4
Circuit basics

The circuit which converts the variable sensor plate capacitance into an output signal should have these characteristics:

- Good linearity
- Shield or guard to isolate the input from stray electric fields
- Insensitivity to stray capacitance to ground on sensor electrodes
- Low noise
- Adequate signal bandwidth
- Correct choice of carrier frequency and waveshape

This chapter discusses the basic circuit building blocks, while Chapter 10 covers these topics in more detail.

4.1 LINEARITY

Capacitive sensors are either area variation, with the plates sliding transversely and no change in spacing as in "Area variation with high-Z amplifier" in Section 3.3.4, or spacing variation where just the spacing changes, as in "Spacing variation with low-Z amplifier" in Section 3.3.1. By looking again at the formula for parallel-plate capacitance

$$C = \varepsilon_0 \varepsilon_r \cdot \frac{A}{d}$$

and the equation for capacitive impedance

$$Z = \frac{1}{2\pi f C}$$

we see that *with area-variation sensors, the capacitance is linear with area* but the impedance is not, and *with spacing-variation sensors, the impedance is linear with spacing* but the capacitance is not. Correct circuit design will linearize either type of sensor.

4.2 MEASURING CAPACITANCE

4.2.1 Circuit comparison

The choice of a capacitive sensor circuit (Table 4.1) should consider the required system accuracy, the cost and space available, and the noise environment. At one extreme, a few components only can handle capacitance measurement with the DC circuit, but at a cost of considerable noise sensitivity and drift. For applications needing more precision in a high noise environment, the synchronous bridge circuit is unsurpassed.

Table 4.1 Capacitance measurement circuit comparison: 0 is good, 5 is bad

Circuit	Function	Sensitive to stray capacity	Sensitive to noise	Needs ADC	Bandpass filter available	Size	Sensitive to shunt resistor
DC	volt = $1/C$	yes	5	no	no	0	5
RC oscillator or 1-shot	freq ≅ $1/RC$ period = RC	yes	5	no	no	1	5
IC oscillator (current-capacitance)	freq ≅ $1/RC$ period ≅ RC	yes	3	no	no	2	5
LC oscillator	freq≅$1/\sqrt{LC}$	yes	2	no	yes	3	0
Sync., single ended	volt ≅ $C1$-$C2$ or $C1/C2$	no	1	yes	yes	4	1
Sync., bridge	volt ≅ $C1$-$C2$ or $1/C1$-$C2$	no	0	yes	yes	5	0

The different capacitance measurement circuits are rated by seven characteristics.

Function

For free-running *RC* oscillators, the output frequency in radians/second is proportional to $1/RC$. The output period is, of course, the reciprocal, or *RC*. With a one-shot *RC* oscillator the ON time varies directly as capacitance.

The two synchronous demodulator circuits can have different functions depending on how the input amplifier is connected; for single-ended circuits the output voltage is a function of a reference capacitor $C1$ and the sense capacitor $C2$.

For synchronous demodulators using a bridge, the output voltage is proportional to $C1 - C2$ or $(C1 - C2)/(C1 + C2)$. In any case, the appropriate function should be chosen to linearize the circuit for spacing-variation or area-variation sensors.

Sensitive to stray capacity

Sensor plates may have signal capacitances in the fractional pF range, and connecting to these plates with 60 pF/m coax would totally obscure the signal. With correct guarding, however, the shield coax and any other stray capacitance can be almost completely nulled out, as shown in Figures 3.11 and 3.12. This guarding is simply a matter of adding a connection with synchronous demodulators, but it is more difficult to guard the oscillator circuits.

Sensitive to noise

One hazard of the oscillator circuits is that the frequency is changed if the capacitor picks up capacitively coupled crosstalk from nearby circuits. The sensitivity of an *RC* oscillator to a coupled narrow noise spike is low at the beginning of a timing cycle but high at the end of a cycle. This time variation of sensitivity leads to beats and aliasing where noise at frequencies which are integral multiples of the oscillator frequency is aliased down to a low frequency. This problem can usually be handled with shields, as shield effectiveness is high for electrostatic fields. Careful power supply decoupling is also needed.

Needs ADC

Frequency is easily converted to a digital number by counting pulses for a fixed time interval, so no ADC may be needed in a typical microcomputer application with *RC* oscillator detectors. Period is similarly converted to digital by counting fixed clock pulses during the measurement interval.

