Current-feedback amplifiers benefit high-speed designs

Current-feedback amplifiers offer significant advantages over conventional high-speed op amps. Like the conventional devices, however, they exhibit nonideal behavior, so some circuit configurations require special care. Understanding the circuit topology will help you achieve successful designs.

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Amplifiers based on the current-feedback topology are now more widely available than ever. They offer designers of high-speed systems some key advantages over conventional op amps (**Ref 1**). First, you can independently vary their gain and bandwidth; second, they have a virtually unlimited slew rate. The absence of slew-rate limiting not only allows for faster settling times, but also eliminates slew-rate-related nonlinearities such as intermodulation distortion. Thus, current-feedback amps are attractive for use in high-quality audio-amplifier applications.

These two advantages are the result of the amps' current-mode operation, which has long been recognized as inherently faster than voltage-mode operation. The effects of stray inductance in a circuit are usually less severe than those of stray capacitance (or the Miller effect), and bipolar transistors can switch currents much more rapidly than voltages. Current amplifiers must still have a voltage output, however, and op-amp designers sidestep some of the problems associated with voltage-mode operation by using gain configurations such as common-collector and cascode configurations, which provide immunity to the Miller effect. Further, thanks to manufacturing processes that ensure symmetrical npn- and pnp-transistor switching characteristics, manufacturers can now create monolithic op amps that achieve high speeds that were previously available only from hybrid devices.

In many ways, current-feedback amps are very similar to their conventional op-amp counterparts (**Ref 2**). For a standard circuit configuration, you derive the transfer functions for current-feedback amplifiers in the same way that you do for conventional op amps. However, if you're going to use a current-feedback amp in your design, you'll have some other considerations to make. For example, you'll have to decide how to use reactive feedback elements, which cause oscillation when connected directly from output to input. Thus, before designing with current-feedback amps, you need a thorough understanding of the current-feedback architecture.

The easiest way to understand the advantages of the current-feedback topology is to compare it with the architecture of a conventional op amp (**Ref 3**). The conventional op amp consists of a high-input-impedance differential stage followed by additional gain stages, *Current-feedback amps don't involve a gain-bandwidth tradeoff, and they have virtually no slew-rate limiting.*

the last of which is a low-output-impedance stage. The op amp's transfer characteristic is:

$$V_{OUT} = a(jf)V_D, \tag{1}$$

where V_{OUT} is the output voltage; $V_D = V_P - V_N$ is the differential input voltage; and a(jf), a complex function of frequency (f), is the open-loop gain **(Fig 1a).** Connecting an external resistor network as shown in **Fig 1b** creates a feedback path; the voltage signal derived from the output is applied to the noninverting input. You can solve for V_D to obtain

$$V_D = V_{IN} - \frac{R_1}{R_1 + R_2}V_{OUT}. \tag{2}$$

By substituting **Eq 2** for V_D in **Eq 1** and solving for the ratio V_{OUT}/V_{IN}, you obtain the familiar closed-loop transfer characteristic for a noninverting amplifier:

$$A(jf) = \frac{V_{OUT}}{V_{IN}} = \left(1 + \frac{R_2}{R_1}\right)\frac{1}{1 + 1/T(jf)}, \tag{3}$$

where $1 + R_2/R_1$ is the ideal gain value, and

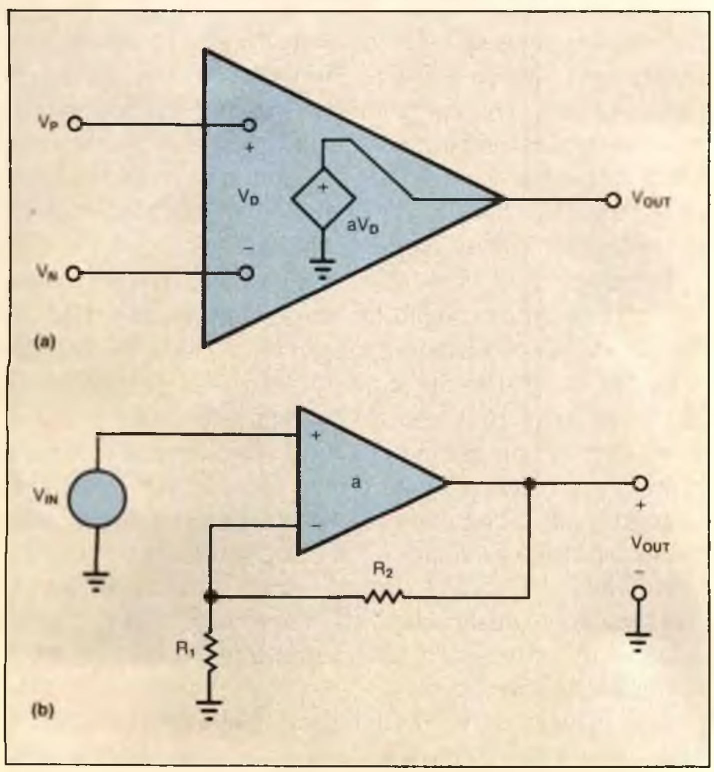

*Fig 1—**The circuit model** of a conventional op amp includes a differential input stage and a gain stage **(a)**; resistive feedback configures the op amp as a noninverting amplifier **(b)**.*

$$T(jf) = \frac{a(jf)}{1 + R_2/R_1} \tag{4}$$

represents the loop gain. The denominator of the loop-gain expression is called the noise gain. In this case, it's equal to $1 + R_2/R_1$. Note that in this example, the noise gain just happens to be equal to the ideal closed-loop gain. It's important not to confuse the two.

Loop gain determines stability

Eq 4 represents the loop gain, because if you break the loop as shown in **Fig 2a** and inject a test signal (V_X) with V_{IN} suppressed, the circuit will first attenuate V_X to produce $V_N = V_X/(1 + R_2/R_1)$, and then amplify V_N to produce $V_{OUT} = -aV_N$. Hence, the gain that a signal experiences when it goes around the loop is $V_{OUT}/V_X = -a/(1 + R_2/R_1)$. The negative of this ratio represents the loop gain, T.

