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DESIGN AND MODELING OF NON-UNIFORMLY DOPED
DEEP-SUBMICRON POCKET MOSFETS FOR LOW-VOLTAGE
LOW-POWER APPLICATIONS

by

Yon Sup Pang

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A Dissertation Submitted to the Faculty of the
DEPARTMENT OF ELECTRICAL AND COMPUTER ENGINEERING
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In the Graduate College
THE UNIVERSITY OF ARIZONA
2000
As members of the Final Examination Committee, we certify that we have read the dissertation prepared by Yon Sup Pang entitled DESIGN AND MODELING OF NON-UNIFORMLY DOPED DEEP-SUBMICRON POCKET MOSFETS FOR LOW-VOLTAGE LOW-POWER APPLICATIONS and recommend that it be accepted as fulfilling the dissertation requirement for the Degree of Doctor of Philosophy.

John R. Brews (Dissertation Director) 11/6/80
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Final approval and acceptance of this dissertation is contingent upon the candidate's submission of the final copy of the dissertation to the Graduate College.

I hereby certify that I have read this dissertation prepared under my direction and recommend that it be accepted as fulfilling the dissertation requirement.

Dissertation Director John R. Brews 11/6/80
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SIGNED: Yen-Luh Pang
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To my parents,

*Sung-Young Pang* and *Sun-Hee Kim*

In memory of

my father
TABLE OF CONTENTS

LIST OF FIGURES ................................................. 9
LIST OF TABLES .................................................. 22
ABSTRACT .......................................................... 23
CHAPTER 1. INTRODUCTION ....................................... 25

CHAPTER 2. MOBILITY MODELS ................................. 36
  2.1 Physics-Based Effective Mobility Model ................. 37
    2.1.1 Relations between Effective Transverse Electric Field and Physical Parameters .......... 37
    2.1.2 Mobility Dependence on Lateral Field .............. 46
    2.1.3 Mobility Dependence on Effective Transverse Field at Low Lateral Field .............. 46
  2.2 Doping-Density-Dependent Mobility Model at Low Transverse Electric Field ........... 50
  2.3 Extraction of Effective Transverse-Field-Dependent Mobility at Low Lateral Electric Field ........... 53
  2.4 Summary ................................................... 57

CHAPTER 3. DESIGN OF 0.1-\(\mu m\) POCKET N-MOSFETS .... 59
  3.1 Short-Channel Effects and Characteristics of Pocket and Conventional Bulk n-MOSFETs ........... 60
  3.2 0.1-\(\mu m\) Pocket n-MOSFET Structures Satisfying Specifications of Off-State Current and Sensitivity of Off-State Current to Channel Length ........... 72
    3.2.1 Computer Algorithms Based on Device Physics and Two-Dimensional Device Simulation Results .......... 72
    3.2.2 Unique Lower Design-Space Boundary and Types of Deep-Submicron Pocket n-MOSFETs .......... 85
    3.2.3 Type A Devices ........................................ 88
    3.2.4 Type B Devices ........................................ 91
  3.3 Comments on Design Philosophy for 0.1-\(\mu m\) Pocket n-MOSFETs ................ 92
  3.4 Summary ................................................... 93

CHAPTER 4. MODELING OF OFF-STATE CURRENT .......... 95
TABLE OF CONTENTS – Continued

4.1 Subthreshold Surface Potential Model for $V_{GS} = 0$ V and $V_{DS} = 0$ V 96
   4.1.1 Model Equations and Surface Potential Patterns .............. 97
   4.1.2 Analytical and Empirical Formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$ ... 104
4.2 Subthreshold Surface Potential Model for $V_{GS} = 0$ V and $V_{DS} > 0$ V 112
   4.2.1 Model Equations and Model Parameter Extraction ............... 114
   4.2.2 Analytical and Empirical Formulas for $\ell_{ps}$, $\ell_{pD}$ and $\ell_c$ for
       $V_{GS} = 0$ V and $V_{DS} = 1$ V ................................ 124
       4.2.2.1 z-Formulation ........................................ 125
       4.2.2.2 Power-Law Formulation ............................... 130
4.3 Modeling of Off-State Current and Model Results .................. 132
   4.3.1 Off-State Current Model .................................. 134
   4.3.2 Results of $I_{off}$ Model and Prediction of Lower Design-Space
       Boundary for $I_{off}$ ........................................ 140
4.4 Simple Model for Off-State Current in 0.1-$\mu$m Pocket n-MOSFETs
   and Design-Space Boundary for Off-State Current ............... 145
   4.4.1 Off-State Current Model and $V_p$ vs. $L_p$ of Lower Design-
       Space Boundary ........................................... 146
   4.4.2 Surface Potential Model and $V_c$ vs. $L_p$ of Lower Design-
       Space Boundary ........................................... 149
   4.4.3 Model Prediction of Barrier Lowering Effect of Pocket n-
       MOSFETs .................................................. 153
4.5 Summary .................................................................. 158

CHAPTER 5. MODELING OF ON-STATE CURRENT ......................... 160

5.1 Characteristics of Deep-Submicron Pocket n-MOSFETs Biased to
   Strong Inversion ................................................. 162
5.2 Model Equations for Above-Threshold Current .................... 176
5.3 Model Results and Parameter Extraction ........................... 188
5.4 Formulation of Model Parameters of $\kappa_p$ and $\ell_{sat}$ ........... 195
   5.4.1 z-Formulation ............................................ 195
   5.4.2 Power-Law Formulation .................................. 198
5.5 Model Result and Prediction of Upper Design-Space Boundary for
   $I_{on}$ .................................................................. 201
5.6 Summary .................................................................. 206

CHAPTER 6. VIABLE DESIGN SPACE AND ADVANTAGES OF
0.1-$\mu$m POCKET N-MOSFETS ........................................... 208

6.1 Viable Design Space for Deep-Submicron Pocket n-MOSFETs ... 209
8

### TABLE OF CONTENTS – Continued

6.2 Advantages of 0.1-\(\mu m\) Pocket n-MOSFETs over 0.1-\(\mu m\) Conventional Bulk n-MOSFETs ................................................................. 222
6.3 Summary ......................................................................................... 231

**CHAPTER 7. CONCLUSIONS** .............................................................. 232

**APPENDIX A. VELOCITY OVERSHOOT IN ULTRASHORT-CHANNEL POCKET N-MOSFET** .......................................................... 234

**APPENDIX B. IMPLEMENTATION OF ALGORITHMS FOR FINDING DESIGN-SPACE BOUNDARY FOR \(I_{\text{off}}\) AND \(\frac{\Delta I_{\text{off}}}{\Delta L/L}\) .................................................................................. 250

**APPENDIX C. ALGORITHM TO FIND SINGLE-DEVICE VALUES OF \(\ell_{p0}, \ell_{c0}\) AND \(K_p\) ........................................................................ 261

**REFERENCES** .................................................................................. 269
LIST OF FIGURES

1.1 (a) Uniformly doped n-MOSFET and (b) pocket n-MOSFET.

2.1 Surface potential $\phi_s$ vs. bulk doping density $N_A$ for two different oxide thicknesses $t_{ox} = 4$ and $10$ nm, temperature $T = 300^\circ K$, and (a) $V_{GS} = 0$ V and (b) $V_{GS} = 1$ V. The results were obtained by solving Equation 2.13.

2.2 Transverse electric fields vs. bulk doping density for $t_{ox} = 4$ nm and $V_{GS} = 0$ V. $\mathcal{E}_{eff} = \mathcal{E}_{N_f} + \mathcal{E}_{D}$.  

2.3 Transverse electric fields vs. bulk doping density for $t_{ox} = 4$ nm and $V_{GS} = 1$ V. $\mathcal{E}_{eff} = \mathcal{E}_{N_f} + \mathcal{E}_{D}$.  

2.4 The dependence of $\mathcal{E}_{eff}$ on both $t_{ox}$ and $N_A$ for $V_{GS} = 0$ V.  

2.5 The dependence of $\mathcal{E}_{eff}$ on both $t_{ox}$ and $N_A$ for $V_{GS} = 1$ V.  

2.6 Electron density in channel vs. effective transverse electric field for $t_{ox} = 4, 10$ nm. (a) $V_{GS} = 0$ V and (b) $V_{GS} = 1$ V.  

2.7 Effective transverse-field-dependent mobility at low lateral electric field vs. effective transverse electric field with bulk doping level as a parameter for $N_f = 0$ cm$^{-2}$ and $T = 300^\circ K$.  

2.8 Results of the unified mobility model combining the transverse-field-dependent and the doping-density-dependent mobility models at a low lateral electric field with bulk doping level as a parameter for $N_f = 0$ cm$^{-2}$ and $T = 300^\circ K$.  

2.9 Uniformly doped long-channel n-MOSFET used for extracting effective transverse-field-dependent mobility for $V_{DS} = 0.05$ V.  

2.10 Comparisons of the effective transverse-field-dependent mobility of the unified mobility model to the extracted mobility for $N_f = 0$ cm$^{-2}$ and $T = 300^\circ K$.  

3.1 Medici results of surface potential $\psi_s$ vs. normalized distance from source (a) of pocket n-MOSFETs with $L = 0.1, 1$ $\mu$m. $N_p = 1.948 \times 10^{18}$ cm$^{-3}$, $N_c = 2.175 \times 10^{17}$ cm$^{-3}$ and $L_p = 0.024$ $\mu$m. and (b) of conventional bulk n-MOSFETs with $L = 0.1, 1$ $\mu$m and $N_A = 2.175 \times 10^{17}$ cm$^{-3}$. $V_{DS} = 0$ V, $V_{GS} = 0$ V, $V_{BG} = 0$ V, $t_{ox} = 4$ nm and $r_j = 0.06$ $\mu$m for all the devices.
3.2 Medici results of surface potential $\psi_s$ vs. normalized distance from source (a) of pocket n-MOSFETs with $L = 0.1 \mu m$, $N_p = 1.948 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu m$. and (b) of conventional bulk n-MOSFETs with $L = 0.1 \mu m$ and $N_A = 2.175 \times 10^{17} \text{ cm}^{-3}$. $V_{GS} = 0$, $V_{BG} = 0$, $t_{ox} = 4 \text{ nm}$, and $r_j = 0.06 \mu m$ for all the devices.

3.3 Subthreshold characteristics of (a) 0.1 and 1 $\mu m$ channel-length pocket n-MOSFETs with $N_p = 1.948 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu m$, and (b) 0.1 and 1 $\mu m$ channel-length conventional bulk n-MOSFETs with $N_A = 2.175 \times 10^{17} \text{ cm}^{-3}$, $V_{DS} = 0.1$ and 1 $V$.

3.4 Schematic illustration of the sensitivity of off-state current to channel length.

3.5 (a) Percent change of $I_{off}$ due to 10% decrease in channel length vs. doping density $N_c$ at center region for $I_{off}(0.9L) = 1 \text{ nA}/\mu m$. $L = 0.1 \mu m$ and $L_p = 0.025, 0.0325$ and 0.04 $\mu m$. (b) Percent change of $I_{off}$ vs. $N_c$ and $N_p$ for $L_p = 0.025$ and 0.04 $\mu m$. This figure (b) is a part of the figure (a).

3.6 (a) Percent change of $I_{off}$ due to 10% decrease in channel length vs. doping density $N_p$ at pocket region for $L_p = 0.025, 0.0325$ and 0.04 $\mu m$ and (b) percent change of $I_{off}$ due to 10% decrease in channel length vs. $L_p$ for $N_c = 1 \times 10^{17}, 2 \times 10^{17}$ and $3 \times 10^{17} \text{ cm}^{-3}$. $I_{off}(0.9L) = 1 \text{ nA}/\mu m$ and $L = 0.1 \mu m$.

3.7 (a) $N_p$ vs. $L_p$ for the 0.1-$\mu m$ pocket n-MOSFETs. $N_c$ in this figure is taken from Figure (b) below. (b) $N_c$ vs. $L_p$ for the 0.1-$\mu m$ pocket n-MOSFETs. $N_p$ in this figure is taken from Figure (a) above. All the 0.1-$\mu m$ pocket devices located at the curves, i.e., at the unique lower design-space boundary satisfy the design specifications of $I_{off} \leq 1 \text{ nA}/\mu m$ and $\Delta I_{off}/\Delta L = 40\%/10\%$. The lines in the figures simply connect the Medici data points.

3.8 (a) Minima of surface potential distributions, i.e., the highest potential barriers to electron injection from source of the 0.1-$\mu m$ pocket n-MOSFETs located at lower design-space boundary for $I_{off} = 0.714 \text{ nA}/\mu m$ and $\Delta I_{off}/\Delta L = 40\%/10\%$, and (b) locations at which the highest potential barriers exist. $V_{GS} = 0 \text{ V}$ and $V_{DS} = 1 \text{ V}$. (a), (b), (c) and (d) in these figures mean the 0.1-$\mu m$ pocket devices (a), (b), (c) and (d) in Figure 3.7, respectively.

3.9 Subthreshold characteristics for $V_{DS} = 1 \text{ V}$ of the 0.1-$\mu m$ pocket n-MOSFETs (a), (b), (c) and (d) in Figure 3.7 (a).
3.10 Potential contours for \( V_{GS} = 0 \text{ V} \) and \( V_{DS} = 1 \text{ V} \) inside the 0.1-\( \mu \text{m} \) pocket n-MOSFETs located at (a) and (d) in Figure 3.7 (a). Dashed lines are depletion-region edges.

3.11 Potential contours for \( V_{GS} = 0 \text{ V} \) and \( V_{DS} = 1 \text{ V} \) inside the 0.1-\( \mu \text{m} \) pocket n-MOSFET located at (e) in Figure 3.7 (a). Dashed line is a depletion-region edge.

4.1 Surface potential distributions between source and drain of (a) 0.1-\( \mu \text{m} \) pocket n-MOSFET with \( N_c = 7.0 \times 10^{17} \text{ cm}^{-3} \), \( N_p = 1.281 \times 10^{18} \text{ cm}^{-3} \) and \( L_p = 0.025 \mu \text{m} \), and (b) 0.2-\( \mu \text{m} \) pocket n-MOSFET with \( N_c = 2.5 \times 10^{17} \text{ cm}^{-3} \), \( N_p = 6.075 \times 10^{17} \text{ cm}^{-3} \) and \( L_p = 0.05 \mu \text{m} \). \( t_{ox} = 4 \text{ nm} \), \( r_j = 0.06 \mu \text{m} \). \( V_{GS} = 0 \text{ V} \) and \( V_{DS} = 0 \text{ V} \) for both pocket devices. Single-device values of \( \ell_{p0} \), \( \ell_c \) and \( K_p \) were used for the model.

4.2 Surface potential \( \psi_{s0} \) profiles between source and drain of 0.1-\( \mu \text{m} \) pocket n-MOSFETs for \( V_{GS} = 0 \text{ V} \) and \( V_{DS} = 0 \text{ V} \). (a) \( L_p = 0.022 \mu \text{m} \), \( N_p = 2.135 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 2.375 \times 10^{17} \text{ cm}^{-3} \), (b) \( L_p = 0.028 \mu \text{m} \), \( N_p = 1.558 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 1.925 \times 10^{17} \text{ cm}^{-3} \), (c) \( L_p = 0.034 \mu \text{m} \), \( N_p = 1.205 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 1.825 \times 10^{17} \text{ cm}^{-3} \) and (d) \( L_p = 0.044 \mu \text{m} \), \( N_p = 8.557 \times 10^{17} \text{ cm}^{-3} \), \( N_c = 2.138 \times 10^{17} \text{ cm}^{-3} \). \( t_{ox} = 4 \text{ nm} \) and \( r_j = 0.06 \mu \text{m} \) for all the pocket devices. Single-device values of \( \ell_{p0} \), \( \ell_c \) and \( K_p \) were used for the model.

4.3 Extracted optimal values as single-device values of the model parameters \( \ell_{p0} \), \( \ell_c \) and \( |K_p| \) vs. \( z = L/L_p N_p N_c \) of the 0.1-\( \mu \text{m} \) pocket n-MOSFETs located at the lower design-space boundary. \( V_{DS} = 0 \text{ V} \) and \( V_{GS} = 0 \text{ V} \).

4.4 \( \log_{10} \ell_{p0} \) vs. \( \log_{10}(L/L_p N_p N_c) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10} \ell_{p0} \) of (a) Type A (0.022\( \mu \text{m} \) \( \leq \) \( L_p \leq \) 0.036\( \mu \text{m} \)) and (b) Type B (0.036\( \mu \text{m} \) \( < \) \( L_p \leq \) 0.044\( \mu \text{m} \)) of 0.1-\( \mu \text{m} \) pocket n-MOSFETs.

4.5 \( \log_{10} \ell_{c0} \) vs. \( \log_{10}(L/L_p N_p N_c) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10} \ell_{c0} \) of (a) Type A (0.022\( \mu \text{m} \) \( \leq \) \( L_p \leq \) 0.034\( \mu \text{m} \)) and (b) Type B (0.034\( \mu \text{m} \) \( < \) \( L_p \leq \) 0.044\( \mu \text{m} \)) of 0.1-\( \mu \text{m} \) pocket n-MOSFETs.

4.6 \( \log_{10}|K_p| \) vs. \( \log_{10}(L/L_p N_p N_c) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10}|K_p| \) of (a) Type A (0.022\( \mu \text{m} \) \( \leq \) \( L_p \leq \) 0.034\( \mu \text{m} \)) and (b) Type B (0.034\( \mu \text{m} \) \( < \) \( L_p \leq \) 0.044\( \mu \text{m} \)) of 0.1-\( \mu \text{m} \) pocket n-MOSFETs.
4.7 Surface potential $\psi_{so}$ profiles between source and drain of 0.1-$\mu$m pocket n-MOSFETs which are the same ones used in Figure 4.2 for $V_{GS} = 0 \ V$ and $V_{DS} = 0 \ V$. The model surface potentials were generated by the complete model equations, including the z-formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$. (a) $L_p = 0.022 \ \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$. $N_c = 2.375 \times 10^{17} \ cm^{-3}$. (b) $L_p = 0.028 \ \mu m$, $N_p = 1.558 \times 10^{18} \ cm^{-3}$. $N_c = 1.925 \times 10^{17} \ cm^{-3}$. (c) $L_p = 0.034 \ \mu m$, $N_p = 1.205 \times 10^{18} \ cm^{-3}$. $N_c = 1.825 \times 10^{17} \ cm^{-3}$. and (d) $L_p = 0.044 \ \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$. $N_c = 2.138 \times 10^{17} \ cm^{-3}$. $t_{ox} = 4 \ nm$ and $r_j = 0.06 \ \mu m$ for all the pocket devices. Global values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$ were used for the model.

4.8 Model surface potential $\psi_c$ for $V_{DS} = 1 \ V$, $V_{GS} = 0 \ V$ checked for pocket n-MOSFETs with (a) $L = 0.1 \ \mu m$, $L_p = 0.022 \ \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$. $N_c = 2.375 \times 10^{17} \ cm^{-3}$. (b) $L = 0.1 \ \mu m$, $L_p = 0.044 \ \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$. $N_c = 2.138 \times 10^{17} \ cm^{-3}$. (c) $L = 1 \ \mu m$, $L_p = 0.022 \ \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$. $N_c = 2.375 \times 10^{17} \ cm^{-3}$. and (d) $L = 1 \ \mu m$, $L_p = 0.044 \ \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$. $N_c = 2.138 \times 10^{17} \ cm^{-3}$, $t_{ox} = 4 \ nm$ and $r_j = 0.06 \ \mu m$ for all the devices. Crude formulas in Equations 4.40 and 4.41 for $\ell_{pS}$, $\ell_{pD}$, $\ell_{c}$, $\ell_{p0}$ and $\ell_{c0}$, and $K_p = 0$ were used for the model.

4.9 Surface potential $\psi_c$ profiles between source and drain of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0 \ V$ and $V_{DS} = 1 \ V$. (a) $L_p = 0.022 \ \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$. $N_c = 2.375 \times 10^{17} \ cm^{-3}$. (b) $L_p = 0.028 \ \mu m$, $N_p = 1.558 \times 10^{18} \ cm^{-3}$. $N_c = 1.925 \times 10^{17} \ cm^{-3}$. (c) $L_p = 0.034 \ \mu m$, $N_p = 1.205 \times 10^{18} \ cm^{-3}$. $N_c = 1.825 \times 10^{17} \ cm^{-3}$, and (d) $L_p = 0.044 \ \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$. $N_c = 2.138 \times 10^{17} \ cm^{-3}$. $t_{ox} = 4 \ nm$ and $r_j = 0.06 \ \mu m$ for all the pocket devices. The 0.1-$\mu$m pocket MOSFETs chosen here are the same ones used in Figure 4.2. Single-device values of $\ell_{pS}$, $\ell_{pD}$, $\ell_{c}$, $\ell_{p0}$ and $\ell_{c0}$ and $K_p$ were used for the model.

4.10 The extracted optimal values as single-device values of the model parameters $\ell_{pS}$, $\ell_{pD}$ and $\ell_{c}$ vs. $z = L L_p N_p N_c$ of the 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary. $V_{DS} = 1 \ V$ and $V_{GS} = 0 \ V$.

4.11 Rms errors of potentials of model surface potential profiles to those of Medici surface potential profiles for $V_{DS} = 1 \ V$ and $V_{GS} = 0 \ V$ of the 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary. The single-device values of $\ell_{pS}$, $\ell_{pD}$ and $\ell_{c}$ in Table 4.5 were used to obtain the model surface potential profiles.
4.12 $\log_{10} \ell_{ps}$ vs. $\log_{10}(LL_p N_p V_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{ps}$ of (a) Type A ($0.022 \mu m < L_p < 0.032 \mu m$) and (b) Type B ($0.032 \mu m < L_p < 0.044 \mu m$) of 0.1-μm pocket n-MOSFETs.

4.13 $\log_{10} \ell_{pd}$ vs. $\log_{10}(LL_p N_p V_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{pd}$ of (a) Type A ($0.022 \mu m < L_p < 0.036 \mu m$) and (b) Type B ($0.036 \mu m < L_p < 0.044 \mu m$) of 0.1-μm pocket n-MOSFETs.

4.14 $\log_{10} \ell_c$ vs. $\log_{10}(LL_p N_p V_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_c$ of (a) Type A ($0.022 \mu m < L_p < 0.036 \mu m$) and (b) Type B ($0.036 \mu m < L_p < 0.044 \mu m$) of 0.1-μm pocket n-MOSFETs.

4.15 Surface potential $\psi_s$ profiles between source and drain for $V_{GS} = 0$ and $V_{DS} = 1$ V of the 0.1-μm pocket n-MOSFETs which are the same ones used in Figure 4.9. The model surface potentials were generated by the complete model equations, including the z-formulas for $\ell_{ps}$, $\ell_{pd}$, $\ell_c$, $\ell_{po}$, $\ell_e0$ and $K_p$. (a) $L_p = 0.022 \mu m$, $N_p = 2.135 \times 10^{18}$ cm$^{-3}$, $N_c = 2.375 \times 10^{17}$ cm$^{-3}$, (b) $L_p = 0.028 \mu m$, $N_p = 1.558 \times 10^{18}$ cm$^{-3}$, $N_c = 1.925 \times 10^{17}$ cm$^{-3}$, (c) $L_p = 0.034 \mu m$, $N_p = 1.205 \times 10^{18}$ cm$^{-3}$, $N_c = 1.825 \times 10^{17}$ cm$^{-3}$, and (d) $L_p = 0.044 \mu m$, $N_p = 8.557 \times 10^{17}$ cm$^{-3}$, $N_c = 2.138 \times 10^{17}$ cm$^{-3}$, $t_{ox} = 4$ nm and $r_j = 0.06 \mu m$. Global values of $\ell_{ps}$, $\ell_{pd}$, $\ell_c$, $\ell_{po}$, $\ell_e0$ and $K_p$ were used for the model.

