Ask The Applications Engineer—7

by James Bryant and Lew Counts

OP-AMP ISSUES—NOISE

Q. *What should I know about op-amp noise?*

A. First, we must note the distinction between noise generated in the op amp and its circuit components and *interference,* or unwanted signals and noise arriving as voltage or current at any of the amplifier's terminals or induced in its associated circuitry.

Interference can appear as spikes, steps, sine waves, or random noise, and it can come from *anywhere:* machinery, nearby power lines, r-f transmitters and receivers, computers, or even circuitry within the same equipment (for example, digital circuits or switching-type power supplies). Understanding it, preventing its appearance in your circuit's neighborhood, finding how it got in, and rooting it out, or finding a way to live with it is a big subject. It's been treated in these pages in the past; those, and a few additional references, are mentioned in the Bibliography.

If all interference could be eliminated, there would still be random noise associated with the operational amplifier and its resistive circuits. It constitutes the ultimate limitation on the amplifier's resolution. That's the topic we'll begin to discuss here.

Q. *O.K. Tell me about random noise in op amps. Where does it come from?*

A. Noise appearing at the amplifier's output is usually measured as a voltage. But it is generated by both voltage- and current sources. All internal sources are generally *referred to the input,* i.e., treated as uncorrelated—or *independent*—random noise generators (see next question) in series or parallel with the inputs of an ideal noisefree amplifier: We consider 3 primary contributors to noise:

- a noise *voltage* generator (like offset voltage, usually shown in series with the noninverting input)
- two *noise-current* generators pumping currents out through the two differential-input terminals (like bias current).
- If there are any resistors in the op-amp circuit, they too generate noise; it can be considered as coming from either current sources or voltage sources (whichever is more convenient to deal with in a given circuit).

Op-amp voltage noise may be lower than 1 nV/$\sqrt{\text{Hz}}$ for the best types. Voltage noise is the noise specification that is more usually emphasized, but, if impedance levels are high, current noise is often the limiting factor in system noise performance. That is analogous to offsets, where offset voltage often bears the blame for output offset, but bias current is the actual guilty party. Bipolar op-amps have traditionally had less *voltage* noise than FET ones, but have paid for this advantage with substantially greater *current* noise—today, FET op-amps, while retaining their low current noise, can approach bipolar voltage-noise performance.

Q. *Hold it! 1 nV/$\sqrt{Hz}$? Where does $\sqrt{Hz}$ come from? What does it mean?*

A. Let's talk about random noise. Many noise sources are, for practical purposes (i.e., within the bandwidths with which the designer is concerned), both white and Gaussian. White noise is noise whose power within a given bandwidth is independent of frequency. Gaussian noise is noise where the probability of a particular amplitude, *X,* follows a Gaussian distribution. Gaussian noise has the property that when the rms values of noise from two or more such sources are added, provided that the noise sources are uncorrelated (i.e., one noise signal cannot be transformed into the other), the resulting noise is not their arithmetic sum but the root of the sum-of-their-squares (RSS).* The RSS sum of three noise sources, V_1, V_2, and V_3, is

$$V_O = \sqrt{V_1^{\,2} + V_2^{\,2} + V_3^{\,2}}$$

Since the different frequency components of a noise signal are uncorrelated, a consequence of RSS summation is that if the white noise in a brick-wall bandwidth of Δf *is* V, then the noise in a bandwidth of 2 Δf *is* $\sqrt{V^2 + V^2} = \sqrt{2}\,V$. More generally, if we multiply the bandwidth by a factor K, then we multiply the noise by a factor $\sqrt{K}$. The function defining the rms value of noise in a Δf = 1 Hz bandwidth anywhere in the frequency range is called the (voltage or current) *spectral density function,* specified in nV/$\sqrt{\text{Hz}}$ or pA/$\sqrt{\text{Hz}}$. For white noise, the spectral density is constant; it is multiplied by the square root of the bandwidth to obtain the total rms noise.

A useful consequence of RSS summation is that if two noise sources are contributing to the noise of a system, and one is more than 3 or 4 times the other, the smaller is often ignored, since

$$\sqrt{4^2} = \sqrt{16} = 4, \text{ while } \sqrt{4^2 + 1^2} = \sqrt{17} = 4.12$$

[difference less than 3%, or 0.26 dB]

$$\sqrt{3^2} = \sqrt{9} = 3, \text{ while } \sqrt{3^2 + 1^2} = \sqrt{10} = 3.16$$

[difference less than 6%, or 0.5 dB]

The source of the higher noise has become the *dominant* source.

