Measuring nanoamperes

Paul Rako - April 26, 2007

Thousands of applications require a circuit to measure a small current. One of the most common is the measurement of photodiode current to infer the light impinging on the diode. Scientific applications, such as CT (computer-tomography) scanners, gas chromatographs, and photo-multiplier and particle and beam monitoring, all require low-level current measurements. In addition to these direct applications, the manufacturers of semiconductors, sensors, and even wires must measure extraordinarily low currents to characterize their devices. Leakage current, insulation-resistance measurements, and other parameters require consistent, accurate measurements to establish data-sheet specifications.

Few engineers realize, however, that the data sheet of a part is a contractual document. It specifies the behavior of the device, and any disputes over the operation of the part always come down to the specs on the data sheet. Recently, a customer of a large analog-IC company threatened legal action against the manufacturer, claiming that the parts he had purchased exhibited far higher operating currents than the submicroampere levels that the company specified. It turned out that the PCB (printed-circuit-board)-assembly house was properly washing the board but that assemblers were picking up the PCB and leaving fingerprints on a critical node. Because it could measure these tiny currents, the semiconductor company proved that its parts were working correctly; the leakage current was due to dirty PCBs.

The difficulty with measuring small currents is that all kinds of other effects interfere with the measurement (see **sidebar**"History of current measurements"). This article looks at two breadboard circuits that must handle surface leakage, amplifier-bias-current-induced errors, and even cosmic rays. As in almost all circuits, EMI (electromagnetic interference) or RFI (radio-frequency interference) can induce errors, but, at these low levels, even electrostatic coupling can cause a problem. As the currents you measure drop into the femtoampere range, the circuits are subject to even more interfering effects. Humidity changes the value of capacitors and causes higher surface leakage. Vibrations induce piezoelectric effects in the circuit. Minor temperature variations, even from a room fan, cause temperature gradients in the PCB that give false readings. Even room light can degrade the accuracy of measurements; light from fluorescent fixtures can enter the glass ends of a detector diode and cause interference (Reference 1).

Small currents require accurate measurement if you want to characterize the performance of quartz-crystal oscillators. Jim Williams, a staff scientist at Linear Technology and longtime *EDN* contributor, shares a circuit he designed for a customer who needed to measure the rms current in a 32-kHz watch crystal (Figure 1). One difficulty with this measurement is that even a FET probe's 1-pF loading can affect the crystal oscillation. Indeed, one of the goals of current measurement is to establish the sizing of the low-value capacitor you use with every crystal oscillator. A further difficulty of this measurement is that it must measure accurately and in real time at 32 kHz, which rules out the use of an integrating capacitor. The signal is a complex ac signal that the system designer must convert to an rms value for evaluation.

"Quartz-crystal rms operating current is critical to long-term stability, temperature coefficient, and reliability," says Williams. The necessity of minimizing introduced parasitics, especially capacitance, complicates accurate determination of rms-crystal current, especially in micropower-crystal types, he says. Figure 2's high-gain low-noise amplifier, he explains, combines with a commercially available closed-core current probe to permit the measurement, and an rms-to-dc converter supplies the rms value. The dashed lines indicate a quartz-crystal test circuit that exemplifies a typical measurement situation. Williams uses the Tektronix CT-1 current probe to monitor crystal current and introduce minimal parasitic loading. A coaxial cable feeds the probe's 50Ω output to A_1; A_1 and A_2 take a closed-loop gain of 1120, and the excess gain over a nominal gain of 1000 corrects for the CT-1's 12% low-frequency gain error at 32.768 kHz.

