Ask The Applications Engineer—22

by Erik Barnes

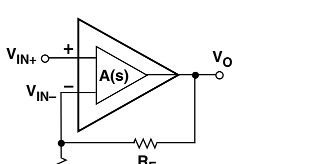
CURRENT FEEDBACK AMPLIFIERS—I

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?

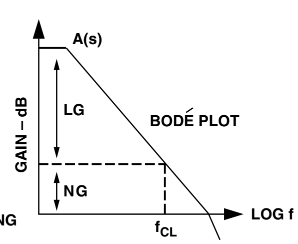
A. Before looking at any circuits, let's define voltage feedback, current feedback, and transimpedance amplifier. *Voltage feedback*, as the name implies, refers to a closed-loop configuration in which the error signal is in the form of a voltage. Traditional op amps use voltage feedback, that is, their 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 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)).



VOLTAGE FEEDBACK AMPLIFIER, NONINVERTING GAIN CONNECTION

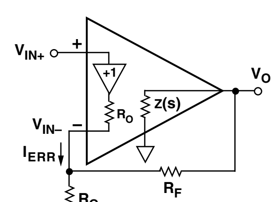


With $V_o = (V_{IN+} - V_{IN-})A(s)$

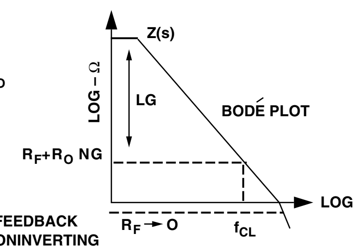
and

substitute and simplify to get:

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 $R_o = 0$

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

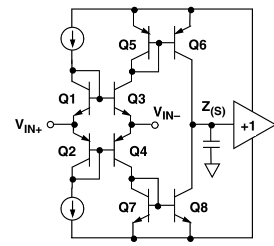
The closed-loop transfer function for the current feedback amplifier is the same as for the voltage feedback amplifier, but the loop gain $(1/LG)$ expression now depends only on R_F, the feedback transresistance—and not $(1 + R_F/R_G)$. Thus, the closed-loop bandwidth of a current feedback amplifier 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:

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. A