EXHIBIT 7

THE ART OF ELECTRONICS

Second Edition

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Winfield Hill ROWLAND INSTITUTE FOR SCIENCE, CAMBRIDGE, MASSACHUSETTS

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Differential amplifiers as comparators

Because of its high gain and stable characteristics, the differential amplifier is the main building block of the *comparator*, a circuit that tells which of two inputs is larger. They are used for all sorts of applications: switching on lights and heaters, generating square waves from triangles, detecting when a level in a circuit exceeds some particular threshold, class D amplifiers and pulse-code modulation, switching power supplies, etc. The basic idea is to connect a differential amplifier so that it turns a transistor switch on or off, depending on the relative levels of the input signals. The linear region of amplification is ignored, with one or the other of the two input transistors cut off at any time. A typical hookup is illustrated in the next section by a temperature-controlling circuit that uses a resistive temperature sensor (thermistor).

2.19 Capacitance and Miller effect

In our discussion so far we have used what amounts to a dc, or low-frequency, model of the transistor. Our simple current amplifier model and the more sophisticated Ebers-Moll transconductance model both deal with voltages, currents, and resistances seen at the various terminals. With these models alone we have managed to go quite far, and in fact these simple models contain nearly everything you will ever need to know to design transistor circuits. However, one important aspect that has serious impact on high-speed and high-frequency circuits has been neglected: the existence of capacitance in the external circuit and in the transistor junctions themselves. Indeed, at high frequencies the effects of capacitance often dominate circuit behavior; at 100 MHz a typical junction capacitance of 5pF has an impedance of 320 ohms!

We will deal with this important subject in detail in Chapter 13. At this point we would merely like to state the problem, illustrate some of its circuit incarnations, and suggest some methods of circumventing the problem. It would be a mistake to leave this chapter without realizing the nature of this problem. In the course of this brief discussion we will encounter the famous *Miller effect* and the use of configurations such as the cascode to overcome it.

Junction and circuit capacitance

Capacitance limits the speed at which the voltages within a circuit can swing ("slew rate"), owing to finite driving impedance or current. When a capacitance is driven by a finite source resistance, you see *RC* exponential charging behavior, whereas a capacitance driven by a current source leads to slew-rate-limited waveforms (ramps). As general guidance, reducing the source impedances and load capacitances and increasing the drive currents within a circuit will speed things up. However, there are some subtleties connected with feedback capacitance and input capacitance. Let's take a brief look.

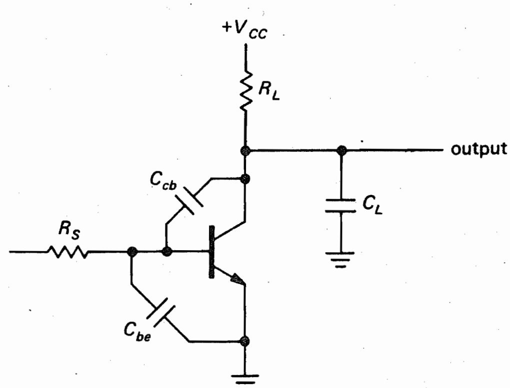


Figure 2.73. Junction and load capacitances in a transistor amplifier.

The circuit in Figure 2.73 illustrates most of the problems of junction capacitance. The output capacitance forms a time constant with the output resistance R_L (R_L includes both the collector and load resistances, and C_L includes both junction and load capacitances), giving a rolloff starting at some frequency $f = 1/2\pi R_L C_L$. The same is true for the input capacitance in combination with the source impedance R_S.

Miller effect

C_{cb} is another matter. The amplifier has some overall voltage gain G_V, so a small voltage wiggle at the input results in a wiggle G_V times larger (and inverted) at the collector. This means that the signal source sees a current through C_{cb} that is $G_V + 1$ times as large as if C_{cb} were connected from base to ground; i.e., for the purpose of input rolloff frequency calculations, the feedback capacitance behaves like a capacitor of value $C_{cb}(G_V + 1)$ from input to ground. This effective increase of C_{cb} is known as the Miller effect. It often dominates the rolloff characteristics of amplifiers, since a typical feedback capacitance of 4pF can look like several hundred picofarads to ground.

There are several methods available to beat the Miller effect. It is absent altogether in a grounded base stage. You can decrease the source impedance driving a grounded emitter stage by using an emitter follower. Figure 2.74 shows two other possibilities. The differential amplifier circuit (with no collector resistor in Q_1) has no Miller effect; you can think of it as an emitter follower driving a grounded base amplifier. The second circuit is the famous cascode configuration. Q_1 is a grounded emitter amplifier with R_L as its collector resistor. Q_2 is interposed in the collector path to prevent Q_1's collector from swinging (thereby eliminating the Miller effect) while passing the collector current through to the load resistor unchanged. V_+ is a fixed bias voltage, usually set a few volts above Q_1's emitter voltage to pin Q_1's collector and keep it in the active region. This fragment is incomplete as shown; you could either include a bypassed emitter resistor and base divider for biasing (as we did earlier in the chapter) or include it within an overall loop with feedback at dc. V_+ might be provided from a divider or zener, with bypassing to keep it stiff at signal frequencies.

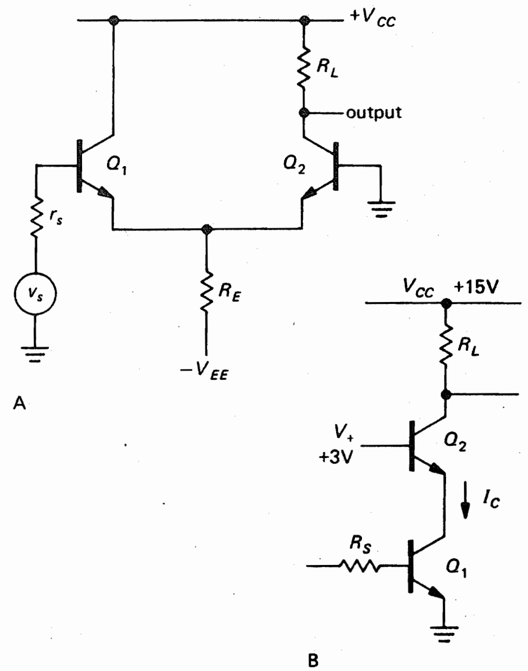


Figure 2.74. Two circuit configurations that avoid Miller effect. Circuit B is the cascode.

EXERCISE 2.14

Explain in detail why there is no Miller effect in either transistor in the preceding differential amplifier and cascode circuits.

Capacitive effects can be somewhat more complicated than this brief introduction might indicate. In particular: (a) The rolloffs due to feedback and output capacitances are not entirely independent; in the terminology of the trade there is pole splitting, an effect we will explain in the next chapter. (b) The input capacitance still has an effect, even with a stiff input signal source. In particular, current that flows through C_{be} is not amplified by the transistor. This base current "robbing" by the input capacitance causes the transistor's small-signal current gain h_{fe} to drop at high frequencies, eventually reaching unity at a frequency known as f_T. (c) To complicate matters, the junction capacitances depend on voltage. C_{be} changes so rapidly with base current that it is not even specified on transistor data sheets; f_T is given instead. (d) When a transistor is operated as a switch, effects associated with charge stored in the base region of a saturated transistor cause an additional loss of speed. We will take up these and other topics having to do with high-speed circuits in Chapter 13.

2.20 Field-effect transistors

In this chapter we have dealt exclusively with *bipolar junction transistors* (BJTs), characterized by the Ebers-Moll equation. BJTs were the original transistors, and they still dominate analog circuit design. However, it would be a mistake to continue without a few words of explanation about the other kind of transistor, the field-effect transistor (FET), which we will take up in detail in the next chapter.

