Chapter 5 - Problems

5.1)

a) Find $\omega_{-3dB} = \frac{1}{R_{out} C_{eq}} = \frac{1}{r_{DS2} // r_{DS4} \times C_L (1 + A_2)}$

where $r_{DS4} = \frac{8000L}{I_{D4}} = \frac{8000 \times 1.2}{0.05} = 192 k\Omega$

$r_{DS2} = \frac{12000L}{I_{D2}} = \frac{12000 \times 1.2}{0.05} = 288 k\Omega$

$\therefore r_{DS4} // r_{DS2} = 115 k\Omega$

$A_2 = g_{m7} (r_{DS6} // r_{DS7})$

$r_{DS6} = \frac{12000 \times 1.2}{0.1} = 144 k\Omega$

$r_{DS7} = \frac{8000 \times 1.2}{0.1} = 96 k\Omega$

$\therefore r_{DS6} // r_{DS7} = 58 k\Omega$

$g_{m7} = \sqrt{2 \mu_n C_{ox} W/L \, I_{DS}} = (2 \times 92 \times 10^{-6} \times 300/1.2 \times 0.1 \times 10^{-3})^{1/2}$

$= 2.15 \, mA/V$

$\therefore A_2 = 2.15 \times 10^{-3} \times 58 \times 10^{3} = 124$

$\therefore \omega_{-3dB} = \frac{1}{115 \times 10^{3} \times 10 \times 10^{-12} \times 125} = \underline{2\pi \times 1.1 kHz}$

b) Find unity-gain frequency, ω_t.

$\omega_t = g_{m1}/C_c$ where $g_{m1} = \sqrt{2 \times 30 \times 10^{-6} \times \frac{300}{1.2} \times 50 \times 10^{-6}}$

$= 0.866 \, mA/V$

$\therefore \omega_t = \frac{0.866 \times 10^{-3}}{10 \times 10^{-12}} = \underline{2\pi \times 14 MHz}$

c) Find slew rate.

$SR = \frac{2 I_{D1}}{C_c} = \frac{2 \times 50 \mu A}{10 pF} = \underline{10 \, V/\mu sec}$

5.2) With $C_c = 4pF$,

$$SR = \frac{2I_{D1}}{C_c} = \frac{2 \times 50\mu A}{4pF} = \underline{25\ V/\mu sec}$$

To double the slew rate while maintaining C_c, we need to double I_{D1} by doubling the width of Q5. In order to prevent this from changing g_{m1} and g_{m2} and hence ω_t, we need to reduce the widths of Q_1 and Q_2 by half.

5.3) Referring to Problem 5.2, if we scale the widths of Q_1 and Q_2 by half, we need to maintain the equality

$$\frac{W/L_7}{W/L_4} = 2\frac{W/L_6}{W/L_5} \qquad (5.28)$$

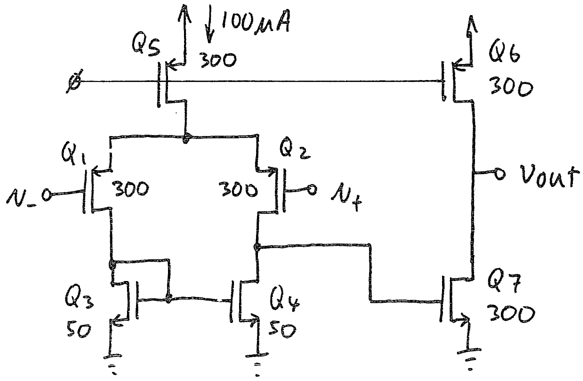
$$\therefore \frac{(W/L)_7}{(W/L)_6} = \frac{2(W/L)_4}{(W/L)_5} = \frac{2(150/1.2)}{(600/1.2)} = \underline{\frac{1}{2}}$$

2 POSSIBILITIES ARE :

1) IF bias current at output stage remains unchanged $\Rightarrow (W/L)_6 = \frac{300}{1.2}$ & $(W/L)_7 = \frac{150}{1.2}$

2) IF bias current at output stage is doubled $\Rightarrow (W/L)_6 = \frac{600}{1.2}$ & $(W/L)_7 = \frac{300}{1.2}$

5.4)



