can do wonders for the CMRR, at least at dc. At increasing frequencies you have the usual problems of presenting matched impedances to the input capacitances.

### *Two-op-amp configuration*

Figure 7.36 shows another configuration that offers high input impedance with only two op-amps. Since it doesn't accomplish the common-mode rejection in two stages, as in the three-op-amp circuit, it requires precise resistor matching for good CMRR, in a manner similar to that of the standard differencing amplifier circuit.

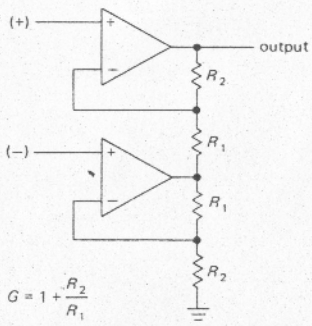


Figure 7.36. Instrumentation amplifier circuit with two op-amps.

### *Special IC instrumentation amplifiers*

There are several interesting instrumentation amplifier configurations available as monolithic (and therefore inexpensive) ICs, some with extremely good performance. They use methods unrelated to the preceding circuits.

☐ *Current-feedback amplifier technique.* This technique, typified by the LM363, AD521, and JFET AMP-05, achieves high CMRR without the need for matched external resistors. In fact, the gain is set by the ratio of a pair of external resistors. Figure 7.37 shows a block diagram of the AMP-01. The circuit employs two differential transconductance amplifier pairs, with a single external resistor setting the gain in each case. One pair is driven by the input signal, and the other is driven by the output signal, relative to the ref terminal. The AMP-05 uses FETs to keep input currents low, whereas the AMP-01 uses bipolar technology to achieve low offset voltage and drift (Table 7.5).

Computer-aided design methods can be extremely useful in precision circuit design; see Section 13.24.

## AMPLIFIER NOISE

In almost every area of measurement the ultimate limit of detectability of weak signals is set by noise – unwanted signals that obscure the desired signal. Even if the quantity being measured is not weak, the presence of noise degrades the accuracy of the measurement. Some forms of noise are unavoidable (e.g., real fluctuations in the quantity being measured), and they can be overcome only with the techniques of *signal averaging* and *bandwidth narrowing*, which we will discuss in Chapter 15. Other forms of noise (e.g., radiofrequency interference and "ground loops") can be reduced or eliminated by a variety of tricks, including filtering and careful attention to wiring configuration and parts location. Finally, there is noise that arises in the amplification process itself, and it can be reduced through the techniques of low-noise amplifier design. Although the techniques of signal averaging can often be used to rescue a signal buried in noise, it always pays to begin with a system that is free of preventable interference and that possesses the lowest amplifier noise practicable.

We will begin by talking about the origins and characteristics of the different

TABLE 7.5. INSTRUMENTATION AMPLIFIERS

| | | | Total supply | | | Maximum input errors[b] | | | | | | Noise | | | | | | | | | | | | | |
|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|---|
| | | | Voltage | | | Offset voltage | | | | Current | | Voltage 0.1-10Hz | | Voltage 10Hz-10kHz | | Current 10Hz-10kHz | CMRR (@dc, min) | | | -3dB bandwidth | | Bandwidth for 1% error | | Settling time to 1% | |
| | | | | | Curr max | RTI[a] | | RTO[a] | | Bias | Offset | RTI[a] | RTO[a] | RTI[a] | RTO[a] | | G=1 | G=1k | Slew rate | G=1 | G=1k | G=1 | G=1k | G=1 | G=1k |
| Type | FET | Curr fdbk | min (V) | max (V) | (mA) | (mV) | (μV/°C) | (mV) | (μV/°C) | (nA) | (nA) | (μV, pp) | | (μV, rms) | | (pA, rms) | (dB) | (dB) | (V/μs) | (kHz) | (kHz) | (kHz) | (kHz) | (μs) | (μs) |
| AMP-01A | – | • | 9 | 36 | 5 | 0.05 | 0.3 | 3 | 50 | 3 | 1 | 0.1 | 13 | 0.5 | - | - | 85 | 125 | 4.5 | | | 570 | 26 | 12 | 50 |
| AMP-05A | • | • | 10 | 36 | 10 | 1 | 10 | 15 | 100 | 0.05 | 0.025 | 4 | 7 | 3 | - | 1 | 90 | 110 | 7.5 | 3000 | 120 | - | - | 5 | 5 |
| LH0036 | – | – | 2 | 36 | 0.6 | 1 | $10^t$ | 5 | $15^t$ | 100 | 40 | - | - | 5 | - | - | 50 | 100 | 0.3 | 350 | 0.35 | - | - | 8 | 600 |
| LH0038[c] | – | – | 10 | 36 | 2 | 0.1 | 0.25 | 10 | $25^t$ | 100 | 5 | 0.2 | - | 0.6 | - | 10 | - | 114 | 0.3 | - | 1.6 | - | - | - | $80^d$ |
| INA101C | – | – | 10 | 40 | 8.5 | 0.025 | 0.25 | 0.2 | 10 | 20 | 20 | 0.8 | - | 1.5 | - | 50 | 80 | 106 | 0.4 | 300 | 2.5 | 20 | 0.2 | 30 | 500 |
| INA102C | – | – | 7 | 36 | 0.8 | 0.1 | 2 | 0.2 | 5 | 30 | 10 | 0.1 | - | 2.5 | 8 | 20 | 90 | 90 | 0.2 | 300 | 0.3 | 30 | 0.03 | 50 | 3300 |
| INA104C | – | – | 10 | 40 | 10 | 0.025 | 0.25 | 0.2 | 10 | 20 | 20 | 0.8 | - | 1.5 | - | 50 | 80 | 106 | 0.4 | 300 | 2.5 | 20 | 0.2 | 30 | 350 |
| INA110B | • | – | 12 | 36 | 4.5 | 0.25 | 2 | 3 | 50 | 0.05 | 0.025 | 1 | 8 | 5 | 10 | - | 80 | 106 | 17 | 2500 | $100^f$ | - | - | 4 | $11^f$ |
| LM363A | – | • | 10 | 36 | 2 | 0.05 | 0.5 | 10 | 250 | 5 | 2 | $0.4^h$ | $100^h$ | 1.2 | 100 | 15 | $100^g$ | 126 | 0.4 | $200^g$ | 30 | $30^g$ | 5 | $20^g$ | 70 |
| AD521 | – | • | 10 | 36 | 5 | 3 | 15 | 400 | 400 | 80 | 20 | 0.5 | 150 | 1.2 | 30 | - | 70 | 100 | 10 | 2000 | 40 | 75 | 6 | 7 | 35 |
| AD522 | – | – | 10 | 36 | 10 | - | 6 | 0.4 | 50 | 25 | 20 | 1.5 | 15 | - | 15 | - | 75 | 100 | 0.1 | 300 | 0.3 | - | - | $500^d$ | $20000^d$ |
| AD524C | – | – | 12 | 36 | 5 | 0.05 | 0.5 | 2 | 25 | 15 | 10 | 0.3 | 15 | 0.5 | 10 | 40 | 80 | 120 | 5 | 1000 | 25 | | | 10 | 50 |
| AD624C | – | – | 10 | 36 | 5 | 0.025 | 0.25 | 2 | 10 | 15 | 10 | 0.2 | 10 | 0.5 | 10 | 40 | 80 | 130 | 5 | 1000 | 25 | - | - | 10 | 50 |
| AD625C | – | – | 10 | 36 | 5 | 0.025 | 0.25 | 2 | 15 | 15 | 5 | 0.2 | - | 0.4 | 7 | 30 | 80 | 120 | 5 | 650 | 25 | - | - | $15^d$ | $75^d$ |
| ICL7605[e] | • | – | 4 | 18 | 5 | 0.005 | 0.2 | - | - | 1.5 | - | 1.7 | - | - | - | - | $100^t$ | $100^t$ | 0.5 | 0.01 | 0.01 | slow | | slow | |

(a) RTI: referred to the input; RTO: referred to the output. Noise and errors can be separated into components generated at both the input and output. The total input-referred noise (or error) is thus given by *RTI+RTO/G*. (b) diff'l input impedance > 1GΩ except LH0038 (5MΩ), AMP-05 (1TΩ), and INA110B (5TΩ). (c) gain range 10-2000. (d) to 0.01%. (e) CAZ type (see section 7.10); 7606 is uncomp. (f) G = 500. (g) G = 10. (h) 0.01Hz to 10Hz. (t) typical.

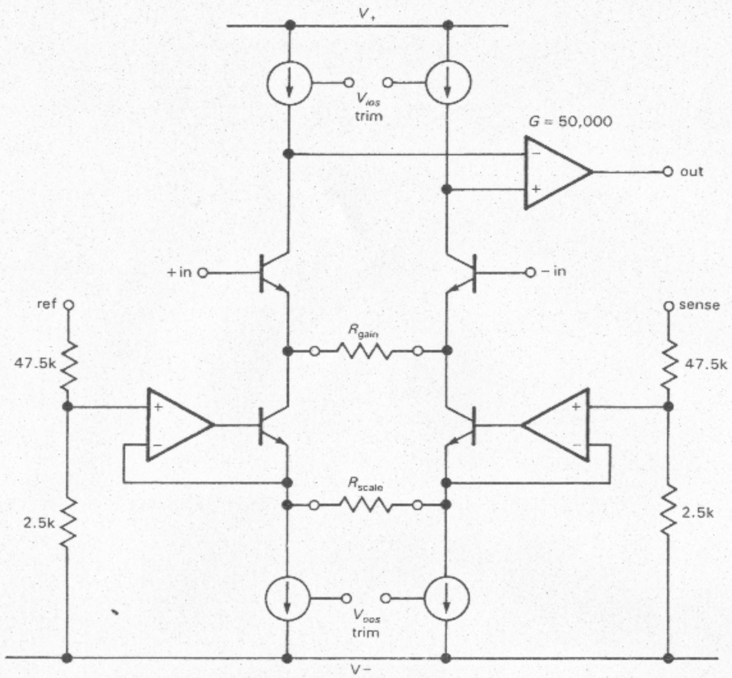


Figure 7.37. Block diagram of the AMP-01 instrumentation amplifier IC.

kinds of noise that afflict electronic circuits. Then we will launch into a discussion of transistor and FET noise, including methods for low-noise design with a given signal source, and will present some design examples. After a short discussion of noise in differential and feedback amplifiers, we will conclude with a section on proper grounding and shielding and the elimination of interference and pickup. See also Section 13.24 (Analog modeling tools).

## 7.11 Origins and kinds of noise

Since the term *noise* can be applied to anything that obscures a desired signal, noise can itself be another signal ("interference"); most often, however, we use the term to describe "random" noise of a physical (often thermal) origin. Noise can be characterized by its frequency spectrum, its amplitude distribution, and the physical mechanism responsible for its generation. Let's next look at the chief offenders.