For synchronous demodulators, an ADC may be needed to convert the output voltage to digital, but the ADC can be easily integrated as shown in Chapter 18.

Bandpass filter available

Wideband noise can be substantially reduced for the synchronous demodulators by adding a bandpass filter tuned to the excitation frequency, but no such option is available for the *RC* and *IC* oscillator solutions. This suggests that these oscillators are not the circuit of choice for very sensitive applications.

Circuit size

The oscillators are much lower in component count than are the synchronous demodulators. This is a considerable advantage for conventional printed-circuit-board construction. For integrated circuits, however, switched capacitor methods (see Chapter 11) easily integrate a synchronous demodulator in 1or 2 mm^2 of area.

Sensitive to shunt resistor

In an application where the circuit may be exposed to contaminants or excessive humidity, resistive paths on the surface of a printed circuit board can affect circuit operation. A very important characteristic of circuit design is sensitivity to resistive or conductive shunts. Correct guarding can handle these problems as well as canceling the effect of stray capacitance, but it is not an available option with many oscillator and one-shot capacitance measurement circuits. An exception is shown in Figure 4.3 on page 52. *LC* oscillators are insensitive to shunt resistance as long as circuit Q remains high enough to sustain oscillation, and correctly designed synchronous demodulators are insensitive to shunts.

4.2.2 Direct DC

The first entry in Table 4.1, direct DC, is the simplest detector circuit. With a very high impedance amplifier, capacitance changes can be measured as DC voltage differences, simply by charging the capacitor to be measured and connecting it to the amplifier input. As the charge is then nearly constant the capacitor voltage will vary as the reciprocal of capacitance or directly with spacing by the relationship $Q = CV$. The time constant RC where R is the amplifier input resistance and C the capacitance being measured must be greater than the time measurement period so as not to introduce a low frequency loss; thus an electrometer-type amplifier with input currents in the femtoampere region is often needed. Electret microphones use this method with a junction FET for an amplifier; the resistance to ground which is needed to keep the capacitor voltage from drifting outside the amplifier's linear range is contributed by the FET's input leakage currents.

For direct DC circuits using an operational amplifier an input resistor of very high value is needed, or a bootstrap circuit can be used to increase the AC input resistance, as shown in Figure 4.1.

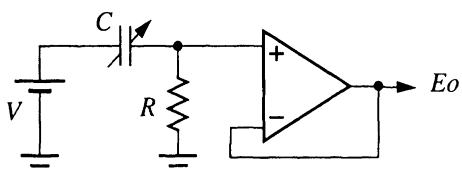

Figure 4.1 DC capacitance circuit

When C varies at frequencies above $1/RC$

$$Eo = \frac{Q}{C} = \frac{C_{avg}V}{C}$$

where Q is the charge on the capacitor C and C_{avg} is the capacitance of C with no displacement.

4.2.3 Oscillator

The direct DC measurement circuit above cannot handle very slow capacitance variations without a very high input impedance amplifier. With a 10 pF capacitor, measuring mV-level signals at 1 Hz requires an amplifier with offset currents in the 0.01 pA range. Also, measurement at DC will admit other unwanted disturbances such as cable noise, thermocouple voltages, power frequency crosstalk, semiconductor $1/f$ noise, and slow variation of component parameters. Circuits which use a high frequency excitation are preferred; for example, the reactive impedance of the unknown capacitor can be measured by using it as the tuning element in an oscillator. Several different types of oscillator can be chosen with different advantages; with an RC oscillator, the frequency is proportional to $1/RC$, but with LC oscillators the frequency is proportional to $1/\sqrt{LC}$ and it is more difficult to linearize. A gyrator circuit (Figure 4.2) which converts capacitance to inductance can be used to save an inductor and to change the output frequency to $1/\sqrt{C1\,C2}$; if $C1$ and $C2$ are both sense electrodes of equal value the response becomes $1/C$.

