12

Noise and stability

Capacitive sensor performance is limited by circuit noise from various sources, by mechanical stability, and by environmental factors. If these limits are understood and handled correctly, capacitive sensor designs can have exceptionally stable and low noise performance.

Circuit and component noise was briefly discussed in Section 10.4, and the level of detail in that chapter should be adequate for almost all applications, but for the small percentage of design jobs which "push the envelope," more information is presented in this chapter and the limiting sensitivity of capacitive sensors is analyzed.

12.1 NOISE TYPES

Five different types of noise can be found in sensor preamplifiers. All five types are present to some degree in typical circuits, but careful design can minimize the noise contribution of all except thermal noise.

Thermal noise

Thermal noise is broadband noise, having an energy which is a function only of the measurement bandwidth and not the frequency of measurement. It is generated by resistive components only; reactances do not generate thermal noise. It is caused by the thermally excited movement of molecules, and disappears at a temperature of absolute zero. Thermal noise generated by resistors follows a simple equation; thermal noise generated by semiconductors can also be predicted by understanding more complicated semiconductor processes. To simplify analysis, all semiconductor noise effects are given as equivalent input voltage and current noise sources.

Shot noise

Shot noise is simply due to the fact that the current through a semiconductor junction is quantized by the electron charge, and electron arrival time has a Gaussian distribution. Shot noise is generated any time that electron flow is gated by a valve which lets through one charge carrier at a time, such as the emission of an electron by a vacuum-tube cathode or a minority carrier in a transistor falling through a potential barrier. In our circuits, shot noise is caused by diodes and transistors. Conductors and resistors do not generate shot noise, as a fraction of a charge carrier can be transmitted by these devices. Shot noise I_s is characterized by this equation

$$I_s = \sqrt{2qI\Delta f} \qquad 12.1$$

$q = 1.60 \times 10^{-19}$ C, electron charge

Δf = measurement bandwidth, Hz

I = DC current through junction

1/*f* noise

Both resistors and semiconductors have a low frequency noise component which varies as approximately $1/f$. This noise is produced by a variety of process-dependent mechanisms and is characterized empirically.

Popcorn noise

Popcorn noise is found only in semiconductors, both bipolar and FET, and produces small, abrupt steps of output voltage at a low and random repetition rate. Popcorn noise was named by early users of operational amplifiers for audio amplifiers. They noticed a low level noise in the audio output which sounded like popcorn popping. Popcorn noise is associated with defects in the semiconductor crystal [Fish, p. 88] and has improved in recent years with increasing purity of semiconductor materials. It is a rapid change in a semiconductor bias current between two extremes (Figure 12.1). The level can be 5–20 times the Gaussian noise level. Popcorn noise will rarely be a problem with newer amplifier types.

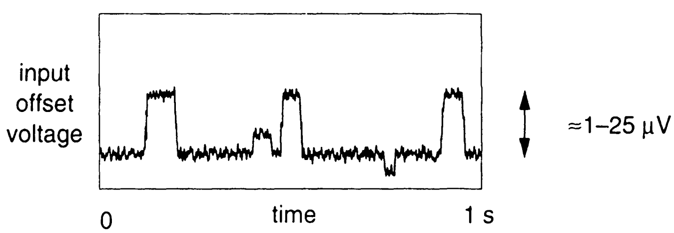

Figure 12.1 Popcorn noise

Crosstalk

Circuit crosstalk is the inadvertent coupling of voltages from other circuits to the sensor amplifier. Digital logic signals are particularly unfriendly, as crosstalk from a number of digital signals can look on an oscilloscope like random noise, and the signals can fall in the bandwidth of the capacitive sensor amplifier. Crosstalk is much easier to handle for high impedance capacitive sensor signals than for low impedance inductive sensor types, as the electric field is almost perfectly attenuated by conductive shielding while low impedance shielding usually must have magnetic field attenuation also.

12.1.1 Resistor noise

Resistors, which are carrying no current, produce thermal (Johnson) noise. DC voltage across a resistor causes an additional low frequency noise component with an approximately $1/f$ spectrum, called excess noise [Fish, 1994, pp. 184–190; Motchenbacher, 1993, pp. 289–296]. The excess noise changes considerably with the resistor size and construction method. The rms noise from these effects is

Thermal noise

$$e_t = \sqrt{4kTR(f_2 - f_1)} \quad \text{V rms} \qquad 12.2$$

where

$k = 1.38 \times 10^{-23}$, Boltzmann's constant, J/K

$T = 273 + {}^\circ\mathrm{C}$, absolute temperature, K ($4kT$ at 25°C = 1.64×10^{-20})

f_1 = lower 3 dB point, Hz

f_2 = upper 3 dB point, Hz

R = resistance, Ω

Excess noise

$$e_x = C_F Vr \sqrt{\log(f_2/f_1)} \quad \text{V rms} \qquad 12.3$$

where

Vr = voltage across resistor

C_F = excess noise factor, empirical

These equations assume a rectangular bandpass filter is used with a lower 3 dB frequency of f_1 and an upper 3 dB frequency of f_2.

