EN390W
Practical Aspects of Grounding, Power, and EMI
Module 2: Shielding and EMI

Lecture Notes

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Review: Noise and Interference

Noise is any electrical signal present in a circuit other than the desired signal.

Interference is the undesirable effect of noise.
If a noise voltage causes improper operation of a circuit, it is interference.
Noise cannot be eliminated, but can be reduced in magnitude, until it no longer causes interference.

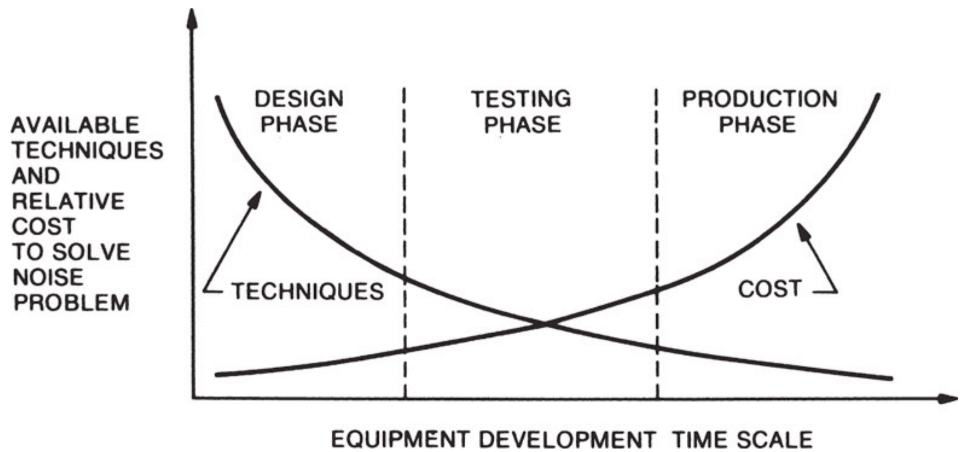


Figure 1-1: As equipment development proceeds, the number of available noise-reduction techniques goes down. At the same time, the cost of noise reduction goes up

Models for Noise Coupling

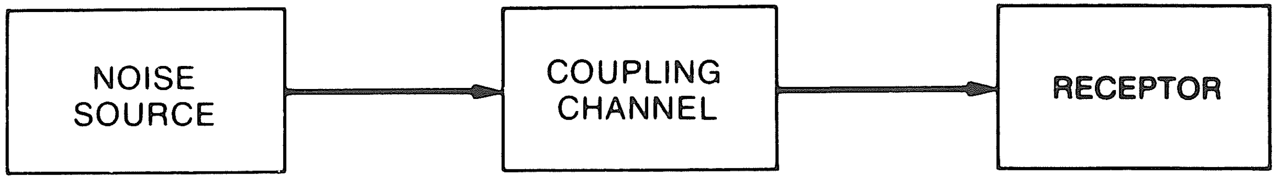


Figure 1-7: Before noise can be a problem, there must be a noise source, a receptor, and a coupling channel

Day 1: Coupling through common impedance

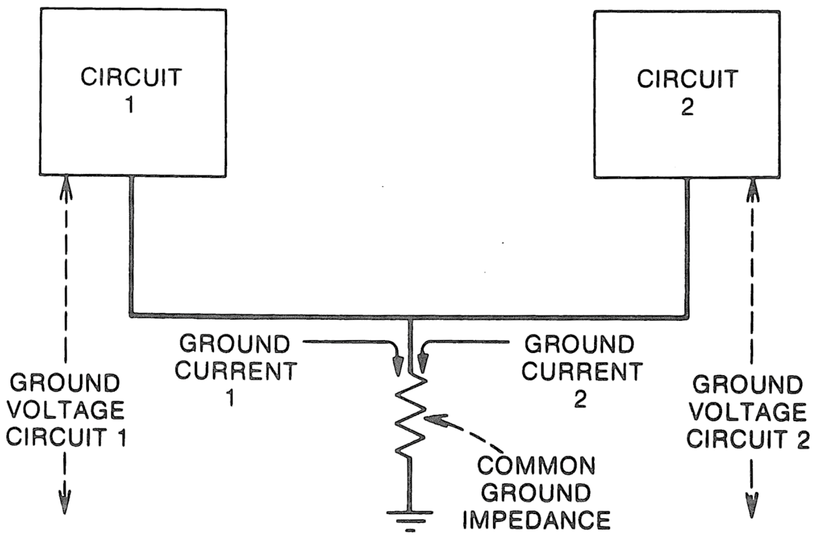


Figure 1-9: When two circuits share a common ground, the ground voltage of each one is affected by the ground current of the other circuit

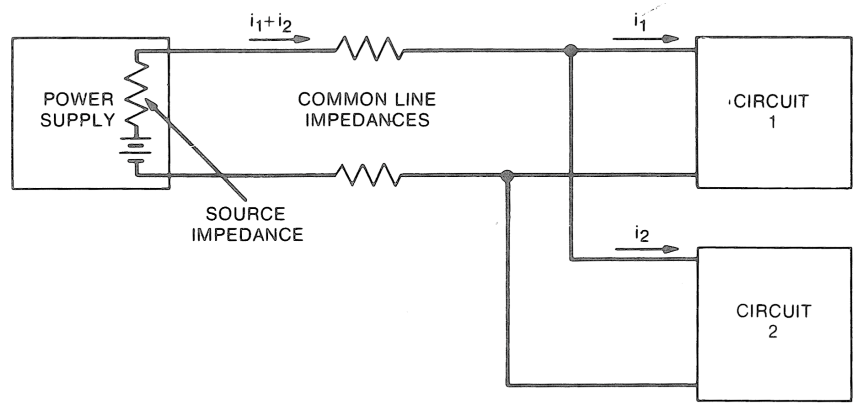


Figure 1-10: When two circuits share a common power supply, current drawn by one circuit affects the voltage at the other circuit

Day 2: Coupling through EM fields

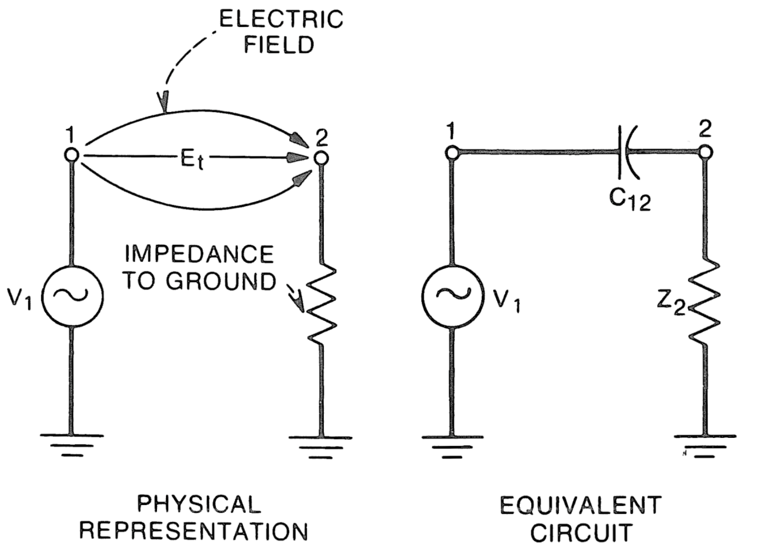


Figure 1-12: When two circuits are coupled by an electric field, the coupling can be represented by a capacitor

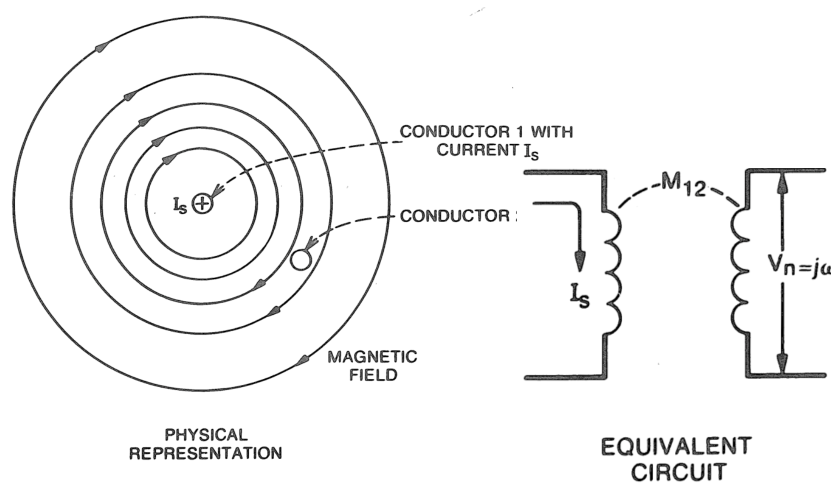


Figure 1-13: When two circuits are coupled by a magnetic field, the coupling can be represented as a mutual inductance

MIL Standards (Source: everyspec.com)

MIL-STD-464C
1 December 2010

SUPERSEDING
MIL-STD-464B
1 October 2010

DEPARTMENT OF DEFENSE
INTERFACE STANDARD

ELECTROMAGNETIC ENVIRONMENTAL EFFECTS
REQUIREMENTS FOR SYSTEMS

MIL-STD-461F
10 December 2007

SUPERSEDING
MIL-STD-461E
20 August 1999

DEPARTMENT OF DEFENSE
INTERFACE STANDARD

REQUIREMENTS FOR THE CONTROL OF
ELECTROMAGNETIC INTERFERENCE
CHARACTERISTICS OF SUBSYSTEMS AND
EQUIPMENT

From MIL-STD-464C:

“The external EME must be determined for each system. When considering the external EMEs (flight deck, airborne, battlefield and so forth), the following areas should be included in the evaluation:

a. Mission requirements. The particular emitters to which the system will be exposed depend upon its intended use. The various parts of MIL-HDBK-235 provide information on the characteristics of many friendly transmitters.
b. Appropriate standoff distance from each emitter. The various parts of MIL-HDBK-235 specify the fields at varying distances.
c. The number of sites and where they are located. The probability of intercept for each emitter and the dwell time should be calculated.
d. If applicable, high power microwave and ultra-wideband emitters should be included. See MIL-HDBK-235-8.
e. Operational performance requirements with options such as survivable only, degraded performance acceptable, or full performance required. “

Requirement Lessons Learned (A.5.3): Without specific design and verification requirements, problems caused by the external EME typically are not discovered until the system becomes operational. By the time interference is identified, the system can be well into the production phase of the program, and changes will be expensive. In the past, the EME generated by the system's onboard RF subsystems produced the controlling environment for many systems; however, with external transmitter power levels increasing, the external transmitters can drive the overall system environment.

Examples of problems observed:

- High-powered shipboard radars have caused interference to satellite terminals located on other ships, resulting in loss of lock on the satellite and complete disruption of communication. The interference disables the satellite terminal for up to 15 minutes, which is the time required to re-establish the satellite link. Standoff distances of up 20 nautical miles between ships are required to avoid the problem.
- A weapon system suffered severe interference due to insufficient channel selectivity in the receiver’s front end. Energy originating from electronic warfare systems and another nearby “sister” channelized weapon system (operating on a different channel but within the same passband) coupled into the victim receiver and was “processed,” severely degrading target detection and tracking capability. An aircraft lost anti-skid braking capability upon landing due to RF fields from a ground radar changing the weight-on-wheels signal from a proximity switch. The signal indicated to the aircraft that it was airborne and disabled the anti-skid system.
- An aircraft experienced uncommanded flight control movement when flying in the vicinity of a high power transmitter, resulting in the loss of the aircraft.
- Aircraft systems have experienced self-test failures and fluctuations in cockpit instruments, such as engine speed indicators and fuel flow indicators, caused by sweeping shipboard radars during flight-deck operations.
- Aircraft on approach to carrier decks have experienced interference from shipboard radars, for example, triggering of false "Wheels Warning" lights, indicating that the landing gear is not down and locked.
- Aircrews have reported severe interference to communications with and among flight deck crew members.

MIL-STD-461F

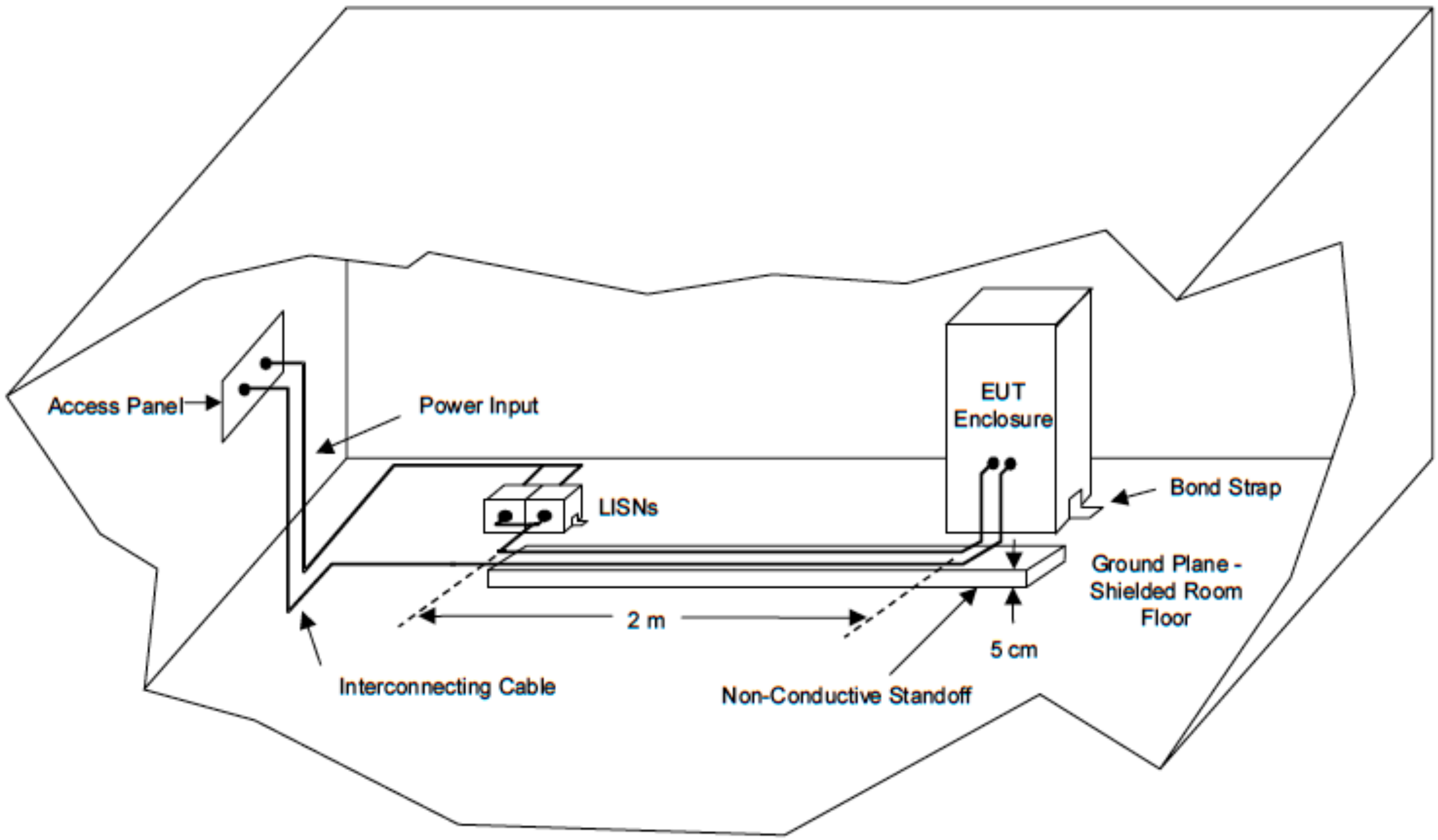


FIGURE 4. Test setup for free standing EUT in shielded enclosure.

MIL-STD-461F

TABLE IV. Emission and susceptibility requirements.

Requirement	Description
CE101	Conducted Emissions, Power Leads, 30 Hz to 10 kHz
CE102	Conducted Emissions, Power Leads, 10 kHz to 10 MHz
CE106	Conducted Emissions, Antenna Terminal, 10 kHz to 40 GHz
CS101	Conducted Susceptibility, Power Leads, 30 Hz to 150 kHz
CS103	Conducted Susceptibility, Antenna Port, Intermodulation, 15 kHz to 10 GHz
CS104	Conducted Susceptibility, Antenna Port, Rejection of Undesired Signals, 30 Hz to 20 GHz
CS105	Conducted Susceptibility, Antenna Port, Cross-Modulation, 30 Hz to 20 GHz
CS106	Conducted Susceptibility, Transients, Power Leads
CS109	Conducted Susceptibility, Structure Current, 60 Hz to 100 kHz
CS114	Conducted Susceptibility, Bulk Cable Injection, 10 kHz to 200 MHz
CS115	Conducted Susceptibility, Bulk Cable Injection, Impulse Excitation
CS116	Conducted Susceptibility, Damped Sinusoidal Transients, Cables and Power Leads, 10 kHz to 100 MHz
RE101	Radiated Emissions, Magnetic Field, 30 Hz to 100 kHz
RE102	Radiated Emissions, Electric Field, 10 kHz to 18 GHz
RE103	Radiated Emissions, Antenna Spurious and Harmonic Outputs, 10 kHz to 40 GHz
RS101	Radiated Susceptibility, Magnetic Field, 30 Hz to 100 kHz
RS103	Radiated Susceptibility, Electric Field, 2 MHz to 40 GHz
RS105	Radiated Susceptibility, Transient Electromagnetic Field

TABLE V. Requirement matrix.

Equipment and Subsystems Installed In, On, or Launched From the Following Platforms or Installations	Requirement Applicability																	
	CE101	CE102	CE106	CS101	CS103	CS104	CS105	CS106	CS109	CS114	CS115	CS116	RE101	RE102	RE103	RS101	RS103	RS105
Surface Ships	A	A	L	A	S	S	S	A	L	A	S	A	A	A	L	A	A	L
Submarines	A	A	L	A	S	S	S	A	L	A	S	L	A	A	L	L	A	L
Aircraft, Army, Including Flight Line	A	A	L	A	S	S	S			A	A	A	A	A	L	A	A	L
Aircraft, Navy	L	A	L	A	S	S	S			A	A	A	L	A	L	L	A	L
Aircraft, Air Force		A	L	A	S	S	S			A	A	A		A	L		A	
Space Systems, Including Launch Vehicles		A	L	A	S	S	S			A	A	A		A	L		A	
Ground, Army		A	L	A	S	S	S			A	A	A		A	L	L	A	
Ground, Navy		A	L	A	S	S	S			A	A	A		A	L	A	A	L
Ground, Air Force		A	L	A	S	S	S			A	A	A		A	L		A	

Legend:

A: Applicable
L: Limited as specified in the individual sections of this standard
S: Procuring activity must specify in procurement documentation

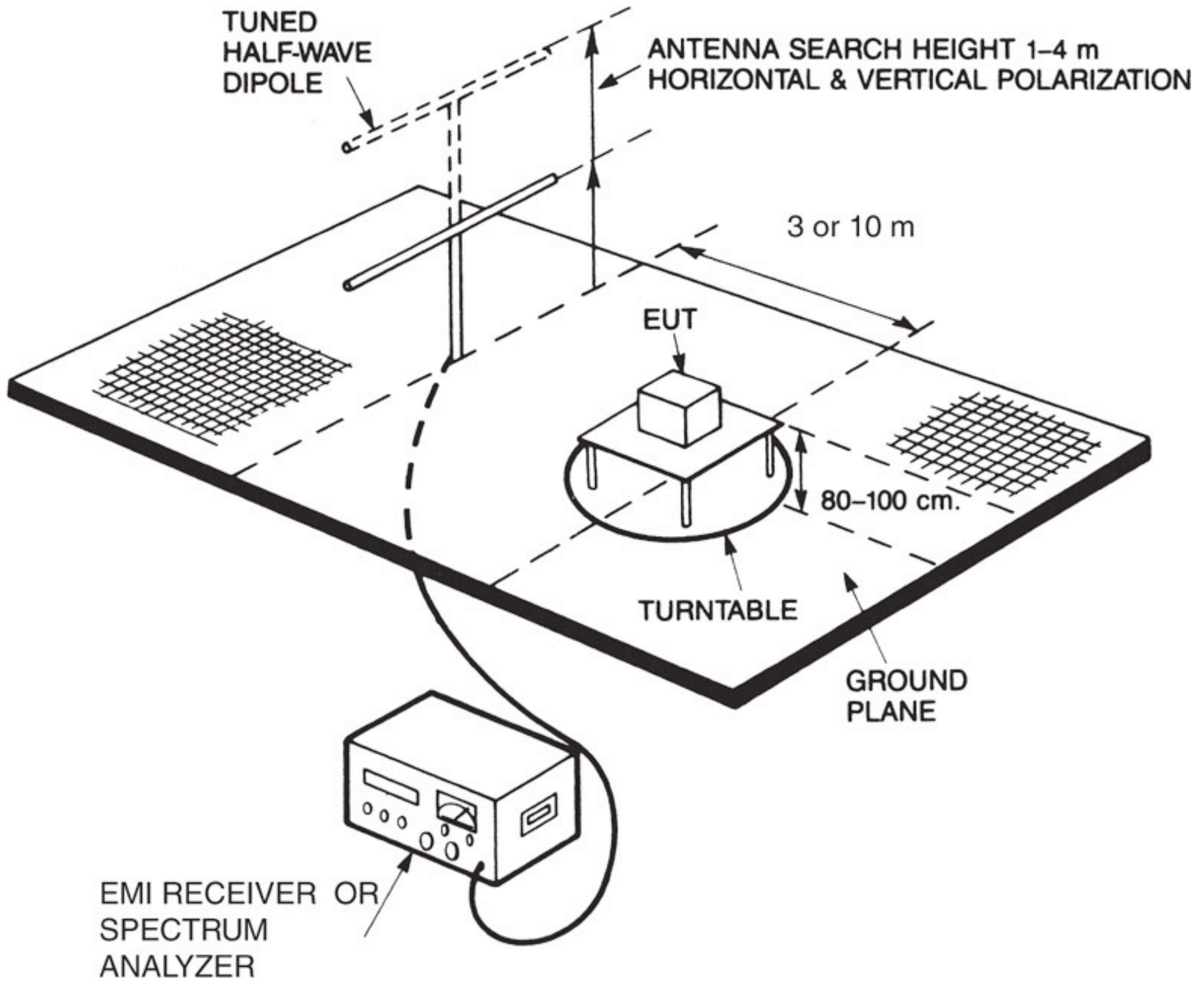


Figure 1-2: Open area test site (OATS) for FCC radiated emission test. The equipment under test (EUT) is on the turntable

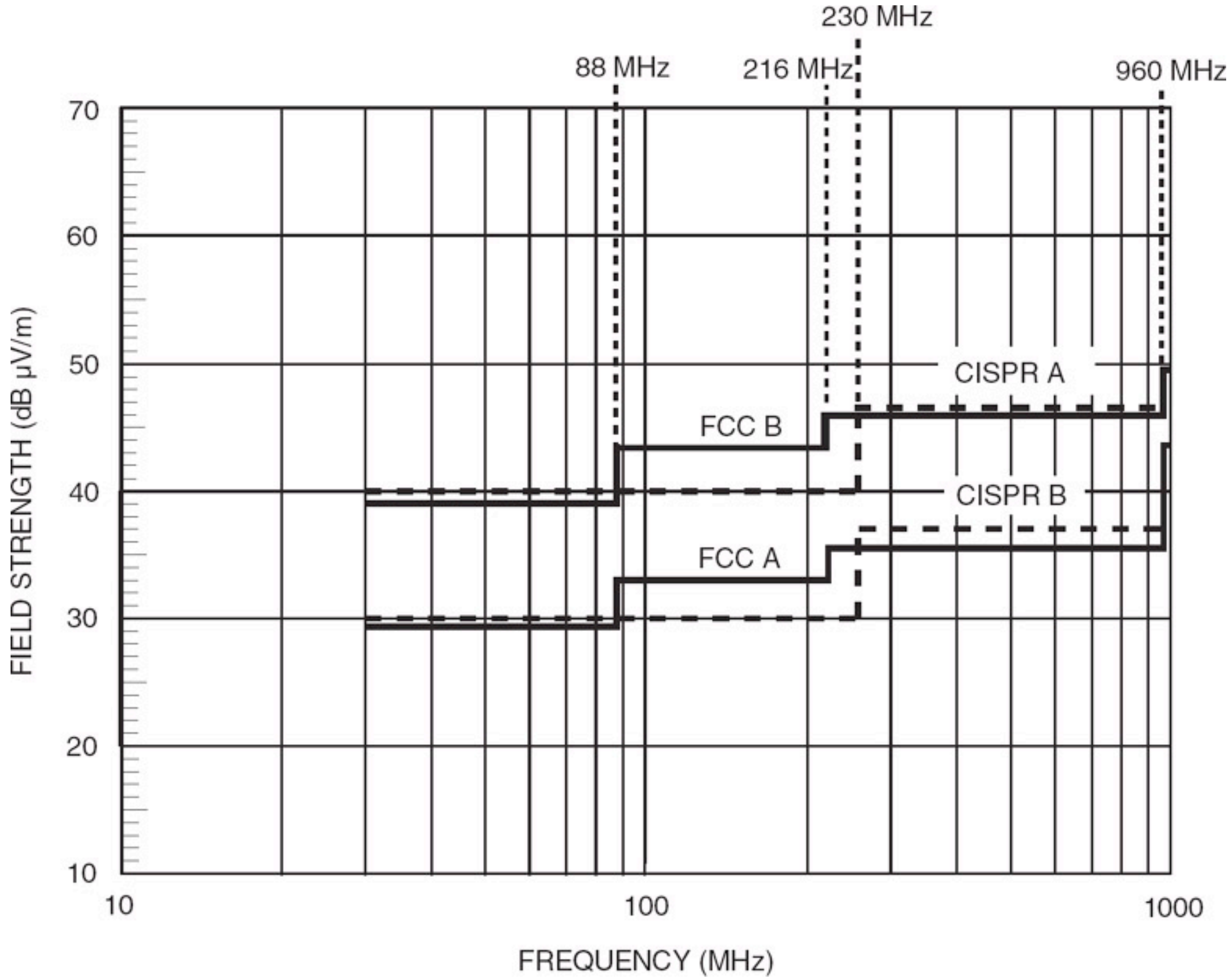


Figure 1-5: Comparison of FCC and CISPR radiated emission limits, measured at a distance of 10 m

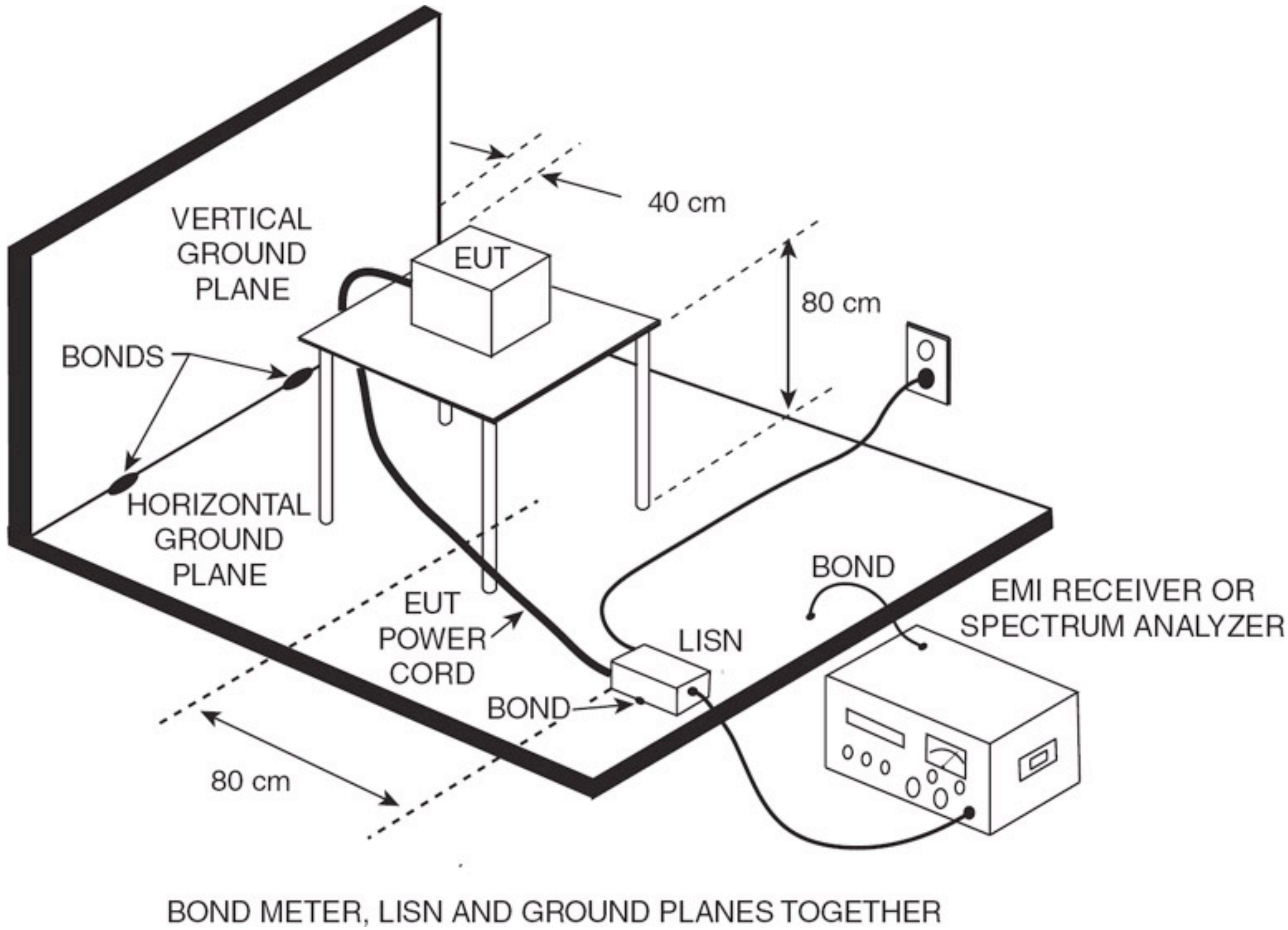


Figure 1-3: Test setup for FCC conducted emission measurements

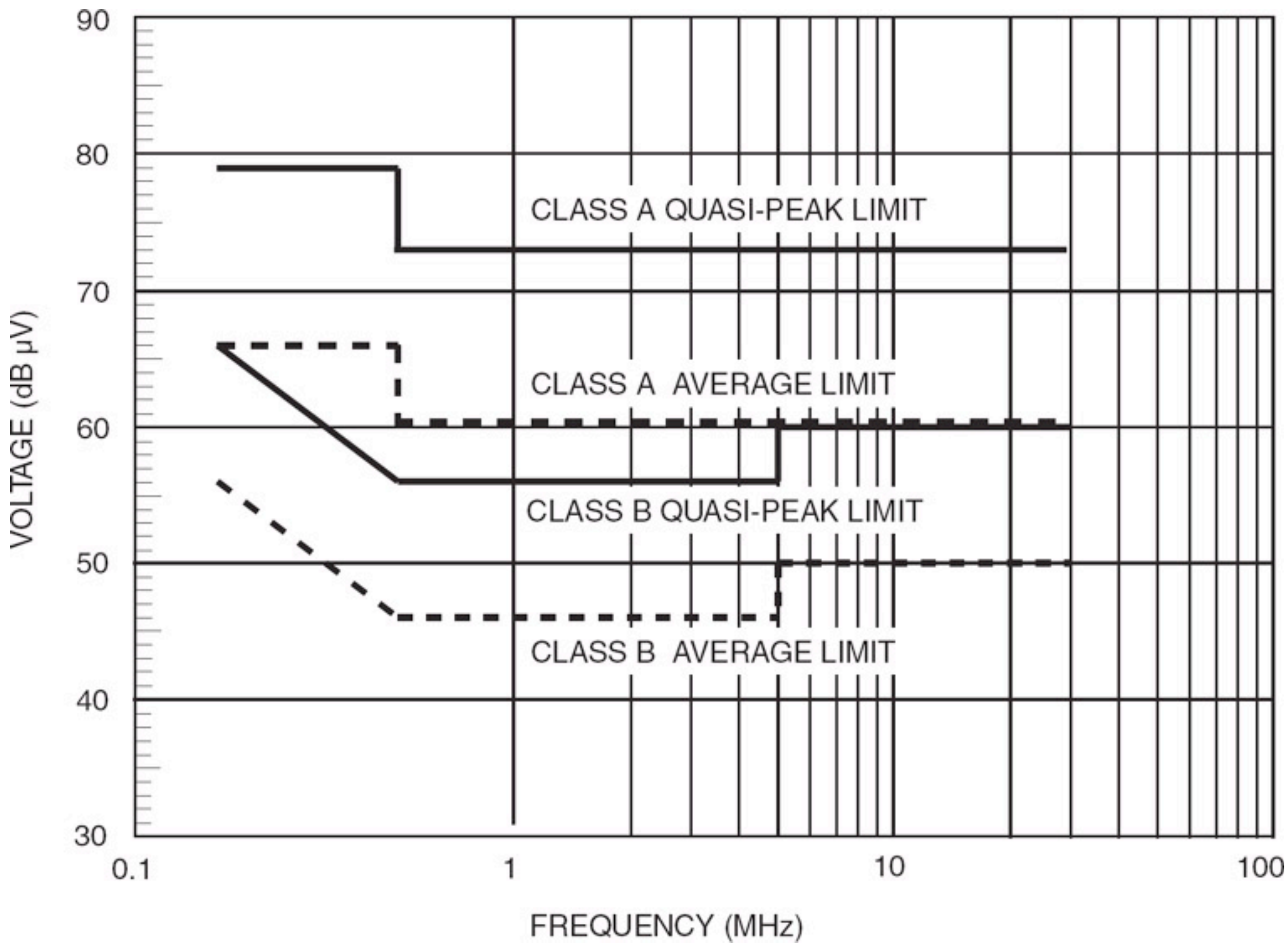


Figure 1-4: FCC/CISPR conducted emission limits

Cabling (chapter 2)

Cables are important because they are usually the longest parts of a system and therefore act as efficient antennas that pick up and/or radiate noise.

Assumptions:

- Shields are made of nonmagnetic materials and have a thickness much less than a skin depth at the frequency of interest.
- The receptor is not coupled so tightly to the source that it loads down the source.
- Induced currents in the receptor circuit are small enough not to distort the original field. (This does not apply to a shield around the receptor circuit.)
- Cables are short compared with a wavelength, so the coupling between circuits can be represented by lumped capacitance and inductance between the conductors. The circuit can then be analyzed by normal network theory.

Three types of couplings are considered.

1. Capacitive or electric coupling, which results from the interaction of electric fields between circuits.
2. Inductive, or magnetic, coupling, which results from the interaction between the magnetic fields of two circuits.
3. Combination of electric and magnetic fields and is appropriately called electromagnetic coupling or radiation.

Cabling Strategies: Capacitive (Electric Field) Coupling

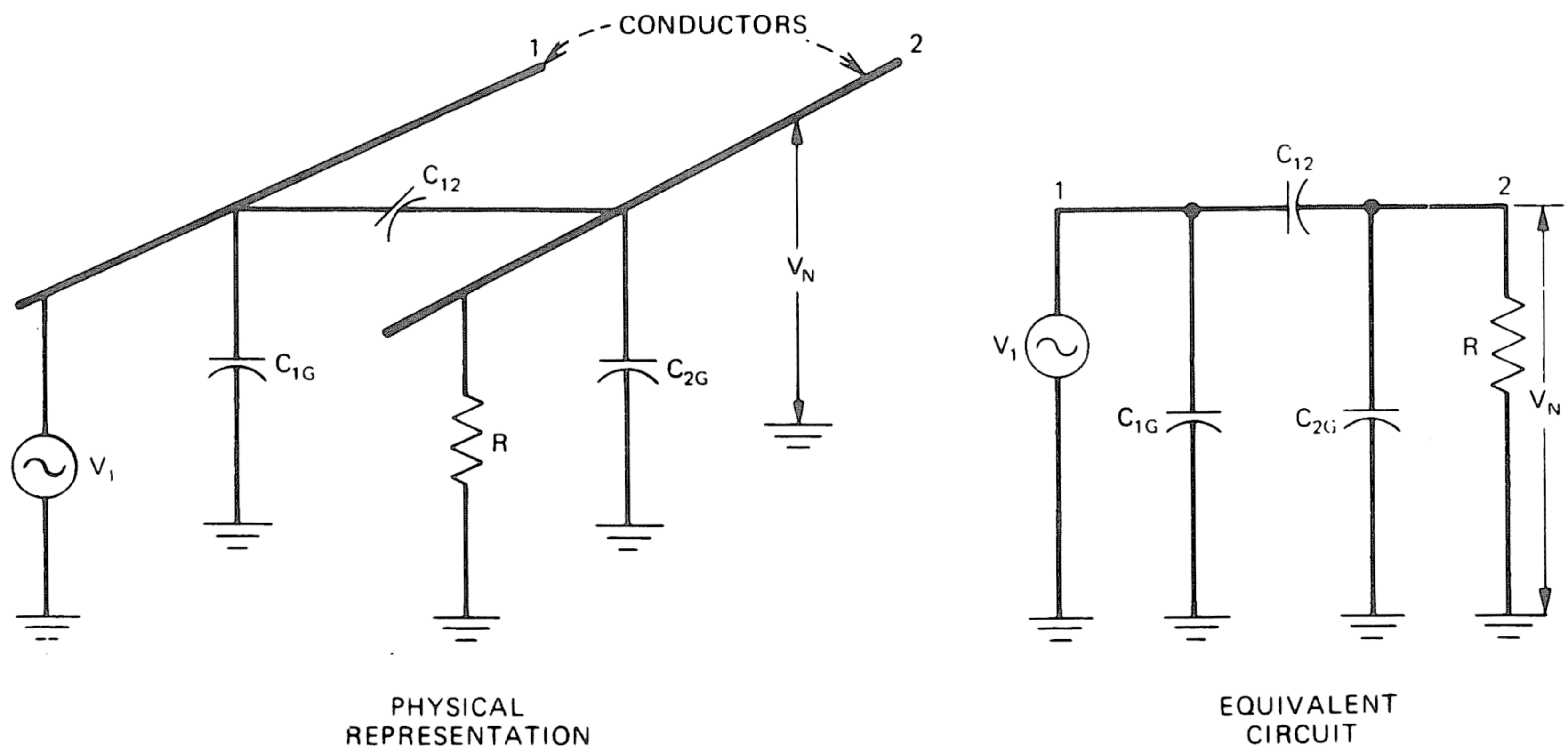


Figure 2-1. *Capacitive coupling between two conductors.*

Above corner frequency, constant fraction of interfering signal coupled

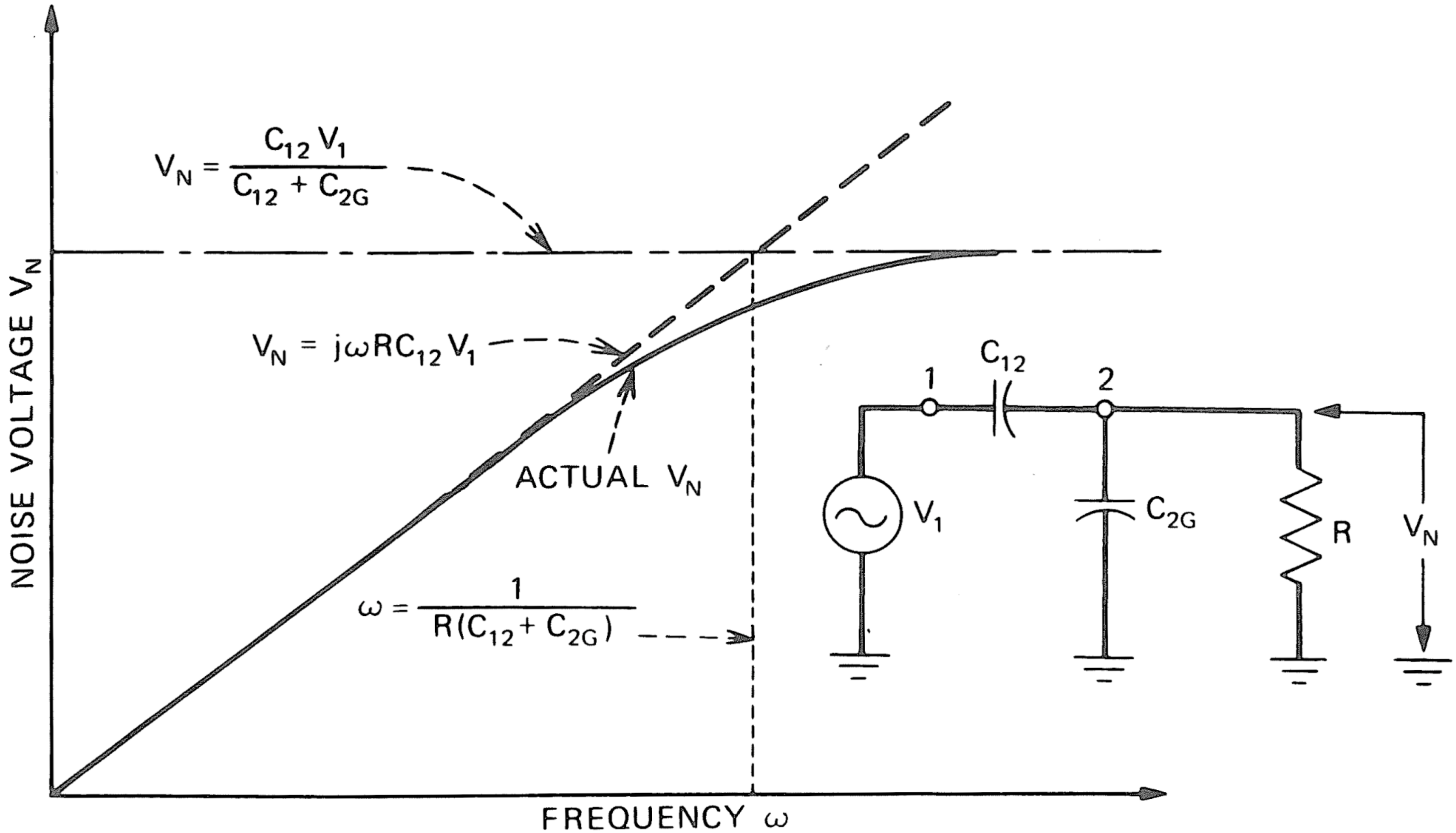


Figure 2-3. *Frequency response of capacitive coupled noise voltage.*

Limited effectiveness of separating conductors

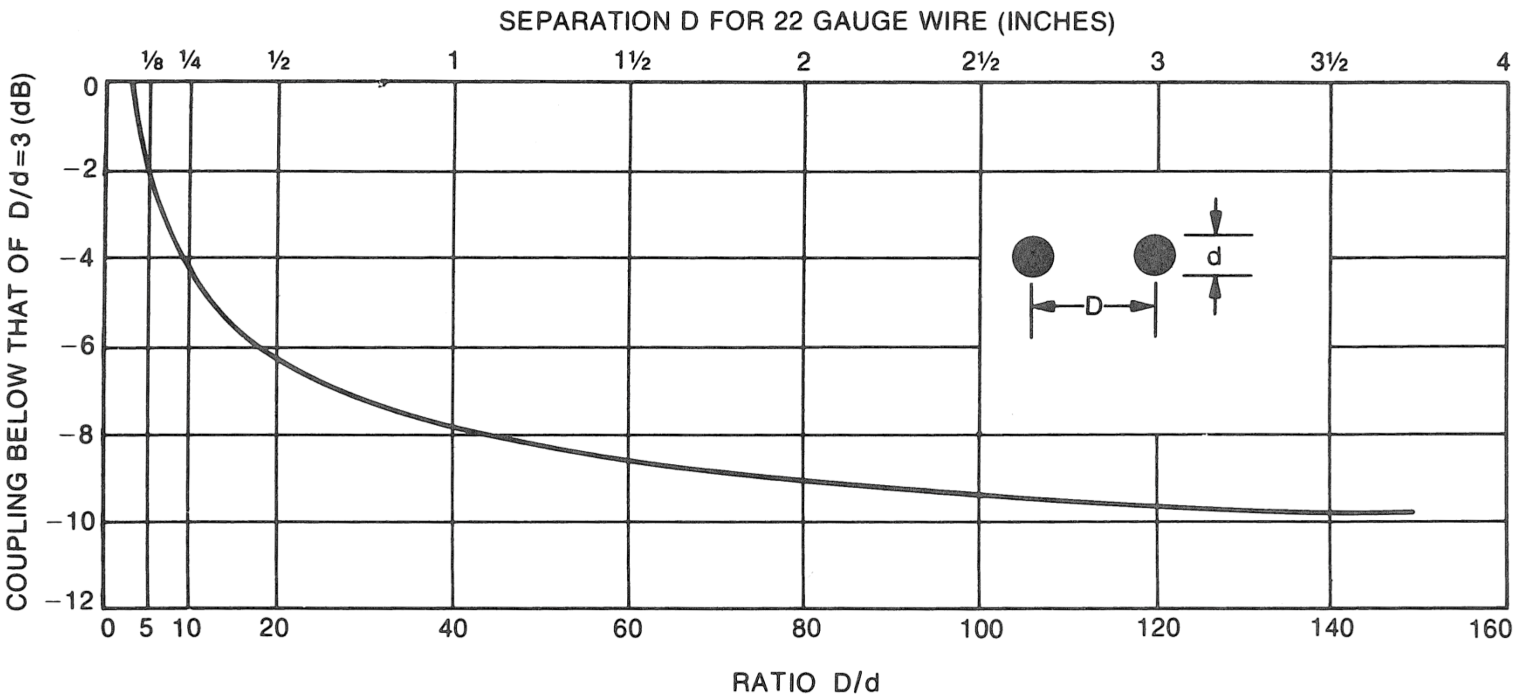


Figure 2-2. *Effect of conductor spacing on capacitive coupling. In the case of 22-gauge wire, most of the attenuation occurs in the first inch of separation.*

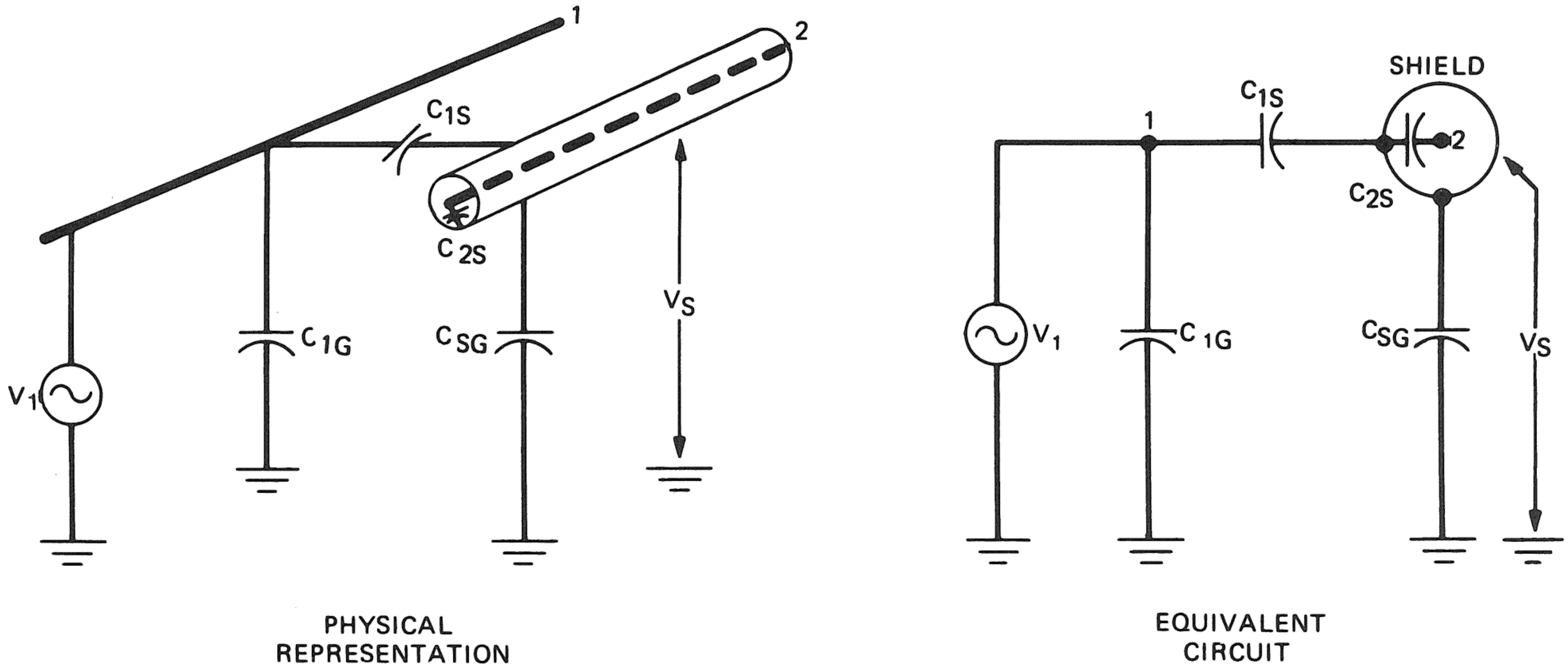
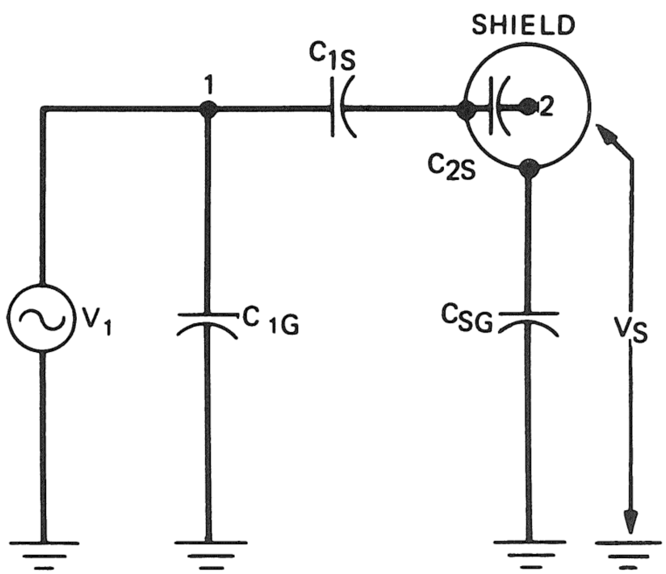


Figure 2-4. *Capacitive coupling with shield placed around receptor conductor.*

One minute quiz: Coupled signal from 60Hz, 1MHz interference sources to the center conductor of a coaxial cable:

In the circuit below, C_{1S} = 5pF, C_{SG} = 20pF, C_{2S} = 1000pF. We observe the voltage V_2 on the center conductor with a 10MΩ || 10pF scope probe.

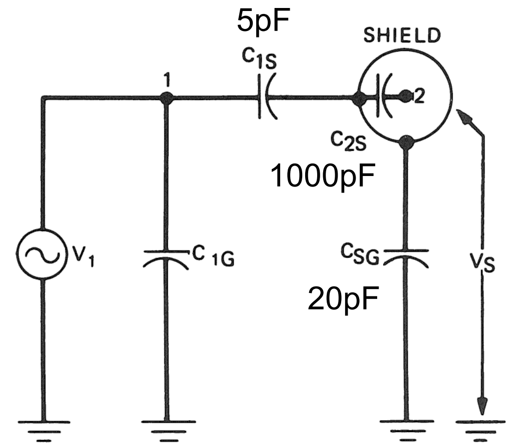


a) Determine the transfer function from v_1 to v_2.

b) Determine the V_2 waveform when v_{in} is a 170V pk sine wave at 60Hz

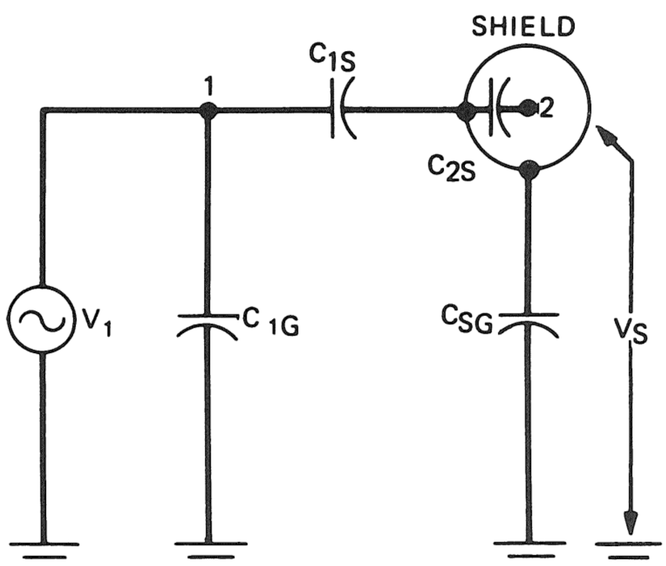
c) Determine the V_2 waveform when v_{in} is a 1V pk square wave at 1 MHz

Coupled signal from 60Hz, 1MHz interference sources to the center conductor of a coaxial cable:



One minute quiz: Coupled signal from 60Hz, 1MHz interference sources to the center conductor of a coaxial cable WITH SHIELD GROUNDED:

In the circuit below, C_{1S} = 5pF, C_{SG} = 20pF, C_{2S} = 1000pF. TO REDUCE INTERFERENCE COUPLED THROUGH THE ELECTRIC FIELD, THE SHIELD IS CONNECTED TO GROUND. We observe the voltage V_2 on the center conductor with a 10MΩ scope probe.

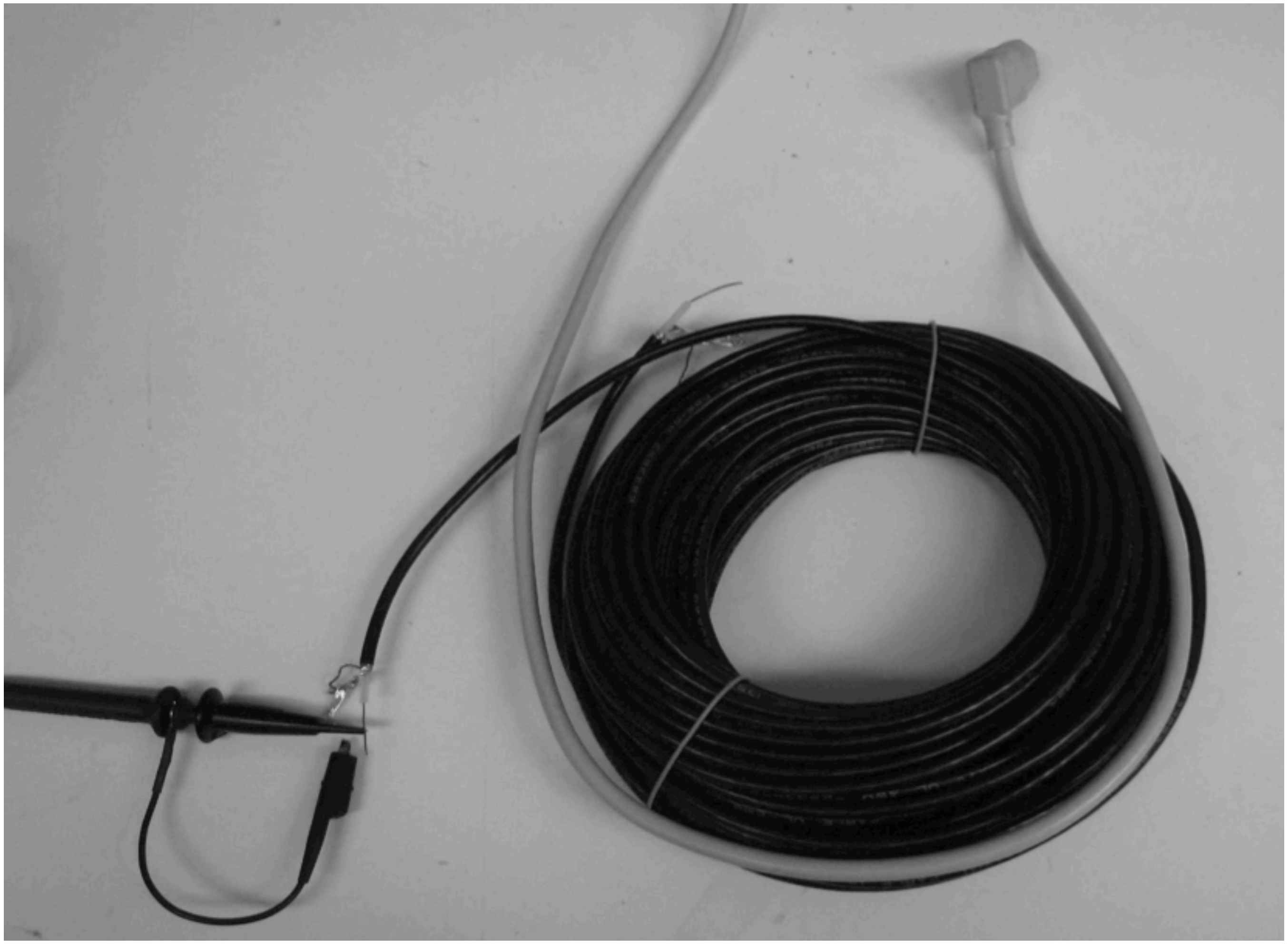


a) Determine the transfer function from v_1 to v_2.

b) Determine the V_2 waveform when v_{in} is a 170V pk sine wave at 60Hz

c) Determine the V_2 waveform when v_{in} is a 1V pk square wave at 1 MHz

Experimental Configuration:



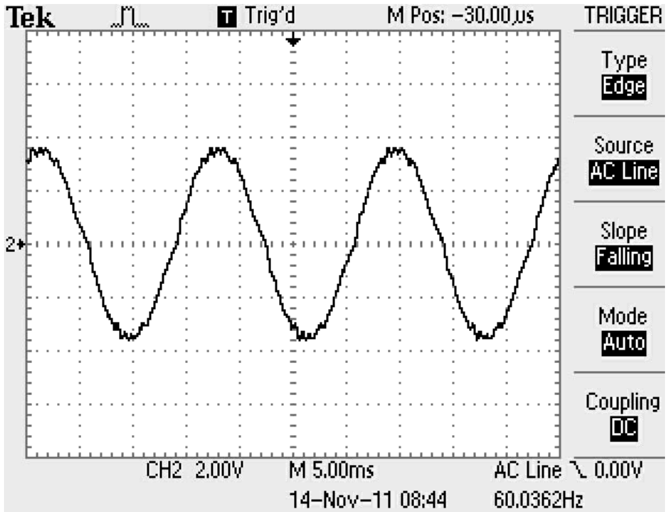
Experimental Results:

Shield Floating

V_1 =
60 Hz
170Vpk

Scope:

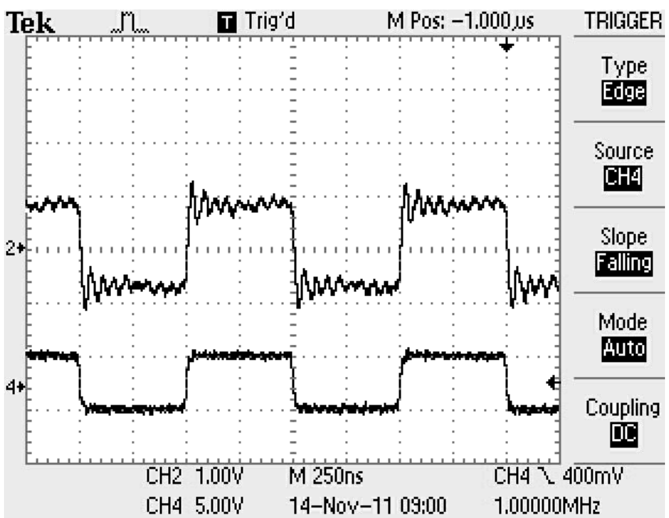
V_2
2V/div



Scope:

V_2
1V/div

V_1
1MHz
5Vpk

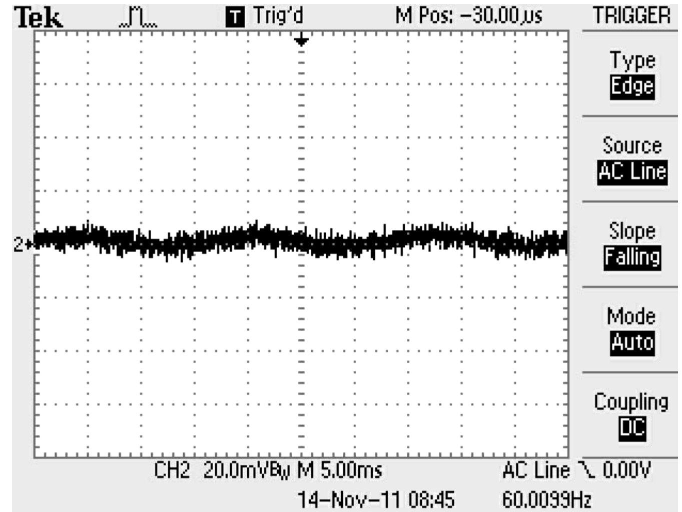


Shield grounded at one point

V_1 =
60 Hz
170Vpk

Scope:

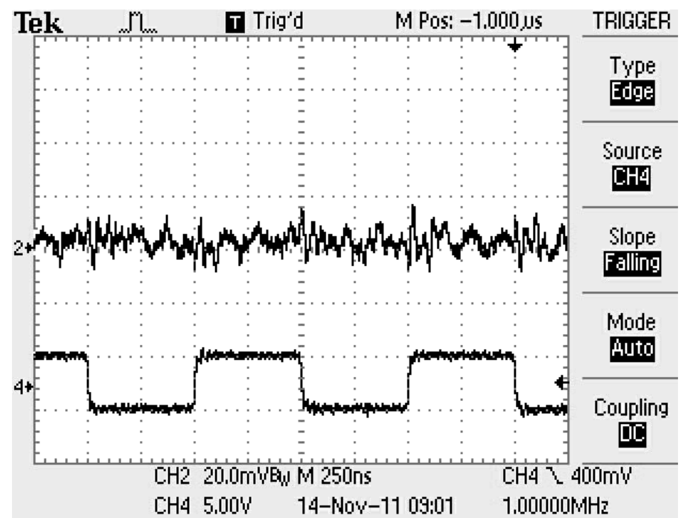
V_2
20mV/div



Scope:

V_2
20mV/div

V_1
1MHz
5Vpk



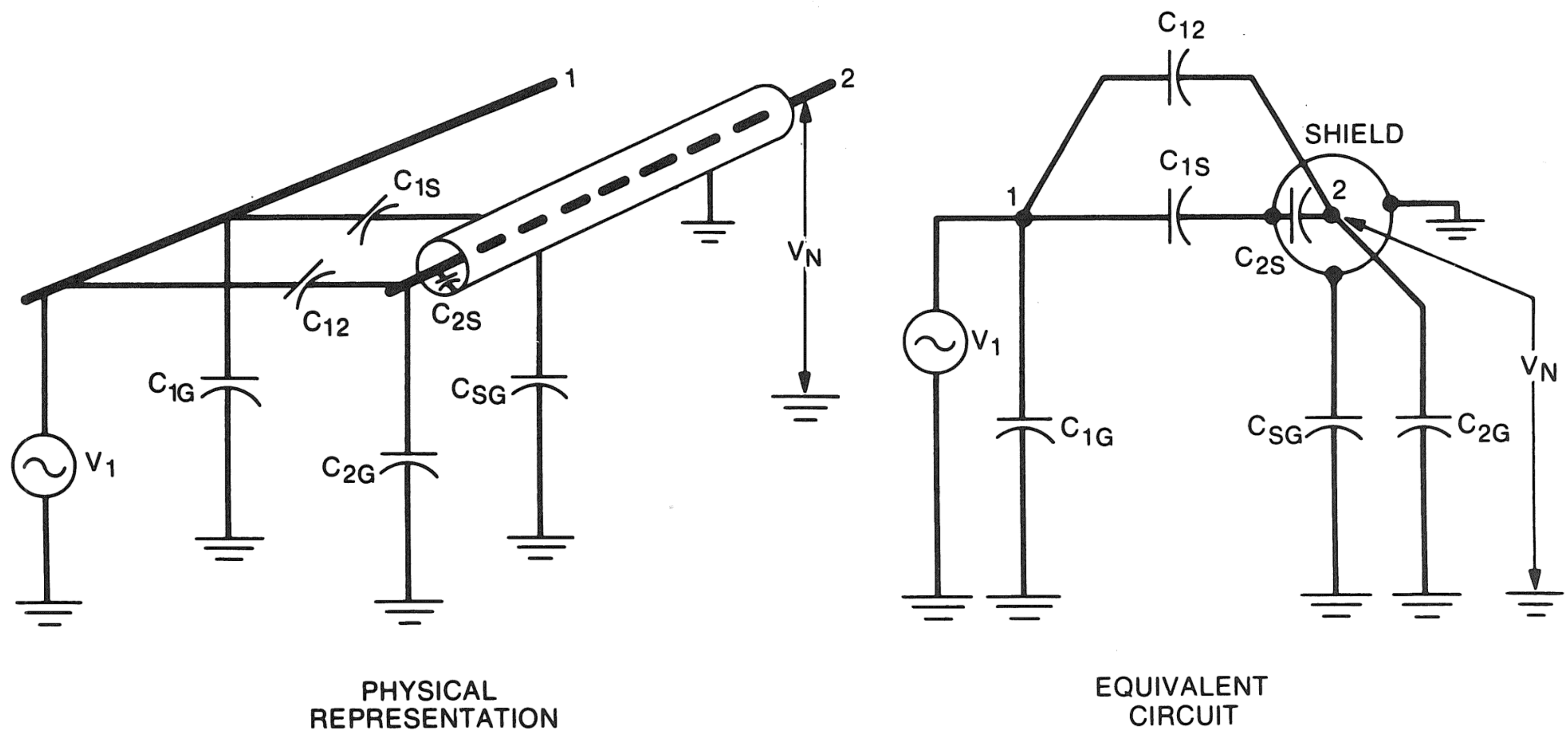


Figure 2-5. *Capacitive coupling when center conductor extends beyond shield; shield grounded at one point.*

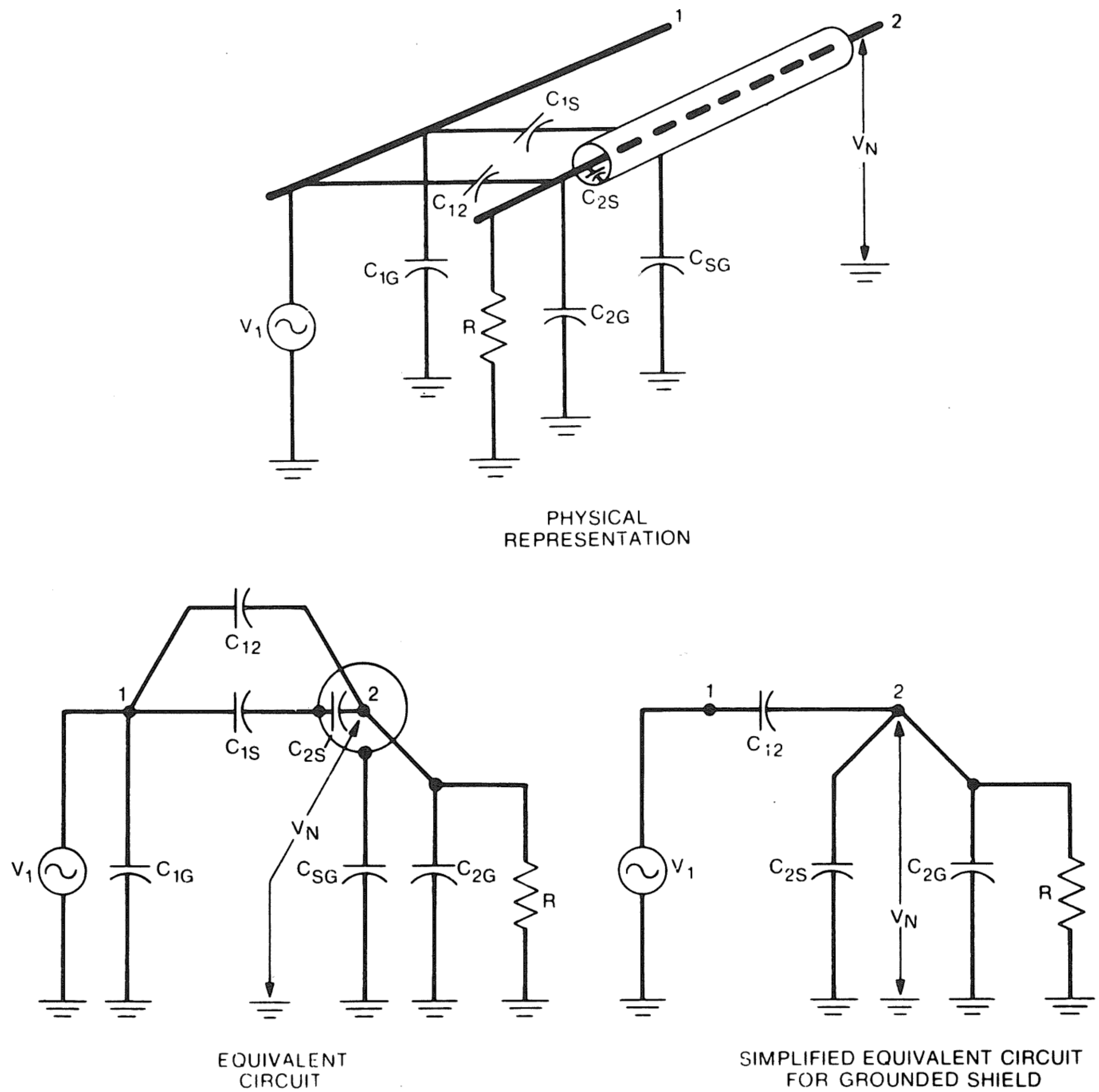


Figure 2-6. *Capacitive coupling when receptor conductor has resistance to ground.*

