22. Resistor noise: reviewing basics, plus a fun quiz

The noise performance of amplifier circuits is greatly affected by the Johnson noise of resistors: the source resistance and feedback resistors. Most everyone seems to know that resistors have noise but may be a bit foggy on some of the details. Here is a bite-sized review in preparation for future discussions on amplifier noise.

The Thevenin noise model for a resistor consists of a noiseless resistor in series with a noise voltage; see **Figure 56.**

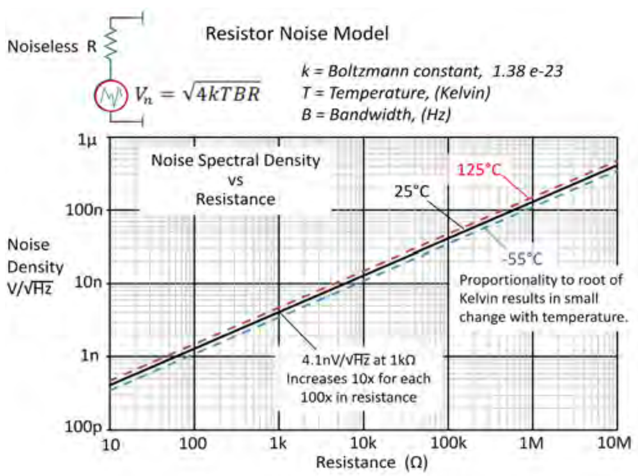


Figure 56: A resistor's Thevenin noise model is a noiseless resistor in series with a noise voltage.

The noise voltage is proportional to the root of the resistance, bandwidth and temperature (Kelvin). TI often quantifies the noise in a 1-Hz bandwidth as its spectral density. The theoretical noise of a resistor is "white," meaning that it is spread uniformly over frequency. It has equal noise voltage in every equal slice of bandwidth.

The noise in each 1-Hz band sums randomly according to the root of the sum of the squares. We often refer to the spectral density in volts/root-Hertz. The numerical value is the same as for a 1-Hz bandwidth. For white noise, it is convenient to multiply by the square root of a bandwidth to sum the random contribution of each 1-Hz band. To measure or quantify the total noise, you need to limit the bandwidth (**Figure 57**). Without a known cutoff frequency, you do not know how much noise you are integrating.

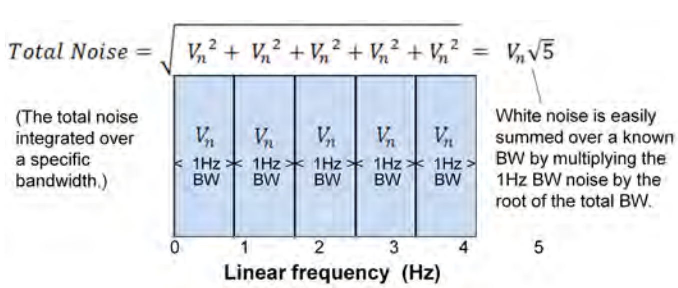


Figure 57: Summing the incremental 1-Hz bandwidths of white noise.

You may instinctively think of spectral plots as having a logarithmic frequency axis – a Bode plot. Note that a Bode plot has more hertz of bandwidth on the right side than the left side. Considering total noise, the right side of a Bode plot may be much more important than the left side.

Resistor noise is also Gaussian, a description of its amplitude distribution, a probability density function. It is Gaussian because it was created by the summation of a gazillion little random events. The central limit theorem explains how this noise becomes Gaussian. The root-mean-square (RMS) voltage of alternating current (AC) noise is equal to ±1 σ of the amplitude distribution (**Figure 58**). For 1-V RMS noise, there is a 68 percent (±1-σ) probability that the instantaneous voltage will be within a ±1-V range. A common misconception is to relate or equate white and Gaussian, but they are unrelated. Filtered resistor noise, for example, is not white but remains Gaussian. Binary noise is definitely not Gaussian, but it can be white. Resistor noise is white **and** Gaussian.