The loop gain provides a measure of how close A is

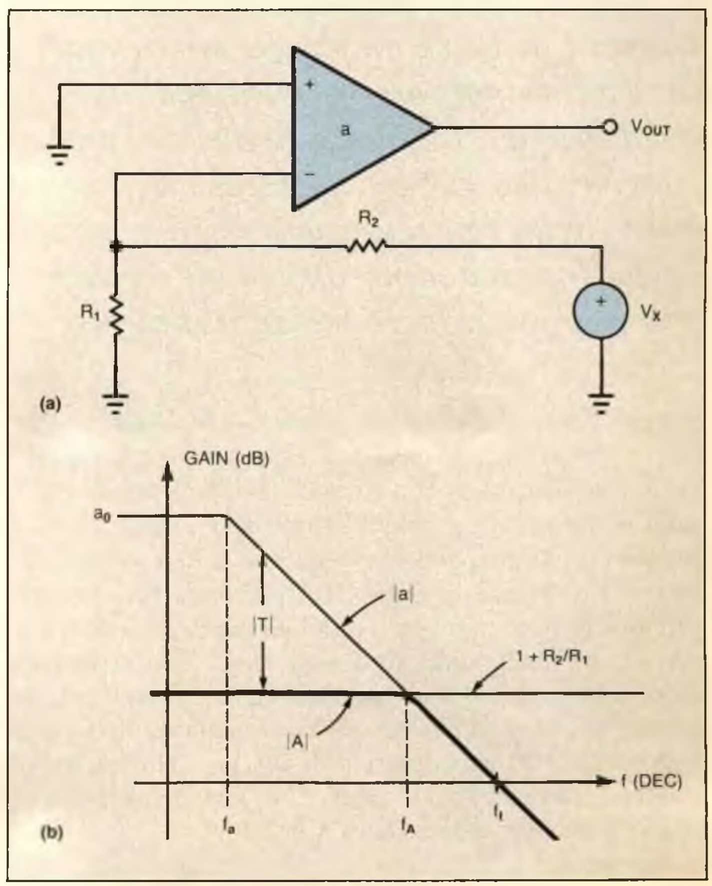

*Fig 2—**You can find the loop gain** by injecting a signal V_X with V_{IN} grounded and solving for V_{OUT}/V_X **(a)**; the loop gain is, graphically, the difference between the open-loop curve and the noise-gain curve.*

to the ideal value of $1+R_2/R_1$. The larger the value of T, the better. To help the user achieve high loop gains over a wide range of closed-loop gains, op-amp manufacturers strive to make the open-loop gain (a) as large as possible. Consequently, V_D will assume extremely small values, because $V_D = V_{OUT}/a$ (see **Eq 1**). As the value of the open-loop gain approaches infinity, V_D approaches zero; that is, the value of V_N approaches that of V_P. This fact is the basis of the familiar op-amp rule: When it's operated with negative feedback, an op amp will ideally provide whatever output voltage and current are needed to force V_N to equal V_P.

Op amps require a familiar tradeoff

In practice, op amps can physically realize large open-loop gains only over a limited frequency range. Beyond this range, the gain rolls off with respect to frequency because of the op amps' internal frequency compensation. Most op amps are designed for a constant rolloff of −20 dB/decade, so the open-loop response can be expressed as

$$a(jf) = \frac{a_0}{1 + j(f/f_a)}, \tag{5}$$

where a_0 represents the dc gain and f_a is the −3-dB frequency of the open-loop response.

By substituting **Eq 5** for a(jf) in **Eq 4** and then substituting **Eq 4** for T(jf) in **Eq 3**, and recognizing the fact that $(1+R_2/R_1)/a_0<1$, you can obtain

$$A(jf) = \frac{1 + R_2/R_1}{1 + j(f/f_A)}, \tag{6}$$

where

$$f_A = \frac{f_t}{1 + R_2/R_1} \tag{7}$$

represents the closed-loop bandwidth and $f_t = a_0 f_a$ represents the open-loop unity-gain frequency—that is, the frequency at which $|a|$ is equal to 1. For instance, the 741 op amp has an f_t equal to 1 MHz.

Eq 7 reveals the familiar gain-bandwidth tradeoff. As you raise the R_2/R_1 ratio to increase the closed-loop gain, you decrease its bandwidth. Moreover, the loop gain also decreases, leading to a greater closed-loop-gain error.

You can see this tradeoff by plotting the frequency response on a graph. From **Eq 4**, $|T|_{dB} = |a|_{dB} - (1+R_2/R_1)_{dB}$. Thus, you can think of the loop gain as the difference between the open-loop gain and the noise gain (**Fig 2b**). The intersection of the two curves is the crossover frequency or −3-dB point, at which T has a magnitude of 1 and a phase shift of −90°.

As you increase the closed-loop gain, the noise-gain curve shifts upward, thus reducing the loop gain. Also, the intersection point will move up the $|a|$ curve, thus decreasing the closed-loop bandwidth. Clearly, the circuit with the widest bandwidth and the highest loop gain is also the one with the lowest closed-loop gain. This circuit is the voltage follower, for which $R_2/R_1 = 0$, so $A = 1$ and $f_A = f_t$.

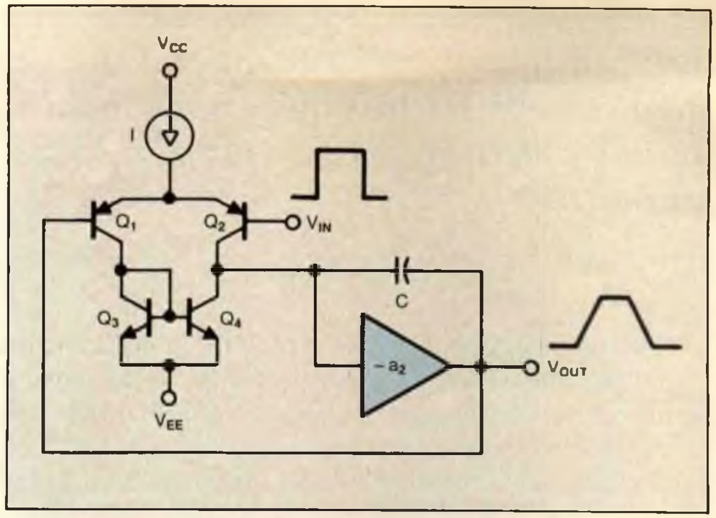

Fig 3—As shown in this simplified slew-rate model, there is limited current to charge and discharge C when the transconductance stage saturates.

