 2 megahertz, decreasing to 2 decibels at 10 megahertz, increasing to a peak of approximately 20 decibels between 20 and 50 megahertz, decreasing to 4 decibels between 100 and 200 megahertz, goes through zero decibel at 260 megahertz, and increases the magnetic coupling by 6 decibels at 400 megahertz. The measured data are of questionable validity above 5 megahertz because of conductor resonances.

3.2.10.4.3 Transmission Loss. The transmission-loss test was made on unshielded and shielded cables with both 1 ohm and 100 k ohm terminations which were matched to the test setup instrumentation.

The distributed-series inductance and shunt capacitance of a conductor form a distributed low-pass filter. This low-pass filter has negligible losses at low frequencies, has a rapidly increasing rate of losses over a relatively narrow band of transition frequencies determined by the distributed electrical constants of the conductor, and has a relatively constant-rate loss with a high, absolute value at high frequencies. The conductor is a useful medium for the transmission of information in electronic systems at low frequencies where the reactive losses of the conductor do not greatly exceed the resistance losses of the conductor. At the high frequencies, the reactive losses become prohibitive and make the conductor useless for the transmission of information in electronic systems.

The unshielded conductor bandwidth is extended for digital systems by utilizing impedance-matched, unshielded conductor pairs, but this application for bulk cable will decrease sharply as integrated circuitry reduces computer size. This bandwidth extension is limited by high-frequency copper losses. At these higher frequencies, only a very limited number of communications systems are capable of accommodating the large conductor losses. The percent of conductor feet utilized for interconnection wiring in future digital and communications systems is expected to be insignificant.

3.2.10.4.3.1 Transmission Loss, Unshielded FCC. Transmission-loss tests were run on one configuration of unshielded cable. Analyses of the resulting information and curves are as given in the following paragraphs.

With 1-ohm terminating impedances (Curve 1, Figure 3-87) at the ends of the 6-foot unshielded cable, the transmission loss down a centrally located conductor is negligible below 50 kilohertz, increases gradually to 40 decibels at 5 megahertz, dips to an irregular 33 decibels between 10 and 150 megahertz, and increases to 38 decibels at 200 megahertz and 42 decibels at 400 megahertz. This relatively short cable would be generally useful below 50 kilohertz, of limited usefulness between 40 and 100 kilohertz, and relatively useless above 100 kilohertz.

With 100,000-ohm terminations (Curve 2, Figure 3-87), the transmission loss is negligible below 500 hertz, increases gradually to a peak of 49 decibels at 2 megahertz, dips slightly to 43 decibels at 5 megahertz, drops to 4 decibels between 10 and 20 megahertz, dips to a negative value (transmission gain) of 20 decibels at 100 megahertz, and dips to a negative value (transmission gain) of 5 decibels at 400 megahertz. The transmission gains are probably false indications because of conductor resonances. This relatively short cable would be generally useful below 500 hertz, of limited usefulness between 500 hertz and 3 kilohertz, and relatively useless above 3 kilohertz.

The transmission losses are proportional to length, and the useful frequency limit is inversely proportional to length. The measured data are of questionable validity above 5 megahertz because of conductor resonances.

3.2.10.4.3.2 Transmission Loss, Shielded FCC. Transmission-loss tests were run on three configurations of shielded cables. Analyses of the resulting information and curves are as given in the following paragraphs.

With 1-ohm terminating impedances (Curve 1, Figure 3-88) at the ends of the 6-foot length of mesh-shielded cable, the transmission loss down a centrally located conductor is negligible below 60 kilohertz, increases gradually to 34 decibels at 5 megahertz, continues irregularly at approximately 30 decibels to 200 megahertz, and increases sharply to 65 decibels at 400 megahertz. This relatively short cable would be generally useful below 60 kilohertz, of limited usefulness between 60 and 130 kilohertz, and relatively useless above 130 kilohertz.

With 100, 000-ohm terminations (Curve 2, Figure 3-88), the transmission loss is negligible below 1 kilohertz, increasing to 58 decibels at 5 megahertz, dropping sharply to a negative value (transmission gain) of 13 decibels at 10 megahertz, decreasing through zero decibels at 84 megahertz, and increasing to 19 decibels at 400 megahertz. The transmission gain is a false indication, probably because of conductor resonance. This relatively short cable would be generally useful below 1 kilohertz, of limited usefulness between 1 and 2. 5 kilohertz, and relatively useless above 2. 5 kilohertz.

The transmission losses are proportional to length, and the useful frequency limit is inversely proportional to length. The measured data are of questionable validity above 5 megahertz because of conductor resonances.

With 1-ohm terminating impedances (Curve 1, Figure 3-89) at the ends of the 6-foot solid 4-79 Permalloy shielded cable with open edges, the transmission loss down a centrally located conductor is negligible below 9 kilohertz, increases gradually to approximately 25 decibels between 2 and 5 megahertz, increases sharply to approximately 73 decibels between 10 and 100 megahertz, and decreases to 35 decibels at 400 megahertz. This relatively short cable would be generally useful below 9 kilohertz, of limited usefulness between 9 kilohertz and 30 kilohertz, and relatively useless above 30 kilohertz.

With 100, 000-ohm terminations (Curve 2, Figure 3-89), the transmission loss is negligible below 1 kilohertz, increases gradually to approximately 55 decibels between 3 and 5 megahertz, dropping sharply to approximately 2 decibels between 9 and 50 megahertz, increasing rapidly to a peak of 56 decibels at 200 megahertz and decreasing to 45 decibels at 400 megahertz. This relatively short cable would be generally useful below 1 kilohertz, of limited usefulness between 1 and 2. 5 kilohertz, and relatively useless above 2. 5 kilohertz.

The transmission losses are proportional to length, and the useful frequency limit is inversely proportional to length. The measured data are of questionable validity above 5 megahertz because of conductor resonances.

With 1-ohm terminating impedances (Curve 1, Figure 3-90) at the ends of the 6-foot length of woven, copper-shielded, round-conductor cable, the transmission loss down the conductor is negligible below 25 kilohertz, gradually increases to approximately 30 decibels between 2 and 50 megahertz, dips to 12 decibels at 100 megahertz, increases to a peak of 32 decibels at 200 megahertz, and decreases to 23 decibels at 400 megahertz. This relatively short cable would be generally useful below 25 kilohertz, of limited usefulness between 25 and 65 kilohertz, and relatively useless above 65 kilohertz.

With 100, 000-ohm terminations (Curve 2, Figure 3-90), the transmission loss is negligible below 700 hertz, gradually increasing to 63 decibels at 5 megahertz, drops sharply to 2 decibels at 10 megahertz, increases to a peak of 12 decibels at 20 megahertz, decreases through zero decibel at 38 megahertz to negative values (transmission gains) of 18 decibels at 100 megahertz, 7 decibels at 200 megahertz, and 20 decibels at 400 megahertz. The transmission gains are probably false indications, because of conductor resonances. This relatively short cable would be generally useful below 700 hertz, of limited usefulness between 700 hertz and 2. 5 kilohertz, and relatively useless above 2. 5 kilohertz.

The transmission losses are proportional to length, and the useful frequency limit is inversely proportional to length. The measured data are of questionable validity above 5 megahertz because of conductor resonances.