# Section 2

# Operational Amplifier and Comparator Theory

## OPERATIONAL AMPLIFIER THEORY

In 1958, the age of the integrated circuit was ushered in by Jack Kilby of Texas Instruments. From the two hand-built circuits which he fabricated, the variety and quantity of integrated circuits have mushroomed at an ever increasing rate. One type of integrated circuit is the operational amplifier that is characterized by its high gain and versatility. Because of its versatility and ease of application, the operational amplifier has become one of the most widely used linear integrated circuits. Operational amplifiers are designed to be used with external components to provide the desired transfer functions.

The rapid evolution and versatility of the operational amplifier is shown by its initial development and use. One of the two hand-built integrated circuits which Jack Kilby built was a phase shift oscillator, the first linear integrated circuit. This was soon followed by the introduction of the uA702 and SN523 operational amplifiers. Even with their lack of short-circuit protection and their requirements for complex compensation they quickly gained acceptance. Among the improved designs which quickly followed was the uA741 single operational amplifier which required no external compensation. Conversely, the uA748 was designed for compensation by external components to change the frequency response for applications requiring wider bandwidth and higher slew rate.

Operational amplifier capabilities and versatility are enhanced by connecting external components to change the operating characteristics. Typical operational amplifier characteristics include frequency response, signal phase shift, gain and transfer function. The external components are placed in one or more feedback networks and/or the circuits that terminate the input.

To adequately evaluate the potential of an operational amplifier for a specific application, an understanding of operational amplifier characteristics is required. Figure 2-1 represents an equivalent operational amplifier circuit and its parameters. The parameters illustrated in Figure 2-1 are as follows:

- Input bias currents ($I_{IB1}$ and $I_{IB2}$) — the current flowing into both operational amplifier inputs. In an ideal condition, $I_{IB1}$ and $I_{IB2}$ are equal.
- Differential input voltage ($V_{DI}$) — the differential input voltage between the noninverting (+) input and the inverting (−) input.
- Input offset voltage ($V_{IO}$) — an internally generated input voltage identified as the voltage that must be applied to the input terminals to produce an output of 0 V.
- Input resistance ($R_I$) — the resistance at either input when the other input is grounded.
- Output voltage ($V_O$) — normal output voltage as measured to ground.
- Output resistance ($R_O$) — resistance at the output of the operational amplifier.
- Differential voltage gain ($A_{VD}$) or open-loop voltage gain ($A_{OL}$) — the ratio of the input voltage to the output voltage of the operational amplifier without external feedback.
- Bandwidth (BW) — the band of frequencies over which the gain ($V_O/V_{DI}$) of the operational amplifier remains within desired limits.

The generator symbol (G) in Figure 2-1 represents the output voltage resulting from the product of the gain and the differential input voltage ($A_{VD}$ $V_{DI}$).

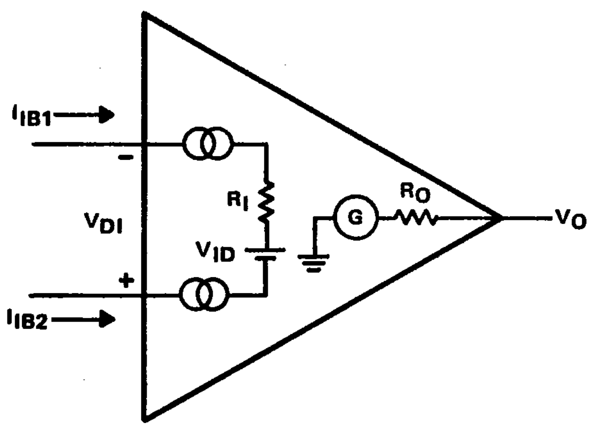



**Figure 2-1. Operational Amplifier Equivalent Circuit**

An ideal operational amplifier (see Figure 2-2) provides a linear output voltage that is proportional to the difference in voltage between the two input terminals. The output voltage will have the same polarity as that of the noninverting (+) input with respect to the voltage at the inverting (−) input. When the noninverting input is more positive than the inverting input, the output voltage will have a positive amplitude. When the noninverting input is more negative than the inverting input, the output voltage will have a negative amplitude.

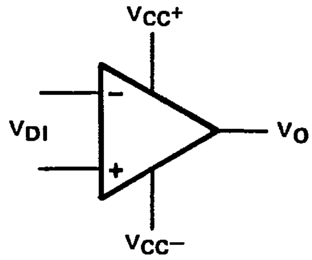



**Figure 2-2. Ideal Operational Amplifier**

An operational amplifier having no external feedback from output to input is described as being in the open-loop mode. In the open-loop mode, the characteristics of the ideal operational amplifier are as follows:

Differential gain = → ∞
Common-mode gain = 0
Input resistance = → ∞
Output resistance = 0
Bandwidth = → ∞
Offset and drift = 0

## MAJOR PERFORMANCE CHARACTERISTICS

The detailed and specific performance characteristics of a particular operational amplifier can be found on the appropriate data sheet. An operational amplifier data sheet will normally provide many nongeneric electrical characteristics. The electrical characteristics provided are for a specified supply voltage and ambient temperature and usually will have minimum, typical, and maximum values.

Major operational amplifier characteristics and their meaning are as follows:

- Input offset current ($I_{IO}$) — the difference between the two input bias currents when the output voltage is zero.
- Common-mode input voltage range ($V_{ICR}$) — the range of the common-mode input voltage (i.e., the voltage common to both inputs).
- Output short-circuit current ($I_{OS}$) — the maximum output current that the operational amplifier can deliver into a short circuit.
- Output voltage swing ($V_{OPP}$) — the maximum peak-to-peak output voltage that the operational amplifier can produce without saturation or clipping occurring. This characteristic is dependent upon output load resistance.
- Large-signal differential voltage gain ($A_{VD}$) — the ratio of the ouput voltage swing to the input voltage swing when the output is driven to a specified large-signal voltage (typically ± 10 V).
- Slew rate (SR) — the time rate of change of the closed-loop output voltage with the operational amplifier circuit having a voltage gain of unity (1).
- Supply current ($I_{CC}$) — The total current that the operational amplifier will draw from both power supplies when unloaded (per amplifier for multiunit packages).
- Common-mode rejection ratio (CMRR) — a measure of the ability of an operational amplifier to reject signals that are present at both inputs simultaneously. The ratio of the common-mode input voltage to the generated output voltage and is usually expressed in decibels (dB).

The preceding paragraphs have discussed basic operational amplifier characteristics. The following paragraphs will provide more detailed information. The specific characteristics that will be discussed are as follows:

- Gain and frequency response
- Difference of input resistance
- Influence of input offset voltage
- Input offset compensation
- Input offset voltage temperature coefficient
- Influence of input bias current
- Influence of output resistance
- Input common-mode range
- Common-mode rejection ratio (CMRR)
- Slew rate
- Noise
- Phase margin
- Output voltage swing
- Feed-forward compensation

### Gain and Frequency Response

Unlike the ideal operational amplifier, a typical operational amplifier has a finite differential gain and bandwidth. Because many of the ideal operational amplifier characteristics cannot be achieved, the characteristics of a typical amplifier differ significantly from those of the ideal amplifier. The open-loop gain of a TL321 operational amplifier is shown in Figure 2-3. At low frequencies, open-loop gain is constant. However, at approximately 6 Hz it begins to roll off at the rate of −6 dB/octave (an octave is a doubling in frequency and decibels are a measure of gain

