20 December 2021

وَمَا أُوتِيتُمْ مِنَ الْعِلْمِ إِلَّا قَلِيلًا

16 جمادى الأولى 1443

Analog IC Design

Lecture 15
Negative Feedback

Dr. Hesham A. Omran

Integrated Circuits Laboratory (ICL)
Electronics and Electrical Communications Eng. (EECE) Dept.
Faculty of Engineering
Ain Shams University

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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- ❑ Recapping previous key results
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- ❑ Phase and gain margin

MOSFET in Saturation

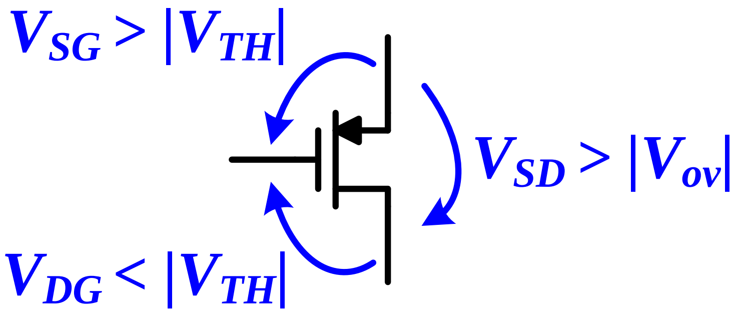
- The channel is pinched off if the difference between the gate and drain voltages is not sufficient to create an inversion layer

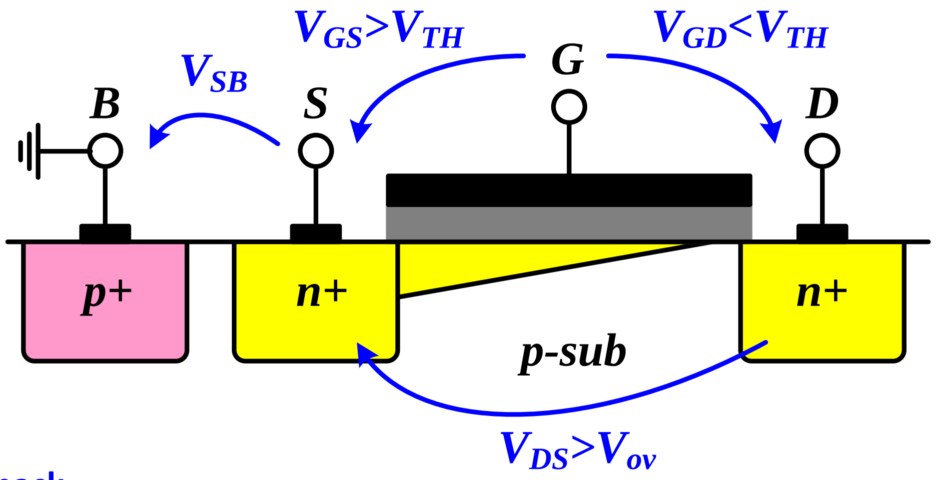
$$V_{GD} \le V_{TH} \quad OR \quad V_{DS} \ge V_{ov}$$

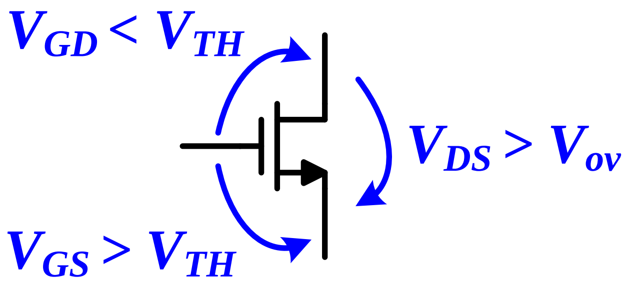
- Square-law (long channel MOS)

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

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







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

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

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

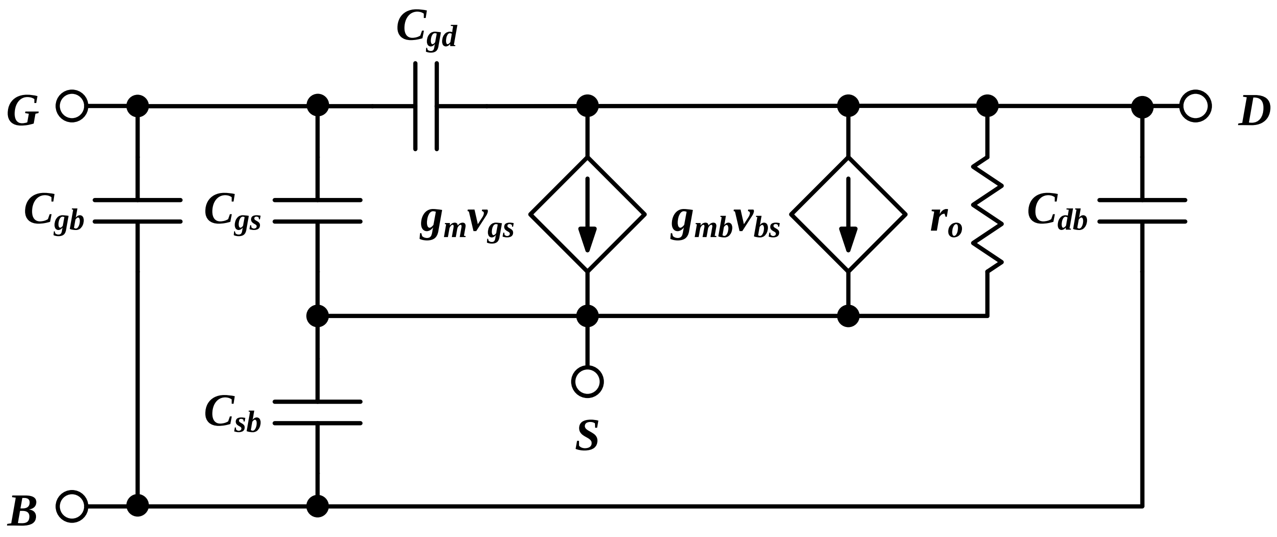
High Frequency Small Signal Model

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

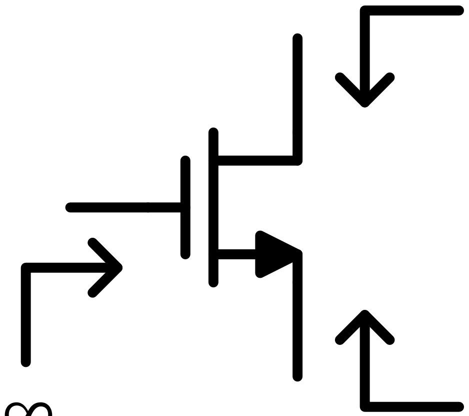
$$g_{mb} = \eta g_m \qquad \eta \approx 0.1 - 0.25$$

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

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



Rin/out Shortcuts Summary



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

H.I.N.

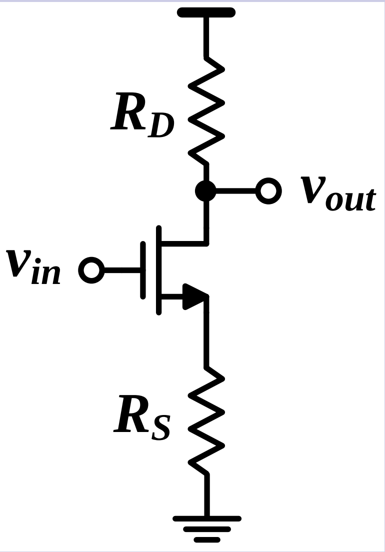
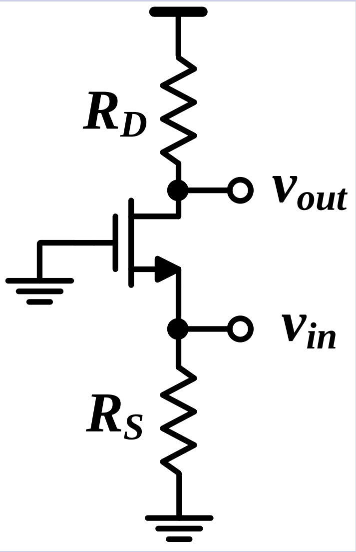
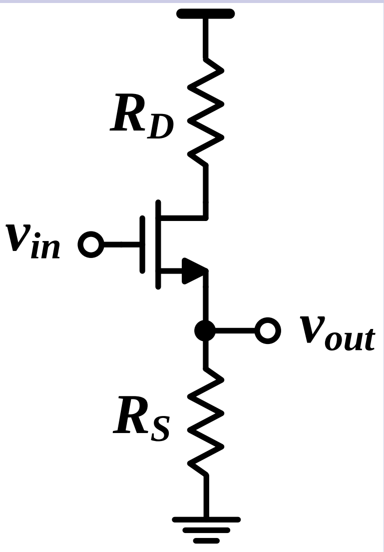
∞

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 Current buffer	Voltage buffer Current amplifier
Rin	∞	$R_S \| \| \dfrac{1}{g_m + g_{mb}}\left(1 + \dfrac{R_D}{r_o}\right)$	∞
Rout	$R_D \| \| r_o[1 + (g_m + g_{mb})R_S]$	$R_D \| \| r_o$	$R_S \| \| \dfrac{1}{g_m + g_{mb}}\left(1 + \dfrac{R_D}{r_o}\right)$
Gm	$\dfrac{-g_m}{1 + (g_m + g_{mb})R_S}$	$g_m + g_{mb}$	$\dfrac{g_m}{1 + R_D/r_o}$

Differential Amplifier

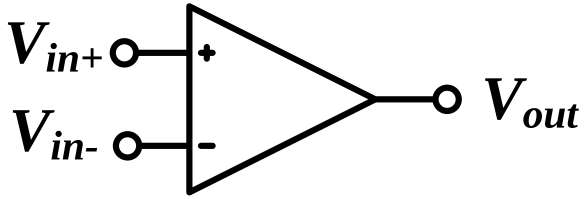
	Pseudo Diff Amp	Diff Pair (w/ ideal CS)	Diff Pair (w/ R_{SS})
A_{vd}	$-g_m R_D$	$-g_m R_D$	$-g_m R_D$
A_{vCM}	$-g_m R_D$	0	$\dfrac{-g_m R_D}{1 + 2(g_m + g_{mb})R_{SS}}$
A_{vd}/A_{vCM}	1	∞	$2(g_m + g_{mb})R_{SS} \gg 1$

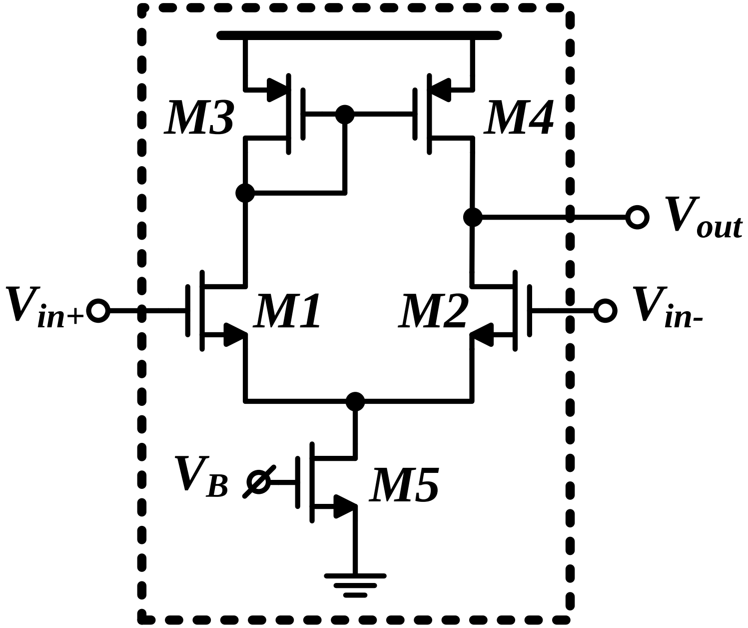
$$A_{vCM2d} = \frac{v_{od}}{v_{iCM}} \approx \frac{\Delta R_D}{2R_{SS}} + \frac{\Delta g_m R_D}{2g_{m1,2} R_{SS}}$$

