Chapter 10
Op Amp Noise Theory and Applications

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TEXAS INSTRUMENTS

Chapter 10

Op Amp Noise Theory and Applications

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10.1 Introduction

The purpose of op amp circuitry is the manipulation of the input signal in some fashion. Unfortunately in the real world, the input signal has unwanted noise superimposed on it.

Noise is not something most designers get excited about. In fact, they probably wish the whole topic would go away. It can, however, be a fascinating study by itself. A good understanding of the underlying principles can, in some cases, be used to reduce noise in the design.

10.2 Characterization

Noise is a purely random signal, the instantaneous value and/or phase of the waveform cannot be predicted at any time. Noise can either be generated internally in the op amp, from its associated passive components, or superimposed on the circuit by external sources. External noise is covered in Chapter 17, and is usually the dominant effect.

10.2.1 rms versus P-P Noise

Instantaneous noise voltage amplitudes are as likely to be positive as negative. When plotted, they form a random pattern centered on zero. Since noise sources have amplitudes that vary randomly with time, they can only be specified by a probability density function. The most common probability density function is Gaussian. In a Gaussian probability function, there is a mean value of amplitude, which is most likely to occur. The probability that a noise amplitude will be higher or lower than the mean falls off in a bell-shaped curve, which is symmetrical around the center (Figure 10–1).

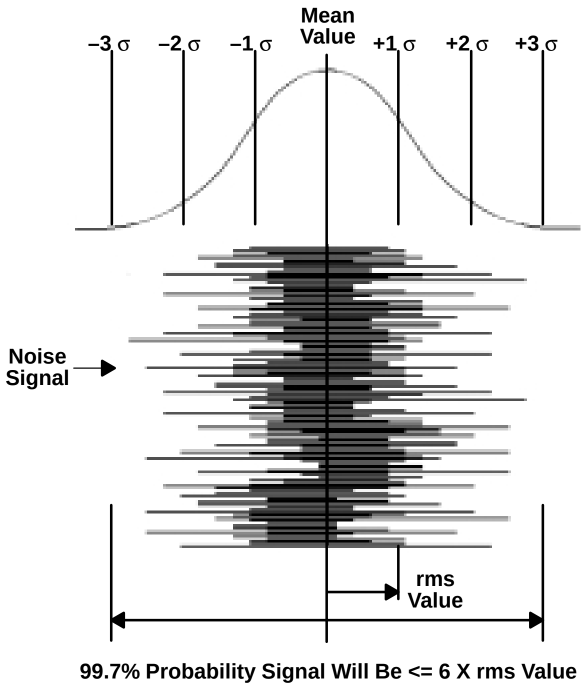

Figure 10–1. Gaussian Distribution of Noise Energy

σ is the standard deviation of the Gaussian distribution and the rms value of the noise voltage and current. The instantaneous noise amplitude is within ±1σ 68% of the time. Theoretically, the instantaneous noise amplitude can have values approaching infinity. However, the probability falls off rapidly as amplitude increases. The instantaneous noise amplitude is within ±3σ of the mean 99.7% of the time. If more or less assurance is desired, it is between ±2σ 95.4% of the time and ±3.4σ 99.94% of the time.

σ^2 is the average mean-square variation about the average value. This also means that the average mean-square variation about the average value, $\overline{i^2}$ or $\overline{e^2}$, is the same as the variance σ^2.

Thermal noise and shot noise (see below) have Gaussian probability density functions. The other forms of noise do not.

10.2.2 Noise Floor

When all input sources are turned off and the output is properly terminated, there is a level of noise called the *noise floor* that determines the smallest signal for which the circuit is useful. The objective for the designer is to place the signals that the circuit processes above the noise floor, but below the level where the signals will clip.

10.2.3 Signal-to-Noise Ratio

The noisiness of a signal is defined as:

$$\frac{S_{(f)}}{N_{(f)}} = \frac{\text{rms signal voltage}}{\text{rms noise voltage}} \tag{10–1}$$

In other words, it is a ratio of signal voltage to noise voltage (hence the name *signal-to-noise ratio*).

10.2.4 Multiple Noise Sources

When there are multiple noise sources in a circuit, the total root-mean-square (rms) noise signal that results is the square root of the sum of the average mean-square values of the individual sources:

$$E_{Totalrms} = \sqrt{e^2_{1rms} + e^2_{2rms} + \ \ldots e^2_{nrms}} \tag{10–2}$$

Put another way, this is the only “break” that the designer gets when dealing with noise. If there are two noise sources of equal amplitude in the circuit, the total noise is not doubled (increased by 6 dB). It only increases by 3 dB. Consider a very simple case, two noise sources with amplitudes of 2 V_{rms}:

$$E_{Totalrms} = \sqrt{2^2 + 2^2} = \sqrt{8} = 2.83\ V_{rms} \tag{10–3}$$

Therefore, when there are two equal sources of noise in a circuit, the noise is $20 \times \log\frac{2.83}{2} = 3.01$ dB higher than if there were only one source of noise — instead of double (6 dB) as would be intuitively expected.

This relationship means that the worst noise source in the system will tend to dominate the total noise. Consider a system in which one noise source is 10 V_{rms} and another is 1 V_{rms}:

$$E_{Totalrms} = \sqrt{10^2 + 1^2} = \sqrt{101} = 10.05\ V_{rms} \tag{10–4}$$

There is hardly any effect from the 1-V noise source at all!

10.2.5 Noise Units

Noise is normally specified as a spectral density in rms volts or amps per root Hertz, $V/\sqrt{Hz}$ or $A/\sqrt{Hz}$. These are not very "user-friendly" units. A frequency range is needed to relate these units to actual noise levels that will be observed.

For example:

- A TLE2027 op amp with a noise specification of $2.5\ nV/\sqrt{Hz}$ is used over an audio frequency range of 20 Hz to 20 kHz, with a gain of 40 dB. The output voltage is 0 dBV (1 V).
- To begin with, calculate the *root Hz* part: $\sqrt{20000 - 20} = 141.35$.
- Multiplying this by the noise spec: $2.5 \times 141.35 = 353.38$ nV, which is the equivalent input noise (EIN). The output noise equals the input noise multiplied by the gain, which is 100 (40 dB).

