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Marshall Space Flight Center Electromagnetic Compatibility Design and Interference Control (MEDIC) Handbook

CDDF Final Report, Project No. 93-15

T.L. Clark, M.B. McCollum, D.H. Trout, and K. Javor

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T.L. Clark, M.B. McCollum, D.H. Trout
Marshall Space Flight Center • MSFC, Alabama

K. Javor,
Sverdrup Technology, Huntsville, AL

National Aeronautics and Space Administration
Marshall Space Flight Center • MSFC, Alabama 35812

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FOREWORD

The Marshall Space Flight Center Electromagnetic Compatibility Design and Interference Control (MEDIC) Handbook is intended to be used primarily by those organizations involved in the electrical design of payload equipment and subsystems. The purpose of this Handbook is to provide practical and helpful information in the design of electrical equipment for electromagnetic compatibility (EMC).

Chapter 1 of this Handbook is an introduction to electromagnetic compatibility (EMC). It includes definitions of terms and units as well as basic electromagnetic interference (EMI) interactions. Chapter 2 is an overview of typical NASA EMI test requirements and associated test setups. It is not intended to be a "how to" of EMI testing, but rather a general overview so that the electrical designer knows what to expect during testing. Chapter 3 contains general design techniques to minimize the risk of EMI and deals with EMI suppression at the board and equipment interface levels. Chapter 4 gives specific EMI test compliance design techniques and retrofit fixes for noncompliant equipment. These techniques and retrofit fixes are specific to a given MSFC EMI test. Chapter 5 explains how to perform special tests useful in the design process or instances of specification noncompliances. Appendix A lists the acronyms and abbreviations used in this document.

The MEDIC Handbook was prepared at the Marshall Space Flight Center (MSFC) by the Electromagnetics and Environments Branch (EL54). Funding for developmental testing was provided by the MSFC Center Director's Discretionary Fund (CDDF), Project No. 93-15.

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1. INTRODUCTION TO ELECTROMAGNETIC COMPATIBILITY

1.1 Electromagnetic Interference

An incompatibility occurs when the operation of one equipment interferes with the operation of another. When the interaction is traced to the transfer of electromagnetic energy from the culprit equipment to the victim, it is termed electromagnetic interference (EMI). In order for this energy transfer to occur, a transfer mechanism or coupling path is necessary (shown in fig. 1-1).

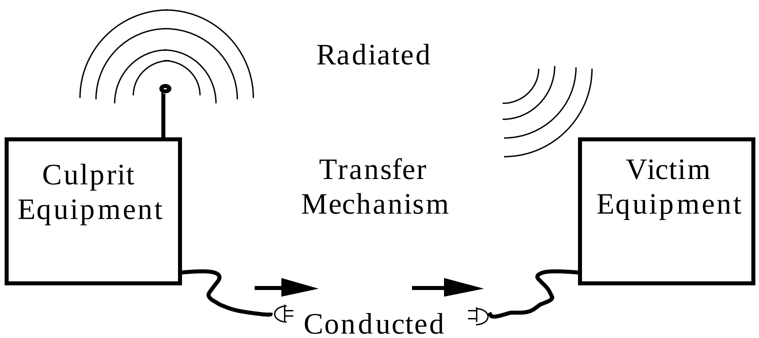


Figure 1-1. Elements of EMI.

At the system level, EMI coupling mechanisms are normally quite lossy, and only a small portion of the energy in the culprit actually transfers to the victim. Thus, the most likely scenario for incompatibility occurs when a relatively high power culprit is located near a very sensitive victim. The transfer mechanisms are a function of culprit-to-victim separation, the spectrum of the signals of the culprit, and spectral sensitivity of the victim. The first EMI incidents occurred when sensitive radio receivers operated near other electronics that intentionally or unintentionally radiated radio frequency (RF) energy.[1-1] A simple, familiar example of such interference is the effect an operating hair dryer or vacuum cleaner has on a television or AM radio.

An inefficient transfer mechanism may also be overcome by a wide disparity in culprit and victim power levels. When a high-power RF transmitter illuminates an ordinary piece of electronics, sufficient energy may be coupled into the victim to interfere with its operation. A famous example of this type of interference is the flight control system of the UH-60 Blackhawk helicopter. When the Blackhawk flies near certain radio transmitters, a loss of flight control occurs and the helicopter could crash.[1-1] Since control electronics are much less sensitive than a receiver, high power-level fields similar to those near a transmitter must be present before an interference situation results. In the Navy environment, helicopters must take off and land while being exposed to radar field levels greater than 200 volts per meter (V/m) as well as high-frequency (HF) and very-high-frequency (VHF) transmitters. It is interesting to note that the sister ship of the UH-60, the Navy *Seahawk*, was commissioned with more stringent electromagnetic shielding and has had no problems.

1.2 Electromagnetic Compatibility

The term electromagnetic compatibility (EMC) denotes the electromagnetically compatible simultaneous operation of different equipment. EMC can be defined by the absence of EMI, but EMC is more than that. Currently, it is common for an EMI test facility to be interchangeably called an

EMC test facility. This is a misnomer. An EMC test is performed at some level of system integration. EMC is ascertained by energizing equipment A, determining proper operation, energizing equipment B, and noting whether or not equipment A continues to operate as before without any degradation. The EMC test results can be summarized as a square matrix of victims and culprits (table 1-1).

Table 1-1. Compatibility matrix.

Culprit/Victim	Equipment A	Equipment B	Equipment C	Equipment D
Equipment A	n/a	EMC	EMC	EMC
Equipment B	EMC	n/a	EMC	EMC
Equipment C	EMC	EMC	n/a	EMC
Equipment D	EMI	EMI	EMI	n/a

Note: In this matrix, the outcome is binary, EMI or EMC. The test is qualitative. In the above example, equipment D has been found to interfere with equipment A, B, and C. The demonstration of EMC at the system level, where all the equipment operates without any interference, is the goal of the EMC program.

In contrast, tests performed in an EMI test facility are quantitative. Emissions measured in volts, amperes, teslas, or volts per meter are compared to specification values. Susceptibility to, or immunity from, specification values of volts, amperes, teslas, or volts per meter is also measured. The successful conclusion of these quantitative tests is a reassuring indication that the final EMC test will have the desired outcome. Failure to meet requirements may indicate a need for redesign, but typically further analysis is first performed to determine if the particular failure is likely to cause an EMC problem. For example, an equipment emission that exceeds the radiated emission (RE) limit by 20 dB at 100 kHz may not be serious if the overall system for which the equipment is destined does not utilize the spectrum below 2 MHz.

1.3 Basic Electromagnetic Interference Interactions

EMI is quantified and controlled by four categories. These categories encompass all the possible permutations of radiated and conducted mechanisms combined with control of emissions from the equipment and with control of susceptibility of the equipment. The four categories are the following:

(a) Conducted emissions (CE)

(b) Conducted susceptibility (CS)

(c) Radiated emissions (RE)

(d) Radiated susceptibility (RS).

1.3.1 Conducted Emissions/Conducted Susceptibility

The simplest example of the CE/CS mechanism is the shared or common bus impedance illustrated in figure 1-2. Noise currents drawn by the current source/sink culprit on the left cause a voltage ripple in the portion of the bus common to both loads. Voltage ripple, not current ripple, is the mechanism for interference and is due to the nonzero impedance of the bus. Two points should be noted: (1) bus impedance elements are depicted only in the portion of the bus feeding both loads, the common impedance path; and (2) impedances in the branches downstream of the common impedance path have no effect in translating culprit current CE into voltage ripple at the victim. (Presumably, the culprit lives with its own induced ripple. If not, it is back to the drawing board for the equipment designer, long before he even thinks about EMI testing.)

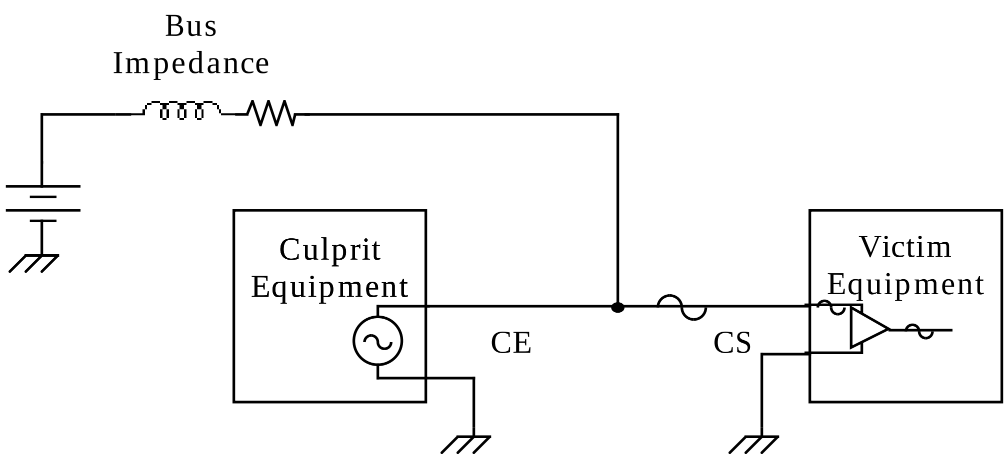


Figure 1-2. CE/CS.

It must be stressed that shared or common bus impedance is a simple interaction, and other factors, such as cable radiated electromagnetic fields, are also considered when developing CE limits.

1.3.2 Radiated Emissions

RE occur principally from currents flowing on equipment-connected wiring and on the equipment enclosure. These currents are not typically a necessary by-product of the intentional signal processing and differential mode signal transmission on equipment interface cables. Instead, they couple parasitically from one portion of the equipment to the exterior and flow in a common mode (CM) path. As shown in figure 1-3, coupling from these currents to the victim occurs in the following ways:

(a) Inductively in wire-wire transformer action

(b) Capacitively, where a fraction of the culprit CM voltage is impressed on the victim circuit

(c) Directly radiating into an antenna with a receiver tuned to the interference frequency.

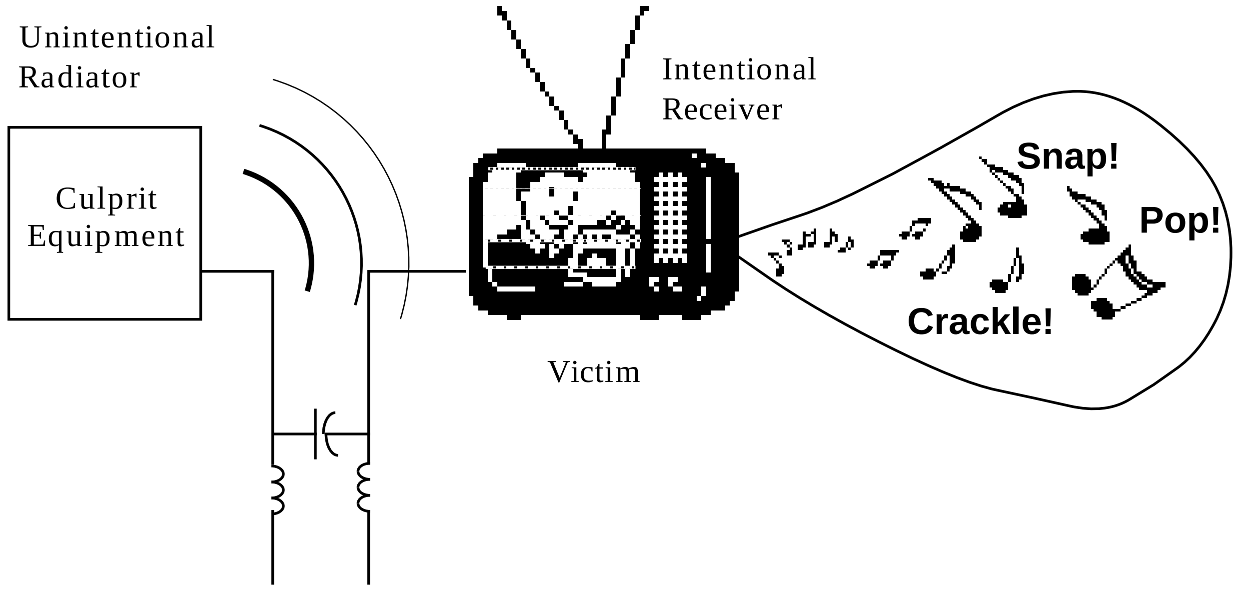


Figure 1-3. RE.

Commercial and military RE limits protect antenna-connected receivers. CE appearing on power lines are controlled per section 1.3.1. Under older versions of MIL-STD-461 and MIL-STD-462, CM CE on all signal lines were also controlled.[1-2, 1-3]

Inductive and capacitive coupling is often called crosstalk. With modern wiring practices, capacitive crosstalk is rare. Inductive crosstalk is easily controlled by properly grounding, bonding, and shielding design. Because crosstalk is a much more serious problem within the equipment enclosure, the equipment designer must consider this during the design process. With modern processor speeds and high-density printed circuit boards and ribbon cable, it is important to plan the layout to minimize crosstalk. In fact, ribbon cable users are now categorizing like and unlike signals for grouping and segregating just as World War II-era aircraft wire harness designers did before dedicated wire returns, twisted wires, and twisted shielded wire pairs were commonly used. Similar problems beget similar solutions, even across 40 plus years.

1.3.3 Radiated Susceptibility

Whereas CE and CS were lumped together based on the common impedance coupling model, RE and RS cannot be so matched. RS occurs when intentionally transmitted RF power is intercepted by wiring associated with a victim circuit operating at low signal levels such that the coupled voltage causes degradation (depicted in fig. 1-4). This occurs for RF field intensities above 1 V/m (without special design), whereas, unintentional RE are always well below 30 mV/m (again, without special design). Thus, there is a huge natural margin between RE and RS such that neither is controlled with respect to the other. As explained above, RE limits protect antenna-connected receivers and RS limits protect non-RF equipment from high power RF transmitters.

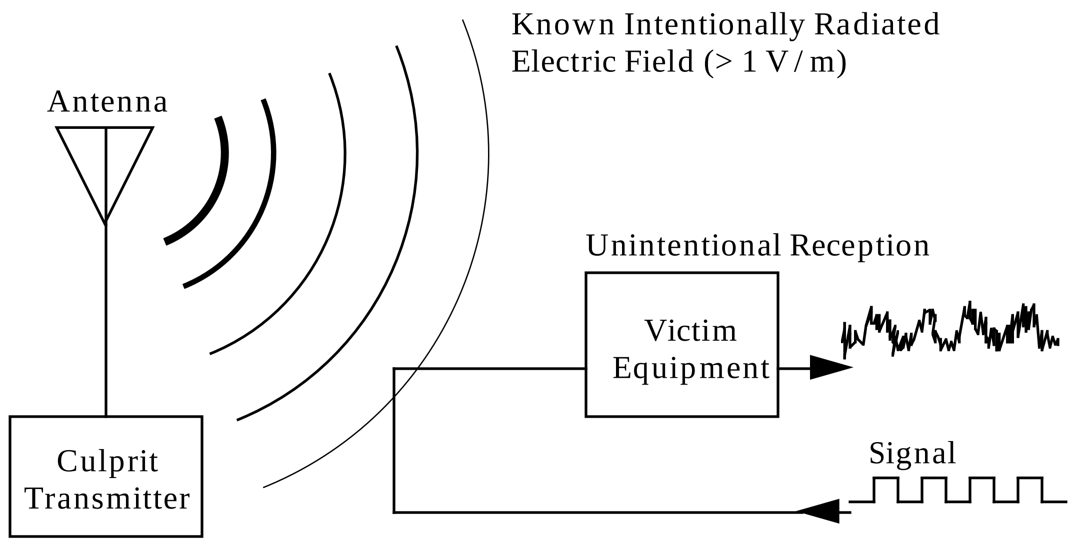


Figure 1-4. RS.

1.4 Common Electromagnetic Interference Terminology

Frequency domain EMI units can be confusing to someone accustomed to working primarily in the time domain. Frequency domain measurements and terminology are simply representative of the class of problems controlled by EMI limits; interference with radio receivers. Sensitivity of tunable radios is measured in dBm or dBμV. Although broadband signals can desensitize the receiver by overloading a wideband receiver front end, channel bandwidth determines both the narrowband and broadband sensitivity.

Section 1.4.1 defines standard units of EMI specifications. For the reader who desires a review of decibel and logarithmic definitions and manipulations, a brief discussion is presented in appendix C.

1.4.1 Standard Units

dBμV:

$$\text{dB}\mu\text{V} = 20 \log \left[\frac{\text{signal strength } (\mu\text{V})}{1\ \mu\text{V}} \right] , \tag{1-1}$$

dBm:

$$\text{dBm} = 10 \log \left[\frac{\text{signal strength (mW)}}{1 \text{ mW}} \right] . \tag{1-2}$$

Note: In the typical 50-Ω EMI measurement system, the following relationship is used to convert between dBμV and dBm:

$$\text{dB}\mu\text{V} = \text{dBm} + 107 \ , \tag{1-3}$$

dBμA:

$$\text{dB}\mu\text{A} = 20 \log \left[\frac{\text{signal strength } (\mu\text{A})}{1\ \mu\text{A}} \right] . \tag{1-4}$$

dBΩ: Watch out! This one is tricky. If a voltage to current relationship (i.e., V = $I\,R$) is being evaluated in log form, then R is a constant of proportionality between voltage and current and takes on the same "20 • log" character:

$$\mathrm{dB}\Omega = 20 \log \left[\frac{\text{resistance } (\Omega)}{1\ \Omega} \right] . \qquad (1\text{-}5)$$

However, if a power relationship (i.e., $P = \frac{V^2}{R}$ or $P = I^2 R$) is being evaluated in log form, then R is a constant of proportionality between power and the square of the voltage or current and takes on the "10 • log" character:

$$\mathrm{dB}\Omega = 10 \log \left[\frac{\text{resistance } (\Omega)}{1\ \Omega} \right] . \qquad (1\text{-}6)$$

Finally, units for narrowly tunable signals differ from those for signals whose spectrum occupancy appears larger than the receiver bandwidth. The typical electronic design engineer is familiar with units such as:

$$\frac{\mu\mathrm{V}}{\sqrt{\mathrm{Hz}}} ,$$

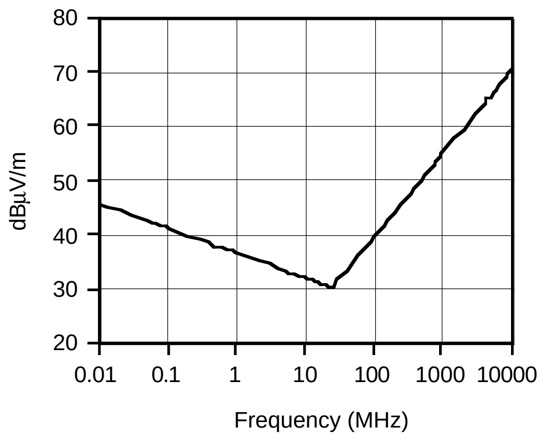
for expressing noise intensity normalized per unit bandwidth. The square root relationship occurs because thermal noise is an incoherent phenomenon. In the EMI measurement community, the unit for broadband signal measurements is:

$$\mathrm{dB}\mu\mathrm{V/MHz}.$$

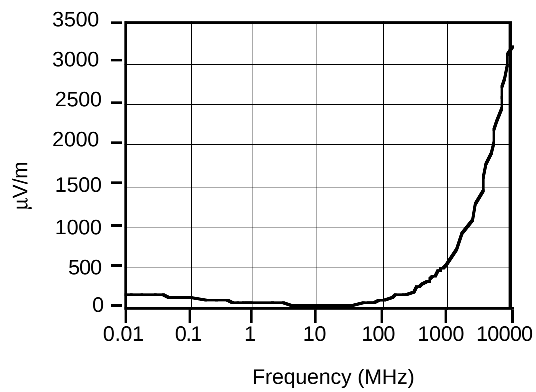
The implication is that the signal measured is a coherent broadband source, i.e., an impulse.

1.4.2 Motivation for the Use of Logarithms and Decibels

Typical radiated EMI measurements encountered within a single equipment qualification test may encompass a dynamic range from 30 μV/m to 30 mV/m (factor of 1,000). CE measurements may range from 10 A to 10 μA (range of 1,000,000). It is difficult to arithmetically handle such numerical ranges but, more importantly, it is very impractical to build instrumentation to display such ranges in a linear mode. Furthermore, the frequency ranges covered by EMI test requirements cannot be conveniently plotted on a linear scale. Figures 1-5 and 1-6 illustrate the problem. Figure 1-5(a) shows an RE limit from MIL-STD-461C.[1-4] Figure 1-5(b) shows the same limit but with the ordinate (y-) axis calibrated in linear, not logarithmic units. The abscissa (x-) axis retains the use of logarithmic units. Severe compression of the low-frequency limit in figure 1-5(b) is evident and it is impossible to interpret the limit curve below about 30 MHz. In figure 1-6, the abscissa (x-) axis has also been linearized. Here it is impossible to determine any low-frequency information from the graph, even though the x-axis has been resized to twice as long as that of figure 1-5.



(a) Ordinate (y-) axis in log units



(b) Ordinate (y-) axis in linear units

Figure 1-5. MIL-STD-461C RE02, part 2.

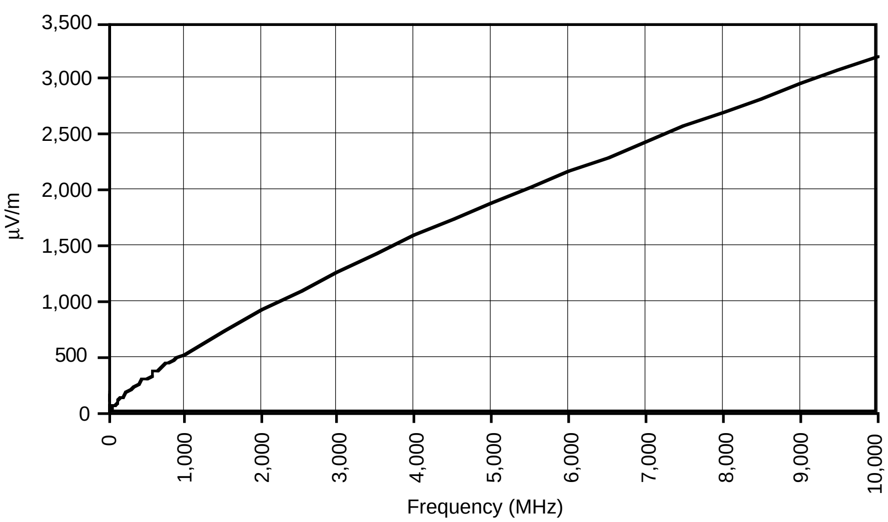


Figure 1-6. MIL-STD-461C RE02, part 2 with linear axes.

The same effect is observed using a measuring device. If a spectrum analyzer must read 0 dBm (1 mW), readings near –90 dBm (1 pW) are only discernible by utilizing a logarithmic display. Figures 1-7 and 1-8 illustrate logarithmic and linear spectrum analyzer displays. A small fraction of the signal dynamic range is exhibited in the linear mode.

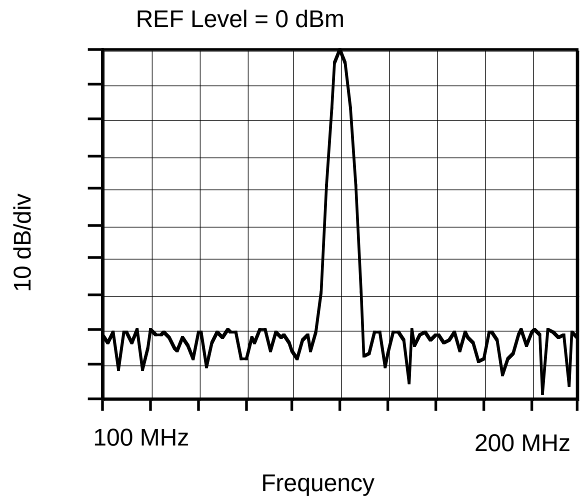


Figure 1-7. Spectrum analyzer log signal display.

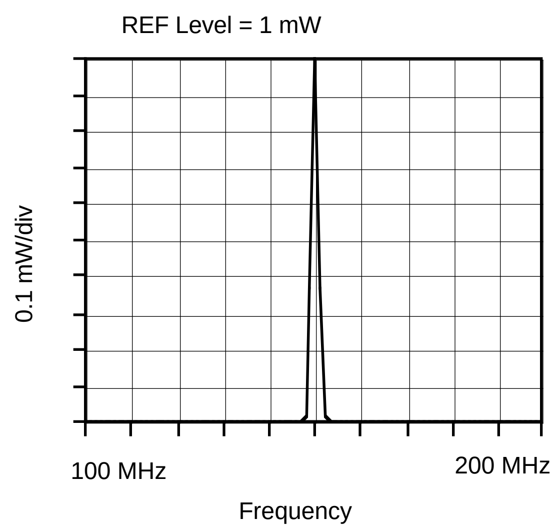


Figure 1-8. Spectrum analyzer linear signal display.

In figure 1-8, the top nine divisions display the contents of the top division of figure 1-7, while the lowest division of figure 1-8 condenses and displays the bottom nine divisions of figure 1-7.

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1-2. MIL-STD-461A, Military Standard, Electromagnetic Interference Characteristics, Requirements for Equipment, August 1968.

1-3. MIL-STD-462, Military Standard, Electromagnetic Interference Characteristics, Measurement of, July 1967.

1-4. MIL-STD-461C, Military Standard, Electromagnetic Emissions and Susceptibility Requirements for the Control of Electromagnetic Interference, August 1986.

FOR FURTHER READING

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2. ELECTROMAGNETIC INTERFERENCE REQUIREMENTS OVERVIEW

2.1 Introduction

This chapter contains an overview of typical NASA EMI test requirements. Each section states the purpose and applicability of an EMI test and gives the general test setup. This chapter is not intended to be a “how to” of EMI testing, but rather a general overview so that the electrical designer knows what to expect during testing. Many different types of experiments fly on various NASA platforms with different sensitivity receivers, intentional and unintentional. For illustration, specification limits from MSFC-SPEC-521B are included in this chapter. MSFC-SPEC-521B, based on MIL-STD-461A, is the specification imposed on Spacelab payloads.[2-1, 2-2] Various platforms, present and future, have (will have) different limits imposed. Differences vary both in frequency range covered and limit levels. The designer interested in exact limits should refer to the contractually imposed specification. Test setups shown in this chapter contain an equipment under test (EUT) that is simply a generic “black box” containing electrical circuits.

2.2 CE01, Conducted Emissions, 30 Hz to 20 kHz

Purpose: The requirement limits low frequency noise currents which can be drawn from a power bus. The test method is suitable for measuring audio frequency (ELF, VF, and VLF) current CE on power leads and signal lines. Current control is imposed because, over part of the frequency range of the requirement, wire resistance will dominate source reactance. This makes it difficult to establish a standard source impedance for all cases. Noise currents generated by the full suite of equipment on a platform can be combined and used to predict platform bus voltage ripple by the integrating activity.

Applicability: This nonintrusive current probe test method is suitable for measuring currents on both alternating current (ac) and direct current (dc) power leads and signal lines. NASA applies the test method and limit only to primary power lines, which are usually dc. The method uses an EMI-type current probe and 10-μF capacitors from each line to ground. The MSFC-SPEC-521B limit is shown in figure 2-1. The test setup is shown in figure 2-2.

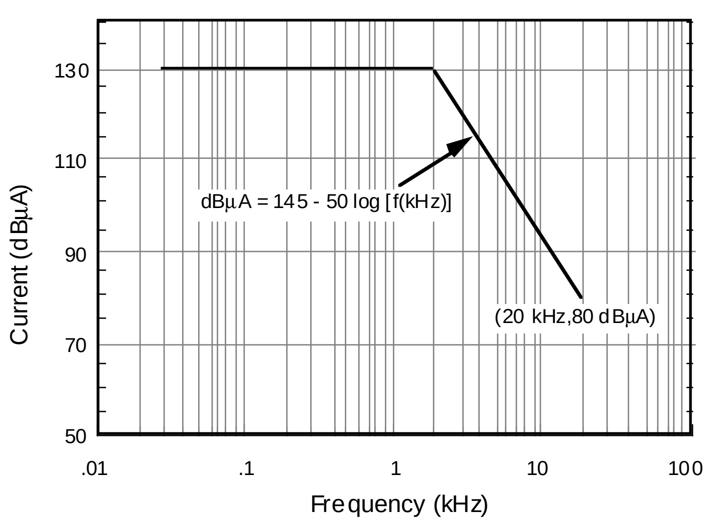


Figure 2-1. CE01 limit for Spacelab 28-Vdc loads.

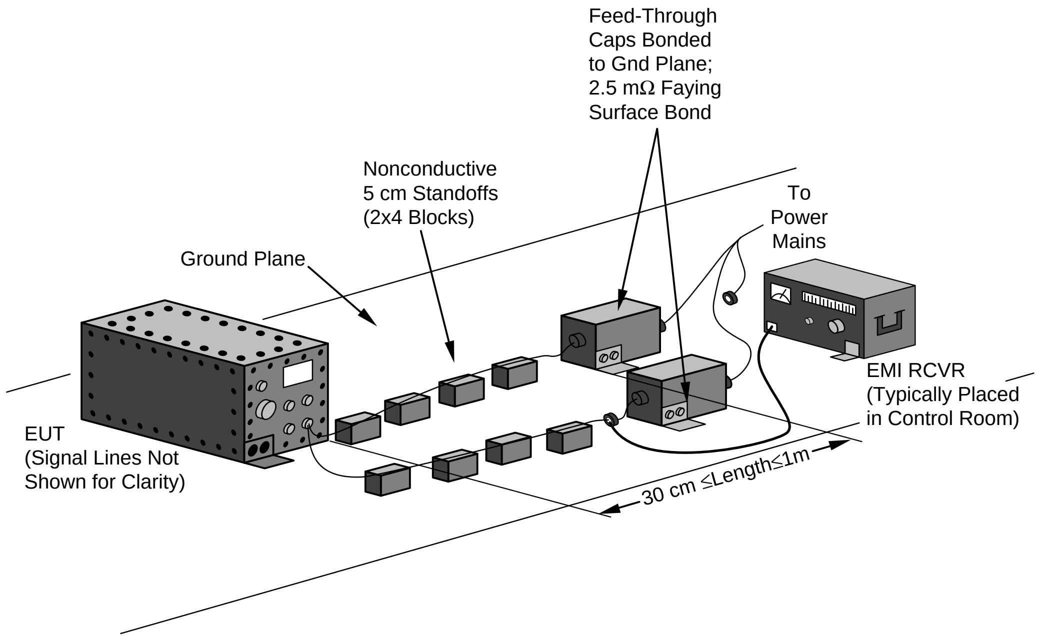


Figure 2-2. CE01/CE03 test setup.

2.3 CE03, Conducted Emissions, 15 or 20 kHz to 50 MHz

Purpose: The requirement limits RF currents drawn from a power bus. The test method is suitable for measuring RF current CE on power leads and signal lines. Current control is imposed rather than voltage control so that (worst case) analyses of resultant bus ripple can be calculated for different installations of the test sample.

Applicability: This nonintrusive current probe test method is suitable for measuring currents on both ac and dc power leads, and signal lines. The method uses an EMI-type current probe and 10-μF capacitors from each power line to ground. MSFC-SPEC-521B limits are shown in figure 2-3 and apply only to power lines. The test setup is the same as for CE01 shown in figure 2-2.

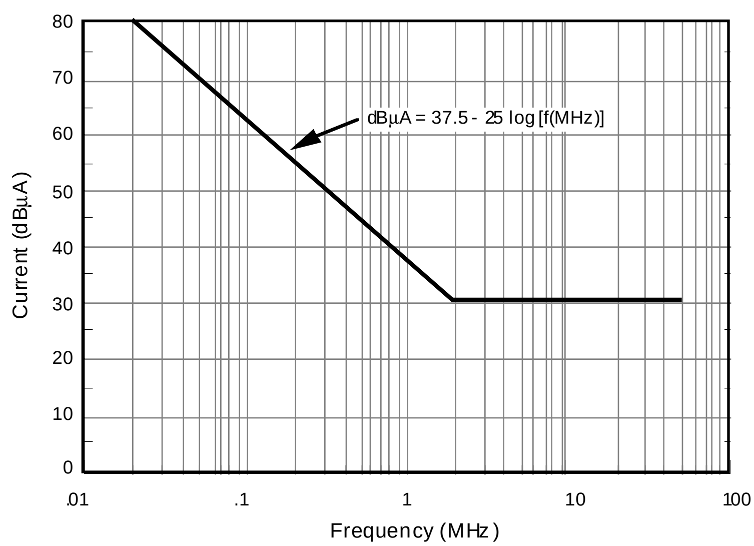


Figure 2-3. CE03 limit.

2.4 CE07 (Also Called TT01), Conducted Emissions, Time Domain Voltage Spikes

Purpose: The purpose of this requirement is to specify and measure, in the time domain, the load-induced effect on power quality caused by cycling the EUT on and off, as well as through any and all of its various modes of operation that could significantly affect the line voltage. The limit is specified as a voltage induced across a specified source impedance (see applicability). The impedance is fixed above a few kHz, but must simulate wire resistance at dc through the low portion of the audio band. Since this is a time-domain test, it is important that the source impedance be specified over the entire range of frequencies which correspond to the transient time duration. The source impedance is specified from dc to 10 MHz, except, as noted above, the dc portion of the impedance is based on the platform power bus resistance. The integrating activity compares transient emission performance to power quality limits and/or the known transient susceptibility of other platform electrical loads.

Applicability: This method is applicable for measuring time-domain spikes (transients). Measurements are to be made line-to-line across a specially designed line impedance simulation network (LISN). The network is intended to model the bus impedance through which common impedance coupling occurs. This requirement is applicable for turn-off transients only when the power switch is contained within the EUT (as opposed to a remotely located power switch or circuit breaker). The limit from MSFC-SPEC-521B is shown in figure 2-4. The limit is based on a desire to protect the Spacelab remote acquisition unit (RAU), which is sensitive to negative going (turn-on) transients. If bus voltage sags below 22 V for more than 80 μs, the RAU will shut down. The LISN models the common impedance of the power bus from fuel cell to the point at which the RAU and the EUT no longer share a common bus. Hence, dc resistance of the LISN is adjustable. The test setup is shown in figure 2-5.

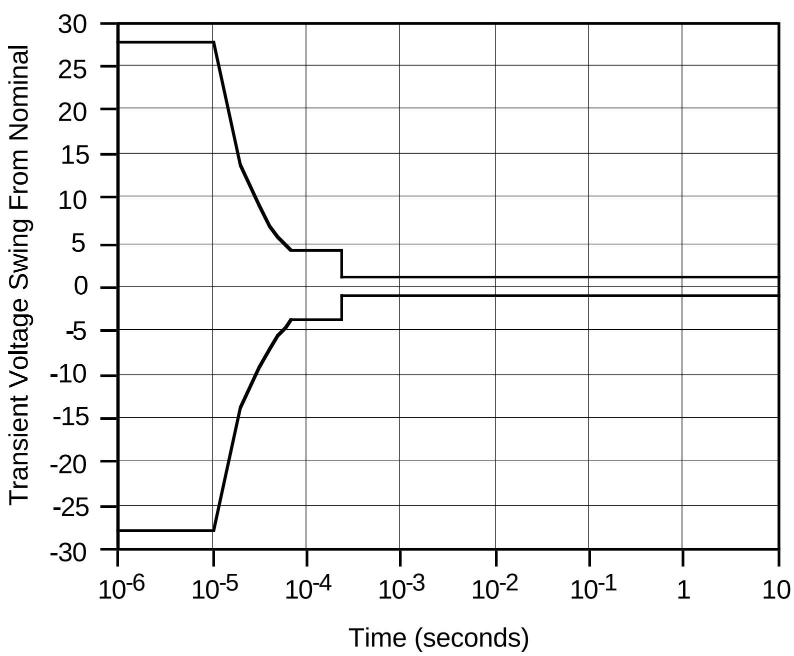


Figure 2-4. TT01 (CE07) limit from MSFC-SPEC-521B.

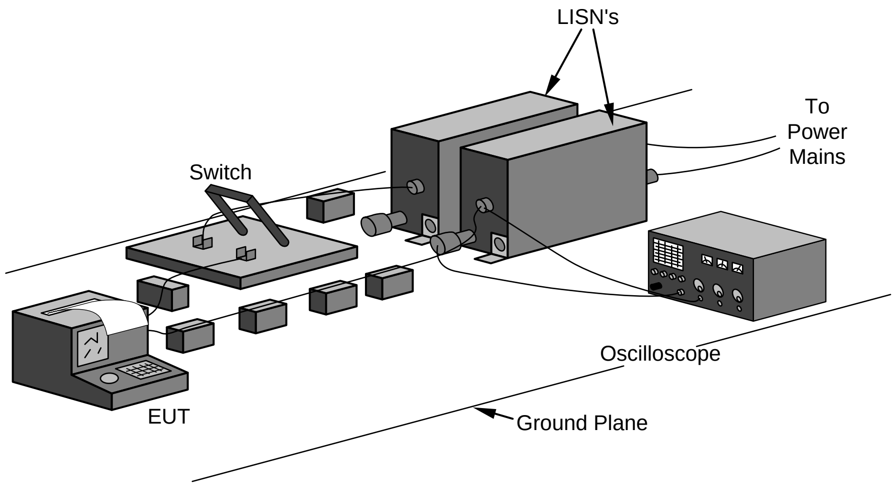


Figure 2-5. TT01 (CE07) test setup (most NASA programs derive a single LISN for this test, rather than using two standard LISN's as shown here for a generic CE07 test).

2.5 RE02, Electric Field Radiated Emissions, 14 kHz to 10 GHz

Purpose: The purpose of this requirement is to limit electric-field radiation from the EUT and associated cabling.

Applicability: The general method is applicable to all types of equipment. Limits and frequency range of the test often depend on use of the EUT. A variety of antennas is used. The most common ones are the 41-in rod, the biconical, and the log periodic. Generic RE02 limits (narrowband and broadband) are shown in figures 2-6 and 2-7. Most NASA programs start with these as point of departure. Sometimes notches are added to protect specific receivers. The test setup is shown in figure 2-8.

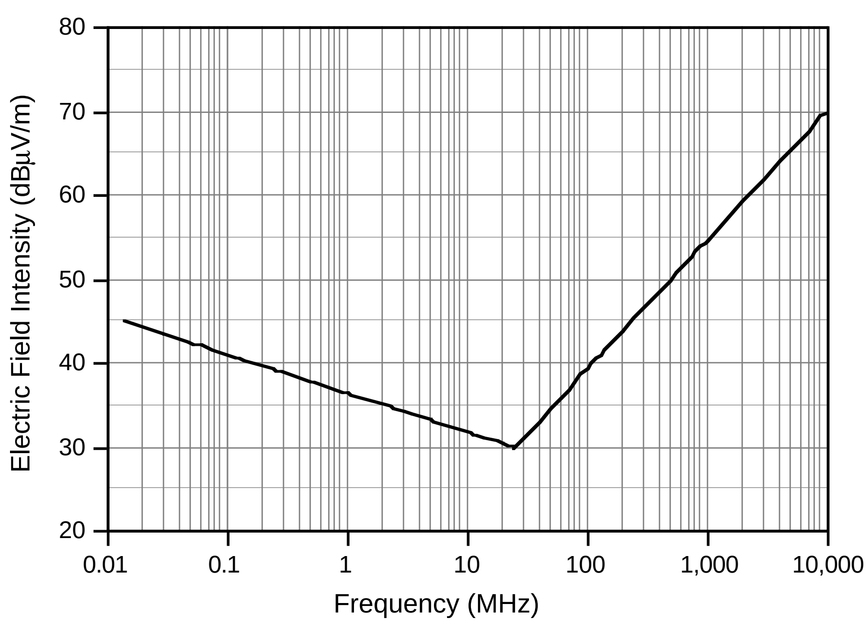


Figure 2-6. RE02 narrowband limit.

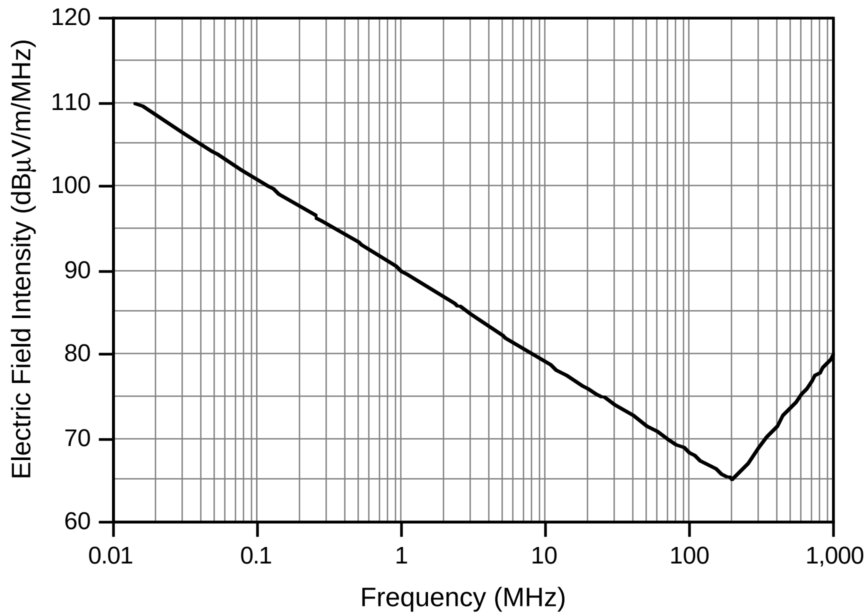


Figure 2-7. RE02 broadband limit.

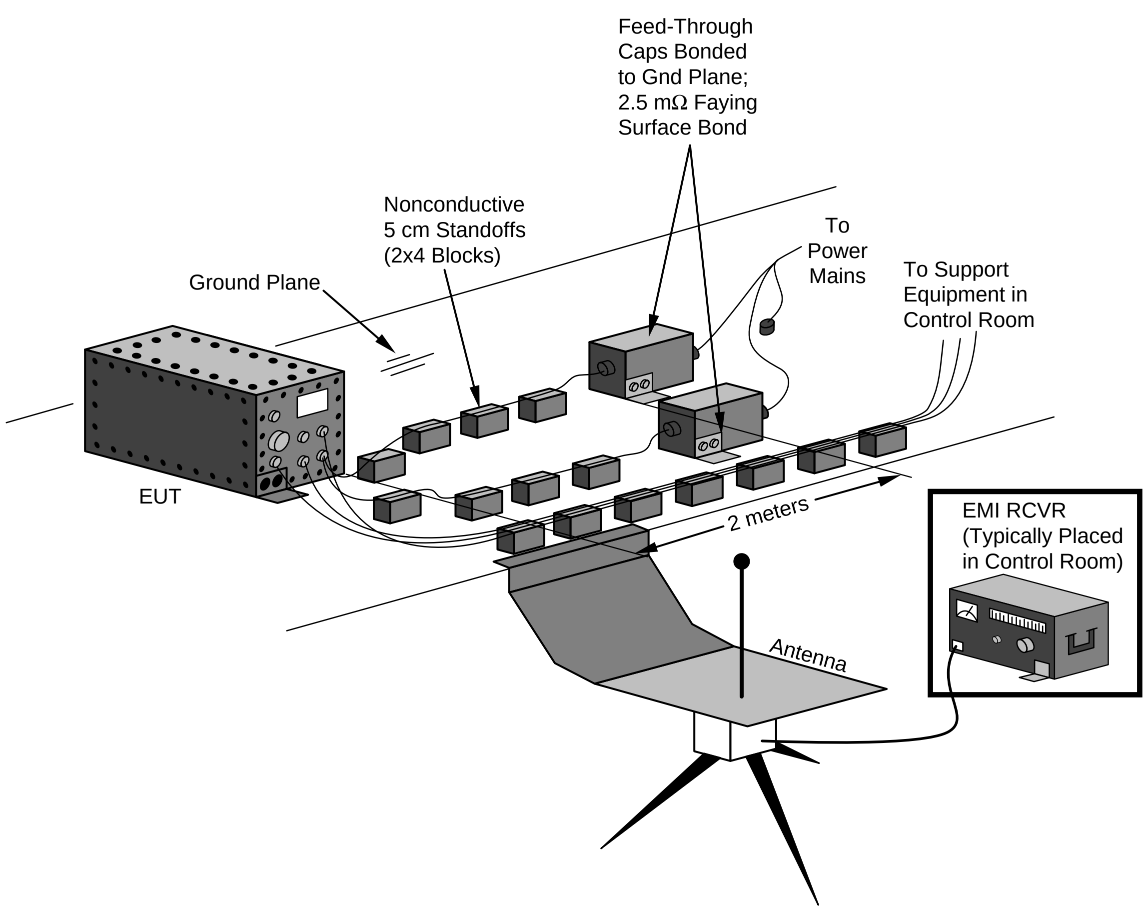


Figure 2-8. RE02 test setup, 41-in rod antenna test (0.01 to 30 MHz)
(antenna 1 m from EUT, counterpoise at least 30 cm wide).

2.6 RE04, Magnetic Field Radiated Emissions, 30 Hz to 50 kHz

Purpose: The purpose of this requirement is to control magnetic field radiation from the EUT and associated cabling.

Applicability: The general method is applicable for measuring magnetic field radiation from equipment, subsystems, cables (including control, pulse, IF, video), power and antenna transmission lines, and interconnecting wiring. The method uses a 5-cm diameter loop antenna held 1 m from the EUT. The RE04 limit currently imposed under MSFC-SPEC-521B is shown in figure 2-9.

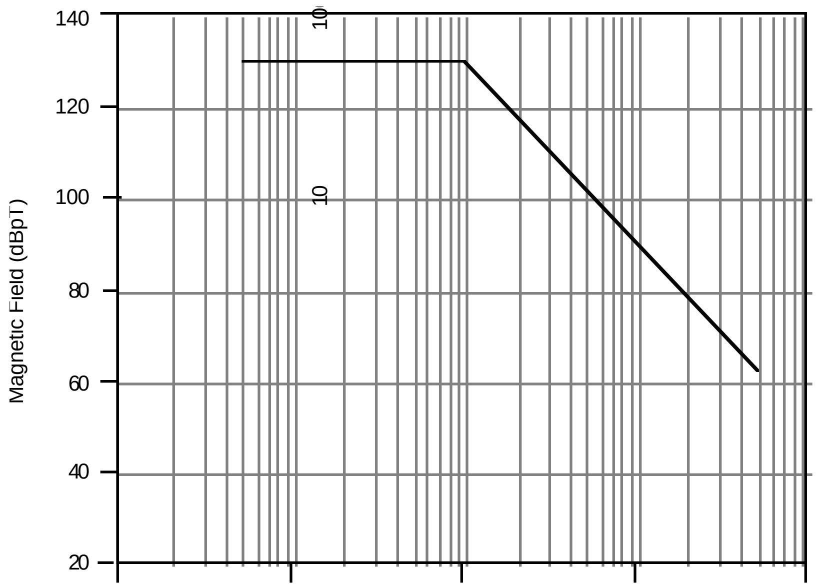


Figure 2-9. RE04 limit of MSFC-SPEC-521B.

2.7 CS01, Conducted Susceptibility, 30 Hz to 50 kHz

Purpose: The purpose of this requirement is to control and determine the susceptibility level of the EUT to audio frequency interference signals on power leads in the audio frequency range.

Applicability: This requirement is imposed on all equipment drawing current from a power bus. The test is rather equipment intensive, requiring several specialized products including an audio amplifier, an injection transformer, etc. The test setup is shown in figure 2-10. MSFC-SPEC-521B imposes 1.5 Vrms for a 28-Vdc bus. The requirement is met when the power source, adjusted to dissipate 50 W in a 0.5-Ω load, cannot develop the required voltage at the EUT power input terminals and does not disrupt the normal operation of the EUT. The power setting at which 50 W is dissipated in a 0.5-Ω load is determined by test in a precalibration setup (fig. 2-11).

