BURR - BROWN®

APPLICATION BULLETIN

Mailing Address: PO Box 11400, Tucson, AZ 85734 • Street Address: 6730 S. Tucson Blvd., Tucson, AZ 85706 • Tel: (520) 746-1111
Telex: 066-6491 • FAX (520) 889-1510 • Product Info: (800) 548-6132 • Internet: www.burr-brown.com/ • FAXLine: (800) 548-6133

NOISE ANALYSIS FOR HIGH SPEED OP AMPS

by Michael Steffes

As system bandwidths have increased, an accurate estimate of the noise contribution for each element in the signal channel has become increasingly important. Many designers are not, however, particularly comfortable with the calculations required to predict the total noise for an op amp, or in the conversions between the different descriptions of noise. Considerable inconsistency between manufacturers in describing noise and, in some cases, incomplete specifications, have contributed to this confusion. A thorough description of the op amp noise model will be developed here with a detailed discussion of the key differences between current and voltage feedback amplifiers. The conversions between several different measures for noise used in the industry will also be described. Broadband effects will be covered for both low frequencies (the 1/f region) and high frequencies (noise power bandwidth).

FUNDAMENTALS OF NOISE ANALYSIS

Random electrical noise (either a current or a voltage) is present in almost every type of component used in a circuit. This noise may be considered as either a frequency domain phenomena or something that occurs over time. The most common models approach noise from the frequency domain first then convert to the time domain using the shape of the noise density curve combined with a noise power bandwidth analysis. This is the approach that will be used here.

Another useful view of op amp noise is to consider the input voltage and current noises to be the time varying component of the input offset voltage and bias currents respectively. Working from the frequency domain to the time domain, as will be done here, develops the required tools to predict the amplitude of this time varying component. Only the random electrical noise generated by the components themselves will be considered. Other sources of "noise" that will **not** be considered here, (but are nevertheless of interest to the system designer), include conducted noise through the power supplies that appear at the output due to finite PSRR, various sources of radiated emission pickup (EMI), micro-phonic effects due to system vibration, and high narrowband noise that is in fact a parasitic oscillation.

The starting point for a frequency domain analysis of noise is the noise density. This is the noise power normalized to 1Hz bandwidth (at a particular center frequency) and is sometimes called the "spot" noise. "White" noise has a flat (or constant) noise power over frequency. Most amplifiers and resistors show a flat noise region that extends over many frequency decades. Most, however, also show an increasing noise power density at low frequencies. This is often called the 1/f region since the noise power density will often increase as the inverse of frequency. "Popcorn" noise is a random long term shift in voltage or current (that sounds like popcorn if fed into an audio speaker). This phenomena doesn't fit well into a frequency domain description and is best observed over time.

To analyze noise in the frequency domain, equivalent noise voltage or current generators are introduced into the circuit that represent the noise density over frequency for that element. These voltages or currents are the square root of the noise power densities. We work with voltages and current in order to avail ourselves of standard circuit analysis techniques. One key caveat to using noise voltages and currents is that they do not add algebraically. Each individual noise source has a random phase to any other (with the exception of correlated sources). This means that, although we can use superposition to get the contribution of each noise term to a particular point in the circuit, the voltages or currents themselves cannot simply be added at that point. It is the powers that are added to get to a total noise power.

OP AMP NOISE MODEL

Figure 1 shows the analysis circuit that will act as a starting point for the subsequent noise analysis. This schematic includes the three equivalent input noise terms for the op amp and the three resistor noise terms that must always be considered for a complete op amp noise analysis. Any particular op amp application circuit can typically be reduced to that of Figure 1 by shorting any input voltage sources and/or opening any input current sources that might be driving the circuit and reducing the remaining impedances to the three elements shown in Figure 1. Reactive elements (capacitors, inductors, transformers) are normally considered to be noiseless. They can, however, strongly influence the frequency response for the noise generators in a circuit. Examples of this will be shown later. At this point consider the elements around the op amp to be purely resistive. Remember that the Johnson noise of the resistor may be represented as either a current or a voltage. The analysis circuit of Figure 1 uses both forms (current noise for R_G and voltage noise for R_F) to simplify later computations.

Most of the discrepancies between different noise analysis come from neglecting as insignificant some of the noise sources in Figure 1. In general, there will in fact be a dominant noise source for a particular op amp in a particular application circuit. However, to maintain the required gen-