Magnetic field coupling

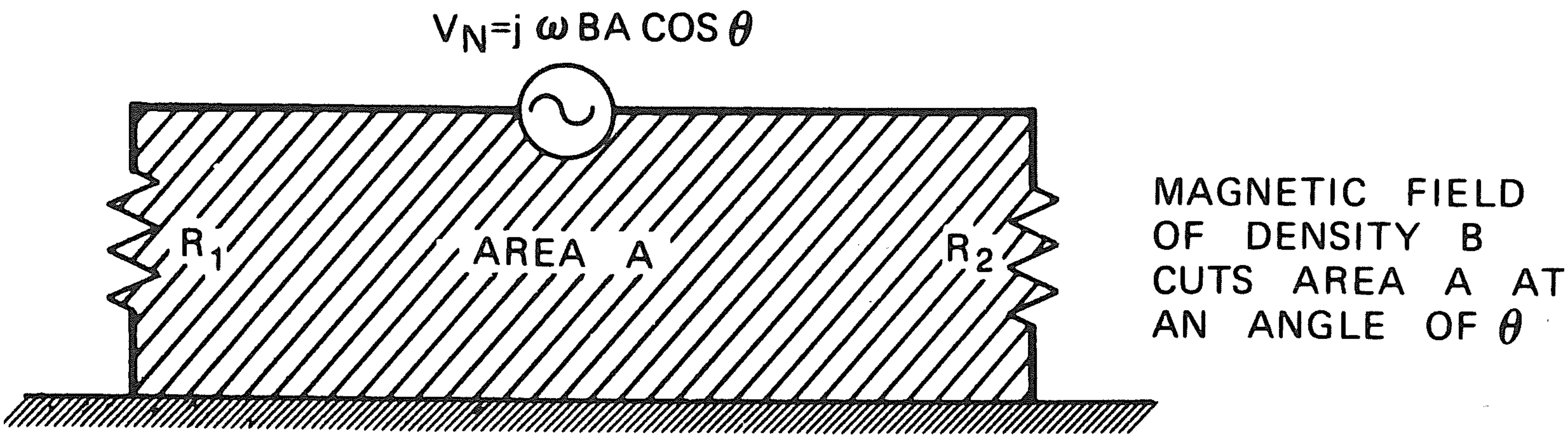


Figure 2-7. *Induced noise depends on the area enclosed by the disturbed circuit.*

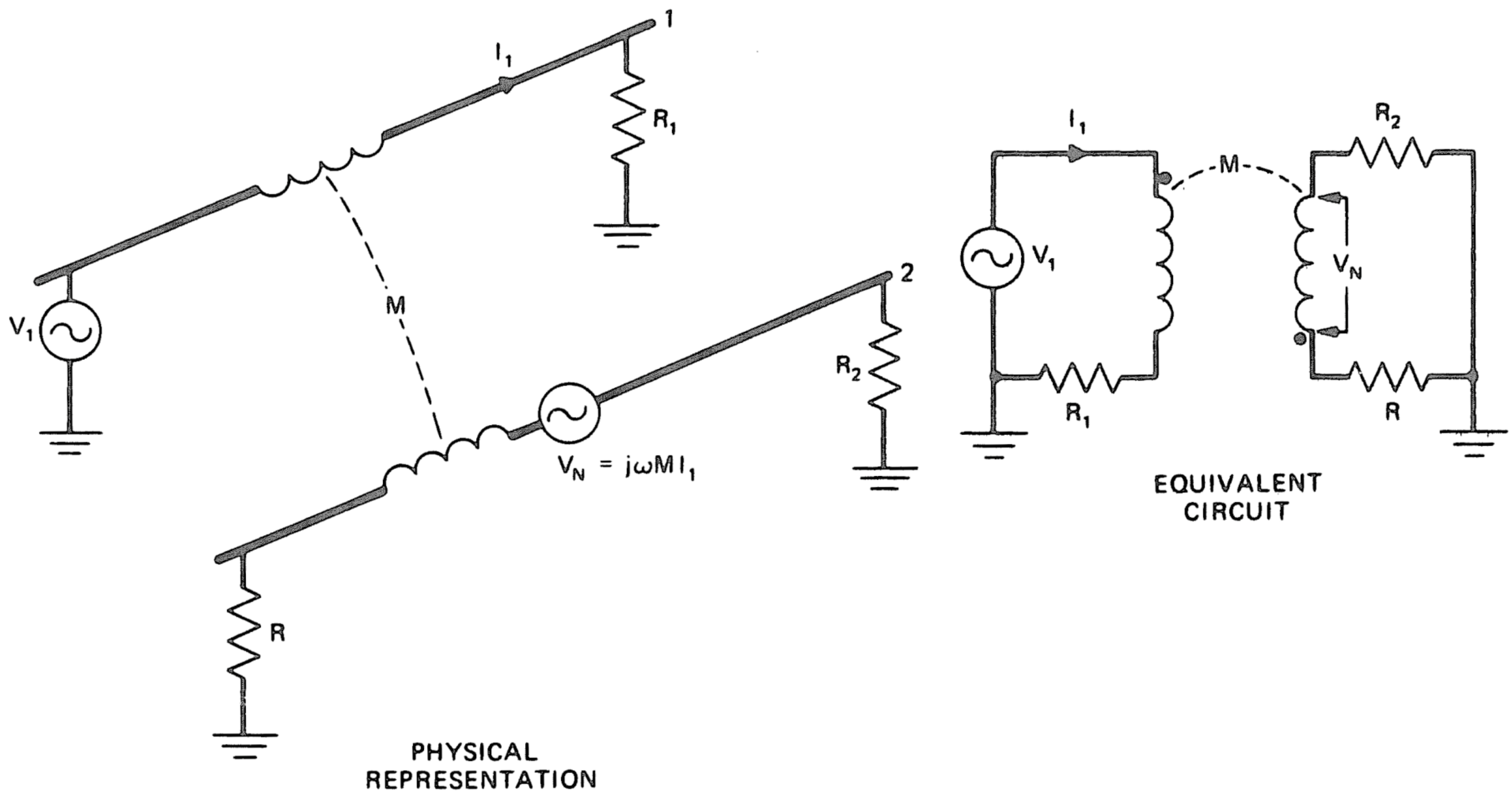


Figure 2-8. *Magnetic coupling between two circuits.*

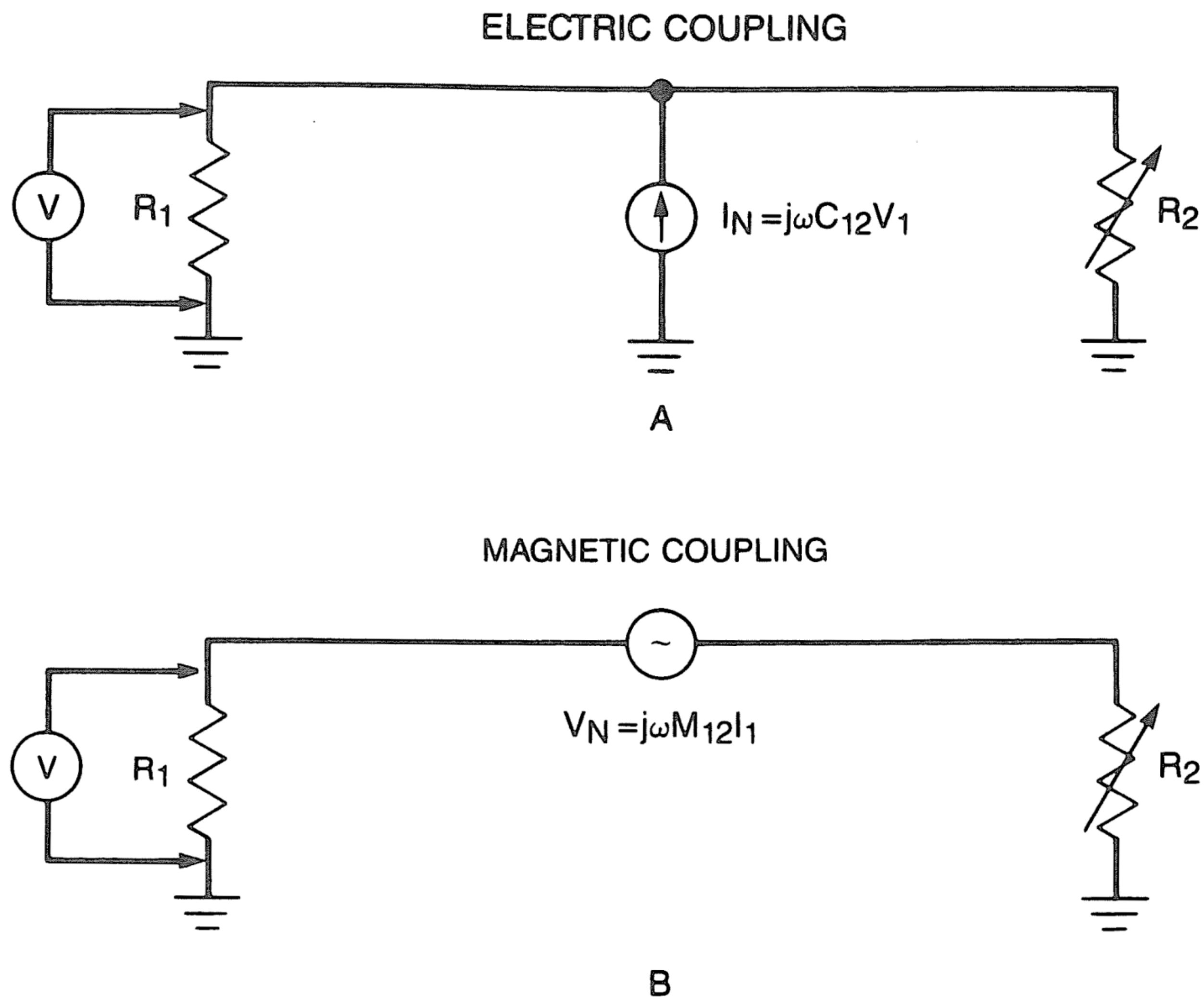
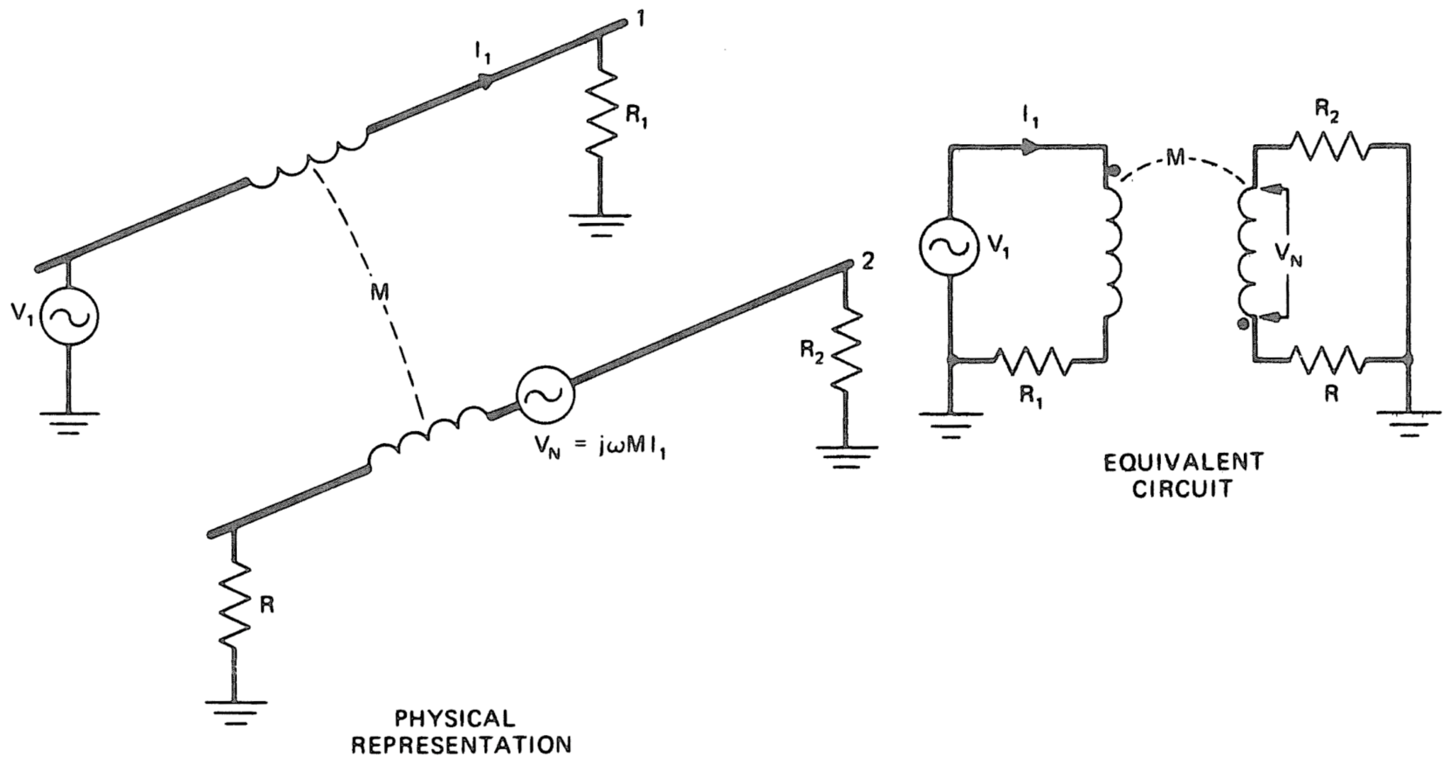


Figure 2-9. (*A*) *Equivalent circuit for electric field coupling*; (*B*) *equivalent circuit for magnetic field coupling.*

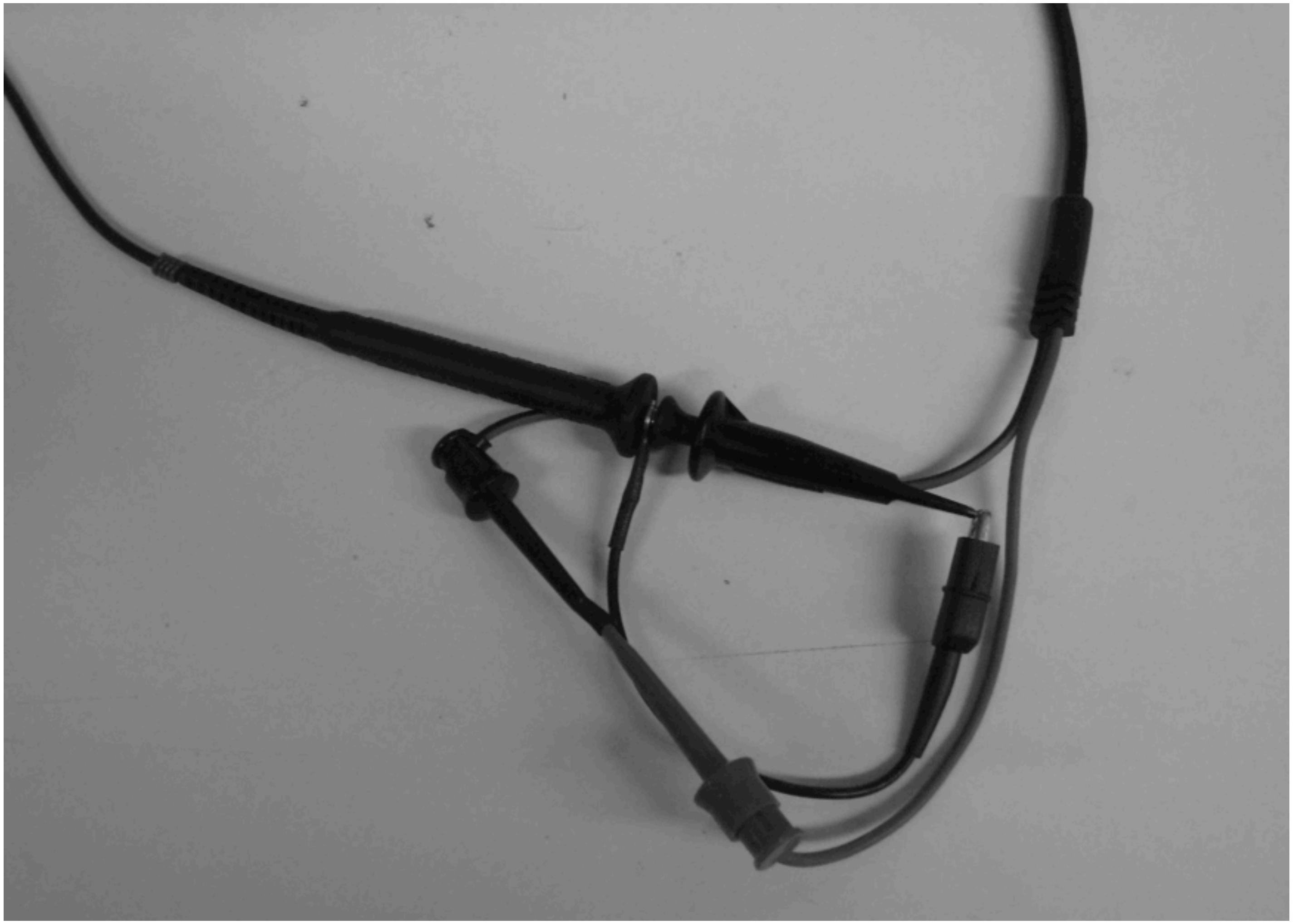
One minute quiz: Magnetic field coupling, adjacent loops

In the circuit below, i_1 is a 10MHz sinusoid with peak amplitude of 50mA. We observe the voltage V_N with a 1MΩ scope probe with R=0Ω and R_2 = 1MΩ. The mutual inductance is M = 30 nHy.

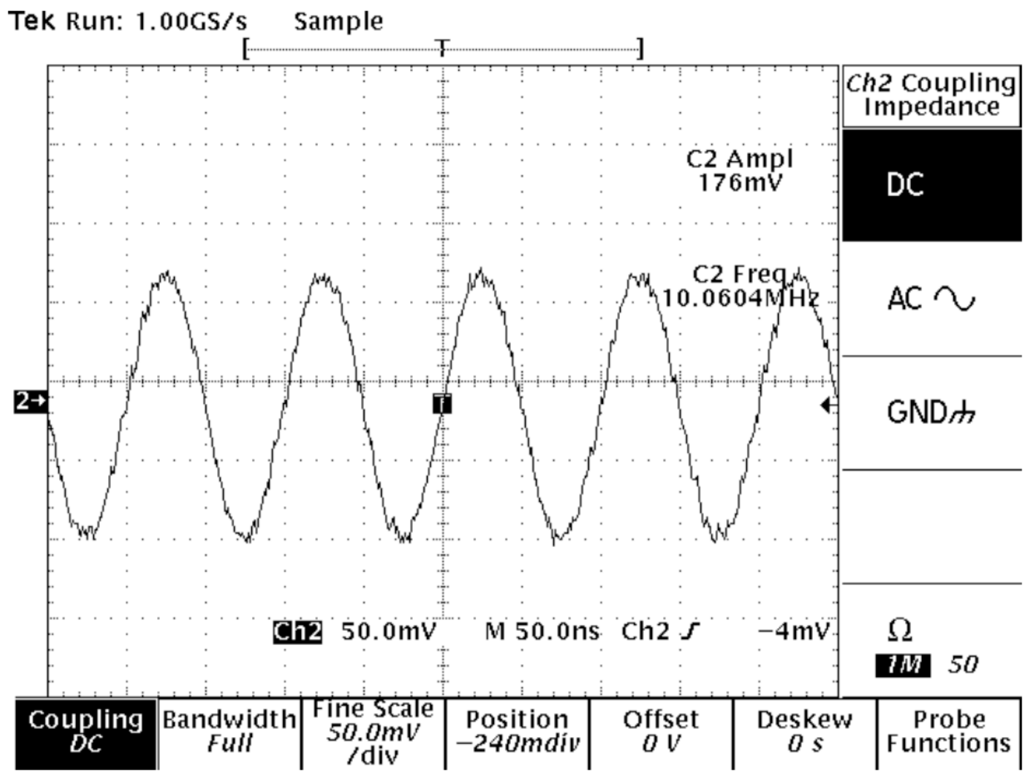


Determine the V_N waveform coupled by the shared magnetic field.

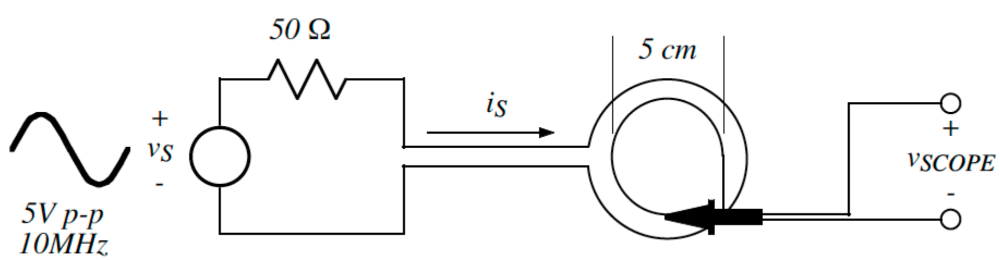
Experimental Configuration:



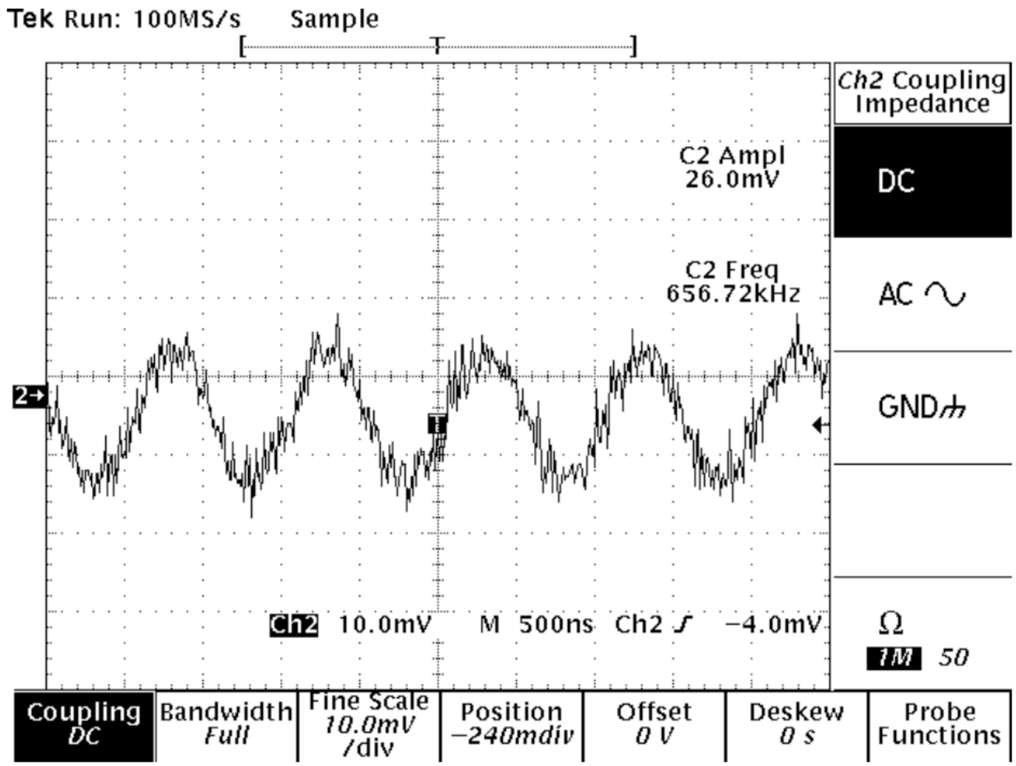
Experimental Results:



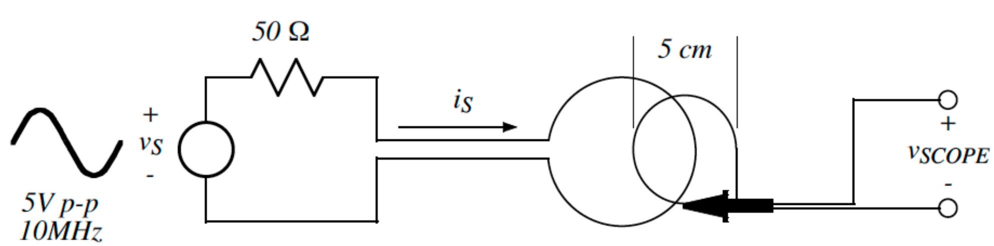
V_{SCOPE}, Scope probe loop enclosed by current loop.



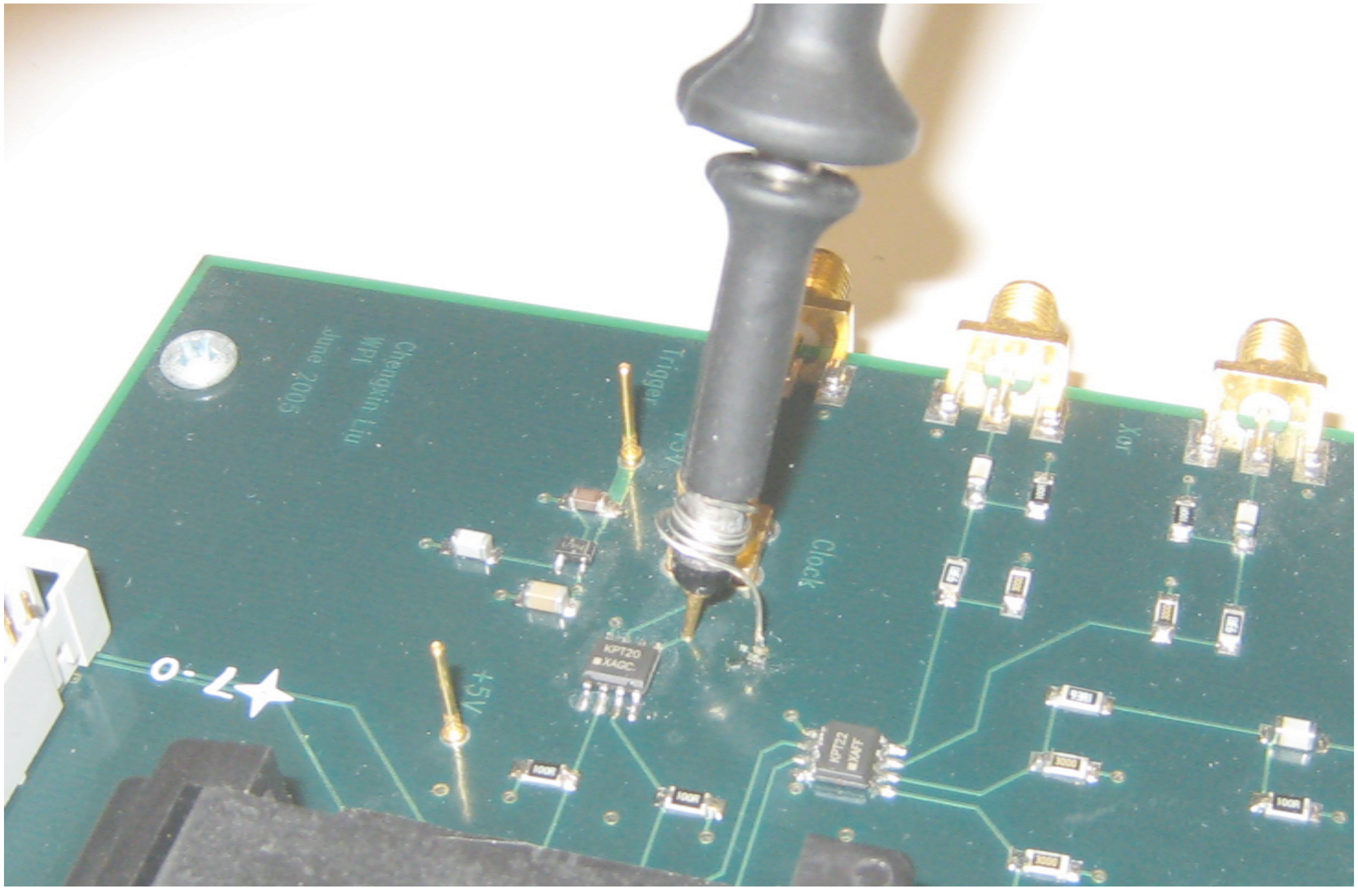
Experimental Results: Reducing loop area ⇒ reduces mutual inductance ⇒ reduces coupled interference



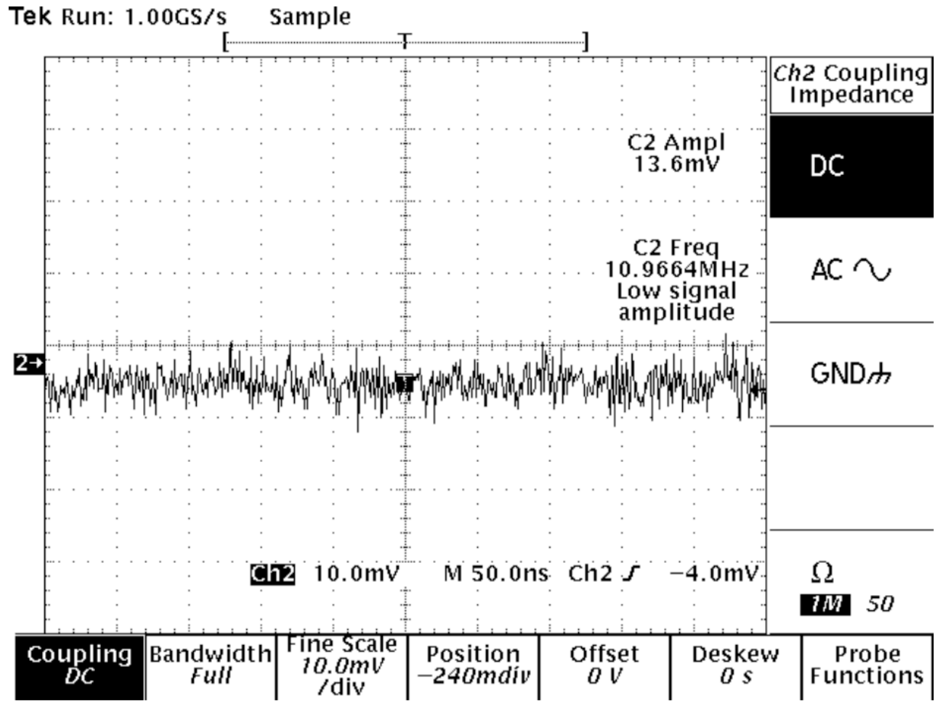
V_{SCOPE}, Scope probe loop not fully linking current loop



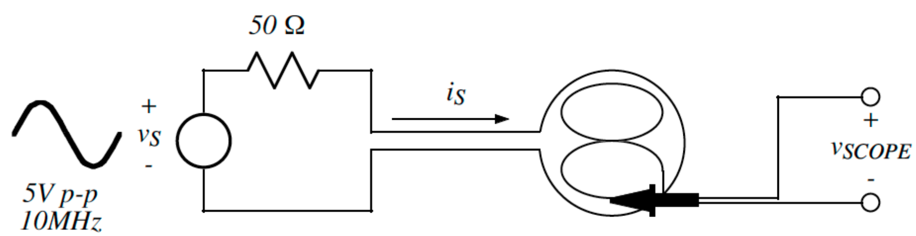
Scope probing “tip” to reduce inductance when probing high speed signals:



Experimental Results: Twisting receiving loop ⇒ changes sign of flux coupled ⇒ partial cancellation of flux ⇒ reduces mutual inductance ⇒ reduces coupled interference



V_{SCOPE}, Scope probe twisted



Calculating mutual inductance

$$\theta_{12} = \int_a^b \frac{\mu I_1}{2\pi r}\, dr = \frac{\mu I_1}{2\pi} \ln\left(\frac{b}{a}\right).$$

$$\theta_{12} = \left[\frac{\mu}{\pi} \ln\left(\frac{b}{a}\right)\right] I_1 .$$

$$M = 4 \times 10^{-7} \ln\left(\frac{b}{a}\right).$$

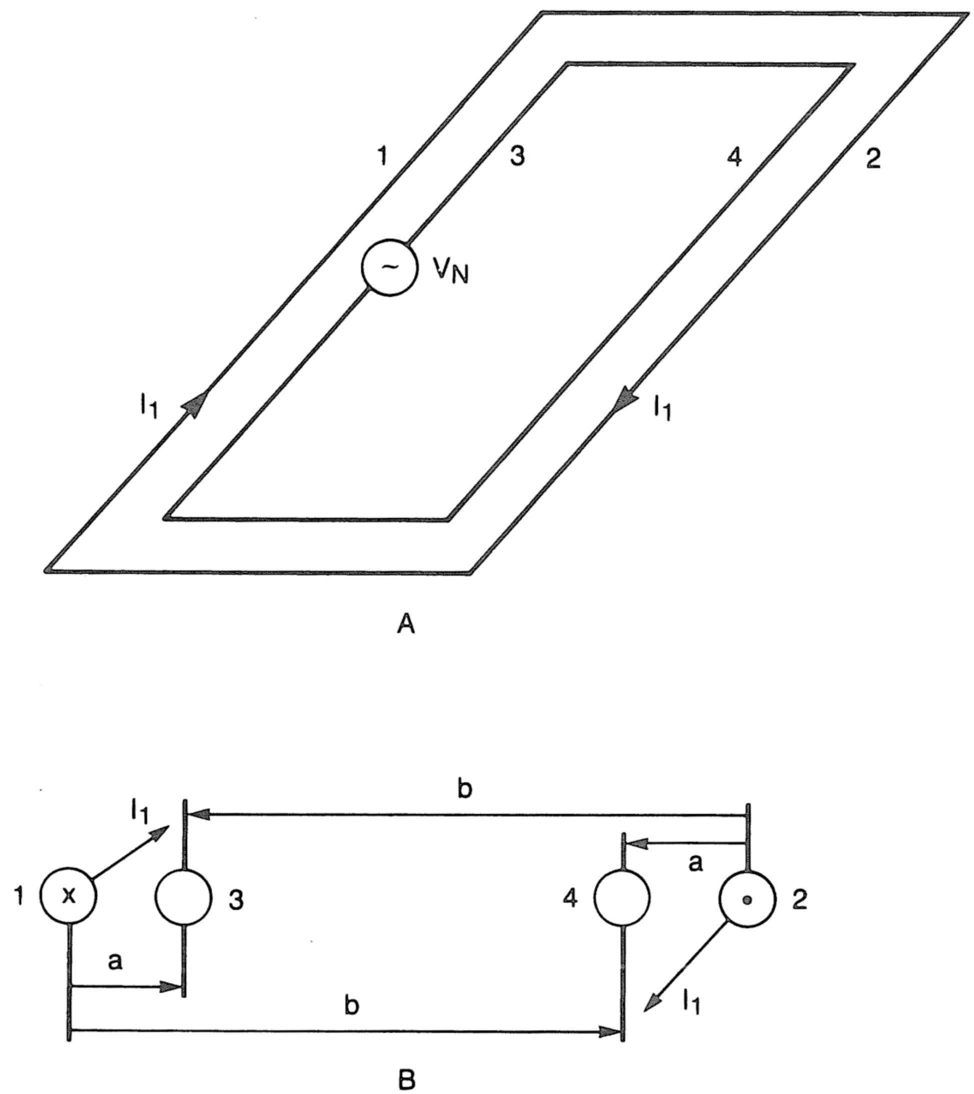


Figure 2-10. *(A) Nested coplanar loops. (B) cross-sectional view of A.*

Effect of ungrounded, nonmagnetic shield: Geometry, medium not changed!

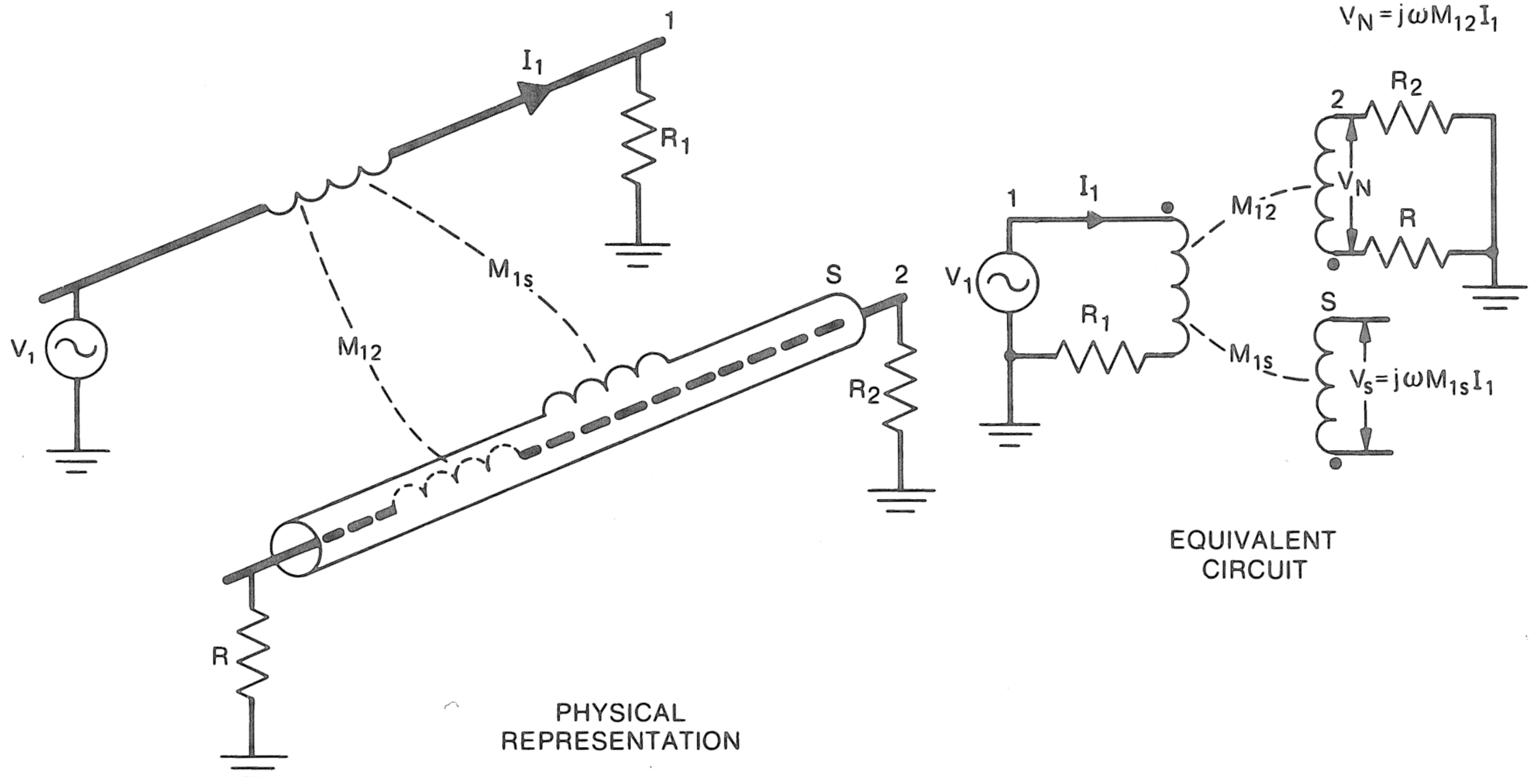


Figure 2-11. *Magnetic coupling when a shield is placed around the receptor conductor.*

A ground connection on one end of the shield does not change the situation. *It follows therefore that a nonmagnetic shield placed around a conductor and grounded at one end has no effect on the magnetically induced voltage in that conductor.*

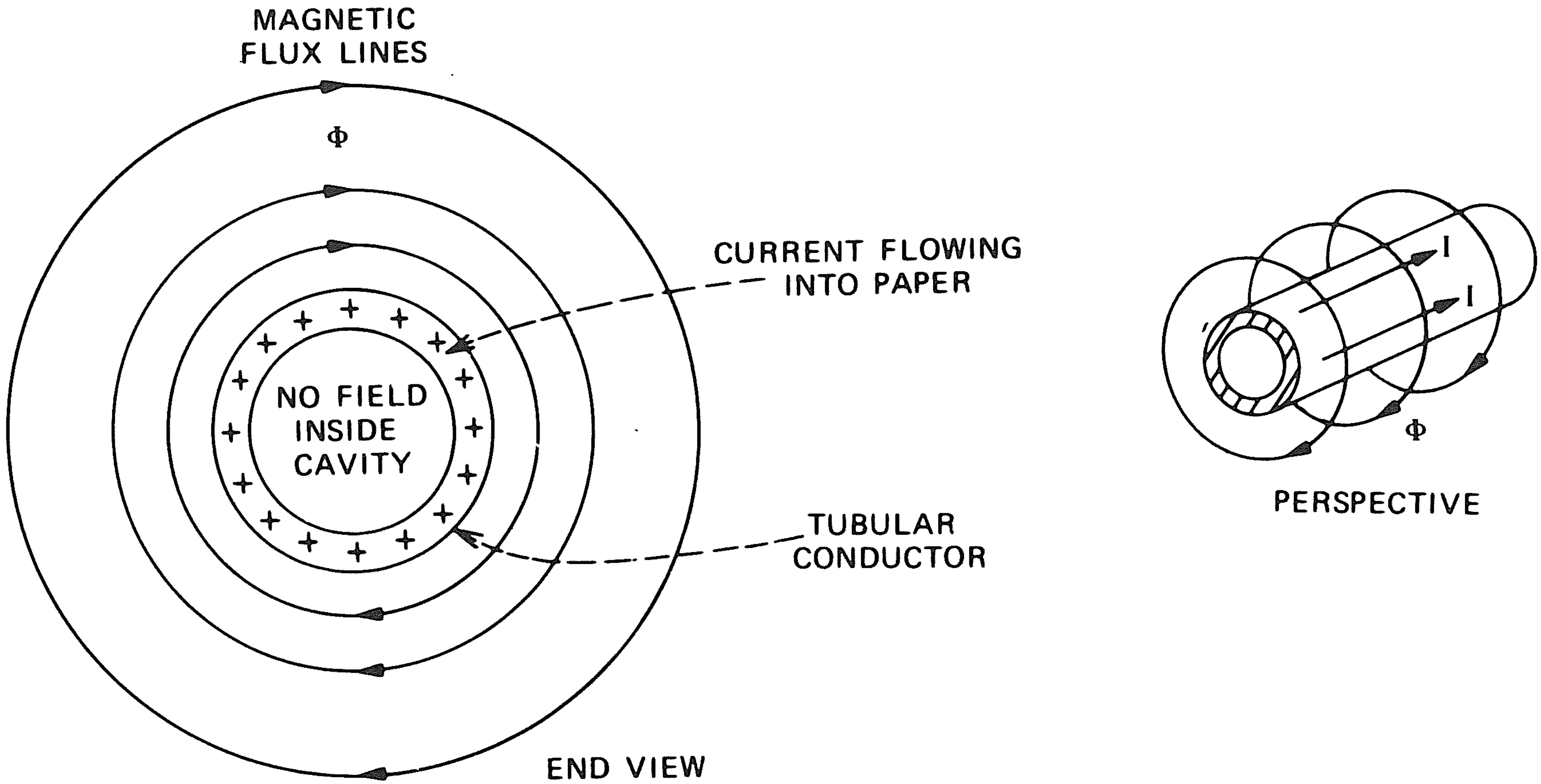


Figure 2-12. *Magnetic field produced by current in a tubular conductor.*

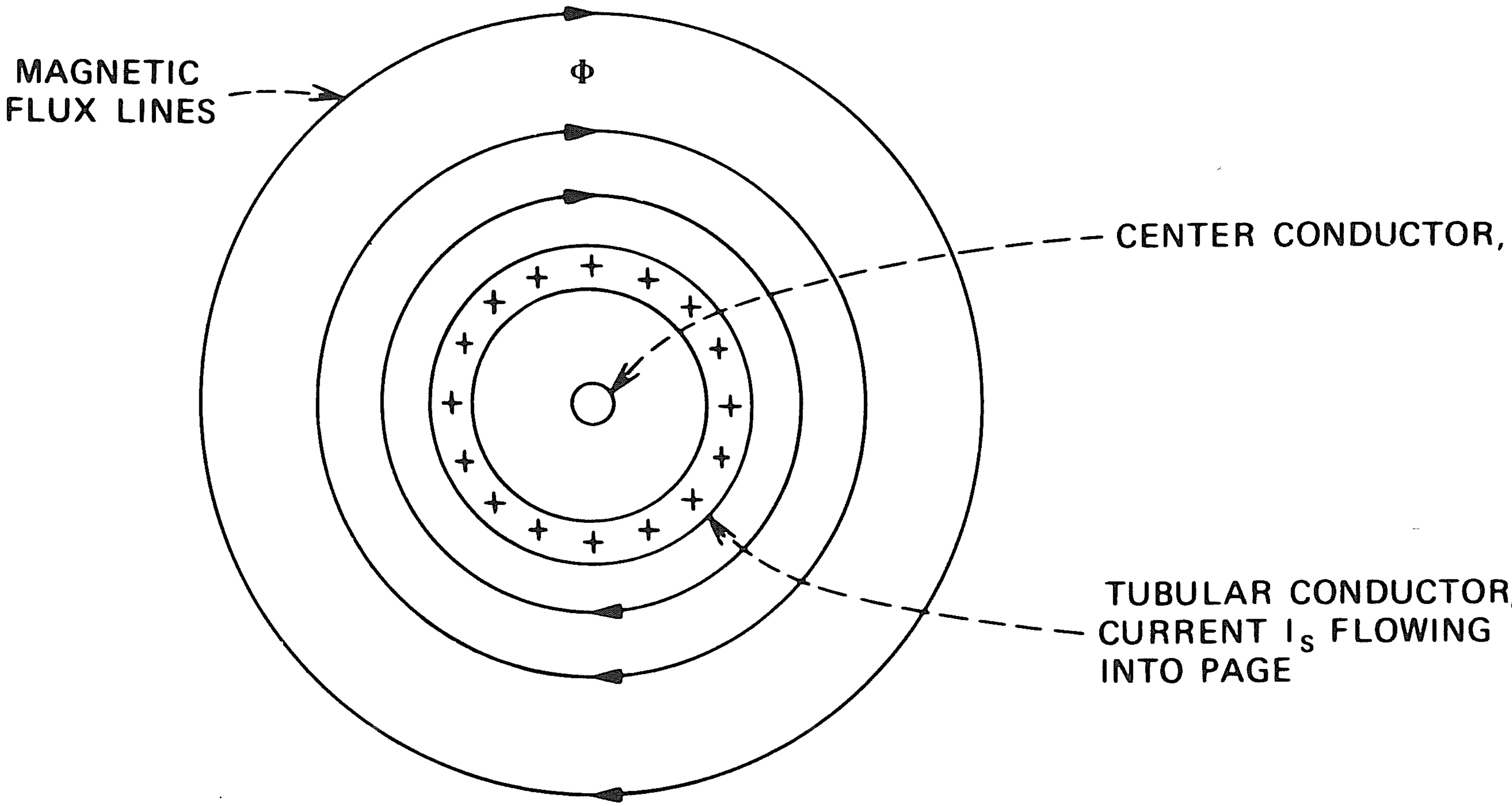
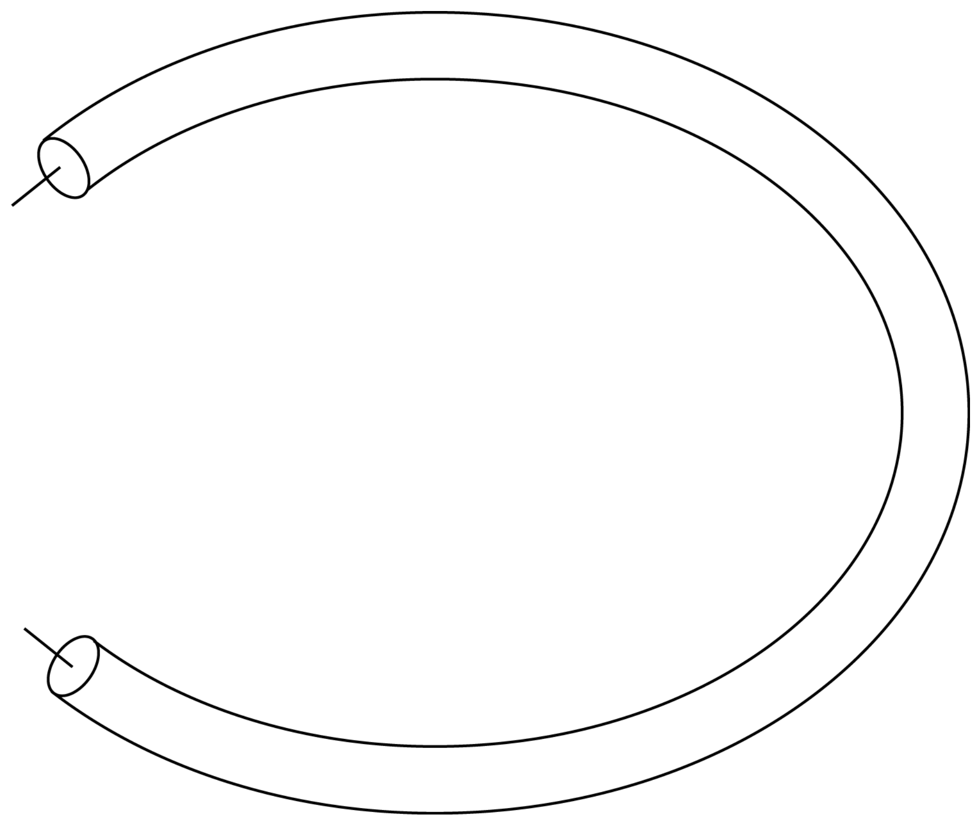


Figure 2-13. *Coaxial cable with shield current flowing.*

No field where center conductor is, so no problem, right? WRONG!!

Remember that important effect for magnetic field is field lines through a loop.

Think about loop of coaxial cable: If any current flows in shield, inner conductor loop also sees all of magnetic field lines



$$V_N = j\omega M I_S \ .$$

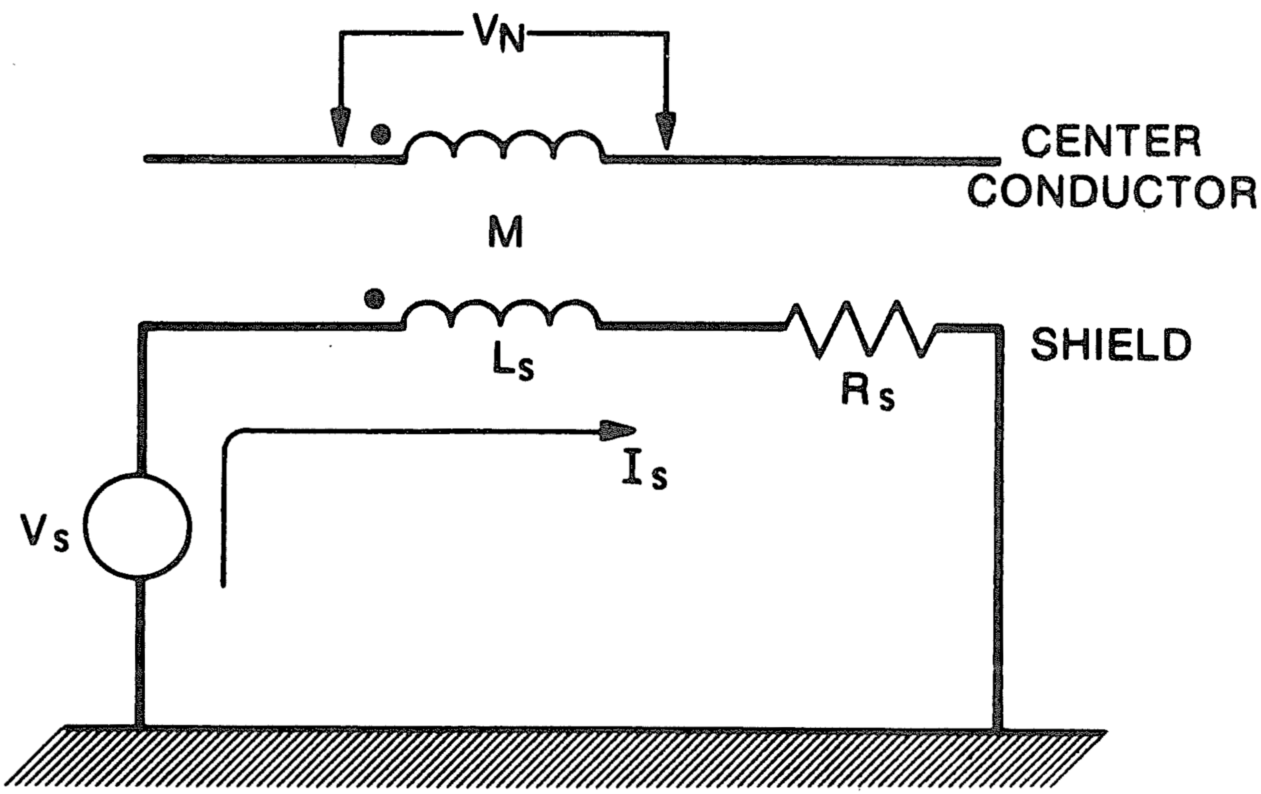


Figure 2-14. *Equivalent circuit of shielded conductor.*

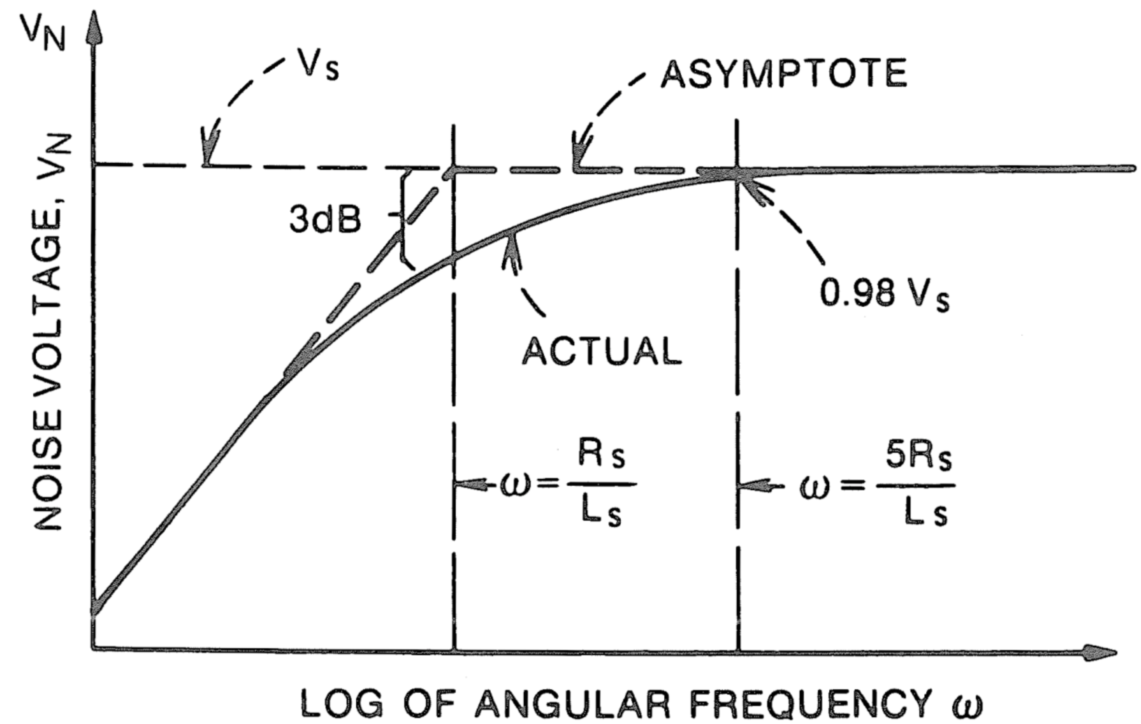


Figure 2-15. *Noise voltage in center conductor of coaxial cable due to shield current.*

Table 2-1 Measured Values of Shield Cutoff Frequency (f_c)

Cable	Impedance (Ω)	Cutoff Frequency (kHz)	Five Times Cutoff Frequency (kHz)	Remarks
Coaxial cable				
RG-6A	75	0.6	3.0	Double shielded
RG-213	50	0.7	3.5	
RG-214	50	0.7	3.5	Double shielded
RG-62A	93	1.5	7.5	
RG-59C	75	1.6	8.0	
RG-58C	50	2.0	10.0	
Shielded twisted pair				
754E	125	0.8	4.0	Double shielded
24 Ga.	—	2.2	11.0	
22 Ga.[a]	—	7.0	35.0	Aluminum-foil shield
Shielded single				
24 Ga.	—	4.0	20.0	

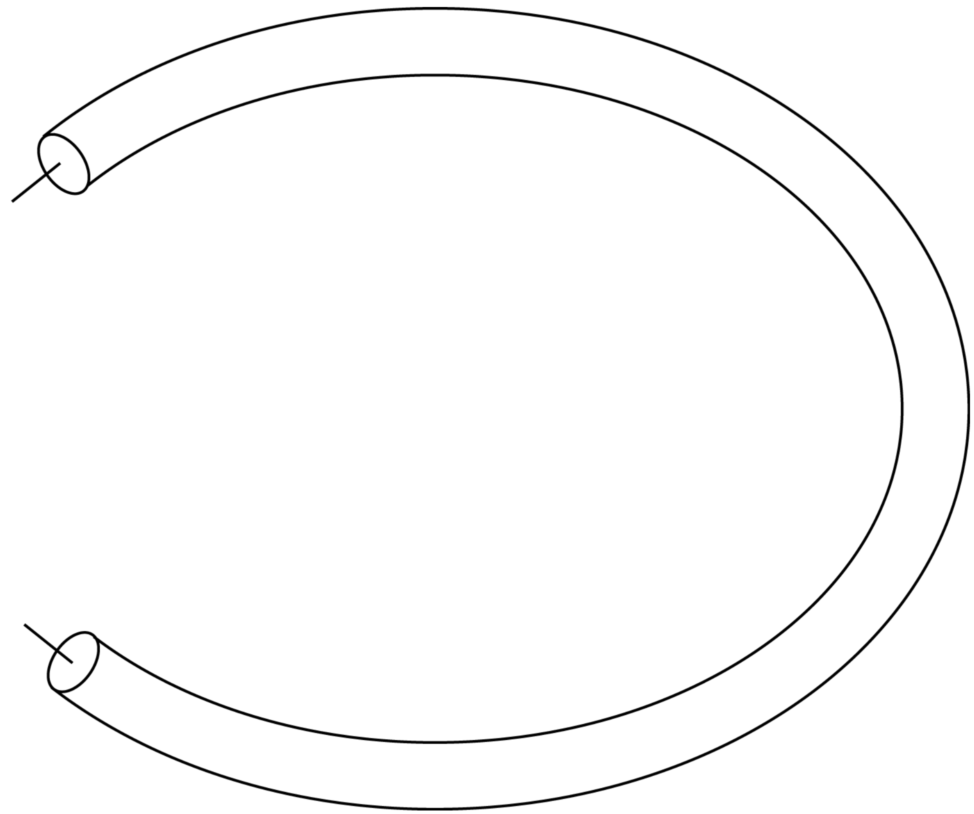
[a]One pair out of an 11-pair cable (Belden 8775).

Fundamental difference between electric field, magnetic field shielding:

Electric field:

Magnetic field:

Magnetic field shielding depends on shield inductance, center conductor inductance being the same! Can we reasonably expect this? Why should they be the same?

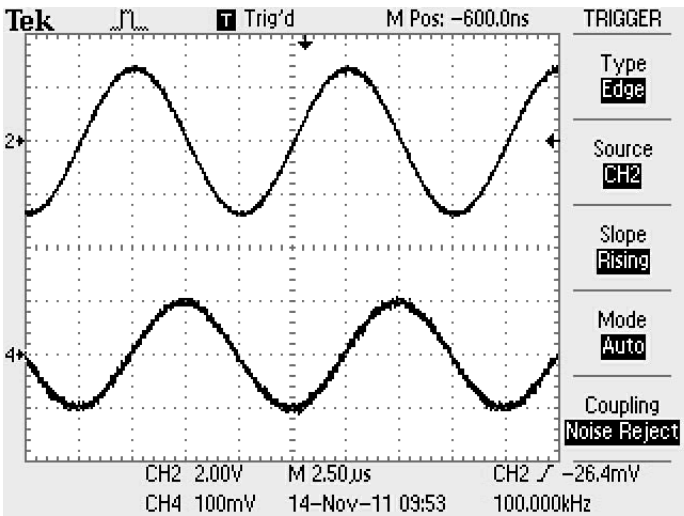


Mutual inductance experiment: Shield grounded at one end only

Current of I_C=10mA peak, 100 kHz on center conductor. What is voltage V_S on shield?

V_S
2V/div

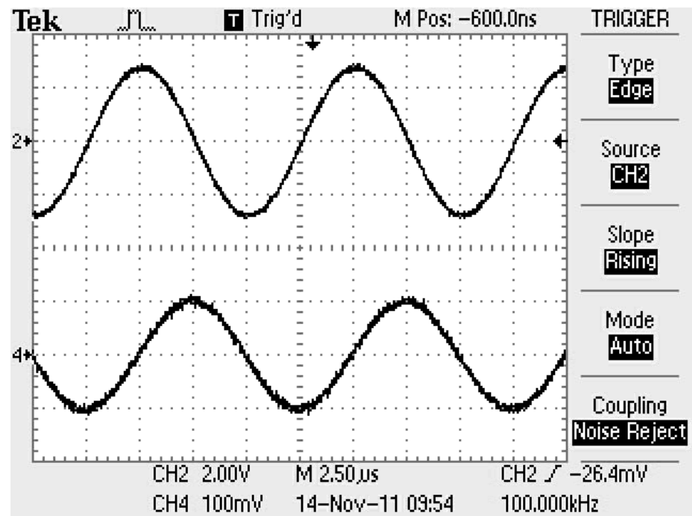
I_C
10mA/div



Current of I_S=10mA peak, 100 kHz on shield What is voltage V_C on center conductor?

V_C
2V/div

I_S
10mA/div



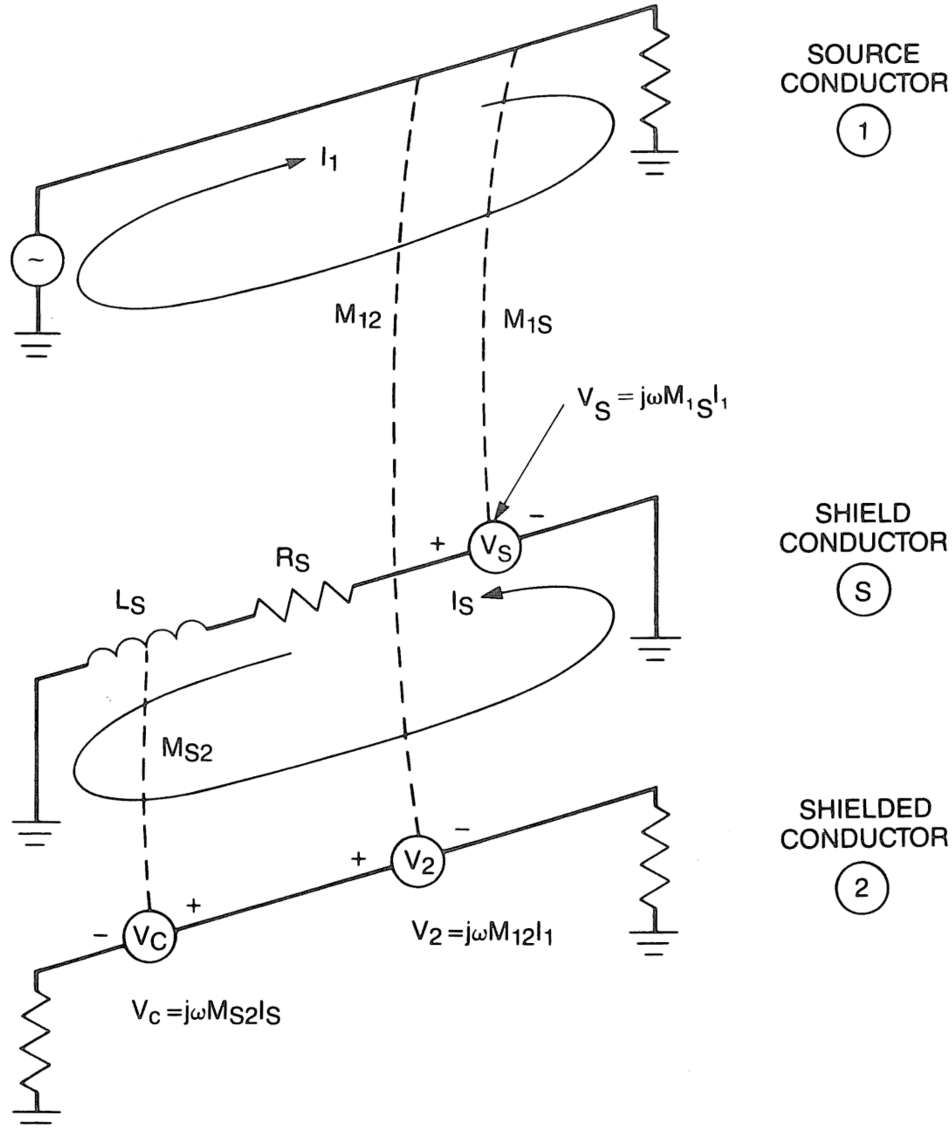


Figure 2-16. *Magnetic coupling to a shielded cable with the shield grounded at both ends.*

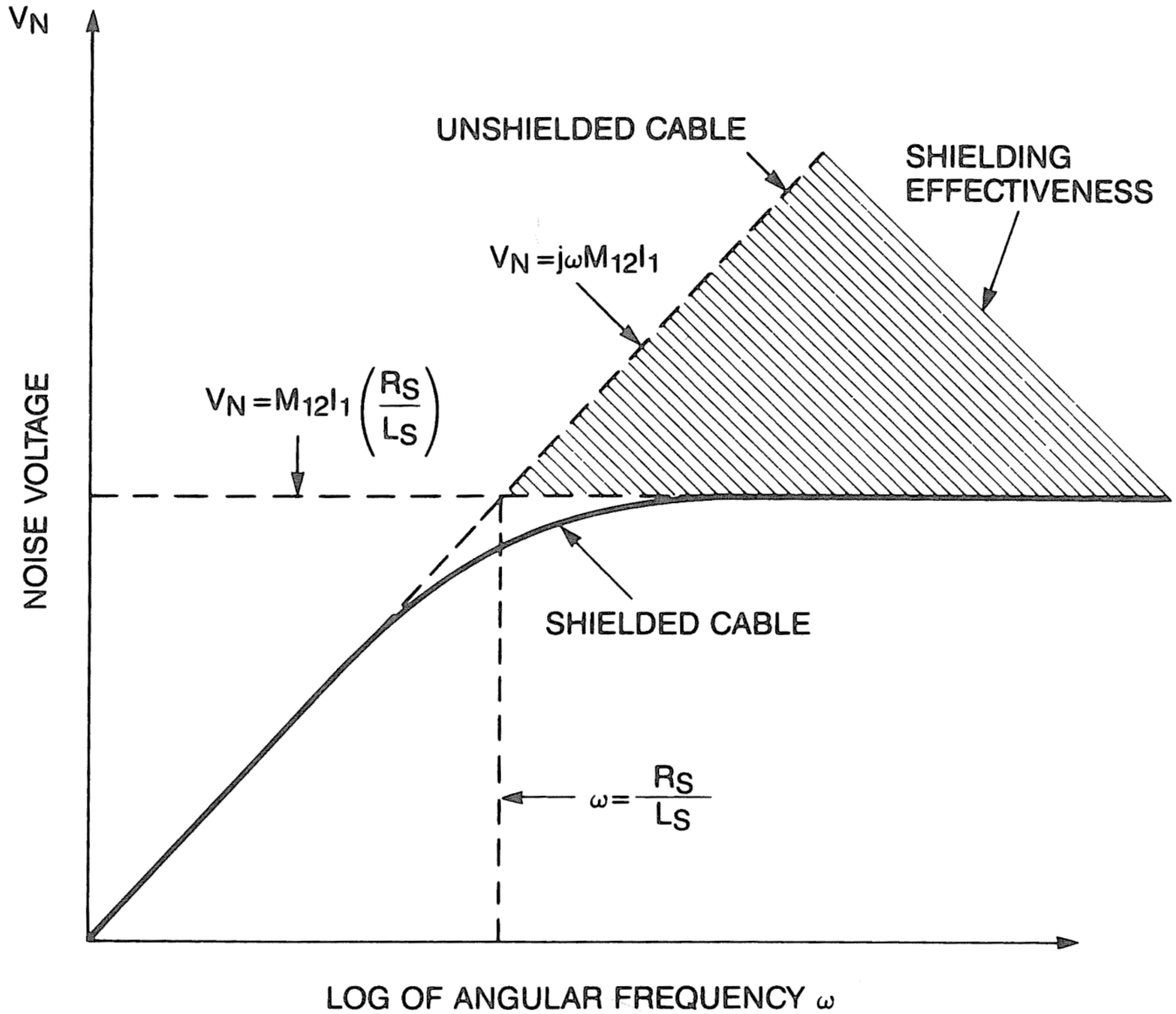


Figure 2-17. *Magnetic field coupled noise voltage for an unshielded and shielded cable (shield grounded at both ends) versus frequency.*

$$V_N = j\omega M_{12} I_1 - j\omega M_{S2} I_S$$

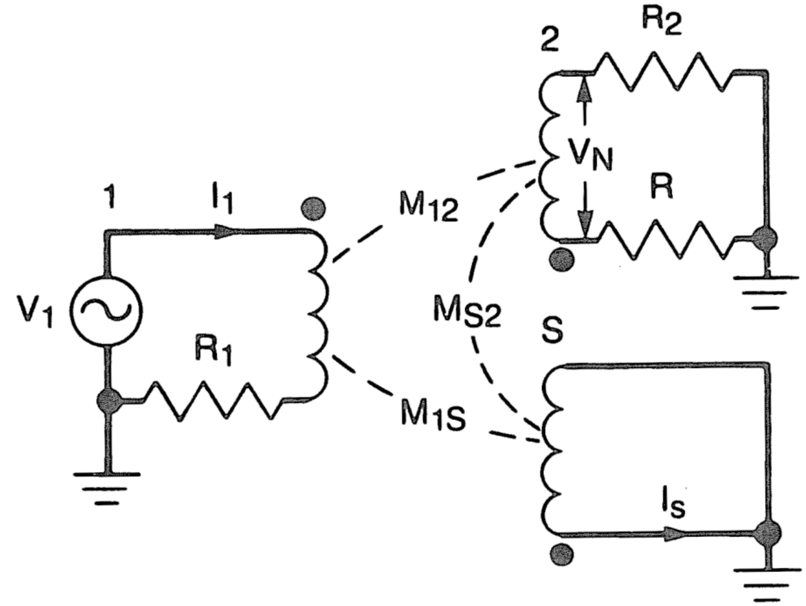


Figure 2-18. *Transformer analogy of magnetic field coupling to a shielded cable when shield is grounded at both ends (M_{S2} is much larger than M_{12} or M_{1S}).*

Previous was for susceptibility when victim. What about radiating field as aggressor?

