Ask The Applications Engineer—22

by Erik Barnes

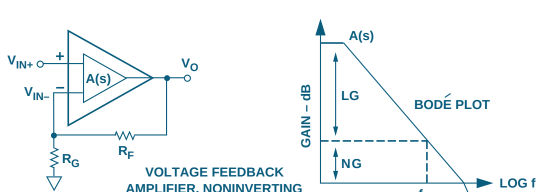
CURRENT FEEDBACK AMPLIFIERS—I

Q. I'm not sure I understand how current-feedback amplifiers work as compared with regular op amps. I've heard that their bandwidth is constant regardless of gain. How does that work? Are they the same as transimpedance amplifiers?

A. Before looking at any circuits, let's define voltage feedback, current feedback, and transimpedance amplifier. *Voltage feedback*, as the name implies, refers to a closed-loop configuration in which the error signal is in the form of a voltage. Traditional op amps use voltage feedback, that is, their inputs will respond to voltage changes and produce a corresponding output voltage. *Current feedback* refers to any closed-loop configuration in which the error signal used for feedback is in the form of a current. A current feedback op amp responds to an error current at one of its input terminals, rather than an error voltage, and produces a corresponding output voltage. Notice that both open-loop architectures achieve the same closed-loop result: zero differential input voltage, and zero input current. The ideal voltage feedback amplifier has high-impedance inputs, resulting in zero input current, and uses voltage feedback to maintain zero input voltage. Conversely, the current feedback op amp has a low impedance input, resulting in zero input *voltage*, and uses current feedback to maintain zero input *current*.

The transfer function of a *transimpedance amplifier* is expressed as a voltage output with respect to a current input. As the function implies, the open-loop "gain", v_O/i_{IN}, is expressed in ohms. Hence a current-feedback op amp can be referred to as a *transimpedance amplifier*. It's interesting to note that the closed-loop relationship of a voltage-feedback op amp *circuit* can also be configured as a transimpedance, by driving its dynamically low-impedance summing node with current (e.g., from a photodiode), and thus generating a voltage output equal to that input current multiplied by the feedback resistance. Even more interesting, since ideally any op amp application can be implemented with either voltage or current feedback, this same I-V converter can be implemented with a current feedback op amp. When using the term *transimpedance amplifier*, understand the difference between the specific current-feedback op amp architecture, and any closed-loop I-V converter circuit that acts like transimpedance.

Let's take a look at the simplified model of a voltage feedback amplifier. The noninverting gain configuration amplifies the difference voltage, $(V_{IN+} - V_{IN-})$, by the open loop gain $A(s)$ and feeds a portion of the output back to the inverting input through the voltage divider consisting of R_F and R_G. To derive the closed-loop transfer function of this circuit, V_o/V_{IN+}, assume



VOLTAGE FEEDBACK AMPLIFIER, NONINVERTING GAIN CONNECTION

that no current flows into the op amp (infinite input impedance); both inputs will be at about the same potential (negative feedback and high open-loop gain)).

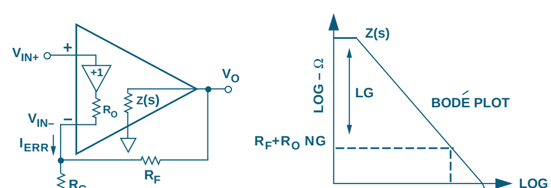
With $$V_o = (V_{IN+} - V_{IN-})A(s)$$

and $$V_{IN-} = \frac{R_G}{R_G + R_F}V_o$$

substitute and simplify to get:

$$\frac{V_o}{V_{IN}} = \left(1 + \frac{R_F}{R_G}\right)\frac{1}{1 + \frac{1}{LG}} \quad where\ LG = \frac{A(s)}{1 + \frac{R_F}{R_G}}$$

The closed-loop bandwidth is the frequency at which the loop gain, LG, magnitude drops to unity (0 dB). The term, $1 + R_F/R_G$, is called the *noise gain* of the circuit; for the noninverting case, it is also the signal gain. Graphically, the closed-loop bandwidth is found at the intersection of the open-loop gain, $A(s)$, and the noise gain, NG, in the Bodé plot. High noise gains will reduce the loop gain, and thereby the closed-loop bandwidth. If $A(s)$ rolls off at 20 dB/decade, the gain-bandwidth product of the amplifier will be constant. Thus, an increase in closed-loop gain of 20 dB will reduce the closed-loop bandwidth by one decade.