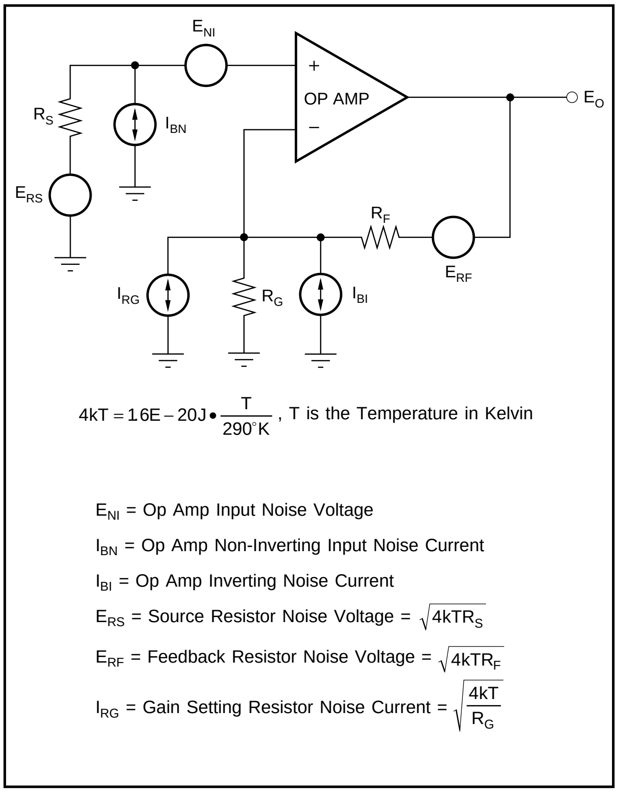

FIGURE 1. Op Amp Noise Analysis Circuit.

erality to handle any op amp in any configuration, all terms will initially be considered, then dropped later where appropriate.

The noise voltage and current sources shown in Figure 1 are treated as spot sources that may have a frequency response of their own. In particular, the op amp input voltage noise, E_{NI}, will typically increase at low frequencies due to 1/f noise effects. The two op amp input current noise terms will also show increasing noise at low frequencies for bipolar input stages. FET input stages have a very low and constant spot input noise currents going to low frequencies, but show

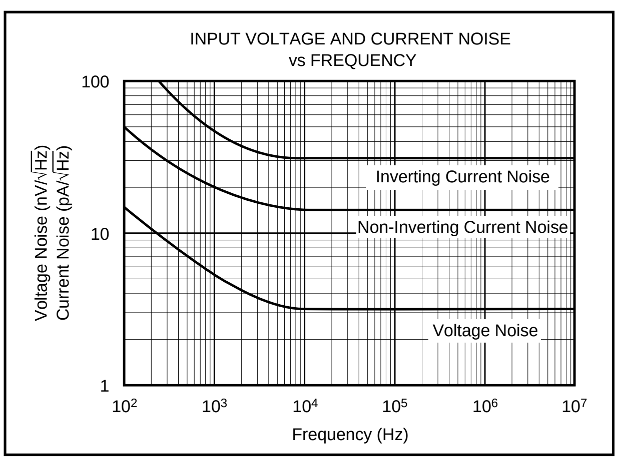

FIGURE 2. Example Input Noise Density Plot for a Bipolar Input Op Amp (OPA658).

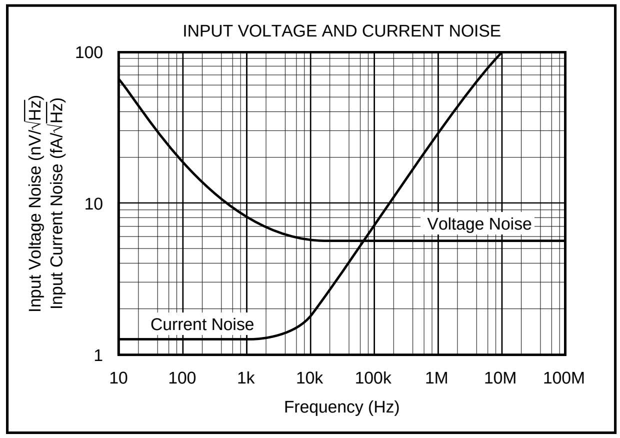

FIGURE 3. Example Input Noise Density Plot for a FET Input Op Amp (OPA655).

an increasing noise at higher frequencies. Figure 2 shows an example input noise density plot for a high speed bipolar input op amp while Figure 3 shows an example for a high speed FET input op amp. Note that Figure 2 shows separate traces for the two input current noise terms while Figure 3 shows a single trace implying that the two are equal (which is normally true for a voltage feedback op amp). One key difference between high speed current feedback op amps (such as the OPA658) and voltage feedback op amps (such as the OPA655) are that the two inputs for the current feedback topology show unequal bias and noise currents. These three input noise terms are intended to model all of the actual internal noise sources of the device. By using an input referred model, the total output noise under any configuration may be calculated.

Normally, the resistor noise terms are considered to have a constant noise voltage (or current) density over frequency. Along with this Johnson noise, there is also a low frequency increase in resistor noise (called “excess” noise) that is dependent on the DC voltage across the resistor. Carbon composition resistors show the highest excess noise with metal film showing a negligible increase in noise at low frequencies (Ref.1, page 171). Wirewound resistor show the lowest excess noise but should never be used in the signal path for high frequency op amps due to their inductive impedance. For this discussion, resistor noise density will be considered flat over frequency (white noise).

CALCULATING THE TOTAL OP AMP OUTPUT SPOT NOISE

As a starting point, calculate the total output spot noise voltage (E_O) of Figure 1. Each term at the output is simply the value of each source at a particular frequency taken to the output by its gain at that frequency. For now, treat all sources as white noise and neglect any frequency response effects for the gain terms in Figure 1. If all of the noise voltage and current sources are uncorrelated (the most common assumption), their powers will add algebraically at the output. This means that the total output spot noise power is the sum of the noise power contributed by each term. Take the square root of this power to get the total output spot noise voltage.