The signal-to-noise ratio can be now be calculated:

- $353.38\ \text{nV} \times 100 = 35.3\ \mu\text{V}$
- Signal-to-noise (dB) =

$$20 \times \log(1\text{V} \div 35.3\ \mu\text{V}) = 20 \times \log(28329) = 89\ \text{dB} \tag{10–5}$$

The TLE2027 op amp is an excellent choice for this application. Remember, though, that passive components and external noise sources can degrade performance. There is also a slight increase in noise at low frequencies, due to the 1/f effect (see below).

10.3 Types of Noise

There are five types of noise in op amps and associated circuitry:

1) Shot noise
2) Thermal noise
3) Flicker noise
4) Burst noise
5) Avalanche noise

Some or all of these noises may be present in a design, presenting a noise spectrum unique to the system. It is not possible in most cases to separate the effects, but knowing general causes may help the designer optimize the design, minimizing noise in a particular bandwidth of interest. Proper design for low noise may involve a "balancing act" between these sources of noise and external noise sources.

10.3.1 Shot Noise

The name *shot noise* is short for Schottky noise. Sometimes it is referred to as *quantum noise*. It is caused by random fluctuations in the motion of charge carriers in a conductor. Put another way, current flow is not a continuous effect. Current flow is electrons, charged particles that move in accordance with an applied potential. When the electrons encounter a barrier, potential energy builds until they have enough energy to cross that barrier. When they have enough potential energy, it is abruptly transformed into kinetic energy as they cross the barrier. A good analogy is stress in an earthquake fault that is suddenly released as an earthquake.

As each electron randomly crosses a potential barrier, such as a pn junction in a semiconductor, energy is stored and released as the electron encounters and then shoots across the barrier. Each electron contributes a little *pop* as its stored energy is released when it crosses the barrier (Figure 10–2).

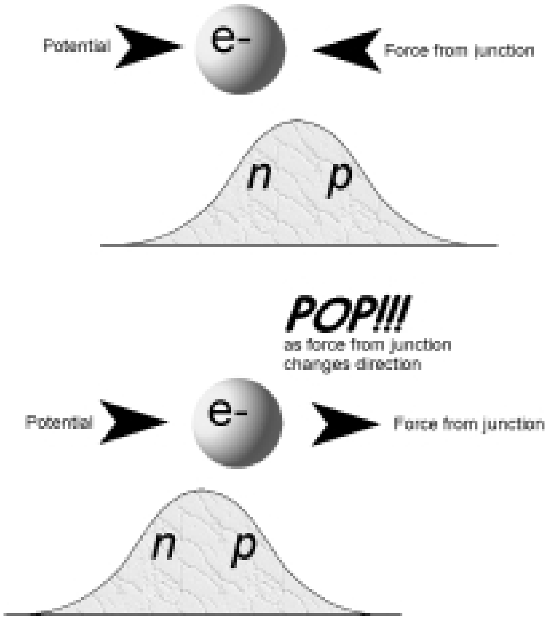

Figure 10–2. Shot Noise Generation

The aggregate effect of all of the electrons shooting across the barrier is the shot noise. Amplified shot noise has been described as sounding like lead shot hitting a concrete wall.

Some characteristics of shot noise:

- Shot noise is always associated with current flow. It stops when the current flow stops.
- Shot noise is independent of temperature.

- Shot noise is spectrally flat or has a uniform power density, meaning that when plotted versus frequency it has a constant value.
- Shot noise is present in any conductor — not just a semiconductor. Barriers in conductors can be as simple as imperfections or impurities in the metal. The level of shot noise, however, is very small due to the enormous numbers of electrons moving in the conductor, and the relative size of the potential barriers. Shot noise in semiconductors is much more pronounced.

The rms shot noise current is equal to:

$$I_{sh} = \sqrt{(2qI_{dc} + 4qI_0)B} \qquad (10\text{–}6)$$

Where:
q = Electron charge (1.6×10^{-19} coulombs)
I_{dc} = Average forward dc current in A
I_o = Reverse saturation current in A
B = Bandwidth in Hz

If the pn junction is forward biased, I_o is zero, and the second term disappears. Using Ohm's law and the dynamic resistance of a junction,

$$r_d = \frac{kT}{qI_{dc}} \qquad (10\text{–}7)$$

the rms shot noise voltage is equal to:

$$E_{sh} = kT\sqrt{\frac{2B}{qI_{dc}}} \qquad (10\text{–}8)$$

Where:
k = Boltzmann's constant (1.38×10^{-23} Joules/°K)
q = Electron charge (1.6×10^{-19} coulombs)
T = Temperature in °K
I_{dc} = Average dc current in A
B = Bandwidth in Hz

For example, a junction carries a current of 1 mA at room temperature. Its noise over the audio bandwidth is:

$$E_{sh} = 1.38 \times 10^{-23} \times 298\sqrt{\frac{2(20000 - 20)}{(1.6 \times 10^{-19}) \times (1 \times 10^{-3})}} = 65\text{ nV} = -144\text{ dBV} \qquad (10\text{–}9)$$

Obviously, it is not much of a problem in this example.

Look closely at the formula for shot noise voltage. Notice that the shot noise voltage is inversely proportional to the current. Stated another way, shot noise voltage decreases as average dc current increases, and increases as average dc current decreases. This can be an elegant way of determining if shot noise is a dominant effect in the op amp circuit being designed. If possible, decrease the average dc current by a factor of 100 and see if the overall noise increases by a factor of 10. In the example above:

$$E_{sh} = 1.38 \times 10^{-23} \times 298 \sqrt{\frac{2(20000 - 20)}{(1.6 \times 10^{-19}) \times (1 \times 10^{-5})}} = 650\ \text{nV} = -124\ \text{dBV} \qquad (10\text{–}10)$$

The shot noise voltage does increase by a factor of 10, or 20 dB.

10.3.2 Thermal Noise

Thermal noise is sometimes referred to as Johnson noise after its discoverer. It is generated by thermal agitation of electrons in a conductor. Simply put, as a conductor is heated, it will become noisy. Electrons are never at rest; they are always in motion. Heat disrupts the electrons' response to an applied potential. It adds a random component to their motion (Figure 10–3). Thermal noise only stops at absolute zero.

