The bias current is \( I_{DQ} = (V_{DD} - V_{DSQ}) / (2 \, \text{k}\Omega) = 2.5 \, \text{mA}. \) This problem is similar to Example 8.3. We have \( K = KP(W/L) / 2 = 10^{-3} \). \( g_m = 2KI_{DQ} = 3.16 \, \text{mS}. \) (This value for \( g_m \) is approximate because the formula was derived assuming \( \lambda = 0 \).) \( r_d \approx 1 / (\lambda I_{DQ}) = 40 \, \text{k}\Omega. \)

\[
R'_L = \frac{1}{\frac{1}{r_d} + \frac{1}{R_D} + \frac{1}{R_L}} = \frac{1}{\frac{1}{40} + \frac{1}{2} + \frac{1}{1}} = 656 \, \Omega
\]

\[
A_{mid} = -g_m R'_L = -(3.16 \times 10^{-3}) \times 656 = -2.07
\]

The frequencies associated with the zero and the poles are

\[
f_z = \frac{g_m}{2\pi C g_d} = 1 \, \text{GHz}
\]

\[
f_{p1} = \frac{1}{2 \pi \left[ C_{gs} R_{sig} + C_{gd} \left( R_{sig} + g_m R'_L R_{sig} + R'_L \right) \right]} = 15.1 \, \text{MHz}
\]

\[
f_{p2} = \frac{C_{gs} R_{sig} + C_{gd} \left( R_{sig} + g_m R'_L R_{sig} + R'_L \right)}{2 \pi \frac{C}{g_m} C_{ds} R'_{sig} R_{sig}} = 2.03 \, \text{GHz}
\]

Because \( f_{p1} \) is much lower than the other two break frequencies, the upper half-power frequency is approximately equal to \( f_{p1} \).

For the simulation we need to know the value of \( V_{bias} \). The equation for the drain current is

\[
I_{DQ} = K(V_{bias} - V_{to})^2 (1 + \lambda V_{DSQ})
\]

\[
2.5 \times 10^{-3} = 10^{-3} (V_{bias} - 1)^2 (1 + 0.01 \times 10)
\]
[Problem 8.14]
\[ B = C_{gs}(R_{sig} + R_L) + C_{gd}R_{sig}(g_m R'_L + 1), \text{ and} \]
\[ A = C_{gs}C_{gd}R_{sig} R'_L \]

(b) Writing current equations at the gate and source we obtain:

\[ \frac{V_g - V_{sig}}{R_{sig}} + sC_{gd}V_g + sC_{qs}(V_g - V_o) = 0 \quad (1) \]

\[ (V_g - V_o)(g_m + sC_{qs}) = V_o/R'_L \quad (2) \]

Solving Equation (2) for \( V_g' \), substituting into Equation (1), and solving for the gain, we eventually obtain:

\[ A(s) = \frac{V_o}{V_{sig}} = \frac{g_m R'_L + sR'_L C_{gs}}{1 + g_m R'_L + B_s + A s^2} \]

in which

\[ B = C_{gs}(R_{sig} + R_L) + C_{gd}R_{sig}(g_m R'_L + 1), \text{ and} \]
\[ A = C_{gs}C_{gd}R_{sig} R'_L \]

(b) Evaluating the gain expression for \( s = 0 \), we obtain the midband gain.

\[ A_{mid} = \frac{g_m R'_L}{1 + g_m R'_L} \]

We have \( K = KP(W/L)/2 = 10^{-3}; g_m = 2\sqrt{K_{IDQ}} = 3.16 \text{ mS}; R_d = \]
(because \( \lambda = 0); R'_L = R_{bias} || R_L = 1 \text{ k}\Omega; \text{ and } A_{mid} = 0.760 \text{ which is equivalent to } -2.39 \text{ dB}. \]

The zero is the root of the numerator of the gain expression.

\[ g_m R'_L + sR'_L C_{gs} = 0 \quad \Rightarrow \quad s_z = -g_m/C_{gs} \]

The corresponding frequency is

\[ f_z = -s_z/2\pi = g_m/(2\pi C_{gs}) = 1.01 \text{ GHz} \]

The poles are the roots of the denominator of \( A(s) \). Evaluating we obtain

\[ 1 + g_m R'_L = 4.16 \]
\[ B = 26.3 \times 10^{-9} \]
\[ A = 2.5 \times 10^{-18} \]
\[ s_{p1} = -161 \times 10^6 \text{ and } s_{p2} = -10.3 \times 10^9 \]
\[ f_{p1} = 25.6 \text{ MHz and } f_{p2} = 1.65 \text{ GHz} \]

Because \( f_z \) and \( f_{p2} \) are much larger than \( f_{p1} \), we expect the half-power frequency to be very nearly equal to \( f_{p1} \). Thus \( f_{3dB} = 25.6 \) MHz.
Problem 8.15

(a) The small-signal equivalent circuit is:

(b) Notice that $V_{gs} = -V_s$. Writing a current equation at the source node we obtain:

$$\frac{V_s - V_{sig}}{R_{sig}} + sC_{gs}V_s + g_mV_s = 0$$

Solving for $V_s$ we obtain

$$V_s = V_{sig} \frac{sC_{gs}R_{sig} + 1 + g_mR_{sig}}{1}$$

Writing a current equation at the drain node we have

$$g_mV_s = sC_{gd}V_o + V_o/R_L$$

Solving for $V_o$ we obtain

$$V_o = V_s \frac{g_mR_L'}{sC_{gd}R_L' + 1}$$
Using Equation (1) to substitute for \( V_s \) in Equation (2) and dividing both sides by \( V_{\text{sig}} \) we have

\[
A(s) = \frac{V_o}{V_{\text{sig}}} = \frac{g_m R'_L}{(sC_{gsR_{\text{sig}}} + 1 + g_m R_{\text{sig}})(sC_{gdR_{\text{sig}}} + 1)}
\]

(c) The poles are the roots of the denominator of \( A(s) \).

\[
s_{p1} = -\left(1 + g_m R_{\text{sig}}\right)/(C_{gsR_{\text{sig}}}^') \quad \text{and} \quad s_{p2} = -1/(C_{gdR_{\text{sig}}}^')
\]

The break frequencies associated with these poles are

\[
f_{p1} = (1 + g_m R_{\text{sig}})/(2\pi C_{gsR_{\text{sig}}}^') \quad \text{and} \quad f_{p2} = 1/(2\pi C_{gdR_{\text{sig}}}^')
\]

(d) The midband gain is obtained by evaluating \( A(s) \) for \( s = 0 \).

\[
A_{\text{mid}} = \frac{g_m R'_L}{1 + g_m R_{\text{sig}}}
\]

(e) We have \( K = KP(W/L)/2 = 10^{-3}; g_m = 2Kl_D Q = 3.16 \text{ mS}; r_d = (\text{because } \lambda = 0); R'_L = R_D||R_L = 1 \text{ k}\Omega; \) and \( A_{\text{mid}} = 2.40 \) which is equivalent to 7.6 dB. The break frequencies are \( f_{p1} = 4.19 \text{ GHz} \) and \( f_{p2} = 318 \text{ MHz} \). Because \( f_{p1} \) is considerably greater than \( f_{p2} \), the upper half-power frequency is approximately equal to \( f_{p2} \).
Problem 8.20

Because the amplifiers are ideal voltage amplifiers, their input impedances are infinite, and the input impedance of each circuit is the Miller impedance. From Equation 8.36, we have

\[ Z_{\text{in, Miller}} = \frac{Z_f}{(1 - A_v)} \]. Evaluating, we obtain

(a) \( Z_{\text{in}} = 909 \, \Omega \)  \quad (b) \( Z_{\text{in}} = 99.0 \, \Omega \)

(c) \( Z_{\text{in}} = \infty \)  \quad (d) \( Z_{\text{in}} = -10 \, \text{k}\Omega \)
The equivalent circuits are shown in Figure 8.26 in the book, except that for the circuit under consideration we have \( R'_L = R_D || r_d || R_L \). The midband voltage gain is \( A_{\text{mid}} = -g_m R'_L \).

The bias current is \( I_{DQ} = \frac{(V_{DD} - V_{DSQ})}{(2 \, \text{k}\Omega)} = 2.5 \, \text{mA} \). We have \( K = KH(W/L)/2 = 10^{-3} \). \( g_m = 2KI_{DQ} = 3.16 \, \text{mS} \). (This value for \( g_m \) is approximate because the formula was derived assuming \( \lambda = 0 \).)

Also we have \( r_d = 1/(\lambda I_{DQ}) = 40 \, \text{k}\Omega \).

\[
R'_L = \frac{1}{1/r_d + 1/R_D + 1/R_L} = \frac{1}{1/40 + 1/2 + 1/1} = 656 \, \Omega
\]

\[
A_{\text{mid}} = -g_m R'_L = -(3.16 \times 10^{-3}) \times 656 = -2.07
\]

Then the Miller capacitance is

\[
C_{\text{Miller}} = C_{gq}(1 - A_{\text{mid}}) = (0.5 \, \text{pF})(1 + 2.07) = 1.54 \, \text{pF}
\]

\[
C_{\text{total}} = C_{gs} + C_{\text{Miller}} = 2.04 \, \text{pF}
\]

\[
f_b = \frac{1}{2\pi R_{\text{sig}} C_{\text{total}}} = 15.6 \, \text{MHz}
\]

In Problem 8.12 we did an exact analysis of this circuit and obtained \( f_b = 15.1 \, \text{MHz} \). (The simulation yielded \( f_b = 14.8 \, \text{MHz} \).)

![Problem 8.42A](image-url)
[Problem 8.49 A]
Problem 8.49 B
Problem 8.59

We elected to cascade three stages. First an emitter-follower to provide high input impedance and low source impedance for the second stage. The second stage is a common-emitter with a partially unbypassed emitter resistor to provide the desired voltage gain. The final stage is an emitter follower providing the low output impedance needed to drive the load capacitance at high frequencies. To reduce the number of components required we use direct coupling. The circuit is: