Audio Electronics

Current Feedback: Fake News or the Real Deal? (Part 2)

A Detailed Examination of Current Feedback Amplifiers

In Part 1 of this article series, we designed, analyzed, simulated, built, and tested simple versions of voltage and current feedback amplifiers and demonstrated some differences in their characteristics. In Part 2, we will take a detailed look at some commercially available current feedback amplifiers and voltage feedback amplifiers.

By
Christopher Paul
(United States)

In the first part of this article series, we addressed the question asked by the article's title. We subjected several circuits to bench tests, simulations, and derivations of equations that described their properties. We also examined the history of the term "current feedback." From all of this, it seems that we might have been asking the wrong question: Not does it exist, but how do we define it? Consider the following possibilities:

1. We can sense either the voltage across the output load, or the current through it. That current can produce a voltage across a small ground-referenced resistor in series with it. Sensing that voltage is referred to as series derived feedback (see Part 1, Figure 7) and sometimes simply as current feedback.
2. Figure 7 from the first part of the article series also addresses shunt applied feedback. In that arrangement an amplifier's non-inverting input is grounded, and its typically high open loop gain drives the voltage difference between the inputs to very small levels. Since the inverting input is now at "virtual ground," the feedback signal looks like a current flowing from the output through the feedback network to ground. This is another claim for current feedback.
3. Using the tools created by R.A. Middlebrook, we discovered in Part 1 that regardless of amplifier input stage structure, the relationship between the impedances of the amplifier's feedback input Z_{FB} and that seen looking from that input into the feedback network (R_{EQ}) determines whether voltage or current feedback predominates.
4. Consider a single transistor input stage whose feedback point is its emitter. Its output is its collector current, and its emitter is biased by a DC current source, which by definition cannot source AC signal current. Its base current is much smaller than its collector's. Therefore, the AC current signal must pass through the feedback network and derive from the amplifier's output. (To see this clearly, replace these elements with their Thevenin equivalent: an attenuated voltage source in series with a single resistor.) Here is yet another claim for current feedback.

It is foolish at this late stage to attempt to restrict the meaning of current feedback to one single thing. The best we can do is remain aware of the context in which the term is used. Let's look at some designs that qualify for definition 4.

A Commercially Available Op-Amp Using Current Feedback

There is at least one op-amp that uses the simple e(mitter)-fed stage introduced in Part 1. In fact, it uses two of them. The part is Analog Device's SSM2019.

This device is a differential to single-ended amplifier (see **Figure 1**). The signal input is applied between IN+ and IN-. A single resistor R_g is connected between the points RG_1 and RG_2 to set gain. The circuit is bilaterally symmetric, and so if the input signal is balanced, the voltage at the center of resistor R_g is at AC ground. In this case, you can think of the 1+ and 1- stages as each operating with an $R_g/2$ resistor to AC ground. Stage 1 drives Stage 2. Stage 2 drives Stage 3. Together, they form a "folded cascode." Through sets of 5 kΩ resistors, Stage 3's ×1 buffered outputs supply differential feedback to Stage 1 and inputs to the Stage 4 differential to a single-ended converter.

Typical distortion at and below gains of 100 is better than 0.01%, and noise is 1nV/√Hz above 100 Hz. To determine the Stage 1 emitter impedances, we can plug this noise level into an equation for thermal noise, R (ohms) = $(\text{volts}/\sqrt{\text{Hz}})^2/(1.62 \times 10^{-20}$ Watt-sec). We must consider that the two input transistors contribute to R in an RMS fashion, and so we can calculate that the resistance at the emitter of each is about 44 Ω. And sure enough, the datasheet graphs are consistent with bandwidth remaining constant as $R_g/2$ rises above that value. At values of 44 Ω, the impedances of the emitters, Z_{FB}, and those seen by the emitters, R_{eq}, are approximately equal. We discussed this type of behavior in Part 1. Let's call it the "constant bandwidth effect."

There is another important effect, independent of the relationship between Z_{FB} and R_{eq}, that arises from this circuit topology. The current available to charge and discharge the capacitor in the signal paths of all op-amps limits their slew rates. Some of that current can come from the DC current sources supplying bias for Stage 1. Portions of these currents can be steered through Stages 1+ and 1- by the voltages at IN+ and IN-. This is the only source of charging current for capacitors in standard voltage feedback amplifier (VFA) op-amps, where feedback is applied to a transistor base rather than the emitter. But when feedback is applied to an emitter through the feedback network, it flows through the entire stage, boosting the charging current as needed. This current only flows when needed and so does not contribute to constant power dissipation. We can call this effect "slew rate enhancement."

