Auto-zero amplifiers ease the design of high-precision circuits

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A wide variety of electronic applications deal with the conditioning of small input signals. These systems require signal paths with very low offset voltage and low offset voltage drift over time and temperature. With standard linear components, the only way to achieve this is to use system-level auto-calibration. However, adding auto-calibration requires more complicated hardware and software and can slow down time to market for new products.

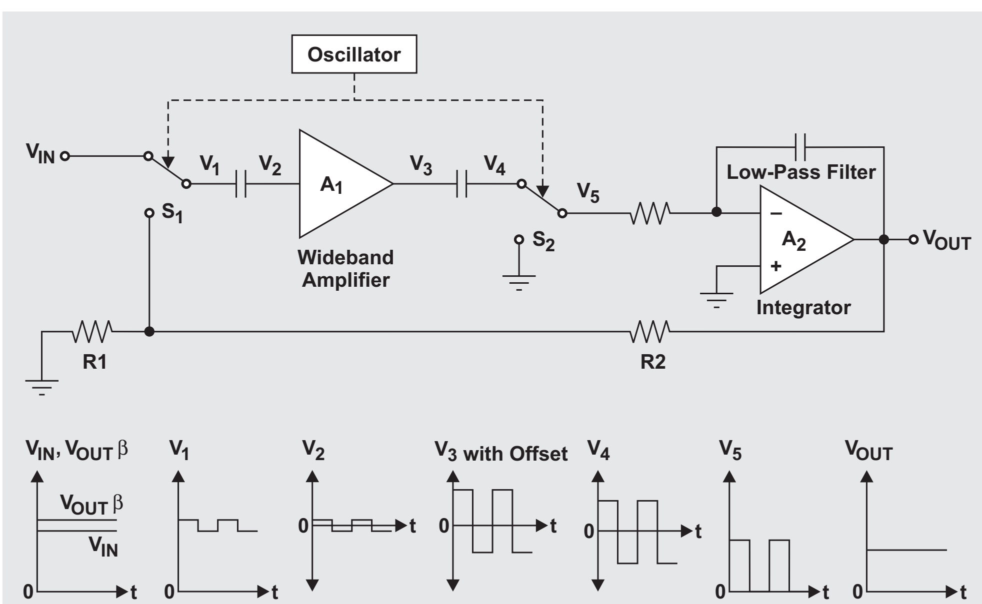
The alternative is to use components with low offset and low drift. The amplifiers with by far the lowest offset and drift available are the auto-zero amplifiers (AZAs). These amplifiers achieve high dc precision through a continuously running calibration mechanism that is implemented on-chip. With a typical input offset of 1 µV, a temperature-related drift of 20 nV/°C, and a long-term drift of 20 nV/month, these amplifiers satisfy even the highest requirements of dc accuracy.

Today's AZAs differ neither in form nor in the application from standard operational amplifiers. There is, however, some hesitation when it comes to using AZAs, as most engineers associate them with the older chopper amplifiers and chopper-stabilized amplifier designs. This stigma has been perpetuated either by engineers who worked with the older chopper amplifiers and remember the difficulties they had with them, or younger engineers who learned about chopper amplifiers in school but probably did not understand them very well.

The original chopper amplifier heralded the beginning of the new era of self-calibrating amplifiers more than 50 years ago. This amplifier provided extreme low values for offset and drift, but its design was complicated and expensive. In addition, ac performance was limited to a few hertz of input bandwidth accompanied by a high level of output noise. Over the years, unfortunately, the term "chopper amplifier" became a synonym for any amplifier with internal calibration capability. Therefore, AZAs, often wrongly designated as chopper or chopper-stabilized amplifiers, are associated with the stigma of the older chopper technique.

This article shows that the auto-zero calibration technique is very different from the chopper technique and is one that, when implemented through modern process technology, allows the economical manufacturing of wideband, high-precision amplifiers with low output noise. The following discussion presents the functional principles of the chopper amplifier, the chopper-stabilized amplifier, and the AZA. It then compares the efficiencies of low-frequency filtering when applied to AZAs and standard operational amplifiers. Finally, three application examples demonstrate the use of an AZA as a signal amplifier and as a calibrating amplifier in dc—and wideband ac—applications.

Figure 1. Chopping principle in the time domain



Note: Labels for amplifiers such as A_{IN}, G_B, and A_M are used to identify amplifiers in figures and to represent amplifier gain in equations.

The chopper amplifier

Figure 1 shows a simplified block diagram of a chopper amplifier.

A dc input signal is chopped into an ac voltage and amplified by an ac-coupled amplifier. A phase-sensitive demodulator converts the output of A_1

back to dc. The demodulator consists of a switch S_2 that is synchronously driven to S_1. An integrator then smooths the switch output and presents the final dc output.

The circuit benefits from high overall dc gain and low baseband noise. The dc gain, being the product of the ac stage and the dc gain of the integrator, easily reaches an open-loop gain of 160 dB and reduces the gain error,

$$\frac{V_{OUT}}{A_1 A_2},$$

to almost zero.

