# Introduction

Current feedback amplifiers (CFA) have sacrificed the DC precision of voltage feedback amplifiers (VFA) in a trade-off for increased slew rate and a bandwidth that is relatively independent of the closed loop gain. Although CFAs do not have the DC precision of their VFA counterparts, they are good enough to be DC coupled in video applications without sacrificing too much dynamic range. The days when high frequency amplifiers had to be AC coupled are gone forever, because some CFAs are approaching the GHz gain bandwidth region. The slew rate of CFAs is not limited by the linear rate of rise that is seen in VFAs, so it is much faster and leads to faster rise/fall times and less intermodulation distortion.

The general feedback theory used in this paper is developed in Intersil Application Note Number AN9415 entitled “Feedback, Op Amps and Compensation.” The approach to the development of the circuit equations is the same as in the referenced application note, and the symbology/terminology is the same with one exception. The impedance connected from the negative op amp input to ground, or to the source driving the negative input, will be called Z\textsubscript{I} rather than Z\textsubscript{I} or Z\textsubscript{F}, because this has become the accepted terminology in CFA papers.

## Development of the General Feedback Equation

Referring to the block diagram shown in Figure 1, Equation 1, Equation 2 and Equation 3 can be written by inspection if it is assumed that there are no loading concerns between the blocks. This assumption is implicit in all block diagram calculations, and requires that the output impedance of a block be much less than input impedance of the block it is driving. This is usually true by one or two orders of magnitude. Algebraic manipulation of Equation 1, Equation 2 and Equation 3 yields Equation 4 and Equation 5 which are the defining equations for a feedback system.

\[
\begin{align*}
V_O & = EA \quad \text{ (EQ. 1)} \\
E & = V_I - \beta V_O \quad \text{ (EQ. 2)} \\
E & = V_O/A \quad \text{ (EQ. 3)} \\
V_O/V_I & = A(1 + A\beta) \quad \text{ (EQ. 4)} \\
E/V_I & = 1/(1 + A\beta) \quad \text{ (EQ. 5)}
\end{align*}
\]

In this analysis the parameter A, which usually includes the amplifier and thus contains active elements, is called the direct gain. The parameter \(\beta\), which normally contains only passive components, is called the feedback factor. Notice that in Equation 4 as the value of A approaches infinity the quantity A\(\beta\), which is called the loop gain, becomes much larger than one; thus, Equation 4 can be approximated by Equation 6.

\[
\frac{V_O}{V_I} = \frac{1}{\beta} \quad \text{ for } A\beta >> 1 \quad \text{ (EQ. 6)}
\]

\(V_O/V_I\) is called the closed loop gain. Because the direct gain, or amplifier response, is not included in Equation 6, the closed loop gain (for \(A >> 1\)) is independent of amplifier parameter changes. This is the major benefit of feedback circuits.

Equation 4 is adequate to describe the stability of any feedback circuit because these circuits can be reduced to this generic form through block diagram reduction techniques [1]. The stability of the feedback circuit is determined by setting the denominator of Equation 4 equal to zero.

\[
1 + A\beta = 0 \quad \text{ (EQ. 7)}
\]

\[
A\beta = -1 = \frac{1}{|-180|} \quad \text{ (EQ. 8)}
\]

Observe from Equation 4 and Equation 8, that if the magnitude of the loop gain can achieve a magnitude of one while the phase shift equals -180 degrees, the closed loop gain becomes undefined because of division by zero. The undefined state is unstable, causing the circuit to oscillate at the frequency where the phase shift equals -180 degrees. If the loop gain at the frequency of oscillation is slightly greater than one, it will be reduced to one by the reduction in gain suffered by the active elements as they approach the limits of saturation. If the value of A\(\beta\) is much greater than one, gross nonlinearities can occur and the circuit may cycle between saturation limits. Preventing instability is the essence of feedback circuit design, so this topic will be touched lightly here and covered in detail later in this application note.

