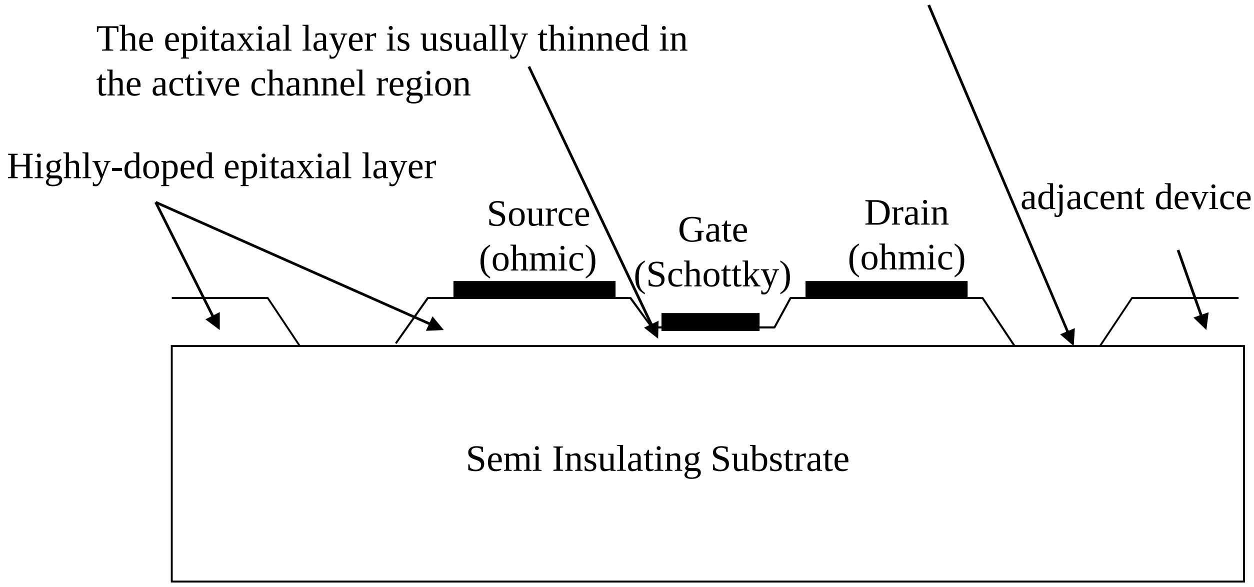
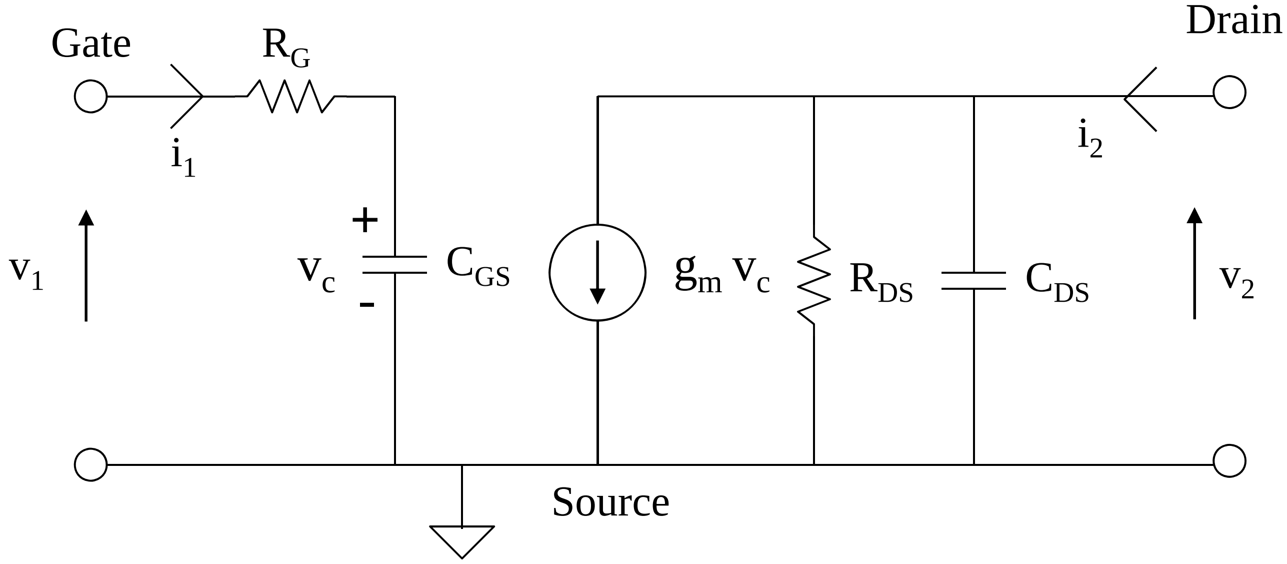
Q1 (a) MESA-isolated MESFET – 4 marks

By removing the highly-doped epitaxial layer between the devices, the MESA-etch process provides isolation between the different devices.



Q1 (b) Small-signal model and y-parameters - 8 marks

Simplified MESFET small-signal circuit, ignoring C_{GD}



Q1 (b) Small-signal model and y-parameters

Definition of y-parameters

$$y_{11} = \left.\frac{i_1}{v_1}\right|_{v_2=0} \qquad y_{21} = \left.\frac{i_2}{v_1}\right|_{v_2=0} \qquad y_{12} = \left.\frac{i_1}{v_2}\right|_{v_1=0} \qquad y_{22} = \left.\frac{i_2}{v_2}\right|_{v_1=0}$$

Analysing the small-signal circuit when v_2=0

$$i_1 = \frac{v_1}{R_G + 1/j\omega C_{GS}} = v_1 \frac{j\omega C_{GS}}{1 + j\omega R_G C_{GS}}$$

$$\Rightarrow y_{11} = \left.\frac{i_1}{v_1}\right|_{v_2=0} = \frac{j\omega C_{GS}}{1 + j\omega R_G C_{GS}}$$

$$v_c = v_1 \frac{1/j\omega C_{GS}}{R_G + 1/j\omega C_{GS}} = v_1 \frac{1}{1 + j\omega R_G C_{GS}}$$

$$i_2 = g_m v_c = v_1 \frac{g_m}{1 + j\omega R_G C_{GS}}$$

Note: there is no current in R_{DS} or C_{DS} because v_2=0

$$\Rightarrow y_{21} = \left.\frac{i_2}{v_1}\right|_{v_2=0} = \frac{g_m}{1 + j\omega R_G C_{GS}}$$

Analysing the small-signal circuit when v_1=0

$$v_1 = 0 \Rightarrow i_1 = 0 \Rightarrow y_{12} = \left.\frac{i_1}{v_2}\right|_{v_1=0} = 0$$

$$i_1 = 0 \Rightarrow v_c = 0 \Rightarrow i_2 = v_2\left(\frac{1}{R_{DS}} + j\omega C_{DS}\right)$$

Note: no current in the transconductance element when v_c=0

$$\Rightarrow y_{22} = \left.\frac{i_2}{v_2}\right|_{v_1=0} = \frac{1}{R_{DS}} + j\omega C_{DS}$$

