17

Proximity detector

A single-plate capacitive proximity detector has been designed and breadboarded to develop the circuit for a low cost detector. The detector has good detection range and small size, and provides a threshold output as well as an analog output which is proportional to changes in capacitance to local objects.

17.1 BLOCK DIAGRAM

Figure 17.1 shows a block diagram of the circuit. The circuit uses two sense electrodes, *A* and *B*, with the capacitance *Cs* varying with target position. A bridge circuit is used so that component variation will cancel as much as possible; *Cr* is the reference plate capacitance. When *Cr* = *Cs* the bridge is balanced and the amplifier output is near zero. Temperature and humidity variation effects are canceled if *Cr*'s construction closely matches *Cs*'s, so both are air-spaced construction. The 1.5 × 2.5 cm electrodes have a capacitance to ground of approximately 1 pF.

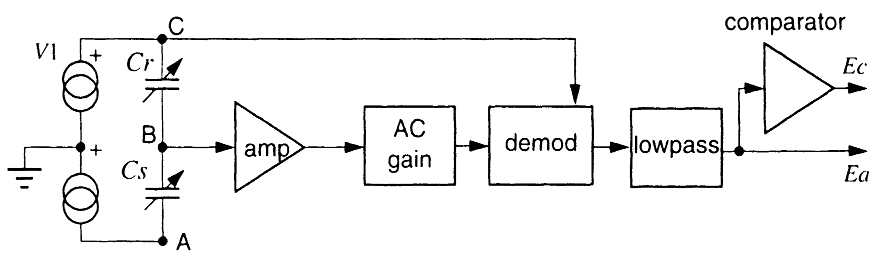

Figure 17.1 Block diagram

17.1.1 Lossy capacitance

The block diagram shows *Cs* and *Cr* as pure capacitors. More generally, *Cs* and *Cr* should be drawn as lossy capacitors (Figure 17.2).

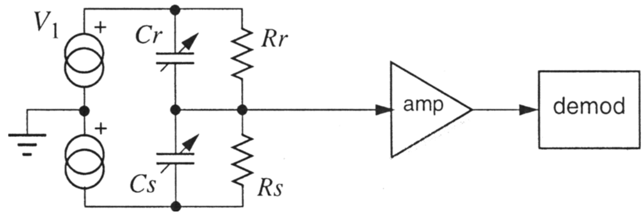

Figure 17.2 Lossy capacitors

The shunt resistance represents the dielectric loss tangent. *Rr* and *Rs* are insignificant for air-dielectric capacitors, but if the sensor environment includes lossy dielectrics like wood, they may become important. If *Rr* and *Rs* are not equal, and assuming sine wave excitation, the effect is to add a quadrature sine wave at 90° to the bridge output. This sine wave does not affect the demodulated output at low levels, since the demodulator circuit and lowpass combination rejects 90° components, but for maximum sensitivity a high gain AC amplifier is needed and the 90° component may saturate the amplifier. This is cured by adding a trim resistor *Rq* to the bridge to correct the *Rr* – *Rs* difference, or by adding a second demodulator sensitive to 90° (and insensitive to 0°) along with a feedback circuit to automatically null the 90° output component.

17.1.2 Construction

With this construction (Figure 17.3), *Cs*, the external capacitance which will be modulated by target position, is balanced in a bridge circuit by *Cr*.

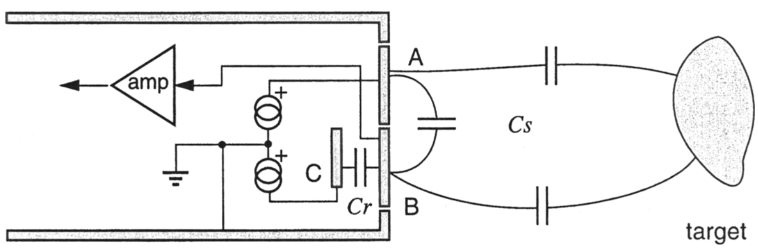

Figure 17.3 Cross section

Monopole response

The face view of the device is shown in Figure 17.4. With this electrode pattern, the response shape will be a nearly uniform sphere, with a small bias toward targets positioned in the direction of the face.