The FET behaves in many ways like an ordinary bipolar transistor. It is a 3-terminal amplifying device, available in both polarities, with a terminal (the *gate*) that controls the current flow between the other two terminals (*source* and *drain*). It has a unique property, though: The gate draws no current, except for leakage. This means that extremely high input impedances are possible, limited only by capacitance and leakage effects. With FETs you don't have to worry about providing substantial base current, as was necessary with the BJT circuit design of this chapter. Input currents measured in *pico*amperes are commonplace. Yet the FET is a rugged and capable device, with voltage and current ratings comparable to those of bipolar transistors.

Most of the available devices fabricated with transistors (matched pairs, differential and operational amplifiers, comparators, high-current switches and amplifiers, radiofrequency amplifiers, and digital logic) are also available with FET construction, often with superior performance. Furthermore, microprocessors and memory (and other large-scale digital electronics) are built almost exclusively with FETs. Finally, the area of micropower design is dominated by FET circuits.

FETs are so important in electronic design that we will devote the next chapter to them, before treating operational amplifiers and feedback in Chapter 4. We urge the reader to be patient with us as we lay the groundwork in these first three difficult chapters; that patience will be rewarded many times over in the succeeding chapters, as we explore the enjoyable topics of circuit design with operational amplifiers and digital integrated circuits.

SOME TYPICAL TRANSISTOR CIRCUITS

To illustrate some of the ideas of this chapter, let's look at a few examples of circuits with transistors. The range of circuits we can cover is necessarily limited, since real-world circuits often use negative feedback, a subject we will cover in the next chapter.

2.21 Regulated power supply

Figure 2.75 shows a very common configuration. R_1 normally holds Q_1 on; when the output reaches 10 volts, Q_2 goes into conduction (base at 5V), preventing further rise of output voltage by shunting base current from Q_1's base. The supply can be made adjustable by replacing R_2 and R_3

characteristic), or it can be amplitude-dependent, producing a nonlinear amplifier (a popular example is a logarithmic amplifier, built with feedback that exploits the logarithmic V_{BE} versus I_C of a diode or transistor). It can be arranged to produce a current source (near-infinite output impedance) or a voltage source (near-zero output impedance), and it can be connected to generate very high or very low input impedance. Speaking in general terms, the property that is sampled to produce feedback is the property that is improved. Thus, if you feed back a signal proportional to the output current, you will generate a good current source.

Feedback can also be *positive*; that's how you make an oscillator, for instance. As much fun as that may sound, it simply isn't as important as negative feedback. More often it's a nuisance, since a negative-feedback circuit may have large enough phase shifts at some high frequency to produce positive feedback and oscillations. It is surprisingly easy to have this happen, and the prevention of unwanted oscillations is the object of what is called *compensation*, a subject we will treat briefly at the end of the chapter.

Having made these general comments, we will now look at a few feedback examples with operational amplifiers.

4.02 Operational amplifiers

Most of our work with feedback will involve operational amplifiers, very high gain dc-coupled differential amplifiers with single-ended outputs. You can think of the classic long-tailed pair (Section 2.18) with its two inputs and single output as a prototype, although real op-amps have much higher gain (typically 10^5 to 10^6) and lower output impedance and allow the output to swing through most of the supply range (you usually use a split supply, most often ± 15V). Operational amplifiers are now available in literally hundreds of types, with the universal symbol shown in Figure 4.1, where the (+) and (−) inputs do as expected: The output goes positive when the noninverting input (+) goes more positive than the inverting input (−), and vice versa. The (+) and (−) symbols don't mean that you have to keep one positive with respect to the other, or anything like that; they just tell you the relative phase of the output (which is important to keep negative feedback negative). Using the words "noninverting" and "inverting," rather than "plus" and "minus," will help avoid confusion. Power-supply connections are frequently not displayed, and there is no ground terminal. Operational amplifiers have enormous voltage gain, and they are *never* (well, hardly ever) used without feedback. Think of an op-amp as fodder for feedback. The open-loop gain is so high that for any reasonable closed-loop gain, the characteristics depend only on the feedback network. Of course, at some level of scrutiny this generalization must fail. We will start with a naive view of op-amp behavior and fill in some of the finer points later, when we need to.

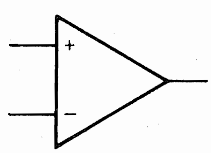


Figure 4.1

There are literally hundreds of different op-amps available, offering various performance trade-offs that we will explain later (look ahead to Table 4.1 if you want to be overwhelmed by what's available). A very good all-around performer is the popular LF411 ("411" for short), originally introduced by National Semiconductor. Like all op-amps, it is a wee beastie packaged in the so-called mini-DIP (dual in-line package), and it looks strange things may happen. (c) Some comparators permit only limited differential input swings, as little as ±5 volts in some cases. Check the specs! See Table 9.3 and the discussion in Section 9.07 for the properties of some popular comparators.

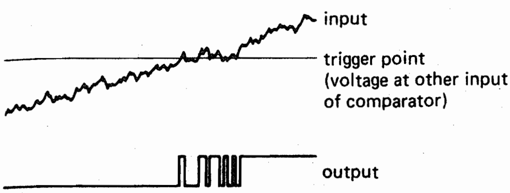


Figure 4.61

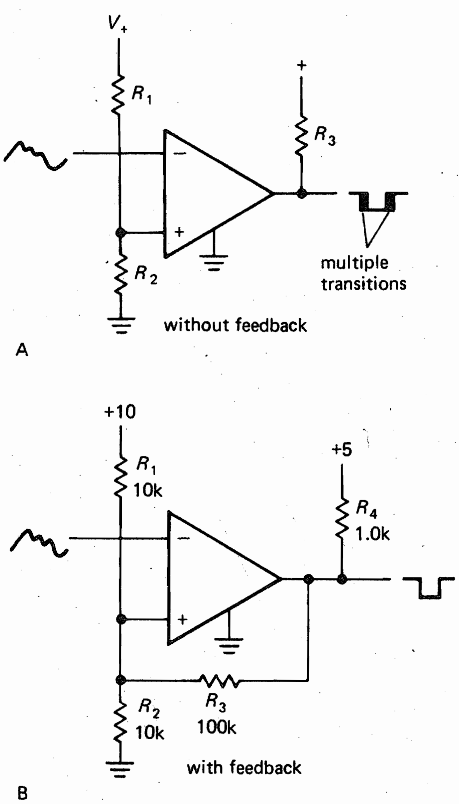


Figure 4.62

4.24 Schmitt trigger

The simple comparator circuit in Figure 4.60 has two disadvantages. For a very slowly varying input, the output swing can be rather slow. Worse still, if the input is noisy, the output may make several transitions as the input passes through the trigger point (Fig. 4.61). Both these problems can be remedied by the use of *positive* feedback (Fig. 4.62). The effect of R_3 is to make the circuit have two thresholds, depending on the output state. In the example shown, the threshold when the output is at ground (input high) is 4.76 volts, whereas the threshold with the output at +5 volts is 5.0 volts. A noisy input is less likely to produce multiple triggering (Fig. 4.63). Furthermore, the positive feedback ensures a rapid output transition, regardless of the speed of the input waveform. (A small "speedup" capacitor of 10–100pF is often connected across R_3 to enhance switching speed still further.) This configuration is known as a Schmitt trigger. (If an op-amp were used, the pullup would be omitted.)