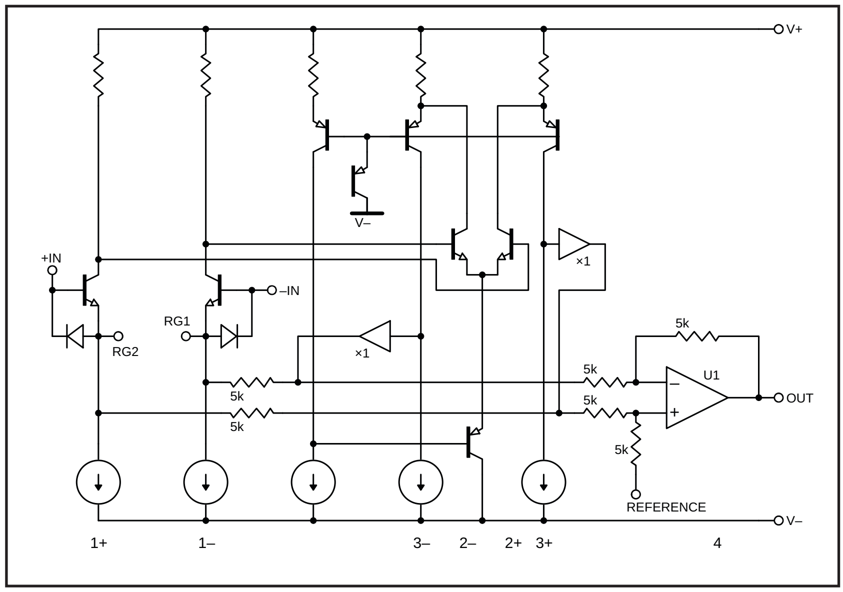


Figure 1: This is the schematic of the SSM2019. Each of its stages is numbered for discussion. (Original image courtesy of Analog Devices)

The CFA

Although we can argue (in accordance with current feedback definitions 3 and 4) that the SSM2019 is a current feedback amplifier, it is not what the industry has come to call a Current Feedback Amplifier (CFA). Such a device has the specific basic structure shown in **Figure 2**.

This design contains two current mirrors, Q5/Q7 and Q6/Q8, and four separate Faux-Darlingtons: Q1/Q3, Q2/Q4, Q9/Q11, and Q10/Q12. (I regret that I have found no better name for this configuration and shall henceforth refer to them as Darlingtons.)

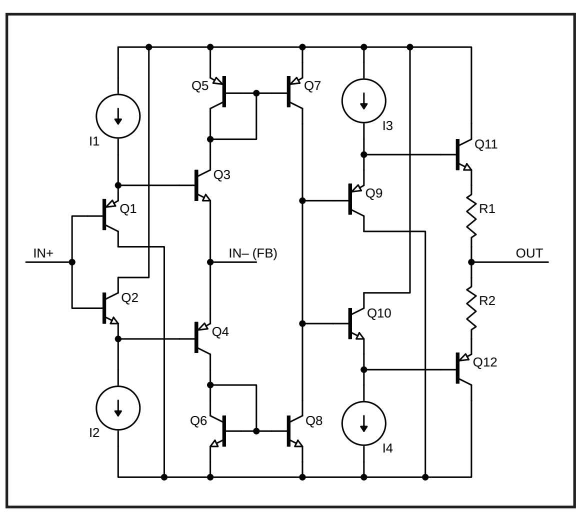
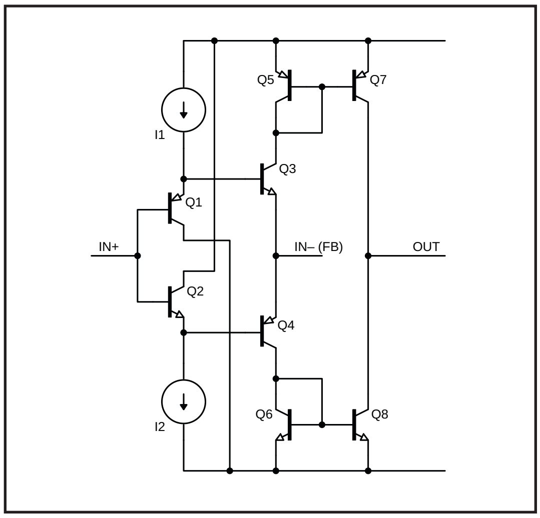


Figure 2: Here is the basic structure of what the industry refers to as a Current Feedback Amplifier (CFA).

Figure 3: The assumed structure of Texas Instruments' "ideal transistor" Operational Transconductance Amplifier, the OPA861.



Each Darlington contains one NPN and one PNP transistor, allowing its input and output to be at the same DC voltage, simplifying the overall design. The input transistor of each pair receives an emitter bias current from a DC current source as shown. The matching capability inherent to ICs enables the bias current of the pair's output transistor to be set as desired (this topology would not work well with discrete transistors). The signal current (I_{FB}) flowing into IN- is routed through Q3 and Q4. Their outputs are added together by the current mirrors, whose outputs have a very high resistance Z. This resistance is only slightly lowered by the two output Darlingtons. Typically, this is the point at which capacitance is controlled to ensure stability when the feedback loop is closed. The transfer function of this device is Z = OUT/I_{FB}. Each of the two input Darlingtons functions like the simple e-fed input stage described in Part 1. Since the same DC current I_{DC} runs through these seriesed devices, the impedance Z_{FB} looking into IN- at room temperature is approximately (26 mV/I_{DC})/2. This design exhibits both the constant bandwidth and the slew rate enhancement effects already discussed.

A CFA Enhancement

The constant bandwidth effect is dependent on R_{eq} being larger than Z_{FB}. To support this effect, we make Z_{FB} as small as possible so that even at high gains, when the feedback shunt resistor that contributes to R_{eq} is very small, this effect is still present. A brute force method is to increase the bias current of the input stage, which of course increases dissipation. But another approach is possible—enclose it in a unity-gain feedback loop. This is what Texas Instruments (TI) did with the OPA683, the OPA684, and the OPA691. It would be nice to be able to describe this circuitry, but TI's datasheets do not provide any schematics. Consider, however, that the OPA683 affords bandwidths of 100 MHz or more and a slew rate of 450V/µS at gains of +5 or less at a supply current of under 1 mA!