CURRENT FEEDBACK AMPLIFIER, NONINVERTING GAIN CONNECTION

Consider now a simplified model for a current-feedback amplifier. The noninverting input is the high-impedance input of a unity gain buffer, and the inverting input is its low-impedance output terminal. The buffer allows an error current to flow in or out of the inverting input, and the unity gain forces the inverting input to track the noninverting input. The error current is mirrored to a high impedance node, where it is converted to a voltage and buffered at the output. The high-impedance node is a frequency-dependent impedance, $Z(s)$, analogous to the open-loop gain of a voltage feedback amplifier; it has a high dc value and rolls off at 20 dB/decade.

The closed-loop transfer function is found by summing the currents at the V_{IN-} node, while the buffer maintains $V_{IN+} = V_{IN-}$. If we assume, for the moment, that the buffer has zero output resistance, then $R_o = 0\Omega$

$$\frac{V_o - V_{IN-}}{R_F} + \frac{-V_{IN-}}{R_G} + I_{err} = 0 \ and\ I_{err} = V_0 / Z(s)$$

Substituting, and solving for V_o/V_{IN+}

$$\frac{V_o}{V_{IN+}} = \left(1 + \frac{R_F}{R_G}\right)\frac{1}{1 + \frac{1}{LG}}, \text{ where } LG = \frac{Z(s)}{R_F}$$

The closed-loop transfer function for the current feedback amplifier is the same as for the voltage feedback amplifier, but the loop gain $(1/LG)$ expression now depends only on R_F, the feedback transresistance—and not $(1 + R_F/R_G)$. Thus, the closed-loop bandwidth of a current feedback amplifier will vary with the value of R_F, but not with the noise gain, $1 + R_F/R_G$. The intersection of R_F and $Z(s)$ determines the loop gain, and thus the closed-loop bandwidth of the circuit (see Bodé plot). Clearly the gain-bandwidth product is not constant—an advantage of current feedback.

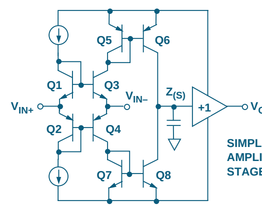
In practice, the input buffer's non-ideal output resistance will be typically about 20 to 40 Ω, which will modify the feedback transresistance. The two input voltages will not be exactly equal. Making the substitution into the previous equations with $V_{IN-} = V_{IN+} - I_{err}R_o$, and solving for V_o/V_{IN+} yields:

$$\frac{V_o}{V_{IN}} = \left(1+\frac{R_F}{R_G}\right)\frac{1}{1+\frac{1}{LG}}, \text{ where } LG = \frac{Z(s)}{R_F + R_o\left(1+\frac{R_F}{R_G}\right)}$$

The additional term in the feedback transresistance means that the loop gain will actually depend somewhat on the closed-loop gain of the circuit. At low gains, R_F dominates, but at higher gains, the second term will increase and reduce the loop gain, thus reducing the closed-loop bandwidth.

It should be clear that shorting the output back to the inverting input with R_G open (as in a voltage follower) will force the loop gain to get very large. With a voltage feedback amplifier, maximum feedback occurs when feeding back the entire output voltage, but the current feedback's limit is a short-circuit current. The lower the resistance, the higher the current will be. Graphically, $R_F = 0$ will give a higher-frequency intersection of $Z(s)$ and the feedback transresistance—in the region of higher-order poles. As with a voltage feedback amplifier, higher-order poles of $Z(s)$ will cause greater phase shift at higher frequencies, resulting in instability with phase shifts > 180 degrees. Because the optimum value of R_F will vary with closed-loop gain, the Bode plot is useful in determining the bandwidth and phase margin for various gains. A higher closed-loop bandwidth can be obtained at the expense of a lower phase margin, resulting in peaking in the frequency domain, and overshoot and ringing in the time domain. Current-feedback device data sheets will list specific optimum values of R_F for various gain settings.

