- Reduce high impedance positive inputs to the minimum allowable value (e.g., replace 1 Meg biasing resistors with 47k ohm, etc.).
- Add small (< 100 pF) capacitors across feedback resistors to reduce amplifier gain at high frequencies (Figure 2.2.3). **Caution: this assumes the amplifier is unity-gain stable.** If not, addition of this capacitor will *guarantee* oscillations. (For amplifiers that are not unity-gain stable, place a resistor in series with the capacitor such that the gain does not drop below where it is stable.)
- Add a small capacitor (size is a function of source resistance) at the positive input to reduce the impedance to high frequencies and effectively shunt them to ground.

2.3 NOISE

2.3.1 Introduction

The noise performance of IC amplifiers is determined by four primary noise sources: thermal noise, shot noise, 1/f, and popcorn noise. These four sources of noise are briefly discussed. Their contribution to overall noise performance is represented by equivalent input generators. In addition to these equivalent input generators, the effects of feedback and frequency compensation on noise are also examined. The noise behavior of the differential amplifier is noted since most op amps today use a differential pair. Finally noise measurement techniques are presented.

2.3.2 Thermal Noise

Thermal noise is generated by any passive resistive element. This noise is "white," meaning it has a constant spectral density. Thermal noise can be represented by a mean-square voltage generator $\overline{e_R^2}$ in series with a noiseless resistor, where $\overline{e_R^2}$ is given by Equation (2.3.1).

$$\overline{e_R^2} = 4k\,TRB \text{ (volts)}^2$$

where: T = temperature in °K

R = resistor value in ohms

B = noise bandwidth in Hz

k = Boltzmann's constant (1.38×10^{-23} W-sec/°K)

The RMS value of Equation (2.3.1) is plotted in Figure 2.3.1 for a one Hz bandwidth. If the bandwidth is increased, the plot is still valid so long as e_R is multiplied by $\sqrt{B}$.

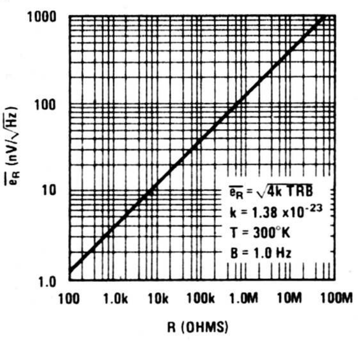


FIGURE 2.3.1 Thermal Noise of Resistor

Actual resistor noise measurements may have more noise than shown in Figure 2.3.1. This additional noise component is known as *excess noise*. Excess noise has a 1/f spectral response, and is proportional to the voltage drop across the resistor. It is convenient to define a *noise index* when referring to excess noise in resistors. *The noise index is the RMS value in μV of noise in the resistor per volt of DC drop across the resistor in a decade of frequency.* Noise index expressed in dB is:

$$NI = 20 \log \left(\frac{E_{ex}}{V_{DC}} \times 10^6 \right) \text{dB}$$

where: E_{ex} = resistor excess noise in μV per frequency decade.

V_{DC} = DC voltage drop across the resistor.

Excess noise in carbon composition resistors corresponds to a large noise index of +10dB to −20dB. Carbon film resistors have a noise index of −10dB to −25dB. Metal film and wire wound resistors show the least amount of excess noise, with a noise index figure of −15dB to −40dB. For a complete discussion of excess noise see Reference 2.

2.3.3 Noise Bandwidth

Noise bandwidth is not the same as the common amplifier or transfer function −3dB bandwidth. Instead, noise bandwidth has a "brick-wall" filter response. The maximum power gain of a transfer function T(jω) multiplied by the noise bandwidth must equal the total noise which passes through the transfer function. Since the transfer function power gain is related to the square of its voltage gain we have:

$$(T_{MAX}^2)B = \int_0^{\infty} |T(j\omega)|^2 \, d\omega \qquad (2.3.2)$$

where: T_{MAX} = maximum value of $T(j\omega)$

$T(j\omega)$ = transfer function voltage gain

B = noise bandwidth in Hz

For a single RC roll-off, the noise bandwidth B is $\pi/2\ f_{-3dB}$, and for higher order maximally flat filters, see Table 2.3.1.

TABLE 2.3.1 Noise Bandwidth Filter Order

Filter Order	Noise Bandwidth B
1	$1.57 f_{-3dB}$
2	$1.11 f_{-3dB}$
3	$1.05 f_{-3dB}$
4	$1.025 f_{-3dB}$
"Brick-wall"	$1.00 f_{-3dB}$

2.3.4 Shot Noise

Shot noise is generated by charge crossing a potential barrier. It is the dominant noise mechanism in transistors and op amps at medium and high frequencies. The mean square value of shot noise is given by:

$$\overline{I_S^2} = 2q\, I_{DC}\, B \text{ (amps)}^2 \qquad (2.3.3)$$

where: q = charge of an electron in coulombs

I_{DC} = direct current in amps

B = noise bandwidth in Hz

Like thermal noise, shot noise has a constant spectral density.

2.3.5 1/f Noise

1/f or flicker noise is similar to shot noise and thermal noise since its amplitude is random. Unlike thermal and shot noise, 1/f noise has a 1/f spectral density. This means that the noise increases at low frequencies. 1/f noise is caused by material and manufacturing imperfections, and is usually associated with a direct current:

$$\overline{I_f{}^2} = K \frac{(I_{DC})^a}{f} B \text{ (amps)}^2 \qquad (2.3.4)$$

where: I_{DC} = direct current in amps

K and a = constants

f = frequency in Hz

B = noise bandwidth in Hz

2.3.6 Popcorn Noise (PCN)

Popcorn noise derives its name from the popcorn-like sound made when connected to a loudspeaker. It is characterized by a sudden change in output DC level, lasting from microseconds to seconds, recurring randomly. Although there is no clear explanation of PCN to date, it is usually reduced by cleaner processing (see Reference 5). In addition, extensive testing techniques are used to screen for PCN units.

2.3.7 Modelling

Every element in an amplifier is a potential source of noise. Each transistor, for instance, shows all three of the above mentioned noise sources. The net effect is that noise sources are distributed throughout the amplifier, making analysis of amplifier noise extremely difficult. Consequently, amplifier noise is completely specified by a noise voltage and a noise current generator at the input of a noiseless amplifier. Such a model is shown in Figure 2.3.2. Correlation between generators is neglected unless otherwise noted.

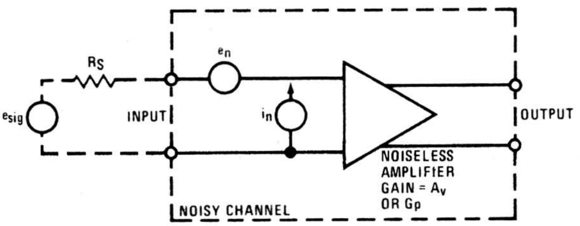


FIGURE 2.3.2 Noise Characterization of Amplifier

Noise voltage e_n, or more properly, equivalent short-circuit input RMS noise voltage, is simply that noise voltage which would appear to originate at the input of the noiseless amplifier if the input terminals were shorted. It is expressed in "nanovolts per root Hertz" (nV/$\sqrt{\text{Hz}}$) at a specified frequency, or in microvolts for a given frequency band. It is measured by shorting the input terminals, measuring the output RMS noise, dividing by amplifier gain, and referencing to the input – hence the term "equivalent input noise voltage." An output bandpass filter of known characteristic is used in measurements, and the measured value is divided by the square root of the bandwidth if data are to be expressed per unit bandwidth.

Figure 2.3.3 shows e_n of a typical op amp. For this amplifier, the region above 1 kHz is the shot noise region, and below 1 kHz is the amplifier's 1/f region.

