Chapter 8

Current-Feedback Op Amp Analysis

Ron Mancini

8.1 Introduction

Current-feedback amplifiers (CFA) do not have the traditional differential amplifier input structure, thus they sacrifice the parameter matching inherent to that structure. The CFA circuit configuration prevents them from obtaining the precision of voltage-feedback amplifiers (VFA), but the circuit configuration that sacrifices precision results in increased bandwidth and slew rate. The higher bandwidth is relatively independent of closed-loop gain, so the constant gain-bandwidth restriction applied to VFAs is removed for CFAs. The slew rate of CFAs is much improved from their counterpart VFAs because their structure enables the output stage to supply slewing current until the output reaches its final value. In general, VFAs are used for precision and general purpose applications, while CFAs are restricted to high frequency applications above 100 MHz.

Although CFAs do not have the precision of their VFA counterparts, they are precise enough to be dc-coupled in video applications where dynamic range requirements are not severe. CFAs, unlike previous generation high-frequency amplifiers, have eliminated the ac coupling requirement; they are usually dc-coupled while they operate in the GHz range. CFAs have much faster slew rates than VFAs, so they have faster rise/fall times and less intermodulation distortion.

8.2 CFA Model

The CFA model is shown in Figure 8–1. The noninverting input of a CFA connects to the input of the input buffer, so it has very high impedance similar to that of a bipolar transistor noninverting VFA input. The inverting input connects to the input buffer's output, so the inverting input impedance is equivalent to a buffer's output impedance, which is very low. Z_B models the input buffer's output impedance, and it is usually less than 50 Ω. The input buffer gain, G_B, is as close to one as IC design methods can achieve, and it is small enough to neglect in the calculations.

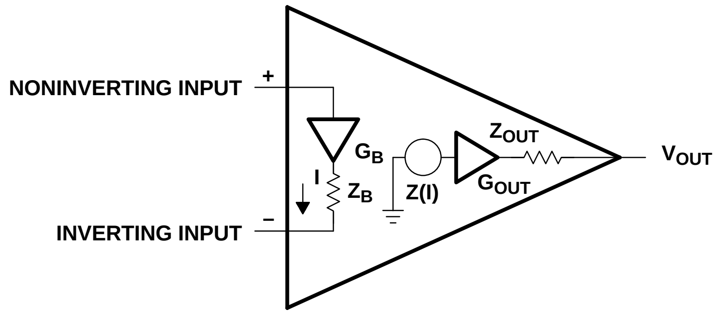


Figure 8–1. Current-Feedback Amplifier Model

The output buffer provides low output impedance for the amplifier. Again, the output buffer gain, G_{OUT}, is very close to one, so it is neglected in the analysis. The output impedance of the output buffer is ignored during the calculations. This parameter may influence the circuit performance when driving very low impedance or capacitive loads, but this is usually not the case. The input buffer's output impedance can't be ignored because affects stability at high frequencies.

The current-controlled current source, Z, is a transimpedance. The transimpedance in a CFA serves the same function as gain in a VFA; it is the parameter that makes the performance of the op amp dependent only on the passive parameter values. Usually the transimpedance is very high, in the MΩ range, so the CFA gains accuracy by closing a feedback loop in the same manner that the VFA does.

8.3 Development of the Stability Equation

The stability equation is developed with the aid of Figure 8–2. Remember, stability is independent of the input, and stability depends solely on the loop gain, Aβ. Breaking the loop at point *X*, inserting a test signal, V_{TI}, and calculating the return signal V_{TO} develops the stability equation.

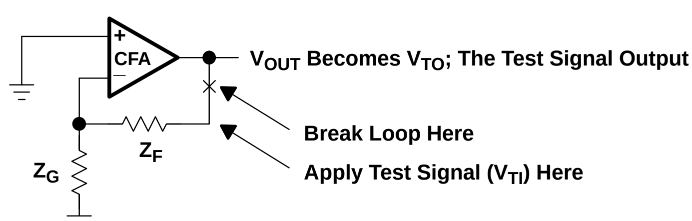


Figure 8–2. Stability Analysis Circuit

The circuit used for stability calculations is shown in Figure 8–3 where the model of Figure 8–1 is substituted for the CFA symbol. The input and output buffer gain, and output buffer output impedance have been deleted from the circuit to simplify calculations. This approximation is valid for almost all applications.

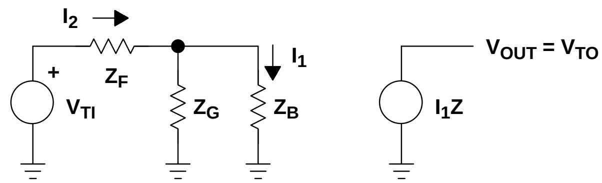


Figure 8–3. Stability Analysis Circuit

The transfer equation is given in Equation 8–1, and the Kirchoff"s law is used to write Equations 8–2 and 8–3.

$$V_{TO} = I_1 Z \tag{8–1}$$

$$V_{TI} = I_2\left(Z_F + Z_G \parallel Z_B\right) \tag{8–2}$$

$$I_2\left(Z_G \parallel Z_B\right) = I_1 Z_B \tag{8–3}$$

Equations 8–2 and 8–3 are combined to yield Equation 8–4.

$$V_{TI} = I_1\left(Z_F + Z_G \parallel Z_B\right)\left(1 + \frac{Z_B}{Z_G}\right) = I_1 Z_F\left(1 + \frac{Z_B}{Z_F \parallel Z_G}\right) \tag{8–4}$$

Dividing Equation 8–1 by Equation 8–4 yields Equation 8–5, and this is the open loop transfer equation. This equation is commonly known as the loop gain.

$$A\beta = \frac{V_{TO}}{V_{TI}} = \frac{Z}{\left(Z_F\left(1 + \frac{Z_B}{Z_F \| Z_G}\right)\right)} \tag{8–5}$$

8.4 The Noninverting CFA

The closed-loop gain equation for the noninverting CFA is developed with the aid of Figure 8–4, where external gain setting resistors have been added to the circuit. The buffers are shown in Figure 8–4, but because their gains equal one and they are included within the feedback loop, the buffer gain does not enter into the calculations.