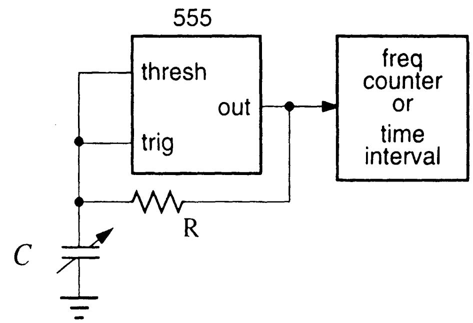

Figure 4.2 *RC* oscillator circuit

Figure 4.2 is a little different from the usual 555 oscillator, but it saves parts, uses a grounded capacitor, produces a 50% duty cycle, and is more linear providing a CMOS-type integrated circuit like the 7555 is used. The normal 555 output does not swing to the power rails and is unstable with temperature in this circuit. The circuit operation is the classic Schmitt-trigger-with-*RC*-feedback, with the Schmitt trigger points accurately determined by the 555 as 1/3 and 2/3 of the power supply. As the output voltage with the CMOS part is accurately driven to the power rails provided that *R* is sufficiently large, the output frequency is unaffected by supply variation. Adding an analog switch allows the measurement capacitor to be switched between several measurement capacitances and a reference capacitor to compensate for changes in resistance and to establish a capacitance-ratio output for greater accuracy.

The output is converted to a digital value using either a frequency counter or a time interval counter with an accurate, usually crystal-controlled, timebase. A frequency counter is appropriate to linearize spacing-variation sensors and a time interval measurement is correct for area-variation sensors.

Guarding the *RC* oscillator

The unguarded *RC* oscillator may have problems with stray capacitance and leakage resistance at the sense capacitor node. The addition of a FET-input op amp as shown in Figure 4.3 produces a low-impedance guard voltage. Adding a guard to the amplifier's *Vee* terminal (Figure 10.9) will further reduce stray capacitance to a fraction of a pF.

Bridge circuit with *RC* oscillator

The advantages of the bridge circuit include a more stable ratiometric response if a matched pair of sense capacitances is used. The *RC* oscillator does not directly accept bridge inputs, but an *RC* oscillator can be configured for a ratiometric response as shown in Figure 4.4.

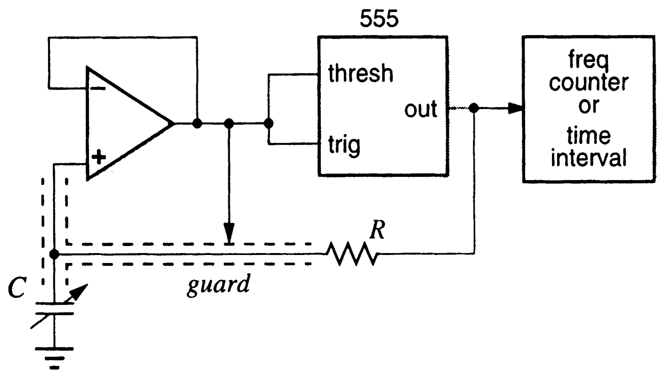

Figure 4.3 Guarded *RC* oscillator

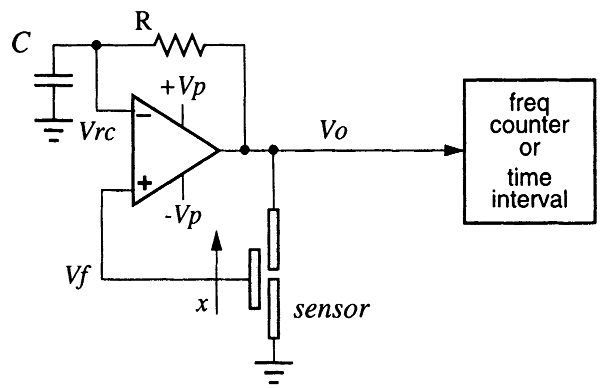

Figure 4.4 Bridge circuit with *RC* oscillator

This circuit shows an *RC* oscillator with the usual output function reversed: frequency is proportional to *RC* instead of $1/RC$. The op amp output *Vo* is multiplied by a constant x which varies linearly between 0 and 1 depending on the sensor plate position, and fed to the positive feedback input, so that $Vf = xVo$. With the positive feedback, the op amp output will be either switched to the positive rail $+Vp$ or the negative rail $-Vp$. The op amp should be a rail-to-rail output type for good accuracy, and a large value resistor or a switch used on the positive op amp input to set the DC level at that point. The op amp negative input, then, oscillates from $+xVp$ to $-xVp$ with a time constant of *RC*, and the oscillation period *T* is

