PDHonline Course E278 (3 PDH)

Operational Amplifiers at Higher Frequencies

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THE OP AMP'S BEHAVIOR AT HIGHER FREQUENCIES

Some of the characteristics of the practical Op Amp are sensitive to changes in operating frequency. In this chapter, we will consider problems caused by the decrease (roll-off) of the open-loop gain at higher frequencies. We will also see how the gain vs. frequency curve of an Op Amp can be tailored either by the circuit designer who uses the Op Amp or by its manufacturer. Other notable frequency-related problems such as reduced output-voltage swing capabilities, limited slew rates, and noise are also described here.

6.1 GAIN AND PHASE SHIFT VS. FREQUENCY

Ideally, an Op Amp should have an infinite bandwidth. This means that, if its open-loop gain is 90 dB with dc signals, its gain should remain 90 dB through audio and on to high radio frequencies. The practical Op Amp's gain, however, decreases (rolls off) at higher frequencies as shown in Fig. 6-1. This gain roll-off is caused by capacitances within the Op Amp circuitry. The reactances of these capacitances decrease at higher frequencies, causing shunt signal current paths, and thus reducing the amount of signal available at the output terminal. Along with decreased gain at higher frequencies, there is an increased phase shift of the output signal with respect to the input (see Fig. 6-2). Normally at low frequencies, there is a 180° phase difference in the signals at the inverting input and output terminals. But at higher frequencies, the output signal lags by more than 180°, and this is called *phase shift*. Thus the Op Amp with the characteristics in Fig. 6-2 has practically no phase shift up to about 30 kHz. Beyond 30 kHz, the output signal starts to lag, and at 300 kHz it lags an additional 40°. This negative or

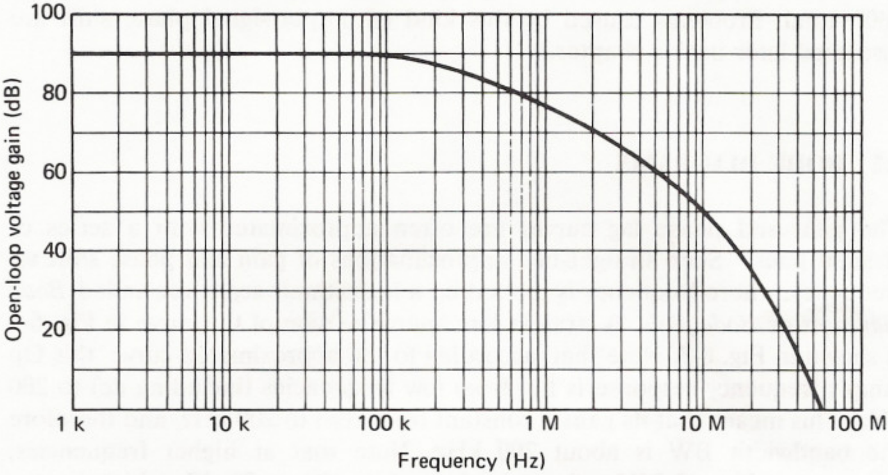


Figure 6-1 Typical open-loop gain vs. frequency curve of an Op Amp.

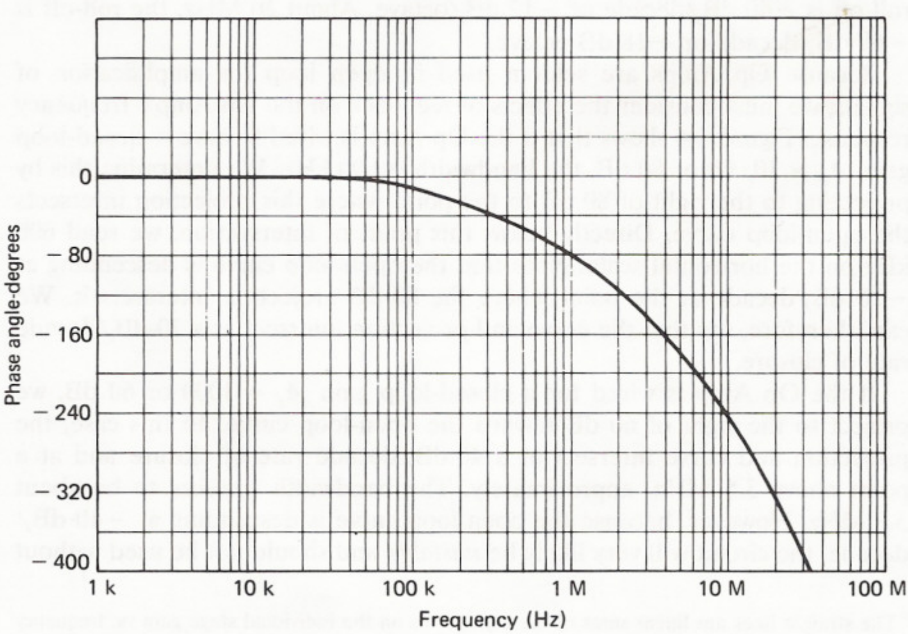


Figure 6-2 Typical open-loop output-signal phase lag vs. frequency.

lagging shift adds to the initially lagging 180°, causing an output signal lag of 220° compared with the signal applied to the inverting input. Similarly, at 1 MHz, the output lags the inverting input's signal by an additional 80°, or a 260° total. Problems caused by this kind of output-signal phase shift are discussed later in this chapter.

6.2 BODE DIAGRAMS

The gain and phase lag curves are often approximated with a series of straight lines.* Such straight-line approximations of gain and phase shift vs. frequency, where frequency is plotted on a logarithmic scale, are called *Bode diagrams* or *Bode plots*. A straight-line approximation of the curve in Fig. 6-1 is shown in Fig. 6-3. Note that, according to the approximated curve, this Op Amp's frequency response is flat from low frequencies (including dc) to 200 kHz. This means that its gain is constant from zero to 200 kHz, and therefore the bandwidth BW is about 200 kHz. Note that at higher frequencies, from 200 kHz to 2 MHz, the gain drops from 90 to 70 dB, which is at a −20-dB/decade† or −6-dB/octave rate. The negative sign refers to the negative (decreasing) slope of the curve. At frequencies from 2–20 MHz, the roll-off is −40 dB/decade or −12 dB/octave. About 20 MHz, the roll-off is −60 dB/decade or −18 dB octave.

Because Op Amps are seldom used in open loop for amplification of signals, we must consider the effects of feedback on the Op Amp's frequency response. Figure 6-4 shows that if the Op Amp is wired to have a closed-loop gain $A_v = 10{,}000$ or 80 dB, the bandwidth is 600 kHz. We determine this by projecting to the right of 80 dB to the point where this projection intersects the open-loop curve. Directly below this point of intersection, we read 600 kHz on the horizontal scale. Note that the open-loop curve is descending at −20 dB/decade at the point where the 80-dB projection intersects it. We can, therefore, say that the curve and projection intersect at a 20-dB/decade rate of closure.