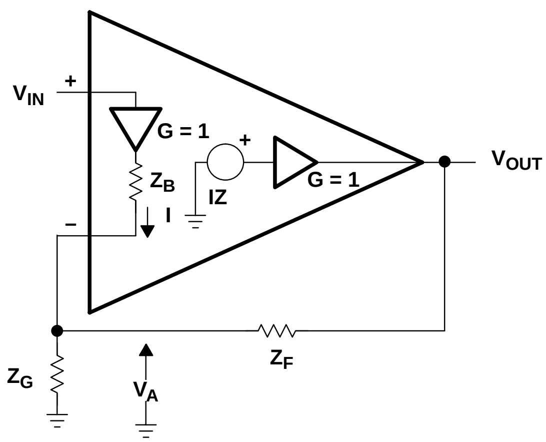


Figure 8–4. Noninverting CFA

Equation 8–6 is the transfer equation, Equation 8–7 is the current equation at the inverting node, and Equation 8–8 is the input loop equation. These equations are combined to yield the closed-loop gain equation, Equation 8–9.

$$V_{OUT} = IZ \tag{8–6}$$

$$I = \left(\frac{V_A}{Z_G}\right) - \left(\frac{V_{OUT} - V_A}{Z_F}\right) \tag{8–7}$$

$$V_A = V_{IN} - IZ_B \tag{8–8}$$

$$\frac{V_{OUT}}{V_{IN}} = \frac{\dfrac{Z\left(1+\dfrac{Z_F}{Z_G}\right)}{Z_F\left(1+\dfrac{Z_B}{Z_F \| Z_G}\right)}}{1+\dfrac{Z}{Z_F\left(1+\dfrac{Z_B}{Z_F \| Z_G}\right)}} \tag{8–9}$$

When the input buffer output impedance, Z_B, approaches zero, Equation 8–9 reduces to Equation 8–10.

$$\frac{V_{OUT}}{V_{IN}} = \frac{\dfrac{Z\left(1+\dfrac{Z_F}{Z_G}\right)}{Z_F}}{1 + \dfrac{Z}{Z_F}} = \frac{1 + \dfrac{Z_F}{Z_G}}{1 + \dfrac{Z_F}{Z}} \tag{8–10}$$

When the transimpedance, Z, is very high, the term Z_F/Z in Equation 8–10 approaches zero, and Equation 8–10 reduces to Equation 8–11; the ideal closed-loop gain equation for the CFA. The ideal closed-loop gain equations for the CFA and VFA are identical, and the degree to which they depart from ideal is dependent on the validity of the assumptions. The VFA has one assumption that the direct gain is very high, while the CFA has two assumptions, that the transimpedance is very high and that the input buffer output impedance is very low. As would be expected, two assumptions are much harder to meet than one, thus the CFA departs from the ideal more than the VFA does.

$$\frac{V_{OUT}}{V_{IN}} = 1 + \frac{Z_F}{Z_G} \tag{8–11}$$

8.5 The Inverting CFA

The inverting CFA configuration is seldom used because the inverting input impedance is very low ($Z_B||Z_F + Z_G$). When Z_G is made dominant by selecting it as a high resistance value it overrides the effect of Z_B. Z_F must also be selected as a high value to achieve at least unity gain, and high values for Z_F result in poor bandwidth performance, as we will see in the next section. If Z_G is selected as a low value the frequency sensitive Z_B causes the gain to increase as frequency increases. These limitations restrict inverting applications of the inverting CFA.

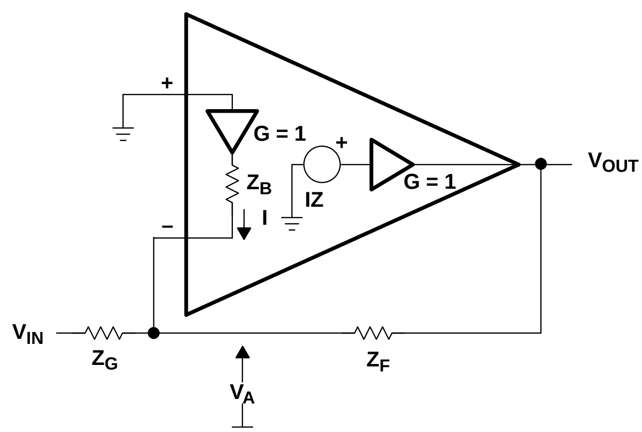


Figure 8–5. Inverting CFA

The current equation for the input node is written as Equation 8–12. Equation 8–13 defines the dummy variable, V_A, and Equation 8–14 is the transfer equation for the CFA. These equations are combined and simplified leading to Equation 8–15, which is the closed-loop gain equation for the inverting CFA.

$$I + \frac{V_{IN} - V_A}{Z_G} = \frac{V_A - V_{OUT}}{Z_F} \tag{8–12}$$

$$IZ_B = -V_A \tag{8–13}$$

$$IZ = V_{OUT} \tag{8–14}$$

$$\frac{V_{OUT}}{V_{IN}} = -\frac{\dfrac{Z}{Z_G\left(1 + \dfrac{Z_B}{Z_F \| Z_G}\right)}}{1 + \dfrac{Z}{Z_F\left(1 + \dfrac{Z_B}{Z_F \| Z_G}\right)}} \tag{8–15}$$

