Electrical and Computer Engineering Department
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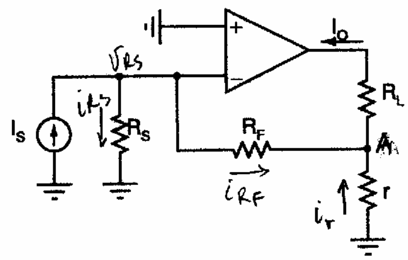
Electronics II - INEL 4202 - SPRING 2002 - Final Exam - Prof. Manuel Toledo
THIS EXAM CONTAINS FOUR PROBLEMS FOR A TOTAL OF 105 POINTS
WORK CLEARLY OR LOOSE POINTS

1. The following circuit shows how to design a current amplifier using an op-amp and current-sampling, current-mixing (*shunt-series*) feedback.



(a) Show, without using feedback theory but assuming an ideal op-amp, that the current gain is given by (15 points)

$$\frac{I_O}{I_S} = 1 + \frac{R_F}{r}$$

(b) Show that feedback theory predicts the gain to be the expression given in part (a) when the op-amp gain is taken to be very large. (HINT: the feedback network is composed of resistors r and R_F.) (5 points)

(c) Using The feedback-analysis method find the closed-loop gain $\frac{I_O}{I_S}$, the input resistance (excluding R_S), and the output resistance (excluding R_L) for the case: open-loop gain of op-amp $= 10^4 V/V$, differential input resistance for op-amp $R_{id} = 100k\Omega$, op-amp output resistance $= 1k\Omega$, $R_S = R_L = 10k\Omega$, $r = 100\Omega$, and $R_F = 1k\Omega$. (20 points)

Answer

(a) Since $v_- = v_+ = 0V$, $v_{RS} = 0$ and $i_{RS} = 0$. $\therefore$ $i_{RF} = I_S$.
and $V_A = -I_S R_F$. KCL ~~at~~ node A yields $I_O = i_r + i_{RF}$,
or

$$I_O = I_S + \frac{-V_A}{r} = I_S + I_S \frac{R_F}{r} \Rightarrow \boxed{\frac{I_O}{I_S} = 1 + R_F/r}$$

(b) For $A\beta >> 1$, $A_f \simeq 1/\beta$. For shunt-series feedback, $\beta = \left.\frac{i_1}{i_2}\right|_{v_1=0}$

$\Rightarrow i_1 = -i_2 \frac{r}{r+R_F} \Rightarrow \frac{i_1}{i_2} = -\frac{1}{1+R_F/r}$

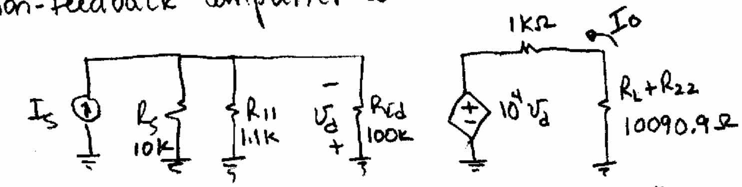
Observing that $I_o = -i_2 \Rightarrow \beta = \frac{1}{1+R_F/r}$ and

$$\boxed{A_f \simeq \frac{1}{\beta} = 1 + R_F/r}$$

(c) $R_{11} = \left.\frac{v_1}{i_1}\right|_{i_2=0} = R_F + r = \underline{1.1\text{k}\Omega}$; $R_{22} = \left.\frac{v_2}{i_2}\right|_{v_1=0} = R_F \| r = \underline{90.9\ \Omega}$

$\beta = \frac{1}{1+R_F/r} = \frac{1}{1+1\text{k}\Omega/100\Omega} = \underline{\frac{1}{11}}$

The non-feedback amplifier is



$I_o = -10^4 v_d / 11090.9\Omega$; $v_d = -100\text{k}\,(I_S)\,\frac{10\text{k}\|1.1\text{k}}{100\text{k} + 10\text{k}\|1.1\text{k}} = -981.3\, I_S$

$\therefore A_I = \frac{I_o}{I_S} = \left(-\frac{10^4}{11090.9\Omega}\right)\left(-981.3\Omega\right) = +\underline{884.75\ \text{A/A}}$

