# Analog IC Design

## Lecture 15
## Negative Feedback

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# Outline

- Recapping previous key results
- General feedback system
- Loop gain
- Why negative feedback?
- Stability of feedback system
- Root locus and Bode plot
- Phase and gain margin

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# MOSFET in Saturation

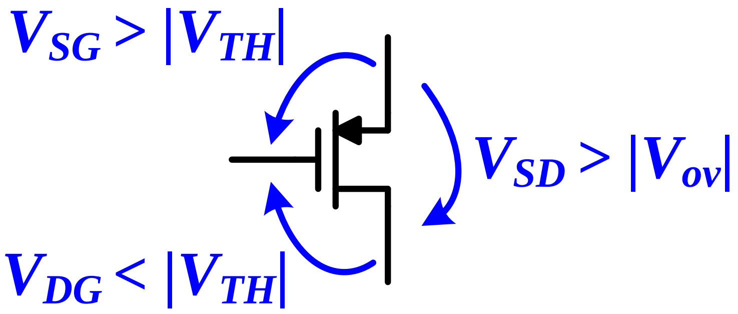
- The channel is pinched off if the difference between the gate and drain voltages is not sufficient to create an inversion layer

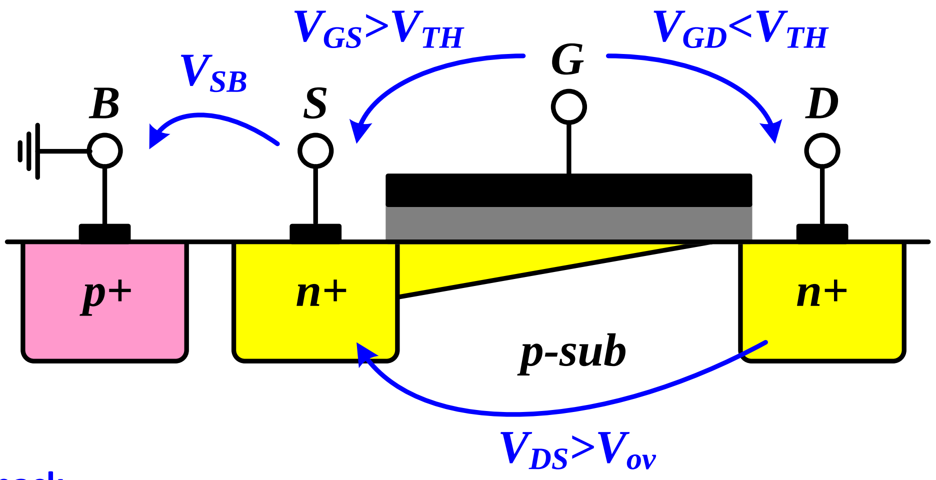
$$V_{GD} \leq V_{TH} \quad OR \quad V_{DS} \geq V_{ov}$$

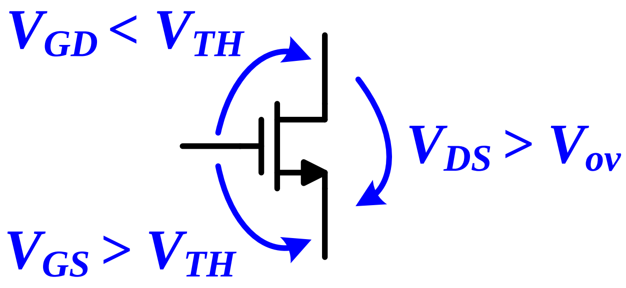
- Square-law (long channel MOS)

$$I_D = \frac{\mu_n C_{ox}}{2}\frac{W}{L} \cdot V_{ov}^2(1 + \lambda V_{DS})$$

$$V_{SB} \uparrow \;\; \Rightarrow \;\; V_{TH} \uparrow$$



# Regions of Operation Summary

**OFF (Subthreshold)**

$V_{GS} < V_{TH}$

**ON**

$V_{GS} > V_{TH}$

**Triode**

$V_{DS} < V_{ov}$

Or

$V_{GD} > V_{TH}$

$$I_D = \mu C_{ox} \frac{W}{L}\left(V_{ov}V_{DS} - \frac{V_{DS}^2}{2}\right)$$

**Pinch-Off (Saturation)**

$V_{DS} \geq V_{ov}$

Or

$V_{GD} \leq V_{TH}$

$$I_D = \frac{\mu C_{ox}}{2}\frac{W}{L}V_{ov}^2(1 + \lambda V_{DS})$$

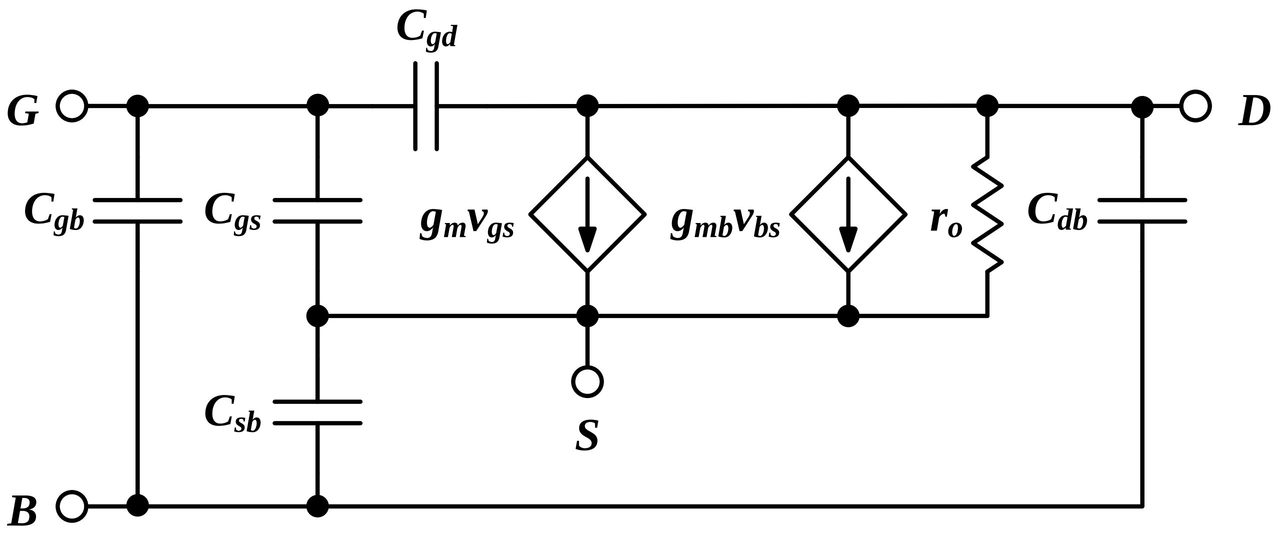
# High Frequency Small Signal Model

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \mu C_{ox} \frac{W}{L} V_{ov} = \sqrt{\mu C_{ox} \frac{W}{L} \cdot 2I_D} = \frac{2I_D}{V_{ov}}$$

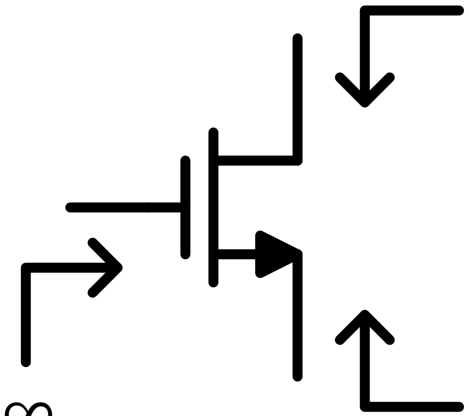
$$g_{mb} = \eta g_m \qquad \eta \approx 0.1 - 0.25$$

$$r_o = \frac{1}{\partial I_D / \partial V_{DS}} = \frac{V_A}{I_D} = \frac{1}{\lambda I_D} \qquad V_A \propto L \leftrightarrow \lambda \propto \frac{1}{L} \qquad V_{DS} \uparrow V_A \uparrow$$

$$C_{gb} \approx 0 \qquad C_{gs} \gg C_{gd} \qquad C_{sb} > C_{db}$$



# Rin/out Shortcuts Summary



$$r_o[1 + (g_m + g_{mb})R_S]$$

H.I.N.

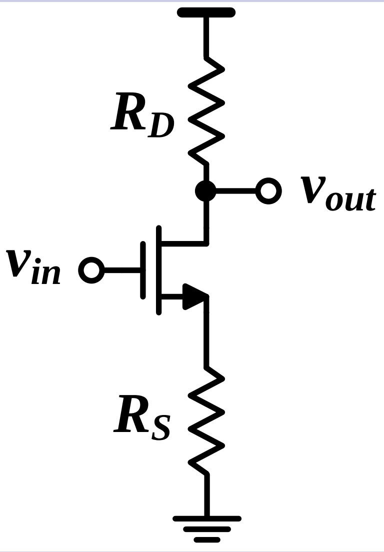
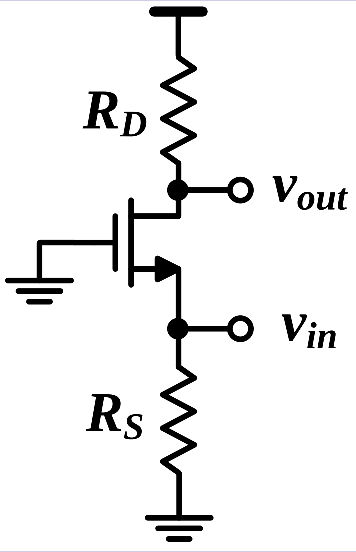
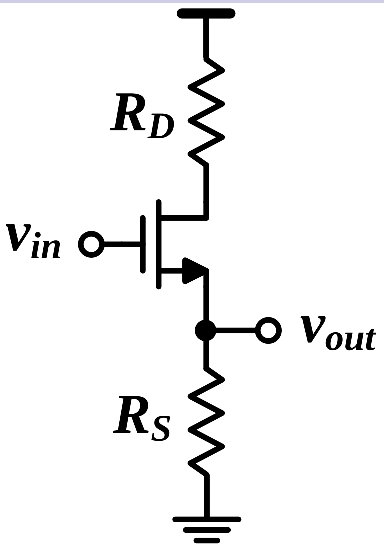
$\infty$

At low frequencies **ONLY**

$$\frac{1}{g_m + g_{mb}}\left(1 + \frac{R_D}{r_o}\right)$$

L.I.N.

# Summary of Basic Topologies

| | CS | CG | CD (SF) |
|---|---|---|---|
| |   |   |   |
| | Voltage & current amplifier | Voltage amplifier<br>Current buffer | Voltage buffer<br>Current amplifier |
| **Rin** | $\infty$ | $R_S \vert\vert \frac{1}{g_m + g_{mb}}\left(1 + \frac{R_D}{r_o}\right)$ | $\infty$ |
| **Rout** | $R_D \vert\vert r_o[1 + (g_m + g_{mb})R_S]$ | $R_D \vert\vert r_o$ | $R_S \vert\vert \frac{1}{g_m + g_{mb}}\left(1 + \frac{R_D}{r_o}\right)$ |
| **Gm** | $\frac{-g_m}{1 + (g_m + g_{mb})R_S}$ | $g_m + g_{mb}$ | $\frac{g_m}{1 + R_D/r_o}$ |

# Differential Amplifier

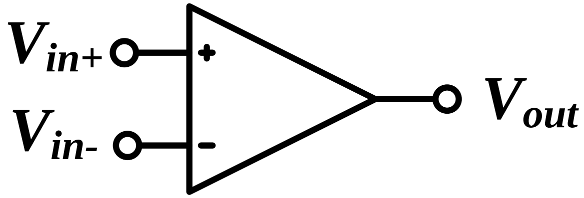
| | Pseudo Diff Amp | Diff Pair (w/ ideal CS) | Diff Pair (w/ $R_{SS}$) |
|---|---|---|---|
| $A_{vd}$ | $-g_m R_D$ | $-g_m R_D$ | $-g_m R_D$ |
| $A_{vCM}$ | $-g_m R_D$ | $0$ | $\frac{-g_m R_D}{1+2(g_m+g_{mb})R_{SS}}$ |
| $A_{vd}/A_{vCM}$ | $1$ | $\infty$ | $2(g_m+g_{mb})R_{SS} \gg 1$ |