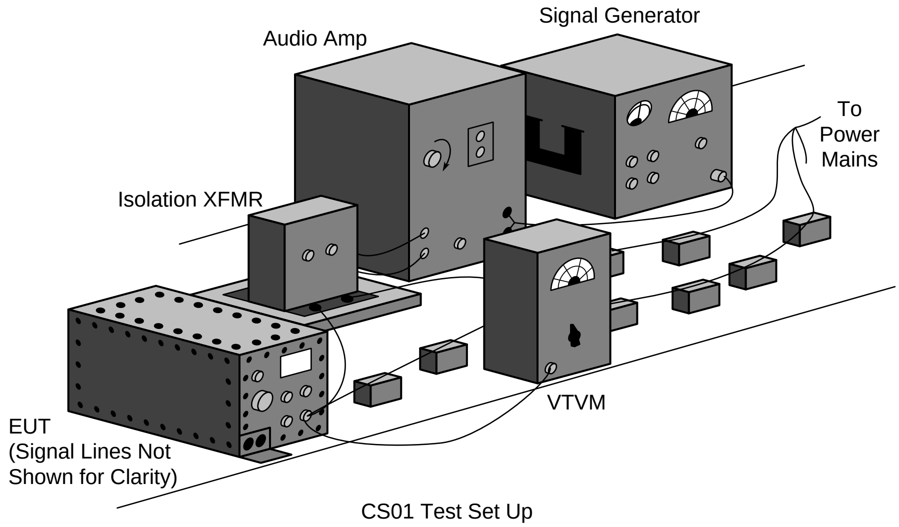


Figure 2-10. CS01 test setup.

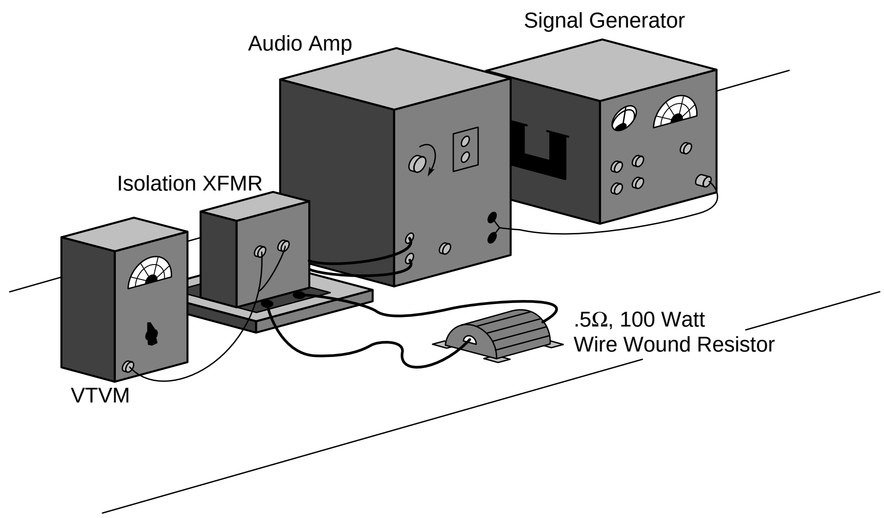


Figure 2-11. CS01 precalibration setup.

2.8 CS02, Conducted Susceptibility, 50 kHz to 400 MHz

Purpose: The purpose of this requirement is to control and measure the susceptibility of the EUT to RF signals injected onto its power input terminals.

Applicability**:** This requirement is imposed on all equipment drawing current from a power bus. The test is rather equipment intensive, requiring several specialized products including an RF signal generator and amplifier, an RF coupler, etc. The test setup is shown in figure 2-12. Typical limits are on the order of 0.1 to 1 V from a 50-Ω source in the frequency range of 50 kHz to 400 MHz. The requirement is met when the signal source, at a setting capable of delivering 1 W into a 50-Ω load, cannot develop the required voltage at the EUT power terminals and does not disrupt normal operation of the EUT.

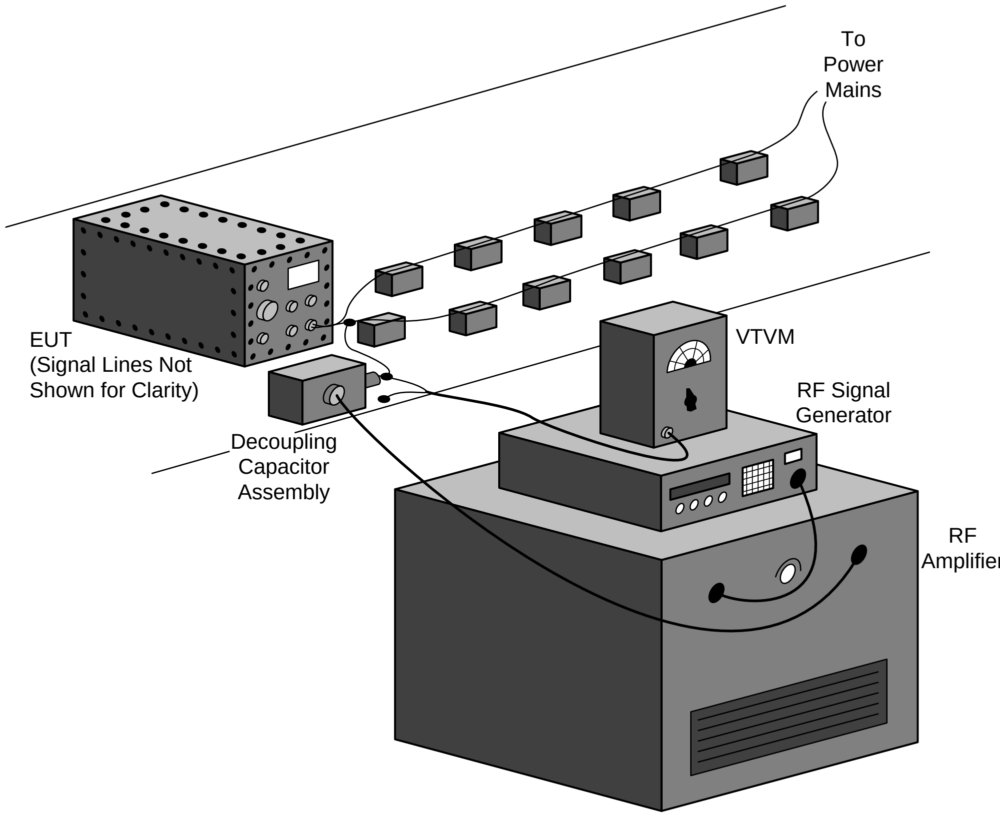


Fig 2-12. CS02 test setup.

2.9 CS06, Conducted Susceptibility, Voltage Spikes

Purpose: The purpose of this requirement is to control susceptibility of the EUT to transient spikes injected onto its ungrounded input power leads.

Applicability**:** This requirement is imposed on all equipment drawing current from a power bus. The spike waveform imposed under MSFC-SPEC-521B is twice the line voltage (100 V maximum) with an on time (10-percent height) of 10 μs superimposed upon the power waveform in both negative and positive polarities. The test setup is shown in figure 2-13. Note: The spike generator required to perform this test is expensive.

Prior to testing, the transient generator is attached across a 5-Ω noninductive resistor. The spike amplitude and duration are observed using an oscilloscope and voltage probe and are adjusted to the required values that are not to be exceeded during the testing.

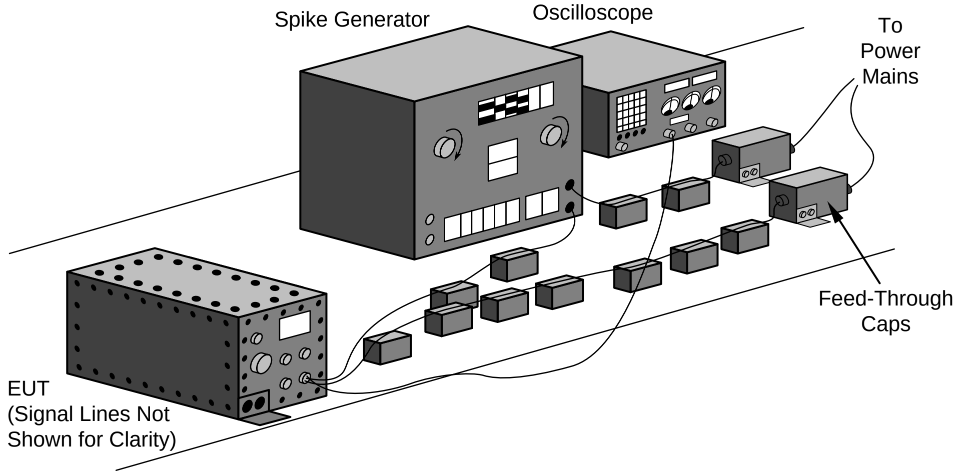


Figure 2-13. CS06 test setup.

2.10 RS02, Magnetic Induction Field Radiated Susceptibility

Purpose: The purpose of this requirement is to control and determine the susceptibility of the EUT to magnetic induction fields. The EUT shall demonstrate no susceptibility to transient spikes and power line frequencies magnetically induced on the signal input and output cable bundles.

Applicability: The spike is only induced into the EUT attached signal cables. To achieve this, tape the spike-carrying wire to the cable under test for a specified parallel length.

The spike waveform is given in the governing specification. For space station, specification SSP 30237 calls out a spike of 240 V (twice line voltage).[2-4] The test spike is injected at the rate of 6 to 10 pulses per second for a period of 5 min. The EUT is subjected to positive and negative spikes. Specialized equipment is required to perform this test (see test setup in fig. 2-14).

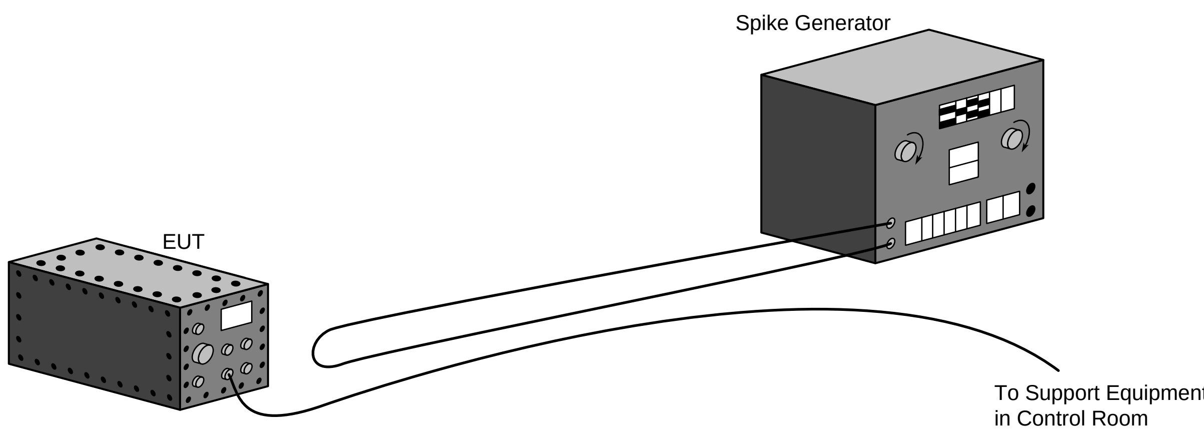


Figure 2-14. RS02 test setup.

2.11 RS03, Electric Field Radiated Susceptibility, 14 kHz to 10 GHz

Purpose: The purpose of this requirement is to control and determine the susceptibility of the EUT to radiated electric fields.

Applicability: RS03 is universally applied to all EUT's. Levels and frequency bands depend on the program, intended use, and placement of EUT relative to high power transmitters. MSFC-SPEC-521B requires the EUT to demonstrate immunity to an electric field strength of at least 2 V/m from 14 kHz to 10 GHz and 13 to 15 GHz.

From 14 kHz to 10 GHz, the electric field is usually amplitude modulated with a 1-kHz sine wave. Above 1 GHz, various pulse and frequency modulations (FM's) are required. MSFC-SPEC-521B requires modulation with a 32-kHz square wave (see test setup in fig. 2-15).

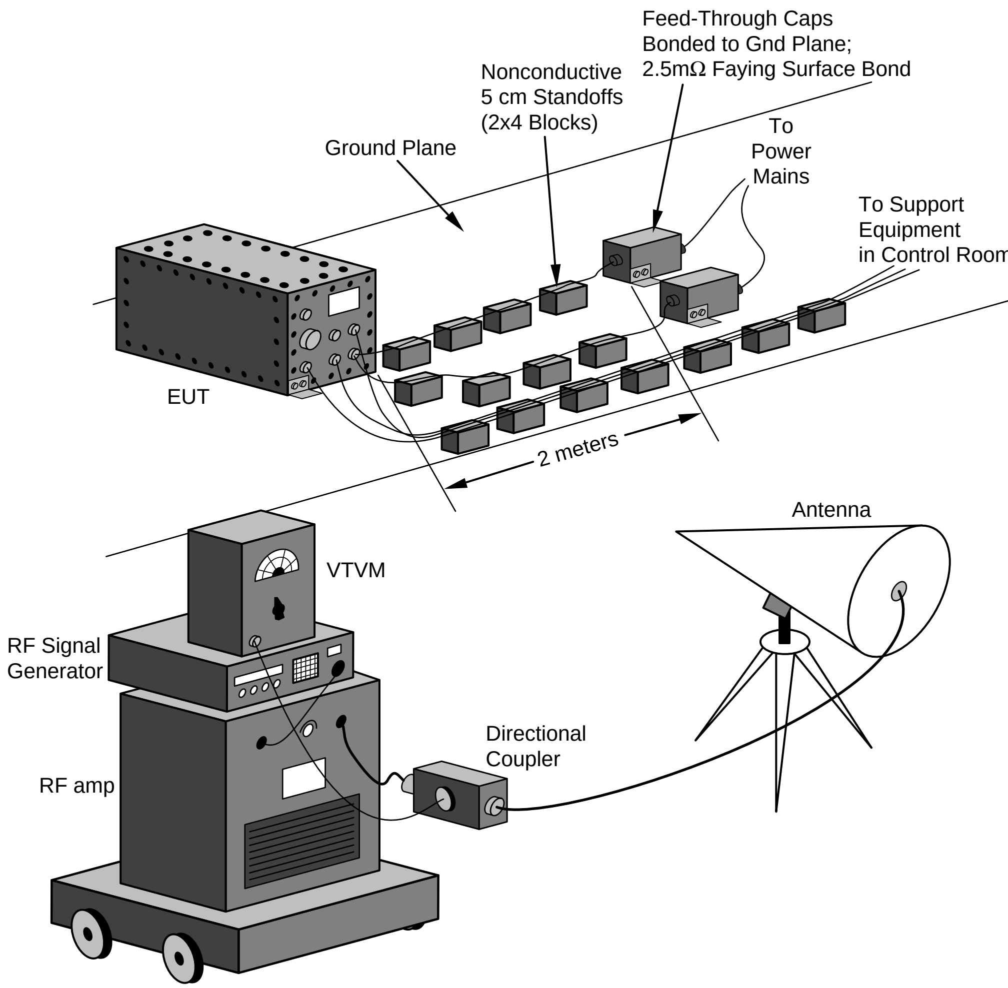


Figure 2-15. RS03 test setup.

REFERENCES

2-1. MIL-STD-461A, EMI Characteristics for Equipment, August 1, 1968.

2-2. MSFC-SPEC-521B, EMC Requirements on Payload Equipment and Subsystems, August 15, 1990.

2-3. SSP 30237, Space Station Electromagnetic Emission and Susceptibility Requirements for Electromagnetic Compatibility, current issue.

3. GENERAL ELECTROMAGNETIC COMPATIBILITY DESIGN GUIDELINES

3.1 Introduction

To design electrical equipment for EMC and to meet EMC requirements, it is necessary to control the emission of undesired electromagnetic energy to and from equipment. The limiting, diverting, or absorbing (referred to in this chapter as suppression) of unwanted electromagnetic energy is accomplished at different design levels. This chapter deals with three various design suppression levels: section 3.2, the board level; section 3.3, equipment interfaces; and section 3.4, enclosures. Because so much aerospace equipment contains and/or is powered by switched-mode power supplies, it is felt that this topic deserves special attention (section 3.5). Grounding is another topic deserving special attention (section 3.6).

3.2 Suppression at the Circuit Board Level

EMI suppression at the circuit board level involves such measures as component selection, limiting signal bandwidths and speeds, board layout, and grounding practices. The following subsections address these suppression measures and offer general design guidelines for EMI suppression.

3.2.1 Component Selection

One of the basic building blocks for any electrical design is selection of the components. Selection of components for EMC ramifications is equally as important as selection for performance. Except for wideband video and circuits employing oscillators, analog circuits are generally much quieter than digital circuits. Because digital circuits are noisier, this section emphasizes the selection of digital components for suppression of EMI.

The most important issue in selecting digital components for low-noise characteristics is rate of change of energy. The noise voltage induced into a victim circuit from a noise source circuit is:

$$\mathrm{V} = -M\ dI/dt\ , \qquad (3\text{-}1)$$

where M is the mutual inductance between the two circuits and the coupling is magnetic in nature. Or:

$$\mathrm{V} = C\ d\mathrm{V}/dt\ , \qquad (3\text{-}2)$$

where C is the capacitance between the two circuits. Coupling is electric in nature.[3-1] Mutual inductance, M, depends on current loop areas of source and victim, orientation, separation distance, and the heights of the circuits above ground. Source and victim current loops are analogous to the primary and secondary windings of a transformer (fig. 3-1). Capacitance, C, depends on the distance between conductors, associated effective areas, and Z, the impedance to ground of the victim circuit. The source and victim conductors act as a parallel plate capacitor (fig. 3-2).

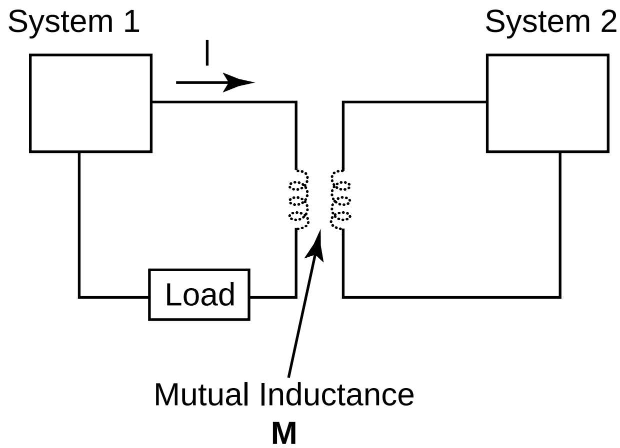


Figure 3-1. Noise coupling via magnetic induction.

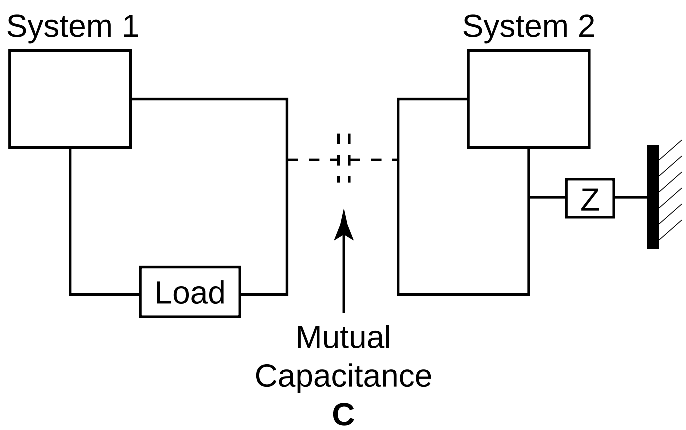


Figure 3-2. Noise coupling via electric induction.

3.2.1.1 Logic Families and *d*V*/dt*

Table 3-1 shows various digital family rises time and voltage rates of change (*d*V*/dt*). The faster the rise time and the higher the voltage swing, the larger the *d*V*/dt*. Using the slowest rise time to achieve the desired function can lower the amount of noise coupling. Another reason for using slower rise time is to limit the higher frequency harmonics of the digital signal. Because the circuit traces on printed circuit boards (PCB's) can act as antennas and radiate noise at higher frequencies, limiting the unnecessary harmonics in a digital signal prevents radiation of these higher frequency harmonics. Section 3.2.1.2 addresses the transformation of time-domain signals into the frequency domain and how slower transition times and lower repetition rates lower and/or eliminate higher frequency harmonics.

Table 3-1. Rise time and voltage rate of change for various logic families.[3-2]

Logic Family	Rise Time (ns)	Voltage Swing (V)	*d*V/*dt* (V/ns)
CMOS 5 V	100	5	0.05
CMOS 12 V	25	12	0.48
CMOS 15 V	50	15	0.30
HCMOS	10	5	0.50
TTL	10	3	0.30
ECL 10 k	2	0.80	0.40
ECL 100 k	0.75	0.80	1.10

3.2.1.2 Fourier Transform and Frequency Spectrum Envelope

Every periodic signal is be represented in the time domain by the Fourier series expansion:[3-3]

$$f(t) = \frac{A_o}{2} + \sum_{n=1}^{\infty}(A_n \cos(n\omega_o t)+B_n \sin(n\omega_o t)) \quad , \qquad (3\text{-}3)$$

where

$$A_o = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\,dt \quad , \qquad (3\text{-}3a)$$

$$A_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\cos(n\omega_o t)\,dt \quad , \qquad (3\text{-}3b)$$

$$B_n = \frac{2}{T}\int_{t_o}^{t_o+T} f(t)\sin(n\omega_o t)\,dt \quad . \qquad (3\text{-}3c)$$

Equation (3-3) means that a periodic signal is a summation of sinusoidal signals of multiple frequencies and amplitudes. Therefore, the signal has corresponding representation in the frequency domain. The Fourier transform converts signals from time domain to frequency domain. Equation (3-4) for this transform is found in reference 3-2:

$$F(\omega) = \int_{-\infty}^{\infty} f(t)\,e^{-j\omega t}\,dt \quad . \qquad (3\text{-}4)$$

A given signal (e.g., a square wave with finite transition times) occupies a frequency spectrum.

In the interest of time and practicality, the Fourier envelope approximation method is used to quickly calculate the worst-case frequency spectrum envelope. For a given periodic square signal with finite rise and fall times, shown in figure 3-3, the frequency spectrum envelope is calculated knowing:

Peak amplitude A (volts, amperes)

Pulse width τ (measured at half-max)

Period T

Rise time τ_r for transition from 0.1 to 0.9 A.

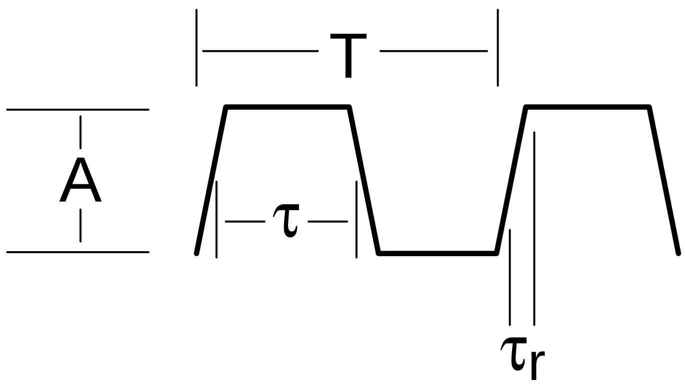


Figure 3-3. Periodic square-wave signal.

The frequency spectrum envelope shown in figure 3-4(a) is calculated using the above information and equations derived from the trigonometric Fourier transform. Amplitude of the signal in frequency domain (A_f) is calculated using:

$$A_f = 2A\frac{\tau}{T}\,, \tag{3-5}$$

where A is peak amplitude in the time domain. Corner frequencies, f_1 and f_2, are calculated using equations (3-6) and (3-7) from reference 3-4:

$$f_1 = \frac{1}{\pi\tau}\,, \tag{3-6}$$

and

$$f_2 = \frac{1}{\pi\tau_r}\,. \tag{3-7}$$

It should be noted here that in practice the signal waveforms are not completely symmetrical. In this case, it is important to use the faster of the two transition times, the rise time or the fall time, in equation (3-7). Figure 3-4(a) shows that between the first corner frequency, f_1, and the second corner frequency, f_2, the amplitude decreases at a rate of 20 dB per decade of frequency.

At frequencies above f_2, the amplitude decreases at a rate of 40 dB per decade of frequency. Figure 3-4(b) shows a frequency spectrum envelope overlaid upon an actual frequency spectrum.

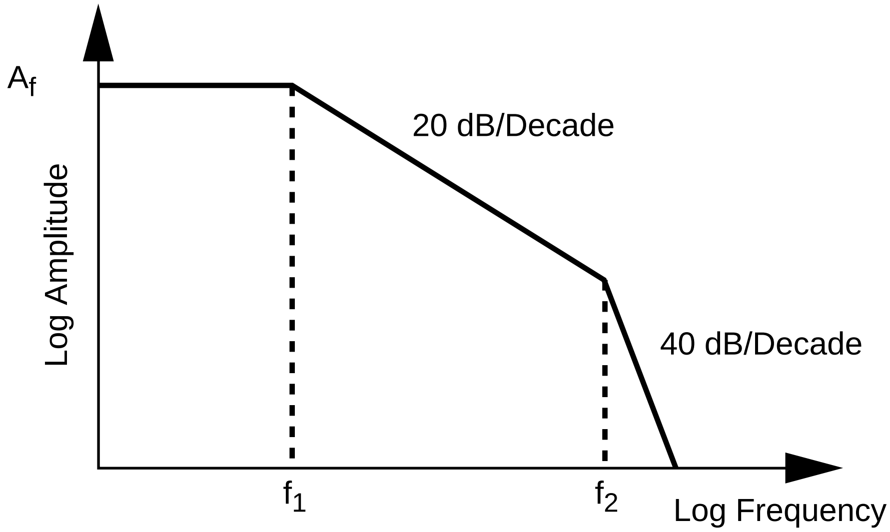


Figure 3-4(a). Frequency spectrum envelope.

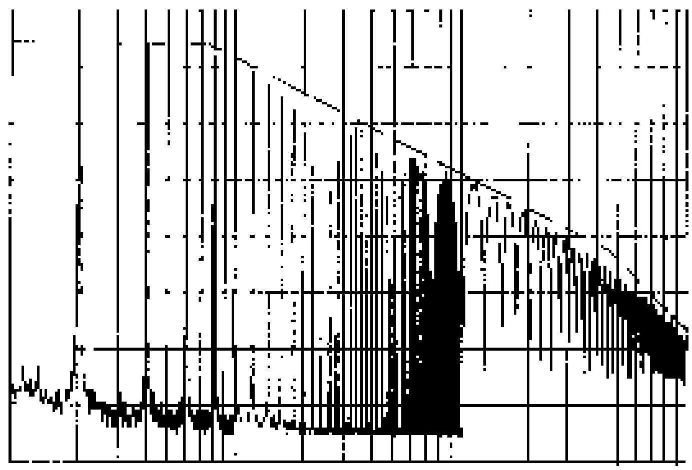


Figure 3-4(b). Frequency spectrum and frequency spectrum envelope.

While this method does not yield an exact frequency spectrum plot, the resulting frequency spectrum envelope does provide a worst-case envelope for a given time-domain signal and other important information. Changes in duty cycle and transition times reduce the frequency spectrum envelope. For a 5-V_p, 500-kHz signal with a 50-percent duty cycle and transition times of 10 ns, f_1 is 318.3 kHz, f_2 is 31.8 MHz, and A_f is 5 V_p. By changing the duty cycle to 30 percent and the transition times to 100 ns, f_2 becomes 3.18 MHz and A_f becomes 3 V_p. This implies that noise amplitudes are reduced and noise frequency amplitudes lowered (table 3-2).

Table 3-2. Frequency spectrum envelope calculations.

A	T (1/f)	τ ($T \times$ duty cycle)	τ_r	A_f	f_1	f_2
5 V	2 μs (1/500 kHz)	1 μs (2 μs $\times$ 50%)	10 ns	5 V	318 kHz	31.8 MHz
5 V	2 μs (1/500 kHz)	0.6 μs (2 μs $\times$ 30%)	100 ns	3 V	531 kHz	3.18 MHz

3.2.1.3 Logic Families and *dI/dt*

As a result of stacking the output stage of the logic circuit in the chip (fig. 3-5(a)), when the logic is switched, the transistors typically turn off slower than they turn on and draw large amounts of transient current from V_{cc} during the transition. This induces transients on the Vcc trace and ground. Notice that the output stage of the TTL circuit in figure 3-5(a) contains a current limiting resistor. The CMOS circuit has no current limiting resistor and, consequently, draws larger currents (*dI/dt* sometimes as high as 5,000 A/s) than TTL.[3-5] One way to limit these surges is through the use of decoupling capacitors. The decoupling capacitor, which will supply the necessary instantaneous currents while the chip is switching, is a capacitor connected between V_{cc} and ground (fig. 3-5(b)). It is important to remember to make the capacitor leads as short as possible to reduce parasitic inductance and to mount the capacitor close to the decoupled chip to reduce loop area.[3-6]

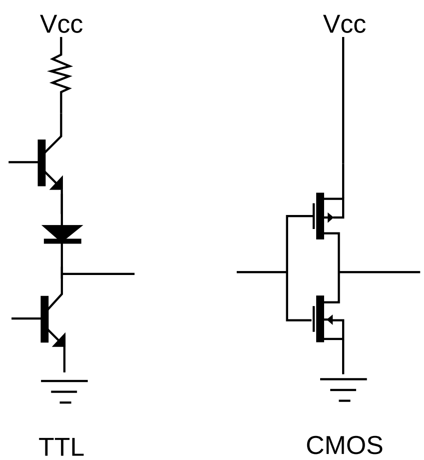


Figure 3-5(a). Logic output drivers.

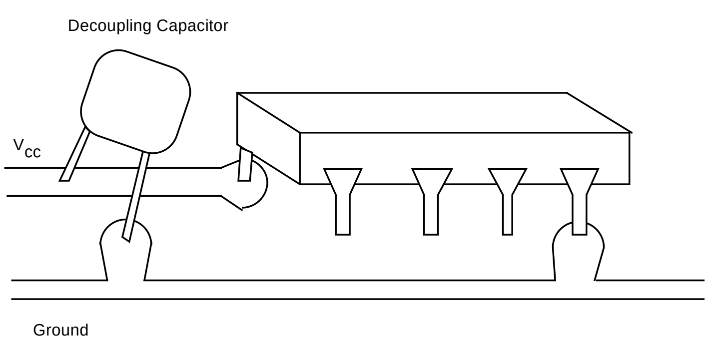


Figure 3-5(b). IC chip and decoupling capacitor.

3.2.1.4 Logic Family Noise Margins

Noise margins are estimated by using data usually provided in vendor data books. This noise margin represents a maximum budget allowable for noise riding on the input signal. Any voltage that exceeds this noise margin is potentially propagated as noise by the logic chip. These parameters, explained below, are used to calculate a conservative noise level immunity. The equations are:

$$V_{hnl} = V_{oh}(\min) - V_{ih}(\min) \ , \tag{3-8}$$

$$V_{lnl} = V_{ol}(\max) - V_{il}(\max) \ , \tag{3-9}$$

where V_{hnl} is the noise level immunity for the logic chip when the logic state is high, V_{lnl} is the noise level immunity for the logic chip when the logic state is low, $V_{oh}(\min)$ is minimum high output generated by the driving gate, $V_{ih}(\min)$ is the minimum high input allowable for the driven gate, $V_{ol}(\max)$ is the maximum low output generated by the driving gate, and $V_{il}(\max)$ is the maximum low input allowable for the driven gate.

Table 3-3 shows typical noise margin for various logic families. It is interesting to note that while CMOS logic has the highest noise immunity, it also generates more noise than other logic families, which can lead to incompatibilities with other logic families.

Table 3-3. Typical noise margin for various logic families.[3-5]

Logic Family	Noise Margin (mV)
TTL	400
CMOS 5 V	1,000
CMOS 15 V	4,500
ECL 10 k	125
ECL 100 k	100

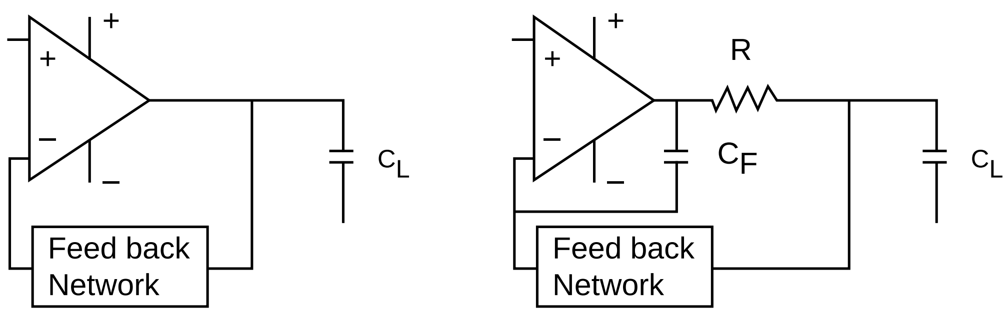
3.2.1.5 Analog Components

Analog circuits in general do not exhibit the *dI/dt* and *d*V*/dt* of digital circuits and, therefore, do not generate excessive emissions. However, analog circuits may unintentionally operate outside their design bandwidths and become EMI sources. In these instances, instability in analog amplifier circuits is usually the culprit. These amplifier circuits may oscillate in the high-frequency range (MHz) due to feedback loop instability, poor decoupling of input stages from power line noise, and output instability due to capacitive loads. Because the designer is much more knowledgeable of the design than the EMC engineer, it is difficult for the EMC engineer to offer specific advice in this area. However, a few points of general advice are offered. Any prototypical amplifier should be checked for high-frequency instability. Poor decoupling may be caused by the parasitic inductance of power leads resonating with decoupling capacitors. Cure this by adding additional resistance in series with the decoupling capacitor or by adding a ferrite bead (addressed in section 3.3.3.2). Output instability due to capacitive loads (10 m of RG58 50-Ω coax cable has approximately 1,000 pF of capacitance) may be cured by using a small value resistor in series and a small direct feedback capacitor. This compensates for the phase lag induced by the capacitive load. The phase lag induced frequency f is given by equation (3-10):

$$\text{Phase lag @ } f = \tan^{-1}(f/f_c) \text{ degrees} \quad , \tag{3-10}$$

where $f_c = 1/(2\pi R_{\text{out}} C_L)$ and R_{out} is the output resistance of the op-amp.

In figure 3-6, the circuit on the left shows the amplifier circuit without the instability correction and the circuit on the right shows resistance and capacitance added to cure amplifier instabilities. R is usually on the order of 10 to 100 Ω and C_F is typically about 20 pF.[3-1]



R and C_F Used to Isolate Large Capacitive Load C_L

Figure 3-6. A cure for instabilities due to capacitive loads.

3.2.2 Layout

A cost-effective approach to meet EMC requirements and prevent interference is to consider the layout of the equipment (board level and box level) at the beginning of the design activity. Two important principles of equipment layout are: (1) partitioning the equipment (board) to control interference and (2) controlling circuit trace layouts on the board to minimize loop areas.

3.2.2.1 Equipment and Board Partitioning

In a typical equipment chassis or on a typical board, there are equipment sections or components that produce interference, that are susceptible to interference, and that are neither interference producers nor susceptible to noise. Partitioning these sections or components is important for achieving EMC internal to the equipment and for meeting equipment-level EMI requirements.

In equipment, partitioning may mean putting sensitive sections in a shielded subenclosure and filtering the interfaces between sensitive and nonsensitive sections (shielding is addressed in section 3.4.1 and filtering in section 3.3). Another way of partitioning is to separate a digital card (interference producer) attached to a motherboard from a low-level analog card (susceptible to interference) by placing nonsensitive analog cards between the two cards on the motherboard. Figures 3-7(a) and 3-7(b) illustrate these two methods of partitioning. Figure 3-8 shows an example of using a shielded subenclosure for partitioning. The power supply in figure 3-8 is in a shielded enclosure to prevent the power supply from interfering with other electronics in the box.

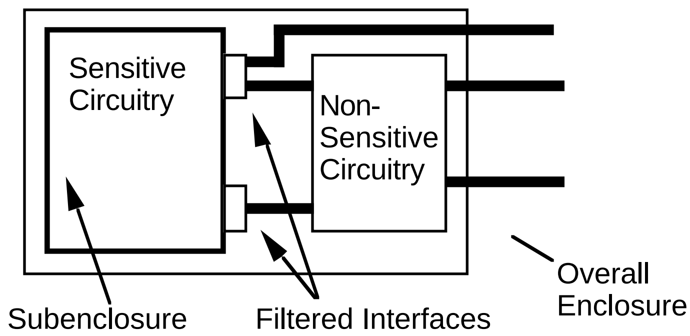


Figure 3-7(a). Partitioning with shielded subenclosure.

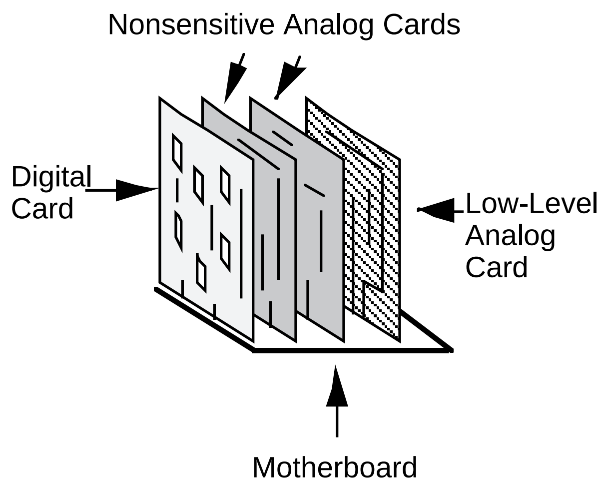


Figure 3-7(b). Partitioning on motherboard.

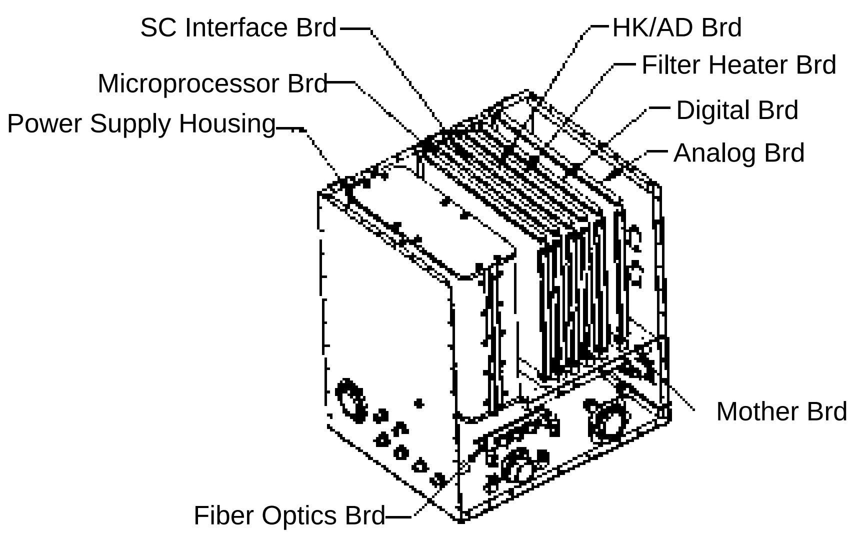


Figure 3-8. Use of shielded subenclosure (external enclosure top and side removed).

In a board layout, there are several ways of partitioning the board to achieve EMC among the board components. Three things to remember in circuit board layout are: (1) separate low-level analog and digital circuitry and use separate isolated ground planes for each; (2) use different areas for low, medium, and high speed logic; and (3) place high-speed components closest to edge connectors and low-speed ones farthest from connectors (to reduce trace impedance and loop areas of high-speed signals). Figures 3-9(a) through 3-9(d) show examples of board partitioning.[3-2]

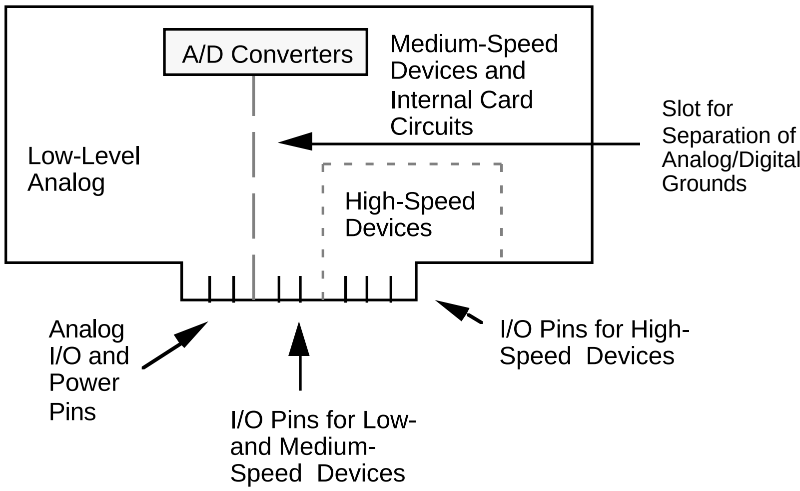


Figure 3-9(a). Board layout showing analog/digital separation.

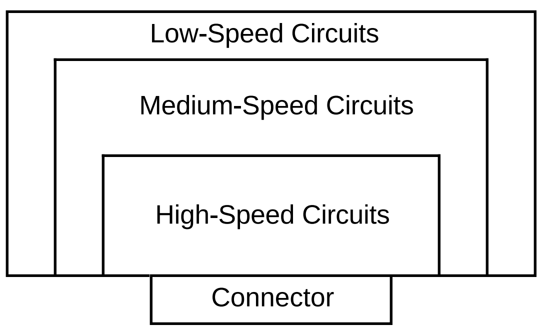


Figure 3-9(b). Suggested board layout for multispeed circuits.

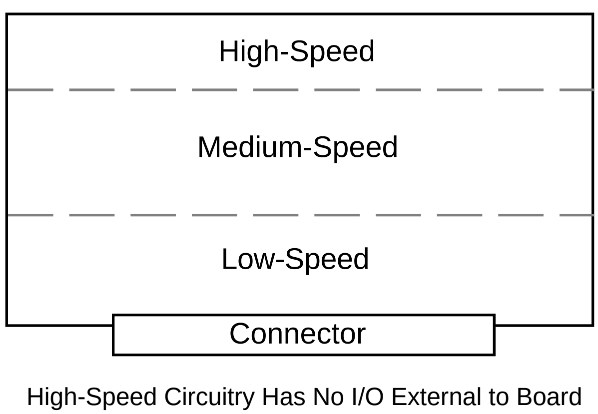


Figure 3-9(c). Suggested board layout for board with only low-speed I/O.

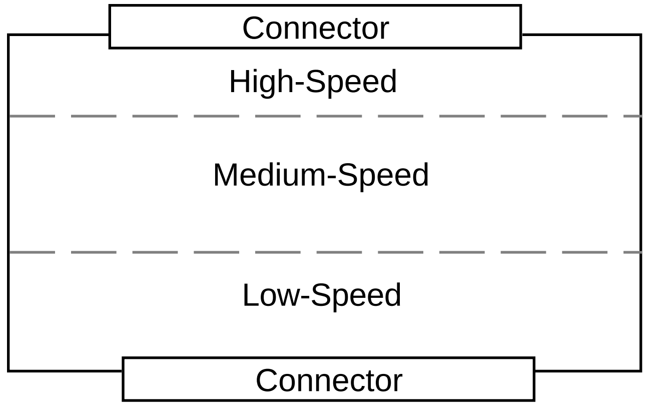


Figure 3-9(d). Suggested board layout with separate connectors.

3.2.2.2 Trace Layouts

The trick in circuit board trace layout is to minimize trace lengths and trace loop areas. This minimizes radiated emissions and susceptibility. The effect of loop areas on interference coupling was explained in section 3.2.1. Minimizing trace lengths reduces trace impedance and prevents the trace from becoming an effective antenna for transmitting or receiving undesired electromagnetic energy. Table 3-4 gives a listing of general rules to apply in trace layout and design.

Table 3-4. General rules for trace design and layout.

• Route power and return traces as closely as possible. Make power and return traces wider than 1 mm when possible.
• Minimize etching of Vcc and returns. Extend supply and ground return traces into large areas (fig. 3-10).
• Dedicate 0-V returns for analog circuits.
• If possible, devote one side of the board for a ground plane (double-sided boards).
• When using high-speed logic, consider raised power distribution (fig. 3-11).
• Long parallel traces provide excellent situations for capacitive coupling interference from one trace to the other. Increasing spacing between traces or adding a 0-V trace between signal traces reduces this coupling.
• Keep high-speed traces away from board edges.
• Allocate 1 in every 10 board connector pins as a 0-V pin.

Traces routed close together look like a transmission line to noise currents on the traces. Using wider traces or larger areas for traces lowers the inductance of the traces.[3-1] The raised power distribution system provides a low-impedance power supply and return trace over a wide frequency range.[3-2] The longer the length of parallel traces, the greater the mutual capacitance and the greater the coupled noise from one circuit to the other. Shortening parallel lengths, increasing space between lengths, or adding a 0-V trace (grounded at both ends) between signal traces reduces this mutual capacitance (equation (3-2)). Because wiring and traces have a finite resistance and a finite inductance, a noisy circuit (digital circuit or an analog circuit carrying noise currents) sharing a return trace with another sensitive analog circuit induces noise voltages into the sensitive analog circuit. Dedicating returns or allocating many return paths reduces currents that cause noise voltages in any one return.

Using multilayer boards in equipment design prevents some EMI problems from occurring. Different types of signal traces are placed on different board layers and are routed perpendicular to signal traces on other layers. Also, whole layers can be dedicated to signal planes or ground planes, minimizing trace impedance. A point of caution is to minimize the number of holes in multilayer board ground plane layers; too many holes raise the impedance of the ground plane.

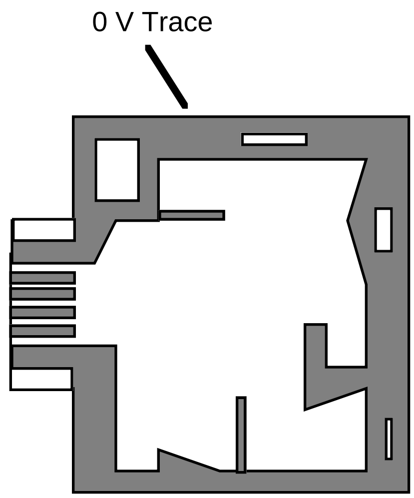


Figure 3-10. Minimized etching of 0-V trace.

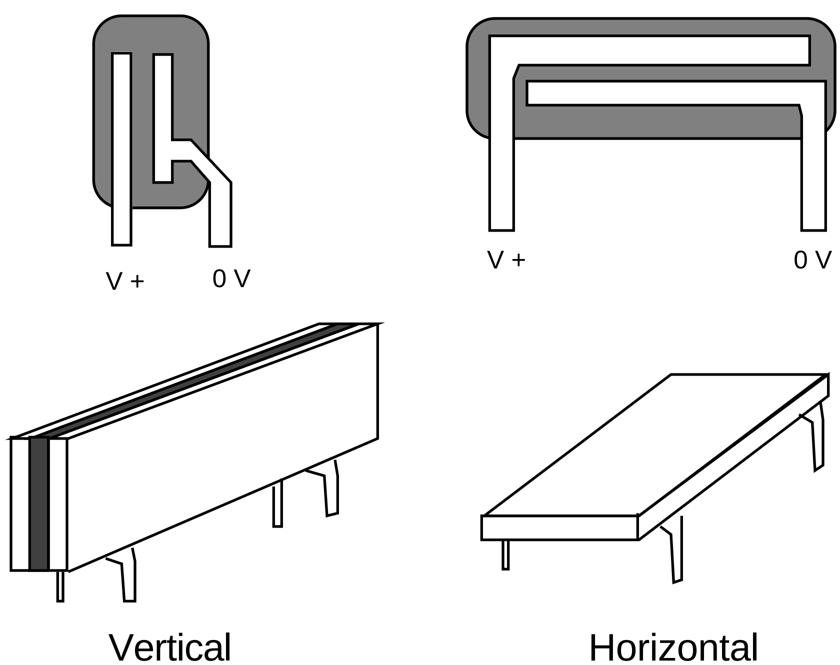


Figure 3-11. Raised power distribution.

3.3 Suppression Through Filtering and Isolation

Filtering and isolation is analogous to shielding (discussed in section 3.4). The filtering and isolation prevents the entry or exit of conducted EMI from equipment, whereas, shielding prevents the entry or exit of radiated EMI from equipment. Filters and isolators are used to attenuate EMI by bypassing, absorbing, or reflecting the noise. Because volumes of work are available on filter design, this section attempts only to give an overview of filtering and isolation and tries to point out short-comings of ideal filter and isolator models.