Current feedback amplifiers have excellent slew-rate capabilities. While it is possible to design a voltage-feedback amplifier with high slew rate, the current-feedback architecture is inherently faster. A traditional voltage-feedback amplifier, lightly loaded, has a slew rate limited by the current available to charge and discharge the internal compensation capacitance. When the input is subjected to a large transient, the input stage will saturate and only its tail current is available to charge or discharge the compensation node. With a current-feedback amplifier, the low-impedance input allows higher transient currents to flow into the amplifier as needed. The internal current mirrors convey this input current to the compensation node, allowing fast charging and discharging—theoretically, in proportion to input step size. A faster slew rate will result in a quicker rise time, lower slew-induced distortion and nonlinearity, and a wider large-signal frequency response. The actual slew rate will be limited by saturation of the current mirrors, which can occur at 10 to 15 mA, and the slew-rate limit of the input and output buffers.



SIMPLIFIED CURRENT FEEDBACK AMPLIFIER, ILLUSTRATING INPUT STAGE AND CURRENT MIRRORS.

Q. What about dc accuracy?

A. The dc gain accuracy of a current feedback amplifier can be calculated from its transfer function, just as with a voltage feedback amplifier; it is essentially the ratio of the internal transresistance to the feedback transresistance. Using a typical transresistance of 1 MΩ, a feedback resistor of 1 kΩ, and an R_o of 40 ohms, the gain error at unity gain is about 0.1%. At higher gains, it degrades significantly. Current-feedback amplifiers are rarely used for high gains, particularly when absolute gain accuracy is required.

For many applications, though, the settling characteristics are of more importance than gain accuracy. Although current feedback amplifiers have very fast rise times, many data sheets will only show settling times to 0.1%, because of thermal settling tails— a major contributor to lack of settling precision. Consider the complementary input buffer above, in which the V_{IN-} terminal is offset from the V_{IN+} terminal by the difference in V_{BE} between Q1 and Q3. When the input is at zero, the two V_{BE}s should be matched, and the offset will be small from V_{IN+} to V_{IN-}. A positive step input applied to V_{IN+} will cause a reduction in the V_{CE} of Q3, decreasing its power dissipation, thus increasing its V_{BE}. Diode-connected Q1 does not exhibit a V_{CE} change, so its V_{BE} will not change. Now a different offset exists between the two inputs, reducing the accuracy. The same effect can occur in the current mirror, where a step change at the high-impedance node changes the V_{CE}, and thus the V_{BE}, of Q6, but not of Q5. The change in V_{BE} causes a current error referred back to V_{IN-}, which—multiplied by R_F—will result in an output offset error. Power dissipation of each transistor occurs in an area that is too small to achieve thermal coupling between devices. Thermal errors in the input stage can be reduced in applications that use the amplifier in the inverting configuration, eliminating the common-mode input voltage.

Q. In what conditions are thermal tails a problem?

A. It depends on the frequencies and waveforms involved. Thermal tails do not occur instantaneously; the thermal coefficient of the transistors (which is process dependent) will determine the time it takes for the temperature change to occur and alter parameters—and then recover. Amplifiers fabricated on the Analog Devices high-speed complementary bipolar (CB) process, for example, don't exhibit significant thermal tails for input frequencies above a few kHz, because the input signal is changing too fast. Communications systems are generally more concerned with spectral performance, so additional gain errors that might be introduced by thermal tails are not important. Step waveforms, such as those found in imaging applications, can be adversely affected by thermal tails when dc levels change. For these applications, current-feedback amplifiers may not offer adequate settling accuracy.

Part II will consider common application circuits using current-feedback amplifiers and view their operation in more detail.