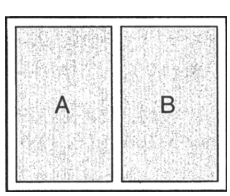

Figure 17.4 Face view, monopole response

Dipole response

If the shielded reference electrode C is instead brought up to the face, a dipole response results (Figure 17.5). Now, Cr and Cs will be equal for any target position which bisects the face in the vertical plane, producing a dipole response for horizontal motion (Figure 17.6).

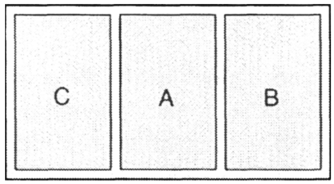

Figure 17.5 Face view, dipole response

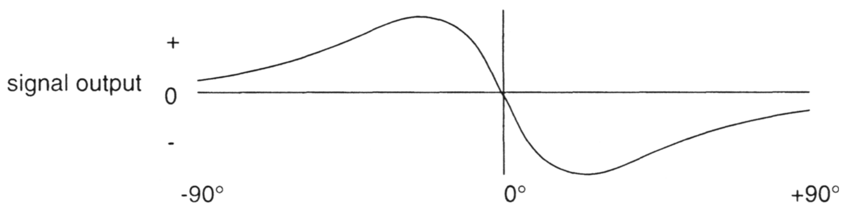

Figure 17.6 Dipole response

This presents the interesting option of adding two more electrodes in the vertical plane which can produce a similar dipole response for vertical motion. Then, with proper switching of electrodes and appropriate amplitude control of the excitation voltage, a target's approximate angular extent and the centroid of its vertical and horizontal position can be determined.

17.1.3 Carrier

The simplest choice of carrier is a square wave in the 10 kHz – 1 MHz range. As the carrier frequency is increased, lower reactance of the sensor capacitance helps minimize amplifier input current noise contribution. A second advantage of increased carrier frequency is that the injected building ground noise near the carrier frequency will decrease, as building ground noise is generally 60 Hz and harmonics and harmonic energy decrease with increasing frequency. Additional discussion on building ground noise can be found in Section 6.3.1.

Carrier frequency cannot be arbitrarily increased to avoid power frequency components, however, as higher frequency, high current amplifiers would be needed. Another factor to be considered in carrier frequency choice is the increasing use of fluorescent lamps which are powered by switching supplies in the 20 kHz – 200 kHz range. For example, the circuit shown in Figure 17.7 uses a 100 kHz carrier, well above power frequencies, but not difficult for conventional operational amplifiers to handle.

17.2 SCHEMATIC

Figure 17.7 shows a schematic of the proximity detector.