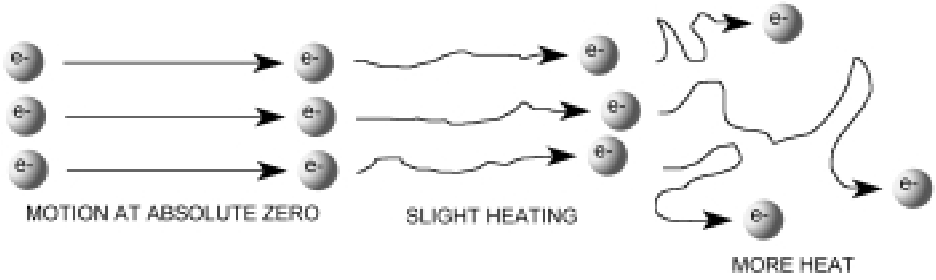

Figure 10–3. Thermal Noise

Like shot noise, thermal noise is spectrally flat or has a uniform power density (it is *white*), but thermal noise is independent of current flow.

At frequencies below 100 MHz, thermal noise can be calculated using Nyquist's relation:

$$E_{th} = \sqrt{4kTRB} \qquad (10\text{–}11)$$

or

$$I_{th} = \sqrt{\frac{4kTB}{R}} \qquad (10–12)$$

Where:

E_{th} = Thermal noise voltage in Volts rms
I_{th} = Thermal noise current in Amps rms
k = Boltzmann's constant (1.38×10^{-23})
T = Absolute temperature (Kelvin)
R = Resistance in ohms
B = Noise bandwidth in Hertz (f_{max}–f_{min})

The noise from a resistor is proportional to its resistance and temperature. It is important not to operate resistors at elevated temperatures in high gain input stages. Lowering resistance values also reduces thermal noise.

For example:

The noise in a 100 kΩ resistor at 25°C (298°K) over the audio frequency range of 20 Hz to 20 kHz is:

$$\begin{aligned} E_{th} &= \sqrt{4kTRB} \\ &= \sqrt{4 \times (1.38 \times 10^{-23}) \times 298 \times 100,000 \times (20,000 - 20)} \\ &= 5.73\ \mu V \\ &= –104.8\ dBV \end{aligned} \qquad (10–13)$$

Decreasing the temperature would reduce the noise slightly, but scaling the resistor down to 1 kΩ (a factor of 100) would reduce the thermal noise by 20 dB. Similarly, increasing the resistor to 10 MΩ would increase the thermal noise to –84.8 dBV, a level that would affect a 16-bit audio circuit. The noise from multiple resistors adds according to the root-mean-square law in Paragraph 10.2.4. Beware of large resistors used as the input resistor of an op amp gain circuit, their thermal noise will be amplified by the gain in the circuit (Paragraph 10.4). Thermal noise in resistors is often a problem in portable equipment, where resistors have been scaled up to get power consumption down.

10.3.3 Flicker Noise

Flicker noise is also called *1/f noise*. Its origin is one of the oldest unsolved problems in physics. It is pervasive in nature and in many human endeavors. It is present in all active and many passive devices. It may be related to imperfections in crystalline structure of semiconductors, as better processing can reduce it.

Some characteristics of flicker noise:

- It increases as the frequency decreases, hence the name *1/f*
- It is associated with a dc current in electronic devices
- It has the same power content in each octave (or decade)

$$E_n = K_v \sqrt{\left(\ln \frac{f_{max}}{f_{min}}\right)} \qquad I_n = K_i \sqrt{\left(\ln \frac{f_{max}}{f_{min}}\right)} \qquad (10–14)$$

Where:
K_e and K_i are proportionality constants (volts or amps) representing E_n and I_n at 1 Hz
f_{max} and f_{min} are the minimum and maximum frequencies in Hz

Flicker noise is found in carbon composition resistors, where it is often referred to as excess noise because it appears in addition to the thermal noise that is there. Other types of resistors also exhibit flicker noise to varying degrees, with wire wound showing the least. Since flicker noise is proportional to the dc current in the device, if the current is kept low enough, thermal noise will predominate and the type of resistor used will not change the noise in the circuit.

Reducing power consumption in an op amp circuit by scaling up resistors may reduce the 1/f noise, at the expense of increased thermal noise.

10.3.4 Burst Noise

Burst noise, also called popcorn noise, is related to imperfections in semiconductor material and heavy ion implants. It is characterized by discrete high-frequency pulses. The pulse rates may vary, but the amplitudes remain constant at several times the thermal noise amplitude. Burst noise makes a popping sound at rates below 100 Hz when played through a speaker — it sounds like popcorn popping, hence the name. Low burst noise is achieved by using clean device processing, and therefore is beyond the control of the designer. Modern processing techniques at Texas Instruments has all but eliminated its occurrence.

10.3.5 Avalanche Noise

Avalanche noise is created when a pn junction is operated in the reverse breakdown mode. Under the influence of a strong reverse electric field within the junction’s depletion region, electrons have enough kinetic energy that, when they collide with the atoms of the crystal lattice, additional electron-hole pairs are formed (Figure 10–4). These collisions are purely random and produce random current pulses similar to shot noise, but much more intense.

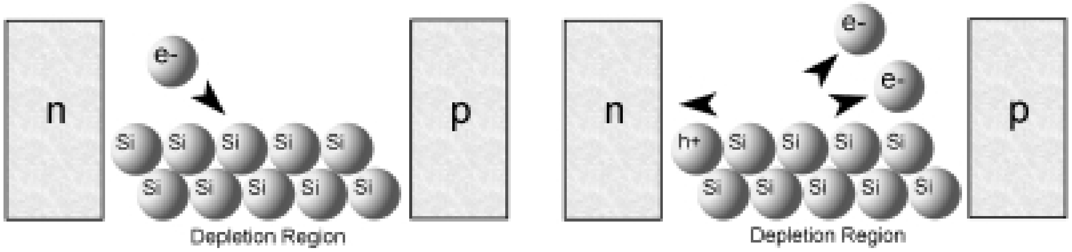

Figure 10–4. Avalanche Noise

When electrons and holes in the depletion region of a reversed-biased junction acquire enough energy to cause the avalanche effect, a random series of large noise spikes will be generated. The magnitude of the noise is difficult to predict due to its dependence on the materials.

Because the zener breakdown in a pn junction causes avalanche noise, it is an issue with op amp designs that include zener diodes. The best way of eliminating avalanche noise is to redesign a circuit to use no zener diodes.

10.4 Noise Colors

While the noise types are interesting, real op amp noise will appear as the summation of some or all of them. The various noise types themselves will be difficult to separate. Fortunately, there is an alternative way to describe noise, which is called *color*. The colors of noise come from rough analogies to light, and refer to the frequency content. Many colors are used to describe noise, some of them having a relationship to the real world, and some of them more attuned to the field of psycho-acoustics.