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

Op-Amp

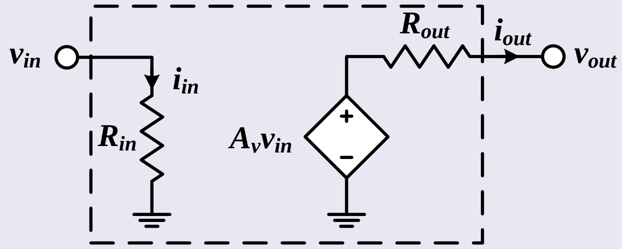
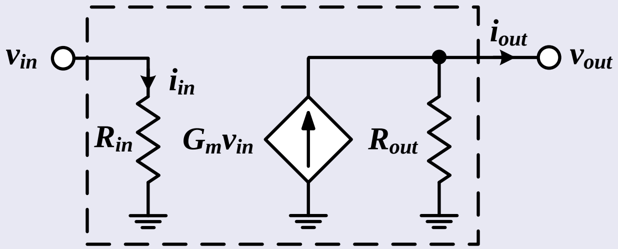
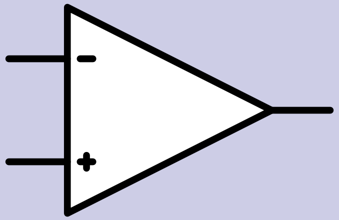
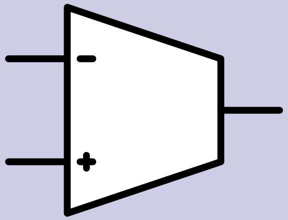
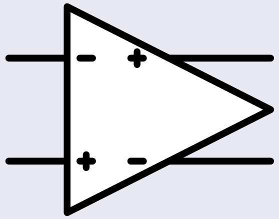
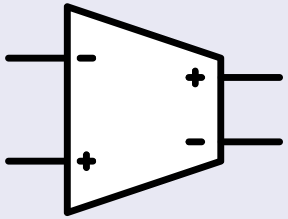
- An op-amp is simply a high gain differential amplifier
  - The gain can be increased by using cascodes and multi-stage amplification
- The diff amp is a key block in many analog and RF circuits
  - DEEP understanding of diff amp is ESSENTIAL





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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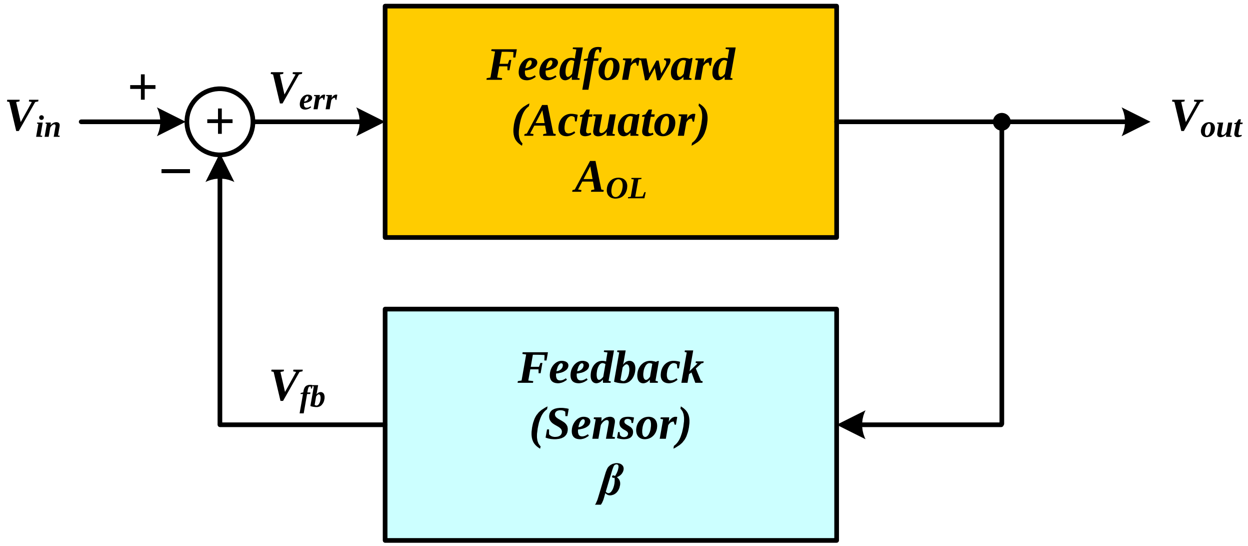
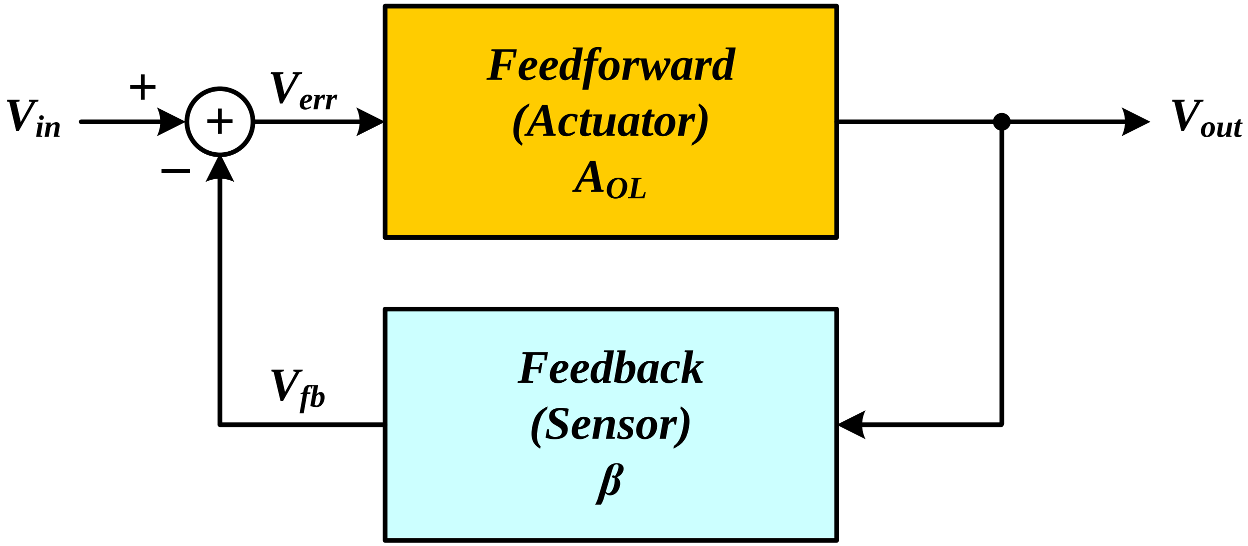
General Feedback System



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

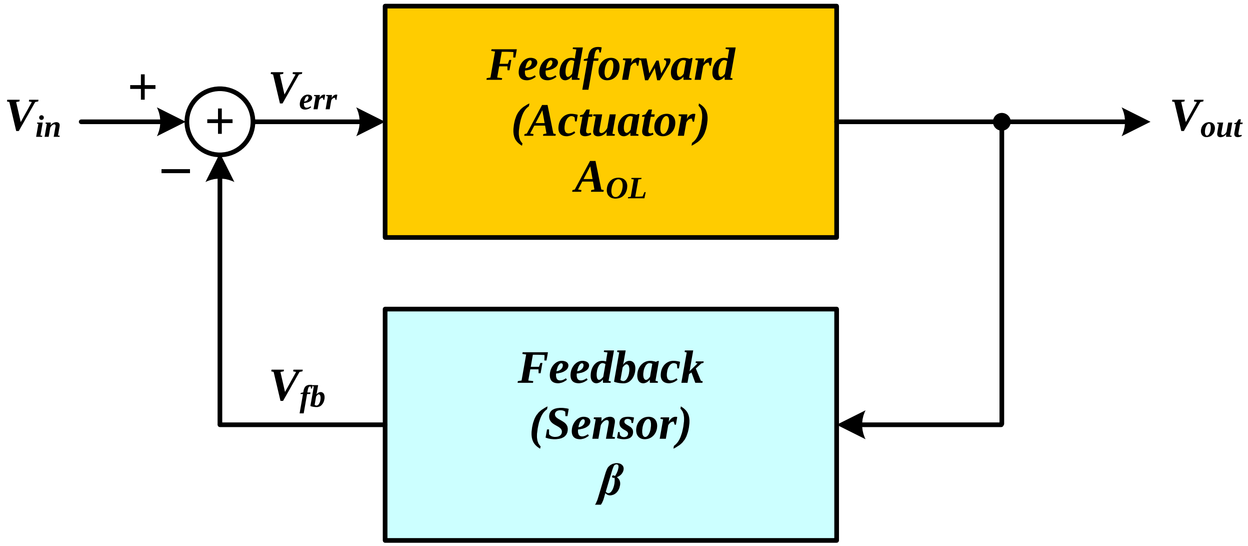
$$V_{out} = A_{OL}\left(V_{in} - V_{fb}\right) = A_{OL}(V_{in} - \beta V_{out})$$

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

- 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

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

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

$V_{out} = V_{err} \times A_{OL}$

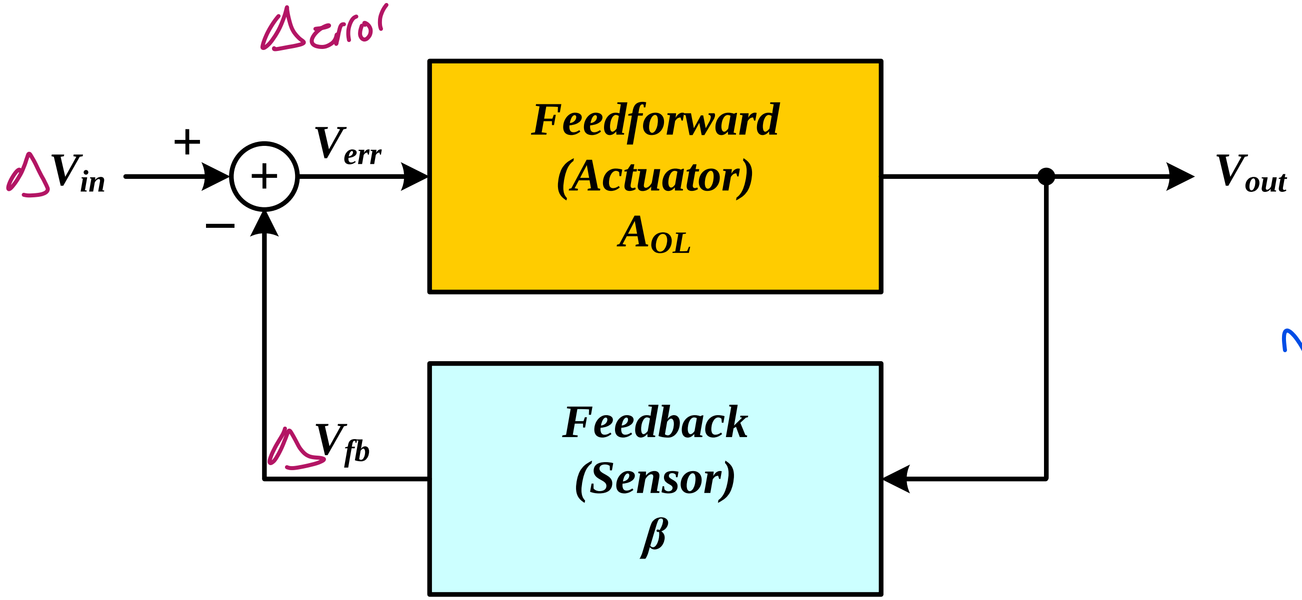
$= 0 \times \infty \rightarrow$ unspecified

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

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

$\left. V_{err} = \frac{V_{out}}{A_{OL} \rightarrow \infty} \right\} \rightarrow 0$

- Negative feedback loop works to minimize the error signal



لما ال Gain يكبر

ال Vin

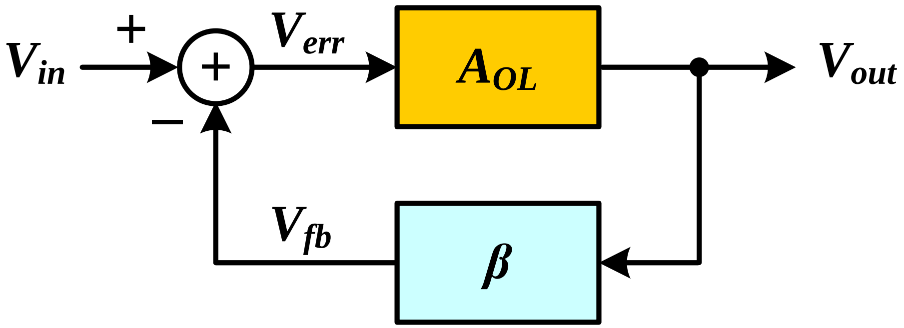
بتصغر

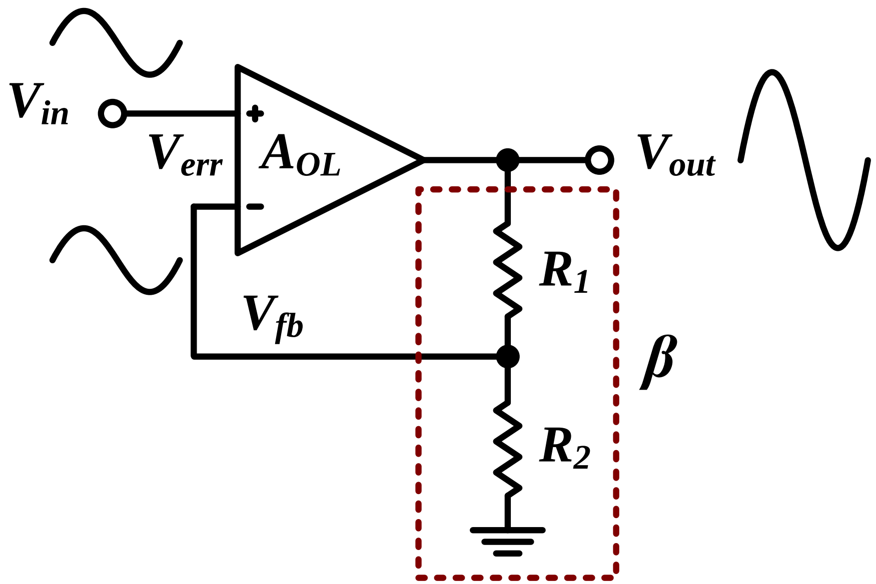
فكرلوا !!

