

Signal Processing

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Amplifiers and Signal Conditioners

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20.1 Introduction

Signals from sensors do not usually have suitable characteristics for display, recording, transmission, or further processing. For example, they may lack the amplitude, power, level, or bandwidth required, or they may carry superimposed interference that masks the desired information.

Signal conditioners, including amplifiers, adapt sensor signals to the requirements of the receiver (circuit or equipment) to which they are to be connected. The functions to be performed by the signal conditioner derive from the nature of both the signal and the receiver. Commonly, the receiver requires a single-ended, low-frequency (dc) voltage with low output impedance and amplitude range close to its power-supply voltage(s). A typical receiver here is an analog-to-digital converter (ADC). Signals from sensors can be analog or digital. Digital signals come from position encoders, switches, or oscillator-based sensors connected to frequency counters. The amplitude for digital signals must be compatible with logic levels for the digital receiver, and their edges must be fast enough to prevent any false triggering. Large voltages can be attenuated by a voltage divider and slow edges can be accelerated by a Schmitt trigger.

Analog sensors are either self-generating or modulating. *Self-generating sensors* yield a voltage (thermocouples, photovoltaic, and electrochemical sensors) or current (piezo- and pyroelectric sensors) whose bandwidth equals that of the measurand. Modulating sensors yield a variation in resistance, capacitance, self-inductance or mutual inductance, or other electrical quantities. *Modulating sensors* need to be excited

or biased (semiconductor junction-based sensors) in order to provide an output voltage or current. Impedance variation-based sensors are normally placed in voltage dividers, or in Wheatstone bridges (resistive sensors) or ac bridges (resistive and reactance-variation sensors). The bandwidth for signals from modulating sensors equals that of the measured in dc-excited or biased sensors, and is twice that of the measurand in ac-excited sensors (sidebands about the carrier frequency) (see [Chapter 21](#)). Capacitive and inductive sensors require an ac excitation, whose frequency must be at least ten times higher than the maximal frequency variation of the measurand. Pallás-Areny and Webster [1] give the equivalent circuit for different sensors and analyze their interface.

Current signals can be converted into voltage signals by inserting a series resistor into the circuit. Graeme [2] analyzes current-to-voltage converters for photodiodes, applicable to other sources. Henceforth, we will refer to voltage signals to analyze transformations to be performed by signal conditioners.

20.2 Dynamic Range

The **dynamic range** for a measurand is the quotient between the measurement range and the desired resolution. Any stage for processing the signal from a sensor must have a dynamic range equal to or larger than that of the measurand. For example, to measure a temperature from 0 to 100°C with 0.1°C resolution, we need a dynamic range of at least $(100 - 0)/0.1 = 1000$ (60 dB). Hence, a 10-bit ADC should be appropriate to digitize the signal because $2^{10} = 1024$. Let us assume we have a 10-bit ADC whose input range is 0 to 10 V; its resolution will be $10 \text{ V}/1024 = 9.8 \text{ mV}$. If the sensor sensitivity is $10 \text{ mV}/^\circ\text{C}$ and we connect it to the ADC, the 9.8 mV resolution for the ADC will result in a $9.8 \text{ mV}/(10 \text{ mV}/^\circ\text{C}) = 0.98^\circ\text{C}$ resolution! In spite of having the suitable dynamic range, we do not achieve the desired resolution in temperature because the output range of our sensor (0 to 1 V) does not match the input range for the ADC (0 to 10 V).

The basic function of voltage amplifiers is to amplify the input signal so that its output extends across the input range of the subsequent stage. In the above example, an amplifier with a gain of 10 would match the sensor output range to the ADC input range. In addition, the output of the amplifier should depend only on the input signal, and the signal source should not be disturbed when connecting the amplifier. These requirements can be fulfilled by choosing the appropriate amplifier depending on the characteristics of the input signal.

20.3 Signal Classification

Signals can be classified according to their amplitude level, the relationship between their source terminals and ground, their bandwidth, and the value of their output impedance. Signals lower than around 100 mV are considered to be low level and need amplification. Larger signals may also need amplification depending on the input range of the receiver.

Single-Ended and Differential Signals

A *single-ended signal* source has one of its two output terminals at a constant voltage. For example, [Figure 20.1a](#) shows a voltage divider whose terminal L remains at the power-supply reference voltage regardless of the sensor resistance, as shown in [Figure 20.1b](#). If terminal L is at ground potential (grounded power supply in [Figure 20.1a](#)), then the signal is single ended and grounded. If terminal L is isolated from ground (for example, if the power supply is a battery), then the signal is single ended and floating. If terminal L is at a constant voltage with respect to ground, then the signal is single ended and driven off ground. The voltage at terminal H will be the sum of the signal plus the off-ground voltage. Therefore, the off-ground voltage is common to H and L; hence, it is called the **common-mode voltage**. For example, a thermocouple bonded to a power transistor provides a signal whose amplitude depends on the temperature of the transistor case, riding on a common-mode voltage equal to the case voltage.

A *differential signal* source has two output terminals whose voltages change simultaneously by the same magnitude but in opposite directions. The Wheatstone bridge in [Figure 20.1c](#) provides a differential

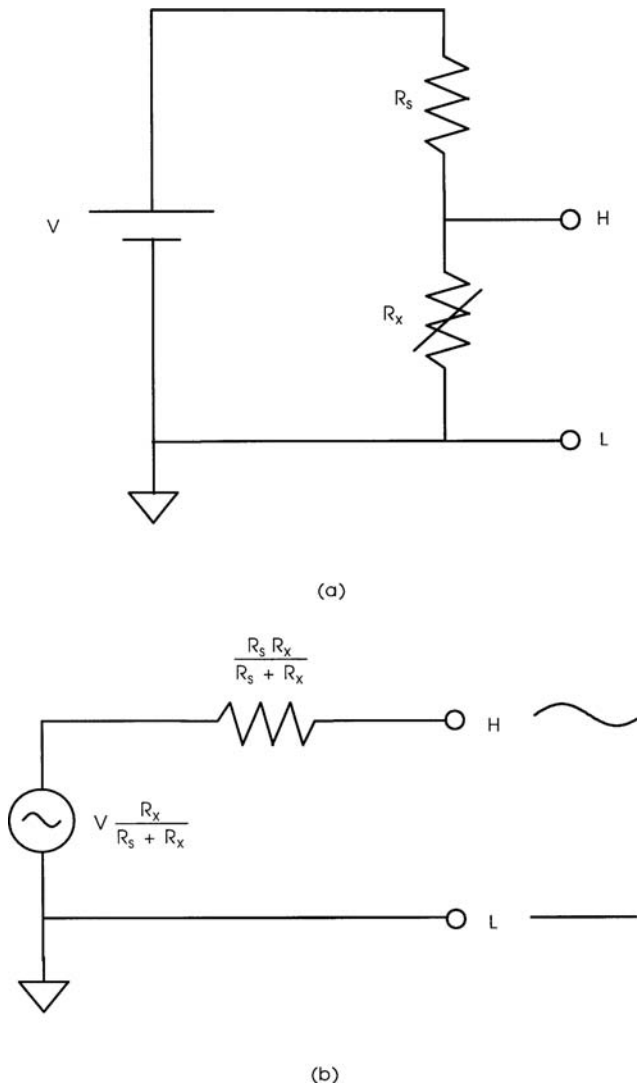


FIGURE 20.1 Classes of signals according to their source terminals. A voltage divider (a) provides a single-ended signal (b) where terminal L is at a constant voltage. A Wheatstone bridge with four sensors (c) provides a balanced differential signal which is the difference between two voltages v_H and v_L having the same amplitude but opposite signs and riding on a common-mode voltage V_c . For differential signals much smaller than the common-mode voltage, the equivalent circuit in (e) is used. If the reference point is grounded, the signal (single-ended or differential) will be grounded; if the reference point is floating, the signal will also be floating.

signal. Its equivalent circuit (Figure 20.1d) shows that there is a differential voltage ($v_d = v_H - v_L$) proportional to x and a common-mode voltage ($V_c = V/2$) that does not carry any information about x . Further, the two output impedances are balanced. We thus have a balanced differential signal with a superimposed common-mode voltage. Were the output impedances different, the signal would be unbalanced. If the bridge power supply is grounded, then the differential signal will be grounded; otherwise, it will be floating. When the differential signal is very small as compared with the common-mode voltage, in order to simplify circuit analysis it is common to use the equivalent circuit in Figure 20.1e. Some differential signals (grounded or floating) do not bear any common-mode voltage.

Signal conditioning must ensure the compatibility between sensor signals and receivers, which will depend on the relationship between input terminals and ground. For example, a differential and grounded

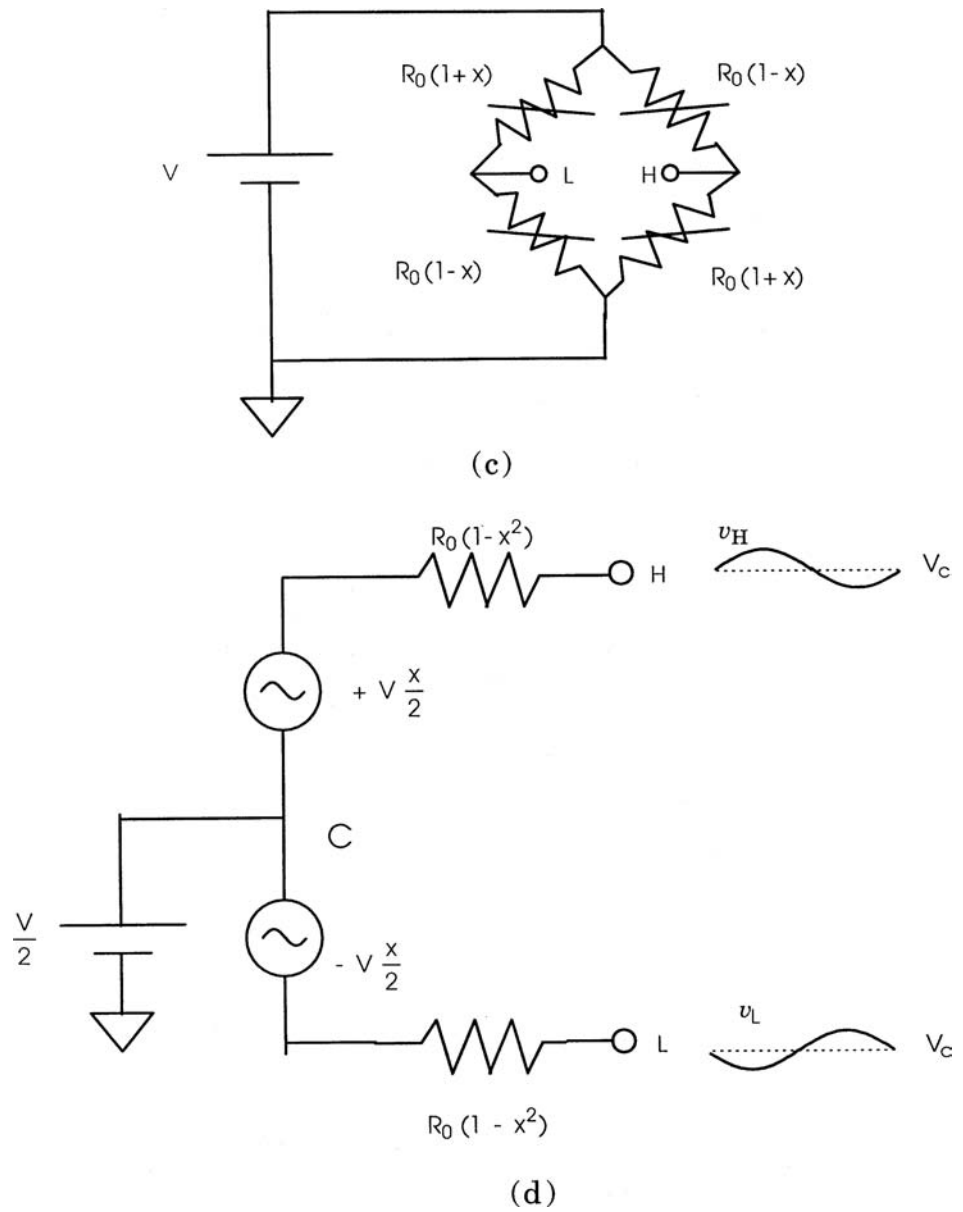


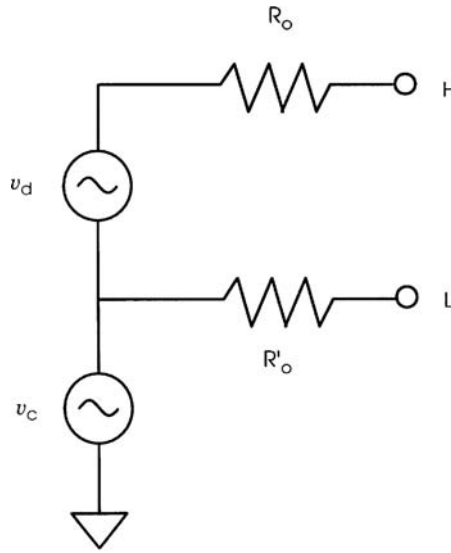
FIGURE 20.1 (continued)

signal is incompatible with an amplifier having a grounded input terminal. Hence, amplifiers must also be described according to their input topology.

Narrowband and Broadband Signals

A *narrowband signal* has a very small frequency range relative to its central frequency. Narrowband signals can be dc, or static, resulting in very low frequencies, such as those from a thermocouple or a weighing scale, or ac, such as those from an ac-driven modulating sensor, in which case the exciting frequency (carrier) becomes the central frequency (see [Chapter 21](#)).

Broadband signals, such as those from sound and vibration sensors, have a large frequency range relative to their central frequency. Therefore, the value of the central frequency is crucial; a signal ranging



(e)

FIGURE 20.1 (continued)

from 1 Hz to 10 kHz is a broadband instrumentation signal, but two 10 kHz sidebands around 1 MHz are considered to be a narrowband signal. Signal conditioning of ac narrowband signals is easier because the conditioner performance only needs to be guaranteed with regard to the carrier frequency.

Low- and High-Output-Impedance Signals

The output impedance of signals determines the requirements of the input impedance of the signal conditioner. Figure 20.2a shows a voltage signal connected to a device whose input impedance is Z_d . The voltage detected will be

$$v_d = v_s \frac{Z_d}{Z_d + Z_s} \quad (20.1)$$

Therefore, the voltage detected will equal the signal voltage only when $Z_d \gg Z_s$; otherwise $v_d \neq v_s$ and there will be a *loading effect*. Furthermore, it may happen that a low Z_d disturbs the sensor, changing the value of v_s and rendering the measurement useless or, worse still, damaging the sensor.

At low frequencies, it is relatively easy to achieve large input impedances even for high-output-impedance signals, such as those from piezoelectric sensors. At high frequencies, however, stray input capacitances make it more difficult. For narrowband signals this is not a problem because the value for Z_s and Z_d will be almost constant and any attenuation because of a loading effect can be taken into account later. However, if the impedance seen by broadband signals is frequency dependent, then each frequency signal undergoes different attenuations which are impossible to compensate for.

Signals with very high output impedance are better modeled as current sources, Figure 20.2b. The current through the detector will be

$$i_d = i_s \frac{Z_s}{Z_d + Z_s} \quad (20.2)$$

In order for $i_d = i_s$, it is required that $Z_d \ll Z_s$ which is easier to achieve than $Z_d \gg Z_s$. If Z_d is not low enough, then there is a *shunting effect*.

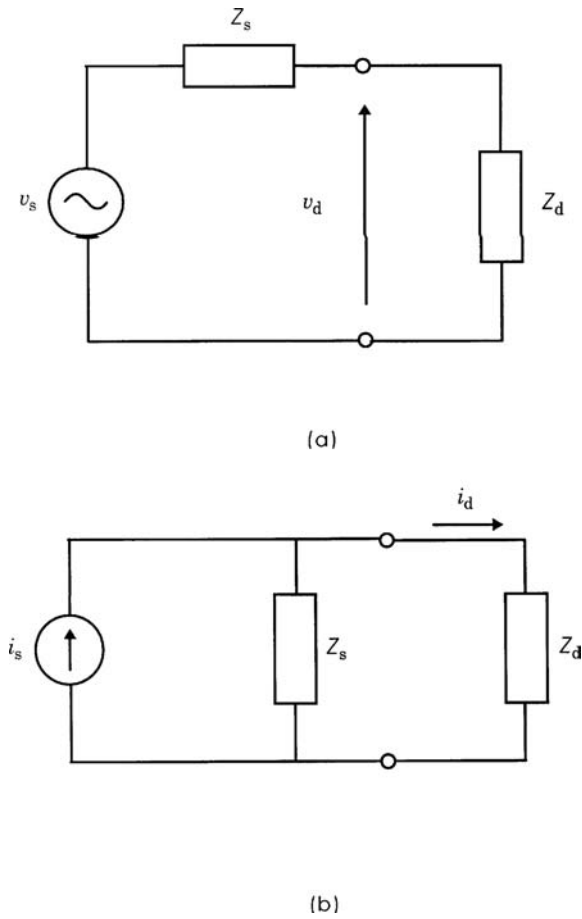


FIGURE 20.2 Equivalent circuit for a voltage signal connected to a voltage detector (a) and for a current signal connected to a current detector (b). We require $Z_d \gg Z_o$ in (a) to prevent any loading effect, and $Z_d \ll Z_s$ in (b) to prevent any shunting effect.

20.4 General Amplifier Parameters

A *voltage amplifier* produces an output voltage which is a proportional reproduction of the voltage difference at its input terminals, regardless of any common-mode voltage and without loading the voltage source. Figure 20.3a shows the equivalent circuit for a general (differential) amplifier. If one input terminal is connected to one output terminal as in Figure 20.3b, the amplifier is single ended; if this common terminal is grounded, the amplifier is single ended and grounded; if the common terminal is isolated from ground, the amplifier is single ended and floating. In any case, the output power comes from the power supply, and the input signal only controls the shape of the output signal, whose amplitude is determined by the *amplifier gain*, defined as

$$G = \frac{v_o}{v_d} \tag{20.3}$$

The ideal amplifier would have any required gain for all signal frequencies. A practical amplifier has a gain that rolls off at high frequency because of parasitic capacitances. In order to reduce noise and reject

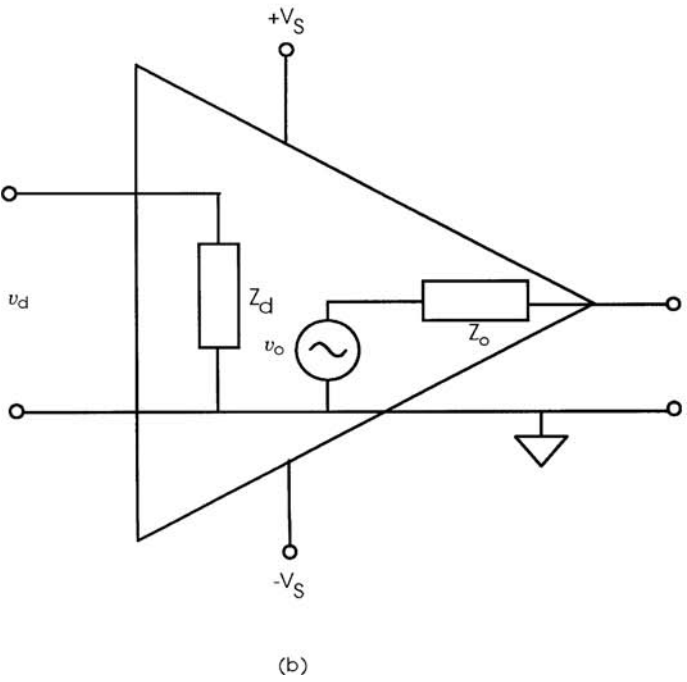
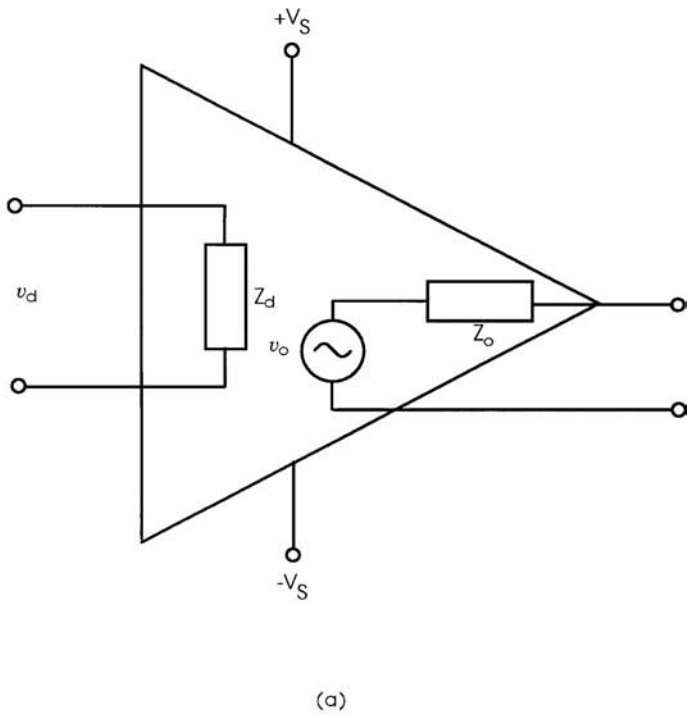


FIGURE 20.3 General amplifier, differential (a) or single ended (b). The input voltage controls the amplitude of the output voltage, whose power comes from the power supply.

interference, it is common to add reactive components to reduce the gain for out-of-band frequencies further. If the gain decreases by n times 10 when the frequency increases by 10, we say that the gain (downward) slope is $20 n$ dB/decade. The corner (or -3 dB) frequency f_0 for the amplifier is that for which the gain is 70% of that in the bandpass. (Note: $20 \log 0.7 = -3$ dB). The *gain error* at f_0 is then 30%, which is too large for many applications. If a maximal error ϵ is accepted at a given frequency f , then the corner frequency for the amplifier should be

$$f_0 = \frac{f(1-\epsilon)}{\sqrt{2\epsilon - \epsilon^2}} \approx \frac{f}{\sqrt{2\epsilon}} \quad (20.4)$$

For example, $\epsilon = 0.01$ requires $f_0 = 7 f$, $\epsilon = 0.001$ requires $f_0 = 22.4 f$. A broadband signal with frequency components larger than f would undergo amplitude distortion. A narrowband signal centered on a frequency larger than f would be amplified by a gain lower than expected, but if the actual gain is measured, the gain error can later be corrected.

Whenever the gain decreases, the output signal is delayed with respect to the input. In the above amplifier, an input sine wave of frequency f_0 will result in an output sine wave delayed by 45° (and with relative attenuation 30% as compared with a sine wave of frequency $f \gg f_0$). Complex waveforms having frequency components close to f_0 would undergo shape (or phase) distortion. In order for a waveform to be faithfully reproduced at the output, the phase delay should be either zero or proportional to the frequency (linear phase shift). This last requirement is difficult to meet. Hence, for broadband signals it is common to design amplifiers whose bandwidth is larger than the maximal input frequency. Narrowband signals undergo a delay which can be measured and corrected.

An ideal amplifier would have infinite *input impedance*. Then no input current would flow when connecting the signal, [Figure 20.2a](#), and no energy would be taken from the signal source, which would remain undisturbed. A practical amplifier, however, will have a finite, yet large, input impedance at low frequencies, decreasing at larger frequencies because of stray input capacitances. If sensors are connected to conditioners by coaxial cables with grounded shields, then the capacitance to ground can be very large (from 70 to 100 pF/m depending on the cable diameter). This capacitance can be reduced by using driven shields (or guards) (see [Chapter 29](#)). If twisted pairs are used instead, the capacitance between wires is only about 5 to 8 pF/m, but there is an increased risk of capacitive interference.

Signal conditioners connected to remote sensors must be protected by *limiting* both *voltage* and *input currents*. Current can be limited by inserting a power resistor (100 Ω to 1 k Ω , 1 W for example), a PTC resistor, or a fuse between each signal source lead and conditioner input. Input voltages can be limited by connecting diodes, zeners, metal-oxide varistors, gas-discharge devices, or other surge-suppression nonlinear devices, from each input line to dc power-supply lines or to ground, depending on the particular protecting device. Some commercial voltage limiters are Thyzorb® and Transzorb® (General Semiconductor), Transil® and Trisil® (SGS-Thomson), SIOV® (Siemens), and TL7726 (Texas Instruments).

The ideal amplifier would also have zero *output impedance*. This would imply no loading effect because of a possible finite input impedance for the following stage, low output noise, and unlimited output power. Practical amplifiers can indeed have a low output impedance and low noise, but their output power is very limited. Common signal amplifiers provide at best about 40 mA output current and sometimes only 10 mA. The power gain, however, is quite noticeable, as input currents can be in the picoampere range (10^{-12} A) and input voltages in the millivolt range (10^{-3} V); a 10 V, 10 mA output would mean a power gain of 10^{14} ! Yet the output power available is very small (100 mW). Power amplifiers are quite the opposite; they have a relatively small power gain but provide a high-power output. For both signal and power amplifiers, output power comes from the power supply, not from the input signal.

Some sensor signals do not require amplification but only *impedance transformation*, for example, to match their output impedance to that of a transmission line. Amplifiers for impedance transformation (or matching) and $G = 1$ are called buffers.

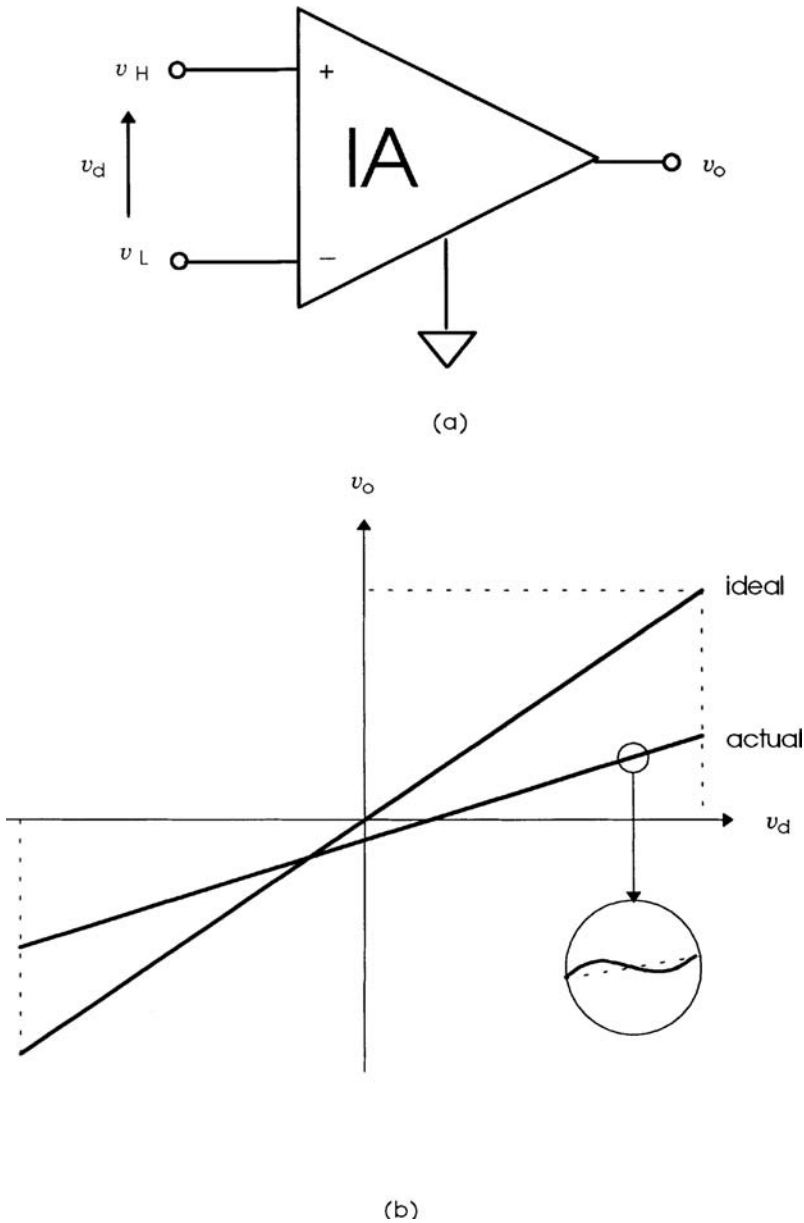


FIGURE 20.4 Instrumentation amplifier. (a) Symbol. (b) Ideal and actual input/output relationship. The ideal response is a straight line through the point (0,0) and slope G .