### *Johnson noise*

Any old resistor just sitting on the table generates a noise voltage across its terminals known as Johnson noise. It has a flat frequency spectrum, meaning that there is the same noise power in each hertz of frequency (up to some limit, of course). Noise with a flat spectrum is also called "white noise." The actual open-circuit noise voltage generated by a resistance $R$ at temperature $T$ is given by



$$V_{\text{noise}}(\text{rms}) = V_{nR} = (4kTRB)^{\frac{1}{2}}$$

where $k$ is Boltzmann's constant, $T$ is the absolute temperature in degrees Kelvin (°K = °C + 273.16), and $B$ is the bandwidth in hertz. Thus, $V_{\text{noise}}(\text{rms})$ is what you would measure at the output if you drove a perfect noiseless bandpass filter (of bandwidth $B$) with the voltage generated by a resistor at temperature $T$. At room temperature (68°F = 20°C = 293°K),

$$4kT = 1.62 \times 10^{-20}\text{V}^2/\text{Hz} - \Omega$$

$$(4kTR)^{\frac{1}{2}} = 1.27 \times 10^{-10} R^{\frac{1}{2}} \quad \text{V/Hz}^{\frac{1}{2}}$$

$$= 1.27 \times 10^{-4} R^{\frac{1}{2}} \quad \mu\text{V/Hz}^{\frac{1}{2}}$$

For example, a 10k resistor at room temperature has an open-circuit rms voltage of 1.3$\mu$V, measured with a bandwidth of 10kHz (e.g., by placing it across the input of a high-fidelity amplifier and measuring the output with a voltmeter). The source resistance of this noise voltage is just $R$. Figure 7.38 plots the simple relationship between Johnson-noise voltage density (rms voltage per square root bandwidth) and source resistance.

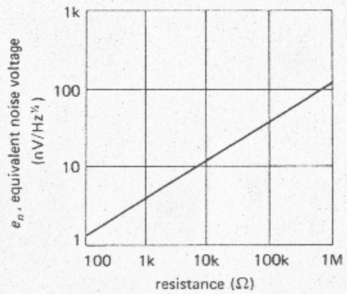


Figure 7.38. Thermal noise voltage versus resistance.

The amplitude of the Johnson-noise voltage at any instant is, in general, unpredictable, but it obeys a Gaussian amplitude distribution (Fig. 7.39),

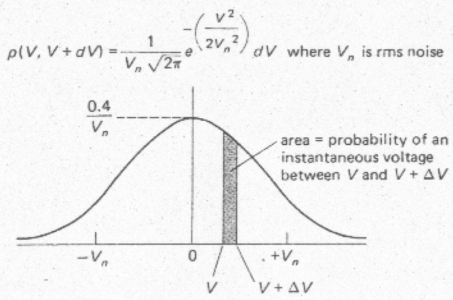


Figure 7.39

where $p(V)dV$ is the probability that the instantaneous voltage lies between $V$ and $V + dV$, and $V_n$ is the rms noise voltage, given earlier.

The significance of Johnson noise is that it sets a lower limit on the noise voltage in any detector, signal source, or amplifier having resistance. The resistive part of a source impedance generates Johnson noise, as do the bias and load resistors of an amplifier. You will see how it all works out shortly.

It is interesting to note that the physical analog of resistance (any mechanism of energy loss in a physical system, e.g., viscous friction acting on small particles in a liquid) has associated with it fluctuations in the associated physical quantity (in this case, the particles' velocity, manifest as the chaotic Brownian motion). Johnson noise is just a special case of this fluctuation-dissipation phenomenon.

Johnson noise should not be confused with the additional noise voltage created by the effect of resistance fluctuations when an externally applied current flows through a resistor. This "excess noise" has a $1/f$ spectrum (approximately) and is heavily dependent on the actual construction of the resistor. We will talk about it later.

### *Shot noise*

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 independent of each other, the fluctuating current is given by

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

where $q$ is the electron charge ($1.60 \times 10^{-19}$ coulomb) and $B$ is the measurement bandwidth. For example, a "steady" current of 1 amp actually has an rms fluctuation of 57nA, measured in a 10kHz bandwidth; i.e., it fluctuates by about 0.000006%. The relative fluctuations are larger for smaller currents: A "steady" current of 1$\mu$A actually has an rms current noise fluctuation, measured over a 10kHz bandwidth, of 0.006%, i.e., −85dB. At 1pA dc, the rms current fluctuation (same bandwidth) is 56fA, 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 (Fig. 2.21), in which negative feedback acts to quiet the shot noise.

EXERCISE 7.4

A resistor is used as the collector load in a low-noise amplifier; the collector current $I_C$ is accompanied by shot noise. Show that the output noise voltage is dominated by shot noise (rather than Johnson noise in the resistor) as long as the quiescent voltage drop across the load resistor is greater than $2kT/q$ (50mV, at room temperature).

### 1/f noise (flicker noise)

Shot noise and Johnson noise are irreducible forms of noise generated according to physical principles. The most expensive and most carefully made resistor has exactly the same Johnson noise as the cheapest carbon resistor (of the same resistance). Real devices have, in addition, various sources of "excess noise." Real resistors suffer from fluctuations in resistance, generating an additional noise voltage (which adds to the ever-present Johnson noise) proportional to the dc current flowing through them. This noise depends on many factors having to do with the construction of the particular resistor, including the resistive material and especially the end-cap connections. Here is a listing of typical excess noise for various resistor types, given as rms microvolts per volt applied across the resistor, measured over one decade of frequency:

| | |
|---|---|
| Carbon-composition | 0.10$\mu$V to 3.0$\mu$V |
| Carbon-film | 0.05$\mu$V to 0.3$\mu$V |
| Metal-film | 0.02$\mu$V to 0.2$\mu$V |
| Wire-wound | 0.01$\mu$V to 0.2$\mu$V |

This noise has approximately a $1/f$ spectrum (equal power per decade of frequency) and is sometimes called "pink noise." Other noise-generating mechanisms often produce $1/f$ noise, examples being base current noise in transistors and cathode current noise in vacuum tubes. Curiously enough, $1/f$ noise is present in nature in unexpected places, e.g., the speed of ocean currents, the flow of sand in an hourglass, the flow of traffic on Japanese expressways, and the yearly flow of the Nile measured over the last 2000 years. If you plot the loudness of a piece of classical music versus time, you get a $1/f$ spectrum! No unifying principle has been found for all the $1/f$ noise that seems to be swirling around us, although particular sources can often be identified in each instance.



### Interference

As we mentioned earlier, an interfering signal or stray pickup constitutes a form of noise. Here the spectrum and amplitude characteristics depend on the interfering signal. For example, 60Hz pickup has a sharp spectrum and relatively constant amplitude, whereas car ignition noise, lightning, and other impulsive interferences are broad in spectrum and spiky in amplitude. Other sources of interference are radio and television stations (a particularly serious problem near large cities), nearby electrical equipment, motors and elevators, subways, switching regulators, and television sets. In a slightly different guise you have the same sort of problem generated by anything that puts a signal into the parameter you are measuring. For example, an optical interferometer is susceptible to vibration, and a sensitive radiofrequency measurement (e.g., NMR) can be affected by ambient radiofrequency signals. Many circuits, as well as detectors and even cables, are sensitive to vibration and sound; they are *microphonic*, in the terminology of the trade.

Many of these noise sources can be controlled by careful shielding and filtering, as we will discuss later in the chapter. At other times you are forced to take draconian measures, involving massive stone tables (for vibration isolation), constant-temperature rooms, anechoic chambers, and electrically shielded rooms.

## 7.12 Signal-to-noise ratio and noise figure

Before getting into the details of amplifier noise and low-noise design, we need to define a few terms that are often used to describe amplifier performance. These involve ratios of noise voltages, measured at the same place in the circuit. It is conventional to refer noise voltages to the input of an amplifier (although the measurements are usually made at the output), i.e., to describe source noise and amplifier noise in terms of microvolts *at the input* that would generate the observed output noise. This makes sense when you want to think of the relative noise added by the amplifier to a given signal, independent of amplifier gain; it's also realistic, because most of the amplifier noise is usually contributed by the input stage. Unless we state otherwise, noise voltages are referred to the input.

### Noise power density and bandwidth

In the preceding examples of Johnson noise and shot noise, the noise voltage you measure depends both on the measurement bandwidth $B$ (i.e., how much noise you see depends on how fast you look) and on the variables ($R$ and $I$) of the noise source itself. So it's convenient to talk about an rms noise-voltage "density" $v_n$:

$$V_n(\text{rms}) = v_n B^{\frac{1}{2}} = (4kTR)^{\frac{1}{2}} B^{\frac{1}{2}}$$

where $V_n$ is the rms noise voltage you would measure in a bandwidth $B$. White-noise sources have a $v_n$ that doesn't depend on frequency, whereas pink noise, for instance, has a $v_n$ that drops off at 3dB/octave. You'll often see $v_n^2$, too, the mean squared noise density. Since $v_n$ always refers to rms, and $v_n^2$ always refers to mean square, you can just square $v_n$ to get $v_n^2$! Sounds simple (and it is), but we want to make sure you don't get confused.

Note that $B$ and the square root of $B$ keep popping up. Thus, for example, for Johnson noise from a resistor $R$

$$v_{nR}(\text{rms}) = (4kTR)^{\frac{1}{2}} \quad \text{V/Hz}^{\frac{1}{2}}$$

$$v_{nR}^2 = 4kTR \quad \text{V}^2/\text{Hz}$$

$$V_n(\text{rms}) = v_{nR}B^{\frac{1}{2}} = (4kTRB)^{\frac{1}{2}} \quad \text{V}$$

$$V_n^2 = v_{nR}^2 B = 4kTRB \quad \text{V}^2$$

On data sheets you may see graphs of $v_n$ or $v_n^2$, with units like "nanovolts per root Hz" or "volts squared per Hz." The quantities $e_n$ and $i_n$ that will soon appear work just the same way.

When you add two signals that are uncorrelated (two noise signals, or noise plus a real signal), the *squared* amplitudes add:

$$v = (v_s^2 + v_n^2)^{\frac{1}{2}}$$

where $v$ is the rms signal obtained by adding together a signal of rms amplitude $v_s$ and a noise signal of rms amplitude $v_n$. The rms amplitudes *don't* add.

### *Signal-to-noise ratio*

Signal-to-noise ratio (SNR) is simply defined as

$$\mathrm{SNR} = 10\log_{10}\left(\frac{V_s^2}{V_n^2}\right) \quad \mathrm{dB}$$

where the voltages are rms values, and some bandwidth and center frequency are specified; i.e., it is the ratio, in decibels, of the rms voltage of the desired signal to the rms voltage of the noise that is also present. The "signal" itself may be sinusoidal or a modulated information-carrying waveform or even a noiselike signal itself. It is particularly important to specify the bandwidth if the signal has some sort of narrowband spectrum, since the SNR will drop as the bandwidth is increased beyond that of the signal: The amplifier keeps adding noise power, while the signal power remains constant.

### *Noise figure*

Any real signal source or measuring device generates noise because of Johnson noise in its source resistance (the real part of its complex source impedance). There may be additional noise, of course, from other causes. The *noise figure* (NF) of an amplifier is simply the ratio, in decibels, of the output of the real amplifier to the output of a "perfect" (noiseless) amplifier of the same gain, with a resistor of value $R_s$ connected across the amplifier's input terminals in each case. That is, the Johnson noise of $R_s$ is the "input signal."

$$\mathrm{NF} = 10\log_{10}\left(\frac{4kTR_s + v_n^2}{4kTR_s}\right)$$

$$= 10\log_{10}\left(1 + \frac{v_n^2}{4kTR_s}\right) \quad \mathrm{dB}$$

where $v_n^2$ is the mean squared noise voltage per hertz contributed by the amplifier, with a noiseless (cold) resistor of value $R_s$ connected across its input. This latter restriction is important, as you will see shortly, because the noise voltage contributed by an amplifier depends very much on the source impedance (Fig. 7.40).

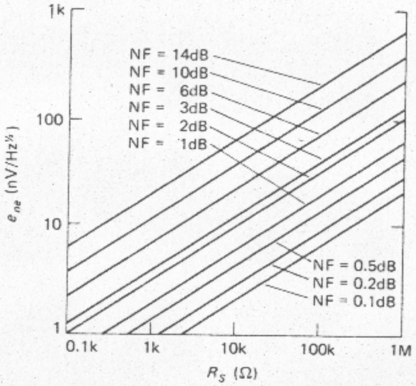


Figure 7.40. Effective noise voltage versus noise figure and source resistance. (National Semiconductor Corp.)

