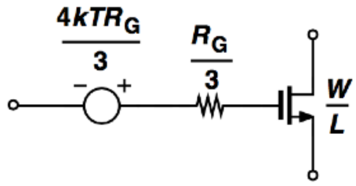
1.

(a). The problem is very similar to Noise Tutorial Problem 1, but here we shall not neglect the gate noise which can be modelled using an additional voltage source at the transistor's input (Razavi, Fig. 2.40c):



The solution without the gate source (check course page for the Tutorial 1, Problem 1 solution at page 3), which includes the two noise source from the transistor channel and load resistor is:

$$\overline{V_{n,in}^2} = \frac{\overline{V_{n,out}^2}}{A_v^2} = \frac{4kT}{g_m^2}(\gamma g_{d0} + \frac{1}{R_L})$$

Then we add the noise contribution from the gate resistance (which is at the input, so not divided by the gain) and get:

$$\boxed{\overline{V_{n,in}^2} = \frac{4kT}{g_m^2}\left(\gamma g_{d0} + \frac{1}{R_L}\right) + 4kT\frac{R_G}{3}}$$

(b). The flicker noise can either be modelled by a current source parallel to the channel noise source, or transferred to a voltage source at the gate input (in series with the gate resistance noise source). The net effect is the same, of course, and the result input-referred noise becomes:

$$\boxed{\overline{V_{n,in}^2} = \frac{4kT}{g_m^2}\left(\gamma g_{d0} + \frac{1}{R_L}\right) + 4kT\frac{R_G}{3} + \frac{K}{WLC_{ox}}\frac{1}{f}}$$

(c). Friis' equation states that "the noise contributed by each stage decreases as the total gain preceding that stage increases, implying that the first few stages in a cascade" (e.g. receiver chain) "are the most critical." (Razavi, just before Example 2.22)

2.

a. See LNA Tutorial Problem 2, pp. 5-6:

$$\boxed{Z_{in} = \frac{V_{in}}{i_{in}} = j\omega\left(L_g + L_s\right) + \frac{1}{j\omega C_{gs}} + \frac{g_m L_s}{C_{gs}}}$$

b. See LNA Tutorial Problem 2, p. 6:

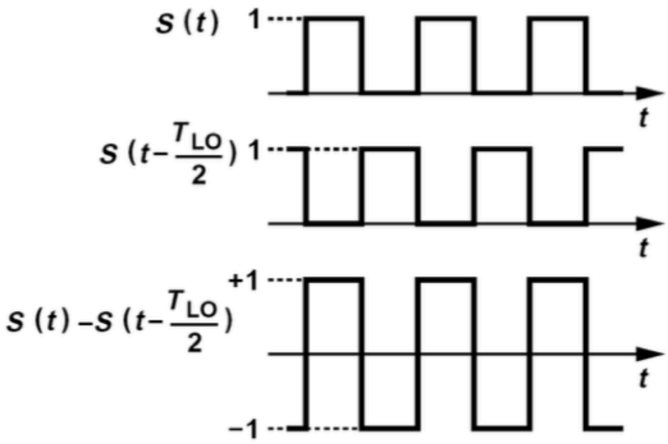
For input matching purpose, the imaginary part of (3) should be zero, which means that L_g + L_s, should be canceled out by C_{gs}. Therefore, at frequency of interest, we have:

$$\omega_o\left(L_g + L_s\right) = \frac{1}{\omega_o C_{gs}} \Rightarrow \omega_o^{\ 2} = \frac{1}{\left(L_g + L_s\right)C_{gs}}$$

And $\dfrac{g_m L_s}{C_{gs}} = R_S = 50\Omega$

3.

(a)



$$V_{out}(t) = I_{RF}R\,[S(t) - S(t - T_{LO}/2)]$$

If $V_{RF} = A_{RF}\cos(\omega_{RF}t)$, then by ignoring the higher order terms:

$$V_{IF} = V_{out} = \frac{4}{\pi}g_{m3}RA_{RF}cos(\omega_{RF}t)cos(\omega_{LO}t) = \frac{2}{\pi}g_{m3}RA_{RF}cos((\omega_{RF} - \omega_{LO})t)$$

Therefore the conversion gain is:

$$\boxed{G_C = \frac{V_{IF}}{A_{RF}} = \frac{2}{\pi}g_{m3}R}$$

(b)

$$\overline{V^2_{n,out,M_3}} = 4kT\gamma g_m R^2$$
$$\overline{V^2_{n,out,R_S}} = 4kTR_S(g_m R)^2$$
$$\overline{V^2_{n,out,R}} = 2 \times 4kTR$$

The noise figure can be written as:

$$NF = \frac{\overline{V^2_{n,out,M_3}} + \overline{V^2_{n,out,R_s}} + \overline{V^2_{n,out,R}}}{G_C^2\overline{V^2_{n,out,R_S}}} = \boxed{\frac{\pi^2}{4}\left(1 + \frac{\gamma}{g_m R_S} + \frac{2}{g_m^2 R_S R}\right)}$$

4.

a. See the Razavi course book, Example 8.14 and Figure 8.26. Here instead we have Q=5 @ 2.45 GHz => Q*(L1+L2)*ω = 154 Ω. (L1 = L2 = 1 nH each!)

gm for the transistors > 154/2 = 77 Ω-1.

b. See the Razavi course book, Example 8.23: -98 dBc/Hz.

5.

Quite similar to Fig 9.30 in the Razavi course book and eq. 9.17 - 9.19, but with M as divider.

The open loop transfer function of the system is:

$$H_o(s) = K_{PFD} Z_{LPF}(s) \frac{K_{VCO}}{s} = \frac{I_o}{2\pi}\left(\frac{1}{sC_p} + R\right)\frac{K_{VCO}}{s}$$

$$H_o(s) = \frac{I_0 K_{VCO}}{2\pi C_p} \frac{1 + sRC_p}{s^2} = k\frac{1 + sRC_p}{s^2} \quad \text{, where } \boxed{k = \frac{I_0 K_{VCO}}{2\pi C_p}}$$

The close-loop transfer function is then:

$$\boxed{H(s)} = \frac{H_o(s)}{1 + \frac{1}{M} H_o(s)} = \boxed{k\frac{1 + sRC_p}{s^2 + sRC_p \frac{k}{M} + \frac{k}{M}}}$$

6.

a.	AB	A is more linear, B more efficient, AB is in-between.
b.	AB	D is a switching amplifier, very nonlinear.
c.	C	When the conduction angle increases, efficiency will be very high with class C, but linearity and output power will drop.
d.	AB	E is about 3.5 times, AB 2 times.