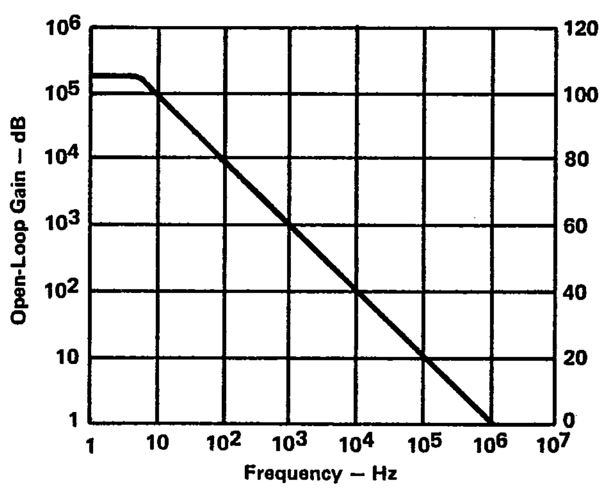


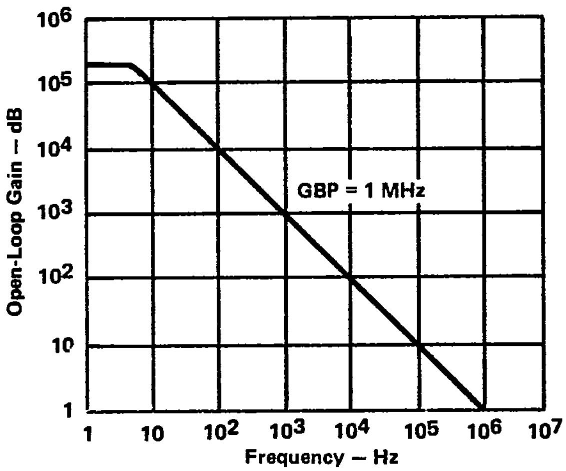

**Figure 2-3. TL321 Operational Amplifier Bandwidth**

calculated by 20 $\log_{10} V_O/V_I$). The frequency at which the gain reaches unity is called unity gain bandwidth and referred to as B1.

When a portion of the output signal is fed back to the input of the operational amplifier, the ratio of the output to input voltage is called closed-loop gain. Closed-loop gain is always less than the open-loop gain. Because gain error is proportional to the ratio of closed-loop gain to open-loop gain, a very high value of open loop gain is desirable.

## Gain-Bandwidth Product

When selecting an operational amplifier for a particular application, gain-bandwidth product is one of the primary factors to consider. The product of closed-loop gain and frequency response (expressed as bandwidth, BW, remains constant at any point on the linear portion of the open-loop gain curve (see Figure 2-4).




Gain-Bandwidth Product = Closed-Loop Gain X Frequency Response

**Figure 2-4. Bandwidth for Operational Amplifier TL321**

The bandwidth is the frequency at which the closed-loop gain curve intersects the open-loop curve as shown in Figure 2-4. The bandwidth may be obtained for any desired closed-loop gain by drawing a horizontal line from the desired gain to the roll-off intersection of the open-loop gain curve. In a typical design, a factor of 1/10 or less of the open loop gain at a given frequency should be used. This ensures that the operational amplifier will function properly with minimum distortion. When the voltage gain of an operational amplifier circuit is increased, the bandwidth will decrease.

## Influence of the Input Resistance

The influence of the input resistance can be determined with Kirchhoff's law. By applying Kirchhoff's law to the circuit in Figure 2-5, we can use the following equations:

$$I_1 = I_2 + I_3 \text{ or}$$

$$\frac{V_I - V_{DI}}{R1} = \frac{V_{DI} - V_O}{R2} + \frac{V_{DI}}{R_I}$$

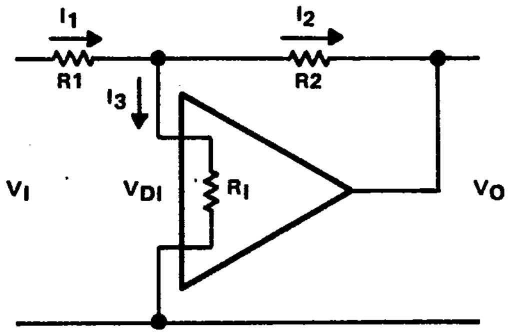



**Figure 2-5. Influence of the Input Resistance**

If the open-loop gain is infinite, the differential input voltage will be zero and the value of input resistance (if it is not zero) will have no influence. Since $V_{DI} = V_O/A_{VD}$, the following equations apply:

$$\frac{V_I - \frac{V_O}{A_{VD}}}{R1} = \frac{\frac{V_O}{A_{VD}} - V_O}{R2} + \frac{\frac{V_O}{A_{VD}}}{R_I}$$

Therefore:

$$\frac{V_I}{V_O} = \frac{1}{A_{VD}} + \frac{1}{\frac{R2}{R1}A_{VD}} + \frac{R1}{R_I A_{VD}} - \frac{1}{\frac{R2}{R1}}$$

or:

$$\frac{V_I}{V_O} = -\frac{1}{\frac{R2}{R1}} + \frac{1}{\frac{R2}{R1}A_{VD}} + \frac{1}{A_{VD}}\left(1 + \frac{R1}{R_I}\right)$$

The previously listed equations indicate that the input resistance (unless it is small relative to R1) will have little or no effect on the ratio of output voltage to input voltage. Therefore, the closed-loop gain for typical applications is independent of the input resistance.

## Influence of Input Offset Voltage

The input offset voltage ($V_{IO}$) is an internally generated voltage and may be considered as a voltage inserted between the two inputs (see Figure 2-6). In addition, it is a differential input voltage resulting from the mismatch of the operational amplifier input stages.

The effect on currents $I_1$ and $I_2$ can be determined by the following equations:

$$\frac{V_I - V_{IO}}{R1} = \frac{V_{IO} - V_O}{R2}$$

If the input voltage ($V_I$) is zero, the equation is as follows:

$$\frac{-V_{IO}}{R1} = \frac{V_{IO} - V_O}{R2}$$

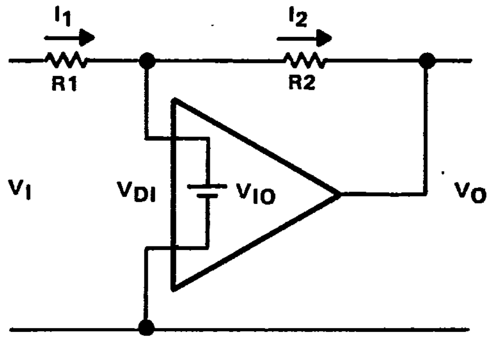



**Figure 2-6. Influence of Input Offset Voltage**

The output voltage is the output offset voltage ($V_{OO}$). The following equation can be used to determine $V_{OO}$:

$$V_{OO} = \left(\frac{R2}{R1} + 1\right) V_{IO}$$

The value of the input offset voltage can be found by dividing the output offset voltage by the closed-loop gain.

**Input Offset Compensation**

An ideal operational amplifier has zero input offset voltage and no drift. However, because of the mismatch of input transistors and resistors on the monolithic circuit, typical operational amplifiers have a low but definite offset voltage. Most operational amplifiers have provisions for connecting an external potentiometer so that the input offset can be adjusted to zero. The exact method used and total resistance of the null adjustment potentiometer is dependent upon the type of operational amplifier circuit. A general-purpose internally compensated operational amplifier (a uA741) may require a 10 kΩ potentiometer. A BIFET or externally compensated operational amplifier may require a 100 kΩ potentiometer. Recommended input offset voltage null adjustment circuits are usually shown in the data sheet.

Methods of nulling the input offset voltage are shown in Figures 2-7 and 2-8. When the offset null pins (N1 and N2) are connected to the emitter of the constant-current generators, a circuit similar to that shown in Figure 2-7 is used. When the null pins are connected to the collectors of the constant-current generator, a circuit similar to that shown in Figure 2-8 is used.

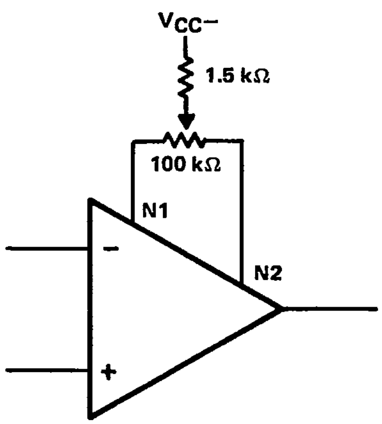



**Figure 2-7. Null Pins Connected to Emitters**

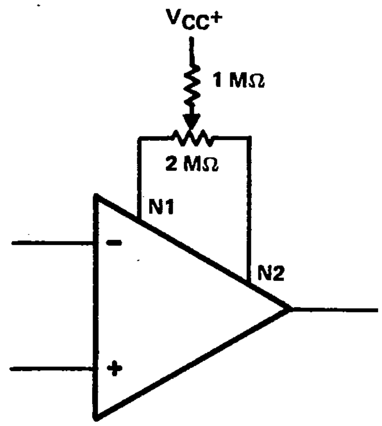



**Figure 2-8. Null Pins Connected to Collectors**

Actual resistor values depend upon the type of operational amplifier used. Consult the appropriate data sheet for complete input offset nulling procedures.

