ECE 264A

Jan 08, 2008 E.G.

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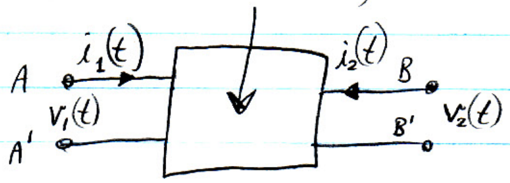
Poles Zeros & Freq. Response

SSM (small signal model) of circ. containing trans. and RLC networks



(1)

A, A' ≡ any two nodes of the circuit

B, B' ≡ " ——— " ——— "

Def. Let $x(t)$ be any real signal

then $\underbrace{\mathcal{L}\{x(t)\}}_{\text{lowercase} \to X(s)} = \int_{-\infty}^{\infty} x(t) e^{-st} dt \equiv$ Bilateral Laplace transform

Let $G(s) = \dfrac{X(s)}{Y(s)}$ where $x(t) = v_1(t), i_1(t), v_2(t), i_2(t)$

$y(t) =$ " ——— "

e.g. $G(s) = \dfrac{V_2(s)}{V_1 s}$ = Laplace transf. of gain from A, A' to B, B'

or $G(s) = \dfrac{V_1(s)}{I_1(s)}$ = " ——— " impedance at A, A' terminals.

Fact $G(s)$ is always rational with real coefficients.

i.e. $$G(s) = \frac{a_n s^n + a_{n-1} s^{n-1} + \dots a_1 s + a_0}{b_n s^n + b_{n-1} s^{n-1} + \dots + b_1 s + b_0} \quad (2)$$

rational = ratio of two polynomial functions

where $a_k, b_k \in \mathbb{R}$

Q Do ∃ practical circ. for which $G(s) \neq$ rational

A Yes! Ex: Delay element: $G(s) = e^{-sT} \equiv T$ second delay

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Can write ② as

$$G(s) = G' \frac{(s-z_1)(s-z_2)\cdots(s-z_n)}{(s-p_1)(s-p_2)\cdots(s-p_n)} \quad ③ \quad \text{where } G', z_k, p_k \in \mathbb{C}$$

z_k = zeros of $G(s)$
p_k = poles of $G(s)$

Fact $G(s)$, $s \in \mathbb{C}$
$G(j\omega)$, $\omega \in \mathbb{R}$ (if it exists) } Contain equiv. info. about ①
$g(t) = \mathcal{L}^{-1}\{G(s)\}$

"impulse response"
freq. response → Also: Magnitude response & phase response
transfer fcn.

Goal: Find simple way to deduce "shape" of $G(j\omega)$ from poles & zeros of $G(s)$

Claim 1 real coeff $\Leftrightarrow$ If $s_0 = \alpha + j\beta$ $(\alpha, \beta \in \mathbb{R})$ is a pole (or zero) of $G(s)$ Then $s_0^* = \alpha - j\beta$ is also a pole (or zero) of $G(s)$

Proof: exercise

Ex $(s-s_0)(s-s_0^*) = s^2 - s\underbrace{(s_0+s_0^*)}_{2\cdot Re\{s_0\}} + \underbrace{s_0 s_0^*}_{|s_0|^2} = s^2 + s(2\zeta\cdot\omega_0) + \omega_0^2$

where $\omega_0 \equiv |s_0| \equiv$ resonant requency
$\zeta = -Re\{s_0\}/\omega_0 \equiv$ damping ratio (Claim 1 $\Rightarrow \zeta, \omega_0 \in \mathbb{R}$)

$\therefore$ Can group conjugate poles & zeros to get

$$G(s) = G' \left[\prod_{\text{real zeros}} (s - z_k)\right] \cdot \left[\prod_{\text{complex zero pairs}} \left(s^2 + s(2\zeta_i\omega_{0i}) + \omega_{0i}^2\right)\right] \cdot \left[\prod_{\text{real poles}} \left(\frac{1}{s-p_k}\right)\right] \cdot \left[\prod_{\text{complex pole pairs}} \left(\frac{1}{s^2 + s(2\cdot\zeta_i\cdot\omega_{0i}) + \omega_0^2}\right)\right] \quad ④$$

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Let ω_{zi}, $i = 1, 2, \dots N \equiv$ the set of (non-zero) real zeros
ω_{pi} $i = 1, 2, \dots N \equiv$ " ——— " poles

$$\therefore\ G(s) = G(0)\, s^{n_0} \left[\prod_{\substack{\text{real}\\\text{zeros}}} \left(1 - s/\omega_{zk}\right)\right]\left[\prod_{\substack{\text{complex}\\\text{zero pairs}}}\left(\frac{s^2}{\omega_{0zi}^2} + \frac{2\zeta_{zi}}{\omega_{0zi}} s + 1\right)\right]$$

$$\cdot \left[\prod_{\substack{\text{real}\\\text{pole}}}\left(\frac{1}{1 - s/\omega_{pk}}\right)\right]\left[\prod_{\substack{\text{complex}\\\text{pole pair}}}\left(\frac{1}{\frac{s^2}{\omega_{opi}^2} + \frac{2\zeta_{pi}}{\omega_{opi}} s + 1}\right)\right]$$

Where $G(0) = \frac{a_0}{b_0}$, n_0 = # of DC-zeros ÷ # of DC-poles

Usually interested in $\underbrace{10 \cdot \lg\left(|G(j\omega)|^2\right)}_{\text{gain in dB}}$ — dB : viser større range, pot en måningsfull måte

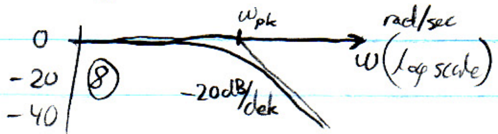
and $\angle G(j\omega) = \tan^{-1}\left(\frac{Im\{G(j\omega)\}}{Re\{G(j\omega)\}}\right)$

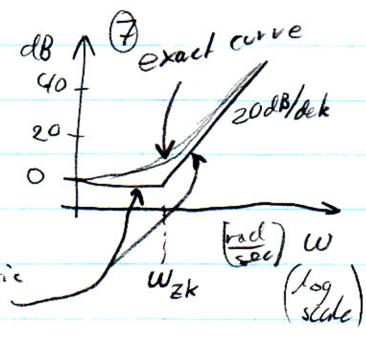
<u>Gain</u> (assume $n_0 = 0$ for now)

$$10 \cdot \lg |G(j\omega)|^2 = 10 \cdot \lg |G(0)|^2 \qquad (6)$$
$$+ \sum_{\substack{\text{real}\\\text{zeros}}} 10 \lg\left(1 + \omega^2/\omega_{zk}^2\right) \qquad (7)$$
$$- \sum_{\substack{\text{real}\\\text{poles}}} 10 \cdot \lg\left(1 + \omega^2/\omega_{pk}^2\right) \qquad (8)$$
$$+ \sum_{\substack{\text{complex}\\\text{zero pairs}}} 10 \cdot \lg\left(\left(1 - \frac{\omega^2}{\omega_{0zi}^2}\right)^2 + 4\frac{\zeta_{zi}^2}{\omega_{0zi}^2} \cdot \omega^2\right) \qquad (9)$$
$$- \sum_{\substack{\text{complex}\\\text{pole pairs}}} \text{(and so on..)} \qquad (10)$$

$\Rightarrow$ only 2 "types" of curves summed together

(7), (8) $\rightarrow 0$ as $\omega \rightarrow 0$
(7) $\rightarrow 20 \cdot \lg(\omega) + \text{const}$ as $\omega \rightarrow \infty$
(8) $\rightarrow -20 \lg(\omega) - \text{const}$ as $\omega \rightarrow \infty$





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New room: Tuesday 1/15 (website)
Homework on Tuesday. Gjør tutorial for Cadence

cont.:
recall (10) $\ldots - \sum_{\substack{\text{complex}\\\text{pole}\\\text{pairs}}} 10 \cdot lg\left[\left(1-\frac{\omega^2}{\omega_{opi}^2}\right)^2 + 4\frac{\zeta_{pi}^2}{\omega_{opi}^2}\omega^2\right]$

<u>Claim 2</u> The max departure of the exact curve for real pole and zero factors from the asymptotic approx is a factor of 0.707 in both (7) and (8).

<u>Proof</u> exercise

$\boxed{Q = \frac{1}{2\cdot\zeta}}$

(7), (10) $\rightarrow$ 0 dB as $\omega \rightarrow 0$

(9) $\rightarrow 40 \cdot lg\, \omega$ − const as $\omega \rightarrow \infty$
(10) $\rightarrow -40 \cdot lg\, \omega$ − const as $\omega \rightarrow \infty$

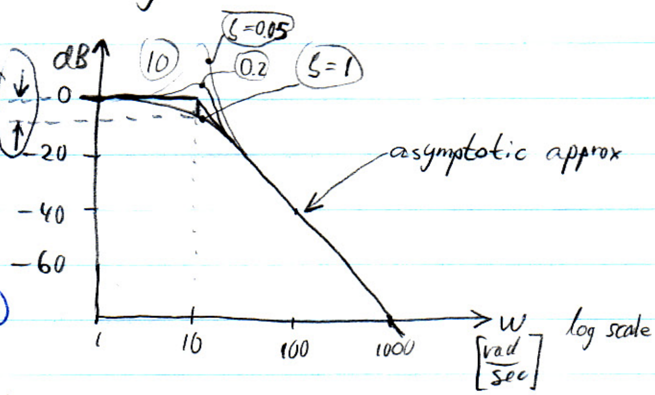
Peaking for $0 \leq \zeta < \frac{1}{\sqrt{2}}$ — $Q > \frac{1}{\sqrt{2}}$

Peak @ DC for $\zeta \geq \frac{1}{\sqrt{2}}$

ζ	Mag. at $\omega = \omega_{opi}$	
0	∞	∞
1/4	+6 dB	2
1/2	0 dB	1
$1/\sqrt{2}$	−3 dB	$1/\sqrt{2}$ (Butterworth)
1	−6 dB	$\frac{1}{2}$ (critically damped)
		Q =

Voltage amplitude at $\omega = \omega_{opi}$

e.g. (picture for $\omega_{opi} = 10 \frac{rad}{sec}$)

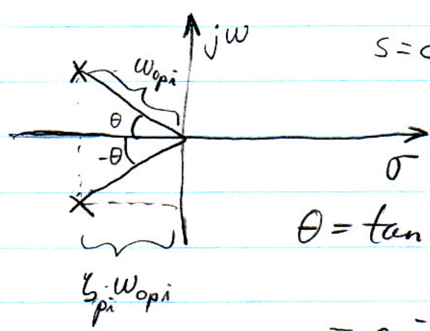


∴ asymptotic approx ≠ good approx near $\omega = \omega_{opi}$ for $\zeta \lessapprox 0.2$

Voltage amplitude at the peak freq.: $\frac{1}{2\zeta\cdot\sqrt{1-\zeta^2}}$ or $\frac{Q^2}{\sqrt{Q^2 - 1/4}}$ or $\frac{1}{\sin(2\cdot\theta)}$

poles accociated with (10) (verify)

$s = \sigma + j\omega$



$\theta = \tan\sqrt{\frac{1}{\zeta_{pi}^2} - 1}$

$= \cos^{-1}(\zeta_{pi}) \quad (\zeta_{pi} < 1)$

$\boxed{\begin{aligned}\omega_r &= \omega_0 \cdot \sqrt{1 - 2\cdot\zeta^2}\\ \omega_d &= \omega_0 \cdot \sqrt{1-\zeta^2}\end{aligned}}$

RLC:	Parallel	Series	
damping:	$\frac{L}{4RC}$	R	Ω
critical damping:	$\frac{1}{2}\sqrt{\frac{L}{C}}$	$2\cdot\sqrt{\frac{L}{C}}$	Ω
damping ratio ζ:	$\frac{1}{2R}\sqrt{\frac{L}{C}}$	$\frac{R}{2}\sqrt{\frac{C}{L}}$	
Q-factor	$R\sqrt{\frac{C}{L}}$	$\frac{1}{R}\sqrt{\frac{L}{C}}$	

② (Start of "New" lecture) ECE 264A Jan 10 E.G.

$$G(s) = \frac{a_m s^m + a_{m-1} s^{m-1} + \cdots + a_1 s + a_0}{b_n s^n + \cdots + b_1 s + b_0} \quad (1)$$

where $a_k, b_k \in \mathbb{R}$ and $s \in \mathbb{C}$

Let $\{\omega_{zk} : k = 1, 2, \ldots, N\}$ = non-zero real zeros of (1) (if any)

$\{\omega_{pk} : k = 1, 2, \ldots, N'\}$ = — " — poles — " —

$\{s_{zk}, s_{zk}^* : k = 1, \ldots N''\}$ = non-real conj. zeros of (1)

$\{s_{pk}, s_{pk}^* : k = 1, \ldots N'''\}$ = — " — poles — " —

$\omega_{oxk} = |s_{xk}|$ $(x = z \text{ or } p)$ ($\equiv$ natural freq. when $x = p$)

$\zeta_{xk} = -\frac{Re\{s_{xk}\}}{\omega_{oxk}}$ ($\equiv$ damping ratio when $x = p$)

(2)

Using (2), (1) becomes

$$G(s) = G(0) \cdot s^{n_0} \left[\prod_{k=1}^{N} (1 - s/\omega_{zk})\right] \left[\prod_{k=1}^{N''} \frac{1}{\omega_{ozk}^2} (s^2 + 2\zeta_{zk}\omega_{ozk}s + \omega_{ozk}^2)\right]$$
$$\cdot \left[\prod_{k=1}^{N'} (1 - s/\omega_{pk})^{-1}\right] \left[\prod_{k=1}^{N'''} \omega_{opk}^2 (s^2 + 2\zeta_{pk}\omega_{opk}s + \omega_{opk}^2)^{-1}\right] \quad (3)$$

where $G(0) = \frac{a_0}{b_0}$, $n_0 = \left(\begin{matrix} \# \text{ zero-freq zeros} \\ - \# \text{ — " — poles} \end{matrix}\right)$

$$G(j\omega) = \underbrace{|G(j\omega)|}_{\text{magnitude}} e^{j \underbrace{\measuredangle G(j\omega)}_{\text{phase}}}$$

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Phase response

Exercise: verify $\measuredangle G(j\omega) = \sum \measuredangle (\text{each factor in ③})\Big|_{s=j\omega}$

where

$$(\text{factor in ③}) \longrightarrow \measuredangle(\text{factor in ③})\Big|_{s=j\omega}$$

$$G(0) = \frac{a_0}{b_0} \longrightarrow \begin{cases} 0 & \text{if } G(0) > 0 \\ \pi & \text{if } G(0) < 0 \end{cases}$$

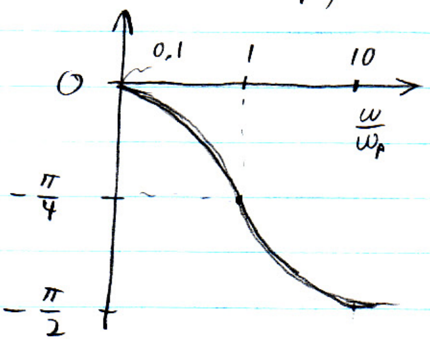
$$s^{n_0} \longrightarrow \frac{\pi}{2} \cdot n_0$$

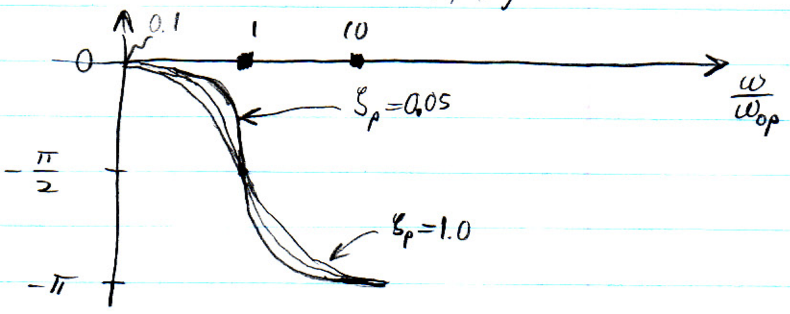
$$\left(1 - s/\omega_{xk}\right)^{\pm 1} \longrightarrow \mp \tan^{-1}\left(\frac{\omega}{\omega_{xk}}\right) \quad (x = p \text{ or } z)$$

$$\left[\frac{1}{\omega_{oxk}^2}\left(s^2 + 2\zeta_{xk}\omega_{oxk}s + \omega_{oxk}^2\right)\right]^{\pm 1} \longrightarrow \pm \tan^{-1}\left(\frac{2\zeta_{xk}\omega_{oxk}\omega}{\omega_{oxk}^2 - \omega^2}\right) \quad x = p \text{ or } z$$

Picture (for **LHP** poles; only the sign changes for LHP zeros. same for RHP zeros)

$\measuredangle\left(\frac{1}{1 - j\omega/\omega_p}\right)$, $\omega_p < 0$ $\qquad$ $\measuredangle\left(\frac{\omega_{op}^2}{s^2 + 2\omega_{op}\zeta_p s + \omega_{op}^2}\right)\Big|_{s=j\omega}$ $\quad$ $Re\{s_p\} < 0$





Poles & zeros - time response

Let $p(s) = a_m s^m + a_{m-1} s^{m-1} + \dots a_0$, $q(s) = b_n s^n + b_{n-1} s^{n-1} + \dots b_0$ $\Rightarrow G(s) = \dfrac{p(s)}{q(s)} = \dfrac{p(s)}{b_1 (s-p_1)(s-p_2)\cdots(s-p_n)}$

Recall "Partial Fraction Expansion" (PFE)

Ex 1 $m < n$ $\underbrace{p_i \neq p_j \text{ for } i \neq j}$

$\Leftrightarrow$ roots of $q(s)$ are all "first order"

Then $G(s) = \sum_{k=1}^{n} \frac{A}{s-p_k}$, $A_k \in \mathbb{R}$ $\underbrace{\text{where } A_k = \lim_{s\to p_k}\left[(s-p_k)G(s)\right]}_{4}$

Ex 2 $m < n$ $p_1, p_2, p_3, \dots p_{n'}$ = 1st order roots of $q(s)$ where $n' \leq n-2$

$4 \leq 7-2$

and $\underbrace{p_{n'+1} = p_{n'+2} = \dots = p_n}_{r \text{ "multiple roots" of } q(s) \text{ where } r = n - n'}$ (7; $3 = 7-4$)

} must 2 like rotter (example no.s)

Then $G(s) = \sum_{k=1}^{n'} \left(\frac{A_k}{s-p_k}\right) + \sum_{k=1}^{r} \frac{B_k}{(s-p_n)^k}$ ($4 \to n'$, $3 \to r$)

where $A_k = 4$

$$B_k = \lim_{s\to p_n} \frac{1}{(r-k)!} \frac{d^{r-k}}{ds^{r-k}}\left[(s-p_n)^r G(s)\right] \quad (5)$$

Similar results hold for any comb. of 1st order and multiple roots of $q(s)$.

Can extend for $m \geq n$

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Cont. from last time (Mid-term calc.: **Maybe**)

let $g(t) = \mathcal{L}^{-1}\{G(s)\}$

PFE $\Rightarrow$ $g(t)$ = weighted sum of terms of types

$$\begin{cases} \mathcal{L}^{-1}\left\{\frac{1}{s-p_k}\right\} = e^{p_k t} \cdot u(t) & (6) \\ \mathcal{L}^{-1}\left\{\frac{1}{(s-p_k)^r}\right\} = \frac{t^{r-1}}{(r-1)!} e^{p_k t} \cdot u(t) & (7) \end{cases}$$

(Note: $(7) = \underbrace{(6) * (6) * (6) * \dots * (6)}_{r \text{ times}}$)

In circuits, rarely have multiple roots of $g(s)$ (i.e., rarely have multiple equal poles)
$\Rightarrow$ only concerned with (6) usually

For real non-zero poles, $(6) \propto \mathcal{L}^{-1}\left\{\frac{1}{1 - s/\omega_{pk}}\right\}$

For non-real poles, we can group conj. pairs for which (6) results in terms of form

$$\mathcal{L}^{-1}\left\{\frac{\omega_{opk}^2}{s^2 + s(2\zeta_{pk}\omega_{opk}) + \omega_{opk}^2}\right\}$$

ω_{opk} is an absolute value (Jan 10, pg. 2)

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Step Response

$$\mathcal{L}\{u(t)\} = \frac{1}{s}$$

Response of "output" with the input $= u(t) \equiv$ "step response"

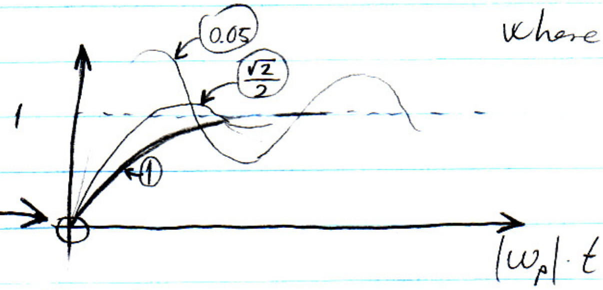
$$\therefore\ g_{step} = \mathcal{L}^{-1}\left\{\frac{1}{s}\cdot G(s)\right\}$$

$$\mathcal{L}^{-1}\left\{\frac{1}{s}\ \frac{1}{1-s/\omega_p}\right\} = u(t)\cdot\left(1-e^{\omega_p t}\right)$$

$$\mathcal{L}^{-1}\left\{\frac{1}{s}\cdot\frac{\omega_{op}^2}{s^2+s(2\zeta_p\omega_{op})+\omega_{op}^2}\right\} = u(t)\cdot\left[1-\frac{1}{\sqrt{1-\zeta_p^2}}e^{-\zeta_p\omega_{op}t}\cdot\sin\left(\omega_{op}\sqrt{1-\zeta_p^2}\cdot t+\theta\right)\right]$$

where $\theta = \tan^{-1}\left(\frac{\sqrt{1-\zeta^2}}{\zeta_p}\right)$

Den deriverte i starten av step-responsen er 0 dersom vi ikke har zeros i t.f.



For $\zeta_p = 0$, poles on img. axis $\Rightarrow$ oscillation.

For $\zeta_p < 0$ poles in RHP $\Rightarrow$ oscillation with exp. increasing amplitude

For $0 < \zeta_p < 1$ "overshoots" but settles to 1.

$$= u(t)\cdot\left[1-\frac{1}{\sqrt{1-\zeta_p^2}}\cdot e^{-\zeta_p\cdot\omega_{op}\cdot t}\cdot\cos\left[\omega_{op}\cdot\sqrt{1-\zeta^2}\cdot t-\tan^{-1}\left(\frac{\zeta_p}{\sqrt{1-\zeta_p^2}}\right)\right]\right]$$

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Dominant pole approx.

Provided have no DC poles or zeros

$$G(s) = G(0) \cdot \frac{(1 - s/z_1)(1 - s/z_2) \dots (1 - s/z_m)}{(1 - s/p_1)(1 - s/p_2) \dots (1 - s/p_n)}$$

Suppose $|p_1| << |p_k|, \quad 2 \le k \le n$

$|p_1| << |z_k|, \quad 1 \le k \le m \qquad (\Rightarrow p_1 \in \mathbb{R} \text{ why?})$

Then $|G(j\omega)| \cong \frac{G(0)}{\sqrt{1 + \frac{\omega^2}{|p_1|^2}}} \qquad \therefore |G(j\omega_{-3dB})| = \frac{G(0)}{\sqrt{2}}$

$\Rightarrow \omega_{-3dB} = |p_1| \qquad$ (3dB Bandwidth)

(New numbering)

Comon source amplifiers freq. response

Aside Course convention

$I_d = I_D + i_d$

$V_{gs} = V_{GS} + v_{gs}$, etc...

v_{gs}: Variations about a constant "bias" value

V_{GS}: a constant "bias" level

V_{gs}: "total" or "absolute" value

$\underbrace{\qquad\qquad}$

Læreboken bruker i_D og v_{GS} for dette.