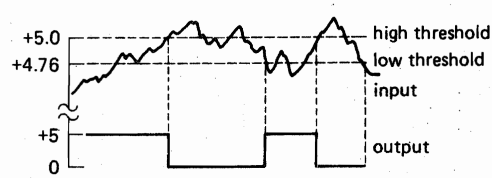


Figure 4.63

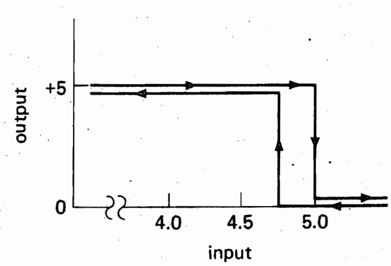
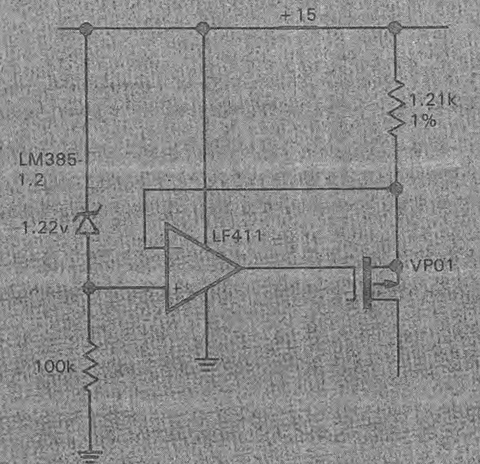


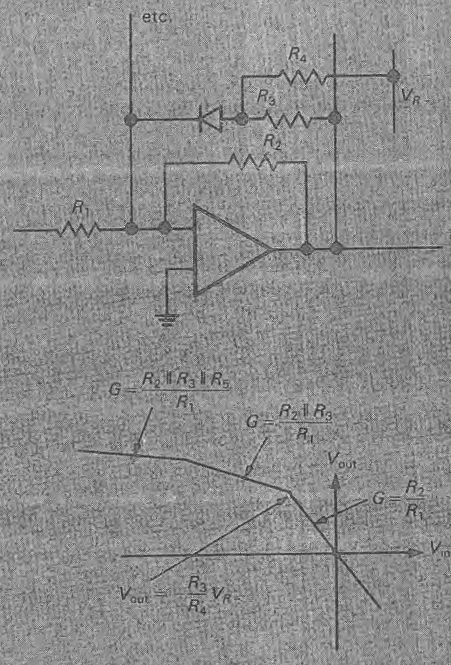
Figure 4.64

The output depends both on the input voltage and on its recent history, an effect called *hysteresis*. This can be illustrated with a diagram of output versus input, as in Figure 4.64. The design procedure

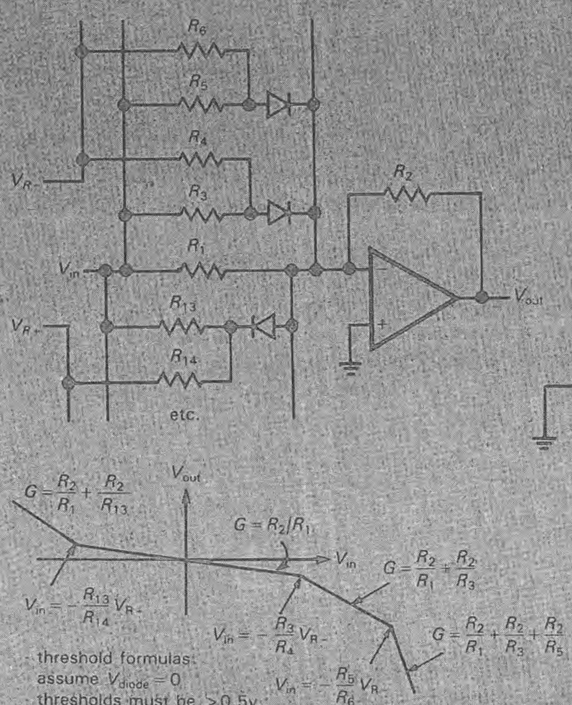


current source using PMOS FET

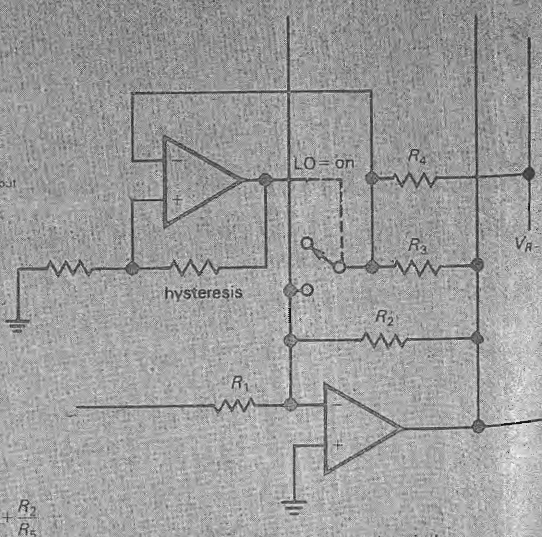
A. precision current source



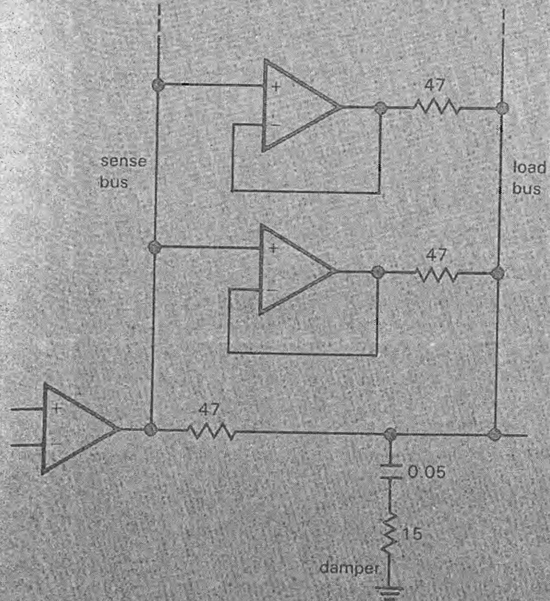
C. as in B, but gain decreases for outputs above threshold



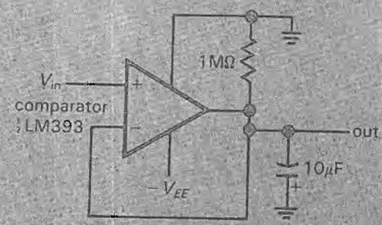
B. inverting piecewise-linear curve amplifier
G increases for inputs above threshold



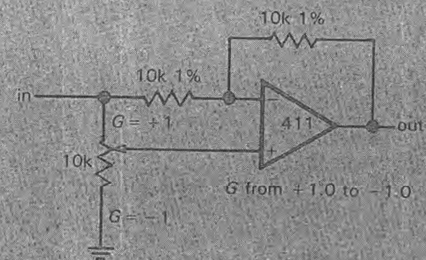
D. as in C, but with comparator and switch acting as a "perfect" diode ($V_D = 0$)



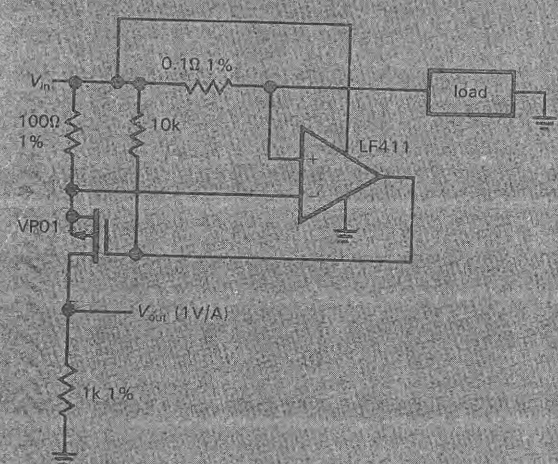
E. higher output current from additional op-amp sections; watch for excess heating



