Common myths about tackling noise in designs

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Noise analysis is crucial in analogue circuit design. Sadly, there is a lot of misinformation about it, leading to confusion, underperformance, costly overdesign and inefficient use of resources. Here are eleven of the most persistent myths relating to noise

1 Lowering resistor values in a circuit improves noise performance

It is well-known that noise voltage increases with higher resistor values, according to the Johnson noise equation $e_{rms} = \sqrt{4kTRB}$, where e_{rms} is the rms of the noise voltage, k is Boltzmann's constant, T is the temperature in Kelvin, R is the resistance and B is bandwidth. This logically leads to the conclusion that reducing resistor values reduces noise. Although often true, this doesn't apply to all cases, since there are specific examples where larger resistors improve noise performance.

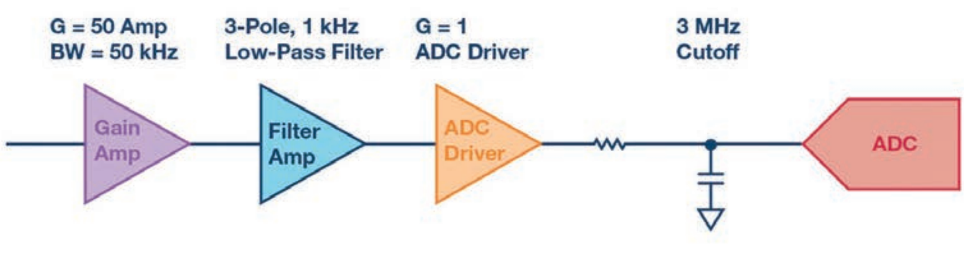
As an example, current is measured by passing it through a resistor and measuring the resulting voltage across its terminals. This voltage is proportional to the resistor value, according to Ohm's law $V = I \times R$, but as per Johnson's equation, the resistor noise is proportional to the square root of its value. Because of this relationship, a 3dB improvement in signal-to-noise ratio (SNR) can be achieved each time the resistor value is doubled. This trend continues right up to the point where the developed voltage is too large or the dissipated power too high.

2 The noise spectral density of all noise sources can be added up and the bandwidth 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! This simplification is only applicable if the bandwidth seen by each noise source is the same. It becomes a dangerous trap if they are different.

Figure 1 shows the implications in an oversampled system. From the noise spectral density (NSD) it would appear that the gain amplifier will dominate the system's total noise. Yet, once bandwidth is taken into account, the rms noise contributed by each stage is very similar.

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



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

3 Every noise source should be included in manual calculations

It may be tempting to consider every noise source in a design, but a designer's time is valuable and this process can be very time-consuming, especially in large designs. Comprehensive noise calculations are best left to simulation software.

But, how does a designer simplify the by-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 add 5%, 2% and 0.5% to the total noise, respectively), it's not worth worrying about noise sources smaller than that until the design is fixed enough to simulate or calculate fully.

4 Pick an ADC driver with a tenth the ADC noise

Analogue-to-digital converter (ADC) data sheets may suggest driving the analogue input with a low-noise ADC driver amplifier with 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 system noise sources preceding the ADC driver are much larger than the ADC driver noise, then choosing a very-low-noise ADC driver provides no system benefit. In other words, ADC driver noise should be commensurate with the rest of the system.

Secondly, even in the simple case where there's just one ADC and an amplifier to drive it, it may still be advantageous to examine the noise trade-off and 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 equivalent to $100\mu V_{rms}$ noise, and an amplifier with $10\mu V_{rms}$ noise as its driver. The total noise when these sources are combined (as the root sum of squares) is $100.5\mu V_{rms}$ – 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 $40\mu V_{rms}$ noise, the total noise would change to $41\mu V_{rms}$. Alternatively, if the 16-bit ADC is retained but the driver is replaced with a lower-power amplifier that contributes $30\mu V_{rms}$ noise, the total noise would be $104\mu V_{rms}$.

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}, at10Hz and wideband noise of 10nV/√Hz. The noise in a 10s 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 10Hz, and zero-drift amplifiers virtually eliminate 1/f noise altogether.