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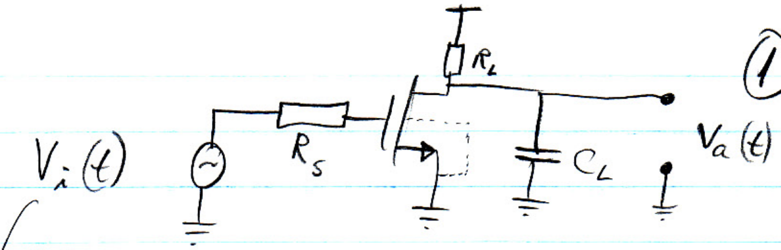
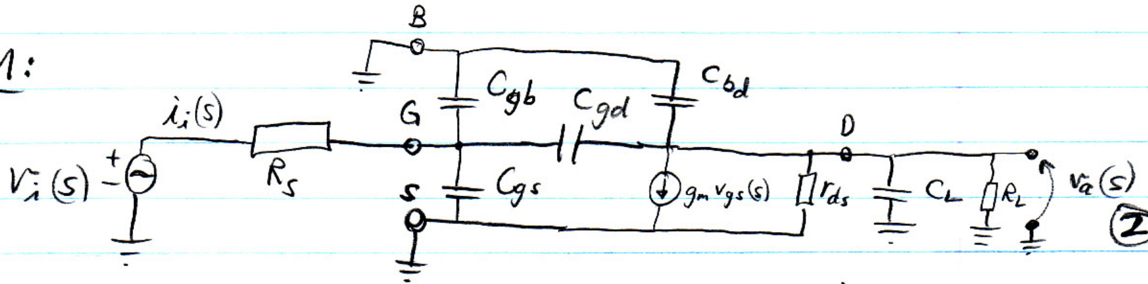


① (R_L & R_s often from active circuits, not actual resistors)

$= V_I + v_i(t)$

SSM:



②

Let $C_{gs}' = C_{gs} + C_{gb}$ ($\approx C_{gs}$ usually)

$$C_d = C_{bd} + C_L$$

$$R_L' = R_L \parallel r_{ds}$$

eg. for $I_D = 200\mu A$, $\frac{W}{L} = 50$, $C_{gs}' = 90fF$, $C_{gd} = 20fF$, $C_d = 80fF$, $g_m = 1.7 \cdot 10^{-3}\,\Omega^{-1}$, $r_{ds} = 500k\Omega$ ③

Note: ② also SSM of diff pair DM ½ circuit

KCL at G: $\frac{v_i(s) - v_{gs}(s)}{R_s} - v_{gs}(s)\cdot s \cdot C_{gs}' - [v_{gs}(s) - v_a(s)]\cdot s \cdot C_{gd} = 0$

$$\therefore v_{gs}(s) = \left[\frac{1}{1/R_s + s(C_{gs}' + C_{gd})}\right]\left(v_a(s)\cdot s \cdot C_{gd} + v_i(s)\cdot \frac{1}{R_s}\right) \quad ④$$

KCL at D: $[v_{gs} - v_a(s)]\cdot s \cdot C_{gd} - g_m v_{gs}(s) - v_a(s)\left[\frac{1}{R_L'} + sC_d\right] = 0$

$$\therefore v_a(s) = \left[\frac{1}{\frac{1}{R_L'} + s(C_d + C_{gd})}\right](sC_{gd} - g_m)\cdot v_{gs}(s) \quad ⑤$$

Let $G(s) = \frac{1}{R_s}\left[\frac{1}{1/R_s + s(C_{gs}' + C_{gd})}\right]$

$A(s) = \frac{sC_{gd} - g_m}{1/R_L' + s(C_d + C_{gd})}$, $F(s) = -s\cdot C_{gd}\left[\frac{1}{1/R_s + s(C_{gs}' + C_{gd})}\right]$ ⑥

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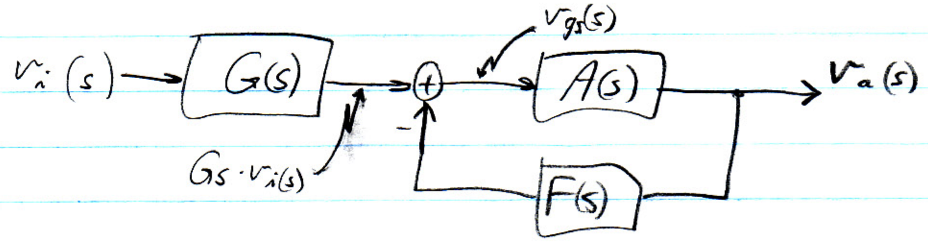
$$\left.\begin{aligned} &(4),(6) \Rightarrow v_{gs}(s) = -F(s)\,v_a(s) + G(s)\,v_i(s) \\ &(5),(6) \Rightarrow v_a(s) = A(s)\cdot v_{gs}(s) \end{aligned}\right\}\ (7)$$

(7) $\Rightarrow$ "Block Diagram" of (2):



(8)

$$\left.\begin{aligned} &(8) \text{ using Masons gain formula} \\ &\qquad\text{OR} \\ &(7) \text{ using algebra} \end{aligned}\right\} \Rightarrow A_v(s) = G(s)\frac{A(s)}{1+A(s)F(s)} \quad (9)$$

$$\equiv \left(\frac{v_a(s)}{v_i(s)}\right)$$

Block Diagrams

SSM (eg. (2)) $\Longleftrightarrow$ system of equations in s (eg. (4) & (5)) $\Leftrightarrow$ block diagram (e.g. (8))

- SSM is symbolic $\Rightarrow$ good for insight
- but components are bidirectional $\Rightarrow$ bad for insight
- can't go directly from SSM to gain & impedances (need syst. of eqs. first)
- Block diagrams provide insight because components are uni-directional $\Rightarrow$ Feedback paths are obvious
- Can go directly from block diagrams to gain & impedances (using Mason's gain formula)

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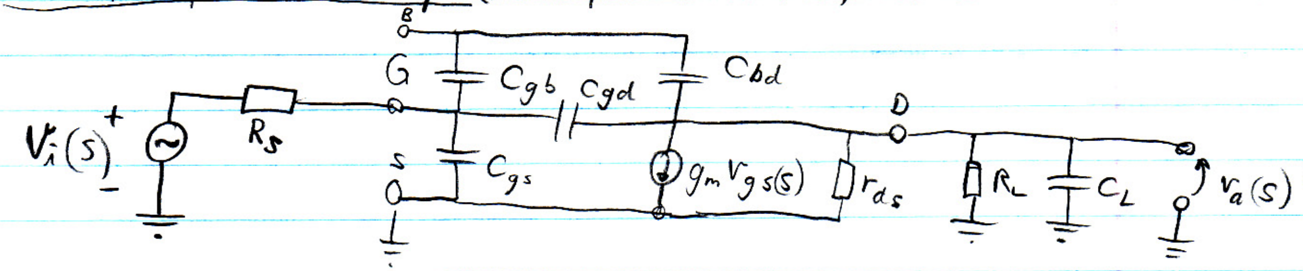
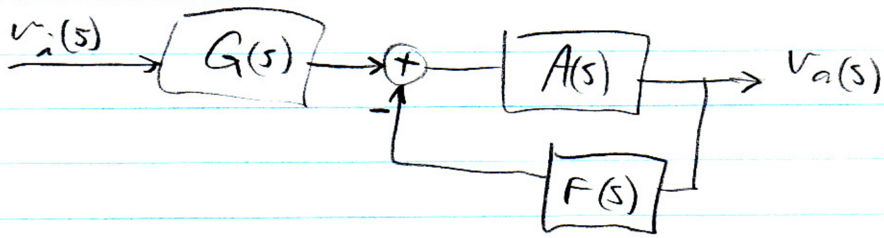
SSM of CS Amp. (recall from last time)



(2)

$$\left.\begin{aligned} v_{gs}(s) &= -F(s)\cdot v_a(s) + G(s)\, v_i(s) \\ v_a(s) &= A(s)\cdot v_{gs}(s) \end{aligned}\right\} \quad (7)$$



(8)

$$A_r(s) = G(s)\cdot\frac{A(s)}{1+A(s)F(s)} \quad (9)$$

Recall Stability $\Leftrightarrow$ no poles in RHP (incl. imag axis?)$_{E.G.}$

(9) $\Rightarrow$ — " — $\Leftrightarrow$ $A(s_0)F(s_0) \neq -1$ for any s_0 with $Re\{s_0\} \geq 0$
(assumes $F(s) \neq 0$ and no pole of $A(s)$ is a zero of $F(s)$)

First consider $A_{r_0} = A_r(j\omega)\big|_{\omega=0}$ "DC-gain"

$$A_{r_0} = G(j\omega)\left.\frac{A(j\omega)}{1+A(j\omega)F(j\omega)}\right|_{\omega=0}$$

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$F(j\omega)\Big|_{\omega=0} = 0$ (makes sence because $F(s)$ represents feedback through C_{gd})

$\therefore\ A_{v_0} = -g_m R_L'$ (10)

Now consider poles & zeros: Let $H(j\omega) = \frac{1}{A_{v_0}} \cdot A_v(j\omega)$

$\therefore A_v(j\omega) = A_{v_0} \cdot H(j\omega)$

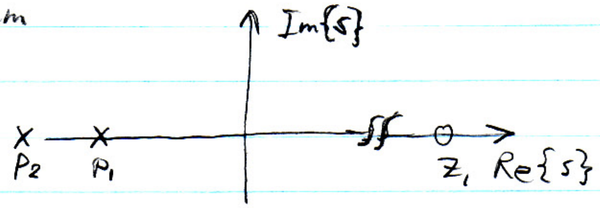
(6), (9) $\Rightarrow H(s) = \dfrac{1 - s/z_1}{1 + a_1 s + a_2 s^2}$

where $\begin{cases} z_1 = g_m / C_{gd} \\ a_1 = R_s\left[C_{gs}' + C_{gd}\left(1 + g_m R_L'\right)\right] + R_L'\left(C_{gd} + C_d\right) \\ a_2 = R_s R_L'\left[C_d C_{gs}' + C_d C_{gd} + C_{gs} C_{gd}\right] \end{cases}$

$\therefore A_v(s) = A_{v_0}\left[\dfrac{1 - s/z_1}{(1 - s/p_1)(1 - s/p_2)}\right]$ where $\left.\begin{aligned} p_1 &= -\frac{1}{2a_2}\left(a_1 + \sqrt{a_1^2 - 4a_2}\right) \\ p_2 &= -\frac{1}{2a_2}\left(a_1 - \sqrt{\ \text{''}\ }\right) \end{aligned}\right\}$ (11)

Usually, device parameters $\Rightarrow a_1^2 > 4a_2$ for CS amp.
$\Rightarrow p_1, p_2$ are real

pole-zero diagram



Hand-analysis: 10~20% accuracy

Ex using (3) with $R_L = 10k\Omega$, $R_s = 5k\Omega$ gives
$A_{v_0} = -15.5$, $f_{z_1} = \frac{z_1}{2\pi} = 13.5\,GHz$, $f_{p_1} = \left|\frac{p_1}{2\pi}\right| = 56\,MHz$

$f_{p_2} = \left|\frac{p_2}{2\pi}\right| = 860\,MHz$

$\left.\begin{aligned} |p_1| &\ll |p_2| \\ |p_1| &\ll |z_1| \end{aligned}\right\} \Rightarrow p_1 = \text{dominant pole}$

$\Rightarrow |A_v(j\omega)| \approx \left|A_{v_0}\left(\dfrac{1}{1 + j\omega/(2\pi \cdot 56MHz)}\right)\right|$ for $\omega \lesssim f_{p_2} \cdot 2\pi$

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$\Rightarrow$ 3dB BW = 56 MHz

Now consider ① with capacitor C_c connected between D and G.

$\Rightarrow C_c$ in parallel with C_{gd} in ②

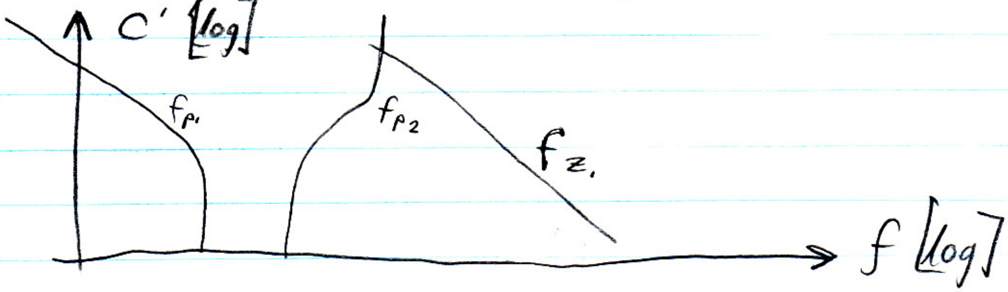
$\Rightarrow$ All eq. so far hold if C_{gd} is replaced by $C' = C_{gd} + C_c$ } (12)

Can show using (11) & (12) (exercise)

$$p_1 \approx -\frac{1}{(C_d + C')R_L' + (C_{gs}' C')R_S + g_m \cdot R_S \cdot R_L' C'} \qquad (13)$$

$$\approx \frac{-1}{g_m R_s R_L' C'} \quad \text{(for large } R_s \text{ \& } R_L\text{)}$$

$$p_2 \approx \frac{-g_m \cdot C'}{C_d C_{gs}' + C'(C_d + C_{gs}')} \qquad Z_1 = \frac{g_m}{C'}$$



$\therefore$ Increasing C' splits poles but moves zero closer to origin

Reset Numbering system!

(4/5)

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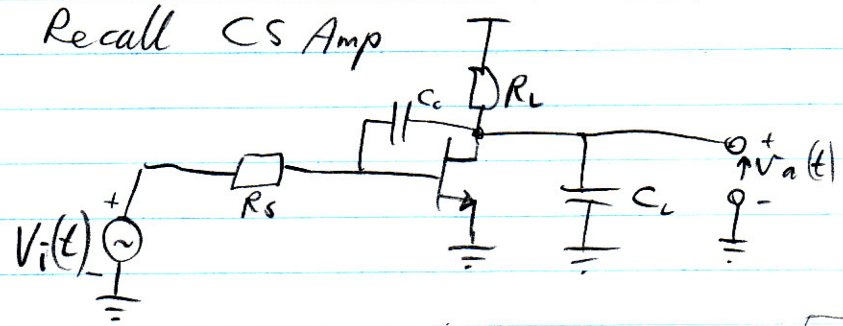
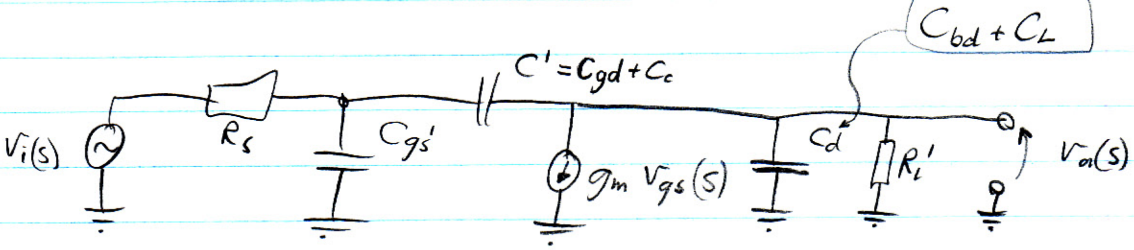
Zero-Value Time Constant Analysis

Ex Recall CS Amp





(2)

Last time: hard work gave us $p_1 \cong \dfrac{-1}{(C_d+C')R_L' + (C_{gs}'+C')R_s + g_m R_L' R_s C'}$ (3)

p_1 = dom. pole so 3dB BW = $\left|\dfrac{p_1}{2\pi}\right|$

Q If only interested in dom. pole ($\Rightarrow$ 3dB BW) is there an easier way?

A Yes – ZVTC Analysis

ZVTC Theorem

β_1 = negative of Σ of all poles

Consider circuit containing only sources, resistors and capacitors and poles $p_1, p_2, \ldots p_n$ where $p_k \neq 0$

Let $\beta_1 = -\sum_{k=1}^{n} {}^{+1}/_{p_k}$

$-\sum(\text{invers av pol}) = \sum \text{tidskonst.}$

Then $\beta_1 = \sum_{k=1}^{k} R_k C_k$ (4)

where C_k = value of k^{th} cap in circuit
and R_k = resistance between nodes of circuit to which C_k is connected (ie. impedance with all caps removed

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Ex (cont...)

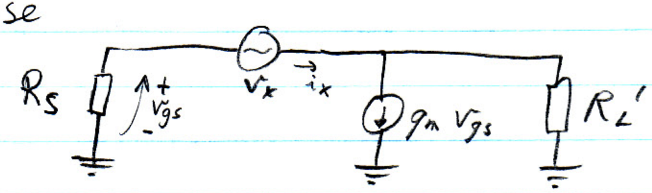
$$|p_1| << |p_2| \Rightarrow \beta_1 \approx \left|1/p_1\right| \quad \therefore \quad p_1 \approx -\frac{1}{\beta_1}$$

Let $C_{gs}' \equiv C_1 \quad C_d \equiv C_2, \quad C' \equiv C_3$ (5)

Inspection of (2) $\Rightarrow$ $R_1 = R_s$

$R_2 = R_L'$

To find R_3 use



$$\left.\begin{aligned} &\Rightarrow v_{gs} = -i_x R_s \\ &v_x = (i_x - g_m \cdot v_{gs}) R_L' - v_{gs} \end{aligned}\right\} R_3 = R_L' + R_S + g_m R_L' R_s$$

$$(4) \Rightarrow p_1 \approx \frac{-1}{\underbrace{R_s C_{gs}'}_{\tau_1} + \underbrace{R_L' C_d}_{\tau_2} + \underbrace{(R_L' + R_s + g_m R_L' R_s) C'}_{\tau_3}} = (3) \quad \checkmark$$

Z.V.T.C. Analysis

Vi vet ikke sikkert vilket RC-ledd som er størst...
Men vi gjør et estimat:

$$\omega_{3dB} \approx |p_{dom.}| \approx \frac{1}{\sum_k C_k \cdot R_k}$$

$$\approx \sum \left|\frac{1}{p_k}\right|$$

"Zero Value" refererer muligens til at alle "de andre" kondensatorne er nullstilte. På norsk heter det "åpen krets tidskonstant metoden".

(Nedre grensefrekvens: "Kortslutnings-tidskonstant-metoden", basert på $\sum_k \frac{1}{C_k \cdot R_k}$)

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ZVTC Theorem (contd.) (same # system)

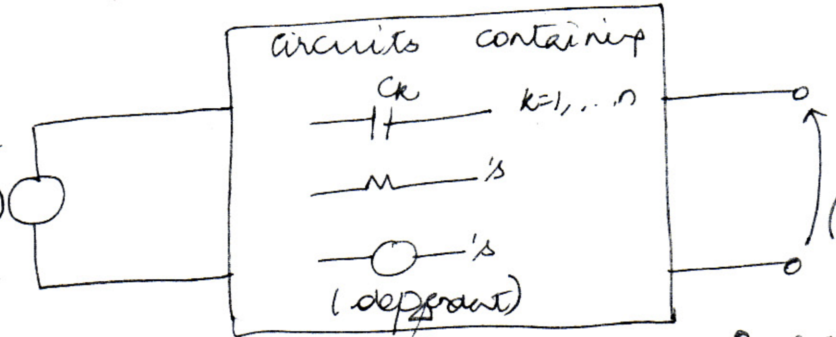
PROOF :-

Have :

Impulse :

$v_i(t) = \delta(t)$

$(v_i(s) = 1)$

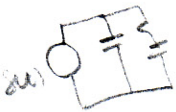


Impulse Response

$v_a(t) = g(t)$

$(v_a(s) = G(s))$

②



∵ For calculating Rs & Cs we need to turn off independent sources

where $G(s) = A_0 \dfrac{p(s)}{(1-s/p_1)(1-s/p_2)\cdots(1-s/p_n)}$ ← some polynomial

Note: In ②, $v_i(t) = \delta(t)$ sets initial condition of each capacitor $C_k, k=1,\ldots n$. After $t=0$, $v_i(t)=0$ (Atleast one cap will get charged)

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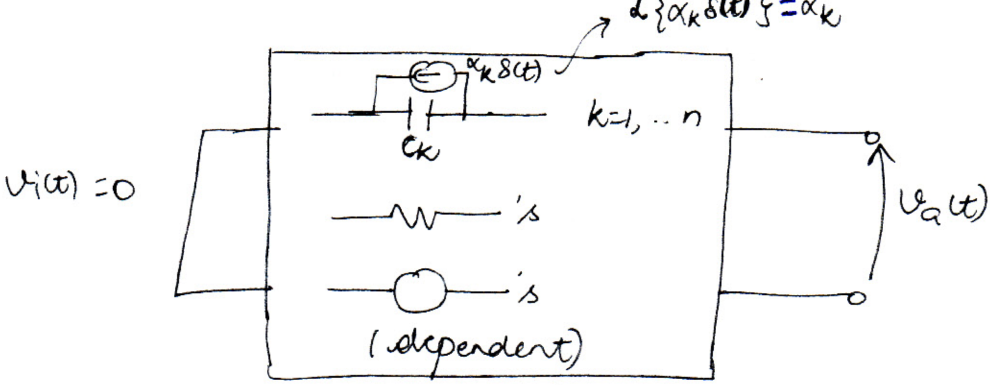
$\Rightarrow$ $g(t)$ = transient of (2) if $v_i(t) = 0$ but with appropriate initial conditions on $C_1, C_2, \ldots, C_n$.

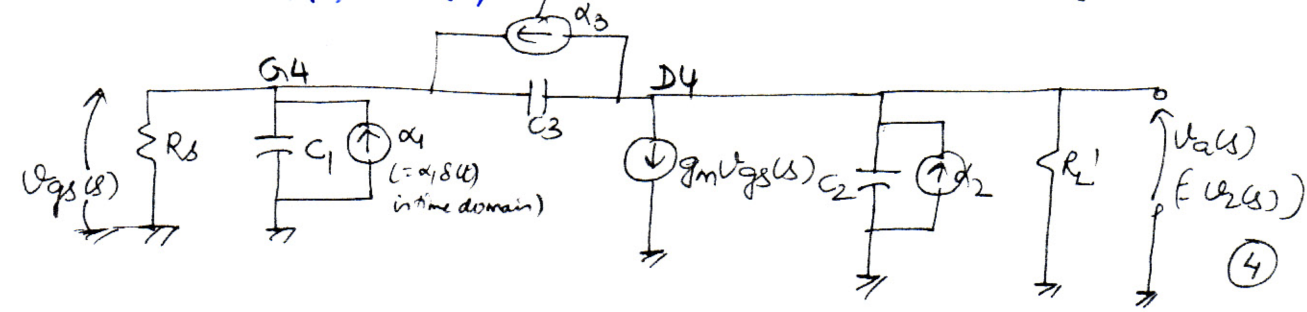
$\therefore$ (2) is equivalent to



(3)

e.g. S.S.M. of C.S. amplifier (last time)

if $v_i(t) = \delta(t)$ produces same $v_a(t)$ as

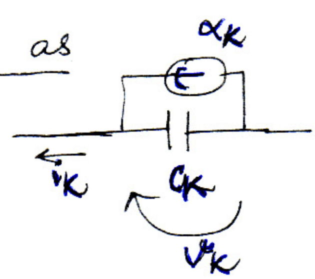


(4)

NOTE: To be specific will use (4) in place of (3) for today

But all results we find generalize to (3).

Define v_k and i_k as



$\therefore$ $i_k(s) = \alpha_k - v_k(s) \cdot sC_k$ — (5)

KCL @ any node $\Rightarrow$ equation of the form:

$$d_1 i_1(s) + d_2 i_2(s) + d_3 i_3(s) + g_1 v_1(s) + g_2 v_2(s) + g_3 v_3(s) = 0$$

where $d_k = 1, 0,$ or -1 (corresponding to current entering, ... leaving a node)

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and g_k = some conductance

e.g. KCL @ gate of (4):

$$-i_1(s) - i_3(s) + \frac{1}{R_S} v_i(s) = 0$$

Can solve all the equations of this form to get:

$$\left.\begin{aligned} i_1(s) &= a_{11} v_1(s) + a_{12} v_2(s) + a_{13} v_3(s) \\ i_2(s) &= a_{21} v_1(s) + a_{22} v_2(s) + a_{23} v_3(s) \\ i_3(s) &= a_{31} v_1(s) + a_{32} v_2(s) + a_{33} v_3(s) \end{aligned}\right\}$$

(6) where a_{jk} = some conductance

Note: To **force** (6) to be linearly independent, may need to add tiny resistors in series w/ some of the caps (add conceptually, not physically)

("tiny" $\Leftrightarrow$ small enough to have negligible effect on ckt.)

In matrix **notation**, (6) is $\underline{i}(s) = \underline{A} \cdot \underline{v}(s)$

$$\underline{i}(s) = \underbrace{\begin{bmatrix} i_1(s) \\ i_2(s) \\ i_3(s) \end{bmatrix}}_{3\times 1}, \quad \underline{A} = \underbrace{\begin{bmatrix} a_{11} & a_{12} & a_{13} \\ a_{21} & a_{22} & a_{23} \\ a_{31} & a_{32} & a_{33} \end{bmatrix}}_{3\times 3} \quad \& \quad \underline{v}(s) = \underbrace{\begin{bmatrix} v_1(s) \\ v_2(s) \\ v_3(s) \end{bmatrix}}_{3\times 1}$$

Ladning på kondensatorne i start øyeblikket?

(Real life: CS Amp: 3 caps, 2 poles)

Using (5),:

$$\underbrace{\begin{bmatrix} \alpha_1 \\ \alpha_2 \\ \alpha_3 \end{bmatrix}}_{\text{Call } \underline{\alpha}} = \underbrace{\begin{bmatrix} (a_{11}+sC_1) & a_{12} & a_{13} \\ a_{21} & (a_{22}+sC_2) & a_{23} \\ a_{31} & a_{32} & (a_{33}+sC_3) \end{bmatrix}}_{\text{Call } \underline{B}} \underline{v}(s) \quad (7)$$

Recall "Cramer's Rule"

$$\Rightarrow v_k(s) = \frac{\Delta_k(s)}{\Delta(s)}$$ where $\Delta(s) = \det(\underline{B})$ (= polynomial in s)

$\Delta_k(s) = \det(\underline{B}$ with kth column replaced by $\underline{\alpha})$

	unit
$\underline{i}(s)$	$A \cdot sec$
$\underline{v}(s)$	$V \cdot sec$
$\underline{A}, \underline{B}$	$\mho = \frac{A}{V} = \text{Siemens}$
α	$C = A \cdot sec$

$$5 \; \& \; 6 \Rightarrow \alpha_1 = i_1(s) + v_1(s) \cdot s \cdot C_1$$

$$= a_{11} \cdot v_1(s) + a_{12} \cdot v_2(s) + a_{13} \cdot v_3(s) + v_1(s) \cdot s \cdot C_1$$

$$= (a_{11} + sC_1) \cdot v_1(s) + a_{12} \cdot v_2(s) + a_{13} \cdot v_3(s)$$

$$\Rightarrow 7$$

Note: $v_a(s) = v_2(s)$ in ④, so $\underbrace{G(s) = \frac{\Delta_2(s)}{\Delta(s)}}_{⑧}$ Jan 22, 2008 9.

(i.e. we can calculate impulse response from cramer's rule)

We know denominator of $G(s)$ has form: $b_0 + b_1 s + b_2 s^2$

$= b_0 (1 + \beta_1 s + \beta_2 s^2)$

(Because of pole-zero cancellation, we have 2nd order polynomial in real life rather than 3rd order polynomial)

$= b_0 (1 - s/p_1)(1 - s/p_2)$

Algebra $\Rightarrow \underbrace{\beta_1 = -\sum_{k=1}^{n} 1/p_k, \quad n=2}_{\text{true for any } n \geq 1}$

⑦, ⑧, defn. of $\det(B) \Rightarrow$

$$\left.\begin{aligned} b_0 &= \Delta(s)\Big|_{c_1=c_2=c_3=0} \\ &= \Delta(0) \\ \& \quad b_0\beta_1 s &\equiv b_1 s = h_1 s c_1 + h_2 s c_2 + h_3 s c_3 \end{aligned}\right\} ⑨$$

where h_1, h_2, h_3 are real #s, i.e. $\in \mathbb{R}$

Let $\Delta_{ij} = (-1)^{i+j} \det(\underline{B}_{ij})$ where $\underline{B}_{ij} = \underline{B}$ with ith row & jth column deleted.