The baseband is defined as the maximum usable input bandwidth. Baseband noise consists of the input offset voltage (also known as dc noise), the 1/f noise, and low-frequency white noise. The reduction of baseband noise happens in several steps:

- Offset and drift in the output integrator stage are nulled by the dc gain of the preceding ac stage.
- dc drifts in the ac stage are also irrelevant because they are isolated from the rest of the amplifier by the coupling capacitors.
- The 1/f noise of the ac amplifier is modulated to higher frequencies via the demodulator.

Figure 2 clarifies the process of noise reduction by demonstrating the effects of chopping in the frequency domain.

The chopping of the input signal constitutes an amplitude modulation (AM), with the chopping frequency, f_{CH}, being the carrier, and the input voltage representing the modulating signal. Both switches, S_1 and S_2, are replaced by the modulators, M_1 and M_2.

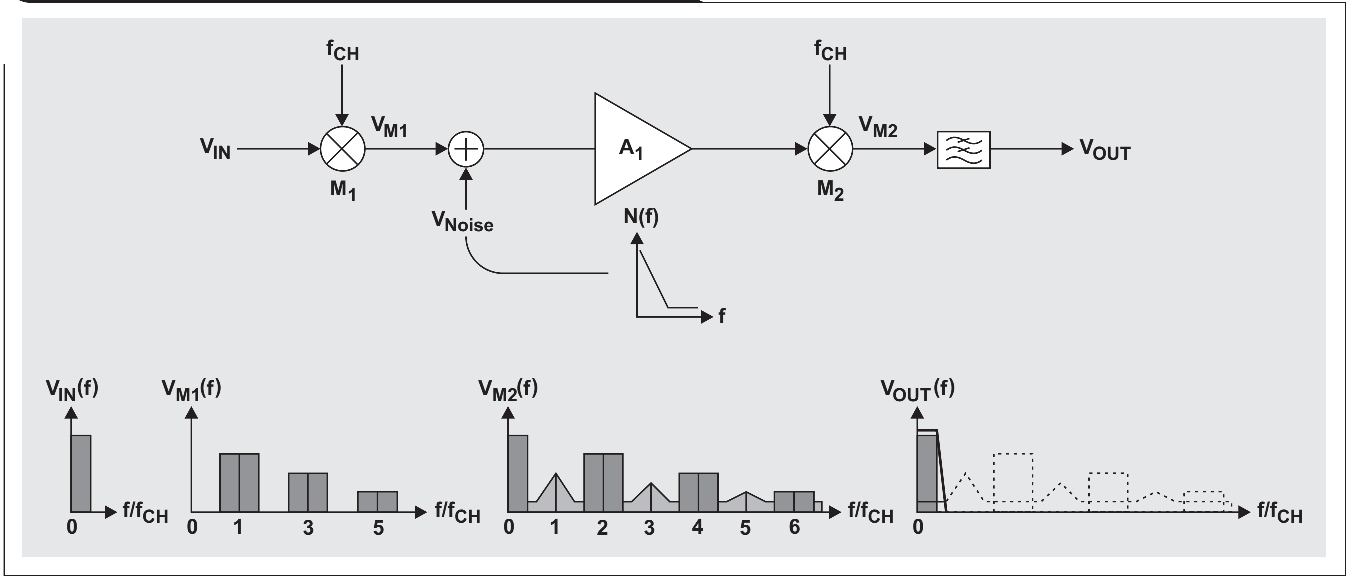
$V_{M1}(f)$ in Figure 2 shows that the modulation of a square wave causes sidebands of the input signal to appear on both sides of the odd harmonics of the chopper frequency. The amplitudes of the harmonics and their sidebands decrease following a 1/n function, with n indicating the order of the harmonic.

The 1/f noise of A_1 present in the baseband adds to the modulated input signal after the first modulation stage, M_1. The combined signal is amplified by A_1 and fed into the demodulator, M_2. The 1/f noise, experiencing its first modulation through M_2, introduces sidebands on both sides of the odd harmonics of f_{CH}. For the modulated input signal, however, M_2 represents the second modulating stage. V_{M1} is now demodulated, causing sidebands of the input signal to occur around the even harmonics of f_{CH}. The input signal reappears in the baseband, and the roll-off of the subsequent low-pass filter limits the baseband to frequencies far below the chopper frequency.

Note that the AM does not change the spectral density of the white noise. The residual baseband noise is therefore limited, low-frequency white noise.

Despite the small values for offset, drift and baseband noise, this approach has some drawbacks. First, the amplifier has a single-ended, noninverting input and cannot accept differential signals without additional circuitry at the front end. Second, the carrier-based approach constitutes a sampled data system, and overall amplifier bandwidth is limited to a small fraction of the chopper frequency. The chopper frequency, in turn, is restricted by ac amplifier gain-phase limitations and errors induced by switch response time. Maintaining good dc performance involves keeping the effects of these considerations small. Chopper frequencies are therefore in the low-kilohertz range, dictating low overall bandwidth.

