D-C AMPLIFIER NOISE REVISITED

Understanding, Measuring, and Testing for Random Noise

A New Op-Amp Noise Fixture for Automatic Benchtop Tests with LTS-2010

by Al Ryan and Tim Scranton

"Like diseases, noise is never eliminated, just prevented, cured, or endured, depending on its nature, seriousness, and the costs/difficulty of treating it."[1]

In **Analog Dialogue** *3-1 (1969)—now out of print—we published an article entitled "Noise and Operational Amplifier Circuits,"[2] intended to be a readable and easily applicable essay on noise principles, with primary emphasis on random noise, in a treatment distinguished more by vigor than by rigor[3]. On this, its 15th anniversary, it seems appropriate to review the salient tutorial, proffer data on modern IC op amps, and discuss the test means embodied in a new test socket assembly for automatic noise testing, employing the LTS-2000 family of benchtop device testers.*

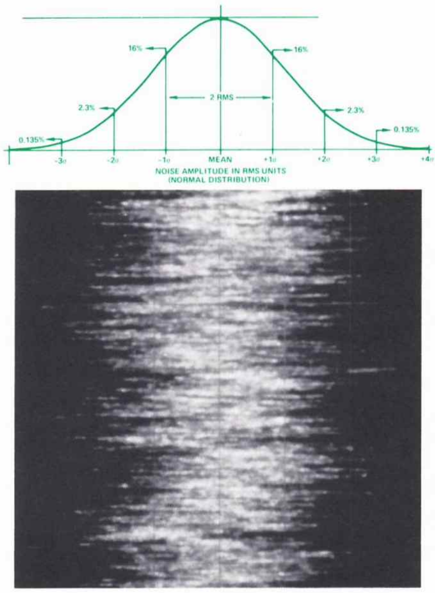

Figure 1. Gaussian amplitude distribution, juxtaposed against random-noise waveform.

WHY BE CONCERNED ABOUT NOISE?

An understanding of noise is needed to know what the real resolution of your system is. A knowledge of noise characterization and testing is needed to evaluate an amplifier or other purchased device with a given set of noise specifications.

As conversion systems employ increased digital resolution, the value of the least-significant bit decreases. For example, the LSB of a 10-bit system with 10V full scale is about 10 millivolts, the LSB of a 12-bit system is about 2.5 millivolts, and the LSB of 16 bits is 153 microvolts. This, by itself, poses significant problems in converter design.

However, the real-world measurement situation is worse, because most signals derived from real-world sources have full-scale amplitudes considerably less than 10 volts and must be amplified. If, for example, the original signal has a full-scale level of 10 mV (not unusual in transducer applications) and thus requires amplification of 1000×, the 12-bit LSB would become about 2.5 μV. If the preamplifier and its associated active and passive circuit elements generate only 1 μV of noise—or 100 pA in 10,000 ohms—(referred to the input) when the signal is sampled, they will significantly affect the accuracy, perhaps neutralizing the resolution.

Noise in data-acquisition systems takes three basic forms, *transmitted noise*—inherent in the received signal, *device noise*—generated within the devices used in data acquisition (preamps, resistors, etc.), and *induced noise*—picked up from the outside world, power supplies, logic, or other analog channels by magnetic, electrostatic, or galvanic coupling.

Transmitted noise must be dealt with, to the degree possible, by reducing noise at the source or in the transmission medium, and otherwise by making use of statistical analysis and filtering to distinguish between properties of the signal and of the noise, often employing techniques of digital signal processing (DSP).[4]

Induced noise is affected by both electrical design choices and physical layout. Purposeful use of design techniques to predict and

[1] *Analog-Digital Conversion Notes*, ed. by D. H. Sheingold, Analog Devices, Inc., $5.95.
[2] Based on work done by L R. Smith at Analog Devices.
[3] A bibliography on random noise will be found on page 30.
[4] Good information and bibliographies on DSP can be found in articles recently published in *Analog Dialogue*: "CMOS ICs for Digital Signal Processing," (17-1, 1983) and "Digital FIR Filters without Tears" (17-2, 1983). Also available free from Analog Devices: "A Cookbook to Digital Filtering and Other DSP Applications," a collection of reprints of papers that originally appeared in EDN magazine during 1983.

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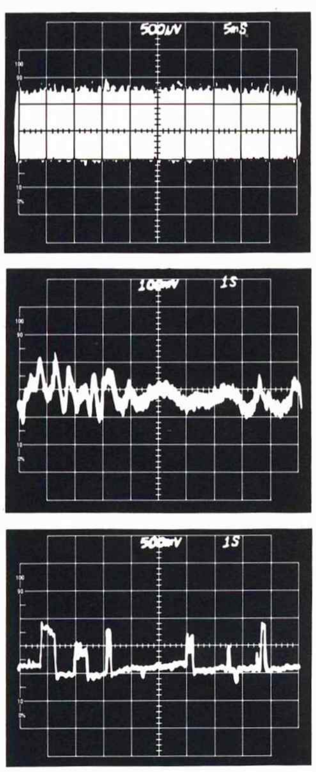
reduce induced noise was discussed in *Analog Dialogue* in a two-part series by Alan Rich.[5]

This article is about the irreducible minimum—*device noise*—its properties, how it affects circuit performance, and how it is measured. Later in the article, we will describe a test fixture that can be used for noise tests employing the Analog Devices LTS-2000 family of automatic benchtop testers.

RANDOM NOISE

Resistors and semiconductor junctions generate *random* noise: the output amplitude at a given instant is uncorrelated with the output amplitude at some other instant, and any value of output is possible at any instant—but it cannot be predicted. When a large number of samples are taken, many stationary random processes have amplitude distributions that appear *Gaussian*, i.e., characterized by the familiar bell-shaped curve (Figure 1). If the distribution is of voltages, and if care has been taken to eliminate dc offsets, then the rms voltage will be equal to the standard deviation of the distribution. Since random noise tends to be a stationary process,* the rms value tends to be constant for a given bandwidth and/or (sufficiently long) averaging interval.

Device-produced random noise, when plotted on an oscilloscope screen, can be identified in terms of several characteristic signatures. The noise output of a single device may combine all three types. Figure 2 (a) shows broad-spectrum "white" noise, which may be thermal (Johnson), or shot (Schottky); it has a constant distribution across the frequency spectrum but looks as if it is heavily oriented to the higher frequencies. Figure 2 (b) shows "pink", or "1/f", or "flicker" noise, dominated by low frequencies, and (c) illustrates "popcorn" noise (so-called because of its sound when presented audibly), characterized by random jumping between two or more levels.



