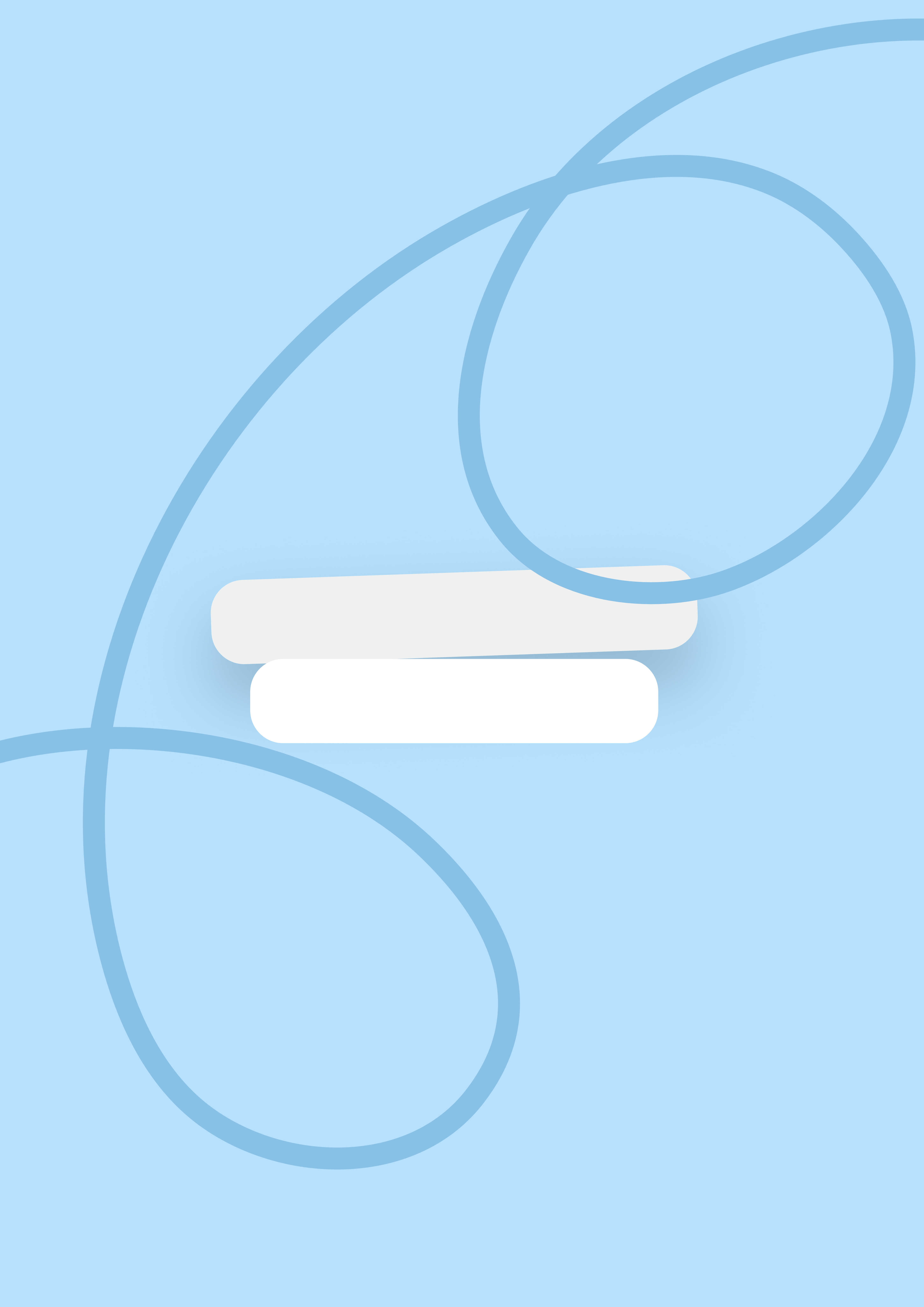

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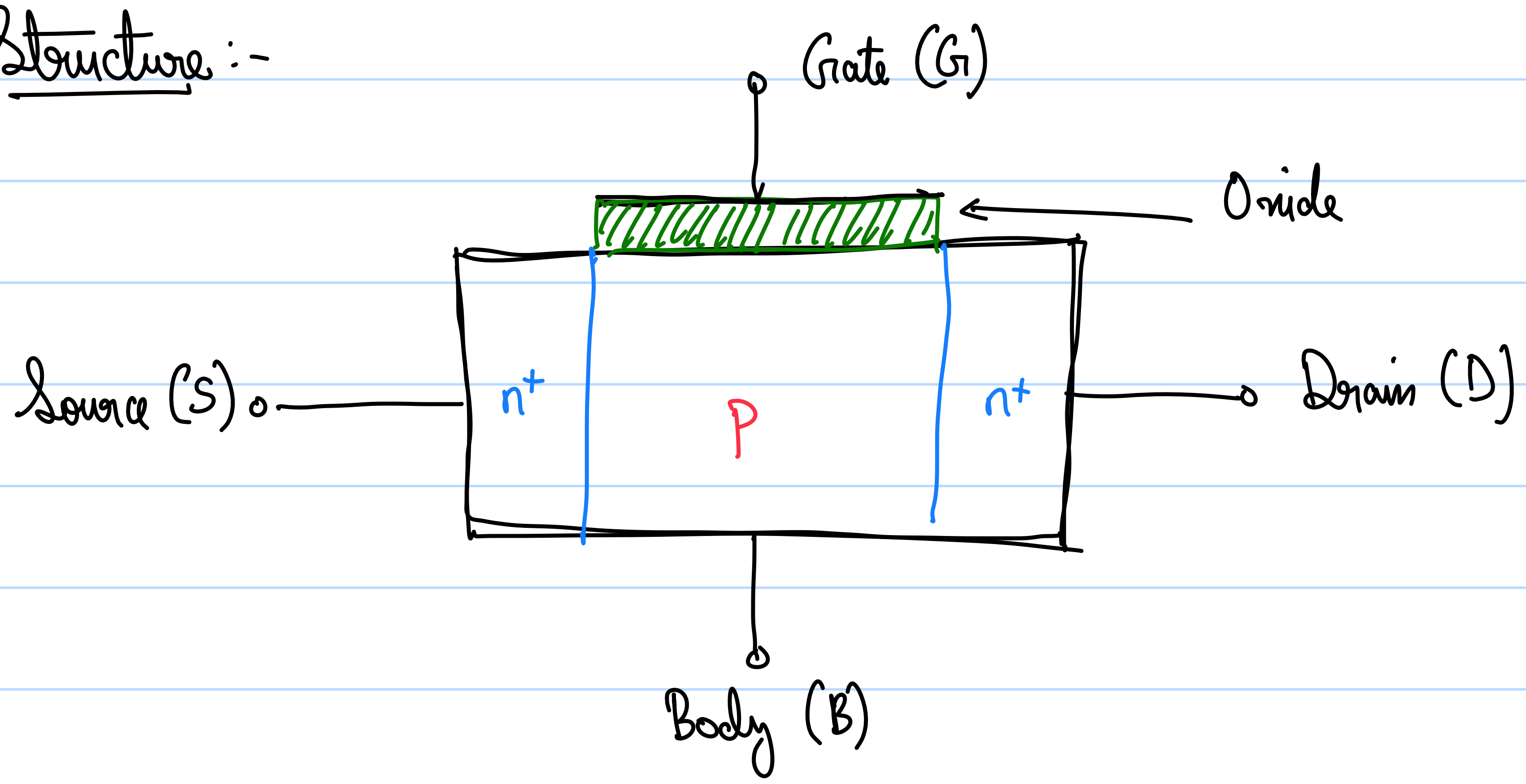
MOSFET

→ **Structure** :-



- If a positive voltage is applied on the Gate plate, electrons from the p type silicon are attracted to the Gate region.