Slew-rate limiting is also a factor

To fully characterize the dynamic behavior of an op amp, you also need to know its transient response. In many applications, the dynamic parameter of greatest concern is the settling time, a characteristic in which slew-rate limiting plays an important role. If you apply a small voltage step to an op amp connected as a unity-gain voltage-follower, the amp's dynamic behavior will be similar to that of an RC network. The input step, ΔV_{IN}, will cause the output to undergo an exponential transition with a magnitude of $\Delta V_O = \Delta V_{IN}$ and a time constant of $\tau = 1/(2\pi f_t)$. For the 741 op amp, $\tau = 1/(2\pi \times 10^6) \approx 170$ nsec.

The rate at which the output changes with time is highest at the beginning of the exponential transition, when its value is $\Delta V_{OUT}/\tau$. Increasing the step magnitude increases this initial rate of change, until the latter saturates at a value called the slew rate (SR). This fact is due to the limited ability of the internal circuitry to charge and discharge the compensation capacitor as well as capacitive loads.

The input stage of a typical op amp is a transconductance block consisting of differential pair Q_1-Q_2 and current mirror Q_3-Q_4 (**Fig 3**). The remaining stages, considered together, comprise an integrator block consisting of an inverting amplifier and the com-

Before designing with current-feedback op amps, you should thoroughly understand their architecture; some circuits require special attention.

pensation capacitor, C. Slew-rate limiting occurs when the transconductance stage saturates, so all the current available to charge and/or discharge C is the bias current (I) of this stage.

For example, for the 741 op amp, $I=20\ \mu A$ and $C=30\ pF$, so $SR=I/C=0.67\ V/\mu sec$. The step magnitude corresponding to the onset of slew-rate limiting is such that $\Delta V_{IN}/\tau=SR$; that is, $\Delta V_{IN}=SR\times\tau=(0.67\ V/\mu sec)\times(170\ nsec)=116\ mV$. As long as the step is less than 116 mV, a 741 op amp configured as a voltage follower will respond with an exponential transition governed by $\tau\approx 170$ nsec, whereas for a greater input step the output will slew at a constant rate of 0.67 V/µsec.

Current-feedback-amp architecture

The architecture of the current-feedback amp differs from the conventional op amp in two respects (**Fig 4**). First, the current-feedback amp's input stage is a unity-gain voltage buffer connected across the inputs of the op amp. Its function is to force V_N to follow V_P, very much as negative feedback forces V_N to follow V_P in a conventional op amp. However, because of the low output impedance of this buffer, current can easily flow in or out of the inverting input. During normal operation, this current is extremely small.

Second, a current-feedback amp has a transimpedance amplifier, which senses the current delivered by the buffer to the external feedback network and produces an output voltage V_{OUT} such that

$$V_{OUT}=z(jf)I_N, \quad (8)$$

where z(jf) represents the transimpedance gain of the amplifier and I_N represents the output current of the inverting input.

To fully appreciate the inner workings of the current-feedback amp, you need to examine the simplified cir-

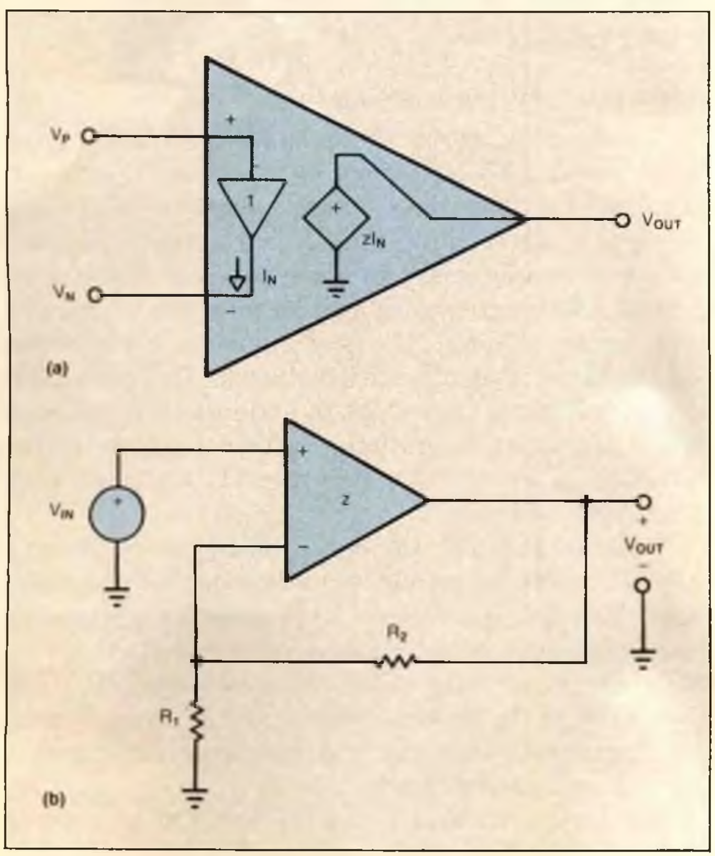

Fig 4—The circuit model of a current-feedback amp *includes a unity-voltage-gain input buffer and a transimpendance block* ***(a)****. Connected as a noninverting amplifier* ***(b)****, the current-feedback amp looks identical to its conventional-op-amp counterpart.*

