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Avionics System Design for High Energy Fields

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Final Report

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16. Abstract

Because of the significant differences in transient susceptibility, the use of digital electronics in flight critical systems, and the reduced shielding effects of composite materials, there is a definite need to define design practices which will minimize electromagnetic susceptibility, to investigate the operational environment, and to develop appropriate testing methods for flight critical systems.

A major part of this report describes design practices which will lead to reduced electromagnetic susceptibility of avionics systems in high energy fields. A second part describes the level of emission that can be anticipated from generic digital devices. It is assumed that as data processing equipment becomes an ever larger part of the avionics package, the construction methods of the data processing industry will increasingly carry out into aircraft. These portions of the report should, therefore, be of particular interest to avionics engineers and designers.

This report includes an extensive bibliography on electromagnetic compatibility and avionics issues.

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PREFACE

This report was prepared by CK Consultants under Contract No. NAS2-12448 with the Federal Aviation Administration. Helpful guidance and technical contributions were received by the author from Mr. Chris Kendall, CK Consultants, and from Mr. William E. Larsen, Technical Representative, of the FAA Field Technical Office, Moffett Field, California. The contract defined a number of tasks relating to the design of digital systems to operate properly in a high energy field environment. However, the information in this report does not represent the last word in EMC design.

A report of this type naturally draws upon very many sources -- so many that it is not possible to individually list them all here. Several, however, were of particular aid in this work:

Some of the figures are reprinted by permission of John Wiley & Sons, Inc. from Noise Reduction Techniques in Electronic Systems by Henry W. Ott, Copyright 1976 by Bell Telephone Laboratories, Incorporated.

Information was extracted from a large number of papers originally published in the Symposium Record of the IEEE International Symposium on Electromagnetic Compatibility from 1974 to 1986.

The MECL System Design Handbook, published by Motorola Semiconductor Products, Inc., was also a highly useful source.

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EXECUTIVE SUMMARY

Until recently, avionic equipment was primarily analog, possessing limited bandwidths and utilizing time averaging indicators. Such equipment was not responsive to transient disturbances unless they exceeded the analog device damage level. Now digital electronics are becoming common-place and their use, even in normally analog systems, will prevail in the near future. However, unlike their analog predecessors, they are very susceptible to transient effects, as well as to discrete-frequency radiation. Digital device performance can be adversely affected before the device damage transient level is reached. The operation of many digital devices is at least 10 times more susceptible to transients than that of their analog counterparts.

Other factors are also involved. Modern aircraft are increasingly using fly-by-wire control systems in which direct mechanical or hydraulic linkages are being replaced by solid state digital systems controlling electrical actuators. There is an increasing use of composite materials on the aircraft, which, while offering strength and weight advantages, give less shielding compared to aluminum.

Because of the significant differences in transient susceptibility, the use of digital electronics in flight critical systems, and the reduced shielding effects of composite materials, there is a definite need to define design practices which will minimize electromagnetic susceptibility, to investigate the operational environment, and to develop appropriate testing methods for flight critical systems.

A major part of this report describes design practices which will lead to reduced electromagnetic susceptibility of avionics systems in high energy fields. Another part describes the levels of emission that can be anticipated from generic digital devices. It is assumed that as data processing equipment becomes an ever larger part of the avionics package, the construction methods of the data processing industry will increasingly carry over into aircraft. These portions of the report should, therefore, be of particular interest to avionics engineers and designers.

The airworthiness specialist should also find this report of use. The detailed technical information developed in the report is summarized in the sections titled RULES FOR DIGITAL SYSTEM DESIGN and EMC GUIDELINES CHECK LIST. These sections will help the airworthiness specialist to perform systems review without subjecting himself unnecessarily to the level of detail contained in the body of the text.

The report includes an extensive bibliography on electromagnetic compatibility and avionics issues.

INTRODUCTION

PURPOSE

This report is useful to the engineer, designer and the airworthiness specialist, since important contributions to safely designed systems can be made by each. To this end, the first part of this section is devoted to engineering information: descriptive material, equations, formulas, tables and graphs. The second part presents an analysis of the radiated emissions to be ejected from digital devices. A third part is in the form of a cookbook: a compendium of rules for digital circuit design which have been found to be effective in producing electromagnetic compatibility, and an EMC design checklist.

The engineer and designer will find the detailed information in the first part of most interest and use. The airworthiness specialist may find the rules and checklist sections and the appendix of more interest.

The report is intended to be a direct source of information and is also intended to point the way to further resources. No set of guidelines can adequately substitute for a good EMC background.

Electromagnetic Compatibility (EMC) is the ability of electrical equipment and systems to function in a given electromagnetic environment without mutual interference. Although nearly every electrical and electronic device is capable of generating or being affected by interference, the proliferation of digital control systems in aircraft and the use of ever-faster digital logic requires greatly increased attention to EMC problems. Concurrently with the introduction of digital control in flight critical systems, composite materials such as kevlar and graphite-epoxy are increasingly being used in airfames, offering significant weight and strength advantages, but providing little or no electromagnetic shielding in comparison with aluminum. It should be noted that embedded metallic meshes and foils adhered to the surfaces can greatly alter the shielding properties of composite materials.

Many of the design practices described here have to do with minimizing circuit board emissions. Not only do these practices enable a system to function without interfering with its own operation, but the very practices which reduce system emissions also work to minimize susceptibility. Control of loop areas, grounding design, by-passing, filtering, and attention to the precise methods of connecting cable shields are all extremely important in both the emission and susceptibility problems.

There is a very high degree of reciprocity between emission or radiation from a circuit, and the susceptibility of that circuit to external fields. Throughout this report "susceptibility" can be substituted for "emission" and "radiation", and the meaning will remain substantially the same.

In an aircraft, some of the most severe threats are from its own on-board communications and radar systems, and from radiation from systems which were never intended to radiate. Although these systems are of significantly lower power than the ground-based high energy threats, their proximity to flight critical systems means that close attention must be directed to the purity of emissions of intentional radiators, and to the minimization of radiation from other sources.

In the next few years there will be a major change in the way avionics equipment is packaged -- at least that part of it which performs data processing. Traditionally, line replaceable units have been built in the style that was developed decades ago for audio and rf units. As data processing becomes a part of avionics, we can expect to see the construction techniques of the data processing industry carry over into aircraft. In other words, we will have card cages containing PC cards which plug into mother boards. Connectors and filtering apparatus will be mounted on back planes, and the entire card cage will be mounted within a shielded and ventilated enclosure. With this in mind, a large part of this report is devoted to the layout of PC cards so as to minimize EMC problems. In the section on Generic Digital Devices Radiated Emissions, most of the analysis was performed with the card cage style of construction in mind.

It is fortuitous that the measures taken to protect one system from its neighbors, and to protect the aircraft from the external threats of broadcast, radar, lightning, and the nuclear electromagnetic pulse, all work together to bring about designs which can be strongly resistant to electromagnetic interference. An integrated design approach will consider all of the threats.

BACKGROUND

EMC problems can be broadly divided into the categories of emission and susceptibility. Each of these categories in turn contain radiation and conduction.

Emission refers to the ability of an electrical or electronic device to act as a generator and radiator of radio frequency energy, in the manner of a radio transmitter, except that the radiation is unintentional. Susceptibility refers to the ability of a device to act as a receiver of radio frequency energy, again unintentionally. Radiation refers to the transmission of radio-frequency energy through space in the form of plane waves, also known as E-H, or transverse electromagnetic (TEM) waves. Radiation as a plane wave is measured at distances greater than one-sixth of a wavelength from the radiating device, in what is referred to as the far field. At closer distances a preponderance of electric field or of magnetic field may exist, depending upon the exact nature of the radiating device and the influence of nearby objects. Conduction refers to the transmission of radio frequency energy along or through metallic elements such as power or I/O (input/output) cables, or even on the surface of a cabinet or chassis. Once energy escapes from an enclosure along a cable, it can turn into radiated energy. Conversely, a radiation field can induce current or voltage in a nearby conductor, and turn into conducted energy. Thus, the distinction between conduction and radiation is not always straight-forward.

In general, emission problems are easier to deal with than susceptibility problems. In the case of emission, the sources and their characteristics, such as rise time and frequency, are under some degree of control by the designer, whereas, in the case of susceptibility, the number of interfering sources and their characteristics are outside his control. But, as mentioned, many of the measures taken to reduce emission also operate to reduce susceptibility, because of the high degree of reciprocity between the "transmit" and "receive" situations. For example, shielding which reduces emissions is equally effective in reducing the strength of incoming interference.

Circuit layout and configuration should provide separation and isolation of susceptible circuits from EMI sources, and confinement of EMI sources to nonsusceptible areas. Low level stages of high susceptibility, such as analog video amplifiers, may have to be enclosed in shields and kept separate from all other circuits, and all leads penetrating the shield must be appropriately filtered. Similar isolation measures should be applied to sources of EMI. For example, power supply or logic circuitry should be shielded and filtered to prevent the EMI it generates from reaching other circuits. Orientation and placement of transformers should be planned to minimize mutual magnetic coupling. Wiring runs should be planned so that susceptible wires are not brought close to EMI-generating circuits, or so that EMI-bearing wires are not brought close to susceptible wires or circuits. Typical EMI sources on PC boards are: digital logic, lamp drivers, relays and relay drivers, deflection amplifiers, crystal oscillators, pulse width modulators, DC/DC switching converters, plasma display drivers, and capacitive discharge circuits. Typical susceptor circuits are: video amplifiers, low level analog circuits, sense circuits, and synchro circuits.

It is most effective to attack EMC at the sources of radiation and conduction, rather than attempt to protect the many potential recipients of this interference. Attention given to reducing emission and conduction at the board level can also ease EMC problems within the system itself. EMC problems should be confronted early in the design phase of a new board or system. The production cost of a well laid-out board is little more than that of a poor layout, but the poor design is likely to require expensive modifications later in the form of add-on shielding, bypassing and filtering, or extensive re-design. Designs become increasingly difficult to modify as they mature and the pressure to meet schedules becomes more severe late in a project.

Radiation produced by most electrical devices is nonionizing radiation, completely different in nature from the ionizing radiation produced by x-rays and nuclear reactions. In most cases the radiation which causes electromagnetic interference is completely harmless to humans.

Many techniques are used in attacking EMC problems. Among these are shielding, filtering, decoupling or bypassing, grounding, coupling reduction, loop area control, and impedance control. Each of these concepts will be treated in detail.

DISCUSSION

DIGITAL SYSTEMS SUSCEPTIBILITY

SHIELDING. Electromagnetic shielding is a complex subject, and for a readable treatment in some depth, the reader is referred to Noise Reduction Techniques in Electronic Systems by Henry W. Ott. (Reference #120 in the bibliography).

In any shielding situation, three loss mechanisms are at work:

(a) absorption
(b) reflection
(c) multiple reflection.

The effectiveness of each mechanism is dependent on the type of field (electric or magnetic) and upon the frequency. The distance between the shielding and the source of radiation is also important.

Some general comments on these mechanisms and their interactions follow:

> For low frequency magnetic fields, absorption is the predominant factor, and reflection loss can be assumed to be zero. Accordingly, glossy, thick magnetic material such as steel or mu-metal are used.
>
> The reflection loss for magnetic fields increases with frequency up to a distance of one-sixth of a wavelength. At larger distances, the reflection loss decreases again.
>
> For high frequencies and at distances greater that one-sixth of a wavelength, good conductors such as copper and aluminum are used for shielding, and the predominant loss mechanism is absorption.
>
> In the case of a magnetic field extending from low to high frequencies, it may be necessary to employ two shields: a magnetic material to take care of the low frequency radiation, and a good conductor for the high frequency emissions.
>
> In the near field (less than lambda/6), the electric field has high reflection loss, while the magnetic field has low reflection loss. For any given frequency, the loss mechanisms become equally effective for the electric and magnetic fields at distances greater than lambda/6.

The above comments illustrate the necessity of keeping in mind the type of field involved, the frequencies of radiation, the wavelength, and the dimensional constraints, in any given shielding problem.

Apertures and seams - any kind of break in shielding integrity - can have a significant effect upon radiation. Ideally, any radiating component or equipment should be completely surrounded by shielding material. In the real world, this is impossible. Large apertures occur where CRT's are used for video display. Knobs, shafts and push buttons inevitably breach the shield. A one-inch hole will permit leakage radiation of -40dB at 60 MHz. Where screws are used to hold sections of shielding together, slot radiators may be found between adjacent screws, unless additional means are taken to assure electrical continuity between the screws.

Any breach in the shielding of cables such as the use of "pigtail" terminations of a shield can cause serious degradation of the shielding effectiveness. For example, a 1- to 2-inch pigtail on a coaxial line can reduce the shielding effectiveness from 70dB to 20dB.

Laboratory measurements have shown large differences in the shielding of nearly identical coaxial cables, dependent upon the size of the carrier wires comprising the braid.

Protective surface coatings are often applied to aluminum surfaces. Anodizing coating are not permissible in shielding materials because they are insulating coatings which do not allow good electrical contact between surfaces. Alodine and several other similar coatings are permissible because they are conductive.

Resistivities of various surface finishes are given in Figure 1. Although the resistivities shown are on the order of thirty times that of aluminum, it is difficult to get good contact between aluminum surfaces because of the nearly instantaneous formation of aluminum oxide, which is non-conductive. In practice, better surface contact is obtained between treated surfaces, and the surfaces are also rendered chemically passive.

Treatment	Resistance/cm^2 (microhms)
Alodine 400	79.81
Alodine 600	79.45
Alodine 1000	80.03
Turco 4178	78.77
Turco 4354	78.81
Iridite 14-2	77.43
Bonderite 710	78.17
Oakite Chromcoat	77.33

FIGURE 1. CONDUCTIVE SURFACE FINISH PROPERTIES

Bonding is used to connect subassemblies together, or to connect a piece of equipment to the ground reference. Bonding straps should have a length not greater that five times the width. The minimum thickness should be 0.5 mm (0.020 in.). The material should be copper, brass, or aluminum straps, not braid. The connections at each end should be clean metal-to-metal contact.

The following summary of shielding properties is taken from Ott's book mentioned earlier:

- Reflection loss is very large for electric fields and plane waves.
- Reflection loss is normally small for low frequency magnetic fields.
- A shield one skin depth thick provides approximately 9 dB of absorption loss.

- Magnetic fields are harder to shield against than electric fields.
- Use a magnetic material to shield against low frequency magnetic fields.
- Use a good conductor to shield against electric fields, plane waves, and high-frequency magnetic fields.
- Actual shielding effectiveness obtained in practice is usually determined by the leakage at seams and joints, not by the shielding effectiveness of the material itself.
- The maximum dimension (not area) of a hole or discontinuity determines the amount of leakage.
- A large number of small holes results in less leakage than a larger hole of the same total area.

GROUNDING. Grounding, like shielding, turns out to be more complex than it might at first seem. Once again, Noise Reduction Techniques in Electronic System by Henry W. Ott is a useful and readable reference.

To suggest the importance and the complexity of the grounding problem, consider that all of the following terms are used:

Digital Ground	Quiet Ground
Analog Ground	Earth Ground
Safety Ground	Hardware Ground
Signal Ground	Single Point Ground
Noisy Ground	Multipoint Ground
Shield Ground	

There are probably more, but this list should suffice to make the point.

Just as with any other portion of a circuit or system, the ground arrangement must be designed, not left to chance. Fortunately, at the board and system level, designing a good grounding system is very cost effective. Other than the one-time engineering cost, very little is added in time or material.

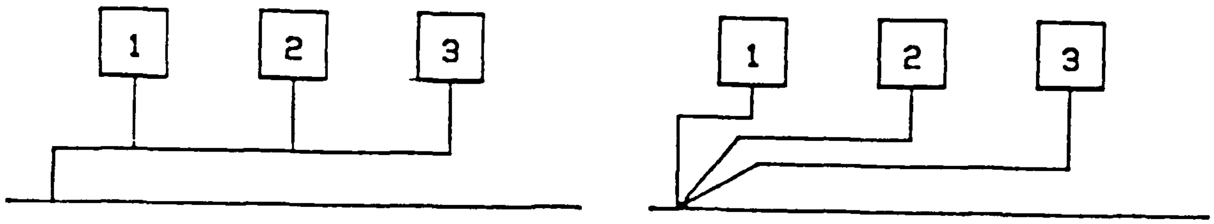
Often a ground system is thought of as an equipotential reference plane which serves as a reference potential. An equipotential plane (or point) is where the voltage does not change regardless of the amount of current supplied to it or drawn from it. In practice, such a plane does not exist. The "equipotential" reference plane often has finite impedances between various ground connection points. When a current flows between these points, the two points are no longer at equal potential, and coupling (common mode) can occur between different parts of the circuit.

It is more useful to consider the various grounds in a circuit as return paths for currents. It is the designer's job to cause these currents to flow where he wants them.

A major effort in digital circuit design is to reduce loop area, that is, conductors and their return paths should be kept close together. It is with loops that magnetic fields are generated or picked up. A poorly designed ground system in which currents are not flowing where intended, can result in many large loop areas - almost guaranteed to cause problems.

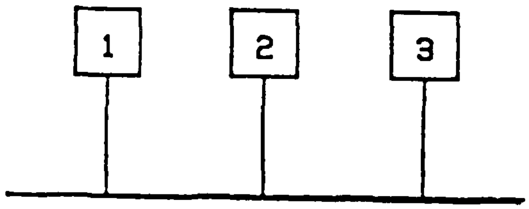
In the design of a grounding system, it is helpful to classify the various grounds by type and by noise-generating capabilities. For example, analog and signal circuits and their returns are generally of low level and are quiet in nature. In contrast, digital circuits often have fast rise time characteristics and high frequency spectral components, if high speed data is being handled. Clock circuits are particularly noisy. Motor and relay circuits are even noisier.

Once circuits have been grouped by type, various grounding topologies can be considered, such as series single point, parallel single point, and multipoint. These schemes are shown in Figure 2.



SERIES CONNECTION PARALLEL CONNECTION

TWO TYPES OF SINGLE POINT GROUNDING CONNECTIONS.



MULTIPOINT GROUNDING CONNECTIONS

FIGURE 2. SERIES, PARALLEL, AND MULTIPOINT GROUNDING

In the series single point connection, the most sensitive circuit returns should be connected closest to the final equipotential point.

In Figure 3 is shown a typical grouping of various grounds.

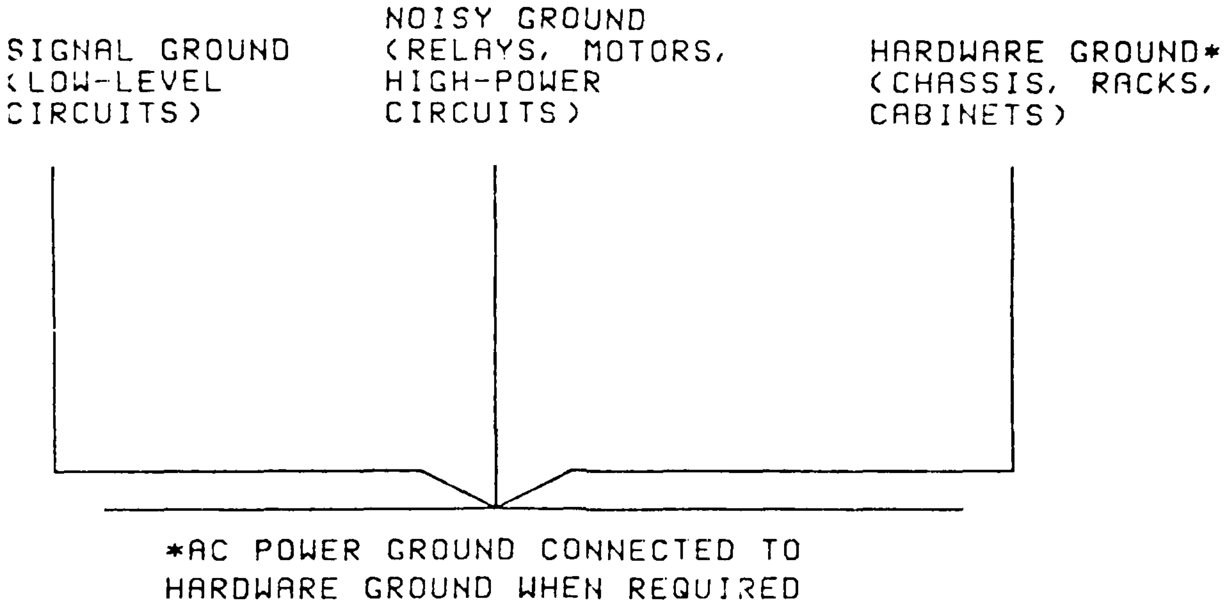


FIGURE 3. GROUPING OF GROUNDS

Grounding for low frequencies and for high frequencies requires different approaches, and where both low and high frequencies are present, ground system designs can become more sophisticated, and compromises may have to be made.

Figure 2, in addition to showing ground grouping, illustrates methods of grounding appropriate below 1 MHz.

Above 1 MHz, multipoint grounding is preferable, as shown in Figure 4. Between 1 and 10 MHz, single point grounding may be used if the longest ground conductor is less than 1/20 of a wavelength. Otherwise, multipoint grounding should be used.

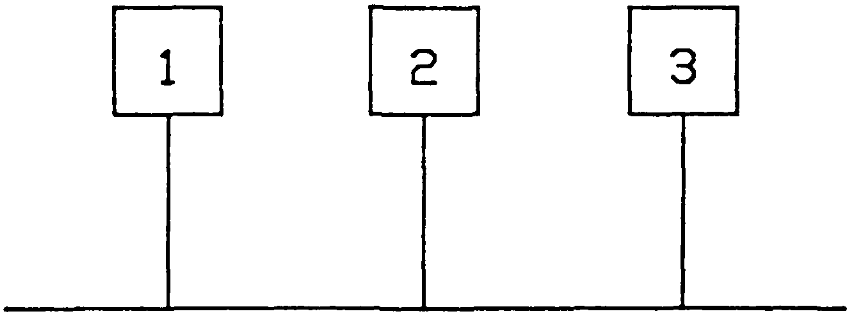


FIGURE 4. MULTIPOINT GROUNDING

At high frequencies, multipoint ground connections from various parts of the circuit must be kept as short as possible. With high speed digital logic and clocks, even 1/4-inch of ground lead length may have a significant impedance (inductive reactance) and cause the "grounded" circuit to become somewhat elevated above ground.

High conductivity of the groundplane to which multipoint grounds are attached is important in order to reduce the impedance between the ground points. This is commonly accomplished by using copper or tin plated copper. Increasing the thickness of the groundplane will not help at high frequencies because most of the current flow is in the surface layer (skin effect).

Grounding considerations for cables and connectors are treated in the cabling section.

Grounding for Loop Control. Figures 2 and 3 presented earlier deal with the problem of reducing the impedance of the ground return scheme. The purpose of reducing the impedance is to avoid, or at least minimize, the flow of several return currents through a shared impedance. When two or more return currents flow through a shared impedance, the noise in each circuit is coupled to the others.

A problem in single reference grounding is that the principle of loop area control is often overlooked. The design of a grounding scheme must take into account not only how the ground currents flow in the return process, but how the signal related to each return flows.

Figure 2 shows how the returns might allow the return currents to flow but is only half the issue. The other half is how the signals flow with respect to how the returns flow. No matter how much care is given to the treatment of the returning currents, if the high sides form large loop areas with shared return flows of current, coupling between co-planar loops will occur.

In Figure 5 a loop area problem is illustrated. A signal circuit connects to daughter board 3 and returns on two separate paths to a central point from both the driver circuit and the daughter board. This arrangement forms co-planar loops with similar circuits in boards 1 and 2. These co-planar loops, brought about by the fact that the signal sides were not routed close to the return sides, can result in magnetic coupling among the daughter boards.

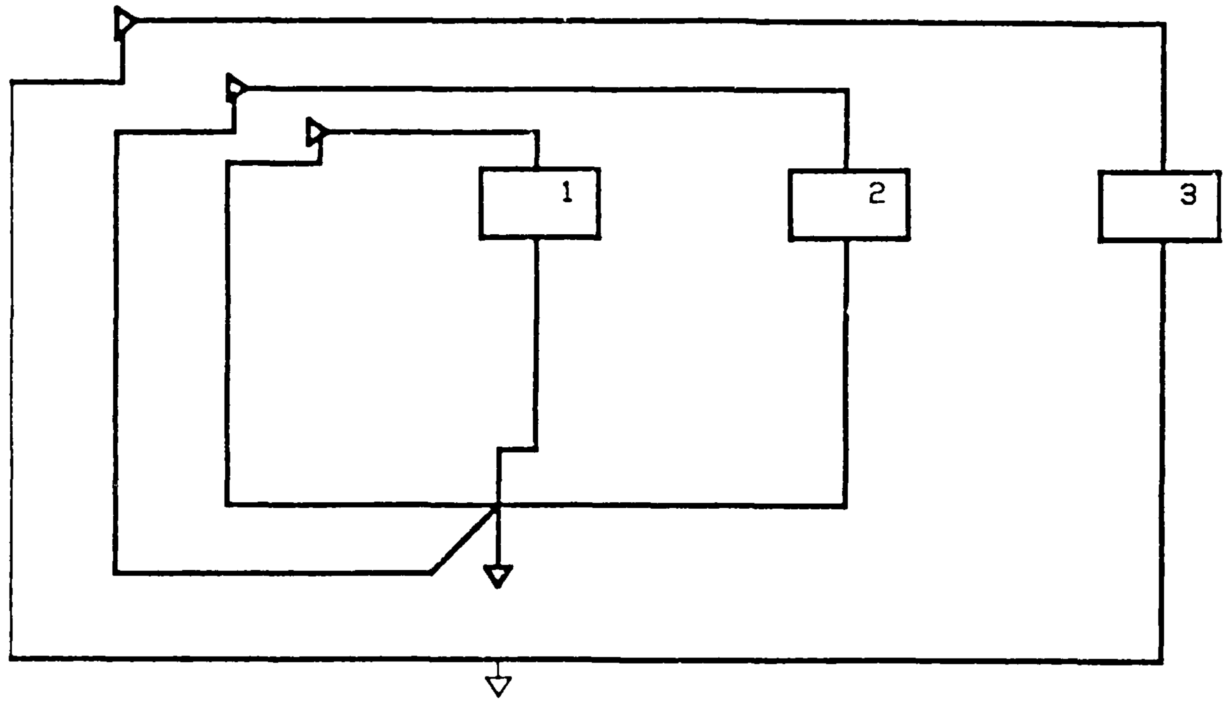


FIGURE 6. THE ISSUE OF LOOP CONTROL

Figure 6 shows how a single ground from each daughter board is routed to a single ground point on the mother board. These long parallel runs are intended to direct the return currents to a single point. Unfortunately, at digital frequencies there will be crosstalk due to inductive and capacitive coupling among these parallel runs.

A better technique is to connect every return pin of each daughter board directly to a common ground plane of the mother board. The digital return is also tied to the ground plane adjacent to the digital signal pin. Now the mutual inductance between the signal trace and the return reduces the return impedance, forces the return current to flow directly under the signal trace, and so reduces loop area.

The same kind of coupling problem created by long parallel return traces can result from long parallel signal traces, for example, in the distribution of signals from a mother board to a daughter board. Just as with the ground traces a common ground plane can force the return currents to flow under their respective signal traces, with the same benefits of reduced return impedance and reduced loop area.

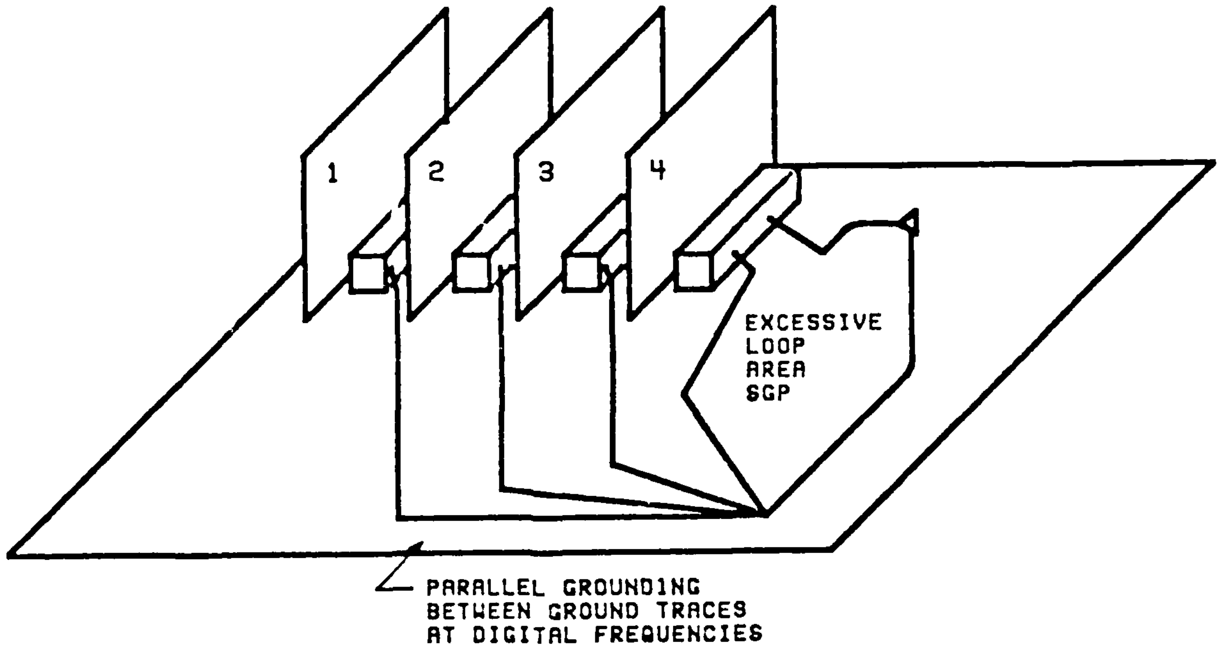


FIGURE 6. SINGLE REFERENCE GROUNDING ON MOTHERBOARD

In the next several figures various aspects of loop area control are illustrated. Figure 7 shows typical grounding of IC's to a ground plane, and poses the question "how does the return current flow?" Figure 8 shows how the return current flows as a mirror image of the signal trace.

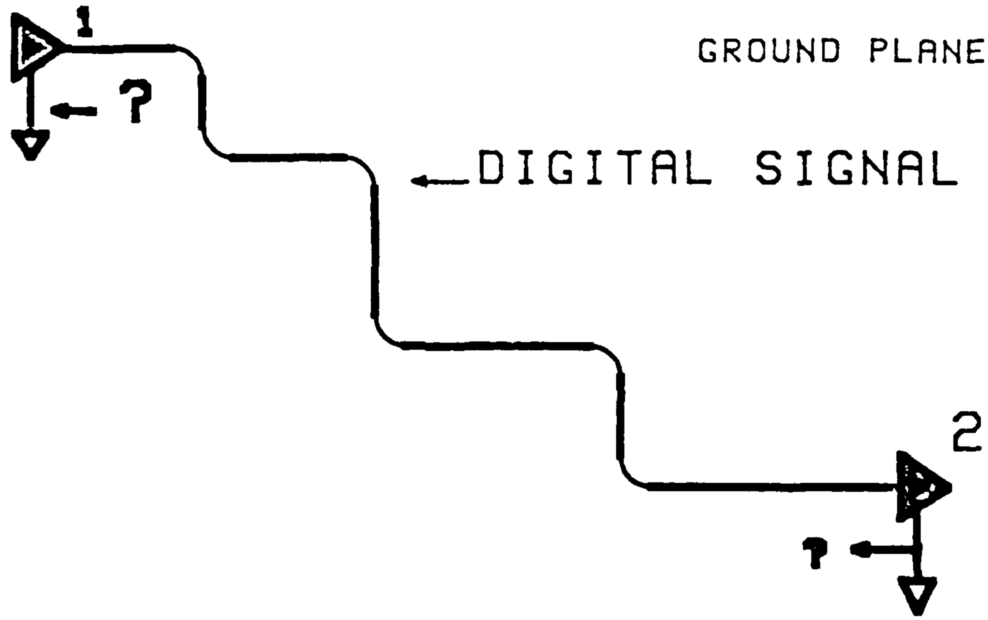


FIGURE 7. HOW DOES THE CURRENT RETURN?

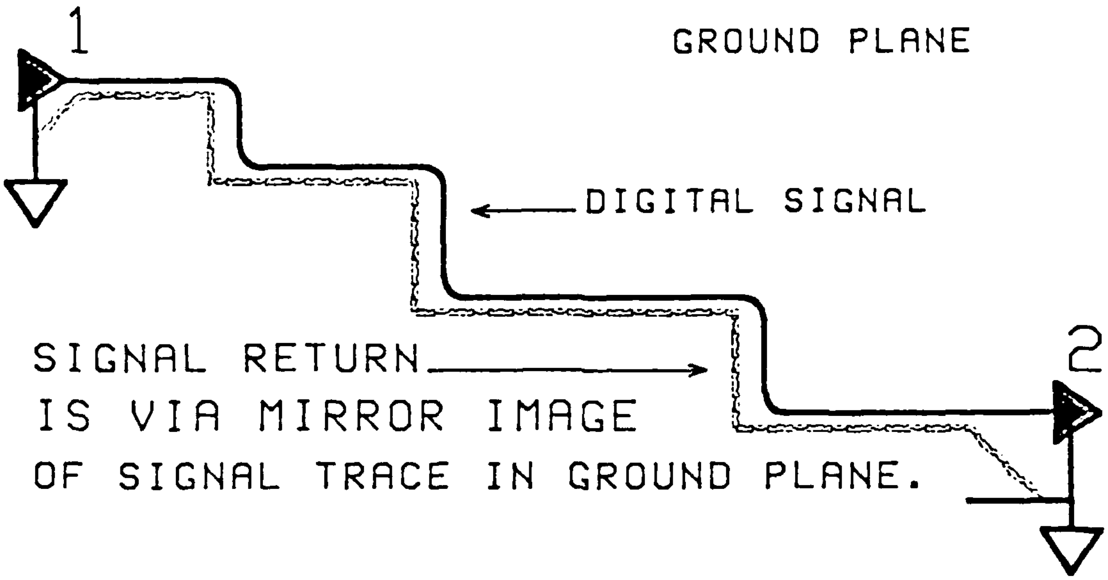


FIGURE 8. SIGNAL RETURN VIA MIRROR IMAGE

In Figure 9 two ground plane areas are shown on a multilayer board, one for digital grounds and one for analog grounds, with a separation area between them. No signal traces on any layer are allowed to cross the separation area. Figure 10 shows the loop area created if signals are allowed to cross the separation area.