A good starting point for discussing stability is finding an easy method to calculate it. Figure 2 shows that the loop gain can be calculated from a block diagram by opening current inputs, shorting voltage inputs, breaking the circuit and calculating the response (\(V_{TO}\)) to a test input signal (\(V_{TI}\)).
\[ \frac{V_{TO}}{V_{TI}} = A\beta \]  

 EQ. 9

**Current Feedback Stability Equation Development**

The CFA model is shown in Figure 3. The non-inverting input connects to the input of a buffer, so it is a very high impedance on the order of a bipolar transistor VFA's input impedance. The inverting input ties to the buffer output; \( Z_B \) models the buffer output impedance, which is usually very small, often less than 50Ω. The buffer gain, \( G_B \), is nearly but always less than one because modern integrated circuit design methods and capabilities make it easy to achieve. \( G_B \) is overshadowed in the transfer function by the transimpedance, \( Z \), so it will be neglected in this analysis.

The output buffer must present a low impedance to the load. Its gain, \( G_{OUT} \), is one, and is neglected for the same reason as the input buffer’s gain is neglected. The output buffer’s impedance, \( Z_{OUT} \), affects the response when there is some output capacitance; otherwise, it can be neglected unless DC precision is required when driving low impedance loads.

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**Equation 10**

\[ V_{TI} = I_2(Z_F + Z_G/Z_B) \]

**Equation 11**

\[ V_{TO} = I_1Z \]

**Equation 12**

\[ I_2(Z_G/Z_B) = I_1Z_B; \text{ For } G_B = 1 \]

Equation 10 and Equation 12 are combined to obtain Equation 13.

\[ V_{TI} = I_1(Z_F + Z_G/Z_B)(1 + Z_B/Z_G) = I_1Z_F (1 + Z_B/Z_G/Z_G) \]  

EQ. 13

Dividing Equation 11 by Equation 13 yields Equation 14 which is the defining equation for stability. Equation 14 will be examined in detail later, but first the circuit equations for the inverting and non-inverting circuits must be developed so that all of the equations can be examined at once.

\[ A\beta = \frac{V_{TO}}{V_{TI}} = \frac{Z}{Z_F + Z_G/Z_B} \]  

EQ. 14

**Developing the Non-Inverting Circuit Equation and Model**

Equation 15 is the current equation at the inverting input of the circuit shown in Figure 6. Equation 16 is the loop equation for the input circuit, and Equation 17 is the output circuit equation. Combining these equations yields Equation 18, in the form of Equation 4, which is the non-inverting circuit equation.
The block diagram equivalent for the non-inverting circuit is shown in Figure 7.

![Figure 7. Block Diagram of the Non-Inverting CFA](image)

**Developing the Inverting Circuit Equation and Model**

Equation 19 is the current equation at the inverting input of the circuit shown in Figure 8. Equation 20 defines the dummy variable \( V_X \), and Equation 21 is the output circuit equation. Equation 22 is developed by substituting Equation 20 and Equation 21 into Equation 19, simplifying the result, and manipulating it into the form of Equation 4.

![Figure 8. Inverting Circuit Diagram](image)

\[
\begin{align*}
\frac{V_{IN}}{V_{OUT}} &= \frac{V_{IN}}{V_{OUT}} - 1 = \frac{V_{X} - V_{OUT}}{Z_{F}} \\
I_{Z} &= -V_{X} \\
I_{Z} &= \frac{V_{OUT}}{Z_{G}} \\
V_{OUT} &= \frac{Z_{G}(1 + Z_{B}/Z_{F})}{Z_{G}}
\end{align*}
\]  
(EQ. 18)

The block diagram equivalent for the inverting circuit is given in Figure 9.

![Figure 9. Inverting Block Diagram](image)

**Stability**

Equation 8 states the criteria for stability, but there are several methods for evaluating this criteria. The method that will be used in this paper is called the Bode plot [2] which is a log plot of the stability equation. A brief explanation of the Bode plot procedure is given in “Feedback, Op Amps and Compensation” [3]. The magnitude and phase of the open loop transfer function are both plotted on logarithmic scales, and if the gain decreases below zero dB before the phase shift reaches 180 degrees the circuit is stable. In practice the phase shift should be \( \leq 140 \) degrees, i.e., greater than 40 degrees phase margin, to obtain a well behaved circuit. A sample Bode plot of a single pole circuit is given in Figure 10.

