

Capacitive Sensors

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Capacitive Sensors

Design and Applications

Larry K. Baxter



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Robert J. Herrick, Series Editor

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Preface

Capacitive sensors can solve many different types of sensing and measurement problems. They can be integrated into a printed circuit board or a microchip and offer non-contact sensing with nearly infinite resolution. They are used for rotary and linear position encoding, liquid level sensing, touch sensing, sensitive micrometers, digital carpenter's levels, keyswitches, light switches and proximity detection. Your telephone and tape recorder probably use electret microphones with capacitive sensing, and your car's airbag may be deployed by a silicon accelerometer which uses capacitive sensing. The use of capacitive sensors is increasing rapidly as designers discover their virtues.

Capacitive sensors can be unaffected by temperature, humidity, or mechanical misalignment, and shielding against stray electric fields is simple compared to shielding an inductive sensor against magnetic disturbances. Capacitive transducer accuracy is excellent, as the plate patterns which determine accuracy can be reproduced photographically with micron precision. The technology is easily integrated, and is displacing traditional silicon-based transducers using piezoresistive and piezoelectric effects, as the sensitivity and stability with temperature are ten times better. Capacitive sensors consume very little power; battery life for small portable products may be several years.

Several factors inhibit the use of capacitive sensors, including the specialized circuits needed and the lack of understanding of the technology, including a widely held superstition that capacitive sensors are nonlinear and cannot operate at extremes of humidity. But many products which embed capacitive encoders have been successful in the market. This book surveys different types of sensors and shows how to build rugged, reliable capacitive sensors with high accuracy and low parts cost. Several product designs which use capacitive sensors are analyzed in detail.

The book is organized as follows.

- **Basics** This section covers theoretical background, different electrode geometries, and basic circuit designs.

- **Applications** Four different uses of capacitive sensors are presented: micrometers, proximity detectors, motion encoders, and some miscellaneous sensors. The theory behind these different uses is discussed.
- **Design** The electrode configurations and the basic circuit designs which were briefly discussed in the Basics section are more fully explored here.
- **Products** This section presents design details of several different products which use capacitive sensors.

Thanks are due to John Ames and Rick Grinnell, who helped with writing and review; Katie Gardner, who handled editing, research, and graphics; and my wife Carol, who helped in many other ways.

*Larry K. Baxter
Gloucester, Massachusetts*

Capacitive Sensors

1

Introduction

Capacitive sensors electronically measure the capacitance between two or more conductors in a dielectric environment, usually air or a liquid. A similar technique is electric field measurement, where the electrostatic voltage field produced by conductors in a dielectric environment is picked up by a probe and a high impedance amplifier.

A high frequency excitation waveform is normally used, as the reactive impedance of small plates in air can be hundreds of megohms at audio frequencies. Excitation frequencies of 100 kHz or more reduce the impedance to a more easily handled kilohm range.

Some systems use the environment as a return path for capacitive currents; a person, for example, is coupled capacitively to the building or the earth with several hundred pF. More often, the return path is a wire.

The range of applications is extreme. At the small end, silicon integrated circuits use the capacitance between micromachined silicon cantilevers to measure micrometer displacements. The dimensions are on the order of 100 μm , and the gap can be a μm or less. At the large end, capacitive personnel and vehicle detectors have plate dimensions of several meters. Capacitive sensors have been used for video-rate signals in one of the early video disk storage technologies. At the sensitive end, one researcher [Jones, p. 589] reports on an instrument with movement sensitivity of 10^{-11} mm.

1.1 SENSOR TRENDS

1.1.1 Market

The world market for sensors [*Control Engineering*, 1993] is over \$2.3 billion, growing at an annual rate of 10%. Consumer products, including imaging sensors, make up 64% of the market, with accelerometers accounting for 19% and a variety of specialized sensors completing the total.

Venture Development Corp., Natick, MA, (508)653-9000, in *The U.S. Market for Proximity, Photoelectric, and Linear Displacement Sensors* (1995), forecasts that market segment's size to grow from \$429.6 million in 1995 to \$548 million in 1999. U.S. consumption in 1999 is forecasted as

Magnetic	1.2%
Ultrasonic	1.9%
Capacitive	3.1%
Magnetic-actuated	4.7%
Inductive	42.2%
Photoelectric	46.9%

1.1.2 Technology

Sensor technology is in the process of a slow migration from discrete “dumb” instruments, expensive and inflexible, to smart, self-calibrating, silicon-based units, and the measurement method of choice for silicon devices as well as for discrete instruments is moving from a variety of transducer technologies, such as magnetic, optical, and piezoelectric, to capacitive.

Capacitive sensors can be used for many different applications. Simple sensors are used for go-no go gaging such as liquid level in reservoirs, where their ability to detect the presence of a dielectric or conductor at a distance allows them to work through a nonconducting window. Similar devices (StudSensor) can find wood studs behind plaster walls. Very high analog accuracy and linearity can be achieved with two- or three-plate systems with careful construction, or multiple-plate geometries can be used with digital circuits to substitute digital precision for analog precision.

Capacitive sensors are also useful for measuring material properties. Materials have different values of dielectric constant as well as dielectric loss, and the values of both properties (particularly dielectric loss, or “loss tangent”) change with temperature and frequency to give a material a characteristic signature which can be measured at a distance. Appendix 2 lists the dielectric properties of different materials vs. frequency.

1.1.3 Integration

Tools are available to fabricate many different types of sensors on silicon substrates, including micromachining which etches silicon wafers into sensor structures like diaphragms and cantilevered beams (Chapter 15). The decision to integrate a sensor can be done in stages, with the sensor and its electronics integrated together or separately, and usually is considered when the volume reaches 250,000 per year for total integration. Advantages of integration on silicon [Keister, 1992] are

- Improved sensitivity
- Differential sensors compensate parasitic effects
- Temperature compensation
- Easier A-D conversion
- Batch fabrication for lower cost

- Improved reliability
- Size reduction
- Less assembly required
- Lower parasitic capacitances for smaller capacitive electrodes

Capacitive technology is particularly useful for integration, as it is temperature-stable and problems of parasitic capacitance, lead length, and high impedance nodes are more easily handled on a silicon chip than on a PC board.

1.1.4 Intelligence

Smart sensors increase their value to the system by handling low level tasks such as calibration, self-test, and compensation for environmental factors. The gage factor of smart sensors can be scaled automatically so that one sensor type can be used in many locations, and sensor-resident software may help with communications by managing its end of a sensor data bus.

1.1.5 Sensor bus

Most sensors use point-to-point connections to the host, with contact closure or TTL level digital signaling or 4-20 mA current loop or 1-5 V voltage signaling. As the sensors proliferate in an application, the difficulty of wiring and testing complex systems encourages a single-shared-wire bus connection. One effort to establish smart sensor bus standards is pursued by the IEEE Instrumentation and Measurement Society TC-9 committee [Ajluni, 1994]. This group is drafting "Microfabricated Pressure and Acceleration Technology." A joint IEEE and National Institute of Science and Technology group is developing guidelines for standardization of protocol and interface standards for system designers and manufacturers.

Another proposal for a smart sensor bus is Fieldbus, from the Smar Research Corp. of Brazil [Ajluni, 1994]. Fieldbus works with WorldFip and ISP protocols, and is defined for all layers of the ISO seven-layer communications structure.

1.2 APPLICATIONS OF CAPACITIVE SENSORS

Some different applications listed below show the variety of uses for capacitive sensors.

1.2.1 Proximity sensing

Personnel detection	Safety shutoff when machine operator is too close.
Light switch	A capacitive proximity sensor with 1 m range can be built into a light switch for residential use.
Vehicle detection	Traffic lights use inductive loops for vehicle detection. Capacitive detectors can also do the job, with better response to slow-moving vehicles.

1.2.2 Measurement

Flow	Many types of flow meters convert flow to pressure or displacement, using an orifice for volume flow or Coriolis effect force for mass flow [Jones, 1985]. Capacitive sensors can then measure the displacement directly, or pressure can be converted to a displacement with a diaphragm.
Pressure	With gases or compressible solids a pressure change may be measured directly as a dielectric constant change or a loss tangent change.
Liquid level	Capacitive liquid level detectors sense the liquid level in a reservoir by measuring changes in capacitance between parallel conducting plates which are immersed in the liquid, or applied to the outside of a nonconducting tank.
Spacing	If a metal object is near a capacitor electrode, the coupling between the two is a very sensitive way to measure spacing.
Scanned multiplate sensor	The single-plate spacing measurement concept above can be extended to contour measurement by using many plates, each separately addressed. Both conductive and dielectric surfaces can be measured.
Thickness measurement	Two plates in contact with an insulator will measure the thickness if the dielectric constant is known, or the dielectric constant if the thickness is known.
Ice detector	Airplane wing icing can be detected using insulated metal strips in wing leading edges.
Shaft angle or linear position	Capacitive sensors can measure angle or position with a multiplate scheme giving high accuracy and digital output, or with an analog output with less absolute accuracy but simpler circuitry.
Balances	A spring scale can be designed with very low displacement using a capacitive sensor to measure plate spacing. A similar technique can measure mass, tilt, gravity, or acceleration.

1.2.3 Switches

Lamp dimmer	The standard metal-plate soft-touch lamp dimmer uses 60 Hz excitation and senses the capacitance of the human body to ground.
Keyswitch	Capacitive keyswitches use the shielding effect of the nearby finger or a moving conductive plunger to interrupt the coupling between two small plates.

Limit switch Limit switches can detect the proximity of a metal machine component as an increase in capacitance, or the proximity of a plastic component by virtue of its increased dielectric constant over air.

1.2.4 Communications

Wireless datacomm Data are capacitively coupled across a short gap in a device which replaces the optical isolator.
RF propagation in the near field can be sensed by a receiver with a capacitive plate antenna.

1.2.5 Computer graphic input

x-y tablet Capacitive graphic input tablets of different sizes can replace the computer mouse as an *x-y* coordinate input device; finger-touch-sensitive, *z*-axis-sensitive, and stylus-activated devices are available.

3

Capacitive sensor basics

Many of the different types of capacitive sensors use similar plate geometry and similar circuits and components. This chapter presents some of the different electrode configurations used to build various types of sensors.

3.1 BASIC ELECTRODE CONFIGURATION

Three different uses of capacitive sensors are to detect material properties, to sense motion or position, and to detect the proximity of conductive or dielectric objects. These applications all need specialized electrode configurations and circuit design.

The simplest electrode configuration, useful for proximity detection, is a single plate (repeated from Chapter 2) (Figure 3.1).