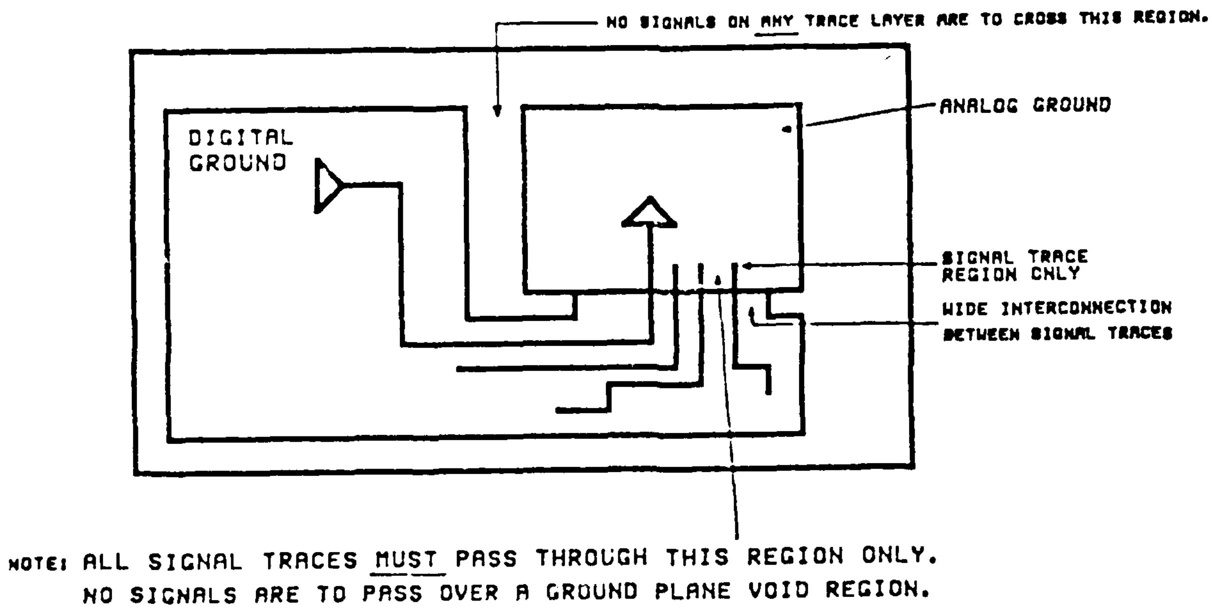


FIGURE 9. DUAL GROUNDPLANES

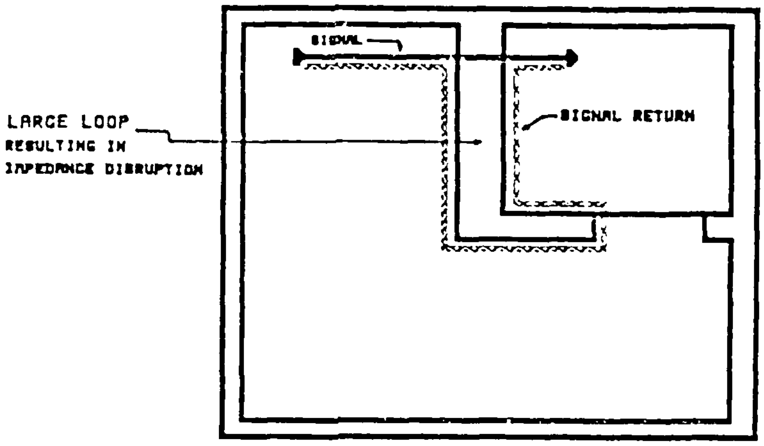


FIGURE 10. SIGNAL TRACE PASSING OVER A VOID REGION

Figure 11 illustrates the flow of return current from a daughter board to a mother board. On both boards, the proximity of the signal trace to the groundplane forces the return current to flow directly under the signal trace. Note that in the connector between the two boards a loop area appears because the ground connection is some distance from the signal trace.

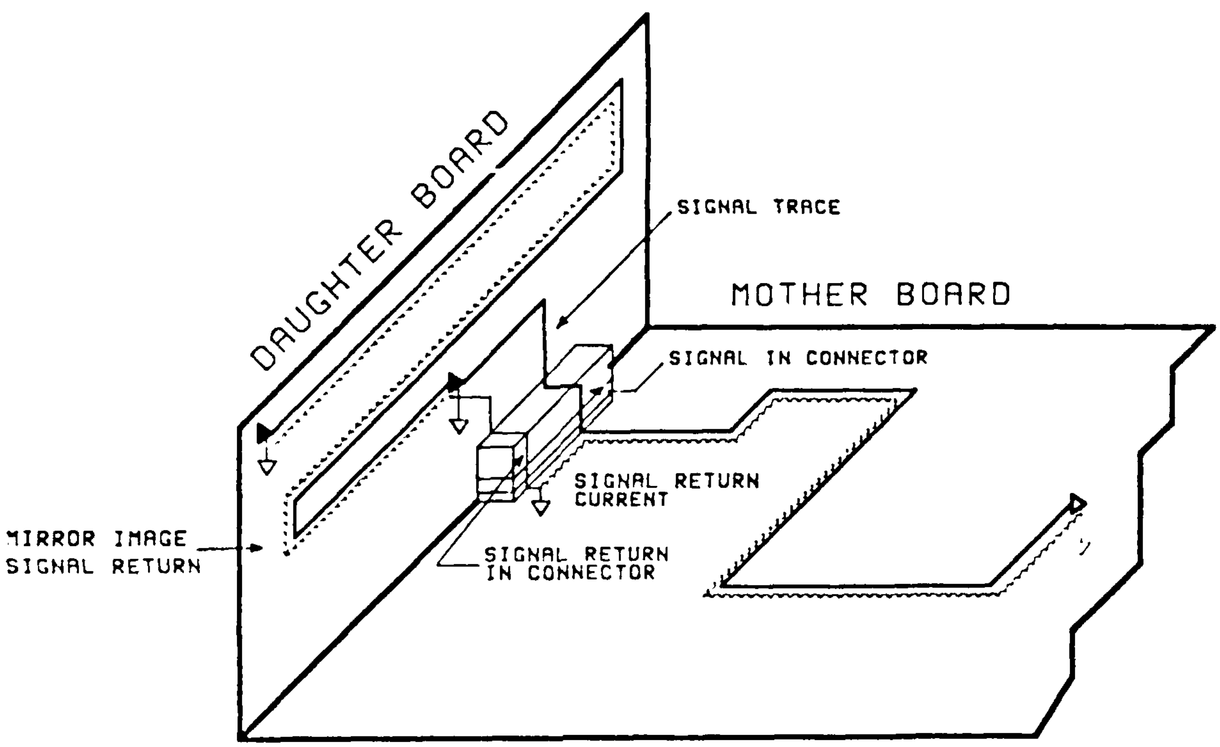


FIGURE 11. DAUGHTER BOARD TO MOTHERBOARD GROUNDING

Figure 12 shows a method of loop area reduction on two-sided boards by running power and return traces orthogonally on opposite sides of the board.

1. TOP OF BOARD HAS ALL VERTICAL TRACES
2. BOTTOM OF BOARD HAS ALL HORIZONTAL TRACES
3. FEEDTHROUGHS WHERE POWER TRACES INTERSECT AND WHERE GROUND TRACES INTERSECT
4. DECOUPLING CAPACITORS BETWEEN POWER AND GROUND AT CONNECTORS AND AT EACH IC
5. SIGNAL LINES FOLLOW VERTICAL/HORIZONTAL PATTERN

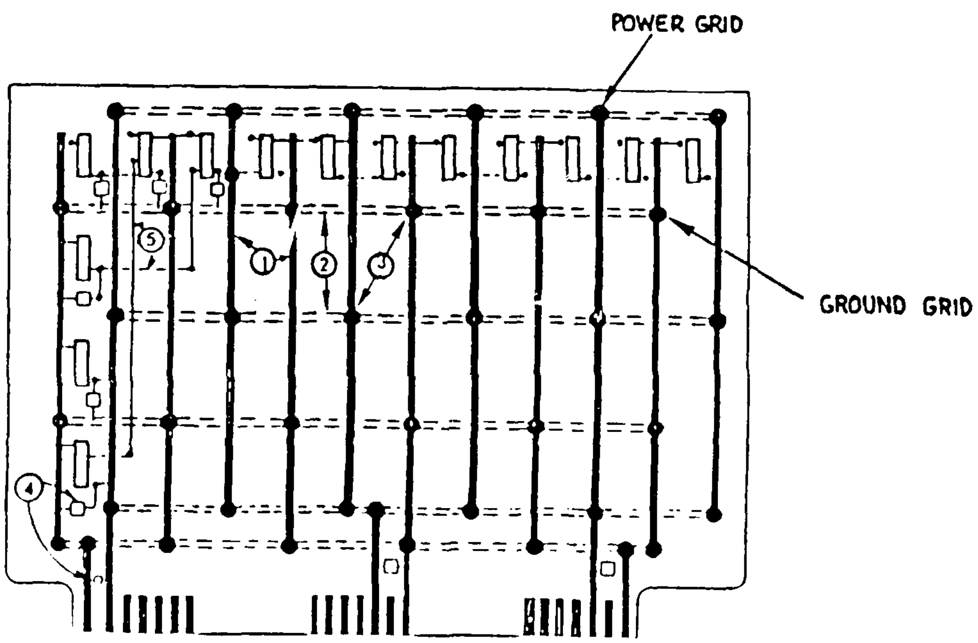
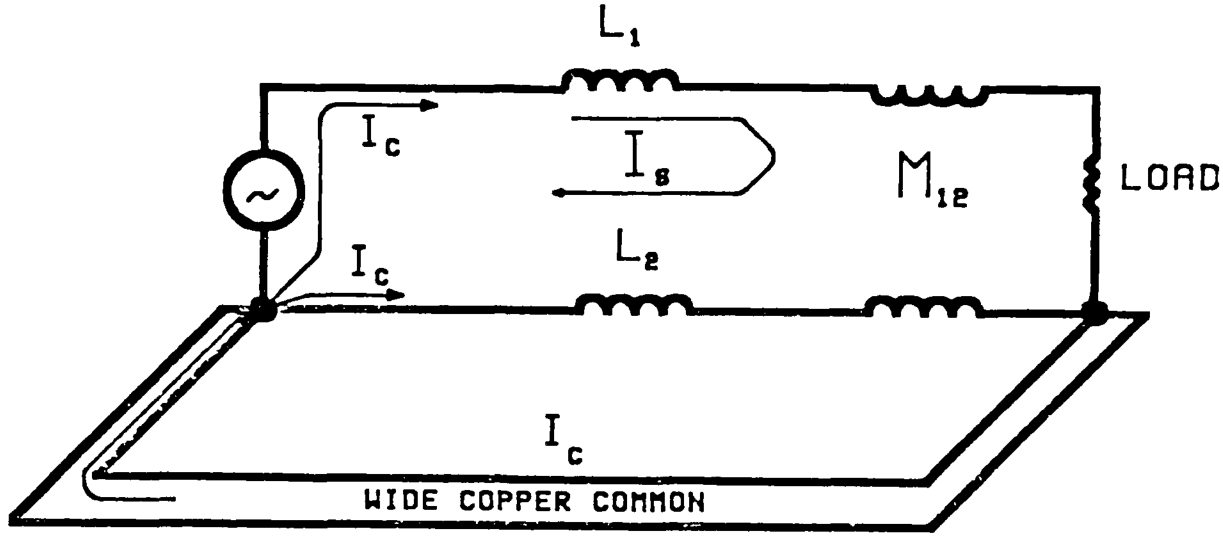


FIGURE 12. POWER DISTRIBUTION ON A TWO-SIDED BOARD

Mutual Inductance as a Common Mode Isolation Device. Normally one thinks in terms of breaking a ground loop by using techniques such as isolation transformers, differential drive circuits, and optical isolators. In digital transmission the best circuit isolator is the transmission line formed by the signal trace above the ground plane. The high mutual inductance between the trace and its ground plane lowers the overall impedance seen by the signal current flow while increasing the impedance to common noise sources shared on the mother board ground plane. Therefore, by going to a transmission line communication of signals between the various boards, one is able to control loop area and decrease the influence of common ground noise at the same time.

Figure 13 shows how the mutual inductance is subtractive for signal currents, and additive for common mode currents, thus tending to reject common mode current flow by forcing it to flow through a higher impedance.



IMPEDANCE TO I_1 : L_1 (SELF INDUCTANCE) + L_2 (SELF INDUCTANCE) $-2M_{12}$ (MUTUAL INDUCTANCE)

IMPEDANCE TO I_C : $L_1 + L_2 + 2M_{12}$

HIGH MUTUAL INDUCTANCE FORCES CURRENT TO RETURN.

FIGURE 13. MIRROR IMAGING OF PARALLEL TRACES

Electrically Long and Short Grounding. One of the major distinctions in grounding technique, including the proper grounding of shielded cables as well as the referencing of various low and high frequency circuits, deals with the length of the circuit itself. In low frequency applications, Figures 2 and 3 are very important in terms of understanding the impedance to the flow of the return currents, especially when the susceptibility threshold of the circuits is extremely low compared to that of digital circuits. Further, in most low frequency cases, the electrical length of the return plane or return trace is extremely short compared to a wavelength of the frequency being transmitted. For example, the quarter wave resonant length for a 20,000 Hz audio signal would be 3.75 kilometers. The likelihood of such a length existing except in AC power distribution is very remote.

Therefore, the signal trace typically appears electrically short and the only impedance concerns are the DC resistance of the trace, the self inductance of the trace, and possibly some skin effect increase in impedance. As long as the signal return trace is properly sized, the impedance within the bandwidth of the circuit is generally relatively small.

A cable shield which is used to block capacitive coupling at low audio frequencies does not require grounding at both ends. In fact, in some cases, grounding a shield at both ends can allow common mode currents to flow in the shield and couple into the circuit. This is especially true for coaxial cables where the common mode current can add to the signal current and cause disruption. However, at frequencies beyond 2 to 3 kHz coaxial cables begin to reject this type of coupling through mutual inductance. Multiconductor shielded cables, however, can often be grounded at one end and thereby prevent common loop coupling currents from flowing on the shield in the first place. The inner circuits have their own reference included and therefore do not see any adverse coupling.

Recent experiments and even experiments done as early as 1956 have shown that current flowing on the shield of a multi-twisted pair conductor will have very little influence if the twisting within the run of the shielded cable is very tight. Typical results have shown that if the twisting inside maintains at least 18 lays per foot, then the influence of the current of the shield becomes virtually nil.

PROPAGATION. Signals transmitted through wires, cables, or on printed circuit boards are subject to noise, distortion and time delay. These effects are in part related to the characteristic impedance of the signal-carrying conductors, the dielectric constant of the board or cable insulating materials, and the source and load terminating impedances.

A very complete discussion of this subject with emphasis on ECL technology is given in the Motorola MECL System Design Handbook (Reference No. 108 in the bibliography).

Velocity of Propagation. The velocity of propagation, vp, is the speed at which data is transmitted through conductors or on a printed circuit board. In air, the velocity of propagation is the speed of light, 3×10^8 meters per second, or about 12 inches per nanosecond. In a dielectric material, the velocity is slower, and is given by

$$vp = \frac{c}{k^{1/2}}$$

where c is the speed of light and k is the effective dielectric constant. Typically k is about 3 for printed circuit boards even though the relative dielectric constant of the board material is near 4.5. The reason for this is that part of the energy flow is in air, and part in the dielectric medium. A typical dielectric constant yields a propagation velocity of 6 to 7 inches per nanosecond.

Propagation Delay. Propagation delay is the time required for a signal to travel the length of a printed circuit board trace at the velocity of propagation, vp. If the length of the trace is 14 inch and the velocity of propagation is 7 inch per nanosecond, the propagation delay is 14/7, or 2 nanoseconds.

Some typical delays are given below.

Transmission Medium	Delay, ns/inch	Delay, ns/foot
Microstrip	0.148	1.78
Stripline	0.188	2.26
Coaxial	0.127	1.52

If a high rise time pulse is transmitted, distortion of the pulse can occur which is a function of the propagation delay and the source and load terminating impedances.

Figure 14 shows an unterminated transmission line. The source impedance is low; the load impedance is very high. At time zero, a high rise time pulse starts from point A. At time T_D later the signal arrives at point B and is reflected because the transmission line is not terminated in its characteristic impedance, Zo.

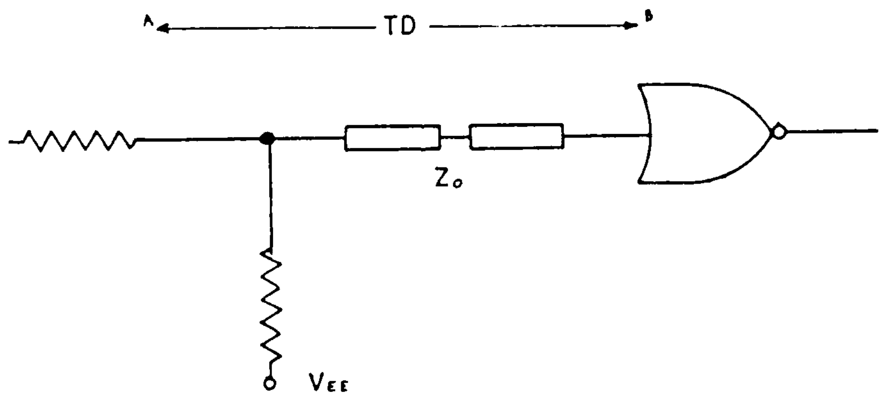


FIGURE 14. UNTERMINATED TRANSMISSION LINE

Because the load impedance is high, the reflection adds to the incoming signal, causing overshoot. The reflected signal travels back to point A, arriving at time $2\ T_D$. Because the source impedance is low, the re-reflected signal is negative, and subtracts, causing undershoot. Successive repetitions of this process occur, at diminishing amplitudes because of line losses, causing ringing.

The undershoot condition decreases the noise immunity level of the system, but can be limited to about 15 percent if the two-way delay of the line is less than the rise time of the pulse.

The maximum line length can be calculated from

$$Lmax = \frac{t_r}{2T_D}$$

where t_r= pulse risetime

T_D = propagation delay

If this condition cannot be met, a solution is to terminate the line in its characteristic impedance.

Impedance Control. The characteristic impedance of a transmission line is determined by the dimensions of the conductors and by the dielectric constant.

For a coaxial line, for example,

$$Zo = \frac{1}{(e_r)^{1/2}} (60) \ln \frac{D}{d}$$

where D is the inner diameter of the outer conductor
where d is the outer diameter of the inner conductor
where e_r is the relative dielectric constant.

For a microstrip line, as illustrated in Figure 15,

$$Zo = \frac{87}{(e_r + 1.41)^{1/2}} \ln \frac{(5.98h)}{(0.8w + t)}$$

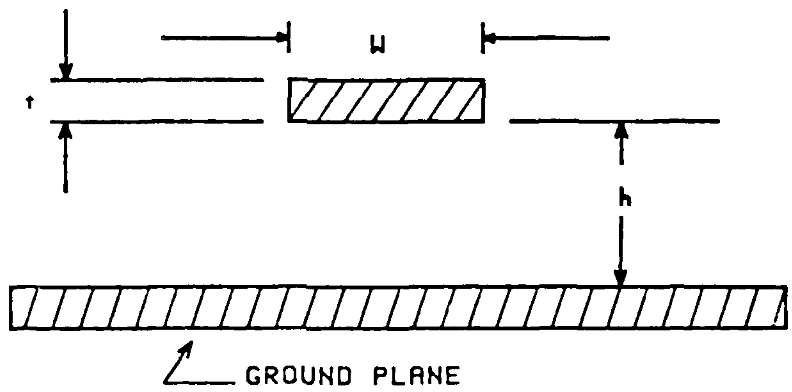


FIGURE 15. MICROSTRIP TRANSMISSION LINE

This particular microstrip impedance formula is valid for:

$$1 < e_r < 15$$

$$0.1 < \frac{w}{h} < 3.0$$

For a stripline, as illustrated in Figure 16,

$$Zo = \frac{60}{(e_r)^{1/2}} \quad Ln \frac{4b}{2.10\ w\ (0.8 + t/w)}$$

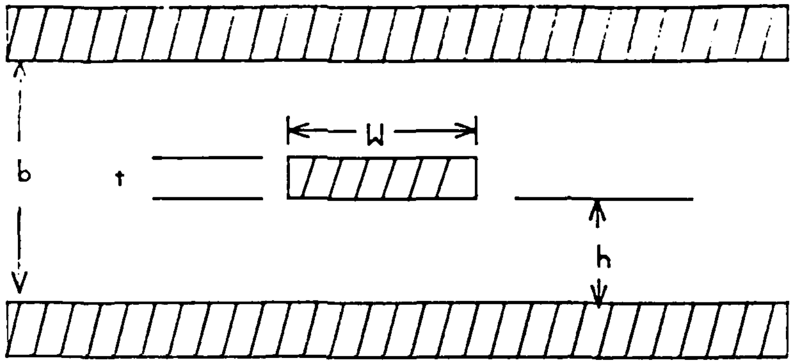


FIGURE 16. STRIPLINE

This stripline impedance formula is valid for:

$$\frac{W}{b-t} > 0.35$$

$$\frac{t}{b} < 0.25$$

Many other formulas for microstrip and stripline impedances can be found in the literature, some of a very complicated nature, and readily usable only via computer program.

What is apparent from the formulas given is the necessity for control of trace width, height above ground, and thickness (determined by plating). Control over the dielectric constant is also required.

Another formula for characteristic impedance is in terms of the lumped circuit inductance and capacitance:

$$Z_o = (L/C)^{1/2}$$

where L is the inductance per unit length
and C is the capacitance per unit length.

For rather typical values of 50nH per foot and 20 pf per foot,

$$Z_o = (50 \times 19^{-9}/20 \times 10^{-12})^{1/2} = 50 \text{ ohms}$$

The inductance, hence line impedance can be increased by decreasing trace width. Conversely, the capacitance can be increased by widening the trace or reducing the height, producing a lower line impedance.

D.C. noise immunity is that signal level at which a system can just distinguish between a 0 and a 1. For typical TTL devices, this noise immunity level is between 0.4 and 1.0 volts.

When noise pulse width is less than the pulse rise time, (typically 10 ns for TTL), the noise immunity level increases beyond the 0.4 to 1.0 volt region. This increased level is referred to as the a.c. noise immunity level.

Knowledge of the noise immunity level of a particular digital system is necessary in order to calculate the required protection level it must be given in the presence of known or estimated threats.

COUPLING. Coupling of noise or signals can occur through a variety of mechanisms:

1. Direct Conduction
2. Common Impedance Coupling
3. Induction (Near Field)

    a. Inductive (Magnetic)
    b. Capacitive (Electric)

4. Radiation (Far Field)

All of these mechanisms can operate singly, or in combination, and conversions from one to another can occur.

Direct Conduction interference travels to or from equipment through interconnecting cables. The typical propagation mode is through power cables, signal cables, and improper ground cables and buses. An example is noise from a switching power supply coupling into the input power lines, and then being conducted via the power line to nearby equipment.

Solutions include decoupling (bypassing) of signal and power line, filtering to allow only the signal frequencies of interest to pass, and shielding to decrease both induction pick-up and radiation with its subsequent conversion into conductive interference.

Common impedance coupling, or common mode coupling occurs when two circuits share an impedance, usually a return conductor. This situation is illustrated in Figure 17.

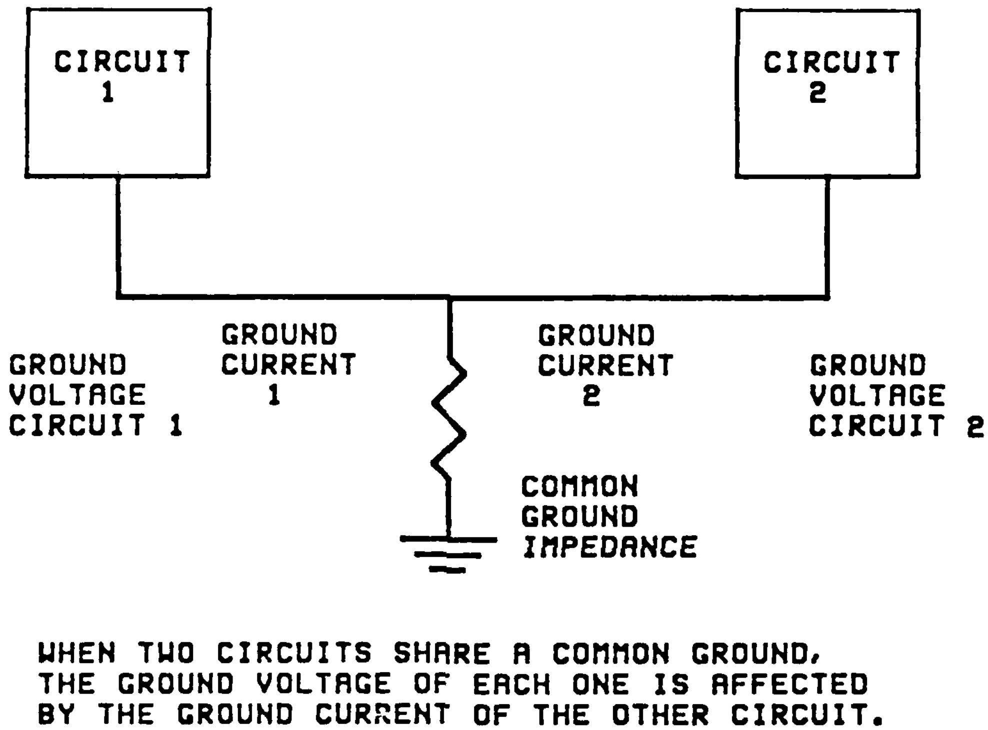


FIGURE 17. COMMON MODE COUPLING

Induction coupling occurs in the near field at distances less than one-sixth of a wavelength (actually λ/π).

Magnetic or inductive coupling occurs when current flows in a wire and the magnetic flux encircles a second wire. The greater the loop area of the current flow, the more opportunity there is for flux linkage to adjacent circuits, and also, the greater the opportunity for flux lines from the adjacent circuits to penetrate the loop area and encircle its conductors.

Electric field or capacitive coupling occurs when voltage exists on, for example, a probe, or capacitor plate. E-field lines can terminate on a nearby object, and energy transfer through capacitive coupling can occur.

An example of magnetic and electric field coupling acting together occurs in the case of near-end and far-end crosstalk. Consider two parallel and adjacent traces above a ground plane on a printed circuit board. The electric coupling has the same phase of coupling for signals traveling in either direction. The magnetic coupling is of opposite phase for signals traveling in opposite directions. The result is that much more signal (typically 20 dB) is transferred between the input ends of two lines (near-end crosstalk) than to their output ends (far-end crosstalk).

This directional property offers a simple way of reducing crosstalk just by paying attention to the direction of signal travel.

Equivalent circuits for low frequency crosstalk are shown in Figure 18.

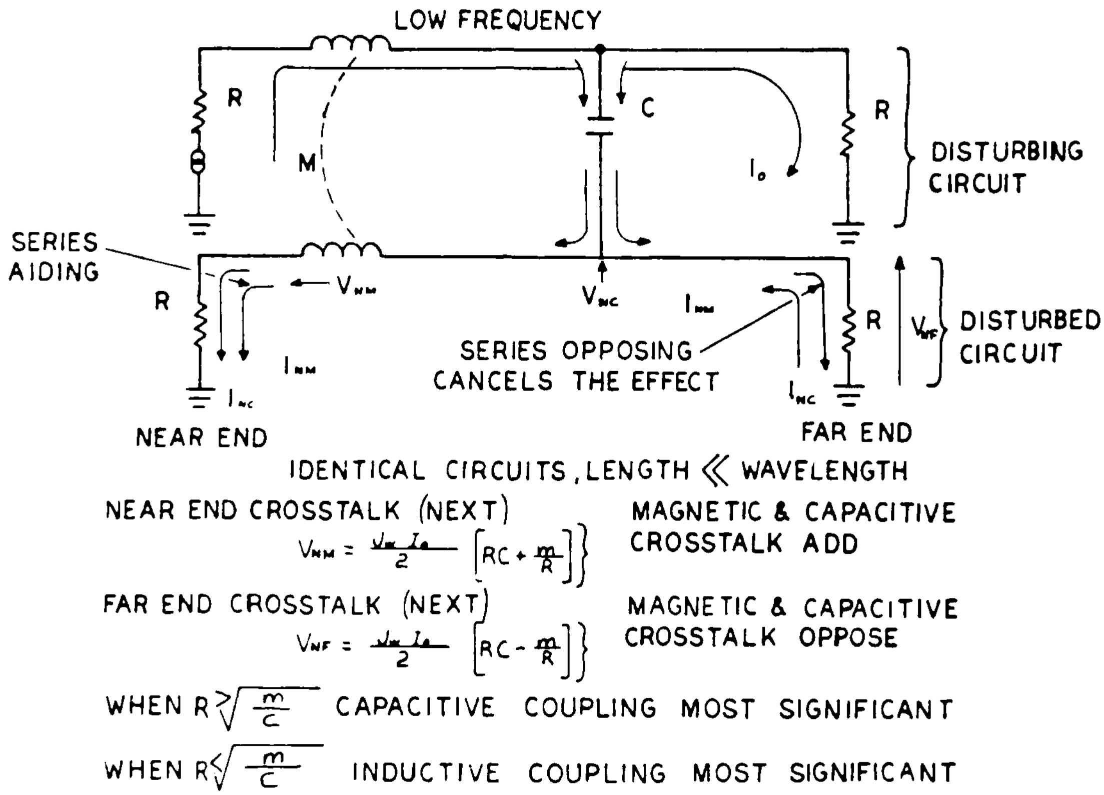


FIGURE 18. LOW FREQUENCY CROSSTALK

As will be shown in the next section, radiated emission from a single trace can be reduced by using wide traces kept close to ground.

Simply increasing the distance between traces will also reduce coupling. Running traces at right angles to each other is also very effective in minimizing crosstalk.

Another technique is to run guard traces parallel to, and on both sides of a critical trace. It has been observed that if the spacing of the guard traces is kept equal to their height above ground the effect on the characteristic impedance is quite small; about 4 percent for a 50 ohm line.

BOARD LEVEL RADIATED EMISSION Radiated emissions from PC board traces are a persistent source of interference. Clocks and their associated fan-outs are particularly bad offenders, because of the high frequencies and fast rise times which are involved in the production of timing pulses.

Radiated emissions from board traces are a function of trace width, height above ground, length, clock frequency, and pulse rise time. Scaling laws for these variables have been derived from EMCad® software. EMCad is a group of programs for EMC analysis based on a rigorous theoretical approach and on laboratory measurements of many different systems.

EMCad data was taken over the following range of variables, in order to derive the scaling laws:

Trace width:	0.007 to .050 inches
Height above ground:	0.004 to .050 inches
Clock frequency:	1 to 40 MHz
Rise time:	2.5 to 10 nanoseconds
Wave form:	Symmetrical trapezoid
Trace length:	0.5 to 6 inches

EMCad predicts board level radiation in relation to regulatory agency specification limits, and sample data is shown in Figures 19 and 20. The regulatory limits can be altered to match any desired specification. Once the predicted or measured level is known, the scaling laws can be used to alter one or more of the variables to bring a particular design into specification.

The scaling laws are useful in a more general sense, as well, to indicate how emission and susceptibility change as a function of length, height, etc.

EMCAD RADIATED EMISSION   ANALYSIS 4   SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF RTCA DO-160B

COMPANY: CKC
PROJECT: FAA DIGITAL SYSTEMS DESIGN
SIGNAL NAME: CLOCK HARMONIC SCALING
DATE (D/M/Y): 4/15/87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2.5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 2 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 1 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 1
DISTANCE BETWEEN SIGNAL AND RETURN: .015 INCHES
TRACE WIDTH: .013 INCHES
DIELECTRIC CONSTANT: 4.5
DISTANCE TO GROUND OR GROUND PLANE: .015 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METER

PREDICTED EMISSION LEVEL

FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
2.0 MHZ	48	51	54	40	14 OUT
25.0 MH	37	40	43	35	8 OUT
127.3 MHZ	30	33	36	46	-10

HIGH RISE CUT OFF FREQUENCY = 173.3 MHZ
NOMINAL CUT OFF FREQUENCY = 244.8 MHZ
WORST CASE CUT OFF FREQUENCY = 345.8 MHZ
LARGEST RECOMMENDED HOLE SIZE = .4337815 METERS

FIGURE 19. RADIATED EMISSION ANALYSIS, 2MHz TRAPEZOID

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF RTCA DO-160B

COMPANY: CKC
PROJECT: FAA DIGITAL SYSTEMS DESIGN
SIGNAL NAME: CLOCK HARMONIC SCALING
DATE (D/M/Y): 4/15/87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2.5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 3 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 1 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 1
DISTANCE BETWEEN SIGNAL AND RETURN: .015 INCHES
TRACE WIDTH: .013 INCHES
DIELECTRIC CONSTANT: 4.5
DISTANCE TO GROUND OR GROUND PLANE: .015 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METER

PREDICTED EMISSION LEVEL

FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
3.0 MHZ	50	53	56	39	16 OUT
25.0 MHZ	40	43	46	35	11 OUT
127.3 MHZ	33	36	39	46	-7

HIGH RISE CUT OFF FREQUENCY = 260.0 MHZ
NOMINAL CUT OFF FREQUENCY = 367.2 MHZ
WORST CASE CUT OFF FREQUENCY = 518.7 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2891879 METERS

FIGURE 20. RADIATED EMISSION ANALYSIS, 3 MHz TRAPEZOID

Scaling Laws. Width of trace: As the trace width decreases, radiated emission increases by

$$dB = 20 \log (w_2/w_1)^2$$

This expression is valid at least up to w = 0.05 inch and heights up to 0.03 inch.

Height above ground: As the height above ground increases, radiated emission increases by

$$dB = 20 \log (h_2/h_1)^2$$

This expression is valid at least up to h = 0.03 inch and w = 0.013 inch.

Length of trace: As the length of the trace increases, the radiated emission increases by

$$dB = 20 \log (L_2/L_1)^{1/2}$$

If the number of fan-out traces from a clock is doubled, the effect is the same as doubling the length of a single trace, 3 dB.

Clock Frequency:

(a) For a change in clock fundamental frequency, and measurement of radiated emission at the fundamental, the radiation increases with an increase in frequency by

$$dB = 20 \log (f_2/f_1)^{1/2}$$

(b) For a comparison of a clock harmonic with its fundamental, the harmonic is weaker by

$$dB = 20 \log n^{1/2}$$

where n is the harmonic number.

(c) For a comparison of the harmonics of clocks having different fundamental frequencies, for example the 30th harmonic of a 1 MHz clock with the 10th harmonic of a 3 MHz clock,

$$dB = 20 \log N_2/N_1$$

where N is the harmonic number for each separate clock.

In this example

$$dB = 20 \log \frac{30}{10} = 9.5 \text{ dB}$$

The 30th harmonic of the 1 MHz clock is weaker than the 10th harmonic of the 3 MHz clock.