Compute the total output noise assuming:

$(1+R_F/R_G) \equiv G_N$ = Noise gain (identically equal to the op amp non-inverting signal gain)

First find the gain to the output for each voltage or current noise term by superposition:

Noise Term	Gain
E_{NI}	G_N
I_{BN}	$R_S \bullet G_N$
E_{RS}	G_N
I_{BI}	R_F
E_{RF}	1
I_{RG}	R_F

Now, substitute in for the resistor noises (Figure 1), multiply each noise term by its gain, square each term (to get to a power), sum them all, and take the square root of that sum to develop the total output spot noise expression.

Equation 1a:

$$E_O = \sqrt{(E_{NI}G_N)^2 + (I_{BN}R_SG_N)^2 + 4kTR_SG_N{}^2 + (I_{BI}R_F)^2 + 4kTR_F + \frac{4kT}{R_G}R_F{}^2}$$

Combining terms at the non-inverting input and recognizing that:

$$4kTR_F + \frac{4kT}{R_G}R^2{}_F = 4kTR_F \bullet G_N$$

will give the total output spot noise voltage for any op amp at a selected frequency.

Equation 1b:

$$E_O = \sqrt{\left(E_{NI}{}^2 + (I_{BN}R_S)^2 + 4kTR_S\right)G_N{}^2 + (I_{BI}R_F)^2 + 4kTR_FG_N}$$

To consider the frequency response effects for any of the gain terms in Equation 1b, substitute in the magnitude of the gain (for R_S, G_N, and/or R_F) at the frequency of interest. Parasitic, or intentional, capacitances across the resistors can strongly influence the spot noise gain for the terms of Equation 1b. One common trick in an inverting op amp configuration is to source match with R_S set to get DC bias current error cancellation (by setting $R_S = R_F \parallel R_G$). To limit the noise added by this DC matching resistor, add a large capacitor across R_S to filter the $4kTR_S \bullet G_N$ and $(I_{BN} \bullet R_S \bullet G_N)$^2 terms in the output spot noise expression.

Several simplifications of Equation 1b are possible. For example, R_S is often set to = $R_F \parallel R_G$ to get bias current cancellation when using a voltage feedback op amp. This is seldom done for current feedback op amps since their input bias currents do not match. Assuming the two input noise currents are equal in the voltage feedback case ($I_{BN} = I_{BI} = I_B$), and making the substitution for R_S will reduce Equation 1b to:

Equation 2:

$$E_O = \sqrt{E_{NI}{}^2G_N{}^2 + 2(I_BR_F)^2 + 2(4kTR_FG_N)}$$

Looking at Equation 2, once E_{NI}, I_B, and G_N are set, the output noise can always be reduced by decreasing the value for R_F (and then R_G and R_S to get the desired gain and source matching). That approach is limited by the increased loading presented by the feedback network, and, in the case of a current feedback op amp, by stability considerations. The value for R_F controls the compensation for a current feedback op amp. Reducing it too much can cause excessive frequency response peaking and possibly oscillations.

CALCULATING THE OP AMP'S TOTAL EQUIVALENT INPUT SPOT NOISE VOLTAGE

Equation 1b gives us an expression for a noise that can be physically measured. Connecting this output into a spectrum analyzer and converting the measured power at a particular frequency (and resolution bandwidth) into spot noise voltage will give the value predicted by Equation 1b. Often, designers would prefer to compare a total equivalent input spot noise voltage to their input signal. This input noise is not the same thing as the input spot voltage noise of the op amp itself. It is instead an abstraction derived by "input referring" the total output spot noise expression (Equation 1b). To "input refer" the total spot output noise, divide it by the gain from the desired signal input point to the output. The "input" can in fact be anywhere in the system. It could be to an inverting input signal—or, to the input of a prior stage (where it would be combined with the input referred noise of that stage to see how much is being added by this 2nd stage). Most often, the input referred spot noise voltage is taken to be the total input referred noise at the non-inverting input, E_N. Dividing Equation 1b by G_N will "input refer" to the non-inverting input.

Equation 3a:

$$E_N = \sqrt{E_{NI}{}^2 + (I_{BN}R_S)^2 + 4kTR_S + \left(\frac{I_{BI}R_F}{G_N}\right)^2 + \frac{4kTR_F}{G_N}}$$

Equation 3a explicitly shows an R_F in the last two terms. Since R_F is relatively fixed for a current feedback op amp, this shows clearly that the input referred contribution of the noise terms that are physically on the inverting side of the circuit will decrease with increasing gain. An equivalent expression, that is more common for voltage feedback amplifiers, substitutes $R_F/G_N = R_F \parallel R_G$ to get:

Equation 3b:

$$E_N = \sqrt{E_{NI}{}^2 + (I_{BN}R_S)^2 + 4kTR_S + \left(I_{BI}(R_F \| R_G)\right)^2 + 4kT(R_F \| R_G)}$$

Imposing the source matching condition ($R_S = R_F \parallel R_G$) and assuming that $I_{BN} = I_{BI}$ on Equation 3b will allow a further simplification to—

Equation 4:

$$E_N = \sqrt{E_{NI}{}^2 + 2\left(I_B(R_F \| R_G)\right)^2 + 2(4kT)(R_F \| R_G)}$$

CONVERTING SPOT INPUT OR OUTPUT NOISE TO A POWER

Most frequency domain oriented applications deal only in powers. Either the input or output spot noise expressions may be converted to dBm. Equation 5 shows this for the output spot noise voltage.