The resistor excess noise factor C_F varies from 0.01–1 μV × V_{DC}^{-1} × (freq decade)$^{-2}$ [Fish, 1994, p. 88]. For a resistor with 1 V drop and 0.1 μV of excess noise in a decade of measurement bandwidth, C_F is 1. C_F is also called NI, or noise index, which is usually expressed in dB.

As an example, assuming $C_F = 10^{-7}$, temperature is 25°C, resistance is 1 MΩ, $Vr = 1$ V, and the 3 dB frequencies of the bandpass are 10 kHz and 20 kHz, the resistor noise is

$$e_t = 12.8 \times 10^{-6} \text{ thermal noise}$$

$$e_x = 3.01 \times 10^{-8} \text{ excess noise}$$

Shot noise

Shot noise will not be seen in a circuit with just a resistor, but if a diode or semiconductor junction is in series, the Gaussian shot noise component will be added, as current transfer through a potential well is quantized by electron charge. For a large value resistor the shot noise contribution is insignificant, less than 60×10^{-12} for the 1 MΩ resistor above.

Excess noise is Gaussian and has a noise spectrum which falls off with frequency

$$e_x = \frac{K_f}{f^{\alpha}} \qquad 12.4$$

where α is usually in the range of 0.8–1.4 [Fish, 1994, p. 85] for semiconductors and 0.8–1.2 for resistors. We use the median value of 1. With $\alpha = 1$, the rms value of the excess noise measured in the bandwidth between 1 Hz and 10 Hz is the same as the rms excess noise measured in a bandwidth of 10 Hz to 100 Hz. Since excess noise is proportional to the number of decades of measurement frequency bandwidth and is independent of where in the spectrum the bandwidth is measured, it is specified as microvolts per volt in one decade. If the circuit bandwidth is not one decade, the noise is scaled as

$$e_t' = e_t \sqrt{\frac{\log(f_2/f_1)}{\log(10)}} \qquad 12.5$$

Philips [*Precision MELF SMD Resistors*, 1992, p. 106] shows excess noise curves for two sizes of surface mount resistors. The small size, 0805, has an excess noise of 0.5 μV/V in a decade at the 100 kΩ value; the larger 1406 size shows a value of 0.15 μV/V for the same conditions. This reference also shows an increase in excess noise with resistance. Excess noise increases as $\sqrt{R}$ for values under 500 kΩ and as about $R^{0.75}$ for values above 500 kΩ. The noise of a 1406 size 5 MΩ precision resistor in one decade of frequency, from this reference, is 2 μV/V. For this resistor,

- Thermal noise = 28 μV
- Excess noise = 10 μV (1 kHz – 10 kHz bandwidth, 5 V DC)

The excess noise performance of all construction methods worsens at high resistance values and small sizes. Also, the shape of the excess-noise-vs.-resistance graph is unpredictable; at 1 kΩ a carbon film resistor [Motchenbacher, 1993, p. 295] has an excess noise of –37 dB, or –97 dB below the applied voltage (add 60 dB as the excess noise C_F is μV/V), while a 1/2 W carbon composition resistor has –33 dB excess noise. At 10 MΩ, however, these figures worsen and reverse, with the carbon comp resistor at –10 dB and the carbon film resistor at +5 dB. The lesson for capacitive sensor preamps is to *design the input circuit so high-value resistors have no voltage drop*. This is not difficult if low bias current op amps or FETs are used.

12.1.2 FET noise

Transistor noise has three components:

- Excess noise is a problem at frequencies of less than 10 Hz–10 kHz. It follows an approximate $1/f$ characteristic.

- Thermal noise is independent of frequency, and is defined as the noise voltage found in 1 Hz of bandwidth. It increases as the square root of bandwidth, and is caused by the thermal noise of the FET channel resistance.
- Shot noise is independent of frequency like thermal noise.

The total noise increases at high frequencies, in excess of 100 kHz–10 MHz, primarily due to current noise increase as thermal noise in the channel resistance is coupled back to the gate by Miller capacitance (Figure 12.2).

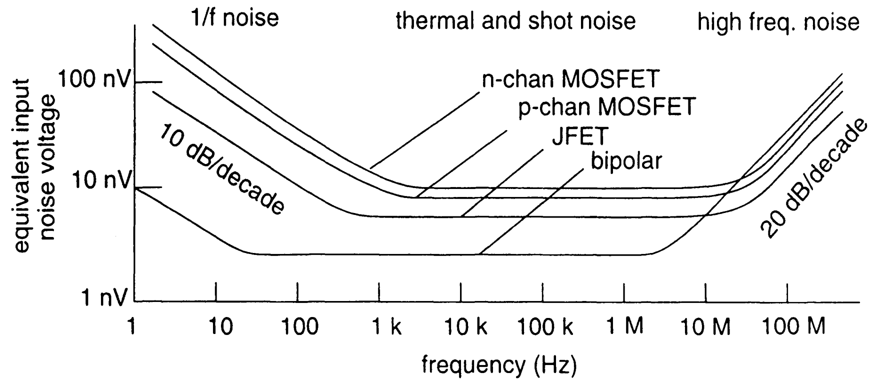

Figure 12.2 Transistor noise

These curves are very approximate; individual transistor processes can result in 10× more noise than shown. GaAs FETs are not shown, but are not useful for capacitive sensors as they have a very high $1/f$ noise corner, in the range of 1 MHz–10 MHz.