Rise Time: Over a range of 1 to 10 nanoseconds rise time, and for clock frequencies of 1 to 30 MHz, there is little difference in harmonic content and consequent radiated emission for a symmetrical trapezoid until rather high frequencies, e.g. 470 MHz are reached. At these frequencies, the radiated emissions are typically well below any of the applicable specifications. It may be of some use to note that at these high frequencies, the rise time scaling law is

$$dB = 20 \log t_1/t_2$$

For each of Figures 21 through 26, a number of other parameters were involved, and these are listed here:

For Figure 21:

Waveform Rise/Fall Time: 10 nanoseconds
Waveform Frequency of Oscillation: 7 MHz
Waveform Amplitude: 5 volts
Length of Traces: .05 inches
Type of Return: Ground plane
Number of Traces: 2
Distance Between Signal and Return: .03 inches
Dielectric Constant: 4.5
Distance to Ground or Ground plane: .03 inches
Test (Measurement) Distance: 10 Meters

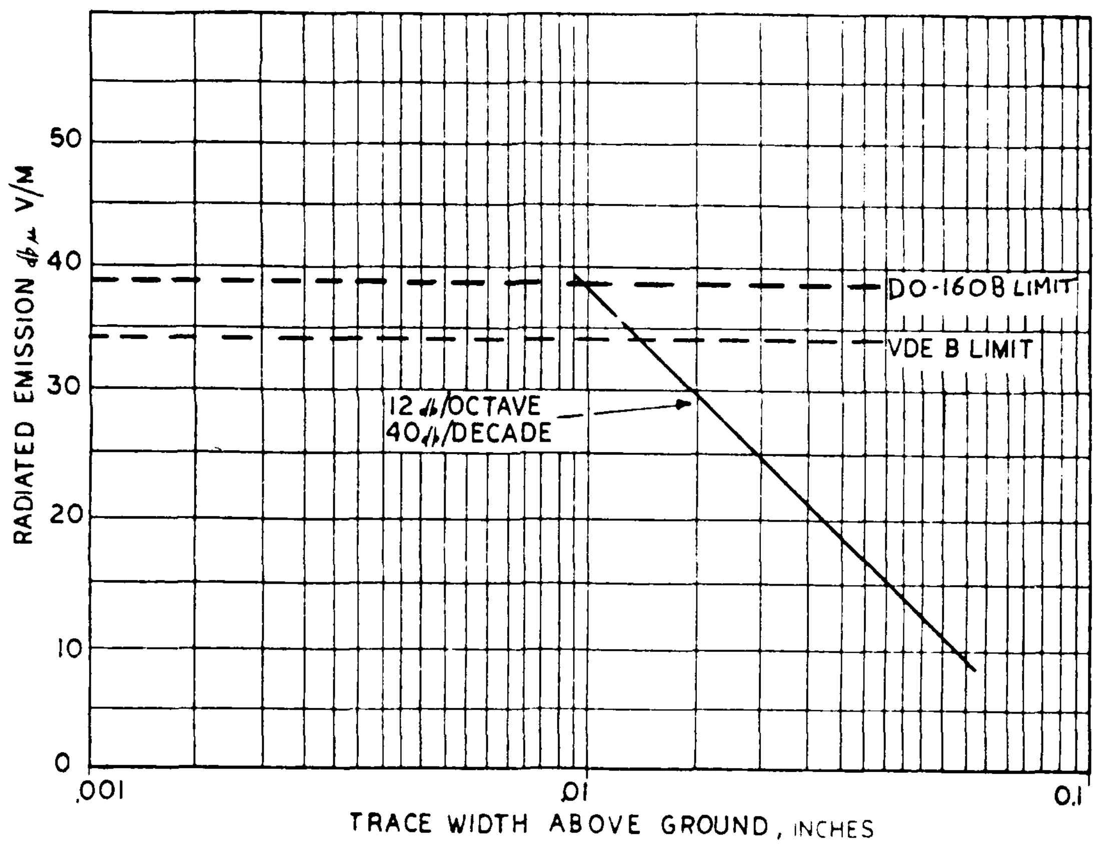


FIGURE 21. RADIATED EMISSION VS. TRACE WIDTH

For Figure 22:

Waveform Rise/Fall Time: 10 nanoseconds
Waveform Frequency of Oscillation: 7 MHz
Waveform Amplitude: 5 volts
Length of Traces: .05 inches
Type of Return: Ground plane
Number of Traces: 2
Trace Width: .013 inches
Dielectric Constant: 4.5
Test (Measurement) Distance: 10 Meters

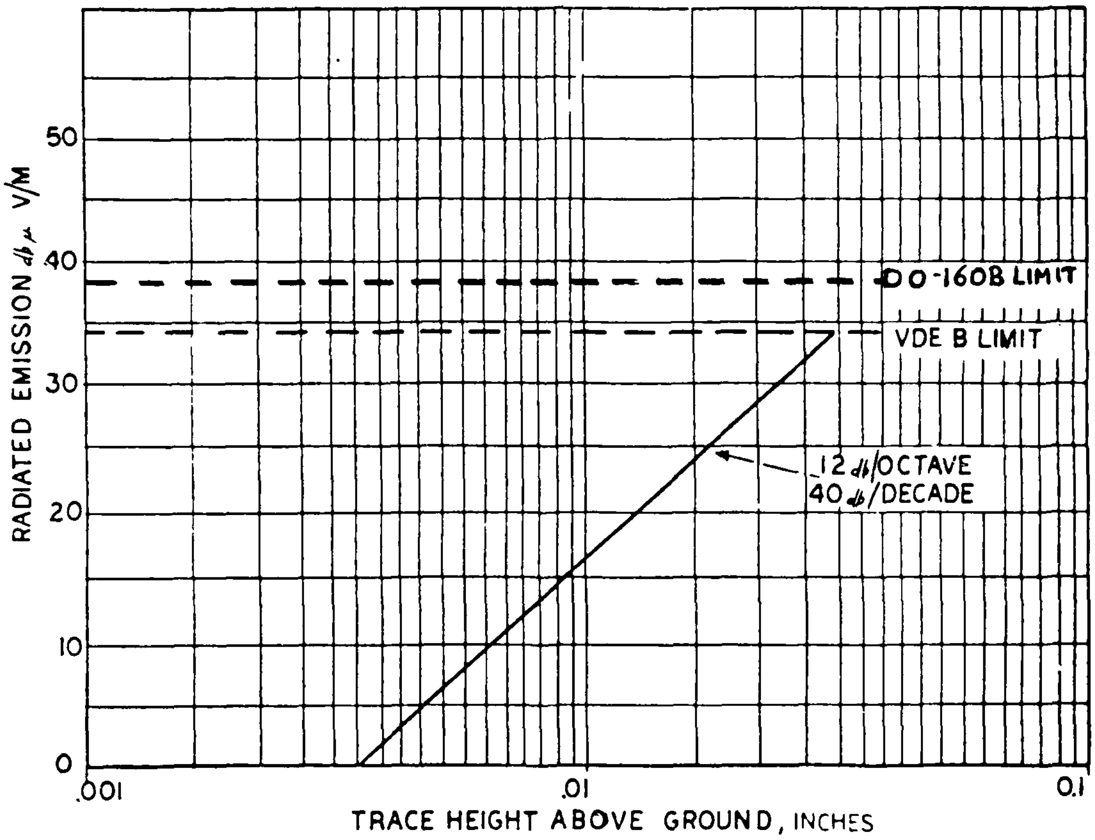


FIGURE 22. RADIATED EMISSION VS. TRACE HEIGHT

For Figure 23:

Waveform Rise/Fall Time: 10 nanoseconds
Waveform Frequency of Oscillation: 10MHz
Waveform Amplitude: 5 volts
Type of Return : Ground plane
Number of Traces: 2
Distance Between Signal and Return: .0075 inches
Trace Width: .013 inches
Dielectric Constant: 4.5
Distance to Ground or Ground plane: .0075 inches
Test (Measurement) Distance: 10 Meters

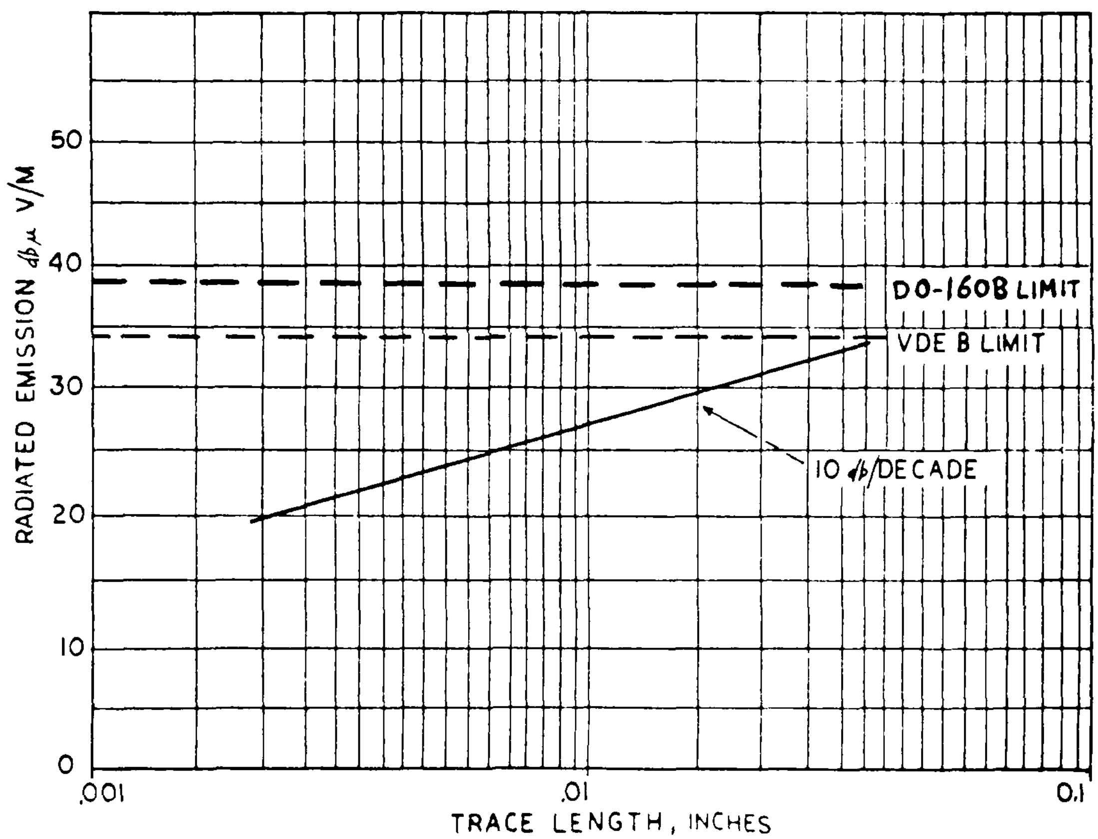


FIGURE 23. RADIATED EMISSION VS. TRACE LENGTH

For Figure 24:

Waveform Rise/Fall Time: 10 nanoseconds
Waveform Amplitude: 5 volts
Length of Traces: .05 inches
Type of Return : Ground plane
Number of Traces: 2
Distance Between Signal and Return: .03 inches
Test (Measurement) Distance: 10 Meters

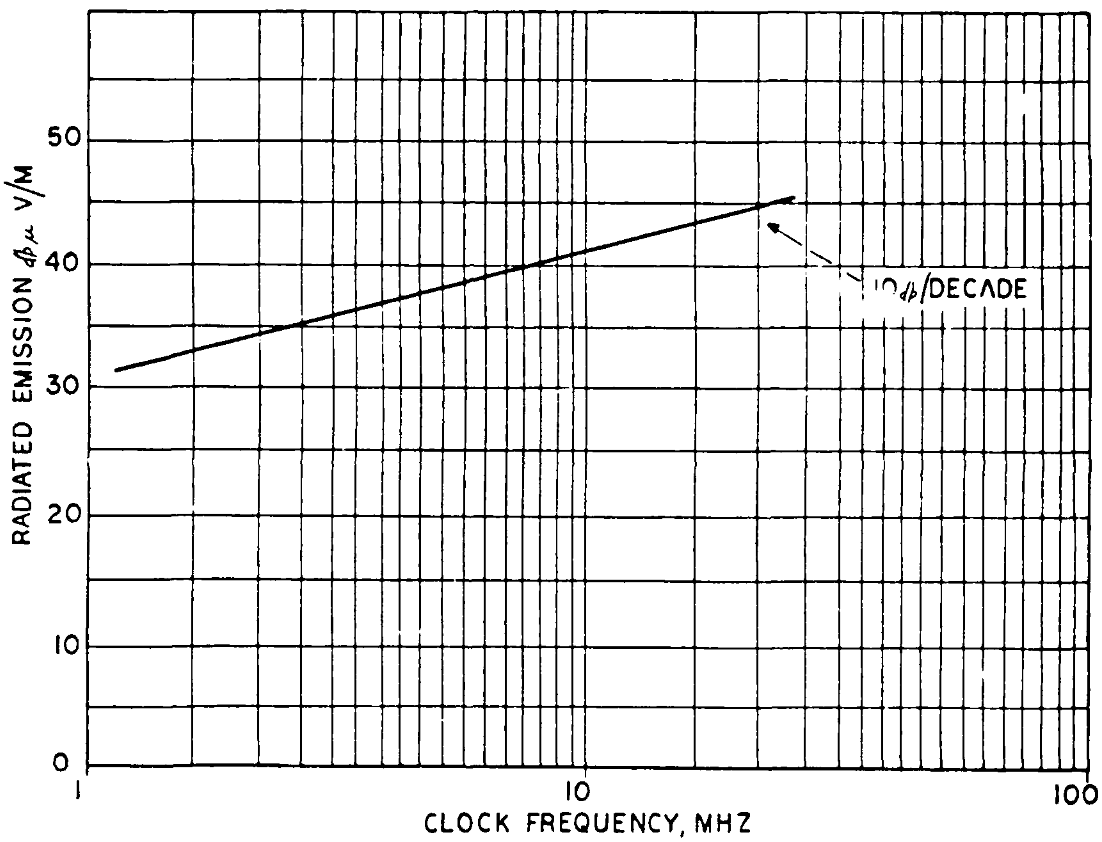


FIGURE 24. RADIATED EMISSION VS. CLOCK FREQUENCY

For Figure 25:

Waveform Rise/Fall Time: 2.5 nanoseconds
Waveform Frequency of Oscillation: 1 MHz
Waveform Amplitude: 5 volts
Length of Traces: 1 inch
Type of Return : Ground plane
Number of Traces: 1
Distance Between Signal and Return: .015 inches
Trace Width: .013 inches
Dielectric Constant: 4.5
Distance to Ground or Ground plane: .015 inches
Test (Measurement) Distance: 10 meters

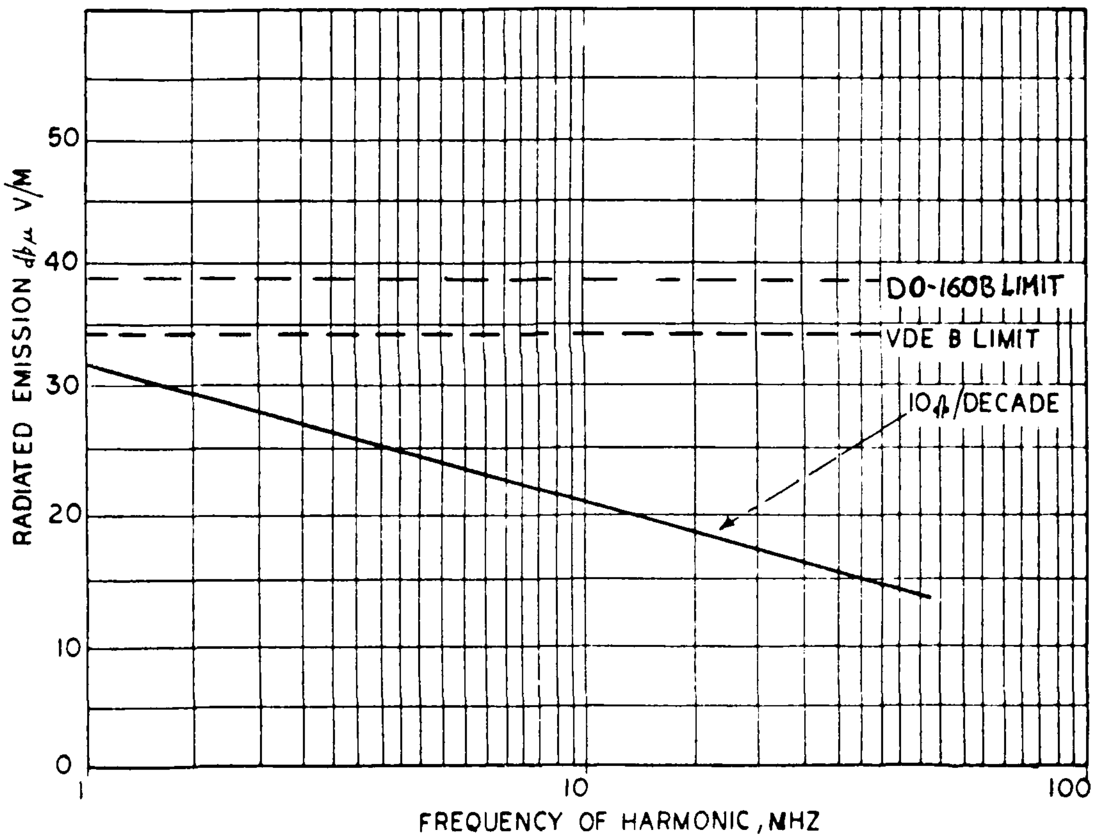


FIGURE 25. RADIATED EMISSION FROM HARMONICS OF 1 MHz CLOCK

For Figure 26:

Waveform Rise/Fall Time: 2.5 nanoseconds
Waveform Amplitude: 5 volts
Length of Traces: 1 inch
Type of Return : Ground plane
Number of Traces: 1
Distance Between Signal and Return: .015 inches
Trace Width: .013 inches
Dielectric Constant: 4.5
Distance to Ground or Ground plane: .015 inches
Test (Measurement) Distance: 10 Meters

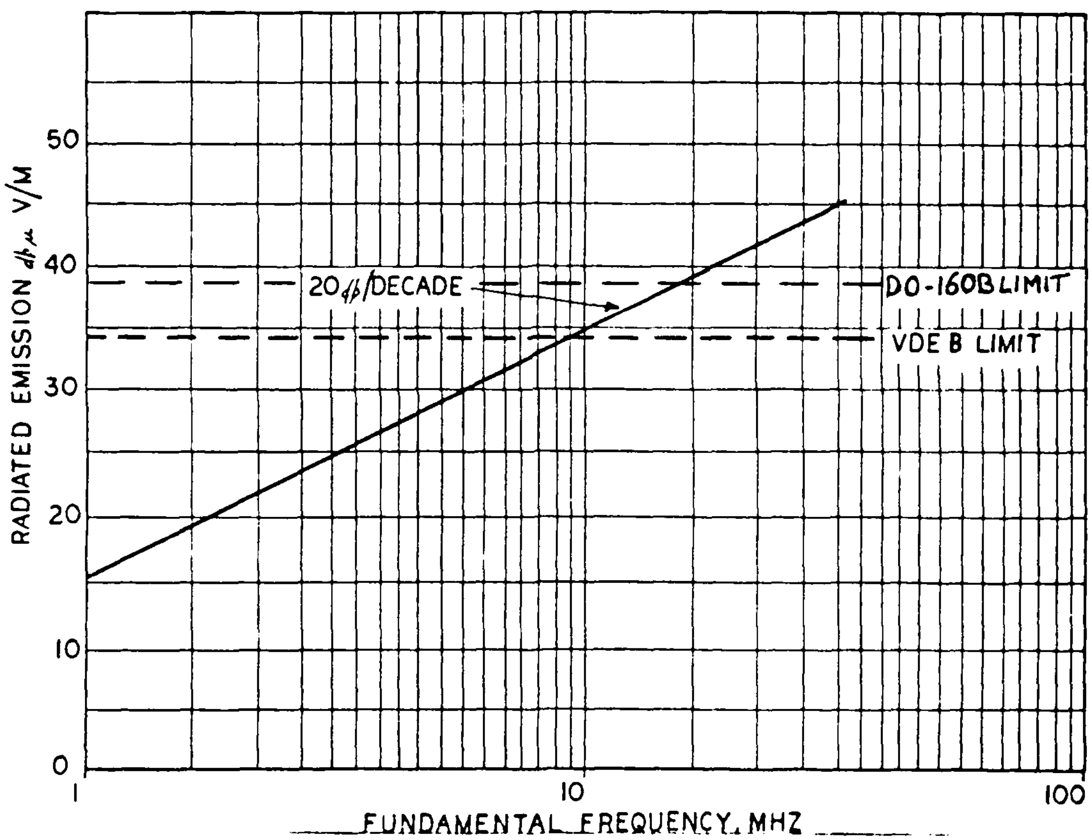


FIGURE 26. RADIATED EMISSION, 30 MHz CLOCK HARMONIC VS FUNDAMENTAL FREQUENCY

Range of Validity for Scaling Laws In general, the scaling laws fail when the characteristic impedance of the microstrip line starts rising rapidly. For a dielectric constant of 4.5 this happens at an impedance of about 125 ohms, or a width to height ratio of 0.3.

For width to height ratios greater than 3, little is gained in reducing radiated emission by further width increases. Furthermore, for excessively wide traces, it can become difficult to drive the line because of high capacitance per unit length, particularly if the line is terminated in high impedance.

Accordingly, the practical range of width to height is from 0.3 to 3.

The frequency scaling laws begin to fail for width to height ratios greater than 2.

The length scaling laws appear valid for any length up to 6" for clock frequencies up to 100 MHz.

Impedance Control. From the scaling law data, it has been observed that for a constant ratio of trace width to trace height, the emitted radiation is constant. A constant width to height ratio represents a constant characteristic impedance, when the trace and its ground plane are treated as a radio frequency transmission line.

It is helpful to consider all potentially radiating traces as transmission lines. Changes in trace width should be avoided, since this causes a change in characteristic impedance. These traces should also be terminated in their characteristic impedances to avoid standing waves. Either a load mismatch or a change in characteristic impedance will cause standing waves to appear on the trace, leading to enhanced radiation.

Analytic determination of the characteristic impedance of a microstrip transmission line is extremely complex. The characteristic impedance, Z_0, is a function of dimensions, dielectric constant and frequency, and the effective dielectric constant is itself frequency dependent.

Plots of Z_0 for various dielectric constants and ratios of w/h are given in Reference Data for Engineers. Numerous papers exist in the literature, several of which are listed in the bibliography. Also, various computer programs exist for the design of microstrip transmission lines. Several formulas for Z_0 are given in the propagation section of this guideline.

Other Considerations On double-sided boards which are notorious for radiated emissions since they lack ground planes, the clock trace and its return can be constructed as a twisted pair by periodically running each trace through a plated-through hole, and crossing it over its mate.

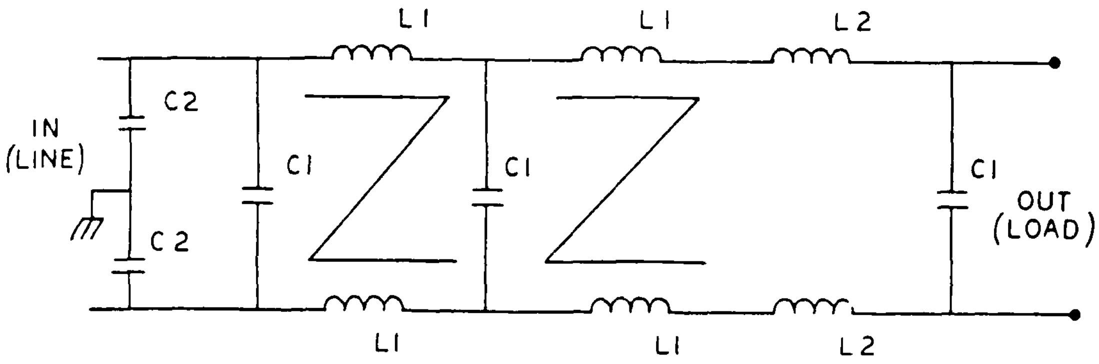
Another technique for reducing board level emissions is to use ground-fill techniques, in which vacant areas of the board are filled with a grid of ground traces. This technique has reduced emissions by as much as 12 dB.

BOARD LEVEL CONDUCTED EMISSION. Conducted susceptibility and emissions are the result of radio frequency energy arriving or departing via power cables or I/O cables. Conducted interference is not a significant part of the high energy threat, since the radiation effects completely overshadow it in most cases. The methods of measurement of conducted susceptibility are described in RTCA DO-160B, Section 20.

Conducted interference is ordinarily controlled by filtering of the power cables at the point of entry to the equipment. In some cases, shielding of the input power cables is also required.

Conducted problems can appear as either differential mode (line to line) or common mode (line to ground). Experience has shown that the differential mode appears up to about 150 kHz. Above this frequency, the common mode predominates.

Elimination of conducted interference is accomplished by common mode and differential mode filtering. A typical powerline filter design is shown in Figure 27.



C1 1.0μf BIPOLAR
C2 .0047μf CERAMIC DISK CAPS, 1KV RATED

L1 6mh COMMON MODE CHOKE
L2 OPTIONAL 500μh AT RATED LOAD INDUCTOR

FIGURE 27. TYPICAL POWERLINE FILTER DESIGN FOR SWITCH MODE POWER SUPPLY

I/O cables can also carry conducted interference. Proper grounding of cable shields as detailed in the cabling section of this report is important. Also, shields of multiconductor cables should never be used as returns for any signal circuit. It may be necessary to filter individual signal wires which are carrying emissions or are susceptible. Ferrite toroids surrounding the cable shield can be used to increase the shield impedance, with resulting decreases in both emission and susceptibility.

MULTILAYER BOARDS. Multilayer boards provide a high level of component packaging density. They also provide important advantages in the control of electromagnetic interference. However, the choice of multilayer construction should not be based solely on EMI considerations, since adequate levels of control are possible with single and two-sided boards.

The cost of multilayer boards is higher. Repair is difficult or impossible. However, if multilayer design is chosen, significant EMI control can be obtained by proper attention to stacking order, through the use of plated-through hole interconnections which can greatly shorten lead lengths, and by appropriate orientation of the traces on one layer to those on adjacent layers.

Ground planes, power planes, and power return planes can all be used to provide shielding and isolation between signal planes.

As a starting point in the design, each signal should be placed in one of the following categories:

. I/O AC power
. I/O signals - digital
. I/O signals - analog
. I/O signals - discrete
. I/O signals - video
. Internal digital
. Internal high current deflection
. Internal analog
. Internal discrete
. Internal video
. Internal D.C. power
. Internal test signals

It is of great importance to keep I/O signals isolated from high speed internal signals for control of both emission and susceptibility.

Whenever possible, signals in each category above should be placed on separate isolated layers.

When more than four levels are used, the stacking order becomes important in achieving isolation among the various categories of signals.

Figure 28 illustrates a possible stacking order for a complex board.

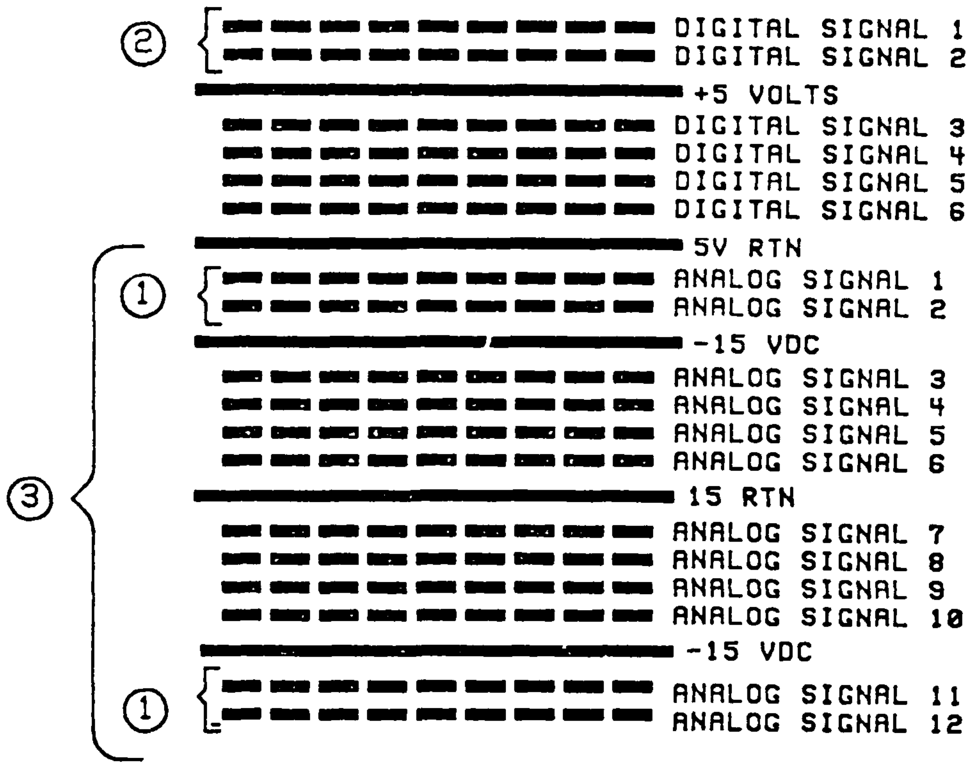


FIGURE 28. MULTI-LAYER BOARD STACKING ORDER

In Figure 28, the following practices, indicated by the circled numbers, are observed:

1. For analog signals, no signal layer can be more than two layers away from a supply, return, or reference plane.

2. Only two digital signal layers are allowed since no reference plane exists above the Digital Signal No. 1 plane.

3. No digital signal planes are allowed between a digital and an analog plane. However, limited bandwidth high level analog planes are allowed between other analog planes.

In this multilayer board design, the +5 V, 5V RTN, +15 V, -15 V and 15 V RTN are all distributed as complete planes. These, planes, when properly decoupled to the ground reference plane, act as shield and reference planes themselves.

The 5 V and 5 V RTN planes are associated with the digital signals, while the $\pm$ 15V and 15 V RTN are analog related.

Each signal layer is oriented at right angles to the preceding layer.

The maximum number of layers between two reference planes is four, so that no signal layer is ever more than two layers away from a reference plane.

The fastest rise/fall time signals (usually the clock) should be on the first layer from either the +5 V or 5 V RTN planes. If possible these signals should be buried between the +5 V and 5 V RTN planes, rather than carried on a top or bottom layer.

Figure 29 illustrates the distribution of clock signals.

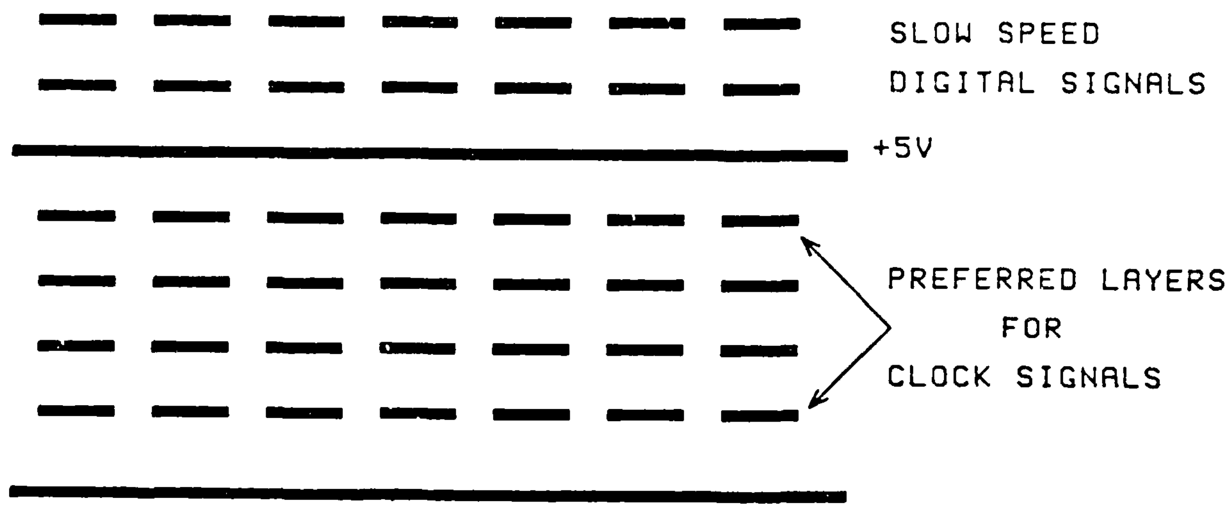


FIGURE 29. THE DISTRIBUTION OF CLOCK SIGNALS

Video signals require special handling because of their high levels and high frequency components. These signals can be routed as parallel adjacent traces on a signal plane, or as a parallel stack between two planes. This is a special case where two adjacent planes are not at right angles to each other.

For video, it is preferable that the surrounding planes be ground reference rather than power distribution planes, to minimize crosstalk.

Figure 30 illustrates the distribution of video signals.

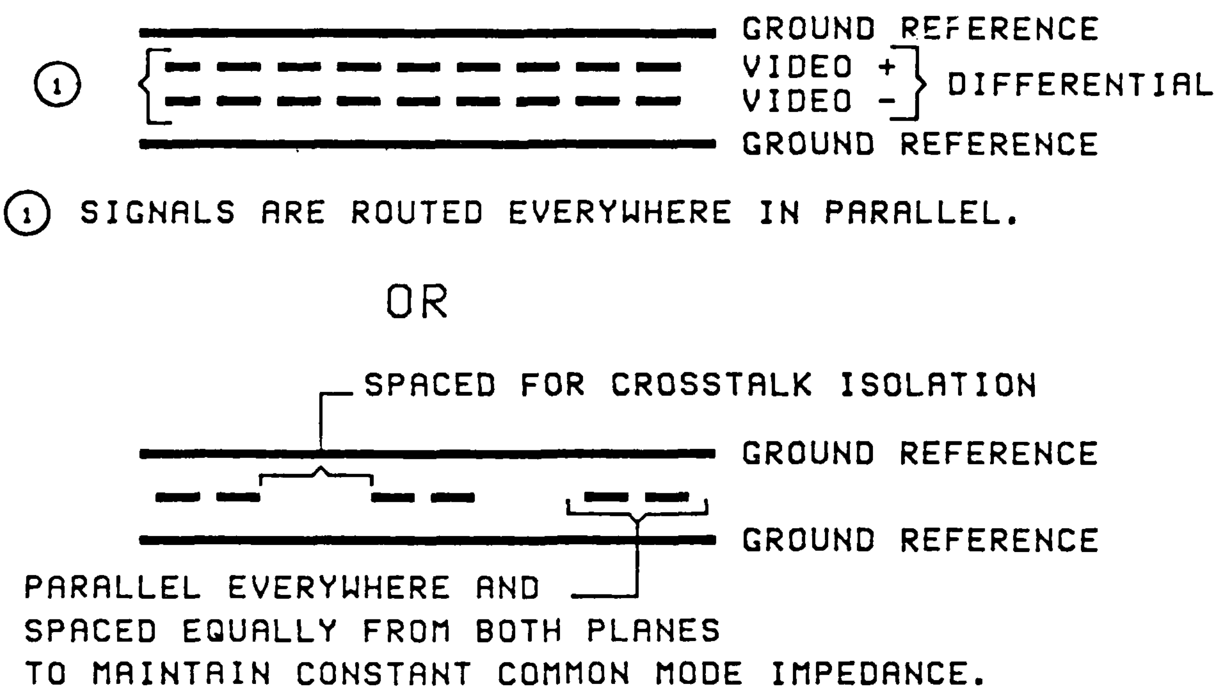


FIGURE 30. VIDEO SIGNAL DISTRIBUTION

On multilayer boards, all signal traces should be kept away from the edge of the board by a distance equal to at least the height above the nearest reference plane. This technique will reduce emission from the edges of the board.

A significant fraction of radiation from boards comes directly from the dual-inline-packages themselves. Obviously, the radiation from any DIPs mounted on an outside layer will not be reduced by multilayer construction.

BACKPLANES AND MOTHER BOARDS Nearly all of the principles applicable to PC board layout also apply to the layout of backplanes and mother boards.

A distinguishing feature of backplanes and mother boards is that they are significantly larger than the individual circuit boards. The length of traces increases, enhancing the possibility of radiated emissions and increased susceptibility. Long, parallel traces increase the chance of crosstalk among different circuits. With incorrect layout, large loop areas can be created, giving the opportunity for magnetic coupling among large co-planar loops on the board.

On double-sided boards, coupling of co-planar loops can be minimized by running the traces on one side of the board at right angles to those on the other side.

The use of ground planes on both backplanes and mother boards is helpful in reducing emissions and decreasing susceptibility. The groundplane can act as a shield, in addition to the shielding provided by the equipment enclosure.

On backplanes not using a ground plane, a ground screen can be constructed as shown in Figure 31.

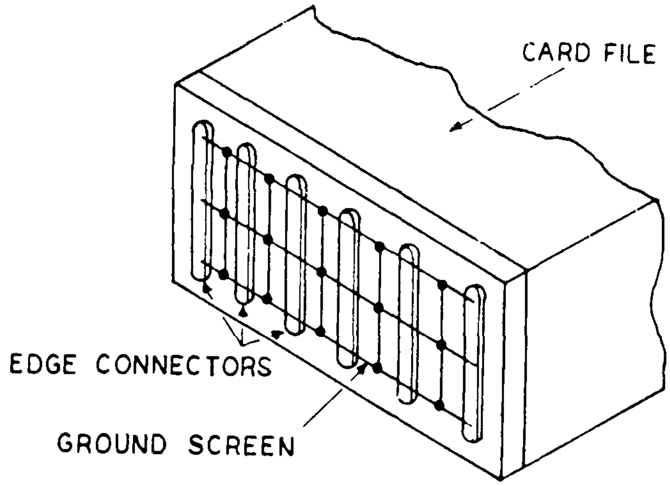


FIGURE 31. GROUND SCREEN CONSTRUCTION

A feature of the ground screen is that numerous ground connections are provided for each edge connector. On critical boards it may be advantageous to provide even more ground connections to the edge connector.