If the Op Amp is wired for a closed-loop gain $A_v = 1000$ or 60 dB, we project to the right of 60 dB toward the open-loop curve. In this case, the projection and curve intersect at a 40-dB/decade rate of closure and at a point above 3.5 MHz, approximately. The bandwidth *appears* to be about 3.5 MHz. However, because the open-loop curve is descending at −40 dB/decade, the circuit will very likely be *unstable* and should not be used without

*The straight lines are linear sums of the asymptotes on the individual stage gain vs. frequency curves.

†A −20-dB/decade gain roll-off means that the open-loop gain decreases 20 dB with a frequency increase by the factor 10, and is equivalent to a −6-dB/octave roll-off. A frequency change by an octave is a change by the factor 2.

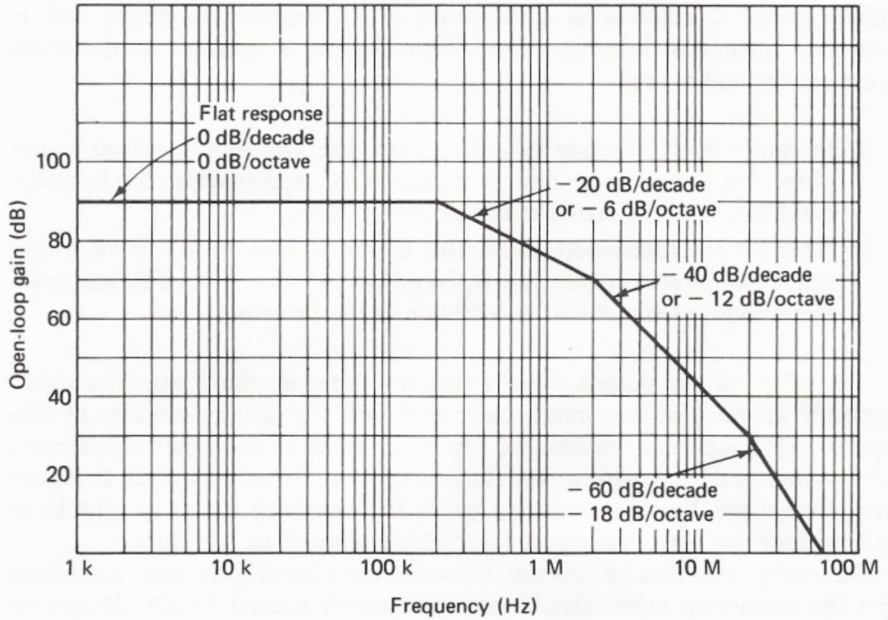


Figure 6-3 Approximation of open-loop gain vs. frequency curve of Fig. 6-1.

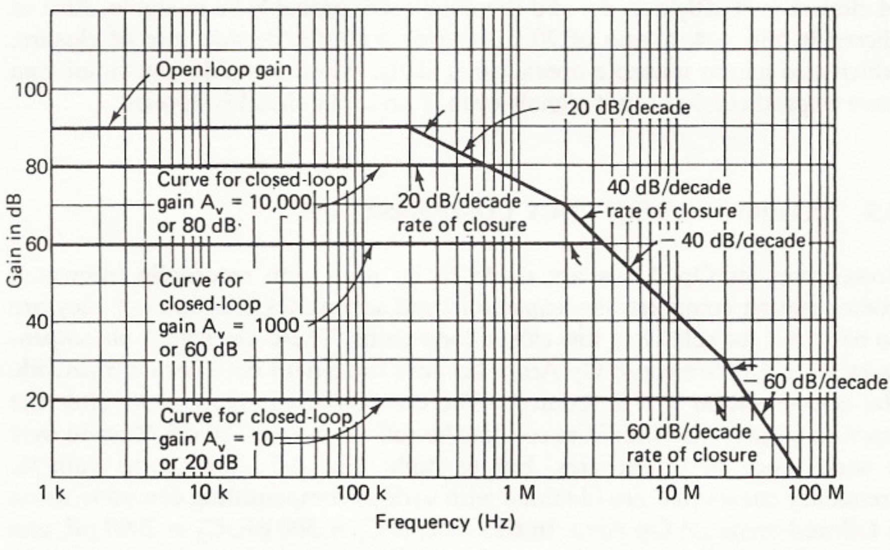


Figure 6-4

modifications. Oscillation is a symptom of an unstable amplifier and is discussed further in later chapters. With Op Amps specifically, the main causes of instability are:

1. In the −40-dB/decade roll-off region, the Op Amp's output signal phase lag is so large that it becomes, or approaches, an in-phase condition with the signal at the inverting input.
2. With lower closed-loop gains, the feedback resistor R_F is relatively small compared to resistance R_1 (see Figs. 3-5 and 3-10). This increases the amount of signal that is fed back to the inverting input.

The effect of the output signal becoming more in phase with the signal normally at the inverting input, combined with the greater amount of this output being fed back, makes the Op Amp oscillate at higher frequencies when wired to have relatively low closed-loop gain. In other words, at higher frequencies and lower closed-loop gains the feedback becomes significant and regenerative and thus meets the requirements of an oscillator.

Generally, the rate of closure between the closed-loop gain projection and the open-loop curve should not significantly exceed 20 dB/decade or 6 dB/octave for stable operation. Therefore the Op Amp with the characteristics in Fig. 6-4 is likely to be unstable if wired for gains below about 70 dB. Thus, as mentioned before, if this Op Amp is wired for 60-dB gain, the rate of closure is 40 dB/decade, and the circuit will probably be unstable. Also as shown in Fig. 6-4, a gain of 20 dB causes a 60-dB/decade rate of closure, which also means unstable operation is likely. When unstable, the circuit can have unpredictable output signals even if no input signal is applied.

6.3 EXTERNAL FREQUENCY COMPENSATION

Some types of Op Amps are made to be used with externally connected compensating components—capacitors and sometimes resistors—if they are to be wired for relatively low closed-loop gains. These are called *uncompensated* or *tailored-response* Op Amps because the circuit designer must provide the compensation if it is required. The compensating components alter the open-loop gain characteristics so that the roll-off is about 20 dB/decade over a wide range of frequencies. For example, Fig. 6-5 shows some gain vs. frequency curves that are obtained with various compensating components on a tailored-response Op Amp. In this case, if C_1 = 500 pF, C_2 = 2000 pF, and C_3 = 1000 pF, then the gain vs. frequency curve 1 applies according to the table, and the roll-off starts at about 1 kHz with a steady −20-dB/decade rate. Thus if feedback components are now added to obtain an 80-dB closed-loop gain, the circuit's frequency response is flat from dc to about 3 kHz. This can be seen if we project to the right of 80 dB to curve 1. The