Figure 3.1 Single plate

This assumes vacuum (or air) dielectric. With D in cm, C , the capacitance to an equivalent ground potential very far away, is in pF. A 1 cm plate will have a capacitance of 0.354 pF. As conductive objects approach the plate, this capacitance increases.

The most useful configuration (also repeated from Chapter 2) is two parallel plates (Figure 3.2).

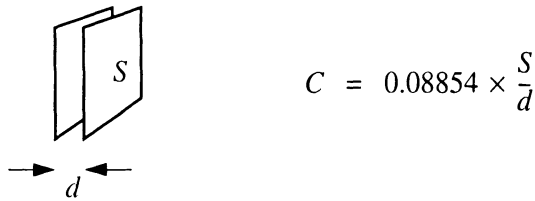


Figure 3.2 Two parallel plates

With dimensions in cm, and an air dielectric, C is in pF. Neglecting fringe effects, the capacitance with 1 cm^2 plate area and 0.1 cm spacing is 0.8 pF . This configuration is useful for measuring dielectric material properties and measuring motion, where either the spacing d is changed or the common area is changed by the transverse motion of one of the plates.

3.2 MOTION SENSING PLATE CONFIGURATIONS

Several more complex plate configurations which optimize linearity or range are used for motion sensing applications.

3.2.1 Spacing variation

For spacing variation motion sensors (Figure 3.3), the capacitance is dependent on the average plate spacing, or more accurately, the integration of $\epsilon/d \cdot dS$ over the common plate area (assuming no fringing effects), where dS is an elementary area and d is the plate spacing. About 10% change in capacitance is produced by a 10% spacing change, so with small spacing this system is very sensitive to spacing changes. The nonlinear relationship between spacing and capacitance change is a problem if capacitance is measured directly, but the output is linear if capacitive impedance is measured instead.

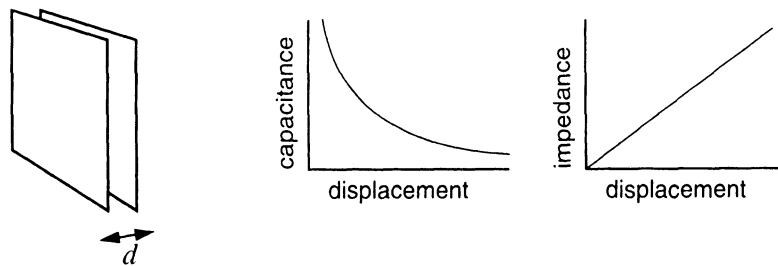


Figure 3.3 Motion sense with spacing variation

This geometry, with both plates equal in size, creates an undesired sensitivity to motion in unwanted axes which is repaired by overlapping or underlapping all edges (Figure 3.4).

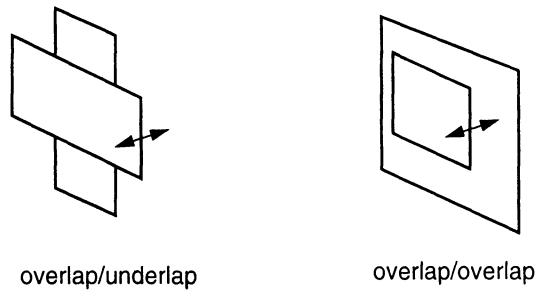


Figure 3.4 Underlap/overlap

The sensitivity of the system to mechanical disturbance in undesirable axes and to circuit drift can be further improved by using an additional electrode in a bridge circuit (Figure 3.5).

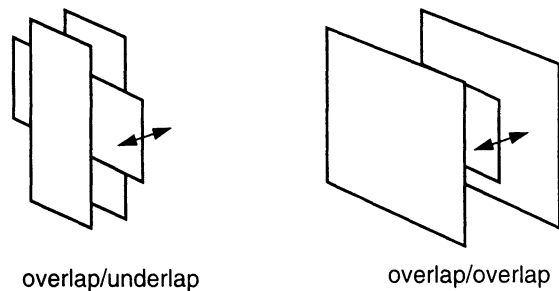


Figure 3.5 Three-plate bridge

The detecting circuit measures the ratio of the center plate's capacitance to each of the side plates. The advantage of this configuration is that the capacitance can be unaffected by translational motion in the two unwanted axes and can also be relatively insensitive to tilt in any of the three tilt axes. Another advantage of the bridge is that no absolute capacitance reference is needed, and two airgap capacitances in a bridge circuit tend to track quite accurately. The output signal is a ratio rather than an absolute value, and thus can be made to be insensitive to variations of power supply voltage, dielectric constant, and most other circuit and system variables.

A drawback with some amplifier types is the parabolic relationship between motion and capacitance and the limited range of motion which can be measured. The maximum useful linear range for spacing-variation motion detection is usually a small fraction of the plate diameter.

3.2.2 Area variation

For transverse motion which varies area with constant spacing (Figure 3.6), motion and capacitance are linearly related and a long range of motion can be measured, but the sensitivity is less, as 10% change in capacity is produced by a transverse displacement of 10% of the large plate diameter dimension rather than 10% of the small spacing dimension. Area-variation motion sensors can have an unwanted sensitivity to spacing and tilt, espe-

cially the two-plate version shown above, but these effects can usually be handled with correct electrode geometry and circuit design.

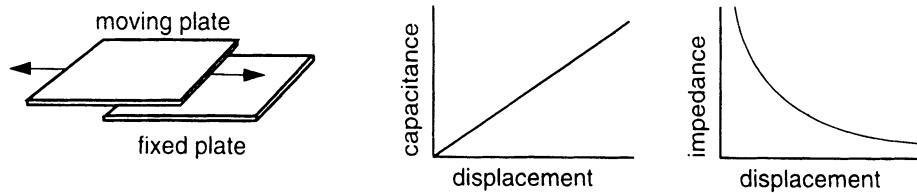


Figure 3.6 Area variation motion sense

Rotary motion

All of the electrode configurations above are shown as linear motion sensors, but they can be easily converted to rotary motion sensors using the familiar $x - y$ to $\rho - \theta$ coordinate transform. Rectangular plate shapes will be transformed to pie-shaped sectors.

Moving shield

A configuration which is less sensitive to motion in unwanted axes is the moving shield, which interposes a grounded moving plate between a pair of fixed capacitor plates (Figure 3.7). The capacitance is proportional to the unshielded area of the plates.

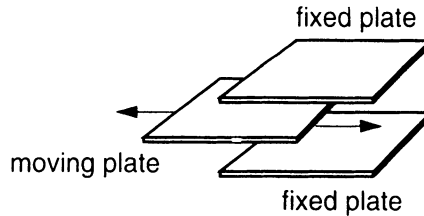


Figure 3.7 Moving shield motion sense

The plate geometries above will almost always need to be protected from external fields with an overall shield or a driven guard electrode, which also serve to control stray capacitance and improve linearity as shown in “Guards and shields” in Section 3.3.3.

3.3 MATCHING THE CIRCUIT TO THE SENSOR

3.3.1 Spacing variation with low-Z amplifier

For the simple parallel-plate spacing sensor used in sensitive capacitive micrometers (Figure 3.8), the output current I_{out} , with A the plate area in meters, x the spacing in meters, vacuum dielectric X_c and the capacitive reactance, in ohms and C , in farads, is

$$I_{out} = \frac{V_I}{X_c} = V_I \omega C = V_I \omega \cdot 8.854 \cdot 10^{-12} \cdot \frac{A}{x} \quad 3.1$$

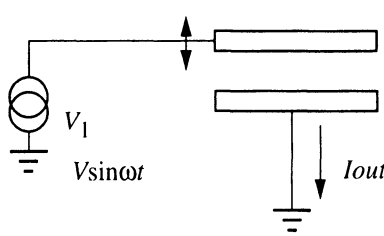


Figure 3.8 Spacing variation, low-Z amplifier

Adding a third plate produces a more stable ratiometric output signal (Figure 3.9).

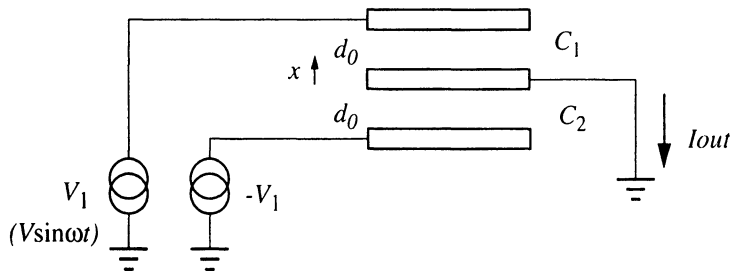


Figure 3.9 Spacing variation, three plates, low-Z amplifier

With the outer plates fixed at a spacing of $2d_0$, when the center plate is moved a distance x the output current is

$$I_{out} = V_I \omega (C_1 - C_2) = V_I \omega \cdot 8.854 \cdot 10^{-12} \cdot A \cdot \frac{2x}{d_0^2 - x^2} \quad 3.2$$

For small displacements, this becomes

$$I_{out} \approx V_I \omega \cdot 8.854 \cdot 10^{-12} \cdot A \cdot \frac{2x}{d_0^2} \quad 3.3$$

3.3.2 Spacing variation with high-Z amplifier

If the center plate feeds a high impedance (high-Z) or feedback amplifier, as in Figure 3.13 on page 44, instead of a low impedance amplifier as above, a voltage output is produced. With a voltage output instead of a current output, the response is no longer nonlinear, the electrostatic force is zero, and the dependence on dielectric constant and area disappears

$$V_{out} = 2V_I \left(\frac{X_{C2}}{X_{C1} + X_{C2}} \right) - V_I = \frac{V_I x}{d_0} \quad 3.4$$

3.3.3 Area variation with low-Z amplifier

A useful area-variation three-electrode geometry for sensing linear displacement is shown in Figure 3.10.

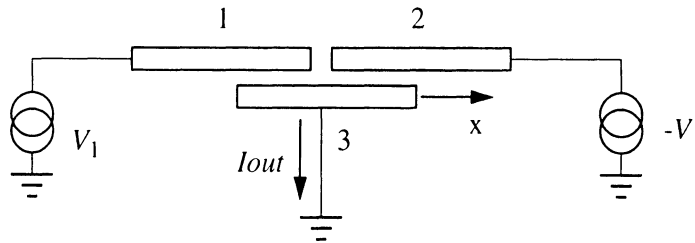


Figure 3.10 Area variation, high-Z amplifier

Here, if electrode 3 is moved to the right a distance x , neglecting fringe effects the inter-electrode capacitance C_{13} will decrease linearly with x while C_{23} increases. I_{out} is

$$I_{out} = C_{13} \frac{dV_1}{dt} - C_{23} \frac{dV_1}{dt} = (C_{13} - C_{23}) \frac{dV_1}{dt} \quad 3.5$$

and will linearly measure displacement x .

