5

SIGNAL CONDITIONING FOR REACTANCE VARIATION SENSORS

To obtain a useful signal from the variation of a capacitance or an inductance, we need at least an ac voltage supply for the component and some method for detecting the variations due to the detected quantity. If the intended application requires the use of an ADC (analog-to-digital converter), the signal at its input must be dc and within a standard amplitude range.

In this chapter we describe several circuits that allow us to obtain a useful signal from variable reactance sensors. Its contents parallel those of Chapter 3, but here we emphasize the new concepts that arise from working with alternating voltages, without repeating concepts similar to those for dc measurements such as those relating to interference. However, we describe variable oscillators, which can also be applied to resistive sensors.

For the case of variable transformer sensors intended for angle measurement, the conversion to digital is performed by means of specific converters that are not usually described in ADC books. For this reason we will discuss them here.

5.1 PROBLEMS AND ALTERNATIVES

After measuring the physical quantities, we consider the following final uses of the measurement: immediate analog presentation, conversion to digital, conversion to a variable frequency signal, voltage telemetry, and current telemetry.

Variable reactance sensors can be classified into the following groups: $C_0 \pm C$ or $L_0 \pm L$, that is, simple capacitive or inductive sensors; $L_0 + L$, L_0, as in the case of eddy current proximity detectors where there are two coils but only one of them changes; differential sensors $C_0 + C$, $C_0 - C$, or $L_0 + L$, $L_0 - L$; and sensors that offer an amplitude-modulated alternating signal, such as LVDTs, synchros, and resolvers.

Signal conditioning for all of these sensors must include a supply of exciting alternating current. For capacitive sensors the values for the capacitances are in general smaller than 100 pF. The supply frequency must then be high enough in order to yield reasonable values for their impedances; typical values range from 10 kHz to 100 MHz. Because their output impedance is very high, we frequently use shielded cables for connecting them. But this adds a capacitance in parallel with that of the sensor, resulting in a reduced sensitivity and a decreased linearity. Furthermore, if there is any relative movement between cable conductors and the insulating dielectric, the error can become very serious. The usual solution is to place the electronic circuits as close as possible to the sensor, thus using short cables and even rigid ones, and to apply driven shield techniques or impedance transformers. The trade-offs of different methods for measuring small capacitances have been reviewed in [14].

When the measurement system requires all measured quantities to be converted to dc voltages, some available options for the sensors working at alternating frequencies are peak detection, rms measurement, and mean value calculation after rectifying. Mean value calculation is the most common, but all are well described in most analog electronics textbooks, so we will not discuss them here.

A common solution is to apply Ohm's law. A change in impedance can be detected by measuring the change in current when a constant alternating voltage is supplied to it, or by the change in the drop in voltage across it when driven by a constant alternating current.

When the quality factor Q for the sensor is not very high, all the above methods imply the measurement of two components of the output signal: the one in phase and the one 90° out of phase with respect to the supply signal; but only the in-phase signal carries useful information about the measured variable. In addition the actual impedance variations are sometimes very small, and usually there are stray capacitances interfering with the changes to be measured. Thus reactance measurement techniques based on constant current or voltage supply and Ohm's law are unusual in sensors. One exception is the circuit in Figure 5.1. This circuit applies the constant current supply method to a capacitive displacement sensor based on the variation of the separation of plates in a parallel plate capacitor.

The interesting point for the circuit in Figure 5.1 is that it provides an output voltage proportional to the measured distance, in spite of the nonlinear relationship between the capacity and the distance. That is, if the capacitance changes in the form

$$C_x = C_0 \frac{1}{1 + x} \tag{5.1}$$

then the output voltage is

$$v_o = -v \frac{Z_x}{Z} = -v \frac{C}{C_0} (1 + x) \tag{5.2}$$

A voltage divider offers an alternative solution to Ohm's law for other variable reactance sensors. But for a linear sensor whose impedance varies in the form of $Z_0(1 + x)$ the output voltage for a voltage divider is nonlinear, as we can easily deduce from Figure 5.2*a*.

$$\frac{v_o}{v} = \frac{Z_0(1 + x)}{Z + Z_0(1 + x)} \tag{5.3}$$

Thus this alternative is not attractive, especially if parasitic impedances shunting the sensor are taken into account, for they will produce a serious output error. For the case of differential sensors, however, if in Figure 5.2*a* we have $Z = Z_0(1 - x)$, then the output is

$$\frac{v_o}{v} = \frac{1 + x}{2} \tag{5.4}$$

Now the output changes linearly with x, though there is a constant term that has a relatively high value. This method is practical, and there are integrated circuits intended for this application, such as the one in Figure 5.2*b*.

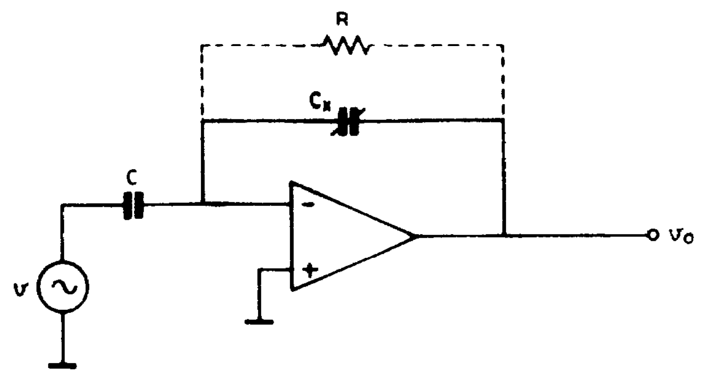

FIGURE 5.1 Linearizing circuit for capacitive sensors. The resistor connected by dashed lines provides a bias current path (Courtesy of Wayne-Kerr).

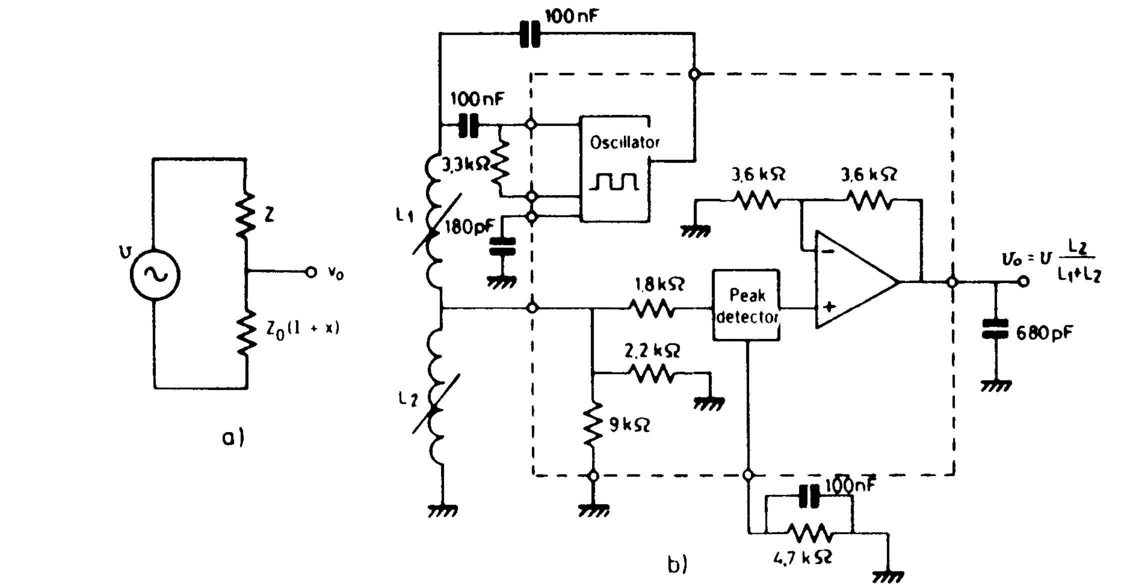

FIGURE 5.2 Voltage divider (*a*) and (integrated) circuit that applies the voltage divider method to differential inductive sensors (Courtesy of Licon Industries Inc.).

5.2 AC BRIDGES

5.2.1 Sensitivity and Linearity

The classical solution to cancel the constant term appearing at the output of a voltage divider, even when it is formed by a differential sensor, is to use a bridge measurement configuration. Assuming that reactive impedances are involved, this bridge must be supplied by a constant current or voltage. When in the bridge only one of the arms has a linear variation with respect to the quantity to be measured, $Z_1 = Z_0(1 + x)$, Figure 5.3*a*, the output is

$$v_o = v \frac{x}{2(2 + x)} \tag{5.5}$$

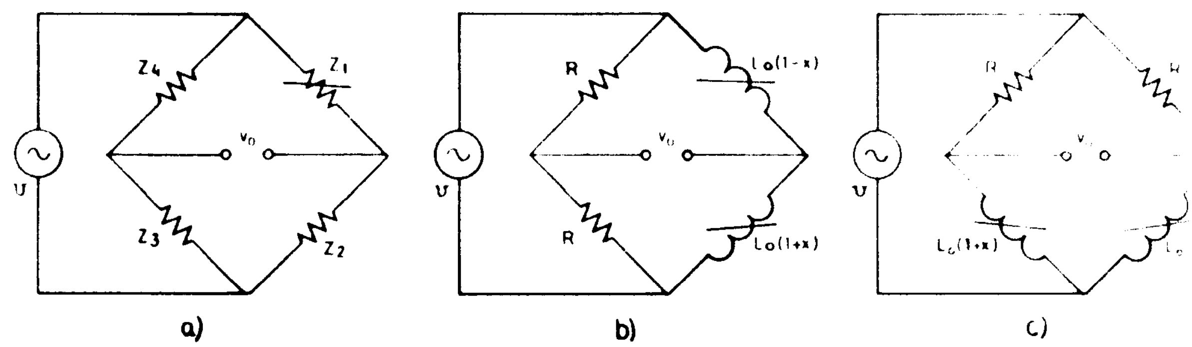

FIGURE 5.3 (*a*) General ac bridge; (*b*) linear ac bridge with resistive arms; (*c*) nonlinear ac bridge with resistive arms.

which shows a nonlinear relationship with x. But in the case of a differential sensor where impedances are placed in adjacent arms, $Z_2 = Z_0(1 - x)$ in Figure 5.3*a*, the output is

$$v_o = -v\frac{x}{2} \tag{5.6}$$

and therefore v_o is proportional to x. Furthermore Section 3.4.4 shows that this configuration will cancel all changes that are simultaneous for both sensors (like the changes due to temperature). This makes ac bridges the most attractive solution for differential sensors.

The impedances for the two remaining bridge arms not occupied by the sensor are chosen depending on the type of sensor. For the case of a differential inductive sensor, resistors can be used. Whenever resistive losses in sensor coils are small, changes in resistances can be neglected, and the circuit in Figure 5.3*b* gives a linear output. When x is small, the circuit gives a double sensitivity but with a poorer linearity.