Can expand $\det(\underline{B})$ as

$$\Delta(s) = \underbrace{(a_{11} + sc_1)\Delta_{11} + a_{21}\Delta_{21} + a_{31}\Delta_{31}}_{\text{inspection of ⑦} \Rightarrow c_1 \text{ only occurs in 1st term}}$$

$$\therefore h_1 = \Delta_{11}\Big|_{c_2=c_3=0} = \Delta_{11}(0)$$

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Similarly, $h_2 = \Delta_{22}(0)$, $h_3 = \Delta_{33}(0)$

$\therefore$ (9) $\Rightarrow$ $$\beta_1 = \frac{\Delta_{11}(0)}{\Delta(0)} \cdot c_1 + \frac{\Delta_{22}(0)}{\Delta(0)} \cdot c_2 + \frac{\Delta_{33}(0)}{\Delta(0)} c_3$$

Now set $c_1 = c_2 = c_3 = 0$

& $i_2 = i_3 = 0$ to get

$$\begin{bmatrix} i_1 \\ 0 \\ 0 \end{bmatrix} = \underline{A} \begin{bmatrix} v_1 \\ v_2 \\ v_3 \end{bmatrix}$$

Cramer's Rule $\Rightarrow$ $$v_1 = \frac{i_1 \det \begin{pmatrix} a_{22} & a_{23} \\ a_{32} & a_{33} \end{pmatrix}}{\det(\underline{A})}$$

$$= i_1 \frac{\Delta_{11}(0)}{\Delta(0)}$$ by defn.

$$\therefore \left. \frac{v_1}{i_1} \right|_{\substack{c_1 = c_2 = c_3 = 0 \\ i_2 = i_3 = 0}} = R_1 = \frac{\Delta_{11}(0)}{\Delta(0)}$$

Similarly for R_2 and R_3

$$\therefore \beta_1 = R_1 c_1 + R_2 c_2 + R_3 c_3$$

1) One RC should dominate

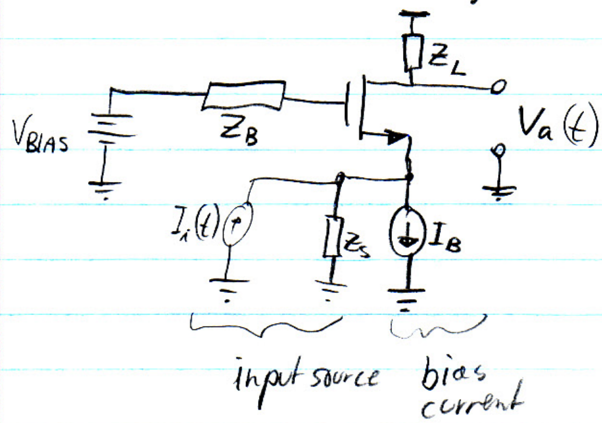
2) One pole should dominate

ECE 264A Jan 24, 2008

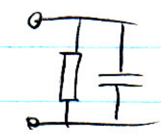
(o.h. tomorrow with TA 11:00 ~ 12^{30} in EBU1, 3329)

Freq resp of common gate (cascode) stage

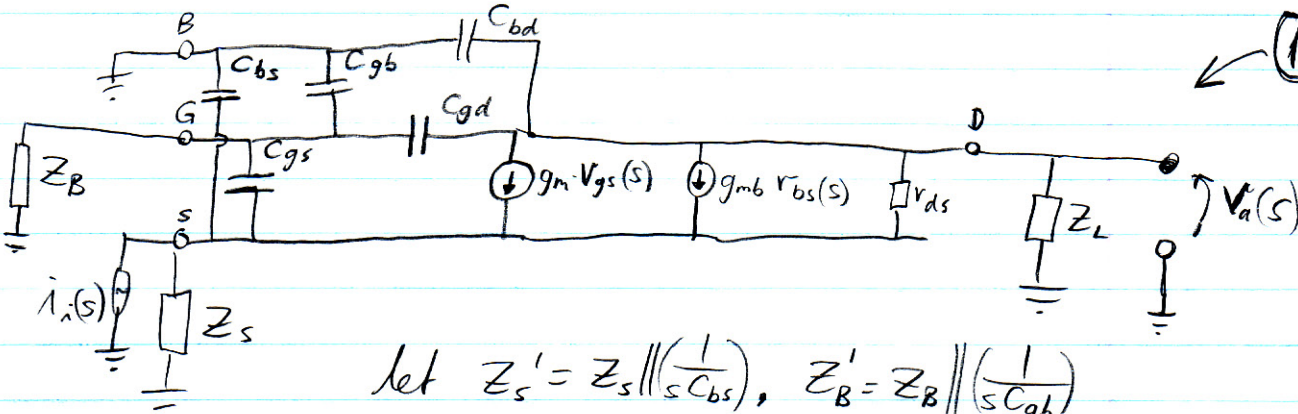
assume Body to ground (lowest potential)



assume

$Z_B, Z_S, Z_L \equiv$ 

SSM



(1)

let $Z_s' = Z_s \| \left(\frac{1}{sC_{bs}}\right)$, $Z_B' = Z_B \| \left(\frac{1}{sC_{gb}}\right)$

$$Z_L' = Z_L \| \frac{1}{sC_{bd}}$$

KCL: (at G)

$$\frac{V_g}{Z_B'} + (V_g - V_d)sC_{gd} + (V_g - V_s)sC_{gs} = 0$$

$$\therefore\ v_g = \underbrace{\left[\frac{1}{Z_B'} + s(C_{gd}\ C_{gs})\right]^{-1}}_{\text{call it } a(s)} \cdot \left[sC_{gs}v_s + sC_{gd}v_d\right] \quad (2)$$

KCL: (at s)

$$\therefore\ v_s = \left[\underbrace{\frac{1}{Z_s'} + g_m + g_{mb} + \frac{1}{r_{ds}} + sC_{gs}}_{\text{call it } b(s)}\right]^{-1}\left[i_i + (g_m + sC_{gs})v_g + \frac{1}{r_{ds}}v_d\right] \quad (3)$$

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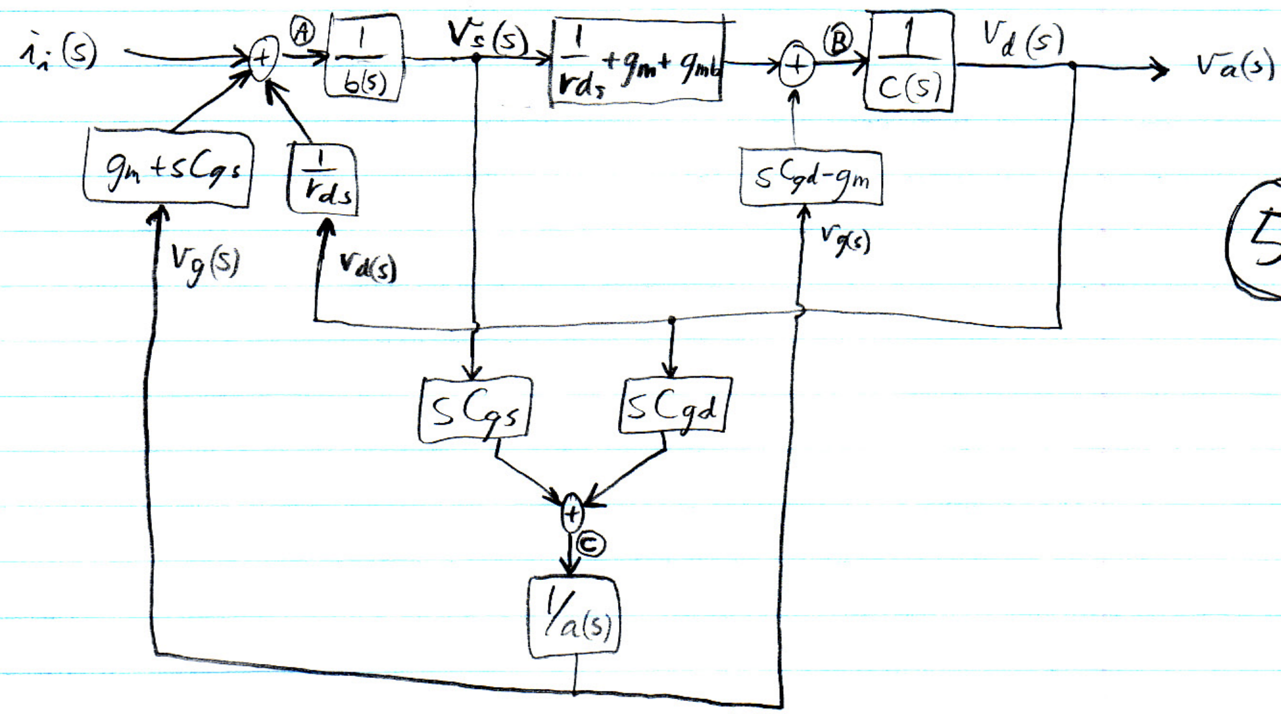
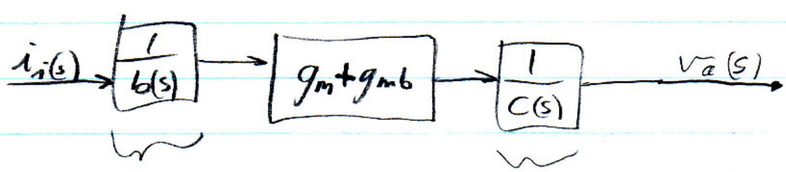
KCL (at D): ...
$$v_d = \underbrace{\left[\frac{1}{Z_L'} + \frac{1}{r_{ds}} + sC_{gd}\right]}_{\text{call it } c(s)}^{-1} \left[\left(\frac{1}{r_{ds}} + g_m + g_{mb}\right) v_s + \left(sC_{gd} - g_m\right) v_g\right] \quad (4)$$

can generate "block diagram" from (2), (3), (4):



(5)

Observations
1) all blocks are unidirectional (in dir. of the arrows)
2) all feedback paths arise from parasitic elements (i.e. if $C_{xy} = 0$, $r_{ds} = \infty$, then no feedback loops)
with $C_{xy} = 0$, $r_{ds} = \infty$, (5) reduces to



(6)

$$\frac{1}{b(s)} = \frac{1}{1/Z_s' + g_m + g_{mb}} \qquad \frac{1}{c(s)} = Z_L'$$

$$\therefore \left.\frac{v_a(s)}{i_i(s)}\right|_{C_{xy}=0,\ r_{ds}=\infty} = \frac{Z_L'(g_m + g_{mb})}{1/Z_s' + g_m + g_{mb}} = \frac{Z_L' Z_s' (g_m + g_{mb})}{1 + (g_m + g_{mb}) Z_s'} \quad (7)$$

$$\rightarrow Z_L' \text{ as } |Z_s'| \rightarrow \infty$$

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Inspection of (6) $\Rightarrow$ (7)

Q: Can we find $\left.\frac{v_a(s)}{i_i(s)}\right|_{C_{xy}\neq 0,\ r_{ds}\neq\infty}$ from (5) by inspection?

A: Yes. Using Mason's Gain Formula (M.G.F.)

Application of MGF to (5)

"Loops":

1: $A \to B \to A$

2: $A \to C \to A$

3: $A \to C \to B \to A$

4: $A \to B \to C \to A$

5: $B \to C \to B$

Note: All loops touch each other

"Forward Paths":

1: $i_i \to A \to B \to v_d$

2: $i_i \to A \to C \to B \to v_d$

Note: deleting either path breaks all loops

"Loop gains":

$$
\left.
\begin{aligned}
L_1 &= \left(\frac{1}{r_{ds}} + g_m + g_{mb}\right)\frac{1}{r_{ds}\, b(s)\, C(s)} \\
L_2 &= sC_{gs}\left(g_m + sC_{gs}\right)\cdot\frac{1}{a(s)\cdot b(s)} \\
L_3 &= sC_{gs}\left(sC_{gd} - g_m\right)\frac{1}{r_{ds}\, a(s)\, b(s)\, C(s)} \\
L_4 &= sC_{gd}\left(g_m + sC_{gs}\right)\left(\frac{1}{r_{ds}} + g_m + g_{mb}\right)\cdot\frac{1}{a(s)\, b(s)\, C(s)} \\
L_5 &= sC_{gd}\left(sC_{gd} - g_m\right)\frac{1}{a(s)\, C(s)}
\end{aligned}
\right\} \quad (8)
$$

"Path gains"

$$
\left.
\begin{aligned}
P_1 &= \left(\frac{1}{r_{ds}} + g_m + g_{mb}\right)\cdot\frac{1}{b(s)\, C(s)} \\
P_2 &= sC_{gs}\left(sC_{gd} - g_m\right)\cdot\frac{1}{a(s)\, b(s)\, C(s)}
\end{aligned}
\right\} \quad (9)
$$

(units: Ω)

Fact M.G.F. (this case) $\Rightarrow$ $$\frac{v_a(s)}{i_i(s)} = \frac{P_1 + P_2}{1 - L_1 - L_2 - L_3 - L_4 - L_5} \quad (10)$$

(Both paths touch all the loops)

$\therefore$ Can easily find $G(s) = \frac{v_a(s)}{i_i(s)}$ but it's really messy to look at

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For negligible capacitance in Z_L, Z_B and Z_s then

$$G(s) \cong \frac{\text{second order poly.}}{\text{third order poly.}}$$ ($a(s)$, $b(s)$ & $c(s)$ are 1^{st} order polys.) } conductance (Siemens)

(Call full version of $G(s)$ (11))

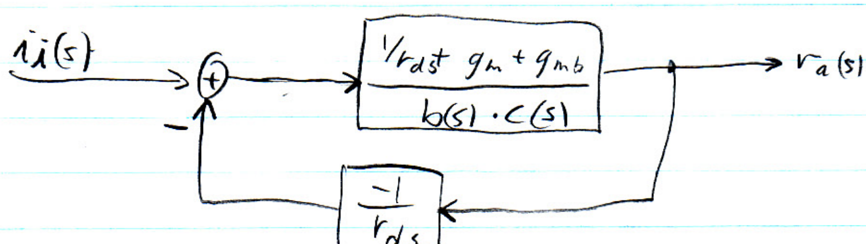
⇓

have third-order system with {2 zeros, 3 poles}

The good news is (11) = "exact" expr.
The bad news is —— " ——
(exact expr. is too complicated to give insight)

<u>Q</u> So what now?
<u>A</u> Plug in numbers (i.e. freq. and comp. values) into boxes in (5) and eliminate highly attenuated paths.

<u>Ex</u> Z_B usually arrises from parasitics
-often $Z_B \neq 0$ causes stability problems in feedback applications
V_{BIAS} = DC, so we can use bypass cap to reduce $|Z_B|$
∴ Suppose $|Z_B| \cong 0$ Then $\left|\frac{1}{a(s)}\right| \approx 0$ and (5) becomes ≈



Now, MGF gives $$G(s)\Big|_{|Z_B|=0} = \frac{\left(\frac{1/r_{ds} + g_m + g_{mb}}{b(s) \cdot c(s)}\right)}{1 - \frac{1}{r_{ds}}\left[\frac{(1/r_{ds} + g_m + g_{mb})}{b(s)\ c(s)}\right]}$$

let $g_m' = g_m + g_{mb}$ $(\approx 1.2 \cdot g_m)$ assume $g_m \gg \frac{1}{r_{ds}}$

then

$$b(s) = \frac{1}{Z_s'} + g_m' + sC_{gs}, \quad \underbrace{c(s) = \frac{1}{Z_L'} + \frac{1}{r_{ds}} + sC_{gd}}_{\text{(as before)}}$$

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$$\therefore \quad G(s)\Big|_{Z_B=0} = \frac{g_m'}{\left(\frac{1}{Z_s'} + g_m' + sC_{gs}\right)\left(\frac{1}{Z_i'} + \frac{1}{r_{ds}} + sC_{gd}\right) - g_m'/r_{ds}} \qquad (13)$$

Sanity check: 1) units V/I $\boxed{\checkmark}$

2) $A(j0)\Big|_{r_{ds}, Z_s = \infty} = Z_i'$ $\boxed{\checkmark}$

(13) $\Rightarrow$ 2^{nd} order behaviour: 2 poles, no zeros

$\Rightarrow$ always stable denom. of (13) has form $\alpha_0 + \alpha_1 \cdot s + \alpha_2 \cdot s^2$

$$\text{with poles} = -\frac{\alpha_1}{2\cdot\alpha_2} \pm \frac{1}{2\cdot\alpha_2}\cdot\underbrace{\sqrt{\alpha_1^2 - 4\alpha_0\alpha_2}}_{<\alpha_1}$$

$\therefore \mathrm{Re}\{\text{poles}\} < 0$

1/6.

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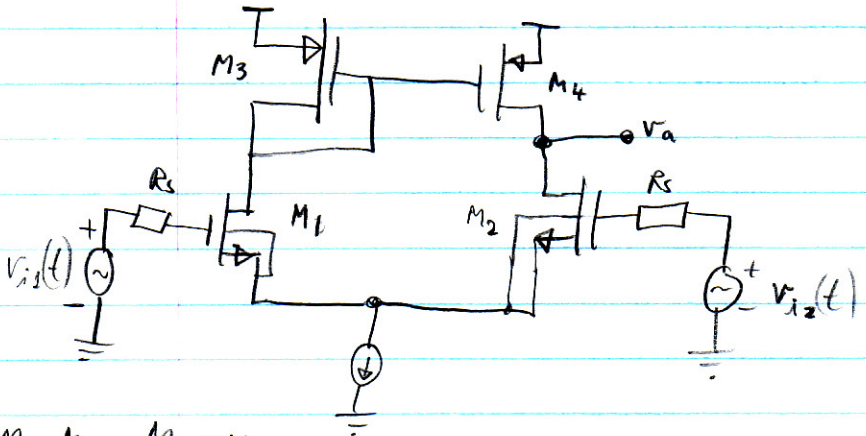
closed book/notes (bring calc.) } mid-term
↳ papir + penn (blyant)

Block diagrams & MGF (continued)

Ex: Diff to single-ended OTA → Operational transconductance amplifier

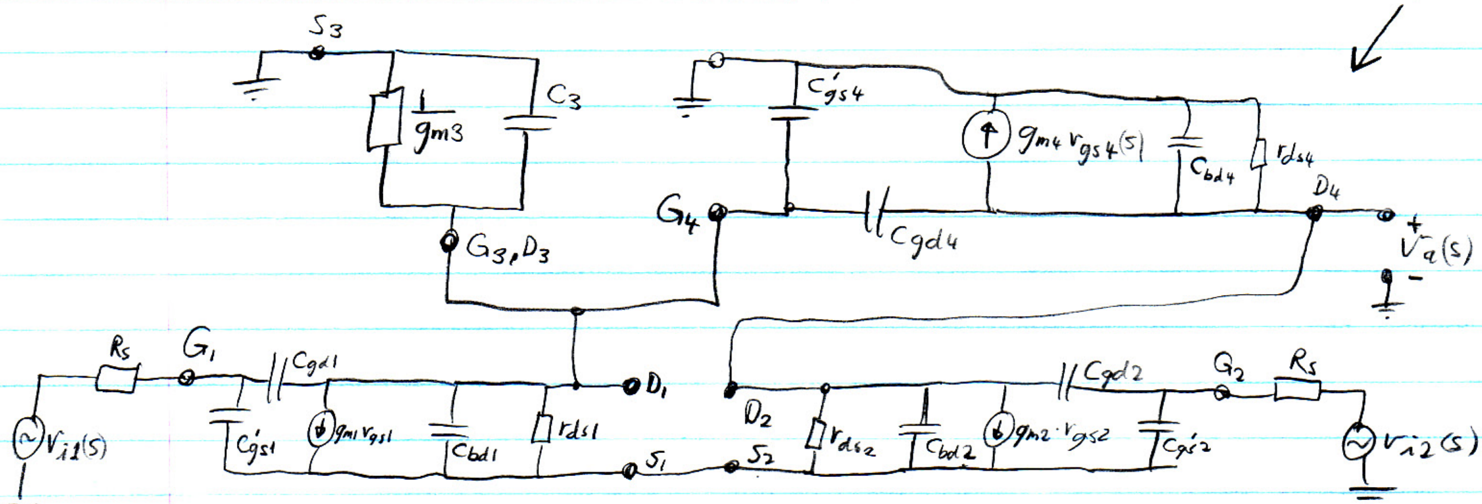


(1)

$M_1 = M_2$, $M_3 = M_4$

SSM of (1)

(2)



where $C'_{gsi} = C_{gsi} + C_{gbi}$, $C_3 = C'_{gs3} + C_{bd3}$

See textbook pp. 180 ~ 181

Want to analyze differential mode (DM) operation: $v_{i1} = \frac{v_{id}}{2}$ $v_{i2} = -\frac{v_{id}}{2}$

Assume R_s = **negligible** for analysis

Want BD. containing nodes: $\underbrace{v_{id}}_{\text{Input}}$, $\underbrace{v_{d1}}_{\text{Drain 1}}$, $\underbrace{v_s \equiv v_{s1} \equiv v_{s2}}_{\text{Sources}}$, $\underbrace{v_a}_{\text{output}}$

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KCL at D_1

$$v_{d1}\left[g_{m3}+s\left(C_3+C'_{gs4}\right)\right]+\frac{v_{d1}-v_s}{r_{ds1}\| C_{bd1}}+\left(v_{d1}-v_a\right)sC_{gd4}+\left(v_{d1}-\frac{v_{id}}{2}\right)sC_{gd1}$$
$$+\left(\frac{v_{id}}{2}-v_s\right)g_{m1}=0$$

where "$r_x \| C_y$" $\equiv \left(\frac{1}{r_x}+sC_y\right)^{-1}$

$$\underbrace{v_{d1}\left[g_{m3}+\frac{1}{r_{ds1}}+s\left(C_3+C'_{gs4}+C_{bd1}+C_{gd4}+C_{gd1}\right)\right]}_{\text{(call it } a(s)\text{)}}-v_s\underbrace{\left[g_{m1}+\frac{1}{r_{ds1}\| C_{bd1}}\right]}_{\approx g_{m1}+sC_{bd}}$$
$$-v_a sC_{gd4}+\frac{1}{2}v_{id}\left(g_{m1}-sC_{gd1}\right)=0$$

$$\therefore\ v_{d1}=\frac{1}{a(s)}\cdot\left[v_s\left(g_{m1}+sC_{bd1}\right)+v_a\cdot s\cdot C_{gd4}-\frac{1}{2}v_{id}\left(g_{m1}-sC_{gd1}\right)\right] \qquad (3)$$

KCL at D_2

...

$$\therefore\ v_a \cong \frac{1}{b(s)}\left[v_s\left(g_{m2}+sC_{bd2}\right)+v_{d1}\left(sC_{gd4}-g_{m4}\right)+\frac{1}{2}v_{id}\left(g_{m2}-sC_{gd2}\right)\right] \qquad (4)$$

where $b(s)=\dfrac{1}{r_{ds4}\| r_{ds2}}+s\left(C_{bd4}+C_{bd2}+C_{gd4}+C_{gd2}\right)$ (used $g_{m2} \gg \frac{1}{r_{ds2}}$)

KCL at S_1, S_2

...

$$\therefore\ v_s=\frac{1}{c(s)}\left[\frac{v_{d1}}{r_{ds1}\| C_{bd1}}+\frac{v_a}{r_{ds2}\| C_{bd2}}\right] \qquad (5)$$

where $c(s)=g_{m1}+g_{m2}+s\left(C'_{gs1}+C'_{gs2}\right)$

used $M_1=M_2$ & $M_3=M_4$

(3) ~ (5) $\Longrightarrow$ block diagram

$$\boxed{\left[\frac{1}{a(s)}\right]=\left[\frac{1}{b(s)}\right]=\left[\frac{1}{c(s)}\right]=\Omega} \leftarrow 1^{st} \text{ order polynomials}$$

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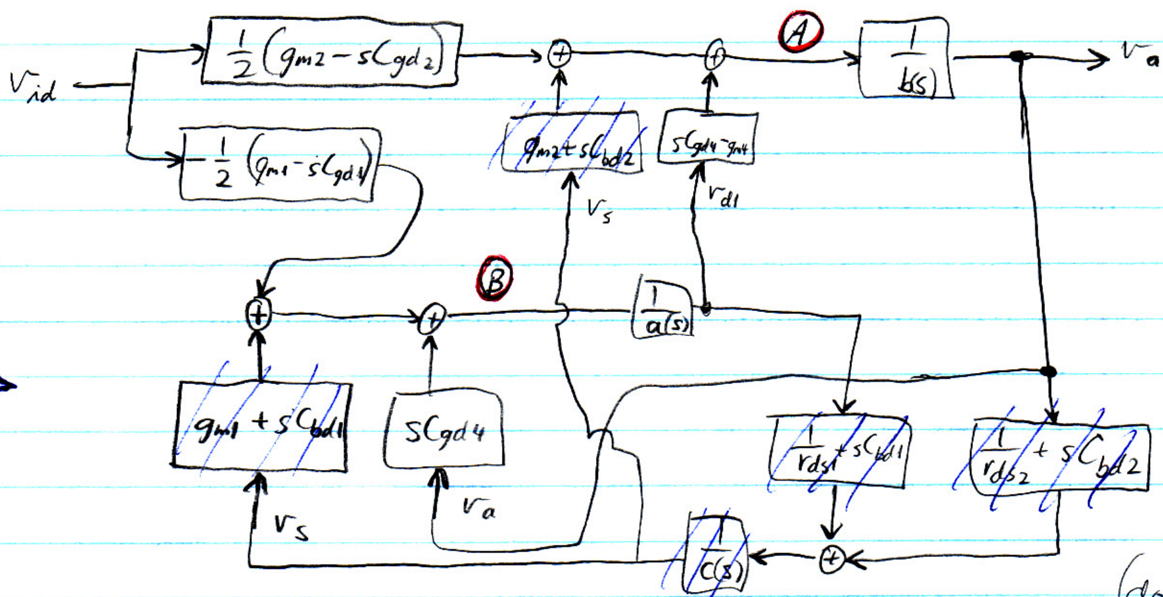

6

(dominant)

Let $p_1, p_2, \ldots$ be the poles $A_v(s) = \frac{v_a(s)}{v_{id}(s)}$ with $|p_1| \le |p_2| \le \ldots$

Provided $\left|\frac{1}{r_{ds1,2}} + j\omega C_{bd1,2}\right|$ is sufficiently small that

$$|v_s(j\omega)| << |v_{d1}(j\omega)|, |v_a(j\omega)| \text{ for } |\omega| \le |p_2| \quad (7)$$

can eliminate the v_s feedback paths in (6)
(asymmetry $\Rightarrow v_s \neq 0$)

We'll assume (7) holds for now. Can later check the assumption by testing with resulting value of p_2.

Sanity check: Find $A_{v_0} = A_v(j\omega)\big|_{\omega=0}$ $\quad \omega = 0 \Rightarrow s = 0$

$$\therefore a(0) = g_{m3} \qquad b(0) = \frac{1}{r_{ds2} \| r_{ds4}}$$

$$\therefore (6) \Rightarrow v_a = \left(\frac{1}{2} g_{m2}\, r_{ds2} \| r_{ds4} - \frac{1}{2} g_{m1} \cdot \frac{-g_{m4}}{g_{m3}} \cdot r_{ds2} \| r_{ds4}\right) v_{id}$$

$$\therefore A_{v_0} = g_{m1} \cdot r_{ds2} \| r_{ds4} \quad \boxed{\checkmark}$$

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Now, apply MGF:

Loop: $A \rightarrow B \rightarrow A$

Forward paths 1. $v_{id} \rightarrow A \rightarrow v_a$

2. $v_{id} \rightarrow B \rightarrow A \rightarrow v_a$

6 $\Rightarrow$ $P_1 = \dfrac{g_{m2} - sC_{gd2}}{2 \cdot b(s)}$, $P_2 = \dfrac{(g_{m1} - sC_{gd1})(g_{m4} - sC_{gd4})}{2 \cdot a(s) \cdot b(s)}$

$$L_1 = \frac{sC_{gd4}(sC_{gd4} - g_{m4})}{a(s)\ b(s)}$$

MGF $\Rightarrow$ $A_v(s) = \dfrac{P_1 + P_2}{1 - L_1}$

$$= \frac{1}{2} \cdot \frac{a(s)(g_{m2} - sC_{gd2}) + (g_{m1} - sC_{gd1})(g_{m4} - sC_{gd4})}{a(s)\,b(s) - sC_{gd4}(sC_{gd4} - g_{m4})} \qquad (8)$$

$\Rightarrow$ 2 zeros, 2 poles

2^{nd} order $\Rightarrow$ can easily find poles & zeros, but results give little insight

Instead, note:

- Feedback depends on $s \cdot C_{gd4}$ (no Miller effect, why?) $\hookrightarrow Z = 0$ (source)
- Usually $C_{gd4} \ll C'_{gsi}, C_{bdi}$
- $\therefore$ Taking $s \cdot C_{gd4} \cong 0$ in (8) = reasonable approximation

Now, MGF $\Rightarrow$ $A_v(s) = P_1 + P_2$ $(L_1 \approx 0)$ $= \dfrac{2^{nd} \text{ order poly}}{a(s)\ b(s)}$

$$= g_{m1}(r_{ds2} \| r_{ds4}) \cdot \frac{(1 - s/z_1)(1 - s/z_2)}{(1 - s/p_1)(1 - s/p_2)}$$

where $p_1 = \dfrac{-1}{(r_{ds2} \| r_{ds4})\, C_2}$, $p_2 = \dfrac{-g_{m3}}{C_1}$

where $C_1 = C'_{gs3} + C_{bd3} + C'_{gs4} + C_{bd1} + C_{gd1}$

$C_2 = C_{bd2} + C_{bd4} + C_{gd2}$

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C_1, C_2 have same order of mag
$r_{ds2} \| r_{ds4} \gg 1/g_{m3}$ (typically) $\Rightarrow$ p_1 = dom pole
p_2 = non dom. pole (next most dominant)

Observation

$\frac{1}{g_{m3}}$ = resistence to small-signal ⏚ at D_1 in (2)

C_1 = cap —— " ——

$r_{ds2} \| r_{ds4}$ = res —— " —— D_2 —"—

C_2 = cap —— " ——

∴ In (2) could have found p_1 and p_2 by calculating R_i, C_i where R_i = res to gnd. at node i
C_i = cap. —— " ——

Then $p_1 = \frac{-1}{\text{largest}\{R_i \cdot C_i\}}$, $p_2 = \frac{-1}{\text{next largest}\{R_i \cdot C_i\}}$

Q : Coincidence ?
A : No. This method is reasonable approx in many cases (HW 2)

Slang : Because of this, people often say that the dominant pole occurs at node D_2 and the non-dominant pole occurs at node D_1.