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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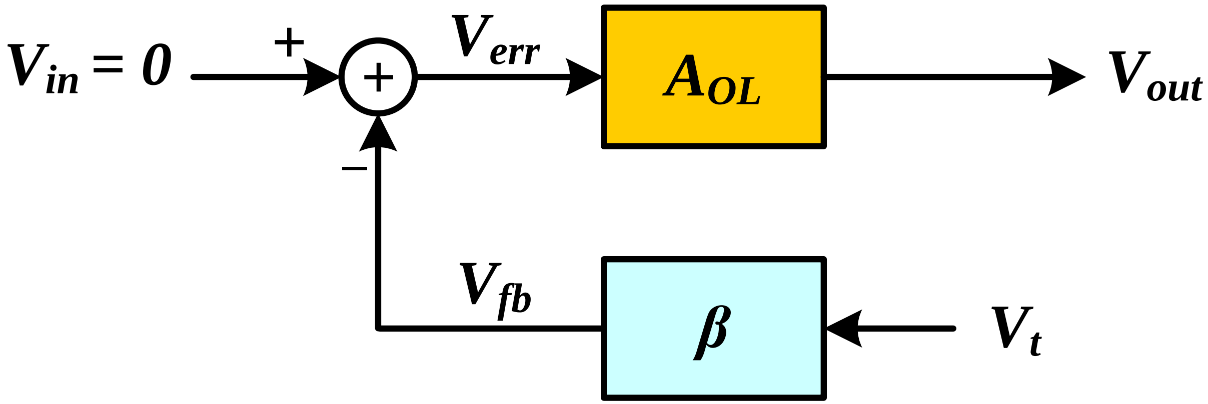
Loop Gain

- 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



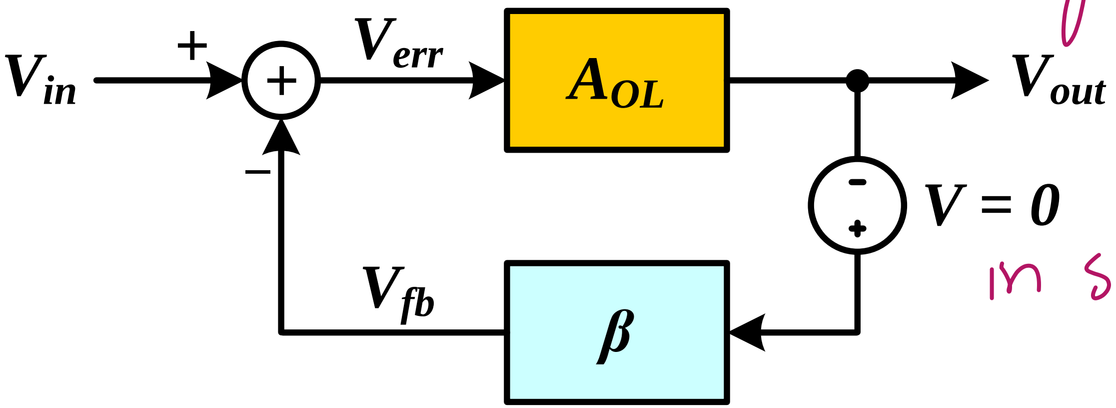
Loop Gain

- 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

Look for "middle break" negative feedback paper for evaluating LG in simulation and lab

Paper is in 70s

also the method is in Razavi



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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

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

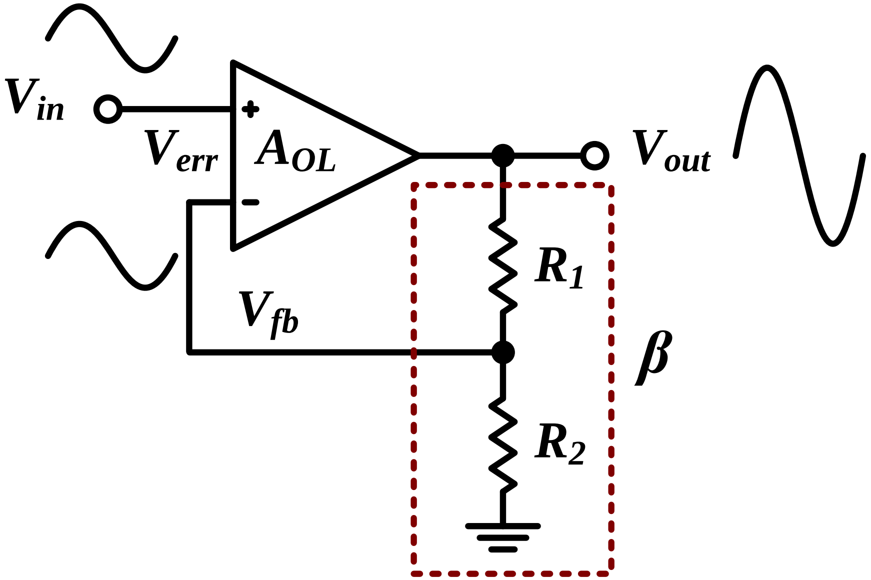
- A_{OL} varies due to PVT, load, and input signal variations
- A_{CL} almost independent of A_{OL} (if $LG \gg 1$)
  - Independent of PVT: stable and robust
  - Independent of load: stable and robust
  - Independent of input level: linear

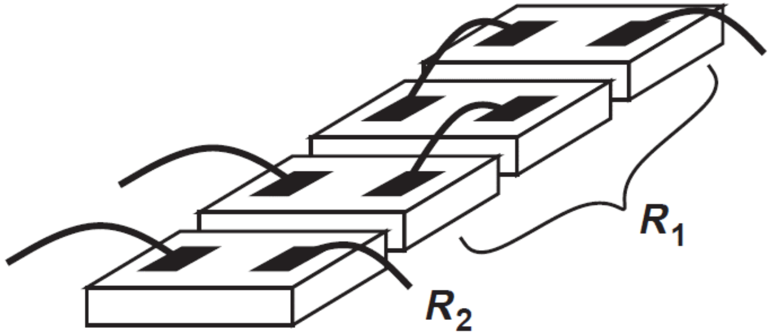
Gain Desensitization

- 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

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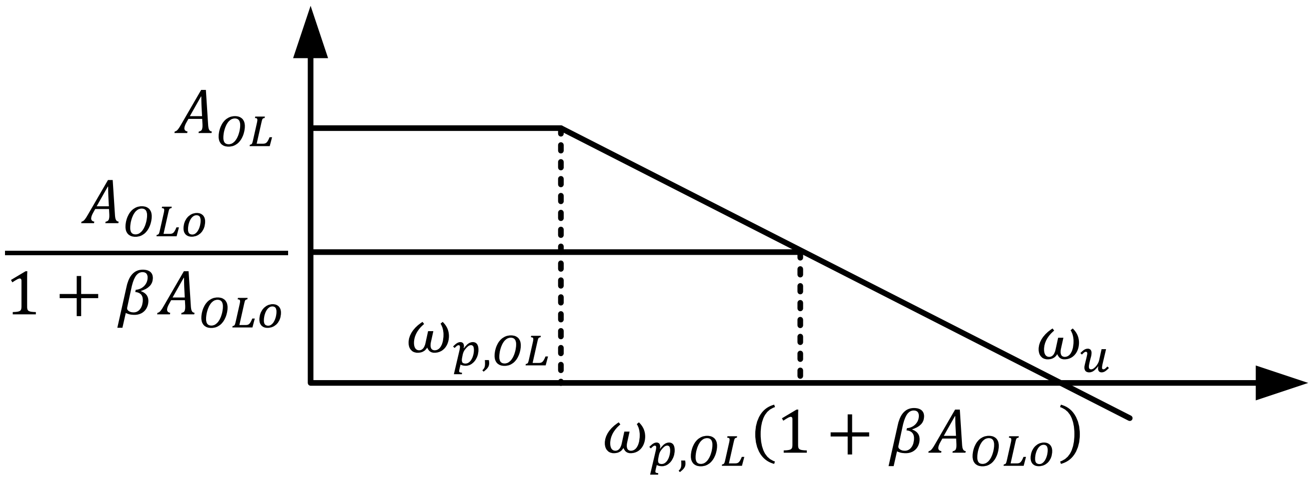
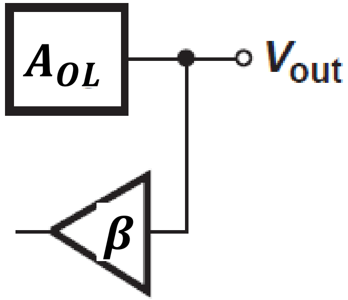
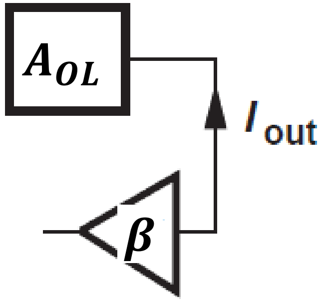


Modification of I/O Impedances

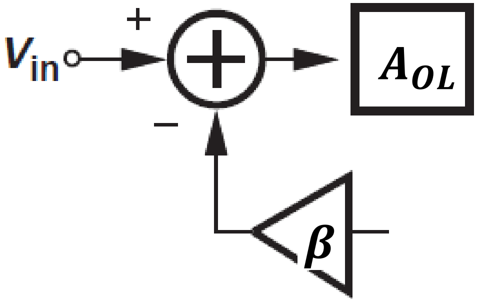
- Shunt sensing/mixing → R decreases
- Series sensing/mixing → R increases



Output impedance falls by 1+ loop gain.



Output impedance rises by 1+ loop gain.



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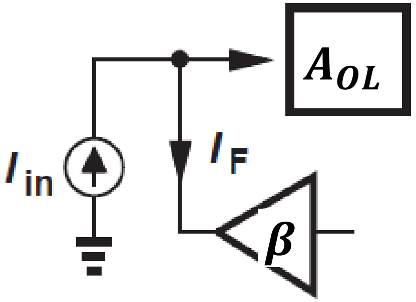


Input impedance falls by 1+ loop gain.

[Razavi, 2017]

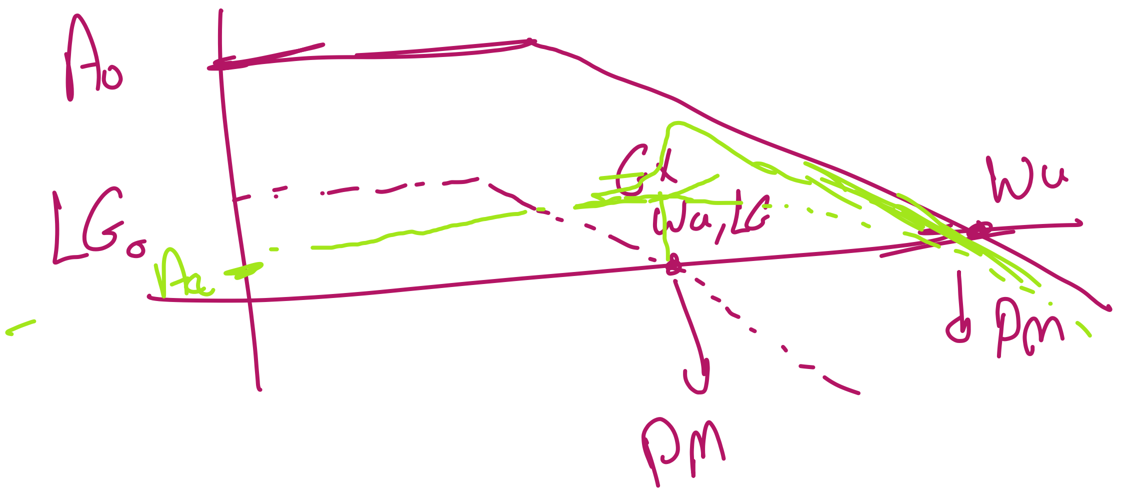
$A_{oL} \omega_{oL} = A_{cL} \omega_L$ → $GBW = \omega_{u,oL}$

important Relation.