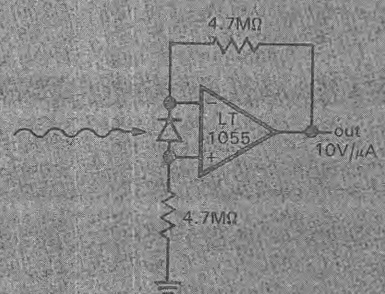
H. negative peak detector



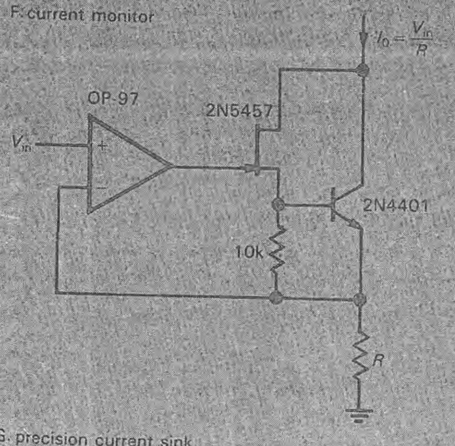
I. continuously variable gain



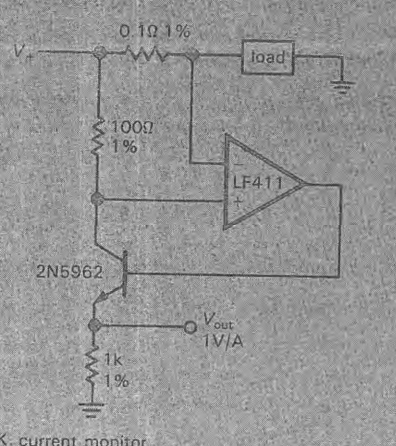
F. current monitor



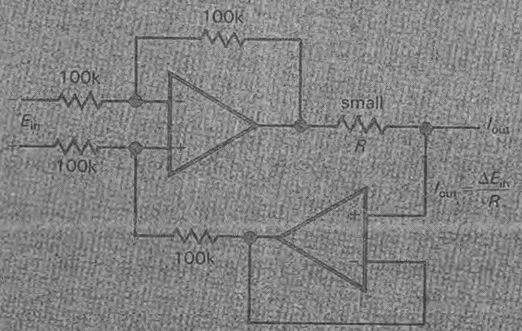
J. photodiode amplifier



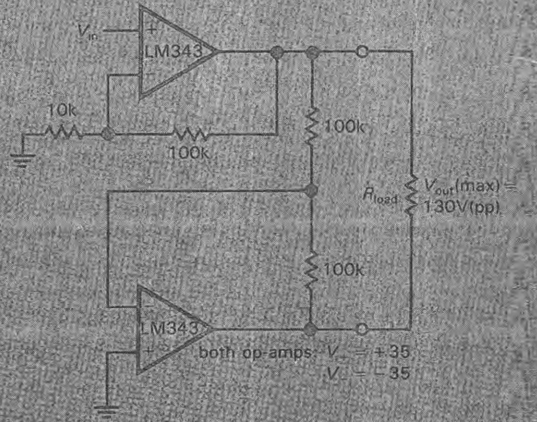
G. precision current sink



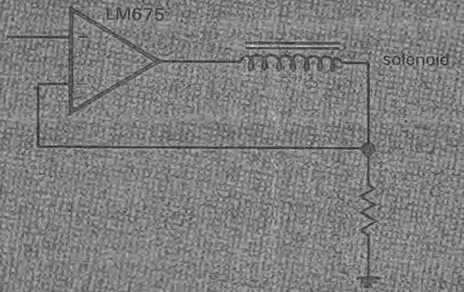
K. current monitor



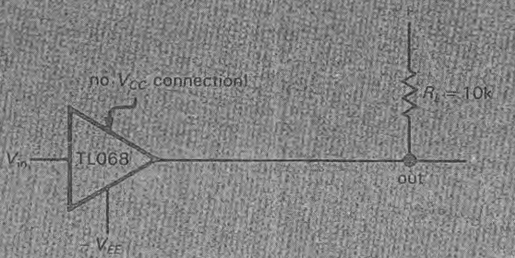
L. precision bipolarity current source



P. high-voltage (bridge) drive to floating load (gain = 22)

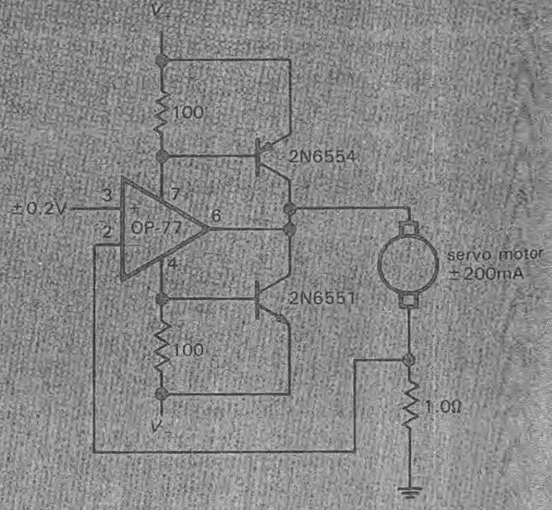


M. active solenoid

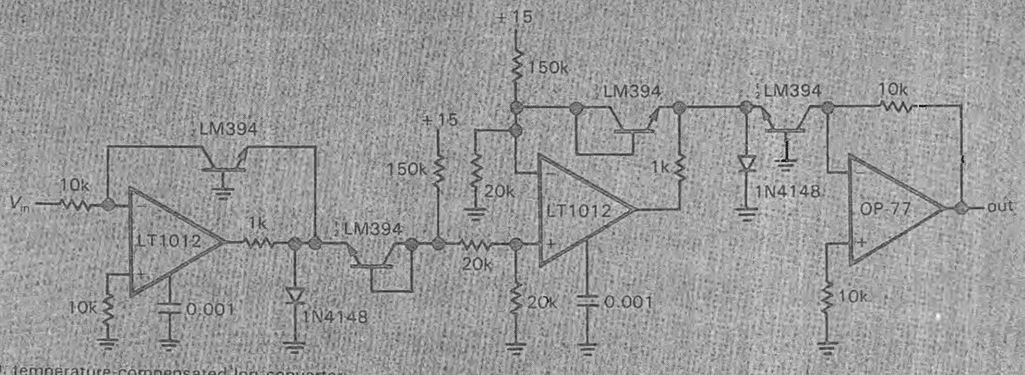


$V_{EE} + 3V < V_{in} < 30V$

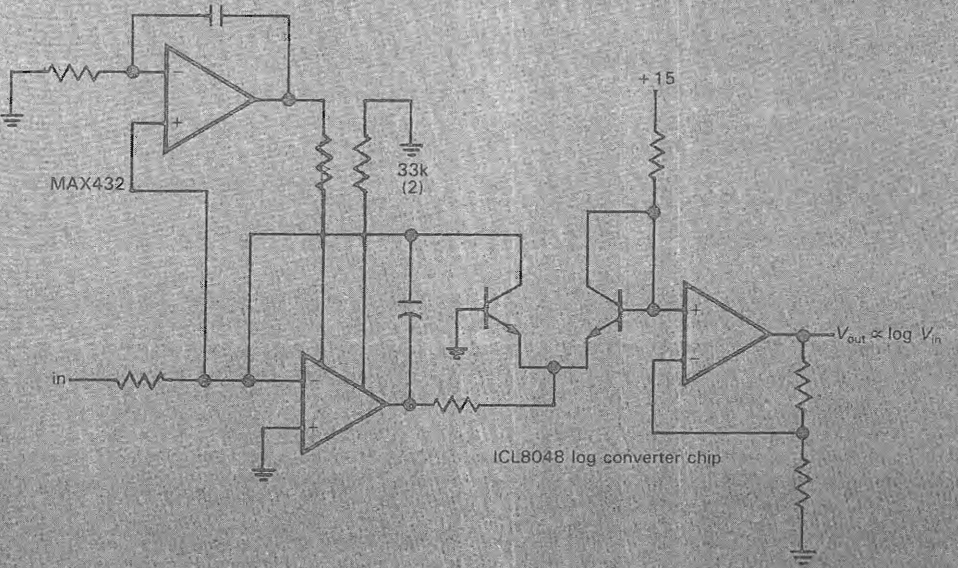
N. unusual 3-pin JFET follower



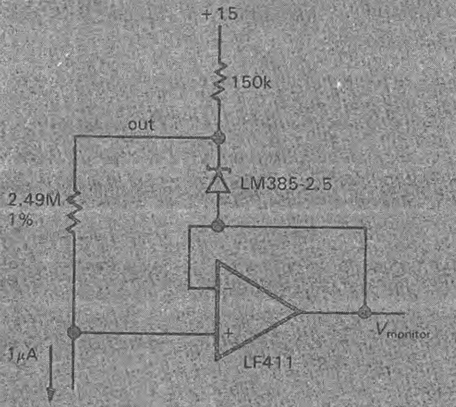
Q. 0.2A servo-motor amplifier



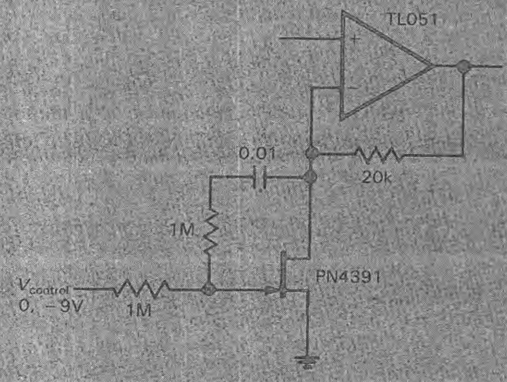
O. temperature-compensated log converter



