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Learn The Limitations Of Low-Pass Sallen-Key Filters

Imperfect Amplifiers, Parasitic Capacitance, And Component Selection All Impact Filter Performance At High Frequencies.

Contributing Author | Dec 16, 1999

Since professors R.P. Sallen and E.L. Key described it in 1955, the Sallen-Key low-pass filter has become one of the most widely used filters in electronic systems. Perhaps because the mathematics can be somewhat daunting, however, little has been written to help working engineers specify the correct components to achieve their objectives. For example, few realize the limitations of Sallen-Key filters at high frequencies.

The following describes the basic operations of a Sallen-Key low-pass filter and offers a simplified way of working with such circuits. Based on laboratory research, it also demonstrates some of this filter's limitations at high frequencies.

larger than the first. But most filters usually require larger Qs than one-half.

Q can be enhanced with an amplifier in positive feedback. With that feedback localized to the filter's cutoff frequency, almost any Q can be realized. Mostly, it's only limited by the physical constraints of the power supply and component tolerances. The Sallen-Key low-pass filter shown in Figure 2 is an example of how an amplifier is used in this manner. C2 is no longer connected to ground, but rather provides a positive feedback path around the amplifier.

The operation can be described qualitatively. At low frequencies, where C1 and C2 appear as open circuits, the signal is simply amplified to the output. R3 and R4 are chosen to give the desired gain. At high frequencies, C1 and C2 appear as short circuits, and the signal is shunted to ground at the amplifier's input. The amplifier amplifies this input to its output, and the signal doesn't appear at V_O. Near the cutoff frequency, where the impedance of C1 and C2 are on the same order as R1 and R2, positive feedback via C2 provides Q enhancement of the signal.

Ideal operation: The standard frequency-domain equation for a second-order low-pass filter is:

where f_C is the corner frequency and Q is the quality factor.

When $f<C$, Equation 1 reduces to $H_{LP} = K$, and the circuit passes signals multiplied by gain factor K. When $f = f_C$, Equation 1 reduces to: $H_{LP} = –jKQ$, and the signals are enhanced by the factor Q. When $f>>f_C$, Equation 1 reduces to:

Deriving the transfer function of the circuit in Figure 2, the Sallen-Key ideal low-pass transfer function is defined by Equation 2.

By letting

Equation 2 follows the same form as Equation 1. With simplifications, you can deal with the equation more easily.

Simplification 1: Set filter components as ratios. Letting R1 = mR, R2 = R, C1 = C, and C2 = nC, results in:

This simplifies things somewhat, but there's interaction between f_C and Q.

The design should start by setting the gain and Q based on m, n, and K, and then selecting C and calculating R to set f_C. It may be observed that K = 1 +\[(m+1)/(mn)\] results in Q = ∞. With larger values, Q becomes negative. In other words, the poles move into the right half of the s-plane and the circuit oscillates. The most frequently designed filters require low Q values, so this

This keeps the gain equal to 1 in the pass band. But again, there's interaction between f_C and Q. Design should start by choosing the ratios m and n to set Q, and then selecting C and calculating R to set f_C.

Simplification 3: Set resistors as ratios and capacitors equal. Letting R1 = mR, R2 = R, and C1 = C2 = C, results in:

The main motivation behind setting the capacitors equal is the limited selection of values in comparison with resistors. Interaction exists between setting f_C and Q. Design should start with choosing m and K to set the gain and Q of the circuit before choosing C and calculating R to set f_C.

Simplification 4: Set filter components equal. Letting R1 = R2 = R and C1 = C2 = C results in:

Now f_C and Q are independent of one another. Design is greatly simplified, although it's simultaneously limited. Q is now determined by the gain of the circuit. The choice of RC sets f_C. The capacitor should be chosen, and the resistor calculated. One minor drawback is that because the gain controls the Q of the circuit, further gain or attenuation may be necessary to achieve the desired signal gain in the passband.

Values of K that are very close to 3 result in high Qs that are sensitive to variations in the component values of R3 and R4. Setting K = 2.9 results in a low Q values, so this should rarely become a design issue.

Non-ideal circuit operation: Up to now, we've assumed that the circuit was ideal, but there comes a time (or actually a frequency) when this is no longer valid. Simple logic tells us that the amplifier must be an active component at the frequencies of interest or else we have problems. But what are these problems?