(a) White noise.

(b) 1/f noise.

(c) Popcorn noise.

Figure 2. Noise signatures.

[5]"Understanding Interference-Type Noise," *Analog Dialogue* 16-3; and "Shielding and Guarding," *Analog Dialogue* 17-1

*The statistical properties of a stationary process are invariant under a shift of the time origin.

Since popcorn noise consists of jumps having approximately equal amplitudes, devices that have popcorn noise tend to have anomalous sharp peaks in the amplitude distribution, indicating that the number of such amplitudes greatly exceeds their probability in a purely random distribution.

White noise is present in all resistors and in semiconductor junctions. Resistor noise is thermal noise; its average power in a given bandwidth depends on temperature. Semiconductor junction noise is known as shot noise; its rms current variation in a given bandwidth is a function of the dc current through the junction.

"Excess noise" comprises all device noise that exceeds theoretical white-noise levels; it may be found when current flows through resistors, especially carbon composition types; it contains both shot noise and flicker components that are due to contact between carbon granules in resistors. For this reason, good practice mandates metal-film or wirewound resistors whenever low noise is required in the presence of significant dc current flow.

Flicker noise in semiconductors is due to random fluctuations in the number of surface recombinations; and "popcorn noise," a negative indicator of semiconductor process quality, is due to random on/off recombination action in the semiconductor material, leading to erratic switching of an affected device's current gain.

QUANTITATIVE MEASURES

The average *Johnson noise power*, generated by thermal agitation of the electrons in a resistor, and dissipated in a circuit containing a matched resistive load of equal magnitude is given by (1):

$$P_n = k\,T\,B \text{ watts} \tag{1}$$

where

k = Boltzmann's Constant = 1.381×10^{-23} joules/kelvin

T = Absolute temperature, kelvins (°C + 273.2°)

B = Bandwidth, $f_2 - f_1$, ("brick wall"*) in hertz.

In a series circuit with a resistor, R, containing a noise generator, E_n, and a matched resistor, of value R, the power dissipated in the load resistor is $(E_n/2)^2/R$, therefore

$$E_n = \sqrt{4\,k\,T\,R\,B} \text{ volts} \tag{2}$$

where E_n is the rms value of voltage generated in source resistance R. Near room temperature, and with more-convenient units (i.e., T = 300 K, and R × B in meg(ohm-Hz)

$$E_n = 0.129\sqrt{R \times B} \text{ microvolts rms} \tag{3}$$

Example: if R = 1kΩ and B = (5 kHz − 4 kHz) = 1 kHz, E_n is equal to 0.129 μV. If R = 100Ω and B = 1 kHz, E_n = 41 nV.

Johnson noise is quite often expressed in terms of an equivalent current source, I_n, in parallel with the resistor,

$$I_n = E_n/R = 0.129\sqrt{\frac{B}{R}} \tag{4}$$

where B is in hertz and R is in megohms.

Shot noise, caused by current flowing through a junction, is normally expressed as an ac current component, added to the dc value of I; it will produce voltage drops in series impedances, such as transistor emitter resistance or op-amp computing resistors:

$$I_n = \sqrt{2\,e\,I\,B} = 5.66 \times 10^{-10}\sqrt{I\,B} \text{ amperes rms} \tag{5}$$

where

*Perfect sharp-cutoff filtering.

e = Unit charge, 1.602×10^{-19} coulombs
I = DC current flowing through the junction, amperes
B = Bandwidth, $f_2 - f_1$ hertz ("brick wall")

Expressed in more convenient units, if I is in microamperes,

$$I_n = 0.566\sqrt{IB} \text{ picoamperes} \tag{6}$$

NOISE DENSITY SPECTRUM

Noise exists in all parts of the frequency spectrum; the noise contribution of a resistor or amplifier varies with the bandwidth over which the observation is made. Characterization of noise in terms of its *spectral density*, i.e., as a function of frequency, makes calculation of noise in differing bandwidths in active circuits easier. A useful way of presenting noise characteristics graphically is via a plot of noise spectral density vs. frequency.

The power spectral density is defined as the derivative of noise power with respect to frequency (watts per hertz), i.e.,

$$p_n = \frac{dP_n}{df} \tag{7}$$

Since power is proportional to the square of rms voltage or current, the respective expressions for voltage and current noise spectral density are:

$$e_n = \sqrt{\frac{dE_n{}^2}{df}} \text{ and } i_n = \sqrt{\frac{dI_n{}^2}{df}} \tag{8}$$

with units of $V/\sqrt{Hz}$ and $A/\sqrt{Hz}$. Figure 3 shows typical spectral density plots of AD741 voltage and current noise, (a) and (b), compared with the low-noise bipolar AD OP-27's e_n and i_n (c) and the the low-drift FET-input AD547's e_n (d). Voltage noise tends to have a lower "corner" frequency (see 1/f noise, next page) than current noise in bipolar amplifiers. Illustrating different ways of presenting the data, the scales of (a) and (b) show mean-square voltage and current spectral density per hertz, while (c) and (d) show rms voltage and current spectral density per $\sqrt{Hz}$; (d) is a semilog plot with a linear voltage scale for greater sensitivity to small noise voltage changes in the vicinity of the corner frequency.

The rms voltage or current in a given frequency band is determined by integrating the expressions in (8), i.e.,

$$E_n\,[f_1 \text{ to } f_2] = \sqrt{\int_{f1}^{f2} e_n{}^2 df} \tag{9}$$

Spot Noise. If $f_2 - f_1$ is quite small, we can assume that e_n is essentially constant (or "white") in that band (or "spot"), and $E_n = e_n\sqrt{f_2 - f_1}$. Using this, we can derive noise spectral density from a series of narrow-band measurements; the individual values of e_n are $E_n/\sqrt{f_2 - f_1}$. Enough points are observed to define a curve. These measurements are not easy to make, especially at low frequencies. By the same token, we can integrate a spectral-density curve of any shape by dividing the spectrum into small enough increments and taking the root sum-of-squares of the individual rms spot-noise contributions.

Graphical integration can be used to good effect, because plotted data based on measurements is seldom describable by easy formulas, and bandwidths aren't "brick-wall." However, approximations are often available that make it unnecessary to perform actual integrations over wide ranges of the noise characteristic.