Current noise is 3–5 orders of magnitude higher in bipolar transistors than in JFETs or MOSFETs, rendering them unacceptable for our high-impedance sensing problems; they are shown for comparison only.

Transistor noise can also be separated into voltage and current noise, and these components behave differently vs. frequency for FET transistors [Motchenbacher, 1993] (Figure 12.3).

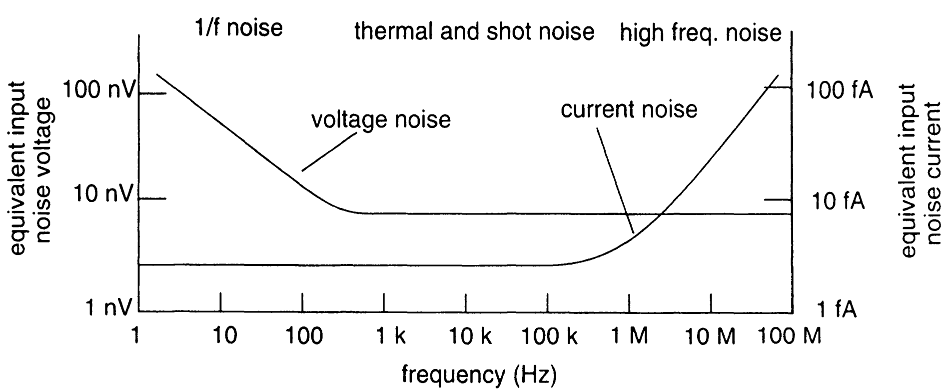

Figure 12.3 FET noise

Bipolar transistors have a total-noise-vs.-frequency characteristic similar to FETs, but the components are interchanged. For bipolars [Motchenbacher, 1993, p.126] the noise voltage shows no $1/f$ noise, except perhaps at high currents and low frequencies, but it increases for frequencies above the transistor f_T, and the noise current shows $1/f$ behavior as well as a high frequency rise. For bipolars and FETs, the $1/f$ portion of the curve above graphs noise *power* changing as $1/f$, so noise voltage changes as roughly $1/\sqrt{f}$ or 10 dB/decade, although process variations may produce a much lumpier curve. The high-frequency slopes are all at the 20 dB/decade rate of a one-pole response.

Table 12.1 Discrete transistor noise measurements

	Bipolar	MOSFET	JFET	JFET op amp
E_n, nV/$\sqrt{\text{Hz}}$, 10 kHz	0.9–3.6	5–6	0.8–6	10–43
I_n, pA/$\sqrt{\text{Hz}}$, 10 kHz	1–4	0.0002–0.0007	0.0035–0.035	0.0001
$1/f$ corner of E_n	none	10 kHz–500 kHz	1 kHz–10 kHz	
High freq. corner of E_n	100 kHz–10 MHz	none	none	
$1/f$ corner of I_n	160–3000 Hz	none	none	
High freq. corner of I_n	none	10–200 kHz	10–500 kHz	
Tested types include	2N4250, 2N4403, 2N4058, 2N4124, 2N4935, 2N4044	2N3631 HRN1030	2N2609, 2N3460,2N3684, 2N5116,2N5266, 2N5394, C413N, UC410	see Table 10.2
Lowest noise type	2N4403	2N3631	C413N	

A sample of low noise transistors was tested by Motchenbacher [1973, pp. 78, 106] and excerpted here in Table 12.1. For a discrete FET, the equivalent input-referred thermal-noise resistance can be estimated as 0.66 times the reciprocal of transconductance, $0.66/g_m$. As an example, a FET transistor with a g_m of 1 S (1 mho) is equivalent to a noiseless FET with a thermal-noise-generating 1 kΩ resistor in series with the output. The input-referred $1/f$ noise of a MOSFET, E_n, is proportional [Takahashi, p. 390] to

$$E_n = \sqrt{\frac{d^2}{\varepsilon^2 WLf}} = \frac{\sqrt{WL}}{C} \qquad 12.6$$

E_n is the rms spot noise at frequency f, d and ε are the thickness and the dielectric constant of the gate insulation, W and L are the width and length of the channel, and C is the gate capacitance. As the WL product increases, noise decreases. Unfortunately, many applications cannot take advantage of a large area because the input capacitance also increases directly with WL. Thermal noise decreases directly as the conductance g_m increases, and as g_m is directly related to W/L, thermal noise will decrease as $W^{3/2}L^{-1/2}$.

