

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 Bode plot). Clearly the gain-bandwidth product is not constant—an advantage of current feedback.

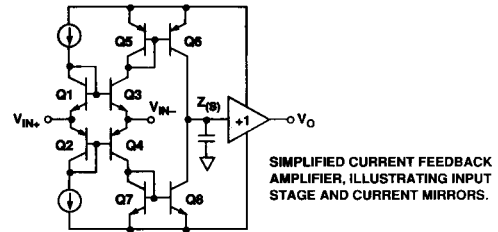
In practice, the input buffer's non-ideal output resistance will be typically about 20 to 40  $\Omega$ , which will modify the feedback transresistance. The two input voltages will not be exactly equal. Making the substitution into the previous equations with  $V_{IN-} = V_{IN+} - I_{err}R_o$ , and solving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



**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 M $\Omega$ , a feedback resistor of 1 k $\Omega$ , 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.* ▶