![Figure 10. Sample Bode Plot](image)

Referring to Figure 10, notice that the DC gain is 20dB, thus the circuit gain must be equal to 10. The amplitude is down 3dB at the break point, \( \omega = 1/RC \), and the phase shift is -45 degrees at this point. The circuit can not become unstable with only a single pole response because the maximum phase shift of the response is -90 degrees.

CFA circuits often oscillate, intentionally or not, so there are at least two poles in their loop gain transfer function. Actually, there are multiple poles in the loop gain transfer function, but the CFA circuits are represented by two poles for two reasons: a two pole approximation gives satisfactory correlation with laboratory results, and the two pole mathematics are well known and easy to understand. Equation 14, the stability equation for the CFA, is given in logarithmic form as Equation 23 and Equation 24.

\[
\begin{align*}
20\log|A| &= 20\log|Z/(Z_{F}(1 + Z_{B}/Z_{F}||Z_{G}))| \\
\phi &= \text{TANGENT}^{-1}(Z/(Z_{F}(1 + Z_{B}/Z_{F}||Z_{G})))
\end{align*}
\]  
(EQ. 23)

(EQ. 24)
The answer to the stability question is found by plotting these functions on log paper. The stability equation, \(20\log|A\beta|\), has the form \(20\log x/y\) which can be written as \(20\log x - 20\log y\). The numerator and denominator of Equation 23 will be operated on separately, plotted independently and then added graphically for analysis. Using this procedure the independent variables can be manipulated separately to show their individual effects. Figure 11 is the plot of Equation 23 and Equation 24 for a typical CFA where \(Z = 1\text{M}\Omega (1/1s + 1)(1/2s + 1)\), \(Z_F = Z_G = 1\text{k}\Omega\), and \(Z_B = 70\Omega\).

If \(20\log[Z_F(1+Z_B/Z_F)||Z_G]|\) were equal to 0dB the circuit would oscillate because the phase shift of \(Z\) reaches -180 degrees before \(20\log[Z]\) decreases below zero. Since \(20\log[Z_F(1+Z_B/Z_F)||Z_G]| = 61.1dB\Omega\), the composite curve moves down by that amount to 58.9dB\Omega where it is stable because it has 120 degrees phase shift or 60 degrees phase margin. If \(Z_B = 0\Omega\) and \(Z_F = R_F\), then \(A\beta = Z/R_F\). In this special case, stability is dependent on the transfer function of \(Z\) and \(R_F\), and \(R_F\) can always be specified to guarantee stability. The first conclusion drawn here is that \(Z_F(1+Z_B/Z_F)||Z_G||\) has an impact on stability, and that the feedback resistor is the dominant part of that quantity so it has the dominant impact on stability. The dominant selection criteria for \(R_F\) is to obtain the widest bandwidth with an accepted amount of peaking; 60 degrees phase margin is equivalent to approximately 10% or, 0.83dB, overshoot. The second conclusion is that the input buffer’s output impedance, \(Z_B\), will have a minor effect on stability because it is small compared to the feedback resistor, even though it is multiplied by \(1/Z_F||Z_G||\) which is related to the closed loop gain. Rewriting Equation 14 as \(A\beta = Z/(Z_F+Z_B(1+R_F/R_G))\) leads to the third conclusion which is that the closed loop gain has a minor effect on stability and bandwidth because it is multiplied by \(Z_G||\) which is a small quantity relative to \(Z_F\). It is because of the third conclusion that many people claim closed loop gain versus bandwidth independence for the CFA, but that claim is dependent on the value of \(Z_B\) relative to \(Z_F\).

