sources this may create a problem, but it can be solved by introducing an intermediate buffer stage with a high input impedance (so as not to load the source) and a low output impedance (to minimize the effects of input bias current). The temperature coefficient of the offset current must produce offset voltages much less than the signal voltage over the anticipated temperature range. To minimize bias-current effects, the resistances to ground from both input terminals should be identical.

(3) Frequency response and compensation must prevent the amplifier from breaking into oscillation. The 741 is internally compensated. High-frequency amplifiers lack internal compensation, but have extra terminals for external compensation circuits.

(4) The maximum rate of change of a sinusoidal output voltage $V_0 \sin \omega t$ is $V_0\omega$. If this exceeds the slew rate of the amplifier, distortion will result. Slew rate is directly related to frequency compensation, and high slew rates are obtained with externally compensated amplifiers using the minimum compensation capacitance consistent with the particular configuration. Data sheets should be consulted.

6.4.3 Operational-Amplifier Circuit Analysis

If ideal behavior can be assumed (as is reasonable for circuits where the open-loop gain is much larger than the closed-loop gain), only the following two rules need be used to obtain the transfer function of an operational-amplifier circuit, as shown schematically in Figure 6.65:

(1) The voltage across the input terminals is zero.

(2) The currents into the input terminals are zero (the inverting terminal is sometimes called the *summing point*).

Applying these to the generalized circuit of Figure 6.65, we see that the voltage at the + input is $v_{i2}Z_3/(Z_2 + Z_3)$, since there is no current flowing into the + input and Z_2, Z_3 form a voltage divider. From Rule 1, the voltage at the − input must also equal $v_{i2}Z_3/(Z_2 + Z_3)$. The current through Z_1 is then the potential difference across it divided by Z_1, or:

$$\left[v_{i1} - v_{i2}\frac{Z_3}{(Z_2 + Z_3)}\right]/Z_1 \qquad (6.33)$$

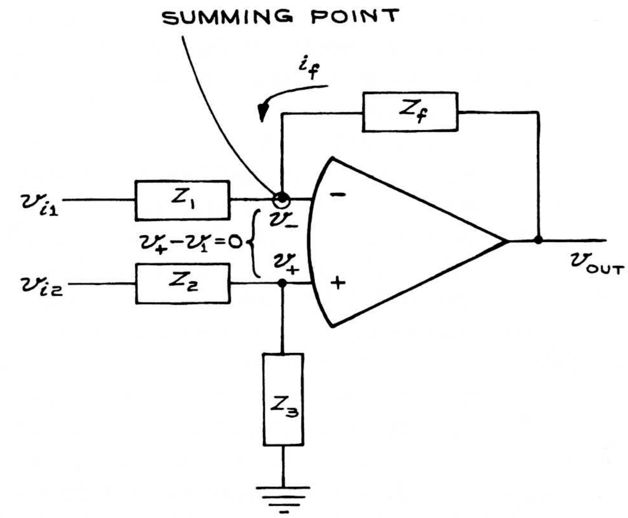

Figure 6.65 The general operational-amplifier circuit. Power supply and null terminals not shown.

This current must be equal in magnitude and opposite in sign to the current from the output through the feedback element Z_f. This current, i_f, is given by the potential difference across Z_f divided by Z_f:

$$\begin{aligned} i_f &= \left[v_{out} - v_{i2}\frac{Z_3}{(Z_2 + Z_3)}\right]/Z_f \\ &= -\left[v_{i1} - v_{i2}\frac{Z_3}{(Z_2 + Z_3)}\right]/Z_1 \end{aligned} \qquad (6.34)$$

Solving for v_{out}, we obtain:

$$-v_{out} = -v_{i1}\frac{Z_f}{Z_1} + v_{i2}(\frac{Z_3}{Z_2 + Z_3})(1 + \frac{Z_f}{Z_1}) \qquad (6.35)$$

Some useful configurations are illustrated in Figure 6.66.