$L = 1.6\,\mu m$

To remove the inherent, systematic offset, the widths of Q_3 and Q_4 should become 150 μm. In that case,

$$V_{eff7} = V_{eff4} = \sqrt{\frac{2 I_{D7}}{\mu_n C_{ox} W/L_7}} = \sqrt{\frac{2 \times 100 \times 10^{-6}}{92 \times 10^{-6} \times 300/1.6}}$$

$$= \underline{0.1077\,V}$$

However, with the current circuit,

$$V_{eff4} = \sqrt{\frac{2 \times 50 \times 10^{-6}}{92 \times 10^{-6} \times 50/1.6}} = 0.1865\,V$$

∴ An input offset voltage will have to be applied in order to decrease V_{gs7} by

$$\Delta V_{gs7} = 0.1865 - 0.1077 = 78.8\,mV$$

For the input-referred offset voltage v_{ioff},

$$v_{ioff} = \Delta V_{gs7} / A_1$$

where A_1 is the first stage's voltage gain

$$A_1 = g_{m1} \times r_{DS2} // r_{DS4}$$

$$g_{m1} = \sqrt{2 \times 30 \times 10^{-6} \times \frac{300}{1.6} \times 50 \times 10^{-6}} = 0.75\,mA/V$$

$$r_{DS4} = 8000 \times 1.6 / 0.05\,mA = 256\,k\Omega$$

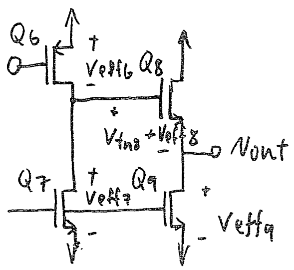
$$r_{DS2} = 12000 \times 1.6 / 0.05\,mA = 384\,k\Omega$$

$$\therefore A_1 = 0.75 \times (256 // 384)$$

$$= 115$$

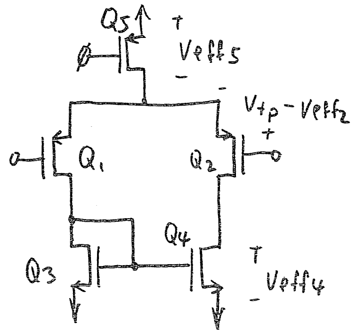
And $\underline{v_{ioff} = \frac{78.8\,mV}{115} = 0.7\,mV}$ (applied to v_+ input).

5.5) Find max. and min Vout and Vin (common mode)



$V_{out\_max} = V_{DD} - V_{eff6} - V_{tn8} - V_{eff8}$

$V_{out\_min} = V_{ss} + V_{eff9}$



$V_{incm\_max} = V_{DD} - V_{eff5} - V_{eff1} + V_{tp1}$

$V_{incm\_min} = V_{ss} + V_{eff3} + V_{tn3} + V_{tp1}$

Calculating all values,

$$V_{eff5} = V_{eff6} = \sqrt{\frac{2 I_{D5}}{\mu_n C_{ox} W/L}} = \sqrt{\frac{2 \times 100 \times 10^{-6}}{30 \times 10^{-6} \times 300/1.6}} = 0.189V$$

similarly,

$V_{eff1} = V_{eff2} = 0.133V$

$V_{eff3} = V_{eff4} = 0.108V$

$V_{eff8} = V_{eff9} = 0.108V$

$\therefore$

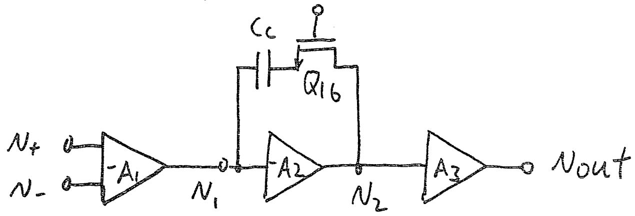
$V_{out\_max} = 5 - 0.189 - 0.8 - 0.108 = 3.9V$

$V_{out\_min} = -5 + 0.108 = -4.9V$

$V_{incm\_max} = 5 - 0.189 - 0.133 - 0.9 = 3.8V$

$V_{incm\_min} = -5 + 0.108 + 0.8 - 0.9 = -4.99V$

5.6) Bad design of compensation network:



In order to understand why oscillations occur on account of this arrangement, we have to keep in mind our use of Q16, operated in the triode region, as a substitute for a feed-forward resistor. When v_{out} is large and positive, v_2 is similarly so. Should the voltage across C_c happen to be small, a large voltage would appear across the drain and source of Q16. If this voltage were large enough to put Q16 into the active mode, capacitor C_c would effectively become disconnected from the output. The result would be an uncompensated op amp that is prone to oscillations.