Guards and shields

Fringe fields will corrupt the measurement of position: as the edge of electrode 3 in Figure 3.10 nears the gap between electrodes 1 and 2, the fringe field will make the measurement nonlinear. Also, stray electric fields can be picked up by the signal electrode 3. One fix is to add a guard electrode which surrounds the pickup and moves with it (Figure 3.11).

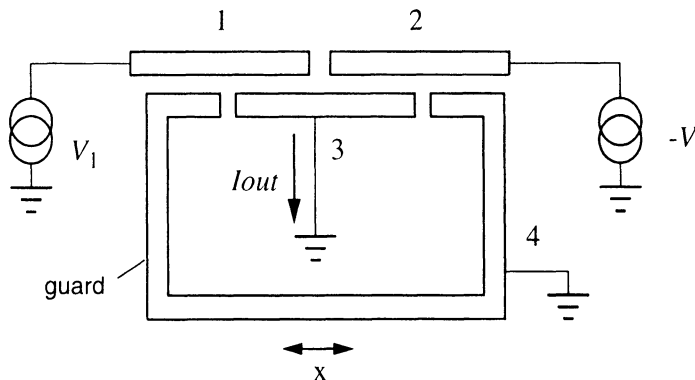
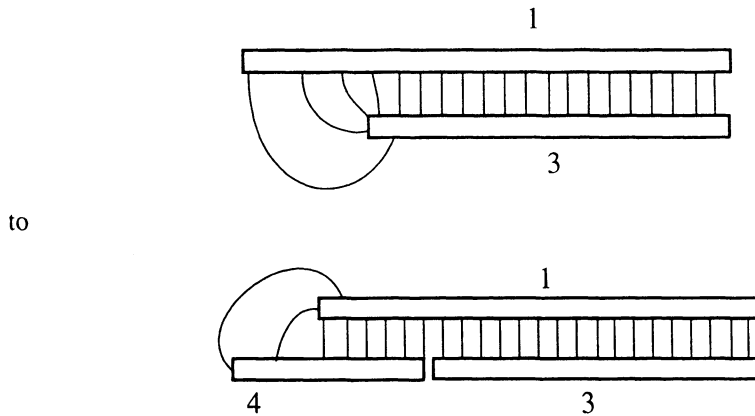


Figure 3.11 Guarded three-electrode displacement sensor

Now, electrode 4 changes the shape of the field lines at the edge of electrode 3 from



The field lines are distorted at the edge of the guard (electrode 4) where this distortion is unimportant, but uniform at the measuring electrode.

The guard electrode also serves as a shield to screen electrode 3 from extraneous fields. Note that the guard electrode does not need to be at the same potential as the signal electrode to be effective [Heerens, pp. 897–898]. It can be at any DC or uncorrelated AC voltage and the same results will be achieved. This is difficult to prove by constructing field lines and applying Poisson’s equation, as the field lines will be distorted by the new voltage, but easy to prove using superposition and an equivalent circuit, as in “Multielectrode capacitors” in Section 2.3.1.

3.3.4 Area variation with high-Z amplifier

The previous linear sensor system (Figure 3.11) can be modified to show another way to use the guard electrode (Figure 3.12).

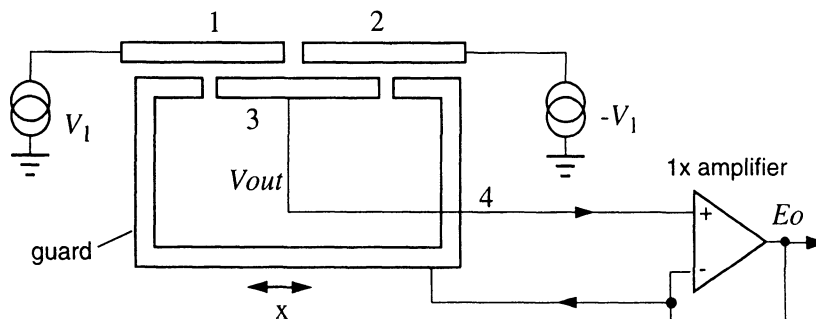


Figure 3.12 Guarded displacement sensor, high impedance amplifier

The output voltage E_o is

$$E_o = V_I \left[\frac{C_{13} - C_{23}}{C_{13} + C_{23}} \right] \quad 3.6$$

Note that this and many other circuits are shown without a way to control the DC voltage at the operational amplifier input; see Section 4.3.4 for more guidance. Without the guard, the use of a high impedance amplifier would mean that stray capacitance from electrode 3 to local grounds would have attenuated the signal, but with the bootstrapped guard these strays are now unimportant. The previous circuit (Figure 3.11) measured $C_{13} - C_{23}$ by amplifying the output current or output charge, using a low input impedance amplifier. This circuit measures $C_{13} - C_{23}$ by amplifying voltage with a high input impedance amplifier. Each circuit has equivalent performance with respect to rejecting the effects of stray capacitance and rejecting external electric fields, but the circuit of Figure 3.12 has a very important advantage of spacing insensitivity. With the current-output circuit of Figure 3.11, the output with $x = 0$ is zero and is insensitive to spacing, but the maximum output level (and hence the gage factor) is a direct function of spacing. For example, at the maximum-output position with the sense electrode 3 directly opposite electrode 2, the output current is proportional to C_{23} , or inversely proportional to the plate spacing; this is a benefit if the plate spacing is the parameter to be measured, but not for the current task of measuring linear displacement orthogonal to the spacing axis. With the high impedance amplifier circuit (Figure 3.12), the output level is ratiometric rather than absolute response, it is sensitive to V_1 instead of dV_1/dt , and it is totally insensitive to spacing, except for second order effects due to unguarded amplifier input capacitance and fringe fields. The output level is, unfortunately, sensitive to rotation in the axis through the paper, but we will find ways around this in later chapters.

3.3.5 Area variation with feedback amplifier

A third arrangement is similar to the high-Z amplifier above, except the feedback is taken around the driven electrodes (Figure 3.13).

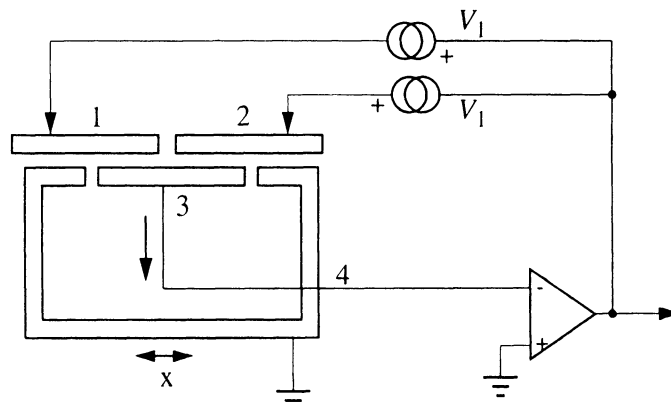



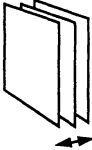
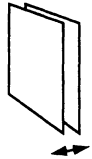
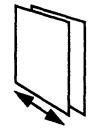
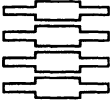
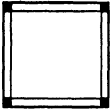


Figure 3.13 Guarded displacement sensor, feedback circuit

The performance of this circuit is very similar to the high impedance amplifier, and it can be made more resistant to stray capacitance. The circuit is similarly insensitive to spacing, as the spacing-variable capacitances C_{13} and C_{23} are inside the feedback loop and in series with the high input impedance of the amplifier. It is discussed further in the next chapter.

3.3.6 Electrode configuration table

Table 3.1 shows the various types of plate and circuit configurations for different applications.

Table 3.1 Capacitive sensor configurations

Application	Plate geometry	Amplifier (Chapter 4)	Circuit	Excitation	See Chap.
Material properties		Low-Z	Single ended or bridge	Variable frequency	10
Micrometer high sensitivity		Low-Z	Bridge	Sine wave	5
Micrometer small movement		Low-Z	Single ended	Square wave 0°/ 180°	5
Motion detect large movement		High-Z or feedback	Single ended	Square wave 0° / 180°	6, 7
Motion detect analog/digital		High-Z or feedback	Sin/cos or tracking	Square wave 0°/ 180°	8
Motion detect 2-axis		High-Z or feedback	Single ended	Square wave 0°/ 90°	19
Proximity detector		Low-Z	Single ended or bridge	Square wave or pseudo-random	6
Silicon sensors		High-Z or feedback	Bridge	Square wave	11, 15

3.4 LIMITS TO PRECISION

3.4.1 Noise

One limit to measurement precision in a capacitive sensor may be the number of charged particles associated with the capacitance. If a test voltage of 1 V is applied to a 1 cm disk, the capacitance of 0.354 pF accumulates a charge $Q = CV$ of 0.354×10^{-12} C. With 6.242×10^{18} electrons/C, we have 2.27×10^6 electrons (or holes) on the disk. In systems where the charges are quantized so that fractional charges are not allowed, as for an isolated electrode, this limits precision to several parts per million. Certainly, the quantization of electron charge will not be a problem for most systems, and fractional charges are usually allowed except for currents through semiconductor junctions or electrodes which are completely floating.

Another limit is the input current noise of the amplifier circuit. With the high impedances of capacitive sensors, current noise is usually more important than voltage noise. With 10 V p-p AC excitation at 100 kHz, the reactance of 0.354 pF is 45 k Ω and the signal current is 10 V/45 k Ω or 2 μ A. Bipolar amplifiers with input current noise of 1 pA in a 1 Hz bandwidth would generate 100 pA in a 10 kHz bandwidth and degrade signal to noise ratio to 4000:1. But FET-input operational amplifiers are available with current noise of less than a femtoamp (10^{-15} A) in 1 Hz, so the signal-to-noise ratio measured in a narrow 1 Hz bandwidth can be greater than $20 \times 10^{-6}/10^{-15}$ or 20×10^9 . Apparently, current noise is not much of a limiting factor either, except for very small capacitive reactance or large bandwidth.

Actual resolution may be several orders of magnitude worse than this, depending on circuit design; shot noise effects, excess noise in semiconductor junctions, and thermal noise in resistors can be larger than amplifier input noise. But careful design can produce current sensors with noise of just a few thousand noise electrons per second. Systems with a sensitivity of 0.01 aF (0.01×10^{-18} F) have been reported [Jones and Richards, 1973]. Chapter 12 has a more detailed discussion of noise problems with capacitive sensor circuits.