3.3.1 Types of Conducted Noise

In order to properly design filters, it is important to understand the types of conducted noise. The first type, known as differential mode (DM) noise, is propagated out one wire and returned on the other. This noise is generated by clock signals or switching waveforms in power supplies. DM noise amplitudes are usually minimal above 2 MHz because line-to-line and line-to-ground capacitance and wiring inductance tend to filter this type noise.[3-7] The other type of conducted noise, common mode (CM) noise, travels in the same direction in both wires and returns through the ground plane or structure. In power and signal systems that have a single reference to ground or single-point ground, CM noise is capacitively coupled to the ground plane or structure. Because of this capacitive coupling, CM noises are generally high frequency (above approximately 2 MHz).[3-7] Figure 3-12 gives examples of DM and CM noise. Because the filter design for these two noise types is different, it is important to understand these types of conducted noises.

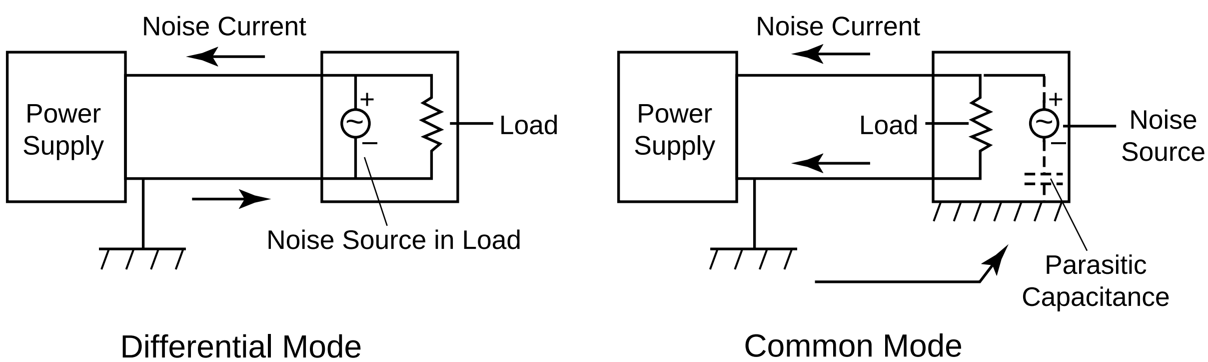


Figure 3-12. DM and CM noise.

3.3.2 Capacitors, Inductors, and Actual Properties

In designing the filter, it is important to note that the capacitor or inductor being used is not an ideal component and will not act as such. A capacitor, even the leadless surface mount type, exhibits parasitic inductance and resistance. “Parasitic” describes the capacitances and inductances that do not appear on engineering drawings, but nevertheless exist and cause odd things to happen to the desired signal or waveform.[3-7] The term “stray capacitance” is a commonly used term that means the capacitance between a conductor and its surroundings. A good example of “stray capacitance” is between a switching transistor and the heat sink upon which it rests, typically 50 to 150 pf. As a general rule, when trying to bypass a certain frequency, try to keep the reactance of the capacitor being used around 0.1 Ω. A lower reactance (0.01 Ω) may tend to self-resonate.

Figure 3-13 shows models of a capacitor and an inductor and includes parasitics. The capacitor parasitics are lead and plate resistance and inductance, dielectric losses, and skin effects losses.

The inductor parasitics are lead and winding resistance, turn-to-turn and turn-to-core capacitance, dielectric losses of insulation, eddy current losses, hysteresis losses, and skin effects losses. One consequence of parasitics is that they cause the inductor or capacitor of a filter to self resonate at its resonant frequency (100 kHz to 20 MHz for capacitors and 2 to 100 MHz for inductors)[3-7] and create EMI problems. Another consequence is that the impedance of the inductor or capacitor is nonideal above the frequency which the parasitic components begin to have an appreciable impedance (fig. 3-14).

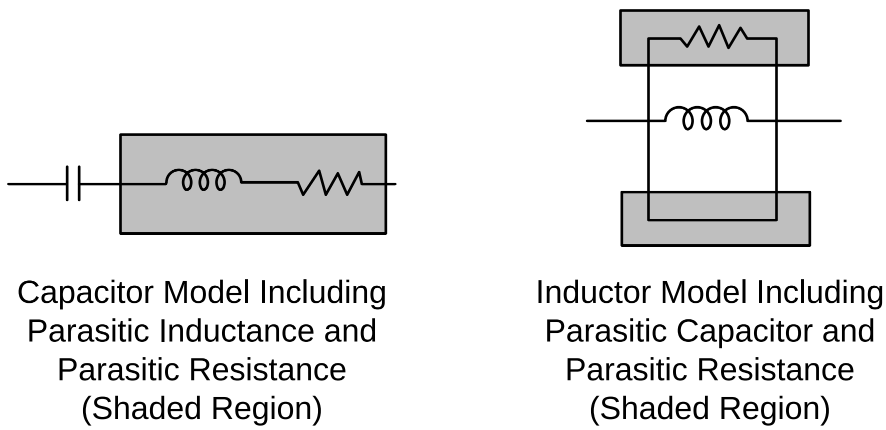


Figure 3-13. Capacitor and inductor models including parasitics.

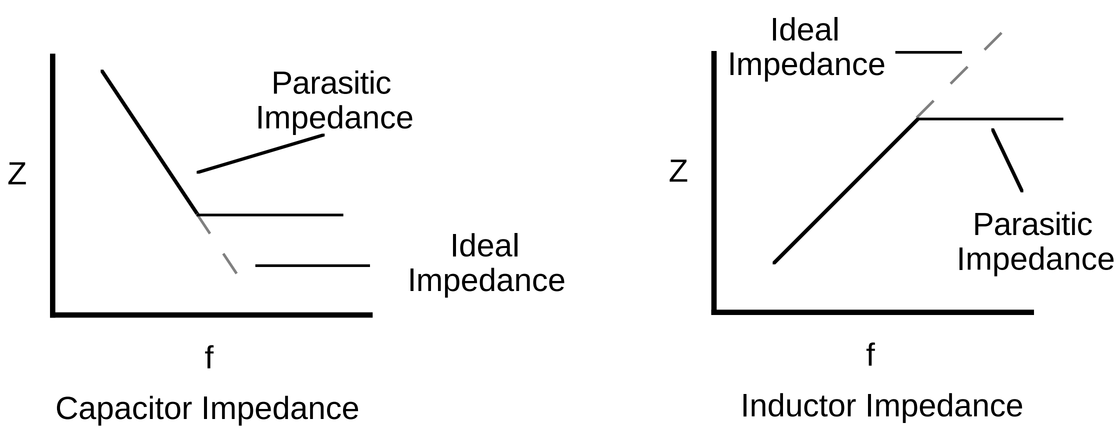


Figure 3-14. Inductor and capacitor impedance.

3.3.3 Filtering Overview

As stated before, the purpose of the EMI filter is to prevent the entry or exit of undesired electromagnetic energy from equipment. Because MSFC EMC-CE and susceptibility requirements apply only to power lines, only power line filtering is addressed in this section.

A filter absorbs the noise energy through the use of lossy elements such as resistors and ferrite components, or reflects the noise energy back to the source through use of reactive elements. Generally, EMI filters are low pass filters with effectiveness depending on the impedances of the elements at either end of the filter.[3-1]

For a filter that attenuates EMI by reflecting noise, the filter should provide a maximum impedance mismatch. If the load impedance is low, the impedance of the filter from the load viewpoint should be high. If the load impedance is high, the impedance of the filter from the load viewpoint should be low. Figure 3-15 gives filter configuration examples for various load and source impedances.[3-8]

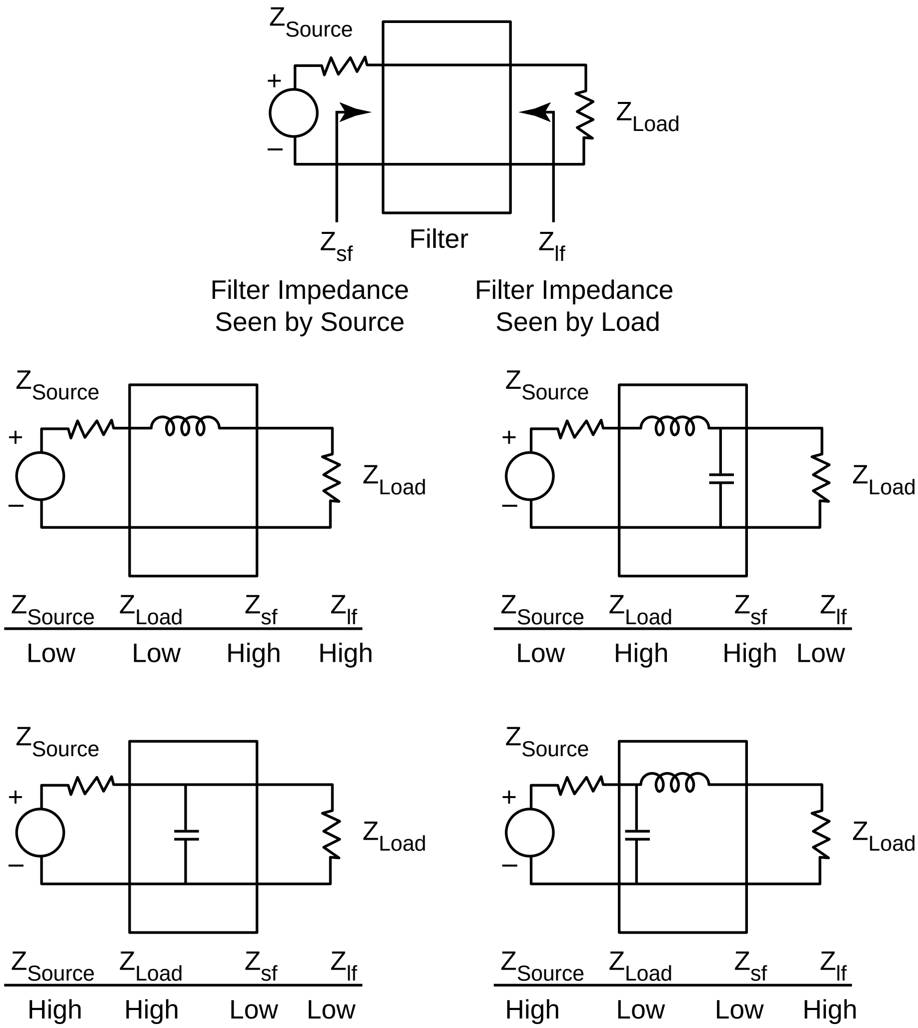


Figure 3-15. Filter configuration examples.[3-7]

EMI filters are single-section filters or several single-section filters cascaded together for more attenuation. It has been demonstrated that a two-section filter has a lower optimum weight than a single-section filter when by design both have identical filtering properties.[3-9] The number of sections and configuration are not limited to this presentation. Chapters 4 and 5 present additional information on filtering.

It is important to remember to isolate the input and output cables of the filter. Isolating input and output cables from each other prevents the cables from coupling to each other and bypassing the filter. Isolation may be accomplished by placing the input cables and the output cables on opposite sides of the filter. However, to properly isolate the cables and prevent noise from bypassing the filter, the filter may have to be shielded by placing it in a shielded subenclosure. Section 3.2.2.1 discusses equipment partitioning and section 3.3.4 has further discussions on isolation and shielding.

3.3.3.1 Filters and Power Supply Stability

When designing a filter for a switched-mode power supply input, it is important to remember that an improperly designed filter may also cause instability problems. The switched-mode power supply has a negative input resistance at low frequencies, and the addition of an input filter may cause the power supply to oscillate.[3-9]

A switched-mode power supply demands constant input power. If the input voltage drops, the power supply compensates by drawing more current. The V-I curve for a power supply, shown in figure 3-16, implies that for a given input voltage, V_a, the power supply draws a given amount of current, I_a. If V_a increases, I_a decreases and the slope of this curve is a negative value ($dV/dI < 0$). If resistance is defined as the rate of change of voltage divided by the rate of change of current at a given point on the V-I curve ($R = dV/dI$), the resistance at that point is negative.[3-10] The work of R.D. Middlebrook[3-11] and others has demonstrated that impedance of the input filter, as seen by the power supply, must be less than the negative input impedance of the power supply to avoid power supply instabilities.

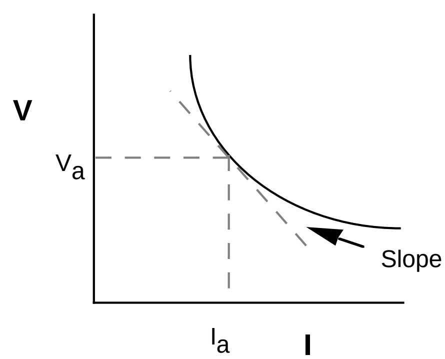
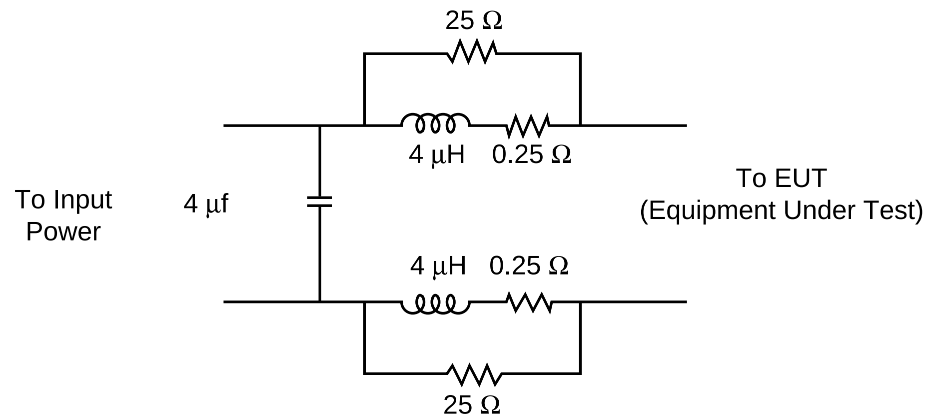
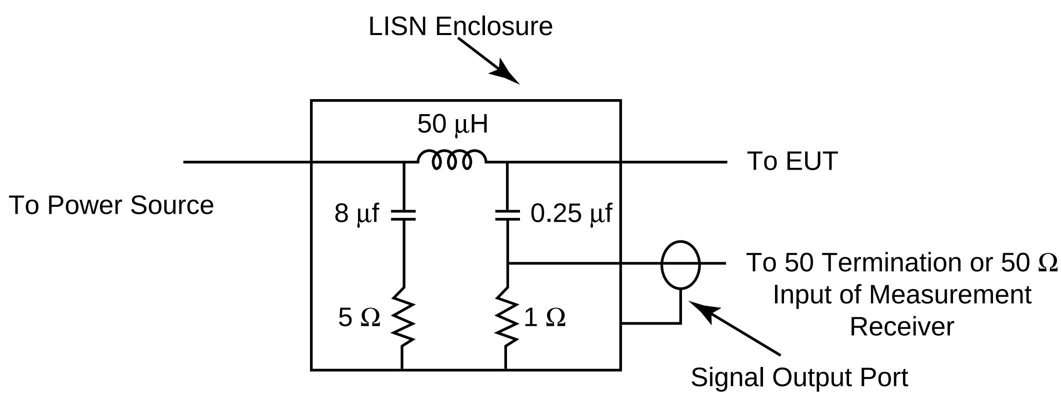


Figure 3-16. Switched-mode power supply V-I curve.

The impedance experienced by the power supply includes the impedances of the power source and the power bus. During EMI testing, this bus impedance is predominately that of the LISN. LISN's are discussed in greater detail in section 5.1.1 and the schematics of two types are shown in figure 3-17. The first LISN is the type used for TT01 testing per MSFC-SPEC-521B[3-20] and the second is the LISN used for MIL-STD-461D[3-22] testing. Note: MIL-STD-461D LISN's are set up with one LISN on the lead wire and one on the return, thereby doubling the impedance of the LISN experienced by the equipment under test. Usually, the LISN impedance is higher than the bus impedance of the spacecraft.



MSFC-SPEC–521B LISN



MIL-STD–461D LISN

Figure 3-17. LISN schematics.

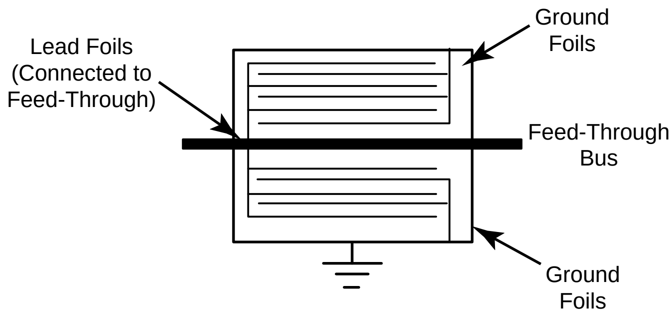
3.3.3.2 Special Filtering Components

Several types of special filtering components are available to the design engineer. Three of these components, the ferrite core (also known as a ferrite bead), the feed-through capacitor, and the three-terminal capacitor, are used in EMI suppression.

Most ferrite cores are available in three different material types: a manganese-zinc core that provides attenuation up to 40 MHz and two nickel-zinc cores that provide attenuation to 200 MHz and higher.[3-12] These metal-oxide materials are blended with iron oxides to form a magnetic ceramic material with high permeability and high electrical resistivity. These cores are used in antenna baluns and CM chokes and are very effective at higher frequencies.

A feed-through capacitor schematic is shown in figure 3-18(a). One electrode of the capacitor is connected to the feed-through housing and the other electrode to the feed-through bus. The construction of the feed-through capacitor allows it to have a resonant frequency generally well above 1 GHz.[3-13] Several CE tests described in chapter 2 require a 10-μF feed-through capacitor on each power and return line.

A three-terminal capacitor schematic is shown in figure 3-18(b). The parasitic lead inductance of the capacitor allows the three-terminal capacitor to act as a "T" filter.



(a) Feed-Through Capacitor Schematic

(b) Three-Terminal Schematic and High Frequency Model

Figure 3-18. Feed-through and three-terminal capacitors.

3.3.3.3 Common Mode Filtering

The various filter configurations shown in figure 3-15 are DM filters. The other type of conducted noise, CM noise, requires a different type filter. CM filters are usually CM chokes or line-to-ground filters such as feed-through capacitors. The CM choke relies on the magnetic properties of ferrite cores to absorb CM noise.

Figure 3-19 shows a schematic of a multiturn CM choke. The cables are wrapped four to five turns around a ferrite core. The magnetic field (H_{dm}) induced by the DM current (I_{dm}) on one side of the core is canceled by the magnetic field induced by the DM current on the return side of the core. Therefore, the DM current is not attenuated. However, for the CM current (I_{cm}) the magnetic fields (H_{cm}) do not cancel, and the series combination of the inductive reactance and resistive losses of the core attenuate the CM noise. Figure 3-20 shows CM choke configurations.

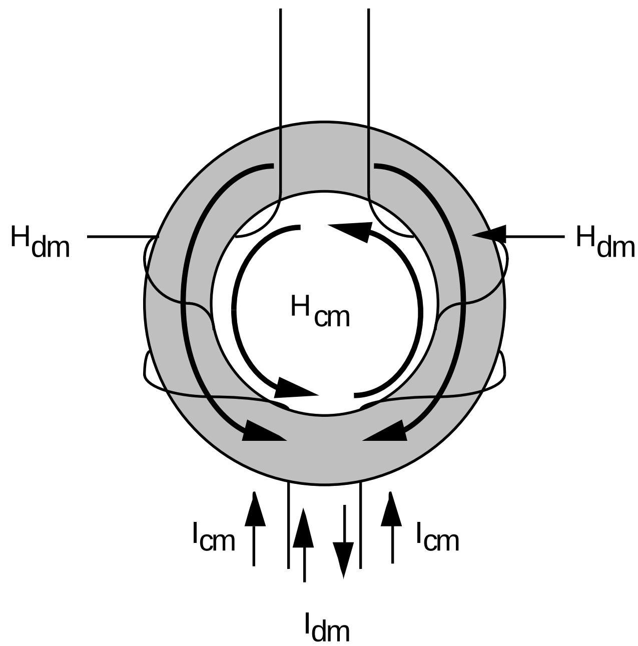


Figure 3-19. CM choke.

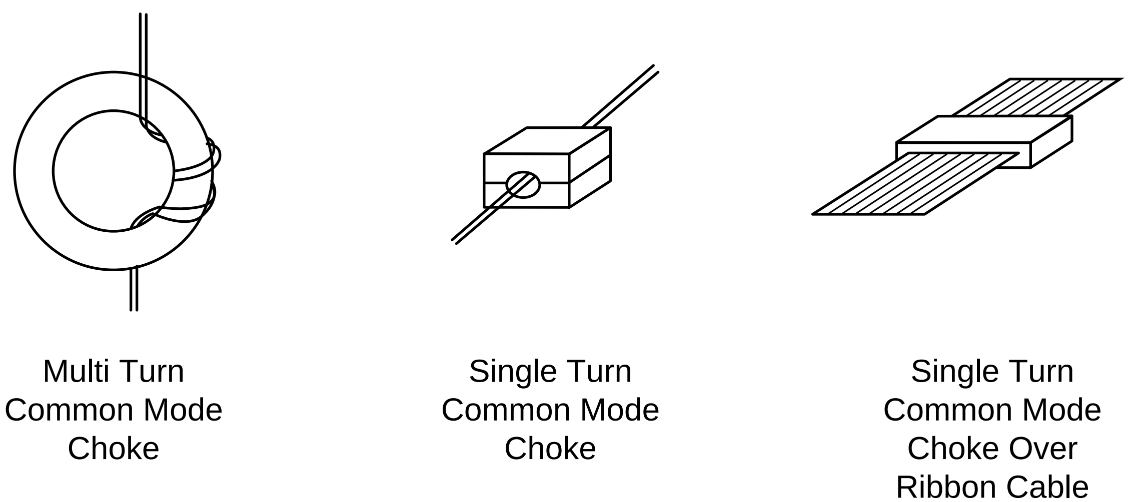


Figure 3-20. CM choke configurations.

3.3.4 Isolation

Isolation is another means of diverting undesired electromagnetic energy. Two methods commonly employed are isolation transformers and opto-isolators. The isolation transformer may be used in ac power circuits, in switched-mode power supplies, and in analog signal circuits such as MIL-STD-1553 data lines.[3-23] The isolation transformer breaks up the ground loop by increasing the impedance of the ground loop. Figure 3-21(a) shows the schematic of a typical isolation transformer. At low frequencies, the capacitance between the primary and secondary windings presents a high impedance in the conducted path. At high frequencies, however, this capacitance impedance is no longer substantial and does not appreciably attenuate CM or DM noise. Addition of a Faraday shield between the primary and secondary windings attenuates high-frequency noise. The primary-to-secondary capacitance is divided between the primary winding and shield and between the shield and secondary winding. For CM reduction, the shield is connected to the transformer housing that is connected to ground. This ground connection impedance, along with the winding to shield capacitance, acts as a voltage divider to reduce CM noise coupled across the transformer. For DM reduction, the shield is connected to the return side of the transformer to short-circuit the DM currents. Figure 3-21(b) shows the schematic of a Faraday shielded isolation transformer for CM reduction. Figure 3-21(c) shows the schematic of a Faraday shielded isolation transformer for DM reduction. Figure 3-21(d) shows the schematic of a triple Faraday shielded isolation transformer that provides common and DM isolation from either side of the transformer.[3-14]

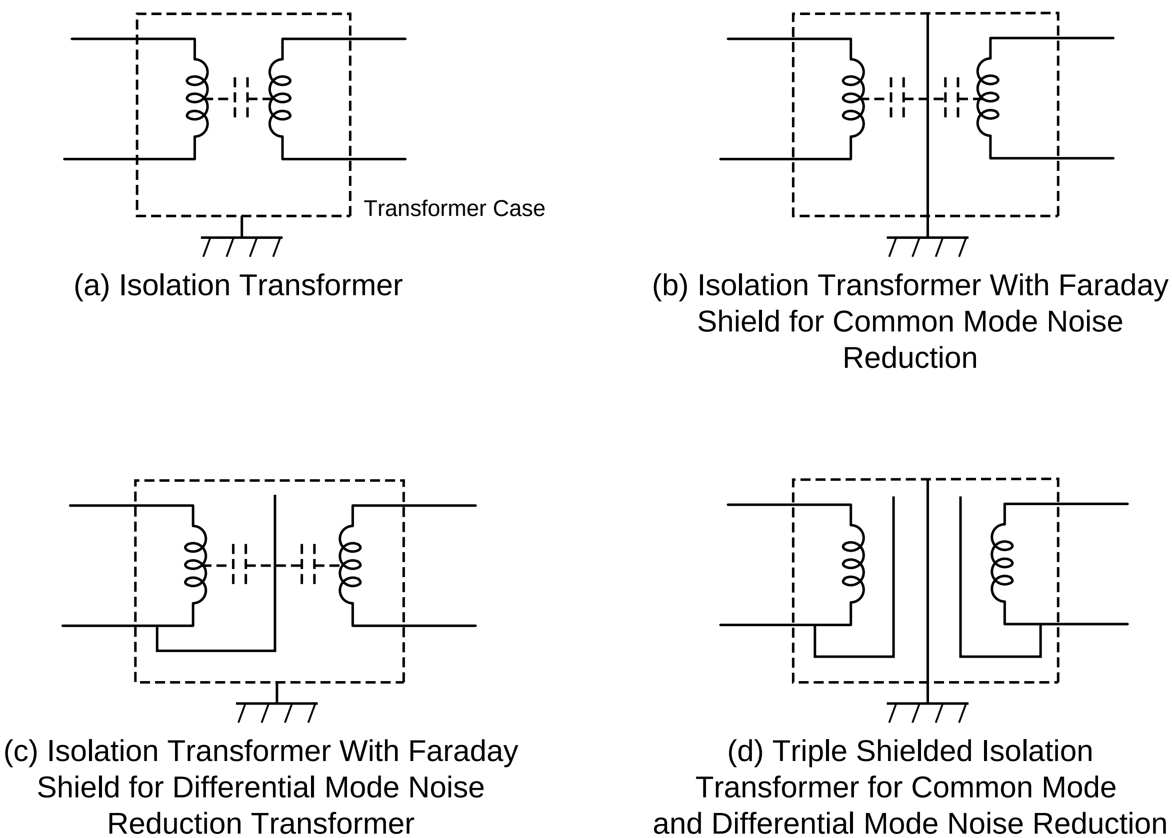


Figure 3-21. Isolation transformer configurations.

Opto-isolators are another method of isolating signals to attenuate conducted EMI. Figure 3-22 shows a schematic of an opto-isolator. Opto-isolators perform over a wide bandwidth (approximately 50 MHz) and work with both logic and analog signals above 100 mV. The limiting factor in high-frequency usefulness of opto-isolators is their input-to-output capacitance (typically 0.1 to 10 pF). This capacitance allows high-frequency noise to bypass the high impedance of the opto-isolator.

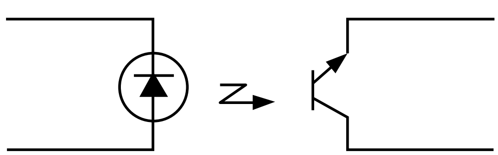


Figure 3-22. Opto-isolator schematic.

3.4 Suppression By Enclosures

EMI suppression by enclosures is another term for EMI shielding. According to reference 3-15, “A shield is a metallic partition placed between two regions of space. It is used to control the propagation of electric and magnetic fields from one of the regions to the other.” This is a simple but effective definition of shielding. A shield is used to divert or absorb unwanted electromagnetic energy. The following subsections offer general information and guidelines for shielding against EMI.

3.4.1 Enclosure Shielding

Most books on shielding delve into a comprehensive coverage of shielding theory that is beyond the scope of this handbook. Guidelines provided here provide a minimum of mathematics and theory.

Shielding of EM fields is accomplished through reflectance or absorption of the fields by a barrier. In most applications, the barrier is a metal, although coated and conductive plastics are being used more frequently in commercial applications. An important point to remember in shielding is that the actual shielding provided by a metal barrier depends on the type electromagnetic field that predominates. Reflection is highly effective against predominately electric fields and plane waves but has little effect on predominately magnetic fields. Absorption is the mechanism in predominately magnetic field attenuation.[3-16] Reflectance increases with surface conductivity of the shield but decreases with frequency.[3-16] Absorption increases with:

- Thickness of the shield
- Conductivity of the shield
- Permeability of the shield
- Frequency of the incident field.[3-6]

Absorption in a metal barrier is exponential in nature, i.e., as an electromagnetic field passes through a metal barrier, the amplitude of the electromagnetic field decays exponentially. At some distance into the metal barrier, the amplitude of the impinging electromagnetic field has decreased to 1/e or 33 percent of the amplitude at the surface of the barrier. The distance at which this occurs is called the skin depth of the metal. The formula for skin depth is given in equation (3-11):

$$\delta = \frac{2.6}{\sqrt{\mu_r\ \sigma_r\ f_{\mathrm{MHz}}}} \quad \text{in mils} \ , \tag{3-11}$$

where μ_r is the permeability of the metal relative to copper, σ_r is the conductivity of the metal relative to copper, and f_{MHz} is the frequency of the electromagnetic field impinging on the metal. The skin depth concept is shown in figure 3-23. Table 3-5 lists plane wave skin depths for copper and aluminum at various frequencies.

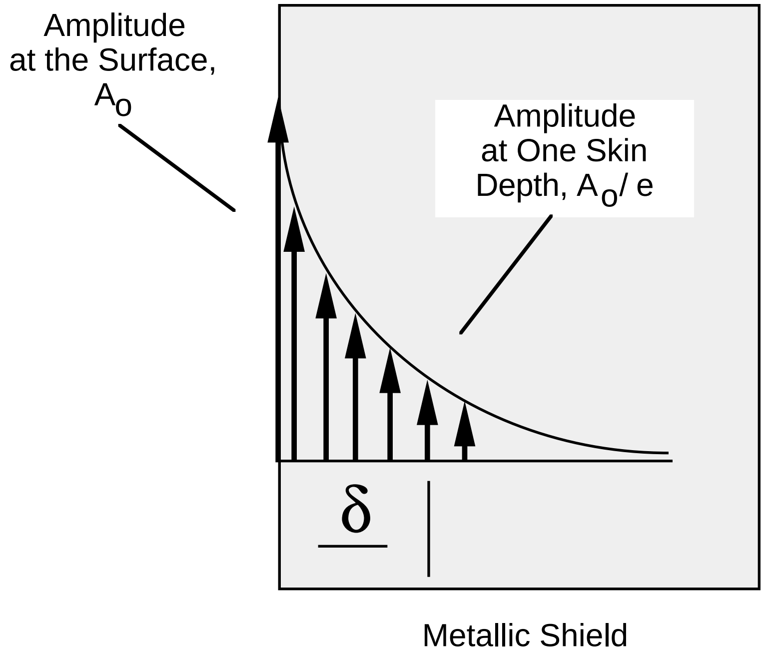


Figure 3-23. Schematic definition of skin depth.

Table 3-5. Skin depths at various frequencies.

Frequency	δ for Copper (mils)	δ for Aluminum (mils)
10 kHz	26	33
100 kHz	8	11
1 MHz	2.6	3
10 MHz	0.8	1
100 MHz	0.26	0.3

The performance of a shield in reducing the electromagnetic energy that passes through it is known as its shielding effectiveness. Equation (3-12) defines the shielding effectiveness (in decibels) for electric fields and magnetic fields:

$$SE_{dB} = 20 \log_{10} \{\frac{E_{in}}{E_{out}}\} \text{ for electric fields }, \tag{3-12a}$$

$$SE_{dB} = 20 \log_{10} \{\frac{H_{in}}{H_{out}}\} \text{ for magnetic fields }, \tag{3-12b}$$

where E_{in} (H_{in}) is the field strength incident on the shield, and E_{out} (H_{out}) is the field strength after passing through the shield.[3-6] Shielding effectiveness is shown in figure 3-24. Note: At one skin depth the SE_{dB} of a metal is at least 8.7 dB and at 2.3 skin depths the SE_{dB} is at least 20 dB.

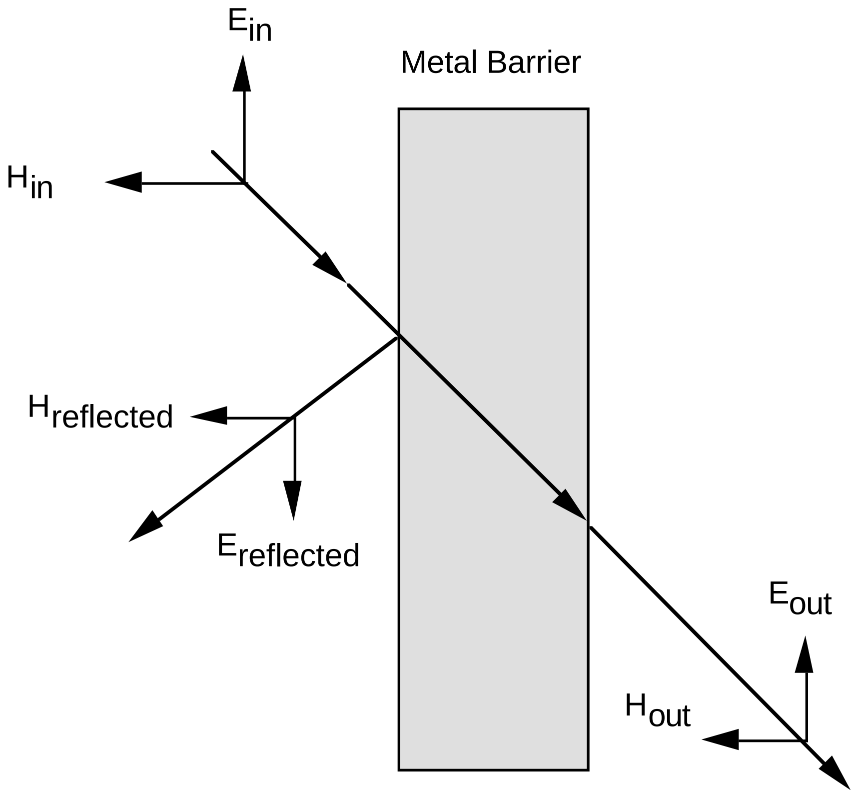


Figure 3-24. Schematic of shielding effectiveness.

The above discussion assumes that the barrier or shielding material is homogeneous and large such that there is no leakage or edge effects. The shielding effectiveness expressed in equation (3-12) is degraded by apertures for connectors, switches, and I/O lines and seams for doors, access panels, and cover plates. These apertures and seams serve as leakage paths for electromagnetic energy; this leakage lowers the SE_{dB} of the barrier. Minimization of these leakage paths is addressed in subsections 3.4.2 and 3.4.3.

Finally, a few shielding rules of thumb:[3-6, 3-17]

- For a predominately electric field or plane wave, use a good conductor (copper or aluminum) to maximize reflection loss.

- For a high frequency magnetic field (frequency >500 kHz), use either a good conductor or a material with a high permeability, μ_r.

- For a low frequency magnetic field (10 kHz > frequency > 500 kHz), use a magnetic material such as steel, for frequency <10 kHz use a material with a high permeability, μ_r, to maximize absorption loss.

- Reflection loss varies with the type of field; absorption loss is independent of the field.

- A metallic shielding material thick enough to support itself usually provides good electric field shielding at all frequencies.

3.4.2 Shield Discontinuities

As stated in section 3.4.1, the shielding effectiveness of an enclosure is degraded by the introduction of discontinuities into the enclosure. These shield discontinuities are the holes, seams, and joints found in nearly all electrical and electronic equipment. Leakage through seams, holes, and joints is usually a greater concern than the shielding effectiveness of the shield material. The methods presented here are equally adequate for minimizing magnetic and electric field leakage; only the types of shielding materials differ.

Discontinuity rules of thumb include:

- The amount of leakage from a discontinuity depends on the maximum linear dimension of the opening and the frequency of the source.

- A slot or rectangular hole may act as a slot antenna when the maximum linear dimension of the slot becomes greater than 1/10 of a wavelength.

- A large number of holes allows less leakage than one large hole of the same total area (fig. 3-25).

- A hole shaped to form a waveguide (the depth of the hole is greater than the diameter of the hole) can offer greater attenuation than a “regular” hole pattern for frequencies lower than the waveguide’s critical frequency. This critical frequency is roughly the frequency at which the maximum linear dimension of the opening of the waveguide equals $\lambda/2$. Below this critical frequency, the waveguide attenuation is dependent on the length of the waveguide and is called waveguide below cutoff. The waveguide below cutoff concept is shown in figure 3-26.

- For seams and joints it is necessary to maintain a continuous metal-to-metal contact along the seam or joint to ensure shielding integrity.

- The preferred seam for preventing EMI leakage is a continuous weld.[3-18] Figure 3-27 shows types of seams in ascending order of preference.

- When bolts or rivets are used to make a bond, the shielding effectiveness depends on the number of rivets or screws per linear inch, the mating pressure at the contact surface, and the cleanliness of the two mating surfaces.

- The higher the number of rivets or screws per linear inch, the greater the shielding effectiveness.[3-18]

- For equipment enclosures that require ventilation, the following materials (in descending order of attenuation) should be used to cover the opening:[3-19]

    Waveguide below cutoff panels (honey-comb panels)

    Perforated metal sheet

    Woven or knitted metal mesh.

Allowing an unfiltered, unshielded cable to leave a shielded enclosure, such as an unfiltered power line, effectively negates all work done to shield the enclosure. The noise on the unfiltered, unshielded cable radiates and the shielding of the enclosure is bypassed. References 3-18 and 3-19 are good sources for more detailed information on shielding.

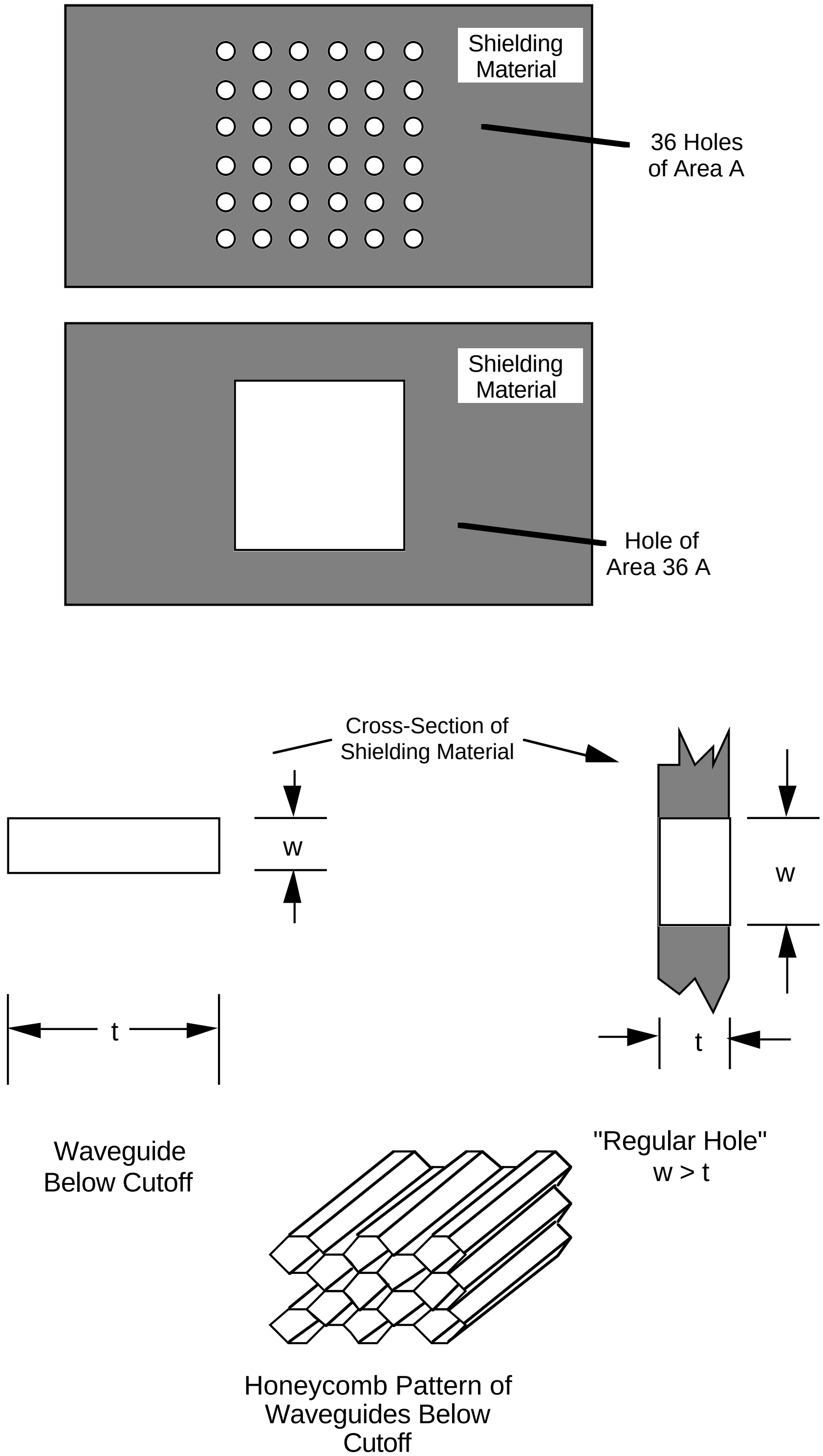


Figure 3-26. Waveguide below cutoff.

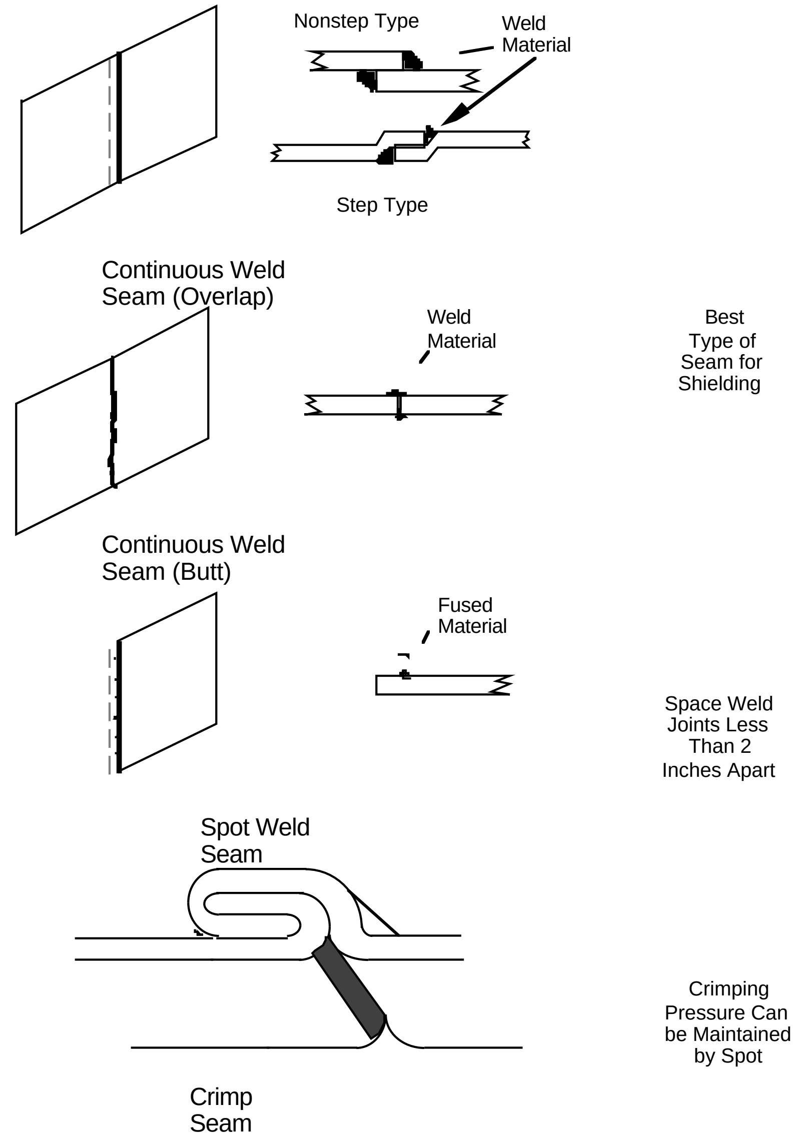


Figure 3-27. Types of seams.

3.4.3 Gaskets

In joints difficult to maintain continuous metal-to-metal contact, i.e. access panels, lids, and hinges, conductive gaskets are used to provide EMI shielding (figs. 3-28(a) and 3-38(b)).

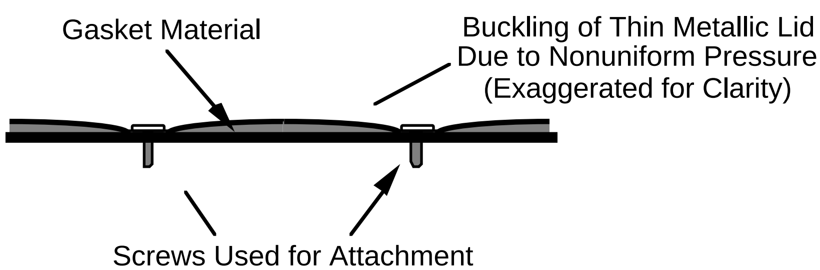


Figure 3-28(a). Example of EMI gasket.

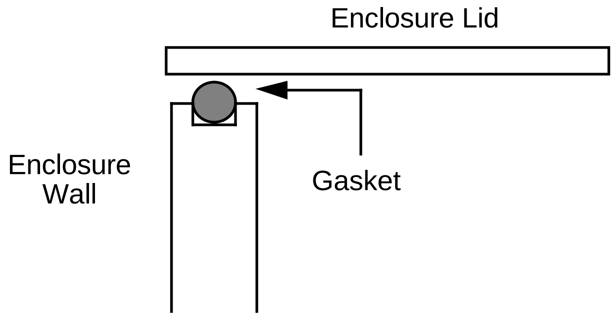


Figure 3-28(b). Use of EMI gasket.

Gasket materials include metallic textile gaskets and knitted wire mesh. Table 3-6 lists several gasket materials and the chief advantages and disadvantages of each. A rule of thumb for conductive gaskets is: the greater the compressibility, the greater the sealability. The gasket must be able to conform to the irregularities of the two mating surfaces under the applied force. However, the contact pressure must be great enough for the gasket to make adequate metal-to-metal contact, even in the presence of nonconductive film on the mating surfaces. Figure 3-29 shows examples of good metal-to-metal contact using EMI gaskets. Figure 3-30 shows examples of uses for conductive gaskets. It is important to remember that contact surfaces must be clean of paint and oil.

Table 3-6. Conductive gasket materials.[3-19]

Material	Chief Advantage	Chief Disadvantage
Compressed knitted wire	Most resilient of all-metal gasket	Certain shapes difficult to make
Beryllium copper gasket	Best break-through on corrosion films	Not truly resilient; Not generally reusable
Imbedded wire gasket	Combines fluid and conductive seals	Requires 0.25-in thickness and 0.5-in width for optimal shielding
Aluminum screen impregnated with neoprene	Thinnest gasket. Combines fluid and conductive seals Can be cut into intricate shapes	Very low resiliency
Soft Metals	Cheapest in small sizes	Cold flows, low resiliency
Metal over rubber	Takes advantage of resilience of rubber	Poor RF properties
Conductive elastomer	Combines fluid and conductive seals	Relatively high cost
Contact gingers (finger stock)	Best suited for sliding contact	Easily damaged
Convoluted Spiral	Can provide conduction at forces as low as 1 psi	Not available in sheet form

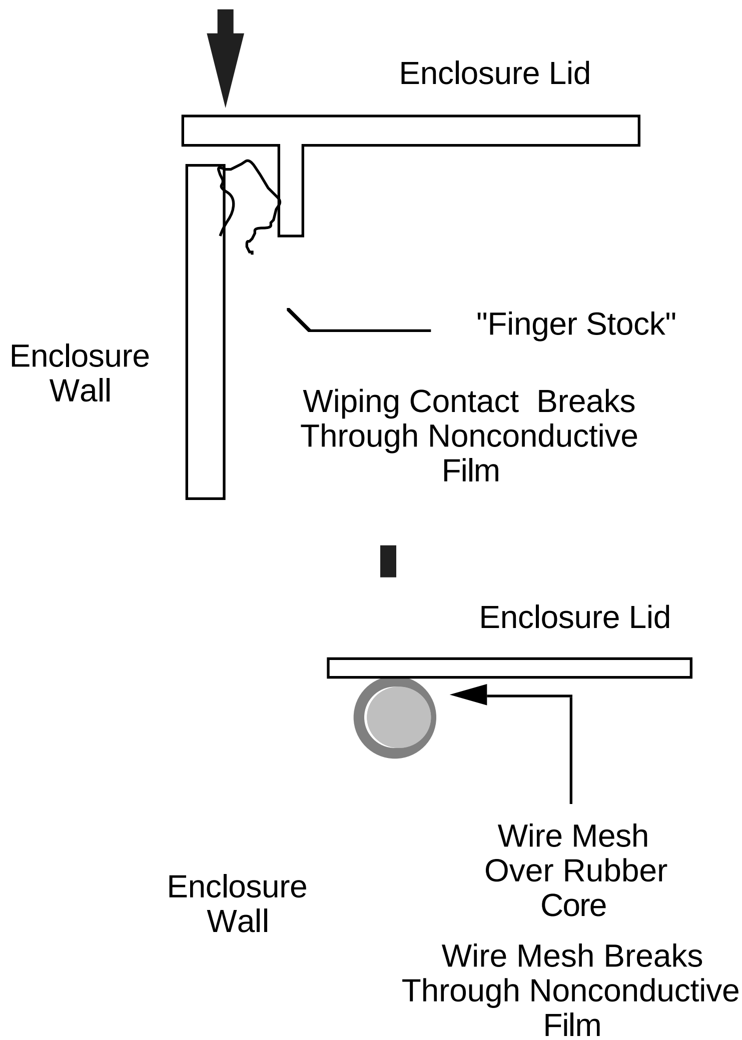
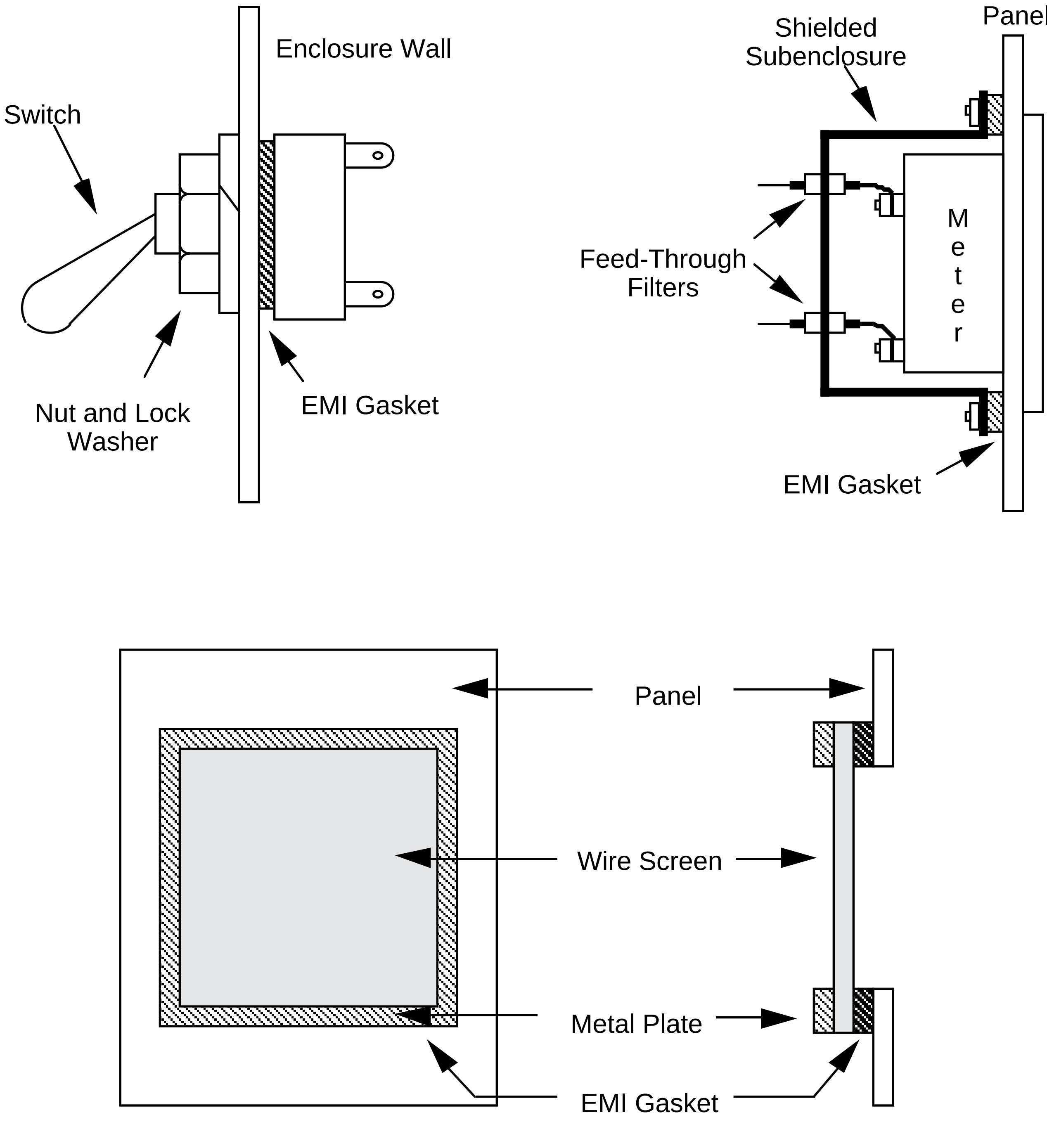


Figure 3-29. Examples of good metal-to-metal contact using EMI gaskets.



Wire Screen Over Opening in Panel

Cross Section of Wire Screen Over Opening in Panel

Figure 3-30. Examples of uses for conductive gaskets.

3.4.4 Cable Shielding

One primary source of RE is unshielded or improperly shielded cables. There are four common types of shielding: braid, flexible conduit, rigid conduit, and spirally wound sheets of high permeability material. Of these four, braid is relatively light weight and easiest to handle. It is important to note that the shielding effectiveness of a cable shield depends on the characteristics of the shield material and the manner in which the shields are terminated.[3-13]

When terminating a shield, it is important that the termination provide a low impedance path for noise currents. Shield terminations fall into two categories: pigtail termination and 360° shield termination (sometimes referred to as RF backshell termination). The 360° shield termination provides a low impedance path and preserves shielding integrity of the enclosure or connector to which the shield is terminated. This type shield termination is much preferred. A pigtail termination is the least preferred method of shield termination because, at RF frequencies, the inductance of the pigtail becomes such that the shielding effectiveness of the cable shield is negated. If, however, pigtail termination is unavoidable, keep the pigtail as short as possible. Figure 3-31 shows examples of pigtail termination and RF backshell termination.

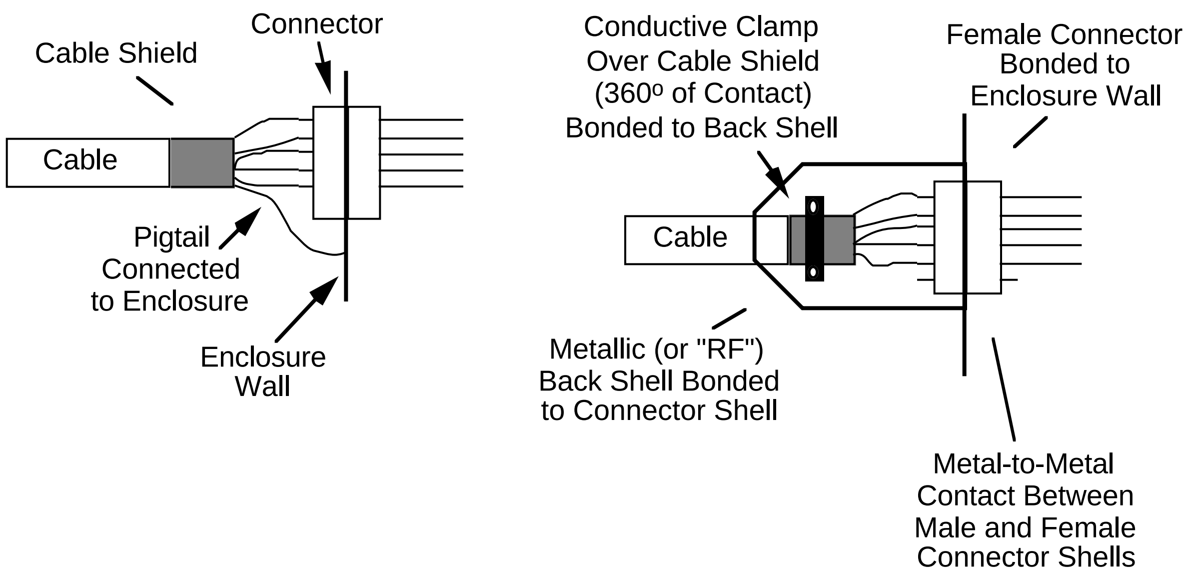


Figure 3-31. Pigtail and RF backshell terminations.

Figure 3-32 shows the preferred methods of shield termination in descending order of preference. Specific requirements on cable shielding and shield termination are found in NASA Handbook NHB 5300.4(3G).[3-24] As a general rule, the cable shield should be grounded at both ends. Also, cable shields should never intentionally carry current. The exception to this rule is coax cable, in which the outer shield serves as the return conductor. Coax should be used only for signals where the lowest signal component is above approximately 100 kHz.

In some applications, double shielding of cables is required to prevent unwanted electromagnetic energy from entering the circuit. Figure 3-33 shows example schematics of how to ground double shielded cables. For some low-frequency, high-load-impedance circuits, grounding the shield at both ends causes low-frequency noise currents on the shield to couple into the circuit. Figure 3-34 is an example of a possible solution for this problem.[3-30]

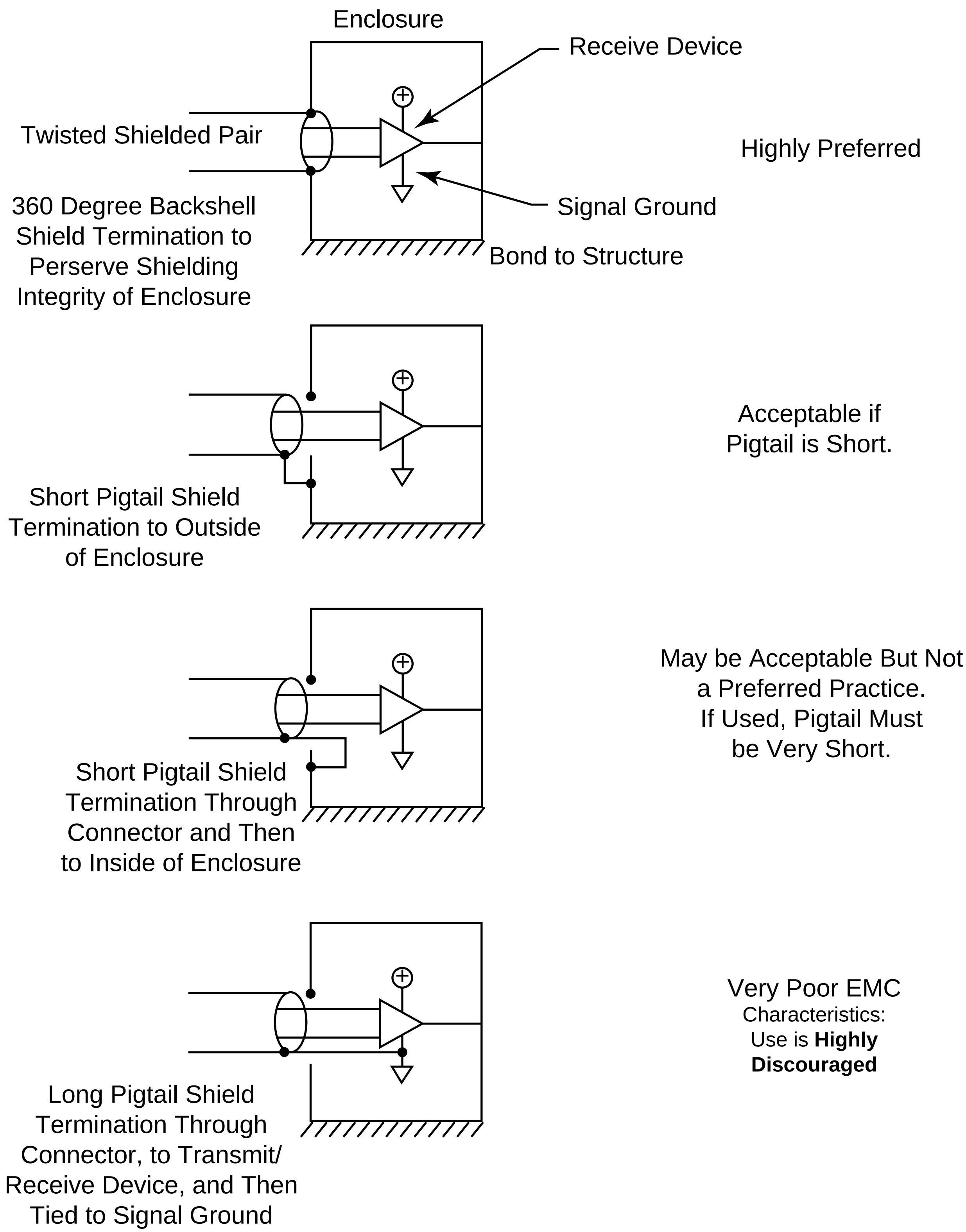


Figure 3-32. Shield termination preferences.

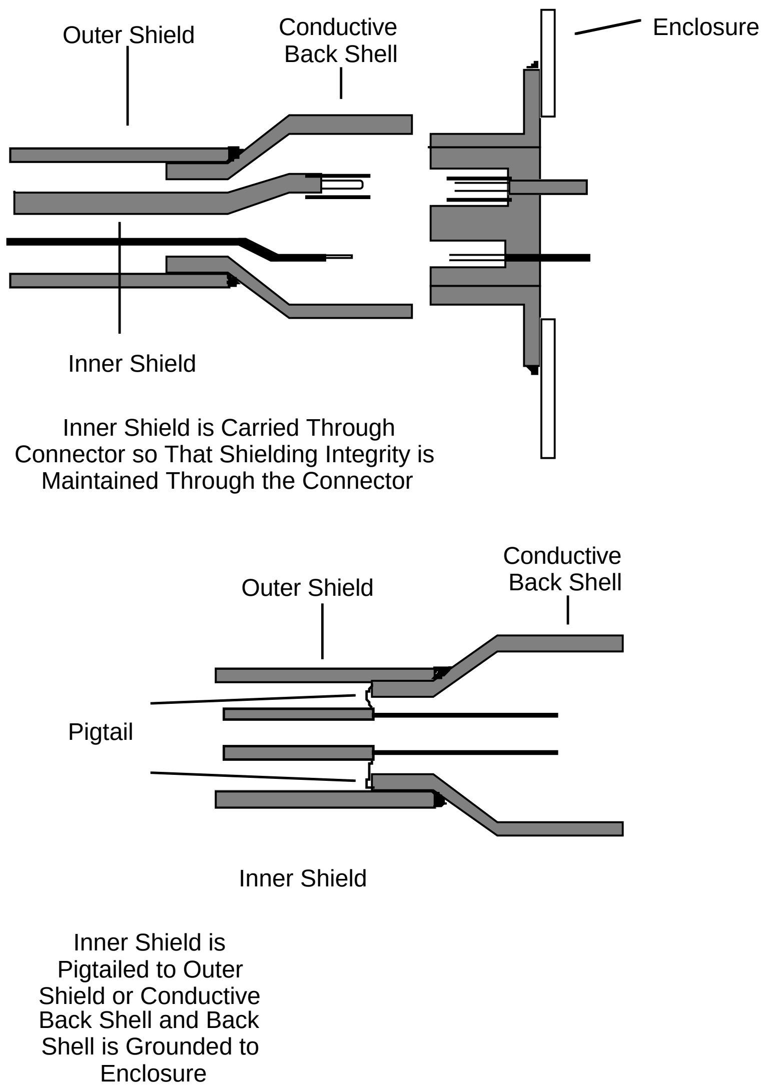


Figure 3-33. Termination of double-shielded cables.

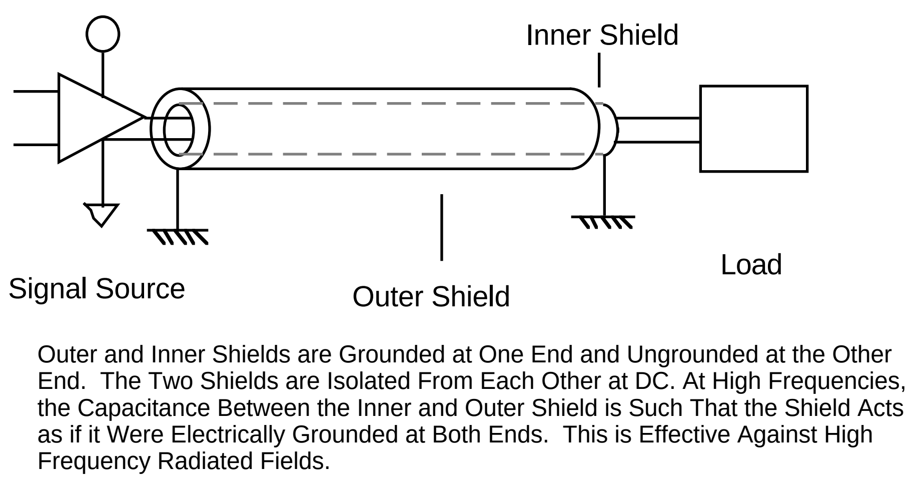


Figure 3-34. Shielding for low-frequency, high-impedance circuits.

3.4.5 Cable and Wiring Classes

Use of power and signal cables is prevalent on all spacecraft and payloads. These cables may act as both transmitting and receiving antennas for radiated EMI and conduits for conducted EMI. Because cables are usually routed to accommodate practical routing paths and equipment location, it is almost impossible to predict and quantify the EMI environment associated with these cables. One way of controlling EMI from cables and wiring is to separate cables and wiring into similar classes of voltage, frequency, and susceptibility levels.

The NASA space shuttle, International Space Station *Alpha*, and U.S. military specifications have requirements or guidelines for wiring classification and separation. For example, the U.S. Air Force Systems Command Design Handbook 1-4 suggests the classification of wiring based on type of electrical power (ac or dc) and frequency susceptibility.[3-18] Also, as a design goal, Design Handbook 1-4 suggests a minimum separation of 2 in (51 mm) between different wire classifications to prevent cable-to-cable coupling. NASA specifications for the Spacelab payloads and space station program (MSFC-SPEC-521B, Electromagnetic Compatibility Requirements on Payload Equipment and Subsystems) and (SSP 30242, Space Station Cable/Wire Design and Control Requirements for Electromagnetic Compatibility) contain requirements for cable classifications and separation.[3-20, 3-21] For programs in which such requirements are not supplied, table 3-7 is a guide. The cables and cable bundles of different classification should be separated by a minimum of 2 in. Figure 3-35 gives examples of the wire types called for in table 3-7.

Table 3-7. Suggested cable classifications.[3-20, 3-21]

Signal Type; Rise, Fall Time (t_r, t_f)	Voltage or Sensitivity Level	Wire Type	Circuit Class
Power (ac, dc)	>6 V	Twisted	Class I
Analog Signals $t_r, t_f > 10\ \mu s$	<6 V	Twisted Shielded	Class II
Analog Signals $t_r, t_f > 10\ \mu s$	≤100 mV	Twisted Double Shielded	Class III
Analog Signals $t_r, t_f < 10\ \mu s$	≤100 mV	Twisted Double Shielded	Class IV
	≥100 mV	Twisted Shielded	
Analog Signals $f > 100$ kHz, Digital Signals	All	Twisted Shielded, Coax	Class IV

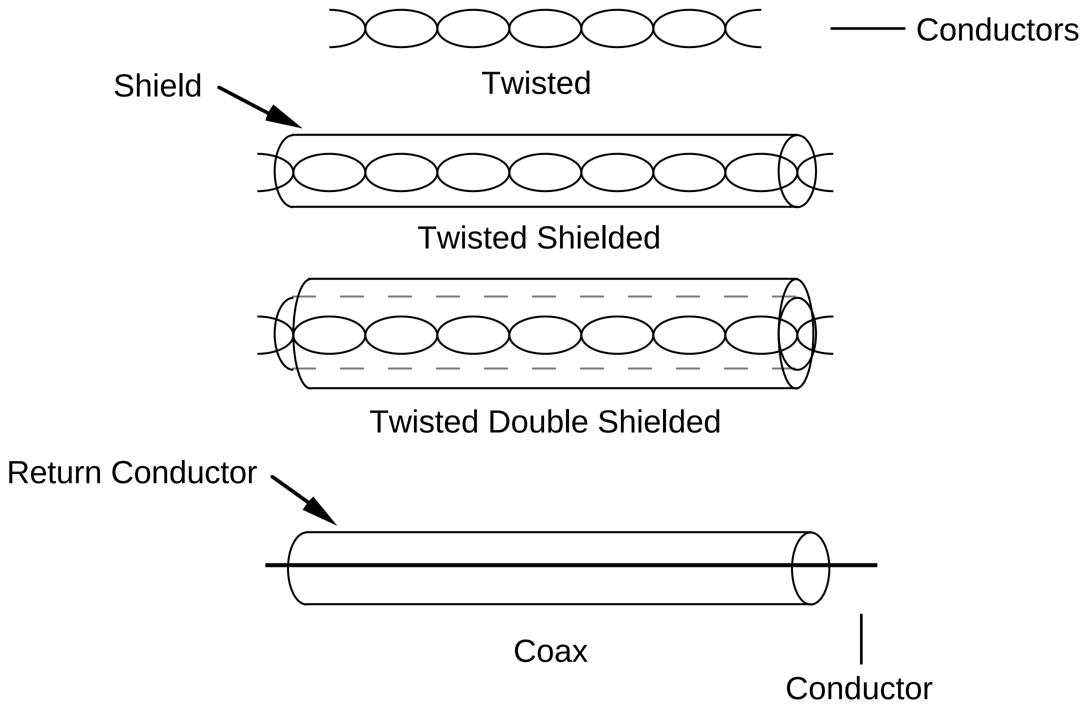


Figure 3-35. Wire types.

Twisting of wire minimizes the loop area of the wire. This minimizes the amount of inductive noise coupling between the circuit and surrounding cabling. The number of twists per foot of cabling is limited by cable size, however, the greater the number of twists per foot, the smaller the loop area of the wire.

3.5 Switched-Mode Power Supplies

Switched-mode power supplies are characterized by high output power per unit volume, low weight, and high efficiency and are ideal for use on NASA experiments and space platforms. If these power supplies, found on almost all NASA satellites and space shuttle experiments and payloads, are improperly designed, they may be a source of EMI that degrades other systems.[3-9]

3.5.1 Power Supply Topologies

The controlling active device in switched-mode power conversion is a switch that is either open or closed. The output voltage is regulated by controlling the ratio between the time interval that the switch is closed and the time that the switch is open (defined as duty ratio or duty cycle). The capacitive and inductive components are added to smooth out the pulsating behavior of the switching transitions. The switching frequencies of these power supplies range from 10’s to 100’s of kHz and higher.

Three basic topologies of switched-mode power conversion most commonly used today are the buck, boost, and buck-boost. The most commonly used converter topology that provides input to output isolation is the push-pull converter. A description of these topologies and potentials as EMI sources is briefly presented in the next four subsections.

3.5.1.1 Buck Converter

The buck converter chops the input voltage and the LC output filter smoothes the output voltage. A schematic of the basic buck converter is shown in figure 3-36. The output voltage, V_o, ideally, is equal to the input voltage, V_i, times the duty cycle, D (equation (3-13)). Because of the inductor in the output side of the converter, the output current is continuous, never falling to zero. But because of the switching transistor being in the input line, the input current is pulsating. This pulsating characteristic of the buck converter is an undesirable side effect (a potential source of conducted EMI). Therefore, the buck converter design necessitates a highly attenuative EMI filter on the input of the converter in order to meet conducted EMI requirements.

$$V_o = V_i D \quad . \tag{3-13}$$

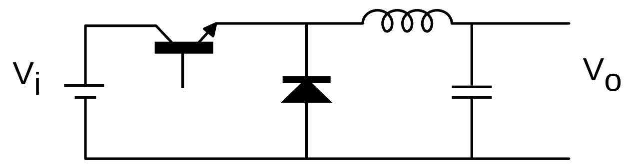


Figure 3-36. Buck converter topology.

3.5.1.2 Boost Converter

As implied by the name, the boost converter performs a step-up voltage conversion. The boost converter topology is shown in figure 3-37. The boost converter is the dual of the buck converter, which performs a step-down voltage conversion. The output voltage, V_o, ideally, is equal to the input voltage, V_i, times the inverse of one minus the duty cycle, D (equation (3-14)). This converter design results in less noise generated at the converter input but more at the output of the converter, i.e., the opposite of what is found in the buck converter.

$$V_o = \frac{V_i}{1\text{-}D} \quad . \tag{3-14}$$

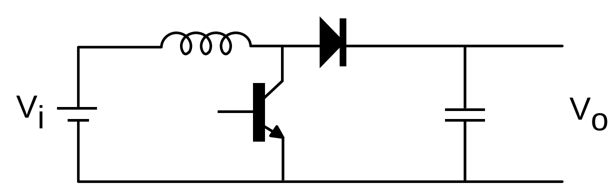


Figure 3-37. Boost converter topology.

3.5.1.3 Buck-Boost Converter

The buck-boost converter is a voltage inverting structure. The buck-boost converter topology is shown in figure 3-38. Its conversion function is a product of the buck and boost gains. Equation (3-15) is the gain equation for an ideal buck-boost converter:

$$V_o = V_i \frac{-D}{1-D} \quad . \tag{3-15}$$

The switch action of the transistor commutates the continuous inductor current alternately between input and output ports. Because both input and output currents are pulsating, the buck-boost converter from an EMI viewpoint is the noisiest type converter.

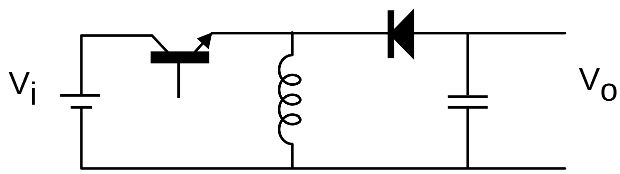


Figure 3-38. Buck-boost converter topology.

3.5.1.4 Push-Pull Converter

The push-pull converter uses two switching transistors to do the power switching. This type converter may be used in situations in which higher power is needed because the configuration of two switches and two diodes allows the average current in each switch-diode set to be reduced by 50 percent from the single switch approach. The transformer provides a dc isolation from input to output to prevent violation of single-point or single-reference ground requirement when the converter load requires a local ground to operate properly. This converter type is derived from the buck converter and, like the buck converter, requires a highly attenuative EMI filter on the converter input. The basic topology is shown in figure 3-39.

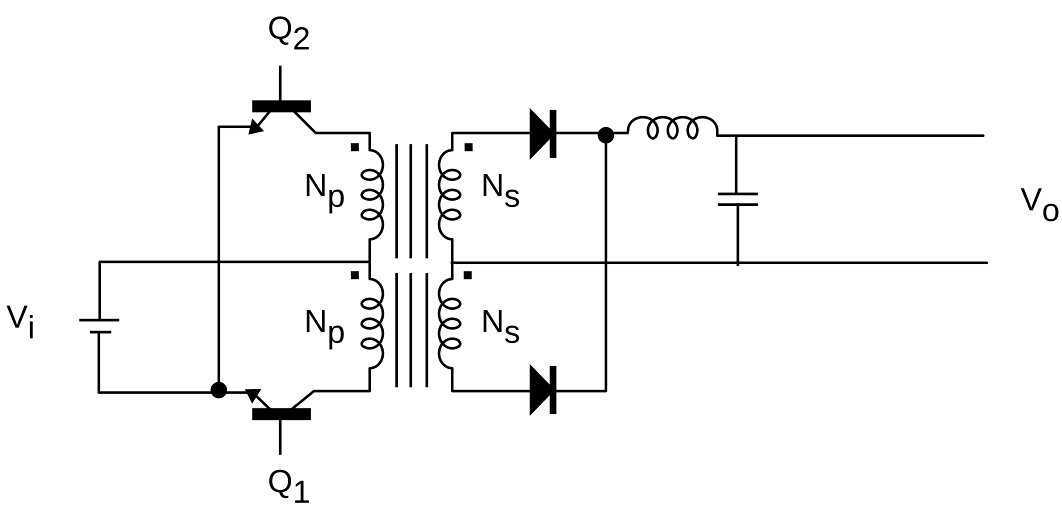


Figure 3-39. Push-pull converter topology.

3.5.2 Electromagnetic Interference From Switching-Mode Conversion

Switch-mode power supplies are potential generators of EMI due to the switching action of the converter. For converters that convert ac to dc, a rectifier is added to convert to dc. This rectification is an additional source of noise. The switching action generates a spectrum of the switching frequency and its harmonics. The main noise sources of switching frequency harmonics are the switched currents and the commutating diode. This noise is a combination of the switching frequency and its associated rise time (approximately 100 ns) and turn on spikes caused by the diode recovery current. This recovery current spike occurs at the end of a diode conduction cycle when reverse voltage is just applied by the transistor.[3-9] It is the combined noise of the transistor and diode that must be filtered in order to meet conducted EMI requirements. The input filter must filter both the common-mode and differential-mode noise generated. Figure 3-40 shows a frequency spectrum envelope of this combined switching and diode recovery noise.

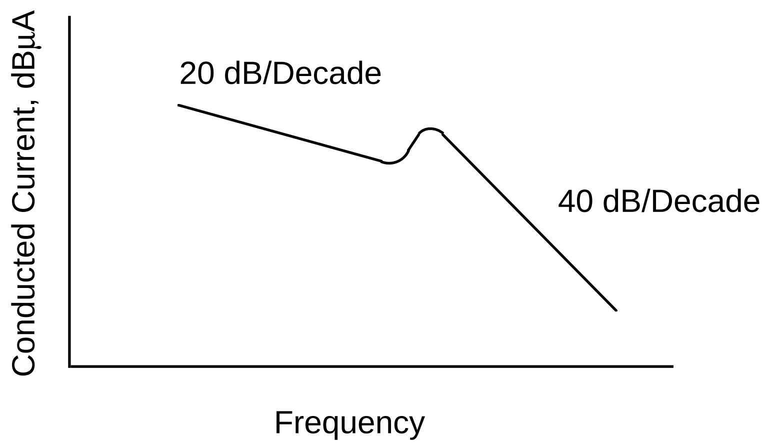


Figure 3-40. Frequency spectrum envelope of switching and diode recovery noise.

MIL-HDBK-241B, “Design Guide for Electromagnetic Interference Reduction in Power Supplies,” is a military handbook by the Department of Defense to help the electrical designer in designing low EMI power supplies.[3-8] It provides basic and fundamental information on EMI reduction as well as information on ensuring power supply stability, component selection, and filter design. This handbook is good source for additional detailed information.

3.6 Grounding

Grounding is defined as referencing an electrical circuit or circuits to Earth or a common reference plane for preventing shock hazards and/or for enhancing operability of the circuit and EMI control. Bonding is defined as the process by which a low impedance path is established for grounding or shielding purposes.[3-25] Because the terms “grounding” and “bonding” are often used interchangeably, it leads to confusion. In this section, only the grounding of electrical circuits, not the grounding of metallic components such as electrical equipment cases, cabling conduit, pipes, and hoses (sometimes referred to as bonding), is addressed. Bonding requirements for most NASA programs are based on MIL-B-5087B, “Bonding, Electrical, and Lightning Protection, for Aerospace Systems.”[3-26] References 3-18 and 3-19 offer good explanations on bonding and bonding concepts.

3.6.1 Grounding Systems

An electrical system is grounded for three reasons: safety, enhanced operability of the circuit, and EMI control. Grounding an electrical power circuit provides a current return path during an electrical fault. This allows the fuse or circuit breaker to operate properly and prevents shock hazards to personnel. This is accomplished by ensuring that the fault current path has an impedance that is small and an ampacity (current carrying capacity) high enough to allow the circuit breaker or other protection device to operate. Additionally, the voltage generated by the fault current between the equipment case and ground must be low enough to meet safety requirements. Voltage generated due to the fault is:

$$V_{fault} = I_{fault} * R_{bond} \quad , \qquad (3\text{-}16)$$

where I_{fault} is the fault current and R_{bond} is the resistance of the equipment ground connection. This resistance includes the resistance of each electrical bond in the ground connection and the resistance of the grounding strap or jumper used in the ground connection. I_{fault} is the maximum amount of current that the electrical power system can source.[3-27]

Some electrical circuits require grounding to a common reference plane ("ground" plane) in order to operate efficiently. Grounding of filter components and other EMI control measures increases EMI suppression. The line-to-ground or feed-through capacitors used to suppress CM noise must have a low impedance path to the source of the CM noise. In order to shunt the CM currents from line to equipment enclosure (preventing noise from escaping onto power lines), the resistance and the reactance of the bonds in the path between noise source and line-to-ground capacitor must be sufficiently low over the bandwidth at which the line-to-ground capacitors operate. It is important to remember that grounding is not a "cure-all" for EMI and improper grounding may aggravate noise problems.[3-19. 3-28]

In regard to EMI control, the objectives of a good grounding scheme are to minimize noise voltages from noise currents flowing through a common impedance and to avoid ground loops.[3-15] These objectives are realized at two levels: (1) platform (vehicle or spacecraft) grounding level and (2) equipment internal grounding level.

A number of system grounding philosophies exists. The basic grounding schemes are the following:

(1) Single point star (star)

(2) Multipoint

(3) Floating ground

(4) Layered single point (single point or single reference).[3-29]

Figures 3-41 to 3-43 are schematics of grounding concepts. The single point star (1) and single reference ground (4) are the most commonly used grounding concepts for NASA projects. The aim of the single point and single reference ground is to reduce low frequency and dc current flow in the ground plane. Adding to the grounding confusion is the fact that the term "single point" may be used to refer to a single point star or a layered single point ground. For consistency, a single point star ground is referred to as a star ground and layered single point ground is referred to as a single point ground. Additional information on grounding schemes is found in references 3-25, 3-28, and 3-29. It is important to remember that one type of ground scheme can be utilized for power signals, another for RF signals, and yet another for analog signals and cable shields. It is important to utilize the various concepts as needed to meet the requirements of safety, enhanced operability, and EMI control.

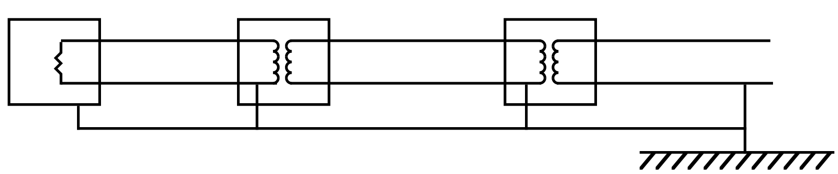


Figure 3-41. Single point star ground.

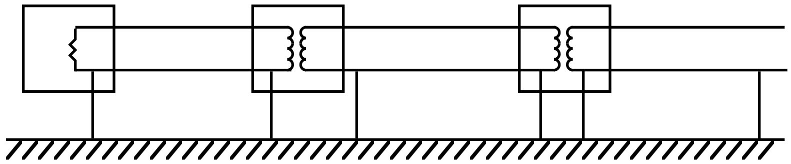


Figure 3-42. Multipoint ground.

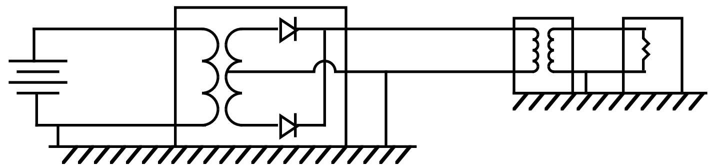


Figure 3-43. Layered single point ground.

3.6.2 Platform Grounding

Platform grounding pertains to the grounding of whole systems and subsystems distributed among many electrical boxes. The platform is a satellite, a launch vehicle, or a payload carrier (e.g., SpaceLab). The structure of the platform is used as the ground plane with few exceptions if the platform is metallic. For platforms composed mostly of nonmetallic materials, the ground plane is a metallic area to which all ground references are attached.

3.6.2.1 Single Point Star Ground (Star)

The star ground scheme (fig. 3-41) is used on platforms composed mostly of nonmetallic materials or metallic platforms with large amounts of noise currents flowing in the ground plane (e.g., platforms in which the structure is the return for the electrical power distribution system). Each isolated electrical system is referenced once to the ground plane at a single point. The major weakness of this grounding system is that the wiring used to make the ground reference connection has a higher reactance than resistance above a few kilohertz. Any noise currents flowing in the ground reference connection develop a voltage between the electrical circuit and the ground plane. The long ground connection can also act as an antenna. RE can couple to this “antenna” and cause the equipment to have some voltage relative to the ground plane, possibly interfering with the correct operation of the equipment.

3.6.2.2 Single Point Ground (Single Reference)

The single reference ground scheme (fig. 3-43) is a derivative of the star ground. Each isolated electrical system is referenced once to the ground plane. In most cases, the ground plane is the vehicle or payload carrier structure. The short jumpers used to reference to ground locally and the metallic structure between the grounding points (if good bonding practices are implemented) have a lower impedance than a wire or cable used to reference the isolated systems in a star ground. This lowers noise voltages caused by noise currents flowing in the ground system.

3.6.2.3 Ground Loop Isolation

It is important to maintain isolation to avoid single point ground violations. These violations result in ground loops that radiate noise or pick up noise from outside sources. In an electrical power distribution system, a switched-mode power supply with transformer isolation is used to prevent ground loops. The power supply output is referenced to ground and any loads powered by the supply are isolated from structure. In figure 3-43, a power supply in one box provides electrical power to a second box. The input of the second box is isolated from ground. Signals sent between boxes can be isolated in a number of various ways. The most common methods are transformer isolation, optical isolation, balanced differential circuits, and single-ended circuits with dedicated returns. Figure 3-44 shows a MIL-STD-1553B data bus between two items of equipment.[3-23] Figure 3-45 shows a control line using optical isolation.[3-15] Figure 3-46 shows a balanced differential data line between two boxes. Figure 3-47 shows single-ended circuit in which current is returned on a dedicated wire instead of the ground plane. Transformer and optical isolation was discussed in section 3.3.4.

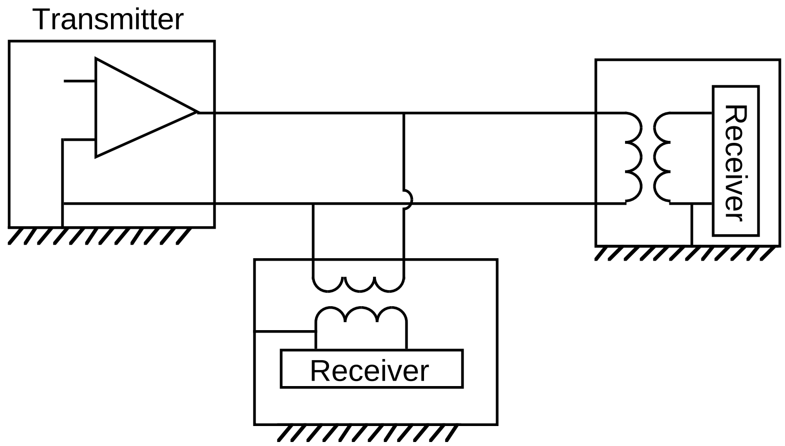


Figure 3-44. MIL-STD-1553B data bus isolation.

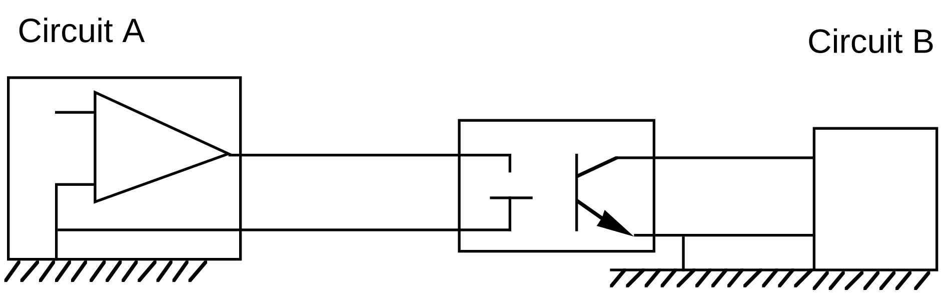


Figure 3-45. Optical isolation.

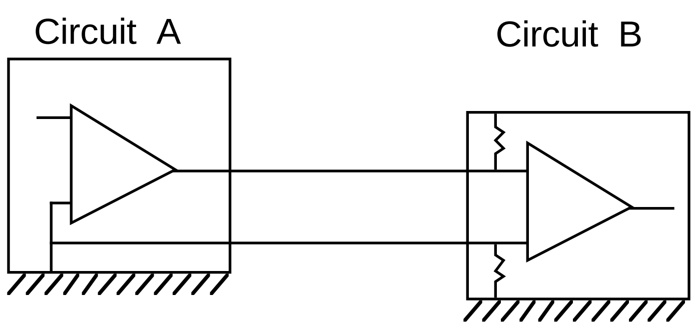


Figure 3-46. Balanced differential data lines.

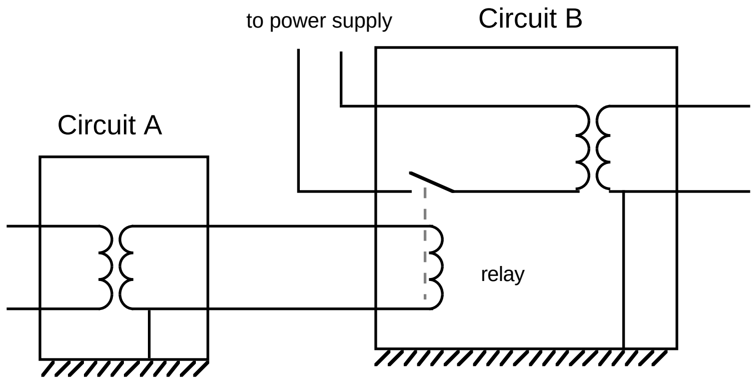


Figure 3-47. Single-ended circuit with dedicated return.

3.6.3 Equipment Internal Grounding

Grounding inside equipment is important for proper operation and reduction of EMI. Ground planes on circuit boards need to be separated in the same manner as the analog and digital circuitry. Typical systems require at least three separate ground systems: analog, digital, and noisy ground.[3-30] Figure 3-48 is an example of three separate grounds on a board. These ground systems may be divided as follows:

(1) Analog/video ground
(2) Digital ground
(3) RF ground
(4) Control ground.[3-29]

The RF signal ground may be considered the "noisy ground" from the viewpoint of the low-level analog system, yet the spectral content of the RF signal is necessary for performance of the circuit. It is important to keep these grounds separate (minimizing capacitive coupling). The connection to the equipment ground should be the only point where the separate ground systems have connection. Figure 3-48 shows the separate grounds utilizing a common connection to ground. This could be a problem in some circuitry. Each separate ground may require its own separate ground connection.

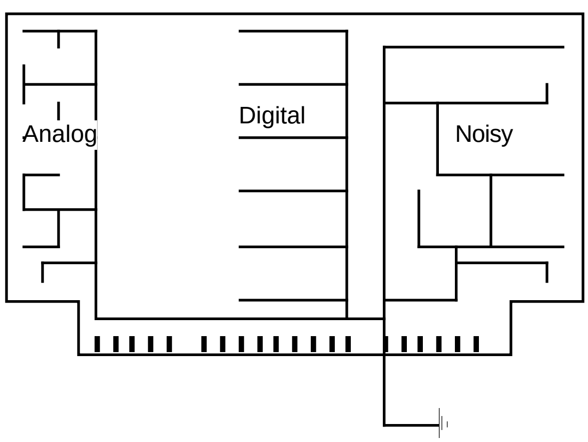


Figure 3-48. Separate ground systems.

The separation of grounds is based on two criteria: (1) signal levels and (2) spectral content. It is important to separate high level and low level returns; It is also important to separate based on frequency of the signals and frequency response of the circuits. Some circuits have an inherent filtering nature while others have no filtering response. Analog circuits such as comparators have some high-frequency filtering due to the limited slew rate of the op-amp.[3-30, 3-31] Digital circuits, by comparison, have very wide band inputs and therefore very little filtering characteristics.

Prevention of common-impedance coupling is the reason for having separate ground systems internal to the box. Allowing high-level noise currents to return through the same conductor as a low-level analog signal creates a voltage drop across the conductor that is seen by the analog circuitry. This noise voltage may interfere with the performance of the analog circuit. Figure 3-49 is a schematic of common-impedance coupling. In this circuit, two separate loads are powered by different sources but utilize a common return. Load 2 is a noisy and/or high current circuit. Load 1 is a sensitive analog signal circuit. Z_{g2} and Z_{g1} are the dc resistances and parasitic inductances of the common conductor. Z_{g2} represents the impedance of the conductor between the two points where load 1 and load 2 are connected to the return. Z_{g1} represents the impedance of the conductor between the connection point of load 1 to the return and the connection to the sources. The voltage drop (V_2) across Z_{g2} equals I_2 times Z_{g2}. The voltage drop (V_1) across Z_{g1} equals the sum of I_1 and I_2 times Z_{g1}. The increased voltage drop due to I_2 can interfere with Load 1.

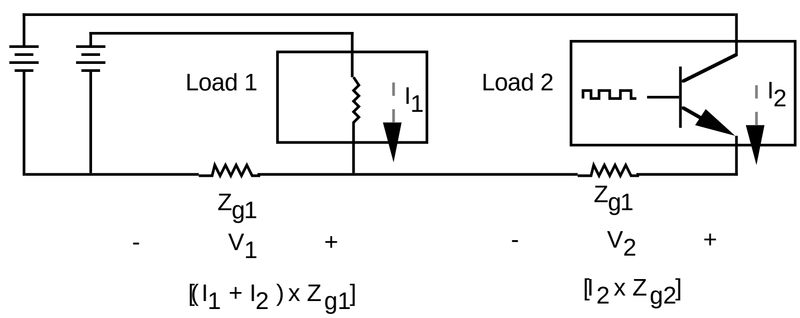


Figure 3-49. Common-impedance coupling.

The ideal way to prevent common-impedance coupling is to use separate returns for each circuit. Since this is not always possible, careful planning of the circuit layout is needed. Figure 3-50 is a schematic of a good rule of thumb to use when sharing returns. Place quiet circuits farthest from the single point ground and the noisy circuits closest to the ground connection. This limits the common-impedance coupling by limiting the impedance of the return path for the noisy circuit. The inverse of this is to place the circuits that are insensitive to common-impedance coupling farther away from the ground connection than the sensitive circuits. The closer the circuit is to the ground point, the smaller the shared impedance to cause a noise voltage.[3-32]

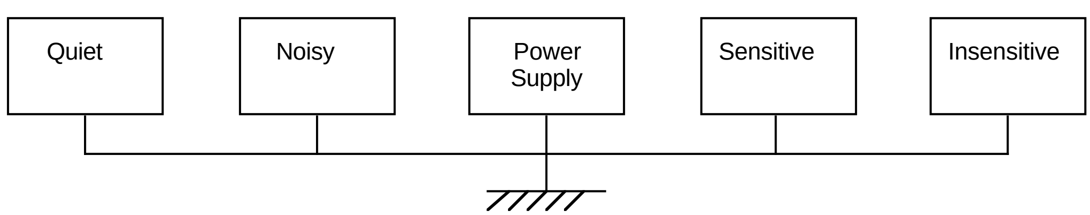


Figure 3-50. Layout rules for sharing returns.

These rules are not "set in stone," i.e., they may be modified as long as the designer considers the potential consequences of each modification and does not forget the intention of a controlled grounding concept: the elimination of common impedance coupling.

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4. ELECTROMAGNETIC COMPATIBILITY DETAILED DESIGN AND PREDICTION TECHNIQUES FOR ELECTROMAGNETIC COMPATIBILITY REQUIREMENT COMPLIANCE

4.1 Introduction

The goal of this chapter is to provide design compliance and prediction techniques specific to each major EMI test. A general description of each type of EMI test is given in chapter 2. The design compliance techniques provided are implemented before or after EMI testing. However, there are more options available if designing to meet each type of EMI test is considered early in the design process. Prediction methods using both hand calculations and computer modeling are also shown. Note: Prediction techniques are generally used to “highlight problem areas as early as possible, to aid in cost-effective design, and to support waiver request.”[4-1] Prediction techniques requiring very detailed and precise modeling are referenced but not included in this handbook.

4.2 Conducted Emissions (CE01/CE03)

CE01 (30 Hz to 20 kHz) and CE03 (20 kHz to 50 MHz) limit the noise currents that can be drawn from the input power lines.

