

## 34 Interference

Consider the following case histories:

- A certain car engine control system system quits while close to a radio transmission tower.
- A development engineer finds that her signal acquisition system is swamped in noise from the on-board microprocessor.
- An airborne computer control system works fine on the bench when the modules are spread out. When the modules are installed in the case, noise from the switching power supply appears on the video output.
- An industrial automation control system is being produced as a commercial product. When tested for compliance with emission regulations, it fails the test. The unit must be redesigned to incorporate filters on the connecting cables.
- An aerial photography camera is installed in an aircraft, powered from the 28VDC bus. Whenever the camera triggers, the aircraft navigation computer resets.
- A film projection system is equipped with a magnetic head to read the sound track on magnetic tape. A nearby motor induces a power line interference signal in the magnetic head. The noise signal is stronger than the desired signal.

These are examples of *interference* – an electronic signal is somehow corrupted by another electrical signal. The interfering signal may be known as *EMI: ElectroMagnetic Interference*. We say that the *signal integrity* is compromised by an interfering signal. The interference may show up as a degradation of some signal-noise ratio, such as background television sync buzz on an audio signal. Or it may be severe enough to cause the system to fail.

As well, the allowable level of emissions from an electronic device are regulated, and must meet the requirements of *EMC: ElectroMagnetic Compliance*.

In other words, it may be necessary to protect an electronic circuit from some external interference signal, or it may be necessary to reduce the interfering signal that the electronic circuit is producing.

In all of the cases cited above, some signal interferes with the proper operation of the device. The precise mechanism is different in each case. However, it is possible to approach each problem with the same general understanding. Simple tests and measurements identify the specific interference mechanism and then specific remedies can be applied to the design. In some cases, there may be more than one mechanism of interference at work, and so it is important to be systematic.

### 34.1 Scope of the Problem

Interference problems are growing more common and severe, for the following reasons:

- Devices progress to smaller sizes, so different sections of the system are in closer proximity. For example, in a cellular telephone the RF transmitter is in close proximity to a microprocessor. Their signals must be kept separate.
- In the days of large equipment and vacuum tubes, interference problems could be solved by using shielding<sup>262</sup> in the form of enclosed metal boxes to separate the interfering circuits. This approach is less useful because of modern demands in size, weight and cost. Shielding, if it is used, must be part of the printed circuit board. The outer case is likely to be plastic, and so is no help as a shield.

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<sup>262</sup>In this context, a *shield* is a physical barrier to some interfering signal. A shield is a metal plate or enclosure that isolates two electric, magnetic or electromagnetic areas in space.

- Many circuits include digital and analog circuitry in same unit. For example, a data acquisition system will typically include an analog front end that conditions a sensor signal and presents it to an A/D converter. The output from the A/D converter is directed to a microprocessor for further processing. The digital signals of the microprocessor must be kept separate from low-level signals in the analog front end.
- Electronic devices are ubiquitous, and so there are more opportunities for interference between different products.
- The clock speeds of digital circuits are extremely high - in the hundreds of MHz. Digital waveforms at these frequencies trigger parasitic component effects, such as stray inductance. Short connecting leads and small patches of a circuit board may become antennas.

Curing these problems can be a challenge, because:

1. **Very small amounts of coupling between a source and receiver of interference may introduce unacceptable levels of interference.** For example, an audio system may require a signal-noise ratio of 80db which is a factor of 10,000 volts/volt. If the reference level is 1 volt, then interfering signals must be kept below  $100\mu\text{V}$ . In a metropolitan area, a short length of wire will pick up radio and television signals that exceed this by a factor of 1000. Consequently, the defenses must attenuate these undesired signals by the same factor.

Another example: depending on the frequency, an unbalanced current of only  $5\mu\text{A}$  in a cable may be sufficient to cause unacceptable radiated electromagnetic interference [281].

2. **Schematic symbols of components do not accurately reflect their true behaviour.** Schematic symbols have a dual meaning: they represent the ideal component and simultaneously represent the real device. For example, an ideal voltage source maintains its output voltage regardless of the output current while a real voltage source has internal resistance. When the output current of the voltage source current is small, the voltage drop across the internal resistance can be ignored. However, that same internal resistance can cause coupling between a source and receiver of interference.

Similarly, capacitors have internal inductance and resistance. These unmarked, extra behaviours are modelled by *parasitic* components: resistors and inductors that appear in series and parallel with an idealized capacitor.

In the interest of simplification, we can often ignore these parasitics. However, when interference is involved - and particularly at high frequency where inductive reactance due to wire length and stray capacitance becomes significant - the parasitics must be taken into account.

3. **A schematic diagram does not reflect the wiring layout.** Many interference mechanisms are a consequence of the routing of wiring and the placement of components. It is necessary to combine an understanding of the schematic and the wiring layout to identify the important parasitic components and coupling mechanisms.

To troubleshoot interference and EMI problems, the engineer must understand the schematic as it is drawn, and must also understand the effect of the parasitic components that are not shown but may have an important effect. The schematic with its parasitic components has been termed *the hidden schematic* [282].

### 34.2 The Mechanism of Interference

A generalized view of interference is shown in figure 890.

There is a *source* for the interfering signal. The signal is conveyed by some *medium* of transmission to the *receiver circuit*<sup>263</sup>.

For example, in the car engine control problem described on page 950:

- The source of interference is a strong radio transmitter.
- The medium of transmission is an airborne radio signal which couples to car wiring. The car wiring - or possibly the metal structure of the car - behaves as a receiving antenna.
- The receiver is the engine control system electronics.

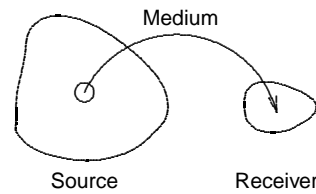


Figure 890: Interference in General

In this generalized model, the interference may be minimized by **eliminating the source, blocking the medium, or by making the receiver resistant to the interference**. In difficult cases, it may be necessary to some combination or all three.

In this particular example, it is not feasible to eliminate the source, so the focus must be on blocking the medium and hardening the receiver.

### 34.3 The Source

In dealing with a problem of interference, it is critical to identify the components in figure 890.

The identification of the source is usually based on the concept of *correlation*. For example, every time the soldering iron switches the radio produces a popping noise. These two events are then said to be correlated. Correlation does not necessarily indicate causality, but it's an important clue to investigate further. For example, if the soldering iron is moved away from the radio and the popping noise decreases in severity, that is a second correlation that points at the same suspect.

As another example, suppose a switching power supply is suspected as being the source of interference in an analog amplifier. Synchronize an oscilloscope on the power supply switching waveform and see if the amplifier noise display is stable. If that's the case, then the two signals are correlated. If there is still some question, operate the circuit from a separate lab power supply and see if the problem persists.

### 34.4 The Coupling Mechanism

Having identified the source, the next step is to determine the coupling mechanism or *medium of transmission*. The usual suspects for coupling are:

- *conductive* (via a power supply or ground path, for example)
- *electric-field*
- *magnetic-field*
- *electro-magnetic field* (radio wave)

Now we'll look at the details of each of these mechanisms.

<sup>263</sup>Some authors refer to the source and the receiver as the *aggressor* and *victim* circuits.

### 34.5 Conductive Coupling

In this case, the source and the receiver share a common impedance. An example audio circuit is shown in figure 891. A preamplifier stage is represented by amplifier  $A$ . A power amplifier stage is represented by the switch and load resistance  $R_L$ .

In figure 891(a), both the amplifier and preamplifier are connected back to the power supply  $E$  by the same wiring. The wiring has resistance, represented by  $R_{1A}$  through  $R_{2B}$ .

The load resistance  $R_L$  is such that significant current flows through it when the switch is closed. This current is supplied via resistor  $R_{1A}$  and returned via resistor  $R_{1B}$ . As a consequence, the voltage  $V_{xy}$  changes when the amplifier draws more or less current.

This change in voltage is communicated to the preamplifier supply terminals. If the preamp has insufficient power supply rejection ratio (section 21.4), then this variation in voltage will find its way into the output signal of the preamp.

A particularly insidious result of this arrangement is the possibility of low-frequency oscillation, called *motorboating*. The preamplifier, power amplifier, and common supply impedance form a loop. The coupling and bypass capacitances provide various phase shifts which result in positive feedback at some low frequency, and the entire arrangement becomes unstable. In this case, we could view the interfering signal as being generated at the power amplifier, coupled by the common power supply impedance, with the preamplifier as the receiver of the interference. Additional capacitance across the supply at  $xy$  will not cure this problem - it simply moves the oscillation to a lower frequency.

Identification of this coupling mechanism may simply be a matter of recognizing that the same signal appears at different points in the circuit, for example on the power supply line. (The measurement technique must be such as to isolate conductive effects from other effects.)

In this particular case, the cure is shown in figure 891(b). The preamplifier power supply lines are moved from the power amplifier connection to the power supply, where the source impedance is low. This eliminates power supply fluctuations caused by the interaction of the power amplifier current and power supply wiring. The power supply resistance wiring of the preamp  $R_{2A}$ ,  $R_{2B}$  has little effect because it is not carrying significant current. In effect, this change blocks the interference coupling mechanism.

Alternatively, the preamplifier could be equipped with a power supply regulator, thereby making it more resistant to variations in power supply voltage.

In this particular example, the coupling impedance is a resistance. It is quite common for the coupling impedance to be an inductance, in which case coupling occurs at a frequency where the impedance is significant. The previous example is for an audio circuit, but the concept applies equally well to circuits which contain digital switching circuitry and sensitive analog circuitry.

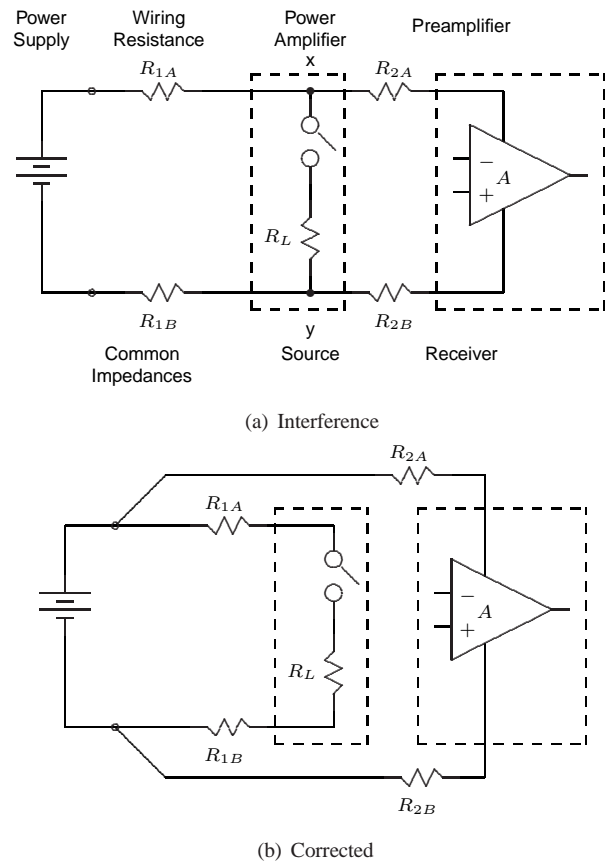


Figure 891: Example, Common Impedance

**Reduce conductive coupling between circuits by providing separate power supply and ground conductors back to the power supply.** In the case of a mixed digital-analog circuit, provide separate analog and digital supply and ground lines. The analog and digital supply meet at the power supply output. The analog and digital ground meet at the power supply common point.

For this to be effective, there must be no other connection points between the analog and digital systems.

### 34.5.1 Implications for PC Board Layout

When laying out a circuit board, if the analog power line and digital power line connect together at the power supply, then a PCBoard autorouter program will think that they're the same net. It will then mindlessly intersperse digital and analog power supply connections, exactly what is not wanted. The analog supply line and digital supply line must be kept separated and given distinct names, such as +5A, +5D. The autorouter then brings the leads back to the power supply area, and a removable jumper wire connects them together. Similarly, the grounds need to be separated and given distinct names, such as AGND, DGND, and brought back to jumpers<sup>264</sup>(figure 892).

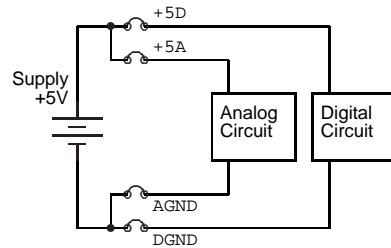


Figure 892: Common Point Grounding

This particular strategy is suitable for circuits where the digital signal edge rates are modest and electromagnetic radiation is not an issue. However, it's not a cure-all. A different strategy, using a ground plane, may be more effective in other situations.

## 34.6 Electric Field Coupling

A time-varying voltage creates a time-varying electric field that will couple into a receiver circuit via any stray capacitance, as indicated in figure 893(a). The degree of coupling increases with the capacitance, which is a function of the surface areas of the source and receiver and the spacing between them. For example, when the two plates of the capacitor have area  $A$  and spacing  $S$ , then the capacitance is

$$C = \frac{\epsilon A}{S} \quad (1516)$$

where  $\epsilon$  is a constant, the permittivity of free space. Then the capacitance and the coupling can be reduced by reducing the areas of the two plates and by increasing the spacing between them.

The coupling circuit is effectively a high-pass RC circuit, for which the Bode plot is given in figure 893(b). As shown in this figure, the coupling from the interfering source  $e$  into the receiver circuit is reduced by increasing the highpass corner frequency. In the figure, the original corner frequency is  $\omega_1$  and the signal at frequency  $\omega$  is attenuated to point  $A$ . If the corner frequency is moved up to  $\omega_2$ , then the attenuation is moved down to point  $B$ , that is, there is less coupling between the source and receiver circuit. The corner frequency is given by

$$\omega_c = \frac{1}{R_{circuit} C_{stray}} \quad (1517)$$

so the corner frequency may be increased by decreasing either the capacitance or resistance. That is, **low impedance circuits are less susceptible to electric-field interference.**

<sup>264</sup>This has other advantages: at a later date, you can separate analog and digital power in the event that they seem to be interfering and must be mutually filtered. The jumper can be used as a test point for board power supply voltage.

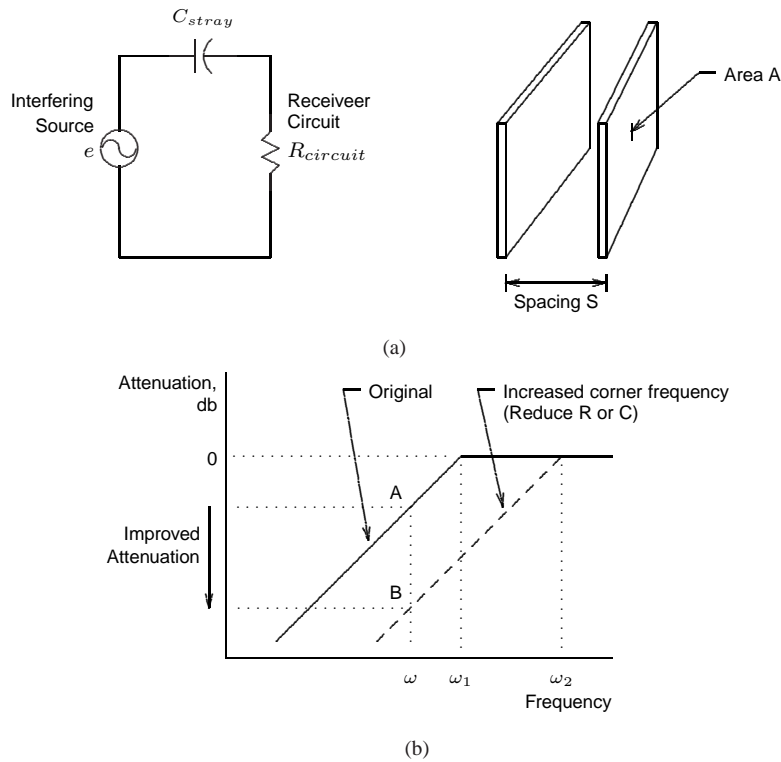


Figure 893: Electric Field Coupling

**34.6.1 Electric-Field Detection**

Electric-field detection is straightforward. The probe is a length of wire, or a flat plate that is coupled to the display device - oscilloscope or spectrum analyser. The source and the probe form the two terminals of the coupling capacitance. The strength of the detected signal depends on the spacing, but not the orientation, of the probe.

**34.6.2 Electric-Field Shielding**

If as shown in figure 894 a metallic plate is interposed between the two plates of a capacitor and this plate is connected to the circuit common, then the electric field stops at the plate. In effect, the shield becomes the second plate of the capacitor and coupled current is short-circuited to ground.

Moreover, an electric field cannot penetrate inside a metallic enclosure. Consequently, an electric-field shield is easily constructed - a thin metal foil is sufficient.

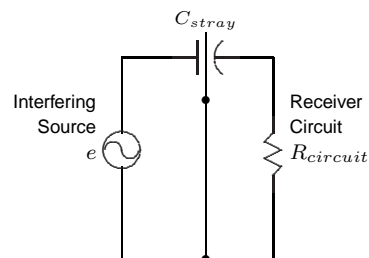


Figure 894: Electric Field Shield

### 34.6.3 Demonstration: Electric Field Pickup

In this case, as shown in figure 895(a), the function generator drives a conductive aluminum plate, approximately 19cm by 19cm. The electric field probe is a stiff piece of wire, about 5cm total length, spaced about 0.5 cm above the aluminum plate. The probe plugs into the end of a BNC cable. The cable leads to an oscilloscope with an input resistance-capacitance of  $1\text{M}\Omega$  in parallel with  $47\text{pF}$ . The generator output is 8 volts p-p sine wave at a frequency of 1MHz.

The measured signal on the oscilloscope is  $100\text{mVp-p}$ . The plate and probe are two electrodes in a 'stray' capacitance. The signal couples through this stray capacitance to the oscilloscope.