Figure 2. Chopping principle in the frequency domain



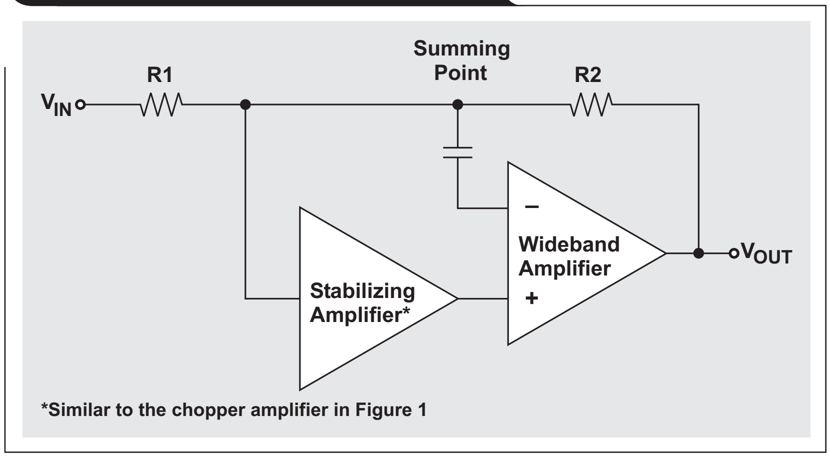
The chopper-stabilized amplifier

The classic chopper-stabilized amplifier solves the chopper amplifier's low-bandwidth problem. It uses a parallel path approach (Figure 3) to provide wider bandwidth while maintaining good dc characteristics. The stabilizing amplifier, a chopper type, biases the fast amplifier's positive terminal to force the summing point to zero.

Fast signals directly drive the ac amplifier, while slow ones are handled by the stabilizing chopper amplifier. The low-frequency cutoff of the fast amplifier must coincide with the high-frequency roll-off of the stabilizing amplifier to achieve smooth overall gain-frequency characteristics. With proper design, the chopper-stabilized approach yields bandwidths of several megahertz with the low-drift characteristic of the chopper amplifier. Unfortunately, because the stabilizing amplifier controls the fast amplifier's positive terminal, the classic chopper-stabilized approach is restricted to inverting operation only.

In addition, the high residual output noise of the chopper amplifier is amplified by the fast amplifier's noise gain. Keeping output noise small dictates additional filter effort, thus increasing complexity and cost of the chopper-stabilized design.

Figure 3. Chopper-stabilized amplifier



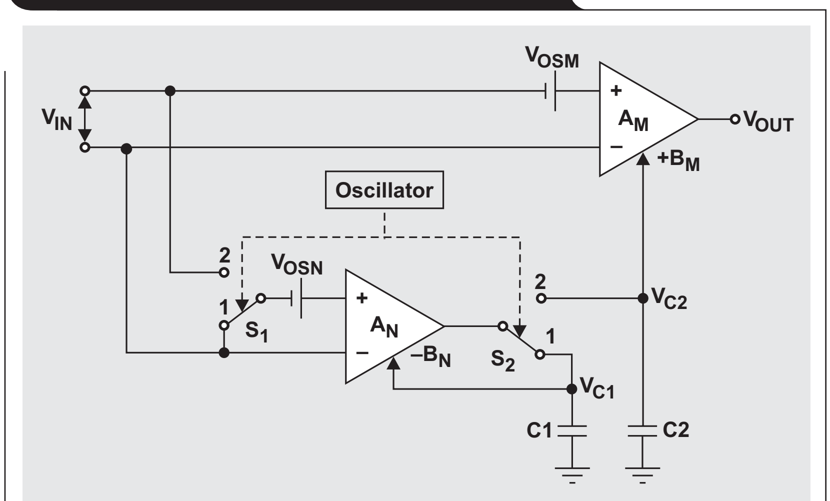
*Similar to the chopper amplifier in Figure 1

The auto-zero amplifier (AZA)

Similar to the chopper-stabilized approach, the AZA uses a main amplifier for wideband signal amplification and a nulling amplifier for offset correction. Figure 4 shows a block diagram of the TLC2654, an AZA developed by Texas Instruments in the mid-80s.

With the calibration path lying in parallel with the signal path, both inputs of the main amplifier are available for differential input operation.

Figure 4. Simplified TLC2654 block diagram



Note: Labels for amplifiers such as A_{IN}, G_B, and A_M are used to identify amplifiers in figures and to represent amplifier gain in equations.

The main amplifier, A_M, and the nulling amplifier, A_N, each have an associated input offset voltage (V_{OSM} and V_{OSN}, respectively) modeled as a dc offset voltage in series with the noninverting input. The open-loop gain of the signal inputs is given as A_M and A_N. Both amplifiers also have additional voltage inputs with the associated open-loop gains of $+B_M$ and $–B_N$.

Offset correction of the overall amplifier occurs within one cycle, f_{AZ}, of the auto-zero clock and is split into two modes of operation: an auto-zero phase and an amplification phase. The oscillator, generating f_{AZ}, initiates the auto-zero phase by driving both switches into position 1. The inputs of the nulling amplifier are shorted together, while its output is connected to capacitor C1. In this configuration A_N measures its input offset voltage and stores it via C1. Mathematically we can express the voltage at C1 as

$$V_{C1} = A_N V_{OSN} - B_N V_{C1},$$

which, by simple rearrangement, is

$$V_{C1} = V_{OSN}\left(\frac{A_N}{1+B_N}\right). \quad (1)$$

This shows that the offset voltage of the nulling amplifier times a gain factor appears at the output of A_N and thus on the C1 capacitor.

In the amplification phase, when both switches are in position 2, this offset voltage remains on C1 and essentially corrects any error from the nulling amplifier. A_N amplifies

V_{C1} by the factor B_N and subtracts it from the amplified input signal,

$$A_N(V_{IN} + V_{OSN}).$$

At the same time, the output of A_N charges capacitor C2 to

$$V_{ON} = V_{C2} = A_N(V_{IN} + V_{OSN}) - B_N V_{C1}.$$

Replacing V_{C1} with Equation 1 results in

$$V_{C2} = A_N\left(V_{IN} + \frac{V_{OSN}}{1+B_N}\right). \quad \textbf{(2)}$$

Equation 2 shows that V_{OSN} has been reduced by a factor 1 + B_N, indicating how the nulling amplifier reduces its own offset voltage error even before correcting the main amplifier. The potential, V_{C2}, now serves the main amplifier as an offset correcting voltage, forcing its output, and thus the output of the complete AZA, to

$$V_{OUT} = A_M(V_{IN} + V_{OSM}) + B_M V_{C2}.$$

Replacing V_{C2} with Equation 2 and combining terms gives us

$$V_{OUT} = V_{IN}(A_M + A_N B_N) + V_{OSM} A_M + V_{OSN}\left(\frac{A_N B_M}{1+B_N}\right). \quad \textbf{(3)}$$

The auto-zero architecture is optimized in such a way that $A_M = A_N$, $B_M = B_N$, and $B_N >> 1$. This allows Equation 3 to be simplified to

$$V_{OUT} = V_{IN} A_N B_N + A_N(V_{OSM} + V_{OSN}). \quad \textbf{(4)}$$

Most obvious is the gain product of both the main and nulling amplifiers. The $A_N B_N$ term in Equation 3 explains why AZAs have extremely high open-loop gain. To understand how V_{OSM} and V_{OSN} relate to the overall effective input offset voltage of the complete amplifier, we should set up the equation for the generic amplifier in Figure 5:

$$V_{OUT} = k(V_{IN} + V_{OS\_Eff}), \quad \textbf{(5)}$$

where k is the open-loop gain of the amplifier and V_{OS_Eff} is its effective offset voltage.

Putting Equation 4 into the form of Equation 5 gives us

$$V_{OUT} = A_N B_N\left(V_{IN} + \frac{V_{OSM} + V_{OSN}}{B_N}\right).$$

From here it is easy to see that k = $A_N B_N$ and

$$V_{OS\_Eff} = \frac{V_{OSM} + V_{OSN}}{B_N}.$$

Thus, the offset voltages of both the main and the nulling amplifiers are reduced by the gain factor B_N. If we consider the open-loop gains of the local amplifiers, A_N and A_M, to be in the region of 10,000 or higher, it quickly becomes evident that even an inherent offset voltage of millivolts is reduced to an effective input offset voltage of microvolts for the complete AZA.

The AZA constitutes a sampled data system. The process of sampling therefore generates frequencies consisting of the sum and difference of the input signal frequency, f_S, and the auto-zero clock frequency, f_{AZ}. The summing frequency, $f_{AZ} + f_S$, can be filtered easily and is therefore of little importance. However, the difference frequency, $f_{AZ} - f_S$, can alias into the baseband if $f_S \geq f_{AZ}/2$. Older AZA designs therefore required the limitation of the input bandwidth to less than half of the auto-zero frequency. Most of the amplifiers available in the mid-80s had typical clock frequencies in the range of only 400 to 500 Hz, thus narrowing the signal bandwidth down to 250 Hz. The TLC2654 was one of the first amplifiers that allowed high-frequency auto-zeroing at 10 kHz, thus extending the input bandwidth up to 5 kHz.

The breakthrough to real wideband operation happened only with the recent introduction of AZAs such as the OPA335. Modern process technology, with gate structures in the submicron region, made the economical integration of complex anti-aliasing circuitry possible. The strong attenuation of alias frequencies enabled wideband operation across the entire amplifier bandwidth.

Figure 5. Generic amplifier with effective offset

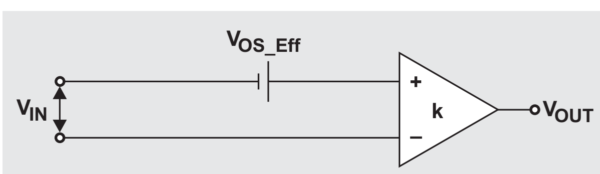


Figure 6 shows the inner structure of the OPA335. The two nulling amplifiers, A_{N1} and A_{N2}, operate in an alternate mode in parallel with the main amplifier, A_M. While A_{N1} nulls its offset during the auto-zero phase, A_{N2} is in the amplification phase, correcting the main amplifier's offset voltage and vice versa.