A more complete description of MOSFET input noise [Motchenbacher, 1993, p.150; Sze, p. 496] is

$$E_{ni}^2 = \frac{0.66}{g_m} \cdot 4kT + \frac{K}{C_{ox}WLf} \qquad 12.7$$

The first term is thermal noise and the second term is $1/f$ noise.

$$I_{ng}^2 = 2qI_{DC} \qquad 12.8$$

E_{ni} = input-referred voltage noise, rms, per $\sqrt{\mathrm{Hz}}$
0.66 = empirically determined noise constant, ranging from 0.5–1
g_m = transconductance, 1/Ω or S (Siemens)
kT = Boltzmann constant × temperature, K; 4.14×10^{-21} at 27 °C
K = flicker noise coefficient, in the range of 10^{-30}
C_{ox} = capacitance per unit area of oxide, farads per μm^2
I_{ng} = gate noise current, rms per $\sqrt{\mathrm{Hz}}$
q = electron charge, 1.602×10^{-19} C
I_{DC} = gate leakage current

A typical low-noise MOSFET with gate leakage of 100 fA, $C_{ox} = 0.7$ fF/μm^2, $W/L = 50$, and a channel length L of 3.2 μm was calculated [Motchenbacher, 1993, p. 151] as

$$E_{ni} = 8.19 \text{ nV}/\sqrt{\mathrm{Hz}}$$

$$I_{ng} = 0.18 \text{ fA}/\sqrt{\mathrm{Hz}}$$

These specifications are typical of n-channel JFET low noise amplifiers (Table 12.2).

Table 12.2 N-channel JFET low noise amplifier transistors

Type	2N5558	2N5105
Process	NJ16	NJ26AL
Gate leakage, typ	10 pA	10 pA
Transconductance g_m	0.0022 S	0.009 S
Input capacitance	3 pF	5 pF
Feedback cap.	1 pF	1.5 pF
Noise voltage, *En*	6 nV/$\sqrt{\mathrm{Hz}}$	1 nV/$\sqrt{\mathrm{Hz}}$
$1/g_m$	454 Ω	111 Ω
Equiv. noise resistor	375 Ω	62.5 Ω

The equivalent noise resistor is calculated from En and In using eq. 12.2.

Low-Z amplifier

The noise equivalent circuit of the inverting or low-Z amplifier is shown in Figure 12.4. If the amplifier gain is high, the output noise voltage assuming uncorrelated noise sources and measured in a bandwidth Δf at a frequency ω higher than the $1/RfCf$ pole is

$$Eo^2 = En^2\left[1 + \frac{Cs + Cg}{Cf}\right]^2 + \frac{4kT\Delta f}{\omega^2 Cf^2 Rf} + \left[\frac{In}{\omega Cf}\right]^2 \qquad 12.9$$

Noise voltages are in volts $\sqrt{\text{Hz}}$ and noise currents are in amps $\sqrt{\text{Hz}}$ rms. The first term is the effect of the input voltage noise, the second is the contribution from the thermal noise of Rf, and the third is the effect of the input current noise. If the resistor Rf is large compared to the sensor impedance, the equation becomes

$$Eo^2 = En^2\left[1 + \frac{Cs + Cg}{Cf}\right]^2 + \left[\frac{In}{\omega Cf}\right]^2 \qquad 12.10$$

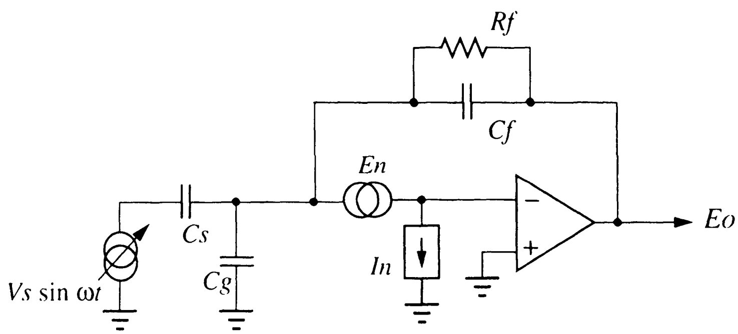

Figure 12.4 Inverting amplifier noise equivalent circuit

12.1.3 High-impedance amplifier

The noninverting or high-Z amplifier noise circuit is shown in Figure 12.5.

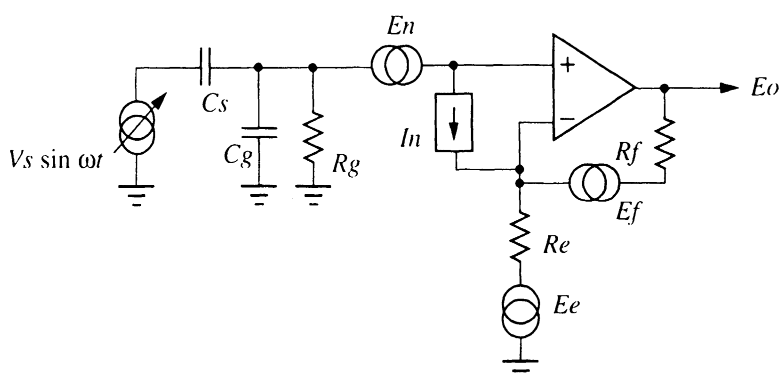

Figure 12.5 Noninverting amplifier noise equivalent circuit

Noiseless feedback resistors Re and Rf have been added to establish a gain $G = 1 + Rf/Re$, and each has a thermal noise generator in series. The output noise of this circuit, assuming the impedance of the parallel combination of Cs, Cg, and Rg is greater than Re and defining $Ct = Cs + Cg$, is

$$Eo^2 = Ef^2 + G^2En^2 + \left(Ee\frac{Rf}{Re}\right)^2 + G^2\frac{4kTRg \cdot \Delta f}{(1+sRgCt)^2} + G^2In^2\frac{Rg^2}{(1+sRgCt)^2}$$