Ask The Applications Engineer—23

by Erik Barnes

CURRENT FEEDBACK AMPLIFIERS—II

Part I (*Analog Dialogue* 30-3) covers basic operation of the current-feedback (CF) op-amp. This second part addresses frequently asked questions about common applications.

Q. *I now have better understanding of how a current feedback op-amp works, but I'm still confused when it comes to applying one in a circuit. Does the low inverting input impedance mean I can't use the inverting gain configuration?*

A. Remember that the inverting mode of operation works *because* of the low-impedance node created at the inverting input. The summing junction of a voltage-feedback (VF) amplifier is characterized by a low input impedance after the feedback loop has settled. A current feedback op amp will, in fact, operate very well in the inverting configuration because of its inherently low inverting-input impedance, holding the summing node at "ground," even before the feedback loop has settled. CF types don't have the voltage spikes that occur at the summing node of voltage feedback op amps in high-speed applications. You may also recall that advantages of the inverting configuration include maximizing input slew rate and reducing thermal settling errors.

Q. *So this means I can use a current feedback op-amp as a current-to-voltage converter, right?*

A. Yes, they can be configured as I-to-V converters. But there are limitations: the amplifier's bandwidth varies directly with the value of feedback resistance, and the inverting input current noise tends to be quite high. When amplifying low level currents, higher feedback resistance means higher signal-to-(resistor-) noise ratio, because signal gain will increase proportionally, while resistor noise goes as $\sqrt{R}$. Doubling the feedback resistance doubles the signal gain and increases resistor noise by a only factor of 1.4; unfortunately the contribution from current noise is doubled, and, with a current feedback op amp, the signal bandwidth is halved. Thus the higher current noise of CF op amps may preclude their use in many photodiode-type applications. When noise is less critical, select the feedback resistor based on bandwidth requirements; use a second stage to add gain.

Q. *I did notice the current noise is rather high in current feedback amplifiers. So will this limit the applications in which I can use them?*

A. Yes, the inverting input current noise tends to be higher in CF op amps, around 20 to 30 pA/$\sqrt{\text{Hz}}$. However, the input voltage noise tends to be quite low when compared with similar voltage feedback parts, typically less than 2 nV/$\sqrt{\text{Hz}}$, and the feedback resistance will also be low, usually under 1 kΩ. At a gain of 1, the dominant source of noise will be the inverting-input noise current flowing through the feedback resistor. An input noise current of 20 pA/$\sqrt{\text{Hz}}$ and an R_F of 750 Ω yields 15 nV/$\sqrt{\text{Hz}}$ as the dominant noise source at the output. But as the gain of the circuit is increased (by reducing input resistance), the output noise due to input current noise will not increase, and the amplifier's input *voltage* noise will become the dominant factor. At a gain, of say, 10, the contribution from the input noise current is only 1.5 nV/$\sqrt{\text{Hz}}$ when referred to the input; added to the input voltage noise of the amplifier in RSS fashion, this gives an input-referred noise voltage of only 2.5 nV/$\sqrt{\text{Hz}}$ (neglecting resistor noise). Used thus, the CF op amp becomes attractive for a low noise application.

Q. *What about using the classic four-resistor differential configuration? Aren't the two inputs unbalanced and therefore not suitable for this type of circuit?*

