Feedback Basics

- Stability
- Feedback concept
- Feedback in emitter follower
- One-pole feedback and root locus
- Frequency dependent feedback and root locus
- Gain and phase margins
- Conditions for closed loop stability
- Frequency compensation
Circuit/System Stability

1. Simply put, a linear system such as an electronic circuits is “stable” if it has no poles (denominator roots or zeros) in the right-half of the complex s-plane.

2. Right-half plane poles correspond to growing exponential terms in the solution of the system/circuit differential equation.

3. To determine stability, we need to check for RH plane poles. This requires either factoring the denominator polynomial or employing a number of other techniques developed to assist in the pencil and paper design of linear systems.
Operational Amplifier From Another View

\[ v_o = A \left( v_i - \Gamma_F v_o \right) \]
\[ (1 + A \Gamma_F) v_o = A v_i \]

\[ v_o = A \left( v_i - v_f \right) \]
\[ v_f = \frac{R_2}{R_1 + R_2} v_o = \Gamma_F v_o \]

\[ v_o = \frac{A}{1 + A \Gamma_F} v_i \]

\[ V_i \rightarrow A \rightarrow V_o \]
\[ V_f \rightarrow \Gamma_F \]

classic feedback structure
Feedback Block Diagram Point of View

Fundamental block diagram

We can split the ideal op amp and resistive divider into 2 separate blocks because they do not interact with, or “load,” each other.

1. The op amp output is an ideal voltage source.
2. The op amp input draws no current.

Fundamental feedback equation

\[ V_o = \frac{A}{1 + A \Gamma_F} V_i \]
Comments on the Feedback Equation

The quantity $A\Gamma_F$ loop-gain and $G = \text{closed-loop gain}$.

The quantity $A = \text{open-loop gain}$ and $\Gamma_F = \text{feedback gain}$.

If the loop gain is much greater than one, i.e.,
$$A\Gamma_F \gg 1$$
(and the system is stable – a topic to be discussed later!) the closed-loop gain approximates $1/\Gamma_F$ and is independent of “$A$”!

$$G = \frac{V_o}{V_i} \approx \frac{1}{\Gamma_F}$$
Feedback in the Emitter Follower

Emitter current equation:

\[ i_e = \frac{1}{R_s \beta_T + 1 + r_e + R_E} v_i \]

Note: \( \beta_T = \beta \), i.e. the BJT “beta”.

Create an “artificial” feedback equation, multiply by \((\beta_T + 1)/R_s\):

\[ i_e = \frac{\beta_T + 1}{R_s} \frac{v_i}{1 + \frac{\beta_T + 1}{R_s}(r_e + R_E)} \]

\[ v_i = \frac{A}{1 + A\Gamma_F} v_i \]
Emitter Follower  Emitter Current

The forward gain term, $A$:

$$A = \frac{\beta_T + 1}{R_s}$$

The feedback term, $\Gamma_F$:

$$\Gamma_F = r_e + R_E$$

If the “loop gain” is large, $A\Gamma_F \gg 1$

$$i_e \approx \frac{1}{\Gamma_F} \frac{1}{r_e + R_E} v_i$$

Dependence on $\beta_T$ is eliminated.

\[ \begin{align*}
    i_e &= \frac{\beta_T + 1}{R_s} \left(1 + \frac{\beta_T + 1}{R_s} (r_e + R_E)\right) \\
    v_i &= \frac{A}{1 + A\Gamma_F} v_i
\end{align*} \]
Loop Gain Sensitivities

For:

\[ G = \frac{A}{1 + A \Gamma_F} \]

Forward gain:

\[
\frac{dG}{dA} = \frac{1}{(1 + A \Gamma_F)^2}
\]

\[
\frac{dG}{G} = \frac{1}{(1 + A \Gamma_F)^2} \frac{1 + A \Gamma_F}{A} dA
\]

\[
= \frac{1}{(1 + A \Gamma_F)} \frac{dA}{A} \rightarrow 0 \quad A \Gamma_F \rightarrow \infty
\]

Feedback gain:

\[
\frac{dG}{d\beta_F} = \frac{-A^2}{(1 + A \Gamma_F)^2}
\]

\[
\frac{dG}{G} = \frac{-A^2}{(1 + A \Gamma_F)^2} \frac{1 + A \Gamma_F}{A} d\Gamma_F
\]

\[
= \frac{-A \Gamma_F}{(1 + A \Gamma_F)} \frac{d\Gamma_F}{\Gamma_F} \rightarrow -\frac{d\Gamma_F}{\Gamma_F} \quad A \Gamma_F \rightarrow \infty
\]

High loop gain makes system insensitive to \( A \), but sensitive to \( \Gamma_F \)!
Feedback - One-pole $A(s)$

Consider the case where:

$$A(s) = \frac{a K_0}{s + a}$$

open-loop pole

Where $\Gamma_F$ and $K_0$ are positive real quantities.

