BSIM3v3.2 MOSFET MODEL
Users’ Manual

by

Weidong Liu, Xiaodong Jin, James Chen, Min-Chie Jeng, Zhihong Liu, Yuhua Cheng, Kai Chen, Mansun Chan, Kelvin Hui, Jianhui Huang, Robert Tu, Ping K. Ko and Chenming Hu

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CHAPTER 1: Introduction

1.1 General Information

BSIM3v3 is the latest industry-standard MOSFET model for deep-submicron digital and analog circuit designs from the BSIM Group at the University of California at Berkeley. BSIM3v3.2 is based on its predecessor, BSIM3v3.1. Its many improvements and enhancements include

- A new intrinsic capacitance model (the Charge Thickness Model), considering the finite charge layer thickness determined by quantum effect, is introduced as capMod 3. It is very accurate in all operating regions.
- Modeling of C-V characteristics at the weak-to-inversion transition is improved.
- The $T_{ox}$ dependence is added into the threshold voltage model.
- The flat-band voltage is added as a new model parameter to accurately model MOSFET's with different gate materials.
- Substrate current dependence on the channel length is improved.
- The non-quasi-static (NQS) model is restructured to improve the model accuracy and simulation efficiency. NQS is added in the pole-zero analysis.
- The temperature dependence is added to the diode junction capacitance model.
- The DC junction diode model now supports a resistance-free diode model and a current-limiting feature.
- Option of using C-V inversion charge equations of capMod 0, 1, 2 or 3 to calculate the thermal noise when noiMod == 2 or 4 is added.
- The small negative capacitance of $C_{gs}$ and $C_{gd}$ in the accumulation-depletion regions is eliminated.
- A separate set of length/width-dependence parameters is introduced in the C-V model to better fit the capacitance data.
- Parameter checking is added to avoid bad values for certain parameters.
- Known bugs are fixed.
1.2 Backward Compatibility

Meticulous care has been taken to accomplish the above model enhancements with high levels of accuracy and fast SPICE convergence properties. In addition, every effort has been made to maintain the backward compatibility with the previous version except for the following change:

- Re-implementation of NQS model;
- Junction diode I-V model;
- Using C-V $Q_{int}$ for BSIM3 thermal noise evaluation;
- Zero-bias $V_{fb}$ for capMod = 1 and 2;
- Removing $P_d$ and $P_s$ clamps;
- Using $Leffcv$ for $AbulkCVfactor$;
- Bug fixes (Refer to “BSIM3v3.2 Model Enhancements” at http://www-device.eecs.berkeley.edu/~bsim3).

1.3 Organization of This Manual

This manual describes the BSIM3v3.2 model in the following manner:

- Chapter 2 discusses the physical basis used to derive the I-V model.
- Chapter 3 highlights a single-equation I-V model for all operating regimes.
- Chapter 4 presents C-V modeling and focuses on the charge thickness model.
- Chapter 5 describes in detail the restrutured NQS (Non-Quasi-Static) Model.
- Chapter 6 discusses model parameter extraction.
- Chapter 7 provides some benchmark test results to demonstrate the accuracy and performance of the model.
- Chapter 8 presents the noise model.
Organization of This Manual

- Chapter 9 describes the MOS diode I-V and C-V models.
- The Appendices list all model parameters, equations and references.
CHAPTER 2: Physics-Based Derivation of I-V Model

The development of BSIM3v3 is based on Poisson's equation using gradual channel approximation and coherent quasi 2D analysis, taking into account the effects of device geometry and process parameters. BSIM3v3.2 considers the following physical phenomena observed in MOSFET devices [1]:

- Short and narrow channel effects on threshold voltage.
- Non-uniform doping effect (in both lateral and vertical directions).
- Mobility reduction due to vertical field.
- Bulk charge effect.
- Velocity saturation.
- Drain-induced barrier lowering (DIBL).
- Channel length modulation (CLM).
- Substrate current induced body effect (SCBE).
- Subthreshold conduction.
- Source/drain parasitic resistances.

2.1 Non-Uniform Doping and Small Channel Effects on Threshold Voltage

Accurate modeling of threshold voltage ($V_{th}$) is one of the most important requirements for precise description of device electrical characteristics. In addition, it serves as a useful reference point for the evaluation of device operation regimes. By using threshold voltage, the whole device operation regime can be divided into three operational regions.
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

First, if the gate voltage is greater than the threshold voltage, the inversion charge density is larger than the substrate doping concentration and MOSFET is operating in the strong inversion region and drift current is dominant. Second, if the gate voltage is smaller than $V_{th}$ the inversion charge density is smaller than the substrate doping concentration. The transistor is considered to be operating in the weak inversion (or subthreshold) region. Diffusion current is now dominant [2]. Lastly, if the gate voltage is very close to $V_{th}$ the inversion charge density is close to the doping concentration and the MOSFET is operating in the transition region. In such a case, diffusion and drift currents are both important.

For MOSFET's with long channel length/width and uniform substrate doping concentration, $V_{th}$ is given by [2]:

$$V_{th} = V_{FB} + \Phi_s + \gamma (\Phi_s - V_{bs}) = V_{Tideal} + \gamma (\sqrt{\Phi_s - V_{bs}} - \sqrt{\Phi_s})$$

where $V_{FB}$ is the flat band voltage, $V_{Tideal}$ is the threshold voltage of the long channel device at zero substrate bias, and $\gamma$ is the body bias coefficient and is given by:

$$\gamma = \frac{\sqrt{2\varepsilon_s q N_a}}{C_{ox}}$$

where $N_a$ is the substrate doping concentration. The surface potential is given by:

$$\Phi_s = 2k_B T \ln \left( \frac{N_a}{n_i} \right)$$
Equation (2.1.1) assumes that the channel is uniform and makes use of the one dimensional Poisson equation in the vertical direction of the channel. This model is valid only when the substrate doping concentration is constant and the channel length is long. Under these conditions, the potential is uniform along the channel. Modifications have to be made when the substrate doping concentration is not uniform and/or when the channel length is short, narrow, or both.

2.1.1 Vertical Non-Uniform Doping Effect

The substrate doping profile is not uniform in the vertical direction as shown in Figure 2-1.

The substrate doping concentration is usually higher near the Si/SiO₂ interface (due to $V_{th}$ adjustment) than deep into the substrate. The distribution of impurity atoms inside the substrate is approximately a half
gaussian distribution, as shown in Figure 2-1. This non-uniformity will make \( \gamma \) in Eq. (2.1.2) a function of the substrate bias. If the depletion width is less than \( X_f \) as shown in Figure 2-1, \( N_a \) in Eq. (2.1.2) is equal to \( N_{ch} \); otherwise it is equal to \( N_{sub} \).

In order to take into account such non-uniform substrate doping profile, the following \( V_{th} \) model is proposed:

\[
V_{th} = V_{Tideal} + K_1 \left( \sqrt{\Phi_s - V_{bs} - \sqrt{\Phi_s}} \right) - K_2 V_{bs}
\]

For a zero substrate bias, Eqs. (2.1.1) and (2.1.4) give the same result. \( K_1 \) and \( K_2 \) can be determined by the criteria that \( V_{th} \) and its derivative versus \( V_{bs} \) should be the same at \( V_{bm} \), where \( V_{bm} \) is the maximum substrate bias voltage. Therefore, using equations (2.1.1) and (2.1.4), \( K_1 \) and \( K_2 \) [3] will be given by the following:

\[
K_1 = \gamma_2 - 2 K_2 \sqrt{\Phi_s - V_{bm}}
\]

\[
K_2 = \frac{(\gamma_1 - \gamma_2) \left( \sqrt{\Phi_s - V_{bs} - \sqrt{\Phi_s}} \right)}{2 \sqrt{\Phi_s} \left( \sqrt{\Phi_s - V_{bm} - \sqrt{\Phi_s}} + V_{bm} \right)}
\]

where \( \gamma_1 \) and \( \gamma_2 \) are body bias coefficients when the substrate doping concentration are equal to \( N_{ch} \) and \( N_{sub} \), respectively:

\[
\gamma_1 = \frac{\sqrt{2q \varepsilon_{si} N_{ch}}}{C_{ox}}
\]
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

\[ \gamma_2 = \frac{\sqrt{2}q\varepsilon_{si}N_{sub}}{C_{ox}} \] (2.1.8)

\( V_{bx} \) is the body bias when the depletion width is equal to \( X_f \). Therefore, \( V_{bx} \) satisfies:

\[ \frac{qN_{ch}X_f^2}{2\varepsilon_{si}} = \Phi_x - V_{bx} \] (2.1.9)

If the devices are available, \( K_1 \) and \( K_2 \) can be determined experimentally. If the devices are not available but the user knows the doping concentration distribution, the user can input the appropriate parameters to specify doping concentration distribution (e.g. \( N_{ch} \), \( N_{sub} \) and \( X_f \)). Then, \( K_1 \) and \( K_2 \) can be calculated using equations (2.1.5) and (2.1.6).

### 2.1.2 Lateral Non-Uniform Doping Effect

For some technologies, the doping concentration near the source/drain is higher than that in the middle of the channel. This is referred to as lateral non-uniform doping and is shown in Figure 2-2. As the channel length becomes shorter, lateral non-uniform doping will cause \( V_{th} \) to increase in magnitude because the average doping concentration in the channel is larger. The average channel doping concentration can be calculated as follows:
Due to the lateral non-uniform doping effect, Eq. (2.1.4) becomes:

\[ V_{th} = V_{th0} + K_1 \left( \sqrt{\Phi_s - V_{bb}} - \sqrt{\Phi_s} \right) - K_2 V_{bb} \]

\[ + K_1 \left( \frac{1 + \frac{N_l x}{L_{eff}}}{1} - 1 \right) \sqrt{\Phi_s} \]

Eq. (2.1.11) can be derived by setting \( V_{bb} = 0 \), and using \( K_1 \propto (N_{\text{eff}})^{0.5} \). The fourth term in Eq. (2.1.11) is used to model the body bias dependence of the lateral non-uniform doping effect. This effect gets stronger at a lower body bias. Examination of Eq. (2.1.11) shows that the threshold voltage will increase as channel length decreases [3].

Figure 2-2. Lateral doping profile is non-uniform.
2.1.3 Short Channel Effect

The threshold voltage of a long channel device is independent of the channel length and the drain voltage. Its dependence on the body bias is given by Eq. (2.1.4). However, as the channel length becomes shorter, the threshold voltage shows a greater dependence on the channel length and the drain voltage. The dependence of the threshold voltage on the body bias becomes weaker as channel length becomes shorter, because the body bias has less control of the depletion region. The short-channel effect is included in the $V_{th}$ model as:

$$V_{th} = V_{th0} + K_1 \left( \sqrt{\Phi_s - V_{bs}} - \sqrt{\Phi_s} \right) - K_2 V_{bs}$$

$$+ K_1 \left( \sqrt{1 + \frac{N L_x}{L_{eff}}} - 1 \right) \sqrt{\Phi_s - \Delta V_{th}}$$

where $\Delta V_{th}$ is the threshold voltage reduction due to the short channel effect. Many models have been developed to calculate $\Delta V_{th}$. They used either numerical solutions [4], a two-dimensional charge sharing approach [5,6], or a simplified Poisson's equation in the depletion region [7-9]. A simple, accurate, and physical model was developed by Z. H. Liu et al. [10]. This model was derived by solving the quasi 2D Poisson equation along the channel. This quasi-2D model concluded that:

$$\Delta V_{th} = \theta_{th} (L) \left( 2(V_{bi} - \Phi_s) + V_{ds} \right)$$

where $V_{bi}$ is the built-in voltage of the PN junction between the source and the substrate and is given by
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

\[ V_{bi} = \frac{K_g T}{q} \ln\left(\frac{N_{ch} N_d}{n_i^2}\right) \]  

(2.1.14)

where \( N_d \) is the source/drain doping concentration with a typical value of around \( 1 \times 10^{20} \) cm\(^{-3}\). The expression \( \theta_{ih}(L) \) is a short channel effect coefficient, which has a strong dependence on the channel length and is given by:

\[ \theta_{ih}(L) = \left[ \exp\left(-L/2l_t\right) + 2 \exp\left(-L/l_t\right) \right] \]  

(2.1.15)

\( l_t \) is referred to as the characteristic length and is given by

\[ l_t = \sqrt{\frac{\varepsilon_{si} T_{ox} X_{dep}}{\varepsilon_{ox} \eta}} \]  

(2.1.16)

\( X_{dep} \) is the depletion width in the substrate and is given by

\[ X_{dep} = \sqrt{\frac{2 \varepsilon_{si} (\Phi_s - V_{bs})}{q N_{ch}}} \]  

(2.1.17)

\( X_{dep} \) is larger near the drain than in the middle of the channel due to the drain voltage. \( X_{dep} / \eta \) represents the average depletion width along the channel.

Based on the above discussion, the influences of drain/source charge sharing and DIBL effects on \( V_{th} \) are described by (2.1.15). In order to make the model fit different technologies, several parameters such as \( D_{vth}, D_{v2}, \)
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

$D_{sub}$, $EtaO$ and $Etab$ are introduced, and the following modes are used to account for charge sharing and DIBL effects separately.

\[
\theta_{th} (L) = D_{vl0} \left[ \exp(-D_{vl1} L / 2l_c) + 2 \exp(-D_{vl1} L / l_c) \right]
\]

(2.1.18)

\[
\Delta V_{th}(L) = \theta_{th}(L)(V_{bi} - \Phi_s)
\]

(2.1.19)

\[
l_c = \sqrt{\frac{\epsilon_{si}/T_{ox} X_{dep}}{\epsilon_{ox}}}(1 + D_{vl2} V_{bs})
\]

(2.1.20)

\[
\theta_{dib}(L) = [\exp(-D_{sub1} L / 2l_{10}) + 2 \exp(-D_{sub1} L / l_{10})]
\]

(2.1.21)

\[
\Delta V_{th}(V_{ds}) = \theta_{dib}(L)(E_{ta0} + E_{tab} V_{ds})V_{ds}
\]

(2.1.22)

where $l_{10}$ is calculated by Eq. (2.1.20) at zero body-bias. $D_{vl1}$ is basically equal to $1/(\eta)^{1/2}$ in Eq. (2.1.16). $D_{vl2}$ is introduced to take care of the dependence of the doping concentration on substrate bias since the doping concentration is not uniform in the vertical direction of the channel. $X_{dep}$ is calculated using the doping concentration in the channel ($N_{ch}$). $D_{vl0}$, $D_{vl1}, D_{vl2}$, $EtaO$, $Etab$ and $D_{sub}$, which are determined experimentally, can improve accuracy greatly. Even though Eqs. (2.1.18), (2.1.21) and (2.1.15) have different coefficients, they all still have the same functional forms. Thus the device physics represented by Eqs. (2.1.18), (2.1.21) and (2.1.15) are still the same.
As channel length $L$ decreases, $\Delta V_{th}$ will increase, and in turn $V_{th}$ will decrease. If a MOSFET has a LDD structure, $N_d$ in Eq. (2.1.14) is the doping concentration in the lightly doped region. $V_{bi}$ in a LDD-MOSFET will be smaller as compared to conventional MOSFET's; therefore the threshold voltage reduction due to the short channel effect will be smaller in LDD-MOSFET's.

As the body bias becomes more negative, the depletion width will increase as shown in Eq. (2.1.17). Hence $\Delta V_{th}$ will increase due to the increase in $I_t$. The term:

$$V_{T_{ideal}} + K_1\sqrt{\Phi_s - V_{bs}} - K_2V_{bs}$$

will also increase as $V_{bs}$ becomes more negative (for NMOS). Therefore, the changes in

$$V_{T_{ideal}} + K_1\sqrt{\Phi_s - V_{bs}} - K_2V_{bs}$$

and in $\Delta V_{th}$ will compensate for each other and make $V_{th}$ less sensitive to $V_{bs}$. This compensation is more significant as the channel length is shortened. Hence, the $V_{th}$ of short channel MOSFET's is less sensitive to body bias as compared to a long channel MOSFET. For the same reason, the DIBL effect and the channel length dependence of $V_{th}$ are stronger as $V_{bs}$ is made more negative. This was verified by experimental data shown in Figure 2-3 and Figure 2-4. Although Liu et al. found an accelerated $V_{th}$ roll-off and non-linear drain voltage dependence [10] as the channel became very short, a linear dependence of $V_{th}$ on $V_{ds}$ is nevertheless a good approximation for circuit simulation as shown in Figure 2-4. This figure shows that Eq. (2.1.13) can fit the experimental data very well.
Furthermore, Figure 2-5 shows how this \( V_{th} \) model can fit various channel lengths under various bias conditions.

**Figure 2-3.** Threshold voltage versus the drain voltage at different body biases.

**Figure 2-4.** Channel length dependence of threshold voltage.
2.1.4 Narrow Channel Effect

The actual depletion region in the channel is always larger than what is usually assumed under the one-dimensional analysis due to the existence of fringing fields [2]. This effect becomes very substantial as the channel width decreases and the depletion region underneath the fringing field becomes comparable to the "classical" depletion layer formed from the vertical field. The net result is an increase in $V_{th}$. It is shown in [2] that this increase can be modeled as:

\[
\frac{\pi q N_e X_{d,\text{max}}^2}{2C_{\text{ox}}W} = 3\pi \frac{T_{\text{ox}}}{W} \Phi:
\]
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

The right hand side of Eq. (2.1.23) represents the additional voltage increase. This change in $V_{th}$ is modeled by Eq. (2.1.24a). This formulation includes but is not limited to the inverse of channel width due to the fact that the overall narrow width effect is dependent on process (i.e. isolation technology) as well. Hence, parameters $K_3$, $K_{3b}$, and $W_0$ are introduced as

$$ (K_3 + K_{3b} V_{he}) \frac{T_{ox}}{W_{eff} + W_0} \Phi_s $$

$(2.1.24a)$

$W_{eff}$ is the effective channel width (with no bias dependencies), which will be defined in Section 2.9. In addition, we must consider the narrow width effect for small channel lengths. To do this we introduce the following:

$$ D_{VT0w} \left( \exp\left(-D_{VT1w} \frac{W_{eff} \cdot L_{eff}}{2l_{n}} \right) + 2 \exp\left(-D_{VT1w} \frac{W_{eff} \cdot L_{eff}}{l_{n}} \right) \right) (V_{bi} - \Phi_s) $$

$(2.1.24b)$

When all of the above considerations for non-uniform doping, short and narrow channel effects on threshold voltage are considered, the final complete $V_{th}$ expression implemented in SPICE is as follows:
Non-Uniform Doping and Small Channel Effects on Threshold Voltage

\[ V_{th} = V_{th0ax} + K_{1ax} \cdot \sqrt{\Phi_s} - V_{aux} - K_{2ax} V_{aux} \]

\[ + K_{1ax} \left( \sqrt{1 + \frac{N_{Lx}}{L_{ox}}} - 1 \right) \left( \Phi_s + \left( K_1 + K_{Lx} V_{aux} \right) \frac{T_{ox}}{W_{ox} + W_0} \right) \Phi_s \]

\[ - D_{VT0x} \left( \exp \left( - D_{VT1x} \frac{W_{ox} L_{ox}}{2l_{ox}} \right) + 2 \exp \left( - \frac{W_{ox} L_{ox}}{l_{ox}} \right) \right) (V_{hs} - \Phi_s) \]

\[ - \exp \left( - D_{VT1x} \frac{W_{ox} L_{ox}}{2l_{ox}} \right) + 2 \exp \left( - \frac{W_{ox} L_{ox}}{l_{ox}} \right) (V_{hs} - \Phi_s) \]

\[ - \left( \exp \left( - D_{sub} \frac{L_{ox}}{2l_{ox}} \right) + 2 \exp \left( - \frac{L_{ox}}{l_{ox}} \right) \right) (E_{sub} + E_{sub} V_{aux} V_{ds}) \]

where \( T_{ox} \) dependence is introduced in the model parameters \( K_1 \) and \( K_2 \) to improve the scalability of \( V_{th} \) model with respect to \( T_{ox} \). \( V_{th0ax}, K_{1ax} \) and \( K_{2ax} \) are modeled as

\[ V_{th0ax} = V_{th0} - K_1 \cdot \sqrt{\Phi_s} \]

and

\[ K_{1ax} = K_1 \cdot \frac{T_{ox}}{T_{oxm}} \]

\[ K_{2ax} = K_2 \cdot \frac{T_{ox}}{T_{oxm}} \]

\( T_{oxm} \) is the gate oxide thickness at which parameters are extracted with a default value of \( T_{ox} \).

In Eq. (2.1.25), all \( V_{bs} \) terms have been substituted with a \( V_{bs eff} \) expression as shown in Eq. (2.1.26). This is done in order to set an upper bound for the body bias value during simulations since unreasonable values can occur if this expression is not introduced (see Section 3.8 for details).
Mobility Model

\[
V_{\text{heff}} = V_{hc} + 0.5[V_{bs} - V_{hc} - \delta_1 + \sqrt{(V_{bs} - V_{hc} - \delta_1)^2 - 4\delta_1 V_{hc}}]
\]

where \(\delta_1 = 0.001\). The parameter \(V_{hc}\) is the maximum allowable \(V_{bs}\) value and is calculated from the condition of \(dV_{th}/dV_{bs} = 0\) for the \(V_{th}\) expression of 2.1.4, 2.1.5, and 2.1.6, and is equal to:

\[
V_{hc} = 0.9 \left( \Phi - \frac{K_1^2}{4K_2^2} \right)
\]

2.2 Mobility Model

A good mobility model is critical to the accuracy of a MOSFET model. The scattering mechanisms responsible for surface mobility basically include phonons, coulombic scattering, and surface roughness [11, 12]. For good quality interfaces, phonon scattering is generally the dominant scattering mechanism at room temperature. In general, mobility depends on many process parameters and bias conditions. For example, mobility depends on the gate oxide thickness, substrate doping concentration, threshold voltage, gate and substrate voltages, etc. Sabnis and Clemens [13] proposed an empirical unified formulation based on the concept of an effective field \(E_{\text{eff}}\) which lumps many process parameters and bias conditions together. \(E_{\text{eff}}\) is defined by

\[
E_{\text{eff}} = \frac{Q_B + (Q_n/2)}{\varepsilon_{si}}
\]

The physical meaning of \(E_{\text{eff}}\) can be interpreted as the average electrical field experienced by the carriers in the inversion layer [14]. The unified formulation of mobility is then given by
Values for $\mu_0$, $E_0$, and $v$ were reported by Liang *et al.* [15] and Toh *et al.* [16] to be the following for electrons and holes:

<table>
<thead>
<tr>
<th>Parameter</th>
<th>Electron (surface)</th>
<th>Hole (surface)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\mu_0$ ($cm^2/V$)</td>
<td>670</td>
<td>160</td>
</tr>
<tr>
<td>$E_0$ (MV/cm)</td>
<td>0.67</td>
<td>0.7</td>
</tr>
<tr>
<td>$v$</td>
<td>1.6</td>
<td>1.0</td>
</tr>
</tbody>
</table>

**Table 2-1. Typical mobility values for electrons and holes.**

For an NMOS transistor with n-type poly-silicon gate, Eq. (2.2.1) can be rewritten in a more useful form that explicitly relates $E_{eff}$ to the device parameters [14]

$$E_{eff} = \frac{V_{gs} + V_{th}}{6T_{ox}}$$

Eq. (2.2.2) fits experimental data very well [15], but it involves a very time consuming power function in SPICE simulation. Taylor expansion Eq. (2.2.2) is used, and the coefficients are left to be determined by experimental data or to be obtained by fitting the unified formulation. Thus, we have
Carrier Drift Velocity

\[ \mu_{\text{eff}} = \frac{\mu_0}{1 + (U_a + U_c V_{\text{beff}}) \left( \frac{V_{\text{gstr}} + 2V_{\text{th}}}{T_{\text{ox}}} \right) + U_b \left( \frac{V_{\text{gstr}} + 2V_{\text{th}}}{T_{\text{ox}}} \right)^2} \]  \hspace{1cm} (2.2.4)

where \( V_{\text{gstr}} = V_{\text{gs}} - V_{\text{th}} \). To account for depletion mode devices, another mobility model option is given by the following

\[ \mu_{\text{eff}} = \frac{\mu_0}{1 + (U_a + U_c V_{\text{beff}}) \left( \frac{V_{\text{gstr}}}{T_{\text{ox}}} \right) + U_b \left( \frac{V_{\text{gstr}}}{T_{\text{ox}}} \right)^2} \]  \hspace{1cm} (2.2.5)

The unified mobility expressions in subthreshold and strong inversion regions will be discussed in Section 3.2.

To consider the body bias dependence of Eq. 2.2.4 further, we have introduced the following expression:

\[ \mu_{\text{eff}} = \frac{\mu_0}{1 + [U_a \left( \frac{V_{\text{gstr}} + 2V_{\text{th}}}{T_{\text{ox}}} \right) + U_b \left( \frac{V_{\text{gstr}} + 2V_{\text{th}}}{T_{\text{ox}}} \right)^2 ](1 + U_c V_{\text{beff}})} \]  \hspace{1cm} (2.2.6)

### 2.3 Carrier Drift Velocity

Carrier drift velocity is also one of the most important parameters. The following velocity saturation equation [17] is used in the model
The parameter $E_{sat}$ corresponds to the critical electrical field at which the carrier velocity becomes saturated. In order to have a continuous velocity model at $E = E_{sat}$, $E_{sat}$ must satisfy:

$$E_{sat} = \frac{2v_{sat}}{\mu_{eff}}$$

### 2.4 Bulk Charge Effect

When the drain voltage is large and/or when the channel length is long, the depletion "thickness" of the channel is non-uniform along the channel length. This will cause $V_{th}$ to vary along the channel. This effect is called bulk charge effect [14].

The parameter, $A_{bulk}$, is used to take into account the bulk charge effect. Several extracted parameters such as $A_0$, $B_0$, $B_1$ are introduced to account for the channel length and width dependences of the bulk charge effect. In addition, the parameter $K$ is introduced to model the change in bulk charge effect under high substrate bias conditions. It should be pointed out that narrow width effects have been considered in the formulation of Eq. (2.4.1). The $A_{bulk}$ expression is given by

$$A_{bulk} = \left(1 + \frac{K}{2\sqrt{\Phi_s - V_{surf}}} \left( \frac{A_0 L_{eff}}{L_{eff} + 2\sqrt{X_s X_{dip}}} \left(1 - A_{s} V_{surf} \left( \frac{L_{eff}}{L_{eff} + 2\sqrt{X_s X_{dip}}} \right)^2 \right) + \frac{B_0}{W_{eff} + B_1} \right) \right) \frac{1}{1 + K_{eta} V_{surf}}$$
2.5 Strong Inversion Drain Current (Linear Regime)

2.5.1 Intrinsic Case \((R_{ds}=0)\)

In the strong inversion region, the general current equation at any point \(y\) along the channel is given by

\[
I_{ds} = WC_{ox} (V_{gs} - A_{bulk} V(y)) v(y)
\]  

(2.5.1)

The parameter \(V_{gsl} = (V_{gs} - V_{th})\), \(W\) is the device channel width, \(C_{ox}\) is the gate capacitance per unit area, \(V(y)\) is the potential difference between minority-carrier quasi-Fermi potential and the equilibrium Fermi potential in the bulk at point \(y\), \(v(y)\) is the velocity of carriers at point \(y\).