CFAs are usually characterized at a closed loop gain \((G_{CL})\) of one. If the closed loop gain is increased then the circuit becomes more stable, and there is the possibility of gaining some bandwidth by decreasing \(Z_F\). Assume that \(A\beta_1 = A\beta_N\) where \(A\beta_1\) is the loop gain at a closed loop gain of one and \(A\beta_N\) is the loop gain at a closed loop gain of \(N\); this insures that stability stays constant. Through algebraic manipulation, Equation 14 can be rewritten in the form of Equation 25 and solved to yield Equation 27 and a new \(Z_{FN}\) value.

\[
\frac{Z_{F1} + Z_{B} (1 + Z_{F1}/Z_G)}{Z_{F1} + Z_{B} G_{CL1}} = \frac{Z}{Z_{FN} + Z_{B} (1 + Z_{FN}/Z_G)}
\]

(EQ. 25)

\[
\frac{Z_{F1} + Z_{B} G_{CL1}}{Z_{F1} + Z_{B} G_{CLN}} = \frac{Z}{Z_{FN} + Z_{B} G_{CLN}}
\]

(EQ. 26)

\[
Z_{FN} = Z_{F1} + Z_{B} (G_{CL1} - G_{CLN})
\]

(EQ. 27)

For the HA5020 at a closed loop gain of 1, if \(Z = 6\text{M}\Omega\), \(Z_{F1} = 1\text{k}\Omega\), and \(Z_B = 75\Omega\), then \(Z_{F2} = 925\Omega\). Experimentation has shown, however, that \(Z_{F2} = 681\Omega\) yields better results. The difference in the predicted versus the measured results is that \(Z_B\) is a frequency dependent term which adds a zero in the loop gain transfer function that has a much larger effect on stability. The equation for \(Z_B\) is given below.

\[
Z_B = h_{ib} + \frac{R_B}{|p_0 + 1|} \left( \frac{1 + s|/\omega_T}{1 + s|/|p_0 + 1|\omega_T} \right)
\]

(EQ. 28)

At low frequencies \(h_{ib} = 500\Omega\) and \(R_B/(|p_0 + 1| = 25\Omega\) which corresponds to \(Z_B = 75\Omega\), but at higher frequencies \(Z_B\) will vary according to Equation 28. This calculation is further complicated because \(\beta_0\) and \(\omega_T\) are different for NPN and PNP transistors, so \(Z_B\) also is a function of the polarity of the output. Refer to Figure 12 and Figure 13 for plots of the transimpedance \((Z)\) and \(Z_B\) for the HA5020 [5]. Notice that \(Z\) starts to level off at 20MHz which indicates that there is a zero in the transfer function. \(Z_B\) also has a zero in its transfer function located at about 65MHz. The two curves are related, and it is hard to determine mathematically exactly which parameter is affecting the performance, thus considerable lab work is required to obtain the maximum performance from the device.

![FIGURE 12. HA5020 TRANSIMPEDEANCE vs FREQUENCY](image-url)
Equation 5, which is rewritten here as Equation 29, expresses the error signal as a function of the loop gain for any feedback system. Consider a VFA non-inverting configuration where the closed loop gain is +1; then the loop gain, $A\beta$, is $a$. It is not uncommon to have VFA amplifier gains of 50,000 in high frequency op amps, such as the HA2841 [6], so the DC precision is then 100% ($1/50,000 = 0.002\%$). In a good CFA the transimpedance is $Z = 6M\Omega$, but $Z_F = 1k\Omega$ so the DC precision is 100% ($1075\Omega/6M\Omega = 0.02\%$). The CFA often sacrifices DC precision for stability.

$$\text{Error} = \frac{V_f}{1 + A\beta} \quad (\text{EQ. 29})$$

The DC precision is the best accuracy that an op amp can obtain, because as frequency increases the gain, $a$, or the transimpedance, $Z$, decreases causing the loop gain to decrease. As the frequency increases the constant gainbandwidth VFA starts to lose gain first, then the CFA starts to lose gain. There is a crossover point, which is gain dependent, where the AC accuracy for both op amps is equal. Beyond this point the CFA has better AC accuracy.

The VFA input structure is a differential transistor pair, and this configuration makes it easy to match the input bias currents, so only the offset current generates an offset error voltage. The time honored method of inserting a resistor, equal to the parallel combination of the input and feedback resistors, in series with the non-inverting input causes the bias current to be converted to a common mode voltage. VFAs are very good at rejecting common mode voltages, so the bias current error is cancelled. One input of a CFA is the base terminal of a transistor while the other input is the output of a low impedance buffer. This explains why the input currents don’t cancel, and why the non-inverting input impedance is high while the inverting input impedance is low. Some CFA’s, such as the HA1120 [7], have input pins which enable the adjustment of the offset current. Newer CFAs are finding solutions to the DC precision problem.