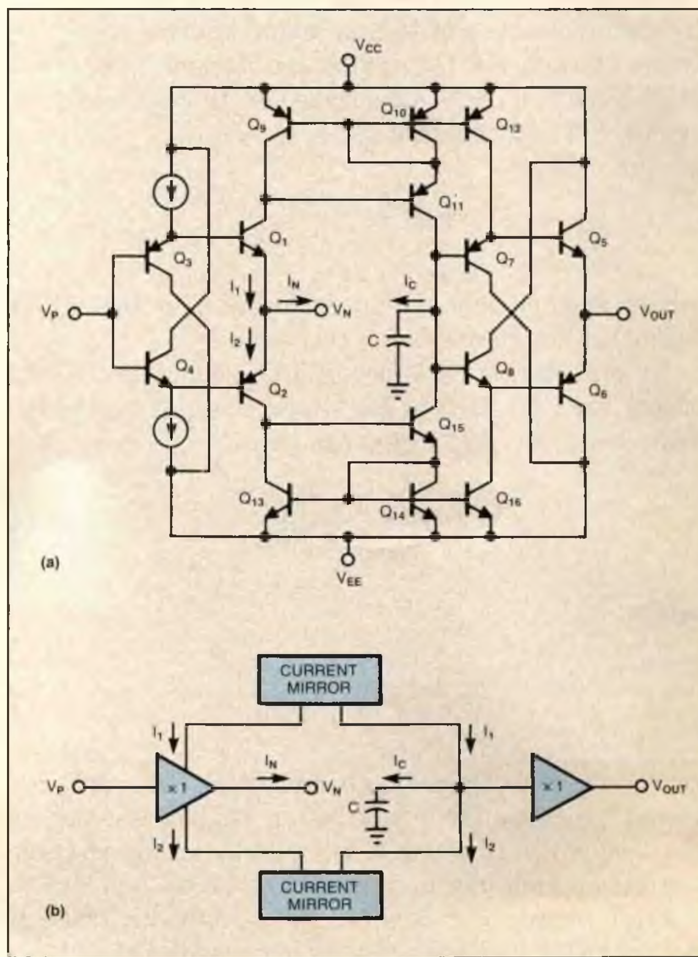

Fig 5—When you look into the actual circuit *inside a current-feedback amp, you'll find both push-pull and Darlington transistor configurations at the input* ***(a)*** *(diagram courtesy Comlinear Corp). The block diagram of the circuit* ***(b)*** *shows the current-feedback amp's basic features.*

cuit diagram of **Fig 5a**. The input buffer consists of transistors Q_1 through Q_4. Q_1 and Q_2 form a low output-impedance push-pull stage. Q_3 and Q_4 provide V_{BE} compensation for the push-pull pair and have a Darlington function, which raises the input impedance.

Summing the currents at the inverting node yields $I_1 - I_2 = I_N$, where I_1 and I_2 are the push-pull transistor currents. Two Wilson current mirrors, consisting of transistors Q_9 through Q_{11} and Q_{13} through Q_{15}, reflect these currents and recombine them at a common node, whose equivalent capacitance to ground is designated "C" in **Fig 5**.

A closer look at the internal circuit

By mirror action, the current through this capacitance is $I_C = I_1 - I_2$; that is, $I_C = I_N$. The voltage developed by C in response to this current is then conveyed to the output by a second buffer, which consists of Q_5 through Q_8. **Fig 5b**'s block diagram summarizes the salient features of the current-feedback amp.

When the amplifier loop is closed, as in **Fig 4b**, and an external signal attempts to imbalance the two inputs, the input buffer will begin sourcing (or sinking) an imbalance current, I_N, to the external feedback network. The Wilson mirrors convey this imbalance to C, causing V_{OUT} to swing in the positive (or negative) direction until the imbalance is neutralized via the negative feedback loop. Thus, I_N plays the role of error signal in the system.

To obtain the closed-loop transfer characteristic, refer again to **Fig 4b**. By summing the currents at the inverting node, you obtain

$$I_N = \frac{V_N}{R_1} - \frac{V_{OUT} - V_N}{R_2}. \quad (9)$$

Because the buffer ensures that $V_N = V_P = V_{IN}$, you can rewrite **Eq 9** as:

$$I_N = \frac{V_{IN}}{R_1 \| R_2} - \frac{V_{OUT}}{R_2}, \quad (10)$$

which confirms that the feedback signal, V_{OUT}/R_2, is now in the form of a current. By substituting **Eq 10** for I_N in **Eq** 8, and solving for the ratio V_{OUT}/V_{IN}, you obtain

$$A(jf) = \frac{V_{OUT}}{V_{IN}} = \left(1 + \frac{R_2}{R_1}\right)\frac{1}{1 + 1/T(jf)}, \quad (11)$$

where A(jf) represents the closed-loop gain of the circuit, and

$$T(jf) = \frac{z(jf)}{R_2} \quad (12)$$

represents the loop gain. As for a conventional op amp, this terminology is derived from the fact that if you break the loop as shown in **Fig 6a**, and inject a test voltage (V_X) with V_{IN} suppressed, the circuit will first convert V_X to $I_N = -V_X/R_2$ and then convert I_N to $V_{OUT} = zI_N$, so $T = z/R_2$, as expected.

To ensure that the circuit will have substantial loop gain, and, therefore, minimal closed-loop gain error, manufacturers strive to make z as large as possible in relation to the expected values of R_2. Consequently, because $I_N = V_{OUT}/z$, the inverting-input current will be very small, even though this input is a low-impedance node because of the buffer. As a current-feedback amp's open-loop gain (z) approaches infinity, its I_N approaches 0, so the amplifier will provide whatever output voltage and current are needed to drive I_N to zero. Thus, the conventional op-amp conditions,

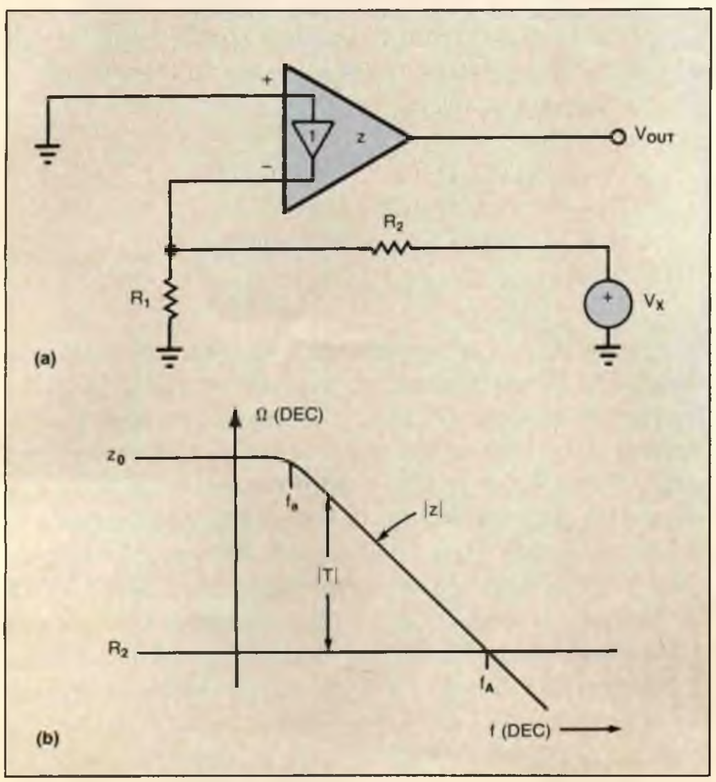

Fig 6—As with conventional op amps, you can use this test circuit (a) to determine the loop gain. The loop gain is represented graphically as the difference between the open loop gain, |z|, and the noise-gain curve, R_2 (b).