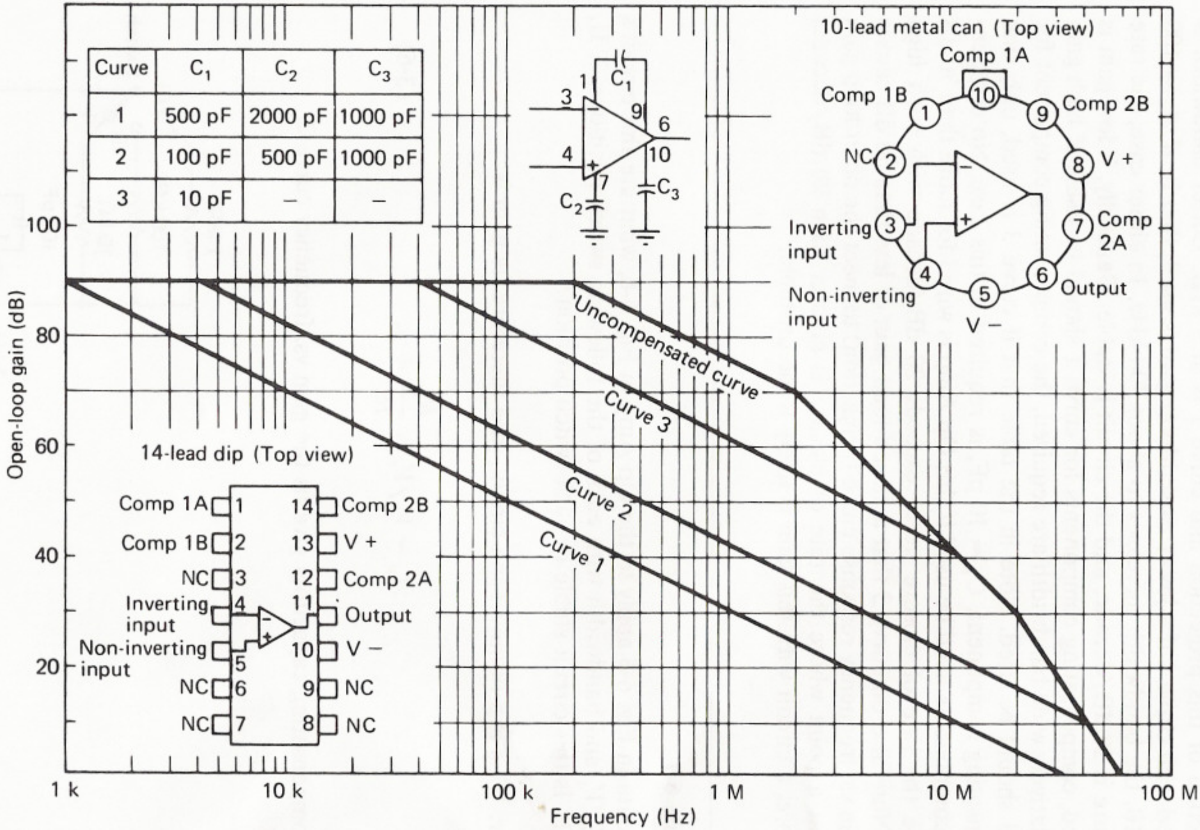


Figure 6-5 Open-loop gain vs. frequency characteristics with various externally connected (outboard) compensating components.

intersection of this projection and curve 1 is at 3 kHz. Better bandwidth is obtained with curve 1 if a lower closed-loop gain is used; that is, if $A_v = 100$ or 40 dB, the bandwidth increases to about 300 kHz. In either case, the rate of closure is 20 dB/decade, and the circuit is stable. Generally, if low gain is required, compensating components for curve 1 should be used. If high gain and relatively wide bandwidth are required, the compensating component for curve 3 should be used. Note in the table that if curve 3 is used, only one compensating component, $C_1 = 10$ pF, is required in this case. No compensating components need be used if this Op Amp is wired for more than 70 dB because the rate of closure never exceeds 20 dB/decade with such high gains. Note also on curve 3 that a closed-loop gain of less than 30 dB causes this gain vs. frequency response curve (projection) to meet the open-loop gain curve at a point where the rate of closure is greater than 20 dB/decade; therefore, a circuit with this gain is likely to be unstable.

Example 6-1

If the data in Fig. 6-5 apply to the Op Amp in Fig. 6-6, what are this circuit's gain V_o/V_s and bandwidth with each of the following switch positions: I, II, and III? Is this circuit stable in each switch position?

Answer. This is a noninverting circuit, and therefore its gain is

$$A_v = V_o/V_s \cong \frac{R_F}{R_1} + 1. \tag{3-6}$$

The compensating capacitors give us the gain vs. frequency curve 2.

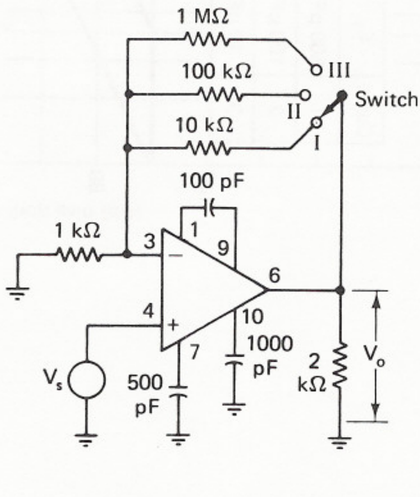


Figure 6-6 For Example 6-1.

With the switch in position I,

$$A_v \cong \frac{10\ \text{k}\Omega}{1\ \text{k}\Omega} + 1 = 11.$$

This is equivalent to about 20.8 dB or about 20 dB. Projecting to the right from 20 dB in Fig. 6-5, we intersect curve 2 at 13 MHz, which means that the bandwidth in this case is about 13 MHz.

With the switch in position II,

$$A_v \cong \frac{100\ \text{k}\Omega}{1\ \text{k}\Omega} + 1 = 101,$$

which is equivalent to about 40 dB. Projecting to the right of 40 dB we intersect curve 2 at about 1.3 MHz, and therefore the bandwidth now is about 1.3 MHz.

With the switch in position III,

$$A_v \cong \frac{1\ \text{M}\Omega}{1\ \text{k}\Omega} + 1 = 1001,$$

which is equivalent to about 60 dB. A projection to the right of 60 dB intersects curve 2 above 130 kHz, which means that the bandwidth now is 130 kHz.

Since all three projections intersect curve 2 at a -20-dB/decade rate of closure, the circuit is stable regardless of the switch position. Note that with these compensating components, if this Op Amp is wired as a voltage follower; that is, with a gain of less than 10 dB, it would be unstable.

6.4 COMPENSATED OP AMPS

Sometimes the relatively broad bandwidth of the uncompensated Op Amps is not needed. For example, in the instrumentation circuit shown in Fig. 5-5 of the previous chapter, the Op Amp is required to amplify relatively slowly changing signals, and therefore it does not require good high-frequency response. In this and similar applications, internally compensated Op Amps can be used. They are usually called *compensated* Op Amps. Also, they are stable regardless of the closed-loop gain and without externally connected compensating components.

The type 741 Op Amp is compensated and has an open-loop gain vs. frequency response as shown in Fig. 6-7. The IC of the 741 contains a 30-pF capacitance that internally shunts off signal current and thus reduces the available output signal at higher frequencies. This internal capacitance, which is an internal compensating component, causes the open-loop gain to

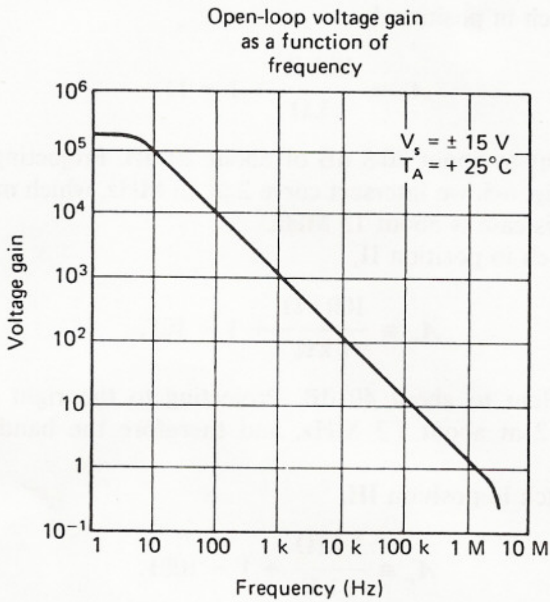


Figure 6-7 Frequency response of the 741 Op Amp.

roll off at a steady 20-dB/decade rate. Therefore, regardless of the closed-loop gain we use, the gain projection will intersect the open-loop gain curve at a 20-dB/decade rate of closure, and this assures us of a stable circuit.

The 741, along with other types of compensated Op Amps, has a 1-MHz *gain-bandwidth product*. This means that the product of the coordinates, gain and frequency, of any point on the open-loop gain vs. frequency curve is about 1 MHz.

Obviously, if the 741 Op Amp is wired for a closed-loop gain of 10^4 or 80 dB, its bandwidth is 100 Hz, as can be seen by projecting to the right from 10^4 to the curve in Fig. 6-7. If the closed-loop gain is lowered, say to 10^2 or to 1, the bandwidth increases to 10 kHz or 1 MHz, respectively. The fact that the 741 has a 1-MHz bandwidth with unity gain explains why the 741 is listed with a *unity gain-bandwidth* of 1 MHz on typical specification sheets.

Example 6-2

If the Op Amp in the circuit of Fig. 6-8 is a type 741 and the circuit is required to amplify signals up to about 10 kHz, what maximum value of POT resistance can we use and still keep the circuit's response flat to 10 kHz?

Answer. The frequency response curve in Fig. 6-7 shows that the bandwidth is greater than 10 kHz if the closed-loop gain is kept under 100. Therefore,

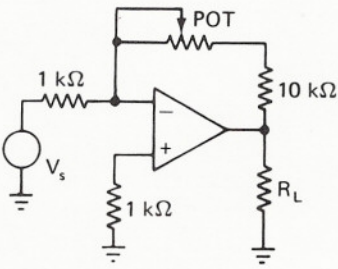


Figure 6-8 For Example 6-2.

we want to keep the ratio R_F/R_1 under 100. Thus since

$$\frac{R_F}{R_1} < 100, \qquad \text{then} \quad R_F < 100R_1 = 100(1\ \text{k}\Omega) = 100\ \text{k}\Omega.$$

If we must keep the total feedback resistance R_F under 100 kΩ, then the POT's maximum resistance is 90 kΩ. Note the 10-kΩ fixed resistor in series with it.

6.5 SLEW RATE

An Op Amp's slew rate is related to its frequency response. Generally, we can expect Op Amps with wider bandwidths to have higher (better) slew rates. The slew rate, as mentioned in Chapter 2, is the rate of output-voltage change caused by a step input voltage and is usually measured in volts per microsecond. An ideal slew rate is infinite, which means that the Op Amp's output voltage should change instantly in response to an input step voltage. Practical IC Op Amps have specified slew rates from 0.1 V/μs to about 100 V/μs which are measured in special circumstances. Some hybrid* Op Amps have slew rates on the order of 1000 V/μs. Unless otherwise specified, the slew rate listed in a data sheet was probably measured with unity gain and open load.

A less than ideal slew rate causes distortion especially noticeable with nonsinusoidal waveforms at higher frequencies. For example, ignoring overshoot, Fig. 6-9 shows typical output waveforms of a voltage follower with square waves of various frequencies applied. In this case, the slew rate is 1 V/μs. When the input frequency is 100 Hz with the waveform in Fig. 6-9b, the output has the waveform shown in part c of this figure. Similarly, when the input is 10 kHz or 1 MHz with the waveform shown in b, the output has the waveform in Fig. 6-9d or e, respectively. Obviously, due to the limited

*Hybrids, as opposed to monolithics (made on one chip), might contain one or more ICs or ICs and discrete components.

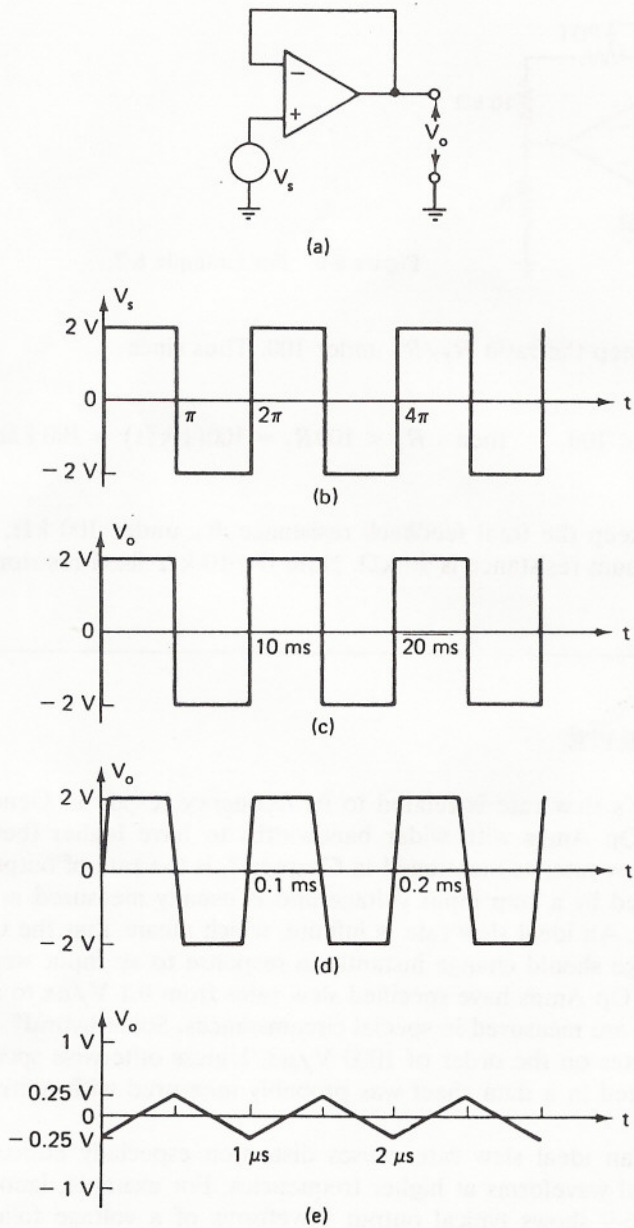
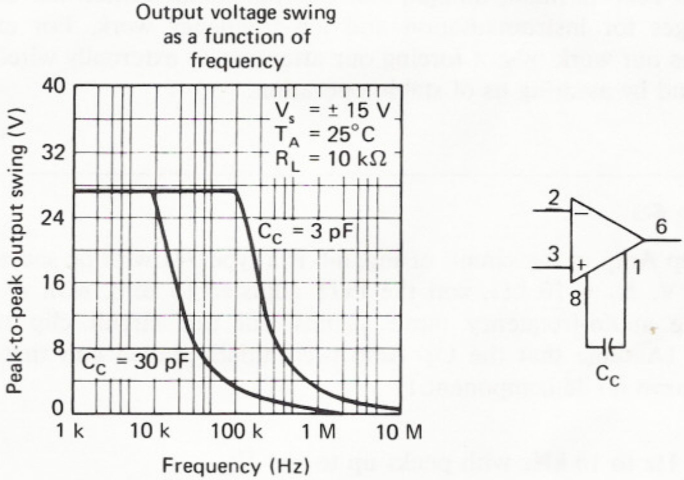


Figure 6-9 Voltage follower with an Op Amp having a slew rate of 1 V/μs and square-wave voltage applied. (c) Output when the frequency of V_s is 100 Hz. (d) Output when the frequency of V_s is 10 kHz. (e) Output when the frequency of V_s is 1 MHz.

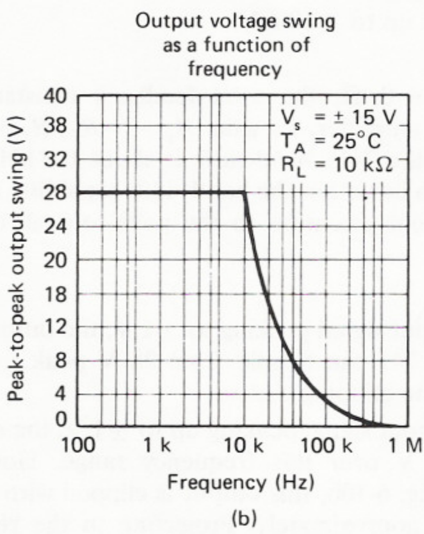
slew rate, the square-wave input is distorted into a sawtooth at higher frequencies.

6.6 OUTPUT SWING CAPABILITY VS. FREQUENCY

An important consideration is an Op Amp's peak-to-peak output signal capability at higher frequencies. Generally, this capability decreases with higher frequencies, as shown in Fig. 6-10. In Fig. 6-10a, an uncompensated



(a)



(b)

Figure 6-10 Peak-to-peak output voltage capability vs. frequency of the (a) uncompensated type 777 Op Amp; (b) compensated type 741 Op Amp.

Op Amp's characteristics are shown for two different compensating capacitors C_c. Note that if $C_c = 30$ pF is used, the peak-to-peak output capability decreases sharply at operating frequencies above 10 kHz. On the other hand, when $C_c = 3$ pF is wired externally (outboard), the bandwidth increases, as it did with smaller compensating capacitors shown in Fig. 6-5. Also, the peak-to-peak output capability does not drop until frequencies on the order of 100 kHz or more are applied.

The output vs. frequency curve in Fig. 6-10b is for the type 741 Op Amp, which is internally compensated. Apparently then, internal compensation not only limits bandwidth, but it also limits the peak-to-peak output capability. We must keep in mind, though, that internal compensation has outstanding advantages for instrumentation and low-frequency work. For example, it simplifies our work by not forcing our attention to externally wired compensation and by assuring us of stable operation.

Example 6-3

If the Op Amp in the circuit of Fig. 6-8 is a type 741 with dc source voltages of ± 15 V, $R_L = 10$ kΩ, and the POT adjusted to zero, with which of the following audio-frequency input signals will this circuit clip the output signals? (Assume that the Op Amp was initially nulled and that the input signals have no dc component.)

(a) 1 Hz to 10 kHz with peaks up to ± 1 V;
(b) 1 Hz to 100 kHz with peaks up to ± 1 V;
(c) 1 Hz to 100 kHz with peaks up to ± 10 mV.

Answer. With the POT resistance 0 Ω, the total feedback resistance is 10 kΩ. Since this is an inverting amplifier, its gain $A_v = -R_F/R_1 = -10$ kΩ/1 kΩ $= -10$. With this gain, the 741's bandwidth is about 100 kHz. See Fig. 6-7. All three signals appear to be within the bandwidth capability of this circuit. Therefore, we can focus our attention to the peak-to-peak output capability vs. frequency.

(a) With the 1-Hz to 10-kHz input signal peaking to ± 1 V, the output will peak to ± 10 V. Since the 741 can handle from 28 V peak to peak (14-V peak) up to 10 kHz, no clipping occurs.
(b) With the 1-Hz to 100-kHz input signal peaking up to ± 1 V, the output *attempts* to peak at ± 10 V over this frequency range. However, according to the curve in Fig. 6-10b, this output is clipped with signal frequencies above 15 kHz approximately. Projecting to the right of 20 V peak to peak (10-V peak), we intersect the curve at roughly

15 kHz. Thus input signals with peaks of 1 V are clipped more or less, depending on how far above 15 kHz the signal is.

(c) With the 1-Hz to 100-kHz input signal peaking up to ± 10 mV, the output peaks at ± 100 mV. No clipping occurs because the 741's peak-to-peak output capability is about 2-V or 1-V peak up to 100 kHz. This capability is well over the ± 100-mV peaks we intend to get.

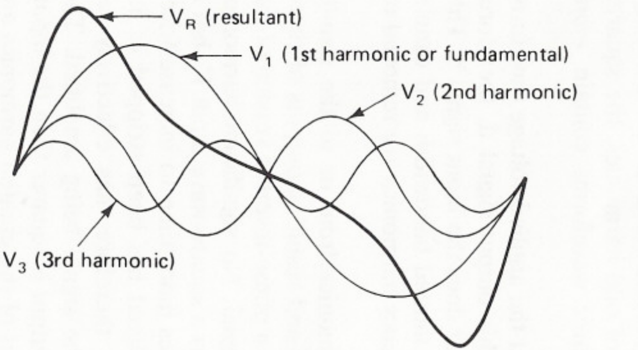
6.7 HARMONICS IN NONSINUSOIDAL WAVEFORMS

The signals processed by Op Amps often are nonsinusoidal waveforms. If such waveforms are repetitious, they can be shown to consist of some *fundamental* sine-wave frequency and one or more of its *harmonics*. Harmonics are also sine waves with frequencies that are *integer* multiples of the fundamental. For example, if a nonsinusoidal waveform repeats every 1 ms, its fundamental frequency $f = 1/T = 1/1$ ms $= 1$ kHz. Its second harmonic is 2 kHz, its third harmonic is 3 kHz, etc. Examples of nonsinusoidal waveforms and their harmonic content are shown in Fig. 6-11.