**Input Offset Voltage Temperature Coefficient**

Input offset voltage temperature coefficient (offset voltage drift) is specified in volts per degree Celsius. The amount of drift that occurs with temperature variations is directly related to how closely matched the input characteristics are when the device is manufactured. BIFET input devices (such as the TL080 family) typically have 10 to 12 μV/°C. The LinCMOS™ operational amplifier family has from 0.7 to 5 μV/°C depending upon the bias mode selected.

**Influence of Input Bias Current**

Both input bias current ($I_3$) and the normal operating currents ($I_1$ and $I_2$) flow through resistors R1 and R2 (see Figure 2-9). A differential input voltage equal to the product of $I_3(R1R2)/R1 + R2$ is generated by this current. The differential input (which is similar to input offset voltage)

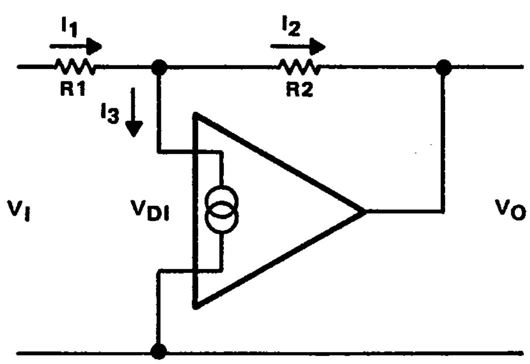



**Figure 2-9. Influence of Input Bias Current**

LinCMOS is a trademark of Texas Instruments also appears as a component of the output that is amplified by the system gain. Methods of correcting for the effects of input bias current are discussed later.

### Influence of Output Resistance

The influence of output resistance is illustrated by Figure 2-10. Output current can be expressed by the following equation:

$$I_O = I_2 + I_L \text{ and } I_2 + I_l = \frac{V_O}{\dfrac{R2R_L}{R2 + R_L}}$$

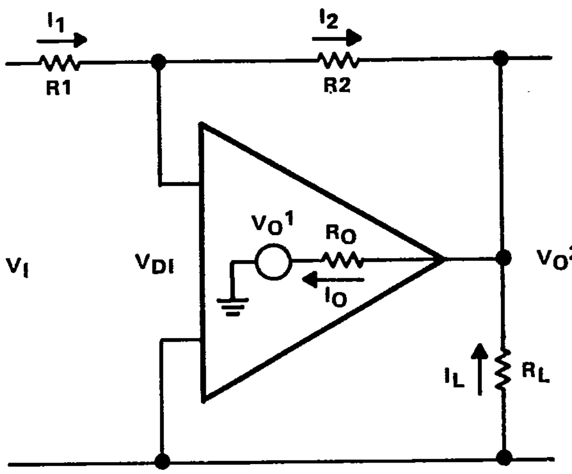



**Figure 2-10. Influence of Output Resistance**

If $V_O1$ is the output voltage of the equivalent ideal amplifier, and $V_O2$ is the output voltage of the actual device, then $V_O2$ can be determined by the following equation:

$$V_O2 = V_O1 - R_OI_O = V_O1 - \frac{R_OV_O2}{\dfrac{R2R_L}{R2 + R_L}}$$

For the ideal case $V_O1 = V_{DI}A_{VD}$; therefore:

$$V_O2 = V_{DI}A_{VD} - R_O\left[\frac{V_O2}{\dfrac{R2R_L}{R2 + R_L}}\right]$$

### Input Common-Mode Range

The input common-mode range may be defined as the maximum range of the input voltage that can be simultaneously applied to both inputs without causing cutoff, clipping, or saturation of the amplifier gain stages. The input stage must be capable of operating within its specifications over the dynamic range of output swing. If it cannot, the amplifier may saturate (or latch-up) when the input limits are exceeded. Latch-up occurs most often in voltage-follower stages where the output voltage swing is equal to the input voltage swing and the operational amplifier is driven into saturation. The specified common-mode voltage range of the input stage must exceed the maximum peak-to-peak voltage swing at the input terminals or the input stage may saturate on peaks. When saturation occurs, an inverting stage no longer inverts. The negative feedback becomes positive feedback and the stage remains in saturation.

### Common-Mode Rejection Ratio (CMRR)

The common-mode rejection ratio may be defined as the ratio of the differential signal gain to the common-mode signal gain and is expressed in decibels.

$$\text{CMRR (dB)} = 20\ (\log_{10})\ \frac{\dfrac{V_O}{V_I}}{\dfrac{V_O}{V_{CM}}}$$

$$\text{or } \frac{\text{(differential signal gain)}}{\text{(common-mode signal gain)}}$$

An ideal operational amplifier responds only to differential input signals and ignores signals common to both inputs. In a typical circuit, however, operational amplifiers have a small but definite common-mode error. Common-mode rejection is important to noninverting or differential amplfiers because these configurations see a common-mode voltage. Depending upon the type of device, dc rejection ratios may range from 90 dB to 120 dB. Generally, bipolar operational amplifiers have higher rejection ratios than FET-input amplifiers.

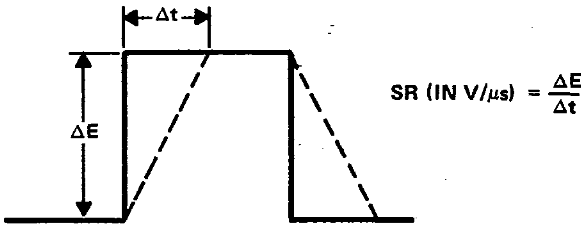
### Influence of Voltage and Current Drift

Input offset voltage, input bias current, and differential offset currents may drift with temperature. Although it is relatively easy to compensate for the effects of these characteristics themselves, correcting for their drift with temperature variations is difficult. However, there is some limited control offered by the design over any drift characteristics. When drift tendencies are expected to be a design problem, device type, construction, and application should be considered.

### Slew Rate

The slew rate may be defined as the maximum rate of change of the output voltage for a step voltage applied to the input (see Figure 2-11). Slew rate is normally measured with the amplifier in a unity gain configuration. Both slew rate and gain bandwidth product are measures of the speed of the operational amplifier.

Slew-rate limiting is accomplished by the limiting of the operational amplifier internal circuit ability to drive capacitive loads. Capacitance limits the slewing ability of the operational




NOTE: Solid line is a square-wave input. Broken line is slewed output.

**Figure 2-11. Effect of Slew Rate**

amplifier at high frequencies. When the current available to charge and discharge the capacitance becomes exhausted, slew-rate limiting occurs.

### Noise

Although not specifically stated as one of the primary characteristics of the ideal operational amplifier, noise-free operation is desirable. Typical operational amplifiers degrade the input signal by adding noise components. Noise components are usually random and determine the ultimate lower limit of signal-handling capability. Noise is usually specified on the data sheet as equivalent input noise, and like the other input factors, is increased by the gain of the stage. There are several potential sources of noise in an operational amplifier. The most common are thermal noise caused by the two source resistances (this noise exists within an ideal operational amplifier), internal noise current, and noise-voltage generators. Under normal audio applications the noise-voltage will be the dominant source of amplifier noise. As the source resistance is increased, the effect of noise-current increases until (at high source resistance) noise current and the bias compensation resistor noise together are the dominant components of amplifier input noise. In specifications, these two parameters are detailed separately. Noise voltage is specified at a low source resistance. Noise current is specified at a high source resistance. Both $V_n$ and $I_n$ are given in terms of density. These are measured with a narrow-bandwidth filter (1 Hz wide) at a series of points across a useful spectrum of the amplifier. Data is usually given in terms of noise voltage versus frequency. Practical data or curves on data sheets are normally given as the following:

$$V_n = e_n/\sqrt{F\ (Hz)}$$

NOTE: Typically a frequency and source resistance will be given in the test conditions included in the device data sheet.