$$T = 2RC \cdot \ln\left(\frac{1+x}{1-x}\right) \qquad 4.1$$

The circuit is reasonably linear for $x \ll 1$, and can be further linearized by replacing the resistor with a current source of *I* amperes; the equation then becomes linear with x

$$T = 2x \cdot \frac{CVp}{I} \qquad 4.2$$

4.2.4 Synchronous demodulator

The most flexible and accurate method of measuring capacitance is to first apply a high frequency signal in the 10 kHz – 1 MHz range through a known impedance to the capacitor under test, then amplify the signal and apply it to a synchronous demodulator. Several variations of the amplifier are available which can appropriately measure either capacitance, C, or impedance, proportional to $1/C$, to produce a linear output, and various input circuit configurations such as bridge or single-ended can be used. With high frequency excitation, electrometer-type very high input impedance amplifiers are not needed, as the capacitive impedance is much lower. Shielding and guarding are easier with a synchronous demodulator than with an oscillator circuit, and a bandpass filter is easily added to limit the noise bandwidth if needed.

The circuit in Figure 4.5 shows a full-wave demodulator, with both positive and negative half-cycles of signal contributing to the output DC level. It can linearize either capacitance- or impedance-variation sensors depending on the configuration of the amplifier.

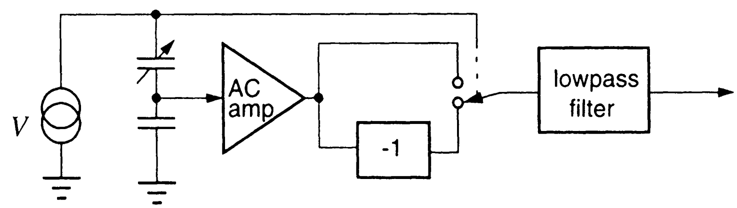

Figure 4.5 Synchronous demodulator circuit

4.3 AMPLIFIERS

With the preferred synchronous demodulator, the amplifier circuit is chosen to produce a linear output for the given sensor configuration, to allow proper guarding and shielding, and to reduce the effect of stray capacitance. Several examples are shown; see also Table 5.1.

4.3.1 High-Z amplifier

The high input impedance amplifier uses a 1× noninverting amplifier configuration as shown in Figure 4.6.

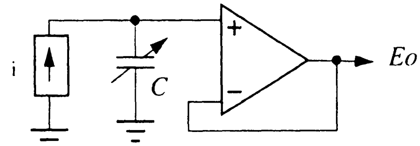

Figure 4.6 High-Z circuit

With an AC current source, this circuit produces an output proportional to the impedance of sensor C, so it produces a linear output with spacing-variation sensors. The input is usually guarded by a shield connected to the output; without this guard, a stray capacitance of only a few pF will seriously attenuate the signal. This stray capacitance is almost completely nulled by a good guard.

4.3.2 Low-Z amplifier

The low input impedance, or virtual ground amplifier (Figure 4.7), will have an input impedance inversely proportional to the gain of the amplifier times the feedback impedance. This connection has similar noise characteristics, better performance with low voltage rails because of its better common mode range, and it can be guarded by use of a grounded shield instead of a floating shield. The output is proportional to $C1 / C2$ or Z_{C2}/Z_{C1} so it can be used to linearize either area or spacing sensors by using the sensor as $C1$ or $C2$.

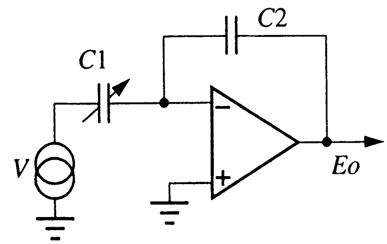

Figure 4.7 Low-Z circuit

4.3.3 Feedback amplifier

The feedback circuit of Figure 3.13 can be redrawn using summing circuits to avoid floating generators and using equivalent variable capacitors to replace sense electrodes (Figure 4.8).