The magnitude of an op amp's loop gain determines the closed-loop-gain error, and its phase determines stability.

$V_N = V_P$ and $I_N = I_P = 0$, hold for current-feedback amps as well.

No gain-bandwidth tradeoff

The transimpedance gain of a practical current-feedback amp rolls off with frequency according to

$$z(jf) = \frac{z_0}{1 + j(f/f_a)}, \qquad (13)$$

where z_0 is the dc value of the transimpedance gain and f_a is the frequency at which rolloff begins. For instance, the data sheets of Comlinear's CLC401 current-feedback amp state that $z_0 \approx 710$ kΩ and $f_a \approx 350$ kHz.

By substituting **Eq 13** for z(jf) in **Eq 12**, and then substituting **Eq 12** for T(jf) in **Eq 11**, and recognizing the fact that $R_2/z_0 < 1$, you obtain

$$A(jf) = \frac{1 + R_2/R_1}{1 + j(f/f_A)}, \qquad (14)$$

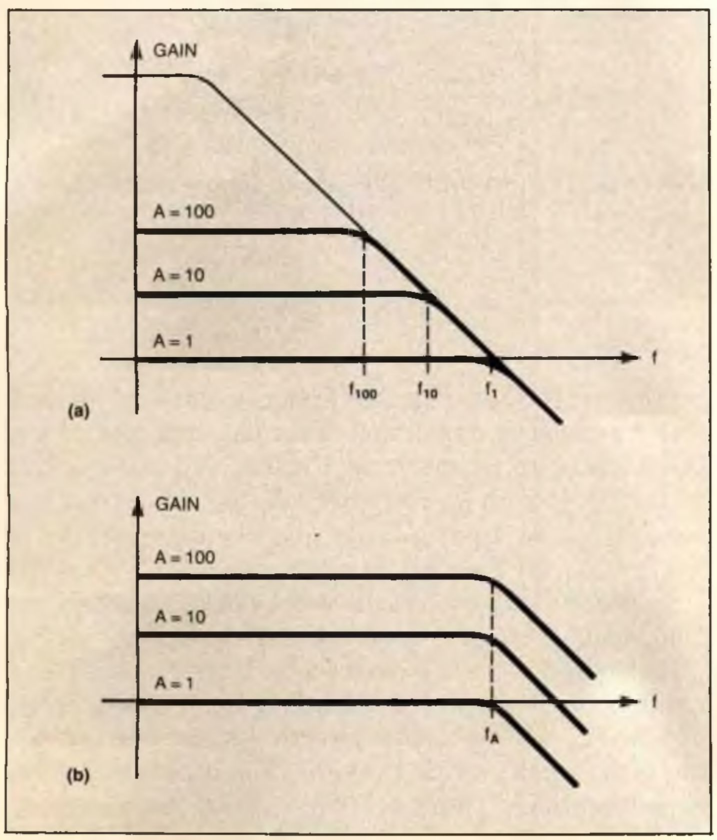

*Fig 7—**The most significant advantage** that current-feedback amps have over conventional op amps can be seen in this simple frequency-response plot. Note the gain-bandwidth tradeoff for conventional op amps in **a** and the absence of such a compromise in **b**.*

where

$$f_A = \frac{z_0 f_a}{R_2} \qquad (15)$$

represents the closed-loop bandwidth. When R_2 is in the kilohm range, f_A is typically in the 100-MHz range. The noise-gain curve is now simply R_2, and f_A can be represented graphically as the frequency at which the R_2 curve meets the $|z|$ curve **(Fig 6b)**.

These closed-loop-gain expressions are formally identical to those for the conventional op amp **Eqs 6** and **7**. However, the bandwidth now depends only on R_2 rather than on the closed-loop gain $1 + R_2/R_1$. Consequently, you can use R_2 to select the bandwidth and R_1 to select the gain. **Fig 7** highlights these frequency-response differences between current-feedback amps and conventional op amps.

The other major advantage of current-feedback amps is their inherent absence of slew-rate limiting. This feature is due to the fact that the current available to charge the internal capacitance at the onset of a step is now proportional to the step, regardless of its size. Indeed, applying the step ΔV_{IN} induces, according to **Eq 10**, an initial current imbalance $I_N = \Delta V_{IN}/(R_1 \| R_2)$, which the Wilson mirrors then convey to the capacitor. The initial rate of charge is, therefore,

$$\begin{aligned} I_C/C &= I_N/C \\ &= \Delta V_{IN}/((R_1 \| R_2)C) \\ &= (\Delta V_{IN}(1 + R_2/R_1))/(R_2C) \\ &= \Delta V_{OUT}/(R_2C), \end{aligned}$$

which indicates an exponential output transition in which the time constant, τ, is equal to R_2C. Like the frequency response, then, the transient response is governed by R_2 alone, regardless of the closed-loop gain. When R_2 is in the kilohm range and C is in the picofarad range, τ will be in the nanosecond range.

The rise time is defined as the amount of time, t_r, that it takes for the output to swing from 10% to 90% of the step size. For an exponential transition, $t_r = \tau \times \ln(0.9/0.1) = 2.2\tau$. For example, the CLC401 has a t_r equal to 2.5 nsec for a 2V output step, indicating an effective τ of 1.14 nsec. The time it takes for the output to settle to within 0.1% of the final value is $t_s = \tau \times \ln 1000$, which is approximately 7τ. For the CLC401, therefore, t_s is approximately 8 nsec, which is in reasonable agreement with the data-sheet value of 10 nsec.