Sometimes it may be necessary to use a shielded twisted pair, or a coaxial line to replace traces. In this case, the shields should be handled as described in the cabling section of this report.

The production costs of installing a coaxial line are obviously much higher than for a trace plated on a board, but for high frequency clock distribution lines, there may be no alternative. Coaxial cables are readily available in characteristic impedances of 50, 75 and 95 ohms. Propagation delays are typically 1.7 ns per foot. On critical traces, such as clock distribution lines, care should be taken to avoid sharp bends or changes in trace width, which cause changes in characteristic impedance, with reflections and consequent radiation. Similarly, attention should be paid to impedance matching where a coaxial cable joins a trace.

If a multilayer backplane is used, critical transmission lines can be handled as stripline, in which the signal trace is sandwiched between two groundplanes.
The power distribution scheme on a backplane should incorporate frequent bypassing. A good practice is to alternate capacitors with values which are 2 orders of magnitude different in value, for example 0.1 μf and 0.001 μf. This subject is covered in the bypassing and filtering sections.

The connectors used to connect boards to mother boards and backplanes should be shielded and should maintain the integrity of shields from board to board. Connectors incorporating built-in bypass capacitors are available and are sometimes helpful in reducing emissions. On a portion of the backplane, or mother board or on the chassis plate below the backplane, an area should be left for the addition of connectors, filtering and bypassing, in case modifications become necessary. The mother board should provide a single central reference ground for D.C. power return, shielding between signal layers, and daughter board returns.

To accomplish this, all return and shield planes within a mother board should be bonded together at every mounting hole and at every connector pin that is used for returning D.C. to the central reference.

Isolation within a mother board can be achieved in three ways:

1. spacing between adjacent traces;
2. addition of guard traces between adjacent traces; and
3. addition of shield planes between stacked traces.

CABLES, CONNECTORS AND METHODS OF TERMINATION. I/O cables and power cables are an important source of emission and susceptibility problems in digital systems. A significant fraction of EMC problems arise from cables. Increased inspection of cables, connectors and methods of termination could reduce this source of trouble in the future.

The causes of radiated and conducted interference and susceptibility from cables are many:

(1) improper types of cables
(2) incorrect grounding in multiconductor cables
(3) no shielding
(4) incorrect connections of shields
(5) defective workmanship in installation
(6) cable deterioration with age
(7) inadequate bypassing and filtering at the cable connector
(8) improper connection of the load
(9) cable resonance from improper termination or shielding

The majority of system cables are of the multiconductor type. Coaxial cables are usually used for transmission of radio frequency energy to antennas. In the future more coaxial cables will probably be used to meet EMC requirements and because with higher data rates, data buses are appropriately treated as radio frequency transmission lines.

In addition to multiconductor and coaxial cables, many other types are also used:

(1) unshielded
(2) twisted pair
(3) tri-lead
(4) shielded twisted pair
(5) triaxial
(6) ribbon

Some of these cable types are shown in Figure 32.

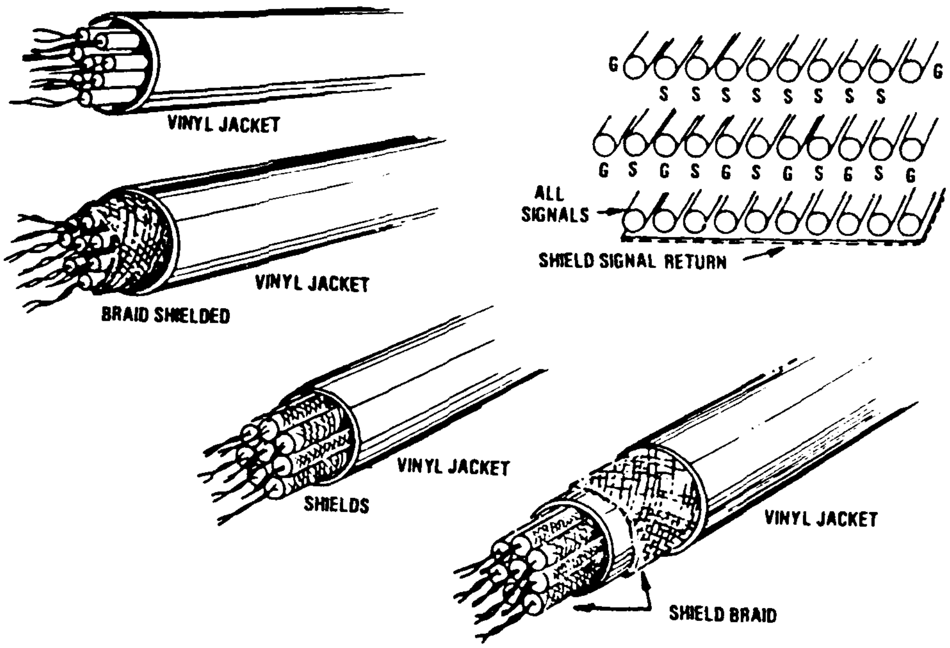


FIGURE 32. I/O CABLE CHOICES

Equivalent Circuit of a Coaxial Cable. The equivalent circuit of a coaxial cable as shown in Figure 33 will now be analyzed. This analysis will also be applicable in most respects to the understanding and proper use of other types of cables.

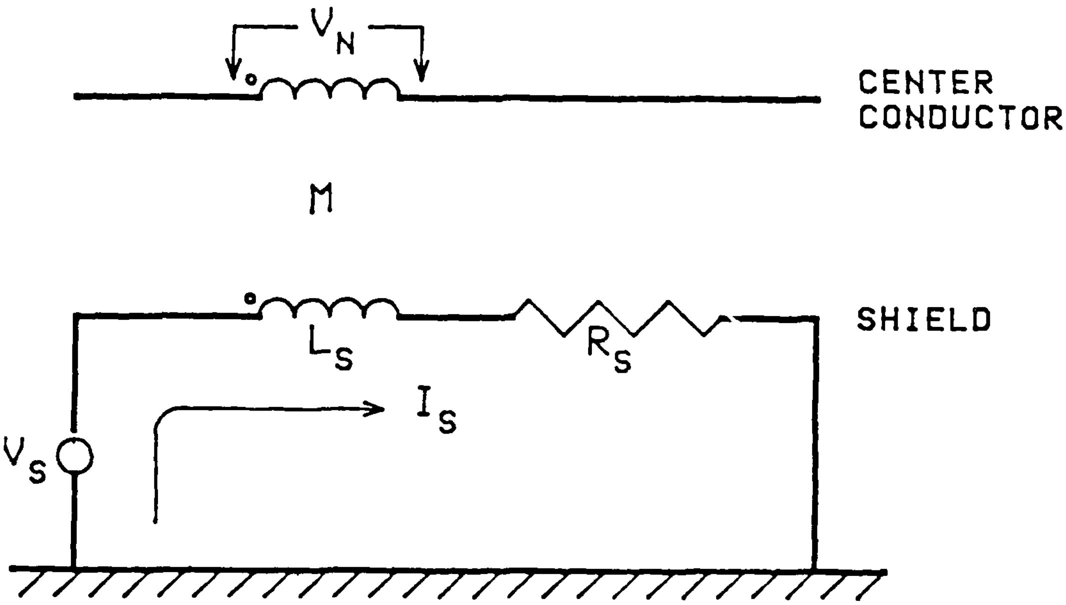


FIGURE 33. EQUIVALENT CIRCUIT OF A SHIELDED CABLE

In Figure 33 V_N is a noise voltage induced on the center conductor by a shield current, I_S. V_S is a source of noise voltage driving the shield. L_S is the self inductance of the shield, and R_S is the series resistance of the shield. M is the mutual inductance between the shield and the center conductor.

Since all of the lines of flux surrounding the shield also surround the center conductor, it can be shown that $M = L_S$, and that

$$V_N = \left(\frac{j}{j + R_S/L_S}\right) V_S \qquad (1)$$

In Figure 34 V_N versus ω on a logarithmic scale is shown.

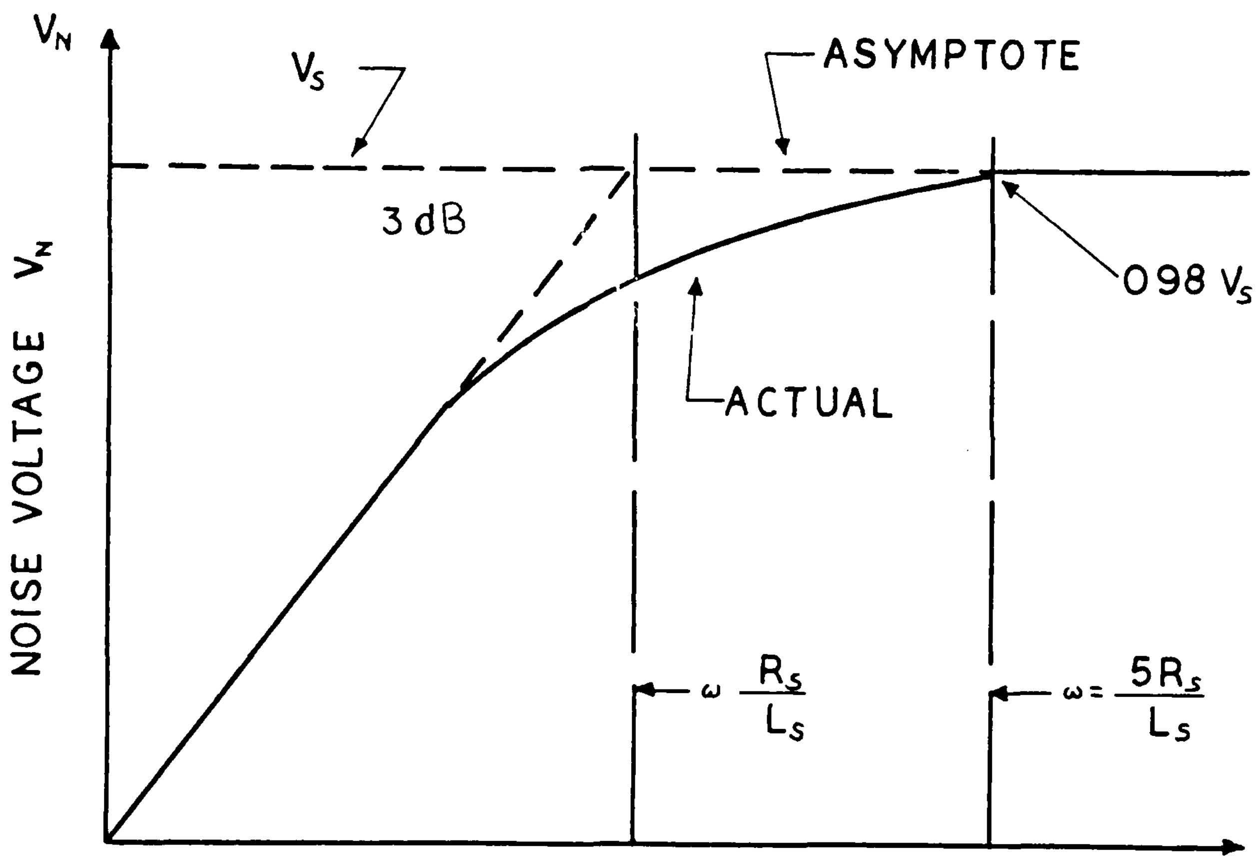


FIGURE 34. NOISE VOLTAGE IN CENTER CONDUCTOR OF COAXIAL CABLE DUE TO SHIELD CURRENT

Low Frequency Performance. The frequency at which the inductive reactance of the shield is equal to its series resistance, marks a division between the very low frequency behavior and the high frequency behavior of a coaxial line, and is called the shield cut-off frequency. This frequency is typically 1-2 kHz, and rarely higher than 7 kHz.

From Equation 1 and Figure 34 it can be seen that the noise voltage induced on the center conductor is zero at D.C. and increases almost to V_S at

$$\omega = 5\ R_S/L_S$$

For avionics digital systems, the performance of shielded cables at D.C. and audio frequencies is of little interest. Accordingly, after a few more general remarks, the balance of this discussion will be directed at the high frequency ($f > 100$ kHz) behavior of shielded cables.

The preceding analysis describes magnetic coupling effects. Shielded cables are extremely effective against electric field coupling at all frequencies, provided that the center conductor does not extend unnecessarily beyond the shield. For effective electric field shielding of cables up to 1/20 wavelength only one end of the shield needs to be grounded. This technique is often used at audio frequencies and decreases the possibility of ground loops. However, it provides no shielding against magnetic fields, and ground loops can still occur from capacitive coupling of the floating end of the shield to ground.

Single-end grounding techniques are rarely appropriate to the design of digital systems.

High Frequency Performance. At high frequencies, skin effect plays an important role in the behavior of a coaxial line. At high frequencies, current tends to flow mainly in the outer layers of a metallic conductor. The phenomena is lucidly explained in Terman's Radio Engineering (McGraw Hill, 3rd edition, pp 19-23). Briefly, some flux lines are within the conductor. The central part of a conductor is surrounded by all the lines of flux, but the outer layers are surrounded by only some of the lines. The inductive reactance of the center is higher than the outside, and most of the current then flows on the outside, where it is surrounded by the fewest lines of flux, and encounters the lowest impedance.

The result is that at high frequencies, a current flowing on the outside of a shield does not penetrate to the inside, and similarly, a current flowing on the inside does not penetrate to the outside. In effect a single shield coaxial line acts like a double shielded (triaxial) line at high frequencies. With braided shields, the isolation of outside and inside is not perfect. The apertures formed by the braid weave permit interpenetration of flux lines, and the apertures act as small mutual inductances to provide coupling. A consequence of this is that the braid weave design and the choice of materials become important in determining shielding effectiveness.

Ordinary braid weaves provide from 60 percent to 90 percent coverage of the center conductor. Copper and tin-coated braids have been found to oxidize with time and lose their shielding effectiveness because of loss of contact among the carrier wires of the braid. If a cable recovers its shielding effectiveness with flexing, oxidation can be assumed to be the problem, and the loss of shielding effectiveness will recur. Silver plated braids have significantly better resistance to oxidation.

Electric Field Shielding. As a rule of thumb, each additional shield braid adds 15-20dB of shielding. For example, a copper braid coaxial line might show on the order of 40 dB shielding effectiveness over an unshielded conductor; a triaxial cable otherwise of the same design might show 60 dB shielding effectiveness. However, in some recent experiments a single braid weave design optimized by computer has shown better performance than two conventional shield braids.

Even better performance is obtained with foil-wrapped shields, drain wires, and braid over foil. Even the manner of folding the foil back on itself for good contact can have a large effect on shielding effectiveness.

All of the preceding description of the high-frequency performance of coaxial line applies to multiconductor shielded cables as well.

Figure 35 shows the radiation field intensity from flat (ribbon) cable with several different shield designs.

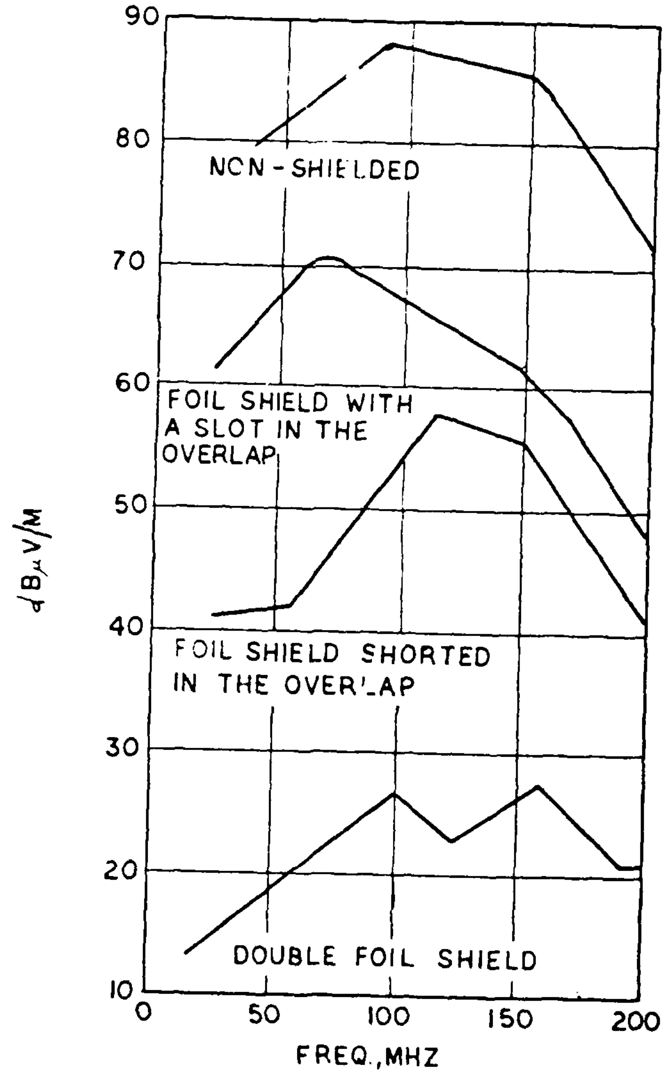


FIGURE 35. FLAT CABLE RADIATION FIELD INTENSITY

In some cases, mainly for radio frequency distribution, the choice of a coaxial cable with a solid metallic outer conductor may be appropriate.

Shield Connections. The choice of shield design can be critical to success or failure in meeting EMC requirements.

The methods by which cable shields are connected to sources, loads, and ground are critically important to shielding effectiveness.

In the past, shields of coaxial and multiconductor cables have often been terminated by the "pig tail" method, shown in Figure 36.

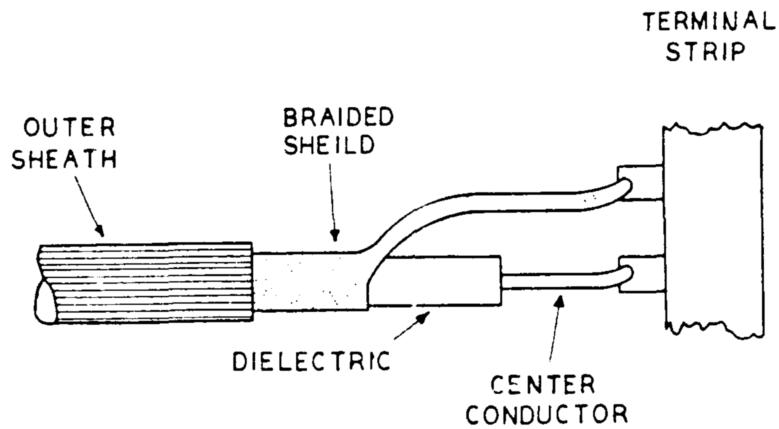


FIGURE 36. PIGTAIL TERMINATION OF A COAXIAL CABLE

This pig-tail termination method is generally unacceptable. Measurements have shown as much as a 40 to 50 dB difference in the 15 to 200 MHz region between pig-tails and a 360° termination of the cable shield to the connector backshell. Figure 37 shows a comparison of several termination methods.

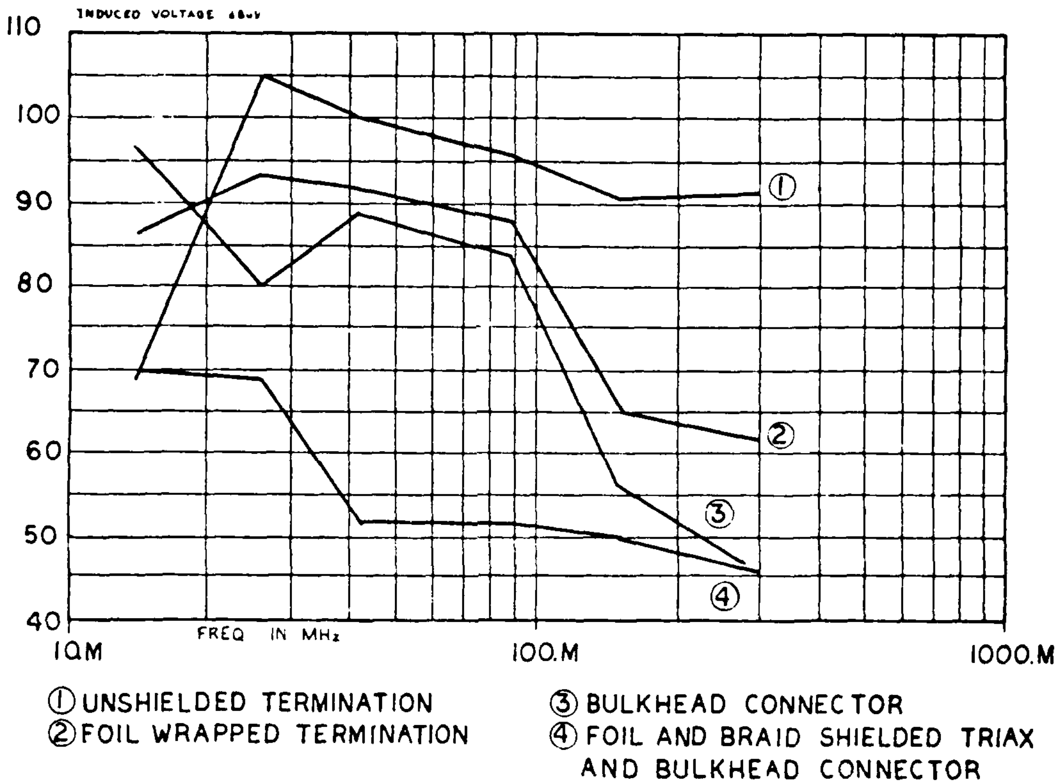


FIGURE 37. INDUCED VOLTAGE WITH VARIOUS CABLE TERMINATIONS TRIAX CABLES, 200 V/M FIELD

For most digital systems applications, the recommended method of handling cable shields is to connect each end of the cable to the cable connector backshell with a full 360° connection. The cable connector in turn mates with a bulkhead connector which has a solid metallic contact with the outer shielding case of the line replaceable unit or other digital device. This method of connection is shown in Figure 38.

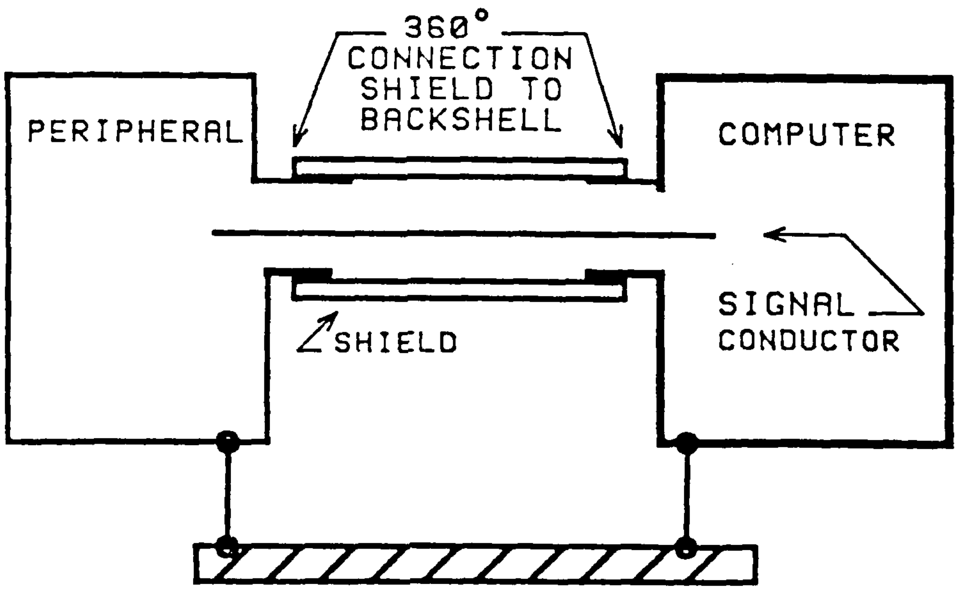


FIGURE 38. RECOMMENDED CABLE SHIELD TERMINATION METHOD

The termination method shown above can be seen to create a ground loop. In high frequency applications, this ground loop is of little significance compared to the effective magnetic field shielding that is achieved. Very little return current flows through the ground reference between the ground connection points. Most of the return current flows through the shield for the following reason (see Figure 12 in the section on Grounding): The reactance of the mutual inductance is subtractive from the impedance presented to the return current by the shield alone. So the impedance via the shield is significantly lower than any alternative path, and most of the current flows on this low impedance path.

Very little flux is produced in the loop area formed by the ground loop, and because the inside and outside of the shield are isolated by skin effect, flux lines which penetrate the ground loop area from outside magnetic field sources cannot induce voltage on the shielded center conductor.

Ground Returns for Multiconductor Cables The shield of a multiconductor cable is treated in the same way as the shield of a coaxial cable. However the shield should not be used as a return path for digital, analog or power functions. Instead one or more of the cable conductors should be used for these purposes. The best (but most costly) practice is to use at least one return conductor for every signal conductor. The mutual inductance between the pairs of conductors produces a low impedance return path just as described earlier.

Some multiconductor cables contain groups of twisted pairs. Using such a pair for signal and return gives even higher mutual inductance, lower impedance returns, and reduces the loop area between the conductors. Shielded twisted pairs in multiconductor cable are even better, but each increment of improvement exacts cost and weight penalties.

Ribbon cable can be regarded as a special class of multiconductor cable. Various methods of providing ground returns for ribbon cable are shown in Figure 39.

CROSSTALK IS AFFECTED BY:

1. FREQUENCY (RISE TIME) OF SIGNAL
2. CABLE LENGTH
3. TERMINATING IMPEDANCES OF ALL WIRES
4. GEOMETRY OF SIGNAL AND GROUND WIRES

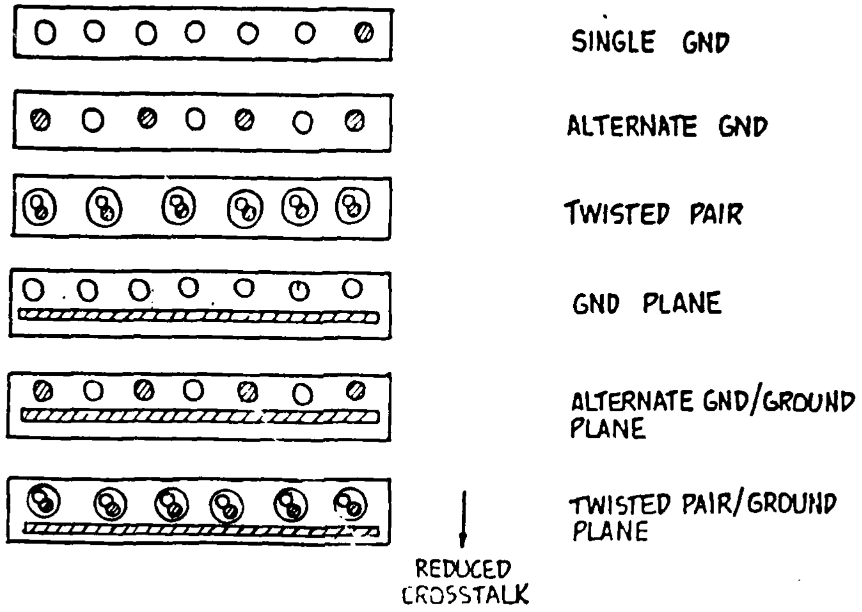
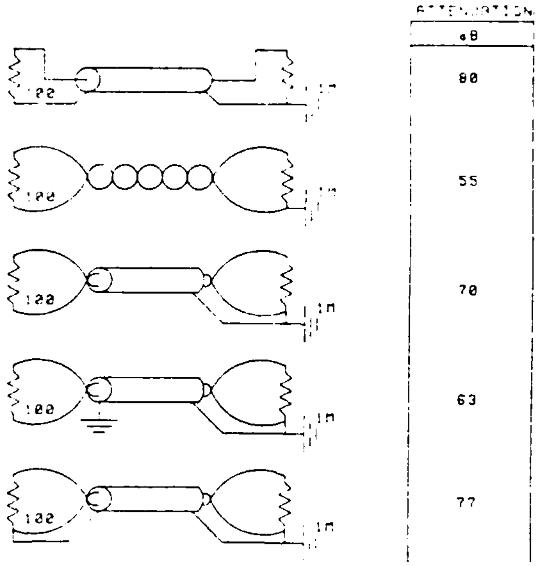


FIGURE 39. RIBBON CABLE

Shielded Twisted Pairs. One method not shown in Figure 39 is to provide two ground conductors between every active conductor. Each active signal conductor then has its own pair of shielding returns, one on either side, which it does not share with any other conductor. This technique is helpful in reducing crosstalk in ribbon connectors.

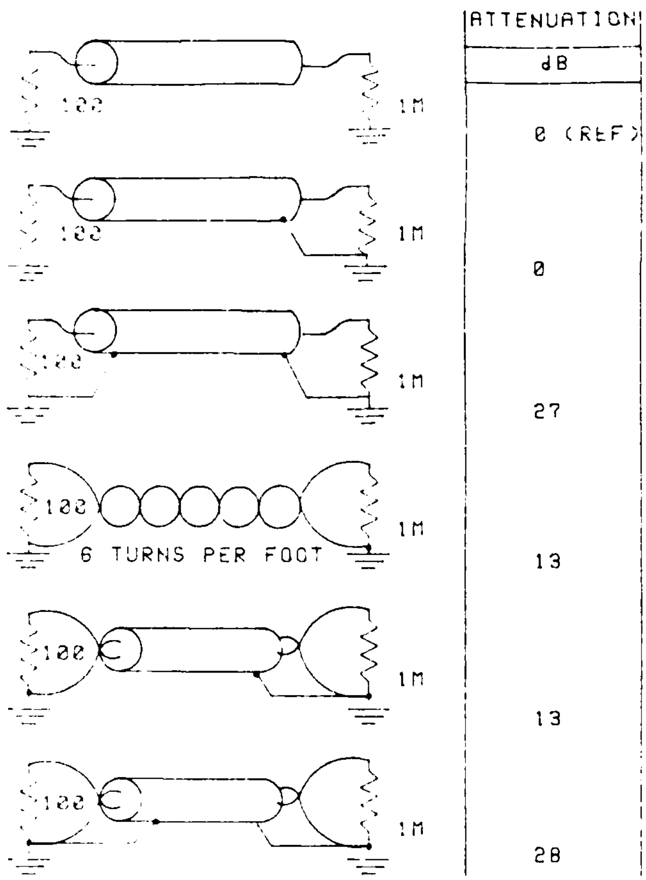
If there are unused conductors in a multiconductor cable, they should be either terminated or left open at both ends (termination is preferable), but not left open at one end. Connecting such a conductor to ground at one end turns it into an antenna at some frequency, with a resulting large increase in both emission and susceptibility.

For many applications below 100 kHz the shielded twisted pair is an effective way of handling signal and data transmission. The methods used for handling shields, terminations, and their connections to ground are, as before, critical to shielding effectiveness. In Figures 40 and 41 are shown the results of a test at 50 kHz under various termination conditions, with the measured attenuation (shielding effectivness) in dB.



FREQUENCY = 50 KILOHERTZ FOR ALL TESTS
RESULTS OF INDUCTIVE COUPLING EXPERIMENT;
ALL CIRCUITS ARE GROUNDED AT ONE END ONLY

FIGURE 40. CABLE TERMINATION METHODS - ONE END GROUNDED



FREQUENCY = 50 KILOHERTZ FOR ALL TESTS
RESULTS OF INDUCTIVE COUPLING EXPERIMENT;
ALL CIRCUITS GROUNDED AT BOTH ENDS.

FIGURE 41. CABLE TERMINATION METHODS - BOTH ENDS GROUNDED

Transfer Impedance. In the literature, the term transfer impedance is sometimes used. Transfer impedance is the ratio of voltage induced on the center conductor to the current on the outside of the shield.

Measurement of transfer impedance can be difficult, but in theory it is an effective method of comparing one cable to another. For typical coaxial cables, transfer impedances are found to be about 10 milliohms per meter of cable length.

Transfer impedance has been used primarily by the nuclear weapons community for measurements of cable susceptibility to the nuclear electromagnetic pulse. The concept has not been widely used in the EMC community. Difficulties have been described by various authors in relating transfer impedance to shielding effectiveness. Shielding effectiveness, which simply relates the emission or susceptibility of a shielded cable to an unshielded conductor, is a straightforward and repeatable method of comparing cables.

Filtering and Bypassing of Cables. General methods of filtering and bypassing are covered in other sections of this guideline. Several techniques are, however, specific to cables and their connectors.

Shielded cables can be passed through a ferrite toroid, forming a single or multiturn choke, effective at high frequencies in suppressing cable shield currents by increasing the shield impedance to currents flowing on the outside of the shield. The flow of current on the inside of the shield is unaffected.

Various manufacturers supply multiconductor cable connectors with built-in bypasses, either of the feedthrough type, or in the form of small disc ceramic capacitors. Such bypassing is often helpful in reducing cable emission and susceptibility. Caution must be exercised so that operation of the equipment is not adversely affected by the additional capacitance. It should also be noted that such capacitors can become self-resonant at frequencies of several hundred megahertz, resulting in very low line to ground impedances at the resonant frequency and ineffective bypassing above the resonance. (See Figure 44 in the Bypassing section).

Useful Frequency Ranges for Various Transmission Lines. The useful ranges of several cable types are given in Figure 42. "Twisted pair" and "shielded twisted pair" include that type of cable when incorporated into a multiconductor cable.

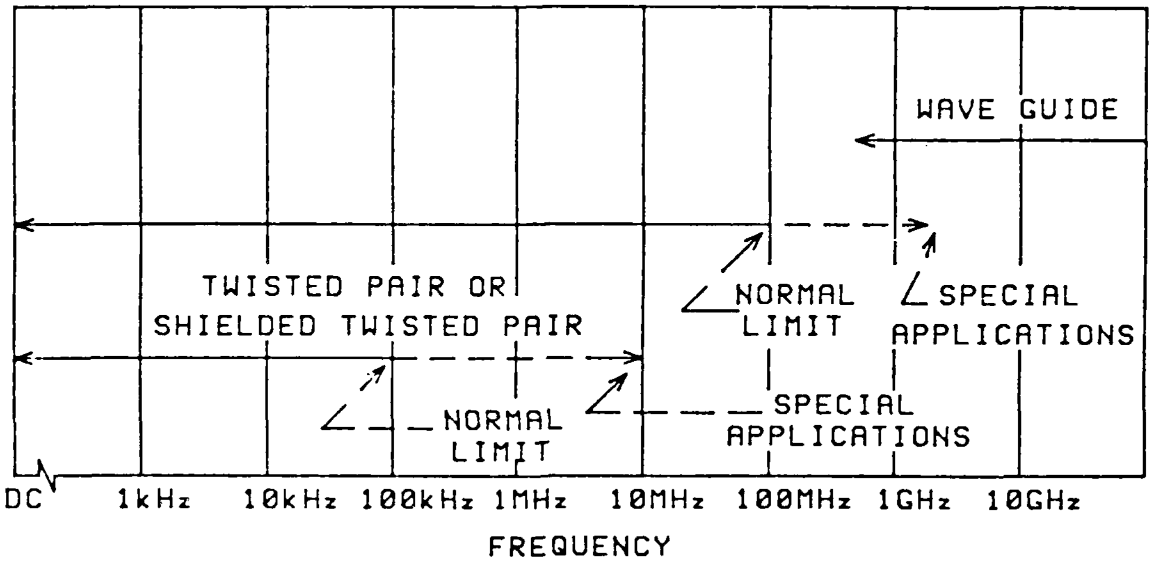


FIGURE 42. USEFUL FREQUENCY RANGE FOR VARIOUS TRANSMISSION LINES

BYPASSING. Bypassing, also referred to as decoupling, is the shunting to ground of high frequency components on a signal line by connection of a capacitor from line to ground. Because the capacitor has an increasingly low impedance at higher frequencies, high frequencies are effectively removed from the signal line, while low frequencies remain relatively unaffected.