Williams investigates the validity of this gain-error correction at one sinusoidal frequency—32.768 kHz—with a seven-sample group of Tektronix CT-1s. He reports that device outputs are collectively within 0.5% of 12% down for a 1-µA, 32.768-kHz sinusoidal input current. Although these results tend to support the measurement scheme, Williams contends that it is worth noting that Tektronix measured the results. "Tektronix does not guarantee performance below the specified –3 dB, 25-kHz low-frequency roll-off. A_3 and A_4 contribute a gain of 200, resulting in total amplifier gain of 224,000. This figure results in a 1V/µA scale factor at A_4 referred to the CT-1's output. A_4's LTC1563-2 32.7-kHz bandpass-filtered output feeds A_5 through an LTC1968-based rms-to-dc converter that provides the circuit's outputs," he says. The signal-processing path, Williams explains, constitutes an extremely narrowband amplifier tuned to the crystal's frequency. Figure 3 depicts typical circuit waveforms. According to Williams, the crystal drive at C_1's output (upper trace), causes a 530-nA rms crystal current that the A_4's output (middle trace) and the rms-to-dc-converter input (lower trace) represent. "Peaking visible in the middle trace's unfiltered presentation derives from parasitic paths shunting the crystal," he says.

Williams' circuit provides several lessons. Measuring nanoamperes is difficult even when using integrating techniques. This problem was far more difficult, because he had to complete the measurement in real time. Further complicating matters was the fact that this ac measurement required a bandwidth of 32 kHz to capture the bulk of energy in the oscillator current waveform. Williams addressed these problems by using a sensor. The Tektronix CT-1 sensor (Reference 2) can cost as much as $500, but, without a good sensor, Williams would not have been able able to recover the signal from all the noise. In addition to good sensitivity, the CT-1 has a 50Ω output impedance that allows for lower noise-signal paths than would a high-impedance output. Another important principle that this example demonstrates is that it is essential to limit the bandwidth of the signal path. By making a narrowband amplifier chain, Williams discarded all the noise contributions from frequencies that were not in his area of interest. Finally, Williams used good low-noise design principles in the circuit. Wiring critical nodes in air minimizes leakage paths, and the LT1028 is perhaps the lowest noise amplifier available from any manufacturer when working from 50Ω source impedance.

Femtoampere bias current

Paul Grohe, an application engineer at National Semiconductor, provides another remarkable example of measuring tiny currents. Years ago, National decided to sell the LMC6001, an amplifier that had a guaranteed bias current of 25 fA, implying that National needed to measure the bias current of each part to verify the specification. The test department could not accommodate test equipment in the setup; all the circuitry had to fit onto a standard probe card. Grohe and engineering colleague Bob Pease built a proof-of-concept fixture to demonstrate the feasibility of a small test circuit that could resolve to 1 fA (Figure 4). Many books and resources discuss using an integrating capacitor to measure small currents (Reference 3). The principle is that a small current can charge a small capacitor and that you can read that voltage to infer the current. In some cases, the current is an external current from a sensor. In this case, the current is leaving the amplifier-input pin. Figure 5 shows a simple theoretical circuit in which the amplifier is measuring its own bias current.

The reality of measuring small currents is far more involved than the figure would suggest. First, Grohe could not use the part itself to measure its own bias current. If he had tried to use the part itself as the integrator, there would have been no way to calibrate the effects of a socket and other leakages associated with the test fixture. Doing so required a separate low-bias-current part as the integrator (Figure 6). Using a CMOS LMC660 amplifier ensured that the bias-current contribution would be less than 2 fA. By employing this technique, Grohe could simply remove any DUT (device under test), and the integrator would then have measured its own bias current as well as all the leakages from the test socket and the PCB on which the integrator was mounted.

Figure 7 shows that Grohe did not insert the DUT into a socket and that none of the pins are in contact with a PCB. To minimize leakage, Grohe brought up just two power pins as long, separate individual sockets that he did not mount to a PCB. Likewise, he hooked the pin to be tested to a socket and a 2-in. flying lead and connected that pin-and-socket combination to the integrating-amplifier input. To keep the DUT from running as an open loop, Grohe soldered together two sockets to bridge the output pins, which are suspended in air. Air currents can carry charged ions that can give false readings, so Grohe enclosed the entire DUT in a shielded copper-clad box.