20.5 Instrumentation Amplifiers

For instrumentation signals, the so-called **instrumentation amplifier** (IA) offers performance closest to the ideal amplifier, at a moderate cost (from \$1.50 up). Figure 20.4a shows the symbol for the IA and Figure 20.4b its input/output relationship; ideally this is a straight line with slope G and passing through the point (0,0), but actually it is an off-zero, seemingly straight line, whose slope is somewhat different from G . The output voltage is

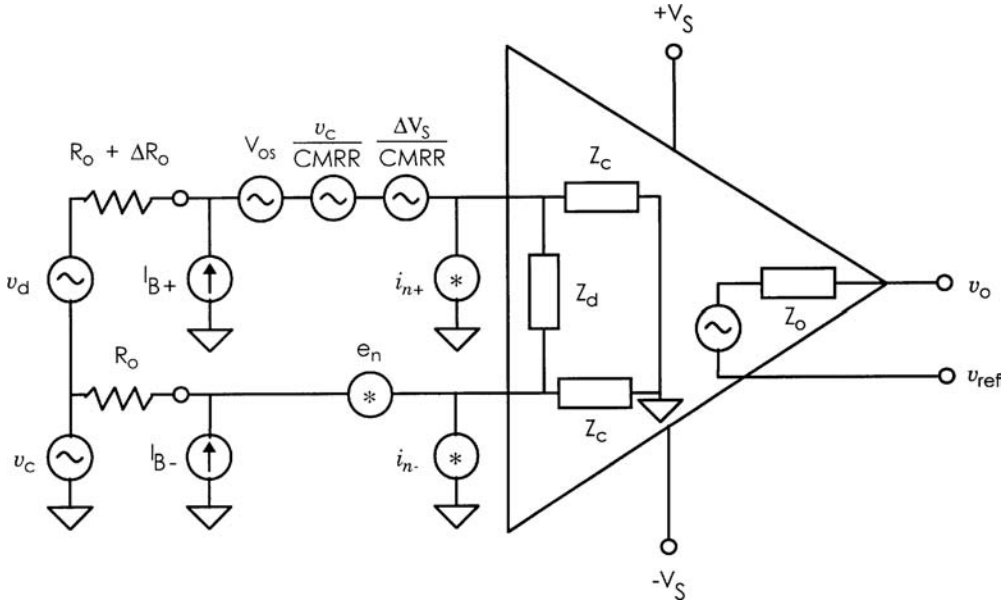


FIGURE 20.5 A model for a practical instrumentation amplifier including major error sources.

$$v_o = v_a + (v_{os} + v_b + v_r + v_n)G + v_{ref} \quad (20.5)$$

where v_a depends on the input voltage v_d , the second term includes offset, drift, noise, and interference-rejection errors, G is the designed gain, and v_{ref} is the reference voltage, commonly 0 V (but not necessarily, thus allowing output level shifting). Equation 20.5 describes a worst-case situation where absolute values for error sources are added. In practice, some cancellation between different error sources may happen.

Figure 20.5 shows a circuit model for *error analysis* when a practical IA is connected to a signal source (assumed to be differential for completeness). Impedance from each input terminal to ground (Z_c) and between input terminals (Z_d) are all finite. Furthermore, if the input terminals are both connected to ground, v_o is not zero and depends on G ; this is modeled by V_{os} . If the input terminals are grounded through resistors, then v_o also depends on the value of these resistors; this is modeled by current sources I_{B+} and I_{B-} , which represent input bias or leakage currents. These currents need a return path, and therefore a third lead connecting the signal source to the amplifier, or a common ground, is required. Neither V_{os} nor I_{B+} nor I_{B-} is constant; rather, they change with temperature and time: slow changes (<0.01 Hz) are called drift and fast changes are described as noise (hence the noise sources e_n , i_{n+} , and i_{n-} in Figure 20.5). Common specifications for IAs are defined in Reference 3.

If a voltage v_c is simultaneously applied to both inputs, then v_o depends on v_c and its frequency. The *common-mode gain* is

$$G_c(f) = \frac{V_o(v_d = 0)}{V_c} \quad (20.6)$$

In order to describe the output voltage due to v_c as an input error voltage, we must divide the corresponding $v_o(v_c)$ by G (the normal- or differential-mode gain, $G = G_d$). The **common-mode rejection ratio** (CMRR) is defined as

$$\text{CMRR} = \frac{G_d(f)}{G_c(f)} \quad (20.7)$$

and is usually expressed in decibels ($\{\text{CMRR}\}_{\text{dB}} = 20 \log \text{CMRR}$). The input error voltage will be

$$\frac{v_o(v_c)}{G_d} = \frac{G_c v_c}{G_d} = \frac{v_c}{\text{CMRR}} \quad (20.8)$$

In the above analysis we have assumed $Z_c \ll R_o$; otherwise, if there were any unbalance (such as that for the source impedance in [Figure 20.5](#)), v_c at the voltage source would result in a differential-mode voltage at the amplifier input,

$$\begin{aligned} v_d(v_c) &= v_c \left(\frac{R_o + \Delta R_o}{Z_c + R_o + \Delta R_o} - \frac{R_o}{Z_c + R_o} \right) \\ &= v_c \frac{Z_c \Delta R_o}{(Z_c + R_o + \Delta R_o)(Z_c + R_o)} \approx v_c \frac{\Delta R_o}{Z_c} \end{aligned} \quad (20.9)$$

which would be amplified by G_d . Then, the effective common-mode rejection ratio would be

$$\frac{1}{\text{CMRR}_e} = \frac{\Delta R_o}{Z_c} + \frac{1}{\text{CMRR}} \quad (20.10)$$

where the CMRR is that of the IA alone, expressed as a fraction, not in decibels. Stray capacitances from input terminals to ground will decrease Z_c , therefore reducing CMRR_e .

The ideal amplifier is unaffected by power supply fluctuations. The practical amplifier shows output fluctuations when supply voltages change. For slow changes, the equivalent input error can be expressed as a change in input offset voltages in terms of the *power supply rejection ratio* (PSRR),

$$\text{PSRR} = \frac{\Delta V_{os}}{\Delta V_s} \quad (20.11)$$

The terms in Equation 20.5 can be detailed as follows. Because of gain errors we have

$$v_a = v_d \left(G + e_G + \frac{\Delta G}{\Delta T} \times \Delta T + e_{\text{NLG}} \right) \quad (20.12)$$

where G is the differential gain designed, e_G its absolute error, $\Delta G/\Delta T$ its thermal drift, ΔT the difference between the actual temperature and that at which the gain G is specified, and e_{NLG} is the nonlinearity gain error, which describes the extent to which the input/output relationship deviates from a straight line (insert in [Figure 20.4b](#)). The actual temperature T_J is calculated by adding to the current ambient temperature T_A the temperature rise produced by the power P_D dissipated in the device. This rise depends on the thermal resistance θ_{JA} for the case

$$T_J = T_A + P_D \times \theta_{JA} \quad (20.13)$$

where P_D can be calculated from the respective voltage and current supplies

$$P_D = |V_{s+}| |I_{s+}| + |V_{s-}| |I_{s-}| \quad (20.14)$$

The terms for the equivalent *input offset error* will be

$$v_{os} = V_{os}(T_a) + \frac{\Delta V_{os}}{\Delta T} \times (T_J - T_a) \quad (20.15)$$

$$v_b = (I_{B+} - I_{B-})R_o + I_{B+}\Delta R_o = I_{os}R_o + I_B\Delta R_o \quad (20.16)$$

where T_a is the ambient temperature in data sheets, $I_{os} = I_{B+} - I_{B-}$ is the offset current, $I_B = (I_{B+} + I_{B-})/2$, and all input currents must be calculated at the actual temperature,

$$I = I(T_a) + \frac{\Delta I}{\Delta T} \times (T_j - T_a) \quad (20.17)$$

Error contributions from finite *interference rejection* are

$$v_r = \frac{v_c}{\text{CMRR}_c} + \frac{\Delta V_s}{\text{PSRR}} \quad (20.18)$$

where the CMRR_c must be that at the frequency for v_c , and the PSRR must be that for the frequency of the ripple ΔV_s . It is assumed that both frequencies fall inside the bandpass for the signal of interest v_d .

The equivalent *input voltage noise* is

$$v_n = \sqrt{e_n^2 B_e + i_n^2 - R_o^2 B_{i+} + i_n^2 - R_o^2 B_{i-}} \quad (20.19)$$

where e_n^2 is the voltage noise power spectral density of the IA, i_{n+}^2 and i_{n-}^2 are the current noise power spectral densities for each input of the IA, and B_e , B_{i+} , and B_{i-} are the respective noise equivalent bandwidths of each noise source. In Figure 20.5, the transfer function for each noise source is the same as that of the signal v_d . If the signal bandwidth is determined as $f_h - f_l$ by sharp filters, then

$$B_e = f_h - f_l + f_{ce} \ln \frac{f_h}{f_l} \quad (20.20)$$

$$B_{i+} = B_{i-} = f_h - f_l + f_{ci} \ln \frac{f_h}{f_l} \quad (20.21)$$

where f_{ce} and f_{ci} are, respectively, the frequencies where the value of voltage and current noise spectral densities is twice their value at high frequency, also known as corner or 3 dB frequencies.

Another noise specification method states the peak-to-peak noise at a given low-frequency band (f_A to f_B), usually 0.1 to 10 Hz, and the noise spectral density at a frequency at which it is already constant, normally 1 or 10 kHz. In these cases, if the contribution from noise currents is negligible, the equivalent input voltage noise can be calculated from

$$v_n = \sqrt{v_{nL}^2 + v_{nH}^2} \quad (20.22)$$

where v_{nL} and v_{nH} are, respectively, the voltage noise in the low-frequency and high-frequency bands expressed in the same units (peak-to-peak or rms voltages). To convert rms voltages into peak-to-peak values, multiply by 6.6. If the signal bandwidth is from f_l to f_h , and $f_l = f_A$ and $f_h > f_B$, then Equation 20.22 can be written

$$v_n = \sqrt{v_{nL}^2 + (6.6e_n)^2 (f_h - f_B)} \quad (20.23)$$

where v_{nL} is the peak-to-peak value and e_n is the rms voltage noise as specified in data books. Equation 20.23 results in a peak-to-peak calculated noise that is lower than the real noise, because noise spectral density is not constant from f_B up. However, it is a simple approach providing useful results.

For signal sources with high output resistors, thermal and excess noise from resistors (see [Chapter 18](#)) must be included. For first- and second-order filters, noise bandwidth is slightly larger than signal bandwidth. Motchenbacher and Connelly [4] show how to calculate noise bandwidth, resistor noise, and noise transfer functions when different from signal transfer functions.

Low-noise design always seeks the minimal bandwidth required for the signal. When amplifying low-frequency signals, if a large capacitor C_i is connected across the input terminals in [Figure 20.5](#), then noise and interference having a frequency larger than $f_0 = 1/2\pi(2R_o)C_i$ ($f_0 \ll f_s$) will be attenuated.

Another possible source of error for any IA, not included in Equation 20.5, is the *slew rate limit* of its output stage. Because of the limited current available, the voltage at the output terminal cannot change faster than a specified value SR. Then, if the maximal amplitude A of an output sine wave of frequency f exceeds

$$A = \frac{SR}{2\pi f} \quad (20.24)$$

there will be a waveform distortion.

[Table 20.1](#) lists some basic specifications for IC instrumentation amplifiers whose gain G can be set by an external resistor or a single connection.

Instrumentation Amplifiers Built from Discrete Parts

Instrumentation amplifiers can be built from discrete parts by using operational amplifiers (op amps) and a few resistors. An *op amp* is basically a differential voltage amplifier whose gain A_d is very large (from 10^5 to 10^7) at dc and rolls off (20 dB/decade) from frequencies of about 1 to 100 Hz, becoming 1 at frequencies from 1 to 10 MHz for common models ([Figure 20.6a](#)), and whose input impedances are so high (up to $10^{12} \Omega \parallel 1 \text{ pF}$) that input currents are almost negligible. Op amps can also be modeled by the circuit in [Figure 20.5](#), and their symbol is that in [Figure 20.4a](#), deleting IA. However, because of their large gain, op amps cannot be used directly as amplifiers; a mere 1 mV dc input voltage would saturate any op amp output. Furthermore, op amp gain changes from unit to unit, even for the same model, and for a given unit it changes with time, temperature, and supply voltages. Nevertheless, by providing external feedback, op amps are very flexible and far cheaper than IAs. But when the cost for external components and their connections, and overall reliability are also considered, the optimal solution depends on the situation.