As mentioned previously, there are three basic modes of operation: below cutoff, above cutoff, and in the area of cutoff. Assuming that the amplifier has adequate frequency response beyond cutoff, the filter works as expected. At frequencies well above cutoff, the high-frequency model depicted in Figure 3 is used to show the expected circuit operation. The assumption made here is that C1 and C2 are effective shorts when compared to the impedance of R1 and R2, so the amplifier's input is at ac ground. In response, the amplifier generates an ac ground at its output limited only by its output impedance, Z_O. The formula shows the transfer function of this particular model.

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Z_O is the closed-loop output impedance. It depends on the loop transmission and the open-loop output impedance, z_o:

where a(f) is the open-loop gain of the amplifier and b is the feedback factor. This feedback factor is constant—set by resistors R3 and R4. But the open-loop gain, a(f), depends on frequency.

With dominant-pole compensation, the amplifier's open-loop gain decreases by 20 dB/decade over the usable frequencies of operation. Assuming z_o is mainly resistive (usually a valid assumption up to a few hundred megahertz), Z_O increases at a rate of 20 dB/decade. The transfer function appears to be a firststray capacitance.

Simulation and lab data: To show the effects described above, a Sallen-Key low-pass filter was simulated in Spice and lab tested using a THS3001 operational amplifier from Texas Instruments. The THS3001 is a high-speed, current-feedback amplifier with an advertised bandwidth of 420 MHz. Choosing R1 = R2 = 1 kΩ, C1 = C2 = 1 nF, R3 = open, and R4 = 1 kΩ results in a low-pass filter with f_C = 159 kHz and Q = 1/2.

Figure 4a depicts the simulation circuit with the Spice model modified so that the output impedance of the amplifier is 0 Ω. In Figure 5, curve (a) shows the frequency response as simulated in Spice. It also reveals that with zero output impedance, the attenuation of the signal continues to increase as frequency rises.

Figure 4b depicts the high-frequency model as exemplified in Figure 3, where the input is at ground and the output impedance controls the transfer function. The Spice model used for the THS3001 includes an LRC network for the output impedance. Again, Figure 5 shows the frequency response as simulated in Spice, but this time it's symbolized by curve (b). The magnitude of the signal at the output is seen to cross curve (a) at about 7 MHz. Above this frequency, the output impedance causes the switch in the transfer function, which is described above.

Look to Figure 4c for the simulation circuit using the Spice model with the LCR output impedance. Figure 5's curve (c) shows the frequency response for this model. With the output impedance, the attenuation caused by the circuit follows curve (a) until it crosses curve (b), at which point it follows curve (b). Figure 4d reveals the circuit as tested in the lab, with curve (d) in Figure 5 showing that the measured data agrees with the simulated data.

Comments about component selection: Until now, the choosing of resistor

In the case of the low-pass Sallen-Key filter, the ratio between the output impedance of the amplifier and the value of filter component R sets the transfer functions seen at frequencies well above cutoff. The larger the resistor's value, the lower the transmission of signals at high frequency. Making R too large may result in C becoming so small that the parasitic capacitors, including the input capacitance of the amplifier, cause errors. The best choice of component values depends on the particulars of your circuit and the tradeoffs you're willing to make.

Here are some general recommendations for capacitors and resistors: Engineers should avoid capacitors with values less than 100 pF. If at all possible, use an NPO type. X7R is okay in a pinch, but avoid Z5U and other low-quality dielectrics. In critical applications, even higher-quality dielectrics, like polyester, polycarbonate, Mylar, etc., may be required. As for resistors, values in the range of a few hundred to a few thousand ohms are the best bet. You also should choose metal-film resistors that possess low temperature coefficients. Finally, use 1%-tolerance capacitors and resistors, preferably those of the surface-mount variety.

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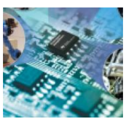

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Driverless Cars: Can We Accurately Regulate an Emerging Technology?

At the International New York Auto Show's Empire State of Mobility Conference, Congressman Dan Lipinski talked about the benefits of self-driving cars and future regulations.

Maria Guerra | Apr 20, 2017

during his presentation *(Fig. 1)*.