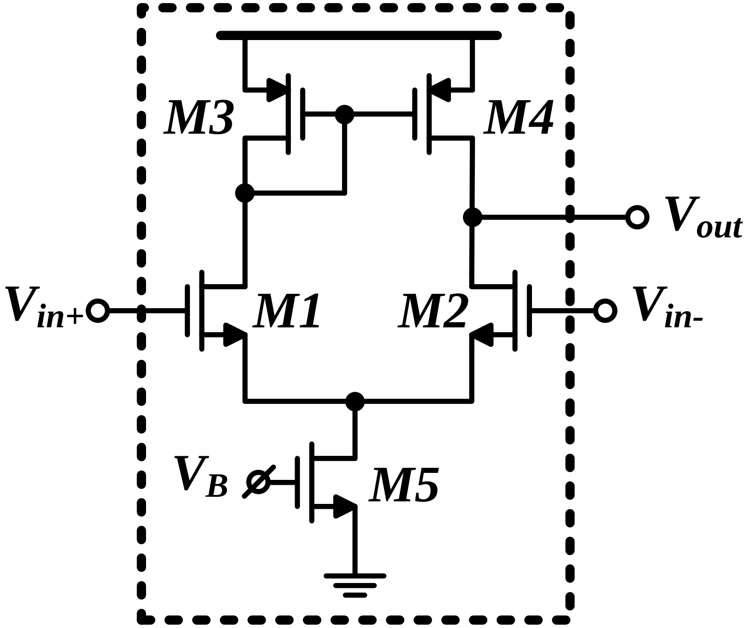
$$A_{vCM2d} = \frac{v_{od}}{v_{iCM}} \approx \frac{\Delta R_D}{2R_{SS}} + \frac{\Delta g_m R_D}{2g_{m1,2}R_{SS}}$$

$$CMRR = \frac{A_{vd}}{A_{vCM2d}}$$

# Op-Amp

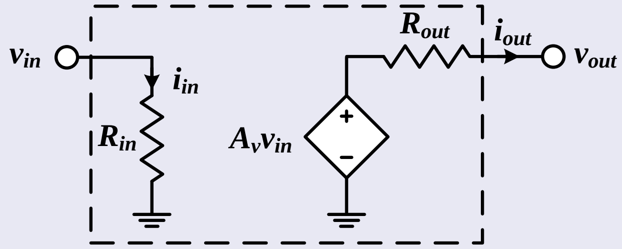
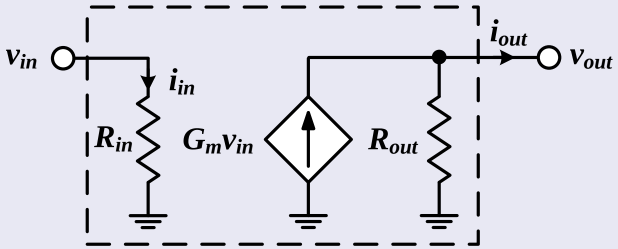
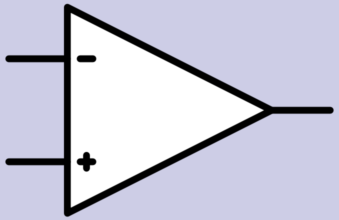
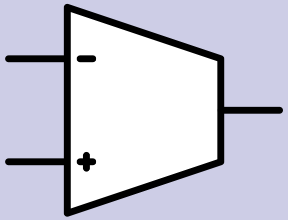
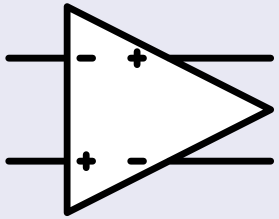
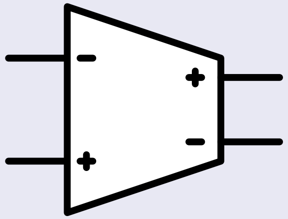
- An op-amp is simply a high gain differential amplifier
  - The gain can be increased by using cascodes and multi-stage amplification
- The diff amp is a key block in many analog and RF circuits
  - DEEP understanding of diff amp is ESSENTIAL



# Op-Amp vs OTA

- In short, an OTA is an op-amp without an output stage (buffer)
- Some designers just use op-amp name and symbol for both

| | Op-amp | OTA |
|---|---|---|
| **Rout** | **LOW** | **HIGH** |
| **Model** |   |   |
| **Diff input, SE output** |  |  |
| **Fully diff** |  |  |

# V-star ($\boldsymbol{V^*}$)

- V-star ($V^*$) is inspired by $V_{ov}$ but calculated from actual simulation data

$$g_m = \frac{2I_D}{V^*} \leftrightarrow V^* = \frac{2I_D}{g_m} = \frac{2}{g_m/I_D}$$

- Figures-of-merit in terms of $V^*$

$$g_m r_o = \frac{2I_D}{V^*} \cdot \frac{1}{\lambda I_D} = \frac{2}{\lambda V^*}$$

$$f_T = \frac{g_m}{2\pi C_{gg}} = \frac{1}{2\pi} \cdot \frac{2I_D}{V^*} \cdot \frac{1}{C_{gg}}$$

$$\frac{g_m}{I_D} = \frac{2}{V^*}$$

- The boundary between weak and strong inversion ($n = 1.2 \rightarrow 1.5$)

$$V_{ov}(SI) = V^*(WI) = 2nV_T \approx 60 \rightarrow 80mV$$

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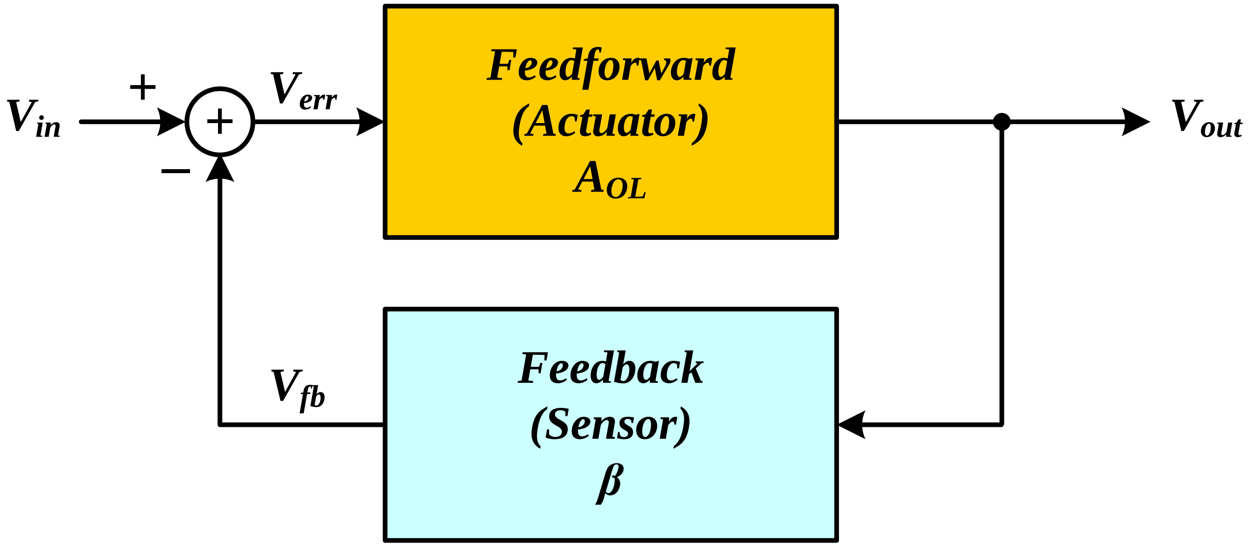
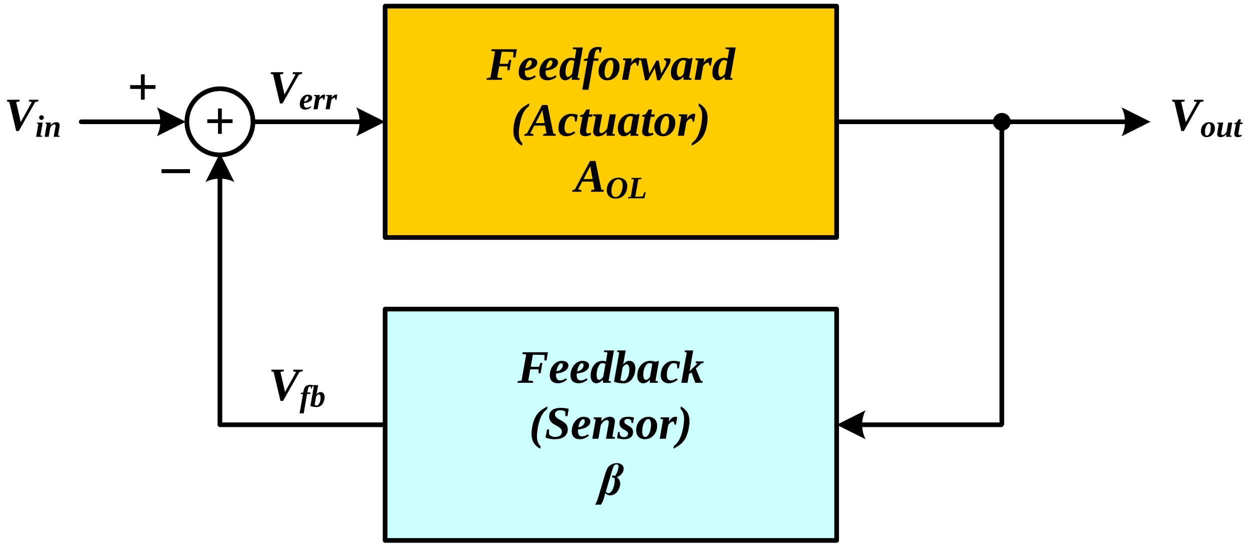
# General Feedback System



# General Feedback System

- Error signal $= V_{err} = V_{in} - V_{fb}$
- Open loop (OL) gain $= A_{OL} = \frac{V_{out}}{V_{err}} \gg 1$
- Feedback factor $= \beta = \frac{V_{fb}}{V_{out}}$
- Closed loop (CL) gain $= A_{CL} = \frac{V_{out}}{V_{in}}$



# Closed-loop Gain

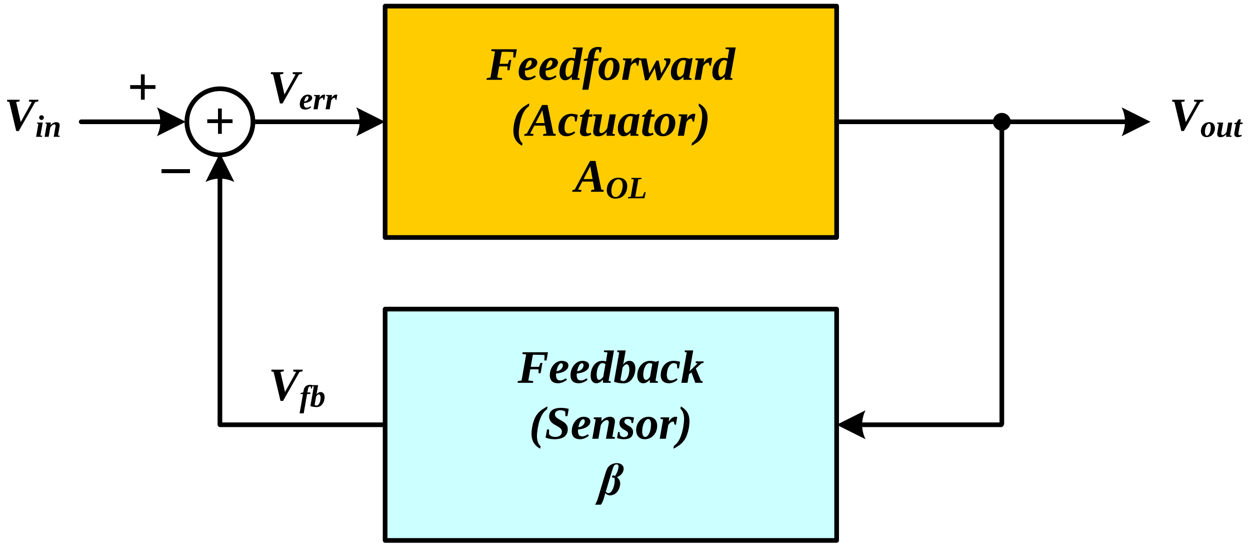
$$V_{out} = A_{OL}(V_{in} - V_{fb}) = A_{OL}(V_{in} - \beta V_{out})$$

$$A_{CL} = \frac{V_{out}}{V_{in}} = \frac{A_{OL}}{1 + \beta A_{OL}} = \frac{A_{OL}}{1 + LG}$$

- Loop gain $= LG = \beta A_{OL} \gg 1$

$$A_{CL} \approx \frac{1}{\beta}$$

- Closed-loop gain is independent of open-loop gain!