$D = 1 + \beta A = 1 + \frac{1}{11} \times 884.75 = \underline{81.43}$

$A_f = \frac{I_o}{I_S} = \frac{A}{D} = \frac{884.75}{81.43} = \boxed{10.87\ \text{A/A}}$

$R_{if} = \frac{R_i}{D} = \frac{10\text{k}\|1.1\text{k}\|100\text{k}}{81.43} = \underline{12\ \Omega} \Rightarrow R_{in} = \left(\frac{1}{R_{if}} - \frac{1}{R_S}\right)^{-1} = \left(\frac{1}{12} - \frac{1}{10\text{k}}\right)^{-1} \simeq \boxed{12\ \Omega}$

$R_{of} = R_o \times D = 11090.9\Omega \times 81.43 = 903\text{k}\Omega$;

$R_{out} = R_{of} - R_L = 903\text{k}\Omega - 10\text{k}\Omega = \boxed{893\text{k}\Omega}$

2. Consider a feedback amplifier for which the open-loop gain $A(s)$ is given by

$$A(s) = \frac{1000}{(1+s/10^4)(1+s/10^5)^2}$$

If the feedback factor β is independent of frequency, find

(a) the frèquency at which the phase shift is 180°, (10 points) and

(b) the critical value of β at which instability will begin.(10 points)

(a) $\phi = \tan^{-1}\left(\frac{\omega}{10^4}\right) + 2\tan^{-1}\left(\frac{\omega}{10^5}\right) = 180^\circ$

We need to find ω. If we express this eq. as

$$\omega = 10^5 \tan\left(90^\circ - \frac{1}{2}\tan^{-1}(\omega/10^4)\right)$$

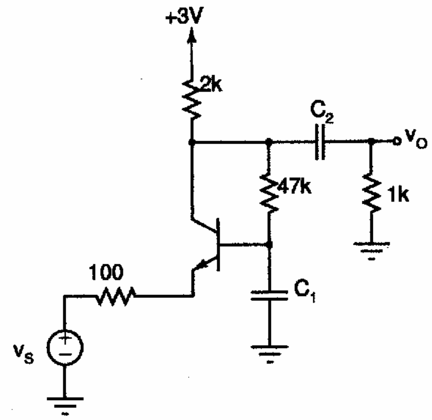
we can try to solve recursively. The solution converges to $\boxed{\omega = 109544.5 \text{ rps}}$. The phase angle at this freq. is indeed 180°, as can be verified by direct substitution.

(b) At $\omega = 109544.5$ rps

$$|A| = \frac{1000}{\sqrt{1+\left(\frac{109544.5}{10^4}\right)^2}\left(\sqrt{1+1.095445^2}\right)^2} = 41.32$$

∴ The critical value of β is $\boxed{\frac{1}{41.32}}$, or $\boxed{0.0242}$

3. For the common-base circuit shown below, assuming that the bias collector current to be about 1 mA, $\beta = 100$, $C_\mu = 0.8pF$, $f_T = 600MHz$, $C_1 = 1\mu F$ and $C_2 = 10\mu F$:



(a) Determine the midband gain v_O/v_S. (5 points)

(b) Use the short-circuit time-constant method to estimate the lower 3-dB frequency, f_L. (15 points)

(c) Find the high-frequency poles, and estimate the upper 3-dB frequency, f_H. (15 points)

(a) Since the circuit is a common-base amplifier, $A_v = \frac{v_c}{v_e} = +g_m R_c$

The overall mid-band gain is

$$\frac{v_o}{v_s} = \left(\frac{1}{25}\right)\left(2k\|1k\|47k\right)\left(\frac{25}{125}\right) = \boxed{+5.26\ V/V}$$

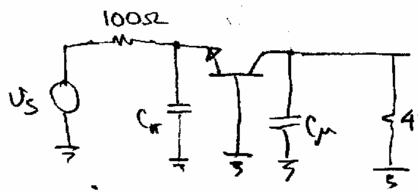
(b) C_1 sees $\left[r_\pi + (\beta+1)100\right] \| \left[47k + 2k\|1k\right] = 12.6k\|47.7k \simeq 9966\Omega$

$$\therefore \omega_{L1} = \frac{1}{9966\Omega \times 10^{-6}F} = \boxed{100.3\ rps}$$