We'll assume that the  $50\Omega$  internal resistance of the signal generator is much smaller than the impedance of  $C_{stray}$ , so it can be ignored. The  $1\text{M}\Omega$  input resistance of the oscilloscope is much larger than the impedance of  $C_{in}$  of the oscilloscope, so it too may be ignored. Then the equivalent circuit is as shown in figure 895(b). The stray capacitance and the input capacitance of the oscilloscope form a capacitive voltage divider.

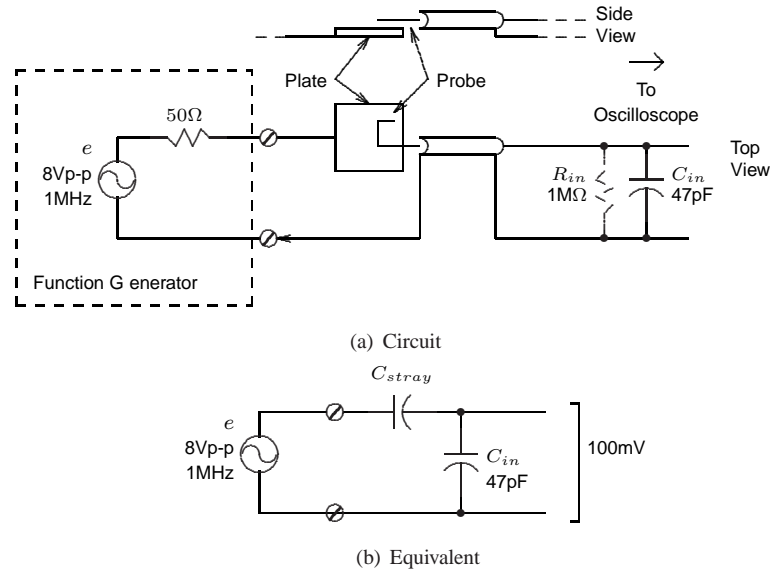


Figure 895: Electric Field Pickup

$$\frac{e_o}{e} = \frac{X_{C_{in}}}{X_{C_{stray}} + X_{C_{in}}} \quad (1518)$$

$$= \frac{1/\omega C_{in}}{1/\omega C_{stray} + 1/\omega C_{in}} \quad (1519)$$

$$= \frac{1}{C_{in}} \frac{C_{in} C_{stray}}{C_{in} + C_{stray}} \quad (1520)$$

Since  $C_{in}$  is much greater than  $C_{stray}$ , we can simplify this to

$$\frac{e_o}{e} \approx \frac{C_{stray}}{C_{in}} \quad (1521)$$

Plugging values for  $e$ ,  $e_o$  and  $C_{in}$  into equation 1521 puts the stray capacitance  $C_{stray}$  at  $0.59\text{pF}$ .

Notice that there is no frequency term in equation 1521, so this divider affects all frequencies equally. A complex waveform such as a square wave will be undistorted passing through this divider. That is confirmed by measurement: when the generator is switched to square and triangle waveforms, the oscilloscope shows the same waveform shape as the generator.

### 34.6.4 Demonstration: Reducing Electric Field Pickup by Lowering Impedance

If the oscilloscope input is shunted by a  $50\Omega$  resistor, the effect is dramatic: the detected signal drops from  $100\text{mV}$  to about  $3\text{mV}$ , an attenuation of

$$\begin{aligned} 20 \log_{10} \frac{V_1}{V_2} &= 20 \log_{10} \frac{100}{3} \\ &= 30.4 \text{ db} \end{aligned}$$

Furthermore, the detected waveform is a differentiated version of the signal-generator waveform. An triangle signal shows becomes a square wave, a square wave becomes alternating positive and negative spikes.

The new equivalent circuit is shown in figure 896. The input capacitance of the oscilloscope is  $47\text{pF}$ , which has an impedance of about  $3.3\text{k}\Omega$  at  $1\text{MHz}$ . Consequently, it's large enough to be ignored compared to  $50\Omega$ . This circuit is an RC highpass network, with a corner frequency around  $6\text{GHz}$ . Frequencies below this will be attenuated by a greater or lesser extent. For waveforms well below the corner frequency, the network will act as a differentiator.

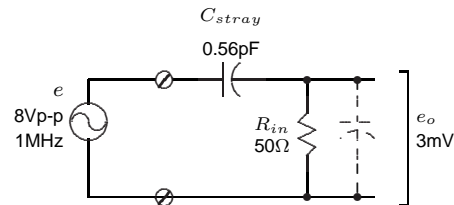


Figure 896: Equivalent Circuit

### 34.6.5 Demonstration: Reducing Electric Field Pickup by Shielding

If a grounded metallic sheet is placed between the plate and probe in figure 895, then the equivalent circuit of figure 894. The electric-field coupling into the probe should drop significantly. For a flat shield approximately the same size as the plate in figure 895, the signal dropped by a factor of 16, some  $24\text{db}$ . Enclosing the probe in a sealed metal container would undoubtedly improve the shielding effect.

**Electric field pickup may be reduced by lowering the impedance of the receiver circuit or placing a grounded metallic shield between the source and the receiver.**

## 34.7 Magnetic Field Coupling

When a wiring *loop* is driven by an alternating current, it creates a time-varying magnetic field. If that AC magnetic field passes through a second loop, a time-varying voltage will be induced in the second loop. For example, the magnetic field might originate in the filter inductor of a switching power supply and be received in a circuit loop that is nearby wiring.

A little theoretical investigation will give us some insight into this effect.

The situation is shown in figure 897. A transmitter coil (shown as one turn here, which could be multiple turns) is driven by an AC source and carries an AC current  $i$ .

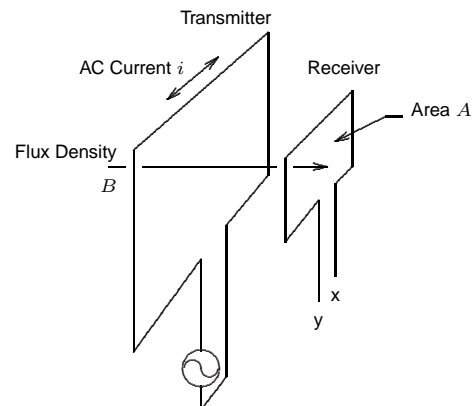


Figure 897: Magnetic Coupling



This current generates a magnetic field:

$$B = \mu_o H \quad (1522)$$

where the variables are

$B$  Flux Density, webers/metre<sup>2</sup>, from the transmitting coil

$\mu_o$  magnetic permeability of space, a constant

$H$  magnetic field intensity, amp-turns.

As well, we can write

$$H = Ni \quad (1523)$$

where

$N$  is the number of turns on the coil, (in this case,  $N=1$ )

$i(t)$  is the current in the coil

Putting  $N = 1$  and combining these two equations, we have:

$$B = \mu_o i(t) \quad (1524)$$

as the time-varying magnetic field intensity in the region of the transmitting coil.

This same field passes through the smaller receiving coil. According to Lenz's law:

$$e_{xy} = N \frac{d\phi}{dt} \quad (1525)$$

where

$e_{xy}$  is the voltage created at the terminals of the coil

$\phi$  is the total flux flowing through the receiver coil

$N$  is the number of turns on the receiver coil (again, one turn)

The flux in the receiver coil is equal to the transmitter magnetic field times the receiver coil area:

$$\phi = BA \quad (1526)$$

Substituting for  $B$  from equation 1524 into equation 1526, and then for  $\phi$  into equation 1525, we have that the receiver terminal voltage is:

$$\begin{aligned} e_{xy} &= A \frac{d}{dt} \mu_o i(t) \\ &= A \mu_o \frac{di(t)}{dt} \end{aligned} \quad (1527)$$

For example, if  $i(t)$  is a sine wave of radian frequency  $\omega$  and peak value  $I$ :

$$\begin{aligned} e_{xy} &= A \mu_o \frac{dI \sin \omega t}{dt} \\ &= A \mu_o \omega (I \cos \omega t) \end{aligned} \quad (1528)$$

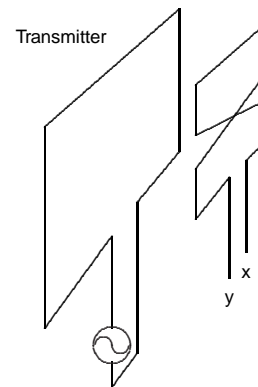


Figure 898: Coil Cancellation

Most real-world configurations are much more complicated and so numerical results and predictions are not as simple as this formula would indicate. However, there are a number of implications of the mathematics that are useful guidelines:

- The generated voltage is proportional to the area of the receiver coil. Pickup can be minimized by reducing this area.
- In the example shown, the receiver coil is smaller than the transmitter coil, so the receiver coil is in a uniform magnetic field and the voltage is proportional to the area of the receiver coil. If the coil sizes are reversed so that the transmitter is smaller than the receiver, then this will reduce the flux flowing through the receiver coil just as if the receiver coil area had been reduced. Consequently, reducing either the transmitter or receiver coil area reduces the induced voltage.
- The induced voltage is proportional to frequency. This coupling is likely to be more of a problem at higher frequencies.
- In the diagram, the coils are parallel. If the coils are at an angle to each other, the induced voltage becomes proportional to the cosine of the relative angle. At  $90^\circ$ , where the transmitter coil and receiver coil are at right angles, the induced voltage is zero. This is a convenient test to determine whether pickup is from an electric or magnetic field. Rotation of the probe will have no effect on electric field pickup.
- If the receiver coil is rotated through  $180^\circ$  with respect to the transmitter coil, the cosine effect continues to hold and the voltage across the terminals reverses polarity. For example, if the the two-loop arrangement in figure 898 is illuminated with a uniform magnetic field, the generated voltages in the two loops will cancel and  $V_{xy}$  will be zero.
- This analysis assumes that there is close proximity of the two coils. The field of the transmitter coil falls off with distance, so the pickup can be reduced by separating the two coils.
- The coupling between the two coils will be reduced if the magnetic field is somehow redirected away from the receiver coil. That is the function of magnetic *shielding*, as discussed below.

### 34.7.1 Magnetic Field Identification

A small coil of wire, connected to an oscilloscope or spectrum analyser, may be used as a *search coil* (probe) to detect magnetic fields. Ideally, the probe is surrounded by an electrostatic shield (with a slot in it) so that electric fields do not generate a signal in the probe [283] The concept is shown in figure 900(a).

However, that is not essential: Smith in [284] describes an unshielded magnetic field probe using a bent paper clip, figure 900(b). The coil is usually terminated through a series  $50\Omega$  resistor at the coil or a  $50\Omega$  resistive input at the oscilloscope or spectrum analyser to avoid reflections in the connecting coaxial cable.

### 34.7.2 Voltage in the Conductor

If the magnetic probe senses the flux due to current in a conductor, then the shape of detected voltage waveform will be the same as the voltage across that segment of the conductor. This follows from equation 1527, where the

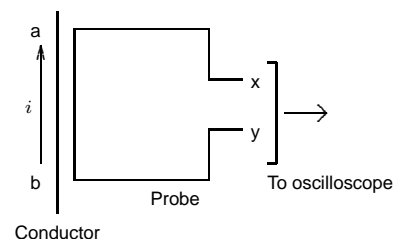


Figure 899: Measuring Voltage in a Conductor

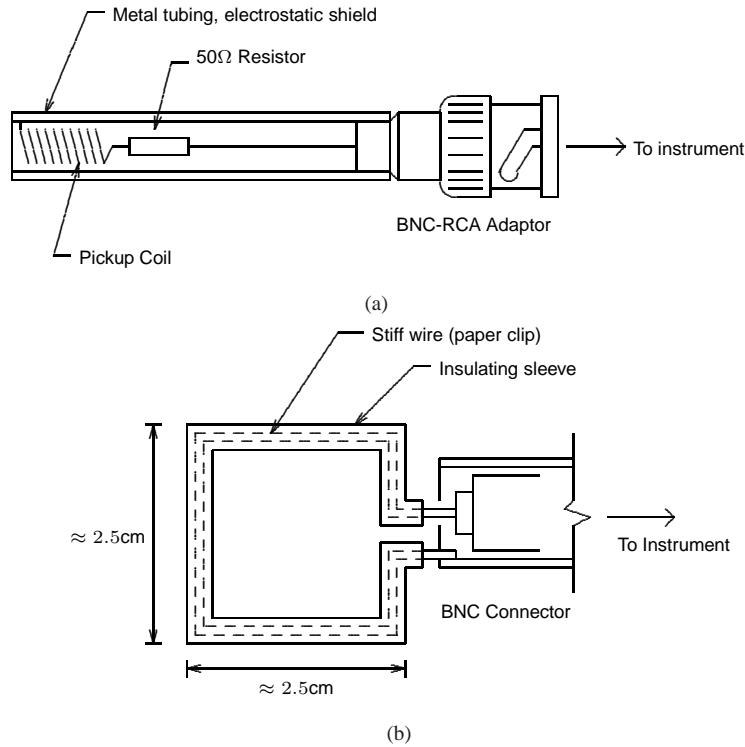


Figure 900: Magnetic Field Probes

probe voltage  $e_{xy}$  is shown to be proportional to the rate of change of the current  $di(t)/dt$  in the probed conductor. If that section  $ab$  of the conductor has an inductance  $L_{ab}$ , then the voltage across it will be

$$\begin{aligned}
 e_{ab} &= L_{ab} \frac{di(t)}{dt} \\
 &\propto e_{xy}
 \end{aligned}
 \tag{1529}$$

where

$e_{ab}$  is the voltage between a and b on the current-carrying conductor

$L_{ab}$  is the inductance of the conductor between a and b

$e_{xy}$  is the open-circuit output voltage of the magnetic probe

If there is perfect magnetic coupling between the conducting wire and the probe, then the magnitude of the two voltages would be the same:  $e_{ab} = e_{xy}$ . This is not likely to be the case, so the value of the probe voltage sets an upper limit on the voltage across the conductor.

The current-carrying wire and the probe constitute a transformer. If the flux of the wire couples perfectly to the probe, then they form a 1:1 transformer. Then voltage across the two parallel wires are experience the same rate of change of flux and the same voltage. In general, there will be some flux leakage and the voltage across the secondary (the probe) will be somewhat less than the voltage in the current conductor.

### 34.7.3 Demonstration: Magnetic Field Pickup

Figure 901(a) shows an arrangement to demonstrate magnetic field pickup. A stiff wire is bent into a square-shaped probe and then plugged into one end of a coaxial cable. The other end goes to the input of an oscilloscope. A signal generator drives current into a separate length of wire to create a magnetic field.

In this example, the probe is approximately 3.5cm on a side. The length of wire is about 35cm long. As the probe is rotated, the detected voltage decreases, going through a null when the probe and wire are at right angles. Since the detected voltage changes with the orientation of the probe, the pickup is magnetic, rather than electric.

The voltage across the probe terminals will be equal to or less than the voltage in a 3.5cm segment of the conductor [285]. In the measurement setup of figure 901(a), the current-carrying conductor was 35cm. The 3.5cm probe produced 20mVp-p, implying a total voltage across the conductor of 200mV. It actually measured some 300mV because the measurement sets a lower bound on the voltage.

The electrical equivalent circuit for the probe is shown in figure 901(b). The inductance of the transformer secondary forms a high-pass filter with the input resistance  $R_i$  of the oscilloscope. When the oscilloscope cable is unterminated, then the effective load resistance  $R_i$  for the transformer is  $1M\Omega$ . Then the 1MHz operating frequency is above the cutoff frequency of the high-pass filter and the detected waveform has the same shape as the signal generator waveform.

When the cable is terminated with  $50\Omega$ , the operating frequency is below the cutoff frequency of the highpass filter. The detected waveform is a differentiated version of the signal generator waveform.

### 34.7.4 Magnetic Field Shielding

Electric-field shielding was discussed in section 34.6.2 on page 955. A thin conductive sheet is effective in blocking an electric field.

Magnetic field shielding is not so straightforward. A ferrous material such as iron functions as a magnetic field shield by providing a high-permeability path for the field lines. Non-ferrous materials such as copper aluminum function as a shield for an AC magnetic field by inducing circulating currents in the material that create an opposing magnetic field. The shielding effectiveness increases with frequency (see reference [286], Part 5). However, non-ferrous materials are not an effective shield for DC or low-frequency magnetic fields. For example, shielding for a 60Hz line-current generated field often requires a special magnetic alloy, commonly referred to as *mu-metal*. This approach tends to be complicated and expensive, and it's usually more cost-effective to shape the magnetic field. For example, changing from an E-core transformer to a toroidal power transformer may reduce the offending field, since a magnetic toroid has very little external magnetic field.