- These electrons can act as a channel for current to pass through and if current density is high enough, acts like a conductive path.

- The density of electrons varies with voltage since wkt $Q = CV$. (the gate is basically a capacitor)

- Note: Current will not flow in the p type area, if channel is present. (Path of least resistance)

- To achieve a strong control of Q with V, C must be minimized, ie, width of oxide layer must be minimized.

- The source terminal provides the charge carriers and the drain absorbs them. MOSFETs are symmetric wrt to S and D.

- If source, drain are n type and substrate is p type : NMOS

- The gate plate was earlier made using aluminium (hence "metal"). Now we use polysilicon, with heavy doping. The oxide is silicon dioxide.

- In proper operation the S and D junctions are in reverse bias. So only depletion region capacitance is taken into account.

→ Operation :

- The gate terminal draws no (low frequency) current since it is insulated by the oxide layer.

- The only current of interest is the one b/w S and D.

- When V_G is slightly positive, a depletion region of immobile charges is created at the gate, due to holes being repelled.

- As V_G increases, the depletion region becomes deeper and deeper.

- When V_G is sufficiently positive, free electrons are attracted to the oxide-silicon interface, forming the required channel of mobile carriers

- This threshold voltage is around 0.3 - 0.5 V. The electrons for the channel are supplied by the drain and source region.

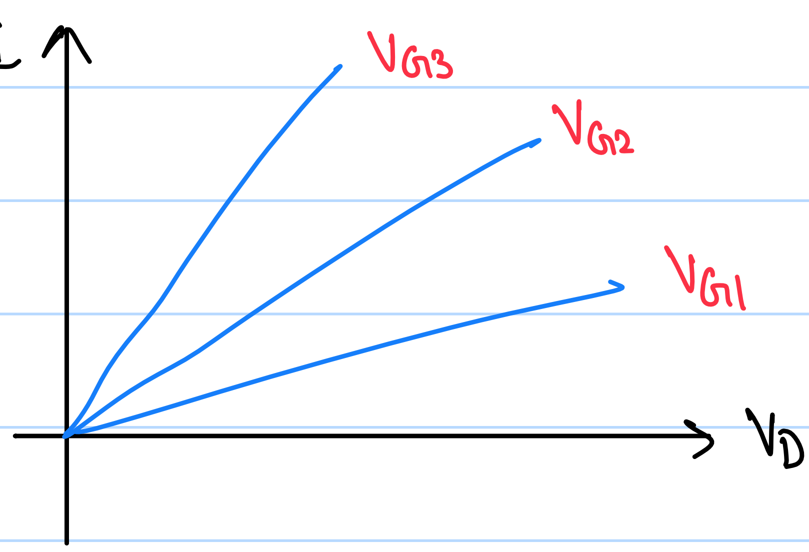
- MOSFET as a Variable Resistor :-

- The channel is not fully conductive and depending on the electron density and the width, will have some resistance.

- The value of this resistance varies with the gate voltage. (Voltage Dependent Resistor)

- The slope of the graph is $1/R_{on}$, where R_{on} is the "on-resistance" of the MOSFET

- If V_D (voltage at drain) varies with constant V_G,



$V_{G3} > V_{G2} > V_{G1}$

$R_{G3} < R_{G2} < R_{G1}$

$\Rightarrow$ $R \propto \frac{1}{V_G}$

$V_G \uparrow$ density $\uparrow$ resistance $\downarrow$

- The current in the channel is majorly due to drift.

- The resistance is directly proportional to the oxide width. ($Q = CV$) (width ↑ C ↓ Q ↓ density ↓ resistance ↑). Therefore, lesser oxide width is needed for more current.

<u>Note</u>: The oxide thickness is defined during fabrication and is constant for all MOSFETS in a generation.

- Circuit designers can control the length of the gate region and the cross-sectional width.

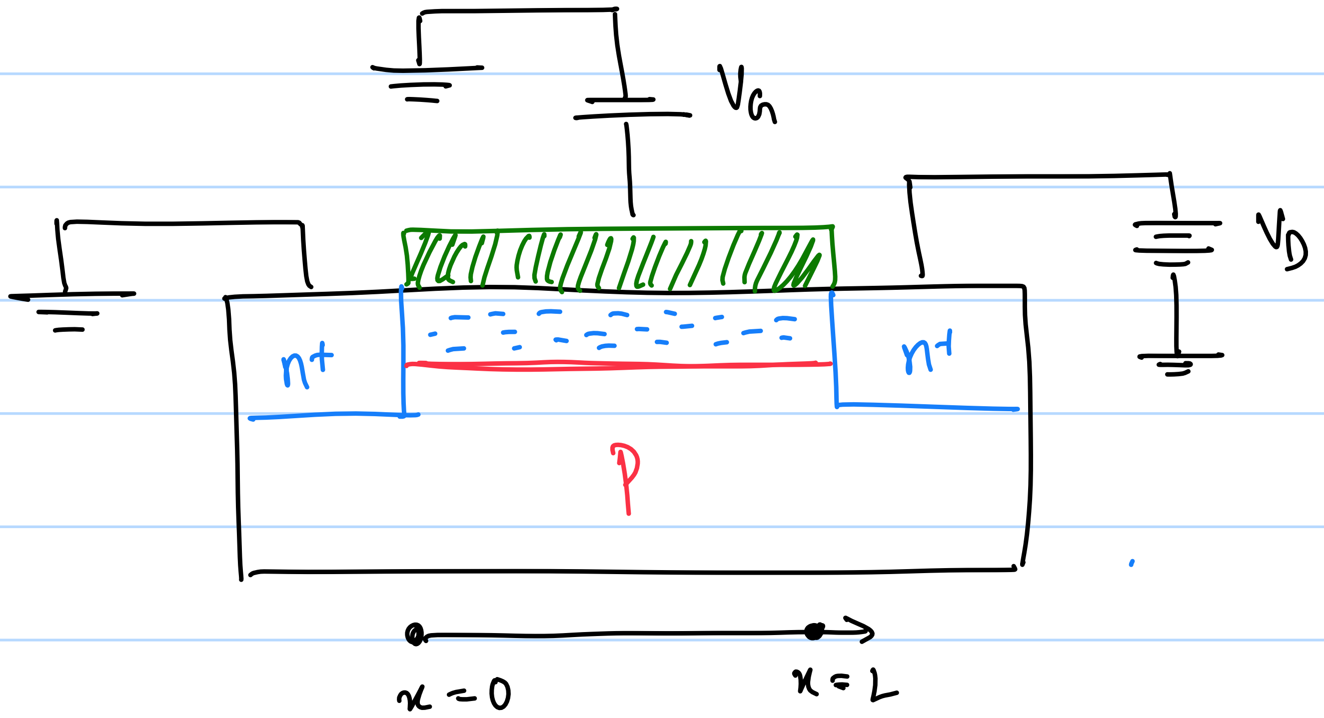
- Cross-Sectional Width $\propto \dfrac{1}{\text{Resistance}}$

