EE 435

Lecture 19

- Moving the RHP pole to LHP in Miller Compensated Amplifier
- Breaking the Loop for Loop Gain Analysis
- Other methods of gain enhancement
Basic Two-Stage Op Amp

\[
A_{FB}(s) \approx \frac{g_{md}(g_{m0} - sC_C)}{s^2C_CC_L + sC_C(g_{m0} - \beta g_{md}) + \beta g_{md}g_{m0}}
\]

Right Half-Plane Zero Limits Performance

- Why does the RHP zero limit performance?
- Can anything be done about this problem?
- Why is this not 3\textsuperscript{rd} order since there are 3 caps?
Why does the RHP zero limit performance?

- accumulate phase shift and slow gain drop with RHP zeros
- effects are dramatic

In this example:
- $p_1 = -1$, $p_2 = -1000$, $z_x = \{\text{none}, +250\}$

- accumulate phase shift and slow gain drop with RHP zeros
- effects are dramatic
Why does the RHP zero limit performance?

- accumulate phase shift and slow gain drop with RHP zeros
- loose phase shift and slow gain drop with LHP zeros
- effects are dramatic

In this example:

- \( p_1 = 1, p_2 = 1000, z_x = \{\text{none}, 250, -250\} \)
Two-stage amplifier
(with RHP Zero Compensation)

What causes the Miller compensation capacitor to create a RHP zero?

At low frequencies, $V_{OUT}/V_d$ is positive but at high frequencies it becomes negative.

Alternately, $C_C$ provides a feed-forward inverting signal from the input to the first stage output which also becomes the second stage output.
Two-stage amplifier
(with RHP Zero Compensation)

What can be done to remove the RHP zero?

\[ A_V = \left( A_0 p_1 p_2 \right) \frac{1}{(s+p_1)(s+p_2)} \]

with Miller Compensation

\[ A_V = \left( A_0 \frac{p_1 p_2}{z} \right) \frac{-s+z}{(s+p_1)(s+p_2)} \]

Alternately, \( C_C \) provides a feed-forward inverting signal from the input to the first stage output which also becomes the second stage output

Break the feed-forward path from the output of the first stage to the output of the second stage at high frequencies
Two-stage amplifier with LHP Zero Compensation

Right Half-Plane Zero Limits Performance

Zero can be moved to Left Half-Plane

$R_C$ realized with single triode region device
Two-stage amplifier with LHP Zero Compensation

\[ A(s) = \frac{g_{md} \left( g_{m5} + sC_c \left[ \frac{g_{m5}}{g_c} - 1 \right] \right)}{s^2 C_c C_L + sC_c g_{m5} + g_{oo} g_{od}} \]

\[ z_1 = \frac{-g_{m5}}{C_c \left[ \frac{g_{m5}}{g_c} - 1 \right]} \]

\( z_1 \) location can be programmed by \( R_C \)

If \( g_c > g_{m5} \), \( z_1 \) in RHP and if \( g_c < g_{m5} \), \( z_1 \) in LHP

\( R_C \) has almost no effect on \( p_1 \) and \( p_2 \)
Two-stage amplifier with LHP Zero Compensation

\[ A(s) = \frac{g_{md}\left(g_{m5} + sC_c \left[ \frac{g_{m5}}{g_c} - 1 \right] \right)}{s^2C_c\,C_L + sC_c\,g_{m5} + g_{oo}\,g_{od}} \]

\[ z_1 = \frac{-g_{m5}}{C_c \left[ \frac{g_{m5}}{g_c} - 1 \right]} \]

\[ p_1 = -\frac{g_{o1} + g_{o5}}{C_c \left( \frac{g_{m5}}{g_{o5} + g_{o6}} \right)} \]

\[ p_2 = -\frac{g_{m5}}{C_L} \]

where should \( z_1 \) be placed?
Two-stage amplifier with LHP Zero Compensation

where should $z_1$ be placed?

$$z_1 = \frac{-g_{m5}}{C_C \left[ \frac{g_{m5}}{g_C} - 1 \right]}$$

$$p_1 = -\frac{g_{o1} + g_{o5}}{C_C \left( \frac{g_{m5}}{g_{o5} + g_{o6}} \right)}$$

$$p_2 = -\frac{g_{m5}}{C_L}$$

Would make situation worse (because ratio between two dominant poles would be reduced!)