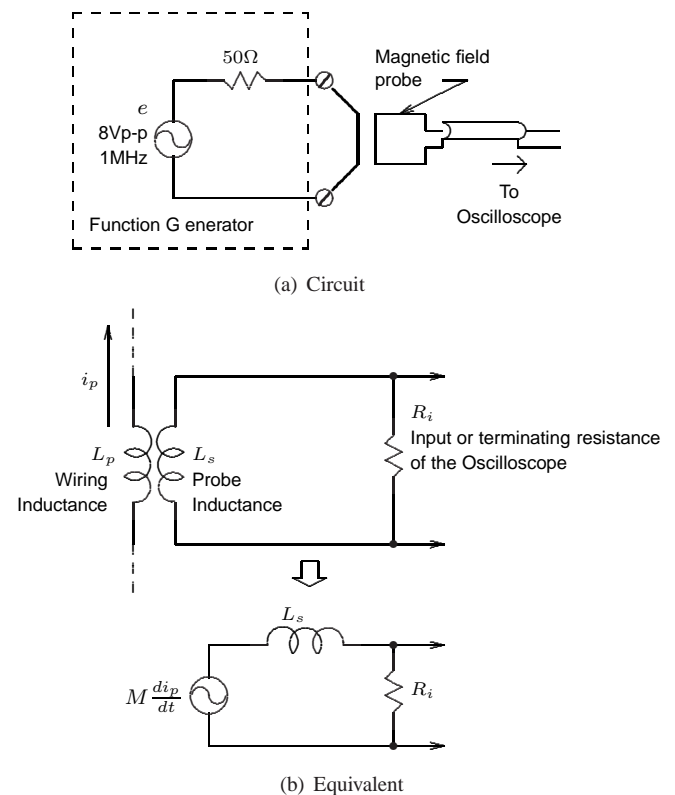


Figure 901: Magnetic Field Pickup

It can be shown that the external magnetic field decreases inside the shielding material to  $1/e$  of its original value in a distance equal to one *skin depth*, where skin depth is given by [287]:

$$\delta = \sqrt{\frac{2}{\omega\mu\sigma}} \quad (1530)$$

where the variables are

$\delta$	Skin depth, metres
$\omega$	Frequency of the magnetic field, radians/sec
$\mu$	Magnetic permeability of the shielding material, given by $\mu = \mu_r\mu_o$
$\mu_r$	Relative permeability, a dimensionless constant, the ratio of the material permeability to free space
$\mu_o$	Magnetic permeability of space, a constant, $4\pi \times 10^{-7}$
$\sigma$	Conductivity of the shield material, $\text{U}/\text{metre}$ given by $\sigma = 1/\rho$ , Siemens/metre
$\rho$	Resistivity of the shield material, $\Omega\text{-metre}$

### Example

A switching power supply, operating at 100kHz, is generating a magnetic field that couples into nearby wiring. What would be the shielding effectiveness of 0.050 inch thick aluminum in blocking this field?

The resistivity of aluminum is found in a table of Properties of Materials to be  $2.6 \mu\Omega\text{-cm}$ .

### Solution

First, convert the resistivity to MKS units:

$$\begin{aligned} \rho &= 2.6 \mu\Omega\text{-cm} \\ &= 2.6 \times 10^{-6} \Omega\text{-cm} \times \frac{1 \text{ M}}{100 \text{ cm}} \\ &= 2.6 \times 10^{-8} \Omega\text{-M} \end{aligned}$$

The conductivity is the reciprocal of the resistivity:

$$\begin{aligned} \sigma &= \frac{1}{\rho} \\ &= \frac{1}{2.6 \times 10^{-8} \Omega\text{-M}} \\ &= 38 \times 10^6 \text{ Siemens/metre} \end{aligned}$$

Since aluminum is not a ferromagnetic material (ie, cannot be magnetized), its relative permeability is close to unity. Then the permeability is simply equal to the permeability of space:

$$\begin{aligned} \mu &= \mu_r\mu_o \\ &= 1 \times 4\pi \times 10^{-7} \\ &= 4\pi \times 10^{-7} \end{aligned}$$

Now we can use equation 1530 to calculate the skin-depth of the magnetic field:

$$\begin{aligned}
 \delta &= \sqrt{\frac{2}{\omega\mu\sigma}} \\
 &= \sqrt{\frac{2}{(2\pi \times 100 \times 10^3) \times (4\pi \times 10^{-7}) \times (38 \times 10^6)}} \\
 &= 2.58 \times 10^{-4} \text{ metres}
 \end{aligned}$$

Now we can relate the skin depth to the attenuation. We convert the material thickness to metres, determine the number of skin-depths in the material, and then calculate the attenuation. The material thickness  $T$  is:

$$\begin{aligned}
 T &= 0.050 \text{ inches} \\
 &= 0.050 \text{ inches} \times \frac{2.54 \text{ cm}}{1 \text{ inch}} \times \frac{1 \text{ metre}}{100 \text{ cm}} \\
 &= 1.27 \times 10^{-3} \text{ metres}
 \end{aligned}$$

Each skin depth contributes a factor of  $1/e$  attenuation, and the overall thickness of the material contains  $T/\delta$  skin depths so the overall attenuation factor  $K$  is:

$$\begin{aligned}
 K &= \left(\frac{1}{e}\right)^{T/\delta} \\
 &= \left(\frac{1}{2.71}\right)^{1.27 \times 10^{-3} / 2.58 \times 10^{-4}} \\
 &= 7.28 \times 10^{-3}
 \end{aligned}$$

That is, magnetic field inside the shield is approximately 100 times less than the field outside the shield. The voltage induced in a coil would reduce by the same factor.

In decibels, this factor is:

$$\begin{aligned}
 K_{db} &= 20 \log_{10} K \\
 &= 20 \log_{10} (7.28 \times 10^{-3}) \\
 &= 42.7 \text{ db}
 \end{aligned}$$

### 34.8 Spectrum of Pulse Waveform

Digital signals, such as those present in a switching power supply or microprocessor, are a frequent source of interference with other circuits. The spectrum of such a signal gives some important clues on minimizing this interference. Once the spectrum is known, then it is possible to design suitable filters to contain the digital noise.

#### 34.8.1 Square Wave Spectrum

A square wave is shown in figure 902(a). The formula for the spectrum of this type of square wave (one with zero rise and fall time) is given in [288] as:

$$C_n = 2A_{AV} \left| \frac{\sin(n\pi t_o/T)}{n\pi t_o/T} \right| \tag{1531}$$

where the variables are:

- $C_n$  amplitude of the nth component of the spectrum, volts
- $A_{AV}$  average value of the waveform, volts
- $n$  harmonic number (multiple of the fundamental)
- $t_o$  time for half the period, seconds
- $T$  period, seconds

The average value of the waveform is:

$$A_{AV} = A \left( \frac{t_o + t_r}{T} \right) \tag{1532}$$

An example spectrum given by equations 1531 and 1532 for a 1MHz square wave ( $T=1\mu\text{sec}$ ) of amplitude 5 volts ( $A=5$ ), is shown in figure 902(b). The vertical axis is expressed in decibels relative to  $1\mu\text{V}$ , that is:

$$\text{Amplitude, db} = 20 \log_{10} \left( \frac{C_n}{1 \times 10^{-6}} \right) \tag{1533}$$

Notice that the harmonics fall at the expected frequencies for a square wave: 1MHz, 3MHz, 5MHz and so on. The amplitude decreases at a rate of 20db/decade.

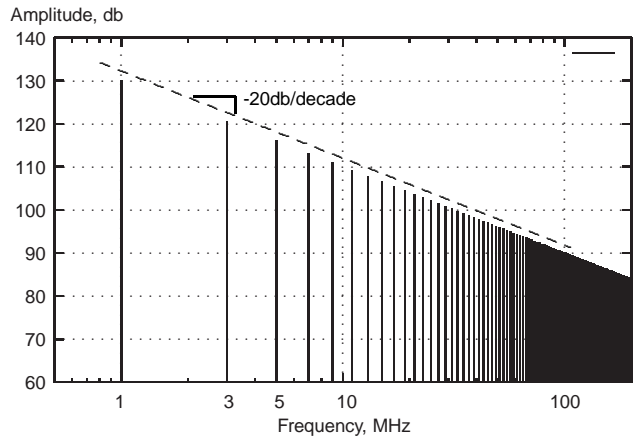
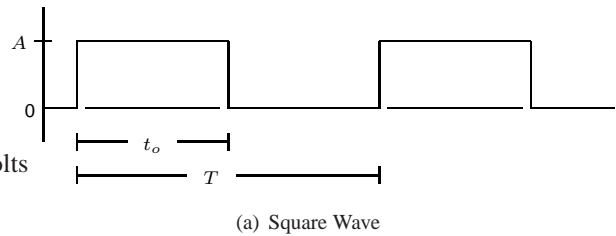


Figure 902: Square Wave Spectrum

**34.8.2 Trapezoidal Wave Spectrum**

Now consider the trapezoidal waveform shown in figure 903(a). This has significant rise time  $t_r$  and fall time  $t_f$ . Assuming that the rise and fall time are equal, the expression for the spectrum [288] is similar to equation 1531, but with an additional term:

$$C_n = 2A_{AV} \left| \frac{\sin(n\pi t_r/T)}{n\pi t_r/T} \frac{\sin(n\pi(t_o + t_r)/T)}{n\pi(t_o + t_r)/T} \right| \tag{1534}$$

where the additional variable  $t_r$  is the rise (and fall) time. The rise time is usually much smaller than the duration of the pulse, so equation 1534 may be simplified to:

$$C_n = 2A_{AV} \left| \frac{\sin(n\pi t_o/T)}{n\pi t_o/T} \frac{\sin(n\pi t_r/T)}{n\pi t_r/T} \right| \tag{1535}$$

where the average value of the waveform is given by equation 1532. Using equation 1535 and the same parameters as the previous square wave example, with a rise time  $t_r$  of 20nsec, the spectrum for the trapezoidal waveform is shown in figure 903(b).

Notice that the amplitude decreases at a rate of 20db/decade up to a certain frequency, and then more rapidly thereafter, at 40db/decade. Finite rise time in the time domain decreases the spectral energy at high frequencies in the frequency domain.

**34.8.3 The Bode Approximation**

In this section we'll look at a method for predicting the amplitude spectrum for a trapezoid waveform.

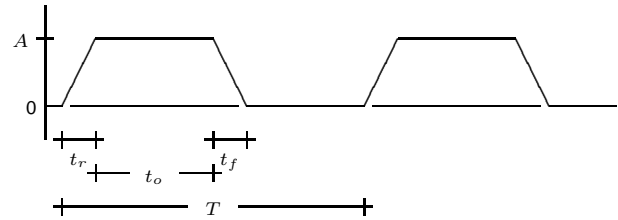
The spectrum of figure 903(b) can be approximated as shown in figure 903(c).

There are two break frequencies, which we'll call  $f_1$  and  $f_2$ . The lower break frequency is determined by the pulse width  $t_o$ ; the upper break frequency  $f_2$  is determined by the pulse rise (and fall) time  $t_r$ .

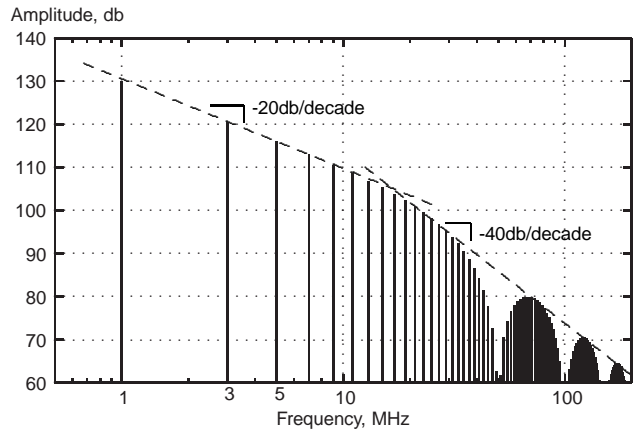
$$f_1 = \frac{1}{\pi t_o}, f_2 = \frac{1}{\pi t_r} \tag{1536}$$

Above  $f_2$ , the amplitude of the spectrum decreases rapidly, and so  $f_2$  is usually taken as the 'upper frequency limit' of the signal.

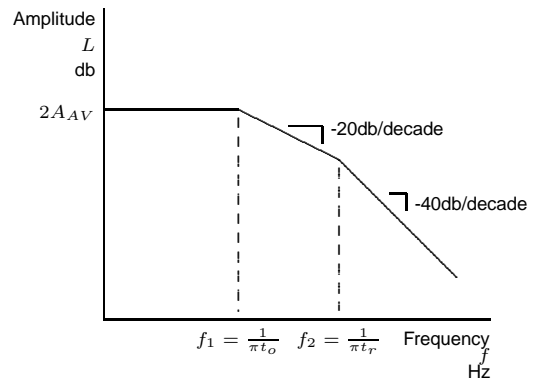
Filtering of noise and containment of electromagnetic radiation is simpler at lower frequencies, so it is desirable to limit this high frequency content. Consequently, **to minimize noise and interference problems digital circuitry should use logic and pulse waveforms with the slowest transition times that are acceptable.**



(a) Square Wave



(b) Trapezoidal Wave



(c) Trapezoidal Spectrum Approximation

Figure 903: Trapezoid Wave Spectra



**Example**

A certain pulse waveform is a square wave with a peak-peak amplitude of 5 volts, a frequency of 1MHz and rise/fall time of 20nsec. Calculate the spectrum amplitude and the two break frequencies.

**Solution**

The amplitude  $L$  is twice the average value  $A_{AV}$ . The duty cycle  $t_o/T$  is 50% or 0.5. Then using equation 1532:

$$\begin{aligned} L &= 2A_{AV} \\ &= 2A \frac{t_o}{T} \\ &= 2 \times 5 \times 0.5 \\ &= 5 \text{ volts} \end{aligned}$$

On a Bode plot this must be expressed in decibels relative to some reference. For example, for a  $1\mu\text{V}$  reference, we have:

$$\begin{aligned} L_{db\mu V} &= 20 \log \frac{L}{1\mu V} \\ &= 20 \log \frac{5}{1 \times 10^{-6}} \\ &= 133.9 \text{ db}\mu V \end{aligned}$$

The break frequencies are given by equation 1536:

$$\begin{aligned} f_1 &= \frac{1}{\pi t_o} \\ &= \frac{1}{\pi \times (0.5 \times 10^{-6})} \\ &= 637 \text{ kHz} \\ \\ f_2 &= \frac{1}{\pi t_r} \\ &= \frac{1}{\pi \times (20 \times 10^{-9})} \\ &= 15.9 \text{ MHz} \end{aligned}$$

Then the amplitude spectrum decreases rapidly above 15.9 MHz

**34.8.4 Derivation: Spectrum of Trapezoidal Waveform**

In this section, we'll derive the break frequency expressions for  $f_1$  and  $f_2$  given by equation 1536.

The amplitude spectrum for a trapezoid digital wave is given by equation 1535. From that starting point, the *envelope* of the trapezoid spectrum is determined by the expression:

$$C_f = 2A_{AV} |\text{sinc}(\pi t_o f) \text{sinc}(\pi t_r f)| \quad (1537)$$

where  $\text{sinc}(x) = \frac{\sin(x)}{x}$  and the discrete harmonics at frequencies  $n/T$  have been replaced by a continuously varying frequency  $f$ .

The sinc function is shown in figure 904, when both axes are plotted to logarithmic scales. For values of  $x$  less than 1, the value of the function is 1. For values of  $x$  more than 1, the function decreases by a factor of 10 when the value of  $x$  increases by a factor of 10. Consequently, the function is proportional to  $1/x$ . At the corner frequency, these two are equal, that is:

$$\begin{aligned} 1/x &= 1 & (1538) \\ x &= 1 \end{aligned}$$

The Bode approximation of a sinc function is thus a constant value of 1 over the range  $0 < x < 1$ , decreasing at 20db/decade for  $x > 1$ . The corner occurs at  $x = 1$ .

Now, in order to convert equation 1537 into a suitable format for a Bode plot, we take 20 times the  $\log_{10}$  of  $C_f$ :

$$\begin{aligned} 20 \log_{10} C_f &= 20 \log_{10} (2A_{AV} |\text{sinc}(\pi t_o f) \text{sinc}(\pi t_r f)|) & (1539) \\ &= 20 \log_{10} (2A_{AV}) + 20 \log_{10} \text{sinc}(\pi t_o f) + 20 \log_{10} \text{sinc}(\pi t_r f) \end{aligned}$$

This function has three parts: a constant value, followed by two sinc functions. Each of the sinc functions is a lowpass filter.

In the case of the first sinc function, the corner frequency  $f_1$  occurs when:

$$\begin{aligned} \pi t_o f_1 &= 1 & (1540) \\ f_1 &= \frac{1}{\pi t_o} \end{aligned}$$

Similarly

$$\begin{aligned} \pi t_r f_2 &= 1 & (1541) \\ f_2 &= \frac{1}{\pi t_r} \end{aligned}$$

Consequently, the frequency spectrum is as shown in figure 903 on page 965.

### 34.9 Components of The Hidden Schematic

In this section, we identify several components of the *hidden schematic*. These are electrical properties of components that are not normally indicated on a schematic diagram, but may be important in the functioning of the circuit.

For example, wires are generally assumed to have zero resistance. However, there are situations where wiring resistance is critical and must be considered. Here we identify some examples of those components and situations.

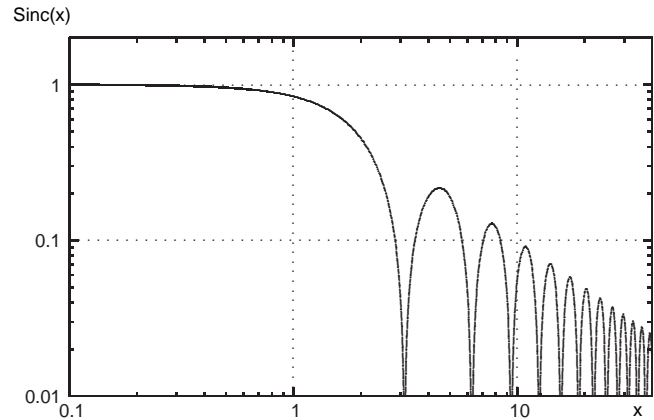


Figure 904: Sinc Function

### 34.9.1 Wiring Resistance

Connecting wire has resistance, which can be significant under certain circumstances. For example, the resistance of #22AWG solid wire is  $52.9\Omega$  per Km. An example where this is important is shown in figure 905(a).

In this example, a 2 amp capacitor-filter full-wave rectified power supply (section 15.2) is found to have 100 millivolt noise 'bumps' superimposed on the DC output voltage. These pulses find their way into the load circuitry as noise.

The noise pulses occur at a rate of 120 per second, twice the line frequency. Increasing the filter capacitance *increases* the amplitude of the noise pulses. What is the cause?

