

## Operational Amplifiers

### Noise Calculations of Op-Amp Circuits

#### Abstract

This application note describes calculating amplifier circuit noise by using the noise parameters from the op-amp datasheet; for engineers new to the subject of op-amp noise, the document guides you in calculating the possible output noise for common amplifier circuits.

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#### Related Literature

For a full list of related documents, visit our website:

[ISL28136](#), [ISL28218](#) device pages

## 1. Introduction

Due to its random nature, noise is computed using root-mean-square (rms) values of noise voltages and noise currents. However, op-amp data sheets commonly specify noise voltage and current in the form of spectral densities. This application note provides the necessary equations to convert noise spectral densities into rms noise.

To distinguish between signal and noise quantities as well as between rms noise and spectral densities, this document uses the following nomenclature:

$V_S$  = rms signal voltage

$E_n$  = rms noise voltage

$e_n$  = noise voltage spectral density

$I_n$  = rms noise current

$i_n$  = noise current spectral density

## 2. Inherent Circuit Noise

An op-amp circuit exhibits internal noise, randomly, because (1) the random generation and recombination of electron-hole pairs in semi-conductors and (2) the thermal agitation of electrons in resistors. From thermal agitation, each vibrating electron inside a resistor constitutes a minuscule current; these currents add up to a net current and, therefore, a net voltage. This voltage, although zero on average, is constantly fluctuating because of the random distribution of the instantaneous magnitudes and directions of the individual currents. As a result, each node voltage and branch current in a circuit is constantly fluctuating around its desired nominal value.

### 2.1 Signal-to-Noise Ratio (SNR)

Noise degrades the quality of a signal by posing a limit on the size of the signal that can be successfully detected. A measure of specifying the signal quality in the presence of noise is the signal-to-noise ratio (SNR). The SNR is commonly defined as the ratio of the rms signal voltage, to the rms noise voltage in dB:

$$(EQ. 1) \quad SNR = 20 \log_{10} \left( \frac{V_S}{E_n} \right) = 20 \log_{10} \left( \frac{\text{rms Signal voltage}}{\text{rms Noise voltage}} \right)$$

In sensor applications, voltage amplifiers amplify small sensor signals to a level where they can be processed by analog-to-digital converters (ADCs). In this case, you must know the SNR because it determines the required resolution of the ADC. Evidently, the lower the SNR value, the lower the ADC resolution, and the more difficult it is to rescue the useful signal from noise.

### 2.2 Noise Gain

In op-amp circuits, noise is commonly assigned to the non-inverting op-amp input, so for the inverting amplifier configuration, the noise gain differs from the signal gain; for the non-inverting amplifier, both gains are identical ([Figure 1](#)).

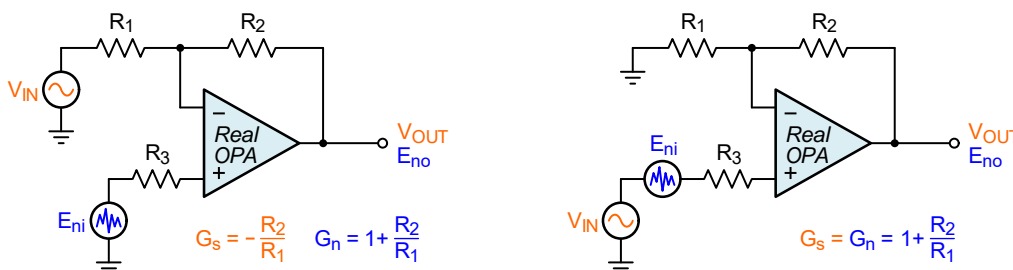


Figure 1. Noise and Signals Gains for the Inverting and Non-Inverting Amplifier

### 2.3 Noise Summation

An amplifier circuit contains multiple noise sources, generated by the op-amp and its surrounding resistors. Due to the random nature of noise, finding the total circuit input requires adding the individual noise sources quadratically:

$$(EQ. 2) \quad E_n = \sqrt{E_{n1}^2 + E_{n2}^2 + \dots + E_{ni}^2}$$

and

$$(EQ. 3) \quad I_n = \sqrt{I_{n1}^2 + I_{n2}^2 + \dots + I_{ni}^2}$$

## 3. Receiver Functional Principle

Op-amp noise is modeled by a noiseless op-amp equipped with two equivalent noise sources at the input: a voltage source with spectral density  $e_n$  and a current source with density (Figure 2). Op-amp noise is a mixture of  $1/f$  noise and broadband or white noise. The spectral densities of  $1/f$  noise, denoted as  $e_{nf}$  and  $i_{nf}$ , decline with frequency at a rate of 0.5dec/dec. The spectral densities of white noise, denoted as  $e_{nw}$  and  $i_{nw}$ , remain constant versus frequency (Figure 3).

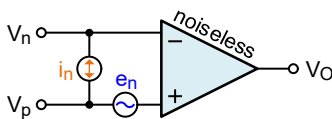


Figure 2. Op-Amp Noise Model

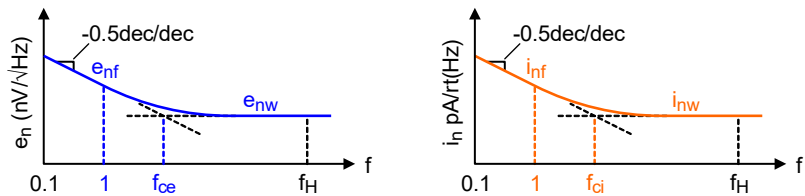


Figure 3. Voltage and Current Noise Spectral Densities

To find the total rms input noise of an op-amp for a given bandwidth, the spectral densities in Figure 3 must be converted into the rms noise voltage and current using Equation 4 and Equation 5:

$$(EQ. 4) \quad E_n = e_{nw} \sqrt{f_{ce} \cdot \ln \frac{f_H}{f_L} + n \cdot f_H - f_L}$$

and

$$(EQ. 5) \quad I_n = i_{nw} \sqrt{f_{ci} \cdot \ln \frac{f_H}{f_L} + n \cdot f_H - f_L}$$

where  $n \cdot f_H - f_L$  is the white noise equivalent bandwidth (see [White Noise Equivalent Bandwidth \(NEB\)](#)), and  $f_{ce}$  and  $f_{ci}$  are the corner frequencies, where the op-amp noise transitions from the  $1/f$  noise region to the white noise region. The corner frequencies can be calculated with:

$$(EQ. 6) \quad f_{ce} = f_x \left[ \left( \frac{e_{nf}(f_x)}{e_{nw}} \right)^2 - 1 \right]$$

and

$$(EQ. 7) \quad f_{ci} = f_x \left[ \left( \frac{i_{nf}(f_x)}{i_{nw}} \right)^2 - 1 \right]$$

where the  $e_{nf(fx)}$  and  $i_{nf(fx)}$  are the  $1/f$  spectral densities at an arbitrary frequency,  $fx$ .

**Note:** The rms noise current,  $I_n$ , is converted into an rms noise voltage,  $E_n$ , when flowing through a resistor:

(EQ. 8)

$$E_n = R \cdot I_n = R \cdot i_{nw} \sqrt{f_{ci} \cdot \ln \frac{f_H}{f_L} + n \cdot f_H - f_L}$$

### 3.1 White Noise Equivalent Bandwidth (NEB)

An interesting phenomenon occurs when running white noise through a first order low-pass filter with  $f_H$  as its -3dB frequency. In this case, the noise behaves as if filtered by a brick-wall filter with a higher cutoff frequency of  $f_c = 1.57f_H$  (Figure 4).