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Full version of MGF

Block diagram defs.:

1) "Path" ≡ route through B.D. (in dir. of arrows) connecting a pair of nodes
2) "Path gain" ≡ product of block gains along path
3) "Loop" ≡ path which starts and ends at same node with no node along path encountered more than once
4) "Loop gain" ≡ path gain of loop
5) "Determinant" ≡ 1 − Σ (all loop gains)
   + Σ (products of loop gains of all loop ~~pairs~~ pairs with no common nodes)
   − Σ (—— " —— triples —— " ——)
   + Σ (. . .)
   " " "
6) "Forward path" ≡ path from input to output containing no full loops

Let $H = \frac{X_{out}}{X_{in}}$ where X_{in} = input of B.D.
X_{out} = output of B.D.

Then $H = \frac{1}{\Delta} \cdot \sum_{k=1}^{L} P_k \cdot \Delta_k$ (MGF)

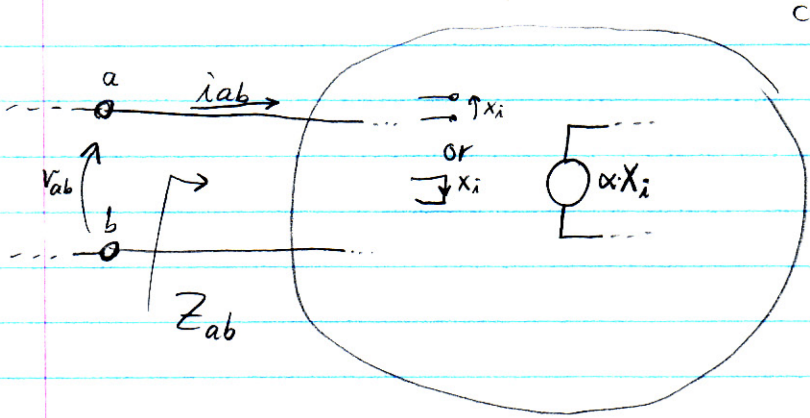
where Δ = determinant
L = # of forward paths
P_k = k^{th} forward path
Δ_k = determinant of B.D. that remains after deleting k^{th} path, P_k

"cofactor" = Δ with loops touching the k^{th} path removed.

ECE 264A Jan 31, 2008 E.G.

Blackman's Impedance Relation

arbitrary SSM circuit containing ≥1 controlled sources



x_i = voltage or current
αx_i = — " —

(e.g. $X_i = v_{gs}$, $\alpha = g_m$
∴ αx_i = current)

BIR $\quad Z_{ab} = Z_{ab}^0 \cdot \dfrac{1+T_{sc}}{1+T_{oc}}$ (1)

$\left(\equiv \dfrac{v_{ab}(s)}{i_{ab}(s)} \right)$

nullstilt!

where $Z_{ab}^0 \equiv$ Impedance between a and b with controlled source removed ($\alpha = 0$)

$T_{sc} \equiv -\alpha \cdot \dfrac{x_i}{x_x}$ when $v_{ab} = 0$ (a, b shorted)

and αx_i is replaced by an independent "test source", X_x

$T_{oc} \equiv$ Same as T_{sc} except with $i_{ab} = 0$ (a, b open circuited)

Note: if have more than one controlled sources, pick any one of them (call it the reference source).

$\boxed{\text{T'ene er "loop gain"}}$

See note of March 18, 2008

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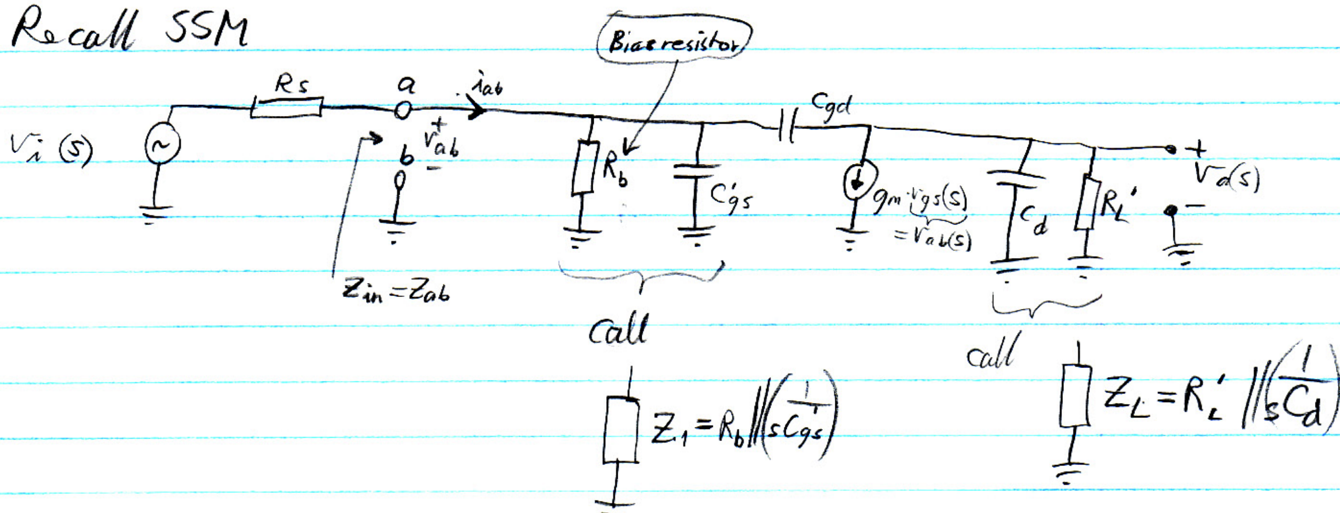
Ex 1 C.S. Amp

Recall SSM



BIR: $\alpha = g_m$, $x_i = v_{ab}$, $x_x = i_x$

$$Z_{ab}^{\circ} = Z_1 \| \left(\frac{1}{sC_{gd}} + Z_L\right) = \frac{Z_1 \cdot (1 + Z_L \cdot s \cdot C_{gd})}{1 + (Z_1 + Z_L) \cdot s \cdot C_{gd}}$$

$$T_{sc} = 0 \quad (\text{because } x_i = v_{ab} = 0)$$

$$T_{oc} = -g_m \cdot \frac{1}{i_x} \cdot \Big[\underbrace{-i_x \left(Z_L \| \left(Z_1 + \tfrac{1}{sC_{gd}}\right)\right)}_{v_d(s)} \cdot \frac{Z_1}{Z_1 + \frac{1}{sC_{gd}}} \Big]$$

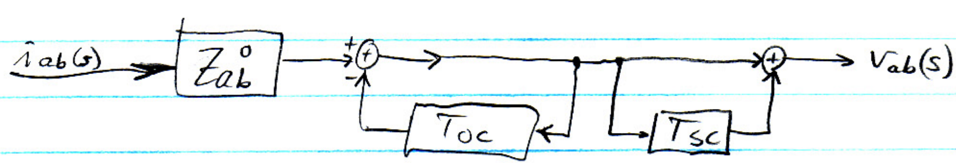
$$= \frac{g_m Z_1 Z_L}{Z_L + Z_1 + \frac{1}{sC_{gd}}} \quad (\text{note} \to 0 \text{ as } C_{gd} \to 0)$$

$$\therefore \; ① \Longrightarrow \; z_{in} = Z_{ab}^{\circ} \cdot \frac{1}{1 + T_{oc}} = Z_1 \cdot \frac{1 + sC_{gd} \cdot Z_L}{1 + [Z_L + \underbrace{Z_1(1 + g_m Z_L)}_{\text{Miller effect}}] \cdot s \cdot C_{gd}} \qquad ②$$

Observations

1) T_{oc} represents a feedback path around the reference source
Why? def ⟹ If output of ref. source has no connection to it's controlling voltage (or current) then $T_{oc} = 0$

Also, MGF and ① give:



(3/6)

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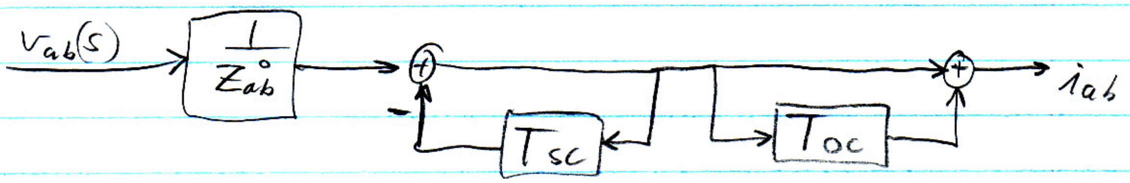
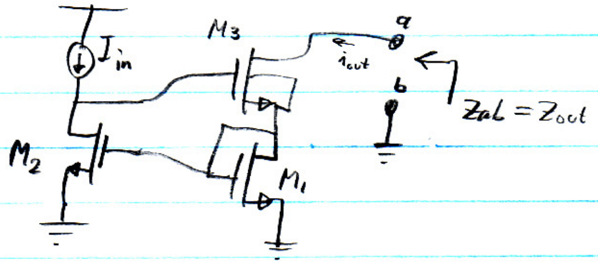
Or...



∴ T_{sc} also arises from a feedback path within the circuit

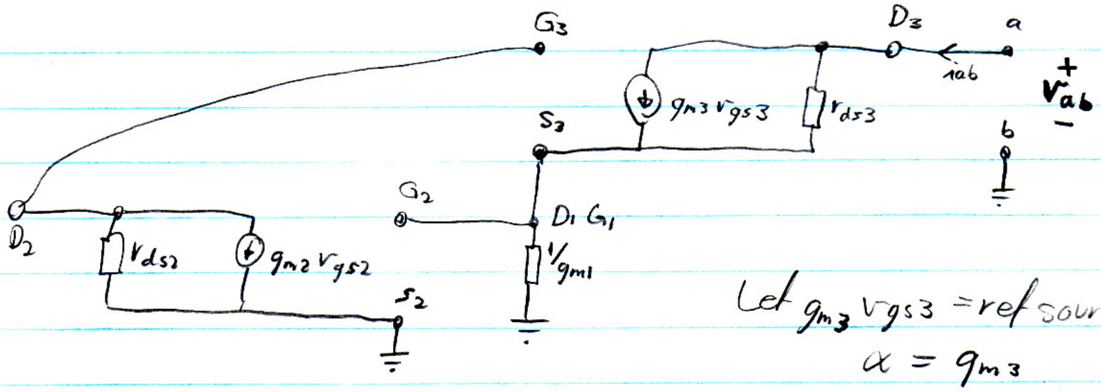
2) Feedback can either increase or decrease Z_{ab}

Ex 2 Wilson current mirror (low freq analysis)



Low freq. SSM (using  $\cong$  $1/g_m$)

(Set $i_{in} = 0$ to find Zout)



③

Let $g_{m3} v_{gs3}$ = ref source ∴

$\alpha = g_{m3}$

$x_i = v_{gs3}$

Inspection of ③ $\Rightarrow Z^{o}_{ab} = r_{ds3} + \frac{1}{g_{m1}} \approx r_{ds3}$

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T_{oc} : $i_{ab}=0 \Rightarrow v_{gs2}=0 \Rightarrow v_{gs3}=0 \Rightarrow x_i=0 \therefore T_{oc}=0$

T_{sc} : with $g_{m3}v_{gs3}$ replaced by i_x and $v_{ab}=0$ have

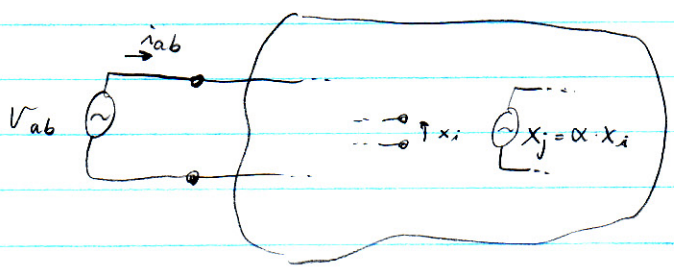
$$\left.\begin{aligned} v_{gs2} &= i_x\left(r_{ds3}\,\|\left(\tfrac{1}{g_{m1}}\right)\right) \approx \frac{i_x}{g_{m1}} \\ \text{KVL} \Rightarrow v_{gs2}+v_{gs3} &= -g_{m2}\cdot v_{gs2}\cdot r_{ds2} \end{aligned}\right\} \Rightarrow v_{gs3} = -\frac{i_x}{g_{m1}}\left(1+g_{m2}r_{ds2}\right)$$

$$\therefore T_{sc} = \frac{g_{m3}}{g_{m1}}\left(g_{m2}\cdot r_{ds2}+1\right) = g_{m3}r_{ds2}+1$$

(for $M_1=M_2=M_3$)

$\therefore$ (1) $\Rightarrow$ $Z_{ab} = r_{ds3}\left(2+g_{m2}\cdot r_{ds2}\right)$ $\therefore$ Feedback increased Z_{ab}

Proof of BIR:



Linearity $\Rightarrow$ can write

$$\left.\begin{aligned} v_{ab} &= A\cdot i_{ab}+B\cdot x_j \\ x_i &= C\cdot i_{ab}+D\cdot x_j \end{aligned}\right\} \quad (4)$$

(two-port representation)

$A(s), B(s), C(s), D(s) \equiv$ transfer functions

Solving (4) for $Z_{ab}=\frac{v_{ab}}{i_{ab}}$ gives $Z_{ab}=A\cdot\dfrac{1-\frac{\alpha(AD-BC)}{A}}{1-\alpha D}$ (5)

(4) $\Rightarrow$ $A=\left.\frac{v_{ab}}{i_{ab}}\right|_{x_j=0}$ = def of Z_{ab}^{0} (6)

Now suppose $v_{ab}=0$ and $x_j=x_x$ (indep. of x_i)

Then (4) $\Rightarrow$ $i_{ab}=-\frac{B}{A}\cdot x_x$

$x_i = C\cdot i_{ab}+D\cdot x_x$

$$\Rightarrow \underbrace{-\alpha\frac{x_i}{x_x}}_{\equiv T_{sc}} = -\frac{\alpha(AD-BC)}{A} \quad (7)$$

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Now suppose $i_{ab}=0$ and $x_j = x_x$ (indep. of x_i)

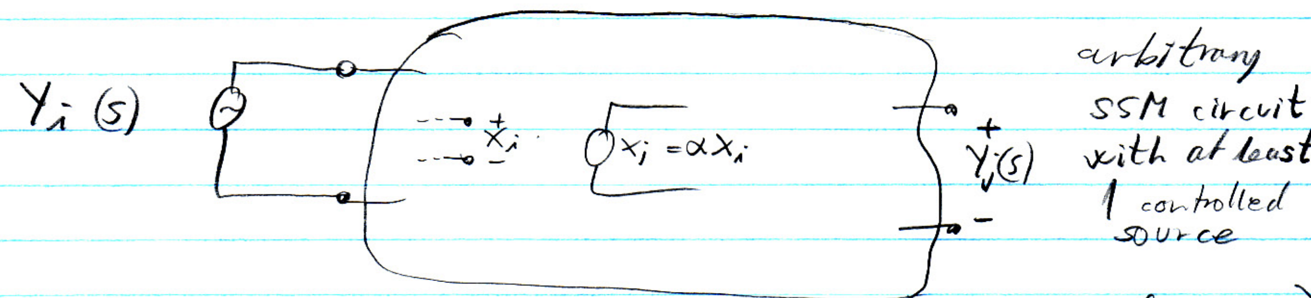
Then 4 similarly $\Rightarrow$ $T_{oc} = -\alpha D$ 8

$\therefore$ 5 ~ 8 $\Rightarrow$ $Z_{ab} = Z_{ab}^{\circ} \cdot \dfrac{1+T_{sc}}{1+T_{oc}}$ $\boxed{\checkmark}$

$\square$

Asymptotic Gain Relation

9



$X_i \neq Y_i$ required

$\left.\begin{array}{l} x_i, x_j \\ Y_i, Y_j \end{array}\right\}$ voltages or currents

AGR: $A(s) = A_\infty \cdot \dfrac{T}{1+T} + A_0 \cdot \dfrac{1}{1+T}$ 10

$\left(= \dfrac{Y_j(s)}{Y_i(s)}\right)$

where $\underbrace{T \equiv -\alpha \cdot \dfrac{x_i}{x_x}}_{\text{"loop gain"}}$ where $Y_i = 0$ and X_j replaced by an independent "test source", X_x

$$A_\infty(s) \equiv \left.\frac{Y_j(s)}{Y_i(s)}\right|_{\alpha \to \infty} \equiv \text{"asymptotic gain"}$$

$$A_0(s) \equiv \left.\frac{Y_j(s)}{Y_i(s)}\right|_{\alpha = 0} \equiv \text{"direct transmission term"}$$

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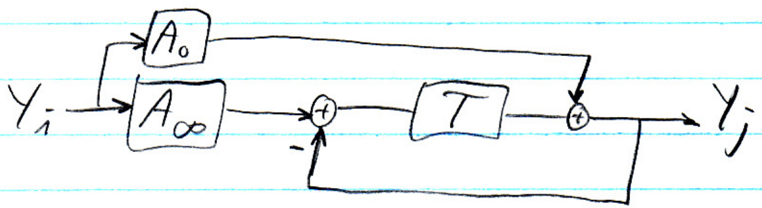
Observations

MGF & 10 $\Rightarrow$ can redraw 9 as



$\Rightarrow$ large T $\Leftrightarrow$ feedback dominates behavior $\Rightarrow$ $A(s) \cong A_\infty(s)$

$A_o(s)$ arises from forward path through feedback network

"Reference source" in BIR & AGF:

See note of March 18, 2008

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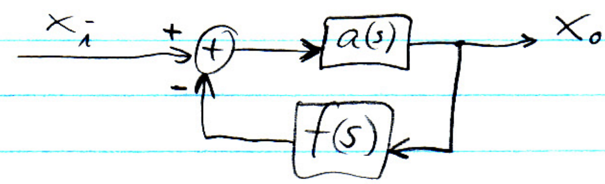
ECE 264A Feb. 12, 2008 E.G.

Avg.: 68 } Mid-term
Med.: 70 }

Feedback

Let $x_i(s)$ = input signal (V or I)
$x_o(s)$ = output — " —

Then a circuit incorporates feedback if its B.D. can be reduced to:

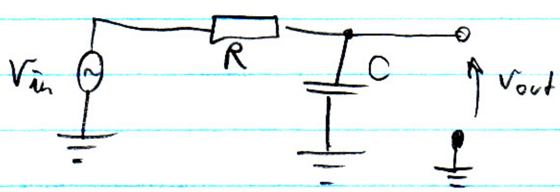


(1)

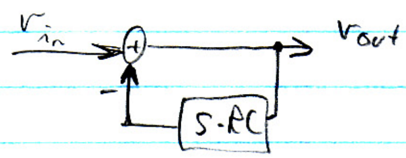
MGF + (1) ⇒

$$A(s) = \frac{a(s)}{1+a(s)f(s)} \quad (2) \quad \text{where } A(s) \equiv \frac{x_o(s)}{x_i(s)}$$

Ex 1 "Passive feedback"



⇒



∴ $a(s) = 1$
$f(s) = sRC$

(more generally, poles ⇒ feedback)

$$\frac{v_{out}}{v_{in}} = \frac{\frac{1}{sC}}{\frac{1}{sC}+R} = \frac{1}{1+sRC} \Rightarrow v_{out} = v_{in} - v_{out} \cdot s \cdot RC$$

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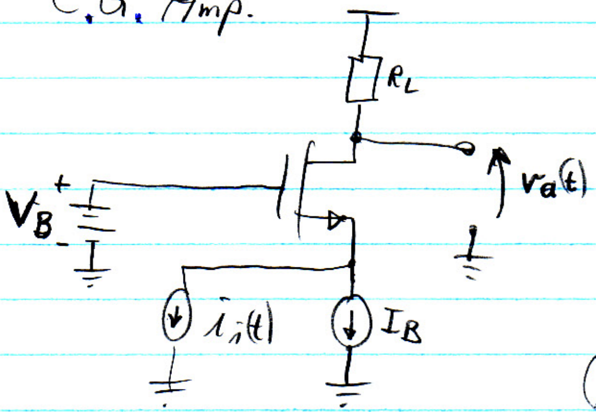
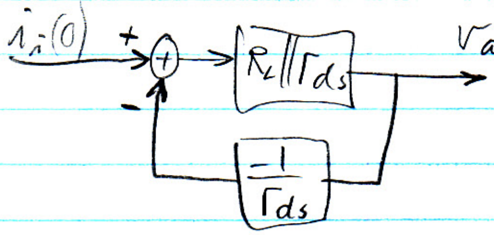
Ex 2 "Parasittic feedback"

C.G. Amp.

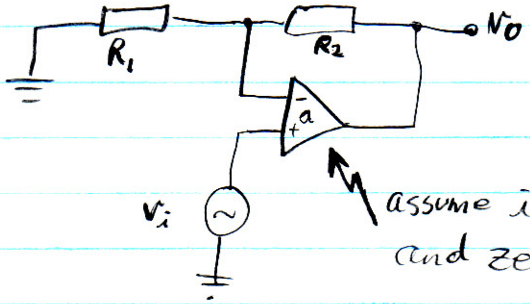


$\Rightarrow$ (found prev.) (low freq. analysis)



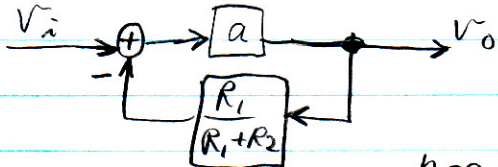
Note: In Ex1 feedback reduces the magnitude of X_o ≡ "neg. feedback"
In Ex2 — " — increases — " — ≡ "pos. feedback"

Ex 3 "Intentional active feedback"



assume infinite inp. impedance and zero output — " —

$$v_o = a\left[v_i - \frac{R_1}{R_1+R_2}\cdot v_o\right]$$



$$A = \frac{a}{1+a\left(\frac{R_1}{R_1+R_2}\right)} \rightarrow 1+\frac{R_2}{R_1} \text{ as } a\to\infty$$

neg. feedback

$\therefore$ $\underbrace{\text{large } a}_{\text{easy in IC's}} \Rightarrow$ A depends only of the $\underbrace{\text{ratio of R's}}_{\text{accurate in IC's}}$

3

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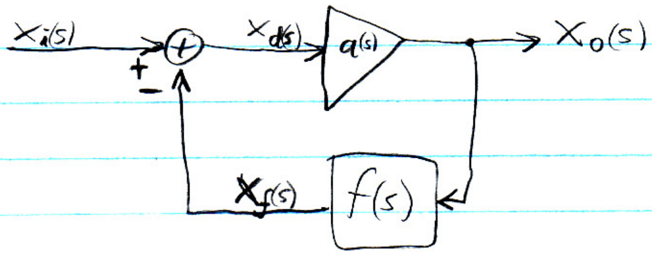
Feb. 12, 2008

Intentional active feedback ⇒ feedback around high gain amp
⇒ many benefits (e.g. 3)

Negative Intentional Active Feedback

Defs.: In the following feedback system



(4)

$a(s) \equiv$ "open loop gain", $f(s) \equiv$ "feedback factor"

$T(s) \equiv a(s)\cdot f(s) \equiv$ "loop gain", $A(s) \equiv$ "closed loop gain"
(recall $A(s) = \frac{a(s)}{1+a(s)\cdot f(s)}$)

Let $A_{ideal} = \lim_{a\to\infty} A = \frac{1}{f}$ (e.g. $A_{ideal} = 1 + \frac{R_2}{R_1}$ in Ex 3)

Asymptotic gain-relation: jan. 31

$A = A_{ideal} \cdot \frac{T}{1+T}$

$\boxed{A = \frac{1}{f}\left(1 - \frac{1}{1+a\cdot f}\right) = \frac{1}{f}\cdot\frac{af}{1+af} = \frac{a}{1+af}}$

$\therefore$ (2) $\Rightarrow$ $A = A_{ideal}\Big(1 - \underbrace{\frac{1}{1+T}}\Big)$ fractional deviation of A from A_{ideal}

Also, MGF & (4) ⇒

$x_d = \frac{1}{1+T}\cdot x_i \to 0$ as $T\to\infty$ $(a\to\infty)$

$x_f = \frac{1}{1+1/T}\cdot x_i \to x_i$ as $T\to\infty$ $(a\to\infty)$

$\therefore x_f$ "tracks" x_i with "error signal" x_d (d = difference)

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Feb. 12, 2008

Benefits: (i) gain desensitivity
(ii) non-linear distortion reduction
(iii) in-loop noise suppression
(iv) broadbanding
(v) input/output impedance adjustment

(Potential) Drawbacks (i) $A < a$
(ii) excessive phase shift can cause instability

$\left(\frac{U}{V}\right)' = \frac{U'V - UV'}{V^2}$

(i) Gain desensitivity

$$\text{②} \Rightarrow \frac{dA}{da} = \frac{1}{(1+af)^2} \Rightarrow \frac{\Delta A}{\Delta a} \cong \frac{1}{(1+af)^2} = \frac{A}{a} \cdot \frac{1}{1+a \cdot f}$$

small Δa

$$\therefore \frac{\Delta A}{A} \cong \left(\frac{1}{1+T}\right) \cdot \frac{\Delta a}{a} \Rightarrow \text{large variation in } a \Rightarrow$$

small — " — A

$\Rightarrow A$ is "stable" WRT variations in a

$$\text{②} \Rightarrow \frac{dA}{df} = -A^2$$

$$\therefore \frac{\Delta A}{\Delta f} \cong -A \cdot \underbrace{\frac{1}{f}}_{A_{ideal}} \cdot \frac{T}{1+T}$$

$$\therefore \frac{\Delta A}{A} \cong -\underbrace{\left(\frac{T}{1+T}\right)}_{\cong 1 \text{ for for } |T| \gg 1} \frac{\Delta f}{f} \Rightarrow A \text{ is } \underline{\text{not}} \text{ "stable" WRT variations}$$

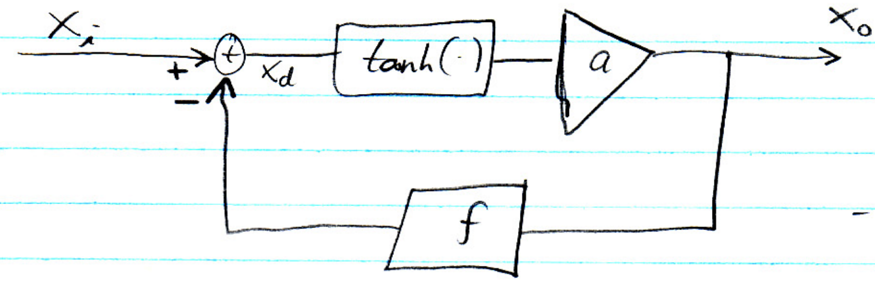
$\Rightarrow$ want high gain amplifier (not nec. well defined) and precise feedback factor.