Noise figure is handy as a figure of merit for an amplifier when you have a signal source of a given source impedance and want to compare amplifiers (or transistors, for which NF is often specified). NF varies with frequency and source impedance, and it is often given as a set of contours of constant NF versus frequency and $R_s$. It may also be given as a set of graphs of NF versus frequency, one curve for each collector current, or a similar set of graphs of NF versus $R_s$, one for each collector current. Note: The foregoing expressions for NF assume that the amplifier's input impedance is much larger than the source impedance, i.e., $Z_{\mathrm{in}} \gg R_s$. However, in the special case of radiofrequency amplifiers, you usually have $R_s = Z_{\mathrm{in}}$ =50 ohms, with NF defined accordingly. For this special case of matched impedances, simply remove the factors "4" from the foregoing equations.

Big fallacy: Don't try to improve things by adding a resistor in series with a signal source to reach a region of minimum NF. All you're doing is making the source noisier to make the amplifier look better! Noise figure can be very deceptive for this reason. To add to the deception, the NF specification (e.g., NF = 2dB) for a transistor or FET will always be for the optimum combination of $R_s$ and $I_C$. It doesn't tell you much about actual performance, except that the manufacturer thinks the noise figure is worth bragging about.

In general, when evaluating the performance of some amplifier, you're probably least likely to get confused if you stick with SNR calculated for that source voltage and impedance. Here's how to convert from NF to SNR:

$$\mathrm{SNR} = 10\log_{10}\left(\frac{v_s^2}{4kTR_s}\right) - \mathrm{NF(dB)}\ (\text{at } R_s) \quad \mathrm{dB}$$

where $v_s$ is the rms signal amplitude, $R_s$ is the source impedance, and NF is the noise figure of the amplifier for source impedance $R_s$.

### □ *Noise temperature*

Rather than noise *figure*, you sometimes see noise *temperature* used to express the noise performance of an amplifier. Both methods give the same information, namely the excess noise contribution of the amplifier when driven by a signal source of impedance $R_s$; they are equivalent ways of expressing the same thing.

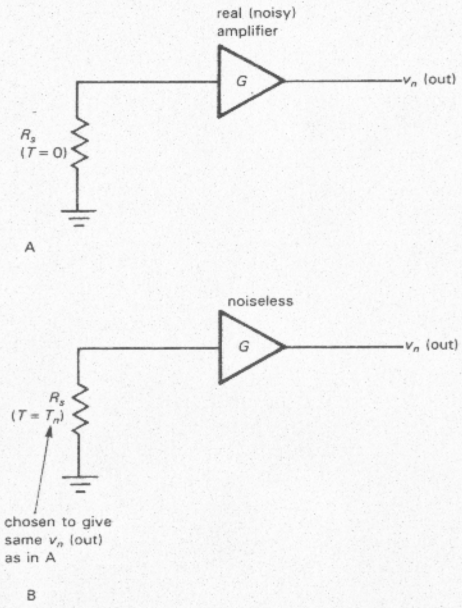


Figure 7.41

Look at Figure 7.41 to see how noise temperature works: We first imagine the actual (noisy) amplifier connected to a *noiseless* source of impedance $R_s$ (Fig. 7.41A). If you have trouble imagining a noiseless source, think of a resistor of value $R_s$ cooled to absolute zero. There will be some noise at the output, even though the source is noiseless, because the amplifier has noise. Now imagine constructing Figure 7.41B, where we magically make the amplifier noiseless, and bring the source $R_s$ up to some temperature $T_n$ such that *the output noise voltage is the same as in Figure 7.41A*. $T_n$ is called the noise temperature of the amplifier, for source impedance $R_s$.

As we remarked earlier, noise figure and noise temperature are simply different ways of conveying the same information. In fact, you can show that they are related by the following expressions:

$$T_n = T\left(10^{\mathrm{NF(dB)}/10} - 1\right)$$

$$\mathrm{NF(dB)} = 10\log_{10}\left(\frac{T_n}{T} + 1\right)$$

where $T$ is the ambient temperature, usually taken as 290°K.

Generally speaking, good low-noise amplifiers have noise temperatures far below room temperature (or, equivalently, they have noise figures far less than 3dB). Later in the chapter we will explain how you go about measuring the noise figure (or temperature) of an amplifier. First, however, we need to understand noise in transistors and the techniques of low-noise design. We hope the discussion that follows will clarify what is often a murky subject!

After reading the next two sections, we trust you won't ever be confused about noise figure again!

### 7.13 Transistor amplifier voltage and current noise

The noise generated by an amplifier is easily described by a simple noise model that is accurate enough for most purposes. In Figure 7.42, $e_n$ represents a noise voltage source in series with the input, and $i_n$ represents an input noise current. The transistor (or amplifier, in general) is assumed noiseless, and it simply amplifies the input noise voltage it sees. That is, the amplifier contributes a total noise voltage $e_a$, referred to the input, of

$$e_a(\mathrm{rms}) = [e_n^2 + (R_s i_n)^2]^{\frac{1}{2}} \quad \mathrm{V/Hz^{\frac{1}{2}}}$$

The two terms are simply the amplifier input noise voltage and the noise voltage generated by the amplifier's input noise current passing through the source resistance. Since the two noise terms are usually uncorrelated, their squared amplitudes add to produce the effective noise voltage seen by the amplifier. For low source resistances the noise voltage $e_n$ dominates, whereas for high source impedances the noise current $i_n$ generally dominates.

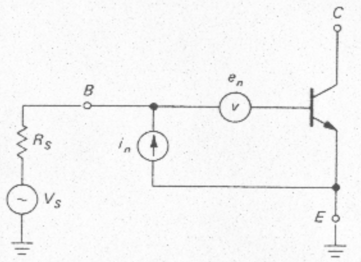

Figure 7.42. Noise model of a transistor.

Just to give an idea of what these look like, Figure 7.43 shows a graph of $e_n$ and $i_n$ versus $I_C$ and $f$, for a 2N5087. We'll go into some detail now, describing these and showing how to design for minimum noise. It is worth noting that voltage noise and current noise for a transistor are in the range of nanovolts and picoamps per root hertz ($\mathrm{Hz^{\frac{1}{2}}}$).

#### *Voltage noise, $e_n$*

The equivalent voltage noise looking in series with the base of a transistor arises from Johnson noise in the base spreading resistance, $r_{bb}$, and collector current shot noise generating a noise voltage across the intrinsic emitter resistance $r_e$. These two terms look like this:

$$e_n^2 = 4kTr_{bb} + 2qI_C r_e^2$$

$$= 4kTr_{bb} + \frac{2(kT)^2}{qI_C} \quad \mathrm{V^2/Hz}$$



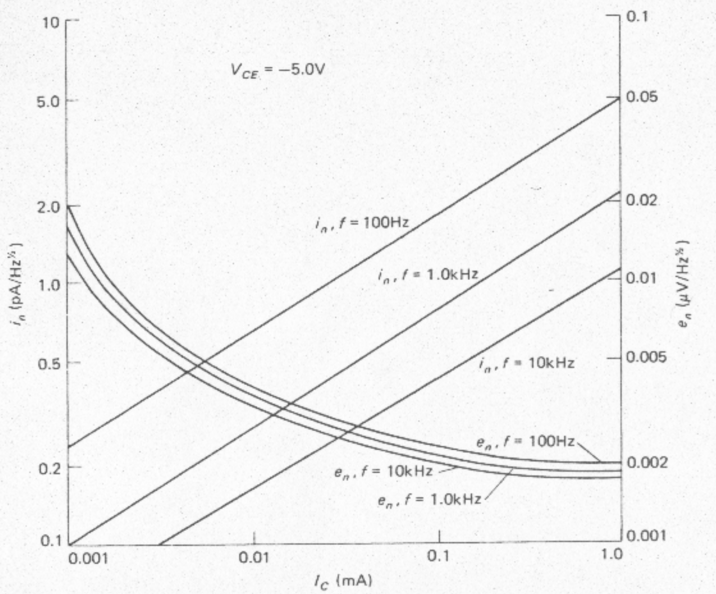

Figure 7.43. Equivalent rms input noise voltage ($e_n$) and noise current ($i_n$) versus ector current for a 2N5087 *npn* transistor. (Courtesy of Fairchild Camera and Instrument Corp.)

Both of these are Gaussian white noise. In addition, there is some flicker noise generated by base current flowing through $r_{bb}$. This last term is significant only at high base current, i.e., at high collector current. The result is that $e_n$ is constant over a wide range of collector currents, rising at low currents (shot noise through an increasing $r_e$) and at sufficiently high currents (flicker noise from $I_B$ through $r_{bb}$). This latter rise is present only at low frequencies, because of its $1/f$ character. As an example, at frequencies above 10kHz the 2N5087 has an $e_n$ of $5\mathrm{nV/Hz^{\frac{1}{2}}}$ at $I_C = 10\mu\mathrm{A}$ and $2\mathrm{nV/Hz^{\frac{1}{2}}}$ at $I_C = 100\mu\mathrm{A}$. Figure 7.44 shows graphs of $e_n$ versus frequency and current for the low-noise LM394 *npn* differential pair, and the low-noise 2SD786 from Toyo-Rohm. The latter uses special geometry to achieve an unusually low $r_{bb}$ of 4 ohms, which is needed to realize the lowest values of $e_n$.

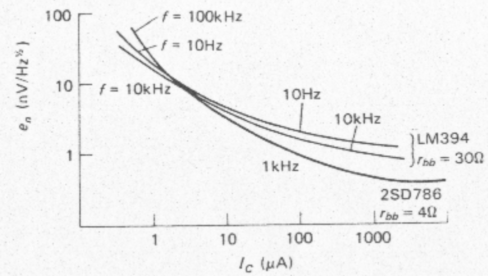

Figure 7.44. Input noise voltage ($e_n$) versus collector current for two low-noise bipolar transistors.

#### *Current noise, $i_n$*

Noise current is important, because it generates an additional noise voltage across the input signal source impedance. The main source of current noise is shot-noise fluctuation in the steady base current, added to the fluctuations caused by flicker

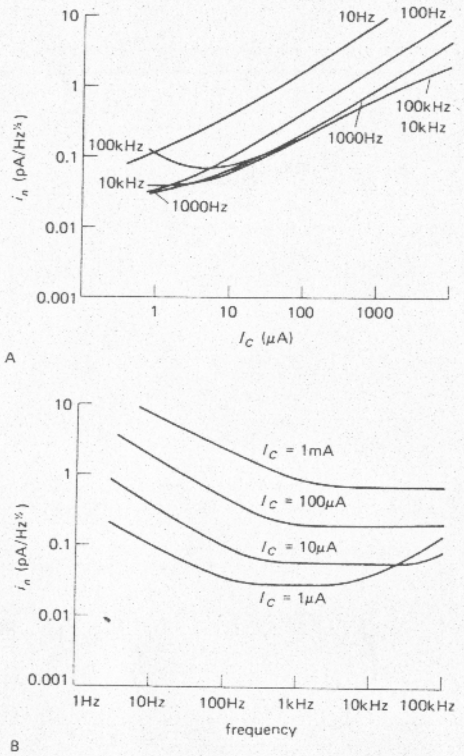


Figure 7.45. Input noise current for the LM394 bipolar transistor.
A. Noise current ($i_n$) versus collector current.
B. Noise current ($i_n$) versus frequency.

noise in $r_{bb}$. The shot-noise contribution is a noise current that increases proportional to the square root of $I_B$ (or $I_C$) and is flat with frequency, whereas the flicker-noise component rises more rapidly with $I_C$ and shows the usual $1/f$ frequency dependence. Taking the example of the 2N5087 again, above 10kHz $i_n$ is about $0.1\text{pA/Hz}^{1/2}$ at $I_C = 10\mu\text{A}$ and $0.4\text{pA/Hz}^{1/2}$ at $I_C = 100\mu\text{A}$. The noise current increases, and the noise voltage drops, as $I_C$ is increased. In the next section you will see how this dictates operating current in low-noise design.

