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Can You Really Get ppm Accuracies from Op Amps?

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Industrial and medical design continually push to improve product accuracy and speed. The analog integrated circuit industry has generally kept up with speed requirements, but it is falling behind on accuracy demands. There is a march toward 1 ppm accurate systems, especially now that 1 ppm linear ADCs are becoming common. This article presents op amp accuracy limitations and how to choose the few op amps that have a chance of 1 ppm accuracy. We will also discuss a few application improvements to existing op amp limitations.

Accuracy is about numbers: how closely a system works to intended numerical value. Precision is about the depth of the numerical value in terms of digits. In this article we will use accuracy as a term that includes all limitations to system measurements, such as noise, offset, gain error, and nonlinearity. Many op amps have some error terms at ppm levels, but none have all the errors at the ppm level. For instance, chopper amplifiers can provide ppm-level offset voltages, dc linearity, and low frequency noise, but they have problematic input bias currents and linearity at frequency. Bipolar amplifiers can provide low wideband noise and good linearity, but their input currents can still cause in-circuit errors (we will hence use the term application for in-circuit). MOS amplifiers have excellent bias currents but are generally deficient in the low frequency noise and linearity areas.

In this article we will use the rough equivalency of 1 ppm nonlinearity in the transfer function as –120 dBc distortion in harmonic distortion.

Non-ppm Amplifier Types

Let's discuss the types of amplifiers we reject as not highly linear. The least linearity is found in so-called video or line driver amplifiers. These are wideband amplifiers with terrible dc accuracies: offsets in the several millivolts and bias currents in the 1 µA to 50 µA range, and usually with poor 1/f noise. Expected accuracies are 0.3% to 0.1% at dc, although the ac distortion can be from –55 dBc to –90 dBc (2000 ppm to 30 ppm linearity).

The next category is older classic op amp designs, such as OP-07, that may have high gain, CMRR, and PSRR, and good offsets and noise, but that cannot achieve better than –100 dBc distortion, especially into a 1 kΩ or heavier load.

Then there are the cheap amplifiers, new or old, that cannot best –100 dBc when loaded more heavily than 10 kΩ.

There is the audio amplifier class of op amps. They are fairly cheap, and their distortions can be very good. However, they are not designed for and do not offer good offsets nor good 1/f noise. They also cannot deliver distortion beyond perhaps 10 kHz.

There are op amps meant to support MHz signals linearly. These are usually bipolar throughout and have large input bias currents and 1/f noise. This application space sees more like –80 dBc to –100 dBc performance, and ppm performance is not practical with these op amps.

Current feedback amplifiers also cannot support deep linearity nor even modest accuracy, no matter how wideband nor huge their slew rates may be. Their input stage has a mess of error sources, and they do not have much gain nor input nor supply rejections. Current feedback amplifiers also have a thermal drift that extends fine settling times greatly.

Then we have the modern general-purpose amplifiers. They typically have a 1 mV offset and microvolts of 1/f noise. They support –100 dBc distortion but usually not when heavily loaded.

Op Amp Error Sources

Figure 1 shows a simplified op amp block diagram with ac and dc error sources added. The topology is a single-pole amplifier with an input g_m that drives a gain node that is buffered as the output. While there are many op amp topologies, the error sources shown apply to them all.
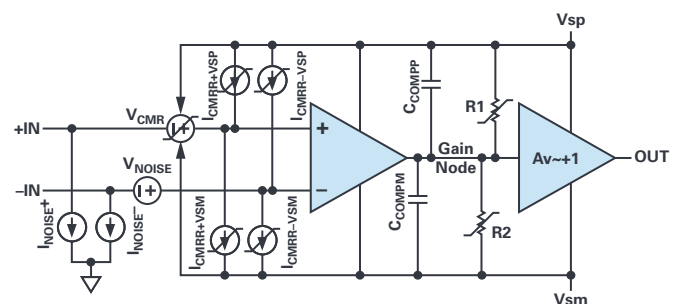

Figure 1. Simplified op amp and sources of error.

Input Noises

We have an input noise voltage V_{NOISE} with wideband and 1/f spectral content. You can't measure a signal accurately if the noise is of similar magnitude or more than a system LSB. For example, if we had a 6 nV/√Hz wideband noise and a 100 kHz system bandwidth, we would have 1.9 µV rms noise at the input. We could filter this noise down: for instance, dropping bandwidth to 1 kHz drops the noise to 0.19 µV rms, or about 1 µV p-p (peak-to-peak). Low-pass filtering in the frequency domain drops noise magnitude, as would averaging the output of an ADC over time.

However, 1/f noise cannot be practically filtered or averaged away because it is so slow. 1/f noise is usually characterized by peak-to-peak voltage noise generated in the 0.1 Hz to 10 Hz spectrum. Most op amps have between 1 µV p-p and 6 µV p-p low frequency noise and are thus not suitable for dc-accurate ppm levels, especially if providing gain.

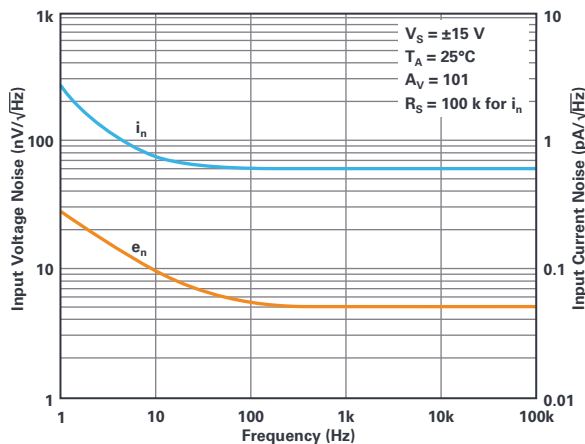Figure 2 shows the current and voltage noise of a good high accuracy amplifier, the LT1468.

Figure 2. LT1468 input voltage and current noise.

At the inputs in Figure 1, we also have bias current noise sources I_{NOISE+} and I_{NOISE-}. They contain both wideband and 1/f spectral content. I_{NOISE} multiplies against application resistors to become more input voltage noise. Generally, the two current noises are uncorrelated and do not cancel with equal input resistors but add in rms fashion. Quite often I_{NOISE} times application resistors exceeds V_{NOISE} in the 1/f region.