Simplifying the circuit with a gain of 1, with $Rf = 0$ and Re = infinity, and again assuming that Rg is large enough so that its noise contribution is negligible, we have

$$Eo^2 \cong En^2 + \left[\frac{In}{\omega(Cs+Cg)}\right]^2 \qquad 12.11$$

If Cf is set equal to Cs so that the gains of the two circuits are equal, and if $Cg = 0$, the high-Z amplifier has a 2× advantage in output signal-to-noise ratio. If, however, a shield is used with a capacitance Cg to the pickup electrode, and it is either grounded for the low-Z amplifier or used as the feedback capacitor for the high-Z amplifier, the output noise for the two circuits is identical. As shields are almost always used, changing from high-Z to low-Z does change gain and output level, but not signal-to-noise ratio.

The feedback amplifier noise performance is similar to the high-Z circuit.

12.2 LIMITING DISPLACEMENT

The just-measurable limiting displacement in a capacitive position sensor x_n is the displacement which produces a voltage change equal to the rms noise voltage in the bandwidth of the amplifier circuit. The just-measurable displacement in meters is defined as

$$x_n = \frac{Eno}{G} \qquad 12.12$$

where G is the responsivity of the sensor in volts per meter of displacement at the amplifier output and Eno is amplifier output noise.

12.2.1 Limiting displacement of three-plate micrometer

Jones [1973] defines the limits of performance of three-plate micrometers. With a low-Z amplifier, we have the circuit shown in Figure 12.6. With frequency high enough so that the crossover resistor En/In is much larger than the impedance of $C1 + Cd$, the capacitance change in $C1$, $\Delta C1$, which produces an output equal to the rms noise is

$$\Delta C1 = (C1 + Cd)\frac{En}{0.707V} \qquad 12.13$$

with En evaluated in the bandwidth of BPF.

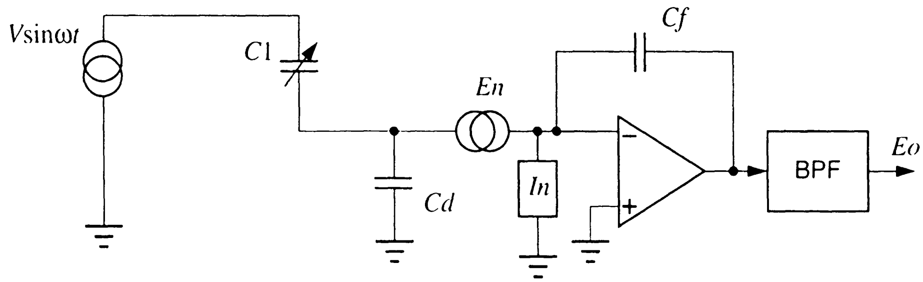

Figure 12.6 Micrometer circuit

With small sensor capacitance and low excitation frequency the current noise *In* may not be small enough to be neglected, and consideration should be given to increasing the excitation frequency. To detect small changes of absolute capacitance, increase excitation voltage *V*, use a low noise transistor, and reduce *C*1 until *C*1 is approximately equal to the stray capacitance *Cd* or until the current noise contribution becomes the dominant noise source.

12.2.2 Maximizing limiting displacement

If $C1 = Cd = 1$ pF and the excitation voltage is 100 V, with a 1 nV/$\sqrt{\text{Hz}}$ amplifier and a bandwidth of 1 Hz, the highly theoretical minimum detectable absolute capacitance is an exhilarating 2×10^{-22} F, 0.0003 aF. This represents a 2×10^{-10} change, so with a plate spacing of 1 mm a displacement of 2×10^{-13} m, 0.04 light wavelengths, is just detectable.

Coupling transformer

A largely theoretical option is to use a wideband input coupling transformer. Ideally, an amplifier with the lowest possible input noise *power* ($In \times En$) should be selected and the sensor impedance matched to the amplifier's *En*/*In* using a transformer. The transformer turns ratio should be the square root of the ratio of the sensor impedance to *En*/*In*. As this amplifier selection criterion would result in very high impedance, several hundred MΩ, the transformer would need an inductance of at least 600 H at a 25 kHz carrier frequency along with very low parasitic capacitance. This is an unrealizable component. Adding a transformer to the drive to increase excitation voltage is preferable, at least until air breakdown happens.

Tuned transformer

A more realizable alternate is a tuned input transformer. This should be used with sine wave drive, or at least with a demodulator which accepts sine wave signals, and it optimally has a turns ratio selected as above. If the transformer inductance is tuned with a shunt capacitor at the signal frequency, any small value of inductance is resonated to a high impedance if the inductor's *Q* is sufficiently high. The tuned transformer circuit has two useful features, as it both reduces noise by bandwidth reduction by a factor proportional to *Q*, and it matches amplifier noise resistance. Jones [1973, p. 594] uses a tuned transformer and a 100 V rms sine wave drive signal to achieve these impressive performance numbers

- Transformer Q100
- Output resolution, 1 Hz bandwidth0.01 aF ($0.01 \cdot 10^{-18}$ F)
- Minimum detectable motion$5 \cdot 10^{-6}$ μm

Tuning the input with a Q of 100 can produce a circuit with an equivalent noise resistance of 100 Ω [Jones, p. 595]. The equivalent noise voltage of this resistance is, from eq. 10.9 above and using a bandwidth of 100 Hz, 12.9 nV, for a 170 dB dynamic range in a 5 V system. This performance is only a factor of 30 from the theoretical limit.