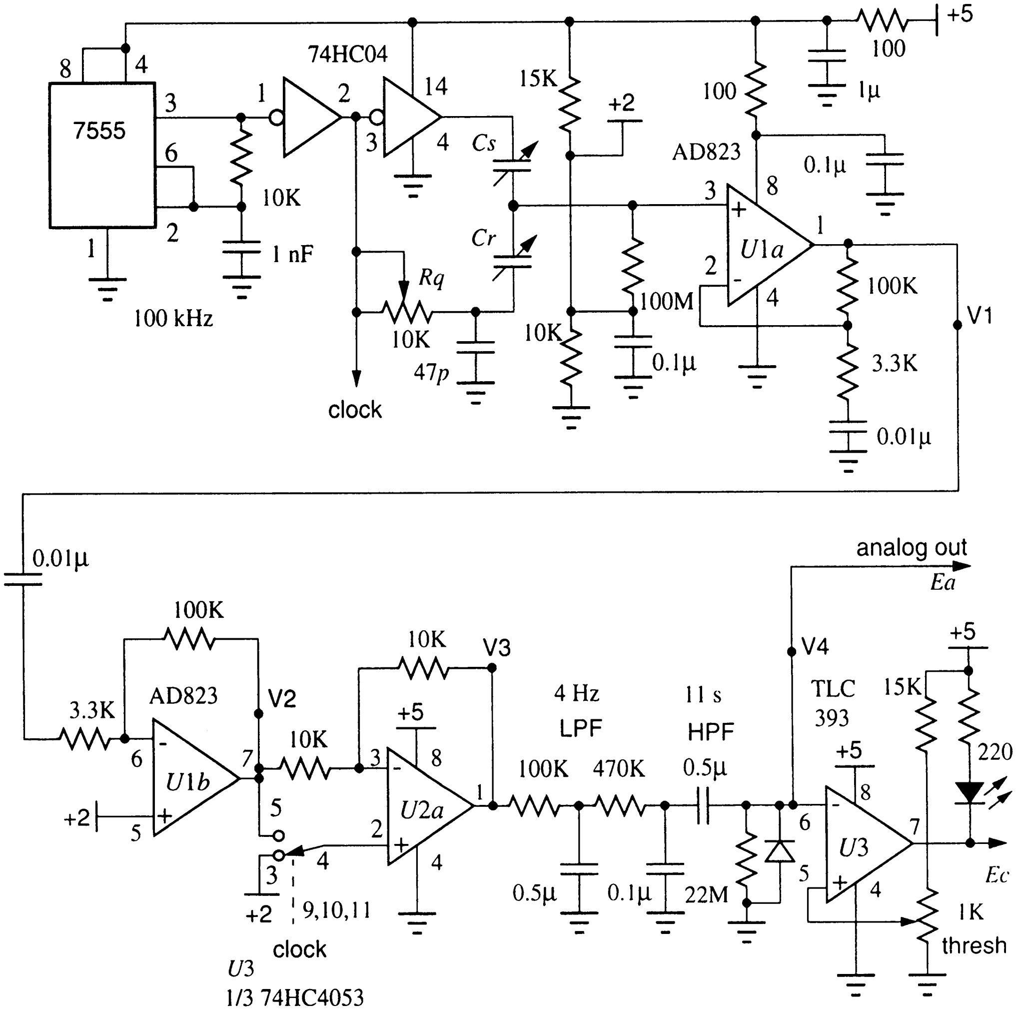

Figure 17.7 Proximity detector schematic

17.2.1 Excitation

A 7555 timer chip generates a square wave clock which is buffered by CMOS inverters to produce a 0° – 180° drive waveform which pulls accurately to the voltage rail. One output feeds *Cs*, the capacitance to be measured, and the other feeds balancing components *Rq* and *Cr* which are adjusted for null.

17.2.2 Input amplifier

For a sensor capacitance which is much less than the amplifier common-mode input capacitance, performance is improved by adding the *Vee* bootstrap circuit as in Figure 1.9

and increasing or bootstrapping the input resistor. For large, high capacity electrodes, the simpler circuit shown will be adequate with the input time constant contributing a useful low-cut filter. The amplifier input capacitance with careful construction will be on the order of 10 pF, which will cause about 20 dB gain loss with the 1 pF sensor capacity. A FET-input integrated amplifier with low current noise is shown, but 2–5× improvement in amplifier noise can be had by using a discrete FET preamp. The amplifier f_T, 15 MHz, limits high frequency response to about 0.5 MHz at 30× gain.

17.2.3 Demodulator

The full-wave demodulation shown in the schematic is preferable to half-wave demodulation as it rejects any remnant low frequency noise components which get by the input highpass circuits.

U2a is connected as an inverter ($V3 = -V2$) with the switch as shown and as a follower ($V3 = V2$) with the switch in the opposite position. This produces a synchronous full-wave demodulator. If the phase shift through the signal path is close to zero, the voltage at *V3* is a chopped waveform with an average DC level proportional to the input AC level with polarity controlled by input phase.

Lowpass, highpass filter

The crude 4 Hz lowpass filter removes the chopping components and also any high frequency in-band noise components, and the 11 s highpass filter rejects very slow DC variations which may be caused by component drift. High value resistors are used to keep filter capacitor size down, so the usual high impedance printed-circuit-board construction techniques should be used. The highpass filter may not be needed for less sensitive applications.

Comparator

A comparator compares the DC output to a preset threshold, and a LED diode shows when the threshold has been exceeded.

Null

For maximum sensitivity, the proximity detector is adjusted so that a null amplifier output is achieved with no object in the field. For a good null:

- Balance the fundamental output by adjusting *Cr*
- Balance the quadrature output by adjusting *Rt*
- Repeat as needed

If the amplifiers had infinite headroom, the goodness of null would not be too critical as the synchronous demodulator responds only to the in-phase fundamental component. With finite headroom, however, other null components such as quadrature and harmonics may be large enough to drive the final amplifier to clipping and compromise the demodulation.