The alternating operation of the nulling amplifiers minimizes output voltage ripple and intermodulation distortion (IMD) by keeping the amplifier's gain bandwidth constant during operation. Proprietary circuit design has made further improvements to the nulling amplifiers. Each amplifier consists of a multistage composite amplifier. This configuration drastically reduces the quiescent current down to 300 µA (versus the 1.5 mA of the TLC2654) while maintaining a high open-loop gain of 130 dB. In addition, the previous external capacitors have been made redundant by achieving the same effective capacity values through Miller equivalence.

Let's return to the process of auto-zeroing. The nulling amplifier, whose switches are in position 1, is in the auto-zero phase, thus charging its capacitor to

$$V_C = G_B(A_{IN}V_{OSN} - A_Z V_C) = V_{OSN}\left(\frac{G_B A_{IN}}{1+G_B A_Z}\right). \quad \textbf{(6)}$$

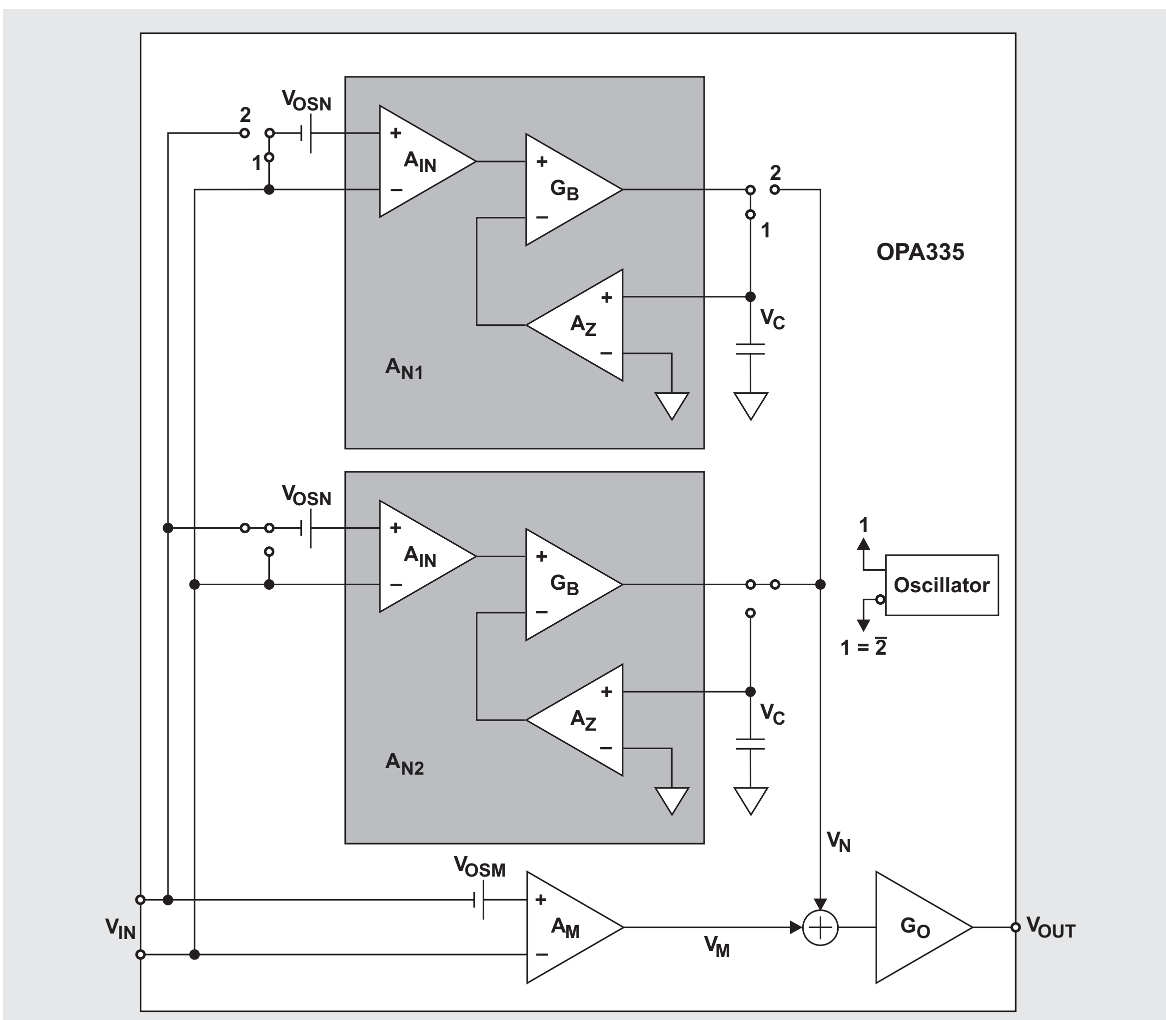
During the amplification phase (with switches in position 2), the output voltage of the nulling amplifier, V_N, adds to the output voltage of the main amplifier, V_M. With

$$V_N = G_B\left[A_{IN}(V_{IN}+V_{OSN}) - A_Z V_C\right],$$

we can replace V_C with Equation 6 to obtain

$$V_N = G_B A_{IN}\left(V_{IN} + \frac{V_{OSN}}{1+G_B A_Z}\right). \quad \textbf{(7)}$$

Figure 6. Simplified OPA335 block diagram



Note: Labels for amplifiers such as A_{IN}, G_B, and A_M are used to identify amplifiers in figures and to represent amplifier gain in equations.