The schematic doesn't tell the whole story. The wiring, shown in figure 905(b), is more informative. The power supply has been wired so that the output is taken from the full-wave bridge, with separate connecting wires to the filter capacitor  $C$ . The wires  $a - a'$  connecting the capacitor to the bridge rectifier are #22AWG, 20cm in length, so each connecting wire has a resistance of:

$$\frac{52.9 \text{ ohms/km}}{1000 \text{ metres/km}} \times 0.20 \text{ metres} = 10.6 \times 10^{-3} \text{ ohms}$$

In a capacitor filtered power supply, the bridge rectifier supplies current to the filter capacitor in short pulses. (Figure 358(c) on page 370 shows an example of the sawtooth-shaped current waveform for a half-wave rectifier.). The average value of these current pulses is equal to the average output current. If the duration of the pulses is short, then the peak charging current into the capacitor is substantially larger than the output DC current. In this particular case, the peak current is 5 amps<sup>265</sup>, in which case the voltage developed across the capacitor connecting wires is

$$\begin{aligned} V &= 10.6 \times 10^{-3} \times 5 \\ &= 0.053 \text{ volts} \end{aligned}$$

There are two connecting wires, so the total voltage developed by the pulse of charging current is 100mV.

Increasing the filter capacitance  $C$  worsens the noise because it reduces the duty cycle of the charging pulses, increases the peak amplitude of the charging current pulse, and thereby increases the magnitude of the noise pulses.

The solution is simply to move the output connections to the terminals of the power supply. Assuming that the filter capacitor has low internal resistance, then the voltage *at its terminals* will be constant, and this is what should be taken to the load circuit.

Notice the steps in solving this particular puzzle:

<sup>265</sup>The power supply example of page 370 has a peak to average current ratio of 28, so 5 amps peak is not at all unlikely in a power supply with 2 amperes output current.

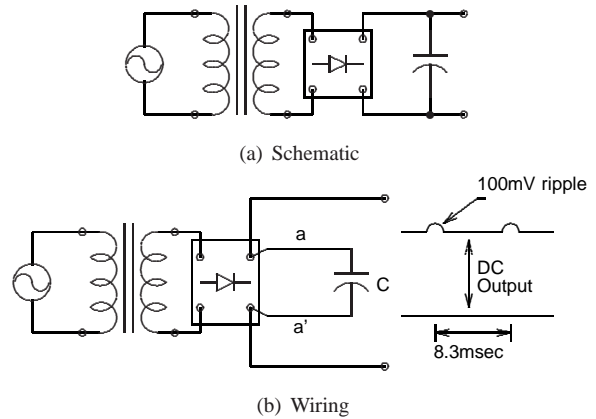


Figure 905: Power Supply Capacitive Filter

- Be aware of the actual *current* waveform in the circuit. An oscilloscope shows the voltage waveform and in this case the shape of the current waveform is quite different from the voltage.
- Treat connecting wires as having resistance.
- Consider the actual wiring arrangement of the circuit.

Incidentally, here is a trick to remembering approximate values of wiring resistance.

First, in English units:

- The resistance of #10AWG wire is about  $1\Omega$  per 1000 feet.
- Resistance increases by a factor of 10 with each increase of 10 wire gauges. So #20 would be about  $10\Omega$  per 1000 feet.

You can interpolate between gauges with the relationship

$$\Delta\text{AWG} = 10 \log_{10} \frac{R_2}{R_1} \quad (1542)$$

where  $\Delta\text{AWG}$  is the change in wire gauge between resistances  $R_1$  and  $R_2$ . This formula is similar to the decibel relationship of powers, so it's not unreasonable to remember.

### Example

What is the approximate resistance of #26AWG copper wire?

### Solution

We know that #10AWG is  $1\Omega$  per 1000 feet. Then #20 is about  $10\Omega$  per 1000 feet and we'll choose that as  $R_1$ , our starting point. The value of  $\Delta\text{AWG}$  is  $26 - 20 = 6$ . Plug those values into equation 1542 and we obtain  $R_2 = 39.8\Omega$  per 1000 feet. (According to a wire table, it's actually  $40.82\Omega$  per 1000 feet, so this approach is fine for estimates.)

If you prefer metric units, #5AWG is about  $1\Omega$  per kilometer, and the same method applies.

### 34.9.2 Resistance of Printed Wiring Traces

The resistance of printed wiring traces depends on their length and cross-sectional area, according to

$$R = \rho \frac{L}{A} \quad (1543)$$

where in CGS units the quantities are:

- $R$  resistance in  $\Omega$
- $\rho$  resistivity of copper,  $1.724\mu\Omega\text{-cm}$
- $L$  length of the conductor in centimetres
- $A$  cross-sectional area of the conductor in square centimetres

In the construction of a printed wiring board, the copper thickness is specified as in ounces, which is the weight of copper plated over a one-foot square area. Boards are typically plated with 1, 1.5 and 2 ounce copper.

Using the density of copper ( $8.92\text{ grams/cm}^3$ ) and working through the various conversion factors, a 1 ounce copper plating is equivalent to a thickness of 0.00135 inches (0.00343 centimetres).

The width of printed circuit traces is specified in 'mils', or thousandths of an inch.

**Example**

A trace 30 mil trace on a printed wiring board is 9 inches long. The board is plated with 1oz copper. If the trace conducts a peak current of 2 amps, what is the end-end voltage drop across this trace?

**Solution**

First, we need to use equation 1543 to calculate the resistance of the trace. The trace width is 0.030 inches, or 0.0762 centimetres. The trace height for 1oz copper is 0.00343 centimetres. The length is 9 inches, or 22.8 centimetres. Plugging these figures into equation 1543, we have

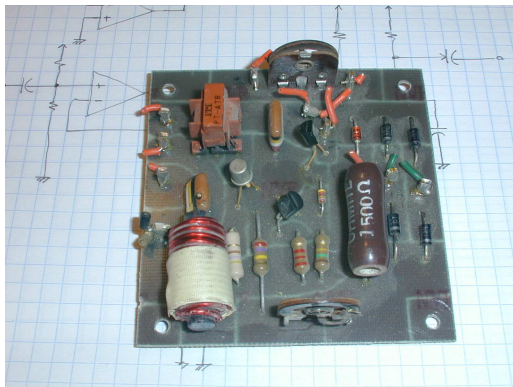
$$\begin{aligned} R &= \rho \frac{L}{A} \\ &= 1.724 \times 10^{-6} \times \frac{22.8}{0.0762 \times 0.00343} \\ &= 0.15\Omega \end{aligned}$$

The voltage drop is

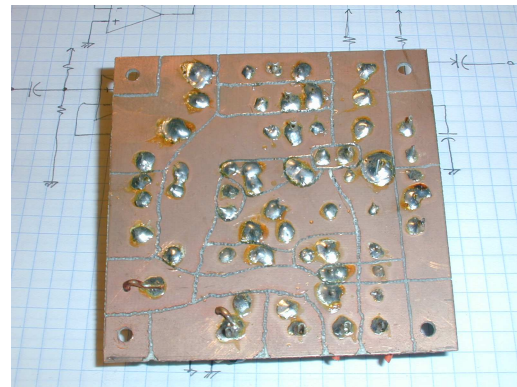
$$\begin{aligned} V &= IR \\ &= 2 \times 0.15 \\ &= 0.3 \text{ volts} \end{aligned}$$

Let us now put this in perspective. For an 8 bit, 5 volt A/D converter, each step is about 19mV, so this voltage drop represents a possible error source of 15 digital counts, or a 6% error. There are many systems where this error would be unacceptable. If this voltage drop is in the form of an AC voltage, then it would show up as jitter in the A/D reading. The wiring of the circuit board must be arranged so that this voltage does not become a source of error - by adding to the signal voltage, for example.

Where traces must have low resistance, they should be made as wide as possible. It is may be possible to design a printed wiring board so that conducting areas occupy most of the board, and the insulating gaps are narrow. This substantially reduces the resistance of the conducting traces.



(a) Component Side



(b) Wiring Side

Figure 906: Reducing Trace Resistance

Figure 906 shows a circuit board that was laid out in this fashion. The conductors are islands of copper that are isolated by narrow channels.

### 34.9.3 Wiring Inductance

Modern digital circuits operate at fast transition rates, in the order of tens of nanoseconds through to fractions of a nanosecond. Even if the associated currents are small, the rapid **change** creates large values of  $di/dt$ .

According to the equation

$$e = L \frac{di}{dt} \quad (1544)$$

this rapidly changing current can create significant voltage across a very small inductance.

Even a short length of wire can have sufficient inductance to generate significant voltage under these circumstances. This is somewhat counterintuitive, so consider the following demonstration.

As shown in figure 907, an #18AWG (Pomona 1440) test lead is connected across the terminals of a signal generator. At DC and low frequency, this shorting strap has a very low impedance and so the measured voltage on the oscilloscope is quite small. However, at a frequency of 1MHz, a short length of wire has an appreciable inductance and consequently a significant voltage appears between the ends of the wire.

The open-circuit voltage of the generator is set to 8Vp-p, 1MHz. The internal  $50\Omega$  resistance of the generator and the inductance of the wire form a voltage divider, with the results shown under *Measured Inductance*.

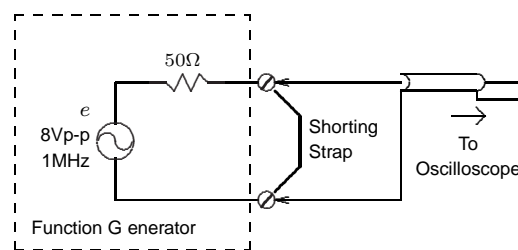


Figure 907: Wiring Inductance

Wire Length Length, Inches	Developed Voltage mVp-p	Reactance $\Omega$	Measured Inductance, $\mu\text{H}$	Predicted Inductance $\mu\text{H}$
12	300	2.05	0.326	0.382
24	600	3.6	0.644	0.856
50	1200	8.8	1.41	1.97

Notice that the developed voltage and the measured inductance of the shorting wire increase with the length of the wire. If the frequency is dropped to 100kHz, the developed voltage drops by a factor of ten, confirming that the inductance is proportional to frequency.

#### A wire has significant inductance at high frequency.

Can this inductance be predicted? Reference [289] gives the following formula for the inductance of a straight wire:

$$L = 2 \times 10^{-4} \left( \ln \frac{4l}{d} - 0.75 \right) \mu\text{H} \quad (1545)$$

where the variables are:

- $L$  inductance in  $\mu\text{H}$
- $l$  length of the conductor, mm
- $d$  diameter of the conductor, mm

The diameter of #18AWG, according to a wire table, is 1.024mm. Plugging the wire lengths (in mm) from the measurement into equation 1545, we obtain the figures under *Predicted Inductance*. The predictions are in the ball-park, which is usually sufficient.

### 34.9.4 Inductance of a Printed Wiring Trace

The inductance of a printed wiring trace is given in [289] as:

$$L = 2 \times 10^{-4} \left[ \ln \left( \frac{2l}{w+h} \right) + 0.2235 \left( \frac{w+h}{l} \right) + 0.5 \right] \mu\text{H} \quad (1546)$$

where the variables are:

- $L$  inductance in  $\mu\text{H}$
- $l$  length of the conductor, mm
- $w$  width of the conductor, mm
- $h$  height of the conductor, mm

#### Example

What is the inductance of a 20cm length of 10mil wide copper trace, printed with 1 oz copper?

#### Solution

First, we need to calculate the dimensions of the conductor in millimetres.

Length  $l$ :

$$\begin{aligned} l &= 20\text{cm} \times \frac{10 \text{ mm}}{1 \text{ cm}} \\ &= 200 \text{ mm} \end{aligned}$$

Width  $w$ : One 'mil' is 1/1000 inch. An inch is 25.4mm. Then 10 mils is:

$$\begin{aligned} w &= 10\text{mil} \times \frac{1 \text{ inch}}{1000 \text{ mil}} \times \frac{25.4 \text{ mm}}{1 \text{ inch}} \\ &= 0.254 \text{ mm} \end{aligned}$$

Height  $h$ : On page 969 we established that 1oz copper plate has a thickness of 0.00343 centimetres, or 0.0343 millimetres.

Now we can calculate the inductance, using equation 1546:

$$\begin{aligned} L &= 2 \times 10^{-4} \left[ \ln \left( \frac{2l}{w+h} \right) + 0.2235 \left( \frac{w+h}{l} \right) + 0.5 \right] \mu\text{H} \\ &= 2 \times 10^{-4} \left[ \ln \left( \frac{2 \times 200}{0.254 + 0.0343} \right) + 0.2235 \left( \frac{0.254 + 0.0343}{200} \right) + 0.5 \right] \mu\text{H} \\ &= 0.309 \mu\text{H} \end{aligned}$$

### The Effect of Length and Width of Conductor on Inductance

Equation 1546 indicates that trace inductance

- increases directly with trace length.
- decreases slowly with trace width, since the log term dominates
- is largely unaffected by trace thickness

For example, quadrupling the trace width in the previous case reduces the inductance to  $0.247\mu\text{H}$ . Consequently **trace inductance is most effectively reduced by shortening the length of a conductor.**

#### 34.9.5 Power Supply Inductance and The Bypass Capacitor

Many systems include both digital and analog circuitry. The high frequency components of a digital signal (figure 902) tend to propagate via power supply lines into other sections of the circuit. It is a major challenge to keep these digital signals from interfering with analog sections of the system.

One of the main lines of defence is the *bypass capacitor*. A bypass capacitor is connected between a power supply line and ground. Ideally, it conducts all frequencies of alternating currents, thereby preventing any variation in power supply voltage. Alternatively, the bypass capacitor can be regarded as a reservoir of charge that can be provided to keep the power supply line at a constant voltage.

An example of a mixed analog-digital system power supply is shown in figure 908(a).

The system consists of digital gates  $U_1$  and  $U_2$ . The same power supply is used by the analog section, represented by the op-amp  $U_3$ .

The digital section is redrawn in figure 908(b) to indicate its essential features. The power supply connections create inductances  $L_1$  and  $L_2$ . The input of logic gate  $U_2$  can be represented by the load capacitance  $C_{in2}$ .

In figure 908(c) the logic gate is represented by a SPDT switch, which connects the output pin to the supply  $V_{CC}$  or to ground.

Suppose that this switch is originally in the lower position, so that the capacitor  $C_{in2}$  is discharged. Then the switch is moved to the upper position. The capacitor must charge through the power supply inductance  $L_1$ . Since inductor current cannot change instantaneously, inductor  $L_1$  appears briefly as an open circuit, and the voltage at  $A$  drops to ground level. Then the capacitor charges to the supply voltage and point  $A$  returns to  $V_{CC}$ . Inductance  $L_1$  and capacitance  $V_{CC}$  form a resonant circuit. Depending on the circuit resistance - which affects the damping -

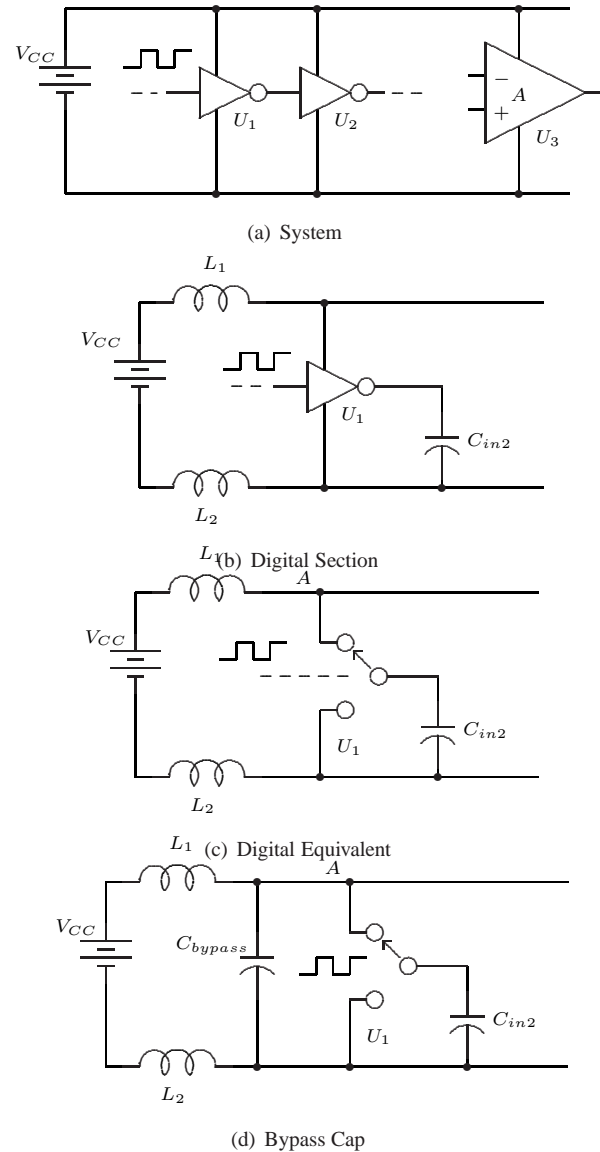


Figure 908: Analog/Digital Power Supply System



the transient may result in ringing at the resonant frequency of the inductor-capacitor. In any case, an ugly voltage transient appears on the power supply line.

Now, back to figure 908(a). This same transient, created by the digital circuitry, will appear at the  $V_{CC}$  terminal of the op-amp. Since the op-amp is unlikely to have significant power supply rejection at the frequency of this transient, the transient will appear at the output of the op-amp and contaminate that signal.

In some cases, the transient voltage at  $A$  may communicate itself sufficiently to the subsequent logic gate  $U_2$  to cause it to switch incorrectly.