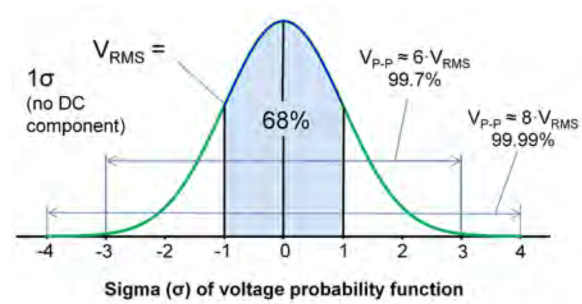


Figure 58: Gaussian noise spikes outside the ±3-times range are rare.

Purists like to rant that Gaussian noise does not have a defined peak-to-peak value – it is infinite, they say. True enough – the tails of a Gaussian distribution reach to infinity, so any voltage is possible. As a practical matter, the likelihood of noise spikes beyond ±3 times the RMS value is pretty small. Many folks use an approximation of six times the RMS for the peak-to-peak value. You can add a large additional guardband by using eight times the RMS without greatly changing the value.

Some fun points to ponder: The noise voltages of two resistors in series sum randomly, and the result is the same noise as for the sum of the resistor values. Similarly, the noise of resistors in parallel results in the noise of the parallel resistance. If it worked out differently, it would be problematic: think about bisecting a physical resistor and combining them in series or parallel. But it all works out.

A large-value resistor lying on your desk will not arc and spark from unlimited self-generated noise voltage. Stray parallel capacitance will limit the bandwidth and the total voltage. Similarly, the high-noise voltage you might imagine on insulators is shunted by parallel capacitance and the resistance of conductors around them.

Fun quiz: What is the total open-circuit noise voltage on a resistor that has a stray parallel capacitance of 0.5 pF? The solution details are here.

To see this original post with comments, click here.

23. Op amp noise: the noninverting amplifier

Building on the discussion of resistor noise in section 22, let us check out some basics of amplifier noise. The noninverting operational amplifier (op amp) configuration is most common for low-noise applications, so I will make that the focus.

Modeling the input source as a voltage noise source with a series resistance (**Figure 59**), you know that the source resistance, R_S, has a noise proportional to the root of its resistance (the straight line in **Figure 60**). The goal of a low-noise amplifier is to contribute minimal additional noise to what the source resistance generates.

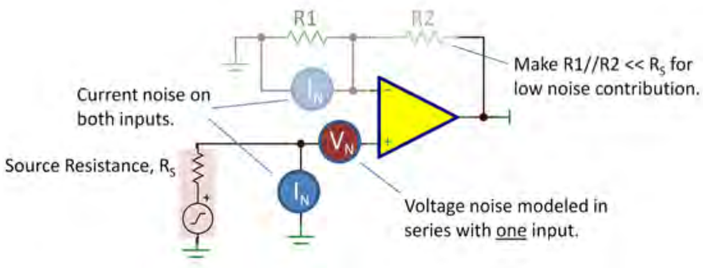


Figure 59: Amplifier noise is modeled as a voltage noise in series with one input and current noise sources connected to each input.

The amplifier noise is modeled as a voltage noise in series with one input and current-noise sources connected to each input; see **Figure 59**. Think of the voltage noise as just a time-varying component of offset voltage. Likewise, the current noise is a time-varying component of input bias current, one on each input. Ignore the current noise at the inverting input in this circuit – you can usually make its noise contribution minimal.

Figure 60 shows the total input-referred noise of the circuit for two op amps – the bipolar junction transistor (BJT)-input OPA209 and the junction FET (JFET)-input OPA140. Each is shown relative to the noise of the source resistance at 25°C. The three sources of noise are summed by root sum of squares for each op amp. You may have seen this graph in some op amp datasheets.