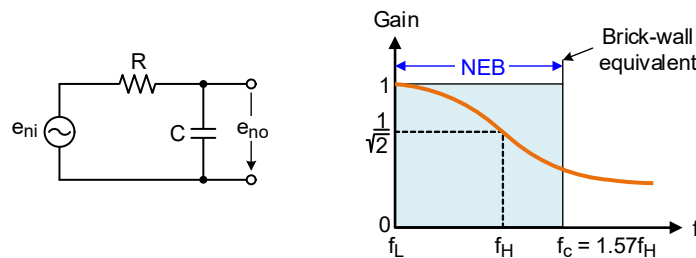


Figure 4. Noise Equivalent Bandwidth (NEB) for 1st Order Low-Pass Filter

This increased noise bandwidth is known as the white noise equivalent bandwidth (NEB). With increasing filter order,  $f_c$  decreases and approaches  $f_H$ , because the steeper roll-off of higher filter orders approaches the brick-wall equivalent. Mathematically, the NEB is expressed through:

(EQ. 9)

$$NEB = f_c - f_L = n \cdot f_H - f_L$$

where  $n$  is the brick-wall factor for a given filter order. The higher the filter order, the closer the value of  $n$  approaches 1. Figure 5 lists the values of  $n$  for the first five orders of passive low-pass filters.

Filter Order	n
1	1.57
2	1.22
3	1.16
4	1.13
5	1.12

Figure 5. Values of n for Higher Order Filters

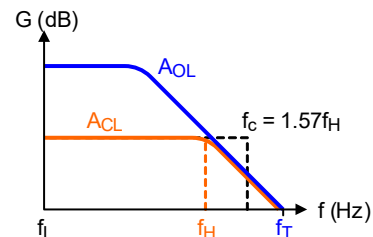


Figure 6. NEB of Amplifier with Resistive Feedback

For amplifiers with purely resistive feedback, the closed-loop gain represents a first order low-pass with a -3dB bandwidth,  $f_H$ . (Figure 6). These amplifiers pass white noise with a cutoff frequency of  $f_c = 1.57 \cdot f_H$ .

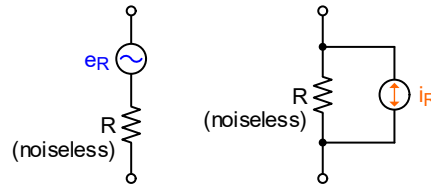
**Note:** Find  $f_H$  by taking the unity-gain bandwidth,  $f_T$ , from the op-amp data sheet and dividing it by the noise-gain,  $1/\beta$ :

(EQ. 10)

$$f_H = f_T / (1/\beta)$$

## 4. Thermal Noise in Resistors

Resistor noise is thermal noise due to the random thermal motion of electrons. It is present in standard resistors as well as in the stray resistances of practical inductors and capacitors. Resistor noise can be modeled either by a noise voltage of spectral density,  $e_R$ , in series, or by a noise current of spectral density,  $i_R$ , in parallel with a noiseless resistor ([Figure 7](#)).



**Figure 7. Thermal (Resistor) Noise Models**

The spectral densities of thermal noise are defined through:

$$(EQ. 11) \quad e_R = \sqrt{4kTR}$$

and

$$(EQ. 12) \quad i_R = \sqrt{\frac{4kT}{R}}$$

where  $k = 1.38 \cdot 10^{-23}$  J/K is Boltzmann's constant,  $T$  is the absolute temperature in kelvins, and  $R$  is the resistance in ohms.

[Equation 11](#) and [Equation 12](#) show that the spectral densities of thermal noise,  $e_R$  and  $i_R$ , are constant with frequency, therefore indicating that thermal noise is a type of white noise with its upper bandwidth limit at  $f_c = n \cdot f_H$ .