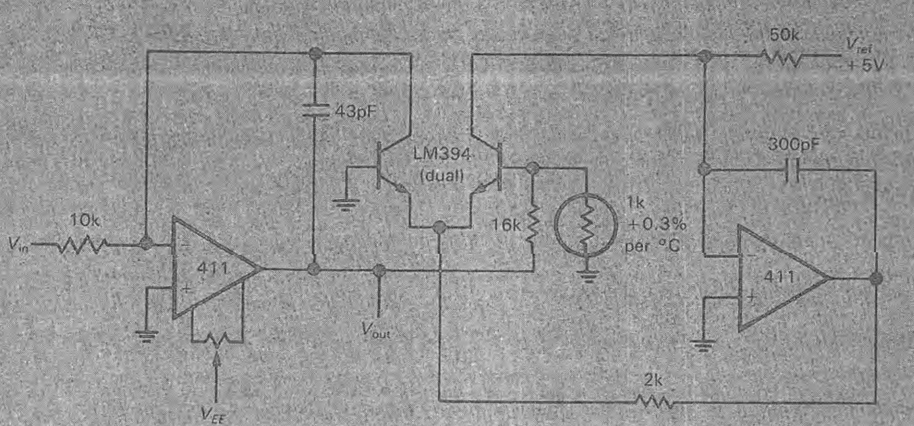
R. log converter with wide input range due to automatic nulling with a chopper amplifier



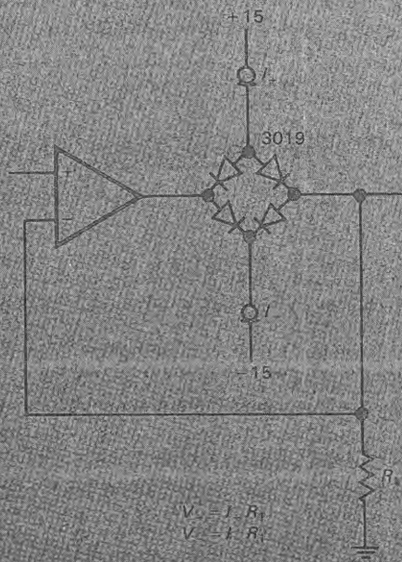
S. current source



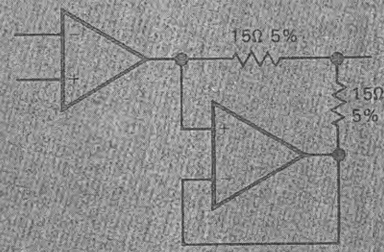
U. low-distortion variable-gain amplifier



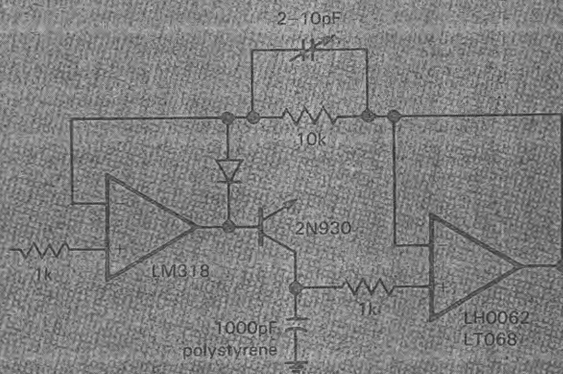
T. fast logarithmic converter



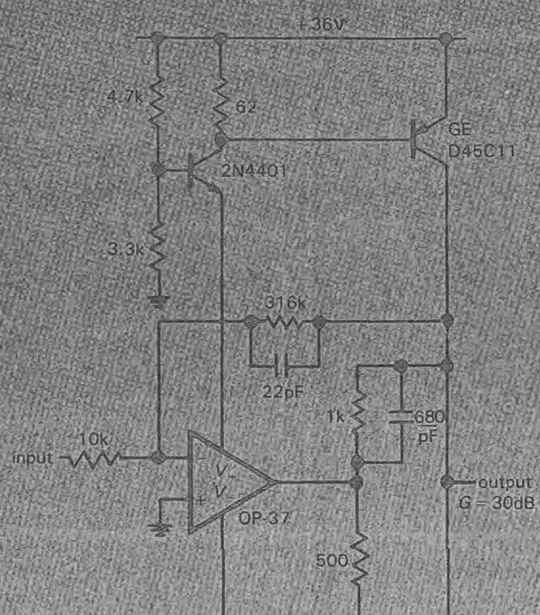
V. precision bipolarity limiter



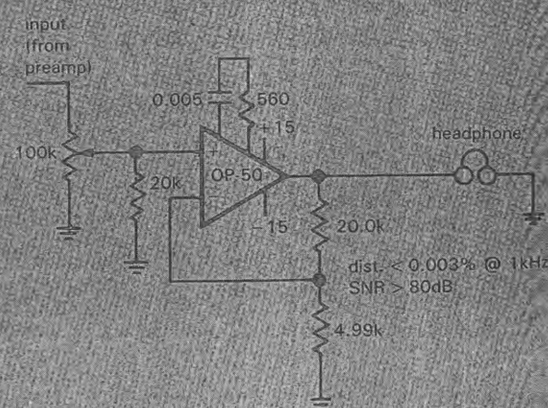
X. obtaining 2× output capability from a dual op-amp



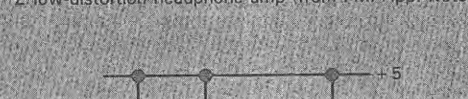
Y. high-speed peak detector



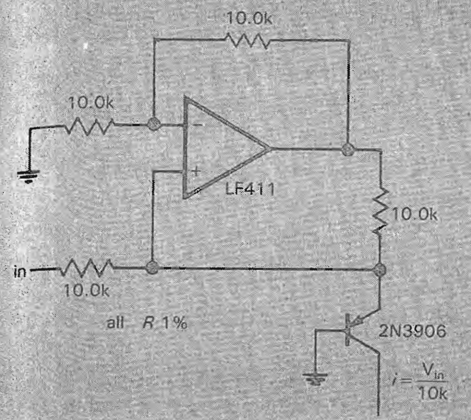
W. low-noise high-voltage op-amp (from PMI App. Note 106)



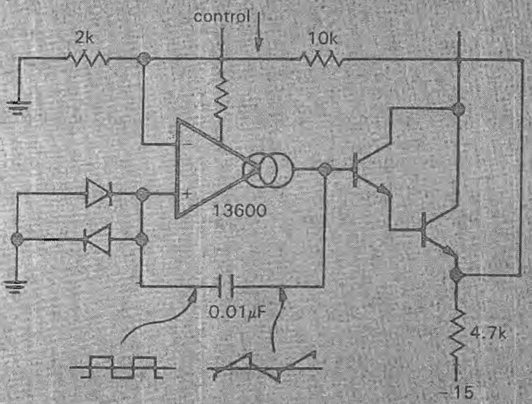
Z. low-distortion headphone amp (from PMI App. Note 106)