When Z_B approaches zero, Equation 8–15 reduces to Equation 8–16.

$$\frac{V_{OUT}}{V_{IN}} = -\frac{\dfrac{1}{Z_G}}{\dfrac{1}{Z} + \dfrac{1}{Z_F}} \tag{8–16}$$

When Z is very large, Equation 8–16 becomes Equation 8–17, which is the ideal closed-loop gain equation for the inverting CFA.

$$\frac{V_{OUT}}{V_{IN}} = -\frac{Z_F}{Z_G} \tag{8–17}$$

The ideal closed-loop gain equation for the inverting VFA and CFA op amps are identical. Both configurations have lower input impedance than the noninverting configuration has, but the VFA has one assumption while the CFA has two assumptions. Again, as was the case with the noninverting counterparts, the CFA is less ideal than the VFA because of the two assumptions. The zero Z_B assumption always breaks down in bipolar junction transistors as is shown later. The CFA is almost never used in the differential amplifier configuration because of the CFA's gross input impedance mismatch.

8.6 Stability Analysis

The stability equation is repeated as Equation 8–18.

$$A\beta = \frac{V_{TO}}{V_{TI}} = \frac{Z}{\left(Z_F\left(1 + \frac{Z_B}{Z_F \| Z_G}\right)\right)} \tag{8–18}$$

Comparing Equations 8–9 and 8–15 to Equation 8–18 reveals that the inverting and non-inverting CFA op amps have identical stability equations. This is the expected result because stability of any feedback circuit is a function of the loop gain, and the input signals have no affect on stability. The two op amp parameters affecting stability are the transimpedance, Z, and the input buffer's output impedance, Z_B. The external components affecting stability are Z_G and Z_F. The designer controls the external impedance, although stray capacitance that is a part of the external impedance sometimes seems to be uncontrollable. Stray capacitance is the primary cause of ringing and overshoot in CFAs. Z and Z_B are CFA op amp parameters that can't be controlled by the circuit designer, so he has to live with them.

Prior to determining stability with a Bode plot, we take the log of Equation 8–18, and plot the logs (Equations 8–19 and 8–20) in Figure 8–6.

$$20\ \text{LOG}\ |A\beta| = 20\ \text{LOG}\ |Z| - 20\ \text{LOG}\left|Z_F\left(1 + \frac{Z_B}{Z_F \| Z_B}\right)\right| \tag{8–19}$$

$$\phi = \text{TANGENT}^{-1}\,(A\beta) \tag{8–20}$$

This enables the designer to add and subtract components of the stability equation graphically.

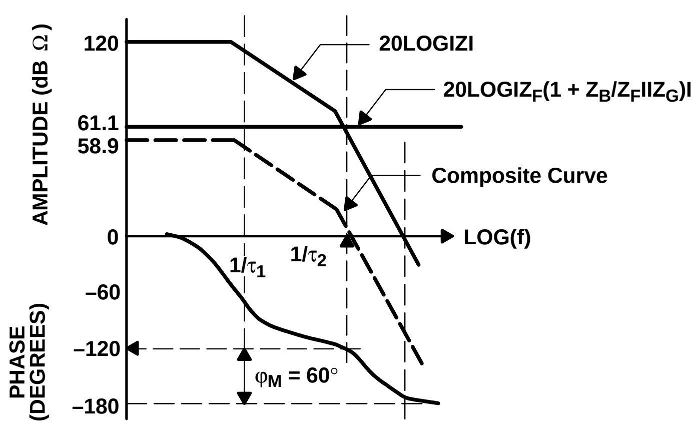


Figure 8–6. Bode Plot of Stability Equation

The plot in Figure 8–6 assumes typical values for the parameters:

$$Z = \frac{1M\Omega}{(1 + \tau_1 S)(1 + \tau_2 S)} \quad (8\text{–}21)$$

$$Z_B = 70\Omega \quad (8\text{–}22)$$

$$Z_G = Z_F = 1k\Omega \quad (8\text{–}23)$$

The transimpedance has two poles and the plot shows that the op amp will be unstable without the addition of external components because 20 LOG|Z| crosses the 0-dB axis after the phase shift is 180°. Z_F, Z_B, and Z_G reduce the loop gain 61.1 dB, so the circuit is stable because it has 60°-phase margin. Z_F is the component that stabilizes the circuit. The parallel combination of Z_F and Z_G contribute little to the phase margin because Z_B is very small, so Z_B and Z_G have little effect on stability.

The manufacturer determines the optimum value of R_F during the characterization of the IC. Referring to Figure 8–6, it is seen that when R_F exceeds the optimum value recommended by the IC manufacturer, stability increases. The increased stability has a price called decreased bandwidth. Conversely, when R_F is less than the optimum value recommended by the IC manufacturer, stability decreases, and the circuit response to step inputs is overshoot or possibly ringing. Sometimes the overshoot associated with less than optimum R_F is tolerated because the bandwidth increases as R_F decreases. The peaked response associated with less than optimum values of R_F can be used to compensate for cable droop caused by cable capacitance.