# Error Signal

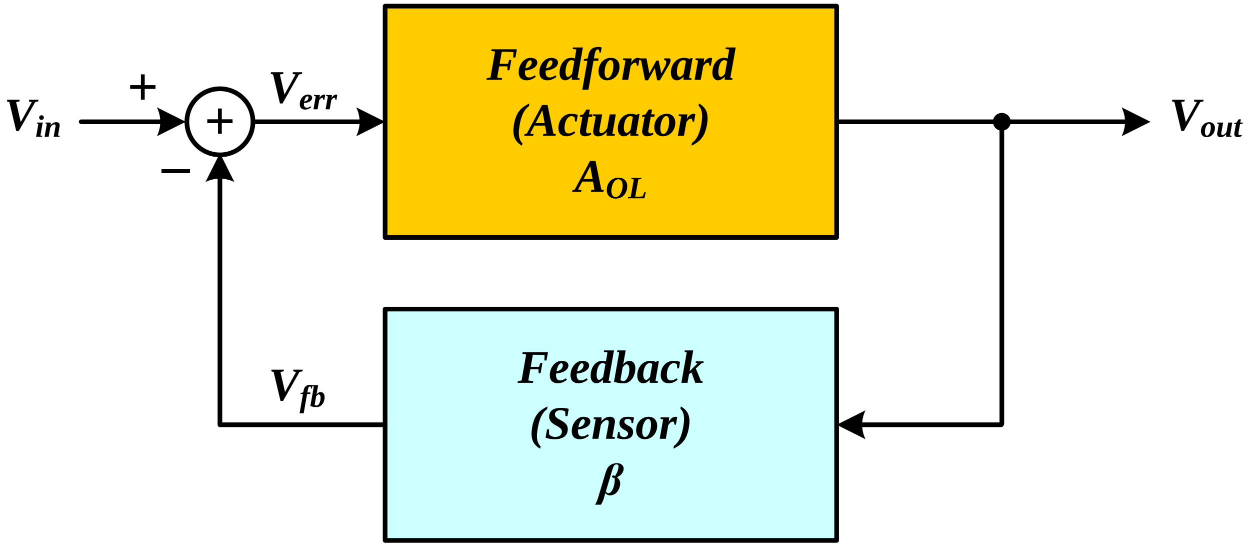
$$V_{err} = V_{in} - V_{fb} = V_{in} - \beta V_{out} = V_{in} - \beta A_{OL} V_{err}$$

$$V_{err} = \frac{V_{in}}{1 + \beta A_{OL}} = \frac{V_{in}}{1 + LG}$$

- Loop gain $= LG = \beta A_{OL} \gg 1$

$$V_{err} = \frac{V_{in}}{1 + LG} \to 0$$

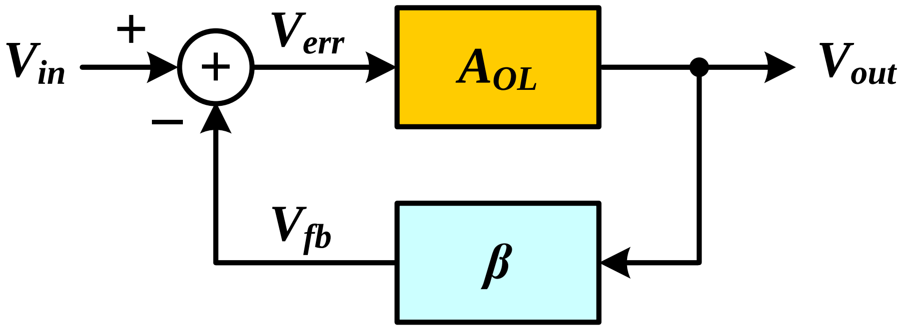
- Negative feedback loop works to minimize the error signal

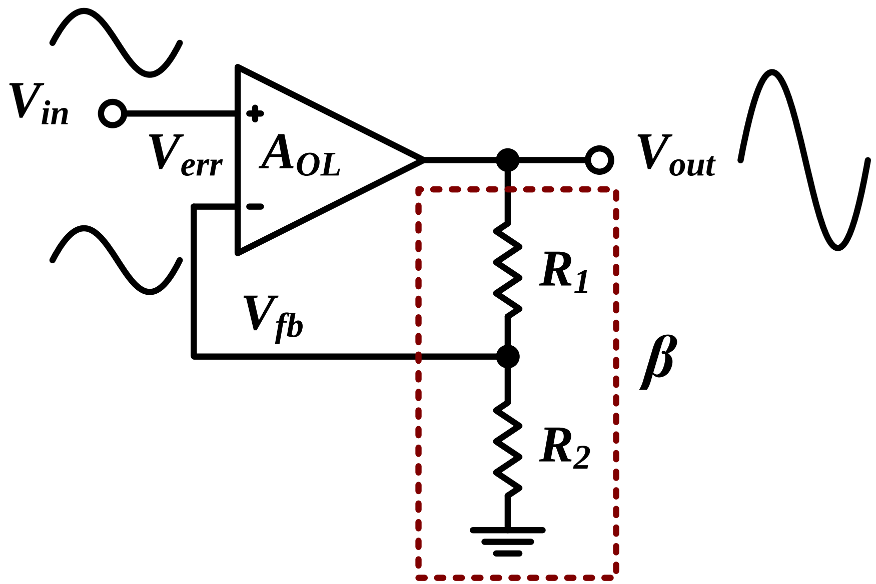


# Feedback Example

- Op-amp functions: (1) subtraction and (2) amplification
- The network $R_1$ and $R_2$ functions: (1) sensing the output voltage and (2) providing a feedback factor $\beta = \frac{R_2}{(R_1+R_2)}$

$$A_{CL} = \frac{V_{out}}{V_{in}} = \frac{A_{OL}}{1+\beta A_{OL}} = \frac{A_{OL}}{1+\frac{R_2}{(R_1+R_2)}\cdot A_{OL}} \approx \frac{R_1+R_2}{R_2} = 1+\frac{R_1}{R_2}$$



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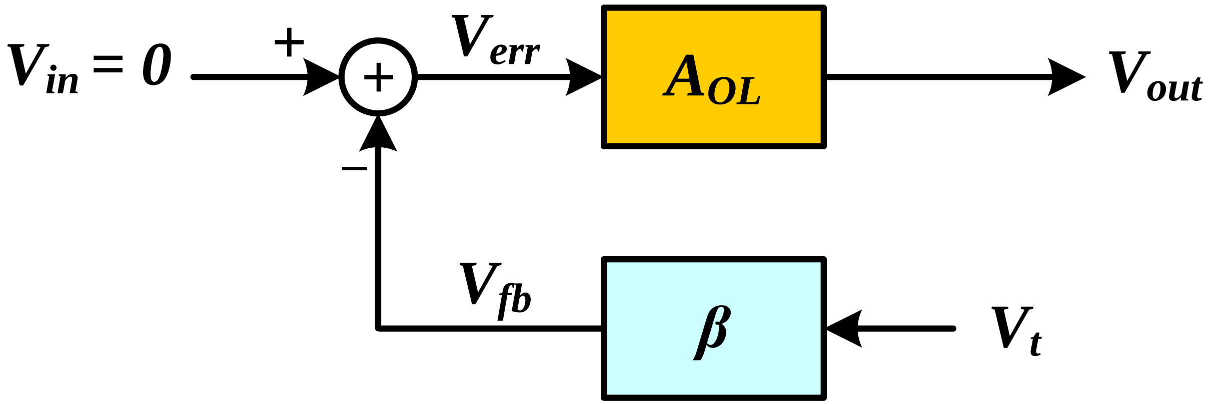
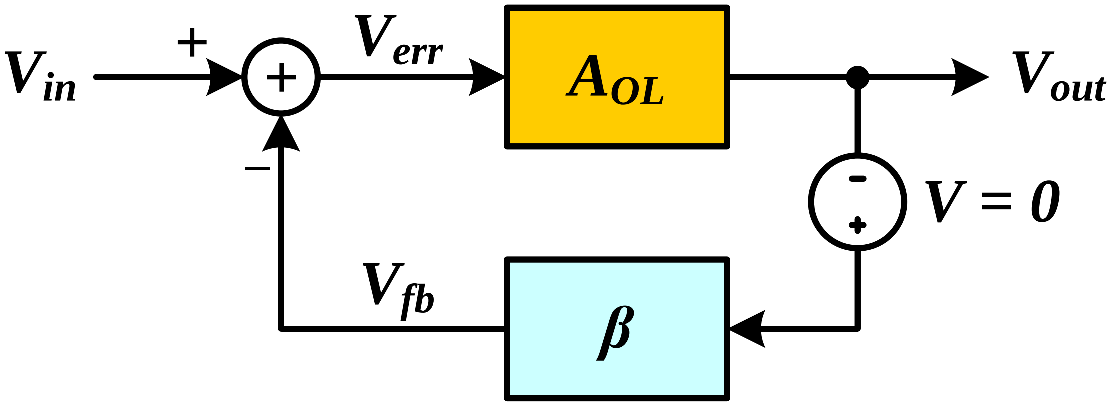
# Loop Gain

- Deactivate the input → Break the loop → Apply a test source → Calculate the gain around the loop
- Loop gain $= LG = -\frac{V_{out}}{V_t} = \beta A_{OL}$
- A.k.a. loop transmission, return ratio …
- But biasing/loading changes when we break the loop!
  - Make sure dc biasing is properly set
  - Add a dummy load



# Loop Gain

- Modern circuit simulators can compute the loop gain without explicitly breaking the loop
  - Use stability (STB) analysis
  - Insert a 0V dc voltage source or iprobe in the loop
    - Polarity matters for Eldo, but not for Spectre
  - Loop gain $= LG = \beta A_{OL}$ is calculated by the simulator



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# Why Negative Feedback?

- We use a very high gain amplifier $(A_{OL})$, but end up with a much smaller gain $A_{CL} = \frac{A_{OL}}{1+\beta A_{OL}} \approx \frac{1}{\beta}$
- We can design high gain amplifiers, but we really do not need all that gain
- High gain is the balance that we use to buy other useful properties
- Negative feedback useful properties
  1. **Gain Desensitization → Stable, linear, and accurate gain**
  2. Bandwidth Extension
  3. Modification of I/O Impedances

# Gain Desensitization

$$A_{CL} = \frac{A_{OL}}{1 + \beta A_{OL}} = \frac{1}{\beta}\left(\frac{1}{\frac{1}{\beta A_{OL}} + 1}\right) \approx \frac{1}{\beta}\left(1 - \frac{1}{\beta A_{OL}}\right) = \frac{1}{\beta}\left(1 - \frac{1}{LG}\right)$$

- Static gain error

$$\epsilon_s = \frac{\left|A_{CL,ideal} - A_{CL,actual}\right|}{A_{CL,ideal}} \approx \frac{1}{\beta A_{OL}} = \frac{1}{LG}$$

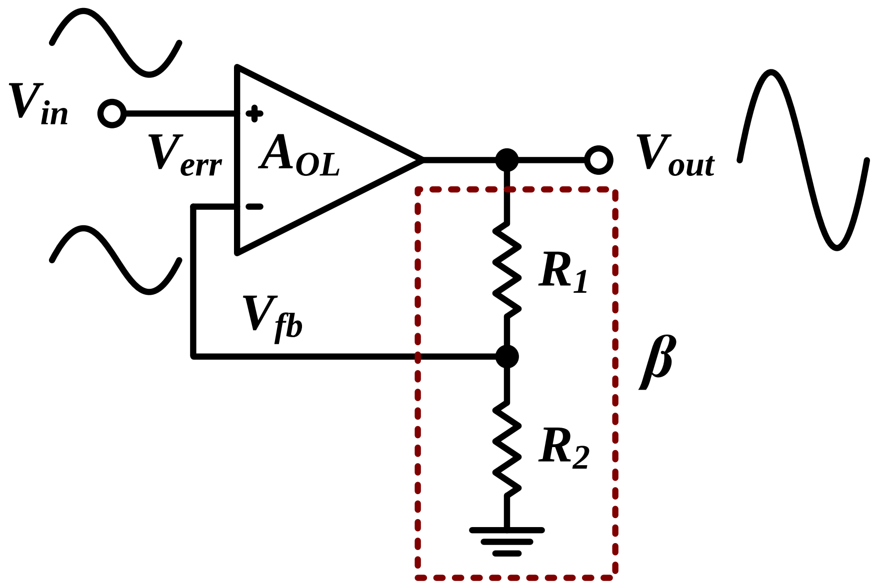
- $A_{OL}$ varies due to PVT, load, and input signal variations
- $A_{CL}$ almost independent of $A_{OL}$ (if $LG \gg 1$)
  - Independent of PVT: stable and robust
  - Independent of load: stable and robust
  - Independent of input level: linear