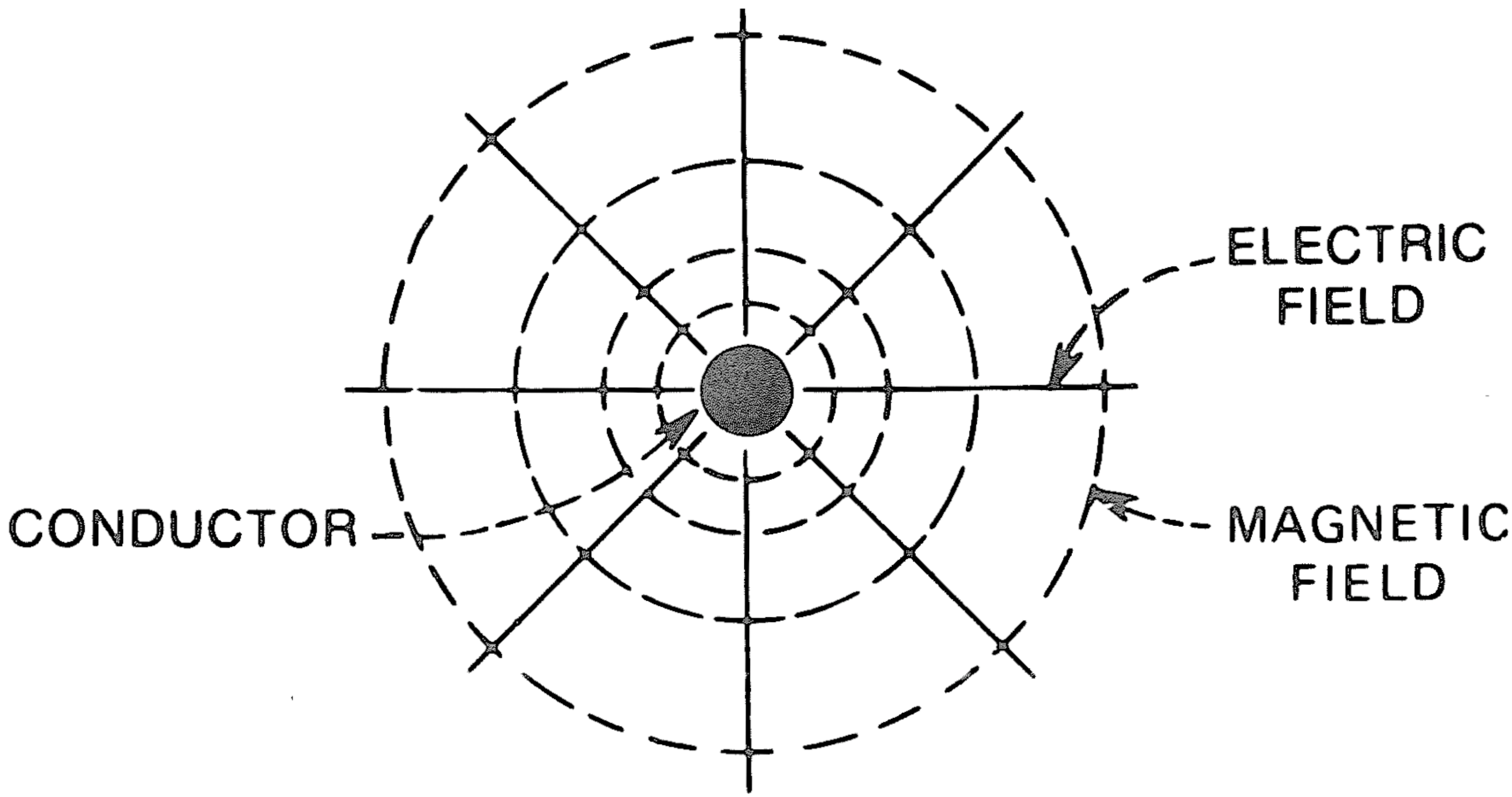


Figure 2-19. *Fields around a current-carrying conductor.*

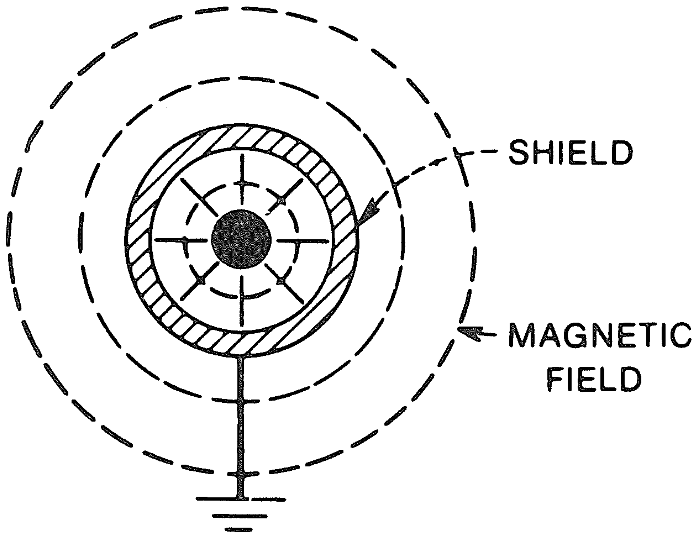


Figure 2-20. *Fields around shielded conductor; shield grounded at one point.*

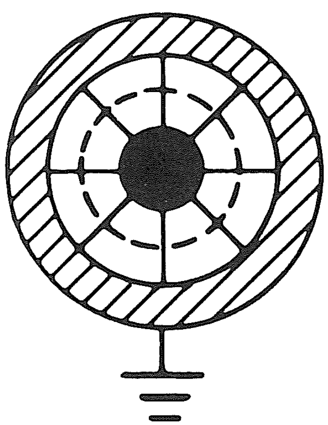


Figure 2-21. *Fields around shielded conductor; shield grounded and carrying a current equal to the conductor current but in the opposite direction.*

Nice result: current flowing in shield helps reduce radiated emissions also!

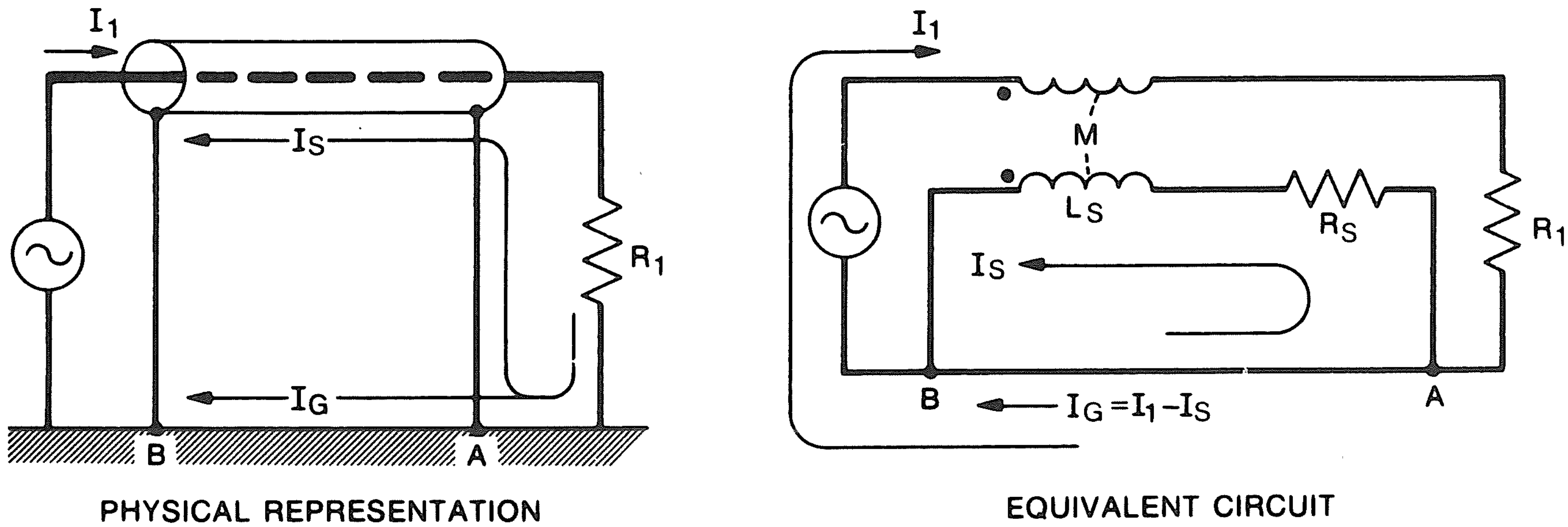


Figure 2-22. *Division of current between shield and ground plane.*

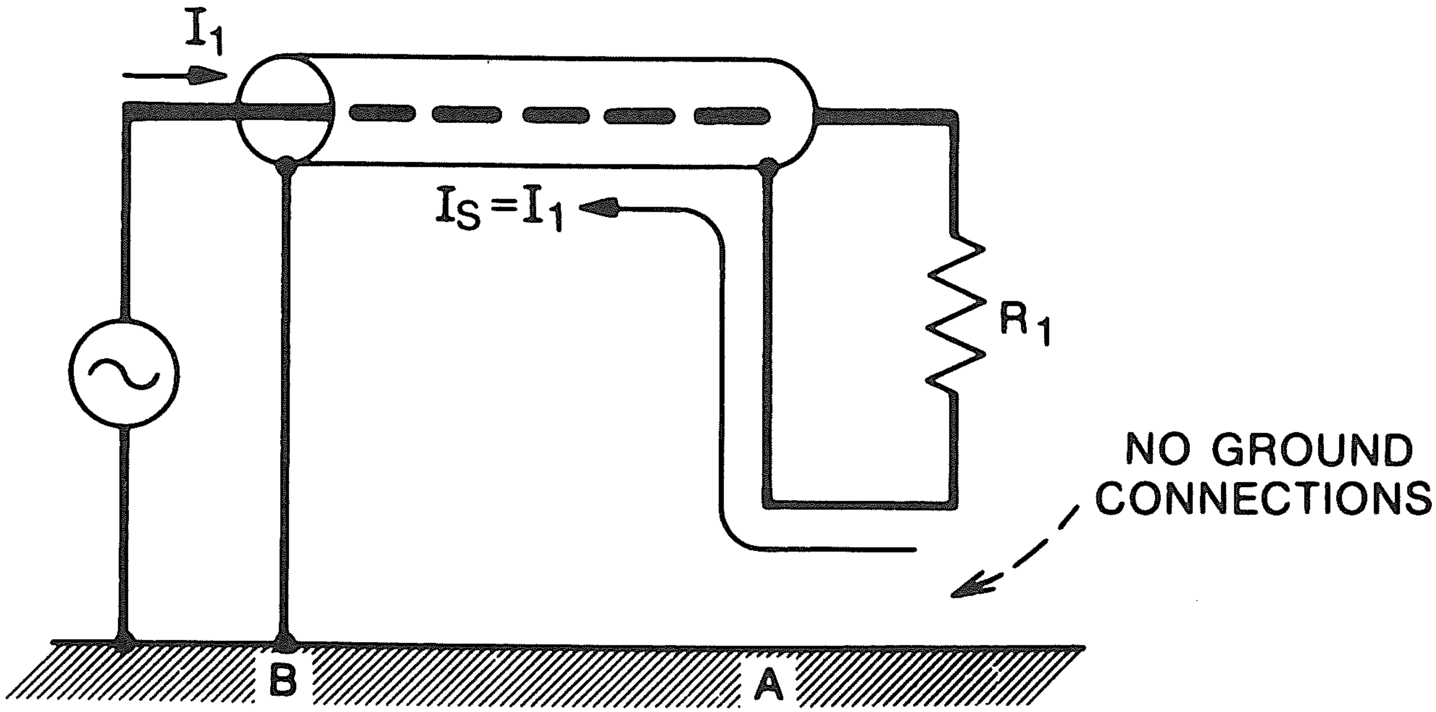


Figure 2-23. *Without ground at far end, all return current flows through shield.*

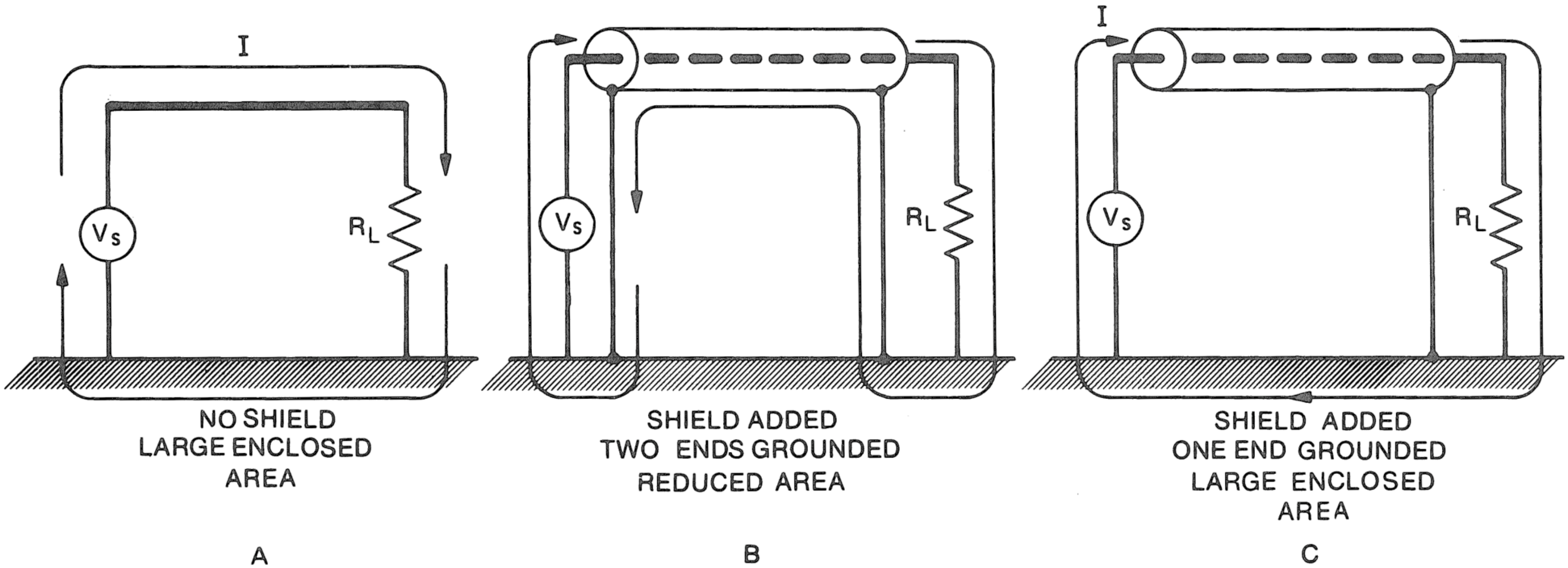


Figure 2-24. *Effect of shield on receptor loop area.*

$$V_{IN} = -j\omega M I_S + j\omega L_S I_S + R_S I_S \,. \tag{2-32}$$

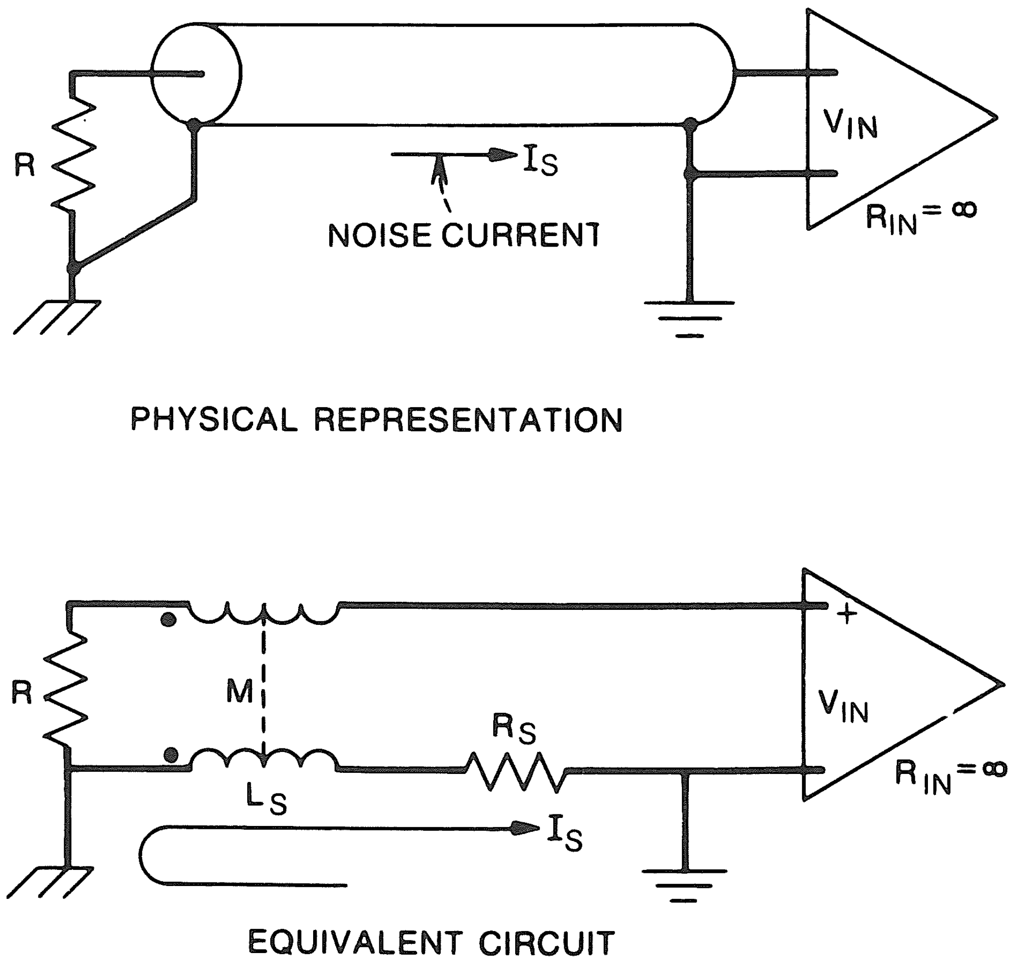


Figure 2-25. *Effect of noise current flowing in the shield of a coaxial cable.*

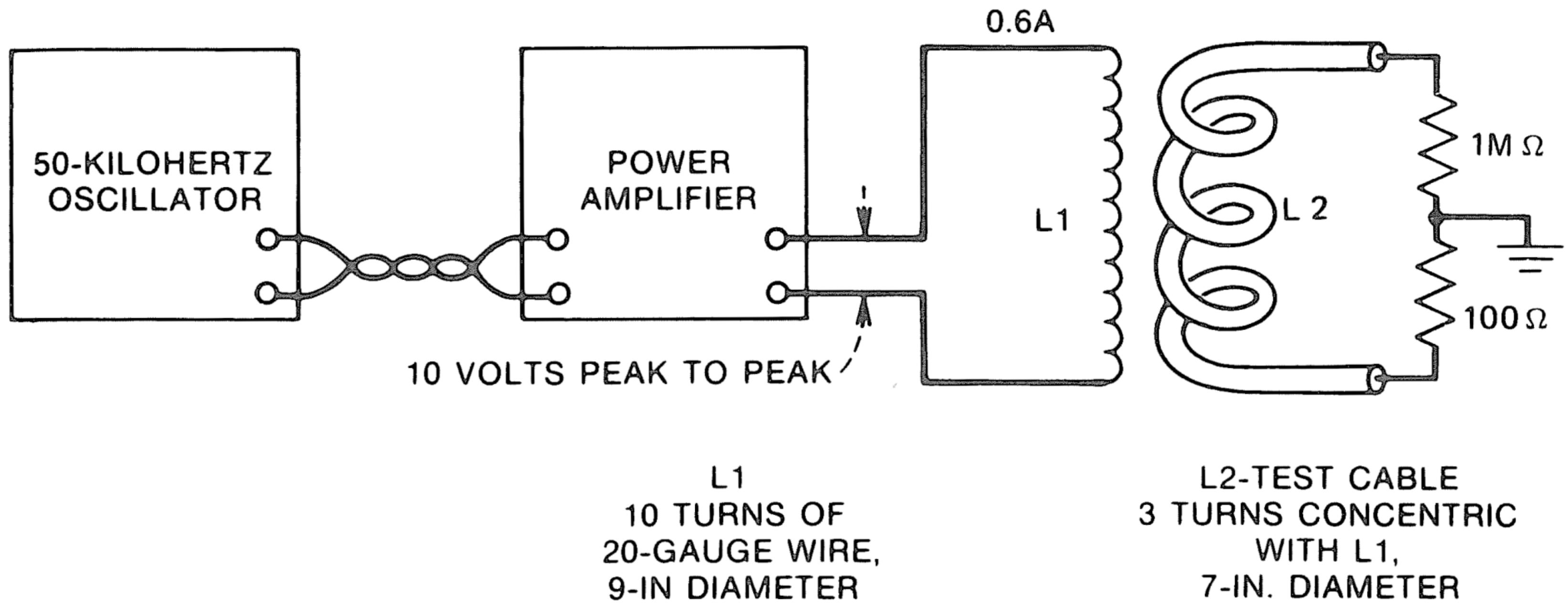


Figure 2-26: Test setup of inductive coupling experiment

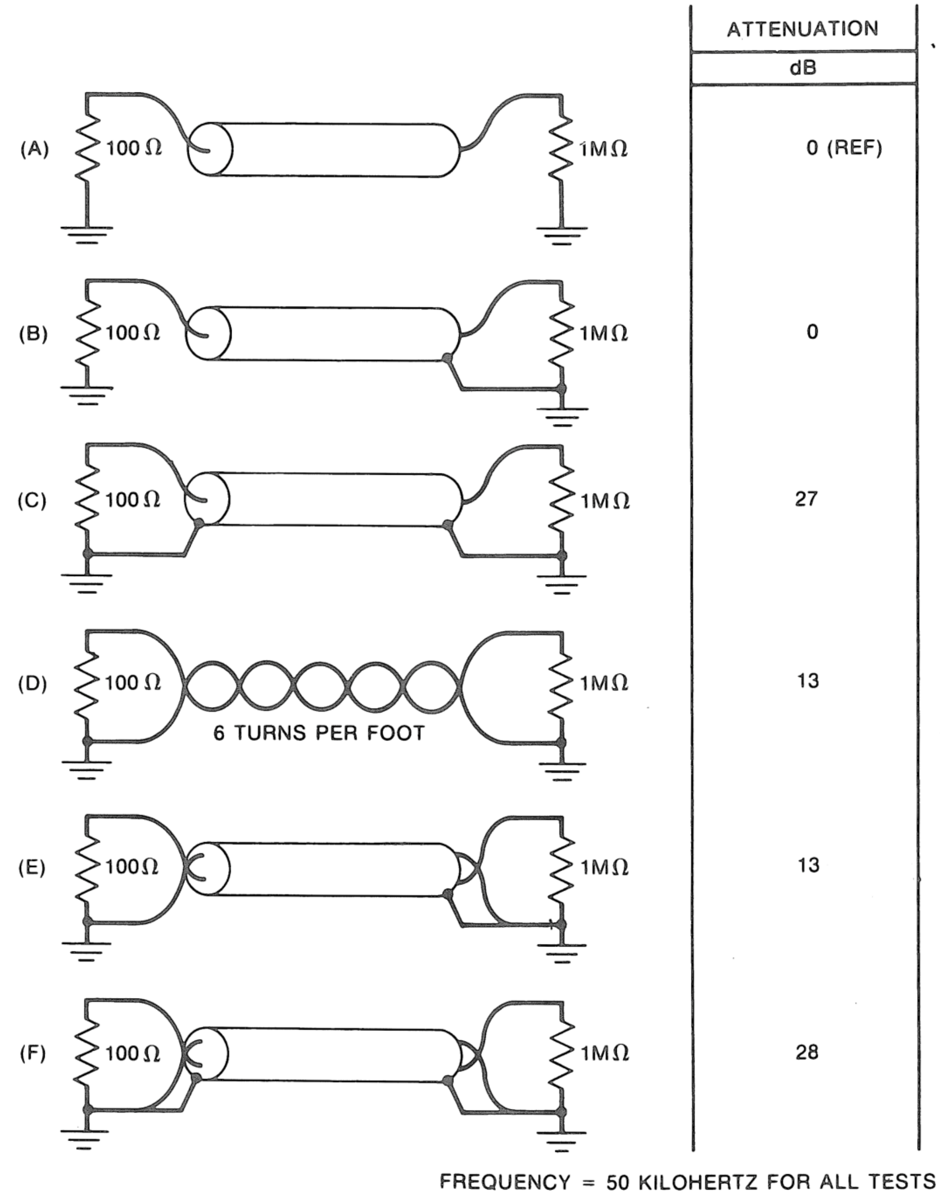


Figure 2-27: Results of inductive coupling experiment; all circuits are grounded at both ends

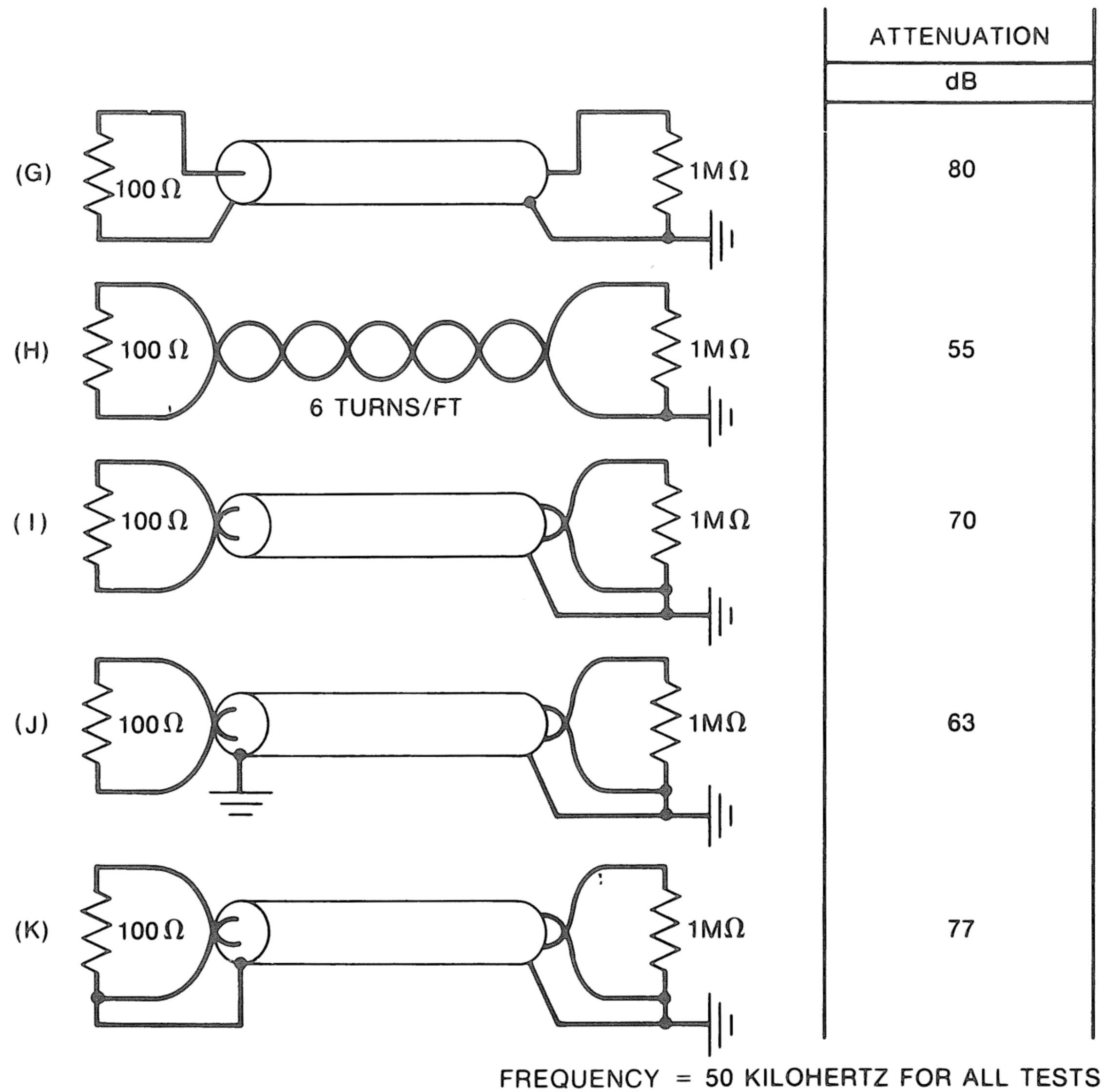


Figure 2-28: Results of inductive coupling experiment; all circuits are grounded at one end only

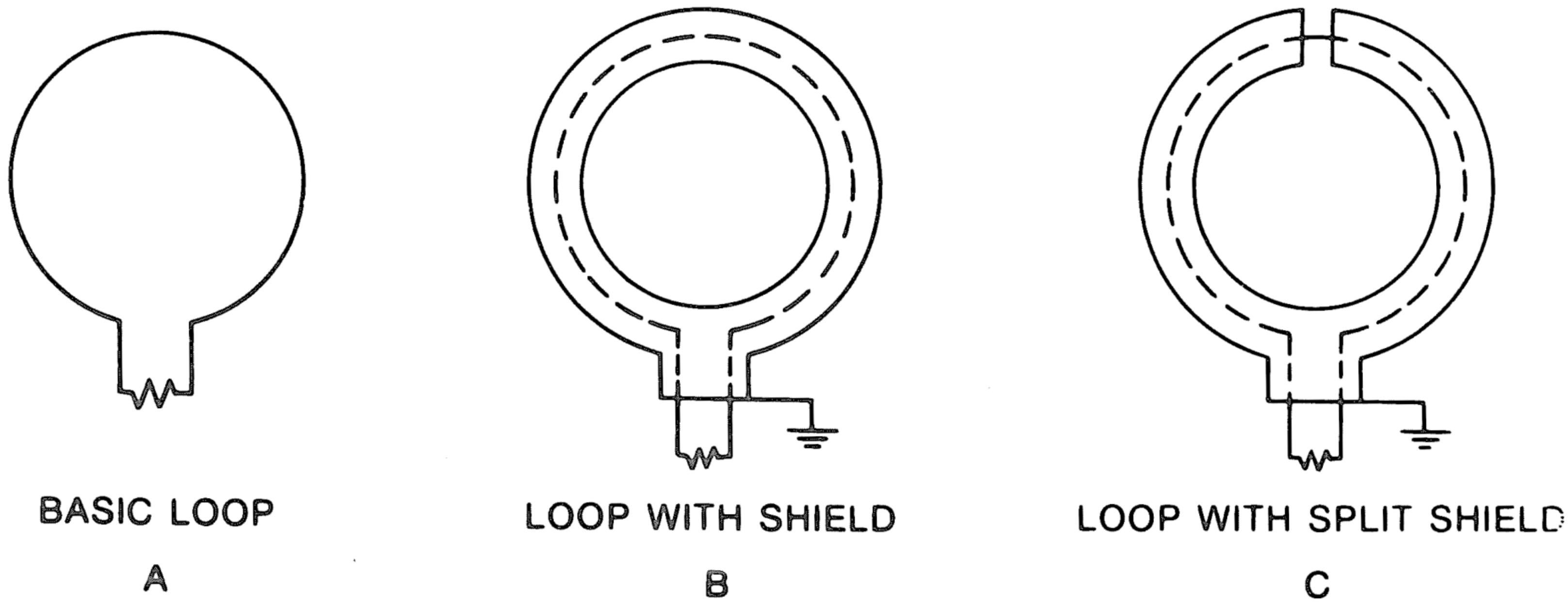


Figure 2-29: Split shield on loop antenna selectively reduces electric field while passing magnetic field

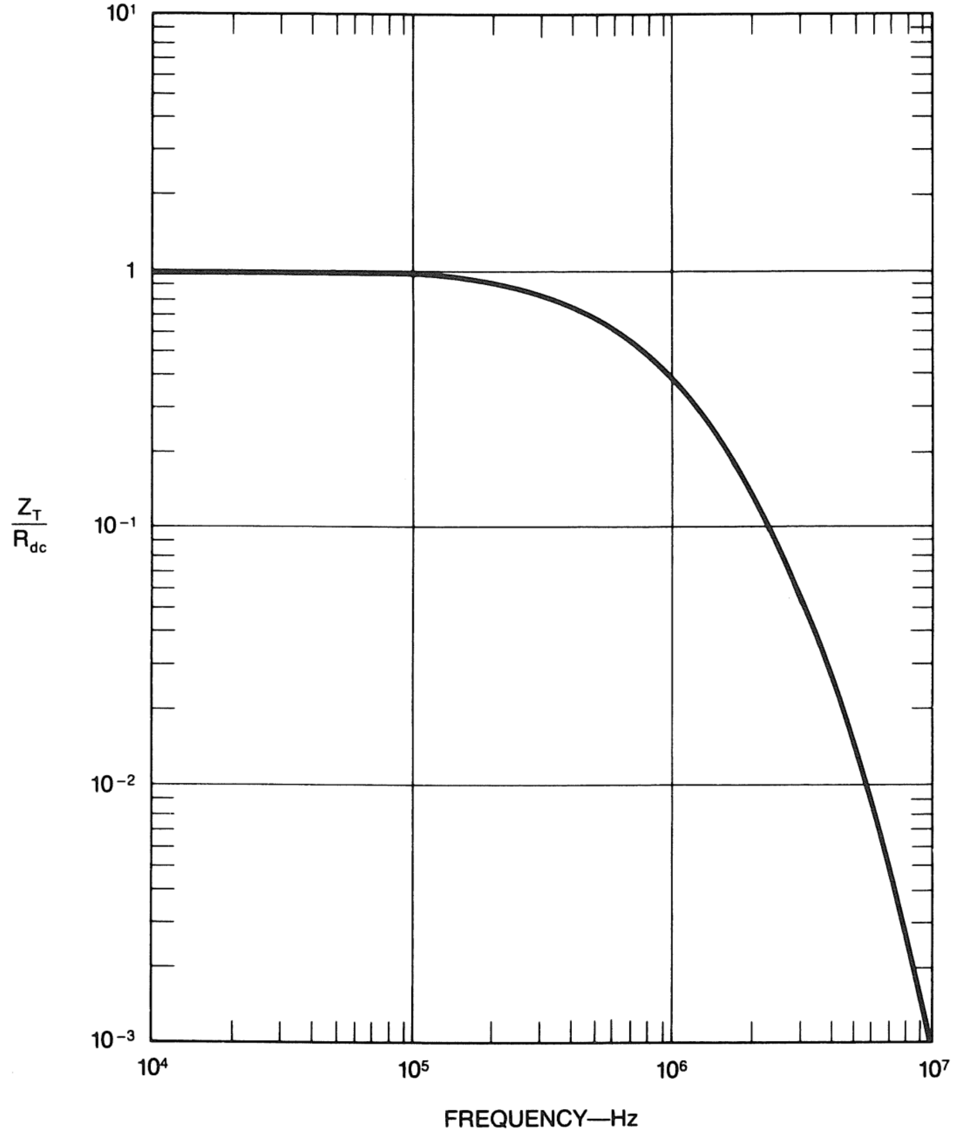


Figure 2-30: Magnitude of normalized transfer impedance for a solid shield

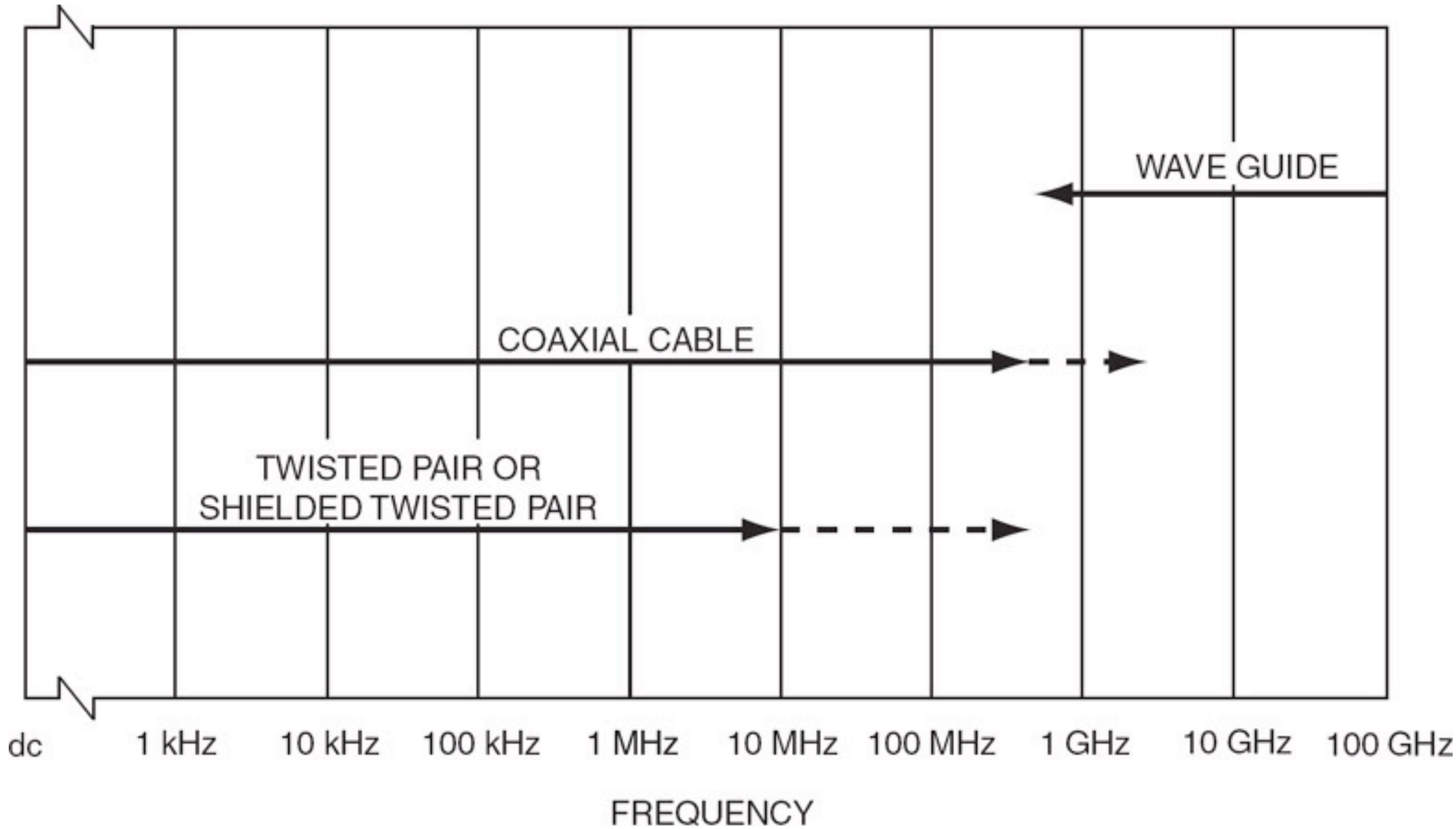


Figure 2-31. *Useful frequency range for various transmission lines.*

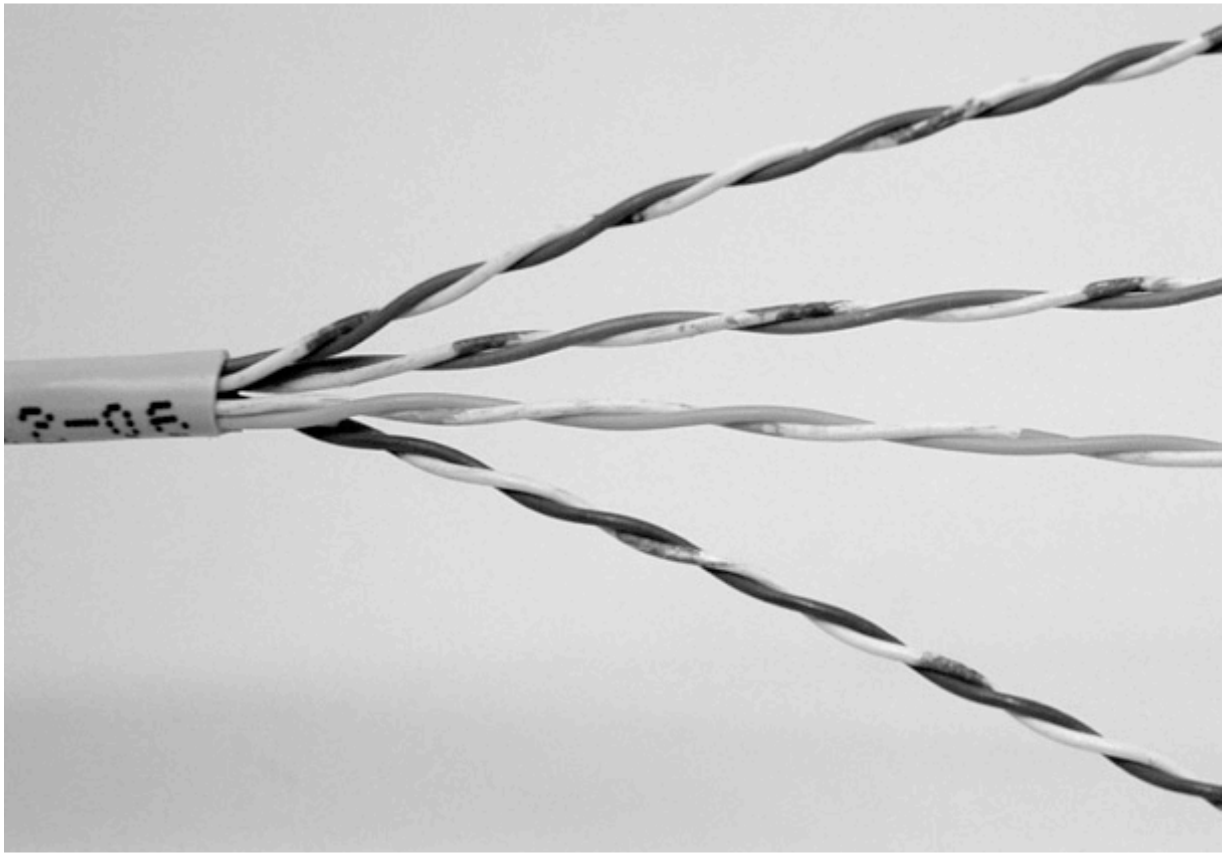


Figure 2-32: Cat 5 Ethernet cable

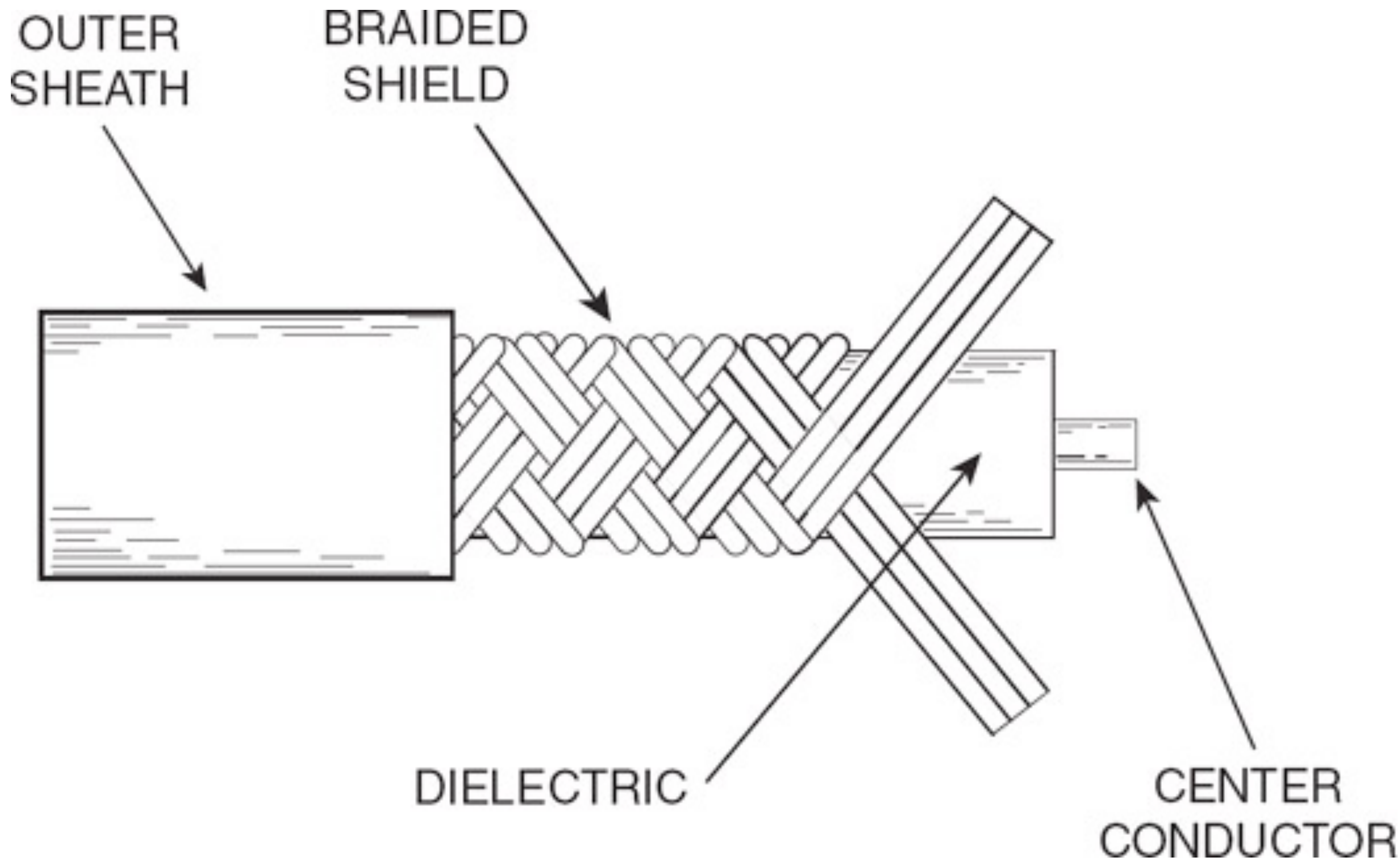


Figure 2-33: Cable with a braid shield

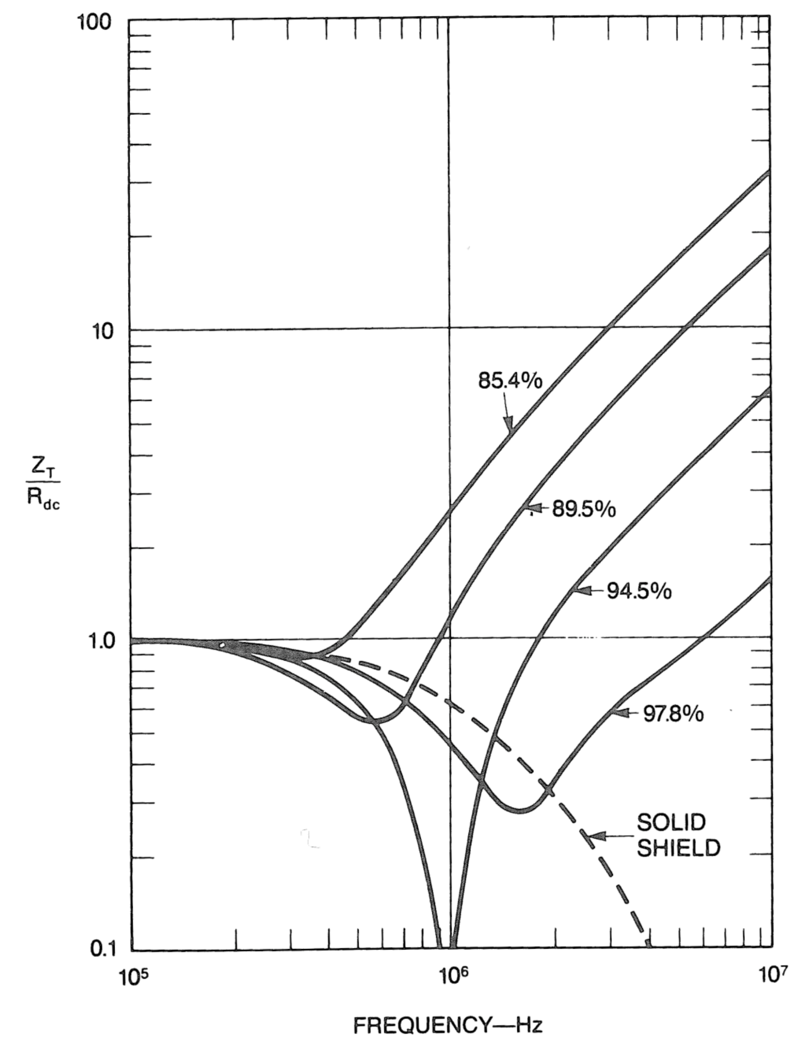


Figure 2-34: Normalized transfer impedance of a braided-wire shield, as a function of percent braid coverage (Vance, 1978, reprinted with permission of John Wiley & Sons, Inc.)

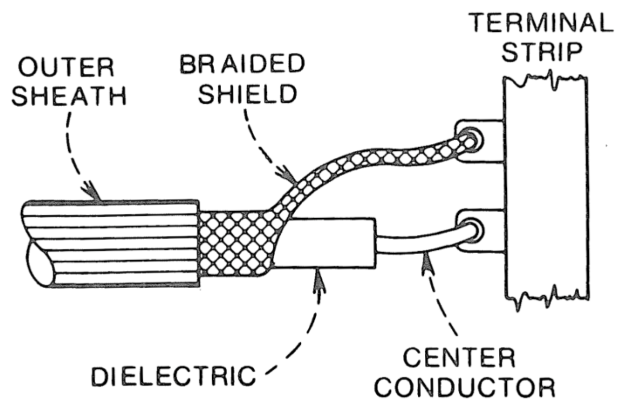


Figure 2-40: Pigtail shield connection concentrates current on one side of the shield

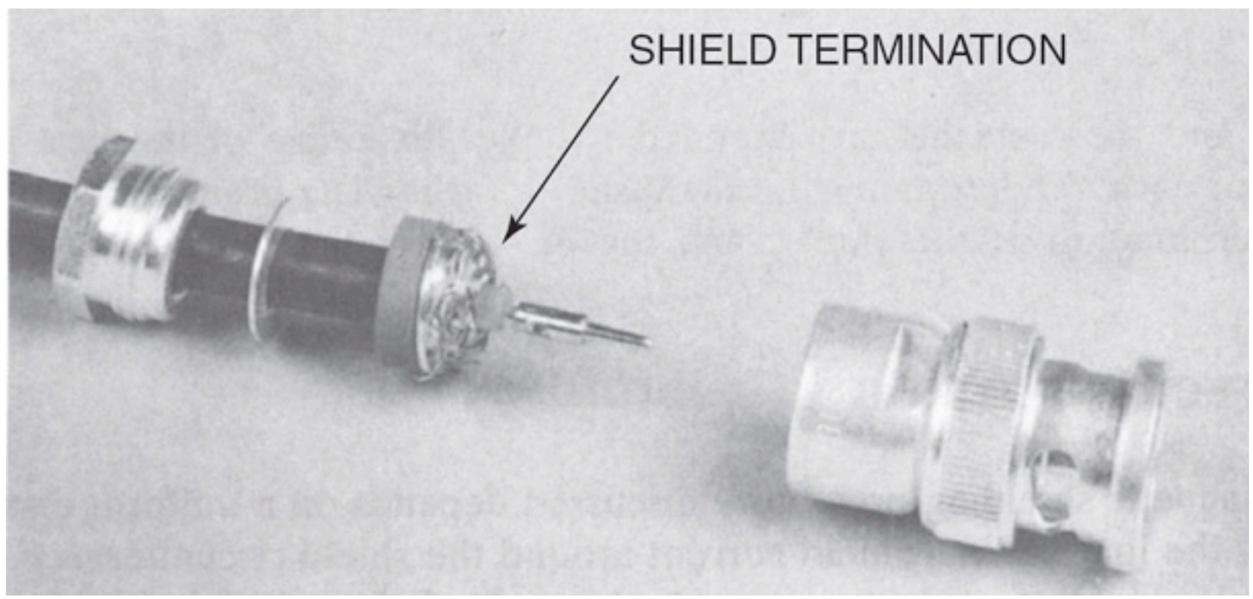


Figure 2-41: Disassembled BNC connector showing a 360° contact to the shield

For example, the coupling to a 3.66-m (12 ft) shielded cable with the shield grounded at both ends with 8-cm pigtail terminations is shown in Fig. 2-44 (Paul, 1980, Fig. 8a). The terminating impedance of the shielded conductor was 50 Ω.
The capacitive (electric field) coupling to the shielded portion of the cable was negligible because the shield was grounded and the shielded conductor's terminating impedance was low (50 Ω°).
As shown in Fig. 2-44, above 100 kHz, primary coupling to cable from inductive coupling to the pigtail.

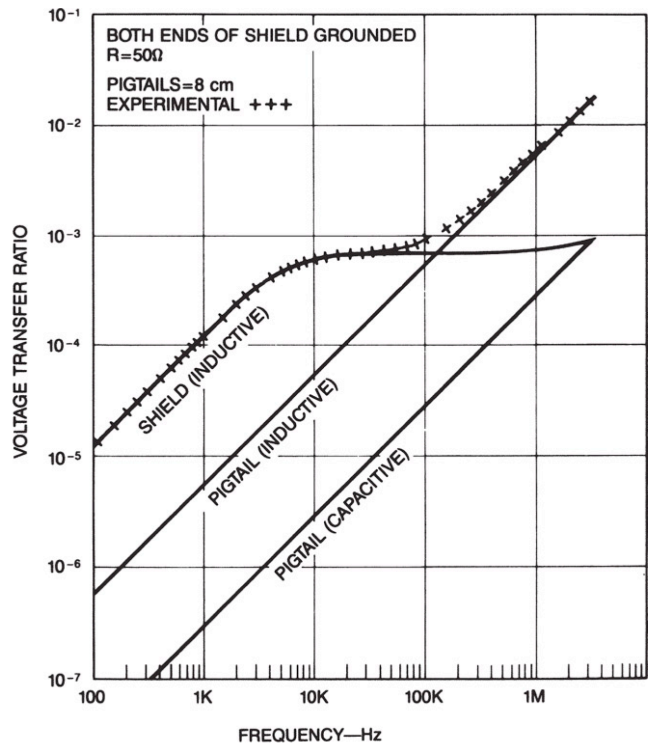


Figure 2-44: Coupling to a 3.7-m shielded cable with an 8-cm pigtail termination. Circuit termination equals 50 Ω (from Paul, 1980, © IEEE)

If the terminating impedance of the shielded conductor is increased from 50 to 1000 Ω, the result is as shown in Fig. 2-45 (Paul, 1980, Fig. 8b).
In this case, the capacitive coupling to the pigtail is the predominant coupling mechanism above 10 kHz.
Coupling at 1 MHz is 40 dB greater than what it would have been if the cable had been completely shielded (no pigtail).

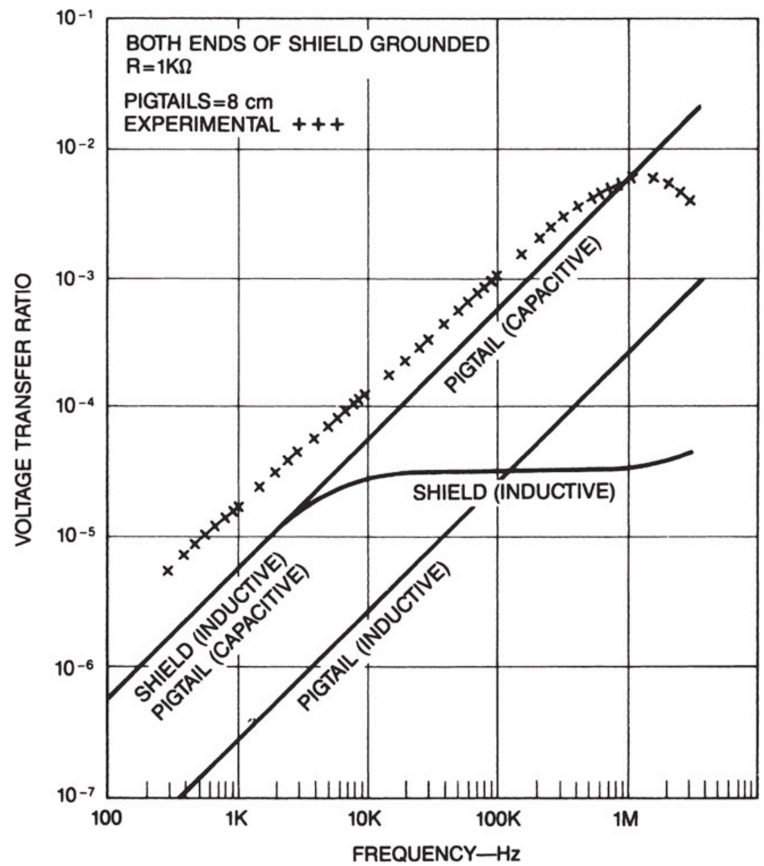


Figure 2-45: Coupling to a 3.7-m shielded cable with an 8-cm pigtail termination. Circuit termination equals 1000 Ω (from Paul, 1980, © IEEE)

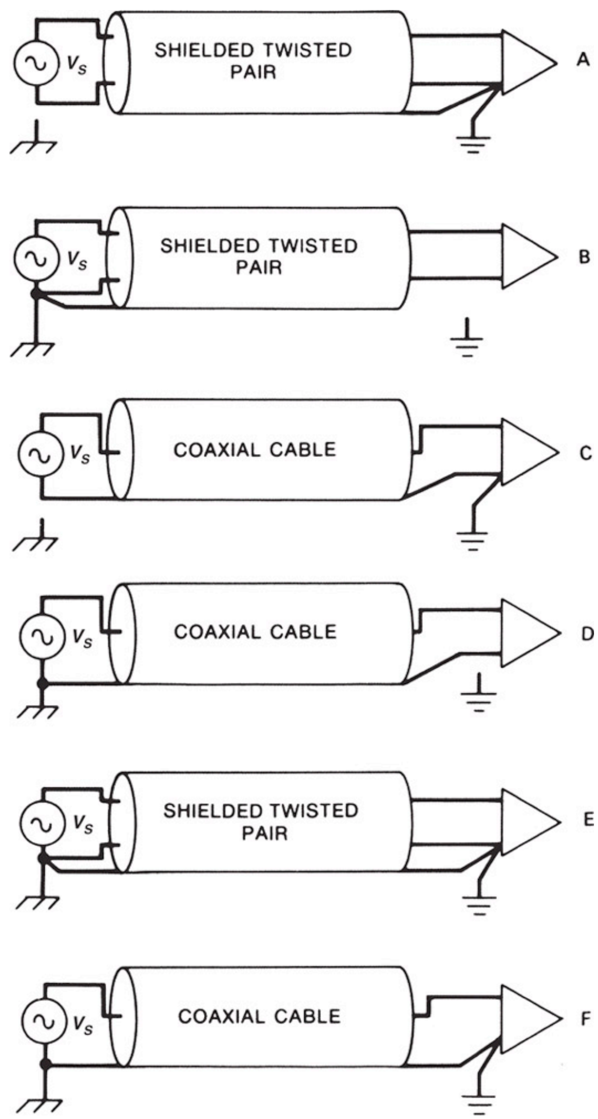


Figure 2-46: Preferred low-frequency shield grounding schemes for shielded twisted pair and coaxial cables

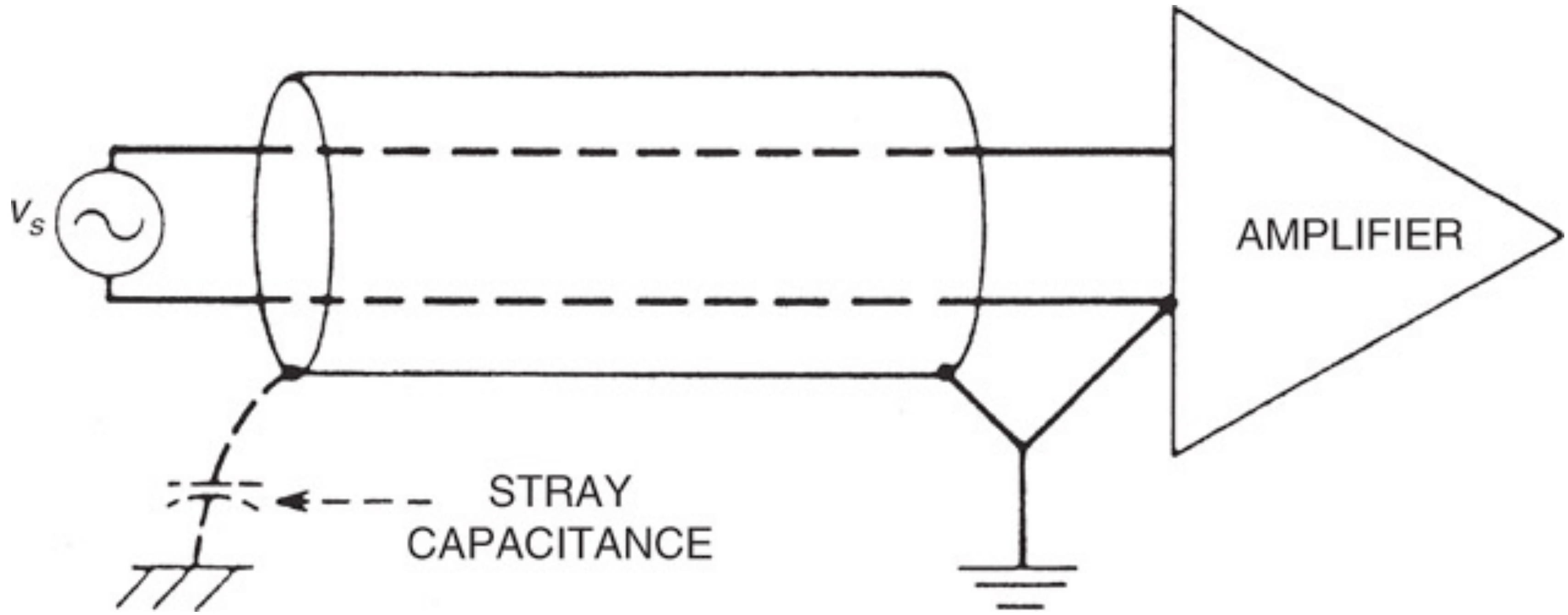


Figure 2-47: At high frequencies, stray capacitance completes the ground loop

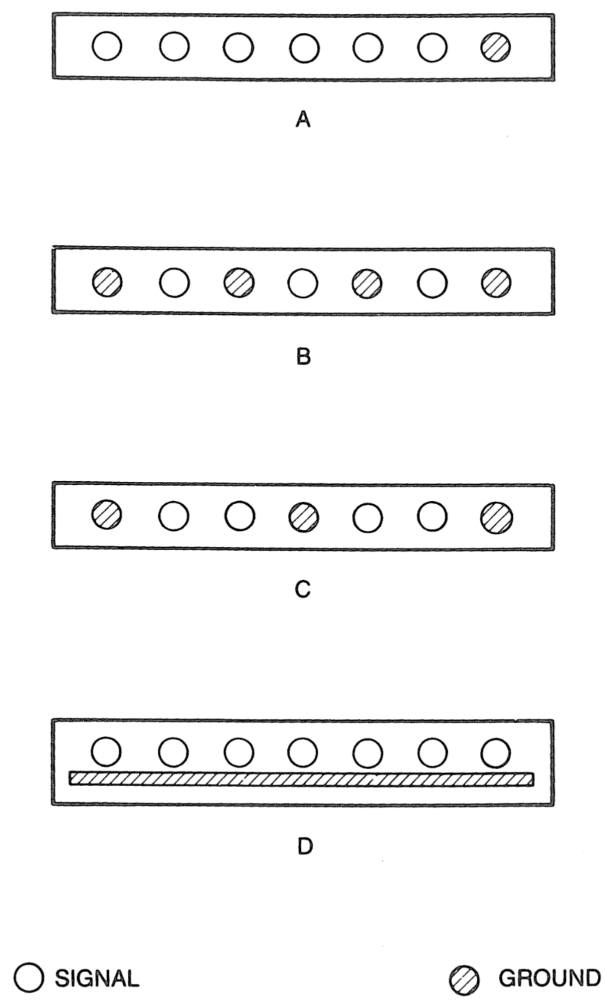


Figure 2-49: Ribbon cable configurations: (A) single ground; *(B)* alternate grounds; (C) ground/signal/signal/ground; *(D)* signal over ground plane

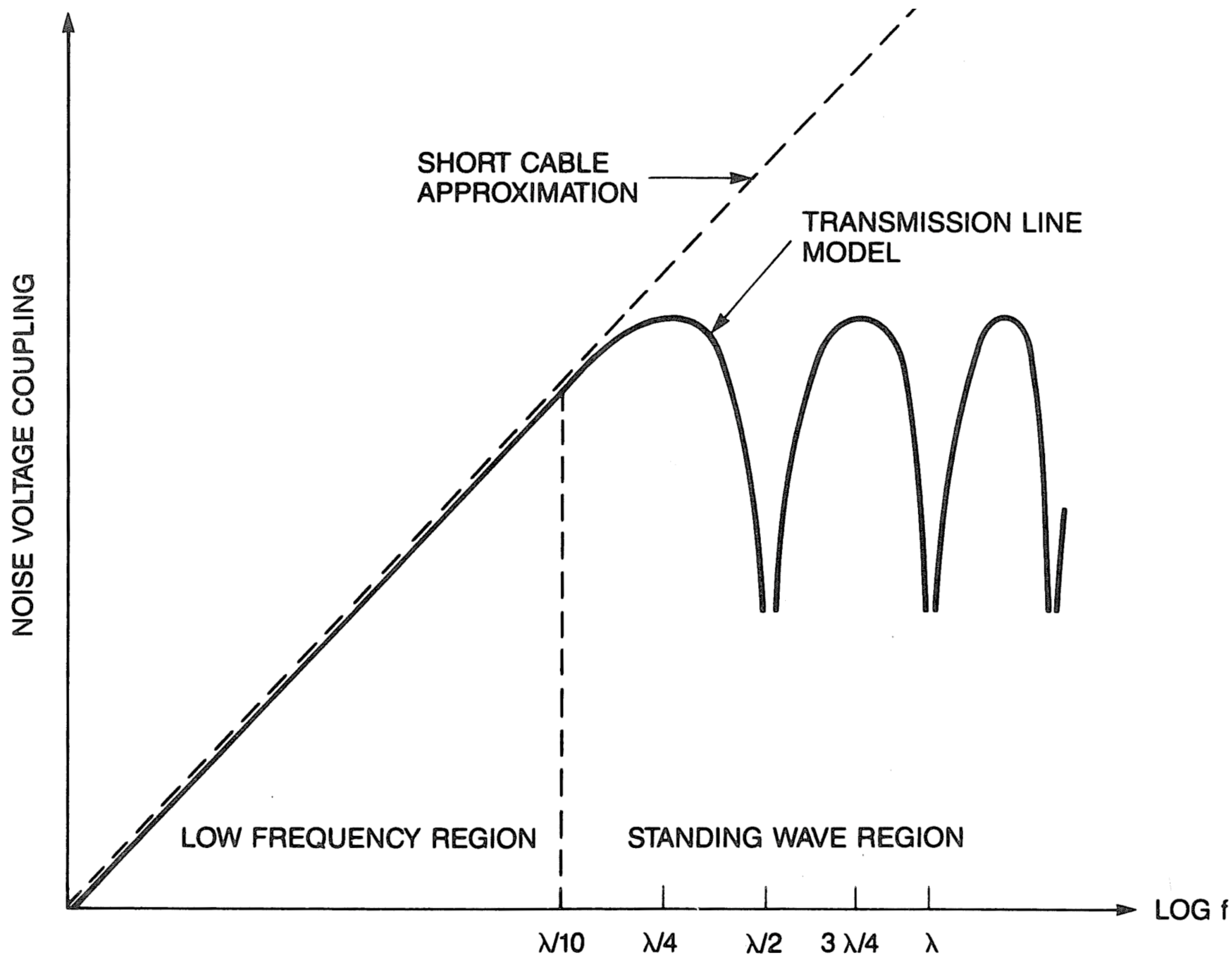


Figure 2-50: Electric field coupling between cables using the short cable approximation (dashed line) and the transmission line model (solid line)

Summary: Cabling

Electric field coupling modeled by inserting a noise current generator in shunt with the receptor circuit.
Magnetic field coupling can be modeled by inserting a noise voltage generator in series with the receptor
Electric fields are much easier to guard against than magnetic fields.
A shield grounded at one or more points shields against electric fields.
The key to reducing magnetic coupling is to decrease the area of the pickup loop.
For a coaxial cable grounded at both ends, virtually all the return current flows in the shied at frequencies above audio.
To prevent magnetic field radiation or pickup, a shield grounded at both ends is useful above audio frequencies.
Any shield in which noise currents flow should not be part of the signal path.
Because of skin effect, at high frequencies, a coaxial cable behaves as a triaxial cable.
The shielding effectiveness of a twisted pair increases as the number of twists per unit length increases.
The magnetic shielding effects listed in this chapter require a cylindrical shield with uniform distribution of shield current over the circumference of the shield.
For a solid-shield cable, the shielding effectiveness increases with frequency.
For a braid-over-foil or double-braid cable, the shielding effectiveness begins to decrease above about 100 MHZ.
For a braided-shield cable, the shielding effectiveness begins to decrease above about 10 MHz.
Most cable shielding problems are caused by improper shield terminations.
At low frequency, cable shields may be grounded at one end only.
At high frequency, cable shields should be grounded at both ends.
Hybrid shield terminations can be used effectively when both low- and high-frequency signals are involved.
Cable shields should be terminated to the equipment enclosure not to the circuit ground.
The major problem with ribbon cables relates to how individual conductors are assigned between signals and grounds.