[Figure 20.6b](#) shows an amplifier built from an op amp with external feedback. If input currents are neglected, the current through R_2 will flow through R_1 and we have

$$v_d = v_s - v_o \frac{R_1}{R_1 + R_2} \quad (20.25)$$

$$v_o = A_d v_d \quad (20.26)$$

Therefore,

$$\frac{v_o}{v_s} = \frac{A_d \left(1 + \frac{R_2}{R_1} \right)}{A_d + 1 + \frac{R_2}{R_1}} = \frac{G_i}{1 + \frac{G_i}{A_d}} \quad (20.27)$$

where $G_i = 1 + R_2/R_1$ is the ideal gain for the amplifier. If G_i/A_d is small enough (G_i small, A_d large), the gain does not depend on A_d but only on external components. At high frequencies, however, A_d becomes smaller and, from Equation 20.27, $v_o < G_i v_s$ so that the bandwidth for the amplifier will reduce for large gains. Franco [5] analyzes different op amp circuits useful for signal conditioning.

TABLE 20.1 Basic Specifications for Some Instrumentation Amplifiers

	AD624A	AMP02F	INA110KP	LT1101AC	Units
Gain range	1–1000	1–1000	1–500	10,100	V/V
Gain error, e_G					
$G = 1$	± 0.05	0.05	± 0.02	n.a.	%
$G = 10$	n.s.	0.40	± 0.05	± 0.04	%
$G = 100$	± 0.25	0.50	± 0.10	± 0.04	%
$G = 1000$	± 1.0	0.70	n.a.	n.a.	%
Gain nonlinearity error e_{NLG}^a					
$G = 1$	± 0.005	0.006	± 0.005	n.a.	%
$G = 10$	n.s.	0.006	± 0.005	± 0.0008	%
$G = 100$	± 0.005	0.006	± 0.01	± 0.0008	
$G = 1000$	± 0.005	0.006	n.a.	n.a.	%
Gain drift $\Delta G/\Delta T$					
$G = 1$	5	50	± 10	n.a.	$\mu\text{V/V}/^\circ\text{C}$
$G = 10$	n.s.	50	± 10	5	$\mu\text{V/V}/^\circ\text{C}$
$G = 100$	10	50	± 20	5	$\mu\text{V/V}/^\circ\text{C}$
$G = 1000$	25	50	n.a.	n.a.	$\mu\text{V/V}/^\circ\text{C}$
V_{os}	$200 + 5/G$	200	$\pm(1000 + 5000/G)$	160	μV
$\Delta V_{os}/\Delta T$	$2 + 50/G$	4	$\pm(2 + 50/G)$	2	$\mu\text{V}/^\circ\text{C}$
I_B	± 50	20	0.05	10	nA
$\Delta I_B/\Delta T$	± 50 typ	250 typ	b	30	$\text{pA}/^\circ\text{C}$
I_{os}	± 35	10	0.025	0.90	nA
$\Delta I_{os}/\Delta T$	± 20 typ	15 typ	n.s.	7.0	$\text{pA}/^\circ\text{C}$
Z_d	$1 \parallel 10$ typ	10 typ	$5000 \parallel 6$ typ	12	$\text{G}\Omega$
Z_c	$1 \parallel 10$ typ	16.5 typ	$2000 \parallel 1$ typ	7	$\text{G}\Omega$
CMRR at dc					
$G = 1$	70 min	80 min	70 min	n.a.	dB
$G = 10$	n.s.	100 min	87 min	82	dB
$G = 100$	100 min	115 min	100 min	98	dB
$G = 1000$	110 min	115 min	n.a.	n.a.	dB
PSRR at dc					
$G = 1$	70 min	80 min	c	n.a.	dB
$G = 10$	n.s.	100 min	c	100	dB
$G = 100$	95 min	115 min	c	100	dB
$G = 1000$	100 min	115 min	n.a.	n.a.	dB
Bandwidth (–3 dB) (typ)					
$G = 1$	1000	1200	2500	n.a.	kHz
$G = 10$	n.s.	300	2500	37	kHz
$G = 100$	150	200	470	3.5	kHz
$G = 1000$	25	200	n.a.	n.a.	kHz
Slew rate (typ)	5.0	6	17	0.1	$\text{V}/\mu\text{s}$
Settling time to 0.01%					
$G = 1$	15 typ	10 typ	12.5	n.a.	μs
$G = 10$	15 typ	10 typ	7.5	n.a.	μs
$G = 100$	15 typ	10 typ	7.5	n.a.	μs
$G = 1000$	75 typ	10 typ	n.a.	n.a.	μs
e_n (typ)					
$G = 1$	4	120	66	n.a.	$\text{nV}/\sqrt{\text{Hz}}$
$G = 10$	4	18	12	43	$\text{nV}/\sqrt{\text{Hz}}$
$G = 100$	4	10	10	43	$\text{nV}/\sqrt{\text{Hz}}$
$G = 1000$	4	9	n.a.	n.a.	$\text{nV}/\sqrt{\text{Hz}}$
v_n 0.1 to 10 Hz (typ)					
$G = 1$	10	10	1	0.9	$\mu\text{Vp-p}$
$G = 10$	n.s.	1.2	1	0.9	$\mu\text{Vp-p}$
$G = 100$	0.3	0.5	1	0.9	$\mu\text{Vp-p}$
$G = 1000$	0.2	0.4	1	0.9	$\mu\text{Vp-p}$

TABLE 20.1 (continued) Basic Specifications for Some Instrumentation Amplifiers

	AD624A	AMP02F	INA110KP	LT1101AC	Units
i_n 0.1 to 10 Hz (typ)	60	n.s.	n.s.	2.3	pAp-p
i_n (typ)	n.s.	400	1.8	20	fA/ \sqrt{Hz}

Note: All parameter values are maximum, unless otherwise stated (typ = typical; min = minimum; n.a. = not applicable; n.s. = not specified). Measurement conditions are similar; consult manufacturers' data books for further detail.

^a For the INA110, the gain nonlinearity error is specified as percentage of the full-scale output.

^b Input current drift for the INA110KP approximately doubles for every 10°C increase, from 25°C (10 pA-typ) to 125°C (10 nA-typ).

^c The PSRR for the INA110 is specified as an input offset $\pm(10 + 180/G)$ μ V/V maximum.

Figure 20.7 shows an IA built from three op amps. The input stage is fully differential and the output stage is a *difference amplifier* converting a differential voltage into a single-ended output voltage. Difference amplifiers (op amp and matched resistors) are available in IC form: AMP 03 (Analog Devices) and INA 105/6 and INA 117 (Burr-Brown). The gain equation for the complete IA is

$$G = \left(1 + 2 \frac{R_2}{R_1} \right) \frac{R_4}{R_3} \quad (20.28)$$

Pallás-Areny and Webster [6] have analyzed matching conditions in order to achieve a high CMRR. Resistors R_2 do not need to be matched. Resistors R_3 and R_4 need to be closely matched. A potentiometer connected to the v_{ref} terminal makes it possible to trim the CMRR at low frequencies.

The *three-op-amp* IA has a symmetrical structure making it easy to design and test. IAs based on an IC difference amplifier do not need any user trim for high CMRR. The circuit in Figure 20.8 is an IA that lacks these advantages but uses only two op amps. Its gain equation is

$$G = 1 + \frac{R_2}{R_1} + \frac{2R_2}{R_G} \quad (20.29)$$

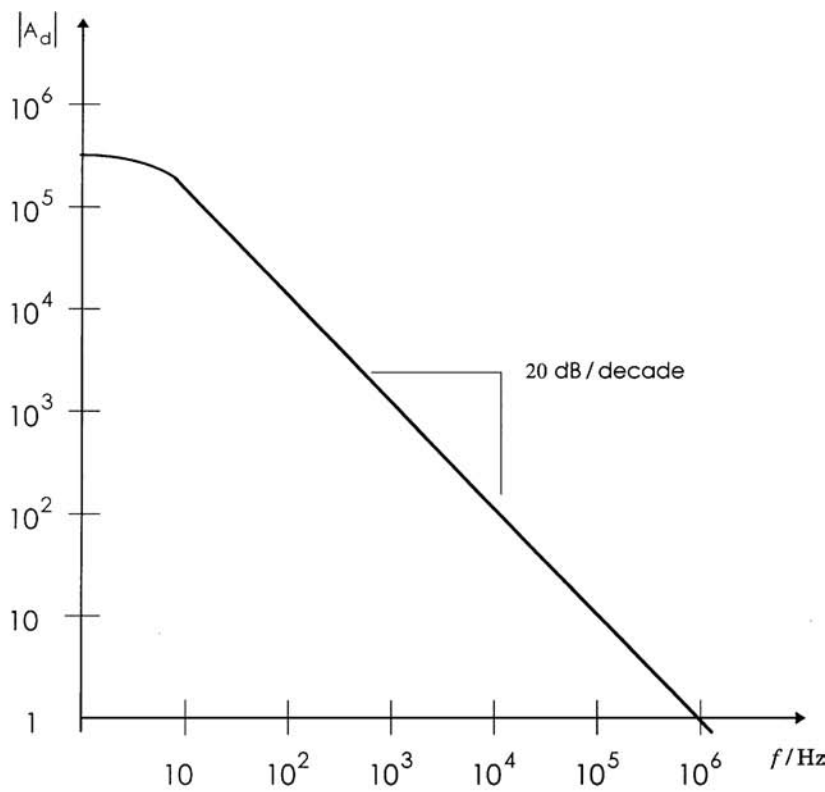
R_1 and R_2 must be matched and R_G should be comparable to R_2 .

Another approach to build an IA is by the *switched-capacitor* technique (Figure 20.9). Switches SW1 and SW2 close together and charge C_S (1 μ F) to the voltage difference $v_H - v_L$; next, SW1 and SW2 open and SW3 and SW4 close, so that C_H (0.1 to 1 μ F) also charges to $v_H - v_L$. Then SW1 and SW2 close again, SW3 and SW4 open, and so on. While C_S is being charged C_H holds the previous voltage difference. Therefore, the maximal frequency for the input signal must be at least ten times lower than the switching frequency. This circuit has a high CMRR because the charge at C_S is almost insensitive to the input common-mode voltage. Furthermore, it converts the differential signal to a single-ended voltage. The LTC 1043 (Linear Technology) includes two sets of four switches to implement this circuit.

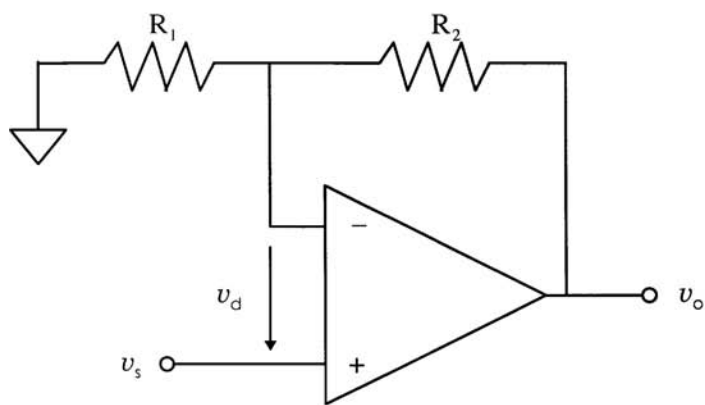
Composite Instrumentation Amplifiers

Instrumentation amplifiers have a very limited bandwidth. They achieve a gain of 10 at 2.5 MHz, at best. Moreover, their inputs must be either dc-coupled or, if ac-coupled with input series capacitors, there must be a path for bias currents; if that path is a resistor from each input to ground, then the common-mode input impedance Z_c decreases and noise may increase.

A larger bandwidth for a given gain can be obtained by cascade connection of two or more amplifiers. However, if the additional gain is provided by a single-ended amplifier after the IA, then the overall CMRR is that of the IA, which is small at high frequencies. The circuit in Figure 20.10a is a broadband IA with large CMRR because the CMRR for the second stage is multiplied by the differential gain for the first stage, which can be very high if implemented by broadband op amps. The overall gain is



(a)



(b)

FIGURE 20.6 (a) Open loop gain for an op amp. (b) Amplifier based on an op amp with external feedback.

