Application Note 6

Noise-Gain Errors, the Dark Side of Feedback Ammeters

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Users of feedback ammeters, hereafter referred to as a trans-impedance amplifier (TIA), appreciate the fact that they have essentially zero voltage drop, unlike the more typical DMM ammeter that typically has ±100 mV drop across it at ±full-scale. TIAs also offer much better sensitivity, with full-scale ranges as low as ±2.0 nA with single-digit femto amps (fA which is 1 x 10^{-15} A) resolution.

There is a dark side to these TIA ammeters, however, and this paper delves more deeply into those dirty little secrets.

Let's start with a quick review of the TIA:

Here we have the simplest instantiation of a TIA ***(figure 1)***. Assuming U1 has extremely high open loop gain, its inverting (-) input is virtually at ground (0 V). Imagine the opamp output is at -1 V and the openloop gain is 106, the input will be +1 µV. Also, assuming U1 has zero inverting input current, all of the input current (I1) must flow through R1. Thus the output is -I1 * R1. This TIA has converted an input current (I1) into an output voltage.

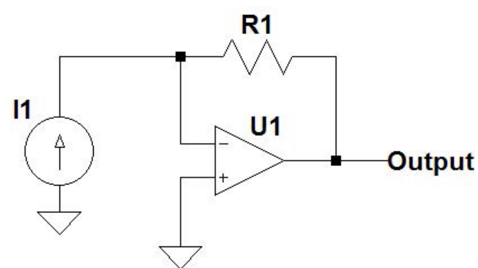

figure 1

As stated in the previous paragraph, its inverting input is virtually at ground, hence proof of the earlier statement that this ammeter has zero voltage drop across its input. If R1 was 10 kΩ, and I1 was 100 µA, the output would be -1 V, and the open loop gain was 106, the input would be -1 µV; pretty darned close to 0 V. Most DMMs don't even have a 100 µA range; the Keysight 3458A does and it's input IR drop is ±75 mV at full-scale. 1 µV vs. 75 mV may, or may not, be significant for a given application, but it is important to recognize that this potential error source exists.

So, if TIAs are so good, why all the scary talk? Noise gain is the answer!

All TIAs suffer from errors due to amplifier noise gain. So, let's review the subject of noise gain.

First of all there are a number of very good references to the subject and here are two:

http://www.analog.com/library/analogdialogue/archives/47-03/raq_91.html

http://www.ti.com/lit/ml/sloa082/sloa082.pdf

It is important to recognize that the word "noise" does not refer to any old noise, but to the undesired signals that appear at the amplifier's non-inverting (+) input. In ***figure 2*** this noise source is represented by V1 which has been added to the schematic of ***figure 1***.

V1 can be from any source, internal to the opamp, external noise pickup, it can be AC or DC.

Notice that I also snuck in V2 and R2 to replace the current source. After all, with a virtual ground at the inverting input of U1, the current through R2 is constant, right? Well, sort of.

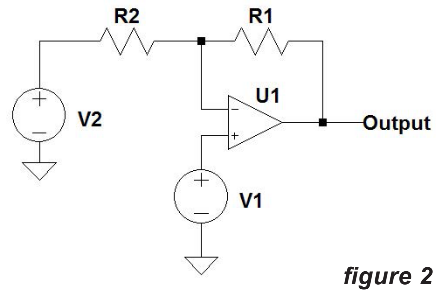

figure 2

For amplifiers with extremely high open-loop gain, the noise-gain circuit is simply the gain that is applied to the noise source V1:

$$NoiseGain = 1 + \frac{R1}{R2} \qquad \textbf{eq 1}$$

In the case of our ideal TIA ***(figure 1)*** with a true current source as its input, R2 = ∞, so the noise gain is 1. If our opamp had $\frac{50\,nV}{\sqrt{Hz}}$ of input voltage noise and one were measuring with a 100-Hz bandwidth, the rms noise would be 500 nV rms.

Or, if U1 had an offset voltage of 100 µV DC, the output offset due to noise gain would be -100 µV DC.