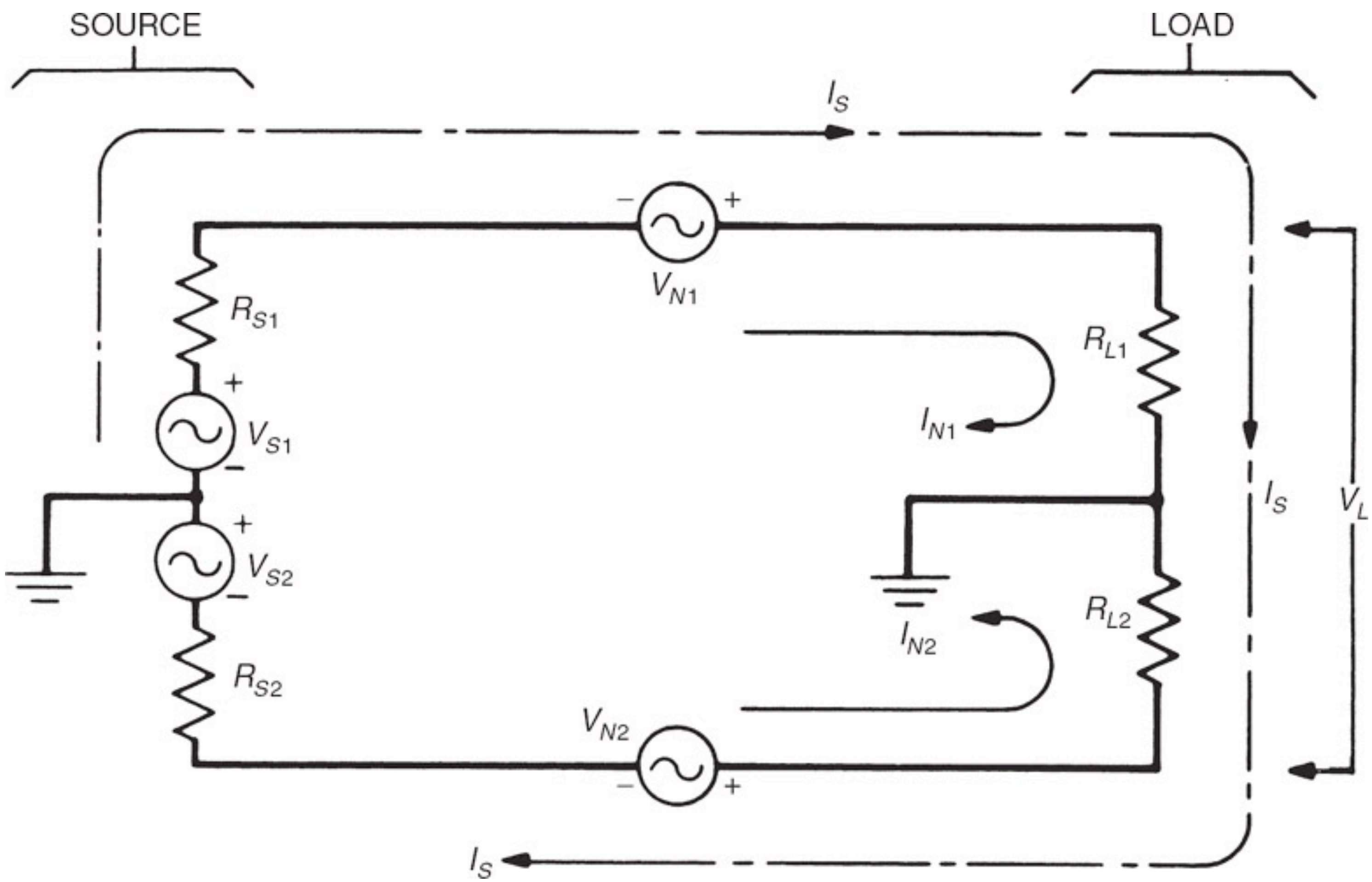


Figure 4-1: For balanced condition: $R_{s1} = R_{s2}$, $R_{L1} = R_{L2}$, $V_{N1} = V_{N2}$, and $I_{N1} = I_{N2}$

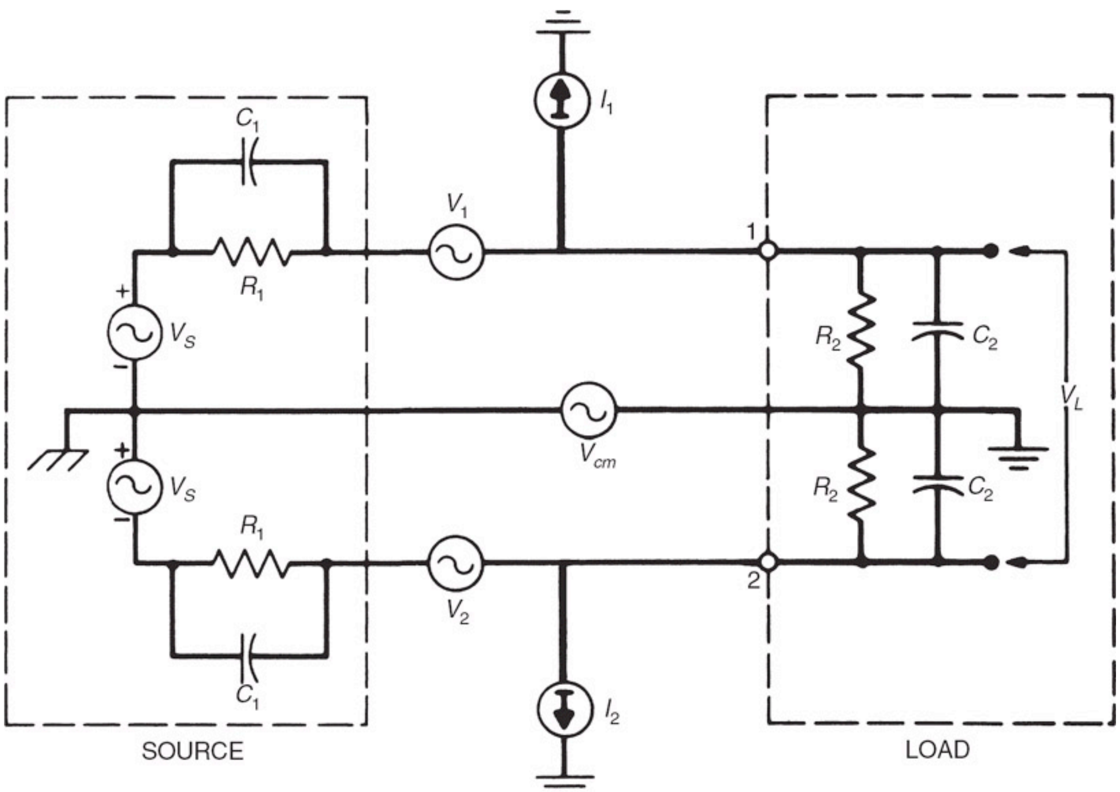


Figure 4-2: A balanced circuit that shows inductive and capacitive noise voltages and a difference in ground potential between source and load

V_1 and V_2 represent inductive pickup voltages
Current generators I_1 and I_2 represent noise that is capacitively coupled into the circuit.
Difference in ground potential between source and load is represented by V_{cm}
If the two signal conductors 1 and 2 are located adjacent to each other, or better yet twisted together, the two inductively coupled noise voltages V_1 and V_2 should be equal and cancel at the load.
Note that inductively coupled noise voltages V_1 and V_2 depend on geometry (loop area and orientation) and not impedance balancing

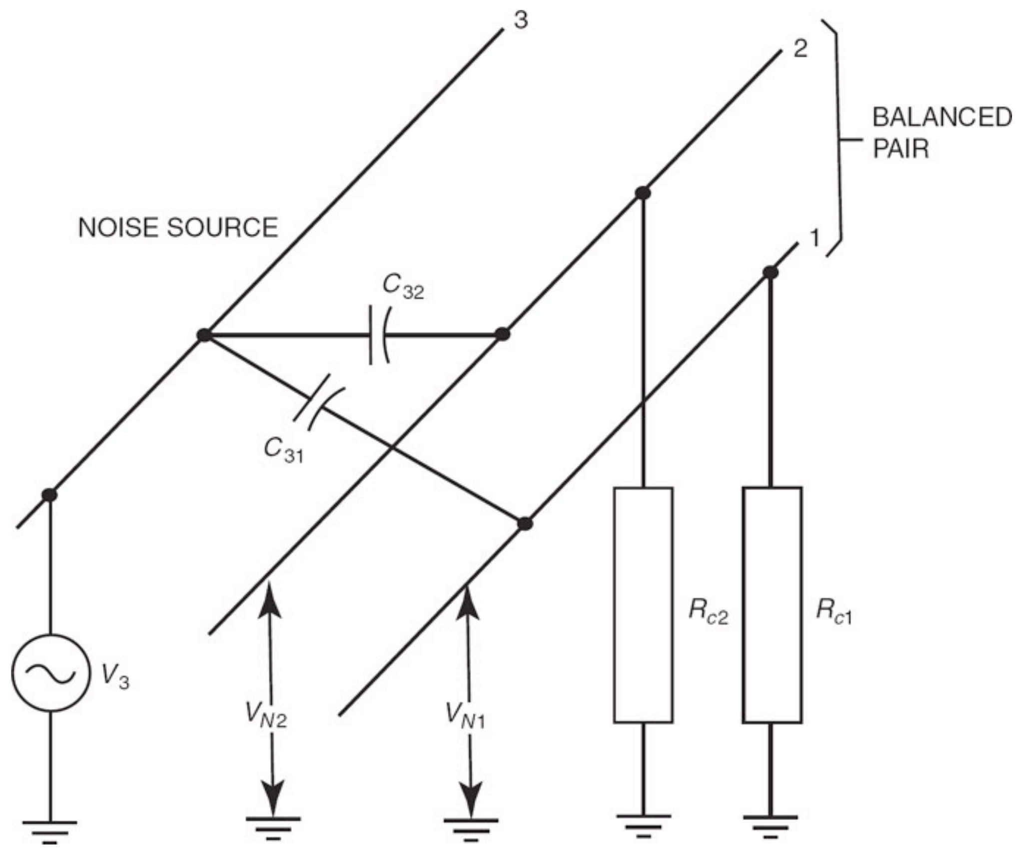


Figure 4-3: Capacitive pickup in balanced conductors

Capacitive coupled noise voltage VN1 induced into conductor 1 from V3 $V_{N1} = j\omega R_{c1}C_{31}V_3$
Capacitive coupled noise voltage VN2 induced into conductor 2 from V3 $V_{N2} = j\omega R_{c2}C_{32}V_3$
If the circuit is balanced, then resistances Rc1 and Rc2 are equal.
If conductors 1 and 2 are located adjacent to each other, or better yet are twisted together, capacitance C31 should be nearly equal C32
Under these conditions, VN1 ≈ VN2, and the capacitively coupled noise voltages cancel in the load.

If the terminations are balanced, then a twisted pair cable can provide protection against capacitive coupling.
Because a twisted pair also protects against magnetic fields, whether the terminations are balanced or not (see Section 2.12), a balanced circuit using a twisted pair will protect against both magnetic and electric fields, even without a shield over the conductors.
Shields may still be desirable, however, because it is difficult to obtain perfect balance, and additional protection may be required.

How to quantify imperfect balance? Common-Mode Rejection Ratio (CMRR)

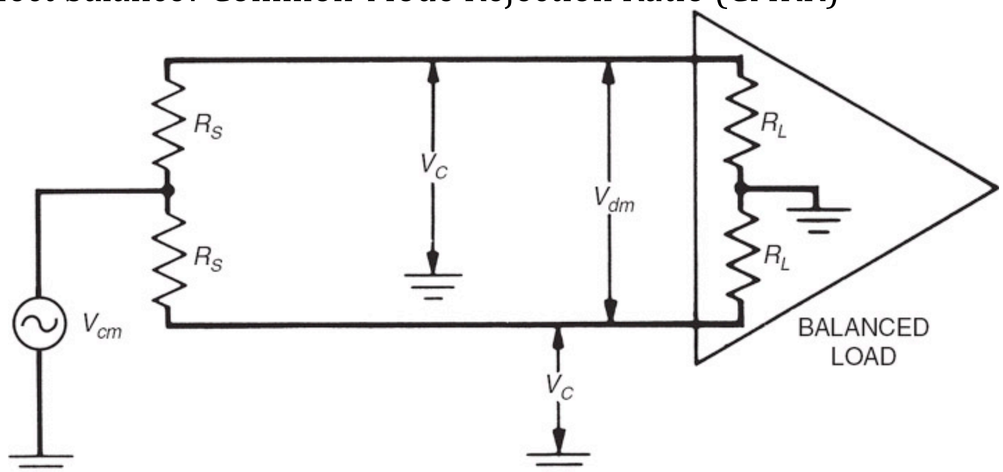
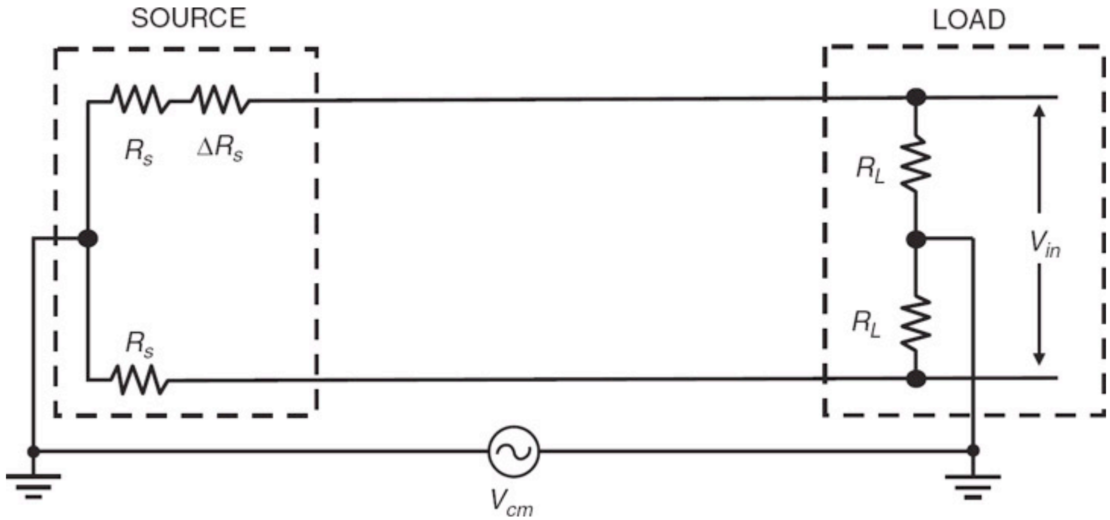


Figure 4-4: Circuit used to define CMRR

Figure 4-4 shows a balanced circuit with a common-mode voltage V_{cm} applied to it.
If the balance were perfect, then no differential-mode voltage V_{dm} would appear across the input of the amplifier.
Because of slight unbalances present in the system, however, a small differential-mode noise voltage V_{dm} will appear across the input terminals of the amplifier as a result of the common-mode voltage $V_{cm} \approx V_{cm}$. The CMRR, or balance, in dB, is defined as

$$\mathrm{CMRR} = 20 \log \left(\frac{V_c}{V_{dm}} \right). \qquad (4\text{-}6)$$

Unbalanced source



$$\mathrm{CMRR} = 20\log\left[\frac{(R_L + R_s + \Delta R_s)(R_L + R_s)}{R_L \Delta R_s}\right]. \qquad (4\text{-}7)$$

If R_L is much greater than $R_s + \Delta R_s$, which is usually the case, then Eq. 4-7 can be rewritten as

$$\mathrm{CMRR} = 20\log\left[\frac{R_L}{\Delta R_s}\right]. \qquad (4\text{-}8)$$

If an unbalanced source (one end of the source grounded) is used with a balanced load, then ΔRs will be equal to the total source resistance Rs

For example, if R_L equals 10 kΩ and ΔR_s equals 10Ω, then the CMRR will be 60 dB.

Detrimental effect of source unbalance on the CMRR of the circuit shown in Fig. 4-5 can be reduced by:

- Reducing the common-mode voltage
- Reducing the source unbalance ΔRs
- Increasing the common-mode load impedance RL

Unbalanced load

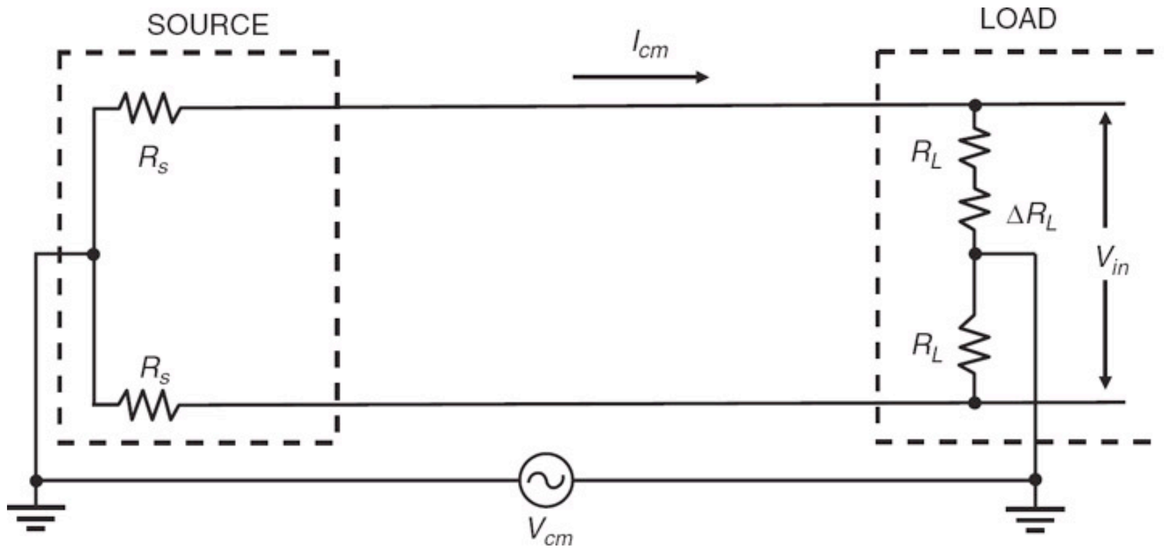


Figure 4-6: Effect of an unbalanced load resistance on the CMRR on a balanced circuit

For R_L much greater than R_s,

$$\mathrm{CMRR} = 20\log\left[\left(\frac{R_L}{R_s}\right)\left(\frac{R_L+\Delta R_L}{\Delta R_L}\right)\right]. \tag{4-10}$$

For example, if $R_S = 100\ \Omega$, $R_L = 10\ \mathrm{k}\ \Omega$, and $\Delta R_L = 100\ \Omega$, then the CMRR = 80 dB.
From Eq. 4-10, we observe that the CMRR is a function of the R_L/R_s ratio;
the larger this is the greater the noise rejection, regardless of the value of ΔR_L. Therefore,
a low source impedance with a high load impedance will provide the largest CMRR.
Ideally, we would like a zero source impedance and an infinite load impedance.

From Eqs. 4-8 and 4-10, we can conclude that a **large load resistance will maximize the CMRR for the case of both source unbalance and load unbalance**.
Problem: can't always do this, for example in RF systems in which $R_s = R_L = Z_0$ impedances must be matched to characteristic impedance.

Cable Balance

For cable, both resistive and reactive balances must be maintained between the two conductors.
Resistance and reactance of each conductor must be equal.
Usually circuit unbalances are greater than the cable unbalances.
When large amounts of common-mode rejection are required, greater than 100 dB, or very long cables are used, the cable imperfections must be considered.

Resistive unbalance of most cables is negligible and can usually be ignored.

Capacitive unbalance is typically in the 3% to 5% range.
At low frequency, this unbalance can usually be ignored because the capacitive reactances will be so much greater than the other impedances in the circuit.
At high frequency, however, the capacitive unbalance may have to be considered.

Inductive unbalances are virtually nonexistent for braid shield cables if properly terminated.
Improper termination of cable shields, that is non-360° contact to the shield, can be a problem, however.
Foil shield cables that contain a drain wire have significantly more inductive unbalance than braid shield cables because of the presence of current in the drain wire.

Effects of balancing and shielding are additive. The shielding can be used to reduce the amount of common-mode pickup in the signal conductors, and the balancing reduces the portion of the common-mode voltage that is converted to differential-mode voltage and coupled into the load.

System Balance

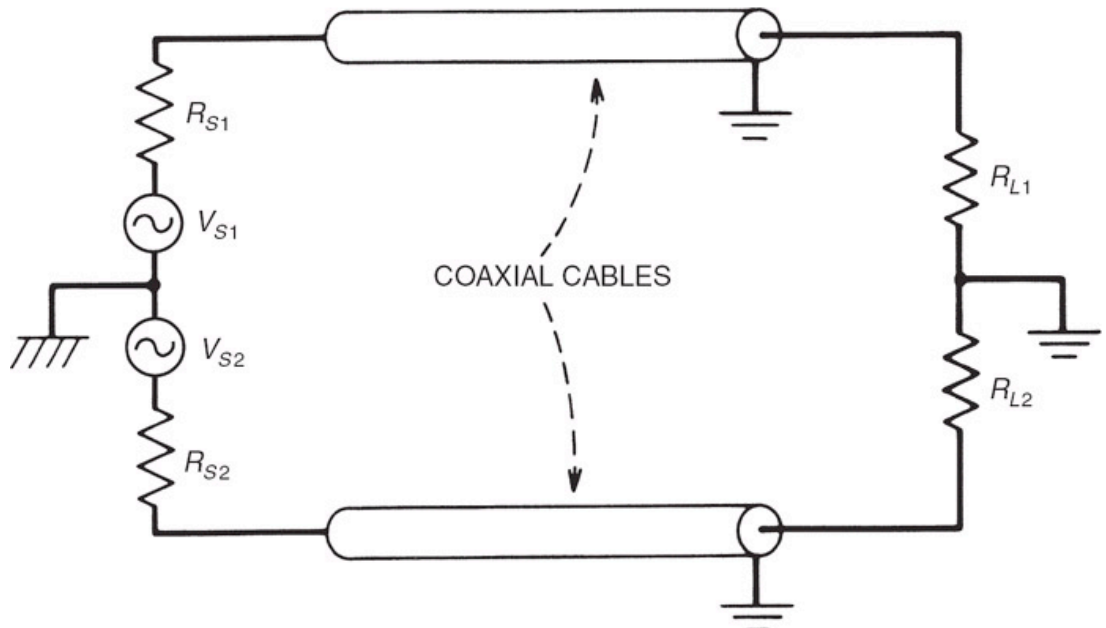


Figure 4-7: Use of coaxial cable in a balanced circuit

With multiple individual components in a system, overall CMRR of the system ≈ CMRR of worst component.

Because a twisted pair cable is inherently a balanced configuration, twisted pair or shielded twisted pair cables are often used as the interconnecting cables in a balanced system.
Coaxial cable is inherently an unbalanced configuration.
If a coaxial cable is to be used in a balanced system, then two cables can be used, as shown in Fig. 4-7.

Differential Amplifier Load

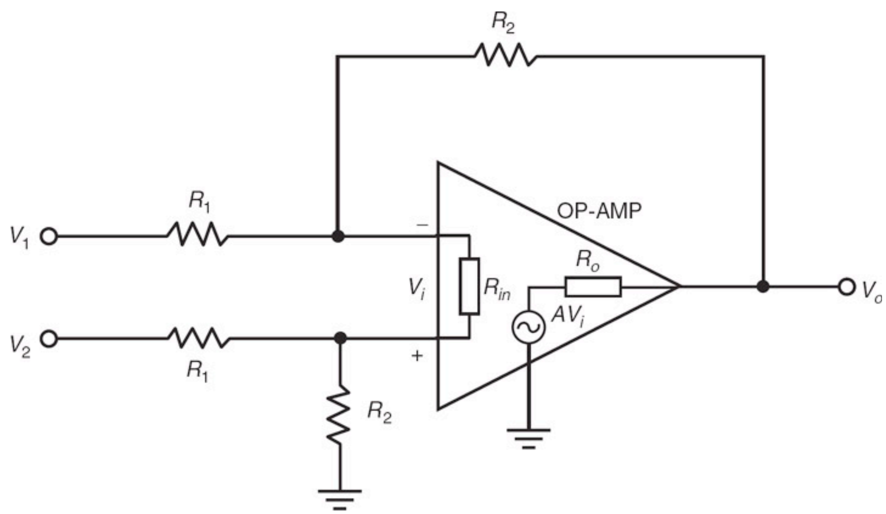


Figure 4-8: Basic differential amplifier circuit

Problem: R1, R2 mismatch

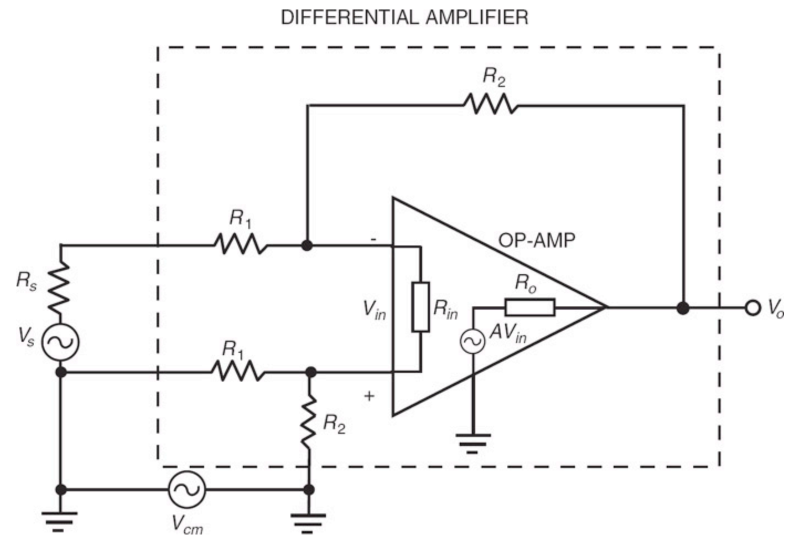


Figure 4-9: Differential amplifier driven from an unbalanced source

Problem: Unbalanced source

Instrumentation amplifier

Improve CMRR by increasing input impedance; tolerates source impedance mismatch

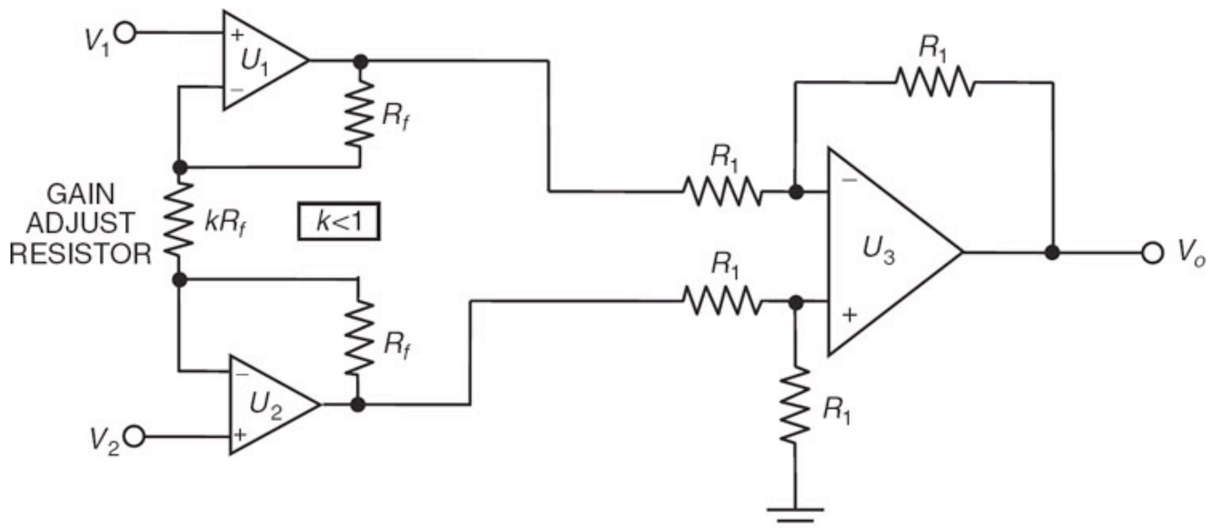


Figure 4-10: Instrumentation amplifier

Table 4-2: CMRR, in dB, for an Instrumentation Amplifier

Resistor Tolerance	A_{dm} = 1	A_{dm} = 10	A_{dm} = 100	A_{dm} =1000
1%	34	54	74	94
0.10%	54	74	94	114
0.01%	74	94	114	134

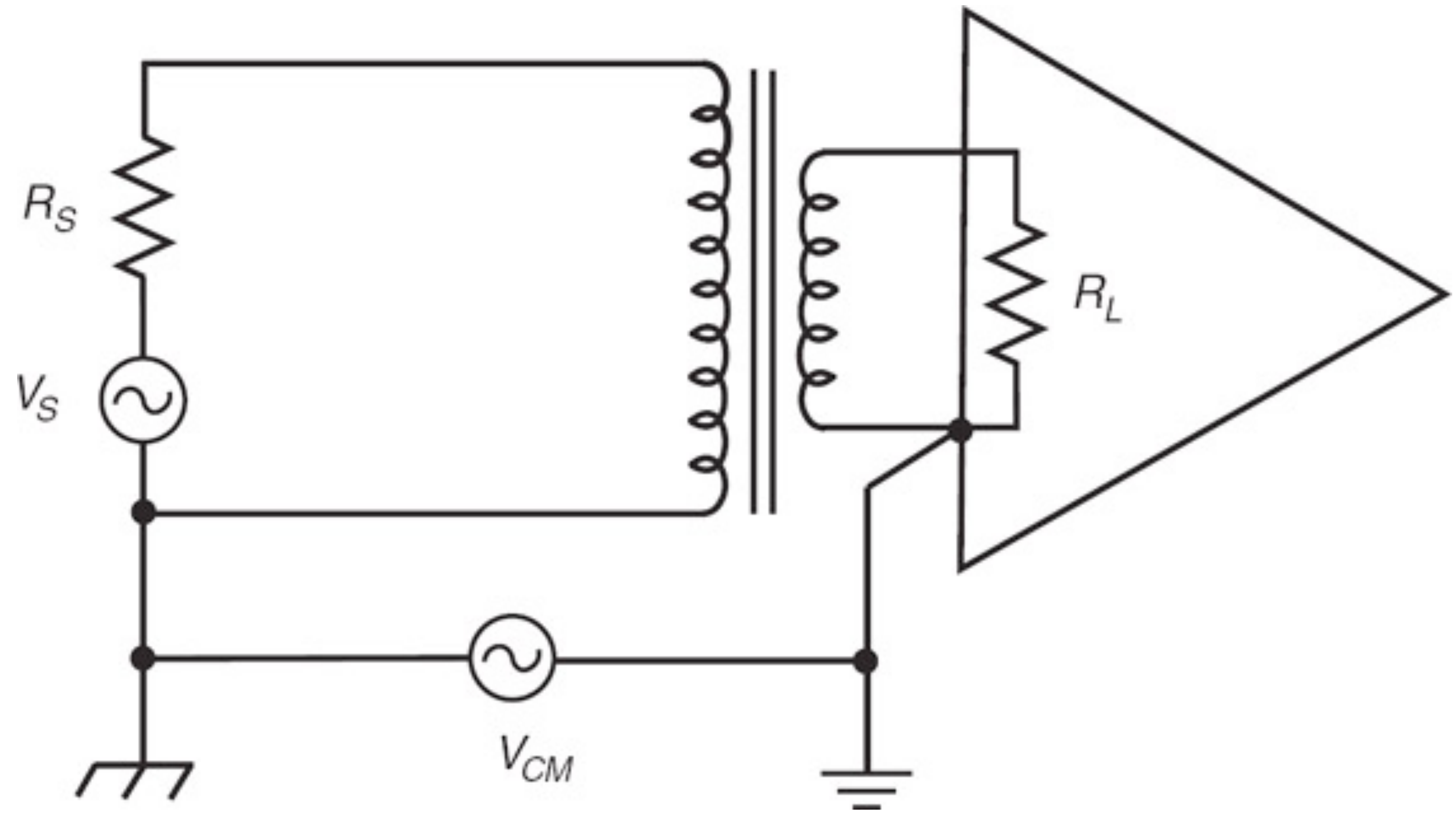


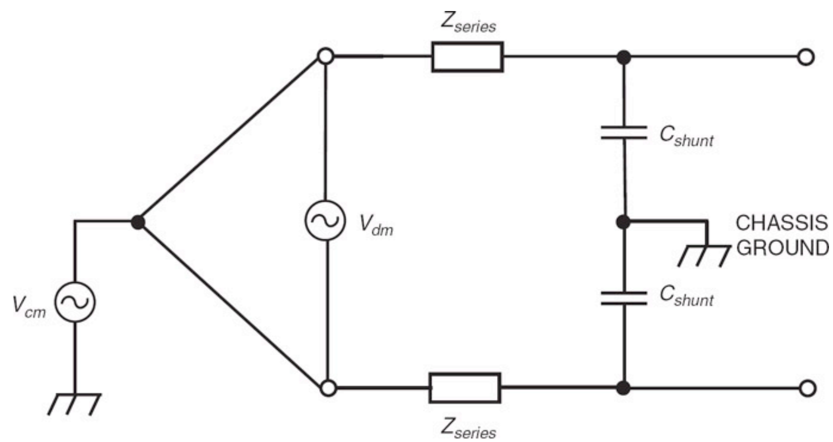
Figure 4-11: A transformer can be used to increase the common-mode load impedance and to provide galvanic isolation

The transformer can be used with a differential amplifier or single-ended amplifier as shown in Fig. 4-11.
Low-frequency CM input impedance will be determined by the insulation resistance (extremely large) between primary and secondary of the transformer.
At high frequency, the interwinding capacitance may also affect the common-mode input impedance.
Transformers also provide galvanic isolation between the source and the load.
Transformers, however, tend to be large and costly but perform well.

Filtering

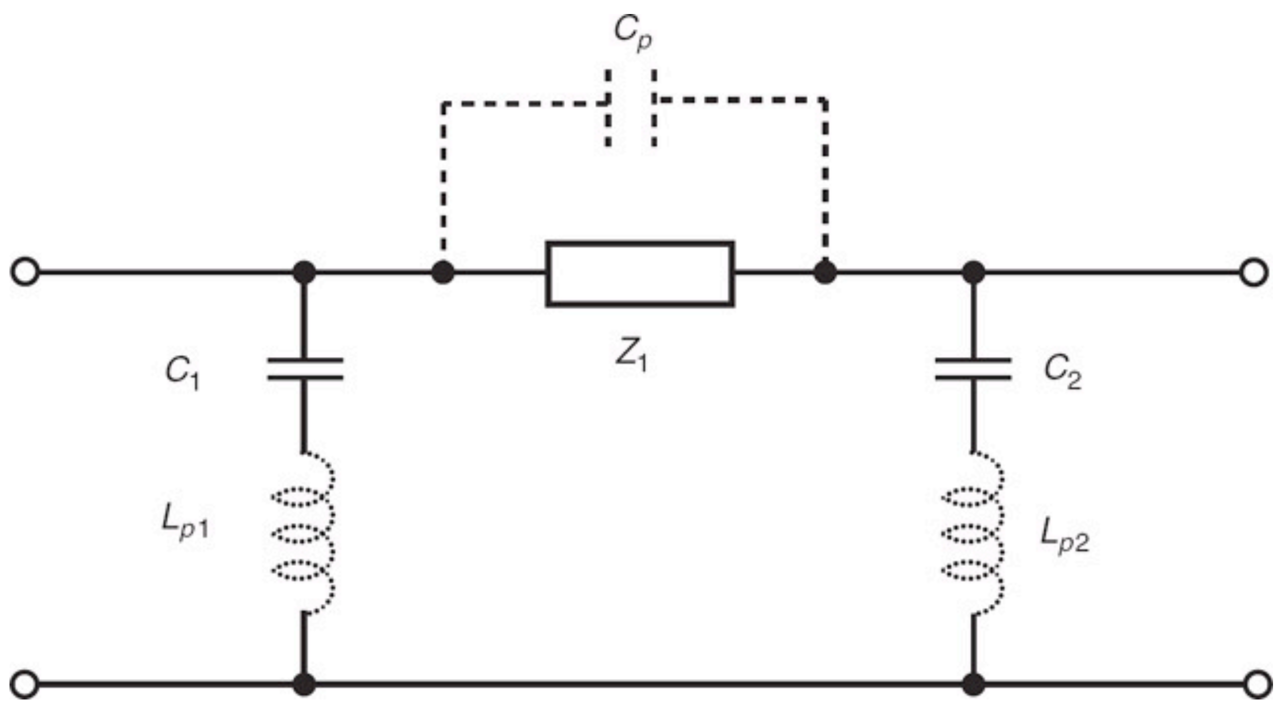
Common-mode filters are usually used to suppress noise on cables while allowing the intended differential-mode signal to pass undisturbed.
CM filters more difficult to design than differential-mode filters for three reasons:

- We usually do not know the source impedance.
- We usually do not know the load impedance.
- The filter must not distort the intentional signal (the differential-mode signal) on the cable.



Shunt element in the filter is almost always a capacitor.
Series element can be a resistor, an inductor, or a ferrite.
If the dc voltage drop can be tolerated, a resistor can be used.
If the dc voltage drop cannot be tolerated, then an inductor or ferrite should be used, both of which will have zero or very small dc voltage drop.
At low frequency (< 10 to 30 MHz), an inductor could be used; at high frequency a ferrite should be used.
Sometimes, a small value resistor can be added in series with the inductor to lower its Q.
Ferrite configured as a CM choke has the added advantage of not affecting the differential-mode signal.

Parasitic effects in filters limit effective frequency range



Summary: Balancing and Filtering

- In a balanced system, both resistive and reactive balance must be maintained.
- In a balanced system, the greater the degree of balance, or CMRR, the less noise that will couple into the system.
- Balancing can be used with shielding, to provide additional noise reduction.
- CM filters can be used but are often more difficult to design than differential-mode filters
- As the result of parasitics, all low pass filters become high pass filters above some frequency.
- Some amplifier circuits will oscillate when driving a capacitive load, unless properly compensated and/or decoupled.
- To minimize noise, the bandwidth of a system should be no more than that necessary to transmit the desired signal.

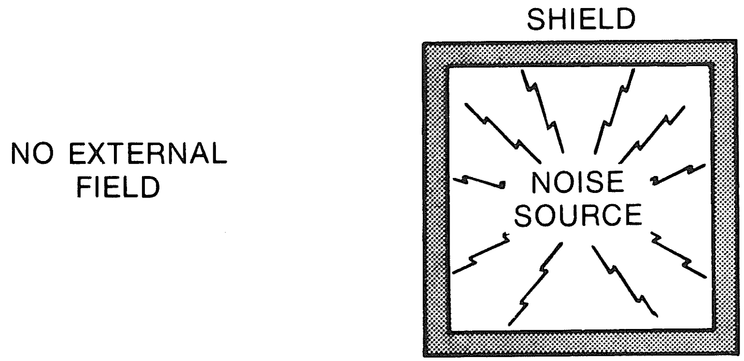


Figure 6-1. *Shield application where a noise source is contained, preventing interference with equipment outside the shield.*

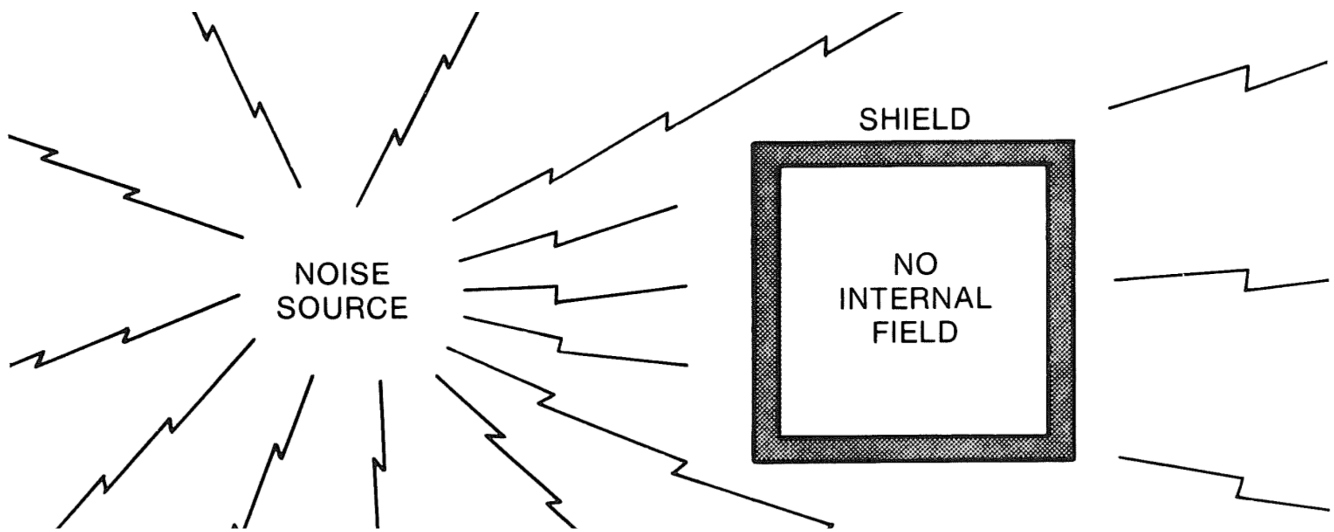


Figure 6-2. *Shield application where interference is prevented by placing a shield around a receptor to prevent noise infiltration.*

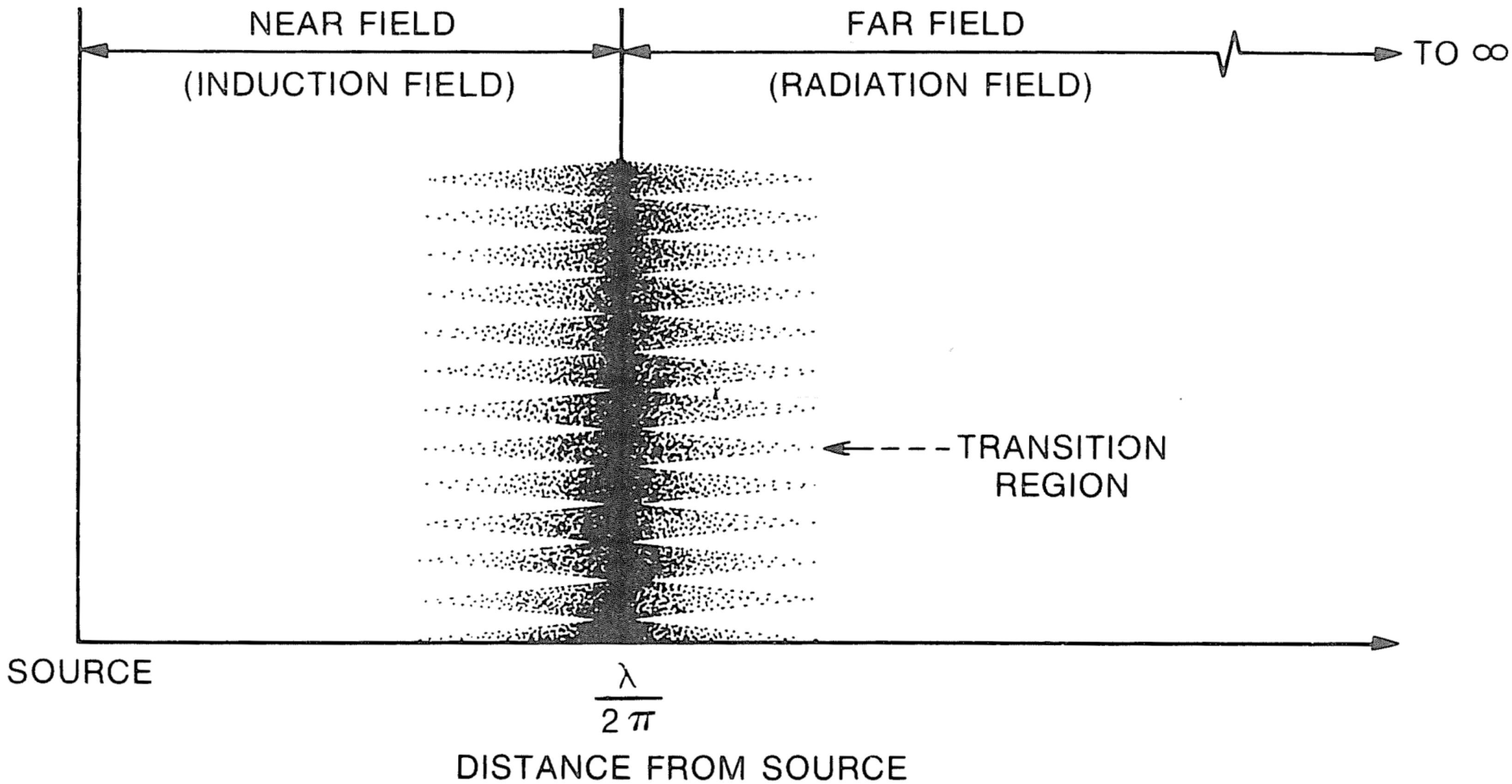


Figure 6-3. *Field character depends on the distance from the source. The transition from the near to far field occurs at $\lambda/2\pi$.*

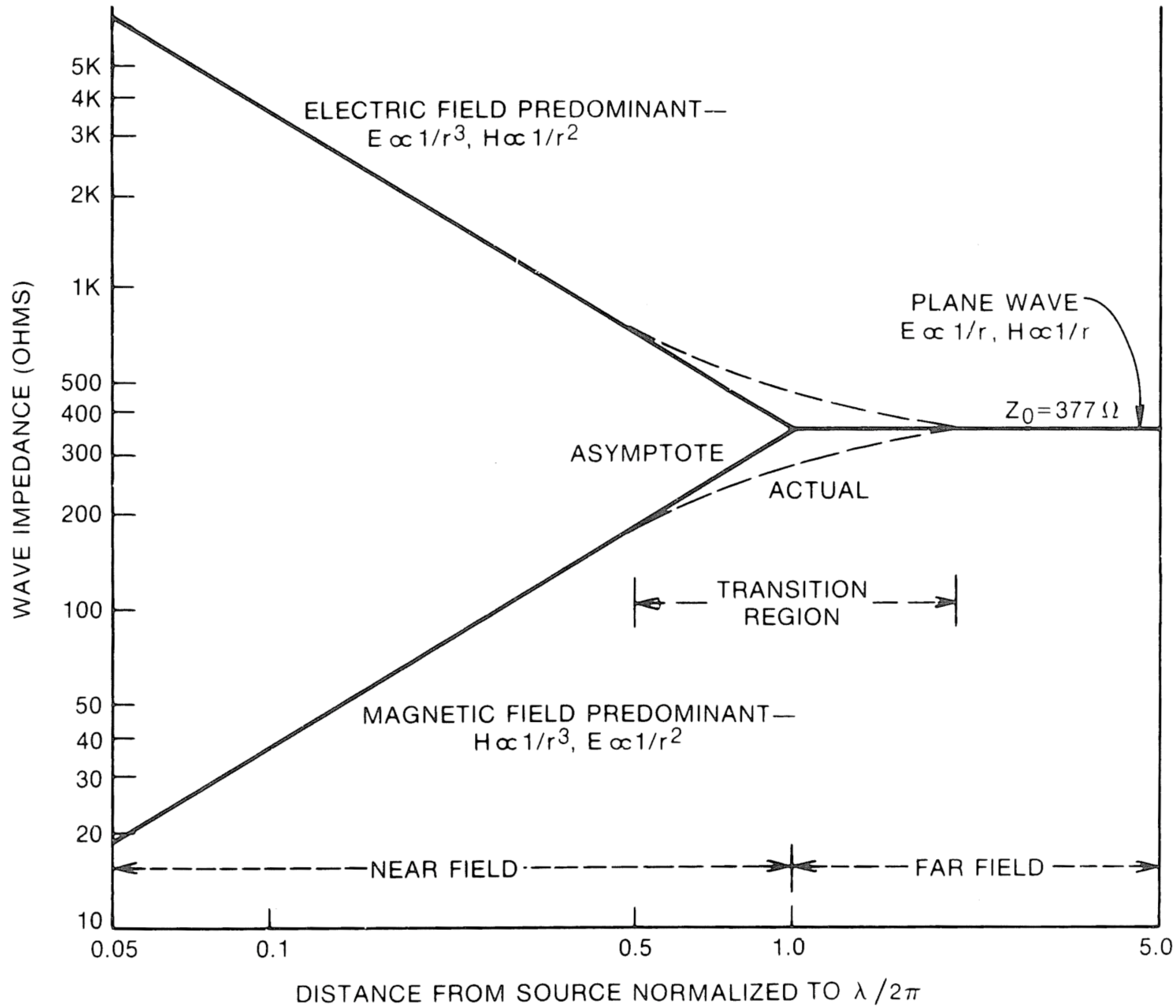


Figure 6-4. *Wave impedance depends on the distance from the source and on whether the field is electric or magnetic.*

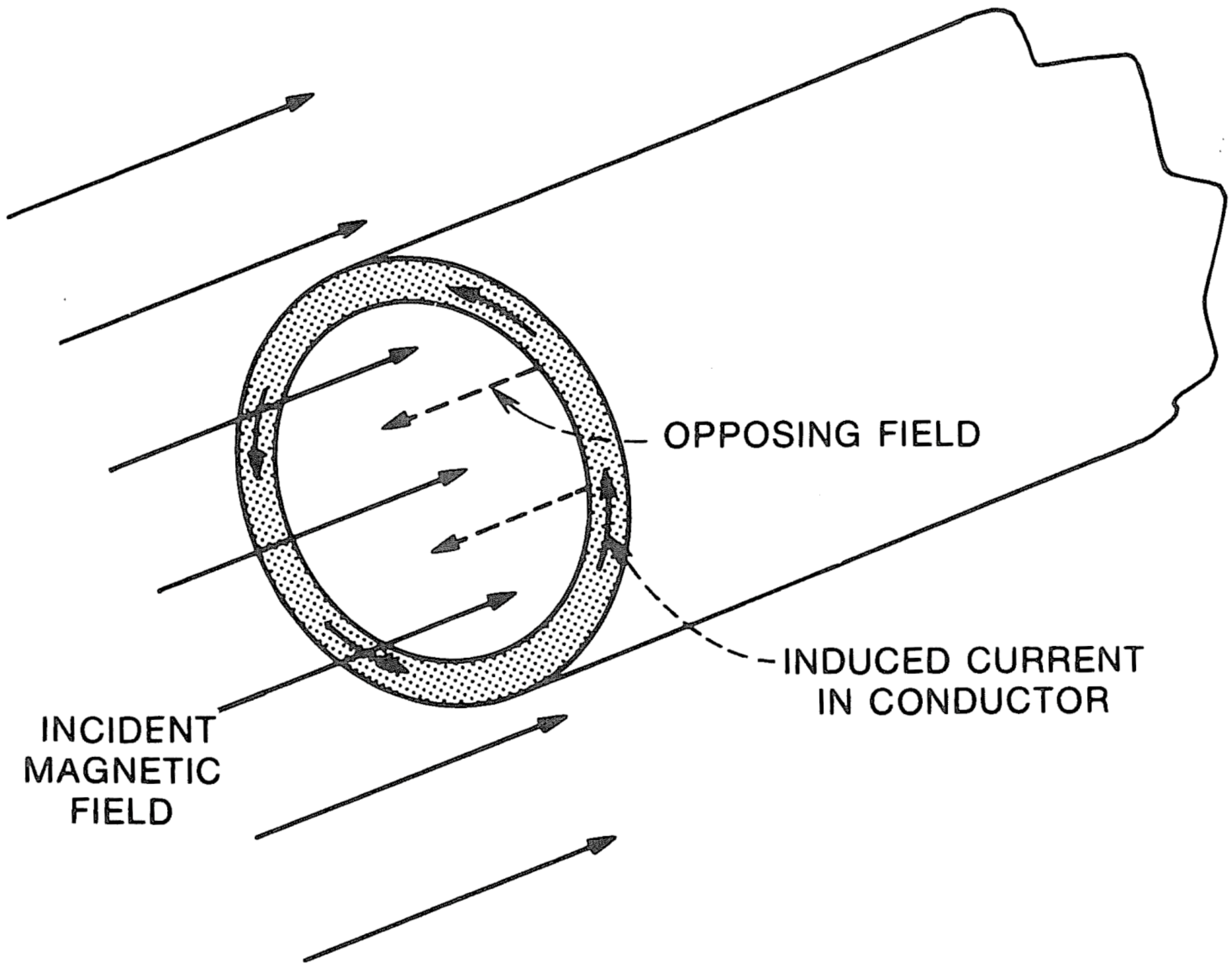


Figure 6-5. *Conducting material can provide magnetic shielding. The incident magnetic field induces currents in the conductor, producing an opposing field to cancel the incident field in the region enclosed by the shield.*

$$\delta = \sqrt{\frac{2}{\omega\mu\sigma}} \quad \text{m} .^{*} \tag{6-11a}$$

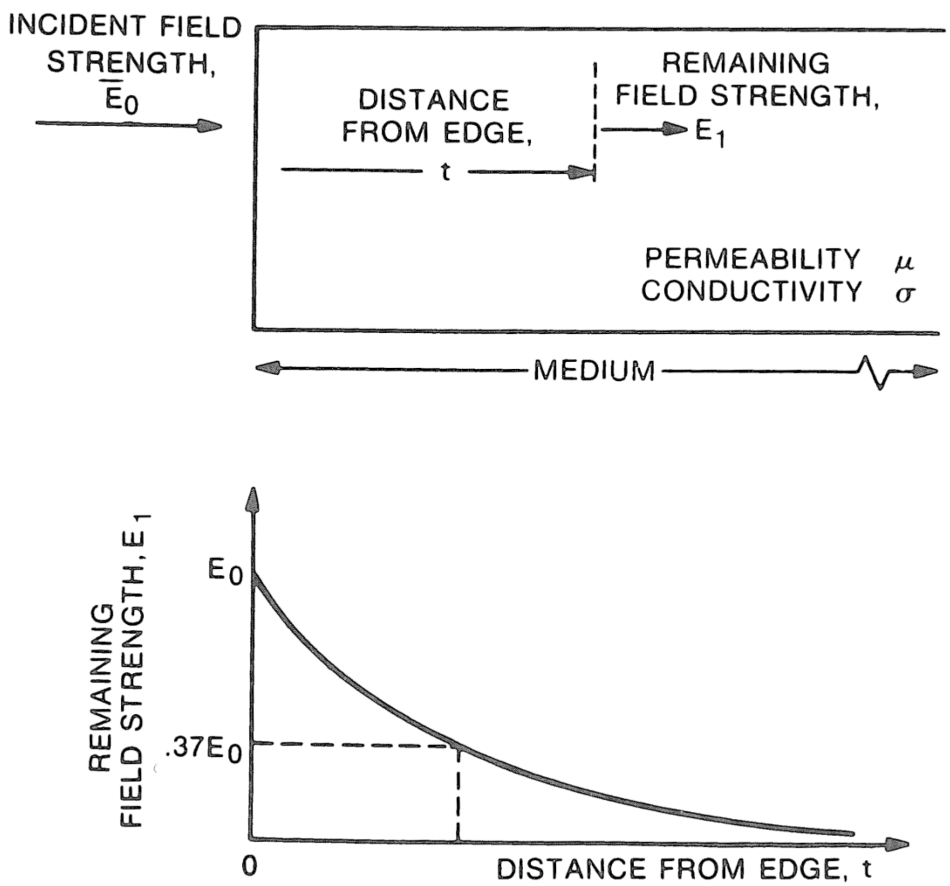


Figure 6-6. *Electromagnetic wave passing through an absorbing medium is attenuated exponentially.*

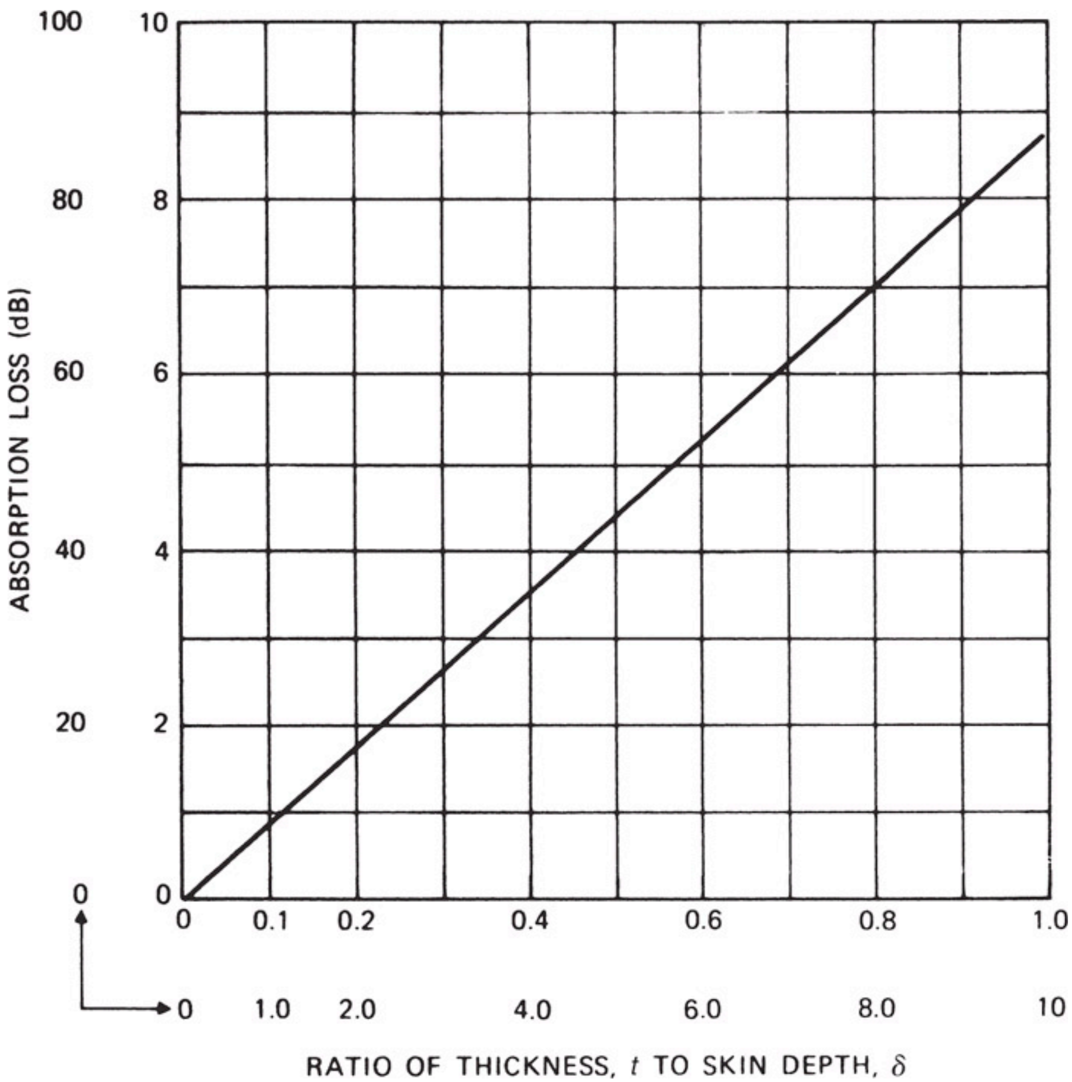


Figure 6-7: Absorption loss is proportional to the thickness and inversely proportional to the skin depth of the shield material. This plot can be used for electric fields, magnetic fields, or plane waves

Table 6-2: Skin Depths of Various Materials

Frequency	Copper (in)	Aluminum (in)	Steel (in)	Mumetal (in)
60 Hz	0.335	0.429	0.034	0.014
100 Hz	0.260	0.333	0.026	0.011
1 kHz	0.082	0.105	0.008	0.003
10 kHz	0.026	0.033	0.003	—
100 kHz	0.008	0.011	0.0008	—
1 MHz	0.003	0.003	0.0003	—
10 MHz	0.0008	0.001	0.0001	—
100 MHz	0.00026	0.0003	0.00008	—
1000 MHz	0.00008	0.0001	0.00004	—

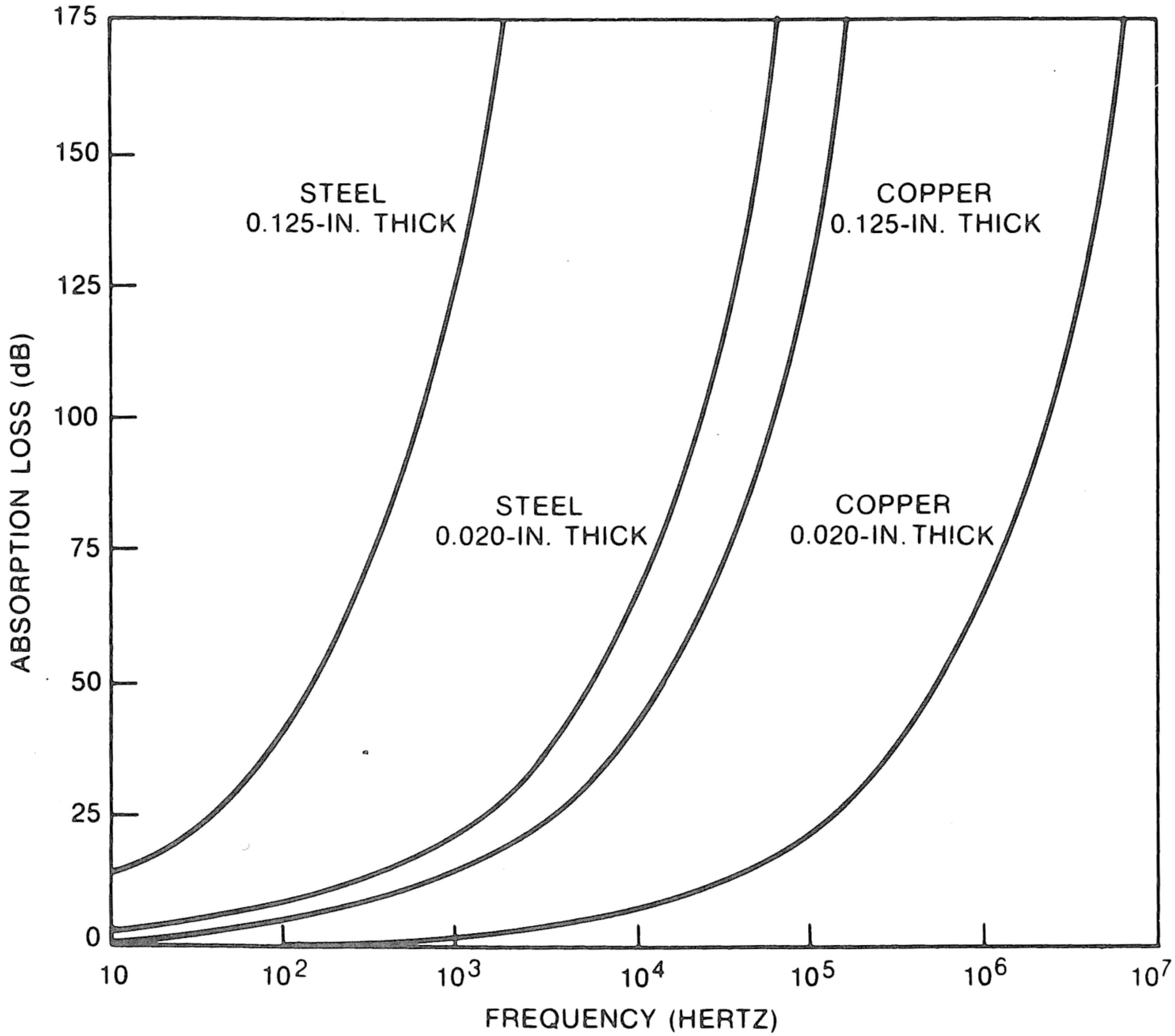


Figure 6-9: Absorption loss increases with frequency and with shield thickness; steel offers more absorption loss than copper of the same thickness

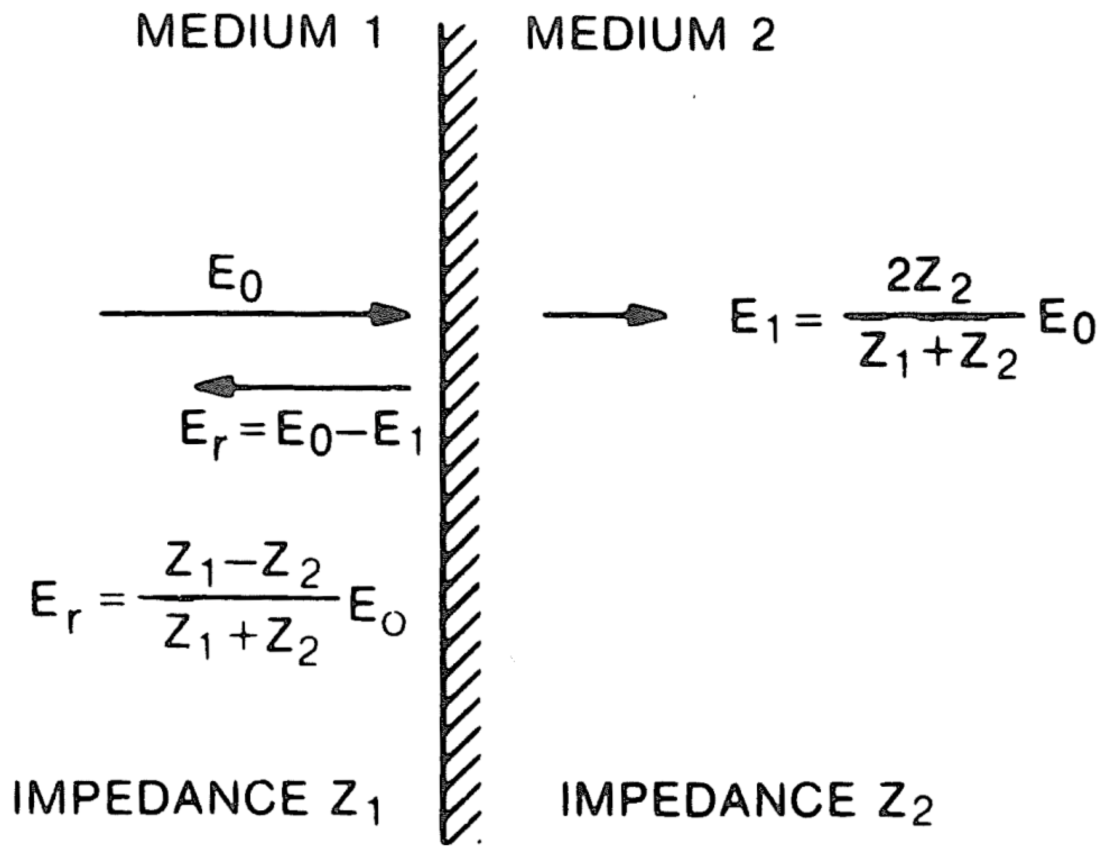


Figure 6-10: An incident wave is partially reflected from, and partially transmitted through, an interface between two media. The transmitted wave is E_1 and the reflected wave is E_r

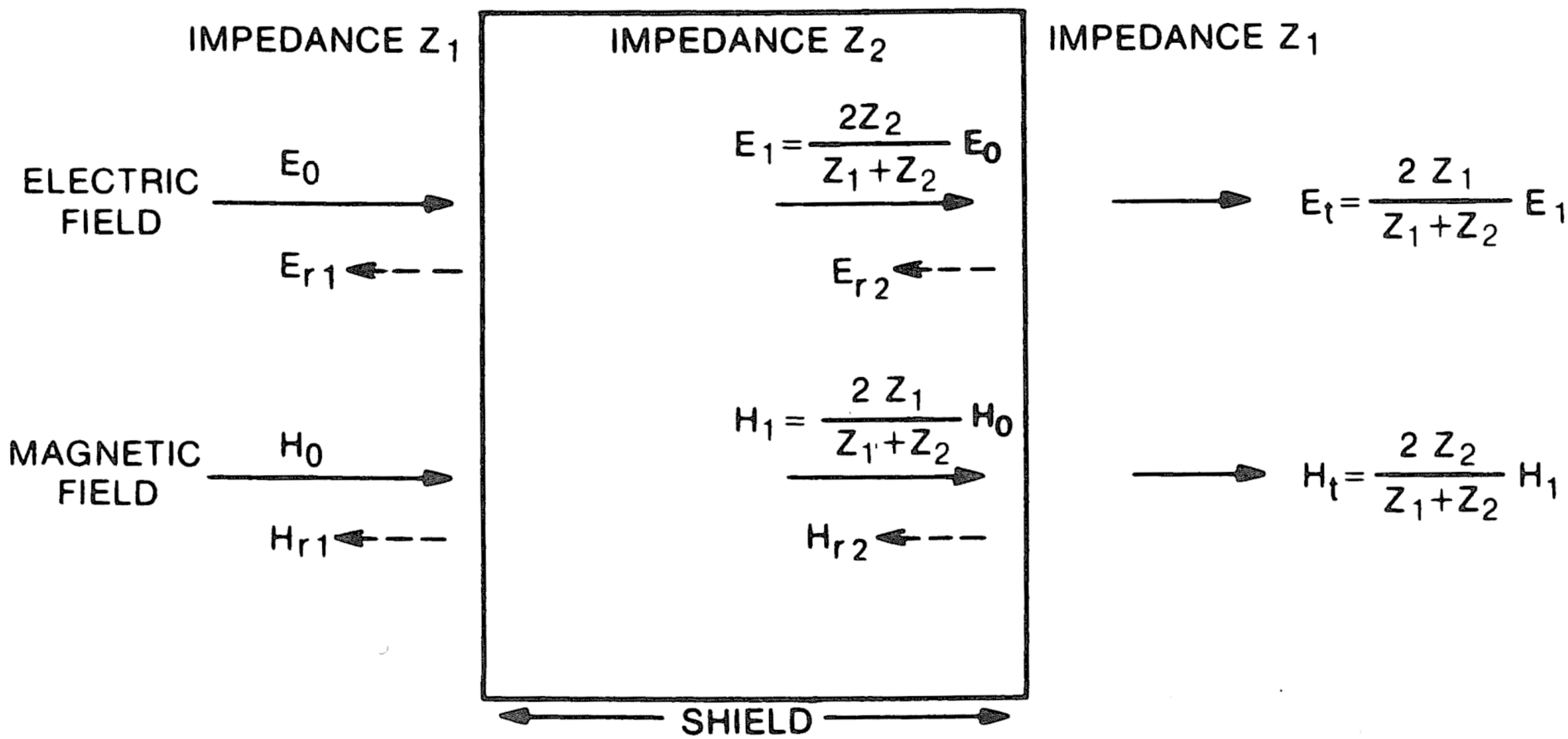


Figure 6-11: Partial reflection and transmission occur at both boundaries of a shield