EXAMPLES OF SPECTRAL DENSITY

Some simple rules for calculating noise over any bandwidth can

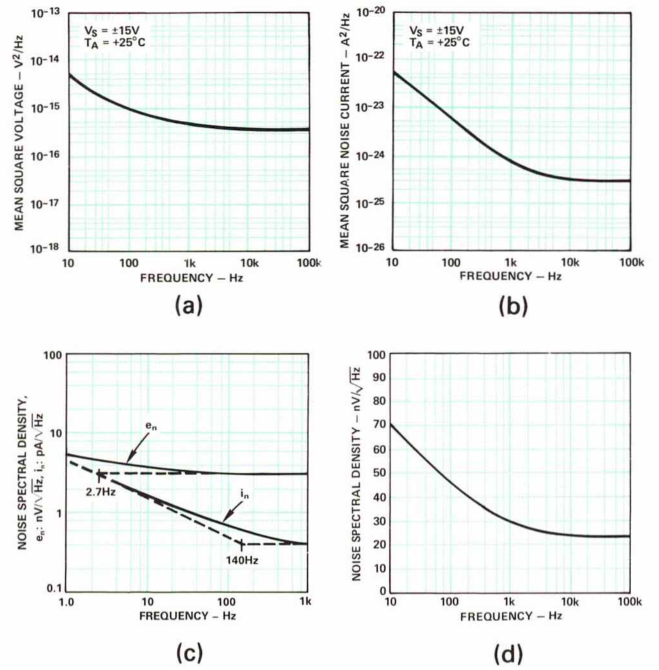


Figure 3. Noise spectral density plots. (a) Voltage-squared/Hz vs. frequency for AD741. (b) Amperes-squared/Hz vs. frequency for AD741, log-log plot. (c) Voltage and current per root hertz for AD OP-27, log-log plot. (d) Voltage per root hertz for AD547, semi-log plot.

be ascertained by the examination of two common forms of noise, white noise and 1/f noise.

White noise. As equations (1) and (5) show, white-noise power is of the form, (constant) $\times$ $(f_2 - f_1)$, which can be recognized as the definite integral of (constant) with respect to frequency between the limits of f_2 and f_1. For white noise, the spectral density is thus a theoretical or measured constant. Similarly, the voltage and current spectral densities, e_n and i_n, are equal to the square-roots of the respective constants ($0.129\sqrt{R}$ microvolts for Johnson voltage, $0.129/\sqrt{R}$ picoamperes for Johnson current (R in megohms), and $0.566\sqrt{I}$ picoamperes for shot current (I in μA). The right-hand portions of the graphs in Figure 3 illustrate the approximately constant level of white noise at the higher frequencies.

From the curves in the white noise regions, wideband noise for any bandwidth is easily calculated as the product of voltage or current spectral density and the square root of the bandwidth $(f_2 - f_1)$. For voltage noise, if e_n is given:

$$E_n = e_n \times \sqrt{f_2 - f_1} \tag{10a}$$

If $(e_n)^2$ is given,

$$E_n = \sqrt{(e_n)^2 \times (f_2 - f_1)} \tag{10b}$$

If $f_1 < 0.1\, f_2$, the error in assuming $f_2 - f_1 = f_2$ is less than 5%.

Example: Calculate the noise in the band, 100 Hz to 10 kHz, for each curve in Figure 3. The results are shown in the table:

Curve (Fig. 3)	Device	Quantity	Value from curve	Calc. rms noise
(a)	AD741	$(e_n)^2$	3.5×10^{-16} V²/Hz	1.86 μV rms
(b)	AD741	$(i_n)^2$	3×10^{-25} A²/Hz	54 pA rms
(c)	AD OP-27	(e_n)	3×10^{-9} V/$\sqrt{Hz}$	0.3 μV rms
(c)	AD OP-27	(i_n)	0.4×10^{-12} A/$\sqrt{Hz}$	40 pA rms
(d)	AD547	(e_n)	23×10^{-9} V/$\sqrt{Hz}$	2.3 μV rms

Although the values of e_n and i_n are not strictly constant over the band, the value of wideband noise is always dominated by the higher frequency; if f_2 is well into the white-noise region, the shape of the curve at the lower frequencies will have negligible effect on *wideband* noise measurements.

Flicker noise, or "1/f" noise, has a noise power spectral density varying inversely with frequency. It is of the form (constant)/f^γ; γ may be any value from 0 to 2, but it is usually close to 1. The shape of the left-hand portion of the noise curves in Figure 3 is due to 1/f noise, and the asymptotic slope—on a log-log plot—at the lowest frequencies is determined by γ. The point at which the projected 1/f asymptote crosses the average white-noise spectral density is the "corner frequency;" it is a measure of the quality of the device process—better devices have lower corner frequencies.

For voltage (and correspondingly for current),

$$e_n = K\sqrt{\frac{1}{f}} \tag{11}$$

K is the actual or extrapolated value of e_n at f = 1 Hz.

To compute E_n for the band, f_1 to f_2, using (9),

$$E_n\,(f_1 \text{ to } f_2) = K\sqrt{\int_{f1}^{f2}\frac{df}{f}}$$

$$= K\sqrt{\ln\left(\frac{f_2}{f_1}\right)} \tag{12}$$

The important result here is that equal amounts of 1/f noise will be generated in frequency bands having equal ratios. Every octave ($f_2 = 2\,f_1$) or decade ($f_2 = 10\,f_1$) in the 1/f-noise region will generate as much noise as every other octave or decade (i.e., $0.83 \times K$ per octave and $1.52 \times K$ per decade);and the noise generated over *m* octaves or decades will be $\sqrt{m}$ times as great as that generated in a single octave or decade.

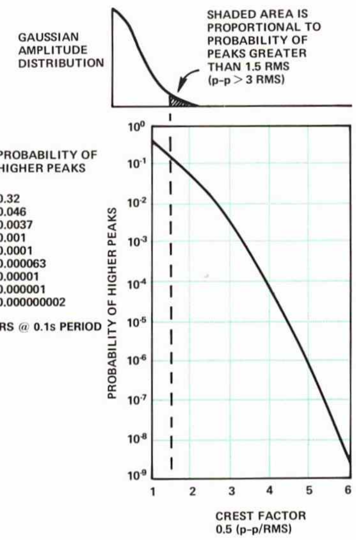
Consider, for example, the rms value of noise in the 9-decade realm below 1 Hz (down to about *1 cycle per 32 years*). If the rms value of noise in the decade, 0.1 Hz to 1.0 Hz, is 1 microvolt, then the noise over 9 decades will be $\sqrt{9} \times 1 = 3.0\,\mu V$! Over such long periods,—unless K increases—flicker noise can be expected to be less significant than drift caused by environmental factors, component aging, and perhaps even component life (however, there is evidence that increased 1/f noise—i.e., increased K—is often part of the end-of-life syndrome).