For ***figure 2***, assume that R1 = R2. Now the rms noise is 1 µV rms. Now, the 100 µV DC, the output offset due to noise gain would be -200 µV DC.

Where did V2 and R2 come from? There are a few places to look:

Since nothing in the real world is ideal, even good current sources, like the leakage current of a reversed biased diode has some resistance, and even worse, capacitance. The noise-gain will increase with frequency due to capacitance from the inverting input to ground. This would include effects due to cable capacitance.

Even when calibrating a TIA-based ammeter, one might be tempted to place a known resistance in location R2 and a known voltage for V2, and claim the current is V2/R2, which it will be, but the noise and offset voltage errors are now increased by 1 + R1/R2. So, while this method works, it is best to ensure that R2 is ≥ R1, or be prepared to do a lot of averaging to reduce noise effects.

Offset voltage errors can be eliminated by separating gain errors from offset errors by the use of reversals. Take two measurements, one with a positive V2 and one with a negative V2:

$$I_{Input} = \frac{1}{2}\left(\frac{V2_{pos} - V2_{neg}}{R2}\right)$$

eq 2

The output will be:

$$V_{Output} = -\left[\left(\frac{+V2}{R2} + Vos_{U1}\left(1 + \frac{R1}{R2}\right)\right) - \left(\frac{-V2}{R2} + Vos_{U1}\left(1 + \frac{R1}{R2}\right)\right)\right] + Vena_{U1}\left(1 + \frac{R1}{R2}\right) + Venb_{U1}\left(1 + \frac{R1}{R2}\right)$$

eq 3

Where:

- $V2/R2$ is the actual input current
- V_{OS} is $U1$'s offset voltage
- V_{ena} and V_{enb} are $U1$'s rms input voltage noise during the time each measurement is taken. Note that the fixed offset voltages (V_{OS}) cancel each other. However the voltage noise (V_{en}) do not because they are uncorrelated, and since they are uncorrelated they root-sum-square (RSS) rather than add directly.

This equation reduces to:

$$V_{Output} = -\left(2\frac{V2}{R2} + \sqrt{{Vena_{U1}}^2 + {Venb_{U1}}^2}\left(1 + \frac{R1}{R2}\right)\right)$$

eq 4

Note that the output voltage, due to input current $(2\frac{V2}{R2})$, doubled in magnitude, while the noise RSSed or only increased by 1.4 times. This is another advantage of reversals, not only did the offset disappear, but the signal-to-noise ratio improved by 3 dB.

As mentioned earlier, averaging will reduce the effects of noise. For Gaussian noise, the noise drops with the square root of the number of samples averaged. One can also choose a lower sample rate in the measurement device, this is another form of averaging and a lower sample rates means measuring the input signal over a longer period, a kind of averaging. Even though it is more work, I sometimes prefer to simply acquire more data at a higher rate and do the analysis in a spreadsheet or some other calculation tool, because this also allows me to assess the variability in a group of measurements, rather than one single measurement; this analysis of variability should be part of the uncertainty analysis anyway. It also lets me plot the data *(highly recommended by this author)* which can frequently reveal trends like settling tails, non-Gaussian noise, and thermal drifts that simple statistics like the mean and the standard deviation will not indicate.

So, what about the value of R1? While the user of a TIA-based ammeter has little choice of this (and I have never seen the actual value of the feedback resistors in a datasheet), the meter's designer does. Similar to the reversals mentioned above, the thermal noise of the resistor goes up with the square root of the resistance, while the signal magnitude increase linearly, so larger resistances have better signal to noise ratios. On the other hand, larger resistances also make the TIA-based ammeter more sensitive to noise-gain errors and more sensitive to lower input resistances. Pick your poison.

Back to the concept of using known voltage sources and known input resistors for calibration purposes, how low can one go? The trade-off here is, for high-value input resistances, say > 1 MegΩ, the accuracies of the standards drop as the resistance increases.

