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Noise 101

A FEW BASIC CONCEPTS, PRESENTED IN THIS FIRST INSTALLMENT OF A TWO-PART SERIES, CAN GET YOU STARTED THINKING ABOUT DESIGNING FOR LOW NOISE.

NOISE REPRESENTS the fundamental limit for many signal-processing applications. As such, it is a key constraint for many electronic designs, particularly interface circuits. Industry trends in applications as disparate as test and measurement, medical imaging, and high-speed data communications demand ever-increasing information densities. Meanwhile, advancements in semiconductor processes enable greater data-processing speeds and functional densities but at the cost of operating supply voltage and, consequently, signal amplitude. The result is increasing pressure on system designs to manage the noise performance of analog front ends.

Approaching a subject as seemingly chaotic as noise in a systematic fashion is a tall order and not one that a few pages can fill. In an effort to extend the scope of this treatment, *EDN* has asked several semiconductor manufacturers with expertise in low-noise devices and circuit design to contribute linkable application notes and technical papers. These resources are available in the Analog Technical Resources section of the *EDN* Web site (www.edn.com). This collection of links can serve as a dynamic information source for further reading on the subject.

If you do your own research into the topic, you'll find that much of the literature groups all unwanted signals—those that couple in from external sources and those whose sources reside within the circuit—under one umbrella heading: *noise*. But the precautions and remedies available to the designer differ substantially between the two types. You cannot afford to ignore either, but the concentration here is on noise sources internal to the signal path. That said, good low-noise-system design requires a clear accounting of interferers within your circuit's operating environment as well (see **sidebar** "External contributors").

RANDOM EVENTS, PREDICTABLE SHAPES

Through various source mechanisms, electronic components produce a combination of three noise spectra. Individual source terms exhibit flatband noise, 1/f noise, or $1/f^2$ noise:

$$p_n(f) = c,$$

$$P_n(f) \propto \frac{1}{f}, \text{ and}$$

$$p_n(f) \propto \frac{1}{f^2},$$

respectively, where $p_n(f)$ is the noise source's power spectral density—the average power in a 1-Hz bandwidth centered at frequency *f*—and c is a constant amplitude (**Reference 1**).

Not to be confused with the shape of the noise spectra, the power spectral density is given as a function in units of watts per hertz, allowing you to calculate the rms noise power in a frequency band by integrating the density over a bandwidth.

$$P(f) = \int_{f_1}^{f_2} p_n(f)df.$$

However, most active circuits process signals as either currents or voltages. For example, bipolar transistors are transconductance devices: They produce an output signal current in response to an input signal voltage. To allow ready comparisons between signals and noise, common practice expresses noise spectral densities in terms of a voltage per root hertz or a current per root hertz

Of the mechanisms that generate the three common noise spectra, the most prevalent produce flatband noise, also called *white* noise because the power is evenly distributed over the entire spectrum in much the same way that white light is evenly distributed over the visible spectrum. Flatband sources generate shot noise and thermal noise, also known as Johnson noise in honor of physicist John Bertrand Johnson, who discovered the phenomenon in 1928. Though their spectra are indistinguishable, the behaviors of shot and Johnson sources differ as functions of circuit operating conditions.

AT A GLANCE

- ▷ Distinguish among device noise, distortion products, and interfering signals. Each has its own set of sources, preventions, and remedies.
- ▷ Active devices and resistors produce a number of noise profiles. Familiarize yourself with the various noise sources and the circuit operating parameters that excite them.
- ▷ Your signal's source impedance can set the context for evaluating noise magnitudes in many applications.
- ▷ You can perform quick and accurate back-of-the-envelope thermal-noise estimates if you keep a couple of scalable values in mind.

A SHOT HEARD 'ROUND THE WORLD

Shot noise derives from the discrete quantum nature of electron flow through a potential barrier. It is most often associated with diodes and bipolar transistors. A current may cross a junction at a steady average rate given by the dc current magnitude, but individual carriers cross as random events only when they have sufficient energy to overcome the junction's potential barrier (**Reference 2**). In the limit, current quantizes to the electron level, so the average current comprises a large number of discrete events.