By keeping Q16 connected to the input of the second stage, most of the large signal swings will appear across the capacitor instead. As such, Q16 will more reliably remain in the triode region of operation.

5.7) The body effect changes the threshold voltages of Q_8, Q_1 and Q_2. The change in V_{tn8} affects only V_{out_max} while the changes in V_{tp1} and V_{tp2} affect both $V_{in\text{-}cm\_max}$ and V_{incm_min}.

From Problem 5.5, $V_{out\_min} = -4.9V$

The other values must be solved iteratively as the threshold voltages are dependent on V_{SB}'s.

For V_{out_max}:

Assume $V_{tn8} = 0.8V$ for first iteration

$$\therefore V_{SB8} = V_{out\_max} - V_{SS} = V_{DD} - V_{eff6} - V_{tn8} - V_{eff8} - V_{SS}$$
$$= 5 - 0.189 - 0.8 - 0.108 - (-5V)$$
$$= 8.9V$$

$$\therefore V_{tn8} = V_{tn0} + \gamma\left(\sqrt{V_{SB8} + 2\phi_F} - \sqrt{2\phi_F}\right)$$
$$= 0.8V + 0.5\left(\sqrt{8.9 + 0.7} - \sqrt{0.7}\right)$$
$$= 1.93V$$

1st iteration

$$\therefore V_{SB8} = 5 - 0.189 - 1.93 - 0.108 - (-5V)$$
$$= 7.77V$$

$$V_{tn8} = 0.8 + 0.5\left(\sqrt{7.77 - 0.7} - \sqrt{0.7}\right)$$
$$V_{tn8} = 1.84V$$

2nd iteration

$$\therefore V_{out\_max} = V_{DD} - V_{eff6} - V_{tn8} - V_{eff8} = 5 - 0.189 - 1.84 - 0.108$$

$\therefore V_{out\_max} = 2.9V$ which is almost 1 volt lower than our original result in P5.5.

For V_{incm_max}:

$$V_{SB1} = V_{eff5} = 0.189V$$

(no iterations required)

$$V_{tp1} = V_{tp0} - \gamma\left(\sqrt{V_{SB1} + 2\phi_F} - \sqrt{2\phi_F}\right)$$
$$= -0.9 - 0.8\left(\sqrt{0.189 + 0.7} - \sqrt{0.7}\right)$$
$$= -0.98V$$

$$\therefore V_{incm\_max} = V_{DD} - V_{eff5} - V_{eff1} + V_{tp1} = 5 - 0.189 - 0.133 - 0.98$$

$V_{incm\_max} = 3.7V$

(cont.)

5.7) (cont.)

For Vincm_min:

Assume $V_{tp1} = -0.9V$

$$V_{BS1} = V_{DD} - (V_{incm\_min} + V_{eff1} - V_{tp1})$$
$$= V_{DD} - (V_{SS} + V_{eff3} + V_{tn3} + \cancel{V_{tp1}} + V_{eff1} - \cancel{V_{tp1}})$$
$$= V_{DD} - V_{SS} - V_{eff3} - V_{tn3} - V_{eff1} = 5-(-5)-0.108-0.8-0.133$$
$$= 8.96V$$

$$\therefore V_{tp1} = V_{tp0} - \gamma\left(\sqrt{V_{SB1} + 2\phi_F} - \sqrt{2\phi_F}\right)$$
$$= -0.9 - 0.8\left(\sqrt{8.96+0.7} - \sqrt{0.7}\right)$$
$$= -2.7\ V$$

$$\therefore V_{incm\_min} = V_{SS} + V_{eff3} + V_{tn3} + V_{tp1} = -5 + 0.108 + 0.8 - 2.7$$
$$= -6.8\ V$$

Note that the min common-mode input voltage is lower than the -5 V supply.

5.8) Show that $1/\omega_{eq} = \sum \frac{1}{\omega_{pi}} - \sum \frac{1}{\omega_{zi}}$.