12.3 SHIELDING

Capacitive sensors often work with sensor capacitance of 1 pF or less. Shielding the sensor plates from extraneous fields is critical for good performance. Even though a synchronous demodulator may reject interfering signals, the sensitivity of the input circuit may be so high that out-of-band signals may saturate the input amplifier. Shielding in general must attenuate both electric and magnetic fields, but because of the very high impedance of capacitive sensor inputs, electric fields are considerably more important. A strong magnetic field which couples an interfering signal to a circuit loop can affect sensor circuits; but, for sensitive PC board circuits, conductors can be paired and "twisted" using vias to avoid loops.

Electric shield effectiveness (S) is measured in dB by

$$S = 20 \log \frac{E_1}{E_2} \qquad 12.14$$

where E_1 is the incident field and E_2 is the attenuated field inside the shield. A conductive shield with thickness d attenuates electric fields by absorption A and by reflection R. The distance that the field penetrates into the shield is measured by skin depth δ, and absorption of a shield of thickness d is calculated as 8.69 d/δ so it increases by 8.69 dB ($1/e$, or 37%) with every skin-depth increase in thickness. The total shield effectiveness is the sum of this absorption and the attenuation due to reflection; when a plane shield is close to the source of the radiation the shield effectiveness S is

$$S = A + R = 8.69\frac{d}{\delta} + 322 - 10 \log\left[f^3 r^2 \frac{\mu_r}{\sigma_r}\right] \quad \text{dB} \qquad 12.15$$

d = thickness, m

δ = skin depth, m

f = frequency, Hz

r = distance from shield to source, m

μ_r = permeability of shield relative to vacuum

σ_r = conductivity of shield relative to copper

Although the absorption term A is independent of the distance of the radiator, the reflection loss measured by the second and third terms is not. The r^2 term is due to the change in the impedance of a dipole radiator, from very high near the dipole to 377 Ω at $\lambda/2\pi$ m away, where λ is the wavelength. The skin depth δ is calculated as

$$\delta = \sqrt{\frac{1}{\pi f \mu \sigma}} \quad \text{m} \qquad 12.16$$

For copper, $\mu = \mu_0 = 4\pi \times 10^{-7}$ and $\sigma = \sigma_c = 5.75 \times 10^7$ mho/m; eq. 12.16 becomes

$$\delta = \frac{0.0664}{\sqrt{f}} \quad \text{m} \qquad 12.17$$

At 25 kHz, the skin depth of copper is 0.420 mm. For 0.420 mm thick (one skin depth) copper and with $r = 1$ m, eq. 12.15 evaluates as

$$S = 8.69 + 322 - 132 = 199 \quad \text{dB} \qquad 12.18$$

or about one part in 10^{10}. With increasing frequency, the absorption term increases as $\sqrt{f}$ while the reflection term decreases, but at 250 kHz S is still 188 dB.

As the source distance increases past $\lambda/2\pi$, eq. 12.15 is inaccurate, and the far-field equivalent is used

$$S = A + R = 8.69\frac{d}{\delta} + 168 - 10 \log\left[f\frac{\mu_r}{\sigma_r}\right] \quad \text{dB} \qquad 12.19$$

The reflection loss of a copper shield in the far field varies from 150 dB at DC to 88 dB at 100 MHz. For more details on shielding effectiveness, see Ott [1988, pp. 159–200].

A graph of eq. 12.19 with $d = 0.42$ mm and $r = 0.1$ m shows the behavior of the two components of the shield attenuation (Figure 12.7).

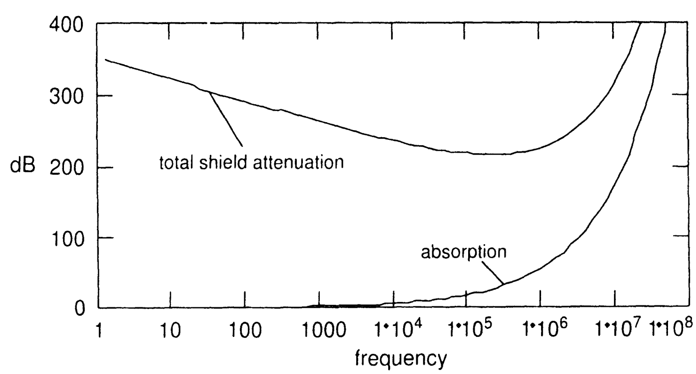

Figure 12.7 Shield attenuation vs. frequency

As shielding effectiveness is only fractionally lower with any reasonably conductive material, shield design is reduced to making sure that the sensor is completely surrounded with shield. If an opening is needed, forming a tube or channel instead of a simple cutout works better. A channel length-to-width ratio of five or more will attenuate low frequency incident fields about as well as the unopened shield.