Also,

$$G(s, \Gamma_F) = \frac{a K_0}{s + a} \cdot \frac{1}{1 + a \Gamma_F K_0} = \frac{a K_0}{s + a(1 + \Gamma_F K_0)}$$

closed-loop pole

stable for all $\Gamma_F$!

Root Locus Plot for $G(s, \Gamma_F)$

$s = \sigma + j\omega$

$\Gamma_F \approx 0$

$\infty \leftarrow \Gamma_F$

stable for all $\Gamma_F$!
Feedback - One-pole \( A(s) \) cont.

\[
A(s) = \frac{aK_0}{s+a}
\]

open-loop (OL)

\[
G(s) = \frac{aK_0}{s + a(1 + \Gamma_F K_0)} = \frac{N(s)}{D(s)}
\]

closed-loop (CL)

\[
GBW_{OL} = GBW_{CL} = aK_0
\]
One-pole Feedback - Root Locus

\[ A(s) = \frac{a K_0}{s+a} \]

\[ G(s) = \frac{a K_0}{s+a(1+\Gamma_F K_0)} = \frac{N(s)}{D(s)} \]

Pole of \( G(s) \): \( D(s) = s + a(1+\Gamma_F K_0) = s + a + a \Gamma_F K_0 = 0 \)

\[ s + a + a \Gamma_F K_0 = 0 \Rightarrow \frac{a \Gamma_F K_0}{s+a} = -1 = 1 \times e^{j\pi \pm 2k\pi} \]

\[ \frac{a \Gamma_F K_0}{|s+a|} = 1 \quad \text{and} \quad -\phi_{(s+a)} = \pi \pm 2k\pi \]

1. \( \phi_{(s+a)} = \pi \)
2. \( \phi_{(s_2+a)} = \pi \)
3. \( \phi_{(s_1+a)} = 0 \neq \pi \pm 2k\pi \)

\[ \infty > \Gamma_F K_0 \geq 0 \quad \text{Not allowed} \]

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Frequency-Dependent Feedback

Consider the case where the open-loop gain is:

$$A(s) = K \frac{(s+a)}{(s+b)(s+c)} \quad a < b < c$$

For convenient Bode plotting we can rewrite $A(s)$ as:

$$A(s) = K \frac{a}{bc} \left[ \frac{(s+1)}{a(s+b+1)(s+c+1)} \right] = K_0 \frac{(s+1)}{(s+b+1)(s+c+1)}$$

where $K_0 = K \frac{a}{bc}$
Frequency-Dependent Feedback

A sketch of the Bode plot would look something like:

Sketching the frequency response is “easy,” if the polynomial is factored into its roots!
Frequency-Dependent Feedback

If the frequency-dependent quantity, \( A(s) \), is embedded in a feedback loop, the poles are not so easily factored:

\[
A(s) = K_0 \frac{(s+a)}{(s+b)(s+c)}
\]

open-loop (OL)

\[
G(s) = \frac{A(s)}{1 + \Gamma_F A(s)}
\]

closed-loop (CL)

\[
G(s) = \frac{K_0 \frac{(s+a)}{(s+b)(s+c)}}{1 + \Gamma_F K_0 \frac{(s+a)}{(s+b)(s+c)}} = K_0 \frac{(s+a)}{(s+b)(s+c) + \Gamma_F K_0 (s+a)}
\]

1. zeros of \( G(s) \) = zeros of \( A(s) \) and are independent of feedback \( \Gamma_F \)
2. poles of \( G(s) \) ≠ poles of \( A(s) \) are a function of the feedback \( \Gamma_F \)
Frequency-Dependent Feedback – Root Locus

\[ G(s) = \frac{K_0(s+a)}{(s+b)(s+c)} \]

Rationalizing this expression leads to:

\[ G(s) = \frac{K_0(s+a)}{1 + \Gamma_F K_0(s+b)(s+c)} = \frac{N(s)}{D(s)} \]