Q1 (c) Determining the small-signal values – 5 marks

$$y_{11} = \frac{j\omega C_{GS}}{1 + j\omega R_G C_{GS}} \Rightarrow \frac{1}{y_{11}} = \frac{1 + j\omega R_G C_{GS}}{j\omega C_{GS}} = -j\frac{1}{\omega C_{GS}} + R_G$$

$$\Rightarrow R_G = \mathbf{Real}\left(\frac{1}{y_{11}}\right) \quad C_{GS} = -\frac{1}{\omega \mathbf{Imag}\left(\frac{1}{y_{11}}\right)}$$

$$y_{12} = 0$$

$$y_{21} = \frac{g_m}{1 + j\omega R_G C_{GS}} \Rightarrow \frac{1}{y_{21}} = \frac{1 + j\omega R_G C_{GS}}{g_m} = \frac{1}{g_m} + j\frac{\omega R_G C_{GS}}{g_m}$$

$$\Rightarrow g_m = \frac{1}{\mathbf{Real}\left(\frac{1}{y_{21}}\right)}$$

$$y_{22} = \frac{1}{R_{DS}} + j\omega C_{DS}$$

$$\Rightarrow R_{DS} = \frac{1}{\mathbf{Real}(y_{22})} \quad C_{DS} = \frac{\mathbf{Imag}(y_{22})}{\omega}$$

$$\varpi = 2\pi f$$

Doing the calculations gives

R_G=7.5Ω, C_{GS}=2.5pF, g_m=0.1S, R_{DS}=100Ω, C_{DS}=0.5pF

Q1 (d) Calculating the gate resistance – 3 marks

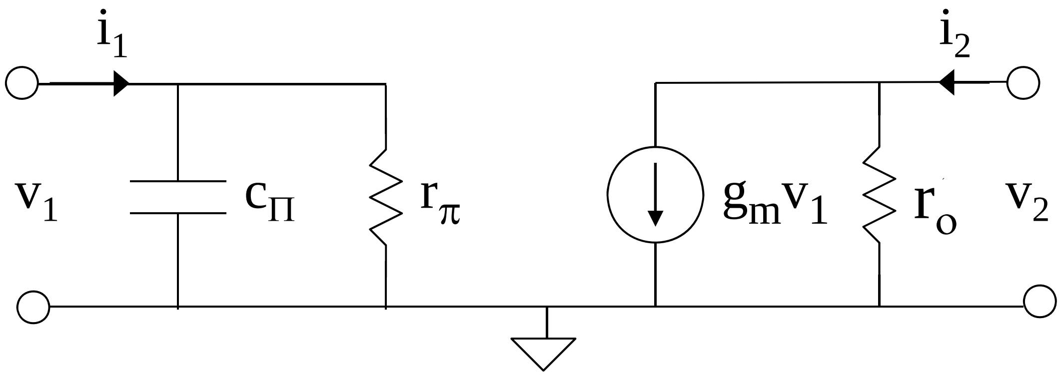
The effective gate resistance of a long single stripe contacted at one side is

$$R_{G,1} = \frac{1}{3}\frac{W}{L}R_{SQ}$$

The effective gate resistance of N parallel stripes each with the original gate length, L, and with the same total width W as the original device and with the gate stripes contacted on both sides is:

$$R_{G,N} = \frac{1}{12N^2}\frac{W}{L}R_{SQ} = \frac{1}{4N^2}R_{G,1} = \frac{1}{4\times 5^2}10 = 0.1\Omega$$

Q2



$$c_\pi = 40\,pF \quad r_\pi = 500\Omega \quad g_m = 0.25S \quad r_o = 200\Omega$$

Q2 (a) Expressions for i_1 and i_2 in terms of v_1 and v_2 – 2 marks

$$i_1 = v_1\left(\frac{1}{r_\pi} + j\omega c_\pi\right) \quad i_2 = g_m v_1 + \frac{v_2}{r_o}$$

Q2 (b) z-parameters – 8 marks

$$z_{11} = \left.\frac{v_1}{i_1}\right|_{i_2=0} \quad z_{21} = \left.\frac{v_2}{i_1}\right|_{i_2=0} \quad z_{12} = \left.\frac{v_1}{i_2}\right|_{i_1=0} \quad z_{22} = \left.\frac{v_2}{i_2}\right|_{i_1=0}$$

$$if\ i_1 = 0\ then\ v_1 = 0 \quad and \quad i_2 = \frac{v_2}{r_o}$$

$$z_{12} = \left.\frac{v_1}{i_2}\right|_{i_1=0} = 0$$

$$z_{22} = \left.\frac{v_2}{i_2}\right|_{i_1=0} = \frac{v_2}{1}\frac{r_o}{v_2} = r_o$$

$$if\ \ i_2 = 0\ then\ g_m v_1 + \frac{v_2}{r_o} = 0 \Rightarrow v_2 = -g_m r_o v_1$$

$$z_{11} = \left.\frac{v_1}{i_1}\right|_{i_2=0} = \frac{v_1}{v_1\left(\frac{1}{r_\pi} + j\omega c_\pi\right)} = \frac{r_\pi}{1 + j\omega c_\pi r_\pi}$$

$$z_{21} = \left.\frac{v_2}{i_1}\right|_{i_2=0} = \frac{-g_m r_o v_1}{v_1\left(\frac{1}{r_\pi} + j\omega c_\pi\right)} = -\frac{g_m r_o r_\pi}{1 + j\omega c_\pi r_\pi}$$

Putting in the small-signal element values and doing the calculations gives:

$$\omega = 2\pi f \qquad f = 0.5GHz$$

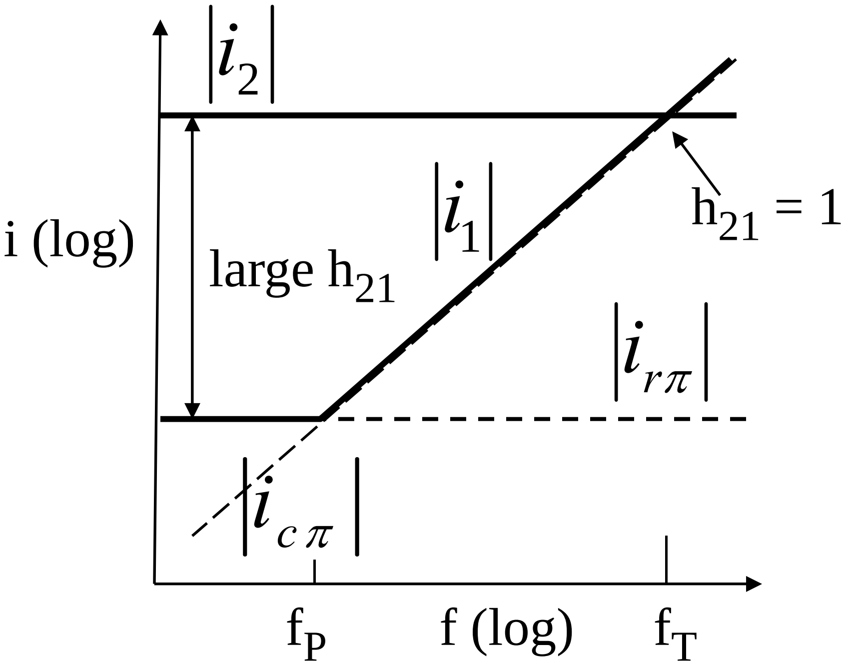
$$z_{11} = 7.96\angle -89^\circ\ \Omega$$

$$z_{12} = 0\ \Omega$$

$$z_{21} = 397.8\angle 91^\circ\ \Omega$$

$$z_{22} = 200\ \Omega$$

Q2(c) Sketch of currents vs. frequency during h_{21} measurement – 4 marks



Note: for h_{21} measurement, v_2 is set to 0 and i_1 and i_2 are measured vs. frequency for an ac signal applied to port 1.

Q2(d)(i) h_{21} at low frequencies – 2 marks

$$i_1 = v_1\left(\frac{1}{r_\pi} + j\omega c_\pi\right) \quad i_2 = g_m v_1 + \frac{v_2}{r_o}$$

$$h_{21} = \left.\frac{i_2}{i_1}\right|_{v_2=0} = \frac{g_m v_1}{v_1\left(\frac{1}{r_\pi} + j\omega c_\pi\right)} = \frac{g_m r_\pi}{1 + j\omega c_\pi r_\pi}$$

At low frequencies $h_{21} \approx g_m r_\pi = 125$

Q2(d)(ii) The input-circuit pole frequency – 2 marks

$$i_{r\pi} = \frac{v_1}{r_\pi} \quad i_{c\pi} = v_1 j\omega c_\pi$$

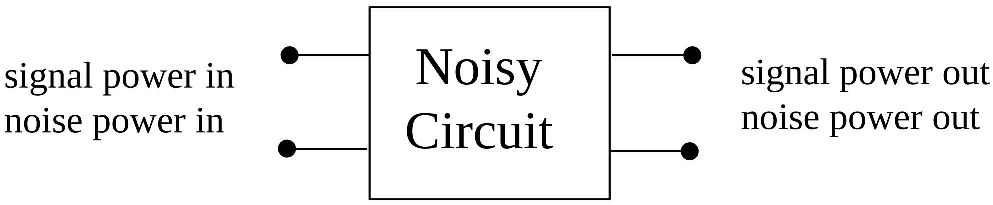
$$|i_{r\pi}| = |i_{c\pi}| \Rightarrow \frac{1}{r_\pi} = 2\pi f c_\pi \Rightarrow f = \frac{1}{2\pi r_\pi c_\pi} = 7.96MHz$$

Q2(d)(iii) The cut-off frequency – 2 marks

For high frequencies $h_{21} \approx \frac{g_m r_\pi}{j\omega c_\pi r_\pi} = \frac{g_m}{j\omega c_\pi} \Rightarrow |h_{21}| \approx \frac{g_m}{\omega c_\pi}$

$$|h_{21}| = 1 \Rightarrow f = f_T = \frac{g_m}{2\pi c_\pi} = 0.99GHz$$

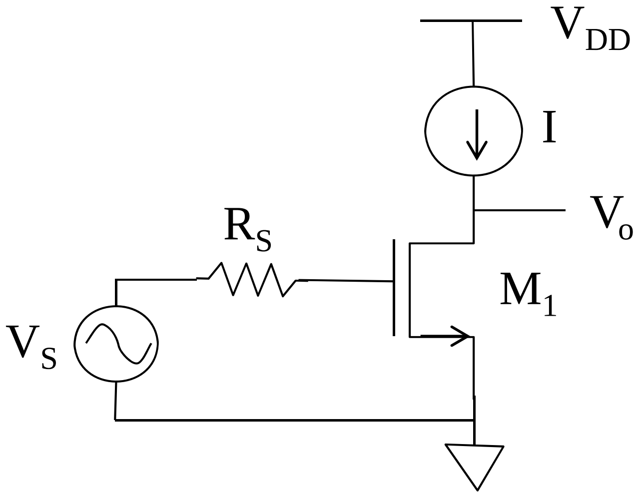
Q3(a) Noise Factor – 2 marks



SNR_{in} = (signal power in)/(noise power in)
SNR_{out} = (signal power out)/(noise power out)

$$\text{Noise Factor, F} = \frac{SNR_{in}}{SNR_{out}} \quad (usually \geq 1)$$

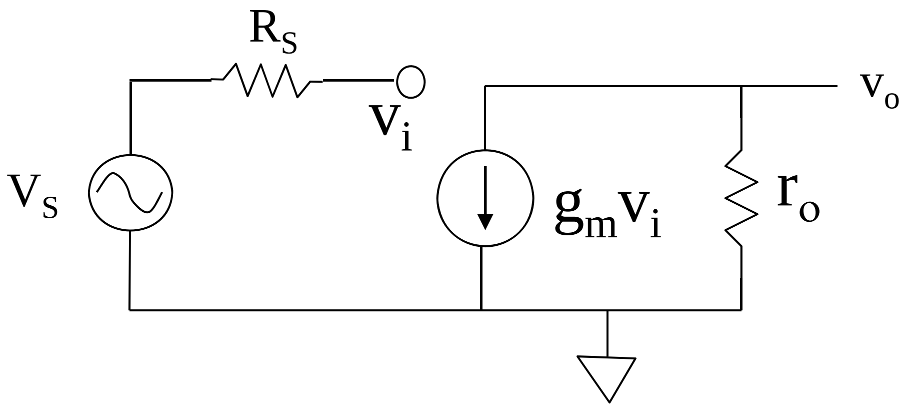
Q3(b) Noise Factor of common-source amplifier – 14 marks



$$\overline{i_n^{\ 2}} = 4kT\gamma g_m$$

$$\overline{v_{n,RS}^{\ \ 2}} = 4kTR_S$$

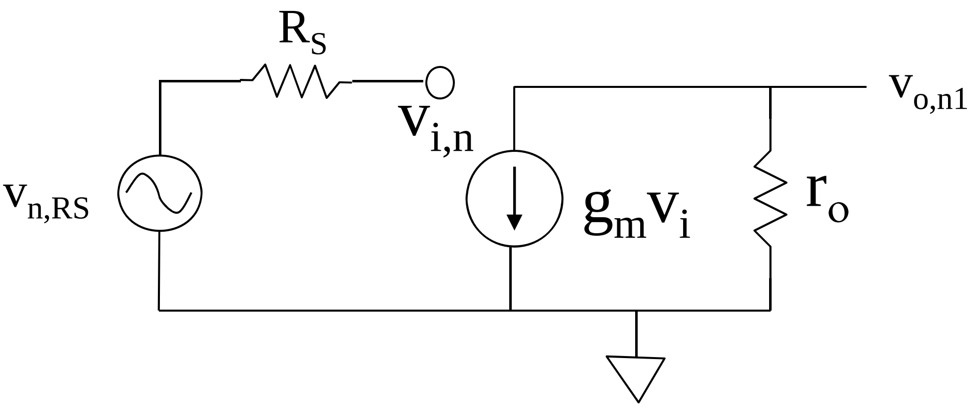
First analyse the circuit with just the signal source and no noise sources



$$v_{i,s} = V_s \Rightarrow v_{i,s}^{\ 2} = V_s^{\ 2}$$

$$v_{o,s} = -g_m r_o v_i = -g_m r_o V_s \Rightarrow v_{o,s}^{\ 2} = \left(g_m r_o\right)^2 V_s^{\ 2}$$

Next analyse the circuit with just the noise from the source resistance



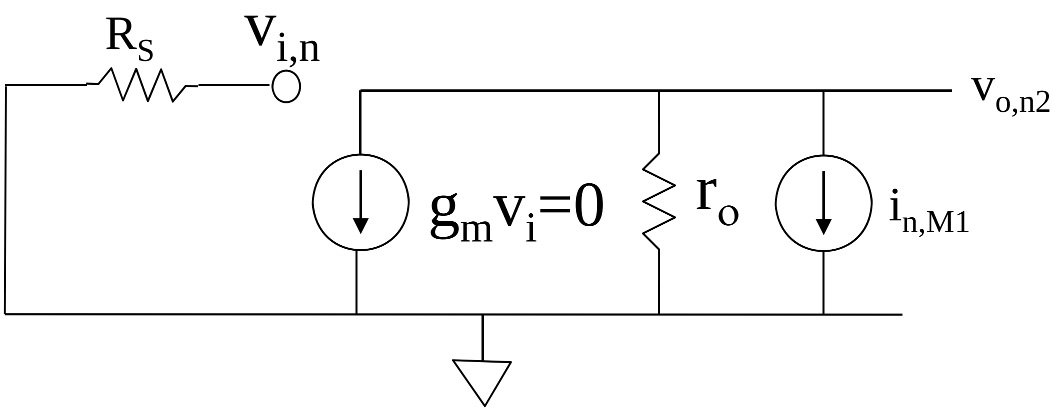
$$v_{i,n} = v_{n,RS} \Rightarrow \overline{v_{i,n}^{\,2}} = \overline{v_{n,RS}^{\,2}} = 4kTR_S \quad (\Delta f = 1Hz)$$

$$v_{o,n1} = -g_m r_o v_{i,n} \Rightarrow \overline{v_{o,n1}^{\,2}} = (g_m r_o)^2 \overline{v_{i,n}^{\,2}} = (g_m r_o)^2 4kTR_S$$

The signal to noise ratio at the input considers only the noise from the source resistance:

$$SNR_i = \frac{v_{i,s}^{\,2}}{\overline{v_{i,n}^{\,2}}} = \frac{V_s^{\,2}}{4kTR_S}$$

Now analyse the circuit with the noise from the transistor on its own



$v_{i,n} = 0$ (ignoring the noise from the source resistance)

$$v_{o,n2} = -r_o i_{n,M1} \Rightarrow \overline{v_{o,n2}^{\,2}} = r_o^{\,2} \overline{i_{n,M1}^{\,2}} = r_o^{\,2} 4kT\gamma g_m \quad (\Delta f = 1Hz)$$

The total noise power at the output can now be determined by adding the contributions at the output from the source resistance and from the transistor

$$\overline{v_{o,n,Total}^{\,2}} = \overline{v_{o,n1}^{\,2}} + \overline{v_{o,n2}^{\,2}} = (g_m r_o)^2 kTR_S + r_o^{\,2} 4kT\gamma\, g_m$$

The signal to noise ratio at the output is:

$$SNR_o = \frac{v_{o,s}^{\,2}}{\overline{v_{o,n,Total}^{\,2}}} = \frac{(g_m r_o)^2 V_s^{\,2}}{(g_m r_o)^2 4kTR_S + r_o^{\,2} 4kT\gamma\, g_m}$$

The noise factor is then:

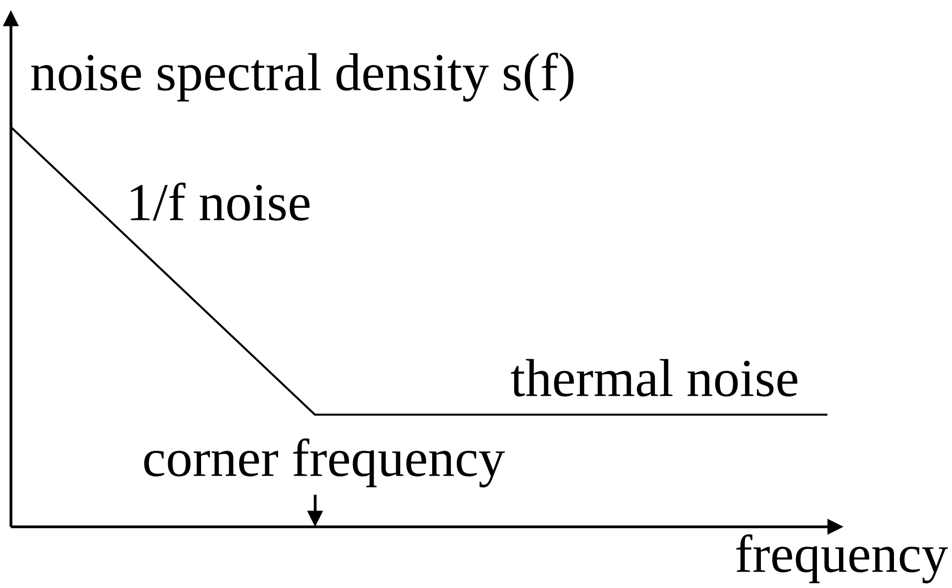
$$\mathrm{F} = \frac{SNR_i}{SNR_o} = \frac{V_s^{\,2}}{4kTR_S} \frac{(g_m r_o)^2 4kTR_S + r_o^{\,2} 4kT\gamma\, g_m}{(g_m r_o)^2 V_s^{\,2}}$$

$$= 1 + \frac{\gamma}{g_m R_S}$$

Q3(c) Noise Figure Calculation – 2 marks

$$R_S = 50\Omega \quad \gamma = \frac{2}{3} \quad g_m = 0.01S \quad \mathrm{F} = 1 + \frac{\gamma}{g_m R_S} = 2.33 = 3.7dB$$

Q3(d) Typical MOSFET noise spectral density – 2 marks



Q4(a) Transistor Stability Considerations – 2 marks

$$s_{11} = 0.65\angle -160° \quad s_{12} = 0.1\angle 30° \quad s_{21} = 4.0\angle 90° \quad s_{22} = 0.4\angle -60°$$

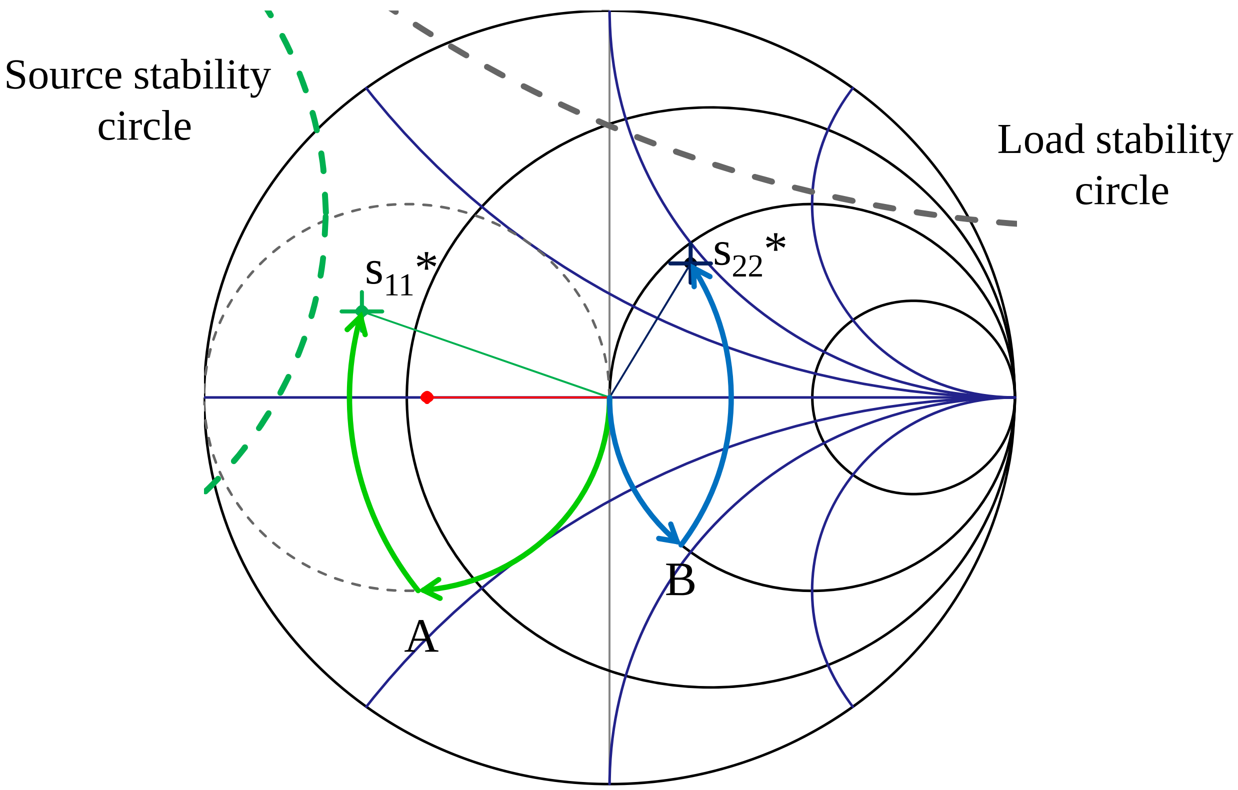
$$\Delta = s_{11}s_{22} - s_{12}s_{21} = 0.18\angle -90° \quad K = \frac{1 - |s_{11}|^2 - |s_{22}|^2 + |\Delta|^2}{2|s_{12}s_{21}|} = 0.56$$

K <1 so the device is only *conditionally stable* and the stability will depend on the source and load reflection coefficients.

Q4(b) Calculations for Source and Load Stability Circles and display on Smith Chart – 4 marks

$$CS_S = \frac{s_{11}^* - \Delta^* s_{22}}{|s_{11}|^2 - |\Delta|^2} = 1.79\angle 165° \quad RS_S = \frac{|s_{12}s_{21}|}{\left||s_{11}|^2 - |\Delta|^2\right|} = 1.02$$

$$CS_L = \frac{s_{22}^* - \Delta^* s_{11}}{|s_{22}|^2 - |\Delta|^2} = 3.78\angle 71° \quad RS_L = \frac{|s_{12}s_{21}|}{\left||s_{22}|^2 - |\Delta|^2\right|} = 3.13$$



In this case the stability circles don’t effect the placement of the source and load reflection coefficients at s_{11}* and s_{22}* for maximum unilateral transducer gain because s_{11}* and s_{22}* are on the stable sides of the stability circles.