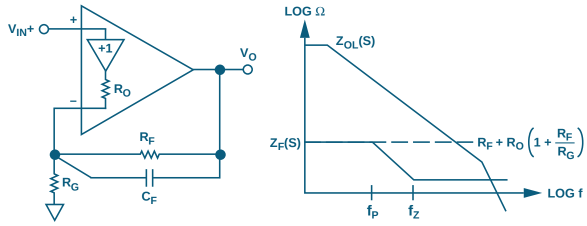
A. I'm glad you asked; this is a common misconception of CF op-amps. True, the inputs are not matched, but the transfer function for the ideal difference amplifier will still work out the same. What about the unbalanced inputs? At lower frequencies, the four-resistor differential amplifier's CMR is limited by the matching of the external resistor ratios, with 0.1% matching yielding about 66 dB. At higher frequencies, what matters is the matching of time constants formed by the input impedances. High-speed voltage-feedback op amps usually have pretty well matched input capacitances, achieving CMR of about 60 dB at 1 MHz. Because the CF amplifier's input stage is unbalanced, the capacitances may not be well matched. This means that small external resistors (100 to 200 Ω) must be used on the noninverting input of some amplifiers to minimize the mismatch in time constants. If careful attention is given to resistor selection, a CF op-amp can yield high frequency CMR comparable to a VF op amp. If higher performance is needed, the best choice would be a monolithic high speed *difference amplifier*, such as the AD830. Requiring no resistor matching, it has a CMR > 75 dB at 1 MHz and about 53 dB at 10 MHz.

Q. *What about trimming the amplifier's bandwidth with a feedback capacitor? Will the low impedance at the inverting input make the current feedback op amp less sensitive to shunt capacitance at this node? How about capacitive loads?*

A. First consider a capacitor in the feedback path. With a voltage feedback op amp, a pole is created in the noise gain, but a pole and a zero occur in the feedback transresistance of a current feedback op amp, as shown in the figure below. Remember that the phase margin at the intersection of the feedback transresistance and the open loop transimpedance will determine closed-loop stability. Feedback transresistance for a capacitance, C_F, in parallel with R_F, is given by

$$Z_F(s) = \left[R_F + R_O\left(1+\frac{R_F}{R_G}\right)\right]\frac{1+\dfrac{sC_FR_FR_GR_O}{R_FR_G+R_FR_O+R_GR_O}}{1+sC_FR_F}$$

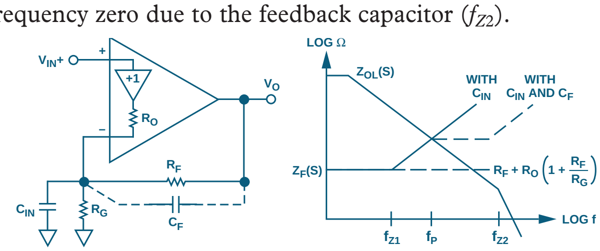
The pole occurs at $1/2\pi R_FC_F$, and the zero occurs higher in frequency at $1/[2\pi(R_F||R_G||R_O)C_F]$. If the intersection of Z_F and Z_{OL} occurs too high in frequency, instability may result from excessive open loop phase shift. If $R_F \rightarrow \infty$, as with an integrator circuit, the pole occurs at a low frequency and very little resistance exists at higher frequencies to limit the loop gain. A CF integrator can be stabilized by a resistor in series with the integrating capacitor to limit loop gain at higher frequencies. Filter topologies that use reactive feedback, such as multiple feedback types, are not suitable for CF op amps; but Sallen-Key filters, where the op amp is used as a fixed-gain block, are feasible. In general, it is not desirable to add capacitance across R_F of a CF op amp.



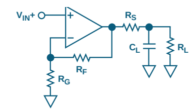
Another issue to consider is the effect of shunt capacitance at the inverting input. Recall that with a voltage feedback amplifier, such capacitance creates a zero in the noise gain, increasing the rate of closure between the noise gain and open loop gain, generating excessive phase shift that can lead to instability if not compensated for. The same effect occurs with a current feedback op amp, but the problem may be less pronounced. Writing the expression for the feedback transresistance with the addition of C_{IN}:

$$Z_F(s)=\left[R_F+R_O\left(1+\frac{R_F}{R_G}\right)\right]\left[1+\frac{sC_{IN}R_FR_GR_O}{R_FR_G+R_FR_O+R_GR_O}\right]$$