Input Common-Mode Rejection and Offset Errors

The next error source is V_{CMRR}. This embodies the common-mode rejection ratio specification where an offset voltage changes in response to the input's level relative to both supply rails (the so-called common-mode voltage, V_{CM}). The symbol used indicates supply interaction at the arrows, and the segmented line through it suggests it's variable but might not be linear. The major effect of CMRR on signals is that the linear part is indistinguishable from a gain error. The nonlinear part will be a distortion. Figure 3 shows the CMRR of an LT6018. The added line intersects extreme points of the CMRR curve just before the curve diverges into overload. The slope of the line gives a CMRR = 133 dB. The CMRR curve diverges from a perfect line by only about 0.5 µV per 30 V span—a very successfully sub-ppm input. Other amplifiers can have much more curvature.
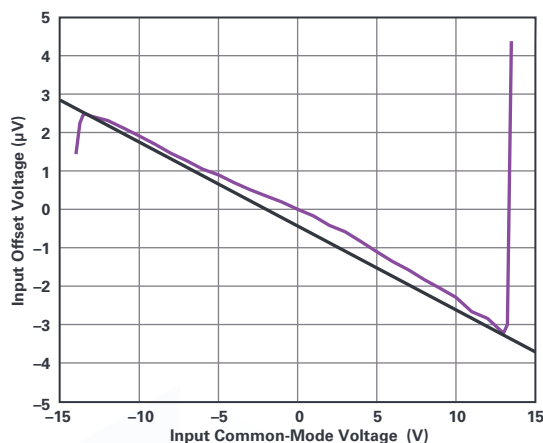

Figure 3. LT6018 input offset voltage vs. V_{CM}.

Offset voltage (VOS) will be lumped into CMRR here. Chopper amplifiers have sub-10 µV input offsets, and that's close to a single-ppm error, relative to typical signals of 2 V p-p to 10 V p-p. Even the best ADCs generally have as much as 100 µV offset. So, the onus of offset is not so much on the op amps; the system will have to auto-zero itself, anyway. Associated

with the input signal's common-mode level is I_{CMRR}, which is the input bias current and its variation with supplies. The broken lines suggest that the bias currents are variable with voltage and also may not be linear. There are four I_{CMRR}s because both inputs can have independent bias currents and level dependencies, and because each input is varied by both supplies independently. The circuit effect of the I_{CMRR}s (which sum to form bias current) is to multiply against application circuit resistances to add to overall circuit offset. Figure 4 shows the bias currents of an LT1468 vs. V_{CM} (the I_{CMR} specification). The slope as shown by the added line is ~8 nA/V, which would be 8 µV/V with a 1 kµΩ applications resistor, or a low ppm error. The deviation from straight line is about 15 nA, which in a 1 kΩ application environment creates 15 µV error over a 26 V span, or a 0.6 ppm nonlinearity.
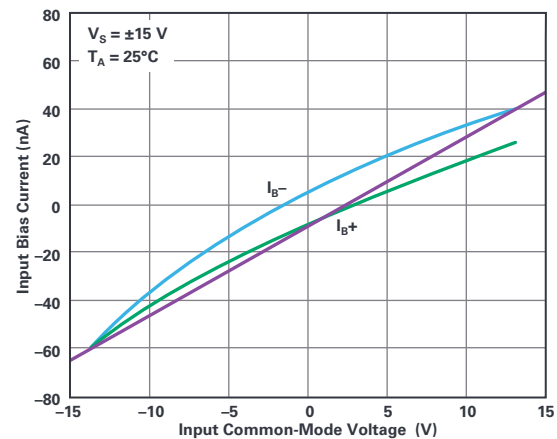

Figure 4. LT1468 input bias current vs. V_{CM}.

Input Stage Distortion

Figure 1 shows the input stage, which is generally a transconductor made from a differential pair of transistors. The top of Figure 5 shows the collector, or drain currents, of various differential amplifier types vs. differential input voltage. We simulate a simple bipolar pair, a translinear circuit that we will call *clever bipolar*, a subthreshold (that is, very large) MOS differential pair, a bipolar pair with emitter resistors (degenerated in Figure 5), and a smaller MOS pair operating out of the subthreshold region and into its square-law regime. All differential amplifiers are simulated with a 100 µA tail current.

Not a lot of information is obvious until we display transconductance vs. V_{IN}, as shown at the bottom of Figure 5. Transconductance (g_m) is the derivative of output current with respect to the input voltage, as generated using the LTspice® simulator. The syntax has d() to be mathematically equal to d()/d(VINP). The non-flatness of g_m is the basic distortion mechanism of op amps at frequency.

At dc, the open-loop voltage gain of the op amp is ~g_m(R1||R2), assuming the output buffer gain is about unity. R1 and R2 represent the output impedances of various transistors in the signal path, each connected to a supply rail or other. This is the basis of limited gain in an op amp. R1 and R2 are not guaranteed to be linear; they are a cause of unloaded distortion or nonlinearity. Aside from linearity, we need gains approaching or exceeding one million for ppm gain accuracies.

Observing the standard bipolar curve, we see it has the greatest transconductance of the group, but that transconductance fades quickly as the input moves from zero volts. This is concerning—a basic requirement for linearity is constant gain or g_m. On the other hand, who cares that the amplifier voltage gain is so high that the differential input would only move microvolts as the output moves volts? Time to introduce C_{COMP}.

C_{COMP} (the parallel of C_{COMPP} and C_{COMPM}) absorbs most of the g_m's output current over frequency. It sets the gain bandwidth product (GBW) of the
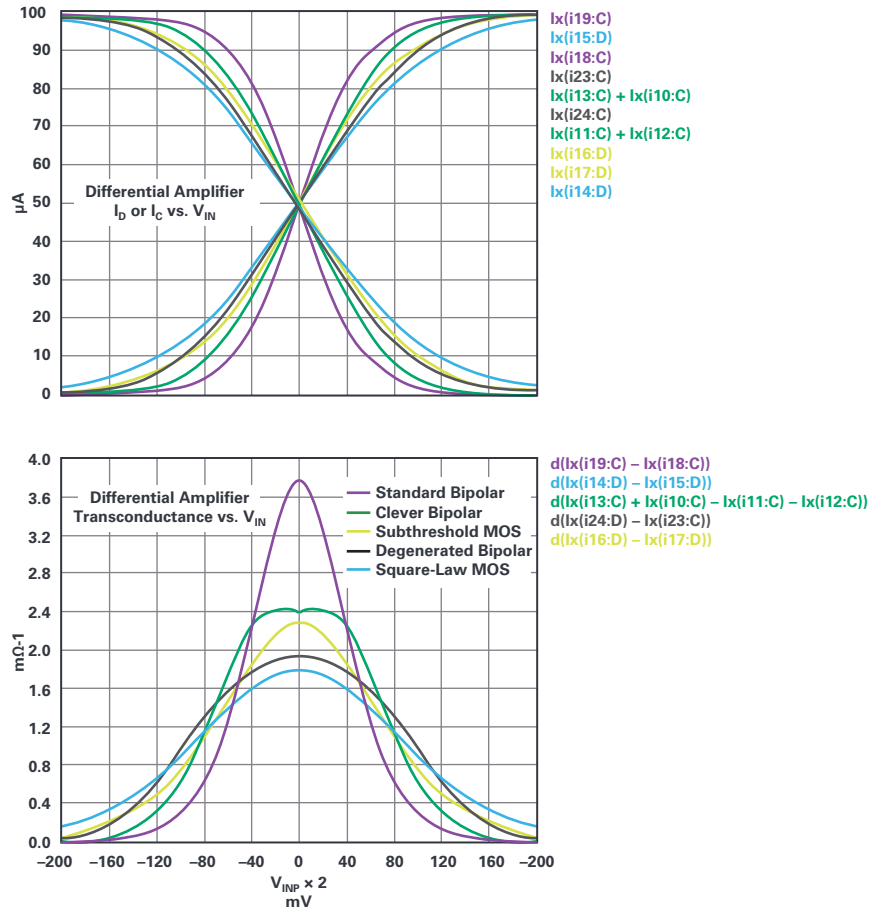
Figure 5. Various differential amplifiers' output current and transconductance vs. input voltage.