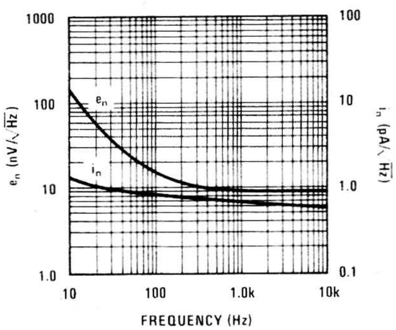


FIGURE 2.3.3 Noise Voltage and Current for an Op Amp

Noise Current, i_n, or more properly, equivalent open-circuit RMS noise current, is that noise which occurs apparently at the input of the noiseless amplifier due only to noise currents. It is expressed in "picoamps per root Hertz" (pA/$\sqrt{\text{Hz}}$) at a specified frequency or in nanoamps in a given frequency band. It is measured by shunting a capacitor or resistor across the input terminals such that the noise current will give rise to an additional noise voltage which is $i_n \times R_{in}$ (or X_{Cin}). The output is measured, divided by amplifier gain, and that contribution known to be due to e_n and resistor noise is appropriately subtracted from the total measured noise. If a capacitor is used at the input, there is only e_n and $i_n X_{Cin}$. The i_n is measured with a bandpass filter and converted to pA/$\sqrt{\text{Hz}}$ if appropriate. Again, note the 1/f and shot noise regions of Figure 2.3.3.

Now we can examine the relationship between e_n and i_n at the amplifier input. When the signal source is connected, the e_n appears in series with the e_{sig} and e_R. The i_n flows through R_s, thus producing another noise voltage of value $i_n \times R_s$. This noise voltage is clearly dependent upon the value of R_s. All of these noise voltages add at the input of Figure 2.3.2 in RMS fashion, that is, as the square root of the sum of the squares. Thus, neglecting possible correlation between e_n and i_n, the total input noise is:

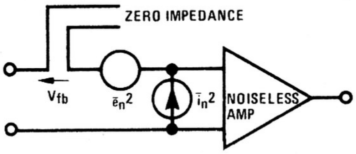
$$\overline{e_N{}^2} = \overline{e_n{}^2} + \overline{e_R{}^2} + \overline{i_n{}^2} R_s{}^2 \qquad (2.3.5)$$

2.3.8 Effects of Ideal Feedback on Noise

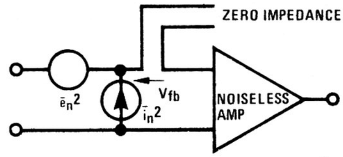
Extensive use of voltage and current feedback are common in op amps today. Figures 2.3.4a and 2.3.4b can be used to show the effect of voltage feedback on the noise performance of an op amp.

Figure 2.3.4a shows application of negative feedback to an op amp with generators $\overline{e_n{}^2}$ and $\overline{i_n{}^2}$. Figure 2.3.4b shows that the noise generators can be moved outside the feedback loop. This operation is possible since shorting both amplifiers' inputs results in the same noise voltage at the outputs. Likewise, opening both inputs gives the same noise currents at the outputs. For current feedback, the same result can be found. This is seen in Figure 2.3.5a and Figure 2.3.5b.

The significance of the above result is that the equivalent input noise generators completely specify circuit noise. *The application of ideal negative feedback does not alter the noise performance of the circuit.* Feedback reduces the output noise, but it also reduces the output signal. *In other words, with ideal feedback, the equivalent input noise is independent of gain.*

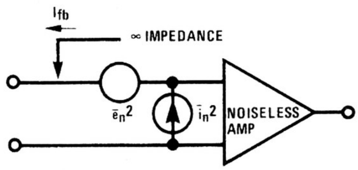


(a) Feedback Applied to Op Amp with Noise Generators

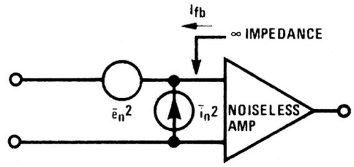


(b) Noise Generators Outside Feedback Loop

FIGURE 2.3.4

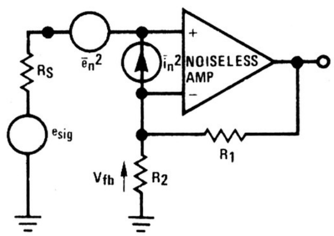


(a) Current Feedback Applied to Op Amp

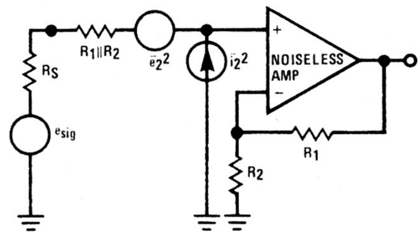


(b) Noise Generators Moved Outside Feedback Loop

FIGURE 2.3.5



(a) Practical Voltage Feedback Amplifier



(b) Voltage Feedback with Noise Generators Moved Outside Feedback Loop

FIGURE 2.3.6

2.3.9 Effects of Practical Feedback on Noise

Voltage feedback is implemented by series-shunt feedback as shown in Figure 2.3.6a.

