11 Myths About Analog Noise Analysis

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Abstract

Noise is a central topic in analog circuit design, directly affecting how much information can be extracted from a measurement as well as the economy with which the required information can be obtained. Unfortunately, there is a large amount of confusion and misinformation regarding noise, which has the potential to cause underperformance, costly overdesign, or inefficienct use of resources. This article addresses 11 of the most persistent myths about noise analysis in analog designs.

1. Decreasing the Resistor Values in the Circuit Always Improves the Noise Performance

It is a well-known relationship that noise voltage increases with higher resistor values according to the Johnson noise equation, $e_{rms} = \sqrt{4kTRB}$, where e_{rms} is the rms voltage noise, k is Boltzmann's constant, T is temperature in Kelvin, R is the resistance, and B is the bandwidth. This leads many engineers to the conclusion that resistor values should be reduced in order to reduce the noise. Although this is often true, it cannot be assumed because there are specific examples where larger resistors improve the noise performance. As an example, in most cases, current is measured by passing it through a resistor and measuring the resulting voltage. The voltage developed is proportional to the resistor value according to Ohm's law, V = I × R, but as shown above, the Johnson noise of the resistor is proportional to the square root of the resistor value. Because of this relationship, a 3 dB improvement in the signal-to-noise ratio can be achieved each time the resistor value is doubled. This trend continues right up to the point where the voltage developed is too large or the power dissipated is too high.

2. The Noise Spectral Density of All Noise Sources Can Be Added Up and the Bandwidth Can Be Taken into Account at the End of the Calculation

It can save time to combine the noise spectral density (nV/√Hz) of multiple noise sources (voltage noise sources are combined as the root sum of squares) rather than computing the rms noise of each noise source separately, but this simplification is only applicable if the bandwidth seen by each noise source is the same. It becomes a dangerous trap if the bandwidths seen by each of the noise sources are different. Figure 1 shows the implications in an oversampled system. It would appear from the noise spectral density that the gain amplifier will dominate the total noise of the system, but once the bandwidth is taken into account, the rms noise contributed by each stage is very similar.

3. It Is Important to Include Every Noise Source in Hand Calculations

It may be tempting to consider every noise source in a design, but a designer's time is valuable and this can be very time consuming in large designs. Comprehensive noise calculations are best left to simulation software. But how does a designer simplify the hand noise calculations needed during the design process? Ignore minor noise sources below a certain threshold. If a noise source is $^1/_5$ e_{rms} value of the dominant noise source (or any other noise source referred to the same point), it contributes less than 2% to the total noise and can reasonably be ignored. Designers argue about where to draw the threshold below which it is not necessary to consider a noise source, but whether that level is $^1/_3$, $^1/_5$, or $^1/_{10}$ (which adds 5%, 2%, and 0.5% to the total noise, respectively), it is not worth worrying about smaller noise sources than that until the design is fixed enough to simulate or calculate fully.
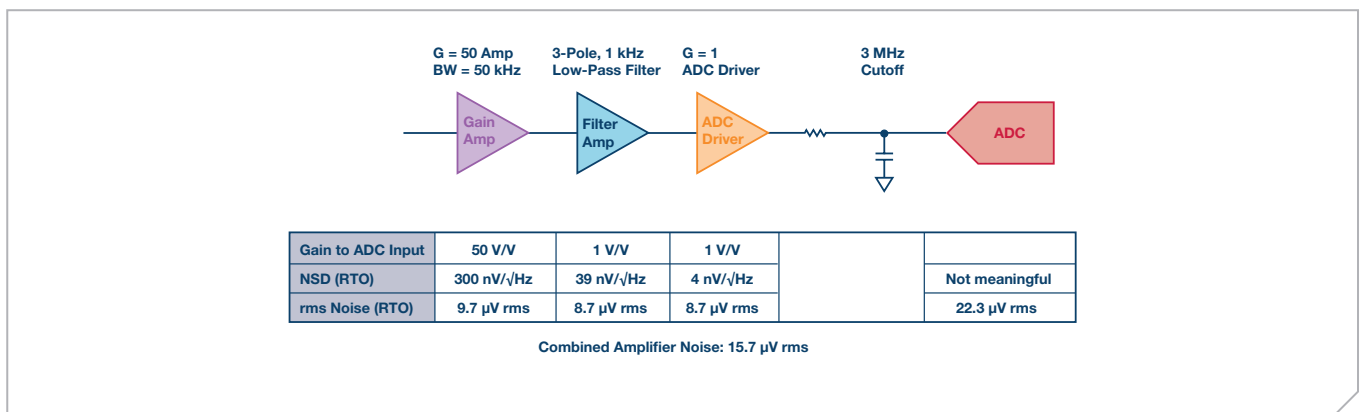

	G = 50 Amp BW = 50 kHz	3-Pole, 1 kHz Low-Pass Filter	G = 1 ADC Driver		3 MHz Cutoff	
Gain to ADC Input	50 V/V	1 V/V	1 V/V			
NSD (RTO)	300 nV/√Hz	39 nV/√Hz	4 nV/√Hz			Not meaningful
rms Noise (RTO)	9.7 µV rms	8.7 µV rms	8.7 µV rms			22.3 µV rms

Combined Amplifier Noise: 15.7 µV rms

Figure 1. Justification for using rms noise rather than spectral density for noise calculations.