Thermal noise is commonly represented as an rms noise voltage. Whether using the  $e_R$  or  $i_R$  spectral density, the resulting rms noise voltage is always:

$$(EQ. 13) \quad E_R = \sqrt{4kTR(n \cdot f_H - f_L)}$$

**Note:** The equation holds true for resistors connected in series or parallel. For example, if two resistors are connected in series,  $R = R_1 + R_2$ , if they are parallel connected,  $R = R_1 \parallel R_2$ .

If resistors are connected in parallel, use the  $i_R$ -model. In this case, each noise current flows through the parallel resistance  $R_p = R_1 \parallel R_2$ , therefore producing the individual noise voltages.

## 5. Total RMS Input Noise, $E_{ni}$

When calculating the total rms input noise of a voltage amplifier,  $E_{ni}$ , convert the initial circuit schematic (Figure 8) into a noise equivalent circuit diagram (Figure 9) and apply the equations from the previous sections.

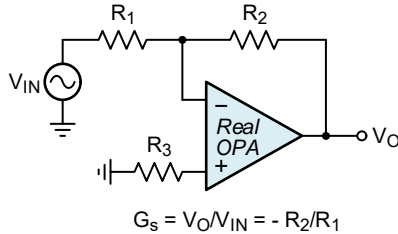


Figure 8. Inverting Amplifier

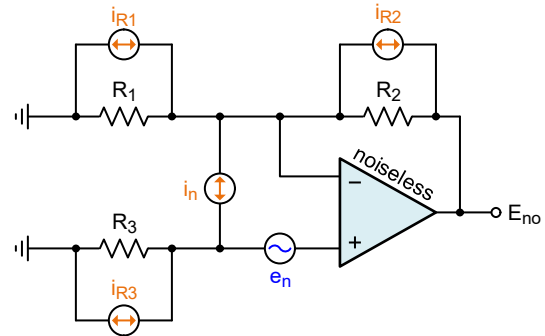


Figure 9. Spectral Noise Equivalent Circuit

Starting with the thermal noise calculation: noise-wise,  $R_1$  and  $R_2$  are parallel connected, therefore generating an rms noise voltage:

$$(EQ. 14) \quad E_{Rp} = \sqrt{4kT(1.57f_H - f_L) \cdot R_1 \parallel R_2}$$

The other thermal noise source is  $R_3$ , with the following rms value:

$$(EQ. 15) \quad E_{R3} = \sqrt{4kT(1.57f_H - f_L) \cdot R_3}$$

Next, calculate the rms voltages that are due to op-amp current noise. At the negative input, this voltage, denoted as  $E_{nn}$ , is the product of the rms noise current,  $I_n$ , and the parallel resistance  $R_1 \parallel R_2$ :

$$(EQ. 16) \quad E_{nn} = R_1 \parallel R_2 \cdot i_{nw} \sqrt{f_{ci} \cdot \ln(f_H/f_L) + 1.57f_H - f_L}$$

At the positive input, the rms noise voltage, denoted as  $E_{np}$ , is the product of  $I_n$ , and  $R_3$ :

$$(EQ. 17) \quad E_{np} = R_3 \cdot i_{nw} \sqrt{f_{ci} \cdot \ln(f_H/f_L) + 1.57f_H - f_L}$$

Finally, calculate the rms input noise of the op-amp itself with the following:

$$(EQ. 18) \quad E_n = e_{nw} \sqrt{f_{ce} \cdot \ln(f_H/f_L) + n \cdot f_H - f_L}$$

Summing up all rms noise voltages above in rms fashion yields the total input noise,  $E_{ni}$ , of the circuit:

$$(EQ. 19) \quad E_{ni} = \sqrt{E_{Rp}^2 + E_{R3}^2 + E_{nn}^2 + E_{np}^2 + E_n^2}$$

