Sedra/Smith
Microelectronic Circuits 5/e

Chapter 5B  MOS Field-Effect Transistors (MOSFETs)
5.5 Small-Signal operation and models

5.5.1 The DC Bias Point

To ensure saturation region operation.

We must have $V_D > V_{OV}$
5.5.2 The Signal Current in the Drain Terminal

\[ v_{GS} = v_{gs} + V_{GS} \]

\[ i_D = \frac{1}{2} k_n (v_{gs} + V_{GS} - V_t)^2 \]

\[ = \frac{1}{2} k_n (V_{GS} - V_t)^2 + k_n (V_{GS} - V_t)v_{gs} + \frac{1}{2} k_n v_{gs}^2 \]  \hspace{1cm} (5.43)

The Current component that is directly proportional to the input signal \( v_{gs} \)

To reduce the nonlinear distortion, the input signal should be kept small

so that \[ \frac{1}{2} k_n \frac{W}{L} v_{gs}^2 << k_n \frac{W}{L} (V_{GS} - V_t)v_{gs} \Rightarrow v_{gs} << 2(V_{GS} - V_t) \]
Example:
For the amplifier in Fig. A, let the MOSFET is specified to have $V_t = 2V$ and $k_n W / L = 2mA/V^2, V_{GS} = 5V$, assume $\lambda = 0$. If $v_{gs} = 0.5 \sin \omega t$ volts
(a) Determine the various components. (b) Find the second harmonic component as a percentage of the amplitude of fundamental
Sol:
(a) $i_D = 1mA/V^2 \times (5 - 2)^2 V^2 + 2mA/V^2 \times 0.5 \sin \omega t \times (5 - 2)V^2$
   $+ 0.25 \sin^2 \omega t \times 1mA$
   $= 9 + 3 \sin \omega t + 0.25(\frac{1}{2} - \frac{1}{2} \cos 2\omega t) \ (mA)$
   $= 9.125 + 3 \sin \omega t - 0.125 \cos 2\omega t \ (mA)$
(b) Harmonic distostion percentage
   $D\% = \frac{0.125}{3} \times 100\% = 4.16\%$
Small-signal operation of the enhancement MOSFET amplifier.

\[ g_m \equiv \frac{\partial I_D}{\partial V_{GS}} \bigg|_{v_{gs}=V_{GS}} \]  

(5.49)

\[ g_m \equiv \frac{i_d}{v_{gs}} = k_n \frac{W}{L} (V_{GS} - V_t) \]  

(5.47)
5.5.3 The voltage gain

\[ v_D = V_{DD} - i_D R_D \]
\[ = V_{DD} - (i_d + I_D) R_D \]
\[ = V_{DD} - I_D R_D - i_d R_D \]

The biasing component is
\[ V_D = V_{DD} - I_D R_D \]

The signal component is
\[ v_{ds} = -i_d R_D = -g_m v_{gs} R_D \]  (5.50)

\[ \therefore A_v = \frac{v_d}{v_{gs}} = -g_m R_D \]  (5.51)

The minus sign, indicates that the \( v_d \) is 180\(^0\) out of phase with respect to the \( v_{gs} \).
Total instantaneous voltages $v_{GS}$ and $v_D$ for the circuit in Fig. 4.34.

$\frac{V}{2} \ll 2(V_{GS} - V_i)$

$\nu_i = \nu_{gs}$

small signal condition
5.5.5 Small-signal Equivalent-Circuit Models

(a) neglecting the dependence of $i_D$ on $v_{DS}$ in saturation (the channel-length modulation effect);

(b) including the effect of channel-length modulation, modeled by output resistance $r_o = |V_A| / I_D$. 

---

S. C. Lin, EE National Chin-Yi University of Technology
5.5.6 The Transconductance $g_m$

$$g_m = \frac{i_d}{v_{gs}} = k_n (v_{GS} - V_t)$$

$$= k_n' \frac{W}{L} (v_{GS} - V_t) = k_n' \frac{W}{L} V_{OV} \tag{5.55}$$

$$I_D = \frac{1}{2} k_n' \frac{W}{L} V_{OV}^2 \Rightarrow V_{OV} = \sqrt{\frac{2I_D}{k_n} \frac{W}{L}}$$

Substituting for $V_{OV}$ in (5.55) by

$$V_{OV} = \sqrt{2I_D / k_n' (W / L)}$$, we obtain

$$g_m = \sqrt{2k_n'} \sqrt{(W / L)} \sqrt{I_D} \tag{5.56}$$

Yet another useful expression for $g_m$, Substituting for $k_n' \frac{W}{L}$

in (5.55) by $\frac{2I_D}{(v_{GS} - V_t)^2}$, we obtain

$$g_m = \frac{2I_D}{v_{GS} - V_t} = \frac{I_D}{V_{OV} / 2} \tag{5.57}$$
Example 5.10 For the amplifier in Fig. a, let the MOSFET is specified to have $V_t = 1.5\text{V}$, $V_A = 50\text{V}$ and $k_n = 0.25\text{mA/V}^2$. Assume the coupling capacitors to be sufficiently large so as to act as short circuit at the signal frequencies of interest.

(a) Find the voltage gain
(b) Find the input resistance
(c) Find the largest allowable input signal.
Sol: (a) \( I_D = \frac{1}{2} k_n \frac{W}{L} (V_{GS} - V_t)^2 \)

\[ = \frac{1}{2} \times 0.25 \times (V_{GS} - 1.5)^2 \]

\( V_{DS} = 15 - I_D R_D = 15 - 10 I_D \)

\( \Rightarrow I_D = 1.06mA \quad V_D = 4.4V \)

\( g_m = k_n \frac{W}{L} (V_{GS} - V_t) \)

\[ = 0.25 \times (4.4 - 1.5) = 0.724mA/V \]

\( r_o = \frac{V_A}{I_D} = \frac{50V}{1.06mA} = 47k\Omega \)

\( \Rightarrow v_o = -g_m v_{gs} (R_D // R_L // r_o) \)

\( A_v = \frac{v_o}{v_i} = -g_m (R_D // R_L // r_o) \)

\[ = -0.725(10 // 10 // 47) = -3.3V/V \]

(b) \( i_i = \frac{(V_i - V_o)}{R_G} \)

\[ = \frac{V_i}{R_G} (1 - \frac{V_o}{V_i}) \]

\[ = \frac{V_i}{R_G} (1 - A_v) \]

\[ = \frac{4.3}{R_i} V_i \]

Thus: \( R_{in} = \frac{V_i}{i_i} = \frac{R_G}{4.3} \)

\[ = 10/4.3 = 2.33M\Omega \]

(c) \( v_{DS(max)} = v_{GS(max)} - V_t \)

\( v_{DS} - |A_v| \hat{v}_i = v_{GS} + \hat{v}_i - V_t \)

\( 4.4 - 3.3 \hat{v}_i = 4.4 + \hat{v}_i - 1.5 \)

\( \hat{v}_i = \frac{V_t}{|A_v|+1} = \frac{1.5}{3.3 + 1} = 0.35V \)
5.5.6 The T Equivalent-Circuit Model

(a) $i_g = 0$

(b) $i_g = 0$

(c) $i_g = 0$

(d) $i_g = 0$

S. C. Lin, EE National Chin-Yi University of Technology
Figure 5.41 (a) The T model of the MOSFET augmented with the drain-to-source resistance $r_o$. (b) An alternative representation of the T model.
5.6 Basic MOSFET Amplifier Configurations
5.6.1 The Three Basic Configurations

(a) Common-Source (CS)

(b) Common-Gate (CG)

(c) Common-Drain (CD) or Source Follower
5.6.2 Characterizing Amplifier

\[ v_i = \frac{R_{in}}{R_{in} + R_{sig}} \cdot v_{sig}, \quad v_o = \frac{R_L}{R_o + R_L} \cdot A_{vo} \cdot v_i \]
Input resistance: \( R_{in} \equiv \frac{v_i}{i_i} \)

Open-circuit voltage gain: \( A_{vo} \equiv \left. \frac{v_o}{v_i} \right|_{R_L=\infty} \)

Voltage gain with \( R_L \): \( A_v \equiv \left. \frac{v_o}{v_i} = A_{vo} \frac{R_L}{R_L + R_o} \right|_{R_L=0} \) (5.67)

Short-circuit current gain: \( A_{is} \equiv \left. \frac{i_o}{i_i} \right|_{R_L=0} \)

Current gain: \( A_i \equiv \left. \frac{i_o}{i_i} \right|_{R_L=0} \)
Output resistance: \( R_o \equiv \frac{v_x}{i_x} \bigg|_{v_i=0, R_L=\infty} \)

Overall voltage gain: \( G_v \equiv \frac{v_o}{v_{\text{sig}}} = \frac{v_i}{v_{\text{sig}}} \cdot \frac{v_o}{v_i} = \frac{R_{in}}{R_{in} + R_{\text{sig}}} A_v \)

\[
= \frac{R_{in}}{R_{in} + R_{\text{sig}}} A_{vo} \frac{R_L}{R_L + R_o} \tag{5.68}
\]
5.6.3 The Common-Source (CS) Amplifier

Characterizing Parameters of CS Amplifier

\[ R_{in} = \infty \]  \hspace{1cm} (5.69)

\[ v_o = -(g_m v_{gs})(R_D // r_o) \]

\[ A_{vo} = -g_m \left( R_D // r_o \right) \]

\[ \approx -g_m R_D \]  \hspace{1cm} (5.71)

\[ R_o = R_D // r_o \]

\[ \approx R_D \]  \hspace{1cm} (5.73)
Concludes:

1. The input resistance is ideally infinite.

2. The $R_o \approx R_c$ is moderate to high in value (typically, in the kilohms to tens kilohms range). Reducing $R_D$ to lower $R_o$ is not a viable proportional, because the $A_v$ is also reduced.

3. The open circuit voltage gain $A_{vo}$ can be high, making the CS configuration the workhorse in MOS amplifier design. However the bandwidth of the CS amplifier is severely limited.
Overall Voltage Gain
\[ A_v = \frac{v_o}{v_i} = -g_m \left( R_D \parallel R_L \parallel r_o \right) \]  
(5.75)

\[ v_i = v_{sig} \]  
(5.74)

\[ G_v \equiv \frac{v_o}{v_{sig}} = \frac{v_i}{v_{sig}} \cdot \frac{v_o}{v_i} = -g_m \left( R_D \parallel R_L \parallel r_o \right) \]  
(4.76)

Performing the Analysis Directly on the Circuit
4.6.4 The Common-Source (CS) Amplifier with an Source Resistance

\[ v_{i} \quad R_{s} \quad R_{D} \quad v_{o} \quad R_{o} \]

\[ R_{s} \quad R_{D} \quad 1/g_{m} \quad V_{gs} \quad i \quad v_{o} \quad R_{o} = R_{D} \]

S. C. Lin, EE National Chin-Yi University of Technology
\[ v_{gs} = \frac{\left(1/ g_m\right)}{(1/ g_m) + R_s} \]
\[ v_i = \frac{1}{1 + g_m R_s} \]
\[ i = \frac{v_i}{(1/ g_m) + R_s} = \frac{g_m}{1 + g_m R_s} v_i \]  \hspace{1cm} (5.78)

\[ v_o = -iR_D = \frac{g_m R_D}{1 + g_m R_s} v_i \Rightarrow A_{vo} = \frac{v_o}{v_i} = \frac{g_m R_D}{1 + g_m R_s} \]  \hspace{1cm} (5.80)

\[ A_v (\text{from gate to drain}) = -\frac{\text{Total resistance in drain}}{\text{Total resistance in source}} \]  \hspace{1cm} (5.81)

\[ R_o = R_D \]

Alternatively, \( A_v \) can be obtained by simply replacing \( R_D \) in (5.80) by \( \left( R_D // R_L \right) \)

\[ A_v = \frac{v_o}{v_i} = \frac{R_D // R_L}{1/ g_m + R_s} = \frac{g_m \left( R_D // R_L \right)}{1 + g_m R_s} \]  \hspace{1cm} (5.83)
5.6.5 The Common-Gate (CG) Amplifier

\[ R_{\text{in}} = 1/ g_m \]  \hspace{1cm} (5.84)

\[ v_o = -iR_D, \quad i = -\frac{v_i}{1/ g_m} \]

\[ A_{vo} \equiv \frac{v_o}{v_i} = g_m R_D \] \hspace{1cm} (5.85)

\[ R_o = R_D \] \hspace{1cm} (5.86)
Overall Voltage Gain

\[ \frac{v_i}{v_{sig}} = \frac{R_{in}}{R_{in} + R_{sig}} = \frac{(1/g_m)}{(1/g_m) + R_{sig}} \]  \hspace{1cm} (5.87)

\[ A_v = \frac{v_o}{v_i} = g_m \left( R_D // R_L \right) \]

\[ G_v \equiv \frac{v_o}{v_{sig}} = \frac{v_o}{v_{sig}} \cdot \frac{v_o}{v_i} = \frac{(1/g_m)}{(1/g_m) + R_{sig}} g_m \left( R_D // R_L \right) \]

\[ = \frac{R_D // R_L}{1/g_m + R_{sig}} \]  \hspace{1cm} (5.88)
5.6.6 The CD Amplifier or Source Follower
The Need for Voltage Buffers

\[ R_{\text{sig}} = 1 \text{M} \Omega \]

\[ v_{\text{sig}} = 1 \text{V} \quad R_L = 1 \text{k} \Omega \]

\[ v_o \approx 1 \text{mV} \]

\[ v_{\text{sig}} = 1 \text{V} \]

\[ R_{\text{sig}} = 1 \text{M} \Omega \]

\[ R_o = 100 \Omega \]

\[ v_o \approx 0.9 \text{V} \]

\[ R_{\text{in}} \text{ very large} \]

S. C. Lin, EE National Chin-Yi University of Technology
\( R_{in} = \infty \)

\[
A_v \equiv \frac{V_o}{V_i} = \frac{R_L}{R_L + \left(1/g_m\right)} \quad (5.89), \quad A_{vo} \equiv \frac{V_o}{V_i}_{R_L=\infty} \approx 1 \quad (5.90)
\]

\( R_o = 1/g_m \quad (5.91) \)

### Overall Voltage Gain

\[
G_v \equiv \frac{V_o}{V_{sig}} = \frac{V_i}{V_{sig}} \cdot \frac{V_o}{V_i} = 1 \cdot A_v
\]

\[
= \frac{R_L}{R_L + \left(1/g_m\right)} \approx 1 \quad (5.92)
\]

\( G_v \) will be lower than unity, but close to unity.
### 5.7 Biasing in MOS amplifier circuits

\[
i_D = \frac{1}{2} \mu_n C_{ox} \frac{W}{L} (V_I - V_t)^2 \\quad \text{Biasing by Fixed } V_{GS}
\]

\[
V_G = V_{GS} + I_D R_S
\]

**Figure 5.51** The use of fixed bias (constant \(V_{GS}\)) can result in a large variability in the value of \(I_D\). Devices 1 and 2 represent extremes among units of the same type.
Figure 5.52  Biasing using a fixed voltage at the gate, $V_G$, and a resistance in the source lead, $R_S$: (a) basic arrangement; (b) reduced variability in $I_D$.
(c) practical implementation using a single supply;
(d) coupling of a signal source to the gate using a capacitor $CC_1$;
(e) practical implementation using two supplies.
**Example 5.12** Design the circuit of Fig.5.53 to establish a $I_D = 0.5\text{mA}$. The MOSFET is specified to have $V_t = 1\text{V}$ and $k_n = 1\text{mA/V}^2$. For simplicity, assume $\lambda = 0$. Calculate the percentage change in value of $I_D$ obtained when the MOSFET is replaced with another unit having the same $k_nW/L$ but $V_t = 1.5\text{V}$.

**Solution**

\[(a) R_D = \frac{V_{DD} - V_D}{I_D} = \frac{15 - 10}{0.5} = 10\text{k}\Omega \]

\[R_S = \frac{V_S}{I_D} = \frac{5}{0.5} = 10\text{k}\Omega \]

\[
I_D = \frac{1}{2} k_n (W/L) V_{OV}^2 \approx \frac{1}{2} \cdot 1 \cdot V_{OV}^2
\]

\[\Rightarrow V_{OV} = 1\text{V} \]
Thus. \( V_{GS} = V_t + V_{OV} = 2V \)
\[ V_G = V_S + V_{GS} = 5V + 2V = 7V \]
we may select \( R_{G1} = 8M\Omega \), and \( R_{G2} = 7M\Omega \)  \\

(b) If the NMOSFET is replaced with another having \( V_t = 1.5V \)

\[ I_D = \frac{1}{2} \cdot 1 \cdot (V_{GS} - V_t)^2 \]

\[ V_G = V_{GS} + I_D R_S \Rightarrow V_{GS} = 7 - 10I_D \]

\[ 2I_D = (7 - 10I_D - 1.5)^2 \Rightarrow I_D = 0.455mA \]

\[ \Delta I_D = 0.455 - 0.5 = -0.045mA \]

which \[ \frac{-0.045mA}{0.5mA} \times 100\% = -9\% \]
5.7.3 Biasing Using a Drain-to-Gate Feedback Resistor

Here the large feedback resistance $R_G$ (Usually in the MΩ range) $\Rightarrow I_G = 0$

We can written

$V_{GS} = V_{DS} = V_{DD} - I_D R_D$

or

$V_{DD} = V_{GS} + I_D R_D$

**Figure 5.54** Biasing the MOSFET using a large drain-to-gate feedback resistance, $R_G$. 
5.7.4 Biasing Using a Constant-Current Source.
\[ I_{D_1} = \frac{1}{2} \left( \mu_n C_{ox} \right) \left( \frac{W}{L} \right)_1 (v_{GS} - V_t)^2 \]

\[ I_{D_1} = I_{ref} = \frac{V_{DD} + V_{SS} - V_{GS}}{R} \]

\[ I = I_{D_2} = \frac{1}{2} \left( \mu_n C_{ox} \right) \left( \frac{W}{L} \right)_2 (v_{GS} - V_t)^2 \]

\[ \frac{I_{D_2}}{I_{D_1}} = \frac{I}{I_{ref}} = \frac{\frac{1}{2} \left( \mu_n C_{ox} \right) \left( \frac{W}{L} \right)_2 (v_{GS} - V_t)^2}{\frac{1}{2} \left( \mu_n C_{ox} \right) \left( \frac{W}{L} \right)_1 (v_{GS} - V_t)^2} \]

\[ I = I_{ref} \left( \frac{W}{L} \right)_2 \left( \frac{W}{L} \right)_1 \]
5.8 Discrete-Circuit MOS Amplifier.
5.8.2 The Common-Source Amplifier

\[ V_D = V_{DD} - R_D I_D \]

\[ V_{GS} = V_t + V_{OV} \]

\[ V_{OV} = \sqrt{2I/k'_n \left(\frac{W}{L}\right)} \]
\[ v_o = -g_m v_{gs} \left( r_o \parallel R_D \parallel R_L \right) \]
Analysis of the circuit is straight ward and proceeds in step-by step manner, from the signal source to the amplifier load.

\[ i_g = 0, \quad R_{in} = R_G \]

\[ \nu_o = -g_m \nu_{gs} (r_o // R_D // R_L) \]

\[ A_v \equiv \frac{\nu_o}{\nu_i} = -g_m (r_o // R_D // R_L) \]

The overall voltage gain from the signal-source to the load will be

\[ G_v = \frac{R_{in}}{R_{in} + R_{sig}} A_v = -\frac{R_G}{R_G + R_{sig}} g_m (r_o // R_D // R_L) \] \hspace{1cm} (5.101)

\[ R_{out} = r_o // R_D \]
5.8.3 The Common-Source Amplifier with a Source Resistance
Small-signal equivalent circuit with $r_o$ neglected.

\[ i_d = i = \frac{v_i}{\left( \frac{1}{g_m} + R_s \right)} \]

\[ R_{\text{in}} = R_G \]

\[ A_v = \frac{v_o}{v_i} = -\frac{g_m (R_D // R_L)}{1 + g_m R_s} \]

\[ A_{v0} = A_v \bigg|_{R_L=\infty} = -\frac{g_m R_D}{1 + g_m R_s} \]

\[ G_v = -\frac{R_G}{R_G + R_{\text{sig}}} \cdot \frac{g_m (R_D // R_L)}{1 + g_m R_s} \]
(a) A common-gate amplifier based on the circuit of Fig. 5.56

\[ i_i = g_m v_i \]

\[ v_i = \frac{1}{g_m} + R_{\text{sig}} \]

\[ R_{\text{out}} = R_D \]
(b) A small-signal equivalent circuit of the amplifier in (a).
(c) The common-gate amplifier fed with a current-signal input.

\[ i_i = i_{\text{sig}} \frac{R_{\text{sig}}}{R_{\text{sig}} + \frac{1}{g_m}} \approx i_{\text{sig}} \]

\[ R_{\text{in}} = \frac{1}{g_m} \]
5.8.5 The Common-Drain or Source-Follower Amplifier.
Small-signal equivalent-circuit model.

\[
A_v = \frac{v_o}{v_i} = \frac{R_L \parallel r_o}{(R_L \parallel r_o) + \frac{1}{g_m}}
\]

\[
A_{vo} = A_v \bigg|_{R_L=\infty} = \frac{r_o}{r_o + \frac{1}{g_m}}
\]

\[
G_v = \frac{R_G}{R_G + R_{sig}} \cdot \frac{R_L \parallel r_o}{(R_L \parallel r_o) + \frac{1}{g_m}}
\]

which approaches unity for \( R_G \gg R_{sig}, r_o \gg \frac{1}{g_m} \) and \( r_o \gg R_L \)
5.8.6 The Amplifier Frequency Response

![Diagram showing the frequency response of an amplifier with annotations for low-frequency band, midband, and high-frequency band]

- **Low-frequency band**
  - Gain falls off due to the effect of $C_{C1}$, $C_S$, and $C_{C2}$

- **Midband**
  - All capacitances can be neglected

- **High-frequency band**
  - Gain falls off due to the effect of $C_{gs}$ and $C_{gd}$

$20 \log |A_M|$ (dB)

$f_L$ to $f_H$ (Hz)
5.9 The Body Effect and other Topics

5.9.1 The role of the substrate-The body effect

\[ V_t = V_{t_0} + \gamma \left[ \sqrt{2 \varphi_f + \nu_{SB}} - \sqrt{2 \varphi_f} \right] \]  \hspace{1cm} (5.107)

where \( V_{t_0} \) is the threshold voltage for \( \nu_{SB} = 0 \)

\( \varphi_f \) is a physical parameter with typically 0.3V

\( \gamma \) is a fabrication-process parameter

given by \( \gamma = \frac{\sqrt{2qN_A \varepsilon_s}}{C_{ox}} \), typically \( \gamma = 0.4V^{1/2} \)  \hspace{1cm} (5.108)
Modeling the body effect

The body effect occurs in a MOSFET when the source is not tied to the substrate. We consider the body effect

\[ i_D = \frac{1}{2} k'_n \frac{W}{L} \left( V_{GS} - \left[ V_t + \gamma \left( \sqrt{2\phi_f} + V_{SB} - \sqrt{2\phi_f} \right) \right] \right)^2 \]

\[ g_m \equiv \frac{\partial i_D}{\partial V_{GS}} = k'_n \frac{W}{L} \left\{ V_{GS} - \left[ V_t + \gamma \left( \sqrt{2\phi_f} + V_{SB} - \sqrt{2\phi_f} \right) \right] \right\} \]

\[ g_{mb} \equiv \frac{\partial i_D}{\partial V_{BS}} = \left. \frac{\partial i_D}{\partial V_{BS}} \right|_{V_{GS}, V_{DS} = \text{constant}} \]

\[ = k'_n \frac{W}{L} \left\{ V_{GS} - \left[ V_t + \gamma \left( \sqrt{2\phi_f} + V_{SB} - \sqrt{2\phi_f} \right) \right] \right\} \times \frac{1}{2} \gamma \left( 2\phi_f + V_{SB} \right)^{-\frac{1}{2}} \]

\[ = g_m \frac{\gamma}{2\sqrt{2\phi_f + V_{SB}}} = \chi g_m \quad \text{where} \quad \chi \equiv \frac{\partial V_t}{\partial V_{SB}} = \frac{\gamma}{2\sqrt{2\phi_f + V_{SB}}} \quad (5.111) \]
Figure 5.62  Small-signal equivalent-circuit model of a MOSFET in which the source is not connected to the body.
5.9.3 Temperature effects

(1) $T \uparrow V_t \downarrow (-2mV/°C) \Rightarrow i_D \uparrow$

(2) $T \uparrow k' \downarrow \Rightarrow i_D \downarrow$ (Dominant)

5.9.4 Breakdown and input protection

(1) As the $v_{DS} \uparrow$, a value is reached at which the PN junction between the drain region and substrate suffers avalanche breakdown. Usually occurs at voltages of $20V \leftrightarrow 150V$

(2) Punch-through in drain to source

(3) When the $v_{GS}$ exceeds about $30V$ the oxide breakdown.

The protection mechanism invariably makes use of clamping diodes.
Thanks For Your Attention!

Q & A