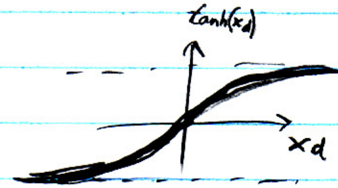
$$\boxed{\frac{\Delta A}{A} = \frac{-T}{1+T} \cdot \frac{\Delta f}{f} \quad \Bigg| \quad \frac{\Delta A}{A} = \frac{1}{1+T} \cdot \frac{\Delta a}{a}}$$

Feb 12, 2008

(ii) Non-linear distortion reduction

Ex 4





$$x_o = a \cdot \tanh(x_i - f \cdot x_o)$$

$$\therefore \tanh^{-1}\left(\frac{x_o}{a}\right) = x_i - f \cdot x_o$$

$$\therefore \frac{x_o}{a} + \frac{x_o^3}{3a^3} + \frac{x_o^5}{5a^5} + \ldots = x_i - f x_o$$

$$\therefore x_o + \underbrace{\frac{x_o^3}{3a^2} + \frac{x_o^5}{5a^4} + \ldots}_{\to 0 \text{ as } a \to \infty} = a(x_i - f x_o)$$

$$\therefore x_o \cong A \cdot x_i \text{ for large } a \text{ and } x_o \to \underbrace{\frac{1}{f} \cdot x_i}_{\text{no distortion}} \text{ as } a \to \infty$$

Heuristics:

1) Follows from gain desensitivity (distortion prior to a $\Leftrightarrow$ input dependent gain variation)
2) Large a $\Rightarrow$ small x_d $\Rightarrow$ small portion of non-linear curve is traversed $\Rightarrow$ $\approx$ linear

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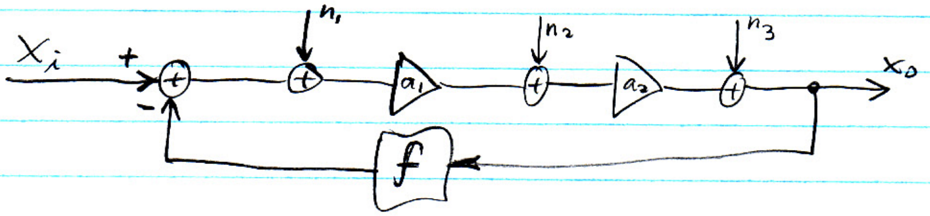
(iii) In-loop noise suppression



Open loop gain: $a = a_1 a_2$

$$MGF \Rightarrow x_o = \frac{a}{1+af}\left(x_i + n_1 + \frac{n_2}{a_1} + \frac{n_3}{a_1 a_2}\right)$$

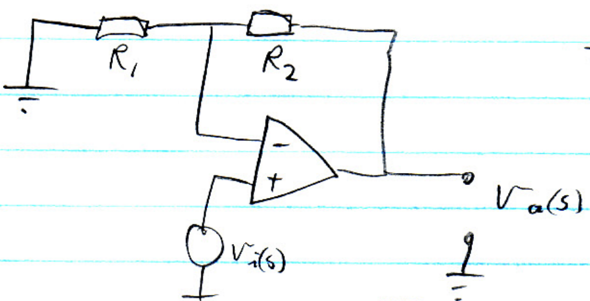
not suppressed — somewhat suppressed — most suppressed

(iv) Broadbanding

Ex 5 suppose $a \cong \frac{a_0}{1 - j\,\omega/\omega_{p0}}$ (dom. pole approx.)

$F \cong$ indep. of ω

e.g.



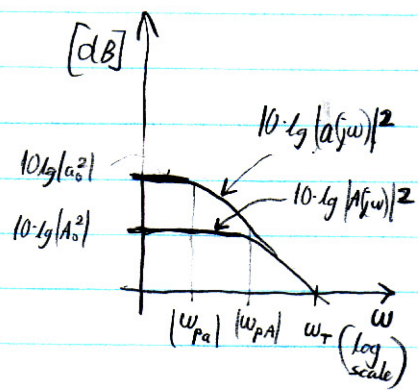
$$f = \frac{R_1}{R_1 + R_2}$$

Then (2) $\Rightarrow$ $\boxed{A = \frac{a}{1+af} = \frac{1}{f}\cdot\frac{af}{1+af} = \frac{1}{f}\cdot\frac{1}{1+\frac{1}{f}\cdot\frac{1}{a}} = A_{ideal}\cdot\frac{1}{1+1/T}}$

$$A(j\omega) = \underbrace{\left(1 + \frac{R_2}{R_1}\right)}_{1/f} \frac{1}{1 + \left(1 + R_2/R_1\right)\left(1 - j\,\omega/\omega_{p0}\right)/a_0}$$

algebra $\Rightarrow A(j\omega) = \frac{A_0}{1 - j\,\omega/\omega_{pA}}$ where $A_0 = \left(1 + \frac{R_2}{R_1}\right)\frac{1}{1 + \left(1 + \frac{R_2}{R_1}\right)/a_0}$ (5)

and $\omega_{pA} = \omega_{p0}\left(1 + a_0 \cdot \frac{R_1}{R_1 + R_2}\right)$ (6)



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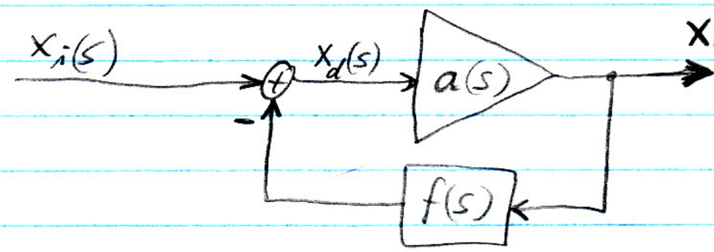
Negative intentional active feedback (cont. from pp. $\frac{3-6}{6}$ last time)

Recall: Can allways write the block diagram of a circuit with feedback as:



(1)

$a(s) \equiv$ "open loop gain"
$f(s) \equiv$ "feedback factor"
$T(s) \equiv a(s) \cdot f(s) \equiv$ loop gain ← (i M.G.F.: $T = -a \cdot f$ ulike definisjoner!)

$A(s) = \frac{x_o(s)}{x_i(s)} \equiv$ "closed loop gain"

$$MGF \Rightarrow A(s) = \frac{a(s)}{1 + a(s) \cdot f(s)} \quad (2)$$

Broadbanding (cont.)

Suppose: $a(j\omega) = \underbrace{a_o}_{\text{DC-gain}} \cdot \frac{1}{1 - j\,\omega/\omega_{pa}}$ (3)

↳ dominant pole approx.

$f(j\omega) = \text{const.}$

(2), (3) $\Rightarrow A(j\omega) = A_o \cdot \frac{1}{1 - j\,\omega/\omega_{pA}}$ where

$A_o = \frac{a_o}{1 + a_o \cdot f}$ (DC closed loop gain)

& $\omega_{pA} = (1 + a_o \cdot f) \cdot \omega_{pa}$

(4)

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$|A_0 \cdot \omega_{pA}| \equiv$ "closed loop gain bandwidth product", $GBW_{closed\ loop}$

$|a_0 \cdot \omega_{pa}| \equiv$ "open loop gain bandwidth product", $GBW_{open\ loop}$

$$④ \Rightarrow \quad A_0 \cdot \omega_{pA} = a_0 \cdot \omega_{pa}$$

$\therefore$ $GBW_{open\ loop} = GBW_{closed\ loop}$ in this case.

$\therefore$ Increasing f (between 0 and 1) decreases A_0, but increases the 3dB BW.

"Unity Gain Frequency"

$\equiv \omega_t$ such that $a(j\omega_t) = 1$, $\omega_t \in \mathbb{R}$

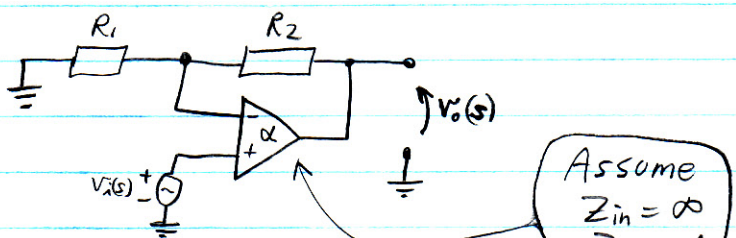
$$\therefore ③ \Rightarrow \quad \left| a_0 \cdot \frac{1}{1 - j\omega_t/\omega_{pa}} \right| = 1$$

$$\Rightarrow (a_0^2 - 1) \cdot \omega_{pa}^2 = \omega_t^2$$

$$\Rightarrow \omega_t \approx a_0 \cdot |\omega_{pa}| \qquad (a_0 \gg 1)$$

$\therefore$ $GBW = \omega_t$ in this case (not allways in other cases)

Ex 1 (from last time)



$$\Rightarrow ① \text{ if } a = \alpha, \quad f = \frac{R_1}{R_1 + R_2}$$

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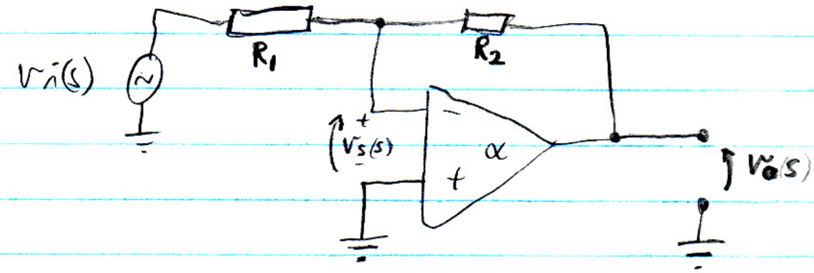
Ex2 (Same except different input location)



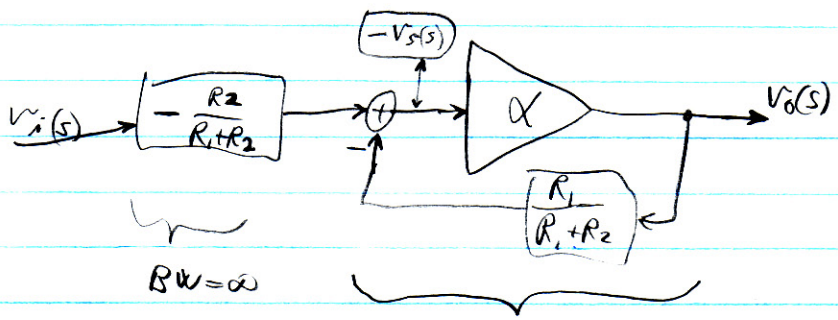
6

$$V_o(s) = \alpha(s) \cdot [-V_s(s)]$$

$$\therefore \; V_s(s) = V_i(s) - R_1 \cdot \left[\frac{V_i(s) - V_o(s)}{R_1 + R_2} \right]$$

$$\therefore \; -V_s(s) = -V_i(s) \underbrace{\left[1 - \frac{R_1}{R_1+R_2} \right]}_{= \frac{R_2}{R_1+R_2}} - V_o(s) \cdot \left[\frac{R_1}{R_1+R_2} \right]$$

$\therefore$ BD of 6 is:



$\equiv$ 1 with $a = \alpha$ & $f = \frac{R_1}{R_1+R_2}$ (Same as Ex. 1.)

$\Rightarrow$ BW = same as Ex 1

But GBW of 6 less than that of Ex 1 by $\frac{R_2}{R_1+R_2}$

Also, $GBW_{\text{Closed loop}} = \frac{R_2}{R_2+R_1} \cdot \omega_t = \frac{R_2}{R_1+R_2} \cdot \underbrace{a_0 \, |\omega_{pa}|}_{GBW_{\text{open loop}}}$

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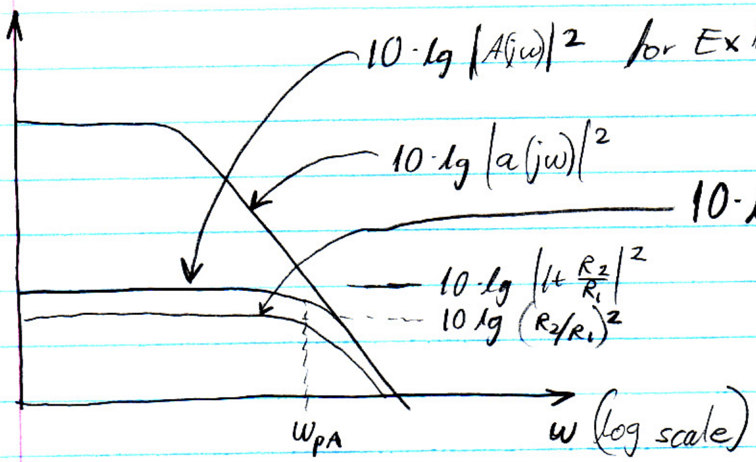
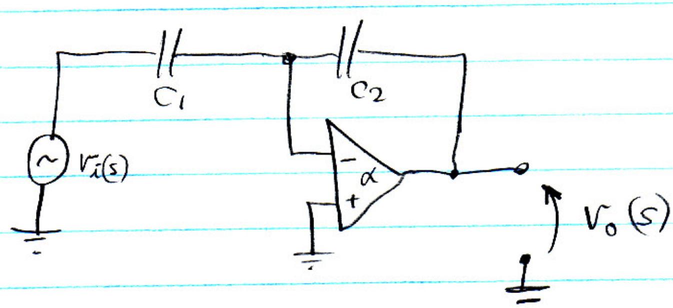


[Inverterende og ikke-inv. koblinger har samme B.W.]

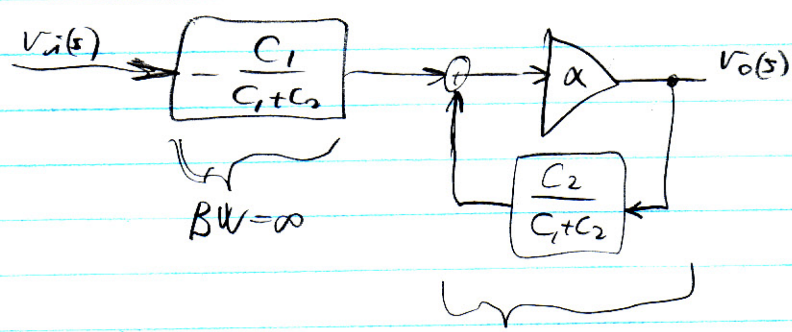
[Inverterende kobling gir lavere gain og GBW produkt i closed loop.]

Ex 3 Same as Ex 2 with R's replaced by C's (important later in switched cap. circuits)



(7)

Exercise show BD of (7) is



$\equiv$ (1) with $a = \alpha$, $f = \frac{C_2}{C_1+C_2} \Rightarrow f = \text{const}\ (\in \mathbb{R})$

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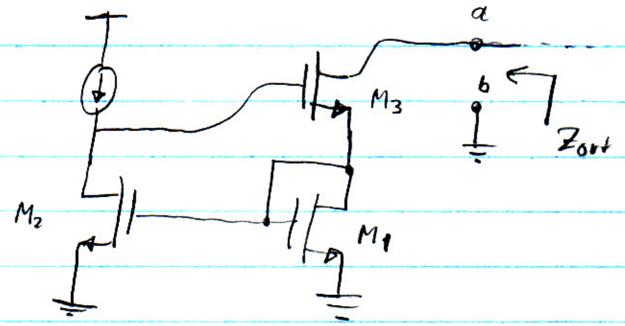
$$\therefore \; (4) \Rightarrow BW = \left(1 + a_0 \frac{C_2}{C_2 + C_1}\right) \cdot |\omega_{pa}|$$

Also $GBW_{\text{closed loop}} = \frac{C_1}{C_1 + C_2} \cdot \omega_t = \frac{C_1}{C_1 + C_2} \cdot a_0 \cdot |\omega_{pa}|$

(V) Impedance transformation

Ex. 4 Wilson current mirror



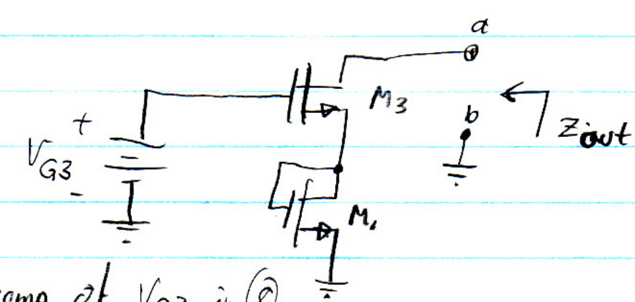
(8)

previously found:

$Z_{out} \approx r_{ds3} \cdot g_{m2} \cdot r_{ds2}$

(low freq. analysis, neglecting body effect)

Now consider



(9)

Where V_{g3} = DC comp. of V_{g3} in (8)

Can show: Z_{out} of (9) $\cong r_{ds3}\left(2 + \frac{1}{g_{m1} \cdot r_{ds3}}\right)$

$$\approx 2 \cdot r_{ds3}$$

Note: (9) is equiv. to (8) with feedback disabled

$\therefore$ feedback increases Z_{out} by $\left(\frac{1}{2} g_{m2} \cdot r_{ds2}\right)$ factor

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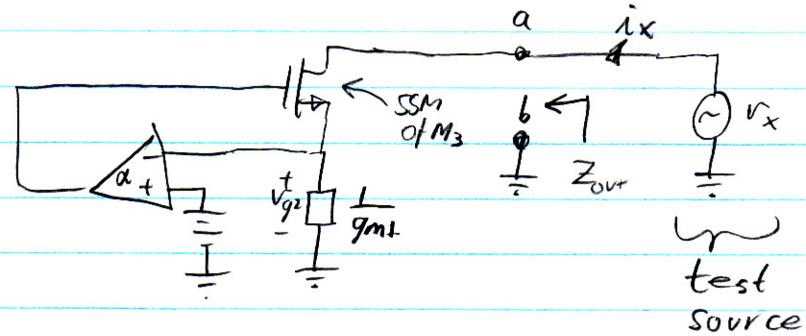
Heuristics:

(8) $\equiv$

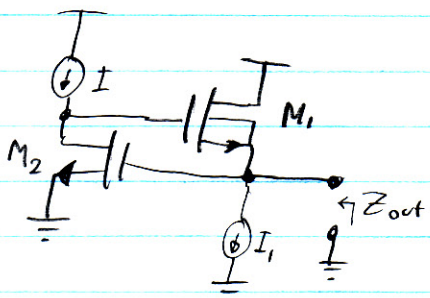


If v_x increases, i_x increases $\Rightarrow v_{g2}$ increases

$\Rightarrow v_{g3}$ decreases

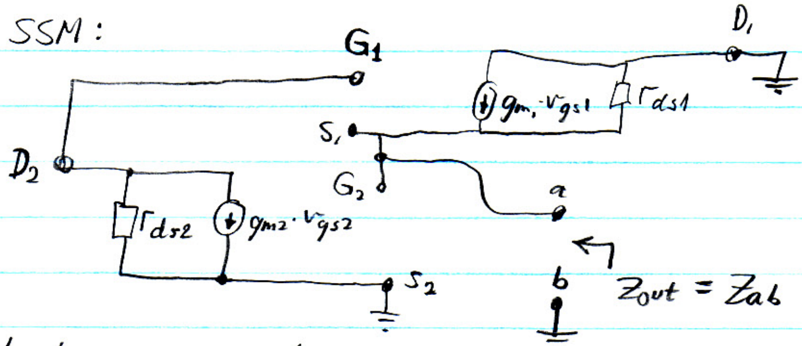
$\Rightarrow$ feedback acts to reduce i_x

Ex 5 Voltage source



(10)

Low freq SSM:



(11)

BIR: Let $g_{m1} \cdot v_{g1} \equiv$ ref. source

$\therefore\ v_{gs1} = X_i,\ \alpha = g_{m1},\ Z_{ab}^{\circ} = r_{ds1}$

Exercise: Show $T_{sc} = 0$, $T_{oc} = g_{m1} \cdot g_{m2} \cdot r_{ds1} \cdot r_{ds2}$

$$\therefore Z_{ab} = r_{ds1} \cdot \frac{1}{1 + g_{m1}\, g_{m2} \cdot r_{ds1} \cdot r_{ds2}} \approx \frac{1}{g_{m1}\, g_{m2}\, r_{ds2}} = \text{small!}$$

Feedback reduced output imp. by a factor of $g_{m2} \cdot r_{ds2}$

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ECE 264A Feb. 19, 2008 E.G.

Stability

Let $A(s) = \dfrac{a_0' + a_1' \cdot s + a_2' s^2 + \ldots + a_m' s^m}{b_0' + b_1' s + b_2' s^2 + \ldots + b_n' \cdot s^n}$ (1)

If $A(0) \neq 0$ or ∞, can write (1) as

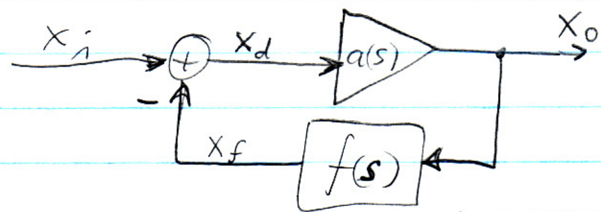
...then:

$A(s) = A_0 \cdot \dfrac{1 + a_1 s + \ldots + a_m \cdot s^m}{1 + b_1 s + b_2 \cdot s^2 + \ldots + b_n \cdot s^n}$ (2)

(other choises of $T(s)$ are possible if we allow $a(s)$ has poles)

$\therefore A(s) = \dfrac{a(s)}{1 + a(s) \cdot f(s)}$ where $a(s) \cdot f(s) = T(s)$
$a(s) = A_0 (1 + a_1 s + \cdots + a_m \cdot s^m)$

$\therefore$ (2) can be impl. as:



Def: A system is stable if (and only if) it satisfies the following property:

For each bounded input signal, $x_i(t)$, the output signal, $x_o(t)$, must also be bounded

($X(t)$ is bounded means $\exists B \in \mathbb{R}$ s.t. $|X(t)| < B \ \forall t$)

"such that"

$\therefore$ "Stability" $\equiv$ "$\underbrace{\text{bounded input, bounded output}}_{\text{"BIBO"}}$ stability"

April 18, 2007

ECE 171A

Bounded input: $R(s) = \dfrac{A_R(s)}{B_R(s)}$ [redacted]

...hvor $B_R(s)$ sine røtter har negativ realdel eller, om de er på den vertikale aksen, multiplisitet maks 1 (ellers får vi terms med rampe-form)

$\Rightarrow$ Stabilt sys. må ha neg. realdel på polene for å garantere bounded output.

2/4

Feb. 19, 2008

<u>Claim 1</u> Let $h(t)$ be the impulse response of an LTI system and let $H(s)$ be its transfer function. Then the system is stable iff either of the following hold:

i) All poles of $H(s)$ are in the LHP (not including imag. axis) (ie. if s_0 is a pole then $Re\{s_0\}<0$)

ii) $\int_{-\infty}^{\infty} h(\tau)\, d\tau < \infty$

iff ≡ "if and only if"

<u>proof</u> Exercise (review material)

<u>Claim 2</u> If an LTI has a transfer function given by 2, then it is stable iff

$$T(s_0) = -1 \Rightarrow Re\{s_0\} < 0$$

proof: Exercise.

Def. "Nyquist plot" ≡ plot of $Im\{T(j\omega)\}$ vs. $Re\{T(j\omega)\}$ as ω increases from $-\infty$ to ∞

<u>Nyquist Criterion</u>:

Provided there are no RHP pole-zero cancellations in $T(s)=a(s)\cdot f(s)$ an LTI system is stable iff the <u>net</u> no. of CCW encirclements of the point $(-1,0)$ by the Nyquist plot equals the no of RHP poles in $T(s)$

$\therefore$ Nyquist crit. $\Rightarrow$ stability test ⟷ {not so useful these days (because of computers)}

will soon see: Nyquist crit. $\Rightarrow$ insight
$\Rightarrow$ concepts of phase & gain margin } very useful

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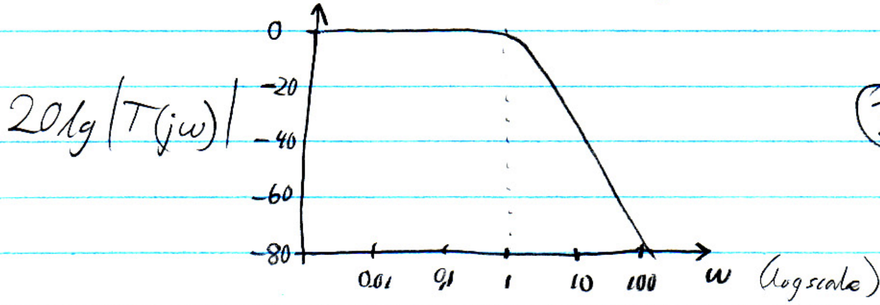
Feb. 19, 2008

Ex1 Suppose $T(s) = \dfrac{1}{(1+s)(1+s/2)}$ poles: -1, -2



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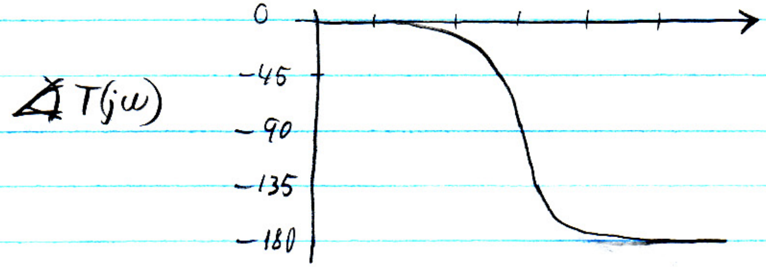
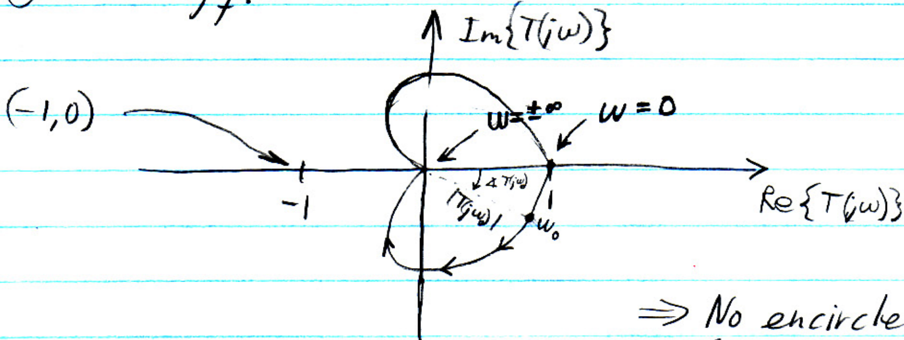


4

3, 4 $\Rightarrow$ Nyq. Plot



$\Rightarrow$ No encirclements of $(-1,0)$
($T(s)$ has no RHP poles by inspection)

Nyq. crit $\Rightarrow$ system with t.f. $A(s) = \dfrac{a(s)}{1+T(s)}$
is stable (provided there are no pole-zero cancellations in $a(s)\,f(s)$) (why?)