We are given that

$$\measuredangle[H(j\omega_t)] \approx \measuredangle[H_{app}(j\omega_t)]$$

$$LS = \measuredangle[H(j\omega_t)] \approx \measuredangle\left[\frac{\prod(1+j\omega_t/\omega_{zi})}{\prod(1+j\,\omega_t/\omega_{pi}}\right]$$
$$= \tan^{-1}\left(\frac{\omega_t}{\omega_{z1}}\right) + \tan^{-1}\left(\frac{\omega_t}{\omega_{z2}}\right) + \ldots + \tan^{-1}\left(\frac{\omega_t}{\omega_{zn}}\right) - \tan^{-1}(\omega_t/\omega_{p1})$$
$$- \ldots - \tan^{-1}(\omega_t/\omega_{pn})$$

But the Taylor Series expansion for $\tan^{-1}x = x - \frac{x^3}{3} + \frac{x^5}{5} - \ldots$

$$\approx x \quad \text{for } x \ll 1$$

$\therefore$ If the op amp is compensated such that the unity gain frequency, ω_t, is much lower than all higher order poles and zeros but much higher than ω_{p1}

$$LS \approx \frac{\omega_t}{\omega_{z1}} + \ldots + \frac{\omega_t}{\omega_{zn}} - 90^0 - \frac{\omega_t}{\omega_{p2}} - \ldots - \frac{\omega_t}{\omega_{pn}} = \omega_t\left(\sum_i \frac{1}{\omega_{zi}} - \sum_i \frac{1}{\omega_{pi}}\right) - 90^0$$

Similarly, $RS = \measuredangle[H_{app}(j\omega_t)] = -\tan^{-1}(\omega_t/\omega_{eq}) - 90^0 \cong -\frac{\omega_t}{\omega_{eq}} - 90^0$

$$\therefore \omega_t\left(\sum_i \frac{1}{\omega_{zi}} - \sum_i \frac{1}{\omega_{pi}}\right) \approx -\omega_t/\omega_{eq}$$

or $$\frac{1}{\omega_{eq}} = \sum_i \frac{1}{\omega_{pi}} - \sum_i \frac{1}{\omega_{zi}} \qquad \text{Q.E.D.}$$

5.9) Given:

$$H(s) = \frac{K}{(1+s/\omega_{p1})(1+s/\omega_{p2})(1+s/\omega_{p3})(1+s/\omega_{p4})}$$

where $\omega_{p1} = 2\pi \times 3\,kHz$, $\omega_{p2} = 2\pi \times 130\,MHz$

$\omega_{p3} = 2\pi \times 160\,MHz$, $\omega_{p4} = 2\pi \times 180\,MHz$

$$\angle H(j\omega) = -90^\circ - \tan^{-1}\left(\frac{\omega/2\pi}{130}\right) - \tan^{-1}\left(\frac{\omega/2\pi}{160}\right) - \tan^{-1}\left(\frac{\omega/2\pi}{180}\right)$$

& Given that $\angle H(j\omega_{eq}) = -135^\circ$ then trial and error gives $\omega_{eq} \cong 2\pi \times 41.3\,MHz$

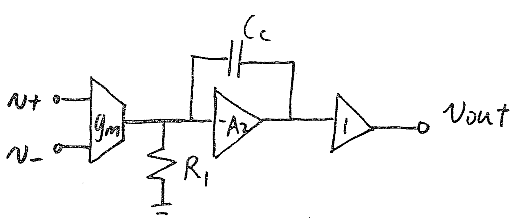
Using Eq. (5.45),

$$\frac{1}{\omega_{eq}} \cong \frac{1}{\omega_{p2}} + \frac{1}{\omega_{p3}} + \frac{1}{\omega_{p4}} = \frac{1}{2\pi}\left(\frac{1}{130\times10^6} + \frac{1}{160\times10^6} + \frac{1}{180\times10^6}\right)$$

$\therefore$ $\omega_{eq} \cong 2\pi \times 51.3\,MHz$

This estimate is about 24% above the true value.