With Eq. (2.3.1) (i.e. before carrier velocity saturates), the drain current can be expressed as

\[
I_{ds} = WC_{ox} (V_{gs} - V_{th} - A_{bulk} V(y)) \frac{\mu_{eff} E(y)}{1 + E_{sat} / E_{sat}}
\]  

(2.5.2)

Eq. (2.5.2) can be rewritten as follows:

\[
I_{ds} = WC_{ox} (V_{gs} - V_{th} - A_{bulk} V(y)) \frac{\mu_{eff} E(y)}{1 + E_{sat} / E_{sat}}
\]
Strong Inversion Drain Current (Linear Regime)

\[
E_{(y)} = \frac{I_{ds}}{\mu_{eff} WC_{ox} (V_{gs} - A_{bulk} V_{(y)}) - I_{ds}/E_{sat}} = \frac{dV(y)}{dy}
\]  

(2.5.3)

By integrating Eq. (2.5.2) from \( y = 0 \) to \( y = L \) and \( V(y) = 0 \) to \( V(y) = V_{ds} \), we arrive at the following

\[
I_{ds} = \mu_{eff} C_{ox} \frac{W}{L} \frac{1}{1 + V_{ds}/E_{sat} L} (V_{gs} - V_{th} - A_{bulk} V_{ds}/2)V_{ds}
\]

(2.5.4)

The drain current model in Eq. (2.5.4) is valid before velocity saturates.

For instances when the drain voltage is high (and thus the lateral electrical field is high at the drain side), the carrier velocity near the drain saturates. The channel region can now be divided into two portions: one adjacent to the source where the carrier velocity is field-dependent and the second where the velocity saturates. At the boundary between these two portions, the channel voltage is the saturation voltage \( V_{dsat} \) and the lateral electrical is equal to \( E_{sat} \). After the onset of saturation, we can substitute \( v = v_{sat} \) and \( V_{ds} = V_{dsat} \) into Eq. (2.5.1) to get the saturation current:

\[
I_{dr} = WC_{ox} (V_{gs} - A_{bulk} V_{sat}) v_{sat}
\]

(2.5.5)

By equating eqs. (2.5.4) and (2.5.5) at \( E = E_{sat} \) and \( V_{ds} = V_{dsat} \), we can solve for saturation voltage \( V_{dsat} \)

\[
V_{dsat} = \frac{E_{sat} L (V_{gs} - V_{th})}{A_{bulk} E_{sat} L + (V_{gs} - V_{th})}
\]

(2.5.6)
2.5.2 Extrinsic Case \((R_{ds} > 0)\)

Parasitic source/drain resistance is an important device parameter which can affect MOSFET performance significantly. As channel length scales down, the parasitic resistance will not be proportionally scaled. As a result, \(R_{ds}\) will have a more significant impact on device characteristics. Modeling of parasitic resistance in a direct method yields a complicated drain current expression. In order to make simulations more efficient, the parasitic resistances is modeled such that the resulting drain current equation in the linear region can be calculated [3] as

\[
I_{ds} = \frac{V_{ds}}{R_{tot}} = \frac{V_{ds}}{R_{ch} + R_{ds}} = \mu_{eff} C_{ox} \frac{W}{L} \frac{1}{1 + V_{ds}/(E_{sat} L)} \left( V_{gst} - A_{bulk} V_{ds}/2 \right) V_{ds} \left( V_{gst} - A_{bulk} V_{ds}/2 \right) \frac{W}{L} \frac{1 + V_{ds}/(E_{sat} L)}{1 + R_{ds} \mu_{eff} C_{ox} \frac{W}{L}}
\]

(2.5.9)

Due to the parasitic resistance, the saturation voltage \(V_{dsat}\) will be larger than that predicted by Eq. (2.5.6). Let Eq. (2.5.5) be equal to Eq. (2.5.9). \(V_{dsat}\) with parasitic resistance \(R_{ds}\) becomes

\[
V_{dsat} = \frac{-b - \sqrt{b^2 - 4ac}}{2a}
\]

(2.5.10)

The following are the expression for the variables \(a\), \(b\), and \(c\):
The last expression for $\lambda$ is introduced to account for non-saturation effect of the device. The parasitic resistance is modeled as:

$$R_{ds} = \frac{R_{ds} \left(1 + P_{rbg} V_{g_{eff}} + P_{rbh} \left(\sqrt{\Phi_{s}} - V_{b_{eff}} - \sqrt{\Phi_{s}}\right)\right)}{\left(10^6 W_{eff}\right)^{W_{r}}}$$

The variable $R_{ds}$ is the resistance per unit width, $W_{r}$ is a fitting parameter, $P_{rbh}$ and $P_{rbg}$ are the body bias and the gate bias coefficients, respectively.

### 2.6 Strong Inversion Current and Output Resistance (Saturation Regime)

A typical I-V curve and its output resistance are shown in Figure 2-6. Considering only the drain current, the I-V curve can be divided into two parts: the linear region in which the drain current increases quickly with the drain voltage and the saturation region in which the drain current has a very weak dependence on the drain voltage. The first order derivative reveals more detailed information about the physical mechanisms which are involved during device operation. The output resistance (which is the reciprocal of the first order derivative of the I-V curve)
Strong Inversion Current and Output Resistance (Saturation Regime)

curve can be clearly divided into four regions with distinct $R_{out}$ vs. $V_{ds}$ dependences.

The first region is the triode (or linear) region in which carrier velocity is not saturated. The output resistance is very small because the drain current has a strong dependence on the drain voltage. The other three regions belong to the saturation region. As will be discussed later, there are three physical mechanisms which affect the output resistance in the saturation region: channel length modulation (CLM) [4, 14], drain-induced barrier lowering (DIBL) [4, 6, 14], and the substrate current induced body effect (SCBE) [14, 18, 19]. All three mechanisms affect the output resistance in the saturation range, but each of them dominates in only a single region. It will be shown next that channel length modulation (CLM) dominates in the second region, DIBL in the third region, and SCBE in the fourth region.

![Graph showing the behavior of MOSFET output resistance with regions marked as Triode, CLM, DIBL, and SCBE.](image)

**Figure 2-6.** General behavior of MOSFET output resistance.
Generally, drain current is a function of the gate voltage and the drain voltage. But the drain current depends on the drain voltage very weakly in the saturation region. A Taylor series can be used to expand the drain current in the saturation region [3].

\[
I_{ds}(V_{gs}, V_{ds}) = I_{ds}(V_{gs}, V_{dsat}) + \frac{\partial I_{ds}(V_{gs}, V_{ds})}{\partial V_{ds}}(V_{ds} - V_{dsat})
\]

\[
\equiv I_{dsat}(1 + \frac{V_{ds} - V_{dsat}}{V_{A}})
\]

where

\[
I_{dsat} = I_{ds}(V_{gs}, V_{dsat}) = W_{v_{sat}} C_{ox}(V_{gst} - A_{bulk} V_{dsat})
\]

and

\[
V_{A} = I_{dsat} \left( \frac{\partial I_{ds}}{\partial V_{ds}} \right)^{-1}
\]

The parameter \( V_{A} \) is called the Early voltage and is introduced for the analysis of the output resistance in the saturation region. Only the first order term is kept in the Taylor series. We also assume that the contributions to the Early voltage from all three mechanisms are independent and can be calculated separately.
2.6.1 Channel Length Modulation (CLM)

If channel length modulation is the only physical mechanism to be taken into account, then according to Eq. (2.6.3), the Early voltage can be calculated by

\[
V_{ACL\text{M}} = I_{dsat} \left( \frac{\partial I_{ds}}{V_{ds}} \right)^{-1} = \frac{A_{bulk} E_{sat} L + V_{gst}}{A_{bulk} E_{sat}} \left( \frac{\partial \Delta L}{V_{ds}} \right)^{-1}
\]

where \( \Delta L \) is the length of the velocity saturation region; the effective channel length is \( L - \Delta L \). Based on the quasi-two dimensional approximation, \( V_{ACL\text{M}} \) can be derived as the following

\[
V_{ACL\text{M}} = \frac{A_{bulk} E_{sat} L + V_{gst}}{A_{bulk} E_{sat}} (V_{ds} - V_{dsat})
\]

where \( V_{ACL\text{M}} \) is the Early Voltage due to channel length modulation alone.

The parameter \( P_{clm} \) is introduced into the \( V_{ACL\text{M}} \) expression not only to compensate for the error caused by the Taylor expansion in the Early voltage model, but also to compensate for the error in \( X_j \) since \( l = \sqrt{X_j} \) and the junction depth \( X_j \) can not generally be determined very accurately. Thus, the \( V_{ACL\text{M}} \) became

\[
V_{ACL\text{M}} = \frac{1}{P_{clm}} \frac{A_{bulk} E_{sat} L + V_{gst}}{A_{bulk} E_{sat}} (V_{ds} - V_{dsat})
\]
2.6.2 Drain-Induced Barrier Lowering (DIBL)

As discussed above, threshold voltage can be approximated as a linear function of the drain voltage. According to Eq. (2.6.3), the Early voltage due to the DIBL effect can be calculated as:

\[ V_{\text{ADIBLC}} = I_{\text{dsat}} \frac{\partial I_{\text{ds}}}{\partial V_{\text{th}}} \frac{\partial V_{\text{th}}}{\partial V_{\text{ds}}}^{-1} \]

\[ V_{\text{ADIBLC}} = \frac{(V_{\text{geo}} + 2\eta)}{\theta_{\text{sat}}(1 + P_{\text{DIBLCV}}V_{\text{geo}})} \left( \frac{A_{\text{sat}}V_{\text{dsat}}}{A_{\text{sat}}V_{\text{dsat}} + V_{\text{geo}} + 2\eta} \right) \]

During the derivation of Eq. (2.6.7), the parasitic resistance is assumed to be equal to 0. As expected, \( V_{\text{ADIBLC}} \) is a strong function of \( L \) as shown in Eq. (2.6.7). As channel length decreases, \( V_{\text{ADIBLC}} \) decreases very quickly. The combination of the CLM and DIBL effects determines the output resistance in the third region, as was shown in Figure 2-6.

Despite the formulation of these two effects, accurate modeling of the output resistance in the saturation region requires that the coefficient \( \theta_{\text{th}}(L) \) be replaced by \( \theta_{\text{rout}}(L) \). Both \( \theta_{\text{th}}(L) \) and \( \theta_{\text{rout}}(L) \) have the same channel length dependencies but different coefficients. The expression for \( \theta_{\text{rout}}(L) \) is

\[ \theta_{\text{rout}}(L) = P_{\text{diblc1}} \{ \exp(-D_{\text{rout}} L / 2l_s) + 2 \exp(-D_{\text{rout}} L / l_s) \} + P_{\text{diblc2}} \]

Parameters \( P_{\text{diblc1}}, P_{\text{diblc2}}, P_{\text{diblcV}} \) and \( D_{\text{rout}} \) are introduced to correct for DIBL effect in the strong inversion region. The reason why \( D_{\text{vto}} \) is not
equal to $P_{diblc1}$ and $D_{vth}$ is not equal to $D_{rout}$ is because the gate voltage modulates the DIBL effect. When the threshold voltage is determined, the gate voltage is equal to the threshold voltage. But in the saturation region where the output resistance is modeled, the gate voltage is much larger than the threshold voltage. Drain induced barrier lowering may not be the same at different gate bias. $P_{diblc2}$ is usually very small (may be as small as 8.0E-3). If $P_{diblc2}$ is placed into the threshold voltage model, it will not cause any significant change. However it is an important parameter in $V_{ADIBL}$ for long channel devices, because $P_{diblc2}$ will be dominant in Eq. (2.6.8) if the channel is long.

### 2.6.3 Current Expression without Substrate Current Induced Body Effect

In order to have a continuous drain current and output resistance expression at the transition point between linear and saturation region, the $V_{Asat}$ parameter is introduced into the Early voltage expression. $V_{Asat}$ is the Early Voltage at $V_{ds} = V_{dsat}$ and is as follows:

$$V_{Asat} = \frac{E_{sat} L + V_{dsat} + 2 R_{ds} V_{sat} C_{ox} W (V_{gst} - A_{bulk} V_{ds} / 2)}{1 + A_{bulk} R_{ds} v_{sat} C_{ox} W}$$  \hspace{1cm} (2.6.9)

Total Early voltage, $V_A$, can be written as

$$V_A = V_{Asat} + \left( \frac{1}{V_{ACLM}} + \frac{1}{V_{ADIBL}} \right)^{-1}$$  \hspace{1cm} (2.6.10)
The complete (with no impact ionization at high drain voltages) current expression in the saturation region is given by

\[ I_{ds} = W_{sat} C_{ox} (V_{gs} - A_{bulk} V_{dsat})(1 + \frac{V_{ds} - V_{dsat}}{V_A}) \]  

(2.6.11)

Furthermore, another parameter, \( P_{VAG} \), is introduced in \( V_A \) to account for the gate bias dependence of \( V_A \) more accurately. The final expression for Early voltage becomes

\[ V_A = V_{Asat} + (1 + \frac{P_{VAG} V_{gs}}{E_{sat} L_{eff}})(\frac{1}{V_{ACL}} + \frac{1}{V_{ADIBL}})^{-1} \]  

(2.6.12)

2.6.4 Current Expression with Substrate Current Induced Body Effect

When the electrical field near the drain is very large (> 0.1MV/cm), some electrons coming from the source will be energetic (hot) enough to cause impact ionization. This creates electron-hole pairs when they collide with silicon atoms. The substrate current \( I_{sub} \) thus created during impact ionization will increase exponentially with the drain voltage. A well known \( I_{sub} \) model [20] is given as:

\[ I_{sub} = \frac{A_i}{B_i} I_{dsat} (V_{ds} - V_{dsat}) \exp(-\frac{B_i l}{V_{ds} - V_{dsat}}) \]  

(2.6.13)
The parameters $A_i$ and $B_i$ are determined from extraction. $I_{sub}$ will affect the drain current in two ways. The total drain current will change because it is the sum of the channel current from the source as well as the substrate current. The total drain current can now be expressed [21] as follows

$$I_{ds} = I_{ds0} + I_{sub}$$

$$= I_{ds0} \left[ 1 + \frac{(V_{ds} - V_{dsat})}{B_i \exp\left(\frac{B_i l}{A_i (V_{ds} - V_{dsat})}\right)} \right]$$

The total drain current, including $CLM$, $DIBL$ and $SCBE$, can be written as

$$I_{ds} = W_{V_{sat}} C_{ox} (V_{gs} - A_p V_{dsat}) \left(1 + \frac{V_{ds} - V_{dsat}}{V_{A}}\right) \left(1 + \frac{V_{ds} - V_{dsat}}{V_{ASCBE}}\right)$$

where $V_{ASCBE}$ can also be called as the Early voltage due to the substrate current induced body effect. Its expression is the following

$$V_{ASCBE} = \frac{B_i}{A_i} \exp\left(\frac{B_i l}{V_{ds} - V_{dsat}}\right)$$

From Eq. (2.6.16), we can see that $V_{ASCBE}$ is a strong function of $V_{ds}$. In addition, we also observe that $V_{ASCBE}$ is small only when $V_{ds}$ is large. This is why $SCBE$ is important for devices with high drain voltage bias. The channel length and gate oxide dependence of $V_{ASCBE}$ comes from $V_{dsat}$ and $l$. We replace $B_i$ with $PSCBE2$ and $A_i/B_i$ with $PSCBE1/L$ to yield the following expression for $V_{ASCBE}$
Subthreshold Drain Current

\[
\frac{1}{V_{ASCBE}} = \frac{P_{SCBE2}}{L} \exp\left(\frac{-P_{SCBE1}}{V_{ds} - V_{dsat}}\right)
\]

(2.6.17)

The variables \( P_{scbe1} \) and \( P_{scbe2} \) are determined experimentally.

### 2.7 Subthreshold Drain Current

The drain current equation in the subthreshold region can be expressed as [2, 3]

\[
I_{ds} = I_{so}(1 - \exp\left(-\frac{V_{ds}}{V_t}\right))\exp\left(\frac{\frac{V_{gs} - V_{th} - V_{off}}{nv_f}}{n}\right)
\]

(2.7.1)

\[
I_{so} = \frac{W}{L} \sqrt{\frac{q\varepsilon_s N_{ch}}{2\phi_s}} V_t^2
\]

(2.7.2)

Here the parameter \( V_t \) is the thermal voltage and is given by \( k_B T/q \). \( V_{off} \) is the offset voltage, as discussed in Jeng's dissertation [18]. \( V_{off} \) is an important parameter which determines the drain current at \( V_{gs} = 0 \). In Eq. (2.7.1), the parameter \( n \) is the subthreshold swing parameter. Experimental data shows that the subthreshold swing is a function of channel length and the interface state density. These two mechanisms are modeled by the following

\[
n = 1 + N_{interface} \left( C_d + C_{dcs} V_{ds} + C_{dch} V_{b} \right) \left( \exp\left(-D_{VT1} \frac{L_d}{2l_D} \right) + 2 \exp\left(-D_{VT1} \frac{L_d}{l_D} \right) \right) + \frac{C_h}{C_{ox}}
\]

(2.7.3)

where the term...
Effective Channel Length and Width

\[ (C_{dsd} + C_{dsc}V_{ds} + C_{dscb}V_{heff}) \left( \exp\left(-D_{vth} \frac{L_{eff}}{2L_i}\right) + 2\exp\left(-D_{vth} \frac{L_{eff}}{L_i}\right) \right) \]

represents the coupling capacitance between the drain or source to the channel. The parameters \( C_{dsd}, C_{dsc} \) and \( C_{dscb} \) are extracted. The parameter \( C_{li} \) in Eq. (2.7.3) is the capacitance due to interface states. From Eq. (2.7.3), it can be seen that subthreshold swing shares the same exponential dependence on channel length as the DIBL effect. The parameter \( N_{factor} \) is introduced to compensate for errors in the depletion width capacitance calculation. \( N_{factor} \) is determined experimentally and is usually very close to 1.

### 2.8 Effective Channel Length and Width

The effective channel length and width used in all model expressions is given below

\[ L_{eff} = L_{drawn} - 2dL \]  \hspace{1cm} (2.8.1)

\[ W_{eff} = W_{drawn} - 2dW \]  \hspace{1cm} (2.8.2a)

\[ W_{eff}' = W_{drawn} - 2dW' \]  \hspace{1cm} (2.8.2b)

The only difference between Eq. (2.8.1a) and (2.8.1b) is that the former includes bias dependencies. The parameters \( dW \) and \( dL \) are modeled by the following
These complicated formulations require some explanation. From Eq. (2.8.3), the variable \( W_{\text{int}} \) models represents the tradition manner from which "delta W" is extracted (from the intercepts of straights lines on a \( 1/R_{ds} \) vs. \( W_{\text{drawn}} \) plot). The parameters \( dW_g \) and \( dW_b \) have been added to account for the contribution of both front gate and back side (substrate) biasing effects. For \( dL \), the parameter \( L_{\text{int}} \) represents the traditional manner from which "delta L" is extracted (mainly from the intercepts of lines from a \( R_{ds} \) vs. \( L_{\text{drawn}} \) plot).

The remaining terms in both \( dW \) and \( dL \) are included for the convenience of the user. They are meant to allow the user to model each parameter as a function of \( W_{\text{drawn}} \), \( L_{\text{drawn}} \) and their associated product terms. In addition, the freedom to model these dependencies as other than just simple inverse functions of \( W \) and \( L \) is also provided for the user. For \( dW \), they are \( W_{\text{int}} \) and \( W_{\text{wn}} \). For \( dL \) they are \( L_{\text{int}} \) and \( L_{\text{wn}} \).

By default all of the above geometrical dependencies for both \( dW \) and \( dL \) are turned off. Again, these equations are provided for the convenience of the user. As such, it is up to the user to adopt the correct extraction strategy to ensure proper use.
2.9 Poly Gate Depletion Effect

When a gate voltage is applied to a heavily doped poly-silicon gate, e.g. NMOS with n⁺ poly-silicon gate, a thin depletion layer will be formed at the interface between the poly-silicon and gate oxide. Although this depletion layer is very thin due to the high doping concentration of the poly-Si gate, its effect cannot be ignored in the 0.1μm regime since the gate oxide thickness will also be very small, possibly 50Å or thinner.

Figure 2-7 shows an NMOSFET with a depletion region in the n⁺ poly-silicon gate. The doping concentration in the n⁺ poly-silicon gate is \( N_{gate} \) and the doping concentration in the substrate is \( N_{sub} \). The gate oxide thickness is \( T_{ox} \). The depletion width in the poly gate is \( X_p \). The depletion width in the substrate is \( X_d \). If we assume the doping concentration in the gate is infinite, then no depletion region will exist in the gate, and there would be one sheet of positive charge whose thickness is zero at the interface between the poly-silicon gate and gate oxide.

In reality, the doping concentration is, of course, finite. The positive charge near the interface of the poly-silicon gate and the gate oxide is distributed over a finite depletion region with thickness \( X_p \). In the presence of the depletion region, the voltage drop across the gate oxide and the substrate will be reduced, because part of the gate voltage will be dropped across the depletion region in the gate. That means the effective gate voltage will be reduced.
The effective gate voltage can be calculated in the following manner. Assume the doping concentration in the poly gate is uniform. The voltage drop in the poly gate ($V_{\text{poly}}$) can be calculated as

$$V_{\text{poly}} = \frac{1}{2} X_{\text{poly}} E_{\text{poly}} = \frac{q N_{\text{gate}} X_{\text{poly}}^2}{2 \varepsilon_{\text{si}}} \quad (2.9.1)$$

where $E_{\text{poly}}$ is the maximum electrical field in the poly gate. The boundary condition at the interface of poly gate and the gate oxide is

$$\varepsilon_{\text{ox}} E_{\text{ox}} = \varepsilon_{\text{si}} E_{\text{poly}} = \sqrt{2q \varepsilon_{\text{si}} N_{\text{gate}}} V_{\text{poly}} \quad (2.9.2)$$

Figure 2-7. Charge distribution in a MOSFET with the poly gate depletion effect. The device is in the strong inversion region.
where $E_{ox}$ is the electrical field in the gate oxide. The gate voltage satisfies

$$V_{gs} - V_{FB} - \Phi_s = V_{poly} + V_{ox} \tag{2.9.3}$$

where $V_{ox}$ is the voltage drop across the gate oxide and satisfies $V_{ox} = E_{ox} T_{ox}$. According to the equations (2.9.1) to (2.9.3), we obtain the following

$$a(V_{gs} - V_{FB} - \Phi_s - V_{poly})^2 - V_{poly} = 0 \tag{2.9.4}$$

where

$$a = \frac{\varepsilon_{ox}^2}{2q\varepsilon_{si} N_{gate} T_{ox}^2} \tag{2.9.5}$$

By solving the equation (2.9.4), we get the effective gate voltage ($V_{g_{eff}}$) which is equal to:

$$V_{g_{eff}} = V_{FB} + \Phi_s + \frac{q\varepsilon_{si} N_{gate} T_{ox}^2}{\varepsilon_{ox}^2} \left( \frac{2E_{ox}^2 (V_{gs} - V_{FB} - \Phi_s)}{q\varepsilon_{si} N_{gate} T_{ox}^2} - 1 \right) \tag{2.9.6}$$

Figure 2-8 shows $V_{g_{eff}} / V_{gs}$ versus the gate voltage. The threshold voltage is assumed to be 0.4V. If $T_{ox} = 40$ Å, the effective gate voltage can be reduced by 6% due to the poly gate depletion effect as the applied gate voltage is equal to 3.5V.
Figure 2-8. The effective gate voltage versus applied gate voltage at different gate oxide thickness.

The drain current reduction in the linear region as a function of the gate voltage can now be determined. Assume the drain voltage is very small, e.g. 50mV. Then the linear drain current is proportional to $C_{ox}(V_{gs} - V_{th})$. The ratio of the linear drain current with and without poly gate depletion is equal to:

$$\frac{I_{ds}(V_{gs\_eff})}{I_{ds}(V_{gs})} = \frac{(V_{gs\_eff} - V_{th})}{(V_{gs} - V_{th})} \quad (2.9.7)$$

Figure 2-9 shows $I_{ds}(V_{gs\_eff}) / I_{ds}(V_{gs})$ versus the gate voltage using Eq. (2.9.7). The drain current can be reduced by several percent due to gate depletion.
Poly Gate Depletion Effect

Figure 2-9. Ratio of linear region current with poly gate depletion effect and that without.
Poly Gate Depletion Effect
The development of separate model expressions for such device operation regimes as subthreshold and strong inversion were discussed in Chapter 2. Although these expressions can accurately describe device behavior within their own respective region of operation, problems are likely to occur between two well-described regions or within transition regions. In order to circumvent this issue, a unified model should be synthesized to not only preserve region-specific expressions but also to ensure the continuities of current and conductance and their derivatives in all transition regions as well. Such high standards are kept in BSIM3v3.2. As a result, convergence and simulation efficiency are much improved.

This chapter will describe the unified I-V model equations. While most of the parameter symbols in this chapter are explained in the following text, a complete description of all I-V model parameters can be found in Appendix A.

3.1 Unified Channel Charge Density Expression

Separate expressions for channel charge density are shown below for subthreshold (Eq. (3.1.1a) and (3.1.1b)) and strong inversion (Eq. (3.1.2)). Both expressions are valid for small $V_{ds}$. 
Unified Channel Charge Density Expression

\[ Q_{ch,0} = Q_0 \exp\left(\frac{V_{gs} - V_{th}}{n \nu_t}\right) \]  
(3.1.1a)

where \( Q_0 \) is

\[ Q_0 = \sqrt{\frac{q \varepsilon s N_{ch}}{2 \phi_s \nu_t}} \exp\left(-\frac{V_{off}}{n \nu_t}\right) \]  
(3.1.1b)

\[ Q_{ch,0} = C_{ox}(V_{gs} - V_{th}) \]  
(3.1.2)

In both Eqs. (3.1.1a) and (3.1.2), the parameters \( Q_{ch,sub,0} \) and \( Q_{ch,0} \) are the channel charge densities at the source for very small Vds. To form a unified expression, an effective \( (V_{gs} - V_{th}) \) function named \( V_{gsteff} \) is introduced to describe the channel charge characteristics from subthreshold to strong inversion

\[ V_{gsteff} = \frac{2 n \nu_t \ln\left(1 + \exp\left(\frac{V_{gs} - V_{th}}{2 n \nu_t}\right)\right)}{1 + 2 n C_{ox} \sqrt{\frac{2 \Phi_s}{q \varepsilon s N_{ch}}} \exp\left(-\frac{V_{gs} - V_{th} - 2V_{off}}{2 n \nu_t}\right)} \]  
(3.1.3)

The unified channel charge density at the source end for both subthreshold and inversion region can therefore be written as

\[ Q_{ch,0} = C_{ox} V_{gsteff} \]  
(3.1.4)

Figures 3-1 and 3-2 show the "smoothness" of Eq. (3.1.4) from subthreshold to strong inversion regions. The \( V_{gsteff} \) expression will be used again in subsequent sections of this chapter to model the drain current.
Unified Channel Charge Density Expression

Figure 3-1. The $V_{\text{gsteff}}$ function vs. $(V_{gs}-V_{th})$ in linear scale.

Figure 3-2. $V_{\text{gsteff}}$ function vs. $(V_{gs}-V_{th})$ in log scale.
Eq. (3.1.4) serves as the cornerstone of the unified channel charge expression at the source for small $V_{ds}$. To account for the influence of $V_{ds}$, the $V_{gs_{eff}}$ function must keep track of the change in channel potential from the source to the drain. In other words, Eq. (3.1.4) will have to include a $y$ dependence. To initiate this formulation, consider first the re-formulation of channel charge density for the case of strong inversion

$$Q_{ch(y)} = C_{ox}(V_{gs} - V_{th} - A_{bulk}V_{F(y)})$$  \hspace{1cm} (3.1.5)

The parameter $V_{F(y)}$ stands for the quasi-Fermi potential at any given point $y$, along the channel with respect to the source. This equation can also be written as

$$Q_{ch(y)} = Q_{ch0} + \Delta Q_{ch(y)}$$ \hspace{1cm} (3.1.6)

The term $\Delta Q_{ch(y)}$ is the incremental channel charge density induced by the drain voltage at point $y$. It can be expressed as

$$\Delta Q_{ch(y)} = -C_{ox}A_{bulk}V_{F(y)}$$ \hspace{1cm} (3.1.7)

For the subthreshold region ($V_{gs} < V_{th}$), the channel charge density along the channel from source to drain can be written as

$$Q_{ch_{sub}(y)} = Q_{0}\exp\left(\frac{V_{gs} - V_{th} - A_{bulk}V_{F(y)}}{n\nu_{t}}\right)$$ \hspace{1cm} (3.1.8)

$$= Q_{ch_{sub0}}\exp\left(-\frac{A_{bulk}V_{F(y)}}{n\nu_{t}}\right)$$
A Taylor series expansion of the right-hand side of Eq. (3.1.8) yields the following (keeping only the first two terms)

$$Q_{chubs(y)} = Q_{chubs0}(1 - \frac{A_{bulk}V_F(y)}{n\nu})$$

(3.1.9)

Analogous to Eq. (3.1.6), Eq. (3.1.9) can also be written as

$$Q_{chubs(y)} = Q_{chubs0} + \Delta Q_{chubs(y)}$$

(3.1.10)

The parameter $\Delta Q_{chubs(y)}$ is the incremental channel charge density induced by the drain voltage in the subthreshold region. It can be written as

$$\Delta Q_{chubs(y)} = - \frac{A_{bulk}V_F(y)}{n\nu} Q_{chubs0}$$

(3.1.11)

Note that Eq. (3.1.9) is valid only when $V_F(y)$ is very small, which is maintained fortunately, due to the fact that Eq. (3.1.9) is only used in the linear regime (i.e. $V_{ds} \leq 2\nu$).

Eqs. (3.1.6) and (3.1.10) both have drain voltage dependencies. However, they are decoupled and a unified expression for $Q_{ch(y)}$ is needed. To obtain a unified expression along the channel, we first assume

$$\Delta Q_{ch(y)} = \frac{\Delta Q_{ch(y)}\Delta Q_{chubs(y)}}{\Delta Q_{ch(y)} + \Delta Q_{chubs(y)}}$$

(3.1.12)

Here, $\Delta Q_{ch(y)}$ is the incremental channel charge density induced by the drain voltage. Substituting Eq. (3.1.7) and (3.1.11) into Eq. (3.1.12), we obtain
where \( V_b = (V_{g_{steff}} + n_v I) / A_{bulk} \). In order to remove any association between the variable \( n \) and bias dependencies \( V_{g_{steff}} \) as well as to ensure more precise modeling of Eq. (3.1.8) for linear regimes (under subthreshold conditions), \( n \) is replaced by 2. The expression for \( V_b \) now becomes

\[
V_b = \frac{V_{g_{steff}} + 2v_I}{A_{bulk}}
\]

A unified expression for \( Q_{ch(y)} \) from subthreshold to strong inversion regimes is now at hand

\[
Q_{ch(y)} = Q_{ch0}(1 - \frac{V_{F(y)}}{V_b})
\]

The variable \( Q_{ch0} \) is given by Eq. (3.1.4).

### 3.2 Unified Mobility Expression

Unified mobility model based on the \( V_{g_{steff}} \) expression of Eq. 3.1.3 is described in the following.
Unified Linear Current Expression

(mobMod = 1) \[ \mu_{\text{eff}} = \frac{\mu_0}{1 + (U_u + U_c V_{b\text{eff}})(\frac{V_{g\text{eff}} + 2V_{th}}{T_{ox}}) + U_b\left(\frac{V_{g\text{eff}} + 2V_{th}}{T_{ox}}\right)^2} \] (3.2.1)

To account for depletion mode devices, another mobility model option is given by the following

(mobMod = 2) \[ \mu_{\text{eff}} = \frac{\mu_0}{1 + (U_u + U_c V_{b\text{eff}})(\frac{V_{g\text{eff}}}{T_{ox}}) + U_b\left(\frac{V_{g\text{eff}}}{T_{ox}}\right)^2} \] (3.2.2)

To consider the body bias dependence of Eq. 3.2.1 further, we have introduced the following expression

(For mobMod = 3) \[ \mu_{\text{eff}} = \frac{\mu_0}{1 + [U_u\left(\frac{V_{g\text{eff}} + 2V_{th}}{T_{ox}}\right) + U_b\left(\frac{V_{g\text{eff}} + 2V_{th}}{T_{ox}}\right)^2](1 + U_c V_{b\text{eff}})} \] (3.2.3)

3.3 Unified Linear Current Expression

3.3.1 Intrinsic case \( (R_{\Delta}=0) \)

Generally, the following expression \[2\] is used to account for both drift and diffusion current
Unified Linear Current Expression

\[ I_{d(y)} = W Q_{ch(y)} \mu_{he(y)} \frac{dV_{F(y)}}{dy} \]  

(3.3.1)

where the parameter \( \mu_{he(y)} \) can be written as

\[ \mu_{he(y)} = \frac{\mu_{eff}}{1 + \frac{E_x}{E_{sat}}} \]  

(3.3.2)

Substituting Eq. (3.3.2) in Eq. (3.3.1) we get

\[ I_{d(y)} = W Q_{ch(y)} (1 - \frac{V_{F(y)}}{V_t}) \frac{\mu_{eff}}{1 + \frac{E_x}{E_{sat}}} \frac{dV_{F(y)}}{dy} \]  

(3.3.3)

Eq. (3.3.3) resembles the equation used to model drain current in the strong inversion regime. However, it can now be used to describe the current characteristics in the subthreshold regime when \( V_{ds} \) is very small (\( V_{ds} \ll V_t \)). Eq. (3.3.3) can now be integrated from the source to drain to get the expression for linear drain current in the channel. This expression is valid from the subthreshold regime to the strong inversion regime

\[ I_{d0} = \frac{W \mu_{eff} Q_{ch0} V_{ds} (1 - \frac{V_{ds}}{2V_t})}{L(1 + \frac{V_{ds}}{V_t})} \]  

(3.3.4)
3.3.2 Extrinsic Case ($R_{ds} > 0$)

The current expression when $R_{ds} > 0$ can be obtained based on Eq. (2.5.9) and Eq. (3.3.4). The expression for linear drain current from subthreshold to strong inversion is:

$$I_{ds} = \frac{I_{dso}}{1 + \frac{R_{ds}I_{dso}}{V_{ds}}}$$

(3.3.5)

3.4 Unified $V_{dsat}$ Expression

3.4.1 Intrinsic case ($R_{ds}=0$)

To get an expression for the electric field as a function of $y$ along the channel, we integrate Eq. (3.3.1) from 0 to an arbitrary point $y$. The result is as follows

$$E_y = \frac{I_{dso}}{\sqrt{(WQ_{ch0}\mu_{eff} - \frac{I_{dso}}{E_{sat}})^2 - \frac{2I_{dso}WQ_{ch0}\mu_{eff}y}{V_b}}}$$

(3.4.1)

If we assume that drift velocity saturates when $E_y=E_{sat}$, we get the following expression for $I_{dsat}$

$$I_{dsat} = \frac{W\mu_{eff}\mu_{eff}Q_{ch0}E_{sat}L V_b}{2L(E_{sat}L + V_b)}$$

(3.4.2)
Unified Vdsat Expression

Let $V_{dS} = V_{dsat}$ in Eq. (3.3.4) and set this equal to Eq. (3.4.2), we get the following expression for $V_{dsat}$

$$V_{dsat} = \frac{E_{satL}(V_{gsleff} + 2V_t)}{A_{bulk}E_{satL} + V_{kneff} + 2V_t}$$  \hspace{1cm} (3.4.3)

**3.4.2 Extrinsic Case ($R_{dS} > 0$)**

The $V_{dsat}$ expression for the extrinsic case is formulated from Eq. (3.4.3) and Eq. (2.5.10) to be the following

$$V_{dsat} = \frac{-b - \sqrt{b^2 - 4ac}}{2a}$$  \hspace{1cm} (3.4.4a)

where

$$a = A_{bulk}^2 W_{eff} V_{sat} C_{ox} R_{DS} + (\frac{1}{\lambda} - 1) A_{bulk}$$ \hspace{1cm} (3.4.4b)

$$b = -\left( (V_{kneff} + 2V_t)(\frac{2}{\lambda} - 1) + A_{bulk}E_{satL} + 3A_{bulk}(V_{kneff} + 2V_t)W_{eff} V_{sat} C_{ox} R_{DS} \right)$$ \hspace{1cm} (3.4.4c)

$$c = (V_{kneff} + 2V_t) E_{satL} + 2(V_{kneff} + 2V_t)^2 W_{eff} V_{sat} C_{ox} R_{DS}$$ \hspace{1cm} (3.4.4d)

$$\lambda = A_1 V_{kneff} + A_2$$ \hspace{1cm} (3.4.4e)
The parameter $\lambda$ is introduced to account for non-saturation effects. Parameters $A_1$ and $A_2$ can be extracted.

### 3.5 Unified Saturation Current Expression

A unified expression for the saturation current from the subthreshold to the strong inversion regime can be formulated by introducing the $V_{gsteff}$ function into Eq. (2.6.15). The resulting equations are the following

$$I_{ds} = \frac{I_{ds0}(V_{dsat})}{1 + \frac{R_d I_{ds0}(V_{dsat})}{V_{dsat}}} \left(1 + \frac{V_{ds} - V_{dsat}}{V_A}\right) \left(1 + \frac{V_{ds} - V_{dsat}}{V_{ASCBE}}\right)$$

where

$$V_A = V_{Asat} + (1 + \frac{P_{vag}V_{gsteff}}{E_{sat}L_{eff}})(\frac{1}{V_{ACLm}} + \frac{1}{V_{ADIBLC}})^{-1}$$

$$E_{sat}L_{eff} + V_{dsat} + 2R_{DS}V_{sat}C_{ox}W_{eff}V_{gsteff}[1 - \frac{A_{bulk}V_{dsat}}{2(V_{gsteff} + 2V_t)}]$$

$$V_{Asat} = \frac{2 / \lambda - 1 + R_{DS}V_{sat}C_{ox}W_{eff}A_{bulk}}{2 / \lambda - 1 + R_{DS}V_{sat}C_{ox}W_{eff}A_{bulk}}$$

$$V_{ACLm} = \frac{A_{bulk}E_{sat}L_{eff} + V_{gsteff}}{P_{CLM}A_{bulk}E_{sat}l_{itl}} (V_{ds} - V_{dsat})$$
3.6 Single Current Expression for All Operating Regimes of Vgs and Vds

The $V_{gs}^{eff}$ function introduced in Chapter 2 gave a unified expression for the linear drain current from subthreshold to strong inversion as well as for the saturation drain current from subthreshold to strong inversion, separately. In order to link the continuous linear current with that of the continuous saturation current, a smooth function for $V_{ds}$ is introduced. In the past, several smoothing functions have been proposed for MOSFET modeling [22-24]. The smoothing function used in BSIM3 is similar to that proposed in [24]. The final current equation for both linear and saturation current now becomes

\[
I_d = \frac{I_{ds}(V_{ds}^{eff})}{1 + \frac{R_{ds} I_{ds}(V_{ds}^{eff})}{V_{ds}^{eff}}} \left(1 + \frac{V_{ds} - V_{ds}^{eff}}{V_A} \right) \left(1 + \frac{V_{ds} - V_{ds}^{eff}}{V_{ASCBE}} \right)
\]

Most of the previous equations which contain $V_{ds}$ and $V_{ds}^{eff}$ dependencies are now substituted with the $V_{ds}^{eff}$ function. For example, Eq. (3.5.4) now becomes
Single Current Expression for All Operating Regimes of $V_{gs}$ and $V_{ds}$

\[ V_{ACLM} = \frac{A_{bulk}E_{sat}L_{eff} + V_{gseff}}{P_{CLM}A_{bulk}E_{sat} \text{litr}} (V_{ds} - V_{dseff}) \]

Similarly, Eq. (3.5.7) now becomes

\[ \frac{1}{V_{ASCBE}} = \frac{P_{sce2}}{L_{eff}} \exp \left( \frac{-P_{sce1} \text{litr}}{V_{ds} - V_{dseff}} \right) \]

The $V_{dseff}$ expression is written as

\[ V_{dseff} = V_{dsat} - \frac{1}{2} \left( V_{dsat} - V_{ds} - \delta + \sqrt{(V_{dsat} - V_{ds} - \delta)^2 + 4\delta V_{dsat}} \right) \]

The expression for $V_{dsat}$ is that given under Section 3.4. The parameter $\delta$ is an extracted constant. The dependence of $V_{dseff}$ on $V_{ds}$ is given in Figure 3-3. The $V_{dseff}$ function follows $V_{ds}$ in the linear region and tends to $V_{dsat}$ in the saturation region. Figure 3-4 shows the effect of $\delta$ on the transition region between linear and saturation regimes.
Single Current Expression for All Operating Regimes of Vgs and Vds

Figure 3-3. $V_{dseff}$ vs. $V_{ds}$ for $\delta=0.01$ and different $V_{gs}$.

Figure 3-4. $V_{dseff}$ vs. $V_{ds}$ for $V_{gs}=3V$ and different $\delta$ values.
3.7 Substrate Current

The substrate current in BSIM3v3.2 is modeled by

\[
I_{\text{sub}} = \frac{\alpha_0 + \alpha_1 \cdot L_{\text{eff}}}{L_{\text{eff}}} (V_{d\text{s}} - V_{d\text{eff}}) \exp \left( - \frac{\beta_0}{V_{d\text{s}} - V_{d\text{eff}}} \right) \frac{I_{d0}}{1 + \frac{V_{d\text{s}} - V_{d\text{eff}}}{V_A}}
\]

(3.7.1)

where parameters \( \alpha_0 \) and \( \beta_0 \) are impact ionization coefficients; parameter \( \alpha_1 \) improves the \( I_{\text{Sub}} \) scalability.

3.8 A Note on \( V_{BS} \)

All \( V_{BS} \) terms have been substituted with a \( V_{BS\text{eff}} \) expression as shown in Eq. (3.8.1). This is done in order to set an upper bound for the body bias value during simulations. Unreasonable values can occur if this expression is not introduced.

\[
V_{BS\text{eff}} = V_{B} + 0.5 \left( V_{B} - V_{K} - \delta_1 + \sqrt{(V_{B} - V_{K} - \delta_1)^2 - 4\delta_1 V_{K}} \right)
\]

(3.8.1)

where \( \delta_1 = 0.001 \).

Parameter \( V_{BC} \) is the maximum allowable \( V_{BS} \) value and is obtained based on the condition of \( dV_{Th}/dV_{BS} = 0 \) for the \( V_{Th} \) expression of 2.1.4.
A Note on Vbs
Accurate modeling of MOSFET capacitance plays equally important role as that of the DC model. This chapter describes the methodology and device physics considered in both intrinsic and extrinsic capacitance modeling in BSIM3v3.2. Detailed model equations are given in Appendix B. One of the important features of BSIM3v3.2 is introduction of a new intrinsic capacitance model (capMod=3 as the default model), considering the finite charge thickness determined by quantum effect, which becomes more important for thinner $T_{ox}$ CMOS technologies. This model is smooth, continuous and accurate throughout all operating regions.

### 4.1 General Description of Capacitance Modeling

BSIM3v3.2 models capacitance with the following general features:

- Separate effective channel length and width are used for capacitance models.
- The intrinsic capacitance models, capMod=0 and 1, use piece-wise equations. capMod=2 and 3 are smooth and single equation models; therefore both charge and capacitance are continuous and smooth over all regions.
- Threshold voltage is consistent with DC part except for capMod=0, where a long-channel $V_{th}$ is used. Therefore, those effects such as body bias, short/narrow channel and DIBL effects are explicitly considered in capMod=1, 2, and 3.
- Overlap capacitance comprises two parts: (1) a bias-independent component which models the effective overlap capacitance between the gate and the heavily doped source/drain; (2) a gate-bias dependent component between the gate and the lightly doped source/drain region.
Geometry Definition for C-V Modeling

- Bias-independent fringing capacitances are added between the gate and source as well as the gate and drain.

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<th>Name</th>
<th>Function</th>
<th>Default</th>
<th>Unit</th>
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<td>capMod</td>
<td>Flag for capacitance models</td>
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<td>(True)</td>
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<td>acde</td>
<td>Exponential coefficient for $X_{dc}$ for accumulation and depletion regions</td>
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<td>Long channel gate capacitance length offset</td>
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</tr>
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Table 4-1. Model parameters in capacitance models.

4.2 Geometry Definition for C-V Modeling

For capacitance modeling, MOSFET's can be divided into two regions: intrinsic and extrinsic. The intrinsic capacitance is associated with the region between the metallurgical source and drain junction, which is defined by the effective length.
Geometry Definition for C-V Modeling

$(L^{active})$ and width $(W^{active})$ when the gate to S/D region is at flat band voltage. 
$L^{active}$ and $W^{active}$ are defined by Eqs. (4.2.1) through (4.2.4).

\[
L^{active} = L^{drawn} - 2\delta L^{eff}
\]  
(4.2.1)

\[
W^{active} = W^{drawn} - 2\delta W^{eff}
\]  
(4.2.2)

\[
\delta L^{eff} = DLC + \frac{L^{lc}}{L^{ln}} \frac{L^{wc}}{W^{ln}} + \frac{L^{lwlc}}{L^{ln}W^{ln}}
\]  
(4.2.3)

\[
\delta W^{eff} = DWC + \frac{W^{lc}}{L^{ln}} \frac{W^{wc}}{W^{wn}} + \frac{W^{lwlc}}{L^{wn}W^{wn}}
\]  
(4.2.4)

The meanings of DWC and DLC are different from those of Wint and Lint in the I-V model. $L^{active}$ and $W^{active}$ are the effective length and width of the intrinsic device for capacitance calculations. Unlike the case with I-V, we assumed that these dimensions have no voltage bias dependence. The parameter $\delta L^{eff}$ is equal to the source/drain to gate overlap length plus the difference between drawn and actual POLY CD due to processing (gate printing, etching and oxidation) on one side. Overall, a distinction should be made between the effective channel length extracted from the capacitance measurement and from the I-V measurement.

Traditionally, the $L^{eff}$ extracted during I-V model characterization is used to gauge a technology. However this $L^{eff}$ does not necessarily carry a physical meaning. It is just a parameter used in the I-V formulation. This $L^{eff}$ is therefore very sensitive to the I-V equations used and also to the conduction characteristics of the LDD.
Methodology for Intrinsic Capacitance Modeling

region relative to the channel region. A device with a large $L_{eff}$ and a small parasitic resistance can have a similar current drive as another with a smaller $L_{eff}$ but larger $R_{ds}$. In some cases $L_{eff}$ can be larger than the polysilicon gate length giving $L_{eff}$ a dubious physical meaning.

The $L_{active}$ parameter extracted from the capacitance method is a closer representation of the metallurgical junction length (physical length). Due to the graded source/drain junction profile the source to drain length can have a very strong bias dependence. We therefore define $L_{active}$ to be that measured at gate to source/drain flat band voltage. If DWC, DLC and the newly-introduced length/width dependence parameters ($L_{lc}$, $L_{wc}$, $L_{wlc}$, $W_{lc}$, $W_{wc}$ and $W_{wlc}$) are not specified in technology files, BSIM3v3.2 assumes that the DC bias-independent $L_{eff}$ and $W_{eff}$ (Eqs. (2.8.1) - (2.8.4)) will be used for C-V modeling, and DWC, DLC,$L_{lc}$, $L_{wc}$, $L_{wlc}$, $W_{lc}$, $W_{wc}$ and $W_{wlc}$ will be set equal to the values of their DC counterparts (default values).

4.3 Methodology for Intrinsic Capacitance Modeling

4.3.1 Basic Formulation

To ensure charge conservation, terminal charges instead of the terminal voltages are used as state variables. The terminal charges $Q_{g}$, $Q_{b}$, $Q_{s}$, and $Q_{d}$ are the charges associated with the gate, bulk, source, and drain terminals, respectively. The gate charge is comprised of mirror charges from these components: the channel inversion charge ($Q_{inv}$), accumulation charge ($Q_{acc}$) and the substrate depletion charge ($Q_{sub}$).
Methodology for Intrinsic Capacitance Modeling

The accumulation charge and the substrate charge are associated with the substrate while the channel charge comes from the source and drain terminals

\[
\begin{align*}
Q_g &= -(Q_{sub} + Q_{inv} + Q_{acc}) \\
Q_b &= Q_{acc} - Q_{sub} \\
Q_{inv} &= Q_s + Q_d
\end{align*}
\]  

(4.3.1)

The substrate charge can be divided into two components - the substrate charge at zero source-drain bias \((Q_{sub0})\), which is a function of gate to substrate bias, and the additional non-uniform substrate charge in the presence of a drain bias \((\delta Q_{sub})\). \(Q_g\) now becomes

\[
Q_g = -(Q_{inv} + Q_{acc} + Q_{sub0} + \delta Q_{sub})
\]

(4.3.2)

The total charge is computed by integrating the charge along the channel. The threshold voltage along the channel is modified due to the non-uniform substrate charge by

\[
V_{th}(y) = V_{th}(0) + (A_{bulk} - 1)\nu_y
\]

(4.3.3)

\[
\begin{align*}
Q_x &= W_{active} \int_{L_{inv}}^{L_{active}} q_s \, dy = -W_{active} C_{ox} \int_{0}^{L_{active}} (V_{gs} - A_{bulk} \nu_y) \, dy \\
Q_g &= W_{active} \int_{L_{inv}}^{L_{active}} q_s \, dy = W_{active} C_{ox} \int_{0}^{L_{active}} (V_{gs} + V_{th} - V_{FB} - \Phi_s - \nu_y) \, dy \\
Q_s &= W_{active} \int_{L_{inv}}^{L_{active}} q_s \, dy = -W_{active} C_{ox} \int_{0}^{L_{active}} (V_{th} - V_{FB} - \Phi_s + (A_{bulk} - 1)\nu_y) \, dy
\end{align*}
\]  

(4.3.4)
Substituting the following

\[ dy = \frac{dV_y}{\varepsilon_y} \]

and

\[ I_{ds} = \frac{W_{\text{active}} \mu_{\text{ef}} C_{ox}}{L_{\text{active}}} \left( V_{gt} - \frac{A_{\text{bulk}}}{2} V_{ds} \right) \]

into Eq. (4.3.4), we have the following upon integration

\[
Q_c = -W_{\text{active}} L_{\text{active}} C_{ox} \left( V_{gt} - \frac{A_{\text{bulk}}}{2} V_{ds} + \frac{A_{\text{bulk}}^2 V_{ds}^2}{12 \left( V_{gt} - \frac{A_{\text{bulk}}}{2} V_{ds} \right)} \right) \\
Q_g = -Q_{\text{sub} 0} + W_{\text{active}} L_{\text{active}} C_{ox} \left( V_{gt} - \frac{V_{ds}}{2} + \frac{A_{\text{bulk}} V_{ds}^2}{12 \left( V_{gt} - \frac{A_{\text{bulk}}}{2} V_{ds} \right)} \right) \\
Q_b = -Q_g - Q_c = Q_{\text{sub}} + Q_{\text{sub} 0} + Q_{\text{acc}}
\]

where
The inversion charges are supplied from the source and drain electrodes such that $Q_{inv} = Q_s + Q_d$. The ratio of $Q_d$ and $Q_s$ is the charge partitioning ratio. Existing charge partitioning schemes are 0/100, 50/50 and 40/60 ($X PART = 0, 0.5$ and $1$) which are the ratios of $Q_d$ to $Q_s$ in the saturation region. We will revisit charge partitioning in Section 4.3.4.

All capacitances are derived from the charges to ensure charge conservation. Since there are four terminals, there are altogether 16 components. For each component

\[ C_{ij} = \frac{\partial Q_i}{\partial V_j} \]  

(4.3.8)

where $i$ and $j$ denote the transistor terminals. In addition

\[ \sum_i C_{ij} = \sum_j C_{ij} = 0 \]

### 4.3.2 Short Channel Model

In deriving the long channel charge model, mobility is assumed to be constant with no velocity saturation. Therefore in saturation region ($V_{ds} \geq V_{dsat}$), the carrier density at the drain end is zero. Since no channel length modulation is assumed, the channel charge will remain a constant throughout the saturation region. In essence, the channel charge in the
Methodology for Intrinsic Capacitance Modeling

saturation region is assumed to be zero. This is a good approximation for long channel devices but fails when $L_{eff} < 2 \mu m$. If we define a drain bias, $V_{dsat,cv}$ in which the channel charge becomes a constant, we will find that $V_{dsat,cv}$ in general is larger than $V_{dsat}$ but smaller than the long channel $V_{dsat}$ given by $V_{gs}/A_{bulk}$. However, in old long channel charge models, $V_{dsat,cv}$ is set to $V_{gs}/A_{bulk}$ independent of channel length. Consequently, $C_f/L_{eff}$ has no channel length dependence (Eqs. (4.3.6), (4.37)). A pseudo short channel modification from the long channel has been used in the past. It involved the parameter $A_{bulk}$ in the capacitance model which was redefined to be equal to $V_{gs}/V_{dsat}$ thereby equating $V_{dsat,cv}$ and $V_{dsat}$. This overestimated the effect of velocity saturation and resulted in a smaller channel capacitance.

The difficulty in developing a short channel model lies in calculating the charge in the saturation region. Although current continuity stipulates that the charge density in the saturation region is almost constant, it is difficult to calculate accurately the length of the saturation region. Moreover, due to the exponentially increasing lateral electric field, most of the charge in the saturation region are not controlled by the gate electrode. However, one would expect that the total charge in the channel will exponentially decrease with drain bias. Experimentally,

$$V_{sat,iv} < V_{dsat,cv} < V_{sat,iv} \bigg|_{L_{min}} \sim \frac{V_{gs}}{A_{bulk}}$$

(4.3.9)

and $V_{dsat,cv}$ is modeled by the following
Methodology for Intrinsic Capacitance Modeling

\[ V_{dsat,cv} = \frac{V_{gs}-V_{th}-voffcv}{A_{bulk} \left( 1 + \left( \frac{CLC}{L_{active}} \right)^{CLE} \right)} \] (4.3.10a)

\[ V_{g_{eff,cv}} = noff \cdot n_{v_i} \ln \left( 1 + \exp \left( \frac{V_{gs}-V_{th}-voffcv}{noff \cdot n_{v_i}} \right) \right) \] (4.3.10b)

Parameters \( noff \) and \( voffcv \) are introduced to better fit measured data above subthreshold regions. The parameter \( A_{bulk} \) is substituted \( A_{bulk0} \) in the long channel equation by

\[ A_{bulk}' = A_{bulk0} \left( 1 + \left( \frac{CLC}{L_{active}} \right)^{CLE} \right) \] (4.3.11)

\[ A_{bulk} = \left( 1 + \frac{K_{int}}{2 \Phi_s - V_{sref}} \left( \frac{A_0L_{eff}}{L_{eff} + 2 \sqrt{X_j X_{dep}}} + \frac{B_0}{W_{eff} + B_i} \right) \right) \frac{1}{1 + K_{eta} V_{sref}} \] (4.3.11a)

In (4.3.11), parameters \( CLC \) and \( CLE \) are introduced to consider channel-length modulation.
4.3.3 Single Equation Formulation

Traditional MOSFET SPICE capacitance models use piece-wise equations. This can result in discontinuities and non-smoothness at transition regions. The following describes single-equation formulation for charge, capacitance and voltage modeling in capMod=2 and 3.

(a) Transition from depletion to inversion region

The biggest discontinuity is the inversion capacitance at threshold voltage. Conventional models use step functions and the inversion capacitance changes abruptly from 0 to $C_{ox}$. Concurrently, since the substrate charge is a constant, the substrate capacitance drops abruptly to 0 at threshold voltage. Both of these effects can cause oscillation during circuit simulation. Experimentally, capacitance starts to increase almost quadratically at ~0.2V below threshold voltage and levels off at ~0.3V above threshold voltage. For analog and low power circuits, an accurate capacitance model around the threshold voltage is very important.

The non-abrupt channel inversion capacitance and substrate capacitance model is developed from the I-V model which uses a single equation to formulate the subthreshold, transition and inversion regions. The new channel inversion charge model can be modified to any charge model by substituting $V_g$ with $V_{gsteff,cv}$ as in the following

$$Q(V_g) = Q(V_{gsteff,cv}) \tag{4.3.12}$$

Capacitance now becomes
Methodology for Intrinsic Capacitance Modeling

\[ C(V_{st})_t = C(V_{steff, CV}) \frac{\partial V_{steff, CV}}{V_{st, eff}} \] (4.3.13)

The "inversion" charge is always non-zero, even in the accumulation region. However, it decreases exponentially with gate bias in the subthreshold region.

(b) Transition from accumulation to depletion region

An effective flatband voltage \( V_{FBeff} \) is used to smooth out the transition between accumulation and depletion regions. It affects the accumulation and depletion charges

\[ V_{FBeff} = vfb - 0.5 \left[ V_3 + \sqrt{V_3^2 + 4 \delta_3 vfb} \right] \quad \text{where} \quad V_3 = vfb - V_{fb} - \delta_3; \quad \delta_3 = 0.02 \] (4.3.14)

\[ vfb = V_{th} - \Phi_s - K_{lox} \sqrt{\Phi_s - V_{bseff}} \] (4.3.15)

In BSIM3v3.2, a bias-independent \( V_{th} \) is used to calculate \( vfb \), whereas if the version parameter VERSION is smaller than 3.2, the bias-dependent \( V_{th} \) is kept for backward compatibility.

\[ Q_{ac} = -W_{active} L_{active} C_{ox} (V_{FBeff} - vfb) \] (4.3.16)

\[ Q_{sub} = -W_{active} L_{active} C_{ox} \frac{K_{lox}^2}{2} \left[ -1 + \sqrt{1 + \frac{4(V_{gs} - V_{FBeff} - V_{steff, CV} - V_{bseff})}{K_{lox}^2}} \right] \] (4.3.17)
(c) Transition from linear to saturation region

An effective $V_{\text{dss}^*}, V_{\text{cveff}}$, is used to smooth out the transition between linear and saturation regions. It affects the inversion charge.

$$V_{\text{cveff}} = V_{\text{dss}^*} - 0.5 \left( V_4 + \sqrt{V_4^2 + 4\delta_4 V_{\text{dss}^*}} \right) \quad \text{where} \quad V_4 = V_{\text{dss}^*} - V_{ds} - \delta_4; \quad \delta_4 = 0.02$$  \hfill (4.3.18)

$$Q_{\text{inv}} = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( V_{\text{goff},cv} \frac{A_{\text{bulk}}}{2} V_{\text{cveff}} + \frac{A_{\text{bulk}}^2 V_{\text{cveff}}^2}{2 \left( V_{\text{goff},cv} \frac{A_{\text{bulk}}}{2} V_{\text{cveff}} \right)} \right)$$  \hfill (4.3.19)

$$\delta Q_{\text{nh}} = W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( \frac{1 - A_{\text{bulk}}}{2} V_{\text{cveff}} - \frac{(1 - A_{\text{bulk}}) A_{\text{bulk}} V_{\text{cveff}}^2}{2 \left( V_{\text{goff},cv} \frac{A_{\text{bulk}}}{2} V_{\text{cveff}} \right)} \right)$$  \hfill (4.3.20)

Below is a list of all the three partitioning schemes for the inversion charge:

(i) The 50/50 charge partition

This is the simplest of all partitioning schemes in which the inversion charges are assumed to be contributed equally from the source and drain nodes.
Methodology for Intrinsic Capacitance Modeling

\[ Q_s = Q_d = 0.5Q_{tot} = -\frac{W_{active}}{2} \frac{L_{active}}{C_m} \left( V_{gs_{eff,cv}} - \frac{A_{bulk'}}{2} V_{cvgf} + \frac{A_{bulk'}^2 V_{cvgf}^2}{12} \right) \]  \hspace{1cm} (4.3.21)

(ii) The 40/60 channel-charge partition

This is the most physical model of the three partitioning schemes in which the channel charges are allocated to the source and drain terminals by assuming a linear dependence on the position \( y \).

\[
\begin{align*}
Q_s &= W_{active} \int_a^{L_{active}} d_y \left( 1 - \frac{y}{L_{active}} \right) dy \\
Q_d &= W_{active} \int_y^{L_{active}} d_y q_e \frac{y}{L_{active}} dy 
\end{align*}
\]  \hspace{1cm} (4.3.22)

\[ Q = \frac{W_{min} L_{min} C_m}{2} \left( V_{gs_{eff,cv}} - \frac{A_{min'}}{2} V_{cvgf} \right)^3 \left( \frac{4}{3} V_{gs_{eff,cv}}^2 (A_{min'} V_{cvgf}) + \frac{2}{3} V_{gs_{eff,cv}} (A_{min'} V_{cvgf})^2 + \frac{2}{15} (A_{min'} V_{cvgf})^3 \right) \]  \hspace{1cm} (4.3.23)

\[ Q = \frac{W_{min} L_{min} C_m}{2} \left( V_{gs_{eff,cv}} - \frac{A_{min'}}{2} V_{cvgf} \right)^3 \left( \frac{5}{3} V_{gs_{eff,cv}}^2 (A_{min'} V_{cvgf}) + V_{gs_{eff,cv}} (A_{min'} V_{cvgf})^2 + \frac{1}{5} (A_{min'} V_{cvgf})^3 \right) \]  \hspace{1cm} (4.3.24)

(iii) The 0/100 Charge Partition

In fast transient simulations, the use of a quasi-static model may result in a large unrealistic drain current spike. This partitioning scheme is developed to artificially suppress the drain current spike by assigning all inversion
charges in the saturation region to the source electrode. Notice that this charge partitioning scheme will still give drain current spikes in the linear region and aggravate the source current spike problem.

\[
Q_s = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( \frac{V_{\text{gs}}}{2} + \frac{A_{\text{bulk}} V_{\text{ceff}}}{4} - \frac{(A_{\text{bulk}} V_{\text{ceff}})^2}{24 \left( V_{\text{gs}} - \frac{A_{\text{bulk}} V_{\text{ceff}}}{2} \right)} \right) 
\]

(4.3.25)

\[
Q_d = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( \frac{V_{\text{gs}}}{2} - \frac{3 A_{\text{bulk}} V_{\text{ceff}}}{4} + \frac{(A_{\text{bulk}} V_{\text{ceff}})^2}{8 \left( V_{\text{gs}} - \frac{A_{\text{bulk}} V_{\text{ceff}}}{2} \right)} \right) 
\]

(4.3.26)

(d) Bias-dependent threshold voltage effects on capacitance

Consistent Vth between DC and CV is important for accurate circuit simulation. capMod=1, 2 and 3 use the same Vth as in the DC model. Therefore, those effects, such as body bias, DIBL and short-channel effects are all explicitly considered in capacitance modeling. In deriving the capacitances additional differentiations are needed to account for the dependence of threshold voltage on drain and substrate biases.

4.4 Charge-Thickness Capacitance Model

Current MOSFET models in SPICE generally overestimate the intrinsic capacitance and usually are not smooth at \( V_{fb} \) and \( V_{th} \). The discrepancy is more
pronounced in thinner $t_{ox}$ devices due to the assumption of inversion and accumulation charge being located at the interface. The charge sheet model or the band-gap($E_g$)-reduction model of quantum effect [31] improves the $\Phi_b$ and thus the $V_{th}$ modeling but is inadequate for CV because they assume zero charge thickness. Numerical quantum simulation results in Figure 4-1 indicate the significant charge thickness in all regions of the CV curves [32].

This section describes the concepts used in the charge-thickness model (CTM). Appendix B lists all charge equations. A full report and analysis of the CTM model can be found in [32].

CTM is a charge-based model and therefore starts with the DC charge thickness, $X_{DC}$. The charge thickness introduces a capacitance in series with $C_{ox}$ as illustrated in Figure 4-2, resulting in an effective $C_{ox}$, $C_{oxeff}$. Based on numerical self-consistent solution of Schrodinger, Poisson and Fermi-Dirac equations, universal and analytical $X_{DC}$ models have been developed. $C_{oxeff}$ can be expressed as
Charge-Thickess Capacitance Model

\[ C_{\text{ox eff}} = \frac{C_{\text{ox}} C_{\text{cen}}}{C_{\text{ox}} + C_{\text{cen}}} \]  

\[ C_{\text{cen}} = \frac{\varepsilon_{\text{si}}}{X_{\text{DC}}} \]

Figure 4-2. Charge-thickness capacitance concept in CTM. \( V_{\text{gse}} \) accounts for the poly depletion effect.

(i) \( X_{\text{DC}} \) for accumulation and depletion

The DC charge thickness in the accumulation and depletion regions can be expressed by [32]

\[ X_{\text{DC}} = \frac{1}{3} L_{\text{debye}} \exp \left[ \frac{acde \left( \frac{N_{\text{sub}}}{2 \times 10^{16}} \right)^{-0.25} V_{\text{gs}} - V_{\text{bn}} - V_{\text{fh}}}{T_{\text{ox}}} \right] \]
where $X_{DC}$ is in the unit of cm and $(V_{gs} - V_{bs} - V_{fb}) / T_{ox}$ has a unit of MV/cm. The model parameter *acde* is introduced for better fitting with a default value of 1. For numerical stability, (4.4.2) is replaced by (4.4.3)

\[
X_{DC} = X_{\text{max}} - \frac{1}{2} \left( X_0 + \sqrt{X_0^2 + 4\delta_x X_{\text{max}}} \right)
\]

where

\[
X_0 = X_{\text{max}} - X_{DC} - \delta_x
\]

and $X_{\text{max}} = L_{\text{debye}}/3$; $\delta_x = 10^{-3} T_{ox}$.

(ii) *$X_{DC}$ of inversion charge*

The inversion charge layer thickness [32] can be formulated as

\[
X_{DC} = \frac{1.9 \times 10^7}{1 + \left( \frac{V_{\text{enff}} + 4(V_{in} - V_{fb} - 2\Phi_B)}{2T_m} \right)^{0.7}} \quad [\text{cm}]
\]

Through $V_{fb}$ in (4.3.30), this equation is found to be applicable to N* or P* poly-Si gates as well as other future gate materials. Figure 4-3 illustrates the universality of (4.3.30) as verified by the numerical quantum simulations, where the x-axe
(iii) Body charge thickness in inversion

In inversion region, the body charge thickness effect is accurately modeled by including the deviation of the surface potential \( \Phi_s \) from \( 2\Phi_B \) [32]

\[
\Phi_\delta = \Phi_s - 2\Phi_B = V_t \ln \left( \frac{V_{st,eff} + 2K_{ox} \sqrt{2\Phi_B}}{\text{moin} K_{ox} V_t^2} + 1 \right)
\]  

(4.4.5)

where the model parameter \( \text{moin} \) (defaulting to 15) is introduced for better fit to different technologies. The inversion channel charge density is therefore derived as

\[
q_{inv} = -C_{ox,eff} \left( V_{st,eff} - \Phi_\delta \right)
\]

(4.4.6)
Extrinsic Capacitance

Figure 4-4 illustrates the universality of CTM model by comparing $C_{gs}$ of a SiON/Ta$_2$O$_5$/TiN gate stack structure with an equivalent $T_{ox}$ of 1.8nm between data, numerical quantum simulation and modeling [32].

![Diagram showing comparison of measured and simulated $C_{gs}$](image)

**Figure 4-4.** Universality of CTM is demonstrated by modeling the $C_{gs}$ of a 1.8nm equivalent $T_{ox}$ NMOSFET with SiON/Ta$_2$O$_5$/TiN gate stack.

### 4.5 Extrinsic Capacitance

#### 4.5.1 Fringing Capacitance

The fringing capacitance consists of a bias independent outer fringing capacitance and a bias dependent inner fringing capacitance. Only the bias independent outer fringing capacitance is implemented. Experimentally, it is virtually impossible to separate this capacitance with the overlap capacitance. Nonetheless, the outer fringing capacitance can be theoretically calculated by

\[
CF = \frac{2\varepsilon_{ox}}{\pi} \ln \left( 1 + \frac{t_{pole}}{T_{ox}} \right)
\]  

(4.5.1)
Extrinsic Capacitance

where \( t_{\text{poly}} \) is equal to \( 4 \times 10^{-7} \) m. \( CF \) is a model parameter.

### 4.5.2 Overlap Capacitance

An accurate model for the overlap capacitance is essential. This is especially true for the drain side where the effect of the capacitance is amplified by the transistor gain. In old capacitance models this capacitance is assumed to be bias independent. However, experimental data show that the overlap capacitance changes with gate to source and gate to drain biases. In a single drain structure or the heavily doped S/D to gate overlap region in a LDD structure the bias dependence is the result of depleting the surface of the source and drain regions. Since the modulation is expected to be very small, we can model this region with a constant capacitance. However in LDD MOSFETs a substantial portion of the LDD region can be depleted, both in the vertical and lateral directions. This can lead to a large reduction of overlap capacitance. This LDD region can be in accumulation or depletion. We use a single equation for both regions by using such smoothing parameters as \( V_{gs,\text{overlap}} \) and \( V_{gd,\text{overlap}} \) for the source and drain side, respectively. Unlike the case with the intrinsic capacitance, the overlap capacitances are reciprocal. In other words, \( C_{gs,\text{overlap}} = C_{sg,\text{overlap}} \) and \( C_{gd,\text{overlap}} = C_{dg,\text{overlap}} \).

(i) Source Overlap Charge

\[
\frac{Q_{\text{overlap,s}}}{W_{\text{active}}} = C_{GSO} \cdot V_{gs} + C_{GS} \left( V_{gs} - V_{gs,\text{overlap}} - \frac{CKAPPA}{2} \left( -1 + \sqrt{1 - \frac{4V_{gs,\text{overlap}}}{CKAPPA}} \right) \right)
\]
Extrinsic Capacitance

(4.5.3)

\[ V_{gs, overlap} = \frac{1}{2} \left( V_{gs} + \delta_1 - \sqrt{\left(V_{gs} + \delta_1\right)^2 + 4\delta_1} \right), \quad \delta_1 = 0.02 \]

where \( \kappaappa \) is a model parameter. \( \kappaappa \) is related to the average doping of LDD region by

\[ \kappaappa = \frac{2e_{\text{eff}} q N_{LDD}}{C_{\alpha}} \]

The typical value for \( N_{LDD} \) is \( 5 \times 10^{17} \text{ cm}^{-3} \).

(ii) Drain Overlap Charge

(4.5.4)

\[ \frac{Q_{\text{overlap,d}}}{W_{\text{active}}} = \text{CGDO} \cdot V_{gd} + \text{CGDI} \left( V_{gd} - V_{gd, overlap} - \frac{\kappaappa}{2} \left( 1 + \sqrt{1 - \frac{4V_{gd, overlap}}{\kappaappa}} \right) \right) \]

(4.5.5)

\[ V_{gd, overlap} = \frac{1}{2} \left( V_{gd} + \delta_1 - \sqrt{\left(V_{gd} + \delta_1\right)^2 + 4\delta_1} \right), \quad \delta_1 = 0.02 \]

(iii) Gate Overlap Charge

(4.5.6)

\[ Q_{\text{overlap,g}} = -\left( Q_{\text{overlap,s}} + Q_{\text{overlap,d}} \right) \]

In the above expressions, if \( \text{CGSO} \) and \( \text{CGDO} \) (the heavily doped S/D region to gate overlap capacitance) are not given, they are calculated according to the following

\[ \text{CGSO} = (DLC \cdot C_{\alpha}) - \text{CGS1} \quad \text{if DLC is given and DLC > 0} \]
Extrinsic Capacitance

\[ CGS0 = 0 \ (\text{if the previously calculated } CGS0 \text{ is less than 0}) \]

\[ CGS0 = 0.6 \times j \times C_{ox} \ (\text{otherwise}) \]

\[ CGD0 = (DLC \times C_{ox}) - CGD1 \ (\text{if DLC is given and } DLC > CGD1/C_{ox}) \]

\[ CGD0 = 0 \ (\text{if previously calculated } CDG0 \text{ is less than 0}) \]

\[ CGD0 = 0.6 \times j \times C_{ox} \ (\text{otherwise}). \]
5.1 Background Information

As MOSFET's become more performance-driven, the need for accurate prediction of circuit performance near cut-off frequency or under very rapid transient operation becomes more essential. However, most SPICE MOSFET models are based on Quasi-Static (QS) assumptions. In other words, the finite charging time for the inversion layer is ignored. When these models are used with 40/60 charge partitioning, unrealistically drain current spikes frequently occur [33]. In addition, the inability of these models to accurately simulate channel charge re-distribution causes problems in fast switched-capacitor type circuits. Many Non-Quasi-Static (NQS) models have been published, but these models (1) assume, unrealistically, no velocity saturation and (2) are complex in their formulations with considerable simulation time.

5.2 The NQS Model in BSIM3v3.2

The NQS model has been re-implemented in BSIM3v3.2 to improve the simulation performance and accuracy. This model is based on the channel charge relaxation time approach. A new charge partitioning scheme is used, which is physically consistent with quasi-static CV model.
5.3 Model Formulation

The channel of a MOSFET is analogous to a bias dependent RC distributed transmission line (Figure 5-1a). In the Quasi-Static (QS) approach, the gate capacitor node is lumped with both the external source and drain nodes (Figure 5-1b). This ignores the finite time for the channel charge to build-up. One Non-Quasi-Static (NQS) solution is to represent the channel as $n$ transistors in series (Figure 5-1c). This model, although accurate, comes at the expense of simulation time. The NQS model in BSIM3v3.2 was based on the circuit of Figure 5-1d. This Elmore equivalent circuit models channel charge build-up accurately because it retains the lowest frequency pole of the original RC network (Figure 5-1a). The NQS model has two parameters as follows. The model flag, nqsMod, is now only an element (instance) parameter, no longer a model parameter.

<table>
<thead>
<tr>
<th>Name</th>
<th>Function</th>
<th>Default</th>
<th>Unit</th>
</tr>
</thead>
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<td>nqsMod</td>
<td>Instance flag for the NQS model</td>
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<td>none</td>
</tr>
<tr>
<td>elm</td>
<td>Elmore constant</td>
<td>5</td>
<td>none</td>
</tr>
</tbody>
</table>

Table 5-1. NQS model and instance parameters.
5.3.1 SPICE sub-circuit for NQS model

Figure 5-2 gives the RC-subcircuit of NQS model for SPICE implementation. An additional node, $Q_{def}(t)$, is created to keep track of the amount of deficit/surplus channel charge necessary to reach the equilibrium based on the relaxation time approach. The bias-dependent resistance $R$ ($1/R=G_{inv}$) can be determined from the RC time constant ($\tau$). The current source $i_{cheq}(t)$ results from the equilibrium channel charge, $Q_{cheq}(t)$. The capacitor $C$ is multiplied by a scaling factor $C_{fact}$ (with a typical value of $1 \times 10^{-9}$) to improve simulation accuracy. $Q_{def}$ now becomes

$$ Q_{def}(t) = V_{def} \times (1 \cdot C_{fact}) $$
5.3.2 Relaxation time

The relaxation time $\tau$ is modeled as two components: $\tau_{\text{drift}}$ and $\tau_{\text{diff}}$. In strong inversion region, $\tau$ is determined by $\tau_{\text{drift}}$, which, in turn, is determined by the Elmore resistance $R_{\text{elm}}$; in subthreshold region, $\tau_{\text{diff}}$ dominates. $\tau$ is expressed by

$$
\frac{1}{\tau} = \frac{1}{\tau_{\text{diff}}} + \frac{1}{\tau_{\text{drift}}}
$$

(5.3.2)

$R_{\text{elm}}$ in strong inversion is calculated from the channel resistance as

$$
R_{\text{elm}} = \frac{L_{\text{eff}}^2}{\mu_{\text{elm}} Q_{\text{ch}}} \approx \frac{L_{\text{eff}}^2}{\mu_{0} Q_{\text{cheq}}}
$$

(5.3.3)
where $elm$ is the Elmore constant of the RC channel network with a theoretical value of 5. The quasi-static (or equilibrium) channel charge $Q_{cheq}(t)$, equal to $Q_{inv}$ of capMod=0, 1, 2 and 3, is used to approximate the actual channel charge $Q_{ch}(t)$. $\tau_{drift}$ is formulated as

$$\tau_{drift} \approx R_{elm} \cdot C_{ox} W_{eff} L_{eff} = \frac{C_{ox} W_{eff} L_{eff}^3}{elm \cdot \mu_0 Q_{cheq}}$$

(5.3.4)

$\tau_{diff}$ has the form of

$$\tau_{diff} = \frac{qL_{eff}^2}{16 \cdot \mu_0 kT}$$

(5.3.5)

### 5.3.3 Terminal charging current and charge partitioning

Considering both the transport and charging component, the total current related to the terminals D, G and S can be written as

$$i_{d,g,s}(t) = I_{d,g,s}(DC) + \frac{\partial Q_{d,g,s}(t)}{\partial t}$$

(5.3.6)

Based on the relaxation time approach, the terminal charge and corresponding charging current can be formulated by
Model Formulation

\[ Q_{\text{def}}(t) = Q_{\text{cheq}}(t) - Q_{ch}(t) \]  

and

\[ \frac{\partial Q_{\text{def}}(t)}{\partial t} = \frac{\partial Q_{\text{cheq}}(t)}{\partial t} - \frac{Q_{\text{def}}(t)}{\tau} \]  

\[ \frac{\partial Q_{\text{d,g,s}}(t)}{\partial t} = D, G, S_{\text{xp}} \frac{Q_{\text{def}}(t)}{\tau} \]

where \( D, G, S_{\text{xp}} \) are the NQS channel charge partitioning number for terminals D, G and S, respectively; \( D_{\text{xp}} + S_{\text{xp}} = 1 \) and \( G_{\text{xp}} = -1 \). It is important for \( D_{\text{xp}} \) and \( S_{\text{xp}} \) to be consistent with the quasi-static charge partitioning number \( X_{\text{part}} \) and to be equal \( (D_{\text{xp}} = S_{\text{xp}}) \) at \( V_{ds} = 0 \) (which is not the case in the previous version), where the transistor operation mode changes (between forward and reverse modes). Based on this consideration, \( D_{\text{xp}} \) is now formulated as

\[ D_{\text{xp}} = \frac{Q_d}{Q_d + Q_s} = \frac{Q_d}{Q_{\text{cheq}}} \]

which is now bias dependent. For example, the derivatives of \( D_{\text{xp}} \) can be easily obtained based on the quasi-static results:

\[ \frac{dD_{\text{xp}}}{dV_i} = \frac{1}{Q_{\text{cheq}}} \left( S_{\text{xp}} C_{di} - D_{\text{xp}} C_{si} \right) \]
where \( i \) represents the four terminals and \( C_{di} \) and \( C_{xi} \) are the intrinsic capacitances calculated from the quasi-static analysis. The corresponding values for \( S_{part} \) can be derived from the fact that \( D_{part} + S_{part} = 1 \).

In the accumulation and depletion regions, Eq. (5.3.9) is simplified as

If \( \text{XPART} < 0.5 \), \( D_{\text{part}} = 0.4 \);
Else if \( \text{XPART} > 0.5 \), \( D_{\text{part}} = 0.0 \);
Else \( D_{\text{part}} = 0.5 \);

### 5.3.4 Derivation of nodal conductances

This section gives some examples of how to derive the nodal conductances related to NQS for transient analysis. By noting that \( \tau = RC \), \( G_{\tau u} \) can be derived as

\[
G_{\tau u} = \frac{C_{\text{fact}}}{\tau}
\]

(5.3.11)

\( \tau \) is given by Eq. (5.3.2). Based on Eq. (5.3.8b), the self-conductance due to NQS at the transistor node \( D \) can be derived as

\[
\frac{dD_{\text{part}}}{dV_d} \cdot (G_{\tau u} \cdot V_{\text{def}}) + D_{\text{part}} \cdot V_{\text{def}} \cdot \frac{dG_{\tau u}}{dV_d}
\]

(5.3.12)

The trans-conductance due to NQS on the node \( D \) relative to the node of \( Q_{\text{def}} \) can be derived as

\[
D_{\text{part}} \cdot G_{\tau u}
\]

(5.3.13)
Other conductances can also be obtained in a similar way.
Parameter extraction is an important part of model development. Many different extraction methods have been developed [23, 24]. The appropriate methodology depends on the model and on the way the model is used. A combination of a local optimization and the group device extraction strategy is adopted for parameter extraction.

6.1 Optimization strategy

There are two main, different optimization strategies: global optimization and local optimization. Global optimization relies on the explicit use of a computer to find one set of model parameters which will best fit the available experimental (measured) data. This methodology may give the minimum average error between measured and simulated (calculated) data points, but it also treats each parameter as a "fitting" parameter. Physical parameters extracted in such a manner might yield values that are not consistent with their physical intent.

In local optimization, many parameters are extracted independently of one another. Parameters are extracted from device bias conditions which correspond to dominant physical mechanisms. Parameters which are extracted in this manner might not fit experimental data in all the bias conditions. Nonetheless, these extraction methodologies are developed specifically with respect to a given parameter's physical meaning. If properly executed, it should, overall, predict
Extraction Strategies

device performance quite well. Values extracted in this manner will now have some physical relevance.

6.2 Extraction Strategies

Two different strategies are available for extracting parameters: the single device extraction strategy and group device extraction strategy. In single device extraction strategy, experimental data from a single device is used to extract a complete set of model parameters. This strategy will fit one device very well but will not fit other devices with different geometries. Furthermore, single device extraction strategy can not guarantee that the extracted parameters are physical. If only one set of channel length and width is used, parameters related to channel length and channel width dependencies can not be determined.

BSIM3v3 uses group device extraction strategy. This requires measured data from devices with different geometries. All devices are measured under the same bias conditions. The resulting fit might not be absolutely perfect for any single device but will be better for the group of devices under consideration.

6.3 Extraction Procedure

6.3.1 Parameter Extraction Requirements

One large size device and two sets of smaller-sized devices are needed to extract parameters, as shown in Figure 6-1.
Extraction Procedure

The large-sized device ($W \geq 10\mu m$, $L \geq 10\mu m$) is used to extract parameters which are independent of short/narrow channel effects and parasitic resistance. Specifically, these are: mobility, the large-sized device threshold voltage $V_{Th,\text{ideal}}$, and the body effect coefficients $K_1$ and $K_2$ which depend on the vertical doping concentration distribution. The set of devices with a fixed large channel width but different channel lengths are used to extract parameters which are related to the short channel effects. Similarly, the set of devices with a fixed, long channel length but different channel widths are used to extract parameters which are related to narrow width.
Extraction Procedure

Regardless of device geometry, each device will have to be measured under four, distinct bias conditions.

1. $I_{ds} \text{ vs. } V_{gs} \text{ @ } V_{ds} = 0.05 \text{V}$ with different $V_{bs}$
2. $I_{ds} \text{ vs. } V_{ds} \text{ @ } V_{bs} = 0 \text{V}$ with different $V_{gs}$
3. $I_{ds} \text{ vs. } V_{gs} \text{ @ } V_{ds} = V_{dd}$ with different $V_{bs}$ ($V_{dd}$ is the maximum drain voltage).
4. $I_{ds} \text{ vs. } V_{ds} \text{ @ } V_{bs} = V_{bb}$ with different $V_{gs}$ ($V_{bb}$ is the maximum body bias).

6.3.2 Optimization

The optimization process recommended is a combination of Newton-Raphson's iteration and linear-squares fit of either one, two, or three variables. This methodology was discussed by M. C. Jeng [18]. A flow chart of this optimization process is shown in Figure 6-2. The model equation is first arranged in a form suitable for Newton-Raphson's iteration as shown in Eq. (6.3.1):

$$f_{exp}(P_{0}, P_{20}, P_{30}) - f_{sim}(P_{1}^{(m)}, P_{2}^{(m)}, P_{3}^{(m)}) = \frac{\partial f_{sim}}{\partial P_{1}} \Delta P_{1}^{m} + \frac{\partial f_{sim}}{\partial P_{2}} \Delta P_{2}^{m} + \frac{\partial f_{sim}}{\partial P_{3}} \Delta P_{3}^{m}$$

The variable $f_{sim}$ is the objective function to be optimized. The variable $f_{exp}$ stands for the experimental data. $P_{10}$, $P_{20}$, and $P_{30}$ represent the desired extracted parameter values. $P_{1}^{(m)}$, $P_{2}^{(m)}$ and $P_{3}^{(m)}$ represent parameter values after the $m$th iteration.
To change Eq. (6.3.1) into a form that a linear least-squares fit routine can be used (i.e. in a form of $y = a + bx_1 + cx_2$), both sides of the Eq. (6.3.1) are divided by $\frac{\partial f_{\text{sim}}}{\partial P_1}$. This gives the change in $P_1$, $\Delta P_1^{(m)}$, for the next iteration such that:

$$\frac{\Delta P}{P_i^{(m)}} < \delta$$
where \( i = 1, 2, 3 \) for this example. The \((m+1)\) parameter values for \( P_2 \) and \( P_3 \) are obtained in an identical fashion. This process is repeated until the incremental parameter change in parameter values \( \Delta P_i^{(m)} \) are smaller than a pre-determined value. At this point, the parameters \( P_1, P_2, \) and \( P_3 \) have been extracted.

### 6.3.3 Extraction Routine

Before any model parameters can be extracted, some process parameters have to be provided. They are listed below in Table 6-1:

<table>
<thead>
<tr>
<th>Input Parameters Names</th>
<th>Physical Meaning</th>
</tr>
</thead>
<tbody>
<tr>
<td>( T_{ox} )</td>
<td>Gate oxide thickness</td>
</tr>
<tr>
<td>( N_{ch} )</td>
<td>Doping concentration in the channel</td>
</tr>
<tr>
<td>( T )</td>
<td>Temperature at which the data is taken</td>
</tr>
<tr>
<td>( L_{drawn} )</td>
<td>Mask level channel length</td>
</tr>
<tr>
<td>( W_{drawn} )</td>
<td>Mask level channel width</td>
</tr>
<tr>
<td>( X_j )</td>
<td>Junction depth</td>
</tr>
</tbody>
</table>

**Table 6-1. Prerequisite input parameters prior to extraction process.**

The parameters are extracted in the following procedure. These procedures are based on a physical understanding of the model and based on local
Extraction Procedure

optimization. (Note: Fitting Target Data refers to measurement data used for model extraction.)

**Step 1**

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Device &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$V_{th0}, K_1, K_2$</td>
<td>Large Size Device (Large W &amp; L).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $V_{th}(V_{bs})$</td>
<td>$I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = 0.05V$ at Different $V_{bs}$</td>
</tr>
<tr>
<td></td>
<td>Extracted Experimental Data $V_{th}(V_{bs})$</td>
</tr>
</tbody>
</table>

**Step 2**

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\mu_0, U_a, U_b, U_c$</td>
<td>Large Size Device (Large W &amp; L).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: Strong Inversion region $I_{ds}(V_{gs}, V_{bs})$</td>
<td>$I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = 0.05V$ at Different $V_{bs}$</td>
</tr>
</tbody>
</table>

**Step 3**

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L_{int}, R_{ds}(R_{dsw}, W, V_{bs})$</td>
<td>One Set of Devices (Large and Fixed W &amp; Different L).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: Strong Inversion region $I_{ds}(V_{gs}, V_{bs})$</td>
<td>$I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = 0.05V$ at Different $V_{bs}$</td>
</tr>
</tbody>
</table>
### Step 4

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$W_{int}, R_{ds}(R_{dsW}, W, V_{bs})$</td>
<td>One Set of Devices (Large and Fixed $L$ &amp; Different $W$). $I_{ds} \text{ vs. } V_{gs} @ V_{ds} = 0.05 \text{V at Different } V_{bs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: Strong Inversion region $I_{ds}(V_{gs}, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>

### Step 5

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$R_{dsW, Prwb, Wr}$</td>
<td>$R_{ds}(R_{dsW}, W, V_{bs})$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $R_{ds}(R_{dsW}, W, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>

### Step 6

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D_{vth} D_{vth}, D_{vth}, Nlx$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $V_{th}(V_{bs}, L, W)$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $V_{th}(V_{bs}, L, W)$</td>
<td></td>
</tr>
</tbody>
</table>
### Extraction Procedure

#### Step 7

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
</table>
| $D_{v1o_w}$, $D_{v1l_w}$, $D_{v2w}$ | One Set of Devices (Large and Fixed $L$ & Different $W$).  
Fitting Target Exp. Data: $V_{th}(V_{bs}, L, W)$ | $V_{th}(V_{bs}, L, W)$ |

#### Step 8

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
</table>
| $K_3$, $K_{3b}$, $W_0$ | One Set of Devices (Large and Fixed $L$ & Different $W$).  
Fitting Target Exp. Data: $V_{th}(V_{bs}, L, W)$ | $V_{th}(V_{bs}, L, W)$ |

#### Step 9

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
</table>
| $V_{off}$, $N_{factor}$, $C_{dso}$, $C_{dscb}$ | One Set of Devices (Large and Fixed $W$ & Different $L$).  
Fitting Target Exp. Data: Subthreshold region $I_{ds}(V_{gs}, V_{bs})$ |
| $I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = 0.05V$ at Different $V_{bs}$ |

#### Step 10

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
</table>
### Extraction Procedure

<table>
<thead>
<tr>
<th>$C_{dsdx}$</th>
<th>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_{ds}$ vs. $V_{gs}$ @ $V_{bs} = V_{bb}$ at Different $V_{ds}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fitting Target Exp. Data: Subthreshold region $I_{ds}(V_{gs}, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>

#### Step 11

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$dWb$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = 0.05V$ at Different $V_{bs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: Strong Inversion region $I_{ds}(V_{gs}, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>

#### Step 12

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$v_{sat}$, $A_0$, $A_{gs}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_{ds}$ vs. $V_{ds}$ @ $V_{bs} = 0V$ at Different $V_{gs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $I_{sat}(V_{gs}, V_{bs})/W$</td>
<td></td>
</tr>
<tr>
<td>$A_1$, $A_2$ (PMOS Only)</td>
<td></td>
</tr>
<tr>
<td>Fitting Target Exp. Data $V_{sat}(V_{gs})$</td>
<td></td>
</tr>
</tbody>
</table>

#### Step 13

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
</table>
### Extraction Procedure

<table>
<thead>
<tr>
<th>$B_0, B_1$</th>
<th>One Set of Devices (Large and Fixed $L$ &amp; Different $W$). $I_{ds}$ vs. $V_{ds}$ @ $V_{bs}$ = 0V at Different $V_{gs}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Fitting Target Exp. Data: $I_{sat}(V_{gs}, V_{bs})/W$</td>
<td></td>
</tr>
<tr>
<td>$dW_g$</td>
<td>One Set of Devices (Large and Fixed $L$ &amp; Different $W$). $I_{ds}$ vs. $V_{ds}$ @ $V_{bs}$ = 0V at Different $V_{gs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $I_{sat}(V_{gs}, V_{bs})/W$</td>
<td></td>
</tr>
<tr>
<td>$P_{sbe1, sbe2}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_{ds}$ vs. $V_{ds}$ @ $V_{bs}$ = 0V at Different $V_{gs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $R_{out}(V_{gs}, V_{ds})$</td>
<td></td>
</tr>
<tr>
<td>$P_{clm}$ $\theta(D_{roul}, P_{diblc1}, P_{diblc2}, L), P_{avg}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_{ds}$ vs. $V_{ds}$ @ $V_{bs}$ = 0V at Different $V_{gs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $R_{out}(V_{gs}, V_{ds})$</td>
<td></td>
</tr>
</tbody>
</table>
### Step 17

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$D_{rout}, P_{diblc}, P_{diblc2}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$).  $\theta(D_{rout}, P_{diblc}, P_{diblc2}, L)$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $\theta(D_{rout}, P_{diblc}, P_{diblc2}, L)$</td>
<td></td>
</tr>
</tbody>
</table>

### Step 18

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$P_{diblcb}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_d$ vs. $V_{gs}$ @ fixed $V_{gs}$ at Different $V_{bs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $\theta(D_{rout}, P_{diblc}, P_{diblc2}, L, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>

### Step 19

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\theta_{dib}(Eta0, EtaB, Dsub, L)$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$). $I_d$ vs. $V_{gs}$ @ $V_{ds} = V_{dd}$ at Different $V_{bs}$</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: Subthreshold region $I_d(V_{gs}, V_{bs})$</td>
<td></td>
</tr>
</tbody>
</table>
### Step 20

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\eta_0, \eta_{ab}, D_{sub}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $\theta_{dib}(\eta_0, \eta_{ab}, L)$</td>
<td>$I_{ds}$ vs. $V_{gs}$ @ $V_{ds} = V_{dd}$ at Different $V_{bs}$</td>
</tr>
</tbody>
</table>

### Step 21

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$K_{eta}$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $I_{sat}(V_{gs}, V_{bs})/W$</td>
<td>$I_{ds}$ vs. $V_{ds}$ @ $V_{bs} = V_{bb}$ at Different $V_{gs}$</td>
</tr>
</tbody>
</table>

### Step 22

<table>
<thead>
<tr>
<th>Extracted Parameters &amp; Fitting Target Data</th>
<th>Devices &amp; Experimental Data</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\alpha_0, \alpha_1, \beta_0$</td>
<td>One Set of Devices (Large and Fixed $W$ &amp; Different $L$).</td>
</tr>
<tr>
<td>Fitting Target Exp. Data: $I_{sub}(V_{gs}, V_{bs})/W$</td>
<td>$I_{ds}$ vs. $V_{ds}$ @ $V_{bs} = V_{bb}$ at Different $V_{ds}$</td>
</tr>
</tbody>
</table>
6.4 Notes on Parameter Extraction

6.4.1 Parameters with Special Notes

Below is a list of model parameters which have special notes for parameter extraction.

<table>
<thead>
<tr>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default Value</th>
<th>Unit</th>
<th>Notes</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vth0</td>
<td>Threshold voltage for large W and L device @ Vbs=0V</td>
<td>0.7 (NMOS) -0.7 (PMOS)</td>
<td>V</td>
<td>nI-1</td>
</tr>
<tr>
<td>K1</td>
<td>First order body effect coefficient</td>
<td>0.5</td>
<td>V(^{1/2})</td>
<td>nI-2</td>
</tr>
<tr>
<td>K2</td>
<td>Second order body effect coefficient</td>
<td>0</td>
<td>none</td>
<td>nI-2</td>
</tr>
<tr>
<td>Vbm</td>
<td>Maximum applied body bias</td>
<td>-3</td>
<td>V</td>
<td>nI-2</td>
</tr>
<tr>
<td>Nch</td>
<td>Channel doping concentration</td>
<td>1.7E17</td>
<td>1/cm(^3)</td>
<td>nI-3</td>
</tr>
<tr>
<td>gamma1</td>
<td>Body-effect coefficient near interface</td>
<td>calculated</td>
<td>V(^{1/2})</td>
<td>nI-4</td>
</tr>
<tr>
<td>gamma2</td>
<td>Body-effect coefficient in the bulk</td>
<td>calculated</td>
<td>V(^{1/2})</td>
<td>nI-5</td>
</tr>
<tr>
<td>Vbx</td>
<td>Vbs at which the depletion width equals xt</td>
<td>calculated</td>
<td>V</td>
<td>nI-6</td>
</tr>
<tr>
<td>Cgso</td>
<td>Non-LDD source-gate overlap capacitance per channel length</td>
<td>calculated</td>
<td>F/m</td>
<td>nC-1</td>
</tr>
<tr>
<td>Cgdo</td>
<td>Non-Ldd drain-gate overlap capacitance per channel length</td>
<td>calculated</td>
<td>F/m</td>
<td>nC-2</td>
</tr>
<tr>
<td>CF</td>
<td>Fringing field capacitance</td>
<td>calculated</td>
<td>F/m</td>
<td>nC-3</td>
</tr>
</tbody>
</table>

Table 6-2. Parameters with notes for extraction.
6.4.2 Explanation of Notes

**nI-1.** If $V_{th0}$ is not specified, it is calculated by

$$V_{th0} = V_{FB} + \Phi_s + K_1 \sqrt{\Phi_s}$$

where the model parameter $V_{FB}=-1.0$. If $V_{th0}$ is specified, $V_{FB}$ defaults to

$$V_{FB} = V_{th0} - \Phi_s - K_1 \sqrt{\Phi_s}$$

**nI-2.** If $K_1$ and $K_2$ are not given, they are calculated based on

$$K_1 = \gamma_2 - 2K_2 \sqrt{\Phi_s - V_{hm}}$$

$$K_2 = \frac{(\gamma_1 - \gamma_2) (\sqrt{\Phi_s - V_{bs}} - \sqrt{\Phi_s})}{2 \sqrt{\Phi_s (\sqrt{\Phi_s - V_{hm}} - \sqrt{\Phi_s}) + V_{hm}}}$$

where $\Phi_s$ is calculated by

$$\Phi_s = 2V_{m0} \ln \left( \frac{N_{ch}}{n_i} \right)$$

$$V_{m0} = \frac{kT_\text{num}}{q}$$
$n_i = 1.45 \times 10^{10} \left( \frac{T_{\text{nom}}}{300.15} \right)^{1.5} \exp \left( 21.5565981 - \frac{E_{g0}}{2V_{m0}} \right)$

\[
E_{g0} = 1.16 - \frac{7.02 \times 10^{-4} T_{\text{nom}}^2}{T_{\text{nom}} + 1108}
\]

where $E_{g0}$ is the energy bandgap at temperature $T_{\text{nom}}$.

**nI-3.** If $N_{ch}$ is not given and $\gamma_1$ is given, $N_{ch}$ is calculated from

$$\gamma_1 = \frac{\gamma_1}{2q\epsilon_{si}}$$

If both $\gamma_1$ and $N_{ch}$ are not given, $N_{ch}$ defaults to $1.7e23$ m$^{-3}$ and $\gamma_1$ is calculated from $N_{ch}$.

**nI-4.** If $\gamma_1$ is not given, it is calculated by

$$\gamma_1 = \frac{\sqrt{2q\epsilon_{si}N_{ch}}}{C_{ox}}$$

**nI-5.** If $\gamma_2$ is not given, it is calculated by

$$\gamma_2 = \frac{\sqrt{2q\epsilon_{si}N_{sub}}}{C_{ox}}$$

**nI-6.** If $V_{bx}$ is not given, it is calculated by

$$\frac{qN_{ch}X_i^2}{2\epsilon_{si}} = \Phi - V_{bx}$$
Notes on Parameter Extraction

nC-1. If $C_{gso}$ is not given, it is calculated by

\[
\begin{align*}
    \text{if } (d_{lc} \text{ is given and is greater than 0),} \\
    & C_{gso} = d_{lc} \times C_{ox} - C_{gs1} \\
    \text{if } (C_{gso} < 0) \\
    & C_{gso} = 0 \\
    \text{else } C_{gso} = 0.6 \times J \times C_{ox} 
\end{align*}
\]

nC-2. If $C_{gdo}$ is not given, it is calculated by

\[
\begin{align*}
    \text{if } (d_{lc} \text{ is given and is greater than 0),} \\
    & C_{gdo} = d_{lc} \times C_{ox} - C_{gd1} \\
    \text{if } (C_{gdo} < 0) \\
    & C_{gdo} = 0 \\
    \text{else } C_{gdo} = 0.6 \times J \times C_{ox} 
\end{align*}
\]

nC-3. If $CF$ is not given then it is calculated using

\[
CF = \frac{2\varepsilon_{ox}}{\pi} \ln \left( 1 + \frac{4 \times 10^{-7}}{T_{ox}} \right)
\]
CHAPTER 7: Benchmark Test Results

A series of benchmark tests [26] have been performed to check the model robustness, accuracy and performance. Although not all the benchmark test results are included in this chapter, the most important ones are demonstrated.

7.1 Benchmark Test Types

Table 7-1 lists the various benchmark test conditions and associated figure number included in this section. Notice that for each plot, smooth transitions are apparent for current, transconductance, and source to drain resistance for all transition regions regardless of bias conditions.

<table>
<thead>
<tr>
<th>Device Size</th>
<th>Bias Conditions</th>
<th>Notes</th>
<th>Figure Number</th>
</tr>
</thead>
<tbody>
<tr>
<td>W/L=20/5</td>
<td>Ids vs. Vgs @ Vbs=0V; Vds=0.05, 3.3V</td>
<td>Log scale</td>
<td>7-1</td>
</tr>
<tr>
<td>W/L=20/5</td>
<td>Ids vs. Vgs @ Vbs=0V; Vds=0.05, 3.3V</td>
<td>Linear scale</td>
<td>7-2</td>
</tr>
<tr>
<td>W/L=20/0.5</td>
<td>Ids vs. Vgs @ Vbs=0V; Vds=0.05, 3.3V</td>
<td>Log scale</td>
<td>7-3</td>
</tr>
<tr>
<td>W/L=20/0.5</td>
<td>Ids vs. Vgs @ Vbs=0V; Vds=0.05, 3.3V</td>
<td>Linear scale</td>
<td>7-4</td>
</tr>
<tr>
<td>W/L=20/5</td>
<td>Ids vs. Vgs @ Vds=0.05V; Vbs=0 to -3.3V</td>
<td>Log scale</td>
<td>7-5</td>
</tr>
<tr>
<td>W/L=20/5</td>
<td>Ids vs. Vgs @ Vds=0.05V; Vbs=0 to -3.3V; W/L=20/5</td>
<td>Linear scale</td>
<td>7-6</td>
</tr>
<tr>
<td>W/L=20/0.5</td>
<td>Ids vs. Vgs @ Vds=0.05V; Vbs=0 to -3.3V</td>
<td>Log scale</td>
<td>7-7</td>
</tr>
<tr>
<td>W/L=20/0.5</td>
<td>Ids vs. Vgs @ Vds=0.05V; Vbs=0 to -3.3V</td>
<td>Linear scale</td>
<td>7-8</td>
</tr>
<tr>
<td>W/L=20/5</td>
<td>Gm/Ids vs. Vgs @ Vds=0.05V, 3-3V; Vbs=0V</td>
<td>Linear scale</td>
<td>7-9</td>
</tr>
</tbody>
</table>
7.2 Benchmark Test Results

All of the figures listed in Table 7-1 are shown below. Unless otherwise indicated, symbols represent measurement data and lines represent the results of the model. All of these plots serve to demonstrate the robustness and continuous behavior of the unified model expression for not only $I_d$, but $G_m$, $G_m/I_d$, and $R_{out}$ as well.
Figure 7-1. Continuity from subthreshold to strong inversion (log scale).

Figure 7-2. Continuity from subthreshold to strong inversion (linear scale).

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Figure 7-3. Same as Figure 7-1 but for a short channel device.

Figure 7-4. Same as Figure 7-2 but for a short channel device.
Benchmark Test Results

Figure 7-5. Subthreshold to strong inversion continuity as a function of $V_{bs}$.

Figure 7-6. Subthreshold to strong inversion continuity as a function of $V_{bs}$.
Benchmark Test Results

Figure 7-7. Same as Figure 7-5 but for a short channel device.

Figure 7-8. Same as Figure 7-6 but for a short channel device.
Figure 7-9. $G_m/I_d$ continuity from subthreshold to strong inversion regions.

Figure 7-10. Same as Figure 7-9 but for a short channel device.
Benchmark Test Results

Figure 7-11. $G_m/I_d$ continuity as a function of $V_{bs}$.

Figure 7-12. Same as Figure 7-11 but for a short channel device.
Figure 7-13. Comparison of $G_{ds}$ with BSIM3v2.

Figure 7-14. Smooth transitions from linear to saturation regimes.
Benchmark Test Results

Figure 7-15. Same as Figure 7-14 but for a short channel device.

Figure 7-16. Continuous and non-negative $R_{out}$ behavior.
8.1 Flicker Noise

8.1.1 Parameters

There exist two models for flicker noise modeling. One is called SPICE2 flicker noise model: the other is BSIM3 flicker noise model [35-36]. The flicker noise model parameters are listed in Table 8-1.

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Noia</td>
<td>noia</td>
<td>Noise parameter A</td>
<td>(NMOS) 1e20</td>
<td>none</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>(PMOS) 9.9e18</td>
<td></td>
</tr>
<tr>
<td>Noib</td>
<td>noib</td>
<td>Noise parameter B</td>
<td>(NMOS) 5e4</td>
<td>none</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>(PMOS) 2.4e3</td>
<td></td>
</tr>
<tr>
<td>Noic</td>
<td>noic</td>
<td>Noise parameter C</td>
<td>(NMOS) -1.4e-12</td>
<td>none</td>
</tr>
<tr>
<td></td>
<td></td>
<td></td>
<td>(PMOS) 1.4e-12</td>
<td></td>
</tr>
<tr>
<td>Em</td>
<td>em</td>
<td>Saturation field</td>
<td>4.1e7</td>
<td>V/m</td>
</tr>
<tr>
<td>Af</td>
<td>af</td>
<td>Flicker noise exponent</td>
<td>1</td>
<td>none</td>
</tr>
<tr>
<td>Ef</td>
<td>ef</td>
<td>Flicker noise frequency exponent</td>
<td>1</td>
<td>none</td>
</tr>
<tr>
<td>Kf</td>
<td>kf</td>
<td>Flicker noise coefficient</td>
<td>0</td>
<td>none</td>
</tr>
</tbody>
</table>

Table 8-1. Flicker noise model parameters.
8.1.2 Formulations

1. For SPICE2 model

\[
\text{Noise density} = \frac{K_f I_{ds,\text{off}}}{C_{ox} L_{\text{eff}}^2 f_{\text{eff}}}.
\]  

(8.1)

where \( f \) is the frequency.

2. For BSIM3 model

If \( V_{gs} > V_{th} + 0.1 \)

\[
\text{Noise density} = \frac{q V_{th} \mu_{\text{eff}} I_{ds}}{C_{ox} L_n^2 f_{\text{eff}}^* \cdot 10^3} \left( \frac{\text{Noia log } \left( \frac{N_0 + 2 \times 10^4}{N_i + 2 \times 10^4} \right) + \text{Noib } (N_i - N_i) + \frac{\text{Noiq}}{2} (N_i^2 - N_i^2)}{N_i^2 + 2 \times 10^4} \right) + \frac{V_{ds} \Delta L_{\text{in}}}{W_{\text{eff}} L_n^2 f_{\text{eff}}^* \cdot 10^3} \frac{\text{Noia + Noib } N_i + \text{Noiq } N_i^2}{(N_i + 2 \times 10^4)^2}
\]  

(8.2)

where \( V_{th} \) is the thermal voltage, \( \mu_{\text{eff}} \) is the effective mobility at the given bias condition, and \( L_{\text{eff}} \) and \( W_{\text{eff}} \) are the effective channel length and width, respectively. The parameter \( N_0 \) is the charge density at the source side given by

\[
N_0 = \frac{C_{ox} (V_{ex} - V_{th})}{q}
\]  

(8.3)

The parameter \( N_i \) is the charge density at the drain end given by
Flicker Noise

\[ N_f = \frac{C_m(V_c - V_m - \min(V_{ds}, V_{dss}))}{q} \]  

(8.4)

\[ \Delta L_{\text{clm}} \] is the channel length reduction due to channel length modulation and given by

\[ \Delta L_{\text{clm}} = \begin{cases} 
L_{\text{itl}} \cdot \log \left( \frac{\frac{V_{ds} - V_{dss} + E_m}{L_{\text{itl}}}}{E_{\text{at}}} \right) & \text{(for } V_{ds} > V_{dss}) \\
0 & \text{(otherwise)}
\end{cases} \]  

(8.5)

\[ E_{\text{at}} = \frac{2 \times V_{\text{at}}}{\mu_n} \]

where

\[ L_{\text{itl}} = \sqrt{3X_{\text{itl}} T_s} \]

(Otherwise)

\[ \text{Noise density} = \frac{S_{\text{limf}} \times S_{\text{w1}}}{S_{\text{limf}} + S_{\text{w1}}} \]  

(8.6)

Where, \( S_{\text{limf}} \) is the flicker noise calculated at \( V_{gs} = V_m + 0.1 \) and \( S_{w1} \) is given by
8.2 Channel Thermal Noise

There also exist two models for channel thermal noise modeling. One is called SPICE2 thermal noise model. The other is BSIM3v3 thermal noise model. Each of these can be toggled through the model flag, noiMod.

1. For SPICE2 thermal noise model

\[
S_{\text{wi}} = \frac{\text{Noia} \cdot V_m^2 \cdot L_{ds} \cdot 4 \times 10^{-16}}{W_{\text{eff}} \cdot L_{\text{eff}} \cdot f_{\text{out}}} \tag{8.7}
\]

where \( G_m, G_{mb}, G_{ds} \) are the transconductances.

2. For BSIM3v3 thermal noise model [37]

\[
\frac{8k_B T}{3} (G_m + G_{mb} + G_{ds}) \tag{8.8}
\]

where \( G_m, G_{mb}, \) and \( G_{ds} \) are the transconductances.

\[
\frac{4k_B T \mu_{\text{v}}}{L_{\text{eff}}} |Q_{in}| \tag{8.9}
\]

\( Q_{in} \) is the inversion channel charge computed from the capacitance models (capMod = 0, 1, 2 or 3).
8.3 Noise Model Flag

A model flag, noiMod, is used to select different combination of flicker and thermal noise models discussed above with possible options described in Table 8-2.

<table>
<thead>
<tr>
<th>noiMod flag</th>
<th>Flicker noise model</th>
<th>Thermal noise model</th>
</tr>
</thead>
<tbody>
<tr>
<td>1</td>
<td>SPICE2</td>
<td>SPICE2</td>
</tr>
<tr>
<td>2</td>
<td>BSIM3v3</td>
<td>BSIM3v3</td>
</tr>
<tr>
<td>3</td>
<td>BSIM3v3</td>
<td>SPICE2</td>
</tr>
<tr>
<td>4</td>
<td>SPICE2</td>
<td>BSIM3v3</td>
</tr>
</tbody>
</table>

Table 8-2. noiMod flag for different noise models.
Noise Model Flag
9.1 Diode IV Model

The diode IV modeling now supports a resistance-free diode model and a current-limiting feature by introducing a new model parameter \( i_{jth} \) (defaulting to 0.1A). If \( i_{jth} \) is explicitly specified to be zero, a resistance-free diode model will be triggered; otherwise two critical junction voltages \( V_{jsm} \) for S/B diode and \( V_{jdm} \) for D/B diode will be computed from the value of \( i_{jth} \).

9.1.1 Modeling the S/B Diode

If the S/B saturation current \( I_{shs} \) is larger than zero, the following equations is used to calculate the S/B diode current \( I_{hr} \).

**Case 1 - \( i_{jth} \) is equal to zero: A resistance-free diode.**

\[
I_{hr} = I_{shs} \left( \exp \left( \frac{V_{hs}}{N V_{tn}} \right) - 1 \right) + G_{mn} V_{hs}
\]

(9.1)

where \( N V_{tn} = NJ \left( k_B T / q \right) \): \( NJ \) is a model parameter, known as the junction emission coefficient.

**Case 2 - \( i_{jth} \) is non-zero: Current limiting feature.**
If $V_{bs} < V_{jsm}$

$$I_{hs} = I_{sh} \left( \exp \left( \frac{V_{bs}}{N V_t} \right) - 1 \right) + G_{min} V_{hs}$$  \hspace{1cm} (9.2)$$

otherwise

$$I_{hs} = j_{th} + \frac{j_{th} + I_{sh}}{N V_t} (V_{hs} - V_{jsm}) + G_{min} V_{hs}$$  \hspace{1cm} (9.3)$$

with $V_{jsm}$ computed by

$$V_{jsm} = N V_t \ln \left( \frac{j_{th}}{I_{sh}} + 1 \right)$$

The saturation current $I_{sh}$ is given by

$$I_{sh} = A_s J_t + P_s J_{sw}$$  \hspace{1cm} (9.4)$$

where $J_t$ is the junction saturation current density, $A_s$ is the source junction area, $J_{sw}$ is the sidewall junction saturation current density, $P_s$ is the perimeter of the source junction. $J_t$ and $J_{sw}$ are functions of temperature and can be written as

$$J_t = J_{ts0} \exp \left( \frac{E_{g0} - E_s}{V_{ts0} V_t} + X T \ln \left( \frac{T}{T_{ts0}} \right) \right)$$  \hspace{1cm} (9.5)$$

$$J_{sw} = J_{sw0} \exp \left( \frac{E_{g0} - E_s}{V_{sw0} V_t} + X T \ln \left( \frac{T}{T_{sw0}} \right) \right)$$
The energy band-gap $E_{x0}$ and $E_x$ at the nominal and operating temperatures are expressed by (9.7a) and (9.7b), respectively:

\[
E_{x0} = 1.16 - \frac{7.02 \times 10^{-4} T_{num}^2}{T_{num} + 1108}
\] (9.7a)

\[
E_x = 1.16 - \frac{7.02 \times 10^{-4} T}{T + 1108}
\] (9.7b)

In the above derivations, $J_{x0}$ is the saturation current density at $T_{num}$. If $J_{x0}$ is not given, $J_{x0} = 1 \times 10^{-4}$ A/m$^2$. $J_{x0m}$ is the sidewall saturation current density at $T_{num}$, with a default value of zero.

If $I_{oh}$ is not positive, $I_{hr}$ is calculated by

\[
I_{hr} = G_{mun} \cdot V_p
\] (9.8)

9.1.2 Modeling the D/B Diode

If the D/B saturation current $I_{sd}$ is larger than zero, the following equations is used to calculate the D/B diode current $I_{bd}$.

Case 1 - $i_{jth}$ is equal to zero: A resistance-free diode.
Diode IV Model

\[ I_{bd} = I_{sbd} \left( \exp \left( \frac{V_{bd}}{N V_{im}} \right) - 1 \right) + G_{mn} V_{bd} \] (9.9)

Case 2 - \( \textit{ijth} \) is non-zero: Current limiting feature.

If \( V_{bd} < V_{jdm} \)

\[ I_{bd} = I_{sbd} \left( \exp \left( \frac{V_{bd}}{N V_{im}} \right) - 1 \right) + G_{mn} V_{bd} \] (9.10)

otherwise

\[ I_{bd} = \text{ijth} + \frac{I_{sbd}}{N V_{im}} (V_{bd} - V_{jdm}) + G_{mn} V_{bd} \] (9.11)

with \( V_{jdm} \) computed by

\[ V_{jdm} = N V_{im} \ln \left( \frac{\text{ijth}}{I_{sbd}} + 1 \right) \]

The saturation current \( I_{sbd} \) is given by

\[ I_{sbd} = A_d J_s + P_d J_{vnn} \] (9.12)

where \( A_d \) is the drain junction area and \( P_d \) is the perimeter of the drain junction. If \( I_{sbd} \) is not positive, \( I_{bd} \) is calculated by

\[ I_{bd} = G_{mn} V_{bd} \] (9.13)
9.1.3 Model Parameter Lists

The diode DC model parameters are listed in Table 9-1.

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Js0</td>
<td>js</td>
<td>Saturation current density</td>
<td>1e-4</td>
<td>A/m²</td>
</tr>
<tr>
<td>Js0sw</td>
<td>jssw</td>
<td>Side wall saturation current density</td>
<td>0</td>
<td>A/m</td>
</tr>
<tr>
<td>NJ</td>
<td>nj</td>
<td>Emission coefficient</td>
<td>1</td>
<td>none</td>
</tr>
<tr>
<td>XTI</td>
<td>xti</td>
<td>Junction current temperature expo-</td>
<td>3.0</td>
<td>none</td>
</tr>
<tr>
<td></td>
<td></td>
<td>nent coefficient</td>
<td></td>
<td></td>
</tr>
<tr>
<td>ijth¹</td>
<td>ijth</td>
<td>Limiting current</td>
<td>0.1</td>
<td>A</td>
</tr>
</tbody>
</table>

Table 9-1. MOS diode model parameters.

9.2 MOS Diode Capacitance Model

Source and drain junction capacitance can be divided into two components: the junction bottom area capacitance $C_{jh}$ and the junction periphery capacitance $C_{jp}$. The formula for both the capacitances is similar, but with different model parameters. The equation of $C_{jh}$ includes the parameters such as $C_j$, $M_j$, and $P_b$. The equation of $C_{jp}$ includes the parameters such as $C_{jw}$, $M_{jw}$, $P_{bw}$, $C_{fw}$, $M_{fw}$, and $P_{bw}$.

9.2.1 S/B Junction Capacitance

The S/B junction capacitance can be calculated by

$$\text{If } P_s > W_{eff}$$
MOS Diode Capacitance Model

\[
\text{Capbs} = A_s C_{ph} + (P_s - W_{eff}) C_{jbxw} + W_{eff} C_{jbxwg}
\]

(9.14)

Otherwise

\[
\text{Capbs} = A_s C_{ph} + P_s C_{jbxwg}
\]

(9.15)

where \( C_{jbx} \) is the unit bottom area capacitance of the S/B junction, \( C_{jbxw} \) is the periphery capacitance of the S/B junction along the field oxide side, and \( C_{jbxwg} \) is the periphery capacitance of the S/B junction along the gate oxide side.

If \( C_j \) is larger than zero, \( C_{jbx} \) is calculated by

if \( V_{bs} < 0 \)

\[
C_{pbn} = C_j \left(1 - \frac{V_{bs}}{P_b}\right)^M
\]

(9.16)

otherwise

\[
C_{pbn} = C_j \left(1 + M_j \frac{V_{bs}}{P_b}\right)
\]

(9.17)

If \( C_{jbxw} \) is larger than zero, \( C_{jbxw} \) is calculated by

if \( V_{bs} < 0 \)
MOS Diode Capacitance Model

\[ C_{j_{hv}} = C_{jw} \left( 1 - \frac{V_{b_{s}}}{P_{hv}} \right)^{-M_{hv}} \]  
\[ \text{otherwise} \]

\[ C_{j_{hv}} = C_{jw} \left( 1 + M_{jw} \frac{V_{b_{s}}}{P_{hv}} \right) \]

If \( C_{j_{hv}} \) is larger than zero, \( C_{j_{hv}} \) is calculated by

\[ C_{j_{hswg}} = C_{jswg} \left( 1 - \frac{V_{b_{s}}}{P_{hswg}} \right)^{-M_{hswg}} \]  
\[ \text{otherwise} \]

\[ C_{j_{hswg}} = C_{jswg} \left( 1 + M_{jswg} \frac{V_{b_{s}}}{P_{hswg}} \right) \]

9.2.2 D/B Junction Capacitance

The D/B junction capacitance can be calculated by
If \( P_d > W_{eff} \)

\[
Capbd = A_J C_{jph} + (P_d - W_{eff}) C_{jbdw} + W_{eff} C_{jbdw}.
\]  
(9.22)

Otherwise

\[
Capbd = A_J C_{jph} + P_d C_{jbdw}.
\]  
(9.23)

where \( C_{jph} \) is the unit bottom area capacitance of the D/B junction. 
\( C_{jbdw} \) is the periphery capacitance of the D/B junction along the field oxide side, and \( C_{jbdw} \) is the periphery capacitance of the D/B junction along the gate oxide side.

If \( C_j \) is larger than zero, \( C_{jph} \) is calculated by

if \( V_{bd} < 0 \)

\[
C_{jph} = C_j \left(1 - \frac{V_{sd}}{P_h}\right)^\nu.
\]  
(9.24)

otherwise

\[
C_{jph} = C_j \left(1 + M_j \frac{V_{bd}}{P_h}\right).
\]  
(9.25)

If \( C_{jsw} \) is large than zero, \( C_{jbdw} \) is calculated by

if \( V_{bd} < 0 \)
MOS Diode Capacitance Model

(9.26)

\[ C_{ph_{dn}} = C_{jn} \left(1 - \frac{V_{bd}}{P_{ph_{nw}}}ight) \]

otherwise

(9.27)

\[ C_{ph_{dn}} = C_{jn} \left(1 + M_{jn} \frac{V_{bd}}{P_{ph_{nw}}}ight) \]

If \( C_{jsw_{x}} \) is larger than zero, \( C_{ph_{dn_{x}}} \) is calculated by

if \( V_{bd} < 0 \)

(9.28)

\[ C_{ph_{dn_{x}}} = C_{jn_{x}} \left(1 - \frac{V_{bd_{x}}}{P_{ph_{nw_{x}}}}\right)^{M_{jn_{x}}} \]

otherwise

(9.29)

\[ C_{ph_{dn_{x}}} = C_{jn_{x}} \left(1 + M_{jn_{x}} \frac{V_{bd_{x}}}{P_{ph_{nw_{x}}}}\right) \]
9.2.3 Temperature Dependence of Junction Capacitance

The temperature dependence of the junction capacitance is modeled. Both zero-bias unit-area junction capacitance \( C_j \), \( C_{jw} \), and \( C_{jwR} \) and built-in potential of the junction \( P_b \), \( P_{bs} \), and \( P_{bsw} \) are temperature dependent and modeled in the following.

For zero-bias junction capacitance:

\[
\begin{align*}
C_j(T) &= C_j(T_{nom})(1 + tc_j \cdot \Delta T) \\
C_{jw}(T) &= C_{jw}(T_{nom})(1 + tc_{jw} \cdot \Delta T) \\
C_{jwR}(T) &= C_{jwR}(T_{nom})(1 + tc_{jwR} \cdot \Delta T)
\end{align*}
\]

For the built-in potential:

\[
\begin{align*}
P_b(T) &= P_b(T_{nom}) - tpb \cdot \Delta T \\
P_{bs}(T) &= P_{bs}(T_{nom}) - tpbs \cdot \Delta T \\
P_{bsw}(T) &= P_{bsw}(T_{nom}) - tpbsw \cdot \Delta T
\end{align*}
\]

In Eqs. (9.30) and (9.31), the temperature difference is defined as...
MOS Diode Capacitance Model

\[ \Delta T = T - T_{num} \]

The six new model parameters in the above equations are described in Table 9-2.

### 9.2.4 Junction Capacitance Parameters

The following table gives a full description of those model parameters used in the diode junction capacitance modeling.

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>Cj</td>
<td>cj</td>
<td>Bottom junction capacitance per unit area at zero bias</td>
<td>5e-4</td>
<td>F/m²</td>
</tr>
<tr>
<td>Mj</td>
<td>mj</td>
<td>Bottom junction capacitance grading coefficient</td>
<td>0.5</td>
<td>none</td>
</tr>
<tr>
<td>Ph</td>
<td>ph</td>
<td>Bottom junction built-in potential</td>
<td>1.0</td>
<td>V</td>
</tr>
<tr>
<td>Cjsw</td>
<td>cjsw</td>
<td>Source/drain sidewall junction capacitance per unit length at zero bias</td>
<td>5e-10</td>
<td>F/m</td>
</tr>
<tr>
<td>Mjsw</td>
<td>mjsw</td>
<td>Source/drain sidewall junction capacitance grading coefficient</td>
<td>0.33</td>
<td>none</td>
</tr>
<tr>
<td>Pbsw</td>
<td>pbsw</td>
<td>Source/drain side wall junction built-in potential</td>
<td>1.0</td>
<td>V</td>
</tr>
<tr>
<td>Cjswg</td>
<td>cjswg</td>
<td>Source/drain gate side wall junction capacitance per unit length at zero bias</td>
<td>Cjsw</td>
<td>F/m</td>
</tr>
<tr>
<td>Mjswg</td>
<td>mjswg</td>
<td>Source/drain gate side wall junction capacitance grading coefficient</td>
<td>Mjsw</td>
<td>none</td>
</tr>
<tr>
<td>Pbswg</td>
<td>pbswg</td>
<td>Source/drain gate side wall junction built-in potential</td>
<td>Pbsw</td>
<td>V</td>
</tr>
<tr>
<td>tpb</td>
<td>tpb</td>
<td>Temperature coefficient of Pb</td>
<td>0.0</td>
<td>V/K</td>
</tr>
<tr>
<td>tpbsw</td>
<td>tpbsw</td>
<td>Temperature coefficient of Pbsw</td>
<td>0.0</td>
<td>V/K</td>
</tr>
</tbody>
</table>
## MOS Diode Capacitance Model

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
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<th>Default</th>
<th>Unit</th>
</tr>
</thead>
<tbody>
<tr>
<td>tpbswg</td>
<td>tpbswg</td>
<td>Temperature coefficient of Pbswg</td>
<td>0.0</td>
<td>V/K</td>
</tr>
<tr>
<td>tcj</td>
<td>tcj</td>
<td>Temperature coefficient of Cj</td>
<td>0.0</td>
<td>1/K</td>
</tr>
<tr>
<td>tcjsw</td>
<td>tcjsw</td>
<td>Temperature coefficient of Cjsw</td>
<td>0.0</td>
<td>1/K</td>
</tr>
<tr>
<td>tcjswg</td>
<td>tcjswg</td>
<td>Temperature coefficient of Cjswg</td>
<td>0.0</td>
<td>1/K</td>
</tr>
</tbody>
</table>

Table 9-2. MOS Junction Capacitance Model Parameters.
A.1 Model Control Parameters

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
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<th>Default</th>
<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>None</td>
<td>level</td>
<td>The model selector</td>
<td>8</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>None</td>
<td>version</td>
<td>Model version selector</td>
<td>3.2</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>None</td>
<td>binUnit</td>
<td>Bining unit selector</td>
<td>1</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>None</td>
<td>param-Chk</td>
<td>Parameter value check</td>
<td>False</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>mobMod</td>
<td>mobMod</td>
<td>Mobility model selector</td>
<td>1</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>capMod</td>
<td>capMod</td>
<td>Flag for capacitance models</td>
<td>3</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>nqsMod(^a)</td>
<td>nqsMod</td>
<td>Flag for NQS model</td>
<td>0</td>
<td>none</td>
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</tr>
<tr>
<td>noiMod</td>
<td>noiMod</td>
<td>Flag for noise models</td>
<td>1</td>
<td>none</td>
<td></td>
</tr>
</tbody>
</table>

\(^a\) nqsMod is now an element (instance) parameter, no longer a model parameter.

A.2 DC Parameters
## DC Parameters

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Vth0</td>
<td>vth0</td>
<td>Threshold voltage @Vbs=0 for Large L.</td>
<td>0.7 (NMOS) -0.7 (PMOS)</td>
<td>V</td>
<td>nI-1</td>
</tr>
<tr>
<td>VFB</td>
<td>vfb</td>
<td>Flat-band voltage</td>
<td>Calculated</td>
<td>V</td>
<td>nI-1</td>
</tr>
<tr>
<td>K1</td>
<td>k1</td>
<td>First order body effect coefficient</td>
<td>0.5</td>
<td>V$^{1/2}$</td>
<td>nI-2</td>
</tr>
<tr>
<td>K2</td>
<td>k2</td>
<td>Second order body effect coefficient</td>
<td>0.0</td>
<td>none</td>
<td>nI-2</td>
</tr>
<tr>
<td>K3</td>
<td>k3</td>
<td>Narrow width coefficient</td>
<td>80.0</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>K3b</td>
<td>k3b</td>
<td>Body effect coefficient of k3</td>
<td>0.0</td>
<td>I/V</td>
<td></td>
</tr>
<tr>
<td>W0</td>
<td>w0</td>
<td>Narrow width parameter</td>
<td>2.5e-6</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>Nlx</td>
<td>nlx</td>
<td>Lateral non-uniform doping parameter</td>
<td>1.74e-7</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>Vbm</td>
<td>vbm</td>
<td>Maximum applied body bias in Vth calculation</td>
<td>-3.0</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>Dvt0</td>
<td>dvt0</td>
<td>first coefficient of short-channel effect on Vth</td>
<td>2.2</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>Dvt1</td>
<td>dvt1</td>
<td>Second coefficient of short-channel effect on Vth</td>
<td>0.53</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>Dvt2</td>
<td>dvt2</td>
<td>Body-bias coefficient of short-channel effect on Vth</td>
<td>-0.032</td>
<td>I/V</td>
<td></td>
</tr>
<tr>
<td>Dvt0w</td>
<td>dvt0w</td>
<td>First coefficient of narrow width effect on Vth for small channel length</td>
<td>0</td>
<td>I/m</td>
<td></td>
</tr>
<tr>
<td>Dvt1w</td>
<td>dvt1w</td>
<td>Second coefficient of narrow width effect on Vth for small channel length</td>
<td>5.3e6</td>
<td>I/m</td>
<td></td>
</tr>
</tbody>
</table>
### DC Parameters

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
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<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Dvt2w</td>
<td>dvt2w</td>
<td>Body-bias coefficient of narrow width effect for small channel length</td>
<td>-0.032</td>
<td>1/V</td>
<td></td>
</tr>
<tr>
<td>μ0</td>
<td>u0</td>
<td>Mobility at Temp = Tnom NMOSFET PMOSFET</td>
<td>670.0</td>
<td>cm²/Vs</td>
<td></td>
</tr>
<tr>
<td>Ua</td>
<td>ua</td>
<td>First-order mobility degradation coefficient</td>
<td>2.25E-9</td>
<td>m/V</td>
<td></td>
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<tr>
<td>Ub</td>
<td>ub</td>
<td>Second-order mobility degradation coefficient</td>
<td>5.87E-19</td>
<td>(m/V)²</td>
<td></td>
</tr>
<tr>
<td>Uc</td>
<td>uc</td>
<td>Body-effect of mobility degradation coefficient</td>
<td>mobMod =1, 2: -4.65e-11 mobMod =3: -0.046</td>
<td>m/V²</td>
<td></td>
</tr>
<tr>
<td>vsat</td>
<td>vsat</td>
<td>Saturation velocity at Temp = Tnom</td>
<td>8.0E4</td>
<td>m/sec</td>
<td></td>
</tr>
<tr>
<td>A0</td>
<td>a0</td>
<td>Bulk charge effect coefficient for channel length</td>
<td>1.0</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>Ags</td>
<td>ags</td>
<td>gate bias coefficient of Abulk</td>
<td>0.0</td>
<td>1/V</td>
<td></td>
</tr>
<tr>
<td>B0</td>
<td>b0</td>
<td>Bulk charge effect coefficient for channel width</td>
<td>0.0</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>B1</td>
<td>b1</td>
<td>Bulk charge effect width offset</td>
<td>0.0</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>Keta</td>
<td>keta</td>
<td>Body-bias coefficient of bulk charge effect</td>
<td>-0.047</td>
<td>1/V</td>
<td></td>
</tr>
<tr>
<td>A1</td>
<td>a1</td>
<td>First non-saturation effect parameter</td>
<td>0.0</td>
<td>1/V</td>
<td></td>
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<tr>
<td>A2</td>
<td>a2</td>
<td>Second non-saturation factor</td>
<td>1.0</td>
<td>none</td>
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</table>
## DC Parameters

<table>
<thead>
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<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rdsw</td>
<td>rds</td>
<td>Parasitic resistance per unit width</td>
<td>0.0</td>
<td>Ω·μm&lt;sup&gt;wr&lt;/sup&gt;</td>
<td></td>
</tr>
<tr>
<td>Prwb</td>
<td>prwb</td>
<td>Body effect coefficient of Rdsw</td>
<td>0</td>
<td>V&lt;sup&gt;-1/2&lt;/sup&gt;</td>
<td></td>
</tr>
<tr>
<td>Prwg</td>
<td>prwg</td>
<td>Gate bias effect coefficient of Rdsw</td>
<td>0</td>
<td>1/V</td>
<td></td>
</tr>
<tr>
<td>Wr</td>
<td>wr</td>
<td>Width Offset from Weff for Rds calculation</td>
<td>1.0</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>Wint</td>
<td>wint</td>
<td>Width offset fitting parameter from I-V without bias</td>
<td>0.0</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>Lint</td>
<td>lint</td>
<td>Length offset fitting parameter from I-V without bias</td>
<td>0.0</td>
<td>m</td>
<td></td>
</tr>
<tr>
<td>dWg</td>
<td>dwg</td>
<td>Coefficient of Weff's gate dependence</td>
<td>0.0</td>
<td>m/V</td>
<td></td>
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<tr>
<td>dWb</td>
<td>dwb</td>
<td>Coefficient of Weff's substrate body bias dependence</td>
<td>0.0</td>
<td>m/V&lt;sup&gt;1/2&lt;/sup&gt;</td>
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<tr>
<td>Voff</td>
<td>voff</td>
<td>Offset voltage in the subthreshold region at large W and L</td>
<td>-0.08</td>
<td>V</td>
<td></td>
</tr>
<tr>
<td>Nfactor</td>
<td>nfactor</td>
<td>Subthreshold swing factor</td>
<td>1.0</td>
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<tr>
<td>Eta0</td>
<td>eta0</td>
<td>DIBL coefficient in subthreshold region</td>
<td>0.08</td>
<td>none</td>
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<tr>
<td>Etab</td>
<td>etab</td>
<td>Body-bias coefficient for the subthreshold DIBL effect</td>
<td>-0.07</td>
<td>1/V</td>
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<tr>
<td>Dsub</td>
<td>dsub</td>
<td>DIBL coefficient exponent in subthreshold region</td>
<td>drout</td>
<td>none</td>
<td></td>
</tr>
<tr>
<td>Cit</td>
<td>cit</td>
<td>Interface trap capacitance</td>
<td>0.0</td>
<td>F/m&lt;sup&gt;2&lt;/sup&gt;</td>
<td></td>
</tr>
<tr>
<td>Cdsc</td>
<td>cdsc</td>
<td>Drain/Source to channel coupling capacitance</td>
<td>2.4E-4</td>
<td>F/m&lt;sup&gt;2&lt;/sup&gt;</td>
<td></td>
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</tbody>
</table>
### DC Parameters

<table>
<thead>
<tr>
<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>$C_{dscb}$</td>
<td>$cdscb$</td>
<td>Body-bias sensitivity of $C_{dsc}$</td>
<td>0.0</td>
<td>F/Vm$^2$</td>
<td></td>
</tr>
<tr>
<td>$C_{dscd}$</td>
<td>$cdscd$</td>
<td>Drain-bias sensitivity of $C_{dsc}$</td>
<td>0.0</td>
<td>F/Vm$^2$</td>
<td></td>
</tr>
<tr>
<td>$P_{clm}$</td>
<td>$pclm$</td>
<td>Channel length modulation parameter</td>
<td>1.3</td>
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<tr>
<td>$P_{diblc1}$</td>
<td>$pdiblc1$</td>
<td>First output resistance DIBL effect correction parameter</td>
<td>0.39</td>
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</tr>
<tr>
<td>$P_{diblc2}$</td>
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<td>Second output resistance DIBL effect correction parameter</td>
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<tr>
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<td>Body effect coefficient of DIBL correction parameters</td>
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<td>1/V</td>
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<tr>
<td>$D_{rou1}$</td>
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<td>L dependence coefficient of the DIBL correction parameter in $R_{ou}$</td>
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<td>$P_{scbe1}$</td>
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<tr>
<td>$P_{scbe2}$</td>
<td>$pscbe2$</td>
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<td>Gate dependence of Early voltage</td>
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<tr>
<td>$\delta$</td>
<td>$delta$</td>
<td>Effective $V_{ds}$ parameter</td>
<td>0.01</td>
<td>V</td>
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<td>$N_{gate}$</td>
<td>$ngate$</td>
<td>poly gate doping concentration</td>
<td>0</td>
<td>cm$^{-3}$</td>
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<tr>
<td>$x_0$</td>
<td>$alpha0$</td>
<td>The first parameter of impact ionization current</td>
<td>0</td>
<td>m/V</td>
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</tr>
<tr>
<td>$x_1$</td>
<td>$alpha1$</td>
<td>Isub parameter for length scaling</td>
<td>0.0</td>
<td>1/V</td>
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<tr>
<td>$\beta_0$</td>
<td>$beta0$</td>
<td>The second parameter of impact ionization current</td>
<td>30</td>
<td>V</td>
<td></td>
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### C-V Model Parameters

<table>
<thead>
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<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Default</th>
<th>Unit</th>
<th>Note</th>
</tr>
</thead>
<tbody>
<tr>
<td>Rsh</td>
<td>rsh</td>
<td>Source drain sheet resistance in ohm per square</td>
<td>0.0</td>
<td>Ω/square</td>
<td></td>
</tr>
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### A.3 C-V Model Parameters

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### C-V Model Parameters

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### A.4 NQS Parameters

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## Temperature Parameters

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### A.6 Temperature Parameters

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## Temperature Parameters

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### Flicker Noise Model Parameters

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### A.7 Flicker Noise Model Parameters

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### A.8 Process Parameters

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### A.9 Geometry Range Parameters

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<td>none</td>
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</tbody>
</table>

### A.10 Model Parameter Notes

**nI-1.** If $V_{th0}$ is not specified, it is calculated by

$$V_{th0} = V_{FB} + \Phi_s + K_1 \sqrt{\Phi_s}$$

where the model parameter $V_{FB} = -1.0$. If $V_{th0}$ is specified, $V_{FB}$ defaults to

$$V_{FB} = V_{th0} - \Phi_s - K_1 \sqrt{\Phi_s}$$

**nI-2.** If $K_1$ and $K_2$ are not given, they are calculated based on

$$K_1 = \gamma_2 - 2K_2 \sqrt{\Phi_s - V_{hm}}$$
\[ K_2 = \frac{(\gamma_1 - \gamma_2)(\sqrt{\Phi_s - V_{bs}} - \sqrt{\Phi_s})}{2\sqrt{\Phi_s}(\sqrt{\Phi_s - V_{hm}} - \sqrt{\Phi_s}) + V_{hm}} \]

where \( \Phi_s \) is calculated by

\[ \Phi_s = 2V_{m0}\ln\left(\frac{N_{ch}}{n_s}\right) \]

\[ V_{m0} = \frac{k_BT_{nom}}{q} \]

\[ n_s = 1.45 \times 10^{10} \left( \frac{T_{nom}}{300.15} \right)^{1.5} \exp\left(21.5565981 - \frac{E_{g0}}{2V_{m0}}\right) \]

\[ E_{g0} = 1.16 - \frac{7.02 \times 10^{-4}T_{nom}^2}{T_{nom} + 1108} \]

where \( E_{g0} \) is the energy bandgap at temperature \( T_{nom} \).

If \( pschel \leq 0.0 \), a warning message will be given.
If \( ijth < 0.0 \), a fatal error message will occur.
If \( T_{oxm} \leq 0.0 \), a fatal error message will occur.
Model Parameter Notes

\textbf{nI-4.} If $N_{ch}$ is not given and $\gamma_1$ is given, $N_{ch}$ is calculated from

$$N_{ch} = \frac{\gamma_1^2 C_{ox}^2}{2q\varepsilon_{si}}$$

If both $\gamma_1$ and $N_{ch}$ are not given, $N_{ch}$ defaults to $1.7e23$ m$^{-3}$ and $\gamma_1$ is calculated from $N_{ch}$.

\textbf{nI-5.} If $\gamma_1$ is not given, it is calculated by

$$\gamma_1 = \frac{\sqrt{2q\varepsilon_{si}N_{ch}}}{C_{ox}}$$

\textbf{nI-6.} If $\gamma_2$ is not given, it is calculated by

$$\gamma_2 = \frac{\sqrt{2q\varepsilon_{si}N_{sub}}}{C_{ox}}$$

\textbf{nI-7.} If $V_{hs}$ is not given, it is calculated by

$$\frac{qN_{ch}x^2}{2\varepsilon_{si}} = \Phi_3 - V_{hs}$$

\textbf{nC-1.} If $C_{gso}$ is not given, it is calculated by

if ($dlc$ is given and is greater than 0),

$C_{gso} = dlc \times C_{ox} - C_{gs1}$

if ($C_{gso} < 0$)

$C_{gso} = 0$

else $C_{gso} = 0.6 \times xj \times C_{ox}$

\textbf{nC-2.} If $C_{gdo}$ is not given, it is calculated by

if ($dlc$ is given and is greater than 0),
Model Parameter Notes

\[ C_{gdo} = dlc \times C_{ox} - C_{gd} \]

if \( C_{gdo} < 0 \)
\[ C_{gdo} = 0 \]
else \( C_{gdo} = 0.6 \times X_j \times C_{ox} \)

nC-3. If \( CF \) is not given then it is calculated using

\[ CF = \frac{2\varepsilon_{ox}}{\pi} \ln \left( 1 + \frac{4 \times 10^{-7}}{T_{ox}} \right) \]

nC-4.

If \( (acde < 0.4) \) or \( (acde > 1.6) \), a warning message will be given.
If \( (moin < 5.0) \) or \( (moin > 25.0) \), a warning message will be given.
If \( (noff < 0.1) \) or \( (noff > 4.0) \), a warning message will be given.
If \( (voffcv < -0.5) \) or \( (voffcv > 0.5) \), a warning message will be given.
APPENDIX B: Equation List

B.1 I-V Model

B.1.1 Threshold Voltage

\[
V_{t} = V_{th0} + K_{1n} \cdot \sqrt{\Phi} - V_{thn} - K_{2n} \cdot V_{thn}
\]

\[
+ K_{1n} \left( \sqrt{1 + \frac{Nl_n}{L_{m}}} - 1 \right) \sqrt{\Phi} + (K_{1n} + K_{2n} \cdot V_{thn}) \frac{T_{on}}{W_{on} + W_{p}} \Phi.
\]

\[
- D_{\gamma n} \cdot \exp \left( - D_{\gamma n} \frac{L_{m}}{2l_n} \right) + 2 \exp \left( - D_{\gamma n} \frac{L_{m}}{l_n} \right) \left( V_{thn} - \Phi \right)
\]

\[
- D_{\gamma n} \cdot \exp \left( - D_{\gamma n} \frac{L_{m}}{2l_n} \right) + 2 \exp \left( - D_{\gamma n} \frac{L_{m}}{l_n} \right) \left( V_{thn} - \Phi \right)
\]

\[
- \left( \exp \left( - D_{\gamma n} \frac{L_{m}}{2l_n} \right) + 2 \exp \left( - D_{\gamma n} \frac{L_{m}}{l_n} \right) \right) \left( E_{on} + E_{off} \cdot V_{thn} \right)
\]

\[
V_{th0n} = V_{th0} - K_{1n} \cdot \sqrt{\Phi},
\]

\[
K_{1n} = K_{1} \cdot \frac{T_{on}}{T_{onm}}
\]

\[
K_{2n} = K_{2} \cdot \frac{T_{on}}{T_{onm}}
\]
I-V Model

\[ l_t = \sqrt{\frac{\varepsilon_x X_{dep}}{C_{ox}}} (1 + D_{VTZ} V_{b eff}) \]

\[ l_{bc} = \sqrt{\frac{\varepsilon_x X_{dep}}{C_{ox}}} (1 + D_{VTZ} V_{b eff}) \]

\[ l_{iv} = \sqrt{\frac{\varepsilon_x X_{dep0}}{C_{ox}}} \]

\[ X_{dep} = \sqrt{\frac{2 \varepsilon_x (\Phi - V_{b eff})}{q N_c}} \]

\[ X_{dep0} = \sqrt{\frac{2 \varepsilon_x \Phi}{q N_c}} \]

\( (\delta t = 0.001) \)

\[ V_{b eff} = V_c + 0.5 [V_{br} - V_c - \delta t + \sqrt{(V_{br} - V_c - \delta t)^2 - 4 \delta t V_c}] \]

\[ V_c = \frac{0.9 (\Phi - K_i)}{4 K_i} \]

\[ V_{th} = V_t \ln \left( \frac{N_{c, NDS}}{n_i^2} \right) \]

**B.1.2 Effective \((V_{gs}, V_{th})\)**

\[ V_{C eff} = \frac{2 n v_t \ln \left[ 1 + \exp \left( \frac{V_c - V_{th}}{2 n v_t} \right) \right]}{1 + 2 n C_{ox} \sqrt{\frac{2 \Phi}{q \varepsilon_x N_c}} \exp \left( - \frac{V_c - V_{th} - 2 V_{qg}}{2 n v_t} \right)} \]
I-V Model

\[ n = 1 + \frac{N_{\text{as}}}{C_i} \left( \frac{C_i + C_{\text{eff}}} {C_{\text{as}}} \right) \left( \exp\left( -L_i \frac{L_i}{2l} \right) + 2 \exp\left( -L_i \frac{L_i}{l} \right) \right) + \frac{C_i}{C_{\text{as}}} \]

\[ C_i = \frac{E_{\text{i}}}{X_{\text{dep}}} \]

B.1.3 Mobility

For \text{mobMod}=1

\[ \mu_{\text{on}} = \frac{\mu_0}{1 + (U_a + U_e V_{\text{eff}} + 2 V_u)\left(\frac{V_{\text{eff}}}{T_{\text{ox}}} + U_b\left(\frac{V_{\text{eff}}^2 + 2 V_u}{T_{\text{ox}}} \right) \right)} \]

For \text{mobMod}=2

\[ \mu_{\text{on}} = \frac{\mu_0}{1 + (U_a - U_e V_{\text{eff}})\left(\frac{V_{\text{eff}}}{T_{\text{ox}}} \right) + U_b\left(\frac{V_{\text{eff}}^2 + 2 V_u}{T_{\text{ox}}} \right) \right)} \]

For \text{mobMod}=3

\[ \mu_{\text{on}} = \frac{\mu_0}{1 + [U_a\left(\frac{V_{\text{eff}} + 2 V_u}{T_{\text{ox}}} \right) + U_b\left(\frac{V_{\text{eff}}^2 + 2 V_u}{T_{\text{ox}}} \right) \right]^2 (1 + U_e V_{\text{eff}} \right)} \]
B.1.4 Drain Saturation Voltage

For $R_{ds} > 0$ or $\lambda \neq 1$:

$$V_{dsat} = \frac{-b - \sqrt{b^2 - 4ac}}{2a}$$

$$a = A_{bulk}^2 W_{eff} V_{sat} C_{ox} R_{DS} + \left(\frac{1}{\lambda} - 1\right) A_{bulk}$$

$$b = -\left(V_{gsteff} + 2V_t\right)\left(\frac{2}{\lambda} - 1\right) + A_{bulk} E_{sat} L_{eff} + 3 A_{bulk} \left(2V_{gsteff} + 2V_t\right) W_{eff} V_{sat} C_{ox} R_{DS}$$

$$c = (V_{gsteff} + 2V_t) E_{sat} L_{eff} + 2(V_{gsteff} + 2V_t)^2 W_{eff} V_{sat} C_{ox} R_{DS}$$

$$\lambda = A_1 V_{gsteff} + A_2$$

For $R_{ds} = 0$ and $\lambda = 1$:

$$V_{dsat} = \frac{E_{sat} L_{eff} (V_{gsteff} + 2V_t)}{A_{bulk} E_{sat} L_{eff} + (V_{gsteff} + 2V_t)}$$

$$A_{bulk} = \left(1 + \frac{K_{Bulk}}{2\Phi_i - V_{th,n}}\right) \left(\frac{A_{L_{eff}}}{L_{n} + 2\sqrt{X_{n}} X_{n,n}}\right) \left(1 - A_{V_{th,n}} \left(\frac{I_{sat}}{L_{n} + 2\sqrt{X_{n}} X_{n,n}}\right) + \frac{B_{n}}{W_{n} + B_{n}}\right) \left(1 + K_{eta} V_{th,n}\right)$$
I-V Model

\[ E_{sat} = \frac{2V_{sat}}{\mu_{eff}} \]

B.1.5 Effective \( V_{ds} \)

\[ V_{dseff} = V_{dsat} - \frac{1}{2} \left( V_{dsat} - V_{ds} - \delta + \sqrt{(V_{dsat} - V_{ds} - \delta)^2 + 4\delta V_{dsat}} \right) \]

B.1.6 Drain Current Expression

\[ I_d = \frac{I_{d,vdseff}}{1 + \frac{R_{ds,Vdseff}}{V_{dseff}}} \left( 1 + \frac{V_{ds} - V_{dseff}}{V_A} \right) \left( 1 + \frac{V_{ds} - V_{dseff}}{V_{ASCBE}} \right) \]

\[ I_{ds} = \frac{W_{eff} \mu_{eff} C_{ox} V_{gseff} (1 - A_{bulk} \frac{V_{dseff}}{2(V_{gseff} + 2V_t)}) V_{dseff}}{L_{eff} [1 + V_{dseff} / (E_{sat} L_{eff})]} \]

\[ V_A = V_{Asat} + (1 + \frac{P_{vac} V_{vseff}}{E_{sat} L_{eff}})(\frac{1}{V_{ACL}} + \frac{1}{V_{ADIBC}})^{-1} \]

\[ V_{ACL} = \frac{A_{bulk} E_{sat} L_{eff} + V_{gseff}}{P_{CLM} A_{bulk} E_{sat} l_{it}} (V_{ds} - V_{dseff}) \]
I-V Model

\[ V_{ADILC} = \frac{(V_{gate} + 2V_f)}{\theta_{out}(1 + P_{DIQLCB}V_{hseff})} \left( 1 - \frac{A_{bulk}V_{dsat}}{A_{bulk}V_{dsat} + V_{gate} + 2V_f} \right) \]

\[ \theta_{out} = P_{DIQLC1}\left[ \exp(-D_{ROUT} \frac{L_{eff}}{2l_0}) + 2\exp(-D_{ROUT} \frac{L_{eff}}{l_0}) \right] + P_{DIQLC2} \]

\[ \frac{1}{V_{ASCBE}} = \frac{P_{sch2}}{L_{eff}} \exp \left( \frac{-P_{sch1} l_{itl}}{V_{ds} - V_{dsat}} \right) \]

\[ E_{sat}L_{eff} + V_{dsat} + 2R_{DS}V_{sat}C_{ox}V_{dsat}V_{gate} + \frac{A_{bulk}V_{dsat}}{2(V_{gate} + 2V_f)} \]

\[ V_{Asat} = \frac{1}{2/\lambda - 1 + R_{DS}V_{sat}C_{ox}V_{ds}A_{bulk}} \]

\[ l_{itl} = \frac{\varepsilon_s T_{ox} X_j}{\varepsilon_{ox}} \]

B.1.7 Substrate Current

\[ I_{sub} = \alpha_0 + \alpha_1 \cdot \frac{L_{eff}}{L_{eff}} (V_{ds} - V_{dsat}) \exp \left( -\frac{\beta_0}{V_{ds} - V_{dsat}} \right) \frac{I_{d,0}}{R_{ds}I_{d,0}} \left( 1 + \frac{V_{ds} - V_{dss}}{V_A} \right) \]
I-V Model

B.1.8 Polysilicon Depletion Effect

\[ V_{poly} = \frac{1}{2} X_{poly} E_{poly} = \frac{qN_{gate} X_{poly}^2}{2\varepsilon_{si}} \]

\[ \varepsilon_{ox} E_{ox} = \varepsilon_{si} E_{poly} = \sqrt{2q\varepsilon_{si} N_{gate} V_{poly}} \]

\[ V_v - V_{FB} - \Phi_s = V_{poly} + V_m \]

\[ a(V_v - V_{FB} - \Phi_s - V_{poly})^2 - V_{poly} = 0 \]

\[ a = \frac{\varepsilon_{ox}^2}{2q\varepsilon_{si} N_{gate} T_{ox}^2} \]

\[ V_{v_{eff}} = V_{FB} + \Phi_s + \frac{q\varepsilon_{ox} N_{gate} T_{ox}^2}{\varepsilon_{ox}^2} \left( \sqrt{1 + \frac{2\varepsilon_{ox}^2 (V_v - V_{FB} - \Phi_s)}{q\varepsilon_{si} N_{gate} T_{ox}^2}} - 1 \right) \]

B.1.9 Effective Channel Length and Width

\[ L_{eff} = L_{drawn} - 2dL \]

\[ W_{eff} = W_{drawn} - 2dW \]
\[ W_{\text{eff}} = W_{\text{drawn}} - 2dW \]

\[ dW = dW^* + dW_{V_{\text{eff}}} + dW_{\Phi} \left( \sqrt{\Phi_s} - V_{h_{\text{eff}}} - \sqrt{\Phi_s} \right) \]

\[ dW^* = W_{\text{int}} + \frac{W_j}{W^L_{\text{Int}}} + \frac{W_w}{W^L_{\text{Wn}}} + \frac{W_{\text{nl}}}{W^L_{\text{LWn}}} \]

\[ dL = L_{\text{int}} + \frac{L_j}{L^L_{\text{Int}}} + \frac{L_w}{L^L_{\text{LWn}}} + \frac{L_{\text{nl}}}{L^L_{\text{LWn}}} \]

**B.1.10 Source/Drain Resistance**

\[ R_{\text{id}} = R_{\text{id,ln}} \left( 1 + P_{\text{nr,1}} V_{\text{steff}} + P_{\text{nr,2}} \left( \sqrt{\Phi_s} - V_{h_{\text{steff}}} - \sqrt{\Phi_s} \right) \right) \left( 10^6 W_{\text{eff}} \right)^{\text{inv}} \]

**B.1.11 Temperature Effects**

\[ V_{\text{th1,ln}} = V_{\text{th1,Tn}} + (K_{T1} + K_{T2} / L_{\text{eff}} + K_{T2} V_{\text{th1}})(T / T_{\text{name}} - 1) \]

\[ \mu_{\text{ln}}(T) = \mu_{\text{ln,Tn}} \left( \frac{T}{T_{\text{name}}} \right)^{\mu_h} \]

\[ V_{\text{th1}}(T) = V_{\text{th1,Tn}} - A_T(T / T_{\text{name}} - 1) \]
Capacitance Model Equations

\[ R_{d,v,T} = R_{d,v,T_{\text{norm}}} + P_{T_{\text{eff}}} \left( \frac{T}{T_{\text{norm}}} - 1 \right) \]

\[ U_{d,v,T} = U_{d,v,T_{\text{norm}}} + U_{d,v}(T/T_{\text{norm}} - 1) \]

\[ U_{p,T} = U_{p,T_{\text{norm}}} + U_{p}(T/T_{\text{norm}} - 1) \]

\[ U_{n,T} = U_{n,T_{\text{norm}}} + U_{n}(T/T_{\text{norm}} - 1) \]

B.2 Capacitance Model Equations

B.2.1 Dimension Dependence

\[ L_{\text{active}} = L_{\text{drawn}} - 2\delta L_{\text{eff}} \]

\[ W_{\text{active}} = W_{\text{drawn}} - 2\delta W_{\text{eff}} \]

\[ \delta L_{\text{eff}} = DLC + \frac{L_{c}}{L_{\text{ln}}} + \frac{L_{c}}{W_{\text{ln}}} + \frac{L_{wlc}}{L_{\text{ln}}W_{\text{ln}}} \]

\[ \delta W_{\text{eff}} = DWC + \frac{W_{c}}{W_{\text{ln}}} + \frac{W_{c}}{W_{\text{ln}}} + \frac{W_{wlc}}{W_{\text{ln}}W_{\text{ln}}} \]
Capacitance Model Equations

B.2.2 Overlap Capacitance

B.2.2.1 Source Overlap Capacitance

(1) for capMod = 0

\[ \frac{Q_{\text{overlap, s}}}{W_{\text{active}}} = \frac{CGS0V_{gs}}{W_{\text{active}}} \]

(2) for capMod = 1

If \( V_{gs} < 0 \)

\[ \frac{Q_{\text{overlap, s}}}{W_{\text{active}}} = \frac{CGS0 \cdot V_{csp} + \frac{CKAPPA \cdot CGS1}{2} \left( -1 + \sqrt{1 - \frac{4V_{csp}}{CKAPPA}} \right)}{W_{\text{active}}} \]

Else

\[ \frac{Q_{\text{overlap, s}}}{W_{\text{active}}} = \left( CGS0 + CKAPPA \cdot CGS1 \right) \cdot V_{gs} \]

(3) for capMod = 2

\[ \frac{Q_{\text{overlap}}}{W_{\text{active}}} = \frac{CGS0 \cdot V_{csp} + CGS1 \left( V_{csp} - V_{\text{overlap}} \right) - \frac{CKAPPA}{2} \left( -1 + \sqrt{1 - \frac{4V_{\text{overlap}}}{CKAPPA}} \right)}{W_{\text{active}}} \]

\[ V_{csp, overlap} = \frac{1}{2} \left( V_{csp} + \delta - \sqrt{(V_{csp} + \delta)^2 + 4\delta} \right) \quad \delta = 0.02 \]
Capacitance Model Equations

B.2.2.2 Drain Overlap Capacitance

(1) for capMod = 0

\[ \frac{Q_{\text{overlap,d}}}{W_{\text{act.inv}}} = \text{CGDO} V_{gd} \]

(2) for capMod = 1

If \( V_{cd} < 0 \)

\[ \frac{Q_{\text{overlap,d}}}{W_{\text{act.inv}}} = \text{CGDO} \cdot V_{cd} + \frac{\text{CKAPPA} \cdot \text{CGD1}}{2} \left( -1 + \sqrt{1 - \frac{4V_{cd}}{\text{CKAPPA}}} \right) \]

Else

\[ \frac{Q_{\text{overlap,d}}}{W_{\text{act.inv}}} = (\text{CGDO} + \text{CKAPPA} \cdot \text{CGD1}) \cdot V_{gd} \]

(3) for capMod = 2

\[ \frac{Q_{\text{overlap,d}}}{W_{\text{act.inv}}} = \text{CGDO} \cdot V_{cd} + \text{CGD} \left( V_{cd} - V_{\text{gd,overlap}} \right) - \frac{\text{CKAPPA}}{2} \left( -1 + \sqrt{1 - \frac{4V_{\text{gd,overlap}}}{\text{CKAPPA}}} \right) \]

\[ V_{\text{gd,overlap}} = \frac{1}{2} \left( V_{gd} + \delta_1 - \sqrt{(V_{gd} + \delta_1)^2 + 4\delta_1} \right) \quad \delta_1 = 0.02 \]
B.2.2.3 Gate Overlap Charge

\[ Q_{\text{overlap,g}} = -\left(Q_{\text{overlap,s}} + Q_{\text{overlap,d}} \right) \]

B.2.3 Intrinsic Charges

(1) \( \text{capMod} = 0 \)

a. Accumulation region \( (V_{\text{gs}} < V_{\text{th1}} + V_{\text{hs}}) \)

\[ Q_v = W_{\text{act,iv}} L_{\text{act,i}} C_{\text{v}} \left( V_{\text{gs}} - V_{\text{hs}} - V_{\text{th1}} \right) \]

\[ Q_{\text{en}} = -Q_v \]

\[ Q_{\text{in}} = 0 \]

b. Subthreshold region \( (V_{\text{gs}} < V_{\text{th}}) \)

\[ Q_{\text{en0}} = -W_{\text{act,iv}} L_{\text{act,i}} C_{\text{v}} \frac{K_{\text{in}}}{2} \left( -1 + \sqrt{1 + \frac{4(V_{\text{gs}} - V_{\text{th1}} - V_{\text{hs}})}{K_{\text{in}}^2}} \right) \]

\[ Q_r = -Q_h \]
Capacitance Model Equations

\[ Q_{nc} = 0 \]

c. Strong inversion \((V_{gs} > V_{th})\)

\[ V_{drw} = \frac{V_{gs} - V_{th}}{A_{bulk}} \]

\[ A_{nu} = A_{nu}\left[1 + \left(\frac{C_L}{L_{eff}}\right)^{CLE}\right] \]

\[ A_{nu(0)} = \left[1 + \frac{K_{d1}}{2\sqrt{\Phi_+ - V_{brev}}} \left(\frac{A_0 L_{eff}}{L_{eff} + 2\sqrt{X_J X_{ds}} + B_0}{B_i} \right)\right] \frac{1}{1 + Keta V_{brev}} \]

\[ V_{th} = V_{FB} + \Phi_s + K_{1m} \sqrt{\Phi_+ - V_{brev}} \]

(i) 50/50 Charge partition

If \( V_{ds} < V_{d_{sat}} \)

\[ Q_{c} = C_{x} W_{dx} L_{dx} \cdot \left(V_{gs} - V_{ps} - \Phi_s - \frac{V_{ds}}{2} + \frac{A_{bulk} V_{ds}^2}{2 \left(V_{gs} - V_{th} - \frac{A_{bulk} V_{ds}}{2}\right)}\right) \]
Capacitance Model Equations

\[ Q_{inv} = -W_{active} L_{active} C_{ox}[V_{gs} - V_{th} - \frac{A_{bulk} \cdot V_{ds}}{2} + \frac{A_{bulk} \cdot V_{ds}^2}{2(V_{gs} - V_{th} - \frac{A_{bulk} \cdot V_{ds}}{2})}] \]

\[ Q_{h} = W_{active} L_{active} C_{ox}[V_{fb} - V_{th} + \Phi_{x} + \frac{(1 - A_{bulk} \cdot V_{ds})}{2} - \frac{(1 - A_{bulk} \cdot A_{bulk} \cdot V_{ds}^2)}{2(V_{gs} - V_{th} - \frac{A_{bulk} \cdot V_{ds}}{2})}] \]

\[ Q_{s} = Q_{d} = 0.5Q_{inv} = -W_{active} L_{active} C_{ox}[V_{gs} - V_{th} - \frac{A_{bulk} \cdot V_{ds}}{2} + \frac{A_{bulk} \cdot V_{ds}^2}{2(V_{gs} - V_{th} - \frac{A_{bulk} \cdot V_{ds}}{2})}] \]

otherwise

\[ Q_{v} = W_{active} L_{active} C_{ox}(V_{gs} - V_{th} - \Phi_{x} - \frac{V_{dsat}}{3}) \]

\[ Q_{i} = Q_{d} = -\frac{1}{3} W_{active} L_{active} C_{ox}(V_{gs} - V_{th}) \]

\[ Q_{b} = -W_{active} L_{active} C_{ox}(V_{fb} + \Phi_{x} - V_{th} + \frac{(1 - A_{bulk} \cdot V_{dsat}}{3}) \]

(ii) 40/60 channel-charge Partition
Capacitance Model Equations

\[
Q_x = C_{ox} W_{active} L_{active} \left[ V_{gs} - V_{th} - \Phi_s - \frac{V_{ds}}{2} + \frac{A_{bulk} V_{ds}^2}{12(V_{gs} - V_{th} - \frac{A_{bulk} V_{ds}}{2})} \right]
\]

\[
Q_{c,x} = -W_{active} L_{active} C_{ox} \left[ V_{gs} - V_{th} - \frac{A_{bulk} V_{ds}}{2} + \frac{A_{bulk}^2 V_{ds}^2}{12(V_{gs} - V_{th} - \frac{A_{bulk} V_{ds}}{2})} \right]
\]

\[
Q_{h} = W_{active} L_{active} C_{ox} \left[ V_{th} - V_{th} + \Phi_s + \frac{(1 - A_{bulk}) V_{ds}}{2} - \frac{(1 - A_{bulk}) A_{bulk} V_{ds}^2}{12(V_{gs} - V_{th} - \frac{A_{bulk} V_{ds}}{2})} \right]
\]

\[
Q_{d} = -W_{active} L_{active} C_{ox} \left[ \frac{V_{c,x} - V_{th}}{2} - \frac{A_{bulk} V_{ds}}{2} \right]
\]

\[
\frac{A_{bulk} V_{ds} \left( V_{c,x} - V_{th} \right)^2}{6} - \frac{A_{bulk} V_{ds} (V_{c,x} - V_{th})}{8} + \frac{(A_{bulk} V_{ds})^2}{40}
\]

\[
Q_s = -(Q_c + Q_h + Q_d)
\]

otherwise

\[
Q_s = W_{active} L_{active} C_{ox} \left( V_{gs} - V_{th} - \Phi_s - \frac{V_{dsut}}{3} \right)
\]
Capacitance Model Equations

\[ Q_d = -\frac{4}{15} W_{active} L_{active} C_{ox} (V_{gs} - V_{th}) \]

\[ Q_c = -(Q_c + Q_b + Q_d) \]

\[ Q_b = -W_{active} L_{active} C_{ox} (V_{fb} + \Phi - V_{th} + \frac{(1 - A_{bulk\,\prime})V_{ds}}{3}) \]

(iii) 0/100 Channel-charge Partition

if \( V_{ds} < V_{dsat} \)

\[ Q_v = C_{ox} W_{active} L_{active} \left( V_{gs} - V_{th} -\Phi - \frac{V_{ds}}{2} + \frac{A_{bulk\,\prime} V_{ds}^2}{12(V_{gs} - V_{th} - A_{bulk\,\prime} V_{ds})} \right) \]

\[ Q_{mv} = -W_{active} L_{active} C_{ox} \left( V_{gs} - V_{th} - \frac{A_{bulk\,\prime} V_{ds}}{2} + \frac{A_{bulk\,\prime}^2 V_{ds}^2}{12(V_{gs} - V_{th} - A_{bulk\,\prime} V_{ds})} \right) \]

\[ Q_h = W_{active} L_{active} C_{ox} \left( V_{fb} - V_{th} + \Phi + \frac{(1 - A_{bulk\,\prime})V_{ds}}{2} - \frac{(1 - A_{bulk\,\prime}) A_{bulk\,\prime} V_{ds}^2}{12(V_{gs} - V_{th} - A_{bulk\,\prime} V_{ds})} \right) \]
Capacitance Model Equations

\[ Q_d = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left[ \frac{V_{gs} - V_{th}}{2} + \frac{A_{\text{bulk}}'}{4} V_{ds} - \frac{(A_{\text{bulk}}')^2 V_{ds}}{24(V_{gs} - V_{th} - \frac{A_{\text{bulk}}'}{2} V_{ds})} \right] \]

\[ Q = -(Q_s + Q_h + Q_v) \]

otherwise

\[ Q_v = W_{\text{active}} L_{\text{active}} C_{\text{ox}} (V_{gs} - V_{th} - \Phi_s - \frac{V_{dsat}}{3}) \]

\[ Q_h = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} (V_{fh} + \Phi_s - V_{th} + \frac{(1 - A_{\text{bulk}}') V_{dsat}}{3}) \]

\[ Q_d = 0 \]

\[ Q_s = -(Q_v + Q_h) \]

(2) capMod = 1

For capMod = 1, if version < 3.2, the \( V_{fh} \) is calculated from

\[ V_{fh} = V_{th} - \Phi_s - K_{\text{int}} \sqrt{\Phi_s - V_{\text{eff}}^2} \]
Capacitance Model Equations

otherwise $V_{fb}$ is calculated from the above equation but under bias independent $V_{th}$.

if ($V_{gs} < V_{fb} + V_{bs} + V_{gs eff}$)

$$Q_{cl} = -W_{active} L_{active} C_{ox} \left( V_{gs} - V_{fb} - V_{bs} - V_{gs eff} \right)$$

else

$$Q_{cl} = W_{active} L_{active} C_{ox} \cdot K_{m} \left( -1 + \frac{4(V_{gs} - V_{fb} - V_{gs eff})}{K_{m}} \right)$$

$$Q_{bh} = -Q_{cl}$$

$$V_{dsat,AV} = \frac{V_{gs eff}}{A_{bulk}}$$

$$A_{bulk} = A_{bulk0} \left( 1 + \left( \frac{C_{L} C}{L_{eff}} \right)^{CLE} \right)$$

$$A_{bulk} = \left( 1 + \frac{K_{m}}{2 \sqrt{\Phi - V_{n eff}}} \left( \frac{A_{h} L_{eff}}{L_{eff} + 2 \sqrt{X_{j} X_{dep}}} + \frac{B_{h}}{W_{eff} + B_{i}} \right) \right) \frac{1}{1 + K_{eta} V_{n eff}}$$
Capacitance Model Equations

\[ V_{	ext{cor}, i} = n_{i} \cdot V \ln \left( 1 + \exp \left( \frac{V_{\text{voltage}} - V_{d} - v_{\text{offset}}}{n_{i}} \right) \right) \]

if \( V_{d} \leq V_{d\text{off}} \)

\[ Q_{c} = Q_{c,i} + W_{c} L_{c} C_{c} \left( V_{\text{gs eff}} - \frac{V_{d}}{2} + \frac{A_{\text{bulk}} \cdot V_{d}^{2}}{12 \left( V_{\text{gs eff}} - \frac{A_{\text{bulk}}}{2} \right)} \right) \]

\[ Q_{p} = Q_{p,i} + W_{p} L_{p} C_{p} \left( \frac{1 - A_{\text{bulk}}}{2} V_{d} - \frac{\left( 1 - A_{\text{bulk}} \right) A_{\text{bulk}} \cdot V_{d}^{2}}{12 \left( V_{\text{gs eff}} - \frac{A_{\text{bulk}}}{2} \right)} \right) \]

(i) 50/50 Channel-charge Partition

\[ Q = Q_{c} - \frac{W_{c} L_{c} C_{c}}{2} \left( V_{\text{gs eff}} - \frac{A_{\text{bulk}}}{2} V_{d} + \frac{A_{\text{bulk}}^{2} \cdot V_{d}^{2}}{12 \left( V_{\text{gs eff}} - \frac{A_{\text{bulk}}}{2} \right)} \right) \]

(ii) 40/60 Channel-charge partition
Capacitance Model Equations

\[ Q_s = -\frac{W_{\text{active}} L_{\text{active}} C_{\text{ox}}}{2 \left( V_{\text{gate eff}} - \frac{A_{\text{bulk}}}{2} V_{ds} \right)^2} \left( V_{\text{gate eff}}^3 - \frac{4}{3} V_{\text{gate eff}}^2 (A_{\text{bulk}}' V_{ds}) + \frac{2}{3} V_{\text{gate eff}} (A_{\text{bulk}}' V_{ds})^2 - \frac{2}{15} (A_{\text{bulk}}' V_{ds})^3 \right) \]

\[ Q_d = -(Q_s + Q_b + Q_t) \]

(iii) 0/100 Channel-charge Partition

\[ Q_s = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( \frac{V_{\text{gate eff}}}{2} + \frac{A_{\text{bulk}}' V_{ds}}{4} - \frac{(A_{\text{bulk}}' V_{ds})^2}{24 \left( V_{\text{gate eff}} - \frac{A_{\text{bulk}}'}{2} V_{ds} \right)} \right) \]

\[ Q_d = -(Q_s + Q_b + Q_t) \]

if \( V_{ds} > V_{d\text{sat}} \)

\[ Q_s = Q_{s\text{eff}} + W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( V_{\text{gate eff}} - \frac{V_{ds\text{sat}}}{3} \right) \]

\[ Q_b = Q_{b\text{eff}} - W_{\text{active}} L_{\text{active}} C_{\text{ox}} \left( V_{\text{gate eff}} - \frac{V_{ds\text{sat}}}{3} \right) \]
Capacitance Model Equations

(i) 50/50 Channel-charge Partition

\[ Q_s = Q_d = - \frac{W_{\text{active}} L_{\text{active}} C_{\text{ox}}}{3} V_{\text{gsteff}} \]

(ii) 40/60 Channel-charge Partition

\[ Q_s = - \frac{2W_{\text{active}} L_{\text{active}} C_{\text{ox}}}{5} V_{\text{gsteff}} \]

\[ Q_d = -(Q_s + Q_h + Q_v) \]

(iii) 0/100 Channel-charge Partition

\[ Q_s = -W_{\text{active}} L_{\text{active}} C_{\text{ox}} \frac{2V_{\text{gsteff}}}{3} \]

\[ Q_d = -(Q_s + Q_h + Q_v) \]

(3) \text{capMod} = 2

For \text{capMod} = 2, if version < 3.2, \( V_{fb} \) is calculated from

\[ V_{fb} = V_{th} - \Phi_s - K_\text{int} \sqrt{\Phi_s - V_{\text{heff}}} \]
Capacitance Model Equations

otherwise $V_{fb}$ is calculated from the above equation but under bias independent $V_{th}$. This is the same as that in $\text{capMod} = 1$.

$$Q_g = -(Q_{inv} + Q_{acc} + Q_{sub0} + \delta Q_{sub})$$

$$Q_b = Q_{acc} + Q_{sub0} + \delta Q_{sub}$$

$$Q_{inv} = Q_s + Q_d$$

$$V_{FBeff} = V_f - 0.5\left(V_s + \sqrt{V_s^2 + 4\delta_2 V_f}ight) \text{ where } V_s = V_f - V_{th} - \delta_2; \quad \delta_2 = 0.02$$

$$Q_{ss} = -W_{iamc}L_{iamc}C_{in}\left(V_{FBeff} - V_f\right)$$

$$Q_{cbl} = -W_{actv}L_{actv}C_{actv}\frac{K_{lat}}{2}\left(-1 + \sqrt{1 + \frac{4(V_c - V_{FBeff} - V_{cbl,U,C} - V_{cbl,L})}{K_{lat}^2}}\right)$$

$$V_{dta,t,CV} = \frac{V_{steff,CV}}{A_{bulk}}$$

$$A_{bulk} = A_{bulk_0}\left(1 + \left(\frac{CLC}{L_{actv}}\right)^{CLF}\right)$$
Capacitance Model Equations

\[
A_{\text{bulk}} = \left(1 + \frac{K_{\text{tr}}}{2\sqrt{V_{\text{tr}} - V_{\text{off}}} \left( \frac{A_{\text{Lg}}}{L_{\text{rg}} + 2V_{\text{th}}X_{\text{th}}} + \frac{B_{\text{th}}}{W_{\text{th}} + B_{\text{th}}} \right)} \right) \frac{1}{1 + K_{\text{eff}}V_{\text{off}}}
\]

\[
V_{\text{on},n} = \text{noff} \cdot nV_i \ln \left(1 + \exp \left(\frac{V_{\text{tr}} - V_{\text{th}} - \text{noff} \cdot nV_i}{\text{noff} \cdot nV_i}\right)\right)
\]

\[
V_{\text{on},p} = V_{\text{on},n} - 0.5 \left(V_g + \sqrt{V_g^2 + 4\delta_4 V_{\text{th},n}} \right) \quad \text{where} \quad V_g = V_{\text{th},n} - V_{\text{th}} - \delta_4; \quad \delta_4 = 0.02
\]

\[
Q_{\text{on}} = -W_{\text{on},n} L_{\text{on},n} C_e \left(V_{\text{on},n} - \frac{A_{\text{bulk}}}{2} V_{\text{on}}\right) + \frac{A_{\text{bulk}}^2 V_{\text{on}}^2}{12 \left(V_{\text{on},n} - \frac{A_{\text{bulk}}}{2} V_{\text{on}}\right)}
\]

\[
\delta Q_{\text{on}} = W_{\text{on},n} L_{\text{on},n} C_e \left(1 - \frac{A_{\text{bulk}}}{2} V_{\text{on}}\right) - \frac{(1 - A_{\text{bulk}})A_{\text{bulk}} V_{\text{on}}^2}{12 \left(V_{\text{on},n} - \frac{A_{\text{bulk}}}{2} V_{\text{on}}\right)}
\]
B.2.3.1 50/50 Charge partition

\[ Q = Q_d = 0.5Q_{	ext{tot}} = \frac{W_{\text{act}} L_{\text{act}} C_m}{2} \left( V_{\text{eff}}^2 + \frac{A_{\text{bulk}}}{2} V_{\text{eff}}^2 \right) \]

B.2.3.2 40/60 Channel-charge Partition

\[ Q = -\frac{W_{\text{act}} L_{\text{act}} C_m}{2 \left( V_{\text{eff}}^2 - \frac{A_{\text{bulk}}}{2} V_{\text{eff}}^2 \right)^2} \left( V_{\text{eff}}^2 - \frac{4}{3} V_{\text{eff}}^2 \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right) + \frac{2}{3} V_{\text{eff}}^2 \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^2 - \frac{2}{15} \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^3 \right) \]

\[ Q_d = -\frac{W_{\text{act}} L_{\text{act}} C_m}{2 \left( V_{\text{eff}}^2 - \frac{A_{\text{bulk}}}{2} V_{\text{eff}}^2 \right)^2} \left( V_{\text{eff}}^2 - \frac{5}{3} V_{\text{eff}}^2 \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right) + V_{\text{eff}}^2 \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^2 - \frac{1}{5} \left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^3 \right) \]

B.2.3.3 0/100 Charge Partition

\[ Q = -W_{\text{act}} L_{\text{act}} C_m \left( \frac{V_{\text{eff}}^2}{2} + \frac{A_{\text{bulk}} \cdot V_{\text{eff}}^2}{4} - \frac{\left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^2}{24 \left( V_{\text{eff}}^2 - \frac{A_{\text{bulk}}}{2} V_{\text{eff}}^2 \right)} \right) \]

\[ Q_d = -W_{\text{act}} L_{\text{act}} C_m \left( \frac{V_{\text{eff}}^2}{2} - \frac{3 A_{\text{bulk}} \cdot V_{\text{eff}}^2}{4} + \frac{\left( A_{\text{bulk}} \cdot V_{\text{eff}} \right)^2}{8 \left( V_{\text{eff}}^2 - \frac{A_{\text{bulk}}}{2} V_{\text{eff}}^2 \right)} \right) \]
(3) \( \text{capMod} = 3 \) (Charge-Thickness Model)

\( \text{capMod} = 3 \) only supports zero bias \( V_{fb} \), i.e. \( V_{fb} \) is calculated from bias independent \( V_{th} \) regardless of the version number. This is different from \( \text{capMod} = 1 \) and 2.

\[
v_{fb} = V_{th} - \Phi, - K_{tr} \Phi, - V_{bep}
\]

For the finite charge thickness (\( X_{DC} \)) formulations, refer to Chapter 4.

\[
Q_{acc} = W L C_{o x e f f} \cdot V_{gbacc}
\]

\[
V_{chn.} = \frac{1}{2} \left[ V_{c} + \sqrt{V_{c}^{2} + 4 \delta_{3} V_{fb}} \right]
\]

\[
V_{o} = V_{fb} \cdot V_{bocf} \cdot V_{c} - \delta_{3}
\]

\[
V_{fro} = v_{fb} - 0.5 \left[ V_{c} + \sqrt{V_{c}^{2} + 4 \delta_{3} v_{fb}} \right]
\]

where \( V_{3} = v_{fb} - V_{fb} \cdot \delta_{3} \); \( \delta_{3} = 0.02 \)

\[
C_{o x e f f} = \frac{C_{ox} C_{cen}}{C_{ox} + C_{cen}}
\]
Capacitance Model Equations

\[ C_{\text{eq}} = \frac{\varepsilon}{X_{DC}} \]

\[ \Phi_{\delta} = \Phi_{+} - 2\Phi_{\beta} = v_i \ln \left( \frac{V_{\text{eff,av}} + 2K_{\text{hn}} \sqrt{2\Phi_{\beta}}}{\text{moin} \ K_{\text{hn}} v_i^2} + 1 \right) \]

\[ Q_{\text{m1}} = -WLC_{\text{eff}} \cdot K_{\text{hn}} \cdot \frac{2}{2} \left[ -1 + \sqrt{1 + \frac{4(V_{\text{ce}} - V_{F\beta\pi} - V_{\text{eff,av}} - V_{\text{in,av}})}{K_{\text{hn}}^2}} \right] \]

\[ V_{\text{eff}} = V_{\text{d}} - \frac{1}{2} \left( V_i + \sqrt{V_i^2 + 4\delta_1 V_{\text{d}}} \right) \]

\[ V_{i} = V_{\text{d}} - V_{\text{d}} - \delta \]

\[ V_{\text{ce}} = \frac{V_{\text{ce},n} - \Phi_{\delta}}{A_{\text{nh}}} \]

\[ Q_{\text{m1}} = -WLC_{\text{eff}} \left[ V_{\text{eff,av}} - \Phi_{\delta} - \frac{1}{2} A_{\text{nh}} V_{\text{eff}} + \frac{A_{\text{nh}} V_{\text{eff}}^2}{12 \left( V_{\text{eff,av}} - \Phi_{\delta} - \frac{A_{\text{nh}} V_{\text{eff}}}{2} \right)} \right] \]
Capacitance Model Equations

\[ \delta Q_{\text{sub}} = WLC_{\text{eqn}} \left[ \frac{1 - A_{\text{bulk}}}{2} V_{\text{eff}} - \frac{(1 - A_{\text{bulk}}) \cdot A_{\text{bulk}} \cdot V_{\text{eff}}^2}{12 \cdot \left( V_{\text{eff}} - \varphi - A_{\text{bulk}} \cdot V_{\text{eff}}/2 \right)} \right] \]

(i) 50/50 Charge Partition

\[ Q_s = Q_n = \frac{1}{2} Q_s = \frac{WLC_{\text{eqn}}}{2} \left[ V_{\text{eff}} - \varphi - \frac{1}{2} A_{\text{bulk}} \cdot V_{\text{eff}} + \frac{A_{\text{bulk}} \cdot V_{\text{eff}}^2}{12 \cdot \left( V_{\text{eff}} - \varphi - A_{\text{bulk}} \cdot V_{\text{eff}}/2 \right)} \right] \]

(ii) 40/60 Charge Partition

\[ Q_s = \frac{WLC_{\text{eqn}}}{2} \left[ V_{\text{eff}} - \varphi - \frac{5}{3} V_{\text{eff}} - \varphi - \frac{A_{\text{bulk}} \cdot V_{\text{eff}}}{2} - \frac{A_{\text{bulk}} \cdot V_{\text{eff}}^2}{15 (A_{\text{bulk}} \cdot V_{\text{eff}})^2} \right] \]

(iii) 0/100 Charge Partition

\[ Q_s = \frac{WLC_{\text{eqn}}}{2} \left[ V_{\text{eff}} - \varphi + \frac{1}{2} A_{\text{bulk}} \cdot V_{\text{eff}} - \frac{A_{\text{bulk}} \cdot V_{\text{eff}}^2}{12 \cdot \left( V_{\text{eff}} - \varphi - A_{\text{bulk}} \cdot V_{\text{eff}}/2 \right)} \right] \]
Capacitance Model Equations

\[ Q_D = -\frac{WLC_{\text{eff}}}{2} \left[ V_{g_{\text{eff},cv}} - \Phi_\delta - \frac{3}{2} A_{\text{bulk}} V_{c_{\text{eff}}} + \frac{A_{\text{bulk}}^{1/2} V_{c_{\text{eff}}}}{4 \left( V_{g_{\text{eff},cv}} - \Phi_\delta - A_{\text{bulk}} V_{c_{\text{eff}}}/2 \right)} \right] \]


APPENDIX D: Model Parameter Binning

Below is the information on parameter binning regarding which model parameters can or cannot be binned. All those parameters which can be binned follow this implementation:

\[ P = P_s + \frac{P_k}{L_{eff}} + \frac{P_w}{W_{eff}} + \frac{P_f}{L_{eff} \times W_{eff}} \]

For example, for the parameter \( k_1 \): \( P_0 = k_1, P_L = Lk_1, P_W = wk_1, P_P = pk_1 \). binUnit is a binning unit selector. If binUnit = 1, the units of \( L_{eff} \) and \( W_{eff} \) used in the binning equation above have the units of microns; otherwise in meters.

For example, for a device with \( L_{eff} = 0.5 \mu m \) and \( W_{eff} = 10 \mu m \). If binUnit = 1, the parameter values for \( \text{vsat} \) are \( 1e5, 1e4, 2e4, \) and \( 3e4 \) for \( \text{vsat}, \text{lvsat}, \text{wvsat}, \) and \( \text{pvsat} \), respectively.

Therefore, the effective value of \( \text{vsat} \) for this device is

\[ \text{vsat} = 1e5 + 1e4/0.5 + 2e4/10 + 3e4/(0.5*10) = 1.28e5 \]

To get the same effective value of \( \text{vsat} \) for binUnit = 0, the values of \( \text{vsat}, \text{lvsat}, \text{wvsat}, \) and \( \text{pvsat} \) would be \( 1e5, 1e-2, 2e-2, 3e-8 \), respectively. Thus,

\[ \text{vsat} = 1e5 + 1e-2/0.5e6 + 2e-2/10e-6 + 3e-8/(0.5e-6 * 10e-6) = 1.28e5 \]
### D.1 Model Control Parameters

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<th>Symbols used in equation</th>
<th>Symbols used in SPICE</th>
<th>Description</th>
<th>Can Be Binned?</th>
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<td>The model selector</td>
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### D.2 DC Parameters

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### DC Parameters

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## DC Parameters

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## DC Parameters

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## D.3 AC and Capacitance Parameters

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### AC and Capacitance Parameters

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### D.4 NQS Parameters

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### D.5 dW and dL Parameters

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## dW and dL Parameters

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### Flicker Noise Model Parameters

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### D.7 Flicker Noise Model Parameters

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#### D.8 Process Parameters

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## D.9 Geometry Range Parameters

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