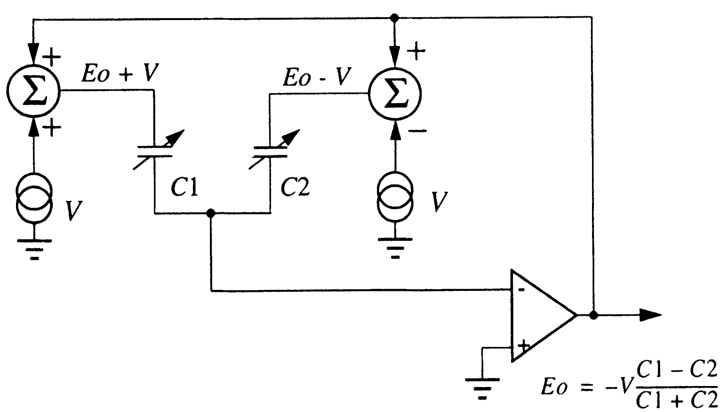

Figure 4.8 Feedback circuit

This circuit has an output proportional to $C1 - C2$ so it has a linear output with area-variation sensors. The input is guarded by ground. This amplifier is particularly good at eliminating the effects of stray capacitance: if we assume we are capacitively coupling to the sense electrodes with a small value capacitor $C3$, the circuit can be redrawn including $C3$ and stray capacitance (Figure 4.9).

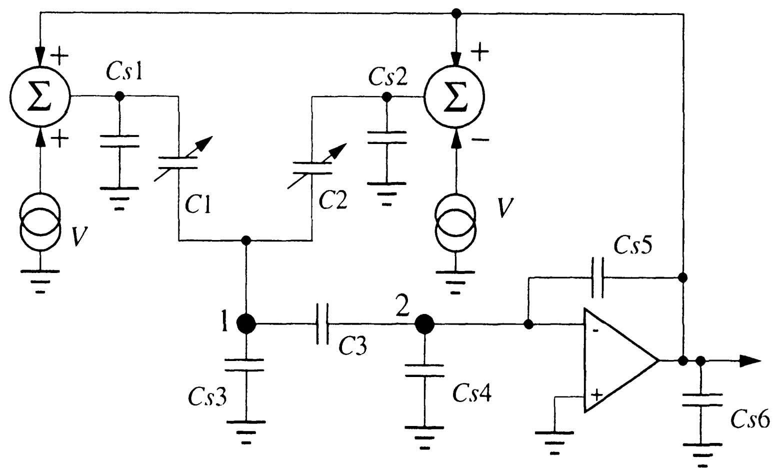

Figure 4.9 Feedback circuit with stray capacitances

$Cs1$, $Cs2$, and $Cs6$ are normally small with respect to the low output impedance of the amplifier and the summers and can be neglected. The circuit of $Cs3$, $C3$, and $Cs4$ will attenuate the signal, but it is inside the amplifier's feedback loop and the only effect the attenuation will have on the circuit is to lower the open loop gain. With a high gain, high frequency amplifier, this will not appreciably affect the closed loop gain. $Cs5$ will directly affect the gain and must be minimized, but this can be handled easily with a two stage amplifier. The feedback circuit has two important advantages over other circuits:

- Insensitivity to stray capacitance
- Guard potential is at ground

In addition, inverting amplifiers are more stable than followers, and amplifier input common mode range, which can be troublesome with followers, is less of a concern. Feedback amplifiers with synchronous demodulators have the best performance and are reccomended for high precision applications.

4.3.4 DC restoration

The circuits of Figures 4.6 – 4.9 will not work for very long, as the amplifier input bias current will cause the output to drift to a power supply rail. This problem can be handled with a high-value impedance across the amplifier input, or across the feedback capacitor for inverting amplifiers, which bleeds off the charge. Some options are:

- Very high value resistor, 100 M or more
- 1 M resistor in T configuration (see Section 10.3.5)
- High impedance FET, or two back-to-back FETs (appropriately high impedance FETs are available only on integrated circuits, not as discrete devices)
- FET switch, momentarily closed at a time when the measurement will not be affected

4.4 SINGLE-ENDED CIRCUITS

Capacitance measurement with a single-ended circuit uses a discrete capacitor, such as an accurate mica or film dielectric component, as a reference to establish the properties of the capacitor under test which is usually an air-dielectric device. One possible circuit is shown in Figure 4.6 where the value of the AC current source is the reference. Another, shown in Figure 4.7, uses a low impedance amplifier and a fixed capacitor as a reference. The output of this circuit is the excitation voltage *V* times the ratio of the impedance of *C*2 to *C*1. With *C*1 as the variable capacitance, as shown, the output is linear with an area-variation sensor. With *C*2 as the variable and *C*1 as a reference, the output is linear with a spacing-variation sensor. The use of a capacitor as the feedback element rather than the usual resistor improves phase shift and noise performance.