4.2.1 Design Considerations

Adequate input power line filtering is the most effective deterrent to CE. The emissions are caused by both common mode (CM) and differential mode (DM) noise currents. Additional information on CM and DM currents is found in sections 3.3.1 and 5.1.3. Because CE measurement is done with a current probe around the individual high and low side of power, the test does not distinguish between CM and DM. General assumptions to distinguish types of noise are shown.

DM noise is generally controlled at lower frequencies (below about 2 MHz). “Above this frequency range component resonances reduce the differential filters ability to attenuate EMI.”[4-2] Differential emissions are predominately caused by noise generated from the fundamental power switching frequency and its associated harmonics. Actually “repetition rates of the signals or waveforms with fast rise and fall times are generally the most significant differential EMI threat.”[4-2]

CM noise, on the other hand, is caused by currents flowing through the ground plane and in the same direction on both the high side and return of power and signal lines. CM noise is generated through parasitic capacitances that create high-frequency current paths and do not exist at dc or lower frequencies.

An EMI filter is used to protect against both CM and DM currents. For maximum filter effectiveness in attenuating these high frequencies, the filter must be enclosed within a Faraday shield bonded to the chassis. Another method of controlling CE is related to controlling the rise time and other component parameters. Information on component selection is given in section 3.2.1.

4.2.1.1 Differential Mode Emissions

Filtering the fundamental switching frequency of the power supply or device is considered first. Since the fundamental switching frequency and its harmonics are on the input power lines during CE01/CE03 testing, filtering is needed. At the lower end of the frequency range (in the kHz range), noise is coupled onto the power line most efficiently by DM coupling; therefore, the method of filtering should be line-to-line. When designing a power supply filter or other device, possible implications the filter design has on other EMI tests should be considered. For instance, there are limits on how large the front end capacitor can be and still meet the transient emission limits (see section 4.3).

In addition, capacitors and inductors in the filters resonate and cause a region of increased CS. Adding small resistors to the filter damps this response. However, since adding dc resistance affects the dc voltage of the circuit, the damping resistors are added in parallel with the front filter inductors. If this technique is used, the inductance is split and part is placed in line with the damping resistor to keep the filter from being bypassed. This technique serves to further reduce the inductance size. Since the inductor is in the same line with the damping resistor, there is minimal dc current flow. For additional information on resistances needed for damping see section 4.5.

Another aspect to consider is distributing the inductance on both the high side and return side of the circuit. Even though the line-to-line filtering is primarily to control DM currents, distributing the inductance keeps one path from being more favorable for CM currents. An example of a buck regulator power supply with a two-stage line-to-line LC filter and damping resistors for high-frequency current is shown in figure 4-1. Information on general power supply topologies is presented in section 3.5.1.

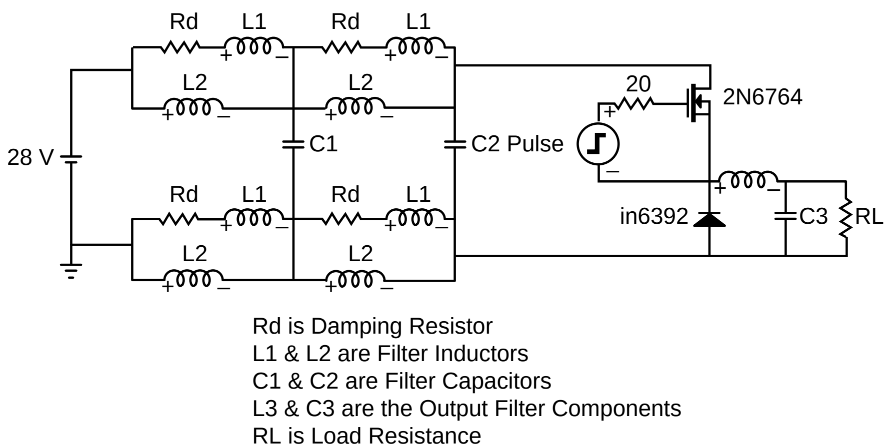


Figure 4-1. Buck regulator power supply with two-stage filter.

4.2.1.2 Common Mode Emissions

The next consideration is CM filtering needed for higher frequency circuit switching devices such as diodes, clocks, etc.

4.2.1.2.1 Heat Sinks and Bypass Filtering

One contributor to CM noise is the path to ground provided by the parasitic capacitances in heat sinks. Since heat sinks are mounted to the chassis, capacitance formed by the component and heat sink, with thermal insulation as the dielectric material, provides a path to ground for switching noise currents. This parasitic or “stray” capacitance is an alternate path for switching currents (spikes) to flow from the chassis through the circuit into the input power lines. Additional information on parasitic capacitances is given in section 3.3.2. This CM noise problem is altered by placing bypass capacitors close to the switching components in the box. Bypass capacitors are capacitors when placed close to the noise source (diodes, transistors, etc.) provide a shorter path back to the source. This decreases noise in the power lines and also decreases the radiating loop area of the CM noise. Damping resistors may also be needed with bypass capacitance to damp the resonances of the capacitor with circuit inductors.

The capacitance of the switching device to the heat sink varies depending on the materials used for mounting and the area of the heat sink. For a T03 type connector, typical capacitances are 50 to 100 pF, but for larger mounting devices the capacitance increases. The size of the bypass capacitor needed varies depending on the heat sink capacitance and the type and level of signal for which suppression is desired.

For example, figure 4-2 shows a buck regulator power supply with stray capacitances between the diode heat sink and structure and the MOSFET heat sink and structure. The high-frequency switching noise from the diode and other heat sink components takes the stray capacitance path to the input power lines where the CE are measured. With bypass capacitors, the noise current path is reduced to a defined area within the equipment (fig. 4-2). Note on figure 4-2 that the bypass capacitors for the MOSFET noise source are close to the source, but, for the diode noise source, are placed opposite from the source. Because violent changes occur in ac voltage at the diode noise source, bypass capacitors interfere with the intentional operation of the signal. For MOSFET's, all of the ac at the source is noise and bypass capacitors are placed close to the source. In general, for any noise source with violent ac voltages as part of the signal, the bypass capacitance should be placed opposite from the source.

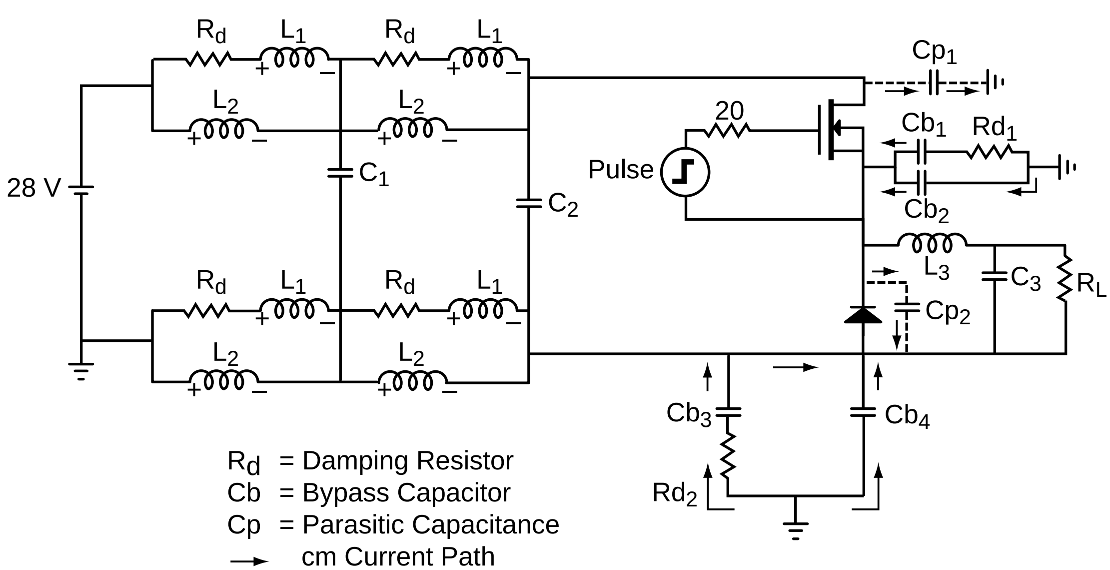


Figure 4-2. Buck regulator power supply with parasitic capacitances.

Figure 4-3 shows that the diode voltage is a clean square wave, but the input current has high-frequency noises associated with it. By adding a 10-nF capacitor from the heat sink to chassis (fig. 4-2), the noise decreases on the input power lines but increases on the diode voltage (fig. 4-4). By changing the bypass capacitance to 100 nF, the high-frequency noise is almost eliminated from the input power line, but is shifted to the diode voltage (fig. 4-5). Most of the high-frequency noise riding on the diode voltage remains in the box and does not show up on EMI tests. The noise does not actually diminish, but is allowed to return to its source through the bypass capacitors. The possibility for radiation from the loop is also reduced by having a shorter loop for current flow.

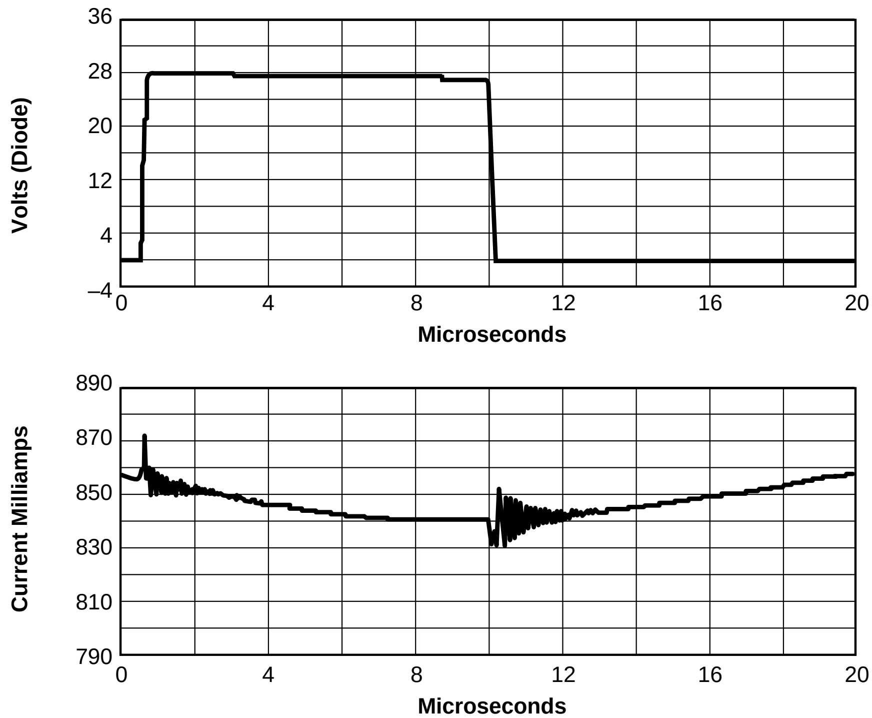


Figure 4-3. Diode voltage and current ripple without bypass capacitance.

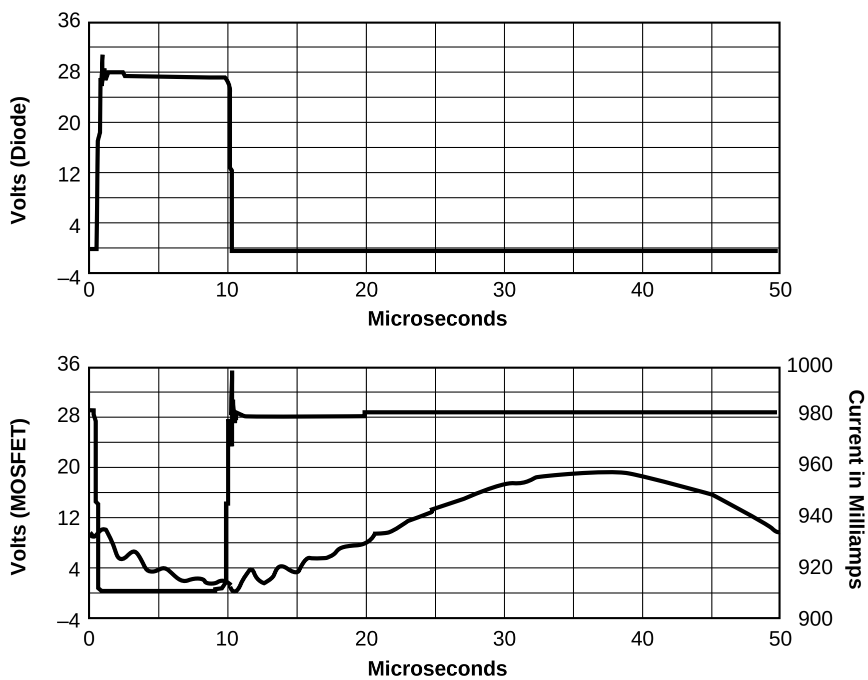


Figure 4-4. Voltage and current ripple with 10-nF bypass capacitance.

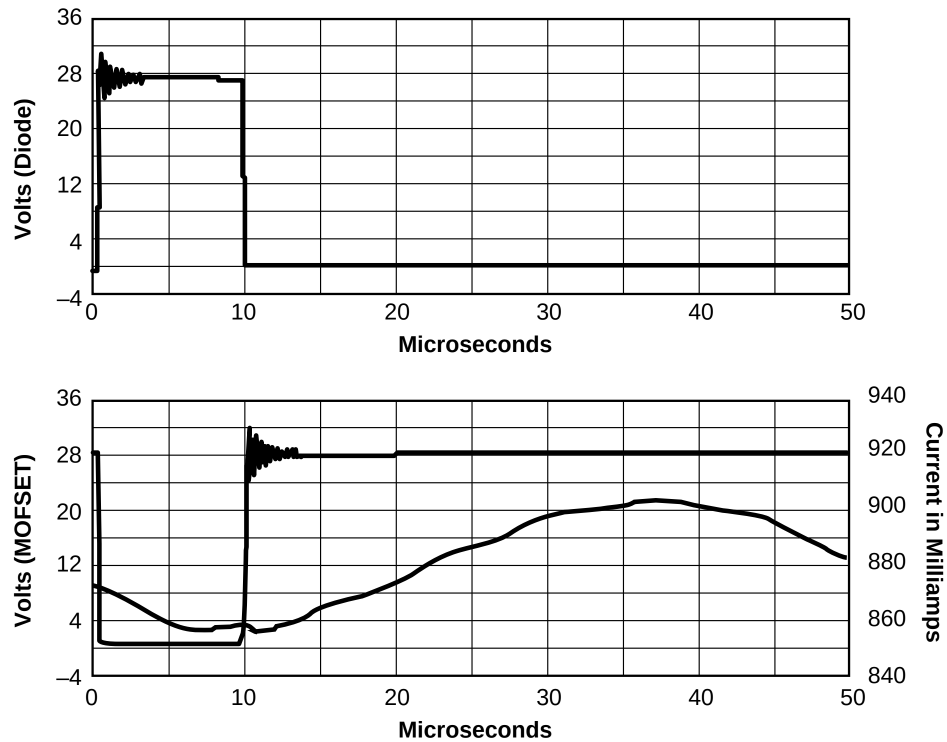


Figure 4-5. Voltage and current ripple with 100-nF bypass capacitance.

4.2.1.2.2 Mounting Washers

Another method to circumvent CM current is to use a special washer for mounting that is insulated on either side and has a copper tab which can be connected to a lead of the power switching device. This connection provides an alternate path for the current to flow that keeps the CM currents off the input power lines. The only company known to this author that makes this special washer is Bergqueist.

4.2.1.2.3 Common Mode Chokes

An alternate method to suppress CM noise is to use a CM filter on the front end of the circuit. This is often referred to as a CM choke. The differential current will not be affected by the impedance, but the CM spike sees a large inductance value that impedes the CM noise from returning through the power lines. An additional consideration, if CM inductors are used, is that damping resistors are not always needed; CM chokes are resistive at high frequencies. For additional information on CM chokes see section 3.3.3.3.

4.2.1.2.4 Damping Resistance

Both the differential filter and CM bypass capacitors require damping at the resonant frequency of the elements. At this resonant frequency, emissions of the circuit are amplified. Since the major problem with resonant frequencies occurs during CS tests, more information on damping resistors is presented in section 4.5.

4.2.1.3 Leakage Current Requirements

Because bypass and front-end CM filters have capacitors from line-to-ground, the maximum capacitance allowed must be considered. Line-to-ground capacitance limits are usually imposed to control the amount of leakage current that flows from box to structure. Although this requirement is applicable for ac-powered boxes since ac leakage is a safety threat, leakage current controls are also placed on dc systems. For methods to determine maximum capacitance to structure for a given leakage current requirement see section 4.2.2.3.

4.2.1.4 Radiation Around Filters

The effectiveness of filtering is reduced by radiated coupling around filters. Even though this is a radiating phenomenon, its effects show up during CE testing. One method to reduce radiation around filters is to decrease radiation within the equipment. Board and trace layout, which affect radiation within equipment, are covered in section 3.2. Another culprit of electric fields within equipment is wide swings in voltage that occur on heat sinks. To control this radiation problem, electrically connect the heat sink to a fixed potential, normally the bulk current return, and place it away from the input filter. Shortening lead lengths, adding ferrite beads for additional high-frequency filtering, and shielding or isolating the filter on another board also reduce the effects of radiation coupling. For additional information on this topic see reference 4-4.

4.2.2 Modeling/Prediction Techniques

4.2.2.1 Differential Mode Filtering

To determine the front-end filter needed for a power supply, Bode plots are used. For instance, a single-stage filter has an early break frequency, but fall off at only 20 dB/decade. An LC filter has twice the break frequency, but falls off at 40 dB/decade, and a double LC filter breaks even higher, but falls off at 80 dB/decade. Figure 4-6 shows a Bode plot for the three cases. A filter is selected by determining the fundamental switching frequency and the amount of filtering needed to reduce the required output current to acceptable levels at the input. Output current considered here is current at the output of the current switching device, before accounting for output filtering.

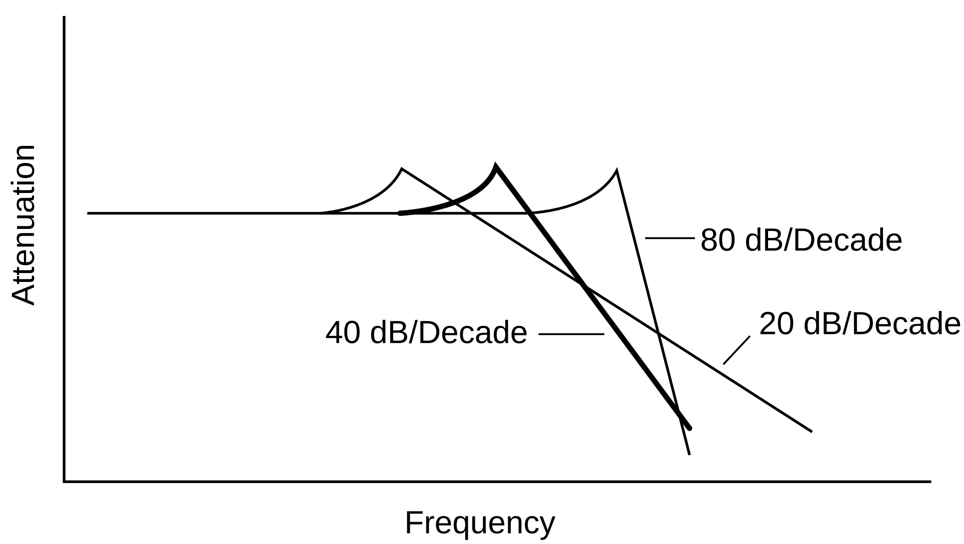


Figure 4-6. Bode plot for three LC filters.

Concentrating the filter design emphasis on the fundamental frequency amplitude is often adequate since many CE specifications fall from 20 to 40 dB/decade to the megaHertz range, and front-end filters usually have this attenuation or more. For example, given a power supply with 5-A ripple current drawn by the power stage, the root-mean-square (rms) current is (5/2.8) = 1.8 A. The rms calculation varies depending on the wave shape of the current. The limit for CE03 at 20 kHz, for instance, is 0.014 A rms per MIL-STD-461.[4-8] Therefore, the attenuation needed is 20*log(1.8/0.014) = 42 dB of filtering. This amount of filtering at the specific frequency is then found on the Bode plot for a circuit with specific component resonance and attenuation.

To design a power filter with specific attenuation at a certain frequency, make $LC \approx (1/x\omega^2)$, where x is the required attenuation, $\omega = 2\pi f$, and L and C are the inductance and capacitance needed for attenuation. Choose the bulk capacitor next based on power staging heat and ripple current requirements (component specifications have this information). Put an identically sized capacitor at the front end of the circuit. Choosing a higher order filter is often advantageous because of saving weight by using more small inductors instead of a few large ones. To decrease the characteristic impedance ($\sqrt{\frac{L}{C}}$) and, thus, the effects of load transients, choose a larger value of capacitance than the inductance. Tips on parallel filter inductors and capacitors with damping resistors are described in section 4.5.2.

Spice analysis is also used to predict CE. In addition, if all the circuit components, including parasitic components are modeled, this tool gives quite accurate results. For more information on Spice analysis see reference 4-2. For more information on power supply filtering see reference 4-4.

4.2.2.2 Common Mode Filtering

To predict the amount of CM filtering needed in a particular circuit, consider the CM current path. For example, if a diode or transistor in the circuit is switching at a certain frequency and is mounted on a heat sink, a typical capacitance from the device to the heat sink and chassis is on the order of 150 pF.

The current path is actually through the stray capacitance between the device and the heat sink, through the chassis impedance and the LISN or power bus impedance (fig. 4-7). The amplitude of the signal in the CM path is determined by using Fourier transforms of the signal of interest. Information on calculation of Fourier transforms is given in section 3.2.1.2. When frequency reaches sufficiently high values, the stray capacitance becomes a more favorable path for the switching noise to return to its source. To predict the amplitude at a certain frequency for a given signal, the change in frequency from the first corner, $f1$, to the frequency of interest is used to determine a delta in dB. This delta is used to determine the amplitude of the signal at the frequency of interest. This type prediction was taken from reference 4-3. For example, consider a transistor switching 120 V at 150 kHz in a switch mode power supply (shown in fig. 4-7 and its frequency domain spectrum in fig. 4-8). The frequency of interest selected is 500 kHz and the signal is a square wave with a 50-percent duty cycle.

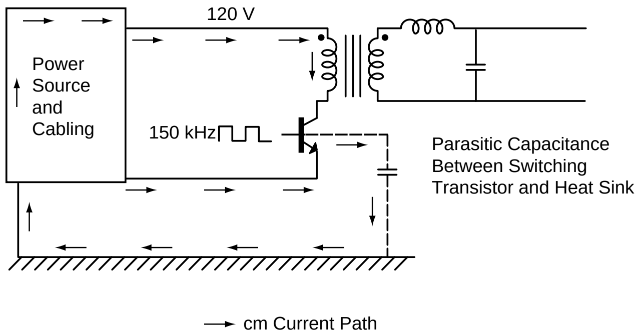


Figure 4-7. Switched-mode power supply with CM noise path.

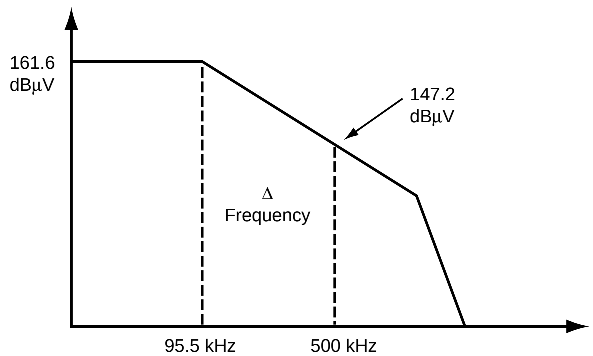


Figure 4-8. Frequency domain spectrum envelope.

Parameters Necessary to Predict CM Filtering:

Stray Capacitance Equivalent Impedance

$$Z = \frac{1}{(2\pi fc)}, \tag{4-1}$$

where f is the frequency of interest and c is the capacitance.

Conversion of the Amplitude to dBμV

$$20*\log(120) = 161.6 \text{ dB}\mu\text{V} \quad \text{(for a 50-percent duty cycle)} . \tag{4-2}$$

Determining the First Break Frequency

For a 3.3-μs pulse width and a 100-ns rise time, the first corner frequency is calculated as follows:

$$f1 = \frac{1}{\pi(\tau)} = \frac{1}{\pi(3.3\mu s)} = 95.5 \text{ kHz} . \tag{4-3}$$

Calculating the Delta to the Frequency of Interest

$$\Delta = \Delta = 20*\log\left(\frac{500 \text{ kHz}}{95.5 \text{ kHz}}\right) = 14.4 \text{ dB} . \tag{4-4}$$

Switching Signal Amplitude

For linear falloff, the frequency change equals the amplitude change; therefore, the signal amplitude equals the signal voltage in dBμV minus the delta change in amplitude at the frequency of interest. Thus, the CM signal amplitude at 500 kHz is:

$$161.6–14.4 = 147.2 \text{ dB}\mu\text{V}. \tag{4-5}$$

Converting the Voltage to Current

Since most CE measurements are specified in dBμA, it is beneficial to convert the CM voltage amplitude to a current amplitude. To determine the current amplitude, the impedance of the CM path is calculated as follows:

$$Z = \frac{1}{2\pi(500 \text{ kHz})(150 \text{ pF})} = 2{,}122\ \Omega = 66.5 \text{ dB}\Omega\ . \tag{4–6}$$

Current amplitude is the CM voltage minus CM impedance (in dB), i.e.,

$$147.2 \text{ dB}\mu\text{V}–66.5 \text{ dB} = 80.7 \text{ dB}\mu\text{A}\ . \tag{4-7}$$

Since the current divides across the high and low sides of the input power lines, 6 dB is subtracted to determine the current amplitude in each side of the power line, i.e.,

$$80.7 \text{ dB}\mu\text{A}–6 \text{ dB} = 74.7 \text{ dB}\mu\text{A}\ . \tag{4-8}$$

This current amplitude is compared to the specification limit at the frequency of interest to determine the amount of filtering needed. However, this CM value adds to whatever differential noise is present at the frequency of interest. This method is used for each switching device.

LISN Measurements of Conducted Emissions

If an LISN measurement is used, the CM current is multiplied by the LISN impedance to obtain a voltage that can be compared to a voltage ripple specification. LISN’s are described in sections 3.3.3.1. and 5.1.3.3. The sample calculation for an LISN with a 50-ohm characteristic impedance follows:

$$\text{CM Voltage} = 74.7 \text{ dB}\mu\text{A}+34 \text{ d}\beta\Omega = 108.7 \text{ d}\beta\mu\text{V}\ , \tag{4-9}$$

where 34 dBΩ represents 50 Ω in dB.

This dB voltage can be converted to volts by the following:

$$A\log\left(\frac{108.7\ (\text{dB}\mu\text{V})}{20}\right) = 0.272 \text{ V}\ . \tag{4-10}$$

This value in volts is compared to whatever CM voltage specification is applicable at 500 kHz. Note: Emissions will vary with the frequency of interest and a spread sheet is useful for broad scale calculations.

4.2.2.3 Leakage Current Calculation

Maximum equivalent resistance for a given leakage current requirement is calculated by taking the allowed voltage ripple and dividing it by the maximum amount of current through structure. Thus, for a 1 Vrms allowable voltage ripple and 5 mA maximum leakage current, the maximum equivalent resistance of the combined capacitance to ground is calculated as follows:

$$\text{Equivalent Resistance} = \frac{1\ \text{Vrms}}{5\ \text{mA}} = 200\ \Omega\ . \tag{4–11}$$

Next, the frequency at which the voltage ripple is specified is considered. If the voltage ripple is specified at 80 kHz, the maximum equipment capacitance to ground is calculated as follows:

$$\frac{1}{2\pi(80\ \text{kHz})C} = 200\ \Omega\ . \tag{4–12}$$

Solving for C gives:

$$C = 0.01\ \mu\text{F}\ . \tag{4-13}$$

This capacitance represents the maximum combined component to chassis capacitance per piece of equipment. Since capacitors add in parallel, this requirement limits design solutions for other conducted emission concerns that are controlled by adding capacitance to structure.

4.2.3 Retrofit Fixes

If a CE03 qualification test is completed or conducted, the emissions tests in chapter 5 are performed and exceedances are identified; there are a few ways to reduce the emissions. The ideal situation is to run a test to differentiate between CM and DM noise (type of emissions will alter the desired fix). If you do not have the equipment for this test (see chapter 5), a rule of thumb is that most differential noise is not well controlled above 2 MHz. For differential noise, line-to-line filtering is most effective; while for CM noise, line-to-ground filtering is more effective.

4.2.3.1 Electromagnetic Interference Filters

EMI filters can reduce both common and DM noise, but there must be sufficient room in the equipment to install the filter and to ensure that it is properly grounded and enclosed such that emissions cannot couple around the filter via radiation. For additional information on EMI filters see section 3.3.3.

4.2.3.2 Ferrite Beads

Another option to control CE is the ferrite bead. These beads are effective against CM emissions and, if the size of bead needed is small enough, are often installed outside the equipment. These beads act as a high impedance for the high-frequency current flowing in the CM direction, but do not attenuate the dc current flow. Using ferrites for the control of EMI is described in detail in section 3.3.3.2.

4.3 Conducted Transient Emissions (TT01/CE07)

TT01 and CE07 are similar requirements that specify and measure, in the time domain, the load-induced effect on power quality caused by cycling the EUT on and off, as well as through any and all of its various modes of operation that might reasonably be expected to significantly affect the line voltage.

4.3.1 Design Considerations

The primary consideration in meeting the transient emission requirement is the size of the front-end filter capacitor. If this capacitor is too large, it will cause the input power line voltage to have a transient representative of the RC time constant of the circuit. Since capacitance is needed to meet the CE03 requirement, this may be a problem.

One way to decrease the impacts of a large front-end capacitor on the input power line is a soft-start device to bring the equipment on line slowly. One method (fig. 4-9) is to use a resistor that is bypassed by a relay. When using this method, however, delay power stage activation until after contact closure. Have the RC time constant less than the relay pickup time to minimize the remaining transient effects when the contacts close. If the relay closes after the bulk capacitor has charged to levels close to the input voltage, than additional circuitry is not required to further delay the relay. In addition, placing the relay close to the bulk capacitor instead of the soft-start circuitry increases the relay delay. Add delay circuitry for the relay if the capacitor is too large to have a reasonable RC time constant. Typical values for the resistor range from 0.5 to 5 ohms.

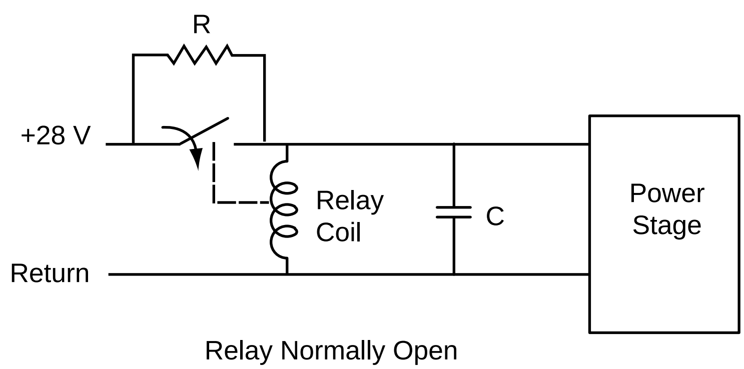


Figure 4-9. Soft-start switch using relay.

Transistor soft-start circuits are also used (fig. 4-10). When using this method, it is important to delay power stage activation until after Q1 is turned on. Also, the $R_T C_T$ time constant is adjusted to tailor the turn-on characteristics. When a current-driven transistor is used (infrequent for space applications), place an inductor in the primary side of the transformer to slowly turn on a transistor, which in turn slowly turns on the filter. Chose the inductor by considering the LC time constant required for a given delay.

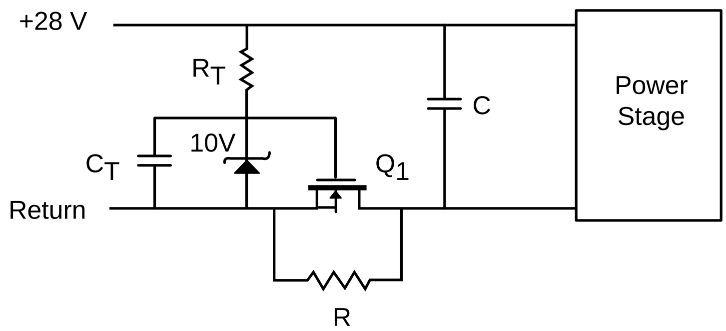
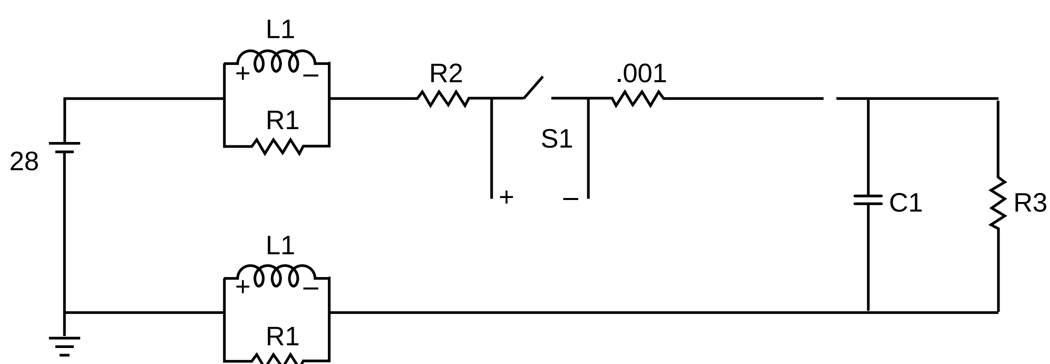


Figure 4-10. Soft-start switch using MOSFET.

4.3.2 Modeling/Prediction Techniques

Transient limits normally envelope the allowed voltage transients on the input power lines. To meet the transient limits, a wide range of front-end filters are effective. For an estimate of the largest front-end capacitor that meets the limits, pick a point on the envelop of the transient limit. Use the RC time constant of the circuit to obtain the maximum capacitance. If the resistance from the power source to the equipment of interest is not known, use the default resistance value for the required LISN. For example, given a voltage source of 28 V (V_s), a line resistance of 0.5 ohms (*R*), and a point on the transient limit of 14 V (V_C) at 10 μs (*t*), substitute using $V_c = V_s - Vs * e^{\left(\frac{-t}{RC}\right)}$ to obtain C = 28.8 μF. This process gives a rough idea of the maximum capacitance allowed to meet transient limits that are curved to simulate capacitance charging times (MSFC-SPEC-521B). Factoring in the inductance of the line in the model and using spice programs make this analysis process more efficient and precise.

A Spice analysis tool is used to model the turn-on spike of a given piece of equipment. This is done by modeling the LISN, a switch, and the front-end characteristics of the EUT. Figure 4-11 shows a model in which the front-end capacitor of the EUT is variable. By altering this value, the effects of the capacitance on the turn-on transient are shown.



R1 = 25 Ohms
R2 = 0.31 Ohms
R3 = 26 Ohms
L1 = 10 Microhenrys
S1 = Switch
C1 = 10–20 Microfarads

Figure 4-11. Transient test setup.

Figures 4-12(a) through 4-12(d) show the Spice modeled turn-on transient traces with varying values for C1. These data are compared to the turn-on transient data from the same model during actual testing (figs. 4-13(a) through 4-13(d)). Comparing the Spice and test data shows a close correlation. The differences are primarily related to how low the bus voltage drops. The drop in the test model is generally not as low as in the Spice model. Some voltage difference is accounted for by the effective series inductors and resistors in the lead lengths of the test setup. Another difference in the data is the lapses in the downward curve in the test data. These lapses are due to bouncing of the turn-on switch during testing and make the test data appear shifted to the right.

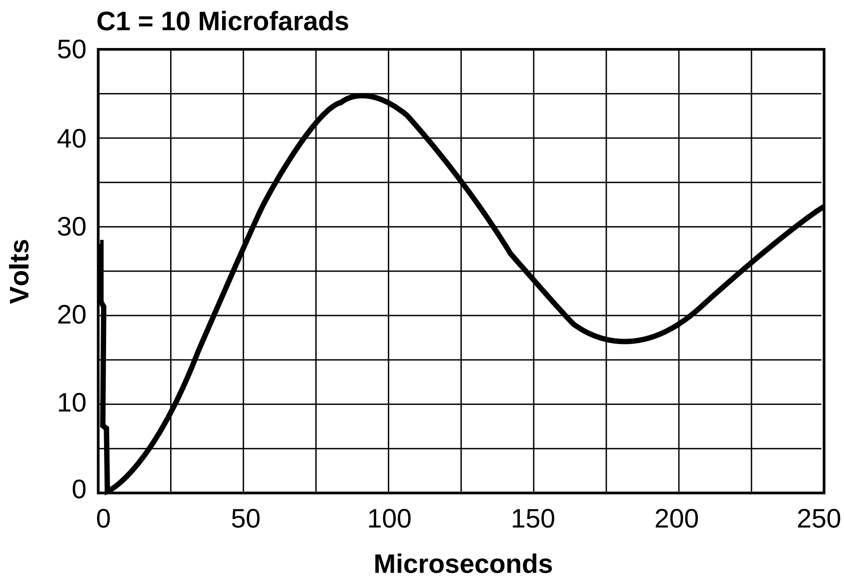


Figure 4-12(a). Predicted turn-on transients (10 μF).

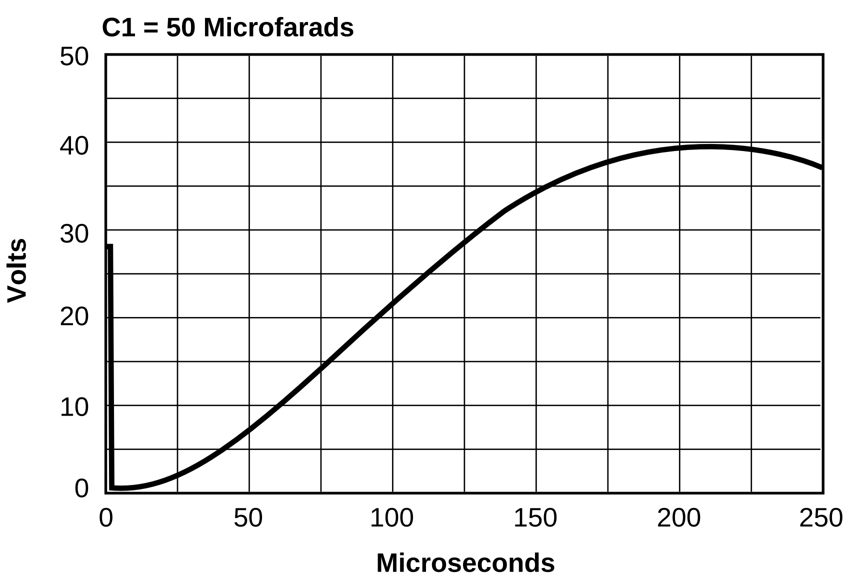


Figure 4-12(b). Predicted turn-on transients (50 μF).

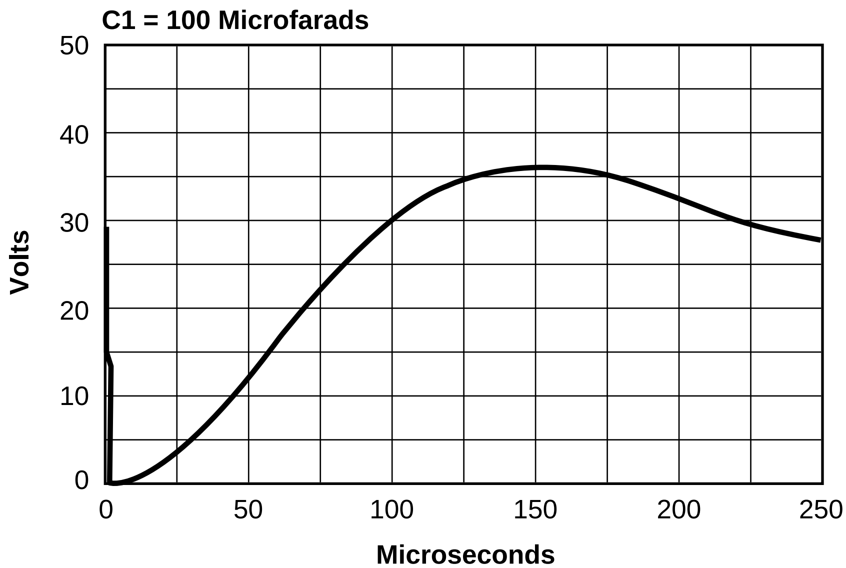


Figure 4-12(c). Predicted turn-on transients (100 μF).

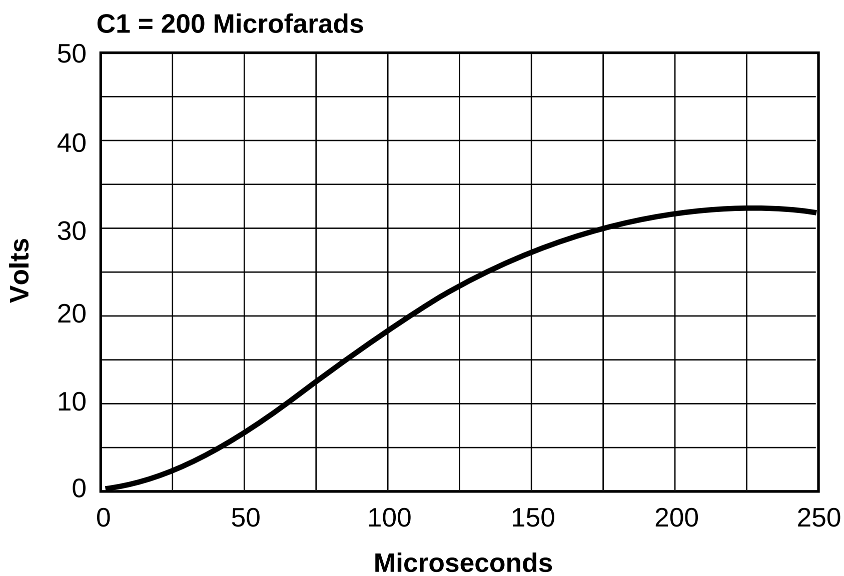


Figure 4-12(d). Predicted turn-on transients (200 μF).

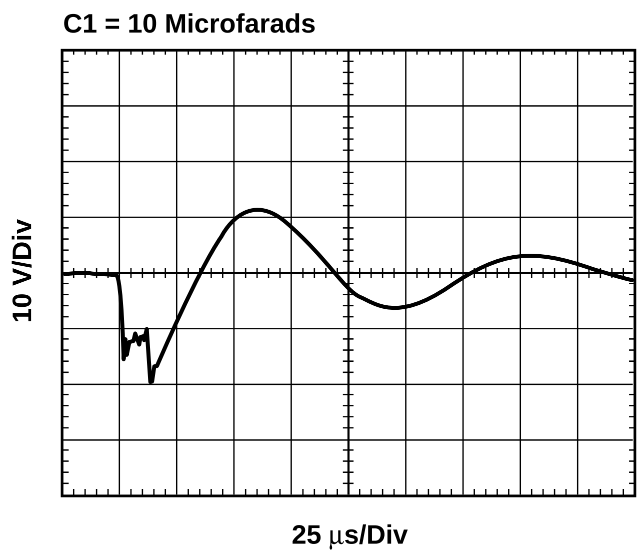


Figure 4-13(a). Turn-on transient test data (10 μF).

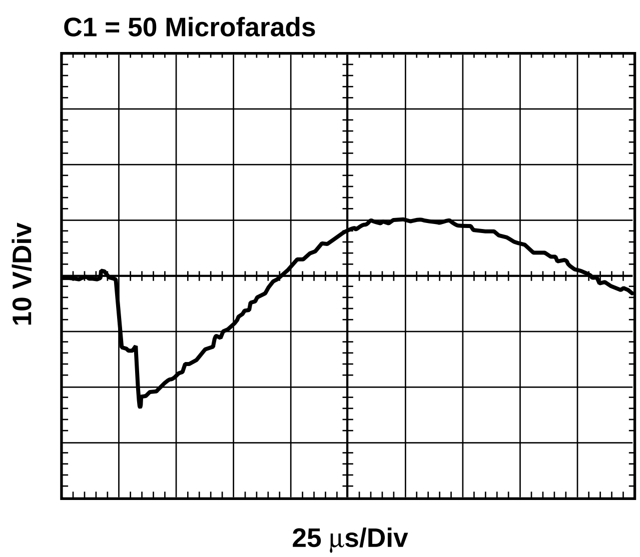


Figure 4-13(b). Turn-on transient test data (50 μF).

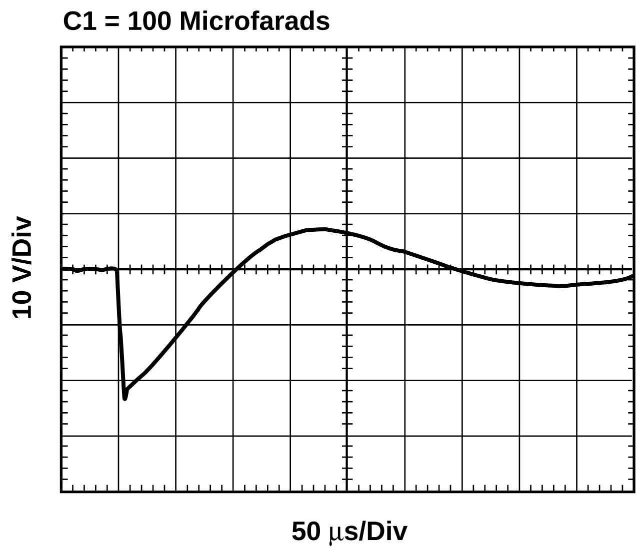


Figure 4-13(c). Turn-on transient test data (100 μF).

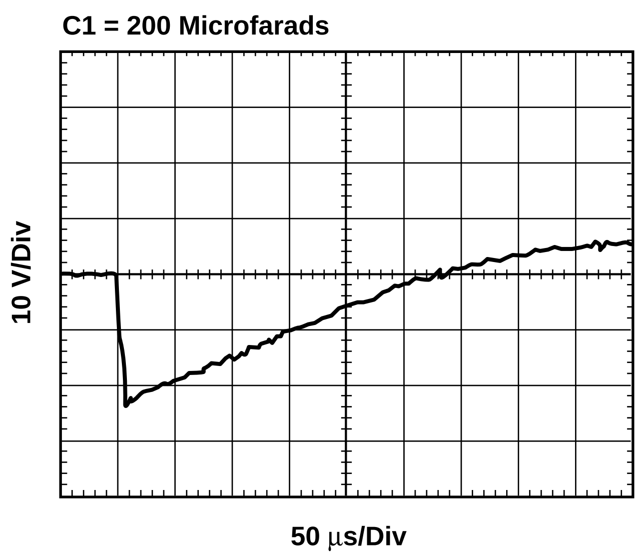


Figure 4-13(d). Turn-on transient test data (200 μF).

Accounting for series inductance and resistance of the leads adds precision to Spice modeling. However, the Spice model without these adjustments is typically a worse case and reasonably used for prediction. The effects of soft-start circuitry are also shown in modeling and test data (fig. 4-14).

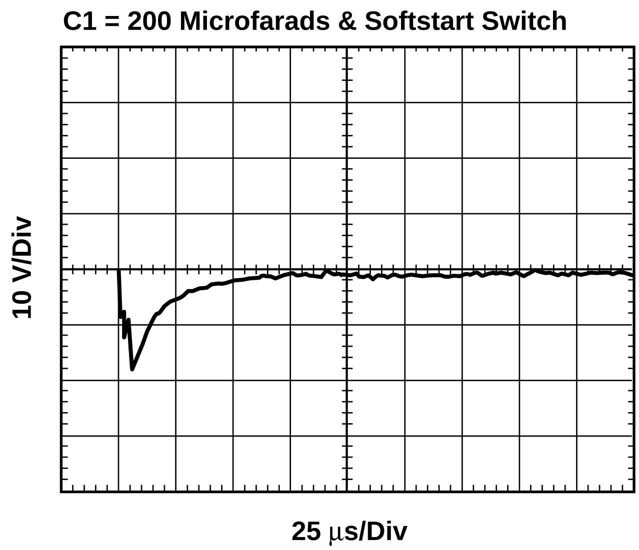


Figure 4-14. Transient test data using soft-start switch.