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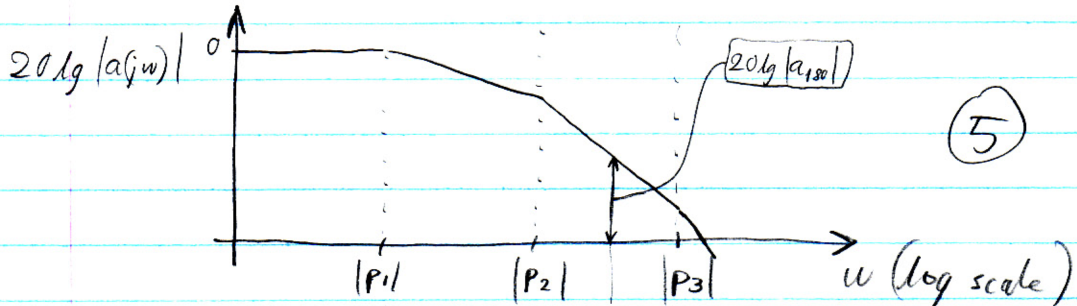
Ex 2 Suppose $a(s) = \dfrac{a_0}{(1 - s/p_1)(1 - s/p_2)(1 - s/p_3)}$ and $f = \text{const.}$

$\therefore T(j\omega) = f \cdot a(j\omega)$

For $|p_1| \approx \frac{1}{10} \cdot |p_2| \approx \frac{1}{100} |p_3|$ and $\mathrm{Re}\{p_i\} < 0 \quad i = 1, 2, 3$



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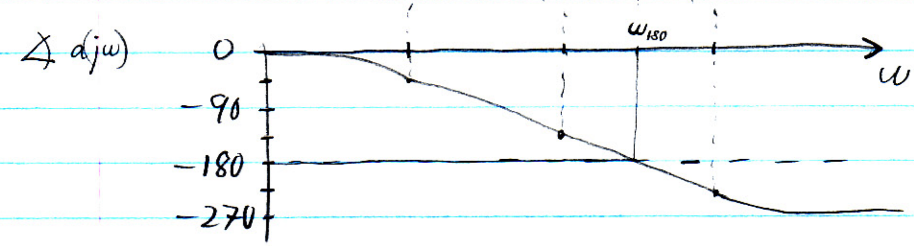

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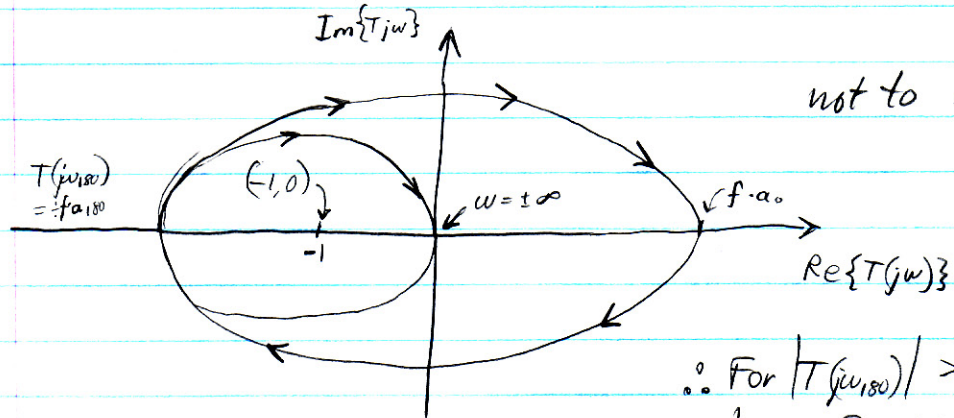

not to scale

$\therefore$ For $|T(j\omega_{180})| > 1$ (the case shown) have 2 c.w. encirclements, but $T(s)$ has no RHP poles

$\therefore$ Nyquist crit $\Rightarrow$ unstable if $|T(j\omega_{180})| > 1$

$\Rightarrow$ stable if $|T(j\omega_{180})| < 1$

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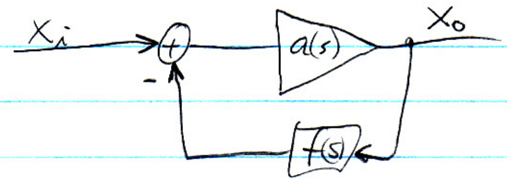
Nyquist Criterion (cont.)

Let $A(s) = \dfrac{a(s)}{1 + a(s)\cdot f(s)}$ $\Leftrightarrow$



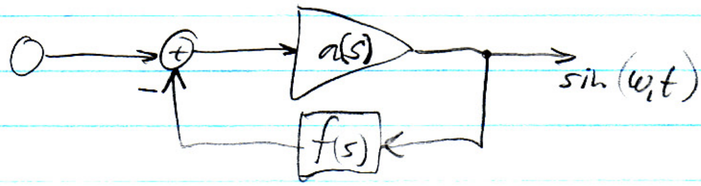
$T(s) = a(s)\cdot f(s)$

Assume: No zero of $f(s)$ is a pole of $a(s)$
Then $T(s_0) = -1 \Leftrightarrow s_0$ is a pole of $A(s)$

<u>Ex. 1</u> Suppose $T(j\omega_1) = -1$ for some $\omega_1 \in \mathbb{R}$
$\Rightarrow A(s)$ has pole on imag. axis $\Rightarrow$ oscillation
$\Rightarrow$ unstable

s.t. = "such that"

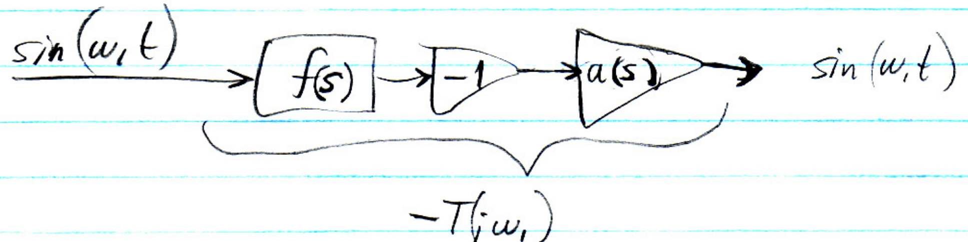
$\therefore$ can choose initial cond. s.t.



(1)

Occurs.

This happens because (1) is equiv. to



$T(j\omega_1) = -1 \Rightarrow |T(j\omega_1)| = 1$

<u>and</u>

$\angle T(j\omega_1) = \pi$

Feb. 21, 2008 E.G.

Ex 2 Suppose $T(s_0) = -1$ for $s_0 = \sigma_0 + jw$, $\sigma_0 \geq 0$

$\Rightarrow$ have (1) with $\sin(\omega_1 t)$ replaced by $\underbrace{e^{\sigma_0 t}}_{\to \infty \text{ as } t \to \infty} \cdot e^{j\omega_0 t}$

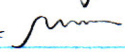
$\Rightarrow$ 0 input $\Rightarrow$ unbounded output

Ex3 Suppose $\underbrace{A(s) \text{ is stable}}_{\text{slang}}$, but for some $\omega_1 \in \mathbb{R}$,

$$|T(j\omega_1)| = 1 - |\varepsilon| \text{ where } |\varepsilon| = \text{small \#}$$

and $\measuredangle T(j\omega_1) = \pi$

Then have ringing for any input step

e.g.: $X_i =$  $X_o =$ 

In many systems, ringing is undesirable because it slows the settling time.

$\Rightarrow$ We need to design with "gain margin" s.t. $|T(j\omega)| \neq 1$ when $\measuredangle T(j\omega) = \pi$ with sufficient "margin" to minimize ringing.

Nyq. crit. = method of evaluating "marginal stability" using simulated or calculated freq. response plots

Recall Nyq. Plot $\equiv$ Plot of $Im\{T(j\omega)\}$ vs. $Re\{T(j\omega)\}$ as ω increases from $-\infty$ to ∞

Nyq. Crit $\equiv$ An LTI system is stable iff the <u>net</u> number of CCW encirclements of $(-1, 0)$ by the Nyq. plot equals the number of RHP poles of $T(s)$.

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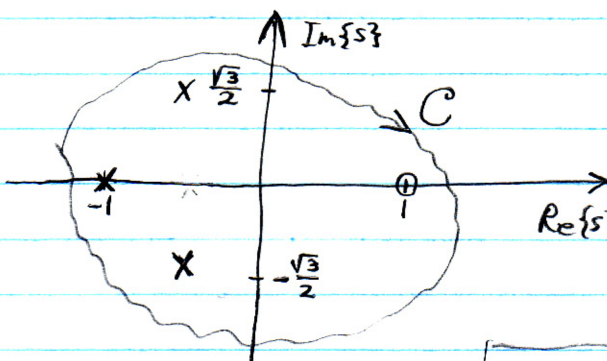
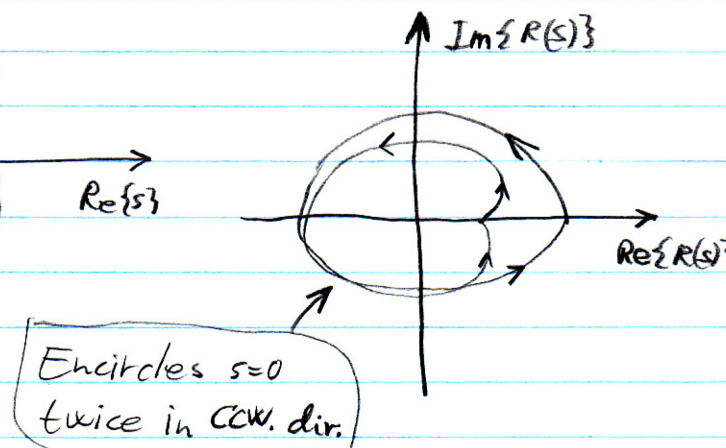
The Nyquist crit. is based on The Encirclement Property (E.P.).

E.P.: Let $R(s)$ be any rational fn. The plot of $R(s)$ along a closed CW path, C, in the complex s-plane encircles the point $s=0$ in a clockwise direction a net no. of times equal to the no. of zeros $\div$ no. of poles within the contour.

<u>Ex. 4</u> Suppose $R(s) = \frac{s-1}{(s+1)(s^2+s+1)}$ poles: $-1, -\frac{1}{2} \pm \frac{\sqrt{3}}{2} j$

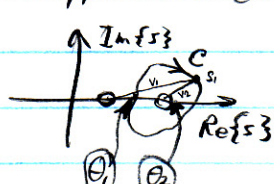
zero: 1





Why does this work?

<u>Ex. 5</u> Suppose $R(s)$ has only 2 (RHP) zeros and no poles



$(V_1, V_2 \in \mathbb{R}, >0)$

$\therefore R(s_1) = V_1 V_2 e^{j(\theta_1+\theta_2)}$

As C is traversed: $\begin{cases} \text{net change in } \theta_1 \text{ is } 0 \\ \text{" —— " } \theta_2 \text{ is } -2\pi \end{cases}$

$\therefore$ (2) $\Rightarrow \measuredangle R(s) = \theta_1 + \theta_2$

$\therefore$ As C is traversed, the net change in $\measuredangle R(s)$ is -2π

$\Rightarrow R(s)\big|_{s \in C}$ circles origin $\underset{(s=0)}{}$ once (in cw dir.)

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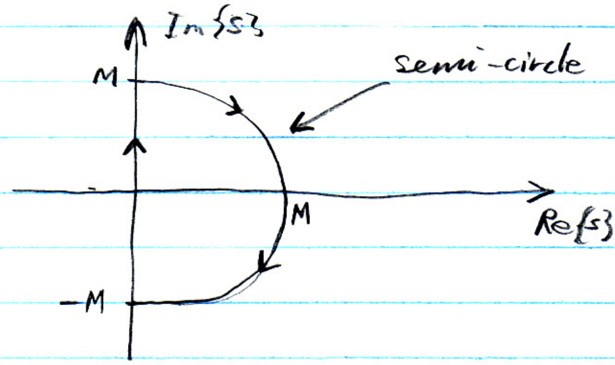
Proof of Nyq. crit.

Restrictions of proof:
1) Rational $T(s)$
2) $|T(j\omega)| < \infty \;\; \forall\, \omega \in \mathbb{R}$
3) $T(s)$ has (# poles) $\geq$ (# of zeros)

(But Nyq. crit also applies without restrictions 1) & 3) and can be modified to handle poles on the imag axis.)

Consider contour C as follows:



As $M \to \infty$, C contains all RHP poles of $A(s)$

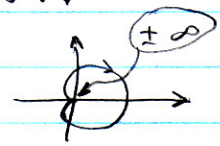
1), 3) $\Rightarrow T(s) = \dfrac{c_0 + c_1 \cdot s + \ldots + c_{N_1} \cdot s^{N_1}}{d_0 + d_1 \cdot s + \ldots + d_{N_2} \cdot s^{N_2}}$ Where $N_1 \leq N_2$

$$\therefore \lim_{|s| \to \infty} T(s) = \begin{cases} \dfrac{c_{N_1}}{d_{N_1}} & \text{if } N_1 = N_2 \\ 0 & \text{if } N_1 < N_2 \end{cases} = \text{const } (\in \mathbb{R})$$

$\Rightarrow$ The plot of $T(s)$ along the semi-circular part of C approaches a single point on real axis as $M \to \infty$

$\Rightarrow$ Nyq. plot is equiv. to plot of $T(s)$ along C as $M \to \infty$

e.g., recall: Nyq. plot of $T(s) = \frac{1}{(1+s)(1+s/2)}$ is



$\Rightarrow$ Can apply encirclement property to Nyq. plot

Let $R(s) = 1 + T(s)$. Then plot of $R(s)$ along C encircles the origin exactly as many times as that of $T(s)$ encircles $(-1, 0)$.

$\therefore$ E.P. $\Rightarrow$ net no. of cw. encirclements of $(-1, 0)$ = # of zeros of $(1+T(s))$ minus # of poles of $1+T(s)$ inside C.

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But:
- zeros of $1+T(s) \equiv$ poles of $A(s)$
- poles of $1+T(s) \equiv$ poles of $T(s)$
- inside C = all RHP as $M \to \infty$

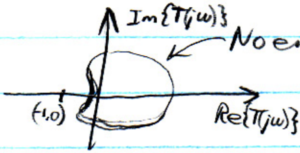
$\Rightarrow$ Nyq. crit.

$\square$

Important special case of Nyquist criterion:

Suppose $\angle T(0) = 0$ and $T(s)$ has no RHP poles
If $|\angle T(j\omega)| = \pi$ for only one pos. value, $\omega = \omega_\pi$,
Then $A(s)$ is stable iff $|T(j\omega_\pi)| < 1$

eg. picture:



Note:
In general, $|T(j\omega_\pi)| > 1 \not\Rightarrow$ instability(!)

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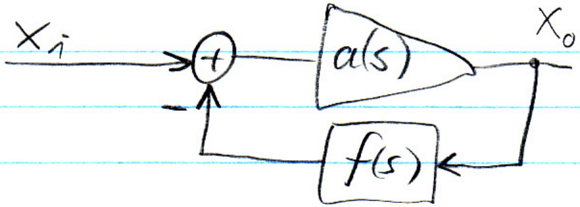
OH: $1 \sim 2^{30}$ this Friday (2/29)

Gain & Phase Margin



$$A(s) = \frac{a(s)}{1 + \underbrace{a(s) \cdot f(s)}_{T(s)}}$$

Assume no pole of $a(s)$ is a zero of $f(s)$.

Nyq. crit $\Rightarrow$ 2 special cases:

i) Suppose $\measuredangle T(0) = 0$, $T(s)$ has no RHP poles and $|\measuredangle T(j\omega)| = \pi$ has no more than one positive solution, $\omega = \omega_\pi$ ①
Then $A(s)$ is stable iff $|T(j\omega_\pi)| < 1$

ii) Suppose $\measuredangle T(0) = 0$, $|T(0)| > 1$, $T(s)$ has no RHP poles, and $|T(j\omega)| = 1$ has no more than one positive solution, $\omega = \omega_u$ ②
Then $A(s)$ is stable iff $-\pi < \measuredangle T(j\omega_u) < \pi$

---

i) $\Rightarrow$ Basis of "Gain margin" (GM) definition (soon)
ii) $\Rightarrow$ Basis of "Phase margin" (PM) definition (soon)

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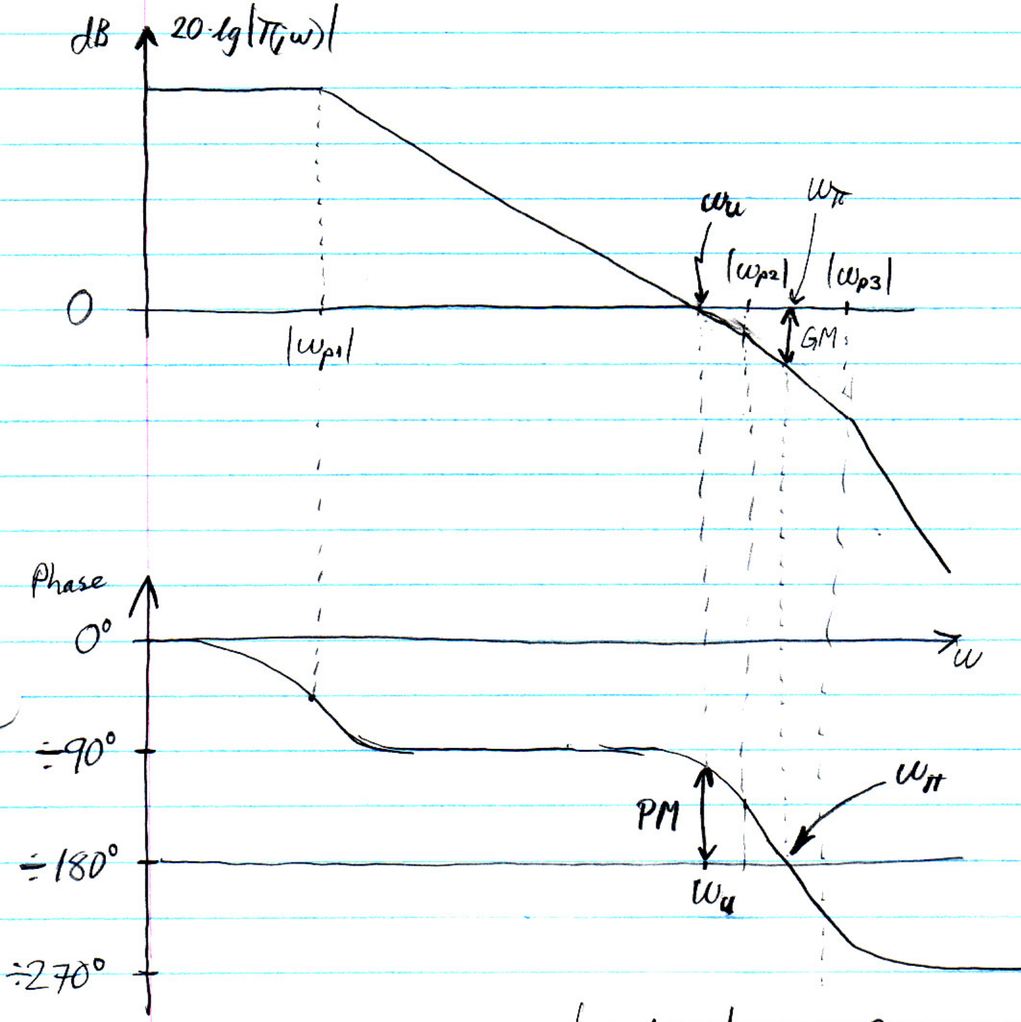
Def. of PM & GM (via 3-pole T(s) example)



i.e. $GM \equiv 20\lg\left|\frac{1}{T(j\omega_\pi)}\right|$ (pos. as shown)

$PM \equiv 180° - |\measuredangle T(j\omega_u)|$ (pos. as shown)

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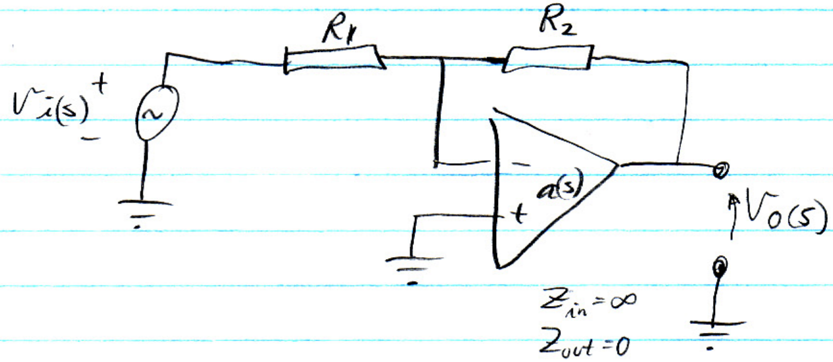
$\therefore$ 1, GM >0 $\Leftrightarrow$ stable (closed loop)

2, PM >0 $\Leftrightarrow$ stable (closed loop)

Also, larger GM (PM) $\Leftrightarrow$ better marginal stability (i.e. step response has less ringing; can tolerate larger component errors without instab.)

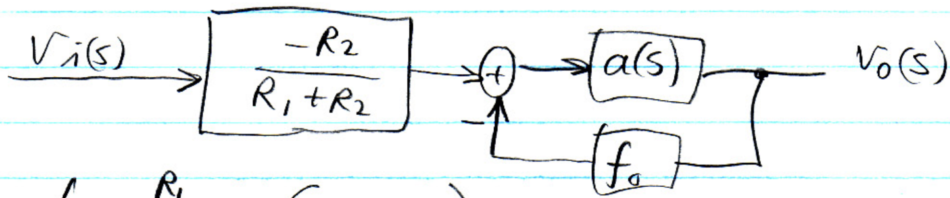
<u>Consider:</u>



3

Previously found:

3 $\equiv$



where $f_0 = \frac{R_1}{R_1+R_2}$ $(0<f_0<1)$

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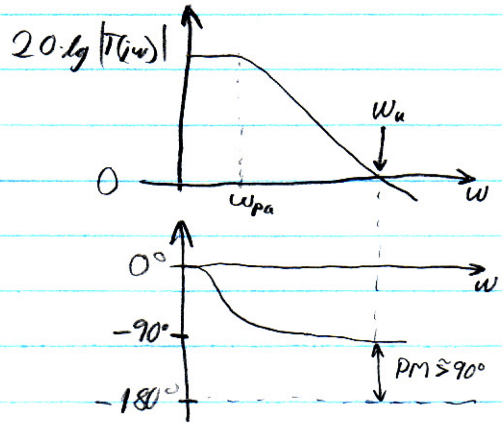
Ex 1 (3) with $a(s) = \dfrac{a_0}{1 - s/\omega_{pa}}$, $\omega_{pa} < 0$ $(\in \mathbb{R})$
$a_0 > 0$ $(\in \mathbb{R})$

$\therefore\ T(j\omega) = \dfrac{a_0 \cdot f_0}{1 - j\,\omega/\omega_{pa}}$



Depending on $a_0 f_0$
$90° < PM < 180°$

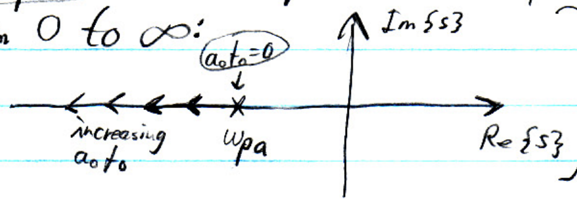
e.g. $\lim_{a_0 f_0 \to 1} PM = 180°$

$\lim_{a_0 f_0 \to \infty} PM = 90°$

Feb. 14

Recall: $A(s) \equiv \dfrac{v_o(s)}{v_i(s)} = A_0 \cdot \dfrac{1}{1 - s/\omega_{pA}}$ where $A_0 \in \mathbb{R}$
$\omega_{pA} = (1 + a_0 f_0) \cdot \omega_{pa}$

Plot of ω_{pA} position in s-plane as $a_0 f_0$ increases from 0 to ∞:

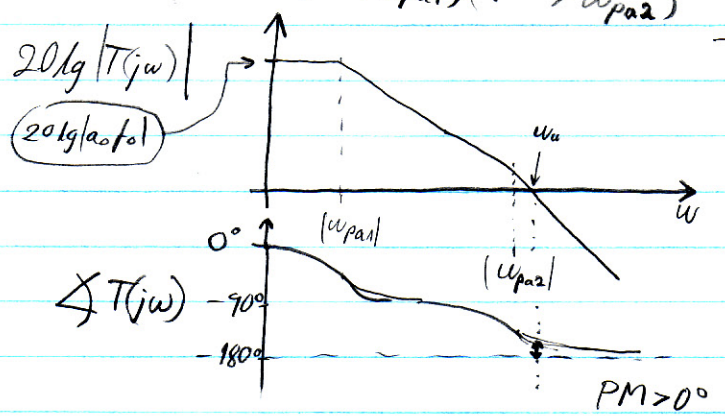


$\equiv$ "root locus" plot

$\therefore$ as expected, it's stable for any choise of $a_0 f_0 > 0$

Ex 2 (3) with $a(s) = \dfrac{a_0}{(1 - s/\omega_{pa1})(1 - s/\omega_{pa2})}$, $\omega_{pai} < 0$ $(\in \mathbb{R})$ $i = 1,2$
$a_0 > 0$ $(\in \mathbb{R})$

e.g.