Equation 5:

$$P_N = 10 \bullet \log\left[\frac{E_O^{\,2}}{50\Omega(0.001)}\right] = \left[13 + 20\log(E_O)\right]\text{dBm}$$

Again, this is the noise power in a 1Hz bandwidth across 50Ωs. A good point of reference here is that a 50Ω resistor at 290ºK will have a spot noise power of –168dBm.

CONVERTING SPOT NOISE TO INTEGRATED NOISE

The spot noise over frequency is a useful means of comparing devices and interpreting spectrum analyzer measurements. It is also the most fundamentally useful starting point for noise analysis. However, very few systems actually have a signal that is defined in 1Hz bandwidths and a conversion from spot noise to the noise over some bandwidth is required. Computing the "integrated noise" is that process of summing all of the noise power over the frequency band of interest. Again, since noise really only adds algebraically as a power, the integration is first done using noise voltages squared, then converted back to voltage by taking the square root. Considering first the output integrated noise, there are really two frequency dependent parts to this calculation. First, the frequency response of the input spot noise term itself and then the frequency response of the gain for that term to the output. Even assuming these are known, there are a couple of interesting nuances here:

1. Starting at Frequency = 0 will not work. First of all, 1/f models go to infinite noise density at F = 0 and secondly, at zero frequency—we must be talking about the beginning of time—so at T = 0 we have infinite noise (big bang theory).
2. What high frequency limit should be used? Often the frequency response of the amplifier itself is used (or more precisely, the frequency response for each term in Equation 1b). This can be pretty painful (from an analysis standpoint) if the separate frequency response shapes for each noise term are meticulously applied. Most actual systems that care about noise impose a bandlimit that is set just prior to the detection stage. It is this bandlimit, either an IF filter for frequency domain applications or a passive low pass for time domain, that sets the system Noise Power Bandwidth (NPB).

   The NPB is often set so that the output spot noise is flat over that frequency (i.e., none of the individual noise sources are allowed to self limit). In an IF chain, the last IF or baseband filter is much lower bandwidth than any of the preceding stages. In a pulse domain application, this last low pass filter is lower bandwidth than the preceding amplifier stages. In these cases, the frequency response that each term has separately to the output may be neglected and the frequency response of this last filter will be the limiting factor.

   Most designers consider a frequency response from a gain and phase standpoint for the V/V transfer function. For noise, it is really the power gain that is important since we can only add noise powers directly. This means that whatever V/V transfer function we have, it must be squared prior to noise integration. There are a couple of simple conversions between standard low pass transfer functions and their equivalent NPB. For a first order low pass filter, the –3dB bandwidth (F_{-3dB}) is related to the noise power bandwidth by:

   Equation 6:

$$NPB = \frac{\pi}{2} \bullet F_{-3dB}\ \text{Hz} \quad \text{(Single Pole Low–Pass)}$$

For a 2nd order low pass, the NPB may also be related to the V/V transfer function (in W_O and Q terms) by:

Equation 7:

$$NPB = \frac{\pi}{2} \bullet F_O \bullet Q\text{Hz} \quad \text{(2nd Order Low–Pass)}$$

where:

$$F_{-3dB} = F_O \bullet \sqrt{\left(1 - \frac{1}{2Q^2}\right) + \sqrt{\left(1 - \frac{1}{2Q^2}\right)^2 + 1}}\ \text{Hz}$$

$$F_O = \frac{W_O}{2\pi}\text{Hz}$$

Both of these are calculating the equivalent brick wall bandwidth that will integrate the same "power" as the original V/V transfer function. W_O and Q may be estimated in several ways from the measured 2nd order frequency response. As one example, a maximally flat Butterworth 2nd order response has a Q = 0.707 and an $F_O = F_{-3dB}$ yielding a NPB = $1.11 \bullet F_{-3dB}$. The results of Equation 7 show that the equivalent NPB increases linearly with Q. This is saying that an increasingly peaked frequency response has the same effect as a broader NPB. Physically, for a flat input spot noise, this peaked response is giving more noise gain (in the peaked region). This can be equivalently accounted for by increasing the NPB. All this effort to get a single number for NPB allows a simple calculation for integrated noise. If the spot output noise is considered flat, the integrated noise is simply:

Equation 8:

$$E_{RMS} = E_O \bullet \sqrt{NPB}$$

This is converting from a frequency domain noise description to a time domain. Multiplying the square root of the measurement channel's noise power bandwidth times the spot noise (assumed flat) will give the RMS noise voltage in that bandwidth. One final conversion that is often of interest is to take this RMS noise voltage to a peak voltage number. The most common "crest" factor to convert RMS to V_{PP} is that for a sinusoid (e.g. $2\sqrt{2} = 2.8$). Noise is distinctly not sinusoidal. Using 6•Vrms will give a V_{PP} limit rarely exceeded.