$S_{11} = 0.65\angle{-160^\circ} \Rightarrow S_{11}^* = 0.65\angle 160^\circ$

$S_{22} = 0.4\angle{-60^\circ} \Rightarrow S_{22}^* = 0.4\angle 60^\circ$

← Radial line, Load Stability Circle

EXAMINATION NUMBER	QUESTION NUMBER EE4011 Summer 2013	DATE
	EE4011 RF IC Design	

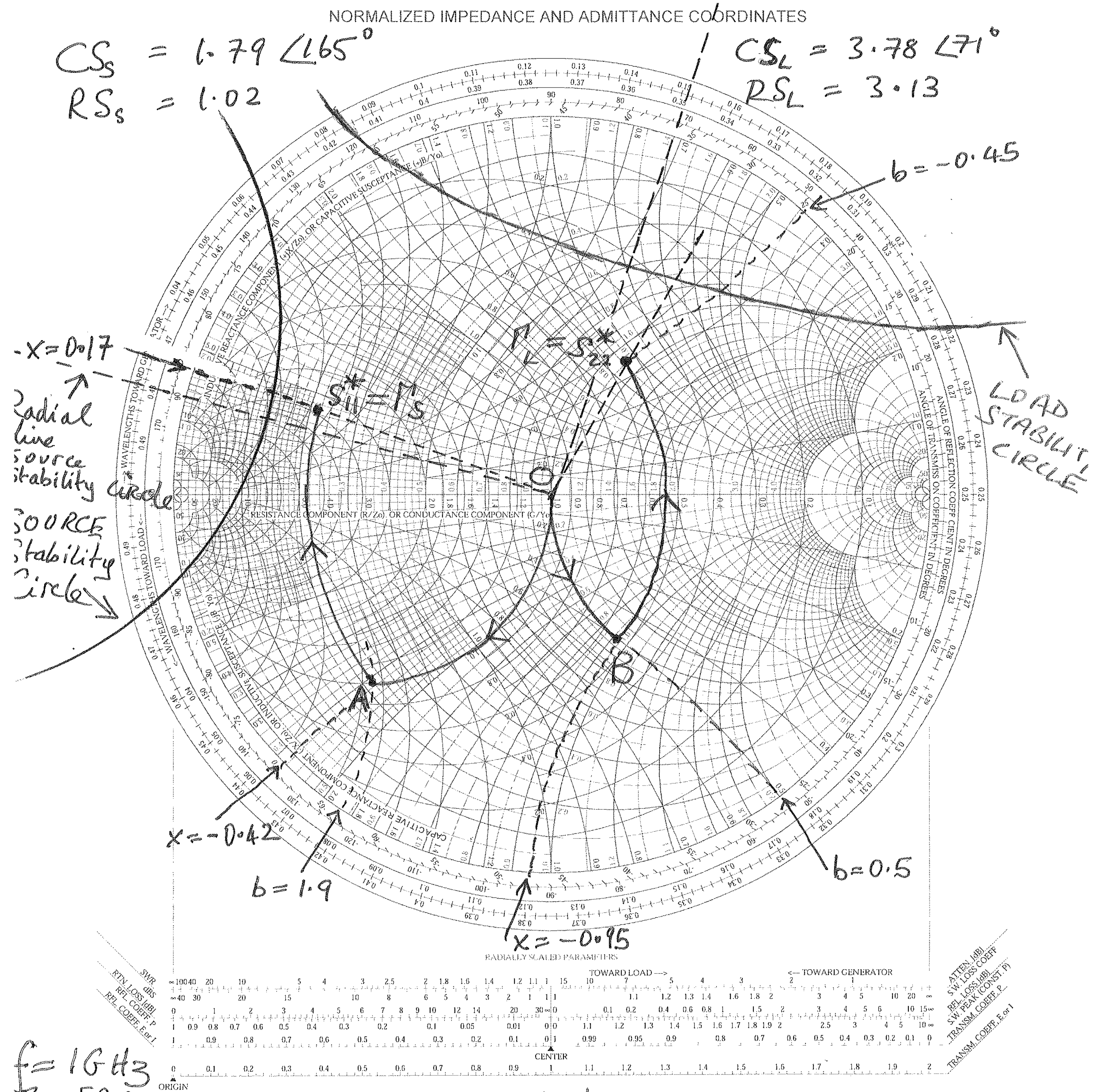
NORMALIZED IMPEDANCE AND ADMITTANCE COORDINATES

$CS_s = 1.79\angle 165^\circ$

$RS_s = 1.02$

$CS_L = 3.78\angle 71^\circ$

$RS_L = 3.13$



$f = 1\,GHz$

$Z_0 = 50\,\Omega$

From Origin to A : $|\Delta b| = 1.9 \Rightarrow C = \frac{|\Delta b|}{2\pi f Z_0} = 6.05pF$

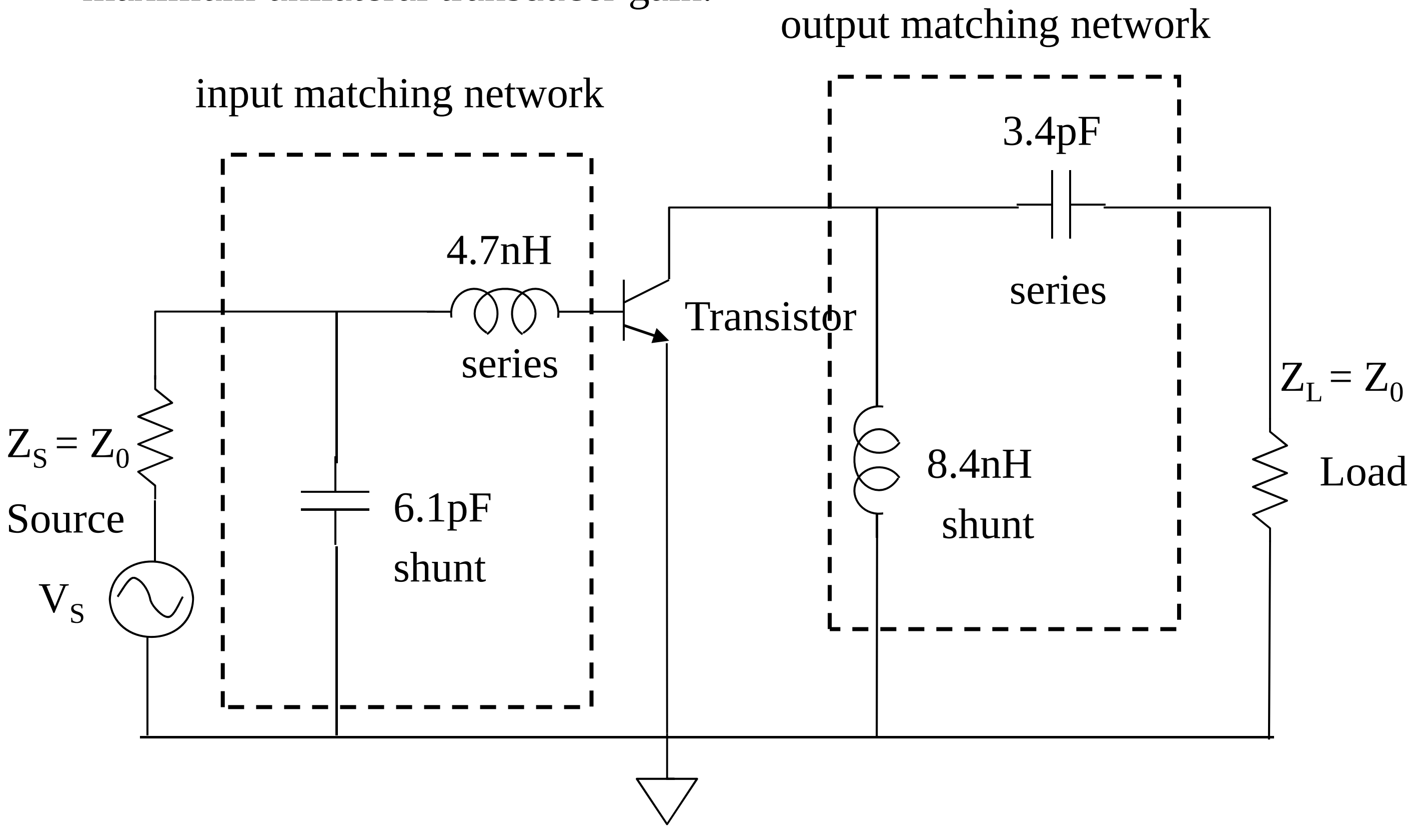
From A to S_{11}^* (Γ_s) : $|\Delta x| = 0.59 \Rightarrow L = \frac{Z_0|\Delta x|}{2\pi f} = 4.70nH$

From Origin to B : $|\Delta x| = 0.95 \Rightarrow C = 1/(2\pi f|\Delta x|Z_0) = 3.35pF$

From B to S_{22}^* (Γ_L) : $|\Delta b| = 0.95 \Rightarrow L = Z_0/(2\pi f|\Delta b|) = 8.38nH$

Q4(c) Input and Output Matching Networks – 10 marks

Using the Smith Chart to design input and output matching networks for maximum unilateral transducer gain.