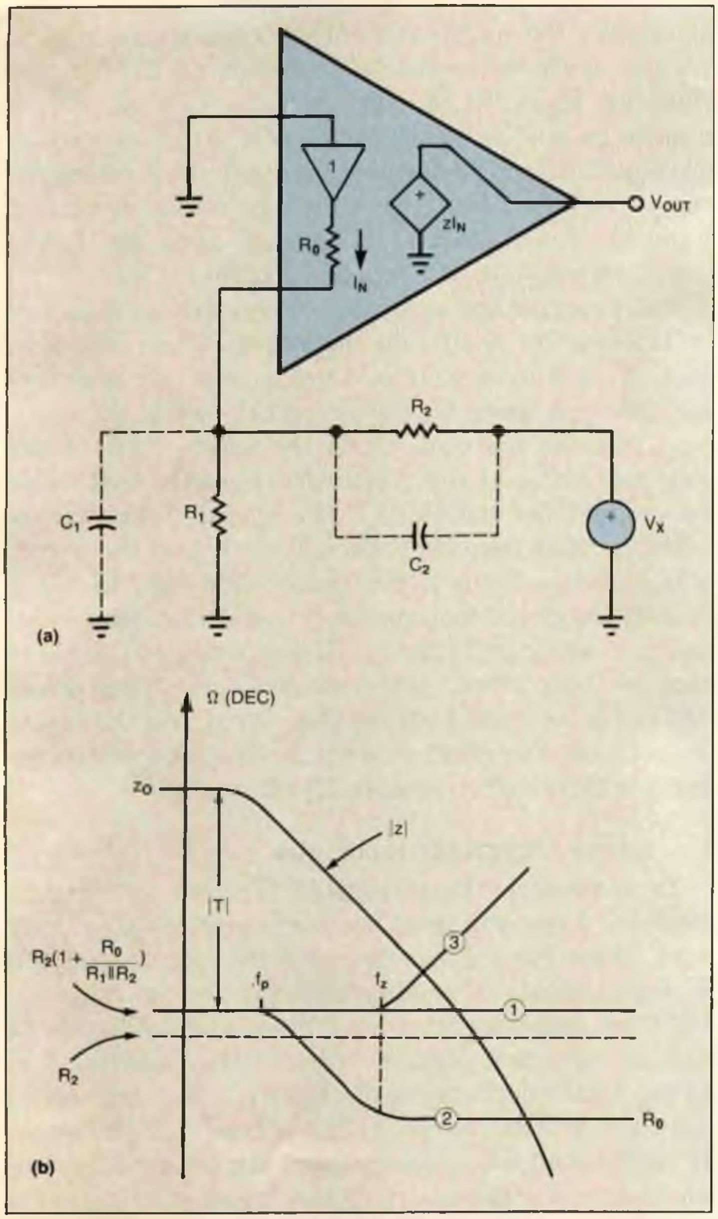

Fig 8—By using the more real-world circuit model in a, you can determine the effects, shown in b, of R_O (curve 1), feedback capacitance (curve 2), and input capacitance (curve 3).

So far, this analysis indicates that once R_2 has been set, the dynamics of the amplifier are unaffected by the closed-loop-gain setting. In practice, you'll find that a current-feedback amp's bandwidth and rise time do vary somewhat with gain, though not as drastically as do those of conventional op amps. The main cause of this nonideal behavior is the input buffer's nonzero output impedance (R_O), which alters the loop gain and, hence, the closed-loop dynamics.

As **Fig 8a** shows, the circuit first converts V_X to a current, $I_{R2} = V_X/(R_2 + R_1 \tau R_O)$, and then divides I_{R2} to produce $I_N = -I_{R2}R_1/(R_1 + R_O)$. Finally, it converts I_N to the voltage $V_{OUT} = V_N$. Eliminating I_{R2} and I_N and letting T equal $-V_{OUT}/V_X$ yields $T = z/Z_2$, where

$$Z_2 = R_2\left(1 + \frac{R_0}{R_1\|R_2}\right). \qquad \textbf{(16)}$$

Thus, the effect of R_O is to increase the noise gain from R_2 to $R_2(1 + R_O/(R_1 \tau R_2))$ (**Fig 8b**, curve 1.) Consequently, both the bandwidth and the rise time will be reduced by a proportional amount.

You can replace R_2 in **Eq** 15 with Z_2 from **Eq** 16, and, after simple manipulation, obtain

$$f_A = \frac{f_t}{1 + \frac{R_0}{R_2}\left(1 + \frac{R_2}{R_1}\right)}, \qquad \textbf{(17)}$$

where $f_t = z_O f_a/R_2$ represents the extrapolated value of f_A in the limit $R_O \rightarrow 0$. **Eq 17** indicates that the bandwidth reduction caused by R_0 will be more pronounced at high closed-loop gains. For example, suppose a current-feedback amp has $R_0 = 50\Omega$, $R_2 = 1.5$ kΩ, and $f_t = 100$ MHz, so $f_A = 10^8/(1 + (50/1500)A_O) = 10^8/(1 + A_O/30)$, where $A_O = 1 + R_2/R_1$. Then, the bandwidths corresponding to $A_O = 1$, $A_O = 10$, and $A_O = 100$ are, respectively, 96.8 MHz, 75.0 MHz, and 23.1 MHz. Note that these values still compare favorably with those of a conventional op amp, whose bandwidth would be reduced, respectively, by 1, 10, and 100.

If you wish, you can predistort the external resistance values to compensate for the bandwidth reduction at high gains. By solving for R_2 in **Eq 17**, you can obtain the required value of R_2 for a given bandwidth (f_A) and gain (A_O), which is

$$R_2 = \frac{z_0 f_a}{f_A} - R_0A_0, \qquad \textbf{(18)}$$

while the required value of R_1 for gain A_O is

$$R_1 = \frac{R_2}{A_0 - 1}. \qquad \textbf{(19)}$$

For example, suppose you want the above amplifier to retain its 100-MHz bandwidth at a closed-loop gain of 10. When $R_2 = 1.5$ kΩ, this device has a $z_O f_a/R_2$ equal to 100 MHz, so it follows that $z_O f_a = 10^8 \times 1500 = 1.5 \times 10^{11}\Omega \times$ Hz. Then, **Eqs** 18 and

In current-feedback amps, you can use one of the feedback resistors to set the gain, and the other to set the closed-loop bandwidth.