In general, low-input-current operational amplifiers (FET) or low-bias-current bipolar operational amplifiers will have lower noise current and tend to be quieter at source impedances above 10 kΩ. Below 10 kΩ, the advantage swings to bipolar operational amplifiers which have lower input voltage noise. When the source impedance is below 10 kΩ, actual source resistance is composed mostly of generator resistance. The noninverting operational amplifier configuration has less noise gain than the inverting configuration for low signal gains and, thus, high signal-to-noise ratio. At high gains, however, this advantage diminishes.

### Phase Margin

Phase margin is equal to 180° minus the phase shift at the frequency where the magnitude of the open-loop voltage gain is equal to unity. Phase margin is measured in degrees and must be positive for unconditional stability. Figure 2-12 illustrates a typical circuit used to measure phase margin.

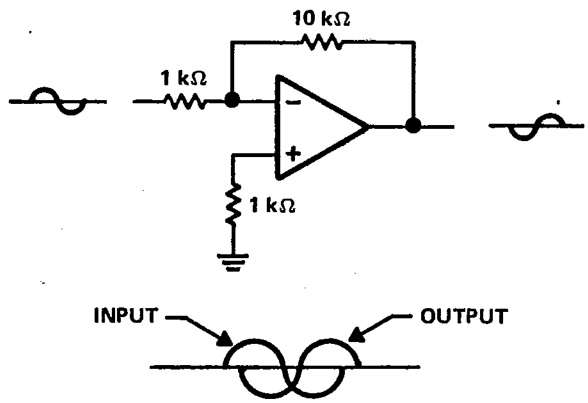



**Figure 2-12. Phase Margin Measurement Circuit**

If the phase difference between the input and output waveform is 120°, 180° minus 120° phase difference leaves 60° as the phase margin. Phase difference may or may not be given on the data sheet. Phase margin will normally be from 50° to 70° on commercially available operational amplifiers. When phase margin decreases to 45°, the operational amplifier becomes unstable and may oscillate.

### Output Voltage Swing ($V_{OPP}$)

$V_{OPP}$ is the peak-to-peak output voltage swing that can be obtained without clipping or saturation. Peak-to-peak swing may be limited by loading effects, operational amplifier frequency capability, load resistance, and power supply used. Load resistances given on the data sheet are usually 2 kΩ or 10 kΩ. With load resistances of 2 kΩ or less, the output decreases due to current limiting. Normally, this will not damage the operational amplifier as long as the specified power-dissipation limits of the package are not exceeded. However, the open-loop gain will be reduced because of excessive loading.

### Feed-Forward Compensation

The TL080 through TL084 BIFET operational amplifiers have been developed by Texas Instruments through state-of-the-art semiconductor technology. The BIFET process allows optimum circuit design with the fabrication of bipolar and FET transistors on a common substrate. This process, along with ion-implanted FET inputs, produces an almost ideal operational amplifier family which typically exhibits a higher input impedance than that obtainable with conventional manufacturing technology. High input impedance and the FET's inherently low input bias currents make the BIFET family ideal for numerous instrumentation and audio amplifier applications.

The TL080 provides for externally controlled compensation on pins 1 and 8. This is an advantage over the internally compensated version (TL081) because it allows the user to obtain slew rates from a typical 12 V/μs (for nominal compensation of 12 pF) to 30 V/μs (for a compensation of 3 pF). This increased slew rate is also reflected in the small signal response where the rise time is decreased from 0.1 μs to less than 50 ns. The power bandwidth can be extended to greater than 1 MHz. This greatly increases the potential for large-signal wide-bandwidth applications and for filters with frequencies at or above 1 MHz. The unity gain bandwidth is identical to the normal compensation mode but the first pole frequency is extended above 10 kHz; thus, gain accuracy is maintained at higher frequencies.

In the feed-forward circuit (see Figure 2-13), a 100 kΩ resistor is shown in parallel with a 3 pF capacitor in the negative feedback loop. A 500 pF feed-forward capacitor is connected from pin 1 to the inverting input of the TL080. The high-frequency response increases from approximately 6 kHz to over 200 kHz. Figure 2-14 is the feed-forward compensation curves.

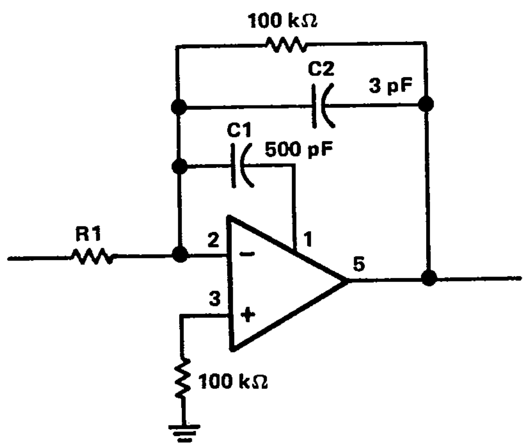



**Figure 2-13. Feed-Forward Compensation**

OPEN-LOOP LARGE SIGNAL DIFFERENTIAL VOLTAGE GAIN vs FREQUENCY

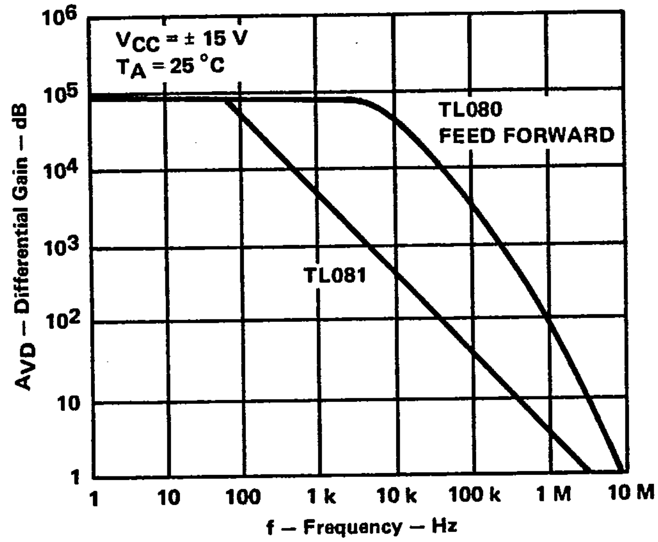



**Figure 2-14. Feed-Forward Compensation Curves**

## BASIC OPERATIONAL AMPLIFIER CIRCUITS

Operational amplifiers, because of their variable characteristics and wide range of adaptablility, can be configured to perform a large number of functions. Operational amplifier applications are frequently limited more by the imagination than by their functional limitations or operating parameters. The basic operational amplifier circuits that are discussed in this section are as follows:

- Noninverting Amplifier
- Inverting Amplifier
- DC Output Offset Amplifier
- Summing Amplifier
- Differentiator
- Active Filter
- Voltage Follower
- Differential Amplifier
- Integrator
- Unity-Gain Active Filter
- Band-Reject Active Filter
- Low-Pass Active Filter

### NONINVERTING OPERATIONAL AMPLIFIER

A noninverting amplifier circuit provides an amplified output that is in phase with the circuit input. Figure 2-15 illustrates a basic noninverting operational amplifier circuit. In the circuit shown in the figure, the output is in phase with the input at low frequencies.

In this circuit, the input signal is applied to the noninverting (+) input of the amplifier. A resistor (R1), which is usually equal to the resistance of the input element, is connected between ground and the inverting (−) input of

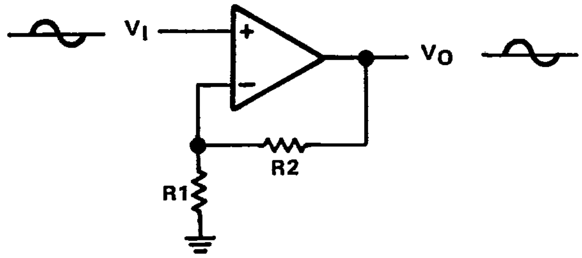



**Figure 2-15. Basic Noninverting Amplifier Circuit**

the amplifier. A feedback loop is connected from the output of the amplifier, through feedback resistor R2, to the inverting input. The voltage gain of a noninverting amplifier circuit is always greater than unity (1). For practical purposes, the input impedance of the noninverting amplifier circuit is equal to the intrinsic input impedance of the operational amplifier.