When $Z_B = 0\ \Omega$ and $Z_F = R_F$ the loop gain equation is; $A\beta = Z/R_F$. Under these conditions Z and R_F determine stability, and a value of R_F can always be found to stabilize the circuit. The transimpedance and feedback resistor have a major impact on stability, and the input buffer's output impedance has a minor effect on stability. Since Z_B increases with an increase in frequency, it tends to increase stability at higher frequencies. Equation 8–18 is rewritten as Equation 8–24, but it has been manipulated so that the ideal closed-loop gain is readily apparent.

$$A\beta = \frac{Z}{Z_F + Z_B\left(1 + \frac{R_F}{R_G}\right)} \quad (8\text{–}24)$$

The closed-loop ideal gain equation (inverting and noninverting) shows up in the denominator of Equation 8–24, so the closed-loop gain influences the stability of the op amp. When Z_B approaches zero, the closed-loop gain term also approaches zero, and the op amp becomes independent of the ideal closed-loop gain. Under these conditions R_F determines stability, and the bandwidth is independent of the closed-loop gain. Many people claim that the CFA bandwidth is independent of the gain, and that claim's validity is dependent on the ratios Z_B/Z_F being very low.

Z_B is important enough to warrant further investigation, so the equation for Z_B is given below.

$$Z_B \cong h_{ib} + \frac{R_B}{\beta_0 + 1}\left(\frac{1 + \frac{s\beta_0}{\omega_T}}{1 + \frac{S\beta_0}{(\beta_0 + 1)\omega_T}}\right) \tag{8–25}$$

At low frequencies h_{ib} = 50 Ω and $R_B/(\beta_0+1)$ = 25 Ω, so Z_B = 75 Ω. Z_B varies in accordance with Equation 8–25 at high frequencies. Also, the transistor parameters in Equation 8–25 vary with transistor type; they are different for NPN and PNP transistors. Because Z_B is dependent on the output transistors being used, and this is a function of the quadrant the output signal is in, Z_B has an extremely wide variation. Z_B is a small factor in the equation, but it adds a lot of variability to the current-feedback op amp.

8.7 Selection of the Feedback Resistor

The feedback resistor determines stability, and it affects closed-loop bandwidth, so it must be selected very carefully. Most CFA IC manufacturers employ applications and product engineers who spend a great deal of time and effort selecting R_F. They measure each noninverting gain with several different feedback resistors to gather data. Then they pick a compromise value of R_F that yields stable operation with acceptable peaking, and that value of R_F is recommended on the data sheet for that specific gain. This procedure is repeated for several different gains in anticipation of the various gains their customer applications require (often G = 1, 2, or 5). When the value of R_F or the gain is changed from the values recommended on the data sheet, bandwidth and/or stability is affected.

When the circuit designer must select a different R_F value from that recommended on the data sheet he gets into stability or low bandwidth problems. Lowering R_F decreases stability, and increasing R_F decreases bandwidth. What happens when the designer needs to operate at a gain not specified on the data sheet? The designer must select a new value of R_F for the new gain, but there is no guarantee that new value of R_F is an optimum value. One solution to the R_F selection problem is to assume that the loop gain, $A\beta$, is a linear function. Then the assumption can be made that $(A\beta)_1$ for a gain of one equals $(A\beta)_N$ for a gain of N, and that this is a linear relationship between stability and gain. Equations 8–26 and 8–27 are based on the linearity assumption.

$$\frac{Z}{Z_{F1} + Z_B\left(1 + \frac{Z_{F1}}{Z_{G1}}\right)} = \frac{Z}{Z_{FN} + Z_B\left(1 + \frac{Z_{FN}}{Z_{GN}}\right)} \qquad (8–26)$$

$$Z_{FN} = Z_{F1} + Z_B\left(\left(1 + \frac{Z_{F1}}{Z_{G1}}\right) - \left(1 + \frac{Z_{FN}}{Z_{GN}}\right)\right) \qquad (8–27)$$

Equation 8–27 leads one to believe that a new value for Z_F can easily be chosen for each new gain. This is not the case in the real world; the assumptions don't hold up well enough to rely on them. When you change to a new gain not specified on the data sheet, Equation 8–27, at best, supplies a starting point for R_F, but you must test to determine the final value of R_F.

When the R_F value recommended on the data sheet can't be used, an alternate method of selecting a starting value for R_F is to use graphical techniques. The graph shown in Figure 8–7 is a plot of the typical 300-MHz CFA data given in Table 8–1.

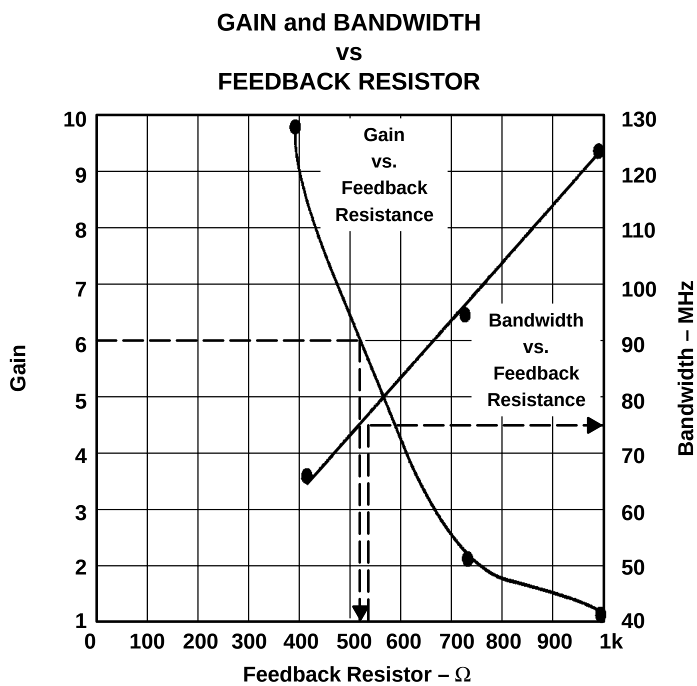


Figure 8–7. Plot of CFA R_F, G, and BW

Table 8–1. Data Set for Curves in Figure 8–7

GAIN (A_{CL})	R_F (Ω)	BANDWIDTH (MHz)
+ 1	1000	125
+ 2	681	95
+ 10	383	65

Enter the graph at the new gain, say A_{CL} = 6, and move horizontally until you reach the intersection of the gain versus feedback resistance curve. Then drop vertically to the resistance axis and read the new value of R_F (500 Ω in this example). Enter the graph at the new value of R_F, and travel vertically until you intersect the bandwidth versus feedback resistance curve. Now move to the bandwidth axis to read the new bandwidth (75 MHz in this example). As a starting point you should expect to get approximately 75 MHz BW with a gain of 6 and R_F = 500 Ω. Although this technique yields more reliable solutions than Equation 8–27 does, op amp peculiarities, circuit board stray capacitances, and wiring make extensive testing mandatory. The circuit must be tested for performance and stability at each new operating point.

8.8 Stability and Input Capacitance

When designer lets the circuit board introduce stray capacitance on the inverting input node to ground, it causes the impedance Z_G to become reactive. The new impedance, Z_G, is given in Equation 8–28, and Equation 8–29 is the stability equation that describes the situation.

$$Z_G = \frac{R_G}{1 + R_G C_G s} \qquad (8–28)$$

$$A\beta = \frac{Z}{Z_B + \frac{Z_F}{Z_G^2 + Z_B Z_G}} \qquad (8–29)$$

$$A\beta = \frac{Z}{R_F\left(1 + \frac{R_B}{R_F \| R_G}\right)\left(1 + R_B \| R_F \| R_G C_G s\right)} \qquad (8–30)$$

Equation 8–29 is the stability equation when Z_G consists of a resistor in parallel with stray capacitance between the inverting input node and ground. The stray capacitance, C_G, is a fixed value because it is dependent on the circuit layout. The pole created by the stray capacitance is dependent on R_B because it dominates R_F and R_G. R_B fluctuates with manufacturing tolerances, so the R_BC_G pole placement is subject to IC manufacturing tolerances. As the R_BC_G combination becomes larger, the pole moves towards the zero frequency axis, lowering the circuit stability. Eventually it interacts with the pole contained in Z, $1/\tau_2$, and instability results.

The effects of stray capacitance on CFA closed-loop performance are shown in Figure 8–8.

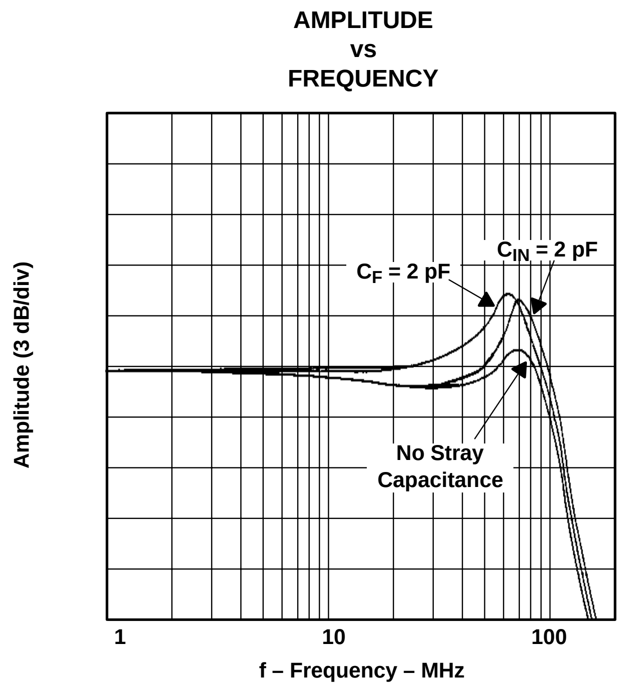


Figure 8–8. Effects of Stray Capacitance on CFAs

Notice that the introduction of C_G causes more than 3 dB peaking in the CFA frequency response plot, and it increases the bandwidth about 18 MHz. Two picofarads are not a lot of capacitance because a sloppy layout can easily add 4 or more picofarads to the circuit.

8.9 Stability and Feedback Capacitance

When a stray capacitor is formed across the feedback resistor, the feedback impedance is given by Equation 8–31. Equation 8–32 gives the loop gain when a feedback capacitor has been added to the circuit.

$$Z_F = \frac{R_F}{1 + R_F C_F s} \tag{8–31}$$

$$A\beta = \frac{Z\left(1 + R_F C_F s\right)}{R_F\left(1 + \frac{R_B}{R_F \| R_G}\right)\left(1 + R_B \| R_F \| R_G C_F s\right)} \tag{8–32}$$

This loop gain transfer function contains a pole and zero, thus, depending on the pole/zero placement, oscillation can result. The Bode plot for this case is shown in Figure 8–9. The original and composite curves cross the 0-dB axis with a slope of –40 dB/decade, so either curve can indicate instability. The composite curve crosses the 0-dB axis at a higher frequency than the original curve, hence the stray capacitance has added more phase shift to the system. The composite curve is surely less stable than the original curve. Adding capacitance to the inverting input node or across the feedback resistor usually results in instability. R_B largely influences the location of the pole introduced by C_F, thus here is another case where stray capacitance leads to instability.

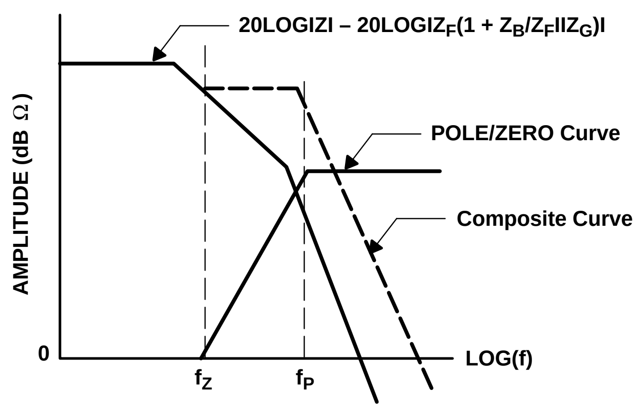


Figure 8–9. Bode Plot with C_F

Figure 8–8 shows that C_F = 2 pF adds about 4 dB of peaking to the frequency response plot. The bandwidth increases about 10 MHz because of the peaking. C_F and C_G are the major causes of overshoot, ringing, and oscillation in CFAs, and the circuit board layout must be carefully done to eliminate these stray capacitances.