Writing [Equation 19](#) in detailed form results in a huge expression:

$$E_{ni} = \sqrt{4kT(1.57f_H - f_L) \cdot R_1 \parallel R_2 + 4kT(1.57f_H - f_L) \cdot R_3 + (R_1 \parallel R_2 \cdot i_{nw})^2 \cdot [f_{ci} \cdot \ln(f_H/f_L) + 1.57f_H - f_L] + (R_3 \cdot i_{nw})^2 \cdot [f_{ci} \cdot \ln(f_H/f_L) + 1.57f_H - f_L] + e_{nw}^2 \cdot [f_{ce} \cdot \ln(f_H/f_L) + n \cdot f_H - f_L]}$$

However, you can simplify the above equation by making the following realistic assumptions that apply to the majority of amplifier circuits:

$f_H > 10 \cdot f_{ce}$ : The upper bandwidth limit is more than 10 times the corner frequency of  $1/f$  and white noise, which makes the term neglectable, therefore leaving only  $n \cdot f_H - f_L$  under the square root.

$f_H \gg f_L$ : The upper bandwidth limit is much larger than the lower bandwidth limit of 0.1Hz typically, which makes  $f_L$  neglectable, therefore reducing the  $(n \cdot f_H - f_L)$  term to  $n \cdot f_H$ .

$n = 1.57$ : Due to a 1<sup>st</sup> order low-pass response, either because of purely resistive feedback, or a simple output filter.

$R_3 = R_1 \parallel R_2 = R_p$ : To minimize the offset due to bias current.

Applying the above assumptions and after factoring out and collecting terms, results in the following:

$$(EQ. 20) \quad E_{ni} = \sqrt{1.57f_H} \cdot \sqrt{8kTR_p + 2R_p^2 i_{nw}^2 + e_{nw}^2}$$

Next, multiplying  $E_{ni}$  with the noise gain yields the total rms output noise,  $E_{no}$ , of the circuit:

$$(EQ. 21) \quad E_{no} = E_{ni} \cdot (1 + R_2/R_1)$$

## 6. Calculation Example 1

Calculate the output signal-to-noise ratio of an inverting amplifier using the op-amp ISL28136.

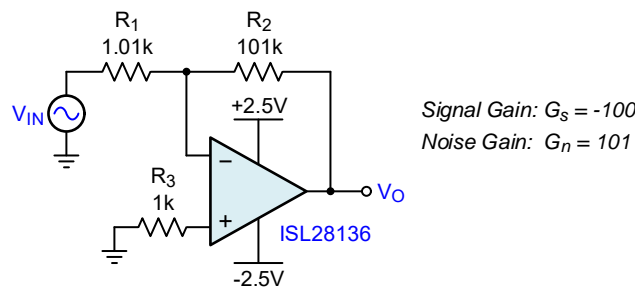


Figure 10. Inverting Amplifier with ISL28136

From the AC specification in the op-amp datasheet we take the following data:

GBW = 5MHz,  $e_{nw} = 15nV/\sqrt{Hz}$ ,  $i_{nw} = 0.35pA/\sqrt{Hz}$ . The corner frequencies of the voltage and current noise densities are calculated with [Equation 6](#) and [Equation 7](#), yielding  $f_{ce} = 30Hz$  and  $f_{ci} = 521Hz$  (see [Figure 11](#)):

$$f_{ce} = 10 \cdot \left[ \left( \frac{30}{15} \right)^2 - 1 \right] = 30\text{Hz}$$

$$f_{ci} = 1 \cdot \left[ \left( \frac{8}{0.35} \right)^2 - 1 \right] = 521\text{Hz}$$