The output of the main amplifier is simply

$$V_M = A_M\left(V_{IN} + V_{OSM}\right). \quad \textbf{(8)}$$

The following output stage amplifies the summing signal by the factor G_O to

$$V_{OUT} = G_O(V_N + V_M). \quad \textbf{(9)}$$

Inserting Equations 7 and 8 into Equation 9 yields

$$V_{OUT} = G_O\left[A_M(V_{OSM} + V_{OSN}) + G_B A_{IN}\left(V_{IN} + \frac{V_{OSN}}{1+G_B A_Z}\right)\right]. \quad \textbf{(10)}$$

During the design process, the auto-zero structure is so optimized that $A_M = A_N$ and $G_B >> 1$. This simplifies Equation 10 to

$$V_{OUT} = G_O G_B A_{IN}\left(V_{IN} + \frac{V_{OSM} + \frac{V_{OSN}}{A_Z}}{G_B}\right),$$

with $G_O G_B A_{IN}$ as the overall open-loop gain and

$$\frac{V_{OSM} + \frac{V_{OSN}}{A_Z}}{G_B}$$

as the effective input offset voltage of the complete AZA.

Output noise filtering

High dc accuracy often requires additional noise filtering. How low a filter's 3-dB frequency can be to provide effective noise filtering is all too often ignored. A comparison between a standard CMOS op amp and an AZA provides valuable insight.

Figure 7 shows that the spectral noise density of a standard op amp consists of two noise characteristics: the 1/f region where noise density decreases with increasing frequency, and the white noise region with constant spectral density. At corner frequency f_C, the magnitude of 1/f noise equals the magnitude of white noise density.

For signal-to-noise ratio calculations we require the rms value of the noise within a defined frequency band. Applying a first-order low-pass filter with a –3-dB cutoff at frequency f_H yields an rms noise voltage of

$$V_{rms} = v_{nw}\sqrt{f_C \ln\frac{f_H}{f_L} + 1.57 f_H - f_L},$$

where v_{nw} is the white noise spectral density, f_C is the corner frequency of the 1/f- and white-noise transition, f_H is the upper frequency of the noise frequency band and filter cutoff, and f_L is the lower frequency of the noise frequency band (here assumed to be 10^{-30} Hz).

Various rms voltages have been calculated with the preceding equation by varying the –3-dB frequency of a first-order low-pass filter. The resulting plot is shown in Figure 8. It is important to notice that lowering the filter's cutoff below $10 f_C$ seems inefficient, since the rms noise hardly decreases.

In contrast to a standard op amp, the continuous offset cancellation of an AZA removes the typical 1/f characteristic and creates the white noise spectral density in Figure 7 instead. Applying the same low-pass filter provides an rms noise of

$$V_{rms} = v_{nw}\sqrt{1.57 f_H}.$$

Plotting the rms values for various cutoff frequencies results in the positive slope in Figure 8. It can be seen that reducing the filter cutoff down to low frequencies is effective in establishing high dc accuracy.

One contributor to high-frequency output noise is clock feedthrough. This term is broadly used to indicate visibility of the auto-zero clock frequency in the amplifier output spectrum. There are typically two types of clock feedthrough. The first is caused by the settling time of the