4.3.3 Retrofit Fixes

After the designed box is complete, a soft-start add-on is often the only realistic fix. It is often not an option to return to the breadboard arena and change the input filter. To reduce complexity, a relay circuit is used. Relay circuits are mounted outside the chassis of the equipment if space inside the equipment does not allow the addition of components. Because transistors use the ground plane as part of the circuit, mounting a transistor soft-start circuit outside the equipment chassis is not realistic.

4.4 Radiated Emissions (RE02/RE04)

RE02 (14 kHz to 10 GHz) limits the electric field radiation, and RE04 (50 Hz to 50 kHz) limits the magnetic field radiation from the EUT and its associated cabling.

4.4.1 Design Considerations

Since RE04 limits are usually controlled by having an enclosed metal equipment housing, this section concentrates on meeting RE02 limits. For more information on enclosure shielding see section 3.4.1.

Current flowing on a conductor results in electromagnetic radiation. Hence, many of the design considerations for RE are similar to those for CE, especially for CM noise control. If emissions are kept off the power line and confined to the box, ordinary metal box enclosures contain most of the noise. Signal lines, however, must also be considered and the appropriate shielding and twisting applied (described in section 3.4).

4.4.1.1 Electric Field Emissions

Designing to control electric field emissions is covered in various sections of chapter 3. One important consideration is reducing electric field emissions is to keep pulse rise times as slow as possible while meeting the circuit application. This approach is described in section 3.2.1.

Another consideration in controlling electric field emissions is circuit shielding, which is done at the equipment level and on input and output cables. Techniques for shielding cables and shielding enclosures are given in section 3.4.

4.4.1.2 Magnetic Field Emissions

To control magnetic field emissions, the primary goal is controlling loop areas. For instance, the input and output power and signal leads are twisted, and the traces and wiring within the equipment are routed close to their returns to minimize loop area. This routing method provides cancellation of magnetic fields with equal magnitude but opposing phase and reduces overall radiated magnetic fields. In addition, using flat wide conductors within equipment, such as traces on PC boards, instead of round ones reduces the radiated field.[4-4]

The most common source of magnetic field emissions in a switch mode power supply is the "high amp-turns" components or magnetics. One method to reduce the leakage flux from magnetic core gap transformers is adding a shorted turn for the leakage flux. This turn goes around the entire magnetic device and causes an opposing current to the leakage flux. "When the flux couples to the shorted turn, a current is induced in the direction such that the resulting flux opposes the incident flux, which changes the pattern of the radiation."[4-4] This change in the radiation pattern reduces the area radiated by the magnetics.

For shielding low-frequency magnetic fields, loss due to reflections is the primary field-shielding mechanism. The incident magnetic field induces a surface current in the shielding material, which in turn re-radiates. The reradiated field is (almost) equal in magnitude and opposite in phase to the incident field. If a discontinuity exists, the currents are disrupted and the reradiated field will not cancel with the incident field. This disruption in current cancellation will degrade the shielding. Additional information on reducing discontinuity is found in chapter 3.

4.4.2 Modeling/Prediction Techniques

While prediction methods are available for predicting RE, there are complicated programs which require tedious circuitry input parameters. More general calculation routines are available in which the calculations are done with pencil and paper or computer spread sheet.

Table 4-1 gives an example of the general type calculation and presents a series of columns which are added to determine the predicted RE from a given component. Basically, the Fourier transform is computed for a given signal. The amplitude at a specific frequency (usually the switching frequency of the component) is found using methods described in chapter 3 and example calculations given in section 4.2.2.2. This amplitude (dBμV) is placed in column 1. A correction factor is added to account for the conversion of the conducted data to free space. This factor is assumed to be –34 dB for a perfect 1/4 λ. The frequency, $f3$, is computed to give a factor that takes into account the actual length of the cables. The calculation is shown in table 4-1 and corrects the 1/4 λ assumption. Finally, the number of leads from the signal is accounted for. The measurement distance is also be factored in; for MSFC test applications it is 1 m and requires no correction factors. The resulting factor is compared to the specification at that frequency to determine the dB of attenuation needed. The attenuation is obtained by using metal equipment housing, shielding, twisting, etc. This prediction method is reasonable for checking the most likely culprits in the circuits (diodes and transistors that switch high levels of current). For additional information regarding EMC radiation prediction see reference 4-3.

Table 4-1. RE prediction analysis.

Frequency	(1) Cn	(2) Voltage to E-Field	(3) Antenna Factor	(4) Number of Leads	(5) Distance	(6) Result
20 MHz	130 dBμV	-34	-2.7	6	0	99.3 dBμV

(1) Frequency domain amplitudes - Fourier transform

(2) Voltage to field intensity level 1 meter away from conductor = –34 dB

(3) -10 log $\frac{f_3}{f_x}$ where: $f_3 = \frac{3\times10^8}{4l}, l$ = wire length in meters: when $f_x > f_3$

the correction factor = 0 dB.

(4) + 10 log N, where N = number of leads

(5) - 20 log D, where D = test distance in meters

(6) Result in dBmV/m, sum (1) through (5).

4.4.3 Retrofit Fixes

When not practical to control radiated EMI by design or layout change, a retrofit fix is needed. The objective is make the least impact on design, packaging, and cost of the product, yet bring it into compliance.

4.4.3.1 Connector Decoupling

Significant radiated coupling occurs to the leads between the last filter element and the chassis connector. By using a connector with integral capacitance as low as 100 pF, significant improvements are realized in the band covering roughly 10 to 220 MHz.

4.4.3.2 Ferrites

Ferrite is a common material for magnetic cores with relative permeabilities ranging from 40 to 10,000. Because of interwinding capacitance, multiple-turn cores have limited usefulness at higher frequencies. Hence, only ferrite beads with a single turn are considered here. Beads are very effective in limiting coupled energy to leads. A convenient property of ferrites is that the impedance is resistive above a given corner frequency. This implies that the coupled energy is dissipated and not reflected as heat. To obtain the required impedance, the general geometry of the bead should be longer. This is more effective than increasing the outer diameter.

4.4.3.3 Ferrite Toroids

Large diameter toroids (1 to 2 in) are used external to the EUT to limit CM currents that cause radiation. These cores are typically used with 3 to 10 turns to increase the net impedance at lower frequencies. Again, the interwinding capacitance is in parallel with the inductor and, by shunting the inductor, limits the high-frequency performance.

4.4.3.4 Clamp-On Ferrites

At higher frequency ranges, acceptable results are often obtained with a single turn. Many manufacturers have developed split cores with plastic retaining housings to clamp the ferrite on a cable. Geometries are available for coaxial or round cables as well as ribbon cables.

4.5 Conducted Susceptibility (CS01/CS02)

CS01 (30 Hz to 50 kHz) and CS02 (50 kHz to 400 MHz) are requirements to control and determine the susceptibility level of the EUT to audio frequency and RF interference signals on the input power leads.

4.5.1 Design Considerations

4.5.1.1 Window of Susceptibility

There is a region of susceptibility, called “window of susceptibility,” between the frequencies of active error checking of the voltage controlled feedback loop of a power supply and the corner frequency of the input power filter. The feedback loop becomes inactive at the point of unity gain crossover. It is considered advantageous to lower the frequency of this unity gain crossover to decrease the chance of instability in the feedback loop. This practice, however, leaves a region without filtering of input noise that leads to failures during CS testing. “There are many circuit configurations and multiloop techniques for closing the attenuation window while retaining absolute stability.”[4-5] For more information on “window of susceptibility” see reference 4-4.

4.5.1.2 Damping Resonances

Again, adding components to meet the EMI requirements for one test causes difficulties in meeting EMI requirements for others. Filtering used to meet CE requirements (CE01/CE03) causes problems in meeting CS01/CS02 requirements. Inductors and capacitors in equipment front-end filters have a series resonant frequency in which the ripple voltage injected on input power lines during CS01/CS02 testing is amplified beyond the limits of the capacitors in the input power filter. Effective resistance present in component leads and traces provide resistance to damp these resonances (especially tantalum capacitors since they are lossy). However, resistances may need to be added to efficiently damp the resonances. If the resonances are not damped when the CS01/CS02 signals are applied at resonant frequency, component failures may result.

One way to add resistance to the input filter for damping, without seriously affecting the dc current flow and thus heat dissipation, is to put a resistor inductor series combination in the filter in parallel with the filter inductive elements. Damping resistors in this configuration are shown in figure 4-1. Since the path with the inductor and series damping resistor carry dc and are sized accordingly, having parallel inductors will help decrease the physical size and weight needed for a given value of inductance. Parallel inductors with series damping resistors are smaller in size with higher values of inductance to attenuate low level noise currents.

To obtain the appropriate value for damping resistance, the combination of series and parallel inductors and capacitors in the circuit is considered. For additional information on computing equivalent inductances and capacitances for the circuit see reference 4-4.

For CM input filtering, damping resistors are often not necessary because the inductors become somewhat resistive at high frequencies. Ferrites and tape-wound inductors are lossy and thus usually have series resistance. These effective resistances act as damping resistors for the circuit.

4.5.2 Modeling/Prediction Techniques

The value of R is optimized by running an ac analysis Spice program to determine which value provides the lowest resonance peak for a given configuration. For the best results, the entire circuit including parasitic effects is modeled. For more information on Spice modeling see reference 4-2.

If Spice analysis routines are not available, more basic equations are used. For damping a parallel combination of inductors and capacitors with series resistances (fig. 4-15(a)), the following equation is applied: $(RL+RC) > \sqrt{\frac{4L}{C}}$. For damping using a parallel resistance as shown in figure 4-15(b), the following equation is applied: $R > \sqrt{\frac{L}{4C}}$.

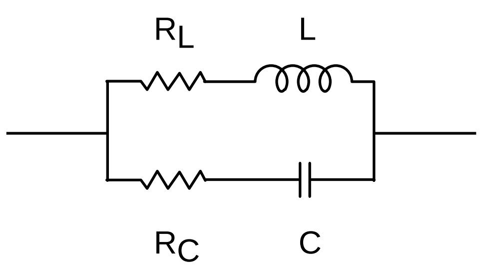


Figure 4-15(a). Damping with series resistance.

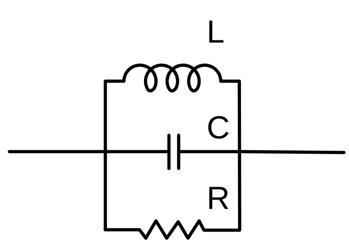
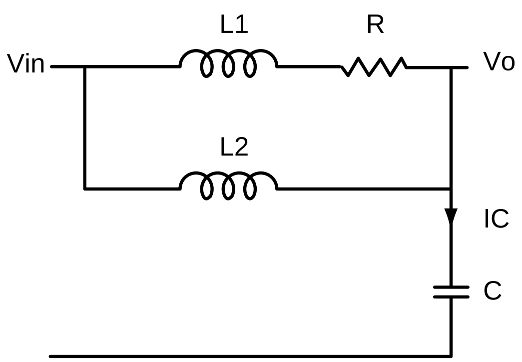


Figure 4-15(b). Damping with parallel resistance.

These equations help predict the damping value needed by using component values. However, it is important to remember that effective resistances in component leads and ferrites aid in damping resonant frequency amplification.

To decrease effects on dc current flow and thus filter efficiency use parallel inductance (or capacitance) in filter design (figs. 4-16(a) and 4-16(b)). Although this method always requires an increase size of the filter, doubling the inductance values usually has less size impact than adding additional capacitors. The parallel inductor can be sized smaller, although it doubles in value, than the original filter inductor because it will not carry dc current, but the parallel capacitor is typically the same value and size as the original capacitor. This assumption is true when the capacitor values are already sized larger than the inductor values to decrease the circuit characteristic impedance and thus its response to load transients. The characteristic impedance is $\sqrt{\frac{L}{C}}$ and increasing C decreases the characteristic impedance, where L and C represent the basic filter inductance and capacitance needed for attenuation (see section 4.2.2.1).

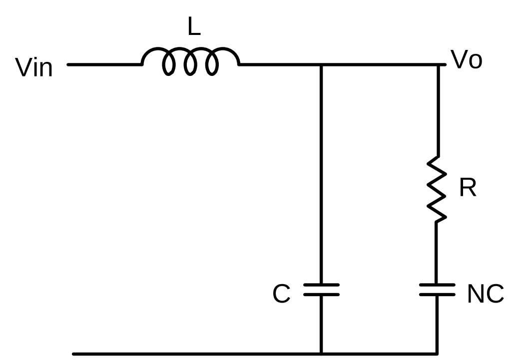
Tables 4-2(a) and 4-2(b) represent the best value damping resistor for a parallel inductor filter and various N values (fig. 4-15(a) and 4-15(b)) to minimize the peak capacitor current and voltage gain respectively. Tables 4-3(a) and 4-3(b) provide the same calculations for parallel capacitors. These resistance values are determined empirically using Spice analysis and resonant current and voltage peak plots for varying resistance values until the lowest resonant peak is achieved. Although increasing N causes smaller resonant peaks, choosing smaller values of N decreases the value needed for the dc current carrying inductor. In general, values of N greater than two cause unreasonable weight and component impacts. Choosing the damping resistor to decrease the resonance by 10 dB is usually sufficient, since the effective series resistance when using tantalum capacitors normally damps the peaks by another 6 dB.



Vin = Input Voltage
Vo = Output Voltage

$L1 = L\left(\frac{N+1}{N}\right)$
L2 = L (N+1)
L is L1 in parallel with L2
N = L2/L1
C = Filter Capacitance
L = Filter Inductance
IC = Current Through C
R = Damping Resistor

Figure 4-16(a). Damping resistor for parallel inductors.



Vin = Input Voltage
Vo = Output Voltage

C = Filter Capacitance
L = Filter Inductance
R = Damping Resistor
N = Integer Number

Figure 4-16(b). Damping resistor for parallel capacitors.

Table 4-2(a). Damping resistor for minimum peak *Ic* (parallel inductors).

N	Best R (minimum Peak Ic)	$\frac{V_O}{V_{IN}}$ @ Peak Ic (dB)	ω @ Peak Ic	$\frac{V_O}{V_{IN}}$ Peak (dB)
1	$\sqrt{\frac{7.2L}{C}}$	9.6	$\sqrt{\frac{1}{1.5LC}}$	9.7
2	$\sqrt{\frac{6L}{C}}$	6.0	$\sqrt{\frac{1}{2LC}}$	6.5
3	$\sqrt{\frac{6L}{C}}$	4.4	$\sqrt{\frac{1}{2.5LC}}$	5.2

Table 4-2(b). Damping resistor for maximum gain (parallel inductors).

N	Best R (minimum Peak $\frac{V_O}{V_{IN}}$)	$\frac{V_O}{V_{IN}}$ @ Max Gain (dB)	ω @ Max Gain	$\frac{I_c \text{ Peak}}{I_c \text{ Best}}$ (dB)
1	$\sqrt{\frac{2.5L}{C}}$	9.5	$\sqrt{\frac{1}{1.4LC}}$	1.03
2	$\sqrt{\frac{3.1L}{C}}$	6.0	$\sqrt{\frac{1}{1.9LC}}$	1.09
3	$\sqrt{\frac{2.7L}{C}}$	4.4	$\sqrt{\frac{1}{2.4LC}}$	1.16

Table 4-3(a). Damping resistor for minimum peak *Ic* (parallel capacitors).

N	Best R (minimum Peak Ic)	$\frac{V_O}{V_{IN}}$ @ Peak Ic (dB)	ω @ Peak Ic	$\frac{V_O}{V_{IN}}$ Peak (dB)
1	$\sqrt{\frac{7.2L}{C}}$	9.6	$\sqrt{\frac{1}{1.5LC}}$	9.7
2	$\sqrt{\frac{0.8L}{C}}$	6.0	$\sqrt{\frac{1}{2LC}}$	6.6
3	$\sqrt{\frac{0.5L}{C}}$	4.5	$\sqrt{\frac{1}{2.5LC}}$	5.3

Table 4-3(b). Damping resistor for maximum gain (parallel capacitors).

N	Best R (minimum Peak $\frac{V_O}{V_{IN}}$)	$\frac{V_O}{V_{IN}}$ @ Max Gain (dB)	ω @ Max Gain	$\frac{I_c \text{ Peak}}{I_c \text{ Best}}$ (dB)
1	$\sqrt{\frac{3L}{C}}$	9.5	$\sqrt{\frac{1}{1.5LC}}$	1.02
2	$\sqrt{\frac{1.5L}{C}}$	6.0	$\sqrt{\frac{1}{2LC}}$	1.07
3	$\sqrt{\frac{1.1L}{C}}$	4.4	$\sqrt{\frac{1}{2.7LC}}$	1.13

4.5.3 Retrofit Fixes

Options available after a CS01/CS02 failure has occurred involve component changes to those with higher voltage and current ratings, or addition of damping resistors (described in section 4.5.1) if it is feasible to alter the board layout.

4.6 Conducted Transient Susceptibility (CS06)

The CS06 test controls the susceptibility of the EUT to transient spikes injected on its ungrounded input power leads.

4.6.1 Design Considerations

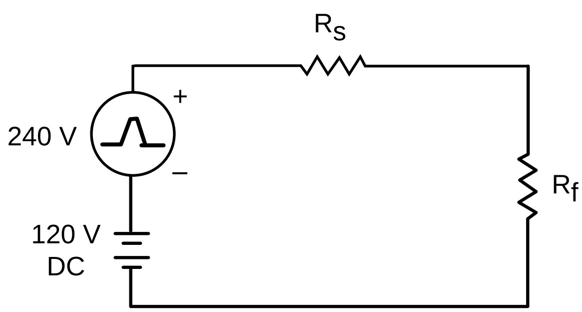
A major consideration in the CS06 test is ensuring that the front-end filter components are rated for the voltage of the CS06 tests. The usual voltage level for a CS06 test is two times the nominal line voltage.

Another important consideration is the source impedance of the signal source. In most military and NASA CS06 testing, the source impedance is on the order of 0.5 Ω. Having this low source impedance puts an additional burden on the designer to have a front-end component or filter with an equivalent resistance (on the same order of the source resistance) to reduce the voltage spike in the sensitive circuitry. This is especially difficult in high voltage systems since the test voltage is twice nominal voltage.[4-6]

4.6.2 Modeling/Prediction Techniques

Simplified modeling is useful in determining the value of equivalent resistance. "The power source is simplified to an ideal voltage source in series with a resistive and/or inductive impedance."[4-6] Input circuitry is normally designed to withstand some voltage level over nominal voltage. This maximum voltage level should be known. For instance, in a 120-V system with a 240-V applied transient, the equipment may have components able to withstand a total voltage of only 160 V. To lower the voltage spike of the CS06 test to this level, the input filter component must represent a sufficiently low equivalent resistance to load down the spike. Referring to figure 4-17, a simple ratio is used to calculate the desired equivalent resistance:

$$\frac{R_f}{(R_f + R_s)} = \frac{40}{240}$$



R_s Is Source Resistance
R_f Is Equivalent Filter Resistance

Figure 4-17. CS06 test circuit model.

The number 40 represents the designed voltage level of the input circuitry over the nominal voltage (160–120 = 40). Solving this equation gives the following ratio:

$$\frac{R_s}{R_f} = 5 \quad .$$

If the source ratio is small, 0.5 Ω, it is difficult to design the filter with this low equivalent resistance. However, for higher source resistances or lower voltage systems, this requirement is more feasibly met.

4.6.3 Retrofit Fixes

As with CS01/CS02, retrofit fixes involve obtaining higher voltage-rated components or changing the input filter design (may be difficult after the design is complete).

4.7 Radiated Susceptibility (RS03)

RS03 (14 kHz to 10 GHz) is a requirement that controls and determines the susceptibility of the EUT to radiated electric fields.

4.7.1 Design Considerations

"An external field can couple either directly with internal circuitry and wiring in DM or with cables to induce a CM current."[4-7] By far the most common entry point of external RE is cabling. Inadequate filtering and/or shielding on I/O lines dominates to at least 10 V/m. This is a direct result of the PCB traces being physically shorter than the I/O cables. In addition, direct coupling through the enclosure, assuming aluminum construction, is unlikely. Methods of cable shield termination heavily influence RS. For additional information on cable shielding and termination see section 3.4. For information on controlling loops in board design see section 3.2.2.

4.7.2 Modeling/Prediction Techniques

Predicting RS of a particular piece of equipment begins with understanding potential exposure to fields. Protection against these fields is implemented the same way that RE from the equipment are controlled. The radiated environment will likely be defined in the EMC specifications on a particular program. These limits are reviewed to understand the degree of needed additional shielding. For a payload in Spacelab, in which the module shielding limits the exposure to fields, the threat of RS is less than for a space station externally mounted payload. Exact calculation of box susceptibility is very cumbersome, but for high electromagnetic fields, if the cable lengths approach $\lambda/2$, additional shielding is necessary.

4.7.3 Retrofit Fixes

Retrofit fixes for RS failures are often limited to adding shielding or improving the connector type for shield termination (back-shell termination versus pigtail). Since RS requirements are written for the system in general and are not specific to the location of each piece of equipment, the location of the box is a consideration in determining if the equipment design needs altering. Also, there may be racks or other metallic barriers between the transmitters of concern and the equipment. This rationale is considered in determining the necessity of changing equipment design.

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4-6. Javor, K.: “Specifying Control of Immunity to Powerline Switching Transients.” IEEE EMC Symposium, August 1994.

4-7. Williams, T.: “EMC for Product Designers.” Butterworth-Heinemann Ltd., Oxford, England, 1992.

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5. DIAGNOSTIC/TROUBLESHOOTING/DESIGN SUPPORT ELECTROMAGNETIC INTERFERENCE TESTING

5.1 Introduction

This chapter explains how to perform special test techniques useful in the design process or when full compliance testing reveals specification noncompliances. Only section 5.2, pertaining to CE testing, is adaptable for pretest, design-to-compliance usage. The balance of the techniques involves EMI test-unique equipment not applicable for use outside an EMI test facility. Since power filter design is an integral part of equipment design, section 5.2 is the most detailed. Sections 5.3 through 5.5 give troubleshooting techniques that the equipment designer may ask the EMI test technician to perform in order to determine the source of the noncompliance.

5.2 Diagnostic Testing for Conducted Emissions

Compliance testing for NASA CE measurements involves the use of several items of test equipment not ordinarily available outside an EMI test facility. These include tunable EMI meters (spectrum analyzers or receivers), current probes, and a specialized feed-through capacitor used as a standardized source impedance. Figure 5-1 shows a typical setup.

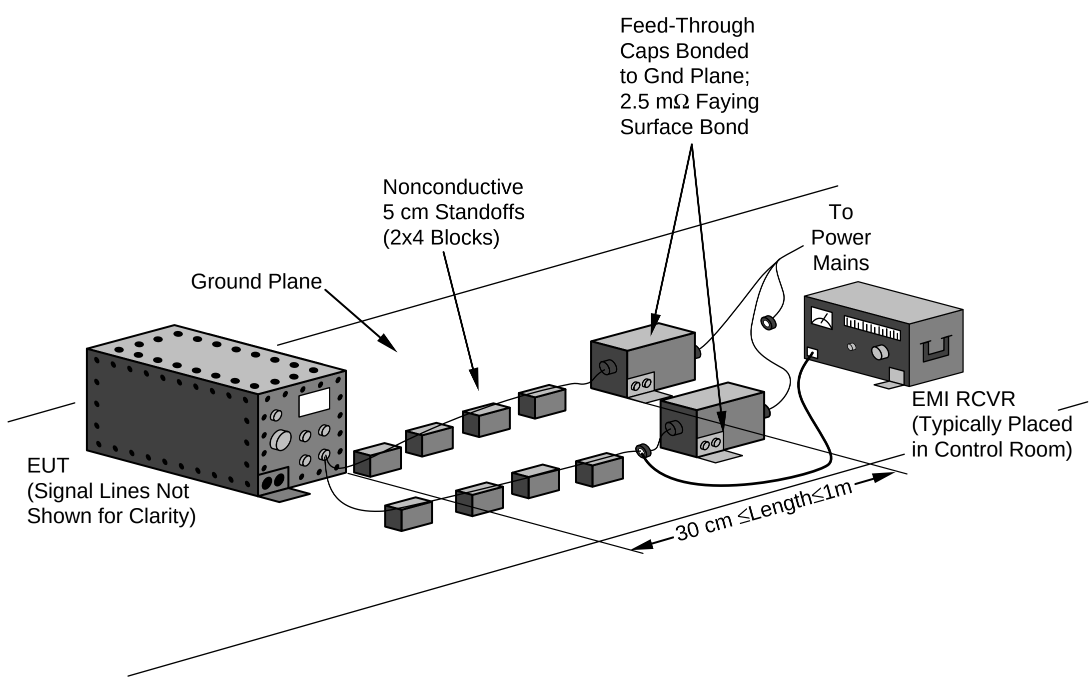


Figure 5-1. Full compliance current CE test setup.

A useful diagnostic CE test (fig. 5-2) uses an ordinary benchtop digital oscilloscope (with either built-in FFT capability or computer interface) to replace the tunable EMI receiver, and either the feed-through capacitor or a high quality replacement to standardize the power source impedance. In order to get the necessary sensitivity, a fairly efficient current probe is necessary, and it is convenient if its voltage output versus current is constant over the frequency range of the CE requirement. The feed-through capacitor is available from several vendors at a reasonable cost (under $200 each, two required). Alternative capacitors described below may be substituted.[1] The current probe used for the test described below had a constant 0.7-V/A output over a frequency range from 15 kHz to above 50 MHz. A sufficiently sensitive oscilloscope current probe might be suitable if the window is large enough for the power conductors and the saturation current is higher than that used by the equipment under test. Otherwise, a wide variety of EMI current probes is available for approximately $500. The advantage of the diagnostic test is lower cost because the designer usually has a digital oscilloscope available and need not procure a spectrum analyzer (low-cost diagnostic analyzer ~$7 to $16 K). The disadvantage is that time-domain data are not easily compared to frequency-domain specifications, hence less useful information is garnered.[2] However, increasing availability of computer interfaces on low-cost oscilloscopes and of inexpensive FFT software running on ubiquitous desk top PC's give a path toward overcoming this disadvantage.

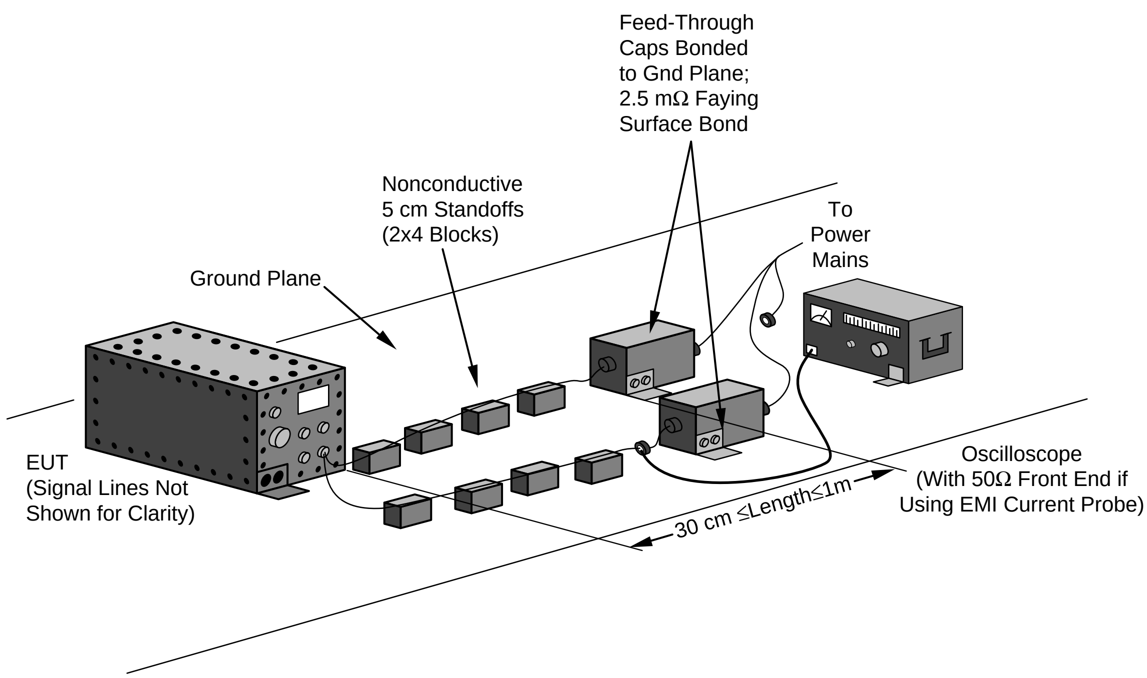


Figure 5-2. Diagnostic CE test setup.

CE test results from a switched mode power supply (SMPS) using FFT oscilloscope measurements are presented. A diagnostic routine resulting in a CE compliant power supply filter combination is flowcharted. Finally, the FFT data are compared to test data taken in the full compliance mode, using a spectrum analyzer. This comparison is to give the user a feel for the accuracy of the diagnostic test.

If use of alternative capacitors and current probes is desired, see the evaluation described in section 5.2.1.

[1]Feedthrough capacitors are available from: Captor 513/667-8484, Fischer Custom Communications 310/644-0728 and Solar Electronics 800/952-5302.

[2] The frequency domain information is necessary both in comparing outages to the specification limit, and also to help in designing an EMI filter to eliminate the outage condition.

5.2.1 Evaluation of Measurement Equipment

The capacitor is evaluated by measuring insertion loss in a 50-Ω signal source/receiver circuit (fig. 5-3). The insertion loss requirement for the standard feed-through capacitor is shown in figure 5-4. In order to meet the curve in figure 5-4, it is important to minimize capacitor lead lengths when performing the test in figure 5-3, and hence, equally important when performing the actual conducted emission test.

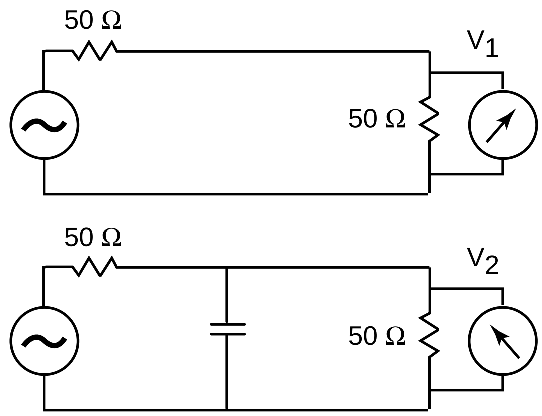


Figure 5-3. Measurement of capacitor insertion loss.

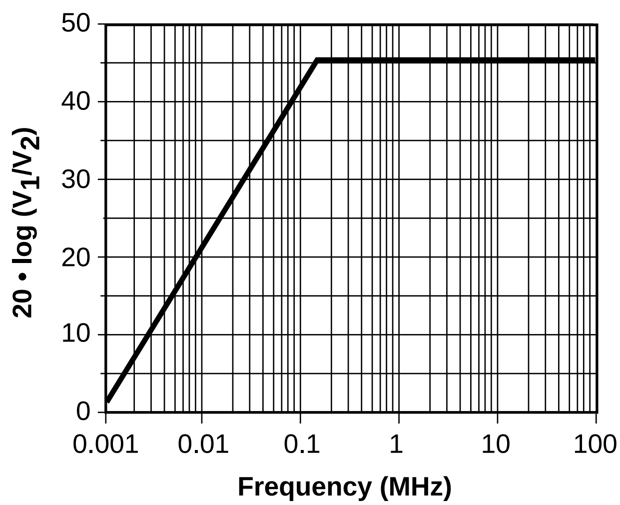


Figure 5-4. Insertion loss requirement on line impedance standardizing capacitor.

One method of simulating the standard 10-μF feed-through capacitor is to mount leaded capacitors on a double-sided printed circuit board (PCB) (fig. 5-5(a)). Different type capacitors are necessary to achieve both the low-frequency 10 μF and the required high-frequency performance. The ground side of the board is connected to the ground plane via a flexible strap of the same width as the PCB. The performance of figure 5-5(b) was achieved only by building a parallel plate capacitor of the grounding strap. Transparent plastic tape was laid on top of the ground strap (conductive tape) and covered with more conductive tape attached to the hot side of the PCB. This high-quality capacitance was not very repeatable, and variations in insertion loss were observed from day to day. The above assembly has the advantage that all parts required are available at a local electronics outlet for less than $30.00. However, a simpler approach is to minimize the cost of the high-quality feed-through capacitor by buying only the voltage and current rating necessary, and to reduce packaging costs by building two capacitors into one housing. A 10-μF feed-through capacitor package (rated at 120 Vac, 400 Vdc, and 10 A) is available for approximately $125.00.[3]

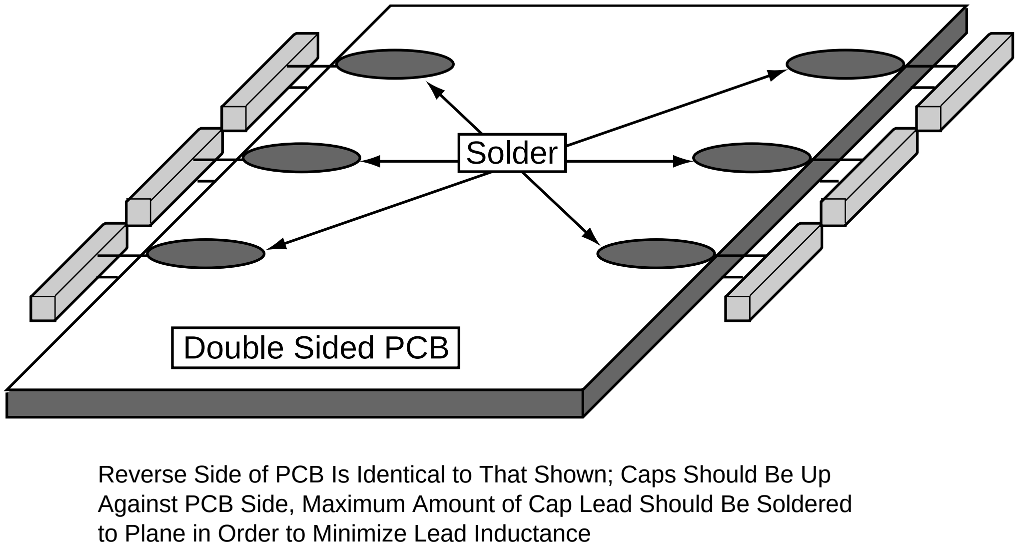


Figure 5-5(a). Possible low-cost construction of a line impedance standardizing capacitor assembly.

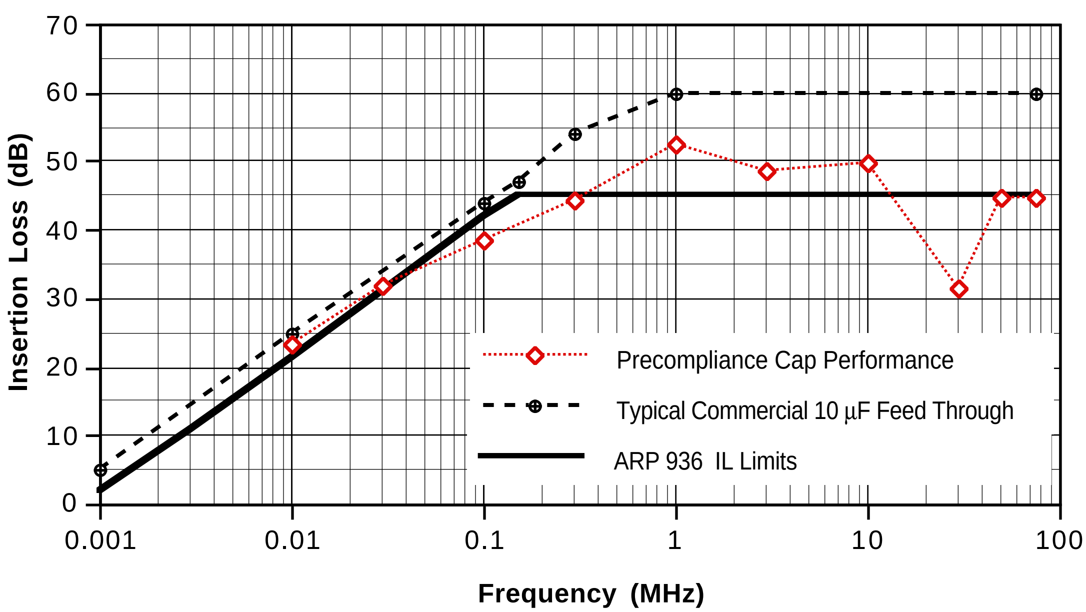


Figure 5-5(b). Performance of capacitor assembly of figure 5-5(a).

[3] Captor Corp., Scott Timms, 513/667-8484.

5.2.2 Conducted Emission Testing

A CE test is performed per figure 5-2. The current probe sequentially measures each line. For high-frequency fidelity, the power-line length is minimized between the capacitor power output connector and the equipment-under-test (EUT) power input connector. Capacitors must be bonded to a ground plane with a low impedance (rf) bond; this means metal-to-metal faying surface bonds. The EUT bond should replicate that achievable in the intended installation. With the EUT operational, voltage measured at the spectrum analyzer, adjusted for the current probe transfer impedance, is directly compared to the specification limit. Or, raw data is compared to the current limit adjusted for the current probe transfer impedance. Either the transfer impedance (dBΩ) is subtracted from the raw data or is added to the current limit (dBμA) to yield the dBμV adjusted limit. When using an oscilloscope, an FFT algorithm must be used to convert the time-domain ripple into frequency-domain data. Because the oscilloscope displays the frequency axis linearly, it is necessary to perform two sweeps at different speeds, one to capture the SMPS fundamental frequency and one other to catch the harmonics.[4] When using a spectrum analyzer, follow manufacturer operation instructions. Be careful to assure that measured noise is due to the EUT and is not a power-line ambient or, especially with an oscilloscope, internal instrument noise. If using a homemade capacitor assembly of limited bandwidth, try limiting the oscilloscope bandwidth appropriately in order to lower the internal oscilloscope noise. Section 5.2.3 on CE filter design goes into more detail on CE testing.

5.2.3 Power-Line Conducted Emission Filter Design

SMPS's generate two types of CE's, designated by the paths they follow: differential mode (DM) and common mode (CM).

5.2.3.1 Differential Mode Emissions

DM noise is the simplest kind. DM noise current flows in the same path and direction as the power frequency current (fig. 5-6(a)).

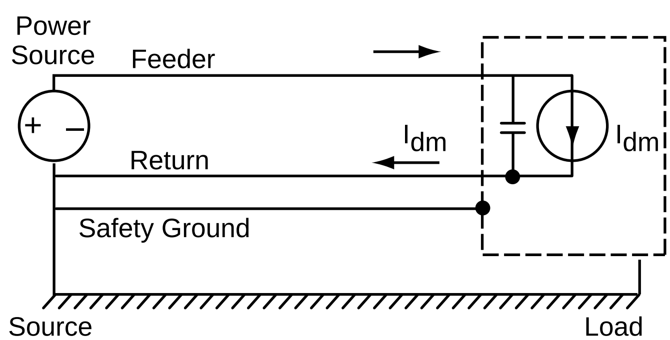


Figure 5-6(a). Single-phase DM noise source.

DM noise, referred to as the normal or longitudinal mode, is characterized by the currents flowing in the feeder and return lines 180° out of phase.

[4]However, the time duration of the record determines the resolution bandwidth, so it cannot be arbitrarily selected. More detailed instructions are presented later.

5.2.3.2 Common Mode Emissions

CM noise is characterized by the noise currents flowing in phase in the feeder and return lines. This propagation mode is shown in figure 5-6(b). When the equipment chassis is isolated from the reference plane, a parasitic capacitance (on the order of tens of pF) is in series with the return path through the reference plane. This high-impedance capacitance results in the safety ground appearing as the lowest-impedance return path and carrying most of the current.

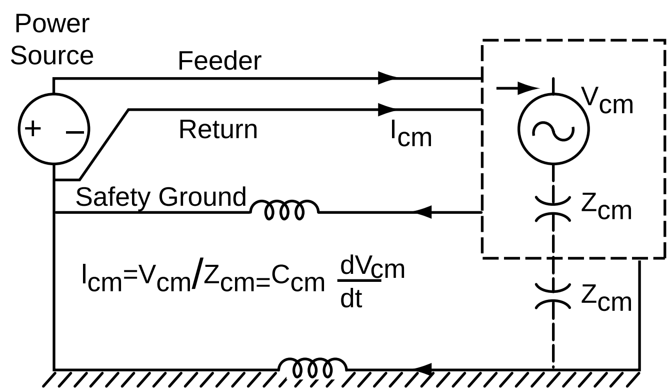
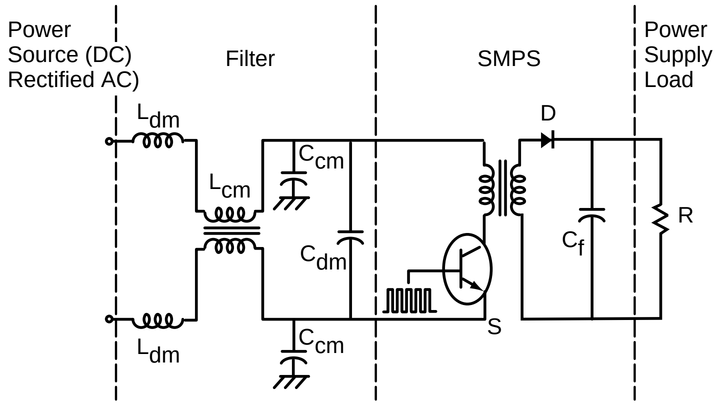


Figure 5-6(b). Single-phase CM noise source.

Reference 5-1 explains in great detail the sources of both DM and CM CE in SMPS. The following discussion is a very brief summary of this reference. SMPS's generate DM CE drawing pulsed currents from the power source. This is an intentional operation of the SMPS. Filtering provided to meet EMI limits cannot materially reduce the SMPS pulsed-current draw; this would adversely affect the performance of the SMPS. Filtering can only provide a local low-impedance current source (bulk storage capacitor) and a small amount of line inductance to force even less current draw from the power line than would be accounted for by LISN/bulk storage capacitor current division.

In contrast, SMPS's generate CM CE via parasitic capacitances between high-voltage, switched-current elements and the power system reference. CM CE are filtered almost without regard for the effect on SMPS performance. Only such secondary effects as leakage inductance causing CM choke saturation or ac leakage current in line-to-ground "Y" capacitors are taken into account.

DM CE are contained by using a line-to-line or "X"-type capacitor providing a low-impedance current source for the power supply switching circuitry and high-impedance inductors facing the power source to raise the impedance of the power source at EMI frequencies. CM CE are contained by using line-to-ground capacitors, "Y"-type, shunting the power supply switching element parasitic capacitances, and a CM inductor facing the power source to raise the impedance of that path. Figure 5-7 illustrates the parts of an EMI filter.



Filter Elements

C_{dm} Line–to–line, or “X” installation large value capacitor, typically electrolytic, provides low source impedance for SMPS; also provides hold up during power surge/sag.

C_{cm} Line–to–ground, or “Y” capacitor for containing common mode currents inside equipment, Y capacitors shunt the parasitic capacitance between the case of switch S and equipment chassis.

L_{cm} Common mode choke, typically presents on the order of 1mH inductance to common mode currents, raising the impedence of this path in order to make the internal CM path more effective.

L_{dm} Differential mode inductor, typically tens of microhenries, raises the impedance of the power source at EMI frequencies and makes the X capacitor a more effective decoupling mechanism.

Power Supply Elements

S Transistor switch whose operation changes power supply input voltage from DC to AC. Frequency and/or duty cycle of switch may be varied by power supply control loop whose function is to supply a fixed output voltage to its load, regardless of load changes or input voltage variation. Parasitic capacitance developed between the case of this switch and equipment chassis is a primary source of common mode noise currents.

T Transformer which provides a power supply secondary side voltage different than primary side. This transformer is much smaller than a 60 or 400 Hz transformer rated for the same power, because it is designed for use at the power supply switching frequency (typically above 20 kHZ).

D Rectification diode to yield DC ouput for secondary.

Cf Secondary filter capacitor. This capacitor is much smaller than a capacitor designed to filter 60 or 400 Hz ripple, because it is designed for use at the power supply switching frequency (typically above 20 kHz).

Figure 5-7. SMPS and filter.

5.2.3.3 Discussion of Conducted Emission Test Procedures [5-2]

NASA CE limits are based on predictions of power bus ripple. Voltage ripple specified in the time domain by the electrical power provider is converted into the frequency domain. This voltage is divided by the power bus wiring impedance to generate a current limit for current type requirements. These limits are based on DM current flow. A typical limit is shown in figure 5-8.

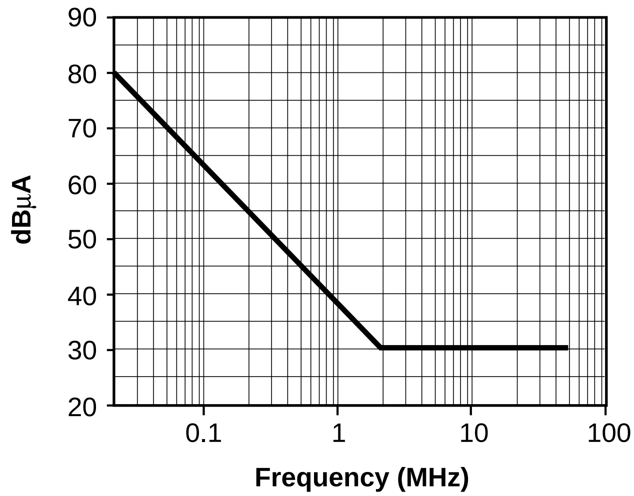


Figure 5-8. Current CE limit, MSFC-SPEC-521B.

The important point of the above discussion is that a single LISN is used to model the source impedance. Although it is a two-wire LISN, only DM current flow is considered. Even though NASA limits are based on DM concerns, the test procedure requires the use of a separate feed-through capacitor in each current-carrying power wire. The present military EMI standards, MIL-STD-461D and MIL-STD-462D, require two 50-μH LISN's, as the commercial standards do. NASA CE testing at the initial release of this handbook still requires the use of feed-through capacitors and current probes, requirements based on MIL-STD-461A/B/C and MIL-STD-462, Notice 2, references 5-3 and 5-4.

It is the use of a two-LISN/feed-through capacitor test setup that makes differentiation between DM and CM CE important. Figure 5-9, showing CE current flow with current return through ground, is compared and contrasted with the two-wire above-ground setup of figure 5-10. In a single-LISN/feed-through capacitor test setup (which models current return on structure), all noise currents flow in the same path, and, while the source impedance of the various sources of CE is quite different, the topology of the EMI filter is the same for all noise sources. In fact, this topology was standardized by the military, under MIL-F-15733.[5-5] The MIL-F-15733 filter topology is summarized as inductance (if used) in series with the power feeder and as capacitance from feeder to ground (not power return). Figure 5-11(a) shows a MIL-F-15733-type filter installed in an

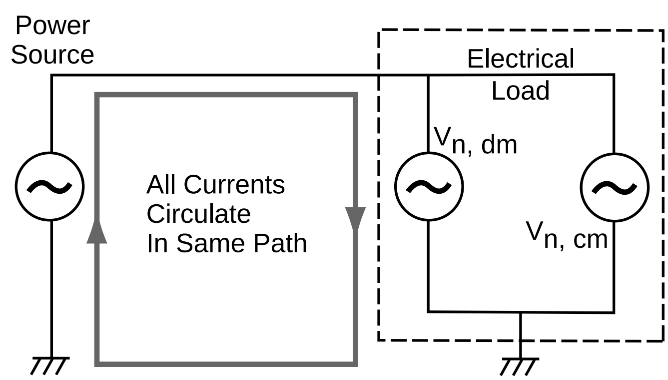


Figure 5-9. Noise current circulation in structure return bus.