amplifier. GBW establishes that, at a frequency f, the amplifier will have an open-loop gain of GBW/f. If the amplifier is outputting 1 V p-p at f = GBW/10 with a closed-loop gain of +1, then we have 100 mV p-p between the inputs. That's ±50 mV from balance. Note that the standard bipolar curve shown in Figure 5 has lost about half its gain at ±50 mV, guaranteeing massive distortion. However, the clever bipolar only lost 13% of its gain, the subthreshold MOS lost 26%, the degenerated bipolar lost 12%, and the square-law MOS lost 15%.

Figure 6 shows the distortion vs. amplitude for the input stage. This will appear (times the noise gain) at the output of the application circuit. You may get more output distortion than this, but not less.
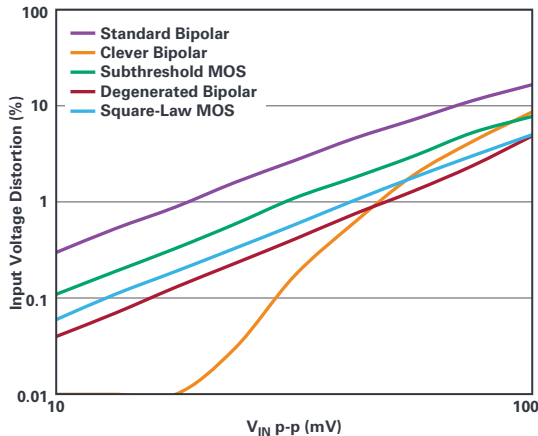

Figure 6. Total harmonic distortion of the input stage vs. differential input voltage.

Excluding the clever bipolar stage, the differential amps show that the distortion is proportional to the square of the input. In a unity-gain application, the output distortion contribution is equal to the input distortion. This is the dominant distortion source for most op amps.

Consider a unity-gain buffer with a bipolar input. For an output of V_{OUT} peak-to-peak volts, the input differential signal would be

$$f_{SIGNAL} \times \frac{V_{OUT}}{GBW} \tag{1}$$

We estimate that

$$\%Distortion_{INPUT} = 0.3\% \times \left(\frac{V_{IN} \, p\text{-}p}{10 \, mV}\right)^2 \tag{2}$$

And

$$\%Distortion_{OUTPUT} = G_{NOISE} \times \%Distortion_{INPUT} = 0.3\% \times \left(\frac{V_{OUT}, \, p\text{-}p \times f_{SIGNAL}}{10 \, mV \times GBW}\right)^2 \tag{3}$$

where G_{NOISE} is the noise gain of the application.

A 1 ppm nonlinearity is like −120 dBc harmonic distortion, and that's 0.0001%. Given an amplifier with a bipolar input stage, a 15 MHz GBW, and outputting 5 V p-p as a buffer, Equation 2 tells us that the maximum frequency for that linearity is just 548 Hz. This assumes the amplifier is at

least that linear at lower frequencies. Of course, when the amplifier provides gain, the noise gain increases and the –120 dBc frequency drops.

The subthreshold MOS input stage supports –120 dBc up to 866 Hz, square-law MOS up to 1342 Hz, and degenerated bipolar up to 1500 Hz. Clever bipolar does not follow the distortion prediction and one must get estimates from the data sheet.

We can use the simpler formula

$$\%\text{Distortion}_{\text{OUTPUT}} = K \times G_{\text{NOISE}} \times \left(\frac{V_{\text{OUT, p-p}} \times f_{\text{SIGNAL}}}{\text{GBW}} \right)^2 \qquad (4)$$

where K is found from the distortion curves of an op amp's data sheet.

As a side-note, there are many op amps with rail-to-rail input stages. Most get this ability from two separate input stages that have a hand-off from one to the other over the input common-mode range. This hand-off generates changes in offset voltage, and potentially bias current, noise, and even bandwidth. It also essentially causes a switching transient at the output. These amplifiers cannot be used for low distortion if the signal ever traverses the crossover region. An inverting application may work, however.

We haven't discussed slew-enhanced amplifiers yet. These designs do not run out of current with large differential inputs. Unfortunately, small differential inputs still cause variations in g_m of similar magnitude to the inputs discussed, and low distortion still demands a large loop gain at frequency.

Since we are looking for ppm-level distortion, we will not operate the amplifier anywhere near its slew rate limit, so, oddly enough slew rate is not an important parameter for ppm linearity at frequency, just GBW.

We've discussed open-loop gain as modeled by a single-pole compensation design. Not all op amps are compensated that way. Generally, open-loop gain is taken from the data sheet curve, and $\text{GBW}/(G_{\text{NOISE}} \times f_{\text{SIGNAL}})$ in the equation is that open-loop gain at frequency.

Gain Node Errors

The next items in Figure 1 to discuss are R1 and R2. These resistors, along with the input g_m, give the amplifier its open-loop dc gain of $g_m \times (R1\|R2)$. These resistors have been drawn with the variable and nonlinear strikethrough in the schematic. Nonlinearities of these resistors embody the amplifier's unloaded distortions. Further, R1 injects influence from the positive supply such that the dc positive power supply rejection ratio (PSRR+) and is approximately equal to $g_m \times R1$. Similarly, R2 is responsible for PSRR–. Note how PSRR is almost equivalent to open-loop gain in magnitude. C_{COMPP} and C_{COMPM} have an analogous injection of supply signals to R1 and R2; they set PSRR+ and PSRR– over frequency.

It is possible that an amplifier with modest gain ($<<10^6$) can be quite linear, but that modest gain will limit gain accuracy.