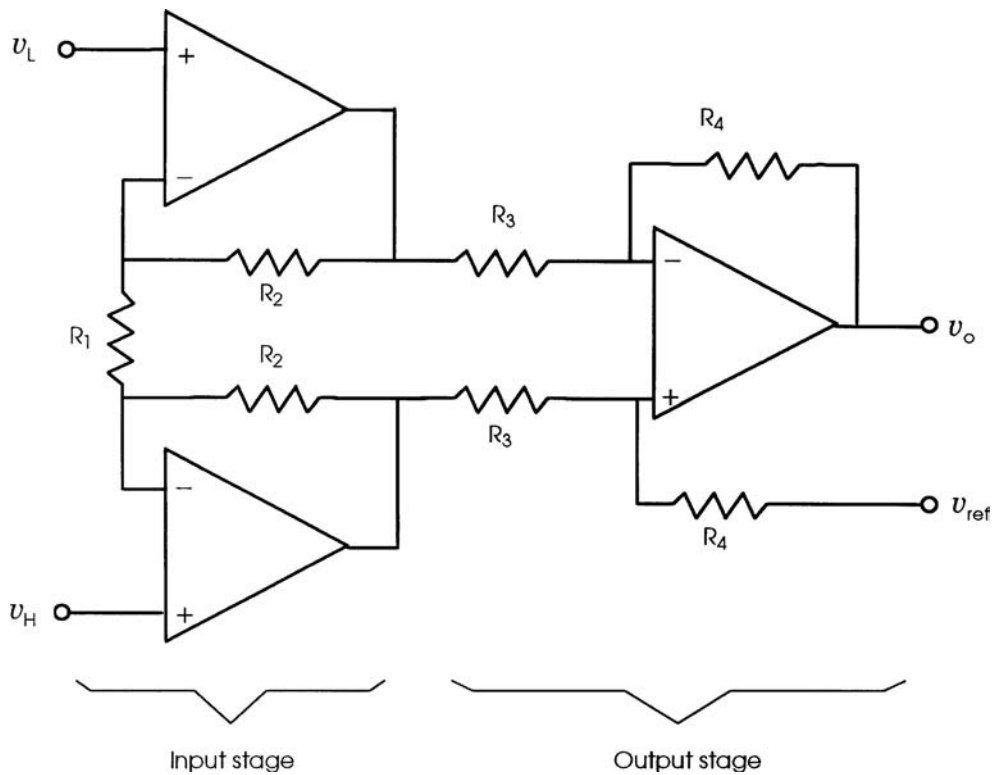


FIGURE 20.7 Instrumentation amplifier built from three op amps. R_3 and R_4 must be matched.

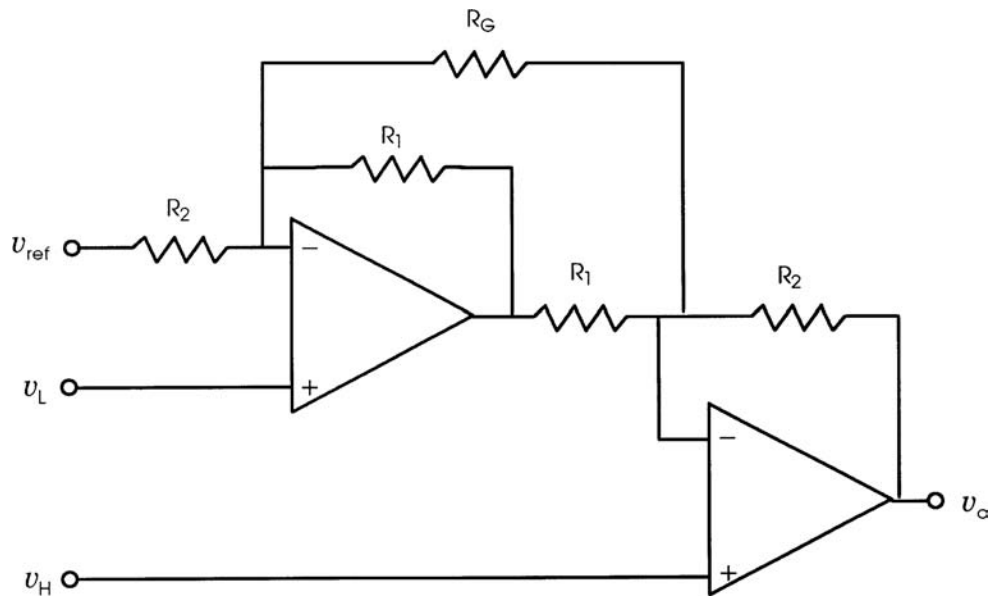


FIGURE 20.8 Instrumentation amplifier built from two op amps. R_1 and R_2 must be matched.

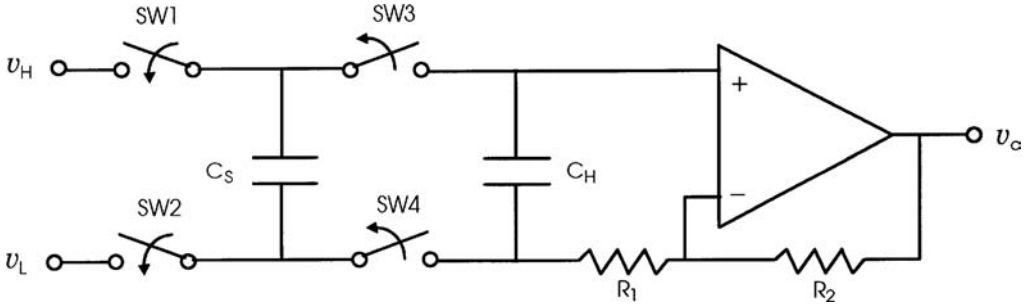


FIGURE 20.9 Instrumentation amplifier based on the switched-capacitor technique. First switches SW1 and SW2 close while SW3 and SW4 are open, and C_S charges to $v_H - v_L$. Then SW1 and SW2 open and SW3 and SW4 close, charging C_H to $v_H - v_L$.

$$G = G_1 G_2 = \left(1 + \frac{2R_b}{R_a} \right) G_{IA} \quad (20.30)$$

An IA can be *ac-coupled* by feeding back its dc output to the reference terminal as shown in [Figure 20.10b](#). The high-pass corner frequency is $f_0 = 1/(2\pi R_0 C_0)$.

20.6 Single-Ended Signal Conditioners

Floating signals (single ended or differential) can be connected to amplifiers with single-ended grounded input. Grounded single-ended can be connected to *single-ended amplifiers*, provided the difference in ground potentials from signal to amplifier is not too large. [Figure 20.11a](#) shows a simple single-ended amplifier based on an IA. However, op amps are better suited than IAs for single-ended amplifiers and signal conditioners performing additional functions.

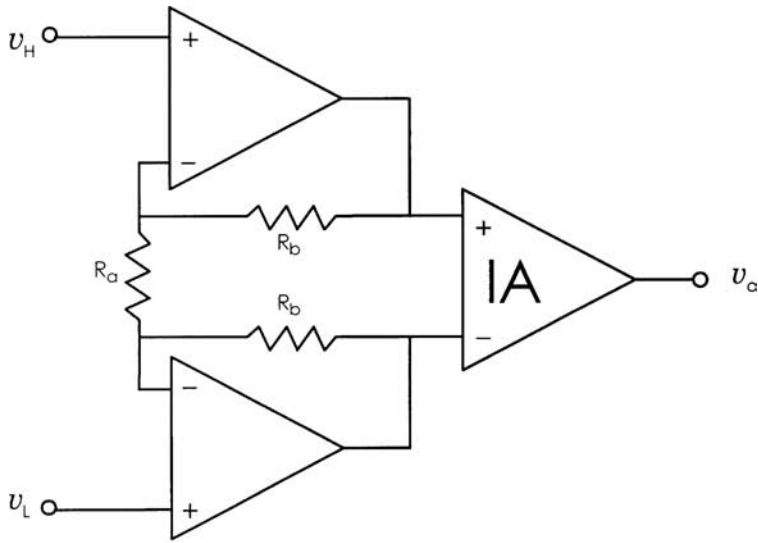
[Figure 20.11b](#) shows an *inverting amplifier* whose gain is $G = -R_2/R_1$, and whose input impedance is R_1 . The capacitor on the dashed line (10 pF or larger) prevents gain peaking and oscillation. If a capacitor C replaces R_2 , input signals are integrated and inverted. If C replaces R_1 instead, input signals are differentiated and inverted. The circuit in [Figure 20.11c](#) has $G = 1$ for dc and signals of low frequency relative to $f_1 = 1/(2\pi R_1 C_1)$ (offset and drift included) and $G = 1 + R_2/R_1$ for high-frequency signals. The circuit in [Figure 20.11d](#) calculates the average for n voltages. The difference between two voltages can be obtained from the difference amplifier in [Figure 20.7](#) (output stage).

Op amps must be carefully selected according to the application. For dc circuits, chopper op amps offer the best performance. For low-impedance signals, op amps with bipolar input transistors are better. For high-impedance signals, op amps with FET input transistors offer lower input currents, but they have larger drift and voltage noise. [Table 20.2](#) lists some parameters for several different op amps. Some manufacturers provide selection guides on floppy disk which suggest the most appropriate model for a set of user-defined values for some parameters.

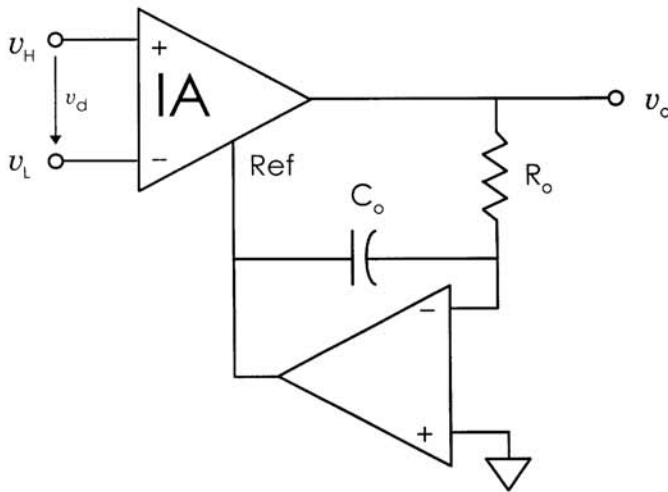
20.7 Carrier Amplifiers

A **carrier amplifier** is a conditioner for extremely narrowband ac signals from ac-driven sensors. A carrier amplifier is made of a sine wave oscillator, to excite the sensor bridge, an ac voltage amplifier for the bridge output, a synchronous demodulator (see [Chapter 24](#)), and a low-pass filter ([Figure 20.12](#)). The NE5520/1 (Philips) are carrier amplifiers in IC form intended for (but not limited to) LVDTs driven at a frequency from 1 to 20 kHz.

Carrier amplifiers make it possible to recover the amplitude and phase of the modulating signal after amplifying the output modulated waveform from the bridge. This is useful first because ac amplifiers



(a)



(b)

FIGURE 20.10 Composite instrumentation amplifiers. (a) Broadband IA with large CMRR; (b) ac-coupled IA.

are not affected by offset, drift, or low-frequency noise, and therefore the bridge output can easily be amplified. Second, the *phase-sensitive demodulator* yields not only the amplitude but also the sign of the measurand. If the measurement range includes positive and negative values for the measurand, phase detection is essential.

A further advantage of carrier amplifiers is their extremely narrow frequency response, determined by the output low-pass filter. In the demodulator, the product of the modulated carrier of frequency f_c by the reference signal, also of frequency f_c , results in a baseband component and components at nf_c ($n \geq 2$). The output low-pass filter rejects components other than the baseband. If the corner frequency for this

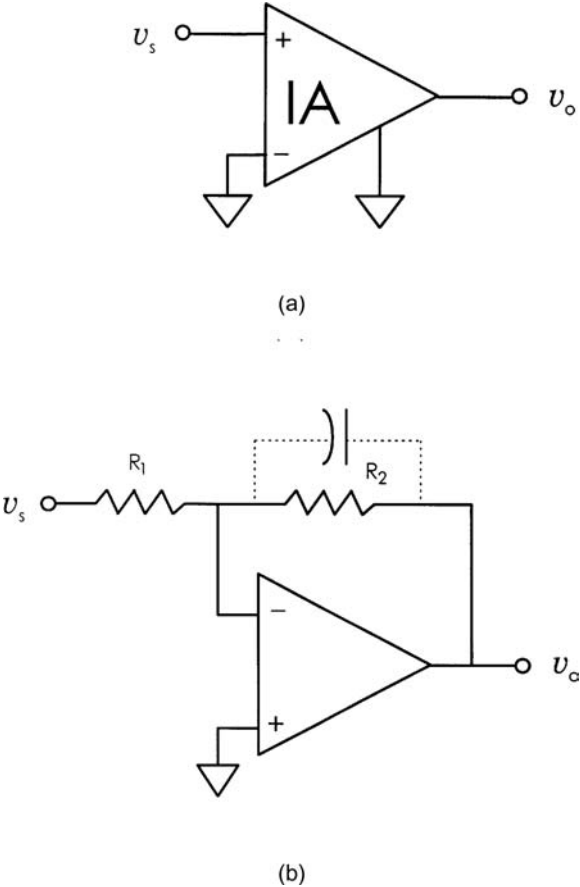


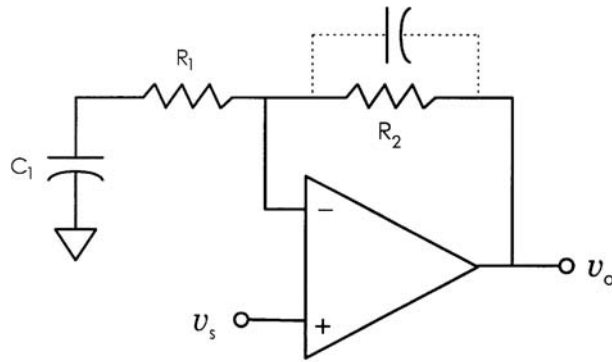
FIGURE 20.11 Single-ended amplifiers and signal conditioners. (a) Noninverting amplifier based on an IA; (b) inverting amplifier based on an op amp; (c) ac amplifier; (d) voltage averager.

filter is f_0 , then the passband for the system is $f_c \pm f_0$. Therefore, any interference of frequency f_i added to the modulated signal will be rejected if falling outside that passband. The ability to discriminate signals of interest from those added to them is described by the *series (or normal) mode rejection ratio* (SMRR), and is usually expressed in decibels. In the present case, using a first-order low-pass filter we have

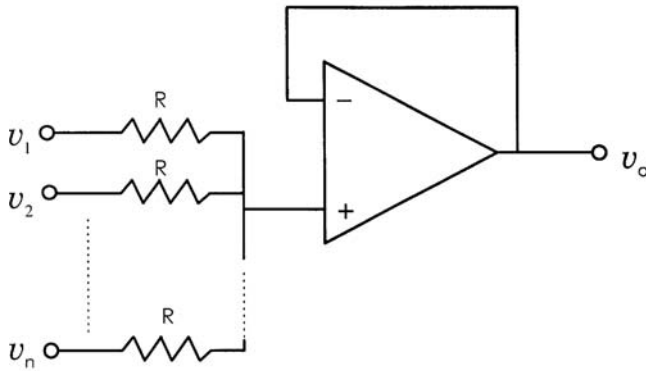
$$\text{SMRR} = 20 \log \frac{v_o(f_c)}{v_o(f_i)} = 20 \log \frac{\sqrt{1 + (f_c - f_i)^2}}{f_0} \approx 20 \log \frac{|f_c - f_i|}{f_0} \tag{20.31}$$

A power-line interference superimposed on a 10 kHz carrier will undergo an 80-dB attenuation if the output low-pass filter has $f_0 = 1$ Hz. The same interference superimposed on the baseband signal would be attenuated by only 35 dB.