Figure 7.45 shows graphs of $i_n$ versus frequency and current, again for the low-noise LM394.

## □ 7.14 Low-noise design with transistors

The fact that $e_n$ drops and $i_n$ rises with increasing $I_C$ provides a simple way to optimize transistor operating current to give lowest noise with a given source. Look at the model again (Fig. 7.46). The noiseless signal source $v_s$ has added to it an irreducible noise voltage from the Johnson noise of its source resistance.

$$e_R^2(\text{source}) = 4kTR_s \qquad \text{V}^2/\text{Hz}$$

The amplifier adds noise of its own, namely,

$$e_a^2(\text{amplifier}) = e_n^2 + (i_nR_s)^2 \qquad \text{V}^2/\text{Hz}$$

Thus the amplifier's noise voltage is added to the input signal, and in addition, its noise current generates a noise voltage across the source impedance. These two are uncorrelated (except at very high frequencies), so you add their squares. The idea is to reduce the amplifier's total noise contribution as much as possible. That's easy, once you know $R_s$, because you just look at a graph of $e_n$ and $i_n$ versus $I_C$, in the region of the signal frequency, picking $I_C$ to minimize $e_n^2 + (i_nR_s)^2$. Alternatively, if you are lucky and have a plot of noise-figure contours versus $I_C$ and $R_s$, you can quickly locate the optimum value of $I_C$.

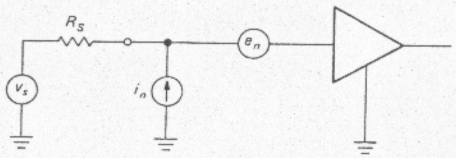


Figure 7.46. Amplifier noise model.

### □ Noise figure example

As an example, suppose we have a small signal in the region of 1kHz with source resistance of 10k, and we wish to make an amplifier with a 2N5087. From the $e_n$-$i_n$ graph (Fig. 7.47) we see that the sum of voltage and current terms (with 10k source) is minimized for a collector current of about 10–20μA. Since the current noise is dropping faster than the voltage noise is rising as $I_C$ is reduced, it might be a good idea to use slightly less collector current, especially if operation at a lower frequency is anticipated ($i_n$ rises rapidly with decreasing frequency). We can estimate the noise figure using $i_n$ and $e_n$ at 1kHz:

$$\text{NF} = 10\log_{10}\left(1 + \frac{e_n^2 + (i_nR_s)^2}{4kTR_s}\right) \text{ dB}$$

For $I_C = 10\mu\text{A}$, $e_n = 3.8\text{nV/Hz}^{1/2}$, $i_n = 0.29\text{pA/Hz}^{1/2}$, and $4kTR_s = 1.65 \times 10^{-16}\text{V}^2/\text{Hz}$ for the 10k source resistance. The calculated noise figure is therefore 0.6dB. This is consistent with the graph (Fig. 7.48) showing NF versus frequency, in which they have chosen $I_C = 20\mu\text{A}$ for $R_s = 10\text{k}$. This choice of collector current is also roughly what you would get from the graph in Figure 7.47 of noise-figure contours at 1kHz, although the actual noise figure can be estimated only approximately from that plot as being less than 2dB.



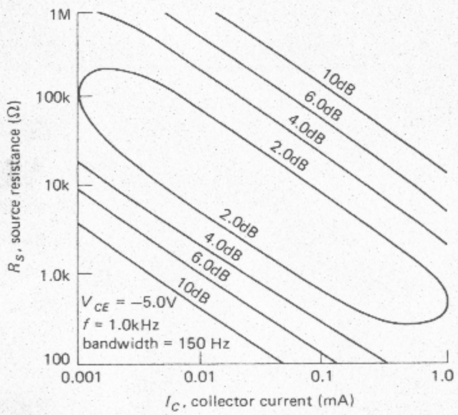


Figure 7.47. Contours of constant narrowband noise figure for the 2N5087 transistor. (Courtesy of Fairchild Camera and Instrument Corp.)

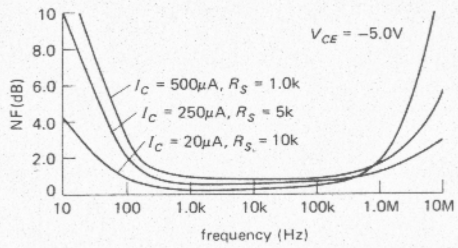


Figure 7.48. Noise figure (NF) versus frequency, for three choices of $I_C$ and $R_S$, for the 2N5087. (Courtesy of Fairchild Camera and Instrument Corp.)

EXERCISE 7.5

Find the optimum $I_C$ and corresponding noise figure for $R_s = 100\text{k}$ and $f = 1\text{kHz}$, using the graph in Figure 7.43 of $e_n$ and $i_n$. Check your answer from the noise-figure contours (Fig. 7.47).

For the other amplifier configurations (follower, grounded base) the noise figure is essentially the same, for given $R_s$ and $I_C$, since $e_n$ and $i_n$ are unchanged. Of course, a stage with unity voltage gain (a follower) may just pass the problem along to the next stage, since the signal level hasn't been increased to the point that low-noise design can be ignored in subsequent stages.

### □ Charting amplifier noise with $e_n$ and $i_n$

The noise calculations just presented, although straightforward, make the whole subject of amplifier design appear somewhat formidable. If you misplace a factor of Boltzmann's constant, you suddenly get an amplifier with 10,000dB noise figure! In this section we will present a simplified noise-estimation technique of great utility.

The method consists of first choosing some frequency of interest in order to get

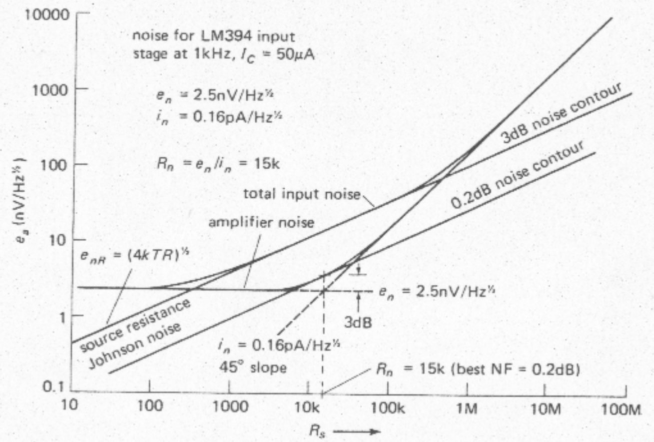


Figure 7.49. Total amplifier input voltage noise ($e_a$) plotted from the $e_n$ and $i_n$ parameters.

values for $e_n$ and $i_n$ versus $I_C$ from the transistor data sheets. Then, for a given collector current, you can plot the total noise contributions from $e_n$ and $i_n$ as a graph of $e_a$ versus source resistance $R_s$. Figure 7.49 shows what that looks like at 1kHz for a differential input stage using an LM394 matched superbeta transistor running 50μA of collector current. The $e_n$ noise voltage is constant, and the $i_nR_s$ voltage increases proportional to $R_s$, i.e., with a 45° slope. The amplifier noise curve is drawn as shown, with care being taken to ensure that it passes through a point 3dB (voltage ratio of 1.4) above the crossing point of individual voltage and current noise contributions. Also plotted is the noise voltage of the source resistance, which also happens to be the 3dB NF contour. The other lines of constant noise figure are simply straight lines parallel to this line, as you will see in the examples that follow.

The best noise figure (0.2dB) at this collector current and frequency occurs for a source resistance of 15k, and the noise figure is easily seen to be less than 3dB for all source resistances between 300 ohms and 500k, the points at which the 3dB NF contour intersects the amplifier noise curve.

The next step is to draw a few of these noise curves on the same graph, using different collector currents or frequencies, or maybe a selection of transistor types, in order to evaluate amplifier performance. Before we go on to do that, let's show how we can talk about this same amplifier using a different pair of noise parameters, the noise resistance $R_n$ and the noise figure NF($R_N$), both of which pop right out of the graph.

### □ *Noise resistance*

The lowest noise figure in this example occurs for a source resistance $R_s$ = 15k, which equals the ratio of $e_n$ to $i_n$. That defines the noise resistance

$$R_n = \frac{e_n}{i_n}$$

You can find the noise figure for a source of that resistance from our earlier expression for noise figure. It is



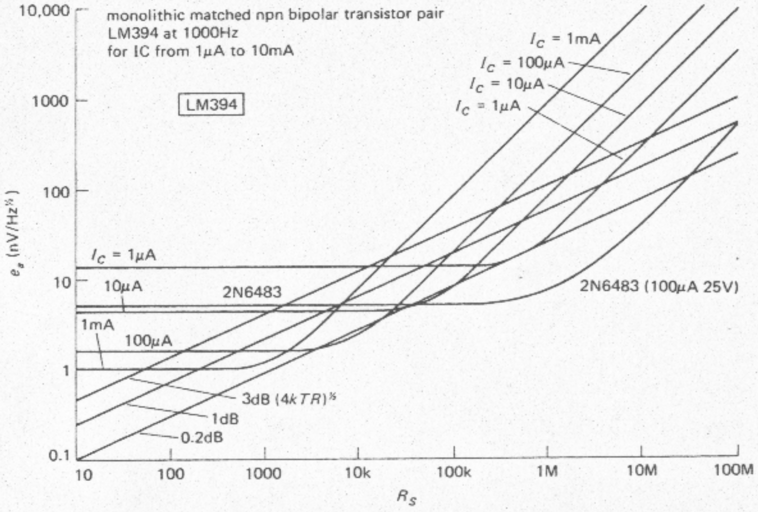


Figure 7.50. Total amplifier input voltage noise ($e_a$) for the LM394 bipolar transistor under various conditions, compared with the 2N6483 JFET.

NF (at $R_n$) =

$$10 \log_{10}\left(1 + 1.23 \times 10^{20}\frac{e_n^2}{R_n}\right) \quad \text{dB}$$

$$\approx 0.2 \quad \text{dB}$$

Noise resistance isn't actually a real resistance in the transistor, or anything like that. It is a tool to help you quickly find the value of source resistance for minimum noise figure, ideally so that you can vary the collector current to shift $R_n$ close to the value of source resistance you're actually using. $R_n$ corresponds to the point where the $e_n$ and $i_n$ lines cross.

The noise figure for a source resistance equal to $R_n$ then follows simply from the preceding equation.

### □ *Charting the bipolar/FET shootout*

Let's have some fun with this technique. A perennial bone of contention among engineers is whether FETs or bipolar transistors are "better." We will dispose of this issue with characteristic humility by matching two of the best contenders and letting them deliver their best punches. In the interest of fairness, we'll let National Semiconductor intramural teams compete, choosing two game fighters.

In the bipolar corner we have the magnificent LM394 superbeta monolithic matched pair, already warmed up, as described earlier. We'll run it at 1kHz, with collector currents from 1μA to 1mA (Fig. 7.50).