Solomon's Partial CFA

To paraphrase from King Solomon's teachings: "That is my CFA, your majesty," shouted Rebecca. "No, it's mine," cried Miriam as she scowled murderously at Rebecca. The wise king considered for a moment. "Since you cannot agree, you both shall have a part of it!" And he proceeded to slice the microchip into two pieces with an extremely precise laser.

History does not record which portion went to whom, let alone how the packaging engineers managed to make useful surface mount components out of them. But one of the pieces proved very useful, and it is shown in **Figure 3**.

Clearly, this is the input of the Figure 2 circuit. But it's also an Operational Transconductance Amplifier (OTA). It accepts a voltage difference between its two "IN" pins and supplies a bidirectional current from its OUT pin. Most likely, something very much like this has been implemented in TI's OPA861. (I can't be certain about this, but some folks I know with respectable credentials claim this to be so.) Not shown in Figure 3 is an enhancement to the basic design which allows input stage bias current to be set

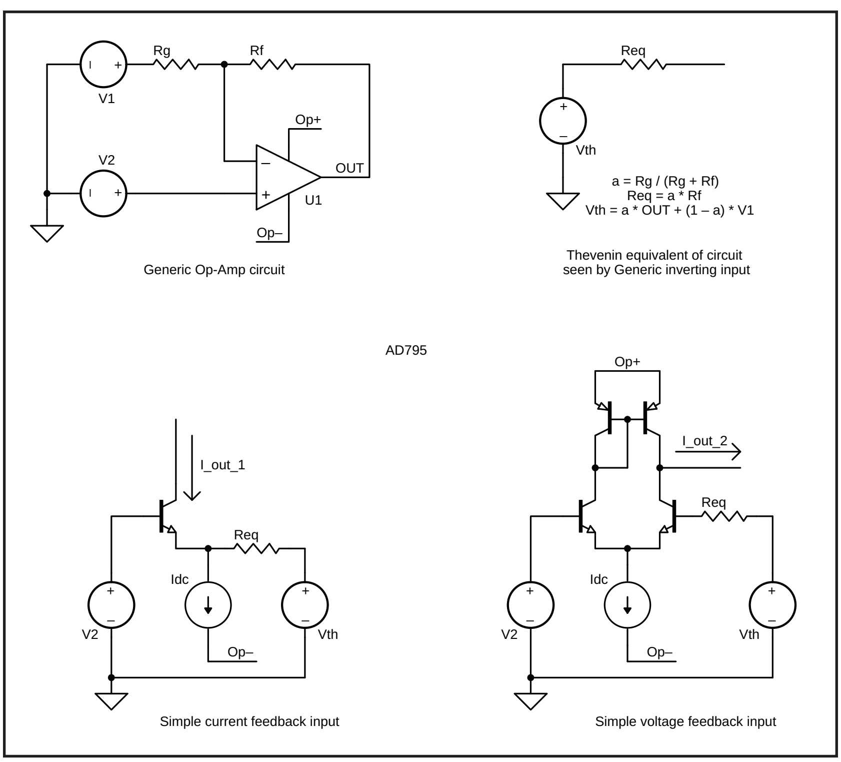


Figure 4: This is a generic op-amp with simplified current and voltage feedback type inputs.

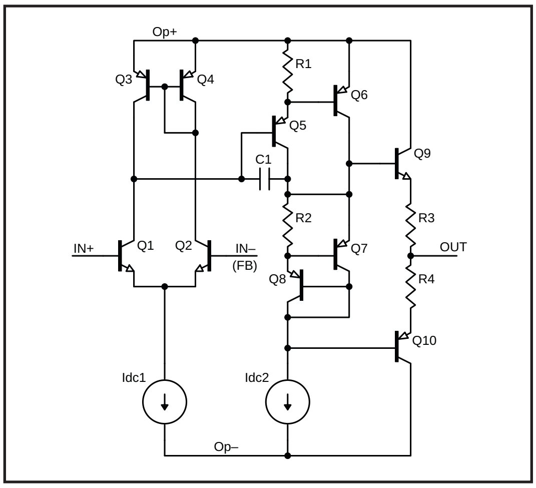


Figure 5: Here is the basic structure of what the industry refers to as a Voltage Feedback Amplifier (VFA).

to trade off quiescent current, gain and bandwidth. This part is marketed as an "ideal transistor," whose IN+ is the "base," IN- the "emitter," and OUT the "collector." Its ideal aspect comes from its zero DC offset "base-emitter" voltage, and the fact that its inputs and outputs support both polarities of voltages and currents. Transconductances from 40 mA/V to more than 100 mA/V can be attained with bandwidths from 50 MHz to 100 MHz. Even if some of the datasheet's applications might not be optimal for audio, it's fun to stretch your analog brain and look at things from the point of view of a device whose relative output and inverting input impedances are the opposite of an "ideal" op-amp's.

Yes Virginia, there is a VFA

Comlinear Corp. was first to offer the CFA commercially back in the 1980s. Before that, there was a topology that didn't need a name because it was pretty much the only game in town. But with the advent of the CFA, it has come to be referred to as a VFA. **Figure 4** helps us see how these names evolved. The generic op-amp, either a VFA or a CFA, works in both inverting and non-inverting configurations. Converting the resistive network into a Thevenin equivalent aids in our understanding of these variants. Simplified examples of both input stages are provided. The output of each is a current. In the VFA, that comes almost exclusively from I_{dc}, which is steered by the difference between voltages V2 and Vth. The output is controlled by this voltage difference, one part of which is fed back from the voltage OUT, hence the term VFA. In the CFA, the DC source cannot be steered; its output passes through the transistor collector save for a small portion that passes through the transistor base. The dynamic (signal) portion of the feedback current, therefore, must come from somewhere other than I_{dc}: OUT, through R_{eq}. Accordingly, this type of op-amp is referred to as a CFA. All CFAs and VFAs are elaborations of these simple circuits, which are distinguished by the type of transistor pin to which feedback is applied.