Carrier amplifiers can be built from a precision sine wave oscillator — AD2S99 (Analog Devices), 4423 (Burr-Brown), SWR300 (Thaler) — or a discrete-part oscillator, and a demodulator (plus the output filter). Some IC demodulators are based on switched amplifiers (AD630, OPA676). The floating capacitor in Figure 20.9 behaves as a synchronous demodulator if the switch clock is synchronous with the carrier, and its duty cycle is small (less than 10%), so that switches SW1 and SW2 sample the incoming modulated waveform for a very short time [7].



(c)



(d)

FIGURE 20.11 (continued)

20.8 Lock-In Amplifiers

A *lock-in amplifier* is based on the same principle as a carrier amplifier, but instead of driving the sensor, here the carrier signal drives the experiment, so that the measurand is frequency translated. Lock-in amplifiers are manufactured as equipment intended for recovering signals immersed in high (asynchronous) noise. These amplifiers provide a range of driving frequencies and bandwidths for the output filter. Some models are vectorial because they make it possible to recover the in-phase and quadrature (90° out-of-phase) components of the incoming signal, by using two demodulators whose reference signals are delayed by 90° . Still other models use bandpass filters for the modulated signal and two demodulating stages. Meade [8] analyzes the fundamentals, specifications, and applications of some commercial lock-in amplifiers.

20.9 Isolation Amplifiers

The maximal common-mode voltage withstood by common amplifiers is smaller than their supply voltage range and seldom exceeds 10 V. Exceptions are the INA 117 (Burr-Brown) and similar difference amplifiers whose common-mode range is up to ± 200 V, and the IA in Figure 20.9 when implemented by high-voltage

TABLE 20.2 Basic Specifications for Operational Amplifiers of Different Technologies

	$V_{os}, \mu\text{V}$	$(\Delta v_{os}/\Delta T)_{av}, \mu\text{V}/^\circ\text{C}$	I_B, pA	$\Delta I_B/\Delta T, \text{pA}/^\circ\text{C}$	I_{os}, pA	$BW_{typ}(G = 1), \text{MHz}$	$e_n(1 \text{ kHz}), \text{nV}/\sqrt{\text{Hz}}$	f_{ce}, Hz	$v_{n(p-p)}, \mu\text{V}$	$i_n(1 \text{ kHz}), \text{fA}/\sqrt{\text{Hz}}$
Bipolar										
$\mu\text{A}741$	6000	15	500000	500	200000	1.5	20	200	—	550
LM358A	3000	20	100000	—	± 30000	1	—	—	—	—
LT1028	80	0.8	180000	—	100000	75	0.9	3.5	0.035	1000
OP07	75	1.3	3000	50	2800	0.6	9.6	10	0.35	170
OP27C	100	1.8	80000	—	75000	8	3.2	2.7	0.09	400
OP77A	25	0.3	2000	25	1500	0.6	9.6	10	0.35	170
OP177A	10	0.1	1500	25	1000	0.6	—	—	0.8	—
TLE2021C	600	2	70000	80	3000	1.2	30	—	0.47	90
TLE2027C	100	1	90000	—	90000	13	2.5	—	0.05	400
FET Input										
AD549K	250	5	0.1	b	0.03 typ	1	35	—	4	0.16
LF356A	2000	5	50	b	10	4.5	12	—	—	10
OPA111B	250	1	1	b	0.75	2	7	200	1.2	0.4
OPA128J	1000	20	0.3	b	65	1	27	—	4	0.22
TL071C	10000	18	200	b	100	3	18	300	4	10
TLE2061C	3000	6	4 typ	b	2 tip	2	60	20	1.2	1
CMOS										
ICL7611A	2	10 typ	50	b	30	0.044	100	800	—	10
LMC660C	6000	1.3 typ	20	b	20	1.4	22	—	—	0.2
LMC6001A	350	10	0.025	b	0.005	1.3	22	—	—	0.13
TLC271CP	10000	2 typ	0.7 typ	c	0.1 typ	2.2	25	100	—	n.s.
TLC2201C	500	0.1 typ	1 typ	d	0.5 typ	1.8	8	—	0.7	0.6
BiMOS										
CA3140	15000	8	50	b	30	4.5	40	—	—	—
CMOS Chopper										
LTC1052	5	0.05	30	e	30	1.2	—	—	1.5	0.6
LTC1150C	5	0.05	100	f	200	2.5	—	—	1.8	1.8
MAX430C	10	0.05	100	g	200	0.5	—	—	1.1	10
TLC2652AC	1	0.03	4 typ	d	2 typ	1.9	23	—	2.8	4
TLC2654C	20	0.3	50 typ	0.65	30 typ	1.9	13	—	1.5	4
TSC911A	15	0.15	70	—	20	1.5	—	—	11	—

Specified values are maximal unless otherwise stated and those for noise, which are typical (typ = typical, av = average; nonspecified parameters are indicated by a dash).

^a Values estimated from graphs.

^b I_B doubles every 10°C .

^c I_B doubles every 7.25°C .

^d I_B is almost constant up to 85°C .

^e I_B is almost constant up to 75°C .

^f I_{B+} and I_{B-} show a different behavior with temperature.

^g I_B doubles every 10°C above about 65°C .

switches (relays, optorelays). Signals with large off-ground voltages, or differences in ground potentials exceeding the input common-mode range, result in permanent amplifier damage or destruction, and a safety risk, in spite of an exceptional CMRR: a 100 V common-mode 60 Hz voltage at the input of a common IA having a 120 dB CMRR at power-line frequency does not result in a $100 \text{ V}/10^6 = 100 \mu\text{V}$ output, but a burned-out IA.

Figure 20.13a shows a signal source grounded at a point far from the amplifier ground. The difference in voltage between grounds v_i not only thwarts signal measurements but can destroy the amplifier. The

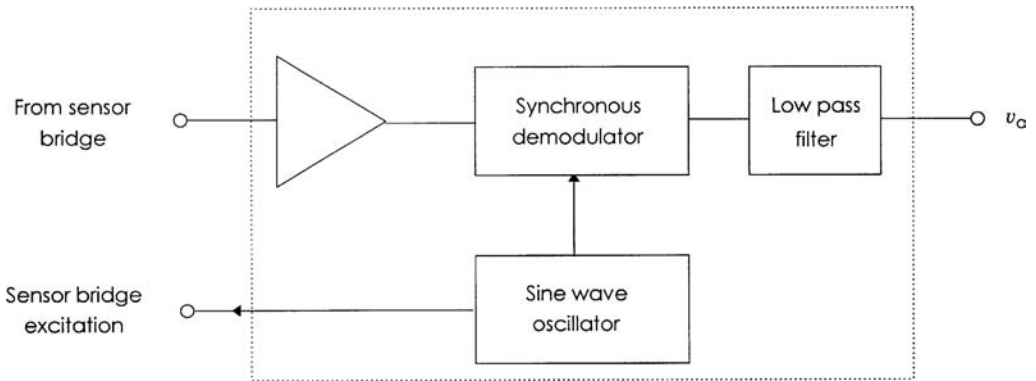


FIGURE 20.12 Elements for a carrier amplifier.

solution is to prevent this voltage from forcing any large current through the circuit and at the same time to provide an information link between the source and the amplifier. Figure 20.13b shows a solution: the signal source and the amplifier have separated (isolated) power supplies and the signal is coupled to the amplifier through a transformer acting as an isolation barrier for v_i . Other possible barriers are optocouplers (IL300-Siemens) and series capacitors (LTC1145-Linear Technology). Those barriers impose a large series impedance (isolation impedance, Z_i) but do not usually have a good low-frequency response, hence the need to modulate and then demodulate the signal to transfer through it. The subsystem made of the modulator and demodulator, plus sometimes an input and an output amplifier and a dc-dc converter for the separate power supply, is called an **isolation amplifier**. The ability to reject the voltage difference across the barrier (isolation-mode voltage, v_i) is described by the **isolation mode rejection ratio** (IMRR), expressed in decibels,

$$\text{IMRR} = 20 \log \frac{\text{OUTPUT Voltage}}{\text{ISOLATION-MODE Voltage}} \quad (20.32)$$

Ground isolation also protects people and equipment from contact with high voltage because Z_i limits the maximal current. Some commercial isolation amplifiers are the AD202, AD204, and AD210 (Analog Devices) and the ISOxxx series (Burr-Brown).

Table 20.3 summarizes the compatibility between signal sources and amplifiers. When grounded, amplifiers and signals are assumed to be grounded at different physical points.

20.10 Nonlinear Signal-Processing Techniques

Limiting and Clipping

Clippers or *amplitude limiters* are circuits whose output voltage has an excursion range restricted to values lower than saturation voltages. Limiting is a useful signal processing technique for signals having the information encoded in parameters other than the amplitude. For example, amplitude limiting is convenient before homodyne phase demodulators. *Limiting* can also match output signals levels to those required for TTL circuits (0 to 5 V). Limiting avoids amplifier saturation for large input signal excursions, which would result in a long recovery time before returning to linear operation.

Limiting can be achieved by op amps with diodes and zeners in a feedback loop. Figure 20.14a shows a positive voltage clipper. When v_s is positive, v_o is negative and R_2/R_1 times larger; the diode is reverse biased and the additional feedback loop does not affect the amplifier operation. When v_s is negative and large enough, v_o forward biases the diode (voltage drop V_f) and the zener clamps at V_z , the output amplitude thus being limited to $v_o = V_f + V_z$ until $|v_s| < (V_f + V_z)R_1/R_2$ (Figure 20.14b). The circuit then acts again as an inverting amplifier until v_s reaches a large negative value.

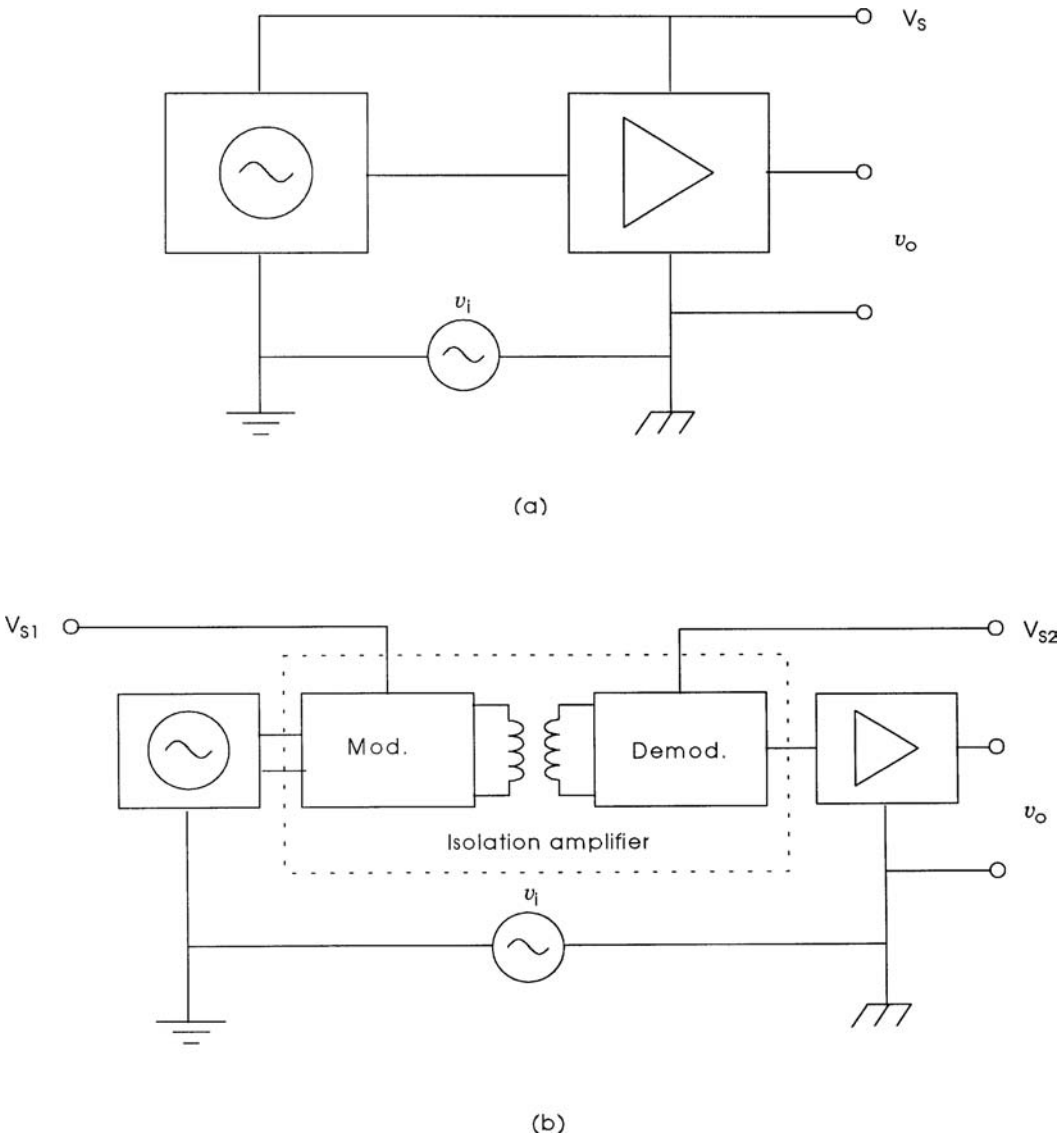


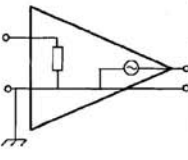
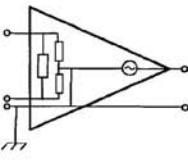
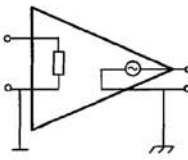
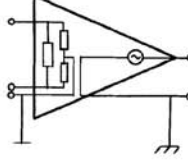


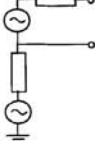
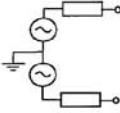
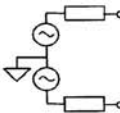
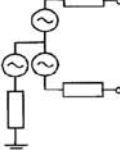
FIGURE 20.13 (a) A large difference in ground potentials damages amplifiers. (b) An isolation amplifier prevents large currents caused by this difference from flowing through the circuit.