The *summer* circuit [Figure 6.66(e)] is an elaboration of the inverting amplifier, with the current into the summing point coming from two external sources (v_A and v_B) canceled by the current from the feedback loop, v_{out}/R_f. If $R_A = R_B = R$, then $v_{out} = -(R_f/R)(v_A + v_B)$. If $R_A = R_B$, then v_{out} is equal to the negative of the weighted sum of voltages v_A and v_B, that is:

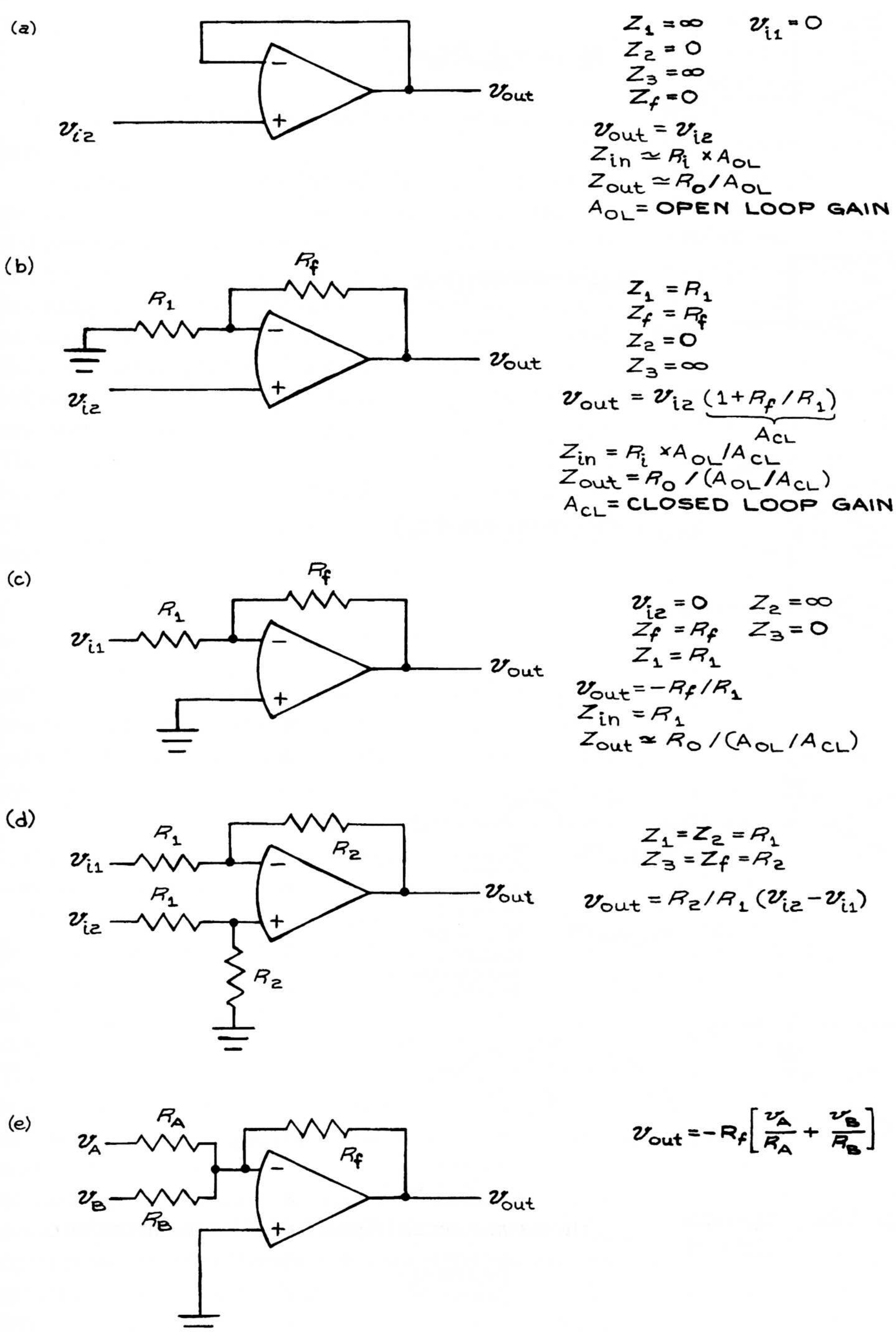

Figure 6.66 Operational-amplifier configurations: (a) follower; (b) follower with gain; (c) inverting amplifier; (d) subtractor; (e) summer; (f) low-pass filter (integrator); (g) high-pass filter (differentiator); (h) logarithmic amplifier; (i) precision rectifier; (j) clamp. The parameters are defined in Figure 6.59.

