ECE 2100 Lecture Notes 3/31/03

A. Stolp
3/30/03

Stuff

Micron Scholarship deadline extended to M, 3/31 (today).
$3000 (Two years tuition)

Exam 3 Monday 4/7/03

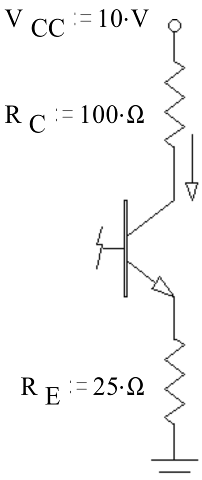
Primarily Ch.4 (BJTs), but may include any earlier material.
Old exams are available on the HW web page.

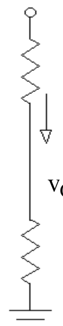
HW #19, due: W, 4/4 Hw 19 handout

HW #20, due: M, 4/7 Ex4.45, Ex4.46
Problems 4.120, 4.119, The transistor of HW 19, problem 1 has C_π = 11pF, C_μ = 2pF & 1pF of stray capacitance between the base & collector, find f_{CH} due to these.

Spice #S2, due: F, 4/4 2 handouts

Load lines

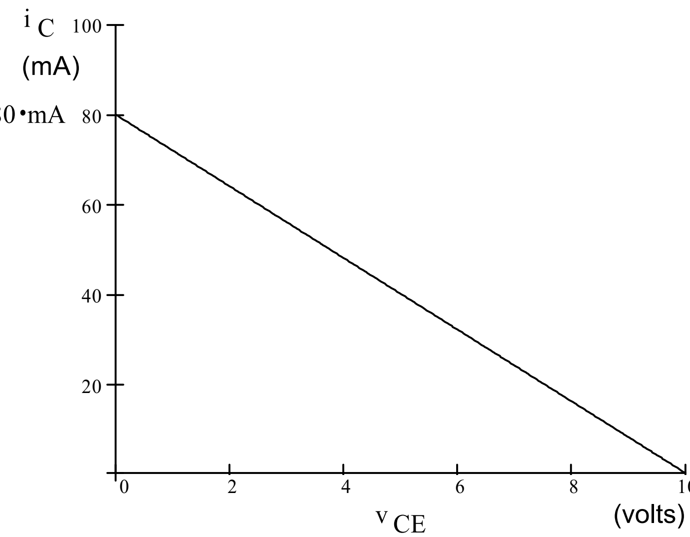


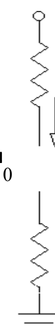


ideal saturation

$$i_C = \frac{V_{CC}}{R_C + R_E} = 80 \cdot mA$$

$v_{CE} = V_{CC}$



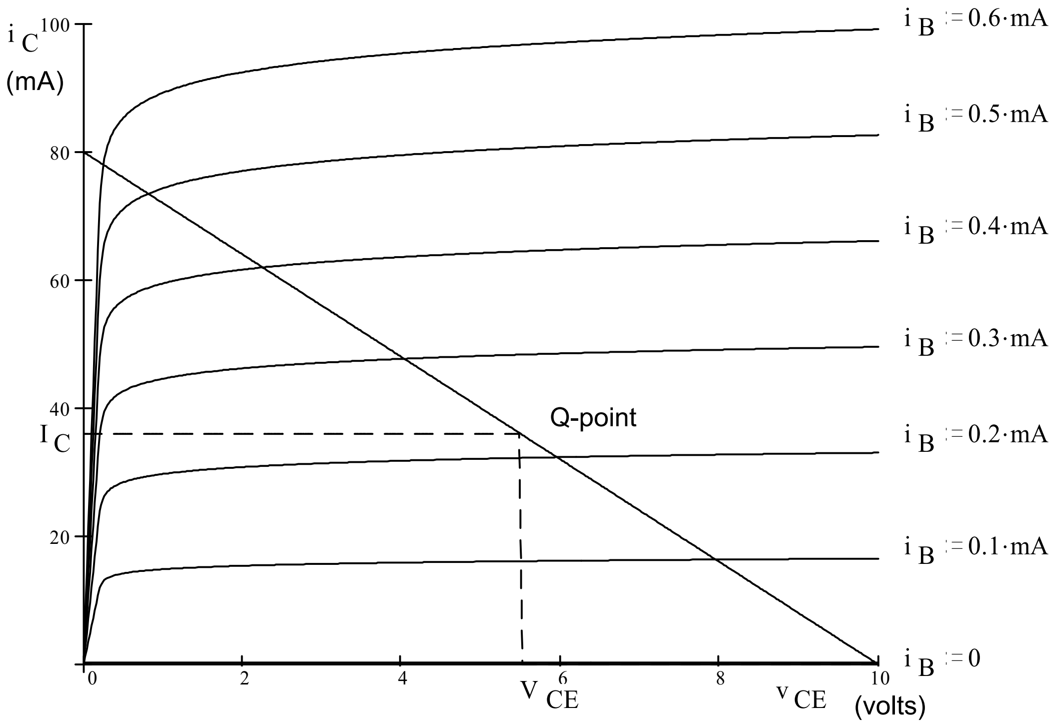


cutoff

$i_C = 0$

$v_{CE} = V_{CC}$

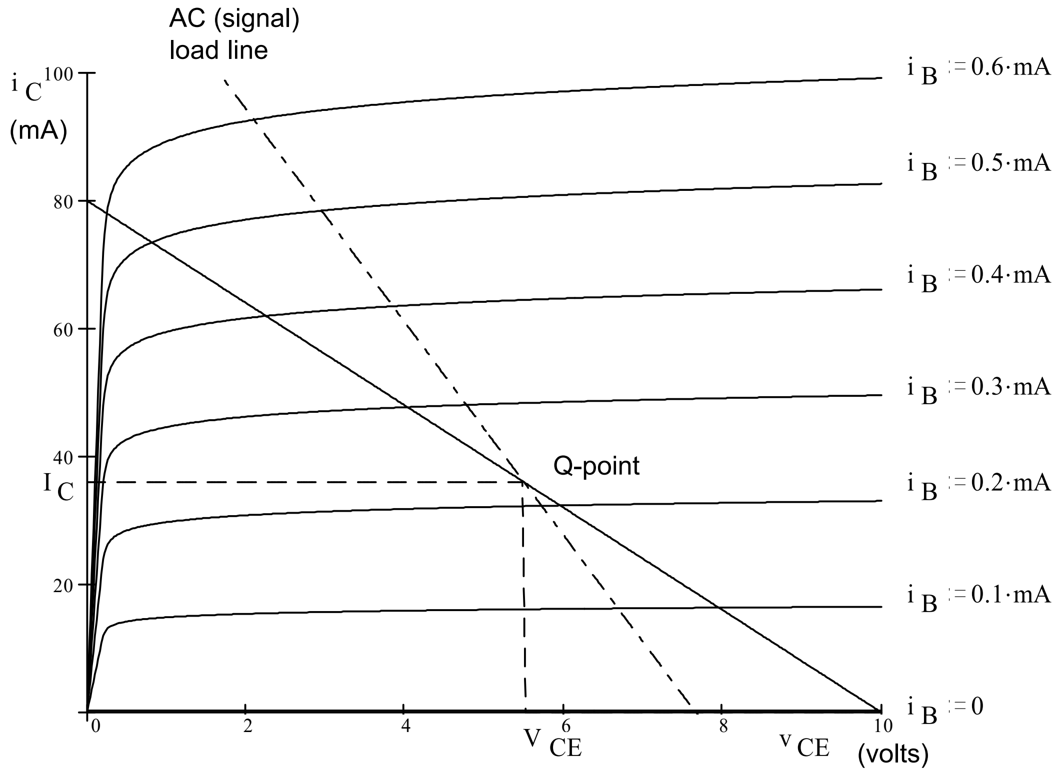
The load line is typically drawn right on top of the transistor's characteristic curves



