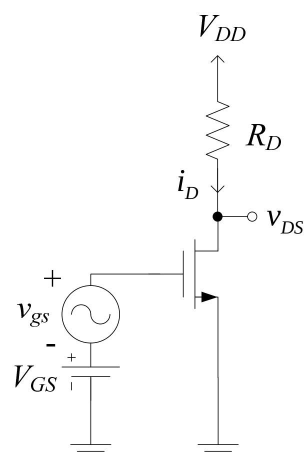
Lecture 28: MOSFET as an Amplifier. Small-Signal Equivalent Circuit Models.

As with the BJT, we can use MOSFETs as AC small-signal amplifiers. An example is the so-called conceptual MOSFET amplifier shown in Fig. 7.2:



(Fig. 7.2)

This is only a "conceptual" amplifier for two primary reasons:

1. The bias with V_{GS} is impractical. (Will consider others later.)
2. In ICs, resistors take up too much room. (Would use another triode-region biased MOSFET in lieu of R_D.)

To operate as a small-signal amplifier, we bias the MOSFET in the saturation region. For the analysis of the DC operating point, we set $v_{gs} = 0$ so that from (7.25) with $\lambda = 0$

$$i_D = \frac{1}{2} k_n' \frac{W}{L} \left(v_{GS} - V_t \right)^2 \qquad (7.25),(1)$$

From the circuit $$V_{DS} = V_{DD} - I_D R_D \qquad (7.26),(2)$$

For operation in the saturation region

$$v_{GD} \le V_t \quad \Rightarrow \quad v_{GS} - v_{DS} \le V_t$$

or

$$v_{DS} \ge v_{GS} - V_t \tag{3}$$

where the total drain-to-source voltage is

$$v_{DS} = \underbrace{V_{DS}}_{\text{bias}} + \underbrace{v_{ds}}_{\text{AC}}$$

Similar to what we saw with BJT amplifiers, we need make sure that (3) is satisfied for the entire signal swing of v_{ds}.

With an AC signal applied at the gate

$$v_{GS} = V_{GS} + v_{gs} \qquad (7.27),(4)$$

Substituting (4) into (1)

$$i_D = \frac{1}{2} k_n' \frac{W}{L} \left(V_{GS} + v_{gs} - V_t \right)^2 = \frac{1}{2} k_n' \frac{W}{L} \left[\left(V_{GS} - V_t \right) + v_{gs} \right]^2 \tag{5}$$

$$= \underbrace{\frac{1}{2} k_n' \frac{W}{L} \left(V_{GS} - V_t \right)^2}_{= I_D \text{ (DC)}} + \underbrace{\frac{2}{2} k_n' \frac{W}{L} \left(V_{GS} - V_t \right) v_{gs} + \frac{1}{2} k_n' \frac{W}{L} v_{gs}^2}_{\text{(time varying)}} \qquad (7.28),(6)$$

The last term in (6) is nonlinear in v_{gs}, which is undesirable for a linear amplifier. Consequently, for linear operation we will require that the last term be "small":

$$\frac{1}{2} k_n' \frac{W}{L} v_{gs}^2 << k_n' \frac{W}{L} \left(V_{GS} - V_t \right) v_{gs}$$

or

$$\underline{v_{gs} << 2 \left(V_{GS} - V_t \right)} \qquad (7.29),(7)$$

If this small-signal condition (7) is satisfied, then from (6) the total drain current is approximately the linear summation

$$i_D \approx \underbrace{I_D}_{\text{DC}} + \underbrace{i_d}_{\text{AC}} \qquad (7.31),(8)$$

where

$$i_d = k_n' \frac{W}{L}\left(V_{GS} - V_t\right) v_{gs}. \qquad (9)$$

From this expression (9) we see that the AC drain current i_d is related to v_{gs} by the so-called transistor transconductance, g_m:

$$\boxed{g_m \equiv \frac{i_d}{v_{gs}} = k_n' \frac{W}{L}\left(V_{GS} - V_t\right) \text{ [S]}} \qquad (7.32),(10)$$

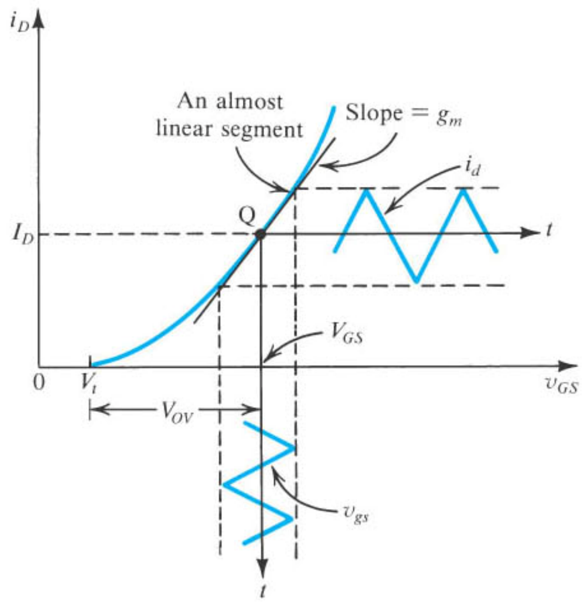
which is sometimes expressed in terms of the overdrive voltage $V_{OV} \equiv V_{GS} - V_t$

$$\boxed{g_m = k_n' \frac{W}{L} V_{OV} \text{ [S]}} \qquad (7.33),(11)$$

Because of the V_{GS} term in (10) and (11), this g_m depends on the bias, which is just like a BJT.