(The matching calculations below are rounded to one decimal place)

Input Matching Element Values – need to get from origin to s_{11}*

Moving from Z_0 (Γ=0) to point A:
Clockwise on conductance circle – shunt capacitor

susceptance at Z_0: b = 0
susceptance at A: b = 1.9

$$C = \frac{|\Delta b|}{2\pi f Z_0} = \frac{|1.9|}{2\pi \times 1 \times 10^9 \times 50} = 6.1pF$$

Moving from A to Γ_S (s_{11}*):
Clockwise on resistance circle – series inductor

reactance at A: x = -0.42
reactance at Γ_S: x = 0.17

$$L = \frac{Z_0|\Delta x|}{2\pi f} = \frac{50 \times |0.59|}{2\pi \times 1 \times 10^9} = 4.7nH$$

Output Matching Element Values – need to get from origin to s_{22}*

Moving from Z_0 (Γ=0) to point B:
Anti-clockwise on resistance circle – series capacitor

reactance at Z_0: x = 0   reactance at B: x = -0.95

$$C = \frac{1}{2\pi f|\Delta x|Z_0} = \frac{1}{2\pi\times1\times10^9\times|-0.95|\times50} = 3.4\,pF$$

Moving from B to Γ_L:
Anti-clockwise on conductance circle – shunt inductor

susceptance at B: b = 0.5
susceptance at Γ_L: b = -0.45

$$L = \frac{Z_0}{2\pi f|\Delta b|} = \frac{50}{2\pi\times1\times10^9\times|0.95|} = 8.4\,nH$$

Q4(d)(i) Maximum Unilateral Transducer Gain – 2 marks

$$G_{TU,\max} = \frac{1}{1-|s_{11}|^2}|s_{21}|^2\frac{1}{1-|s_{22}|^2} = 33 = 15.2\,dB$$

Q4(d)(ii) Unilateral Figure of Merit and comment – 2 marks

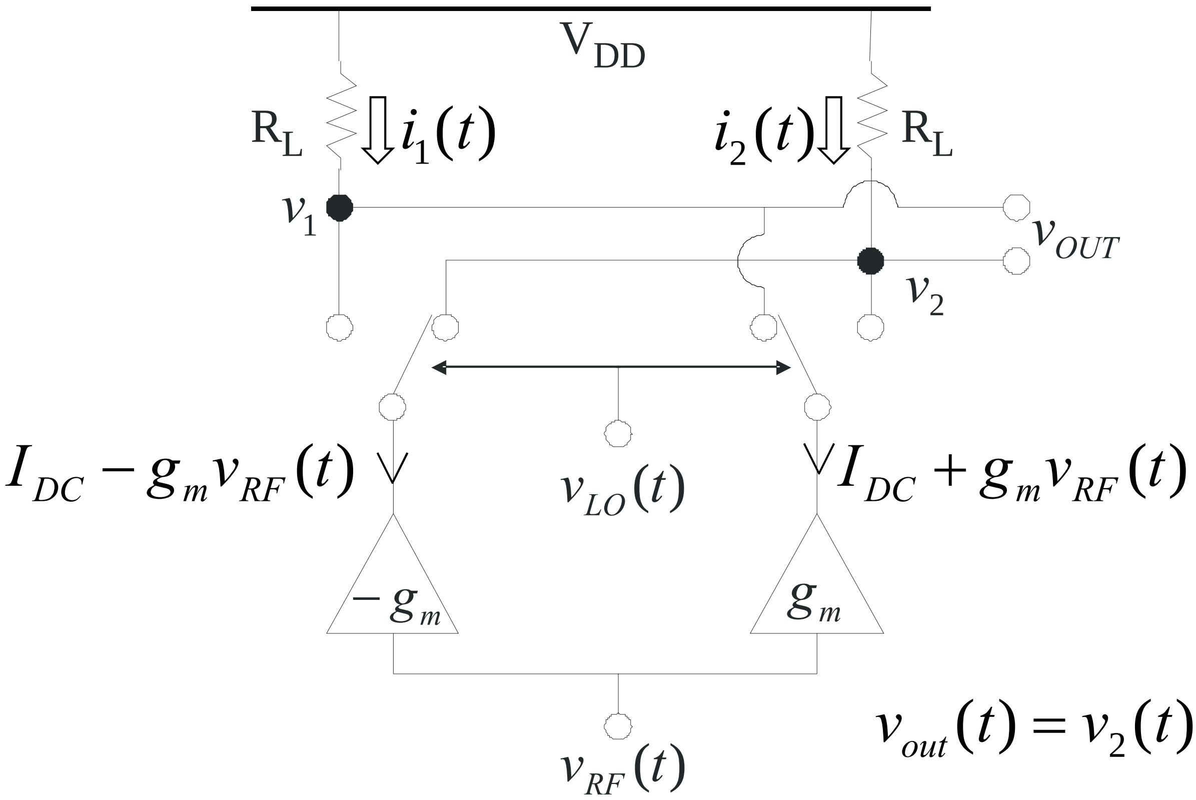
$$M = \frac{|s_{11}||s_{12}||s_{21}||s_{22}|}{\left(1-|s_{11}|^2\right)\left(1-|s_{22}|^2\right)} = 0.214$$

$$\frac{1}{(1+M)^2} < \frac{G_T}{G_{TU,\max}} < \frac{1}{(1-M)^2}$$

$$\frac{1}{(1+M)^2} = 0.68 = -1.7\,dB \qquad \frac{1}{(1-M)^2} = 1.67 = 2.1\,dB$$

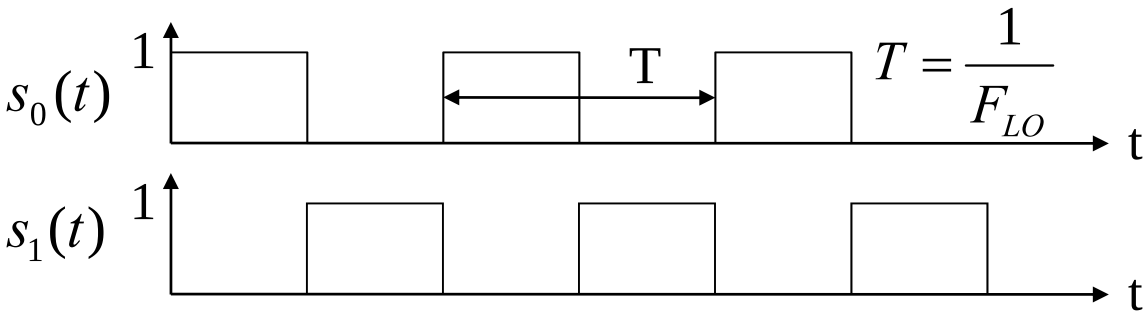
These calculations indicate that the error in calculating gain with the unilateral approximation is as high as 2dB and is not acceptable.

Question 5(a) Double-Balanced Mixer Derivation – 12 marks



$$v_{out}(t) = v_2(t) - v_1(t)$$

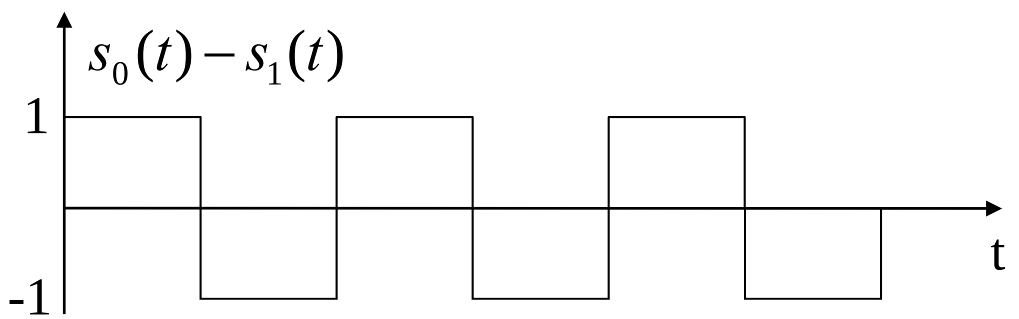
The LO drive can be represented as two non-overlapping square waves



$$i_1(t) = \left(I_{DC} + g_m v_{RF}(t)\right)s_0(t) + \left(I_{DC} - g_m v_{RF}(t)\right)s_1(t)$$

$$i_2(t) = \left(I_{DC} - g_m v_{RF}(t)\right)s_0(t) + \left(I_{DC} + g_m v_{RF}(t)\right)s_1(t)$$

$$i_1(t) - i_2(t) = 2g_m v_{RF}(t)\left(s_0(t) - s_1(t)\right)$$



$$
\begin{aligned}
v_{out}(t) &= v_2(t) - v_1(t) = \left(V_{DD} - i_2(t)R_L\right) - \left(V_{DD} - i_1(t)R_L\right) \\
&= R_L\left(i_1(t) - i_2(t)\right) \\
&= 2g_m R_L v_{RF}(t)\left(s_0(t) - s_1(t)\right) \\
&= 2g_m R_L v_{RF}(t)\frac{4}{\pi}\left[\sin(\varpi_{LO}t) + \frac{1}{3}\sin(3\varpi_{LO}t) + \cdots\right] \\
&= \frac{8g_m R_L}{\pi} v_{RF}(t)\left[\sin(\varpi_{LO}t) + \frac{1}{3}\sin(3\varpi_{LO}t) + \cdots\right]
\end{aligned}
$$