4.16 DIBL effect for $V_{GS} = 0$ V of 0.1-μm pocket n-MOSFET with $L_p = 0.022 \mu m$, $N_p = 2.135 \times 10^{18}$ cm$^{-3}$, $N_c = 2.375 \times 10^{17}$ cm$^{-3}$, $t_{ox} = 4$ nm and $r_j = 0.06 \mu m$. The subthreshold surface potential model for $\psi_s(x)$ was used to generate the potential profiles.

4.17 Rms errors of potentials of model surface potential profiles to those of Medici surface potential profiles for $V_{DS} = 1$ V and $V_{GS} = 0$ V of the 0.1-μm pocket n-MOSFETs located at the lower design-space boundary. Either the global or the single-device values of $\ell_{ps}$, $\ell_{pd}$, $\ell_c$, $\ell_{po}$, $\ell_e0$ and $K_p$ were used to obtain the model surface potential profiles.

4.18 Comparisons of global values with single-device values of the parameter $\ell_{ps}$ of 0.1-μm pocket n-MOSFETs. (a) $\ell_{ps}$ vs. $L_p$, (b) $\ell_{ps}$ vs. $N_p/N_p0$, (c) $\ell_{ps}$ vs. $N_c/N_c0$ and (d) $\ell_{ps}$ vs. $z/z_0$. The global parameter values were calculated using power-law and z-formulas. $N_p0 = 10^{18}$ cm$^{-3}$, $N_c0 = 10^{17}$ cm$^{-3}$ and $z_0 = 10^{24}$ cm$^{-4}$. 

4.19 Comparisons of global values with single-device values of the parameters $\ell_{pD}$ and $\ell_c$ of 0.1-$\mu m$ pocket n-MOSFETs. (a) $\ell_{pD}$ vs. $L_p$ and (b) $\ell_c$ vs. $L_p$. The global parameter values were calculated using power-law and $z$-formulas.

4.20 $\log_{10} L_{eff}$ vs. $\log_{10}(L_{p} N_p N_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} L_{eff}$ of (a) Type A ($0.022\mu m \leq L_p \leq 0.034\mu m$) and (b) Type B ($0.034\mu m < L_p \leq 0.044\mu m$) of 0.1-$\mu m$ pocket n-MOSFETs.

4.21 Calculation of $I_{eff}$.

4.22 Model results of (a) $N_p$ vs. $L_p$ and (b) $N_c$ vs. $L_p$ at the lower design-space boundary for $I_{eff} = 0.714$ nA/$\mu m$ and $\frac{\Delta I_{eff}/I_{eff}}{\Delta L_{eff}} = 40\% / 10\%$ of the 0.1-$\mu m$ pocket n-MOSFETs are compared with the two-dimensional simulation results shown in Figure 3.7. The lines in the figures are the model results.

4.23 Model predictions of design-space boundaries for the worst-case specification of $I_{eff} = 1$ nA/$\mu m$ in 0.1-$\mu m$ pocket n-MOSFETs with $L_p = 0.022 - 0.033 \mu m$. (a) Two-dimensional design-space boundary $N_p$ vs. $L_p$ with $N_c$ as a parameter, and (b) three-dimensional design-space boundary.

4.24 Model predictions of design-space boundary of $N_p$ vs. $L_p$ with $N_c$ and $I_{eff} \leq 1$ nA/$\mu m$ as parameters of 0.1-$\mu m$ pocket n-MOSFETs.

4.25 Comparison of global parameter values with single-device values of $L_{eff}$ of 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary.

4.26 Comparison of the model lower design-space boundary for $I_{eff}$ in 0.1-$\mu m$ pocket n-MOSFETs with the Medici lower design-space boundary.

4.27 Comparison of global parameter values with single-device values of $\zeta$ of 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary.

4.28 Comparisons of global parameter values with single-device values of (a) $\ell_{p0}$ and (b) $\ell_{c0}$ of 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary.

4.29 Comparison of the model with the Medici $N_c$ vs. $L_p$ of the lower design-space boundary for $I_{eff}$ in 0.1-$\mu m$ pocket n-MOSFETs. Twelve Medici values of $\psi_{30}(max)(L/2)(\text{Medici})$ were used to find the model $N_c$ vs. $L_p$.

4.30 Medici values of $\psi_{30}(max)(L/2)(\text{Medici})$ of the twelve 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary.
4.31 Surface potential $\psi_{s0}$ profiles between source and drain for $V_{DS} = 0 \, V$, $V_{GS} = 0 \, V$ and three different values of $N_c$ of 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.104 \times 10^{18} \, \text{cm}^{-3}$ and $L_p = 0.022 \, \mu\text{m}$.

4.32 Comparisons of global values with single-device values of (a) $\ell_{ps}$ and (b) $\ell_{pD}$ of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$. Fourth-degree polynomials in terms of $L_p$ were used to formulate the model parameters.

4.33 Comparison of global values with single-device values of $\ell_c$ of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$. Fourth-degree polynomial in terms of $L_p$ was used to formulate the model parameter.

4.34 Variation of surface potential profile between source and drain of pocket n-MOSFET with channel length $L$ for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$. $L_p = 0.024 \, \mu\text{m}$. $N_p = 1.906 \times 10^{18} \, \text{cm}^{-3}$ $N_c = 2.175 \times 10^{17} \, \text{cm}^{-3}$. $t_{ox} = 4 \, \text{nm}$ and $r_j = 0.06 \, \mu\text{m}$. The subthreshold surface potential model for $\psi_s(x)$ was used to generate the potential profiles.

5.1 (a) The unique upper design-space boundary $N_p$ vs. $L_p$ for $I_m = 0.2 \, \text{mA/}\mu\text{m}$ in 0.1-$\mu$m pocket n-MOSFETs, and (b) $N_c$ vs. $L_p$ of all the devices located at the design-space boundary. This figure (b) is a repeat of Figure 3.7 (b) for the pocket devices satisfying the channel-length sensitivity and the off-current specifications.

5.2 (a) Potential (b) electric field and (c) electron density profiles between source and drain at the Si-SiO$_2$ interface $(x, 0)$ for $V_{DS} = 1 \, V$ and $V_{GS} = 1 \, V$, and (d) $I_D - V_{DS}$ characteristic at $V_{GS} = 1 \, V$ of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 2.175 \times 10^{17} \, \text{cm}^{-3}$ and $L_p = 0.024 \, \mu\text{m}$.

5.3 Medici $I_D - V_{DS}$ characteristics of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 2.175 \times 10^{17} \, \text{cm}^{-3}$ and $L_p = 0.024 \, \mu\text{m}$. The model C was used for the two-dimensional device simulations.

5.4 Medici $I_D - V_{DS}$ characteristics at $V_{GS} = 1 \, V$ of 0.1-$\mu$m pocket n-MOSFETs with (a) $L_p = 0.022 \, \mu\text{m}$, $N_p = 2.995 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 2.375 \times 10^{17} \, \text{cm}^{-3}$, (b) $L_p = 0.028 \, \mu\text{m}$, $N_p = 2.276 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 1.925 \times 10^{17} \, \text{cm}^{-3}$, (c) $L_p = 0.038 \, \mu\text{m}$, $N_p = 1.580 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 1.875 \times 10^{17} \, \text{cm}^{-3}$, (d) $L_p = 0.044 \, \mu\text{m}$, $N_p = 1.307 \times 10^{18} \, \text{cm}^{-3}$. $N_c = 2.138 \times 10^{17} \, \text{cm}^{-3}$, and of 0.1-$\mu$m conventional bulk n-MOSFET with $N_A = 1.098 \times 10^{18} \, \text{cm}^{-3}$. All the devices are located at the design-space boundary in Figure 5.1. $t_{ox} = 4 \, \text{nm}$ and $r_j = 0.06 \, \mu\text{m}$. The model C was used for the two-dimensional device simulations.
5.5 Comparisons between 0.1-\mu m pocket n-MOSFETs located at the upper design-space boundary for \( I_{on} \) in Figure 5.1 and a 0.1-\mu m conventional bulk n-MOSFET with \( N_A = 1.098 \times 10^{18} \text{ cm}^{-3} \) and the same \( I_{on} = 0.2 \text{ mA/\mu m} \), \( t_{ox} = 4 \text{ nm} \) and \( r_j = 0.06 \mu m \) as those of the pocket n-MOSFETs regarding (a) \( g_m \) and (b) \( v_{eff} \). \( v_{eff} \) was calculated using Equations 5.1 and 5.2. The model C was used for all the Medici simulations. \( V_{DS} = 1 \text{ V} \) and \( V_{GS} = 1 \text{ V} \).

5.6 (a) Electric potential and (b) longitudinal electric field profiles at the Si surface for the 0.1-\mu m pocket n-MOSFET with \( N_p = 2.713 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 2.175 \times 10^{17} \text{ cm}^{-3} \) and \( L_p = 0.024 \mu m \), and the 0.1-\mu m conventional bulk n-MOSFET with \( N_A = 1.098 \times 10^{18} \text{ cm}^{-3} \). The devices have the same \( I_{on} = 0.2 \text{ mA/\mu m} \). \( V_{DS} = 1 \text{ V} \) and \( V_{GS} = 1 \text{ V} \).

5.7 Pocket n-MOSFET structure illustrating the gradual channel and velocity saturation regions.

5.8 Model \( I_D - V_{DS} \) characteristics at \( V_{GS} = 1 \text{ V} \) compared to Medici data for the 0.1-\mu m pocket n-MOSFETs with (a) \( L_p = 0.022 \mu m \), \( N_p = 2.995 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 2.375 \times 10^{17} \text{ cm}^{-3} \), (b) \( L_p = 0.028 \mu m \), \( N_p = 2.276 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 1.925 \times 10^{17} \text{ cm}^{-3} \), (c) \( L_p = 0.038 \mu m \), \( N_p = 1.580 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 1.875 \times 10^{17} \text{ cm}^{-3} \), and (d) \( L_p = 0.044 \mu m \), \( N_p = 1.307 \times 10^{18} \text{ cm}^{-3} \), \( N_c = 2.138 \times 10^{17} \text{ cm}^{-3} \). Single-device values of \( \kappa_p \), \( \ell_{sat} \) and \( K_{sat} = 2.3 \) were used for the model.

5.9 The extracted optimal values as single-device values of the model parameters (a) \( \kappa_p \) and (b) \( \ell_{sat} \) vs. \( z = LL_pN_pN_c t_{ox} r_j \) of the 0.1-\mu m pocket n-MOSFETs located at the upper design-space boundary for \( I_{on} = 0.2 \text{ mA/\mu m} \). \( K_{sat} = 2.3 \) and \( V_{GS} = 1 \text{ V} \).

5.10 Rms errors of drain voltages of model \( I_D - V_{DS} \) curves to those of Medici \( I_D - V_{DS} \) curves at \( V_{GS} = 1 \text{ V} \) of the 0.1-\mu m pocket n-MOSFETs located at the upper design-space boundary. The single-device values of \( \kappa_p \) and \( \ell_{sat} \) in Table 5.2. and \( K_{sat} = 2.3 \) were used to obtain the model \( I_D - V_{DS} \) curves.

5.11 \( \log_{10} \kappa_p \) vs. \( \log_{10}(LL_pN_pN_c t_{ox} r_j) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10} \kappa_p \) of (a) Type A (0.022\mu m \leq L_p \leq 0.034\mu m) and (b) Type B (0.034\mu m < L_p \leq 0.044\mu m) of the 0.1-\mu m pocket n-MOSFETs.

5.12 \( \log_{10} \ell_{sat} \) vs. \( \log_{10}(LL_pN_pN_c t_{ox} r_j) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10} \ell_{sat} \) of (a) Type A (0.022\mu m \leq L_p \leq 0.034\mu m) and (b) Type B (0.034\mu m < L_p \leq 0.044\mu m) of the 0.1-\mu m pocket n-MOSFETs.
5.13 \( I_D - V_{DS} \) characteristics at \( V_{GS} = 1 \, \text{V} \) generated by the complete model equations, including the z-formulas for \( \kappa_p \), \( \ell_{sat} \) and \( K_{sat} = 2.3 \). and compared to Medici \( I_D - V_{DS} \) for the 0.1-\( \mu \)m pocket n-MOSFETs with (a) \( L_p = 0.022 \, \mu \text{m}, \, N_p = 2.995 \times 10^{18} \, \text{cm}^{-3}, \, N_c = 2.375 \times 10^{17} \, \text{cm}^{-3} \), (b) \( L_p = 0.028 \, \mu \text{m}, \, N_p = 2.276 \times 10^{18} \, \text{cm}^{-3}, \, N_c = 1.925 \times 10^{17} \, \text{cm}^{-3} \), (c) \( L_p = 0.038 \, \mu \text{m}, \, N_p = 1.580 \times 10^{18} \, \text{cm}^{-3}, \, N_c = 1.875 \times 10^{17} \, \text{cm}^{-3} \), and (d) \( L_p = 0.044 \, \mu \text{m}, \, N_p = 1.307 \times 10^{18} \, \text{cm}^{-3}, \, N_c = 2.138 \times 10^{17} \, \text{cm}^{-3} \). The 0.1-\( \mu \)m pocket n-MOSFETs chosen here are the same ones used in Figure 5.8. Global values of \( \kappa_p \) and \( \ell_{sat} \) were used for the model.

5.14 Rms errors of drain voltages of model \( I_D - V_{DS} \) curves to those of Medici \( I_D - V_{DS} \) curves at \( V_{GS} = 1 \, \text{V} \) of the 0.1-\( \mu \)m pocket n-MOSFETs located at the upper design-space boundary. Either the global or the single-device values of \( \kappa_p \) and \( \ell_{sat} \) were used to obtain the model \( I_D - V_{DS} \) curves. The z- and power-law formulas were used to calculate the global parameter values. The power-law formulas for \( \kappa_p \) and \( \ell_{sat} \) are provided in the subsequent subsection. \( K_{sat} = 2.3 \).

5.15 Comparisons of global values with single-device values of the parameter \( \ell_{sat} \) of 0.1-\( \mu \)m pocket n-MOSFETs. (a) \( \ell_{sat} \) vs. \( L_p \), (b) \( \ell_{sat} \) vs. \( N_p/N_{p0} \), (c) \( \ell_{sat} \) vs. \( N_c/N_{c0} \) and (d) \( \ell_{sat} \) vs. \( z/z_0 \). The global parameter values were calculated using power-law and z-formulas.

5.16 Model result of \( N_p \) vs. \( L_p \) at the upper design-space boundary for \( I_{on} = 0.2 \, \text{mA/\mu m} \) in the 0.1-\( \mu \)m pocket n-MOSFETs. The values of \( N_c \) are shown in Figure 5.1.(b).

5.17 Model predictions of design-space boundaries for \( I_{on} = 0.2 \, \text{mA/\mu m} \) in 0.1-\( \mu \)m pocket n-MOSFETs. (a) Two-dimensional design-space boundary \( N_p \) vs. \( L_p \) with \( N_c \) as a parameter, and (b) three-dimensional design-space boundary. The solid line in Figure (a) is the upper design-space boundary for \( I_{on} = 0.2 \, \text{mA/\mu m} \) in Figure 5.16.

5.18 Model predictions of design-space boundaries for different specifications of \( I_{on} \) in the 0.1-\( \mu \)m pocket n-MOSFET with \( N_c = 2 \times 10^{17} \, \text{cm}^{-3} \).

6.1 (a) The zone pointed by an arrow is the viable design space \( N_p \) vs. \( L_p \) for 0.1 \( \mu \)m channel-length pocket n-MOSFETs. The devices located within the unique viable design space meet all the design specifications of \( I_{on} \geq 0.2 \, \text{mA/\mu m}, \, I_{off} \leq 1 \, \text{nA/\mu m} \) and \( S = \frac{I_{on}/I_{off}}{S_{L/L}} \leq 40\%/10\% \). (b) \( N_c \) vs. \( L_p \) for all the devices located within the viable design space. This figure (b) is a repeat of Figure 3.7 (b) for the pocket devices satisfying both the sensitivity and the off-current specifications.
6.2 (a) The viable design space for 0.15 μm channel-length pocket n-MOSFET. The devices within the viable design space meet all the design specifications of $I_{on} \geq 0.2 \, mA/\mu m$, $I_{off} \leq 1 \, nA/\mu m$ and $S = \frac{I_{on}/I_{off}}{L/L_{T}} \leq 40%/10\%$. (b) $N_c$ vs. $L_p$ for all the devices within the viable design space.

6.3 (a) The viable design space for 0.2 μm channel-length pocket n-MOSFET. The devices within the viable design space meet all the design specifications of $I_{on} \geq 0.2 \, mA/\mu m$, $I_{off} \leq 1 \, nA/\mu m$ and $S = \frac{I_{on}/I_{off}}{L/L_{T}} \leq 40%/10\%$. (b) $N_c$ vs. $L_p$ for all the devices within the viable design space.

6.4 Potential contours for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$ inside 0.1-μm pocket n-MOSFETs located at (a), (f), (c) and (d) in Figure 6.1. Dashed lines are depletion-region edges.

6.5 Potential contours for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$ inside 0.1-μm pocket n-MOSFETs located at (b) and (g) in Figure 6.1. Dashed lines are depletion-region edges.

6.6 (a) Off-state current and (b) channel-length sensitivity of off-state current variations with pocket length for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$ of 0.1-μm pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ and $S = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$), respectively.

6.7 Threshold voltage variation with pocket length for $V_{DS} = 1 \, V$ of 0.1-μm pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ and $S = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$), respectively.

6.8 (a) Minima of surface potential distributions, i.e. the highest potential barriers to electron injection of the 0.1-μm pocket n-MOSFETs located at lower and upper design-space boundaries for $S = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$) and $I_{on} = 0.2 \, mA/\mu m$, respectively, and (b) locations at which the highest potential barriers exist. $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$.

6.9 On-state current variation with pocket length for $V_{GS} = 1 \, V$ and $V_{DS} = 1 \, V$ of 0.1-μm pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ and $S = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$), respectively.

6.10 (a) Transconductance and (b) electron effective velocity variations with pocket length for $V_{GS} = 1 \, V$ and $V_{DS} = 1 \, V$ of 0.1-μm pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ and $S = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$), respectively.
6.11 (a) $I_{off}$ vs. $L$ for $V_{GS} = 0$ V and $V_{DS} = 1$ V and (b) $V_{T}$ vs. $L$ for $V_{DS} = 1$ V of typical pocket n-MOSFETs with $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$, $N_p = 1.906 \times 10^{18} \text{ cm}^{-3}$, and $L_p = 0.024 \mu\text{m}$, and of conventional bulk n-MOSFETs with $N_A = 2.175 \times 10^{17} \text{ cm}^{-3}$. .... 223

6.12 Comparisons of the 0.1-$\mu$m pocket n-MOSFETs and a 0.1-$\mu$m conventional bulk n-MOSFET with $N_A = 1.098 \times 10^{18} \text{ cm}^{-3}$ regarding (a) sensitivity of off-state current to channel length and (b) off-state current for $V_{GS} = 0$ V and $V_{DS} = 1$ V. The devices have the same $I_{on} = 0.2$ mA/$\mu$m. 226

6.13 Comparisons of the 0.1-$\mu$m pocket n-MOSFETs and a 0.1-$\mu$m conventional bulk n-MOSFET with $N_A = 7.297 \times 10^{17} \text{ cm}^{-3}$ regarding (a) transconductance and (b) electron effective velocity for $V_{GS} = 1$ V and $V_{DS} = 1$ V. The devices have the same $I_{off} = 0.714$ nA/$\mu$m. 228

6.14 Comparisons of the 0.1-$\mu$m pocket n-MOSFETs and a 0.1-$\mu$m conventional bulk n-MOSFET with $N_A = 7.297 \times 10^{17} \text{ cm}^{-3}$ regarding (a) sensitivity of off-state current to channel length for $V_{GS} = 0$ V and $V_{DS} = 1$ V, and (b) on-state current for $V_{GS} = 1$ V and $V_{DS} = 1$ V. The devices have the same $I_{off} = 0.714$ nA/$\mu$m. 229

A.1 Comparison of the three different models A. B and C with respect to $I_D - V_{DS}$ characteristics at $V_{GS} = 1$ V of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu\text{m}$. 240

A.2 Comparisons of the three different models A. B and C with respect to (a) potential (b) electric field (c) electron density and (d) electron velocity profiles between source and drain at the Si-SiO$_2$ interface for $V_{DS} = 1$ V and $V_{GS} = 1$ V of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$, and $L_p = 0.024 \mu\text{m}$. $v_{sat} = 9.2 \times 10^6 \text{ cm/s}$. 241

A.3 Comparisons of the three different models A. B and C for the identical low-field mobility model, i.e., the generalized mobility curve mobility model with respect to (a) electron velocity profile at the Si-SiO$_2$ interface for $V_{DS} = 1$ V and $V_{GS} = 1$ V, and (b) $I - V$ characteristics at $V_{GS} = 1$ V of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu\text{m}$. 243

A.4 (a) Surface longitudinal electric field and (b) surface electron velocity vs. normalized distance from source for $V_{DS} = 1$ V and $V_{GS} = 1$ V, and (c) $I - V$ characteristics at $V_{GS} = 1$ V of the pocket n-MOSFETs with $L = 0.1, 1 \mu\text{m}$, $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu\text{m}$. 245
LIST OF TABLES

2.1 Parameter values of doping-density-dependent mobility model in Equation 2.29. \( T = 300^\circ K \). ........................................................... 51

3.1 Pocket MOSFET. Minimum and maximum values of surface potential of 0.1-\( \mu m \) and 1-\( \mu m \) pocket n-MOSFETs, and percent changes (numeral values in parentheses) of \( \psi_{s(max)} \) from the surface potential value of the long-channel model for \( V_{GS} = 0 \) V and \( V_{DS} = 0 \) V. .... 65

3.2 Conventional MOSFET. Minimum values of surface potential of 0.1-\( \mu m \) and 1-\( \mu m \) conventional bulk n-MOSFETs, and percent changes (numeral values in parentheses) of \( \psi_{s(min)} \) from the surface potential value of the long-channel model for \( V_{GS} = 0 \) V and \( V_{DS} = 0 \) V. .... 65

3.3 Pocket MOSFET. Minimum surface potentials of 0.1-\( \mu m \) and 1-\( \mu m \) pocket n-MOSFETs for \( V_{DS} = 0.1, 1 \) V and \( V_{GS} = 0 \) V. ................................. 68

3.4 Conventional MOSFET. Minimum surface potentials of 0.1-\( \mu m \) and 1-\( \mu m \) conventional bulk n-MOSFETs for \( V_{DS} = 0.1, 1 \) V and \( V_{GS} = 0 \) V. ................................. 68