The next issue was selecting an integrating capacitor. Initially, Grohe felt that the best capacitor would be an air-dielectric capacitor, so he fashioned two large plates, measuring about 4×5 in., for the integrator capacitor. The size of this capacitor accounts for the size of the second copper-clad box on which the DUT box is mounted. Using a large capacitor proved to be a bad idea. The large area provides an ample target for cosmic radiation, creating ionized charges that interfere with the measurement (Figure 8). Grohe then minimized the capacitor's size while still using a good dielectric. It occurred to him that RG188 coax cable uses Teflon insulation. A 2-in. section of this cable provided the 10 pF for the integration capacitor (Figure 9). As a further benefit, the outside braiding would serve as shield. Grohe therefore hooked it to the low-impedance-output side of the amplifier. With the switch to this capacitor, the cosmic rays struck only once every 30 seconds or so. Grohe took the integrated measurement for 15 seconds and, by taking five measurements, negated their effect. He then discarded any single outlying measurement. Any ionizing radiation sources, even an old watch with a radium dial, can cause cosmic-ray problems. Note that Grohe pried up the input pin of the amplifier to prevent leakage from the PCB.

Before taking a measurement, you need to reset the integrating capacitor to zero. Using a semiconductor switch is impractical, because of leakage currents and the 5- to 20-pF capacitance most analog switches offer. That capacitance exhibits the varactor effect, as well; it changes with applied voltage, further complicating measurement. To minimize these problems, Grohe used a Coto-reed relay. Knowing that the coil might couple to the internal reed when the relay was open, he specified a relay with an electrostatic shield. Much to his dismay, there still was a large jump in the measurement when the relay opened due to charge injection. It turns out that you can also look at a reed relay as a transformer, with the reed assembly representing a single turn. This phenomenon explains the failure of the electrostatic shield to prevent the interference. Magnetic fields inducing voltages in the high-impedance side of the circuit caused the charge injection. The relay does not open instantaneously, and the pulse needed to energize the coil makes a significant current injection just before the relay opens. Grohe minimized this problem by characterizing the absolute minimum voltage swing needed to operate the relay he had installed. It turned out that the relay would pull in with 3.2V and drop out with 2.7V. He used a set of resistor taps on an LM317 adjustable regulator to control the output between these two values. By choosing not to energize the relay with a full 5V, he minimized the jump in the integrator output and made it repeatable. He then nulled out the jump by injecting a small current into the second gain-stage amplifier.

The gain stages are two low-noise amplifiers—the LMV751 or perhaps a chopper amplifier, such as the LM2011, would be suitable. Grohe sent this gained-up signal to a digital scope, which could record data and subtract the slope of the calibration run from the test runs to give a valid measurement. Grohe used two LS123-style one-shot circuits—one to trigger the relay and another to provide a suitable and repeatable time delay that triggered the digital scope.

Grohe also understood that good low-noise-design principles also include the power rails to the parts, so he chose not to power the relay or digital circuits from the same power he used for the integrator and DUT. He used a handful of fixed and variable regulators to provide ±5V for the DUT and integrator, 8V for the relay-drive circuit, and a separate 5V for the digital circuits.

Using this circuit, Grohe was easily able to resolve 1 fA of current and found that most of the LMC6001 parts he tested had less than 5 fA of bias current, far exceeding the spec. He used this breadboard as the basis for a production-test circuit mounted on a standard probe card. (See reference 4, reference 5, and reference 6 for more about his design, including a video of the system.)

Grohe would not use this circuit to measure femtoampere currents in his lab. “I would wheel out the Keithley 2400 electrometer,” he says (Editor's note). “We would have used that instrument to test the LMC6001 in manufacturing had the fab allowed us to use external test equipment.” His faith in Keithley is well-placed. The company offers free on its Web site an excellent article on measuring attoamperes (Reference 7), as well as a book on delicate measurements (Reference 8).