Now consider figure 908(d), in which a *bypass capacitor*  $C_{bypass}$  is provided near the digital device. This capacitor is chosen to be large compared to  $C_{in2}$ . It acts as a reservoir of charge that can be delivered rapidly into the load capacitance, without materially affecting the supply voltage. The charge on  $C_{bypass}$  is then restored via  $L_1$  and the power supply over a longer period of time. The net effect is to substantially reduce the magnitude of switching transients on the power supply line. (It would help significantly to provide the analog circuitry with its own power supply lines, independent of the digital power supply lines. However, since the power supply undoubtedly has some finite internal inductance, separate leads are not a perfect solution and bypass capacitors must be provided at the digital circuitry.)

For this to work, the bypass capacitor itself must not contain significant inductance, and that is the subject of the next section.

### 34.9.6 Parasitic Inductance in Capacitors

The reactance of an ideal capacitor decreases with frequency, without limit. Unfortunately, the connecting leads of a real-world capacitor form an inductance in series with the capacitor. The reactance of this inductance rises with frequency until it dominates, and the capacitor ceases to function as such.

Figure 909 illustrates this effect in the time domain.

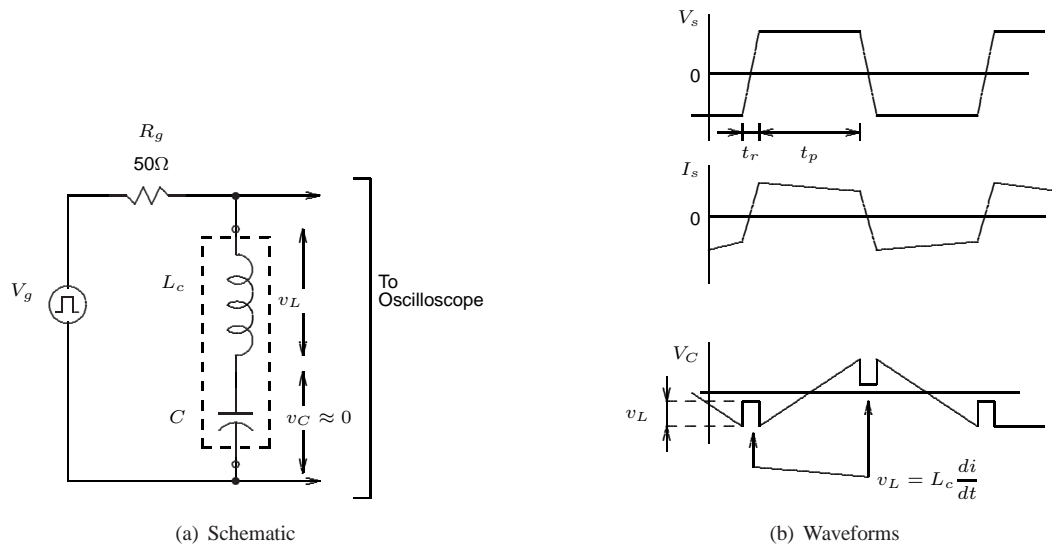


Figure 909: Capacitor Parasitic Inductance

A square wave generator with internal resistance  $R_g$  connects to a capacitor  $C$ . At low frequencies, this would result in the usual exponential charging waveform of an RC low-pass filter. However, the frequency of the generator is set to a large enough value that the capacitor cannot change substantially in voltage over one half of the pulse cycle. Consequently, it's a reasonable approximation to take the capacitor voltage as zero.

The capacitor includes parasitic inductance  $L_p$ . When the generator voltage transitions from its negative value to its positive value, there is an abrupt change of capacitor current. This change in current induces a voltage in the parasitic inductance during the transition time  $t_r$ . This inductive spike is shown in idealized form as a 'notch' in the capacitor waveform  $V_c$ .

(In practice, the situation is more complex. The lead inductance and the capacitance form a resonant circuit. The generated pulse is a damped sinusoid with a decay time that depends on the damping in the circuit.)

To minimize the parasitic capacitance of bypass capacitors, their lead length and the length of the connecting traces to the device being bypassed, should be as short as possible. Surface mount capacitors, with their zero lead length, are most effective.

### Example

In the circuit diagram of figure 909, an 11 volt peak-peak, 1MHz square wave generator voltage is driven through a  $50\Omega$  resistance into a 100nF capacitor. The rise (and fall) time of the square wave is 17nsec. The capacitor lead lengths are 2.7cm each. The inductive spikes across the capacitor measure 200mV peak-peak. Determine the parasitic inductance of the capacitor.

### Solution

Assume that the voltage across the capacitor is approximately zero volts. Then the capacitor current changes by an amount equal to the step change in voltage divided by the generator resistance.

$$\begin{aligned}\Delta i &= \frac{\Delta V_g}{R_g} \\ &= \frac{11 \text{ volts}}{50\Omega} \\ &= 0.22\text{A}\end{aligned}$$

This change in current occurs during the rise time of the generator. Using

$$v_L = L_p \frac{\Delta i}{\Delta t}$$

We can solve this equation for the inductance as:

$$\begin{aligned}L_p &= v_L \frac{\Delta t}{\Delta i} \\ &= (200 \times 10^{-3}) \frac{17 \times 10^{-9}}{0.22} \\ &= 15.4 \times 10^{-9} \text{ H}\end{aligned}$$

### 34.9.7 Capacitor Resonance

The capacitance and parasitic inductance form a series resonant circuit, along with a very small series resistance. In the frequency domain, the overall impedance is as shown with the solid line in figure 910. At low frequencies, the total impedance is essentially capacitive, since the inductance is small. The total impedance decreases with frequency. At the resonant frequency, the capacitive and inductive reactances are equal and cancel, leaving only the series resistance. Above the resonant frequency, the total impedance is inductive, since the capacitance is small. The total impedance increases with frequency. **A bypass capacitor is useable in the region below resonance. The connecting leads of a bypass capacitor should be as short as feasible to minimize parasitic inductance.**

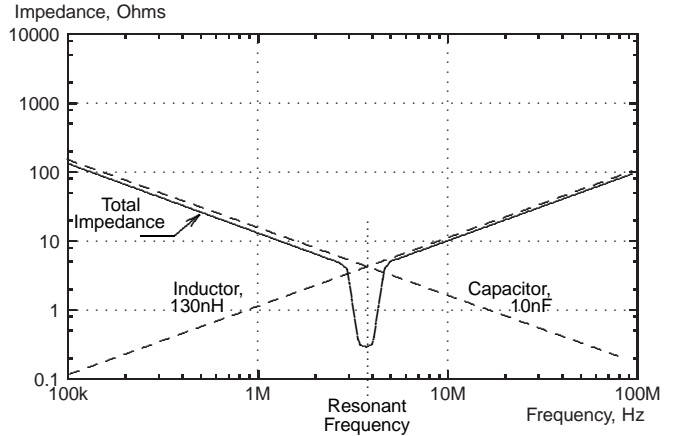


Figure 910: Bypass Capacitor Resonance

### 34.9.8 Broadbanding the Bypass Capacitor

A digital waveform spectrum contains a wide range of frequencies, as seen on page 964. Ideally, then, the bypassing capacitor should have a broadband response. This can be obtained by placing two capacitors of different values in parallel. Then the resonant frequencies are staggered and the lower impedance capacitor will define the total impedance.

Figure 911 shows the effect of a 100nF capacitor in parallel with a 1nF capacitor. The lower of the individual impedances (dotted lines) define the total impedance (solid line), which is now maintained at a low value over a wider range of frequencies.

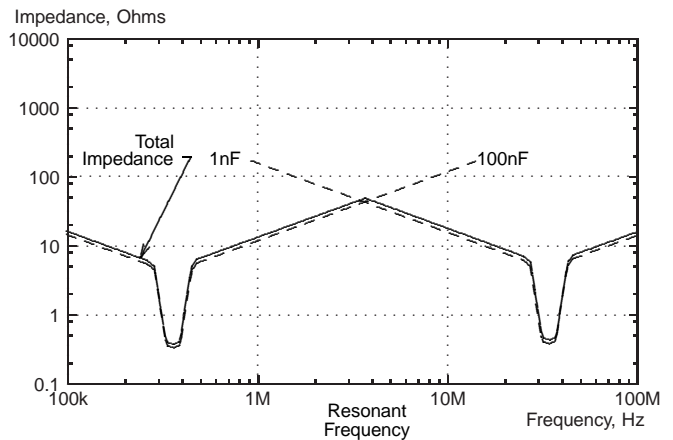


Figure 911: Broadbanding the Bypass Response

## 34.10 Common Mode and Differential Signals

The concept of *differential* and *common-mode* signals (section 21.5) is important in the control of interference. In this section, we'll show an example where the input circuitry of an amplifier can reject noise by (a) setting up a differential input and (b) rejecting the common-mode input signal.

**34.10.1 Rejection of Ground Conductor Noise**

There is nothing magical about a *ground* connection. It connects between various points of a circuit and, for convenience, is taken as the reference point for certain voltage measurements. However, currents flow through the various ground connections and thereby create voltage gradients that put different so-called ground points at different potentials.

It is an illustrative exercise to measure the potential between the ground terminals of two AC outlets that are separated in space. There can be millivolts or even volts of potential difference between them.

This potential difference between ground points is an issue in the situation shown in figure 912(a). A sensor of some type produces a signal  $e_s$  and is connected to ground at point A. The signal is connected to an amplifier which is connected to ground at point B. By virtue of circulating currents in the ground path, points A and B are separated by a potential  $e_n$ .

Redrawing figure 912(a) as figure 912(b), it can be seen that the signal and noise voltages are in series. The input voltage is the sum of  $e_s$  and  $e_n$ . The signal-noise ratio is entirely dependent on the relative amplitudes of the two voltages: there is no rejection of the noise signal.

Figure 912(c) shows way to reduce the effect of the noise. Providing that the signal is AC (as in an audio system, for example) a transformer at the input of the amplifier will reject the noise voltage. By superposition, the noise voltage appears equally at both transformer primary leads, so it cancels. The signal voltage appears differentially across the transformer primary, so it creates a voltage in the secondary.

Stray capacitance may exist between the transformer primary and secondary windings, and this electrostatic coupling will insinuate a noise signal into the input of the amplifier. High quality audio transformers include an electrostatic shield between the windings that can be grounded to prevent this coupling.

In effect, the sensor voltage  $e_s$  is a differential signal and the noise voltage  $e_n$  is the common mode signal. The system rejects the common-mode signal.

Figure 912(d) shows an approach similar to figure 912(c). In this case, a differential amplifier or instrumentation amplifier rejects the common-mode noise signal  $e_n$ . Since the input is direct coupled (no transformer) this approach will work down to zero frequency (DC).

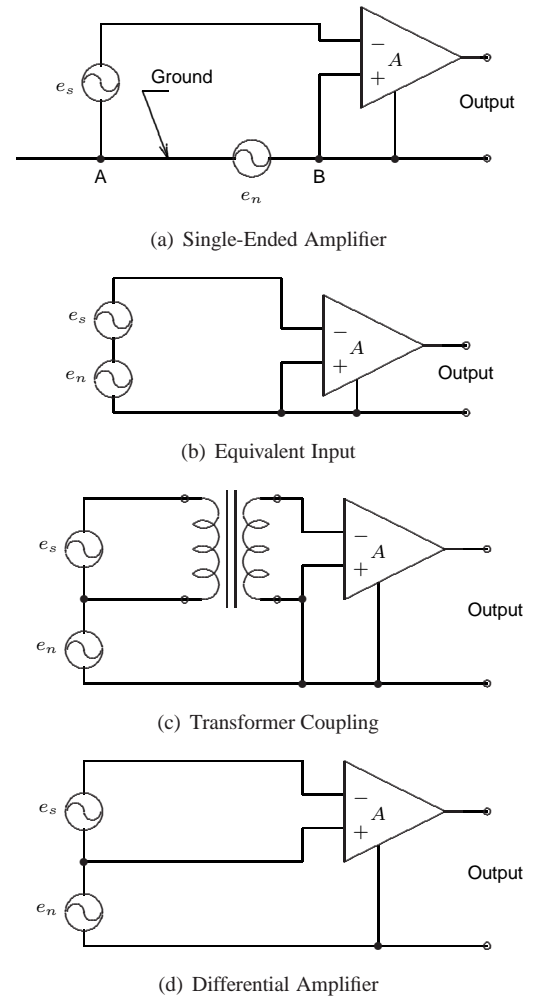


Figure 912: Differential Amplifier Application

### 34.10.2 Differential Amplifier Enhancements

To further reduce noise pickup, there are additional improvements that can be made to the differential input circuit.

The differential arrangements of figures 912(c) and 912(d) allow the signal conductors to be equal in length and physically close together. This minimizes the loop area in the input circuit of the amplifier, and therefore reduces pickup of any AC magnetic field. For even better magnetic field rejection, the two signal leads can be twisted together as shown in figure 913(a). Any field passing through the two input conductors will generate equal and opposite voltages in two adjacent loops of the conductor, so that magnetically induced voltages cancel.

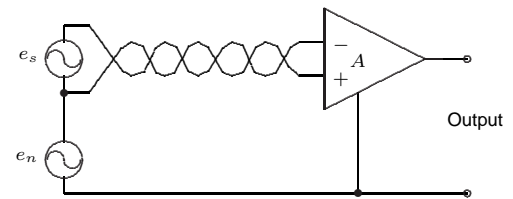
When Shakespeare wrote *When sorrows come, they come not single spies but in battalions*<sup>266</sup>, he could well have had interference problems in mind. Not only is magnetic field pickup an issue, but there may be an interfering electric field, and there may be interfering radio signals.

Electric field interference couples via stray capacitance into the input leads. If these capacitances are precisely equal and at a frequency where the differential amplifier has sufficient common-mode rejection, then the differential amplifier will eliminate the interfering signal. However, cable is readily available with a twisted pair and an outer electrostatic shield, and this provides an additional level of defense. A twisted-shielded pair is shown in figure 913(b). As shown in that diagram, a capacitively coupled electric field signal  $e_{n2}$  will be directed to ground by the shield.

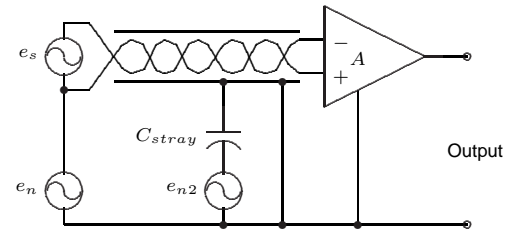
This same shield acts as a defense against radio-frequency interference. This is important, because differential and instrumentation amplifiers have little or no common-mode rejection at radio frequencies. Radio signals cannot penetrate a sealed metal box. The cable shield is connected to the metallic enclosure in such a way that it becomes an extension of the enclosure and prevents radio signal access into the internal circuitry.

Notice that the shield is connected to the enclosure – and ultimately, the reference potential – **at one point only**. If it is connected to ground at two points, then it and the ground path become a complete loop. This is undesirable: AC magnetic fields may induce currents in this loop, or the shield may provide an additional path for ground currents. In either case, these currents may develop a voltage across the shield that creates interference.

The scheme shown in figure 913(b) is widely used in audio systems for microphone signals. A dynamic (moving coil) microphone generates signals in the millivolt range and requires a signal-noise ratio of at least 60db. A microphone cables is often long and therefore act as an excellent antenna, and may pass through regions around lighting equipment with large magnetic interfering fields. Consequently, stringent measures are required to minimize noise pickup. Twisted, shielded pair cable is standard, with connection of the shield to the equipment enclosure and the microphone casing. The input circuitry is differential, either using a transformer or instrumentation amplifier.



(a) Twisted Pair Wiring



(b) Twisted-Shielded Pair Wiring

Figure 913: Differential Amplifier Enhancements

<sup>266</sup>Words uttered by King Claudius in Hamlet, Act 4, Scene 5.

**34.10.3 Guarding**

There is one further enhancement that may be useful under some circumstances: the *guard* [136], [290]. According to [291]:

*Guarding involves driving a shield, at low impedance, with a potential essentially equal to the common-mode voltage on the signal wire contained within the shield. Guarding has many useful purposes: it reduces common-mode capacitance, improves common-mode rejection, and eliminates leakage currents in high-impedance measurement circuits.*

There is a certain amount of stray capacitance between the shield and the signal-carrying conductors of a balanced-shielded cable. If these stray capacitances are not equal, then the common-mode voltage will drive more current through one than the other. This creates a differential voltage on the internal signal wires.

If the shield is driven to the same potential as the common-mode voltage, then regardless of the magnitude of the stray capacitances there will no current through them, thereby improving rejection of the common-mode signal. Figure 914 shows a method of accomplishing this [136].

The resistors  $R_{3A}$ ,  $R_{3B}$  establish a voltage equal to the common mode input voltage  $E_{CM}$  at the midpoint of the divider. Amplifier A3 buffers this voltage and provides the necessary drive to the cable. This technique is limited in effectiveness by the frequency response of the op-amps, so it should be regarded as an option only at low frequencies.