The basic structure of a complete VFA can be seen in **Figure 5**. DC current Idc1 supplies Q1 and Q2. The voltage difference between IN+ and IN- steers Idc1 between these two transistors. The current mirror consisting of Q3 and Q4 recombines the currents, which supplies the input of the Darlington consisting of Q5 and Q6. C1 provides frequency compensation to ensure stability. The Darlington's output and Idc2 supply current to the voltage reference consisting of Q7, Q8 and R2. This reference, along with R3 and R4, establishes the bias current for the output stage comprising R3, R4, Q9, and Q10.

VFA Enhancements

Several techniques have been applied to VFAs to enhance slew rate without increasing quiescent power dissipation. Analog Devices employs what it calls a "Quad Core" design, which can be seen in **Figure 6**. The "central" resistor in the middle of and equidistant to Q1, Q2, Q3, and Q4 is the key to its operation. (The complete bias circuitry is not shown.) When the voltages at the + and – inputs are equal, no current flows through this resistor. But if the voltage difference increases because of inadequate slewing, the input stage becomes increasingly unbalanced. If the voltage at the – pin were to exceed that at the + input current through the components and paths indicated by the arrows would increase. The increase could much more than double the currents supplied by only constant current sources (CCS). These "unbalance" currents would charge the capacitors, increasing slew rate.

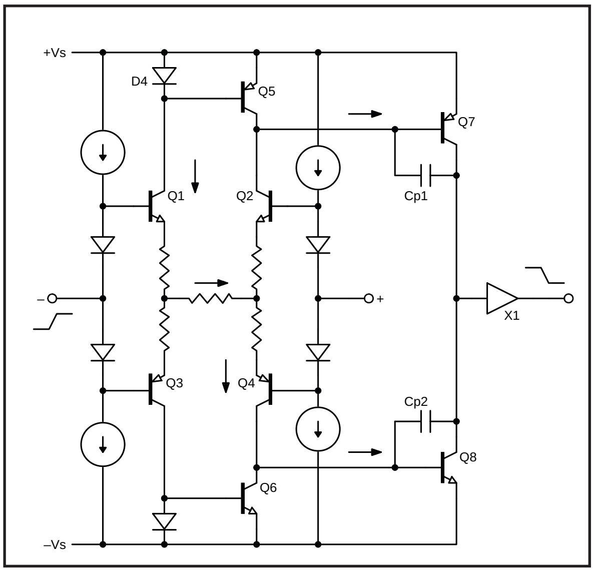
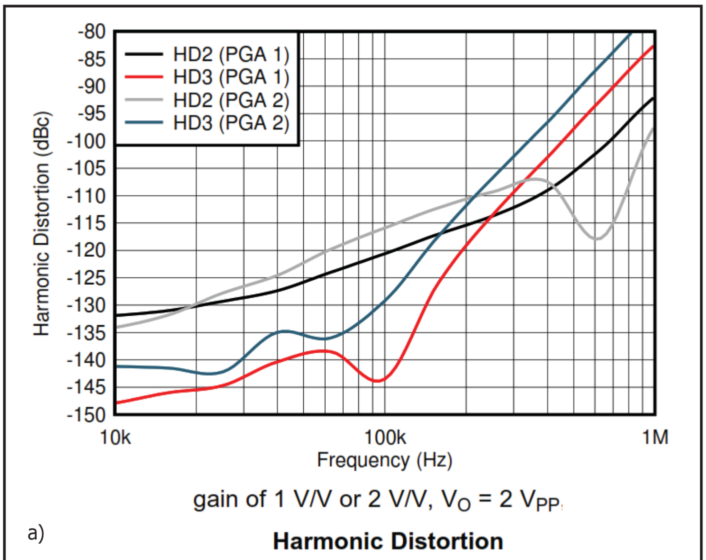


Figure 6: This is the basic schematic of Analog Device's "Quad Core" input stage. (Original image courtesy of Analog Devices)

One of the op-amps employing this technique is the AD8038. With a supply current of only 1.5 mA per amplifier, it achieves a gain bandwidth product of 350 MHz and a slew rate of 425/µs! Another approach to enhancing slew rates is to sense the difference in input voltages and increase the input stage "DC" bias current as this difference increases. Neither scheme adds to quiescent dissipation. Manufacturers use these and other approaches to achieve similar results.