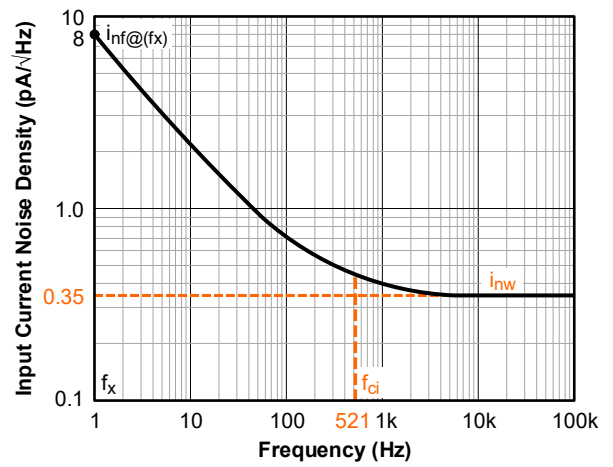
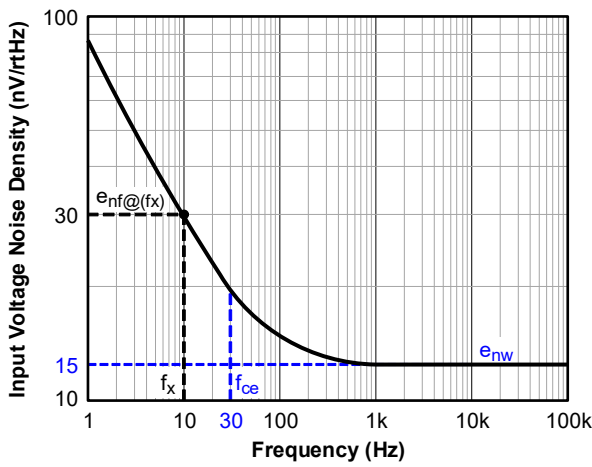


Figure 11. Inverting Amplifier with ISL28136

Table 1. Detailed Calculations

Bandwidth Limits	
$f_L = 0.1\text{Hz}$	Assumed lower bandwidth limit
$f_H = \frac{\text{GBW}}{G_n} = \frac{5\text{MHz}}{101} = 49.5\text{kHz}$	Upper bandwidth limit for 1/f noise
$\text{NEB} = 1.57 \cdot f_H - f_L = 77.7\text{kHz}$	White noise equivalent bandwidth
Thermal Noise Contribution	
$E_{Rp} = \sqrt{4kT \cdot \text{NEB} \cdot R_1 \parallel R_2}$ $E_{Rp} = \sqrt{4 \cdot 1.38 \cdot 10^{-23} \cdot 298\text{K} \cdot 77.7\text{kHz} \cdot 1\text{k}\Omega} = 1.13\mu\text{V}$	Thermal noise of $R_1$ and $R_2$ at $T = 298$ Kelvin ( $25^\circ\text{C}$ ) using <a href="#">Equation 13</a>
$E_{R3} = E_{Rp} = 1.13\mu\text{V}$	Thermal noise of $R_3$ is the same as in $R_1 \parallel R_2$ due to $R_3 = R_1 \parallel R_2$ .
Noise Voltage due to Op-amp Current Noise	
$E_{nn} = R_p \cdot i_{nw} \sqrt{f_{ci} \cdot \ln(f_H/f_L) + \text{NEB}}$ $E_{nn} = 1\text{k}\Omega \cdot 0.35 \frac{\text{pA}}{\sqrt{\text{Hz}}} \cdot \sqrt{521\text{Hz} \cdot \ln\left(\frac{49.5\text{kHz}}{0.1\text{Hz}}\right) + 77.7\text{kHz}} = 102\text{nV}$	Voltage noise due to op-amp noise current flowing through $R_1 \parallel R_2$ using <a href="#">Equation 8</a>
$E_{np} = E_{nn} = 102\text{nV}$	Voltage noise due to op-amp noise current flowing through $R_3$ is the same as in $R_1 \parallel R_2$ due to $R_3 = R_1 \parallel R_2$ .
Op-amp Input Noise Voltage	
$E_n = e_{nw} \sqrt{f_{ce} \cdot \ln(f_H/f_L) + \text{NEB}}$ $E_n = 15 \frac{\text{nV}}{\sqrt{\text{Hz}}} \cdot \sqrt{30\text{Hz} \cdot \ln\left(\frac{49.5\text{kHz}}{0.1\text{Hz}}\right) + 77.7\text{kHz}} = 4.192\mu\text{V}$	RMS value of op-amp voltage noise using <a href="#">Equation 4</a> and <a href="#">Equation 5</a>