The slope of the load line is $-\frac{1}{R_C + R_E}$

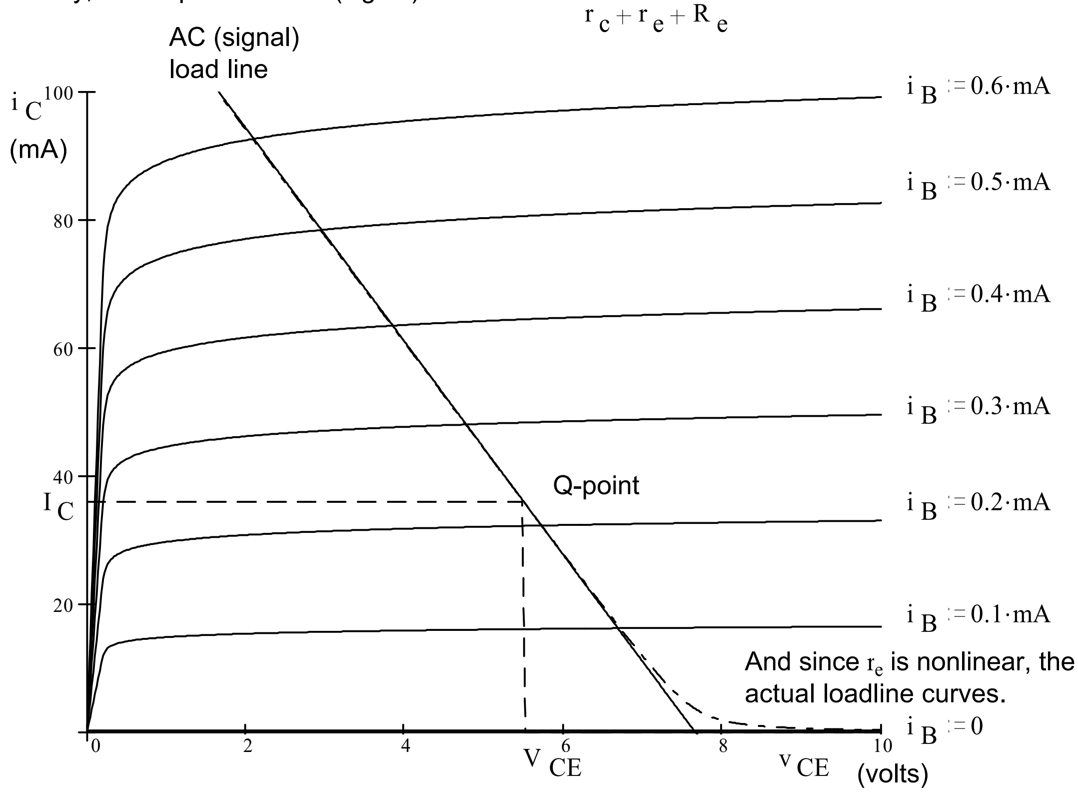
You can also draw a load line for the signal

The slope of the AC (signal) load line is $-\dfrac{1}{r_c + R_e}$



proportional to L- proportional to L+

Actually, the slope of the AC (signal) load line is $-\dfrac{1}{r_c + r_e + R_e}$

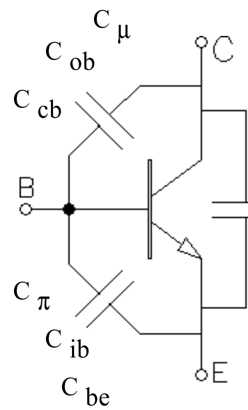


Similar analysis can be used with diode curves (see p157 & p273 in your textbook.
Load lines can be helpful in visualizing the output limits and the optimal Q-point, but that's about all.