The power supply terminals can be a source of distortion. When the output stage drives a heavy load, that load current flows from one of the supplies. At frequency, the distant source of the supply may have little remote regulation, such that the op amp's bypass capacitors are the real supply source. The supply current drops across the bypass capacitors. These drops are dependent on the ESR, ESL, and the reactance, and they cause a supply disturbance. Since the output is class AB, only half of the output current waveform modulates the supply, creating even harmonic distortion. PSRR over frequency attenuates the supply disturbance. As an example, if we observed 50 mV p-p supply disturbance and wish the PSRR-induced input disturbance to be less than 5 µV p-p, we need a PSRR of 80 dB at signal frequency. Estimating that PSRR(f)~Avol(f), an amplifier with a 15 MHz GBW would have enough PSRR at frequencies below 1500 Hz.

Output Stage Distortions

The last item in Figure 1 is the output stage, which is considered a buffer for this discussion. A typical output stage transfer function is shown in Figure 7.
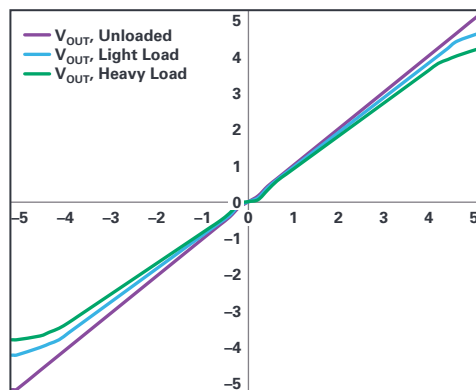

Figure 7. Transfer function of an output buffer with different loads.

For the different loads, we see four kinds of error. The first is clipping: although this hypothetical output stage has a gain of 1 nominally, it's not quite a rail-to-rail output stage. Even the unloaded output clips, in this case, 100 mV from each supply rail. The output clips at successively lower voltages as the load is increased (decreased load resistance). Obviously, clipping is a disaster for distortion and the output swing must be reduced to avoid it.

The next error is gain compression, which we see as curvature in the transfer function at the extremes of signal. The compression happens at earlier voltages as the load is increased, and like clipping no ppm-level distortion is generally possible in this regime. This compression is generally due to a small output stage that is struggling to output required current. A good rule of thumb is that the linear, uncompressed maximum output current available from the amplifier is only about 35% of output short-circuit current.

Another glaring source of distortion is the crossover region around $V_{\text{IN}} = 0$. Unloaded, the crossover kink may not be apparent, but with increasing loading we get something like the exaggerated kink of the green curve. Eliminating crossover distortion generally demands robust supply current.

The last distortion is harder to perceive. Because there are some bits of amplifier circuitry that output positive voltages and currents, and other bits for negative signals, there is no guarantee that they have the same gain, especially when loaded. Figure 7 shows a lesser gain for negative signals when loaded.

All these distortions are reduced by loop gain. If the output stage had 3% distortion, we would need a loop gain of 30,000 to get to a –120 dBc level. That of course happens below a frequency of GBW/(30,000 × G_{NOISE}), generally in the 1 kHz regime for a 15 MHz amplifier.

Some output stages' distortions are frequency-dependent, but many are not. The open-loop gain suppresses the output stage distortion, but that gain falls with frequency. If the output distortion is constant with frequency, then the gain loss creates an output distortion that increases linearly with frequency. Meanwhile, input distortion causes a total output distortion that increases with frequency. In this case, the input distortion will probably dominate the total closed-loop output distortion, masking the output stage distortion contribution.

On the other hand, if the output stage distortion does vary, say, linearly with frequency, the falling loop gain creates another output distortion that varies as frequency squared, additional to and indistinguishable from input distortion.

Low power op amps generally have starved output stages whose quiescent currents are low. These amplifiers' output stages may well dominate output distortion, rather than the input stage. It is somewhat true that it takes at least 2 mA supply current to make a low distortion op amp.

Required Specifications for ppm-Level Accuracies

In practical level-shift, attenuate/gain, and active filter circuits, we have some basic op amp requirements for an amplifier supporting ±5 V signal while working in a 1 kΩ environment and achieving 1 ppm linearity shown in Table 1.

Table 1. List of Op Amp Errors and Magnitude Needed for ppm Accuracy

Characteristic	Magnitude	Comment
V_{NOISE}	<6 nV/√Hz	Wideband input voltage noise
V_{NOISE} 0.1 Hz to 10 Hz	<1 ppm, p-p re. full-scale	Low frequency input voltage noise
I_{NOISE}	<6 pA/√Hz	Wideband input current noise
I_{NOISE} 0.1 Hz to 10 Hz	<10 nA, p-p re. full-scale	Low frequency input current noise
V_{OS}	<200 µV	Input offset voltage; usually digitally corrected
CMRR	>100 dB	Input common-mode rejection ratio; <10 ppm gain error
CMRR Linearity	>120 dB	The curvature of CMRR
I_{BIAS}	<200 nA	Input bias current; times a 1 kΩ resistor, <V_{OS}
I_{BIAS} vs. V_{CM} (I_{CMR})	<10 nA/V	Times a 1 kΩ resistor, less than 10 ppm gain error
I_{BIAS} vs. V_{CM} Linearity	<1 nA to 5 nA	Times a 1 kΩ resistor, less than 1 ppm re. full-scale
PSRR	>90 dB over some BW	Power supply rejection ratio; <1.6 µV V_{OS} shift per 50 mV supply variation
GBW	>1000× signal BW	Gain-bandwidth product; generally necessary for low distortion
Linear Output Current	>15 mA	Generally necessary for low distortion (= 35% of output short-circuit current)
DC V_{OS} vs. V_{OUT} Linearity	<1 ppm nonlinearity	Without dc linearity, you can't achieve ac linearity

So, we see the limitations of op amps in the ppm accuracy world—can we do anything to improve them?

Noise: Obviously the first step is to select an op amp with input noise voltage no higher than an application resistors' combined noise. One could reduce the overall impedance of applications circuits to reduce their noises. Of course, as the applications' impedance drops, the signal currents through them increase and can increase load-induced distortion. In any case, there's no reason to reduce the output noise of an op amp stage much below the input noise of the stage it drives.

Current noise will multiply against application impedances as more voltage noise. MOS inputs are attractive in having very low current noise, but they often have more 1/f voltage noise than bipolar inputs. Bipolar inputs have pA/√Hz levels of current noise that make non-trivial applications noise, but the 1/f current content can produce applications voltage noise greater than the 1/f voltage noise of the amplifier. A general rule of thumb is that the application impedance should be less than V_{NOISE}/I_{NOISE} of the amplifier to avoid I_{BIAS}-dominated application noise. The lower the V_{NOISE} of a bipolar amplifier, the higher the I_{NOISE} will be.