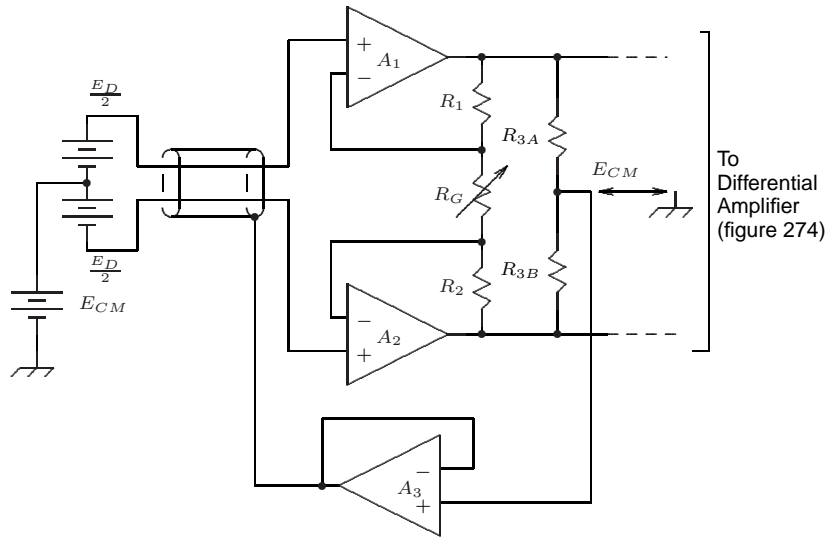


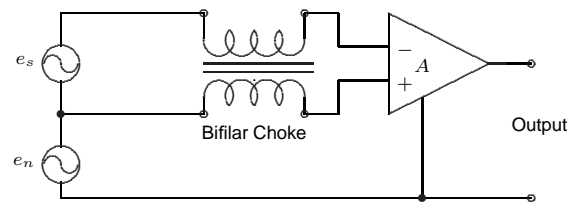
Figure 914: Guarding

**34.10.4 Bifilar Choke**

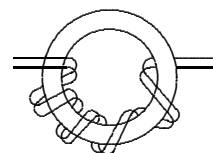
Most differential amplifiers are limited to the frequency regime below 100kHz. It is common for one or more radio-frequency signals to appear as a common-mode noise signal at the input of an amplifier. Amplitude modulated RF signals are detected in an amplifier nonlinearity (such as the base-emitter junctions of a BJT differential amplifier pair) and then appear as noise in the amplified output signal.

The *bifilar choke* is often effective against this type of high-frequency common-mode signal. The choke is inserted in the input leads as shown in figure 915(a).

The winding of a bifilar choke is shown in figure 915(b). The two signal-carrying conductors are wound in the same direction around the core. The differential current (the desired signal due



(a) Circuit



(b) Winding

Figure 915: Bifilar Choke

to  $e_s$ ) flows in opposite directions in the two conductors so their magnetic fluxes cancel and the choke has no inductance and no effect.

The common-mode current (the undesired signal due to  $e_n$ ) is flowing in the same direction in both conductors. It generates a magnetic flux in the core and so the core has significant inductance and opposes the flow of this current.

In the presence of strong radio fields, as in the downtown of a large city, the bifilar choke may be useful in the *output* wiring of an amplifier. Many amplifier systems include a negative feedback connection between the output of the amplifier and the input. If there are long leads attached to the output of the amplifier (the speaker leads of an audio amplifier, for example), they behave as an antenna, bring the RF signal to the negative feedback path, and then to the input stage of the amplifier. The unwanted RF signal then becomes part of the signal path of the system.

If the amplifier behaved as an ideal voltage source at all frequencies, the output circuit of the amplifier would absorb the unwanted signal. However, the amplifier gain and the effectiveness of negative feedback are usually absent at radio frequencies.

### 34.11 EMC: Electromagnetic Compliance

Alternating currents in a conductor generate alternating electric and magnetic fields launch radio waves – an *electromagnetic field* – into space, that is, the current and conductor behave as a *transmitting antenna*. The wiring of nearby electronic devices act in a reverse fashion: these conductors intercept the electromagnetic field, this creates an alternating current in the conductor, and these conductors act as a *receiving antenna*.

If the received electromagnetic signal is sufficiently large compared to the normal currents in the receiving device, they can interfere with its operation. For example, personal computers and other digital devices can radiate electromagnetic signals of significant magnitude over a wide range of the frequency spectrum. To prevent interference with radio receivers and other electronic apparatus, the emitted signals must be reduced and contained or the receiving equipment must be made less susceptible to interfering signals<sup>267</sup>.

To minimize the social disruption that is caused by mutual interference between electronic devices, governments regulate the maximum allowable field strength of emitted electromagnetic waves. For example, in the United States of America, the allowable maximum emissions are specified in the Federal Communications Commission *Part 15 Rules* [292], [293]. In Europe, the corresponding standard is *EMC Directive 89/336/EEC*. In addition to specifying allowable emissions, the European standard specifies the required immunity of devices to external electromagnetic fields [294].

Other industries and organizations may have their own requirements. For example, the allowable emission levels for certain automotive manufacturers (eg, General Motors, with standard GMW3097) are substantially stricter than the FCC Part 15 Rules<sup>268</sup>.

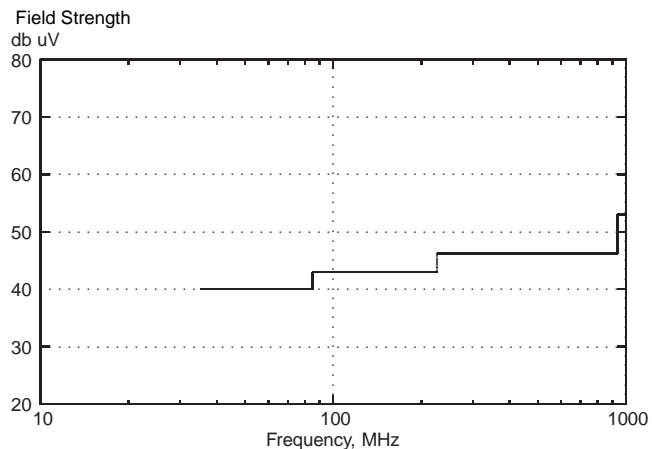


Figure 916: FCC Part 15 Class B Emission Limits

<sup>267</sup>In one case, a student using a pacemaker inadvertently placed himself in the beam of a laboratory microwave transmitter. Interference from the microwave signal stopped the pacemaker and the patient collapsed. Fortunately, this moved him out of the radio field, the pacemaker restarted again, and the person survived.

<sup>268</sup>Electronic devices in automobiles exist in close proximity to each other, and automotive radios are more sensitive than domestic radios.

The allowable emission limits for FCC Regulation Part 15 are shown in figure 916. There are two limits: the higher one, Class A, for is for industrial equipment. The lower limit, Class B, is required for domestic equipment since television and radio receivers are likely to be in close proximity to electronic equipment in a home environment. In both cases, the equipment emission in microvolts/metre must not exceed the limits shown.

A calibrated spectrum analysis from the equipment under test can be overlaid on figure 916 to determine whether the equipment meets the compliance requirement.

### 34.11.1 EMI from Differential and Common Mode Currents

Both differential and common mode currents cause electromagnetic interference.

A differential mode current is a current that flows in a loop on the printed circuit board. The source current is balanced by the return current.

A common mode current is a current that is not balanced by a return current. Typically this occurs in a cable that is attached to the board, in which the outgoing and return currents are not completely balanced. In this situation, the common-mode current flows through stray capacitances associated with the cable, the chassis, and earth grounds.

The magnitude of the field due to differential current in a loop and common-mode current in a conductor are given by [295]:

$$E = \frac{\mu_o I A \pi f^2}{2rC} \quad (1547)$$

$$E = \frac{\mu_o I l f}{2r} \quad (1548)$$

where the variables are

- $E$  Field strength, volts/metre
- $\mu_o$  Magnetic permeability of space, a constant,  $4\pi \times 10^{-7}$
- $I$  Current in the loop, amps.
- $A$  Area of the circuit loop, square metres
- $f$  Frequency, Hz
- $r$  Distance of measurement from coil centre, metres
- $C$  Speed of light,  $3 \times 10^8$  metres/second
- $l$  Length of the radiating conductor, metres

#### Example

The FCC Part 15 limit on emissions for a 'Part B' device, used in a residential location, are not to exceed  $150\mu\text{V/metre}$  in the frequency band 88 to 216MHz, measured at a distance of 3 metres.

Determine the (a) maximum allowable (differential) current in a circuit loop 1cm x 1cm, and (b) the maximum allowable common-mode current in a cable 1 metre long.

#### Solution

(a) **Differential current:** The area of the circuit loop is 0.01 metres x 0.01 metres, or  $10^{-4}$  square metres. According to equation 1547, the radiated field increases with frequency, so the worst case is at a frequency of 216 MHz. Rearranging equation 1547 to solve for differential current, we have:

---

This forces lower radiation limits.



$$\begin{aligned}
 I &= E \left[ \frac{2rC}{\mu_o A \pi f^2} \right] \\
 &= 150 \times 10^{-6} \left[ \frac{2 \times 3 \times (3 \times 10^8)}{(4\pi \times 10^{-7}) \times (10^{-4}) \times 3.141 \times (216 \times 10^6)^2} \right] \\
 &= 14\text{mA}
 \end{aligned}$$

(b) **Common-Mode Current:** Equation 1548 indicates that the common-mode current increases with frequency, so the worst case frequency is again 216 MHz. Rearranging equation 1548 to solve for common-mode current, we have:

$$\begin{aligned}
 I &= E \left[ \frac{2r}{\mu_o l f} \right] \\
 &= 150 \times 10^{-6} \left[ \frac{2 \times 3}{(4\pi \times 10^{-7}) \times 1 \times (216 \times 10^6)} \right] \\
 &= 3.3\mu\text{A}
 \end{aligned}$$

Notice that the same field strength is created with a much smaller common-mode current than differential current. This is an indication that *a cable attached to an electronic circuit has the potential of creating a substantial interfering electromagnetic field, even with very small amounts of common-mode current.* Common-mode currents should be measured with a current probe and if necessary reduced with a common-mode choke.

### 34.11.2 Common Mode Current, EMI, and the Bifilar Choke

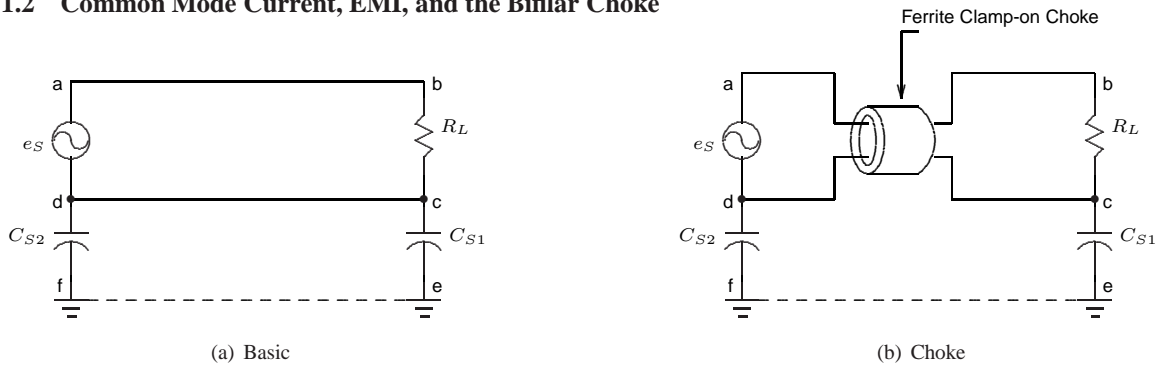


Figure 917: Common-Mode Output Signal

Common mode current, generated by digital circuitry, is a major source of EMI (electromagnetic interference). As an example, consider the circuit shown in figure 917(a), in which signal source  $e_S$  drives AC current through a load resistor  $R_L$ .

In an ideal world, all the current would flow around the smaller loop  $abcd$ . Since the conductors  $ab$  and  $cd$  can be located close together, the area of loop  $abcd$  the radiated magnetic field are both minimized. However, the return conductor  $cd$  has a certain amount of resistance and inductance. As a consequence, some of the return current will flow via the stray capacitances  $C_{S1}$  and  $C_{S2}$  and the ground conductor  $ef$ . This creates a much larger loop  $abeef$  which then radiates a significant interfering signal.

Since the ground current appears in the signal lead  $ab$  but not in the return lead  $cd$ , it appears as a common-mode current in the cable  $abcd$ . The magnitude of the common-mode current can be detected with a clamp-on current probe that encloses both conductors  $ab$  and  $cd$ .

Unfortunately, a very small common-mode current can cause unacceptable electromagnetic radiation. Ott [281] cites  $5\mu\text{A}$  as the maximum allowable common-mode current in a one-metre cable at a frequency of 50MHz, to meet EMI requirements.

One possible cure is shown in figure 917(b). A common-mode ferrite choke<sup>269</sup> is clamped around the cable [296]. The differential currents in the conductor cancel any magnetic field, so the choke has no effect on those currents. An common-mode current does create a net magnetic field, and so the choke opposes that current. In effect, the choke forces the currents  $ab$  and  $cd$  to be equal, so that the common mode current in the ground path  $ef$  and the radiated interference is zero.

### 34.11.3 Differential Drive and Common-Mode EMI Current

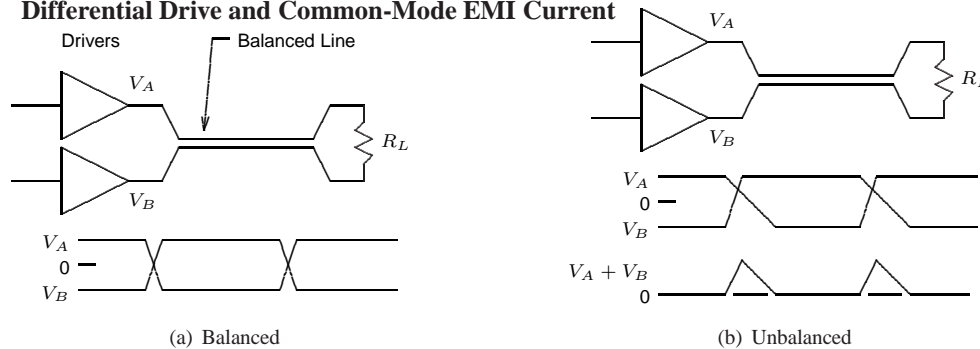


Figure 918: Unbalanced Differential Drive

Figure 918(a) shows another possible example of common-mode output current. The two amplifiers drive a balanced line in antiphase. The peak-peak voltage developed across the load resistance is twice the peak value of each amplifier output. The voltages are opposite and equal at all times, so the average voltage across the load is zero.

Now consider figure 917(b) where the amplifiers have a longer fall time than rise time. The voltages are not equal during the transition times. Triangular pulses of common-mode voltage appear at the load resistance. As in figure 917(a), this common mode voltage will drive current through stray capacitances and the ground path, creating an interfering electromagnetic signal.

Again, a coaxial cable choke will reduce this common-mode current.

### 34.11.4 Evaluating EMC

For a complete circuit board with many digital signals, it is not simple to predict the level of interfering electromagnetic field caused by differential currents in the various circuit loops. Digital signals radiate at a variety of frequencies and the resultant field is a vector sum of a large number of individual fields. The usual approach is to

- Construct the board in a manner that minimizes differential radiation, for example by keeping the area of circuit loops to a minimum [286], [297].
- Measure the overall radiated field, and if it exceeds the allowable level, then
- Determine the magnetic and electric fields at various points on the board to determine the *hot spots* that must be remediated.

On the other hand, a current probe can be used to measure the common-mode current in a cable, and thereby predict its contribution to the radiated electromagnetic field.

<sup>269</sup>Also known as a *balun*, for *balanced to unbalanced* transformer. In this application, the operation is the reverse: to balance an unbalanced current.

### 34.12 Electromagnetic Field Coupling

An electronic circuit may interact with an electromagnetic field as a source or as a receiver.

As a source, the circuit must not radiate a signal with sufficient field strength to interfere with other victim circuits. This is the EMC (electromagnetic compliance) requirement discussed in section 34.11.

As a receiver, an external electromagnetic field couples into the connecting wires of the circuit which act as receiving antennas. This field becomes an electrical signal in the circuit which disturbs the circuit operation. This phenomenon of external an external radio signal disturbing the operation of a circuit is generally known as RFI for *radio frequency interference*.

For example, it is very common in metropolitan areas for television radio-frequency signals to be received by sound reinforcement audio equipment. The radio signal is demodulated by the audio circuitry and is heard as *sync buzz* in the output. An examination of the audio signal shows a demodulated video signal riding on the audio.

As another example, radio frequency interference appears in low-frequency measurement circuits as a variable DC output voltage. If a measurement circuit shows a DC output voltage that varies with the placement of the circuit or the presence of a human body, RFI should be investigated as a cause<sup>270</sup>.

The susceptibility of an electronic circuit to external radio signals may be reduced by three measures:

- **Choose components that are less susceptible to radio-frequency signals.** The translation of a high-frequency RF signal to a DC signal occurs as the result of non-linearity in the active devices. There is some evidence that JFET's and MOSFET's are less susceptible to this RFI demodulation process. All other things being equal, a JFET or MOSFET op-amp is a better choice where RFI may be a problem.
- **Filter all leads connected to the circuit.** Radio-frequency signals tend to be common-mode in nature, so common-mode chokes are most effective. For example, in audio circuitry, radio signals tend to ride on the shield conductor of input signal cables, and thereby enter the sensitive preamplifier circuitry.
- **Enclose the circuit and its cables in a metallic shield.** An unbroken metal shield can attenuate an external radio signals by more than 100db, so this measure is very effective when it can be applied.

Shielding is compromised by running one or more cables into the enclosure – these must be filtered at the point of entry or shielded in such a way that the cable shield is an extension of the enclosure, as shown in figure 919(a).

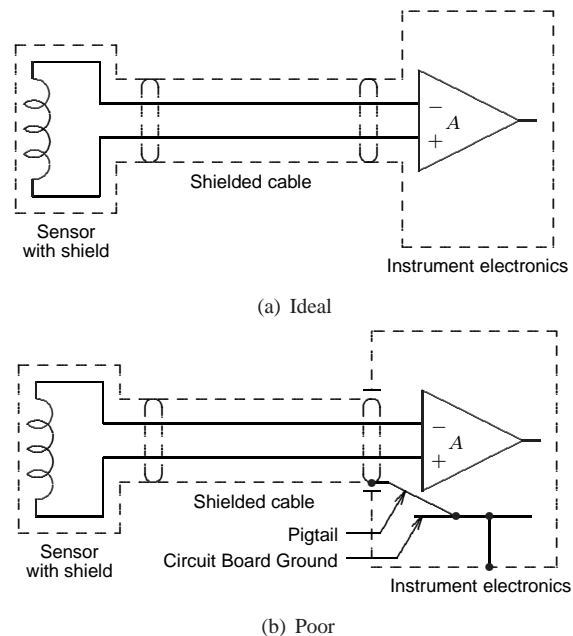


Figure 919: Shielded Enclosure

<sup>270</sup>The other candidate for this behaviour is an unintended high-frequency oscillation in the circuit. A high frequency oscilloscope, properly connected to the circuit through  $\times 10$  probes (section 34.13.4) should show whether oscillation is occurring. In some pathological cases, attaching a scope probe puts an additional capacitance into the circuit that stops the oscillation, and this adds to the diagnostic challenge.