Much of the address focused on the benefits of driverless-car technology for our society and how new modes of mobility will transform the transportation system. He also emphasized the need for appropriate federal regulations, noting that they could safely speed up the rollout of this emerging technology.

1. Illinois Congressman Dan Lipinski delivers his keynote at The Empire State of Mobility Conference.

Proponents of driverless-vehicle technology say that such vehicles will likely improve road safety, as most accidents result from human driver errors. They also point out that driverless cars have the potential to reduce traffic congestion

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Say Farewell to Offset Voltage with a Zero-Drift Op Amp

Sponsored by: Texas Instruments Zero-crossover devices help eliminate the problem of offset-voltage performance degradation in low-voltage precision applications.

Paul Pickering | Apr 20, 2017

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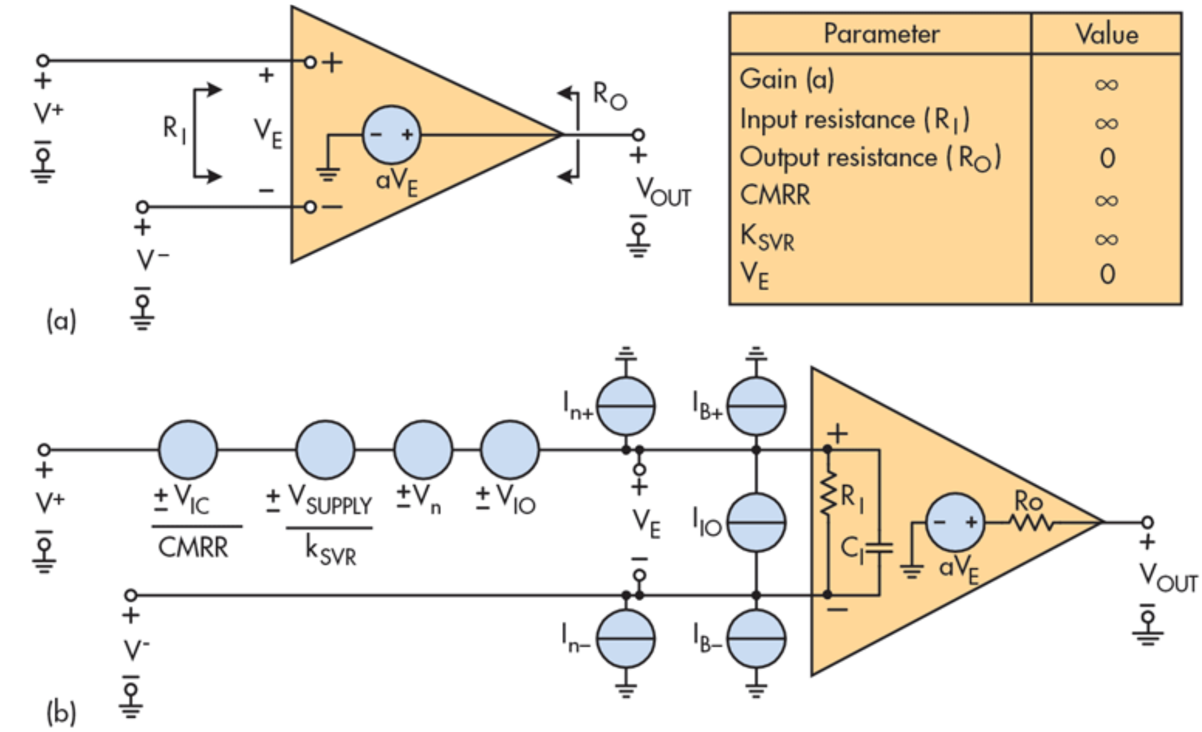
As analog engineers, it's tempting to fantasize about one day coming close to designing the ideal operational amplifier we encountered in college: infinite bandwidth, zero output resistance, and so on. Although it's a good first-order approximation to simplify analysis, the ideal op amp will remain forever just out of reach, together with ending world hunger, eliminating the National Debt, and the Cubs winning another World Series. Okay, scratch that last one.



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. The model of a real-world op amp (b) is considerably more complex than its deal equivalent (a). (Source: Texas Instruments)

Offset Voltage

Offset voltage (V_{OS}) is the differential dc voltage required between the input pins of an op amp to make its output zero. Our ideal op amp has zero volts across its negative or positive in polarity and vary from die to die in the same wafer lot.