The trade-off here is the uncertainty associated with the resistance increases for high-value resistances. A good example ***(figure 3)*** is the uncertainty of a Keysight 3458A's resistance measurement function. This is a darned good meter with very impressive specs. Note how above 100 kΩ things rise drastically. Even though most lab standards are more accurate, they still follow a similar uncertainty versus resistance curve.

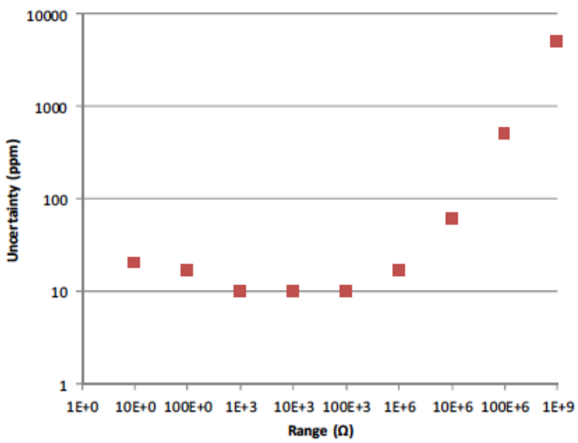

figure 3

So, when choosing an input resistor to use when calibrating a TIA ammeter, low resistances cause higher noise gain errors and high resistances have higher uncertainty. How does one find a reasonable compromise?

The real-world TIA does not have only resistors for feedback, they have capacitors in parallel with the resistors to improve stability when there are higher capacitance input sources. This, along with other circuit components, complicate the simple noise gain equation. These capacitors affect the frequency response and the shape of the noise gain, so all of the above is only an approximation, but the effect is still real, lower input impedance sources increase the noise gain, and thus the measurement noise. Just as car manufacturers state that "your mileage may vary", your noise gain may vary! It will be necessary to try things, make measurements, make changes, and try again.

Here in ***figure 4*** are simulation results (solid line) for the Model 100 Series' 20 µA range along with some actual measured values (violet squares). The noise gain axis is simply 1 + R1/R2, the noise ratio axis is the output noise at a given noise gain divided by the noise at a noise gain of one. Note that, as mentioned above, the output noise gain isn't a simple 1 + R1/R2 due to feedback and input capacitances, but it definitely increases with noise gain.

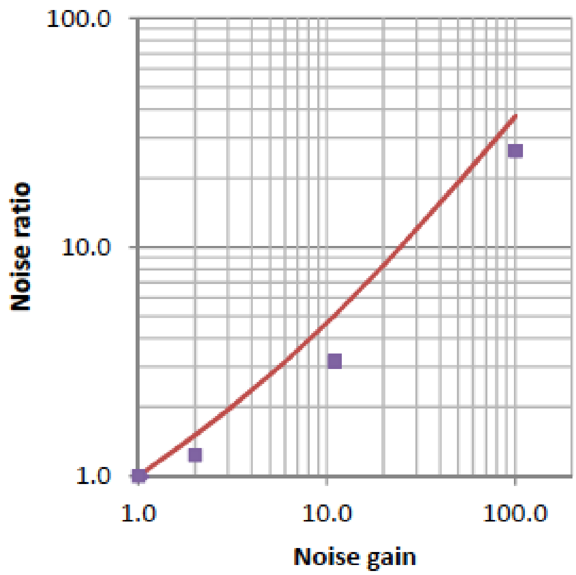

figure 4

The SPICE simulation used an opamp with the same input voltage noise and bandwidth, and the same feedback resistor with shunt capacitor.

The model and measured correlate reasonably well; well enough to prove the point about noise gain being affected by the TIA input source resistance.

Figure 5 is a photo of the AMETRIX Instruments Model 101 with a switchable input resistance used for the actual noise-gain measurements.