A negative voltage clipper can be designed by reversing the polarity of the diode and zener. To limit the voltage in both directions, the diode may be substituted by another zener diode. The output is then limited to $|v_o| < V_{z1} + V_{f2}$ for negative inputs to $|v_o| < V_{f1} + V_{z2}$ for positive inputs. If $V_{z1} = V_{z2}$, then the voltage limits are symmetrical. Jung [9] gives component values for several precision limiters.

Logarithmic Amplification

The dynamic range for common linear amplifiers is from 60 to 80 dB. Sensors such as photodetectors, ionizing radiation detectors, and ultrasound receivers can provide signals with an amplitude range wider than 120 dB. The only way to encompass this wide amplitude range within a narrower range is by amplitude compression. A logarithmic law compresses signals by offering equal-output amplitude changes in response to a given ratio of input amplitude increase. For example, a scaling of 1 V/decade means that the output would change by 1 V when the input changes from 10 to 100 mV, or from 100 mV

TABLE 20.3 Compatibility between Signal Sources and Conditioners

<div>Conditioner input</div> <div>Signal source</div>				
	Incompatible unless grounds are very close	Compatible if CMRR is large	Compatible	Compatible
	Compatible	Compatible	Compatible	Compatible
	Incompatible unless grounds are very close	Compatible if CMRR is large	Compatible for large Z_i	Compatible
	Incompatible	Compatible	Compatible for large Z_i	Compatible
	Compatible	Compatible	Compatible	Compatible
	Incompatible	Compatible if CMRR is large	Compatible for large Z_i	Compatible

Note: When grounded, signals sources and amplifiers are assumed to be grounded at different points. Isolation impedance is assumed to be very high for floating signal sources but finite (Z_i) for conditioners.

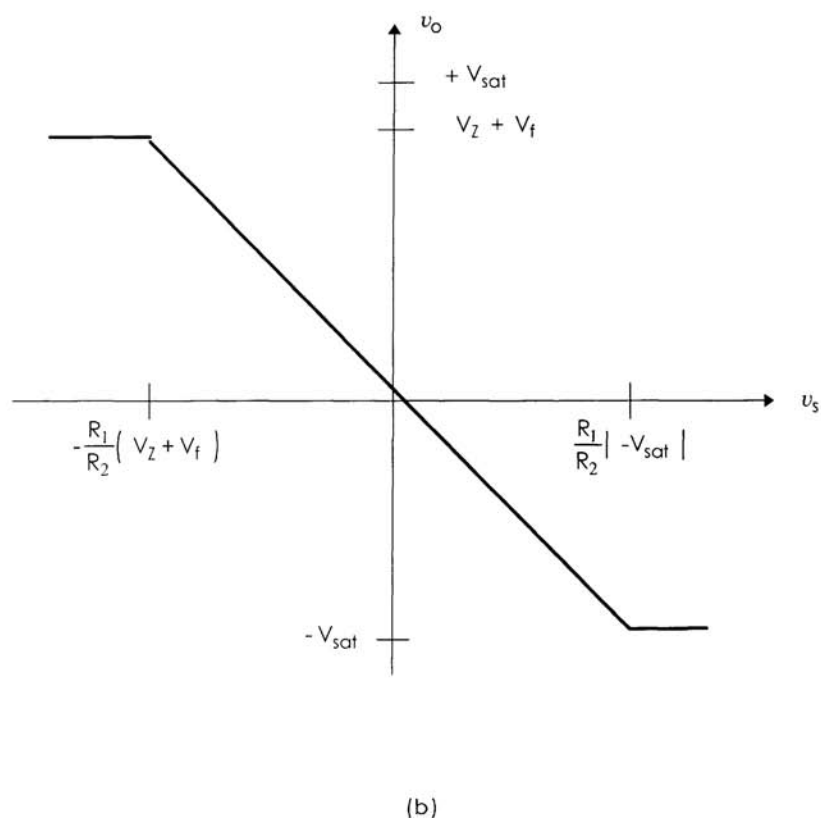
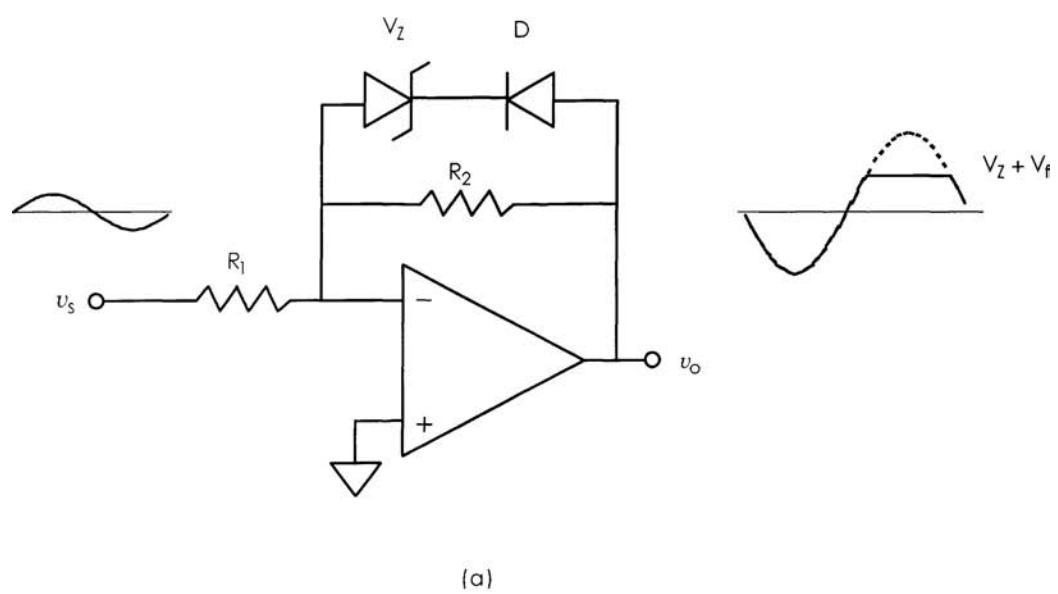
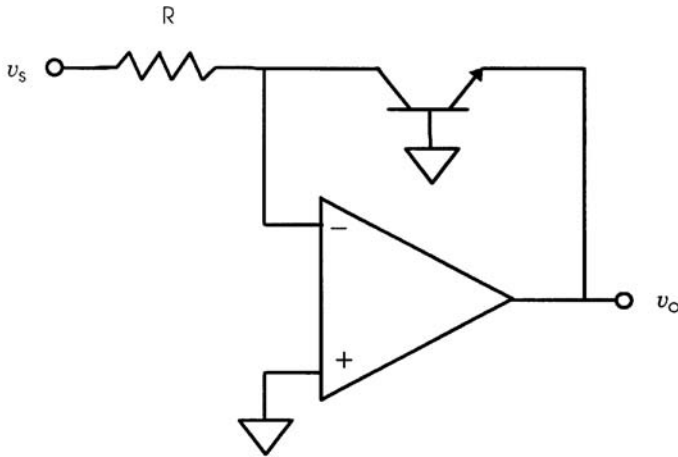
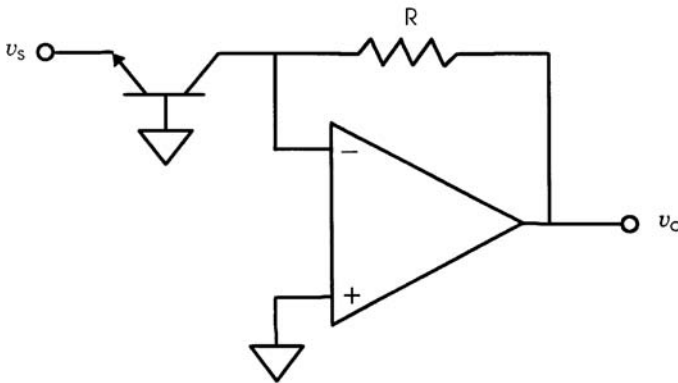


FIGURE 20.14 Voltage limiter. (a) Circuit based on op amp and diode network feedback. (b) Input/output relationship.



(a)



(b)

FIGURE 20.15 Basic circuit for logarithmic (a) and antilog or exponential (b) conversion using the transdiode technique. Practical converters include additional components for error reduction and protection.

to 1 V. Therefore, *logarithmic amplifiers* do not necessarily amplify (enlarge) input signals. They are rather converters providing a voltage or current proportional to the ratio of the input voltage, or current, to a reference voltage, or current.

Logarithmic conversion can be obtained by connecting a bipolar transistor as a feedback element of an op amp, Figure 20.15a. The collector current i_C and the base-emitter voltage have an exponential relationship. From the Ebers–Moll model for a transistor, if $v_{CB} = 0$, then

$$i_C = I_s \left(e^{v_{BE}/v_T} - 1 \right) \quad (20.33)$$

where $v_T = kT/q = 25$ mV at room temperature, and I_s is the saturation current for the transistor. In Figure 20.15a the input voltage is converted into an input current and the op amp forces the collector

current of the transistor to equal the input current, while maintaining $v_{CB} \approx 0$ V. Hence, provided $i_C \gg I_S$, for $v_s > 0$,

$$v_o = \frac{v_T}{\log e} \log \frac{v_s}{RI_s} \quad (20.34)$$

The basic circuit in Figure 20.15a must be modified in order to provide temperature stability, phase compensation, and scale factor correction; reduce bulk resistance error; protect the base-emitter junction; accept negative input voltages and other improvements. Wong and Ott [10] and Peyton and Walsh [11] describe some common circuit techniques to implement these and additional functions. The LOG100 (Burr-Brown) is a logarithmic converter using this so-called transdiode technique. The AD640 (Analog Devices) and TL441 (Texas Instruments) use different techniques.

Figure 20.15b shows a basic *antilog* or *exponential converter* for negative input voltages. The transistor and the resistor have interchanged positions with respect to Figure 20.15b. For $v_s < 0$,

$$v_o = I_S R e^{v_s/v_T} \quad (20.35)$$

Positive voltages require an input *pnp* transistor instead.

Multiplication and Division

Analog multiplication is useful not only for analog computation but also for modulation and demodulation, for voltage-controlled circuits (amplifiers, filters), and for linearization [12]. An *analog multiplier* (Figure 20.16a) has two input ports and one output port offering a voltage

$$v_o = \frac{v_x v_y}{V_m} \quad (20.36)$$

where V_m is a constant voltage. If inputs of either polarity are accepted, and their signs preserved, the device is a *four-quadrant multiplier*. If one input is restricted to have a defined polarity but the other can change sign, the device is a *two-quadrant multiplier*. If both inputs are restricted to only one polarity, the device is a *one-quadrant multiplier*. By connecting both inputs together, we obtain a voltage squarer (Figure 20.16b).

Wong and Ott¹⁰ describe several multiplication techniques. At low frequencies, one-quadrant multipliers can be built by the log–antilog technique, based on the mathematical relationships $\log A + \log B = \log AB$ and then $\text{antilog}(\log AB) = AB$. The AD538 (Analog Devices) uses this technique. Currently, the most common multipliers use the transconductance method, which provides four-quadrant multiplication and differential ports. The AD534, AD633, AD734, AD834/5 (Analog Devices), and the MPY100 and MPY600 (Burr-Brown), are *transconductance multipliers*. A digital-to-analog converter can be considered a multiplier accepting a digital input and an analog input (the reference voltage). A multiplier can be converted into a *divider* by using the method in Figure 20.16c. Input v_x must be positive in order for the op amp feedback to be negative. Then

$$v_o = -V_m \frac{R_2}{R_1} \frac{v_z}{v_x} \quad (20.37)$$

The log–antilog technique can also be applied to dividing two voltages by first subtracting their logarithms and then taking the antilog. The DIV100 (Burr-Brown) uses this technique. An analog-to-digital converter can be considered a divider with digital output and one dc input (the reference voltage). A multiplier can also be converted into a square rooter as shown in Figure 20.16d. The diode is required to prevent circuit latch-up [10]. The input voltage must be negative.