Table 1: Example of the effect of 1/f noise on 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 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 there is no such thing as 0Hz. As the frequency decreases to 0Hz, the period extends, approaching infinity. The implication is that there's 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 aperture time, which is how long the output of the device is being watched. If a device is turned on and its output watched for 100s, the lowest observed frequency artifact would be 0.01Hz, which means the lowest observed frequency noise is also 0.01Hz.

Here's an example: consider a DC-1kHz circuit where the output is continuously monitored. If a certain amount of 1/f noise is observed in the first 100s, from 0.01Hz to 1kHz (five decades of frequency), then the amount of noise observed in 30 years – which is about 1nHz (12 decades) – can be calculated as $\sqrt{12/5} = 1.55$, or 55% more noise than was observed in the first 100s. This somewhat banal increase even assumes the worst case: that 1/f noise continues to increase down to 1nHz, for which so far there's 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 ageing effects and long-term drift rather than 1/f noise. Many engineers set a minimum frequency such as 0.01Hz for noise calculations in DC circuits in order to keep the calculations practical.

7 Noise equivalent bandwidth is a multiplier for noise

The noise equivalent bandwidth (NEB) is a useful simplification for noise calculations. Some noise from beyond the circuit's bandwidth can 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 circuit creates. The NEB is larger than the -3dB bandwidth, and it has been calculated for common filter types and orders; for example, it is 1.57 times larger than the -3dB bandwidth (BW) for a one-pole low-pass filter or, in equation form, $NEB_{1\text{-pole}} = 1.57 \times BW_{3dB}$. However, there's persistent confusion about where to put that multiplication factor in the noise equation. Remember that the NEB is an adjustment for bandwidth not noise, so it goes under the square root as follows:

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

8 Amplifiers with lowest voltage noise are the best choice

When choosing an op-amp, voltage noise is often the only noise specification considered by the designer, yet it's important not to overlook current noise as well. Except in special cases such as input bias current compensation, current noise is typically the shot noise of the input bias current = $\sqrt{2 \times q \times I_B}$. Current noise is converted to voltage by the source resistance, so with large resistance in front of the amplifier input, current noise can be a larger noise contributor than 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 10kΩ resistor in series with the input. The voltage noise of the ADA4898-1 is 0.9nV/√Hz, the 10kΩ resistor has 12.8nV/√Hz, and the 2.4pA/√Hz current noise times the 10kΩ resistor is 24nV/√Hz – the largest noise source in the system.

In cases like these, where 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, although 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 0.9nV/√Hz voltage noise, a JFET input amplifier such as the AD8033 or the ADA4817-1 could have been chosen.

9 Best noise performance is achieved by taking 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 is that this reduces the maximum signal the system can accommodate. In some cases, rather than taking a large gain in the first stage, improving the sensitivity of the measurement but limiting the dynamic range, it may be better to limit the gain taken in the first stage and digitise with high resolution to maximise both sensitivity and dynamic range.

10 All resistor types have the same noise for a given resistance

The Johnson noise of resistors is fundamental, expressed as a simple noise equation for a 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 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 $1\mu V_{rms}/V_{dc}$ per decade. This means that if there is $1V_{dc}$ across a resistor with a 0dB NI, the excess noise in a given frequency decade is $1\mu V_{rms}$.

Carbon- and thick-film resistors have some of the highest NI, ranging up to +10dB, and it's better to avoid them in noise-sensitive parts of the signal path. Thin-film resistors are generally much better at around -20dB, and metal foil and wirewound resistors can be below -40dB.

11 Given enough acquisitions, averaging reduces noise indefinitely

Averaging is considered to be a way to reduce noise by the square root of the number of averages. This is conditionally true when the noise spectral density is flat. However, this relationship breaks in the 1/f range and 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 an m number of 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 white noise by $\sqrt{n}$. 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 (since there's no such thing as 0Hz).

The more averages are taken, the lower these maximum and minimum frequencies move in 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 applies to long integration times for an integrating ADC such as multi-slope. Beyond this mathematical exercise, there are other practical limits, too. For example, if quantisation 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. EW