ISL28134, READ2302G

Composite Amplifier Design for High Gain Applications

Abstract

Industrial signal conditioners with high signal gain often use composite amplifiers. A composite amplifier is an amplifier whose design is composed of a high-precision and a high-speed or fast-slewing operational amplifier (op-amp). Some of the performance improvements of the composite design versus the individual op-amps include low output offset (especially at high gain) high-gain accuracy, increased signal bandwidth, and low-distortion of large output signals.

This application note discusses the design of a composite amplifier using the precision op-amp ISL28134, and the fast slewing op-amp READ2302. The key parameters of both amplifiers are listed in Table 1.

Table 1.   Key Parameters

Parameter	Symbol	A1 = ISL28134	A2 = READ2302
Supply Range	V_{CC}	2.25V to 6.0V	2.25V to 6.0V
Operating Temperature Range	T_A	-40°C to +125°C	-40°C to +105°C
Offset Voltage	V_{OS}	0.2µV	6mV
DC Open-Loop Gain	a	174dB	90dB
Slew Rate	SR	1V/µs	8V/µs
Gain Bandwidth	f_c	3.5MHz	6MHz
Phase Margin	Φ_m	75°	65°

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1.   Circuit Overview

Figure 1 shows the composite amplifier with the precision op-amp, A1, and the fast switching op-amp, A2. The composite open-loop gain is the product of the individual open-loop gains: $A_{OLc} = A_{OL1} \cdot A_{OL2}$.
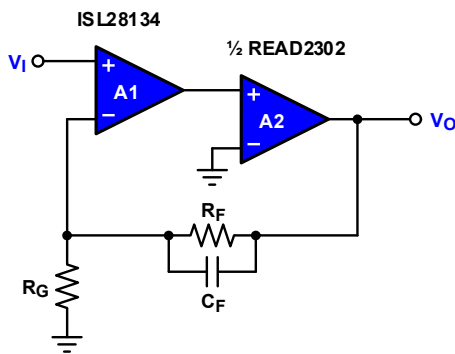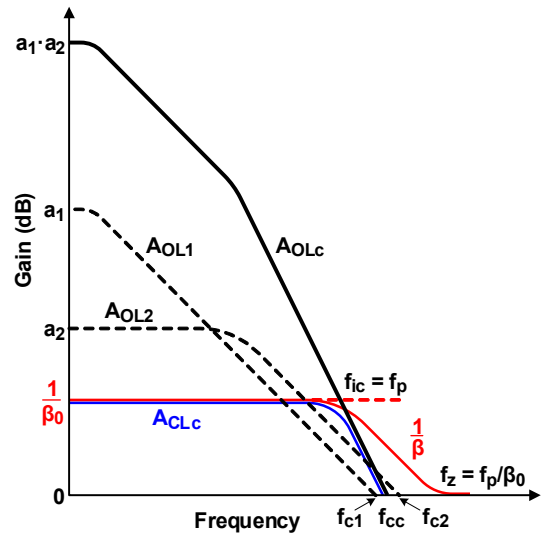

Figure 1. Composite Amplifier

Figure 2. Frequency Response

Figure 2 shows A_{OL1} and A_{OL2} rolling off at a rate of -20dB/decade. A_{OLc} however, starts rolling off at the rate of A_{OL1}, with -20dB/decade, which increases to -40dB/decade, when A_{OL2} starts rolling off. This faster roll-off leads to the crossing of AOLc with AOL2 prior to reaching unity gain. The two unity crossover frequencies, f_{c1} and f_{c2}, therefore determine the unity crossover frequency of the composite case, f_{cc}, using: $f_{cc} = \sqrt{f_{c1} \cdot f_{c2}}$.

The single feedback path, using R_G and R_F, sets the ideal gain of the composite amplifier to $1/\beta_0 = 1 + R_F/R_G$. This gain crosses A_{OLc} during its 40dB roll-off, where the composite phase shift is Φ = -180°.

Here the loop-gain $T = \left|A_{OLc}\right| e^{-j180°} \cdot \beta_0 = -1$ bears the risk of self-sustaining oscillations.