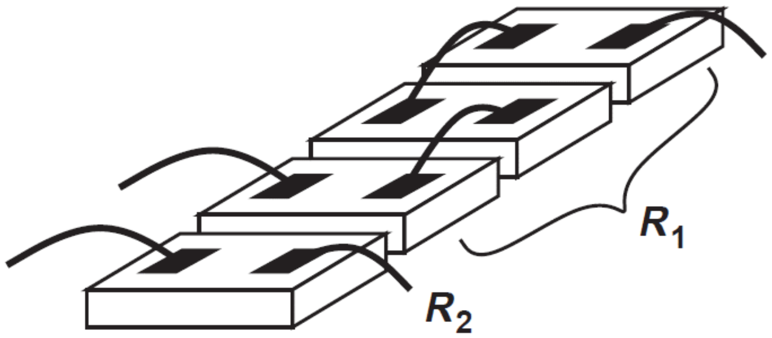
# Gain Desensitization

- In IC design, we cannot control absolute values due to PVT, load, and input signal variations
- But we can precisely control ratios of MATCHED components

$$A_{CL} = \frac{Y}{X} = \frac{A_{OL}}{1 + \beta A_{OL}} = \frac{A_{OL}}{1 + \frac{R_2}{(R_1 + R_2)} \cdot A_{OL}} \approx \frac{R_1 + R_2}{R_2} = 1 + \frac{R_1}{R_2}$$

- $R_1 = 3R\ and\ R_2 = R \Rightarrow A_{CL} = 4 \Rightarrow$ Stable, linear, and **accurate**






[Razavi, 2017]

# Bandwidth Extension

- Assume the op-amp (OL) is a first order system

$$A_{OL}(s) = \frac{A_{OLo}}{1 + \frac{s}{\omega_{p,OL}}}$$

- The CL transfer function is also a first order system

$$A_{CL}(s) = \frac{A_{OL}(s)}{1 + \beta A_{OL}(s)} = \frac{\frac{A_{OLo}}{(1 + \beta A_{OLo})}}{1 + \frac{s}{(1 + \beta A_{OLo})\omega_{p,OL}}} = \frac{A_{CLo}}{1 + \frac{s}{\omega_{p,CL}}}$$

  - But the pole is at a much higher frequency

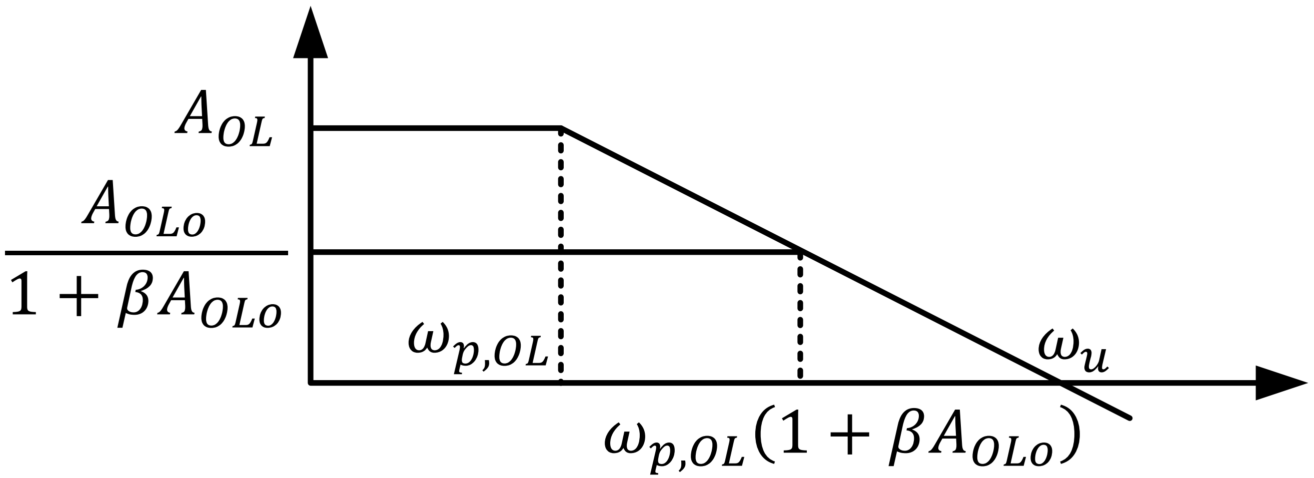
$$\omega_{p,CL} = (1 + LG_o)\omega_{P,OL}$$

- CL DC gain reduced by $(1 + LG_o)$
- CL bandwidth extended by $(1 + LG_o)$
- GBW (and UGF) remains constant

# Bandwidth Extension

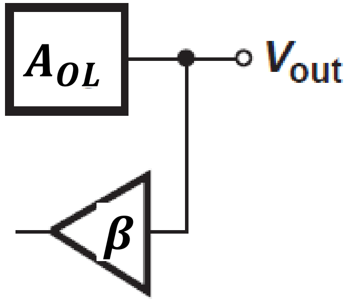
- CL DC gain reduced by $(1 + LG_o)$
- CL bandwidth extended by $(1 + LG_o)$
- GBW (and UGF) remains constant

$$A_{CL}(s) = \frac{\dfrac{A_{OLo}}{(1 + LG_o)}}{1 + \dfrac{s}{(1 + LG_o)\omega_{p,OL}}} = \frac{A_{CLo}}{1 + \dfrac{s}{\omega_{p,CL}}}$$

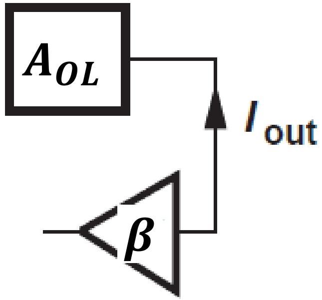


# Modification of I/O Impedances

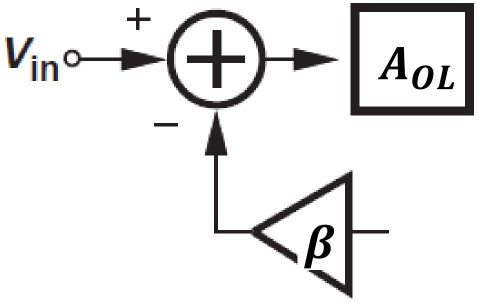
- ❑ Shunt sensing/mixing → R decreases
- ❑ Series sensing/mixing → R increases



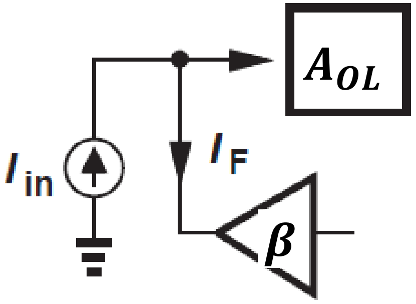

**Output impedance falls by 1+ loop gain.**




**Output impedance rises by 1+ loop gain.**




**Input impedance rises by 1+ loop gain.**




**Input impedance falls by 1+ loop gain.**



[Razavi, 2017]

# The Price We Pay

- Negative feedback useful properties
  1. Gain Desensitization → Stable, linear, and accurate gain
  2. Bandwidth Extension
  3. Modification of I/O Impedances
- The price we pay to buy these useful properties
  1. Gain reduction
  2. The risk of instability

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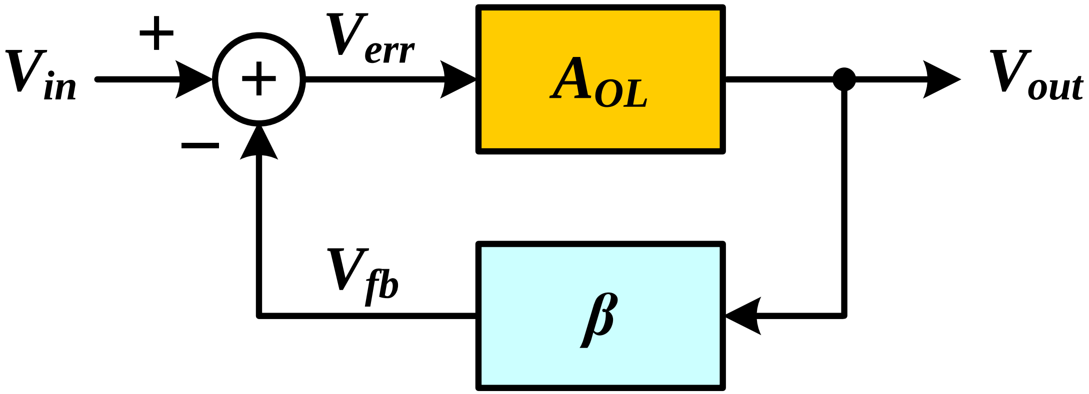
# Stability of Feedback System

$$A_{CL}(s) = \frac{V_{out}(s)}{V_{in}(s)} = \frac{A_{OL}(s)}{1 + \beta A_{OL}(s)}$$