12.4 GROUNDING

Noise reduction and low noise transistors can be sabotaged by a bad grounding scheme. A good ground, like security and freedom, cannot be won or lost; it must be won and rewon many times. A proper treatment of grounding is found in Ott [1988] and Morrison [1977], but a brief review is presented here for designs where analog capacitive sensors share a circuit board with digital logic.

12.4.1 Ideal ground

A poor ground is contaminated by voltage gradients caused by return currents from the analog circuit or from nearby digital circuits; a good ground is all at the same potential. Signal coupling from the same circuit can cause high frequency oscillation, low frequency oscillation (motorboating), or hysteresis, while coupling from a digital circuit can look like repetitive impulse noise with simple digital circuits or Gaussian noise for more complex digital circuits.

At first glance digital noise due to crosstalk may be indistinguishable from Gaussian preamplifier noise, but digital noise can be identified as it increases with frequency at about 6 dB/octave. This high frequency peaking is because digital noise generally couples through the reactive ground impedance or a reactive capacitive coupling; each of these coupling mechanisms results in 6 dB/octave spectral shaping. Digital noise is the most troublesome for low level circuits. Capacitive coupling is easily handled by separation, guarding, or shielding, but reducing ground coupling is more difficult.

There are three ways to establish an ideal unipotential ground: zero impedance, zero current flow, or correct connection of grounds.

Zero impedance

In any real circuit, some current must flow in the ground connections, so reducing the impedance of the ground connections as much as possible will improve performance. For capacitive sensors, which are much more sensitive to electric fields than magnetic sensors, a continuous ground plane under the preamplifier circuits is best. Failing that, a well-connected mesh of wires or circuit traces should be used. For low frequency signals, below, say, 20 kHz, the resistance of the plane is important, but at higher frequencies, skin effect predominates and any thickness over 0.03 mm is adequate. If a ground plane is used without special routing consideration for return currents, a 20 dB advantage over a two-layer board will be realized, but improvements past that figure will usually require rerouting of return currents away from the affected ground.

Zero current flow

Zero current flow can be approximated by ensuring that all signal paths are terminated in infinite impedance receivers. This will guarantee that no current circulates in the ground, and the ground will be ideal. This can be approximated by using high impedance receivers for all signals, but this technique fails at high frequencies where transmission lines need to be terminated properly for good signal integrity. Also, some loads may be low impedance, and providing a high impedance receiver for each load would add circuit complexity. Another factor which mitigates against zero current flow is the ground current caused by signals coupling to ground through parasitic capacitances.

Correct connection

Even when zero current flow cannot be arranged, correct circuit design can guarantee that currents do not flow in the low level analog ground system, or if they do, they flow through zero impedance connections. This is demonstrated by first analyzing a poor connection (Figure 12.8).

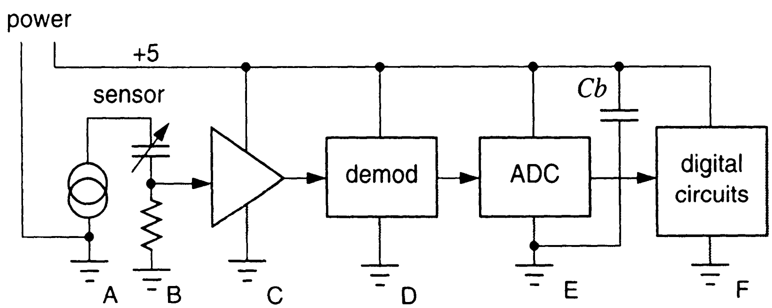

Figure 12.8 Poor ground connection

This circuit violates all the rules:

- No instructions for PC layout are included; the worst case PC ground connection, a single trace connecting grounds *A* through *F*, may result
- Power supply current to all circuits returns through connection *ABC*. A voltage drop equal to the ground resistance between *A* and *C* times the power supply current is added to the sensor voltage.
- Digital power current through bypass capacitor *Cb* flows through the ADC ground
- Analog circuits use noisy digital power

A better connection is shown in Figure 12.9. This circuit is considerably improved. No current flows through connection *A-B-C* to degrade the low-level analog performance except power supply returns from those circuits. The noisy digital +5 voltage has been filtered by *Rb-Cb*2, and *Cb* now keeps the digital power supply bypass currents in a small local loop. The ground connection is ambiguous, however, and should be formalized so the PC board layout is deterministic, as shown in Figure 12.10.