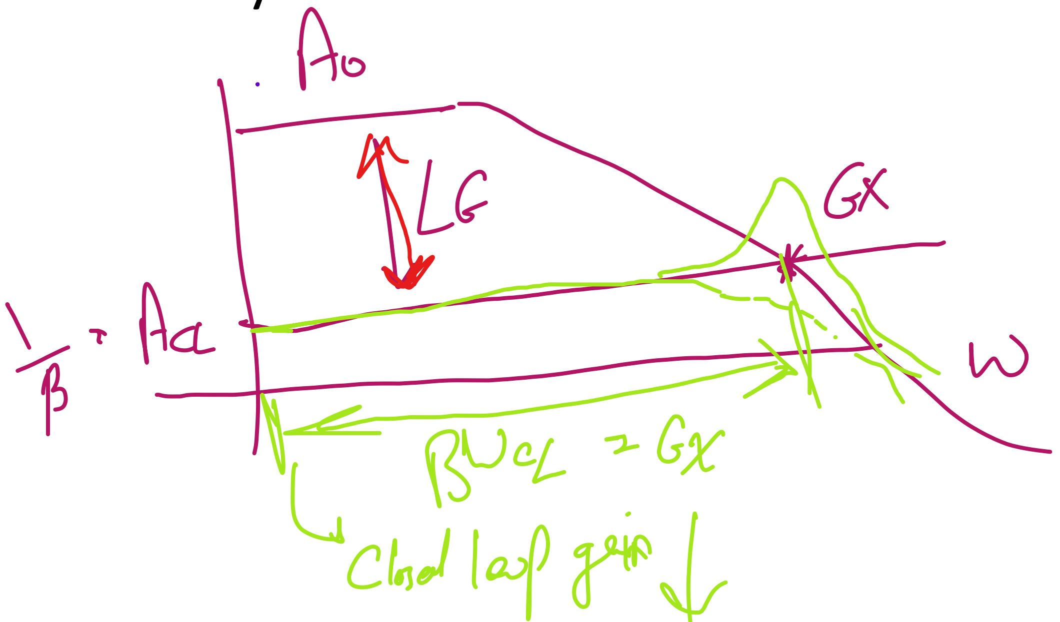
The Price We Pay

- 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

$A_{oL} = LG|_{\beta=1}$



Dr. prefers this one.



$LG = \beta A$

$\log A - \log \frac{1}{\beta}$

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closed loop gain Plot

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$$LG_0 = \beta A_0$$

$$GX = \beta \omega_{u,OL}$$

$$= \frac{\omega_u}{1/\beta} = \frac{\omega_{u,OL}}{A_{CL}}$$

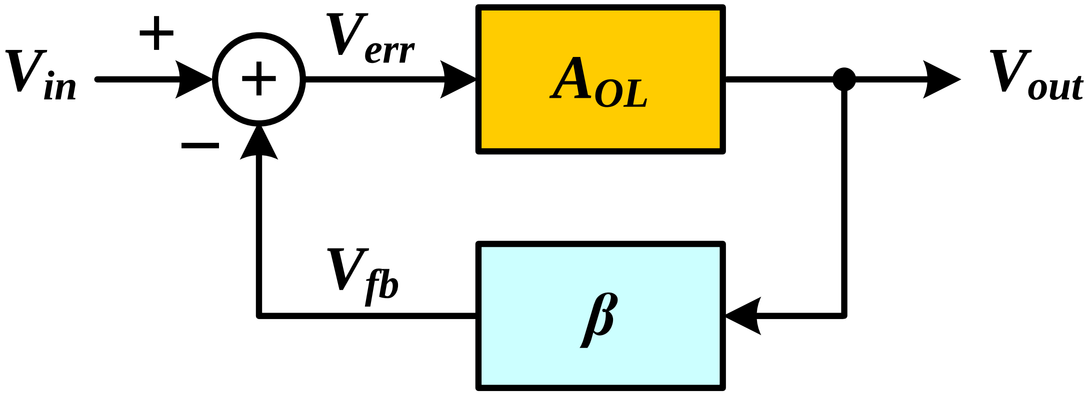
$$GX = \beta \omega_{CL}$$

Stability of Feedback System

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

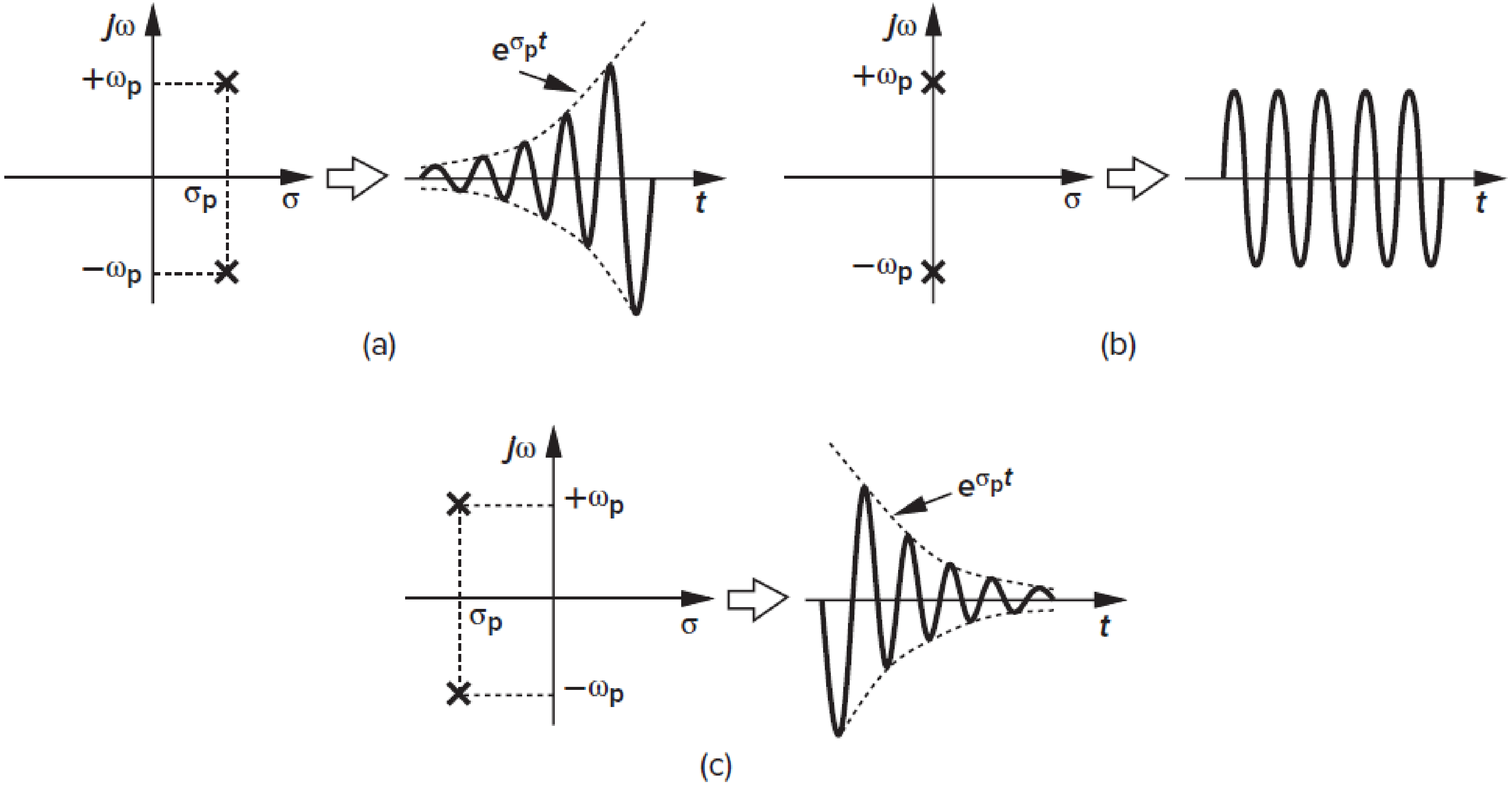
- Barkhausen's Oscillation Criteria

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



Stable vs Unstable System: Pole-Zero Plot

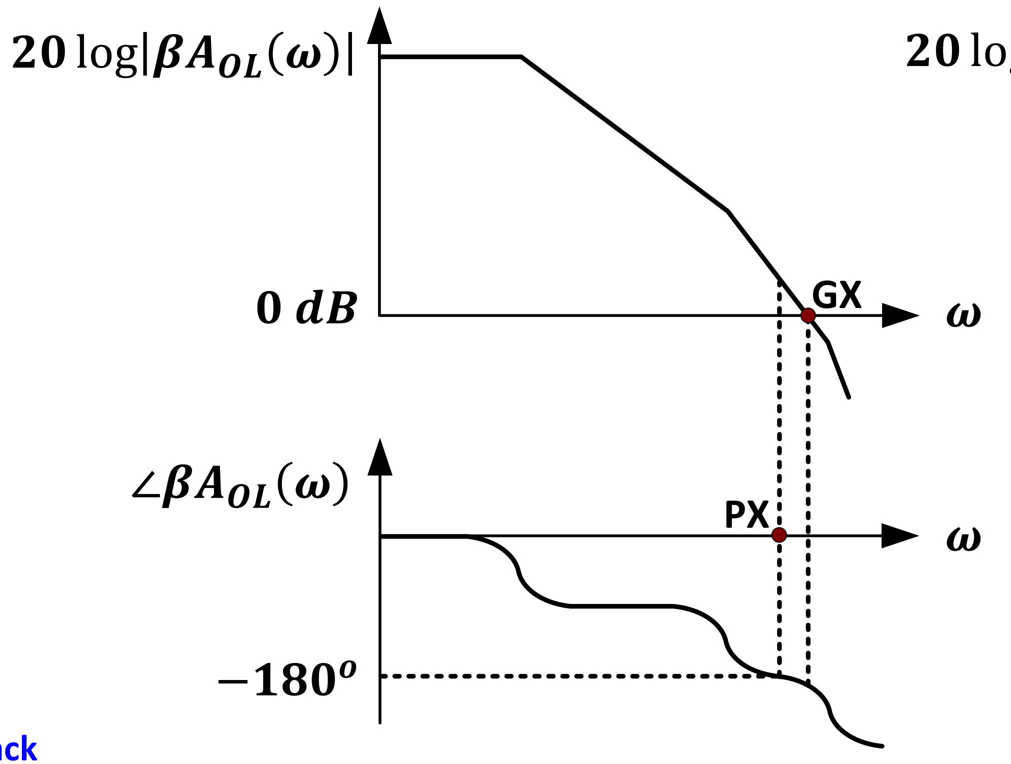
Laplace domain	Time domain
$\frac{1}{s-a}$	e^{at}

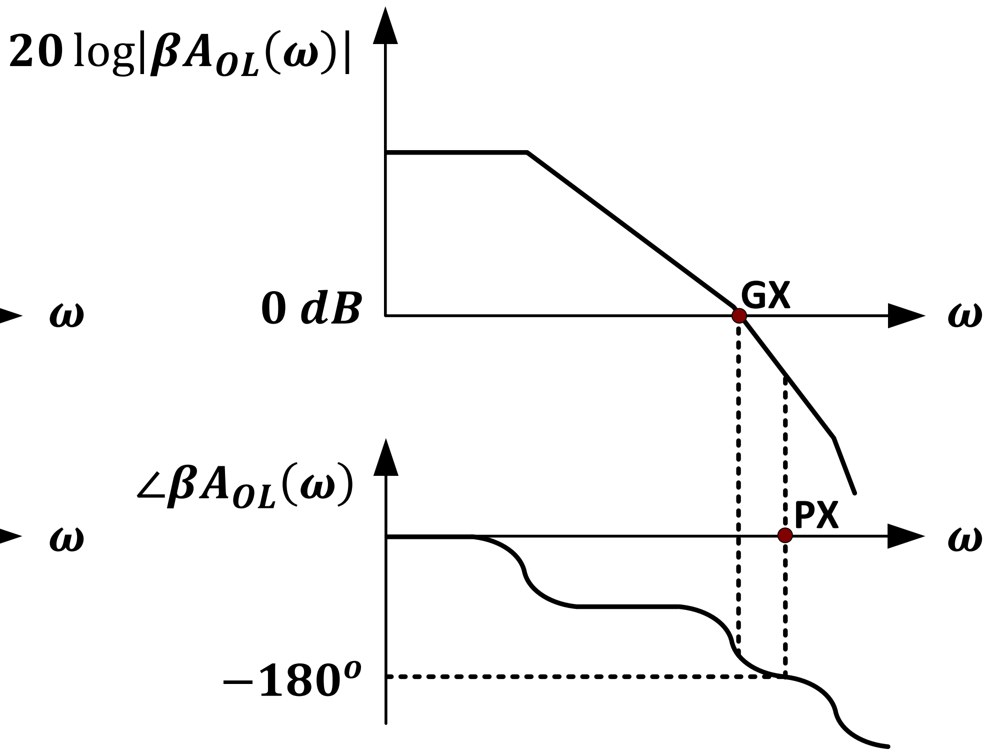


[Razavi, 2017]