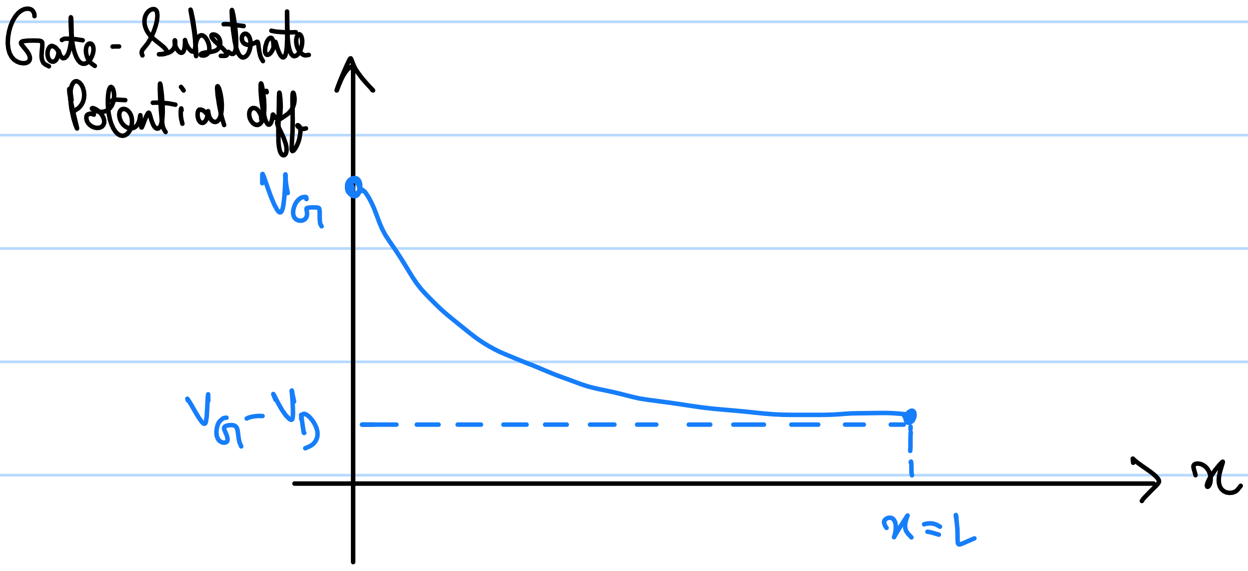
- <u>Channel Pinch-Off</u> :-

- If the drain voltage is sufficiently positive, the current through the transistor will remain constant for increasing drain voltage.

- This is due to channel pinch-off.

- Since $V_S < V_D$ always, there will be a voltage gradient from the source to drain.



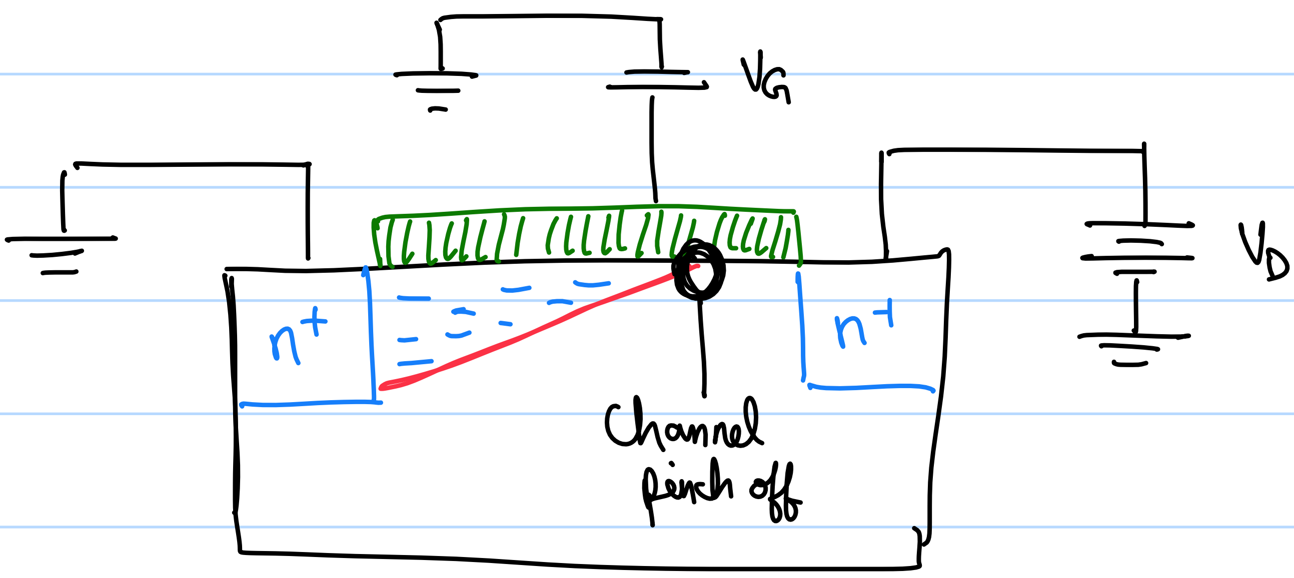


- If $V_G - V_D < V_{TH}$, then the channel will not exist near the drain.

$$V_G - V_D < V_{TH}$$

$$\Rightarrow \quad V_D > V_G - V_{TH}$$

$\therefore$ If $V_D > V_G - V_{TH}$, the channel will not exist near the drain



- Current will still pass through, as any electron that reaches the pinched region will be swept off into the drain (carrier injection)

- This behaviour is similar to that of BJT in forward active region.