The FET entry is the 2N6483 monolithic $n$-channel JFET matched pair, known far and wide for its stunning low-noise performance, reputed to exceed that of bipolar transistors. According to its data sheet, it was trained only for 100μA and 400μA drain currents (Fig. 7.51).

And the winner? Well, it's a split decision. The FET won points on lowest minimum noise figure, NF($R_n$), reaching a phenomenal 0.05dB noise figure, and dipping well below 0.2dB from 100k to 100M source impedance. For high source

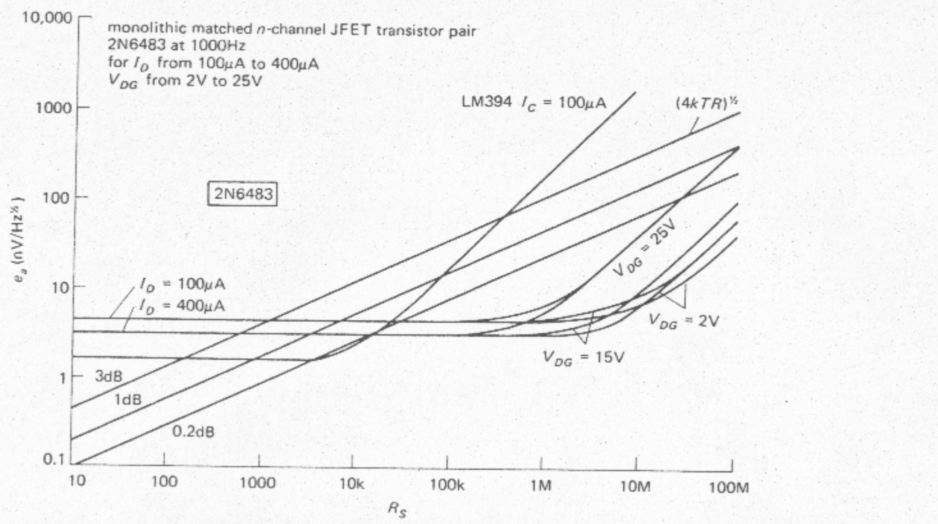


Figure 7.51. Total amplifier input voltage noise ($e_a$) for the 2N6483 JFET compared with the LM394 bipolar transistor.

impedances, FETs remain unbeaten. The bipolar transistor is best at low source impedances, particularly below 5k, and it can reach a 0.3dB noise figure at $R_s = 1\text{k}$, with suitable choice of collector current. By comparison, the FET cannot do better than 2dB with a 1k source resistance, owing to larger voltage noise $e_n$.

Just as in boxing, where the best fighters haven't yet had a chance to compete in a world championship, there are some younger contenders for the best low-noise transistor. For example, the 2SJ72 and 2SK147 complementary JFETs from Toshiba use a meshed-gate geometry to achieve a phenomenal $e_n$ of $0.7\text{nV}/\sqrt{\text{Hz}}$ at $I_D = 10\text{mA}$ (equivalent to Johnson noise from a 30Ω resistor!). But these are *JFETs*, with their low input current (hence low $i_n$), and thus the noise resistance is about 10k. When used as an amplifier with a source impedance equal to their noise resistance (i.e., $R_s = 10\text{k}$), their performance is unbeatable – the noise temperature is just 2°K!

Before you go out and buy a bushel of these remarkable JFETs, consider the remarks of the critics, who claim they are muscle-bound – they have high input and feedback capacitance (85pF and 15pF, respectively), which limits their usefulness at high frequencies. Their relative, the 2SK117, is better in this regard, at the expense of higher $e_n$. These same critics argue that the Toyo-Rohm bipolar complementary pair, the 2SD786 and 2SB737, with $e_n$ as low as $0.55\text{nV}/\sqrt{\text{Hz}}$, can offer even better performance at moderate source impedances and frequencies.

### □ Low source impedances

Bipolar transistor amplifiers can provide very good noise performance over the range of source impedances from about 200 ohms to 1M; corresponding optimum collector currents are generally in the range of several milliamps down to a microamp. That is, collector currents used for the input stage of low-noise amplifiers generally



tend to be lower than in amplifier stages not optimized for low-noise performance.

For very low source impedances (say 50Ω), transistor voltage noise will always dominate, and noise figures will be poor. The best approach in such cases is to use a transformer to raise the signal level (and impedance), treating the signal on the secondary as before. High-quality signal transformers are available from companies such as James and Princeton Applied Research. As an example, the latter's model 116 FET preamp has voltage and current noise such that the lowest noise figure occurs for signals of source impedance around 1M. A signal around 1kHz with source impedance of 100 ohms would be a poor match for this amplifier, since the amplifier's voltage noise is much larger than the signal source's Johnson noise; the resultant noise figure for that signal connected directly to the amplifier would be 11dB. By using the optional internal step-up transformer, the signal level is raised (along with its source impedance), thus overriding amplifier noise voltage and giving a noise figure of about 1.0dB.

At radiofrequencies (e.g., beginning around 100kHz) it is extremely easy to make good transformers, both for tuned (narrowband) and broadband signals. At these frequencies it is possible to make broadband "transmission-line transformers" of very good performance. We will treat some of these methods in Chapter 13. It is at the very low frequencies (audio and below) that transformers become problematic.

Three comments: (a) The voltage rises proportional to the turns ratio of the transformer, whereas the impedance rises proportional to the square of the ratio. Thus a 2:1 voltage step-up transformer has an output impedance four times the input impedance (this is mandated by conservation of energy). (b) Transformers aren't perfect. They have trouble at low frequencies (magnetic saturation) and at high frequencies (winding inductance and capacitance), as well as losses from the magnetic properties of the core and from winding resistance. The latter is a source of Johnson noise, as well. Nevertheless, when dealing with a signal of very low source impedance, you may have no choice, and transformer coupling can be very beneficial, as the preceding example demonstrates. Exotic techniques such as cooled transformers, superconducting transformers, and SQUIDs (superconducting quantum interference devices) can provide good noise performance at low impedance and voltage levels. With SQUIDs you can measure voltages of $10^{-15}$ volt! (c) Again, a warning: Don't attempt to improve performance by adding a resistor in series with a low source impedance. If you do that, you're just another victim of the noise-figure fallacy.

### □ High source impedances

If the source impedance is high, say greater than 100k or so, transistor current noise dominates, and the best device for low-noise amplification is a FET. Although their voltage noise is usually greater than that of bipolar transistors, the gate current (and its noise) can be exceedingly small, making them ideally suited for low-noise high-impedance amplifiers. Incidentally, it is sometimes useful to think of Johnson noise as a current noise $i_n = v_n/R_s$. This lets you compare source noise contributions with amplifier current noise (Fig. 7.52).

## 7.15 FET noise

We can use the same amplifier noise model for FETs, namely a series noise voltage source and a parallel noise current source. You can analyze the noise performance with exactly the same methods used for bipolar transistors. For example, see the graphs in the section on bipolar/FET shootout.

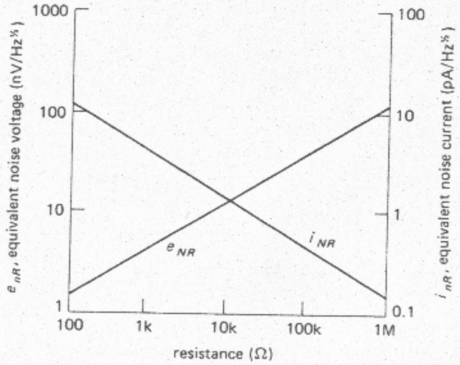


Figure 7.52. Thermal noise voltage density versus resistance at 25° C. The equivalent short-circuit current noise density is also shown.

***Voltage noise of JFETs***

For JFETs the voltage noise $e_n$ is essentially the Johnson noise of the channel resistance, given approximately by

$$e_n^2 = 4kT\left(\frac{2}{3}\frac{1}{g_m}\right) \qquad \text{V}^2/\text{Hz}$$

where the inverse transconductance takes the place of resistance in the Johnson-noise formula. Since the transconductance rises with increasing drain current (as $\sqrt{I_D}$), it is generally best to operate FETs at high drain current for lowest voltage noise. However, since the $e_n$ is Johnson noise, which goes only as $1/\sqrt{g_m}$, and that in turn goes as $\sqrt{I_D}$, $e_n$ is finally proportional to $I_D^{-1/4}$. With such a mild dependence of $e_n$ on $I_D$ it doesn't pay to run at a drain current so high that other properties of the amplifier are degraded. In particular, a FET running at high current gets hot, which (a) decreases $g_m$, (b) increases offset voltage drift and CMRR, and (c) raises gate leakage dramatically; the latter effect can actually *increase* voltage noise, since there is some contribution to $e_n$ from flicker noise associated with the gate leakage current.

There is another way to increase $g_m$, and therefore decrease JFET voltage noise: By paralleling a pair of JFETs you get twice the $g_m$, but of course this is at twice the $I_D$. But now if you run the combination at the previous $I_D$, you still improve $g_m$ by a factor of $\sqrt{2}$ over the single-JFET value, without increasing total drain current. In practice you can simply parallel a number of matched JFETs, or look for a large-geometry JFET like the 2SJ72 and 2SK147 mentioned earlier.

There is a price to pay, however. All the capacitances scale with the number of paralleled JFETs. As a result, high-frequency performance (including noise figure) is degraded. In practice you should stop paralleling additional transistors when the circuit's input capacitance roughly matches the source's capacitance. If you care about performance at high frequencies, choose JFETs with high $g_m$ and low $C_{rss}$; you might consider the ratio $g_m/C_{rss}$ a high-frequency figure of merit. Note that circuit configurations can also play an important role; e.g., the cascode circuit can be used to eliminate the Miller effect (gain multiplication) on $C_{rss}$.

MOSFETs tend to have much higher voltage noise than JFETs, with $1/f$ noise predominating, since the $1/f$ knee is as high as 10kHz to 100kHz. For this reason you wouldn't normally choose a MOSFET for low-noise amplifiers below 1MHz.

***Current noise of JFETs***

At low frequencies the current noise $i_n$ is extremely small, arising from the shot noise in the gate leakage current (Fig. 7.53):

$$i_n = (3.2 \times 10^{-19} I_G B)^{\frac{1}{2}} \qquad \text{A(rms)}$$

In addition, there is a flicker-noise component in some FETs. The noise current rises with increasing temperature, as the gate leakage current rises. Watch out for the rapidly increasing gate leakage in $n$-channel JFETs that occurs for operation at high $V_{DG}$ (see Section 3.09).



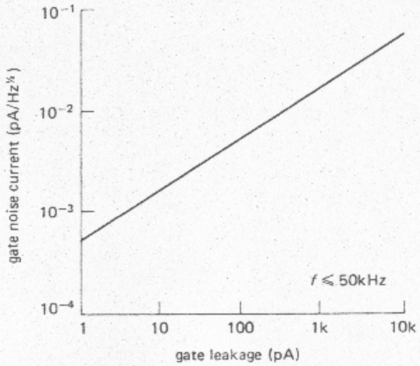


Figure 7.53. Input noise current versus gate leakage current for JFETs. (Courtesy of National Semiconductor Corp.)

At moderate to high frequencies there is an additional noise term, namely the real part of the input impedance seen looking into the gate. This comes from the effect of feedback capacitance (Miller effect) when there is a phase shift at the output due to load capacitance; i.e., the part of the output signal that is shifted 90° couples through the feedback capacitance $C_{rss}$ to produce an effective resistance at the input, given by

$$R = \frac{1 + \omega C_L R_L}{\omega^2 g_m C_{rss} C_L R_L^2} \qquad \text{ohms}$$

As an example, the 2N5266 $p$-channel JFET has a noise current of 0.005pA/Hz$^{\frac{1}{2}}$ and a noise voltage $e_n$ of 12nV/Hz$^{\frac{1}{2}}$, both at $I_{DSS}$ and 10kHz. The noise current begins climbing at about 50kHz. These figures are roughly 100 times better in $i_n$ and 5 times worse in $e_n$ than the corresponding figures for the 2N5087 used earlier.