Example: What rms noise can be expected in the band, 0.1 to 10 Hz, for the devices plotted in Figure 3 (a), (b), and (c), extrapolating the curves to be asymptotic to a 1/f slope? The results are shown in the table.

Curve	K (Extrapolated)	Calculated E_n	Corner Frequency
(a)	$0.22\,\mu V/\sqrt{Hz}$	$0.47\,\mu V$ rms	100 Hz
(b)	$22\,pA/\sqrt{Hz}$	47 pA rms	2 kHz
(c) e_n	$4.9\,nV/\sqrt{Hz}$	12 nV rms (white)	2.7 Hz
i_n	$4.7\,pA/\sqrt{Hz}$	10 pA rms	140 Hz

Noise amplitude and Crest Factor. The rms value of noise is a consistent measure and is especially useful when dealing with signal phenomena having stationary properties, such as ac waveforms. However, in many cases, the accuracy or resolution of data is affected by the *instantaneous* value of noise, for example, sampled dc measurements and one-shot responses. For these, we must know the expected peak or peak-to-peak values of noise.

Although all values of noise amplitude are theoretically possible, there is a rapid decrease in the likelihood of large values, corresponding to the residual area under the distribution curve (Figure 4) beyond a given value. For example, the probability of a *crest factor* (peak/rms) exceeding 3.3 is 0.1%; for a c.f. exceeding 6, it is 2×10^{-9}. Figure 4 is a table and plot of the percentage of time that noise can be expected to exceed a given nominal peak-to-peak value ($2 \times$ c.f.), for a Gaussian distribution.



P-P / RMS	PROBABILITY OF HIGHER PEAKS
2	0.32
4	0.046
6	0.0037
6.6	0.001
7.8	0.0001
8	0.000063
8.8	0.00001
9.8	0.000001
12*	0.000000002

* 1 ½ YEARS @ 0.1s PERIOD

Figure 4. Crest factor of gaussian noise. For a measurement period of 0.1 seconds, a crest factor of 6 can be expected to occur once in 1 ½ years.

Combining noise. Noise from uncorrelated sources adds as the square-root of the sum of the squares of the individual rms values,

$$E_n = \sqrt{(E_{n1})^2 + (E_{n2})^2 + \ldots + (I_{n1}\,R_1)^2 + \ldots} \tag{13}$$

Examples of uncorrelated sources are different resistors, transistors, diodes, amplifiers, etc. However, the magnitudes of the noise voltage and the current that it causes to flow through a resistor (or any impedance) are correlated.

Since noise values are random with time, they are also random with frequency; thus, it is reasonable to consider that rms noise in different "brick-wall" frequency bands is uncorrelated and may be combined in root-square fashion. In fact, the integral performs just that function, using infinitesimal bands and the "spot" value of spectral density at each. This suggests that, in the example of Figure 5, we can treat the total noise given by integrating the spectral distribution characteristic over the broad frequency range as the noise obtained by separately integrating over the numbered regions and vector-summing (root sum-of-squares) the results.

NOISE AND AMPLIFIERS

The model of a differential operational amplifier for noise, lumping the effects of all the internal sources, and referring noise to the input, is very similar to that for offsets and drift. As Figure 6 shows, it consists of an ideal noiseless amplifier, a noise-voltage generator in series with either of the inputs, and two uncorrelated noise-current generators, one in parallel with each input. Basic noise-test circuits resemble those for offset voltage and bias current.

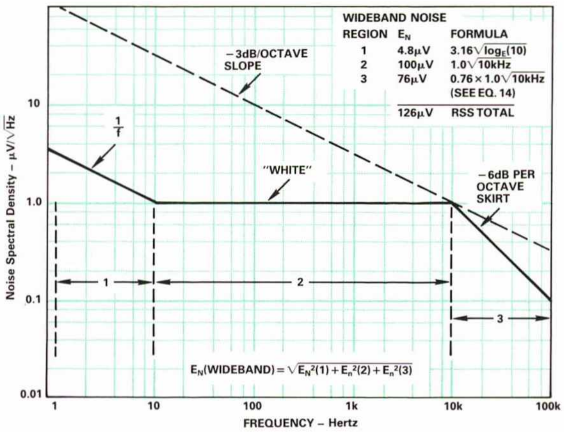


Figure 5. Wideband noise is equal to the root-square sum of components in (1), (2), and (3). For any upper-frequency limit, total noise can be well-approximated by summing just those components in the vicinity of a −3 dB/octave line lowered tangent to the curve. In this example, (1) can be ignored for the wideband case, but not for a cutoff at 20 Hz.

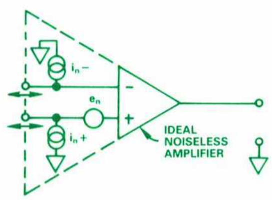


Figure 6. Operational amplifier noise model.

In instrumentation amplifiers, in addition to the input noise sources (especially relevant at high gains), there is also a noise source associated with the output stage; its effect, referred to the input, is seen principally when the amplifier is set for low gains.

Like offsets and drifts, voltage noise is amplified by the "noise gain," (the closed-loop gain of the amplifier with its feedback circuit). Current noise, flowing through impedances, produces noise voltage, just as bias current produces offset voltage.

In addition to the effects lumped within the amplifier, each external resistor's noise contributes to the overall noise total, as though it were in series with a voltage noise generator or in parallel with a current noise generator (the choice depends on which makes the analysis easier). The output noise, in a given bandwidth, depends on the contributions (to the output) of all independent noise sources. Since we assume the amplifier—and all the associated components—to be operating in the linear region, we may calculate the effects separately and combine them by superposition, but in root-square fashion, (they are uncorrelated random signals). In the example of Figure 7, the individual contributions are:

E_A	Amplifier voltage noise	Multiplied by $G = 1 + R_2/R_1$
I_{N-}	Current noise at − input	Multiplied by R_2
I_{N+}	Current noise at + input	Multiplied by $R_c \times G$
E_{Rc}	Voltage noise due to R_c	From eq. (2), multiplied by G
I_{R1}	Current noise due to R_1	From eq. (3), multiplied by R_2
E_{R2}	Voltage noise due to R_2	From eq. (2), direct

When performing root-square summation, it is worthwhile to remember that any noise source less than ¼ of the largest noise source contributes less than 3% to the total and can probably be ignored ($\sqrt{1 + 0.25^2} = 1.03$). Table 1 illustrates a typical calculation for the bandwidth, 100 Hz to 10 kHz, using values previously determined from the 741 curves in Figure 3 (a) and (b), the AD OP-27 curves in (c), and the AD547 voltage noise curve in (d).