The output voltage of the noninverting amplifier circuit can be determined by the following equation:

$$V_O = \left[1 + \frac{R2}{R1}\right](V_I)$$

The voltage gain of the noninverting amplifier circuit can be determined by the following equation:

$$A_V = \frac{V_O}{V_I} = 1 + \frac{R2}{R1}$$

## INVERTING AMPLIFIERS

An inverting amplifier circuit, as illustrated in Figure 2-16, provides an output that is 180 degrees out of phase with the input signal.

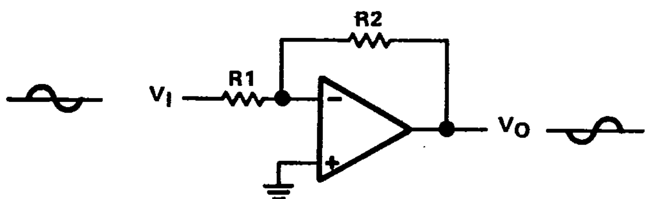



**Figure 2-16. Basic Inverting Amplifier Circuit**

In this circuit, the input signal is applied through a resistor (R1) to the inverting input (−) of the operational amplifier. The noninverting input (+) is connected to ground. A feedback loop is connected from the output, through feedback resistor R2, to the inverting input. The voltage gain of an inverting amplifier circuit can be less than, equal to, or greater than unity (1). Resistor R1 determines the input impedance of the inverting amplifier circuit. The input impedance is much lower than for a noninverting amplifier circuit.

The output voltage for an inverting amplifier can be determined by the following equation:

$$V_O = \left[-\frac{R2}{R1}\right](V_I)$$

NOTE: The minus sign in the equation indicates the 180° phase reversal.

The voltage, or closed-loop, gain can be determined by the following equation:

$$A_{CL} = \frac{V_O}{V_I} = -\left[\frac{R2}{R1}\right]$$

## DC OUTPUT OFFSETS

When the input voltage to an operational amplifier is zero, the ideal output voltage is also zero. However, the ideal condition cannot be realistically achieved because of dc offset. DC offset may be caused by the internal input offset and input bias currents. It may also be caused by an input signal offset voltage. With no signal into the amplifier shown in Figure 2-17, input bias current flows through resistors R1 and R2. Because of the voltage drop across R1 and R2, these input currents will produce an offset voltage. Since the noninverting input is grounded, the voltage appears as input offset and is amplified by the operational amplifier.

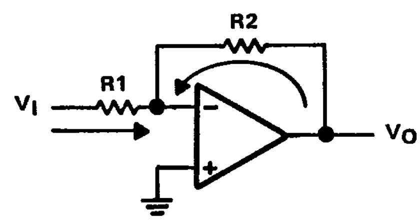



**Figure 2-17. Inverting Amplifier with Input Bias Currents**

The method commonly used to correct for a dc offset condition is to place an additional resistor (R3) between the noninverting input and ground as shown in Figure 2-18.

The value of resistor R3 is calculated as the parallel combination of R1 and R2 as follows:

$$R3 = \frac{R1R2}{R1 + R2}$$

A voltage is developed across R3 that is equal to the voltage across the parallel combination of R1 and R2. Ideally, the voltages appear as common-mode voltages and are cancelled. However, in a typical operational amplifier, the bias currents are not exactly equal. Because of this difference, a small dc offset voltage remains.

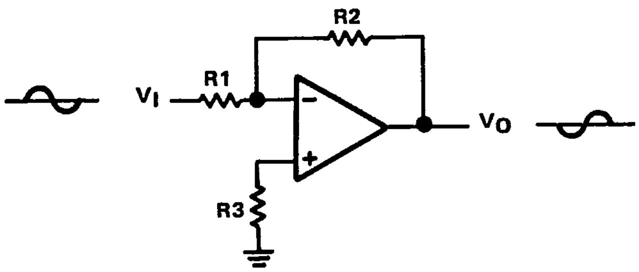



**Figure 2-18. Inverting Amplifier with DC Offset Correction**

The remaining source of output offset voltage is due to the internal input offset voltage. This may be nulled in several ways. The most common method is to connect a potentiometer across the offset null terminals available on many operational amplifiers. Depending upon the type of circuit and chip construction, the center arm is connected to the $V_{CC+}$ rail or $V_{CC-}$ rail. The terminals on the operational amplifier used for this purpose are usually labeled N1 and N2. For complete information on a specific device, consult the appropriate data sheet.

## VOLTAGE FOLLOWER

The voltage or source follower is a unity-gain, noninverting amplifier with no resistor in the feedback loop (see Figure 2-19). The output is exactly the same as the input. The voltage follower has a high input impedance which is equal to the operational amplifier intrinsic input impedance.

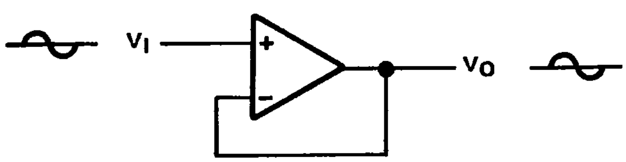



**Figure 2-19. Basic Voltage Follower Circuit**

The function of the voltage follower circuit is identical to a emitter follower for a bipolar transistor or a source follower on an FET transistor. The main purpose of the circuit is to buffer the input signal from the load. The input impedance is high and the output impedance is low.

## SUMMING AMPLIFIER

If several input resistors are connected to the inverting input of the operational amplifier, as shown in Figure 2-20, the result is an amplifier which sums the separate input voltages.

The output voltage of the summing amplifier circuit can be determined by the following equation:

$$V_O = -R4\left(\frac{V_I1}{R1} + \frac{V_I2}{R2} + \frac{V_I3}{R3}\right)$$

If feedback resistor R4 and input resistors R1, R2, and R3 are made equal, the output voltage can be determined by the following equation:

$$V_O = -(V_I1 + V_I2 + V_I3)$$

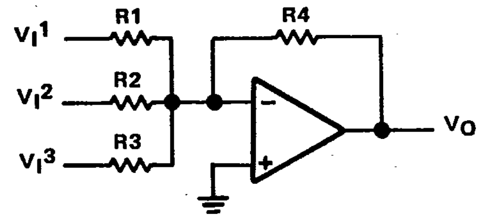



**Figure 2-20. Basic Summing Amplifier**

## DIFFERENCE AMPLIFIER

In a difference amplifier circuit, input voltages $V_I1$ and $V_I2$ are applied simultaneously to the inverting and noninverting inputs of the operational amplifier (see Figure 2-21).

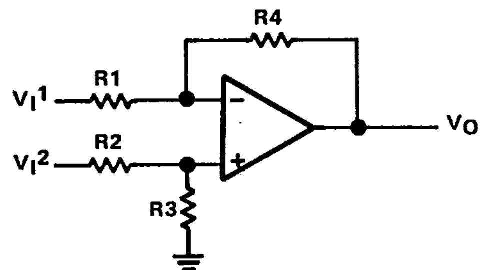



**Figure 2-21. Basic Difference Amplifier Circuit**

When all four resistors are equal, the output voltage is equal to the difference between $V_I2$ and $V_I1$. This circuit is called a unity-gain analog subtractor. Mathematically, the output voltage is stated as follows:

$$V_O = V_I2 - V_I1$$

## DIFFERENTIATOR

The operational amplifier differentiator is similar to the basic inverting amplifier circuit except that the input component is a capacitor rather than a resistor (see Figure 2-22).