VFA-CFA Shootout

I won't claim that the devices I've chosen to compare are the very best of their kind. I'm not sure that "bests" exist. Certain parameters are superior in other parts. But I do believe that like Ralph's Pretty Good Grocery Store in Lake Wobegone (a fictional town that served as the setting for stories from longtime radio broadcast *A Prairie*



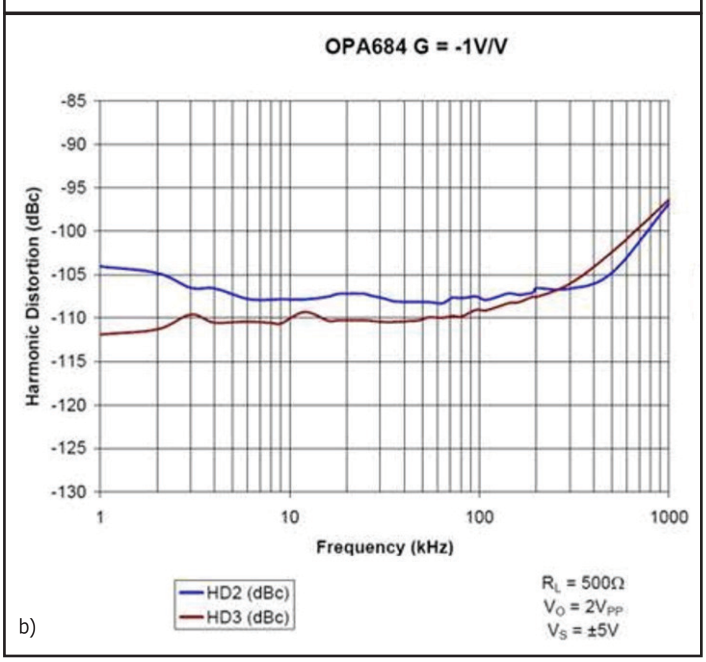


Figure 7a: Harmonic distortion of the VFA OPA837; Figure 7b: Harmonic distortion of the CFA OPA684 (Image courtesy of Texas Instruments)

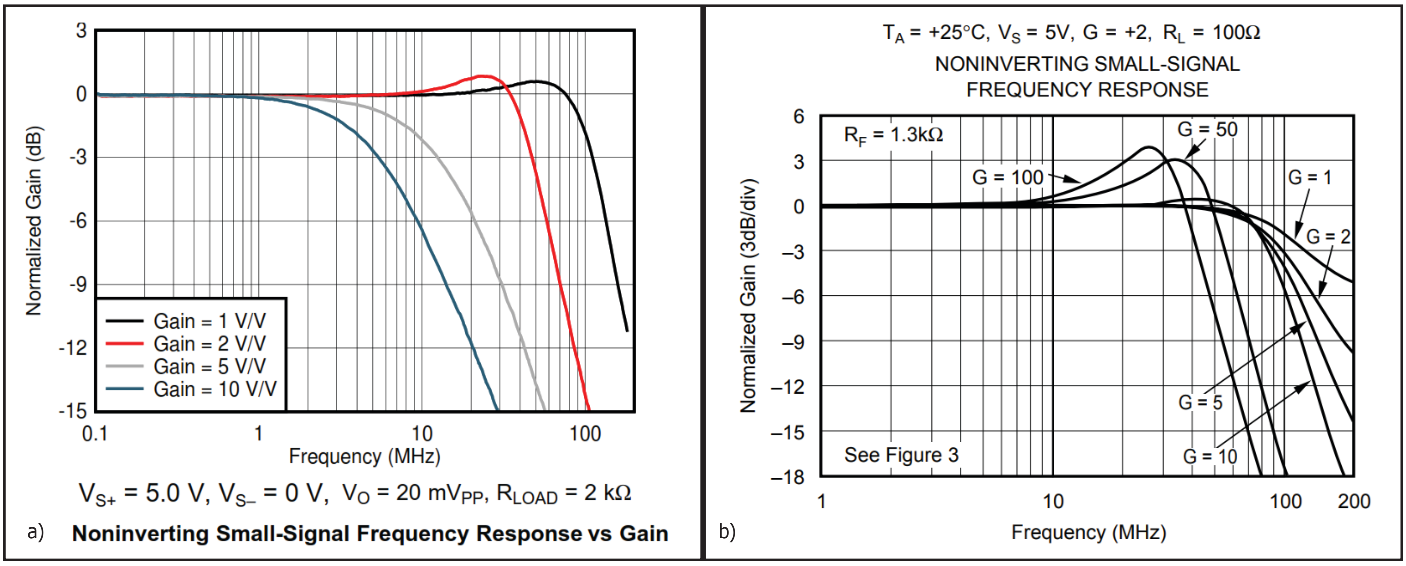


Figure 8a: Non-inverting small-signal frequency response of the VFA OPA837; Figure 8b: Non-inverting small-signal frequency response of the CFA OPA684 (Image courtesy of Texas Instruments)

Home Companion), these parts are pretty good—maybe even pretty darn good.

We are fortunate to have several graphs that display characteristics of interest to audiophiles for the OPA387 VFA and the OPA684 CFA. Unless explicitly stated otherwise, I'll restrict my comments to the audio frequency range. The graphs shown in **Figures 7-12** are courtesy of TI.

Traditionally, one of the first things we look at is distortion performance. With a closed loop gain (CLG) of unity, we see total harmonic distortion (THD) better than -100 dB (0.001%), although this VFA at audio frequencies is perhaps 30 dB better. We expect the relative distortions at higher CLGs to move toward favoring the CFA because of its much slower fall in open loop gain (OLG), as shown in Figure 5 from Part 1, but that data is unavailable. (Note: The CFA distortion graph generated by TI was never updated in the datasheet.) Next, we note the VFA's inverse relationship between bandwidth and CLG in comparison to the CFA's bandwidth change of a mere factor of three as CLG varies by a factor of 100. Moving on, the CLG output impedance and power supply rejection ratio (PSRR) of both units and the CFA's common-mode rejection ratio (CMRR) closely track the shapes of their OLGs. Interestingly, the VFA's CMRR has a much wider bandwidth than its OLG. Regarding audio frequency output impedance, the VFA out-performs the CFA because of its greater loop gain. As for speed, the CFA is about 10 times