To ensure circuit stability, capacitor C_F is connected parallel to R_F, which introduces a pole and a zero into the gain function:

$$\frac{1}{\beta} = \frac{1}{\beta_0} \cdot \frac{1 + jf/f_z}{1 + jf/f_p} \quad \text{with} \quad f_p = \frac{1}{2\pi C_F R_F} \quad \text{and} \quad f_z = \frac{1}{2\pi C_F (R_F \| R_G)}$$

The now complex 1/β function generates a positive phase angle that reduces the total phase shift of the loop-gain, T.

However, to fully restore circuit stability while preserving maximum signal bandwidth, the pole and zero frequencies must be sufficiently apart to create enough positive phase angle. This in turn requires that $R_F \gg R_G$. Therefore, feedback phase compensation for maximum bandwidth lends itself to high-gain applications. Here, the optimum C_F value for maximum signal bandwidth is:

$$\text{(EQ. 1)} \qquad C = \frac{1}{2\pi R_F \sqrt{f_{c1} f_{c2} \beta_0}}$$

The resulting composite closed-loop gain, A_{CLc}, is close to the 1/β line, up to the A_{OLc} roll-off, and then follows A_{OLc} down to unity.

2.   Detailed Derivations for the Practical Amplifier

The following sections derive the equations and key parameters for the practical amplifier design using the ISL28134 as precision amplifier, A1, and the READ2302 as fast slewing, A2.

2.1   Unity Gain Bandwidth

Figure 3 shows A_{OL1} and A_{OL2} approaching unity gain at -20dB/decade, while A_{OLc} is rolling off at -40dB/decade. The unity gain bandwidth of the composite open-loop gain is at the frequency where the product of $1 / A_{OL1} \cdot A_{OL2} = 1V/V$.

Denoting these gains as $1/A_x$ and A_x, the gain-bandwidth products of A_{OL1} and A_{OL2} defines them through: $A_x = f_{c2}/f_{cc}$ and $A_x = f_{cc}/f_{c1}$. Setting both equations equal and solving for f_{cc} gives: $f_{cc} = \sqrt{f_{c1}f_{c2}}$.

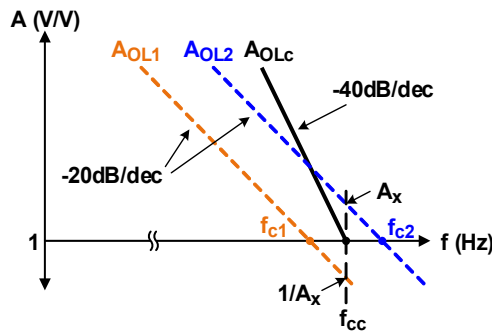**Figure 3. Composite Unity Gain Bandwidth**

Therefore, the unity-gain bandwidth of our composite amplifier is $f_{cc} = \sqrt{3.5MHz \cdot 6MHz} = 4.58MHz$.

2.2   Compensation Capacitor, C_F, for Maximum Bandwidth or Maximum Phase Margin?

Figure 4 depicts the loop-gain $T = A_{OLc} \cdot \beta_0$. From op-amp theory we recall that the phase shift between input voltage, V_I, and feedback voltage, V_{FB}, should never reach -180°, due to the possibility of oscillations. However, long before this angle, gain peaking and signal overshoot occur. A commonly applied reference value for minimum phase margin and therefore, circuit stability, is $\Phi_m = 45°$, as here the gain peaking is still below 3dB (Figure 5).
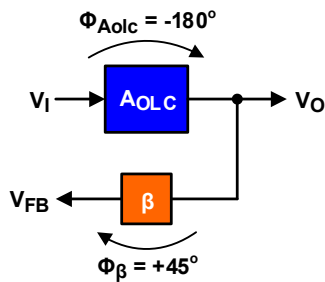
Figure 4. Loop Gain: $A_{OLc} \cdot \beta$