Input offset voltage is primarily due to the inherent mismatch of the input transistors and components during die fabrication. Stresses placed on the die during packaging also make a minor contribution. These factors cause a mismatch of the bias currents flowing through the input circuit of the op amp, resulting in a voltage differential at the input terminals.

In a typical application, the differential voltage is multiplied by the closed-loop gain of the op amp, which is determined by the external components. Over the years, better matching of components and improved package materials have slowly reduced offset voltages. A trimming procedure during manufacturing is used to reduce the offset voltage, and op amps that exceed the maximum allowable offset voltage after trimming are rejected.

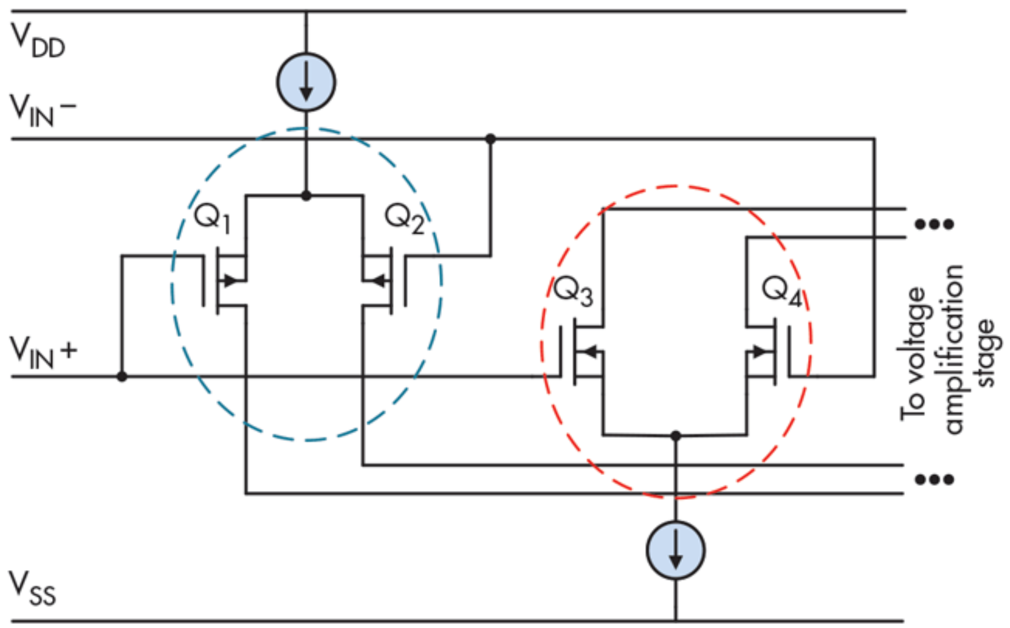
The offset voltage of an op amp drifts over temperature. Offset voltage drift (dV_{OS}/dT) is expressed in terms of the number of microvolts increase per degree of temperature change (µV/°C).

Offset Voltage and Crossover Distortion in Rail-to-Rail Op Amps

Early op-amp designs required dual power supplies (±15 V was typical). The supplies' input and output voltages could swing to within a volt or so of the rails before running out of headroom. As system voltages declined and battery-powered equipment proliferated, the preferred choice for many precision applications became a low-voltage single-supply op amp that could swing to the supply rail on both input and output, referred to as an RRIO device.

A state-of-the-art device such as Texas Instruments' new OPAx320-Q1 family of CMOS precision low-voltage RRIO op amps can operate from a single supply of only 1.8 V, or alternatively, a ±0.9-V split supply. The input common-mode range extends 100 mV beyond the supply rails, both positive and negative. The

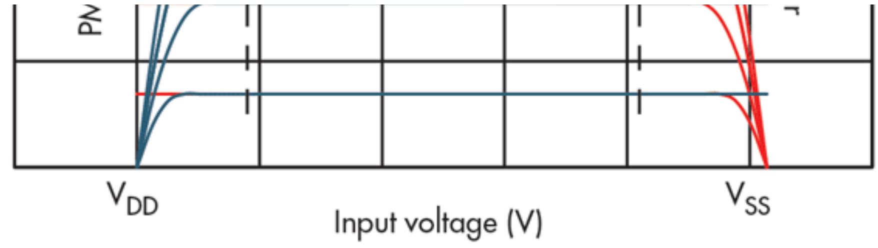
—one PMOS transistor pair (blue) and one NMOS transistor pair (red)—compared to the single pair of a non-RR design.