3.5 Two-dimensional simulation results of the off-state current for \( V_{DS} = 1 \) V, the DIBL rate for \( V_{GS} = 0 \) V and \( V_{DS} = 0 \) V, the subthreshold slope for \( V_{DS} = 1 \) V of 0.1 and 1 \( \mu m \) channel-length pocket and conventional bulk n-MOSFETs. ........................................ 71

4.1 Structural parameters and \( z = L_{p}N_{p}N_{c} \) of the 0.1-\( \mu m \) pocket n-MOSFETs located at the lower design-space boundary, and the single-device values of \( \ell_{p0} \), \( \ell_{c0} \) and \( K_{p} \) of the pocket devices. \( V_{DS} = 0 \) V and \( V_{GS} = 0 \) V. .......................................................... 105

4.2 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{p0} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs. .......................................................... 109

4.3 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{c0} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs. .......................................................... 109

4.4 The coefficients of the second-degree polynomial of \( \log_{10} |K_{p}| \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs. .......................................................... 109

4.5 Structural parameters and \( z = L_{p}N_{p}N_{c} \) of 0.1-\( \mu m \) pocket n-MOSFETs located at the lower design-space boundary, and single-device values of \( \ell_{ps} \), \( \ell_{pD} \) and \( \ell_{c} \) of the pocket devices. \( V_{DS} = 1 \) V and \( V_{GS} = 0 \) V. .......................................................... 122

4.6 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{ps} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs. .......................................................... 125
4.7 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{PD} \) of two types of 0.1-\( \mu \)m pocket n-MOSFETs. ................................. 127
4.8 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{c} \) of two types of 0.1-\( \mu \)m pocket n-MOSFETs. ................................. 127
4.9 Rms errors of the global values to the single-device values of \( \ell_{ps} \), \( \ell_{PD} \) and \( \ell_{c} \) for the twelve 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary. ................................. 132
4.10 Structural parameters of 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary and single-device values of \( L_{eff} \) of the pocket devices. \( V_{DS} = 1 \) V and \( V_{GS} = 0 \) V. ................................. 139
4.11 The coefficients of the second-degree polynomial of \( \log_{10} L_{eff} \) of two types of 0.1-\( \mu \)m pocket n-MOSFETs. ................................. 139
4.12 Model values of \( I_{gs} \) compared to Medici data for the 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary. ................................. 143
4.13 Structural parameters of 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary and single-device values of \( L_{eff} \) of the pocket devices to be used for developing an analytical formula for \( L_{eff} \) of the simple model. \( V_{DS} = 1 \) V and \( V_{GS} = 0 \) V. ................................. 146
4.14 Structural parameters of 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary and single-device values of \( \zeta \) of the pocket devices to be used for developing an analytical formula for \( \zeta \). \( V_{DS} = 0 \) V and \( V_{GS} = 0 \) V. ................................. 150
5.1 Transconductance [mS/\( \mu \)m], effective electron velocity [cm/s] and on-state current [A/\( \mu \)m] at \( V_{GS} = 1 \) V and \( V_{DS} = 1 \) V of the 0.1- and 1-\( \mu \)m pocket n-MOSFETs with \( t_{ox} = 4 \) nm, \( L_{p} = 0.024 \) \( \mu \)m, \( N_{p} = 2.713 \times 10^{18} \) cm\(^{-3} \), \( N_{c} = 2.175 \times 10^{17} \) cm\(^{-3} \) and \( r_{j} = 0.06 \) \( \mu \)m. ................................. 171
5.2 Structural parameters of the 0.1-\( \mu \)m pocket n-MOSFETs located at the upper design-space boundary in Figure 5.1. and the best single-device values of \( \kappa_{p} \) and \( \ell_{sat} \) of the pocket devices. \( K_{sat} = 2.3 \) and \( V_{GS} = 1 \) V. ................................. 193
5.3 The coefficients of the second-degree polynomial of \( \log_{10} \kappa_{p} \) of the two types of the 0.1-\( \mu \)m pocket n-MOSFETs. ................................. 196
5.4 The coefficients of the second-degree polynomial of \( \log_{10} \ell_{sat} \) of the two types of the 0.1-\( \mu \)m pocket n-MOSFETs. ................................. 196
5.5 Rms errors of the global values to the single-device values of \( \kappa_{p} \) and \( \ell_{sat} \) for the twelve 0.1-\( \mu \)m pocket n-MOSFETs located at the upper design-space boundary. ................................. 201
6.1 Comparisons of a 0.1-μm pocket n-MOSFET located in the middle of the viable design space in Figure 6.1 with 0.1-μm conventional bulk n-MOSFETs. "Pocket" is the 0.1-μm pocket device with $N_p = 1.627 \times 10^{18}$ cm$^{-3}$, $L_p = 0.032$ μm and $N_c = 1.825 \times 10^{17}$ cm$^{-3}$. "Convent 1" is the 0.1-μm conventional device with $N_A = 9.1 \times 10^{17}$ cm$^{-3}$ and the same $I_{on}$ as that of the pocket device. "Convent 2" is the 0.1-μm conventional device with $N_A = 8.925 \times 10^{17}$ cm$^{-3}$ and the same $I_{off}$ as that of the pocket device.

A.1 Transconductance and electron effective velocity for $V_{DS} = 1$ V and $V_{GS} = 1$ V of the 0.1- and 1-μm pocket n-MOSFETs with $t_{ox} = 4$ nm, $L_p = 0.024$ μm, $N_p = 2.713 \times 10^{18}$ cm$^{-3}$, $N_c = 2.175 \times 10^{17}$ cm$^{-3}$ and $r_j = 0.06$ μm.

A.2 Transconductance [$\mu S/\mu m$], electron effective velocity [cm/s] and on-state current [A/μm] at $V_{GS} = 1$ V and $V_{DS} = 1$ V of the 0.1- and 1-μm pocket n-MOSFETs with $t_{ox} = 4$ nm, $L_p = 0.024$ μm, $N_p = 2.713 \times 10^{18}$ cm$^{-3}$, $N_c = 2.175 \times 10^{17}$ cm$^{-3}$ and $r_j = 0.06$ μm.

A.3 Transconductance, electron effective velocity and on-state current at $V_{GS} = 1$ V and $V_{DS} = 1$ V of the 0.1-μm pocket n-MOSFET with $t_{ox} = 4$ nm, $L_p = 0.024$ μm, $N_p = 2.713 \times 10^{18}$ cm$^{-3}$, $N_c = 2.175 \times 10^{17}$ cm$^{-3}$ and $r_j = 0.06$ μm. The model A and the generalized mobility curve mobility model were used for the two-dimensional device simulation.
ABSTRACT

Laterally non-uniformly doped deep-submicron pocket n-MOSFETs satisfying specifications of off-state (leakage) current, on-state (drive) current, sensitivity of off-state current to channel length and 1 V power-supply voltage have been designed for low-voltage low-power applications. To determine a viable range of the deep-submicron pocket n-MOSFET structural parameters - the dopant concentration at the center region ($N_c$), the dopant concentration at the pocket region ($N_p$) and the length of the pocket region ($L_p$), unique viable design spaces locating the deep-submicron devices that meet all the device specifications have been constructed for the 0.1, 0.15 and 0.2 μm channel-length pocket n-MOSFETs, using computer algorithms developed on the basis of device physics and two-dimensional device simulation results, and implemented in the programming language of the two-dimensional device simulator, Medici. For known $N_c$ vs. $L_p$, the pocket n-MOSFETs for low-power applications are located in an upper area of higher $N_p$ vs. $L_p$ of the viable design space while the devices for high-performance applications are located in a lower area of lower $N_p$ vs. $L_p$ of the viable design space. Well-designed deep-submicron pocket n-MOSFETs prove to be promising candidates to improve short-channel effects as well as switching performance in comparing the 0.1-μm pocket n-MOSFETs located within the viable design space to 0.1-μm conventional bulk n-MOSFETs selected to meet the same specifications. The 0.1-μm
pocket n-MOSFETs located within the viable design space can be partitioned into two types of pocket devices based on gate controllability of channel- and depletion-layer charges.

Analytical models for subthreshold and above-threshold currents in the deep-submicron pocket n-MOSFETs have been developed for the first time to generate the off-state and the on-state currents, and the design-space boundaries. The models are based on solutions of the drift-diffusion current transport and the 1-D Poisson’s equations, the charge sheet approximation, subthreshold surface potential models based on solutions of the quasi-two-dimensional Poisson’s equation, a quasi-two-dimensional velocity saturation model, and analytical formulas for model parameters as functions of the structural parameters of the pocket n-MOSFETs. Mobility models used for the modeling of the currents include the lateral-field-dependent mobility, the transverse-field-dependent mobility and the doping density-dependent mobility at low transverse fields. The analytical models provide explicit relations between process, device and model parameters of the deep-submicron pocket n-MOSFETs, reduce time and cost of the two-dimensional device simulation, and deal with device structures in a simple way. Some algorithms developed for generating $I_D-V_{DS}$ characteristics and constructing the design-space boundaries for the on- and the off-state currents in the deep-submicron pocket n-MOSFETs are described in this dissertation.
CHAPTER 1

INTRODUCTION

Modern electronics and computer industry demands low-power, high-density, and high-speed chips. The Metal-Oxide-Semiconductor Field-Effect Transistor (MOSFET) has been an excellent semiconductor device to meet the demand. Nowadays scaling of MOSFETs down to the deep-submicrometer or nanometer (sub-100 nm) regime has made possible much denser and faster chips than before. However, as MOSFETs are scaled down, the power-supply voltage, which has a direct effect on circuit switching performance, i.e., speed, needs to be lowered in order to satisfy reliability requirements [1] and to lower both active and standby power dissipation. The lower limit of the power-supply voltage is determined by acceptable circuit switching performance. The following is an expression for the sum of the active and standby power dissipation in CMOS circuits [1].

\[
P_{\text{total}} = I_{\text{off}} V_{DD} + K C_L f V_{DD}^2
\]  

(1.1)

where \( V_{DD} \) is the power-supply voltage, \( I_{\text{off}} \) is the off-state current, \( K \) is the switching factor, \( C_L \) is the total load capacitance, and \( f \) is the clock frequency. The first and second terms in Equation 1.1 are the standby and active power consumption, respectively. Short-channel effects [1]–[26], e.g., surface DIBL (Drain-Induced Barrier Lowering), bulk punchthrough, threshold voltage roll-off, etc., of deep-submicron
MOSFETs can cause the standby power to significantly increase due to a substantial increase in the off-state current. The short-channel effects that are not predicted by long-channel models have been obstacles to the design of the deep-submicron MOSFETs because the design of the MOSFETs satisfying specifications of subthreshold leakage current (off-state current), sensitivity of off-state current to channel length or drain voltage, and circuit noise immunity [32] pays a price of degradation of device performance, which is called a design trade-off.

Drain engineering or lateral channel engineering utilizing halo or pocket implant [3]–[20] in deep-submicron MOSFETs even in sub-100 nm gate-length MOSFETs [2], [5] has been reported to be fairly effective in suppressing short-channel effects such as threshold voltage roll-off and bulk punchthrough, and body effect [33] since S. Ogura et al. [6] introduced in 1982 a half micron n-MOSFET with double-implanted LDD (Lightly Doped Drain) and self-aligned p-type pockets around the lightly doped drain and source. Historically, several studies on the bulk punchthrough [26]–[30], [34], [35] as a short-channel effect had been conducted through two-dimensional numerical analyses and experiments. The significant increase in off-state current in deep-submicron MOSFETs under bulk punchthrough condition can cause the standby power to become substantial. Recent experimental results [6], [7], [11]–[14] show that the halo or pocket implants in deep-submicron MOSFETs are effective in preventing the bulk punchthrough. However, the pocket
Implantation can cause on-state current to degrade [18] because the carrier mobility degrades at the heavily doped pocket regions and the threshold voltage of the pocket MOSFET becomes higher than the MOSFET without the pocket implant.

Advanced device-manufacturing technologies, e.g., Large-Angle-Tilt Implantation (LATI) [11] and Self-aligned Pocket Implantation (SPI) [15] have made possible optimization of the halo or pocket implant by controlling the dose, depth, and extension of the halo implant. CMOSFETs utilizing the pocket implantation have been used for VLSI or ULSI DRAM (Dynamic Random Access Memory), microprocessor, and low-power LSI for battery-operated portable computers and communication equipment [12]–[16].

The goals of the work for this dissertation are to design deep-submicron pocket n-MOSFETs satisfying device specifications by controlling short-channel effects while maintaining acceptable device performance, to establish a viable design space that is a portion of a design space locating the pocket devices meeting all the required design specifications of the on-state current, the off-state current and the sensitivity of off-state current to channel length, to develop models for the off-state and on-state currents in the deep-submicron pocket n-MOSFETs, which can provide a solution to the problem of time-consuming tasks of two-dimensional device simulation and be used to generate the design-space boundaries for the off- and the on-state currents, and to explore advantages of the 0.1-μm pocket n-MOSFETs located within the viable design space over 0.1-μm conventional bulk n-MOSFETs. Well-designed
deep-submicron pocket n-MOSFETs can be better than deep-submicron conventional bulk n-MOSFETs in short-channel immunity as well as in performance.

Figure 1.1 illustrates structures of the conventional bulk and pocket n-MOSFETs. Heavily doped pocket regions are located around the source and the drain, whereas a lightly doped center region is located at the center of the pocket MOSFET. The pocket device is idealized for studying its DC characteristics only. This idealized structure completely surrounds the junctions with the heavily doped pocket regions, and so is not useful for optimizing the junction capacitance for AC studies.

The design parameters of the pocket n-MOSFET are defined below.

- \( L_p \) = length of pocket region
- \( L = \) channel length \( (=2L_p + L_c) \) or \( L_c = \) length of center region \( (=L - 2L_p) \)
- \( N_p = \) pocket dopant concentration
- \( N_c = \) center dopant concentration

The channel length \( L \) means the distance between the edges of the \( n^+ \)-regions of the source and drain or the metallurgical source-to-drain separation of the MOSFET. The other structural parameters have been selected to be typical of an advanced low-power low-voltage technology: \( t_{ox} \) (oxide thickness) = 4 \( nm \), \( r_j \) (junction depth) = 0.06 \( \mu m \), and \( \varphi_{w(npoly)} \) (work function of \( n^+ \)-poly contacts for gate, source and drain) = 4.05 \( eV \). The substrate has a neutral contact of which the work function is the same as that of the substrate material.
Figure 1.1: (a) Uniformly doped n-MOSFET and (b) pocket n-MOSFET.
The design specifications required for the deep-submicron pocket n-MOSFETs are as follows:

- $I_{\text{off}}$ (off-state current) $\leq 1 \, \text{nA/\mu m}$ for $V_{GS} = 0 \, \text{V}$ and $V_{DS} = 1 \, \text{V}$.

- $\frac{\Delta I_{\text{off}}}{\Delta L/L}$ (sensitivity of $I_{\text{off}}$ to $L$) $\leq \frac{40\%}{10\%}$ $= 4$ for $V_{GS} = 0 \, \text{V}$ and $V_{DS} = 1 \, \text{V}$. $\frac{\Delta I_{\text{off}}}{\Delta L/L} = \frac{40\%}{10\%}$ at the design-space boundary for $I_{\text{off}}$, i.e., $I_{\text{off}}(L) = 0.714 \, \text{nA/\mu m}$ and $I_{\text{off}}(0.9L) = 1 \, \text{nA/\mu m}$.

- $I_{\text{on}}$ (on-state current) $\geq 0.2 \, \text{mA/\mu m}$ for $V_{GS} = 1 \, \text{V}$ and $V_{DS} = 1 \, \text{V}$. $I_{\text{on}}(L) = 0.2 \, \text{mA/\mu m}$ at the design-space boundary for $I_{\text{on}}$.

- $V_{DD}$ (power-supply voltage) $= 1 \, \text{V}$.

The specifications of $I_{\text{off}}$ and $I_{\text{on}}$ are requirements of the digital memory and logic technology, and depend on applications. Some typical values of $I_{\text{off}}$ and $I_{\text{on}}$ can be found in [19] and [38]. $I_{\text{off}} = 1 \, \text{nA/\mu m}$ is the common maximum leakage at room temperature in the semiconductor industry. $I_{\text{on}} = 0.2 \, \text{mA/\mu m}$ is considered as a nominal value at room temperature for $V_{DD} = 1 \, \text{V}$. Based on the data in [38], $I_{\text{off}} = 1 \, \text{nA/\mu m}$ is a specification for high-performance applications, while $I_{\text{on}} = 0.2 \, \text{mA/\mu m}$ is the one for low-power applications. The channel-length sensitivity of off-state current is evaluated as a change in the off-state current from $I_{\text{off}}(L)$ to $I_{\text{off}}(0.9L) = 1 \, \text{nA/\mu m}$ (the worst-case specification of the off-state current) due to 10% reduction of channel length $L$. The channel-length sensitivity specification.
is important for uniformity of manufactured devices where \( L \) is controlled only to \( L - \Delta L \). The tolerance in channel length has been taken as a nominal 10% change, but this value obviously depends on the manufacturer's control of his own process. It is therefore only a representative number. Likewise, the selected value of a change of 40% in \( I_{\text{off}} \) in response to a change of 10% in channel length is only representative, and the actual tolerance of any specific circuit might be different depending on the application (e.g., battery operated or not) and on the design style (e.g., dynamic vs. static CMOS). In this dissertation the development of algorithms considering sensitivity as a design specification, even if it is not specific to a particular situation, is important in finding a unique design-space boundary for \( I_{\text{off}} \). The algorithms developed here can be applied for different choices of sensitivity, and of on- and off-currents.

To design the deep-submicron pocket n-MOSFETs meeting all the required design specifications from two-dimensional device simulations, the pocket devices satisfying the boundary specification of \( \frac{\Delta I_{\text{off}}/I_{\text{off}}}{\Delta L/L} = \frac{40\%}{10\%} \) will first be explored using the computer algorithms developed on the basis of device physics and two-dimensional device simulation results to automatically and accurately find the values of \( N_p \) and \( N_c \) corresponding to each of given \( L_p \) values for fixed values of \( L, t_{ox} \) and \( r_f \). Then, the values of \( N_p \) and \( N_c \) with which the deep-submicron pocket devices meet the boundary specification of the sensitivity of off-state current to channel length make up the unique lower design-space boundary for the channel-length
sensitivity of off-state current and the off-state current on a plot of $N_p$ vs. $L_p$ of the deep-submicron pocket devices. In order to explore the deep-submicron pocket n-MOSFETs satisfying the boundary specification of $I_{on} = 0.2 \, mA/\mu m$, a similar algorithm is used to find another plot of $N_p$ vs. $L_p$ for fixed values of $L$, $t_{ox}$ and $r_j$, and the same $N_c$ vs. $L_p$ as the lower design-space boundary for the deep-submicron pocket devices. The values of $N_p$ and $N_c$ with which the deep-submicron pocket devices satisfy the boundary specification of the on-state current make up the unique upper design-space boundary for the on-state current on the plot of $N_p$ vs. $L_p$. The zone constructed between the lower and the upper design-space boundaries is the unique viable design space locating the deep-submicron pocket n-MOSFETs satisfying all the design specifications. Amongst all the devices meeting the specifications, those devices located within the viable design space far from the design-space boundaries are the most immune to uncontrolled departures of device parameter values from the design specifications. The deep-submicron pocket n-MOSFETs as well-designed devices located within the viable design space will be selected and compared with deep-submicron conventional bulk n-MOSFETs in short-channel immunity and performance.

Although pocket or halo doping significantly improves short-channel effects in deep-submicron or sub-100 nm MOSFETs, the maximum pocket doping is limited by hot-carrier effects such as hot-electron degradation and avalanche breakdown, i.e., an increase in the impact ionization rate [7], [21], and gate-induced drain
leakage (GIDL) [2], [16], [26] due to quantum mechanical band-to-band tunneling induced by a transverse field in the gate-drain overlap region. However, to achieve GIDL in the overlap region, the gate-to-drain voltage must be so high as to cause band overlap in the drain. When the drain is heavily doped and gate-to-drain voltage is the maximum $1 \text{ V}$ ($\sim$ the band gap), no GIDL can occur. It is primarily a problem in the LDD type of structure, and not in our structure. Based on recent reports on the fabrication of a 0.04-$\mu m$ gate-length n-MOSFET [22] and experimental data on pass transistor designs using pocket implant [16], we make assumptions that hot-carrier effects are negligible in designing deep-submicron ($L = 0.1 - 0.2 \mu m$) pocket n-MOSFETs for the maximum $1 \text{ V}$ power-supply voltage. A leakage current that results from the generation of free carriers in space-charge regions is ignored. Parasitic drain/source resistance is also assumed to be zero. Sub-100 nm channel-length n-MOSFETs can be in effective velocity overshoot, which is a controversial issue [22], [73]. Velocity overshoot effects are excluded in constructing the viable design spaces for deep-submicron pocket n-MOSFETs in this dissertation. The error due to this neglect is estimated to be a secondary factor in the Appendix A.

In Chapter 2, an effective mobility model to be used for the modeling of currents in n-MOSFETs is described. The effective mobility model includes the lateral-field-dependent mobility, the transverse-field-dependent mobility and the doping density-dependent mobility at low transverse fields. Relations between physical
parameters and the effective transverse electric field are presented. A novel and simple method for extracting the effective transverse-field-dependent mobility from the drift-diffusion current transport equation and some Medici [39] data is provided, and the extracted mobility is compared with model results. In Chapter 3, computer algorithms developed on the basis of device physics and two-dimensional device simulation results to automatically and accurately find the unique lower design-space boundary for the off-state current and the sensitivity of off-state current to channel length of 0.1-$\mu m$ pocket n-MOSFETs are presented, and a qualitative model for the pocket devices are provided. Characteristics of the 0.1-$\mu m$ pocket n-MOSFETs biased to weak inversion are discussed in this chapter. In Chapter 4 a model for the subthreshold current in deep-submicron pocket n-MOSFETs is provided. The current model together with short-channel subthreshold surface potential models is used to generate off-state currents in the 0.1-$\mu m$ pocket n-MOSFETs. Some analytical and empirical formulas of characteristic or decay lengths for the short-channel surface potential models are introduced. The subthreshold surface potential model is used to generate potential profiles between source and drain of the pocket devices, and to predict a DIBL effect and a barrier lowering effect due to the scaling down of channel length of the pocket n-MOSFETs. Model predictions of lower design-space boundaries for $I_{off}$ are also presented. Chapter 5 presents the unique upper design-space boundary for the on-state current in the 0.1-$\mu m$ pocket n-MOSFETs, and a model for the above-threshold current in the deep-submicron pocket n-MOSFETs.
The current model, including the drift-diffusion current transport equation, the charge-sheet model, a quasi-two-dimensional velocity saturation model and analytical formulas for model parameters, is used to generate $I_D - V_{DS}$ characteristics at $V_{GS} = 1 \, V$ and design-space boundaries for $I_{on}$ in the 0.1-$\mu$m pocket n-MOSFETs. Characteristics of the 0.1-$\mu$m pocket n-MOSFETs biased to strong inversion are discussed. In Chapter 6, the unique viable design spaces locating deep-submicron devices that meet all the design specifications are constructed for 0.1, 0.15 and 0.2 $\mu$m channel-length pocket n-MOSFETs, using the computer algorithms developed for finding automatically and accurately the design-space boundaries for $I_{off}$ ($= 0.714 \, nA/\mu m$), $\frac{\Delta I_{off}/I_{off}}{\Delta L/L} (= 40\%/10\%)$, and $I_{on} (= 0.2 \, mA/\mu m)$. The 0.1-$\mu$m pocket MOSFETs located within the unique viable design space are compared with 0.1-$\mu$m conventional bulk n-MOSFETs selected to meet the same specifications in terms of short-channel immunity and performance. Finally, conclusions on this work are drawn in Chapter 7.
CHAPTER 2

MOBILITY MODELS

A physics-based effective mobility model is studied, and will be used for the modeling of $I_{on}$ and $I_{off}$ in deep-submicron pocket n-MOSFETs in the subsequent chapters. The mobility model includes the lateral-field-dependent mobility, the transverse-field-dependent mobility and the doping density-dependent mobility at low transverse fields in subthreshold regime. In order to reduce short-channel effects, the deep-submicron MOSFETs require thinner oxide thickness and heavier channel dopant concentration, which result in higher transverse electric field and lower effective mobility. In Section 2.1, relations between the effective transverse electric field and physical parameters are presented. The electron lateral-field-dependent mobility model [41]-[43] and a semi-empirical model [44], [45] for the electron effective transverse-field-dependent mobility in strong inversion layers of conventional, uniformly doped n-MOSFETs are introduced. The semi-empirical mobility model contains fitting parameters for scattering mechanisms that are selected to accurately reproduce the experimentally measured effective transverse-field-dependent mobility versus the effective transverse electric field [46], [47]. In Section 2.2, an empirical relation [48] for electron mobility in bulk silicon is modified
to formulate the electron effective transverse-field-dependent mobility model continuous from the above-threshold regime to the subthreshold regime. In Section 2.3, the effective mobility at low lateral electric field, i.e., the effective transverse-field-dependent mobility is extracted from Medici using the drift-diffusion current transport equation and some Medici [39] data by a novel and simple mobility extraction method, and compared to the model result. This chapter is summarized and conclusions are drawn in Section 2.4.