4. Pick an ADC Driver With $1/10$ the Noise of the ADC

Analog-to-digital converter (ADC) data sheets may suggest driving the analog input with a low noise ADC driver amplifier that has something like $1/10$ the noise of the ADC. However, this is not always the best choice. In a system, it is often worth examining the trade-off of the ADC driver noise from a system level.

First, if the noise sources in the system preceding the ADC driver are much larger than the ADC driver noise, then choosing a very low noise ADC driver will not provide any system benefit. In other words, the ADC driver noise should be commensurate with the rest of the system.

Secondly, even in the simple case where there is just an ADC and an amplifier to drive it, it may still be advantageous to examine the noise trade-off and to determine the effects on the system. The reason for this can be made clearer with a numerical example. Consider a system that uses a 16-bit ADC with an SNR value that equates to 100 µV rms noise, and an amplifier with 10 µV rms noise as the ADC driver. The total noise when these sources are combined as the root sum of squares is 100.5 µV rms, very close to the noise of the ADC alone. The following two options that bring the amplifier and ADC into closer balance can be considered as well as the effects on system performance. If the 16-bit ADC is replaced with a similar 18-bit ADC that specifies SNR equivalent to 40 µV rms noise, the total noise would change to 41 µV rms. Alternatively, if the 16-bit ADC is retained, but the driver is replaced with a lower power amplifier that contributes 30 µV rms noise, the total noise would be 104 µV rms. One of these trade-offs may be a better choice for system performance than the original combination. It's just a matter of evaluating the trade-offs and their effects on the overall system.

5. 1/f Noise Must Always Be Considered in DC-Coupled Circuits

1/f noise is a menace to very low frequency circuits because it defies many of the usual noise rejection techniques like low-pass filtering, averaging, and long integrations. However, many dc circuits are dominated by white noise sources to the point where it is not useful to calculate the 1/f noise because it does not add to the total noise. To see this effect, consider an amplifier with a 1/f noise corner, f_{nc}, at 10 Hz and a wideband noise of 10 nV/$\sqrt{Hz}$. The noise in a 10 sec acquisition is computed for various bandwidths with and without the 1/f noise to determine the effect of leaving it out. In this case, wideband noise begins to dominate when the bandwidth is 100 times f_{nc}, and 1/f noise is not significant when the bandwidth is more than 1000 times f_{nc}. Good, modern bipolar amplifiers can have noise corners well below 10 Hz, and zero-drift amplifiers virtually eliminate 1/f noise altogether.

Table 1. Example of the Effect of 1/f Noise vs. Circuit Bandwidth

BW (Hz)	BW/f_{nc}	Wideband (nV rms)	1/f noise (nV rms)	Total Noise (nV rms)	Increase due to 1/f
100	10	100	220	240	140%
300	30	170	250	310	77%
1000	100	320	290	430	36%
3000	300	550	330	640	16%
10000	1000	1k	360	1.1k	6%
30000	3000	1.7k	400	1.8k	3%
100000	10000	3.2k	440	3.2k	1%

6. Since the 1/f Noise Increases at Lower Frequencies, DC Circuits Have Infinite Noise

Although dc is a useful concept for circuit analysis, the truth is that if dc is considered to be operating at 0 Hz, then there really is no such thing. As the frequency gets lower and lower, approaching 0 Hz, the period gets longer and longer, approaching infinity. The implication is that there is a minimum frequency that can be seen, even in a circuit that theoretically responds to dc. This minimum frequency depends on the length of the acquisition, or the aperture time, which is how long the output of the device is being watched. If an engineer turns on a device and watches the output for 100 seconds, the lowest frequency artifact they could observe would be 0.01 Hz. This also means the lowest frequency noise that can be observed in this case is 0.01 Hz as well.

To extend this with a numerical example, consider a dc to 1 kHz circuit where the output is monitored continuously. If a certain amount of 1/f noise is observed in the circuit in the first 100 seconds, from 0.01 Hz to 1 kHz (5 decades of frequency), then the amount of noise observed in 30 years, which is about 1 nHz (12 decades), can be calculated as $\sqrt{12/5} = 1.55$, or 55% more noise than was observed in the first 100 seconds. This somewhat banal increase even assumes the worst case: that 1/f noise continues to increase down to 1 nHz, for which there is, so far, no measured evidence. In theory, when the aperture time is not well-defined, the 1/f noise could be calculated down to a frequency equal to one over the lifetime of the circuit. In practice, these very long timeline variations are dominated by aging effects and long-term drift rather than 1/f noise. Many engineers set a minimum frequency such as 0.01 Hz or 1 mHz for noise calculations in dc circuits in order to keep the calculations practical.