Stable vs Unstable System: Bode Plot

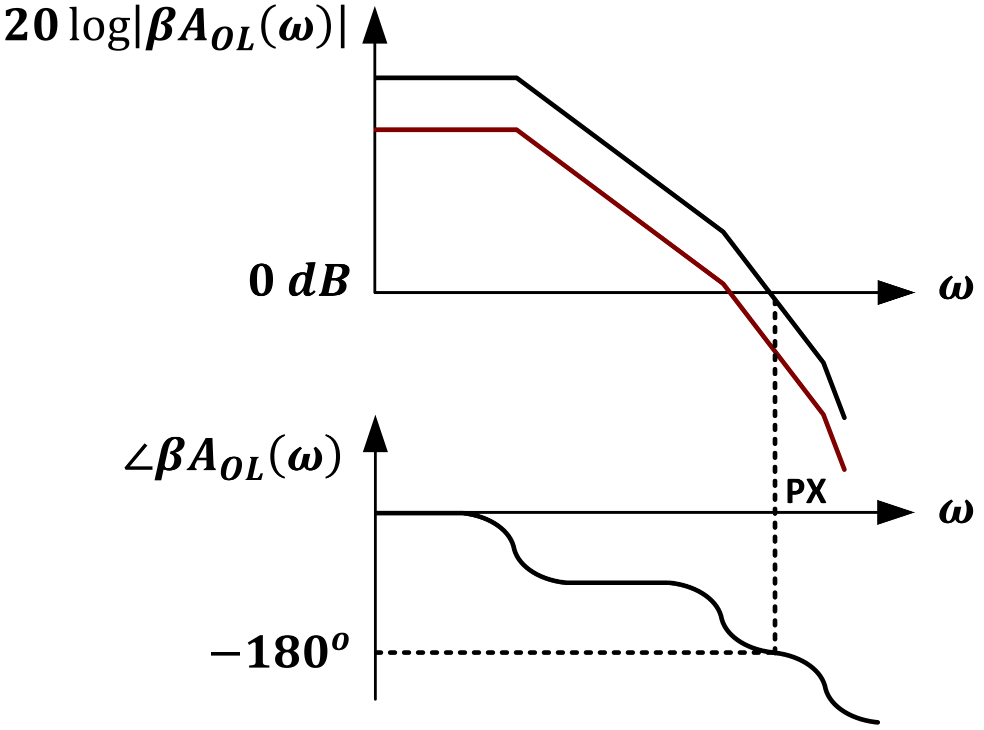
- Gain crossover frequency (GX): @ $|\beta A_{OL}(s)| = 1$
  - Same as ω_u
- Phase crossover frequency (PX): @ $\angle\beta A_{OL}(s) = -180$
- For a stable system: GX < PX





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

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



Outline

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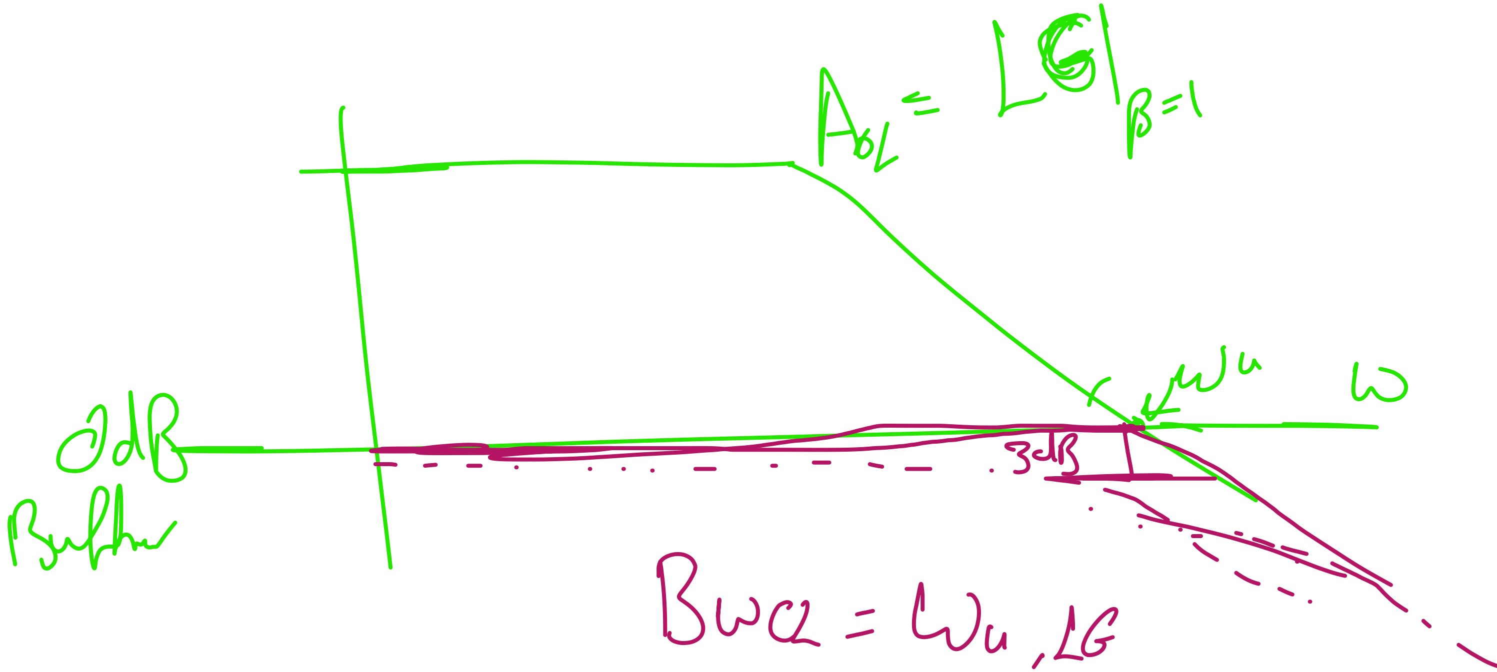
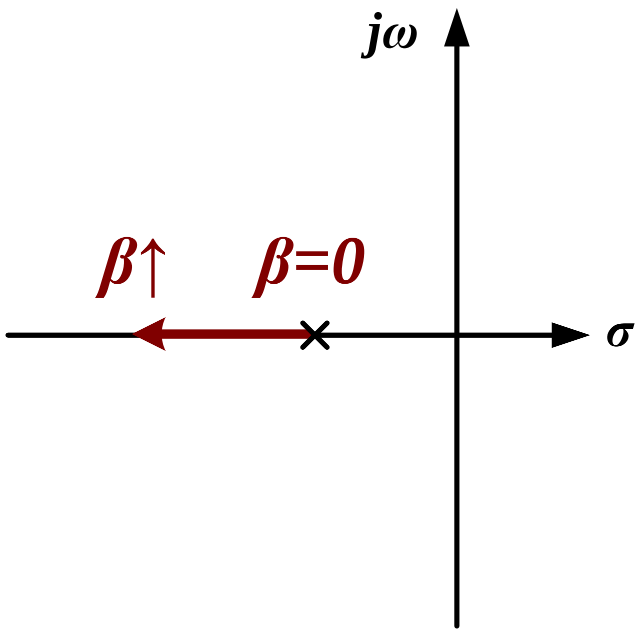
- ❑ Loop gain
- ❑ Why negative feedback?
- ❑ Stability of feedback system
- ❑ Root locus and Bode plot
- ❑ Phase and gain margin



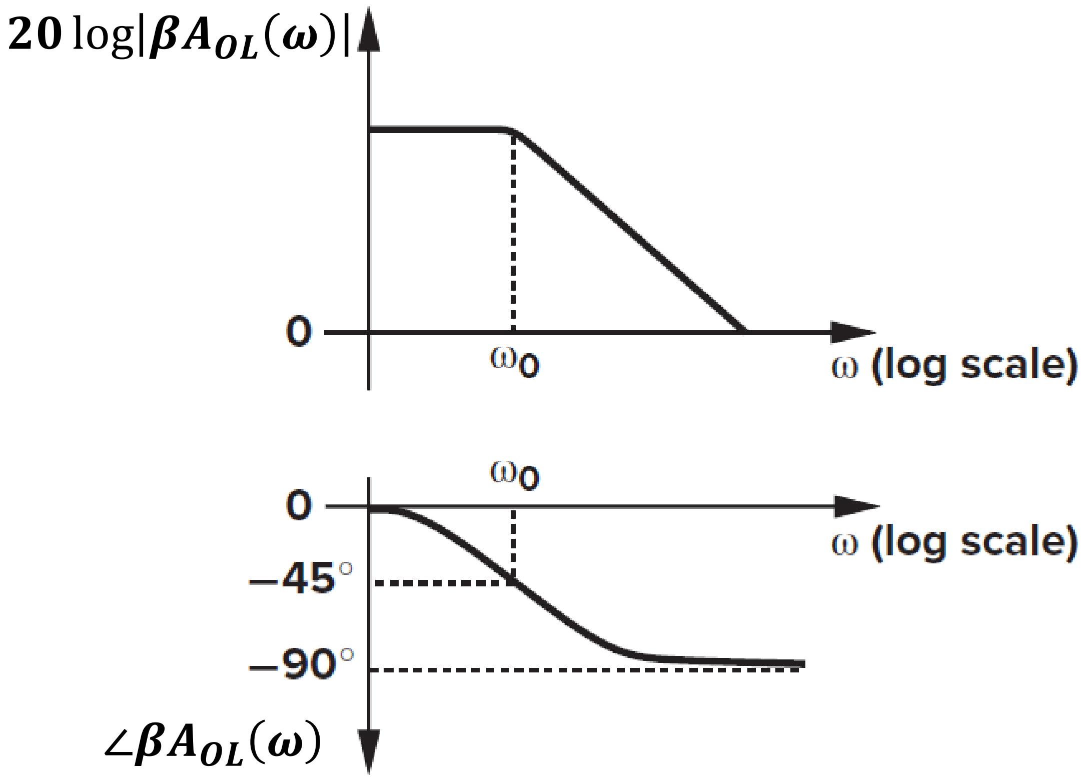
Single-Pole System: Root Locus

- Note that β does not affect poles of LG (assuming β 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

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



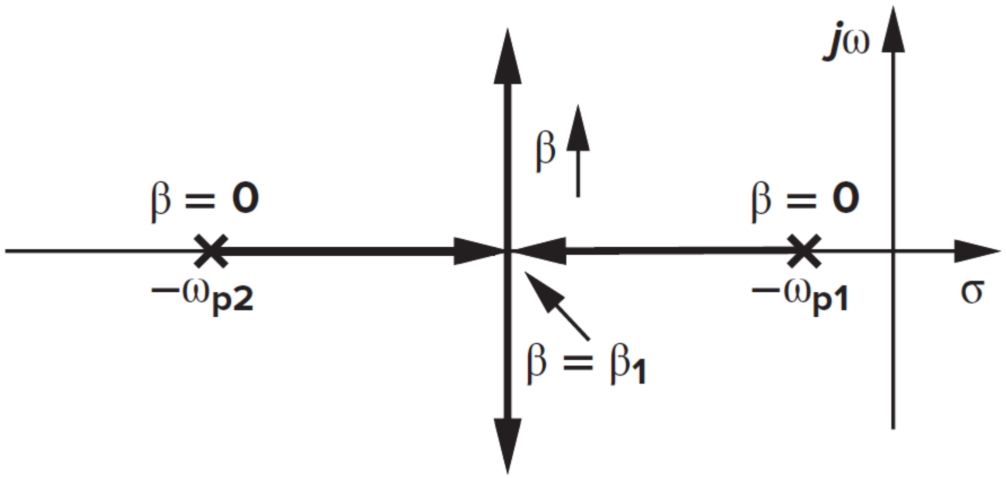
[Razavi, 2017]