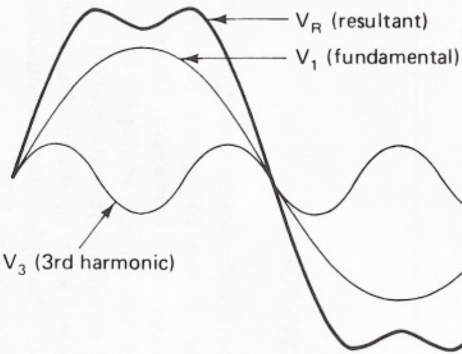
The waveforms b through d of Fig. 6-11 are of particular interest to us. Note in these cases that the *resulting* nonsinusoidal waveforms V_R are somewhat square and that they consist of *odd*-numbered harmonics only. Especially note that the greater the number of odd harmonics, the squarer the resulting voltage V_R is. Generally, squarer waveforms contain *more* odd-numbered harmonics.

Referring back to Fig. 6-9, we observed that the applied voltage waveform b is squarer than the output waveform d. This output signal d, therefore, contains fewer higher-frequency harmonics that does the input signal b. This is caused by the fact that an Op Amp has a limited bandwidth and cannot pass high frequencies. Thus, the higher-frequency harmonics are removed or attenuated resulting in the less square output.

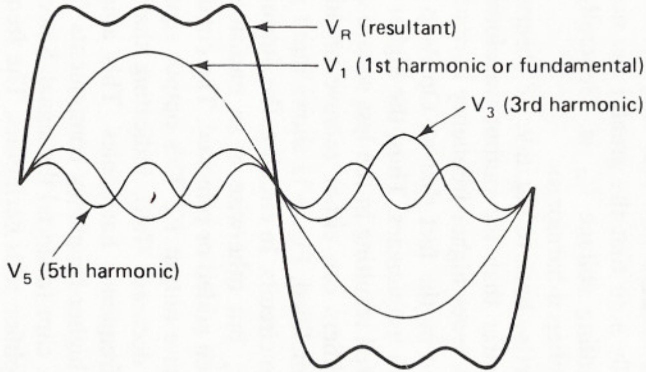
Amplifiers can either remove or add harmonics from or to the signals being amplified. Fig. 6-12 shows input signals and resulting outputs of three Op Amp circuits. In circuit a, the output V_o is a phase-inverted version of the input V_s but otherwise is an undistorted signal. No significant harmonics have been added or removed. The circuit b has a square-wave input V_s but a less square output V_o. This output signal takes more time to increase (rise) and to decrease (fall), indicating that the signal has been stripped of its higher-frequency harmonics. This amplifier, therefore, has effectively removed higher-frequency components from the signal being amplified. The circuit c clips (distorts) the signal. Since the output is squarer than the input, this amplifier adds harmonics. The frequencies of the created harmonics are mainly odd-integer multiples of the input signal frequency.



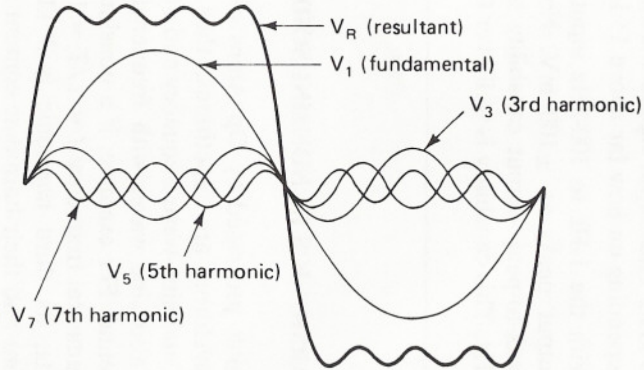
(a) $V_R = V_1 + V_2 + V_3 = \sin \omega t + \frac{1}{2} \sin 2\omega t + \frac{1}{3} \sin 3\omega t$



(b) $V_R = V_1 + V_3 = \sin \omega t + \frac{1}{3} \sin 3\omega t$



(c) $V_R = V_1 + V_3 + V_5 = \sin \omega t + \frac{1}{3} \sin \omega t + \frac{1}{5} \sin 5\omega t$



(d) $V_R = V_1 + V_3 + V_5 + V_7 = \sin \omega t + \frac{1}{3} \sin 3\omega t + \frac{1}{5} \sin 5\omega t + \frac{1}{7} \sin 7\omega t$

Figure 6-11 Examples of nonsinusoidal waveforms V_R and their harmonics.

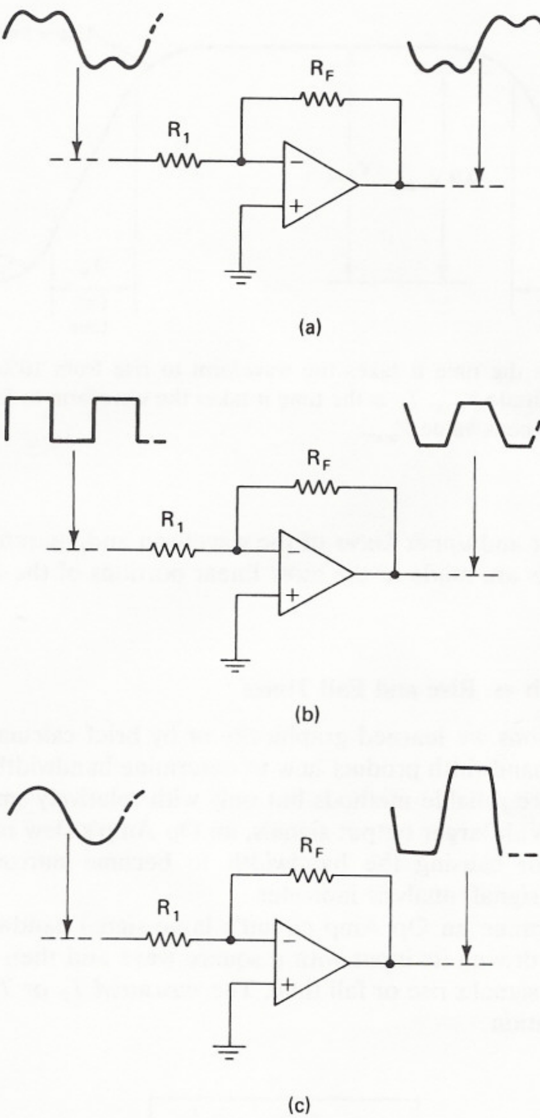


Figure 6-12 Amplifier (a) neither adds nor removes harmonics significantly; amplifier (b) removes higher-frequency harmonics; and amplifier (c) adds harmonics.