The equivalent circuit of a capacitor is shown in Figure 43. Series and shunt losses are represented by R_1 and R_2. Series inductance is represented by L which has a very significant effect upon the performance of the capacitor.

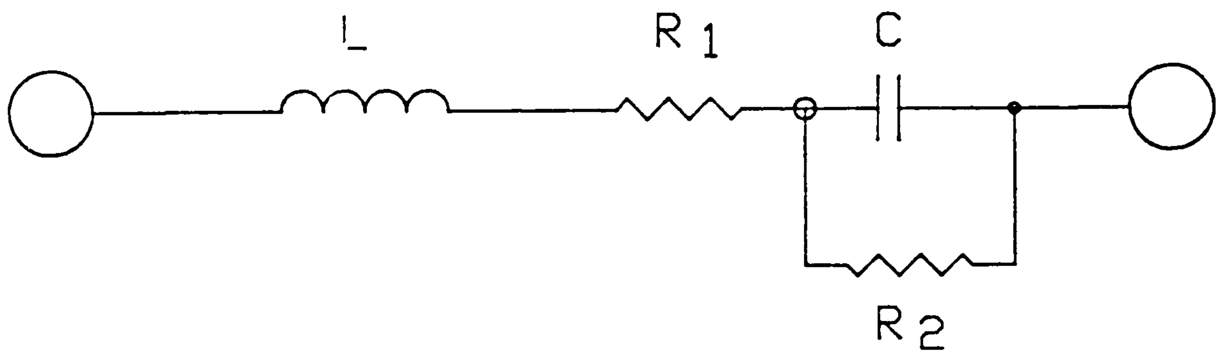


FIGURE 43. EQUIVALENT CIRCUIT OF A CAPACITOR

The series inductance can be attributed to lead inductance and the internal capacitor structure. At some frequency the series combination of C and L becomes resonant, providing very low impedance and effective shunting. Above this self-resonant frequency the impedance is an increasing inductive reactance, and the bypassing becomes ineffective.

Self-resonance can occur at relatively low frequencies. For example, a typical 0.1 μF paper capacitor is self-resonant at 2.5 MHz. The impedance of such a capacitor is shown in Figure 44.

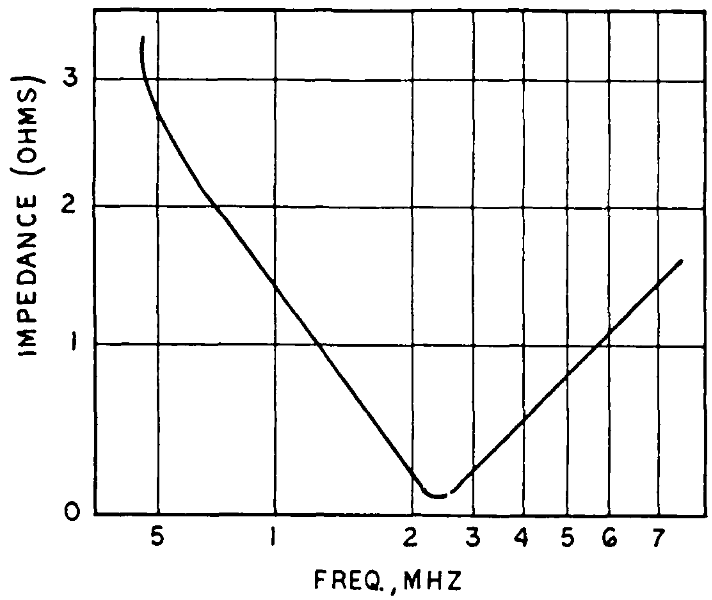


FIGURE 44. EFFECT OF FREQUENCY ON THE IMPEDANCE OF A 0.1μf PAPER CAPACITOR

The approximate usable frequency ranges for a variety of capacitor types ar shown in Figure 45.

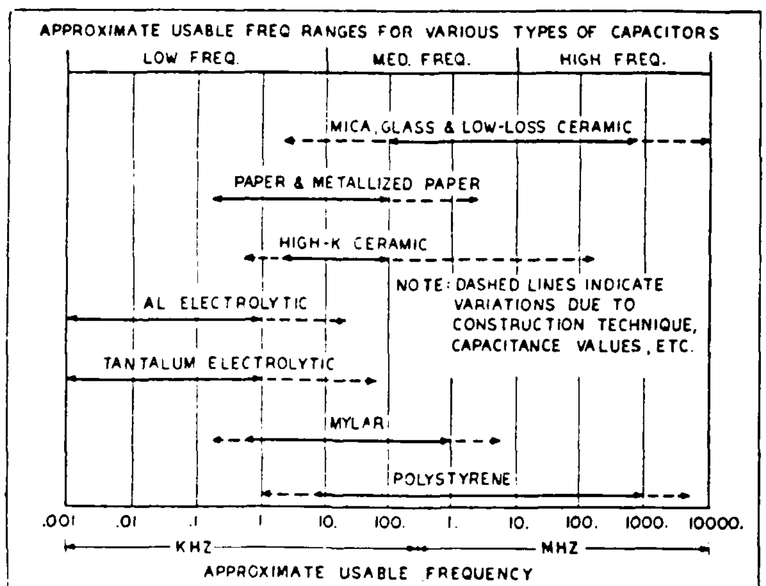


FIGURE 45. APPROXIMATE USABLE FREQUENCY RANGES FOR VARIOUS TYPES OF CAPACITORS

Because no single capacitor can provide satisfactory bypassing over a wide range of frequencies, an effective practice is to use two different capacitors in parallel. Good results are obtained when the capacitors are different in value by two orders of magnitude. For example, a 0.1 μf paper capacitor paralleled by a 0.001 μf ceramic will give good bypassing from audio frequencies to 35 MHz, where the small capacitor becomes self-resonant.

BOARD COMPONENT LOCATION. Component location can be important in minimizing radiation from printed circuit boards, and also in avoiding unwanted coupling among circuits on the board.

Clock chips and quartz crystals should be centrally located among the circuits they serve. This keeps the clock traces short. The traces themselves should be wide and near ground. A width to height ratio of at least 1:1 is desirable. The case of a quartz crystal should be laid down horizontally against a ground plane, rather than standing up vertically. Crystal cases should be grounded, not floating.

Components capable of generating magnetic fields should be located carefully with respect to other components which are susceptible to magnetic fields. Relays, inductors and toroids are cases in point. Mounting such components so that their fields are at right angles can reduce coupling.

High current devices should be located near their source of power. This practice will minimize the length of traces carrying high current.

The various types of grounds should be separated. More details on this subject are to be found in the grounding section.

Great care should be given to the reduction of loop area. Control of loop area is one of the most important aspects of EMI control in all electronic systems.

Bypassing should be arranged so that at least one bypass capacitor occurs for every three or four IC's. More bypassing may be necessary.

It is good practice to leave room on the board for additional capacitors in case add-on EMI control becomes necessary.

BOARD COMPONENT DENSITY. Higher board component density can be achieved by using narrow traces run close together. Unfortunately, this also increases the radiation from the traces and increases the crosstalk between traces.

One helpful aspect of high component density is that trace length is reduced, causing a decrease in both radiation, crosstalk and susceptibility.

Magnetic field coupling can occur between closely spaced relays, toroids and inductors, as mentioned in the preceding section. In some cases electric field coupling can also occur.

Thermal problems can occur from heat dissipated by IC's and by resistors, given a sufficiently great component density.

SOURCE OF NOISE. Fast rise time signals are the principal source of noise in digital systems.

High frequency clocks, digital logic circuits, switching power supplies, and rectifiers are examples of devices which generate fast rise time signals. Clocks and logic circuits generally produce radiated emissions. Switching power supplies and rectifiers cause conducted emissions.

Other possible sources of noise are relay contacts, switches, and thermal devices. All of these switching mechanisms can produce extremely fast rise time signals with radiation up to many megahertz, although the duty cycle or repetition rate is usually low. Diodes should be connected across relay coils to reduce noise from the collapse of field in the coil.

The highest frequency related to a fast rise time pulse is given approximately by.

$$f_{max} = \frac{1}{\pi t_r}$$

where t_r is the pulse risetime.

Note that the frequency given by the approximation above is not, in fact, the highest frequency. However, harmonics above f_{max} fall off at a rate of 40 dB per decade or 12 dB per octave.

The frequency spectrum of a signal is related to its time domain description through the Fourier transform. The time domain representation of a symmetrical trapezoidal pulse is shown in Figure 46, and the corresponding frequency domain representation, or Fourier transform is shown in Figure 47.

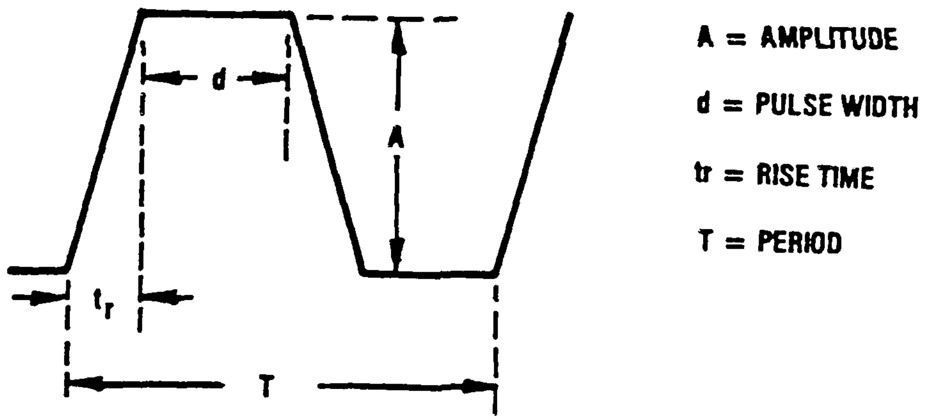


FIGURE 46. SYMMETRICAL TRAPEZOIDAL PULSE

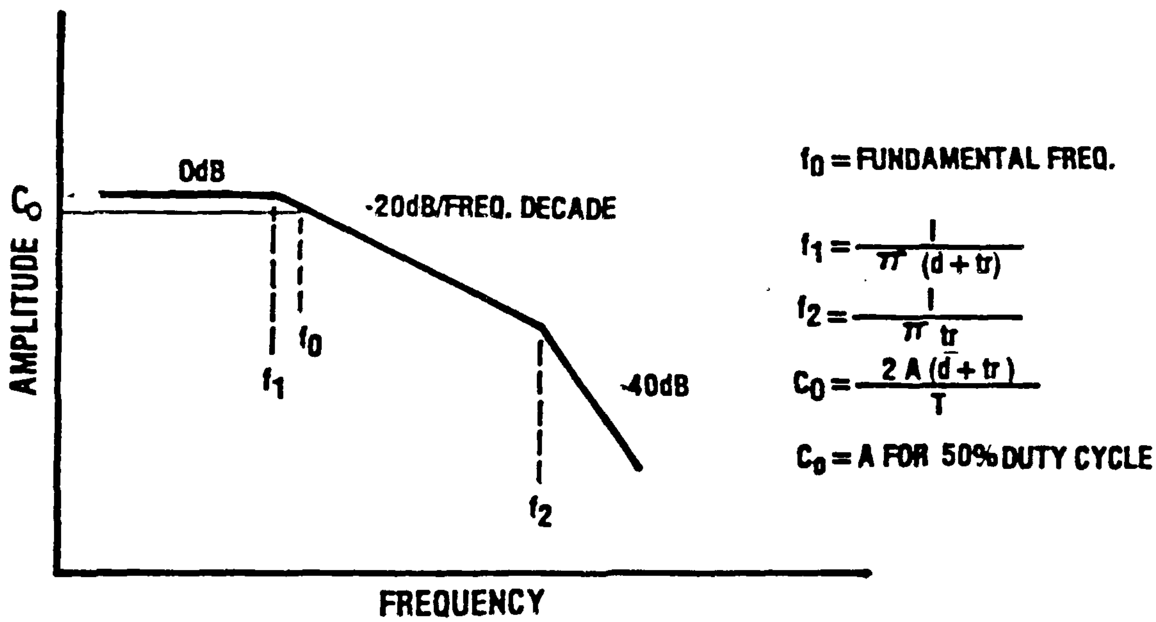


FIGURE 47. FOURIER TRANSFORM OF SYMMETRICAL TRAPEZOIDAL PULSE

ESD TESTING PRACTICES

INTRODUCTION. ESD, or electrostatic discharge, is the familiar phenomenon we have all experienced in stroking the back of a cat, or touching a metal door knob on a cold winter day after crossing a carpeted room. Some typical methods of generating electrostatic voltages are shown in Figure 48.

Means of Static Generation	Electrostatic Voltages	
	10 to 20 Percent Relative Humidity	65 to 90 Percent Relative Humidity
Walking across carpet	35,000	1,500
Walking over vinyl floor	12,000	250
Worker at bench	6,000	100
Vinyl envelopes for work instructions	7,000	600
Common poly bag picked up from bench	20,000	1,200
Work chair padded with polyurethane foam	18,000	1,500

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2 May 1980

FIGURE 48. TYPICAL ELECTROSTATIC VOLTAGES

Transfer of electronic charge is at the heart of this effect, and the accumulation of charge on the human body, or upon objects, can generate voltages up to several tens of kilovolts. When the accumulated charge is eventually redistributed via a discharge arc, the fast rise time pulse that results can have adverse effects on susceptible electronics equipments, ranging from upset to damage.

Aircraft are well suited for the generation of ESD. Relative humidity is often low, so that accumulated charge cannot slowly and harmlessly leak off without an arc; the decrease of atmospheric pressure with altitude somewhat decreases the insulating properties of air; and the motion of the aircraft through the air can cause an accumulation of charge. Finally, the movements of passengers and crew can cause charge accumulation in the same way as on the ground. Discharges from passengers or crew to metallic objects in the cabin can generate a field which may couple to cables routed behind decorative panels.

DESCRIPTION OF THE ESD PULSE. The typical ESD pulse is a double exponential, similar in shape to a lightning pulse, but with much faster rise and fall times. The 10 percent to 90 percent rise time is often taken to be 2 nanoseconds. The time to decay to 50 percent of the peak value is about 200 nanoseconds. This pulse is shown in Figure 49, along with definitions of the resistive phase, the time of arc breakdown, and the decay phase. For easier analysis the pulse can be approximated by a triangular waveform as shown in Figure 50. The frequency domain characterization of the pulse is also shown in Figure 50, and it can be seen that significant amounts of energy appear well into the VHF region. The ESD pulse is inherently broadband, and susceptible equipment must be protected against it from very low frequencies to several hundred megahertz.

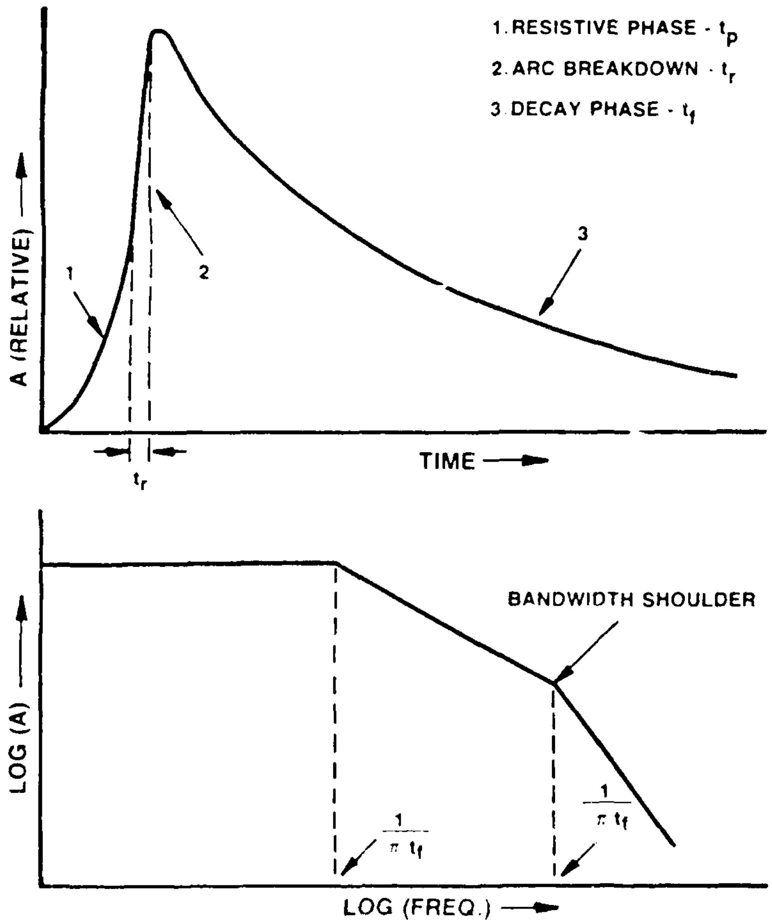
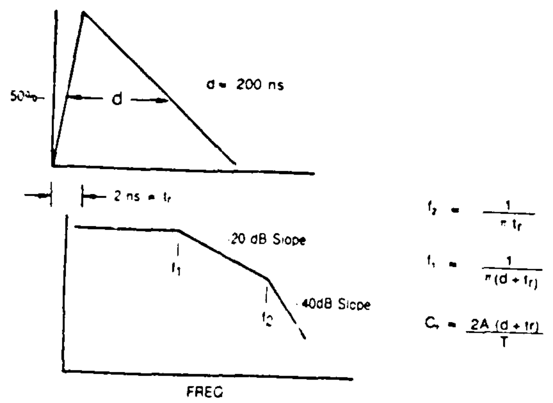


FIGURE 49. ESD VOLTAGE BREAKDOWN FACTORS

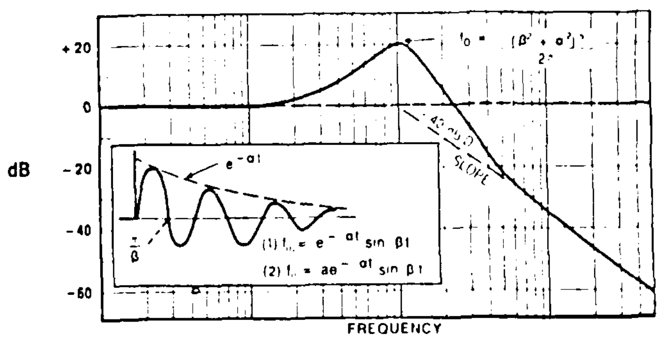


Let f_0 = 1 sec, T = 1
A = 12KV

FREQ	C_n	C_n dB µV	C_n db µV/MHz
1 Hz	4.8 mv	73.7	193.7
f_1 = 1.58MHz		73.7	193.7
f_2 = 159 MHz		33.7	153.7
1.0 GHz			

FIGURE 50. ESD VOLTAGE PREDICTION

This broadband pulse is capable of exciting resonances in nearby structures, and the resulting response is a damped sinusoid, as shown in Figure 51. The frequency of resonance is determined by the susceptible structure itself.



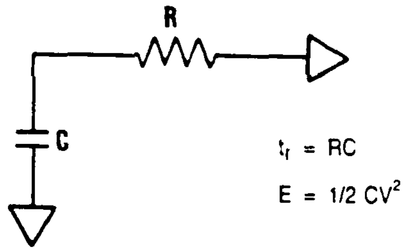
(1) $F_{(\omega)} = \dfrac{\beta}{(\alpha^2 + \beta^2) - \omega^2 + j\,2\alpha\omega}$

where $\omega = 2\pi f$

(2) $F_{(\omega)} = \dfrac{1}{\frac{2\alpha}{a} + j\frac{(\alpha^2 + \beta^2 - \omega^2)}{a\omega}}$

FIGURE 51. TIME TO FREQUENCY DOMAIN TRANSFORM FOR DAMPED SINUSOID WAVEFORM

DIRECT DISCHARGE ESD. An arc from the hand to a susceptible device is called direct discharge. The human body, for ESD purposes, can be modeled as a simple series R-C circuit, as shown in Figure 52.

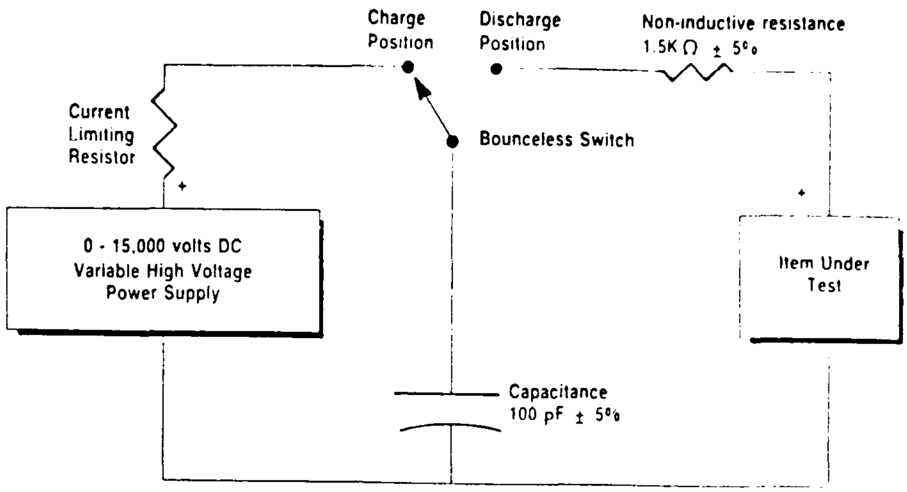


PARAMETER	LOW	HIGH	TYPICAL
R	5	100 kΩ	1-4 kΩ
C	25 pf	.003 uf	100-150 pf
V	1 kv	30 kv	8-12 kv
t_r	125 p sec	300 u sec	1-6 n sec
E	100 uJ	1.3 J	3-10 mJ

FIGURE 52. HUMAN ESD MODELING

For the simulation of an electrostatic discharge, an ESD simulator like that shown in Figure 53 is used. The energy stored in the human body or in the ESD simulator is given by 1/2 CV^2, where C is the capacitance and V is the voltage. This energy is in joules or watt-seconds and for a 100 pf capacitor charged to 12 kV, the energy delivered to the device under test is 0.007 joules. This energy can be compared to the energy specification of the device, if given, to see if there is danger of damage. The time required for the voltage to decay to 1/e (37 percent) of its initial value is given by t = RC, in this case 0.15 microseconds.

In many cases upset rather than damage is the result of this kind of discharge, and this condition is best checked during actual operation of the system under test. Repeated discharges will sometimes cause upset when a single discharge does not. A single discharge may occur in the time between data bits, whereas a repeated discharge has a good chance of occurring at the time of a data bit transmission and can cause a loss or alteration of data.



NOTE: Test voltages are measured across the capacitance. The capacitor shall be discharged through the series resistor into the item under test by maintaining the bounceless switch to the discharge position for a time no shorter than required to decay the capacitor voltage to less than 1 percent of the test voltage or 5 seconds, whichever is less. Power supply voltage shall be within a tolerance of ±5 percent of test voltage.

FIGURE 53. ESD TEST CIRCUIT

Many solid state devices are very susceptible to permanent damage from ESD incidental to handling. Careful training of installation and maintenance personnel, and adherence to handling procedures is required to avoid damage. CMOS integrated circuits are known to be particularly subject to damage by mishandling, both as isolated components and when installed on circuit boards. For the proper handling procedures for ESD sensitive components, much useful information can be obtained from ESD Control in the Manufacturing Environment, published by the Department of Defense Reliability Analysis Center. Military Handbook DODHDBK-263, Electrostatic Discharge Control Handbook for Protection of Electrical and Electronic Parts. Assemblies and Equipment is also of use.

Figures 54 through 56 list a wide variety of electronic devices, grouped by their sensitivity to the voltage of an electrostatic discharge. Figures 57 and 58 show the failure mechanisms that occur in various devices from ESD exposure.

Testing for direct discharge effects is performed as shown in Figure 59. In the laboratory, the AC power would be provided as shown. On board an aircraft, it is probably more convenient to power the ESD simulator from batteries. Systems are best checked by operating them and watching for upset or damage when the ESD pulse is applied.

The effects of ESD on a digital system can range from missing data, to a condition requiring reset of the system, to damage requiring the replacement of parts.

CLASS 1: SENSITIVITY RANGE 0 TO ≤1000 VOLTS

- Metal Oxide Semiconductor (MOS) devices including C, D, N, P, V and other MOS technology without protective circuitry, or protective circuitry having Class 1 sensitivity
- Surface Acoustic Wave (SAW) devices
- Operational Amplifiers (OP AMP) with unprotected MOS capacitors
- Junction Field Effect Transistors (JFETs) (Ref.: Similarity to MIL-STD-701 Junction field effect, transistors and junction field effect transistors, dual unitized)
- Silicon Controlled Rectifiers (SCRs) with $Io<0.175$ amperes at 100° Celcius (°C) ambient temperature (Ref.: Similiarity to MIL-STD-701: Thyristors (silicon controlled rectifiers)
- Precision Voltage Regulator Microcircuits: Line or Load Voltage Regulation <0.5 percent
- Microwave and Ultra-High Frequency Semiconductors and Microcircuits: Frequency >1 gigahertz
- Thin Film Resistors (Type RN) with tolerance of ≤0.1 percent; power>0.05 watt
- Thin Film Resistors (Type RN) with tolerance of >0.1 percent; power ≤0.05 watt
- Large Scale Integrated (LSI) Microcircuits including microprocessors and memories without protective circuitry, or protective circuitry having Class 1 sensitivity (Note: LSI devices usually have two to three layers of circuitry with metallization crossovers and small geometry active elements)
- Hybrids Utilizing Class 1 parts

FIGURE 54. LIST OF ESD PARTS BY PART TYPE (Class 1)

CLASS 2: SENSITIVITY RANGE >1000 TO ≤4000 VOLTS

- MOS devices or devices containing MOS constituents including C, D, N, P, V, or other MOS technology with protective circuitry having Class 2 sensitivity
- Schottky diodes (Ref.: Similiarity to MIL-STD-701: Silicon switching diodes (listed in order of increasing trr)
- Precision Resistor Networks (Type RZ)
- High Speed Emitter Coupled Logic (ECL) Microcircuits with propagation delay ≤1 nanosecond
- Transistor-Transistor Logic (TTL) Microcircuits (Schottky, low power, high speed, and standard)
- Operational Amplifiers (OP AMP) with MOS capacitors with protective circuitry having Class 2 sensitivity
- LSI with input protection having Class 2 sensitivity
- Hybrids utilizing Class 2 parts

FIGURE 55. LIST OF ESD PARTS BY PART TYPE (Class 2)

CLASS 3: SENSITIVITY RANGE >4000 TO ≤15,000 VOLTS

- Lower Power Chopper Resistors (Ref.: Similarity to MIL-STD-701: Silicon Low Power Chopper Transistors)
- Resistor Chips
- Small Signal Diodes with power ≤1 watt excluding Zeners (Ref.: Similarity to MIL-STD-701: Silicon Switching Diodes (listed in order of increasing trr)
- General Purpose Silicon Rectifier Diodes and Fast Recovery Diodes (Ref.: Similarity to MIL-STD-701: Silicon Axial Lead Power Rectifiers, Silicon Power Diodes (listed in order of maximum DC output current), Fast Recovery Diodes (listed in order of trr)
- Low Power Silicon Transistors with power ≤ 5 watts at 25°C (Ref.: Similarity to MIL-STD-701: Silicon Switching Diodes (listed in order of increasing trr), Thyristors (bi-directional triodes), Silicon PNP Low-Power Transistors (Pc ≤ 5 watts T_A = 25°), Silicon RF Transistors)
- All other Microcircuits not included in Class 1 or Class 2
- Piezoelectric Crystals
- Hybrids utilizing Class 3 parts

DOD-HDBK-263
2 May 1980

FIGURE 56. LIST OF ESD PARTS BY PART TYPE (Class 3)

Part Constituent	Part Type	Failure Mechanism	Failure Indicator
MOS Structures	MOS FET (Discretes) MOS ICs Semiconductors with metalization cross-overs Digital ICs (Bipolar and MOS) Linear ICs (Bipolar and MOS) MOS Capacitors Hybrids Linear ICs	Dielectric breakdown from excess voltage and subsequent high current	Short (high leakage)
Semiconductor Junctions	Diodes (PN, PIN, Schottky) Transistors, Bipolar Junction Field Effect Transistors Thyristors Bipolar ICs, Digital and Linear Input Protection Circuits on: Discrete MOS FETs MOS ICs	Microdiffusion from microplasma-secondary breakdown from excess energy or heat Current filament growth by silicon and aluminum diffusion (electromigration)	

FIGURE 57. PART CONSTITUENTS SUSCEPTIBLE TO ESD

Part Constituent	Part Type	Failure Mechanism	Failure Indicator
Film Resistors	Hybrid ICs: Thick Film Resistors, Thin Film Resistors Monolithic IC-Thin Film Resistors Encapsulated Film Resistors	Dielectric breakdown voltage dependent-creation of new current paths Joule heating-energy dependent-destruction of minute current paths	Resistance shift
Metallization Strips	Hybrid ICs Monolithic ICs Multiple Finger Overlay Transistors	Joule heating-energy dependent metallization burnout	Open
Field Effect Structures and Nonconductive Lids	LSI and Memory ICs employing nonconductive quartz or ceramic package lids especially ultraviolet EPROMS	Surface inversion or gate threshhold voltages shifts from ions deposited on surface from ESD	Operational degradation
Piezoelectric Crystals	Crystal Oscillators Surface Acoustic Wave Devices	Crystal fracture from mechanical forces when excessive voltage is applied	Operational degradation
Closely Spaced Electrodes	Surface Acoustic Wave Devices Thin metal unpassivated, unprotected semiconductors and microcircuits	Arc Discharge melting and fusing of electrode metal	Operational degradation

FIGURE 58. PART CONSTITUENTS SUSCEPTIBLE TO ESD (Con't)

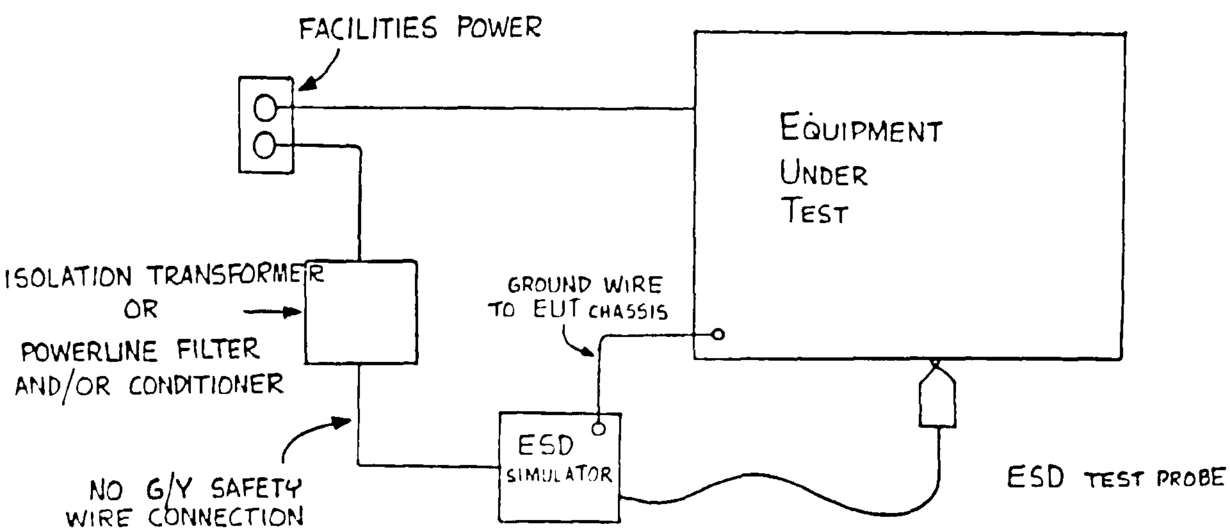


FIGURE 59. ESD DIRECT DISCHARGE EQUIPMENT CONFIGURATION

RADIATED INDUCED ESD.

Prediction of Electric Fields. In the following example, we predict the effects of an electrostatic discharge from the human hand to a metallic object in the cockpit of a commercial transport. A cabin attendant comes forward, touches a metallic object located one meter away from a susceptible item of equipment, and causes a discharge to occur.

In this analysis we start with the triangular pulse approximation of Figure 50, and use the chart in Figure 60 to guide the calculations.

Referring to the frequency column of Figure 60, we enter 1.58 MHz as the frequency at which we first calculate the radiated field of a 12 kV discharge. In column 1 we enter 193.7 dBuV/MHz, the Fourier coefficient obtained in Figure 50. In column 2 we enter -34 dB, which is a standard transfer function value to convert from a voltage on a conductor to the electric field generated one meter away. In column 3 we enter a figure in dB for the antenna factor, to determine how efficient the source is as a radiator of energy. Basically this calculation compares the level of radiation from a short antenna to that of a quarter wavelength antenna at the same frequency. In this case, the entire human body is the major part of the radiating circuit, and can be taken to be 2 meters. Using Note 3 in Figure 60, we calculate the antenna factor as 13.8 dB. Finally, in columns 4,5 and 6 we enter decibel values for the number of conductors carrying the signal, for the distance from the discharge to the susceptible nearby equipment, and for the height above ground of the radiating object. Values of zero dB have been entered in each of these columns, indicating one conductor, a distance of 1 meter, and a height above ground of 0.5 meters.

FREQ	1. Cn dBμv/MHz	2. T_F	3. Ant. F.	4. N	5. D	6 Height	7. Result
1 58MHz	193.7	-34	-13.8	0	0	0	145.9

1. Freq versus ampt. in dB μv/MHz from trransform
2. T_F = Transfer function that converts voltage on lead to E-field 1 meter from radiating source = −34dB
3. Ant Factor for when radiator is less than $\lambda/4$ in length $-10 \text{ Log } f_3/f_x$ where $f_3 \Rightarrow 3 \times 10^8/4\ell$
   ℓ = length of radiating lead, meters
4. # of leads carring same signal = $+ 10 \text{ Log } N$
5. Distance Factor $\Rightarrow -20 \text{ Log } D/1$ meter
6. Height above ground surface compared to .5 meters = $10 \text{ Log } .5/H$ (meters)

FIGURE 60. PREDICTION OF RADIATED EMISSIONS

The decibel values in columns 1 through 6 are now added to obtain the result in column 7, 145.9 dBuV/MHz. This value can be converted to a radiated field strength:

$$E = 10^{(145.9/20)} \qquad \mu V/\text{meter}/\text{MHz}$$

or

$$E = 19.7 V/\text{meter}/\text{MHz}$$

To arrive at the magnitude of the field threatening a broadband device, we must integrate the frequency domain plot of Figure 50 over the frequency range of interest. Piecewise integration from dc to 1.58 MHz yields a value of 31.2 volts per meter. This is the field value that would be seen by a device which cuts off sharply at 1.58 MHz. Continuing the integration from 1.58 MHz to 159 MHz gives an additional field value of 1,550 volts per meter as the contribution from this portion of the spectrum. The total field is the sum of the two piecewise integrations, and is then 31.2 + 1,550 = 1,581 volts per meter per 159 MHz. This is the field which would be seen by a broadband device having response from very low frequencies up to 159 MHz. Above 159 MHz, the field strength decreases at a rate of 40 dB per decade, and the contribution from this part of the spectrum is negligible.

Prediction of Voltage Induced on a Susceptible Circuit The next part of the problem is to calculate how much voltage might be induced on a susceptible circuit by the field calculated above. For the susceptible circuit we will use a printed circuit board into which a loop area has been inadvertently designed. The loop acts as an antenna which responds to the field calculated above. We assume loop dimensions of 4 inches by 6 inches, which could be created by a trace around the periphery of the board. The induced voltage as a function of field strength is given by:

$$V(\text{volts}) = 2\pi \, E \, N \, A/\lambda \, \cos\theta$$

E is the field strength; N is the number of turns in the receiving loop; A is the area of the loop in square meters; λ is the wavelength; and θ is the angle of the loop with respect to the radiating field, and is usually taken to be zero for a worst case estimate.