White noise is in the middle of a *spectrum* that runs from purple to blue to white to pink and red/brown. These colors correspond to powers of the frequency to which their spectrum is proportional, as shown in Table 10–1.

Table 10–1. Noise Colors

COLOR	FREQUENCY CONTENT
Purple	f^2
Blue	f
White	1
Pink	$\frac{1}{f}$
Red/Brown	$\frac{1}{f^2}$

There are an infinite number of variations between the colors. All inverse powers of frequency are possible, as are noises that are narrowband or appear only at one discrete frequency. Those, however, are primarily external sources of noise, so their presence is an important clue that the noise is external, not internal. There are no pure colors; at high frequencies, all of them begin to roll off and become pinkish. The op amp noise sources described above appear in the region between white noise and red/brown noise (Figure 10–5).

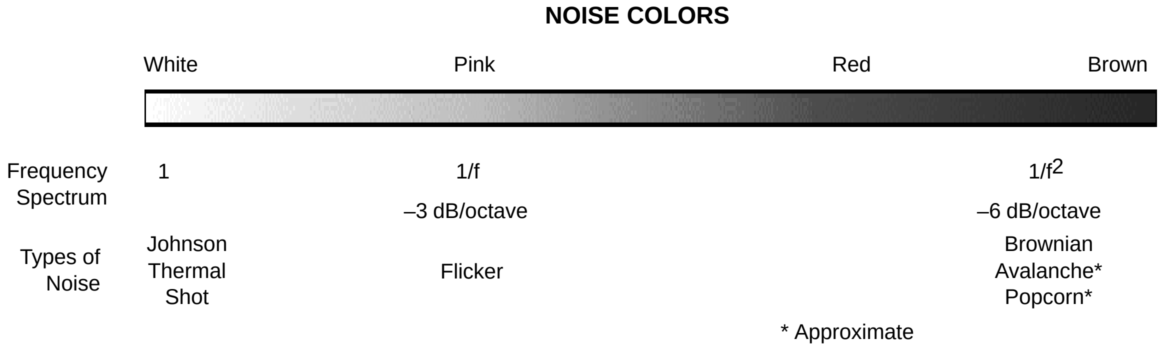

Figure 10–5. Noise Colors

10.4.1 White Noise

White noise is noise in which the frequency and power spectrum is constant and independent of frequency. The signal power for a constant bandwidth (centered at frequency f_o), does not change if f_o is varied. Its name comes from a similarity to white light, which has equal quantities of all colors.

When plotted versus frequency, white noise is a horizontal line of constant value.

Shot and thermal (Johnson) noise sources are approximately white, although there is no such thing as pure white noise. By definition, white noise would have infinite energy at infinite frequencies. White noise always becomes pinkish at high frequencies.

Steady rainfall or radio static on an unused channel approximate a white noise characteristic.

10.4.2 Pink Noise

Pink noise is noise with a 1/f frequency and power spectrum excluding dc. It has equal energy per octave (or decade for that matter). This means that the amplitude decreases logarithmically with frequency. Pink noise is pervasive in nature — many supposedly random events show a 1/f characteristic.

Flicker noise displays a 1/f characteristic, which also means that it rolls off at 3 dB/octave.

10.4.3 Red/Brown Noise

Red noise is not universally accepted as a noise type. Many sources omit it and go straight to brown, attributing red characteristics to brown. This has more to do with aesthetics than it does anything else (if brown noise is the low end of the spectrum, then pink noise should be named tan). So if pink noise is pink, then the low end of the spectrum should be red. Red noise is named for a connection with red light, which is on the low end of the visible light spectrum. But then this noise simulates Brownian motion, so perhaps it should be called Brown. Red/brown noise has a –6 dB/octave frequency response and a frequency spectrum of $1/f^2$ excluding dc.

Red/brown noise is found in nature. The acoustic characteristics of large bodies of water approximate red/brown noise frequency response.

Popcorn and avalanche noise approximate a red/brown characteristic, but they are more correctly defined as pink noise where the frequency characteristic has been shifted down as far as possible in frequency.

10.5 Op Amp Noise

This section describes the noise in op amps and associated circuits.

10.5.1 The Noise Corner Frequency and Total Noise

Op amp noise is never specified as shot, thermal, or flicker, or even white and pink. Noise for audio op amps is specified with a graph of equivalent input noise versus frequency. These graphs usually show two distinct regions:

- Lower frequencies where pink noise is the dominant effect
- Higher frequencies where white noise is the dominant effect

Actual measurements for the TLV2772 show that the noise has both white and pink characteristics (Figure 10–6). Therefore, the noise equations for each type of noise are not able to approximate the total noise out of the TLV2772 over the entire range shown on the graph. It is necessary to break the noise into two parts — the pink part and the white part — and then add those parts together to get the total op amp noise using the root-mean-square law of Paragraph 10.2.4.

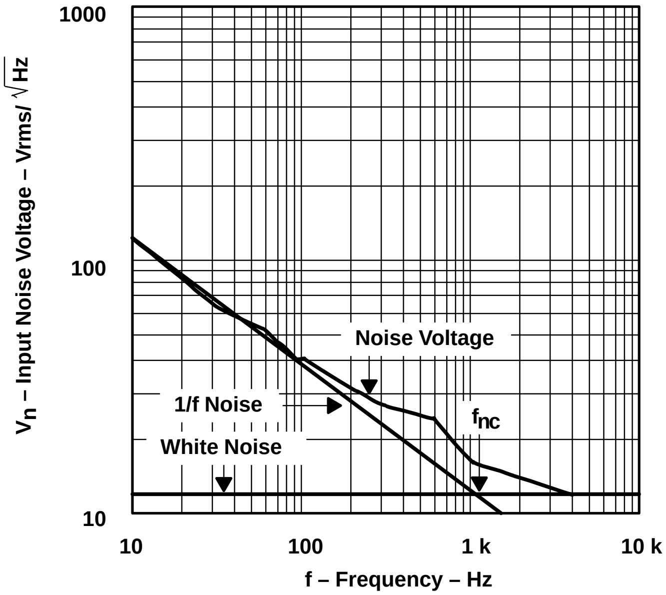

Figure 10–6. TLV2772 Op Amp Noise Characteristics

10.5.2 The Corner Frequency

The point in the frequency spectrum where 1/f noise and white noise are equal is referred to as the noise corner frequency, f_{nc}. Note on the graph in Figure 10–6 that the actual noise voltage is higher at f_{nc} due to the root-mean-square addition of noise sources as defined in Paragraph 10.2.4.

f_{nc} can be determined visually from the graph in Figure 10–6. It appears a little above 1 kHz. This was done by:

- Taking the white noise portion of the curve, and extrapolating it down to 10 Hz as a horizontal line.
- Taking the portion of the pink noise from 10 Hz to 100 Hz, and extrapolating it as a straight line.
- The point where the two intercept is $f_{nc,}$ the point where the white noise and pink noise are equal in amplitude. The total noise is then $\sqrt{2}$ x white noise specification (from Paragraph 10.2.4). This would be about $17\ \text{nV}/\sqrt{\text{Hz}}$ for the TLV2772.

This is good enough for most applications. As can be seen from the actual noise plot in Figure 10–6, small fluctuations make precise calculation impossible. There is a precise method, however:

- Determine the 1/f noise at the lowest possible frequency.

- Square it.
- Subtract the white noise voltage squared (subtracting noise with root-mean-squares is just as valid as adding).
- Multiply by the frequency. This will give the noise contribution from the 1/f noise.
- Then divide by the white noise specification squared. The answer is f_{nc}.

For example:

The TLV2772 has a typical noise voltage of $130\ \text{nV}/\sqrt{\text{Hz}}$ at 10 Hz (from 5-V plot on data sheet).

The typical white noise specification for the TLV2772 is $12\ \text{nV}/\sqrt{\text{Hz}}$ (from data sheet)

$$1/f\ \text{noise}^2@10\text{Hz} = \left[\left(\frac{130\ \text{nV}}{\sqrt{\text{Hz}}}\right)^2 - \left(\frac{12\ \text{nV}}{\sqrt{\text{Hz}}}\right)^2\right] \times 10\ \text{Hz} = 167560(\text{nV})^2 \tag{10–15}$$

$$f_{nc} = \frac{1/f\ \text{noise}^2@10\text{Hz}}{\text{white noise}^2} = \frac{167560(\text{nV})^2}{\left(\frac{12\ \text{nV}}{\sqrt{\text{Hz}}}\right)^2} = 1164\ \text{Hz} \tag{10–16}$$

Once the corner frequency is known, the individual noise components can be added together as shown in Paragraph 10.2.2. Continuing the example above for a frequency range of 10 Hz to 10 kHz:

$$E_n = E_{\text{whitenoise}} \sqrt{f_{nc} \times 1\text{n}\frac{f_{max}}{f_{min}} + (f_{max} - f_{min})} \tag{10–17}$$

$$E_n = \frac{12\ \text{nV}}{\sqrt{\text{Hz}}} \sqrt{1164\ \text{Hz} \times 1\text{n}\frac{10^4}{10} + \left(10^4\ \text{Hz} - 10\ \text{Hz}\right)} = 1.611\ \mu\text{V} = -116\ \text{dBV} \tag{10–18}$$

The example above presupposed that the bandwidth includes f_{nc}. If it does not, all of the contribution will be from either the 1/f noise or the white noise. Similarly, if the bandwidth is very large, and extends to three decades or so above f_{nc}, the contribution of the 1/f noise can be ignored.

10.5.3 Op Amp Circuit Noise Model

Texas Instruments measures the noise characteristics of a large sampling of devices. This information is compiled and used to determine the typical noise performance of the device. These noise specifications refer the input noise of the op amp. Some noise portions can be represented better by a voltage source, and some by a current source. Input voltage noise is always represented by a voltage source in series with the noninverting input. Input current noise is always represented by current sources from both inputs to ground (Figure 10–7).

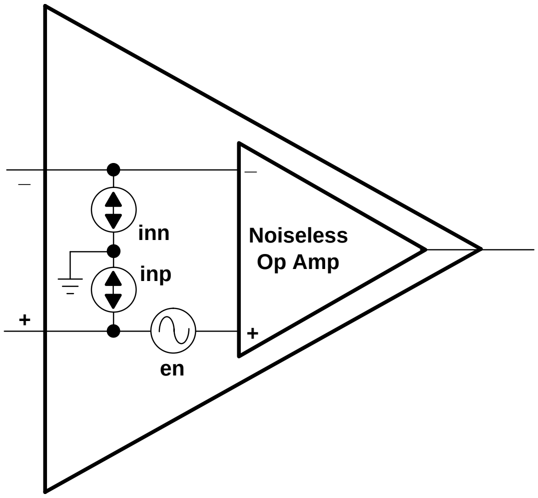

Figure 10–7. Op Amp Circuit Noise Model

In practice, op amp circuits are designed with low source impedance on the inverting and noninverting inputs. For low source impedances and CMOS JFET inputs, only the noise voltage is important; the current sources are insignificant in the calculations because they are swamped in the input impedances.

The equivalent circuit, therefore, reduces to that shown in Figure 10–8:

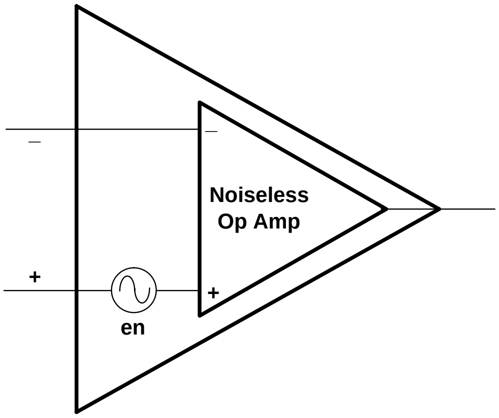

Figure 10–8. Equivalent Op Amp Circuit Noise Model

10.5.4 Inverting Op Amp Circuit Noise

If the previous circuit is operated in an inverting gain stage, the equivalent circuit becomes that shown in Figure 10–9:

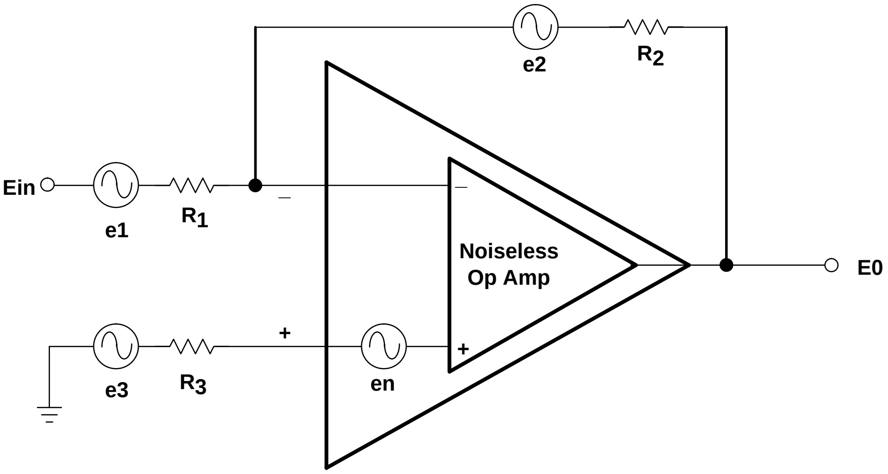

Figure 10–9. Inverting Op Amp Circuit Noise Model

The additional voltage sources e1 through e3 represent the thermal noise contribution from the resistors. As stated in Paragraph 10.3.2, the resistor noise can also be discounted if the values are low. Resistor noise will be omitted in the examples that follow. R3 is also not usually present, unless low common-mode performance is important. Deleting it and connecting the noninverting input directly to (virtual) ground makes the common mode response of the circuit worse, but may improve the noise performance of some circuits. There will be one less noise source to worry about. Therefore, the equivalent circuit becomes that shown in Figure 10–10:

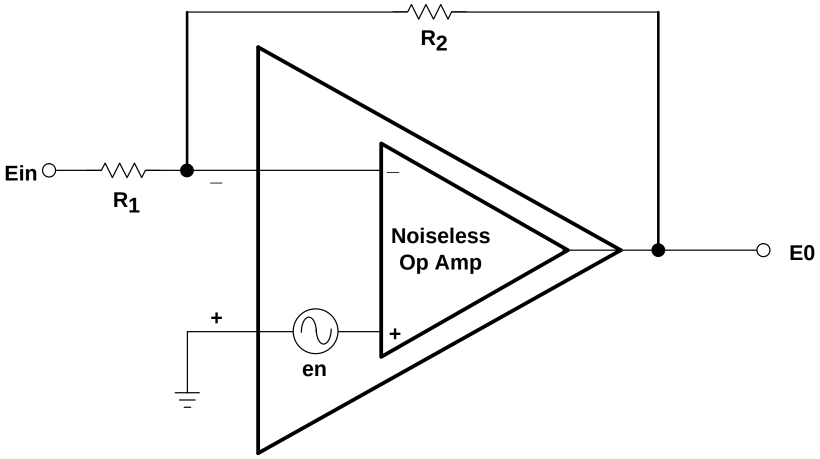

Figure 10–10. Inverting Equivalent Op Amp Circuit Noise Model

This simplifies the gain calculation:

$$E_0 = \sqrt{\left(E_{in}\frac{R_2}{R_1}\right)^2 + \left(e_n\left(1 + \frac{R_2}{R_1}\right)\right)^2} \tag{10–19}$$

where e_n = the total noise over the bandwidth of interest.

10.5.5 Noninverting Op Amp Circuit Noise

Taking the simplified equivalent op amp circuit from Paragraph 10.5.2 as the base, the noise equivalent of a noninverting op amp circuit is shown in Figure 10–11:

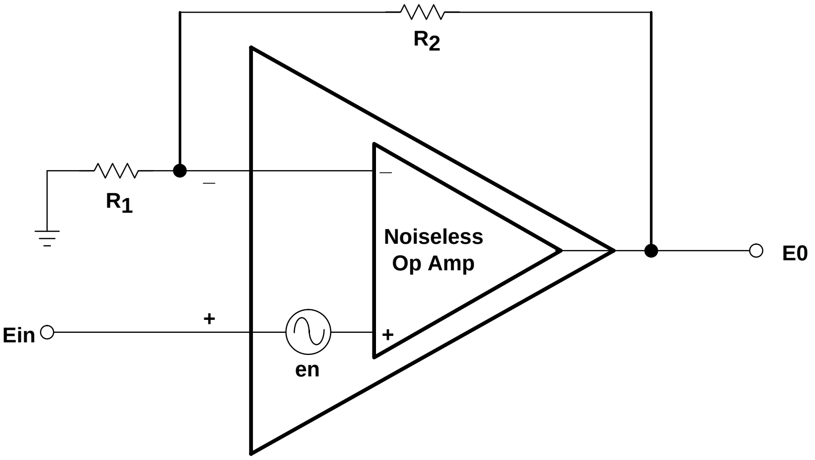

Figure 10–11. Noninverting Equivalent Op Amp Circuit Noise Model

The gain of this circuit is:

$$E_0 = \sqrt{\left(E_{in}\left(1 + \frac{R_2}{R_1}\right)\right)^2 + \left(e_n\left(1 + \frac{R_2}{R_1}\right)\right)^2} \tag{10–20}$$

10.5.6 Differential Op Amp Circuit Noise

Taking the simplified equivalent op amp circuit from Paragraph 10.5.2 as the base, the noise equivalent of a differential op amp circuit is shown in Figure 10–12:

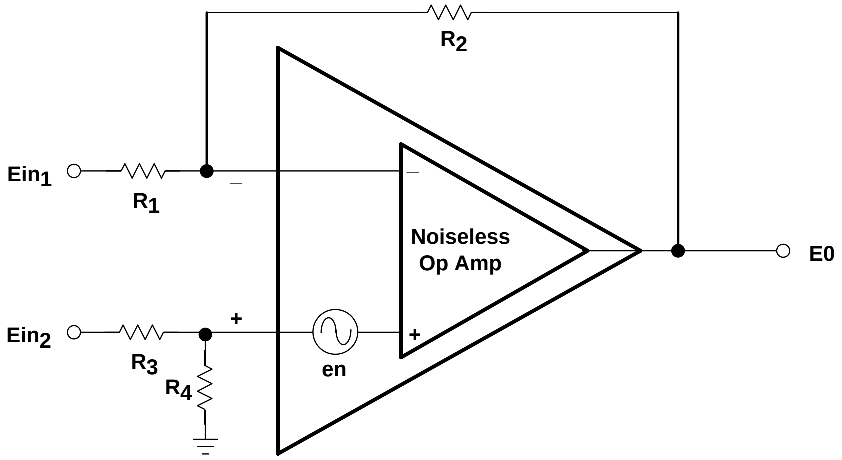

Figure 10–12. Differential Equivalent Op Amp Circuit Noise Model

Assuming that R1 = R3 and R2 = R4, the gain of this circuit is:

$$E_0 = \sqrt{\left((E_{in2} - E_{in1})\frac{R_2}{R_1}\right)^2 + \left(e_n\left(1 + \frac{R_2}{R_1}\right)\right)^2} \qquad (10–21)$$

10.5.7 Summary

The previous examples, though trivial, illustrate that noise always adds to the overall output of the op amp circuit. Reference 1 provides a much more in-depth derivation of op amp noise in circuits, including resistive effects.