High-frequency response

In general, capacitors that are placed in the circuit intentionally, those you can see, cause low-frequency poles. The unseen capacitors inside the parts and between the leads and the board traces cause high-frequency poles. These unseen capacitors have many names, Your textbook uses C_μ and C_π

This capacitance causes the most trouble in common-emitter amplifiers because of it's location. It is connected between the input an the output, so it's effects are multiplied by the voltage gain. (The miller effect.)

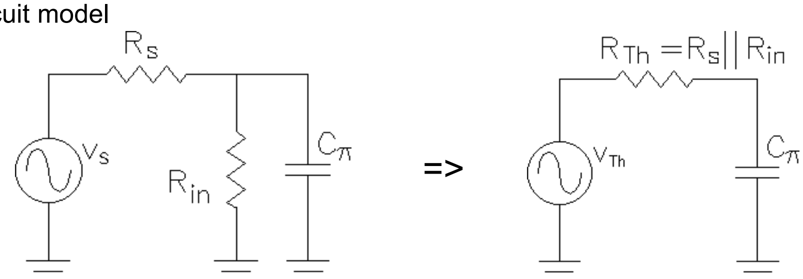


This cacacitance is usually small and unimportant.

This capacitance varies with the base current so much that it is not even specified on data sheets. f_T is given instead. f_T is the frequency where so much current flows through C_π that the effective β is reduced to 1.

$$f_T = \frac{1}{2\cdot\pi\cdot\left(C_\pi + C_\mu\right)\cdot r_e} \quad \text{= freq. where } \beta \underset{-}{\sim} 1$$

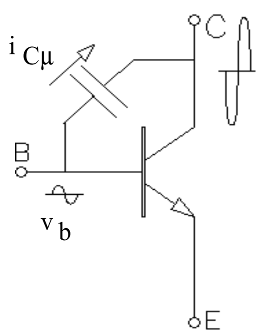
Input circuit model



$$f_{CH1} = \frac{1}{2\cdot\pi\cdot C_\pi}\cdot\left(\frac{1}{R_s} + \frac{1}{R_{in}}\right)$$

Miller Effect

In a common-emitter amplifier:



$$v_c = -\left|A_v\right|\cdot v_b$$

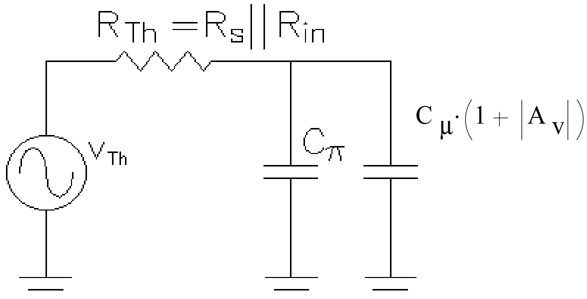
$$i_{C\mu} = \frac{v_b - v_c}{\left(\frac{1}{j\cdot\omega\cdot C_\mu}\right)} = \left[v_b - \left(-\left|A_v\right|\cdot v_b\right)\right]\cdot\left(j\cdot\omega\cdot C_\mu\right)$$

$$= v_b\cdot\left(1 + \left|A_v\right|\right)\cdot\left(j\cdot\omega\cdot C_\mu\right)$$

$$= v_b\cdot j\cdot\omega\cdot C_\mu\cdot\left(1 + \left|A_v\right|\right)$$

If you wanted to make an equivalent amount of current flow to ground, you'd need a capacitor that was $(1+A_v)$ times as big. This is the Miller effect.