Q. *O.K. How about current noise?*

A. The current noise of simple (i.e. not bias-current-compensated) bipolar and JFET op-amps is usually within 1 or 2 dB of the Schottky noise (sometimes called the "shot noise") of the bias current; it is not always specified on data sheets. Schottky noise is current noise due to random distribution of charge carriers in the current flow through a junction. The Schottky noise current, I_n, in a bandwidth, B, when a current, I, is flowing is obtained from the formula

$$I_n = \sqrt{2 I q B}$$

Where q is the electron charge (1.6×10^{-19} C). Note that $\sqrt{2 I q}$ is the spectral density, and that the noise is white.

This tells us that the current noise spectral density of simple bipolar transistor op-amps will be of the order of 250 fA/$\sqrt{\text{Hz}}$, for I_b = 200 nA, and does not vary much with temperature—and that the current noise of JFET input op-amps, while lower (4 fA/$\sqrt{\text{Hz}}$ at I_b, = 50 pA), will double for every 20°C chip temperature increase, since JFET op-amps' bias currents double for every 10°C increase.

Bias-compensated op-amps have much higher current noise than one can predict from their input currents. The reason is that their net bias current is the *difference* between the base current of the input transistor and the compensating current source, while the noise current is derived from the *RSS sum* of the noise currents.

Traditional voltage-feedback op-amps with balanced inputs almost always have equal (though uncorrelated) current noise on both

[*Note the implication that noise *power* adds linearly (sum of squares).]

their inverting and non-inverting inputs. Current-feedback, or transimpedance, op-amps, which have different input structures at these two inputs, do not. Their data sheets must be consulted for details of the noise on the two inputs.

The noise of op-amps is Gaussian with constant spectral density, or "white", over a wide range of frequencies, but as frequency decreases the spectral density starts to rise at about 3 dB/octave. This low-frequency noise characteristic is known as "$1/f$ noise" since the noise *power* spectral density goes inversely with frequency (actually $1/f^{\gamma}$). It has a –1 slope on a log plot (the noise *voltage* (or *current)* $1/\sqrt{f}$ spectral density slopes at –1/2). The frequency at which an extrapolated –3 dB/octave spectral density line intersects the midfrequency constant spectral density value is known as the "$1/f$ corner frequency" and is a figure of merit for the amplifier. Early monolithic IC op-amps had $1/f$ corners at over 500 Hz, but today values of 20–50 Hz are usual, and the best amplifiers (such as the AD-OP27 and the AD-OP37) have corner frequencies as low as 2.7 Hz. $1/f$ noise has equal increments for frequency intervals having equal ratios, i.e., per *octave* or per *decade*.

Q. *Why don't you publish a noise figure?*

A. The noise figure (NF) of an amplifier (expressed in dB) is a measure of the ratio of the amplifier noise to the thermal noise of the source resistance.

$V_n = 20 \log \{ [V_n(\text{amp})+V_n(\text{source})]/V_n(\text{source})\}$

It is a useful concept for r-f amplifiers, which are almost always used with the same source resistance driving them (usually 50 Ω or 75 Ω), but it would be misleading when applied to op amps, since they are used in many different applications with widely varying source impedances (which may or may not be resistive).

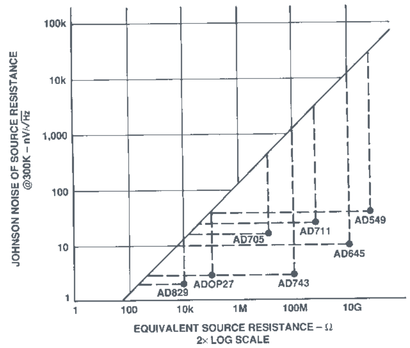
Q. *What difference does the source impedance make?*

A. At temperatures above absolute zero all resistances are noise sources; their noise increases with resistance, temperature, and bandwidth (we'll discuss basic resistance noise, or *Johnson noise*, in a moment). *Reactances* don't generate noise, but noise currents through them will develop noise voltages.

If we drive an op-amp from a source resistance, the equivalent noise input will be the RSS sum of the amplifier's noise voltage, the voltage generated by the source resistance, and the voltage caused by the amplifier's I_n flowing through the source impedance. For very low source resistance, the noise generated by the source resistance and amplifier current noise would contribute insignificantly to the total. In this case, the noise at the input will effectively be just the voltage noise of the op-amp.