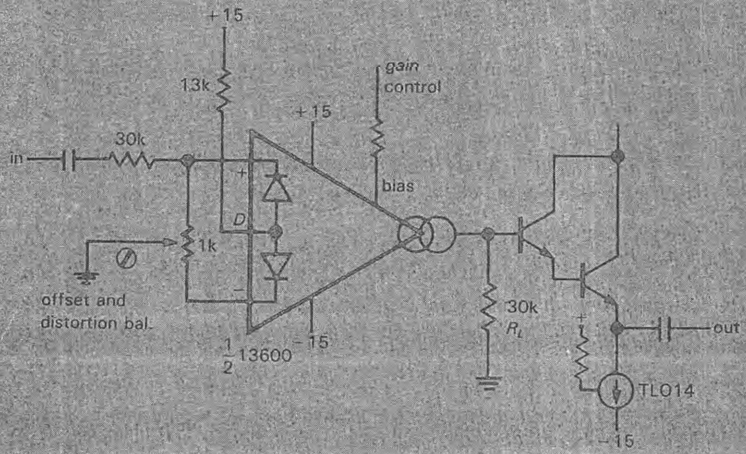
AA. zero-crossing detector, single supply



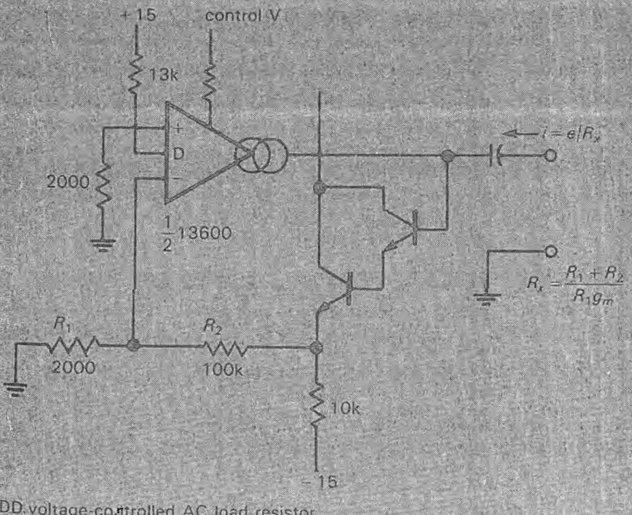
BB. Howland-style current source for transconductance voltage-current control circuits (1μA to 1mA)



EE. voltage-contolled oscillator with transconductance amplifier



CC. transconductance linearized voltage-controlled amplifier (VCA)



DD. voltage-controlled AC load resistor

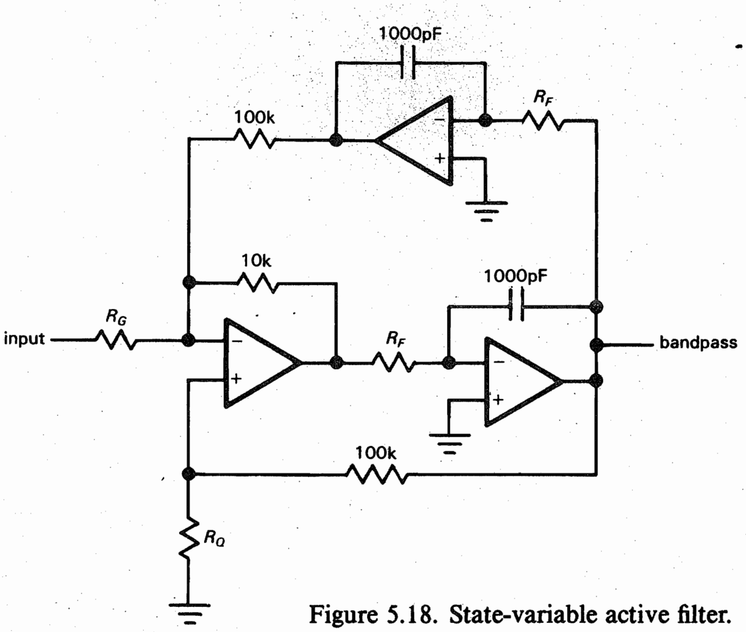


Figure 5.18. State-variable active filter.

but it is popular because of its improved stability and ease of adjustment. It is called a state-variable filter and is available as an IC from National (the AF100 and AF150), Burr-Brown (the UAF series), and others. Because it is a manufactured module, all components except R_G, R_Q, and the two R_Fs are built in. Among its nice properties is the availability of high-pass, low-pass, and bandpass outputs from the same circuit; also, its frequency can be tuned while maintaining constant Q (or, alternatively, constant bandwidth) in the bandpass characteristic. As with the VCVS realizations, multiple stages can be cascaded to generate higher-order filters.

Extensive design formulas and tables are provided by the manufacturers for the use of these convenient ICs. They show how to choose the external resistor values to make Butterworth, Bessel, and Chebyshev filters for a wide range of filter orders, for low-pass, high-pass, bandpass, and band-reject responses. Among the nice features of these hybrid ICs is integration of the capacitors into the module, so that only external resistors need be added.

Bandpass filters

The state-variable circuit, in spite of its large number of components, is a good choice for sharp (high-Q) bandpass filters. It has low component sensitivities, does not make great demands on op-amp bandwidth, and is easy to tune. For example, in the circuit of Figure 5.18, used as a bandpass filter, the two resistors R_F set the center frequency, while R_Q and R_G together determine the Q and band-center gain:

$$R_F = 5.03 \times 10^7/f_0 \text{ ohms}$$

$$R_Q = 10^5/(3.48Q + G - 1) \text{ ohms}$$

$$R_G = 3.16 \times 10^4 Q/G \text{ ohms}$$

So you could make a tunable-frequency, constant-Q filter by using a 2-section variable resistor (pot) for R_F. Alternatively, you could make R_Q adjustable, producing a fixed-frequency, variable-Q (and, unfortunately, variable-gain) filter.

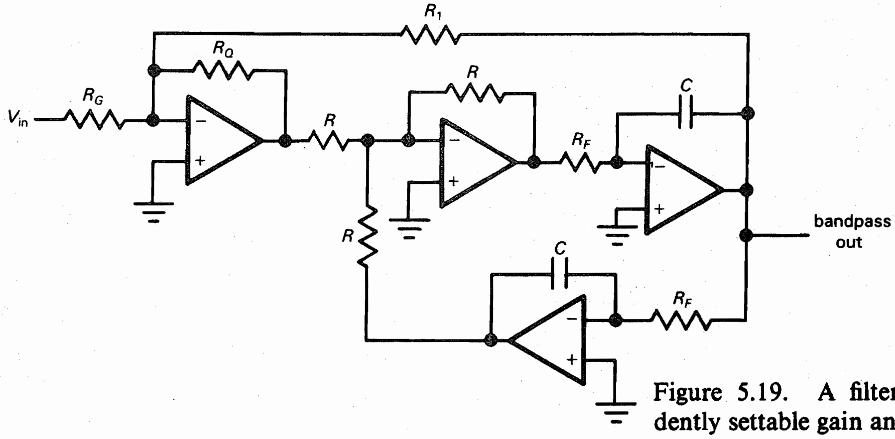


Figure 5.19. A filter with independently settable gain and Q.

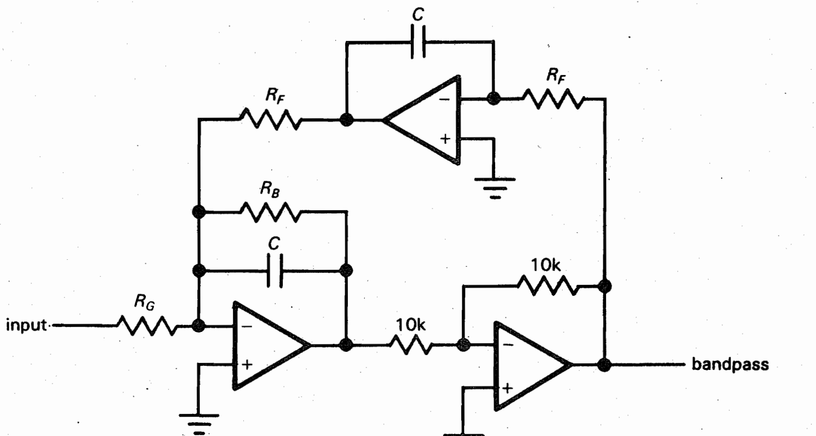


Figure 5.20. Biquad active filter.