The numerator is factored, but the denominator is not. We have a new quadratic \( D(s) \) for \( G(s) \). It could be factored using the quadratic formula, but we will do it another way! Factoring a polynomial is equivalent to finding its roots, i.e. the values of \( s \) that make the polynomial equal zero.

\[ D(s) = (s+b)(s+c) + \Gamma_F K_0(s+a) = 0 \]
Frequency-Dependent Feedback – Root Locus

\[(s + b)(s + c) + \Gamma_F K_0 (s + a) = 0\]

Since the poles of \(G(s)\) will not equal any of the poles of \(A(s)\), we can divide by \((s + b)(s + c)\) and obtain:

\[1 + \Gamma_F K_0 \frac{(s + a)}{(s + b)(s + c)} = 0\]

\[\Gamma_F K_0 \frac{(s + a)}{(s + b)(s + c)} = -1\]

The “loop-gain” terms, \(\Gamma_F\) and \(K_0\), are positive real numbers, so for a root, say \(s = -r_1\), to exist, the value of the frequency-dependent terms must be real and negative when evaluated at \(s = -r_1\)!
Frequency-Dependent Feedback – Root Locus

\[ \Gamma_F K_0 \frac{(s + a)}{(s + b)(s + c)} = -1 = 1 * e^{j\pi \pm 2k \pi} \text{ where } k = 0, 1, ... \]

Working with the complex numbers in polar form:

\[ \Gamma_F K_0 \frac{|s + a|}{|s + b||s + c|} = 1 \quad \text{and} \quad \phi_{(s_i + a)} - \phi_{(s_i + b)} - \phi_{(s_i + c)} = \pi \pm 2k \pi \]

we add the angles for zeros to the point “s” in the complex plane and subtract the angles for poles.
**Frequency-Dependent Feedback – Root Locus**

A sketch of the operations looks like:

\[
\begin{align*}
|s_i + c| & = \phi(s_i + c) \\
|s_i + a| & = \phi(s_i + a) \\
|s_i + b| & = \phi(s_i + b)
\end{align*}
\]

Move point \( s_i \) until the phase angle relation is satisfied.

The only values of \( s_i \) that satisfy the phase angle relationship lie on the negative \( \sigma \) - axis.

Root Locus Protocol

\[
\frac{|s_i + a|}{|s_i + b||s_i + c|} = \frac{1}{\Gamma_F K_0}
\]

\[
\phi(s_i + a) - \phi(s_i + b) - \phi(s_i + c) = \pi \pm 2k \pi
\]
Frequency-Dependent Feedback - Root Locus

Only $D(s)$ root locations where the angles with respect to open-loop pole/zero locations of $A(s)$ are odd multiples of $\pm 180^\circ$ are candidates.

\[
D(s) = (s + b)(s + c) + \Gamma_F K_0(s + a) = 0
\]

\[
\phi_{tot} = \phi_{(s+a)} - \phi_{(s+b)} - \phi_{(s+c)} = \pi \pm 2k \pi
\]

1. All roots of $D(s)$ must lie along the “green” loci.
2. Bandwidth of $G(s) >$ bandwidth of $A(s)$
3. If $s_2 = -a$ there is a zero/pole cancellation

$G(s)$ stable for all $\Gamma_F$!
Frequency-Dependent Feedback – Root Locus

Now, to find the gain $\Gamma_F K_0$ required for a given root or pole of $G(s)$, measure the distances from each critical point and form the ratio:

$$\frac{|s+a|}{|s+b||s+c|} = \frac{1}{(\Gamma_F K_0)} \Rightarrow (\Gamma_F K_0)_r = \frac{|s+b||s+c|}{|s+a|} \bigg|_{s=-s_r}$$

Where $s = -s_r$ is on the root-locus

The only allowed poles {e.g. “$s_r$”} of $G(s)$ are on the “green” loci.
Root Locus Two-pole \( A(s) \)

Note: no zero at \( s = -a \)

\[
A(s) = \frac{K_0 bc}{(s+b)(s+c)}
\]

\[
G(s) = \frac{K_0 bc}{(s+b)(s+c)+\Gamma_F K_0 bc} = \frac{N(s)}{D(s)}
\]

complex conjugate poles of \( G(s) \)

\( \infty > \Gamma_F \geq 0 \)

\( G(s) \) stable for all \( \Gamma_F \)!
Root Locus Three-pole $A(s)$

$$A(s) = \frac{K_0 bcd}{(s+b)(s+c)(s+d)}$$

$$G(s) = \frac{K_0 bcd}{(s+b)(s+c)(s+d)+\Gamma F K_0 bcd} = \frac{N(s)}{D(s)}$$

$G(s)$ NOT stable for all $\Gamma_F$!
The Root Locus Method

This graphical method for finding the roots of a polynomial is known as the root locus method. It was developed before computers were available. It is still used because it gives valuable insight into the behavior of feedback systems as the loop gain is varied. Matlab (control systems toolbox) will plot root loci.