19 yield $R_2=(1.5\times 10^{11}/10^8)-(50\times 10)=1\ k\Omega$, and $R_1=1000/(10-1)=111\Omega$, respectively.

Current-feedback amps have higher-order poles

In addition to the dominant pole at f_a, the open-loop response of a practical current-feedback amp also has poles above the crossover frequency. As **Fig 8b** shows, the effect of these poles is to cause a steeper gain rolloff at higher frequencies, further reducing the closed-loop bandwidth. Moreover, the additional phase shift caused by these poles decreases the phase margin somewhat, thus causing a small amount of peaking in the frequency response and creating some ringing in the transient response.

Like the real current-feedback-amp bandwidth characteristics, the transient response also strays from the ideal. The rise time of a practical current-feedback amp increases somewhat with the step size, primarily because of the transistor's current-gain degradation at high current levels. For instance, the rise time of the CLC401 changes from 2.5 to 5 nsec as the step size changes from 2 to 5V. Despite their second-order limitations, current-feedback amps provide dynamics superior to those of conventional op amps.

Consider other feedback configurations

This discussion has focused so far on the noninverting configuration, but you can use current-feedback amps in most other resistive feedback configurations, such as the inverting amplifier, the summing and difference amplifier, current-to-voltage and voltage-to-current converters, and KRC active filters (**Ref 3**).

You should take special care, however, with circuits in which the feedback network includes reactive elements, whether they're intentional or parasitic. Consider first the effect of feedback capacitance (C_2) in parallel with R_2 in the basic circuit of **Fig 8a**. By replacing R_2 in **Eq 16** with $Z=R_2\parallel(1/sC_2)$, you obtain a noise gain of $Z_2=Z(1+R_0/(R_1\parallel Z))$. After expanding the equation (and performing some algebraic manipulation), you'll find that the noise-gain curve now has a pole at $f_p=1/(2\pi R_2C_2)$, and a zero at $f_z=1/(2\pi(R_0\parallel R_1\parallel R_2)C_2)$, as curve 2 of **Fig 8b** shows.

This new pole and zero move the crossover frequency or intersection point into the region where the loop gain, T, will have increased negative phase shift (remember that there are higher-order poles in the open-loop transfer function). It is the phase shift of the loop-gain curve at the crossover frequency that determines amplifier stability. If the overall shift reaches $-180°$ at that frequency, then $T=-1$, and the circuit will oscillate. Even if the phase shift fails to reach $-180°$, the closed-loop response may still exhibit intolerable peaking and ringing. Hence, when you use current-feedback amps, you must avoid applying direct capacitive feedback between the output and the input. To minimize the effect of stray feedback capacitances, manufacturers often provide R_2 internally.

Use unique integrator topologies

To synthesize the integrator function in current-feedback form, you must use configurations that don't have direct capacitance between the output and the inverting input. (The integrator function provides the basis for dual-integrator-loop filters and oscillators as well as for other popular circuits.) One possibility is to use the Deboo integrator (**Ref 3**), which belongs to the class of KRC filters. It has a drawback, however: If you desire lossless integration, you must make sure the circuit resistances are tightly matched.

The alternative circuit shown in **Fig 9** provides indirect feedback and also features active compensation, a highly desirable feature for coping with Q-enhancement problems in dual-integrator-loop filters (**Ref 3**). By using standard op-amp-analysis techniques, you can see that the unity-gain frequency of this integrator is $f_0=(R_2/R_1)/(2\pi RC)$. The availability of current-feedback amps in dual monolithic packages, such as the OP-260 from Precision Monolithics, makes this circuit cost-effective.

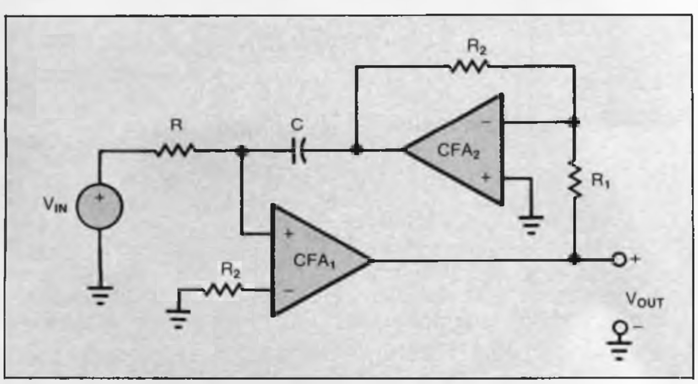

***Fig 9—To implement an integrator,** you must use circuit configurations that involve indirect feedback, such as this actively compensated current-feedback integrator.*

Compensate for stray input capacitance

Next, consider the effect of input capacitance (C_1) in parallel with R_1 in the basic circuit of **Fig 8a**. By replacing R_1 in **Eq 17** with Z, and letting $Z=R_1\parallel(1/$

Beware of the instability problems that direct capacitive feedback and stray input capacitance can cause.

sC_1), you obtain a noise gain of $Z_2 = R_2(1 + R0/(Z\tau R_2))$. After expanding the equation and performing more algebraic manipulation, you find that, as curve 3 of **Fig 8b** shows, the noise-gain curve now has a zero at

$$f_z = 1/(2\pi(R_0 \tau R_1 \tau R_2)C_1).$$

Again, recall that T is equal to $|Z|$ in decibels *minus* the noise gain in decibels. Likewise, the phase of T is equal to the phase of Z minus the phase of the noise-gain curve. So, the positive phase shift contributed by the new zero in the noise-gain curve looks like negative phase shift to T. If C_1 is sufficiently large, the phase of T at the crossover frequency will again approach $-180°$, placing the circuit on the verge of instability. This fact is of particular concern in current-mode-DAC output buffering, where C_1 is the output capacitance of the DAC, typically in the range of a few tens to a few hundreds of picofarads, depending on the DAC type.