**Stability Calculations for Input Capacitance**

When there is a capacitance from the inverting input to ground, the impedance $Z_G$ becomes $R_G/(R_GC_Gs+1)$, and Equation 14 can be written in the form of Equation 30. Then the new values for $Z_G$ are put into the equation to yield Equation 31. Notice that the loop gain has another pole in it: an added pole might cause an oscillation if it gets too close to the pole(s) included in $Z$. Since $Z_G$ is small it will dominate the added pole location and force the pole to be at very high frequencies. When $C_G$ becomes large the pole will move in towards the poles in $Z$, and the circuit may become unstable.

$$A\beta = \frac{Z}{(Z_B + Z_F/Z_G + Z_B)} \quad (\text{EQ. 30})$$

If $Z_B = R_B$, $Z_F = R_F$, and $Z_G = R_G||C_G$, Equation 30 becomes:

$$A\beta = \frac{Z}{R_F(1 + R_B/R_F||R_G)(R_B||R_F||R_G||C_Gs + 1)} \quad (\text{EQ. 31})$$

The CFA’s transimpedance, which is also a function of frequency, shows up in both the loop gain and closed loop gain equations, Equations 18 and 22. The gain setting impedances, $Z_F$ and $Z_G$, do not appear in the loop gain as a ratio unless they are multiplied by a secondary quantity, $Z_B$, so $Z_F$ can be adjusted independently for maximum bandwidth. This is why the bandwidth of CFAs are relatively independent of closed loop gain. When $Z_B$ becomes a significant portion of the loop gain the CFA becomes more of a constant gain bandwidth device.
Stability Calculations for Feedback Capacitance

When a capacitor is placed in parallel with the feedback resistor, the feedback impedance becomes \( Z_F = R_F/(R_F+1) \). After the new value of \( Z_F \) is substituted into Equation 30, and with considerable algebraic manipulation, it becomes Equation 32.

\[
A(\beta) = \frac{Z(Z_F C_F s + 1)}{R_F(1 + (R_B/R_F)||R_G)/(R_B/R_F)||R_G||R_G C_F s + 1)} \quad (EQ. 32)
\]

The new loop gain transfer function now has a zero and a pole; thus, depending on the placement of the pole relative to the zero oscillations can result.

The loop gain plot for a CFA with a feedback capacitor is shown in Figure 14. The composite curve crosses the 0dB axis with a slope of -40dB/decade, and it has more time to accumulate phase shift, so it is more unstable than it would be without the added poles and zeros. If the pole occurred at a frequency much higher than the highest frequency pole in \( Z \) then the \( Z \) pole would have a chance to roll off the gain before any phase shift from \( Z \) could add to the phase shift from the pole. In this case, \( C_F \) would be very small and the circuit would be stable. In practice almost any feedback capacitance will cause ringing and eventually oscillation if the capacitor gets large enough. There is the case where the zero occurs just before the \( A(\beta) \) curve goes through the 0dB axis. In this case the positive phase shift from the zero cancels out some of the negative phase shift from the second pole in \( Z \); thus, it makes the circuit stable, and then the pole occurs after the composite curve has passed through 0dB.

Calculations and Compensation for \( C_G \) and \( C_F \)

\( Z_G \) and \( Z_F \) are modified as they were in the previous two sections, and the results are incorporated into Equation 30, yielding Equation 33.

\[
A(\beta) = \frac{Z(Z_F C_F s + 1)}{R_F(1 + (R_B/R_F)||R_G)/(R_B/R_F)||R_G||R_G C_F s + 1)} \quad (EQ. 33)
\]

Notice that if the zero cancelled the pole in equation that the circuit AC response would only depend on \( Z \), so Equation 34 is arrived at by doing this. Equation 35 is obtained by algebraic manipulation.

\[
(R_F C_F s + 1) = (R_B/R_F)||R_G/(C_F + C_G)s + 1) \quad (EQ. 34)
\]

\[
R_F C_F = C_G R_G/R_B/(R_G + R_B) \quad (EQ. 35)
\]
References


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