2. The composite input stage of a rail-to-rail op amp contains both PMOS and NMOS differential pairs. (Source: Texas Instruments)

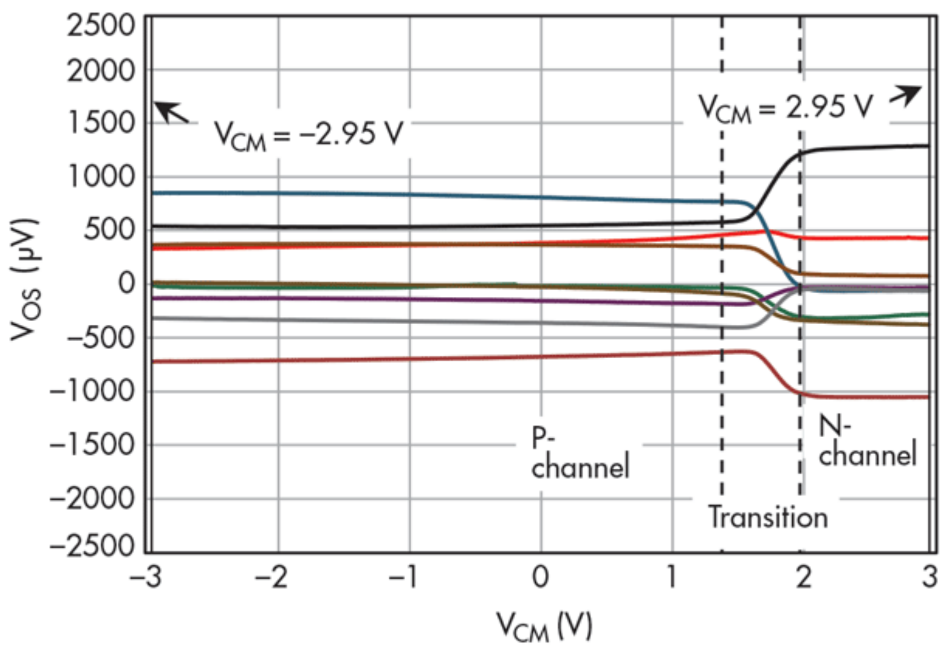
The PMOS transistors can operate with common-mode input voltages from V_{SS} to (V_{DD}-1.8V), and the NMOS transistors can operate with common-mode input voltages from (V_{DD}-1.8V) to V_{DD}. The two input transistor pairs are independent, so their input offset voltages, temperature coefficients, and noise are uncorrelated.

During the transition from the PMOS pair to the NMOS pair, and vice versa, there's a crossover region at ≈1.8 V below the positive rail where both inputs are conducting *(Fig. 3)*. Within this region, the value of V_{OS} can change—this is a source of distortion known as input crossover distortion.



3. Crossover distortion occurs during the transition region, when both sets of transistors are conducting. (Source: Texas Instruments)

For example, the input common-mode voltage (V_{CM}) range of the OPAx316 family of RRIO devices extends 200 mV beyond the supply rails. As *Figure 4* illustrates, there's a pronounced shift in V_{OS} as operation changes from the PMOS to NMOS transistors, and the shift can be in either direction.



4. Offset voltage and common-mode voltage are compared for nine typical OPA316 RRIO op amps. (Source: Texas Instruments)

This difference shows up in the rail-to-rail op amp data sheet as two or more common-mode rejection ratio (CMRR) specifications.

occurs when the OPA316 is in its transition region for almost the entire range of the test.