$\times$ $\times$ $z_1$ $\circ$
$p_2$ $p_1$ $z_1$

$\times$ $\times$ $\circ$
$p_2$ $p_1$ $z_1$

$\times$ $\circ$ $\times$
$p_2$ $z_1$ $p_1$

$\times$ $\times$
$p_2$ $p_1$
Two-stage amplifier with LHP Zero Compensation

where should $z_1$ be placed? Would make situation worse (because ratio between two dominant poles would be reduced!)

Other parasitic poles, at higher frequencies are present and not too much larger than $p_2$!
Two-stage amplifier with LHP Zero Compensation

\[
z_1 = \frac{-g_{m5}}{C_c \left[ \frac{g_{m5}}{g_c} - 1 \right]}
\]

- \(z_1\) often used to cancel \(p_2\)
- Can reduce size of required compensation capacitor
  - a) eliminates RHP zero
  - b) increases spread between \(p_1\) and \(p_3\)
- Improves phase margin
- Design formulations easily extend to this structure
Two-stage amplifier with LHP Zero Compensation

Analytical formulation for compensation requirements not easy to obtain  
(must consider at least 3rd-order poles and both T(s) and poles not  
mathematically tractable)  

$C_C \left[ \frac{-g_{m5}}{g_C - 1} \right]$  

$z_1 = \frac{-g_{m5}}{C_C \left[ \frac{g_{m5}}{g_C} - 1 \right]}$  

$C_C$ often chosen to meet phase margin (or settling/overshoot) requirements  
after all other degrees of freedom used with computer simulation from magnitude  
and phase plots
Basic Two-Stage Op Amp with LHP zero

8 Degrees of Freedom
\[ \{ P, \theta, V_{EB1}, V_{EB3}, V_{EB5}, V_{EB6}, V_{EB7}, R_C, C_C \} \]
1 constraint (phase margin)

with zero cancellation of \( p_2 \)

7 Degrees of Freedom
\[ \{ P, \theta, V_{EB1}, V_{EB3}, V_{EB5}, V_{EB6}, V_{EB7}, R_C, C_C \} \]
2 constraints (phase margin), \( z_1 = p_2 = \frac{-g_{m5}}{C_C g_{m5}} \left( \frac{g_{m5}}{g_C} - 1 \right) \)
**Basic Two-Stage Op Amp with LHP zero**

with zero cancellation of $p_2$

7 Degrees of Freedom

\[
\{ P, \theta, V_{EB1}, V_{EB3}, V_{EB5}, V_{EB6}, V_{EB7}, R_C, C_C \}
\]

2 constraints (phase margin), $z_1 = p_2 = \frac{-g_{m5}}{C_C \left[ \frac{g_{m5}}{g_C} - 1 \right]}

**Design Flow:**

1. Ignore $R_C$ and design as if RHP zero is present
2. Pick $R_C$ to cancel $p_2$
3. Adjust $p_1$ (i.e. change/reduce $C_C$) to achieve desired phase margin
Basic Two-Stage Op Amp with LHP zero

Realization of $R_C$

$$R_C = \frac{L}{\mu C_{OX} W V_{EB}}$$

Transistors in triode region

Very little current will flow through transistors (and no dc current)

$V_{DD}$ or GND often used for $V_{XX}$ or $V_{YY}$

$V_{BQ}$ well-established since it determines $I_{Q5}$

Using an actual resistor not a good idea (will not track $gm5$ over process and temp)
Two-Stage Amplifiers

Practical Considerations

• Loop Gain
  – Loading of A and $\beta$ networks
  – Breaking the Loop (with appropriate terminations)
  – Biasing of Loop
  – Simulation of Loop Gain

• Open-loop gain simulations
  – Systematic Offset
  – Embedding in closed loop
Loop Gain is a Critical Concept for Compensation of Feedback Amplifiers

- Sometimes it is not obvious where the actual loop gain is at in a feedback circuit

- The A amplifier often causes some loading of the β amplifier and the β amplifier often causes some loading of the A amplifier

- Often try to “break the loop” to simulate or even calculate the loop gain or the gains A and β

- If the loop is not broken correctly or the correct loading effects on both the A amplifier and β amplifier are not included, errors in calculating loop gain can be substantial and conclusions about compensation can be with significant error
Loop Gain - \( A\beta \)

(for voltage-series feedback configuration)

The loop is often broken on the circuit schematic to determine the loop gain.
Loop Gain - $A\beta$

Breaking the loop to obtain the loop gain

$$\beta = \frac{R_1}{R_1+R_2}$$

Note terminations where the loop is broken – open and short

$$v_{LP} = v_{IN}A_V\frac{R_1}{R_1+R_2}$$

$$\frac{v_{LP}}{v_{IN}} = A_{LOOP} = A\beta$$
But what if the amplifier is not ideal?