C_2 sees $1k + 2k\|47k \Rightarrow \omega_{L2} = \frac{1}{10\mu F \times 2.92k} = \boxed{34.3\ rps}$

$$\therefore \omega_L \simeq \omega_{L1} + \omega_{L2} = \boxed{134.6\ rps}$$

(c) At high freqs, the circuit looks as follows



C_π sees $\frac{1}{g_m} \| 100\Omega = 25\Omega // 100\Omega = 20\Omega$

C_μ sees 657Ω.

To find C_π, observe that $f_T = \dfrac{\beta_o}{2\pi r_\pi (C_\mu + C_\pi)} = \dfrac{100}{(5k)\pi(.8pF + C_\pi)}$

$$C_\pi = \frac{100}{5000\pi(600)(10^6)} - .8pF = 9.8pF$$

So that $\omega_\pi = \dfrac{1}{9.8pF \times 20\Omega} = 5.1 \times 10^9$ rps

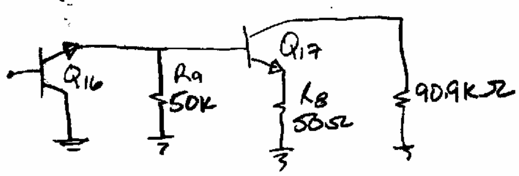
and $\omega_\mu = \dfrac{1}{.8pF \times 657\Omega} = 1.9 \times 10^9$ rps

Since they are less than 2 octaves away,

$$\omega_H = \frac{1}{\frac{1}{\omega_\pi} + \frac{1}{\omega_\mu}} = \boxed{1.3 \times 10^9 \text{ rps}} = \boxed{220\text{ MHz}}$$

4. Consider a variation on the design of the 741 second stage in which $R_8 = 50\Omega$. What will be the corresponding input resistance R_{i2} and trans-resistance R_{m2} of the second stage? (10 points)

From class, we know that $I_{13B} \cong 550\mu A$. $\Rightarrow r_{o13B} = \frac{50V}{550\mu A} = 90.9k\Omega$ is the load of Q_{17}. The second stage is as follows, for a.c.



$$g_{m17} = \frac{550\mu A}{25mV} = \frac{.55}{25}\frac{A}{V}$$

To find the gain, we must first estimate the collector current of Q_{16}. From $V_{E17} \approx 50\Omega(550\mu A) = 0.0275V = 27.5mV$ and

$v_{BE17} = V_T \ln\left(\frac{I_{C17}}{I_S}\right) = 25mV \ln\left(\frac{550\mu A}{10^{-14}A}\right) = 618mV$, we get that

$V_{E16} = 618mV + 27.5mV = 645.8mV$. Thus $I_{R9} = \frac{.6458V}{50k\Omega} = 12.9\mu A$.

Since $I_{B17} = \frac{550\mu A}{\beta_N} = \frac{550\mu A}{200} = 2.75\mu A$, $I_{E16} = 15.65\mu A$ and

$g_{m16} = 6.26 \times 10^{-4} A/V$.

The input resistance is

$$R_{in2} = r_{\pi16} + (\beta_N+1)\left[50k \,//\, (r_{\pi17} + (\beta_N+1)50\Omega)\right] =$$
$$= 319.5k\Omega + (201)\left[50k \,//\, (9k + 201 \times 50)\right] = \boxed{3.1M\Omega}$$

$$R_{M2} = R_{iN2} A_{v2} = 3.1M\Omega \left(\frac{g_{m16}\left[50k//(9k+201\times50)\right]}{1+g_{m16}\left[50k//(9k+201\times50)\right]}\right)\left(\frac{-g_{m17}(90.9k)}{1+g_{m17}(50)}\right)$$

$$= -3.1\times10^6 \times 0.897 \times 952 = \boxed{-2.6G\Omega}$$