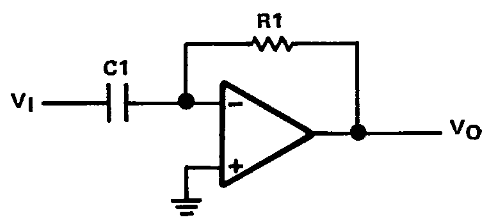



**Figure 2-22. Basic Differentiator Circuit**

The output voltage of the differentiator circuit can be determined by the following equation:

$$V_O = -R1C1\frac{\Delta V_I}{\Delta t}$$

In this equation $\Delta V_I/\Delta t$ is the change in input voltage divided by a specified time interval. A problem with the basic differentiator circuit is that the reactance of input capacitor C1 ($1/2\pi fC1$) varies inversely with frequency. This causes the output voltage to increase with frequency and makes the circuit susceptible to high-frequency noise. To compensate for this problem, a resistor is connected in series with the capacitor on the inverting input (see Figure 2-23).

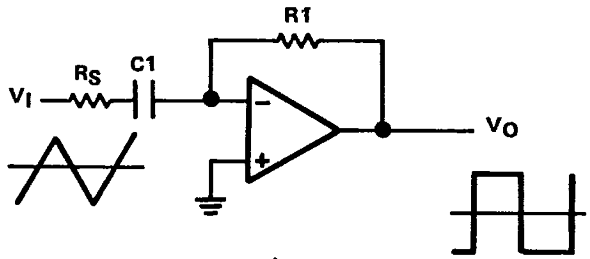



**Figure 2-23. Differentiator with High Frequency Noise Correction**

However, this circuit functions as a differentiator only on input frequencies which are less than those which can be determined by the following equation:

$$fC1 = \frac{1}{2\pi R1C1}$$

The time constant (R1C1) should be approximately equal to the period of the input signal to be differentiated. In practice, series resistor $R_S$ is approximately 50 Ω to 100 Ω.

## INTEGRATOR

An operational amplifier integrator circuit can be constructed by reversing the feedback resistor and input capacitor in a differentiator circuit (see Figure 2-24).

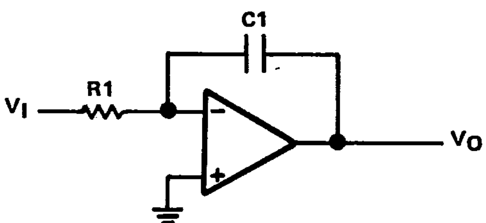



**Figure 2-24. Basic Integrator Circuit**

The resistor (R1) is the input component and the capacitor (C1) is the feedback component. However, if the low-frequency gain of the circuit is not limited, the dc offset (although small), would be integrated and eventually saturate the operational amplifier. A more practical integrator circuit is shown in Figure 2-25.

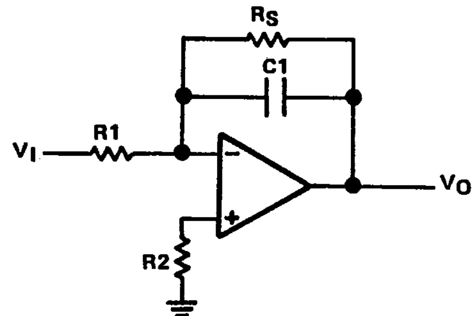



**Figure 2-25. Typical Integrator Circuit**

In this circuit, a shunt resistor ($R_S$) is connected across feedback capacitor C1 to limit the low-frequency gain of the circuit. The dc offset (due to the input bias current) is minimized by connecting resistor R2 between the noninverting input and ground. Resistor R2 is equal to the parallel combination of R1 and shunt resistor $R_S$. The shunt resistor helps limit the circuit low-frequency gain for input frequencies greater than those determined by the following equation.

$$fC1 = \frac{1}{2\pi R_S C1}$$

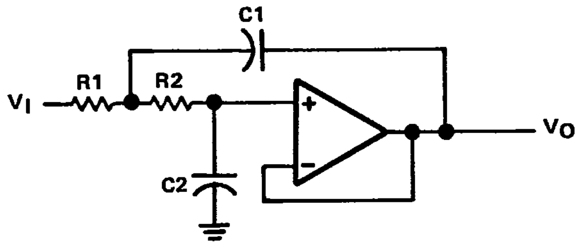
## ACTIVE FILTERS

Filters are often thought of as discrete networks consisting of resistors, capacitors, and inductors (passive components). Because the components are passive, the energy from a passive filter is always less than the energy applied by the input signal. The attenuation (or insertion losses) limit the effectiveness of passive filters and make some applications impractical. However, resistors and capacitors can be combined with operational amplifiers to form active filters which operate without signal loss.

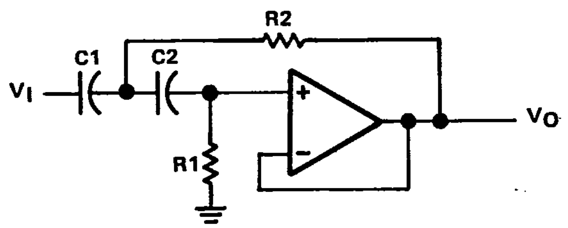
Depending upon the circuit type, low-pass filters as well as high-pass, bandpass, or band-reject filters can be designed with a roll-off characteristic of 6 to 50 dB or greater per octave. Some of the more common active filters that use operational amplifiers are discussed in the following paragraphs.

### Unity-Gain Active Filters

The unity-gain active filter is the simplest to design. It combines an operational amplifier connected in a unity gain configuration with RC filter networks. It can be either a low-pass filter [Figure 2-26(a)], or a high-pass filter [Figure 2-26(b)], depending upon the positions of its discrete resistors and capacitors.




(a) LOW-PASS FILTER




(b) HIGH-PASS FILTER

**Figure 2-26. High-Pass and Low-Pass Filter Circuits**

The −3 dB (cutoff) frequency of the filter can be determined from the following equations:

Low-frequency cutoff $f_o = \frac{1}{2\pi\ C2\sqrt{R1R2}}$

High-frequency cutoff $f_o = \frac{1}{2\pi\ R2\sqrt{C1C2}}$

The Q of the circuit can be calculated using the following formulas for a low or high pass filter:

Low-pass filter $Q = 1/2\sqrt{C1/C2}$
High-pass filter $Q = 1/2\sqrt{R1/R2}$

These formulas are valid for a value of Q greater than 10.

### Low-Pass Active Filters

Figure 2-27 illustrates the response curve typical of a low-pass active filter using a general-purpose operational amplifier. Outside the passband, the attenuation is computed at 12 dB per octave. However, at high frequencies the attenuation of the filter is less than predicted. In simple theory, the operational amplifier is considered to be perfect, and, for a typical general-purpose operational amplifier, this perfection proves to be acceptable up to 100 kHz. However, above 100 kHz the output impedance and other characteristics of the amplifier can no longer be ignored. The combined effect of these factors causes a loss of attenuation at high frequencies. General-purpose operational amplifiers are most effective in the audio frequency range. For higher frequency applications, a broad band amplifier such as the LM318 or TL291 should be used.

When the frequency spectrum of the input signal is especially wide, the high-frequency rejection characteristic must be considered. This is true when the input to the filter

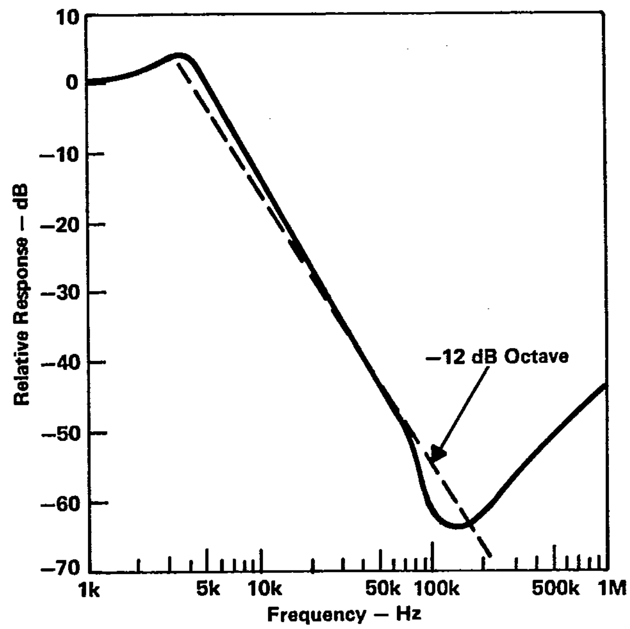



**Figure 2-27. Response Curve of a Low-Pass Active Filter**

is a rectangular signal. Figure 2-28 shows the response of a low-pass active filter to a 1-MHz square-wave signal.