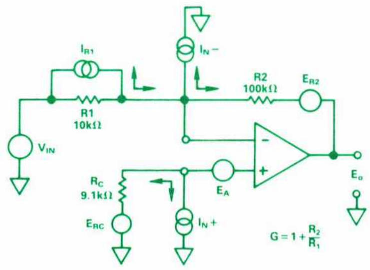


Figure 7. Noise sources in a gain-of-10 op-amp circuit.

Anticipating the reader's question, curves for current noise in FET-input op amps are not often published, because the noise is so low. In junction FETs, current noise is basically equal to the shot noise produced by the leakage current. For the AD547K, the maximum bias current is 25 pA at room temperature. Using (6), $i_n = 0.566\sqrt{0.000025\ \mu A}\ pA/\sqrt{Hz} = 2.8\ fA/\sqrt{Hz}$.

Table 1. Op Amp Noise Calculations in 100 Hz to 10 kHz BW (Δf = 9900 Hz)

Type	Source	RMS Noise	Factor	Output Contribution (μV)
AD741	E_A	1.86 μV	×11	20
	I_{N-}	54 pA	×0.1 M	5.5
	I_{N+}	54 pA	×0.009 M X11	5.5
	E_{Rc}	1.22 μV	×11	13.5
	I_{R1}	128 pA	×0.1 M	12.8
	E_{R2}	4.1 μ V	×1	4.1
	Total	(root sum-of-squares)		29
AD OP-27	E_A	0.3 μV	×11	3.3
	I_{N-}	40 pA	×0.1 M	4
	I_{N+}	40 pA	×0.009 M X11	4
	E_{Rc}	1.22 μV	×11	13.5
	I_{R1}	128 pA	×0.1 M	12.8
	E_{R2}	4.1 μV	×1	4.1
	Total	(root sum-of-squares)		20
AD547K	E_A	2.3 μV	×11	25.3
	I_{N-}	0.28 pA	×0.1 M	0.03
	I_{N+}	0.28 pA	×0.009M X11	0.03
	E_{Rc}	1.22 μV	×11	13.5
	I_{R1}	128 pA	×0.1 M	12.8
	I_{R2}	4.1 μV	×1	4.1
	Total	(root sum-of-squares)		31.6

In this example, it can be readily seen that, with an inverting gain of 10, the effective overall noise—*referred to the input signal*—is dominated by the AD741 and AD547's voltage noise—with significant contributions from the resistors—and amounts to about 3 μV in both cases. On the other hand, with the much quieter AD OP-27, the resistors are the principal source of noise—with about 2 μV, referred to the input signal.

Noise Figure. Even if there were no other sources of noise, the Johnson noise due to the input signal's source resistance would constitute a minimum. The noise contributed by the amplifier circuit, in relation to the noise generated by the source resistance, provides a means of characterizing the noise referred to the signal input. The circuit's *noise figure* is the logarithmic expression of the ratio of the total noise to that contributed by the source resistance. If no additional noise were produced, the total noise would be identical with the source noise and the ratio would be unity, thus the noise figure would be 0 dB. In this particular example, if we were to let the input resistance (R_1) represent the source resistance, the noise figure would become 7.1 (= $20 \log_{10}(29/12.8)$), 3.9, and 7.8 dB for the three cases.

Equivalent noise resistance. The amplifier circuit's noise may be characterized in terms of the Johnson noise, generated by the additional resistance that would be required to generate an equivalent amount of noise in the presence of an otherwise noise-free circuit; for the above examples, 41, 14, and 51 kilohms.

DYNAMICS

The noise, in the calculations above, is considered to be broadband white noise, with brick-wall filtering and no frequency-shaping elements (e.g., capacitors) in the circuits. Real filters, though, tend to have reduced response near the extremities of the pass band and graduated responses ("skirts") beyond the cutoff frequency; they are likely to pass more noise than a filter having the same cutoff frequency and perfectly sharp cutoff characteristics.

A simple circuit that can introduce us to both real-world considerations is shown in Figure 8. It consists of a resistor, R, considered as a current source, in parallel with a capacitor, C. The noise voltage generated by this circuit will be identical to that obtained using R as a noise voltage generator and a first-order low-pass filter with time constant RC, and cutoff frequency, $1/(2\pi RC)$.

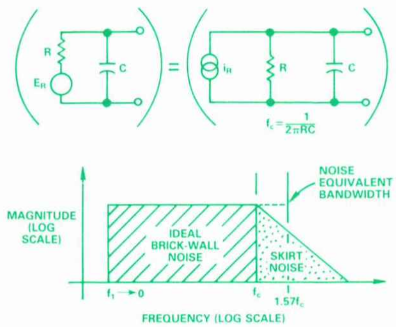


Figure 8. White noise voltage with first-order lag filtering is similar to noise produced by a resistor with a capacitor in parallel and the same RC.

To determine the output voltage in accordance with (9), multiply the resistor's noise current spectral density, i_n, which is flat with frequency, by the magnitude of the impedance of the parallel R-C to determine e_n, then square it, integrate from f_1 to f_2, and take the square-root. Assuming that f_1 is much smaller than f_c, and desiring to determine all the noise within the passband of the filter, we integrate over all f, 0 to ∞. Since the impedance magnitude is

$$|Z| = \frac{R}{\sqrt{1 + \left(\frac{f}{f_c}\right)^2}}$$

$$E_n = i_n R \sqrt{\int_{f1}^{f2} \frac{df}{1 + \left(\frac{f}{f_c}\right)^2}}$$

$$E_n = i_n R \sqrt{f_c} \sqrt{\tan^{-1}\left(\frac{f}{f_c}\right)\Big|_0^{\infty}} \qquad (14)$$

$$E_n = i_n R \sqrt{f_c} \sqrt{\frac{\pi}{2}}$$

$$= 1.25\, i_n R \sqrt{f_c}$$

Thus, the first-order filter's 6-dB-per-octave skirt causes a 25% increase over the noise measured with an ideal sharp-cutoff filter.