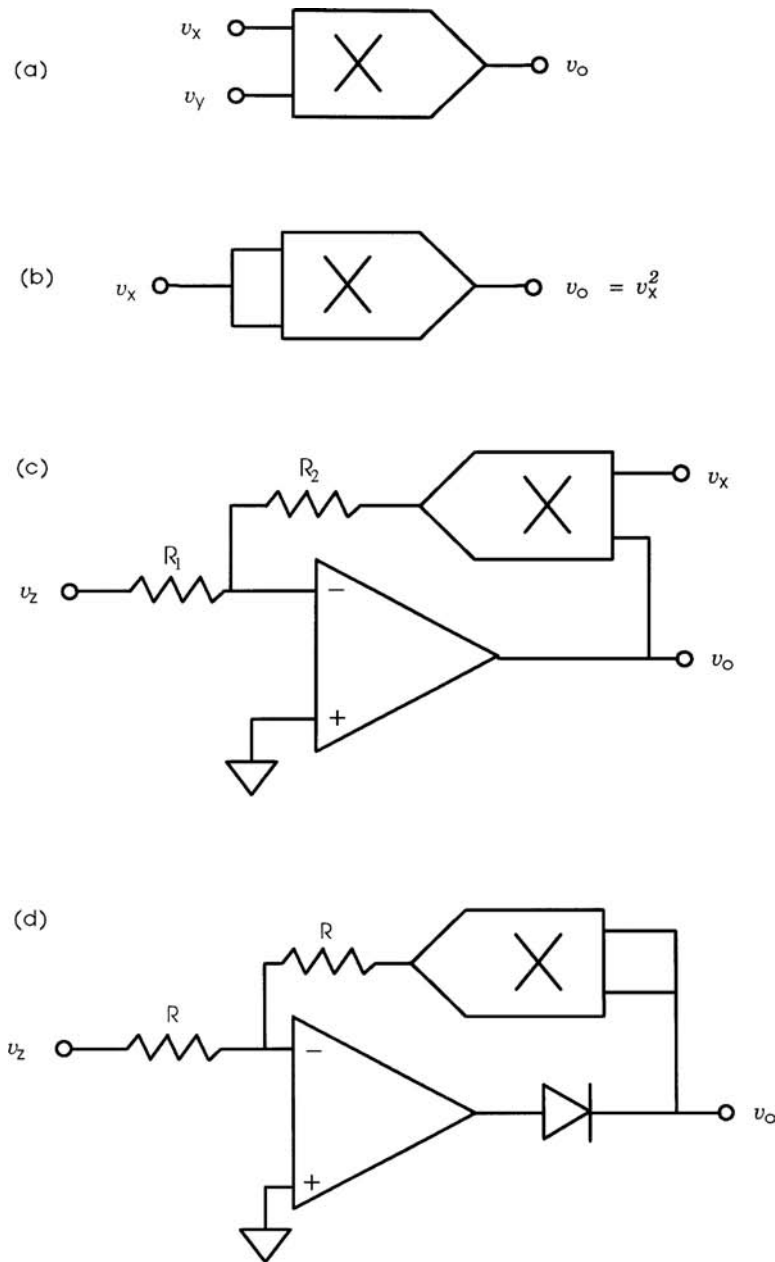


FIGURE 20.16 (a) Symbol for an analog multiplier. (b) Voltage squarer from an analog multiplier. (c) Two-quadrant analog divider from a multiplier and op amp feedback. (d) Square rooter from a multiplier and op amp feedback.

20.11 Analog Linearization

Nonlinearity in instrumentation can result from the measurement principle, from the sensor, or from sensor conditioning. In pressure-drop flowmeters, for example, the drop in pressure measured is proportional to the square of the flow velocity; hence, flow velocity can be obtained by taking the square root of the pressure signal. The circuit in Figure 20.16d can perform this calculation. Many sensors are linear only in a restricted measurand range; other are essentially nonlinear (NTC, LDR); still others are

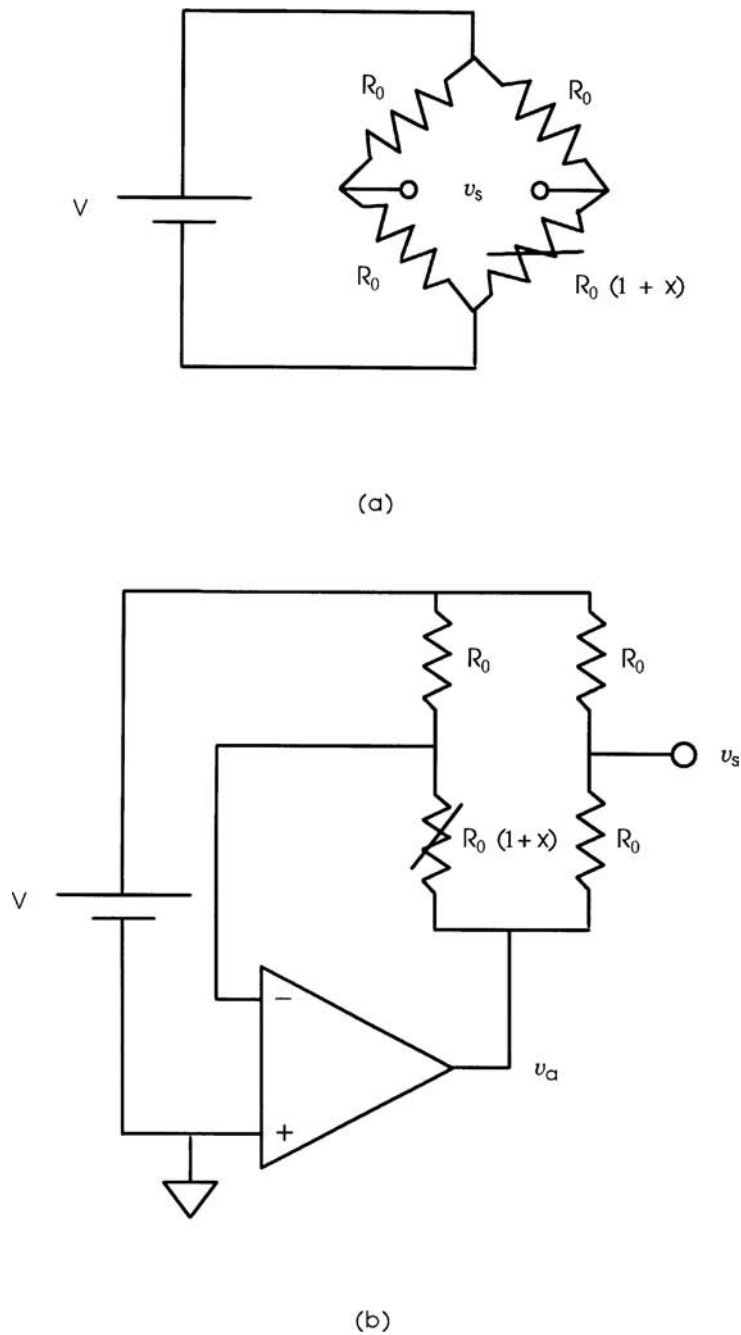


FIGURE 20.17 (a) A Wheatstone bridge supplied at a constant voltage and including a single sensor provides a nonlinear output voltage. (b) By adding an op amp which forces a constant current through the sensor, the output voltage is linearized.

linear in some ranges but nonlinear in other ranges of interest (thermocouples). Linearization techniques for particular sensors are described in the respective chapters.

Nonlinearity attributable to sensor conditioning is common, for example, when resistive (linear) sensors are placed in voltage dividers or bridges. The Wheatstone bridge in Figure 20.17a, for example, includes a linear sensor but yields a nonlinear output voltage,

TABLE 20.4 Special-Purpose Integrated Circuit Signal Conditioners

Model	Function	Manufacturer
4341	rms-to-dc converter	Burr-Brown
ACF2101	Low-noise switched integrator	Burr-Brown
AD1B60	Intelligent digitizing signal conditioner	Analog Devices
AD2S93	LVDT-to-digital converter (ac bridge conditioner)	Analog Devices
AD594	Thermocouple amplifier with cold junction compensation	Analog Devices
AD596/7	Thermocouple conditioner and set-point controllers	Analog Devices
AD598	LVDT signal conditioner	Analog Devices
AD636	rms-to-dc (rms-to-dc converter)	Analog Devices
AD670	Signal conditioning ADC	Analog Devices
AD698	LVDT signal conditioner	Analog Devices
AD7710	Signal conditioning ADC with RTD excitation currents	Analog Devices
AD7711	Signal conditioning ADC with RTD excitation currents	Analog Devices
IMP50E10	Electrically programmable analog circuit	IMP
LM903	Fluid level detector	National Semiconductor
LM1042	Fluid level detector	National Semiconductor
LM1819	Air-core meter driver	National Semiconductor
LM1830	Fluid detector	National Semiconductor
LT1025	Thermocouple cold junction compensator	Linear Technology
LT1088	Wideband rms-to-dc converter building block	Linear Technology
LTK001	Thermocouple cold junction compensator and matched amplifier	Linear Technology
TLE2425	Precision virtual ground	Texas Instruments

$$v_s = V \left(\frac{1+x}{2+x} - \frac{1}{2} \right) = \frac{Vx}{2(2+x)}$$

(20.38)

The nonlinearity arises from the dependence of the current through the sensor on its resistance, because the bridge is supplied at a constant voltage. The circuit in [Figure 20.17b](#) provides a solution based on one op amp which forces a constant current V/R_0 through the sensor. The bridge output voltage is

$$v_s = \frac{V + v_a}{2} = V \frac{x}{2}$$

(20.39)

In addition, v_s is single ended. The op amp must have a good dc performance.

20.12 Special-Purpose Signal Conditioners

Table 20.4 lists some signal conditioners in IC form intended for specific sensors and describes their respective functions. The decision whether to design a signal conditioner from parts or use a model from Table 20.4 is a matter of cost, reliability, and availability. Signal conditioners are also available as subsystems (plug-in cards and modules), for example, series MB from Keithley Metrabyte, SCM from Burr-Brown, and 3B, 5B, 6B, and 7B from Analog Devices.

Defining Terms

- Carrier amplifier:** Voltage amplifier for narrowband ac signals, that includes in addition a sine wave oscillator, a synchronous demodulator, and a low-pass filter.
- Common-mode rejection ratio (CMRR):** The gain for a differential voltage divided by the gain for a common-mode voltage in a differential amplifier. It is usually expressed in decibels.
- Common-mode voltage:** The average of the voltages at the input terminals of a differential amplifier.
- Differential amplifier:** Circuit or device that amplifies the difference in voltage between two terminals, none of which is grounded.

Dynamic range: The measurement range for a quantity divided by the desired resolution.

Instrumentation amplifier: Differential amplifier with large input impedance and low offset and gain errors.

Isolation amplifier: Voltage amplifier whose ground terminal for input voltages is independent from the ground terminal for the output voltage (i.e., there is a large impedance between both ground terminals).

Isolation Mode Rejection Ratio (IMRR): The amplitude of the output voltage of an isolation amplifier divided by the voltage across the isolation impedance yielding that voltage.

Signal conditioner: Circuit or device that adapts a sensor signal to an ensuing circuit, such as an analog-to-digital converter.

Voltage buffer: Voltage amplifier whose gain is 1, or close to 1, and whose input impedance is very large while its output impedance is very small.

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Further Information

B.W.G. Newby, *Electronic Signal Conditioning*, Oxford, U.K.: Butterworth-Heinemann, 1994, is a book for those in the first year of an engineering degree. It covers analog and digital techniques at beginners' level, proposes simple exercises, and provides clear explanations supported by a minimum of equations.

P. Horowitz and W. Hill, *The Art of Electronics*, 2nd ed., Cambridge, U.K.: Cambridge University Press, 1989. This is a highly recommended book for anyone interested in building electronic circuits without worrying about internal details for active components.

M.N. Horenstein, *Microelectronic Circuits and Devices*, 2nd ed., Englewood Cliffs, NJ: Prentice-Hall, 1996, is an introductory electronics textbook for electrical or computer engineering students. It provides many examples and proposes many more problems, for some of which solutions are offered.

J. Dostál, *Operational Amplifiers*, 2nd ed., Oxford, U.K.: Butterworth-Heinemann, 1993, provides a good combination of theory and practical design ideas. It includes complete tables which summarize errors and equivalent circuits for many op amp applications.

T.H. Wilmshurst, *Signal Recovery from Noise in Electronic Instrumentation*, 2nd ed., Bristol, U.K.: Adam Hilger, 1990, describes various techniques for reducing noise and interference in instrumentation. No references are provided and some demonstrations are rather short, but it provides insight into very interesting topics.

Manufacturers' data books provide a wealth of information, albeit nonuniformly. Application notes for special components should be consulted before undertaking any serious project. In addition, application notes provide handy solutions to difficult problems and often inspire good designs. Most manufacturers offer such literature free of charge. The following have shown to be particularly useful and easy to obtain: *1993 Applications Reference Manual*, Analog Devices; *1994 IC Applications Handbook*, Burr-Brown; *1990 Linear Applications Handbook* and *1993 Linear Applications Handbook*, Vol. II, Linear Technology; *1994 Linear Application Handbook*, National Semiconductor; *Linear and Interface Circuit Applications*, Vols. 1, 2, and 3, Texas Instruments.

R. Pallás-Areny and J.G. Webster, *Analog Signal Processing*, New York: John Wiley & Sons, 1999, offers a design-oriented approach to processing instrumentation signals using standard analog integrated circuits, that relies on signal classification, analog domain conversions, error analysis, interference rejection and noise reduction, and highlights differential circuits.