3.4.2 Stability

Environmental

The environmental stability of air-dielectric capacitors is quite good. The dielectric constant of dry air at 0°C and 760 mm pressure is 1.000590, and that of gaseous water (steam) at 110°C is 1.00785, so the effects of temperature and humidity are minimal; Section 12.5 considers these effects in more detail. But absolute capacitance measurement is difficult to do accurately, as stray capacitances must be considered; careful grounding and shielding are needed. If stray capacitances are well guarded, the variation of capacitance due to dielectric constant variation and fringe fields can usually be ignored.

Mechanical

The mechanical stability of the supporting structure can be a concern. For spacing-variation micrometers discussed in Chapter 5, structural stability to μ m dimensions is critical, but for area-variation systems or systems with carefully designed reference capac-

itors, ratiometric bridge amplifiers, and spacing-insensitive circuit design, structural stability is not so much of a problem.

Electrical

With ratiometric circuits, the output is sensitive only to a ratio of capacitances. Very precise components are not needed, and correct circuit design mitigates or cancels the effects of power supply variation and component variation.

6

Proximity detectors

Proximity detectors sense the presence of nearby targets, usually without requiring any contact or wiring to the target or any particular target material properties. Various sensors are available for proximity detection and measurement, including capacitive, inductive, optical, ultrasonic, and magnetic sensors. These detectors are used for many industrial applications and they are available with a variety of analog and digital outputs.

6.1 APPLICATIONS

Typical applications of proximity sensors include [Soloman et al., 1994]:

Motion detection	Detection of rotating motion Zero-speed indication Speed regulation
Motion control	Shaft travel limiting Movement indication Valve open/closed Conveyer system control Transfer lines Assembly line control Packaging machine control
Process control	Automatic filling Product selection Machine control Fault condition indication Broken tool indication

	Product present
	Bottle fill level
	Product count
Sequence control	Verification and counting
	Product selection
	Return loop control
	Product count
Liquid level detection	Tube high-low liquid level
	Overflow limit
	Dry tank
Material level control	Low level limit
	Overflow limit
	Material present

6.2 TECHNOLOGY

Proximity sensors are used as inputs for industrial control systems. Sensors for this application are available with many different outputs to connect to different control systems, including two- and three-wire AC, two-, three-, and four-wire DC, normally open and normally closed switches, TTL logic level, AC/DC, and several four-wire analog output versions including linear 4–20 mA current and 0–5 V voltage output versions. A two-wire analog variable-resistance interface (NAMUR) is often used; these devices are small and easy to wire in a system. NAMUR sensors are calibrated to sink 1.55 mA with a nominal 8.2 V DC supply at the nominal sensing range.

Proximity switches can replace mechanical limit switches for difficult environments. Different versions are used to detect ferrous and nonferrous materials, including glass, cardboard, and plastics. LED lamps are often provided for a visual indication of output state.

6.2.1 Inductive sensors

Inductive sensors generate a local magnetic field at a 2 kHz–500 kHz frequency and detect ferrous materials as an increase in flux for constant excitation, or conductive nonferrous metals are detected as a decrease in flux due to their shielding effect or due to induced lossy eddy currents. Detection range is a function of the size of the magnetic system and is in the range of 1 mm–100 mm; for most industrial sensors the maximum detection range is about the same as the diameter of the magnetic coil. Inductive sensors are rugged, cheap, and reliable, but cannot detect dielectric materials. The range and field topology is similar to capacitive sensors with the limited range similarly due to the inability to focus the magnetic field in air. More power is needed for inductive sensors than capacitive sensors.

6.2.2 Magnetic sensors

Magnetic sensors are similar to inductive sensors, except a DC magnetic field generated by a permanent magnet is used instead of the AC field.

As pickup coils are sensitive only to AC magnetic flux, a Hall effect device which is sensitive to DC magnetic fields is often used. The Hall effect, discovered by Mr. E. F. Hall in 1879, is created in a conductive sheet. With a linear current flowing in one axis, a linear field-dependent voltage is measured in the other axis when a magnetic field is induced through the sheet. Silicon implementations are usually packaged in a plastic transistor case with built-in linear amplifier; output sensitivity is 0.25–2 mV/G with a 25 kHz bandwidth. Some versions are available from vendors such as Allegro Microsystems in Worcester, MA, with Schmitt trigger outputs and high current drivers. A drawback of commercially available Hall sensors is the large and poorly controlled offset voltage which is compensated by the use of AC coupling or computer calibration strategies.

Another magnetic sensor with response to DC uses a saturable high permeability material. Permalloy, for example, has a very high permeability of 100,000 or more, a rectangular hysteresis loop, and low saturation flux density, and can be biased to switch states in response to low level DC magnetic fields.

A switching response is also available with a magnet-actuated reed relay, but sensitivity is not as good.

6.2.3 Optical sensors

Many different types of optical sensors are available, from simple slotted optical gate types to reflective sensors. For very accurate long-range object distance measurement, a coherent laser beam can be bounced off a reflective target, and optical interference effects used to measure position differences by counting interference cycles, with one cycle representing 680 nm for red laser light. At the extreme in complexity, a tunable laser can be used to provide an absolute distance capability as well as the differential distance capability.

Optical sensors can be focused, and they can be very accurate, but often need lenses and special target preparation.

6.2.4 Ultrasonic sensors

Ultrasonic sensors radiate a short ultrasonic pulse in the 20 kHz–500 kHz range. The pulse bounces off a local object and the echo is detected, often by the transducer which launched the pulse. Operation depends on the transmission of air and the sonic reflectivity of the target, which is a function of the orientation and material of its surfaces. Ultrasonics is quite useful in sea water, which attenuates E and H fields but transmits sound well. Soft materials such as cloth and foam do not reflect well.

Sound velocity in air is about 344 m/s, but it is affected by several factors [Eshbach, p. 1092]:

Temperature	+ 0.18%/ °C
Relative humidity	+ 0.4% from 0 to 20% RH
Pressure	little effect below 50 atm
Frequency	- 0.243% for 41 kHz to 1.5 MHz

Some ultrasonic sensors can be adjusted to detect objects only between a preset maximum and minimum range [Turck, p. E4]. As the sound is easily focused, maximum range is higher than inductive and capacitive sensors; maximum range for industrial control units is 6–8 m, while security applications make use of longer range units, with up to 15 m range.

6.3 CAPACITIVE PROXIMITY DETECTORS

Proximity sensing is a simple and effective application of capacitive sensors. Capacitive proximity sensors are used for many applications in plant control, and generally are supplied as a small (1×5 cm) cylinder with a pair of electrodes on one end and wire leads on the other. The usual output is a contact closure or TTL-level pulse when an object comes within detection range, about 1 cm. A small amount of hysteresis is added to guarantee dither-free output.

Capacitive proximity sensors are noncontact, can detect small objects, and work with either conducting or insulating objects, such as an unprepared surface of a mechanism or a moving conveyerized object. Similar sized conductive objects are all detected at the same range, but detection of insulating objects depends on a lossy dielectric or a dielectric constant sufficiently different from unity; different insulators are detected at different ranges depending on these parameters and on excitation frequency.

Commercial capacitive proximity sensors specify detection ranges of up to 40 mm for a 34 mm diameter cylinder. One manufacturer [Turck] lists cylinder-style capacitive proximity sensors with diameters between 25 and 80 mm with detection distance approximately proportional to diameter. Several other mechanical configurations are available such as rectangular form factors intended for bolt-in replacements for mechanical limit switches. The output is either a switched AC or DC voltage, analog signals, or a contact closure, as for inductive sensors. Two concentric electrodes are used on the end face, as shown in Figure 6.1.

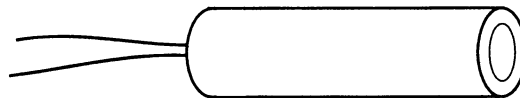


Figure 6.1 Cylindrical capacitive proximity detector

The active face of capacitive proximity sensors is composed of two concentric metal rings. In one manufacturer's implementation the mutual capacitance between the rings appears as the capacitive feedback element of a high frequency Colpitts oscillator. With no object present, the oscillator is tuned to a point just below oscillation; as a target approaches, the coupling between electrodes increases and the circuit oscillates. The oscillator output feeds a switch or an analog amplitude measurement circuit which generates the digital or analog output voltage.

Dielectric materials are effective at actuating the sensor, with effectiveness increasing with increasing dielectric constant. Conductive materials have a different detection

range if they are connected to earth ground; the capacitance of the target to ground produces a decrease in coupling but an increased capacitance to ground. A 40 mm diameter model, the CP40 from Turck, lists these specifications (see Table 6.1).

Table 6.1 CP40 capacitive sensor

Nominal sensing range	20 mm
Max. switching frequency	100 Hz
Input voltage	10–65 VDC
Output contacts	1 NO, 1 NC
Sensing range	
Metal	40 mm
Water	40 mm
PVC	20 mm
Wood (varies with RH)	20–32 mm
Cardboard	8 mm
Glass	12 mm

Turck also lists the dielectric constants shown in Table 6.2.

Table 6.2 Dielectric constants

Material	Dielectric constant	Material	Dielectric constant
Acetone	19.5	Ethylene glycol	38.7
Acrylic resin	2.7–4.5	Fired ash	1.5–1.7
Air	1.000264	Flour	1.5–1.7
Alcohol	25.8	Freon R22, 502	6.11
Ammonia	15–25	Gasoline	2.2
Aniline	6.9	Glass	3.7–10
Aqueous solution	50–80	Glycerine	47
Bakelite	3.6	Hard paper	4.5
Benzene	2.3	Marble	8.0–8.5
Cable sealing compound	2.5	Melamine resin	4.7–10.2
Carbon dioxide	1.000985	Mica	5.7–6.7
Carbon tetrachloride	2.2	Nitrobenzene	36
Celluloid	3.0	Nylon	4–5
Cement powder	4.0	Oil saturated paper	4.0
Cereal	3–5	Paraffin	1.9–2.5
Chlorine liquid	2.0	Paper	1.6–2.6
Ebonite	2.7–2.9	Perspex	3.2–3.5
Epoxy resin	2.5–6.0	Petroleum	2.0–2.2
Ethanol	24	Phenol resin	4–12

Table 6.2 Dielectric constants *Continued*

Material	Dielectric constant	Material	Dielectric constant
Polyacetal	3.6–3.7	Shell lime	1.2
Polyamide	5.0	Silicon varnish	2.8–3.3
Polyester resin	2.8–8.1	Soybean oil	2.9–3.5
Polyethylene	2.3	Styrene resin	2.3–3.4
Polypropylene	2.0–2.3	Sugar	3.0
Polystyrene	3.0	Sulfur	3.4
Polyvinyl chloride resin	2.8–3.1	Teflon	2.0
Porcelain	3.5–4	Toluene	2.3
Powdered milk	3.5–4	Transformer oil	2.2
Pressboard	3.7	Turpentine oil	2.2
Quartz glass	3.7	Urea resin	5–8
Rubber	2.5–3.5	Vaseline	2.2–2.9
Salt	6.0	Water	80
Sand	3–5	Wood, dry	2–7
Shellac	2.5–4.7	Wood, wet	10–30

6.3.1 Limits of proximity detection

Single or dual electrodes

Either a single or a dual electrode structure can be used for proximity detectors. A single electrode produces dielectric flux lines which terminate on the surrounding building structure or on the surface of the earth, and single electrodes can have higher values of detection distance relative to electrode size than dipole electrodes. A hazard is that the earth or building ground potential is available to AC-powered instruments only as the third wire in the power cable, and that connection is usually contaminated by noise, but the use of a high carrier frequency usually successfully avoids this noise component.

The dipole electrode pair is simpler and works as well for low ratios of electrode size to detection distance. Dipole geometry is used in commercially available capacitive proximity sensors.

Earth ground

Capacitive proximity detectors may need to function in environments which are corrupted with extraneous electric fields. One cause of extraneous fields in industrial environments is due to high impedance ground connections.

The “earth ground” in textbooks is a perfectly conducting infinite plane. In real life, earth is not particularly conductive, with a resistivity of 500 Ω -cm or more [Morrison, p. 138]. Sand and gravel can have bulk resistivity of 10 k Ω -cm or more, and a 1 in diameter rod driven to a 10 ft depth can have a resistance of 30 – 50 Ω . Treatment with magnesium sulphate will reduce resistance [Fink et al., pp. 19–52].

Current flow in the earth is due to lightning strikes, power transmission lines, and also earth return currents in power distribution systems. Lightning strikes generate by far

the largest earth currents; maximum current can exceed 20,000 A. Power transmission line earth currents are restricted to the area immediately beneath the lines and will not trouble most installations.

Earth current returns from power distribution will be a factor in any industrial situation. The power distribution inside a building is set up so that the “hot” AC connection is made through a black wire, in the U.S., and the return current is nominally through the white wire. Conductive equipment enclosures are earthed through the service neutral, the third (green) wire for three-wire connections. The service neutral is unfused, and it has two functions: to provide a return path for small mA-level leakage currents to the equipment enclosures to return to earth ground for safety, and to provide a return for the much larger fault current which flows if a hot wire is shorted to an enclosure. The fault current then results in a blown fuse instead of a dangerous potential on the enclosure.

Leakage currents to the service neutral result from motor or power transformer insulation leakage and from charging current due to small primary-side capacitors used for EMI control. The service neutral potential in an industrial building relative to earth ground is determined by the sum of these mA-level leakage currents flowing through the service neutral ohmic resistance multiplied by the resistance of the local connection to earth. This potential is usually in the 1–15 V range.

6.3.2 Maximum detection range

It is interesting to calculate the approximate detection range of a carefully designed capacitive proximity detector.

From “Monopole proximity detector, parametric plot, log” (Figure 2.40), the falloff in capacitance with 16 mm diameter electrodes is at a rate of 1 decade / 18 mm of spacing. The capacitance at 1 mm target distance is

$$C_1 = \pi r^2 \cdot \frac{\epsilon_0}{d} = 1.78 \quad \text{pF} \quad 6.1$$

For this geometry

$$C = 1.78 \cdot 10^{-d/0.018} \quad \text{pF, m} \quad 6.2$$

The theoretical limit of detectable capacitance variation from “Limiting displacement of three-plate micrometer,” Section 12.2.1, is 3×10^{-22} F for a detection limit of 0.176 m, using a narrow 1 Hz bandwidth. Most proximity detectors have a motion detection bandwidth of 100 Hz; for a 100 Hz bandwidth the theoretic limit degrades to 0.158 m.

The limit of experimentally verified capacitive detection with high voltage (100 V rms) excitation [Jones and Richards, 1973] is 0.05 aF, 0.05×10^{-18} F. Substituting this value for C and solving for d , we have

$$d = 0.018 \log \frac{1.75 \cdot 10^{-12}}{0.05 \cdot 10^{-18}} = 0.136 \quad \text{m} \quad 6.3$$

That limit is, not unexpectedly, difficult to achieve in practice because of the several effects discussed as follows.

Dielectric constant of air

One limit to the maximum detection distance is due to the change of the dielectric constant ϵ_r of air. At extreme distances, target movement is indistinguishable from ϵ_r variations. This section calculates the effect of environmental variations on maximum detection distance.

The dielectric constant of air changes slightly with pressure, temperature, and humidity. At standard temperature and pressure, the dielectric constant changes with temperature as $2 \times 10^{-6}/^\circ\text{C}$ for dry air, increasing to $7 \times 10^{-6}/^\circ\text{C}$ for moist air. At 20°C , the dielectric constant change with relative humidity is 7×10^{-5} for an RH change from 40 to 90%. A change of pressure of 1 atm changes the dielectric constant by 10^{-4} . Over a distance of a few tens of meters and a time span of a few hours, the expected variation of these parameters might be typically as shown in Table 6.3.

Table 6.3 Change in dielectric constant of air

	Change	Coefficient	Effect
Temperature	5 $^\circ\text{C}$	5 ppm / $^\circ\text{C}$	25 ppm
Relative humidity	10%	1.4 ppm / %RH	14 ppm
Pressure	0.05 atm	100 ppm / atm	5 ppm
Total			44 ppm

The total of the atmospheric variations above will cause the 1.78 pF capacitor of eq. 2.13 to change by 44 ppm, or by 7.8×10^{-5} pF. Using this new value for C and recalculating the limit of detection of the 16 mm circular plate proximity detector (eq. 6.3) we arrive at

$$L = 0.08 \quad \text{m}$$

so the maximum detection distance has been cut down by a factor of 20 by these environmental effects. A bridge circuit which uses an airgap capacitor as a reference will compensate for these atmospheric variations, but becomes sensitive to temperature and RH gradients, and constructing an air-spaced capacitor with less than 44 ppm drift is not a simple project.

Local motion

Another limit to the sensitivity of proximity detectors is the effect of small movements of local objects which change the mutual capacitance between the measuring electrodes. This motion may be temperature-induced. The temperature coefficient of aluminum, brass, and steel is in the range of 10–20 ppm/ $^\circ\text{C}$. Depending on the orientation of nearby conductors relative to the measurement electrodes, the effect of a nearby conductor moving only a few μm could considerably decrease detection distance.

Focusing field lines

Sensitivity is also limited by the inability to focus capacitance. The field lines can be controlled locally by use of guarding and shielding electrodes, but once in free space, they spread uncontrollably in response to Poisson's law. Wave-propagated signals like light, sound, or electromagnetic radiation can be accurately focused at considerable distance, but that advantage is not available in capacitive sensors; hence a nearby moving object such as a machine operator may drastically limit the usable maximum distance of a proximity detector. Some improvement in performance is available by positioning grounds appropriately; if the sensor is detecting moving dielectric objects on a conveyer belt, say, the best electrode setup is with the two detection plates on either side of the belt. If that is not possible, with both electrodes on one side, a ground on the other improves field focus by a factor of two or three.

Experimental circuit

An experimental capacitive proximity sensor was built to verify the theoretical limiting range calculated above; it is detailed in Chapter 17.

6.3.3 Proximity sensing equivalent circuit

The proximity sensor electrodes are shown in Figure 6.2.

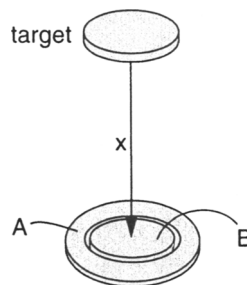


Figure 6.2 Proximity sensor electrodes

A simplified block diagram (Figure 6.3) of the standard two-electrode proximity detector shows the effect of the various capacitances.

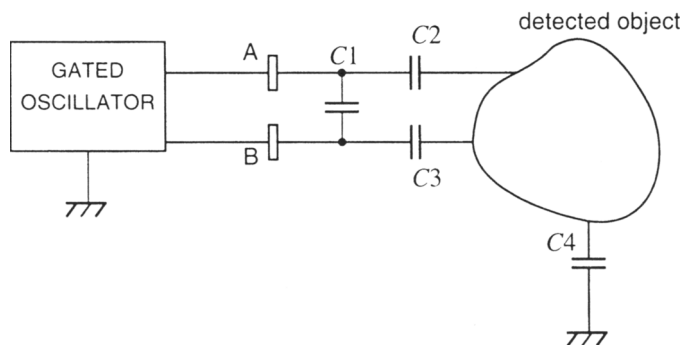


Figure 6.3 Proximity sensor block diagram

The gated oscillator is commonly used for proximity detectors, but it does not have as good a performance as a synchronous detector in noisy environments. The equivalent circuit, for a synchronous detector implementation with electrode A excited, is illustrated in Figure 6.4.

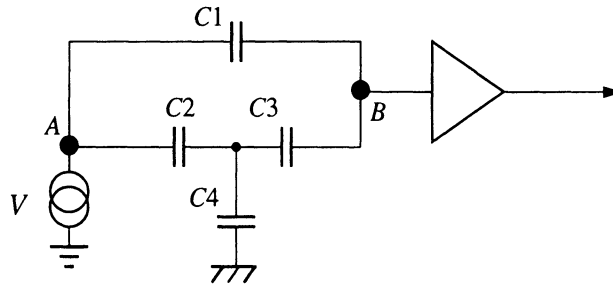


Figure 6.4 Proximity detector equivalent circuit

The synchronous detector and the keyed oscillator circuit can both be analyzed using this equivalent circuit.

The capacitances to be detected, C_2 and C_3 , are contaminated by stray capacitance C_1 and C_4 . As target distance increases, C_2 and C_3 become much smaller than C_1 , and the additional shunting effect of C_4 further decreases the signal so that the change in signal with target movement may become a tiny fraction of a percent. Three possible capacitive proximity detector targets may be considered.

Metallic, floating

The equivalent circuit of the floating metallic target is shown in eq. 2.13. C_2 and C_3 decrease by about a decade for every sense-plate-diameter increment of target motion.

Metallic, grounded

If the target is a conductive grounded object C_4 is a short circuit and C_1 will decrease as the target approaches due to the shielding effect; the change in C_1 is the measured variable, or a circuit which measures C_2 , the capacitance to ground, can be used.

Dielectric

With a dielectric target C_4 can be considered open and the values of C_2 and C_3 are modulated by the target's proximity. C_1 increases as a dielectric target approaches due to the increase in permittivity.