Input circuit model



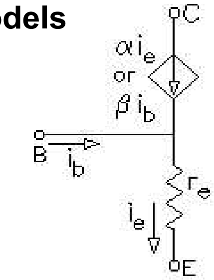
$$f_{CH} = \frac{1}{2\cdot\pi\left[C_\pi + C_\mu\cdot\left(1 + \left|A_v\right|\right)\right]}\cdot\left(\frac{1}{R_s} + \frac{1}{R_{in}}\right)$$

The Miller effect will amplify any capacitance between the base and the collector, not just the capacitance within the transistor, so place leads and circuit traces carefully. If you're modeling a circuit in SPICE you'll have to model these "stray" capacitances if you want your high-frequency results to be any good.

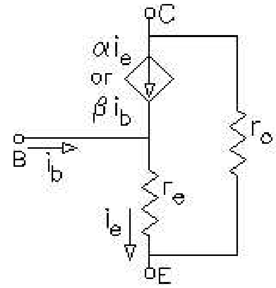
BJT transistor small-signal models

"T" Model p261, p286, p289, p291

We've been using these models in one form or another all along, without really naming them

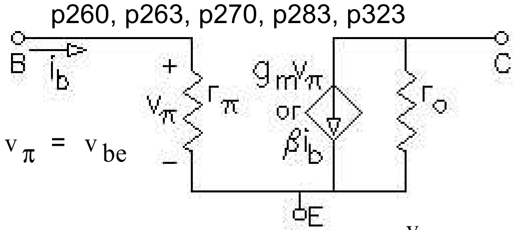


with r_o

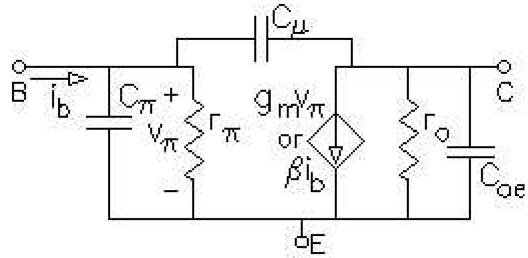


Hybrid-π model

p260, p263, p270, p283, p323



$v_\pi = v_{be}$



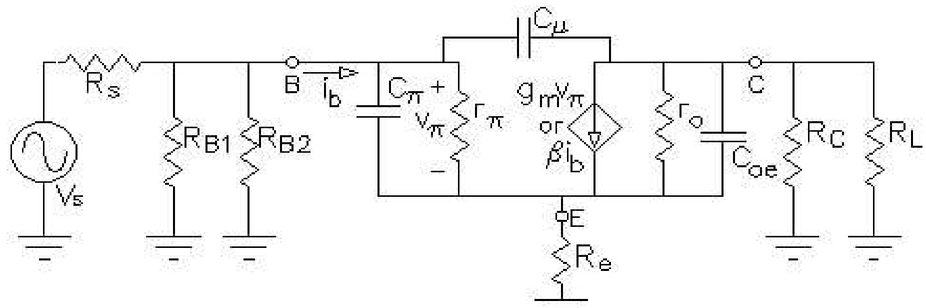
with internal capacitances, p323

$$r_\pi = (\beta + 1)\cdot r_e \qquad i_b = \frac{v_\pi}{r_\pi}$$

$$\beta\cdot i_b = \beta\cdot\frac{v_\pi}{r_\pi} = \beta\cdot\frac{v_\pi}{(\beta+1)\cdot r_e}$$

$$g_m = \frac{\beta}{(\beta+1)\cdot r_e} = \frac{\alpha}{r_e} \simeq \frac{1}{r_e}$$ = transconductance

In a circuit:



Common Base p288

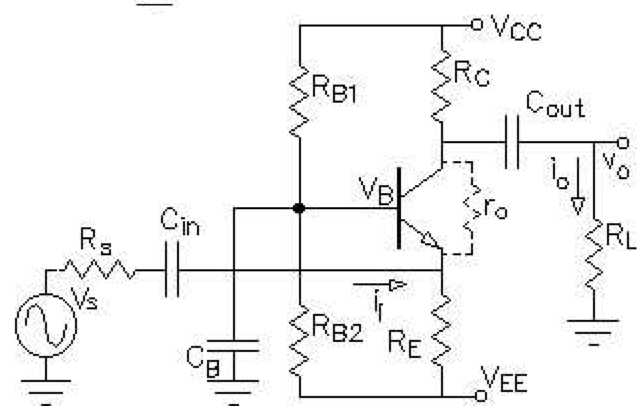
Bias just like a common emitter, only the base is hooked to signal-common and the signal is fed in at the emitter.