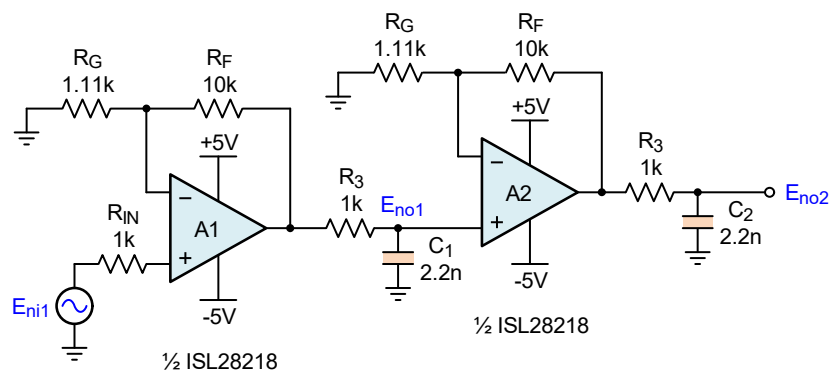
**Table 1. Detailed Calculations (Continued)**

<b>Total Input Noise per Equation 19</b>	
$E_{ni} = \sqrt{E_{Rp}^2 + E_{R3}^2 + E_{nn}^2 + E_{np}^2 + E_n^2} = \sqrt{2E_{Rp}^2 + 2E_{nn}^2 + E_n^2}$ $E_{ni} = \sqrt{(1.599\mu\text{V})^2 + (0.144\mu\text{V})^2 + (4.192\mu\text{V})^2} = 4.489\mu\text{V}$	Total RMS Input Noise if Equation 19 is applied with $R_3 = R_1 \parallel R_2$ .
<b>Total Input Noise using simplified Equation 20</b>	
$E_{ni} = \sqrt{NEB \cdot (8kTR_p + 2R_p^2 i_{nw}^2 + e_{nw}^2)}$ $E_{ni} = \sqrt{77.7\text{kHz} \cdot [32.9 \cdot 10^{-18} + 245 \cdot 10^{-21} + 225 \cdot 10^{-18}]} = 4.479\mu\text{V}$	Total RMS Input Noise if simplified Equation 20 is applied. Error is only 0.23%.
<b>Total Output Noise</b>	
$E_{no} = E_{ni} \cdot G_n = 4.489\mu\text{V} \cdot 101 = 453.4\mu\text{V}$	Total RMS Output Noise
<b>Output SNR</b>	
$\text{SNR} = 20 \cdot \log_{10} \left( \frac{V_{O-pp}}{2\sqrt{2} E_{no}} \right) = 20 \cdot \log_{10} \left( \frac{4.6\text{V}}{2\sqrt{2} \cdot 453.4\mu\text{V}} \right) = 71\text{dB}$	Output SNR for Output FSR = 4V <sub>pp</sub>

The dominant noise source in the highlighted  $E_{ni}$  term is the  $4.2\mu\text{V}$ , which is caused by the high input noise voltage,  $e_{nw}$ , of the op-amp. A rule-of-thumb states that if a dominant noise source is three-times higher than the other noise sources in the system, the other noise sources can be neglected, and noise reduction efforts should focus on the dominant noise source. In other words, to get a higher SNR, you need an op-amp with lower voltage noise. When this is achieved, further noise reduction can be obtained by reducing the resistor values.

## 7. Calculation Example 2

To reduce the output noise from the previous circuit example without narrowing the overall bandwidth, select the dual low-noise op-amp ISL28218 for a two-stage amplifier design. This allows you to use symmetrical lower gains, and each stage has a gain of 10 and also lower resistor values.

