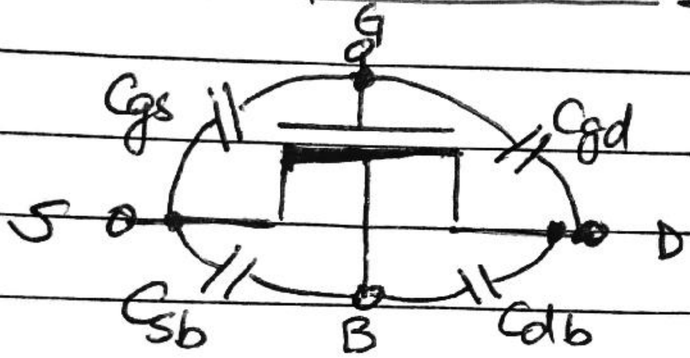
FREQUENCY RESPONSE OF TRANSISTOR CKTS

* High-Frequency (HF) MOS Small-Signal (SS) model.

HF SS Model = LF SS Model + Parasitic Caps.

Parasitic Caps in a MOS in Saturation



where

$$C_{gs} = \frac{2}{3} WL C_{ox} + C_{ov}$$

C_{ov} is the overlap capacitance : $L_{ov} \cdot W \cdot C_{ox}$

$$C_{gd} \approx C_{ov}$$

The drain-to-bulk (C_{db}) and the source-to-bulk (C_{sb}) capacitances are depletion capacitance due to reverse bias diodes.

$$C_{sb} = (A_s + WL) C_{js} + P_s C_{jsw}$$

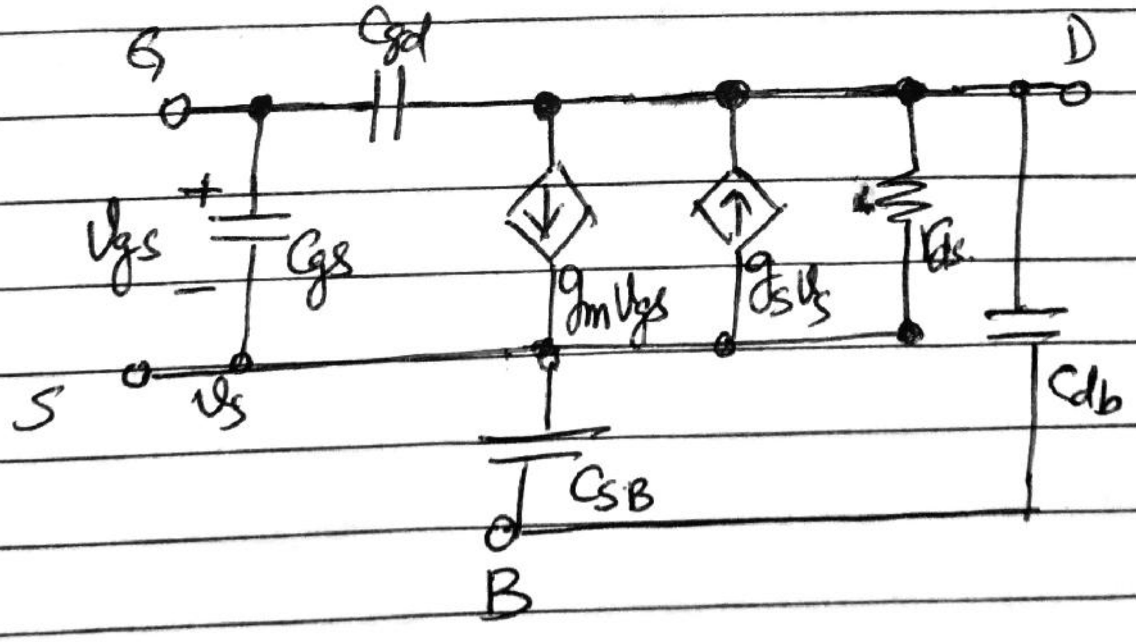
where A_s is the area of source junction, P_s is the effective perimeter of the source junction and C_{js} & C_{j-sw} are the area and perimeter capacitance density.

<u>NOTE:</u> C_{sb} includes the channel area because the channel is connected to the source junction and there is depletion underneath.