DDC112

Grohe and Pease’s integration approach is not limited to laboratory setups. Texas Instruments has created a line of parts that can measure in the femtoampere range and provide a digital output to boot. The line includes a single-channel DDC101 as well as the improved-sensitivity, dual-channel DDC112, which provides for external integrating capacitors. The four- and eight-channel DDC114 and DDC118 have a charge sensitivity of 12 pC (Reference 9). The sample rate for these 20-bit parts reaches 3 kHz.

You must be cognizant of physics to attempt these measurements. If the DDC112 can measure 12 pC of charge and you want to measure 12 pA of current, you need to set the integrating time to 1 second, the maximum the DDC114 allows. It is impossible to obtain a 3-kHz update rate if the part’s integration interval is a full second. However, using the part configured in this fashion yields a 20-bit value at the end of the conversion. In other words, the DDC (direct digital converter) can resolve femtoampere currents, although at reduced accuracy. The input bias of the part is 20 fA, but your system’s software can calibrate out this value, so the part should still be able to resolve to very low levels. Bear in mind that this type of sensitivity makes it difficult to calibrate the system just once in the factory and then have it work for all time. As temperature increases, the bias current increases, doubling every 10°C, and leakages as well as sensor drift can develop on your board. Providing the means for field calibration at power-up or more frequently is always a good idea when measuring currents in the femtoampere range. Texas Instruments offers evaluation boards for these parts that you can get up and running in hours, measuring currents too small for even a good handheld digital voltmeter (Figure 10).

Read more In-Depth Technical Features

According to Jim Todsen, product-line manager for oversampling converters at TI and patent holder on the technology that the part uses, the DDC line's development started with the Burr-Brown ACF2101—a dual switched integrator front end that provides a single-chip option for the current-t--voltage function. The benefit of a dual integrator, Todsen explains, is that it is always collecting input current. While one integrator is sampling the input, the other side is presenting its integrated value to the ADC, and this process continues for as long as you need measurements. "After the ACF2101 converts the input current to a voltage," he says, "a discrete high-resolution ADC digitizes it. The DDC112 brought together both the current-to-voltage function of the ACF2101 and the digitization of the high-resolution ADC in one chip." He attributes this achievement to advances in wafer processing that allow high levels of mixed-signal integration as well as TI's development of a high-speed delta-sigma core that can provide the required speed and resolution to measure the front-end signals. "In addition," he notes, "we took advantage of having all the circuit elements under our control to optimize for very-low-leakage inputs and very stable performance over long integration periods."

These applications should convince you of the difficulty of measuring small currents. They should also convince you of the value of using proven parts and equipment—whether Analog Devices' AD549, National Semiconductor's LMC660, TI's DDC114 integrated circuits, Keithley's 2400 parameter-measurement unit (Editor's note), or Agilent's 4156 parameter-measurement unit—in this demanding application. Remember, though, that these remarkable parts and instruments are not magic boxes. You can take advantage of them only by removing noise sources and leakage paths from your board or test setup. Understanding op-amp specifications for voltage and current noise will help you select the right part (Reference 10). In the meantime, if your boss wants to know why you need $5 or $10 for a chip or thousands of dollars for an electrometer, you can now explain that, with the challenges entailed in measuring small currents, this equipment is a bargain.

Editor's note

After *EDN* published this article, some alert readers wrote in to say that the Keithley 2400 is a source meter, not an electrometer. For details, see "Keithley & TEGAM on measuring nanoamperes" from Paul Rako's blog, .

References

1. Long, James, "Sidebands be gone, or let there be (no) light,"*EDN*, Oct 12, 2006, pg 40.
2. "AC current probes".
3. Mancini, Ron, "The nuances of op-amp integrators,"*EDN*, March 18, 2004, pg 28.
4. Pease, Bob, "What's All This Teflon Stuff, Anyhow?" Feb 14, 1991.
5. Pease, Bob, "What's All This Femtoampere Stuff, Anyhow?" Sept 2, 1993.
6. www.national.com/nationaltv.
7. Daire, Adam, "Counting Electrons: How to measure currents in the attoampere range," Keithley Instruments Inc, September 2005.
8. www.keithley.com/wb/141.
9. "Quad Current Input 20-Bit Analog-to-Digital Converter," Texas Instruments Inc, June 2005.
10. Brisebois, Glen, "Op Amp Selection Guide for Optimum Noise Performance,"*Linear Technology Design Note 355*, January 2005.

History of current measurements

Measuring small currents is difficult. Measuring extremely small currents is extremely difficult and runs you into two problems: physics and noise. A recent newsgroup posting exemplifies the physics issue (Reference A). The message-poster asked for help measuring 1-fA currents and said he required a 1-MHz bandwidth, as well. Several respondents pointed out his problem with physics. A femtoampere is 6242 electrons/sec. There is no sense in measuring it with an instrument capable of 1 million events/sec. One respondent told the original poster: "You can't usefully use a 1-MHz bandwidth with a femtoamp of current, at least not if it has full-shot noise, which I imagine it does. If your current consists of N electrons per second, counting statistics predict that your SNR will drop to 0 dB in a measurement bandwidth of N/2 Hz. A femtoamp is only 6200 electrons/sec, so assuming that you want at least a 20-dB SNR (because below that [value], you haven't got a measurement really), your bandwidth is going to be limited to 31 Hz."

This answer alludes to the other problem with measuring extremely small currents: noise. The noise can manifest itself as a voltage noise or a current noise. In addition, there is noise from triboelectric effects, electrostatics, piezoelectric effects, and dielectric absorption. Even the mismatch of metals on your board can cause thermocouple effects that interfere with these low-level measurements (Reference B). You cannot simply slap a 1-GΩ resistor across the source and expect to turn that 1 fA into 1 μV. Experienced engineers know that 1 μV is difficult to measure in and of itself, but the more horrendous problem is the huge amount of voltage noise that large resistances generate. The voltage spectral density in a resistor is N=(4kTR)½ (with units of V/Hz), where k is Boltzmann's constant, 1.38×10^{-23}; T is absolute temperature in degrees Kelvin (approximately 300 at room temperature); and R is the resistance (Reference C). A 100-kΩ resistance has a voltage spectral density of 41 A 1-GΩ resistor has 41 With a 1-MHz bandwidth, the rms noise voltage that a 1-GΩ resistor creates is equivalent to V=N which means that the 41 becomes 41 mV of rms noise. The futility of trying to discern 1 μV in 41 mV of noise should be apparent to any analog-signal electrical engineer.

Limiting the bandwidth is therefore a central principle when measuring small currents. In centuries past, you could measure small currents with analog meters, which limited the bandwidth by virtue of the mass in the needle and armature. A Simpson 260 has a 50-μA scale that would easily let you repeatedly and accurately measure a few microamperes (Reference D). You should note that an analog meter directly measures current. The current passes though windings on the armature, which creates a magnetic field. You use this field to rotate the needle assembly. When you use an old analog meter to measure voltage, you are still really measuring current. You insert a precise resistor in series with the meter to convert the voltage into a current. A modern electronic voltmeter uses an integrating ADC to measure voltage. To measure current, the meter inserts a small shunt resistance into the current path so that it can measure the voltage drop across the shunt; it then converts that value to an equivalent current. Old analog meters, such as the mirrored galvanometer, could resolve 10 fA, so do not disparage these mechanical techniques.

One class of analog meters directly measures voltage. English chemist Edward Weston's 1894 patent of a metered electrometer was one such instrument. These instruments use an electrostatic force between plates to react against a spiral spring to give a voltage reading and draw only a minuscule leakage current from the source (Figure A, Reference E).

Once you address the problem of physics—that is, once your current is high enough and your bandwidth low enough—then you have to address noise. The mechanical meters of previous centuries did not have electronics to generate or amplify noise, and the mass of the meter movement damped out all the high-frequency noise. This technique is fine for dc current measurements, but measuring tiny ac currents demands the use of amplifiers.