The shot noise is given by

$$I_n = \sqrt{2qI_D\Delta_f}$$

in amperes rms, where q is the electron charge (1.6×10^{-19} C), I_D is the forward junction current, and Δ_f is the measurement bandwidth (**Figure 1**). As you can see from the expression, the shot noise is proportional to the square root of the junction current and independent of temperature. Both facts are noteworthy. Increasing the bias current may imply a larger shot noise in absolute terms, but circuits can take advantage of relationships that grow linearly with the bias—much faster than does the noise—a recurring theme in low-noise design. For example, g_m, the small signal transconductance of a bipolar transistor is linear in the collector current:

$$g_m = \frac{qI_C}{kT},$$

where I_C is the collector current, k is Boltzmann's constant (1.38×10^{-23}J/K), and T is the temperature in Kelvin.

You can also express the shot noise as a noise *voltage* by multiplying the shot current by the dynamic junction impedance. In this form, the shot noise appears to have a temperature dependence, but that situation is due to the fact that the dynamic junction impedance—the reciprocal of the transconductance in the case of a bipolar transistor—is linear in temperature.

There is also a shot-noise term associated with the reverse junction leakage, but the scaling current in this case is orders of magnitude smaller than forward currents. So, though you can devise circuits that exhibit the reverse-current shot

EXTERNAL CONTRIBUTORS

The path to a low-noise implementation of a given function does not begin with a search for the lowest noise amplifiers or data converters. Instead, it starts at a higher level of abstraction with a thorough understanding of your circuit's goals, your application's requirements, and the electrical and thermal environments in which your product will operate.

Just as you need to be clear about your design goals and the conditions imposed by the operating environment, good low-noise design requires that you distinguish between noise and other unwanted signal components, most notably distortion products and interference signals.

Though caused by widely differing mechanisms, noise processes generate incoherent signals and, collectively, do so over a frequency range from beyond your application's upper frequency limit to values approaching dc, including, for example, drift.

Conversely, unwanted signal components that exhibit coherence do not derive from device noise sources but from interferers or distortion mechanisms. The spectral fingerprint of each extraneous signal helps point to its origin. The challenge is often separating the voices in the chorus, so to speak.

Early on in your design cycle, assess your product's intended operating environment and determine the types, spectra, signal magnitudes, and coupling methods of nearby interferers. They may include RF sources, inductively coupled power transients, or even heat sources that can induce drift. Consider how your design can reduce the coupling coefficient by taking advantage of simple and common techniques, such as shielding, using balanced external signal lines, and good ground-system design.

noise, most practical circuits have several other noise sources that swamp the term.

WHERE THERE'S HEAT

Unlike shot noise, which originates from the carrier conduction behavior through a potential barrier, Johnson noise derives from random carrier motion within the device, producing an rms noise power given by

$$P = \int_{f_1}^{f_2} kT\,df = kT\Delta_f,$$

where Δ_f is the measurement bandwidth in hertz. This term is often called thermal noise because the carrier motion is thermally excited. Johnson noise has a gaussian amplitude distribution in the time domain and is evenly distributed across the spectrum. Thermal noise's spectral breadth and its sources' ubiquity lead it to dominate other noise types in many applications.

Thermal carrier agitation requires only a population of carriers within a conductive region. As such, you can observe Johnson noise in passive as well as active devices. The thermal voltage of a resistance, e_n, is a function of the resistance, temperature, and measurement bandwidth.

$$E_n = \sqrt{4kTR\Delta_f},$$

in volts rms, where R is the resistance in ohms (**Figure 2**). Dividing through by the resistance yields the Norton equivalent noise source,

$$I_n = \sqrt{\frac{4kT\Delta_f}{R}},$$

in amperes rms.