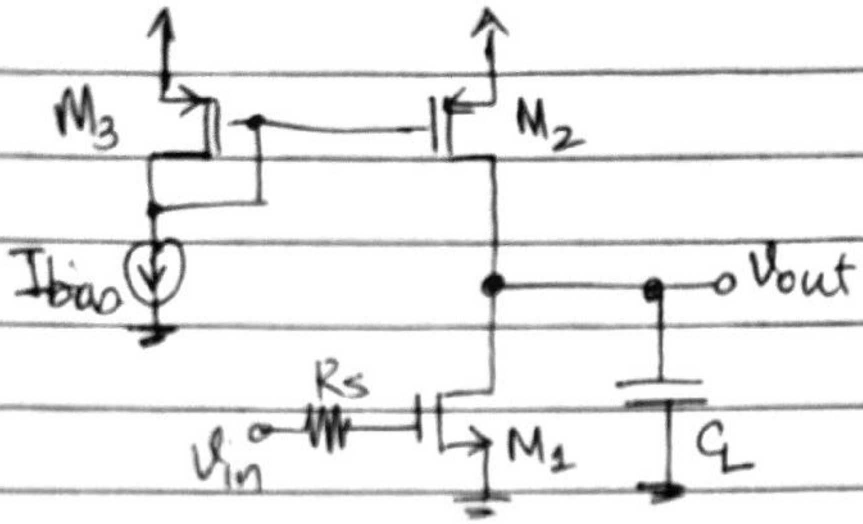
Similarly,

$$C_{db} = A_d C_{jd} + P_d C_{j-sw}$$

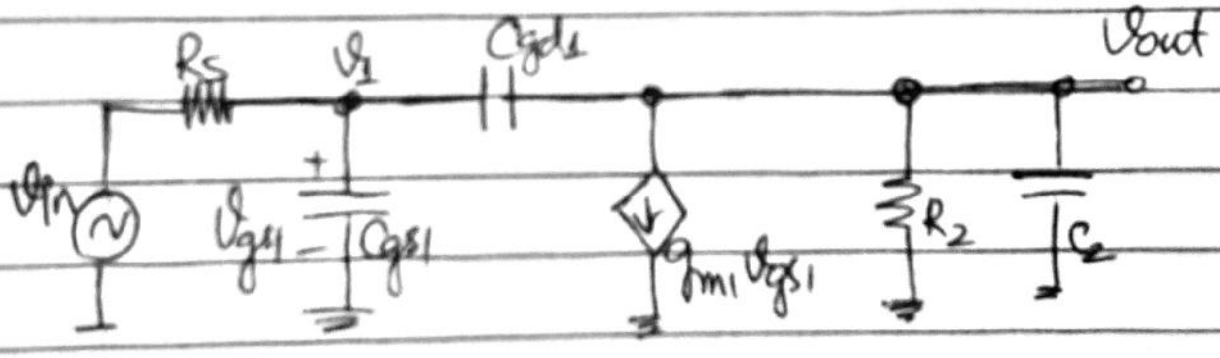
<u>The HF model in the active region:</u>



FREQUENCY RESPONSE OF A COMMON-SOURCE AMPLIFIER



HF small-signal model



$$C_2 = C_{db1} + C_{db2} + C_{gd2} + C_L$$

$$R_2 = r_{ds1} \parallel r_{ds2}$$

Now we can ~~[illegible]~~ apply node equations as we did in resistive network by substituting the capacitors with its equivalent Laplace impedance

KCL @ V_1

$$V_1 \left(G_s + sC_{gs1} + sC_{gd1}\right) - v_{in} G_s - v_{out}\, s C_{gd1} = 0 \quad \text{—— (1)}$$

KCL @ v_{out}

$$v_{out} \left(G_2 + sC_{gd1} + sC_2\right) - V_1 s C_{gd1} + g_{m1} V_1 = 0 \quad \text{—— (2)}$$

where $v_1 = v_{gs1}$

On solving (1) & (2) we get

$$A(s) = \frac{v_{out}}{v_{in}} = \frac{-g_{m1} R_2 \left(1 - s \frac{C_{gd1}}{g_{m1}}\right)}{1 + sa + s^2 b}$$

$$a = R_s \left[C_{gs1} + C_{gd1}\left(1 + g_{m1} R_2\right)\right] + R_2\left(C_{gd1} + C_2\right)$$

&

$$b = R_s R_2 \left(C_{gd1} C_{gs1} + C_{gs1} C_2 + C_{gd1} C_2\right)$$

First sanity check: for $s = 0$ the result should be that of LF model.