EXERCISE 5.4

Calculate resistor values in Figure 5.18 to make a bandpass filter with $f_0 = 1\text{kHz}$, $Q = 50$, and $G = 10$.

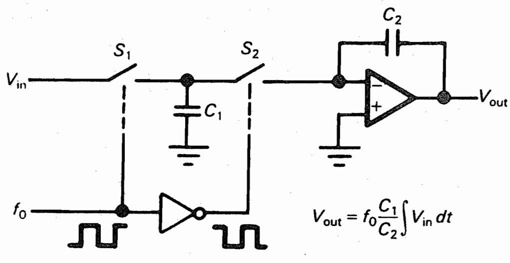
Figure 5.19 shows a useful variant of the state-variable bandpass filter. The bad news is that it uses four op-amps; the good news is that you can adjust the bandwidth (i.e., Q) without affecting the midband gain. In fact, both Q and gain are set with a single resistor each. Q, gain, and center frequency are completely independent and are given by these simple equations:

$f_0 = 1/2\pi R_F C$
$Q = R_1/R_Q$
$G = R_1/R_G$
$R \approx 10\text{k}$ (noncritical, matched)

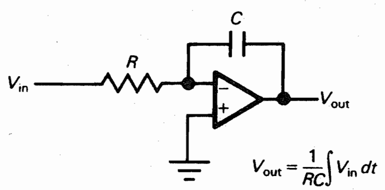
Biquad filter. A close relative of the state variable filter is the so-called biquad filter, shown in Figure 5.20. This circuit also uses three op-amps and can be constructed from the state-variable ICs mentioned earlier. It has the interesting property that you can tune its frequency (via R_F) while maintaining constant *bandwidth* (rather than constant Q). Here are the design equations:

$f_0 = 1/2\pi R_F C$
$\text{BW} = 1/2\pi R_B C$
$G = R_B/R_G$

The Q is given by f_0/BW and equals R_B/R_F. As the center frequency is varied (via R_F), the Q varies proportionately, keeping the bandwidth Qf_0 constant.



A



B

Figure 5.25. A. Switched-capacitor integrator B. conventional integrator.

There's another way to implement the integrators that are needed in the state-variable or biquad filter. The basic idea is to use MOS analog switches, clocked from an externally applied square wave at some high frequency (typically 100 times faster than the analog signals of interest), as shown in Figure 5.25. In the figure, the funny triangular object is a digital *inverter*, which turns the square wave upside down so that the two MOS switches are closed on opposite halves of the square wave. The circuit is easy to analyze: When S_1 is closed, C_1 charges to V_{in}, i.e., holding charge $C_1 V_{in}$; on the alternate half of the cycle, C_1 discharges into the virtual ground, transferring its charge to C_2. The voltage across C_2 therefore changes by an amount $\Delta V = \Delta Q/C_2 = V_{in}C_1/C_2$. Note that the output voltage *change* during each cycle of the fast square wave is proportional to V_{in} (which we assume changes only a small amount during one cycle of square wave), i.e., the circuit is an integrator! It is easy to show that the integrators obey the equations in the figure.

EXERCISE 5.6
Derive the equations in Figure 5.25

There are two important advantages to using switched capacitors instead of conventional integrators. First, as hinted earlier, it can be less expensive to implement on silicon: The integrator gain depends only on the *ratio* of two capacitors, not on their individual values. In general it is easy to make a matched pair of anything on silicon, but very hard to make a similar component (resistor or capacitor) of precise value and high stability. As a result, monolithic switched-capacitor filter ICs are very inexpensive – National's universal switched-capacitor filter (the MF10) costs \$2 (compared with \$10 for the conventional AF100) and furthermore gives you *two* filters in one package!

The second advantage of switched-capacitor filters is the ability to tune the filter's frequency (e.g., the center frequency of a bandpass filter, or the −3dB point of a low-pass filter) by merely changing the frequency of the square wave ("clock") input. This is because the characteristic frequency of a state-variable or biquad filter is proportional to (and depends only on) the integrator gain.

Switched-capacitor filters are available in both dedicated and "universal" configurations. The former are prewired with on-chip components to form bandpass or low-pass filters, while the latter have various intermediate inputs and outputs brought out so you can connect external components to make anything you want. The price you pay for universality is a larger IC package and the need for external resistors. For example, National's self-contained MF4 Butterworth low-pass filter comes in an 8-pin DIP (\$1.30), while their MF5 universal filter comes in a 14-pin DIP (\$1.45), requiring 2 or 3 external resistors (depending on which filter configuration you choose). Figure 5.26 shows just how easy it is to use the dedicated type.