An early method of measuring ac was the vibrating-reed electrometer. Anyone familiar with the operation of a condenser microphone will appreciate the principles this instrument uses. In a condenser microphone, a bias current charges a capacitor plate to a high level, and small sound vibrations of the plate diaphragm then create a usable voltage across the capacitor. A signal path amplifies this voltage, providing a low-noise representation of the sound. In a vibrating-reed electrometer, a mechanical arrangement moves the plate of the capacitor. Any small voltage you are measuring on the plate creates a voltage signal across the capacitor that you can amplify to represent the charge. The input impedance of this instrument is enormous, because the input pin is simply a metal plate suspended in air. The only bias current into or out of this pin is due to leakage across the surfaces touching the pin or across the air gap in the capacitor. The process realizes a further benefit, because it modulates the output signal at the vibrating-reed frequency. Because it is an ac signal, the dc noise of the amplifier is not a factor in the signal path. All amplifiers, even tube amplifiers, exhibit worse noise at low frequencies. Electronic amplifiers exhibit a classic noise curve (Figure B). The part of the curve at which the low-frequency, or "flicker," noise bends and becomes horizontal is the 1/f corner. If the vibrating reed operates above this frequency, ac-coupled amplifiers can filter out the noise below the 1/f corner. Once the process has sufficiently amplified the signal, synchronous demodulation, a peak-detecting circuit, or some other circuit that measures the envelope of the modulation frequency and rejects the carrier can turn it into a representation of the input charge. Before you discard this concept as antiquated, note that several studies are under way to use MEMS (microelectromechanical systems) to fabricate the vibrating reeds, likely giving new life to this method (Reference F).

In the late 1960s, Philbrick Researchers used this same ac-amplification principle in its varactor- bridge electrometer (Reference G). In this clever circuit, the input voltage changes the bias across two varactor capacitors. Varactor-diode capacitors change value with applied voltage. Figure C shows the how the two capacitors create a small ac voltage when the input voltage makes one varactor capacitance bigger and the other smaller. This process amplifies the ac voltage with all the benefits of rejecting the 1/f noise and then synchronously demodulates it with a diode ring. Analog Devices' 310 is an improved model of this varactor-bridge electrometer (Reference H). The company followed with the 311, adapting it for inverting applications, and then produced the 312, for noninverting applications (Reference I). ADI transferred the production of these and its other modules to Intronics, which still supports the modules for users who do not want to employ more modern parts (Reference J).

Even back in the late 1960s, Analog Devices alluded to low input-bias devices, such as the JFET and the MOSFET. Reference K and Reference L provide an excellent history of the development of these technologies and further information on electrometer design, respectively. In 1967, the quality of JFETs and MOSFETs was such that engineers were using varactor-bridge and tube circuits. As JFETs improved, amplifiers started taking advantage of the low current noise and low bias current inherent in JFET technology. Figure D, the classic current-detection circuit, shows that the bias current of the amplifier is a fundamental limitation of the resolution of low currents. Because the input of a JFET is a reversed-biased diode, JFETs exhibit leakage currents on the order of picoamperes. It is an unfortunate fact that this leakage doubles with every 10°C rise in temperature, so JFETs are not optimal in high-temperature environments. In the 1970s, analog companies made parts that used discrete JFETs in hybrid-type packages. In the 1975, National Semiconductor introduced the LF156, a part that integrated the JFETs into the amplifier itself. There is a further benefit of JFETs even over MOSFETs. Due to their structure in a monolithic die, JFETs are buried devices; the active areas are all below the surface of the die. This characteristic is desirable, because the surface of the die always contains lattice defects that contribute to noise. The downside of the buried JFET is that diffusion, not lithography, controls its parameters, so making perfectly matched pairs of JFETs is difficult.