8.10 Compensation of C_F and C_G

When C_F and C_G both are present in the circuit they may be adjusted to cancel each other out. The stability equation for a circuit with C_F and C_G is Equation 8–33.

$$A\beta = \frac{Z\left(1 + R_F C_F s\right)}{R_F\left(1 + \frac{R_B}{R_F \| R_G}\right)\left(R_B \| R_F \| R_G \left(C_F + C_G\right) s + 1\right)} \quad (8–33)$$

If the zero and pole in Equation 8–33 are made to cancel each other, the only poles remaining are in Z. Setting the pole and zero in Equation 8–33 equal yields Equation 8–34 after some algebraic manipulation.

$$R_F C_F = C_G\left(R_G \| R_B\right) \quad (8–34)$$

R_B dominates the parallel combination of R_B and R_G, so Equation 8–34 is reduced to Equation 8–35.

$$R_F C_F = R_B C_G \quad (8–35)$$

R_B is an IC parameter, so it is dependent on the IC process. R_B it is an important IC parameter, but it is not important enough to be monitored as a control variable during the manufacturing process. R_B has widely spread, unspecified parameters, thus depending on R_B for compensation is risky. Rather, the prudent design engineer assures that the circuit will be stable for any reasonable value of R_B, and that the resulting frequency response peaking is acceptable.

8.11 Summary

Constant gain-bandwidth is not a limiting criterion for the CFA, so the feedback resistor is adjusted for maximum performance. Stability is dependent on the feedback resistor; as R_F is decreased, stability is decreased, and when R_F goes to zero the circuit becomes unstable. As R_F is increased stability increases, but the bandwidth decreases.

The inverting input impedance is very high, but the noninverting input impedance is very low. This situation precludes CFAs from operation in the differential amplifier configuration. Stray capacitance on the inverting input node or across the feedback resistor always leads to peaking, usually to ringing, and sometimes to oscillations. A prudent circuit designer scans the PC board layout for stray capacitances, and he eliminates them. Breadboarding and lab testing are a must with CFAs. The CFA performance can be improved immeasurably with a good layout, good decoupling capacitors, and low inductance components.

DESIGN NOTES

Current Feedback Amplifier "Do's and Don'ts" – 46

William H. Gross

Introduction

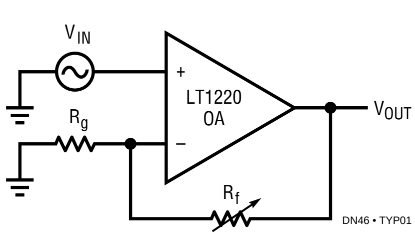
The introduction of current feedback amplifiers, such as the LT®1223, has significantly increased the designer's ability to solve difficult high speed amplifier problems. The current feedback architecture has very high slew rate and the small signal bandwidth is fairly constant for all gains. Current feedback amplifiers are used in broadcast video systems, radar systems, IF and RF stages, RGB distribution systems and many other high speed circuits.

As with any new circuit, there are several new rules that must be kept in mind to prevent problems. Because current feedback amplifiers act very much the same as regular op amps, it is important to note the differences and show how some standard op amp circuits should be implemented.

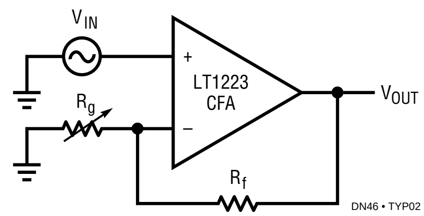
The most important thing to remember about current feedback amplifiers is that the impedance at the inverting (negative) input sets the bandwidth and therefore the stability of the amplifier. It should be resistive, not capacitive. To slow the amplifier down, increase the resistance driving the inverting input. If the amplifier peaks too much due to capacitive loading or anything else, increase the value of the feedback resistors.

The best way to demonstrate how to use current feedback amplifiers is to show some example circuits. To make it as painless as possible, I will show the traditional op amp implementation next to the current feedback amplifier version.

Op Amp Adjustable Gain Amp

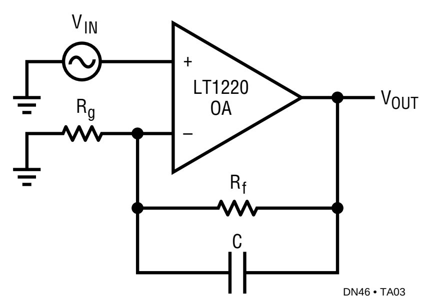


Current Feedback Amp Adjustable Gain Amp

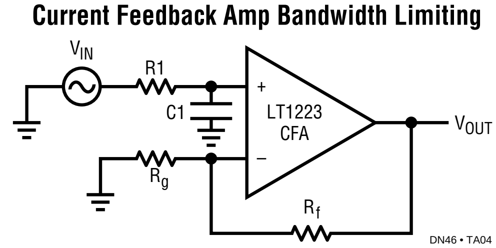


With a standard op amp you can vary the gain of the amplifier with either R_f or R_g. The only real restriction on the values is the loading affect the resistors have on the amplifier output. With a current feedback amplifier the value of R_f should not be varied. Do not make R_f the variable resistor or the bandwidth will be reduced at maximum gain and the circuit will oscillate when R_f is very small.