Figure 7. Spectral noise density

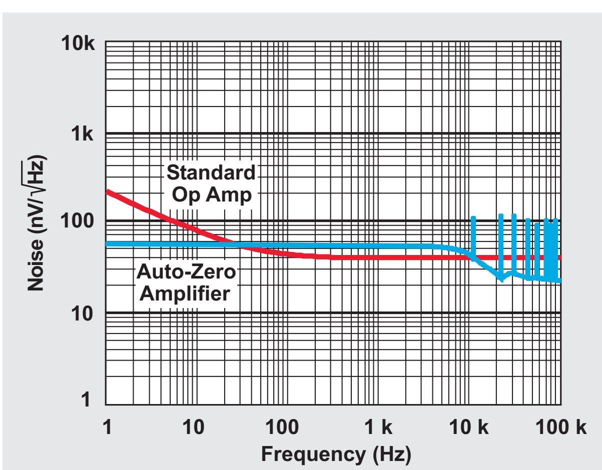


Figure 8. rms noise voltage

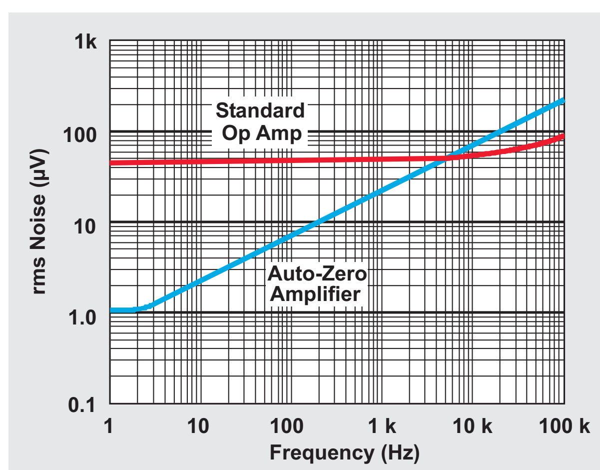
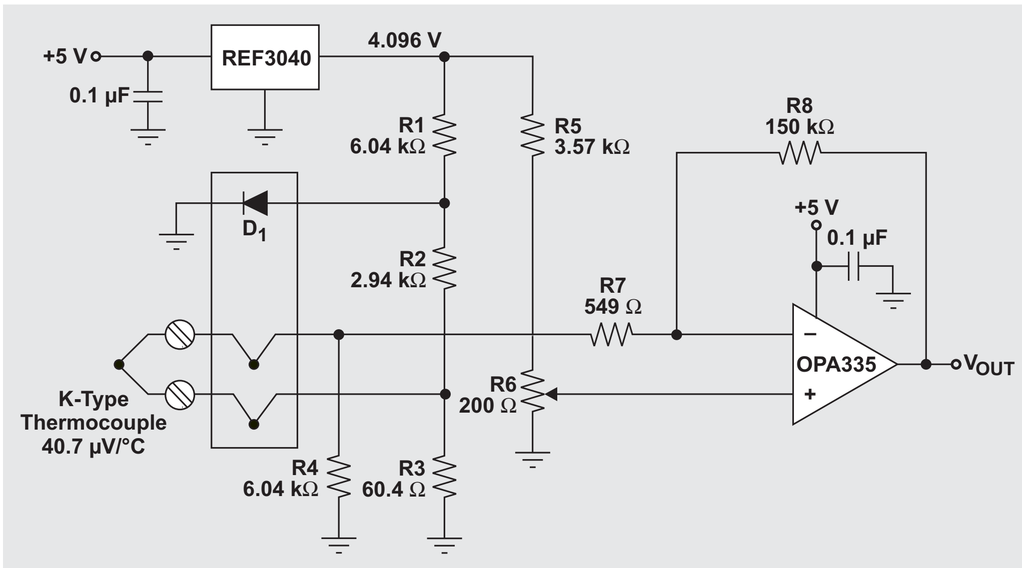


Figure 9. Temperature measurement via thermocouple



internal sampling capacitors and is input-referred. The second is caused by the small amount of charge injection occurring during the sampling and holding of the amplifier's input offset voltage.

The OPA335, however, has remarkably little noise. Although zero correction occurs at a 10-kHz rate, there is virtually no fundamental noise energy present at that frequency. For all practical purposes, any glitches have energy at 20 MHz or higher and are easily filtered if required. Most applications are not sensitive to such high-frequency noise, and no filtering is required.

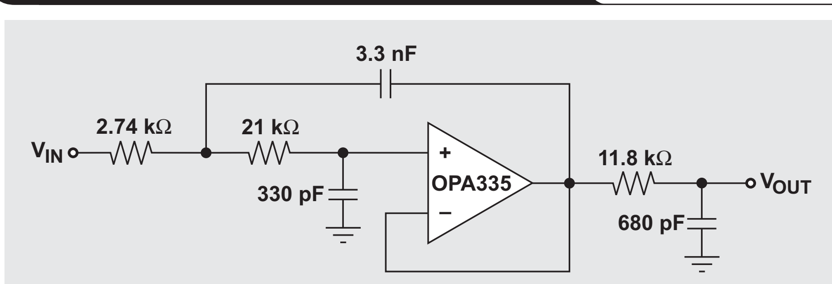
Applications

The temperature measurement circuit in Figure 9 is a low-frequency application that allows the OPA335 to be switched directly into the signal path.

A precision voltage reference provides the 4.096-V bridge supply. The forward voltage of diode D1 has a negative temperature coefficient of –2 mV/ºC and provides cold-junction compensation via the resistor network R1 to R3. The zero adjustment for a defined minimum temperature is achieved via R6, while R7 and R8 set the gain for the output amplifier. The single-supply amplifier providing an open-loop gain of 130 dB allows 16-bit or better accuracy at high gain in low-voltage applications. Auto-zeroing removes 1/f noise and provides typical values of 1 µV of input offset and 20 nV/ºC of offset drift over temperature. Thus, AZAs ideally suit single-supply precision applications where high accuracy, low drift, and low noise are imperative.

The third-order low-pass filter in Figure 10 has a corner frequency of 20 kHz, which is twice as high as the auto-zero clock frequency. Aliasing and intermodulation noise are highly attenuated, which permits input signal operation across its entire gain bandwidth. In addition, the amplifier's output provides rail-to-rail drive capability, allowing for a high signal-to-noise ratio at low supply voltages.

Figure 10. A 20-kHz, third-order low-pass filter



In wideband applications with bandwidths in the tens of megahertz, the AZA provides dc accuracy to a wideband amplifier. Figure 11 shows the required circuit configuration in the form of a composite amplifier design.