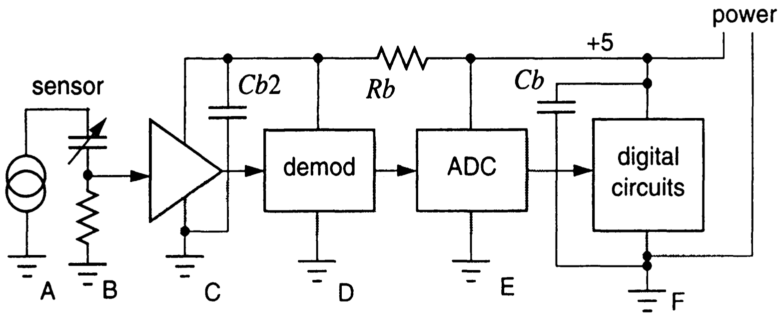

Figure 12.9 Better ground connection

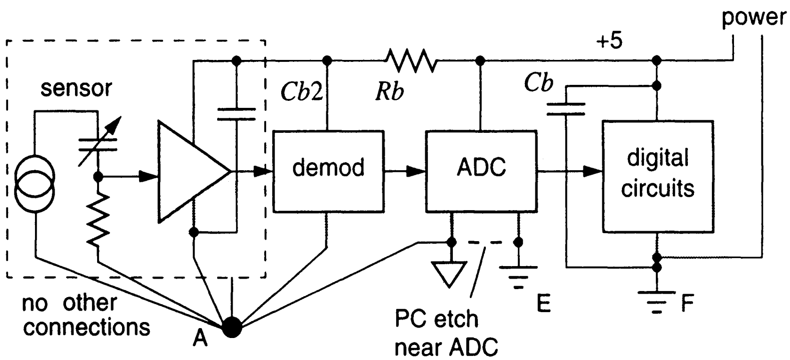

Figure 12.10 Best ground connection

Here the ground connections are unambiguous, an analog ground symbol is added so the PC layout process is forced to keep analog and digital grounds separate, a single ground node is used for the analog circuits, and a single connection between analog and digital grounds is specified. A grounded shield connects to the analog ground node.

12.5 STABILITY

Capacitive sensors are constructed with two or more conducting electrodes and an insulating support. These components will change size in response to environmental factors, and long-term stability may be a concern. These effects were briefly discussed in "Limits to precision" in Section 3.4, and the effect of environmental variables on air is found in "Dielectric constant of air" in Section 6.3.2. The stability of carefully constructed micrometers and tiltmeters [Jones, 1973] was seen in Section 9.13 to be about one part in 10^9 per day.

12.5.1 Temperature

Conductors

Most common metals and alloys have a temperature coefficient of linear expansion in the range of 9–32 μm/m/°C (5–18 μin/in/°F). Silicon, at 5 μm/m/°C, is quite stable, and specially formulated alloys such as Invar have even lower coefficients. Achieving the theoretical limiting displacement calculated above as 3×10^{-10} would require a stability of 3×10^{-4} µm/m. With a stable material such as silicon, a temperature change as low as 6×10^{-5}°C will cause an expansion equal to the limiting displacement if the plate spacing is determined directly by the silicon.

Invar [Harper, 1993, p. 5.44] is an alloy composed of 36% Ni, 0.4% Mn, 0.1% C, with the remainder Fe. Its coefficient of linear expansion between 0 and 38°F is 0.877 µm/m/°C, but at higher and lower temperatures its stability degrades to a pedestrian 10–15 µm/m/°C. Its coefficient is affected by annealing and cold working and may be artificially reduced to zero by these methods, but time and temperature will cause the coefficient to return to about 0.877.

Insulators

Insulators have a much wider range of thermal expansion than metals (Table 12.3).

Table 12.3 Temperature coefficient of some insulators

Pyrex glass	3.2 ppm/°C
Diamond	1.1
Fused quartz, SiO_2	0.5
Silicon nitride	2.0
Soda lime glass	9.2
Borosilicate glass	3.3
Acrylic	50–90
Epoxy	45–65
Polyamide	55
Polyimide	20
Polyester	30
TFE	100

Data from Harper [1970, p. 2.65] and *CRC Handbook for Applied Engineering Science* [1973, pp. 152–153].

For applications in high-temperature environments, TFE and polyimide have upper temperature limits of 260°C.

12.5.2 Stability with time

Long-term stability of materials is a concern for very sensitive applications. Standard meter bars machined from platinum-iridium alloy and bronze were found to change by three parts and 20 parts in 10^{-7}, respectively, in a 52 yr period. Gage blocks stable to one part in 10^{-7}/yr have been built from heat treated stainless steel and titanium carbide, and 70-30 annealed brass is also very good. Light alloys are unstable.

12.5.3 Stability with humidity

Metals have insignificant water absorption and chemical absorption and are generally quite stable to any chemical challenge other than acid. Most insulators, however, absorb water from the air during high humidity and expand. In the case of wood crosswise to the grain, this expansion is at its maximum, reaching 1/8 in per 1.5 ft in my kitchen cabinets. Glass and fused silica do not absorb water; a few plastics also do not absorb water, such as TFE (Teflon™) and polyimide (Kapton™), and some absorb very little, like mica. For the plastics, polyimide would be a good choice of insulator for critical capacitive sensor applications, with low temperature coefficient and low water absorption, excellent dielectric strength (5400 V/mil in thin layers), low dielectric constant (3.5), high volume resistivity (10^{18}), reasonably low loss tangent (0.003), and excellent tensile strength and abrasion resistance.

Another possible stability problem is in applications where insulators are exposed to chemicals or chemical vapor; for example, plastics like ABS or styrene will expand in the presence of organic solvents. Good choices for fluid chemical resistance are polyethylene, TFE, fluorinated ethylene propylene (FEP), and polyimide.