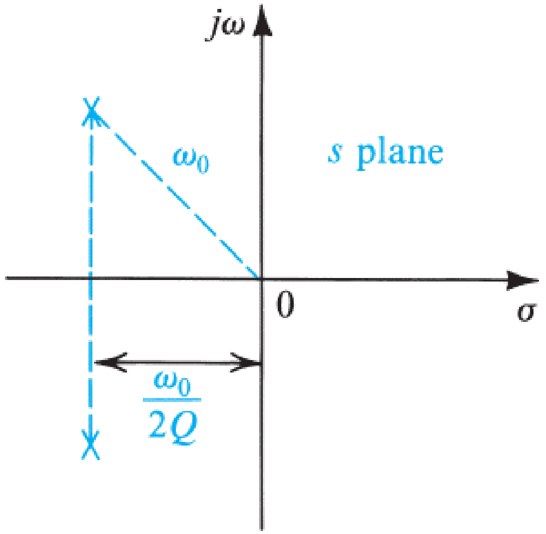
Two-Pole System: Root Locus

- 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)

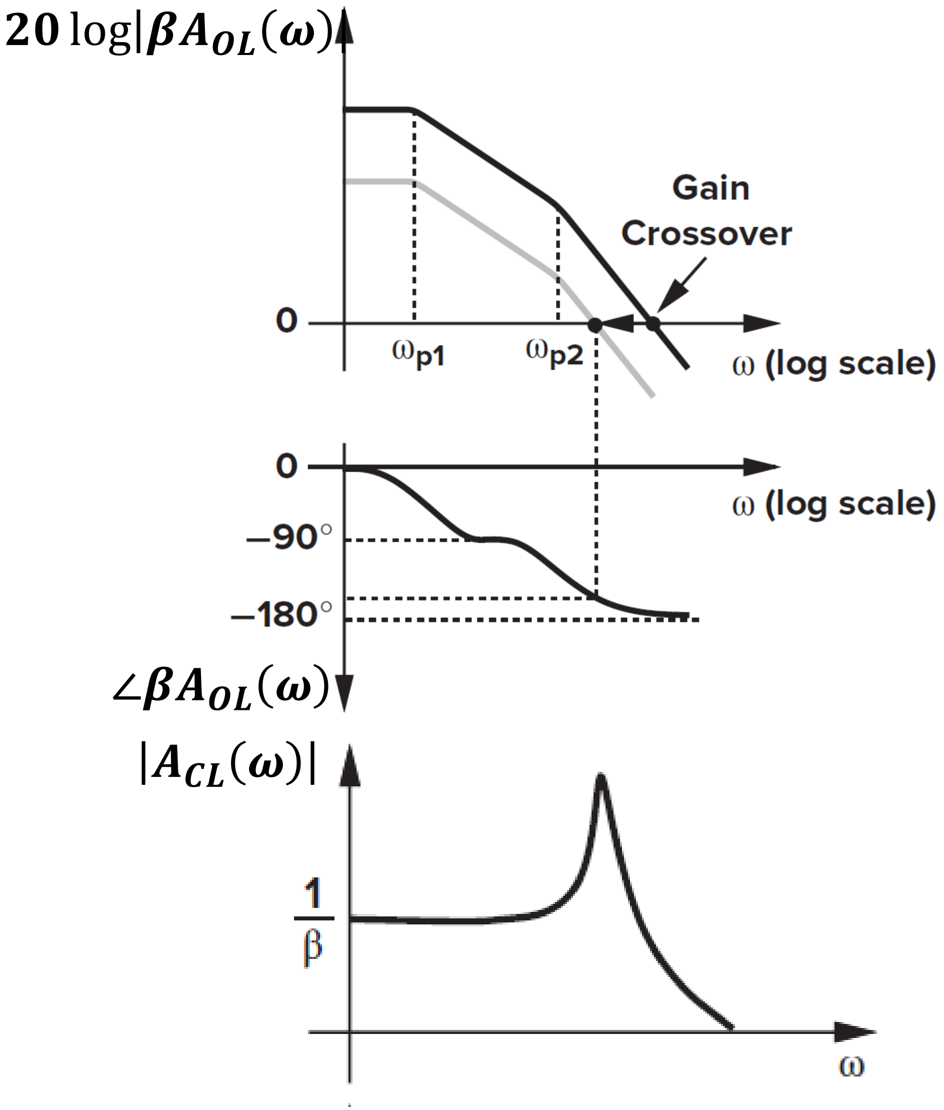




[Sedra/Smith, 2015] & [Razavi, 2017]

Two-Pole System: Bode Plot

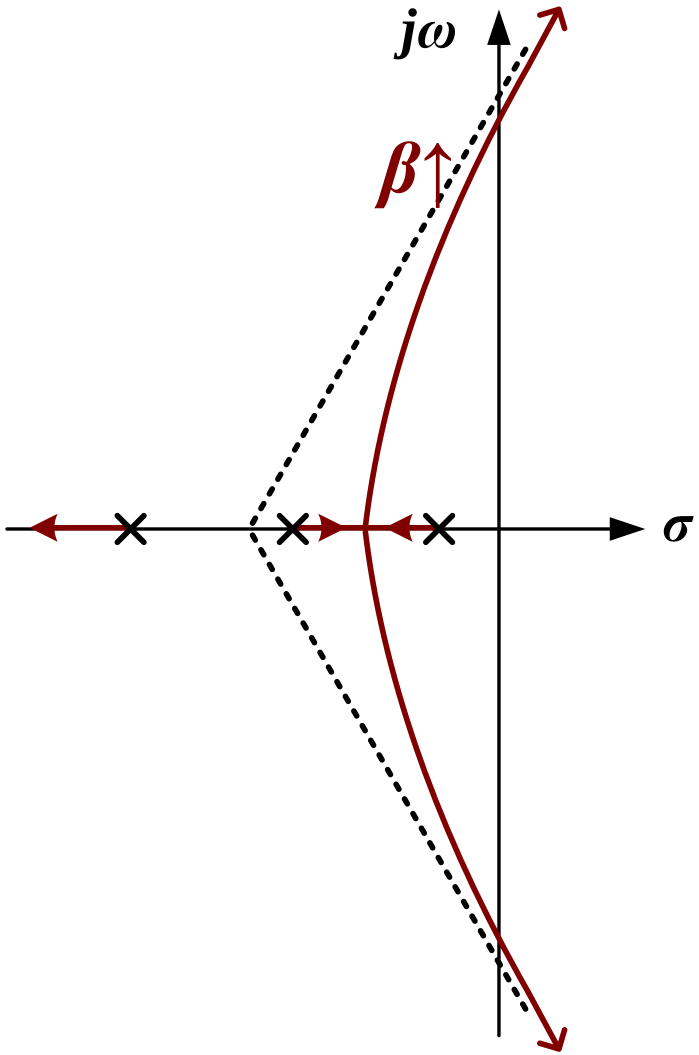
- Phase shift always < 180^o
  - Unconditionally stable
- $\zeta > 1$ ($Q < 0.5$): Overdamped
- $\zeta = 1$ ($Q = 0.5$): Critical damped
- $\zeta < 1$ ($Q > 0.5$): Underdamped
  - Overshoot in step response
- $\zeta < 1/\sqrt{2} = 0.707$
  - $Q > 1/\sqrt{2} = 0.707$
  - Peaking in frequency response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



[Razavi, 2017]

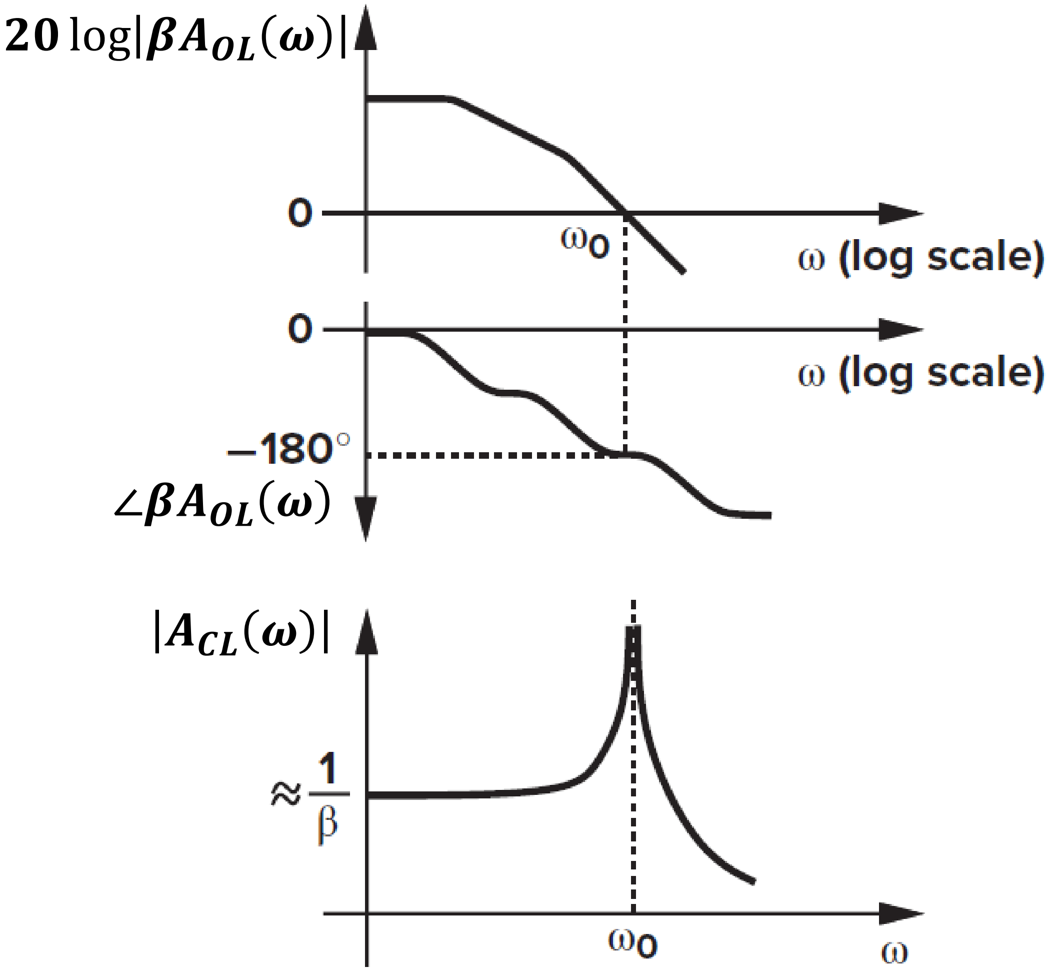
Three-Pole System: Root Locus

- 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?