4.5 BRIDGE CIRCUITS

4.5.1 Wheatstone bridge

The standard Wheatstone bridge circuit, shown in Figure 4.10, is often used for low noise instrumentation.

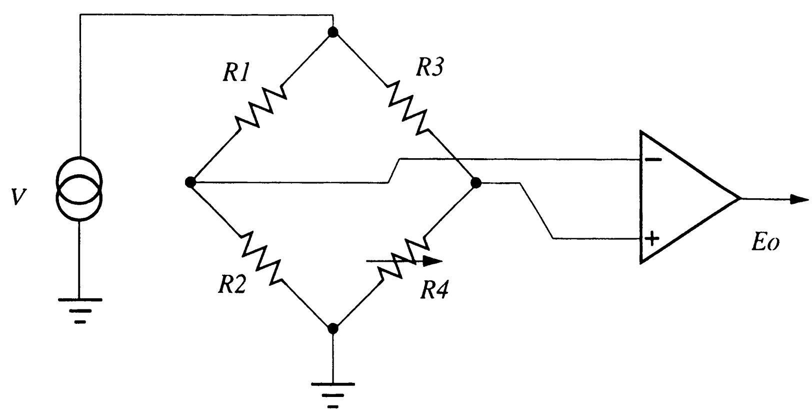

Figure 4.10 Resistor bridge

If the bridge is balanced, $R1/R2 = R3/R4$ and the output is zero. Then large amplifier gain can be used to amplify small differences in one or both legs of the bridge. The advantage of the bridge is that two identical components can be used for $R1$–$R2$ and $R3$–$R4$ so that thermal drifts in these components tend to track and do not affect the balance.

4.5.2 Capacitance bridge

Several variations of capacitance bridges are possible. With a balanced drive replacing the differential amplifier and with capacitances replacing resistors, we have a useful capacitance bridge (Figure 4.11).

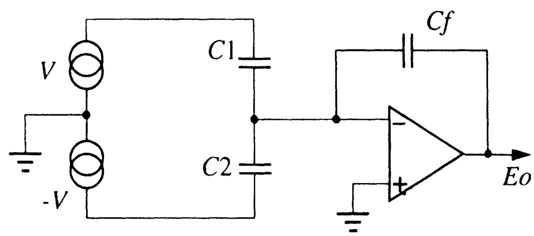

Figure 4.11 Capacitance bridge

$$Vout = -V \cdot \frac{C1 - C2}{Cf} \qquad 4.3$$

The amplifier gain should be high for this equation to be accurate. Any stray capacitance to ground at the amplifier input does not affect the output voltage, but does cut down amplifier available gain at high frequencies, so it should be minimized. $C1$ and $C2$ should be of identical construction for the bridge balance to be stable. The gage factor, or the output voltage divided by input displacement, is a function of $Cf/C1 \parallel C2$, but the null accuracy is a function only of $C1 / C2$. The ratio $C1 / C2$ can be made stable if the capacitors have identical construction.

This type of bridge relies on a balanced drive rather than on good common-mode rejection of the amplifier, and it is generally the preferred circuit. A center-tapped transformer, an inverting amplifier, or a CMOS logic inverter can be used to supply the balanced drive.

4.5.3 Bridge vs. single-ended

The bridge circuit is essential for sensitive circuits which are amplifying a small difference of capacitance, as in Chapter 5, as it is much easier to stabilize a ratio of similar components than a ratio of dissimilar components. The ratiometric output is insensitive to power supply or other circuit variations; with a single-ended circuit these changes would cause a DC error in the output. Also, for circuits with high linear range and low amplifier gain, as in Chapter 7, the bridge is preferred as it can be designed to be insensitive to many circuit parameters. Some additional design details for integrated devices can be found in Kung et al. [1988, 1992].

4.6 EXCITATION

4.6.1 Sine wave

Sine wave excitation is useful for systems needing high frequency (above 1 MHz) carriers, for example, systems which must measure very low capacitance. Sine waves are also preferred for circuits which need high accuracy. Compared to square wave excitation, amplifier slew rate problems are lessened by a factor of 10 and lower frequency amplifiers can be used. Sine wave excitation is essential for high gain bridge circuits, as a good null is more easily achieved without the presence of harmonic energy. But accurate sine wave generation is difficult, and sine wave demodulation uses analog multipliers and other more expensive and less accurate parts compared to square wave circuits.