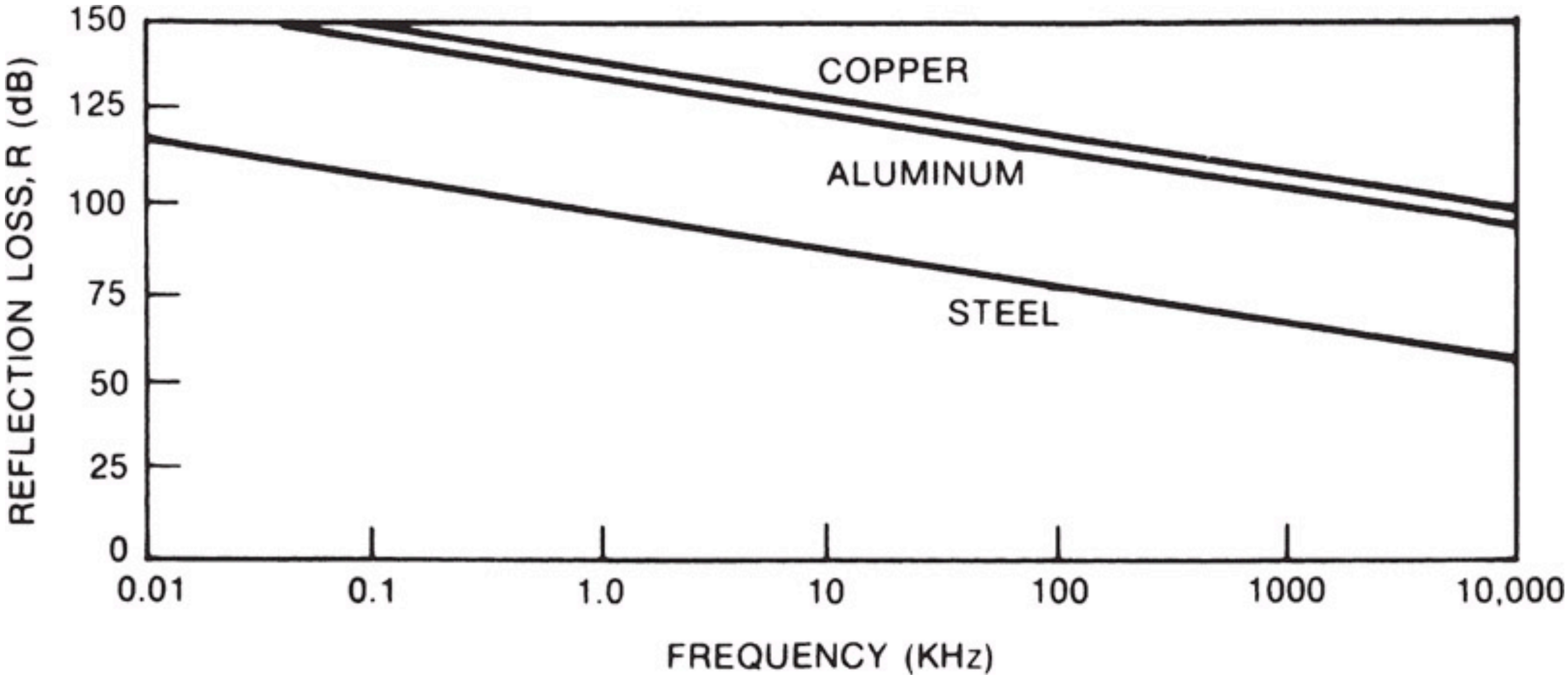


Figure 6-12: Reflection loss for plane waves is greater at low frequencies and for high conductivity material

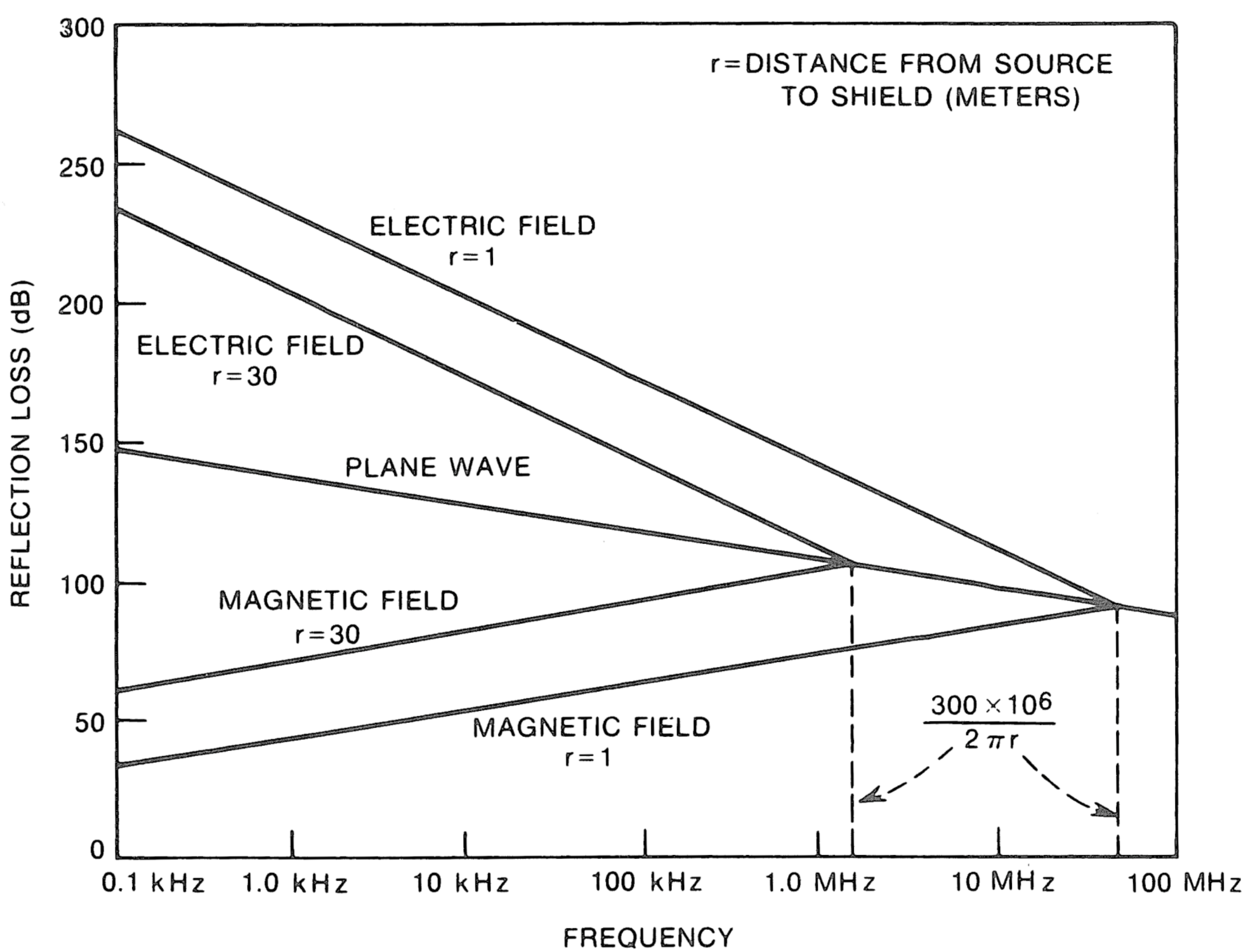


Figure 6-13: Reflection loss in a copper shield varies with frequency, distance from the source, and type of wave

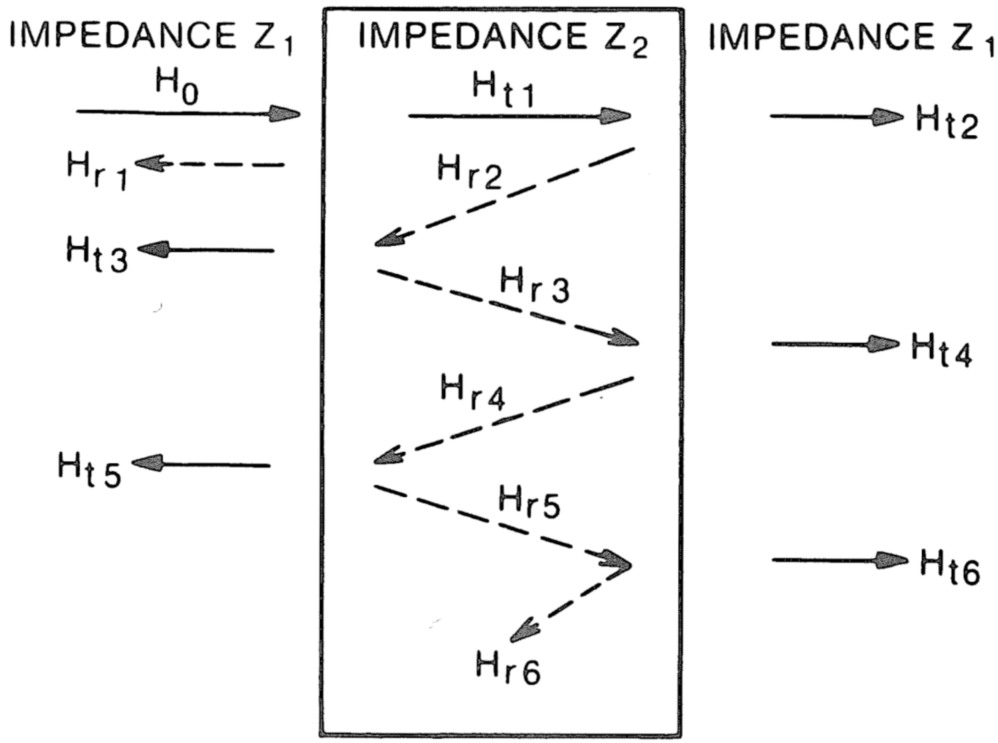


Figure 6-14: Multiple reflections occur in a thin shield; part of the wave is transmitted through the second boundary at each reflection

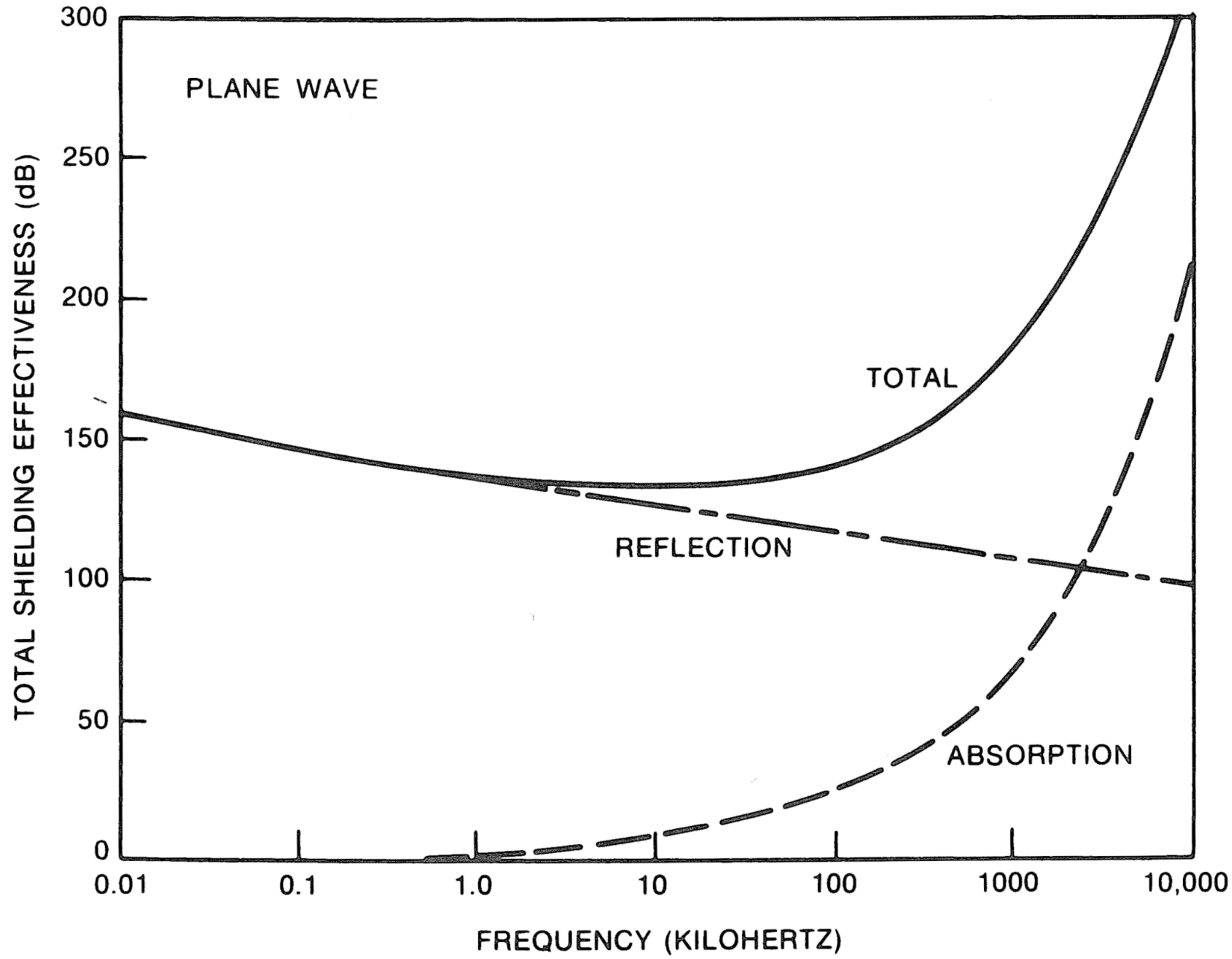


Figure 6-16: Shielding effectiveness of a 0.02-in thick copper shield in the far field

At *low-frequency*, reflection loss in the primary shielding mechanism for electric fields.
At *high-frequency*, absorption loss is the primary shielding mechanism.

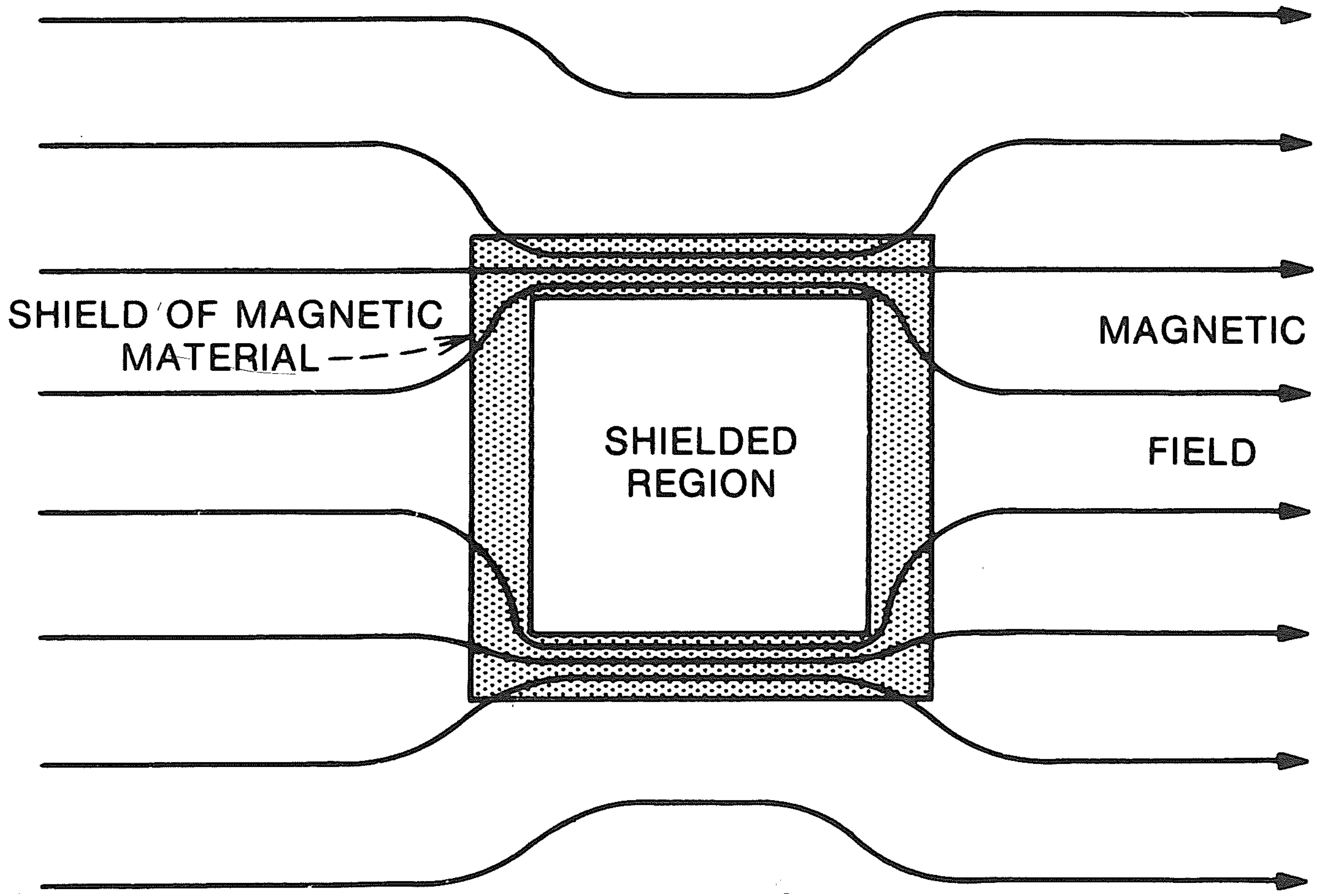


Figure 6-17: Magnetic material used as a shield by providing a low-reluctance path for the magnetic field, diverting it around the shielded region

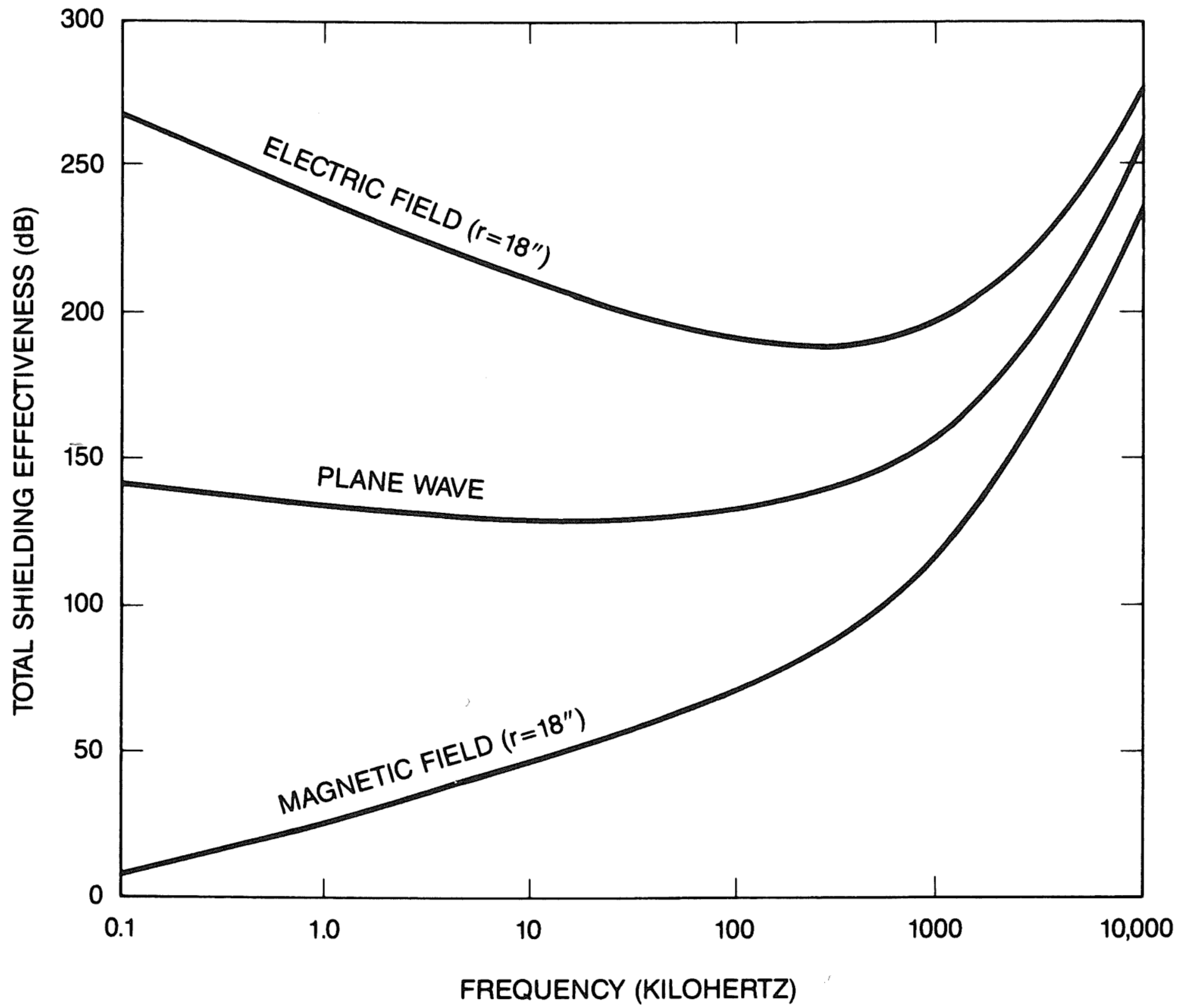


Figure 6-18: Electric field, plane wave, and magnetic field shielding effectiveness of a 0.02-in-thick solid aluminum shield

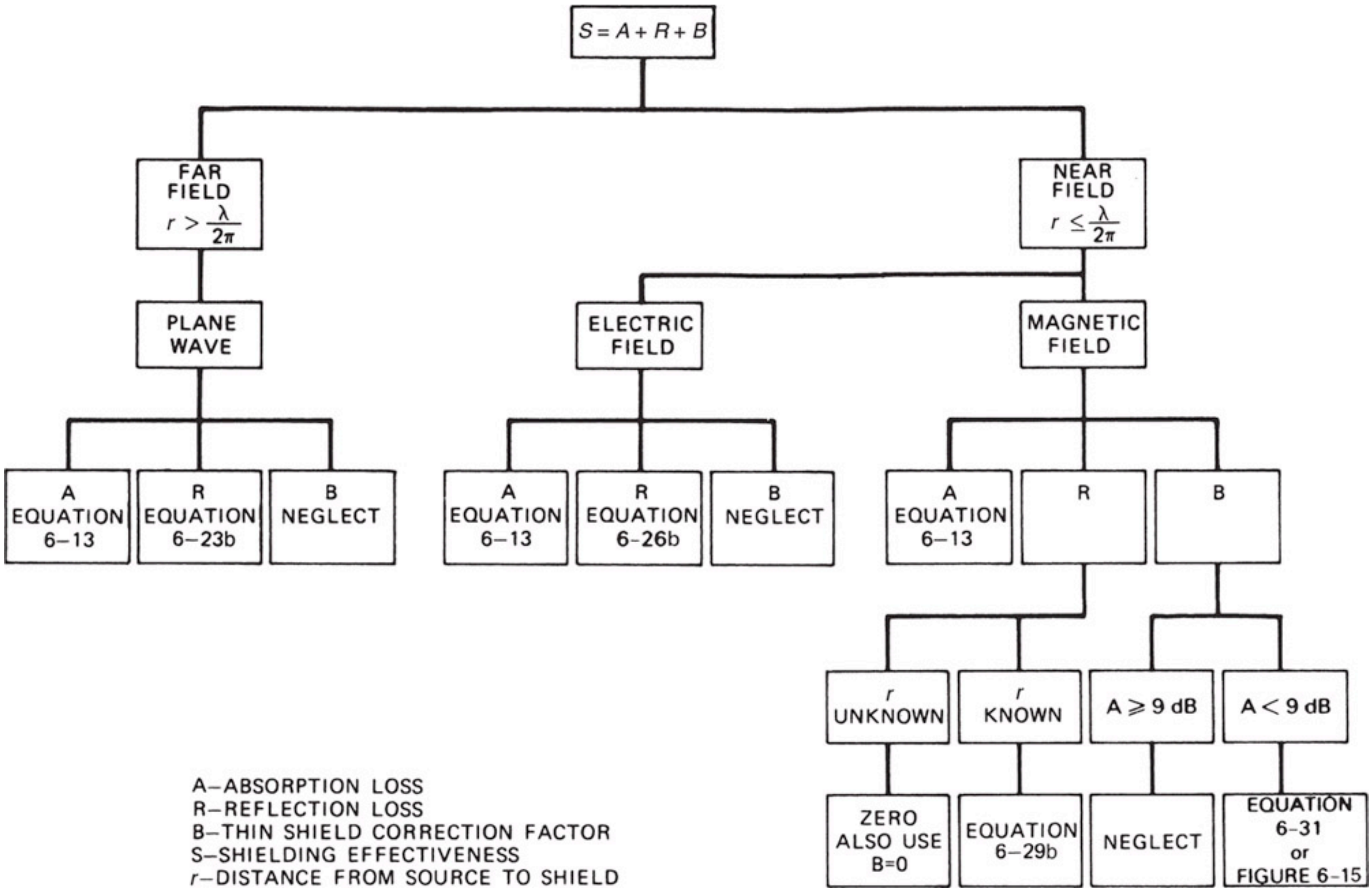


Figure 6-19: Shielding effectiveness summary shows which equations are used to calculate shielding effectiveness under various conditions

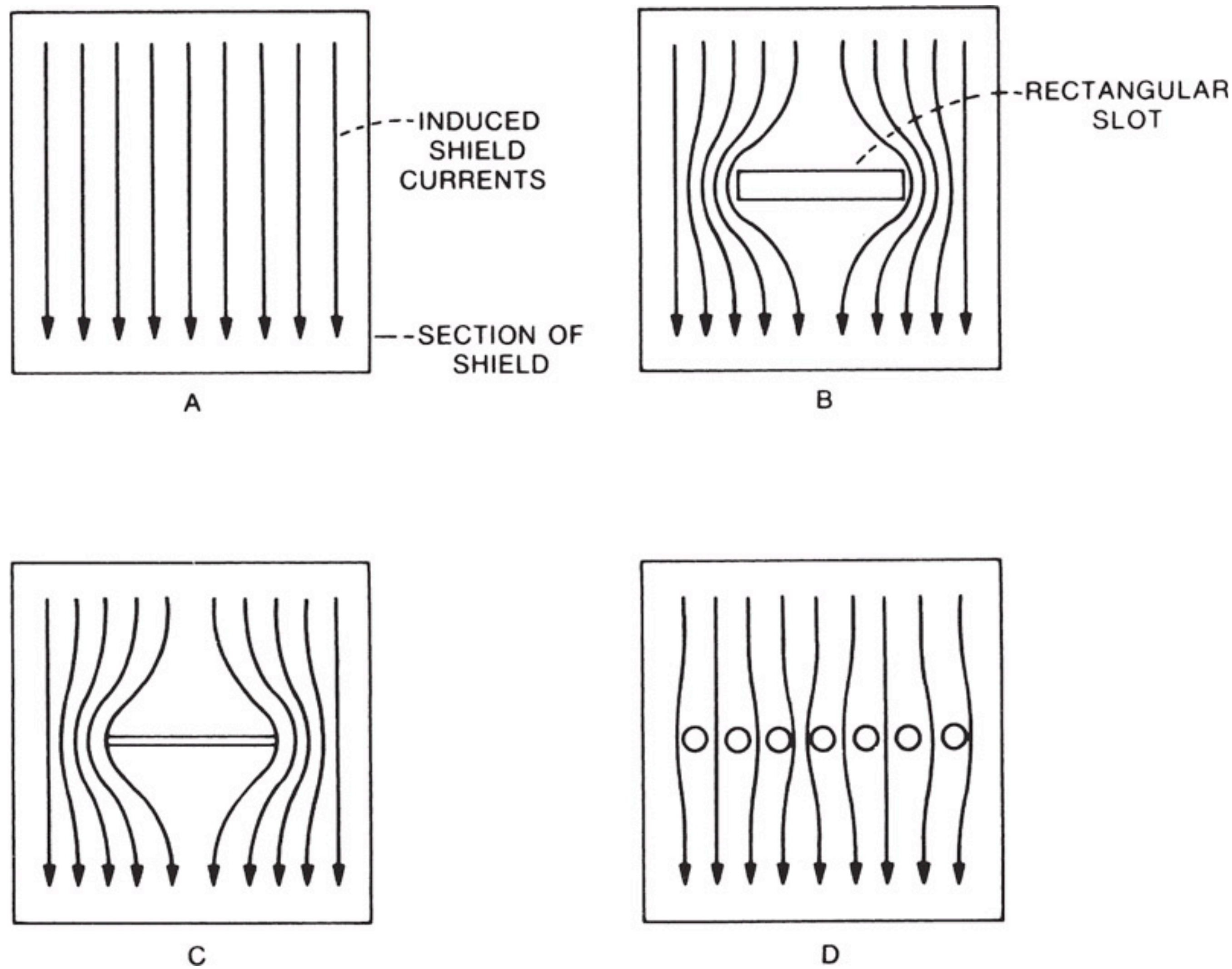


Figure 6-25: Effect of shield discontinuity on magnetically induced shield currents

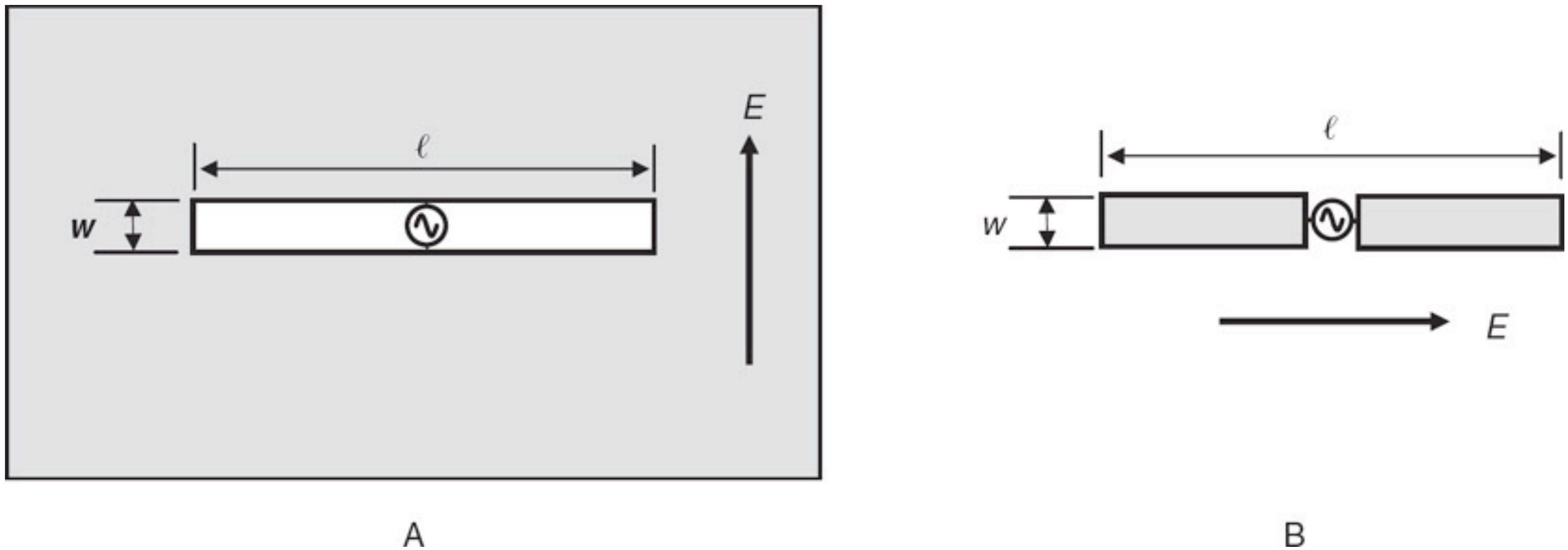


Figure 6-26: (A) A slot antenna, and (B) its complementary dipole antenna

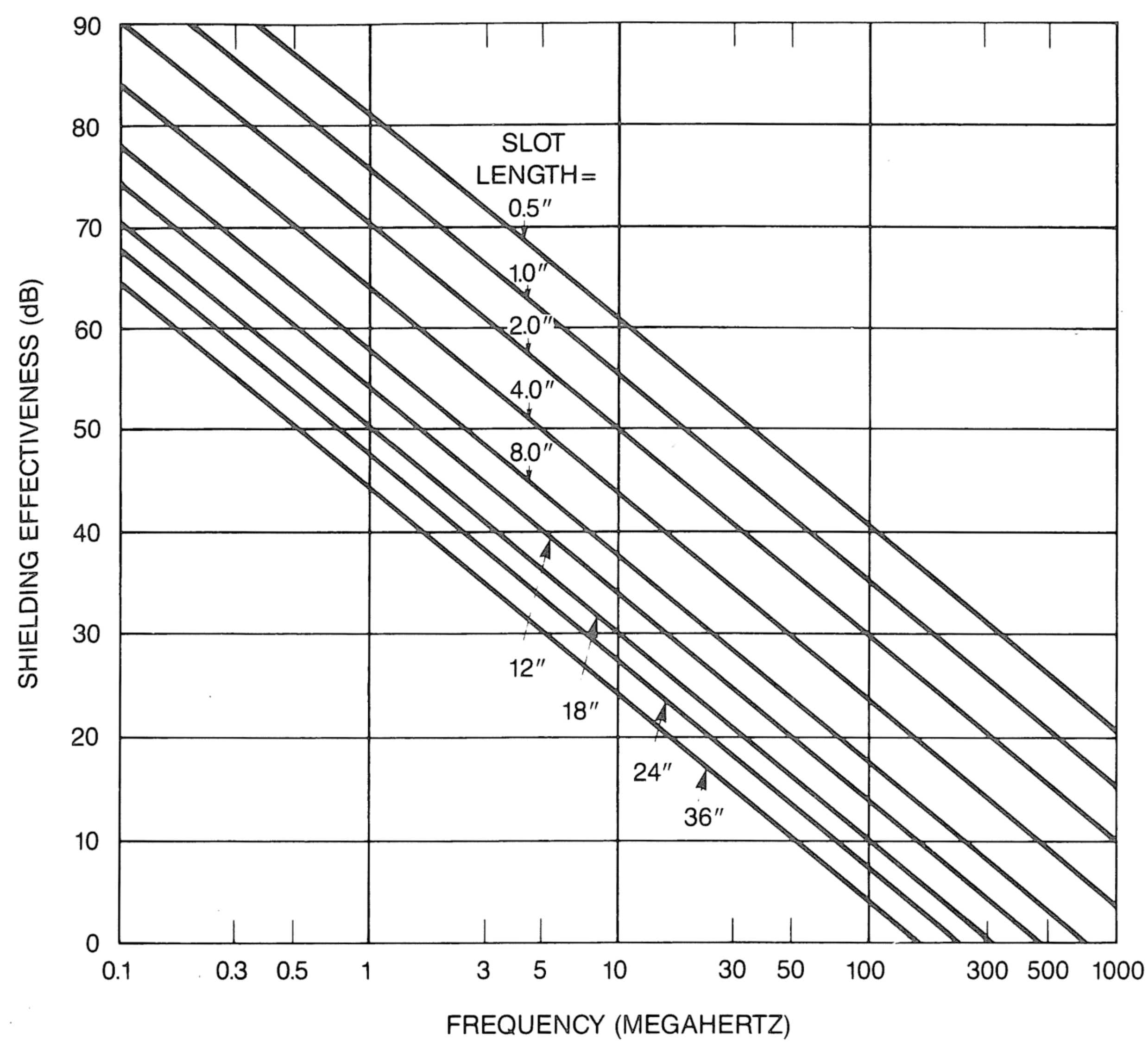


Figure 6-27: Shielding effectiveness versus frequency and maximum slot length for a single aperture

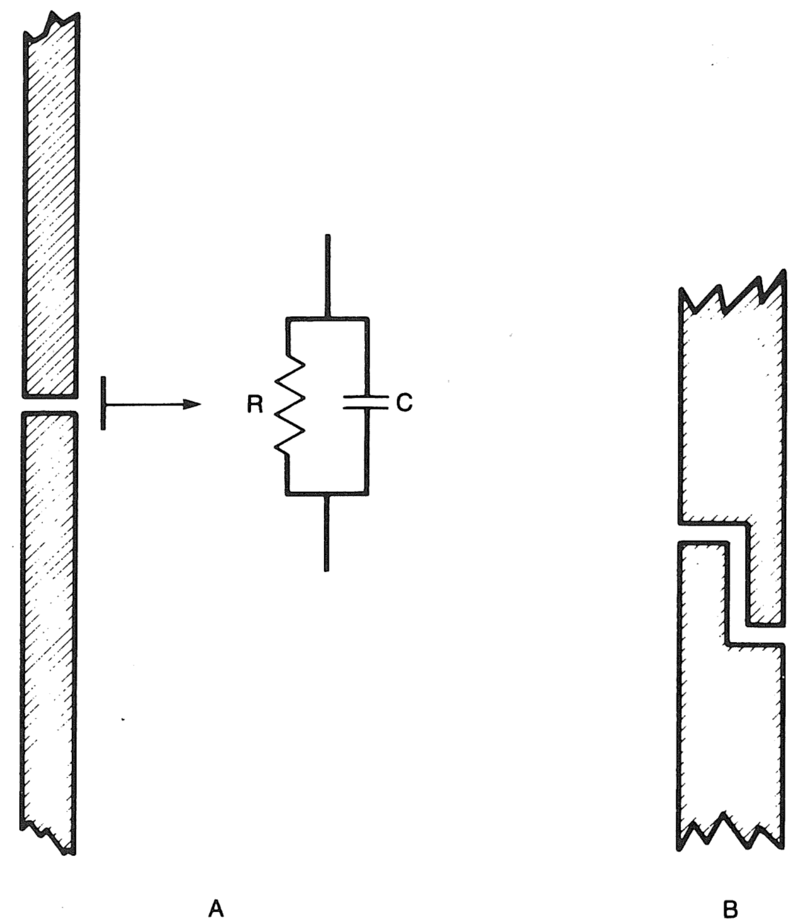


Figure 6-31: (A) The impedance of a seam consists of a resistive and capacitive component; (B) seam overlap increases the capacitance across the joint

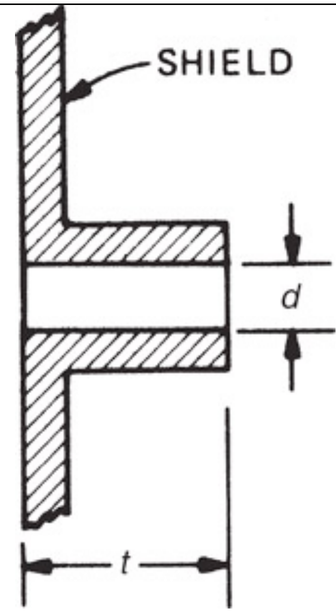


Figure 6-32: Cross section of a hole formed into a waveguide with diameter *d* and depth *t*

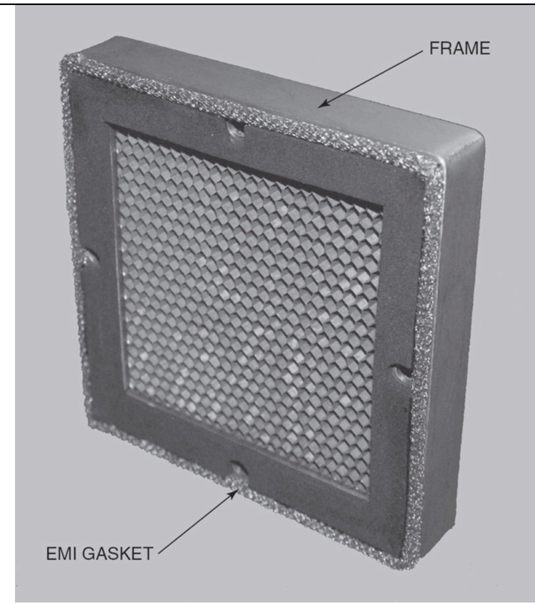


Figure 6-33: Honeycomb ventilation panel, rear view showing EMI gasket used to make electrical contact to chassis that it is mounted on (courtesy of MAJR Products Corp.)

The classic example of the application of this principal is a honeycomb ventilation panel as shown in Fig. 6-33. When mounted, the entire perimeter of the panel must make electrical contact to the chassis. The maximum dimension of the holes is usually 1/8-in, and the panels are usually 1/2-in., thick, giving a *t/d* ratio of 4 and 128 dB additional shielding from the waveguide effect.

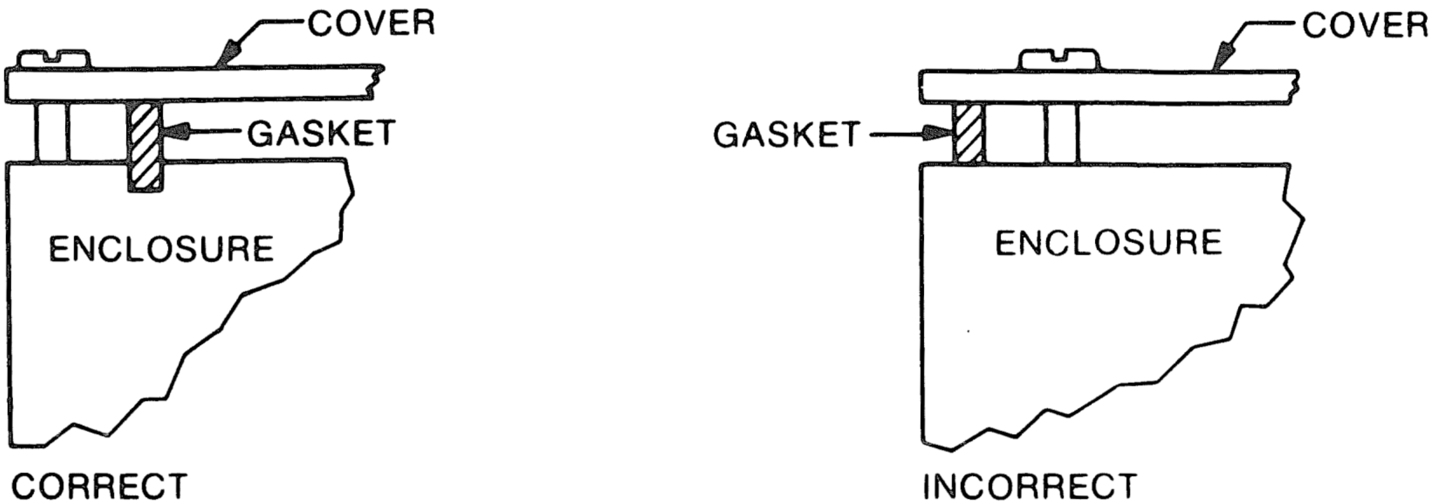


Figure 6-35: EMI gaskets, correct and incorrect installation

The gasket should be in a slot and on the inside of the screws to protect against leakage around the screw holes. For electrical continuity across the joint or seam, the metal should be free of paints, oxides, and insulating films. The mating surfaces should be protected from corrosion with a conductive finish.

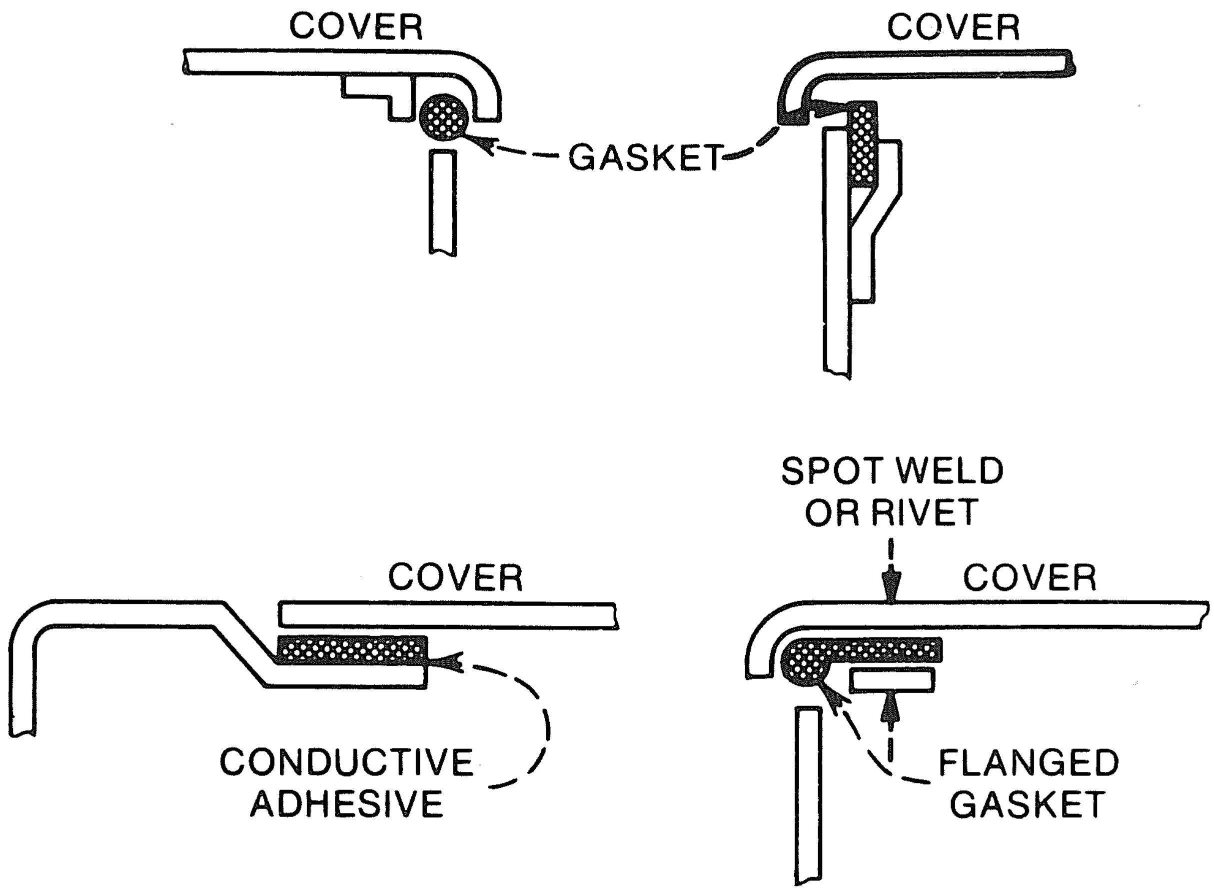
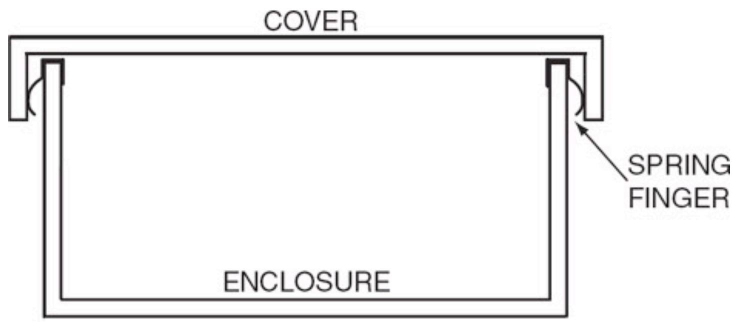


Figure 6-36: Suitable ways to install EMI gaskets in sheet metal enclosures

In the seam designs shown in both Figs. 6-35 and 6-36, the EMI gaskets are in compression and require screws or some other means to provide the necessary pressure to properly compress the gasket.

With the seam designs shown in Fig. 6-37, however, the gaskets are mounted in shear, and no fasteners are required to provide the necessary compression.
Performance of the seam is not a function of the exact positioning of the cover or front panel.



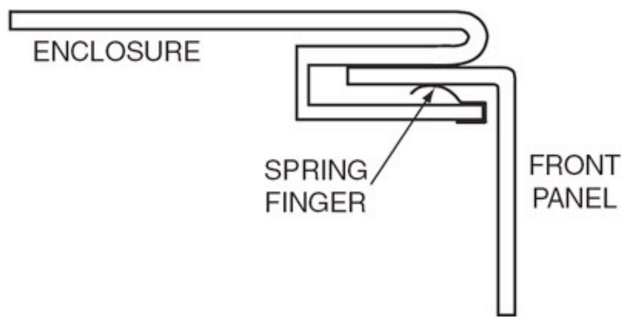


Figure 6-37: EMI gasket mounting methods that require no fasteners. Gasket is mounted in shear not compression

For optimum shielding, the enclosure should be thought of as "electrically watertight," with EMI gaskets used in place of normal environmental gaskets.

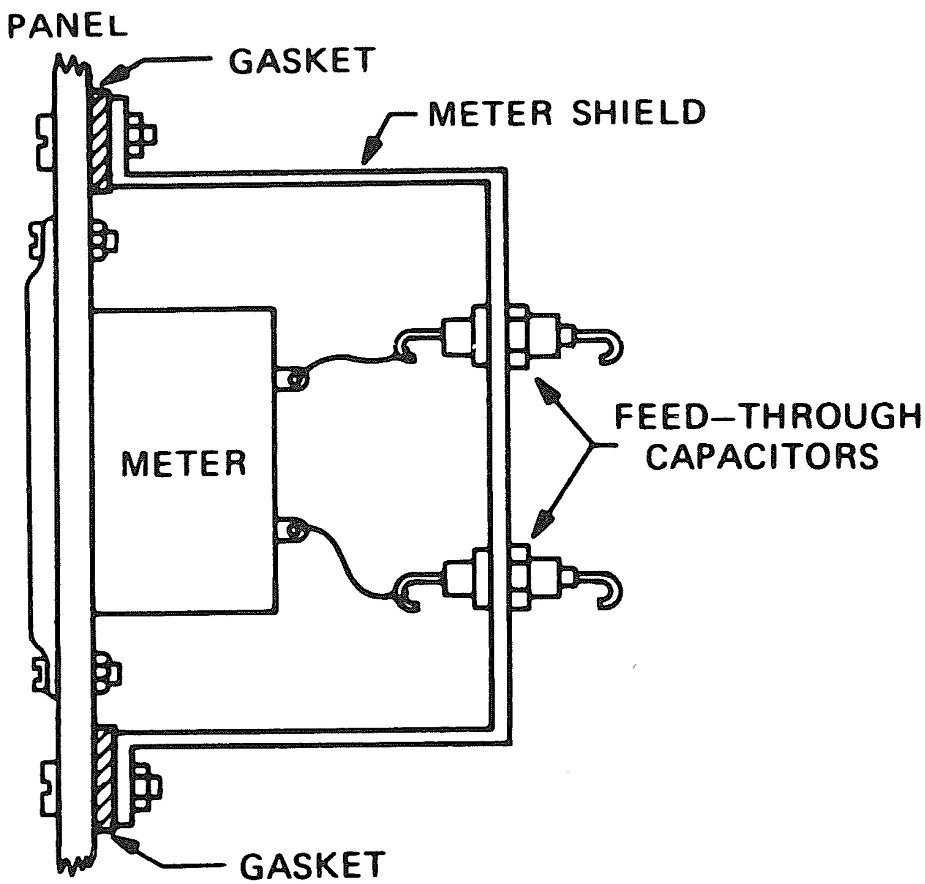


Figure 6-39: Method of mounting a meter or display in a shielded panel

Internal shields

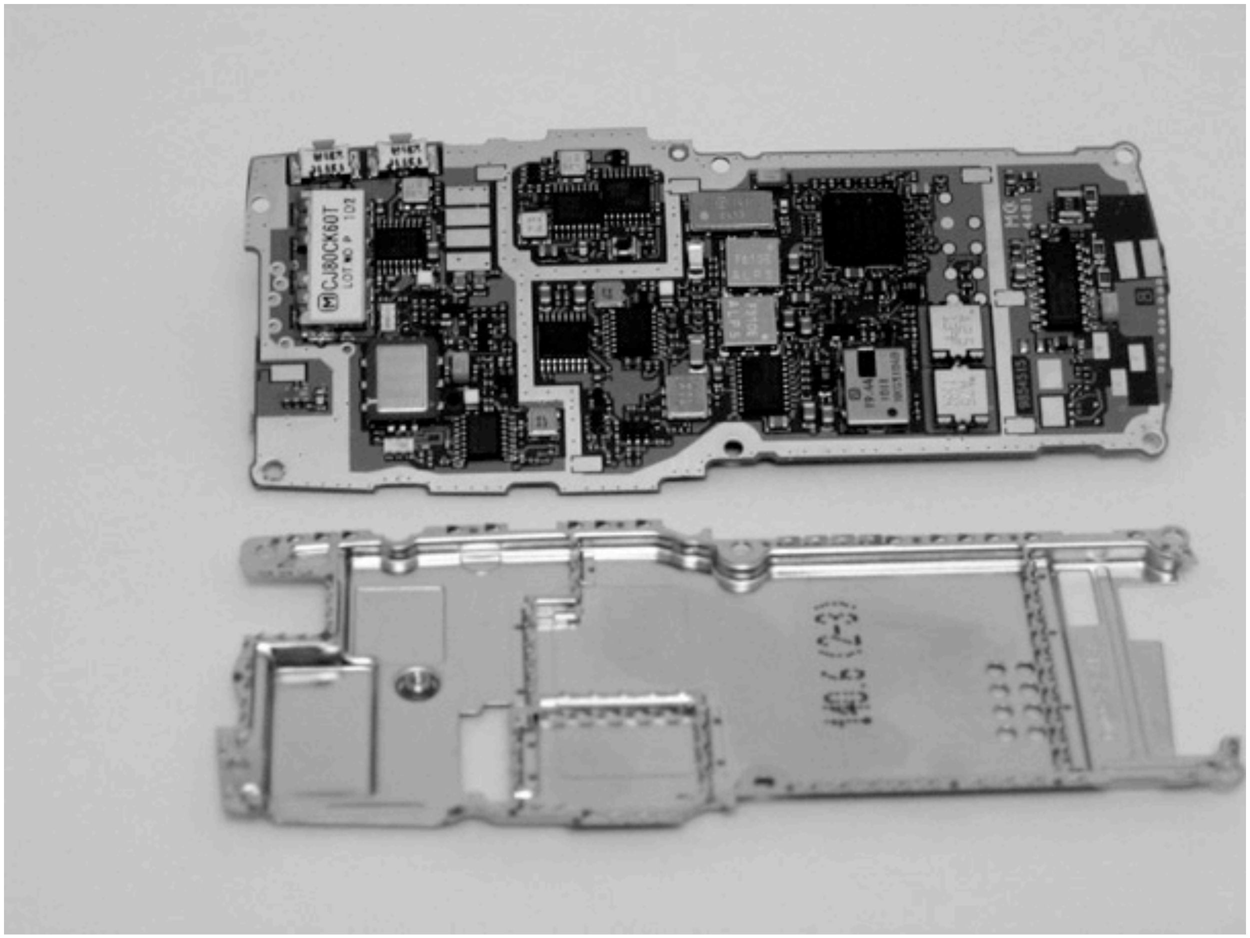


Figure 6-41: A PCB (top) with its board level shield (bottom)

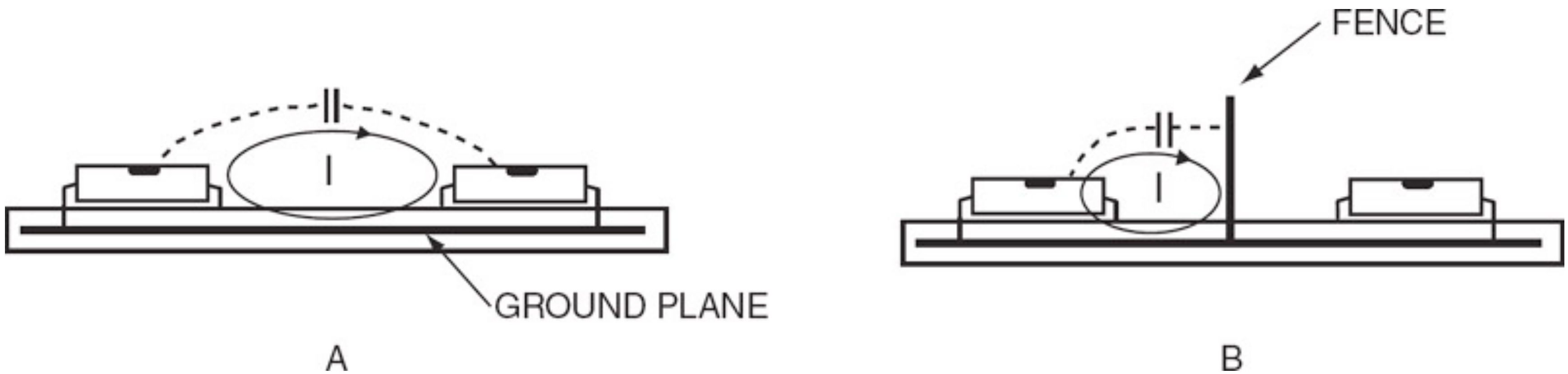


Figure 6-42: *(A)* Parasitic capacitance coupling between integrated circuits (ICs), *(B)* parasitic coupling blocked by the use of a PCB fence. Current *I* represents noise coupling through parasitic capacitance

Shielding can even be applied within a PCB by using power and/or ground planes to shield layers located between them. The high-speed signals are routed on layers 3 and 4, and they are shielded by the planes on layers 2 and 5. A 1-oz copper plane in a PCB will be greater than three skin depths thick at all frequencies above 30 MHz. For a 2 oz copper plane, this will be true above 10 MHz.

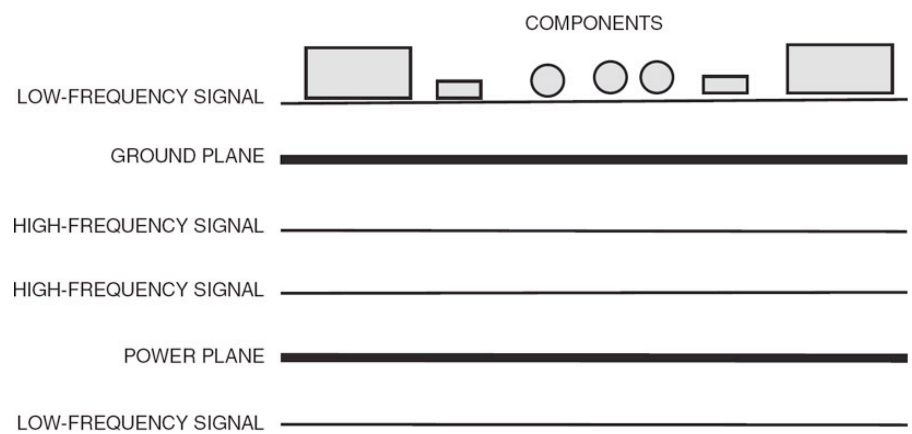


Table 6-7 Qualitative Summary of Shielding Effectiveness

Material	Frequency (kHz)	Absorption Loss[a] All Fields	Reflection Loss: Magnetic Field[b]	Reflection Loss: Electric Field	Reflection Loss: Plane Wave
Magnetic ($\mu_r = 1000$, $\sigma_r = 0.1$)	<1	Bad–poor	Bad	Excellent	Excellent
	1–10	Averaged–good	Bad–poor	Excellent	Excellent
	10–100	Excellent	Poor	Excellent	Good
	>100	Excellent	Poor–average	Good	Average–good
Non magnetic ($\mu_r = 1$, $\sigma_r = 1$)	<1	Bad	Poor	Excellent	Excellent
	1–10	Bad	Average	Excellent	Excellent
	10–100	Poor	Average	Excellent	Excellent
	>100	Average–good	Good	Excellent	Excellent

Key	*Attenuation*
Bad	0–10 dB
Poor	10–30 dB
Average	30–60 dB
Good	60–90 dB
Excellent	>90 dB

[a]Absorption loss for 1/32-in. thick shield.
[b]Magnetic field reflection loss for a source distance of 1 m. (Shielding is less if distance is less than 1 m, and more if distance is greater than 1 m.)

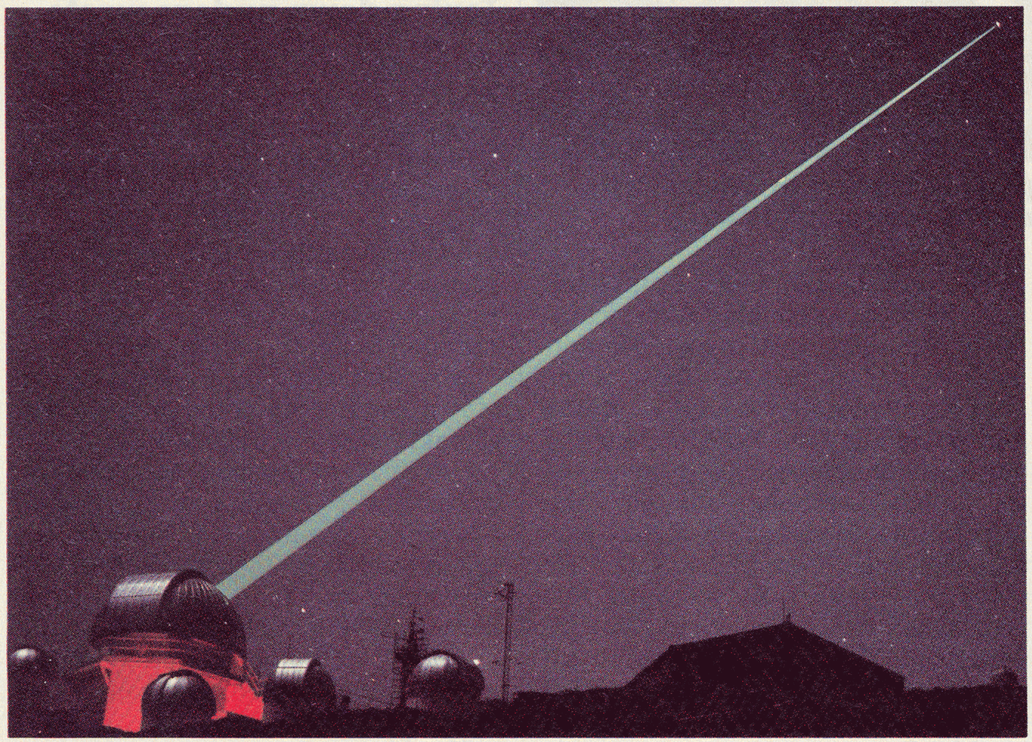
Shielding: Summary

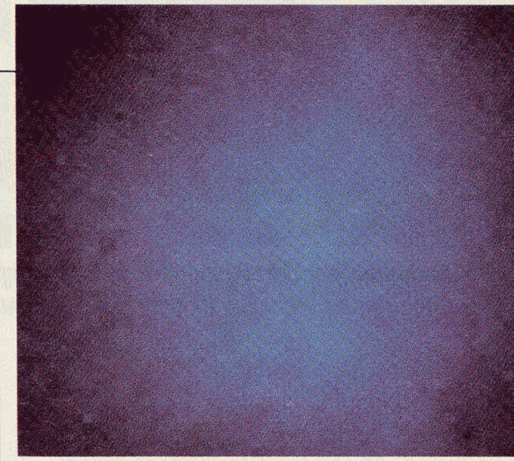
- All cables that enter or leave a shielded enclosure should be shielded or filtered.
- Shielded cables that enter a shielded enclosure should have their shields bonded to the enclosure.
- Reflection loss is large for electric fields and plane waves.
- Reflection loss is normally small for low-frequency magnetic fields.
- A shield one-skin depth thick provides approximately 9 dB of absorption loss.
- Reflection loss decreases with frequency; Absorption loss increases with frequency.
- Magnetic fields are harder to shield against than electric fields.
- Use a material with a high relative permeability to shield against low-frequency magnetic fields.
- Use a highly conductive material to shield against electric fields, plane waves, and high-frequency magnetic fields.
- Absorption loss is a function of the square root of the permeability times the conductivity.
- Reflection loss is a function of the conductivity divided by the permeability.
- Increasing the permeability of a shield material increases absorption loss and decreases reflection loss.
- Aperture control is the key to high-frequency shielding.
- The maximum linear dimension, not the area, of an aperture determines the amount of leakage.
- In order to minimize leakage, electrical contact must exist across the seams of shielded enclosures.
- Shielding effectiveness decreases proportional to the square root of the number of apertures.
- For most shield materials the absorption loss predominates above 1 MHz.
- Low-frequency shielding effectiveness depends primarily on the shield material.
- Aperture control is just as important in conductive coated plastic enclosures as in metal shields.
- Shielding not only can be done at the enclosure level, but also at the module level and at the PCB level.
- Shields do not have to be grounded to be effective.

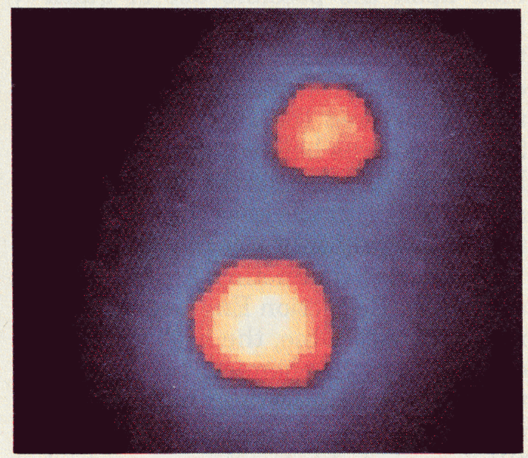
A System Integration “War Story”: Adaptive Optics

Science Newsfront

Edited by DAWN STOVER





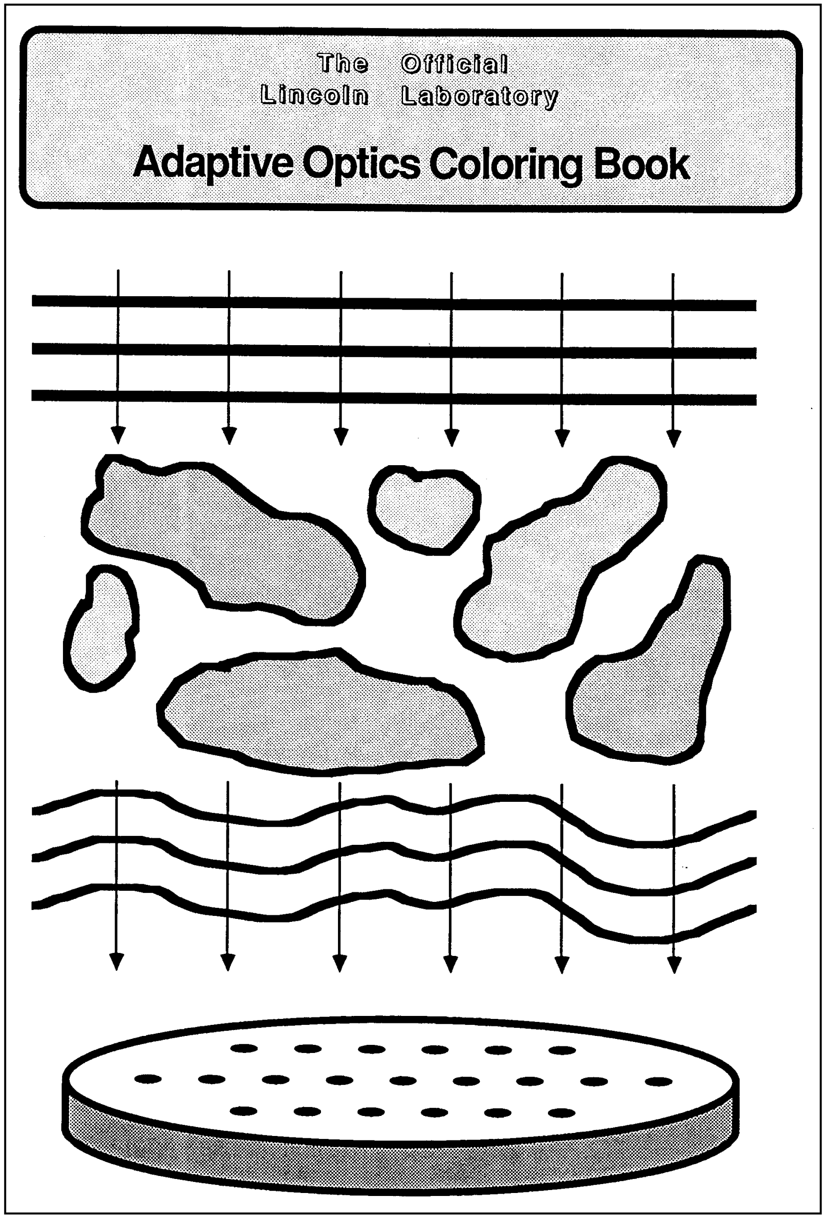


A laser beam fired into space helps astronomers measure optical distortions caused by turbulence in the atmosphere; a deformable mirror then corrects for the distortions. A distant double star that normally looks blurred and very faint (upper right) can be seen clearly when corrected with these adaptive optics (right).