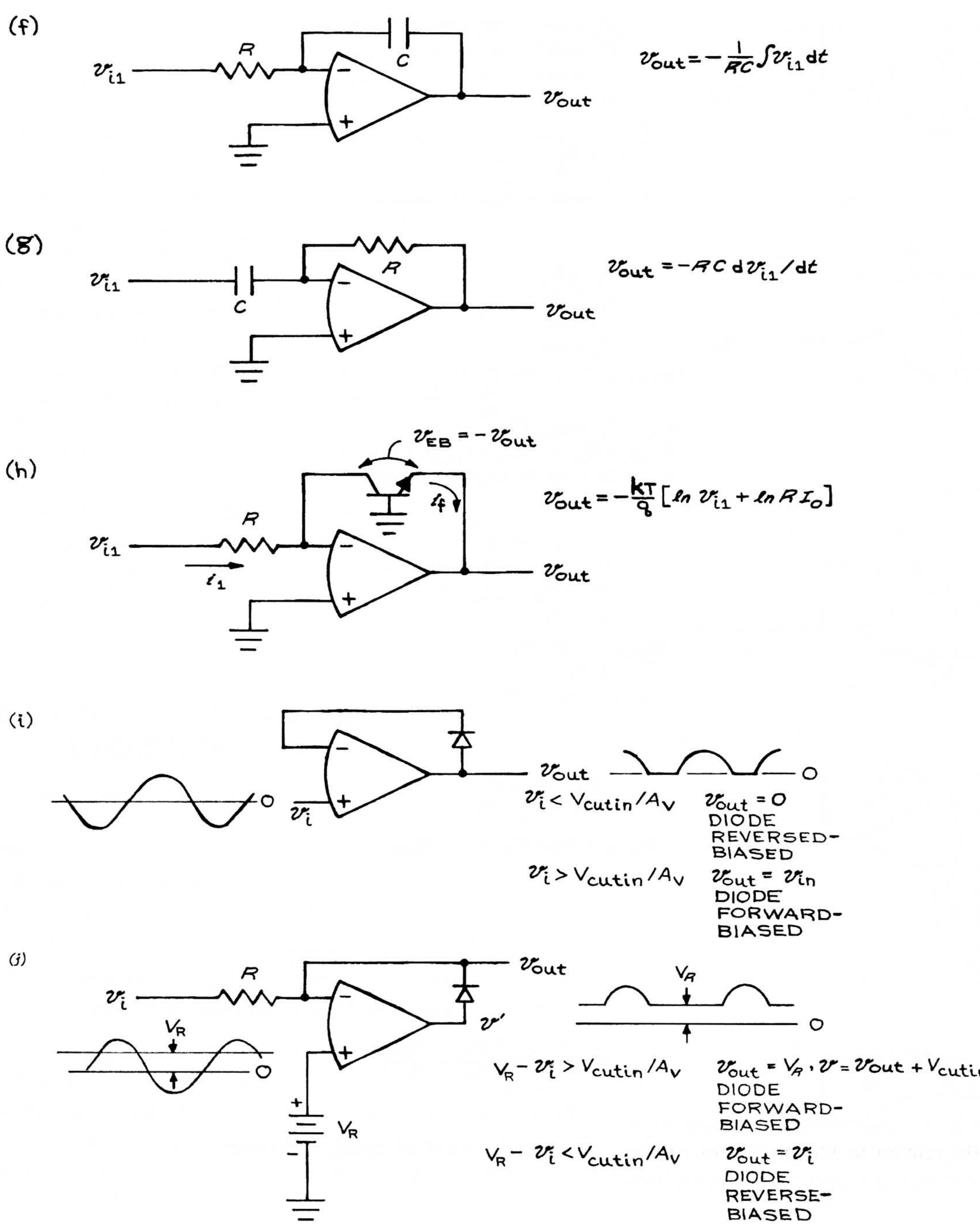

Figure 6.66 *(cont.)*

$$v_{out} = -[(\frac{R_f}{R_A})v_A + (\frac{R_f}{R_B})v_B] \quad (6.36)$$

Any number of external signal sources can be summed in this way.