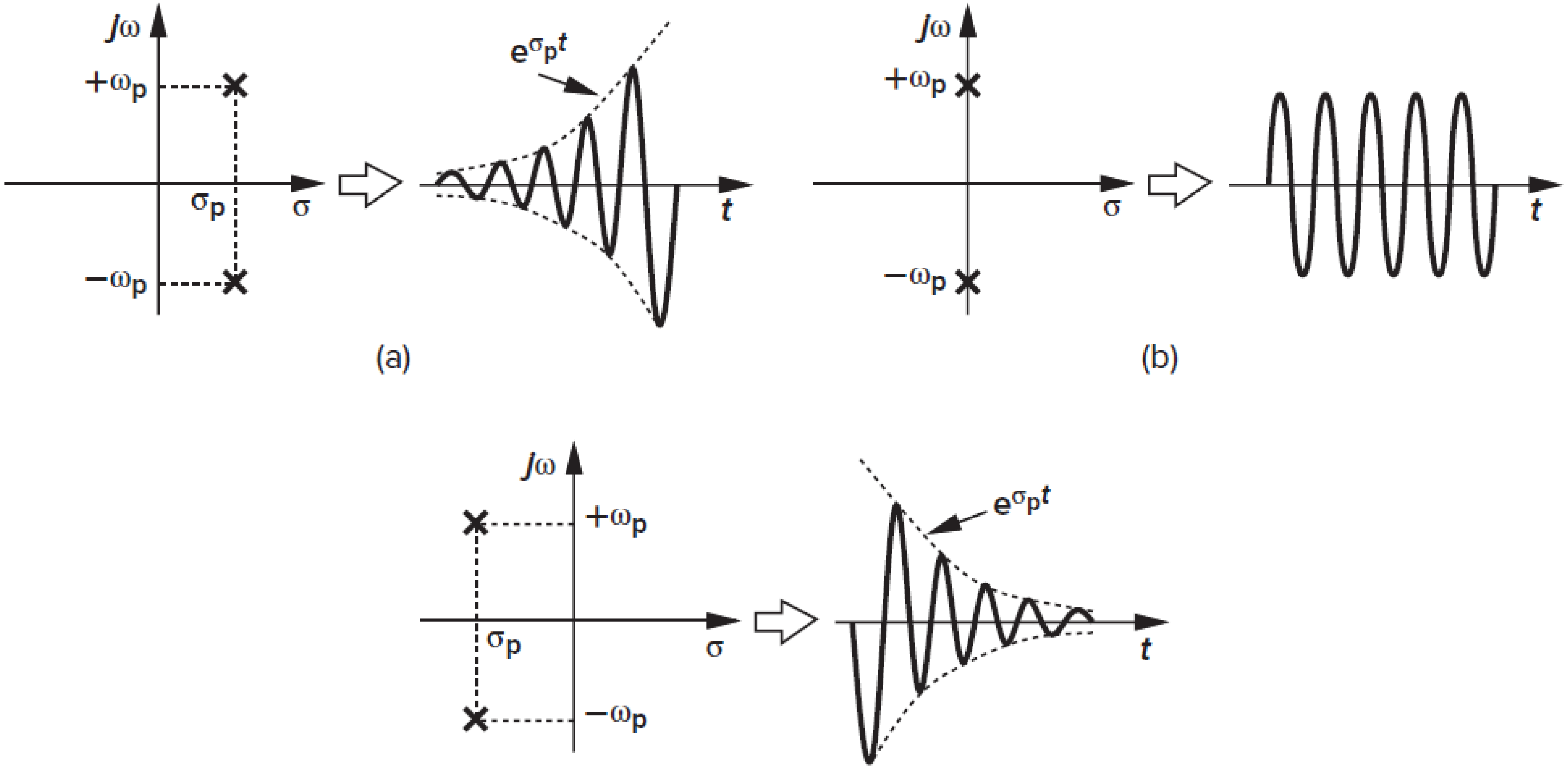
- Barkhausen's Oscillation Criteria

$$|\beta A_{OL}(s)| = 1$$
$$\angle \beta A_{OL}(s) = -180$$



# Stable vs Unstable System: Pole-Zero Plot

| Laplace domain | Time domain |
|---|---|
| $\frac{1}{s-a}$ | $e^{at}$ |



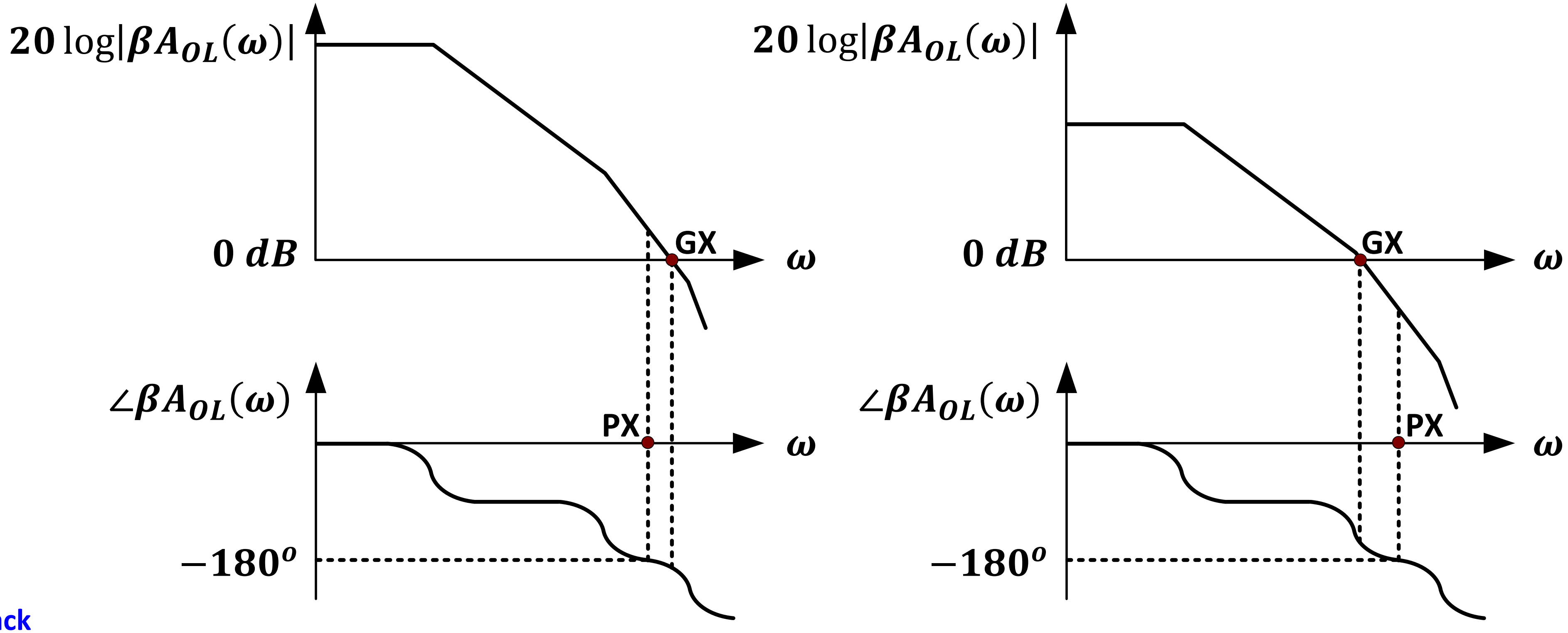

(a) (b) (c)



[Razavi, 2017]

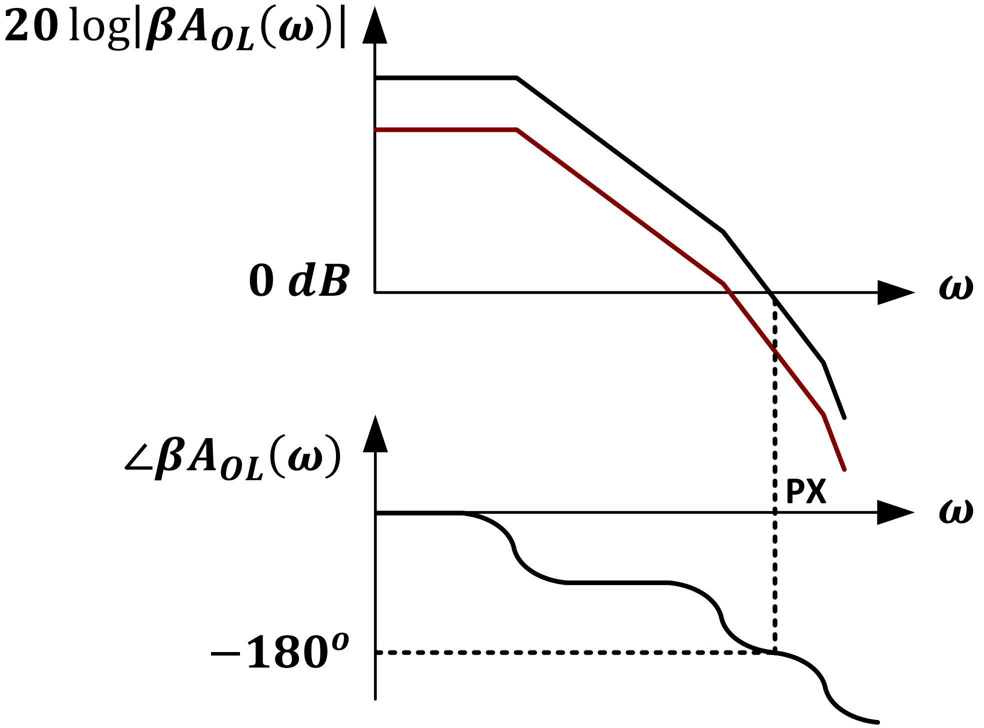
# Stable vs Unstable System: Bode Plot

- Gain crossover frequency (GX): @ $|\beta A_{OL}(s)| = 1$
  - Same as $\omega_u$
- Phase crossover frequency (PX): @ $\angle\beta A_{OL}(s) = -180$
- For a stable system: GX < PX



# Effect of Feedback Factor $(\boldsymbol{\beta})$

- We assume $\beta$ is independent of frequency
  - $\angle\beta A_{OL}$ is independent of $\beta$ → PX is independent of $\beta$
- Increasing $\beta$ shifts mag up → GX increases → bad for stability
- Worst-case stability corresponds to $\beta$ = 1 → $\beta A_{OL} = A_{OL}$
  - Unity-gain feedback → buffer → smallest CL gain



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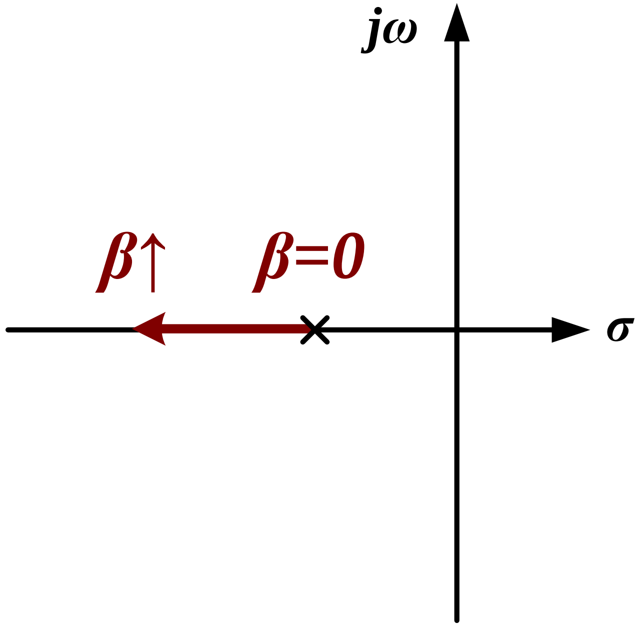
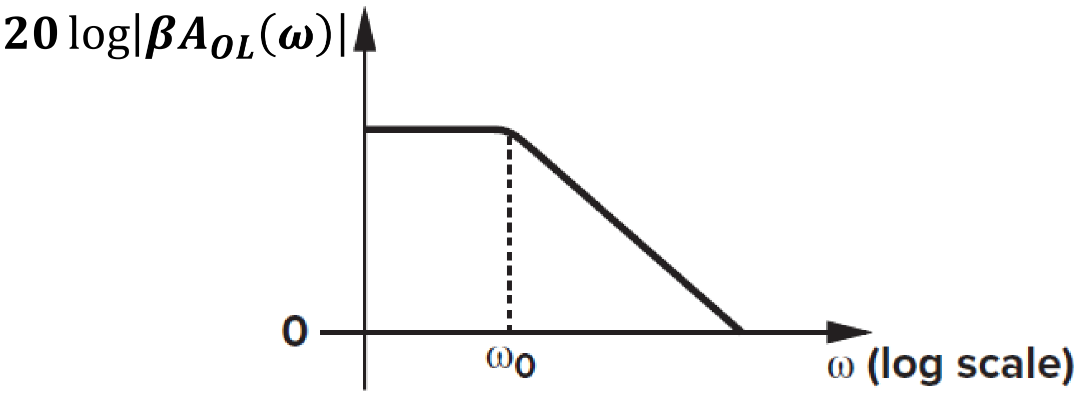
# Single-Pole System: Root Locus

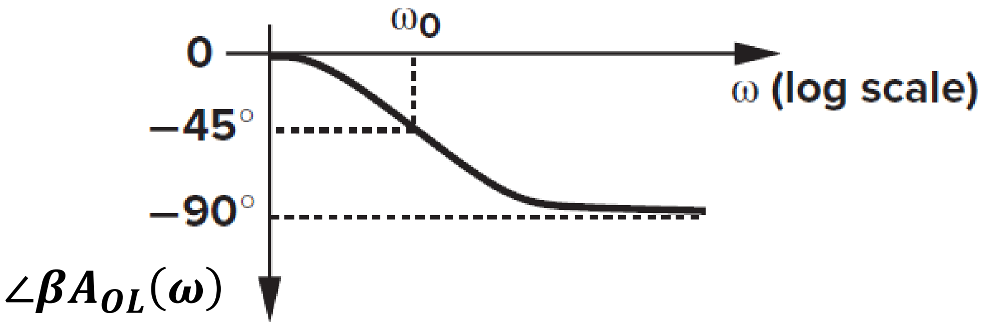
- Note that $\beta$ does not affect poles of $LG$ (assuming $\beta$ is real!)
  - But it affects poles of $A_{CL}$: Roots of the characteristic equation $(1 + \beta A_{OL})$
- The locus exists on real axis to the left of an odd number of poles and zeros.
- The locus starts at the open-loop poles and ends at the open-loop zeros or at infinity.
- For first-order system, pole always in LHP: Unconditionally stable



# Single-Pole System: Bode Plot

$$A_{CL}(s) = \frac{A_{OL}(s)}{1+\beta A_{OL}(s)} = \frac{\frac{A_{OLo}}{(1+\beta A_{OLo})}}{1+\frac{s}{(1+\beta A_{OLo})\omega_{p,OL}}} = \frac{A_{CLo}}{1+\frac{s}{\omega_{p,CL}}}$$





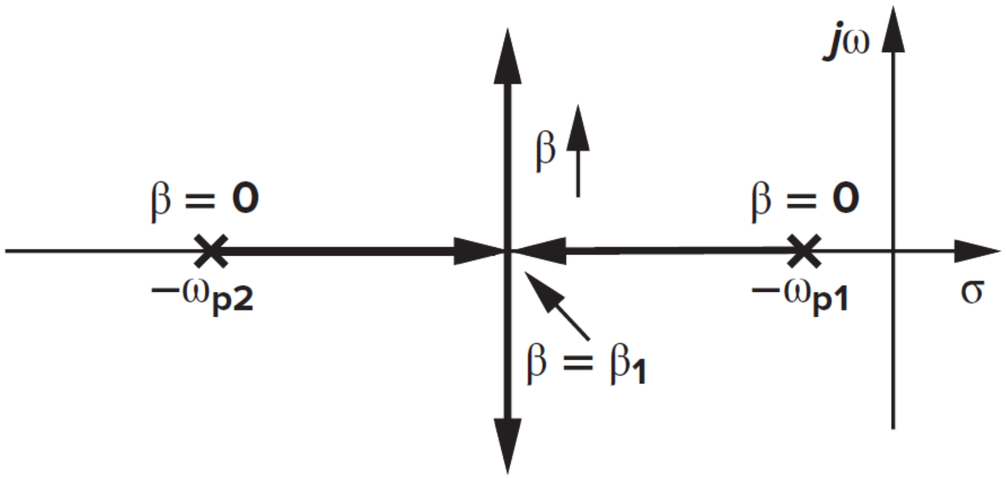

[Razavi, 2017]