6.7-1 Rise and Fall Times of Nonsinusoidal Waveforms

The terms *rise time* T_R and *fall time* T_F are commonly used to describe the characteristics of nonsinusoidal waveforms. As shown in Fig. 6-13, the rise time T_R is defined as the time it takes a voltage to rise from 10% to 90% of its peak-to-peak amplitude $V_{\max}$. Similarly, the fall time T_F is the time it takes the waveform to fall from 90% to 10% of $V_{\max}$. These definitions

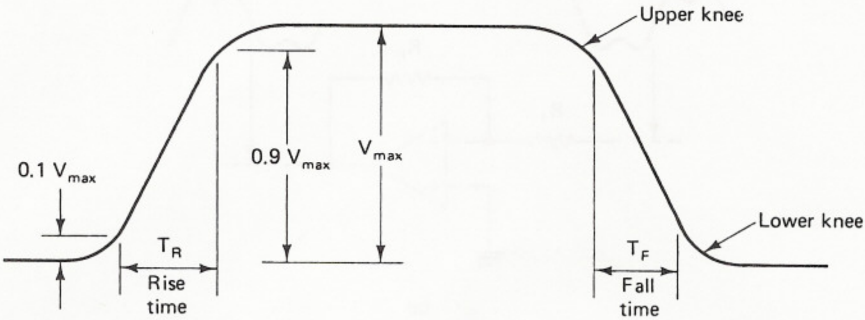


Figure 6-13 T_R is the time it takes the waveform to rise from 10% to 90% of its peak-to-peak amplitude V_{max}. T_F is the time it takes the waveform to fall from 90% to 10% of its peak-to-peak value V_{max}.

exclude the lower and upper *knees* of the waveform and, therefore, the T_R or T_F measurements are made in the most linear portions of the rise or fall.

6.7-2 Bandwidth vs. Rise and Fall Times

In previous sections we learned graphically or by brief calculations with an Op Amp's gain-bandwidth product how to determine bandwidths of Op Amp circuits. These are reliable methods but only with relatively small amplitude output signals. With larger output signals, an Op Amp's slew rate becomes a dominating factor causing the bandwidth to become narrower than the graphical (small-signal) analysis indicates.

We can determine an Op Amp circuit's large-signal bandwidth BW two ways. One is by driving its input with a square wave and then observing the resulting output signal's rise or fall time. The *measured* T_F or T_F can then be used in the equation

$$\boxed{\mathrm{BW} \cong \frac{0.35}{T_R} = \frac{0.35}{T_F}.} \qquad (6\text{-}1)^*$$

The second method is to calculate the rise or fall time using the Op Amp's specified slew rate. The *calculated* T_R or T_F can then be used with the foregoing equation to determine the bandwidth BW.

*See Appendix I for derivation.

Example 6-4

A square wave, having an extremely fast rise and fall time, is applied to the input of an Op Amp circuit. Without saturating the Op Amp, the resulting output is a trapezoidal waveform, like that of Fig. 6-9d. By adjustment of the sweep time on a triggered sweep oscilloscope, we observe the leading edge of the trapezoid as shown in Fig. 6-14. If the SWEEP TIME control is in the CAL 10-μs/DIV position, what is this circuit's bandwidth?

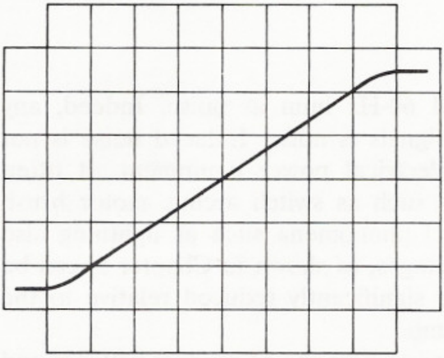


Figure 6-14 Rise time T_R measured on an oscilloscope.

Answer. The rise time T_R takes about six horizontal divisions. Since each division is 10 μs.

$$T_R = 6(10\ \mu\text{s}) = 60\ \mu\text{s}.$$

Using this measured T_R in Eq. (6-1) we find that the bandwidth

$$\text{BW} \cong \frac{0.35}{60\ \mu\text{s}} \cong 5.8\text{ kHz}.$$

Example 6-5

A 741 is to output unclipped signals as large as 20 V peak to peak. What bandwidth can we expect with this output amplitude? The 741 has a 0.5 v/μs slew rate.

Answer. Since the 741 has a slew rate of about 0.5 V/μs, the total time T_{tot} it takes the output voltage to rise by 20 V is

$$T_{\text{tot}} = \frac{20\text{ V}}{0.5\text{ V}/\mu\text{s}} = 40\ \mu\text{s}.$$

The rise time T_R is about 80% of this because the first and last 10% of T_{tot} is not included in the definition of T_R. In this case, then,

$$T_R \cong 0.8(40\ \mu\text{s}) = 32\ \mu s,$$

and, therefore, the bandwidth

$$\text{BW} \cong \frac{0.35}{32\ \mu\text{s}} \cong 10.9\ \text{kHz}. \tag{6-1}$$

6.8 NOISE

In Chapter 5 we referred to induced 60-Hz hum as noise. Indeed, any unwanted signal mixing with desired signals is noise. Induced noise is not limited to the 60 Hz from nearby electrical power equipment. It often originates in other man-made systems such as switch arcing, motor brush sparking, and ignition systems. Natural phenomena such as lightning also cause induced noise. Induced noise voltages, as shown in Chapter 5, can be made common mode and thus can be significantly reduced relative to the desired signals at the load of an Op Amp.

The term *noise* is also commonly used to describe ac random voltages and currents generated within conductors and semiconductors. Such noise, associated with Op Amps and with amplifiers in general, limit their signal sensitivity. If very weak signals are to be amplified, very high closed-loop gain must be used to bring the signals up to useful levels. With a very high gain, however, the noise is amplified along with the signals to the point where nearly as much noise as signal appears at the output. If fed to a speaker, random noise causes a hissing, frying sound.

There are three main types of noise phenomena associated with Op Amps and with solid-state amplifiers in general. These are: *thermal* or *Johnson* noise, *shot* or *Schottky* noise, and *flicker* or $1/f$ noise.

Thermal noise is caused by the random motion of charge carriers within a conductor which generate noise voltages V_n within it. The rms value of this thermally generated noise voltage V_n can be predicted with the equation

$$\boxed{V_n = \sqrt{4KTR(\text{BW})}\,,} \tag{6-2}$$

where K is Boltzmann's constant; 1.38×10^{-23} joules/°K,

T is the temperature in degrees Kelvin—the Celsius temperature plus 273°,

R is the resistance of the conductor in question, and

BW is the bandwidth in hertz.