The high-frequency-cutoff problem is resolved by using a simple RC filter ahead of the active filter. The combination of an RC filter and an active filter having superior low-frequency performance will significantly improve high-frequency cutoff. In addition, an impedance adapter should be inserted between the two filters shown in Figure 2-29.

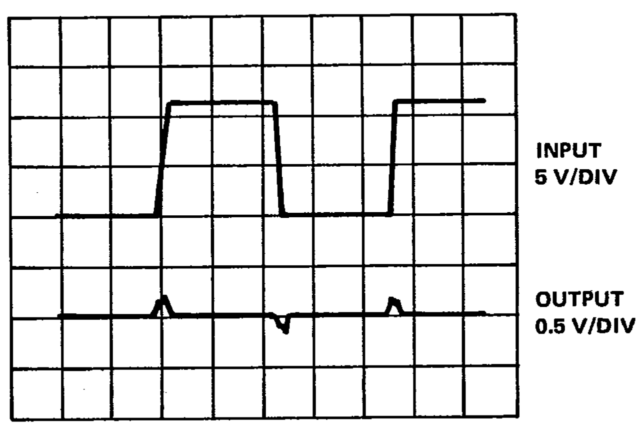



**Figure 2-28. Response of a Low-Pass Active Filter to High Frequency Signals**

### Band-Reject Active Filters

In addition to the previously described functions, an active filter may be used to perform a band-reject function. A filter with a band-reject characteristic is frequently referred to as a notch filter. A typical circuit using a uA741 in unity-gain configuration for this type of active filter is shown in Figure 2-30.

The filter response curve shown in Figure 2-31 is a second-order band-reject filter with a notch frequency of

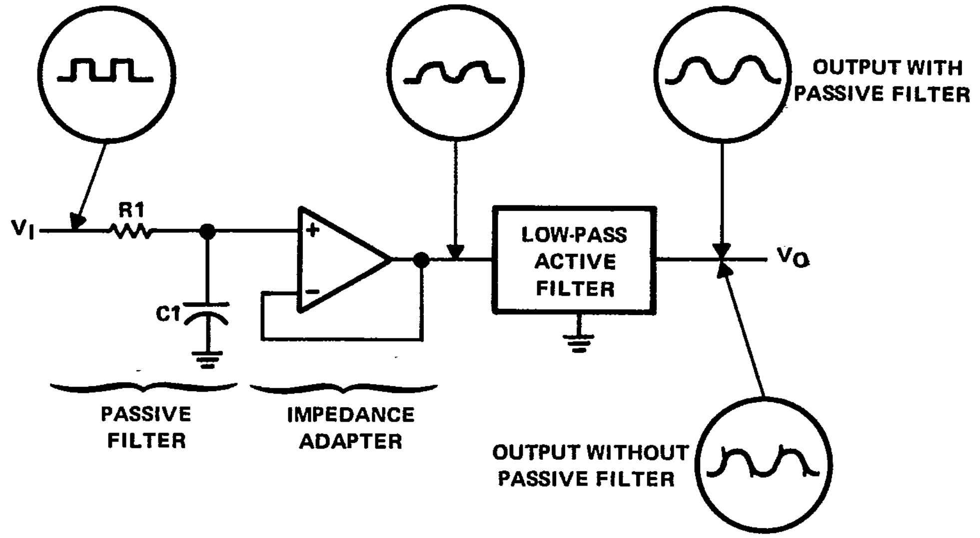



Figure 2-29. Use of a Passive Filter Preceding an Active Filter

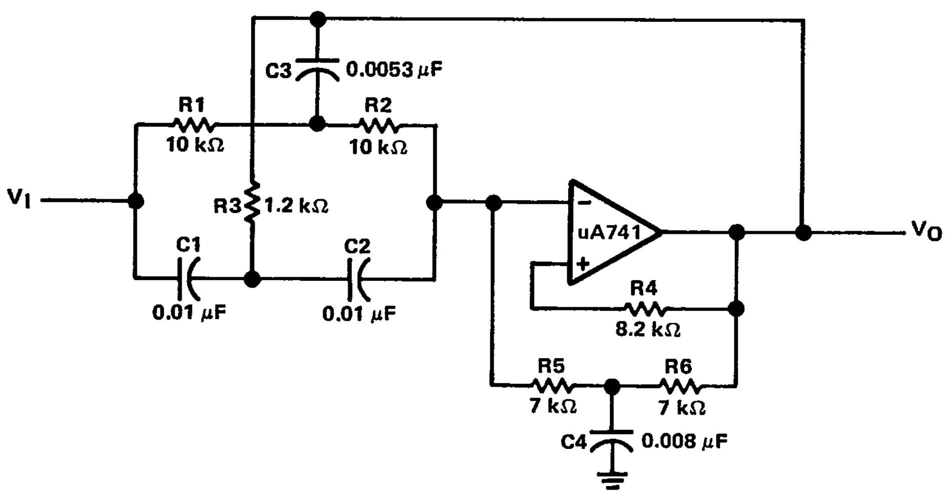



Figure 2-30. Band-Reject Active Filter

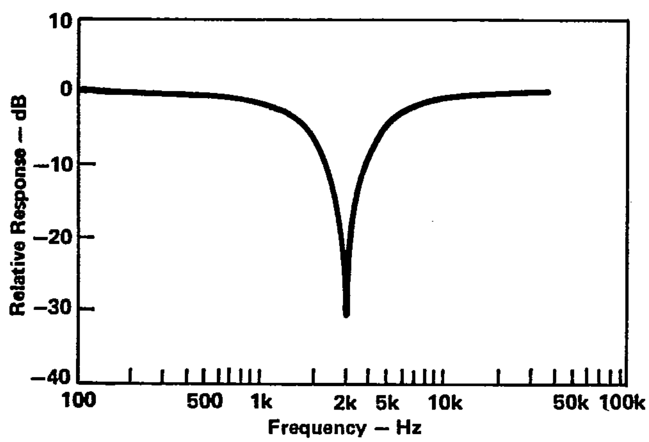



Figure 2-31. Response Curve of Band-Reject Active Filter

3 kHz. The resulting Q of this filter is about 23, with a notch depth of −31 dB. Although three passive T networks are used in this application, the operational amplifier has become a sharply tuned low-frequency filter without the use of inductors or large-value capacitors.

## Bandpass Active Filters

A bandpass filter permits a range of frequencies to pass while attenuating frequencies above and below this range. The center frequency ($f_0$) is the frequency at which the maximum voltage gain occurs. The bandwidth of this type of filter is the difference between the upper and lower frequencies at the points where the voltage gain is 0.707 times the maximum value, or 3 dB lower than the response at the center frequency. As shown in Figure 2-32 $f_L$ is called the lower 3-dB frequency and $f_H$ is called the upper 3-dB frequency. Bandwidth is determined by the following equation:

$$\text{Bandwidth} = f_H - f_L$$

The bandpass filter bandwidth and center frequency are related to each other by the Q, which is defined as follows:

$$Q = \frac{f_0}{f_H - f_L} \text{ or } \frac{f_0}{BW}$$

Bandpass filter responses like those shown in Figure 2-32 can be built with operational amplifiers. The filter circuit shown in Figure 2-33 uses only one operational amplifier and is most often used for Q's of 10 or less.