Op Amp Bandwidth Limiting

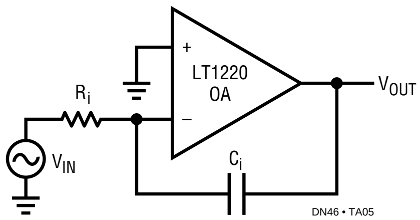


Current Feedback Amp Bandwidth Limiting

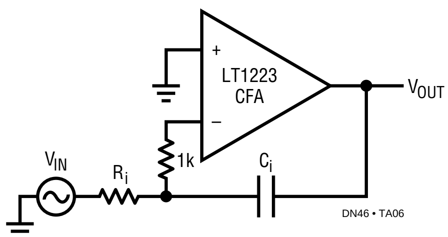


It is very common to limit the bandwidth of an op amp by putting a small capacitor in parallel with R_f. This works with all unity gain stable op amps; DO NOT PUT A SMALL CAPACITOR FROM THE INVERTING INPUT OF A CURRENT FEEDBACK AMPLIFIER TO ANYWHERE, ESPECIALLY NOT TO THE OUTPUT. The capacitor on the inverting input will cause peaking or oscillations. If you need to limit the bandwidth of a current feedback amplifier, use a resistor and capacitor at the non-inverting input (R1 and C1). This technique will also cancel (to a degree) the peaking caused by stray capacitance at the inverting input. Unfortunately, this will not limit the output noise the way it does for the op amp.

Op Amp Integrator

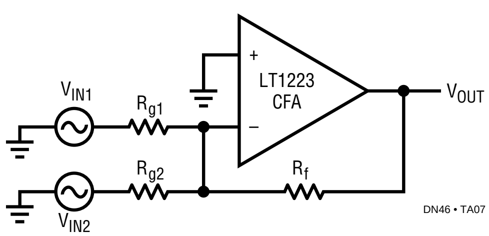


Current Feedback Amplifier Integrator



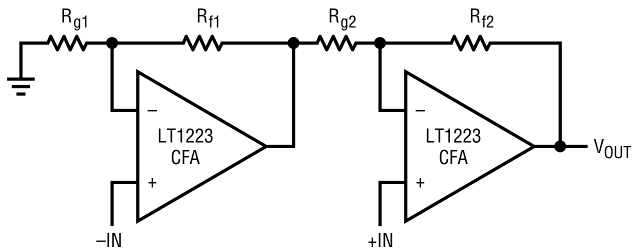
The integrator is one of the easiest circuits to make with an op amp. However, the circuit must be modified before a current feedback amplifier can be used. Since we remember that the inverting input wants to see a resistor, we can add one to the standard circuit. This generates a new summing node where we can apply capacitive feedback. The new current feedback amplifier compatible integrator works just like you would expect; it has excellent large signal capability and accurate phase shift at high frequencies.

Current Feedback Amplifier Summer (DC Accurate)



There is no I_{OS} spec on current feedback amplifiers because there is no correlation between the two input bias currents. Therefore we will not improve the DC accuracy of the inverting amplifier by putting an extra resistor in the non-inverting input. This is also true of input bias current canceled op amps where the I_{OS} spec is the same as the I_B spec, such as the LT1220.

Two Amplifier Instrumentation Amp



TRIM R_{g2} FOR GAIN, THEN TRIM R_{g1} FOR CMRR. VOLTAGE GAIN, G, IS V_{OUT} DIVIDED BY DIFFERENCE BETWEEN +IN AND –IN.

OP AMP DESIGN EQUATIONS:

$R_{f1} = R_{g2}$; $R_{f2} = (G–1)\ R_{g2}$; $R_{g1} = R_{f2}$

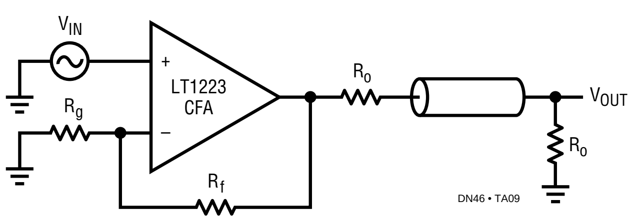
CURRENT FEEDBACK AMP DESIGN EQUATIONS:

$R_{f1} = R_{f2}$; $R_{g1} = (G–1)\ R_{f2}$; $R_{g2} = \frac{R_{f2}}{G–1}$

DN46 • TA08

The two amplifier instrumentation amp is easily modified for current feedback amplifiers. The only necessary change is to make the feedback resistor of each amplifier the same and therefore make the gain setting resistors different. This way the bandwidth of both amps is the same and the common mode rejection at high frequencies is better than that of the op amp circuit. In the op amp circuit one amplifier has maximum bandwidth, since it runs at about unity gain, while the other is limited to its gain bandwidth product divided by the gain.

Cable Driver



The cable driver circuit is the same for both types of amplifiers. But because most op amps do not have enough output drive current, they are not often used for heavy loads like cables. When driving a cable it is important to properly terminate both ends if even modest high frequency performance is required. The additional advantage of this is that it isolates the capacitive load of the cable from the amplifier so it can operate at maximum bandwidth.

For literature on our Current Feedback Amplifiers, call **1-800-4-LINEAR**. For applications help, call (408) 432-1900, Ext. 2593

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Ask The Applications Engineer—22

by Erik Barnes

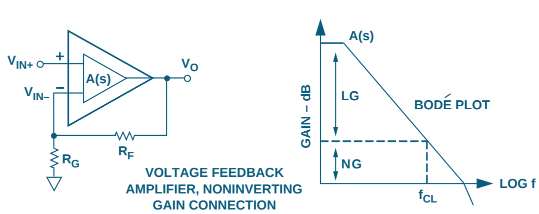
CURRENT FEEDBACK AMPLIFIERS—I

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)).