When $s = 0$ $A(s) = -g_{m1} R_2$ ← confirmed.

Since the denominator $D(s)$ is second-order, it has two poles. and if assume they are far apart i.e. $\omega_{p1} \ll \omega_{p2}$, the denominator can be expressed as

$$D(s) = \left(1 + \frac{s}{\omega_{p1}}\right)\left(1 + \frac{s}{\omega_{p2}}\right) \cong 1 + \frac{s}{\omega_{p1}} + \frac{s^2}{\omega_{p1}\omega_{p2}}$$

then

$$\omega_{p1} \cong \frac{1}{a} = \frac{1}{R_S\left[C_{gs1} + \underbrace{C_{gd1}(1 + g_m R_2)}_{\text{Miller Cap.}}\right] + R_2 \underbrace{(C_{gd1} + C_2)}_{\text{Typically dominant.}}}$$

$$\omega_{p2} \cong \frac{1}{\omega_{p1} b} \cong \frac{g_{m1} C_{gd1}}{C_{gs1}C_{gd1} + C_{gs1}C_2 + C_{gd1}C_2}$$

And the high frequency zero

$$\omega_z = -g_{m1}/C_{gd1}$$

Since $\omega_{p1} \ll \omega_{p2}, \omega_z$, a dominant-pole approximation may be applied for frequencies $\omega \ll \omega_{p2}, \omega_z$

$$A(s) \cong \frac{A_0}{1 + \frac{s}{\omega_{p1}}} = \frac{-g_m R_2}{1 + s/\omega_{p1}}$$

For special case where the load capacitance and therefor C_2 is large;

$$\omega_{p1} \cong \frac{1}{R_2 C_2}$$

$$\omega_{p2} \cong \frac{C_{gd1}}{C_{gs1} + C_{gd1}} \cdot \frac{g_{m1}}{C_2}$$

BODE PLOTS

Once the s-domain transfer function is formulated, plotting the transfer function versus frequency is an essential analytic tool.

If we consider the transfer function of the previous example with two widely separated poles:

$\omega_{p1} << \omega_{p2}$ | real poles.

$$A(s) = \frac{A_0}{(1 + s/\omega_{p1})(1 + s/\omega_{p2})}$$

where A_0 is the DC ($\omega = 0$) gain

ie $A_0 = -g_m R_2$

Since $A(s)$ is a complex number, the transfer function is analyzed by plotting both Magnitude $|A(j\omega)|$ & the Phase $\angle A(j\omega)$.

With the assumption of widely separated poles, the magnitude & phase can be plotted for different ~~fre~~ frequency regions.

(1) For $\omega \ll \omega_{p1}$, $A(j\omega) = -A_0$

$\therefore A(j\omega) \cong -g_m R_2$

or $|A(\omega)| = -g_m R_2$ OR $g_m R_2$

& $\angle A(\omega) = 0^\circ$ 180°

(2) At $\omega = \omega_{p1}$, $|A(\omega)| = \dfrac{g_m R_2}{\sqrt{2}}$

$\angle A(\omega) = -180^\circ - 45^\circ$

(3) For $\omega_{p1} \ll \omega \ll \omega_{p2}$

$$|A(\omega)| \cong \frac{g_m R_2 \cdot \omega_{p1}}{\omega}$$

If we express the magnitude in decibels (dB)

$$|A(\omega)|_{dB} \cong 20 \log (g_m R_2 \cdot \omega_{p1}) - 20 \log (\omega)$$

Thus the magnitude response is decreasing -20 dB for every decade increase in ω or known as <u>20 dB-per-decade</u>

In addition, in this region, we have $\angle A(\omega) \simeq -180^\circ - 90^\circ = -270^\circ$.

(4) $\omega = \omega_{p2}$

$$|A(\omega_{p2})| = \frac{g_m R_2 \cdot \omega_{p1}}{\sqrt{2} \cdot \omega_{p2}}$$

$$\& \angle A(\omega_{p2}) = -180^\circ - 135^\circ$$

(5) Finally for $\omega \gg \omega_{p2}$

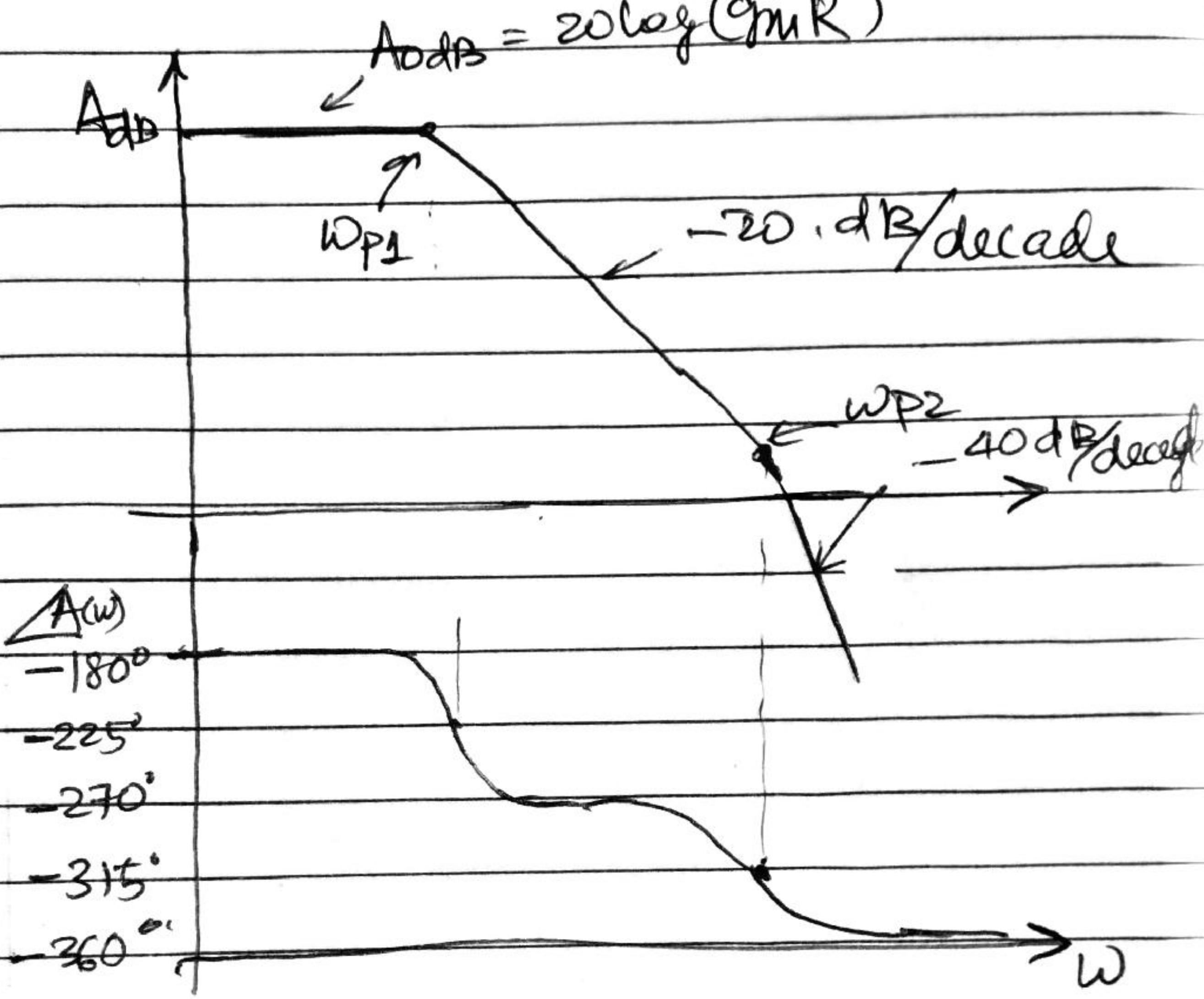
$$|A(\omega)| = \frac{g_m R_2 \cdot \omega_{p1} \cdot \omega_{p2}}{\omega^2}$$

In dB, $|A(\omega)|_{dB} \simeq 20 \log(g_m R_2 \cdot \omega_{p1} \cdot \omega_{p2}) - 40 \log(\omega)$

$\Rightarrow$ 40 dB-per-decade fall

$\& \angle A(\omega) = -180^\circ - 180^\circ$

THE ~~B~~ PLOT



AodB = 20log(gmR)
AdB
ωp1
-20 dB/decade
ωp2
-40 dB/decade
∠A(ω)
-180°
-225°
-270°
-315°
-360°
ω