An effective way of combining the broadband field generated by the pulse, with the response of the loop from dc to 159 MHz is to plot the field and the response in dB on semilog paper, as in Figure 61. No values are shown above 159 MHz because the induced voltage above that point is negligible. The generated E field is shown in the uppermost curve, and has the frequency distribution as in Figure 50. The loop response with an applied field of 1 volt per meter is shown in the lower curve, and shows a linear response with frequency with a slope of 20 dB per decade. The curve for the induced voltage in the loop is found by combining the applied field curve and the response curve, and from 1.58 MHz up shows a steady value of 7.9 millivolts per MHz. When the curve in Figure 61 is integrated over the bandwidth to 159 MHz, we find that the total voltage to which a broadband device would be exposed is about 1.25 volts, which is in the upset range for TTL devices.

In the analysis it was assumed that the receiving loop was unshielded. Shielding of the susceptible circuit will normally provide enough attenuation to reduce the induced voltage to a non-threatening level. However, an increase in the original 12 kV discharge voltage, or a failure of shielding, a resonance condition in the susceptible circuit, or a decrease in the distance from the discharge to the susceptible equipment, can increase the induced voltage to the upset or damage levels.

Testing for radiated induced ESD is performed on installed systems using an ESD simulator while operating the systems under test and looking for upset conditions or damage conditions.

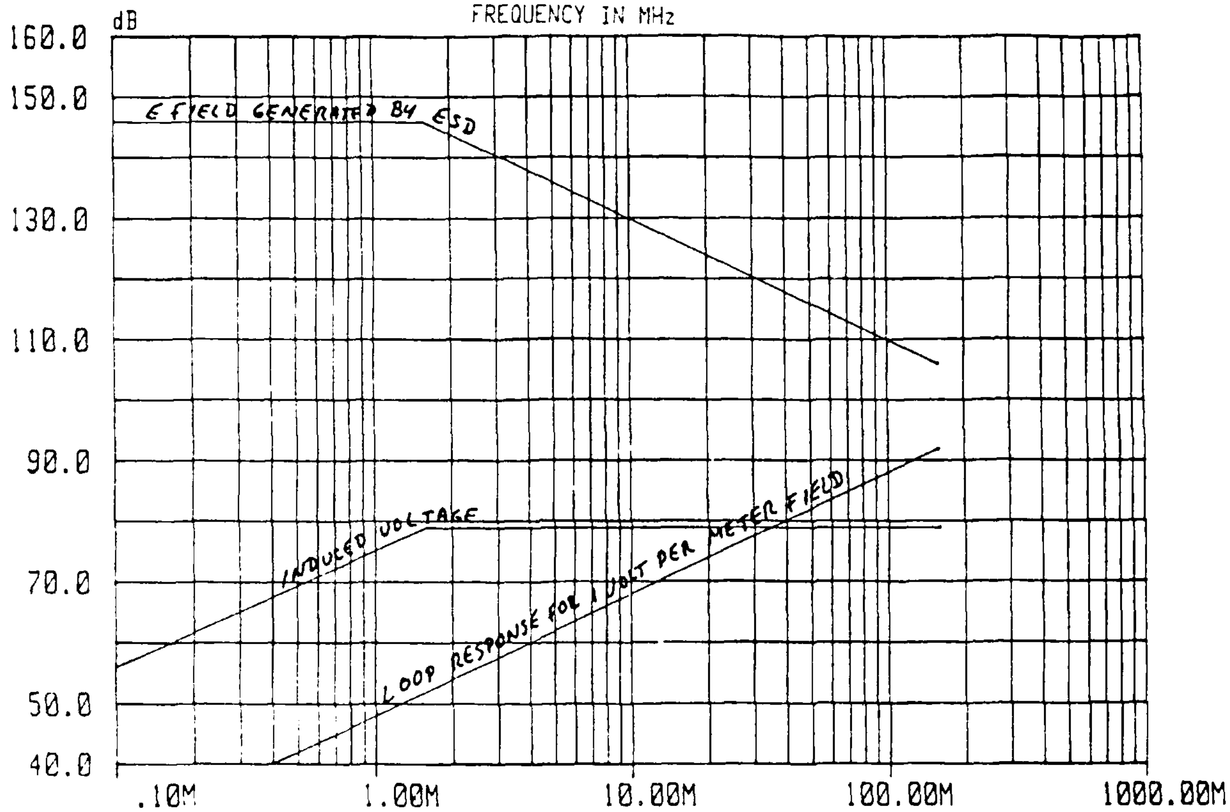
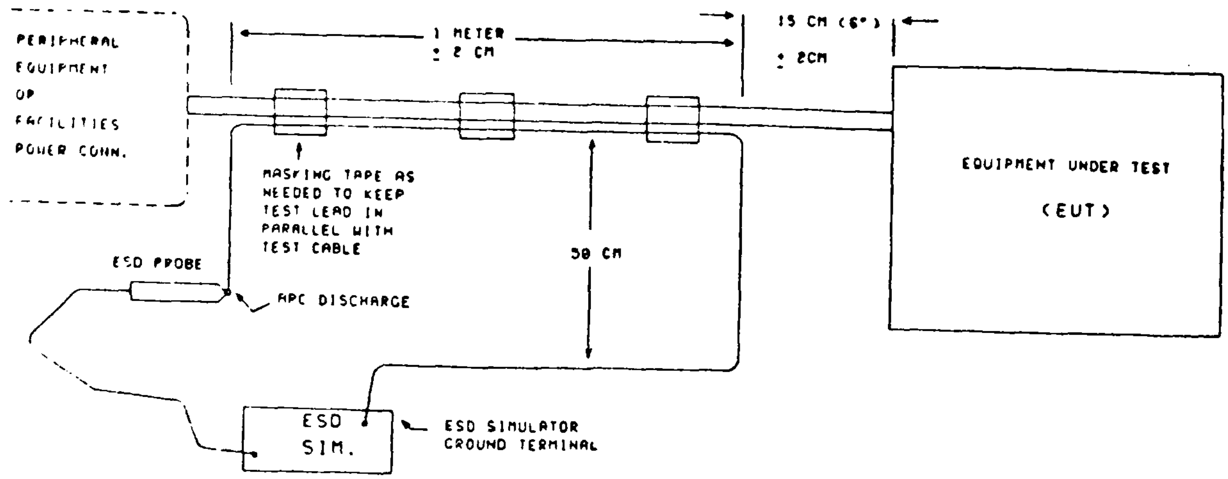


FIGURE 61. INDUCED VOLTAGE ON A LOOP FROM ESD

ESD INDUCED VOLTAGES ON CABLES. When a discharge occurs to a cabinet which is connected to a shielded cable, some of the discharge current flows to ground via the cable shield. The noise of the discharge can be coupled to the cable center conductor through the transfer impedance of the cable, or by direct coupling to an exposed center conductor, in the case of improperly terminated shields, or directly through the shield itself. Figure 62 shows a test arrangement used for the ESD testing of cables. In an installed system, the ESD simulator voltage would be applied near the connectors at each end of a cable, and to as much of the shielding between as is accessible.

In tests by the 3M Company, voltages larger than 500 volts could be induced on a center conductor by ESD if one end of the shield was unterminated. When the shield was terminated with a pigtail connection, the induced voltage was reduced to 16 volts, and when the proper 360 degree backshell termination was made, a further reduction to just over a volt occurred. These results point out both the threatening nature of ESD, and the necessity of proper cable shield termination, as outlined earlier in this report.



NOTES:
1. TEST LEAD PLUS LOOP SHOULD BE 1 METER AWAY FROM ANY CONDUCTING SURFACE INCLUDING CEMENT FLOOR. A WOOD TABLE IS AN IDEAL TEST PLATFORM.
2. WHEN TEST CABLE IS LESS THAN 1.15 METERS IN LENGTH, THEN TEST PARALLEL SHALL BE SHORTER AND STARTED A MINIMUM OF 5 CM (2") AWAY FROM EUT CONNECTOR.

FIGURE 62. ESD TEST CONFIGURATION (CABLE INDUCED)

PREDICTION OF ESD PHENOMENA

INTRODUCTION. A detailed analysis was given in the preceding section for the prediction of the voltage which would be induced in a susceptible circuit when an electrostatic discharge took place nearby. This procedure will be briefly reviewed.

The procedure falls into four parts. First, the field at a distance of one meter from the discharge is calculated. Later, if desired, the field strength can be adjusted for distances other than one meter. A starting point for this field calculation is the information contained in Figure 50.

Next, the susceptibility of the threatened circuit is calculated. Then the field and the susceptibility curves are combined to produce a resultant curve of induced voltage versus frequency. Graphical combining of the curves is most readily performed on a semilog scale as shown in Figure 61.

The voltage induced in the susceptible circuit is found by converting from decibel microvolts per meter per megahertz to volts per meter per megahertz. Finally, the average value of the induced voltage over the frequency range is found by integrating over the entire spectrum of interest. After appropriate adjustments are made for shielding or other attenuation factors, the induced voltage can be compared to the upset and damage levels of the threatened circuit or system.

Typical upset levels are 0.4 volts for TTL, 5 to 25 millivolts for video circuits, and 25 millivolts for analog circuits.

Much of this analysis can be more easily performed by EMCad software than by the manual procedure detailed above. EMCad is a group of IBM PC compatible programs useful in solving a wide variety of electromagnetic compatibility problems.

We first use EMCad to predict the field strength from a discharge. Next, we use a second program to predict the response of a susceptible circuit to a field of one volt per meter. Finally, just as in the manual procedure, the applied field and the response are combined to predict the induced voltage on the susceptible circuit. At this time, the software will not accomplish this last step, which still must be done manually.

Using EMCad a number of examples of the effects of ESD upon cables and harnesses will be worked out. In each case, we start with an unshielded cable or harness. Assuming upset levels of 0.4 volts, typical for TTL, we can then determine what amount of cable or cabinet shielding is required to prevent upset.

EXAMPLES

Example No. 1. An electrostatic discharge of 12 kV is applied to a 2-meter length of cable. One meter away, a susceptible wiring harness 2 feet in length, responds to the field generated by the discharge. What is the voltage as a function of frequency induced on the harness, and what is the total voltage induced on a broadband device?

Using EMCad we first predict the field strength generated by the discharge at a distance of one meter. For this prediction a number of inputs are needed:

Type of Pulse: Triangular
Pulse Risetime: 2 nanoseconds
Pulse Fall Time: 200 nanoseconds
Pulse Repetition Rate: 1 Hz
Pulse Amplitude: 12 kilovolts
Cable Length: 2 meters
Shielding of Cable: None
Number of Conductors in Cable: 1
Distance from signal to return: 2 inches
Distance from signal to ground: 2 inches
Wire Size: 18 AWG
Test Distance: 1 meter

The resulting field as a function of frequency is shown in the upper curve of Figure 63. The frequency is shown on the logarithmic horizontal axis. The field strength is in dB microvolts per meter per megahertz, on the linear vertical axis.

Next, using a second program, we predict the susceptibility of a nearby circuit to an applied field of 1 volt per meter. The inputs needed for this prediction are:

Harness Length: 24 inches
Height above ground: 2 inches
Wire Size: 18 AWG
Source Resistance: 50 Ohms
Load Resistance: 50 Ohms

The voltage induced on the harness is shown as the lower curve in Figure 63. The induced voltage is expressed in dB microvolts per megahertz, and it should be noted that this is the induced voltage which would be produced by application of a 1 volt per meter field.

At 100 kHz, the applied field is 120 dB microvolts per meter per megahertz, or one volt per meter per megahertz. The resulting induced voltage at that frequency is 28 dB microvolts per meter per megahertz, or 25 microvolts per megahertz.

Note that both the upper and lower curves have well defined regions in which the slopes are either 10, 20, or 40 dB per decade. Once the starting point of the induced voltage curve has been determined, the slopes of the two curves can be added to rapidly find the slope of the resultant curve.

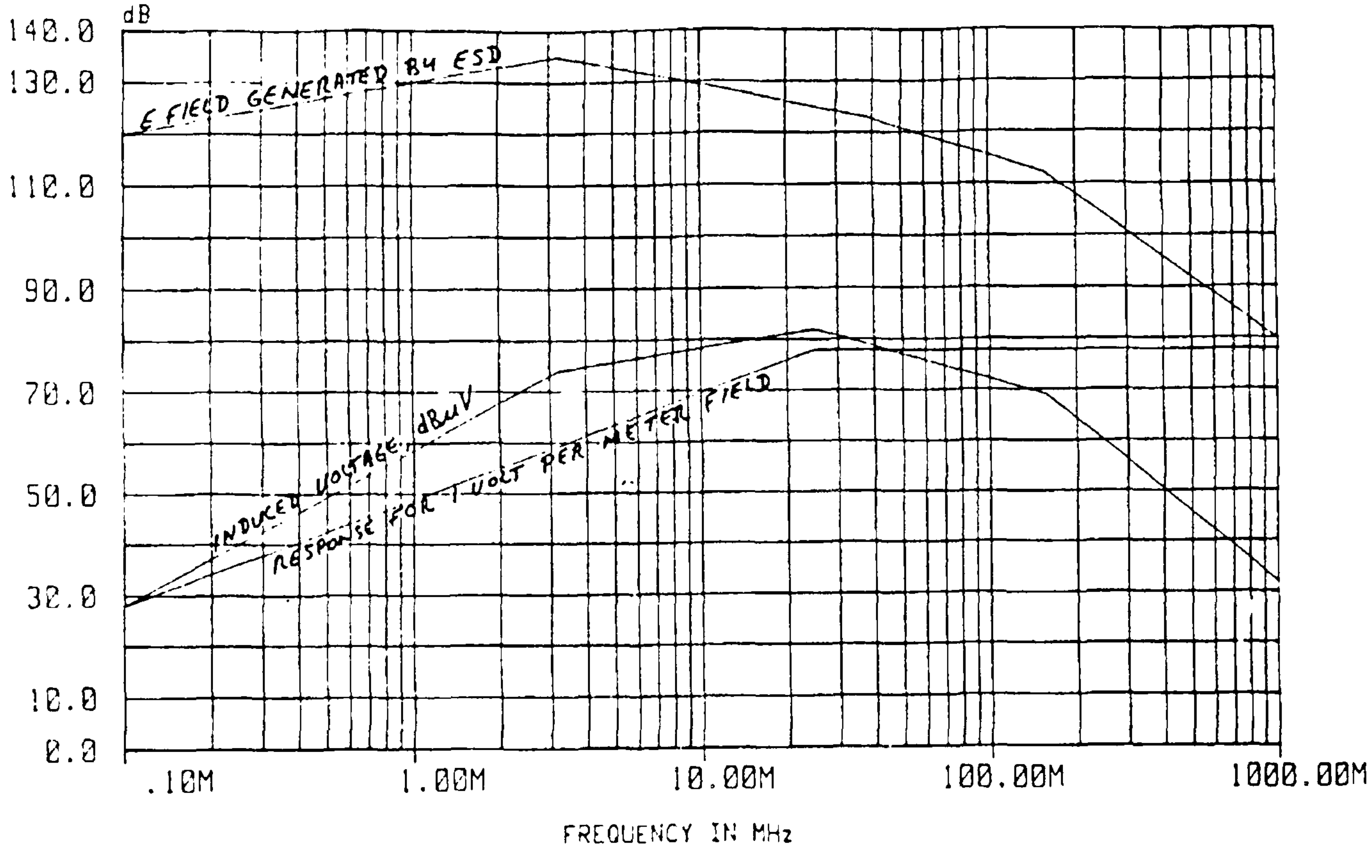


FIGURE 63. VOLTAGE INDUCED ON 2-FT. HARNESS FROM A 2 METER ESD SOURCE

The resultant curve is shown between the other two curves on the left of Figure 63. At 40 MHz it crosses below the 1 volt per meter response curve. It can be seen that the maximum induced voltage occurs in the 25 MHz region, where the value is 78 dBuV/MHz or about 8 mV/MHz.

Piecewise integration of each section of the induced voltage curve, followed by addition of the values obtained, yields a total voltage of 1.4 volts. This is the voltage that would be induced on a broadband device with a response to 1 GHz. Even if the response cuts off at 160 MHz, the induced voltage will be about 0.75 volts, which is highly likely to cause upset in a TTL device.

Further Examples. The preceding calculation has been repeated for cable lengths of 3, 4, and 5 meters. That is, the electrostatic discharge is applied to cables of those lengths. All of the other input parameters were kept constant. The characteristics of the harness also have been kept the same as in the first example. The results are shown in Figures 64, 65, and 66.

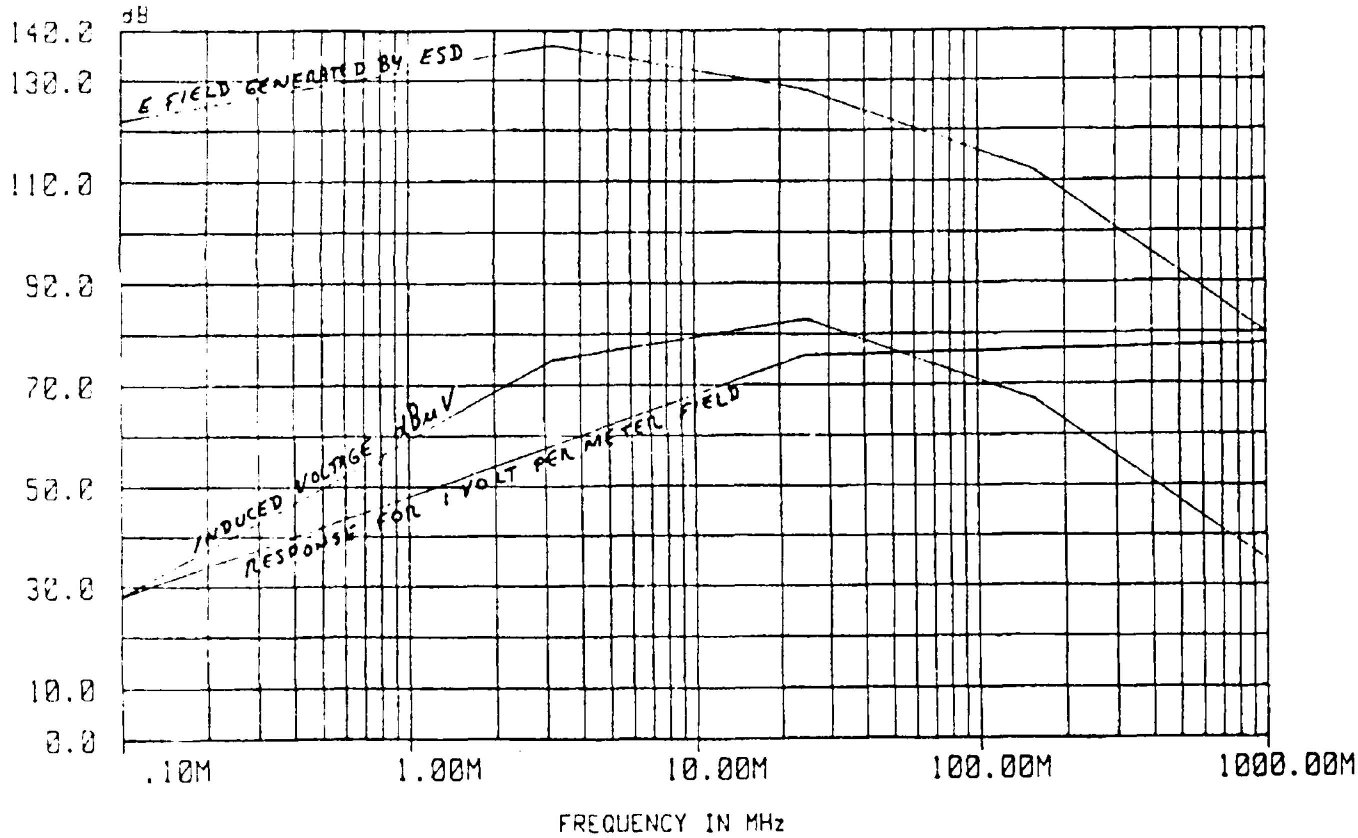


FIGURE 64. VOLTAGE INDUCED ON 2-FOOT HARNESS FROM A 3-METER ESD SOURCE

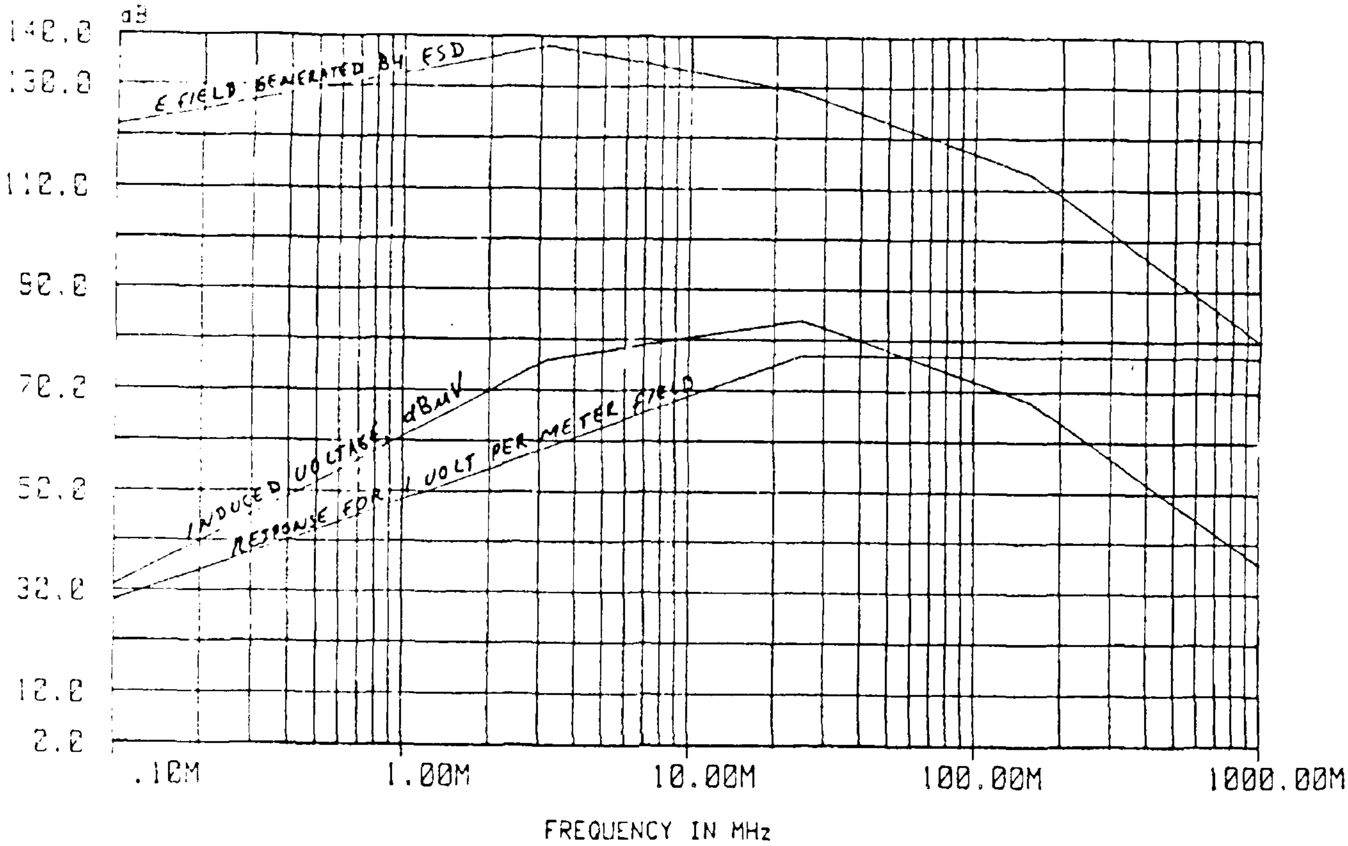


FIGURE 65. VOLTAGE INDUCED ON 2-FOOT HARNESS FROM A 4-METER ESD SOURCE

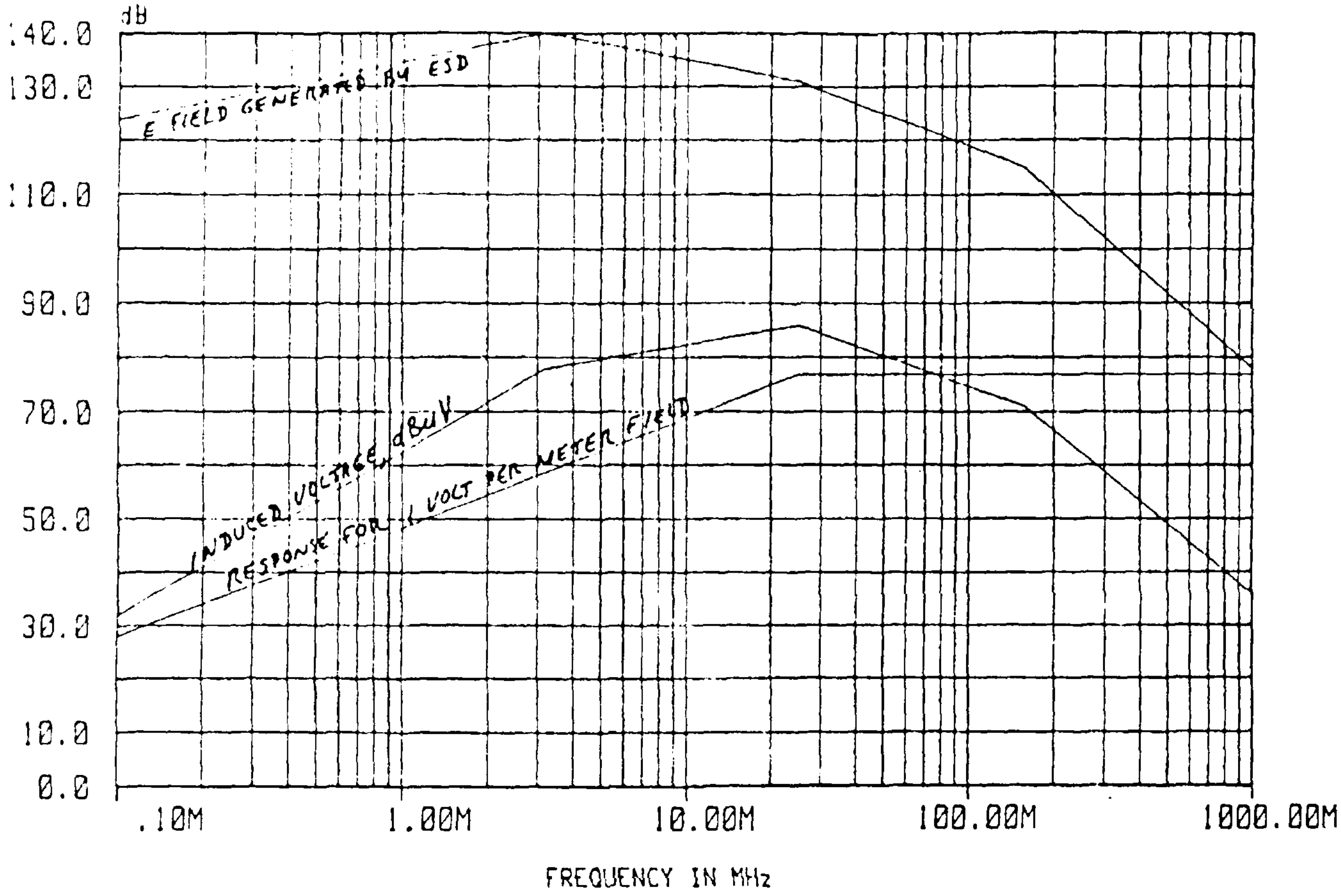


FIGURE 66. VOLTAGE INDUCED ON 2-FOOT HARNESS FROM A 5-METER ESD SOURCE

The shape of the induced voltage curve is nearly the same in every case. The main effect of increasing the length of cable exposed to the discharge is to increase the radiated field level by a few dB with each increment of length. For frequencies up to 10 or 20 MHz, the excited cable is short compared to wavelength, and is a relatively inefficient radiator. At the point in the frequency spectrum where the cable is of significant length to radiate efficiently, the ESD pulse is beginning to decrease in amplitude. The net result of these two effects is to create an induced voltage peak at about 25 MHz in each case.

In all of these examples, the susceptible harness was unshielded. Shielding can be accomplished by putting the harness in a braided cable shield, or enclosing it within a cabinet. In either case, the shielding will be sufficient to reduce the induced voltage to a negligible level. For example, a standard braid will provide 70 dB of shielding to one megahertz, decreasing at 20 dB per decade above one megahertz, until a residual value of 20 dB is reached up to several gigahertz, as shown in Figure 67.

Any ordinary cabinet with ventilating holes will also provide adequate shielding, as long as slot radiators, caused by lack of contact on seams between fasteners, are avoided.

In the absence of shielding, ESD represents a serious hazard. As a final example, we will predict the voltage induced on a coaxial cable which has been improperly terminated in a pigtail connection rather than a full 360-degree connection to the backshell.

In this example, a 12 kV discharge is applied to a 2-meter length of cable as in the first example. The resulting field is shown in the upper curve of Figure 68. The response of the one inch of exposed center conductor is also shown, as is the induced voltage response curve, which is, as before, the combination of the other two.

The induced voltage curve is reproduced in Figure 69. The voltage induced is on the order of 70 millivolts for a broadband device with response to about 300 MHz. This level could be troublesome for wideband video or analog circuits, although it is probably below the upset level for logic circuits.

When the curve for braided shielding of Figure 67 is combined with that of Figure 69, the lower curve in Figure 69 is obtained, showing the dramatic improvement in ESD response when the cable shield is properly terminated. Now there is complete protection against even stronger ESD fields than were generated in this example.