With FETs you can achieve good noise performance for input impedances in the range of 10k to 100M. The PAR model 116 preamp has a noise figure of 1dB or better for source impedances from 5k to 10M in the frequency range from 1kHz to 10kHz. Its performance at moderate frequencies corresponds to a noise voltage of 4nV/Hz$^{\frac{1}{2}}$ and a noise current of 0.013pA/Hz$^{\frac{1}{2}}$.

### 7.16 Selecting low-noise transistors

As we mentioned earlier, bipolar transistors offer the best noise performance with low source impedances, owing to their lower input voltage noise. Voltage noise, $e_n$, is reduced by choosing a transistor with low base spreading resistance, $r_{bb}$, and operating at high collector current (as long as $h_{FE}$ remains high). For higher source impedances the current noise can be minimized instead by operating at lower collector current.

At high values of source impedance, FETs are the best choices. Their voltage noise can be reduced by operating at higher drain currents, where the transconductance is highest. FETs intended for low-noise applications have high k values (see Section 3.04), which usually means high input capacitance. For example, the low-noise 2N6483 has $C_{iss}$ = 20pF, whereas the 2N5902 low-current FET has $C_{iss}$ = 2pF.

Figures 7.54 and 7.55 show comparisons of the noise characteristics of a number of popular and useful transistors.

### □ 7.17 Noise in differential and feedback amplifiers

Low-noise amplifiers are often differential, to obtain the usual benefits of low drift and good common-mode rejection. When you calculate the noise performance of a differential amplifier, there are three points to keep in mind: (a) Be sure to use the individual collector currents, not the sum, to get $e_n$ and $i_n$ from data sheets. (b) The $i_n$ seen at each input terminal is the same as for a single-ended amplifier configuration. (c) The $e_n$ seen at one input, with the other input grounded, say, is 3dB

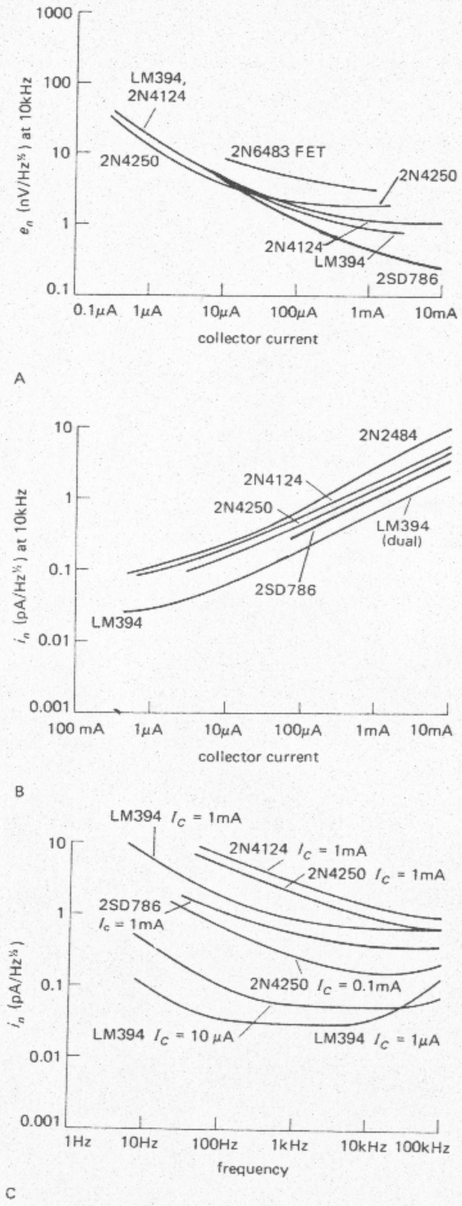


Figure 7.54. Input noise for some popular transistors.
A. Input noise voltage ($e_n$) versus collector current.
B. Input noise current ($i_n$) versus collector current.
C. Input noise current ($i_n$) versus frequency.

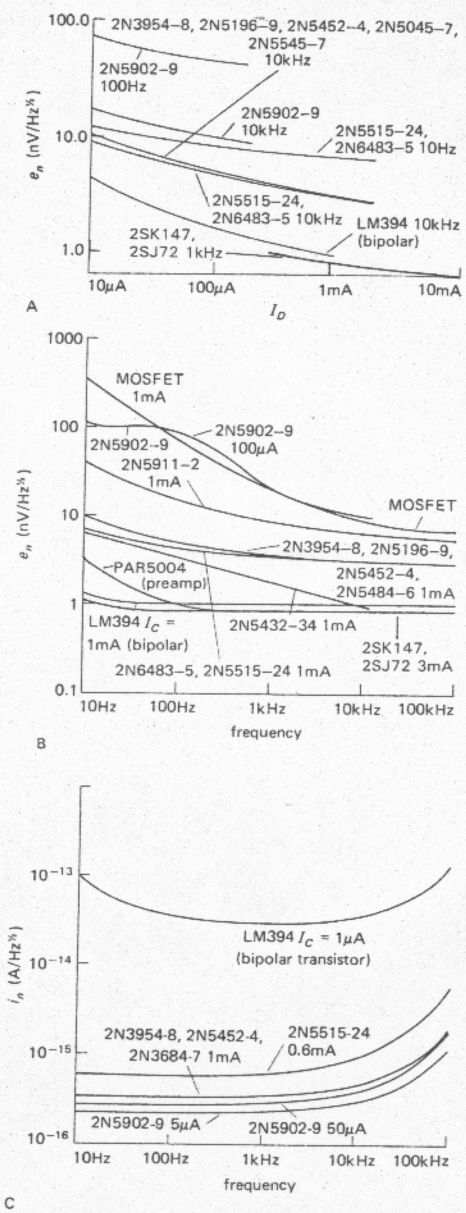


Figure 7.55. Input noise for some popular FETs.
A. Input noise voltage ($e_n$) versus drain current ($I_D$).
B. Input noise voltage ($e_n$) versus frequency.
C. Input noise current ($i_n$) versus frequency.





larger than the single-transistor case, i.e., it is multiplied by $\sqrt{2}$.

In amplifiers with feedback, you want to take the equivalent noise sources $e_n$ and $i_n$ out of the feedback loop, so you can use them as previously described when calculating noise performance with a given signal source. Let's call the noise terms brought out of the feedback loop $e_A$ and $i_A$, for *amplifier* noise terms. Thus the amplifier's noise contribution to a signal with source resistance $R_s$ is

$$e^2 = e_A^2 + (R_s I_A)^2 \quad \text{V}^2/\text{Hz}$$

Let's take the two feedback configurations separately.

### □ Noninverting

For the noninverting amplifier (Fig. 7.56) the input noise sources become

$$i_A^2 = i_n^2$$
$$e_A^2 = e_n^2 + 4kTR_\parallel + (i_n R_\parallel)^2$$

where $e_n$ is the "adjusted" noise voltage for the differential configuration, i.e., 3dB larger than for a single-transistor stage. The additional noise voltage terms arise from Johnson noise and input-stage noise current in the feedback resistors. Note that the effective noise voltage and current are now not completely uncorrelated, so calculations in which their squares are added can be in error by a maximum factor of 1.4.

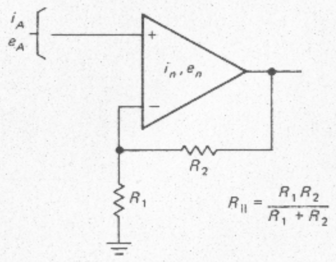


Figure 7.56

For a follower, $R_2$ is zero, and the effective noise sources are just those of the differential amplifier alone.

### □ Inverting

For the inverting amplifier (Fig. 7.57) the input noise sources become

$$i_A^2 = i_n^2 + 4kT\frac{1}{R_2}$$
$$e_A^2 = e_n^2 + R_1^2\left(i_n^2 + 4kT\frac{1}{R_2}\right)$$
$$= e_n^2 + R_1^2 i_A^2$$

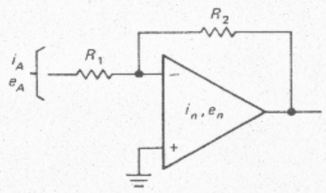


Figure 7.57

### Op-amp selection curves

You now have all the tools necessary to analyze op-amp input circuits. Their noise is specified in terms of $e_n$ and $i_n$, just as with transistors and FETs. You don't get to adjust anything, though; you only get to use them. The data sheets may need to be taken with a grain of salt. For example, "popcorn noise" is typified by jumps in offset at random times and duration. It is rarely mentioned in polite company. Figure 7.58 summarizes the noise performance of some popular op-amps.

### Wideband noise

Op-amp circuits are generally dc-coupled and extend to some upper frequency limit $f_{\text{cutoff}}$. Therefore it is of interest to know the total noise voltage over this band, not merely the noise power density. Figure 7.59 presents some graphs showing the rms noise voltage in a band extending

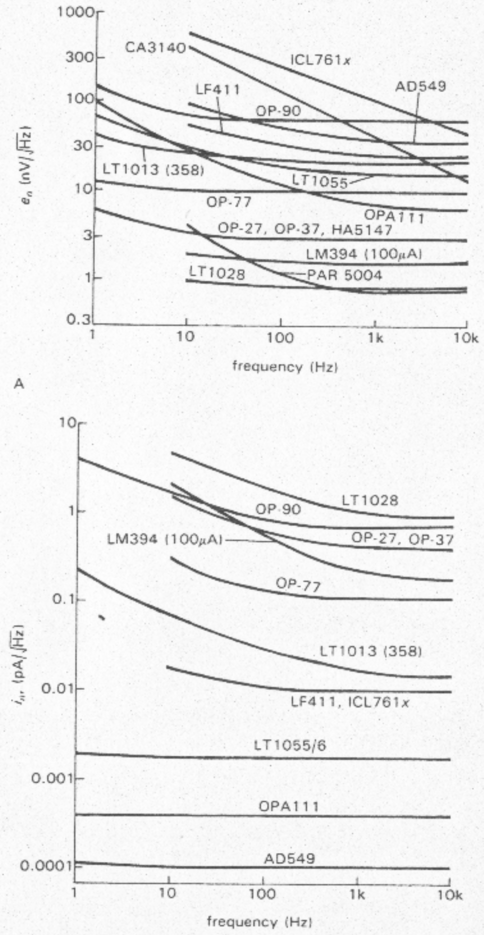


Figure 7.58. Input noise for some popular op-amps.
A. Input noise voltage ($e_n$) versus frequency.
B. Input noise current ($i_n$) versus frequency.

from dc to the indicated frequency; they were calculated by integrating the noise power curves for the various op-amps.

### *Choosing a low-noise op-amp*

It is simple to choose an op-amp to minimize noise in some frequency range, given the signal impedance seen from the op-amp, $R_{sig}$ (which includes the effects of feedback components, as given in the foregoing expressions). Generally speaking, you want op-amps with low $i_n$ for high signal impedances, and op-amps with low $e_n$ for low signal impedances. Assuming the signal source is at room temperature, the total input-referred squared noise voltage density is just

$$e_A^2 = 4kTR_{sig} + {e_n}^2 + (i_n R_{sig})^2$$

where the first term is due to Johnson noise, and the last two terms are due to op-amp noise voltage and current. Obviously the Johnson noise sets a lower bound to the input-referred noise. In Figure 7.60 we've plotted the quantity $e_A$ (at 10Hz) as a function of $R_{sig}$ for the quietest op-amps we could find. For comparison we included our jellybean JFET LF411 and the micropower bipolar OP-90. The latter, although an excellent micropower op-amp, has high noise voltage (because the front end operates at low collector current, hence high $r_e$ and therefore high Johnson noise) and also high noise current (because the bipolar input has substantial base current); it shows just how good the premium low-noise op-amps really are.