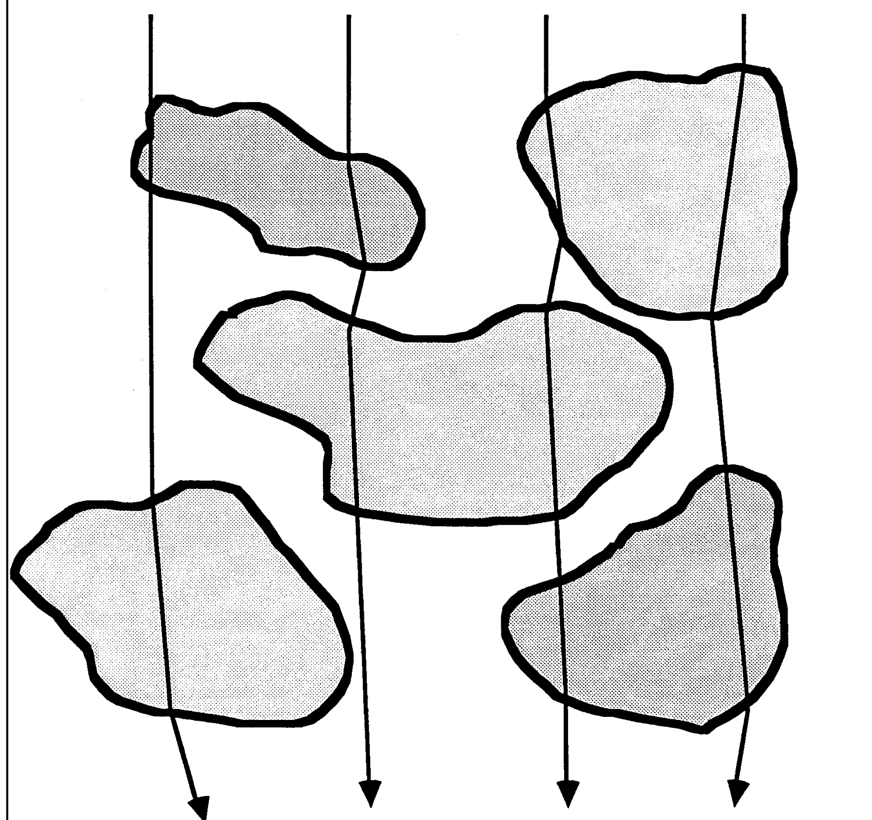
Laser guide star

tions of the laser beam reflect back to earth, providing scientists with infor-

well as a real star, Fugate says. Reflected from a distance of only 6.2

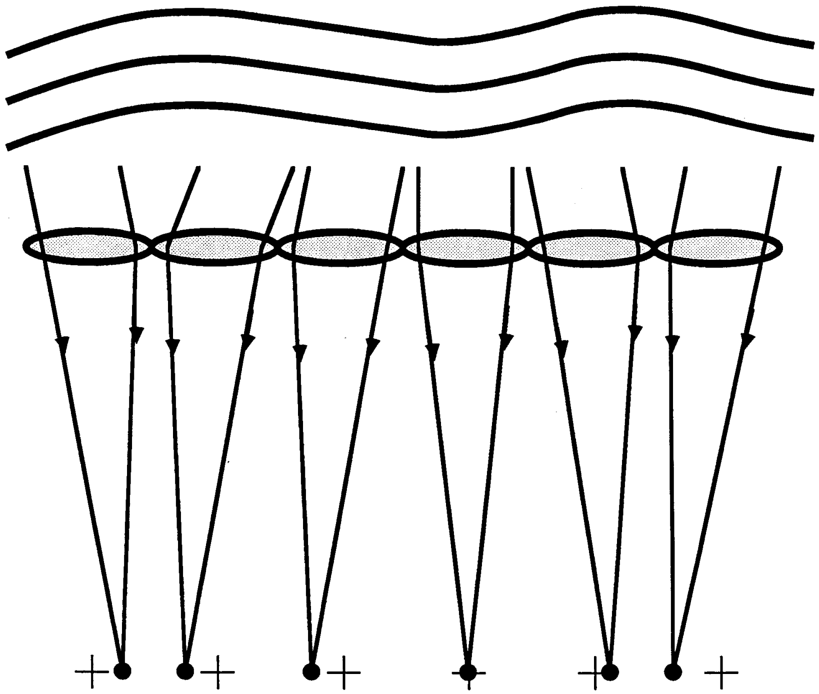


Atmospheric turbulence causes light from distant sources to bend in random ways, so that the images formed of small objects are blurred. This bending is caused by small temperature differences in the atmosphere between the source and the receiver.



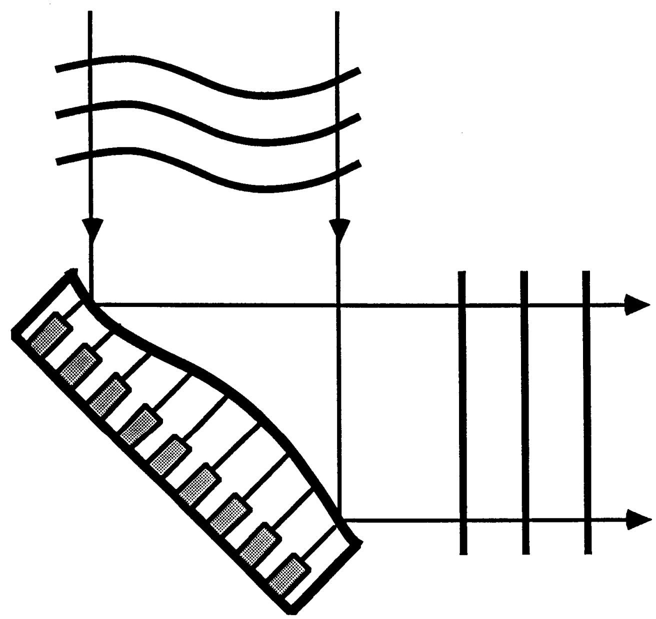
Color the nasty turbulence cells purple.

A device called a WAVEFRONT SENSOR can be used to measure the distortion caused by the turbulence cells. Some wavefront sensors use lots of small lenses to form multiple images of a distant light source.



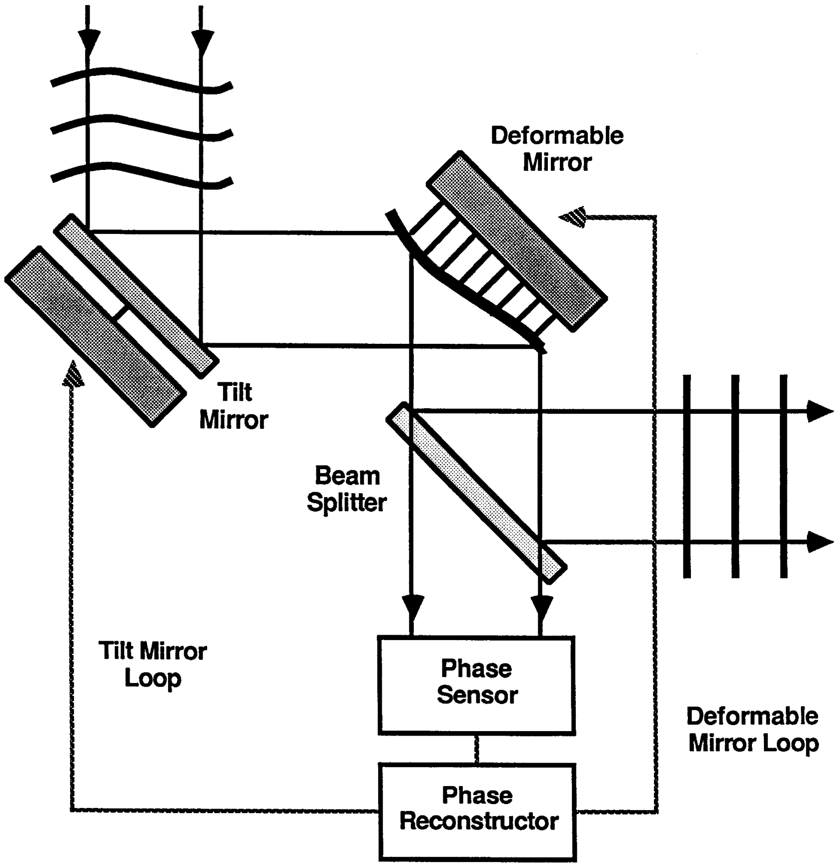
The distance between these spots is directly related to the amount that the input light rays are bent. Color the lenses yellow and the light beams blue.

Once the wavefront distortion has been correctly measured, it can be corrected by bouncing the light off of a device called a DEFORMABLE MIRROR. Small pushing rods attached to the mirror move the surface up and down to remove the wavefront bending caused by turbulence.



The deformable mirror has to move every time the wavefront changes (about 1000 times every second). Color the mirror red before it moves again.

An ADAPTIVE OPTICS SYSTEM combines the wavefront sensor and the deformable mirror with lots of lenses, mirrors, and fancy electronics.



All of these components are very expensive.
Color each of the system components dark green.

High Speed video signal chain:

Analog and Digital Ground: What is the same? What is different?

Same: Ground pin on any IC is a local reference for determining value of a voltage.

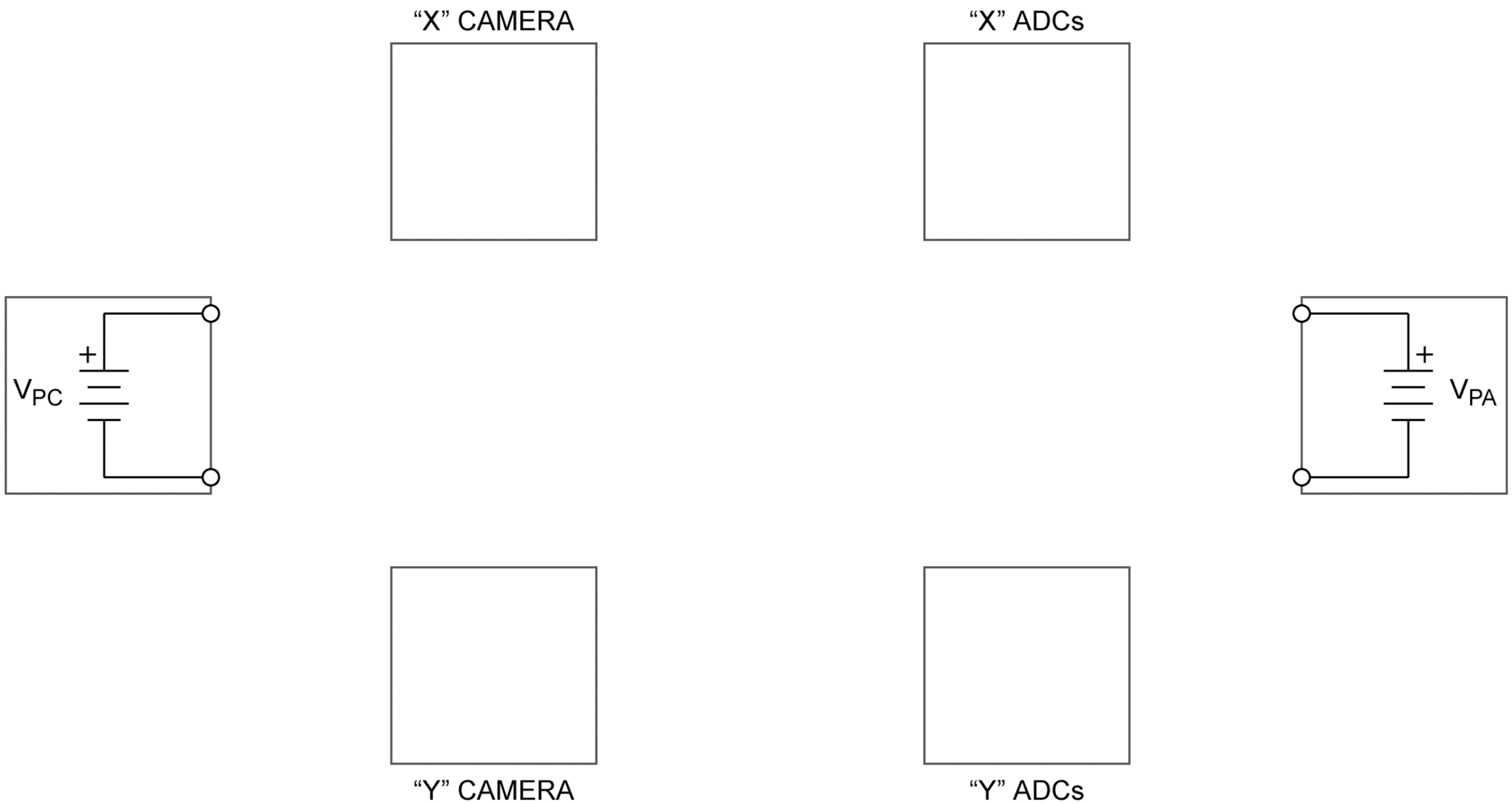
- Input: voltage on an input pin (digital logic or analog) is compared to local ground
- Output: voltage driven is defined relative to local ground
- Exception: True differential pair (e.g. op-amp input, LVDS pair)
- But even then there will be some common mode dependence (quantified by CMRR)

Different: Currents flowing in ground return paths tend to be very different.

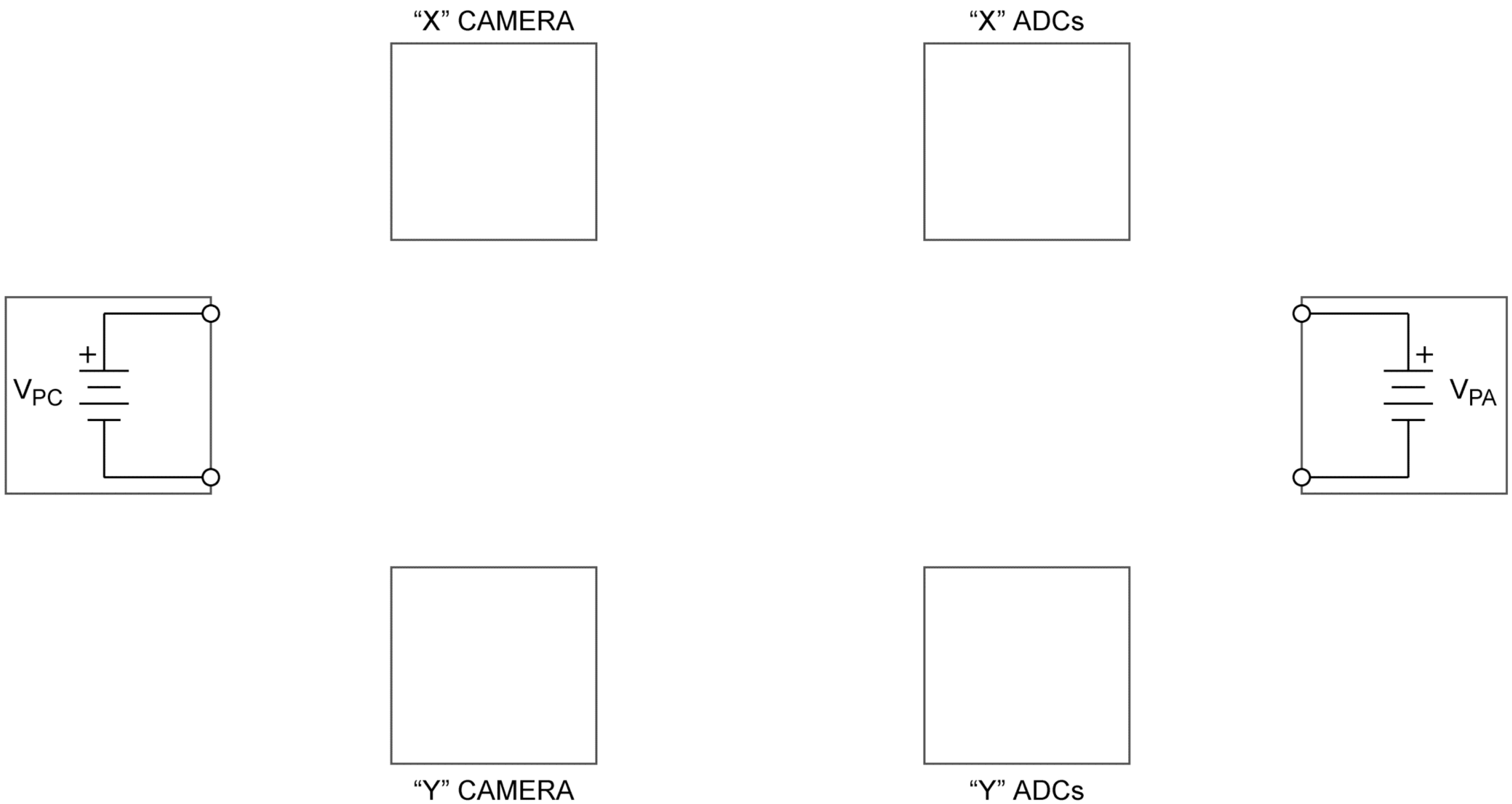
- Digital currents tend to have much higher transient content and depending on system may be much larger in magnitude than analog currents
- Digital inputs are usually much more noise tolerant: depending on logic family, 10s or even 100s of mV of noise will not affect operation
- Analog currents tend to be “smoother” (less transient content) and often are related to signal in a benign way (e.g. larger signal amplitude corresponds to larger current, so any current-induced error as a fraction of signal is not as much of a problem)
- Analog signals are often much more sensitive to voltage errors (mV or even μV of noise can cause performance degradation)
- Impedance levels of analog signals are constrained by factors other than noise performance

High Speed video signal chain: Analog and Digital Ground

System Integration: X and Y cameras, ADCs, power supplies



Solution:



Summary: A System Integration “War Story”: Adaptive Optics

Need to know lots of techniques because there will often be some you want to use but can’t
Example: prefer fully differential twisted shielded pair but needed 75Ω coax
Prefer high input impedance for good CMRR but needed to match cable impedance

Design power and ground distribution at the beginning; explicitly show all cabling
Follow all current through all possible return paths

Design user interface for debugging as well as normal operation
Corollary: if user interface not designed for debugging, need to look at potential problems directly!

HUGE cost of fix in field
Example: Cost of extra power supply vs. weeks of lost customer time, goodwill

RF and Transient Immunity

Radio frequency susceptibility usually involves *audio rectification*: unintentional detection (rectification) of high-frequency rf energy by a non-linear element, in a low-frequency circuit.
Demodulated rf signal produces a dc or low-frequency offset voltage in the affected system
For audio rectification to be a problem the following two things must happen:

- First, the rf energy must be picked up.
- Second, it must be rectified.

Eliminate either of the above, and audio rectification does not occur.
The rf energy is usually picked up by the cables—and in some very high-frequency cases, by the circuit itself. In most cases, the detection occurs in the first p-n junction that the rf energy encounters. In rare cases, the detection can be caused by the rectifying properties of a bad solder joint or a poor ground connection.
The most critical circuits are usually low-level analog circuits such as amplifiers and linear voltage regulator circuits.

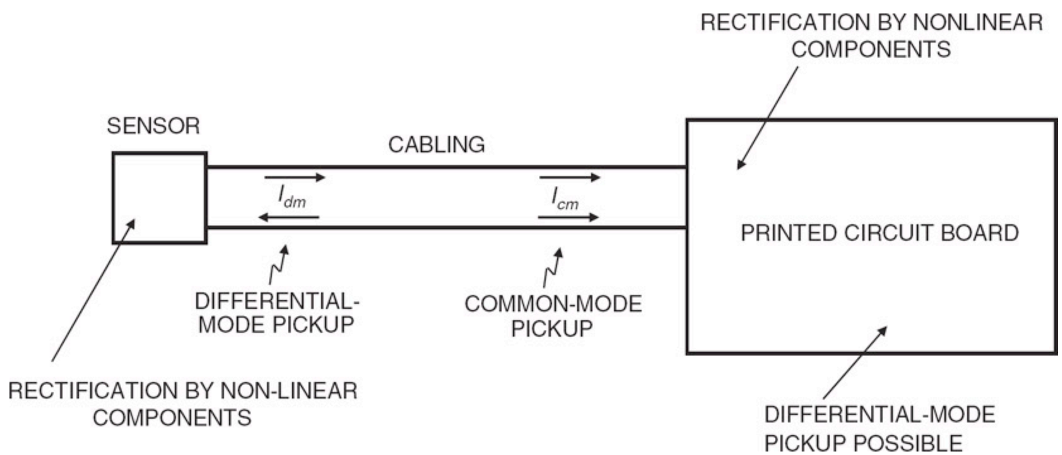


Figure 14-1: Radio frequency immunity example

RFI suppression should start at the device level and can then be supplemented with enclosure and cable-level protection.

The most critical circuits are the ones that operate at the lowest signal levels, and those located closest to the input/output (IO) cables.

Keep all critical signal loop areas as small as possible, especially the input circuit and the feedback circuit of low-level amplifiers.

Sensitive ICs should be protected with rf filters directly at their inputs.

A low-pass R-C filter consisting of a series impedance (ferrite bead, resistor, or inductor) and a shunt capacitor should be used at the input to the sensitive device, as shown in Fig. 14-2, to divert the rf currents away from the device and prevent audio rectification.

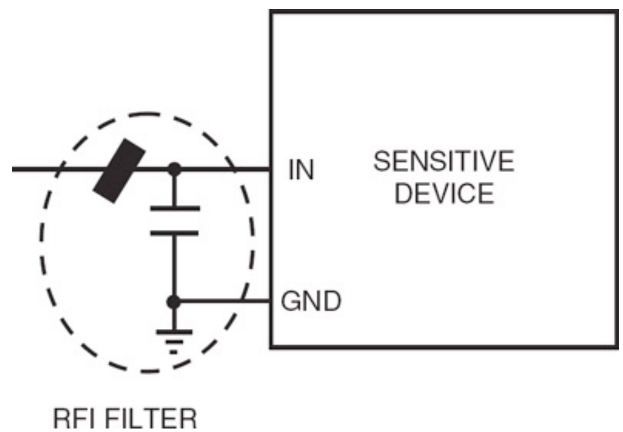


Figure 14-2: RFI filter on the input to a sensitive device

An effective RFI filter can be made using a series element with an impedance of 50 to 100 Ω and a shunt element (usually a capacitor) with an impedance of a few ohms or less, both determined at the frequency of interest.

RFI filter examples

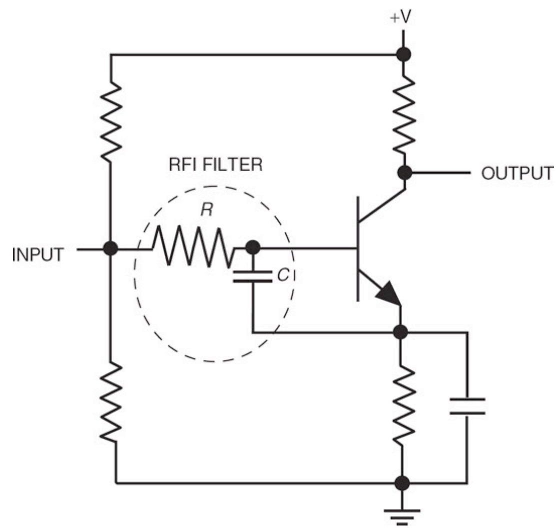


Figure 14-3: RFI filter applied to a transistor amplifier's base-to-emitter junction

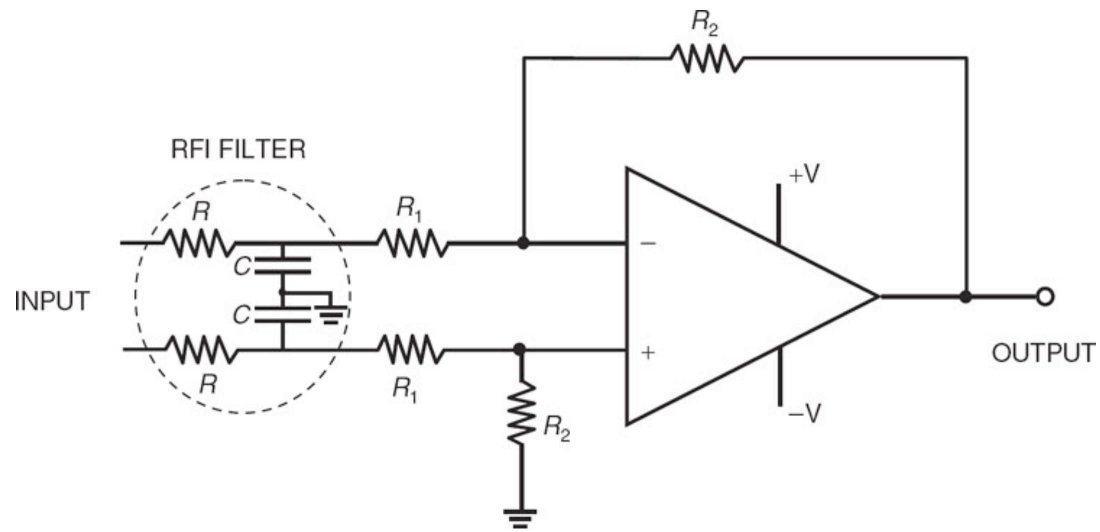


Figure 14-4: RFI filter applied to the input of an operational amplifier

Caution: Check that filter impedance does not compromise desired input impedance in signal-of-interest band

Voltage regulators, which include three terminal regulators, have also demonstrated susceptibility to rf fields.
Add rf filter capacitors (≈1000 pF) directly on the input and output of the regulator.
A small ferrite bead added to the regulator's input and output leads will increase the effectiveness of the filter.
These capacitors are in addition to the larger value capacitors required for proper operation and/or stability of the regulator and should be connected directly to the common pin of the regulator as shown in Fig. 14-7.

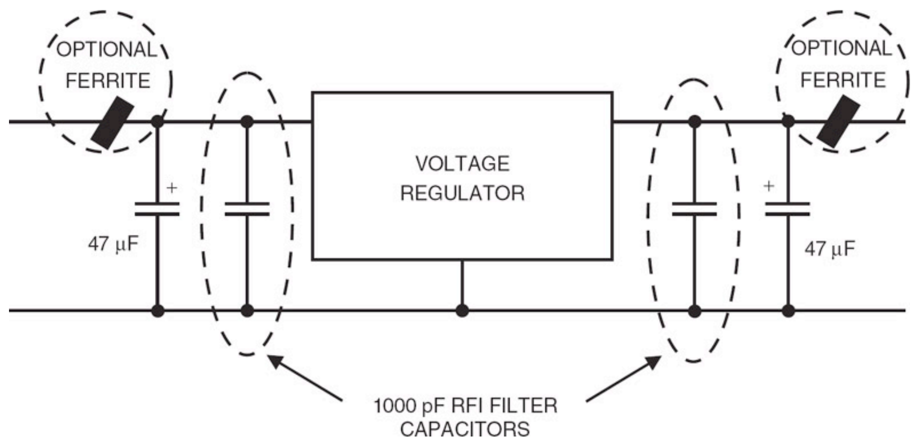


Figure 14-7: Protecting a voltage regulator from rf interference

Cable Suppression Techniques

In most cases, rf energy is picked up by the cables; important to minimize this pickup.
In the case of shielded cables, use a good quality high-coverage braid, multiple braids, or braid over foil with proper termination, no pigtails or drain wires (see Section 2.1.5).
Do not use a spiral-shielded cable for rf immunity.
Although cable connectors with metal backshells that make 360° contact are the ideal shield termination methods, other less-expensive approaches can also be effective.
Figure 14-8 shows one possible way to terminate a shielded cable effectively by using a metal cable clamp.
The cable clamp should be located as close as to where the cable enters the enclosure as possible.

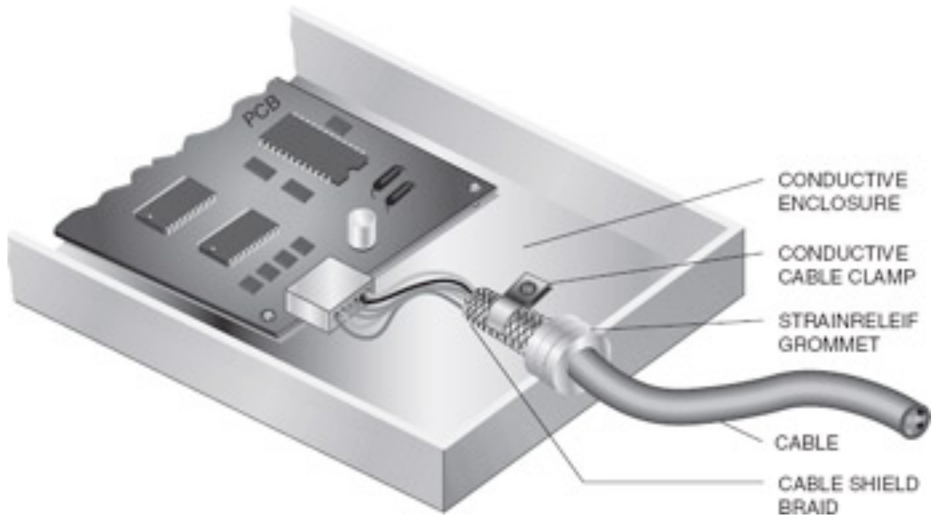


Figure 14-8: A simple method of effectively terminating a shielded cable without a pigtail or coaxial connector

Table 14-3: Characteristics of High-Voltage Transients

Transient	**Voltage**	**Current**	**Rise Time**	**Pulse Width**	**Pulse Energy**
ESD	4-8 kV	1-10s A	1 ns	60 ns	1-10s mJ
EFT (Single Pulse)	0.5-2 kV	10s A	5 ns	50 ns	4mJ
EFT (Burst)	0.5-2 kV	10s A	n/a	15 ms	100s mJ
Surge	0.5-2 kV	100s A	1.25 µs	50 µs	10-80 J

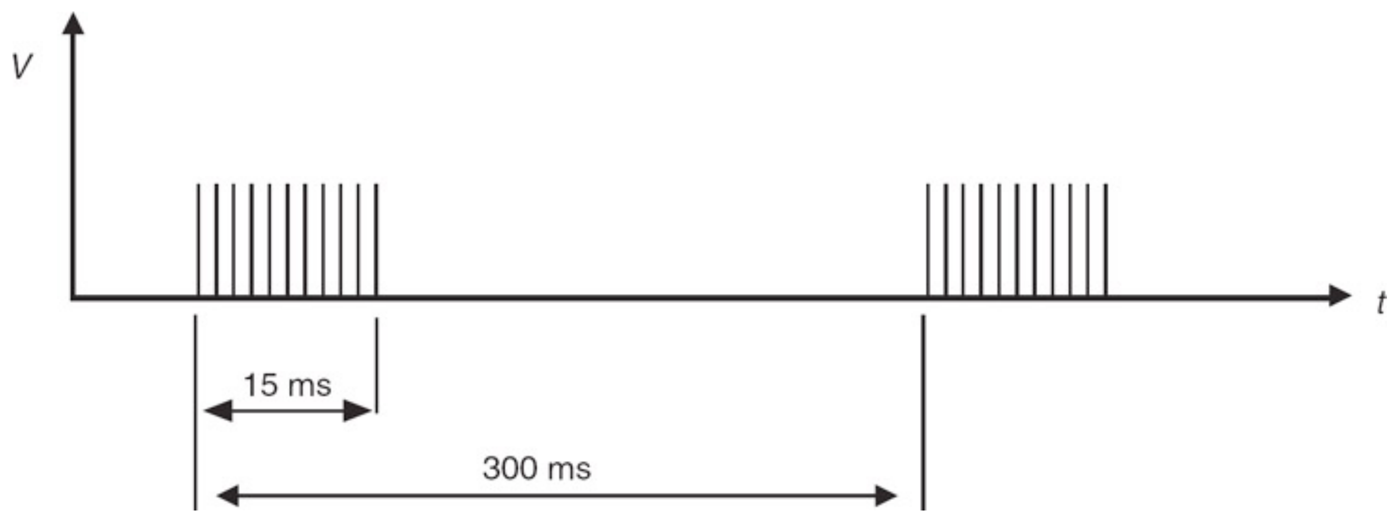


Figure 14-9: EFT test impulse

Some desirable characteristics of transient protection networks are listed below.

- Limit the voltage
- Limit the current
- Divert the current
- Operate fast
- Be capable of handling the energy
- Survive the transient
- Have a negligible affect on the system operation
- Fail safe
- Have minimal cost and size
- Require minimal or no maintenance

In most cases, not all the above objectives can be met simultaneously.

General principle of transient protection (single stage example):

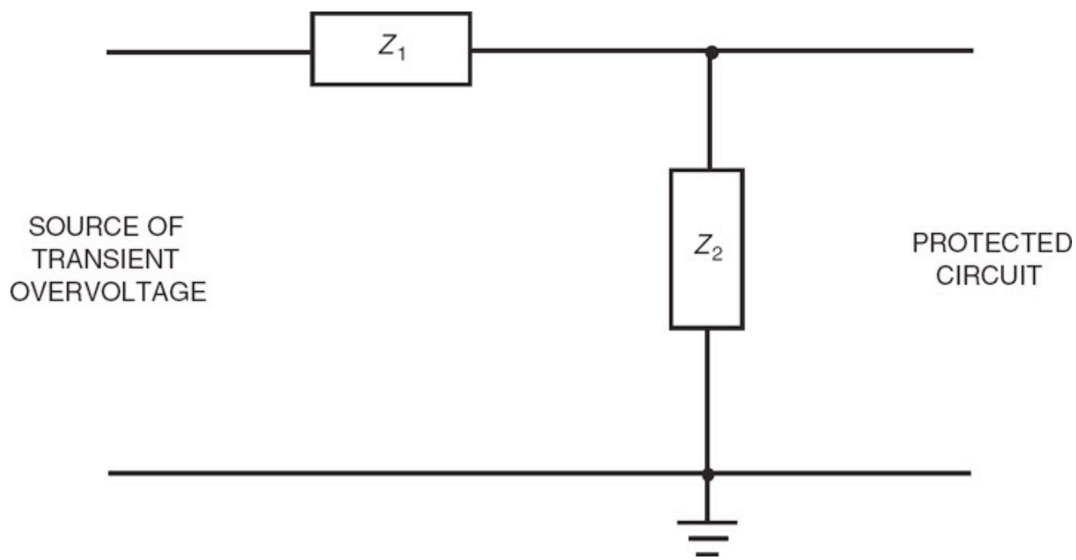


Figure 14-10: Single-stage transient voltage protection network

Signal Line Suppression

I/O signal cable protection can be achieved with the addition of transient voltage suppression (TVS) diodes where the cables enter the product.
TVS diode is similar to a zener diode but with a larger p-n junction area that is proportional to its transient power rating.
Note that increased junction area also increases the capacitance of the diode, which could have a detrimental effect on the normal operation of the circuit.
The three most important parameters of TVS diodes are as follows:

- Reverse standoff voltage
- Clamping voltage
- Peak pulse current

TVS diodes must have a low inductance connection to the chassis to effectively divert the transient energy away from the susceptible circuits.

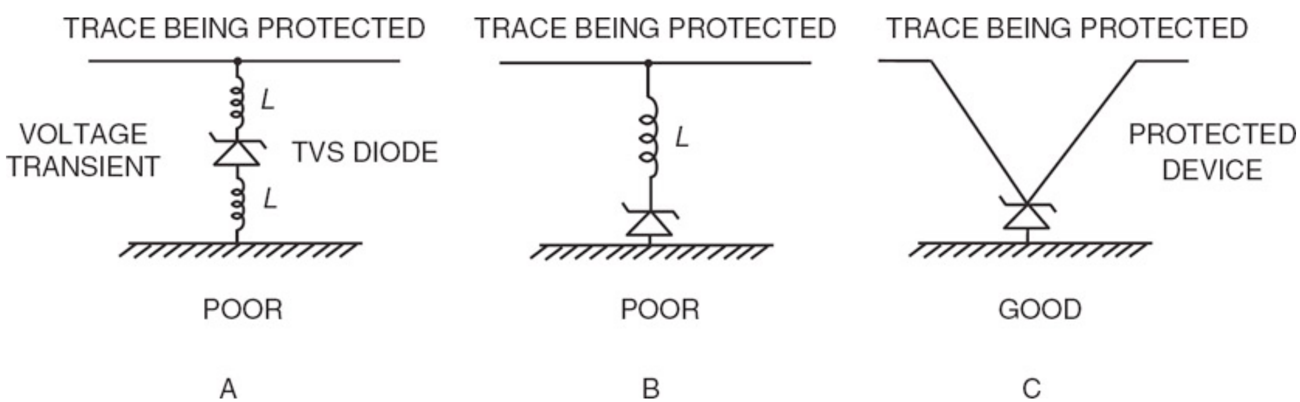


Figure 14-11: Mounting configurations for TVS diodes. (C) shows proper mounting

Possible to combine transient current entry protection and sensitive device hardening.
Notice that the transient suppressor on the input cable is grounded to the chassis, because its purpose is to divert the transient current away from the PCB.
The protective filter on the sensitive device's input, however, is grounded to the circuit ground, because its purpose is to minimize or eliminate any transient voltage from appearing between the device's input pin and ground pin.

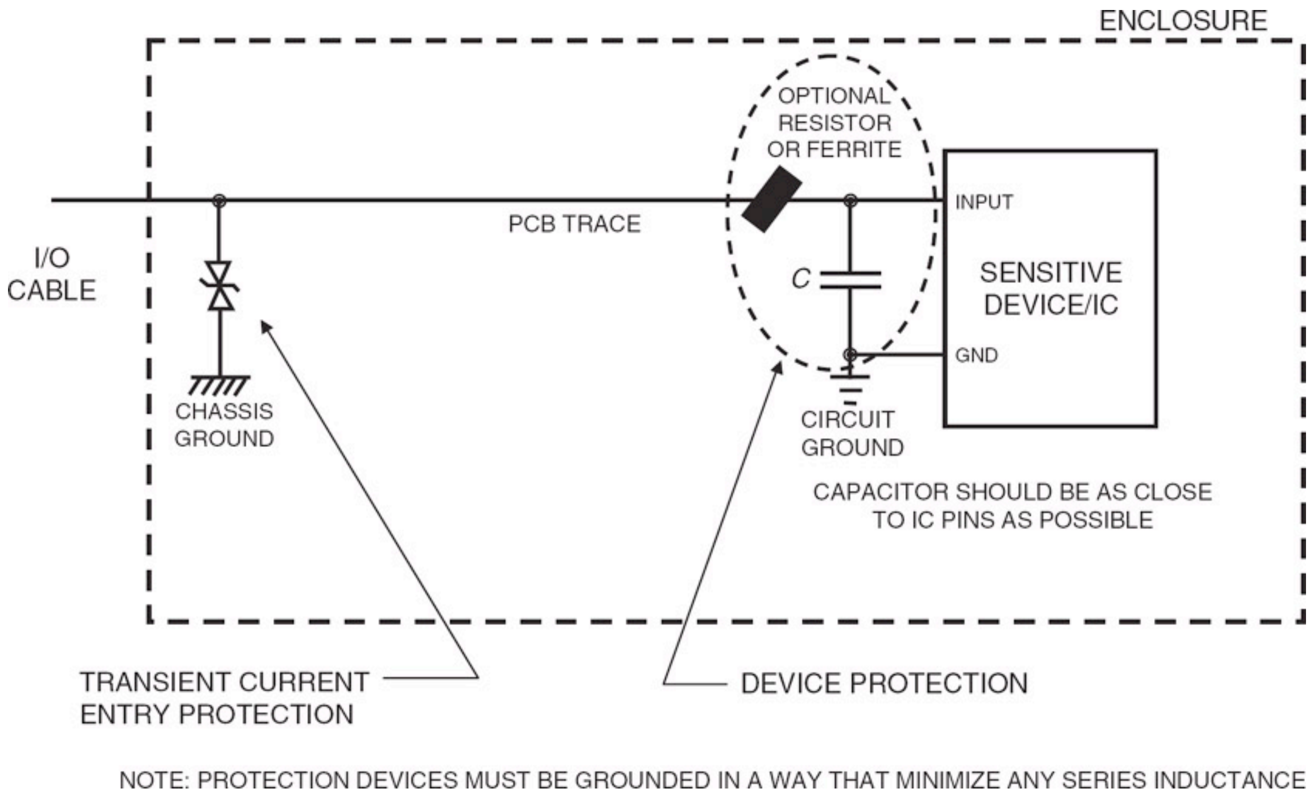


Figure 14-13: Transient protection device grounding

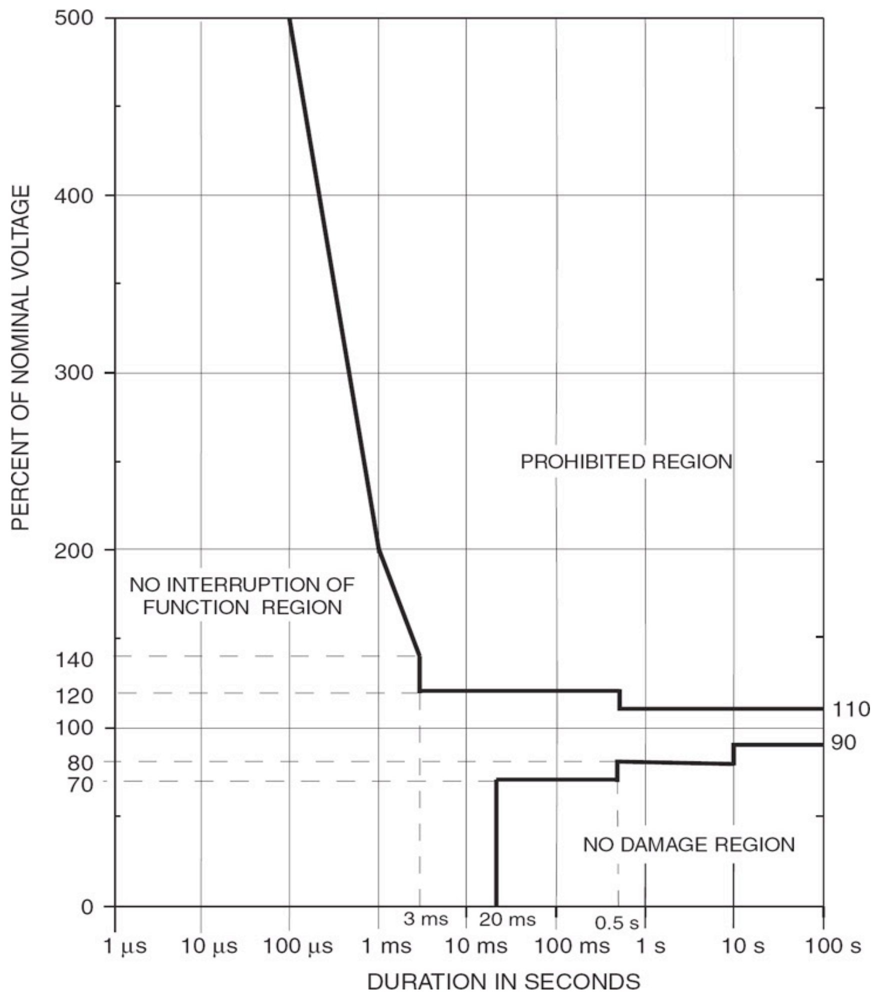


Figure 14-21: The CBEMA curve (revised in 2000) defines the power line voltage tolerance envelope applicable to single-phase, 120-V equipment. *(© Information Technology Industry Council, 2000)*

MIL Standards (Source: everyspec.com)

MIL-STD-1275D
29 August 2006
SUPERSEDING
MIL-STD-1275C
23 June 2006

DEPARTMENT OF DEFENSE
INTERFACE STANDARD

CHARACTERISTICS OF
28 VOLT DC
ELECTRICAL SYSTEMS IN
MILITARY VEHICLES

MIL-STD-704F
12 MARCH 2004
SUPERSEDING
MIL-STD-704E
1 MAY 1991

DEPARTMENT OF DEFENSE
INTERFACE STANDARD

AIRCRAFT ELECTRIC POWER
CHARACTERISTICS

TABLE III. AC normal operation characteristics - 60 Hertz (see 5.2.3).

Steady state characteristics	Limits
Steady state voltage	105.0 to 125.0 Volts, RMS
Voltage unbalance	3.0 Volts, RMS maximum
Voltage modulation	2.5 Volts, RMS maximum
Voltage phase difference	116° to 124°
Distortion factor	0.05 maximum
Distortion spectrum	Figure 12
Crest factor	1.31 to 1.51
DC component	+0.10 to – 0.10 Volts
Steady state frequency	59.5 to 60.5 Hz
Frequency modulation	0.5 Hz
Transient characteristics	**Limits**
Peak voltage	±271.8 Volts
Voltage transient	Figure 8
Frequency transient	Figure 10

MIL-STD-704: 60 Hz

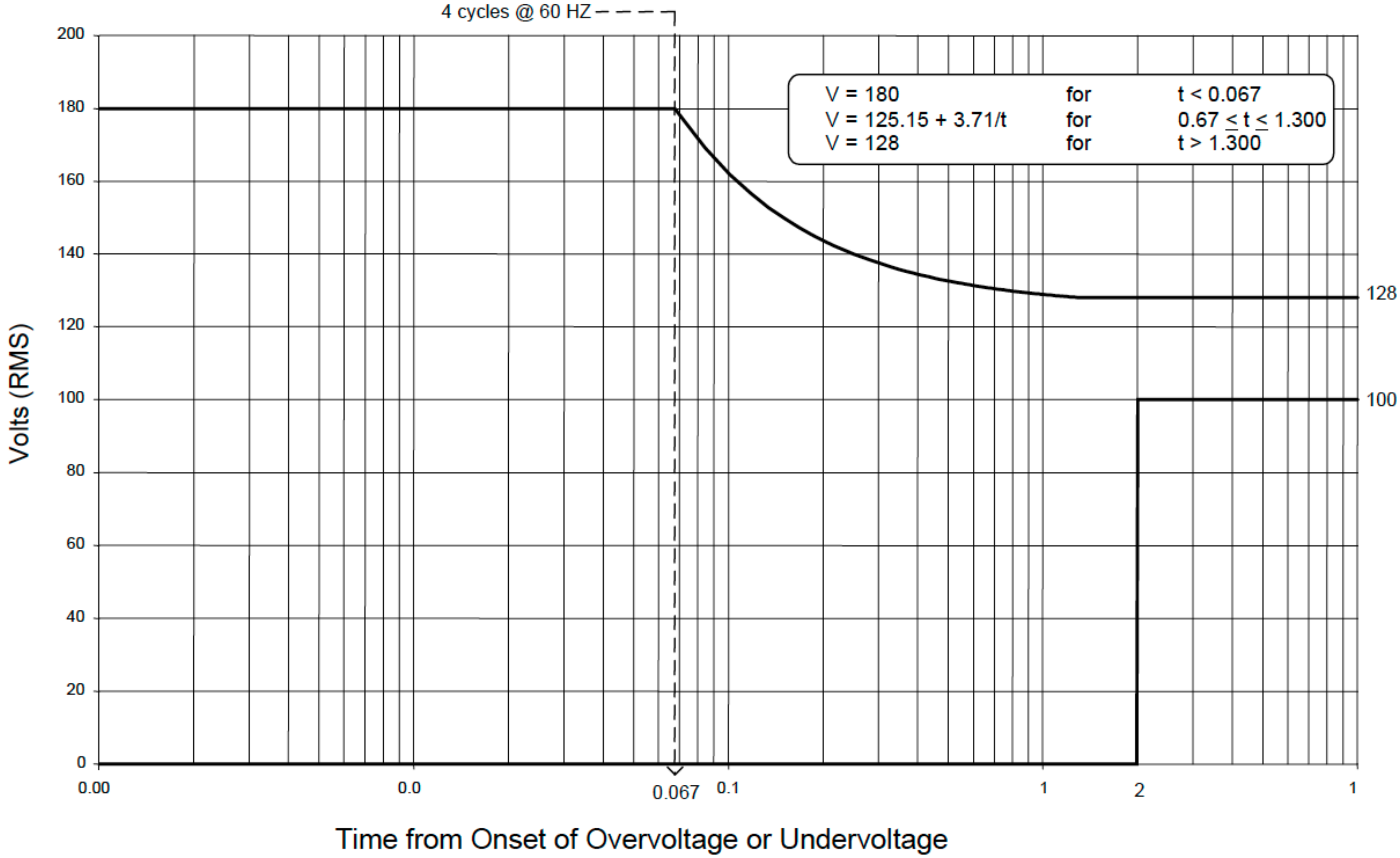


TABLE IV. DC normal operation characteristics (see 5.3.2.1, 5.3.4.1).

Steady state characteristics	Limits	
	28 Volt DC system	270 Volt DC system
Steady state voltage Distortion factor Distortion spectrum Ripple amplitude	22.0 to 29.0 Volts 0.035 maximum Figure 15 1.5 Volts maximum	250.0 to 280.0 Volts 0.015 maximum Figure 18 6.0 Volts maximum
	Limits	
Transient characteristics	28 Volts DC system	270 Volts DC system
Voltage transient	Figure 13	Figure 16

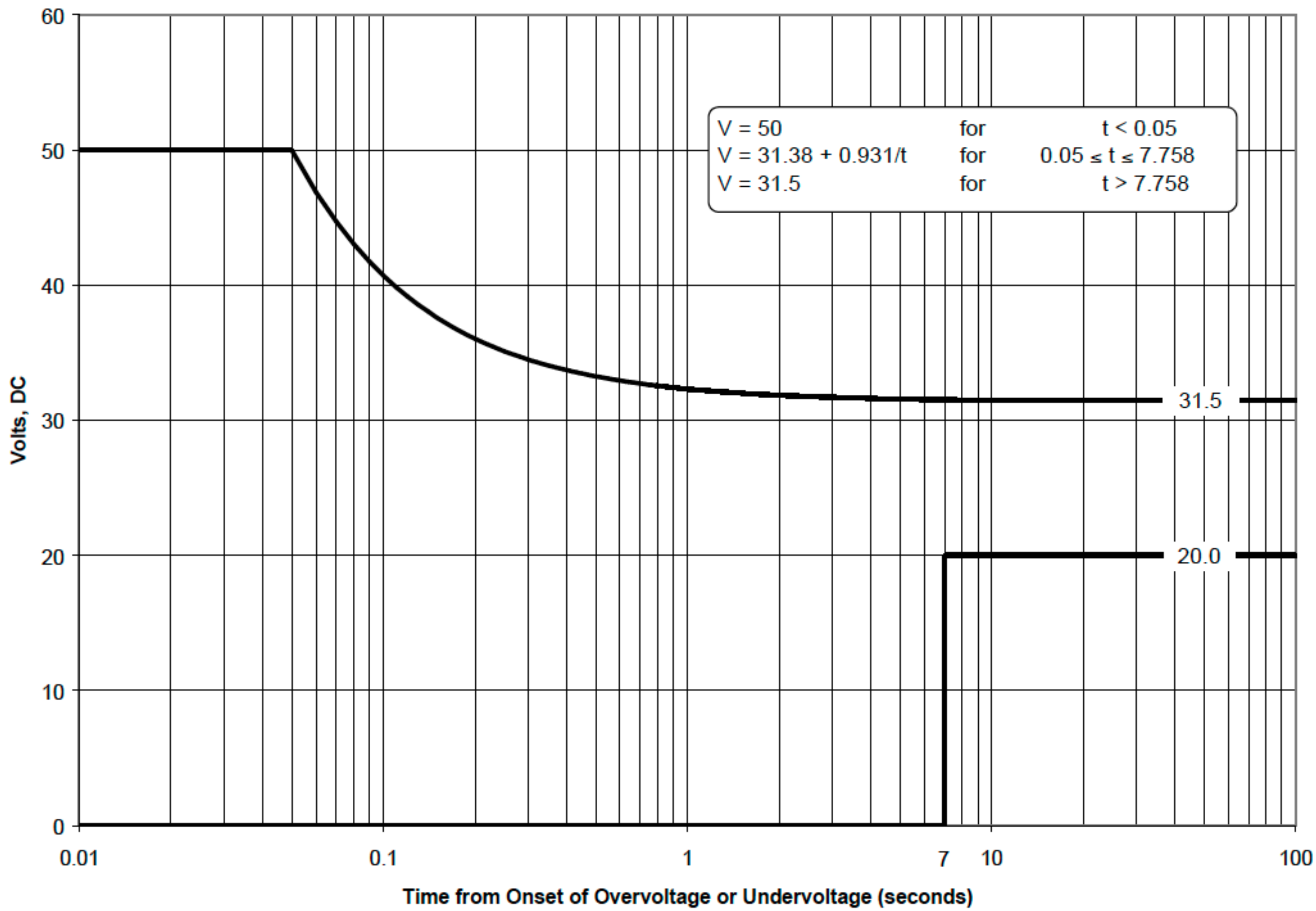


FIGURE 14. Limits for overvoltage and undervoltage for 28 volts DC system.

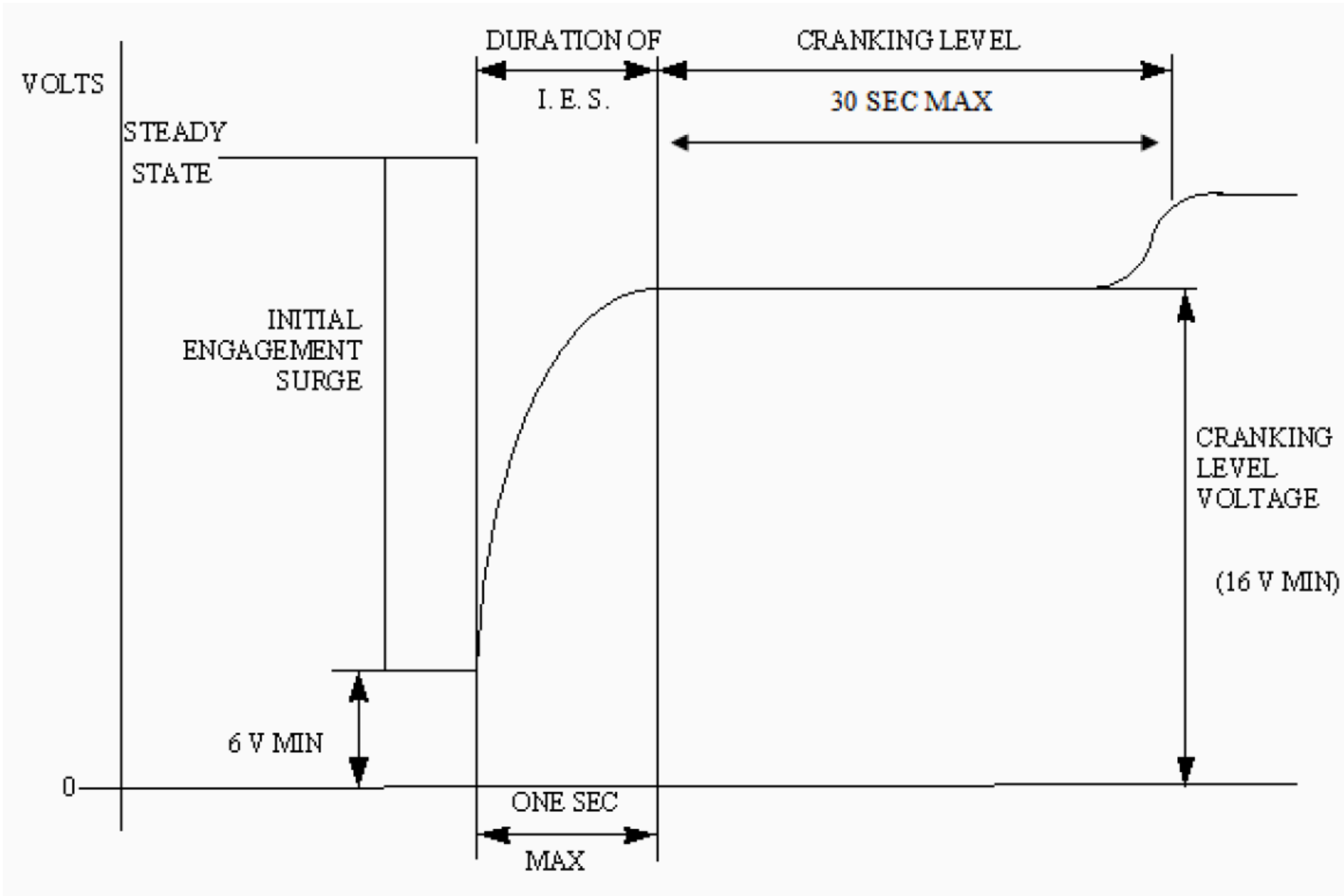


FIGURE 4. Starting Disturbances.

5.1.4.4 Spikes. All spikes resulting from system operation shall fall within the envelope shown in Figure 8.

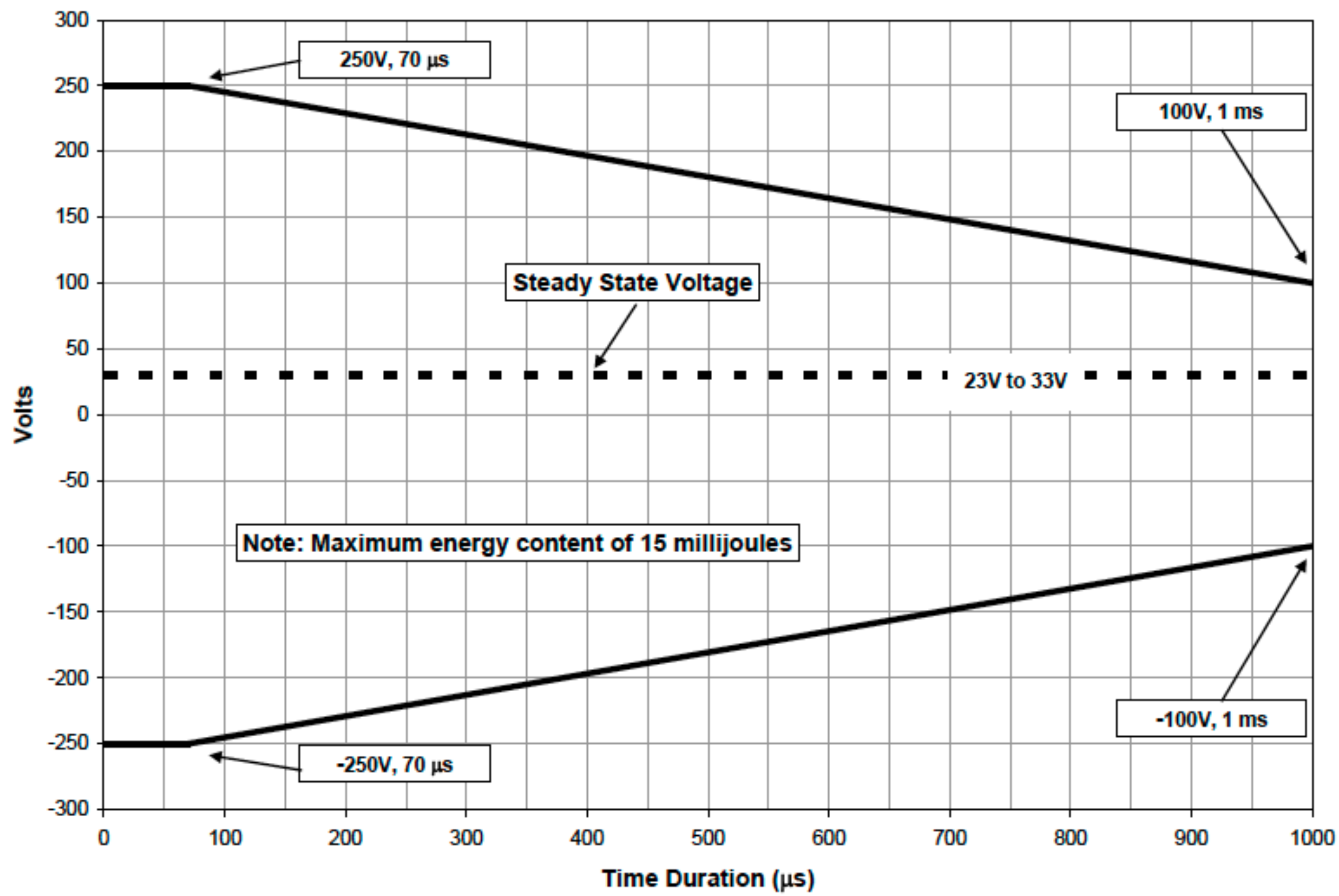


FIGURE 8. Envelope of Spikes in Generator-only Mode for 28 VDC Systems.

Summary: RF and Transient Immunity

For audio rectification to be a problem, the following two things must occur:

- First, the rf energy must be picked up.
- Second, the rf energy must be rectified.

RFI mitigation techniques should be applied at the device level, to the cables, and to the enclosure.
Keep loop areas of sensitive signals as small as possible.
Add RFI filters on inputs to sensitive devices.
A 1000-pF capacitor is effective over the frequency range of 80 MHz to 1000 MHz.
Use rf decoupling, even if the circuit operates at low frequency.
Use multilayer PCBs with ground and power planes.
Use balanced circuitry whenever possible, especially on the low-level inputs to amplifiers.
Add high-frequency capacitive filters on the inputs and outputs of voltage regulators.
Use high-quality shielded cables, high-coverage braid, or braid-over-foil shields.
Properly terminate shield to enclosure, no pigtails or drain wires.
For ribbon cables, limit the signal to ground conductor ratio to 3 to 1.
The maximum linear dimension of apertures in shielded enclosures should be less than 1/20 wavelength at the highest frequency of concern.
Even thin shields made of aluminum or copper are effective at frequencies above 500 kHz.
Steel shields should be used at frequencies below 500 kHz.
The three most common high voltage transients effecting equipment are as follows:

- Electrostatic discharge
- Electrical fast transient
- Lightning surge

Power-line disturbances, such as voltage dips, interruptions, and surges, are a fact of life, and should be protected against.
The CBEMA curve (Fig. 14-21) or applicable MIL-STD can be used as a power-line voltage tolerance immunity guideline for equipment.

Digital Circuit Radiation (Chapter 12)

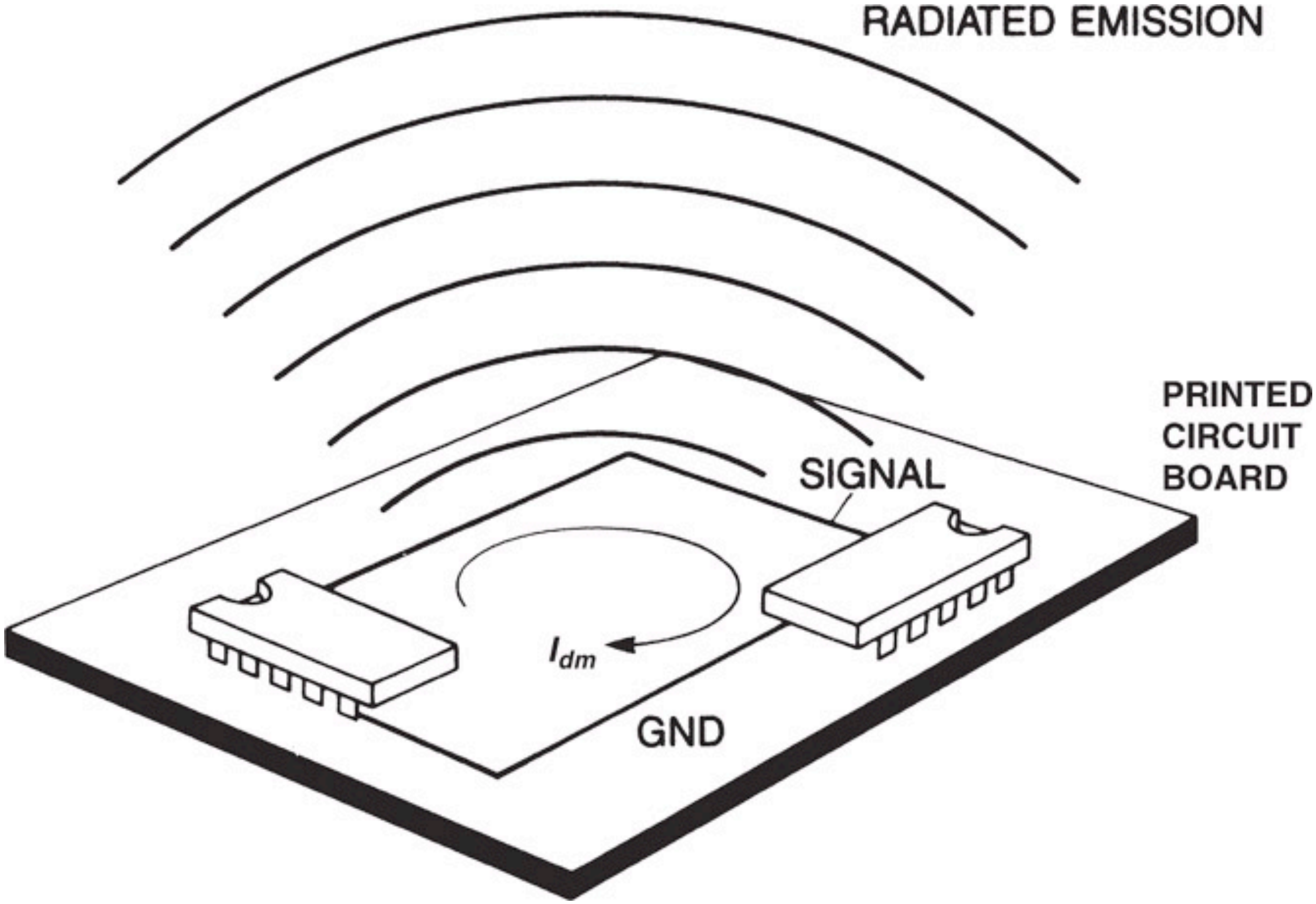


Figure 12-1: Differential-mode radiation from printed circuit board (PCB)

Differential-mode radiation is the result of the normal operation of the circuit and results from current flowing around loops formed by the conductors of the circuit

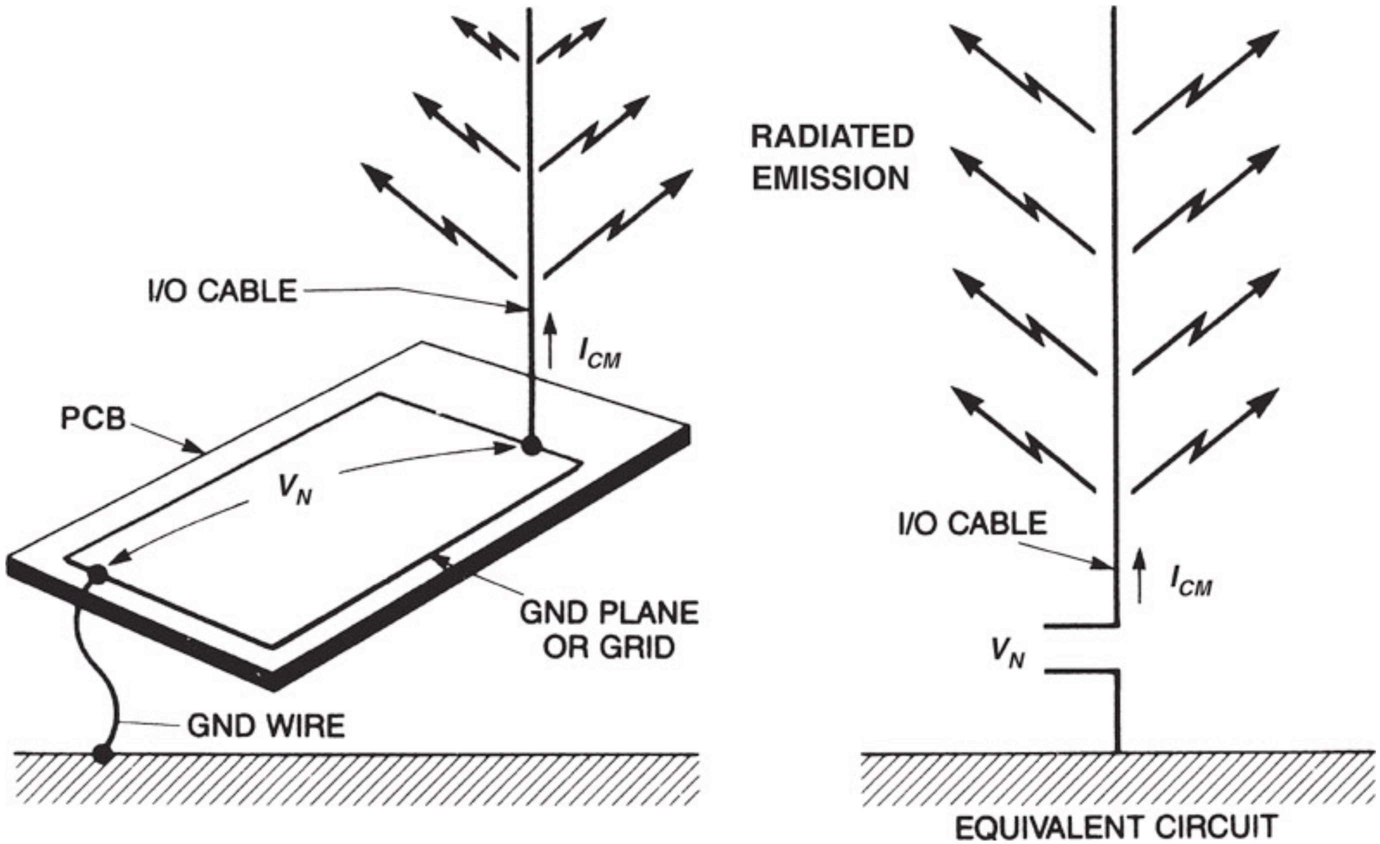


Figure 12-2: Common-mode radiation from system cables

- Common-mode radiation is the result of *parasitics in the circuit* and results from undesired voltage drops in the conductors.
- The differential-mode current that flows through the ground impedance produces a voltage drop in the digital logic ground system.
- When cables are then connected to the system, they are driven by this common-mode ground potential, forming antennas, which radiate predominately electric fields

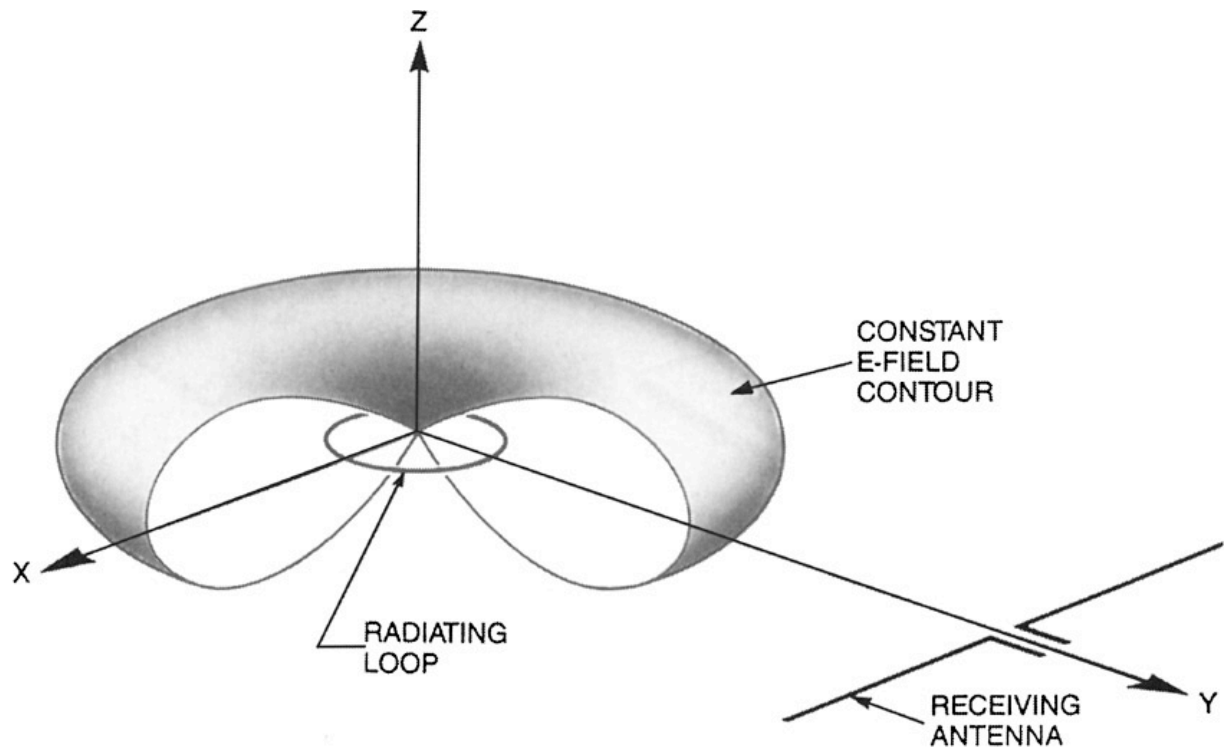


Figure 12-3: Free-space radiation pattern for a small loop antenna

$$E = 263 \times 10^{-16} (f^2 A I_{dm}) \left(\frac{1}{r}\right). \tag{12-2}$$

Differential-mode (loop) radiation can be controlled by

1. Reducing the magnitude of the current
2. Reducing the frequency or harmonic content of the current
3. Reducing the loop area

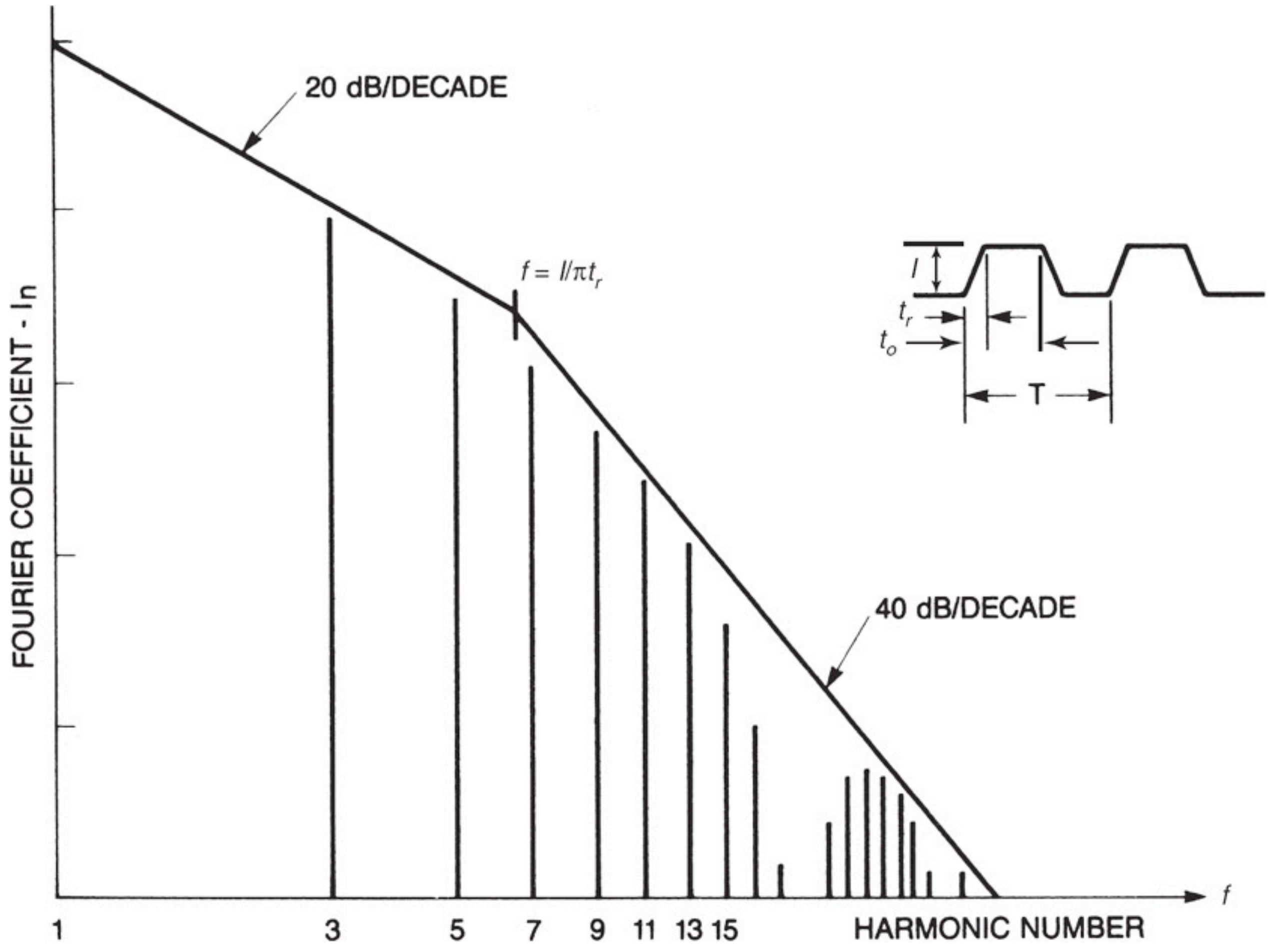


Figure 12-4: Envelope of Fourier spectrum of a 50% duty cycle trapezoidal wave

The frequency-squared term in Eq. 12-2 represents an increase in emission with frequency of 40 dB per decade. The result of combining Eq. 12-2 and the envelope of the Fourier spectrum is that the radiated emission increases 20 dB per decade for frequencies less than $1/\pi\ t_r$, and it remains constant above this frequency as shown in Fig. 12-5.