If the RF waveform is of the form:

$$V_{RF}(t)\cos(\varpi_{RF}t)$$

Then, use of cos(A)sin(B) expressions leads to

$$
\begin{aligned}
v_{out}(t) &= \frac{8g_m R_L v_{RF}(t)}{\pi}\left[\sin(\varpi_{LO}t) + \frac{1}{3}\sin(3\varpi_{LO}t) + \cdots\right] \\
&= \frac{8g_m R_L V_{RF}\cos(\varpi_{RF}t)}{\pi}\left[\sin(\varpi_{LO}t) + \frac{1}{3}\sin(3\varpi_{LO}t) + \cdots\right] \\
&= \frac{4g_m R_L V_{RF}}{\pi}\left[\begin{array}{l}\sin\left((\varpi_{RF} + \varpi_{LO})t\right) - \sin\left((\varpi_{RF} - \varpi_{LO})t\right) \\ + \frac{1}{3}\sin\left((\varpi_{RF} + 3\varpi_{LO})t\right) - \frac{1}{3}\sin\left((\varpi_{RF} - 3\varpi_{LO})t\right) + \cdots\end{array}\right]
\end{aligned}
$$

In this expression for the output voltage, there are no terms at DC or at the LO or RF frequencies so these have all been eliminated. The largest two terms are the LO and RF sum and difference frequencies as desired and then the higher order terms are the sum and difference between the RF signal and the odd harmonics of the LO.

Question 5(b) – Voltage Conversion Gain – 4 marks

$$V_T = \frac{kT}{q} = 25.8mV$$

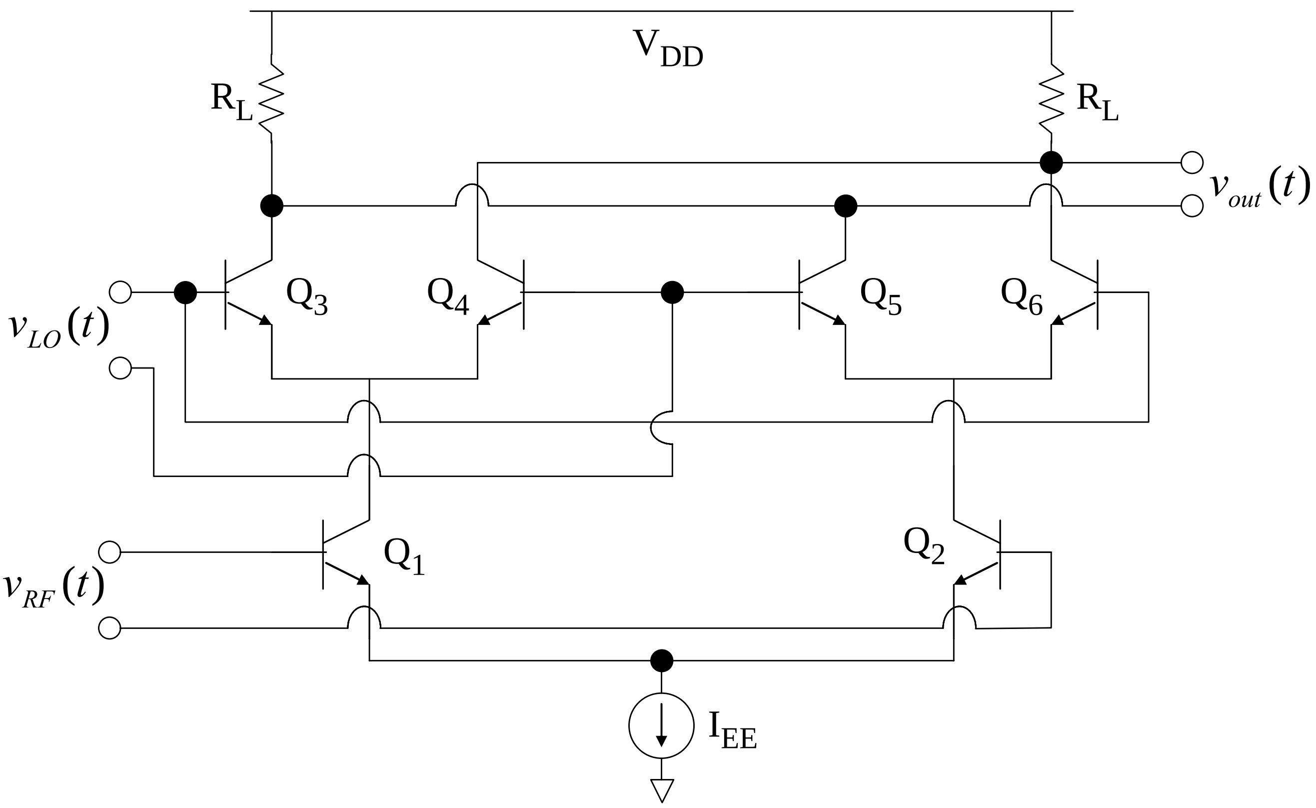
$$g_m = \frac{I_C}{V_T} = 9.675mS$$

$$A_{CF} = \frac{4g_m R_L}{\pi} = 24.65$$

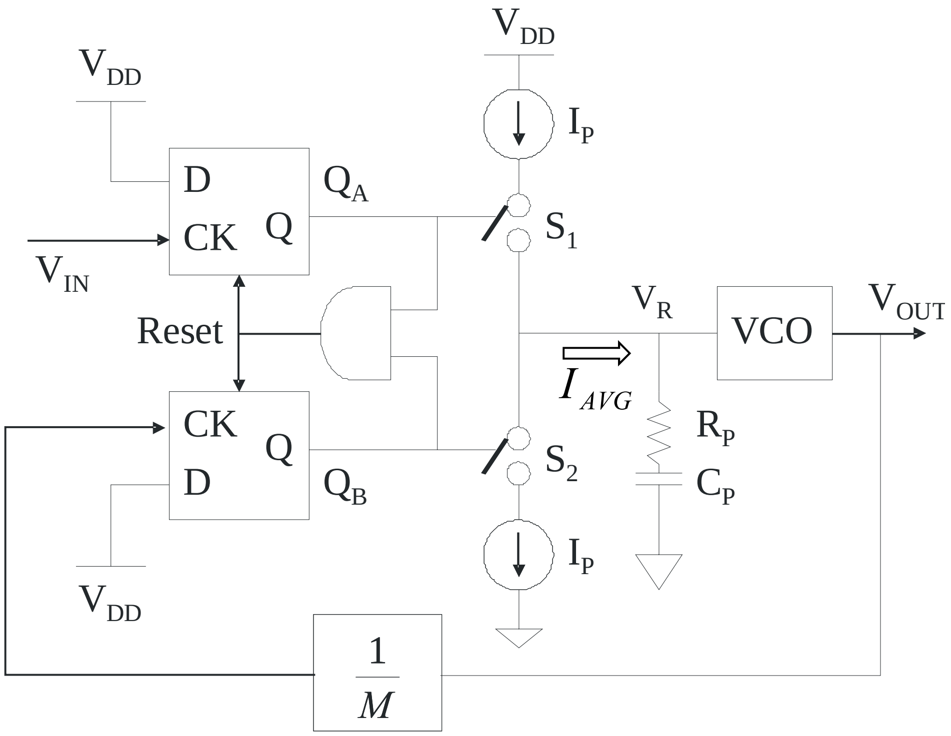
Note: If the total tail current is 0.5mA under small-signal conditions there will be 0.25mA flowing through each transistor in the diff pairs so I_C=0.25mA for the g_m calculation.