We can think of this in two ways. First, the first-order lag filter adds an amount of rms noise equal to $0.76\, e_n\sqrt{f_c}$, i.e., $\sqrt{1.25^2 - 1}$, root-sum-square, to the brick-wall value of $e_n\sqrt{f_c}$. Second, we can think of the filter as being equivalent to a brick-wall filter having an increased bandwidth, or *noise equivalent bandwidth* (N.E.B.) of $(\pi/2)f_c = 1.57\, f_c$. Noise equivalent bandwidth, a useful concept, differs for various filters, depending on order (1st, 2nd, 3rd, etc.), type (high-pass, low-pass, Butterworth, Chebyshef, etc.), and spectral density characteristics of the noise being filtered, e.g., white, pink, shaped.

In similar fashion, overall frequency content of output noise is determined by the shaping effect of the circuit's dynamic elements on the spectral density of the noise from each source. The noise spectral density at the output of the amplifier due to the amplifier's voltage noise (in the absence of current noise) is the product of the closed-loop gain transfer function and the noise spectrum. The output noise produced by noise current in an inverting op amp is determined by multiplying the noise-current spectrum by the impedance spectrum of the feedback element. The overall output noise spectrum is determined by adding the noise spectral densities of output voltage due to all sources, taken independently, at each frequency, in root-square fashion.

This is easier than it sounds, because in practice very little error is incurred by allowing the largest value of noise at any frequency to represent the r.s.s. sum at that frequency. In the few exceptional cases, the largest values can be r.s.s.-added; for this kind of calculation, a calculator that performs vector addition is useful. A graphical technique is helpful in visualization.

ANOTHER EXAMPLE

Consider the example of Figure 9, a dynamic version of one of the examples given previously, employing the 741-type op amp and assuming a feedback capacitance, in parallel with R_2, of 100 pF. The AD741 has a −6 dB/octave rolloff and unity-gain bandwidth of 1 MHz; assume an input capacitance (amplifier and wiring) of 10 pF from the inverting input to common, and that R_c is bypassed by a 0.1 μF capacitor. We wish to determine the overall noise spectral density and the wideband rms noise.

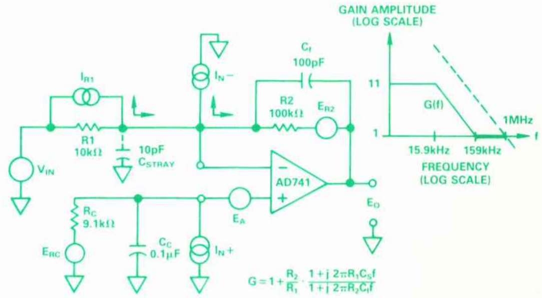


Figure 9. Dynamic version of Figure 7.

Table 2 describes the noise sources, the spectral density of the noise generated by each source, and its gain-bandwidth-output function. For the circuit shown, the noise gain is 11 V/V at dc, breaking at 15.9 kHz, dropping at 6 dB/octave to unity at frequencies above 159 kHz, and rolling off with the amplifier's gain at 1 MHz.

Figure 10 is a plot of the individual contributions of all six sources to the output, plus a plot of the combined noise spectral density, obtained by r.s.s. summation at various critical frequencies (e.g., 1 Hz, 100 Hz, 175 Hz, 300 Hz, 1 kHz, 2 kHz, 16 kHz, 160 kHz). Note that, at low frequencies, the amplifier's voltage and current 1/f noises are the most-significant contributors, while at high frequencies, e_A and R_1 are the most-significant contributors. R_2's contribution is insignificant, and R_c is a minor contributor, principally in the 100-200 Hz range.

Table 2. Calculations for spectral density plots of Figure 10.

Noise Source	Frequency or Band Hz	Generated Spectral Density	Output Spectral Density, $\mu V\sqrt{Hz}$ 1 Hz	100 Hz (e_A corner)	175 Hz (0.1 μF corner)	2 kHz (I_N corner)	15.9kHz (C_f corner)	159 kHz (C_s corner)	1 MHz (AD741 corner)
e_A	1	220 nV/$\sqrt{Hz}$	2.4	0.35	0.29	0.22	0.2	0.02	0.02
	10	71							
	100	32							
	175	26							
	2k +	20							
i_{N-}	1	22 pA/$\sqrt{Hz}$	2.2	0.23	0.19	0.071	0.055	0.0055	0.0009
	10	7.1							
	100	2.3							
	175	1.8							
	2k	0.71							
	10k +	0.55							
i_{N+}	Same as for i_{N-}		2.2	0.23	0.19	0.006	–	–	–
R_c	Broadband	12.2 nV/$\sqrt{Hz}$	0.134	0.134	0.134	0.012	–	–	–
R_1	Broadband	1.29 pA/$\sqrt{Hz}$	0.129	0.129	0.129	0.129	0.129	0.013	0.002
R_2	Broadband	41 nV/$\sqrt{Hz}$	0.041	0.041	0.041	0.041	0.041	0.004	–

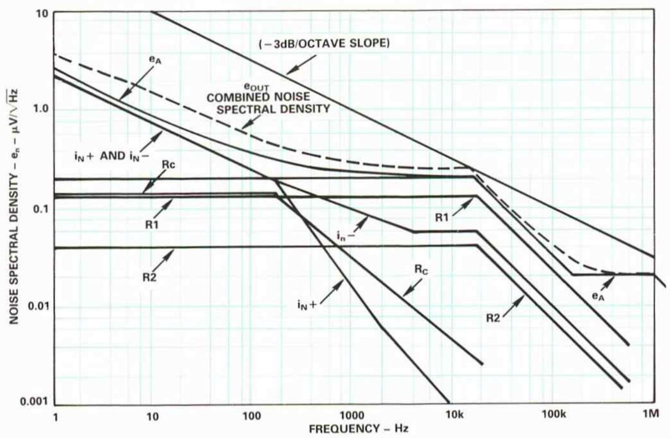


Figure 10. Output voltage contributions of noise sources in the circuit of Figure 9 and combined noise spectral-density.