Examining this equation, we see that thermal noise increases with higher temperatures, larger resistances, and wider bandwidths.

Although a constant average dc current may be maintained in a semiconductor, it will have random variations. These variations have an rms value referred to as noise current I_n. Noise generated in this manner is called *shot noise*. Its rms value can be predicted with the equation

$$\boxed{I_n = \sqrt{2qI_{\text{dc}}(\text{BW})}\,,} \tag{6-3}$$

where q is the charge of an electron: 1.6×10^{-19} coulombs,
I_{dc} is the average dc current in the semiconductor, and
BW is the bandwidth.

Here again, wider bandwidths generate more noise, which is to say, we can expect less noise in narrow-bandwidth amplifiers. Of course, the noise current I_n, flowing through a resistance R, will generate noise voltage RI_n.

In addition to shot noise, semiconductors have low-frequency noise called *flicker* or $1/f$ noise. The term $1/f$ describes the inverse nature of this noise with respect to frequency; that is, the amount of flicker noise is greater at lower frequencies f.

6.9 EQUIVALENT INPUT NOISE MODEL

An Op Amp contains many active and passive components that generate and add noise to its output. These noise sources can be represented by voltage and current noise generators at the input of the amplifier's equivalent circuit, as shown in Fig. 6-15. This equivalent is sometimes called an *amplifier noise model*. The net effect of these input noise generators is an *equivalent*

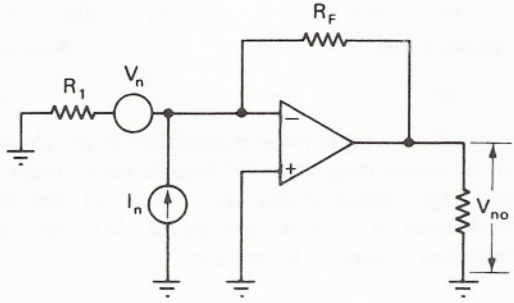


Figure 6-15 An amplifier's inherent noise voltage and current can be shown as voltage and current generators at its inputs.

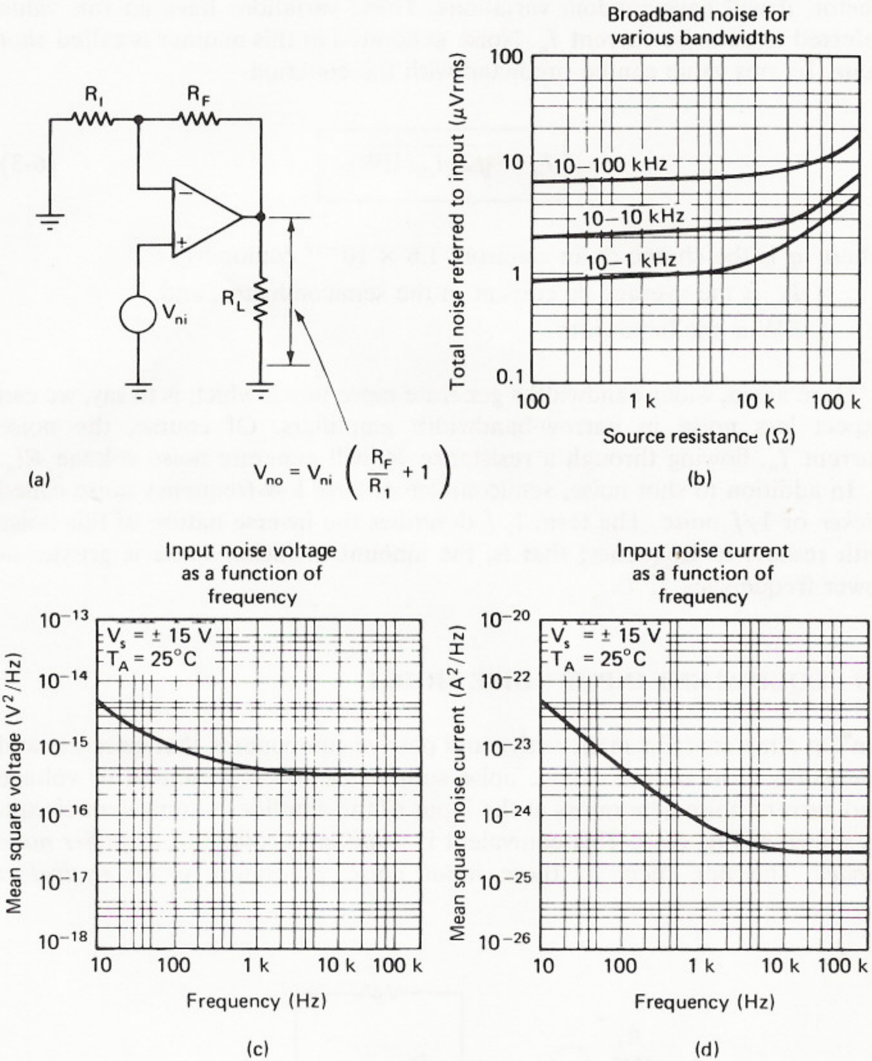


Figure 6-16 The total effective input noise is amplified by the closed-loop gain of the amplifier seeing it as a noninverting type. (b) Typical noise levels higher with wider bandwidths and with larger signal source resistances. (c) The rms noise voltages squared typically developed at various frequencies. (d) The rms noise currents squared typically developed at various frequencies.

input-noise voltage V_{ni}. It is this equivalent input noise V_{ni} that is amplified by the stage gain, and with it we can predict the output noise V_{no} that a given Op Amp stage might have.

If we neglect the thermal noise of the amplifier's signal source, the equivalent input noise V_{ni} is related to the amplifier's specified input noise voltage V_n and noise current I_n by the following equation:

$$\boxed{V_{ni}^2 \cong V_n^2 + (R_E I_n)^2,} \tag{6-4}$$

where V_{ni}^2 is the mean-square equivalent input noise voltage.

V_n^2 is the specified mean-square input noise voltage.

I_n^2 is the specified mean-square input noise current, and R_E is the parallel equivalent of R_1 and R_F.

It can be shown that equivalent input noise voltage V_{ni} sees the Op Amp circuit as a noninverting type as shown in Fig. 6-16. Broadband input noise vs. source resistance curves are sometimes shown as in Fig. 6-16b, and the general-purpose Op Amp's specified noise voltage V_n and input noise current I_n vs. frequency characteristics are shown in Fig. 6-16c and d. Larger source resistances appreciably add to thermally generated noise as shown in part b of the figure. The $1/f$ phenomenon adds low-frequency noise as shown in Fig. 6-16c and d.

The output noise voltage V_{no} is a function of the effective input noise voltage V_{ni} and is approximated with the equation

$$\boxed{V_{no}^2 \cong (A_v V_{ni})^2} \tag{6-5a}$$

or

$$\boxed{V_{no} \cong A_v \sqrt{V_n^2 + (R_E I_n)^2},} \tag{6-5b}$$

where A_v is the closed-loop gain.