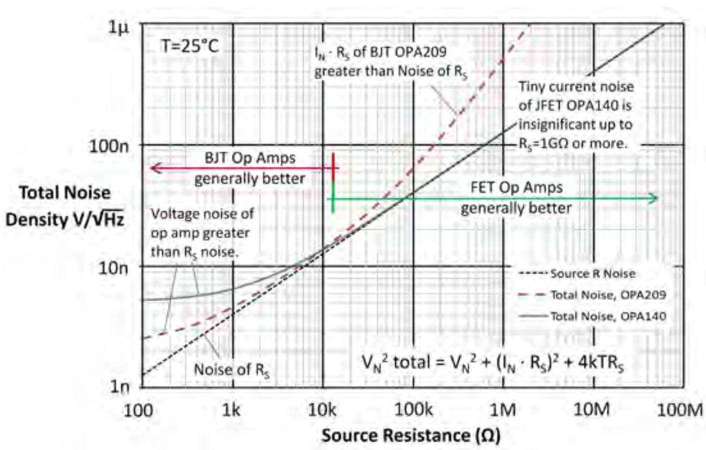


Figure 60: Using the OPA209, voltage noise dominates at low source resistance and current noise dominates at high source resistance.

As the source resistance decreases, its attendant Johnson noise decreases (by the inverse of the root of the resistance), and at some point the amplifier's voltage noise dominates. The total noise flattens to a value equal to the voltage noise of the amplifier. As the source resistance increases, the current noise flowing through the source resistance creates noise that increases linearly, rising more rapidly and eventually exceeding the noise of the source resistor. So, with high source resistance, the current noise effects dominate.

The greatest challenges in a low-noise amplifier design often come with low source resistance – 2 kΩ and lower. The lower source-resistance noise in this region demands amplifiers with very low-voltage noise. In general, BJT-input amplifiers excel in this range. Notice also that the total noise of the OPA209 in **Figure 60** dips nearest to that of the source resistance at a "sweet spot." This source resistance of best noise performance occurs at $R_S = V_N/I_N$.

FET-input amplifiers contribute little additional noise, with source resistance above 20 kΩ or so. The current noise of a FET op amp does not generally play an important role until you reach multigigaohm source resistance. A guideline: Below 10-kΩ source resistance, low-noise BJT amplifiers generally provide lower noise. Above approximately 10 kΩ, FET or complementary metal-oxide semiconductor (CMOS) op amps will likely have an advantage.

The feedback network, R1 and R2, also contributes noise but you can generally make this insignificant. How? The short answer is that if the parallel combination of R1 and R2 is one-tenth of R_S (or less), they will add less than 10 percent (<1 dB) to the total noise. This is true regardless of the ratio of the resistors that set the closed-loop gain. The noise of feedback components is assumed to be zero in **Figure 60**.

Of course, there is much more to know, but an understanding of this frequent case is a good start. Want more? I recommend "Operational Amplifier Noise: Techniques and Tips for Analyzing and Reducing Noise," written by my colleague Art Kay.

Point to ponder: The OPA140 has a very broad resistance range above 10 kΩ where noise performance is excellent. Is there a way to adapt a lower source resistance to take advantage of this region of operation?

To see this original post with comments, click here.

24. Op amp noise: but what about the feedback resistors?

In section 23 I explored noninverting amplifier noise, but I dodged the issue of the feedback network's noise contribution. So what about the noise from R1 and R2 in **Figure 61**?

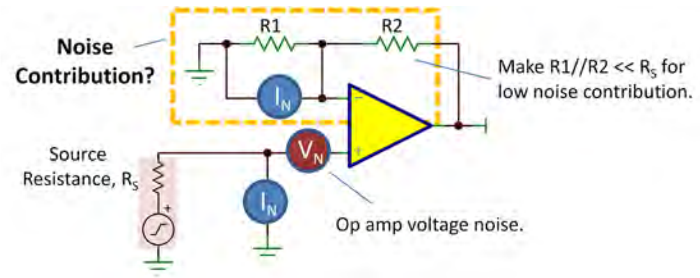


Figure 61: The inverting input comprises the feedback resistors' thermal noise and the op amp's current noise reacting with R1 and R2 components.

The noise contribution at the inverting input comprises the thermal noise of the feedback resistors and the operational amplifier's (op amp's) current noise reacting with R1 and R2 components. You can calculate the output contribution of these noise sources using basic op amp assumptions:

- R1's thermal-noise voltage is amplified to the output by the inverting gain of the circuit, –R2/R1.
- R2's thermal noise contributes directly to the output noise.
- The inverting input current noise flows through R2, resulting in an output noise contribution of I_N×R2.