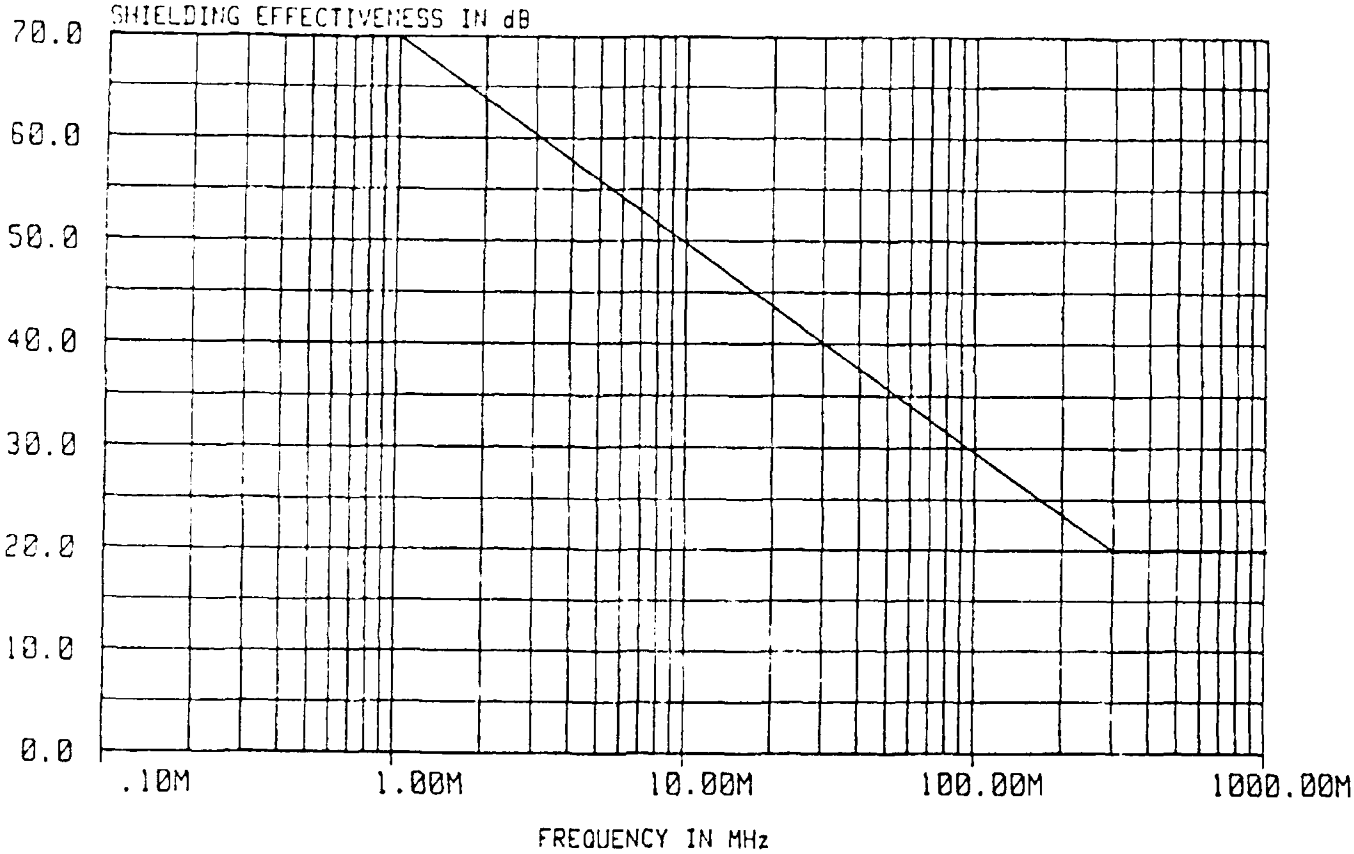


FIGURE 67. TYPICAL BRAIDED SHIELD CABLE
SHIELDING EFFECTIVENESS

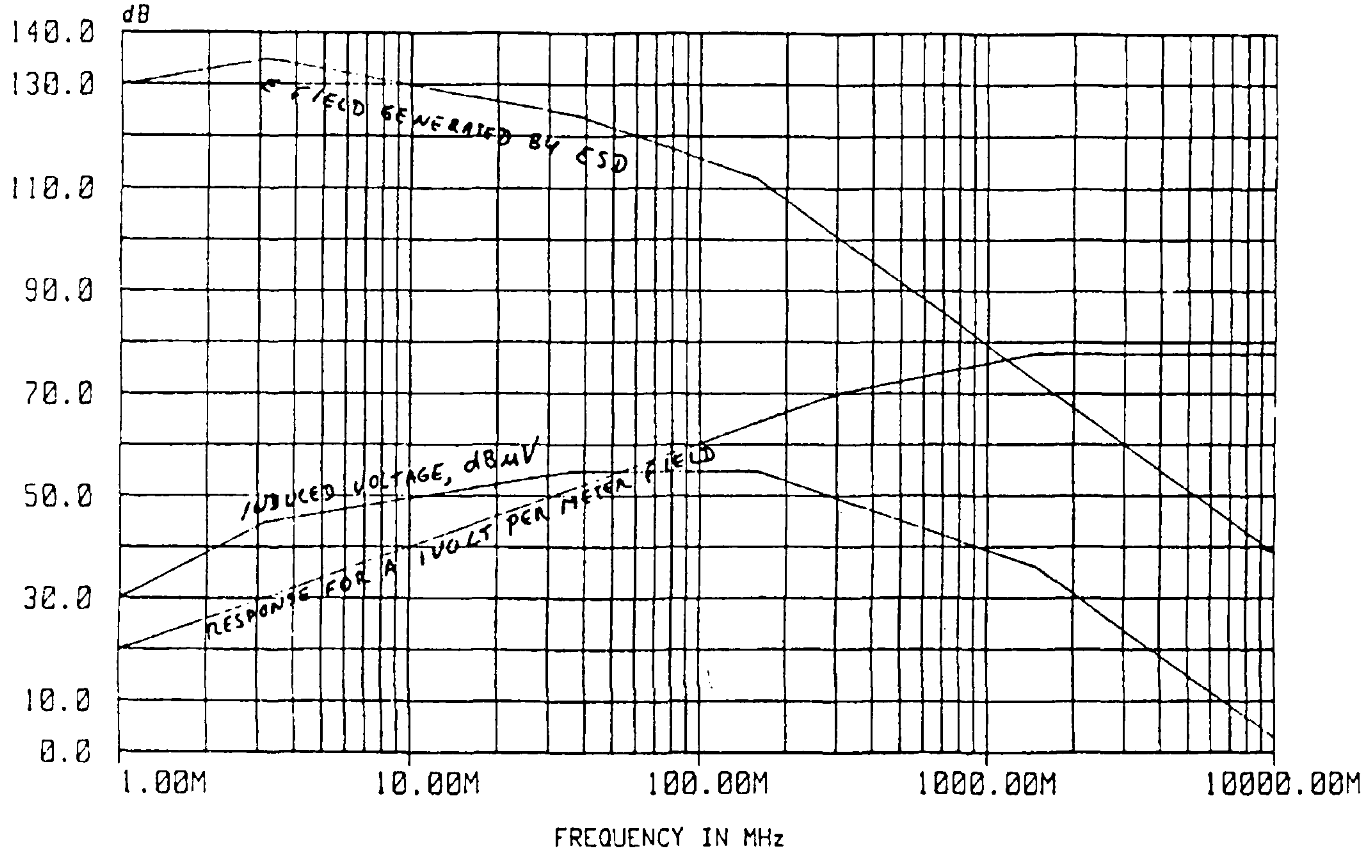


FIGURE 68. INDUCED VOLTAGE ON A 1-INCH PIGTAIL FROM A 2-METER ESD SOURCE

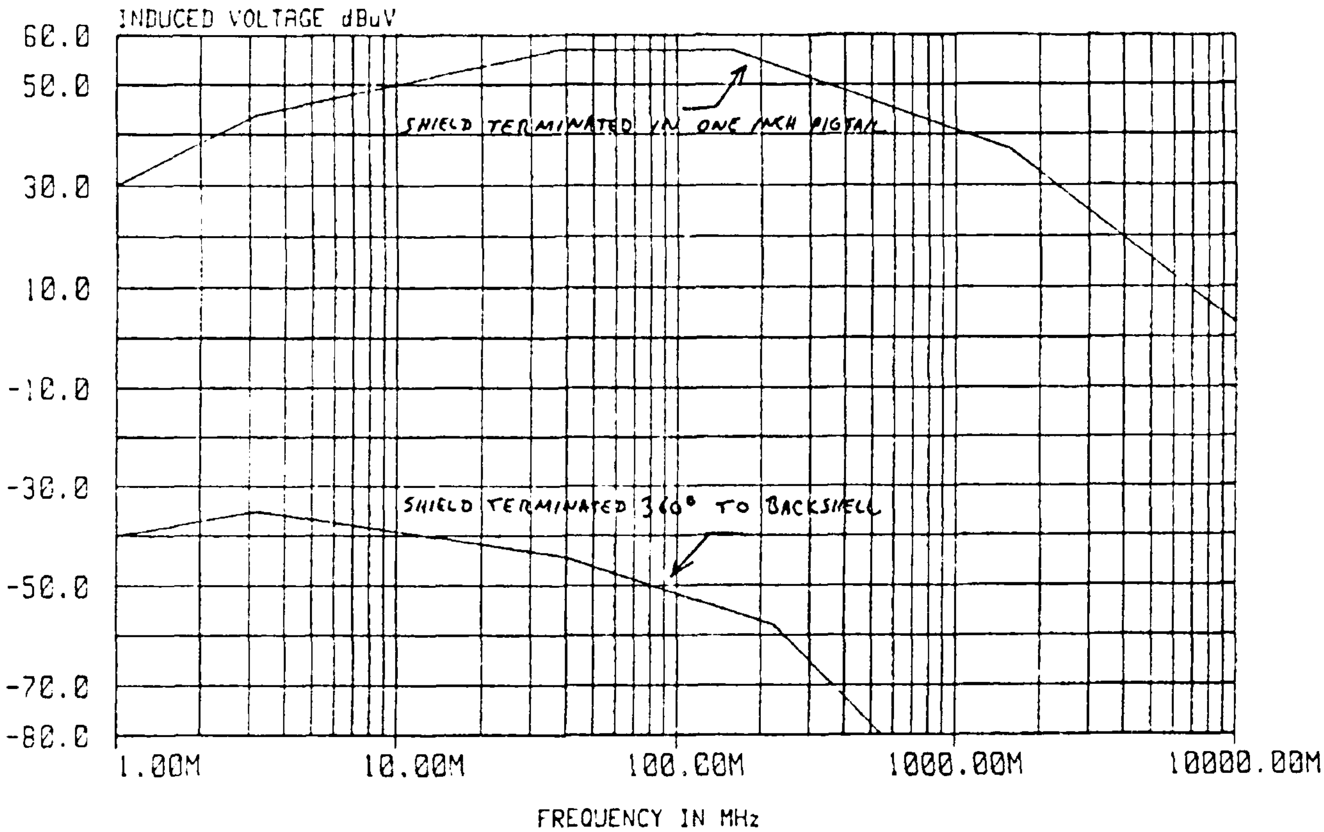


FIGURE 69. COMPARISON OF ESD INDUCED VOLTAGE ON A PIGTAIL AND 360° BACKSHELL

ESD TESTING RECOMMENDATIONS. RTCA DO-160B Environmental Conditions and Test Procedures for Airborne Equipment specifies several tests for transient voltages in Section 17, Voltage Spike and Section 19, Induced Signal Susceptibility. In addition to those tests, it is recommended that ESD tests be performed because of the flight critical nature of new systems, and the high susceptibility of broadband logic devices to ESD phenomena.

The specific ESD test recommendations for flight critical systems are as follows.

Equipment, systems, and installed systems should be ESD tested according to the following table:

Failure Mode	ESD Test Level	Simulator	Circuit Values
Soft	8 kV	150 pf	1200 ohms
Hard	12 kV	150 pf	1200 ohms
Damage	25 kV	150 pf	1200 ohms

A soft failure is one which causes an alteration of data or missing data. A hard failure is one which requires a reset of equipment. Damage requires repair or replacement of a system, subsystem, or component.

All tests should start at the 2-kV level, and should be advanced in 1-kV increments until a failure occurs, or the specified level is reached without failure.

GENERIC DIGITAL DEVICES EMISSIONS

CASE SHIELDING AND POWER LINE FILTERING Many different types of digital devices are used to control the timing and processing of digital data. Pulse rise times of 2 to 10 nanoseconds are common. The fast rise pulses generated and processed in logic devices are the principal sources of noise in digital systems.

Although the number of possible system designs for data processing and control in aircraft is extremely large, most of these systems share features which can be characterized in terms of their potential for electromagnetic noise generation. Some of the dominant features are the type of logic family used, the physical arrangement of PC cards and mother boards, and the shielding enclosures. A rather typical arrangement will be described in the following section. This arrangement is typical of existing practices in data processing equipment, and is expected to appear increasingly in avionics as data processing becomes an ever larger part of the avionics package.

Description Of A Generalized Data Processing Unit. A typical data processing unit might consist of a card cage containing up to 16 PC cards, each connected to a mother board. One of the cards will contain the high frequency clock used to generate timing pulses. Distribution of the clock pulses takes place on microstrip lines on the cards. These lines extend through connectors to the mother board and then to other cards.

The microstrip lines are typically designed to have a characteristic impedance of 93 ohms. On a glass-epoxy G-10 material board with a thickness of 0.032 inches and a dielectric constant of 4.5, the trace width will be about 0.015 inches.

Each PC card could have as much as 3 inches of trace length, the length on the mother board could be about 14 inches, and for a worst case analysis, we can assume 16 independent traces, each having a total length of 20 inches.

An off-to-on voltage of 5 volts has been assumed as typical of most logic circuits.

For the model to be analyzed, clock frequencies of 8 and 25 MHz were chosen. In the current state of digital systems development, these frequencies are neither very high or very low, and may be considered representative.

The frequency spectrum of a pulse is related to its time domain description through the Fourier transform. The mathematics of the Fourier transform may be found in most advanced calculus or engineering mathematics texts. Some of the features of the Fourier transform of a symmetrical trapezoidal pulse are shown in Figure 47. For many typical rise times and flat tops, harmonics up to the second or third will decrease at a rate of 20 dB per decade, relative to the fundamental frequency. Higher harmonics decrease at a rate of 40 dB per decade, and can rapidly become of negligible amplitude unless microstrip line or cable resonances occur. Although Figure 47 suggests a continuous noise spectrum, for clock pulses, which are the dominant noise source, noise will appear only at harmonics (integer multiples) of the fundamental clock frequency.

In this model, any of the parameters can be changed through application of the scaling laws developed in an earlier part of this report dealing with board level radiated emissions.

Table I shows the rise times associated with various commonly used logic families. These rise times were taken from data sheets in the National Semiconductor Corporation catalogs. While not every device in a given logic family has the rise time indicated, the values shown are representative.

TABLE 1. TYPICAL RISE TIMES FOR DIGITAL DEVICES

Logic Family	Rise Time, ns.
TTL	10
LSTTL	10
ALS	2
SCHOTTKY	5
HCMOS	4
ECL	2
AS	3
FAST	2

Analysis of the described model was performed using EMCad® software, a computer program which has been found to yield accurate results for electromagnetic compatibility analysis over a long period of time and for many different types of systems.

Radiated emission analyses for the various digital logic families are given in Tables 2 through 6 for the 8 MHz clock. Tables 7 through 11 show the results for the 25 MHz clock.

TABLE 2. RADIATED EMISSIONS FOR ALS, ECL AND FAST LOGIC,
8 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION LEVEL----			
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	89	92	95	37	58 OUT
25.0 MHZ	84	87	90	35	55 OUT
25.0 MHZ	84	87	90	35	55 OUT
147.6 MHZ	77	80	83	47	36 OUT
159.2 MHZ	76	79	82	47	35 OUT

HIGH RISK CUT OFF FREQUENCY = 7.5 GHZ
NOMINAL CUT OFF FREQUENCY = 10.5 GHZ
WORST CASE CUT OFF FREQUENCY = 14.9 GHZ
LARGEST RECOMMENDED HOLE SIZE = 1.008601E-02 METERS. (0.397")

TABLE 3. RADIATED EMISSIONS FOR AS LOGIC, 8 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 3 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	89	92	95	37	58 OUT
25.0 MHZ	84	87	90	35	55 OUT
25.0 MHZ	84	87	90	35	55 OUT
106.1 MHZ	78	81	84	45	40 OUT
147.6 MHZ	74	77	80	47	33 OUT

HIGH RISK CUT OFF FREQUENCY = 7.5 GHZ
NOMINAL CUT OFF FREQUENCY = 10.5 GHZ
WORST CASE CUT OFF FREQUENCY = 14.9 GHZ
LARGEST RECOMMENDED HOLE SIZE = 1.008601E-02 METERS. (0.397")

TABLE 4. RADIATED EMISSIONS FOR HCMOS LOGIC, 8 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS CONDUCTED EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 4 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	89	92	95	37	58 OUT
25.0 MHZ	84	87	90	35	55 OUT
25.0 MHZ	84	87	90	35	55 OUT
79.6 MHZ	79	82	85	43	43 OUT
147.6 MHZ	71	74	77	47	31 OUT

HIGH RISK CUT OFF FREQUENCY = 7.5 GHZ
NOMINAL CUT OFF FREQUENCY = 10.5 GHZ
WORST CASE CUT OFF FREQUENCY = 14.9 GHZ
LARGEST RECOMMENDED HOLE SIZE = 1.008601E-02 METERS. (0.397")

TABLE 5. RADIATED EMISSIONS FOR SCHOTTKY LOGIC, 8 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	89	92	95	37	58 OUT
25.0 MHZ	84	87	90	35	55 OUT
25.0 MHZ	84	87	90	35	55 OUT
63.7 MHZ	80	83	86	41	45 OUT
147.6 MHZ	69	72	75	47	31 OUT

HIGH RISK CUT OFF FREQUENCY = 7.5 GHZ
NOMINAL CUT OFF FREQUENCY = 10.5 GHZ
WORST CASE CUT OFF FREQUENCY = 14.9 GHZ
LARGEST RECOMMENDED HOLE SIZE = 1.008601E-02 METERS. (0.397")

TABLE 6. RADIATED EMISSION FOR TTL AND LSTTL LOGIC, 8 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B
COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 10 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION LEVEL----			
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	89	92	95	37	58 OUT
25.0 MHZ	84	87	90	35	55 OUT
25.0 MHZ	84	87	90	35	55 OUT
31.8 MHZ	83	86	89	37	53 OUT
147.6 MHZ	63	66	69	47	23 OUT

HIGH RISK CUT OFF FREQUENCY = 7.5 GHZ
NOMINAL CUT OFF FREQUENCY = 10.5 GHZ
WORST CASE CUT OFF FREQUENCY = 14.9 GHZ
LARGEST RECOMMENDED HOLE SIZE = 1.008601E-02 METERS. (0.397")

TABLE 7. RADIATED EMISSIONS FOR ALS, ECL AND FAST LOGIC, 25 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS CONDUCTED EMISSIONS-FAA
SIGNAL NAME: 25 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 25 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
25.0 MHZ	94	97	100	35	65 OUT
25.0 MHZ	94	97	100	35	65 OUT
147.6 MHZ	87	90	93	47	46 OUT
159.2 MHZ	86	89	92	47	45 OUT

HIGH RISK CUT OFF FREQUENCY = 23.3 GHZ
NOMINAL CUT OFF FREQUENCY = 32.9 GHZ
WORST CASE CUT OFF FREQUENCY = 46.5 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.227524E-03 METERS. (.127")

TABLE 8. RADIATED EMISSIONS FOR AS LOGIC, 25 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B
COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 25 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 3 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 25 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
25.0 MHZ	94	97	100	35	65 OUT
25.0 MHZ	94	97	100	35	65 OUT
106.1 MHZ	88	91	94	45	49 OUT
147.6 MHZ	84	87	90	47	43 OUT

HIGH RISK CUT OFF FREQUENCY = 23.3 GHZ
NOMINAL CUT OFF FREQUENCY = 32.9 GHZ
WORST CASE CUT OFF FREQUENCY = 46.5 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.227524E-03 METERS. (.127")

TABLE 9. RADIATED EMISSIONS FOR HCMOS LOGIC, 25 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B
COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 25 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 4 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 25 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION LEVEL----			
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
25.0 MHZ	94	97	100	35	65 OUT
25.0 MHZ	94	97	100	35	65 OUT
79.6 MHZ	89	92	95	43	53 OUT
147.6 MHZ	81	84	87	47	40 OUT

HIGH RISK CUT OFF FREQUENCY = 23.3 GHZ
NOMINAL CUT OFF FREQUENCY = 32.9 GHZ
WORST CASE CUT OFF FREQUENCY = 46.5 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.227524E-03 METERS. (.127")

TABLE 10. RADIATED EMISSIONS FOR SCHOTTKY LOGIC, 25 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 25 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 25 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
25.0 MHZ	94	97	100	35	65 OUT
25.0 MHZ	94	97	100	35	65 OUT
63.7 MHZ	90	93	96	41	55 OUT
147.6 MHZ	79	82	85	47	39 OUT

HIGH RISK CUT OFF FREQUENCY = 23.3 GHZ
NOMINAL CUT OFF FREQUENCY = 32.9 GHZ
WORST CASE CUT OFF FREQUENCY = 46.5 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.227524E-03 METERS. (.127")

TABLE 11. RADIATED EMISSIONS FOR TTL AND LSTTL LOGIC, 25 MHz CLOCK

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM PRINTED CIRCUIT BOARD TRACES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB USING PROGRAM RTCA DO-160B
COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 25 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 10 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 25 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF TRACES: 20 INCHES
TYPE OF RETURN (ADJACENT, STACKED, GROUND PLANE): GROUND PLANE
NUMBER OF TRACES: 16
DISTANCE BETWEEN SIGNAL AND RETURN: .032 INCHES
TRACES WIDTH: .015 INCHES
DIELECTRIC CONSTANT 4.5
DISTANCE TO GROUND OR GROUND PLANE: .032 INCHES
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
25.0 MHZ	94	97	100	35	65 OUT
25.0 MHZ	94	97	100	35	65 OUT
31.8 MHZ	93	96	99	37	63 OUT
147.6 MHZ	73	76	79	47	32 OUT

HIGH RISK CUT OFF FREQUENCY = 23.3 GHZ
NOMINAL CUT OFF FREQUENCY = 32.9 GHZ
WORST CASE CUT OFF FREQUENCY = 46.5 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.227524E-03 METERS. (.127")

TABLE 12. RADIATED EMISSIONS FOR HIGH RESOLUTION VIDEO DISPLAY UNIT

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: HIGH RESOLUTION VIDEO
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 40 MHZ
WAVEFORM AMPLITUDE (V OR A): 30 VOLTS
LENGTH OF INTERCONNECT: 4 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): UNSHIELDED
NUMBER OF WIRES IN INTERCONNECT: 1
DISTANCE BETWEEN SIGNAL AND RETURN: 2 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 20 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
40.0 MHZ	92	95	98	38	60 OUT
63.7 MHZ	90	93	96	41	55 OUT
738.2 MHZ	58	61	64	58	6 OUT

HIGH RISK CUT OFF FREQUENCY = 19.3 GHZ
NOMINAL CUT OFF FREQUENCY = 27.3 GHZ
WORST CASE CUT OFF FREQUENCY = 38.6 GHZ
LARGEST RECOMMENDED HOLE SIZE = 3.886009E-03 METERS. (0.153")

An analysis was also performed for a high resolution video display unit. This unit is characterized by a 40-MHz bandwidth, a peak video signal of 30 volts, 4 inches of unshielded connecting wire, and a rise time of 5 nanoseconds. The results of this analysis are given in Table 12.

Results of Analysis. For all rise times between 2 and 10 nanoseconds, there is no difference in the radiated emissions up to 25 MHz, for the 8-MHz clock. Differences show up in the 150-MHz region, but still represent factors of only 4 in field strength (12-13 dB).

The 25-MHz clock shows consistently about 10-dB higher emissions, as would be expected from the frequency scaling laws developed earlier.

Since the higher frequency components decrease in amplitude at a rate of 40 dB per decade, all of the logic families are either within specification or close to it at about 1500 MHz.

The necessary case shielding can be found directly from the analysis tables. The shielding requirement is exactly the dB value by which the model fails to meet the specification, as shown in the "Status" column. For a 2-nanosecond rise time and the 8-MHz clock, 58 dB of shielding is required at 8 MHz, and slightly less than that at higher frequencies. For the same rise time, but using a clock frequency of 25 MHz, 65 dB of shielding is required. These shielding values can be obtained with well designed enclosures.

Any wiring or cables which penetrate the shield will themselves have to be shielded, or if not shielded, must be filtered to the same 58 or 65 dB level that the shielding must provide.

CASE APERTURE SIZING. It will be noted that the last data line in the tables gives the largest recommended hole size. This is the dimension of the largest allowable single hole in the shielding enclosure which will just allow the unit to meet the radiated emission specification. Since leakage increases as the square root of the number of the holes, it is evident that for the model analyzed, the shielding integrity will have to be very good. Very small holes, or a screen-mesh type of material will have to be used.

The largest hole size also suggests the spacing that can be permitted between screws on an enclosure seam. An enclosure seam, shorted at each end by screws, forms a very effective slot antenna. In the model analyzed, a seam held together with screws will not meet specifications, since the spacings of 0.397 and 0.127 inches are too small to be practical. Instead, some form of continuous contact between the seam edges will have to be provided, via finger stock or gasketing. For durability, a woven wire gasketing material is preferable to finger stock.

Allowable hole sizes and distances between screws are determined in the following way: When the hole or distance between screws is equal to a half wavelength, the shielding effectiveness is assumed to be zero, on the basis that the hole or seam becomes resonant and radiates all of the energy incident upon it with 100% efficiency. Next, the field strength is assumed to decrease linearly with increasing wavelength. This leads to a 20 dB per decade decrease in the field strength as the frequency of excitation goes down. A convenient rule of thumb derived from these considerations is that a hole of one inch diameter will have a shielding effectiveness of 40 dB at 60 MHz.

A common way of providing ventilation is to use a square array of round holes in an enclosure wall. Figure 70 shows such an array, in which the hole diameter is d, the center to center spacing of the holes is c, and the length and width of the array is L. For rectangular rather than square patterns, the value of L is the geometric mean of the length and width of the pattern. The thickness of the wall material is t. If the hole diameter is less than one-sixth of a wave length, the magnetic field shielding effectiveness is:

$$S = 20 \log (c^2 l/d^3) + 32t/d + 3.8\text{dB}$$

with c = 0.375", d = 0.125", L = 6", and t = 0.032", the shielding effectiveness is 64.7 dB up to 15 GHz. This hole pattern would be appropriate for the model analyzed above.

INTERCONNECT CABLING. The interconnect cabling between subsystems is of concern because radiated and conducted emissions can take place from these cables.

Cable Radiated Emissions. As in the case of the radiated emission from PC board traces, analyses were performed for a clock signal of 8 MHz, using rise times of 2, 3, 4, 5, and 10 nanoseconds. The clock signal was assumed to be transmitted on 24 inches of number 18 AWG unshielded twisted pair wire. The results are shown in Tables 13 through 17. For comparison, an analysis was also run with the 2 ns. signal on coaxial line. The result of this analysis is shown in Table 18.

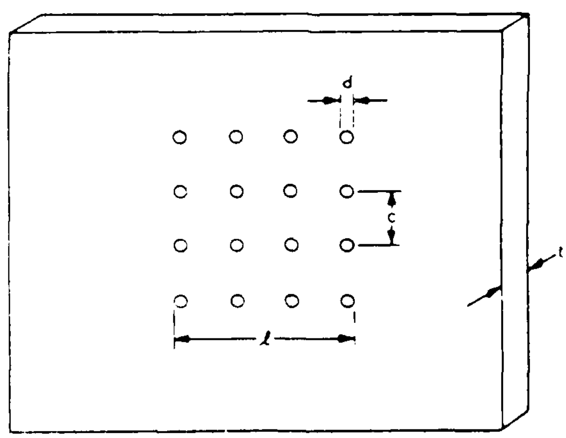


FIGURE 70. SHIELDING CONTAINING A SQUARE ARRAY OF ROUND HOLES

TABLE 13. INTERCONNECTING CABLE EMISSIONS FOR 2 ns RISE TIME PULSE ON TWISTED PAIR

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B
COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): TWISTED
NUMBER OF WIRES IN INTERCONNECT: 2
DISTANCE BETWEEN SIGNAL AND RETURN: .05 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	63	66	69	37	32 OUT
25.0 MHZ	58	61	64	35	29 OUT
25.0 MHZ	58	61	64	35	29 OUT
123.0 MHZ	51	54	57	46	12 OUT
159.2 MHZ	49	52	55	47	8 OUT

HIGH RISK CUT OFF FREQUENCY = 363.9 MHZ
NOMINAL CUT OFF FREQUENCY = 514.0 MHZ
WORST CASE CUT OFF FREQUENCY = 726.1 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2065952 METERS.

TABLE 14. INTERCONNECTING CABLE EMISSIONS FOR 3 ns RISE TIME PULSE ON TWISTED PAIR

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 3 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): TWISTED
NUMBER OF WIRES IN INTERCONNECT: 2
DISTANCE BETWEEN SIGNAL AND RETURN: .05 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	63	66	69	37	32 OUT
25.0 MHZ	58	61	64	35	29 OUT
25.0 MHZ	58	61	64	35	29 OUT
106.1 MHZ	52	55	58	45	13 OUT
123.0 MHZ	50	53	56	46	10 OUT

HIGH RISK CUT OFF FREQUENCY = 363.9 MHZ
NOMINAL CUT OFF FREQUENCY = 514.0 MHZ
WORST CASE CUT OFF FREQUENCY = 726.1 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2065952 METERS.

TABLE 15. INTERCONNECTING CABLE EMISSIONS FOR 4 ns RISE TIME PULSE ON TWISTED PAIR

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-27-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 4 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): TWISTED
NUMBER OF WIRES IN INTERCONNECT: 2
DISTANCE BETWEEN SIGNAL AND RETURN: .05 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	63	66	69	37	32 OUT
25.0 MHZ	58	61	64	35	29 OUT
25.0 MHZ	58	61	64	35	29 OUT
79.6 MHZ	53	56	59	43	17 OUT
123.0 MHZ	48	51	54	46	8 OUT

HIGH RISK CUT OFF FREQUENCY = 363.9 MHZ
NOMINAL CUT OFF FREQUENCY = 514.0 MHZ
WORST CASE CUT OFF FREQUENCY = 726.1 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2065952 METERS.

TABLE 16. INTERCONNECTING CABLE EMISSIONS FOR 5 ns RISE TIME PULSE ON TWISTED PAIR

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 5 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): TWISTED
NUMBER OF WIRES IN INTERCONNECT: 2
DISTANCE BETWEEN SIGNAL AND RETURN: .05 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	63	66	69	37	32 OUT
25.0 MHZ	58	61	64	35	29 OUT
25.0 MHZ	58	61	64	35	29 OUT
63.7 MHZ	54	57	60	41	19 OUT
123.0 MHZ	46	49	52	46	6 OUT

HIGH RISK CUT OFF FREQUENCY = 363.9 MHZ
NOMINAL CUT OFF FREQUENCY = 514.0 MHZ
WORST CASE CUT OFF FREQUENCY = 726.1 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2065952 METERS.

TABLE 17. INTERCONNECTING CABLE EMISSIONS FOR 10 ns RISE TIME PULSE ON TWISTED PAIR

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF NB

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 10 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): TWISTED
NUMBER OF WIRES IN INTERCONNECT: 2
DISTANCE BETWEEN SIGNAL AND RETURN: .05 INCHES
DISTANCE TO GROUND OR SHIELD: 2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	63	66	69	37	32 OUT
25.0 MHZ	58	61	64	35	29 OUT
25.0 MHZ	58	61	64	35	29 OUT
31.8 MHZ	57	60	63	37	27 OUT
123.0 MHZ	40	43	46	46	0 OUT

HIGH RISK CUT OFF FREQUENCY = 363.9 MHZ
NOMINAL CUT OFF FREQUENCY = 514.0 MHZ
WORST CASE CUT OFF FREQUENCY = 726.1 MHZ
LARGEST RECOMMENDED HOLE SIZE = .2065952 METERS.

TABLE 18. INTERCONNECTING CABLE EMISSIONS FOR 2 ns RISE TIME PULSE ON COAXIAL CABLE

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS PREDICTS THE RADIATED EMISSIONS FROM CABLES AND INTERCONNECT WIRES AND IS BASED ON THE EMISSION REQUIREMENTS OF MIL-STD-461 A BASIC RE02 NB USING PROGRAM ERE3001B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS EMISSIONS-FAA
SIGNAL NAME: 8 MHZ CLOCK
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 NANOSEC
WAVEFORM FREQUENCY OF OSCILLATION: 8 MHZ
WAVEFORM AMPLITUDE (V OR A): 5 VOLTS
LENGTH OF INTERCONNECT: 24 INCHES
TYPE (UNSHIELDED, TWISTED, SHIELDED, COAX): COAX
NUMBER OF WIRES IN INTERCONNECT: 1
DISTANCE BETWEEN SIGNAL AND RETURN: .2 INCHES
DISTANCE TO GROUND OR SHIELD: .2 INCHES
WIRE SIZE (1-40 AWG OR DIAMETER): 18 AWG
TEST (MEASUREMENT) DISTANCE: 1 METERS

	----PREDICTED	EMISSION LEVEL----			
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
8.0 MHZ	26	29	32	37	-5
25.0 MHZ	31	34	37	35	2 OUT
25.0 MHZ	31	34	37	35	2 OUT
123.0 MHZ	38	41	44	46	-2
159.2 MHZ	38	41	44	47	-3

WORST CASE CUT OFF FREQUENCY = 32.2 MHZ
LARGEST RECOMMENDED HOLE SIZE = 4.658532 METERS.

TABLE 19. CONDUCTED EMISSIONS, 1 AMP 400 Hz POWER SUPPLY

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS IS BASED ON THE CONDUCTED EMISSION REQUIREMENTS OF RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS CONDUCTED EMISSIONS-FAA
SIGNAL NAME: POWER SUPPLY
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 MICROSEC

WAVEFORM FREQUENCY OF OSCILLATION: 400 HZ

WAVEFORM AMPLITUDE (V OR A): 1 AMPS

OUTPUT (DIFFERENTIAL, COMMON, TOTAL) TOTAL

RIPPLE FILTER (Y/N) NO

	----PREDICTED	EMISSION LEVEL----			
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
159.2 KHZ	60	66	78	54	24 OUT
2.0 MHZ	7	13	25	20	5 OUT
10.0 MHZ	-21	-15	-3	20	-23
50.0 MHZ	-49	-43	-31	20	-51

TABLE 20. CONDUCTED EMISSIONS, 3 AMP 400 Hz POWER SUPPLY

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS IS BASED ON THE CONDUCTED EMISSION REQUIREMENTS OF RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS CONDUCTED EMISSIONS-FAA
SIGNAL NAME: POWER SUPPLY
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 MICROSEC

WAVEFORM FREQUENCY OF OSCILLATION: 400 HZ

WAVEFORM AMPLITUDE (V OR A): 3 AMPS

OUTPUT (DIFFERENTIAL, COMMON, TOTAL) TOTAL

RIPPLE FILTER (Y/N) NO

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
159.2 KHZ	70	76	88	54	33 OUT
2.0 MHZ	16	22	34	20	14 OUT
10.0 MHZ	-12	-6	6	20	-14
50.0 MHZ	-40	-34	-22	20	-42

TABLE 21. CONDUCTED EMISSIONS, 10 AMP 400 Hz POWER SUPPLY

EMCAD RADIATED EMISSION ANALYSIS 4 SYMMETRICAL TRAPEZOIDAL WAVEFORM

THIS EMCAD ANALYSIS IS BASED ON THE CONDUCTED EMISSION REQUIREMENTS OF RTCA DO-160B

COMPANY: CK CONSULTANTS
PROJECT: DIGITAL SYSTEMS CONDUCTED EMISSIONS-FAA
SIGNAL NAME: POWER SUPPLY
DATE (M/D/Y): 4-28-87
ANALYSIS PERFORMED BY: RAM

THE INPUT VARIABLES USED FOR THIS ANALYSIS ARE AS FOLLOWS

WAVEFORM RISE/FALL TIME: 2 MICROSEC

WAVEFORM FREQUENCY OF OSCILLATION: 400 HZ

WAVEFORM AMPLITUDE (V OR A): 10 AMPS

OUTPUT (DIFFERENTIAL, COMMON, TOTAL) TOTAL

RIPPLE FILTER (Y/N) NO

	----PREDICTED	EMISSION	LEVEL----		
FREQUENCY	HIGH RISK	NOMINAL	WORST CASE	SPEC LIMIT	STATUS
159.2 KHZ	80	86	98	54	44 OUT
2.0 MHZ	27	33	45	20	25 OUT
10.0 MHZ	-1	5	17	20	-3
50.0 MHZ	-29	-23	-11	20	-31

The radiated emissions from interconnecting cables were on the order of 32 dB out of specification at 8 MHz for the unshielded twisted pair. This value can be seen in the "Status" column of Tables 13 through 17. Shielding of this value (32 dB) will be required to meet the specifications of DO-160B. This value of shielding is readily attainable. However, an 8-MHz clock signal would probably not be distributed on 24 inches of unshielded twisted pair, but this model was chosen as a deliberately poor design to produce a worst case result.

The improvement shown in Table 18, where the distribution is via a coaxial cable, is striking. Here the signal is within the specification except at 25 MHz, where it is out by 2 dB. The coaxial line is assumed to be RG-58/U, or equivalent, with 70 dB shielding effectiveness at 1 MHz, sloping at 20 dB per decade above 1 MHz, so that at 10 MHz the cable has degraded to 50 dB of shielding effectiveness, and at 100 MHz to 30 dB. RG-58/U is, then, suitable for connections between subsystems within enclosures, but a higher quality cable would be required for interconnection between systems, when the cable is not enclosed by other shielding. A foil plus braid cable, or a semi-rigid line with a solid metallic outer conductor would be required for cables running outside of enclosures.

Cable Conducted Emissions. Conducted emissions, particularly those arising from switching power supplies, are also a source of noise, and have been analyzed. A 400 Hz supply is assumed. With typical bridge rectifiers, in the absence of an input transformer, rise times are typically 2 microseconds. Three analyses were performed, for power supply input current values of 1, 3, and 10 amps. The results are shown in Tables 19, 20, and 21.

From Tables 19, 20, and 21 it can be seen that conducted emissions for even modest power supply currents are out of specification, and therefore line filtering is a necessity. At 150 kHz, where the requirements of DO-160B start, conducted emissions are generally dominated by the common mode, and appear as line to ground signals.

The line filter shown in Figure 27 earlier in this report would be appropriate to reduce common mode conducted emission. At one amp, this filter will provide 37 dB of common mode filtering at 150 kHz, and more than 100 dB at 25 MHz. Although the conducted emissions are higher for the higher current supplies, the filter provides more attenuation because of the reduced input impedance of these supplies. At 3 amps the attenuation is 42 dB and at 10 amps better than 45 dB. The attenuation of this filter is sufficient in all three cases to bring the conducted emissions within specification. Power supply currents higher than 10 amps will require higher values of L_1 and C_2.

CONCLUSIONS

1. This report gives background theory and practice for minimizing the response of digital systems to interfering electromagnetic fields. These fields can be of many different types: steady state or pulsed, one-time or repetitive, predominantly electric, or predominantly magnetic, near field or far field, and extending from very low to very high frequencies.

2. Until further radiated field intensity data is collected, a protection level of 200 volts per meter is recommended for flight critical systems.

3. The microwave landing system should be tested for immunity to the high energy threat.

4. Protection of the navigation band of 108.0 - 117.9 MHz should be upgraded through renewed attention to receiver design, and in view of new threats in Europe from increased FM transmitter power and increased frequency allocations.

5. Test procedures should be modified to include modulation of test signals at all frequencies and in all modulation modes used in flight critical systems. For example, 30 Hz is used in VOR systems, and 90 and 150 Hz in the ILS.

6. Alternative measurement procedures should be developed to simulate 200-volt per meter far field tests to avoid the extreme costs associated with a wide band high power test facility.

7. There will be a need for periodic re-verification of aircraft compliance to susceptibility requirements.

8. It may be necessary to define new prohibited airspace until both the extent of the electromagnetic threat and systems susceptibility are better understood. (Recently the West Germans requested a safe flight route for a visiting Tornado fighter. Are we able to provide such information?)

9. International coordination must be brought about through coordination by the FAA with its counterpart organizations in foreign countries.

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GLOSSARY

Absorption - Absorption is the loss of energy in the transmission of waves over radio or wire paths due to conversion into heat or other forms of energy. In wire transmission, the term is usually applied only to loss of energy in extraneous media.

AC Noise Immunity - The relative immunity of a device to noise pulses. As pulse width increases, AC noise immunity approaches the DC noise margin. As the pulse width is reduced toward the propagation delay time, the circuits become less able to respond and a larger noise voltage is required. See Noise-Energy Immunity; DC Noise Margin.

Attenuation - Attenuation is the general term used to denote a decrease in magnitude in transmission from one point to another. It may be expressed as a ratio or by extension of the term in decibels.

ALS - Advanced Low Power Schottky Logic

AS - Advanced Schottky Logic

B-Field - See Magnetic Flux Density

Balanced - In communication practice, the term balanced signifies (1) electrically alike and symmetrical with respect to ground or (2) arranged to provide conjugacy between certain sets of terminals.

Bandpass Filter - A bandpass filter is a wave filter which has a single transmission band, neither of the cutoff frequencies being zero or infinite.