If the source resistance is higher, the Johnson noise of the source resistance may dominate both the op-amp voltage noise and the voltage due to the current noise; but it's worth noting that, since the Johnson noise only increases with the square root of the resistance, while the noise voltage due to the current noise is directly proportional to the input impedance, the amplifier's current noise will always dominate for a high enough value of input impedance. When an amplifier's voltage and current noise are high enough, there may be no value of input resistance for which Johnson noise dominates.

This is demonstrated by the figure nearby, which compares voltage and current noise noise for several Analog Devices op amp types, for a range of source-resistance values. The diagonal line plots vertically the Johnson noise associated with resistances on the horizontal scale. Let's read the chart for the ADOP27: The horizontal line indicates the ADOP27's voltage noise level of about 3 nV/$\sqrt{\text{Hz}}$ is equivalent to a source resistance of less than about 500 Ω. Noise will not be reduced by (say) a 100-Ω source impedance, but it will be increased by a 2-kΩ source impedance. The vertical line for the ADOP27 indicates that, for source resistances above about 100 kΩ, the noise voltage produced by amplifier's current noise will exceed that contributed by the source resistance; it has become the dominant source.



Remember that any resistance in the non-inverting input will have Johnson noise and will also convert current noise to a noise voltage; and Johnson noise in feedback resistors can be significant in high-resistance circuits. All potential noise sources must be considered when evaluating op amp performance.

Q. *You were going to tell me about Johnson noise.*

A. At temperatures above absolute zero, all resistances have noise due to thermal movement of charge carriers. This is called Johnson noise. The phenomenon is sometimes used to measure cryogenic temperatures. The voltage and current noise in a resistance of R ohms, for a bandwidth of B Hz, at a temperature of T kelvins, are given by:

$$V_n = \sqrt{4kTRB} \text{ and } I_n = \sqrt{4kTB/R}$$

Where k is Boltzmann's Constant (1.38×10^{-23} J/K). A handy rule of thumb is that a 1-kΩ resistor has noise of 4 nV/$\sqrt{\text{Hz}}$ at room temperature.

All resistors in a circuit generate noise, and its effect must always be considered. In practice, only resistors in the input(s) and, perhaps, feedback, of high-gain, front-end circuitry are likely to have an appreciable effect on total circuit noise.

Noise can be reduced by reducing resistance or bandwidth, but temperature reduction is generally not very helpful unless a resistor can be made very cold—since noise power is proportional to the *absolute* temperature, T = °C + 273°. ▶

(to be continued)

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Ask The Applications Engineer—8

by James Bryant

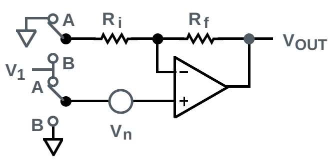
OP-AMP ISSUES

(Noise, continued from the last issue, 24-2)

Q. What is "noise gain"?

A. So far we have considered noise sources but not the gain of the circuits where they occur. It is tempting to imagine that if the noise voltage at the input of an amplifier is V_n and the circuit's signal gain is G, the noise voltage at the output will be GV_n; but this is not always the case.

Consider the basic op-amp gain circuit in the diagram. If it is being used as an inverting amplifier (B), the non-inverting input will be grounded, the signal will be applied to the free end of R_i and the gain will be $-R_f/R_i$. On the other hand, in a non-inverting amplifier (A) the signal is applied to the non-inverting input and the free end of R_i is grounded; the gain is $(1 + R_f/R_i)$



	SIGNAL GAIN	AMPLIFIER NOISE GAIN
A:	$\left(1+\frac{R_f}{R_i}\right)$	$\left(1+\frac{R_f}{R_i}\right)$
B:	$-\frac{R_f}{R_i}$	$\left(1+\frac{R_f}{R_i}\right)$

The amplifier's own voltage noise is always amplified in the non-inverting mode; thus when an op-amp is used as an inverting amplifier at a gain of G, its voltage noise will be amplified by the noise gain of (G+ 1). For the precision attenuation cases, where G<1, this may present problems. (A common example of this is an active filter circuit where stopband gain may be very small but stop-band noise gain is at least unity.)

Only the amplifier voltage noise—and any noise developed by the noninverting-input current noise flowing in any impedance present in that input (for example, a bias-current compensation resistor)—is amplified by the noise gain. Noise in R_i, either Johnson noise or arising from inverting input noise current, is amplified by G in the same way as the input signal, and Johnson noise voltage in the feedback resistor is not amplified but is buffered to the output at unity gain.

Q. What's "popcorn" noise?

A. Twenty years ago this column would have spent a great deal of space discussing popcorn noise, which is a type of low frequency noise manifesting itself as low level (but random amplitude) step changes in offset voltage occurring at random intervals. When played through a loudspeaker it sounds like cooking popcorn—hence the name.