For differential capacitive sensors, and even for single capacitance sensors when nonlinearity is not of great concern, the errors due to parasitic impedances to ground are very large if resistors are used for the remaining bridge arms. These errors can be reduced by using a bridge with two tightly coupled inductive arms having an accurate winding ratio and a central terminal, forming what is known as a Blumlein or transformer bridge [1].

It consists of a transformer (Figure 5.4*a*) or autotransformer (Figure 5.4*b*) with a central terminal. This yields three terminals able to form the two fixed arms of a bridge. When the oscillator (device 1) is applied to the external terminals of the windings, we have a voltage transformer, and the detector (device 2) is connected between the central terminals of the sensor and the transformer. Conversely, if we place the input oscillator as device 2 and the detector as device 1, we have a current comparator, usually designed as current transformer. An analysis of this can be found in [2].

For capacitive sensor conditioning, the central terminal is usually grounded. By so doing, stray capacitances to ground C_p have a negligible influence on bridge balance. This is because in a voltage transformer, coils

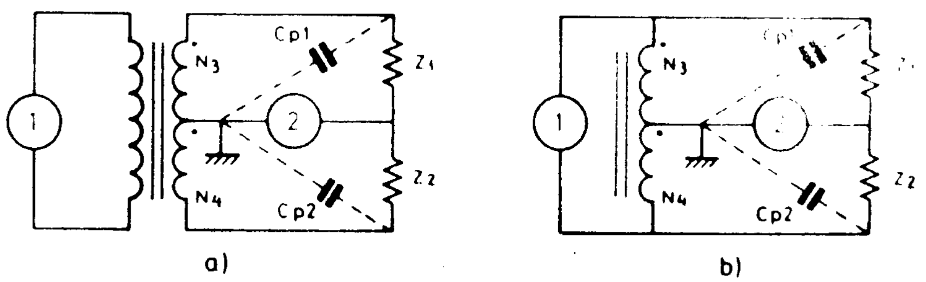

FIGURE 5.4 Blumlein bridges: (*a*) using a transformer; (*b*) using an autotransformer. The oscillator and detector can be either device 1 or 2.

N_3 and N_4 are the ones that determine the voltage ratio across Z_1 and Z_2, assuming that the voltage generator has a low output impedance. For a current transformer, assuming that when in balance the flux density in the core is zero, there is no drop in voltage across capacitances C_p, and therefore these have no effect. This allows us to detect very small changes in capacity even in the presence of much larger stray capacitances. Reference [3] gives some data and results obtained by this technique, the most relevant being the detection of changes of 0.1 fF in 50-pF capacitors in the presence of 1-nF stray capacitances.

To this very important advantage we have to add that the voltage or current ratio are very constant both with time and temperature, as they only depend on N_3/N_4. Furthermore this ratio can be changed very accurately through a broad range of values by placing intermediate terminals in the transformer.

When a three-winding transformer is used, there is a galvanic isolation between the oscillator and detector. This permits both to be grounded without requiring differential measurements when both grounds are different.

Figure 5.5 shows the equivalent circuit for the bridges in Figure 5.4, where Z_d is the detector input impedance and v_s and Z_s are, respectively, the Thevenin equivalent source voltage and impedance. For a voltage transformer having $N_3 = N_4$, we have $Z_s = Z_1 \| Z_2$ and

$$v_s = \frac{v}{2}\frac{Z_2 - Z_1}{Z_1 + Z_2} \tag{5.7}$$

Suppose that the impedance variations are linear. Then for a capacitive differential sensor based on the change in plate separation, $Z_1 = Z_0(1 - x)$, $Z_2 = Z_0(1 + x)$. If a high input impedance detector is used, the output is

$$v_o = v_s = V\frac{x}{2} \tag{5.8}$$

In contrast, for a differential capacitive sensor based on the variation of effective plate area, $Z_1 = Z_0/(1 - x)$, $Z_2 = Z_0/(1 + x)$. It is better to use a low input impedance detector because this gives

$$I_0 = -\frac{vx}{Z_0} \tag{5.9}$$

The corresponding equations for current transformers are much more involved and show that with capacitive sensors it is possible for resonant effects to appear [2]. This makes current transformers less attractive, so they are mainly used for differential inductive sensors whose impedance is so high that stray capacitances should be considered and their effects canceled by means of a transformer ratio bridge.

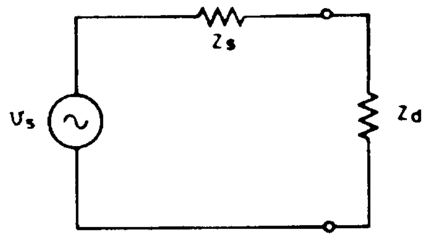

FIGURE 5.5 Equivalent circuit for Blumlein bridges.

5.2.2 Capacitive Bridge Analog Linearization

The increasing interest in capacitive sensors has lead to the development of circuits that yield the advantages of bridges for interference canceling, or ratio measurements. In addition these circuits overcome the nonlinearity of some transformer bridges and are simpler to build. Some of these circuits are generically called pseudobridges [4].

When the sensor is a single capacitor, the circuit in Figure 5.6*a* provides a linear output for a choice of parameters that may change the capacitance. The output is

$$v_o = v \frac{Z_3/Z_4 - Z_2/Z_1}{1 + Z_3/Z_4} \tag{5.10}$$

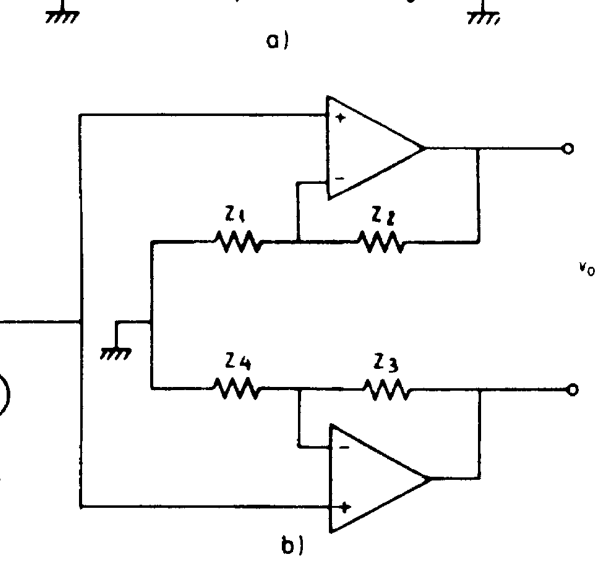

FIGURE 5.6 Capacitive pseudobridges: (*a*) for single capacitive sensors; (*b*) for differential capacitive sensors. In this second case the output voltage is floating.

When the measured variable produces a change in plate separation, then the sensor must be placed at Z_2. When permittivity or area change, then the sensor must be placed at Z_1. In both cases the output is linear as in (5.6) and (5.8). Z_3 and Z_4 can be resistors. Because Z_1 and Z_2 are capacitors (one fixed and the other variable—the sensor), it is necessary to add parallel resistors to provide bias current for the op amp. These resistors may change the output voltage as given by (5.10).

Figure 5.6*b* shows a better solution, which provides a differential output

$$v_o = v\left(\frac{Z_2}{Z_1} - \frac{Z_3}{Z_4}\right) \tag{5.11}$$

If Z_1 or Z_2 is the variable capacitor, then either Z_1 and Z_4 or Z_2 and Z_3 form a differential capacitor. Voltage v_o varies linearly with x. In addition in some cases it is possible to ground one electrode, which simplifies shielding. Resistors added to provide bias currents for op amps or required to stabilize them do not affect the output as long as they are matched.

In this circuit, and also in the previous one, the gain for the op amps at the working frequency must be high enough so that the analysis leading to equations (5.10) and (5.11) is valid.

5.2.3 AC Amplifiers: Power Supply Decoupling

The performance of currently available op amps allows us to easily amplify 10 MHz signals by 10 with a single-stage amplifier. These characteristics are good enough for most ac bridges, so low-cost components suffice in most applications.

Because the central terminal of the ratio arm in ac bridges is usually grounded, one of the output signal terminals is also grounded. Therefore no differential amplifier is required, in contrast with the case for dc bridges. When an inverting amplifier like the one in Figure 5.7*a* is used as detector for the circuit in Figure 5.5, then we have the advantage that the output signal is independent from any parasitic capacitances Z_p shunting the bridge output.

On the other hand, the amplified voltage depends on v_s and Z_s because the output is

$$v_o = -v_s \frac{Z}{Z_s} \tag{5.12}$$

and this can result in a nonlinear dependence on the measured variable, even when the source voltage v_s varies linearly with the variable.

The noninverting amplifier configuration shown in Figure 5.7*b* displays just the opposite characteristics. That is, parasitic impedances affect the amplified signal

$$v_o = v_s \frac{Z_p}{Z_s + Z_p}\left(1 + \frac{Z_2}{Z_1}\right) \tag{5.13}$$

However, if these impedances are high enough, v_o does not depend on Z_s, and therefore if v_s is linear with x, v_o will also be linear. Whatever the case, we can choose impedances to achieve a restricted bandpass appropriate for the signal to be amplified and to reduce excessive noise in the following signal-conditioning stages.

Working at ac frequencies requires us to consider several parameters that limit the performance of op amps. First, input impedances are far below dc values. This is due to input capacitances that exceed 3 pF for the component alone. At 1 MHz this implies an impedance of about 50 kΩ. The presence of sockets and connecting cables further reduces this value.

Other limitations are due to the presence of parasitic capacitances in passive components, particularly in resistors. Figure 5.8 shows a simple dc-coupled inverting amplifier that illustrates the problem. If R_2 = 1 MΩ, then a mere C = 1 pF reduces the −3-dB bandwidth to 160 kHz, even if the op amp has a larger bandwidth. We must therefore avoid high-value resistors and reduce parasitic capacitances.

This bandwidth limitation due to capacitance C can be more restrictive than the one determined by op-amp slew rate (SR). To avoid distortion due to slew rate, the maximal frequency for a signal having a peak amplitude V_p should not exceed

$$f_M = \frac{SR}{2\pi V_p} \tag{5.14}$$

But at the same time, according to the circuit in Figure 5.8, the maximal speed of response at the output when the op amp is considered ideal is

$$\frac{dv_o}{dt} = \frac{v_i}{R_1 C} \tag{5.15}$$

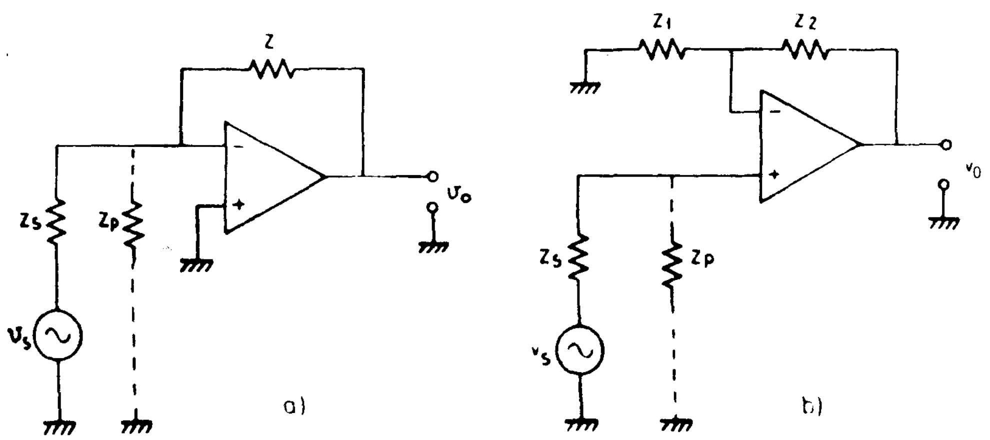

FIGURE 5.7 (*a*) Inverting and (*b*) noninverting amplifier for ac bridges.

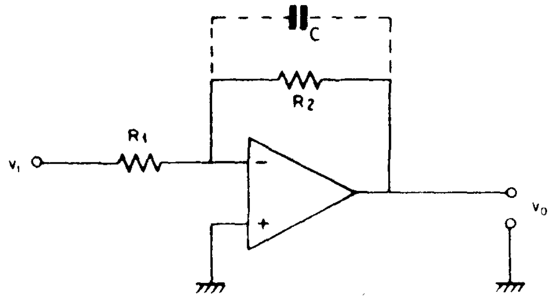

FIGURE 5.8 Stray capacitance C in an amplifier working at medium frequencies reduces the bandwidth.

For an input $v_e = V_p \sin \omega t$, assuming that $v_o = v_i R_2/R_1$, we have

$$f'_M = \frac{1}{2\pi R_2 C} \tag{5.16}$$

Therefore C and R_2 must both be very low. This requires R_1 to be as low as permitted by the output capacity of the signal generator and by the maximal acceptable loading effect.

When working at high frequency it is useful to add power supply decoupling, as shown in Figure 5.9. The objective is to provide a low impedance path to the reference terminal for possible transients coupled to power supply lines.

The problem to be solved is that described by Figure 3.47. The presence of a common impedance can result in large transient voltage fluctuations on the supply terminals of a given component. Op amps have a limited ability to reject these fluctuations, as described by their PSRR (power supply rejection ratio), which decreases at high frequencies. For the OPA 605, for example, which is a 200-MHz bandwidth op amp, the PSRR is 100 dB at dc, 70 dB at 1 kHz, and 30 dB at 1 MHz. It has been shown for the μA741 that it can even oscillate if a large transient voltage is applied to one of its supply terminals.

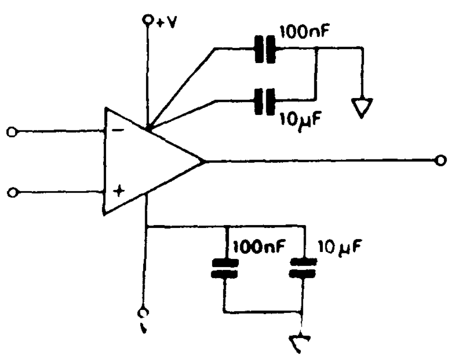

FIGURE 5.9 Decoupling for an op amp minimizes power supply transients.

The goal when decoupling is to reduce the amplitude of transient voltages on op-amp supply lines by forming a voltage divider with the impedance of the supply lines and a conveniently placed capacitor. Usually ceramic capacitors of about 100 nF are used because of their low inductance, shunted by 10 μF tantalum units. Their mounting leads must be as short as practicable, and they must be placed near the amplifier case. In some instances the manufacturer may recommend a circuit different from the one in Figure 5.9.

5.2.4 Electrostatic Shields: Driven Shields

Op-amp characteristics also influence the efficiency of electrostatic shields. Capacitive sensors have an impedance so high that we cannot neglect the parasitic capacitances between the sensor and its environment. Because these parasitic impedances change with movement of the sensor with respect to nearby conductors, they can cause serious errors.

An electric shield has been defined in Section 3.6.1 as a conductive surface enclosing the component or circuit of interest and connected to a fixed potential. The goal of shielding a capacitive sensor is to keep the capacitance constant in the presence of any changes in its electric environment.

There are several options for connecting the shield. For the case shown in Figure 5.10*a*, the total capacitance will be $C_t = C + C_1$, and it will remain constant while the capacitance to ground C_G will change depending on the relative position of the conductors. If the ground terminal is part of the measurement circuit, this causes an error. This kind of shield is therefore used in those cases where we can connect a terminal of the component C to ground. But if this is not allowed, then it is better to use a double shield like the one in Figure 5.10*b*.

Shielding thus keeps parasitic capacitances constant but does not reduce them. In fact shielding increases the value of parasitic capacitance, particularly when it is extended to the cables connecting the sensor to the amplifier (coaxial cables) as is usually done. This increase in capacitance decreases the sensitivity. Instead of grounding the shield, the value for the parasitic capacitance can be reduced by connecting it to a voltage close to that of the

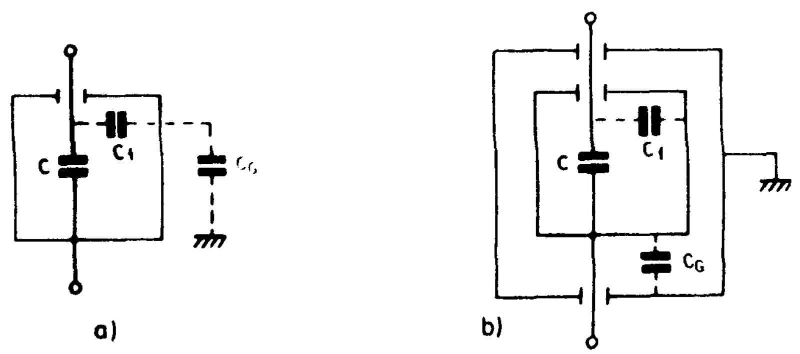

FIGURE 5.10 Single (*a*) and double (*b*) shield for a capacitive sensor C.

conductors inside it. This technique is known as a driven shield and requires the use of electronic circuits before or included in the ac amplifier.

In order to describe how driven shields work, we will examine the case for the coaxial cable in Figure 5.11. When the cable shield is grounded as shown in Figure 5.11*a*, the cable capacitance shunts the sensor and the input capacitance of the amplifier. If instead the shield is connected to a voltage close to that of the internal conductor, Figure 5.11*b*, then we have a driven shield. We can analyze the circuit using the model in Figure 5.11*c*. The corresponding equations are

$$V_o = A(V_i - V_o) = A(I_2 Z_c - V_o) \tag{5.17}$$

$$V_s = I_1(Z + Z_s) - I_2 Z + V_o \tag{5.18}$$

$$0 = -I_1 Z + I_2(Z + 2Z_c) - I_3 Z_c \tag{5.19}$$

$$V_o = (I_3 - I_2)Z_c \tag{5.20}$$

where $Z = Z_d \| (1/Cs)$. From these equations we deduce that the input impedance is

$$\frac{V_s}{I_1} = (A + 1)Z \| Z_c \tag{5.21}$$

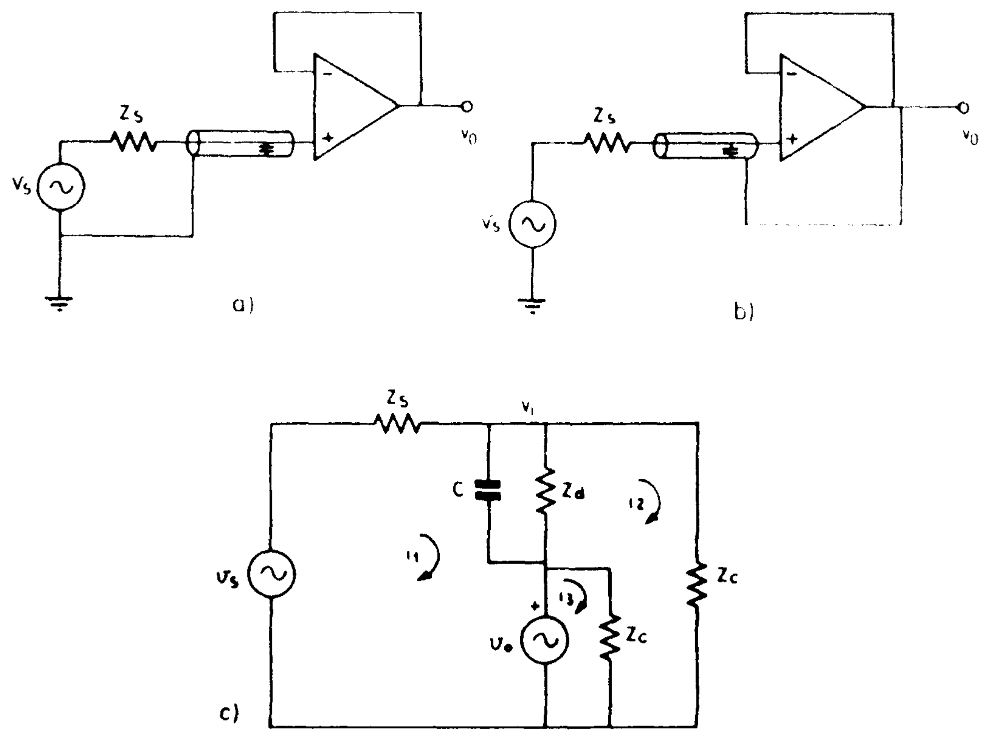

FIGURE 5.11 (*a*) Common electric shield or guard; (*b*) driven shield; (*c*) equivalent circuit for analyzing a driven shield.

That is, the capacitive impedance (and the differential input impedance as well) are multiplied by $A + 1$. Therefore the effective input capacitance for the cable is reduced by a factor somewhat higher than the op-amp open loop gain. This decreases from values higher than 10^6 at dc, to a value between 1 and 10 at 1 MHz. The higher the value for A at the working frequency, the larger is the reduction in effective parasitic capacitance.

5.3 CARRIER AMPLIFIERS

5.3.1 Fundamentals and Structure

A carrier amplifier is required for all sensors whose output is an amplitude-modulated ac signal and that include the zero value in their measurement range, thus implying a change in sign for the measured quantity. That is the case, for example, for LVDTs and also for all those sensors placed in a voltage divider or an ac bridge. Resistance voltage dividers and bridges can be also supplied by ac signals. Then we have the advantage of easier amplification because there is no interference due to component drifts (Section 7.1.1) or to thermoelectromotive forces in the junctions of dissimilar metals (Section 6.1.1).

A carrier amplifier is a circuit that performs the functions of ac amplification, demodulation, and low-pass filtering, including the necessary oscillator, as shown in Figure 5.12. Carrier amplifiers are available in monolithic form (NE 5521, Signetics), but it is common to build them from discrete parts. When the signal from the reference oscillator is used to drive the measured system rather than to supply the sensor, the designations coherent or lock-in amplifier are preferred. When the signal is to be further processed by a computer, it is possible to implement a digital lock-in amplifier by sampling in synchrony with the reference signal; no analog processing other than amplification is then necessary [15].

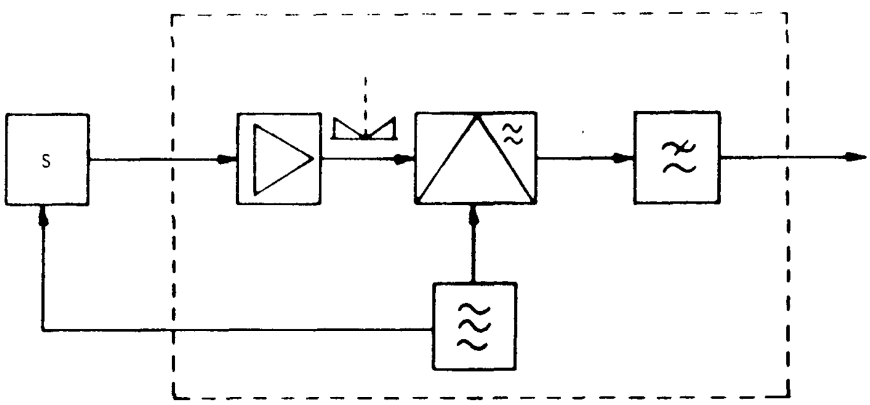

FIGURE 5.12 In a carrier amplifier, the oscillator drives the sensor and the demodulator.

The amplitude modulation in ac sensors arises from the product of the supply voltage times the variable to be measured. For example, for a bridge including a single linear sensor in one of its arms, the source voltage is

$$v_s(t) = v(t)\,\frac{x(t)}{4} \tag{5.22}$$

where we have assumed that the four bridge arms have the same nominal resistance and that $x \ll 1$. If the voltage supply for the bridge $v(t)$ is the output of a sinusoidal oscillator, $v(t) = V_0 \cos \omega_0 t$, and the variable to be measured $x(t)$ is assumed to be in principle sinusoidal, $x(t) = A\cos(\omega_s t + \phi)$, then we have

$$v_s = \frac{V_0 A}{8}\{\cos[(\omega_0 - \omega_s)t - \phi] + \cos[(\omega_0 + \omega_s)t + \phi]\} \tag{5.23}$$

which is a suppressed-carrier amplitude-modulated signal, as shown in Figure 5.13.

V_0 must be highly stable because possible fluctuations would be interpreted as produced by x. Also the bandwidth for x must be at least five times smaller than ω_0 in order to require only simple demodulation methods. Otherwise, very high-order filters would be required to reject any carrier-induced ripple. The bandwidth of the ac amplifier must be at least $0.2\omega_0$.

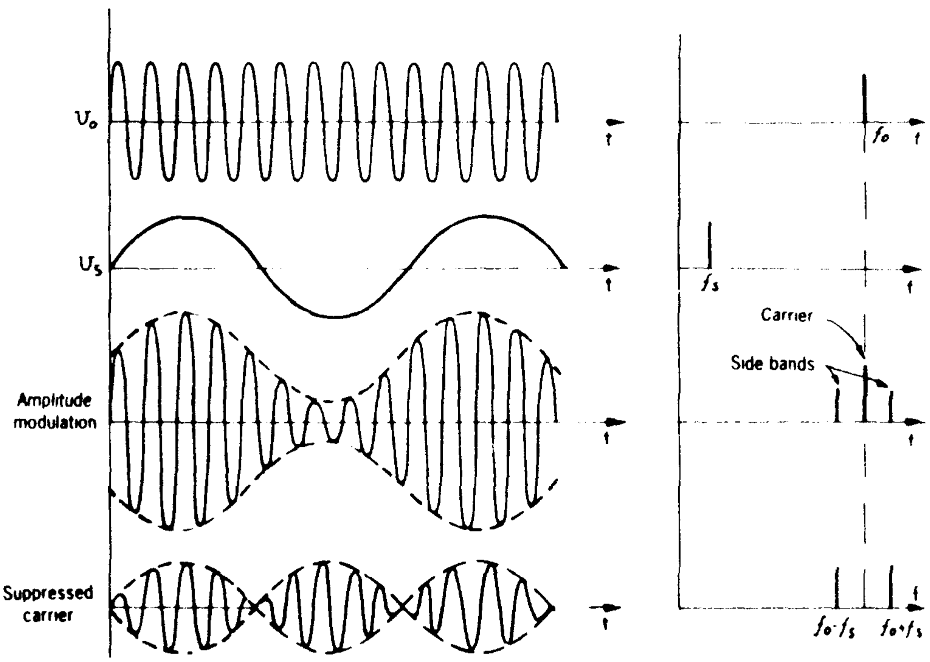

FIGURE 5.13 Amplitude modulation with and without carrier suppression. Waveforms and corresponding spectra for the case where the modulating and carrier signals are both sinusoidal.

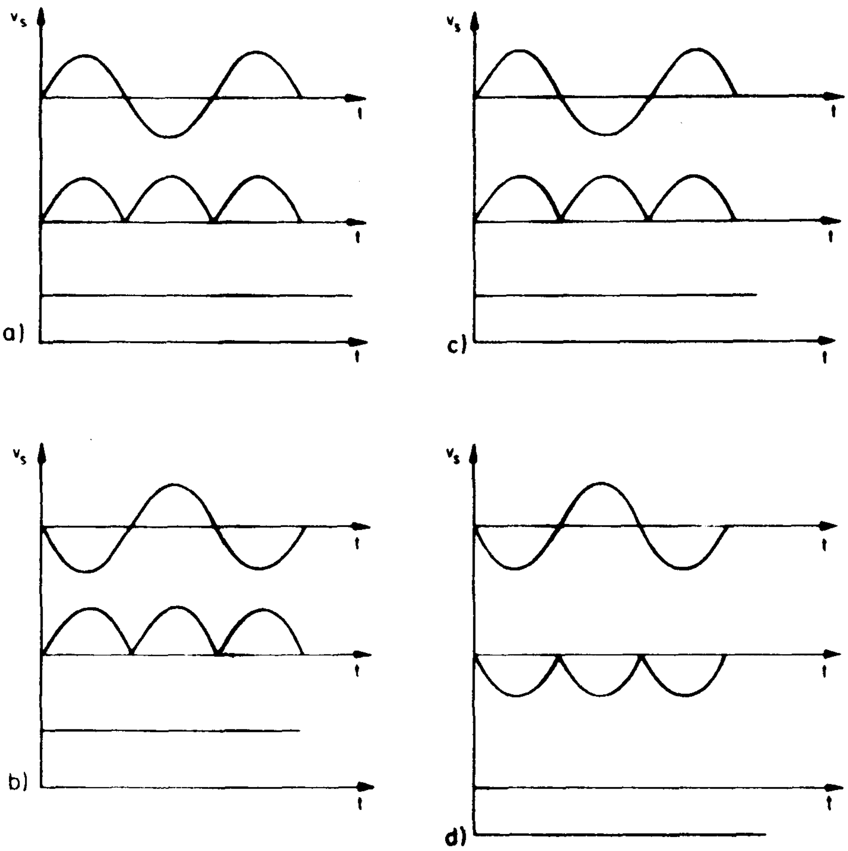

FIGURE 5.14 Phase recovery problem in amplitude demodulation. In cases (*a*) and (*b*) the signal v_s is demodulated by simple rectification and low-pass filtering, and the information about the sign is lost. In cases (*c*) and (*d*) this information is recovered by means of phase-sensitive demodulation.

Typical carrier frequencies for inductive sensors range from 5 to 10 kHz, and the maximal accepted bandwidth for the modulating wave is in general from 0 to 500 or 1500 Hz. The carrier for capacitive sensors ranges usually from 10 to 500 kHz, with a maximal accepted modulating signal of up to 25 kHz for the highest carrier frequencies.

The demodulation in carrier amplifiers must be synchronous. Otherwise, if, for example, simple envelope detection were performed (rectification followed by low-pass filtering), the information about the sign of x would be lost. This situation is described in Figure 5.14. In cases *a* and *b* the same output is obtained in spite of the respective inputs having opposite signs. In cases *c* and *d* outputs with different signs are obtained, corresponding to the respective signal $v_s(t)$.

Phase-sensitive (coherent or synchronous) demodulation consists of multiplying the modulated signal $v_s(t)$ by a reference ac voltage $V_r \cos \omega_0 t$, in phase with $v(t)$, and then filtering the resulting signal with a low-pass filter as shown in Figure 5.15. At the output of the multiplier a voltage v_m is obtained

$$
\begin{aligned}
v_m &= v_s V_r \cos \omega_0 t \\
&= \frac{V_0 A V_r}{8}\{\cos[(\omega_0 - \omega_s)t - \phi] + \cos[(\omega_0 + \omega_s)t + \phi]\} \cos \omega_0 t \\
&= \frac{V_0 A V_r}{16}\{\cos[\omega_s t + \phi] + \cos[-\omega_s t - \phi] + \cos[(2\omega_0 \pm \omega_s)t \pm \phi]\}
\end{aligned} \tag{5.24}
$$

The low-pass filter suppresses the high-frequency component, so at its output we have

$$
v_d = \frac{V_0 A V_r}{8} \cos(\omega_s t + \phi) = \frac{V_0 V_r}{8} x \tag{5.25}
$$

This equation shows that the demodulated signal has the same phase as x.

For the low-pass filter and the oscillator providing the carrier signal, several solutions are available. The filter can be active as the system includes other components that require a stabilized supply. There are commercially available monolithic models but their cost and delivery times may suggest an in-house design using filter design handbooks [6].

The same may be true for oscillators. The amplitude stability required is very high for a sinusoidal output signal. Then a good option is those circuits based on the Wien bridge [7]. For other waveforms, there are many options, but for square waves the op amp's slew rate can be a serious limitation.

5.3.2 Phase-sensitive Detectors

The key element in a carrier amplifier is the demodulator, which is called a *phase-sensitive demodulator* because it can detect polarity changes.