$\Rightarrow 0° < PM < 180°$

$\Rightarrow$ always stable (at least barely)

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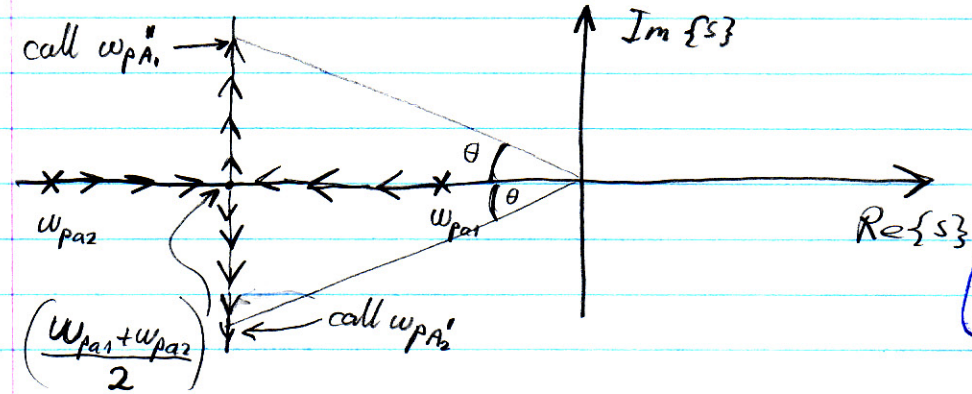
Root Locus plot (verify) (pos. of ω_{pA_i})



$$\frac{a(s)}{1+f\cdot a(s)} = \frac{a_0\cdot\omega_{pa1}\cdot\omega_{pa2}}{s^2-(\omega_{pa1}+\omega_{pa2})\cdot s+(1+a_0\cdot f)\cdot\omega_{pa1}\omega_{pa2}}$$

Recall $\theta = \cos^{-1}\zeta$ where ζ = damping ratio

$\therefore$ $\zeta = \cos\theta$, so

$\zeta \to 0$ as $\theta \to 90°$

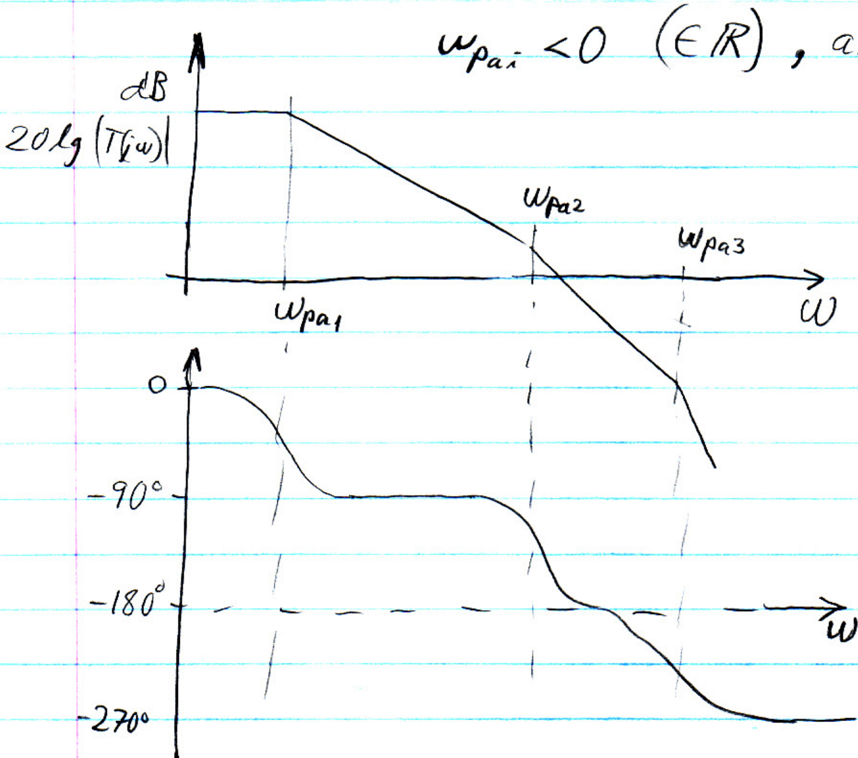
(i.e. as $a_0 f_0 \to \infty$)

(i.e. as PM $\to 0$)

$\therefore$ Smaller PM $\Rightarrow$ more ringing.

Ex. 3 ③ with $a(s) = \dfrac{a_0}{(1-s/\omega_{pa1})(1-s/\omega_{pa2})(1-s/\omega_{pa3})}$

$\omega_{pai} < 0$ $(\in \mathbb{R})$, $a_0 > 0$ $(\in \mathbb{R})$



$\Rightarrow -90° < PM < 180°$

$\Rightarrow$ can be unstable

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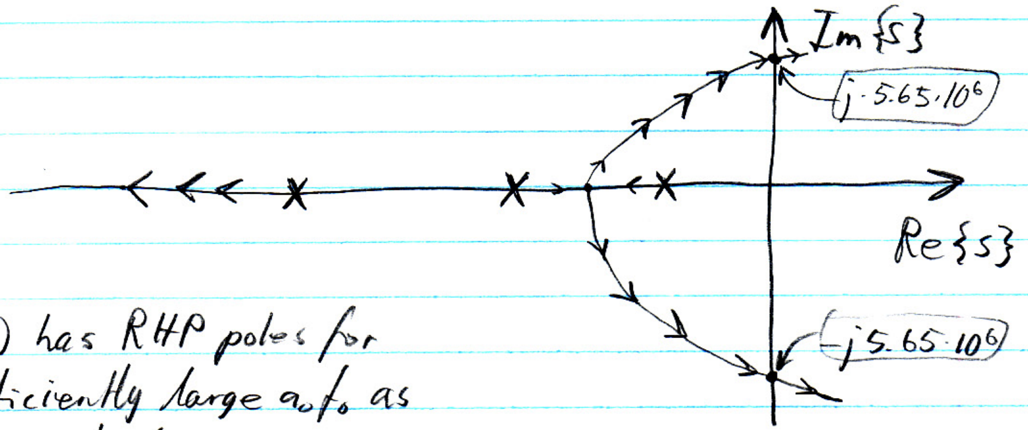
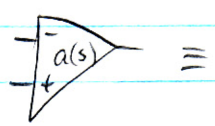
e.g. Suppose $\omega_{pa1} = -10^6 \frac{rad}{sec}$, $\omega_{pa2} = -2 \cdot 10^6 \frac{rad}{sec}$, $\omega_{pa3} = -10^7 \frac{rad}{sec}$
Then the root locus plot is as follows:

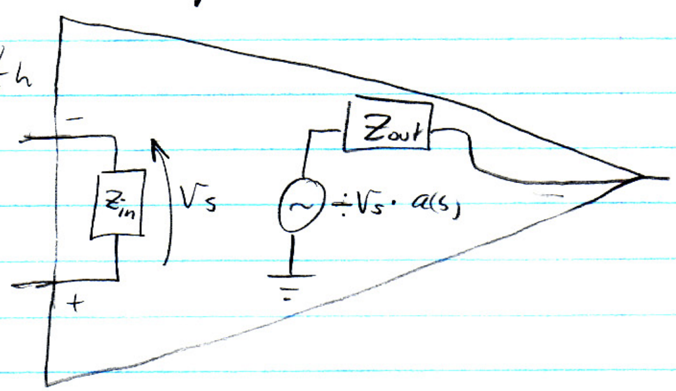


$\therefore$ $A(s)$ has RHP poles for sufficiently large $a_o f_o$ as expected.

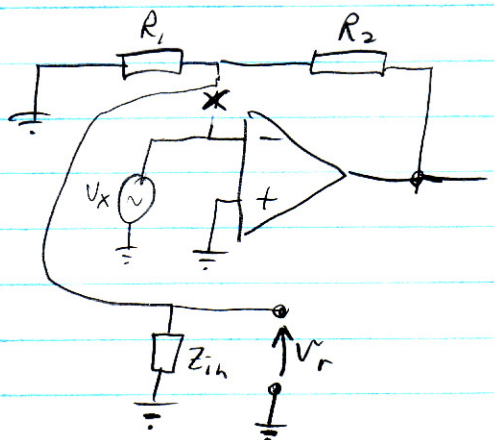
Using SPICE to estimate $T(j\omega)$

Ex 4 Consider ③ with





Suppose we measure $T(j\omega)$ using:



(4)

have:

1) Broken feedback loop
2) Applied test source after break
3) Loaded node prior to break as if it were not broken

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First suppose $Z_{in} = \infty$, $Z_{out} = 0$

Inspection $\Rightarrow$ $\frac{v_r}{v_x} = -a(s) \cdot \frac{R_1}{R_1+R_2} = -a(s) \cdot f_o = -T(s)$

$\therefore T(s) = -\frac{v_r(s)}{v_x(s)}$

$\therefore$ Can use this approach to "measure" $T(j\omega)$
Often works, but must be careful in general.

Suppose $Z_{in} \neq \infty$, $Z_{out} \neq 0$

AGR $\Rightarrow$ $A(s) = A_\infty \cdot \frac{T}{1+T} + A_0 \cdot \frac{1}{1+T}$

Before $A_0 = 0$ because $Z_{out} = 0$
Now, $A_0 \neq 0$

We still get $T(j\omega)$ using 4 but analysis will miss poles contributed by $A_0(s)$

$$N^{(1+T)}_{CW,(0,0)} = Z_{(1+T)} - P_{(1+T)} =$$

$$= P_{cl.l.} - P_T = N^{T}_{CW,(-1,0)}$$

$$N^{T}_{CCW,(-1,0)} = P_T - P_{cl.l.}$$

RHP poles & RHP zeros in general

(map from s to either T(s) or to (1+T(s)).

SEC:

)

$$\boxed{\begin{array}{l}\text{For stability, } P_{cl.l} = 0 \\ \Rightarrow N^{T}_{CCW,(-1,0)} = P_T\end{array}}$$

$$\boxed{N^{T}_{CCW,(-1,0)} = \text{"\# RHP poles, open loop"} \div \text{"\#RHP poles, closed loop"}}$$

(pg. 471 ~ 479 in Dorf & Bishop)

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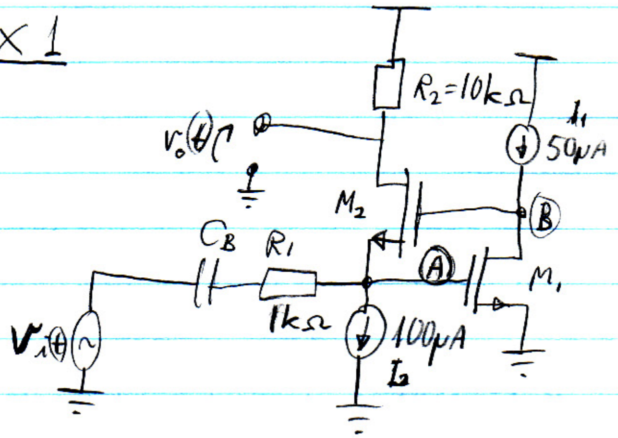
ECE 264A Feb. 28, 2008 E.G.

Using SPICE to estimate $T(j\omega)$

Ex 1



1

$M_1: \frac{W}{L} = 15\mu m/1\mu m$

$M_2: \frac{W}{L} = 20\mu m/1\mu m$

C_B = large DC-blocking cap (assume $C_B \approx \infty$)

Observations

(i) A is a low impedance node:

- inside the loop bandwidth ⇒ A = almost virtual ground (HW)
- even with feedback disabled, impedance of A $\approx 1k\Omega \,\|\, \left(\frac{1}{g_{m2}}\right)$

(ii) Similar reasoning ⇒ driving point resistance between nodes A and B is relatively small ($\approx 1/g_{m1}$)

(ii) ⇒ Reasonable to neglect C_{gd1}, C_{gs2} 2

Apply A.G.R. with $g_{m1} \cdot v_{gs1} \equiv$ ref. source

$$\therefore \quad A(s) = A_\infty(s) \cdot \frac{T(s)}{1+T(s)} + \underbrace{A_o(s)}_{} \cdot \frac{1}{1+T(s)}$$

→ gain of C.G. amplifier formed by M_2 (but with high impedance gate connection)

$\frac{1}{g_{m2}} \approx 1k\Omega$, so $A_o(0) \approx 5$

The point is: $T(s)$ does not give the whole stability "picture" because $A_o(s)$ might have marginal stability.

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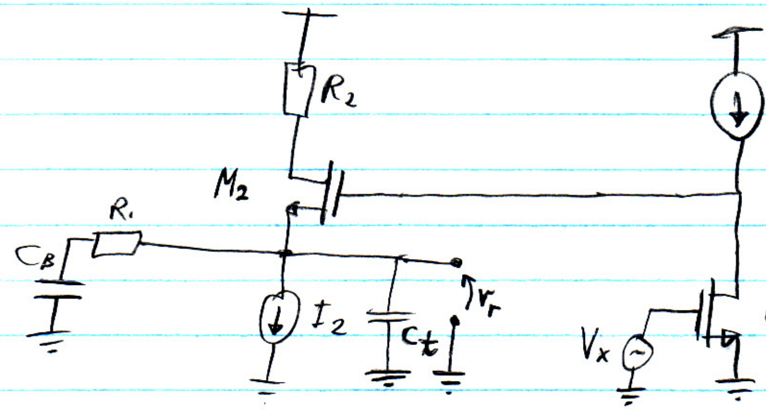
Good approach 1) Measure $T(s)$ to estimate the degree of stab.
2) If simulations show more ringing than expected, consider $A_o(s)$

② ⇒ Can "measure" $T(j\omega)$ using:



③

where
$C_t \equiv C_{gs1}$ (termination impedance),
DC-value of V_x = DC-value of node Ⓐ in ①

⇒
have "broken" feedback loop and applied test signal
③ ⇒ $T(j\omega) = -\frac{v_r(j\omega)}{v_x(j\omega)}$

(same $T(j\omega)$ as in A.G.R. if ② holds - verify)

⇒
simple test to find PM, GM, or Nyquist plot
simulations ⇒ PM ≈ 90°

In general, this allways works, but must:
1) Bias input node following break point as in the closed loop circuit.
2) Must terminate the node (prior to breakpoint) as in the closed loop circuit.

HW6 ⇒ Method of avoiding need to terminate output node.

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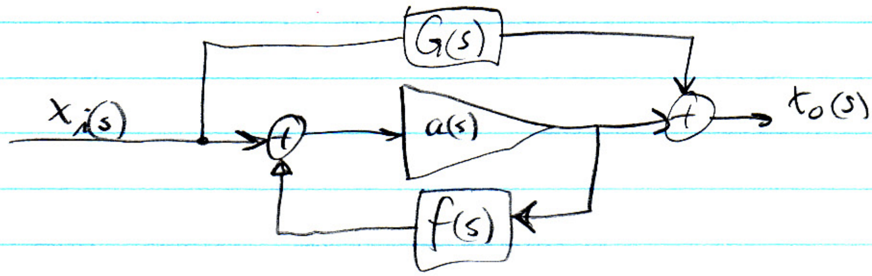
Theory behind breaking feedback loops

Ex



④

$$H(s) \equiv \frac{x_o(s)}{x_i(s)} = \underbrace{\frac{a(s)}{1+a(s)f(s)}}_{\text{call } A(s)} + G(s)$$

Both $G(s)$ and $A(s)$ { • have poles (in general) • affect stability }

In circuits, often { • $A(s) \Rightarrow$ desired behavior (dominant) • $G(s) \Rightarrow$ parasitic signal-path (non-dominant) }

$\therefore$ Feedback loop in ④ $\Rightarrow A(s)$

Knowing $A(s) \not\Rightarrow$ insight into how $a(s)$ or $f(s)$ can be modified to improve performance.

But

Knowing $T(j\omega) \Longrightarrow$ " ———— || ———— ... ... ———— || ———— "

(via Nyq. crit., etc..)

Can "break" any point of feedback loop in ④ and "measure" $T(j\omega)$ via simulation.

The problem: Usually have a circuit, not a B.D. to analyze. Previously have used A.G.R. for specific circuits to justify measuring $T(j\omega)$ directly from circuit. (i.e. without first conv. to a B.D.)

Does this work in general?

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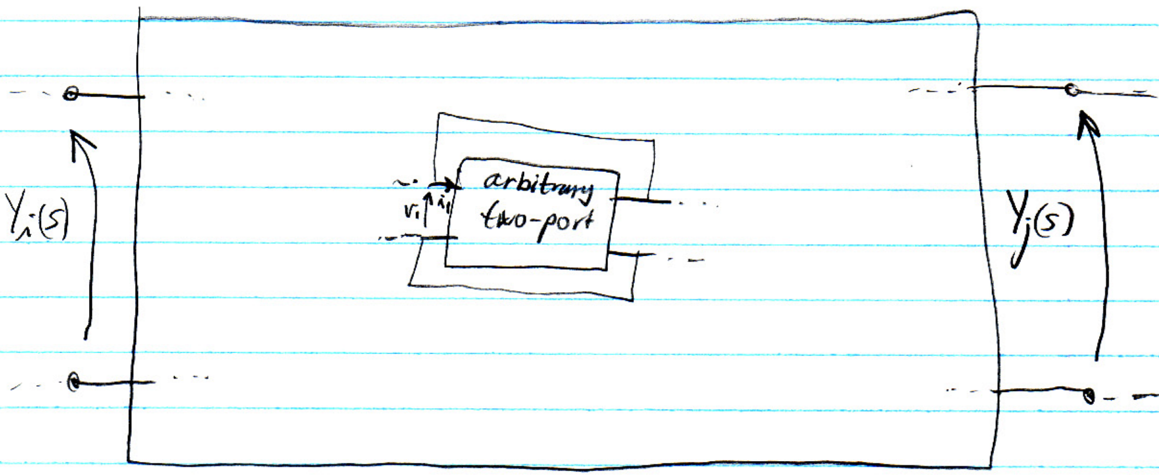
Arbitrary LTI circuit containing a feedback loop



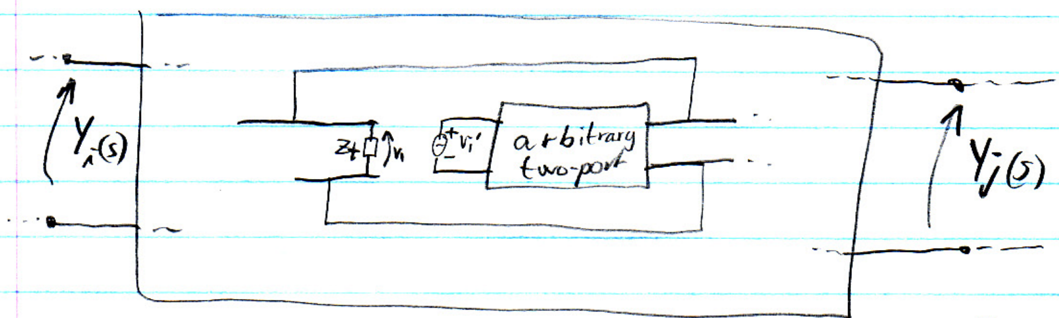
(5)

(Y is voltage, but could be current!)

Let $H(s) = \dfrac{Y_j(s)}{Y_i(s)}$

Note: Can draw (5) s.t. $Y_i(s)$ = current and/or $Y_j(s)$ = current

Can redraw (5) as:



(6)

where $v_1' = \alpha \cdot v_1$ with $\alpha \equiv 1$, $Z_t \equiv \dfrac{v_1(s)}{i_1(s)}$ (input imp. of two-port depends on loading of $Y_i(s)$ & $Y_j(s)$ ports)

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Now, suppose $v_i'(s)$ replaced by independent source, v_x
LTI-system $\Rightarrow$ $v_i = C(s)\cdot Y_i(s) + D(s)\cdot v_x(s)$

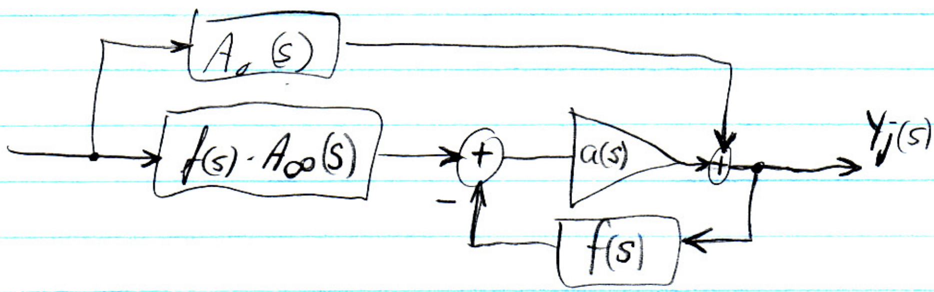
Feedback loop "enabled" $\Leftrightarrow$ $\underbrace{D(s) \not\equiv 0}_{7}$ (eg. $v_i \equiv Y_i$ $\Rightarrow$ 7 violated)

Provided 7 holds, can apply A.G.R.
Recall A.G.R. $\Rightarrow$ $H(s) = A_\infty(s)\cdot\frac{T(s)}{1+T(s)} + A_0(s)\cdot\frac{1}{1+T(s)}$ 8

where $T(s) = -\frac{v_i(s)}{v_x(s)}$ for $Y_i(s)=0$ and $v_i'(s)$ replaced by $v_x(s)$.

$A_0(s) = Y_j(s)/Y_i(s)$ with $x=0$, $A_\infty = \lim_{x\to\infty}\frac{Y_j(s)}{Y_i(s)}$

$\therefore$ Block Diagram of 6:



9

Where $a(s)$, $f(s)$ are any functions s.t. $a(s)\cdot f(s) = T(s)$

Observations

1) $T(j\omega)$ in 8 same as that found by breaking feedback loop in 5, term. the break with $Z_t(s)$, and injecting v_x (or i_x) at input of broken loop
$\Rightarrow$ Method always works provided 7 holds
2) Can use Nyq. crit. (or P.M., G.M.) to assess relative stability of F.B. loop in 8 (If F.B. loop has marginal stab. then 9 (ie 5) "——— " ——— ")

ECE 264A March 04, 2008 E.G.

Note: For the version of the Nyquist Plot and Criterion presented previously, must have at least as many poles as zeros.

(More $\underbrace{\text{zeros}}^{\text{than poles}}$ $\Rightarrow$ inf. B.W. !)

Compensation

The problem: Given an amplifier and feedback network, how do we adjust or add components to achieve a desired stability margin?

Most often $T(s)$ satisfies:

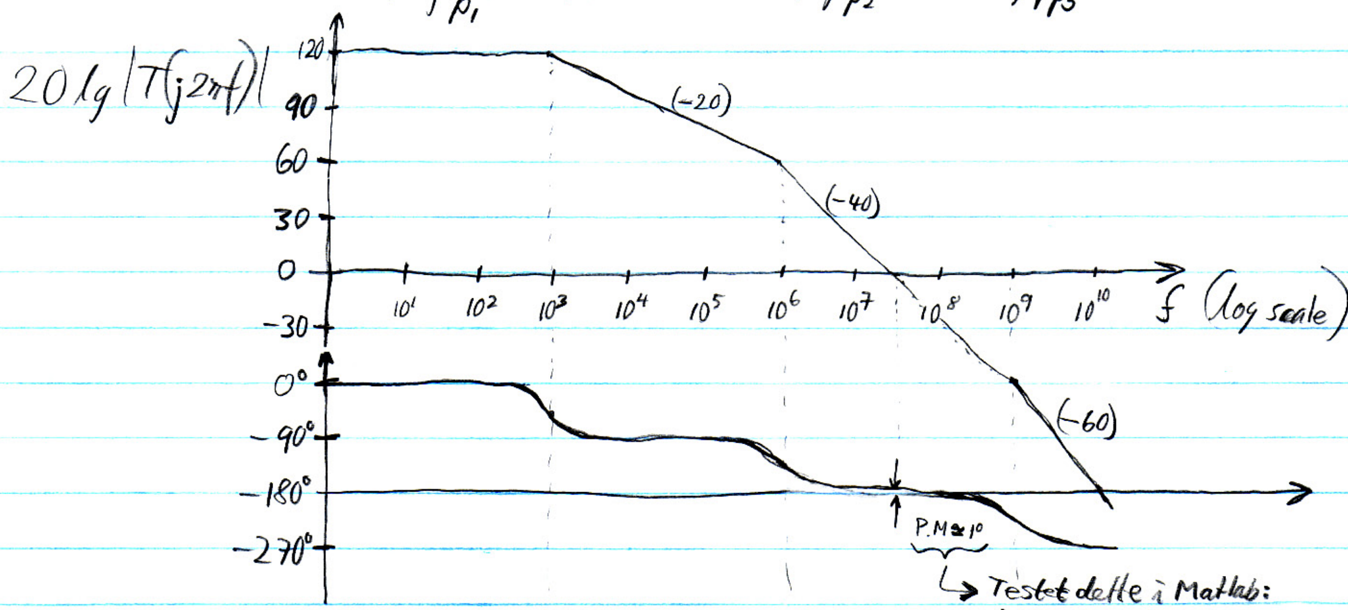
$$\left.\begin{array}{ll} (i) & |\measuredangle T(0)| < \pi \quad \text{(neg. feedback at DC)} \\ (ii) & T(s) \text{ has no RHP (incl. } j\omega \text{ axis) poles} \\ (iii) & |T(j\omega)| = 1 \text{ has } \underline{\text{one}} \text{ pos. solution, } \omega = \omega_u \end{array}\right\} \quad (1)$$

(1) $\Rightarrow$ PM, GM valid indicators of stab. margin.

Ex 1: $$T(s) = T_0 \cdot \frac{1}{(1 - s/\omega_{p1})(1 - s/\omega_{p2})(1 - s/\omega_{p3})} \qquad (2)$$

with

$T_0 = 10^6$, $f_{p_1} = \frac{\omega_{p1}}{2\pi} = -10^3\,Hz$, $f_{p2} = -10^6\,Hz$, $f_{p3} = -10^9\,Hz$



↳ Testet dette i Matlab: Fikk $\underline{0.004°}$

March 04, 2008 E.G.

Observations

i) Both $\measuredangle T(j\omega)$ and $|T(j\omega)|$ decrease **monotonically**

ii) PM $\gtrsim 45°$ requires $|f_{p2}| > f_u$

$\therefore$ Only 3 ways (compensation methods) to increase P.M. given $T(s)$ has the form of (2)

1) Reduce T_0 $\Rightarrow$ ω_u decreases, but $\measuredangle T(j\omega)$ unchanged
$\Rightarrow$ PM increases
But: Trades C.L. accuracy for PM
(feedback benefits requires large T_0)

2) Reduce $|f_{p1}|$
$\Rightarrow$ ω_u decreases, but $\measuredangle T(j\omega)$ almost unchanged for $\gtrsim 4 \cdot |f_{p1}|$ (2 octaves = 0.6 decades)
$\Rightarrow$ P.M. increases
But: Trades C.L. BW for PM

3) Increase $|f_{p2}|$
$\Rightarrow$ ω_u increases until $|f_{p2}| > f_u$, then remains unchanged
but $\measuredangle T(j\omega_u)$ increases
$\Rightarrow$ P.M. increases (up to 90°)
But: Usually not possible without simultaneously reducing $|f_{p1}|$.