These noise sources are uncorrelated, so they "add" by the root sum of the squares.

But there is a more intuitive way to look at this. It can be handy to refer to noise sources as if they all occur at the noninverting input. Output noise contributions are divided by the noninverting gain. This referred-to-input (RTI) approach makes it easy to compare noise sources to the input signal.

The noise occurring at the inverting input relates to the parallel combination of R1 and R2. When referred to the noninverting input, the combined RTI thermal noise of R1 and R2 is equal to the thermal noise of R1//R2. The current-noise RTI contribution at the inverting input is equal to I_N×(R1//R2). It is all about R1//R2.

Noise contribution of R1 and R2 and inverting current noise (equation 2):

$$\text{Output Noise}^2 = [V_{NR1} \cdot (R2/R1)]^2 + (V_{NR2})^2 + (I_N \cdot R2)^2 \qquad (2)$$

Dividing by non-inverting gain to refer to input (equation 3):

$$\text{RTI Noise}^2 = \underbrace{(V_{NR1//R2})^2}_{\textit{Thermal noise of R1//R2}} + (I_N \cdot R1//R2)^2 \qquad (3)$$

This result reveals an important factor for a low-noise design. Make R1//R2 < R_S and the noise contribution at the inverting input is negligible. If R1//R2 = R_S, then the feedback network contributes equal noise to that of the source resistance. That may be too much for some designs.

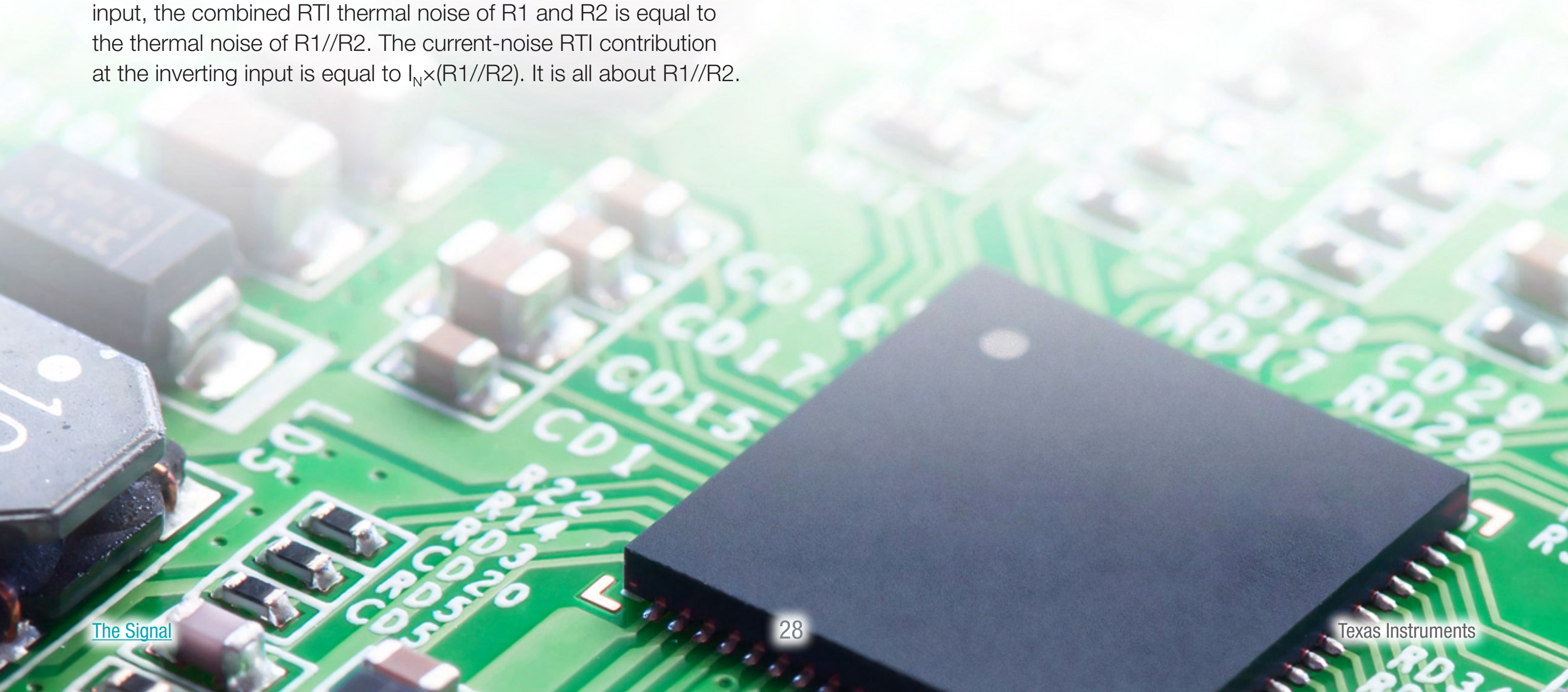
In high gains, it is easy to keep the parallel resistance low – R1 can be made much less than Rs and R2 is big. At moderate gains it gets more difficult. G = 2 is the worst case when R1 and R2 are equal. If you want to make the parallel resistance 100 Ω, for example, R1 and R2 need to be 200 Ω.