Equation (5.25) shows that phase-sensitive (synchronous) demodulation can be performed by multiplying the modulated signal by a reference voltage synchronous with the carrier and then filtering with a low-pass circuit. In the range of frequencies common for sensors, this product can be performed by

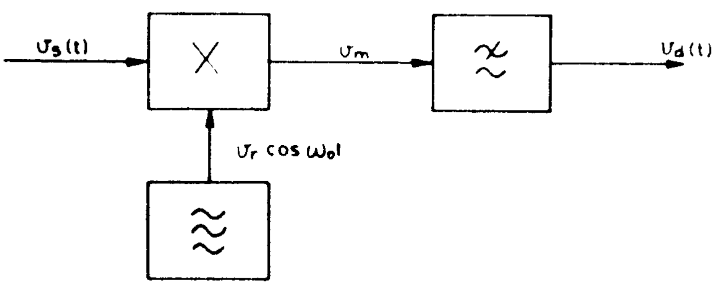

FIGURE 5.15 Phase-sensitive (synchronous or coherent) demodulation. It is assumed that the modulated signal V_s has been obtained from a carrier in phase with the reference voltage.

a four-quadrant multiplier (MC1495, AD633, etc.), but this is in general an expensive solution.

A simpler solution consists of using the switching phase detectors, where the reference signal is not a sine wave but a square wave in phase with the carrier, that is,

$$v_r(t) = V\,\mathrm{sgn}[\cos\omega_0 t] \tag{5.26}$$

where the function "sign" is defined as

$$\begin{aligned} \mathrm{sgn}[x] &= 1 \qquad \text{when } x > 0 \\ &= -1 \qquad \text{when } x < 0 \end{aligned}$$

By expanding (5.26) in a Fourier series, we have

$$v_r(t) = \frac{4V}{\pi}\left(\cos\omega_0 t - \frac{\cos 3\omega_0 t}{3} + \frac{\cos 5\omega_0 t}{5} + \cdots\right) \tag{5.27}$$

If now we multiply equation (5.27) by the modulated signal v_s, we obtain

$$\begin{aligned} v_s(t)v_r(t) = \frac{V_0 A V}{2\pi}\Big\{&\cos(\omega_s t + \phi) + \frac{1}{2}\cos[(2\omega_0 + \omega_s)t + \phi] \\ &+ \frac{1}{2}\cos[(2\omega_0 - \omega_s)t - \phi] - \frac{1}{6}\cos[(2\omega_0 + \omega_s)t + \phi] \\ &- \frac{1}{6}\cos[(4\omega_0 + \omega_s)t + \phi] - \frac{1}{6}\cos[(4\omega_0 - \omega_s)t - \phi] + \cdots\Big\} \end{aligned} \tag{5.28}$$

By low-pass filtering this signal, the terms of frequency $2\omega_0$ and higher are suppressed, so the detected signal is

$$v_d = \frac{V_0 A V}{2\pi}\cos(\omega_s t + \phi) = \frac{V_0 V}{2\pi}x \tag{5.29}$$

Therefore the phase of x is preserved at the output, the same as in equation (5.25). But now we have two definite advantages. First, the output still depends on the voltage amplitude of the reference signal V, but now this is a square wave and therefore it is easier to keep it constant than for a sine wave. Second, the product can be computed using a simple polarity detector (gain of $+1$ or -1) which is much less expensive than an analog multiplier.

Figure 5.16 shows one of the first circuits used for synchronous demodulation based on two matched diodes. The requirement for matched diodes, however, makes it more difficult to use, so it has been replaced by the diode-

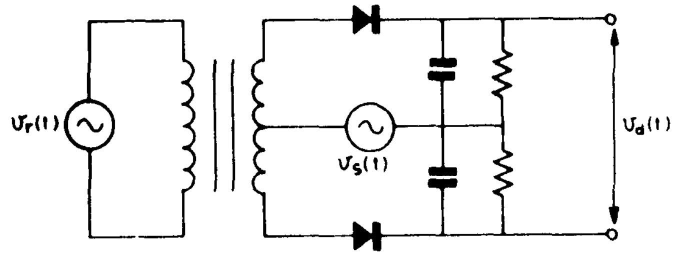

FIGURE 5.16 Phase-sensitive demodulator based on two matched diodes.

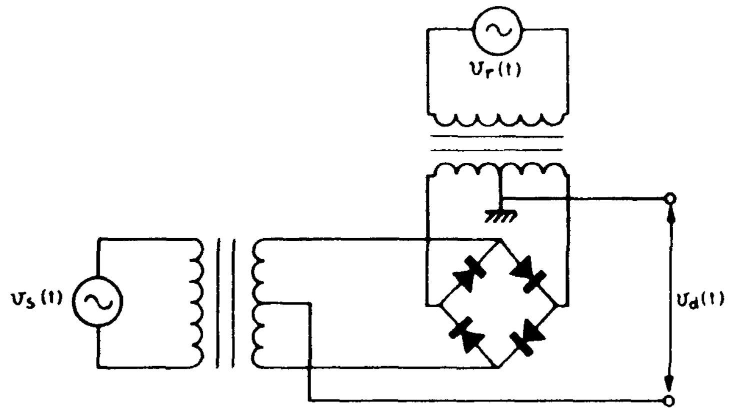

FIGURE 5.17 Diode-ring phase-sensitive demodulator or double-balanced mixer.

ring demodulator or double-balanced mixer shown in Figure 5.17. The behavior of this circuit depends on whether the applied signal is lower than 25 mV. If not, it is necessary to add in series with each diode a resistor larger than the conducting diode resistance [8]. The bandwidth for these demodulators is very large, but because they use transformers, they find limited use at frequencies lower than about 500 Hz. Nevertheless, they are very common in RF circuits and very low-cost units are available.

At low frequencies it is easier to use solid state demodulators like the one in Figure 5.18. They are based on a differential transistor pair where the emitter current is controlled by the signal to be demodulated and an output stage. We use the notation in Figure 5.18, the Ebers-Moll model in the active region, and $I_E \gg I_{ES}$. Then we have for each transistor of the differential pair,

$$I_E = -I_{ES} \exp\left(\frac{qv_{BE}}{kT}\right) \tag{5.30}$$

$$I_C = -\alpha_F I_E \tag{5.31}$$

where α_F is the collector-emitter current gain and $q/kT \approx 25$ mV at 25°C. Also we have

$$
\begin{aligned}
I_0 &= -(I_{E1} + I_{E2}) \\
&= I_{ES} \cdot \left[\exp\left\{\frac{qv_{BE1}}{kT}\right\}\right] \cdot \left[1 + \exp\left\{\frac{q(v_{BE1} - v_{BE2})}{kT}\right\}\right] \\
&= -I_{C1}\,\frac{[1 + \exp\{q(v_{BE1} - v_{BE2})/kT]}{\alpha_F}
\end{aligned}
\tag{5.32}
$$

Given that $v_{BE1} - v_{BE2} = v_r$, from the preceding equation we have

$$I_{C1} = \frac{\alpha_F I_0}{1 + \exp(qv_r/kT)} \tag{5.33}$$

and in a similar way

$$I_{C2} = \frac{\alpha_F I_0}{1 + \exp(-qv_r/kT)} \tag{5.34}$$

where we have assumed that both transistors are equal and therefore have the same α_F.

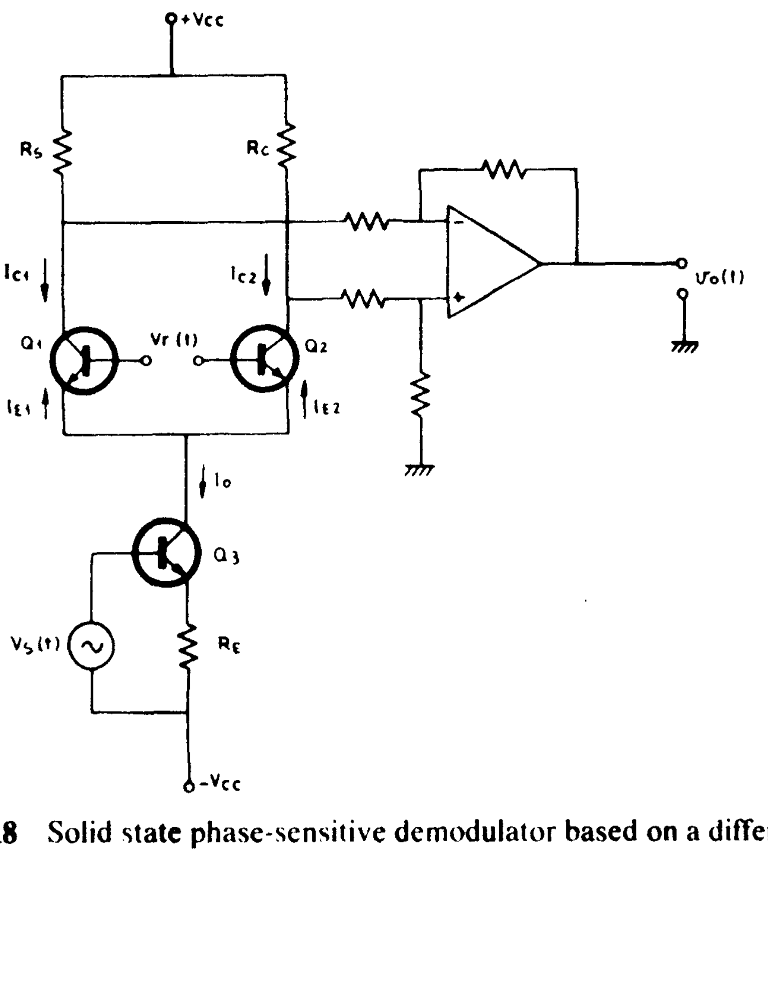

FIGURE 5.18 Solid state phase-sensitive demodulator based on a differential transistor pair.

In the third transistor we have

$$I_0 \approx \frac{\alpha_F(v_s - V_{CC})}{R_E} \tag{5.35}$$

The respective output voltages at Q1 and Q2 collectors are

$$v_{o1} = V_{CC} - I_{C1}R_C \tag{5.36}$$

$$v_{o2} = V_{CC} - I_{C2}R_C \tag{5.37}$$

The output signal is the difference between these two voltages, and therefore

$$\begin{aligned} v_D &= v_{o1} - v_{o2} = (I_{C2} - I_{C1})R_C \\ &= \frac{(v_s - V_{CC})\alpha_F^2 R_C}{R_E}\left[\frac{1}{1 + \exp(-qv_r/kT)} - \frac{1}{1 + \exp(qv_r/kT)}\right] \end{aligned} \tag{5.38}$$

The term inside the square brackets can be approximated by

$$\begin{aligned} \frac{1}{1 + \exp(-x)} - \frac{1}{1 + \exp(x)} &= \frac{\exp(x) - \exp(-x)}{2 + \exp(-x) + \exp(x)} \\ &\approx \frac{1 + x - 1 + x}{2 + 1 + 1} \approx \frac{x}{2} \end{aligned} \tag{5.39}$$

which holds when $x \ll 1$. By using this approximation in (5.38), and therefore by assuming very small values for $v_r q/kT$, we finally have

$$v_D \approx \frac{\alpha_F^2 R_C q}{2R_E kT}(v_s - V_{CC})v_r \tag{5.40}$$

The modulated signal v_s and the reference signal v_r are thus multiplied as desired. The circuit will only work properly if the pair $Q1$–$Q2$ is perfectly matched and without any offset, and therefore it should be in integrated form. There are several integrated circuits suitable for this function, for example, CA3028 (RCA) and NE510 (Signetics).