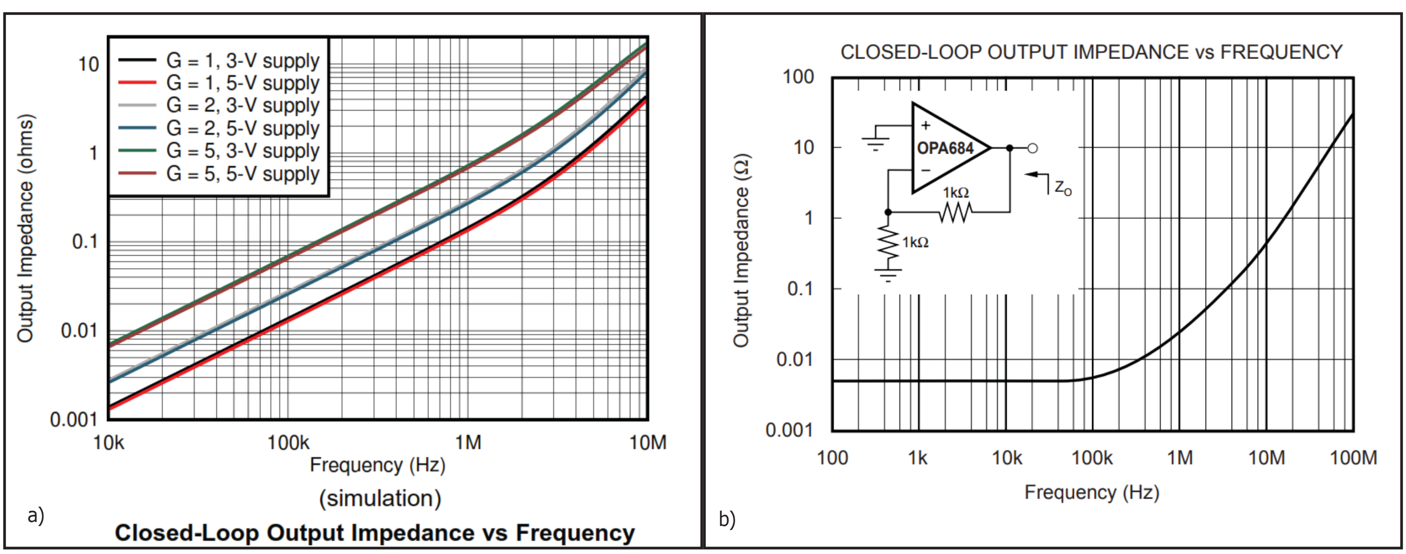


Figure 9a: Closed-loop output impedance of the VFA OPA837; Figure 9b: Closed-loop output impedance of the CFA OPA684 (Image courtesy of Texas Instruments)

faster at a gain of 2 than the VFA (note the difference in time scales.) As with bandwidth, this difference would be even more extreme at higher CLGs. Note also the comment about the need to limit the slew rate of the signal applied to the VFA to preclude slewing-induced distortion. This is of little concern at audio frequencies for this device. Finally, the voltage noise of the two are about equal, with an edge given to the CFA. But VFA current noise is better by a factor of 2 to 4.

Additional Comparisons

This VFA has a maximum offset voltage of 200 µV vs. the CFA's 4.3 mV. When it comes to bias current, the VFA's maximum is 718 nA vs. the CFA's 12 µA at its non-inverting input and 18.5 µA at its inverting one. The CFA's offset current isn't even specified. Its non-inverting single-ended input impedance is typically 50 kΩ || 2 pF, while the VFA's differential impedance is typically 180 kΩ || 0.5 pF. Our VFA draws 865 µA from a supply of up to 5 V, while the CFA