A sinusoidal input to the *low-pass filter* [Figure 6.66(f)] gives $v_{out} = (j\omega RC)v_{i1}$. For a nonsinusoidal input, the current into the summing point from v_{i1} is v_{i1}/R, because the inverting input is at ground (rule 1). Often one speaks of the inverting input being at *virtual ground.* Although there is no direct connection from the inverting input to ground, the condition in which both inputs are at the same potential results in the inverting input assuming a potential of zero when the noninverting input is at ground potential. The current through the feedback capacitor of this circuit is i_f, and the voltage across the capacitor, v_{out} is equal to $(1/C)\int i_f dt$. Since i_f and v_{i1}/R are equal in magnitude and opposite in sign:

$$v_{out} = (1/C)\int i_f dt = -(1/RC)\int v_{i1} dt. \quad (6.37)$$

and the circuit acts as an *integrator* of the input voltage. Integration times of several minutes or even hours are possible with high quality, low-leakage capacitors and operational amplifiers with small offset voltages and bias currents.

For the *high-pass filter* [Figure 6.66(g)], $v_{out} = -RC(dv_{il}/dt)$ and the circuit acts as a differentiator.

The integrator and differentiator circuits are simple forms of active filters. More complex networks involving only capacitors and resistors can be used to obtain poles in the left half of the complex *s*-plane, which in the past were only obtained with inductors in passive filter networks. The operational amplifier allows the use of reasonable resistor and capacitor values even at frequencies as low as a few hertz. An additional advantage is the high isolation of input from output due to the low output impedance of most operational-amplifier circuits. The limitations of active filters are directly related to the properties of the operational amplifier. Inputs and outputs are usually single-ended and cannot be floated as passive filters can. The input and output voltage ranges are limited, as is the output current. Offset currents and voltages, bias currents, and temperature drifts can all affect active filter performance.

A good introduction to active filters is the *Active Filter Cookbook* by D. Lancaster.

The circuits in Figure 6.66(b) to (g) are examples of the use of operational amplifiers in analog computation. With these circuits, the mathematical operations of addition, subtraction, multiplication by a constant, integration, and differentiation can be performed. Multiplication or division of two voltages is accomplished by a logarithmic amplifier, an adder (for multiplication) or subtractor (for division), and an exponential amplifier.

Both the *logarithmic* and *exponential* amplifiers rely on the exponential *I–V* characteristics of a *p–n* junction. When this junction is the emitter–base junction of a transistor, the collector current I_C with zero collector–base voltage is:

$$I_C = I_0[\exp(qV_{EB}/\eta kT) - 1] \quad (6.38)$$

where I_0 is a constant for all transistors of a given type, V_{EB} is the emitter–base voltage, q is the electron charge and η is a constant, 1 for Ge and 2 for Si. Typically, I_0 is 10 to 15 nA for silicon planar transistors.

For $I_C \geqslant 10^{-8}$ A, the equation reduces to $I_C = I_0\exp(qV_{EB}/\eta kT)$. For the configuration shown in Figure 6.66(h), $i_1 = v_{il}/R = I_0\exp(qv_{out}/\eta kT)$ and $v_{out} = -(\eta kT/q)(\ln v_{i1} + \ln RI_0)$. Since k, T, q, η, R, and I_0 are constants, v_{out} will be proportional to v_{i1}. Practical logarithmic amplifiers can operate over three decades of input voltages. They do, however, need temperature-compensating circuits that generally require the use of matched pairs of transistors. In addition to their arithmetic use, logarithmic amplifiers are frequently used for transforming signal amplitudes that cover orders of magnitude to a linear scale. Exponential (antilogarithmic) amplifier circuits are logarithmic circuits with the input and feedback circuit elements interchanged.

Logarithmic and *exponential* amplifiers are examples of nonlinear circuits – the output is not linearly related to the input. Operational amplifiers are used extensively in nonlinear circuits. Because the cutin voltage of simple diodes is a few tenths of a volt (0.2 V for Ge and 0.6 V for Si), they cannot be used to rectify millivolt-level a.c. voltages. A diode in the feedback loop of an operational amplifier [see Figure 6.66(i)] results in a rectifier with a cutin voltage equal to the diode cutin voltage divided by the open-loop gain of the amplifier. With a slight modification, the rectifier circuit can be converted to a *clamp* [see Figure 6.66(j)] in which the output follows the input for voltages greater than a reference voltage V_R, but equals V_R for input voltages less than V_R.