Amplifiers with a digitally selectable gain of +1 or −1 are becoming an increasingly more frequent solution for synchronous demodulation using equation (5.26) than the methods in Figures 5.16, 5.17, and 5.18. Figure 5.19 shows one of the many available circuits to implement a switched-gain amplifier. The logic inverter indicates that both switches work with opposite phase. When S1 is open and S2 is closed we have

$$v_d = -v_s + 2v_s\left(\frac{R_c}{R_c + R_o}\right) \tag{5.41}$$

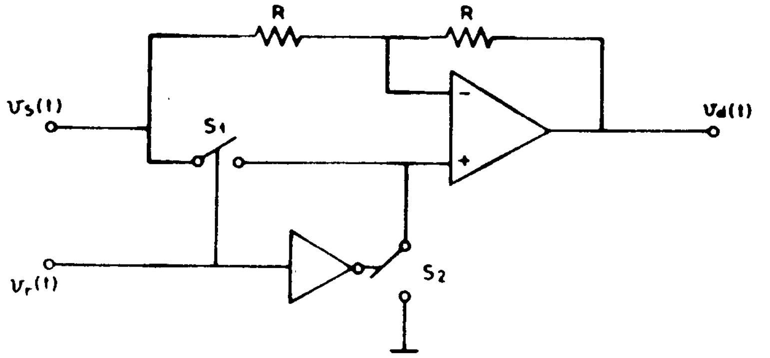

FIGURE 5.19 Switched-gain amplifier with selectable +1 or −1 gains.

where R_o and R_c are, respectively, the resistances for the switch open and closed. For a CMOS switch, for example, they are about 1 GΩ and 100 Ω. Consequently $v_d \approx -v_s$.

When S_1 is closed and S_2 is open, we have

$$v_d = -v_s + 2v_s\left(\frac{R_o}{R_c + R_o}\right) \tag{5.42}$$

so that now $v_d \approx v_s$. If the control signals for S_1 and S_2 are obtained from the carrier for v_s, then we can perform the function indicated by equation (5.26).

The interest of this and other related applications has lead to the development of monolithic circuits that integrate the amplifier and the necessary switches for gain commutation. Figure 5.20 shows two examples.

5.3.3 Application to LVDTs

The output signal for LVDTs is also an amplitude modulated wave, but in some instances its amplitude is so high that no amplification is required and passive demodulators work well enough [9]. Nevertheless, if we are interested in detecting the sense of the core displacement with respect to the central position, then synchronous demodulation is also required.

The simplest solution consists of obtaining a continuous voltage from each secondary winding, then rectifying and subtracting. The sign of the output voltage will indicate the core position. The rectification can be half-wave, Figure 5.21*a*, or full-wave, Figure 5.21*b*. No reference voltage from the primary winding is required as contrasted to phase detectors based on multiplication. This is due to the particular form for the output signal that is given by three or four terminals while the output of ac bridges comes from two terminals. An additional requirement here is that the output or display device must be differential. A shortcoming is that the diodes must work with voltages larger than their threshold, and this is not always possible at the end of the range of some LVDTs. If we must implement circuits that work as ideal diodes, then we would lose the principal advantage of this method, namely, its simplicity.

7

SIGNAL CONDITIONING FOR SELF-GENERATING SENSORS

Generating sensors offer a voltage or a current whose amplitude, frequency, and output impedance determine the characteristics required for the signal-conditioning stage.

When the sensor output voltage or current is small, an amplification is required. The amplification is different from that described in preceding chapters because generating signals are not in the form of a bridge output.

Besides being very small, the voltages to be amplified sometimes have a very low frequency. This prevents the use of ac-coupled high-gain amplifiers because the capacitors required would not be practical. We might consider dc amplifiers, but then we have offset voltage and bias and offset current drifts with time and temperature. High-gain dc amplifiers are based on op amps, so we will first discuss the problems and solutions for their drifts.

In other cases the signal to be processed is not small, but it comes from a high output impedance source. Then because of parasitic capacitances, we need amplifiers having special characteristics or with a structure different from the conventional one.

To obtain high resolution even when drift is not a problem, we must deal with internal noise in amplifiers. Internal noise is inherent to all electronic devices, but we will consider here only op amps and instrumentation amplifiers, which are the most common devices for sensor signal conditioning.

7.1 CHOPPER AND LOW-DRIFT AMPLIFIERS

7.1.1 Offset and Drifts in Op Amps

In an ideal op amp the output voltage is zero when both input voltages are zero. The input currents are then also zero. In a real op amp, neither of these conditions holds. In addition to being different from zero when the input voltage is zero, the input currents are not equal each other. Their difference is called *offset current*. This is due to the imbalance between input transistors (bipolar or field-effect transistors (FETs)). This imbalance also requires an *offset voltage* between the input terminals for the output voltage to be zero.

We can analyze the effects of these voltages and currents by considering a simple amplifier, for example, the inverting amplifier in Figure 7.1. The output voltage is

$$V_o = -\frac{R_2}{R_1} V_i + \left(1 + \frac{R_2}{R_1}\right) V_{io} - I_1 R_2 + I_2 R_3 \left(1 + \frac{R_2}{R_1}\right) \tag{7.1}$$

Resistor R_3 is not necessary for the amplification function itself. However, if its value is chosen such that $R_3 = R_1 \| R_2$, then the equation (7.1) reduces to

$$V_o = -\frac{R_2}{R_1} V_i + \left(1 + \frac{R_2}{R_1}\right) V_{io} + I_{io} R_2 \tag{7.2}$$

where I_{io} designates the offset current ($I_{io} = I_2 - I_1$). The errors for a noninverting amplifier are the same, but the signal voltage gain is $1 + R_2/R_1$.

To refer the error to the input, from (7.2) we have

$$V_o = -\frac{R_2}{R_1} \left(V_i - V_{io} \frac{R_1 + R_2}{R_2} - I_{io} R_1\right) \tag{7.3}$$

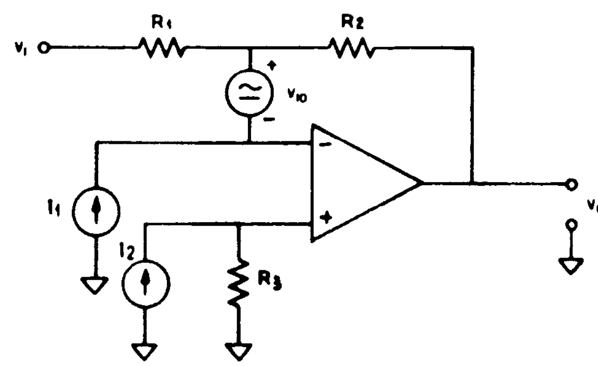

FIGURE 7.1 Offset voltage and bias currents in an op amp.

This shows that the error will increase if high input impedance and high gain are desired (R_1 and R_2/R_1 large). Also, if all resistors are reduced by the same factor, the error due to V_{io} does not change while that due to I_{io} does. We conclude that we must use resistors having a value as small as possible.

The importance of the error terms in (7.3) depends on the values of V_{io} and I_{io}. These values depend on the op-amp technology and quality. As a rule bipolar-input op amps display lower offset voltage and drifts, but FET-input op amps have lower bias and offset currents. Table 7.1 gives some typical value for these parameters.

The effect of offset voltage and currents in a measurement system can be very serious. For example, if a μA741C op amp is used to measure the signal of a type J thermocouple whose sensitivity is 50.2 μV/°C, the offset voltage drift is 15 μV/°C; a mere 1°C change in its temperature reduces the resolution in temperature measurement to 0.3°C. Suppose that the op-amp temperature remained constant. Since $V_{io} = 3$ mV, a 60°C error would result if there were not an initial zero adjustment.

The specifications for thermal drifts in op amps depend on the manufacturer. These drifts are not constant at all temperatures. Rather, they show a nonlinear variation, which is not necessarily monotonic over all the temperature ranges of interest. In particular, the slope usually increases toward the ends of the operating temperature range. If the specifications give the slope at 25°C, this may result in a value lower than the one at the working temperature. Therefore the best approach is to use the worst-case specifications available for the op amp in developing specifications for your op-amp circuit design.

Offset voltages and bias and offset currents also change with time. Specifications reflect this, but drifts are not cumulative. If, for example, the monthly drift is specified, the drift in one year will not be 12 times larger. A good approach is to take a quadratic sum so that the drift in a quarter of year is calculated as $\sqrt{3}$ times the monthly drift and the drift in a year is $\sqrt{12}$ times the monthly drift.

TABLE 7.1 Offset Voltage, Bias Currents, Offset Current, and Their Drifts for a Bipolar op Amp (μA 741C), a FET-input op Amp (TL 071C), and Two Precision (Bipolar) op Amps (AD OP-07 and AD707C)

Parameter	Unit	μA 741C	TL071C	AD OP-07	AD707C
V_{io} maximum at 25°	mV	6	10	0.075	0.015
I_1, I_2, maximum at 25°C	nA	500	0.2	4	1
I_{io} maximum at 25°C	nA	200	0.05	3.8	1
$\Delta V_{io}/\Delta T$	μV/°C	15	10	1.3	0.1
$\Delta I/\Delta T$	pA/°C	—	×2 each 10°C	35	25
$\Delta I_{io}/\Delta T$	pA/°C	500	×2 each 10°C	35	25

A solution for the problem of the op-amp initial offset voltage consists of using the terminal provided in some op-amp models for internal offset adjustment. But this is not always the best solution because this adjustment interacts with bias currents and their imbalance and also with the thermal drift of the offset voltage. Typically for each millivolt of offset voltage canceled, there is an increased thermal drift of 3 μV/°C. Therefore it may be better to add an external voltage to the reference terminal and match input resistances as indicated to go from (7.1) to (7.2). Figure 7.2 shows two circuits that perform both functions. In Figure 7.2*a* we must choose

$$R_3 = \frac{R_1R_2}{R_1 + R_2} - R_4 \tag{7.4}$$

In Figure 7.2*b*, where it has been assumed that R_3 is not adjustable (e.g., it may be determined by the signal source), R_4 must be

$$R_4 = R_3 - \frac{R_1R_2}{R_1 + R_2} \tag{7.5}$$

The use of an external network for offset nulling also allows compensation for offset voltages due to causes other than the op amp, such as a constant difference in potential between two grounds. The compensation range provided by nulling terminals in op amps does not permit compensation for the large offset voltages that sometimes arise due to this cause.