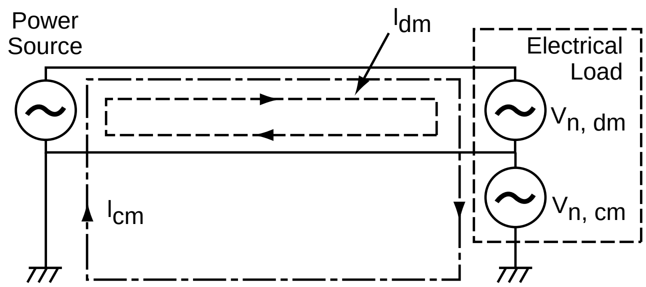


Figure 5-10. Circulation path of noise currents in above-ground current return bus.

equipment with a grounded power return (objective of MIL-F-15733 design). Figure 5-11(b) shows two filters of the type shown in figure 5-11(a) installed in equipment using ungrounded power return. Contrast figure 5-11(b) with figure 5-7, showing a filter optimized for ungrounded power return. Figure 5-7 incorporates two features impossible to provide in a MIL-F-15733 filter: large line-to-line "X"-type capacitor and a CM choke. The X-capacitor provides the low-impedance current source for the SMPS. The CM choke provides as much as a millihenry of CM inductance, whereas the MIL-F-15733-type filter must limit its DM inductance to less than 100 μH, due to core saturation, size, and filter impedance constraints.

Because filter topology is important in the ungrounded power lead configuration (modeled by the two LISN setup[5]), it is important for the filter designer to accurately isolate and measure DM and CM CE. No compliance standard addresses this issue. In the balance of this section, test data are presented showing CM and DM emissions above the limit.

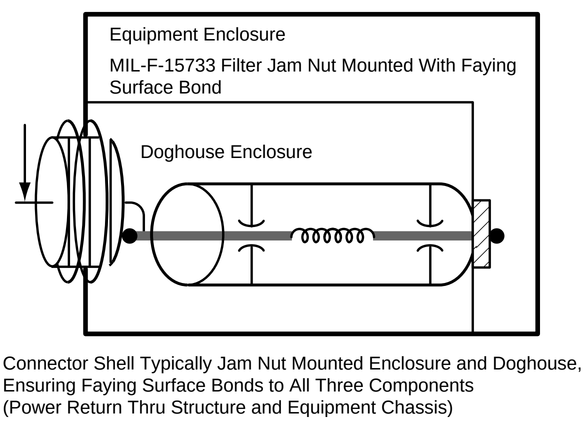


Figure 5-11(a). Installation of single MIL-F-15733 EMI filter in equipment using structure for power current return.

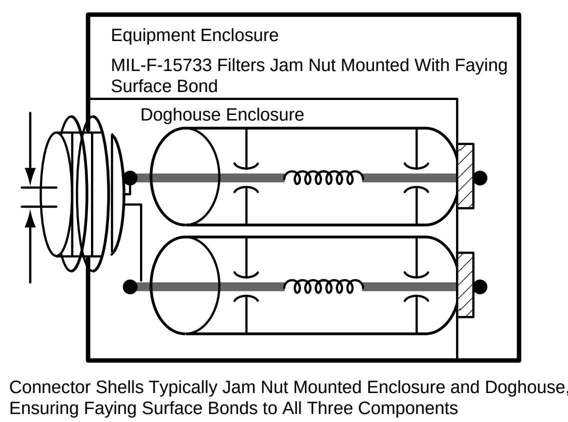
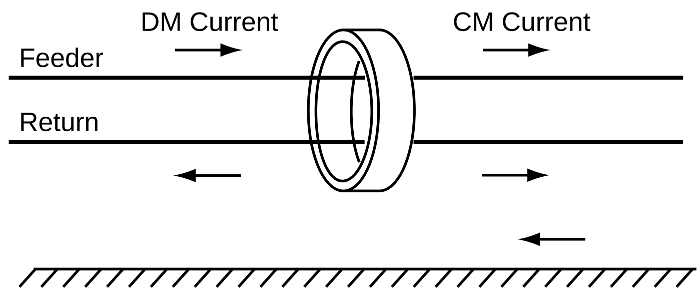


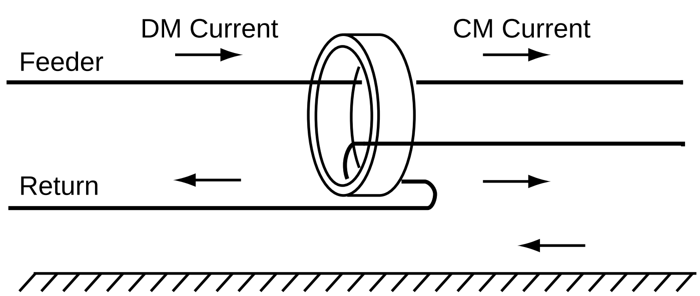
Figure 5-11(b). Installation of MIL-F-15733 filters in equipment using above-ground current return.

[5]It should be noted that using two LISN's to model above ground current return is not very accurate. Consider the DM source impedance presented by two LISN's. It is the sum of the series impedances, or 100 Ω at high frequencies. The CM impedance is each LISN in parallel with the other, or 25 Ω. The DM impedance is four times the CM impedance. A typical two-wire line (black and white) would have a much lower DM than CM source impedance, while a three-wire line (black, white, and green) might have roughly equal CM and DM impedances.

Measuring CE from the LISN/feed-through capacitor, it is quite possible that installation of a filter element resulting in excellent rejection of one mode would not be recognized, due to predominance of the unrejected mode. An erroneous conclusion that the filter element was ineffective could be reached. When designing to MIL-STD-461C and MIL-STD-462, and derivative specifications (NASA), a current probe is used to separate modes. Measurement of CM and DM currents using current probes is diagrammed in figure 5-12(a) and (b). Mode isolation techniques make possible the algorithmic approach to filter design in sections 5.2.3.4 through 5.2.3.6.



(a) CM Current Measurement (DM Rejection)



(b) DM Current Measurement (CM Rejection)

Figure 5-12. Mode selection/rejection with current probes.

5.2.3.4 Filter Design Troubleshooting Flowchart

Troubleshooting includes the following four steps:

(1) Measure emissions on each line per standard compliance techniques (case history plots figures 5-13.1(a) through 5-13.1(d)). Assuming that outages exist, place current probe around both feeders per figure 5-12(a).

(2) Measure emissions from the current probe (case history plots figures 5-13.2(a) and 5-13.2(b)). If outages still exist, they are guaranteed to be CM. If no outages exist, the excessive emissions must be DM. Skip to step 4.

(3) Employ CM filtering techniques (Y-capacitors and CM choke) to bring CM emissions to the desired level. The CM isolation technique is very useful in checking the effectiveness of CM components; in other words, for filter optimization. In the following case history, the CM filter was

first implemented as Y-capacitors alone (data plot fig. 5-13.3); this was helpful but insufficient; the CM choke was required (data plot fig. 5-13.4). Conversely, sometimes a filter is still effective with much less attenuation; this could lead to cost and/or space/weight savings.

(4) Reconnect EMI receiver/analyzer/FFT oscilloscope to the current probe in the standard CE compliance setup. If no outages exist, the filter is a success. If outages still exist, they must be DM CE, and DM techniques are confidently employed to reduce the emissions (X-capacitors and DM inductors). The same optimization rationale applies here; in the absence of CM noise, the DM section is finely tuned to provide just the right amount of attenuation. In the case history, an X-capacitor was added first (data plots figs. 5-13.5(a) and 5-13.5(b)), with a reduction in CE but not limit compliance; the addition of a DM inductor brought compliance (data plots figs. 5-13.6(a) and 5-13.6(b)).

(5) Special Instructions on the Use of an FFT Oscilloscope—DSO's with built-in FFT capability are much easier to use than those requiring a PC to perform the processing. This is because the conversion is almost real time, and the effect of built-in windowing functions is easy to assess. Time-domain windows are important when the periodicity of the waveform is not clear and the recorded sample period cannot be adjusted to be an exact multiple of the waveform period. The Hamming window function is typically available, the Blackman window is better. Both of these window functions trade frequency for amplitude accuracy. This is exactly what is desired for a pre-compliance or diagnostic scan. If windows are not available, then the test engineer must observe the waveform and select an integral number of periods for processing. Regardless of whether the processing is performed onboard or after data have been ported to a PC, the record length must support the resolution bandwidth desired. In some DSO's, record length may be limited by memory or available resolution (number of data points taken when porting the display to memory). Long record lengths then will limit the achievable frequency scan. Memory of 50 kbytes or more is sufficient for most cases. If the memory is much less, then multiple scans have to be performed at various sweep speeds. Record length must be the reciprocal of the specification resolution bandwidth (RBW), or that necessary to resolve spectral components. For a scan above 100 kHz, a record length of 100 μs should be sufficient (10 kHz RBW). In general, the RBW should be no more than one-tenth the lowest tuned frequency. However, for diagnostic purposes, the RBW may be selected to be one-half the SMPS switching rate; this allows for resolution of one spectral component from another. The faster the switching, the shorter the record length necessary. In some cases, especially having to do with short-duration time-domain waveforms occurring at waveform leading and trailing edges, it is difficult to record an entire period with the necessary sweep speed to accurately record the waveform. If the record length is less than the period of the waveform, the FFT algorithm overestimates the spectral content by the ratio of the waveform period to the record length. For more detailed information on this topic, see K. Javor, "Measurement of Frequency Domain Conducted Emissions Using An Oscilloscope," 1995 EMC/ESD International Symposium Record. That reference evaluates the conducted emissions of the same SMPS evaluated herein, but using an entry-level oscilloscope with much more limited capabilities than that used for this investigation.

5.2.3.5 Filter Design Case History

CE measurements were made on a switched-mode power supply. This is the same power supply investigated in reference 5-1. Baseline measurements of the unfiltered supply are compared to typical CE limits in figures 5-13.1(a) through 5-13.1(d). The current probe used for this test was chosen to have a flat response curve over the frequency range 15 kHz to 50 MHz. The transfer impedance is 0.7 V/A (–3 dBΩ). The limit used is from MSFC-SPEC-521B CE03 and shown in figure 5-8. The adjusted limit penciled on the test data is then 3 dB below the current limit, but the units are dBμV. Since data points are transferred from a log-log graph to a semilog graph, the penciled limit line is not entirely accurate. However, desired accuracy is increased by transferring more data points.

Figures 5-13.2(a) and 5-13.2(b) are CM test data from the unfiltered supply, identical in configuration to figure 5-13.1 measurement, but uses CM isolation to reject DM emissions. Note: All CM emissions above the noise floor are on the higher frequency plot (6.25 MHz per horizontal division). Further CM plots will be taken only at this setting. The limit drawn is directly from figure 5-8. Since both wires are measured together in the CM test, the limit should be relaxed 6 dB from that shown.

Figure 5-13.3 shows the effects of CM filtering using two 1,000-pF Y-capacitors between each converter input and ground. Figure 5-13.3 is comparable to figure 5-13.2. The Y-capacitors greatly improved the high-frequency CM emissions but the lower frequency CM emissions require further reduction.

Figure 5-13.4 measures the performance of a CM filter including an added CM choke (six turns AWG 20 on a Supermalloy™ core). Figure 5-13.4 is comparable to figures 5-13.2 and 5-13.3. The effectiveness of the CM filter is clearly seen at all frequencies. Because CM emissions are below the limit, it is guaranteed that any further individual conductor-based measurements above limits are DM in nature. Begin DM filter design now.

Figures 5-13.5(a) and 5-13.5(b) show performance of combined CM and X (DM) 20-μF capacitor connected between the converter inputs. The converter fundamental switching frequency and harmonics are reduced below the specification limit. Because the fundamental is still close to the limit, another filter stage was added. The total DM filter consists of the 20-μF X-capacitor and a 100-μH DM choke in the the 28-Vdc conductor. Figures 5-13.6(a) and 5-13.6(b) show total elimination of DM emissions. Figures 5-13.7(a) through 5-13.7(d) oscilloscope plots indicate complete compliance. Figures 5-13.8(a) through 5-13.8(d) are a final verification check using an EMC spectrum analyzer. Figure 5-14 shows the completed power supply/filter schematic.

5.2.3.6 Conclusion

SMPS filter design via conduction mode isolation is pursued in a logical manner devoid of guesswork. Because the contribution of each filter element is clearly visible, the method presented herein lends itself nicely to filter optimization.

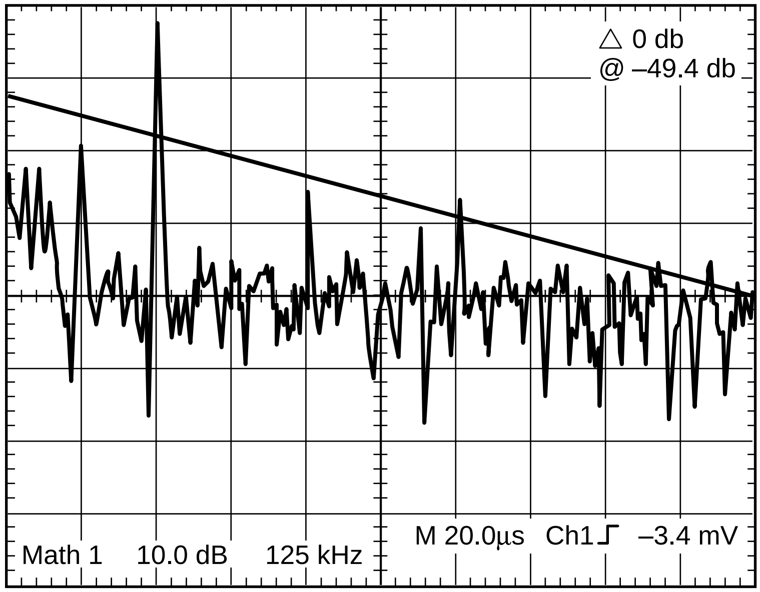


Figure 5-13.1(a). Baseline measurements on unfiltered SMPS, 28 Vdc, low frequency (Y-axis 10 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

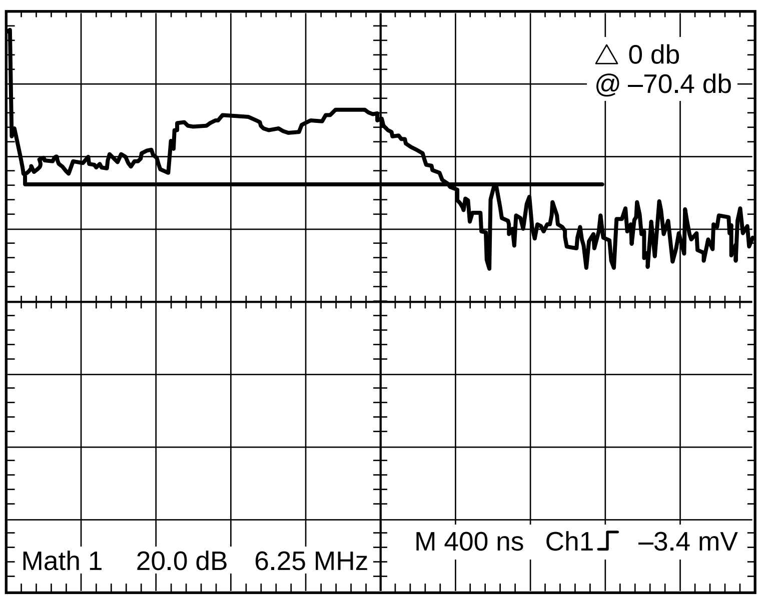


Figure 5-13.1(b). Baseline measurements on unfiltered SMPS, 28 Vdc, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

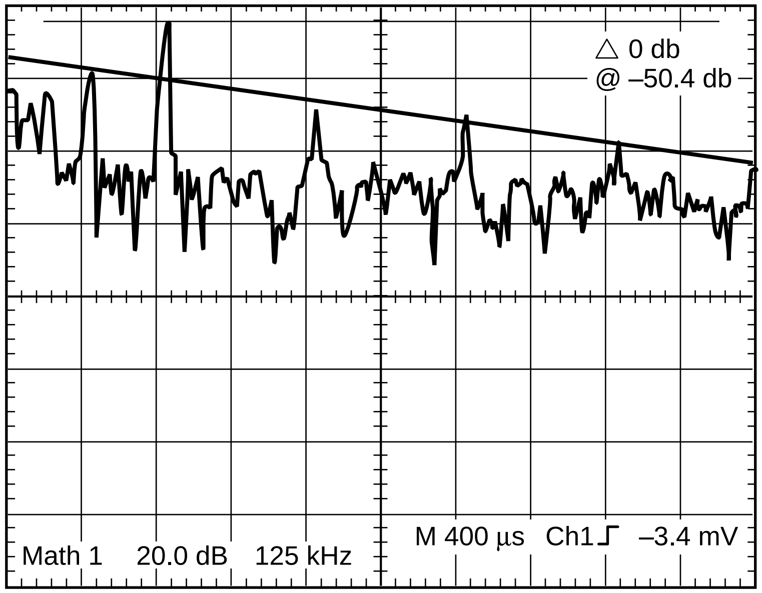


Figure 5-13.1(c). Baseline measurements on unfiltered SMPS, 28 VRTN, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

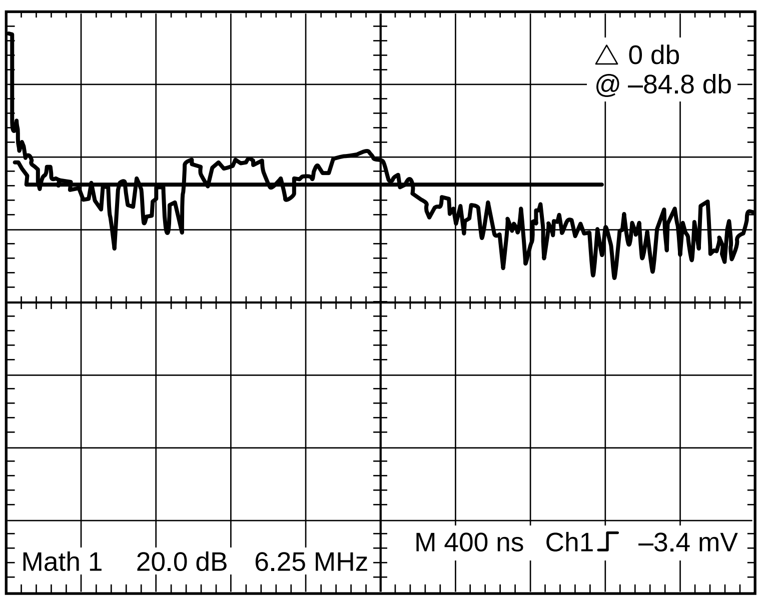


Figure 5-13.1(d). Baseline measurements on unfiltered SMPS, 28 VRTN, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

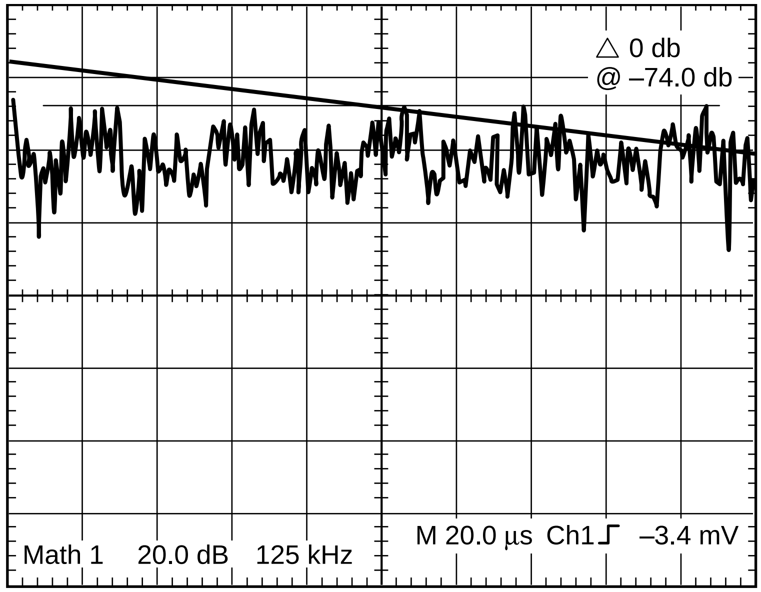


Figure 5-13.2(a). Baseline measurements on unfiltered SMPS, CM CE, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

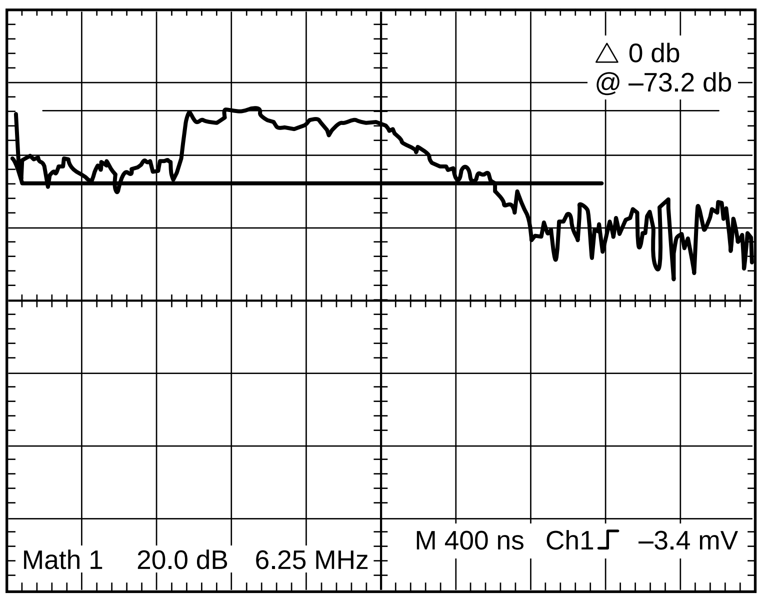


Figure 5-13.2(b). Baseline measurements on unfiltered SMPS, CM CE, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

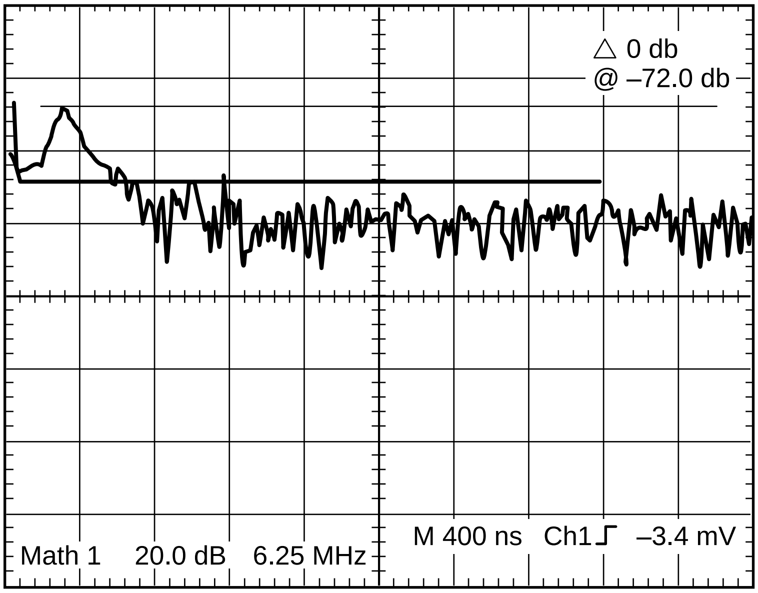


Figure 5-13.3. CM filtering: 2,000 pF Y caps installed, CM data, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

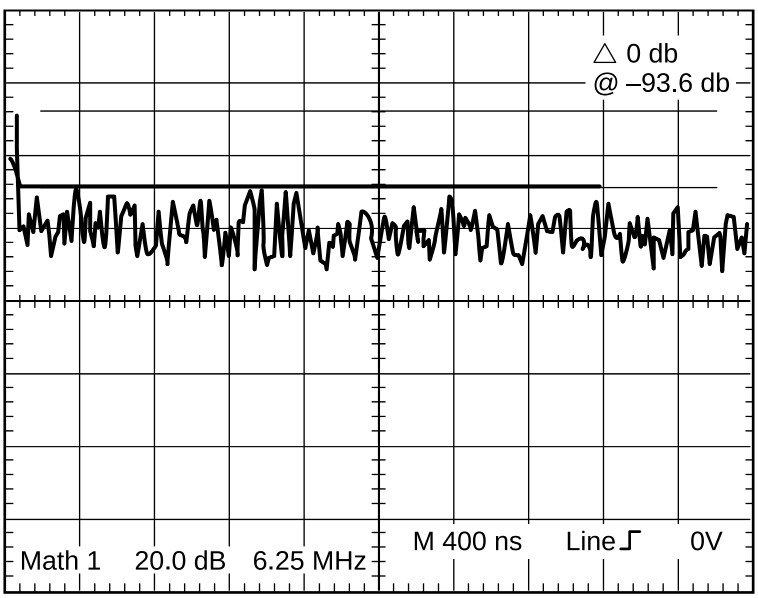


Figure 5-13.4. CM filtering: CM choke installed in addition to 2,000 pF Y caps installed, CM Data, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

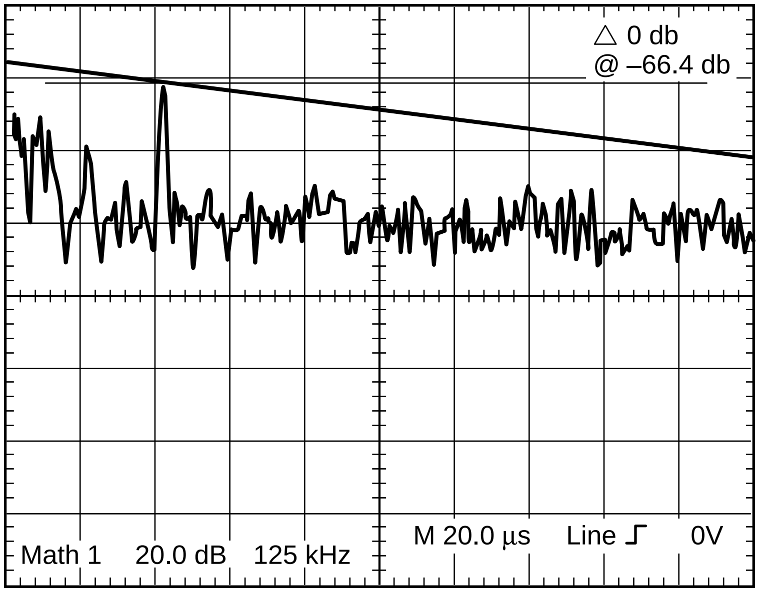


Figure 5-13.5(a). CM filtering: as above plus 20-μF line-line X-capacitance, for DM filtering, 28-Vdc input, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

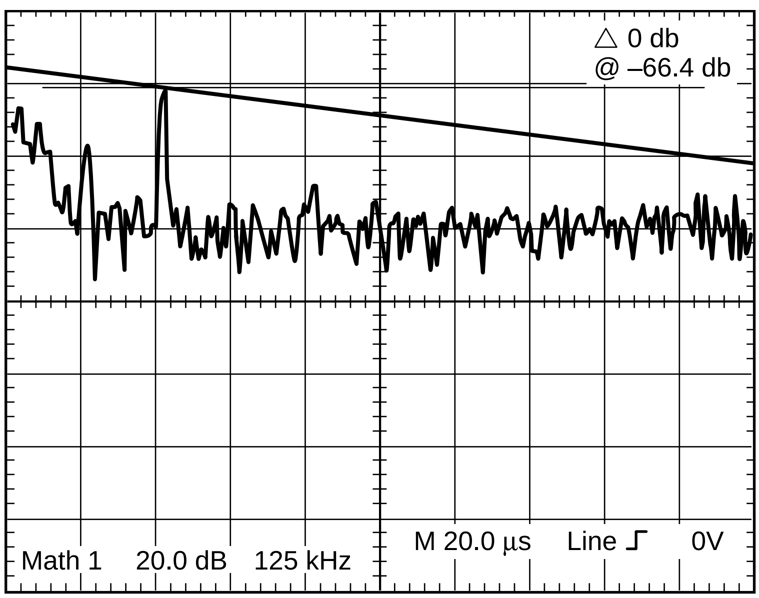


Figure 5-13.5(b). CM filtering: as above plus 20-μF line-line X-capacitance, for DM filtering, 28 VRTN, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

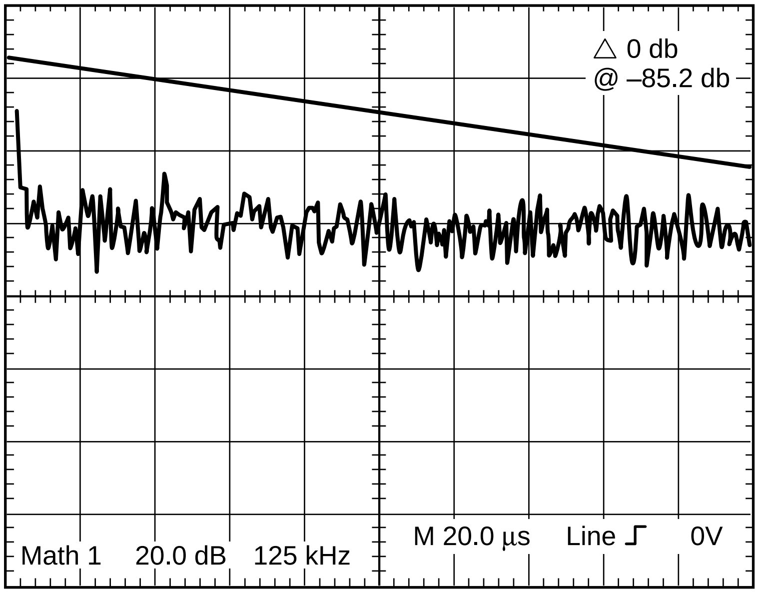


Figure 5-13.6(a). CM filter plus complete DM filter; as above plus 100-μH choke in 28-Vdc Line, 28-Vdc input , low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

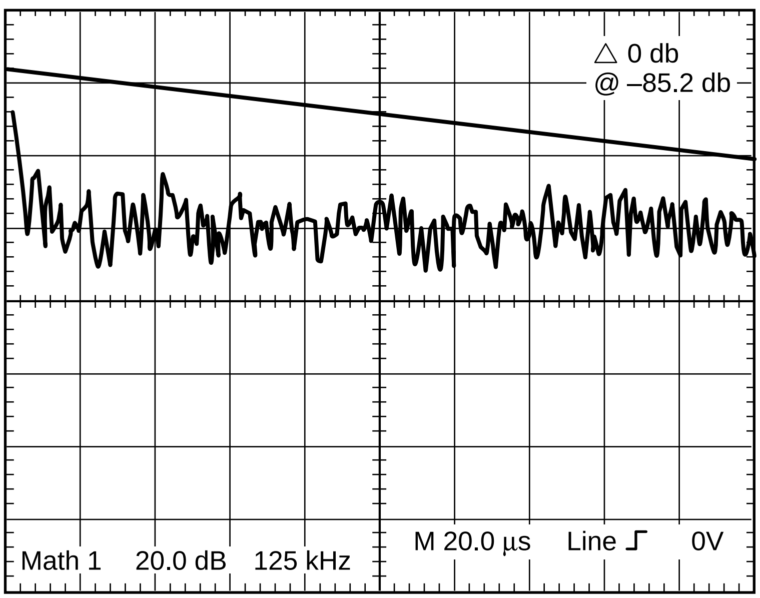


Figure 5-13.6(b). CM filter plus complete DM filter; as above plus 100-μH choke in 28-Vdc line, 28 VRTN, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

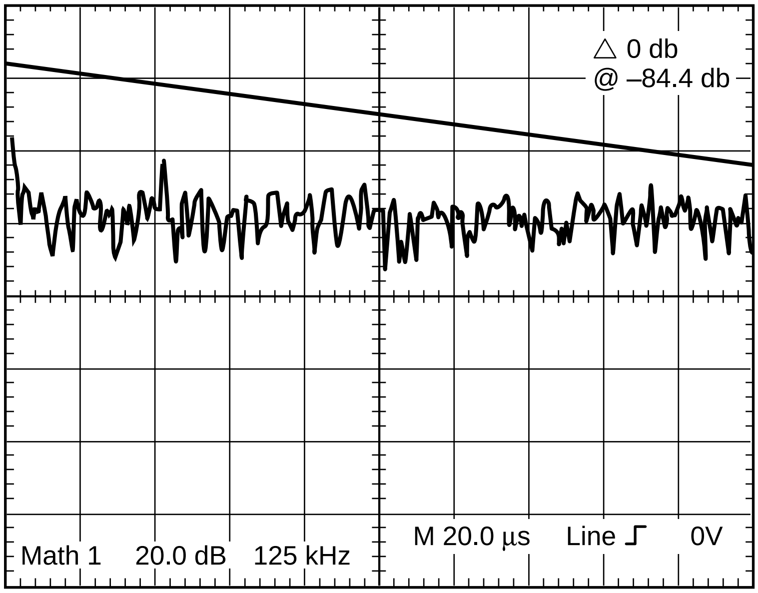


Figure 5-13.7(a). Final compliance check 28 Vdc, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

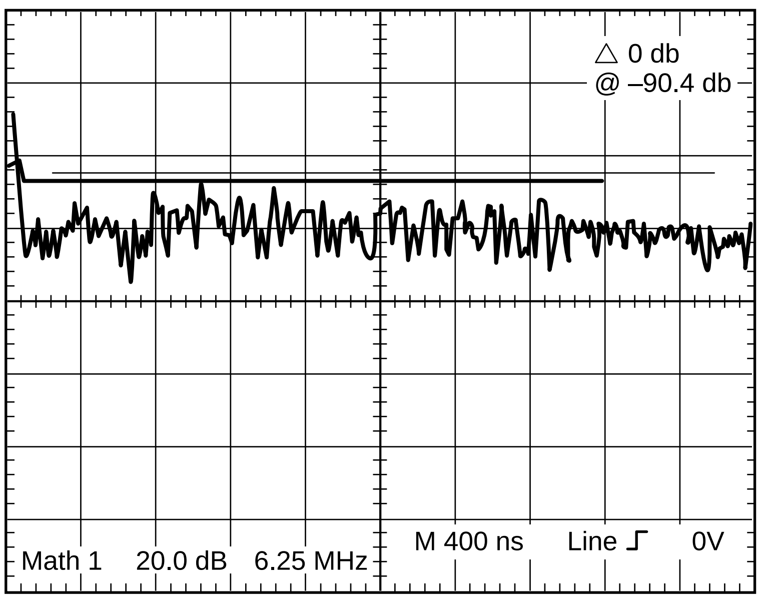


Figure 5-13.7(b). Final compliance check 28 Vdc, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

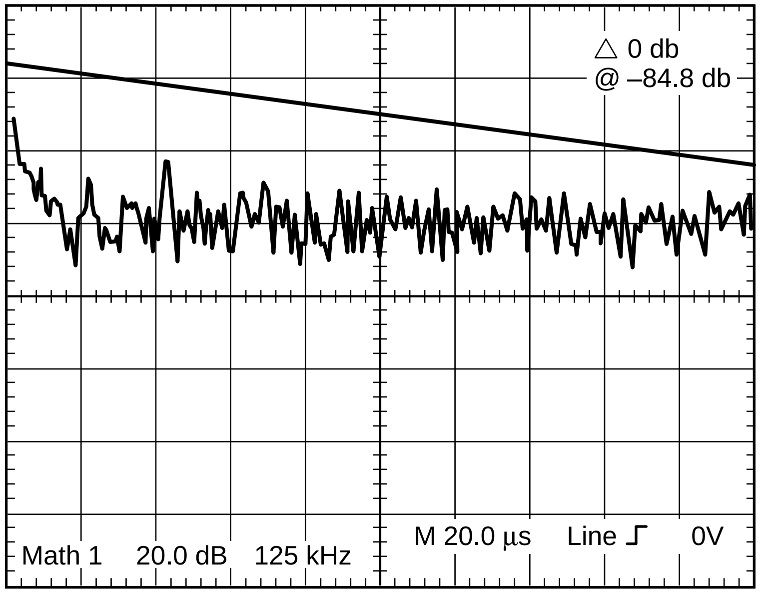


Figure 5-13.7(c). Final compliance check 28 VRTN, low frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 125 kHz/div, starting at dc).

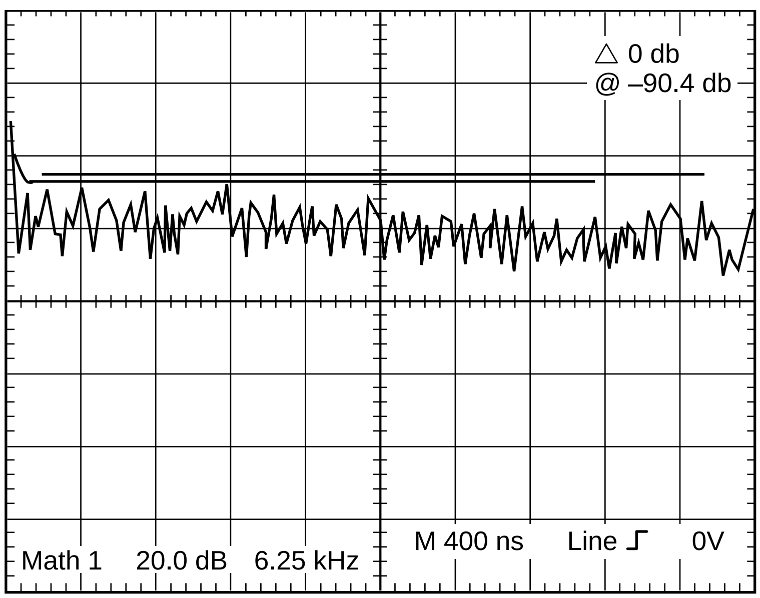


Figure 5-13.7(d). Final compliance check 28 VRTN, high frequency (Y-axis 20 dB/div, compare amplitudes to limit line; X-axis is linear, 6.25 MHz/div, starting at dc).

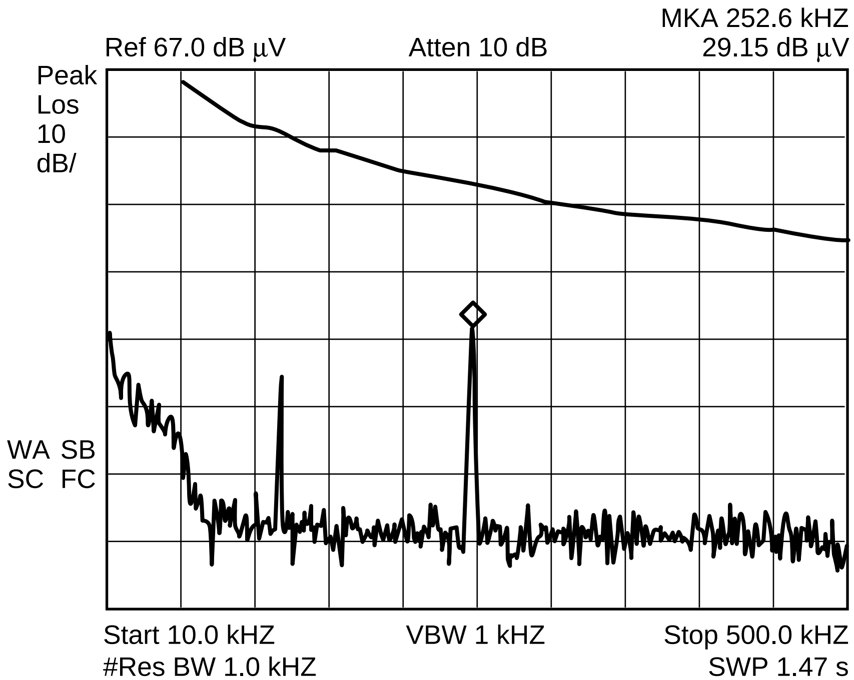


Figure 5-13.8(a). Verification using EMC spectrum analyzer, 28 Vdc, low frequency.

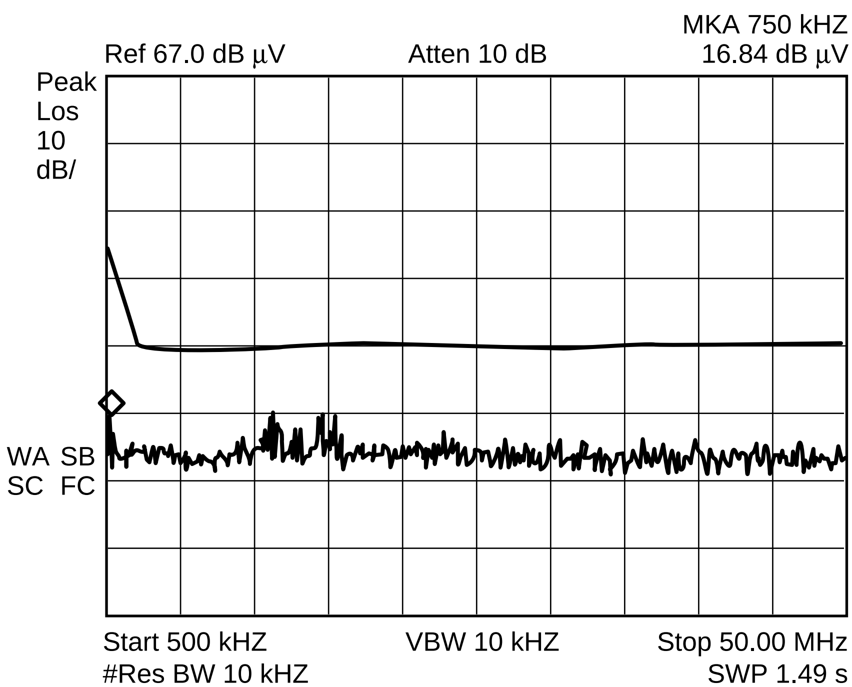


Figure 5-13.8(b). Verification using EMC spectrum analyzer, 28 Vdc, high frequency.

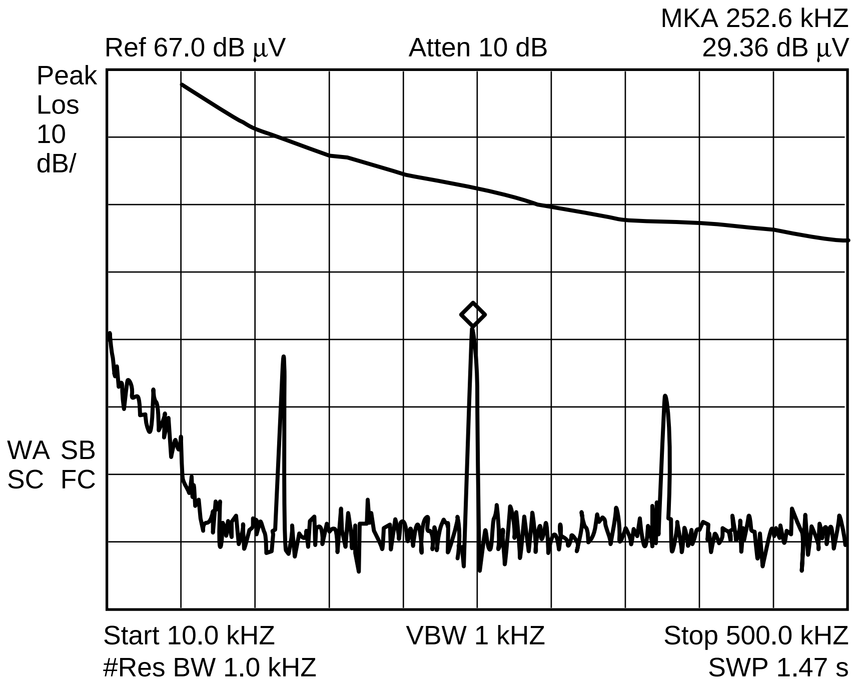


Figure 5-13.8(c). Verification using EMC spectrum analyzer, 28 VRTN, low frequency.

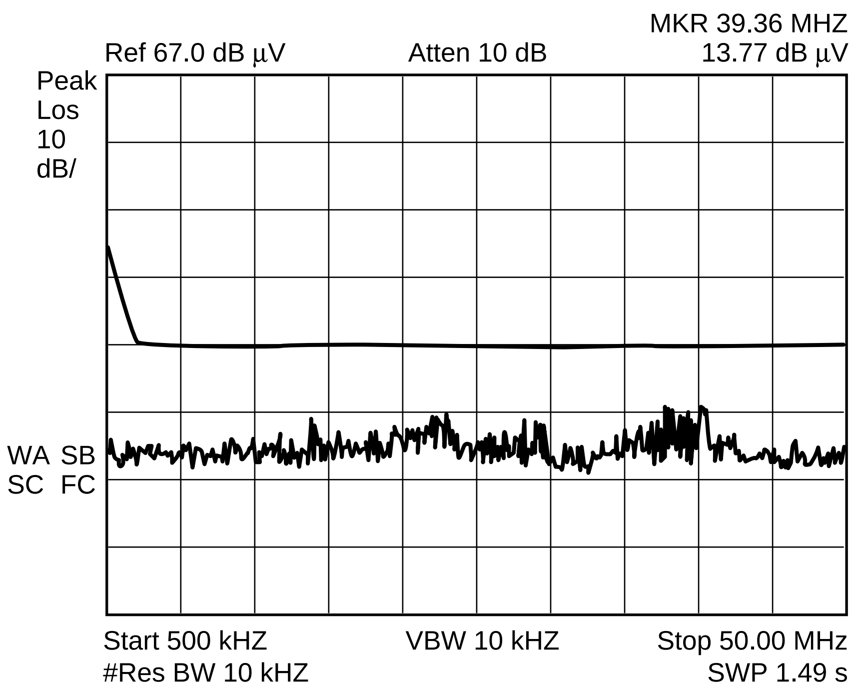
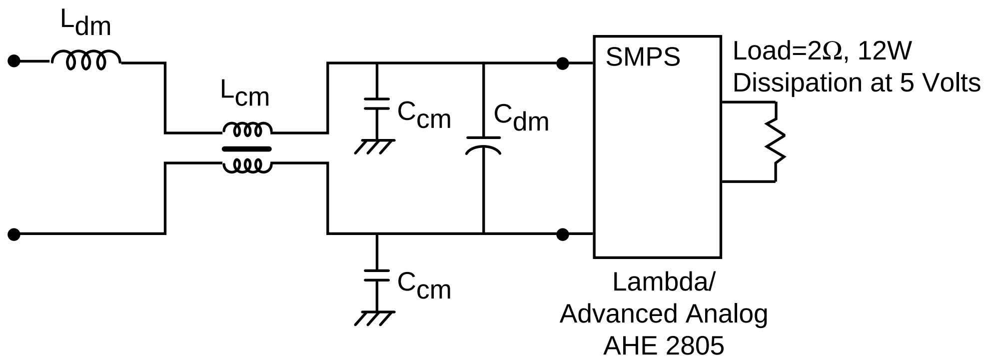


Figure 5-13.8(d). Verification using EMC spectrum analyzer, 28 VRTN, high frequency.



Filter Elements

L_{dm}—100 µH, rated at 2 Amps power frequency current

L_{cm}—six turns of AWG 20 on Supemalloy™ core

C_{cm}—2,000 pF ceramic caps, lead length is negligible (<5 mm)

C_{dm}—20 µF electrolytic

Figure 5-14. Final filter configuration schematic.

5.3 Radiated Emissions Diagnostics

RE are caused by the flow of current on conducting surfaces. These currents include intentional signals on wires in cables, unintentional noise flowing on the same cables, and currents flowing on equipment enclosure surface. The following techniques assist in determining the source of radiated emissions and in curing the problems.

5.3.1 Low-Frequency Specification Outages

If emission frequency is such that dimensions of the test setup are small relative to a wavelength, then equipment-connected cables are the primary suspects. Efficiency of a circuit as a radiator is proportional to its length when the length is short with respect to a wavelength. Even though the intentional signal is a baud rate (10 kHz or slower), MHz signals may parasitically couple to the wire or the outside of a cable shield. Using a current probe with suitable bandwidth is indicated here. Various models of current probes (fig. 5-15.1) are available, ranging from 20 Hz to 1 GHz. Established laboratories have several models. Current probes are single-turn primary, multiple-turn secondary transformers that output a voltage into a 50-Ω load (spectrum analyzer/receiver front end) corresponding to the current flowing through the current probe window. The probe is provided with a transfer impedance (voltage out per unit current through window) versus frequency curve by the manufacturer. EMI technicians are well versed in its use. When the frequency of the RE failure is such that the cable-under-test (CUT) approaches a quarter wavelength, it is important to slide the probe up and down the cable to search for resonances (peaks). In fact, such resonances give rise to RE. A rule of thumb, CE in excess of those shown in figure 5-15.2 are candidates for becoming RE culprits.

Figure 5-15.2 stops at 400 MHz because emissions at this frequency and above are more likely to emanate from the equipment enclosure itself. If significant CE are found at a frequency corresponding to out-of-specification RE, then mitigating steps in the next two sections are worthwhile.