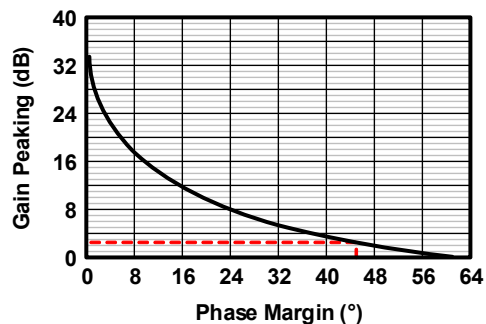**Figure 5. Gain Peaking vs Phase Margin**

In the composite amplifier however, the -180° are already consumed by the two open-loop gains. To restore phase margin, the feedback factor must be modified to produce enough positive phase angle (such as 45°) to ensure circuit stability. This is accomplished by connecting a capacitor parallel to R_F and is known as feedback phase compensation (Figure 6 on page 4).
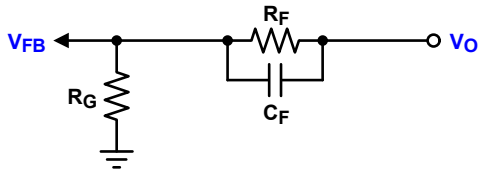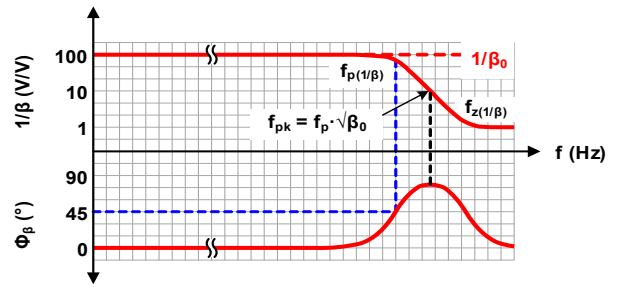
Figure 6. Feedback Phase Compensation

Figure 7. 1/β-Frequency Response and Feedback Factor, β, Phase Response

The resulting complex feedback factor then becomes:

$$\beta = \frac{R_G}{R_G + R_F \| X_C} = \beta_0 \cdot \frac{1 + jf/f_{z(\beta)}}{1 + jf/f_{p(\beta)}} \quad \text{with} \quad \beta_0 = \frac{R_G}{R_G + R_F} \, , \quad f_{z(\beta)} = \frac{1}{2\pi C_F R_F} \quad \text{and} \quad f_{p(\beta)} = \frac{1}{2\pi C_F R_F \| R_G}$$

Because Bode plots use the 1/β rather than the β representation, we establish the reciprocal of β with:

$$\frac{1}{\beta} = \frac{1}{\beta_0} \cdot \frac{1 + jf/f_{p(\beta)}}{1 + jf/f_{z(\beta)}} \quad \text{with} \quad \frac{1}{\beta_0} = 1 + \frac{R_F}{R_G} \, , \quad f_{p(1/\beta)} = \frac{1}{2\pi C_F R_F} \quad \text{and} \quad f_{z(1/\beta)} = \frac{1}{2\pi C_F R_F \| R_G}$$

Note: The zero pole and zero frequencies of the β response reverse for the 1/β response: $f_{z(\beta)} \rightarrow f_{p(1/\beta)}$ and $f_{p(\beta)} \rightarrow f_{z(1/\beta)}$. Figure 7 shows the frequency response of 1/β and the corresponding phase response of β. From the previous equation for $f_{p(1/\beta)}$ and $f_{z(1/\beta)}$, we find that $f_{z(1/\beta)} / f_{p(1/\beta)} = 1 + R_F/R_G = 1/\beta_0$ so $f_{z(1/\beta)} = f_{p(1/\beta)} / \beta_0$. Therefore, for $R_F = 100\, R_G$, $1/\beta_0 = 101$ and $f_{z(1/\beta)} = 101 \cdot f_{p(1/\beta)}$.

With f_p and f_z being two frequency decades apart, Φ_β, which is also the phase margin Φ_m, is 45° at f_p. The maximum phase margin however, occurs at f_{pk}, which is the geometric mean of f_p and f_z:

$$f_{pk} = \sqrt{f_{p(1/\beta)} \cdot f_{z(1/\beta)}}$$

Substituting $f_{z(1/\beta)}$ with $f_{p(1/\beta)} / \beta_0$ then makes Equation 2.