INPUT VOLTAGE RANGE	TEST CONDITIONS	MIN	TYP	MAX	UNIT
V_{CM} common-mode voltage	V_S = 1.8 V to 2.5 V	(V–) –0.2		(V+)	V
	V_S = 2.5 V to 5.5 V	(V–) –0.2		(V+) + 0.2	V
CMRR common-mode rejection ratio	V_S = 1.8 V, (V–) – 0.2 V < V_{CM} < (V+) – 1.4 V, T_A = –40°C to 125°C	70	86		dB
	V_S = 5.5 V, (V–) – 0.2 V < V_{CM} < (V+) – 1.4 V, T_A = –40°C to 125°C	76	90		dB
	V_S = 1.8 V, V_{CM} = –0.2 V to 1.8 V, T_A = –40°C to 125°C	57	72		dB
	V_S = 5.5 V, V_{CM} = –0.2 V to 5.7 V, T_A = –40°C to 125°C	65	80		dB

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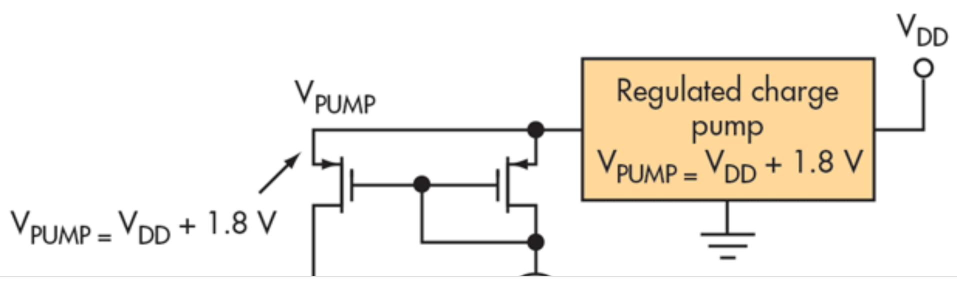
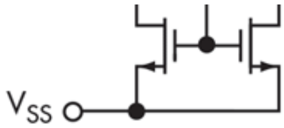
5. *RRIO op amps often have multiple CMRR figures to account for the change from PMOS to NMOS operation. (Source: Texas Instruments)*

How can we eliminate this crossover distortion? A "zero-crossover" op amp adds an internal charge pump to the input circuitry that raises the top of the differential input to 1.8 V above V_{DD}, the power-supply voltage positive side. Since the PMOS transistors can now operate over the full input range (which hasn't changed), we can eliminate the NMOS transistors and hence the crossover distortion. *Figure 6* shows the improved input stage with the added charge pump.





6. Adding a charge pump makes it possible to eliminate the NMOS differential pair and with it, the crossover distortion. (Image Source: Texas Instruments)

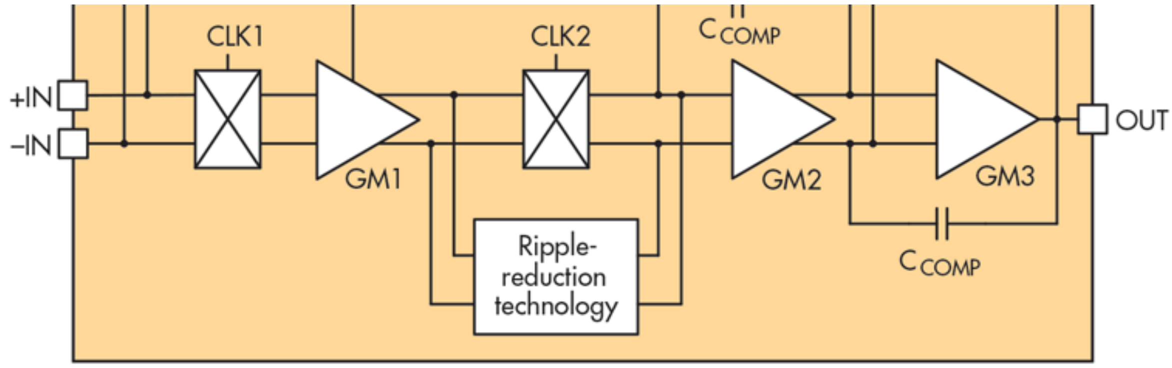
Eliminating Offset Voltage: The Zero-Drift Amplifier

Putting another nail in the offset-voltage coffin, zero-drift amplifiers employ a self-correcting technology that provides ultra-low V_{OS} and near-zero drift. This suits them for both general-purpose and precision applications.

Zero-drift amplifiers are good fits for a wide variety of general-purpose and precision applications that benefit from stability in the signal path. The excellent offset and drift performance of these amplifiers make them especially useful early in the signal path, where high gain configurations and interfacing with microvolt signals are common.