7. The Noise Equivalent Bandwidth Is a Multiplier For the Noise

The noise equivalent bandwidth (NEB) is a useful simplification for noise calculations. Some noise from beyond the bandwidth of the circuit is able to get into the circuit because the gain above the cutoff frequency is not zero. The NEB is the cutoff frequency of a calculated ideal brick wall filter that would let in the same amount of noise that the actual circuit does. The NEB is larger than the –3 dB bandwidth and it has been calculated for common filter types and orders, for example, it is 1.57× larger than the –3 dB bandwidth for a 1-pole, low-pass filter or, in equation form, $NEB_{1-pole} = 1.57 \times BW_{3dB}$. However, there seems to be consistent confusion about where to put that multiplication factor in the noise equation. Remember that the NEB is an adjustment for the bandwidth, not the noise, so it goes under the square root, as follows:

$$e_{rms} = NSD \times \sqrt{NEB_{1-pole}} = NSD \times \sqrt{1.57 \times BW_{3dB}}$$

8. The Amplifier with the Lowest Voltage Noise Is the Best Choice

When choosing an op amp, the voltage noise is often the only noise specification considered by the designer. It is important not to overlook the current noise as well. Except in special cases such as input bias current compensation, the current noise is typically the shot noise of the input bias current: $i_n = \sqrt{2 \times q \times I_B}$. The current noise is converted to a voltage via the source resistance, so when there is a large resistance in front of the amplifier input, the current noise can be a larger noise contributor than the voltage noise. The typical case where current noise is a problem is when a low noise op amp is used with a large resistance in series with the input. For example, consider the ADA4898-1 low noise op amp with a 10 kΩ resistor in series with the input. The voltage noise of the ADA4898-1 is 0.9 nV/$\sqrt{Hz}$, the 10 kΩ resistor has 12.8 nV/$\sqrt{Hz}$, and the 2.4 pA/$\sqrt{Hz}$ current noise times the 10 kΩ resistor is 24 nV/$\sqrt{Hz}$, the largest noise source in the system. In cases like this, where the current

noise dominates, it is often possible to find a part with lower current noise and thereby reduce the noise of the system. This is especially true for precision amplifiers, but there are high speed FET input op amps that can help in high speed circuits as well. For example, instead of choosing the ADA4898-1 and not getting the benefit of the 0.9 nV/√Hz voltage noise, a JFET input amplifier such as the AD8033 or the ADA4817-1 could have been chosen.

9. The Best Noise Performance Is Achieved By Taking Lots of Gain in the First Stage

It is often suggested that the gain should be taken in the first stage for better noise performance, which is true because then the signal will be larger compared to the noise of subsequent stages. However, the drawback of taking gain is that it reduces the maximum signal that the system can accommodate. In some cases, rather than taking a large amount of gain in the first stage, which improves the sensitivity of the measurement but limits the dynamic range, it may be better to limit the amount of gain taken in the first stage and digitize with high resolution in order to maximize both sensitivity and dynamic range.

10. All Resistor Types Have the Same Noise for a Given Resistance

The Johnson noise of resistors is fundamental, giving rise to a simple equation for the noise of a certain resistor at a certain temperature. However, Johnson noise is the least amount of noise that can be observed in a resistor, and it does not mean that all resistor types are created equal with respect to noise. There is also excess noise, which is a source of 1/f noise in resistors that is highly dependent on the resistor type. Excess noise, somewhat confusingly also called current noise, is associated with the way current flows in a discontinuous medium. It is specified as a noise index (NI) in dB referred to 1 µV rms/V_{dc} per decade. This means that if there is 1 V_{dc} across a resistor with a 0 dB NI, the excess noise in a given frequency decade is 1 µV rms. Carbon and thick film resistors have some of the highest NI, ranging up to roughly +10 dB, and it is better to avoid them in noise sensitive parts of the signal path. Thin film resistors are generally much better at around −20 dB, and metal foil and wirewound resistors can be below −40 dB.

11. Given Enough Acquisitions, Averaging Reduces the Noise Indefinitely

Averaging is considered to be a way to reduce the noise by the square root of the number of averages. This is conditionally true when the NSD is flat. However, this relationship breaks in the 1/f range and in a few other cases. Consider the case of averaging in a system sampling at a constant frequency f_s, such that n samples are averaged and decimated by n, and some number m decimated samples are returned. Taking n averages moves the effective sampling rate after decimation to f_s/n, reducing the effective maximum frequency seen by the system by a factor of n and reducing the white noise by √n. However, it also took n times longer to obtain m samples, so the lowest frequency that can be seen by the system is also reduced by a factor of n (remember, there is no such thing as 0 Hz). The more averages are taken, the lower these maximum and minimum frequencies move on the frequency band. Once the maximum and minimum frequencies are both within the 1/f range, the total noise depends only on the ratio of these frequencies, so increasing the number of averages provides no further benefit to the noise. The same logic holds for long integration times for an integrating ADC such as multislope. Beyond this mathematical exercise, there are other practical limits as well. For example, if quantization noise is the dominant noise source such that the output of the ADC with a dc input voltage is a constant code with no flicker, then any number of averages will return the same code.

References

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