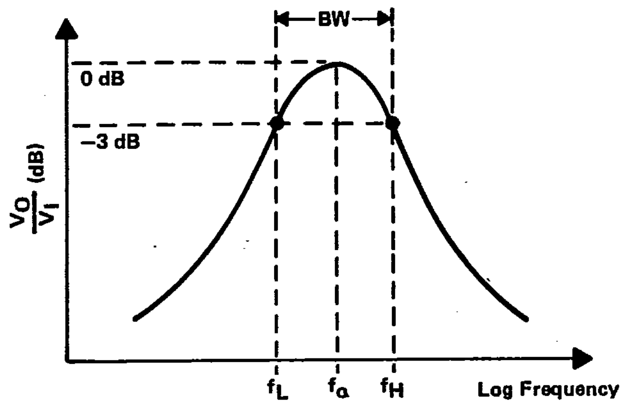



**Figure 2-32. Active Bandpass Filter Response Curve**

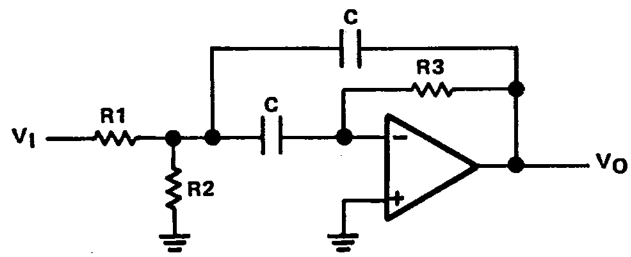



**Figure 2-33. Active Bandpass Filter**

Given the center frequency ($f_O$), the Q, and the desired gain (G), choose a convenient value of capacitance C. For typical audio filters, C is often 0.01 to 0.1 $\mu$F. The component values are easy to find from the following equations.

$$R1 = \frac{Q}{2\pi f_O GC}$$

$$R2 = \frac{Q}{(2Q^2 - G)2\pi f_O C}$$

$$R3 = \frac{2Q}{2\pi f_O C}$$

Best performance is obtained when the gain is somewhat greater than the square root of the Q. For instance, if the filter is designed for a Q of 16, then the gain should be greater than four.

## CHOOSING THE RIGHT OPERATIONAL AMPLIFIER

The operational amplifier, because of its versatility and ease of application, is the most widely used linear integrated circuit today. Because of this useful linear building block, many electronic circuits are much less complex. Due to the popularity of the operational amplifier, many different types are available that offer a variety of features. Which device to use for a specific application is a question that must be answered. If the characteristics of the selected device are not adequate, total system performance may be less than desired. If the selected device is too complex for the job, system cost may be increased unnecessarily. The following paragraphs provide a summary of the various types of operational amplifiers. To assist in the selection of the most effective operational amplifier for a specific application, the features and key applications are presented.

### GENERAL-PURPOSE OPERATIONAL AMPLIFIERS — BIPOLAR

Since their inception, many operational amplifier designs have used bipolar transistors as the differential amplifier pair at the operational amplifier inputs. Because these input transistors operate from constant-current sources, an additional pair of matched transistors are used to obtain closely matched base-emitter voltages for a predictable ratio of currents for the constant-current generators. Phase-shift is controlled by frequency compensation that is internal to the amplifier. Amplfier phase-shift must be less than 135° at the frequency where the open-loop gain curve and closed-loop gain curve intersect. In bipolar operational amplifiers, the phase-shift is typically set with an internal capacitance of approximately 30 pF. The output stage should be designed to have a wide range of voltage swing with medium current capability.

The bipolar operational amplifier is usually operated in a class-B configuration. The key features of a bipolar operational amplifier are as follows:

- Input impedance of $10^6$ Ω.
- Typical slew rates from 0.5 to 1 V/$\mu$s.
- Typical unity-gain bandwidth of 1 MHz.
- Noise levels of approximately 25 to 30 nV/$\sqrt{\text{Hz}}$.

Table 2-1 is a selection guide showing the major parameters to be considered in choosing bipolar operational amplifiers for a particular circuit design.

**Table 2-1. Bipolar Operational Amplifier Comparison Chart**

| PARAMETER | DEVICE | | | | | | | | | UNITS |
|---|---|---|---|---|---|---|---|---|---|---|
| | OP-07 | 741 | TL321 | SE5534A | LM358 | LM318 | MC1458 | RC4136 | RC4558 | |
| $V_{IO}$ | ·30 | 1 | 2 | 0.5 | 2 | 2 | 1 | 0.5 | 0.5 | mV |
| $I_{IO}$ | 0.5 | 20 | 5 | 10 | 5 | 30 | 20 | 5 | 5 | nA |
| $I_B$ | ±1.2 | 80 | 45 | 400 | 45 | 150 | 80 | 40 | 40 | nA |
| SR | 0.2 | 0.5 | 0.5 | 6 | 0.5 | 70 | 0.5 | 1 | 1 | V/μs |
| B1 | 0.6 | 1 | 1 | 10 | 1 | 15 | 1 | 3 | 3 | MHz |

†Test conditions are $V_{CC}$ = ±15 V. All values are typical.
‡Unity-gain bandwidth

## BIFET OPERATIONAL AMPLIFIERS

BIFET operational amplifiers combine JFET input transistors with bipolar transistors in a monolithic integrated circuit. The ion-implantation process used in making BIFET devices results in closely matched transistors. This permits true class-AB operation in the output stage which results in near zero crossover distortion and low total harmonic distortion.

In addition to high input impedance ($10^{12}$ Ω) and input bias currents in the picoampere range, most BIFET operational amplfiers have slew rates of approximately 13 V/μs and a typical unity-gain bandwidth of 3 MHz. However, BIFET operational amplifiers have higher offset voltages and input noise than bipolar operational amplifiers.

Some BIFET operational amplifiers are power-adjustable. This allows the user to select (with an external resistor) the operating current levels. While this causes a tradeoff with power dissipation, it gives greater control over slew rate or signal bandwidth. An example of such a device is the TL066 BIFET operational amplifier. The TL066 can be adjusted for a no-signal supply current of 5 to 200 μA. Slew rate and bandwidth will also change depending upon the level of operating current. Except for the adjustable feature, the TL066 is similar to the TL061. The key application for power adjustable operational amplifiers is in battery-operated and telecommunication equipment where power consumption is an important factor. Table 2-2 is a selection guide listing the major parameters to be considered when choosing a BIFET operational amplifier for a particular application.

**Table 2-2. BIFET Operational Amplifier Comparison Chart**

| PARAMETER | DEVICE SERIES | | | | UNITS |
|---|---|---|---|---|---|
| | TL080 | TL070 | TL060 | TL087 | |
| $V_{IO}$ | 5 | 3 | 3 | 0.1 | mV |
| $I_B$ | 30 | 30 | 30 | 60 | pA |
| NOISE | 25 | 18 | 42 | 18 | nV/Hz |
| SR | 13 | 13 | 3.5 | 13 | V/μs |
| B1 | 3 | 3 | 1 | 3 | MHz |

Test conditions are $V_{CC}$ = ±15 V. All values are typical.

## LinCMOS™ OPERATIONAL AMPLIFIERS

The linear silicon-gate CMOS integrated circuit process was first developed by Texas Instruments and designated by the trademark LinCMOS. The LinCMOS technology combines the high speed of the bipolar device with the low power, low voltage, and high input impedance of the CMOS device. The LinCMOS device provides better offset and voltage swing characteristics than most bipolar devices. In addition, the LinCMOS device overcomes the stability and bandwidth limitations imposed on linear designs by metal-gate CMOS.

### ULTRASTABLE OFFSETS

The primary disadvantage of using conventional bipolar metal-gate CMOS for linear applications is the unavoidable threshold-voltage shifts that take place with time and with changes in temperature and gate voltage. These shifts (caused by the movement of sodium ions within the device transistor) are frequently more than 10 mV/V of applied gate voltage. However, LinCMOS technology overcomes this problem by replacing the metal gates with phosphorus-doped polysilicon gates that bind the sodium ions. The result is linear integrated circuits with low (2 to 10 mV) input-offset voltages that vary no more than a few microvolts from their original values.

The TLC251 and TLC271 series of general-purpose operational amplfiers have low input offset voltages that typically vary only 0.1 μV per month and 0.7 μV per degree Celsius. The extremely low offset voltage can be reduced even further by using the offset null pins on the device. Unlike metal-gate CMOS devices, the input-offset voltage of LinCMOS devices is not sensitive to input-overdrive voltages.

### WIDE BANDWIDTHS

In addition to providing stable offset voltages, LinCMOS technology produces integrated circuits with bandwidths that are two to three times wider than those of metal-gate CMOS devices. This occurs because the silicon gate in LinCMOS transistors is formed during the same processing step that forms the source and drain. As a result, the source, gate, and drain are self-aligned. In contrast, metal gates are formed after the source and drain regions are diffused, necessitating a built-in overlap to ensure source, gate, and drain alignment.