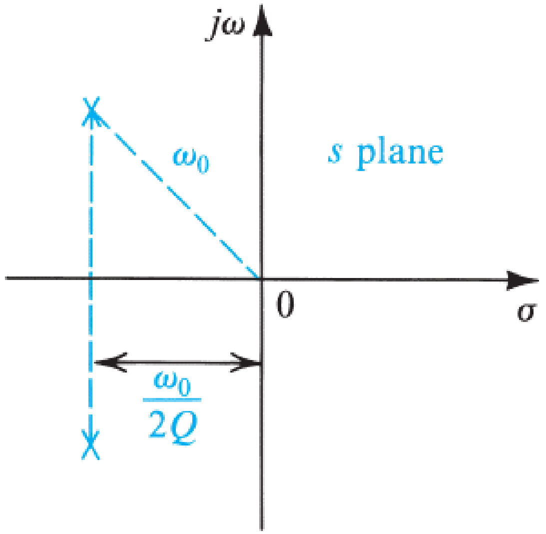
# Two-Pole System: Root Locus

- Poles always in LHP: Unconditionally stable

$$A_{CL}(s) = \frac{A_{CLo}}{1 + \frac{1}{Q}\frac{s}{\omega_o} + \frac{s^2}{\omega_o^2}} = \frac{A_{CLo}}{1 + 2\zeta\frac{s}{\omega_o} + \frac{s^2}{\omega_o^2}}$$

  - $\zeta > 1$ ($Q < 0.5$): Overdamped (real and distinct CL poles)
  - $\zeta = 1$ ($Q = 0.5$): Critical damped (real and equal CL poles)
  - $\zeta < 1$ ($Q > 0.5$): Underdamped (complex conjugate CL poles)



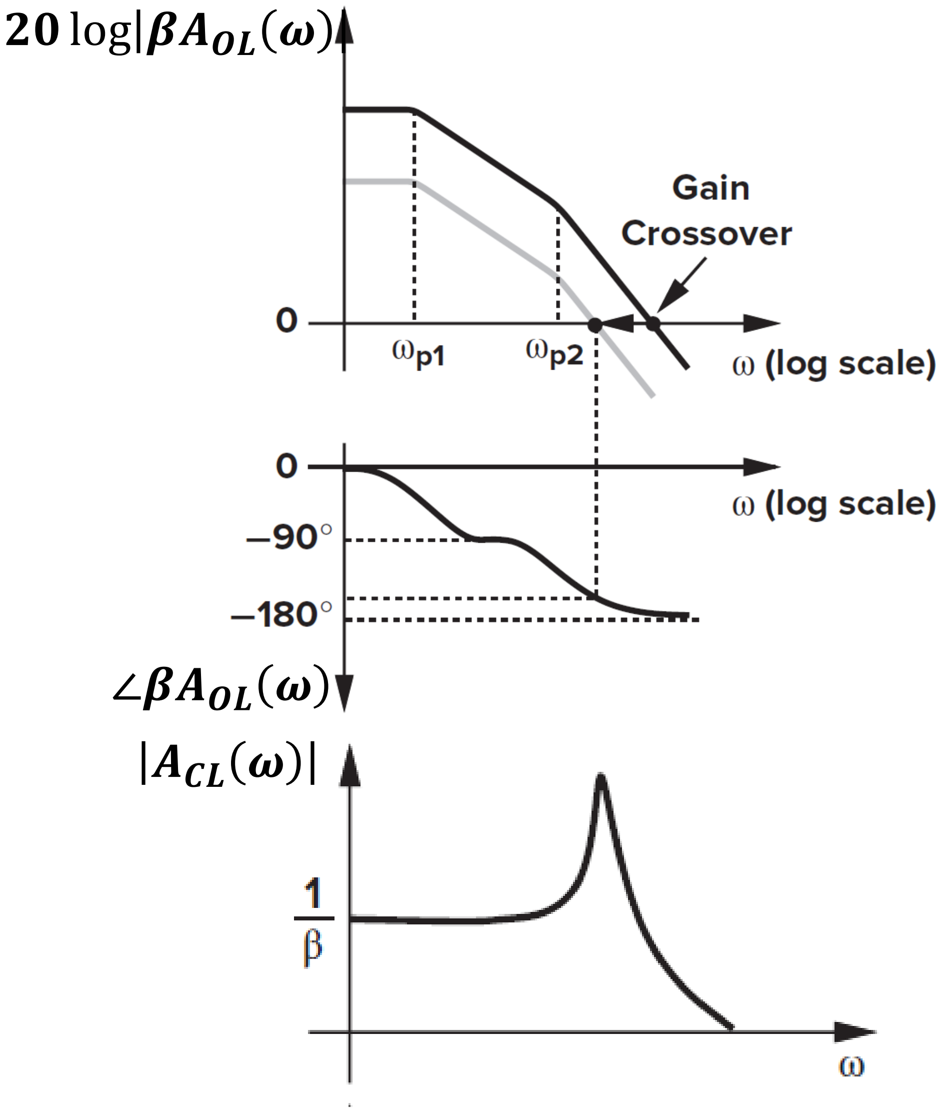






[Sedra/Smith, 2015] & [Razavi, 2017]

# Two-Pole System: Bode Plot

- Phase shift always < $180^o$
  - Unconditionally stable
- $\zeta > 1$ ($Q < 0.5$): Overdamped
- $\zeta = 1$ ($Q = 0.5$): Critical damped
- $\zeta < 1$ ($Q > 0.5$): Underdamped
  - Overshoot in step response
- $\zeta < 1/\sqrt{2} = 0.707$
  - $Q > 1/\sqrt{2} = 0.707$
  - Peaking in frequency response
  - Ringing in step response

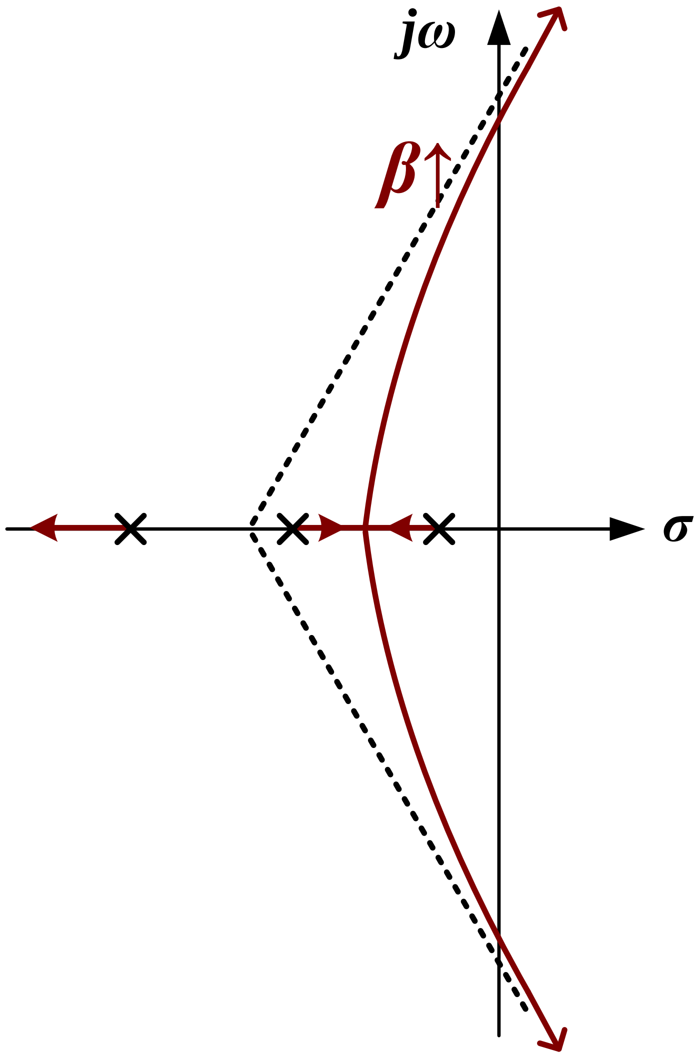





[Razavi, 2017]

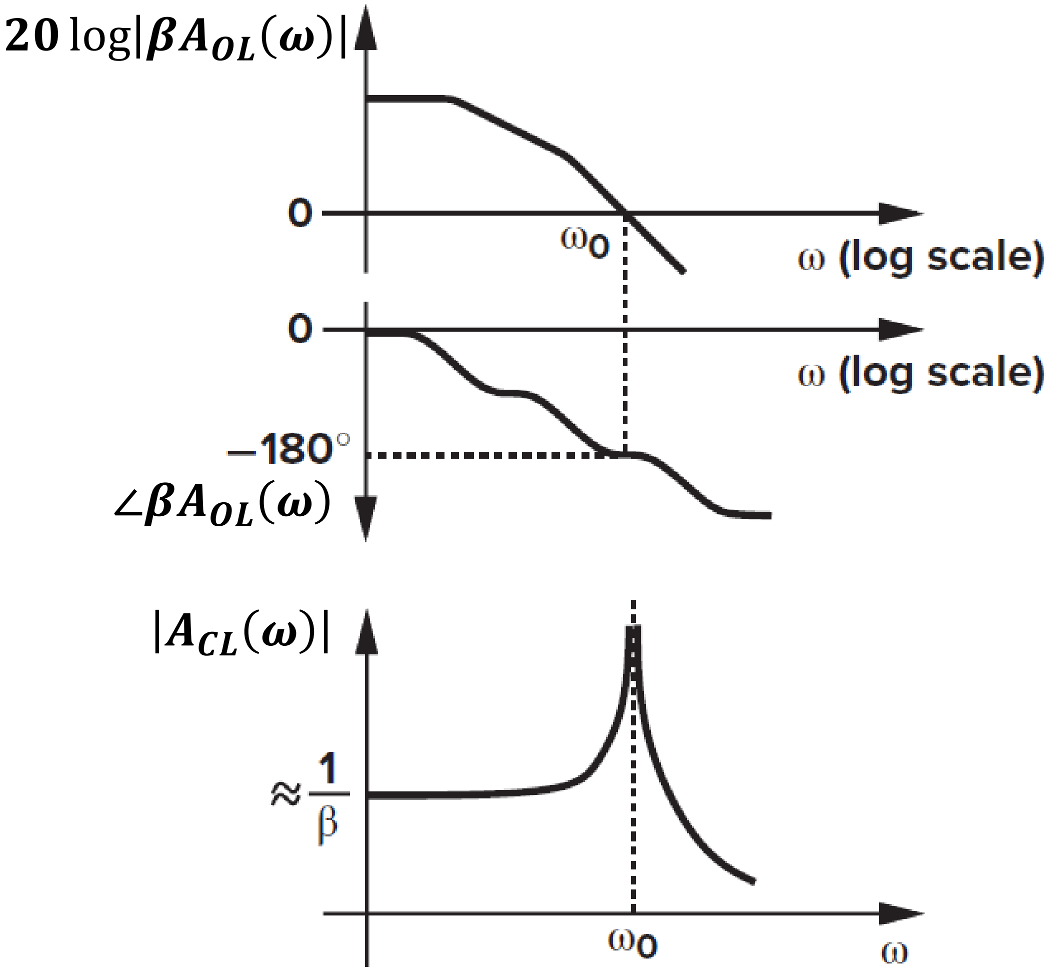
# Three-Pole System: Root Locus

- Poles cross $j\omega$ axis at a specific value of $\beta = \beta_{crit}$
  - The system becomes unstable
- Pole are NOT always in LHP
  - Conditionally stable: $\beta < \beta_{crit}$



# Three-Pole System: Bode Plot

- Instability (oscillation) condition can be satisfied





[Razavi, 2017]

# Stability Summary

- First-order system
  - Unconditionally stable
- Third order system
  - Conditionally stable: Set $\beta < \beta_{crit}$
- Second-order system
  - Unconditionally stable
  - But may suffer from CL peaking/ringing if close to oscillation condition
  - How much margin is needed?