For the feedback amplifier:

\[
\begin{aligned}
\mathbf{v}_{\text{OUT}} (G_O + G_L + G_2) &= v_x G_2 + A_V v_1 G_O \\
v_x (G_1 + G_2 + G_{IN}) &= \mathbf{v}_{\text{OUT}} G_2 + v_{IN} G_{IN} \\
v_{IN} &= v_1 + v_x
\end{aligned}
\]

Solving, we obtain

\[
A_{FB} = \frac{\mathbf{v}_{\text{OUT}}}{\mathbf{v}_{IN}} = \frac{G_{IN} G_2 + A_V \left( G_O \left[ G_1 + G_2 \right] \right)}{(G_O + G_L) \left[ G_1 + G_2 + G_{IN} \right] + G_2 \left( G_1 + G_{IN} \right) + A_V G_2 G_O}
\]

What is the Loop Gain? Needed to obtain the Phase Margin!
But what if the amplifier is not ideal?

\[ A \beta \]

What is the Loop Gain? Needed to obtain the Phase Margin!

\[
A_{FB} = \frac{v_{OUT}}{v_{IN}} = \frac{G_{IN}G_2 + A_V (G_O [G_1 + G_2])}{(G_O + G_L) [G_1 + G_2 + G_{IN}] + G_2 (G_1 + G_{IN}) + A_V G_2 G_O}
\]

Remember: \( A_{FB} = \frac{F_1(s)}{1 + A\beta} \) Characteristic Polynomial Determined by \( D(s) = 1 + A\beta \)

Whatever is added to “1” in the denominator is the loop gain
Loop Gain - $\mathbf{A\beta}$

But what if the amplifier is not ideal?

The Loop Gain is

$$A_{LP} = A_V \left[ \frac{G_2G_O}{\left(G_O + G_L\right)[G_1+G_2+G_{IN}]+G_2 \left(G_1+G_{IN}\right)} \right]$$
Loop Gain - $A\beta$

But what if the amplifier is not ideal?

The Loop Gain is

$$A_{LP} = A_V \left[ \frac{G_2 G_O}{(G_O + G_L)[G_1 + G_2 + G_{IN}] + G_2 (G_1 + G_{IN})} \right]$$

This can be rewritten as

$$A_{LP} = \left( A_V \left[ \frac{G_O (G_1 + G_2)}{(G_O + G_L)[G_1 + G_2 + G_{IN}] + G_2 (G_1 + G_{IN})} \right] \right) \left[ \frac{G_2}{G_1 + G_2} \right]$$

This is of the form

$$A_{LP} = (A_{VL}) \left[ \frac{G_2}{G_1 + G_2} \right]$$

where $A_{VL}$ is the open loop gain including loading of the load and $\beta$ network!
Loop Gain - $A\beta$

But what if the amplifier is not ideal?

The Loop Gain is

$$A_{LP} = A_{V} \left[ \frac{G_2 G_O}{(G_O + G_L)[G_1 + G_2 + G_{IN}] + G_2 (G_1 + G_{IN})} \right]$$

$$A_{VL} = A_{V} \left[ \frac{G_O (G_1 + G_2)}{(G_O + G_L)[G_1 + G_2 + G_{IN}] + G_2 (G_1 + G_{IN})} \right]$$

Note that $A_{VL}$ is affected by both its own input and output impedance and that of the $\beta$ network

This is a really “messy” expression

Any “breaking” of the loop that does not result in this expression will result in some errors though they may be small
Loop Gain - $A\beta$

(for voltage-series feedback configuration)

But what if the amplifier is not ideal?

- Most authors talk about breaking the loop to determine the loop gain $A\beta$
- In many if not most applications, breaking the loop will alter the loading of either the $A$ amplifier or the $\beta$ amplifier or both
- Should break the loop in such a way that the loading effects of $A$ and $\beta$ are approximately included
- Consequently, breaking the loop will often alter the actual loop gain a little
- Q-point must not be altered when breaking the loop
- In most structures, broken loop only gives an approximation to actual loop gain
- Sometimes challenging to break loop in appropriate way

\[ V_{\text{IN}} \Rightarrow A(v_1) \Rightarrow v_{\text{OUT}} \]

\[ V_{\text{OUT}} \]

Breaking the Loop

\[ v_1 \]

\[ A_v v_1 \]

\[ R_0 \]

\[ R_L \]

\[ R_1 \]

\[ R_2 \]
Loop Gain - $A\beta$

(for voltage-series feedback configuration)

But what if the amplifier is not ideal?

Breaking the Loop

Standard Small-Signal Loop Gain Circuit

Standard Loop Gain Circuit including Biasing
Loop Gain - $A\beta$

(for voltage-series feedback configuration)

But what if the amplifier is not ideal?

$$A_{LP} = A_V \left[ \frac{G_2 G_O}{(G_O + G_L)[G_1+G_2]+G_2(G_1)} \right]$$

Loop Gain from Terminated Loop

$$A_{LP} = A_V \left[ \frac{G_2 G_O}{(G_O + G_L)[G_1+G_2+G_{IN}]+G_2(G_1+G_{IN})} \right]$$

Real Loop Gain
Loop Gain - $A\beta$
(for voltage-series feedback configuration)

But what if the amplifier is not ideal?

Better Standard Small-Signal Loop Gain Circuit

Better Loop Gain Circuit including Biasing
Loop Gain - $A\beta$
for four basic amplifier types

voltage-series feedback

Feedback Amplifier

current-series feedback

Loop Gain Amplifier
Loop Gain - $A\beta$
for four basic amplifier types

Current-shunt feedback

Feedback Amplifier

Voltage-shunt feedback

Loop Gain Amplifier
Open-loop gain simulations

- Must first adjust $V_{XX}$ to trim out any systematic offset
- Always verify all devices are operating in the desired region of operation
- If an ac input is applied to $V_{IN}$, no information about linearity or signal swing will be obtained
- If any changes in amplifier circuit are made, $V_{XX}$ must be trimmed again
- Include any loading including loading of beta network (with proper termination)
Open-loop gain simulations
(with a closed-loop test bench)

- Stabilizes the effect of the systematic offset voltage
- Test β network may not be related to actual β at all
- Loading of actual β network included in “Load with Termination”
- Input and output buffers eliminate any loading effects of the test β network
- $A_V$ must be calculated from measurements of $V_{OUT}$ and $V_A$
- Test β network must be chosen so overall network is stable

Why not just use actual β network for test β network?

Actual β network may even be unstable before compensation is complete
Why not just simulate the frequency response of the actual feedback amplifier and look at the magnitude of the gain to see if that is what we want?

Isn’t that what we really want anyway?

If the amplifier is overly underdamped or oscillatory, won’t that show up anyway?

Remember, the small-signal analysis will have the same magnitude response for minimum-phase and non-minimum phase systems!
Tools for Helping with Amplifier Compensation

Numerous tools but generally require analytical models

Based upon testbenches using actual circuit schematics (though behavioral descriptions can be included)

STB (in Spectre)

The Spectre STB analysis provides a way to simulate continuous time loop gain, phase margin and gain margin without breaking the feedback loop.

In the stability analysis you are required to choose a probe from which the loop gain measurements are taken. The probes, described below, can be found in the analogLib

Many sources on line discussing STB analysis.
(One youtube video is listed below (without assessment of either validity or quality)

https://youtu.be/L8wJhENPZNc
Other Methods of Gain Enhancement

Methods used so far:

- Increasing the output impedance of the amplifier: cascode, folded cascode, regulated cascode

- Increasing the transconductance (current mirror op amp) but it didn’t really help because the output conductance increased proportionally

- Cascading gives a multiplicative gain effect (thousands of architectures but compensation is essential) practically limited to a two-level cascade because of too much phase accumulation
Other Methods of Gain Enhancement