While no integrated circuit process is entirely free from the problem, high levels of popcorn noise result from inadequate processing techniques. Today its causes are sufficiently well understood that no reputable op-amp manufacturer is likely to produce op-amps where popcorn noise is a major concern to the user. {Oat-bran noise is more likely to be an issue in situations where cereal data is concerned[:-)]}

Q. Pk-pk noise voltage is the most convenient way to know whether noise will ever be a problem for me. Why are amplifier manufacturers reluctant to specify noise in this way?

A. Because noise is generally Gaussian, as we pointed out in the last issue. For a Gaussian distribution it is meaningless to speak of a maximum value of noise: if you wait long enough any value will, in theory, be exceeded. Instead it is more practical to speak of the rms noise, which is more or less invariant—and by applying the Gaussian curve to this we may predict the probability of the noise exceeding any particular value. Given a noise source of V rms, since the probability of any particular value of noise voltage follows a Gaussian distribution, the noise voltage will exceed a pk-pk value of 2 V for 32% of the time, 3 V for 13% of the time, and so on:

Pk-pk value	% of time pk-pk value is exceeded
2 × rms	32%
4 × rms	4.6%
6 × rms	0.27%
6.6 × rms	0.10%
8 × rms	60 ppm
10 × rms	0.6 ppm
12 × rms	2×10^{-9} ppm
14 × rms	2.6×10^{-12} ppm

So if we define a peak value in terms of the probability of its occurrence, we may use a peak specification—but it is more desirable to use the rms value, which is generally easier to measure. When a peak noise voltage is specified, it is frequently 6.6 × rms, which occurs no more than 0.1% of the time.

Q. How do you measure the rms value of low-frequency noise in the usually specified band, 0.1 to 10 Hz? It must take a long time to integrate. Isn't this expensive in production?

A. Yes, it is expensive, but—Although it's necessary to make many careful measurements during characterization, and at intervals thereafter, we cannot afford the time it would take in production to make an rms measurement. Instead, at very low frequencies in the 1/*f* region (as low as 0.1 to 10 Hz), the peak value is measured during from one to three 30-second intervals and must be less than some specified value. Theoretically this is unsatisfactory, since some good devices will be rejected and some noisy ones escape detection, but in practice it is the best test possible within a practicable test time and is acceptable if a suitable threshold limit is chosen. With conservative weightings applied, this is a reliable test of noise. Devices that do not meet the arbitrary criteria for the highest grades can still be sold in grades for which they meet the spec.

Q. What other op-amp noise effects do you encounter?

A. There is a common effect, which often appears to be caused by a noisy op amp, resulting in missing codes. This potentially serious problem is caused by ADC input-impedance modulation. Here's how it happens:

Many successive-approximation ADCs have an input impedance which is modulated by the device's conversion clock. If such an ADC is driven by a precision op amp whose bandwidth is much lower than the clock frequency, the op amp cannot develop sufficient feedback to provide a stiff voltage source to the ADC input port, and missing codes are likely to occur. Typically, this effect appears when amplifiers like the OP-07 are used to drive AD574s.

It may be cured by using an op amp with sufficient bandwidth to have a low output impedance at the ADC's clock frequency, or by choosing an ADC containing an input buffer or one whose input impedance is not modulated by its internal clock (many sampling ADCs are free of this problem). In cases where the op amp can drive a capacitive load without instability, and the reduction of system bandwidth is unimportant, a shunt capacitor decoupling the ADC input may be sufficient to effect a cure.

Q. Are there any other interesting noise phenomena in high-precision analog circuits?

A. The tendency of high-precision circuitry to drift with time is a noise-like phenomenon (in fact, it might be argued that, at a minimum, it is identical to the lower end of *1/f* noise). When we specify long-term stability, we normally do so in terms of µV/1,000 hr or ppm/1,000 hr. Many users assume that, since there are, on the average, 8,766 hours in a year, an instability of *x*/1,000 hr is equal to 8.8 *x*/yr.

This is not the case. Long-term instability (assuming no long-term steady deterioration of some damaged component within the device), is a "drunkard's walk" function; what a device did during its last 1,000 hours is no guide to its behavior during the next thousand. The long-term error mounts as the square-root of the elapsed time, which implies that, for a figure of *x*/l,000 hr, the drift will actually be multiplied by $\sqrt{8.766}$, or about 3× per year, or 9× per 10 years. Perhaps the spec should be in µV/1,000 $\sqrt{hr}$.

In fact, for many devices, things are a bit better even than this. The "drunkard's walk" model, as noted above, assumes that the properties of the device don't change. In fact, as the device gets older, the stresses of manufacture tend to diminish and the device becomes more stable (except for incipient failure sources). While this is hard to quantify, it is safe to say that—provided that a device is operated in a low-stress environment—its rate of long-term drift will tend to reduce during its lifetime. The limiting value is probably the 1/*f* noise, which builds up as the *square-root of the natural logarithm* of the ratio, i.e., $\sqrt{\ln 8.8}$ for time ratios of 8.8, or 1.47 × for 1 year, 2.94 × for 8.8 years, 4.4 × for 77 years, etc.

A READER'S CHALLENGE:

Q. A reader sent us a letter that is just a wee bit too long to quote directly, so we'll summarize it here. He was responding to the mention in these columns *(Analog Dialogue* 24-2, pp. 20-21) of the shot effect, or Schottky noise (Schottky was the first to note and correctly interpret shot effect—originally in vacuum tubes[1]). Our reader particularly objected to the designation of shot noise as solely a junction phenomenon, and commented that we have joined the rest of the semiconductor and op-amp engineering fraternity in disseminating misinformation.

In particular, he pointed out that the shot noise formula—

$$I_n = \sqrt{2q\,IB} \text{ amperes,}$$

where I_n is the rms shot-noise current, *I* is the current flowing through a region, *q is* the charge of an electron, and *B* is the bandwidth—does not seem to contain any terms that depend on the physical properties of the region. Hence (he goes on) shot noise is a *universal* phenomenon associated with the fact that any current, *I*, is a flow of electrons or holes, which carry discrete charges, and the noise given in the formula is just an expression of the graininess of the flow.

He concludes that the omission of this noise component in any circuit carrying current, including purely resistive circuits, can lead to serious design problems. And he illustrates its significance by pointing out that this noise current, calculated from the flow of dc through any ideal resistor, becomes equal to the thermal Johnson noise current at room temperature when only 52 mV is applied to the resistor—and it would become the dominant current noise source for applied voltages higher than about 200 mV.

A. Since designers of low-noise op amps have blithely ignored this putative phenomenon, what's wrong? *The assumption that the above shot noise equation is valid for conductors.*

Actually, the shot noise equation is developed under the assumption that the carriers are independent of one another. While this is indeed the case for currents made up of discrete charges crossing a barrier, as in a junction diode (or a vacuum tube), it is not true for metallic conductors. Currents in conductors are made up of very much larger numbers of carriers (individually flowing much more slowly), and the noise associated with the flow of current is accordingly very much smaller—and generally lost in the circuit's Johnson noise.

Here's what Horowitz and Hill[2] have to say on the subject:

"An electric current is the flow of discrete electric charges, not a smooth fluidlike flow. The finiteness of the charge quantum results in statistical fluctuations of the current. *If the charges act independently of each other,** the fluctuating current is . . .

$$I\,noise\ (\text{rms}) = I_{nR} = (2\,qI_{dc}\,B)^{1/2}$$

where *q* is the electron charge (1.60×10^{-19} C) and *B* is the measurement bandwidth. For example, a "steady" current of 1 A actually has an rms fluctuation of 57 nA, measured in a 10-kHz bandwidth; i.e., it fluctuates by about 0.000006%. The relative fluctuations are larger for smaller currents: A "steady" current of 1 µA actually has an rms current-noise fluctuation, over 10 kHz, of 0.006%, i.e., –85 dB. At 1 pA dc, the rms current fluctuation (same bandwidth) is 56 fA, i.e., a 5.6% variation! Shot noise is 'rain on a tin roof.' This noise, like resistor Johnson noise, is Gaussian and white.

"The shot noise formula given earlier assumes that the charge carriers making up the current act independently. That is indeed the case for charges crossing a barrier, as for example the current in a junction diode, where the charges move by diffusion; *but it is not true for the important case of metallic conductors, where there are long-range correlations between charge carriers. Thus the current in a simple resistive circuit has far less noise than is predicted by the shot noise formula.** Another important exception to the shot-noise formula is provided by our standard transistor current-source circuit, in which negative feedback acts to quiet the shot noise."

*Italics ours

[1]Goldman, Stanford, *Frequency Analysis, Modulation, and Noise.* New York: McGraw-Hill Book Company, 1948, p. 352.

[2]Horowitz, Paul and Winfield Hill, *The Art of Electronics,* 2nd edition. Cambridge (UK): Cambridge University Press, 1989, pp. 431-2.