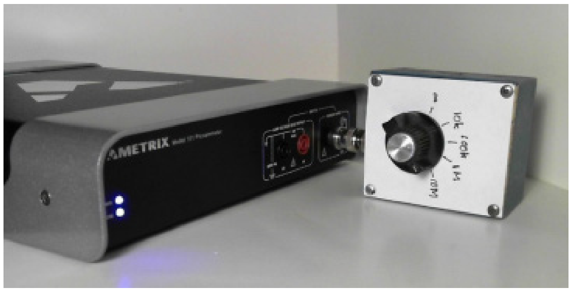

figure 5

This physical arrangement does not represent a typical application; it's only purpose is to study the effects of various less-than-ideal input source impedances on the quality of measurement in a TIA-based ammeter. This arrangement ensured minimal normal mode noise, mains pickup, for example, while allowing for convenient source impedance selection. Unfortunately adding cables and fixtures will add extra noise due to pickup, but in this study it was desired to keep those other sources out of the measurement.

Another source of error when using low values for R2 is thermal EMFs. If, for example, R1 was 1 MegΩ and full-scale current was 200 µA (i.e. full-scale output is 2 V), and R2 was 10 kΩ, a V2 of a mere 20 mV would drive the output to 2 V. In other words, the output is very sensitive to V2.

The various connections to the calibration voltage source (V2) don't show up on the schematic, there are numerous electrical junctions that contain different materials and each one acts as a small voltage source, like a thermocouple. The material of R2's leads, copper PCB traces to the feedback ammeter's input terminals, and the input connector's phosphor bronze contacts, will all create some thermal EMFs. While this is not considered noise gain, low values for R2 creates an error source. The message in this paragraph is that all critical electrical connections and equipment should be low thermal-EMF types and will have to be in a stable thermal environment. Surrounding low thermal EMF connections with foam will help; placing the resistor R2 in a thermally lagged box will help. While these thermal EMFs are "eliminated" by reversals, these thermal EMFs tend to change during the test time due to breezes from HVAC systems cycling, people walking by, etc. Be aware of your surroundings.

Charge injection due to cables dangling in air, or vibration of the work table can and will cause measurement errors. Even charge injection due to people walking by or moving in the area of the resistors and cables can have negative effects on the measurement. The solution is to work on a heavy, rigid bench with proper ESD mats. Please tape down sensitive cables so they don't sway and wiggle. Listen to your grandmother and sit still.

Up until this point, things have sounded very predictable and scientific, but in actual application there is some art involved.

Every measurement situation is different, so please consider all of the above to be guidelines. In reality there is some kind of multi-dimensional thing ***(figure 6)*** that vaguely describes any given measurement setup with all of its variables, and somewhere there is probably a minimum uncertainty point; getting near that point is the goal.

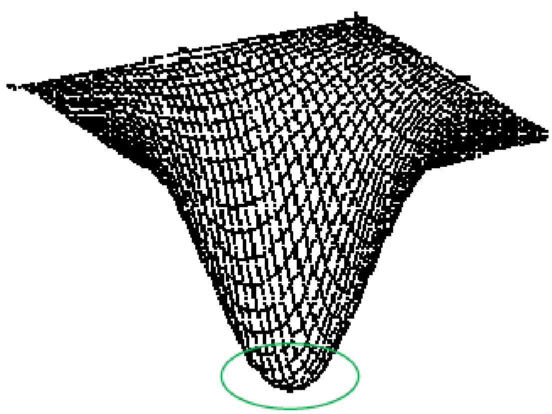

figure 6

Figure out which error source dominates the uncertainty and optimize it, and then move on to the next major contributor. Take data and keep track of the changes and how those changes affect the measurement. Note which changes degrade and which improve the measurements, and how and why they do it. It may take some iteration of the various inputs into the system to find the minimum.

Usually the multi-dimensional blob that describes any given measurement situation has a fairly wide and somewhat flat minimum, so getting exactly at the bottom may not be all that important; close may be close enough.

In conclusion, feedback ammeters are sensitive to low input source impedances, so keep the source impedances as high as possible, and if possible, use low-sample rates and averaging to reduce the effects of inherent opamp input noise. Be aware of the other error sources too, and find a combination of voltage sources, input resistances, cables, etc. that minimize the uncertainty.

High noise gain does absolutely nothing to improve the desired signal's gain, high noise gain only degrades the signal-to-noise ratio.

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