The noise generators can be moved outside the feedback loop as shown in Figure 2.3.6b if the thermal noise of $R_1 \| R_2$ is included in $e_2{}^2$. In addition, the noise generated by $i_n \times (R_1 \| R_2)$ must be added even though the (–) input is a virtual ground (see Appendix 6). The above effects can be easily included if $R_1 \| R_2$ is considered to be in series with R_S.

$$\overline{e_2{}^2} = \overline{e_n{}^2} + 4k\,T\,(R_s + R_1 \| R_2)$$

$$\overline{i_2{}^2} = \overline{i_n{}^2}$$

Example 2.3.1

Determine the total equivalent input noise per unit bandwidth for the amplifier of Figure 2.3.6a operating at 1 kHz from a source resistance of 1 kΩ. R_1 and R_2 are 100 kΩ and 1 kΩ respectively.

Solution:

Use data from Figure 2.3.1 and Figure 2.3.3.

1. Thermal noise from $R_s + R_1 \| R_2 \approx 2k$ is $5.65\,nV/\sqrt{Hz}$.
2. Read e_n from Figure 2.3.3 at 1 kHz; this value is $9.5\,nV/\sqrt{Hz}$.
3. Read i_n from Figure 2.3.3 at 1 kHz; this value is $0.68\,pA/\sqrt{Hz}$. Multiply this noise current by $R_s + R_1 \| R_2$ to obtain $1.36\,nV/\sqrt{Hz}$.
4. Square each term and enter into Equation (2.3.5).

$$e_N = \sqrt{\overline{e_2{}^2} + \overline{i_2{}^2}\,(R_s + R_1 \| R_2)^2} \quad nV/\sqrt{Hz}$$

$$e_N = \sqrt{\overline{e_n{}^2} + 4\,k\,T\,(R_s + R_1 \| R_2) + \overline{i_n{}^2}\,(R_s + R_1 \| R_2)^2}$$

$$e_N = \sqrt{(9.5)^2 + (5.65)^2 + (1.36)^2}$$

$$e_N = 11.1\,nV/\sqrt{Hz}$$

This is total RMS noise at the input in one Hertz bandwidth at 1 kHz. If total noise in a given bandwidth is desired, one must integrate the noise over a bandwidth as specified. This is most easily done in a noise measurement set-up, but may be approximated as follows:

1. If the frequency range of interest is in the flat band, i.e., between 1 kHz and 10 kHz in Figure 2.3.3, it is simply a matter of multiplying e_N by the square root of the noise bandwidth. Then, in the 1 kHz-10 kHz band, total noise is:

$$e_N = 11.1\sqrt{9000}$$

$$= 1.05 \mu V$$

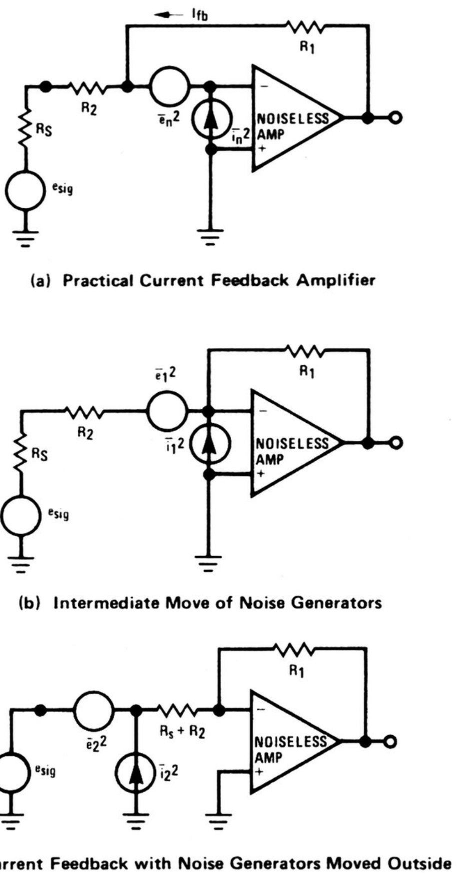
2. If the frequency band of interest is not in the flat band of Figure 2.3.3, one must break the band into sections, calculating average noise in each section, squaring, multiplying by section bandwidth, summing all sections, and finally taking square root of the sum as follows:

$$e_N = \sqrt{\overline{e_R}^2 B + \sum_1^i (\overline{e_N}^2 + \overline{i_n}^2 R_s^2)_i B_i} \qquad (2.3.6)$$

where: i is the total number of sub-blocks

For details and examples of this type of calculation, see application note AN-104, "Noise Specs Confusing?"