**Figure 12. Two-Stage Amplifier with ISL28218**

The key parameters of the ISL28218 are:  $GBW = 3.2\text{MHz}$ ,  $e_{nw} = 5.5\text{nV}/\sqrt{\text{Hz}}$ ,  $i_{nw} = 0.38\text{pA}/\sqrt{\text{Hz}}$ .

The closed-loop bandwidth of each stage is:

$$(EQ. 22) \quad f_{Acl} = \frac{GBW}{G_n} = \frac{3.2\text{MHz}}{10} = 320\text{kHz}$$

The external low-pass filters however, provide an earlier cutoff at:

$$(EQ. 23) \quad f_c = \frac{1}{2\pi RC} = \frac{1}{2\pi \cdot 1k\Omega \cdot 2.2nF} = 72.4kHz = f_H$$

And therefore, determine the equivalent noise bandwidth, NEB.

**Note:** The NEB of the two gain stages differ, because the noise of the first stage experiences two low-pass functions (of Filter 1 and Filter 2), while the second stage sees only one low-pass. The NEB is therefore:

$$(EQ. 24) \quad NEB_1 = 1.22 \cdot f_H = 1.22 \cdot 72.4kHz = 88.3kHz$$

The 2<sup>nd</sup> stage faces only one low-pass function, therefore making the NEB:

$$(EQ. 25) \quad NEB_2 = 1.57 \cdot f_H = 1.57 \cdot 72.4kHz = 113.6kHz$$

To reduce the mathematical effort, apply [Equation 20](#) using to calculate the rms input noise for each stage under the condition that:

$$(EQ. 26) \quad E_{ni1} = \sqrt{NEB_1 \cdot (8kTR_P + 2R_P^2 i_{nw}^2 + e_{nw}^2)} = 2.367\mu V$$

$$(EQ. 27) \quad E_{ni2} = \sqrt{NEB_2 \cdot (8kTR_P + 2R_P^2 i_{nw}^2 + e_{nw}^2)} = 2.685\mu V$$

Each input noise contributes individually to the overall output noise. We denote  $E_{no21}$  as the output noise due to  $E_{ni1}$ , and  $E_{no22}$ , as the output noise due to  $E_{ni2}$ :

$$(EQ. 28) \quad E_{no21} = E_{ni1} \cdot G_{n1} \cdot G_{n2} = 2.367\mu V \cdot 10 \cdot 10 = 236.7\mu V$$

$$(EQ. 29) \quad E_{no22} = E_{ni2} \cdot G_{n2} = 2.685\mu V \cdot 10 = 26.85\mu V$$

The total rms output noise is:

$$(EQ. 30) \quad E_{no} = \sqrt{E_{no1}^2 + E_{no2}^2} = 238.2\mu V$$

While maintaining the same signal bandwidth, of about 48kHz, we have reduced the rms output noise from 453 $\mu$ V down to 238 $\mu$ V. Assuming the same 4V<sub>pp</sub> full scale voltage range at the output, the new SNR is:

$$(EQ. 31) \quad SNR = 20 \cdot \log_{10} \left( \frac{V_{O-pp}}{2\sqrt{2} E_{no}} \right) = 20 \cdot \log_{10} \left( \frac{4.6V}{2\sqrt{2} \cdot 238.2\mu V} \right) = 76.7dB$$

which is an improvement of nearly 6dB, or twice the resolution.

---

## 8. Conclusion

Noise calculations can be tedious. Applying correct, realistic assumptions drastically reduces the mathematical effort. While this application note is not a replacement for the vast amount of technical documentation on noise, this document helps engineers new to the subject of op-amp noise quickly assess the possible output noise for common amplifier circuits.

## 9. Revision History

Rev.	Date	Description
1.00	Aug.4.20	Initial release

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