2.1 Physics-Based Effective Mobility Model

2.1.1 Relations between Effective Transverse Electric Field and Physical Parameters

As the gate voltage is increased, the inversion-layer mobility decreases because the surface scattering becomes substantial. The mobility degradation at high transverse fields has been observed from the universal relationship [2], [46], [47] between the inversion-layer mobility (the effective transverse-field-dependent mobility $\mu_{\text{eff0}}$) and the effective transverse electric field $E_{\text{eff}}$ [44], [45], [49]. The effective transverse electric field is an important parameter in the modeling of the universality of the inversion-layer mobility. $E_{\text{eff}}$ depends directly on the inversion- and the depletion-layer charge densities, which are dependent on the applied gate voltage, the oxide thickness and the surface bulk doping density.
The effective transverse or normal electric field, \( E_{\text{eff}} \), is defined as

\[
E_{\text{eff}} \equiv \frac{E_s + E_D}{2} = \frac{qN_I}{2\varepsilon_s} + \frac{qN_A W_D}{\varepsilon_s} = E_N + E_D \tag{2.1}
\]

where

\[
E_s = \frac{qN_I}{\varepsilon_s} + \frac{qN_A W_D}{\varepsilon_s} \tag{2.2}
\]

\[
E_D = \frac{qN_A W_D}{\varepsilon_s} \tag{2.3}
\]

\[
W_D = L_D \sqrt{2} \sqrt{3\phi_s - 1} \tag{2.4}
\]

\[
\beta = \frac{1}{V_{th}} = \frac{q}{kT} \tag{2.5}
\]

\[
L_D = \sqrt{\frac{\varepsilon_s V_{th}}{qN_A}} \tag{2.6}
\]

\[
\phi_s = \frac{E_i(0) - E_i(\infty)}{-q} \tag{2.7}
\]

\( E_s \) is the transverse electric field at surface of silicon, \( E_D \) is the transverse electric field at depletion-layer region, \( E_N \) is the transverse electric field in inversion layer, \( N_I \) is the electron density per unit area in inversion layer, \( W_D \) is the depth of depletion-layer region, \( N_A \) is the bulk dopant concentration, \( q = 1.602 \times 10^{-19} \text{ C} \) is the electronic charge, \( V_{th} \) is the thermal voltage (= 0.025851 V at 300°K), \( L_D \) is the extrinsic Debye length, \( \varepsilon_s = (11.7)(8.854 \times 10^{-14}) \text{ F/cm} \) is the permittivity of Si, and \( \phi_s \) is the surface potential related to band bending on the energy band diagram. \( E_i(0) \) and \( E_i(\infty) \) are the intrinsic Fermi levels at the surface of silicon and in the silicon bulk, respectively.
To find the variations of $\phi_s$ with $N_A$, $\varepsilon_{\text{eff}}$ with $N_A$ and $N_I$ with $\varepsilon_{\text{eff}}$ for $V_{GS} = 1$ and 0 V, we review important equations formulated from the long-channel MOSFET theory [23], [50], [51]. Using the coordinate system in Figure 1.1, solve the following one-dimensional Poisson's equation for $\varepsilon_s$ at the Si surface ($y = 0$) subject to the boundary conditions of $\varepsilon_y(\infty) = 0$ and $\phi(\infty) = 0$ at the Si bulk ($y = \infty$) by integrating the Poisson's equation from the bulk towards the surface.

$$\frac{d^2\phi}{dy^2} = \frac{-d\varepsilon_y}{dy} = -\frac{q}{\varepsilon_s}[-N_A - n(y) + p(y)]$$ (2.8)

where

$$n(y) = \frac{n_i^2}{N_A}e^{\phi(y)}$$ (2.9)

$$p(y) = N_Ae^{-\phi(y)}$$ (2.10)

Using the solution of $\varepsilon_s$, and the continuity of electric displacement or Gauss's law at the Si-SiO$_2$ interface,

$$\varepsilon_{ox}\varepsilon_{ox} = \varepsilon_s\varepsilon_s$$ (2.11)

where

$$\varepsilon_{ox} = \frac{\phi_G - \phi_s}{t_{ox}}$$ (2.12)

we can derive the following equation.

$$C_{ox}(\phi_G - \phi_s) = qN_ALD\sqrt{2}\sqrt{\beta}\phi_s - 1 + \left(\frac{n_i}{N_A}\right)^2 e^{3\phi_s}$$ (2.13)
where

\[ C_{ox} = \frac{\varepsilon_{ox}}{t_{ox}} \quad (2.14) \]

\[ \phi_G = V_{GB} - V_{FB} \quad (2.15) \]

\[ V_{FB} = -\psi_{NPOLY} - V_{th} \ln \left( \frac{N_A}{n_i} \right) - \frac{qN_f}{C_{ox}} - \int_0^{t_{ox}} \frac{\rho_{ox}(y)}{t_{ox}} dy \quad [33] \quad (2.16) \]

\[ \psi_{NPOLY} = \chi(Si) + \frac{E_G}{2q} + \frac{V_{th}}{2} \ln \left( \frac{N_C}{N_V} \right) - \phi_{w(npoly)} \quad [39] \quad (2.17) \]

\( C_{ox} \) is the oxide capacitance. \( \varepsilon_{ox} = (3.9)(8.854 \times 10^{-14}) \) \( F/cm \) is the permittivity of SiO2. \( V_{GB} = V_{GS} \) is the gate-to-substrate voltage, \( V_{FB} \) is the flat band voltage that is a voltage applied between the gate and the silicon bulk to achieve the flat-band condition in the silicon. \( \psi_{NPOLY} \) is the surface potential at the gate Schottky contact with the zero reference at the mid-bandgap, i.e., the intrinsic Fermi level.

\( n_i = 6.1737 \times 10^9 \ cm^{-3} \quad [39] \) is the intrinsic carrier concentration of Si at 300°K, \( \chi(Si) = 4.05 \ V \quad [33] \) is the electron affinity potential of Si with the zero reference at the vacuum level, \( E_G = 1.124 \ eV \quad [33] \) is the energy gap of Si at 300°K, \( N_C = 2.8 \times 10^{19} \ cm^{-3} \) is the effective density of states in the conduction band of Si at 300°K.

\( N_V = 1.04 \times 10^{19} \ cm^{-3} \) is the effective density of states in the valence band of Si at 300°K, and \( \phi_{w(npoly)} = 4.05 \ V \) is the work function potential of the \( n^+ \)-poly gate contact with the zero reference at the vacuum level. Note that the substrate has a neutral contact of which the work function is the same as that of the substrate material, i.e., the \( p \)-type silicon. We assume that the fixed interface charge density, \( qN_f \), and the oxide volume charge density, \( \rho_{ox}(y) \), in the oxide layer are zeros.
Then, $V_{FB}$ is just equal to the difference in the work functions of the $n^+$-poly gate contact and the $p$-type silicon.

For heavily doped silicon n-MOSFETs, the above analyses can include quantum mechanical effects of the electron motion in inversion layers by using a model for the intrinsic carrier concentration at the Si-SiO$_2$ interface proposed by M. J. Dort et al. [52]. The quantization of the electron motion causes a splitting [53] of energy levels in the conduction band and displacement of the electron distribution from the Si-SiO$_2$ interface. The bandgap widening due to the two quantum mechanical effects was modeled in [52], and the increase in the bandgap causes an exponential decrease in the intrinsic carrier concentration of the quantum mechanical model from the intrinsic carrier concentration of the conventional model.

The effective transverse field $E_{\text{eff}}$ is a function of the surface potential $\phi_s$, which depends strongly on inversion- and depletion-layer charges and should meet the requirement of Equation 2.11. Solving Equation 2.13 for $\phi_s$ for a given substrate doping density $N_A$ using a numerical method yields the dependence of $\phi_s$ on $N_A$ of long-channel n-MOSFETs for $V_{GS} = 0$ and 1 V with $t_{ox}$ as a parameter in Figure 2.1. As $N_A$ is increased at small values of $N_A$ for fixed values of $t_{ox}$ and $V_{GS}$, $\phi_s$ increases because the inversion-layer charge density decreases and the depletion-layer charge density increases. Note that the inversion-layer charge density is greater than the depletion-layer charge density at the small values of $N_A$. On
the other hand, as $N_A$ is increased at large values of $N_A$, $\phi_s$ decreases because the inversion layer is not formed and the depletion-layer charge density increases.

\begin{figure}[h]
\centering
\includegraphics[width=\textwidth]{figure2_1.png}
\caption{Surface potential $\phi_s$ vs. bulk doping density $N_A$ for two different oxide thicknesses $t_{ox} = 4$ and 10 nm, temperature $T = 300^\circ K$, and (a) $V_{GS} = 0$ V and (b) $V_{GS} = 1$ V. The results were obtained by solving Equation 2.13.}
\end{figure}

In order to evaluate $\mathcal{E}_{eff}$ at a given $\phi_s$, the electron density, $N_I$, in the inversion layer of the long-channel n-MOSFET must be found. Using the definition of $\mathcal{E}_{eff}$ and the Maxwell-Boltzmann statistics, the electron density can be expressed by

\begin{equation}
N_I = n_s W_I = n_s \frac{V_{th}}{\mathcal{E}_{eff}} = \left( \frac{n_I^2 e^{3\phi_s}}{N_A} \right) \left( \frac{V_{th}}{qN_s + qN_A W_D} \right)
\end{equation}

where $W_I$ is the inversion-layer depth under the gate. Solving the resulting quadratic equation of Equation 2.18 for $N_I$, we obtain

\begin{align}
N_I &= -N_A W_D + \sqrt{(N_A W_D)^2 + 2(n_I L_D)^2 e^{3\phi_s}} \\
&= -N_A W_D + \frac{\varepsilon_s \mathcal{E}_s}{q} \quad (2.19)
\end{align}
Equation 2.19 or 2.20 is a result obtained after applying Gauss's law to a Gaussian box formed from the silicon surface ($\mathcal{E}(0) = \mathcal{E}_s$) to the silicon bulk ($\mathcal{E}(\infty) = 0$). Equation 2.19 can be used to calculate the electron density of the MOSFET biased to strong inversion, i.e., $(N_AW_D)^2 \ll 2(n_iL_D)^2e^{\phi_s}$. but the form of the expression needs to be changed for the MOSFET biased to weak inversion or the onset of strong inversion. i.e., $(N_AW_D)^2 \gg 2(n_iL_D)^2e^{\phi_s}$ to prevent possible numerical errors. If \( \frac{2(n_iL_D)^2e^{\phi_s}}{(N_AW_D)^2} < 10^{-6} \) (at weak inversion or near the onset of strong inversion), use the binomial series expansion to obtain the following approximate expression for \( N_I \):

\[
N_I \approx \frac{(n_iL_D)^2}{N_AW_D}e^{\phi_s} \tag{2.21}
\]

From Equations 2.1, 2.4, 2.19 and 2.21 we can calculate the effective transverse field \( \mathcal{E}_{\text{eff}} \), the field \( \mathcal{E}_N \) due to the electron charge in inversion layer, and the field \( \mathcal{E}_D \) due to the depletion-layer charge. Figures 2.2 and 2.3 show the variations of the three transverse electric fields with the bulk doping density for \( t_{ox} = 4 \) nm, and \( V_{GS} = 0 \) and \( 1 \) V, respectively. The transverse electric field due to the electron charge in inversion layer is dominant at low bulk doping levels while the transverse electric field due to the depletion-layer charge becomes dominant as the bulk doping level becomes higher. e.g., \( N_A > 3 \times 10^{17} \ cm^{-3} \) for \( V_{GS} = 1 \) V. Figures 2.4 and 2.5 show that the effective transverse electric field depends on the oxide thickness, the bulk doping density and the gate voltage. As the oxide thickness is decreased, both the oxide field and the Si surface field increase at a given gate voltage, and
Figure 2.2: Transverse electric fields vs. bulk doping density for $t_{ox} = 4 \text{ nm}$ and $V_{GS} = 0 \text{ V}$. $E_{\text{eff}} = E_N + E_D$.

Figure 2.3: Transverse electric fields vs. bulk doping density for $t_{ox} = 4 \text{ nm}$ and $V_{GS} = 1 \text{ V}$. $E_{\text{eff}} = E_N + E_D$.

Figure 2.4: The dependence of $E_{\text{eff}}$ on both $t_{ox}$ and $N_A$ for $V_{GS} = 0 \text{ V}$.

Figure 2.5: The dependence of $E_{\text{eff}}$ on both $t_{ox}$ and $N_A$ for $V_{GS} = 1 \text{ V}$. 
thereby $E_{\text{eff}}$ increases as $t_{ox}$ is decreased since $E_{\text{eff}}$ is directly proportional to the Si surface field $E_s$. At a large value of $N_A (\geq 10^{17}$ or $10^{18}$ cm$^{-3}$), the electron charge density in channel region becomes smaller because of a higher threshold voltage. $E_{\text{eff}}$ increases markedly as $N_A$ is increased further since the depletion-layer charge density is directly proportional to $\sqrt{N_A}$, whereas the electron charge density in channel is directly proportional to $\frac{1}{N_A}$ ($\alpha$ is a positive real number) at large values of $N_A$.

![Graph](image)

Figure 2.6: Electron density in channel vs. effective transverse electric field for $t_{ox} = 4, 10$ nm. (a) $V_{GS} = 0$ V and (b) $V_{GS} = 1$ V.

Figures 2.6 (a) and (b) show the variations of electron density $N_t$ in channel region with $E_{\text{eff}}$ for $t_{ox} = 4$ and 10 nm at $V_{GS} = 0$ and 1 V, respectively. The electron density in channel increases as $t_{ox}$ is decreased for a fixed $E_{\text{eff}}$, or as $E_{\text{eff}}$
or $V_{th}$ is decreased for a fixed $t_{ox}$ because the threshold voltage, $V_{th}$, decreases for both cases.

2.1.2 Mobility Dependence on Lateral Field

Historically, many mobility models have been developed to fit experimental data. The following mobility expression [41]-[43] derived by Caughey and Thomas in 1967 is still valid in describing the effective mobility $\mu_{eff}$ of electrons in inversion layer of n-MOSFET, which depends on lateral and transverse fields, fixed interface charge density, temperature, and doping density.

$$\mu_{eff}(|E_z|, E_{eff}, N_f, T, N_A) = \frac{\mu_{eff0}(E_{eff}, N_f, T, N_A)}{\sqrt{1 + \left(\frac{\mu_{eff0}|E_z|}{v_{sat}}\right)^2}}$$ (2.22)

where $E_z$ is the lateral electric field parallel to the Si-SiO$_2$ interface at Si surface. $E_{eff}$ is the effective transverse electric field at the Si surface. $qN_f$ is the fixed interface charge density per unit area. $T$ is the absolute temperature. $v_{sat}$ ($= 9.2 \times 10^6 \text{ cm/s}$ [45]) is the saturation velocity of electrons and $\mu_{eff0}$ is the effective transverse-field-dependent mobility of electrons in inversion layer at a low $E_z$.

2.1.3 Mobility Dependence on Effective Transverse Field at Low Lateral Field

The semi-empirical equation called the transverse-field-dependent mobility model for $\mu_{eff0}$ described in [44] takes into account the effects [44], [54] of phonon, surface roughness and screened Coulomb scattering. The phonon (lattice) scattering at surface or bulk is collisions between carriers and thermally agitated lattice atoms.
The surface roughness scattering is collisions between carriers and defects on rough semiconductor surface depending on gate oxide processing. The screened Coulomb scattering is collisions between carriers and ionized impurity atoms screened by the mobile carriers. The scattering due to fixed interface charges is collisions between carriers and the fixed interface charges existing at the Si-SiO₂ interface. The functional dependence of each mobility [44] due to each scattering effect on physical parameters for $\mu_{\text{eff0}}$ can be expressed by

$$
\mu_{\text{ph}} = f_{\text{ph}}(\mathcal{E}_{\text{eff}}, N_I, N_f, T) \quad (2.23)
$$

$$
\mu_{\text{sr}} = f_{\text{sr}}(\mathcal{E}_{\text{eff}}) \quad (2.24)
$$

$$
\mu_{\text{coul}} = f_{\text{coul}}(\mathcal{E}_{\text{eff}}, N_I, N_A, T) \quad (2.25)
$$

where $\mu_{\text{ph}}$ is the mobility related to the scattering by phonons and fixed interface charges, $\mu_{\text{sr}}$ is the mobility related to the surface roughness scattering that is dominant at a high transverse field region, and $\mu_{\text{coul}}$ is the mobility related to the screened Coulomb scattering [44], [55] that is dominant at a low transverse field region. By using the Matthiessen's rule [56] combining the aforementioned mobility components, the effective transverse-field-dependent mobility $\mu_{\text{eff0}}$ can be obtained.

$$
\mu_{\text{eff0}} = \frac{1}{1/\mu_{\text{ph}} + 1/\mu_{\text{sr}} + 1/\mu_{\text{coul}}} \quad (2.26)
$$
Figure 2.7: Effective transverse-field-dependent mobility at low lateral electric field vs. effective transverse electric field with bulk doping level as a parameter for $N_f = 0 \text{ cm}^{-2}$ and $T = 300^\circ K$.

$\mu_{eff0}$ is evaluated for given $N_A$ and $\phi_s$, $qN_f = 0 \text{ C/cm}^2$, and $T = 300^\circ K$ using Equations 2.1, 2.4, 2.19, 2.21, 2.23 - 2.25, i.e., the mobility equations for the scattering effects from [44], and Equation 2.26. Figure 2.7 shows $\mu_{eff0}$ vs. $E_{eff}$ with $N_A$ as a parameter at the onset of strong inversion and strong inversion. The model results of $\mu_{eff0}$ were already verified with Takagi’s experimental data [46], [47] for $qN_f \neq 0$ case, which are valid over a range of $E_{eff}$ from the onset of strong inversion to strong inversion. The roll-off of the effective mobility $\mu_{eff0}$ at the low transverse
field region for a fixed bulk doping level results from the screened Coulomb scattering due to doping impurities, or carrier-density fluctuation effect [57] due to the fixed interface charges at the Si-SiO₂ interface if \( q_N f \neq 0 \). However, as the bulk doping density is increased, the screened Coulomb scattering dominates the roll-off of \( \mu_{eff} \) at the low field region because the scattering due to the increased dopant atoms becomes significant. Integrating the Poisson's equation of Equation 2.8 from the semiconductor bulk towards the semiconductor surface \( (y = 0) \) with the boundary conditions of \( \frac{d\phi}{dy} = 0 \) and \( \phi = 0 \) in the bulk, and using Equations 2.1 and 2.4 at \( \alpha_s = 2|\psi_B| = 2V_{th} \ln \left( \frac{N_A}{n_i} \right) \), which is a common criterion for the onset of strong inversion of the long-channel MOSFET, give the effective transverse electric field at the onset of strong inversion, \( E_{eff0} \).

\[
E_{eff0} = \frac{\varepsilon_s + \varepsilon_D}{2} \bigg|_{\phi_s = 2|\psi_B|} \\
= \frac{1}{2} \left\{ \frac{V_{th}}{L_D} \sqrt{2} \left( e^{-\beta(2|\psi_B|)} + \beta(2|\psi_B|) - 1 + \left( \frac{n_i}{N_A} \right)^2 \left( e^{\beta(2|\psi_B|)} - \beta(2|\psi_B|) - 1 \right) \right) \\
+ \frac{V_{th}}{L_D} \sqrt{2} \beta(2|\psi_B|) - 1 \right\} \\
\approx \frac{V_{th}}{L_D} \frac{1}{\sqrt{2}} \left( \sqrt{\beta(2|\psi_B|)} - 1 + \left( \frac{n_i}{N_A} \right)^2 e^{\beta(2|\psi_B|)} + \sqrt{\beta(2|\psi_B|) - 1} \right) \\
= \frac{V_{th}}{L_D} \frac{1}{\sqrt{2}} \left( \sqrt{\beta(2|\psi_B|)} + \sqrt{\beta(2|\psi_B|) - 1} \right) \\
(2.27)
\]

The long-channel threshold voltage at the onset of strong inversion is given by

\[
V_T = V_{FB} + 2|\psi_B| + \frac{qN_A}{C_{ox}} L_D \sqrt{2} \sqrt{\beta(2|\psi_B|) - 1 + \left( \frac{n_i}{N_A} \right)^2 e^{\beta(2|\psi_B|)}} \\
(2.28)
\]
The values of $S_{ref0}$ and $V'_{cs} = V'T$ calculated using Equations 2.27 and 2.28 determine the values of $\mu_{ef0}$ at the starting points of the mobility curves in Figure 2.7.