(EQ. 2) $\qquad f_{pk} = f_{p(1/\beta)} / (\sqrt{\beta_0})$

These frequency-phase relations of Figure 7 therefore allow for two options of phase compensation: one aiming for maximum bandwidth, and a second one establishing maximum phase margin. Both options are shown in Figures 8 and 9.
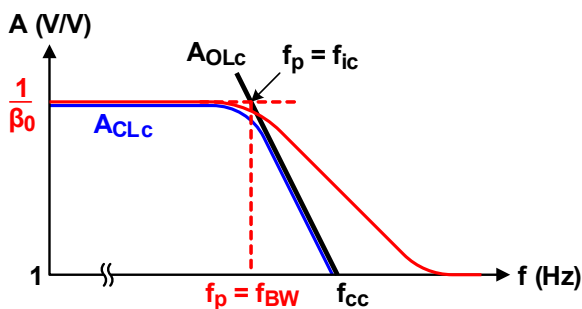
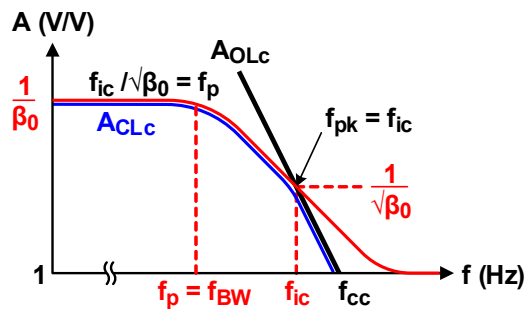

Figure 8. Compensation for Maximum Bandwidth

Figure 9. Compensation for Maximum Phase Margin

2.2.1 Phase Compensation for Maximum Bandwidth, $f_{BW\text{-}max}$

Placing the pole frequency at the intercept of $1/\beta_0$ and A_{OLc} generates the minimum phase margin of 45°. The frequency at the intercept, f_{ic}, is defined by the second order roll-off of A_{OLc} and its unit crossover frequency as: $f_{ic} = f_{cc} \cdot \sqrt{\beta_0}$. Therefore, expressing $f_p = f_{ic}$ through their detailed equations:

$$\frac{1}{2\pi C_F R_F} = f_{cc} \cdot \sqrt{\beta_0}$$, and solving for C_F, gives the capacitor value for maximum bandwidth:

(EQ. 3) $$C_{F(BW-max)} = \frac{1}{2\pi R_F f_{cc} \cdot \sqrt{\beta_0}} = \frac{1}{2\pi R_F \sqrt{f_{c1}f_{c2}\beta_0}}$$

therefore, confirming [Equation 1 on page 2](#).

The -3dB signal bandwidth is the actual pole frequency:

(EQ. 4) $$f_{BW} = \frac{1}{2\pi R_F C_{F(BW-max)}}$$

For a practical amplifier circuit with R_F = 10kΩ and R_G = 100Ω, $C_{F(BW\text{-}max)}$ = 35pF. Selecting the standard value of 39pF, results in a signal bandwidth of f_{BW} = 408kHz.

2.2.2 Phase Compensation for Maximum Phase Margin, $\Phi_{m\text{-}max}$

To ensure maximum phase margin at the intercept of $1/\beta$ and A_{OLC}, the $1/\beta$ function is moved left towards lower frequencies, until f_{pk}, the frequency at maximum phase margin, coincides with f_{ic}, the frequency at which A_{OLc} equals half the $1/\beta_0$ value in dB, or $\sqrt{1/\beta_0}$ in V/V. Here, f_{ic} is defined by the second order roll-off of A_{OLc}, the unit crossover frequency, f_{cc}, and the gain $\sqrt{1/\beta_0}$ as: $f_{ic} = f_{cc} \cdot \sqrt{\sqrt{\beta_0}} = f_{cc} \cdot \sqrt[4]{\beta_0}$.