The rms noise can be approximated in 3 "brick-wall" regions: 1/f, from below 1 Hz to about 400 Hz (extrapolated corner), with K = 3.9 μV/$\sqrt{Hz}$, white noise with e_n = 250 nV/$\sqrt{Hz}$, from that point to the 15.9 kHz rolloff, and white noise at 20 nV/$\sqrt{Hz}$, from about 159 kHz to the 1-MHz amplifier rolloff. The contributions and sum, referred to output and signal input, are:

		R.T.O.	R.T.I.
1/f noise (400Hz down to 0.1 Hz): $3.9\sqrt{\ln(4000)}$	=	11 μV rms	1.1 μV rms
Wideband noise (to 15.9 kHz and including the skirt from 15.9 kHz)): $0.25\sqrt{15.9\text{ kHz}} \times 1.25$	=	39 μV rms	3.9 μV rms
Total noise in passband:		41 μV rms	4.1 μV rms
H-F out-of-band noise from 159 kHz to 1 MHz and including the NEB of the 1-MHz skirt: $0.02\sqrt{(1.57 \times 1 - 0.159)\text{MHz}}$	=	24 μV rms	2.4 μV rms
Total rms noise:		48 μV rms	4.8 μV rms

NOISE TESTING

In order for users to predict that an amplifier will be sufficiently quiet for a given application, with a guarantee that goes beyond the "typical" curves given in the data sheet, the manufacturer must guarantee a set of maximum noise specifications. Here is an example, the specifications of noise for the AD OP-07:

Band	Typical	Maximum	Units	Remarks
0.1 Hz–10 Hz	0.35	0.6	μV p-p	Voltage noise
f_o = 10 Hz	10.3	18.0	nV/$\sqrt{Hz}$	V spectral density
f_o = 100 Hz	10.0	13.0	nV/$\sqrt{Hz}$	V spectral density
f_o = 1 kHz	9.6	11.0	nV/$\sqrt{Hz}$	V spectral density
0.1 Hz–10 Hz	14	30	pA p-p	Current noise
f_o = 10 Hz	0.32	0.80	pA/$\sqrt{Hz}$	I spectral density
f_o = 100 Hz	0.14	0.23	pA/$\sqrt{Hz}$	I spectral density
f_o = 1 kHz	0.12	0.17	pA/$\sqrt{Hz}$	I spectral density

Noise is usually specified in this way, peak-to-peak at low frequency and either spectral density or band-limited rms noise at higher frequencies. The manufacturer further specifies whether these specs have been 100% tested in production or on a sampling basis. Whether the noise is tested in characterization, in production, or by the user to verify noise performance of an incoming lot of devices, the test definition and circuitry should be standardized.

Figure 11 shows a typical test scheme, and the measurement circuitry employed. The device under test (DUT) is connected for a noise gain of (say) 10. A pair of 1-megohm resistors in series with the input terminals permit current noise to be measured, and a pair of shorting switches permit voltage noise and the current noise at either input to be measured individually. The output of the DUT is applied to a bandpass filter with a further gain of 10; the high-pass makes the noise measurement independent of the DUT's dc offset. The low-pass filter is a two-pole Butterworth. With the high-pass, the filter is programmable for 7 different frequency bands. Measurements are performed by either the rms-to-dc converter or the system voltmeter (for discrete sampling).

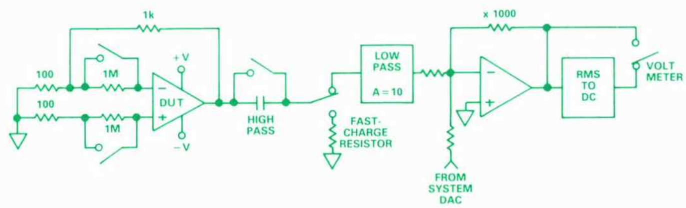


Figure 11. Noise-test circuit for op amps.

The bandpass filter has a noise effective bandwidth (N.E.B.) comparable to the band in question, e.g., 10 to 1000 Hz. The test system must take into account the DUT's reduction of bandwidth with increased gain, if it is a factor in the measurement. A wideband true-rms voltmeter measures the output of the filter and either interprets it directly as rms noise in the band, or calculates noise spectral density. At lower frequencies, the output is sampled, and calculations are performed to determine the rms, the peak-to-peak amplitude, and the presence of popcorn noise.

Both white noise and 1/f noise are random and have benign statistics. However, popcorn noise with sufficient amplitude to be readily detectable is of concern because it is an indicator of defective processing. Also, its "burst" nature (over periods from microseconds to seconds) is particularly annoying in many applications. When looked at with an oscilloscope, it is readily visible as a shift between levels (Figure 2c). How does a *computer* determine within a short time if substantial popcorn noise is present?

In one easy-to-automate approach, the criterion used to identify popcorn noise is: for successively sampled values of the output of the DUT and filter, are there any differences (jumps) that exceed 4 × rms (4 σ)? If so, these jumps, which have a probability of less than 0.01%, are taken to be "hits" of popcorn-noise (Figure 12).

NOISE TESTING WITH THE LTS-2000 FAMILY

The LTS-2000 family of device testers are powerful low-cost computer-based self-contained benchtop test systems that can test a wide gamut of linear, digital, active, and passive devices, including amplifiers, d/a and a/d converters, digital logic ICs, discrete semiconductors, and passive components. Calibration is automatic and tests are performed quickly and automatically in accordance with either menu-driven or BASIC-programmed software pro-

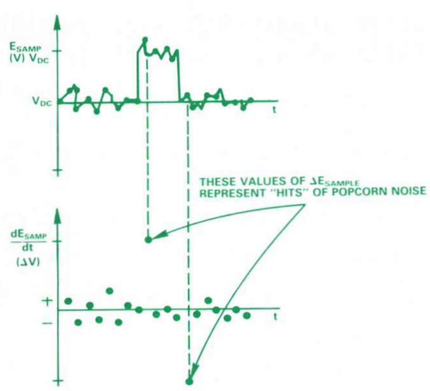


Figure 12. Examining the sampled signal for popcorn noise by differencing.

grams. Documentation is available as pass/fail, quantitative, and statistical information.

Flexible hardware, as well as software, programs the system for testing specific devices. Plug-in family boards provide the rapid-change circuitry essential to testing such families as DACs, ADCs, amplifiers, etc. Circuitry for performing general- or special-purpose tests (such as noise) is wired to a device-test socket assembly, which plugs into the family board. Devices under test are plugged into test-socket cards, which are wired for specific devices and plugged into the socket assembly.

The LTS-0613 Noise-Test Socket Assembly (soon to be announced*), in conjunction with the op-amp family board, enables users of the LTS-2000 family to perform noise testing on op amps—a capability available for the first time in benchtop ATE.