4.6.2 Square wave

Square wave drives are available virtually free in any system using CMOS logic gates, as the CMOS rail-to-rail output voltage can be quite accurate. Square wave demodulation is generally done with CMOS switches and op amps, and it is easy to integrate for single-chip systems. It is also better suited to low gain systems, such as motion detectors with wide linear operation. For more information, see Section 10.6. For circuits using square wave modulation the designer must be careful to avoid the unstable and nonlinear effects which are produced when an amplifier runs into slew rate limiting, and amplifier bandwidth must be a factor of ten higher than for sine wave circuits to retain good waveshape.

For both sine and square wave amplifiers, some attention needs to be paid to the amplifier phase shift characteristics. Uncompensated phase shift will typically reduce the demodulator gain.

4.7 FILTERING

4.7.1 Lowpass filter

The circuit of Figure 4.5 includes a lowpass filter to remove demodulation components. The need for this filter can be seen from the spectrum of the signals, shown in Figure 4.12 with sine wave excitation at 20 kHz and motion in the DC-5 kHz frequency range.

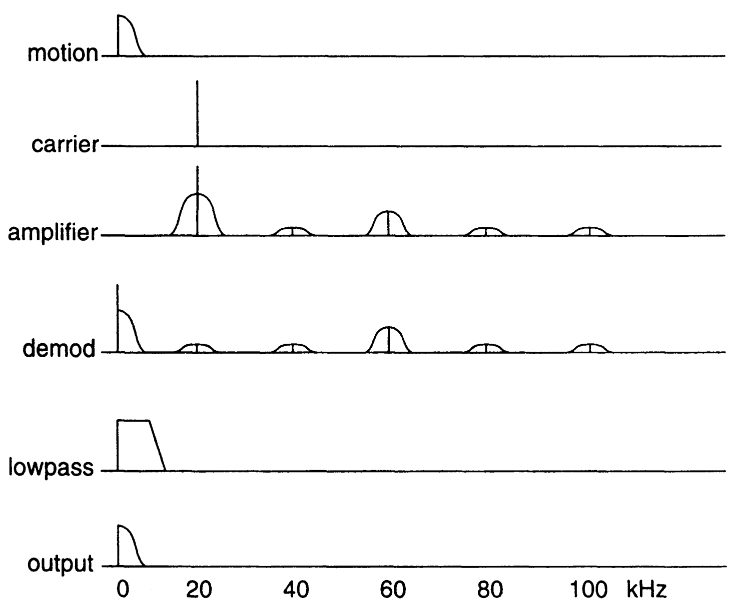

Figure 4.12 Signal spectra

The carrier is assumed to be a sine wave. With changing plate spacing, the amplifier output will be, as shown above, a double sideband replica of the motion signal centered on the 20 kHz carrier, with sidebands produced by the plate spacing variation. Harmonics of 20 kHz are generated by nonlinearities in the amplifier or the carrier or in the multiplication process; often the ±1 circuit shown is used to replace a linear multiplier and the lowpass input is rich with harmonics. With the ±1 circuit, the signal input is linear and the carrier input is hard limited, but the signal linearity and stability can be better than with a conventional two-port linear multiplier.

The demodulator output shows the regenerated motion waveform near DC along with carrier-frequency feedthrough caused by demodulator imbalance, and the upper harmonics generated by the amplifier distortion.

The lowpass filter removes the offending harmonic components and high frequency noise. Also, interfering noise signals at a frequency ω2 which couple to a sensor excited with ω1 will produce signals of new frequencies at ±(ω2–ω1), but if ω2–ω1 is larger than the lowpass bandwidth and the amplifier does not saturate, the lowpass filter output will reject the interference.

4.7.2 Bandpass filter

A further improvement in rejecting noise can be made by inserting a bandpass filter before the amplifier. The amplifier, especially if it has high gain, can generate excessive spurious frequencies with out-of-band noise or slew rate limiting, or in extreme cases it can saturate. These effects are eased with a bandpass filter, centered on the carrier and sized as narrow as possible without affecting the signal bandwidth, preceding the amplifier. If a sharp-cutoff bandpass filter is not used, single-pole highpass and lowpass filters will help.