The AZA functions as an integrator operating in the "bias path" of the wideband amplifier. The signal path still runs from V_{IN} via R_G and R_F to V_{OUT}. The integrator has two functions. At low frequencies, it provides high gain to the offset-cancellation loop, reducing the input offset of the wideband amplifier down to the input offset of the AZA. At high frequencies, a large time constant (R_{INT}, C_{INT}) ensures that the integrator's closed-loop gain quickly decreases to prevent signal transfer to the noninverting input of the wideband amplifier.

Note that the amplifier's input noise is amplified by the noninverting closed-loop gain of the integrator. Thus, at high frequencies, the OPA335 operates as a voltage follower (gain = 1), passing its input noise on to the wideband amplifier. To eliminate this noise, a low-pass filter (R2, C2) with low-frequency cutoff is added to the output of the AZA. The same precaution is taken for the OPA353. Here the low-pass filter (R1, C1) limits the output noise of the wideband amplifier.

To compensate for the signal voltage drop across R1, the feedback loop is closed by connecting the right side of R_F to the filter output. The internal feedback loop via C_F establishes stability at high frequencies by compensating the phase shift of the low-pass filter.

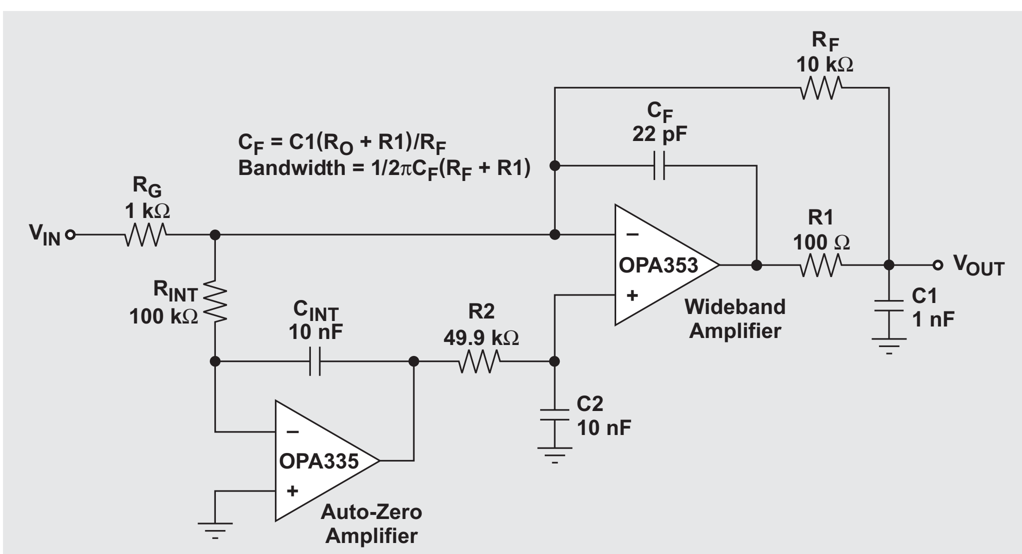
Summary

True chopper- and chopper-stabilized amplifiers perform offset correction through amplitude modulation. These amplifiers are not available as integrated circuits but require multiple amplifier integrated circuits instead. The circuit design is therefore complicated and time- and cost-intensive. Despite the extreme low values for input offset voltage and drift, ac performance is limited to a small fraction of the chopper frequency and is accompanied by high levels of output noise.

AZAs perform offset correction by a sample-and-hold method. Older-generation amplifier designs benefited from integrated circuit design. Aliasing and IMD, however, narrowed the input bandwidth down to half the auto-zero clock frequency. These devices required supply voltages of 10 V minimum and had quiescent currents in the range of milliamperes. In addition, external capacitors were needed to store the offset-correcting voltages.

Today's AZAs are by far the most sophisticated precision amplifiers available. Advancements in process technologies have drastically lowered the effects of aliasing and IMD, thus enabling true differential signal operation across the entire gain bandwidth of the amplifier. Proprietary circuit design has reduced the amount of supply voltage and quiescent current significantly, allowing for low-power operation in high-gain, high-precision applications. The integration of the external storage capacitors in combination with the given performance enhancements makes AZAs as easy to use as standard CMOS op amps.

Figure 11. Auto-zeroed wideband amplifier



In addition to operational amplifiers, the auto-zero technique has also been implemented in a wide range of other signal conditioning components such as:

- the OPA734/735 op amps with extended supply range from 2.7 to 12 V;
- the OPA380 wideband transimpedance amplifier;
- the INA326 instrumentation amplifier for single-supply applications;
- the INA330 instrumentation amplifier for constant temperature control; and
- the PGA309, a fully integrated pressure sensor conditioning system on-chip.

Future design ideas aim to shape the noise floor of extremely high-gain-bandwidth amplifiers by shifting energy from the baseband to higher, out-of-band frequencies. This would approach the ideal op amp—an interesting concept that may not be far in the future.

Related Web sites

amplifier.ti.com
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Replace *partnumber* with INA326, INA330, OPA335, OPA353, OPA380, OPA734, OPA735, PGA309, REF3040, or TLC2654

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