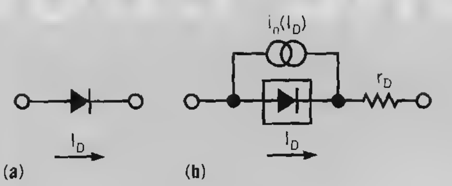


Figure 1 **Junction shot noise models as a current source with a noise current that scales as a function of junction current.**

Normalizing the rms noise voltage or current to a 1-Hz bandwidth gives the spectral density—e_n and i_n, respectively—with corresponding units of volts per root hertz and amperes per root hertz. Depending on the applications you are most often designing for, handy numbers to keep in mind are the voltage-noise spectral density of a 50Ω resistor—about 0.9 nV/$\sqrt{\text{Hz}}$—or of a 1-kΩ resistor—4 nV/$\sqrt{\text{Hz}}$. Because the noise spectral density is proportional to the square root of the resistance, you can easily scale these values for impedances appropriate to your circuit. Also, noting that these figures represent the rms noise over a 1-Hz bandwidth, you can scale to bandwidths appropriate to your application in similar fashion, by multiplying through by the square root of the bandwidth. **Table 1** shows the voltage noise spectral density for characteristic impedances for several applications.

The ability to quickly calculate the rms amplitude of noise sources helps you to identify the dominant source that sets your circuit's performance limit. In cases in which several sources exhibit similar amplitudes, you need to calculate their total (see **sidebar** "Random sums").

As was the case with the shot current, increasing the absolute noise amplitude can improve circuit performance if the signal amplitude increases faster as a result. So, for example, if you increase the load resistor in a gm-R stage, its absolute thermal noise increases, but the stage gain increases linearly with R, the noise rises only as root-R.

If try to entirely rid your circuit of resistors and their thermal noise by implementing switched-capacitor circuits, you'll find a thermal-noise term associated with them as well. Capacitors do not generate noise in and of themselves but scale noise terms generated elsewhere in the circuit:

$$E_n = \sqrt{\frac{kT}{C}},$$

in volts rms, where C is the capacitance in farads. For example, the charge uncertainty on a capacitor due to thermal-carrier motion is the analog to thermal noise in a resistor. In switched-capacitor circuits, the kT/C term can force a trade-off between noise performance on the one hand and implementation density, signal bandwidth, and power dissipation on the other (**Reference 3**).

NOT ALL THAT FLICKERS

Flicker noise occurs in all active devices and depends on the dc bias current:

$$I_n = \sqrt{m\frac{I^a}{f^b}\Delta_f},$$

where m is a device-dependent factor, a

RANDOM SUMS

The rms amplitudes of multiple sources combine to determine a circuit's noise performance. Assuming that the noise sources are uncorrelated, they sum as

$$\sum_{1}^{m=i} e_{n_m} = \sqrt{e_{n_1}^2 + e_{n_2}^2 + \cdots + e_{n_i}^2}.$$

Because one term often dominates rms sums, the signal source impedance can often set the context for noise evaluations. If your circuit's noise sources are predominantly resistances, you can make quick comparisons by expressing nonresistive noise sources in terms of an equivalent noise resistance. So, for example, an amplifier with an input noise voltage spectral density of 4 nV/$\sqrt{\text{Hz}}$ represents an equivalent noise resistance of about 1 kΩ. In a circuit with a 50Ω signal source, you can by casual inspection determine that the amplifier noise will dominate that of the source impedance. Indeed, ignoring other noise sources, the total noise spectral density in this example is only 0.1 nV/$\sqrt{\text{Hz}}$ greater than that of the amplifier. You can combine the noise resistances by adding them directly:

$$\text{let } e_{n_1} = \sqrt{4kTR_1}, \text{ and}$$
$$e_{n_2} = \sqrt{4kTR_2}$$

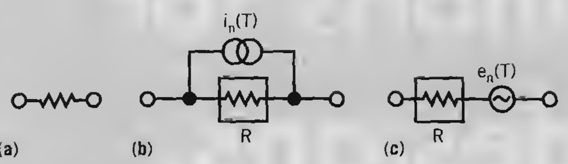
The total noise is

$$\begin{aligned} e_n &= \sqrt{e_{n_1}^2 + e_{n_2}^2} \\ &= \sqrt{4kTR_1 + 4kTR_2} \\ &= \sqrt{4kT(R_1 + R_2)}. \end{aligned}$$

Using the values from the above example, the amplifier's equivalent input noise resistance is 1 kΩ, and the source impedance is 50Ω.

$$\begin{aligned} e_n &= \sqrt{4 \cdot 1.38 \times 10^{-23}\,\text{J/K} \cdot 293\text{K} \cdot 1050(\Omega)} \\ &= 4.1\frac{\text{nV}}{\sqrt{\text{Hz}}}. \end{aligned}$$