Jargon: 1) $\equiv$ "gain compensation"
2) $\equiv$ "dom. pole compensation"
2)+3) $\equiv$ "pole splitting compensation" or "miller compensation"

See 3/5 Jan. 17

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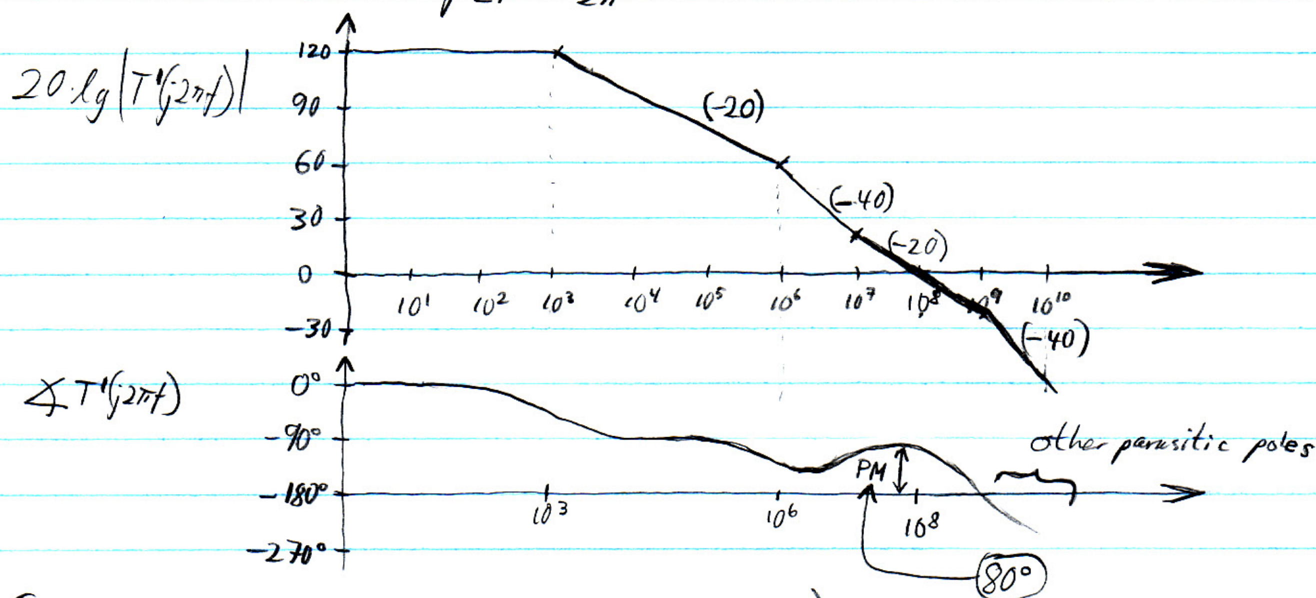
Ex 2 Suppose we can add LHP zero to $T(s)$ given by ②

e.g. let $T'(s) = \underbrace{T(s)}_{\text{same as in Ex 1.}} \cdot (1 - s/\omega_{z1})$

with $f_{z1} = \frac{\omega_{z1}}{2\pi} = -10^7\,Hz$



(zero approx at $1{,}2 \cdot \omega_u$ ← rule of thumb)

Observations

i) LHP zero added pos. phase shift without significantly **changing f_u** (f_u gikk fra $10^{7.5}$ til 10^8 Hz)

⇒ Increased PM (fra ~0° til 80°)

⇒ Adding LHP zero ≡ compensation strategy

ii) A RHP zero would have decreased PM → to −89,4° ⇒ unstable!

called "lead compensation"

Lead compensation is used in 2-stage CMOS op-amps (soon).

Vil nesten ikke endre ω_u men likevel påvirke fasen til en viss grad. Denne tommelfinger-regelen er ikke brukt her. (Her er $|z| < \omega_u$.)

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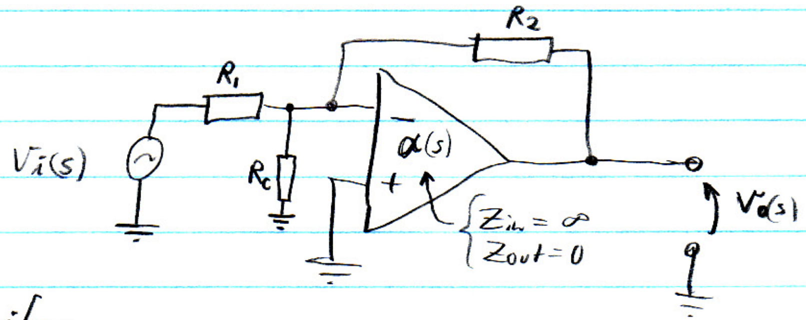
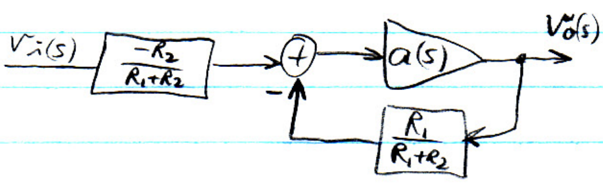
Ex 3 Gain compensation



③

Can verify:
③ ⇒

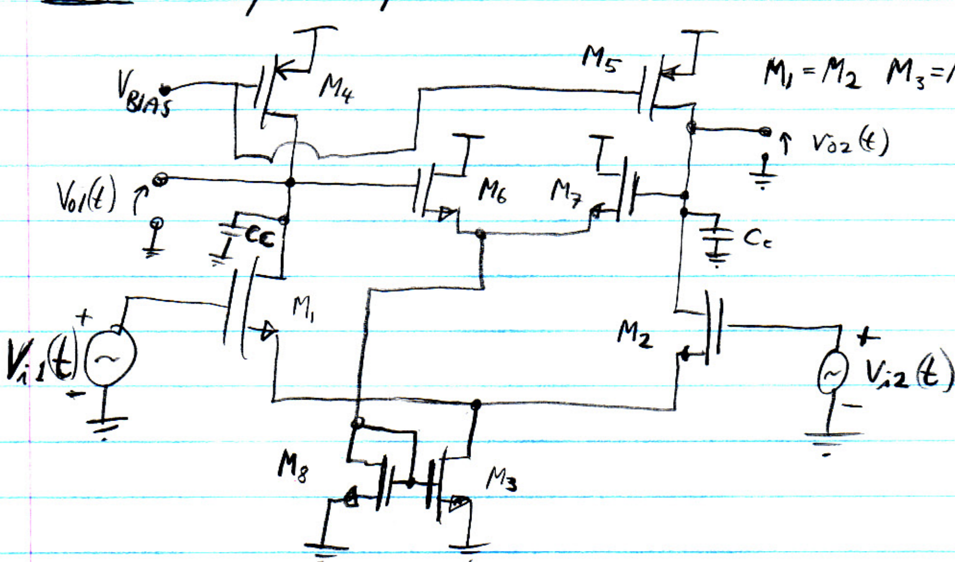


Where
$$a(s) = \frac{R_1 \| R_2 \| R_C}{R_1 \| R_2} \cdot \alpha(s)$$

e.g. Suppose $\alpha(s)$ = 3-pole, no-zero transfer fcn.
⇒ $T(s)$ given by ②
⇒ decreasing R_C ⇒ smaller T_0 but same $\angle T(j\omega)$ as before

e.g. suppose with $R_C = \infty$, $T(s)$ = same as Ex 1
Then PM ≅ 4° (very poor relative stability) { Jeg fikk 0.004° PM i Matlab...
e.g. suppose $R_1 = R_2$ and $R_C = \frac{R_1}{1000}$
Then PM ≃ 35°
But went from $T_0 = 120\,dB$ to $T_0 = 66\,dB$

Ex 4 Dom. pole compensation in CMFB



$M_1 = M_2$ $M_3 = M_8$ $M_4 = M_5$
$M_6 = M_7$

Exercise: Consider how C_c implements dom. pole compensation!

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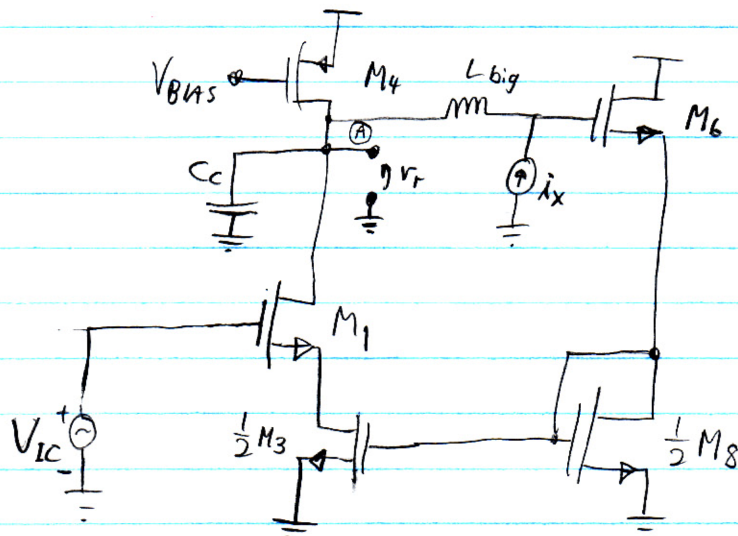
Ex 4 from last time

(recall diff pair with CMFB diagram)

CM ½-circuit & $T(j\omega)$ measurement config



(L_{big} ⇒ zero-freq voltages at two terminals of L_{big} are same, but L_{big} blocks non-zero frequency components)

valid if $C_{gs6} \ll C_c$

Note: trans impedance from node A to all other nodes is small *

$\Rightarrow \div \frac{1}{C_c \cdot R_A}$ = pole of $T(s)$ (C_c has little effect on other poles of $T(s)$)

where R_A = small-sig. res. from node A to small-sig ground

⇒ A = dom. pole node

⇒ can reduce $|f_{p1}|$ (only) by increasing C_c

Strøm injisert i node A vil i liten grad påvirke spenningene på andre noder i kretsen. $\frac{V_{andre}}{I_A}$ = liten = $Z_{trans.}$

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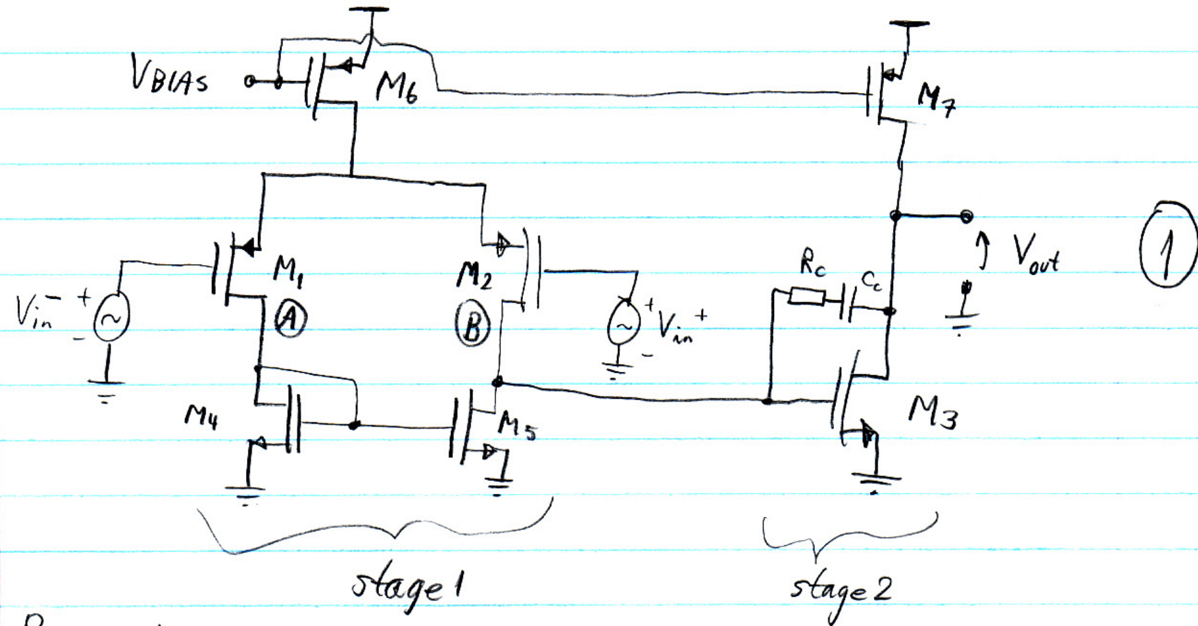
Ex Two-stage op-amp with lead compensation



Remarks:

1) Typ. designed s.t. pole assoc. with A

i.e. $p_A = \frac{-g_{m4}}{C_A}$ ← cap from A to small-sig. ground

is s.t. $|p_A| \gg \omega_u$ ← unity-gain freq.

2) A source-follower buffer stage is sometimes added to 1

3) Could replace pMOSs by nMOSs and vice versa, but pMOSs have better 1/f noise performance so 1 as shown has better 1/f noise (why?)

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First consider with $R_c = 0$

Using prev. results and Remark 1):

(1) has 2 significant poles, p_1 & p_2 and 1 significant zero, z_1 given by:

$$p_1 \cong \div \frac{1}{g_{m3} R_1 R_2 C_c} \quad (2) \quad \text{where} \quad R_1 \equiv r_{ds2} \| r_{ds5}, \quad R_2 \equiv r_{ds3} \| r_{ds7}$$

and $C_c \gg C_{gd}$ is assumed

$$p_2 \cong \div \frac{g_{m3} C_c}{C_{d3} C'_{gs3} + C_c (C_{d3} + C'_{gs3})} \quad (3)$$

where C_{d3} = total cap. from output to ground from M_3, M_7, and any load.

$$\cong \div \frac{g_{m3}}{C_{d3}} \quad \text{(often)}$$

$$z_1 \cong \frac{g_{m3}}{C_c} \quad (4)$$

p_1 = dom. pole (sets open loop BW)

p_2 = non-dom pole (with p_1 sets open-loop phase shift at unity gain freq.)

z_1 = RHP-zero (adds neg. phase shift - just like a LHP pole)

eg. $g_{m3} = 1 \cdot 10^{-3}\ \Omega^{-1}$, $R_1 = R_2 = 100\text{k}\Omega$, $A_{V01} = 70$ (DC-gain of stage 1)

$C_{d3} = C'_{gs3} = 350\text{fF}$, $C_c = 1\text{pF}$ } (5)

(2) - (5) $\Rightarrow$

$$p_1 = -100 \cdot 10^3 \tfrac{rad}{sec} \quad (-16\text{kHz})$$
$$p_2 = -1.2 \cdot 10^9 \tfrac{rad}{sec} \quad (-191\text{MHz})$$
$$z_1 = 10^9 \tfrac{rad}{sec} \quad (159\text{MHz})$$

A_{V02} ($\equiv$ DC-gain of stage 2) $= -g_{m3} \cdot R_2 = -100$

$$\therefore \quad A_v(s) \cong 7000 \cdot \frac{(1 - s/10^9)}{(1 + s/10^5)(1 + s/1.2\cdot 10^9)}$$

$\therefore \quad \omega_u \cong 7.1 \cdot 10^8 \frac{rad}{sec}$ (113 MHz) (verify)

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$$\measuredangle A_v(j\omega_u) = \tan^{-1}\left(-\frac{7.1\cdot10^8}{10^9}\right) - \tan^{-1}\left(\frac{7.1\cdot10^8}{10^5}\right) - \tan^{-1}\left(\frac{7.1\cdot10^8}{1.2\cdot10^9}\right)$$

$$\cong \underbrace{-35.4^\circ}_{\text{zero}} \underbrace{-90^\circ}_{\text{dom. pole}} \underbrace{-30.6^\circ}_{\text{non-dom pole}} = -156^\circ$$

If $f(s)=1$, then $T(s) = A_v(s)$ so

$$P.M. = 180^\circ - 156^\circ = \underline{24^\circ} \quad (\text{low } P.M.)$$

__Note:__ 1) Increasing C_c reduces neg. phase shift from p_2 at ω_u but increases ——"—— z_1 at ω_u (because ω_u decreases but so does z_1)

2) Situation is actually worse than calculated because of pole at A

__Heuristics:__ As ω increases, feed fwd. path through C_c begins to dominate the gain path through M_3. Gain path through M_3 has negative polarity, but feed-fwd. path through C_c has phase $\to 0$ as $\omega \to \infty$ $\Rightarrow$ As $\omega \to \infty$, have pos. feedback

__Now consider with $R_c > 0$__

Can show, $R_c \neq 0 \Rightarrow$ have $z_1 \cong \dfrac{1}{C_c(1/g_{m3} - R_c)}$ (6)

have a 3rd pole, p_3

Usually:

1) $|p_3| \gg |p_1|, |p_2|, |z_1|$, so can ignore p_3
2) p_1, p_2 are hardly affected by R_c

} (7)

(6) $\Rightarrow R_c > \frac{1}{g_{m3}} \Rightarrow z_1$ "moves" to LHP

$\Rightarrow z_1$ contributes __pos__ phase shift

Fact: (7) breaks down for very large R_c.

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Typical rule of thumb: Choose R_c s.t. $|z_1| \cong 1.2 \cdot \omega_u$ (8)

(6), (8) $\Rightarrow$ $1.2 \cdot \omega_u \cong \left| \frac{1}{C_c(1/g_{m3} - R_c)} \right|$

$$R_c = \frac{1}{g_{m3}} + \frac{1}{1.2 \cdot \omega_u \cdot C_c} \quad (9) \quad \left(\text{for } R_c > \frac{1}{g_{m3}}\right)$$

e.g. (using (5)) (9) $\Rightarrow$ $R_c = 2.17\, k\Omega$

Now $\measuredangle A_v(j\omega) = \tan^{-1}\left(\frac{\omega_u}{1.2 \cdot \omega_u}\right) - 90° - 30.6° \cong -80.8°$

$\therefore$ new P.M. for $f(s) = 1$: P.M. $\cong 99.2°$

$\Rightarrow$ Can afford to reduce C_c ($\Rightarrow$ increases BW) to reduce P.M. (typ. want P.M. > 65°)

<u>Closed loop system:</u>

- critically damped if P.M. = 76°
- gain peaking if P.M. < 66°

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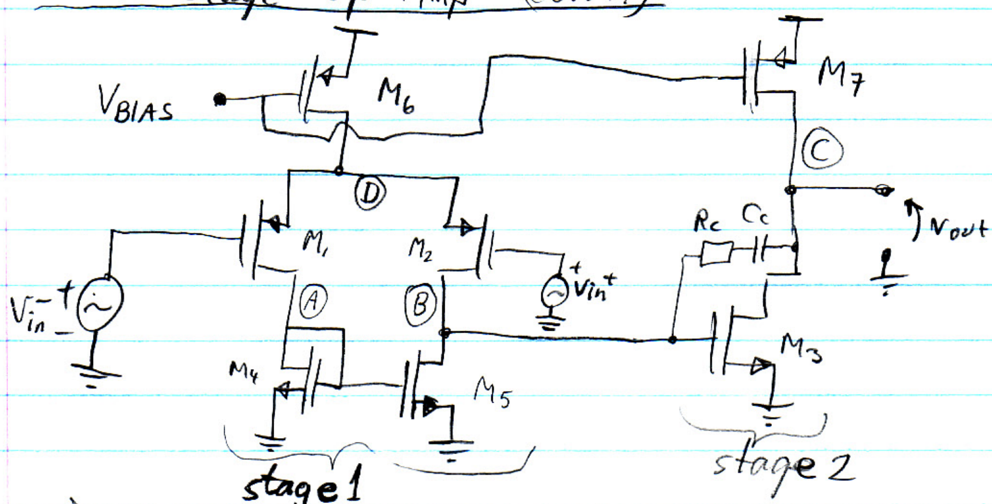
Final exam : WLH 2204 (Thurs, March 20)
Extra Office Hours on Wed. of next week

Two-stage Op-Amp (cont.)



(1)

1) Common mode rejection of stage 1 $\Rightarrow$ usually can neglect **pole** associated with node **D** when considering diff. mode of op-amp.

2) Usually designed s.t. pole associated with A i.e. $p_A \cong -\frac{g_{m4}}{C_A}$ is s.t. $|p_A| \gg \omega_u$ (2)

Then $M_2, M_5, M_3, M_7, R_c, C_c$ determine P.M. when (1) is used in feedback system.

3) Prev. found :

$$A_v(j\omega) \equiv \frac{V_{out}(j\omega)}{V_{in}^+(j\omega) - V_{in}^-(j\omega)} \cong \underbrace{g_{m1} \cdot R_1}_{\text{DC-gain of stage 1}} \cdot \underbrace{g_{m3} \cdot R_2}_{\text{DC-gain of stage 2}} \frac{(1 - j\,\omega/z_1)}{(1 - j\omega/p_1)(1 - j\omega/p_2)} \quad (3)$$

where $R_1 \equiv r_{ds2} \| r_{ds5}$, $R_2 \equiv r_{ds3} \| r_{ds7}$

$$p_1 \cong -\frac{1}{g_{m3} R_1 R_2 C_c} \quad \text{(dom. pole)} \quad (4)$$

$$p_2 \cong \frac{-g_{m3}}{C_{d3} + C_{gs3}} \quad \text{(first non-dom. pole)} \quad (5)$$

and

$$z_1 \cong \frac{1}{C_c(1/g_{m3} - R_c)} \quad \text{(dom. zero)} \quad (6)$$

Note: (4)–(6) hold for typ. design parameters

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general: zero is placed near ω_u

4) $p_1 \Rightarrow$ 3dB BW and introduces -90° phase at ω_u
$p_2, z_1 \Rightarrow$ PM $\underbrace{\text{(i.e. P.M. varies with } p_2, z_1\text{; not with } p_1\text{)}}_{\text{assumes } |p_2|, |z_1| \gtrsim 10\cdot|p_1|}$

5) Rc chosen s.t. z_1 in LHP $\Rightarrow$ introduces pos. phase shift at ω_u

Jargon: PM of an open-loop op-amp $\equiv 180^\circ + \angle A_v(j\omega_u)$
why? When connected as voltage follower, i.e. $f=1$, (hardest non-attenuating config to compensate) loop gain $\equiv T(j\omega) \equiv A_v(j\omega)$

Usually, ① designed s.t. $\underbrace{|p_2|, |z_1|}_{\text{same order of mag.!}} > \omega_u$

$$\therefore ③ \Rightarrow |A_v(j\omega_u)| \cong \left| g_{m1} g_{m3} R_1 R_2 \cdot \frac{1}{1 - j\,\omega_u/p_1} \right|$$

$$\cong g_{m1} g_{m3} R_1 R_2 \cdot \left| \frac{p_1}{j\omega_u} \right|$$

$$\therefore ④ \Rightarrow \cong \left| \frac{g_{m1}}{C_c \cdot j \cdot \omega_u} \right|$$

$$\therefore |A_v(j\omega_u)| = 1 \Rightarrow \omega_u \cong \frac{g_{m1}}{C_c} \qquad ⑦$$

$$P.M. = 180^\circ - \tan^{-1}\frac{\omega_u}{z_1} + \tan^{-1}\frac{\omega_u}{p_2} - \underbrace{90^\circ}_{\text{contrib. by } p_1} \qquad ⑧$$

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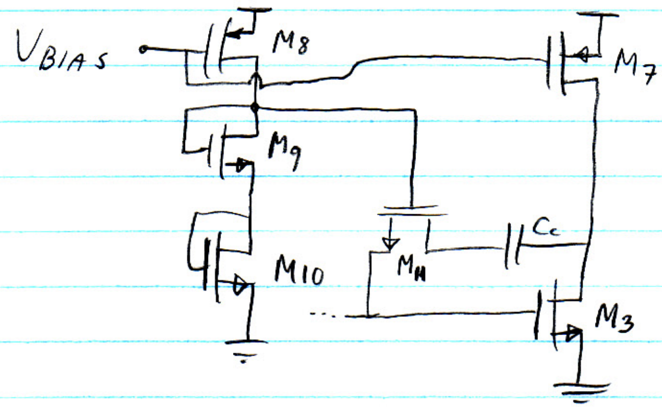
Problem: How to maintain constant P.M. across temp./process/supply voltage variations?

5 - 8 $\Rightarrow$ P.M. depends on $\underbrace{g_{m1}, g_{m3}, C_{d3}, C_{gs3}, C_c, R_c}_{\text{These parameters vary and dont track each other}}$

A Solution: Replace stage 2 of 1 by:



with $\frac{W_7/L_7}{W_3/L_3} = \frac{W_8/L_8}{W_{10}/L_{10}}$ (10)

(9)

M_{11} biased by M_8-M_{10} $\equiv R_C$

No DC-current $\Rightarrow M_{11}$ in triode

$$\Rightarrow R_C \cong \frac{1}{\mu_n C_{ox}(W_{11}/L_{11})\underbrace{(V_{GS_{11}} - V_{T_n})}_{\text{call } V_{eff\,11}}}$$

$$g_{m3} = \mu_n C_{ox}(W_3/L_3) \cdot V_{eff\,3}$$

$$\therefore R_C\, g_{m3} = \frac{(W_3/L_3)\, V_{eff\,3}}{(W_{11}/L_{11})\, V_{eff\,11}} \quad (11)$$

$$\therefore (6) \Rightarrow z_1 = \frac{g_{m3}}{C_c \cdot \left[1 - \left[\frac{(W_3/L_3)\, V_{eff\,3}}{(W_{11}/L_{11})\, V_{eff\,11}}\right]\right]} \quad (12)$$

$$(7), (12) \Rightarrow \frac{\omega_u}{z_1} = \frac{g_{m1}}{g_{m3}} \cdot \left[1 - \left[\frac{(W_3/L_3)\, V_{eff\,3}}{(W_{11}/L_{11})\, V_{eff\,11}}\right]\right]$$

$$(5), (7) \Rightarrow \left|\frac{\omega_u}{p_2}\right| = \frac{g_{m1}}{g_{m3}} \cdot \frac{C_{d3} + C_{gs3}}{C_c}$$

$$\frac{g_{m1}}{g_{m3}} = \frac{\sqrt{2 \cdot \mu_p \cdot C_{ox} \cdot (W_1/L_1) \cdot I_{D1}}}{\sqrt{2 \mu_n \cdot C_{ox} (W_3/L_3) I_{D3}}}$$

$\boxed{\mu_0 \cdot C_{ox} \approx K'}$

Facts:

1) $\mu_p/\mu_n \cong$ const. for a given process
2) $I_{D1}/I_{D3} \cong$ const. because derived from a common bias network

$\Rightarrow \frac{g_{m1}}{g_{m3}} \cong$ const.

3) $\frac{C_{d3} + C_{gs3}}{C_c} \neq$ const., but does not vary much over process & temp. because dominated by oxide capacitor

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$\therefore$ P.M. $\cong$ const. provided $V_{eff3}/V_{eff11} \cong$ const.

10 $\Rightarrow$ $V_{eff10} = V_{eff3}$
$V_{eff9} = V_{eff11}$

$$\therefore \frac{V_{eff3}}{V_{eff11}} = \frac{V_{eff10}}{V_{eff9}} = \frac{\sqrt{\dfrac{2 \cdot I_{D10}}{\mu_n \cdot C_{ox}(W_{10}/L_{10})}}}{\sqrt{\dfrac{2\, I_{D9}}{\mu_n \cdot C_{ox}(W_9/L_9)}}} = \sqrt{\frac{W_{10}/L_{10}}{W_9/L_9}} = \left\{\begin{array}{l}\text{ratio of}\\ \text{like quantities}\end{array}\right.$$

$\therefore$ P.M. $\cong$ indep. of process & temperature

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The "reference source" in Blackman's Impedance Relation (BIR) and the Asymptotic Gain Formula (AGF):

The values of the variables in BIR (Z^0_{ab}, T_{sc}, T_{oc}) or in AGF (A_∞, A_0, T) depend on which reference source we use.

Of course, the resulting impedance (Z_{ab}) or gain (A) is the same, independent of our choise of reference source.

BIR & AGF was covered on Jan. 31 in class.