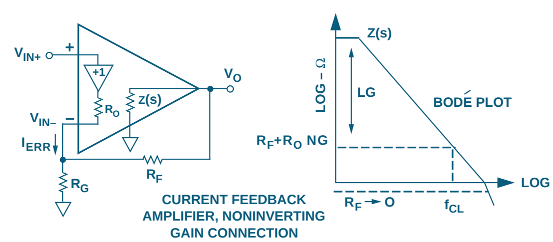
With $$V_o = (V_{IN+} - V_{IN-})A(s)$$

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.



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.

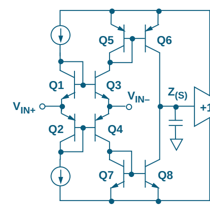
In practice, the input buffer's non-ideal output resistance will be typically about 20 to 40 Ω, which will modify the feedback transresistance. The two input voltages will not be exactly equal. Making the substitution into the previous equations with $V_{IN-} = V_{IN+} - I_{err}R_o$, and solving for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

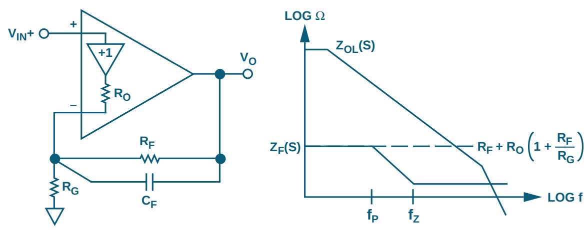
A. I'm glad you asked; this is a common misconception of CF op-amps. True, the inputs are not matched, but the transfer function for the ideal difference amplifier will still work out the same. What about the unbalanced inputs? At lower frequencies, the four-resistor differential amplifier's CMR is limited 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

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

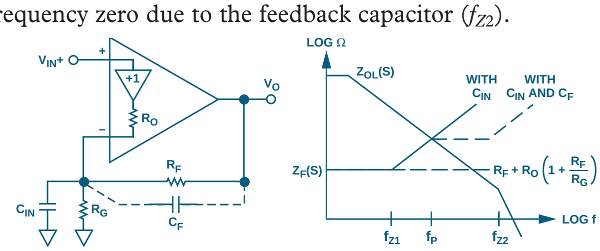
The pole occurs at $1/2\pi R_FC_F$, and the zero occurs higher in frequency at $1/[2\pi(R_F||R_G||R_O)C_F]$. If the intersection of Z_F and Z_{OL} occurs too high in frequency, instability may result from excessive open loop phase shift. If $R_F\rightarrow\infty$, as with an integrator circuit, the pole 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}:

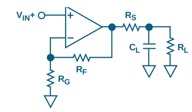
$$Z_F(s) = \left[R_F + R_O \left(1 + \frac{R_F}{R_G} \right) \right] \left[1 + \frac{sC_{IN} R_F R_G R_O}{R_F R_G + R_F R_O + R_G R_O} \right]$$

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 $R_F = 750\ \Omega$, $R_G = 750\ \Omega$, 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.

Ask The Applications Engineer—23

by Erik Barnes

CURRENT FEEDBACK AMPLIFIERS—II

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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?*

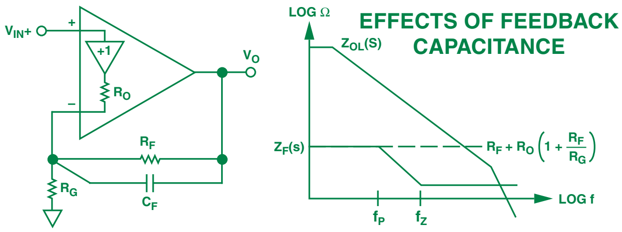
A. I'm glad you asked; this is a common misconception of CF op amps. True, the inputs are not matched, but the transfer function for the ideal difference amplifier will still work out the same. What about the unbalanced inputs? At lower frequencies, the four-resistor differential amplifier's CMR is limited 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. Both VF and CF amplifiers can further benefit from additional hand-trimmed capacitors at the expense of signal bandwidth. 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

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

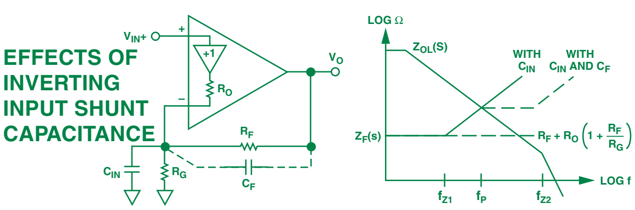
The pole occurs at $1/2\pi R_FC_F$, and the zero occurs higher in frequency at $1/[2\pi(R_F||R_G||R_O)C_F]$. If the intersection of Z_F and Z_{OL} occurs too high in frequency, instability may result from excessive open loop phase shift. If $R_F\rightarrow\infty$, as with an integrator circuit, the pole 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.



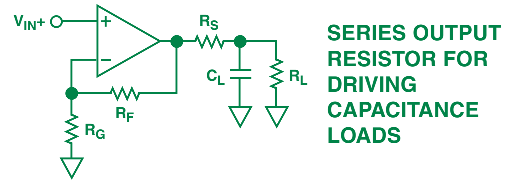
Another issue to consider is the effect of shunt capacitance at the inverting input. Recall that with a voltage feedback amplifier, such capacitance creates a zero in the noise gain, increasing the rate of closure between the noise gain and open loop gain, generating excessive phase shift that can lead 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. A