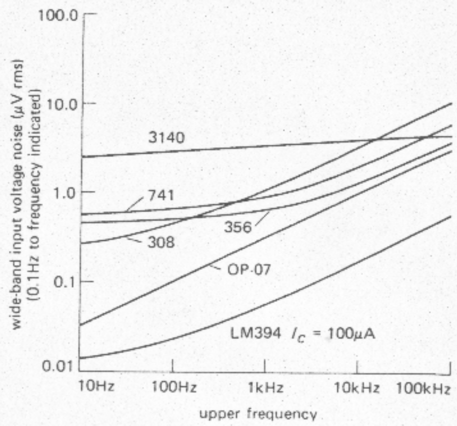


Figure 7.59. Wideband noise voltage for some popular op-amps.



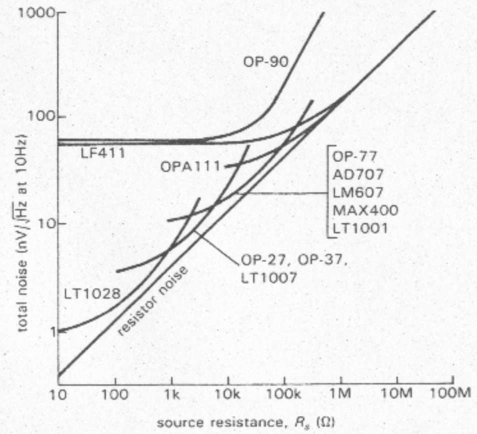


Figure 7.60. Total noise (source resistor plus amplifier, at 10Hz) for high-performance op-amps.

### □ *Low-noise preamps*

In addition to the low-noise op-amps, there are some nice low-noise IC *preamplifiers*. Unlike op-amps, these generally have fixed voltage gain, though in some models you can attach an external gain-setting resistor. People sometimes call these "video amplifiers" because they often have bandwidths into the tens of megahertz, though they can be used for low-frequency applications as well. Examples are the Plessey SL561B and several models from Analog Systems. These amplifiers typically have $e_n$ less than $1\text{nV}/\sqrt{\text{Hz}}$, achieved (at the expense of high input noise *current*, $i_n$) by running the input transistor at relatively high collector current.

## □ NOISE MEASUREMENTS AND NOISE SOURCES

It is a relatively straightforward process to determine the equivalent noise voltage and current of an amplifier, and from these the noise figure and signal-to-noise ratio for any given signal source. That's all you ever need to know about the noise performance of an amplifier. Basically the process consists of putting known noise signals across the input, then measuring the output noise signal amplitudes within a certain bandwidth. In some cases (e.g., a matched input impedance device such as a radio-frequency amplifier) an oscillator of accurately known and controllable amplitude is substituted as the input signal source.

Later we will discuss the techniques you need to do the output voltage measurement and bandwidth limiting. For now, let's assume you can make rms measurements of the output signal, with a measurement bandwidth of your choice.

### □ 7.18 Measurement without a noise source

For an amplifier stage made from a FET or transistor and intended for use at low to moderate frequencies, the input impedance is likely to be very high. You want to know $e_n$ and $i_n$ so that you can predict the SNR with a signal source of arbitrary source impedance and signal level, as we discussed earlier. The procedure is simple:

First, determine the amplifier's voltage gain $G_V$ by actual measurement with a signal in the frequency range of interest. The amplitude should be large enough to override amplifier noise, but not so large as to cause amplifier saturation.

Second, short the input and measure the rms noise output voltage, $e_s$. From this you get the input noise voltage per root hertz from

$$e_n = \frac{e_s}{G_V B^{\frac{1}{2}}} \qquad \text{V/Hz}^{\frac{1}{2}}$$

where $B$ is the bandwidth of the measurement (see Section 7.21).

Third, put a resistor $R$ across the input, and measure the new rms noise output voltage, $e_r$. The resistor value should be large enough to add significant amounts of current noise, but not so large that the input impedance of the amplifier begins to dominate. (If this is impractical, you can leave the input open and use the amplifier's input impedance as $R$.) The output you measure is just

$$e_r^2 = [e_n^2 + 4kTR + (i_nR)^2]BG_V^2$$

from which you can determine $i_n$ to be

$$i_n = \frac{1}{R_s}\left[\frac{e_r^2}{BG_V^2} - (e_n^2 + 4kTR)\right]^{\frac{1}{2}}$$

With some luck, only the first term in the square root will matter (i.e., if current noise dominates both amplifier voltage noise and source resistor Johnson noise).

Now you can determine the SNR for a signal $V_s$ of source impedance $R_s$, namely

$$\text{SNR} = 10\log_{10}\left(\frac{V_s^2}{V_n^2}\right)$$

$$= 10\log_{10}\left[\frac{V_s^2}{[e_n^2 + (i_nR_s)^2 + 4kTR_s]B}\right]$$

where the numerator is the signal voltage (presumed to lie within the bandwidth $B$) and the terms in the denominator are the amplifier noise voltage, amplifier noise current applied to $R_s$, and Johnson noise in $R_s$. Note that increasing the amplifier bandwidth beyond what is necessary to pass the signal $V_s$ only decreases the final SNR. However, if $V_s$ is broadband (e.g., a noise signal itself), the final SNR is independent of amplifier bandwidth. In many cases the noise will be dominated by one of the terms in the preceding equation.

## □ 7.19 Measurement with noise source

The preceding technique of measuring the noise performance of an amplifier has the advantage that you don't need an accurate and adjustable noise source, but it requires an accurate voltmeter and filter, and it assumes that you know the gain versus frequency of the amplifier, with the actual source resistance applied. An alternative method of noise measurement involves applying broadband noise signals of known amplitude to the amplifier's input and observing the relative increase of output noise voltage. Although this technique requires an accurately calibrated noise source, it makes no assumptions about the properties of the amplifier, since it measures the noise properties right at the point of interest, at the input.

Again, it is relatively straightforward to make the requisite measurements. You connect the noise generator to the amplifier's input, making sure that its source impedance $R_g$ equals the source impedance of the signal you ultimately plan to use with the amplifier. You first note the amplifier's output rms noise voltage, with the noise source attenuated to zero output signal. Then you increase the noise source rms amplitude $V_g$ until the amplifier's output rises 3dB (a factor of 1.414 in rms output voltage). The amplifier's input noise voltage in the measurement bandwidth, for this source impedance, equals this value of added signal. The amplifier therefore has a noise figure

$$\text{NF} = 10\log_{10}\left(\frac{V_g^2}{4kTR_g}\right)$$

From this you can figure out the SNR for a signal of any amplitude with this same source impedance, using the formula from Section 7.12

$$\text{SNR} = 10\log_{10}\left(\frac{V_s^2}{4kTR_s}\right) - \text{NF}(R_s)\ \text{dB}$$

There are nice calibrated noise sources available, most of which provide means for attenuation to precise levels in the microvolt range. Note: Once again, the preceding formulas assume $R_{\text{in}} \gg R_s$. If, on the other hand, the noise-figure measurement is made with a *matched* signal source, i.e., if $R_s = Z_{\text{in}}$, then omit the factors "4" in the preceding expressions.

Note that this technique does not tell you $e_n$ and $i_n$ directly, just the appropriate combination for a source of impedance equal to the driving impedance you used in the measurement. Of course, by making several such measurements with different noise source impedances, you could infer the values of $e_n$ and $i_n$.

A nice variation on this technique is to use resistor Johnson noise as the "noise source." This is a favorite technique used by designers of very low noise radiofrequency amplifiers (in which, incidentally, the signal source impedance is usually 50Ω and matches the amplifier's input impedance). It is usually done the following way: A dewar of liquid nitrogen holds a 50 ohm "termination" (a fancy name for a well-designed resistor that has negligible inductance or capacitance) at the temperature of boiling nitrogen, 77°K; a second 50 ohm termination is kept at room temperature. The amplifier's input is connected alternately to the two resistors (usually with a high-quality coax relay), while the output noise power (at some center frequency, with some measurement bandwidth) is measured with an RF power meter. Call the results of the two measurements $P_C$ and $P_H$, the output noise power corresponding to cold and hot source resistors, respectively. It is then easy to show that the amplifier's noise temperature, at the frequency of the measurement, is just

$$T_n = \frac{T_H - YT_C}{Y - 1}$$

where $Y = P_H/P_C$, the ratio of noise powers. Noise figure is then given by the formula of Section 7.12, namely

$$\text{NF(dB)} = 10\log_{10}\left(\frac{T_n}{290} + 1\right)$$

EXERCISE 7.6

Derive the foregoing expression for noise temperature. Hint: Begin by noting that $P_H = \alpha(T_n + T_H)$ and $P_C = \alpha(T_n + T_C)$, where $\alpha$ is a constant that will shortly disappear. Then note that the noise contribution of the amplifier, stated as a noise temperature, *adds* to the noise temperature of the source resistor. Take it from there.

EXERCISE 7.7

Amplifier noise temperature (or noise figure) depends on the value of signal source impedance, $R_s$. Show that an amplifier characterized by $e_n$ and $i_n$ (as in Fig. 7.46) has minimum noise temperature for a source impedance $R_s = e_n/i_n$. Then show that the noise temperature, for that value of $R_s$, is given by $T_n = e_ni_n/2k$.

### □ *Amplifiers with matched input impedance*

This last technique is ideal for noise measurements of amplifiers designed for a matched signal source impedance. The most common examples are in radiofrequency amplifiers or receivers, usually meant to be driven with a signal source impedance of 50 ohms, and which themselves have an input impedance of 50 ohms. We will discuss in Chapter 13 the reasons for this departure from our usual criterion that a signal source should have a small source impedance compared with the load it drives. In this situation $e_n$ and $i_n$ are irrelevant as separate quantities; what matters is the overall noise figure (with matched source) or some specification of SNR with a matched signal source of specified amplitude.

Sometimes the noise performance is explicitly stated in terms of the *narrowband* input signal amplitude required to obtain a certain output SNR. A typical radiofrequency receiver might specify a 10dB SNR with a 0.25μV rms input signal and 2kHz receiver bandwidth. In this case the procedure consists of measuring the rms receiver output with the input driven by a

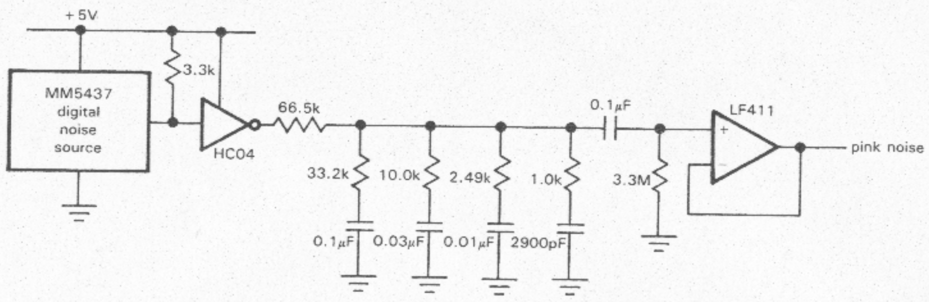


Figure 7.61. Pink noise source (−3dB/octave, ±0.25dB from 10Hz to 40kHz).

matched sine-wave source initially attenuated to zero, then increasing the (sine-wave) input signal until the rms output rises 10dB, in both cases with the receiver bandwidth set to 2kHz. It is important to use a meter that reads true rms voltages for a measurement where noise and signal are combined (more about this later). Note that radiofrequency noise measurements often involve output signals that are in the audiofrequency range.