Helping Op Amps Achieve Best Performance

Reducing Input Errors

Aside from selecting an op amp with superior CMRR, designers can use op amps in inverting circuits rather than noninverting. In inverting circuits, the inputs hug ground or some reference and do not induce CMRR errors at all. Not all applications circuits can be made inverting, and often no negative power supply is available for the negative signal excursions. Figure 8 shows two-pole Sallen-Key filters in noninverting and inverting realizations.
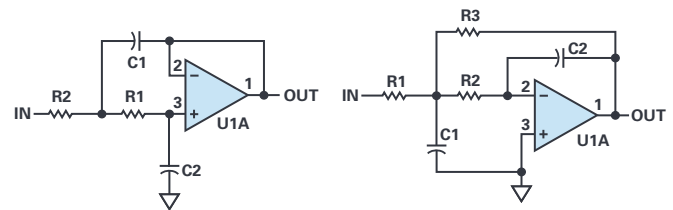

Figure 8. Noninverting (left) and inverting (right) Sallen-Key active filters.

I_{CMR} errors can be cancelled if both inputs have application resistors such that each input's bias current gets cancelled as an output error by the appropriate resistances. For instance, if an amplifier is set up at a gain of 10 with 900 Ω feedback and 100 Ω ground resistors, placing a series 90 Ω at the positive input will cancel perfectly equal bias currents at the output. The bias currents of most bipolar op amps are so well-matched that it is useful to choose 0.1% rather than common 1% resistors to achieve the best I_{CMR} rejection. In Figure 8, the compensating resistors would be placed in series with each –input. They should probably be bypassed across. The extra input resistor contributes more noise, unfortunately.

Inverting gain allows us to use an op amp that has rail-to-rail inputs without the signal traversing a switching point—assuming we bias the supplies and common-mode input level to avoid that switching voltage.

Supply Considerations

Output currents will modulate the local supply voltage. That supply signal will find its way to the input through PSRR. The induced input will produce an output signal that runs around its loop. At 1 kHz, a 1 µF local bypass capacitor has 159 Ω of impedance, much less than the impedance of the line between the supply source plus that of the source itself. Thus, the local bypass capacitor won't really be effective below 100 kHz. At 1 kHz, the remote supply controls regulation. At 1 kHz the amplifier may have, say, 90 dB supply rejection. Noting that most of the current from the op amp's supply terminals is made of even harmonics of the signal, we want the gain from the output to the offending supply to be lower than 30 dB to make our 120 dBc goal. The gain of 30 dB requires that the supply impedance should be <30× the load impedance. So, a 500 Ω load requires a supply with less than 17 Ω impedance. This is practical, but it does not allow a series isolating resistor or inductor between the supply and op amp. Things are tighter at 10 kHz; PSRR would drop from 90 dB to 70 dB, and the supply impedance would have to drop to 1.7 Ω. Doable, but tight. A large local bypass would help.
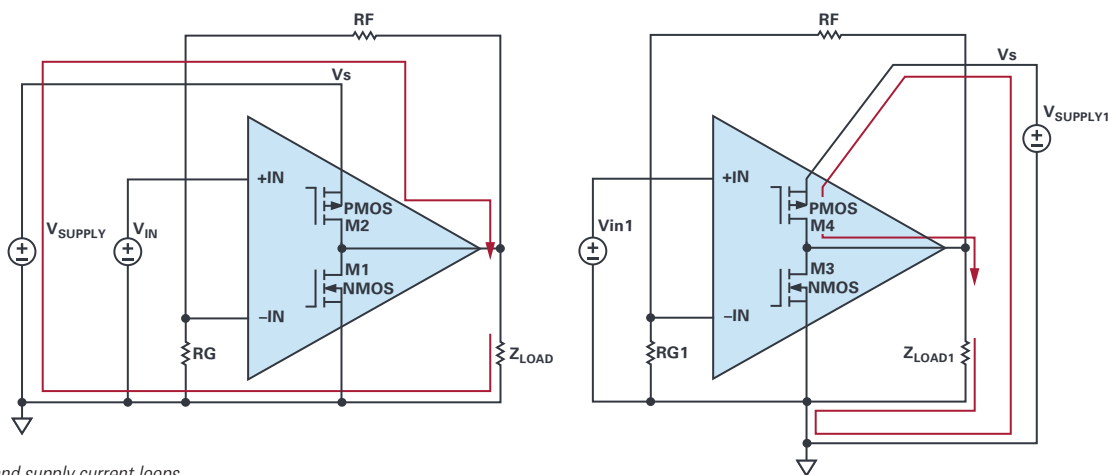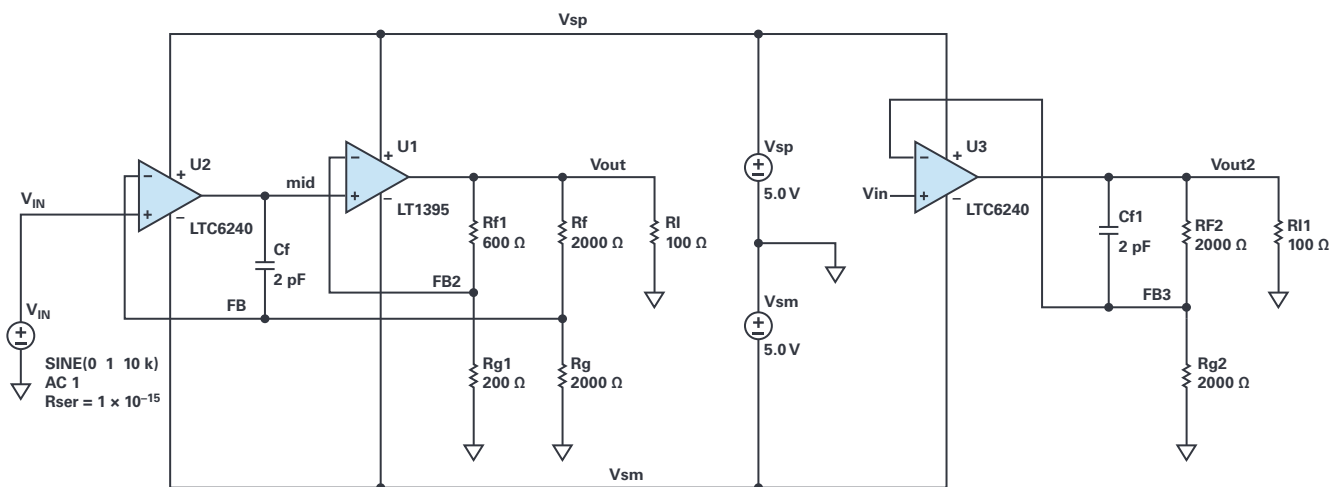
Figure 9. Load and supply current loops.

Figure 10. Composite amplifier vs. single amplifier distortion test.