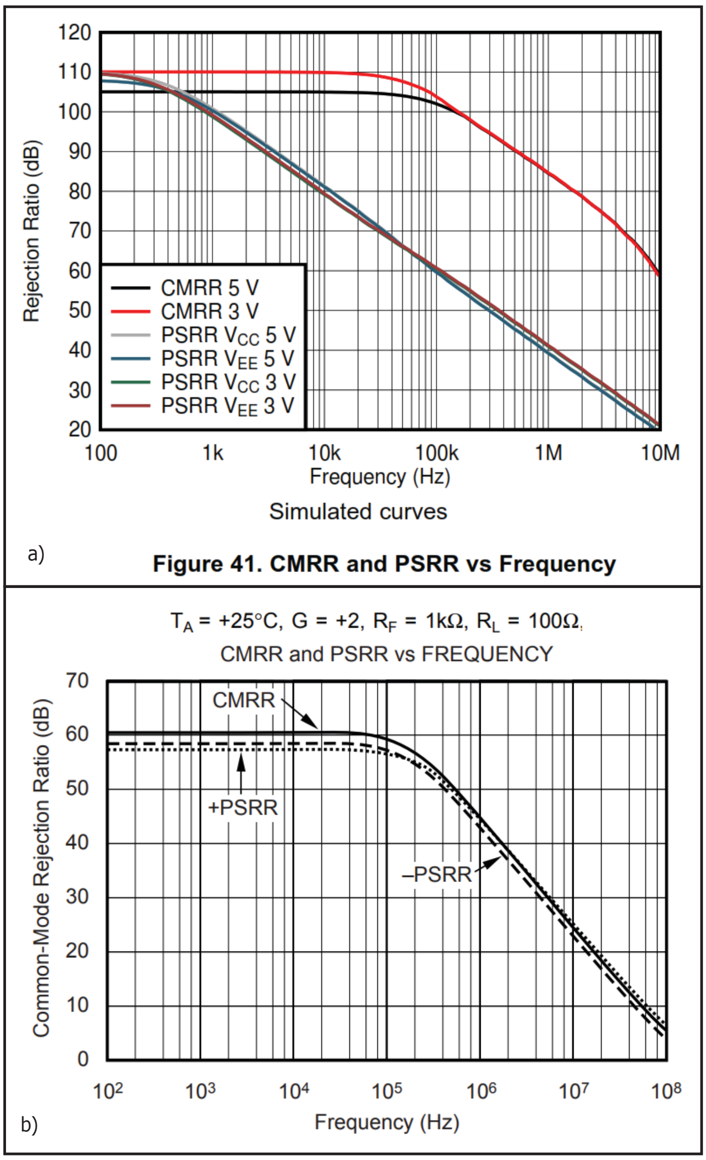


Figure 10a: CMRR and PSRR of the VFA OPA837; Figure 10b: CMRR and PSRR of the CFA OPA684 (Image courtesy of Texas Instruments)

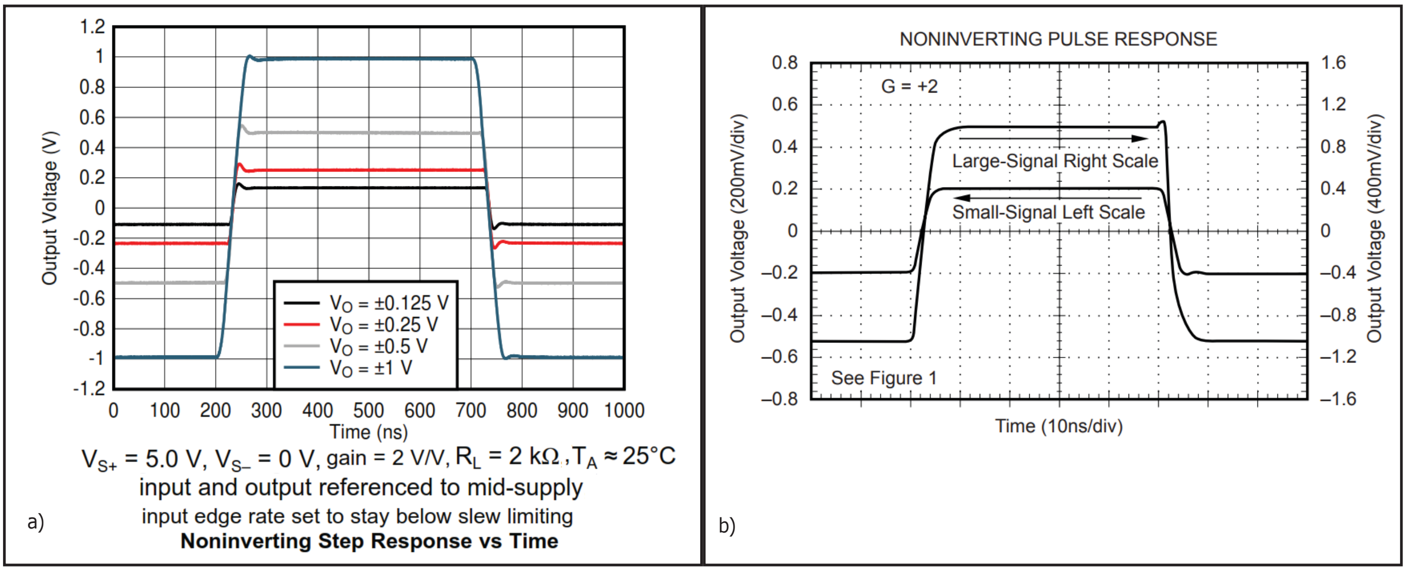


Figure 11a: Non-inverting step response of the VFA OPA837; Figure 11b: Non-inverting step response of the CFA OPA684 (Image courtesy of Texas Instruments)

consumes 1.85 mA from ±6 V rails.

A CFA's bandwidth and slew rate are highly dependent on the value of R_f (see Figure 4). The constraint on R_f's value forces a commensurate restraint on R_g for a given gain, which limits the usefulness of CFAs in inverting modes of operation. There are no such limitations on a VFA.

Summing Up

Current feedback is real. Unfortunately, it has multiple definitions. We must be aware of the context in which the term is used. The industry has taken to calling a specific type of amplifier a Current Feedback Amplifier (see Figure 2), and the reason for this name has been discussed.

In Part 1, problems were noted with the traditional associations of voltage feedback with series applied feedback and current feedback with shunt applied feedback. It is this author's opinion that we may retain series/shunt derived/applied to describe signal routing. But we must avoid associating the signal routing type with exclusively voltage or current feedback. This is because "Applied feedback" is "applied" to both CFA and VFA amplifiers.

The feedback type accepted by an amplifier can be distinguished by the source of the output current of its input stage. If none of this comes from the amplifier output, voltage feedback is the obvious choice. If all or much of it derives from the amplifier output, then we have current feedback.