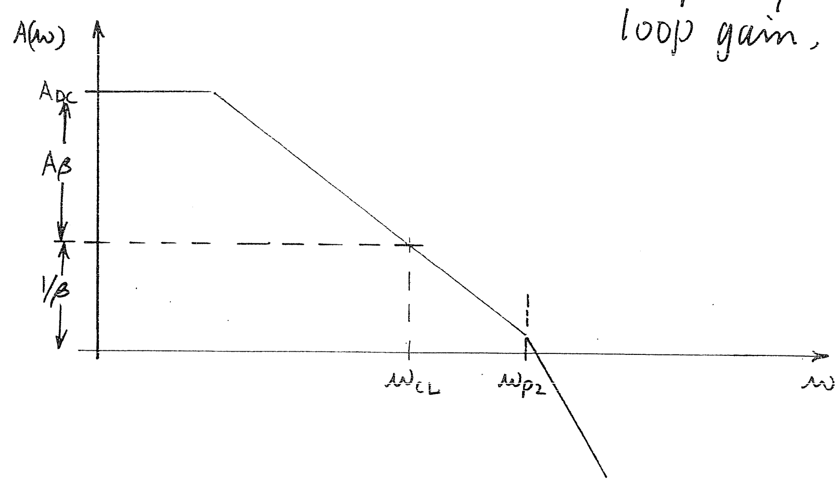
5.10) Two stage opamp:



$g_m = 0.775\,mA/V$

$\omega_{p2} = 60\,MHz$

ω_{CL} is defined as the frequency at which the loop gain, $A\beta$, is unity.



(cont.)

5.10 (cont.)

For a closed loop phase margin of 55° eqn (5.53) states

$$\omega_t = \tan(90^\circ - PM)\,\omega_{eq}$$

$$= \tan(90^\circ - 55^\circ)\,\omega_{eq}$$

$$\omega_t = 0.7\,\omega_{eq}$$

For $\omega_{eq} = 2\pi \times 60\,MHz$, $\omega_t = 2\pi \times 42\,MHz$

Now including a feedback factor β into eqn (5.46), we have

$$A(s) = \frac{\omega_{ta}\,\beta}{s(1 + s/\omega_{eq})} = \frac{g_m\beta / C_c}{s(1 + s/\omega_{eq})}$$

Here $\beta = \dfrac{R_1}{R_1 + R_2} = \dfrac{1}{6}$

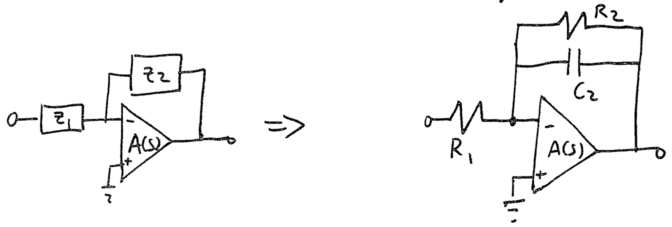
We require that $|A(j\omega_t)| = 1$ leading to

$$|A(j\omega_t)| = 1 = \frac{g_m\beta / C_c}{|\omega_t|\sqrt{1 + \left(\frac{\omega_t}{\omega_{eq}}\right)^2}}$$

$$\therefore \quad C_c = \frac{g_m\beta}{\omega_t\sqrt{1 + 0.7^2}} = \frac{0.775\times10^{-3} \times \frac{1}{6}}{2\pi \times 42\times10^{6} \times 1.221}$$

$$C_c = \underline{0.4\,pF}$$

5.11) Find C_2 that provides compensation.



In Problem 5.10, we found that

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

This can be generalized to $\beta = \frac{Z_1}{Z_1 + Z_2}$. We now wish to use the feedback network for compensation.

For $Z_1 = R_1$, $Z_1 = R_2 \,//\, \frac{1}{sC_2} = \frac{R_2}{1 + sR_2C_2}$

$$\beta = \frac{R_1}{R_1 + \frac{R_2}{1 + sR_2C_2}} = \frac{R_1(1 + sR_2C_2)}{sR_1R_2C_2 + R_1 + R_2}$$ new zero added

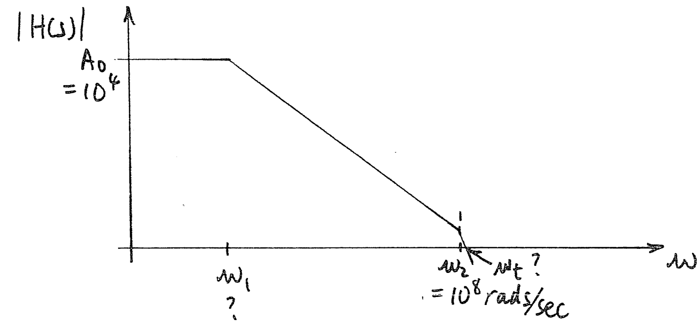
From Section 5.2, lead compensation is achieved by placing a zero at 1.2 times the unity loop gain frequency, ω_t.

$$\therefore \quad \omega_z = \frac{1}{R_2C_2} \equiv 1.2 \times \omega_t$$

$$\therefore \quad C_2 = \frac{1}{1.2 \times 0.7\omega_{p2} \times R_2}$$

$$= 63\,fF$$

5.12) Find ω_1 and ω_t.