Comparator circuits are used to compare an input signal with a reference voltage level. The output is driven to V^+ or V^- depending on whether the input is less than or greater than the reference level [see Figure 6.67(a)]. Using an amplifier in such a saturated mode is very poor practice and practical comparators with internally clamped outputs less than $|V^+|$ and $|V^-|$ are available. The LM311 is a common example. The transition between the two output states can be accelerated by the use of positive feedback [Figure 6.67(b)]. Such a circuit is called a *Schmitt trigger* and finds wide application in signal conditioning when it is necessary to convert a slowly changing input voltage into an output waveform with a very steep edge. The price paid for the fast transition is *hysteresis* – where the threshold voltage for a positive transition of the output is different from the threshold voltage for a negative transition. Hysteresis inhibits the output from oscillating between the two output states when the input is very close to the threshold level.

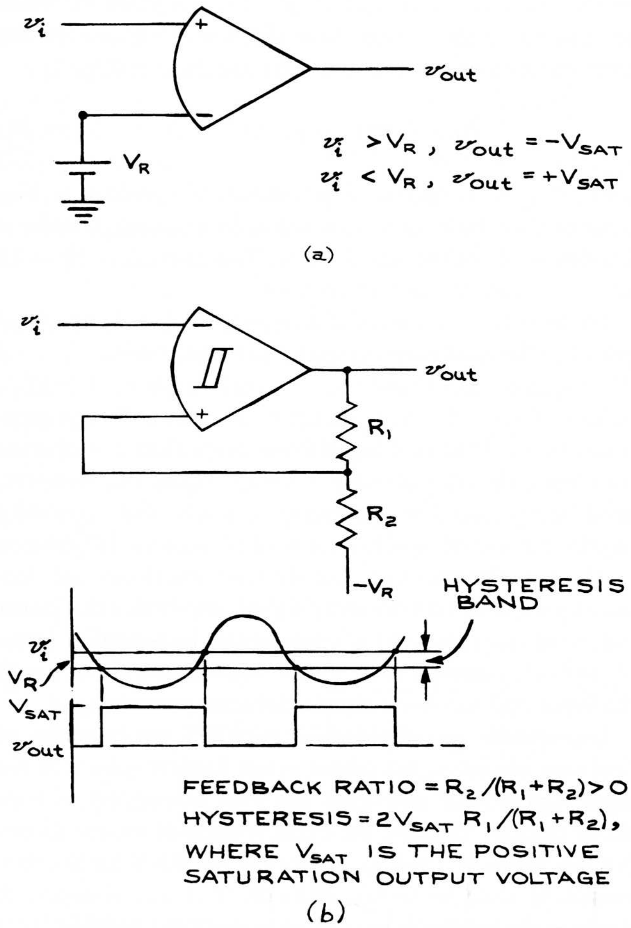

Figure 6.67 (a) Simple comparator ($-V_{sat}$ and $+V_{sat}$ are the minimum and maximum voltages the amplifier can deliver; V_R is a constant reference voltage); (b) Schmitt trigger (comparator with positive feedback).

6.4.4 Instrumentation and Isolation Amplifiers

Instrumentation amplifiers are used for amplifying low-level signals in the presence of large common-mode voltages, as, for example, in thermocouple and bridge circuits. The instrumentation amplifier is different from the operational amplifier; the feedback network of the instrumentation amplifier is integral to it and a single external resistor sets the gain. The output of the instrumentation amplifier is usually single-ended with respect to ground and is equal to the gain multiplied by the differential input voltage. Compared to an operational amplifier in the differential configuration with equivalent gain, the instrumentation amplifier has a higher common-mode rejection ratio and higher input impedance.

Instrumentation amplifiers are often made up of three operational amplifiers, as shown in Figure 6.68. Amplifiers A1 and A2 together provide a differential gain of $1 + 2R_1/R_G$ and a common-mode gain of unity. Amplifier A3 is a unit-gain differential amplifier. Amplifiers A1 and A2 operate as followers with gain, and resistors R_1 do not affect the input impedances. Because amplifier A3 is driven by the low output impedances of Al and A2, the resistors R_o can have relatively small values in order to optimize CMRR and frequency response while minimizing the effects of input offset currents.

The three-amplifier configuration is limited to common-mode voltages of ± 8 V in most applications. Typical high-performance instrumentation amplifiers are the 1NA110 (Burr-Brown), LH0038 (National Semiconductor), and AD624 (Analog Devices). Typical specifications for high-performance amplifiers are given in Table 6.24.

Instrumentation amplifiers cannot be used when the common-mode voltage exceeds the power-supply voltage.