2.2 Doping-Density-Dependent Mobility Model at Low Transverse Electric Field

The transverse-field-dependent mobility model described in the previous section is valid for MOSFETs operating at the onset of strong inversion or at strong inversion. Therefore, the mobility model applies to the modeling of on-state drive current rather than that of off-state leakage current. A doping-density-dependent mobility model [48] is selected to model the subthreshold current, and to formulate the transverse-field-dependent mobility model continuous from the strong inversion regime to the weak inversion regime. The doping-density-dependent mobility model takes into account the effects of lattice, ionized impurity and electron-electron scattering in the case of the mobility of electrons. The empirical expression [39], [48] for the doping-density-dependent mobility in bulk silicon derived by N. D. Arora et al. is given by

$$\mu_{n0} = \alpha_1 T_n^{\beta_1} + \frac{\alpha_2 T_n^{\beta_2}}{1 + \left( \frac{N_A}{\alpha_3 T_n^3} \right) \alpha_4 T_n^{\beta_4}}$$

(2.29)

where $T_n = T/300^oK$ is normalized temperature, and $\alpha_i$ and $\beta_i$ ($i = 1 - 4$) are fitting parameters. The original values [39], [48] of the fitting parameters are given in Table 2.1. In order to formulate the effective mobility model continuous at the point of the onset of strong inversion, i.e., $\phi_s = 2|\psi_B| (V_{GS} = V'T)$ or $\mathcal{E}_{eff} = \mathcal{E}_{ref0}$
Table 2.1: Parameter values of doping-density-dependent mobility model in Equation 2.29. $T = 300^\circ K$.

(Equation 2.27) at $T = 300^\circ K$, optimal fitting parameter values in Equation 2.29 have been found using an optimization technique [48]. The errors between $\mu_{n0}$ of Equation 2.29 and $\mu_{eff0}$ of Equation 2.26 at the onset of strong inversion are within 2.2 % over a wide range of the bulk doping density. The modified values of the fitting parameters are shown in Table 2.1. The unified mobility model created by combining the transverse-field-dependent and the doping-density-dependent mobility models has been confirmed to be continuous over a wide range of the bulk doping density, e.g., $3.9 \times 10^{15} \text{ cm}^{-3} \leq N_A \leq 4 \times 10^{18} \text{ cm}^{-3}$. Figure 2.8 shows the results of the unified mobility model, i.e., the variations of the effective mobility at a low lateral electric field with the effective transverse electric field and the bulk doping density as a parameter.

The transverse-field-dependent mobility model can also generate the values of $\mu_{eff0}$ at low effective transverse fields in weak inversion. However, there are no
Figure 2.8: Results of the unified mobility model combining the transverse-field-dependent and the doping-density-dependent mobility models at a low lateral electric field with bulk doping level as a parameter for $N_J = 0$ and $T = 300\, ^\circ K$.

experimental data to support the very low mobilities suggested by the mobility model. Because the transverse fields in weak inversion are low, surface scattering should be weak and the use of the doping-density-dependent mobility in bulk silicon in Figure 2.8 is not unreasonable. However, the bulk mobility incorporates screening by electrons in $p$-type silicon and there are few electrons at the surface in weak inversion, so the approach is not expected to be very accurate.
2.3 Extraction of Effective Transverse-Field-Dependent Mobility at Low Lateral Electric Field

The effective transverse-field-dependent mobility at low lateral electric field is extracted from Medici using the drift-diffusion current transport equation and some Medici data by a novel and simple extraction method, and compared to the result of the unified mobility model. The mobility extraction is necessary to check that the effective transverse-field-dependent mobility of Medici fits the Takagi's experimental data, i.e., the universal mobility curve in [46], [47].

In order to extract the effective transverse-field-dependent mobility, the gradual channel approximation [33] especially for gate voltages at and after the onset of strong inversion, and the current transport equation are used. Under the assumption of the gradual channel approximation, magnitudes of transverse fields in the direction perpendicular to the Si surface are much greater than magnitudes of lateral fields in the direction of current flow. The three-dimensional transport equation consisting of the drift current and the diffusion current components is given by

\[ \vec{J}_D = qn\mu\vec{E} + qD_n\vec{V}n \]  (2.30)

\[ = -qn\mu\vec{\nabla}\phi + q\mu V_{th}\vec{\nabla}n \]

where \( \vec{J}_D \) is the current density per unit area, \( D_n = \mu V_{th} \) is the Einstein relation that is valid by assuming Maxwell-Boltzmann statistics, \( n \) is the volume electron
density and $\mu(x, y, z)$ is the local-field-dependent microscopic mobility. Integration of Equation 2.30 from the Si bulk to the Si surface using the assumptions of the charge-sheet model [50] and assuming that the parameters in Equation 2.30 are independent of the position $z$ yields the following transport equation.

$$I_D \approx qN_i \mu_{eff} E_z(x, y = 0) + qD_n \frac{dN_i}{dx}$$ (2.31)

$$= -qN_i \mu_{eff} \frac{d\phi_s}{dx} + q\mu_{eff} V_{th} \frac{dN_i}{dx}$$

where $\mu_{eff}(x)$ is the effective mobility defined as

$$\mu_{eff}(x) \equiv \frac{\int_0^{\infty} n(x, y) \mu(x, y) \frac{\partial \phi(x, y)}{\partial x} dy}{\int_0^{\infty} n(x, y) \frac{\partial \phi(x, y)}{\partial x} dy}$$ (2.32)

$I_D$ is the current per unit width, $D_n = \mu_{eff} V_{th}$ is the Einstein relation, and $E_z(x, y = 0)$ is the lateral electric field at the Si surface. Note that the averaged $x$-component of field in deriving Equation 2.31 was replaced with $E_z(x, y = 0)$ at the Si surface since $E_z(x, y)$ is almost independent of the position $y$ within the inversion layer of a long-channel n-MOSFET [45]. From Equation 2.31 we can directly calculate the effective mobility $\mu_{eff}(x)$ without integration. Rearranging Equation 2.31 for $\mu_{eff}(x)$, we obtain

$$\mu_{eff}(x) = I_D \left[qN_i E_z(x, 0) + qV_{th} \frac{dN_i}{dx} \right]^{-1}$$ (2.33)

We can evaluate Equation 2.33 with reasonable numerical values of $\Delta N_i$ and $\Delta x$ in the middle of the channel $x = L/2$ at the Si surface in a long-channel MOSFET in Figure 2.9 by excluding the effect due to field penetration from the drain or source.
depletion region. From Equation 2.33 we have the expression for $\mu_{\text{eff}}(L/2)$.

$$\mu_{\text{eff}}(L/2) = I_D \left[ qN_i(L/2)E_x(L/2, 0) + qV_{th} \frac{N_i(L/2 + \Delta x) - N_i(L/2)}{\Delta x} \right]^{-1}$$  \hspace{1cm} (2.34)$$

where $E_x(L/2, 0)$ is the lateral electric field measured in the middle of the channel at the Si surface. Note that as the saturation velocity approaches infinity mathematically, the effective mobility $\mu_{\text{eff}}(L/2)$ becomes the same as the effective transverse-field-dependent mobility $\mu_{\text{eff}0}(L/2)$.

Figure 2.9: Uniformly doped long-channel n-MOSFET used for extracting effective transverse-field-dependent mobility for $V_{DS} = 0.05 \text{ V}$.

A procedure to extract the effective transverse-field-dependent mobility at a low lateral field is described:

1. For a given bulk doping level $N_A$ of the long-channel n-MOSFET in Figure 2.9, set a range of gate voltages over which the effective transverse-field-dependent
mobility $\mu_{eff0}$ vs. the effective transverse field $E_{eff}$ in Figure 2.8 can be predicted. The effective transverse electric field and gate voltage equal to the threshold voltage at the onset of strong inversion are given in Equations 2.27 and 2.28, respectively.

2. Set $V_{DS} = 0.05 \text{ V}$. Select $L = 1 - 10 \mu \text{m}$ depending on the bulk doping density.

3. Let $v_{sat} = \infty$, e.g., $v_{sat} = 1 \times 10^{30} \text{ cm/s}$, then $\mu_{eff0} \approx \mu_{eff}$. Set $\Delta x = 0.1 \text{ nm}$.

4. Perform Medici simulations. Obtain output data of $I_D < 0$, $N_i(L/2)$, $N_i(L/2 + \Delta x)$, $E_x(L/2) < 0$, $\phi_s(L/2)$ at the Si surface for each $V_{GS}$ and $V_{DS} = 0.05 \text{ V}$.

5. Calculate $\mu_{eff0}$ and $E_{eff}$ using Equations 2.34, 2.1, 2.4, and the Medici output data for a given $N_A$.

Figure 2.10 shows comparisons of $\mu_{eff0}$ of the unified mobility model with $\mu_{eff0}$ extracted from Medici using the Medici data and Equations 2.34, 2.1 and 2.4 for three different doping levels. The extracted effective transverse-field-dependent mobilities are in good agreement with the unified mobility model results. The mobility extraction technique described in this section is useful to extract the effective transverse-field-dependent mobility of any different model used for two-dimensional device simulations and to compare the extracted results with the universal mobility curves of $\mu_{eff0}$ vs. $E_{eff}$ obtained from measurements.
Figure 2.10: Comparisons of the effective transverse-field-dependent mobility of the unified mobility model to the extracted mobility for $N_f = 0 \text{ cm}^{-2}$ and $T = 300^\circ K$.

2.4 Summary

The relations between the effective transverse electric field and the physical parameters have been presented together with some important model equations based on the long-channel MOSFET theory. Most important and independent parameters determining the effective transverse field, in turn, the effective transverse-field-dependent mobility and the threshold voltage of long-channel MOSFETs are
the bulk or the channel doping density and the oxide thickness. The transverse-field-dependent mobility model accounting for the effects of phonon, surface roughness and screened Coulomb scattering has been combined with the doping-density-dependent mobility model accounting for the effects of lattice, ionized impurity and electron-electron scattering. The resulting unified mobility model gives the effective transverse-field-dependent mobility at a low lateral field as a function of $E_{\text{eff}}$ for the MOSFET operating at subthreshold, threshold and above-threshold regions. The unified mobility model will be used to model currents in deep-submicron pocket MOSFETs in the subsequent chapters. A mobility extraction method has been proposed, and the effective transverse-field-dependent mobility has been successfully extracted from Medici using the current transport equation and some Medici simulation data. Using the mobility extraction technique, the relation between the effective transverse-field-dependent mobility of any model to be used for two-dimensional simulations and the effective transverse field can be found to compare the extracted results with the universal mobility curves of $\mu_{\text{eff}0}$ vs. $E_{\text{eff}}$ obtained from measurements.
CHAPTER 3

DESIGN OF 0.1-μm POCKET N-MOSFETS

A heavily doped pocket implant surrounding drain/source region minimizes surface drain-induced barrier lowering (DIBL). It also prevents bulk punchthrough called subsurface DIBL [3] although this is not a major consideration in this dissertation. In Section 3.1, short-channel effects and characteristics of pocket and conventional bulk n-MOSFETs are studied by comparing the pocket devices with the conventional bulk devices in terms of DIBL, subthreshold slope and off-state current. A deep-submicron pocket n-MOSFET proves to be effective in controlling the short-channel effects. In Section 3.2, 0.1-μm channel-length pocket MOSFET structures satisfying the specifications of the off-state current, $I_{off} \leq 1 \, nA/μm$, and the sensitivity of off-state current to channel length, $\frac{ΔI_{off}/I_{off}}{ΔL/L} = 40\%$, at the unique lower design-space boundary for the off-state current are explored by using the two-dimensional device simulator Medici [39]. Procedures, i.e., algorithms to automatically and accurately find the doping densities, $N_p$ and $N_c$, of the device structures to be located at the lower design-space boundary were developed and are described in the section. A qualitative model for the deep-submicron pocket devices built based on the comparison of pocket length ($L_p$) to drain-junction depletion depth is presented. In Section 3.3 comments on design philosophy for the
0.1-$\mu$m pocket n-MOSFETs is presented. Finally, this chapter is summarized in Section 3.4.

3.1 Short-Channel Effects and Characteristics of Pocket and Conventional Bulk n-MOSFETs

In this section, short-channel effects which must be controlled in designing deep-submicron n-MOSFETs are studied through two-dimensional device simulations. The heavily doped pocket implants around drain and source of deep-submicron pocket n-MOSFETs prove to be effective in suppressing the short-channel effects by comparisons of the deep-submicron pocket n-MOSFETs with deep-submicron conventional bulk n-MOSFETs without pocket implants. The characteristics of the deep-submicron pocket n-MOSFETs biased to weak inversion are presented in this section.

As MOSFETs are scaled down to deep-submicrometer or sub-100 nm regime, short-channel effects give rise to an increase in the off-state current and to a decrease in the threshold voltage, thereby increasing the standby power consumption in CMOS integrated circuits. The following are typical short-channel effects observed from measurements and two-dimensional simulations.

- Subthreshold current $I_{D(sub)}$ for $V_{GS} < V_T$ increases as $V_{DS}$ is increased.
\begin{itemize}
  \item Sensitivity of threshold voltage or off-state current to channel length or drain voltage increases as channel length is decreased.
  \item Subthreshold slope increases as channel length is decreased.
\end{itemize}

The subthreshold slope is defined as measuring how many volts of gate swing a MOSFET takes to get one decade of reduction in $I_D$. Mathematically [23],

$$S \equiv (\ln 10) \left( \frac{kT}{q} \right) \left[ \frac{dV_{GS}}{d(\ln I_D)} \right]$$

(3.1)

The subthreshold slope in a long-channel approximation is strongly dependent on the oxide and depletion-layer capacitances, and given by [23], [36]

$$S = 2.3V_{th} \left( 1 + \frac{\epsilon_s}{\epsilon_{ox}} \frac{t_{ox}}{W_D} \right) [V/dec]$$

(3.2)

where $W_D$ is the depletion-layer depth controlled by gate.

The surface or subsurface DIBL [25] that contributes to the short-channel effects has two-dimensional nature. For $V_{DS} = 0$, the distance between source and drain becomes short enough to increase the source-to-drain coupling of electric field lines as channel length is decreased. Therefore, the highest energy barrier becomes lowered: the minimum potential between source and drain is higher than for long-channel case. For $V_{DS} > 0$, more field lines from the drain junction penetrate into the source junction through the gate-controlled channel or depletion region. Therefore, the highest energy barrier to injection of majority carriers from source is lowered further: the minimum potential between source and drain becomes even
higher. DIBL for $V_{DS} > 0$ causes the carriers in source to be injected into the surface or even into the bulk if the DIBL effect is larger, thereby increasing the off-state current. The surface DIBL depending on MOSFET design parameters increases as the bulk or channel doping density is decreased, the channel length is decreased, the junction depth is increased, the oxide thickness is increased or the back gate bias is decreased [25].

Figure 3.1 shows two-dimensional simulation results of surface potential distributions between source and drain of pocket and conventional bulk n-MOSFETs with 0.1- and 1-$\mu$m channel lengths for $V_{GS} = 0$ V and $V_{DS} = 0$ V. Notice that

![Figure 3.1: Medici results of surface potential $\psi_s$ vs. normalized distance from source (a) of pocket n-MOSFETs with $L = 0.1$, 1 $\mu$m, $N_p = 1.948 \times 10^{18}$ cm$^{-3}$, $N_c = 2.175 \times 10^{17}$ cm$^{-3}$ and $L_p = 0.024$ $\mu$m, and (b) of conventional bulk n-MOSFETs with $L = 0.1$, 1 $\mu$m and $N_A = 2.175 \times 10^{17}$ cm$^{-3}$. $V_{DS} = 0$ V, $V_{GS} = 0$ V, $V_{BG} = 0$ V, $t_{ox} = 4$ nm and $r_j = 0.06$ $\mu$m for all the devices.](image)
the potentials in the center of both devices are the same for $L = 1 \mu m$ because the doping level at the center region of the pocket n-MOSFET is identical with the doping level of the conventional bulk n-MOSFET, both devices have the same values of $t_{ox}$, $r_j$ and $V_{BG}$ (back gate bias), and the pocket length of the pocket MOSFET is even shorter than the channel length, i.e., $L_p = 0.024 \mu m \ll L$. The threshold voltage in both devices is controlled by the potential minimum, which for the pocket n-MOSFET is located in the pocket region while for the conventional device the potential minimum is located in the center of the device. The identical zero bias voltages applied to the source and drain electrodes result in the symmetrical potential distribution and zero drain-to-source current.

To evaluate the subthreshold surface potential for $V_{DS} = 0$ V of the long-channel n-MOSFETs, we assume that the mobile carrier (electron) charge density is negligible compared to the depletion-layer charge under gate since the MOSFETs are biased to subthreshold regime at $V_{GS} = 0$, and then we can use the following long-channel surface potential model equation [50] derived from the one-dimensional Poisson's equation. The assumption is justified by negligible errors between the surface potential values calculated from Equations 2.13 and 3.3.

$$C_{ox}(\phi_G - \phi_s) = qN_AL_D\sqrt{2\beta\phi_s} - 1$$

(3.3)
A closed-form solution $\phi_s$ to a quadratic equation derived from Equation 3.3 is as follows.

$$
\phi_s = V_{th} \left[ \left( \frac{-\alpha + \sqrt{\alpha^2 + 4(3\phi_G - 1)}}{2} \right)^2 + 1 \right] \quad (3.4)
$$

where

$$
\alpha = \frac{3qN_A L D \sqrt{2}}{C_{ox}} \quad (3.5)
$$

The surface potential $\psi_s$ in Figure 3.1 is defined as

$$
\psi_s = \frac{E_i(0) - E_f}{-q} \quad (3.6)
$$

where $E_i(0)$ is the intrinsic Fermi level at the surface of semiconductor, and $E_f$ is the extrinsic Fermi level of the semiconductor. Using the definitions of $\phi_s$ and $\psi_s$ in Equations 2.7 and 3.6, the relation between $\psi_s$ and $\phi_s$ is derived.

$$
\psi_s = \phi_s - |\psi_B| \quad (3.7)
$$

where $\psi_B$ is the bulk potential of the p-type semiconductor defined as

$$
\psi_B \equiv \frac{E_i(\infty) - E_f}{-q} = -V_{th} \ln \left( \frac{N_A}{n_i} \right) \quad (3.8)
$$

The maximum and minimum of the surface potential, $\psi_s(\text{max})$ and $\psi_s(\text{min})$, in the center and pocket regions, respectively, of the pocket n-MOSFETs, and a percent decrease in $\psi_s(\text{max})$ of the 0.1-$\mu$m pocket device from the surface potential of the long-channel model are summarized in Table 3.1, while the minimum of the surface
potential of the conventional bulk n-MOSFETs, and a percent increase in $\psi_{s(min)}$ of the 0.1-$\mu m$ conventional device from the surface potential of the long-channel model are summarized in Table 3.2. The values of the surface potential of the long-channel model in Tables 3.1 and 3.2 were calculated using Equations 3.4, 3.7, 2.14–2.17 and the constants and parameter values given in Subsection 2.1.1 of Chapter 2.

<table>
<thead>
<tr>
<th></th>
<th>Two-dimensional model</th>
<th>Long-channel model</th>
</tr>
</thead>
<tbody>
<tr>
<td>$L = 0.1 \mu m$</td>
<td>0.2394 (673%)</td>
<td>0.2416 (678%)</td>
</tr>
<tr>
<td>$L = 1 \mu m$</td>
<td>0.2862 (-7.23%)</td>
<td>0.3085 (0%)</td>
</tr>
</tbody>
</table>

Table 3.1: Pocket MOSFET. Minimum and maximum values of surface potential of 0.1-$\mu m$ and 1-$\mu m$ pocket n-MOSFETs, and percent changes (numeral values in parentheses) of $\psi_{s(max)}$ from the surface potential value of the long-channel model for $V_{GS} = 0 \, V$ and $V_{DS} = 0 \, V$.

<table>
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<th>Two-dimensional model</th>
<th>Long-channel model</th>
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</thead>
<tbody>
<tr>
<td>$L = 0.1 \mu m$</td>
<td>0.4228 (37.05%)</td>
<td>0.3085 (0%)</td>
</tr>
<tr>
<td>$L = 1 \mu m$</td>
<td>0.3085</td>
<td>0.3085</td>
</tr>
</tbody>
</table>

Table 3.2: Conventional MOSFET. Minimum values of surface potential of 0.1-$\mu m$ and 1-$\mu m$ conventional bulk n-MOSFETs, and percent changes (numeral values in parentheses) of $\psi_{s(min)}$ from the surface potential value of the long-channel model for $V_{GS} = 0 \, V$ and $V_{DS} = 0 \, V$.

$\psi_{s(max)}$ of the 1-$\mu m$ pocket n-MOSFET is the same as that of the long-channel model, whereas $\psi_{s(max)}$ of the 0.1-$\mu m$ pocket n-MOSFET is lower than the long-channel value. $\psi_{s(min)}$ of the 0.1- or 1-$\mu m$ pocket n-MOSFET is even higher than
that of the one-dimensional long-channel model because the two-dimensional po-
tential distribution at the source or drain junction has affected $\psi_{s(min)}$ of the two-
dimensional model. Note that $\psi_{s(max)}$ exists in the center region while $\psi_{s(min)}$ exists
in the pocket region. $\psi_{s(min)}$ of the 1-$\mu$m conventional n-MOSFET is just that of the
long-channel model while $\psi_{s(min)}$ of the 0.1-$\mu$m conventional n-MOSFET is 37.05%
higher than the long-channel value. The 0.1-$\mu$m conventional device is expected to
have large surface DIBL as well as subsurface DIBL. The long-channel MOSFET
theory predicts that the 0.1-$\mu$m conventional MOSFET is under punchthrough
condition since the sum (=0.154 $\mu$m) of the depletion-region widths of source and
drain junctions in the semiconductor bulk exceeds the channel length 0.1 $\mu$m. We
note that as the channel length is shortened from 1 $\mu$m to 0.1 $\mu$m, $\psi_{s(min)}$ of the
conventional MOSFET increases by 37.05% while $\psi_{s(min)}$ of the pocket MOSFET
even decreases by 0.91%. Therefore, when drain bias is applied, more leakage cur-
rent due to much lower barrier to carrier injection from the source in the 0.1-$\mu$m
conventional MOSFET will flow than in the 0.1-$\mu$m pocket MOSFET.

Now we examine short-channel effects of the 0.1-$\mu$m pocket and conventional
bulk MOSFETs for $V_{DS} > 0$ and thus non-zero drain-to-source current from two-
dimensional simulation results. Figure 3.2, Tables 3.3 and 3.4 show
that as the channel length is decreased from 1 $\mu$m to 0.1 $\mu$m, the minimum of the
potential distribution of the pocket device becomes 7.57% higher at $V_{DS} = 1$ V.
while the minimum of the potential distribution of the conventional bulk device
Figure 3.2: Medici results of surface potential $\psi_s$ vs. normalized distance from source (a) of pocket n-MOSFETs with $L = 0.1, 1 \mu m$, $N_p = 1.948 \times 10^{18} \text{cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{cm}^{-3}$ and $L_p = 0.024 \mu m$, and (b) of conventional bulk n-MOSFETs with $L = 0.1, 1 \mu m$ and $N_A = 2.175 \times 10^{17} \text{cm}^{-3}$, $V_{GS} = 0$, $V_{BG} = 0$, $t_{ox} = 4 nm$, and $r_j = 0.06 \mu m$ for all the devices.
Table 3.3: Pocket MOSFET. Minimum surface potentials of 0.1-\(\mu\)m and 1-\(\mu\)m pocket n-MOSFETs for \(V_{DS} = 0.1, 1\) V and \(V_{GS} = 0\) V.