$\because$ $\omega_z = \omega_t$

$\therefore$ $H(s) = \dfrac{A_0(1 + s/\omega_t)}{(1+s/\omega_1)(1+s/\omega_2)}$

from zero

and $\measuredangle H(j\omega_t) = 180° - PM = -100° = +45° - \tan^{-1}\left(\frac{\omega_t}{\omega_1}\right) - \tan^{-1}\left(\frac{\omega_t}{\omega_2}\right)$

$\because$ $A_0 >> 1$

$\therefore$ $\omega_t >> \omega_1$ by the constant gain bandwidth product

$\therefore$ $\tan^{-1}\left(\frac{\omega_t}{\omega_1}\right) \approx 90°$

$\therefore$ $-100° \approx +45° - 90° - \tan^{-1}\left(\frac{\omega_t}{\omega_2}\right)$

$\frac{\omega_t}{\omega_2} \approx \tan(55°)$

$\therefore$ $\omega_t \approx 1.43\,\omega_2 = 1.43 \times 10^8$ rads/sec

$|H(j\omega_t)| = A_0\sqrt{2} \times \dfrac{1}{\sqrt{1+\left(\frac{\omega_t}{\omega_1}\right)^2} \times \sqrt{1+\left(\frac{\omega_t}{\omega_2}\right)^2}} \equiv 1$

$\left[1+\left(\frac{\omega_t}{\omega_1}\right)^2\right]\left[1+\left(\frac{\omega_t}{\omega_2}\right)^2\right] = 2A_0^2$

$1+\left(\frac{\omega_t}{\omega_1}\right)^2 = \frac{2\times10^8}{1+1.43^2}$

$\omega_1 = \frac{1.43\times10^8 \text{ rads/sec}}{8.1\times10^3}$

$\omega_1 = 1.8\times10^4$ rads/sec

5.13) Given $A(s) \approx \frac{A_o(1+s\tau_z)}{s\tau_1(1+s\tau_2)}$, the closed loop transfer function is given by

$$A_{CL}(s) = \frac{A(s)}{1+A(s)\beta} \approx \frac{A_o(1+s\tau_z)/(s\tau_1(1+s\tau_2))}{1+A_o\beta(1+s\tau_z)/(s\tau_1(1+s\tau_2))}$$

$$= \frac{A_o(1+s\tau_z)}{s\tau_1(1+s\tau_2)+A_o\beta(1+s\tau_z)}$$

$$\therefore A_{CL}(s) = \frac{A_o(1+s\tau_z)}{\tau_1\tau_2\left[s^2+s\left(\frac{1}{\tau_2}+A_o\beta\frac{\tau_z}{\tau_1\tau_2}\right)+\frac{A_o\beta}{\tau_1\tau_2}\right]}$$

Equating the coefficients of the denominator polynomial to the standard second-order pole polynomial,

$$s^2+\frac{\omega_o}{Q}s+\omega_o^2,$$

we find that

$$\frac{\omega_o}{Q} = \frac{1}{\tau_2}+A_o\beta\frac{\tau_z}{\tau_1\tau_2}, \qquad \omega_o^2 = \frac{A_o\beta}{\tau_1\tau_2}$$

$$\therefore \omega_o = \sqrt{\frac{A_o\beta}{\tau_1\tau_2}}$$

$$\therefore Q = \frac{\omega_o}{\frac{1}{\tau_2}+A_o\beta\frac{\tau_z}{\tau_1\tau_2}}$$

$$= \frac{\omega_o\tau_2}{1+A_o\beta\,\tau_z/\tau_1}$$

Since τ_z is roughly τ_1/A_o we cannot make any approximations here.