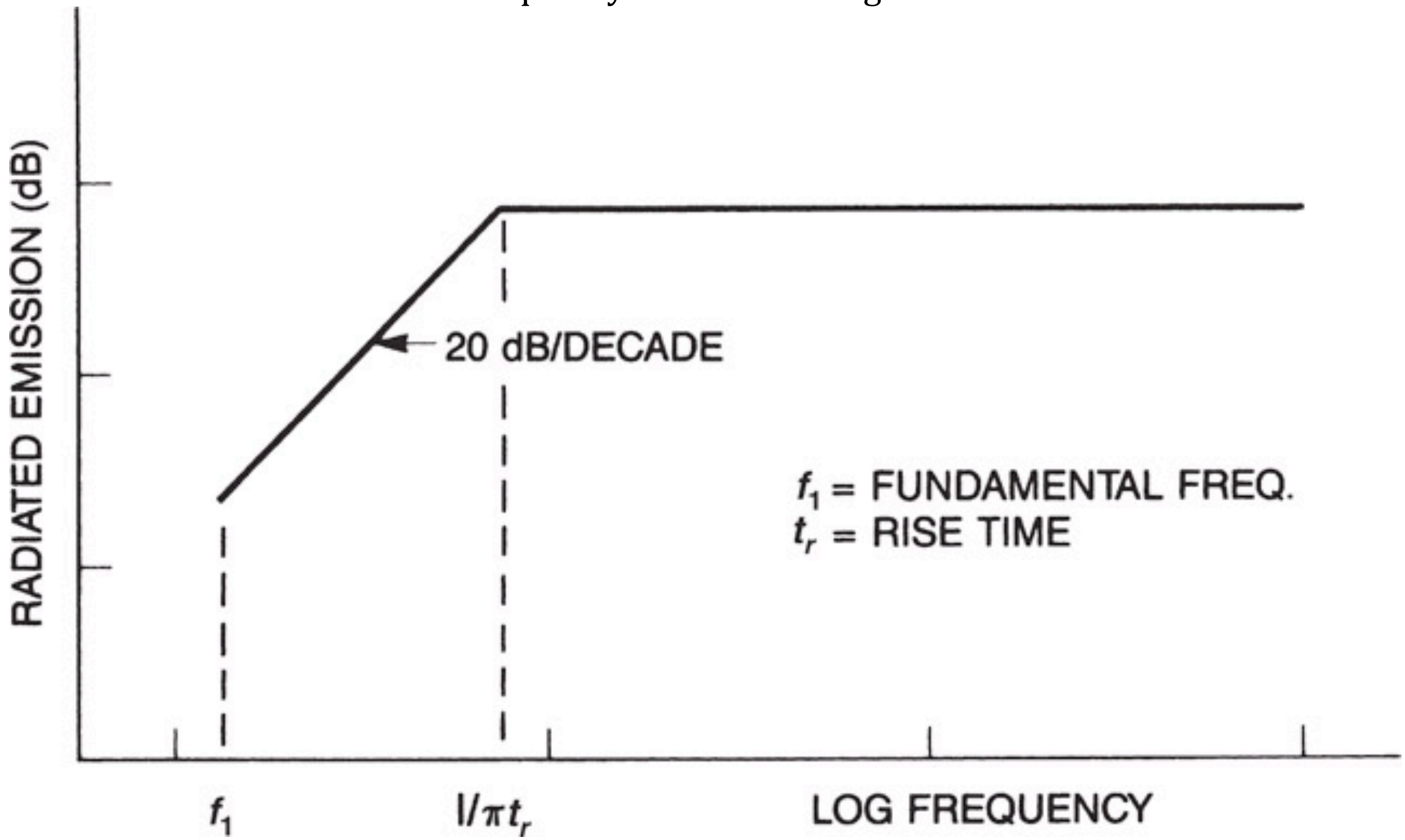


Figure 12-5: Differential-mode radiated emission envelope versus frequency

As an example, the radiated emission measured at 3 m from a 6-MHz clock with a 4-ns rise time flowing around a loop with an area of 10 cm2 (1.5 in2) is shown in Fig. 12-6. The loop is being driven by a square wave of current with a peak amplitude of 35 mA.

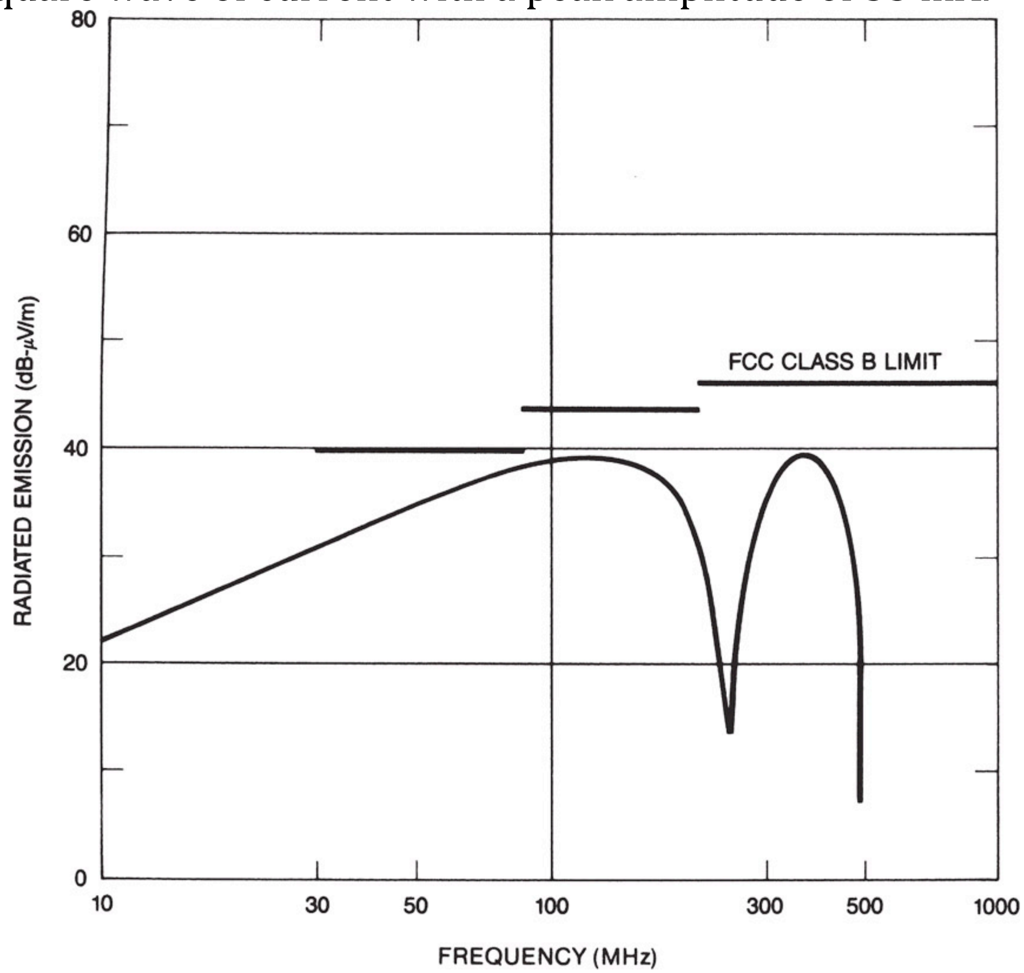


Figure 12-6: Radiated emission spectrum and FCC Class B limit. The spectrum is for a 6-MHz, 4-ns rise time, 35-mA clock signal in a 10-cm^2 loop

Controlling Differential-Mode Radiation

Figure 12-7A shows the radiated emission spectrum from a typical digital circuit.
Figure 12-7B shows the emission from the same circuit with only the clock signal operating.
The maximum emission is about the same in both cases, yet 95% or more of the circuitry was turned off for the case of Fig. 12-7B.
In almost all cases, the emission from the clock harmonics exceeds the emission from all the other circuits.

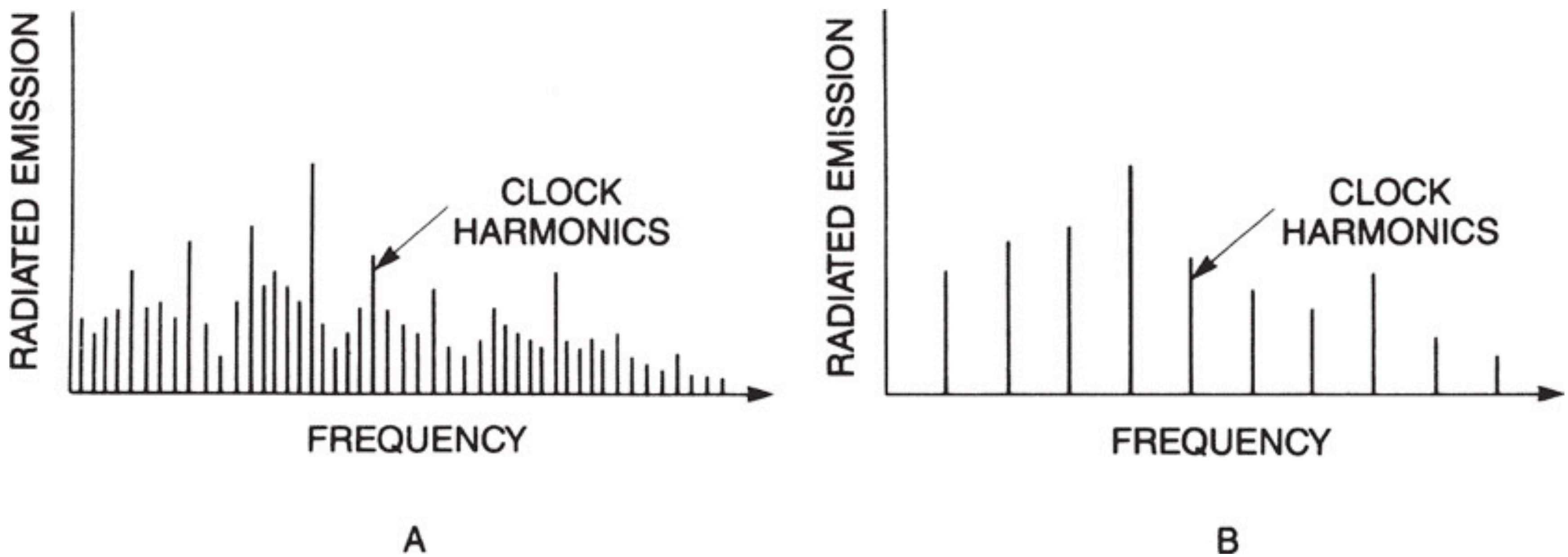


Figure 12-7: Typical radiated emission spectrum from a digital circuit: *(A)* with all circuits operational; *(B)* with only clock circuit operational

Controlling CM radiation

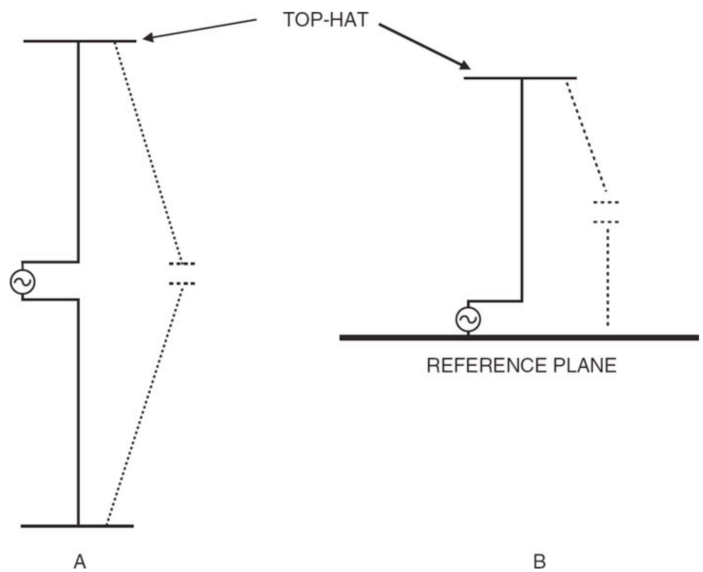


Figure 12-11: Capacitor loaded dipole(A) and monopole(B)

When the observation is made at a distance r perpendicular to the antenna axis

$$E = \frac{12.6 \times 10^{-7}(flI_{cm})}{r}. \qquad (12\text{-}7)$$

Equation 12-7 shows that the radiation is proportional to the frequency, the length of the antenna, and the magnitude of the common-mode current on the antenna.

Common-mode (dipole) radiation can be controlled by the following methods:

1. Reducing the magnitude of the common-mode current (none of which is required for the normal operation of the circuit)
2. Reducing the frequency or harmonic content of the current
3. Reducing the antenna (cable) length

The common-mode radiation mechanism is much more efficient than the differential-mode. A common-mode current of a few µA can cause the same amount of radiation as a few mA of differential-mode current.

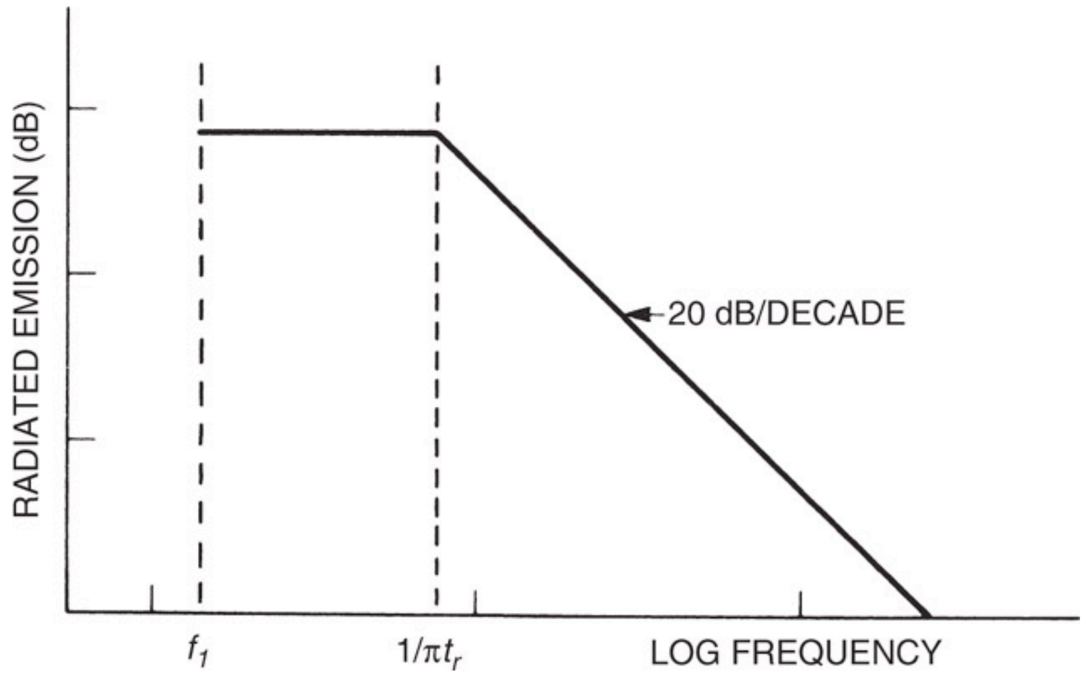


Figure 12-12: Common-mode radiated emission envelope versus frequency

Table 12-1: Maximum Allowable Common-Mode Current in a 1-m-Long Cable at 50 MHz

Regulation	Limit	Distance	Maximum CM Current
FCC Class A	90 µV/m	10 m	15 µA
FCC Class B	100 µV/m	3 m	5 µA
MIL-STD 461	16 µV/m	1 m	0.25 µA

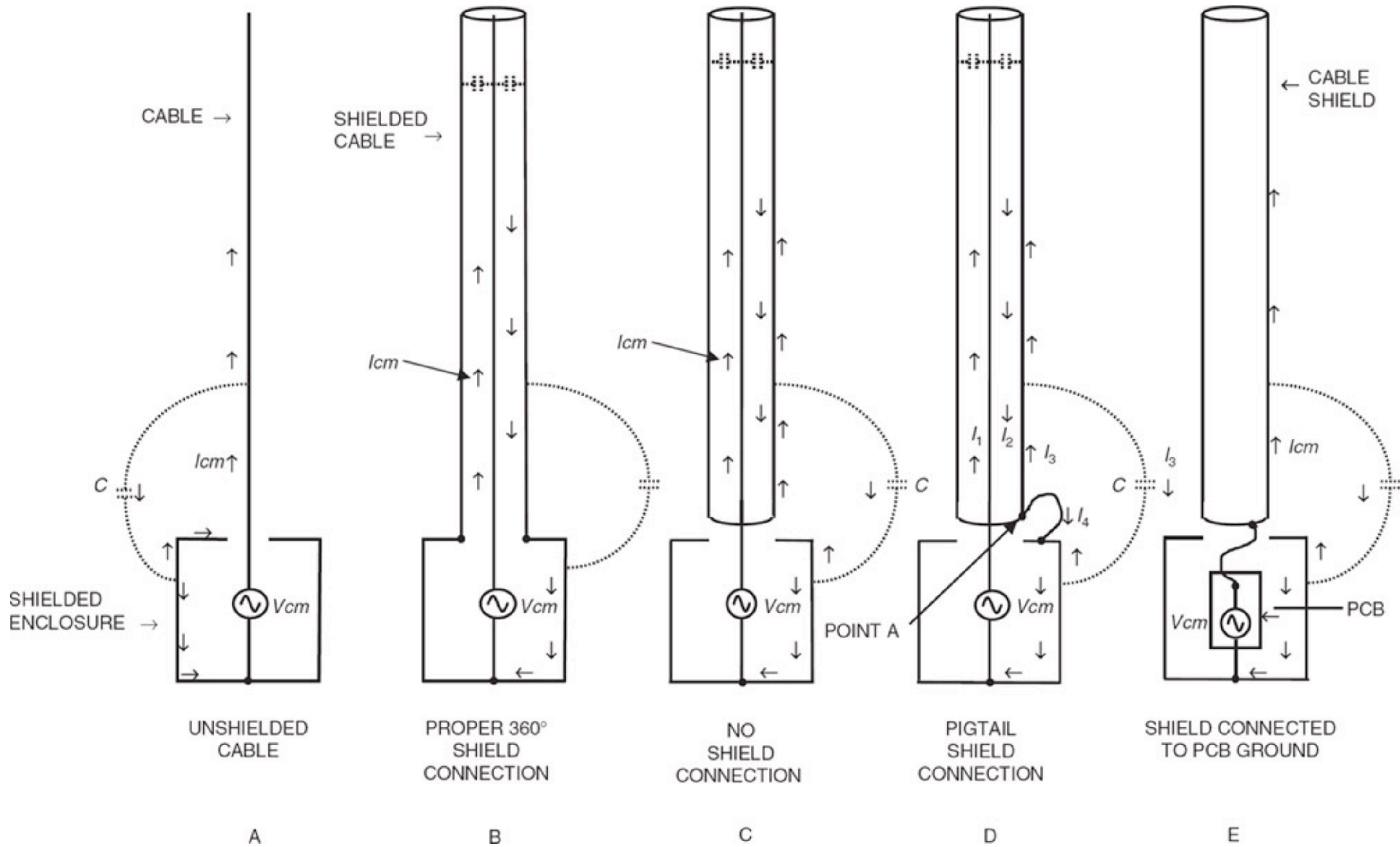


Figure 12-13: The effect of cable shield termination on the common-mode cable current, and hence on the radiated emission from the cable

Summary: Digital Circuit Radiation

- Emission control should be considered during the initial design and layout of a product.
- The differential-mode radiated emission is proportional to frequency squared, loop area, and differential-mode current in the loop.
- The primary way to control the differential-mode radiation is to reduce the loop area.
- PCB technology (the ability to print smaller loops) is not keeping up with the increased radiation that results from the frequency-squared term in the differential-mode radiation equation.
- When loops cannot be made small enough, other unconventional techniques such as dithered clocks and canceling loops may be necessary.
- The most critical signals are those:
    - That have the highest frequency
    - That are periodic
- CM radiated emission is proportional to frequency, cable length, and CM current in the cable.
- The harmonic content of a square wave is determined by rise time, not fundamental frequency.
- The primary way to control CM radiation is by reducing, or eliminating, the CM current on the cables.
- Only a few μA, or less, of CM current is required on a cable to fail radiate emission requirements.
- Common-mode radiation can be reduced by
    - Reducing the ground noise voltage
    - Filtering the I/O cables
    - Shielding the I/O cables
- Both CM and DM radiation can be decreased by reducing the frequency and/or slowing down the rise time of the signals.
- I/O connector backshells and cable filter capacitors must be connected to the enclosure ground, not the circuit ground.
- Cable shields should make 360° connections to the enclosure.
- Most common-mode radiated emission problems occur below 300 MHz, and most differential-mode radiated emission problems occur above 300 MHz.

Conducted Emissions (chapter 13)

Control the radiation from the public alternating current (ac) power distribution system, which results from noise currents conducted back onto the power line. Normally, these currents are too small to cause interference directly with other products connected to the same power line; however, they are large enough to cause the power line to radiate and possibly become a source of interference.
The conducted emission limits exist below 30 MHz, where most products themselves are not large enough to be very efficient radiators, but where the ac power distribution system can be an efficient antenna.
Conducted emission requirements, therefore, are really radiated emission requirements in disguise.
The FCC and the European Union require that both quasipeak (narrowband) and average (broadband) conducted emission measurements be made. The respective limits are shown in Fig. 18-12.

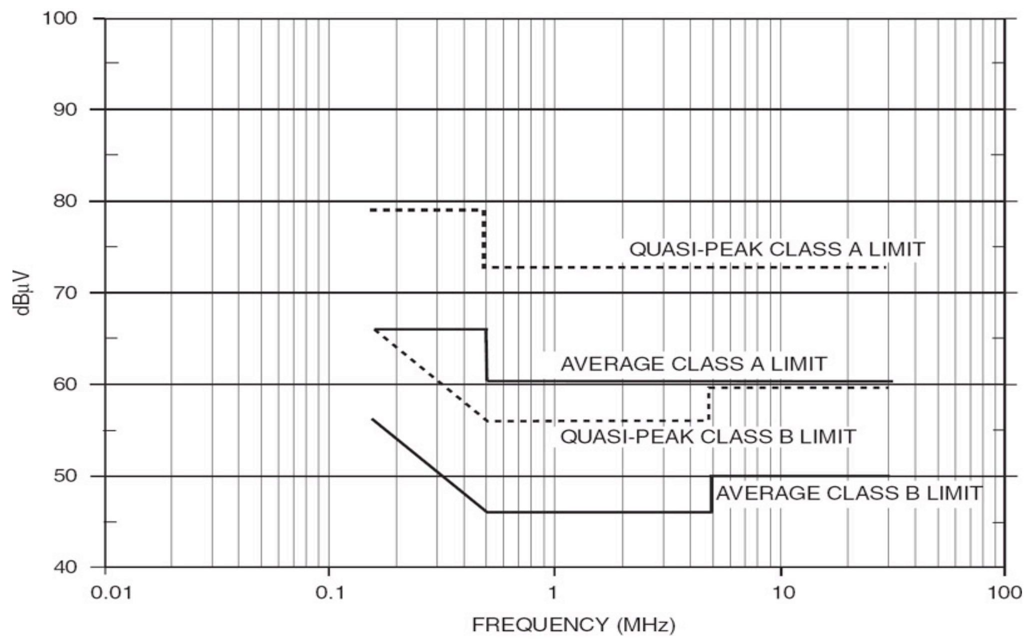


Figure 18-12: FCC/CISPR quasipeak and average conducted emission limits

Figure 13-1 is a plot of the maximum and minimum impedance of the ac power line in the frequency range of 100 kHz to 30 MHz (Nicholson and Malack, 1973).

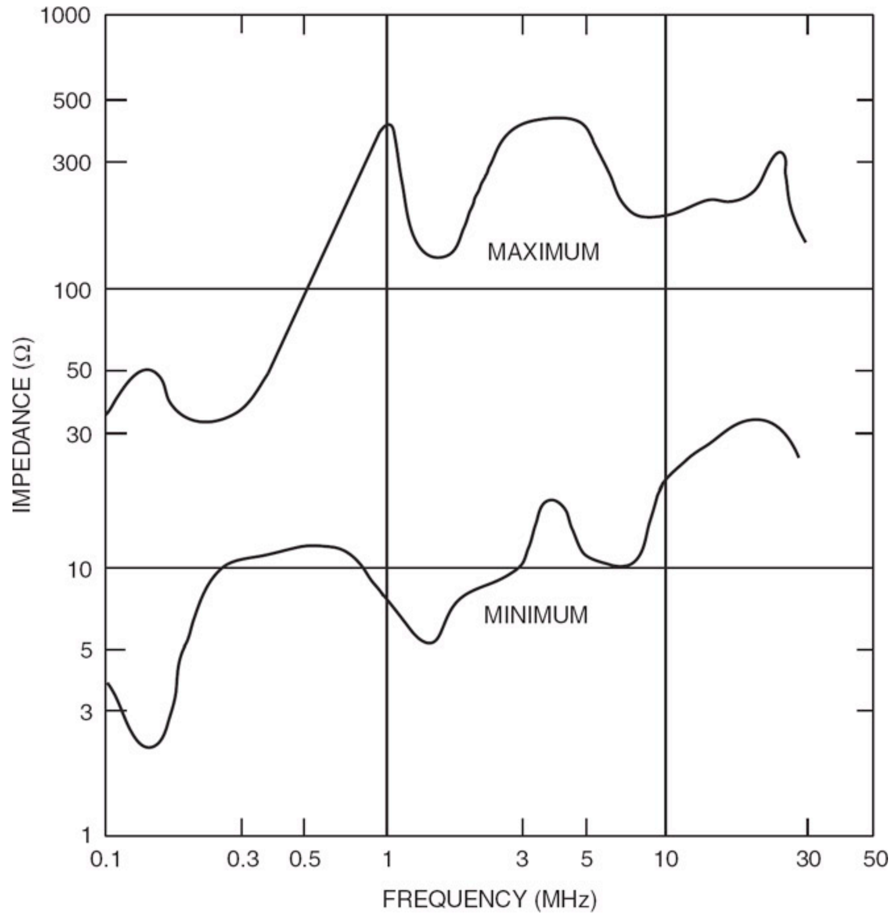


Figure 13-1: Measured impedance of the 115-V ac power line. © *IEEE 1973, reproduced with permission*

Measurement: LISN

During the conducted emission test, a line impedance stabilization network (LISN) is placed between the product and the actual power line in order to present a known impedance to the product's power line terminals over the frequency range of 150 kHz to 30 MHz.

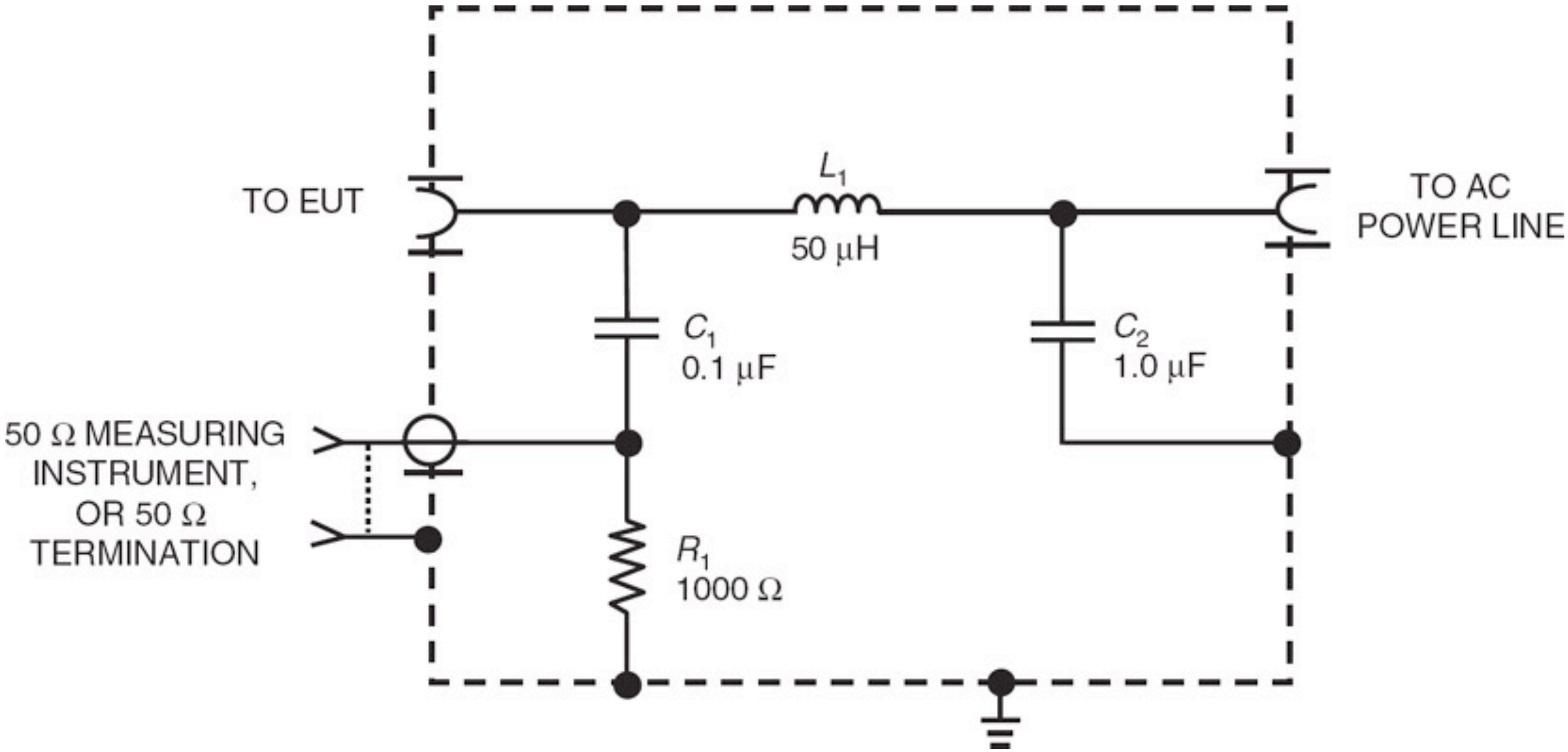


Figure 13-2: Circuit of a 50-μH LISN used for conducted emission testing

MIL-STD-461: LISN

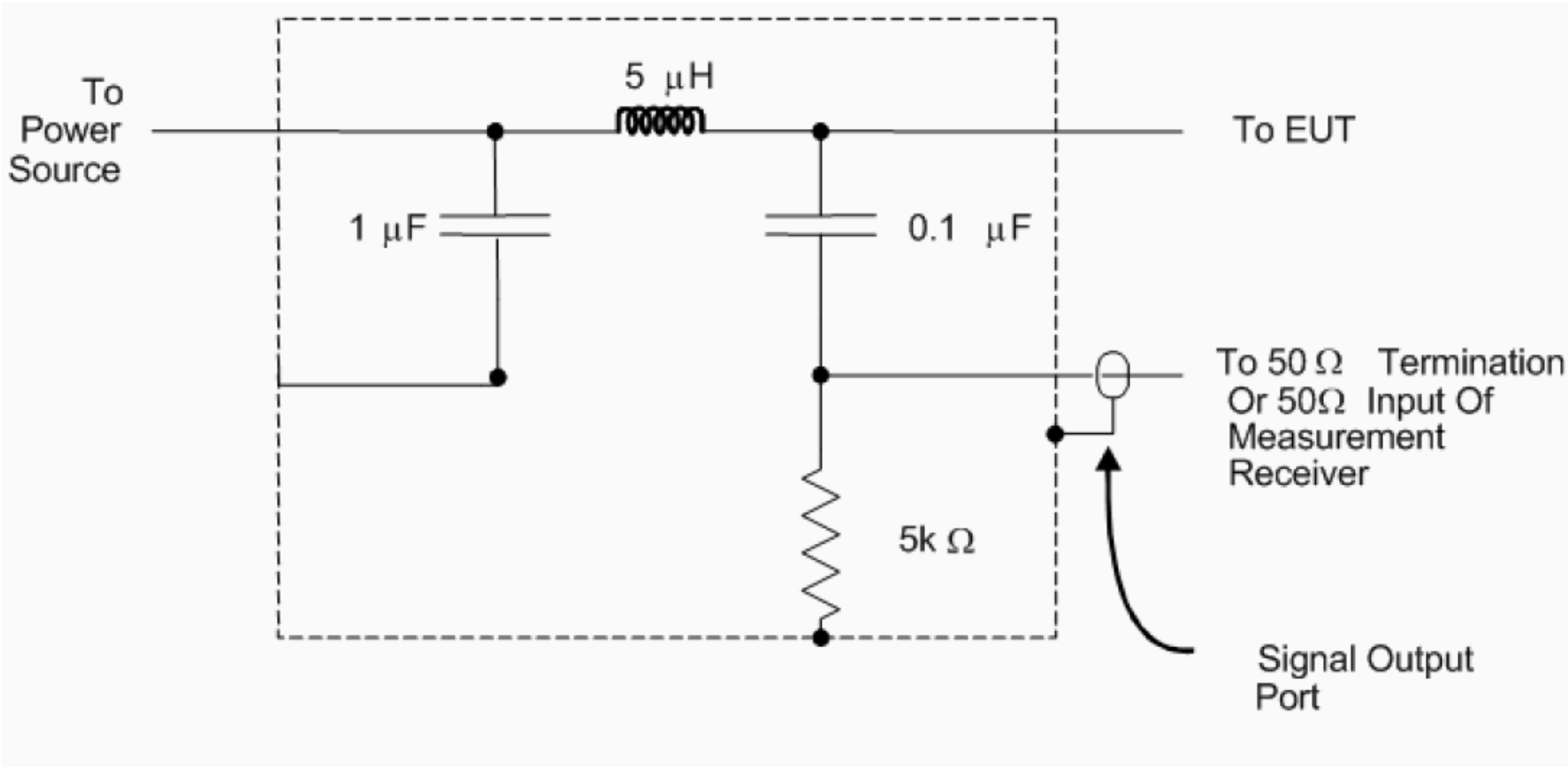


FIGURE A-2. 5 µH LISN schematic.

MIL-STD-461: LISN Impedance

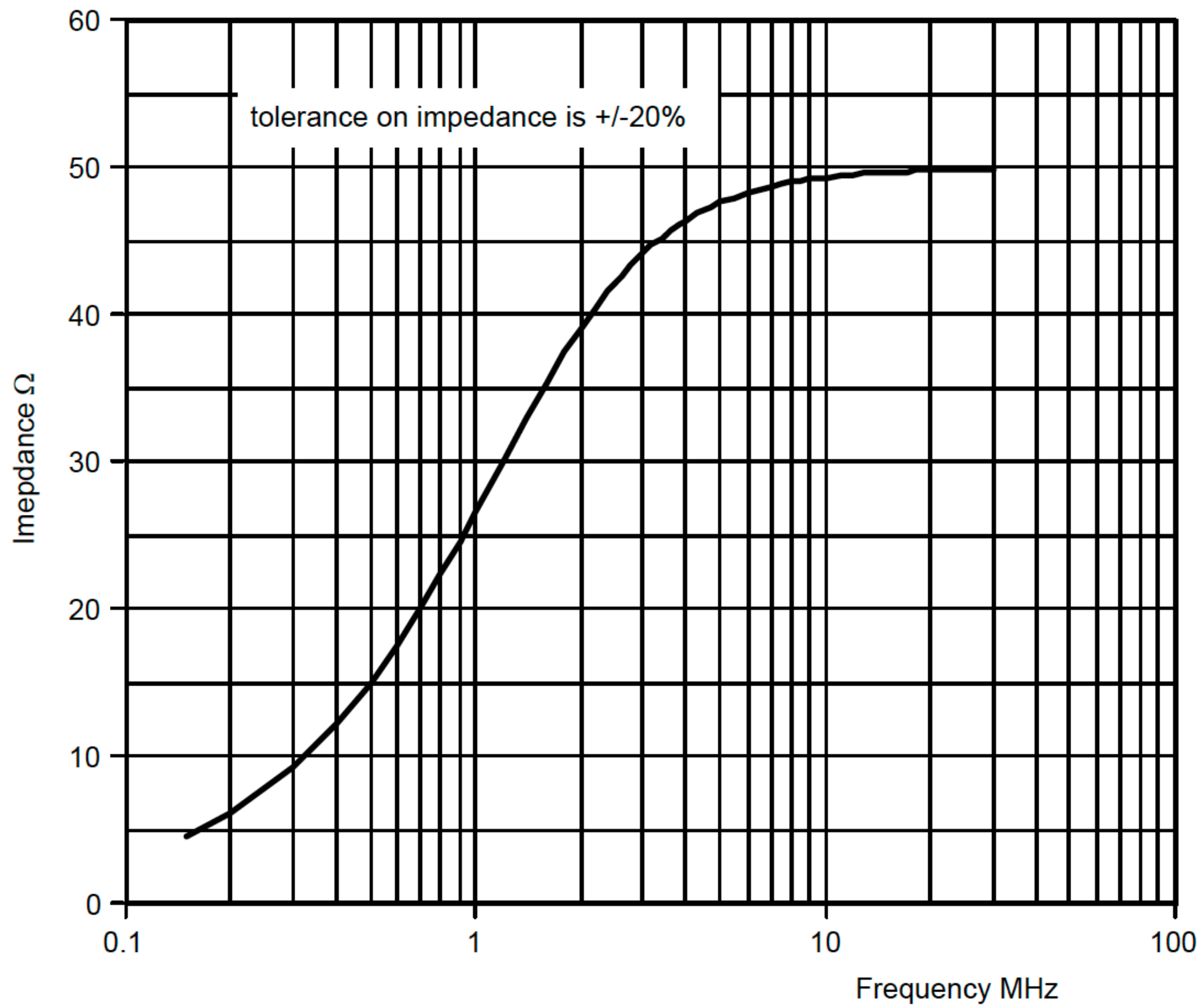


FIGURE A-3. 5 μH LISN impedance.

Power Supply Issues; Example: Flyback converter SMPS

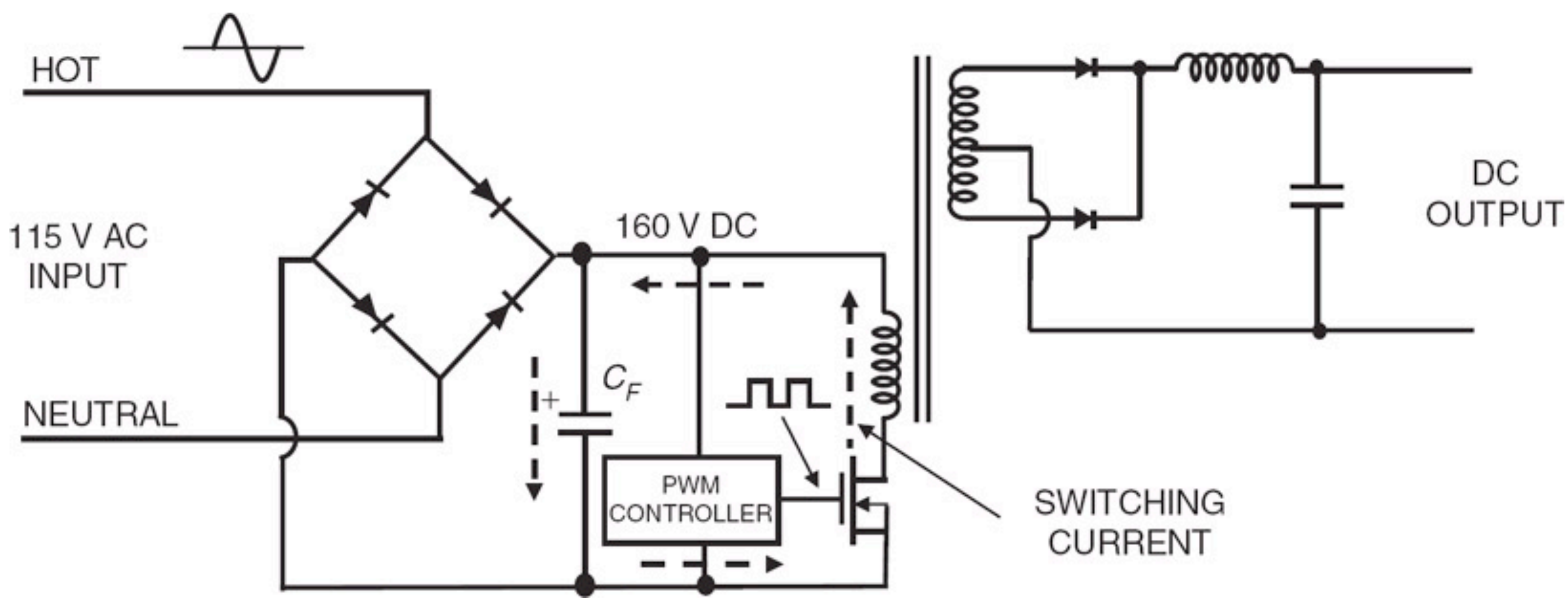


Figure 13-5: Simplified circuit of a flyback converter switched-mode power supply

Full wave rectification of the power line voltage feeding a capacitive input filter results in current spikes on the power line at the peaks of the voltage cycle.
Current waveform has a large amount of harmonic distortion.
The resultant current waveshape is rich in odd harmonics (3rd, 5th, 7th, 9th, etc.) and can cause overheating of the power company's transformers.
In three-phase power distribution systems, produces excessive neutral conductor currents.
Figure 13-4 shows typical voltage and current waveforms present on the input of a switched-mode power supply.
The pulsating ac current drawn not only contains many higher frequency harmonics but also has a much larger peak amplitude than a sine wave would have for the same power rating.

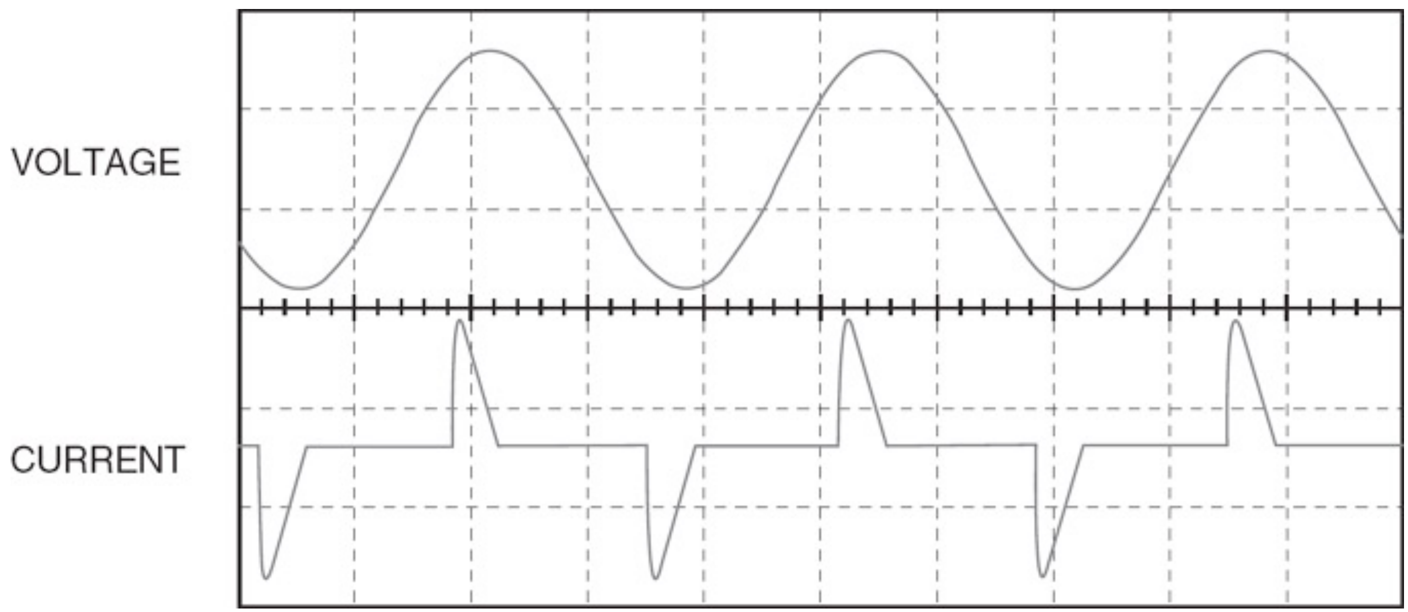


Figure 13-4: SMPS input waveforms. Top trace, voltage; bottom trace, current

Power Line Filter

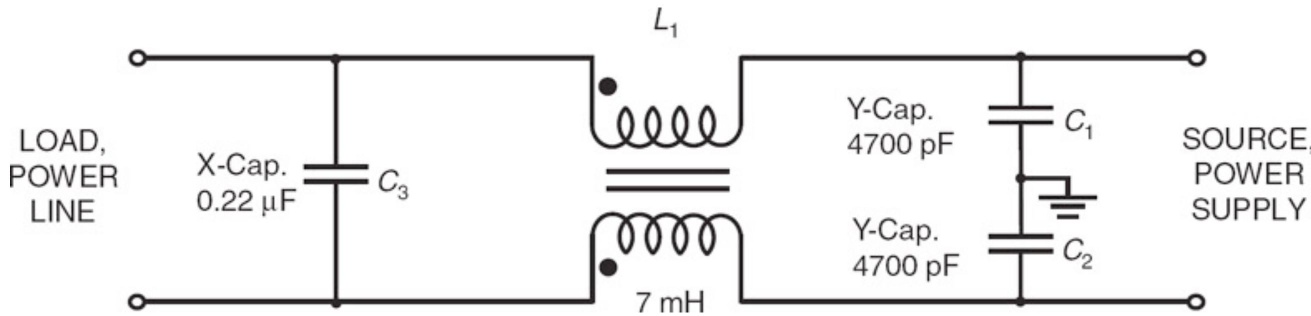


Figure 13-16: Generic power-line filter topology, including typical component values

Very important for filter to be mounted and grounded properly:
Any wire between the filter's case and the enclosure will decrease the effectiveness of the filter as a result of its inductance. Even short wires are too inductive and should be avoided.

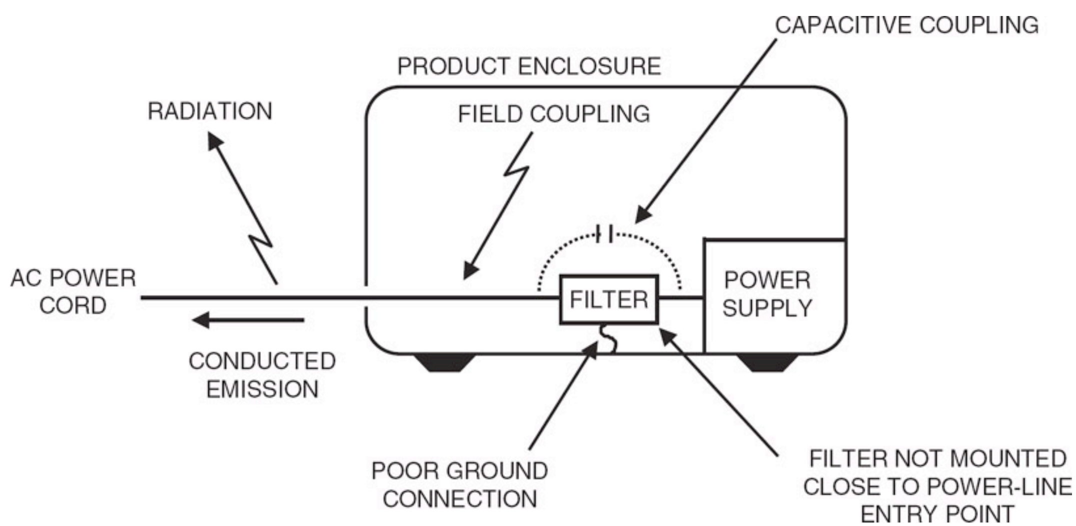


Figure 13-22: Improper power-line filter mounting and grounding

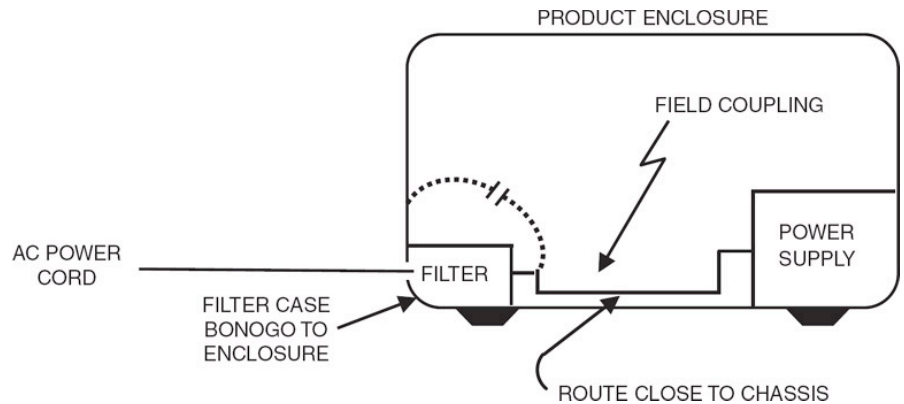


Figure 13-23: A properly mounted, and grounded power-line filter

Integral ac power cord connector (as shown in Fig 13-24) forces the filter to be mounted where the power cord enters the enclosure, and when the filter's metal flange is screwed of riveted to the enclosure (on an unpainted, conductive surface) the Y-capacitors will be properly grounded.
The importance of proper power-line filter mounting and wiring cannot be overemphasized.

Figure 13-24: Commercial power-line filter with integral power cord connector

Magnetic Field Emissions

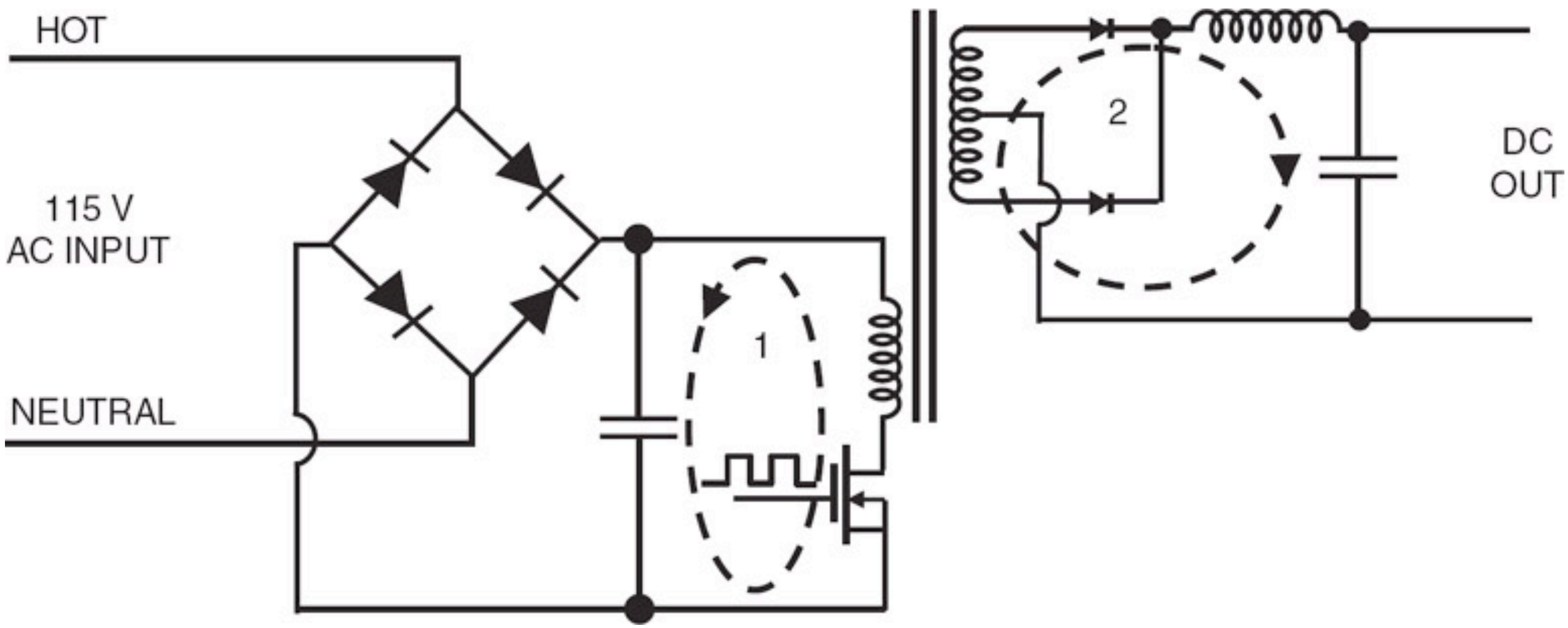


Figure 13-28: Critical loop areas in a switched-mode power supply. (1) The switching transistor loop (primary loop) and (2) the rectifier loop (secondary loop)

Reducing Harmonics

The use of an inductive input filter will spread out the current waveform, and in many cases reduce the harmonics sufficiently to make the product compliant.

The inductor limits the *di/dt*, and therefore it slows down the rise of the current waveform. In addition, it spreads out the current pulse and by decreases the peak amplitude, which reduces the total harmonic content of the waveform.

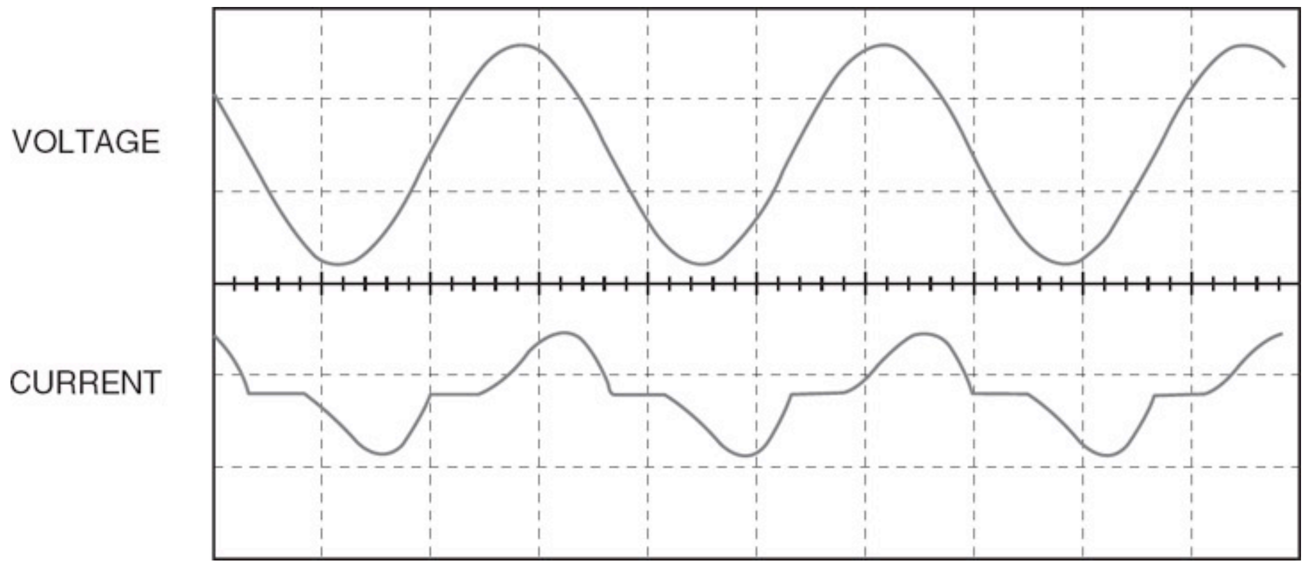


Figure 13-38: Input waveforms for a switched-mode power supply with an inductive input filter. Top trace, voltage; bottom trace, current

Active Power Factor Correction

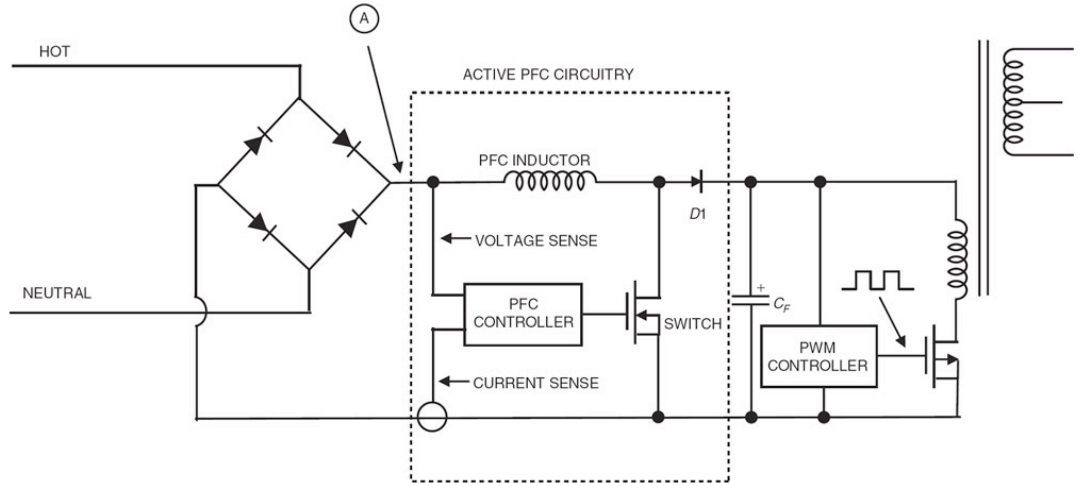


Figure 13-39: A switched-mode power supply with active power factor correction

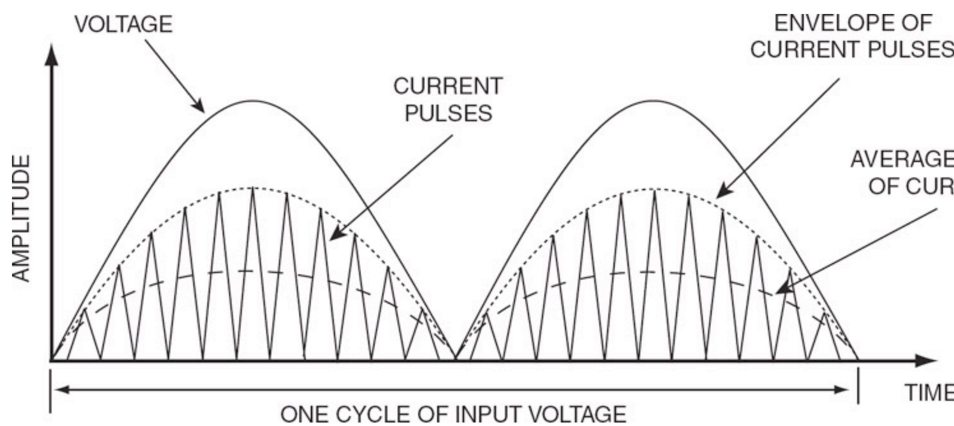


Figure 13-40: Active power factor corrector circuit input voltage and current waveforms. Waveforms shown are at point A in Fig. 13-39

Summary: Conducted emissions

The ac power line impedance from100 kHz to 30 MHz can vary from about 2 to 450 Ω.

The effectiveness of a power-line filter is as much a function of how and where the filter is mounted, and how the leads are run to it, as it is of the electrical design of the filter.

When the power-line filter is on the same PCB as the power supply, magnetic field coupling, and high ground inductance in series with the Y-capacitors can significantly reduce the effectiveness of the filter.

To reduce magnetic fields in power supplies:

- Minimize the area of all loops containing a large di/dt.
- Use a toroidal core transformer instead of an E-core transformer.
- Add a shorted turn to an E-core transformer.
- Shield the unit with steel or other magnetic material, not aluminum.

A bridge capacitor can be effective in reducing certain types of common-mode conducted emission in switched-mode power supplies.

Intentionally dithering (varying) the frequency of the power supply is another method of reducing the conducted emissions, both common - and differential-mode.

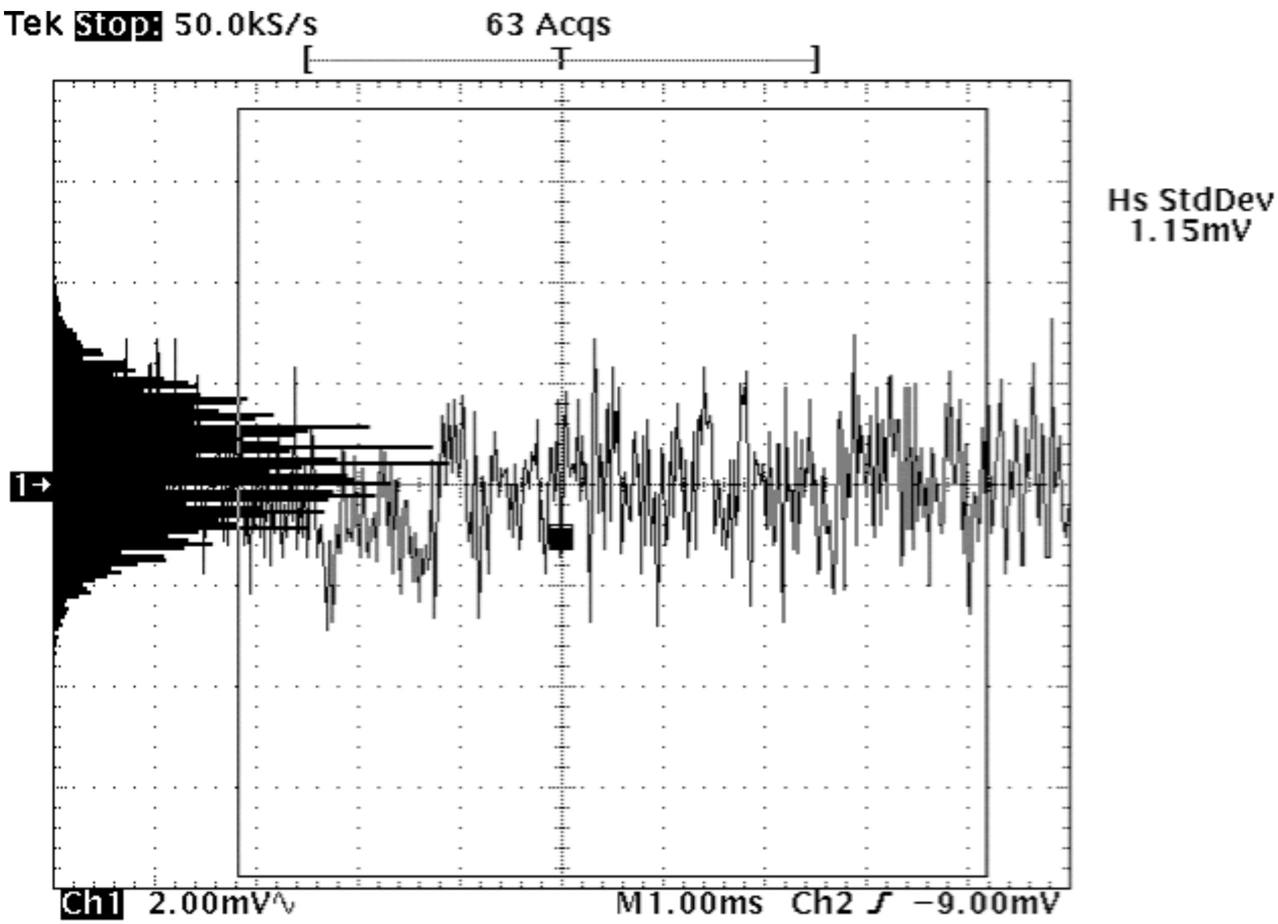
The primary generator of harmonics on the ac power line is a full wave rectifier followed by a capacitor input filter.

The following methods can be used to reduce harmonic currents on the ac power line in switched-mode power supplies and variable speed motor drives:

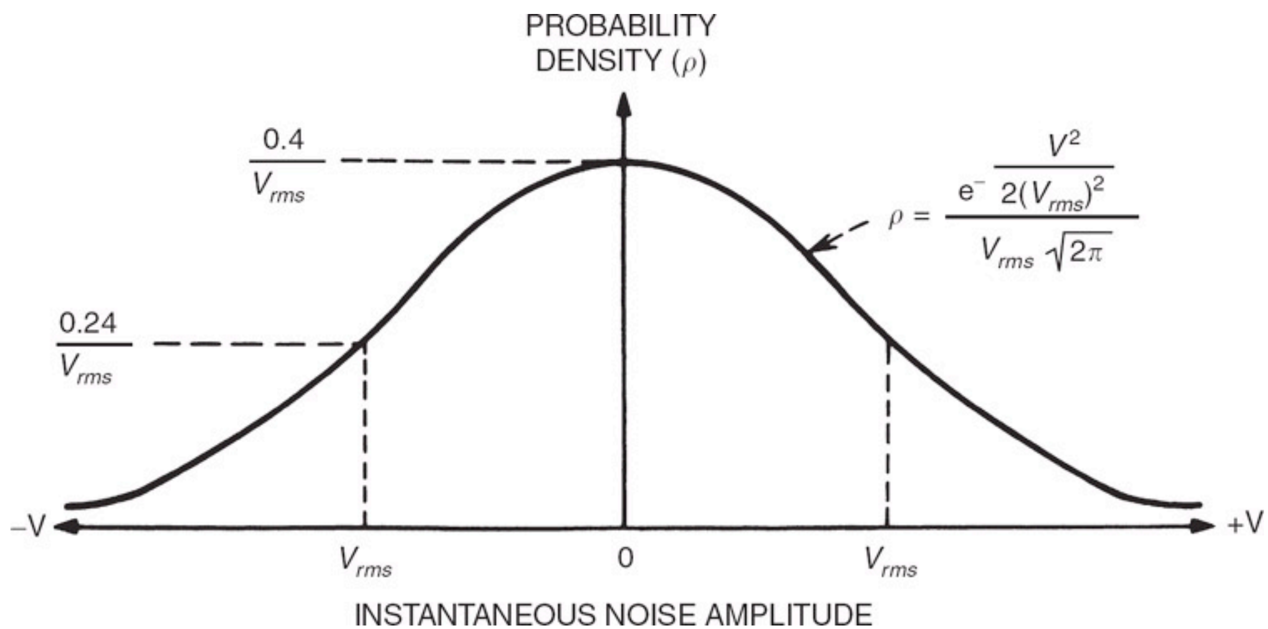
- Use an inductive input filter instead of a capacitive input filter,
- Use active power factor correction circuitry, or
- Add a power factor correction inductor (line reactor) on the ac power line.