From a layout point of view, it's important to see where the output current loops go, as seen in Figure 9.

The diagram on the left of Figure 9 shows positive supply current driven into a load, coming from a supply, then returning through ground back to the load. There can be voltage drops all along the ground path, such that even-harmonic supply current drops voltage from the signal source to the output, and from the feedback divider to output or input ground. This ground is not that ground. The right side of Figure 9 shows a better way to route the supply currents. The supply current is dressed away from input and feedback nodes.

At higher frequencies, above 100 kHz, the magnetic radiation of the supply lines can be a source of distortion. The even-harmonic currents of the supply can magnetically couple to the input or feedback network, dramatically increasing distortion with frequency. Careful layout is essential at these frequencies. Some amplifiers have nonstandard pinout; they keep the supply pins away from the inputs, and a few even offer an extra output terminal on the input side for to avoid magnetic interactions.

Reducing Load-Dominated Distortion

Many op amps' output stages become dominant distortion sources when loaded heavily. There are a couple of tricks to improve loaded distortion. One is the composite amplifier: one amplifier drives the output and another amplifier controls it, as seen in Figure 10.

This from an LTspice simulation. The LTC6240 and LT1395 have macromodels that include distortion playback. Most macromodels do not attempt to display distortion, and even if they do the simulated value can be inaccurate. This writer was able to look at the text of the macromodels, and yes, in these macromodels distortion was modeled fairly well.

On the right of Figure 10 is an LTC6240 providing a gain of 2 while driving 100 Ω—a difficult load for this amplifier. On the left of Figure 10 is a composite amplifier with another LTC6240 at the input, and a nice wideband current-feedback amplifier (CFA) to drive the same load as the standalone amplifier. The idea of the composite amplifier is that the output op amp already has a modestly low distortion, and that distortion can be reduced further by the input amplifier's loop gain over frequency. We have the same closed-loop gain of 2 for both standalone and composite amplifiers, but in the composite amplifier the LT1395 can be set up with its own gain (4 as set by Rf1 and Rg1) to reduce the output swing of the control amplifier. Since input-induced distortion increases as the square of output amplitude, there is a further reduction of distortion for the control op amp.

Figure 11 shows the spectrum of 10 kHz, 4 V p-p outputs.

Harmonic distortion would be measured as each harmonic level (dB) minus the level of the fundamental (at 10 kHz). As seen in the figure bottom the input signal has about −163 dBc distortion, good enough to trust the simulation. V(out2) is from the unassisted LTC6240 and has −78 dBc distortion. Not bad, but certainly not ppm-level.
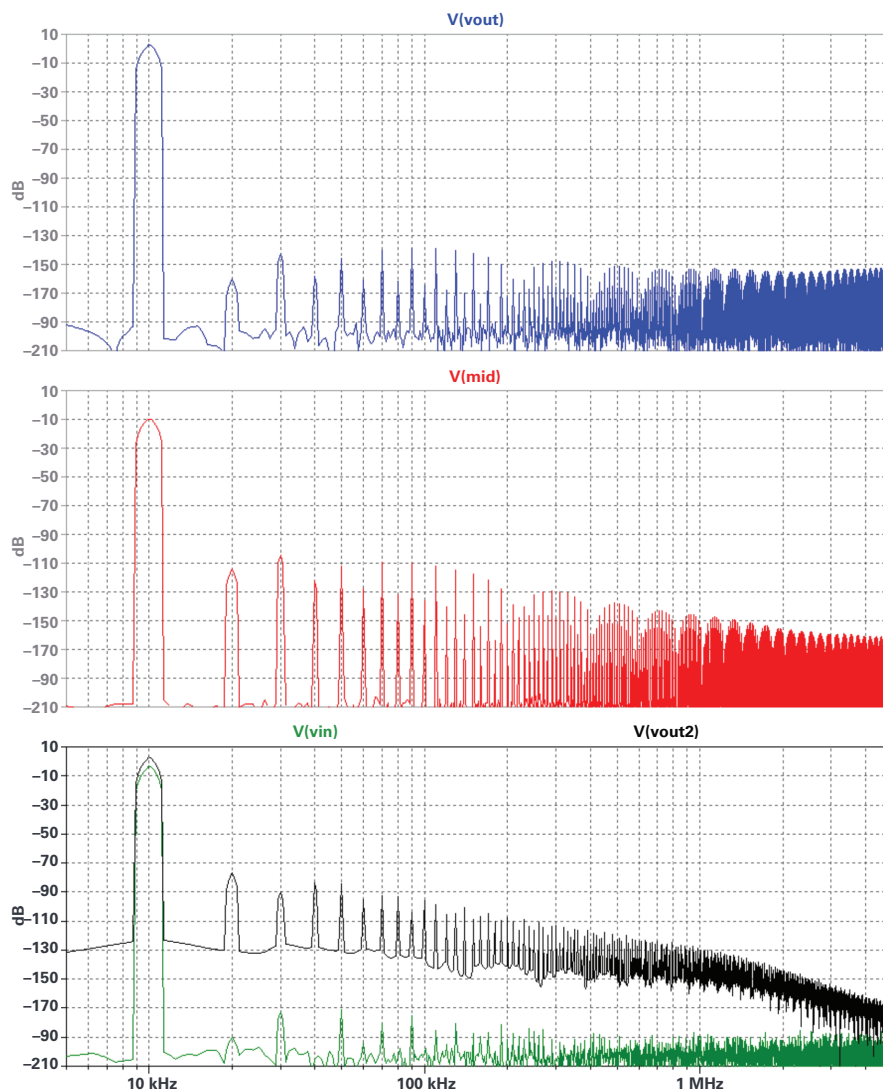
Figure 11. Composite and normal amplifier distortion spectra.

The top of Figure 11 shows the composite amplifier distortion at −135 dBc—pretty darn spectacular. Can we trust such a good result? For verification, the distortion at the schematic node mid is shown in the middle. If the output of a composite amplifier has near-zero distortion, but the output amplifier itself does have finite distortion, the feedback process will place the negative of the output amplifier's distortion at its input (mid). The distortion at mid is −92 dBc, and it actually matches the LT1395 data sheet curve! I would still wonder if the physical LTC6240 input CMRR or I_{CMR} curvature is expressed in the macromodel—they may yet increase real circuit distortion.

Unfortunately, few macromodels include distortion. You'd have to read the header in macromodel .cir files to see if it is supported. Some simulation is required to see if distortion matches the data sheet curves.

Compensation of a composite amplifier can be a little tricky, but in our example, we have a second amplifier whose bandwidth is over 10 times that of the input amplifier's, and just a little Cf compensates the circuit. In this compensation scheme, if the control amplifier has a bandwidth of BW at an overall gain, the output amplifier should have a bandwidth of >3 × BW, and the overall bandwidth will be conservatively set to ~BW/3.