Voltage gain is similar to the common emitter: $\dfrac{r_c}{r_e + R_E}$

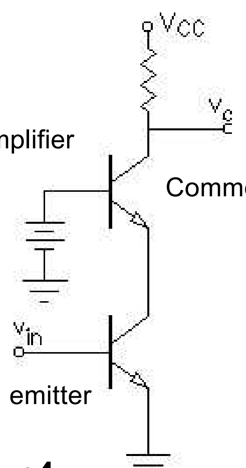
Current gain is 0

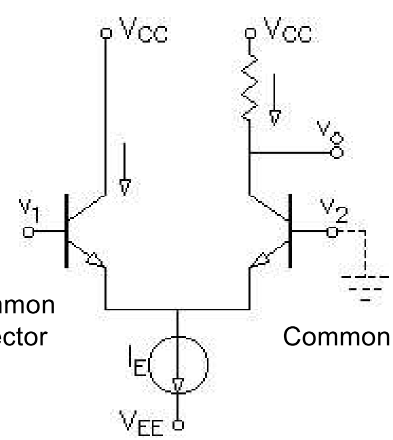
R_{in} is very low

But NO MILLER EFFECT, so very good frequency response, which is the main use for this amplifier.



Rarely used alone. Usually part of Cascode or Differential amplifier



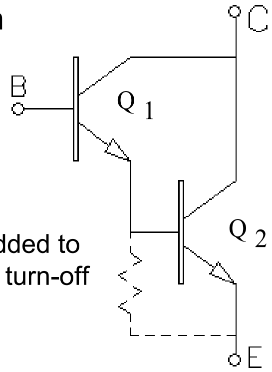


Differential amplifier

Special multiple-transistor connections,

often wired together in a single package

Darlington



R is often added to improve the turn-off speed.

For the pair taken together:

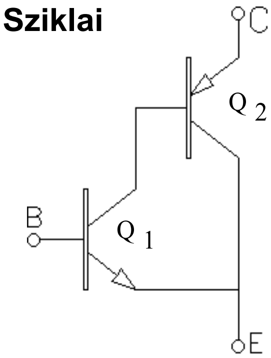
$V_{BE} = 1.4 \cdot V$

$\beta = \beta_1 \cdot \beta_2$

Saturation:

$V_{CE} = 0.9 \cdot V$

Sziklai



For the pair taken together:

$V_{BE} = 0.7 \cdot V$

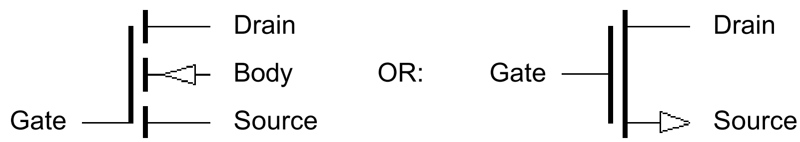
$\beta = \beta_1 \cdot \beta_2$

Saturation:

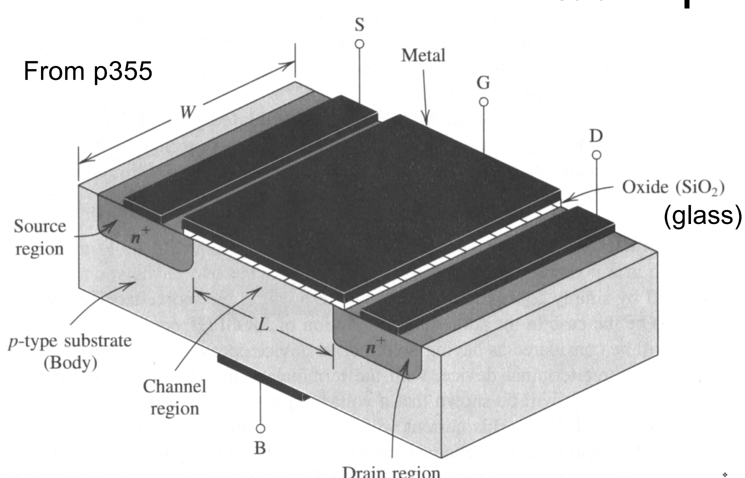
$V_{CE} = 0.9 \cdot V$

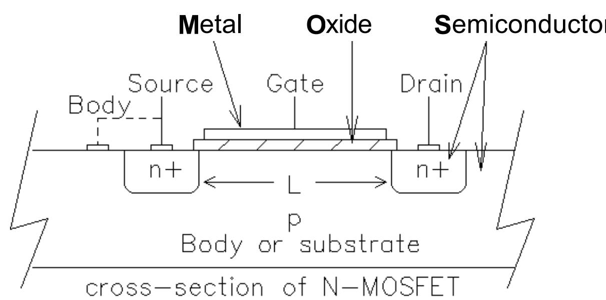
MOSFETS

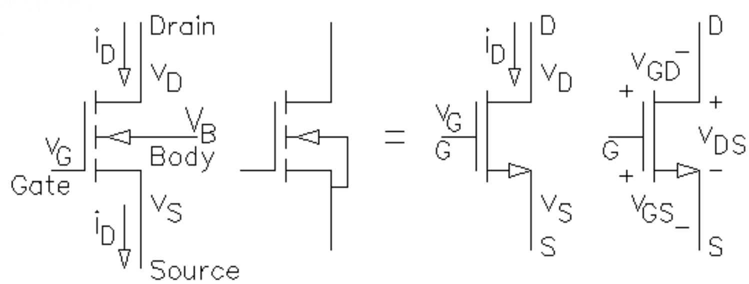
n-channel enhancement:



From p355







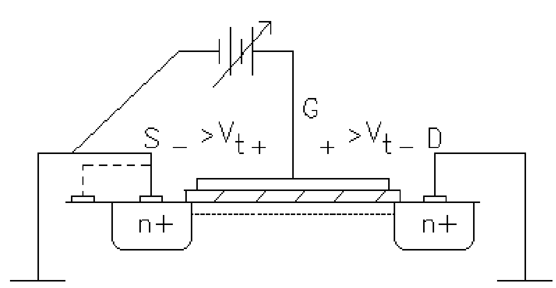
Threshold Voltage

Need some minimum voltage to induce a channel (n) in the p substrate

This minimum voltage is V_t, the threshold voltage

Usually: $1V < V_t < 3V$ but is very variable (like β)

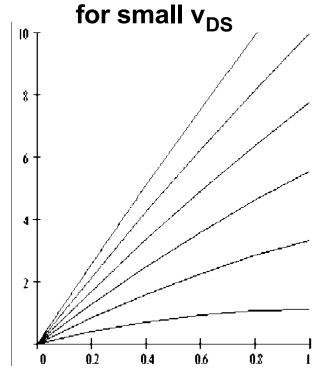
Below this voltage the FET is off



Like the BJT, FETs also have several different regions of operation

Ohmic region of operation (for small v_{DS})

After the threshold voltage, the conductance of the channel will be proportional to the V_{GS}.



$$r_{DS} = \frac{1}{k'_n \cdot \frac{W}{L} \cdot (v_{GS} - V_t)}$$ **<-- IMPORTANT EQUATION** for small v_{DS}

k'_n = process transconductance parameter = 20 to 100 $\mu A/V^2$ $\quad k'_p$ = 8 to 40 $\mu A/V^2$