With square wave excitation and wideband amplifiers, achieving an accurate null is more difficult, as all frequencies within the passband must be equalized in amplitude and phase. With either a sine wave drive or with a narrow bandpass filter, only one frequency need be considered. The circuit above uses a square wave drive, but each amplifier acts as a bandpass filter in the 5–500 kHz range with the upper frequency cutoff due to the amplifiers f_T. On the bench, the null at the amplifier output suffered from a quadrature component due to the small 20 ns delay between the 0 and 180° 5 V square wave drives, which was compensated by adjusting *Rq*.

Gage factor

The gain, or gage factor, is not too critical for a proximity detector. The circuit above is not accurate, as the phase shift through the amplifiers causes an uncompensated gain error in the demodulator output. A second similar amplifier circuit could be used to feed the carrier input to the demodulator to equalize delays, but this approach is not much better because of the extreme sensitivity of the phase shift to the filter center frequency and *R* and *C* values. Another approach would be to use a bandpass filter which has less phase shift, such as an FIR (finite impulse response) type, or a filter with more stable phase shift such as a crystal or ceramic resonator.

17.2.4 Power supply

The power supply can be any standard 5 V supply, but should be low noise, less than 10 mV. Note that most voltage regulators have high noise (over 100 $\mu V/\sqrt{Hz}$) and low output impedance measured in milliohms, so a simple capacitor does not do a good job at removing regulator output noise. A 0.005 Ω regulator output impedance with the normal 10 µF filter capacitor produces a 50 ns time constant, nowhere near long enough for useful filtering, and the current to slew a 10µ capacitor 5 mV at 50 kHz carrier frequency is only about 10 mA. Adding the 100 Ω resistor as shown reduces power supply noise to a few $\mu V/\sqrt{Hz}$.

17.2.5 Noise and stability

Although the amplifier used was not a particularly low noise device, the circuit's noise bandwidth is very low and noise was not a limiting factor. The amplifier output noise at *V*3 is equal to the input noise (25 $nV/\sqrt{Hz}$ in a 500 kHz bandwidth) times the amplifier gain of 1000, or 18 mV rms. Power line noise at 60 Hz was much larger than the amplifier noise at the input stage, but it was effectively removed by the bandpass amplifiers.

The 18 mV noise component at the amplifier output is attenuated by the lowpass filter following the demodulator. The lowpass corner frequency is about 5 Hz, so the attenuation is the ratio of 500 kHz to 5 Hz or 100,000:1 for a filtered noise of 0.18 µV.

Stability was a much more serious problem. The demodulated DC output at the lowpass filter was affected by temperature, small movements of electrical components, and even air currents. At null, the amplifier stability is not a factor, so the presumed cause of the instability is the rather unstable electrode construction of glued-up printed circuit board stock and the unattached hookup wire electrode connections. The lowpass/highpass filter configuration rejects slow DC variations, however. If a true DC response were necessary, more stable mechanics and a better solution to the bandpass phase response problem would be needed.

Layout

The general layout shown in Figure 3.9 was used. Standard AM radio IF construction rules were followed, with the amplifiers in a linear array so the output could be as far as possible from the input, but with maximum amplifier gain, capacitive feedback from amplifier output to input caused oscillation at about 20 kHz. A copper ground plane fixed the problem.

17.3 PERFORMANCE

The limiting detection distance of the experimenter's hand was about 250 cm. As this was determined by stability rather than noise, and as noise was at least a factor of 1000 lower than limit signal, the range with careful construction could be extended by 60 dB. As the response of 2 cm electrodes (close to our 1.5 × 2.5 cm electrodes) is seen in Figure 2.39 to decrease by 20 dB for each 18 mm of separation, a 60 dB sensitivity improvement results in a detection range increase to only 304 cm.

An improved version would concentrate on improved mechanical stability. Also, the tradeoff between increasing predemodulation gain (better DC stability, but added risk of saturating the amplifier with quadrature signals or added noise) and increasing postdemodulation gain should be carefully evaluated for each application.