Current feedback is accomplished by shunt-shunt feedback as shown in Figure 2.3.7a.



(a) Practical Current Feedback Amplifier

(b) Intermediate Move of Noise Generators

(c) Current Feedback with Noise Generators Moved Outside Feedback Loop

FIGURE 2.3.7

$\overline{e_n}^2$ and $\overline{i_n}^2$ can be moved outside the feedback loop if the noise generated by R_1 and R_2 are taken into account.

First, move the noise generators outside feedback R_1. To do this, represent the thermal noise generated by R_1 as a noise current source (Figure 2.3.7b):

$$\overline{i_{R1}}^2 = 4k\,T \frac{1}{R_1} B$$

so: $$\overline{e_1}^2 = \overline{e_n}^2$$

and: $$\overline{i_1}^2 = \overline{i_n}^2 + 4k\,T \frac{1}{R_1} B$$

Now move these noise generators outside $R_s + R_2$ as shown in Figure 2.3.7c to obtain $\overline{e_2}^2$ and $\overline{i_2}^2$:

$$\overline{e_2}^2 = \overline{e_n}^2 + 4k\,T\,(R_s + R_2)\,B \qquad (2.3.7)$$

$$\overline{i_2}^2 = \overline{i_n}^2 + 4k\,T \frac{1}{R_1} B \qquad (2.3.8)$$

$\overline{e_2}^2$ and $\overline{i_2}^2$ are the equivalent input generators with feedback applied. The total equivalent input noise, e_N, is the sum of the noise produced with the input shorted, and the noise produced with the input opened. With the input of Figure 2.3.7c shorted, the input referred noise is $\overline{e_2}^2$. With the input opened, the input referred noise is:

$$\left(\frac{i_2 R_1}{A_V}\right)^2 = \overline{i_2}^2 (R_s + R_2)^2$$

The total equivalent input noise is:

$$e_N = \sqrt{\overline{e_2}^2 + \overline{i_2}^2 (R_s + R_2)^2}$$

Example 2.3.2

Determine the total equivalent input noise per unit bandwidth for the amp of Figure 2.3.7a operating at 1 kHz from a 1 kΩ source. Assume R_1 is 100 kΩ and R_2 is 9 kΩ.

Solution

Use data from Figures 2.3.1 and 2.3.3.

1. Thermal noise from $R_s + R_2$ is $12.7\,nV/\sqrt{Hz}$.
2. Read e_n from figure 2.3.3 at 1 kHz; this value is $9.5\,nV/\sqrt{Hz}$. Enter these values into Equation (2.3.7).
3. Determine the thermal noise current contributed by R_1:

$$i_{R1} = \sqrt{4k\,T \frac{1}{R_1} B} = \sqrt{\frac{1.61 \times 10^{-20}}{100k}} = 0.401\,pA/\sqrt{Hz}$$

4. Read i_n from Figure 2.3.3 at 1 kHz; this value is $0.68\,pA/\sqrt{Hz}$. Enter these values into Equation (2.3.7).

$$e_N = \sqrt{\overline{e_n}^2 + (R_s + R_2)^2 \left(\overline{i_n}^2 + 4k\,T \frac{1}{R_1} B\right) + 4k\,T\,(R_s + R_2)\,B}$$

$$e_N = \sqrt{(9.5)^2 + (10k)^2 (0.68^2 + 0.401^2) + (12.7)^2}\ nV/\sqrt{Hz}$$

$$e_N = 17.7\,nV/\sqrt{Hz}$$

For the noise in the bandwidth from 1 kHz to 10 kHz, $e_N = 17.7\,nV\sqrt{9000} = 1.68\,\mu V$. If the noise is not constant with frequency, the method shown in Equation (2.3.6) should be used.