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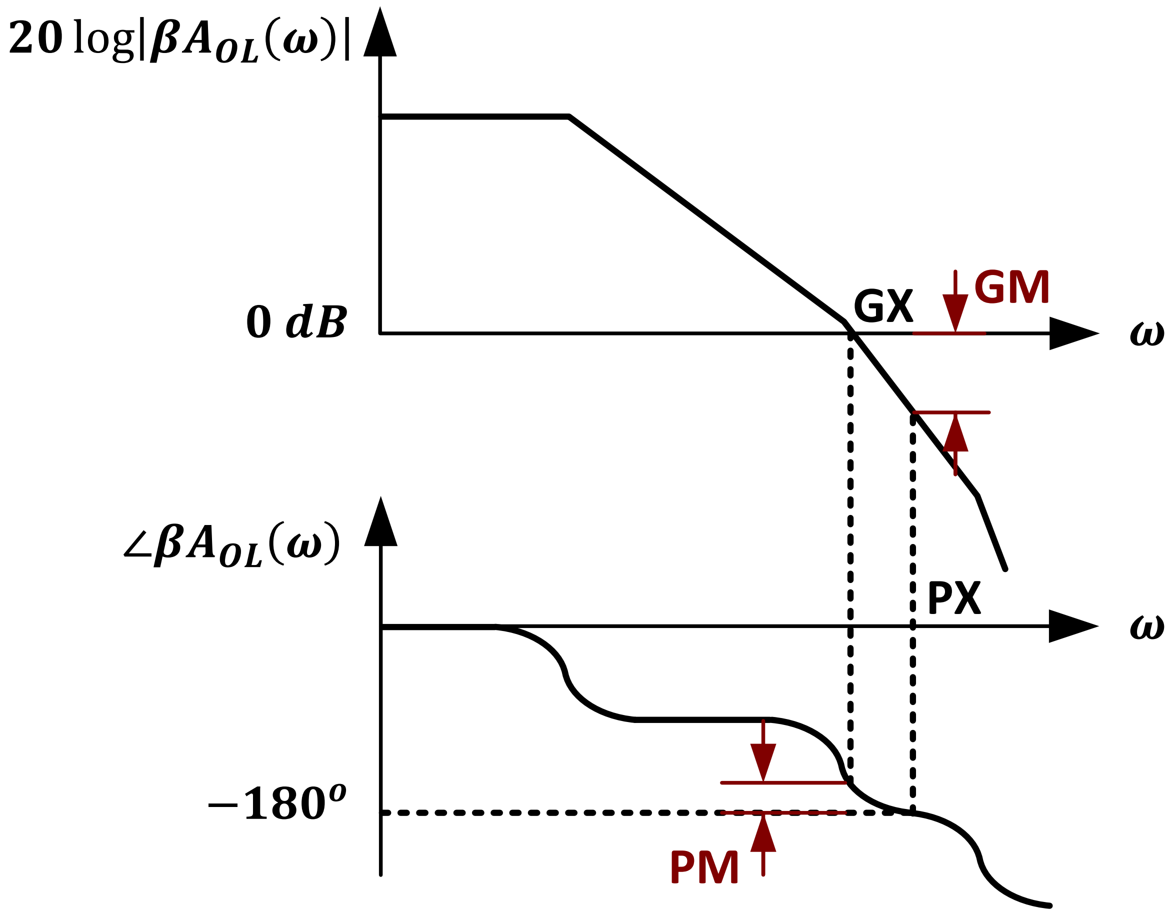
Phase and Gain Margin

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

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

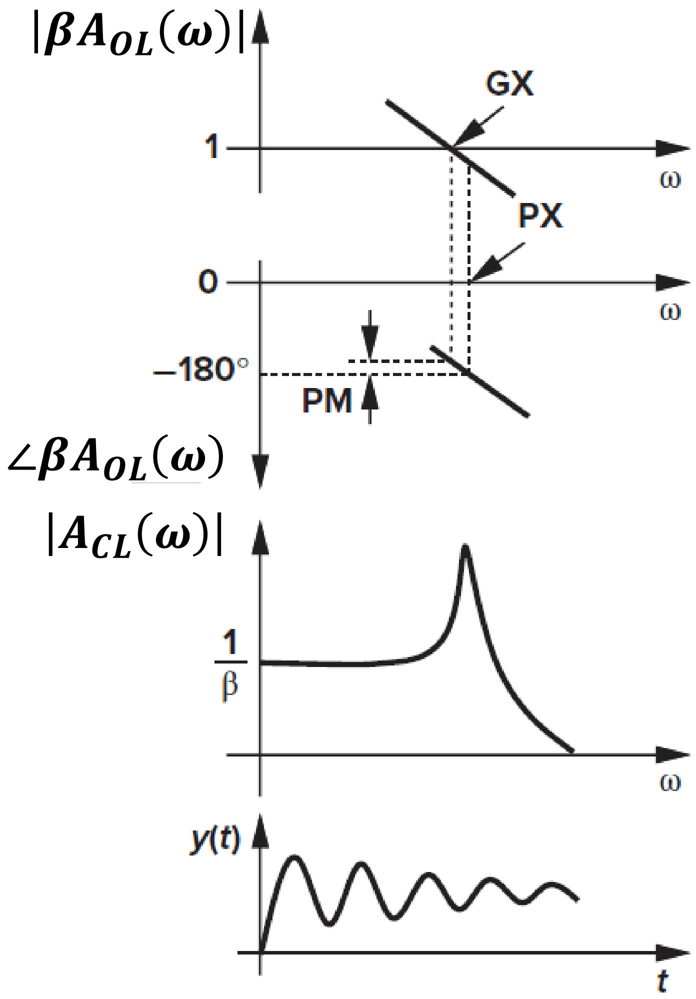


Phase Margin (PM)

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

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]

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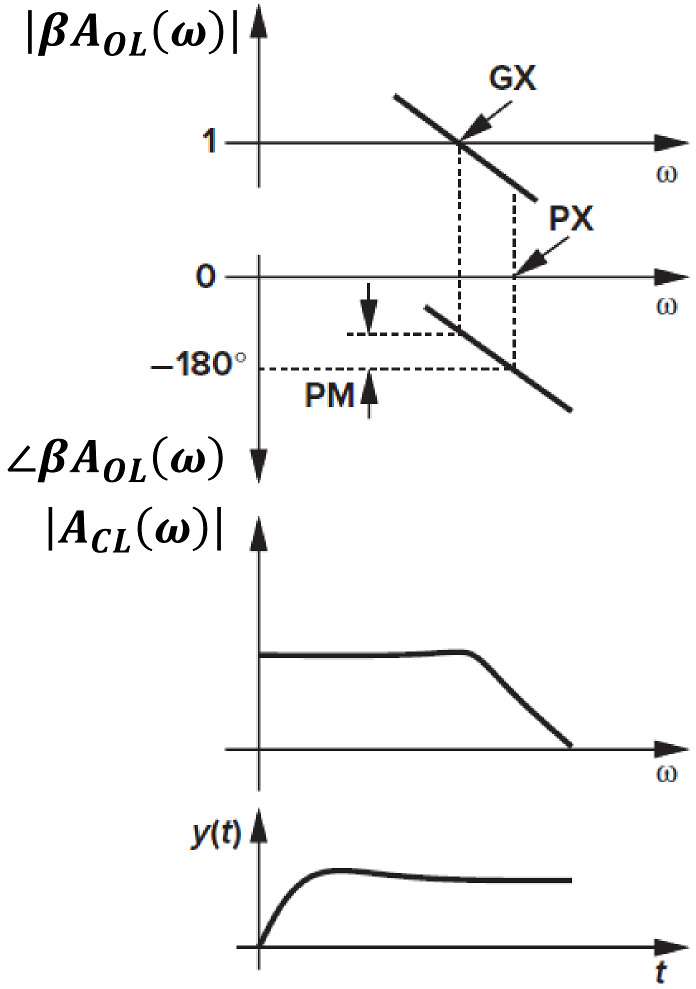
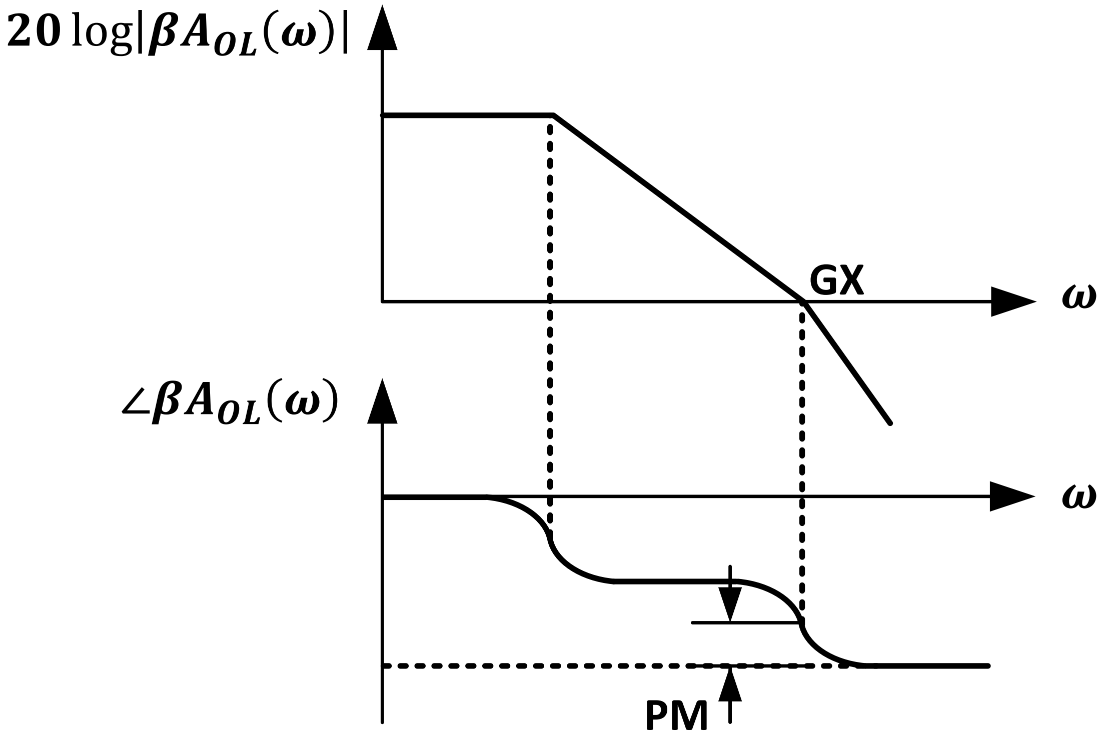


 [Razavi, 2017]

Phase Margin: Ultimate GBW

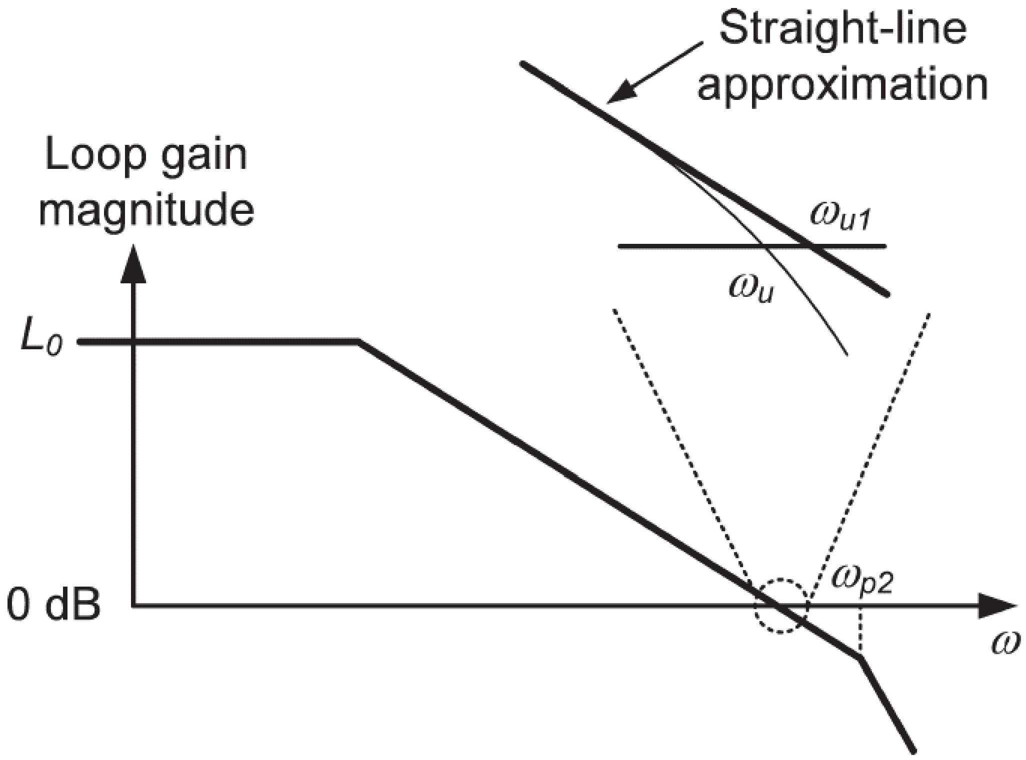
- If $\omega_{p2} = \omega_u$: PM = 45°
  - Typically inadequate (peaking/ringing)
- Thus ω_{p2} should be $> \omega_u$ → $\omega_{p1} \ll \omega_u < \omega_{p2}$
  - ω_{p1} defines OL BW and ω_{p2} defines ultimate GBW (max CL BW)