An inferior arrangement is shown in figure 919(b). In this case, the cable shield is connected by a length of wire (the *pigtail*) to a ground on the circuit board and then by another length of wire to the enclosure. A radio-frequency signal will travel in common-mode form on the shield of the cable, onto the pigtail, and then via various stray capacitances into the instrument electronics. In figure 919(b) the cable shield should be directly connected to the enclosure shield. When the shield cannot be grounded in this fashion (because it violates the single-point grounding rule, for example), then the pigtail should be fitted with a radio frequency ferrite choke and bypassed to the enclosure by a capacitor. The capacitor is chosen so that it is a short circuit to RF interference and an open circuit to the desired (lower frequency) signals.

Any openings in the shield – to allow for cooling air or a display unit, for example – may also allow radio energy into the enclosure. To maintain the shielding effect, openings should be small compared to the lowest radio frequency wavelength. An approximate formula for the shielding effectiveness of an aperture [286] is:

$$K_{db} = 20 \log_{10} \left( \frac{\lambda}{2d} \right) \quad (1549)$$

where the variables are

$K_{db}$	Shielding effectiveness, db
$\lambda$	Wavelength of the interfering radio signal, metres
$d$	Largest dimension of the aperture, metres

The wavelength is given by:

$$c = f\lambda \quad (1550)$$

where:

$c$	Velocity of an electromagnetic wave, $3 \times 10^8$ metres/sec
$\lambda$	Wavelength, metres
$f$	Frequency, Hz

### 34.13 Measurement Technique

Ideally, our measurement instruments should show an accurate rendition of what is taking place in the circuit under test. However – and this is particularly true when troubleshooting noise problems – the observed measurement may be totally misleading and incorrect. There are two principal reasons:

- *The measuring instrument affects the operation of the circuit.* For example, the capacitance of the measurement probe cable may reduce the observed noise level in the circuit, or may cause the circuit to burst into oscillation.
- *Noise signals couple into the test leads of the measuring instrument.* For example, when you short-circuit the oscilloscope test lead to its ground, the measured signal should go to zero. In many situations it shows a substantial amplitude because one or more noise signals are coupling into the leads of the oscilloscope.

Great care must be taken to validate measurements - make sure they are real results and not the result of some error in the measurement technique.

This section shows various methods of connecting an oscilloscope into a circuit. The oscilloscope is the primary measurement tool for identifying noise problems, and the connection into the circuit is critical in making accurate measurements. The equivalent circuit of each of the following circuits, showing components in the hidden schematic, helps to identify strengths and weaknesses of the measurement technique.

**34.13.1 Single Lead Connection**

An example of poor measurement technique is shown in figure 920. This arrangement is often seen in a university electrical engineering lab, where test leads are in short supply. A signal source  $e_s$  is connected by a single test lead to an oscilloscope. Completion of the measurement loop is via the AC line ground connection of the source (via a power supply, for example) and the AC line ground connection of the oscilloscope. This setup will work after a fashion, but it has several problems:

- If there are significant currents  $i_{AC}$  flowing through the Power Line Ground connection, then there will be a 60Hz voltage signal  $v_{AC}$  between the two ground pins. This AC noise voltage appears in series with the signal, so it reduces the signal-noise ratio.
- The measurement circuit and the Power Line Ground form a loop with a large area. A magnetic field in the vicinity will generate a voltage in the loop and again, noise voltage appears in series with the signal and reduces the signal-noise ratio.
- The unshielded test lead functions as one electrode of a stray capacitance  $C_{stray}$ , coupling electric field signal  $e_n$  into the signal path. The degree to which this is a problem depends on the magnitude of the noise voltage, the stray capacitance, and the impedances of the source and load.
- The return path for signal current is extremely long, and so it will contain significant inductance. If the signal is a pulse waveform, then it will exhibit significant overshoot and ringing at the pulse edges. For pulse waveforms, this is a completely unacceptable arrangement.

**34.13.2 Twin Lead Connection**

Another common measuring setup is shown in figure 921(a). The oscilloscope is connected by a signal and ground lead to the circuit under test.

The test leads are connected to the oscilloscope via a BNC adaptor, figure 921(b).

The merits and problems of this circuit are:

- Figure 921(a) shows the circuit under test is isolated from ground, which is preferred since it prevents current in the Power Line Ground conductor from inducing noise into the circuit. Most lab power supplies

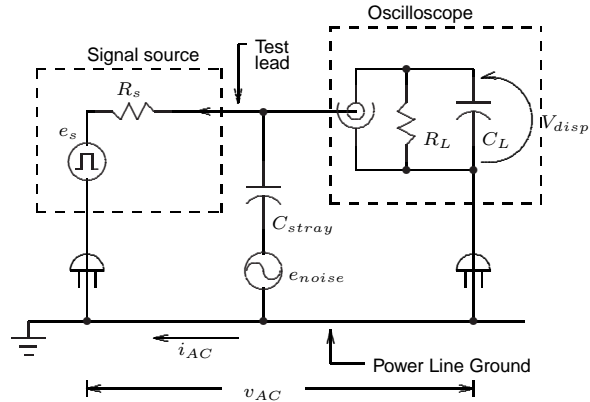
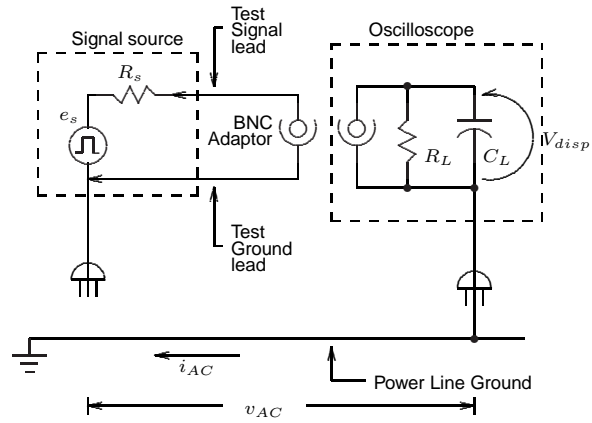
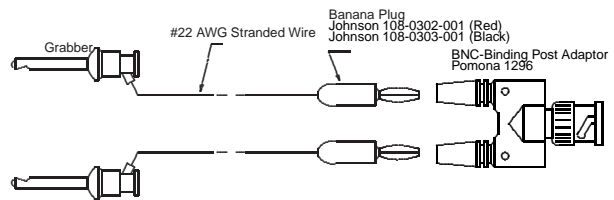


Figure 920: Measurement Using AC Ground



(a) Measurement Setup



(b) BNC Adaptor

Figure 921: Twin Unshielded Leads

are isolated from ground unless the positive (red) or negative (black) terminal is explicitly jumpered to the earth ground (green) terminal.

For this reason, the setup of figure 921(a) is an improvement over that of figure 920.

- To avoid clutter in the diagram, both the noise source  $e_{noise}$  and coupling capacitance  $C_{stray}$  are omitted from figure 921(a), but they are still present and potentially problematic.
- The inductance of the return path is reduced, but not enough to meet the requirement for accurate rendition of digital pulse waveforms.
- This is a very poor technique for assessing the noise level in a circuit. In addition to the possible electrostatic field pickup via stray capacitance into the unshielded signal lead, the signal and ground test leads form a large loop. A time-varying magnetic field will induce voltage into this loop.
- The input capacitance of the oscilloscope loads the circuit under test. Connecting this capacitance into the circuit under test can have a significant effect on the measurement. The internal resistance  $R_s$  of the source on the one hand and the cable capacitance plus the oscilloscope input capacitance on the other, form an RC lowpass filter that will shunt high frequencies to ground. This may show up as a degraded rise time in pulses.

In the frequency domain, the input impedance of the oscilloscope decreases at high frequencies, as illustrated by the following example:

### Example

The equivalent input circuit of a certain oscilloscope is shown as  $1\text{M}\Omega \parallel 50\text{pF}$ . What is the input impedance of the oscilloscope at 1MHz?

### Solution

$$\begin{aligned} X_c &= \frac{1}{2\pi fC} \\ &= \frac{1}{2 \times 3.14 \times (1 \times 10^6) \times (50 \times 10^{-12})} \\ &= 3\text{k}2\Omega \end{aligned}$$

At 1MHz, the oscilloscope input impedance is only  $3\text{k}2\Omega$ . The  $1\text{M}\Omega$  resistive component is irrelevant by comparison.

**The input impedance of an oscilloscope decreases significantly at high frequencies.**

The twin-lead oscilloscope connection is physically convenient and suitable only for non-critical measurement of analog and low-edge-rate digital signals where the signals are large compared to noise.

### 34.13.3 Coaxial Cable Connection

When noise pickup is a potential problem, the signal may be conveyed to the oscilloscope by means of a coaxial cable as shown in figure 922. This is also the arrangement when the oscilloscope is equipped with a so-called  $\times 1$  oscilloscope probe.

- The coaxial cable shields against electrostatic noise pickup, so this source of error is reduced.
- The ground return path is shorter than that of figure 920 and (probably) figure 921 so the transient response to pulse is less likely to show overshoot and ringing.
- However, the coaxial cable has a certain amount of capacitance that appears in parallel with the input capacitance  $C_L$  of the oscilloscope. This further decreases the rise time of pulses and lowers the oscilloscope impedance at high frequencies.

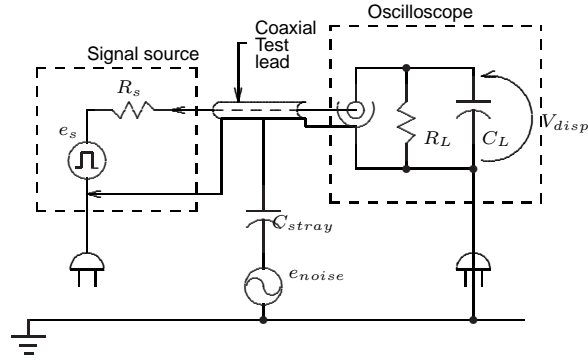


Figure 922: Measurement Using Coaxial Cable

### 34.13.4 $\times 10$ Probe Connection

An oscilloscope  $\times 10$  probe reduces the loading on the circuit under test at the expense of the signal magnitude. Resistors  $R_P$  and  $R_L$  form a voltage divider, so that the signal is reduced by a factor of 10 and the input resistance increased from  $1\text{M}\Omega$  to  $10\text{M}\Omega$ . The probe capacitance  $C_P$  together with the cable capacitance and oscilloscope input capacitance form a capacitive voltage divider. For the attenuation ratio to be independent of frequency, the impedances of the capacitors must have the same ratio as the resistances (section 4.15). The cable capacitance  $C_T$  and input capacitance  $C_L$  are in parallel, so that section of the capacitive divider is  $75\text{pF}$ . Then the probe capacitance  $C_P$  is one tenth that value,  $8.3\text{pF}$ .

Without the  $\times 10$  probe, the input capacitance would be  $75\text{pF}$ . With the  $\times 10$  probe, the input capacitance is  $8.3\text{pF}$  in series with  $75\text{pF}$ , or about  $7.5\text{pF}$ . The  $\times 10$  probe reduces the input capacitance by a factor of 10. This reduces the loading on the measurement circuit, and as a result a  $\times 10$  probe is less likely to affect the rise-time of pulse waveforms.

Notice that the measurement ground point is connected to the oscilloscope probe, not to the oscilloscope. This is important, because it reduces the length of the ground return and the inductance in the measurement ground lead. Furthermore, the area in the loop formed between the probe (point A) and its ground (point B) is susceptible to pickup of any AC magnetic field. Consequently, this area should be minimized by keeping points A and B close together, and minimizing the length of the ground lead.

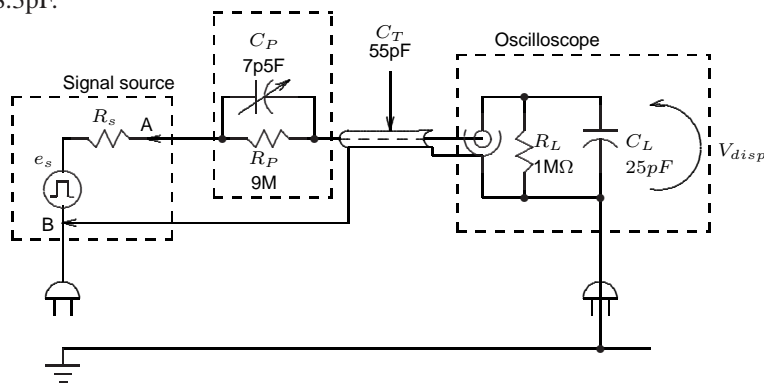


Figure 923: Measurement Using  $\times 10$  Probe

### 34.13.5 Active Probe Connection

The  $\times 10$  probe is useful for many routine analog and pulse measurements. However, where even the resistance and capacitance of a  $\times 10$  probe cannot be tolerated, there is the *active probe*.

Typically, an active probe consists of a discrete component unity-gain amplifier, using a JFET or mosfet input device. In [111], Pease shows an active probe circuit with an input resistance of  $10^{11}\Omega$  in parallel with a capacitance of only 0.29pF.

On the downside, an active probe must be supplied with power, there is a requirement for additional connections to the probe and the maximum signal amplitude is limited to something in the order of  $\pm 10V$ .

## 34.14 Printed Wiring Board Layout

Complete books [30] have been written on printed wiring board layout, and so this section will necessarily mention only some of the highlights. We will assume that the circuit includes both analog and digital circuitry, and the primary objective is to keep digital noise from corrupting analog signals. Recall that the rapid transitions of digital currents create noise spikes across wiring inductance, and these noise spikes may couple into the analog circuitry.

As an illustration of the difficulty involved, consider that the system being designed uses an 8 bit A/D converter with a 3 volt span. Then one step on the A/D converter is:

$$\begin{aligned} V_{step} &= \frac{V_{span}}{2^N} \\ &= \frac{3}{2^8} \\ &= 11.7\text{mV} \end{aligned}$$

As a rough indicator, noise induced in the analog system previous to the A/D converter must be less than this amount.

Now consider that the digital system contains pulses of current equal to 20mA which transition in 10nSec. Using

$$e = L \frac{di}{dt}$$

an 11mV signal will be created by as little as 5.5nH, approximately the inductance of a 5 inch printed wiring trace. The problem grows more severe as the number of A/D bits increase and the edge rate of the digital logic increases. To prevent this problem:

- Separate the digital circuitry and analog circuitry into two distinct areas of the printed wiring board.
- Provide each digital integrated circuit with its own decoupling capacitor, located as close as possible to the device. The decoupling capacitors should have the lowest possible lead inductance.
- If a four-layer (or greater) circuit board is allowable, provide the digital logic with one plane dedicated to ground and a second plane dedicated to power supply positive voltage. The power and ground planes are on the outer layers, the signal lines run on the two inner layers. This has two benefits:
  - The inductance of the power supply and ground lines is at an absolute minimum. Rapidly transitioning currents in these layers will run in such a path that minimizes the inductance (just as current tends to choose the path of lowest resistance). This will then generate the smallest possible noise spikes. If allowed, the currents will mirror each other, with the source current running directly above the return current and the loop area at a minimum. Notice that any slot in the copper plane, that forces the current to take a circuitous path away from the mating current in the other plane, will raise the inductance of that path and the noise voltage.



- Noise voltages generate by loops in the signal path are shielded by the ground and power plane, and this minimizes electromagnetic radiation interference.
- If a two-layer board is required, then provide separate power and ground leads for the analog and digital circuits. The power leads meet only at the power supply positive, and the ground leads only at the power supply ground. Ensure that there is adequate decoupling, so that noise on the digital supply and ground cannot couple into the analog supply and ground.

Low-frequency decoupling can be obtained by providing the digital and analog circuitry each with its own three-terminal regulator. However, integrated-circuit regulators do not function well at the frequency of digital noise pulses. High frequency decoupling requires ceramic capacitors, possibly in an RC or LC lowpass filter configuration.

- Assuming that power and ground lines must be run as separate traces on a two-layer board, route the traces to minimize the enclosed loop area [298], [299].

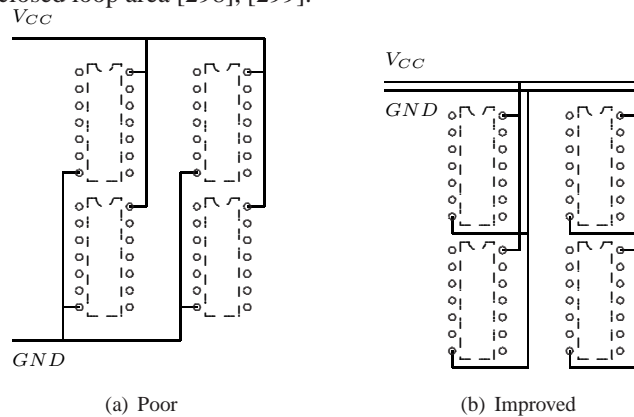


Figure 924: PCB Power Trace Layout

An example of a poor power trace layout is shown in figure 924(a). (The circuit is shown with 16-pin dual inline packages, for which pin 8 is  $GND$ , pin 16 is  $V_{CC}$ . The same principles apply to other package shapes.) The  $V_{CC}$  and  $GND$  lines are some distance from each other and enclose a large loop area. Consequently, the inductance of the leads will be higher than necessary, and this will create electrical noise.