To avoid losing bandwidth, we can use the boosted amplifier trick. It provides less distortion improvement than the composite approach but loses no bandwidth nor settling time. Figure 12 shows a test schematic.

The right side of Figure 12 shows U2, our standalone LTC6240, and the left side shows two LTC6240 amplifiers. U1 controls the output and has a gain of two, like the standalone; U2 has a gain of three. U2's output voltage at the boost node is larger than U1's, so U2 drives current into the output. R_{BOOST} and U2's gain are configured such that U2 drives 96% of the load current into Rl, leaving a light load for U1 and thereby improving its distortion. One needs to assure that U2 has enough output headroom for its extra swing.

The LTC6240 has input-dominated distortion for loads in the kΩ range, but is output stage distortion dominated with our 100 Ω load.

Figure 13 shows the spectral results.

Again, we have −78 dBc distortion at 10 kHz for the standalone amplifier. The boosted amplifier delivers −106 dBc; not nearly as good as the composite amplifier, but almost 30 dBc better than the standalone. However, the boosted amplifier suffers only a modest bandwidth reduction.
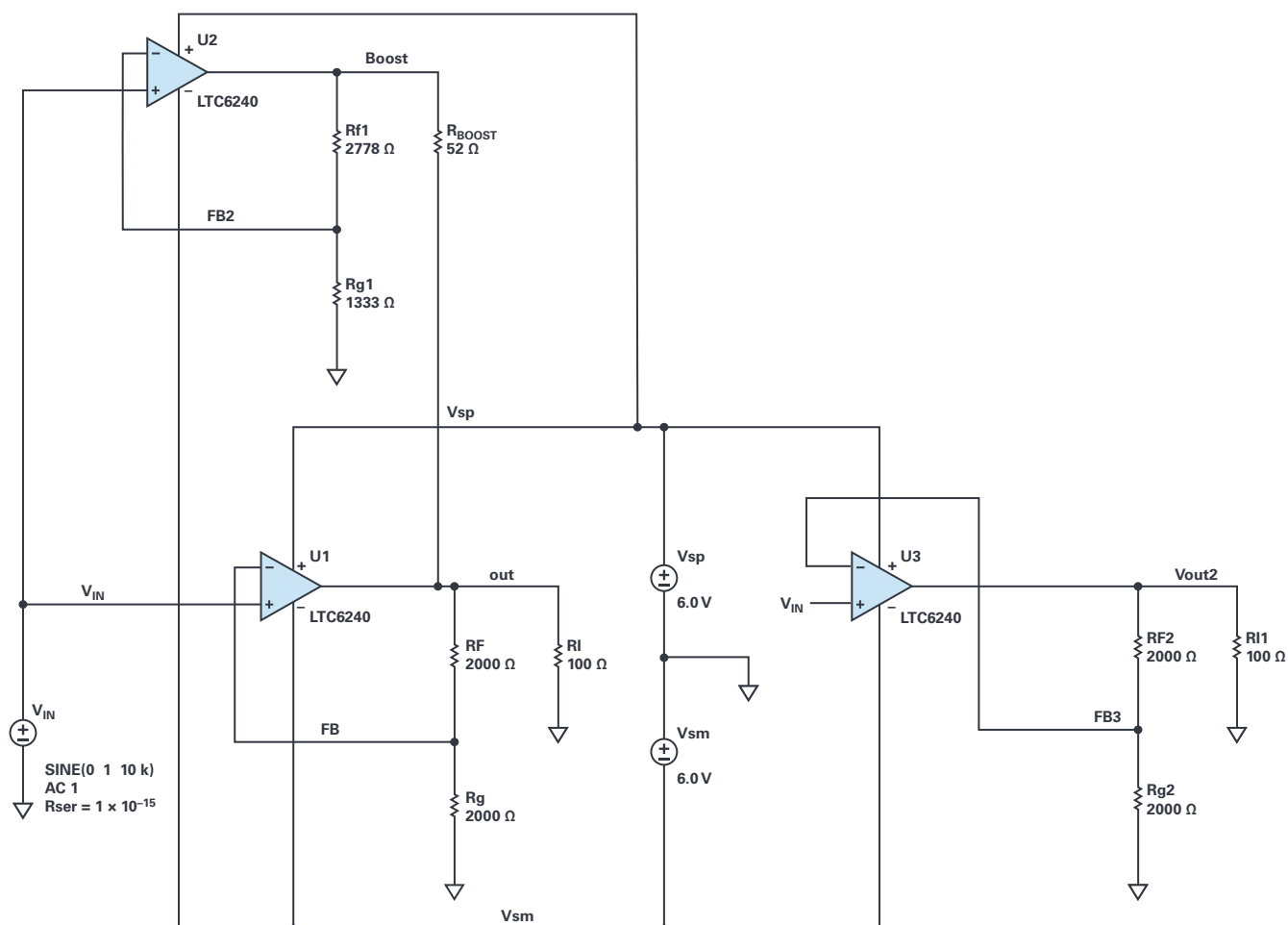
Figure 12. Boosted vs. standalone amplifier simulation setup.

Note that R_BOOST is tweaked; if we vary it as 52 ± 2 Ω, the boosted distortion degrades by 10 dBc, although little change thereafter happens for up to ±10 Ω. It would appear U1 likes to have some modest load of the expected polarity; ideal (no load) or excess boost current causes more distortion.

Ideally, U2 would have the same group delay as U1 so that the boost signal occurs at the same time as the output. U2 has 50% more gain than U1 and thus has less closed-loop bandwidth, suggesting the boost output lags the main output at frequency. U1's bandwidth could be reduced to be equal to U2's by installing a resistor across U1's inputs. This would increase the noise gain of U1 to be equal to U2 and achieve equality between the group delays. The simulator showed no improvement at 10 kHz; U1 gave best distortion with no delay balancing. Knowing whether or not this is true at higher frequencies will require trying it. If the amplifiers were of current-feedback type, Rf1 and Rg1 could be reduced to bring U2's bandwidth up to U1's.

Recommended ppm-Quality Amplifiers

Table 2 shows the salient specifications for some suggested amplifiers approaching ppm linearity.

Entries are in red to alert the reader that a parameter may violate ppm-level distortion. The easy to use winners of the group are the AD8597, ADA4807, ADA4898, LT1468, LT1678, and LT6018.