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



Asymptotic vs Actual UGF

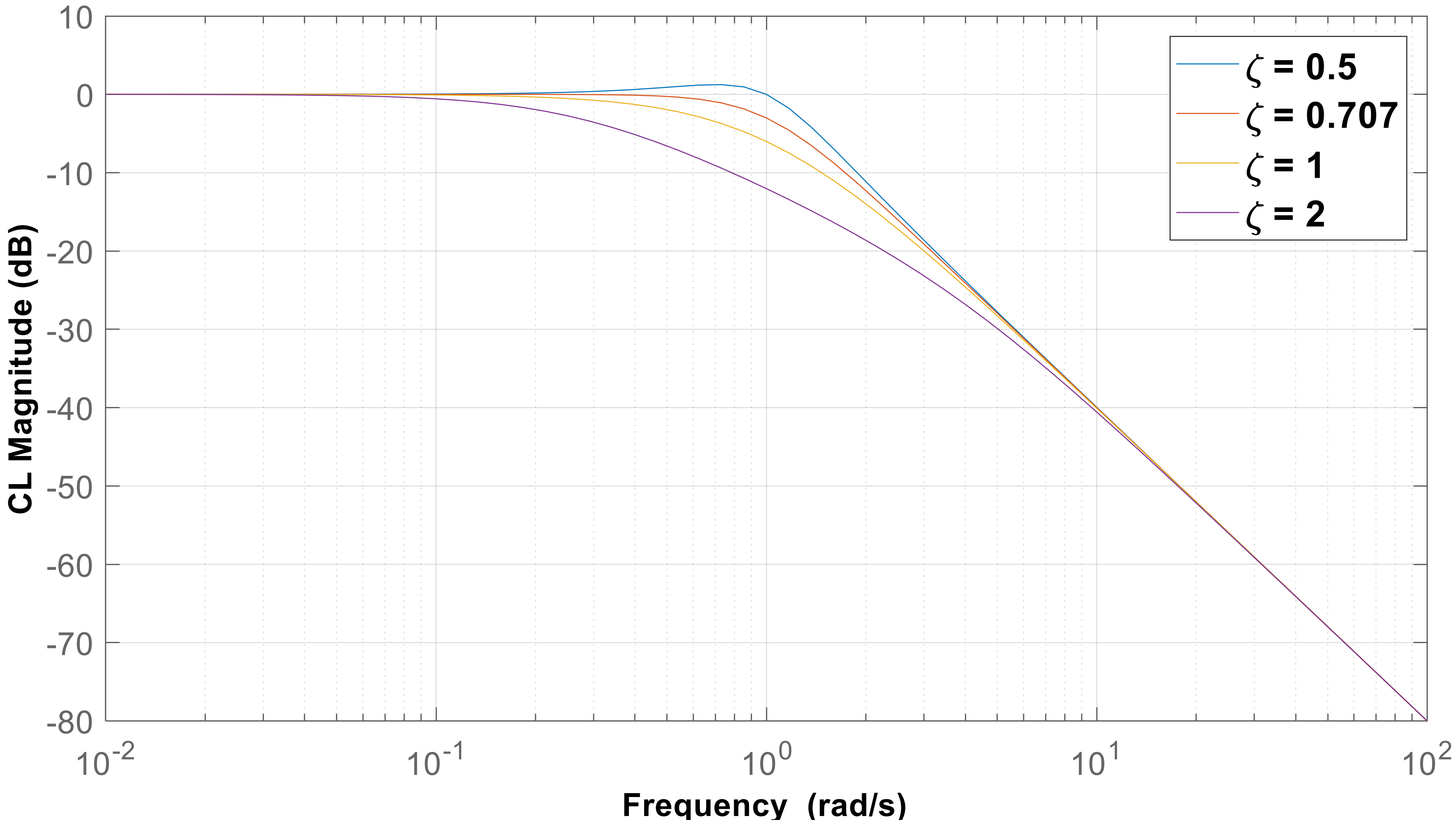
- For $\omega < \omega_u$ the Bode plot is similar to a 1st order system
- For a true 1st order system: GBW = UGF $= \omega_u = \omega_{u1}$
- But the 2nd pole causes some bending below the asymptote
  - Actual ω_u slightly < ω_{u1}: UGF < GBW



[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 <u>without peaking</u> occurs at $\zeta = Q = 0.707$
  - Maximally flat response
  - $\omega_{p2} = 2\omega_{u1}$ and $PM \approx 65^o$

ω_{p2}/ω_{u1}	Q	ω_u/ω_{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
∞	—	1	90

[Jespers & Murmann, 2017]

Optimum Phase Margin (Freq Response)

- Maximum CL BW <u>without peaking</u> 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 $PM \approx 76^o$

ω_{p2}/ω_{u1}	Q	ω_u/ω_{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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9	0.333	0.994	83.7
10	0.316	0.995	84.3
∞	—	1	90

[Jespers & Murmann, 2017]

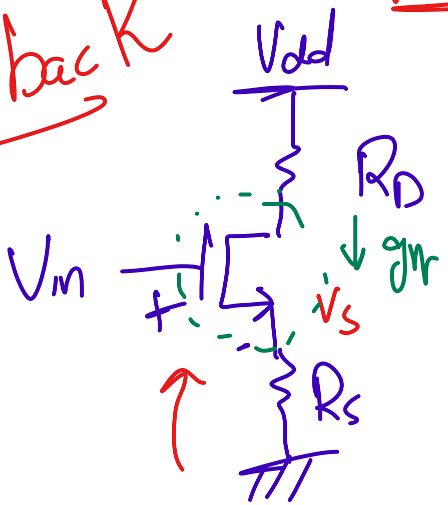
Second-Order System Speed-up

- First order system corresponds to $\omega_{p2}/\omega_{u1} \rightarrow \infty$
  - 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 (ε_d)	t_s/τ ($\omega_{p2}/\omega_{u1} \rightarrow \infty$)	t_s/τ ($\omega_{p2}/\omega_{u1} = 4$)	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

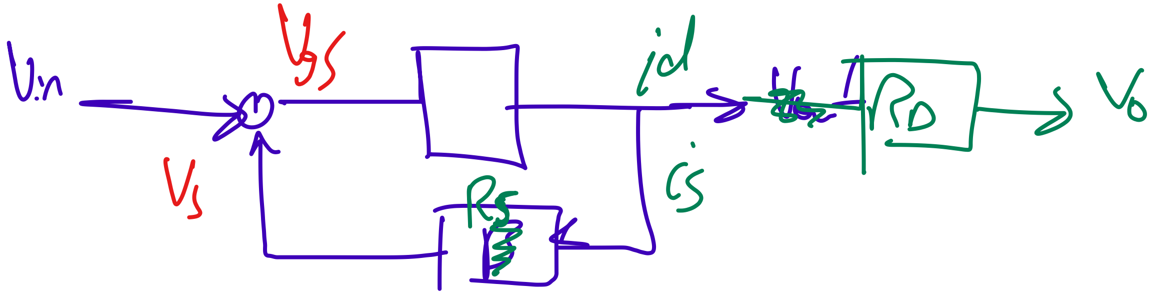
[Jespers & Murmann, 2017]

Local feedback — output is current



$\downarrow g_m V_{gs}$ sensing here

$V_s =$



"Series sensing"

Sense current

Thank you!

mixing happens here

"Series" Voltage mixing

"Series - series" feedback

$$A_{CL} = \frac{A_{OL}}{1+\beta A_{OL}} = \frac{g_m}{1+g_m R_s} = \frac{i_{out}}{V_{in}} = G_m$$

"Voltage - Current"

returning sensing

A → transconductance

β → conductance

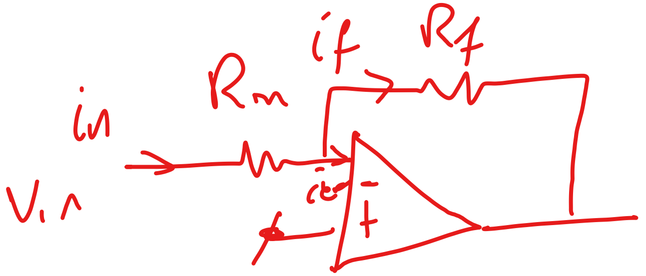
unitless.

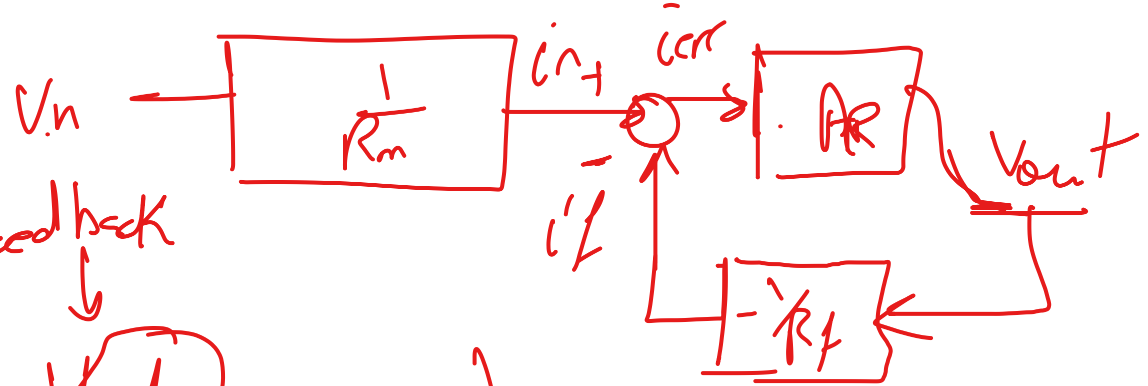
$A_{OL}\beta$ is always (unitless)

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.

Lec 2
Shunt-shunt
Feedback



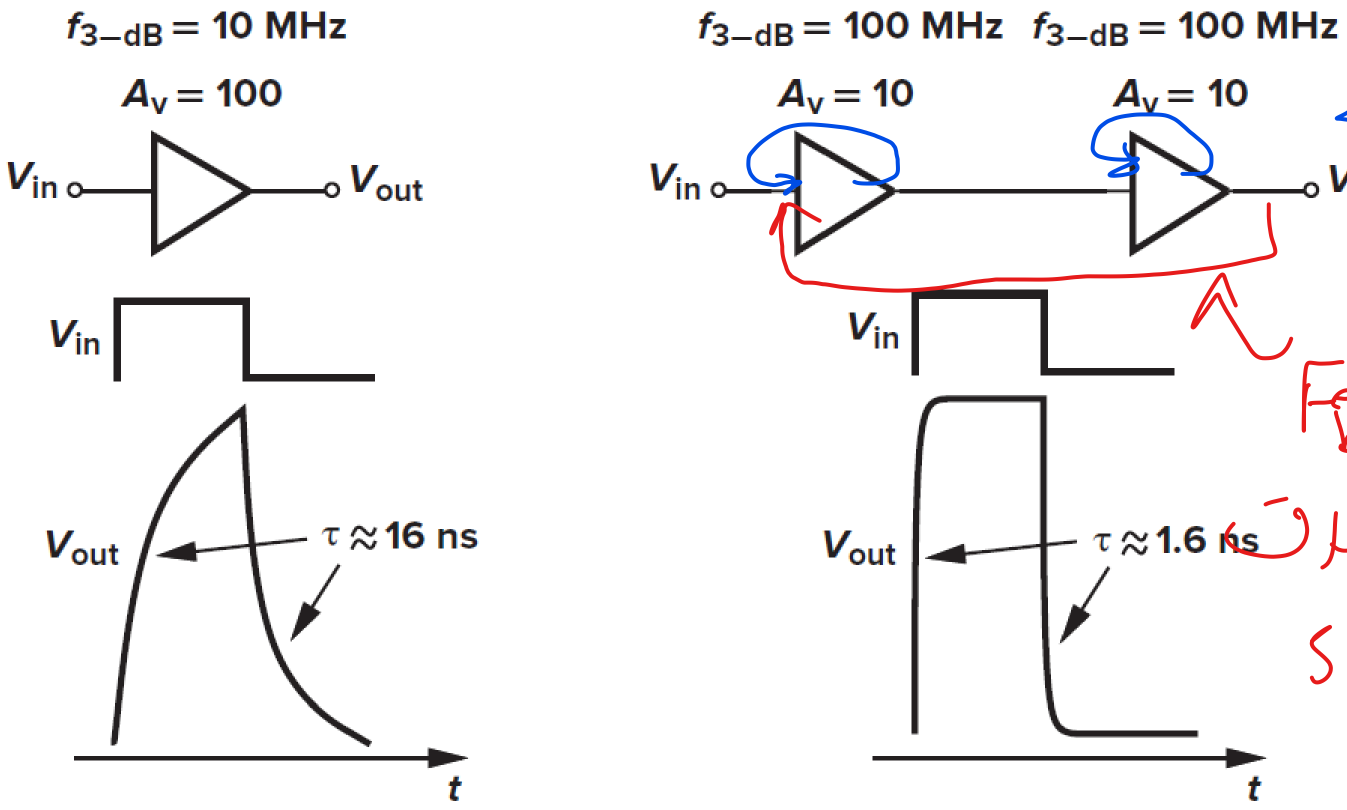


$i_{error} = i_n - i_f$, find $\frac{V_{out}}{V_{in}} = \frac{V_{out}}{i_n R_m} \approx \frac{\frac{1}{\beta}}{R_m} = \frac{-R_f}{R_{out}}$ #

Bandwidth Extension

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

and noise increases (might be a good option)



كل واحدة

ال Gain

كل نفسها

لو ال Feedback

كدة هتعمل فرق

في ال stability