When used in conjunction with the LTS-2100 Op-Amp Family Board, the LTS-0613 tests noise, as well as the full range of parameter tests that the LTS-0600 Op-Amp Socket Assembly has been providing, for single, dual, and quad packages. It permits measurement of noise spectral density and peak-to-peak values for both voltage and current noise over a range of seven different bandwidths; the associated software also provides a routine for detecting popcorn noise.

The LTS-0613's circuit is quite similar to the block diagram of Figure 12, and—in fact—performance of the tests is similar to the technique described above. Automatic switching is available to choose among the four op amps in a quad. The high-pass filter, which serves the function of blocking any DUT offsets, can be initially switched—prior to testing—to have low values of resistance so that the blocking capacitor can be rapidly charged; without it, the wait for the circuit to settle before performing tests at the low end (0.1 to 10 Hz) would greatly reduce the speed of testing.

The DUT is wired for a gain of 10, a compromise that allows its noise to be amplified to a level significantly higher than that of the filter and gain stages, while retaining as large a bandwidth as possible for measurement of noise in the higher-frequency bands. Gain is also taken in the filter, which employs low-noise amplifiers, and in a pair of post-filter amplifiers. The overall gain from the DUT input is 100,000 V/V.

Relay-switchable filter time constants permit tests over the following bands: 0.1 to 10 Hz; 0.1 to 100 Hz; 1 to 100 Hz; 1 Hz to 1 kHz; 10 Hz to 1 kHz; 10 Hz to 10 kHz; and 100 Hz to 10 kHz. Output to the system is software-switchable between an analog true-rms circuit, for wideband measurements, and individual samples, for the low frequencies. Sampling permits both accurate low-frequency rms computations and examination for popcorn hits.

The software is set up as a subroutine to be called from the user's BASIC test program. All variables (such as test limits, bandwidths, choice of voltage noise, current noise, or both) are listed in documentation that accompanies the socket assembly. The user simply establishes the appropriate variables in the test program and then calls the subroutine. The flow chart in Figure 13 is a diagram of the subroutine's procedure.

When the testing has been completed, the system prints the results in a form similar to the datalog shown in Figure 14, which depicts the test results (vs. limits) of 16 critical parameters for an AD OP-07E. ▶

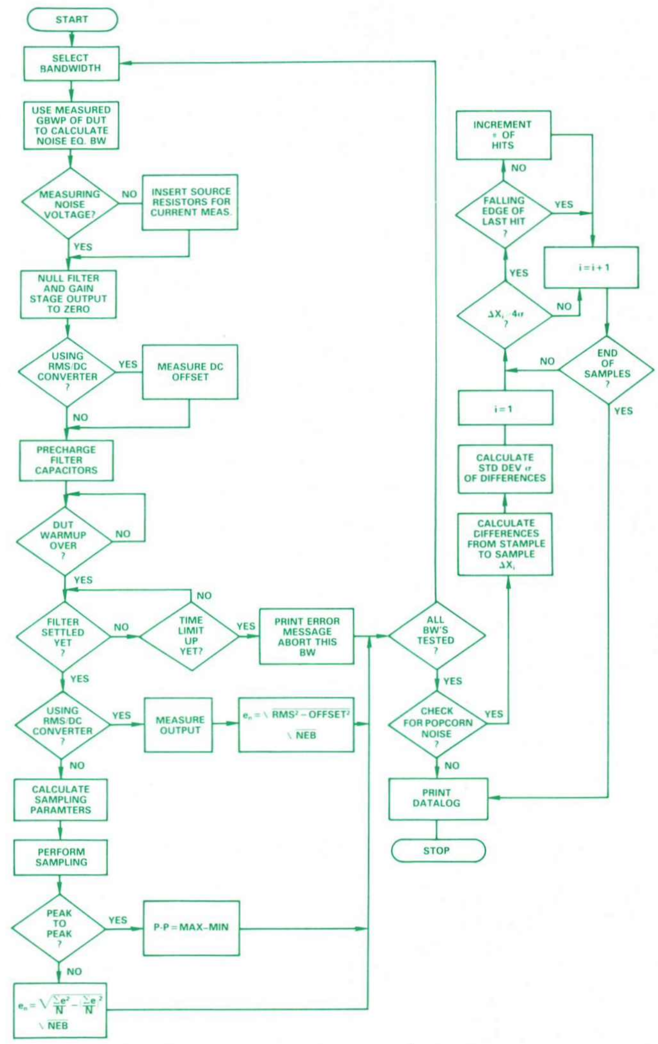


Figure 13. The LTS-0613 noise-test subroutine.

```
DEVICE 2 TESTING
T# 1       3.13 MA       +I SUPPLY                   [ 0.1 TO 4 MA ]
T# 2      -3.13 MA       -I SUPPLY                   [ -4 TO -0.1 MA ]
T# 3      33.38 UV       VOS 50 OHMS                 [ -75 TO 75 UV ]
T# 4       -2.4 NA       +I BIAS                     [ -4 TO 4 NA ]
T# 5       -3.0 NA       -I BIAS                     [ -4 TO 4 NA ]
T# 6         .7 NA       I OFFSET                    [ -3.8 TO 3.8 NA ]
T# 7     7654.8 K        AOL                         [ 300 TO 1E12 K ]
T# 8      140.0 DB       CMRR                        [ 106 TO 1E12 DB ]
T# 9        3.2 UV/V     PSRR                        [ -20 TO 20 UV/V ]
T# 10     14.01 V        +V OUT                      [ 10.5 TO 15 V ]
T# 11    -13.62 V        -V OUT                      [ -15 TO -10.5 V ]
T# 12      9.15 nV/RH   NOISE VOLT (100-10K HZ)      [ 0 TO 13 nV/RH ]
T# 13      9.47 nV/RH   NOISE VOLT (10-1K HZ)        [ 0 TO 18 nV/RH ]
T# 14       .23 pA/RH   NOISE CUR (10-1K HZ)         [ 0 TO 0.8 pA/RH ]
T# 15     13.86 pA      P-P NOISE I (.1-10 HZ)       [ 0 TO 30 pA ]
T# 16         0 HITS    POPCORN NOISE                [ 0 TO 0 HITS ]
PASS BIN 1
```

Figure 14. Data-log of test results on OP-07E, employing the LTS-0613. Columns are test number, actual measurement, description, and pass limits.

*For technical data, use the reply card.