There are some amplifiers that have input problems that must be dealt with (noninverting applications might be an issue) but that can still provide good distortion: AD797, ADA4075, ADA4610, ADA4805, ADA4899, and LTC6228.
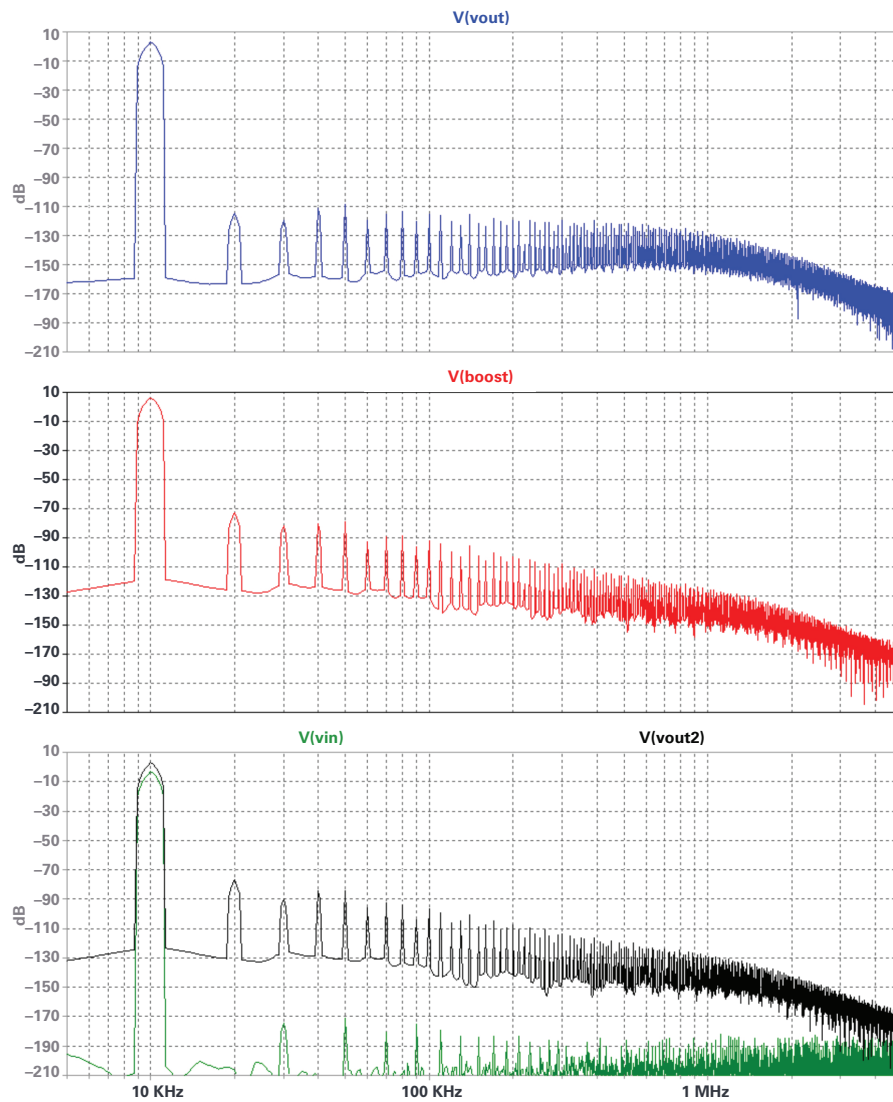
Figure 13. Boosted and normal amplifier distortion spectra.

Table 2. Salient Specifications for Some Suggested Amplifiers Approaching ppm Linearity

Device #	V_NOISE nV/√Hz	V_NOISE 0.1 Hz to 10 Hz μV p-p	I_NOISE pA/√Hz	I_NOISE 0.1 Hz to 10 Hz pA p-p	V_OS μV (max)	CMRR dB (min)	CMRR Nonlinearity μV/V	I_BIAS nA (max)
AD797	0.9	0.05	2	220	60 to 180	110 to 114		2000 or 3000
AD8597	1.1	0.08	2.4	190	120	105	0.1	200
ADA4075	2.8	0.06	1.2	60	1000	106	1.5	100 to 150
ADA4610	7.3	0.45	Very small		800 to 1800	96		0.025 to 1500 (hot)
ADA4805	5.2	0.1	0.7	140	125	103		800
ADA4807	3.1	1.6	0.7	370	125	103	0.2	800
ADA4898	0.9	0.05	2.4	130	125	103		400
ADA4899	1.0	0.4	5.2	4800	230	98		1000
LT1468	5	0.3	0.6	3	150 to 400	96		10, 40
LT1678	3.9	0.09	0.7	26	350	96		35
LT6018	1.2	0.03	0.75 or 3	110 or 750	75 to 95	120	0.02	150 to 900
LTC6228	0.9	0.94	6.3	9000 or 20,000	95 to 250	94	0.1	4000 or 44,000

Table 3. Op Amp Comparison Continued

Device #	I_{BIAS} vs. V_{CM} nA/V	I_{BIAS} vs. V_{CM} Nonlinearity pA/V	PSRR dB (min)	GBW MHz	Linear Output Current mA (min)	DC V_{OS} vs. V_{OUT} Nonlinearity ppm	Distortion: AV = 2, 2 V p-p Out 10 kHz dBc	Macromodel Shows Distortion?
AD797			110 or 114	110	±30		−120	Simulation model is too optimistic
AD8597	5	0.2	118	14	±20		~−120	Compare with data sheet
ADA4075	2	3000	100	6.5	~±15		~−130	Optimistic
ADA4610	Very small	0.1	100 or 103	12	~±30		~−114	None
ADA4805	2.2	4000	100	30	~±30	30	−125	None
ADA4807	0.7	~140	98	17	±50		~−130	None
ADA4898			98	120	±40		~−120	None
ADA4899			84	280			−117	None
LT1468	3.5	600	100	55	±15		~−120	Yes
LT1678			1000.7	10	~±10		~−120	Simulation model is too optimistic
LT6018			128	12	~±15	0.02	~−115	Yes
LTC6228	300	28 or 140	95	800	±20	0.2	−120	

Conclusions

Sadly, commercially available ppm-accurate amplifiers are difficult, if not impossible, to find. There are ppm-linear amplifiers, but attention must be paid to the amplifier's input currents that create distortion against application impedances. Those impedances can be lowered, but driving them in feedback runs the risk of creating distortion at the op amp input. By using an op amp with particularly low input currents and variations, the application impedance can be raised to obtain the best distortion from the op amp, but this will raise system noise. Careful op amp selection and applications circuitry optimization are required for ppm linearity and noise.

About the Author

Barry Harvey has worked as an analog IC designer, designing high speed op amps, voltage references, mixed-signal circuits, video circuits, DSL line drivers, DACs, sample-and-hold amplifiers, multipliers, and more. He has an M.S.E.E. from Stanford University. He holds more than 20 patents and has published about as many articles and papers. Barry's hobbies include repairing used test equipment, playing guitar, and working on Arduino-related projects. He can be reached at barry.harvey@analog.com.