Whatever the case, the nulling of the output voltage when the circuit input is held at the reference voltage must be done after the amplifier has reached its normal operating temperature. Further, temperature gradients in active components must be avoided, and passive components should be of the kind with low-temperature coefficient. Power supplies must be well regulated; otherwise, their fluctuations would show up at the circuit output.

Recent improvements in manufacturing of matched electronic components, presently based on a computer-controlled adjustment that relies on

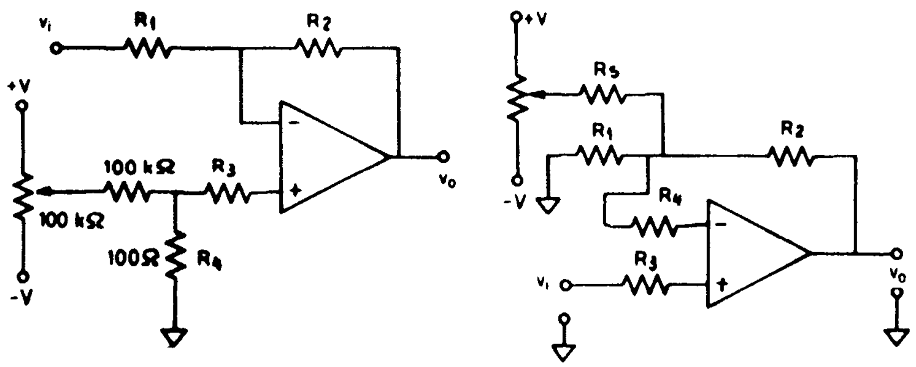

FIGURE 7.2 Offset voltage nulling and bias current compensation in (*a*) an inverting and in (*b*) a noninverting amplifier.

high current pulses, have resulted in op amps with an almost perfectly symmetrical input stage [1]. In the OP-07, the first model of this kind, the input stage is followed by two additional stages besides the output one, instead of only one stage as in conventional op amps (μA741). This results in a lower temperature coefficient for the offset voltage. It can be as low as 0.2 μV/°C and 0.2 μV/month, with V_{io} = 10 μV.

This performance is close to what would be expected from an ideal op amp, although bias currents are somewhat high because they use bipolar input transistors. Nevertheless, its application to low resistance circuits is straightforward without requiring any error corrections as errors are very small. Figure 7.3 shows a circuit for temperature measurement that for a thermocouple with 5 μV/°C sensitivity displays an error of 0.05°C/year, and a 1°C error when the op-amp temperature changes from −25°C to +75°C [1].

7.1.2 Chopper Amplifiers

Before manufacturing techniques made low-drift amplifiers available, some different solutions for drift problems were applied. The most common for many years was to modulate an ac signal with the input voltage, then amplify the ac signal, and later demodulate. In carrier amplifiers, where the carrier is a sinusoidal signal, the modulation process takes place in the sensor. But this is not the case for generating sensors. Here the output of an oscillator is used to modulate a square wave by means of a repetitive switch or chopper, hence the name "chopper amplifiers."

In a chopper amplifier a repetitive switch alternately connects the input of an ac amplifier to the nearly constant voltage to be measured and to a reference voltage (usually ground), as shown in Figure 7.4. The resulting square wave is high-pass-coupled to an ac amplifier whose drifts will not affect the input. The amplified signal is then synchronously demodulated and low-pass-filtered in order to reduce any ripple due to oscillator frequency and its harmonics. For the demodulation process to be simple, any possible high-frequency signal at the input is blocked by low-pass filtering (with respect to the oscillator frequency).

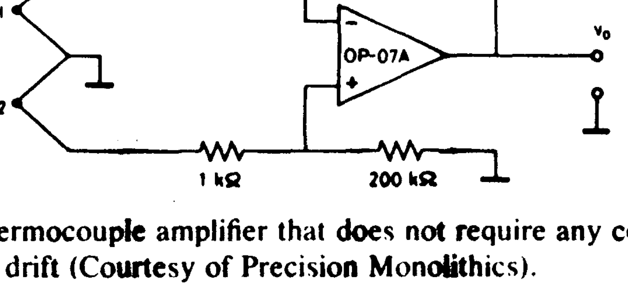

FIGURE 7.3 Thermocouple amplifier that does not require any compensation for offset voltage and drift (Courtesy of Precision Monolithics).

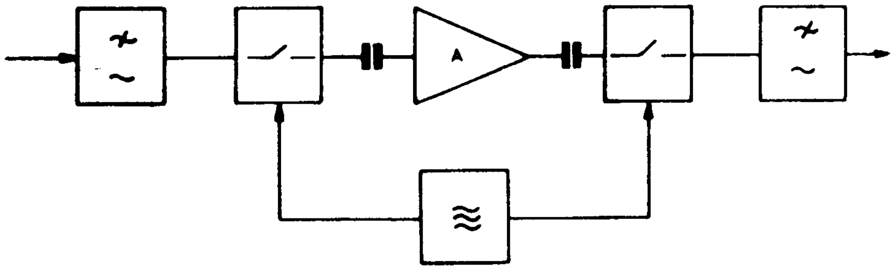

FIGURE 7.4 Principle of operation for a chopper amplifier.

From a mathematical point of view, the process can be described in the following way: The modulation performed by the input switch implies the multiplication of the signal to be amplified by the oscillator's signal. The result is a signal whose spectrum consists of the oscillator frequency and two side bands that are symmetrical with respect to the oscillator frequency and whose amplitude depends on that of the input signal. After amplifying the modulated signal, the demodulation performed by the output switch is again a multiplication by the oscillator signal. In this product all components having the same frequency of both signals will yield a dc component whose amplitude will depend on that of the ac amplifier output. After low-pass filtering, that dc component will be the only component present at the output.

During the amplification process, only the amplifier's high-frequency noise will be superimposed on the signal, thus becoming a source of error. But neither offset voltage drifts nor offset currents drifts will affect the signal that, thanks to the modulation, is now in the frequency band around the oscillator frequency. The only important dc and low-frequency errors will be those due to the input switch (usually based on FETs) that will not normally exceed those of a conventional op amp.

By using this technique, drifts as low as 0.1 to 1 μV/°C and 1 μV/month for the offset voltage and 1 pA/°C for offset currents are achieved. Circuits in Figure 7.2 are still of interest in order to reduce errors due to bias currents.

These performances, exceeding by far those available in conventional op amps are not, however, the only ones to be considered in the amplification process. Further, to achieve these, we must pay the cost of decreased performance for other parameters that are important in some applications. We mention the following parameters:

1. The described circuit allows us to implement an inverting amplifier (when input and output switches operate in opposite phases) or a noninverting amplifier (when both switches operate in phase). But it does not allow us to implement a differential amplifier. Therefore it cannot be applied, for example, to sensor bridges where supply voltage is not floating and where none of the input terminals is grounded.

2. Signal bandwidth is limited because the maximal frequency for the input signal must be much lower than the switching frequency. A 100-Hz bandwidth, however, is easy to obtain, and this is appropriate for output signal conditioning for many sensors.
3. Input impedance is low, usually lower than 1 MΩ. This is due to the limitation on its value imposed by resistors included in the input low-pass filter. Filter component values cannot be chosen very high because noise and drifts increase with the ohmic value of the resistors, and in general the quality for these decreases. Therefore these chopper amplifiers cannot be applied to high output impedance sensors such as pH electrodes.
4. Recovery time after overload can last several seconds because of large capacitors in the filters. This factor can seriously limit the commutation frequency from one channel to another in a data acquisition system.

Present low-drift op amps do better than chopper amplifiers in both performance and cost. For example, their low-frequency noise can be 0.35 μVp-p in the band from 0.1 to 10 Hz, which is 5 to 100 times lower than that of modular and monolithic chopper amplifiers. Further they are free from cross-modulation between input and switching signals and also from interference due to these signals. For all these reasons chopper amplifiers are not preferred in new designs.

7.1.3 Chopper-stabilized Amplifiers

Chopper amplifiers have a narrow band because the frequency for the input signal must be much lower than the chopping frequency. We can, however, design a drift-free dc-coupled wideband amplifier by separating low-frequency from high-frequency components in the input signal, amplifying them separately, and later adding them at the output (Figure 7.5) [2]. This circuit is called a *chopper-stabilized amplifier*. It is clear that the presence of a chopper amplifier does not result in an improved stability, but it overcomes the need for a high-stability dc amplifier. By arranging two of these circuits in parallel, one of them being inverting, it is possible to build a differential amplifier.

The drift improvement obtained is of the order of the gain in the chopper channel, as voltage and output offsets for the main amplifier do not directly add to the input signal. They add now to the chopper amplifier output, after the low-frequency band of the input signal has been amplified. The resulting improvement is a factor of about 100.

Band splitting and separate amplification lead finally to a frequency response like the one shown in Figure 7.6*a*. Note that at low frequency the chopper response predominates whereas at high frequency the ac amplifier

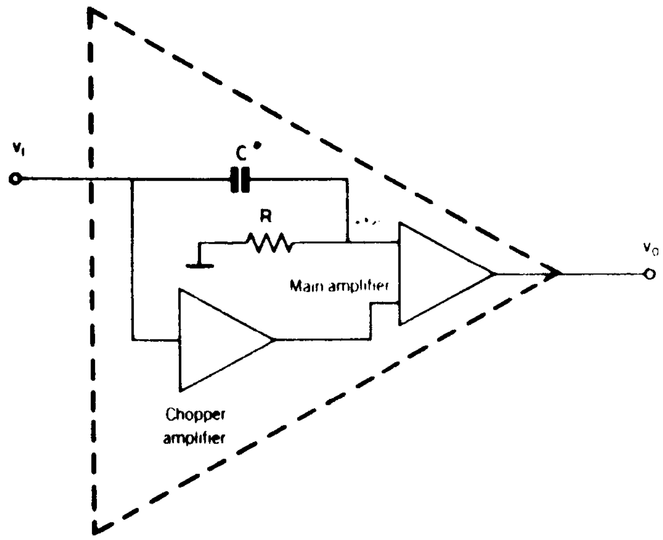

FIGURE 7.5 Wideband chopper-stabilized amplifier.

response does. However, since the amplifier is not used in an open loop but with a large negative feedback, as in an op amp (Figure 7.6*b*), the frequency response for the circuit is flat up to frequencies of about 1 MHz.