6.3.4 Stray fields

Stray electrostatic or electromagnetic fields can interfere with capacitive proximity detection in two ways: by saturating the input amplifier or by being mistaken for the exciting signal.

Input amplifier saturation

An ambient electric field can result from capacitive coupling to a 60 Hz power signal or from the electric field associated with electromagnetic radiation from an RF transmitter. The ambient field may saturate the input amplifier by driving the output to the power rails and the detector will be blocked. Regulatory bodies specify that electronic equipment should not exhibit degraded performance in the presence of a 3 or 10 V/m maximum electric field.

Amplifier saturation is normally not a serious problem, except in very strenuous industrial environments, but it may be avoided by preceding the amplifier with a bandpass filter circuit tuned to the excitation frequency (Figure 6.5).

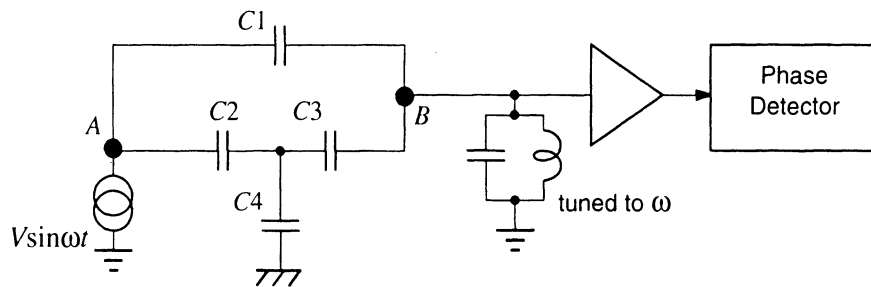


Figure 6.5 Tuned-input proximity detector circuit

Capacitively coupled in-band noise

If, for example, a 20 kHz excitation frequency is used and a local 60 Hz power signal with an irregular waveform is capacitively coupled to the sense electrodes, the 333rd harmonic of 60 Hz is within 20 Hz of the excitation frequency. If the proximity detector's circuits cannot reject this signal, the maximum detection range will be considerably reduced. A resonant circuit or a bandpass filter can help, but a high Q is difficult to achieve and may require tuning. A synchronous detector (below) replaces the bandpass filter with a lowpass filter. This is a very good trade as a 10 Hz lowpass filter is easy to build while a 10 Hz bandpass filter around a 20 kHz carrier is difficult.

Shielding

A metallic shield almost completely attenuates stray E -fields, but proximity detectors cannot usually be completely shielded. Partial shields may help reject local fields.

6.3.5 Phase detectors

Several effective phase detector circuit configurations can be used to virtually eliminate the effects of coupled in-band noise.

Synchronous phase detector

As the detection circuit is in the same enclosure as the transmitter, the detector knows the exact frequency and approximate phase of the signal to be detected. The phase

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is known exactly if the equivalent circuit is purely capacitive, but in practice resistive elements such as amplifier input impedance will contribute a small phase shift, usually in the range of 10–20°. This phase shift can often be ignored or compensated, and a synchronous detector used (Figure 6.6).

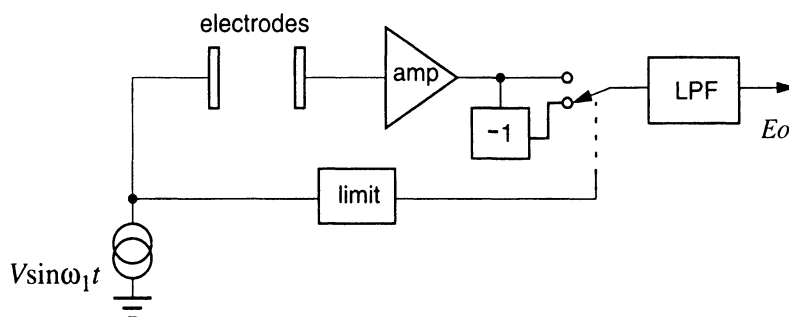


Figure 6.6 Synchronous phase detector

Here, the excitation voltage for the electrodes is also applied to a switch which chooses between the amplified signal electrode voltage and its negative. If phase shifts are small, the result at the LPF output is a DC representation of the magnitude of the electrode voltage at the switching frequency, with switch transients removed by the lowpass filter. The sinusoidal excitation signal, hard limited to a square wave, activates the switch, or a square wave signal feeds both switch and electrodes.

If the phase shifts in the two paths from excitation to multiplier are unequal, the performance degrades until, when the difference in phase shift becomes 90°, the circuit is completely inoperable. Phase shifts can be matched by adding a passive delay line to the control input, usually the shorter path, or by adding a second matched electrode-amplifier with reference capacitors for very accurate tracking.

Pseudorandom carrier

A fix for proximity circuits which need maximum performance in the presence of interfering spectral noise components is to use a pseudorandom excitation voltage, a two-level signal which changes phase in a predictably random way, which is applied to the sense plates and also to the demodulator. The maximal length shift register connection [Rhee, 1989, pp. 265–268] generates a repeating sequence of $2^n - 1$ binary outputs. For example, a 10 bit maximal length shift register outputs $2^{10} - 1$ or 1023 random-appearing binary symbols before repeating. The number of 1s and 0s is almost equal (it differs by 1) and the longest length of a repeated 0 or 1 symbol is $n - 1$. This pseudorandom number (PRN) code can be used directly as a carrier with, say, a 50 kHz clock and a 50 kHz/1023 repetition frequency, and gives good rejection of in-band spectral noise components. With a band-limited amplifier, performance would suffer because of the spectrum-broadening effect of a long run of adjacent 1s or 0s, but a simple digital phase modulator can be added to the PRN output to guarantee a maximum run length of two symbols. The added circuitry for a pseudorandom carrier is minimal; a 10 bit PRN is built with a 10-stage shift register and a quad exclusive-OR gate.

6.4 CAPACITIVE LIMIT SWITCHES

An important type of proximity detector is the limit switch which senses a moving vane and activates a switch closure. For the general case of proximity detector the target cannot be altered, but a limit switch can be designed to use a grounded, dielectric, or floating target, as the target is built as part of the switch. The limit switch should be accurate, inexpensive, and reliable. An example is the optical gate, where a light-emitting diode is paired with a phototransistor to detect any opaque object moved through its 2.5 mm gap.

A capacitive limit switch has several advantages over the optical gate, including much lower power dissipation, lower cost, and more stable operation (Figure 6.7).

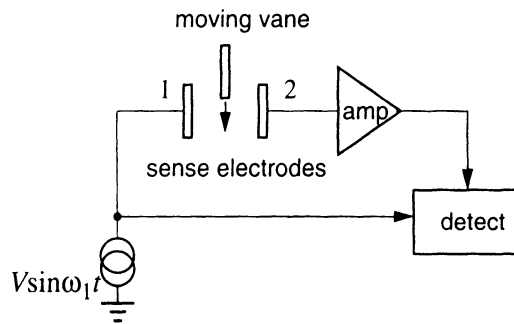


Figure 6.7 Limit switch block diagram

6.4.1 Vanes

The capacitive limit switch detects the difference in capacitive coupling between electrodes 1 and 2 caused by the movement of the vane. The vane may be grounded, and decrease capacitance; floating or dielectric, and increase capacitance; or it may be separately excited.

Grounded vane

To detect a grounded vane, the optimum electrode geometry maximizes coupling when the vane is not present (Figure 6.8).

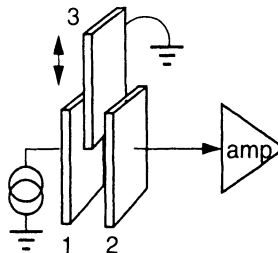


Figure 6.8 Grounded vane electrode geometry

With the grounded vane limit switch, the modulation depth can be very good, especially if the vane is larger than the electrodes; the minimum capacity can be 2–3% of the maximum capacity. The lateral position of the vane does not affect the measurement.

Floating vane

With a floating vane, detection is best done with electrodes which have minimum capacity before the vane is introduced (Figure 6.9).

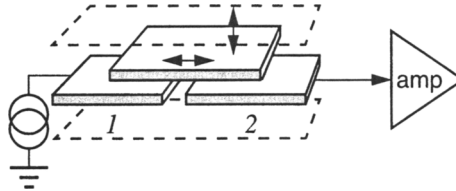


Figure 6.9 Floating vane electrode geometry

Floating vane modulation depth is not as good as grounded vane and, if detecting lateral movement, the vertical position of the vane considerably changes the capacity. Grounded shields installed at the position of the dotted lines will reduce stray capacity and increase modulation depth. These shields will minimize stray capacitance between electrodes 1 and 2, but not affect the coupled capacitance when the moving vane is in position. Motion in the vertical axis can also be detected.

Dielectric vane

The dielectric vane uses an electrode geometry similar to the grounded vane configuration, and the presence of the vane increases electrode capacitance proportional to the vane's dielectric constant and the percent of gap occupied.

Separately excited vane

If the vane can be connected to a high frequency AC source, the performance is improved considerably. In Figure 6.8, for example, this would be done by removing electrode 1 and attaching the exciting signal to electrode 3. Normally this is considered the last choice because of the inconvenience of the extra wire.

Summary of vane performance

A very approximate comparison of the performance of these different limit switch geometries is given in Table 6.4.

Table 6.4 Vane performance

Type	Modulation depth	Sensitivity to gap variation
Grounded vane	97%	low
Floating vane	60%	high
Excited vane	99%	low

6.4.2 Limit switch circuits

The challenge for limit switch circuit design is to reliably detect small changes of capacitance and be insensitive to board contamination and stray resistance or capacitance. For 1 cm square plates and a 2.5 mm gap the capacity is, from eq. 2.13, 0.35 pF, so the detector needs to resolve a change from an open gap at 0.35 pF to a grounded-vane capacitance of 0.05 pF or so. If the electrodes are connected with a 1 cm 0.015 in trace width on 1/16 in glass-epoxy with a ground plane on the opposite side, the PC trace will add a stray capacitance of 0.5 pF. The capacity of an IC pin adds about 4–10 pF to ground. A simple method is needed to detect a fraction of a pF in the presence of much larger strays; the guarding techniques used for the synchronous demodulator circuits would work well, but these circuits are more complex. One method is to use an AC reference.

AC reference

An AC reference, shown in Figure 6.10, is used rather than a DC threshold so that the detection threshold is at ground.