With [Equation 4](#), making $f_{pk} = f_{ic}$ results in $f_p / \sqrt{\beta_0} = f_{cc} \cdot \sqrt[4]{\beta_0}$.

Further substitution of f_p yields $1 / (2\pi R_F C_F \sqrt{\beta_0}) = f_{cc} \cdot \sqrt[4]{\beta_0}$, and solving for C_F gives the capacitor value for maximum phase margin:

(EQ. 5) $$C_{F(\Phi max)} = \frac{1}{2\pi R_F f_{cc} \cdot \sqrt{\beta_0 \sqrt{\beta_0}}} = \frac{1}{2\pi R_F \sqrt{f_{c1}f_{c2}\beta_0 \sqrt{\beta_0}}}$$

Again, the -3dB signal bandwidth is the actual pole frequency:

(EQ. 6) $$f_{BW} = \frac{1}{2\pi R_F C_{F(\Phi max)}}$$

For a practical amplifier circuit with R_F = 10kΩ and R_G = 100Ω, $C_{F(BW\text{-}max)}$ = 111pF. Selecting the standard value of 100pF, results in a signal bandwidth of f_{BW} = 161kHz.

2.3 Bandwidth Comparison

When using the READ2302 in a normal, non-inverting amplifier configuration, its gain-bandwidth product of 6MHz limits the signal bandwidth to 60kHz, when operating at a gain of 100V/V. In strong contrast, the composite amplifier, which is designed for maximum bandwidth extends this bandwidth by a factor of almost 7 to about 408kHz. Even when optimized for maximum phase margin, the composite almost triples the bandwidth to about 160kHz. We can therefore conclude that the composite amplifier significantly increases signal bandwidth over the standard configuration.

2.4 Output Offset Voltage

In a non-inverting G = 100 amplifier application, the READ2302 rather large input offset of 6mV generates a 600mV output offset. In comparison, analysis of the composite amplifier output offset in Figure 10 defines the output voltage with:
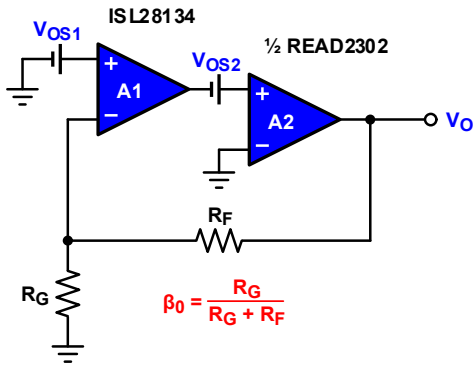
$$V_O = a_2[V_{OS2} + a_1(V_{OS1} - V_O \cdot \beta_0)]$$

Multiplying and collecting terms gives:

$$V_O = \frac{V_{OS1}}{\frac{1}{a_1 a_2} + \beta} + \frac{V_{OS2}}{\frac{1}{a_2} + a_1\beta}$$

Because the DC open-loop gains of both op-amps are much larger than 1 (a_1, a_2 >> 1), the output voltage due to offset is:

$$V_O = \frac{1}{\beta}\left(V_{OS1} + \frac{V_{OS2}}{a_1}\right)$$

Figure 10. Output Offset Evaluation Model

In practical terms, V_{OS1} = ±4µV, V_{OS2} = ±6mV, a_1 = 5·10^8 V/V (174dB), and 1/β = 101V/V, therefore generating an output offset of about only 400µV at a gain of 100. This shows that in the composite design the DC shortcomings of the high-speed amplifier, A2, are reduced by the open-loop gain of the precision amplifier, A1, which is beneficial in high-gain applications.

2.5 Gain Accuracy and Gain Error

The closed-loop gain, A_{CL}, can be expressed through the product of the ideal gain, 1/β, and the gain accuracy, 1/(1+1/T).

$$A_{CL} = \frac{1}{\beta} \cdot \frac{1}{1 + 1/T} \quad \text{with } T = A_{OL}\beta$$

The gain error is defined as the difference between ideal and actual gain, referred to the ideal gain:

$$E_G = \frac{1/\beta - A_{CL}}{1/\beta} = 1 - \frac{1}{1 + 1/T}$$

Because both functions contain loop gain T, and therefore A_{OL}, in their denominators, gain accuracy and gain error are frequency dependent. Rather than calculating the complex equations for a complex and composite A_{OL}, it is easier to approximate the gain error if loop-gain T >> 1. The following steps simplify the gain error expression to:

$$E_G = 1 - \frac{1}{1 + 1/T} = 1 - \frac{T}{1 + T} = \frac{1 + T - T}{1 + T} = \frac{1}{1 + T}$$

Then for T >> 1 follows: $E_G \approx \frac{1}{T}$ and $E_G(\%) \approx \frac{100}{T}$

In a Bode plot, loop-gain is the difference in dB between the open-loop gain and the ideal gain: T(dB) = A_{OL}(dB) - 1/β(dB). Therefore, to determine the bandwidth for 1% gain error, we identify the frequency where the difference between A_{OL} and 1/β is 40dB (100V/V).

This procedure is applied to A_{OL2} and A_{OLc} in Figure 11 on page 7. It clearly shows that the large open-loop gain of the composite amplifier ensures gain errors of ≤1% up to 47kHz, which is almost 80 times the 600Hz range of the non-inverting amplifier using a single READ2303 op-amp.
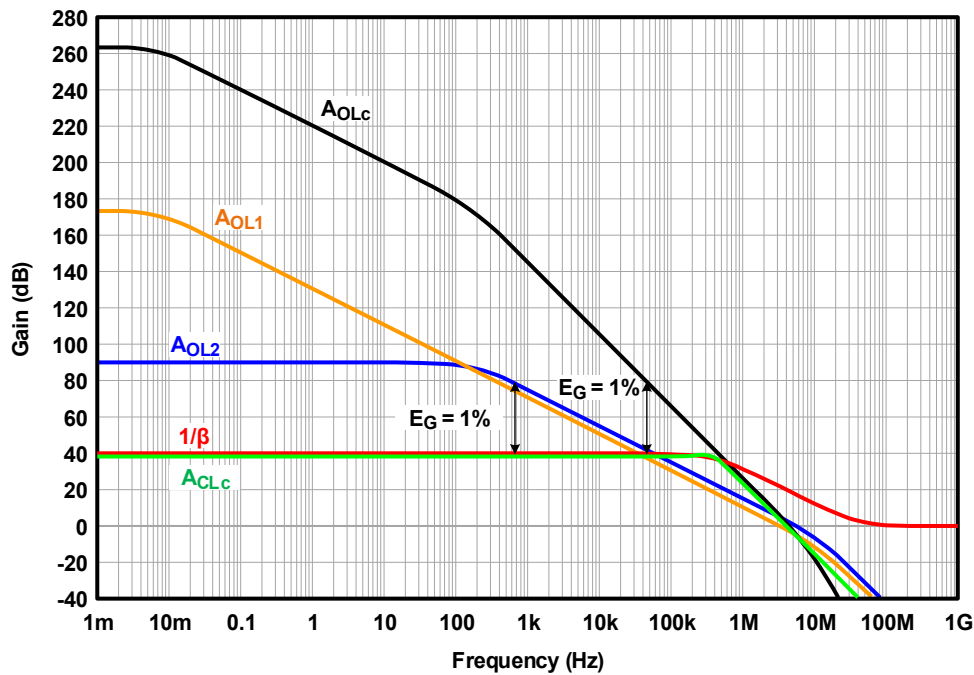
Figure 11. Comparison of 1% Gain Error Bandwidths

3. Conclusion

While there is a variety of composite amplifier designs using different stabilizing techniques, the design discussed here uses feedback phase compensation. It is one of the simpler compensation techniques and mainly applied in high-gain applications.

Composite amplifiers combine the DC performance of a precision op-amp with the AC performance of a fast-slewing or high-speed op-amp.

4. Revision History

Rev.	Date	Description
1.00	Oct.2.19	Initial release

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