A zero occurs at $1/[2\pi(R_F||R_G||R_O)C_{IN}]$, shown in the next figure (f_{Z1}). This zero will cause the same trouble as with a VF amplifier, but the corner frequency of the zero tends to be higher in frequency because of the inherently low input impedance at the inverting input. Consider a wideband voltage feedback op amp with R_F = 750 Ω, R_G = 750 Ω, and C_{IN} = 10 pF. The zero occurs at $1/[2\pi(R_F||R_G)C_{IN}]$, roughly 40 MHz, while a current feedback op-amp in the same configuration with an R_O of 40 Ω will push the zero out to about 400 MHz. Assuming a unity gain bandwidth of 500 MHz for both amplifiers, the VF amplifier will require a feedback capacitor for compensation, reducing the effect of C_{IN}, but also reducing the signal bandwidth. The CF device will certainly see some additional phase shift from the zero, but not as much because the break point is a decade higher in frequency. Signal bandwidth will be greater, and compensation may only be necessary if in-band flatness or optimum pulse response is required. The response can be tweaked by adding a small capacitor in parallel with R_F to reduce the rate of closure between Z_F and Z_{OL}. To ensure at least 45° of phase margin, the feedback capacitor should be chosen to place a pole in the feedback transresistance where the intersection of Z_F and Z_{OL} occurs, shown here (f_P). Don't forget the effects of the higher frequency zero due to the feedback capacitor (f_{Z2}).



Load capacitance presents the same problem with a current feedback amplifier as it does with a voltage feedback amplifier: increased phase shift of the error signal, resulting in degradation of phase margin and possible instability. There are several well-documented circuit techniques for dealing with capacitive loads, but the most popular for high speed amplifiers is a resistor in series with the output of the amplifier (as shown below). With the resistor outside the feedback loop, but in series with the load capacitance, the amplifier doesn't directly drive a purely capacitive load. A CF op amp also gives the option of increasing R_F to reduce the loop gain. Regardless of the approach taken, there will always be a penalty in bandwidth, slew rate, and settling time. It's best to experimentally optimize a particular amplifier circuit, depending on the desired characteristics, e.g., fastest rise time, fastest settling to a specified accuracy, minimum overshoot, or passband flatness.



Q. *Why don't any of your current feedback amplifiers offer true single-supply operation, allowing signal swings to one or both rails?*

A. This is one area where the VF topology is still favored for several reasons. Amplifiers designed to deliver good current drive and to swing close to the rails usually use common-emitter output stages, rather than the usual emitter followers. Common emitters allow the output to swing to the supply rail minus the output transistors' V_{CE} saturation voltage. With a given fabrication process, this type of output stage does not offer as much speed as emitter followers, due in part to the increased circuit complexity and inherently higher output impedance. Because CF op amps are specifically developed for the highest speed and output current, they feature emitter follower output stages.

With higher speed processes, such as ADI's XFCB (extra-fast complementary bipolar), it has been possible to design a common-emitter output stage with 160-MHz bandwidth and 160-V/µs slew rate, powered from a single 5-volt supply (AD8041). The amplifier uses voltage feedback, but even if, somehow, current feedback had been used, speed would still be limited by the output stage. Other XFCB amplifiers, with emitter-follower output stages (VF or CF), are much faster than the AD8041. In addition, single-supply input stages use PNP differential pairs to allow the common-mode input range to extend down to the lower supply rail (usually ground). To design such an input stage for CF is a major challenge, not yet met at this writing.

Nevertheless, CF op amps can be used in single-supply applications. Analog Devices offers many amplifiers that are specified for +5- or even +3-volt operation. What must be kept in mind is that the parts operate well off a single supply *if the application remains within the allowable input and output voltage ranges*. This calls for level shifting or ac coupling and biasing to the proper range, but this is already a requirement in most single-supply systems. If the system must operate to one or both rails, or if the maximum amount of headroom is demanded in ac-coupled applications, a current feedback op amp may simply not be the best choice. Another factor is the rail-to-rail output swing specifications when driving heavy loads. Many so-called rail-to-rail parts don't even come close to the rails when driving back- terminated 50- or 75-Ω cables, because of the increase in V_{CESAT} as output current increases. If you really need true rail-to-rail performance, you don't want or need a current feedback op amp; if you need highest speed and output current, this is where CF op amps excel. ▶