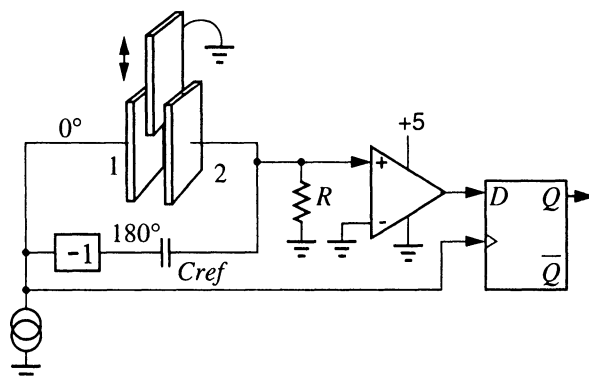


Figure 6.10 Limit switch, AC reference

The waveforms at the excitation terminals and the comparator input are as shown in Figure 6.11.

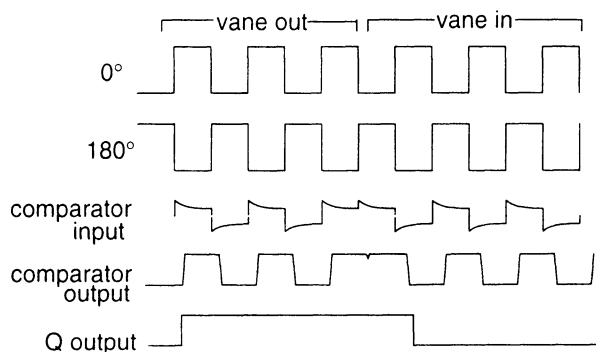


Figure 6.11 Limit switch, AC reference, waveforms

A 5 V 10–200 kHz logic signal can be used for the 0° clock, and a logic inverter generates the 180° clock. If the equivalent capacity between electrodes 1 and 2 varies from 0.05 to 0.35 pF as the vane is moved, C_{ref} , the reference capacitor, is chosen to be halfway between the maximum and minimum value of the sense capacitor, or 0.2 pF. R is a large-value resistor, over $22\text{ M}\Omega$, and the comparator is a totem-pole-output CMOS type such as Texas Instruments' dual, the TLC3702. The TLC3702 works on a 5 V supply, the input common mode voltage range includes ground, and the input bias current is 5 pA at 25°C temperature.

It is important with a single-supply comparator or amplifier not to exceed the common mode input range with a signal that goes more negative than ground; the usual specification is to limit negative excursions to less than 0.2 V. This is done automatically if stray capacitance is large, as the peak signal is lower than 0.4 V, but if the peak signal is higher than this, a dual power supply or an input bias voltage is needed. Also, the comparator needs to have a minimum delay which is greater than the latch's hold time to avoid a race condition.

As the logic signals can be shared with many switches, the parts to build a switch are just one or two discretes, a comparator, and a latch.

The redeeming virtue of this circuit is that shunt resistance and shunt capacitance strays across the comparator input do not affect the threshold voltage, just the gain. As the comparator has high gain, there is no change in operation with a stray 10 pF or $10\text{ M}\Omega$ shunt impedance. If the comparator has nonzero input voltage offset, this is not exactly true; the 3702's 5 mV maximum offset will have a small effect, on detection threshold; adding an AC amplifier fixes it. With 10 pF stray, the signal swing across the comparator input is determined by the capacitive divider to be $5 \times 0.2 / 10.2$ or about 100 mV, so 5 mV offset change will change the threshold by 5%.

Simpler AC reference

If a microcomputer is available, the circuit can be further simplified. The $0^\circ/180^\circ$ square wave signals can be replaced by a single interrogation pulse (Figure 6.12).

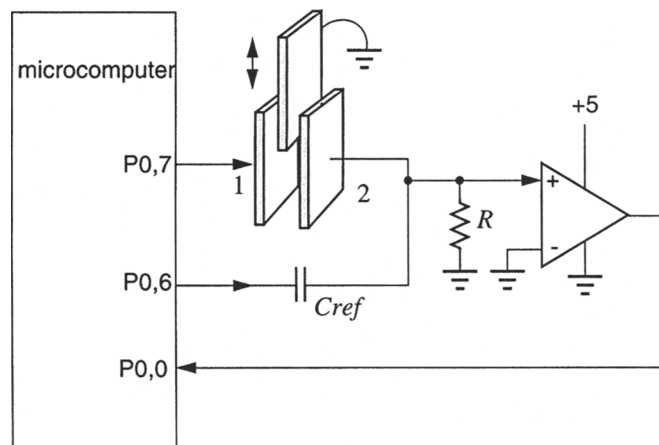


Figure 6.12 Limit switch, single pulse detector

With a microcomputer having an available general-purpose input-output port, this instruction sequence for an 8 bit microcomputer excites and reads the switch.

During power on initialization:

```
out    port0    40H
```

To read the switch:

```
out    port0    80H    ;switch both P0,7 and P0,6
nop                                ;delay by a time interval
nop                                ;longer than the comparator delay
in     port0,0          ;read the comparator; 1= vane out
out    port0    40H    ;reset excitation voltages
```

Noise rejection is handled by multiple reads and discarding a single reading in disagreement with its neighbors.

In Figure 6.13 port 0,7 outputs a positive-going edge at the same time as port 0,6 outputs a negative-going edge. If C_{12} , the coupling capacitance between electrodes 1 and 2, is greater than the reference capacitor, C_{ref} , the comparator output will be a zero. C_{ref} can be built as a printed circuit board capacitance so its characteristics track with C_{12} . Here, too, stray capacitance across the load resistor does not affect operation if the comparator input offset is small.

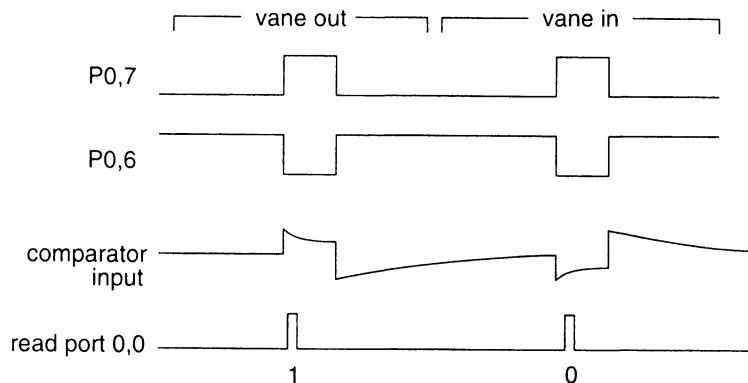


Figure 6.13 Limit switch, pulse reference, waveforms

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 distribu-

tion. 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

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

$q = 1.60 \times 10^{-19}$ 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.

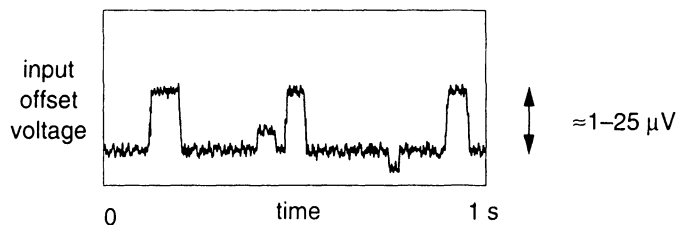


Figure 12.1 Popcorn noise

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

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

where

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

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

f_1 = lower 3 dB point, Hz

f_2 = upper 3 dB point, Hz

R = resistance, Ω

Excess noise

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

where

V_r = 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 – $1 \mu\text{V} \times V_{\text{DC}}^{-1} \times (\text{freq decade})^{-2}$ [Fish, 1994, p. 88]. For a resistor with 1 V drop and $0.1 \mu\text{V}$ 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 $1 \text{ M}\Omega$, $V_r = 1 \text{ V}$, 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 M Ω resistor above.

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

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

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

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

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 $\mu\text{V/V}$ in a decade at the 100 k Ω value; the larger 1406 size shows a value of 0.15 $\mu\text{V/V}$ 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 k Ω and as about $R^{0.75}$ for values above 500 k Ω . The noise of a 1406 size 5 M Ω precision resistor in one decade of frequency, from this reference, is 2 $\mu\text{V/V}$. 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 k Ω 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 $\mu\text{V/V}$), while a 1/2 W carbon composition resistor has –33 dB excess noise. At 10 M Ω , 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).

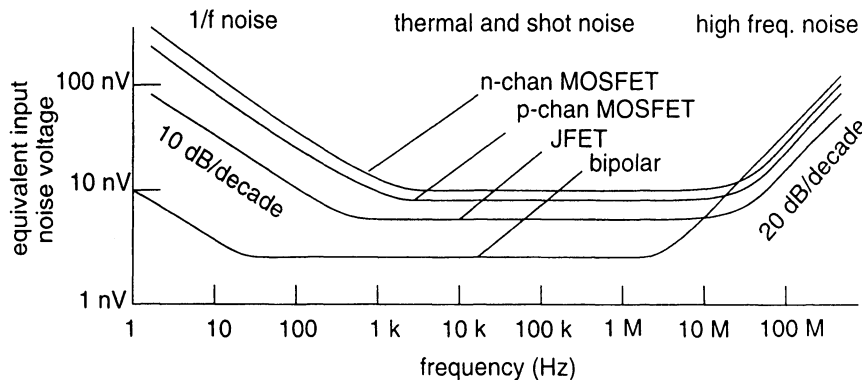


Figure 12.2 Transistor noise

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).