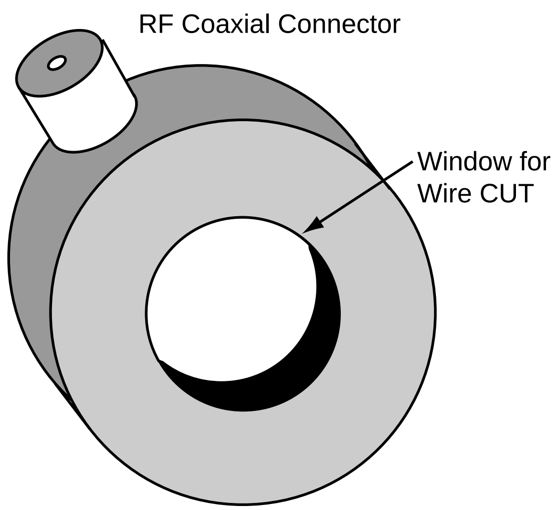


Figure 5-15.1. Current probe.

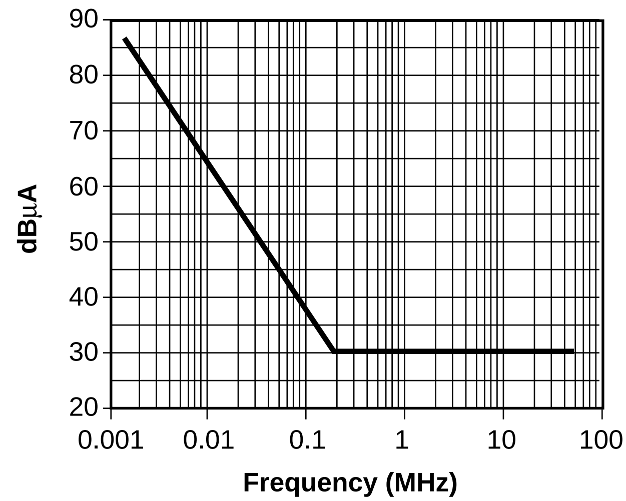


Figure 5-15.2. CE current to limit RE from cables.

5.3.1.1 Attenuating Common Mode Currents on Cable Overshields

If excessive CE are found on a cable overshield, cable shield terminations are the primary suspect. If the out-of-specification condition occurs at 100 kHz or higher, pigtailed terminations are ineffective. A 360˚ peripheral shield termination effect, either through a special EMI backshell or a reasonable facsimile, is necessary. Bringing the cable shield over the connector shell assembly and attaching with a hose clamp is a quick way to determine if a better shield termination mitigates the problem. Sometimes this is difficult to do if the cable shields are cut short and pigtailed to a connector strain relief. A diagnostic check is to remove plastic coating from each and every cable shield and wrap or wind steel wool around the exposed shields. The steel wool should be long enough to spirally wrap it up and over the cable connector. Hose clamps are used to secure the wool over the connector and cable shield ends. If this technique significantly reduces RE, a 360˚ peripheral shield termination is definitely indicated. If cables exit the test chamber, the shield treatment at the exit point must also be as just described in order to stop both support equipment EMI and ambient RFI from entering the test chamber and reradiating.

Theory of cable shielding at medium frequency and beyond (300 kHz +) is quite different than for instrumentation type shields. The shield must not only have a low impedance to ground (at both ends) but also provide for physical separation of currents interior to the shield versus exterior ones. The shield acts as a “container” of EMI; in fact, it is an extension of the equipment enclosure. The shield is analogous to a tunnel or enclosed walkway between two buildings; its purpose is to preserve and extend the environment of the two buildings in the space between them. The cable must not leak and allow mixing of two separate environments. Since currents flow on the inside of the cable shield due to EUT generated noise, the shield termination must prevent those currents from finding a way to the outside. A 360˚ peripheral termination accomplishes this separation. The skin depth of the shield material (presumably resulting in low shield transfer impedance) provides the separation function on the cable shield.

If cable terminations are good, but excessive cable RE are evident, the leakage must occur at some other point of the enclosure. Look for visible seams and apertures. Surface current probes and near-field magnetic probes are common troubleshooting test equipment in an EMI test facility. The surface current probe is similar to the more typical windowed probe; i.e., it outputs a voltage according to the surface current flowing immediately underneath it. A near-field magnetic probe senses the magnetic field in the immediate vicinity of the surface current. Since the magnetic field drops rapidly with distance from the source current, the sensor quickly “homes in on” the culprit leak.

If an enclosure inhomogeneity is found, numerous solutions are available. In troubleshoot mode, EMI tape (aluminum or copper tape with conductive adhesive) is used to cover the seam to determine if a fix is useful. The real solution is to use an EMI gasket, or fingerstock, or to improve tolerances on the mating surfaces. If the aperture is intentionally placed, i.e., a CRT screen or other user interface, optically transparent shielding is necessary. Many manufacturers provide these products, which use a mesh of fine wires to provide good shielding with some optical degradation (typically, moiré patterns). Another technique uses a thin deposition of metal on glass to provide shielding. Here, there are no moiré patterns, but the shielding effectiveness (in dB) is proportional to the thickness of the deposition as is the amount of light loss (in dB) through the glass. Also suspect are apertures for knobs, keyboards, air intake/exhaust, etc. Shielded knobs and switches exist, as do many air/EMI filters for fan intake/exhaust. Keyboard emission problems generally have to be solved at the printed circuit board level.

5.3.1.2 Attenuating Common Mode Currents on Unshielded Cables

If an unshielded cable is determined to be the source of RE, via use of a current probe, the parasitic coupling of EMI to the cable must be attenuated. A current probe measures only the net current flowing through its window. If the cable is designed correctly, the intentional signal is not measured by the probe (both signal and return lines pass through the probe window). Any CE measured in-band to the RE failure are unintentionally present on the cable and, therefore, are filtered or otherwise removed from the cable without affecting the intentional current. Techniques for achieving this are CM chokes implemented as RF beads and line-to-ground capacitance providing the capacitors do not load the intended signal. If these after-the-fact-bandaid approaches do not work, it may be necessary to redesign the cable interface at the equipment enclosure. A bandaid approach is to shield the cable. Another approach is to determine how the EMI is coupled to the cable and to perform an isolation at the circuit level (source suppression). This might involve redesigning the PCB layout, or choosing interface circuitry which inherently provides isolation (opto-isolators, transformer coupling, etc).

5.3.2 Higher-Frequency Outages

When the frequency of the outage is such that the wavelength approximates the EUT enclosure dimensions, seams and apertures in the enclosure are suspected sources of emissions. A general rule is seams and apertures are shorter than one tenth wavelength in order to avoid leakage.

Since cables are easier to troubleshoot, by virtue of pushing them onto the ground plane or shading them from the antenna, this check is typically performed first. If cabling is not indicated as the problem source, the EUT enclosure must be investigated. Obvious leakage points are air vents and data I/O ports. For troubleshooting purposes, seal the I/O ports with EMI tape (available in every EMI lab) and rotate air vents away from the antenna if not already the case. If emissions are not significantly reduced and air vents are suspected, try building a waveguide beyond the cut-off intake/exhaust tube electrically bonded to the EUT enclosure. The tube should be long with respect to diameter, and the diameter should be small relative to wavelength. If this is not possible with one tube, many narrow tubes are necessary. Since use of a tube or tubes increases back pressure, this is strictly a diagnostic test method to be used for a short period of time. If air vent protection is necessary as a final fix, metallic mesh screens or metallic “honeycomb” type EMI protection is available from a number of sources. Figure 5-15.3 shows typical honeycomb material. As in tubing, shielding effectiveness depends on length-to-diameter ratio (larger is better), diameter-to-wavelength ratio (smaller is better), and bonding achieved between honeycomb and EUT enclosure.

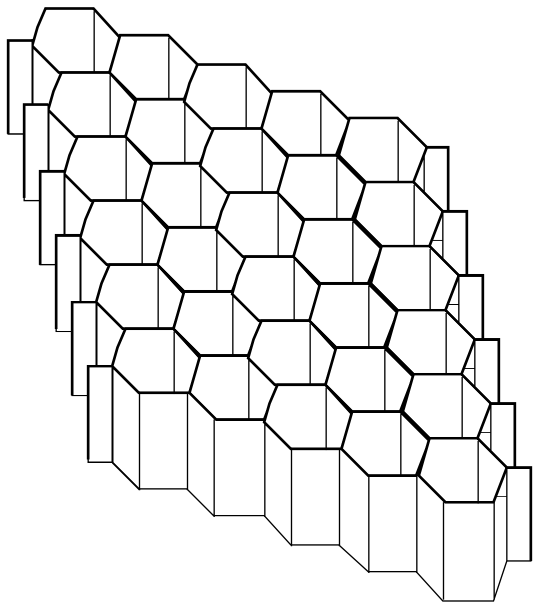


Figure 5-15.3. Honeycomb air vent protection.

5.4 Immunity to Radio Frequency Field Disturbances Diagnostics

Electronic equipment operates in a complex electromagnetic environment (EME) made up of sensitive receivers and powerful transmitters. The ability of the equipment to operate without degrading the reception of weak radio signals has been covered previously (conducted and radiated emissions sections). This section deals with the capability of equipment to operate in a high field intensity EME, due to immediate proximity of high power RF transmitters.

Very often the nonantenna-connected type EUT is not very responsive to the transmitted carrier, but is rather responsive to amplitude modulation (AM) of the carrier. This is because carrier frequency is out-of-band to the intentional operation of the victim circuitry and its sensitivity is therefore degraded. Furthermore, the threshold of sensitivity of non-RF electronics is generally much higher than that of sensitive RF electronics. (An exception to this rule is thermocouple-connected amplifiers, typically both high sensitivity and high-input impedance, leading to extremely low thresholds of susceptibility to RF fields). Diode (P/N junction) detection of carrier modulation results in audio rectification of carrier modulation. The audio rectified signal (typically 1 kHz or below) is easily interpreted by victim circuitry as an intentional signal and processed accordingly.

A case history in this type problem involved a furnace used to grow crystals in microgravity environment. Temperature control circuitry was remote from the furnace and temperature sensors were thermocouples. Necessary temperature stability dictated that the thermocouple voltages had to be accurate to the microvolt level. Wires emanating from the furnace led to a control unit over 1-m distant. The first stage of electronics was op-amps, with no filtering installed. The next stage of amplification included RF filtering. The furnace was supposed to operate in a radiated field intensity of 1 V/m from 14 kHz to 1 GHz or higher. The furnace was susceptible to much lower than 1 V/m. The first stage of amplification was entirely unprotected. Although the cable was well shielded, enough EMI entered the unprotected first stage to become rectified and interpreted as signal in the second (RF filtered) stage. The lesson is to protect the very first section of the cable interface.

5.4.1 Troubleshooting Low-Frequency Susceptibility Problems

If frequencies of the susceptibilities are such that dimensions of the test setup are small relative to wavelength, equipment-connected cables are the primary entry points. Efficiency of a circuit as a pick up is proportional to its length, when the length is short with respect to a wavelength. Placing the suspect cable(s) on the ground plane is a good way to determine the weak link. With cables as close as possible to the ground plane, pick up of EMI is decreased radically and threshold of susceptibility should rise accordingly. If this is the case, the shielding (shield termination and/or filtering/bonding of the interface circuitry) is suspect. Shielding troubleshooting follows the same lines as radiated emission control with one important exception. For radiated susceptibility (RS) control, bonding of shield terminations to the ground plane is of utmost importance. Since the shield is terminated to an equipment enclosure, not the ground plane, the EUT enclosure itself must make a good, low RF impedance bond to the ground plane. In the RE case, the shield must terminate, appropriately, to the EUT enclosure, but the enclosure itself could be floated from the ground plane and not hurt (probably help) RE performance. Not so for RS control. Currents are induced to flow on all metallic surfaces by the impinging RF field. These currents must find a low-impedance path to ground or they will induce RF voltages in circuit elements. Even though the EUT enclosure is a seamless copper box and the cable shield a protruding copper pipe, if the enclosure is floated, the induced RF currents place a potential between the enclosure and ground plane. If the victim circuits are unbalanced and ground referenced, noise is introduced into the signal reference.[1] Based on this, it is extremely important that the EUT bond to the ground plane precisely model that in the planned system installation, otherwise, results of RS testing are invalid.

If the designer suspects RS performance is important, or that the EUT is likely to fail RS tests, there is a relatively inexpensive benchtop test for precompliance checkout of RS performance. This test involves a technique known as bulk current injection (BCI).

5.4.1.1 Bulk Current Injection

BCI is a lumped element model of field-to-wire coupling. As such, it is most applicable at low frequencies, in which the CUT is short relative to wavelength. Figure 5-16.1 stops at 400 MHz, since fields above this frequency are just as likely to penetrate the equipment enclosure as to couple to the cables and since serious concern exists with the validity of the test at higher frequencies. An injection clamp similar to that used in CE measurements injects currents onto the CUT. Figure 5-16.1 shows the induced current expected as a function of frequency due to a 1 V/m field impinging on a 2-m CUT. Computation from Faraday's law and typical cable installation geometries yield the result that 1.5 mA of current flows on a cable in response to 1 V/m of incident field intensity at frequencies in which the cable is at least one half wavelength long. At lower frequencies, the induced current drops at 20 dB per decade. If field intensity is different from 1 V/m, the dBμA and dBm curves are adjusted as 20•log (actual field intensity in V/m). If the cable is longer than 2 m, the low-frequency breakpoint is extended in direct ratio to the length extension. If the BCI clamp insertion loss differs from that plotted, the dBm curve shifts accordingly. Figure 5-16.1 shows why the test is a good precompliance tool: power requirements at the clamp are compatible with a signal generator

[1]Notice that circuitry which is considered balanced at the frequency of its intended operation may be quite unbalanced at higher frequencies, where parasitic effects dominate.

output and no expensive amplifier is necessary. Figure 5-16.1 assumes a clamp covering 2 to 400 MHz. Injected current at lower frequencies is so low that only tunable RF electronics operating in-band to the susceptibility signal are expected to respond. A typical BCI clamp is shown in figure 5-16.2. A BCI test setup is shown in figure 5-16.3. If this test is performed at an EMI test facility, a refinement is available: use a current measurement probe and spectrum analyzer for over-current control. Regardless of predicted clamp drive power from figure 5-16.1, actual injected current should not exceed the dBμA curve by more than 6 dB.

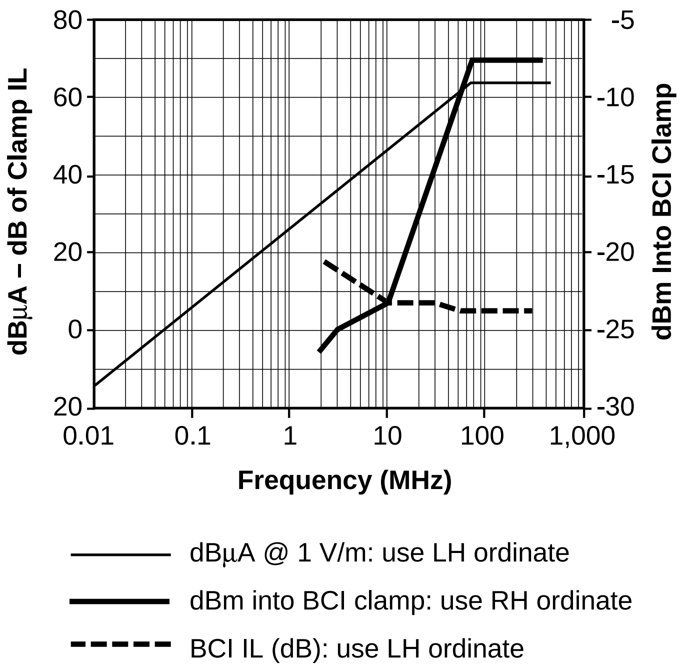


Figure 5-16.1. Converting 1-V/m field to bulk current drive.

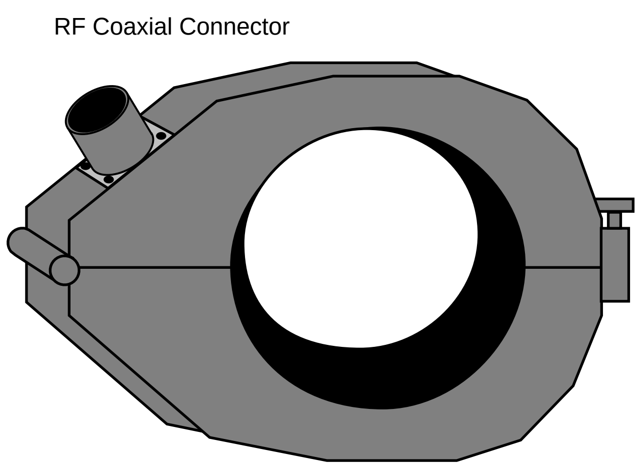


Figure 5-16.2. Typical current injection clamp.

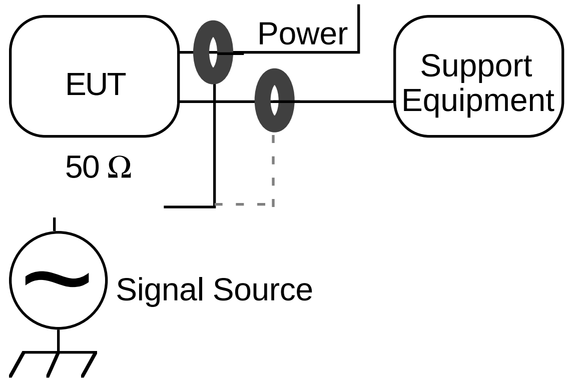


Figure 5-16.3. BCI test setup.

5.4.2 Higher-Frequency Susceptibility

When susceptibility frequency is such that the wavelength approximates the EUT enclosure dimensions, seams and apertures in the enclosure itself are suspected entry points. A general rule is that seams and apertures are shorter than one tenth wavelength to avoid leakage.

Since cables are easier to troubleshoot, by pushing them onto the ground plane or shading them from the antenna, this check is typically performed first. If cabling is not indicated as the problem, the EUT enclosure is investigated. Obvious leakage points are air vents and data input/output ports. For troubleshooting purposes, seal the I/O ports with EMI tape (every EMI lab has this available) and rotate air vents away from the antenna if not already rotated. If the threshold of susceptibility is not increased and air vents are suspected, try building a waveguide beyond the cutoff intake/exhaust tube electrically bonded to the EUT enclosure. The tube should be long with respect to diameter and diameter should be small relative to wavelength. If this is not possible with one tube, many narrow tubes are necessary. Because use of a tube or tubes necessarily increases back pressure, this is strictly a diagnostic test method to be used for a short time period. If air vent protection is necessary as a final fix, metallic mesh screens or metallic “honeycomb” EMI protection is available from a number of sources. Figure 5-15.3 shows typical honeycomb material. As in tubing, shielding effectiveness depends on the length-to-diameter ratio (larger is better), diameter-to-wavelength ratio (smaller is better), and bonding achieved between honeycomb and EUT enclosure.

5.5 Checking Transient Emissions, and Immunity to Conducted Switching Transients

5.5.1 Conducted Transient Sources and Characteristics

Power bus transients arise from load switching, lightning, electrical faults, or electromagnetic pulse (EMP). Only switching transients are of interest for orbiting payloads. Specification of switching transient emissions (designated CE07 or TT01) is a fairly accurate representation of real power bus switching transients. CE07 or TT01 is easy to measure, and the simplifications supplied herein allow the test to be performed outside the EMI test facility. The spike susceptibility requirement (CS06) requires expensive, special spike generating equipment (not found outside an EMI test facility). Instructions provided in this section are used to assesss immunity to spikes. The spike test procedure provided is a highly accurate model of switching transients. Demonstrated compatibility with properly chosen spike amplitudes yields a high degree of confidence that the EUT is immune to power bus transients. However, immunity to spikes described in this section does not predict or guarantee immunity to CS06 spikes. These spikes are of much lower source impedance and are much more difficult to filter.

5.5.1.1 How Does a Switching Transient Occur?

Figure 5-17.1 shows the elements of a power distribution system: a power source, distribution wiring, and load. The power source is simplified to an ideal voltage source in series with a resistive and/or inductive impedance. The distribution wiring contributes both resistance and inductance. The load, at turn on or turn off, provides a rapid change of current through the power source and wiring impedance. This simple model ignores, other than the load, any capacitive effects. Source parallel capacitance (especially in a dc supply) contributes to source stiffness which is easily modeled in the transient case by using a smaller series source impedance. Line-to-line or line-to-ground wiring capacitance is easily accounted for by modeling the distribution wiring as an inductance bypassed by a resistor, i.e., a lumped element model of a transmission line (LISN). Figure 5-17.2 shows a model for both calculating and measuring switching transients. In figure 5-17.2, the LISN models the distribution wiring impedance.

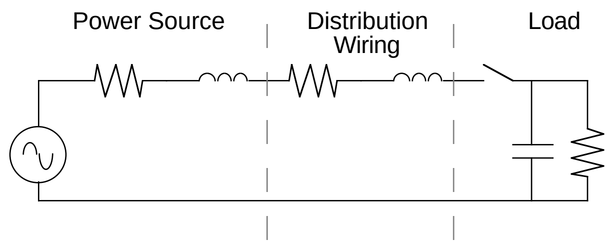


Figure 5-17.1. Model of electrical power distribution system.

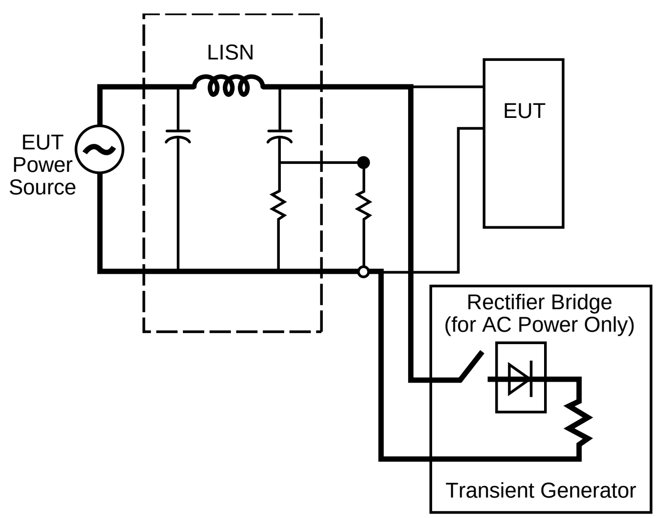


Figure 5-17.2. Proposed spike generator (heavy lines show flow of high current to spike generating load).

The 50-μH, 50-Ω LISN has been arbitrarily selected to serve as worst-case model of wiring impedance. There is some intuitive rationale for the selection. Consider that the inductance of a wire above a ground plane is roughly one microhenry per meter (for typical geometries). Fifty microhenries account for a wire length of 50 m, which is certainly a reasonable worst case for this type power distribution (only feasible in metallic vehicles). A two-wire line has an inductance about one-tenth of the wire-above-ground; therefore 50 μH represents about 500 m of wiring in a commercial setting (also a worst-case model).

The transient generating mechanism is the switching on/off of a heavy power bus load. If this load is the EUT, TT01, or CE07 type, requirements are imposed to bound both the amplitude and duration of the generated transient. If the immunity of the EUT to other power bus load induced spikes is to be assessed, a heavy load must be switched from the LISN while the EUT is in steady-state operation and shares the same LISN as a power bus source impedance. The LISN models the common impedance to the EUT and switched load. Qualitative analysis of the on/off transients is presented next.

5.5.1.1.1 The Turn-On or Negative-Going Transient

In order to measure turn-on transient emissions from the EUT, instructions of the contractually imposed TT01 or CE07 procedures are followed with only the substitution of the 50-μH LISN as a line-above-ground power source impedance. This facilitates a single channel unbalanced oscilloscope measurement. The switch described herein is an excellent substitute for mechanical switches (it does not bounce or arc). For immunity assessment, the initial condition in figure 5-17.2 is that the load switch is open; no current is flowing in the switched load. The EUT is on and in steady-state operation. Upon switch closure, current attempts to flow through the load. LISN inductance opposes the change in the current by dropping the source voltage across itself. LISN output voltage momentarily dips to near zero and then gradually increases as the inductor relaxes. The transient time constant is a function of LISN inductance and RC time constant of the load, with oscillations due to inductor-capacitor energy transfer. The source impedance of the transient is the impedance of the switched load. In this example, the supply voltage is 28 Vdc, and the load bank is 7 Ω paralleled by 100 μF, drawing 4 A after the capacitor charges. (The rectifier diode and ac power source shown in figure 5-17.2 are not applicable in the dc case.) The turn-on transient for these conditions is shown in figure 5-17.3.

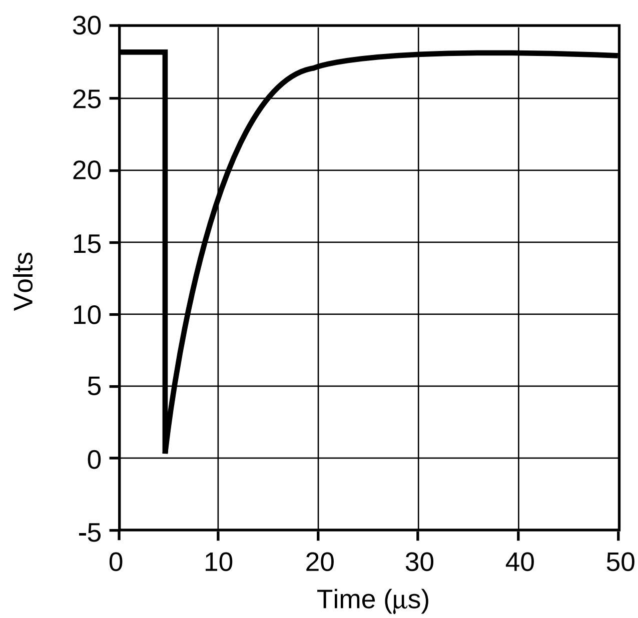


Figure 5-17.3. Turn-on transient (turning on 4-A load in parallel with 100-μf capacitor).

5.5.1.1.2 The Turn-Off or Positive-Going Transient

In order to measure turn-off transient emissions from the EUT, the instructions of the contractually imposed TT01 or CE07 procedures are followed, with only the substitution of the 50-μH LISN as a line-above-ground power source impedance. The switch described herein is an excellent substitute for mechanical switches because it does not bounce or arc and switches so rapidly that a worst-case turn-off spike is guaranteed. For immunity assessment, the initial condition is the load is on long enough to achieve steady-state 4-Adc current flow. The switch is abruptly opened. The LISN 50-μH inductor tries to maintain the 4-A current flow through itself, by raising the voltage at the output of the LISN relative to the input. (Incidentally, this phenomenon answers the oft raised question about spike tests: “Does the specified spike amplitude include the line voltage, or is it superimposed on the line voltage?” The line inductance superimposes the spike voltage on the power line voltage or it would not have the desired effect of maintaining the current through the inductance). If the inductor were the only element to consider, the spike induced by turning off the load would be infinite in amplitude. However, reality imposes line-to-line and other stray capacitance which tends to snub the spike. One benefit of the 50-Ω LISN is that the 50-Ω dummy load provides a stronger snubbing effect than any stray capacitance, yielding repeatable, predictable spikes. If our switch is fast enough (the one described herein is), we are in the 50-Ω frequency domain of the LISN and the spike voltage is the switched current multiplied by 50 Ω. In this example, we should see a 200-V spike (50 Ω × 4 A). The time constant is independent of the load impedance; it has been switched out of the circuit. The time constant is the ratio of the 50-μH inductor and the 50-Ω dummy load (one microsecond). The source impedance is 50 Ω. The qualitatively predicted waveform is shown in figure 5-17.4.

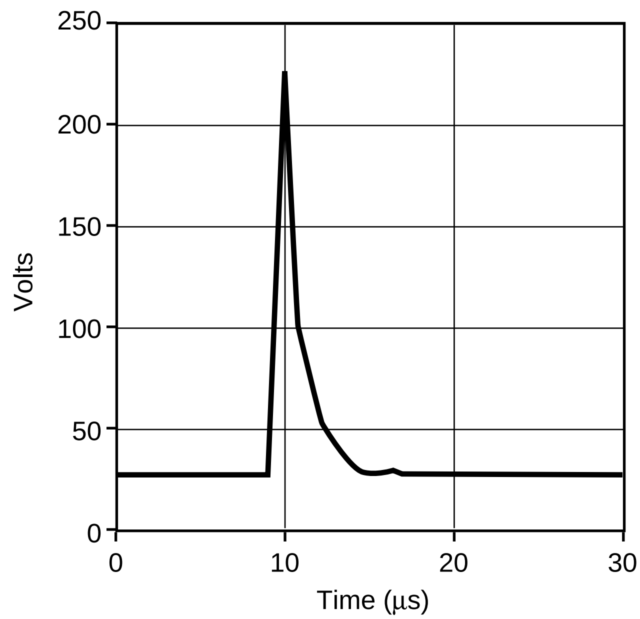


Figure 5-17.4. Turn-off transient (turning off 4-A load on 28-Vdc bus).

5.5.2 The Switch

The circuit for performing the immunity test switching function is shown in figure 5-17.5. A 100-μF capacitor lengthens the turn-on transient but has no effect on the turn-off transient. Figures 5-17.3 and 5-17.4 are open-circuit measurements. While the turn-on transient is low impedance and difficult to load, the 50-Ω turn-off transient is easily loaded (reduced amplitude, increased duration) by another load on the LISN (such as the EUT).

To assure waveform repeatability, it is necessary for the switch transition time spectrum to be in the 50-Ω region of the LISN. While other methods are possible, this example uses a MOSFET and FET driver circuit to achieve the desired transition time.

To measure spike emissions, only the FET and FET driver portions of the figure 5-17.5 circuit are necessary. Pin 2 of the FET driver is mechanically switched either to V_{CC} or ground, depending on whether the FET is to be switched on or off, respectively. The FET source is connected to the 28-V return wire of the EUT and the FET drain is connected to ground (power supply return). The switched load is not part of this test.

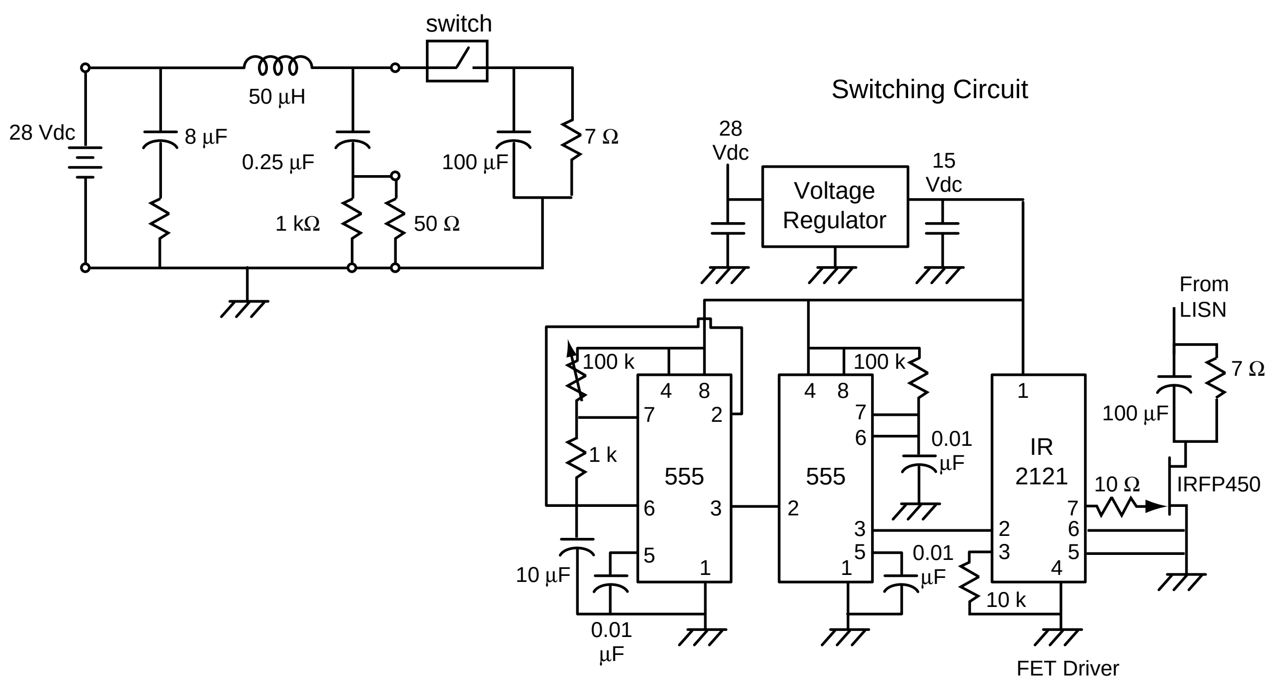


Figure 5-17.5. Transient generating circuit for 28-Vdc loads.

5.5.3 An Important Note About Power Source Rating

For spike emission measurements, the power source must be rated to provide the required EUT current. For immunity testing, however, the extra current requirement is not the steady-state current of the transient generating load. For switched load, the necessary current sourcing capability of the power source is the amount of current drawn by the switched load divided by the on/off duty cycle. If the duty cycle is 10 percent, then the power source need only be capable of providing additional average current of 10 percent of switched current. The line-to-ground capacitance on the input side of the LISN assists in this derating. For a dc power source, any amount of capacitance bypasses the LISN input to further augment derating. Thus, a 1-Adc power source quite adequately provides for a 10-A switching transient capability. Power dissipation of the switched load and transistor is derated proportionally to the square of the switched current. In actual practice, a duty cycle of 1 percent is easily achievable because an “ON” time of 1 ms is more than adequate to achieve steady-state current draw and a pulse repetition rate of 1 to 10 pps is traditional to test spike immunity.

5.5.4 Test Specification and Procedures

The circuit of figure 5-17.5 is adequate for any test in which peak line voltage plus spike amplitude sum is less than 400 V. The amount of switched current depends on both switched load resistance and on/off duty cycle. In this example, 4-A loads were switched to generate spikes.

To define spike immunity limits, both the amplitude and duration inrush and turn-off waveforms must be quantified. Inrush current, provided by a capacitive load, determines the transient duration. The steady-state current is drawn by the parallel resistive load. The “ON” time must be sufficient to fully charge the capacitor. LISN characteristics are equally important parameters to define the spike amplitude, duration, and source impedance. This example used the ubiquitous ANSI C63.4 50-μH, 50-Ω LISN. Using the traditional 5-μH LISN results in shorter transients for both turn-on/off transients.

Although the transient amplitude is not affected, the turn-on transient time constant varies as the square root of the inductance changes, and the turn-off transient time constant is directly proportional to the inductance. For example, switching 4 A off from a 5-μH LISN yields a 200-V spike 0.1-μs wide. If different turn-off spike source impedances are desired, these are achieved by using an LISN dummy load other than 50 Ω. A lower source impedance means more switched current to achieve a given spike amplitude. The opposite effect is achieved by raising the source impedance. If source impedance is raised, one must be careful that the characteristic impedance is achieved at a frequency below that corresponding to the spectrum of the switch rise and fall times. Regardless of the amplitude and time duration selected, the load the turn-off transient is developed for proving specification compliance must be specified. Here the load is open-circuit (oscilloscope probe). If it is desired to measure into a matched load, the turn-off transient amplitude is halved and the time duration doubled.

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APPENDIX A

ACRONYMS AND ABBREVIATIONS

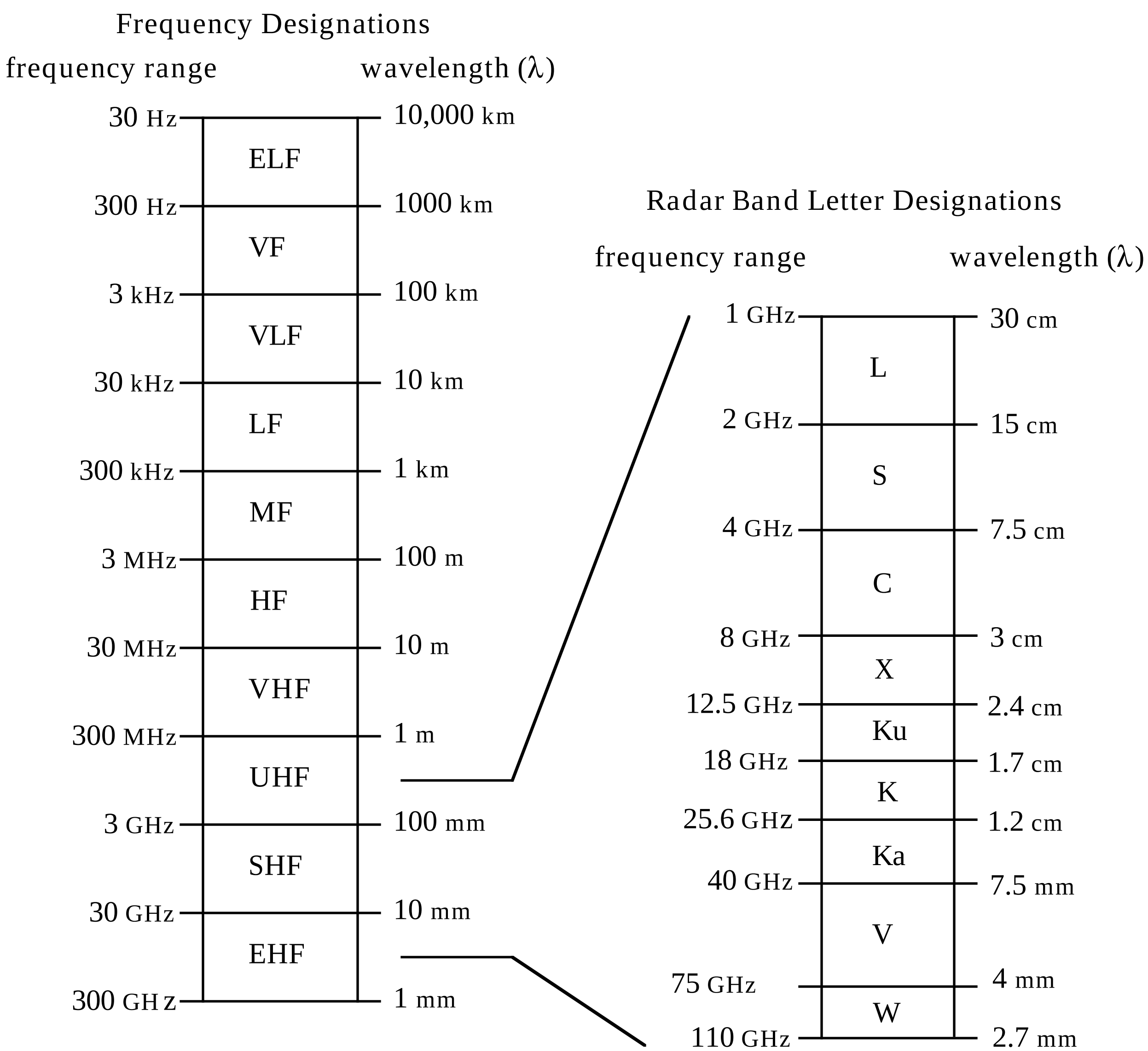
A	ampere
A/m	ampere per meter
ac	alternating current
AM	amplitude modulation
AWG	American wire gauge
BB	broadband
BCI	bulk current injection
C	capacitor
CE	conducted emissions
cm	centimeter
CM	common mode
CS	conducted susceptibility
CUT	cable under test
CW	continuous wave
dB	decibel
dBm	decibel above 1 milliwatt
dBV	decibel above 1 volt
dBW	decibel above 1 watt
dc	direct current
DSO	digital storage oscilloscope
DM	differential mode
ELF	extremely low frequency
EHF	extremely high frequency
EM	electromagnetic
EMC	electromagnetic compatibility
EME	electromagnetic environment
EMI	electromagnetic interference
EMP	electromagnetic pulse
ESD	electrostatic discharge
EUT	equipment under test
FFT	fast Fourier transform
FM	frequency modulation
GHz	gigaHertz

Gnd	ground (electrical)
H	Henry
HF	high frequency
Hz	Hertz
I/O	input/output
kHz	kiloHertz
km	kilometer
IF	intermediate frequency
L	inductor
LC	inductive/capacitive
LH	left hand
LISN	line impedance stabilization network
	line impedance simulation network
LF	low frequency
LO	local oscillator
m	meter
MEDIC	MSFC EMC Design and Interference Control
MF	medium frequency
MHz	megaHertz
mm	millimeter
MSFC	Marshall Space Flight Center
N/A	not applicable
NASA	National Aeronautics and Space Administration
NB	narrowband
pF	picoFarad
PM	phase modulation
PC	personal computer or printed circuit
PCB	printed circuit board
PRF	pulse repetition frequency
PAM	pulse amplitude modulation
PCM	pulse code modulation
PWM	pulse width modulation
RAU	remote acquisition unit
RBW	resolution bandwidth
RC	resistive/capacitive
RCVR	receiver

RE	radiated emissions
RF	radio frequency
RFI	radio frequency interference
RH	right hand
rms	root-mean-square
RS	radiated susceptibility
s	second
S/N	signal-to-noise ratio
SHF	super high frequency
SMPS	switched mode power supply
T	tesla
TT	turn-on/off transient
UHF	ultra high frequency
V	volt
V/m	volt per meter
VBW	video bandwidth
VF	voice frequency
VHF	very high frequency
VLF	very low frequency
VTVM	vacuum tube volt meter
W	watt
WWII	World War II
XFMR	transformer
ε (epsilon)	permitivity
λ (lambda)	wavelength
μ (mu)	permeability or prefix micro
Ω (omega)	ohm
μF	microFarad

APPENDIX B

FREQUENCY BANDS



ELF: Extremely Low Frequency
VF: Voice Frequency
VLF: Very Low Frequency
LF: Low Frequency
MF: Medium Frequency

HF: High Frequency
VHF: Very High Frequency
UHF: Ultra High Frequency
SHF: Super High Frequency
EHF: Extremely High Frequency

APPENDIX C

LOGARITHMS

C.1 Review of Logarithm Rules

The base 10 logarithm function is mathematically defined as follows; if

$$c = 10^{a}\ , \quad \text{then} \quad a = \log_{10} c \quad \text{or} \quad a = \log c\ .$$

Use of logarithms accomplishes two goals which simplify mathematical calculation: Replaces multiplicative processes by additive processes and compresses large numbers into smaller ones. Mathematical rules that govern these transformations are shown in table 1-2.

Table 1-2. Transformation of mathematical operations between linear and log form.

linear calculation	logarithmic calculation
$c = a \bullet b$	$\log c = \log a + \log b$
$c = a \div b$	$\log c = \log a - \log b$
$c = a^n$	$\log c = n \bullet \log a$
$c = \frac{1}{a}$	$\log c = - \log a$

In the physical world, these mathematical constructs are implemented through the use of the decibel defined as a power ratio:

$$\text{decibel (dB)} = 10 \bullet \log_{10} \frac{P_1}{P_2}\ . \qquad \text{(C-1)}$$

If the two power quantities are electrical power dissipated in resistors, the equation (1-7) is expressed as:

$$\text{dB} = 10 \log \frac{P_1}{P_2} = 10 \log \left[\frac{V_1^2/R_1}{V_2^2/R_2} \right]\ . \qquad \text{(C-2)}$$

The rules in table 1-2 allow equation (C-2) to be rearranged as:

$$\text{dB} = 20 \log \frac{V_1}{V_2} - 10 \log \frac{R_1}{R_2}\ . \qquad \text{(C-3)}$$

The objective is to evaluate the power dissipated, or the voltage across the same resistor, under different circumstances. Under the condition that:

$$R1 = R2\ , \qquad \text{(C-4)}$$

equation (C-3) is simplified to the familiar relationship:

$$dB = 20 \log \frac{V_1}{V_2} \ . \tag{C-5}$$

Similar manipulations derive the dB relationship for current through a resistor:

$$dB = 20 \log \frac{I_1}{I_2} \ . \tag{C-6}$$

When V_2 and I_2 are replaced by 1 μV and 1 μA, respectively, the definitions of dBμV (dB relative to one microvolt) and dBμA (dB relative to one microamp) are obtained as given in equations (1-1) and (1-4). When P_2 in equation (1-7) is replaced by 1 mW, the definition of the dBm (dB relative to one milliwatt) is obtained as given in equation (1-2). Repeating equations (1-1) throught (1-4):

dBμV: $$dB\mu V = 20 \ \log \left[\frac{\text{signal strength } (\mu V)}{1 \ \mu V} \right] \tag{1-1}$$

dBm: $$dBm = 10 \ \log \left[\frac{\text{signal strength (mW)}}{1 \text{ mW}} \right] \tag{1-2}$$

dBμA: $$dB\mu A = 20 \ \log \left[\frac{\text{signal strength } (\mu A)}{1 \ \mu A} \right] \tag{1-4}$$

Equation (1-3) is the relationship between dBμV and dBm in a 50-Ω resistor:

$$dB\mu V = dBm + 107 \ . \tag{1-3}$$

An exercise using decibels and logarithms is to derive the relationship of equation (1-3). Begin with:

$$P = \frac{V^2}{R} \ , \tag{C-7}$$

take the logarithm of both sides, and multiply by ten to yield the decibel equation:

$$10 \log P = 20 \log V - 10 \log R \ . \tag{C-8}$$

But the units, respectively, of the parameters in equation (C-7) are watts, volts, and ohms. The desired units are dBm, dBμV, and dBΩ. So, convert equation (C-7) to units of milliwatts and microvolts (resistance remains in ohms):

$$\frac{P(mW)}{1000} = \frac{\left[\frac{V(\mu V)}{10^6} \right]^2}{R} \ . \tag{C-9}$$

Reduce equation (C-9) to:

$$P(\mathrm{mW}) = \frac{V(\mu V)^2 \bullet 10^{-9}}{R} \quad . \tag{C-10}$$

Replace the resistance parameter with 50 Ω to yield:

$$P(\mathrm{mW}) = V(\mu V)^2 \bullet \ 2 \bullet 10^{-11} \quad . \tag{C-11}$$

Take the logarithm of both sides and use the definitions of equations (1-1) through (1-4) to yield the desired result, equation (1-3).

This is a rather long process, albeit mathematically simple. The power and advantage of using logarithms are nicely demonstrated by revisiting the process entirely in logarithms (or decibels). Start again with equations (C-7) and (C-8):

$$P = \frac{V^2}{R} \ , \tag{C-7}$$

and

$$10 \log P = 20 \log V - 10 \log R \quad . \tag{C-8}$$

Convert from watts and volts to dBm and dBμV.

Note: 1000 mW/W translates into:

$$\mathrm{dBm} = \mathrm{dB_{Watts}} + 30 \quad , \tag{C-12}$$

10^6 μV/V translates into:

$$\mathrm{dB}\mu\mathrm{V} = \mathrm{dB_{Volts}} + 120 \quad , \tag{C-13}$$

and equation (C-8) may be immediately converted to:

$$\mathrm{dBm} - 30 = \mathrm{dB}\mu\mathrm{V} - 120 - 10 \log R \quad . \tag{C-14}$$

Evaluate equation (C-14) using 50 ohms for R and collect terms on the right-hand side of the equation results in equation (1-3).

C.2 Logarithm Mnemonics

If the reader works with decibels on a daily basis, relationships in table C-1 enable the reader to solve problems more quickly than using a calculator.

Table C-1. Useful logarithmic relationships.

number	logarithm (number)	10 log (number)	20 log (number)
1	0	0	0
1.1 (= 10 ÷ 9)	0.05 (log 10 - log 9)	0.5	1 20 • (log 10 - log 9)
1.25 (=5÷4)	0.1 (log 5 - log 4)	1	2 20•(log 5 - log 4)
1.4 ($\sqrt{2}$)	0.15 ($\frac{1}{2}$ log 2)	1.5	3
1.5 (=3÷2)	0.177	1.77	3.5
2	0.3	3	6
2.5 (=5÷2)	0.4 (log 5 - log 2)	4	8
3	0.477/0.5	4.7/5	9.5/10
4 (=2^2)	0.6 (2 • log 2)	6	12
5	0.7	7	14
6 (= 2 • 3)	0.77 (log 2 + log 3)	7.7	15
7 (= 10 ÷ 1.4)	0.85 (log 10 - $\frac{1}{2}$ log 2)	8.5	17
8 (= 2^3)	0.9 (3 • log 2)	9	18
9 (=3^2)	0.95 (2 • log 3)	9.5/10	19/20
10	1	10	20

Note: Shaded entries, derived from unshaded entries, are of secondary importance.

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