The feedback network then imposes a 400-Ω load on the op amp – too low in most circumstances. It gets easy again very close to G = 1 when R1 is big and R2 is small. This case is not common because you generally want significant gain in the first low-noise stage.

To address a common concern: There is no inherent noise penalty in making R2 a high resistance. If you can achieve higher gain by increasing R2 and decreasing R1 while maintaining a constant parallel resistance, noise performance remains constant.

You can download an Excel file to calculate the noise of this commonly used input-amplifier stage, including the op amp and source-resistance noise. It shows the percentage contribution of each noise source and graphs the total noise over a range of source resistance. It also calculates noise figure, which is the noise (in decibels) that the amplifier adds to thermal noise of the source. This is a handy measure of the noise performance of the amplifier. Tinker with it and you will quickly get a feel for the issues and trade-offs.

To see this original post with comments, click here.



25. 1/f noise: the flickering candle

The one-over-f (1/f) low-frequency noise region of amplifiers seems just a bit mysterious. It is also called flicker noise, like a flickering candle. Seen on an oscilloscope with a slow sweep, it has a wandering baseline (**Figure 62**) because the high-frequency noise rides on larger low-frequency content. Pink noise, another metaphoric name, also suggests a stronger low-frequency component. Flicker noise seems ever-present in physical systems and life science. Weather/climate patterns, for example, have a 1/f component. I will not attempt to explain why it is found in semiconductors – deep subject!

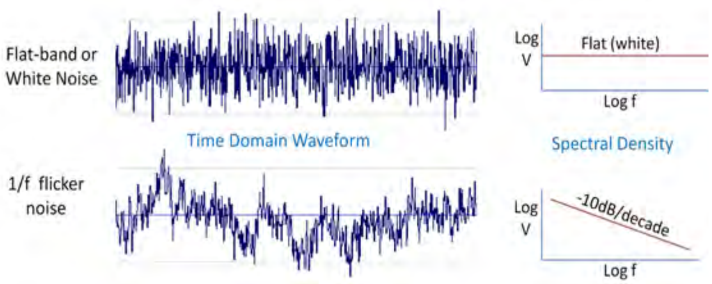


Figure 62: White noise (top) compared with 1/f noise (bottom).

The spectrum of flicker noise has a nominal slope of -10 dB/decade, half that of a single resistor-capacitor (RC) pole. Note that it's the square of the voltage (or power) that declines at a 1/f rate. Noise voltage falls at 1/sqrt(f). The actual slope can vary somewhat, but this does not greatly change its behavior or the conclusions.

A measured spectrum of flicker noise generally looks lumpy, with dips and valleys. You need to average for long periods to get a reasonably smooth plot. The period of 0.1-Hz noise content is 10 seconds, so for a good measurement down to 0.1 Hz you need to average many 10-second periods – five minutes or more. For 0.01-Hz data, take a long lunch. If you repeat the measurement it will likely look different. Noise is noisy and 1/f noise seems noisier than most other noise (did I write that?).