<table>
<thead>
<tr>
<th>(L = 0.1) (\mu)m</th>
<th>(L = 1) (\mu)m</th>
</tr>
</thead>
<tbody>
<tr>
<td>(V_{DS} = 0.1) V</td>
<td>(V_{DS} = 1) V</td>
</tr>
<tr>
<td>(V_{x(min)}) [V]</td>
<td>0.2412</td>
</tr>
</tbody>
</table>

Table 3.4: Conventional MOSFET. Minimum surface potentials of 0.1-\(\mu\)m and 1-\(\mu\)m conventional bulk n-MOSFETs for \(V_{DS} = 0.1, 1\) V and \(V_{GS} = 0\) V.

<table>
<thead>
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<th>(L = 1) (\mu)m</th>
</tr>
</thead>
<tbody>
<tr>
<td>(V_{DS} = 0.1) V</td>
<td>(V_{DS} = 1) V</td>
</tr>
<tr>
<td>(V_{x(min)}) [V]</td>
<td>0.4333</td>
</tr>
</tbody>
</table>

increases by 54.26\% at \(V_{DS} = 1\) V. Therefore, the well-designed pocket MOSFET has much less DIBL effect than the conventional bulk MOSFET without pocket implants in the deep-submicron regime. The minimum surface potential, i.e., the highest energy barrier to electron injection from the source in the 0.1-\(\mu\)m pocket n-MOSFET always exists at a fixed position in the pocket region near the source over the range of \(V_{DS} = 0 - 1\) V in Figures 3.1 (a) and 3.2 (a) as the long-channel (\(L = 1\) \(\mu\)m) pocket or conventional MOSFET has a fixed position of the minimum surface potential over the range of \(V_{DS} = 0 - 1\) V. But the highest energy barrier to electron injection from the source of the 0.1-\(\mu\)m conventional bulk n-MOSFET moves closer towards the source noticeably as the drain voltage is increased from 0.1 V to 1 V. This result indicates that the 0.1-\(\mu\)m conventional device is in bulk punchthrough and then has an increase in punchthrough current.
Such a conventional device is unacceptable for applications, and heavier doping would have to be used.

The two-dimensional simulation results in Figure 3.3 and Table 3.5 show that the 0.1-\(\mu\)m pocket n-MOSFET has a low subthreshold slope like a long-channel subthreshold slope and much less DIBL than the 0.1-\(\mu\)m conventional bulk n-MOSFET. In Figure 3.3 and Table 3.5, the subthreshold slope of the long-channel (1-\(\mu\)m) pocket MOSFET is slightly higher than the 1-\(\mu\)m conventional MOSFET because the average depletion depth controlled by the gate of the pocket device might be smaller than that of the conventional device and then the subthreshold slope for the pocket device calculated by Equation 3.2 is higher than for the conventional device. The high subthreshold slopes of the 0.1-\(\mu\)m conventional MOSFET departing markedly from the long-channel subthreshold slopes indicate that the device is in bulk punchthrough. The punchthrough leads to a significant increase in the subthreshold current that is not proportional to \(\exp(V_{GS}/V_{th})\) since the gate of the short-channel MOSFET has lost the control of the channel depletion-layer charge.

One of methods [3], [27] to evaluate DIBL is to measure an increase in subthreshold current \(I_{D(sub)}\) as \(V_{DS}\) is increased, e.g., from \(V_{DS} = 0.1\) to 1 V at a fixed \(V_{GS} = 0\) V. Mathematically,

\[
R = \frac{\Delta V_{DS}}{\Delta \log_{10}(I_{D(sub)})} \text{ [V/dec]} \tag{3.9}
\]
Figure 3.3: Subthreshold characteristics of (a) 0.1 and 1 μm channel-length pocket n-MOSFETs with $N_p = 1.948 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \ \mu\text{m}$. and (b) 0.1 and 1 μm channel-length conventional bulk n-MOSFETs with $N_A = 2.175 \times 10^{17} \text{ cm}^{-3}$. $V_{DS} = 0.1$ and 1 V.
Thus, a large $R$ means a small DIBL effect. Table 3.5 shows the short-channel

<table>
<thead>
<tr>
<th></th>
<th>$I_{off}$ [$\mu A/\mu m$] for $V_{DS} = 1, \text{V}$</th>
<th>$R$ [V/dec] for $V_{GS} = 0, \text{V}$</th>
<th>$S$ [mV/dec] for $V_{DS} = 1, \text{V}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.1 $\mu m$ pocket MOSFET</td>
<td>0.567</td>
<td>1.809</td>
<td>92.66</td>
</tr>
<tr>
<td>1 $\mu m$ pocket MOSFET</td>
<td>0.115</td>
<td>10.815</td>
<td>79.49</td>
</tr>
<tr>
<td>0.1 $\mu m$ conventional MOSFET</td>
<td>9982</td>
<td>0.870</td>
<td>172.50</td>
</tr>
<tr>
<td>1 $\mu m$ conventional MOSFET</td>
<td>0.224</td>
<td>38.290</td>
<td>71.48</td>
</tr>
</tbody>
</table>

Table 3.5: Two-dimensional simulation results of the off-state current for $V_{DS} = 1\, \text{V}$, the DIBL rate for $V_{GS} = 0\, \text{V}$ and the subthreshold slope for $V_{DS} = 1\, \text{V}$ of 0.1 and 1 $\mu m$ channel-length pocket and conventional bulk n-MOSFETs.

effects of the deep-submicron n-MOSFETs quantitatively. As the channel length is decreased from 1 $\mu m$ to 0.1 $\mu m$, the off-state current in the pocket n-MOSFET increases by 393 $\%$, while the current in the conventional n-MOSFET increases by $4.46 \times 10^6$ $\%$.

The above discussion suggests that the well-designed deep-submicron pocket n-MOSFET has better short-channel immunity compared to the conventional bulk n-MOSFET as discussed qualitatively and quantitatively. However, a more heavily doped conventional n-MOSFET could have better short-channel behavior than the conventional device considered in this section.
3.2 0.1-μm Pocket n-MOSFET Structures Satisfying Specifications of Off-State Current and Sensitivity of Off-State Current to Channel Length

3.2.1 Computer Algorithms Based on Device Physics and Two-Dimensional Device Simulation Results

In designing a deep-submicron pocket MOSFET, it is important to accurately find relations between process and device parameters from measurement or two-dimensional simulation. In this section the doping densities $N_p$ and $N_c$ at the pocket and center regions corresponding to each of given pocket lengths of the 0.1-μm pocket n-MOSFETs, and relations between the three design variables, $N_p$, $N_c$ and $L_p$, for the pocket n-MOSFETs are found by using the two-dimensional device simulator Medici with the constraints of the off-state current $I_{off} \leq 1 \, nA/\mu m$ and the sensitivity of off-state current to channel length i.e., $40\%$ increase in the off-state current from $I_{off}(L) = 0.714 \, nA/\mu m$ to $I_{off}(0.9L) = 1 \, nA/\mu m$ due to $10\%$ decrease in the channel length (Figure 3.4). Note that the decrease in the channel length applies to the center region, i.e., $0.9L = 2L_p + (L_c - 0.1L)$. The values of $N_p$, $N_c$ and $L_p$ as solutions to a set of nonlinear equations for the two-dimensional simulation will make up the lower design-space boundary for the off-state current. The oxide thickness and junction depth of the pocket MOSFETs are fixed at $t_{ox} = 4 \, nm$ and $r_j = 0.06 \, \mu m$. A plot of $I_{off}$ vs. $L$ in Figure 3.4 shows schematically how the off-state current depends on the channel
Figure 3.4: Schematic illustration of the sensitivity of off-state current to channel length.

At long channel lengths the current is proportional to $1/L$, but as the channel length is reduced to the point where short-channel effects due to DIBL begin, the off-state current increases more rapidly than $1/L$. The off-state current is less sensitive to $L$ for large $L$ but increases rapidly as $L$ enters the short-channel regime. The slope of this curve is the channel-length sensitivity of the off-state current in the device. One of the specifications of the device is that the sensitivity cannot be too large, so that the off-state current will be well controlled in a chip containing many devices scaled down to the short-channel regime.

The two-dimensional Medici simulation using most of the maximum grid points of 10000 inherently takes a long time to produce final outputs. To automatically and accurately find the parameter values of $N_p$ and $N_c$ as the solutions with which
the 0.1-μm pocket devices satisfy the boundary design specification of the sensitivity of $I_{off}$ to $L$ within any desired error tolerance in a finite simulation time, computer algorithms are developed, based on the properties of the variation of $I_{off}$ with $N_p$ for a fixed $N_c$ and the variation of the channel length sensitivity of $I_{off}$ with $N_c$ for $I_{off}(0.9L) = 1 \text{ nA/μm}$. The algorithms are described by using the ALGOL programming language. As studied in Chapter 2, as the doping density of a uniformly doped MOSFET is increased, the drain-to-source current is reduced since the threshold voltage becomes higher and the mobility of carriers in channel decreases. Likewise, $I_{off}$ in a pocket MOSFET will be reduced as $N_p$ is increased if all the other design parameters are fixed. The following procedure POCKET.DOPING is implementation of the algorithm to find, using Medici, the pocket dopant concentration for a given dopant concentration at the center region and a given pocket length so that $I_{off}(0.9L)$ for the channel length 0.9$L$ is equal to the maximum of the design specification of $I_{off(spec)}(0.9L) = 1 \text{ nA/μm}$.

```
procedure POCKET_DOPING($N_c, L, L_p, N_{p0}, \Delta N_{p0}$):
  begin
    $I_{off(spec)}(0.9L) \leftarrow 1 \times 10^{-9} \text{ A/μm};$
    comment: Assign proper values to variables of initial off-state current, initial pocket doping level, initial incremental pocket doping level, and error tolerance of off-state current.
    $I_{off}^0(0.9L) \leftarrow 0; \text{ IOFF.TOL} \leftarrow 5 \times 10^{-3}; i \leftarrow 0; N_p^1 \leftarrow N_{p0}$ (initial guess);
  end
```
$$\Delta N_p^i \leftarrow \Delta N_{p0} \text{ (initial guess: a fraction of } N_p^i, \text{ e.g., } 5 \times 10^{17} \text{ cm}^{-3})$$

while $$\left| \frac{I_{off}^i(0.9L) - I_{off(\text{spec})}(0.9L)}{I_{off(\text{spec})}(0.9L)} \right| \geq \text{IOFF.TOL} \right \} \text{ do}

\begin{align*}
\text{begin} \\
\quad i \leftarrow i + 1; \\
\text{end}
\end{align*}

Compute $$I_{off}(0.9L)$$ with given $$N_p^i, N_c, L_p$$ and $$L$$ for $$i \geq 1$$:

if $$I_{off}^i(0.9L) < I_{off(\text{spec})}(0.9L)$$ then

\begin{align*}
\text{begin} \\
\quad \text{if } I_{off}^{i-1}(0.9L) > I_{off(\text{spec})}(0.9L) \text{ then} \\
\text{begin} \\
\quad \Delta N_p^i \leftarrow \Delta N_p^i/2; \quad \Delta N_p^{i+1} \leftarrow \Delta N_p^{i+1}/2; \quad I_{off}^i(0.9L) \leftarrow I_{off}(0.9L); \\
\text{end} \\
\text{else} \\
\text{begin} \\
\quad \Delta N_p^{i+1} \leftarrow \Delta N_p^i, \quad \Delta N_p^{i+1} \leftarrow \Delta N_p^{i+1}; \quad I_{off}^i(0.9L) \leftarrow I_{off}(0.9L); \\
\text{end} \\
\text{else} \\
\text{begin} \\
\quad \text{if } [I_{off}^i(0.9L) > I_{off(\text{spec})}(0.9L)] \cdot \text{AND.} \quad [I_{off}^{i-1}(0.9L) < I_{off(\text{spec})}(0.9L)] \\
\text{AND.} \quad [I_{off}^{i-1}(0.9L) > 0.] \text{ then} \\
\text{begin} \\
\text{end}
\end{align*}
\[ N_{p}^{i+1} \leftarrow N_{p}^{i} + \Delta N_{p}^{i}/2; \Delta N_{p}^{i+1} \leftarrow \Delta N_{p}^{i}/2; I_{off}^{i}(0.9L) \leftarrow I_{off}^{i}(0.9L); \]

\[ \text{end} \]

\[ \text{else} \]

\[ \begin{align*} 
N_{p}^{i+1} & \leftarrow N_{p}^{i} + \Delta N_{p}^{i}; \Delta N_{p}^{i+1} \leftarrow \Delta N_{p}^{i}; 
I_{off}^{i}(0.9L) & \leftarrow I_{off}^{i}(0.9L); 
\end{align*} \]

\[ \text{end} \]

\[ \text{end} \]

\[ \text{end}/*end of while loop*/ \]

\[ \text{comment: } I_{off}^{i}(0.9L) \text{ is within the error tolerance IOFF.TOL.} \]

\[ \text{return} \]

\[ N_{p}^{i}; \]

\[ \text{end}/*end of procedure POCKET.DOPING*/ \]

The procedure POCKET.DOPING is used to find from two-dimensional simulation a pocket doping density of \( N_{p} \) for given values of \( N_{c}, \) \( L_{p} \) and \( L \) of a pocket MOSFET satisfying \( I_{off}(0.9L) = 1 \) \( nA/\mu m, \) and then \( I_{off}(L) \) is computed by Medici using this value of \( N_{p} \) that is appropriate to the given values of \( N_{c}, L_{p} \) and \( L. \) The sensitivity of off-state current to channel length is calculated by

\[ \frac{\Delta I_{off}/I_{off}}{\Delta L/L} = \frac{[I_{off}(0.9L) - I_{off}(L)]/I_{off}(L)}{10 \%} \quad (3.10) \]

Figures 3.5 and 3.6 show the variations of percent change in \( I_{off} \) with \( N_{c}. \)
Figure 3.5: (a) Percent change of $I_{off}$ due to 10 % decrease in channel length vs. doping density $N_c$ at center region for $I_{off}(0.9L) = 1 \text{nA/\mu m}$, $L = 0.1 \mu m$ and $L_p = 0.025$, $0.0325$ and $0.04 \mu m$. (b) Percent change of $I_{off}$ vs. $N_c$ and $N_p$ for $L_p = 0.025$ and $0.04 \mu m$. This figure (b) is a part of the figure (a).
Figure 3.6: (a) Percent change of $I_{off}$ due to 10 % decrease in channel length vs. doping density $N_p$ at pocket region for $L_p = 0.025$, $0.0325$ and $0.04 \mu m$, and (b) percent change of $I_{off}$ due to 10 % decrease in channel length vs. $L_p$ for $N_c = 1 \times 10^{17}$, $2 \times 10^{17}$ and $3 \times 10^{17} \text{cm}^{-3}$. $I_{off}(0.9L) = 1 \text{nA}/\mu \text{m}$ and $L = 0.1 \mu m$.

$N_p$ and $L_p$ due to the 10 % decrease in the channel length $L = 0.1 \mu m$. Two-dimensional simulation shows that both $I_{off}(L)$ and $N_p$ of the 0.1-$\mu m$ pocket n-MOSFETs decrease as $N_c$ increases for fixed $L_p$ and $I_{off}(0.9L) = 1 \text{nA}/\mu \text{m}$. Then we find the results in Figures 3.5 and 3.6 from $\Delta I_{off}/I_{off} = (I_{off}(0.9L)-I_{off}(L))/I_{off}(L)$ for $L = 0.1 \mu m$ of the definition of the sensitivity of off-state current to channel length in Equation 3.10. We see that as $N_c$ approaches $N_p$ (uniformly doped n-MOSFET case) or $N_c$ becomes higher than $N_p$ (reversed pocket n-MOSFET case), the channel-length sensitivity of $I_{off}$ increases monotonically. We also see that as $L_p$ increases, the channel-length sensitivity depends more strongly on the doping densities $N_p$ and $N_c$. 
All acceptable pocket n-MOSFETs satisfy $N_c < N_p$. In the cases of $N_c = N_p$ and $N_c > N_p$, the device has a small off-state current ($I_{off}(L)$) and a high channel-length sensitivity. However, the channel-length sensitivity is a control issue which cannot allow a very high sensitivity. Therefore, $N_c = N_p$ and $N_c > N_p$ are not acceptable in designing 0.1-μm n-MOSFETs even though the off-state current is small. We are forced to use $N_c < N_p$ to gain an acceptable channel-length sensitivity at the cost of a large off-state current. If the channel length is shortened into the sub-100 nm channel-length regime for fixed $N_c$, $N_p$ and $L_p$, the channel-length sensitivity will go up with an increase in the off-state current due to larger short-channel effect. To bring the sensitivity down again to the specification, $N_c$ should be even lower while $N_p$ should be even higher. Thus, more pronounced pocket n-MOSFET is necessary.

Based on the variations of the sensitivity with $N_c$ in Figure 3.5, another algorithm to automatically and accurately find both $N_c$ and $N_p$ for given $L$ and $L_p$ so that $I_{off}(0.9L) = 1$ nA/μm, $I_{off}(L) = 0.714$ nA/μm, and \( \frac{I_{off}(0.9L) - I_{off}(L)}{I_{off}(L)} = 40\% \) due to 10\% decrease in the channel length, i.e., due to the reduction of the length of the center region is developed, and implemented in the procedure CENTER_POCKET_DOPING. A Medici program implementing the algorithms of CENTER_POCKET_DOPING and POCKET_DOPING is in Appendix B.

```
procedure CENTER_POCKET_DOPING(L, Lp):
begin
```
\[
S_{\text{lof}} \triangleq \frac{I_{\text{lof}}(0.9L) - I_{\text{lof}}(L)}{I_{\text{lof}}(L)}.
\]

\[
S_{\text{lof}}(\text{spec}) \leftarrow 0.4 \ (40 \%);
\]

**comment:** Assign proper values to variables.

\[
S_{\text{lof}}^0 \leftarrow 0.; \ S\text{SENS.TOL} \leftarrow 5 \times 10^{-3}; \ j \leftarrow 0;
\]

\[
N_c^1 \leftarrow N_{c0} \ (\text{initial guess}); \ N_p^1 \leftarrow N_{p0} \ (\text{initial guess});
\]

\[
\Delta N_c^1 \leftarrow \Delta N_{c0} \ (\text{initial guess: a fraction of } N_c^1, \ e.g., \ 5 \times 10^{16} \ \text{cm}^{-3});
\]

\[
\Delta N_p^1 \leftarrow \Delta N_{p0} \ (\text{initial guess: a fraction of } N_p^1, \ e.g., \ 5 \times 10^{17} \ \text{cm}^{-3});
\]

\[
\textbf{while} \ \left| \frac{S_{\text{lof}}^j - S_{\text{lof}}(\text{spec})}{S_{\text{lof}}(\text{spec})} \right| \geq \text{SENS.TOL} \ \text{do}
\]

\[
\begin{align*}
&j \leftarrow j + 1; \\
\text{comment: Call POCKET.DOPING to find } N_{\text{p}}^i \text{ of } L-\mu m \text{ pocket MOSFET} \\
&\text{satisfying } I_{\text{lof}}(0.9L) = I_{\text{lof}}(0.9L) = 1 \ nA/\mu m.
\end{align*}
\]

\[
N_{p}^{i+1} \leftarrow \text{POCKET.DOPING}(N_c^i, L, L_p, N_p^i, \Delta N_p^i);
\]

**comment:** The index \( i \) is determined by

the procedure POCKET.DOPING.

\[
\Delta N_{p}^{i+1} \leftarrow \Delta N_p^i; \ I_{\text{lof}}^i(0.9L) \leftarrow I_{\text{lof}}^i(0.9L);
\]

Compute \( I_{\text{lof}}^j(L) \) with given \( N_{\text{p}}^i, N_{c}^i, L_p, \) and \( L \) for \( j \geq 1; \)

\[
S_{\text{lof}}^j \leftarrow \frac{I_{\text{lof}}^j(0.9L) - I_{\text{lof}}^j(L)}{I_{\text{lof}}^j(L)};
\]

\[
\textbf{if} \ S_{\text{lof}}^j < S_{\text{lof}}(\text{spec}) \ \text{then}
\]

\[
\begin{align*}
&\textbf{begin} \\
&\textbf{if} \ S_{\text{lof}}^{j-1} > S_{\text{lof}}(\text{spec})\ \text{then}
\end{align*}
\]
\begin{verbatim}
begin
\[ N^j_{c} + \Delta N^j_{c}/2: \Delta N^j_{c} + \Delta N^j_{c}/2: S^j_{t,off} \leftarrow S^j_{t,off}; \]
end

else
begin
\[ N^j_{c} + \Delta N^j_{c}: \Delta N^j_{c} + \Delta N^j_{c}/2: S^j_{t,off} \leftarrow S^j_{t,off}; \]
end
end
else
begin
\textbf{if} \ [S^{j}_{t,off} > S^{j}_{t,off(spec)}] \ \textbf{AND} \ [S^{j-1}_{t,off} < S^{j}_{t,off(spec)}] \ \textbf{AND} \ [S^{j-1}_{t,off} > 0.] \ \textbf{then}
begin
\[ N^j_{c} + \Delta N^j_{c}/2: \Delta N^j_{c} + \Delta N^j_{c}/2: S^j_{t,off} \leftarrow S^j_{t,off}; \]
end
else
begin
\[ N^j_{c} + \Delta N^j_{c}: \Delta N^j_{c} + \Delta N^j_{c}/2: S^j_{t,off} \leftarrow S^j_{t,off}; \]
end
\end/*end of while loop*/
\end{verbatim}
Using the Medici program implementing the algorithms of the procedures CENTER_POCKET_DOPING and POCKET_DOPING, we find \( N_p \) and \( N_c \) of the 0.1-\( \mu \)m pocket n-MOSFETs satisfying the specifications of \( I_{off} \leq 1 \) nA/\( \mu \)m and \( \Delta I_{off}/I_{off} = 40 \% \) with 10 \% decrease in \( L \). Then, the values of \( N_p \) and \( N_c \) for each \( L_p \) make up the lower design-space boundary for the off-state current and the sensitivity of off-state current to channel length. Figure 3.7 shows the results for fourteen 0.1-\( \mu \)m pocket n-MOSFETs with \( L_p = 0.02 \) – 0.045 \( \mu \)m.

The algorithm above can be summarized using functional expressions as follows. The off-state current is a function of \( L_p \), \( N_p \), and \( N_c \) when all other parameters are held constant. Setting the specification \( I_{off}^{\text{spec}}(0.9L) = I(L_p, N_c, N_p) \) in the inner loop (the procedure POCKET_DOPING) of the procedure CENTER_POCKET_DOPING determines a function \( N_p(N_c, L_p) \). Then setting the sensitivity specification \( S_{I_{off}^{\text{spec}}} = S(L_p, N_c, N_p) = S[L_p, N_c, N_p(N_c, L_p)] = S(L_p, N_c) \) in the outer loop determines the function \( N_c(L_p) \). Combining this result with \( N_p(N_c, L_p) \) from the inner loop we find \( N_p[N_c(L_p), L_p] = N_p(L_p) \). The two functions \( N_p(L_p) \) and \( N_c(L_p) \) are plotted in Figure 3.7 (a) and (b).