Considering noise power bandwidth in a frequency domain application is slightly different. Most IF filters are multiple pole so that their NPB approach their F_{-3dB} bandwidths. To convert to integrated noise power, simple add 10•log (NPB) to the spot noise power of Equation 5 where the NPB is usually the same as the F_{-3dB} bandwidth.

Now consider the frequency response for the input noise terms themselves. If the bandwidth of interest for the system includes the low frequency 1/f noise region, a significant increase in integrated noise is possible due to this increased spot noise at lower frequencies. Equation 9 shows the spot noise equation for 1/f noise.

Equation 9:

$$E_T = E_N \bullet \sqrt{1 + \frac{F_{3dB}}{F}}, \text{ spot noise voltage over frequency}$$

where:

E_N = Flatband Spot Noise Voltage

F_{3dB} = Frequency Where Total Spot Noise Power Has Doubled

F = Frequency

Squaring Equation 9 and taking the integral from some arbitrarily low frequency up to the NPB will give the total power contributed over that band. Dividing by the frequency band of integration and taking the square root will give us an equivalent white noise that will integrate to the same power as the actual source described by Equation 9. The integral is set up in Equation 10a, solved in 10b, and taken back to an equivalent spot input voltage noise in Equation 10c.

Equation 10a:

$$E_{EQ}{}^2 = \frac{1}{F_2 - F_1} \int_{F_1}^{F_2} E_N{}^2 \left(1 + \frac{F_{3dB}}{F}\right) df$$

where:

$F_1 \rightarrow$ Lower Frequency Limit

$F_2 \rightarrow$ NPB

Equation 10b:

$$E_{EQ}{}^2 = \frac{E_N{}^2}{F_2 - F_1}\left[(F_2 - F_1) + F_{3dB}\, ln\left(\frac{F_2}{F_1}\right)\right] = E_N{}^2\left[1 + \frac{F_{3dB}}{F_2 - F_1} ln\frac{F_2}{F_1}\right]$$

Equation 10c:

$$E_{EQ} = E_N \sqrt{1 + \frac{F_{3dB}}{F_2 - F_1} ln\frac{F_2}{F_1}}, \text{ equivalent white noise voltage}$$

As the maximum frequency of interest far exceeds the 1/f noise corner (F_{3dB}), Equation 10c approaches E_N.

To illustrate this, let F_1 be an arbitrarily low 10Hz (remember we can't use $F_1 = 0$ here) with F_{3dB} = 10kHz and sweep F_2 to yield the following result for the radical in Equation 10c.

F_2	$\sqrt{1 + \frac{F_{3dB}}{F_2 - F_1} ln\left(\frac{F_2}{F_1}\right)}$
1kHz	6.9
10kHz	2.8
100kHz	1.4
1MHz	1.06
10MHz	1.006

TABLE I.

As Table 1 shows, when F_2 (= NPB) far exceeds the 1/f noise corner frequency, the effects of this increasing noise at low frequency may be safely neglected. When this is not the case, each of the three input noise terms for the op amp should be recomputed using Equation 10c to include the integrated effects of low frequency noise. These results are then placed into the total output spot noise expression of Equation 1b. This result may then be multiplied by the $\sqrt{NPB}$ to arrive at an integrated noise voltage that correctly includes the low frequency effects.

COMPUTING NOISE FIGURE

Noise figure is a common description of noise effects for RF and IF amplifiers. It is defined as 10•log (Signal/Noise at the input divided by the Signal/Noise at the output). All terms are spot powers (noise and signal). The input noise power is defined as the noise power delivered from some source impedance to the input—often a matching impedance termination. This is somewhat complicated for an op amp since the input termination is set by the user where it is often fixed at a matching value for an RF or IF amplifier. Figure 4 shows a non-inverting op amp with an arbitrary input termination (R_T) showing the noise sources and definition points for noise figure analysis

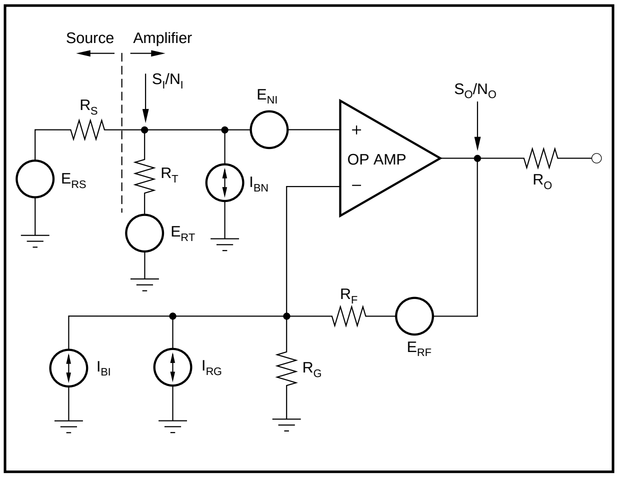

FIGURE 4. Non-Inverting Op Amp Noise Figure Analysis Circuit.

Letting R_T remain an arbitrary input termination yields a very general noise figure expression for the non-inverting op amp configuration (NF^+) shown in Equation 11 (where $R_P \equiv R_T \parallel R_S$).