To calculate total noise, V_B, over a bandwidth (f_1 to f_2), integrate the 1/f function which results in the natural logarithm of the frequency ratio, f_2/f_1.

$$V_B^2 = v_a^2 f_a \int_{f1}^{f2} \frac{1}{f}\, df = v_a^2 f_a \cdot ln\left(\frac{f_2}{f_1}\right); \quad \boxed{V_B = v_a \sqrt{f_a \cdot ln\left(\frac{f_2}{f_1}\right)}} \qquad (4)$$

Where v_a is the flicker spot noise density at frequency f_a.

Points to ponder:

- Each decade of frequency (or other constant ratio of frequencies) contributes equally to total noise. Each successive decade has lower noise density but more bandwidth.
- From the spectral plot, you might infer that 1/f noise grows boundlessly as you measure for increasingly long periods. It does, but very slowly. Noise from 0.1 to 10 Hz doubles (approximately) with a lower bandwidth extended to 3.17e-8 Hz (a one-year period). Add another six percent for 10 years.
- It is challenging, but not impossible, to filter 1/f noise. Flicker noise from 0.1 Hz to 1 kHz (four decades) filtered to 10 Hz (two decades) only reduces the noise by 3 dB. Resistor values must be low for low noise, which makes capacitor values large for a low-frequency cutoff.

Amplifier noise is a combination of 1/f noise and flat (white) noise. The flat noise continues at a low frequency but 1/f noise dominates (**Figure 63**). The 1/f noise continues at a high frequency but flat noise dominates. The two blend at the corner frequency, adding randomly to make a 3-dB increase.

Amplifier noise is summed over the f_1 to f_2 bandwidth by integrating the 1/f and flat noise separately over the bandwidth, and then combined by the root sum of squares.

Other points to ponder are:

- An N-times increase in flicker-noise density increases the corner frequency by N^2.
- The total noise from a decade below to a decade above the corner frequency is dominated by the flat-band noise (68 percent) even though the 1/f noise region "looks bigger."

You can download an Excel file here that calculates integrated 1/f noise and flat-band noise, producing a graph and data similar to **Figure 63**. Tinker with it to get a better feel for the issues.

Amplifiers with bipolar junction transistor (BJT)-input stages (OPA211) generally have lower 1/f noise, but new-generation analog integrated circuit (IC) processes have greatly improved junction FET (JFET) and complementary metal-oxide semiconductor (CMOS) transistors. The OPA140 (JFET) and OPA376 (CMOS) operational amplifiers, for example, have corner frequencies of 10 Hz and 50 Hz, respectively. Chopper amplifiers virtually eliminate 1/f noise by correcting offset-voltage changes.

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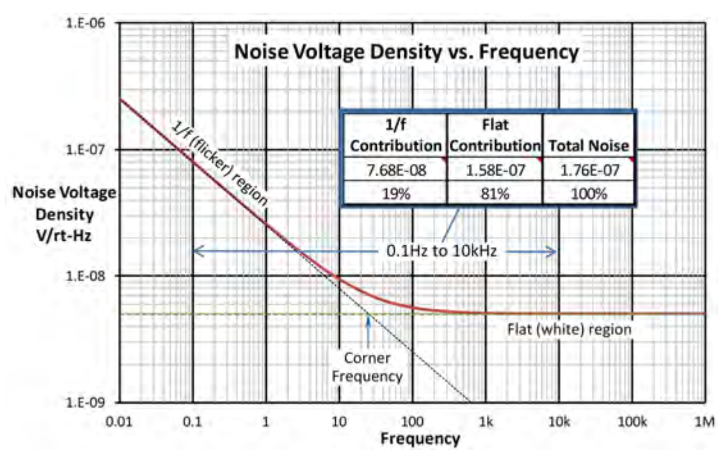

Figure 63: Amplifier noise in this graph is a combination of 1/f noise and flat (white) noise.

26. Chopper op amps: are they really noisy?

Chopper operational amplifiers (op amps) offer very low offset voltage and dramatically reduce low-frequency 1/f (flicker) noise. How do they do it?