10.6 Putting It All Together

This example is provided for analysis only — actual results depend on a number of other factors. Expanding on the techniques of Paragraph 10.2.5:

A low-noise op amp is needed over an audio frequency range of 20 Hz to 20 kHz, with a gain of 40 dB. The output voltage is 0 dBV (1V). The schematic is shown in Figure 10–13:

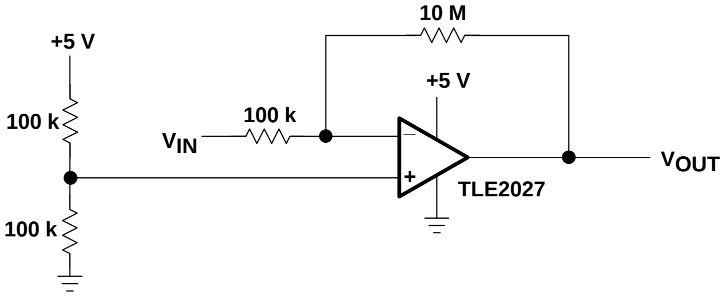

Figure 10–13. Split Supply Op Amp Circuit

It would be nice to use a TLE2027 with a noise figure of $2.5\ nV/\sqrt{Hz}$. The data sheet, however, reveals that this is a ±15-V part, and that noise figure is only specified at ±15 V. Furthermore, the specification for V_{OM+} and V_{OM-} (see Chapter 11) show that it can only swing to within approximately 2 V of its voltage rails. If they are +5 V and ground, the op amp is close to clipping with a 1-V output signal. This illustrates a common fallacy: the designer chooses an op amp based on one parameter only, without checking others that affect the circuit. An expert analog designer must develop an attention to details or be prepared to spend a lot of time in the lab with false starts and unexpected problems.

So, the only choice is to select a different op amp. The TLC2201 is an excellent choice. It is a low-noise op amp optimized for single supply operation. Figure 10–14 appears right on the first page of the data sheet, which should be extremely significant to the designer.

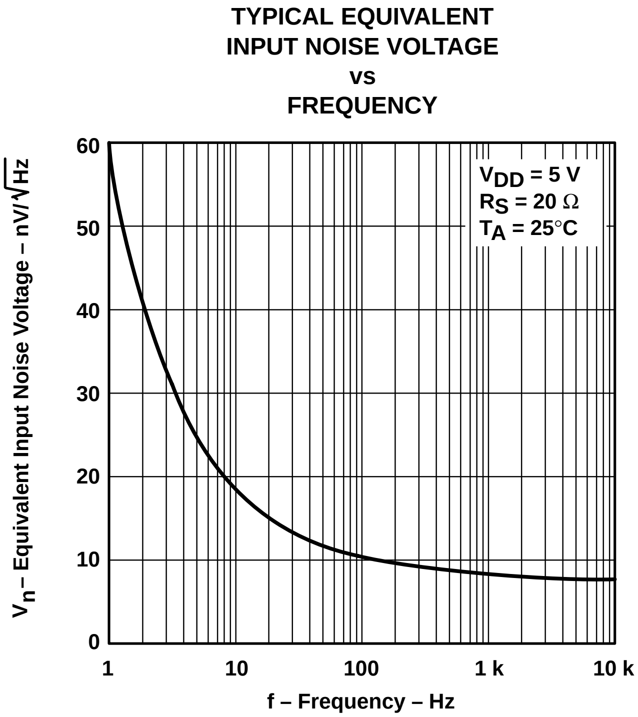

Figure 10–14. TLC2201 Op Amp Noise Performance

The first circuit change in this example is to change the TLE2027 to a TLC2201. Visually, the corner frequency f_{nc} appears to be somewhere around 20 Hz (from Paragraph 10.5.2), the lower frequency limit of the band we are interested in. This is good, it means for all practical purposes the 1/f noise can be discounted. It has 8 $nV/\sqrt{Hz}$ noise instead of 2.5 $nV/\sqrt{Hz}$, and from Paragraph 10.2.5:

- To begin with, calculate the *root Hz* part: $\sqrt{20000 - 20} = 141.35$.
- Multiplying this by the noise spec: $8 \times 141.35 = 1.131\ \mu V$, which is the equivalent input noise (E_{IN}). The output noise equals the input noise multiplied by the gain, which is 100 (40 dB).

The signal-to-noise ratio can be now be calculated:

- $1.131\ \mu V \times 100 = 113.1\ \mu V$
- Signal-to-noise (dB) =

$$20 \times \log(1V \div 113.1\ \mu V) = 20 \times \log(8842) = 78.9\ dB \qquad (10–22)$$

Pretty good, but 10 dB less than would have been possible with a TLE2027. If this is not acceptable (lets say for 16-bit accuracy), one is forced to generate a ±15-V supply. Let’s suppose for now that 78.9 dB signal-to-noise is acceptable, and build the circuit.

When it is assembled, it oscillates. What went wrong?

To begin with, it is important to look for potential sources of external noise. The way the schematic in Figure 10–14 is drawn provides a visual clue to the culprit: a long connection from the half-supply voltage reference to the high-impedance noninverting input. Added to that is a 50-kΩ source impedance, which does not effectively swamp external noise sources from entering the noninverting input. There is a big difference between simply providing a correct dc operating point, and providing one that has low impedances where they are needed. Most designers know the “fix”, which is to decouple the noninverting input as shown in Figure 10–15:

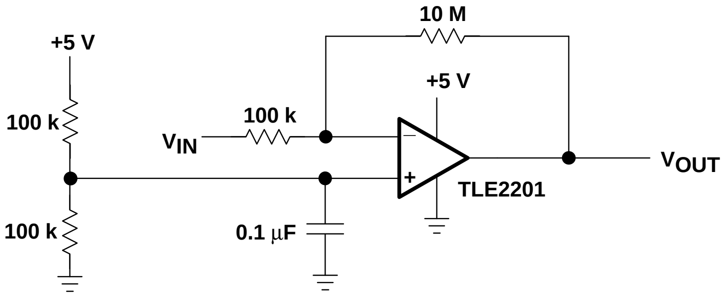

Figure 10–15. TLC2201 Op Amp Circuit

Better — it stopped oscillating. Probably a nearby noise source radiating into the noninverting input was providing enough noise to put the circuit into oscillation. The capacitor lowers the input impedance of the noninverting input and stops the oscillation. There is much more information on this topic in Chapter 17, including layout effects and component selection. For now, it is assumed that all of these have been taken into account.