TABLE 2.3.2 Equivalent Input Noise Comparison

NON-INVERTING AMPLIFIER					INVERTING AMPLIFIER				
A_V	R_s	R_1	R_2	e_N (nV$\sqrt{\text{Hz}}$)	A_V	R_s	R_1	R_2	e_N (nV$\sqrt{\text{Hz}}$)
101	1k	100k	1k	11.1	100	1k	100k	0	10.3
11	1k	100k	10k	17.3	10	1k	100k	9k	17.7
2	1k	100k	100k	46.0	2	1k	100k	49k	49.5
1	1k	100k	∞	80.2	1	1k	100k	99k	89.1

Example 2.3.3

Compare the noise performance of the non-inverting amplifier of Figure 2.3.6a to the inverting amplifier of Figure 2.3.7a.

Solution:

The best way to proceed here is to make a table and compare the noise performance with various gains.

Table 2.3.2 shows only a small difference in equivalent input noise for the two amplifiers. There is, however, a large difference in the flexibility of the two amplifiers. The gain of the inverting amplifier is a function of its input resistance, R_2. Thus, for a given gain and input resistance, R_1 is fixed. This is not the case for the non-inverting amplifier. The designer is free to pick R_1 and R_2 independent of the amplifier's input impedance. Thus in the case of unity gain, where $R_2 = \infty$, R_1 can be zero ohms. The equivalent input noise is:

$$e_N = \sqrt{e_n{}^2 + 4k\,T\,R_s + i_r{}^2\,R_s{}^2}$$

$$e_N = 10.3\,\text{nV}/\sqrt{\text{Hz}}$$

There is now a large difference in the noise performance of the two amplifiers. Table 2.3.2 also shows that *the equivalent input noise for practical feedback can change as a function of closed loop gain A_V. This result is somewhat different from the case of ideal feedback.*

Example 2.3.4

Determine the signal-to-noise ratio for the amplifier of Example 2.3.2 if e_{SIG} has a nominal value of 100mV.

Solution:

Signal to noise ratio is defined as:

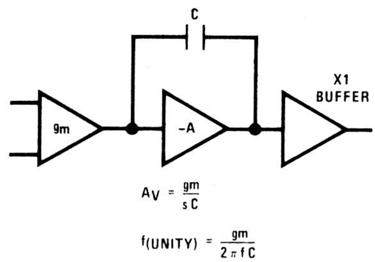
$$S/N = 20\log\frac{e_{SIG}}{e_N} \qquad (2.3.9)$$

$$= 20\log\frac{100\,\text{mV}}{1.68\,\mu\text{V}} = 95.5\,\text{dB}$$

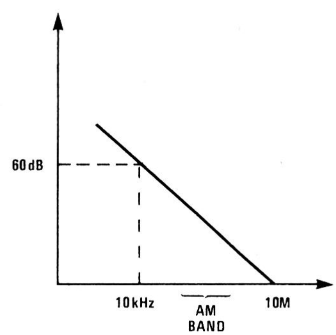
2.3.10 RF Precautions

A source of potential RF interference that needs to be considered in AM radio applications lies in the radiated wideband noise voltage developed at the speaker terminals. The method of amplifier compensation (Figure 2.3.8a) fixes the point of unity gain cross at approximately 10MHz (Figure 2.3.8b). A wideband design is essential in achieving low distortion performance at high audio frequencies, since it allows adequate loop-gain to reduce THD. (Figure 2.3.8b shows that for a closed-loop gain of 34dB there still exists 26dB of loop-gain at 10kHz.)

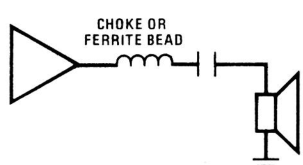
The undesirable consequence of a single-pole roll-off, wideband design is the excess gain beyond audio frequencies, which includes the AM band; hence, noise of this frequency is amplified and delivered to the load where it can radiate back to the AM (magnetic) antenna and sensitive RF circuits. A simple and economical remedy is shown in Figure 2.3.8c, where a ferrite bead, or small RF choke is added in series with the output lead. Experiments have demonstrated that this is an effective method in suppressing the unwanted RF signals.