In the frequency-dependent feedback example, we noted that increasing feedback increases the closed-loop bandwidth – i.e. the low and high frequency break points moved in opposite directions as $\Gamma_F$ increases. Higher $\Gamma_F$, as a trade-off, reduces the closed-loop mid-band gain.
Stability - Gain and Phase Margins

\[ G(j\omega) = \frac{V_o(j\omega)}{V_i(j\omega)} = \frac{A(j\omega)}{1 + A(j\omega)\Gamma_F} \]

loop-gain

\[ A(j\omega)\Gamma_F = 1 e^{j\pm k\pi} \Rightarrow \text{oscillation or instability} \]

For stable closed-loop amplifier:

\[ |A(j\omega_{180})\Gamma_F| < 1 \]

or

\[ 20 \log_{10}|A(j\omega_{180})\Gamma_F| < 0 \text{ dB} \]

or

\[ \phi(j\omega_1) = \text{arg}[A(j\omega_1)\Gamma_F] > -180^\circ \]

\[ GM = 0 \text{ dB} - 20 \log_{10}|A(j\omega_{180})\beta_F| \]

\[ GM > 0 \text{ dB} \Rightarrow \text{stable amplifier} \]

\[ PM = \phi(j\omega_1) - (-180^\circ) = \phi(j\omega_1) + 180^\circ \]

\[ PM > 0^\circ \Rightarrow \text{stable amplifier} \]
Three-poles

$$\rho = \text{damping ratio}$$

$$\rho \omega_0$$

$$G(j\omega) = \frac{v_o}{v_i} = \frac{A(j\omega)}{1 + A(j\omega) \Gamma_F}$$

Effect of PM on closed-loop gain
overshoot or peaking

- $$\rho = 0.5$$
- $$\rho = 0.707$$ (maximally flat response)
- $$\rho = 1$$
- $$\rho = 3.33$$

For acceptable overshoot: $$PM > 50^\circ$$
Frequency-Domain

Effect of PM on closed-loop gain

\[ |G(j\omega)| \text{dB} \]

- \( \rho = 0.5 \)
- \( \rho = 0.707 \) (maximally flat response)
- \( \rho = 1 \)
- \( \rho = 3.33 \)
- \(-12 \text{ dB/octave}\)

Time-Domain

For best compromise of overshoot vs. \( t_{\text{rise}} \) & \( t_{\text{settle}} \).

\[ \rho < 0.5 \quad PM < 55^\circ \]

\[ \rho > 1 \quad PM > 75^\circ \]
Alternative Stability Analysis

1. Investigating stability for a variety of feedback gains $\Gamma_F$ by constructing Bode plots for the loop-gain $A(j\omega)\Gamma_F$ can be tedious and time consuming.

2. A simpler approach involves constructing Bode plots for $A(j\omega)$ and $1/\Gamma_F$ separately.

\[
20 \log |A(j\omega)\Gamma_F| = 20 \log |A(j\omega)| - 20 \log \frac{1}{\Gamma_F}
\]

when
\[
20 \log |A(j\omega)|_{\omega=\omega_1} = 20 \log \frac{1}{\Gamma_F} \Rightarrow |A(j\omega_1)\Gamma_F| = 1
\]

RECALL: $A(j\omega)\Gamma_F \gg 1 \Rightarrow G(j\omega) = \frac{A(j\omega)}{1 + A(j\omega)\Gamma_F} \approx \frac{1}{\Gamma_F}$ closed-loop gain
Alternative Stability Analysis - cont.

\[ |A(jf)| = \frac{10^5}{(1 + jf/10^5)(1 + jf/10^6)(1 + jf/10^7)} \]

-20 dB /dec
\[ A(jf) = 20 \log |A| \Gamma_F \]

-20 dB /dec
\[ 20 \log |A(jf)| \Gamma_F \]

-60 dB /dec
\[ 20 \log \left| A(jf) \right| - 20 \log \frac{1}{\Gamma_F} \]

stable iff
\[ 20 \log 1/\Gamma_F > 60 \text{ dB} \]
or
\[ 20 \log \Gamma_F < -60 \text{ dB} \]
or
\[ \Gamma_F < 0.001 \]
Alternative Stability Analysis - cont.