The circuit is still slightly noisier than the 78.9 dB signal-to-noise ratio given above, especially at lower frequencies. This is where the real work of this example begins: that of eliminating component noise.

The circuit in Figure 10–15 has 4 resistors. Assuming that the capacitor is noiseless (not always a good assumption), that means four noise sources. For now, only the two resistors in the voltage divider that forms the voltage reference will be considered. The capacitor, however, has transformed the white noise from the resistors into pink (1/f) noise. From Paragraph 10.3.2 and 10.2.5, the noise from the resistors and the amplifier itself is:

$$E_{Totalrms} = \sqrt{5.73\ \mu V^2 + 5.73\ \mu V^2 + 113.1\ \mu V^2} = 113.1\ \mu V_{rms} \tag{10–23}$$

Signal-to-noise (dB) =

$$20 \times \log(1V \div 113.1\ \mu V) = 20 \times \log(8842) = 78.9\ \text{dB} \tag{10–24}$$

So far, so good. The amplifier noise is swamping the resistor noise, which will only add a very slight pinkish component at low frequencies. Remember, however, that this noise voltage is multiplied by 101 through the circuit, but that was previously taken into account for the 78.9 dB signal-to-noise calculation above.

Reducing the value of the resistors to decrease their noise is an option. Changing the voltage divider resistors from 100 kΩ to 1 kΩ while leaving the 0.1 μV capacitor the same, changes the corner frequency from 32 Hz to 796 Hz, right in the middle of the audio band.

Note:

Resist the temptation to make the capacitor larger to move the pinkish effect below the lower limits of human hearing. The resulting circuit must charge the large capacitor up during power up, and down during power down. This may cause unexpected results.

If the noise from the half-supply generator is critical, the best possible solution is to use a low-noise, low-impedance half-supply source. Remember, however, that its noise will be multiplied by 101 in this application.

The effect of the 100-kΩ resistor on the inverting input is whitish, and will appear across the entire bandwidth of the circuit. Compared to the amplifier noise, it is still small, just like the noise from the noninverting resistors on the input. The noise contribution of resistors will be discounted.

Of much more concern, however, is the 10-MΩ resistor used as the feedback resistor. The noise associated with it appears as a voltage source at the inverting input of the op amp, and, therefore, is multiplied by a factor of 100 through the circuit. From Paragraph 10.3.2, the noise of a 10-MΩ resistor is –84.8 dBV, or 57.3 μV. Adding this and the 100-kΩ resistor noise to the amplifier noise:

$$E_{Totalrms} = \sqrt{5.73\ \mu V^2 + 113.1\ \mu V^2} = 126.8\ \mu V_{rms} = -77.9\ \text{dBV} \tag{10–25}$$

Signal-to-noise (dB) =

$$20 \times \log(1V \div 126.8\ \mu V) = 20 \times \log(7887) = 77.9\ \text{dB} \tag{10–26}$$

The noise contribution from the 10-MΩ resistor subtracts 1 dB from the signal-to-noise ratio. Changing the 10-MΩ resistor to 100 kΩ, and the input resistor from 100 kΩ to 1 kΩ preserves the overall gain of the circuit. The redesigned circuit is shown in Figure 10–16:

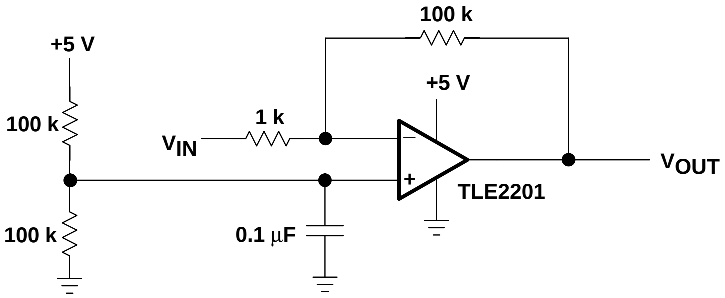

Figure 10–16. Improved TLC2201 Op Amp Circuit

For frequencies above 100 Hz, where the 1/f noise from the op amp and the reference resistors is negligible, the total noise of the circuit is:

$$E_{Totalrms} = \sqrt{0.57\ \mu V^2 + 5.73\ \mu V^2 + 113.1\ \mu V^2} = 113.2\ \mu V_{rms} = -78.9\ \text{dBV} \qquad (10–27)$$

Signal-to-noise (dB) =

$$20 \times \log(1V \div 113.2\ \mu V) = 20 \times \log(8830) = 78.9\ \text{dB} \qquad (10–28)$$

Proper selection of resistors, therefore, has yielded a signal-to-noise ratio close to the theoretical limit for the op amp itself. The power consumption of the circuit, however, has increased slightly, which may be unacceptable in a portable application. Remember, too, that this signal-to-noise ratio is only at an output level of 0 dBV, an input level of –40 dBV. If the input signal is reduced, the signal-to-noise ratio is reduced proportionally.

Music, in particular almost never sustains peak levels. The average amplitude may be down 20 dB to 40 dB from the peak values. This erodes a 79 dB signal-to-noise ratio to 39 dB in quiet passages. If someone “cranks up the volume” during the quiet passages, noise will become audible. This is done automatically with automatic volume controls. The only way a designer can combat this is to increase the voltage levels through the individual stages. If the preceding audio stages connecting to this example, for instance, could be scaled to provide 10 dB more gain, the TLC2201 would be handling an output level of 3.16 V instead of 1 V, which is well within its rail-to-rail limit of 0 V to 4.7 V. This would increase the signal-to-noise gain of this circuit to 88.9 dB — almost the same as would have been possible with a TLE2027 operated off of ±15V! But noise in the preceding stages would also increase. Combatting noise is a difficult problem, and there are always tradeoffs involved.

10.7 References

(1) Texas Instruments Application Report, *Noise Analysis in Operational Amplifier Circuits*, SLVA043A, 1999

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