Many analog companies make JFET amplifiers, including Texas Instruments, NEC, Linear Technology, Intersil, Fairchild, and Toshiba, but the industry reached a significant milestone in 1985 when Analog Devices' colleagues Lew Counts and JoAnn Close designed the AD549. This amplifier used monolithic JFETs, but they used a patented Topgate technique that electrically isolated the top- and back-side gates to achieve input-bias currents, on the order of 40 fA (Reference M). Furthermore, Counts and Close used a bootstrapped input structure so that the common-mode rejection of the part was exemplary. The data sheet (Reference N) is well worth reading for the applications section that describes the low-noise and low-leakage design techniques necessary when measuring femtoampere currents. Counts, now a fellow at Analog Devices, relates, "Even a DVM [digital voltmeter] will measure microamperes, and nanoampere measurements are almost routine. The ability to measure picoamperes with an op amp was something new." Indeed, the ADM1191 power monitor can resolve down to microamperes while measuring voltage.

Although JFETs have low leakage current, it is hard to beat CMOS MOSFETs for leakage. Because the gate is insulated with a layer of glass, the leakage is orders of magnitude lower than that of the JFET. The measurable leakage in CMOS parts is not due to the input MOSFET but rather the ESD (electrostatic-discharge) diodes required on every manufacturable part. These diodes have reverse leakage that does not exactly match between the top and bottom diode. That small difference enters or exits the input pin. This input bias is not only small, but also may be either into or out of the pin, depending on the part. Even parts of the same lot may have different polarities of bias current. JFET parts also have ESD diodes, but the leakage of the JFET usually masks the small leakage mismatch of the ESD diodes. Like JFET leakage, the unfortunate fact is that the ESD-diode leakage also increases greatly with temperature, approximately doubling every 10°C.

The drawback of MOS parts compared with JFETs was the earlier described surface noise. As analog semiconductor processes improved, the reduction in surface defects allowed the creation of CMOS parts that have low noise, such as National's LMV751 and Maxim's MAX4475. National's LMC660 has a typical bias current of 2 fA. Another part, Maxim's MAX4238 CMOS-input chopper-stabilized amplifier, has an input- bias current of 1 pA. The chopper action removes the 1/f noise, similar to the operation of a vibrating-reed electrometer or a varactor-bridge circuit. Be careful when using any chopper-stabilized amplifier with a high-impedance source; however, there could be ac-bias current due to charge injection of the internal chopper switches that may interfere with the measurement. As always, be cognizant of your source impedance and the voltage and current noise characteristics of your amplifier (Reference O).

References

A. "Amplifying a sub femtoamp of current," *sci.electronics.design*, July 24, 2006.
B. Frenzel, Louis E, "Accurately Measure Nanoampere And Picoampere Currents," *Electronic Design*, Feb 15, 2007.
C. "Noise analysis—resistor example".
D. www.simpson260.com.
E. Weston, Edward, Electrometer patent, No.530145.
F. Lee, J, LF Velásquez-García, A Seisha, AI Akinwande, "A MEMS Electrometer for Gas Sensing".
G. www.philbrickarchive.org.
H. Smith, LR, "A parametric Operational Amplifier," *Analog Dialog, Volume 1, No. 2*, June 1967, pg 6.
I. "New Varactor Bridge Op-Amps Models 302, 303," *Analog Dialog, Volume 1, No. 3*, September 1967, pg 9.
J. www.intronics.com/products/analogproducts.
K. Jung, Walt, "Op Amp History," *Op Amp Applications*, ISBN 0-916550-26-5, 2002.
L. Demrow, Bob, et al, "Op Amps as Electrometers or—The World of fA," *Best of Analog Dialog*, May 2, 1971, pg 48.
M. Counts, Lewis, and JoAnn Close, "Very Low Input Current JFET Amplifier," Patent No. 4639683, Jan 27, 1987.
N. www.analog.com/en/prod/0,,759_786_AD549,00.html.
O. "Femtoamp (fA) measurements," *Test & Measurement World*, May 1, 2002.