It’s possible to optimize system performance with standard continuous-time amplifiers plus a system-level auto-calibration mechanism. However, the method requires complicated hardware and software, and increases development time, cost, and board space.

The more efficient solution is to use a zero-drift amplifier such as the OPAx388 family of precision operational amplifiers, available in single, dual, and quad versions. These ultra-low-noise, fast-settling, zero-drift, zero-crossover devices have rail-to-rail input and output operation. The parts feature only 0.25 µV of offset and 0.005 µV/°C of drift over temperature, together with a 1-kHz noise figure of 7.0nV/√Hz and no 1/f noise (140 nV p-p from 0.1 to 10 Hz).

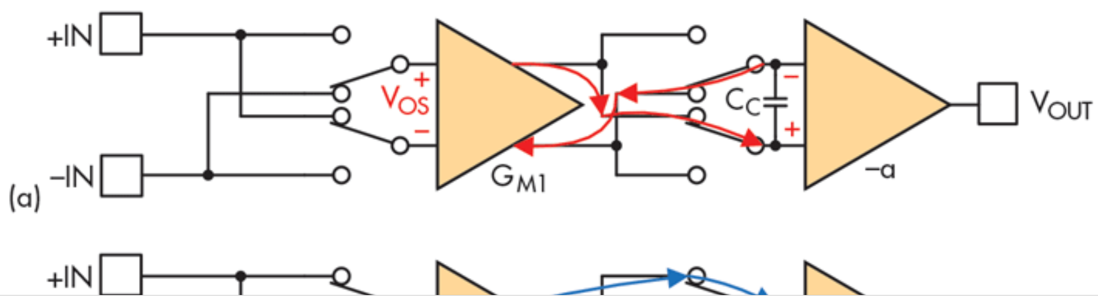


7. Shown is the internal block diagram for the OPA388. (Source: Texas Instruments)

Figure 7 shows the internal architecture of the OPA388. It includes three transconductance amplifiers in the main signal path, plus two chopping networks (CLK1 and CLK2). CLK1 provides initial modulation of the differential input signal up to the chopping frequency; CLK2 both demodulates the input signal and simultaneously modulates the offset and 1/f noise from GM1.

The ripple-reduction block senses the modulated ripple after CLK2 and demodulates it back to dc. It also includes a synchronous notch filter at the switching frequency to attenuate any residual error. The output of the block is then applied to GM1 to null out its offset and 1/f noise.

GM_FF is another transconductance amplifier that provides a feedforward path for high frequencies. The low-noise charge pump supplying the input stage provides zero-crossover performance as mentioned earlier.



Successive modulation and demodulation null out the error in the input transconductance stage. (Source: Texas Instruments)

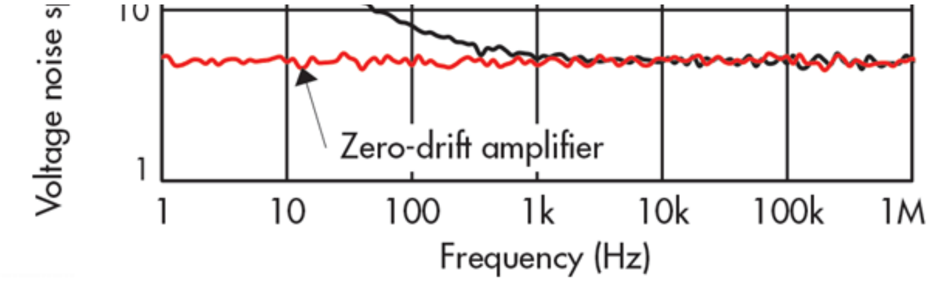
Figure 8 shows a simplified switching sequence. Fig. 8a shows the first half cycle: Both sets of switches are configured to flip the input signal twice, but the offset flips once. This keeps the input signal in phase, but reverses the offset error polarity.

Figure 8b shows the second half cycle. Here, both sets of switches are configured to pass the signal and offset error through unaltered. Effectively, the input signal is never out of phase, remaining unchanged from end to end. Since the offset error from the first clock phase and second clock phase are opposite in polarity, the error is averaged to zero when it's added back into GM1.