Equation 11:

$$NF^+ = 10\log\left[1+\frac{R_S}{R_T}+\frac{R_S}{4kT}\left[\left(\frac{E_{NI}}{R_P}\right)^2+\left(I_{BN}\right)^2+\left(\frac{I_{BI}R_F}{R_P G_N}\right)^2+\frac{4kTR_F}{G_N {R_P}^2}\right]\right]$$

Two special cases for R_T greatly simplify Equation 11. Letting $R_T = R_S$ to get an input impedance match will produce Equation 12.

Equation 12:

$$NF^+ = 10\log\left[2+\frac{{E_{NI}}^2+\left(I_{BN}\frac{R_S}{2}\right)^2+\left(\frac{I_{BI}R_F}{G_N}\right)^2+\frac{4kTR_F}{G_N}}{kTR_S}\right]$$

$$\text{with } R_S = R_T$$

Letting R_T = infinity will not provide an input match, but will provide a lower noise figure as shown by Equation 13.

Equation 13:

$$NF^+ = 10\log\left[1+\frac{{E_{NI}}^2+\left(I_{BN}R_S\right)^2+\left(\frac{I_{BI}R_F}{G_N}\right)^2+\frac{4kTR_F}{G_N}}{4kTR_S}\right]$$

$$\text{with } R_T = \infty$$

Generally speaking, anything that will reduce the input referred voltage noise of the op amp (Equation 3a) will decrease the noise figure (an ideal noise figure = 0dB when the output SNR = input SNR). Changing R_S will trade off the effects of source resistor noise and op amp produced noise. It is sometimes suggested to operate at an optimum source resistance where the noise figure is minimized. This is effective when using reactive impedance transformations (such as a transformer) but is not really helping the noise if the increased source resistance is a real (noisy) resistor (Ref.2).

The noise figure for an op amp operated in the inverting mode is considerably more complicated. Figure 5 shows the noise figure analysis circuit for the inverting op amp topology. This circuit includes a resistor to ground on the non-inverting input, R_T, which should be set to a relatively low value for the lowest noise. It also includes a matching resistor to ground at the input to allow the input impedance to be set separately from the signal gain. The input impedance looking into this inverting op amp configuration is the parallel combination of $R_G \parallel R_M$.

Since the input impedance cannot be infinite, the main application for this circuit would be providing an input impedance matched to R_S. If R_M is constrained (given an R_G and R_S) to give an input impedance matched to R_S, an inverting op amp noise figure expression (NF^-) can be derived as shown in Equation 14.

Equation 14:

$$NF^- = 10\log\left[2\left(2\frac{R_G}{R_S}-1\right)+\frac{\left({E_{NI}}^2+\left(I_{BN}R_T\right)^2+4kTR_T\right)A_T^2+\left(\frac{I_{BI}R_F}{G_I}\right)^2+\frac{4kTR_F}{\left(G_I\right)^2}}{kTR_S}\right]$$

$$\text{where } A_T \equiv \frac{1+G_I\left(1-\frac{R_S}{2R_G}\right)}{G_I} = \text{non–inverting gain}$$

Since R_M has been constrained by the values for R_G and R_S, it does not appear in the noise figure expression. Using the same amplifier with a fixed R_F in either a non-inverting or inverting configuration will typically yield a $NF^- > NF^+$ at low gains with the inverting noise figure dropping below the non-inverting noise figure at higher gains. The non-inverting noise figure vs. gain asymptotically approaches a limit set by the noise terms that are physically on the non-inverting input. Those same noise terms are actually attenuated at higher inverting gains when input referred to the inverting input. Taking a simple example: let $R_S = R_G = 50\Omega$ with $R_F = 200\Omega$ (and R_M = infinity). Looking just at the op-amp's non-inverting input noise voltage, it will have a gain to the output that is (1 + 400/100) = 5. The magnitude of the