“Rule of Thumb” – Closed-loop amplifier will be stable if the $20 \log \frac{1}{\Gamma_F}$ line intersects the $20 \log |A(jf)|$ curve on the -20 dB/dec segment.

Using “Rule of Thumb”:

$$\arg A(jf_1) > -135^\circ$$

$$\Rightarrow PM > 45^\circ$$
Frequency Compensation

“Rule of Thumb” – Closed-loop amplifier will be stable if the $20\log(1/\upi_F)$ line intersects the $20\log|A(j\omega)|$ curve on the $-20$ dB/dec segment.

Using “Rule of Thumb”:

$$\arg A(jf_1) > -135^\circ$$

$$\Rightarrow \quad \text{PM} > 45^\circ$$

Ex: LM 741 op amp is frequency compensated to be stable with $60^\circ$ PM for $|G(j\omega)| = 1/\upi_F = 1$.

frequency compensation – modifying the open-loop $A(s)$ so that the closed-loop $G(s)$ is stable for any desired value of $|G(j\omega)|$.

desired implementation - minimum on-chip or external components.
Compensation – What if 1st pole is shifted lower?

\[ |A(j\omega)| \]

\[ |A_{\text{comp}}(j\omega)| \]

-20 dB /dec

-60 dB /dec

-40 dB /dec

-20 dB /dec

-40 dB /dec

-60 dB /dec

\[ \frac{10^5}{(1 + j\omega/10^5)(1 + j\omega/10^6)(1 + j\omega/10^7)} \]

\[ \approx \]

\[ \frac{10^5}{(1 + j\omega/10^3)(1 + j\omega/10^6)(1 + j\omega/10^7)} \]

\[ 20\log \frac{1}{\Gamma_{F1}} = 85 \text{ dB} \]

\[ 20\log \frac{1}{\Gamma_{F2}} = 45 \text{ dB} \]
Frequency Compensation Using Miller Effect

\[ C_1 = C_\pi + C_\mu (1 + g_m R_2) \]

\[ R_1 = R_\beta \| r_\pi \]  
BJT Miller Capacitance

\[ V_o = \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{1 + s[C_1 R_1 + C_2 R_2 + C_{\text{comp}}(g_m R_1 R_2 + R_1 + R_2)] + s^2[C_1 C_2 + C_{\text{comp}}(C_1 + C_2)] R_1 R_2} \]

\[ = \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{(1 + \frac{s}{\omega_{p1c}})(1 + \frac{s}{\omega_{p2c}})} = \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{1 + s\left(\frac{1}{\omega_{p1c}} + \frac{1}{\omega_{p2c}}\right) + \frac{s^2}{\omega_{p1c}\omega_{p2c}}} \]

where \( \omega_{p2c} \gg \omega_{p1c} \)

with \( C_{\text{comp}} = 0 \)

\[ \omega_{p1} = \frac{1}{R_1 C_1} \]

\[ \omega_{p2} = \frac{1}{R_2 C_2} \]
Compensation Using Miller Effect - cont.

\[
\frac{V_o}{I_i} = \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{1 + s\left(\frac{1}{\omega_{plc}} + \frac{1}{\omega_{p2c}}\right) + \frac{s^2}{\omega_{plc} \omega_{p2c}}} \approx \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{1 + s\left(\frac{1}{\omega_{plc}}\right) + s^2\left(\frac{1}{\omega_{plc} \omega_{p2c}}\right)}
\]

and

\[
\frac{V_o}{I_i} = \frac{(sC_{\text{comp}} - g_m) R_1 R_2}{1 + s\left[C_1 R_1 C_2 R_2 C_{\text{comp}}\left(g_m R_1 R_2 + R_1 + R_2\right)\right] + s^2\left[C_1 C_2 + C_{\text{comp}}(C_1 + C_2)\right] R_1 R_2}
\]

If \(\omega_{p2c} \gg \omega_{plc}\)

\[
\omega_{plc} = \frac{1}{C_1 R_1 + C_2 R_2 + C_{\text{comp}}\left(g_m R_1 R_2 + R_1 + R_2\right)} \approx \frac{1}{C_{\text{comp}} g_m R_1 R_2}
\]