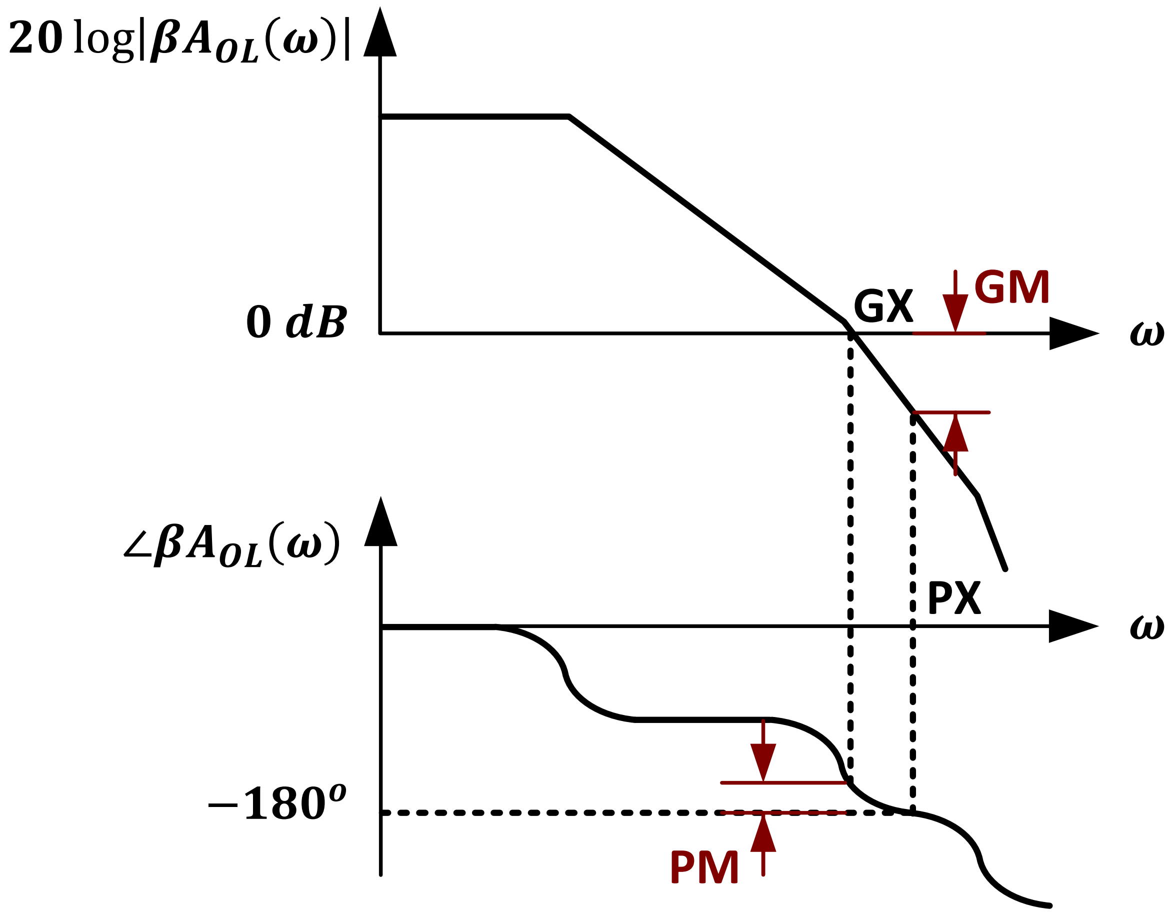
# Outline

- ❑ Recapping previous key results
- ❑ General feedback system
- ❑ Loop gain
- ❑ Why negative feedback?
- ❑ Stability of feedback system
- ❑ Root locus and Bode plot
- ❑ Phase and gain margin

# Phase and Gain Margin

$$PM = 180^o - |\angle\beta A_{OL}(GX)| = 180^o - \tan^{-1}\left(\frac{GX}{\omega_{p1}}\right) - \tan^{-1}\left(\frac{GX}{\omega_{p2}}\right)$$

$$GM = 0 - 20\,log|\beta A_{OL}(PX)|$$

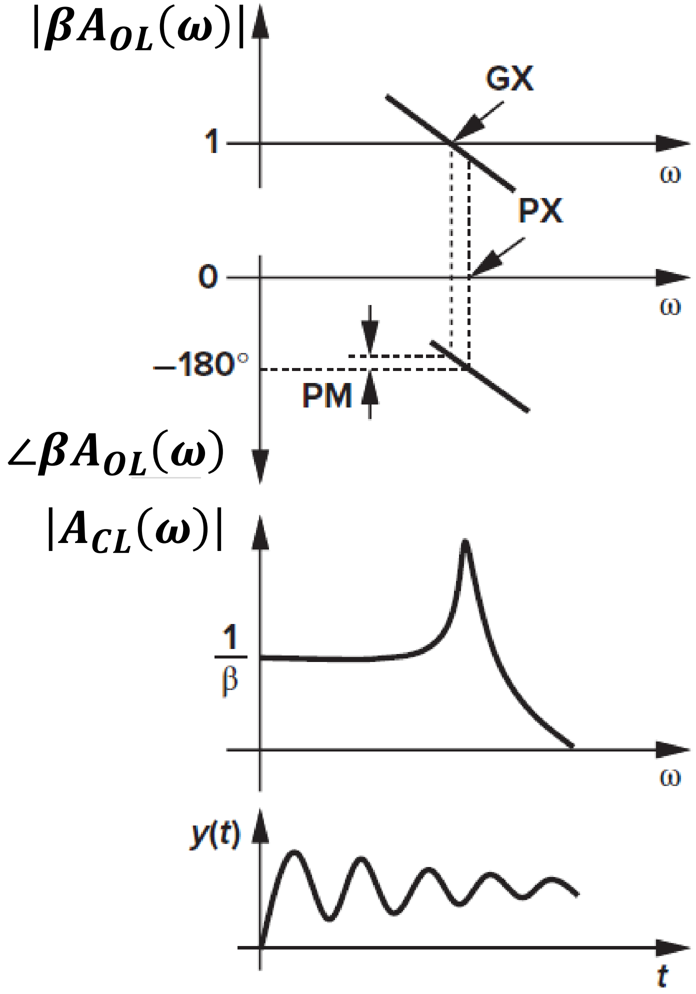


# Phase Margin (PM)

$$PM = 90^o - \tan^{-1}\left(\frac{GX}{\omega_{p2}}\right)$$

PM > 0 → stable
But low PM means:
→ frequency domain peaking
→ time domain ringing

Frequency domain peaking
→ noise amplification
Time domain ringing
→ poor settling time





[Razavi, 2017]

# Phase Margin (PM)

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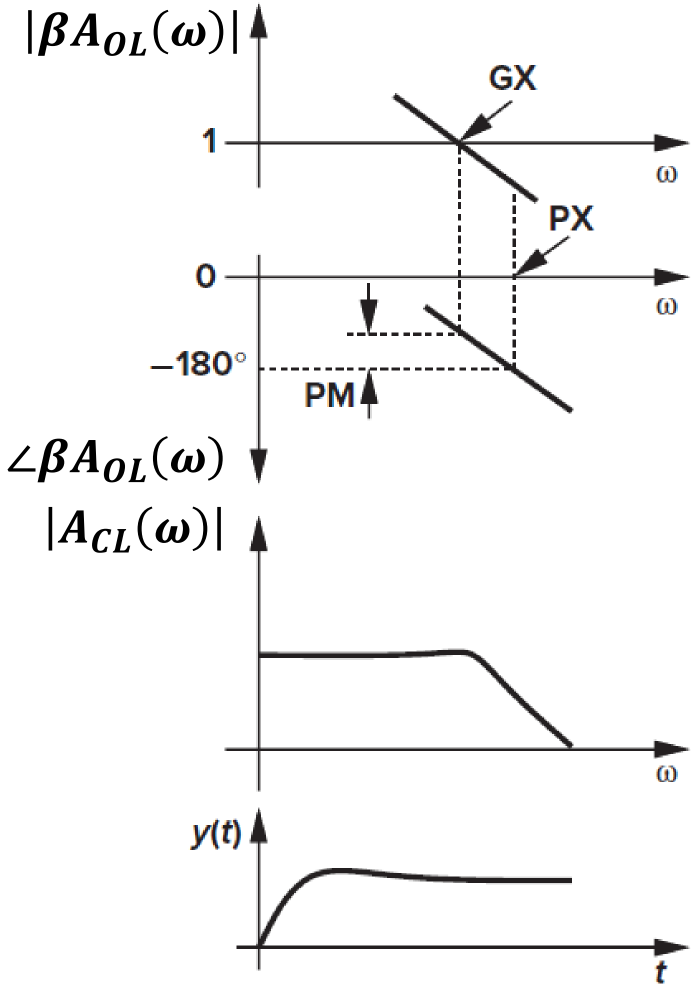
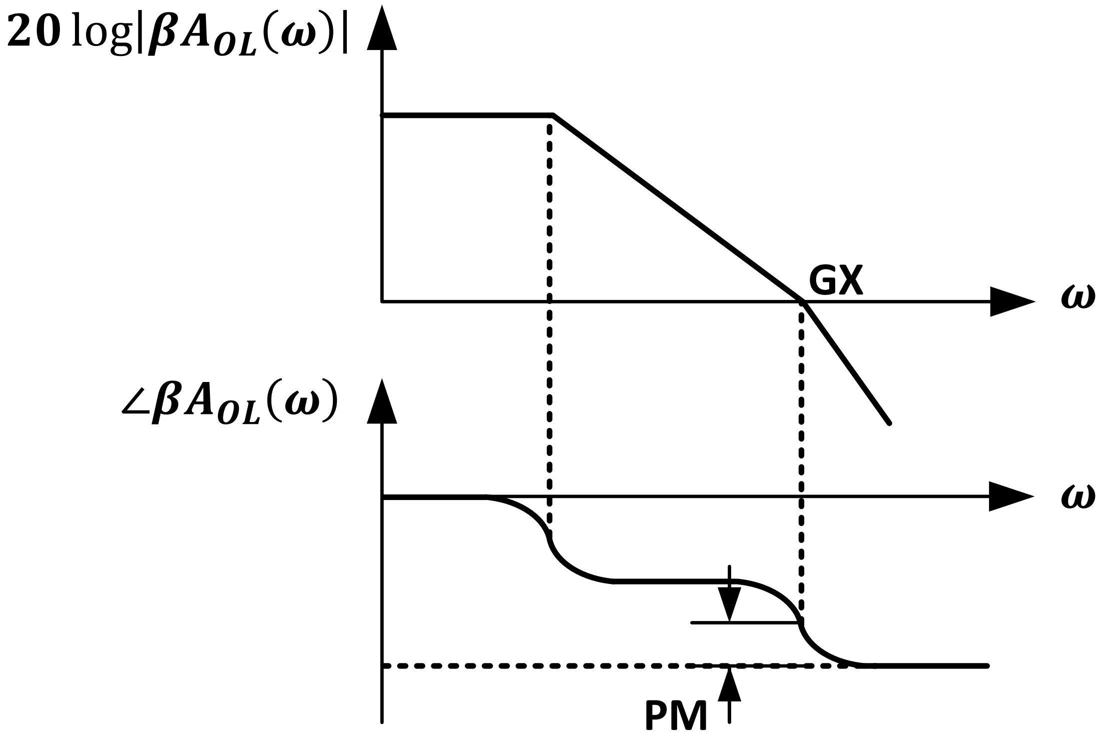



 [Razavi, 2017] 

# Phase Margin: Ultimate GBW

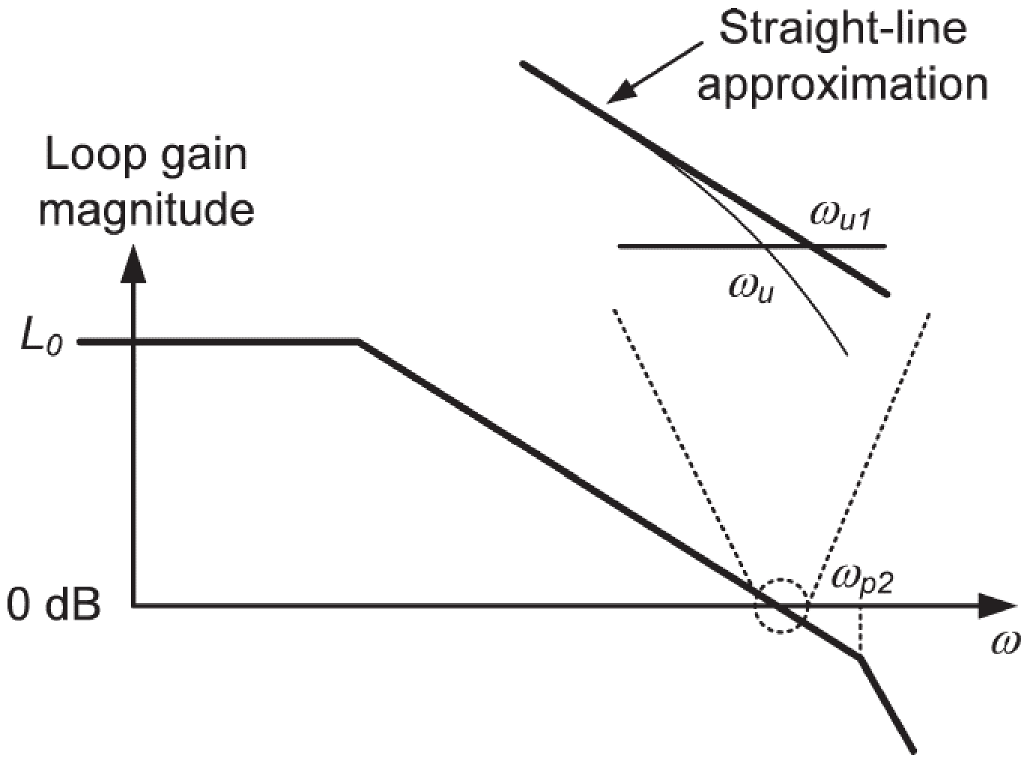
- If $\omega_{p2} = \omega_u$: PM = 45°
  - Typically inadequate (peaking/ringing)
- Thus $\omega_{p2}$ should be $> \omega_u$ → $\omega_{p1} \ll \omega_u < \omega_{p2}$
  - $\omega_{p1}$ defines OL BW and $\omega_{p2}$ defines ultimate GBW (max CL BW)