7.1.4 Autozero Amplifiers

Chopper amplifiers and chopper-stabilized amplifiers so far discussed are but an implementation based on circuits previously implemented by tubes and electromechanical switches. They are modular or hybrid integrated circuits, and therefore their cost tends to be high. In order to take advantage of the cost reduction provided by monolithic integrated circuits, other circuit structures have been developed. At the same time these have improved the recovery time after overload and the switching frequency transients present at the output of chopper amplifiers.

The method usually applied in monolithic op amps to achieve a very low drift consists of periodically measuring the offset voltage and then subtract-

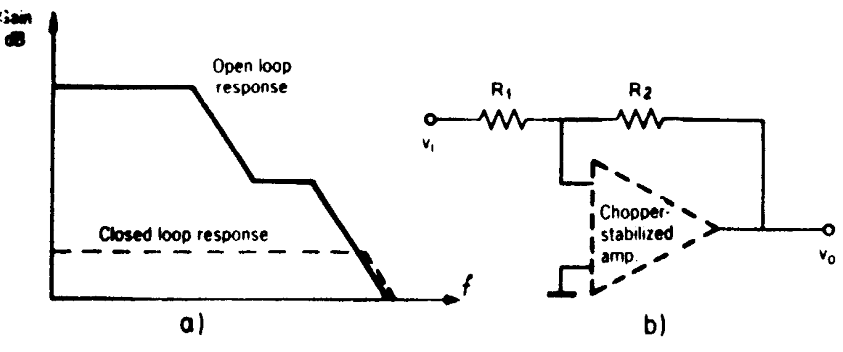

FIGURE 7.6 (*a*) Open loop frequency response for a chopper-stabilized amplifier and (*b*) practical working circuit using negative feedback.

ing it from the signal of interest. While the offset voltage is being measured, a hold circuit presents the signal of interest to the output. Figure 7.7 shows the diagram for one of these circuits.

During the autozero phase, Figure 7.7*a*, switch S_1 short-circuits the amplifier input, while the voltage at the inverting input (common mode voltage) remains applied. The output for the first stage will be then that due to the offset voltage and the CMRR. This output is reduced by means of a negative feedback loop, and at the same time capacitor C_A (external in this model) and switch S_2 form a sampling circuit that holds the voltage required to null the offset voltage.

During the input sampling phase, switch S_1 connects the input signal to the amplifier in differential mode as usual, while S_2 connects the amplified signal to the output stage, free of the offset voltage because of the action in the previous phase and of C_A. Capacitor C_B, also external, holds the voltage necessary to give the correct output during the next autozero phase. If the switching from one phase to the other is performed quickly enough, these commutations do not have any influence on the output waveform. The real circuit includes additional elements intended to improve high-frequency response and stability, but the circuit described is the one that yields the desired low-frequency characteristics.

There are several models available that use the preceding method or similar ones. Some of these include the capacitors; others accept an external clock to determine the time duration for each phase. This permits, for example, synchronization of several nearby units to avoid interference between oscillators having close frequencies, which could result in an increased output noise.

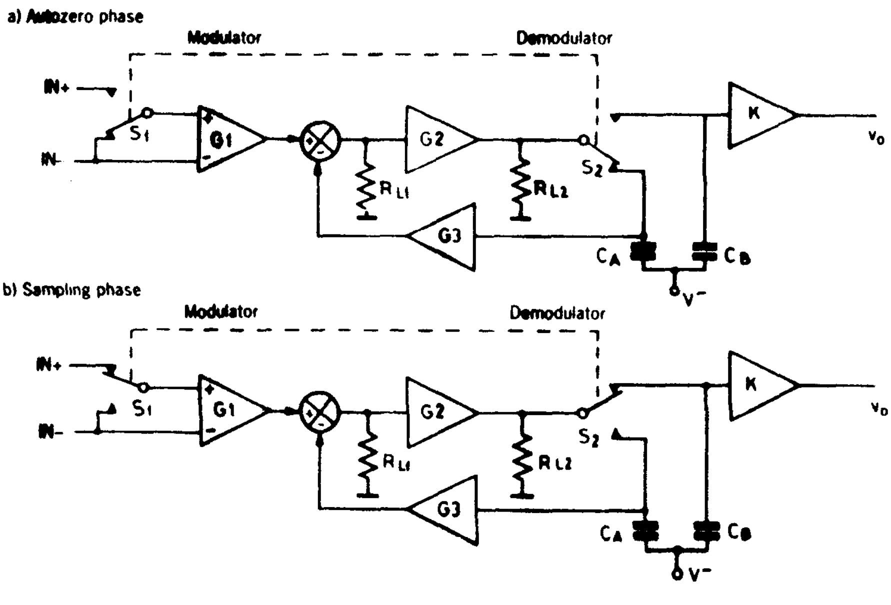

FIGURE 7.7 Simplified circuit for an autozero op amp, (*a*) during the autozero phase and (*b*) during the sampling phase (Courtesy of Linear Technology Corp.).

Op amp models ICL 7652, LTC 1052, Max 420, Si 7652, and TSC 900 include autozero circuits. Typical specifications are 5 μV offset voltage with 0.05 μV/°C drift, 1 to 100 pA bias currents, and CMRR and PSRR ranging from 110 to 120 dB. Some models including the hold capacitors are Max 530, LTC 1050, and TSC901. The already mentioned LTC 1052, the ICL 7650S, and the Max 421/423 have an improved speed of response after overload.

This last technique for drift correction can also be applied to circuits implemented with discrete components if use is made of a sample and hold circuit. Figure 7.8 shows a practical implementation for this method in a high-gain amplifier. The input is periodically connected to the reference voltage, and the resulting output value is held on a sample and hold circuit. To avoid the output transient during the time when the input is at zero volts, another sample and hold circuit can be added in series with the output.

7.1.5 Offsets and Drifts in Instrumentation Amplifiers

Whenever a low-frequency signal coming from a differential output sensor is to be amplified, such as that from load cells based on a strain gage bridge, an instrumentation amplifier is required. This can be implemented by the circuit in Figure 3.40 using low-drift op amps, or by an IC model. In any case the specifications for the offset voltage and its drifts are somewhat different from those discussed in Section 7.1.1 for op amps.

The presence of the two stages shown in Figure 3.40, one with a selectable $G + 1$ gain and the second one with a fixed gain k, results in an output offset voltage,

$$V_{oo} = (V_{o2} - V_{o1})(G + 1)k + V_{o3}(k + 1) \tag{7.6}$$

where V_{oi} are the respective offset voltages for each op amp. The output offset voltage depends therefore on the present gain G plus a fixed term due to V_{o3}. If the contribution from V_{o3} is small, then the input-referred offset voltage is almost gain independent. But the same is not necessarily true for all drifts.

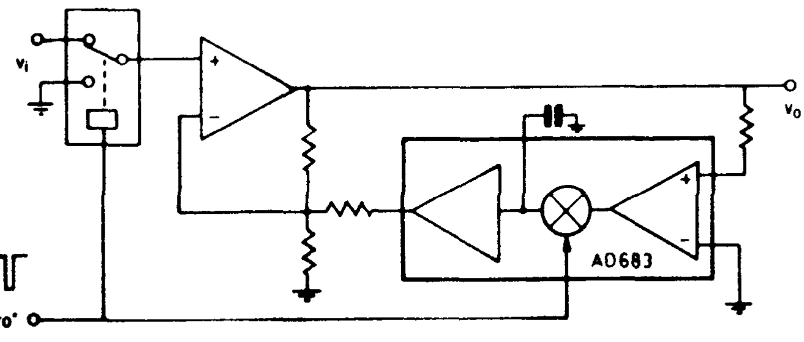

FIGURE 7.8 Circuit for automatic drift correction in an amplifier using discrete components (Courtesy of Analog Devices, Inc.).

For the AD 624C, for example, the maximal offset voltage for the input stage is 25 μV, and its maximal thermal drift 0.25 μV/°C. Hence its contribution to the output is obtained by multiplying these values by the gain. The offset voltage for the second stage is 2 μV, and its thermal drift is 10 μV/°C. These voltages are directly added to the output, independent of the gain. Since the drifts are larger, they will predominate at low-gain settings. The change in the total output offset voltage due to fluctuations in supply voltages is attenuated at least 80 dB for gain 1, 110 dB for gain 100, and 115 dB for gain 1000. This model has two adjustments for the offset voltage. One acts only on the input stage and is used for high gains, and another one acts on the output stage and is used for unity gain amplification.

7.2 ELECTROMETER AMPLIFIERS

Signals coming from current sources or from high output impedance voltage sources—for example, semiconductor-junction-based nuclear radiation detectors, photoelectric cells, photomultiplier tubes, ionization cells (e.g., for vacuum measurement), and piezoelectric sensors—require a measurement system featuring a low input current. When low-frequency components are of interest, then a current-to-voltage converter based on a low-drift op amp is required. Otherwise, as in piezoelectric sensors that do not have dc response or in radiation detectors that only have to detect incoming pulses, either an electrometer or a charge amplifier can be used.

An electrometer amplifier, or just an electrometer, is an expensive measuring system having an input resistance larger than 1 TΩ and an input current lower than approximately 1 pA. Small currents can be measured with an electrometer amplifier by two different methods: by directly measuring the drop in voltage across a high-value resistor, Figure 7.9*a*, or by means of a current-to-voltage conversion based on an electrometer grade op amp, Figure 7.9*b*.

In the first method, when R has a high value, it is not possible to measure high-frequency phenomena because the capacitance of the sensor together

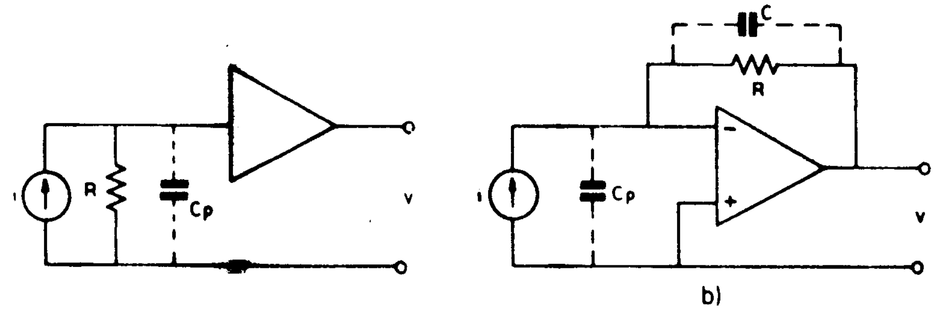

FIGURE 7.9 Methods for measuring small currents using an electrometer amplifier: (*a*) detecting the drop in voltage produced in a resistor; (*b*) performing a current-to-voltage conversion.