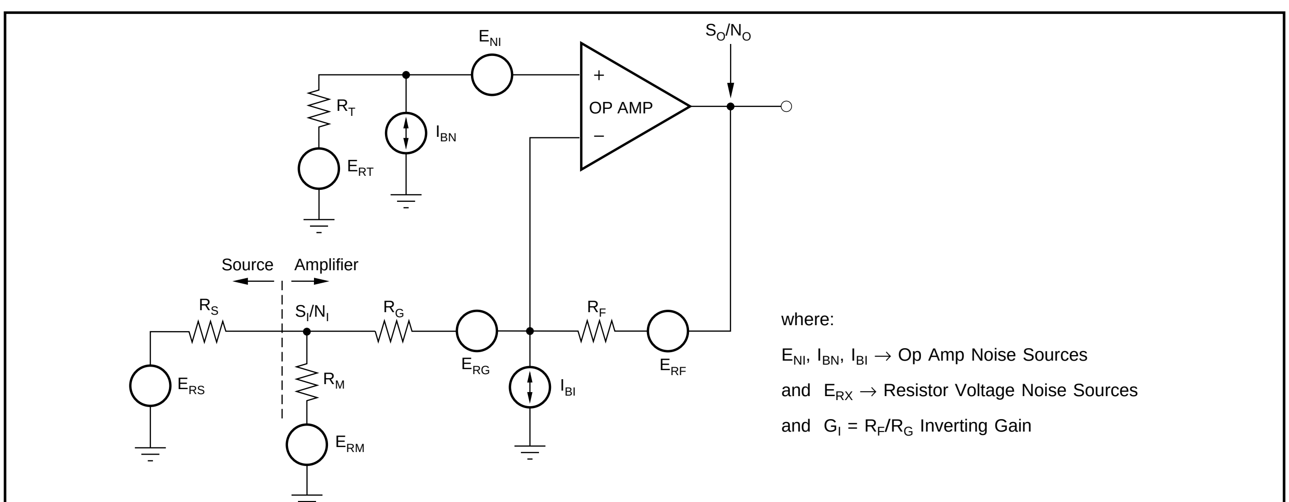

FIGURE 5. Inverting Op Amp Noise Figure Analysis Circuit.

inverting gain from the input reference point of Figure 3 is 400/50 = 8. Input referring the non-inverting input noise voltage to the inverting input yields a gain of 5/8 = 0.625 in this case. This effect gives a lower achievable noise figure in the inverting configuration than in the non-inverting for a matched input impedance case at higher gains.

EXAMPLE NOISE CALCULATIONS

To show how the analysis described here can be applied, let's consider two widely different applications using the OPA658 and OPA655 noise characteristics shown in Figure 2 and 3 respectively. First, apply the wideband current feedback OPA658 to an IF amplifier application where I/O impedances are matched to 50Ω and a gain of 10dB to the load over a frequency band from 5 to 50MHz is desired. A non-inverting circuit that achieves this is shown in Figure 6.

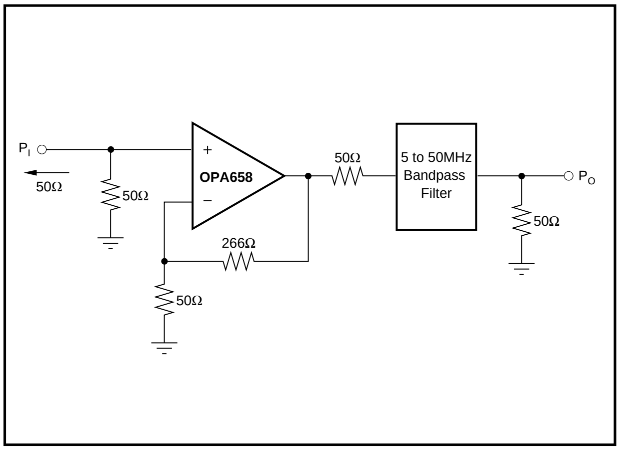

FIGURE 6. I_F Amplifier Example Circuit.

Since we are only interested in a frequency band well above the 1/f corner frequencies for any of the noise terms shown in Figure 2, those low frequency effects may be neglected. The OPA658 is operating at a voltage gain of 6.32 to get the desired 10dB gain to the matched load. At this gain (and with the feedback resistor reduced from the recommended 402Ω to 266Ω) the OPA658 will have over 200MHz bandwidth. The IF filter will therefore set the noise power bandwidth independently of the amplifiers frequency response.

The first step will be to compute the total output spot noise using Equation 1b. Note that the source resistor should be set to 25Ω for this calculation. Using the flatband numbers from Figure 2, Equation 15 shows this calculation of total output spot noise voltage (at the output pin of the amplifier).

Equation 15:

$$E_O = \sqrt{\left[(3.2\text{nV})^2 + (12.6\text{pA} \bullet 25\Omega)^2 + 16\text{E}-21 \bullet 25\Omega\right](6.32)^2 + (32\text{pA} \bullet 266\Omega)^2 + 16\text{E}-21 \bullet 266\Omega \bullet 6.32}$$
$$= 23\text{nV}/\sqrt{\text{Hz}}$$

To input refer this voltage noise, simple divide by the voltage gain to get 23nV/6.32 = 3.64nV/$\sqrt{\text{Hz}}$. To convert this spot input noise voltage to a spot noise power, use Equation 5 to get –156dBm. To get a noise floor for signal detection at the input, add 10•log ((50-5)MHz) to this to get –79dBm. Normally, to compute a minimum detectable signal, RF engineers add 3dB to this number. Notice that it was the system defining noise power bandwidth that is physically after this stage that is used for this calculation. The last common measure of noise used in an IF application would be the Noise Figure. Substituting into Equation 12 we can compute that:

Equation 16:

$$NF^+ = 10\log\left(2 + \frac{(3.64\text{nV})^2}{(4\text{E}-21) \bullet 50}\right) = 18.3\text{dB}$$

Notice that the noise term used in this calculation is simply the total input referred voltage noise calculated earlier using a source impedance equal to $R_S/2$. How would this noise figure change for the inverting configuration? Holding R_G = 50Ω, and increasing R_F to 316Ω to get the same gain magnitude, and holding R_T = 25Ω on the non-inverting input will yield an inverting noise figure equal to (using Equation 14).