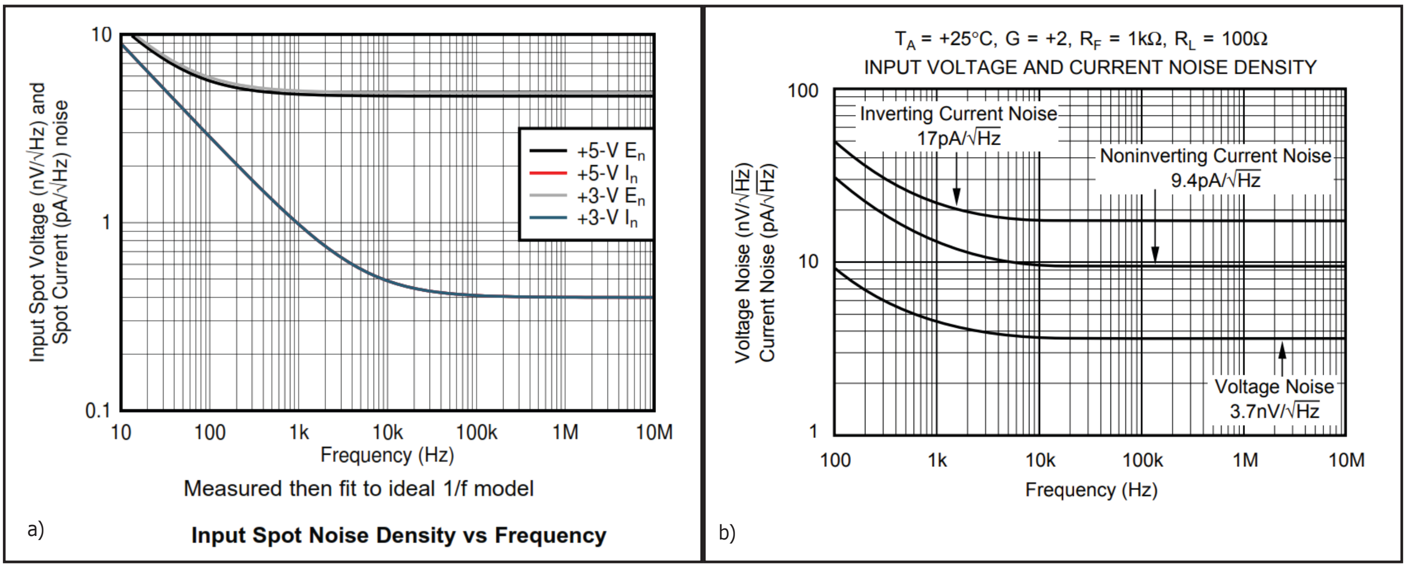


Figure 12a: Input noise density of the VFA OPA837; Figure 12b: Input noise density of the CFA OPA684 (Image courtesy of Texas Instruments)

An exemplary CFA and an exemplary VFA have been considered. These devices are generally representative of members of these two amplifier classes. Certain performance characteristics of interest to audiophiles have been presented.

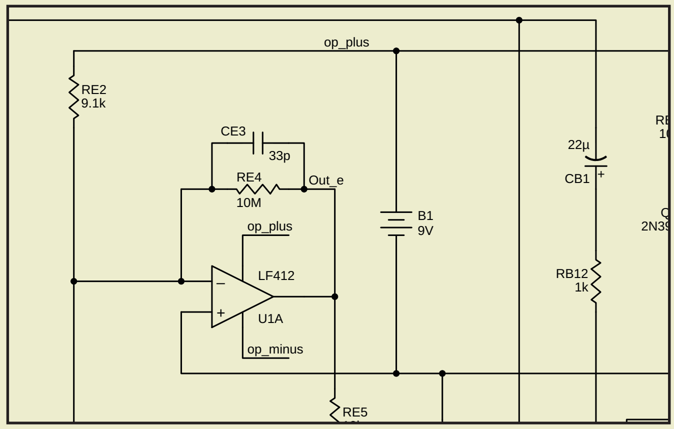
Amplifiers that employ current feedback can readily exhibit constant bandwidth with varying closed loop gain. All that is needed is for the network impedance R_{eq} seen by the inverting input to be kept larger than that input's impedance Z_{FB}. This effect can also be exhibited in voltage feedback amplifiers by adding a suitable resistor between their inputs (see Part 1, Figure 6). However, there is no good reason to do this, as it reduces bandwidth. Regardless of these relative impedances, current feedback amplifiers inherently benefit from slew rate enhancement, putting the current flowing from the output into their inverting inputs to good use. Slew rates of voltage feedback amplifiers can be increased by additional specialized circuitry within the amplifier.

Voltage feedback amplifiers' generally higher audio frequency loop gains usually afford them lower distortion and higher CMRRs and PSRRs. Their input structures typically allow them to benefit from lower input noise, bias currents, and better DC characteristics. They also lack the inverting mode limitations of current feedback amplifiers. The slew rates and bandwidths of modern units, though perhaps not as great as those of current feedback alternatives, may not be barriers to use in audio designs. Readers can draw their own conclusions as to which class is better suited to a particular audio application.

About the Author

Christopher Paul was born in the US the year before the Nobel prize was awarded for the invention of the transistor, the development of which he claims no credit for. Chris has worked for companies whose products are in the fields of communications, electronic article surveillance, and enterprise hand-held computers. He holds two M.Sc. degrees in electrical engineering from Brooklyn Polytechnic (now part of New York) University.

Figure Correction for Part 1 of "Current Feedback: Fake News or the Real Deal?"



Part 1 of the "Current Feedback: Fake News or the Real Deal?" article, which ran in the July 2018 issue of *audioXpress*, contained an incomplete figure. The label "Out_e" was left off of Figure 2. It should have been associated with the output of op-amp U1a. Here is the portion of Figure 2 that was missing.