5.14) Given $R_c = 0$, Equations (5.66) ~ (5.68) describe the denominator polynomial, $D(s)$, as

$$D(s) = 1 + sa + s^2 b \quad \text{where}$$
$$a = (C_2 + C_c)R_2 + (C_1 + C_c)R_1 + g_{m7}R_1R_2C_c$$
$$b = R_1R_2(C_1C_2 + C_1C_c + C_2C_c)$$

When $R_c \neq 0$, the admittance sC_c becomes $\frac{sC_c}{1+sR_cC_c}$. Thus, we can obtain the new denominator polynomial, $D'(s)$, by simply substituting C_c with $\frac{C_c}{1+sR_cC_c}$.

$$\therefore a' = \left(C_2 + \frac{C_c}{1+sR_cC_c}\right)R_2 + \left(C_1 + \frac{C_c}{1+sR_cC_c}\right)R_1 + g_{m7}R_1R_2\frac{C_c}{1+sR_cC_c}$$

$$= \frac{1}{1+sR_cC_c}\Big[(C_2(1+sR_cC_c)+C_c)R_2 + (C_1(1+sR_cC_c)+C_c)R_1$$
$$+ g_{m7}R_1R_2C_c\Big]$$

$$\underline{a' = \frac{1}{1+sR_cC_c}\left[a + s(R_2C_2R_cC_c + R_1C_1R_cC_c)\right]}$$

$$b' = R_1R_2\left[C_1C_2 + C_1\frac{C_c}{1+sR_cC_c} + C_2\frac{C_c}{1+sR_cC_c}\right]$$

$$\underline{b' = \frac{1}{1+sR_cC_c}\left[b + sR_1C_1R_2C_2R_cC_c\right]}$$

$$\therefore D'(s) = 1 + sa' + s^2b' = \frac{1}{1+sR_cC_c}\underbrace{\Big[(1+sR_cC_c) + s(a + s(R_2C_2R_cC_c + R_1C_1R_cC_c)) + s^2(b + sR_1C_1R_2C_2R_cC_c)\Big]}_{\triangleq D''(s)}$$

this term is moved to the numerator

The poles of the system are determined by the roots of $D''(s)$.

$$D''(s) = 1 + s(a + R_cC_c) + s^2(b + R_2C_2R_cC_c + R_1C_1R_cC_c)$$
$$+ s^3 R_1C_1R_2C_2R_cC_c$$

$\because$ $R_c \ll R_1$ or R_2

$\therefore$ any time constant with R_c is much less than time constants with R_1 or R_2

$\therefore$ $a \gg R_cC_c$ and $b \gg R_2C_2R_cC_c + R_2C_1R_cC_c$

(cont.)

5.14 (cont.)

$\therefore D''(s) \approx 1 + sa + s^2 b + s^3 R_1 C_1 R_2 C_2 R_C C_C$

$\because$ system poles are spaced far apart with $\omega_{p1} << \omega_{p2} << \omega_{p3}$

$\therefore (1 + s/\omega_{p1})(1 + s/\omega_{p2})(1 + s/\omega_{p3}) \approx 1 + \frac{s}{\omega_{p1}} + \frac{s^2}{\omega_{p1}\omega_{p2}} + \frac{s^3}{\omega_{p1}\omega_{p2}\omega_{p3}}$

Equating these coefficients, we have the original approximations,

$\omega_{p1} \approx 1/a$

$\omega_{p2} \approx a/b$

$\therefore$ Equations (5.70) and (5.71) still hold true.

Q.E.D.

5.15) Find R_B and V_{eff} at 70°C.

From Eq. (5.108)

$R_B = 1/g_{m13} = 1/(\mu_n C_{ox} W/L\, V_{eff}) = 1/(92\times10^{-6} \times 10/1.2 \times 0.25)$

$\underline{R_B = 5.2\,k\Omega}$

To determine the effect of temperature on V_{eff}, note that

$\mu_n \propto T^{-3/2}$ (see pg 250)

Also $R_B \propto \frac{1}{\mu_n V_{eff}} \propto \frac{1}{T^{-3/2} V_{eff}}$ and assuming R_B is a constant

$\therefore T_1^{-3/2} V_{eff1} = T_2^{-3/2} V_{eff2}$

Let $T_1 = 300K$ (27°C), $T_2 = 343K$ (70°C), $V_{eff1} = 0.25V$

$\therefore V_{eff2} = 0.25 \times \left(\frac{300}{343}\right)^{-3/2} = \underline{0.31V}$

Results from HSPICE:

at 27°C: $V_{eff} = V_{Dsat} = \underline{0.26V}$

at 70°C: $V_{eff} = V_{Dsat} = \underline{0.31V}$

$\therefore$ Results are consistent with our calculations.