Equation 17:

$$NF^- = 10\log\left[2(2-1) + \frac{\left[(3.2\text{nV})^2 + (12.6\text{pA} \bullet 25\Omega)^2 + 16\text{E}-21 \bullet 25\Omega\right](.66)^2 + \left(\frac{32\text{pA} \bullet 316\Omega}{6.32}\right)^2 + \frac{16\text{E}-21 \bullet 316\Omega}{(6.32)^2}}{(4\text{E}-21)50\Omega}\right]$$
$$= 15.9\text{dB}$$

which is slightly less than the non-inverting configuration noise figure.

Now consider a DC coupled time domain example using the OPA655 FET input op amp. In this case, a very good pulse response is desired with minimal integrated noise over the bandwidth necessary to achieve the pulse settling times. Set the OPA655 up for a gain of +5 (using an R_F = 10kΩ) and follow it by a 5MHz cutoff 2nd order Butterworth filter. Figure 7 shows this application.

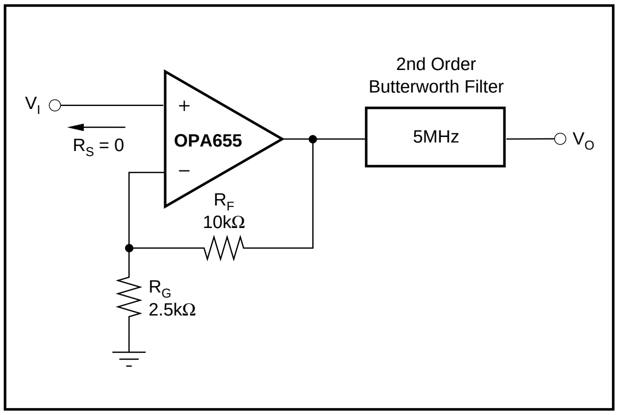

FIGURE 7. Low Frequency Pulse Amplifier.

This voltage feedback op amp has a gain bandwidth product of 240MHz. At a non-inverting signal gain of +5, we will achieve > 48MHz single pole bandwidth to the output of the op amp. The system bandwidth will be set by the low pass Butterworth filter following this stage. First calculate the noise power bandwidth for this filter. Using Equation 7 (and a Q = 0.707 and F_O = F–3dB = 5MHz):

NPB = 1.11*5MHz = 5.55MHz

Now include 1/f effect by computing the equivalent flatband noise for the non-inverting input noise voltage density shown in Figure 3. Using Equation 10c with F_{3dB} = 2kHz, F_1 = 10Hz, and F_2 = 5.55MHz, and E_N = 5.6nV/√Hz yields an equivalent constant input voltage noise :

Equation 18:

$$E_{EQ} = 5.6\text{nV}\sqrt{1+\frac{2\text{kHz}}{5.55\text{MHz}-10\text{Hz}}\, ln\frac{5\text{MHz}}{10}} = 5.613\text{nV}$$

Even at this relatively limited frequency band, the contribution of the noise in the 1/f region is negligible. Figure 3 shows the bias current noise increasing with frequency. Multiplying the 10MHz value of 0.1pA times the feedback resistor will still only yield a 1nV contribution at the output well out of band. As is normally the case, the current noise contribution to the total noise may be neglected for this FET input op amp. Equation 19 computes the total output spot noise considering only the voltage and resistor noise terms.

Equation 19:

$$E_O = \sqrt{(5.6\text{nV}\bullet 5)^2 + 16\text{E}-21\bullet 10\text{k}\Omega\bullet 5} = 40\text{nV}/\sqrt{\text{Hz}}$$

The resistor values were chosen here to add an almost equal noise power at the output as the op amp's input noise voltage. The total integrated noise at the output of the filter is then:

Equation 20:

$$\text{Integrated } E_O = \left(40\text{nV}/\sqrt{\text{Hz}}\right)\sqrt{5.55\text{MHz}} = 94\mu V_{rms}$$

And finally, the peak-peak noise excursion will almost always be less than 6•94µV = 0.56mV due to the noise added by this stage to the original source.

SUMMARY

Once the complete output spot noise equation for an op amp is developed (Equation 1b), all other descriptions or simplifications may be derived. This equation is also important in that it describes what is actually being measured in any noise measurement. Other numbers that may be reported are simply computations from this measurement. A pedantic output noise computation would also include the individual frequency responses to the output for each noise source. In many cases, these can be ignored if the stage in question is followed by a bandlimiting filter. Low frequency 1/f effects can be handled by computing an equivalent white noise source that will integrate to the same power over the band of interest (Equation 10c). This increasing low frequency noise makes a negligible contribution to integrated noise when the high frequency bandlimit is 100X the 1/f noise corner frequency. The noise figure for an op amp may be predicted using the equations developed here (Equation 11 to 14). Resistively source matching for optimum noise figure will lose the signal to noise ratio battle. Shifting the source impedance reactively can, however, effectively improve the noise figure. The inverting op amp configuration can provide a lower noise figure at higher gains. In general, start a noise analysis including every term, then drop those that are clearly contributing negligible noise in the application.

REFERENCES

1. *Low Noise Electronic Design*; Motchenbacher & Fitchen, Wiley 1973.
2. National Semiconductor Application Note AN104-1.