◦ I/V Characteristics :-

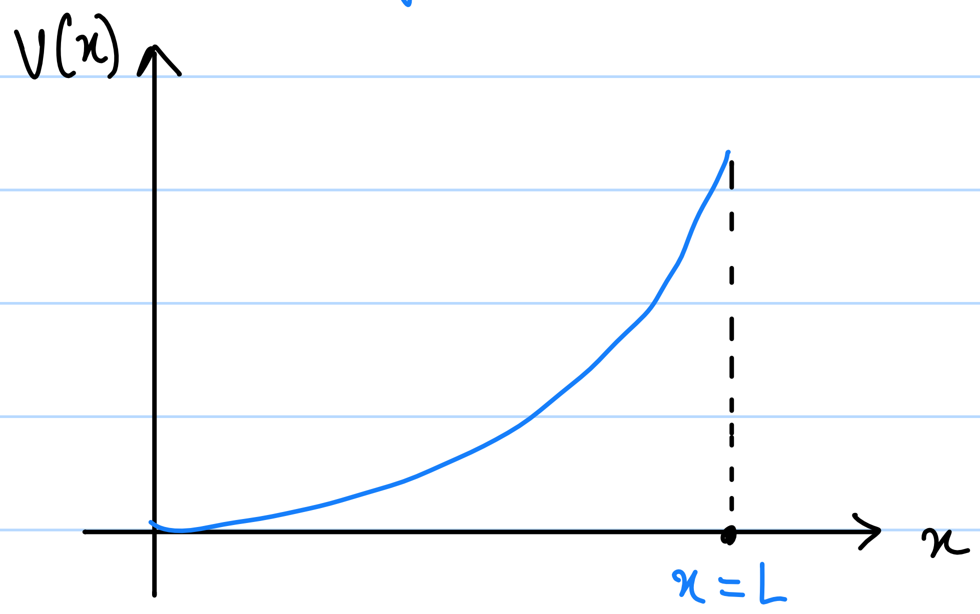
1) Channel Charge Density :

Charge in the channel is given by,

$$Q = W C_{ox} (V_{GS} - V_{TH})$$

C_{ox} – Capacitance per unit area of the oxide.

Channel potential varies w/ length as,



$$\Rightarrow Q(x) = W C_{ox} (V_{GS} - V(x) - V_{TH})$$

Coulomb/meter

2) Drain Current :-

- Since drain current is due to the same phenomena that creates regular drift current, we can say that,

$$I_D = Q \cdot v_D$$

$v_D \rightarrow$ Drift velocity

$Q \rightarrow$ Linear charge density

$$v_D = -\mu_n E = -\mu_n \frac{dV}{dx}$$

$$\Rightarrow \quad I_D = W C_{ox} \left[V_{GS} - V(x) - V_{TH} \right] \mu_n \frac{dV}{dx}$$

Since I_D must remain constant along the channel (KCL), the variation in $V(x)$ and $\frac{dV}{dx}$ is such that $(V_{GS} - V(x) - V_{TH}) \frac{dV}{dx}$ is independent of x.

Shift dx to LHS and integrate LHS from $x = 0$ to L and RHS from $V(x) = 0$ to V_{DS}

Solving the differential equation, we will get,

$$I_D = \frac{1}{2} \mu_n C_{ox} \frac{W}{L} \left(2(V_{GS} - V_{TH}) V_{DS} - V_{DS}^2 \right)$$

$\rightarrow$ Parabolic relation b/w I_D and V_{DS}

<u>Observations</u>:

1. $I_D \propto \mu_n C_{ox} \frac{W}{L} \rightarrow I_D \propto \mu_n, \; I_D \propto C_{ox}, \; I_D \propto W, \; I_D \propto \frac{1}{L}$

2. $I_{D,max} = \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{TH})^2$ at $V_{DS} = V_{GS} - V_{TH}$

3. If $V_{DS} \ll 2(V_{GS} - V_{TH})$,

$$I_D \approx \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{TH}) V_{DS}$$

$\rightarrow$ Linear behaviour in I_D vs V_{DS}.

On resistance,

$$R_{on} = \frac{1}{\mu_n C_{ox} \frac{W}{L}(V_{GS} - V_{TH})}$$ $\rightarrow$ If $V_{GS} \rightarrow V_{TH}$, $R_{on} \rightarrow \infty$

3) Triode And Saturation Region :-

- The equation for I_D implies that I_D begins to fall when $V_{DS} > V_{GS} - V_{TH}$

- $V_{DS} < V_{GS} - V_{TH}$ is termed as triode region.
  $V_{DS} << V_{GS} - V_{TH}$ is termed as deep triode. linear behaviour of $I_D - V_{DS}$

- We know that pinch-off occurs at $V_{DS} = V_{GS} - V_{TH}$. Further increase in V_{DS} shifts the pinch-off point towards the drain.

- Since the equation we have for I_D is valid only if channel is fully present,

$$I_D = \frac{1}{2} \mu_n C_{ox} \frac{W}{\underline{\underline{L_1}}} (V_{GS} - V_{TH})^2$$

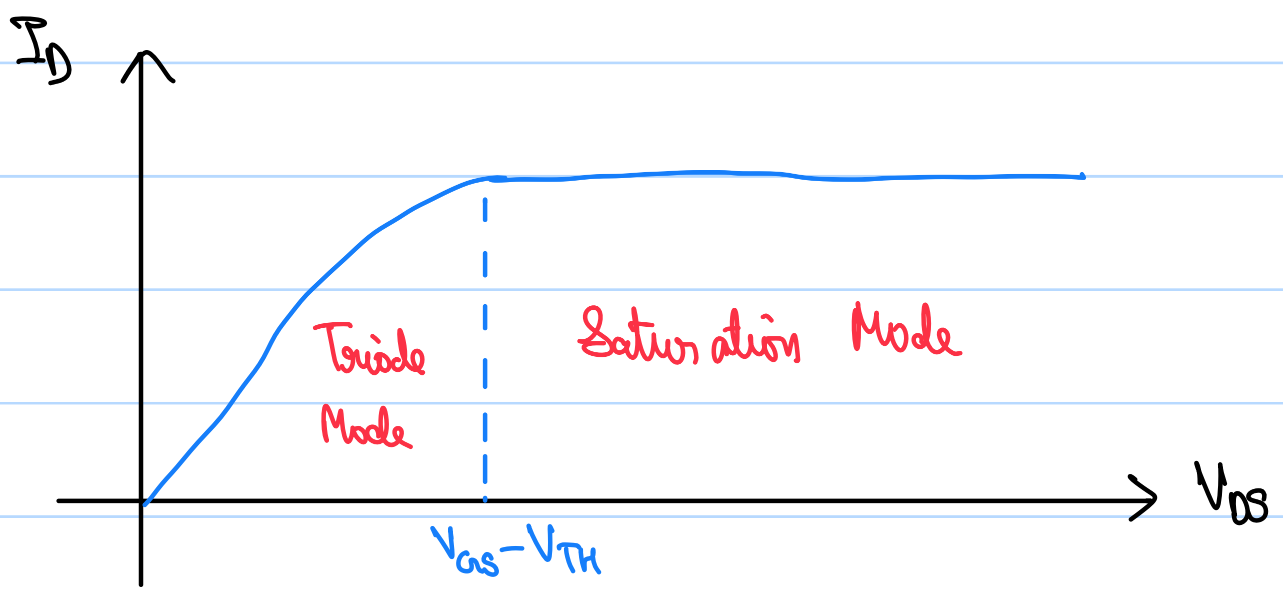
Since we are changing the limits of integration to:
$x = 0 \rightarrow L_1$, $V = 0 \rightarrow (V_{GS} - V_{TH})$, where L_1 is length till the pinch-off point

- This equation is independent of V_{DS}. So it follows that at $V_{DS} > V_{GS} - V_{TH}$, I_D is independent of V_{DS}.

- The quantity $V_{GS} - V_{TH}$ is termed as overdrive voltage.

- We can estimate L_1 as L.

- This region is termed as saturation mode of the MOSFET.



- The MOSFET will act as a current source in saturation mode.

- Due to the square-law dependance of I_D on $V_{GS} - V_{TH}$, MOSFETs are termed as square-law devices.

- Because of the square law dependance, the MOSFET in saturation acts like a Voltage Dependent Current Source.

- <u>Comparision Between BJT and MOSFET</u>:

- MOSFETs do not draw any bias current.

- BJTs have an exponential relation wrt to the base voltage, whereas MOSFETs have a square law relation.

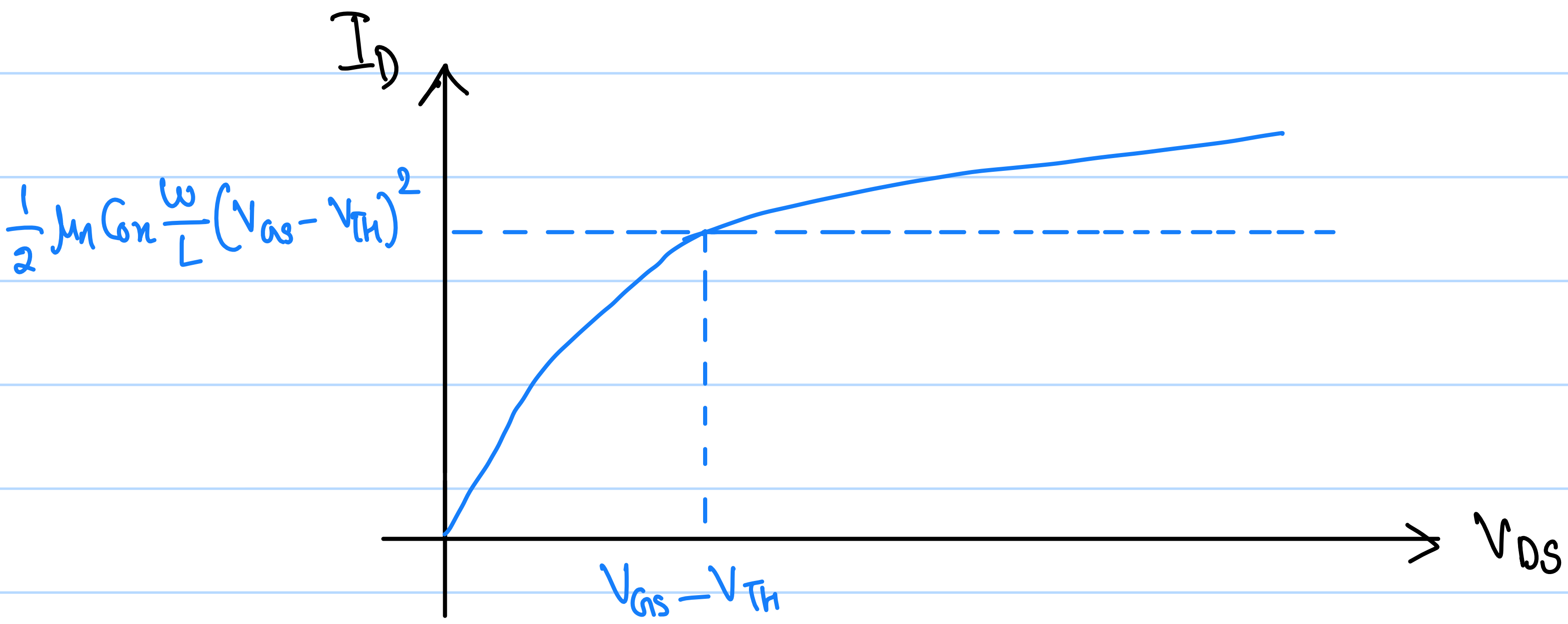
- The dimension ratios of MOSFETs can vary widely as per circuit requirements.

4) Channel Length Modulation:

- In the earlier IV graph, we get a constant current for $V_D > V_{GS} - V_{TH}$.

- This is partly due to the approximation of L_1 as L.

- But L_1 is a V_{DS} dependent quantity, so I_D actually still continuously increases after the overdrive voltage, but in a somewhat linear manner.



- This phenomenon is termed as Channel Length Modulation

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})^2(1+\lambda V_{DS})$$

λ - Channel Length Modulation Coefficient

- λ is a parameter that can be controlled by the designer, since

$$\lambda \propto \frac{1}{L} \longrightarrow$$ If L is large enough, the relative change in L becomes smaller.

- λ limits the output impedance of MOSFET based devices (Sim to r_o in BJT)

→ **MOS Transconductance :**

- While behaving like a voltage dependent current source, the MOSFET can be characterized by its transconductance.

$$g_m = \frac{dI_D}{dV_{GS}}$$

$$\Rightarrow \quad g_m = \mu_n C_{ox} \frac{W}{L} (V_{GS} - V_{TH})$$

$$\Rightarrow \quad g_m = \sqrt{2 \mu_n C_{ox} \frac{W}{L} I_D}$$

$$\Rightarrow \quad g_m = \frac{2 I_D}{V_{GS} - V_{TH}}$$

- **Observations :**

1. $g_m \propto \frac{W}{L}$ for a given $V_{GS} - V_{TH}$ and viceversa

2. $g_m \propto \sqrt{\frac{W}{L}}$ for a given I_D and vice versa.

3. $g_m \propto I_D$ for a given $V_{GS} - V_{TH}$ and $g_m \propto \frac{1}{V_{GS} - V_{TH}}$ for a given I_D.

→ Velocity Saturation:

- In normal conductors, at high enough electric fields, the carrier mobility starts to degrade, which leads to a constant drift velocity. This is velocity saturation.

- MOSFETs can experience velocity saturation at V_{DS} as low as 1V, due to which I_D no longer holds the square law behavior.

$$\Rightarrow I_D = V_{sat} Q$$
$$= V_{sat} W C_{ox} (V_{GS} - V_{TH}) \longrightarrow \text{Linear dependance}$$

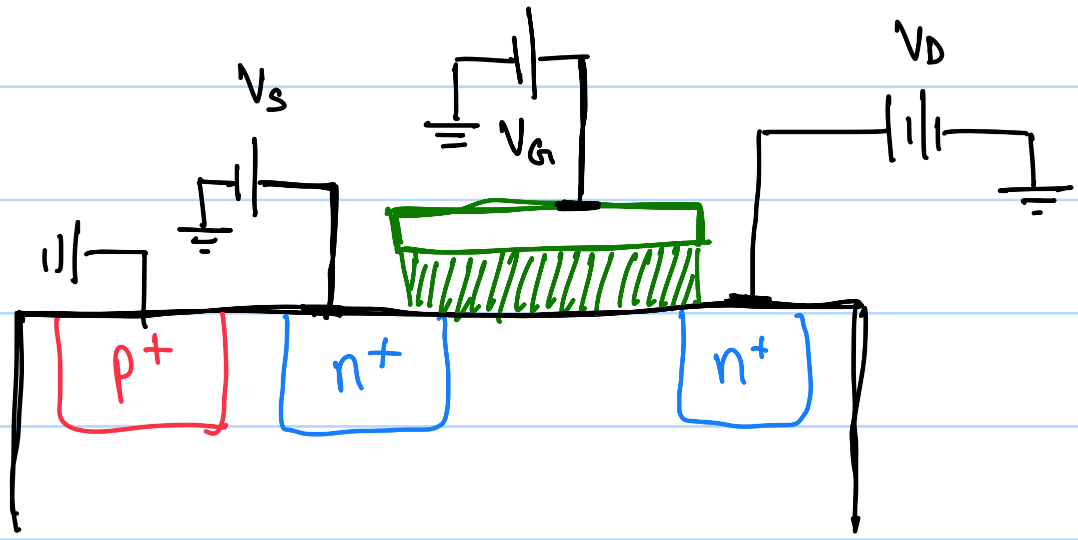
$$\Rightarrow \boxed{g_m = V_{sat} W C_{ox}}$$

→ Second Order Effects:

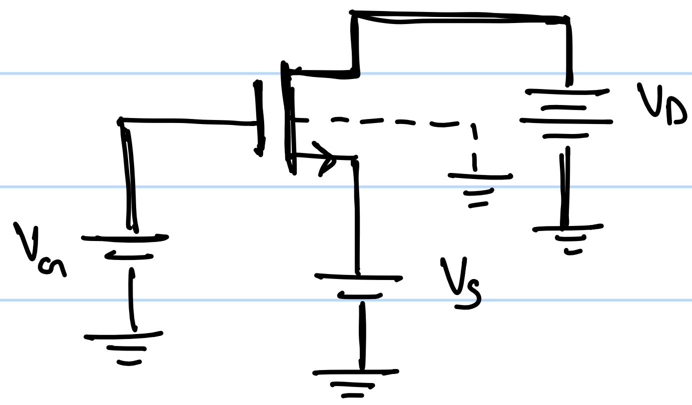
- Body Effects:

In all the previous models, the source and the substrate both are assumed to be grounded.

- However that need not be the case,



2)



- Here the source is at a higher voltage than the substrate. The source - substrate junction is still in reverse bias. So the overall working of the MOSFET is not disturbed.

- The higher substrate voltage has an effect on the threshold voltage however. This is termed as "Body effect" and is formulated as,

$$V_{TH} = V_{TH0} + \gamma\left(\sqrt{|2\phi_F + V_{SB}|} - \sqrt{|2\phi_F|}\right)$$

$\hookrightarrow$ V_{TH} at $V_{SB} = 0$ $\searrow$ $V_{TH} \propto \sqrt{V_{SB}}$

- γ and ϕ_F are transistor dependent quantities

- **Subthreshold Conduction:**

- In the IV characteristics, we have assumed that I_D only exists when $V_G > V_{TH}$, ie, when channel is fully formed.

- However, channel formation is a gradual process and a small I_D exists even when $V_G < V_{TH}$. This is subthreshold conduction.

→ **Modelling of MOSFETs :-**

- **Large Signal Model:**

- From the IV characteristics, we have 2 equations,

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left[2(V_{GS}-V_{TH})V_{DS} - V_{DS}^2\right] \quad \text{Triode Region}$$

$$I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})^2(1+\lambda V_{DS}) \quad \text{Saturation Region}$$

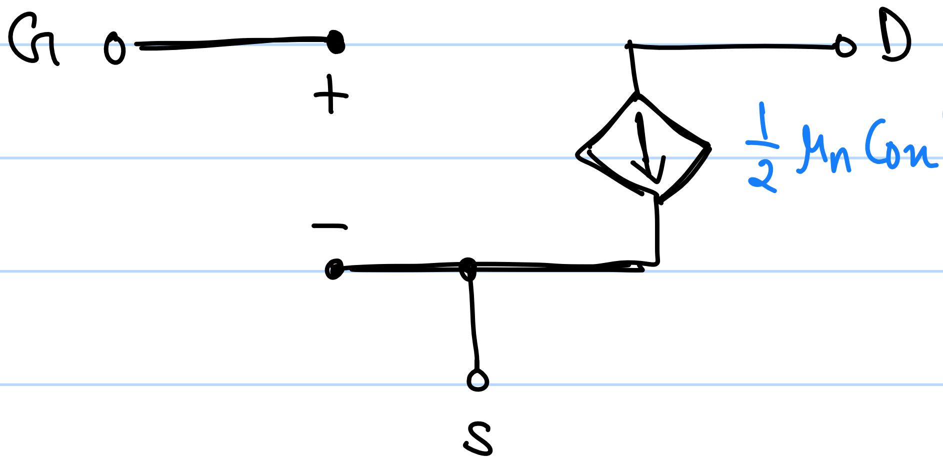
- In the saturation region, we have a voltage-controlled current source (Not ideal since V_{DS} dependence is present)

- In the triode region, we still have a voltage controlled current source, with a different relation.

- If $V_{DS} \ll 2(V_{GS}-V_{TH})$, the transistor will act like a voltage dependent resistor.

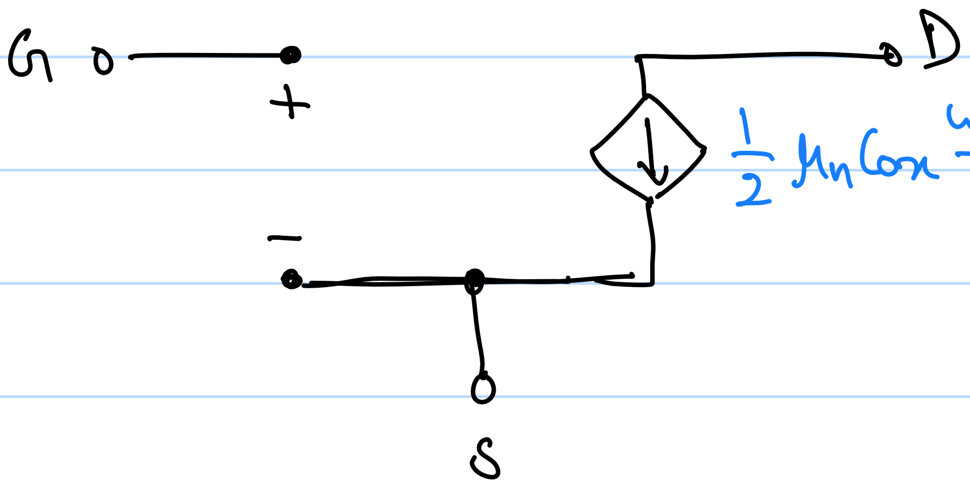
- In all the 3 cases, the gate terminal is open since gate current is zero.

a) Saturation



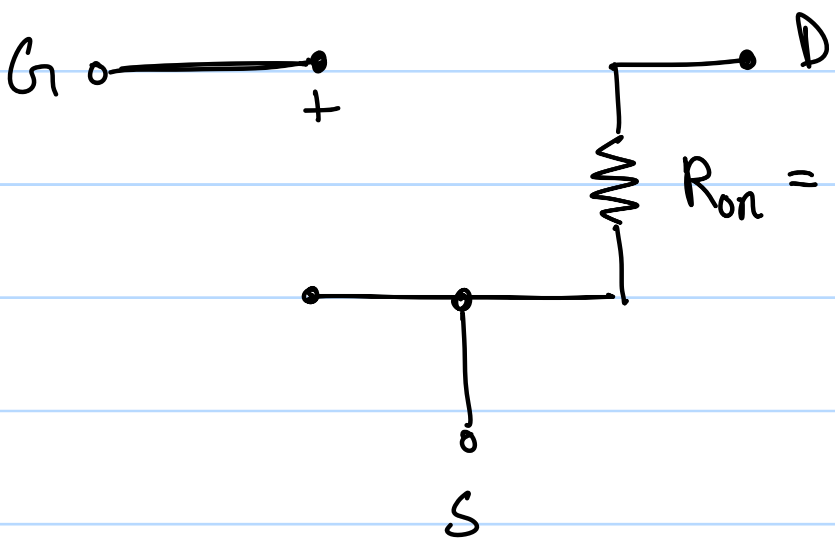
$$\frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})^2(1+\lambda V_{DS})$$

b) Triode



$$\frac{1}{2}\mu_n C_{ox}\frac{W}{L}\left(2(V_{GS}-V_{TH})V_{DS}+V_{DS}^2\right)$$

c) Deep Triode

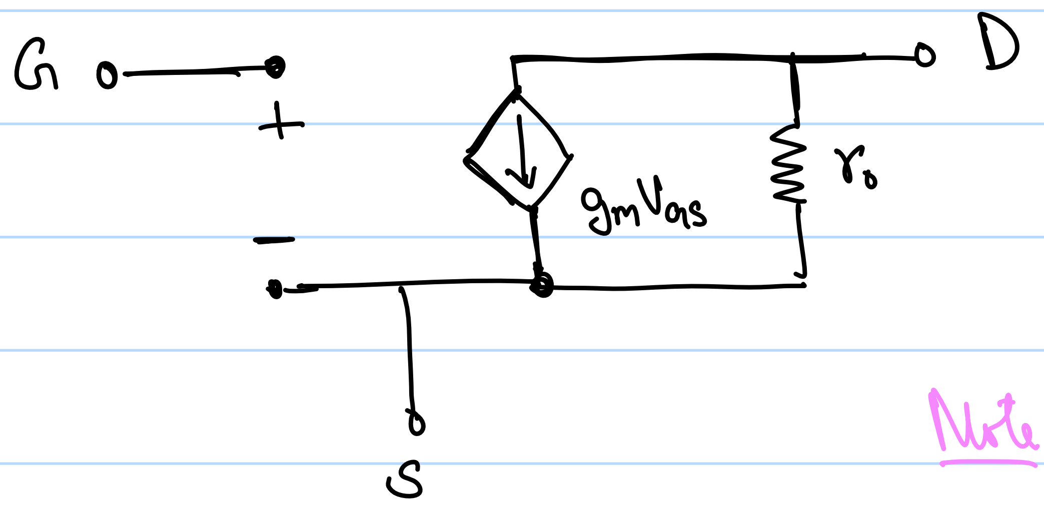


$$R_{on}=\frac{1}{\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})}$$

Small Signal Model :-

- The small signal model of a MOSFET can be derived in a similar way as what was done in BJTs.

- The small signal model is,



Note: Capital letters are for large signal components, Small letters are for small signal components.

Where,

$$r_o = \left(\frac{dI_D}{dV_{DS}}\right)^{-1}$$

$$= \frac{1}{\frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})^2 \cdot \lambda}$$

$$\Rightarrow \boxed{r_o \approx \frac{1}{\lambda I_D}}$$

Since Channel length modulation is small for small signals, variation in I_D is negligible.

- **Derivation For i_D :-**

For a small signal,

$$V_{GS} = V_{GS0} + v_{in}(t)$$

$\longrightarrow$ DC component

$$\Rightarrow \quad I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS}-V_{TH})^2(1+\lambda V_{DS})$$

$$\Rightarrow \quad I_D = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS0}+v_{in}(t)-V_{TH})^2(1+\lambda V_{DS})$$

$$I_{D0} = \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(V_{GS0}-V_{TH})^2(1+\lambda V_{DS})$$

I_{D0} is the drain current due to the DC component / Bias voltage.

By principle of superposition

$$\Rightarrow \quad i_D = I_D - I_{D0}$$

$$= \frac{1}{2}\mu_n C_{ox}\frac{W}{L}(1+\lambda V_{DS})\left((V_{GS0}+v_{in}(t)-V_{TH})^2 - (V_{GS0}-V_{TH})^2\right)$$

$$(V_{GS0}+v_{in}(t)-V_{TH})^2 - (V_{GS0}-V_{TH})^2$$

$$= (V_{GS0}+v_{in}(t)-V_{TH}+V_{GS0}-V_{TH})(\cancel{V_{GS0}}+v_{in}(t)-\cancel{V_{TH}}-\cancel{V_{GS0}}+\cancel{V_{TH}})$$

$$= (2(V_{GS0}-V_{TH})+v_{in}(t))(v_{in}(t))$$

If $\boxed{\text{amp}(v_{in}(t)) << 2(V_{GS0}-V_{TH})}$ $\xrightarrow{\approx 200\text{ mV}}$ Condition of small signal operation

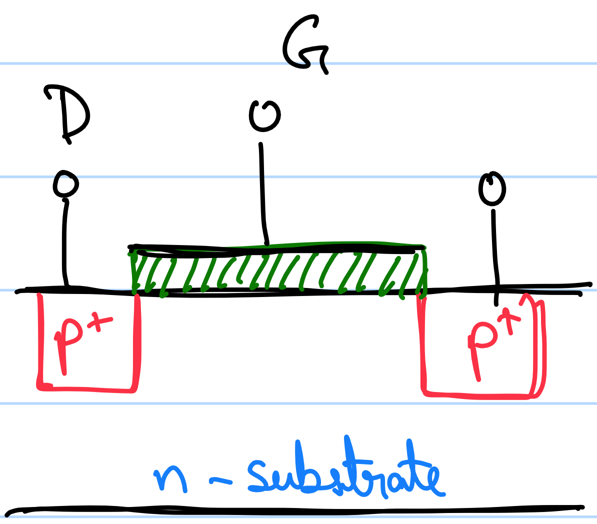
$$\Rightarrow \quad i_D = \frac{1}{\cancel{2}}\mu_n C_{ox}\frac{W}{L}(\cancel{2}(V_{GS0}-V_{TH})v_{in}(t))(1+\lambda V_{DS})$$

$$\Rightarrow \quad \boxed{i_D = g_m v_{in}(t)}$$

wkt,

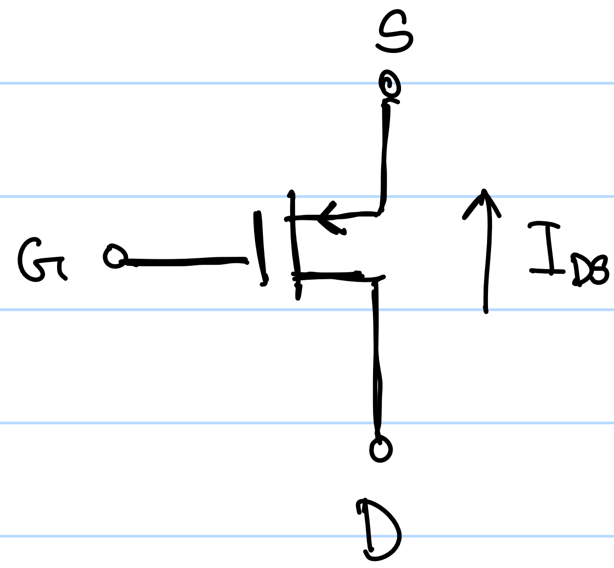
$$g_m = \mu_n C_{ox}\frac{W}{L}(V_{GS0}-V_{TH})(1+\lambda V_{DS})$$

→ PMOS:-



- A PMOS has a channel made up of holes

- To create such a channel, we need a negative V_{GS}.
  $V_{GS} < V_{TH}$, where $V_{TH} < 0$.



- Triode region if the drain voltage is near the source potential and saturation region if V_D falls to $V_G - V_{TH} = V_G + |V_{TH}|$

- In saturation,

$$I_{D,sat} = \frac{-1}{2} \mu_p C_{ox} \frac{W}{L} (V_{GS} - V_{TH})^2 (1 - \lambda V_{DS})$$

In triode,

$$I_{D,tri} = \frac{-1}{2} \mu_p C_{ox} \frac{W}{L} \left[2(V_{GS} - V_{TH}) V_{DS} - V_{DS}^2 \right]$$

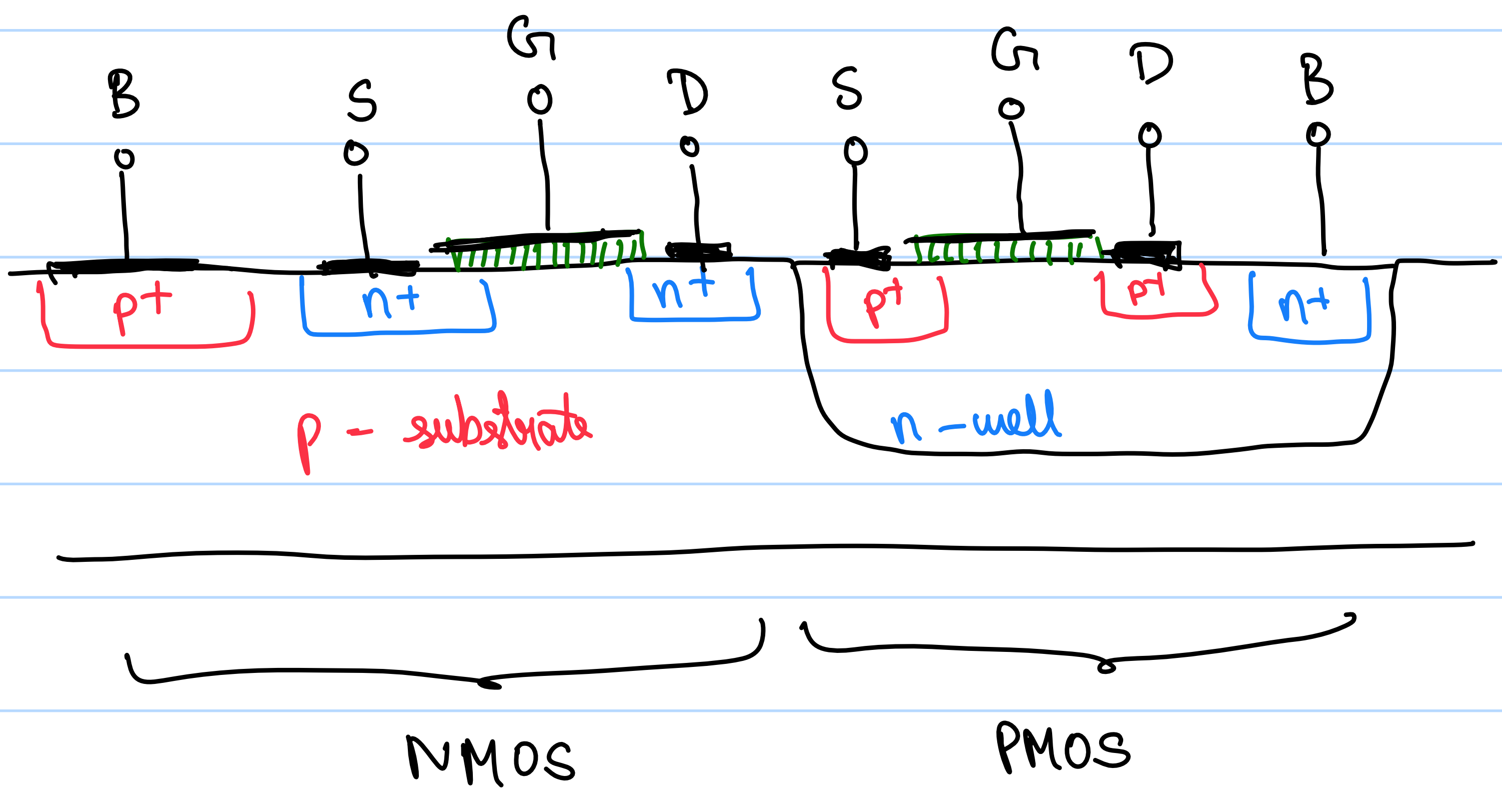
- PMOS small signal is the same as NMOS small signal.

- Generally, $\mu_p < \mu_n$. Therefore NMOS is preferred.

→ CMOS Technology :-

- PMOS and NMOS require 2 different substrates to function properly.

- If we wish to integrate them together, we use a Complementary MOS or CMOS structure.



- The CMOS device is made using a p type substrate, suitable for NMOS, with n type "wells" suitable for PMOS.

→ Comparision of BJT and MOSFET :-

- IV Characteristics : Exponential Quadratic

- Regions of Operation

Active : $V_{CB} > 0$ Saturation : $V_{DS} > V_{GS} - V_{TH}$

Saturation : $V_{CB} < 0$ Triode : $V_{DS} < V_{GS} - V_{TH}$

- Base Current : Finite Zero (for small frequencies)

- Increase in I_{DS} during Amplification :

Early Effect Channel Length Modulation

- Current Flow Mechanism :

Diffusion Drift

+ MOSFET can act like a voltage dependent resistor. BJTs cannot.