Figure 924(b) shows an improved arrangement. This arrangement requires a double-sided board or wire jumpers where the traces cross. However, the supply and ground traces are much closer together and the enclosed area is a minimum, reducing electrical noise.

### 34.14.1 A/D Converter Layout and Grounding

Many instrumentation circuits consist of an analog section that conditions an input signal, followed by a digital system that provides various digital functions such as storage and display. Linking these two is an analog-digital converter.

The analog and digital sections must have a common ground. How should the analog-digital converter be connected to these circuits and the circuit common, in order to minimize the injection of digital noise into the analog circuit?

Some possible measures are:

- The noise problem increases in difficulty as the step size of the A/D converter decreases, so a 12-bit A/D converter is more likely than an 8-bit converter to be accompanied by noise issues. Use the least precise A/D converter that will satisfy the requirements. (Lower precision A/Ds are also significantly less expensive.)

- Integrating type A/D converters (single and dual slope, for example) are slow but inherently noise resistant. If speed is not a requirement, then these are a preferred choice.
- If an integrating A/D cannot be used, then limit the analog bandwidth to match the signal. Any excess bandwidth in the analog circuitry provides a window for noise.
- High-frequency clocks and high speed edges in the digital circuitry are more likely to cause noise injection into the analog circuitry. Use the slowest logic and lowest acceptable clock speed.
- If the budget permits, use a four-layer circuit board with power supply and ground dedicated to two of the layers. This reduces the impedance of the power and ground conductors and helps contain electromagnetic interference.

In addition to these measures, it is important to understand how the A/D converter should be connected to the supply and ground lines.

An equivalent diagram for a high-speed A/D converter [300] is shown in figure 925.

The analog section contains buffer amplifiers and comparator switches. The digital section may be pictured as a  $n$ -way collection of SPDT switches that connect the digital output lines to the digital supply  $DV_{DD}$  or  $DV_{SS}$ <sup>271</sup>.

Any transition on the digital supply line or digital ground line will couple through the stray capacitances  $C_{S1}$  and  $C_{S2}$  into the analog circuitry. Consequently, the  $DV_{DD}$  and  $DV_{SS}$  lines must be kept stable - or track the analog supply and ground.

Furthermore, when an A/D output logic level changes, the output voltage must charge or discharge the input capacitance of the subsequent logic load, shown as  $C_L$  in figure 925. These pulses of current must be supplied via the  $DV_{DD}$  or  $DV_{SS}$  power supply and ground lines, and may cause voltage excursions that couple through the stray capacitances  $C_{S1}$  and  $C_{S2}$ .

One possible arrangement of power wiring is shown in figure 926(a). This arrangement uses two separate supplies – one for analog circuits and one for digital circuits – and ties the two of them together at single point. (In practice, the analog supply and digital supply would each be a 3-terminal regulator that derives power from some common power supply.) Unfortunately, spikes of current in the digital circuit loop, in the inductance of the ground line at A will generate spikes of voltage that will couple through stray capacitance  $C_{S2}$  into the analog circuitry. Furthermore, digital noise on the digital supply line at B will also couple through stray capacitance  $C_{S1}$  into the analog circuitry.

A better arrangement is shown in figure 926(b). The digital circuit and supply share no conductors with the analog section, so voltage spikes induced on the supply and ground lines are less likely to couple into the A/D converter. The analog and digital grounds of the A/D converter are tied together so voltage cannot be created across stray capacitance  $C_{S2}$ . One possible problem remains: the digital section of the A/D converter must source spikes of current into the load capacitance  $C_L$ , and this will tend to generate spikes across the analog supply line. The filter capacitor  $C_F$  provides a low-impedance source for these pulses of current and inductor  $L_F$  prevents transients at the  $DV_{DD}$  terminal from coupling into the analog section of the A/D.

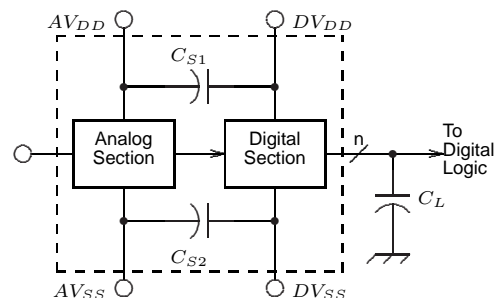
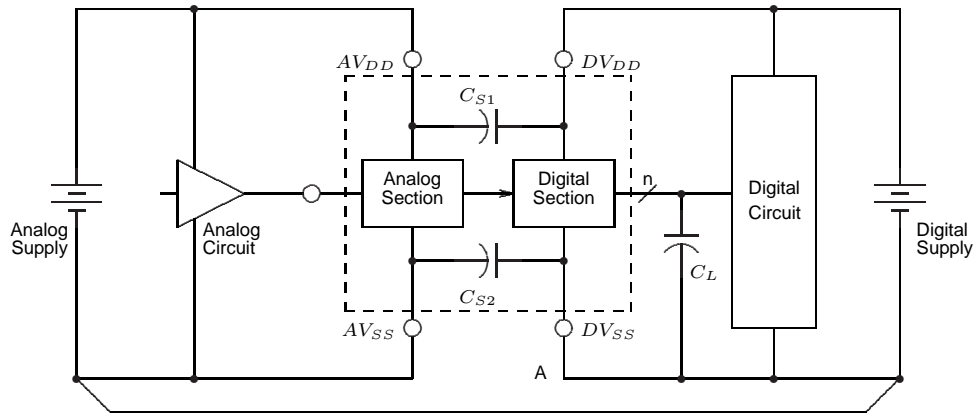
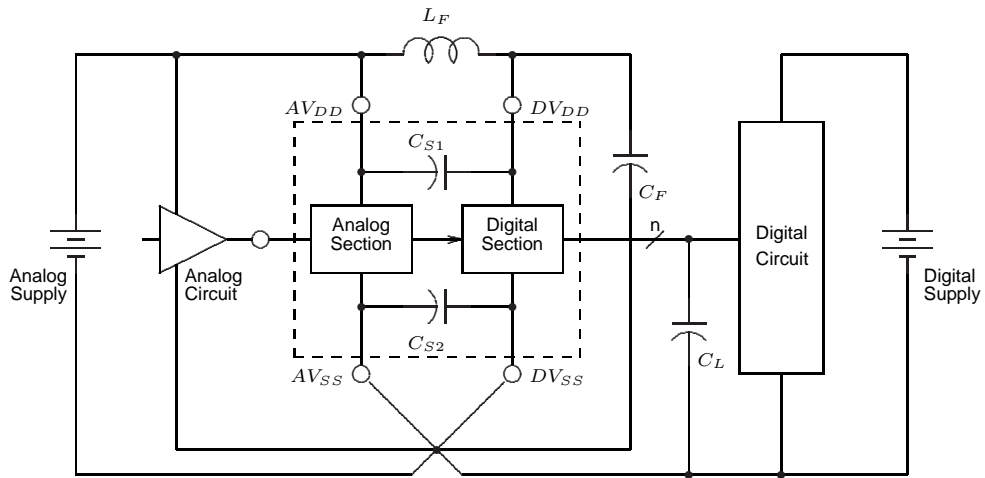


Figure 925: A/D Converter Internal Wiring

<sup>271</sup>The terms  $DV_{DD}$  and  $DV_{SS}$  are used when the A/D converter uses CMOS construction.  $DV_{DD}$  is the positive supply (+5 volts, for example) and  $DV_{SS}$  is the negative supply (usually the digital ground).



(a) Poor



(b) Improved

Figure 926: A/D Converter System Wiring

Notice that the noise problem worsens with increasing load capacitance  $C_L$ . If many logic devices must be driven by the A/D logic output, logic buffers should be used to minimize the load capacitance seen by logic outputs of the A/D converter.

### 34.14.2 Ground Plane and the A/D Converter

We will use the term *ground plane*, as a generic phrase for *area of copper on the PC board*. Large conductive areas for ground and power supply are attractive because they lower the impedance of the supply and ground. Under the right circumstances, a ground plane can reduce electromagnetic emissions from the circuit board.

There are two schools of thought on the design of ground planes.

On the one hand, authors such as Banyai and Gerke [282] recommend splitting the ground plane into areas that isolate signals - between the analog and digital sections, for example. This technique is often recommended for the design of ground planes around an A/D converter [301], [302]. The analog circuitry and digital circuitry each have their own separate ground plane, joined by a narrow conductor at the A/D mecca ground point. This narrow bridge (figure 927) has a high impedance for high frequency noise, preventing it from transferring from the digital ground plane to the analog ground plane. Care must be taken that traces from the digital circuit do not overlap into the analog circuit area.

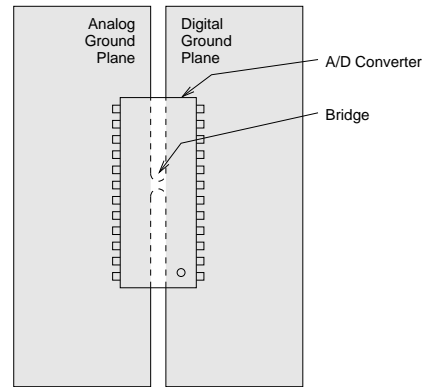


Figure 927: Segmented Ground

Ott [303] maintains that better control of emissions can be achieved with a single ground plane. The analog and digital circuits are each kept in separate areas, and care is taken that traces do not cross from one area into the other. Separate power planes (or traces) are used for the analog and digital supplies.

For high-resolution A/D conversion using multiple A/D converters, the situation is more complicated. All A/D converter grounds should attach to the mecca ground point, but this is physically impossible with multiple chips and a segmented ground plane. Ott cites this as a situation that requires a single continuous ground plane with careful component placement. Johnson [301] recommends segmentation and low-impedance ground, using external metal if necessary.

Simulation will eventually take the uncertainty out of this design issue. In the meantime, it is necessary to prototype a circuit, measure its performance, and be prepared to modify the design.

## 34.15 Case Histories: Solved

Here are the cures that were applied in the case histories of section 34 at the beginning of the chapter.

- *A certain car engine control system system quits while close to a radio transmission tower.*

The engine control system was susceptible to external electromagnetic radiation. External radio signals were picked up on cables attached to the system, and then carried into the electronics. Operational amplifiers rectified the RF signal and created DC output levels that caused the system to malfunction.

It was a surprise to find that the metal body of the car was an insufficient barrier to these signals. However, there was an opening at the front of the car, filled by a plastic grille. The opening was many wavelengths at various strong RF signals, and so it provided an opportunity for entry of the offending RF signal. The cure: Place the engine control system in a metal box and ensure that every line going into the box has an RF filter.

- *A development engineer finds that her signal acquisition system is swamped in noise from the on-board microprocessor.*

The board layout had intermingled the analog and digital circuitry, and the grounds were co-mingled. The cure: re-lay out the board, moving analog and digital circuitry into separate sections, well away from each

other. Organize the central ground point to be at the system A/D converter. Provide separate analog and digital grounds.

- *An airborne computer control system works fine on the bench when the modules are spread out. When the modules are installed in the case, noise from the switching power supply appears on the video output.*

Some exploration with electric field and magnetic field probes identified that the offending field was primarily electric in nature. (The switching power supply used a toroidal inductor, so the magnetic field was contained, as expected.) The cure: Place an electrostatic shield between the power supply and video circuitry. (As a temporary measure, the shield was simply a sheet of aluminum cooking foil glued to cardboard, then connected to the system ground by a wire lead.)

- *An industrial automation control system is being produced as a commercial product. When tested for compliance with emission regulations, it fails the test.*

Probing with a current probe, oscilloscope and spectrum analyser indicates a hot spot on the circuit board associated with the microprocessor clock signal. The clock lines are long and unterminated, and show evidence of ringing at high frequency. Cure #1: A terminating network is patched into the clock line to damp down the oscillation. This reduces the offending noise signal.

A measurement of the common-mode current on an external cable indicates that it is a major source of emissions. Cure #2: A clamp-on ferrite core is placed on the cable so that all the lines pass through the core of the cable. This reduces the noise emission to the point where the product is compliant with regulations.

- *An aerial photography camera is installed in an aircraft, powered from the 28VDC bus. Whenever the camera triggers, the aircraft navigation computer resets.*

Inspection of the aircraft wiring showed that the camera and nav computer were on the same electrical circuit. A bench test of the camera revealed that the starting current of the camera DC motor was in the order of 20 amps. This current surge created a large voltage drop across the aircraft wiring, a case of conducted interference. The cure: The camera connection was moved from the circuit that supplied the navigation computer and provided with separate wiring and circuit breaker back to the main 28VDC bus. A line filter was installed into the 28VDC power supply line at the camera to further contain transients in the area of the camera.

- *A film projection system is equipped with a magnetic head to read the sound track on magnetic tape. A nearby motor induces a power line interference signal in the magnetic head. The noise signal is stronger than the desired signal.*

Examination of the noise waveform with an oscilloscope shows a strong component at 60Hz and a smaller component at 120Hz. Probing the area with a search coil shows conclusively that the field is magnetic. The cure: Ideally, one would construct a metal shield (steel or mu-metal, to be effective at these frequencies) and enclose either the motor or pickup coil. Unfortunately, the physical arrangement and time constraints made this approach very difficult. Instead, two tuneable notch filters were constructed, one for each of the offending frequencies, similar to the circuits shown in section 17.14.1 on page 495. These were inserted in the audio signal path and tuned to give maximum attenuation of the noise signals. There was a minor effect on the low-frequency component of the program material, but overall the result was quite satisfactory.

### 34.16 Exercises

1. A 20MHz square-wave clock pulse waveform has an amplitude of 3.3 volts with transition times of 5nsec. Calculate the spectrum amplitude and the break frequencies.

2. By what factor is the resistance of a trace reduced if its width is increased by a factor of 16?
3. The figure shows a driver circuit of the type described in section 34.11.3 (page 983), coupled to the load resistance via a transformer.

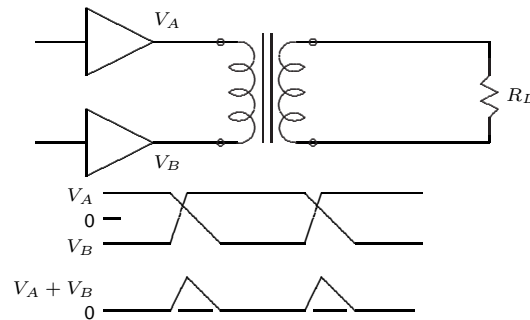


Figure 928: Unbalanced Differential Drive

When the common-mode current is measured in the cable driving  $R_L$ , it is found to be excessive.

- (a) Given that the transformer has a certain capacitance between its primary and secondary windings, use this and any other stray capacitances to explain why this common-mode current occurs.
  - (b) It's suggested that the common-mode current could be reduced by using a transformer which has an electrostatic shield between primary and secondary. Would this work? Explain.
4. A certain oscilloscope has an input resistance of  $1\text{M}\Omega$  in parallel with a capacitance of  $20\text{pF}$ . A 2 foot long piece of RG-174 coaxial cable ( $30.8\text{pF}$  per foot) connects the oscilloscope input into a test circuit.
    - (a) What is the effective capacitive load seen by the test circuit?
    - (b) What is the impedance of the load seen by the test circuit at a frequency of  $100\text{kHz}$ ?
    - (c) If the internal resistance of the test circuit is  $10\text{k}\Omega$ , what is the rise time of the measurement circuit?
    - (d) The oscilloscope used for this measurement has a 3db bandwidth of  $60\text{MHz}$ . Assuming a  $10\text{k}\Omega$  source resistance for the circuit, what is the 3db bandwidth of the measurement of this circuit?
    - (e) If a  $\times 10$  probe circuit is constructed at the test circuit end of the cable, what will be the effective capacitive load on the test circuit?
  5. Draw up a table showing the allowable amounts of common-mode current for the different frequency bands of the FCC Part 15 rules.
  6. The example of section 34.11.1 on page 981 calculated an allowable level of common-mode current in a cable. Assuming that this common-mode current is created by a 5 volt RMS AC signal driving current through a stray capacitance, calculate the magnitude of this capacitance. How does this value compare with the value of capacitor components?
  7. A certain vendor of shielding products provides a copper strip with fingers that mount into corresponding holes in a printed circuit board. The shielding effect is specified as '70 db attenuation of a magnetic field at  $200\text{kHz}$ '. The material thickness is 0.005 inches.
 

Could this level of attenuation be expected from ordinary copper sheeting, resistivity  $\rho = 1.589\mu\Omega\text{-cm}$ , permeability  $\mu_r \approx 1$ ? Explain.

8. In a laboratory setting, the temperature of a small oven is sensed by means of a thermistor. The resultant voltage signal is converted to digital format and recorded by a computer.

The temperature of the oven is controlled by a simple on-off thermostatic control. When the heater switches, the temperature recording system shows a large transient error.

- (a) Identify the possible sources of interference in this situation.
  - (b) How would you determine which of these possibilities is the offending problem?
  - (c) For each possible source of the interference, give the appropriate solution.
9. In pursuit of the lowest possible part cost in a switching power supply, the project supervisor has pointed out that a ferrite rod inductor is one-third the cost of an equivalent toroidal wound inductor. Why is a toroidal inductor a better choice, in spite of its higher part cost?
10. An airborne data acquisition system is being designed. It consists of inductive sensors which produce a low-level voltage signal, amplifier circuitry, analog-digital conversion, followed by a data-logging micro-processor system.

The system is being custom-built for a one-off experiment. The project deadline is very important. Cost is a secondary consideration.

The system circuit board can be built on one, two or four layers. Which would you recommend? Explain your choice.

The project leader has expressed a preference for a plastic enclosure in order to minimize the instrument package weight. What is your recommendation, and why?