W = channel width

L = channel length

$K = k'_n \cdot \frac{W}{L}$ = gain factor (combined in books that are less interested in IC design)

$k'_n = \mu_n \cdot C_{ox}$ $\quad \mu_n$ = electron mobility = $580 \cdot \frac{cm^2}{V \cdot s}$ $\quad \mu_p := 230 \cdot \frac{cm^2}{V \cdot s}$ ~ 40%μ_n

$C_{ox} = \frac{\text{oxide capacitance}}{\text{unit area}}$ = ε_{ox}/t_{ox} = permittivity / thickness = capacitance / (unit area)

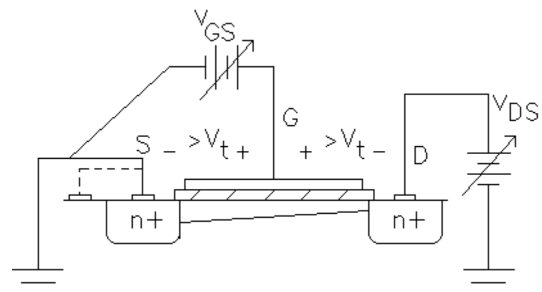
Actually this is only true for small V_{DS}. otherwise the channel won't be constant size from end to end.

Triode Region (aka, linear, also includes ohmic region)

But to get some current to flow from the Drain to the Source, you need some V_{DS}

But that means that the drain-end of the channel starts to close off again,

As long as V_{GD} and V_{GS} are both above V_t, then both ends of the channel will still be open.



This will be true if: $v_{GD} > V_t$ $\quad v_{GD} = v_{GS} - v_{DS}$ $\quad v_{DS} < v_{GS} - V_t$

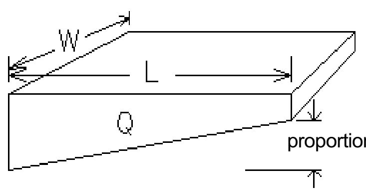
Derivation (not really important to you, as you will not be held accountable for this)

Capacitors: $C = \frac{\text{charge}}{\text{voltage}} = \frac{Q}{V}$ or $Q = C \cdot V$ $\quad C_{ox}$ is capacitance per unit area, so

$Q = C_{ox} \cdot V \cdot L \cdot W$ if the voltage is constant from side to side

left side

$V = V_{GS} - V_t$



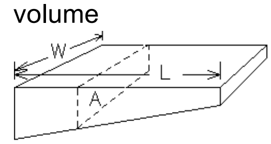
right side

$V = V_{GS} - V_t - V_{DS}$ $\quad$ Average voltage = $V_{GS} - V_t - \frac{1}{2} \cdot V_{DS}$

$$Q = C_{ox} \cdot \left(V_{GS} - V_t - \frac{1}{2} \cdot V_{DS}\right) \cdot L \cdot W$$

Remember drift current? $J_{drift} = q \cdot n \cdot \mu_n \cdot E$ $\quad I_{drift} = i_D = A \cdot q \cdot n \cdot \mu_n \cdot E = \frac{Q}{L} \cdot \mu_n \cdot E$ because: $A_{ave} \cdot L \cdot q \cdot n = Q$

volume



finally, average $E = \frac{V_{DS}}{L}$ $\quad$ So: $i_D = \frac{Q}{L} \cdot \mu_n \cdot E = C_{ox} \cdot \left(V_{GS} - V_t - \frac{1}{2} \cdot V_{DS}\right) \cdot W \cdot \mu_n \cdot \frac{V_{DS}}{L}$

$k'_n = \mu_n \cdot C_{ox}$

rearrange: Drain current = $i_D = k'_n \cdot \frac{W}{L} \cdot \left[(v_{GS} - V_t) \cdot v_{DS} - \frac{1}{2} \cdot v_{DS}^2\right]$ **<-- IMPORTANT EQUATION**