Figure 2



Any physical implementation of a resistor as it appears on your schematic (a) includes a Johnson-noise voltage (b), which alternatively may be represented by its Norton-equivalent noise current (c).

is a constant in the range of 0.5 to 2, and b is a constant in the range of 0.8 and 1.2 (**Reference 4**). The inverse dependence on frequency gives this term its most common name: 1/f noise. Johnson observed 1/f noise in vacuum tubes in 1925 (**Reference 5**). Although they clearly derive from differing mechanisms, 1/f noise terms appear in semiconductors, metal films, electrolytic solutions, and in non-electronic forms in mechanical and biological systems, to name just a few. The detailed source mechanisms are not completely known; several models exist to explain the phenomenon. But, in general terms, 1/f noise's origins are attributed in semiconductor devices to the effects of contaminants and defects in the crystal structure. In MOS structures, 1/f noise is associated with oxide surface states that periodically trap and release carriers. Over the decades, advances in semiconductor processes and fabrication practices have reduced flicker noise in otherwise-similar devices.

The frequency at which a device's 1/f noise exceeds its thermal noise is the 1/f corner. The corner frequency is a function of the operating conditions—most notably temperature and bias current—and of the fabrication process. Under "typical" operating conditions, precision bipolar processes offer the lowest 1/f corners: around 1 to 10 Hz. The corner for devices fabricated in high-frequency bipolar processes is often 1 to 10 kHz. The 1/f corner frequency in MOSFETs goes as the reciprocal of the channel length, with typical values of 100 kHz to 1 MHz. Devices built on III-V processes, such as gallium-arsenide FETs and indium-gallium-phosphorous heterojunction-bipolar transistors, offer extremely wide bandwidths but yield higher frequency 1/f corners in the region of 100 MHz.

In addition to oxide traps, MOSFETs exhibit generation/recombination noise, which is a carrier-trap phenomenon in the bulk semiconductor that causes a fluctuation in the number of carriers in the conduction channel and, thus, an apparent fluctuation in the channel resistance. This mechanism generates a Cauchy spectral distribution, which some literature calls a Lorentzian distribution.

Burst, or "popcorn," noise generates fluctuations between two potential states. The burst-noise rms amplitude is proportional to current and stays level up to its corner frequency, at which point it falls at a rate of $1/f^2$. Different burst-noise mechanisms within the same device can

TABLE 1—THERMAL VOLTAGE NOISE AT ROOM TEMPERATURE

Resistance (Ω)	Noise voltage spectral density ($nV/\sqrt{Hz}$)	Applications
1	0.127	Reference
50	0.899	Instrumentation and RF
75	1.10	CATV
100	1.27	Categories 5 and 6, platinum RTD
150	1.56	Professional microphones
600	3.12	Professional audio, DSL, plain-old telephone system
1000	4.02	Reference
4700	8.72	Consumer audio
10,000	12.7	Reference
47,000	27.6	Consumer audio
1 million	127	Reference

Notes:
Temperature=293K. Boltzmann's Constant=1.38E-23J/K.

exhibit different corner frequencies. When superimposed on the flicker noise, burst noise can cause bumps in the flicker noise's otherwise-straight spectral slope. Neither flicker nor burst noise result in Gaussian amplitude distributions, which makes it difficult to extrapolate reliable trends from a small set of measurements.☐

REFERENCES

1. Liu, Shih-Chii, Jörg Kramer, Giacomo Indiveri, Tobias Delbrück, and Rodney Douglas, *Analog VLSI: Circuits and Principles*, 2002, Bradford Books, MIT Press, pg 314.

2. Gray, Paul R, and Robert G Meyer, *Analysis and Design of Analog Integrated Circuits, Third Edition*, 1993, John Wiley and Sons, pg 716.

3. Israelsohn, Joshua, "Pour your own programmable analog," *EDN*, June 12, 2003, pg 38.

4. Lundberg, Kent H, "Noise sources in bulk CMOS," http://web.mit.edu/klund/www/CMOSnoise.pdf.

5. Johnson, John B, "The Schottky effect in low frequency circuits," *Physical Review*, July 1925, pg 71.

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