Bonded - Electrically connected by means of welding, brazing, compressing or other mechanical means not likely to be affected by heat or corrosion and which will give a good low-impedance (< 0.1 ohm) electrical path.

Bonding - Electrically connecting two mechanical parts together such that the voltage potential across the joints is very small (millivolts).

Broadband Emissions - Term used to describe emissions which have a bandwidth greater than some reference bandwidth. The reference bandwidth may be that of a victim receptor or that of an EMI measurement receiver.

Broadband Emissions are subclassified as Coherent and Incoherent. A coherent broadband emission source has harmonics that are related to the fundamental frequency in amplitude and phase. A typical example of a coherent broadband source is a digital logic signal. An incoherent broadband emission source has harmonics that are not related in amplitude and phase (random) to the fundamental harmonic. Typical examples include lightning and noisy diodes.

When specifying broadband sources, the reference bandwidth must be specified since a larger reference bandwidth will result in a higher specified level. Broadband signals are therefore specified per reference bandwidth. For example, MILSTD-461 broadband signals are generally specified per one megahertz bandwidth, so that a coherent broadband noise source that results in a detected signal level of 100 microvolts (i.e., the sum of all the harmonics over a 1 MHz bandwidth is equal to 100 microvolts) is expressed as 100 microvolts per MHz bandwidth (100 uV/MHz). If the ratio of this level is compared to 1 uV/MHz, one would get:

$$20 \text{ Log } (100/1) = 40 \text{ dBuV/MHz}$$

Capacitive Coupling - The effect which one circuit or subsystem has on another by means of the capacitance that exists between them. The amount of coupling is a function of the noise frequency and the capacitive impedance between the circuits.

Characteristic Impedance - (Assuming Transverse Electromagnetic Wave) The ratio of the root mean square voltage to the current flowing from the source when the conductor they appear on is terminated in the impedance which results in minimum standing wave ratio on the line. Used in transmission line discipline.

Circular Mils (abbr. Cir Mil) - The diameter of a cylinder squared. 1 mil = .001 inch. When multiplied by $\pi/4$, equals the area of the circle in square mils.

Common Impedance Coupling - The means by which two or more circuits sharing a common impedance share a common voltage drop across the impedance, thereby coupling it to the secondary source(s). Conducted interference.

Conducted Emissions - Frequencies that are generated within the system and appear on the power line through internal radiation, capacitive or inductive coupling, or direct conduction through the power supply and signal lines.

Conducted Interference - Interference that propagates to and from circuits through interconnecting conductors. Complement: radiated interference.

Conducted Susceptibility - The response of a system to the presence of noise on the power and signal lines. The noise can be either line-to-line (differential mode) or line-to-ground (common mode).

Crosstalk - The interference energy transferred from a source to an unintentional load by conductive, inductive or capacitive coupling. Usually of concern when source and susceptor are less than 1/6 wavelength apart. Referred to as Near End or Far End Crosstalk depending on relation of where it is measured, with respect to the source.

Culprit - Signal or circuit that is the source of interference.

dB - Decibel - One tenth of a Bel. If P_1 and P_2 designate two power levels and N their decibel ratio, then

$$N = 10 \text{ Log } (P_2/P_1) \text{ decibels}$$

If P1 is referenced at 1 mW, N becomes dBm at a specified impedance (usually 50 or 600 ohms).

If the ratios of voltages or currents can be expressed as the square roots of the corresponding power ratios, i.e., the impedances associated with P_1 and P_2 are the same, then the ratio N can be expressed as

$$N = 20 \text{ Log } (I_2/I_1) \text{ decibels}$$

$$N = 20 \text{ Log } (V_2/V_1) \text{ decibels}$$

where I_2/I_1 and V_2/V_1 are the given current and voltage ratios, respectively.

dBm - A unit of expression of power level in decibels where the power level is compared to a reference power level of one milliwatt into 50 ohms (sometimes referenced to 600 ohms). For example, a power level of 100 milliwatts is equal to

$$10 \text{ Log } (100 \text{ milliwatts}/1 \text{ milliwatt}) = 20 \text{ dBm}$$

dBuA - A unit of expression of current level in decibels where the current level is compared to a reference level of one microamp. For example, a current level of one milliamp is equal to 20 Log (1 milliamp/1 microamp) = 60 dBuA

dBuV - A unit of expression of voltage level in decibels where the voltage level is compared to a reference level of one microvolt. For example, a voltage level of 1 volt is equal to

$$20 \text{ Log } (1 \text{ volt}/1 \text{ microvolt}) = 120 \text{ dBuV}$$

dBuV/m - A unit of expression of electric field strength level where the field strength level is compared to a reference level of one microvolt per meter (1 uV/m). For example, a field strength of 1 volt/meter is equal to 120 dBuV/m (see dBuV).

dBuV/MHz - A unit of expression of a broadband voltage level where the voltage level is compared to a reference level of one microvolt per megahertz bandwidth (see Broadband Emissions).

DC Noise Margin - The difference between the maximum input voltage [V_{IL} (max)] that will still be interpreted as a "0" and the maximum output voltage that can be provided by the source in the "0" state [V_{OL} (max)]. Referred to as the low-level signal line noise margin, V_{NSL} (min) = V_{IL} (max) - V_{OL} (max).

Decoupling - Filtering, the removing of the coupling effect usually referring to conductive coupling. Decoupling capacitors are used on the power supply to isolate the effects of the load on the distribution system or vice-versa (also Bypass Capacitor).

Dielectric - A material capable of passing electric lines of force but not conductive.

Dual In-line Package - DIP - A pin configuration, common on integrated circuits, where the pins are aligned in two parallel, and opposing rows.

E-Field - See Electric Field Intensity

Earth - With respect to charge, the earth is considered an infinite sink, and, therefore, the lowest possible zero potential reference.

ECL - Emitter coupled logic, a high speed logic family.

Electric Field Intensity - Deprecated name for Electric Field Strength, the magnitude of the electric field vector at a point in space, caused by current flow.

Electromagnetic Compatibility (EMC) - The capability of electronic equipment to be operated as designed in the intended operational electromagnetic environment.

Electromagnetic Interference (EMI) - Degradation of the electromagnetic environment by an unwanted disturbance.

Electrostatic Discharge - (ESD) - Arc discharge across an air gap caused by the electric potential across the gap exceeding the dielectric breakdown voltage of air in the gap.

Emitter - Source of electromagnetic signal.

FAST - Fairchild Advanced Schottky Logic (trademark).

Far Field Radiation - Electromagnetic radiation that occurs at a distance greater than approximately 1/6 wavelength from the source of interference. At these distances the electromagnetic field decreases at a rate inversely proportional to the distance from the source and the ratio of E to H = 377 ohms in free space.

Ground - 1. Voltage reference which is eventually related back to the potential of the earth. Because of path resistance, usually not at earth potential. Ground is usually of "local" value such that all grounded equipment in immediate area is the "same" potential, irrespective of earth.
2. Low impedance return path for signal or power currents. Deprecated version of "Return."
3. Electron source or sink for the dissipation of electrostatic discharge (ESD) currents.

Ground plane - An equipotential voltage reference to which electronic units are referenced so that the voltage potential between interconnected units is less than or equal to an acceptable noise level.

HCMOS - High Speed Complementary Metal Oxide Semiconductor.

H-Field - See Magnetic Field Intensity.

I.C. - Integrated Circuit

Inductive Coupling - The detrimental effect which one circuit or subsystem has on another by means of the mutual inductance that exists between them. The amount of coupling is a function of the frequency, the loop area of the systems, and the length and proximity of parallel interacting wiring runs.

Loop - The current caused by a difference in voltage between two points. Occurs only when there are two or more paths between the two points.

Low Frequency - Those frequencies whose wavelength is 16 times greater than the largest circuit dimension (rule of thumb). Also 0 to 10 kHz. (Use whichever is smaller).

LSTTL - Low Power Schottky Transistor-Transistor Logic.

Magnetic Field Intensity - Deprecated name for magnetic field strength, the magnitude of the magnetic field vector at a point in space caused by current flow.

Magnetic Flux Density - The field intensity integrated over the area in question.

Matched Transmission Line - A transmission line is said to be matched at any transverse section if there is no wave reflection at that section (i.e., the line impedance equals the load impedance).

Multilayer - Referring to a printed circuit board which is constructed by laminating two or more layers of etched conductor and substrate together, and connecting layers by plated holes. The conductive layers have dedicated functions such as power, ground, signals laid out in the "X" or "Y" direction, or shield.

Multistacked - Multistacked PCB's are complete double- or single-sided PCB's which are riveted together through holes intended to conduct current. Rivets replace plated through-holes and hold board layers together. These boards may be repaired (where multilayer cannot) by removing the rivets. Not used very often in commercial electronics.

Multiwire (TM) - A PCB fabrication technique where power and ground are etched on a single- or double-sided PCB, then coated with adhesive; wire is laid onto the adhesive using a computer numerical controlled wiring machine. The holes are then drilled and plated to make connection to the wires. Multiwire can achieve higher density than multilayer boards and it is simple to make changes. It is relatively expensive and not usually economical for production applications.

Near Field Radiation - Electromagnetic radiation that occurs at a distance less than approximately 1/6 wavelength from the source of interference. Electric fields (E-Fields) in the near field attenuate inversely proportional to the cube of the distance for high-impedance sources and $1/r^2$ for low-impedance sources. Near field magnetic (H-Fields) attenuate as the inverse square of the distance from the source for high impedance sources and $1/r^3$ for low-impedance sources.

Noise Energy Immunity - EN_2, equals the noise voltage required to cause a circuit to malfunction squared $(Vn)^2$, times the noise pulse-width (PW), divided by the parallel equivalent of the input under evaluation and the impedance of the wiring and circuit connected to it (R_o). (Motorola App. Note 707)

$$EN = (Vn)^2/R_o \times (PW)$$

Not usually applied to power and ground leads.

Noise Immunity - See AC Noise Immunity, also Noise-Energy Immunity.

Noise Margin - See DC Noise Margin, also AC Noise Immunity.

PCB - Printed Circuit Board. A substrate with conductors etched out of a conductor bonded to it, used to support and interconnect circuit components. May also refer loosely to multiwire and wire wrap boards.

Pickup - Interference from a nearby circuit or system. Also crosstalk.

Printed Circuit Board - PCB, also Printed Wiring Board (PWB). A substrate of rigid material with conductive paths (traces) for current etched in or fastened to it for the purpose of supporting and interconnecting circuit components.

Propagation Delay - The delay in moving a signal from Point A to Point B through a medium or circuit. The delay is caused by the energy storing characteristics of the medium (such as capacitance or inductance) that need to be satisfied before the next increment of medium in the path can be excited. Each increment requires a finite delay.

Radiated Emissions - The undesired electromagnetic fields generated by electronic or electric circuitry.

Radiated Susceptibility - The response of a system to the presence of electromagnetic waves.

Receptor - Signal or circuit which responds to electromagnetic noise.

Return - The low impedance conductor(s) used to direct current from the load back to the source. Does not have to be ground or earth potential (i.e., balanced line). See Earth, Ground.

Schottky Logic - Logic circuits whose transistors have a metal-semi-conductor diode built in between the collector and base. This diode prevents the transistor from saturating, thereby greatly increasing its switching speed.

Shielding - Reducing EMI by interposing a conductive or absorptive blockage (shield) between a noise and a receptor.

Skin Depth - The distance from a conductor's outside surface to where the current density in the conductor falls to 1/e of its surface value.

Susceptibility - System or subsystem that is not resistant to electromagnetic interference.

System - A group of components, circuits or equipment connected together to provide a complementary function, and thereby having to operate without interference in a mutual electromagnetic environment.

T.E.M. - See Transverse Electromagnetic Mode.

Trace - The conductor bonded to a printed circuit board substrate that conducts current between two or more points.

Transmission Line - A combination of conductors arranged in a mutual field, having the characteristics of distributed reactance, and uniform impedance along its length when properly terminated.

TTL - Transistor-Transistor Logic.

SYMBOLS

1. c velocity of light in meters per second
2. dB decibels
3. E electric field in volts per meter
4. e_r relative dielectric constant (also k)
5. f frequency in Hertz
6. h height
7. H magnetic field in amps per meter
8. I current in amps
9. k relative dielectric constant (also e_r)
10. L inductance in Henrys
11. ℓ length
12. N,n integer
13. ns nanosecond
14. pf picofarad
15. t time
16. T_D delay time
17. t_r rise time
18. uf microfarad
19. uH microhenry
20. us microsecond
21. uV microvolt
22. V volt
23. w width
24. λ lambda : wavelength
25. π pi:3.14
26. ω omega : radian frequency
27. Z impedance
28. Z_0 characteristic impedance

RULES FOR DIGITAL SYSTEMS DESIGN

INTRODUCTION. The principles presented herein are reduced to rules or guidelines. The organization of this section parallels that of the section on Digital Systems Susceptibility, so that if the reader wishes to investigate the basis for a particular rule, they may do so by referring to the parallel section of the preceding text.

Shielding

1. For shielding of low frequency magnetic fields, use thick lossy materials such as steel or mu-metal.

2. For high frequency magnetic fields use copper or aluminum shield material.

3. Apertures for ventilation, and slots and seams in shielding should be kept to dimensions less than a half wavelength at the highest frequency of emission.

4. Connections between shielding materials should be through direct metal-to-metal contact, not just through screws.

5. Screws should be a maximum of 3 inches apart. Closer spacing is helpful, if it is practicable.

6. Protective coatings should be reviewed carefully for conductivity. Anodized materials should not be used in shielding applications.

7. Bonding straps should be of solid material rather than braid, if possible.

8. Bonding straps should have a length not greater than five times the width, and a minimum thickness of 0.020 inch.

9. Bonding straps should make clean metal-to-metal contact at both ends.

10. The application of paint to shielding surfaces should be reviewed carefully to make sure that contact between shielding surfaces is not impaired.

11. A hole one inch in diameter provides 40 dB of shielding at 60 MHz, and degrades at 20 dB per decade with increasing frequency.

12. A shield one skin depth thick provides approximately 9 dB of absorption loss. Skin depth at 500 kHz is approximately 0.004 inch.

13. A large number of small holes gives better shielding than a smaller number of large holes.

14. All mating seams used in shielding should have a minimum overlap of 1.2 inches.

15. No conductor should enter a shielded enclosure unless it is bypassed at the point of entry.

16. No shielded cable should enter a shielded enclosure unless the shield is connected to the enclosure shield, or bypassed to it.

17. All metal components or assemblies within a shielded enclosure should be bonded together.

18. Piano-type hinges should not be used for achieving electrical bond.

19. Shields can be mounted directly on boards. However, such shields should not be used as signal returns or jumpers.

20. Shielding material should be at least 0.032 inch thick aluminum or copper.

21. The resistance between sections of shielding should be 2.5 milliohms or less.

22. Metex metallic woven mesh material or beryllium-copper finger stock is useful in gasketing and bonding applications.

23. Do not splice cable shielding.

Grounding

1. Grounds should be located so as to minimize the loop area between a signal path and the return path.

2. Logic grounds, which are noisy, should be kept separate from analog and signal grounds.

3. The most sensitive signal returns should be connected closest to the final equipotential point.

4. Every ground indicated on a schematic diagram should be precisely defined in location, not left to chance.

5. The length of a ground conductor should be kept shorter than 1/20 of a wavelength.

6. For grounding at high frequencies, the ground connections should be made as short and direct as possible to a common reference plane.

7. The common reference plane should be of highly conductive material - copper or silver.

8. Safety grounds should go directly from the equipment back to the point of power entry. The safety ground should connect to the equipment within 1 1/2 inches of the power entry point.

9. The hardware ground should run from racks and cabinets to the point of attachment of the safety ground.

10. At high frequencies, the surface area of ground planes and conductors is important since most of the current flows on the surface due to skin effect.

11. Grounding schemes should be reviewed with the object of minimizing the flow of two or more return currents through a common grounding impedance.

12. In the grounding of a daughter board every return connection should be tied as directly as possible to the mother board ground reference plane.

13. In board-to-board connectors, signals and their returns should be on adjacent pins, to minimize loop area.

14. If there are vacant areas on a ground plane or power plane, no signal conductors should be allowed to cross the vacant area.

15. On single sided boards, power and return lines should run parallel with each other inside all IC's, not outside, to reduce loop area.

16. On two-sided boards, power and return traces should be run at right angles on opposite sides of the board to form a grid pattern.

17. Unused areas on boards should be filled with ground plane material.

18. Signals and their returns should be routed in parallel, either on the same side or on opposite sides of a board to reduce return impedance and loop area.

19. In a PCB connector, <u>at least</u> two return pins should be used, one at either end of the connector.

20. Do not run a ground trace completely around the perimeter of a board unless ground gridding is also part of the design.

21. Do not use metal card edge guides as grounds.

<u>Propagation</u>

1. For high frequency signals where pulse distortion is important, signal paths should be treated as transmission lines.

2. Transmission lines should be designed to keep their dimensions constant, to avoid sharp bends, and should be terminated in their characteristic impedance.

3. Constant plating thickness and uniform dielectric constant are important in maintaining the characteristic impedance of a transmission line.

Coupling

1. Avoid placing magnetic components (relays, inductors, toroids, transformers) close to each other, to avoid magnetic field coupling.

2. Signal traces should be separated from each other by guard traces or ground areas, to avoid transmission line coupling effects.

3. Where possible, use distance to separate signal traces and reduce coupling.

4. Where possible, run signal traces at right angles to reduce magnetic and electric coupling effects.

5. Bypassing and filtering can be used to reduce coupling.

Board Radiation

1. Keep clock and other high speed traces short, wide, and close to ground. A width-to-height ratio of 1:1 is usually favorable. Lengths should be less than 1/20 of a wavelength.

2. Keep pulse rise times as slow as possible.

3. Keep clock frequencies as low as possible.

4. Limit signal trace width to height ratios to less than 3:1. Ratios greater than 3:1 do not appreciably reduce radiation or improve susceptibility.

5. Use trace width to height ratios greater than 0.3:1. Below 0.3:1, impedances rise rapidly and radiation and susceptibility increase.

6. Keep traces in from the edge of the board by a distance equal to at least the height above ground.

7. If a change in direction for a trace is needed, change in steps of 45°,

8. Terminate critical lines in their characteristic impedance.

9. Use as low bandwidth components, signals and circuits as possible to reduce harmonic content.

10. Use low current devices where possible, to reduce magnetic field strength

11. Make power distribution traces as wide as possible.

12. Minimize the use of sockets in high frequency portions of a board.

13. Trace lengths can sometimes be reduced by a 180° re-orientation of an IC.

Board Conducted Emission and Susceptibility

1. Below 150 kHz, conducted interference is usually differential mode (line to line).

2. Above 150 kHz, conducted interference is usually common mode (line-to-ground).

3. Filtering and bypassing are the common ways of suppressing conducted interference.

4. Power cables are the primary source of conducted interference although any I/O cable can also contribute.

5. In I/O cables, every signal conductor should have its own return conductor.

6. A cable shield should be grounded at both ends,

7. Never use a cable shield as a return for any signal of any kind.

8. In some cases it may be necessary to use a shield on the input power cable.

9. Ferrite beads on individual signal leads are sometimes effective in reducing high frequency conducted interference.

10. Toroids around cables can act as inductors, raising shield impedance, thus reducing shield currents.

11. Do not locate common mode chokes over ground planes (to prevent crosstalk around the choke).

Multilayer Boards

1. Ground plane, power planes, and power return planes can all be used to provide shielding and isolation between signal planes.

2. Power planes and power return planes must be heavily bypassed to the neighboring ground planes, in order to provide adequate shielding and isolation.

3. Plated-through holes can be used to advantage to reduce trace length.

4. Plated-through holes should be avoided on clock traces.

5. I/O signals should be kept isolated from high speed internal signals for control of both emission and susceptibility.

6. Signals should be grouped into like categories and placed, if possible, on separate isolated layers.

7. For analog signals, no signal layer should be more than two layers away from a supply, return, or reference.

8. There should be no more than two digital signal layers above a reference plane, as outer layers of a multilayer board.

9. No digital signal plane should be placed between another digital signal plane and an analog plane.

10. The traces on each signal layer should be oriented at right angles to traces on adjacent signal layers.

11. The maximum number of layers between two reference planes is four, so that no signal layer is ever more than two layers away from a reference plane.

12. The fastest rise/fall time signals should be buried between power and return planes if possible, rather than on a top layer.

13. Video signals can be routed as parallel adjacent traces on a single plane, or as a parallel stack on two planes.

14. Signal traces should be kept in from the edge of a board by a distance equal to the trace height above ground.

15. Significant amounts of radiation come directly from the integrated circuits themselves. Multiple power input and return connections on the IC are helpful in reducing emissions.

16. It is preferred that high current deflection signals that drive a CRT yoke be routed on three adjacent planes. The top plane should carry the positive voltage signal. The middle plane should carry the reference signal return and the bottom plane should be routed on top of each other throughout the entire run. The width of all three traces must be at least 1.5 times the height of separation between any two planes.

17. Ground plane material should surround every plated-through hole in multi-layer construction, to minimize loop-area.

Backplanes and Mother Boards

1. Because backplanes and mother boards are large, radiation and susceptibility from long traces can become severe.

2. Ground planes should be used on both backplanes and mother boards.

3. Just as on smaller boards, traces on one side of the board should be run at right angles to those on the other side.

4. All of the rules applicable to PC boards apply also to backplanes and mother boards.

5. Any high speed lines should be treated as transmission lines, with attention to dimensions and proper termination.

6. It may be advantageous to use multilayers, which permits construction of stripline, where the signal traces are sandwiched and effectively shielded.

7. Extend the mother board about 2 inches beyond the rear of the card cage. This area can be used for filtering and termination of I/O connectors. Bypass capacitors for I/O cables can be installed in this area. Cable shield attachment to the ground plane can occur in this area also.

8. All return and shield planes within a mother board should be bonded together at every mounting hole.

9. DC power and return pins should be equally spaced within connectors.

10. For digital signals, power and return pins should be spaced about every 0.5 inch.

11. Isolation within a mother board can be achieved by spacing between adjacent traces, by adding guard traces on either side of a critical trace, or by adding shield planes between stacked traces.

Cables Connectors and Methods of Termination

1. Shielded, twisted pairs are effective up to frequencies of a few megahertz.

2. For audio frequencies, a shield can be grounded at one end only.

3. For any frequencies higher than audio, cable shields should be grounded at both ends.

4. Pig-tail termination of shielded cables is poor practice.

5. Cable shields should have a 360° connection of the shield to the backshell.

6. Shielded cables are extremely effective against electric fields, less so against magnetic fields.

7. A single braided shield gives about 40 dB of shielding compared to an unshielded wire. Each additional shield gives 15 - 20 dB more shielding.

8. Copper and tin coated braids oxidize and lose shielding effectiveness with time.

9. Best performance of coaxial cables is obtained with foil wrapped shields with drain wires and braid over foil shields, and with solid outer metallic shields.

10. With multiconductor cables, each signal conductor should have its own separate return conductor.

11. Multiconductor cables containing twisted pairs are an effective way of conducting signals and their returns.

12. Cable shields should never be used as signal returns.

13. The precise methods of termination of cables can have a 20 to 30 dB effect on shielding effectiveness.

14. Multiconductor cable connectors can be procured with built-in bypass capacitors, helpful in reducing cable emissions.

15. Ribbon cables can be considered to be special forms of multiconductor cable. Ribbon cables are available with twisted pair, shielded twisted pair and ground plane conductors.

16. Metallized polyester film shields should be avoided.

17. Avoid the use of floating BNC connectors. If it is necessary to use a floating BNC connector, the shield connection should be bypassed to ground with a 0.001 uf 1KV capacitor.

Bypassing

1. Self-resonance in bypass capacitors must be considered. Above the self resonant frequency, the bypass capacitor is ineffective.

2. Different types of capacitors are effective at different frequencies. The type of capacitor should be carefully matched to the range of frequencies it is intended to bypass.

3. Bypass capacitors should be used on power traces for at least every third or fourth IC. More bypassing may be necessary.

4. All power leads should be bypassed where they first enter the board.

5. Use two bypass capacitors in parallel, different in value by two orders of magnitude.

6. Feedthrough bypass capacitors tend to have high self resonant frequencies because of short lead length, and can be effectively incorporated into connectors.

7. Bypassing of every signal conductor in a I/O cable can sometimes eliminate the need for a shielded cable, and may be less expensive.

8. The size of a bypass capacitor is partly dependent upon the amount of transient current that the load draws. The capacitor must be capable of supplying this transient current.

9. Do not use a plated-through hole to connect a capacitor to the circuit it bypasses.

10. With tubular capacitors, connect the outside foil end to ground, to provide shielding.

11. Make provision for bypassing of every integrated circuit, in the event this becomes necessary.

12. Tantalum capacitors are more effective at high frequencies than aluminum electrolytic capacitors.

Filtering

1. Differential mode filtering is effective in suppressing conducted emissions up to 150kHz.

2. Common mode filtering is effective in suppressing conducted emissions above 150 kHz.

3. Ferrite beads can be used on single conductors to reduce high frequency conducted emission.

4. Cables can be passed one or more times through ferrite toroids to raise shield impedance and reduce shield currents.

5. Good practice is to enclose filter components in shielded compartments. The shielded filter should be located where the filter power or I/O cable first enters the equipment. The length from entry to return of bypasses to ground should not exceed 1 1/2 inches.

6. If an analog board requires LC filtering on the DC power leads, care must be exercised to ensure that the filter does not ring. The Q of the inductor must be low enough to prevent ringing. If possible, only capacitive filtering should be used.

Board Component Location

1. Clock quartz crystals should be centrally located on a board to minimize trace lengths to the circuits fed by the clock.

2. Crystal housings should be flat against the board and grounded.

3. Components with magnetic fields, such as relays, inductors, and toroids should be separated or mounted at right angles to minimize coupling.

4. High current devices should be located close to the sources of power.

5. The various types of grounds should be separated to reduce common mode coupling.

6. Components should be located so that their ground returns do not enclose large areas.

7. Provision should be made for additional filtering and bypassing components, if they are needed.

8. Components should be located so as to keep all trace lengths as short as possible.

9. Run groups of traces that are incompatible in frequency or function through different connectors. Alternatively, group these trace connections within a single connector.

10. Do not pre-assign connector pins. Allow the layout to determine pin sequence.

11. Locate the highest frequency components nearest the connectors, unless this conflicts with locating centrally. The object is to minimize trace lead length. Next locate high current devices near the connector. Finally, locate high bandwidth and latching circuits such as flip-flops, counters and registers to achieve minimum input and power lead length.

12. Keep low level inputs and high level outputs separated by two inches or more to avoid feedback.

13. Power traces should start from a common point near the input connector, and fan-out separately to the individual loads, to reduce common mode coupling.

14. High impedance input devices should be located close to their sources with short input traces.

15. CAD/CAM automatic layout systems should not be used without reference to EMC requirements.

Board Component Density

1. Component density may be limited by thermal considerations.

2. High density implies narrow trace widths. Width to height ratios of less than 1:1 may lead to high radiation and increased susceptibility.

3. Crosstalk may be increased by running traces close together and mounting components close together.

4. Electric and magnetic fields from components may couple and cause crosstalk on dense boards.

Sources of Noise

1. Fast rise time signals from clocks and digital logic circuits are the principle sources of electrical noise.

2. Integrated circuit chips themselves are a major source of radiation. Multiple connections of power sources and returns in the chip are helpful.

3. Switching power supplies and rectifiers are sources of conducted emissions.

4. Video circuits are often sources of noise because of fast rise times and high amplitudes.

5. Relay contacts, switches and thermal devices are all sources of fast rise time, low duty cycle pulses.

6. Signals which contain ringing, overshoot, or undershoot are potential sources of EMI, and should be controlled by impedance matching and damping.

7. Never leave circuits unterminated.

8. Motors with brushes, or any arcing high voltage circuit may require filtering to eliminate broadband noise.

System Level Subassembly Location

1. Circuit boards should be grouped by function (analog, digital or power), frequency, and current consumption.

2. Power supplies should be located near the point of primary power entry and near any power line filtering.

3. Power supplies and other components with magnetic field should be kept away from CRT's, CRT deflection yokes, and other magnetically sensitive components.

EMC GUIDELINES CHECKLIST

Suppression of EMI Sources

a) Are internal subassemblies and embedded peripherals shielded and filtered to prevent undesired coupling?

b) Are diodes or other suppression components being used across relay and solenoid circuits?

c) Are RC circuits being used across AC switch or relay contacts?

d) Are solid-state switches with controlled rise and fall times being used instead of mechanical switches whenever possible?

e) Are suppression networks being used directly across DC motor brushes?

f) Are electrostatically shielded isolation transformers being used?

g) Are twisted pair, triax or shielded wires being used when required?

Connectors

a) Are signal and power circuits routed in separate cable harnesses and connectors?

b) Have provisions been made for termination of shields at both ends?

Bonding

a) Have all bare metal mating surfaces been cleaned prior to mating?

b) Where a protective film or plating is required, is it a good conductor?

c) Is the fastening method providing sufficient pressure to hold the surfaces in contact in the presence of deforming stresses, shocks and vibrations associated with the equipment and its environment?

d) Are means other than threads of screws or bolts relied upon to establish an RF bond path?

e) Have compatible metals been used throughout the enclosure?

f) Are subassemblies and/or subenclosures bonded to main enclosures?

Grounding

a) Has there been a check for ground loops and floating grounds?

b) Have good electrical bonding practices been used for ground terminations?

Primary Power Distribution

a) Has the internal primary power wiring been tightly twisted?

b) Has the internal primary wiring been physically separated and isolated from all other wiring?

c) Has the internal primary wiring been shielded as recommended?

d) Has the safety ground been terminated immediately upon entering the system, as close as possible to any EMI filter?

Shielding

a) Has the enclosure been mechanically designed to assure sufficient pressure between mating parts?

b) Has each opening in the enclosure been analyzed to determine the need for gaskets, waveguide-below-cutoff techniques, screening, etc.?

c) Have compatible metals been used in the enclosure?

d) Have provisions been made for peripheral shielding?

Filtering

a) Has space been allocated for a power line filter?

b) Has the power line filter been installed immediately at the power line entrance?

c) Are the outputs of the power line filter isolated from the inputs?

SUPPLEMENTAL DATA

Figures 71 through 75 contain supplemental information.

Frequency of Interest	1/20 Wavelength Distance	
10 MHz	[illegible]	[illegible] feet
27 MHz	[illegible]	[illegible] feet
35 MHz	.43 [illegible]	1.4 feet
50 MHz	.3 meters	(1[illegible] inches)
80 MHz	.19 meters	[illegible]
100 MHz	.15 meters	([illegible] inches)
160 MHz	.094 meters	(3.[illegible] inches)
200 MHz	[illegible] meters	([illegible] inches)

FIGURE 71. 1/20 WAVELENGTH AT VARIOUS FREQUENCIES

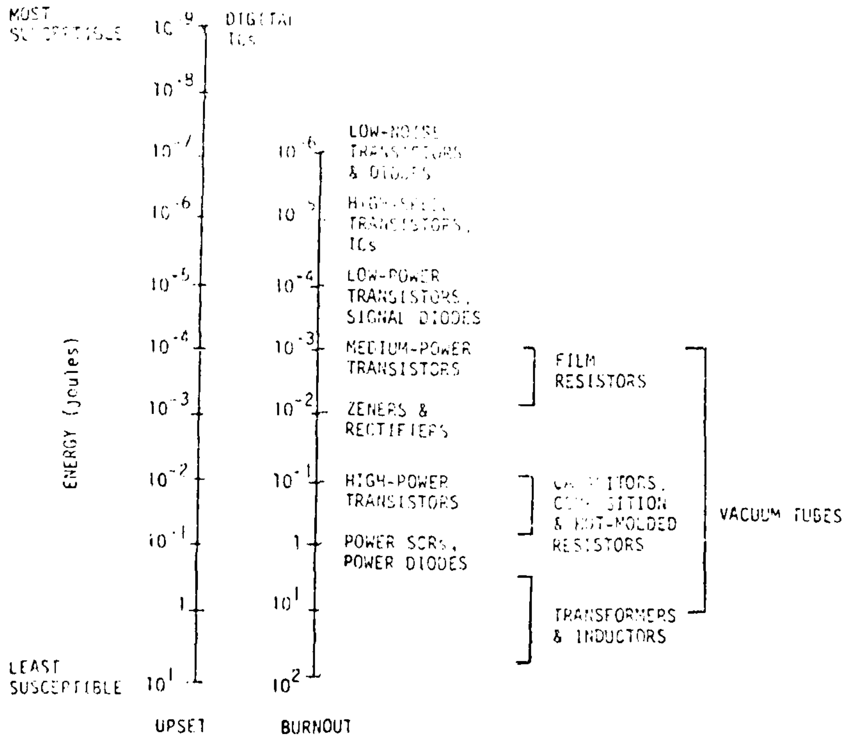

FIGURE 72. UPSET AND BURNOUT ENERGIES FOR VARIOUS CIRCUIT ELEMENTS

Voltage Ratio	dB	dB	Voltage Ratio
1	0	0	1
2	6	1	1.12
3	9.5	2	1.26
4	12	3	1.41
5	14	4	1.58
6	16	5	1.78
7	17	6	2.00
8	18	7	2.24
9	19	8	2.51
10	20	9	2.82
		10	3.16

FIGURE 73. DECIBELS - VOLTAGE RATIOS

Noise Source	Amplitude	Typical Risetime	Typical Pulse Width	Breakpoint 1		Breakpoint 2	
				Amplitude	Frequency	Amplitude	Frequency
Lightning	20 KA	1 microsec	50 microsec	240 dBuA/MHz	12.5 kHz	210dBuA/MHz	318 kHz
EMP Pulse	50 KV	10 nanosec	10 microsec	233 dBuV/MHz	64 kHz	179dBuV/MHz	32 MHz
Electro-static discharge	15 KV	5 nanosec	400 nanosec	195 dBuV/MHz	1.5 MHz	163dBuV/MHz	64 MHz
1 MHZ TTL Logic	5V	10 nanosec	10 nanosec	130 dBuV	1 MHz	99dBuV	32 MHz

FIGURE 74. TYPICAL BREAKPOINT FREQUENCIES FOR PULSE TYPE NOISE SOURCES

LOGIC FAMILY (GATES)	RISE/FALL TIME (T_r)	PRINCIPAL HARMONIC CONTENT $F_2 = (1/\Pi T_r)$	POSSIBLE SIGNIFICANT SPECTRUM	MAXIMUM NON-TRANSMISSION LINE TRACE LENGTH
74L xxx	31-35 ns	10 MHz	100 MHz	150cm (59")
74C xxx	25-60 ns	13 MHz	130 MHz	115cm (45")
CD 4xxx (CMOS)	25 ns	13 MHz	130 MHz	115cm (45")
74HC xxx	13-15 ns	25 MHz	250 MHz	60cm (24")
74 xxx (Flip-Flop)	10-12 ns 15-22 ns	32 MHz 21 MHz	270 MHz 210 MHz	47cm (18.5) 71cm (28")
74LS xxx (Flip-Flop)	9.5 ns 13-15 ns	34 MHz 25 MHz	340 MHz 250 MHz	44cm (17") 60cm (24")
74H xxx	4-6 ns	80 MHz	800 MHz	22.5cm (9")
74S xxx	3-4 ns	106 MHz	1.1 GHz	14cm (5.6")
74F xxx	1.5-6.5 ns	212 MHz	2.1 GHz	7cm (2.8")
ECL 10K	1.5 ns	212 MHz	2.1 GHz	7cm (2.8")
ECL 100K	0.75	424 MHz	4.2 GHz	3.5cm (1.4")

T_r depends greatly on load capacitance, supply Voltage and IC complexity. Consult the Manufacturer's specifications.

"1.7 ns/foot propagation for epoxy fiber, $E_r = 4.7$

FIGURE 75. APPROXIMATE HARMONIC SPECTRUM OF DIGITAL LOGIC FAMILIES

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