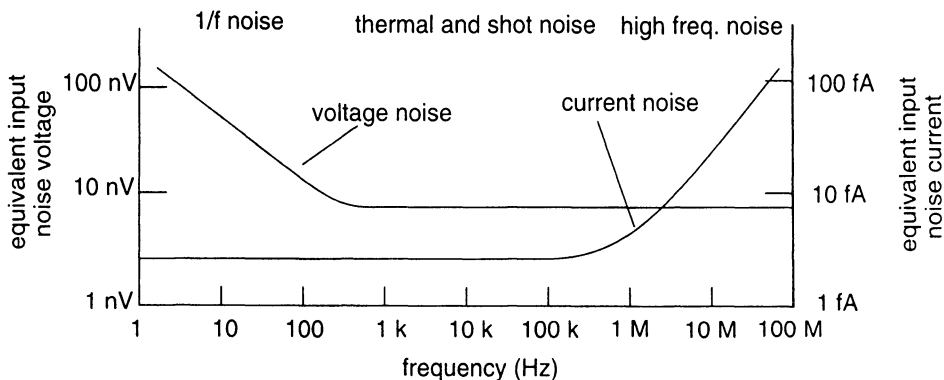


Figure 12.3 FET noise

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 k Ω resistor in series with the output. The input-referred $1/f$ noise of a MOSFET, E_n , is proportional [Takahashi, p. 390] to

$$E_n = \sqrt{\frac{d^2}{\epsilon^2 WLf}} = \frac{\sqrt{WL}}{C} \quad 12.6$$

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

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

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

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

E_{ni} = input-referred voltage noise, rms, per $\sqrt{\text{Hz}}$

0.66 = empirically determined noise constant, ranging from 0.5–1

g_m = transconductance, $1/\Omega$ or S (Siemens)

kT = Boltzmann constant \times 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 $\sqrt{\text{Hz}}$

q = electron charge, 1.602×10^{-19} C

I_{DC} = gate leakage current

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

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

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

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, E_n	$6 \text{ nV}/\sqrt{\text{Hz}}$	$1 \text{ nV}/\sqrt{\text{Hz}}$
$1/g_m$	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

$$E_o^2 = E_n^2 \left[1 + \frac{C_s + C_g}{C_f} \right]^2 + \frac{4kT\Delta f}{\omega^2 C_f^2 R_f} + \left[\frac{I_n}{\omega C_f} \right]^2 \quad 12.9$$

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 R_f , and the third is the effect of the input current noise. If the resistor R_f is large compared to the sensor impedance, the equation becomes

$$E_o^2 = E_n^2 \left[1 + \frac{C_s + C_g}{C_f} \right]^2 + \left[\frac{I_n}{\omega C_f} \right]^2 \quad 12.10$$

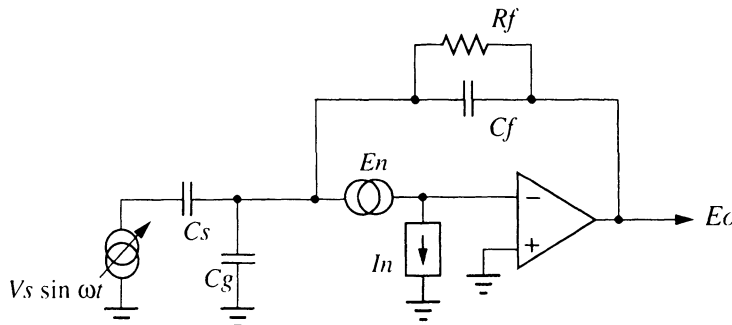


Figure 12.4 Inverting amplifier noise equivalent circuit

12.1.3 High-impedance amplifier

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

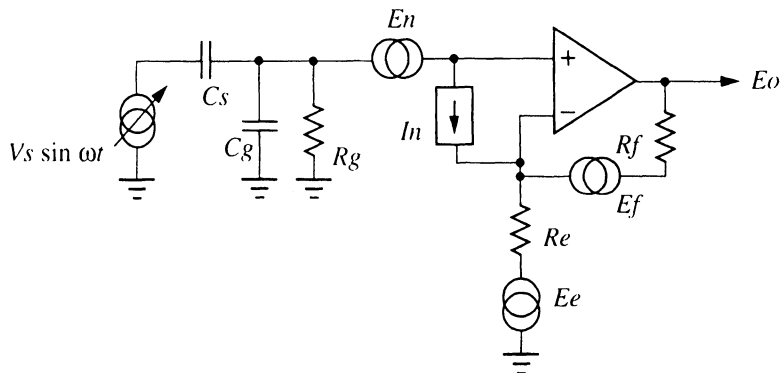


Figure 12.5 Noninverting amplifier noise equivalent circuit

Noiseless feedback resistors R_e and R_f have been added to establish a gain $G=1+R_f/R_e$, and each has a thermal noise generator in series. The output noise of this circuit, assuming the impedance of the parallel combination of C_s , C_g , and R_g is greater than R_e and defining $C_t = C_s + C_g$, is

$$E_o^2 = E_f^2 + G^2 E_n^2 + \left(E_e \frac{R_f}{R_e} \right)^2 + G^2 \frac{4kTR_g \cdot \Delta f}{(1 + sR_g C_t)^2} + G^2 I_n^2 \frac{R_g^2}{(1 + sR_g C_t)^2}$$

Simplifying the circuit with a gain of 1, with $R_f = 0$ and $R_e = \text{infinity}$, and again assuming that R_g is large enough so that its noise contribution is negligible, we have

$$E_o^2 \cong E_n^2 + \left[\frac{I_n}{\omega(C_s + C_g)} \right]^2 \quad 12.11$$

If C_f is set equal to C_s so that the gains of the two circuits are equal, and if $C_g = 0$, the high-Z amplifier has a 2× advantage in output signal-to-noise ratio. If, however, a shield is used with a capacitance C_g 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

$$x_n = \frac{E_{no}}{G} \quad 12.12$$

where G is the responsivity of the sensor in volts per meter of displacement at the amplifier output and E_{no} 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 E_n/I_n is much larger than the impedance of $C_1 + C_d$, the capacitance change in C_1 , ΔC_1 , which produces an output equal to the rms noise is

$$\Delta C_1 = (C_1 + C_d) \frac{E_n}{0.707V} \quad 12.13$$

with E_n evaluated in the bandwidth of BPF.

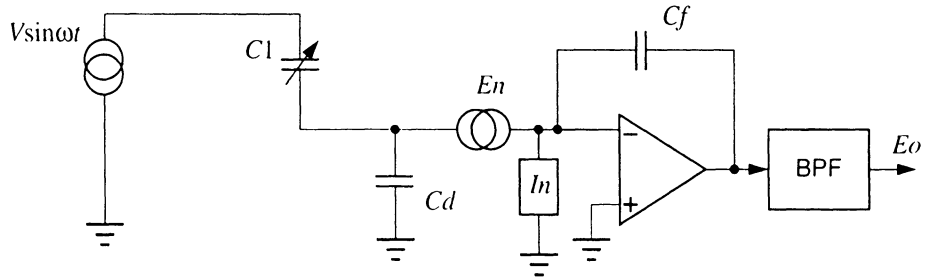


Figure 12.6 Micrometer circuit

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 $C1$ until $C1$ 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 $1 \text{ nV}/\sqrt{\text{Hz}}$ 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 Q 100
- Output resolution, 1 Hz bandwidth 0.01 aF ($0.01 \cdot 10^{-18}$ F)
- Minimum detectable motion $5 \cdot 10^{-6}$ μ m

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

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

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 $8.69 d/\delta$ 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

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

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 377Ω at $\lambda/2\pi$ m away, where λ is the wavelength. The skin depth δ is calculated as

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

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

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

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

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

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

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

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).

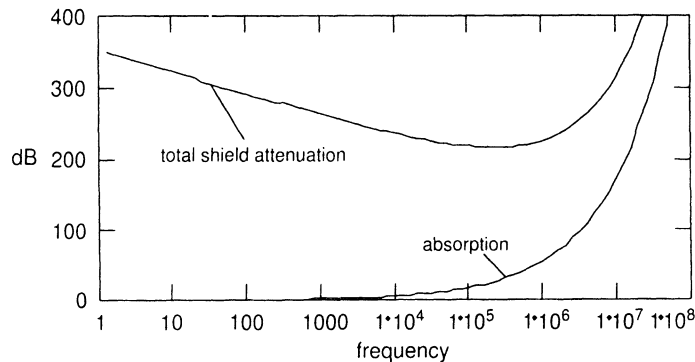


Figure 12.7 Shield attenuation vs. frequency

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).

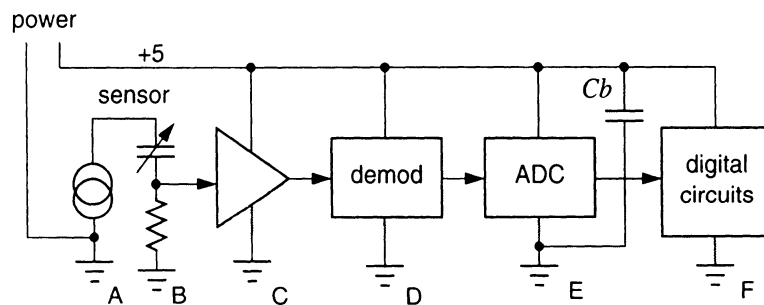


Figure 12.8 Poor ground connection

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 C_b 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 R_b - C_b , and C_b 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.

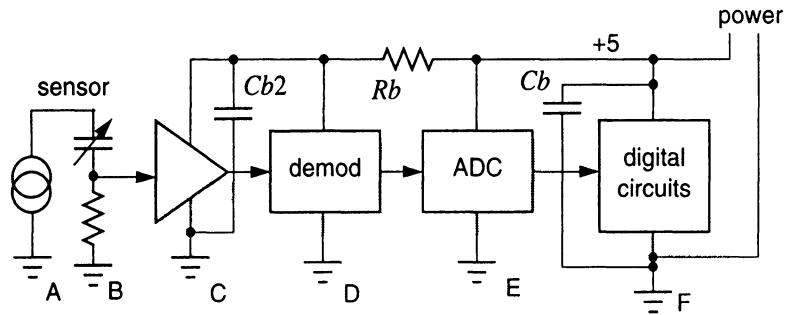


Figure 12.9 Better ground connection

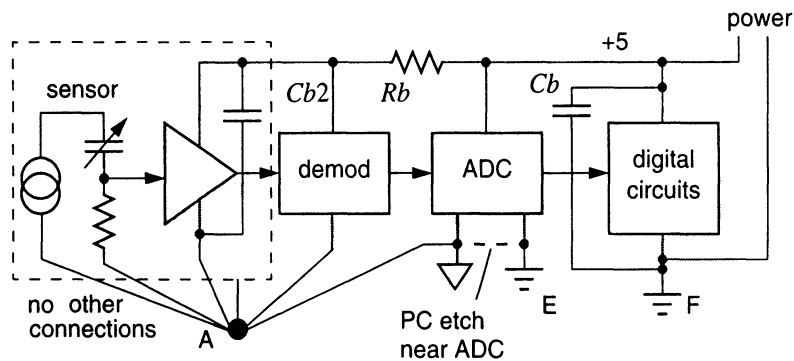


Figure 12.10 Best ground connection

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 $9\text{--}32\ \mu\text{m/m}^\circ\text{C}$ ($5\text{--}18\ \mu\text{in/in}^\circ\text{F}$). Silicon, at $5\ \mu\text{m/m}^\circ\text{C}$, is quite stable, and

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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 $3 \times 10^{-4} \mu\text{m/m}$. With a stable material such as silicon, a temperature change as low as $6 \times 10^{-5}^\circ\text{C}$ 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 $0.877 \mu\text{m/m}/^\circ\text{C}$, but at higher and lower temperatures its stability degrades to a pedestrian 10–15 $\mu\text{m/m}/^\circ\text{C}$. 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 $10^{-7}/\text{yr}$ 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.