Frequency domain peaking
→ noise amplification
Time domain ringing
→ poor settling time



# Asymptotic vs Actual UGF

- For $\omega < \omega_u$ the Bode plot is similar to a 1st order system
- For a true 1st order system: GBW = UGF $= \omega_u = \omega_{u1}$
- But the 2nd pole causes some bending below the asymptote
  - Actual $\omega_u$ slightly < $\omega_{u1}$: UGF < GBW

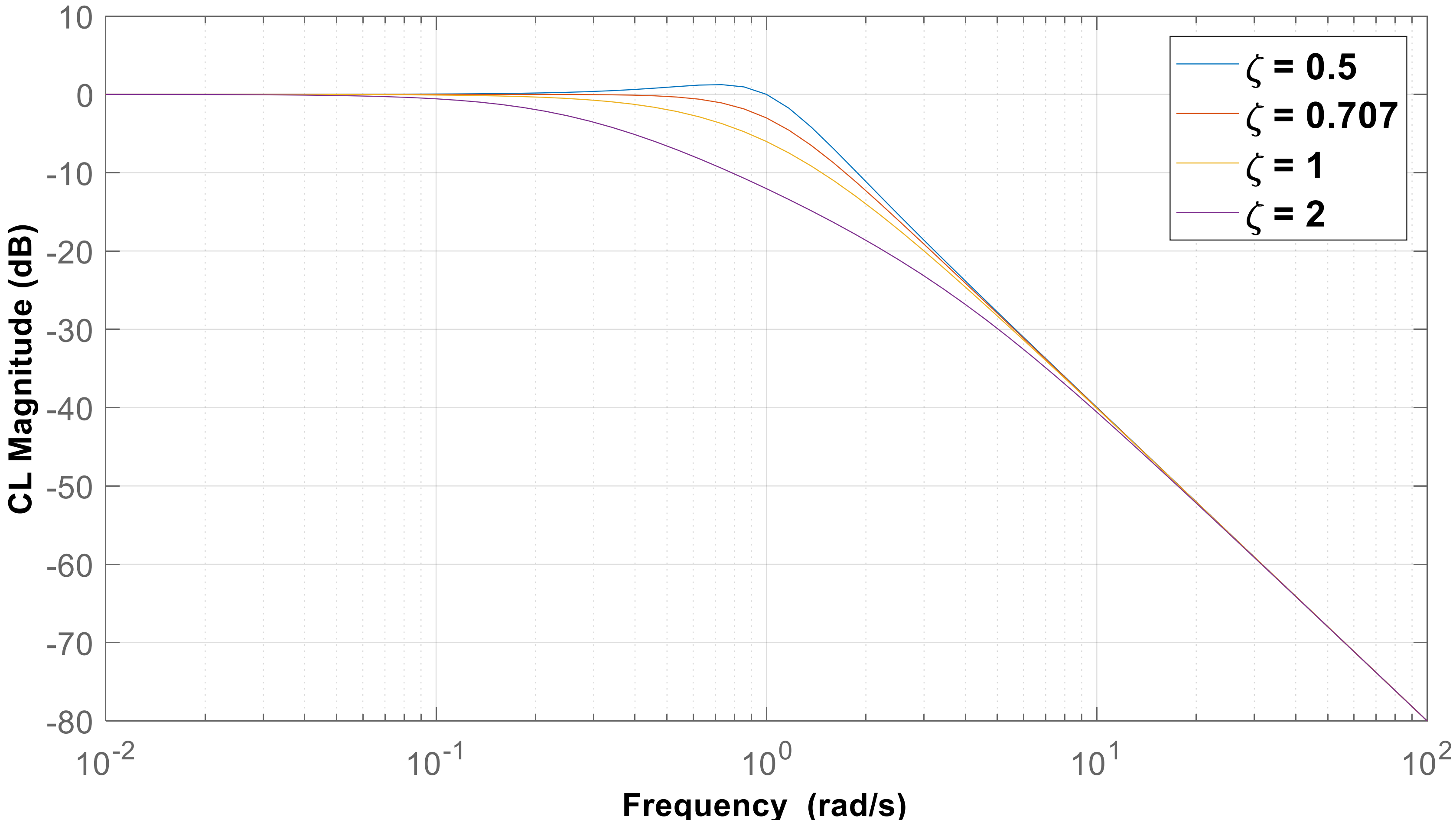




[Jespers & Murmann, 2017]

# Optimum Phase Margin (Freq Response)

- Maximum CL BW <u>without peaking</u> occurs at $\zeta = Q = 0.707$
  - Maximally flat response



# Optimum Phase Margin (Freq Response)

- Maximum CL BW without peaking occurs at $\zeta = Q = 0.707$
  - Maximally flat response
  - $\omega_{p2} = 2\omega_{u1}$ and $PM \approx 65^o$

| $\omega_{p2}/\omega_{u1}$ | $Q$ | $\omega_u/\omega_{u1}$ | Phase margin (°) |
|---|---|---|---|
| 1 | 1 | 0.786 | 51.8 |
| 2 | 0.707 | 0.910 | 65.5 |
| 3 | 0.577 | 0.953 | 72.4 |
| **4** | **0.500** | **0.972** | **76.3** |
| 5 | 0.477 | 0.981 | 78.9 |
| 6 | 0.408 | 0.987 | 80.7 |
| 7 | 0.378 | 0.990 | 81.9 |
| 8 | 0.354 | 0.992 | 82.9 |
| 9 | 0.333 | 0.994 | 83.7 |
| 10 | 0.316 | 0.995 | 84.3 |
| $\infty$ | — | 1 | 90 |

[Jespers & Murmann, 2017]

# Optimum Phase Margin (Freq Response)

- Maximum CL BW without peaking occurs at $\zeta = Q = 0.707$
  - Maximally flat response
  - $\omega_{p2} = 2\omega_{u1}$ and $PM \approx 65^o$
- But $\zeta < 1$: Underdamped system
  - Overshoot exists in transient response

# Optimum Phase Margin (Tran Response)

- Fastest settling <u>without overshoot</u> occurs at $\zeta = 1$ ($Q = 0.5$)
  - Critical damped system
  - $\omega_{p2} = 4\omega_{u1}$ and $PM \approx 76^o$

$\omega_u$ real or actual

| $\omega_{p2}/\omega_{u1}$ | $Q$ | $\omega_u/\omega_{u1}$ | Phase margin (°) |
|---|---|---|---|
| 1 | 1 | 0.786 | 51.8 |
| 2 | 0.707 | 0.910 | 65.5 |
| 3 | 0.577 | 0.953 | 72.4 |
| **4** | **0.500** | **0.972** | **76.3** |
| 5 | 0.477 | 0.981 | 78.9 |
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| 10 | 0.316 | 0.995 | 84.3 |
| ∞ | — | 1 | 90 |

$\omega_{p2} \neq \omega_{p1}$

1st order system



[Jespers & Murmann, 2017]

# Second-Order System Speed-up

- First order system corresponds to $\omega_{p2}/\omega_{u1} \to \infty$
  - Too much overdamping
- The settling of 2nd order system with optimum PM is faster
  - Critical damped system is faster than overdamped system
- But we must take some extra margin to account for variations

| Dynamic settling error ($\varepsilon_d$) | $t_s/\tau$ ($\omega_{p2}/\omega_{u1} \to \infty$) | $t_s/\tau$ ($\omega_{p2}/\omega_{u1} = 4$) | Speedup (%) |
|---|---|---|---|
| 10% | 2.3 | 1.9 | 15.5 |
| 1% | 4.6 | 3.3 | 27.9 |
| 0.1% | 6.9 | 4.6 | 33.1 |
| 0.01% | 9.2 | 5.9 | 36.2 |



[Jespers & Murmann, 2017]

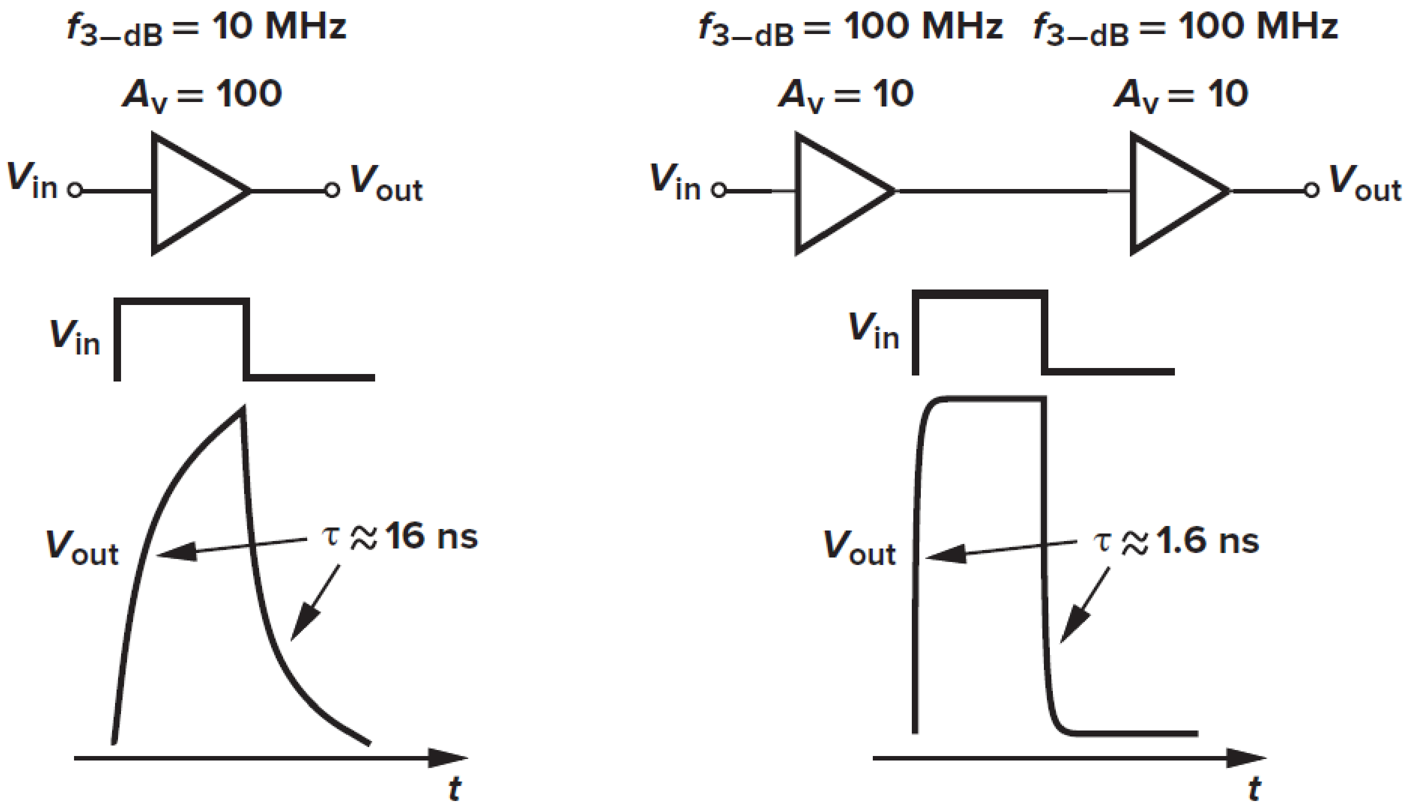
# Thank you!

# References


- A. Sedra and K. Smith, "Microelectronic Circuits," Oxford University Press, 7th ed., 2015
- B. Razavi, "Design Of Analog CMOS Integrated Circuit," 2nd ed., McGraw-Hill, 2017.
- T. C. Carusone, D. Johns, and K. W. Martin, "Analog Integrated Circuit Design," 2nd ed., Wiley, 2012.
- P. Jespers and B. Murmann, Systematic Design of Analog CMOS Circuits Using Pre-Computed Lookup Tables, Cambridge University Press, 2017.

# Bandwidth Extension

- Cascade of feedback amplifiers provides the same gain and a much faster response
  - But power consumption and area doubled






[Razavi, 2017]