(a) Typical Compensation



(b) Source of RF Interference

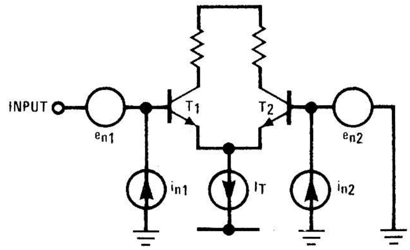


(c) Reduction of RF Interference

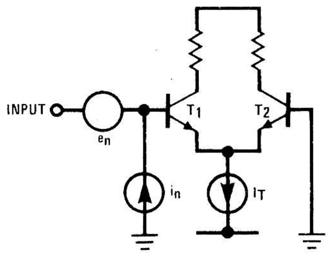
FIGURE 2.3.8

2.3.11 Noise in the Differential Pair

Figure 2.3.9a shows a differential amplifier with noise generators e_{n1}, i_{n1}, e_{n2}, and i_{n2}.



(a) Differential Pair with Noise Generators



(b) Differential Pair with Generators Input Referred

FIGURE 2.3.9

To see the intrinsic noise of the pair, short the base of T_2 to ground, and refer the four generators to an input noise voltage and noise current as shown in Figure 2.3.9b. To determine e_n, short the input of 9(a) and 9(b) to ground. e_n is then the series combination of e_{n1} and e_{n2}. These add in an RMS fashion, so:

$$e_n = \sqrt{e_{n1}{}^2 + e_{n2}{}^2}$$

Both generators contribute the same noise, since the transistors are similar and operate at the same current; thus, $e_n = \sqrt{2}\, e_{n1}$, i.e., 3dB more noise than a single ended amplifier. This can be significant in critical noise applications (see Section 2.7).

In order to find the input noise current generator, i_n, open the input and equate the output noise from Figure 2.3.9a and Figure 2.3.9b. The result of this operation is $i_n = i_{n1}$. Thus, from a high impedance source, the differential pair gives similar noise current as a single transistor.

2.3.12 Noise Measurement Techniques

This section presents techniques for measuring e_n, i_n, and e_N. The method can be used to determine the spectral density of noise, or the noise in a given bandwidth. The circuit for measuring the noise of an LM387 is shown in Figure 2.3.10.

The system gain, V_{OUT}/e_n, of the circuit in Figure 2.3.10 is large — 80dB. This large gain is required since we are trying to measure *input referred noise generators* on the order of $5\text{nV}/\sqrt{\text{Hz}}$, which corresponds to $50\mu\text{V}/\sqrt{\text{Hz}}$ at the output. R_1 and R_2 form a 100:1 attenuator to provide a low input signal for measuring the system gain. The gain should be measured in both the e_n and i_n positions, since LM387 has a 250k bias resistor which is between input and ground. The LM387 of Figure 2.3.10 has a closed loop gain of 40dB which is set by feedback elements R_5 and R_6. 40dB provides adequate gain for the input referred generators of the LM387. The output noise of the LM387 is large compared to the input referred generators of the LM381; consequently, noise at the output of the LM381 will be due to the LM387. To measure the noise voltage e_n, and noise current $i_n \times R_3$, a wave analyzer or noise filter set is connected. In addition the noise in a given bandwidth can be measured by using a bandpass filter and an RMS voltmeter. If a true RMS voltmeter is not available, an average responding meter works well. When using an average responding meter, the measured noise must be multiplied by 1.13 since the meter is calibrated to measure RMS *sine waves*. The meter used for measuring noise should have a crest factor (ratio of peak to RMS value) from 3 to 5, as the peak to RMS ratio of noise is on that order. Thus, if an average responding meter measures 1mV of noise, the RMS value would be 1.13mV_{RMS}, and the peak-to-peak value observed on an oscilloscope could be as high as 11.3mV (1.13mV x 2 x 5).

Some construction tips for the circuit of Figure 2.3.10 are as follows:

1. R_4 and R_6 should be metal film resistors, as they exhibit lower excess noise than carbon film resistors.
2. C_1 should be large, to provide low capacitive reactance at low frequency, in order to accurately observe the 1/f noise in e_n.