The lower design-space boundary for the 0.1-\( \mu \)m pocket n-MOSFETs in Figure 3.7 (a) and (b) is unique for \( I_{off}(0.9L) = 1 \) nA/\( \mu \)m, \( I_{off}(L) = 0.714 \) nA/\( \mu \)m,
Figure 3.7: (a) $N_p$ vs. $L_p$ for the 0.1-$\mu$m pocket n-MOSFETs. $N_c$ in this figure is taken from Figure (b) below. (b) $N_c$ vs. $L_p$ for the 0.1-$\mu$m pocket n-MOSFETs. $N_p$ in this figure is taken from Figure (a) above. All the 0.1-$\mu$m pocket devices located at the curves, i.e., at the unique lower design-space boundary satisfy the design specifications of $I_{off} \leq 1\ nA/\mu m$ and $\frac{\Delta I_{eff}/I_{eff}}{\Delta L/L} = 40%/10\%$. The lines in the figures simply connect the Medici data points.
\[
\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40%/10\%, \quad L = 0.1 \mu m, \quad t_{ox} = 4 nm \quad \text{and} \quad r_j = 0.06 \mu m.
\]
The plots in Figures 3.5 and 3.6 confirm the uniqueness by showing that there is only one set of doping densities \(N_p\) and \(N_c\) for a given \(L_p\) and the boundary specification of the channel-length sensitivity of off-state current, \(40%/10\%\). In other words, whenever any boundary specifications of the sensitivity and the off-state current are given, the values of the design variables, \((N_p, N_c, L_p)\), for a lower design-space boundary are uniquely determined by the algorithms CENTER.POCKET.DOPING and POCKET.DOPING. Two 0.1-\(\mu m\) pocket devices with different pocket lengths can have the same doping levels at their center regions as shown in Figure 3.7 (b). It is because the two pocket devices exist at a crossing point at a low sensitivity region in the plots of the variation of the channel-length sensitivity of \(I_{off}\) with \(N_c\) as shown in Figure 3.5. However, the two pocket n-MOSFETs must have different pocket doping levels because it is impossible physically for the two pocket devices with the identical \(N_c\) and the different \(L_p\) to meet \(I_{off}(0.9L) = 1 nA/\mu m\) with the same \(N_p\).

A uniformly doped n-MOSFET is a structure simplified from the pocket n-MOSFET. Thus, when \(L_p = 0\) or \(L_p = L/2\), the pocket device is reduced to the conventional device. However, there is no 0.1-\(\mu m\) conventional n-MOSFET located at the lower design-space boundary in Figure 3.7. Two-dimensional device simulations show that 0.1-\(\mu m\) conventional devices with the fixed \(t_{ox} = 4 nm\) and \(r_j = 0.06 \mu m\) never meet both the specifications of \(\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40%/10\%\)
and $I_{\text{off}} = 0.714 \text{ nA/} \mu\text{m}$ for the lower design-space boundary. According to the two-dimensional device simulations, only one conventional n-MOSFET with $N_A = N_p = N_c = 7.297 \times 10^{17} \text{ cm}^{-3}$ has $I_{\text{off}} = 0.714 \text{ nA/} \mu\text{m}$ but $\frac{\Delta I_{\text{off}}}{\Delta L} = 630\% / 10\%$. Thus, the conventional device cannot satisfy both boundary specifications of $\frac{\Delta I_{\text{off}}}{\Delta L} = 40\% / 10\%$ and $I_{\text{off}}(L) = 0.714 \text{ nA/} \mu\text{m}$. To locate the uniformly doped 0.1-$\mu\text{m}$ conventional n-MOSFET with $N_A = 7.297 \times 10^{17} \text{ cm}^{-3}$ for $I_{\text{off}} = 0.714 \text{ nA/} \mu\text{m}$ at the design-space boundary in Figure 3.7, the constraint on the channel-length sensitivity of off-state current has to be released.

3.2.2 Unique Lower Design-Space Boundary and Types of Deep-Submicron Pocket n-MOSFETs

To explain why at the lower design-space boundary for $I_{\text{off}}$ in Figure 3.7 (a) $N_p$ decreases as $L_p$ increases, consider a solution of the one-dimensional Poisson's equation. We can derive the following expression for the junction depletion depth $W_D$ inside the drain-end pocket region. Thus,

\begin{align*}
W_D &= \sqrt{2L_D \sqrt{\epsilon_0 (\psi_{bi} + V_{DS})}} - 1 \\
L_D &= \sqrt{\frac{\epsilon_0 V_{th}}{q N_p}} \\
\psi_{bi} &= \psi_{B(D)} + V_{th} \ln \left( \frac{N_p}{n_i} \right)
\end{align*}

(3.11) (3.12) (3.13)

where $L_D$ is the extrinsic Debye length of the pocket, $\psi_{bi}$ is the built-in potential across the $pn$ junction of the drain, $\psi_{B(D)}$ is the bulk potential at the drain region,
$\beta = 1/V_{th}$ and $V_{th}$ is the thermal voltage. Equation 3.11 implies that as $N_p$ drops at a fixed $L_p$, the junction-depletion depth inside the pocket increases, and then more drain-field lines penetrate into the center region. Additionally, the electron mobility becomes higher as $N_p$ drops. These mechanisms lead to an increase in $I_{off}$. Therefore, as $N_p$ is decreased, $L_p$ must be increased in order to keep $I_{off}$ constant at the design-space boundary by reducing the penetration of drain-field lines into the center region.

Figure 3.8: (a) Minima of surface potential distributions, i.e., the highest potential barriers to electron injection from source of the 0.1-$\mu$m pocket n-MOSFETs located at lower design-space boundary for $I_{off} = 0.714$ nA/$\mu$m and $\Delta I_{off}/I_{off(L)} = 40\% / 10\%$, and (b) locations at which the highest potential barriers exist. $V_{GS} = 0$ V and $V_{DS} = 1$ V. (a), (b), (c) and (d) in these figures mean the 0.1-$\mu$m pocket devices (a), (b), (c) and (d) in Figure 3.7, respectively.

Figure 3.8 shows the minimum surface potential $\psi_{s(min)}$, which is the highest potential barrier to electron injection from the source region, and the location $x_m$
where the minimum of the surface potential distribution exists in the 0.1-\(\mu\)m pocket n-MOSFETs located at the lower design-space boundary. \(x_m\) is measured from the source end \(x = 0\) at the Si surface. \(\psi_{s(min)}\) and \(x_m\) depend on doping densities, geometrical parameters of pocket n-MOSFETs, drain and gate bias voltages. \(\psi_{s(min)}\) directly determines the off-state current in the 0.1-\(\mu\)m pocket n-MOSFETs, and the threshold voltage when \(\psi_{s(min)} = V_{th} \ln \left( \frac{N_x}{n_i} \right)\) or \(\phi_{s(min)} = 2V_{th} \ln \left( \frac{N_x}{n_i} \right)\) [4], [31]. The 0.1-\(\mu\)m pocket n-MOSFETs located at the lower design-space boundary have not only the same \(\psi_{s(min)}\) but the same \(I_{off} = 0.714\) nA/\(\mu\)m. Figure 3.8 (b) shows that \(x_m\) increases as \(L_p\) is increased. The distance \(x_m\) from the source is shorter than \(L_p\) but comparable to \(L_p\). Thus, \(x_m\) of the typical pocket n-MOSFET is always located within the source-end pocket region. As discussed in Section 3.1, \(\psi_{s(min)}\) and \(x_m\) are not very sensitive to the drain voltage, which means that short-channel effects are controlled at the source-end pocket region. Figure 3.9 shows subthreshold characteristics for \(V_{DS} = 1\) V of the 0.1-\(\mu\)m pocket n-MOSFETs (a), (b), (c) and (d) located at the design-space boundary in Figure 3.7 (a). As the pocket length is increased, the subthreshold current for \(V_{GS} > 0\) V increases and the threshold voltage decreases slightly, while the off-state current for \(V_{GS} = 0\) V of all the pocket devices are the same.

Figure 3.10 and 3.11 show potential contours and depletion-region edges for \(V_{GS} = 0\) V and \(V_{DS} = 1\) V inside the 0.1-\(\mu\)m pocket n-MOSFETs (a), (d) and (e) in Figure 3.7 (a). After examining Figures 3.7 – 3.11, a qualitative model for the
deep-submicron pocket n-MOSFETs located within the viable design space can be built based on the comparison of the pocket length with the drain-junction depletion depth inside the drain-end pocket and the location of the highest potential barrier to electron injection from source. Figure 3.10 suggests that there are two types of deep-submicron pocket n-MOSFETs: one type named Type A of the device (a), and the other named Type B of the device (d).

3.2.3 Type A Devices

Type A device has a pocket length shorter than the depletion depth $W_D$ (Equation 3.11) of the drain junction. As a result, the drain-field lines penetrate into
Figure 3.10: Potential contours for $V_{GS} = 0\, \text{V}$ and $V_{DS} = 1\, \text{V}$ inside the 0.1-\textmu m pocket n-MOSFETs located at (a) and (d) in Figure 3.7 (a). Dashed lines are depletion-region edges.
Figure 3.11: Potential contours for $V_{GS} = 0$ V and $V_{DS} = 1$ V inside the 0.1-$\mu$m pocket n-MOSFET located at (e) in Figure 3.7 (a). Dashed line is a depletion-region edge.

the center region. The amount of the penetration of the drain-field lines depends strongly on $N_c$ and $L_c (= L - 2L_p)$. The gate control region of the 0.1-$\mu$m pocket n-MOSFET (a) becomes almost a triangle, wide at the interface between the center and the drain-end pocket regions but tapering to a point in the depletion layer of the source-end pocket region. Such a structure appears to suffer from the Yau charge-sharing model [59], and would have a threshold voltage at the center region lowered a great deal by the influence of the drain. However, the penetration of the drain-field lines is weakened significantly within the strong pocket around the source, and the minimum of the surface potential distribution or the highest
potential barrier to electron injection exists at the source-end pocket. Therefore, the control of the current as well as short-channel effects actually takes place in the source-end pocket of the device. In Type A device, the doping densities at the center and pocket regions need to drop as the pocket length increases at the fixed channel length as shown in Figure 3.7 because the pocket shields the drain, and the device requires the drain-field penetration into the center region in order to maintain the same $\psi_{s\langle\text{min}\rangle}$ and $I_{\text{off}}$ at the lower design-space boundary.

3.2.4 Type B Devices

On the other hand, in Type B device (d), the penetration of the drain-field lines into the center region does not occur because the device has $L_p$ longer than or comparable to $W_D$ within the drain-end pocket. The gate controls the entire center region and portions of the source- and drain-end pockets. The highest potential barrier to electron injection still exists at the source-end pocket region in this type of structure, too. As shown in Figure 3.7 (b), $N_c$ increases as $L_p$ increases in Type B device because $N_p$ decreases and the average doping must be maintained to keep the same $\psi_{s\langle\text{min}\rangle}$ and $I_{\text{off}}$ at the lower design-space boundary.

The 0.1-$\mu$m pocket n-MOSFET (e) in Figure 3.11 are situated near the minimum in $N_c$ from Figure 3.7 (b). This device is right at the boundary between Type A and Type B.
3.3 Comments on Design Philosophy for 0.1-$\mu$m Pocket n-MOSFETs

The lower design-space boundary and a qualitative model for the 0.1-$\mu$m pocket n-MOSFETs were presented in the preceding section. Figure 3.7 (a) shows that $N_p$ decreases as $L_p$ increases in order to satisfy the design specifications. The reason for this inverse relationship between these parameters was explained in Subsection 3.2.2. The $N_c$ vs. $L_p$ plot in Figure 3.7 (b) exhibits double-valued behavior and a minimum value for $N_c$. The reason for this behavior for the deep-submicron pocket n-MOSFETs was explained in Subsections 3.2.3 and 3.2.4. The gate of Type A device controls the source-end pocket region while the gate of Type B device controls the entire center region and portions of the source- and drain-end pockets. As $L_p$ increases in Figures 3.7 and 3.10, we are forced to switch from Type A to Type B because as $L_p$ increases, $L_p$ becomes comparable to or longer than the junction-depletion depth $W_D$ inside the drain-end pocket region and the penetration of the drain-field lines into the center region decreases.

The interdependence of $N_p$, $N_c$ and $L_p$ in Figure 3.7 originated from the specifications of the off-state current and the sensitivity of off-state current to channel length. We can find the $N_p$ vs. $L_p$ for a fixed $N_c$ of typical deep-submicron pocket n-MOSFETs like the curve in Figure 3.7 (a). However, the $N_c$ vs. $L_p$ in Figure 3.7 (b) resulted solely from the specification of the sensitivity of off-state current to channel length. Further discussion is made in Chapter 4.
3.4 Summary

The heavily doped pocket implants around drain and source of pocket n-MOSFETs suppress short-channel effects, and enable MOSFET dimensions to be scaled down to deep-submicron or sub-100 nm regime. Computer algorithms were developed on the basis of device physics and two-dimensional device simulation results for finding automatically and accurately the lower design-space boundary for $I_{off}$ and $\frac{\Delta I_{off}/I_{off}}{\Delta L/L}$ of deep-submicron pocket n-MOSFETs, and implemented in the Medici programming language. The lower design-space boundary for the 0.1-μm pocket devices is unique for given boundary specifications of the off-state current and the channel-length sensitivity of off-state current. A qualitative model for the 0.1-μm pocket n-MOSFETs was also provided.

This chapter has been devoted to the short-channel effects, and the exploration of the 0.1-μm pocket n-MOSFETs satisfying both of the design specifications of $I_{off} \leq 1 \, nA/\mu m$ and $\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40%/10\%$. Comparisons between Type A and Type B pocket devices will be made in Chapter 6. Also, advantages of the 0.1-μm pocket n-MOSFETs located at the design-space boundary over 0.1-μm conventional bulk n-MOSFETs will be discussed in Chapter 6.

In Chapter 4, analytical models for the off-state current and the subthreshold surface potential are provided. The complete model equations, including some analytical and empirical formulas for model parameters, will be used to generate
surface potential profiles, off-state currents and the lower design-space boundary for the 0.1-\(\mu m\) pocket n-MOSFETs fast and correctly.
CHAPTER 4

MODELING OF OFF-STATE CURRENT

In Section 4.1, the model for the subthreshold surface potential of the deep-submicron pocket n-MOSFETs for $V_{GS} = 0 \, V$ and $V_{DS} = 0 \, V$ is developed from a quasi-two-dimensional Poisson’s equation, and comparisons of model results to Medici data are presented. The surface potential of the deep-submicron pocket MOSFETs depends strongly on the position $x$, the characteristic (or decay) lengths $\ell_{p0}$ and $\ell_{c0}$ at the pocket and the center regions, and a potential parameter $\mathcal{K}_p$ at the pocket region. Analytical and empirical formulas for $\ell_{p0}$, $\ell_{c0}$ and $\mathcal{K}_p$ are proposed and verified with optimal data of the parameters extracted by fitting model surface potential profiles to Medici surface potential profiles. Section 4.2 presents the model for the subthreshold surface potential of the deep-submicron pocket n-MOSFETs for $V_{GS} = 0 \, V$ and $V_{DS} > 0 \, V$ developed from the quasi-two-dimensional Poisson’s equation. The surface potential model is used to evaluate the effect of DIBL due to the penetration of field lines from drain junction into the center or the pocket region of the deep-submicron pocket devices for $V_{DS} > 0 \, V$. Also, analytical and empirical formulas for $\ell_{pS}$, $\ell_{pD}$ and $\ell_{c}$, which are the characteristic lengths at the source-end pocket, the drain-end pocket and the center regions, respectively, as parameters of the subthreshold surface potential model for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$, are
proposed and verified with extracted optimal data of the parameters. In Section 4.3, a subthreshold current model equation for $V_{GS} < V_T$ is derived from the diffusion current component of the transport equation. The subthreshold current model together with the aforementioned surface potential models generates the off-state current at $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$, and $N_p$ vs. $L_p$ and $N_c$ vs. $L_p$ of the lower design-space boundary for the off-state current in the 0.1-$\mu m$ pocket n-MOSFETs. Section 4.4 presents simplification of model equations for the off-state current in the 0.1-$\mu m$ pocket n-MOSFETs and generation of $N_p$ vs. $L_p$ and $N_c$ vs. $L_p$ of the lower design-space boundary for the off-state current using the simple model equations. Finally, this chapter is summarized and conclusions are drawn in Section 4.5.

4.1 Subthreshold Surface Potential Model for $V_{GS} = 0 \, V$ and $V_{DS} = 0 \, V$

A closed form of a solution of the quasi-two-dimensional Poisson's equation can correctly generate subthreshold surface potential profiles of two-dimensional device simulations for the deep-submicron pocket MOSFETs. The subthreshold surface potential model for $V_{GS} = 0 \, V$ and $V_{DS} = 0 \, V$ is necessary to develop a model for the subthreshold surface potential for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$ and an off-state current model to be discussed in the subsequent sections. A general expression for analytical and empirical formulas of model parameters of the subthreshold surface potential and the off-state current models is proposed in this section.
4.1.1 Model Equations and Surface Potential Patterns

Recently, analytical surface potential models [4], [24] based on solutions of the quasi-two-dimensional Poisson’s equation for deep-submicron conventional and pocket MOSFETs have been proposed to predict threshold voltage shift of the deep-submicron devices. The surface potential model for the deep-submicron conventional MOSFET [24] fits two-dimensional device simulations of Medici well with a modification of a model parameter of the characteristic (decay) length [60] depending on doping density and geometrical parameters. However, the surface potential model for the deep-submicron pocket MOSFET [4] neither fits two-dimensional device simulations nor satisfies the boundary condition of the continuity of electric field at the boundary of pocket and center regions. In this dissertation, an improved subthreshold surface potential model based on solutions of the quasi-two-dimensional Poisson’s equation [4, 24, 37] for $V_{DS} = 0 \ V$ of the deep-submicron pocket MOSFETs is developed by considering the following boundary conditions.

- Potential $\psi_{s0}(x)$ for $V_{GS} = 0 \ V$ and $V_{DS} = 0 \ V$ is continuous at $x = 0$, $L_p$, $L - L_p$ and $L$.

- Electric field $-\frac{d\psi_{s0}(x)}{dx}$ is continuous at $x = L_p$ and $L - L_p$.

- $\lim_{L_c \to \infty} \psi_{s0}(x = L/2) = \psi_{s(LC)}$ (long-channel surface potential at center region).

- $\psi_{s0}(0) = \psi_{s0}(L) = \psi_{NPOLY}$
Note that assuming the work function of source/drain contact is identical with that of source/drain region, the surface potentials at \( x = 0 \) and \( L \) are exactly the same as the potential given in Equation 2.17 for the \( n^+ \)-type polycrystalline silicon contact. To obtain an accurate subthreshold surface potential \( \psi_{s0}(x) \) between the source and the drain of the deep-submicron pocket n-MOSFET, we have to solve the two-dimensional Poisson’s equation given in Equation 4.1, neglecting the mobile electron charge in the channel depletion region.

\[
\frac{\partial^2 \psi}{\partial x^2} + \frac{\partial^2 \psi}{\partial y^2} = \frac{q N_A(x, y)}{\epsilon_s}
\]  

(4.1)

where \( N_A(x, y) \) is either \( N_p \) or \( N_c \) at a position. Medici provides the solution of Equation 4.1. To obtain a closed form of an approximate solution \( \psi_{s0}(x) \) of the two-dimensional Poisson’s equation, we can solve a quasi-two-dimensional Poisson’s equation in the channel depletion region. Let \( W_D(x) \) be the depletion-layer depth under gate. First, consider a Gaussian box of height \( W_D(x) \) extending from the Si surface to the depletion-layer edge in the Si bulk and of length \( \Delta x \) in the channel depletion region. Applying Gauss’s law to the Gaussian box, we obtain

\[
\epsilon_s W_D(x) \left( \lim_{\Delta x \to 0} \frac{-|\mathcal{E}_s(x + \Delta x)| + |\mathcal{E}_s(x)|}{\Delta x} \right) - C_{ox}(\psi_G(x) - \psi_s(x)) = -q N_A(x) W_D(x)
\]

(4.2)

where \( N_A(x) \) is either \( N_p \) or \( N_c \). Thus,

\[
\epsilon_s W_D(x) \left( \frac{d|\mathcal{E}_s(x)|}{dx} \right) + C_{ox}(\psi_G(x) - \psi_s(x)) = q N_A(x) W_D(x)
\]

(4.3)
where

\[ |\mathcal{E}_s(x)| = \frac{d\psi_s}{dx} \quad (4.4) \]

\[ \psi_G(x) = V_{GS} - V_{FB}(x) - |\psi_B(x)| \quad (4.5) \]

The first and second terms in Equation 4.2 are the electric fluxes entering the plane of the Gaussian box located at \( x + \Delta x \) and exiting the plane located at \( x \), respectively. The third term \((-C_{ox}(\psi_G(x) - \psi_s(x)))\) in the equation is the electric flux entering the top plane of the Gaussian box. The electric flux or the field passing through the bottom of the Gaussian box located in the Si bulk is zero. The depletion depth \( W_D(x) \) is invariable within either the pocket or the center region. However, an accurate solution of the two-dimensional Poisson's equation (Equation 4.1) reveals that the depletion-layer depth is variable between source and drain even in a uniformly doped deep-submicron MOSFET.

Assuming that \( W_D(x) \) is invariable in each region for fixed values of the doping densities at the pocket and the center regions, the gate voltage and the gate oxide thickness, the general solution of the second-order differential equation of \( \psi_s(x) \) in Equation 4.3 consists of exponential terms as the general solution of the homogeneous differential equation and a constant term as the particular solution of the differential equation in Equation 4.3 at each of the pocket and the center regions.
Thus, the subthreshold surface potential, \( \psi_{s0}(x) \), for \( V_{DS} = 0 \) \( V \) is

\[
\psi_{s0}(x) = \begin{cases} 
K_p + f_1 e^{-x/\ell_{p0}} + f_2 e^{-(L_p - x)/\ell_{p0}}, & 0 \leq x \leq L_p \\
\psi_{s(LC)} + f_3 \cosh \left( \frac{L - L_p - x}{\ell_{c0}} \right), & L_p \leq x \leq L - L_p \\
K_p + f_1 e^{-(L - x)/\ell_{p0}} + f_2 e^{-(L_p - L + x)/\ell_{p0}}, & L - L_p \leq x \leq L
\end{cases}
\] (4.6)

where \( \psi_{s(LC)} \) is the long-channel surface potential at the center region calculated from Equations 3.4 and 3.7 in Chapter 3, and

\[
\ell_{p0} = \sqrt{\frac{\epsilon_s}{\epsilon_{ox}} W_{D(p)}} \quad (4.7)
\]

\[
\ell_{c0} = \sqrt{\frac{\epsilon_s}{\epsilon_{ox}} W_{D(c)}} \quad (4.8)
\]

\[
K_p = \psi_{s(LC)}(p) \quad (4.9)
\]

The parameters \( \ell_{p0} \) and \( \ell_{c0} \) are the characteristic (decay) lengths for \( V_{DS} = 0 \) \( V \) at the pocket and the center regions, respectively. \( \psi_{s(LC)}(p) \) is the long-channel surface potential at the pocket region. \( W_{D(p)} \) and \( W_{D(c)} \) are the depletion-layer depths under gate at the pocket and the center regions, respectively.