As with a conventional op amp, you can stabilize the current-feedback amp by using feedback capacitance (C_2) to introduce sufficient negative phase shift in the noise-gain curve (positive phase shift to T), thus compensating for the effect of the input capacitance (C_1). For a phase margin of 45°, choose the value of C_2 so that the noise-gain pole, $f_p = 1/(2R_2C_2)$, coincides with the crossover frequency, f_A **(Fig 10a)**. Using linearized Bode-plot reasoning **(Ref 3)**, also known as straight-line approximation, you find that:

$$f_A = (z_0 f_a f_z/(R_0 + R_1))^1,$$

where $f_z = 1/(2\pi(R_0 \tau R_2)C_1)$. Setting $f_p = f_A$ yields

$$C_2 = \left[\frac{R_0}{2\pi R_2 z_0 f_a} C_1\right]^{1/2} \quad (20)$$

To cope with impractically low values of C_2, it's convenient to drive C_2 with a voltage divider as in **Fig 10b**; this action will scale the value of C_2 to a more practical value:

$$C_c = \left(1 + \frac{R_B}{R_A}\right)C_2. \quad (21)$$

(Note that this circuit configuration will provide an additional zero in the noise-gain curve that lies to the right of the compensation pole, f_A, in **(Fig 10a)**.

For this technique to be effective, R_B must be much

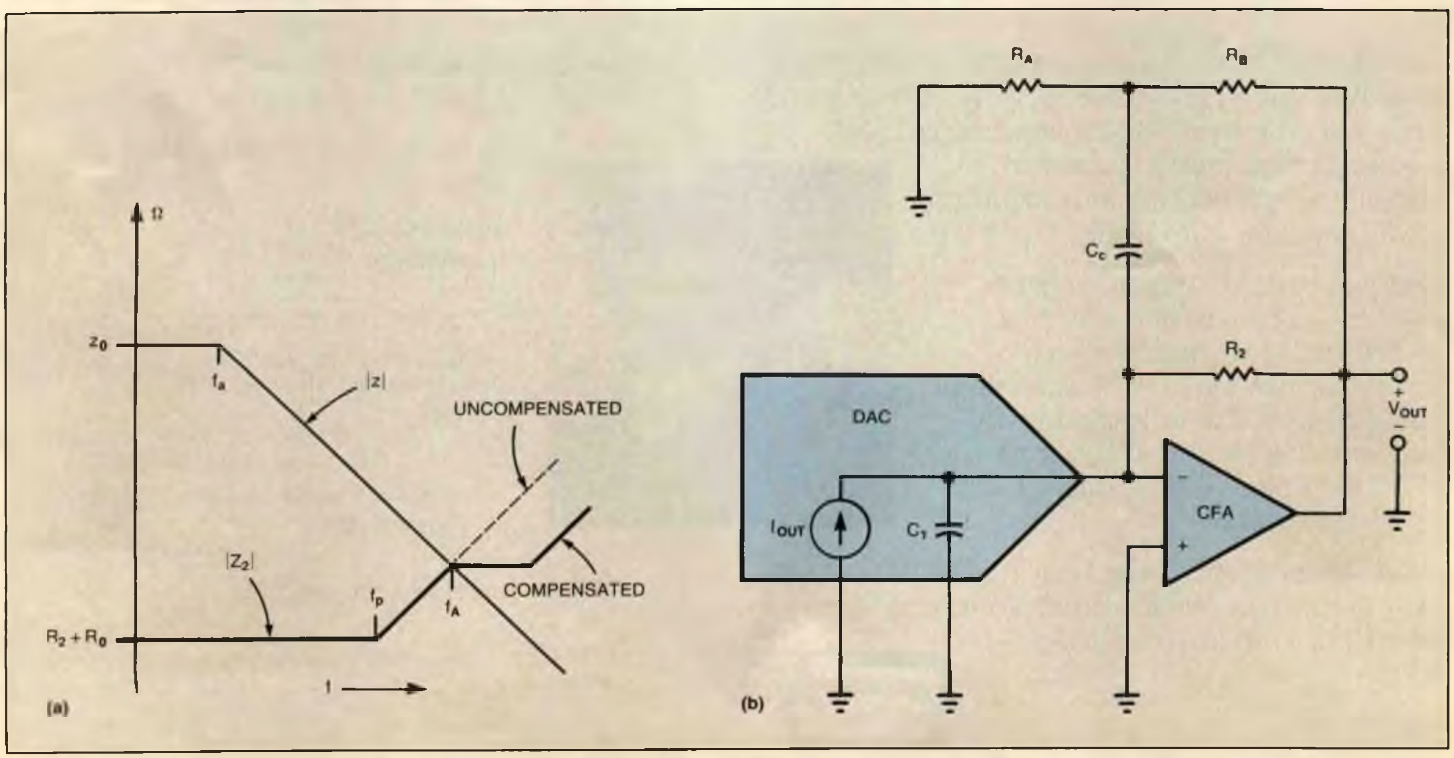

***Fig 10—To compensate for input capacitance,** you should add a pole at f_A (**a**) that will add positive phase shift to the loop gain, thereby stabilizing the circuit. Use the circuit in **b** to achieve practical compensation-capacitor values.*

less than R_2. For example, suppose that a DAC in which C_1 equals 100 pF feeds the current-feedback amp considered earlier. **Eq 20** yields:

$$C_2 = (50\times 100\times 10^{-12} \div (2\pi\times 1.5\times 10^3\times 1.5\times 10^{11}))^{1/2} = 1.88\text{ pF}.$$

To scale C_2 to a more practical value, you can use $R_A = 50\Omega$ and $R_B = 500\Omega$ (**Ref 4**). **Eq 20** then yields $C_c = (1 + 500/50)\times 1.88\text{ pF} = 21\text{ pF}$. You may need to fine-tune this estimate to optimize the transient response. EDN

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Sergio Franco is a professor of electrical engineering at San Francisco State University, where he teaches microelectronics courses and acts as an industry consultant. He has taught at the university for the past eight years. Previously, he was employed at Zeltron, Zanussi's Electronics Institute (Udine, Italy). Sergio received a BS in physics from the University of Rome (Italy), an MS in physics from Clark University (Worcester, MA), and a PhD in computer science from the University of Illinois at Urbana. He is a member of the IEEE. In his spare time, Sergio enjoys classical music, gardening, and mountain hiking.

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High 488 Medium 489 Low 490