Question 5(c) – Gilbert Cell 4 marks

A Gilbert Cell Double Balanced Mixer



Question 6



Q6(a) – 10 marks

If the V_{IN} and V_{OUT} waveforms have the same frequency but a phase offset $\Delta\phi$ then the average current supplied to the loop filter is

$$I_{AVG} = \frac{\Delta\varphi}{2\pi} I_P$$

The impedance of the loop filter to ground is given by:

$$Z = R_P + \frac{1}{C_P s}$$

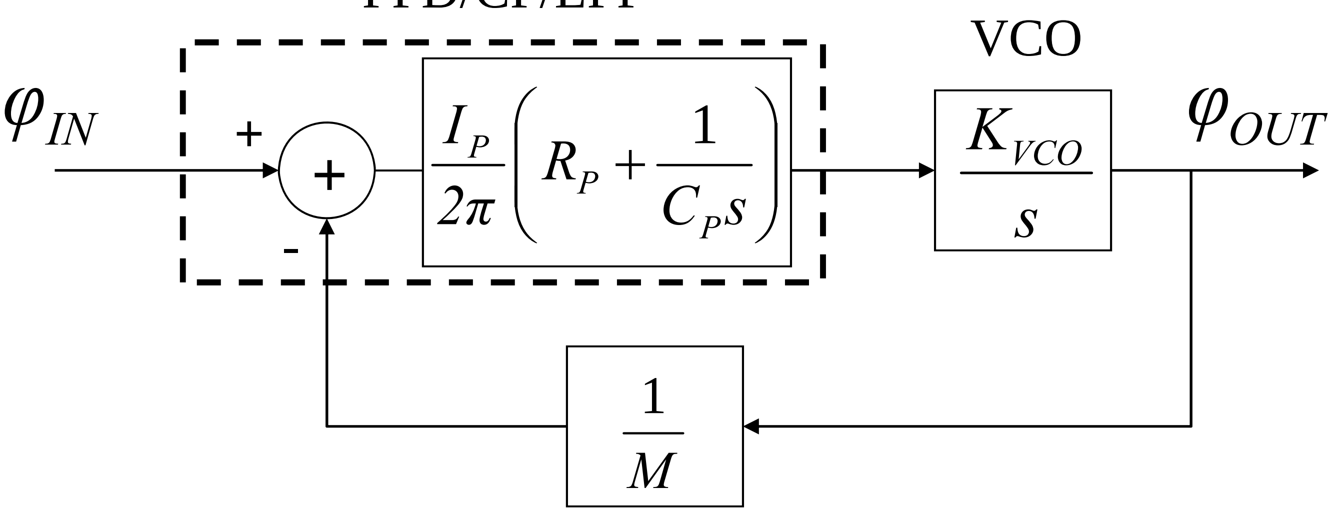
The transfer function of the PFD/CP/LPF can then be approximated by:

$$PFD\_CP\_LPF(s) = \frac{I_P}{2\pi}\left(R_P + \frac{1}{C_P s}\right)$$

The transfer function of the VCO is given by:

$$VCO(s) = \frac{K_{VCO}}{s}$$

Question 6(a)



Open Loop Response

$$H(s)=\frac{I_P K_{VCO}}{2\pi s}\left(R_P+\frac{1}{C_P s}\right)$$

Closed Loop Response

$$\varphi_{OUT}(s)=\left(\varphi_{IN}(s)-\frac{\varphi_{OUT}(s)}{M}\right)H(s)$$

$$\Rightarrow H_{Closed}(s)=\frac{H(s)}{1+\frac{H(s)}{M}}=\frac{\frac{I_P K_{VCO}}{2\pi s}\left(R_P+\frac{1}{C_P s}\right)}{1+\frac{I_P K_{VCO}}{2\pi M s}\left(R_P+\frac{1}{C_P s}\right)}$$

$$=\frac{\frac{I_P K_{VCO}}{2\pi C_P}\left(R_P C_P s+1\right)}{s^2+\frac{I_P K_{VCO}}{2\pi M}R_P s+\frac{I_P K_{VCO}}{2\pi C_P M}}$$

Question 6(b)

$$H_{Closed}(s) = \frac{\frac{I_P K_{VCO}}{2\pi C_P}\left(R_P C_P s + 1\right)}{s^2 + \frac{I_P K_{VCO}}{2\pi M} R_P s + \frac{I_P K_{VCO}}{2\pi C_P M}} \equiv \frac{\frac{I_P K_{VCO}}{2\pi C_P}\left(R_P C_P s + 1\right)}{s^2 + 2\varsigma \omega_n s + \omega_n^2}$$

$$\Rightarrow \omega_n = \sqrt{\frac{I_P K_{VCO}}{2\pi C_P M}} \quad \varsigma = \frac{R_P}{2}\sqrt{\frac{I_P C_P K_{VCO}}{2\pi M}} \quad \tau = \frac{1}{\varsigma \omega_n} = \frac{4\pi M}{I_P R_P K_{VCO}}$$

Using

I_P=3mA, C_P=50pF, R_P=20kΩ, K_{VCO}=200MHz/V, M=250

Note: K_{VCO}=200MHz/V = 1.26 x 10^9 rad/s/V

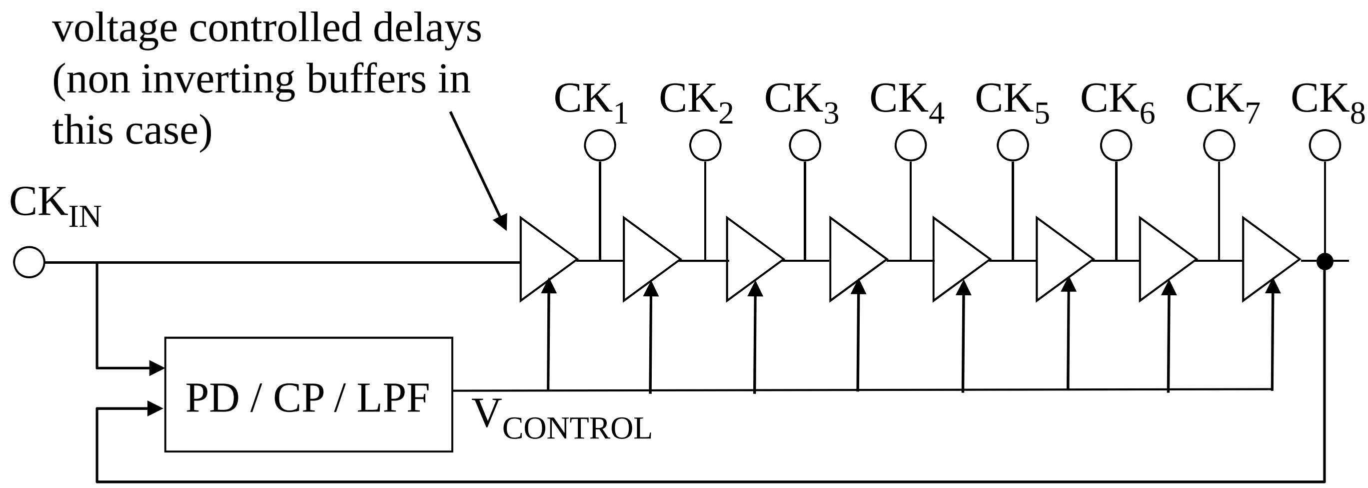
(i) The natural frequency – 2 marks

$$\omega_n = \sqrt{\frac{I_P K_{VCO}}{2\pi C_P M}} = 2.19 \times 10^6 \text{ rad/s} = 349kHz$$

(ii) The damping factor – 2 marks

$$\varsigma = \frac{R_P}{2}\sqrt{\frac{I_P C_P K_{VCO}}{2\pi M}} = 1.1$$

Question 6(c)   A Delay Locked Loop (DLL) – 6 marks



The DLL uses a phase-detector, charge pump and low pass filter similar to a Phase Locked Loop but the output voltage from the low pass filter drives a set of voltage-controlled delay blocks instead of a VCO. Assuming that the delay circuits are identical, then the same time delay is created between each of the clock phases. Voltage controlled delays can be formed from CMOS inverters where the current flowing in the inverters is controlled by an applied voltage in a voltage-controlled current-source type configuration.

Q7 – Essay type question based on Continuous Assessment