Physically, this transconductance g_m equals the slope of the i_D-v_{GS} characteristic curve at the Q point:

$$g_m \equiv \left.\frac{\partial i_D}{\partial v_{GS}}\right|_{v_{GS} = V_{GS}} \qquad (7.34),(12)$$



(Fig. 7.11)

Lastly, it can be easily show that for this conceptual amplifier in Fig. 7.2,

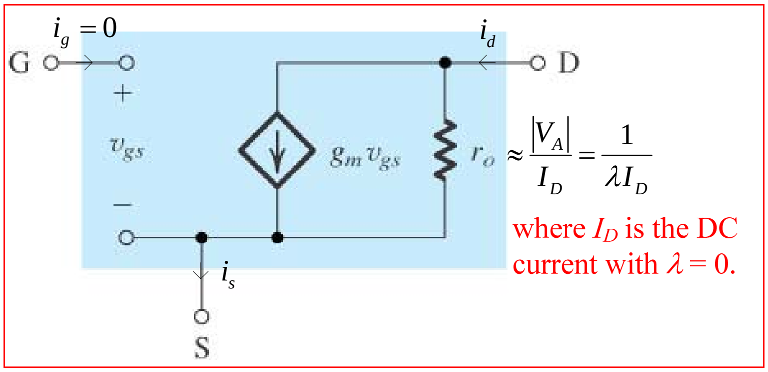
$$A_v \equiv \frac{v_{ds}}{v_{gs}} = -g_m R_D \qquad (7.36),(13)$$

Consequently, $A_v \propto g_m$, which is the same result we found for a similar BJT conceptual amplifier [see (7.78)].

MOSFET Small-Signal Equivalent Models

For circuit analysis, it is convenient to use equivalent small-signal models for MOSFETs – as it was with BJTs.

In the saturation mode, the MOSFET acts as a voltage controlled current source. The control voltage is v_{gs} and the output current is i_d, which gives rise to this small-signal π model:



(Fig. 7.13b)

Things to note from this small-signal model include:

1. $i_g = 0$ and $v_{gs} \neq 0 \Rightarrow$ infinite input impedance.
2. r_o models the finite output resistance. Practically speaking, it will range from $\approx 10\text{ k}\Omega \to 1\text{ M}\Omega$. Note that it depends on the bias current I_D.
3. From (10) we found

$$g_m = k_n' \frac{W}{L}(V_{GS} - V_t) \tag{14}$$

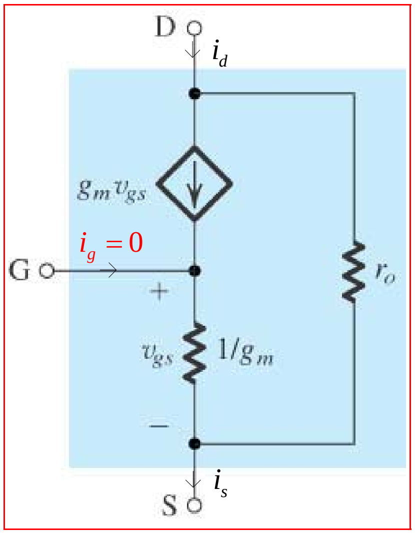
Alternatively, it can be shown that

$$g_m = \frac{2I_D}{V_{OV}} = \frac{I_D}{(V_{GS} - V_t)/2} \quad (7.42),(15)$$

which is similar to $g_m = I_C / V_T$ for BJTs.

One big difference from BJTs is $V_T \approx 25$ mV while $V_{eff} = 0.1$ V or greater. Hence, for the same bias current g_m is much larger for BJTs than for MOSFETs.

A small-signal T model for the MOSFET is shown in Fig. 7.17:



(Fig. 7.17a)

Notice the direct connection between the gate and both the dependent current source and $1/g_m$. While this model is correct, we've **added the explicit boundary condition that $i_g = 0$** to this small-signal model.

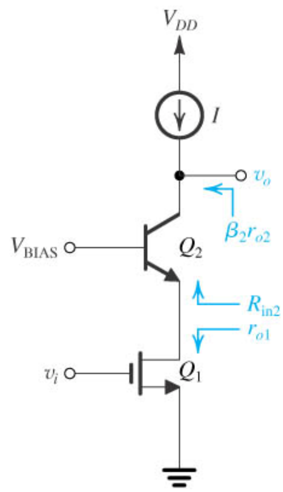
It isn't necessary to do this because the currents in the two vertical branches are both equal to $g_m v_{gs}$, which means $i_g = 0$. But adding this condition $i_g = 0$ to the small-signal model in Fig. 7.17a makes this explicit in the circuit calculations. (The T model usually shows this direct connection while the π model usually doesn't.)

MOSFETs have many advantages over BJTs including:

1. High input resistance
2. Small physical size
3. Low power dissipation

4. Relative ease of fabrication.

One can combine advantages of both technologies (BJT and MOSFET) into what are called BiCMOS amplifiers:



(Sedra and Smith, 5th ed.)

Such a combination provides a very large input resistance from the MOSFET and a large output impedance from the BJT.