Compensation Miller Capacitance

\[
\omega_{p2c} = \frac{\omega_{plc} \omega_{p2c}}{\omega_{plc}} = \frac{C_1 R_1 + C_2 R_2 + C_{\text{comp}}\left(g_m R_1 R_2 + R_1 + R_2\right)}{[C_1 C_2 + C_{\text{comp}}(C_1 + C_2)] R_1 R_2} \approx \frac{C_{\text{comp}} g_m}{C_1 C_2 + C_{\text{comp}}(C_1 + C_2)} \approx \frac{g_m}{C_1 + C_2}
\]

If \(C_{\text{comp}} \gg C_2\)
Compensation Using Miller Effect - cont.

\[ f_{p1c} = \frac{\omega_{p1c}}{2\pi} \approx \frac{1}{2\pi \left[ C_{comp} g_m R_1 R_2 \right]} \]
\[ f_{p2c} = \frac{\omega_{p2c}}{2\pi} \approx \frac{g_m}{2\pi \left[ C_1 + C_2 \right]} \]

\[ A(jf) = \frac{10^5}{(1 + \frac{jf}{10^5})(1 + \frac{jf}{10^6})(1 + \frac{jf}{10^7})} = \frac{10^5}{(1 + \frac{jf}{f_{p1c}})(1 + \frac{jf}{f_{p2c}})(1 + \frac{jf}{f_{p3c}})} \]

Using Miller effect compensation: \( f_{p1} \rightarrow f_{p1c} \ll f_{p1} \) and \( f_{p2} \rightarrow f_{p2c} \gg f_{p2} \) (pole-splitting)

Also \( f_{p3} \) is unaffected by the compensation \( \Rightarrow f_{p3c} = f_{p3} \)

\[ A_{comp}(jf) = \frac{10^5}{(1 + \frac{jf}{10^2})(1 + \frac{jf}{f_{p2c}})(1 + \frac{jf}{10^7})} = \frac{10^5}{(1 + \frac{jf}{f_{p1c}})(1 + \frac{jf}{f_{p2c}})(1 + \frac{jf}{f_{p3c}})} \]
**Miller Compensation Example**

Given:

\[
A(jf) = \frac{10^5}{(1 + jf/10^5)(1 + jf/10^6)(1 + jf/10^7)} = \frac{10^5}{(1 + \frac{jf}{f_{p1}})(1 + \frac{jf}{f_{p2}})(1 + \frac{jf}{f_{p3}})}
\]

where \( C_1 = 100 \text{ pF}, C_2 = 5 \text{ pF}, g_m = 40 \text{ mS}, R_1 = 100/2\pi \text{ k} \Omega \) and \( R_2 = 100/\pi \text{ k} \Omega \)

Design Objective: determine \( C_{\text{comp}} \) s.t. \( f_{p1c} = 10^2 \text{ Hz} \)

\[
A_{\text{comp}}(jf) = \frac{10^5}{(1 + jf/10^5)(1 + jf/f_{p2c})(1 + jf/10^7)} = \frac{10^5}{(1 + \frac{jf}{f_{p1c}})(1 + \frac{jf}{f_{p2c}})(1 + \frac{jf}{f_{p3c}})}
\]

\[
f_{p1c} = 100 \text{ Hz} \approx \frac{1}{2\pi [C_{\text{comp}} g_m R_1 R_2]} \Rightarrow C_{\text{comp}} = 78.5 \text{ pF}
\]

\[
f_{p2c} = \frac{g_m}{2\pi (C_1 + C_2)} \approx 60 \text{ MHz}
\]
Compensation Using Miller Effect - cont.

\[ A(jf) = \frac{10^5}{(1 + jf/10^5)(1 + jf/10^6)(1 + jf/10^7)} \]

\[ A_{comp}(jf) \approx \frac{10^5}{(1 + jf/10^2)(1 + jf/6*10^7)(1 + jf/10^7)} \]

PM > 45° for \[ |G(jf)| = \frac{1}{\Gamma_F} \geq 1 \]
Summary

Feedback has many desirable features, but it can create unexpected - undesired results if the full frequency-dependent nature (phase-shift with frequency) of the feedback circuit is not taken into account.

We next will use feedback to convert a well-behaved stable circuit into an oscillator. Sometimes, because of parasitics, feedback in amplifiers turns them into unexpected oscillators.

The root-locus method was used to help us understand how feedback can create the conditions for oscillation and instability.