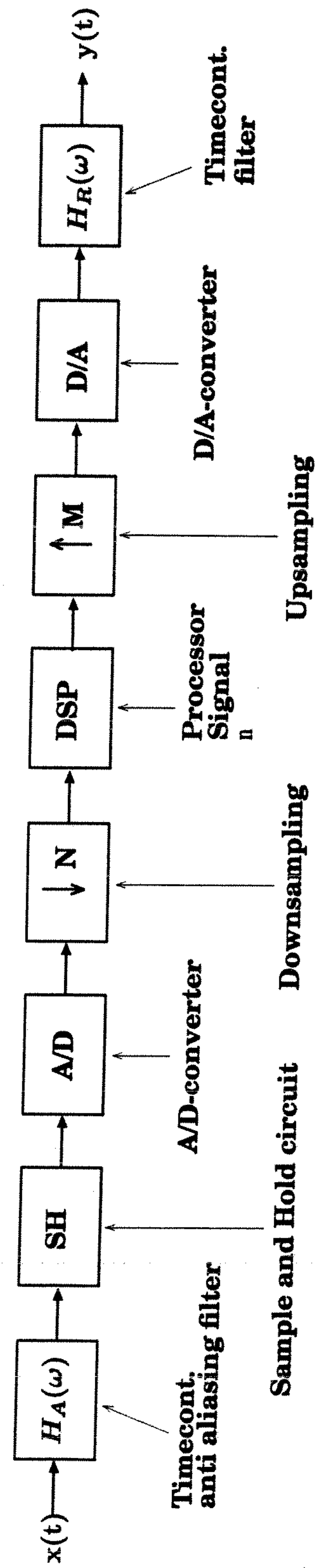
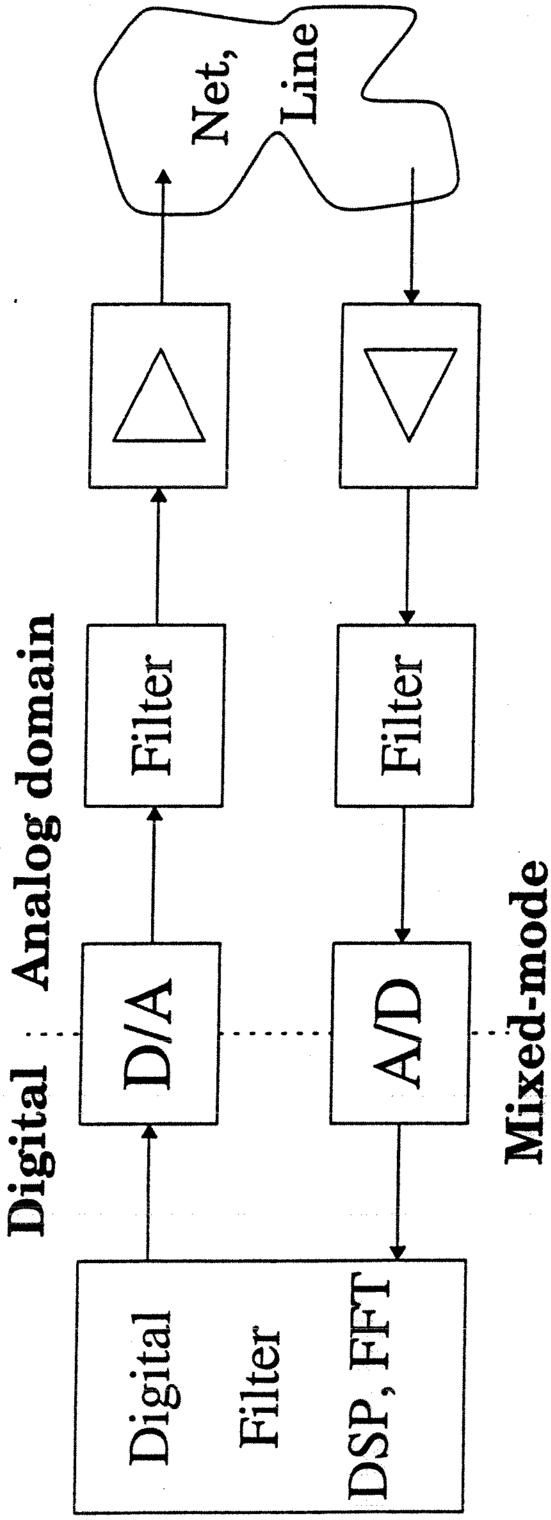
Blockdiagram. Digital Signal Processing of time continuous signals.





Traditional view of the telecommunication system.

ANALOG PERFORMANCE METRICS

TIME DOMAIN	FREQUENCY DOMAIN
• STEP ANSWER g(t)	• SNR (SIGNAL NOISE RATIO)
○ RISE TIME (10-90%)	• THD (TOTAL HARMONIC DISTORSION)
○ SETTLING TIME (5%)	• SNDR (SIGNAL NOISE DISTORSION RATE)
	• SFDR (SPURIOUS-FREE DYNAMIC RANGE)
○ SLEW RATE $\left.\frac{dg(t)}{dt}\right\|_{max}$	• BANDWIDTH
• OVERSHOOT/GLITCH	
• POWER DISSIPATION	• POWER DISSIPATION

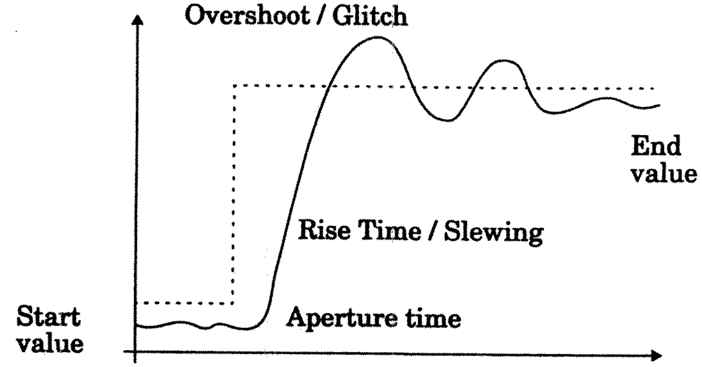


Figure 1.2: Time-domain characteristics.

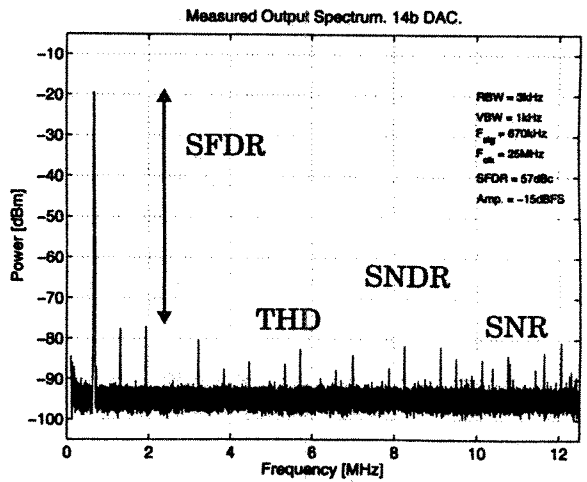


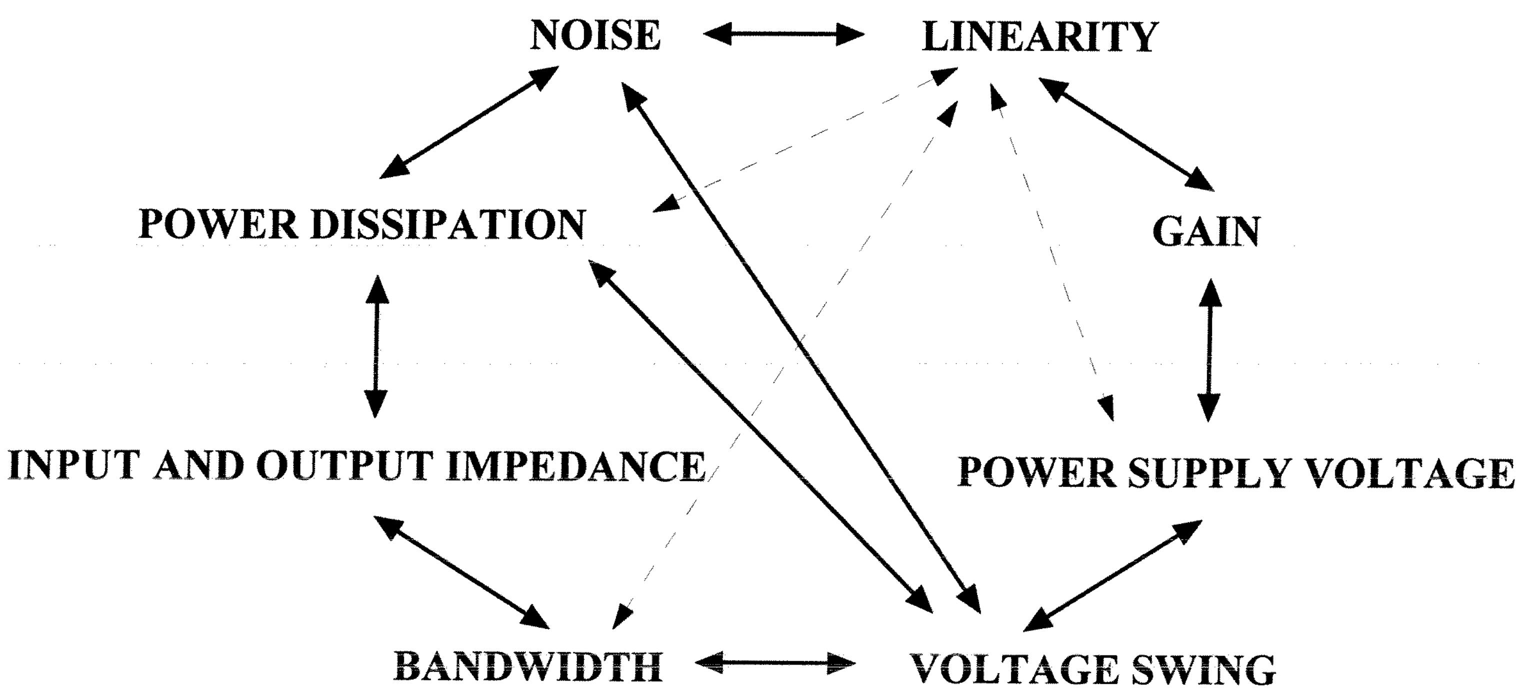
Figure 1.3: Frequency-domain characteristics.

DIGITAL PERFORMANCE METRICS

TIME DOMAIN

- CLOCK FREQUENCY
- NUMBER OF DIGITS
- POWER DISSIPATION

DESIGNOCTAGONE



NOISE AND DISTORSION METRICS

$$SNR = 10\log\frac{P_s}{P_n} \quad [dB]$$

$$THD = 10\log\frac{P_s}{P_{d,harm}} \quad [dB]$$

$$SNDR = 10\log\frac{P_s}{P_{n+d}} \quad [dB]$$

$$SFDR = 10\log\frac{P_s}{P_{d,\max}} \quad [dB]$$

P_s = signal power
P_n = noise power
$P_{d,harm}$ = power of all harmonic distorsion
P_{n+d} = total power of noise and distorsion
$P_{d,\max}$ = power of the strongest spurious peak

Diode Equations

Reverse-Biased Diode (Abrupt Junction)

$C_j = \dfrac{C_{j0}}{\sqrt{1+\dfrac{V_R}{\Phi_0}}}$	$Q = 2C_{j0}\Phi_0\sqrt{1+\dfrac{V_R}{\Phi_0}}$
$C_{j0} = \sqrt{\dfrac{qK_s\varepsilon_0}{2\Phi_0}\dfrac{N_D N_A}{N_A+N_D}}$	$C_{j0} = \sqrt{\dfrac{qK_s\varepsilon_0 N_D}{2\Phi_0}}$ if $N_A >> N_D$
$\Phi_0 = \dfrac{kT}{q}\ln\left(\dfrac{N_A N_D}{n_i^2}\right)$	

Forward-Biased Diode

$I_D = I_S e^{V_D/V_T}$	$I_S = A_D q n_i^2\left(\dfrac{D_n}{L_n N_A}+\dfrac{D_p}{L_p N_D}\right)$
$V_T = \dfrac{kT}{q} \cong 26$ mV at 300 °K	

C_j = diode depletion capacitance

Φ_o = built-in voltage PN-junction

N_D = doping concentration electrons

N_A = doping concentration holes

I_S = scale current

V_T = thermal voltage

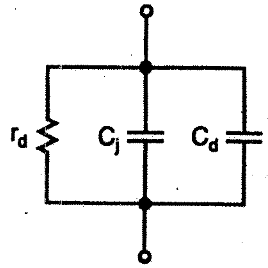
D_n = diffusion constant of electrons, p-side

D_p = diffusion constant of holes, n-side

L_n = diffusion length of electrons, p-side

L_p = diffusion length of holes, n-side

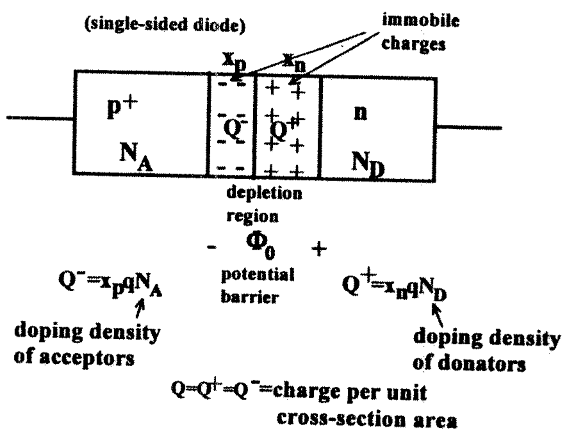
Small-Signal Model of Forward-Biased Diode



$r_d = \dfrac{V_T}{I_D}$	$C_T = C_d + C_j$
$C_d = \tau_T\dfrac{I_D}{V_T}$	$C_j \cong 2C_{j0}$
$\tau_T = \dfrac{L_n^2}{D_n}$	

τ_T = transit time for diode

PN-junction



Constants

$q = 1.602 \times 10^{-19}$ C	$k = 1.38 \times 10^{-23}$ JK^{-1}
$n_i = 1.1 \times 10^{16}$ carriers/m³ at T = 300 °K	$\varepsilon_0 = 8.854 \times 10^{-12}$ F/m
$K_{ox} \cong 3.9$	$K_s \cong 11.8$
$\mu_n = 0.05$ $m^2/V \cdot s$	$\mu_p = 0.02$ $m^2/V \cdot s$

q = elektron charge

k = Bolzmann's constant

n_i = carrier concentration intrinsic silicon

ε_o = permittivity for free space

K_{OX} = relative permittivity for SiO_2

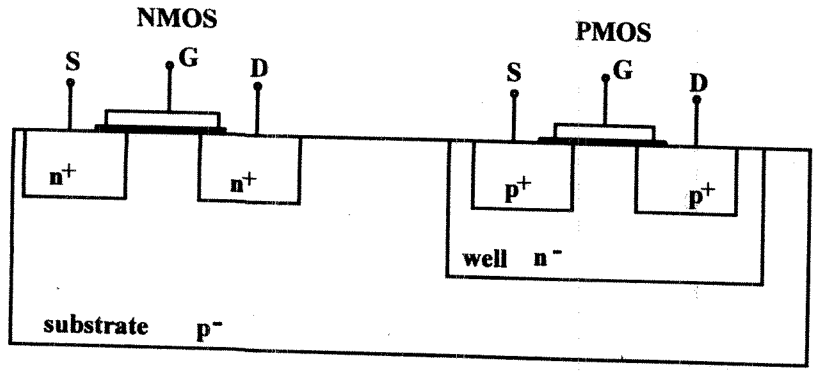
K_S = relative permittivity for Si

μ_n = mobility of electrons

μ_p = mobility of holes

Table 1.1 **Important SPICE parameters for modelling diodes**

SPICE Parameter	Model Constant	Brief Description	Typical Value
IS	I_S	Transport saturation current	10^{-17} A
RS	R_d	Series resistance	30 Ω
TT	τ_T	Diode transit time	12 ps
CJ	C_{j0}	Capacitance at 0-V bias	0.01 pF
MJ	m_j	Diode grading coefficient exponent	0.5
PB	Φ_0	Built-in diode contact potential	0.9 V



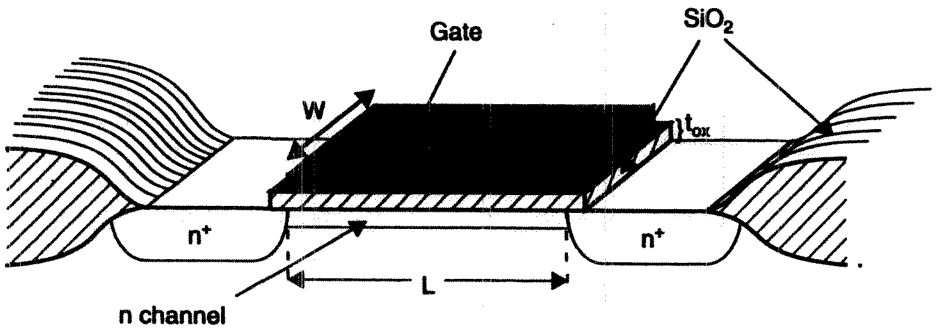


Fig. 1.10 The important dimensions of a MOS transistor.

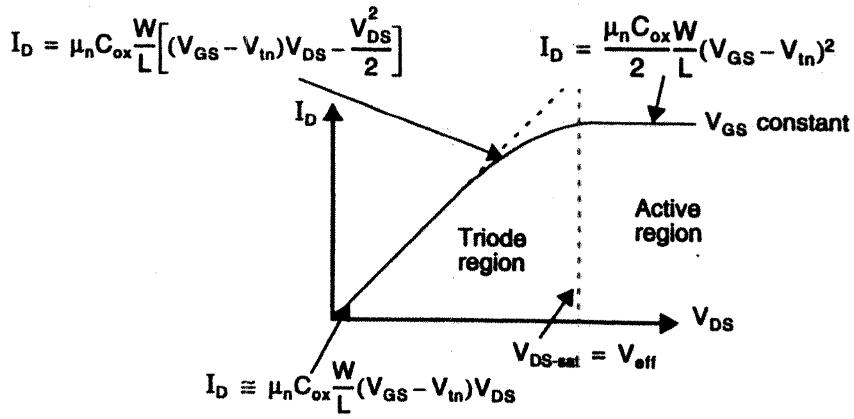


Fig. 1.14 The I_D versus V_{DS} curve for an ideal MOS transistor. For $V_{DS} > V_{DS\text{-sat}}$, I_D is approximately constant.

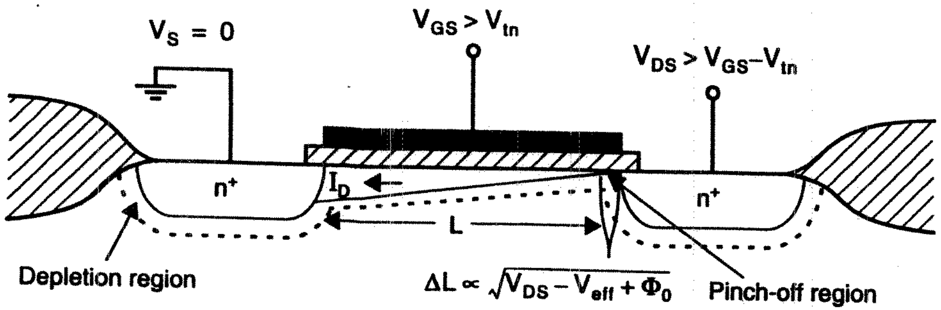


Fig. 1.15 Channel length shortening for $V_{DS} > V_{eff}$.

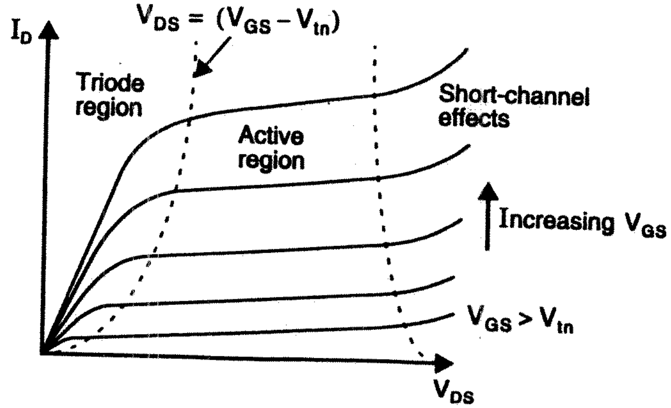


Fig. 1.16 I_D versus V_{DS} for different values of V_{GS}.

MOS Transistor Equations

The following equations are for n-channel devices—for p-channel devices, put negative signs in front of all voltages. These equations do not account for short-channel effects (i.e., $L < 2L_{min}$).

Triode Region ($V_{GS} > V_{tn}$, $V_{DS} \le V_{eff}$)

$I_D = \mu_n C_{ox}\left(\frac{W}{L}\right)\left[(V_{GS}-V_{tn})V_{DS} - \frac{V_{DS}^2}{2}\right]$	
$V_{eff} = V_{GS} - V_{tn}$	$V_{tn} = V_{tn\text{-}0} + \gamma(\sqrt{V_{SB}+2\phi_F} - \sqrt{2\phi_F})$

Active (or Pinch-Off) Region ($V_{GS} > V_{tn}$, $V_{DS} \ge V_{eff}$)

$I_D = \frac{\mu_n C_{ox}}{2}\frac{W}{L}(V_{GS}-V_{tn})^2[1+\lambda(V_{DS}-V_{eff})]$	
$\lambda \propto \frac{1}{L\sqrt{V_{DS}-V_{eff}+\Phi_0}}$	$V_{tn} = V_{tn\text{-}0} + \gamma(\sqrt{V_{SB}+2\phi_F} - \sqrt{2\phi_F})$
$V_{eff} = V_{GS} - V_{tn} = \sqrt{\frac{2I_D}{\mu_n C_{ox} W/L}}$	

$\phi_F = \frac{kT}{q}\ln\left(\frac{N_A}{n_i}\right)$	$\gamma = \frac{\sqrt{2qK_{si}\varepsilon_0 N_A}}{C_{ox}}$
$C_{ox} = \frac{K_{ox}\varepsilon_0}{t_{ox}}$	

V_{tn} = threshold voltage n-chanal

C_{ox} = gate capacitance — t_{ox} = thicknes of insulating SiO_2

W = gate width — L = effective gate lenght

Φ_F = Fermi potential — γ = body-effect constant

λ = output impedance $[V^{-1}]$

V_{tp} = threshold voltage p-chanal (negative)

$C_{S\text{-}SW}$ = capacitance between source and sidewalls

$C_{D\text{-}SW}$ = capacitance between drain and sidewalls

N_B = doping concentration bulk

Table 1.2 A reasonable set of MOS parameters for a typical 0.8-μm technology

SPICE Parameter	Model Constant	Brief Description	Typical Value
VTO	V_{tn}:V_{tp}	Transistor threshold voltage (in V)	0.7:–0.9
UO	μ_n:μ_p	Carrier mobility in bulk (in $cm^2/V{\cdot}s$)	500:175
TOX	t_{ox}	Thickness of gate oxide (in m)	1.8×10^{-8}
LD	L_D	Lateral diffusion of junction under gate (in m)	6×10^{-8}
GAMMA	γ	Body-effect parameter	0.5: 0.8
NSUB	N_A:N_D	The substrate doping (in cm^{-3})	3×10^{16}:7.5×10^{16}
PHI	$\lvert 2\phi_F \rvert$	Surface inversion potential (in V)	0.7
PB	Φ_0	Built-in contact potential of junction to bulk (in V)	0.9
CJ	C_{j0}	Junction-depletion capacitance at 0-V bias (in F/m^2)	2.5×10^{-4}:4.0×10^{-4}
CJSW	$C_{j\text{-}sw0}$	Sidewall capacitance at 0-V bias (in F/m)	2.0×10^{-10}:2.8×10^{-10}
MJ	m_j	Bulk-to-junction exponent (grading coefficient)	0.5
MJSW	$m_{j\text{-}sw}$	Sidewall-to-junction exponent (grading coefficient)	0.3

Typical Values for a 0.8-μm Process

$V_{tn} = 0.8\ V$	$V_{tp} = -0.9\ V$
$\mu_n C_{ox} = 90\ \mu A/V^2$	$\mu_p C_{ox} = 30\ \mu A/V^2$

$C_{ox} = 1.9 \times 10^{-3}\ pF/(\mu m)^2$	$C_j = 2.4 \times 10^{-4}\ pF/(\mu m)^2$
$C_{j\text{-}sw} = 2.0 \times 10^{-4}\ pF/\mu m$	$C_{gs(overlap)} = 2.0 \times 10^{-4}\ pF/\mu m$
$\phi_F = 0.34\ V$	$\Phi_0 = 0.9\ V$
$\gamma = 0.5\ V^{1/2}$	$t_{ox} = 0.02\ \mu m$
$N_B = 6 \times 10^{21}$ impurities$/m^3$	

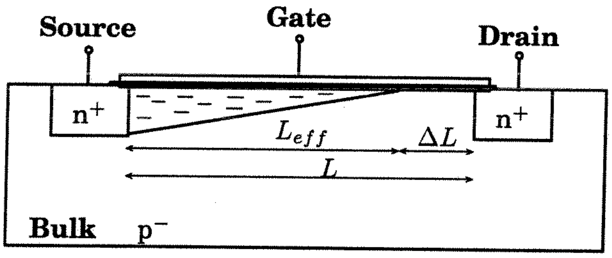
Derivation. RELATION BETWEEN I_D AND V_{DS} WHEN CHANNEL LENGTH MODULATION

At *pinch-off* limit:

$$I_{D,sat} = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)(V_{GS} - V_{tn})^2 \qquad (1)$$

When $V_{DS} > V_{DS,sat}$:

$$I_D = \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L_{eff}}\right)(V_{GS} - V_{tn})^2 \qquad (2)$$



Assumptions: $V_{GS} > V_{tn}$, $V_{DS} > V_{GS} - V_{tn} = V_{eff}$. $N_D >> N_A \Rightarrow \Delta L = x_p + x_n \approx x_p$.

$$L_{eff} = L - \Delta L \qquad (3)$$

$$\begin{aligned}\Delta L \approx x_p &= \frac{Q_-}{q \cdot N_A} = \frac{1}{q \cdot N_A}\sqrt{2qK_s\epsilon_0(\phi_0 + V_R)\frac{N_A N_D}{N_A + N_D}} = \sqrt{\frac{2K_s\epsilon_0(\phi_0 + V_R)}{q} \cdot \frac{N_D}{N_A(N_A + N_D)}} \\ &\approx \sqrt{\frac{2K_s\epsilon_0(\phi_0 + V_R)}{q} \cdot \frac{1}{N_A}} = k_{ds}\sqrt{V_R + \phi_0} = k_{ds}\sqrt{V_{DS} - V_{eff} + \phi_0} \qquad (4)\end{aligned}$$

$k_{ds} = \sqrt{\frac{2K_s\epsilon_0}{qN_A}}$ $\quad V_R = V_{DS} - V_{DS,sat} = V_{DS} - V_{eff}$ $\quad V_{eff} = V_{GS} - V_{tn}$

– V_{DS} increase $\Rightarrow \Delta L$ increase $\Rightarrow L_{eff}$ decrease $\Rightarrow I_D$ increase.

– Determine the relation between I_D and V_{DS}!

$$(2) \Rightarrow I_D = \frac{A}{L_{eff}}(V_{GS} - V_{tn})^2 \qquad (5)$$

$$(1) \Rightarrow I_{D,sat} = \frac{A}{L}(V_{GS} - V_{tn})^2 \qquad (6)$$

Taylorseries around $I_{D,sat}$:

$$I_D = I_{D,sat} + \left.\frac{\partial I_D}{\partial L_{eff}}\right|_{L_{eff}=L} \cdot \left.\frac{\partial L_{eff}}{\partial V_{DS}}\right|_{V_{DS}=V_{DS,sat}} \cdot \Delta V_{DS} \qquad (7)$$

Derivating (5) and (3) gives:

$$\begin{aligned} \frac{\partial I_D}{\partial L_{eff}} &= -\frac{A}{L_{eff}^2}\left(V_{GS}-V_{tn}\right)^2 \\ \frac{\partial L_{eff}}{\partial V_{DS}} &= \frac{\partial(L-\Delta L)}{\partial V_{DS}} = -\frac{\partial \Delta L}{\partial V_{DS}} = -\frac{k_{ds}}{2}\frac{1}{\sqrt{V_{DS}-V_{eff}+\phi_0}} \\ \Delta V_{DS} &= V_{DS}-V_{DS,sat} = V_{DS}-V_{eff} \end{aligned} \tag{8}$$

(6), (7) and (8) give:

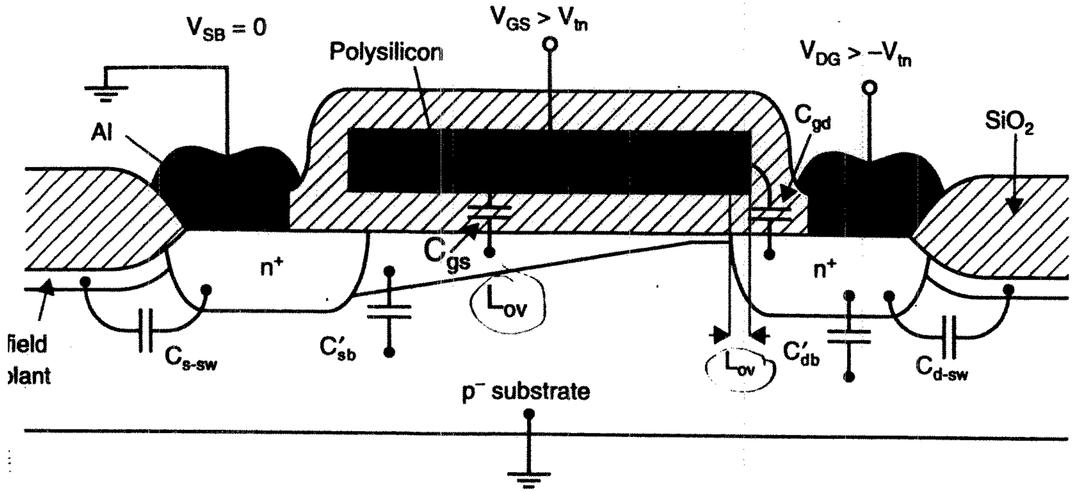
$$\begin{aligned} I_D &= \frac{A}{L}\left(V_{GS}-V_{tn}\right)^2 + \left(-\frac{A}{L^2}\left(V_{GS}-V_{tn}\right)^2\right)\left(-\frac{k_{ds}}{2}\frac{1}{\sqrt{V_{DS,sat}-V_{eff}+\phi_0}}\right)\left(V_{DS}-V_{eff}\right) \\ &= \frac{A}{L}\left(V_{GS}-V_{tn}\right)^2\left(1+\frac{k_{ds}}{2L}\cdot\frac{1}{\sqrt{\phi_0}}\cdot\left(V_{DS}-V_{eff}\right)\right) \\ &= \frac{\mu_n C_{ox}}{2}\left(\frac{W}{L}\right)\left(V_{GS}-V_{tn}\right)^2\left(1+\lambda(V_{DS}-V_{eff})\right) \\ &= \alpha \cdot V_{eff}^2\left(1+\lambda(V_{DS}-V_{eff})\right) \end{aligned}$$

I.e.:

$$I_D = \alpha \cdot V_{eff}^2\left(1+\lambda(V_{DS}-V_{eff})\right) \tag{9}$$

where $\alpha = \frac{\mu_n C_{ox}}{2}\cdot\frac{W}{L}$ and channellength-modulation constant

$\lambda = \frac{k_{ds}}{2L\cdot\sqrt{\phi_0}}$; $\quad k_{ds} = \sqrt{\frac{2K_s\epsilon_0}{qN_A}}$



. 1.20 A cross section of an n-channel MOS transistor showing the small-signal capacitances.

Table 1.2 A reasonable set of MOS parameters for a typical 0.8-μm technology

SPICE Parameter	Model Constant	Brief Description	Typical Value
VTO	$V_{tn}:V_{tp}$	Transistor threshold voltage (in V)	0.7:−0.9
UO	$\mu_n:\mu_p$	Carrier mobility in bulk (in cm²/V·s)	500:175
TOX	t_{ox}	Thickness of gate oxide (in m)	1.8×10^{-8}
LD	L_D	Lateral diffusion of junction under gate (in m)	6×10^{-8}
GAMMA	γ	Body-effect parameter	0.5: 0.8
NSUB	$N_A:N_D$	The substrate doping (in cm⁻³)	$3 \times 10^{16}:7.5 \times 10^{16}$
PHI	$\lvert 2\phi_F \rvert$	Surface inversion potential (in V)	0.7
PB	Φ_0	Built-in contact potential of junction to bulk (in V)	0.9
CJ	C_{j0}	Junction-depletion capacitance at 0-V bias (in F/m²)	$2.5 \times 10^{-4}:4.0 \times 10^{-4}$
CJSW	$C_{j\text{-}sw0}$	Sidewall capacitance at 0-V bias (in F/m)	$2.0 \times 10^{-10}:2.8 \times 10^{-10}$
MJ	m_j	Bulk-to-junction exponent (grading coefficient)	0.5
MJSW	$m_{j\text{-}sw}$	Sidewall-to-junction exponent (grading coefficient)	0.3

Matching

Differentiate the saturation current

$$I_D = K'\frac{W}{L}(V_{GS} - V_T)^2(1 + \lambda V_{DS}) \text{ this gives}$$

$$\frac{\Delta I_D}{I_D} = \frac{\Delta K'}{K'} + \frac{\Delta W}{W} - \frac{\Delta L}{L} + 2\frac{\Delta V_{GS} - \Delta V_T}{V_{GS} - V_T} + \frac{\Delta V_{DS}\lambda + V_{DS}\lambda}{1 + \lambda V_{DS}}$$

Rule of thumb 1

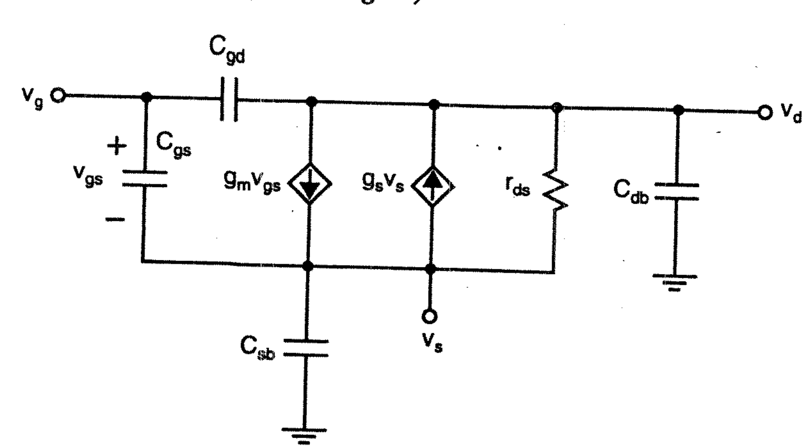
$V_{GS} - V_T = V_{eff} > 0.15–0.25$ to obtain good V_T-matching.

Rule of thumb 2

Choose $L \geq 1.5 L_{min}$ for good β-matching.

Choose $W \gg W_{min}$ for high gain and good β-matching.

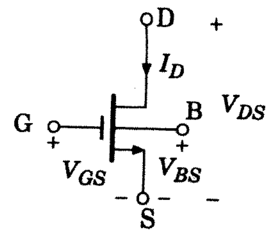
Small-Signal Model (Active Region)

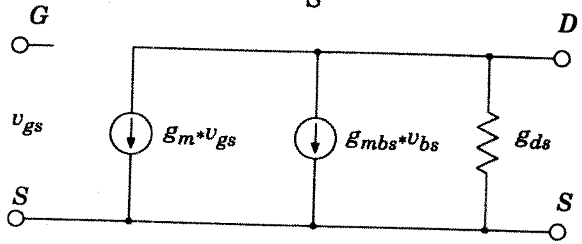


$g_m = \mu_n C_{ox}\left(\frac{W}{L}\right)V_{eff}$	$g_m = \sqrt{2\mu_n C_{ox}(W/L)I_D}$
$g_m = \frac{2I_D}{V_{eff}}$	$g_s = \frac{\gamma g_m}{2\sqrt{V_{SB}+\|2\phi_F\|}}$
$r_{ds} = \frac{1}{\lambda I_D}$	$g_s \cong 0.2g_m$
$\lambda = \frac{k_{rds}}{2L\sqrt{V_{DS}-V_{eff}+\Phi_0}}$	$k_{rds} = \sqrt{\frac{2K_s\varepsilon_0}{qN_A}}$
$C_{gs} = \frac{2}{3}WLC_{ox}+WL_{ov}C_{ox}$	$C_{gd} = WL_{ov}C_{ox}$
$C_{sb} = (A_s+WL)C_{js}+P_sC_{j\text{-}sw}$	$C_{js} = \frac{C_{j0}}{\sqrt{1+V_{SB}/\Phi_0}}$
$C_{db} = A_dC_{jd}+P_dC_{j\text{-}sw}$	$C_{jd} = \frac{C_{j0}}{\sqrt{1+V_{DB}/\Phi_0}}$

Small-signal model (Linear region)

$$I_D = \beta\left[(V_{GS}-V_T)V_{DS}-\frac{V_{DS}^2}{2}\right]$$



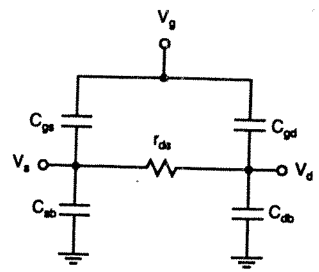


$$g_{ds} = \frac{\partial I_D}{\partial V_{DS}} = \beta[V_{GS}-V_T-V_{DS}]$$

$$g_m = \frac{\partial I_D}{\partial V_{GS}} = \beta V_{DS} \approx 0$$

$$g_{mbs} = \frac{\partial I_D}{\partial V_{BS}} = \eta g_m \approx 0$$

Small-Signal Model in Triode Region (for $V_{DS} << V_{eff}$)



$r_{ds} = \frac{1}{\mu_n C_{ox}\left(\frac{W}{L}\right)V_{eff}}$	
$C_{gd} = C_{gs} \cong \frac{1}{2}WLC_{ox}+WL_{ov}C_{ox}$	$C_{sb} = C_{db} = \frac{C_{j0}(A_s+WL/2)}{\sqrt{1+\frac{V_{sb}}{\Phi_0}}}$

DETERMINE g_m AND g_{ds}

$$g_m = \frac{2I_{DQ}}{V_{GSQ} - V_t} \qquad g_{ds} = \frac{1}{r_{ds}} = \lambda I_{DQ} \qquad \frac{g_m}{g_{ds}} = \frac{2}{\lambda(V_{GSQ} - V_t)}$$

Regard a 0.35 µm CMOS-process ($L_{min} = 0.35$ µm):

	NMOS L=1 µm	PMOS L=1 µm	
λ	0.03	0.05	[1/V]
V_t	0.47	0.62	[V]
μ	400	130	[cm²/Vs]
C_{ox}	$4.5\cdot10^{-7}$	$4.5\cdot10^{-7}$	[F/cm²]

For good matching choose $V_{GSQ} - V_t$ in the interval [0.15,0.25] V.
Choose e.g.. $V_{GSQ} - V_t = 0.25$ V

NMOS: $\frac{g_m}{g_{ds}} = \frac{2}{0.03\cdot0.25} \approx 267$

PMOS: $\frac{g_m}{g_{ds}} = \frac{2}{0.05\cdot0.25} = 160$

Determine g_m: $g_m = \mu C_{ox}\frac{W}{L}(V_{GSQ} - V_t)$

Suppose: $W = 10$ µm, $L = 1$ µm och $V_{GSQ} - V_t = 0.25$ V

NMOS: $g_m = 400\cdot4.5\cdot10^{-7}\cdot10\cdot0.25 = 0.45\cdot10^{-3}$ [S]

PMOS: $g_m = 130\cdot4.5\cdot10^{-7}\cdot10\cdot0.25 \approx 0.15\cdot10^{-3}$ [S]

Determine g_{ds}:

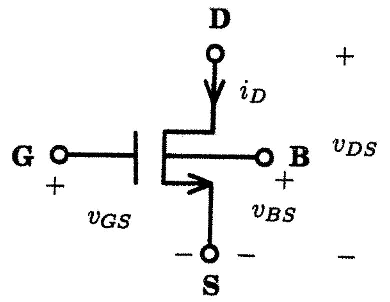
NMOS: $g_{ds} = \frac{g_m}{267} \approx 1.69\cdot10^{-6}$ [S] $\Rightarrow r_{ds} \approx 0.59$ MΩ

PMOS: $g_{ds} = \frac{g_m}{160} \approx 0.9\cdot10^{-6}$ [S] $\Rightarrow r_{ds} \approx 1.09$ MΩ

DERIVATION OF A SMALL SIGNAL EQUIVALENT FOR NMOS-TRANSISTOR

Regard i_D as a function of v_{GS}, v_{BS} and v_{DS}:

$$i_D = f(v_{GS}, v_{BS}, v_{DS}) \tag{1}$$



i_D, as well as v_{GS}, v_{BS} and v_{DS} consists of an DC-part, I_{DQ}, V_{GSQ}, V_{BSQ} and V_{DSQ} (the quiscent point) and an AC-part i_d, v_{gs}, v_{bs} and v_{ds} (the small signal).

$$i_D = I_{DQ} + i_d, v_{GS} = V_{GSQ} + v_{gs}, v_{BS} = V_{BSQ} + v_{bs}, v_{DS} = V_{DSQ} + v_{ds} \tag{2}$$

That is:

$$I_{DQ} + i_d = f(V_{GSQ} + v_{gs}, V_{BSQ} + v_{bs}, V_{DSQ} + v_{ds}) \tag{3}$$

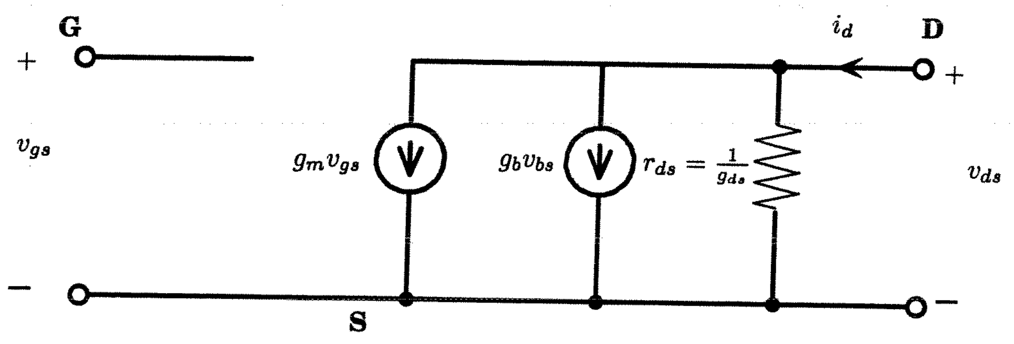
Taylor's formula for 3 variables gives:

$$I_{DQ} + i_d = \underbrace{f(V_{GSQ}, V_{BSQ}, V_{DSQ})}_{I_{DQ}} + \underbrace{\left.\frac{\partial f}{\partial v_{GS}}\right|_Q}_{g_m} \cdot v_{gs} + \underbrace{\left.\frac{\partial f}{\partial v_{BS}}\right|_Q}_{g_b} \cdot v_{bs} + \underbrace{\left.\frac{\partial f}{\partial v_{DS}}\right|_Q}_{g_{ds}} \cdot v_{ds} \tag{4}$$

That is:

$$i_d = g_m \cdot v_{gs} + g_b \cdot v_{bs} + g_{ds} \cdot v_{ds} \tag{5}$$

Equation (5) gives following equivalent small signal scheme:



To calculate the values of g_m, g_b and g_{ds} we first have to determine in which region the transistor is operating.

(CONT. ↦)

We presume that the transistor operates in the SATURATION region, where following relations applies: (Notice that $v_{BS} = -v_{SB}$.)

$$i_D = \alpha(v_{GS} - V_t)^2(1 + \lambda(v_{DS} - V_{eff})) \tag{6}$$

$$V_t = V_{t0} + \gamma\left(\sqrt{2|\phi_F| - v_{BS}} - \sqrt{2|\phi_F|}\right) \tag{7}$$

To determine g_m we derivate (6) with respect to v_{GS}:

$$\frac{\partial i_D}{\partial v_{GS}} = 2\alpha(v_{GS} - V_t)(1 + \lambda(v_{DS} - V_{eff}))$$

The value of this derivative in the Q-point gives g_m:

$$g_m = 2\alpha(V_{GSQ} - V_t)(1 + \lambda(V_{DSQ} - V_{eff})) = \frac{2I_{DQ}}{V_{GSQ} - V_t} \tag{8}$$

Neglecting the channel-length modulation gives the approximation:

$$g_m \approx 2\sqrt{\alpha I_{DQ}}$$

To determine g_{ds} we derivate (6) with respect to v_{DS} and calculate the value in the Q-point.

$$\frac{\partial i_D}{\partial v_{DS}} = \lambda \cdot \alpha(v_{GS} - V_t)^2$$

That is:

$$g_{ds} = \lambda \cdot \alpha(V_{GSQ} - V_t)^2 \approx \lambda I_{DQ} \tag{9}$$

To determine g_b we have to use the chain-rule for derivatives, as we don't explicit have i_D as a function of v_{BS}.

$$\frac{\partial i_D}{\partial v_{BS}} = \frac{\partial i_D}{\partial V_t} \cdot \frac{\partial V_t}{\partial v_{BS}} = -2\alpha(v_{gs} - V_t)(1 + \lambda(v_{DS} - V_{eff})) \cdot \left(-\frac{\gamma}{2}(2|\phi_F| - v_{BS})^{-1/2}\right)$$

That is:

$$g_b = -\underbrace{2\alpha(V_{GSQ} - V_t)(1 + \lambda(V_{DSQ} - V_{eff}))}_{g_m} \cdot \left(-\frac{\gamma}{2}(2|\phi_F| - V_{BSQ})^{-1/2}\right) = g_m \cdot \frac{\gamma}{2\sqrt{2|\phi_F| - V_{BSQ}}} = \eta \cdot g_m \tag{10}$$

Conclusion:

$$\begin{cases} g_m = \dfrac{2I_{DQ}}{V_{GSQ} - V_t} \approx 2\sqrt{\alpha I_{DQ}} & \alpha \sim \dfrac{W}{L}; \quad g_m \sim \sqrt{\dfrac{W}{L}} \\ g_{ds} = \lambda \cdot \alpha(V_{GSQ} - V_t)^2 \approx \lambda I_{DQ} & \lambda \sim \dfrac{1}{L}; \quad g_{ds} \sim \dfrac{1}{L} \\ g_b = g_m \cdot \dfrac{\gamma}{2\sqrt{2|\phi_F| - V_{BSQ}}} = \eta \cdot g_m & \eta \approx 0.2; \quad g_b \sim \sqrt{\dfrac{W}{L}} \end{cases}$$