Fundamental Noise Sources (chapter 8, 9)

Describing Noise: Time Domain



Gaussian White Noise Probability Density

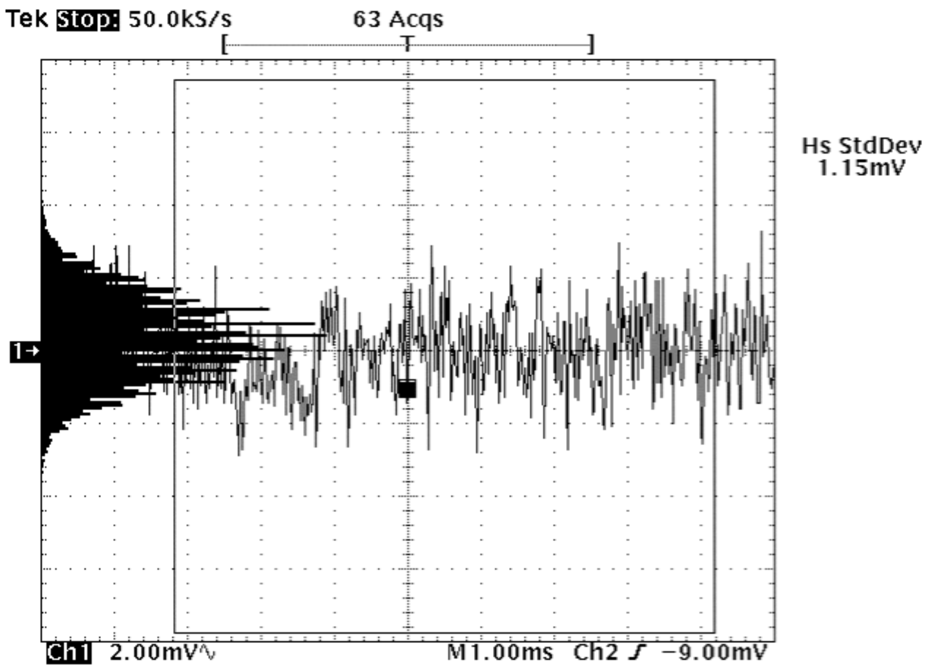


If scope / measurement instrument does not have histogram capability, estimate rms standard deviation from peak-to-peak measurement and crest factor correction.

Table 8-1: Crest Factors for Thermal Noise

Percent of Time Peak Exceeded	**Crest Factor (peak/rms)**
1	2.6
0.1	3.3
0.01	3.9
0.001	4.4
0.0001	4.9

Describing Noise: Frequency Domain



Intuition: Reducing bandwidth will reduce rms noise. Quantitatively?

Equivalent Noise Bandwidth

Noise bandwidth is defined for a system with uniform gain throughout the passband and zero gain outside the passband.
Figure 8-7 shows this ideal response for a low-pass circuit and a bandpass circuit.

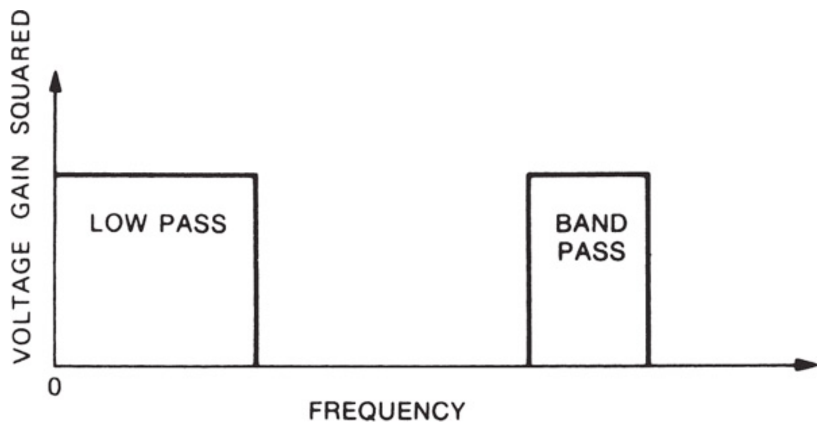


Figure 8-7: Ideal bandwidth of low-pass and band-pass circuit elements

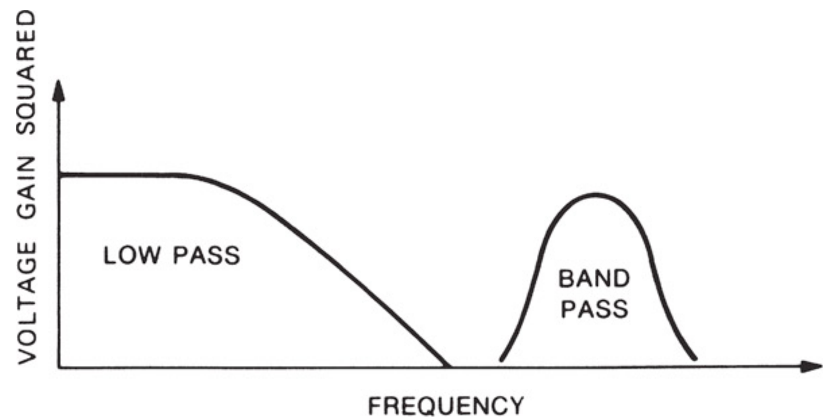
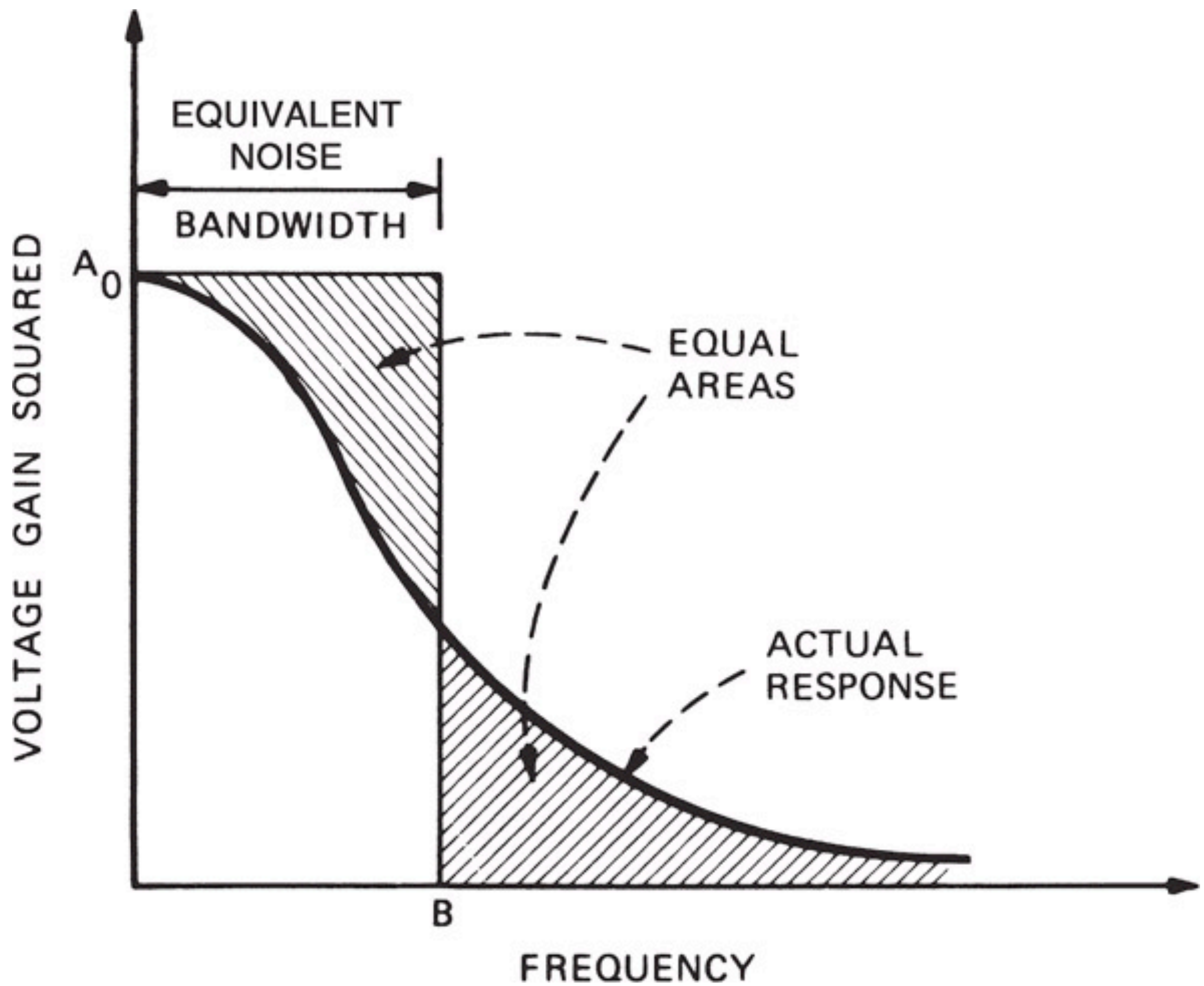


Figure 8-8: Actual bandwidth of low-pass and band-pass circuit elements

Practical circuits do not have these ideal characteristics but have responses similar to those shown in Fig. 8-8.
The problem then is to find an equivalent noise bandwidth that can be used in equations to give the same results as the actual nonideal bandwidth does in practice.

Equivalent Noise Bandwidth



Noise BW for different transfer functions

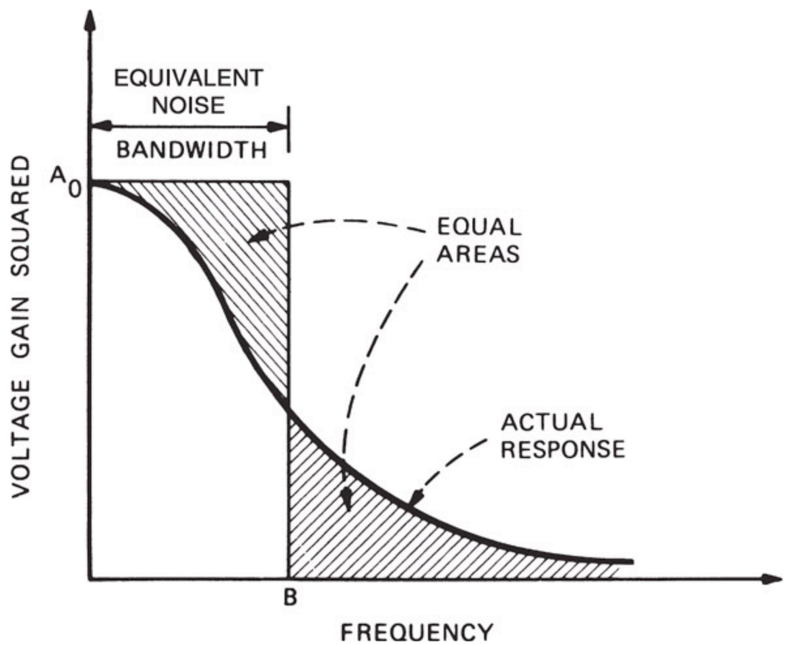


Table 8-2: Ratio of the Noise Bandwidth B to the 3-db Bandwidth f_0

Number of Poles	***B/f₀***	**High-Frequency Rolloff (dB per octave)**
1	1.57	6
2	1.22	12
3	1.15	18
4	1.13	24
5	1.11	30

Example: A noise source of 1.6E-16 V^2/Hz passes through a system with a single-pole (-6dB/decade rolloff) bandwidth of 65 kHz. What is the rms noise at the output?

Where do noise density expressions for resistor, active devices come from?

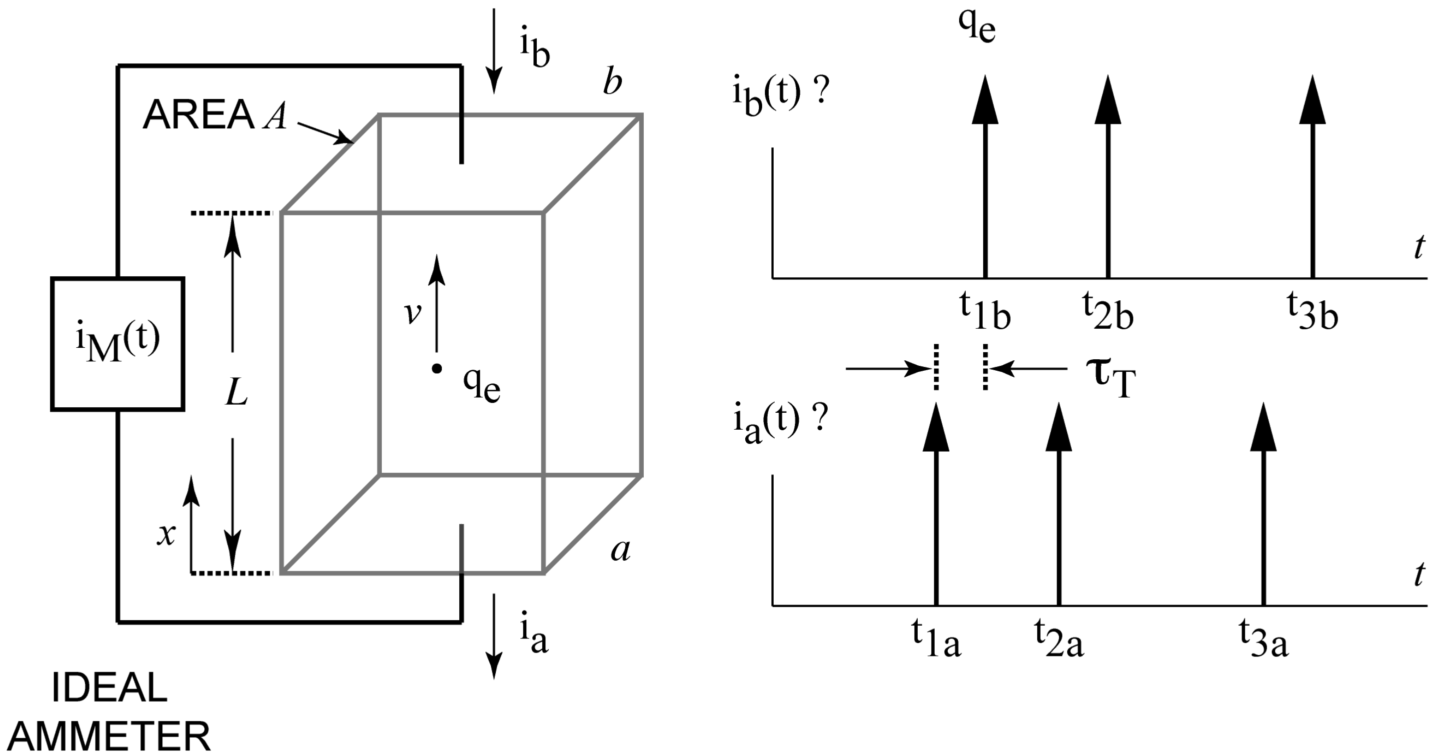
Shot Noise

- Current noise density for bipolar transistor DC collector current I_{DC}

$$i_n^{\ 2} = 2q_e I_{DC}$$

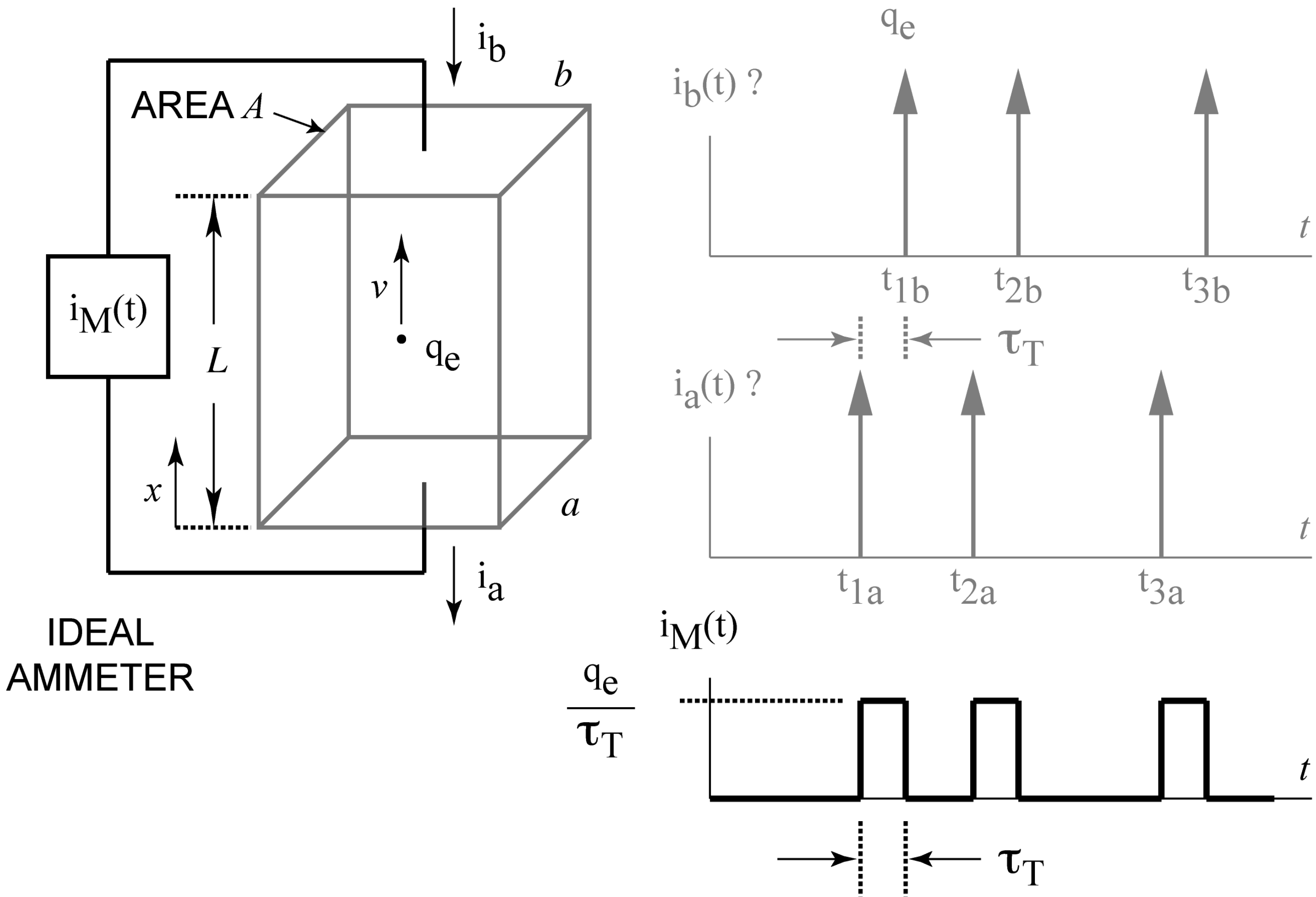
- Where does this come from?
- Key assumption:
  - Electron arrivals independent events

Shot Noise



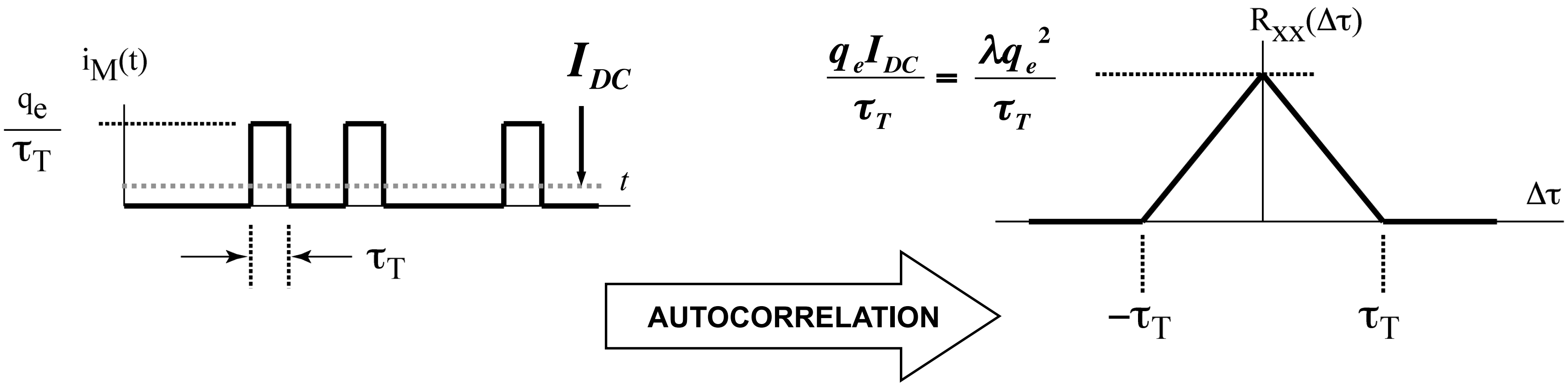
- **What is current measured by ammeter?**

Ramo-Shockley Theorem



- **Current measured by ammeter:**
  - **–Randomly arriving pulses with area q_e**

Poisson Process

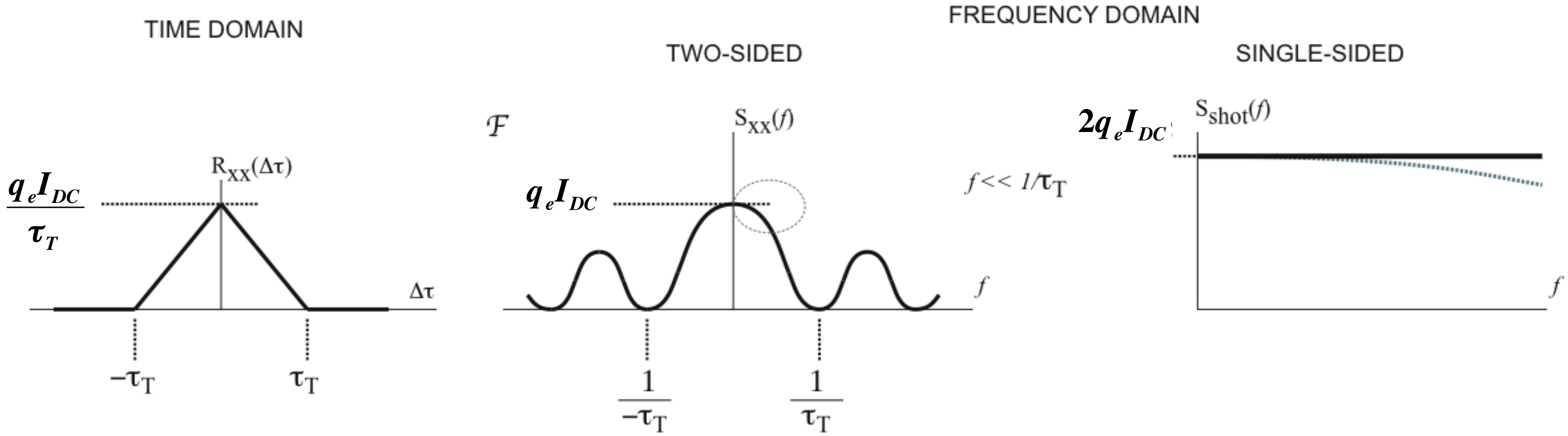


- **Average arrival rate λ [sec^{-1}]**
- **Average DC current:**

$$I_{DC} = \lambda q_e$$

- **Autocorrelation: time domain description of random process**

Shot Noise Power Spectral Density



- **Wiener-Khinchine theorem**
  - **Autocorrelation → frequency domain p.s.d**
- **Frequency domain**
  - **For frequencies < 1/τ_T**

$$i_n^2 = 2q_e I_{DC}$$

Shot Noise Power Spectral Density

- **Key Points:**
  - **Discrete nature of charge is essential**
  - **Carrier transits are independent events**
  - **Carriers do not interact with each other or with any medium**
  - **Temperature not a factor**

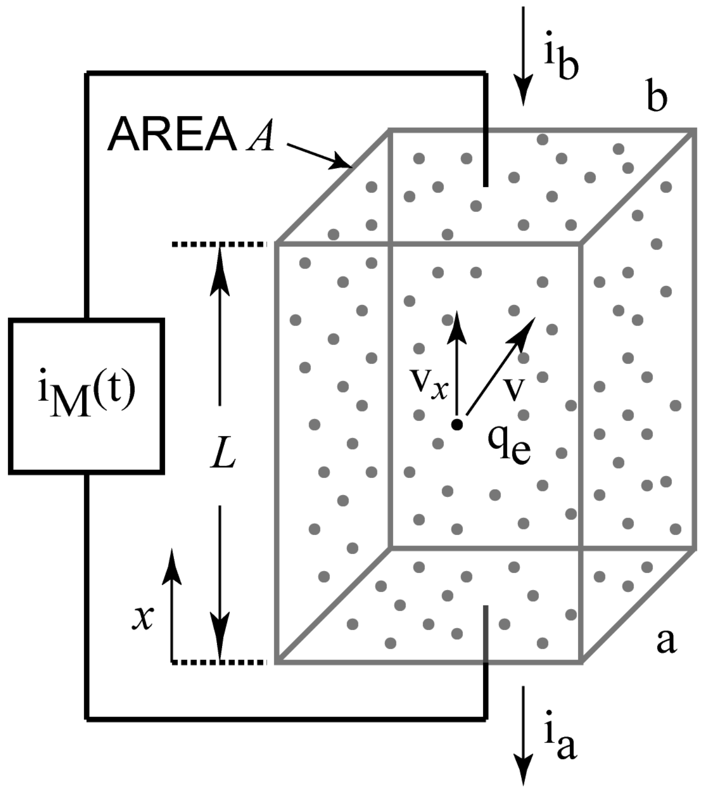
Thermal Noise

- Current noise density for resistor

$$i_n^2 = \frac{4kT}{R}$$

- Where does this come from?
- Assumption:
  - Carriers in thermal equilibrium

Thermal Noise in Resistor



- **Assumption:**
  - **Carriers in thermal equilibrium**
- **Random velocity vectors** v
- **Only** v_x **component contributes to current**

Boltzmann's Constant k

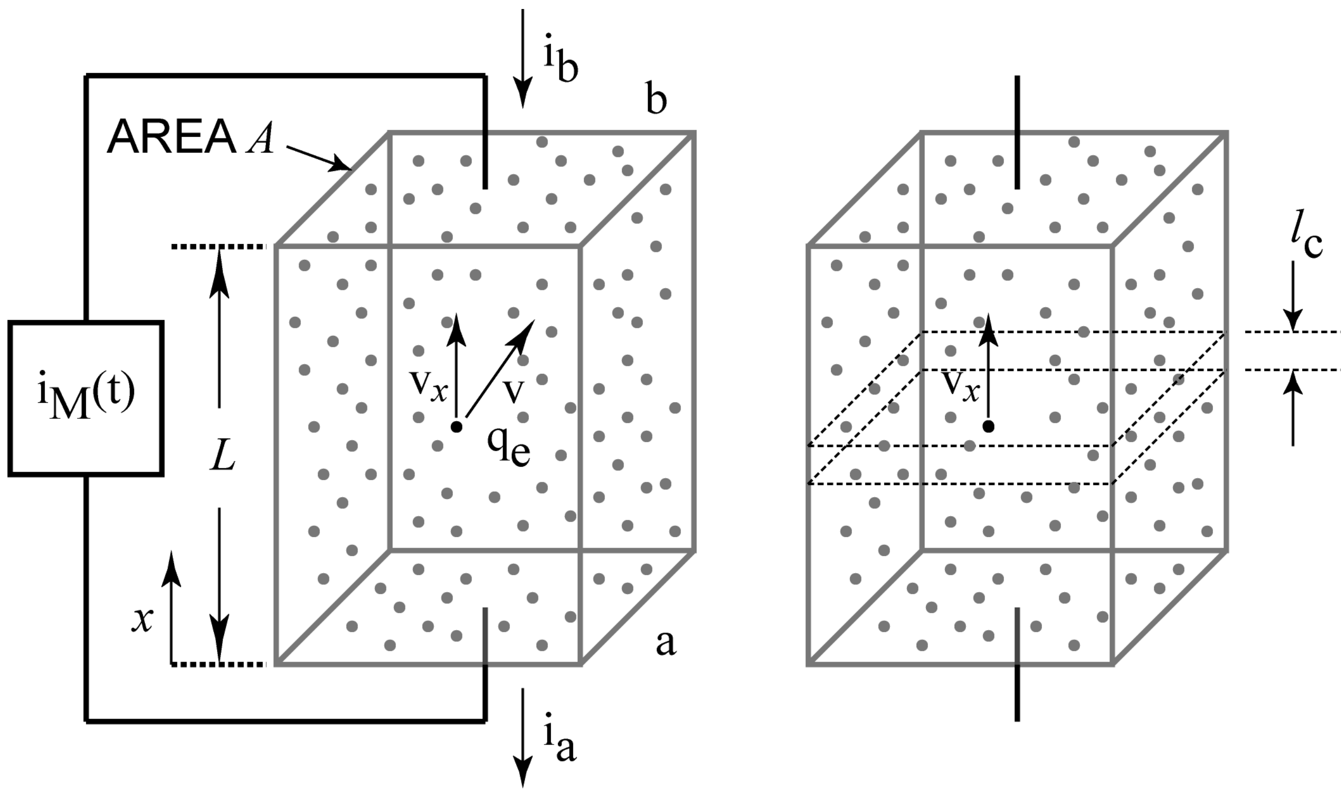
- k = 1.38 E-23 J/K Meaning?
- Thermodynamics: Equipartition theorem
  - Independent energy storage modes in a system at equilibrium have average energy of kT/2
  - Equivalent statements:

"Temperature in this room is 293K"

"Average kinetic energy (in each of x, y, z directions) for each air molecule in this room is 2.02E-21 joule"

$$\frac{kT}{2} = \frac{\left(1.38E-23\left[J/K\right]\right)\left(293\left[K\right]\right)}{2} = 2.02E-21\left[J\right]$$

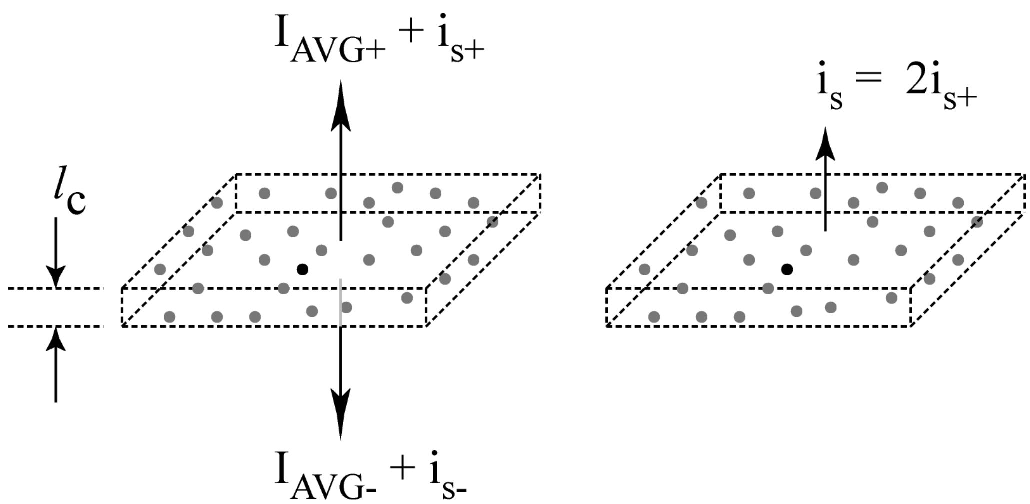
Thermal Noise



- **Approximate collision statistics (Si):**

Mean free path	l_c	**0.1 μm**
Mean free time	τ_c	**1 ps**
Velocity (rms)	v_x	**0.1 μm/ps**

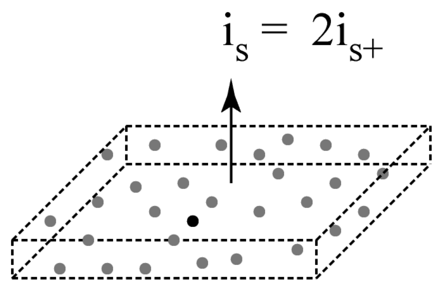
Thermal Noise

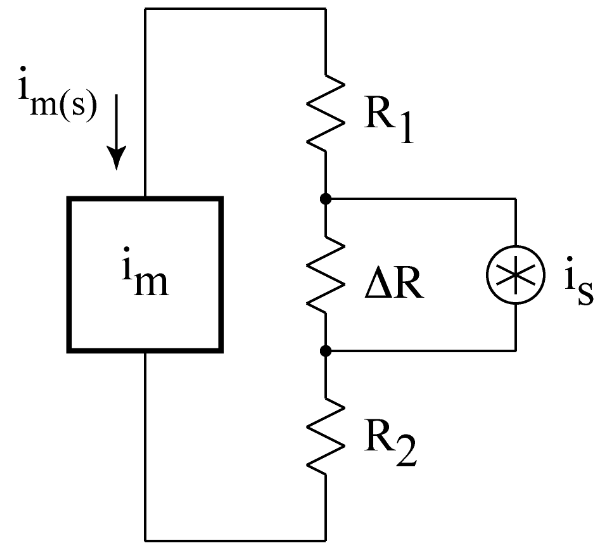


- **Consider "slice" equal to mean free path l_c**
- **During one mean free time τ_c**
  - **On average, half of carriers exit each way:** $I_{AVG+} = I_{AVG-}$
- **Shot noise components $i_{s+} = -i_{s-}$ correlated**
  - **Noise current from "slice" $i_s = 2i_{s+}$**

Sarpeshkar, Delbruck, and Mead, "White noise in MOS transistors and resistors," *IEEE Circuits & Devices Magazine*, Nov. 1993

Thermal Noise



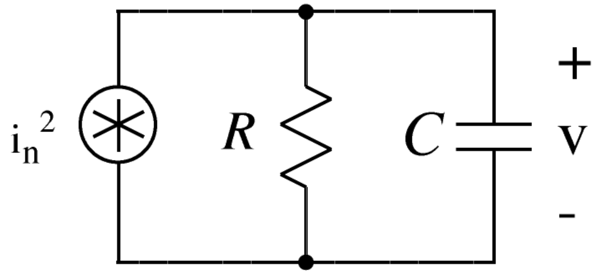


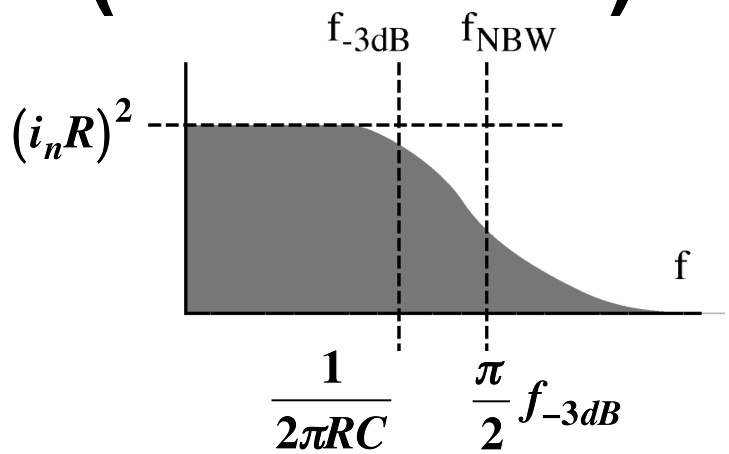
- **Sum (independent) contributions from slices**
- **Noise current seen by external ammeter $i_{m(s)}$ reduced by current divider factor: ΔR of slice, total resistance $R = R_1 + \Delta R + R_2$**
- **Relating to R using mobility definition gives**

$$i_n^{\,2} = \frac{4kT}{R}$$

Sarpeshkar, Delbruck, and Mead, "White noise in MOS transistors and resistors," *IEEE Circuits & Devices Magazine*, Nov. 1993

Thermal Noise (Alternative)





- **Equipartition, rms energy in capacitor:**

$$\frac{1}{2} C v^2 = \frac{kT}{2} \quad \Rightarrow \quad v^2 = \frac{kT}{C}$$

- **Integrate noise p.s.d. over noise bandwidth:**

$$v^2 = (i_n R)^2 \left(\frac{\pi}{2} \frac{1}{2\pi RC} \right) \quad \Rightarrow \quad v^2 = i_n{}^2 \frac{1}{4RC}$$

- **Equate:**

$$\frac{kT}{C} = i_n{}^2 \frac{1}{4RC} \quad \Rightarrow \quad \boxed{i_n{}^2 = \frac{4kT}{R}}$$

Thermal Noise Power Spectral Density

- Key Points:
  - Discrete nature of charge is not essential
    - Can also be derived from equipartition only (e.g. kT/C noise)
  - Carrier scattering: interact with medium, thermal equilibrium
  - Carrier transits are not independent due to interaction with medium
  - Temperature is important to determine carrier average kinetic energy / velocity

Resistor Noise Models

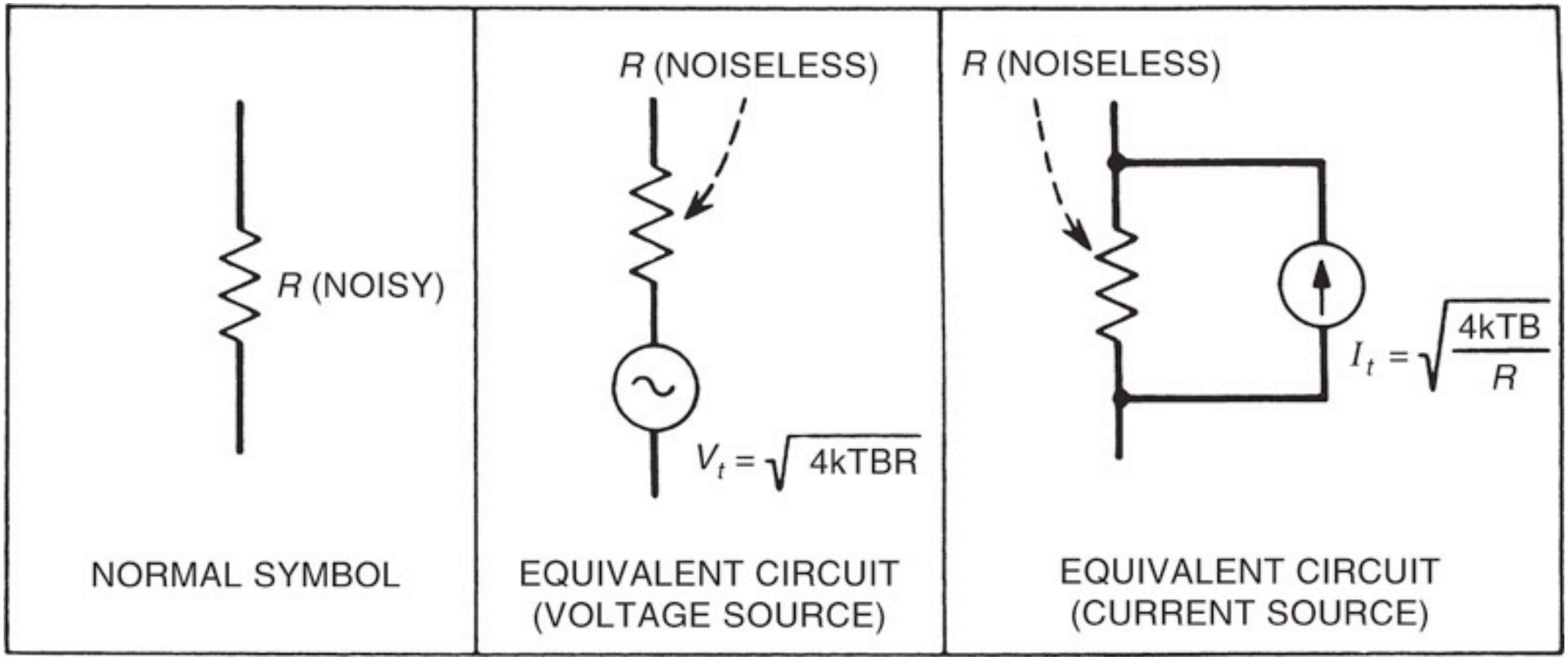


Figure 8-2: Thermal noise in a resistor (left) can be represented in an equivalent circuit as a voltage source (center) or a current source (right)

Note: Ideal capacitor or inductor does not contribute thermal noise.
For a lossy (real) L or C, the real part of impedance will have a corresponding resistive thermal noise term.

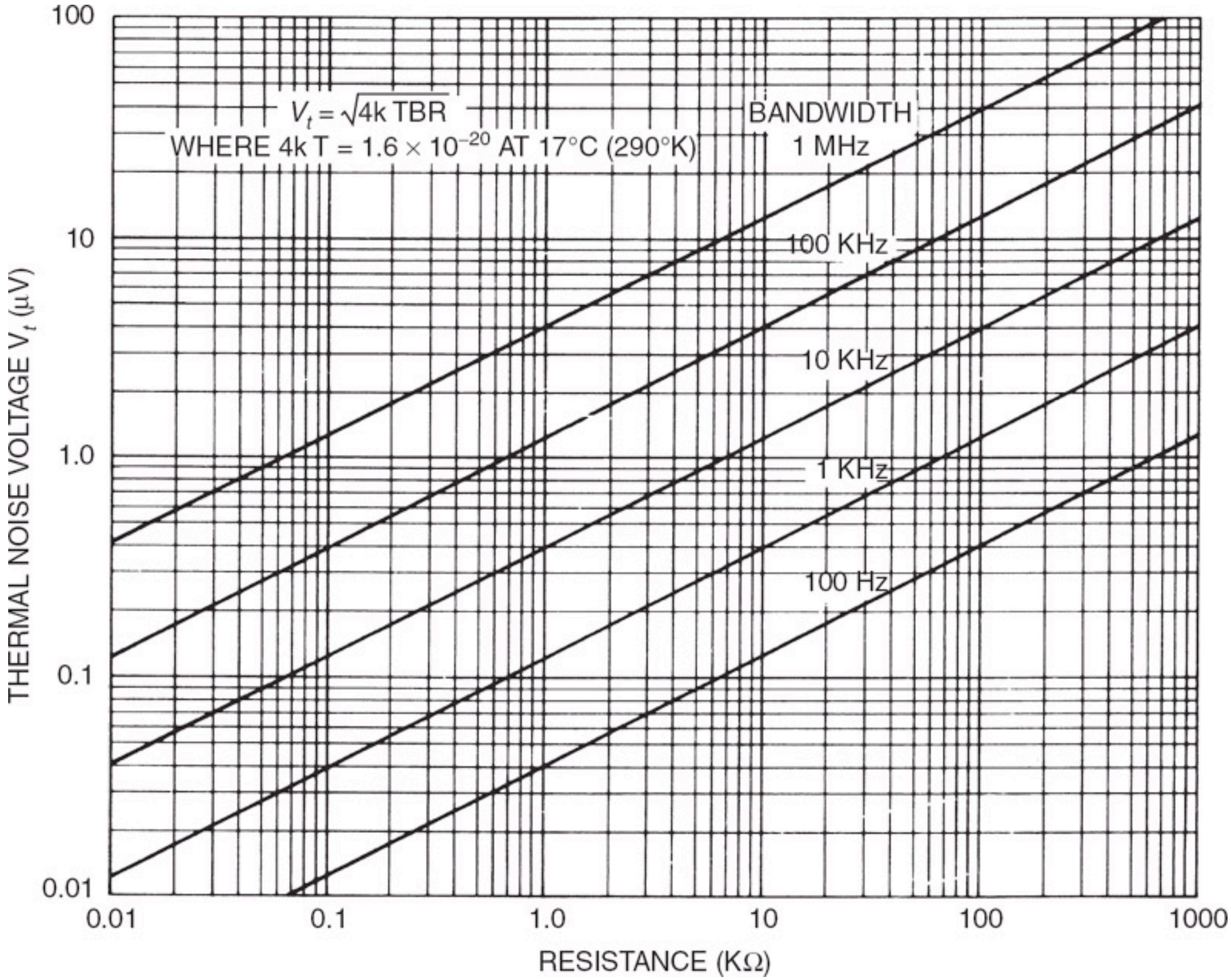


Figure 8-1: Thermal noise voltage as a function of resistance and bandwidth

Amplifier Noise Models

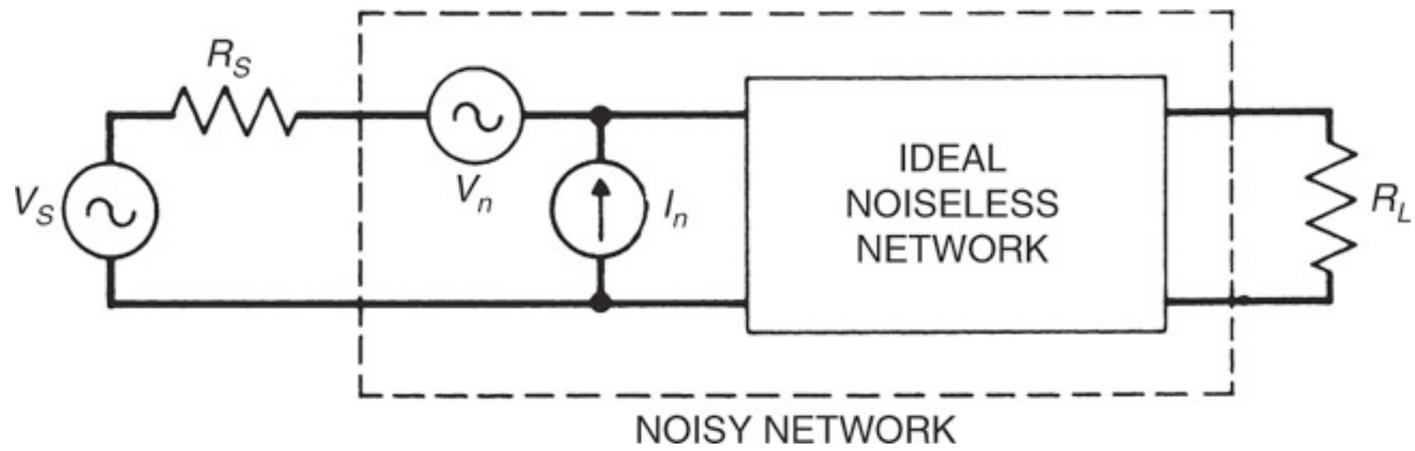


Figure 9-4: A noisy network modeled as an ideal noiseless network with the addition of an input noise voltage and input noise current source

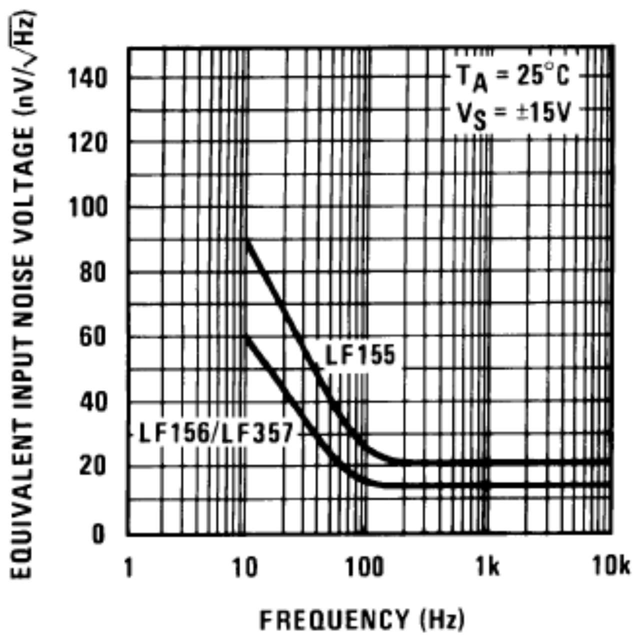
Example: LF356 op-amp noise specification

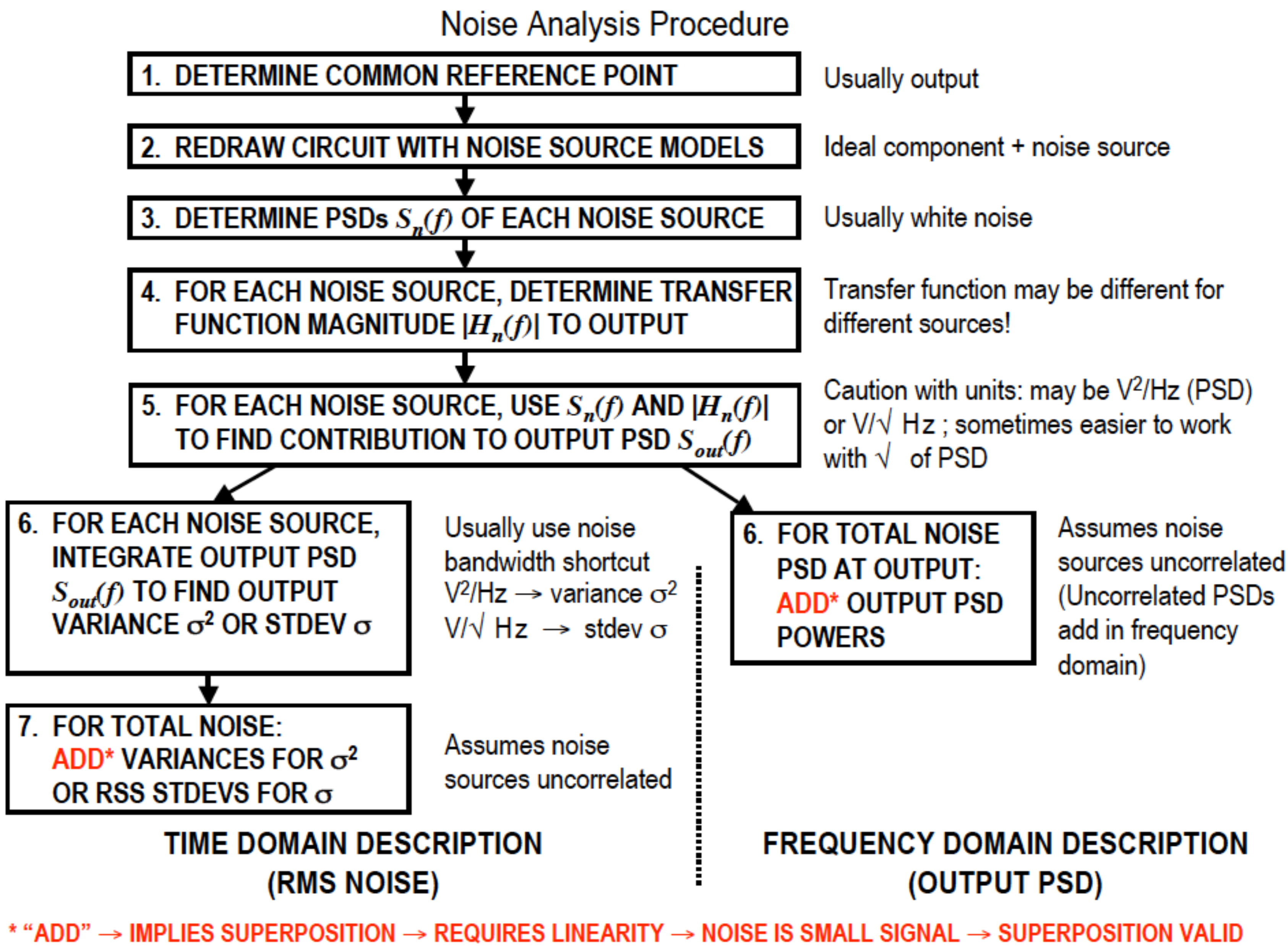
AC Electrical Characteristics

$T_A = T_J = 25^{\circ}C$, $V_S = \pm 15V$

Symbol	Parameter	Conditions	LF155/355	LF156/256/356B	LF156/256/356/LF356B	LF257/357	Units
			Typ	Min	Typ	Typ	
e_n	Equivalent Input Noise Voltage	$R_S=100\Omega$					
		f=100 Hz	25		15	15	$nV/\sqrt{Hz}$
		f=1000 Hz	20		12	12	$nV/\sqrt{Hz}$
i_n	Equivalent Input Current Noise	f=100 Hz	0.01		0.01	0.01	$pA/\sqrt{Hz}$
		f=1000 Hz	0.01		0.01	0.01	$pA/\sqrt{Hz}$

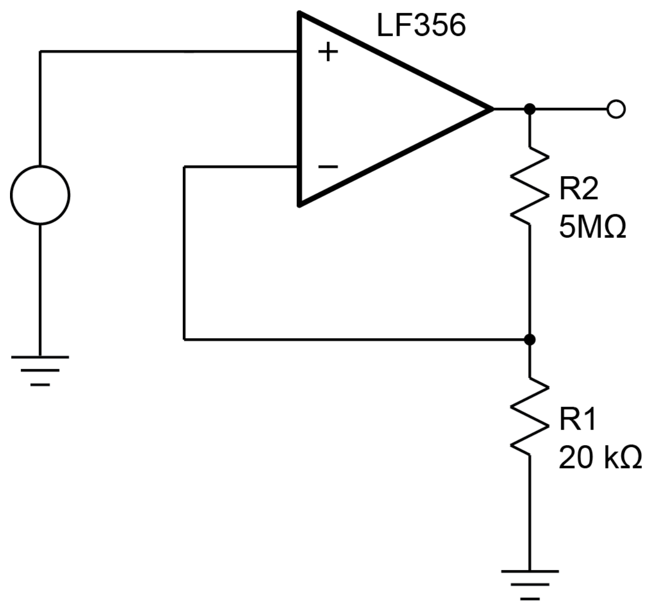
Equivalent Input Noise Voltage



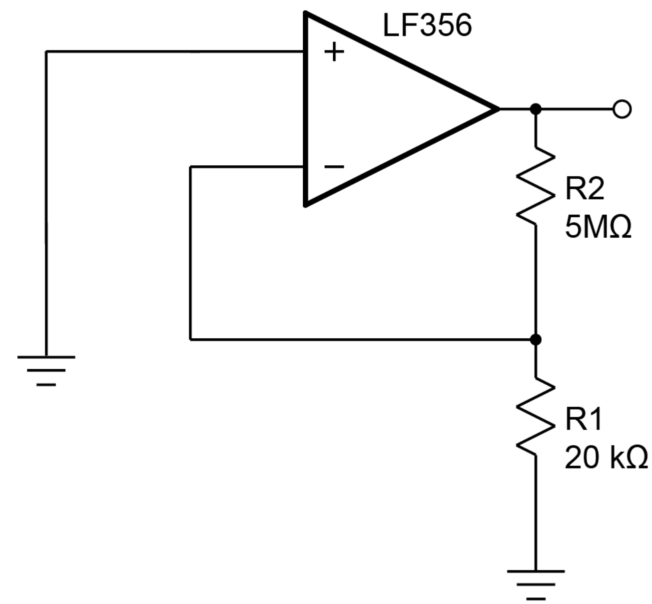

Noise Analysis Procedure
1. DETERMINE COMMON REFERENCE POINT
Usually output
2. REDRAW CIRCUIT WITH NOISE SOURCE MODELS
Ideal component + noise source
3. DETERMINE PSDs $S_n(f)$ OF EACH NOISE SOURCE
Usually white noise
4. FOR EACH NOISE SOURCE, DETERMINE TRANSFER FUNCTION MAGNITUDE $|H_n(f)|$ TO OUTPUT
Transfer function may be different for different sources!
5. FOR EACH NOISE SOURCE, USE $S_n(f)$ AND $|H_n(f)|$ TO FIND CONTRIBUTION TO OUTPUT PSD $S_{out}(f)$
Caution with units: may be V^2/Hz (PSD) or $V/\sqrt{Hz}$; sometimes easier to work with $\sqrt{}$ of PSD
6. FOR EACH NOISE SOURCE, INTEGRATE OUTPUT PSD $S_{out}(f)$ TO FIND OUTPUT VARIANCE σ^2 OR STDEV σ
Usually use noise bandwidth shortcut
$V^2/Hz \rightarrow$ variance σ^2
$V/\sqrt{Hz} \rightarrow$ stdev σ
7. FOR TOTAL NOISE: ADD* VARIANCES FOR σ^2 OR RSS STDEVS FOR σ
Assumes noise sources uncorrelated
6. FOR TOTAL NOISE PSD AT OUTPUT: ADD* OUTPUT PSD POWERS
Assumes noise sources uncorrelated (Uncorrelated PSDs add in frequency domain)
TIME DOMAIN DESCRIPTION (RMS NOISE)
FREQUENCY DOMAIN DESCRIPTION (OUTPUT PSD)
* "ADD" → IMPLIES SUPERPOSITION → REQUIRES LINEARITY → NOISE IS SMALL SIGNAL → SUPERPOSITION VALID

Example: Noise of LF356 gain of 250 op-amp amplifier

Amplifier:

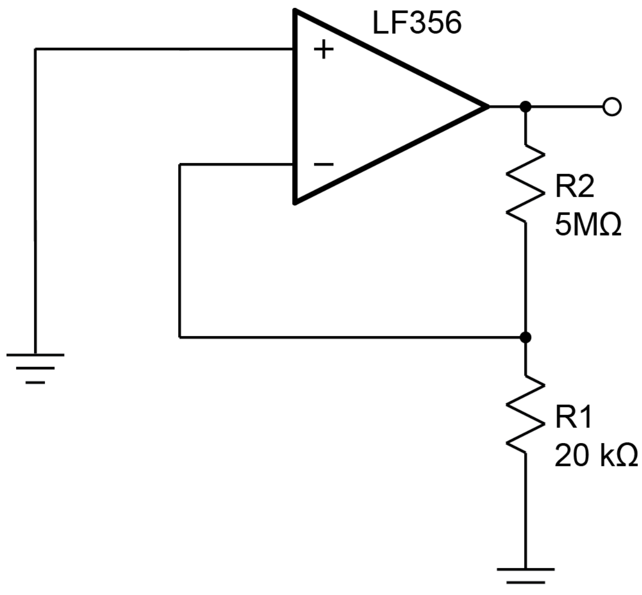


Steps 1, 2: Noise model:



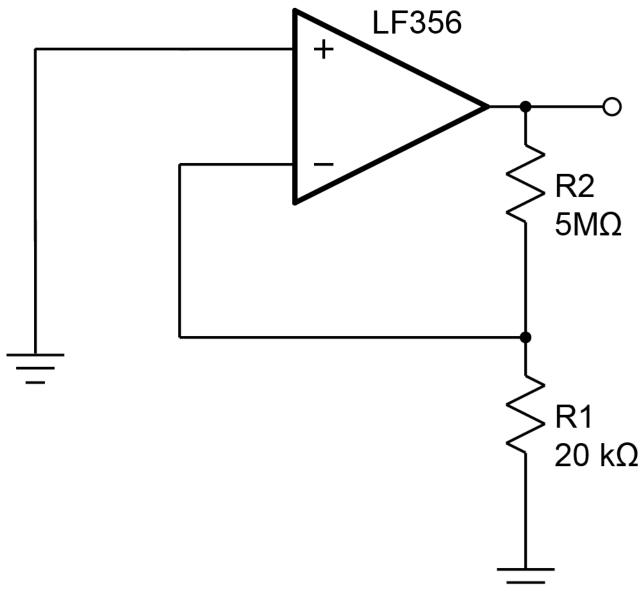
Example: Noise of LF356 gain of 250 op-amp amplifier

Steps 3, 4, 5, 6: Op-amp voltage noise e_n

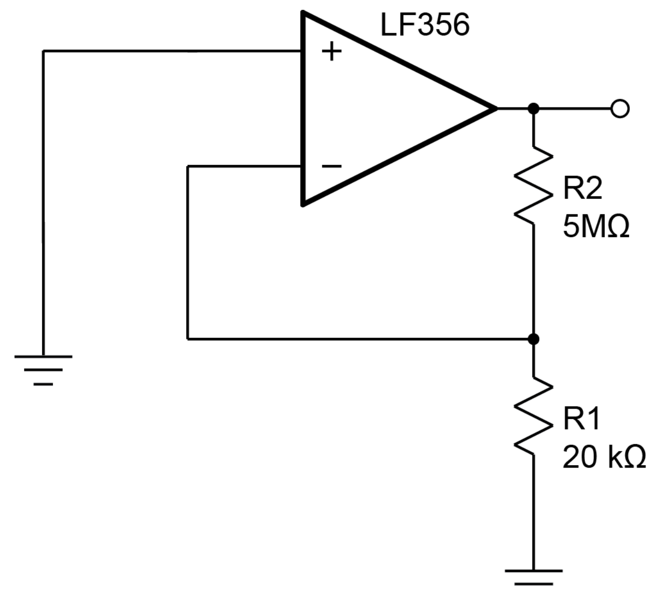


Example: Noise of LF356 gain of 250 op-amp amplifier

Steps 3, 4, 5, 6: Op-amp current noise i_{n+}

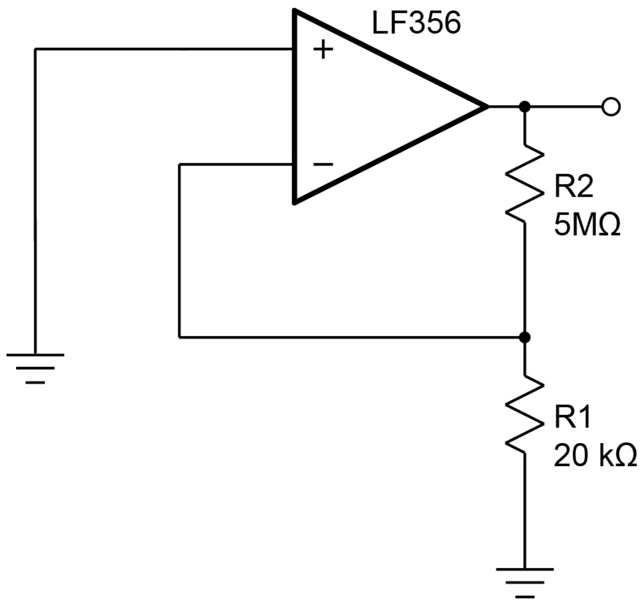


Steps 3, 4, 5, 6: Op-amp current noise i_{n-}

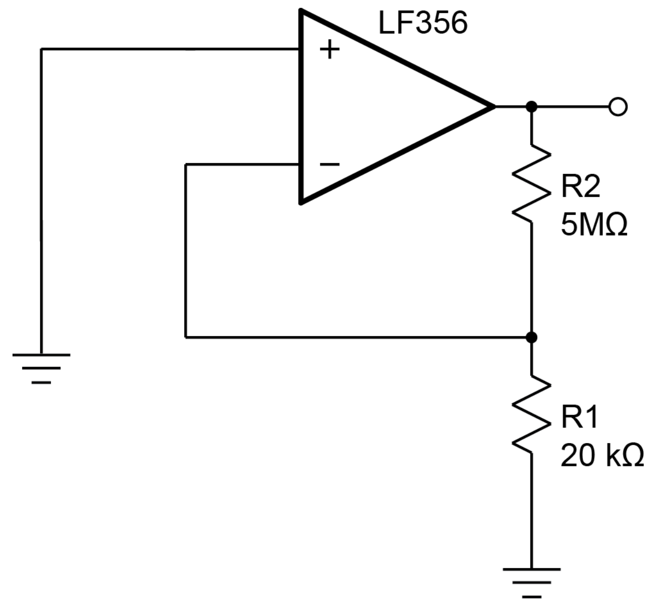


Example: Noise of LF356 gain of 250 op-amp amplifier

Steps 3, 4, 5, 6: Thermal noise R1



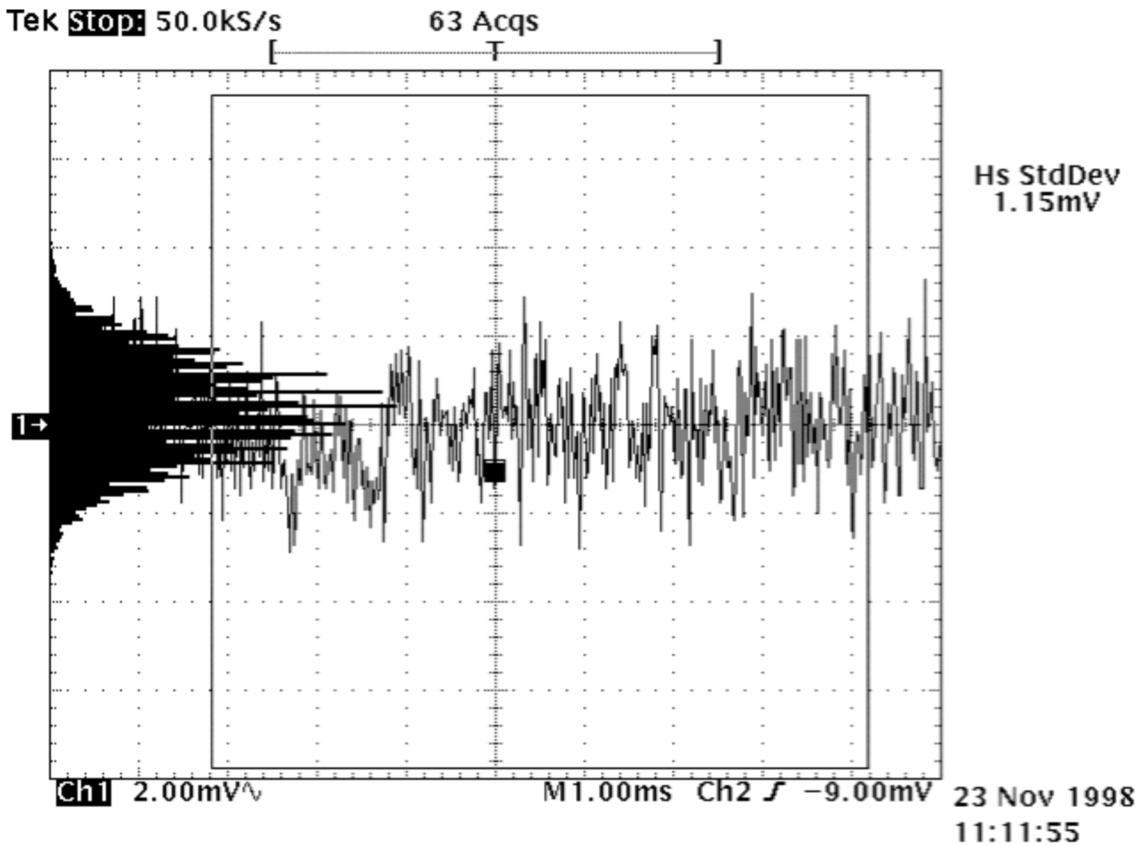
Steps 3, 4, 5, 6: Thermal noise R2



Step 7: Add contributions of noise sources

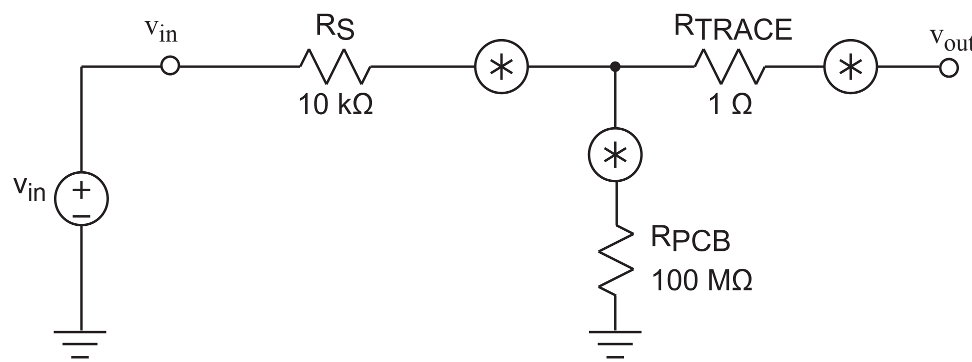
Noise Source	Density at Source	Gain to Output	Density at Output	Integrated over Noise Bandwidth
e_n				
i_{n+}				
i_{n-}				
R_1				
R_2				

Comparison with Measured Result

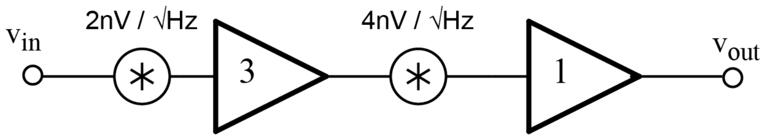


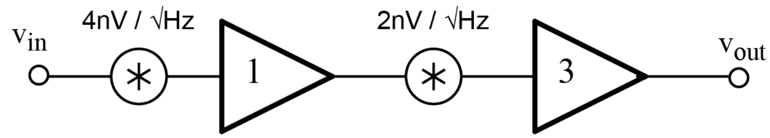
Noise: Do I need to worry about everything?!?

One Minute Quiz: Find effect of each resistor thermal noise source at v_{out}.



Note: Order matters for noise!





Noise Summary

Thermal noise is present in all elements that contains resistance.
A reactance does not generate thermal noise.
The thermal noise in any connection of passive elements is equal to the thermal noise that would be generated in a resistance equal to the real part of the equivalent network impedance.
Shot noise is produced by current flow across a potential barrier.
Noise having equal power in each unit of bandwidth (such as thermal and shot noise) is refered to as white noise.
1/f noise has equal power per decade (or octave) of bandwidth.
Contact noise (1/f noise) is present whenever current flows through a nonhomogeneous material.
1/f noise is only a problem at low frequencies.
The noise bandwidth is greater than the 3-dB bandwidth.
As the number of poles (time constants) increase, the noise bandwidth approaches the 3-dB bandwidth.
The crest factor for thermal noise is normally assumed to be three to four.
Uncorrelated noise voltages add on a power basis; therefore

$$V_{\text{total}} = \sqrt{V_1^2 + V_2^2 + \cdots V_m^2}.$$

For best noise performance a low-source resistance should be used (assuming the source voltage remains constant).
In cascaded gain stage systems, changing the order of stages can change noise performance
If the gain of the first stage of a system is high, the total system noise is determined by the noise of the first stage.
Noise performance of functional blocks such as op-amps can be specified as an equivalent input noise voltage V_n and input noise current I_n.