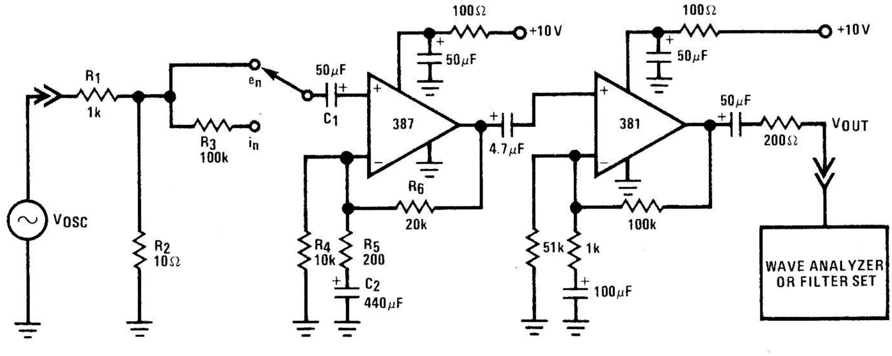


FIGURE 2.3.10 Noise Test Setup for Measuring e_n and i_n of an LM387

3. C_2 should be large to maintain the gain of 80 dB down to low frequencies for accurate 1/f measurements.
4. The circuit should be built in a small grounded metal box to eliminate hum and noise pick-up, especially in i_n.
5. The LM387 and LM381 should be separated by a metal divider within the metal box. This is to prevent output to input oscillations.

Typical LM387 noise voltage and noise current are plotted in Figure 2.3.11.

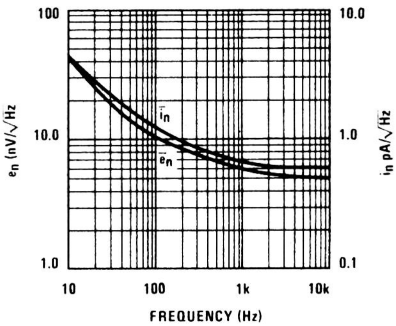


FIGURE 2.3.11 LM387 Noise Voltage and Noise Current

Many times we do not care about the actual spectral distribution of noise, rather we want to know the noise voltage in a given bandwidth for comparison purposes. For audio frequencies, we are interested only in a 20 kHz bandwidth. The noise voltage is often the dominant noise source since many systems use a low impedance voltage drive as the signal. For this common case we use a test set-up as shown in Figure 2.3.12.

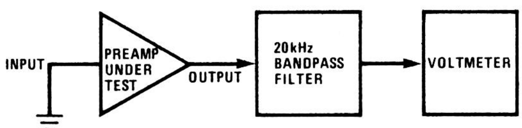


FIGURE 2.3.12 Test Setup for Measuring Equivalent Input Noise for a 20 kHz Bandwidth

Example 2.3.5

Determine the equivalent input noise voltage for the preamp of Figure 2.3.12. The gain, A_V, of the preamp is 40 dB and the voltmeter reads 0.2 mV. Assume the voltmeter is average responding and the 20 kHz low-pass filter has a single R-C roll-off.

Solution:

Since the voltmeter is average responding, the RMS voltage is V_{RMS} = 0.2 mV x 1.13 = 0.226 mV. Using an average responding meter causes only a 13% error. The filter has a single R-C roll-off, so the noise bandwidth is $\pi/2$ x 20 kHz = 31.4 kHz, i.e., the true noise bandwidth is 31.4 kHz and not 20 kHz. Since RMS noise is related to the square root of the noise bandwidth, we can correct for this difference:

$$V_{OUT} = \sqrt{\frac{0.226}{\pi/2}} = 0.18\,\text{mV}$$

The equivalent input noise is:

$$\frac{V_{OUT}}{A_V} = \frac{0.18\,\text{mV}}{100} = 1.8\,\mu\text{V in a 20 kHz bandwidth.}$$

If this preamp had an NAB or RIAA playback equalization, the output noise, V_{OUT}, would have been divided by the gain at 1 kHz.

Typical values of noise, measured by the technique of Figure 2.3.12, are shown in Table 2.3.3. For this data, B = 10 kHz and R_s = 600 Ω.

TABLE 2.3.3 Typical Flat Band Equivalent Input Noise

Type	e_N (μV)
LM381	0.70
LM381A	0.50
LM382	0.80
LM387	0.80
LM387A	0.65

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