Figure 64 shows the input stage of a chopper op amp. The amplifier is a relatively conventional transconductance stage with differential input and differential output current. Chopping is accomplished with commutating switches on the input and output that synchronously reverse the polarity. Since both differential input and output are reversed simultaneously, the net effect on the output capacitor, C1, is a constant signal-path polarity.

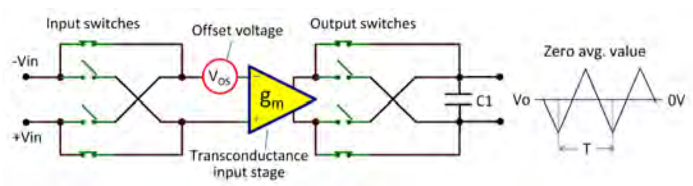

Figure 64: Input stage of a chopper op amp.

The offset voltage of the transconductance stage is inside the input switching network, so its contribution to output is periodically reversed by the output switches. The output current caused by offset voltage causes the voltage on C1 to ramp up and down at an equal rate. Internal logic assures equal up and down ramp times, so the average output voltage on C1 is zero. Thus, zero offset!

Early-generation choppers provided only modest filtering of triangular chopping noise, causing them to be branded as wickedly noisy devices, used only when very low offset voltage was crucial. (And this is how big, noisy motorcycles came to be.) Particularly troublesome was that the pre-chopping offset voltage determined the magnitude of the triangle waveform, so chopping noise could vary considerably from unit to unit.

New-generation choppers are dramatically quieter, incorporating a switched-capacitor filter with multiple notches aligned with the chopping frequency and its odd harmonics. This is accomplished by integrating a charge on C1 for a full cycle before transferring its charge to the next stage of the op amp. Integrated over a full up-down cycle, its net value is zero – perfectly averaged. In the frequency domain, this creates a sinc(x) or sin(x)/x filter response with nulls that precisely align with the fundamental and all harmonics of the triangle wave (**Figure 65**).

In its final implementation, eight switches in the output-commutation network alternately charge two C1 capacitors. This enables integration of the input signal on one capacitor, while charge on the other capacitor transfers to the next stage of the op amp.

Since 1/f (flicker) noise is merely a slow time-varying offset voltage, choppers virtually eliminate this increased noise-spectral density in the low-frequency range. The chopping shifts the baseband signal to the chopping frequency, beyond the input stage's 1/f region. Thus, the low-frequency signal range of the chopper has a noise-spectral density equal to that of the amplifier's high-frequency range.

I made this all sound neat and tidy. Zero offset ... perfect! Of course, there is still some residual offset error produced by switching charge injection and the mismatch of capacitance and parasitics. The gain of the input stage (discussed here) greatly reduces offset contributed by later op amp stages. In general, a wider amplifier bandwidth requires faster chopping, which increases residual offset errors. The residual offset tends to be very stable with temperature and through product life, an important attribute for these devices.

Now I do not claim that modern chopper op amps eliminate the need for standard op amps: far from it. But new-generation choppers are now useful in a much wider range of applications. They provide very low and stable offset voltage, virtually no flicker noise, and very near the behavior of a standard op amp.

To see this original post with comments, click here.

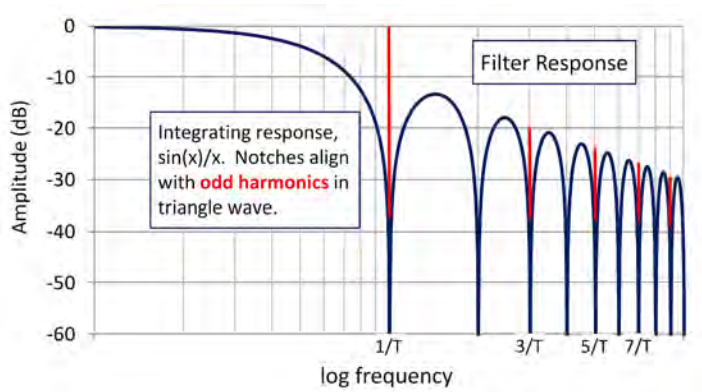

Figure 65: New-generation chopper op amps incorporate a switched-capacitor filter with multiple notches aligned with the chopping frequency and its harmonics.