Recall:

\[ A_{v0} = \frac{-g_{MQC}}{g_{OQC} + g_{OCC}} \]

\[ GB = \frac{g_{mQC}}{C_L} \]

Two Strategies:
1. Decrease denominator of \( A_{v0} \)
2. Increase numerator of \( A_{v0} \)

Previous approaches focused on decreasing denominator or increasing numerator with current mirror

Consider now increasing numerator with excitation
Other Methods of Gain Enhancement

Consider now increasing numerator by changing the excitation

\[
A_{V_0} = \frac{-\left(g_{mQC} + g_{mCC}\right)}{g_{OQC} + g_{OCC}}
\]

\[
GB = \frac{g_{mQC} + g_{mCC}}{C_L}
\]
$g_{meq}$ Enhancement with Driven Counterpart Circuit

\[ A_{v0} = \frac{g_{m1} + g_{m3}}{g_{o1} + g_{o3}} \]

\[ GB = \frac{g_{m1} + g_{m3}}{C_L} \]

Both gain and GB enhancement
$g_{meq}$ Enhancement with Driven Counterpart Circuit

Needs CMFB Circuit to $V_{B1}$ or $V_{B2}$
$g_{meq}$ Enhancement with Driven Counterpart Circuit

$$A_{V_0} = \frac{1}{2} \frac{g_{m1} + g_{m3}}{g_{o1} + g_{o3}}$$

$$GB = \frac{1}{2} \frac{g_{m1} + g_{m3}}{C_L} \left( \frac{1}{V_{EB1}} + \frac{1}{V_{EB3}} \right)$$

$GB$ and $A_{V_0}$ improved!
Other Methods of Gain Enhancement

Increasing the output impedance of the amplifier
cascode, folded cascode, regulated cascode

Increasing the transconductance
(current mirror op amp) but it didn’t really help because
the output conductance increased proportionally

Driving the counterpart circuit does offer some improvements in gain

Cascading gives a multiplicative gain effect
(thousands of architectures but compensation is essential)
practically limited to a two-level cascade because of too much
phase accumulation
Other Methods of Gain Enhancement

\[ A_{V0} = \frac{-g_{MQC}}{g_{OQC} + g_{OCC}} \]

Two Strategies:
1. Decrease denominator of \( A_{V0} \)
2. Increase numerator of \( A_{V0} \)

Consider again decreasing the denominator

\[ A_{V0} = \frac{-g_{MQC}}{g_{OQC} + g_{OCC} - g_{OX}} \]

Is it possible to come up with circuits that will provide a subtraction of conductance in the denominator?
Other Methods of Gain Enhancement

\[ V_{OUT}(s) = \frac{-g_{MQC}}{sC_L + g_{OQC} + g_{OCC} - g_{MCC}} \]

\[ A_V(s) = \frac{-g_{mF1}}{sC_L + g_{OQC} + g_{OCC} - g_{MCC}} \]

\[ A_V(s) = \frac{-g_{mF1}}{sC_L + g_{OQC} + g_{OCC} - g_{MCC}} \]
Gain Enhancement with Regenerative Feedback

The gain can be made arbitrarily large by selecting $g_{mP1}$ appropriately.

The GB does not degrade!

But if not careful, maybe $g_{mP1}$ will get too large!
Gain Enhancement with Regenerative Feedback

$$A_{V0} = \frac{-g_{mF1}}{sC_L + g_{oF1} + g_{oP1} - g_{mP1}}$$

$$A_{V0} = \frac{g_{mF1}}{g_{oF1} + g_{oP1} - g_{mP1}}$$

$$BW = \frac{g_{oF1} + g_{oP1} - g_{mP1}}{C_L}$$

$$GB = \frac{g_{mF1}}{C_L}$$

The gain can be made arbitrarily large by selecting $g_{mP1}$ appropriately.

The GB does not degrade!

But - can we easily build circuits with this property?
Gain Enhancement with Regenerative Feedback

But - can we easily build circuits with this property?

But – the inverting amplifier may be more difficult to build than the op amp itself!
Gain Enhancement with Regenerative Feedback

But - can we easily build circuits with this property?

But – the inverting amplifier may be more difficult to build than the op amp itself!

YES – simply by cross-coupling the outputs in a fully differential structure
End of Lecture 19