## □ 7.20 Noise and signal sources

Broadband noise can be generated from the effects we discussed earlier, namely Johnson noise and shot noise. The shot noise in a vacuum diode is a classic source of broadband noise that is especially useful because the noise voltage can be predicted exactly. More recently, zener diode noise has been used in noise sources. Both of these extend from dc to very high frequencies, making them useful in audiofrequency and radiofrequency measurements.

An interesting noise source can be made using digital techniques, in particular by connecting long shift registers with their input derived from a modulo-2 addition of several of the later bits (see Section 9.33). The resultant output is a pseudorandom sequence of 1's and 0's that after low-pass filtering generates an analog signal of white spectrum up to the low-pass filter's break-point, which must be well below the frequency at which the register is shifted. These things can be run at very high frequencies, generating noise up to 100kHz or more. The "noise" has the interesting property that it repeats itself exactly after a time interval that depends on the register length (an $n$-bit maximal-length register goes through $2^n - 1$ states before repeating). Without much difficulty that time can be made to be very long (months or years), although most often a period of a second is long enough. For example, a 50-bit register shifted at 10MHz will generate white noise up to 100kHz or so, with a repeat time of 3.6 years. A design for a pseudorandom noise source based on this technique is shown in Section 9.36.

Some noise sources can generate pink noise as well as white noise. Pink noise has equal noise power per *octave*, rather than equal power per hertz. Its power density (power per hertz) drops off at 3dB/octave. Since an $RC$ filter drops off at 6dB/octave, a more complicated filter is necessary to generate a pink spectrum from a white noise input. The circuit shown in Figure 7.61 uses a 23-bit pseudorandom white noise generator chip to generate pink noise, accurate to ±0.25dB from 10Hz to 40kHz.

Versatile signal sources are available with precisely controlled output amplitude (down to the microvolt range and below) over frequencies from a fraction of a hertz to gigahertz. Some can even be programmed via a digital "bus." An example is the Hewlett-Packard model 8660 synthesized signal generator, with output frequencies from 0.01 to 110MHz, calibrated amplitudes from 10nV to 1 volt rms, handsome digital display and bus interface, and nifty accessories that extend the frequency range to 2.6GHz and provide modulation and frequency sweeping. This is a bit more than you usually need to do the job.



## □ 7.21 Bandwidth limiting and rms voltage measurement

### □ *Limiting the bandwidth*

All the measurements we have been talking about assume that you are looking at the noise output only in a limited frequency band. In a few cases the amplifier may have provision for this, making your job easier. If not, you have to hang some sort of filter on the amplifier output before measuring the output noise voltage.

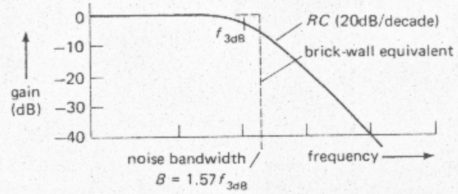


Figure 7.62. Equivalent "brick-wall" noise bandwidth for $RC$ low-pass filter.

The easiest thing to use is a simple $RC$ low-pass filter, with 3dB point set at roughly the bandwidth you want. For accurate noise measurements, you need to know the equivalent "noise bandwidth," i.e., the width of a perfect "brick-wall" low-pass filter that lets through the same noise voltage (Fig. 7.62). This noise bandwidth is what should be used for $B$ in all the preceding formulas. It is not terribly difficult to do the mathematics, and you find

$$B = \frac{\pi}{2} f_{3dB} = 1.57 f_{3dB}$$

For a pair of cascaded $RC$s (buffered so they don't load each other), the magic formula becomes $B = 1.22 f_{3dB}$. For the Butterworth filters discussed in Section 5.05, the noise bandwidth is

$B = 1.57\ f_{3dB}$ (1 pole)
$B = 1.11\ f_{3dB}$ (2 poles)
$B = 1.05\ f_{3dB}$ (3 poles)
$B = 1.025\ f_{3dB}$ (4 poles)

If you want to make band-limited measurements up at some center frequency, you can just use a pair of $RC$ filters (Fig. 7.63), in which case the noise bandwidth is as indicated. If you have had experience with contour integration, you may wish to try the following exercise.

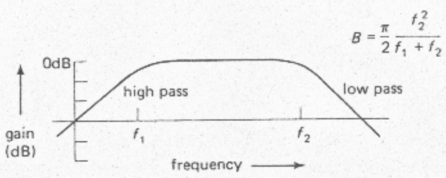


Figure 7.63. Equivalent "brick-wall" noise bandwidth for $RC$ bandpass filter.

EXERCISE 7.8

Optional exercise: Derive the preceding result, beginning with the response functions of $RC$ filters. Assume unit power per hertz input signal, and integrate the output power from zero to infinity. A contour integral then gets you the answer.

Another way to make a bandpass filter for noise measurements is to use an $RLC$ circuit. This is better than a pair of cascaded high-pass and low-pass $RC$ filters if you want your measurement over a bandpass that is narrow compared with the center frequency (i.e., high $Q$). Figure 7.64 shows both parallel and series $RLC$ circuits and their exact noise bandwidths. In both cases the resonant frequency is given by $f_0 = 1/2\pi\sqrt{LC}$. You might arrange the bandpass filter circuit as a parallel $RLC$ collector (or drain) load, in which case you use the expression as given. Alternatively, you might interpose the filter as shown in Figure 7.65; for noise bandwidth purposes the circuit is exactly equivalent to the parallel $RLC$, with $R = R_1 \| R_2$.

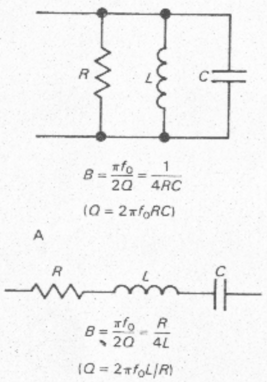


Figure 7.64. Eqivalent "brick-wall" noise bandwidth for $RLC$ bandpass filter.

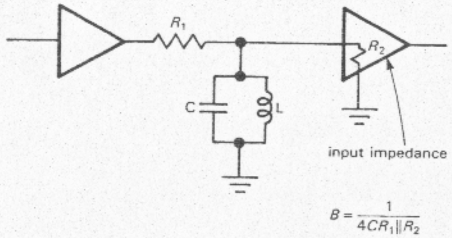


Figure 7.65

### □ *Measuring the noise voltage*

The most accurate way to make output noise measurements is to use a true rms voltmeter. These operate either by measuring the heating produced by the signal waveform (suitably amplified) or by using an analog squaring circuit followed by averaging. If you use a true rms meter, make sure it has response at the frequencies you are measuring; some of them only go up to a few kilohertz. True rms meters also specify a "crest factor," the ratio of peak voltage to rms that they can handle without great loss of accuracy. For Gaussian noise, a crest factor of 3 to 5 is adequate.

You can use a simple averaging-type ac voltmeter instead, if a true rms meter is unavailable. In that case, the values read off the scale must be corrected. As it turns out, all averaging meters (VOMs, DMMs, etc.) already have their scales adjusted, so what you read isn't actually the *average*, but rather the rms voltage *assuming a sine-wave signal*. For example, if you measure the power-line voltage in the United States, your meter will read something close to 117 volts. That's fine, but if the signal you're reading is Gaussian noise, you have to apply an additional correction. The rule is as follows: To get the rms voltage of Gaussian noise, multiply the "rms" value you read on an averaging ac voltmeter by 1.13 (or add 1dB). Warning: This works fine if the signal you are measuring is pure noise (e.g., the output of an amplifier with a resistor or noise source as input), but it won't give accurate results if the signal consists of a sine wave added to noise.

A third method, not exactly world-famous for its accuracy, consists of looking at the noise waveform on an oscilloscope: The rms voltage is 1/6 to 1/8 of the peak-to-peak value (depending on your subjective reading of the pp amplitude). It isn't very accurate, but at least there's no problem getting enough measurement bandwidth.

## 7.22 Noise potpourri

Herewith a collection of interesting, and possibly useful, facts.

*1.* The averaging time required in an indicating device to reduce the fluctuations of



a rectified noise signal to a desired level for a given noise bandwidth is

$$\tau \approx \frac{1600}{B\sigma_2} \text{ seconds}$$

where $\tau$ is the required time constant of the indicating device to produce fluctuations of standard deviation $\sigma$ percent at the output of a linear detector whose input is noise of bandwidth $B$.

*2.* For band-limited white noise the expected number of maxima per second is

$$N = \sqrt{\frac{3(f_2^5 - f_1^5)}{5(f_2^3 - f_1^3)}}$$

where $f_1$ and $f_2$ are the lower and upper band limits. For $f_1 = 0$, $N = 0.77f_2$; for narrowband noise ($f_1 \approx f_2$), $N \approx (f_1 + f_2)/2$.

*3.* rms-to-average (i.e., average magnitude) ratios:

| | |
|---|---|
| Gaussian noise: | rms/avg $= \sqrt{\pi/2} = 1.25$ $= 1.96$dB |
| Sine wave: | rms/avg $= \pi/2^{\frac{3}{2}} = 1.11$ $= 0.91$dB |
| Square wave: | rms/avg $= 1 = 0$dB |

*4.* Relative occurrence of amplitudes in Gaussian noise. Figure 7.66 gives the fractional time that a given amplitude level is exceeded by a Gaussian noise waveform of amplitude 1 volt rms.

# INTERFERENCE: SHIELDING AND GROUNDING

"Noise" in the form of interfering signals, 60Hz pickup, and signal coupling via power supplies and ground paths can turn out to be of far greater practical importance than the intrinsic noise sources we've just discussed. These interfering signals can all be reduced to an insignificant level (unlike thermal noise) with proper layout and construction. In stubborn cases the cure may involve a combination of filtration of input and output lines, careful

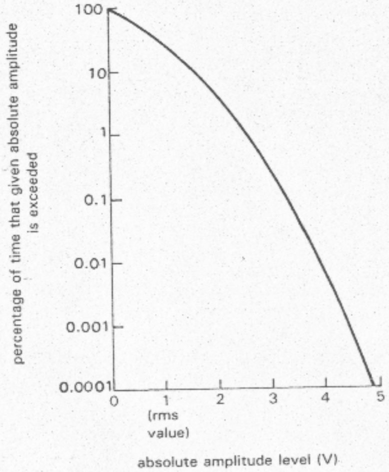


Figure 7.66. Relative occurrence of amplitudes in Gaussian noise.

layout and grounding, and extensive electrostatic and magnetic shielding. In these sections we would like to offer some suggestions that may help illuminate this dark area of the electronic art.

## 7.23 Interference

Interfering signals can enter an electronic instrument through the power-line inputs or through signal input and output lines. In addition, signals can be capacitively coupled (electrostatic coupling) onto wires in the circuit (the effect is more serious for high-impedance points within the circuit), magnetically coupled to closed loops in the circuit (independent of impedance level), or electromagnetically coupled to wires acting as small antennas for electromagnetic radiation. Any of these can become a mechanism for coupling of signals from one part of a circuit to another. Finally, signal currents from one part of the circuit can couple to other parts through voltage drops on ground lines or power-supply lines.