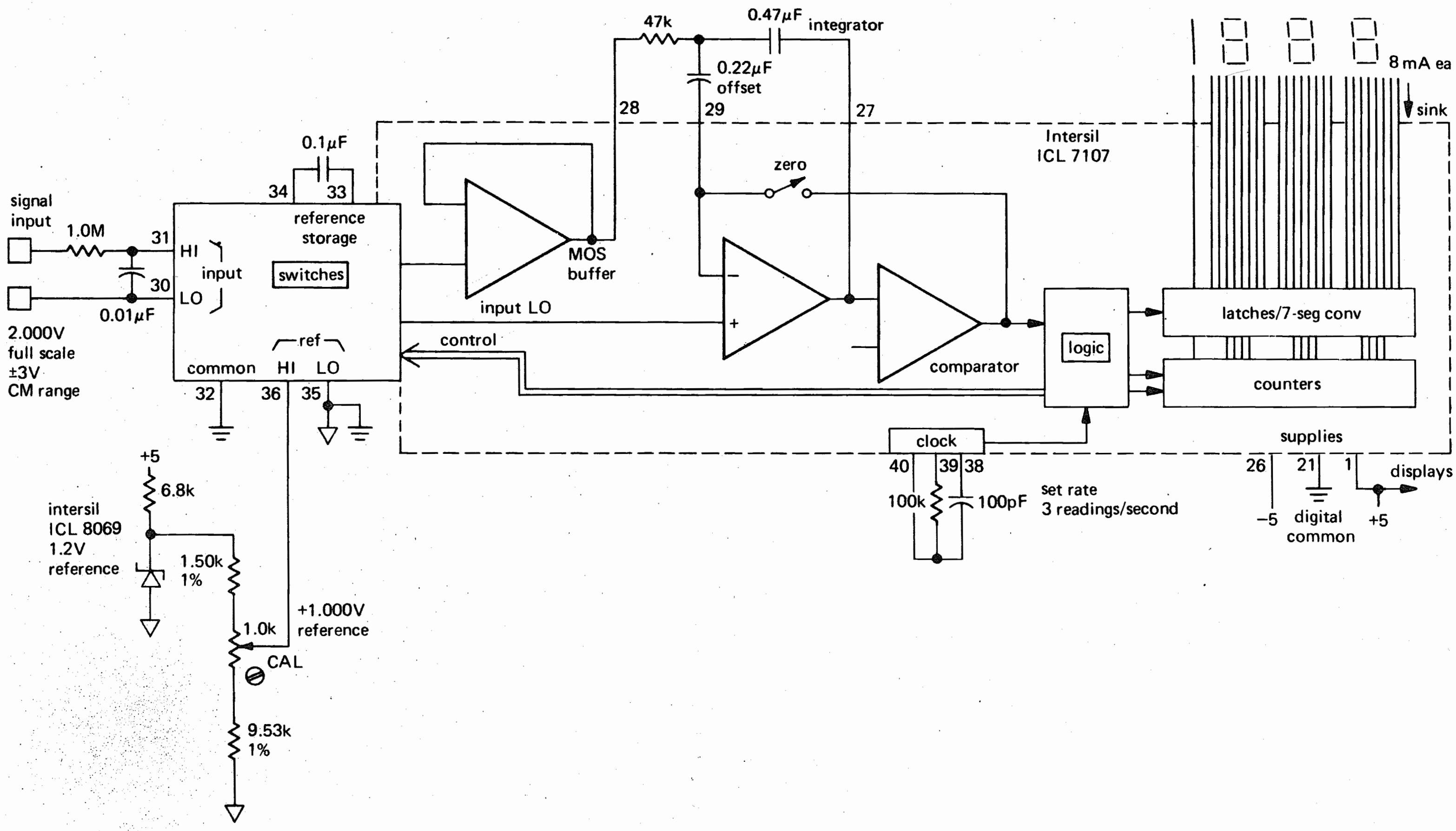


Figure 9.65. Single-chip $3\frac{1}{2}$-digit dual-slope DVM.

digital-voltmeter CMOS LSI chip, the only external components being the integrator and clock RCs, an accurate voltage reference, and the display itself. The ICL 7107 includes an automatic zeroing cycle in its operation, and it even generates all the 7-segment multiplexed outputs to drive a 4-digit LED display directly. By using an external attenuator at the input (or a reference of a different voltage), you can generate other full-scale voltage ranges. Dual-slope conversion is well suited to DVM operation because it provides good accuracy (including auto-zeroing) and 60Hz rejection in an averaging instrument at low cost; the converter chip used here costs less than \$20.

□ 9.26 Coulomb meter

The circuit in Figure 9.66 is a charge-balancing current integrator, or "coulomb meter." This instrument can be used to measure the integrated current (total charge) over some time period; it might find application in electrophoresis or electrochemistry. The action begins in the lower left-hand corner, where the current to be integrated flows through a precision 4-wire power resistor, generating a proportional voltage. IC_2 is a relatively low cost (under \$5) precision single-supply op-amp with low initial voltage offset ($80\mu V$ max) and low drift of offset with time and temperature (less than $2\mu V$ per degree and $0.5\mu V$ per month). It generates an output current, programmed by the current being measured, to drive the charge-balancing integrator, IC_3. Five decades of input sensitivity are selectable via the rotary switch at the input, with $200\mu A$ collector current in Q_1 corresponding to full-scale input in any range. Q_1 is a MOSFET (rather than a BJT), to eliminate control current error.

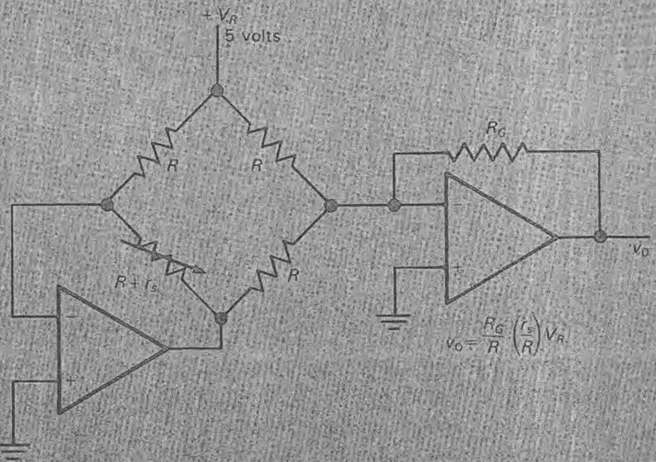
The charge-balancing circuitry is a standard delta-sigma scheme, with p-channel enhancement-mode MOSFET Q_2 doling out parcels of charge as directed by the state of flip-flop IC_{5a} after each clock cycle. IC_{5b} acts as a monostable, incrementing binary scaler IC_7 for each clock cycle during which Q_2 conducts. This circuit doesn't count for a fixed number of clock cycles, but simply integrates until it is stopped. The 4-digit counters IC_9 and IC_{10} deep track of the total charge, driving an 8-digit LED display.

If the current being measured ever exceeds the full-scale current of the range selected, Q_2 will be unable to balance Q_1's current even if it is ON continuously, and the measured charge registered in the counters will be in error. IC_{4a} checks for this overrange condition, lighting the LED if the integrator output rises past a fixed reference voltage (chosen comfortably larger than the output excursions of the integrator under normal conditions).

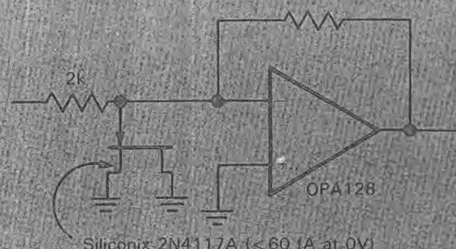
□ *Design calculations*

There are several interesting decisions that have to be made in designing a circuit like this. For instance, most of the CMOS logic is operated from +15 volts in order to simplify switching of Q_2. Since the 4-digit counters require +5 volts, a 4049 is used to interface the high-level CMOS logic signals to the counters. IC_4 is operated with single supply so that its output goes between ground and +15, for simple connection to IC_{5a}. The reference voltage for the integrator and comparator is put at about +4.7 volts by zener D_2, in order to allow headroom for Q_1; a simple zener is fine, since no accuracy is needed here. Note that a precision reference rides on the +4.7 volt level used to scale the current switched into the integrator. The REF-02's operating current is conveniently used to bias the zener.

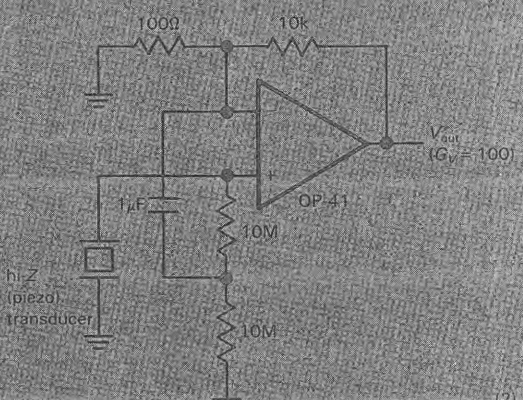
The choice of switch (Q_2) can critically affect the overall precision of the instrument. If it has too much capacitance, the additional charge residing on its drain will



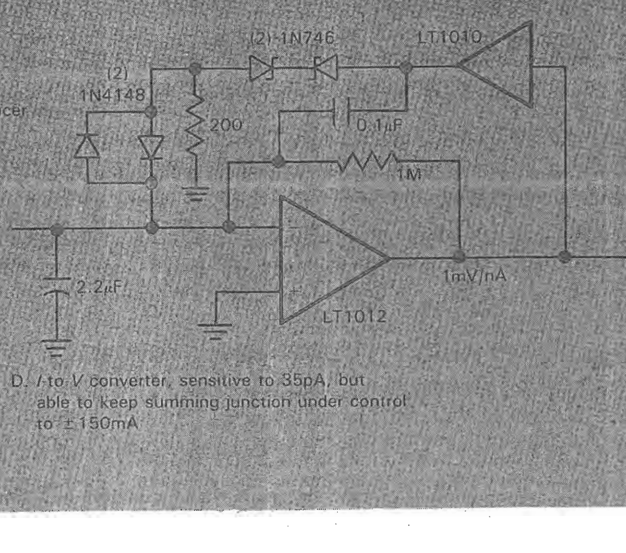
A. linearizes bridge response, with minimum V_{OS} effect; note that many bridges (e.g., semiconductor strain gauge) have high TCs of R



C. charge amplifier input protection with low leakage



B. high-Z_{in} bootstrapped amplifier for piezo transducer



D. I-to-V converter, sensitive to 35pA, but able to keep summing junction under control to ±150mA

Figure 15.45