Device		V_{OS} (μV)	dV_{OS}/dT (μV/°C)
OPA388 (zero-drift)	typ	0.25	0.005
	max	5	0.05
OPA316 (continuous-time)	typ	500	2
	max	2500	10

9. A zero-drift op amp has considerably better offset performance and drift than a comparable continuous-time device. (Source: Texas Instruments)

Figure 9 compares the offset voltage and drift of the continuous-time OPA316 and the zero-drift OPA388. Notice that the V_{OS} and dV_{OS}/dT are three orders of magnitude smaller on the zero-drift amplifier. The 1/f noise shows a similar improvement *(Fig. 10)*.

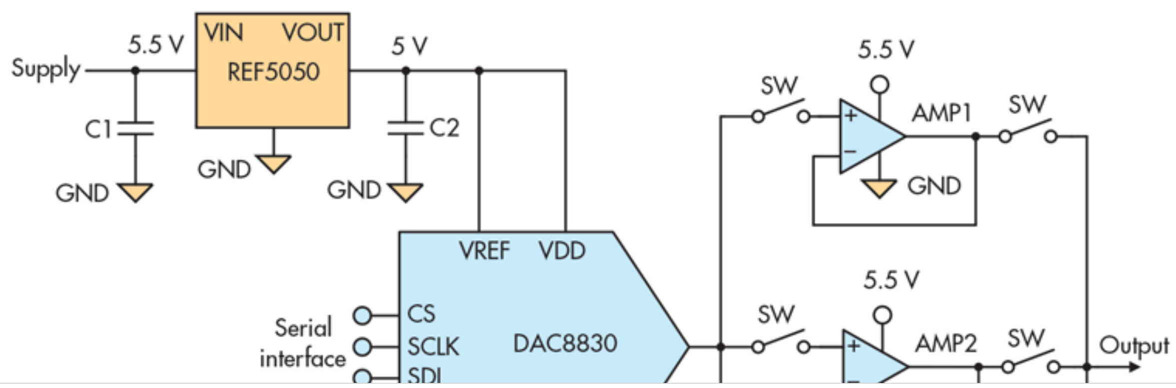


10. The zero-drift topology practically eliminates 1/f noise. (Source: Texas Instruments)

Application Example

Texas Instruments released a reference design that demonstrates the capability of the OPA388 to buffer the analog output of a digital-to-analog converter (DAC) for data acquisition, wireless infrastructure, and test-and-measurement applications. The design shows how the OPA388's zero-crossover and zero-drift feature operation can minimize the integral nonlinearity (INL) of the system, as well as make use of the full-scale range of the DAC, an ultra-low power 16-bit DAC8830.

Figure 11 shows the reference design block diagram. The REF5050 is a low-noise, low-drift, high-precision voltage reference that's able to both sink and source current. The device has excellent line and load regulation, very low drift of 3 ppm/°C, and accuracy of 0.05%.



11. This precision DAC reference design uses the OPA388 to buffer a DAC output. (Source: Texas Instruments)

The 16-bit DAC8830 has an LSB voltage of 76 µV when used with a 5-V reference such as the REF5050. The datasheet recommends the use of a low-offset op amp to eliminate the need for offset trimming.

Several op amps can meet the requirements, but the crossover region of a traditional RRIO device will negatively affect the INL, the most important specification for high-precision applications. The zero-crossover capability of the OPA388 allows the INL to remain below 1 LSB across the whole operating range.

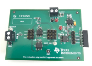
Conclusion

Rail-to-rail op amps are widely used in low-voltage precision applications, but traditional input topologies degrade offset-voltage performance in certain operating regions. Zero-crossover devices remedy this shortcoming. Specifically, zero-drift op amps solve the problem of offset drift by adding a switching component to the design.

The advantages of zero-drift op amps over continuous-time devices are particularly pronounced at low frequencies. The op amps are commonly used in applications such as precision strain gauge and weigh scales, current shunt measurement, and thermocouple-, thermopile-, and bridge-sensor interfaces.

Texas Instruments has a broad family of op amps with zero-drift or zero-crossover operation, including the OPA388, which features both capabilities.

Zero-Crossover Amplifiers: Features and Benefits



High-Precision Reference Design for Buffering a DAC Signal

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