Applying all the boundary conditions described above to Equation 4.6, we can find \( f_1, f_2 \) and \( f_3 \) in Equation 4.6 as follows.

\[
f_2 = \frac{\psi_{s(LC)} - K_p}{2} \quad (4.10)
\]

\[
+ \frac{1}{2} \left[ (\psi_{NPOLY} - K_p)/ \cosh \left( \frac{L_p}{\ell_{p0}} \right) - (\psi_{s(LC)} - K_p) \right] \left[ 1 - \left( \frac{\ell_{p0}}{\ell_{c0}} \right) \tanh \left( \frac{L_p}{2\ell_{c0}} \right) \right]
\]

\[
f_1 = \psi_{NPOLY} - K_p - f_2 e^{-L_p/\ell_{p0}} \quad (4.11)
\]

\[
f_3 = \frac{(\psi_{NPOLY} - K_p)/ \cosh \left( \frac{L_p}{\ell_{p0}} \right) - (\psi_{s(LC)} - K_p)}{\cosh \left( \frac{L_p}{2\ell_{c0}} \right) + \left( \frac{\ell_{p0}}{\ell_{c0}} \right) \sinh \left( \frac{L_p}{2\ell_{c0}} \right) \tanh \left( \frac{L_p}{\ell_{p0}} \right)} \quad (4.12)
\]
Equations 4.6 - 4.12 completely describe the surface potential model for $V_{DS} = 0 \, V$ of short-channel as well as long-channel pocket n-MOSFETs. However, $\ell_{p0}$, $\ell_{c0}$ and $K_p$ in Equations 4.7 - 4.9 are not accurate for deep-submicron pocket devices because $W_{D(p)}$ and $W_{D(c)}$ are not constant at the pocket and the center regions. Also, surface potential profiles between source and drain of 0.1-$\mu$m pocket n-MOSFETs generated by the subthreshold surface potential model for $\psi_{s0}(x)$ do not agree with Medici simulation results. Therefore, $\ell_{p0}$, $\ell_{c0}$ and $K_p$ are treated as model fitting parameters, yet the physical meanings of the model parameters in Equations 4.7 – 4.9 are still effective. It is important to find accurate values of the model parameters to generate a correct pattern of the potential distribution between source and drain and to develop accurate formulas for the model parameters. An algorithm to extract the optimal values of the model parameters $\ell_{p0}$, $\ell_{c0}$ and $K_p$ accurately and fast by fitting surface potential profiles generated by the model for $\psi_{s0}(x)$ in Equations 4.6 and 4.10 – 4.12 to Medici surface potential profiles was developed by combining dynamic adjustments of solution brackets with the secant method [61]. The algorithm is described in Appendix C.

A single-device value of a model parameter in this chapter is an optimal value, which is extracted by fitting a model surface potential profile to a Medici surface potential profile. Figure 4.1 shows two patterns of the potential distribution of 0.1-$\mu$m and 0.2-$\mu$m pocket n-MOSFETs, and comparisons of the model surface potential profiles to the Medici surface potential profiles. The pocket MOSFET
Figure 4.1: Surface potential distributions between source and drain of (a) 0.1-\(\mu m\) pocket n-MOSFET with \(N_c = 7.0 \times 10^{17} \text{ cm}^{-3}\), \(N_p = 1.281 \times 10^{18} \text{ cm}^{-3}\) and \(L_p = 0.025 \mu m\), and (b) 0.2-\(\mu m\) pocket n-MOSFET with \(N_c = 2.5 \times 10^{17} \text{ cm}^{-3}\), \(N_p = 6.075 \times 10^{17} \text{ cm}^{-3}\) and \(L_p = 0.05 \mu m\). \(t_{ox} = 4 \text{ nm}\), \(r_j = 0.06 \mu m\), \(V_{GS} = 0 \text{ V}\) and \(V_{DS} = 0 \text{ V}\) for both pocket devices. Single-device values of \(\ell_p0\), \(\ell_e0\) and \(K_p\) were used for the model.

with a surface potential minimum in the center region shown in Figure 4.1 (a) is controlled by the center region, while the device with surface potential minima in the pocket regions shown in Figure 4.1 (b) is controlled by the pocket region.

Next, we observe the patterns of potential distributions of the 0.1-\(\mu m\) pocket n-MOSFET structures satisfying the specifications of \(I_{off} \leq 1 \text{ nA}/\mu m\) and \(\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = \frac{1075}{10^{10}}\) and located at the lower design-space boundary in Figure 3.7, and compare the model to the Medici surface potentials. From the results in Figure 4.2 for the 0.1-\(\mu m\) pocket devices (a), (b), (c) and (d) located at the design-space boundary,
Figure 4.2: Surface potential $\psi_{so}$ profiles between source and drain of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0$ V and $V_{DS} = 0$ V.  
(a) $L_p = 0.022$ $\mu$m, $N_p = 2.135 \times 10^{18}$ $cm^{-3}$, $N_c = 2.375 \times 10^{17}$ $cm^{-3}$,  
(b) $L_p = 0.028$ $\mu$m, $N_p = 1.558 \times 10^{18}$ $cm^{-3}$, $N_c = 1.925 \times 10^{17}$ $cm^{-3}$,  
(c) $L_p = 0.034$ $\mu$m, $N_p = 1.205 \times 10^{18}$ $cm^{-3}$, $N_c = 1.825 \times 10^{17}$ $cm^{-3}$, and  
(d) $L_p = 0.044$ $\mu$m, $N_p = 8.557 \times 10^{17}$ $cm^{-3}$, $N_c = 2.138 \times 10^{17}$ $cm^{-3}$.  
$tox = 4$ $nm$ and $r_j = 0.06$ $\mu$m for all the pocket devices.  
Single-device values of $\ell_{po}$, $\ell_{co}$ and $K_p$ were used for the model.
we know that all the acceptable pocket MOSFETs are controlled by the pocket region, and the minimum surface potentials $\psi_{s0}(\text{min})$ always exist at the pocket regions while the maximum surface potentials $\psi_{s0}(\text{max})$ always exist at the center regions.

The values of $N_p$, $N_c$ and $L_p$ in Figure 4.2 are taken from the curves in Figure 3.7. Notice that after applying the algorithm in Appendix C for finding the single-device values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$ of the twelve 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary by minimizing the errors of the surface potentials, $\psi_{s0}(L/6)$, $\psi_{s0}(\text{min})$ and $\psi_{s0}(\text{max})$, between the model and Medici results to a parameter extraction process, each of the relative errors of the model surface potentials, $\psi_{s0}(L/6)$, $\psi_{s0}(\text{min})$ and $\psi_{s0}(\text{max})$, with reference to the corresponding Medici surface potentials is within 0.1%. The extracted optimal data of the model parameters $\ell_{p0}$, $\ell_{c0}$ and $K_p$ for the 0.1-$\mu$m pocket MOSFETs with $L_p = 0.022 - 0.044$ $\mu$m vs. a new parameter $z = LL_pN_pN_c$ [cm$^{-4}$] are shown on logarithmic scales in Figure 4.3, and the values are listed in Table 4.1. These extracted optimal data of the parameters will be used as the single-device values in order to formulate the analytical and empirical formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$.

### 4.1.2 Analytical and Empirical Formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$

The surface potential in deep-submicron pocket MOSFETs for $V_{GS} = 0$ V and $V_{DS} = 0$ V is a function of the position $x$, the characteristic lengths $\ell_{p0}$ and $\ell_{c0}$, and the fitting potential parameter $K_p$ in the pocket region for fixed values of structural
<table>
<thead>
<tr>
<th>$L_p$ [µm]</th>
<th>$N_p$ [cm$^{-3}$]</th>
<th>$N_c$ [cm$^{-3}$]</th>
<th>$z$ [cm$^{-4}$]</th>
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<tr>
<td>0.022</td>
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<tr>
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</table>

<table>
<thead>
<tr>
<th>$L_p$ [µm]</th>
<th>$\ell_{p0}$ [µm]</th>
<th>$\ell_{c0}$ [µm]</th>
<th>$K_p$ [V]</th>
</tr>
</thead>
<tbody>
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<td>0.022</td>
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<td>$-7.159 \times 10^{-2}$</td>
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Table 4.1: Structural parameters and $z = LL_p N_p N_c$ of the 0.1-µm pocket nMOSFETs located at the lower design-space boundary, and the single-device values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$ of the pocket devices. $V_{DS} = 0$ V and $V_{GS} = 0$ V.
parameters of the pocket devices. Thus,

$$\psi_{s0}(x) = f(x, \ell_p0, \ell_c0, K_p)$$

(4.13)

$\ell_p0$, $\ell_c0$ and $K_p$ described in the preceding subsection are the important parameters of $\psi_{s0}(x)$ and should be formulated as accurately as possible in order to correctly predict the surface potential distribution and the off-state current to be discussed in a subsequent section. Global parameter values are calculated from the analytical and empirical formulas for the model parameters. Such a formula for each model parameter is fitted to the single-device values available for the model parameter. To
approximate closed-form formulas to the single-device values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$, the analytical and empirical formulas, which are based on logarithmic and exponential functions, and the approximation theory [62] of numerical analysis, are proposed. Let the parameter $z = LL_pN_pN_c$ with a dimension of $[cm^{-1}]$ be a new composite variable of the empirical formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$. Then, the following is the analytical and empirical formula proposed for a parameter $\ell$ ($\ell = \ell_{p0}$, $\ell_{c0}$ or $K_p$).

\[
\ell = 10^{\log_{10} \ell} = \alpha z^{\beta_0 (\log_{10} z)^{n-1}} z^{\beta_1 (\log_{10} z)^{n-2}} z^{\beta_2 (\log_{10} z)^{n-3}} \ldots z^{\beta_{n-1}} \quad (4.14)
\]

where $\beta_0, \ldots, \beta_{n-1}$ and $\alpha$ are real numbers. Evidently the parameter $\ell$ in Equation 4.14 is an overall fit to the single device parameter, and usually will not provide as good a surface potential for a selected device as the single device parameter. We call the parameter in Equation 4.14 a global parameter.

The function in Equation 4.14 is transformed (mapped) into a polynomial of degree $n$ in terms of $\log_{10} z$ by taking the logarithmic function for both sides of the equation.

\[
\log_{10} \ell = \beta_0 (\log_{10} z)^n + \beta_1 (\log_{10} z)^{n-1} + \beta_2 (\log_{10} z)^{n-2} + \ldots + \beta_{n-1} (\log_{10} z) + \log_{10} \alpha \quad (4.15)
\]

The higher the degree of the polynomial, the better fit to the single-device values. Thus, for $n + 1$ single-device values, the $n$th-degree polynomial will contain all the $n + 1$ single-device values. However, there exist possibilities of overflow or underflow in finding the values of the coefficients $\beta_0, \ldots, \beta_{n-1}$ and $\log_{10} \alpha$ of a higher degree.
polynomial of \( \log_{10} z \) in Equation 4.15, e.g., the degree \( n > 5 \). Therefore, the following second-degree polynomials of \( \log_{10} z \) are tried to fit \( \ell_{p0} \), \( \ell_{c0} \) and \( K_p \) to the corresponding set of the single-device values. The global values generated by the formulas based on the second-degree polynomials of \( \log_{10} z \) for the model parameters will prove to be in good agreement with the corresponding single-device values.

\[
\log_{10} \ell_{p0} = \beta^0_{p0}(\log_{10} z)^2 + \beta^1_{p0}(\log_{10} z) + \log_{10} \alpha_{p0} \tag{4.16}
\]
\[
\log_{10} \ell_{c0} = \beta^0_{c0}(\log_{10} z)^2 + \beta^1_{c0}(\log_{10} z) + \log_{10} \alpha_{c0} \tag{4.17}
\]
\[
\log_{10} \vert K^p \vert = \beta^0_{K^p}(\log_{10} z)^2 + \beta^1_{K^p}(\log_{10} z) + \log_{10} \alpha_{K^p} \tag{4.18}
\]

Now, \( \ell_{p0} \), \( \ell_{c0} \) and \( K_p \) are directly obtained by the inverse transformation of \( \log_{10} \ell_{p0} \), \( \log_{10} \ell_{c0} \) and \( \log_{10} \vert K^p \vert \) into \( \ell_{p0} \), \( \ell_{c0} \) and \( K_p \). Thus,

\[
\ell_{p0} = 10^{\log_{10} \ell_{p0}} = \alpha_{p0} z^{\beta^0_{p0}(\log_{10} z)} z^{\beta^1_{p0}} \tag{4.19}
\]
\[
\ell_{c0} = 10^{\log_{10} \ell_{c0}} = \alpha_{c0} z^{\beta^0_{c0}(\log_{10} z)} z^{\beta^1_{c0}} \tag{4.20}
\]
\[
K_p = -10^{\log_{10} \vert K^p \vert} = -\alpha_{K^p} z^{\beta^0_{K^p}(\log_{10} z)} z^{\beta^1_{K^p}} \tag{4.21}
\]

Note that the first equations in Equations 4.19 - 4.21 are recommended in evaluating the model parameters in order to avoid possible overflow or underflow in using the second equations in Equations 4.19 - 4.21. Using a MATLAB function [63] implemented by utilizing a technique of polynomial regression [64] for computing a polynomial to fit the single-device values on the logarithmic scale by minimizing the distance of the corresponding values of the polynomial from the single-device data.
points, the coefficients of the second-degree polynomials of Equations 4.16, 4.17 and 4.18 are found, and tabulated in Tables 4.2, 4.3 and 4.4. Note that each

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<th></th>
<th>Type A</th>
<th>Type B</th>
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<tr>
<td></td>
<td>(0.022 ( \mu m ) \leq L_p \leq 0.036 ( \mu m ))</td>
<td>(0.036 ( \mu m ) \leq L_p \leq 0.044 ( \mu m ))</td>
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<td>(-8.994788 \times 10^2)</td>
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<td>(-1.611809)</td>
<td>(-1.415758)</td>
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<td>( \beta_{p0}^1 )</td>
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<td>(7.115284 \times 10)</td>
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Table 4.2: The coefficients of the second-degree polynomial of \( \log_{10} \ell_{p0} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs.

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<tr>
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<th>Type A</th>
<th>Type B</th>
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<tbody>
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<td>( \beta_{c0}^1 )</td>
<td>(2.336033 \times 10^2)</td>
<td>(-1.413567 \times 10^2)</td>
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Table 4.3: The coefficients of the second-degree polynomial of \( \log_{10} \ell_{c0} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs.

<table>
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<tr>
<th></th>
<th>Type A</th>
<th>Type B</th>
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<tbody>
<tr>
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<td>(0.034 ( \mu m ) \leq L_p \leq 0.044 ( \mu m ))</td>
</tr>
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<td>( \log_{10} \alpha_{Kp} )</td>
<td>(-1.645301 \times 10^4)</td>
<td>(1.268558 \times 10^5)</td>
</tr>
<tr>
<td>( \beta_{Kp}^0 )</td>
<td>(-2.649795 \times 10)</td>
<td>(2.044801 \times 10^2)</td>
</tr>
<tr>
<td>( \beta_{Kp}^1 )</td>
<td>(1.320538 \times 10^3)</td>
<td>(-1.018621 \times 10^4)</td>
</tr>
</tbody>
</table>

Table 4.4: The coefficients of the second-degree polynomial of \( \log_{10} |K_p| \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs.

set of the single-device values of \( \ell_{p0} \), \( \ell_{c0} \) and \( K_p \) as well as all single-device values of the 0.1-\( \mu m \) pocket n-MOSFETs to be discussed in the subsequent sections is divided into two groups because of the two types of the 0.1-\( \mu m \) pocket n-MOSFETs
discussed in the preceding chapter. It should be emphasized that the partition of the model parameters into the two groups resulted solely from the characteristic $N_c$ vs. $L_p$ in Figure 3.7 (b). Figures 4.4 - 4.6 show the second-degree polynomial fits to the sets of the single-device values of $\log_{10} \ell_{p0}$, $\log_{10} \ell_{c0}$ and $\log_{10} |K_p|$ of the 0.1-$\mu$m pocket n-MOSFETs with $0.022 \ \mu m \leq L_p \leq 0.044 \ \mu m$. Now, the surface potential equation of Equation 4.6 in the preceding subsection and Equations 4.16 - 4.21 accurately model any of the surface potential distributions for $V_{GS} = 0 \ \text{V}$ and $V_{DS} = 0 \ \text{V}$ of the 0.1-$\mu$m pocket MOSFETs located at the lower design-space boundary. Note that the model parameters $\ell_{p0}$, $\ell_{c0}$ and $K_p$ are calculated using the first equations in Equations 4.19 - 4.21 after Equations 4.16 - 4.18 are evaluated. Figure 4.7 shows comparisons of the model surface potential profiles with the Medici surface potential profiles of the 0.1-$\mu$m pocket n-MOSFETs (a), (b), (c) and (d) located at the lower design-space boundary in Figure 3.7. The model results were generated by the complete model equations that include the analytical and empirical formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$. For the best one of the surface potential profiles of the twelve 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary, the relative errors of the surface potentials $\psi_{s0}(L/6)$, $\psi_{s0(min)}$ and $\psi_{s0(max)}$ between the Medici and the model results using the global parameters are within 0.14 \%, while being within 3.30 \% for the worst one, cf. within 0.1 \% for the single-device parameters.
Figure 4.4: $\log_{10} \ell_{p0}$ vs. $\log_{10}(LL_p N_p N_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{p0}$ of (a) Type A ($0.022\mu m \leq L_p \leq 0.036\mu m$) and (b) Type B ($0.036\mu m < L_p \leq 0.044\mu m$) of 0.1-$\mu m$ pocket n-MOSFETs.

---

Figure 4.5: $\log_{10} \ell_{c0}$ vs. $\log_{10}(LL_p N_p N_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{c0}$ of (a) Type A ($0.022\mu m \leq L_p \leq 0.034\mu m$) and (b) Type B ($0.034\mu m < L_p \leq 0.044\mu m$) of 0.1-$\mu m$ pocket n-MOSFETs.
Figure 4.6: $\log_{10} |K_p|$ vs. $\log_{10}(L_p N_p N_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} |K_p|$ of (a) Type A ($0.022 \mu m < L_p < 0.034 \mu m$) and (b) Type B ($0.034 \mu m < L_p < 0.044 \mu m$) of 0.1-$\mu m$ pocket n-MOSFETs.

4.2 Subthreshold Surface Potential Model for $V_{GS} = 0 \ V$ and $V_{DS} > 0 \ V$

The subthreshold surface potential model for $V_{GS} = 0 \ V$ and $V_{DS} > 0 \ V$ is necessary to correctly predict the effects of DIBL in the deep-submicron pocket MOSFETs. A closed form of the potential model for $V_{DS} > 0 \ V$ is used to derive an expression for the location $x_m$ at which the minimum surface potential $\psi_s(\text{min})(x_m)$ exists. The model for $\psi_s(\text{min})(x_m)$ at $V_{GS} = 0 \ V$ and $V_{DS} = 1 \ V$ will be used for the modeling of the off-state current to be discussed in the subsequent section.
Figure 4.7: Surface potential $\psi_{so}$ profiles between source and drain of 0.1-$\mu$m pocket n-MOSFETs which are the same ones used in Figure 4.2 for $V_{GS} = 0 \ V$ and $V_{DS} = 0 \ V$. The model surface potentials were generated by the complete model equations, including the z-formulas for $\ell_{p0}$, $\ell_{c0}$ and $K_p$. (a) $L_p = 0.022 \ \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$, $N_c = 2.375 \times 10^{17} \ cm^{-3}$, (b) $L_p = 0.028 \ \mu m$, $N_p = 1.558 \times 10^{18} \ cm^{-3}$, $N_c = 1.925 \times 10^{17} \ cm^{-3}$, (c) $L_p = 0.034 \ \mu m$, $N_p = 1.205 \times 10^{18} \ cm^{-3}$, $N_c = 1.825 \times 10^{17} \ cm^{-3}$, and (d) $L_p = 0.044 \ \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$, $N_c = 2.138 \times 10^{17} \ cm^{-3}$. $t_{ox} = 4 \ nm$ and $r_j = 0.06 \ \mu m$ for all the pocket devices. Global values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$ were used for the model.
4.2.1 Model Equations and Model Parameter Extraction

Assuming that \( W_D(x) \) is invariable in each region, the general solution given in Equation 4.6 of the quasi-two-dimensional Poisson's equation (Equation 4.3) are used for developing the subthreshold surface potential model for \( V_{GS} = 0 \) \( V \) and \( V_{DS} > 0 \) \( V \). The general solution for the model is given below in a modified form of Equation 4.6.

\[
\psi_s(x) = \begin{cases} 
K_p + g_1 e^{-x/\ell_{ps}} + g_2 e^{-(L_p-x)/\ell_{ps}}, & 0 \leq x \leq L_p \\
\psi_{sc} + g_3 \cosh(L/2-x)/\ell_c + g_4 \sinh(L/2-x)/\ell_c, & L_p \leq x \leq L - L_p \\
K_p + g_5 e^{-(L-x)/\ell_{pd}} + g_6 e^{-(L_p-L+x)/\ell_{pd}}, & L - L_p \leq x \leq L
\end{cases} \tag{4.22}
\]

where \( \ell_{ps} \) and \( \ell_{pd} \) are the characteristic or decay lengths at the pocket implants near source and drain, respectively, and \( \ell_c \) is the characteristic or decay length at the center region for \( V_{DS} > 0 \) \( V \). Note that the general solution in Equation 4.6 cannot be used for the subthreshold surface potential model for \( V_{DS} > 0 \) \( V \) because the surface potential profile for \( V_{DS} > 0 \) \( V \) is not symmetrical and \( \ell_{ps}, \ell_{pd} \) and \( \ell_c \) depend on drain voltage. To obtain a complete solution for the pocket n-MOSFETs, the expressions for \( g_1, \ldots, g_6 \) and \( \psi_{sc} \) in Equation 4.22 should be found by applying the following boundary conditions.

- Potential \( \psi_s(x) \) is continuous at \( x = 0, L_p, L - L_p \) and \( L \).
- Electric field \( -\frac{d\psi_s(x)}{dx} \) is continuous at \( x = L_p \) and \( L - L_p \).
- For \( V_{DS} = 0 \) \( V \), \( \psi_s(L/2) = \psi_{so}(L/2) \).
• \( \lim_{L_c \to \infty} \psi_s(L/2) = \psi_s(L_C) \) (long-channel surface potential at center region).

• \( \psi_s(0) = \psi_{N\text{POLY}} \) and \( \psi_s(L) = \psi_{N\text{POLY}} + V_{DS} \), assuming the resistance in source/drain region is zero.

\( \psi_{sc} \) and \( g_1, \ldots, g_6 \) can be represented by functions of the parameters of the pocket MOSFET. Thus,

\[
\psi_{sc} = f_p(\ell_p, \ell_c, K_p, \psi_s(L_C)) \tag{4.23}
\]

\[
g_i = g_p(\psi_{sc}, \ell_p, \ell_D, \ell_c, K_p, V_{DS}) \tag{4.24}
\]

\[
g_i = g_p(\ell_p, \ell_c, \ell_p, \ell_D, \ell_c, K_p, \psi_s(L_C), V_{DS}), \quad i = 1, \ldots, 6
\]

Applying the continuity of potential at \( x = 0 \), \( L_p \), \( L - L_p \) and \( L \), and the continuity of electric field at \( x = L_p \) and \( L - L_p \) to Equation 4.22, we obtain

\[
g_1 = \psi_{N\text{POLY}} - K_p - g_2 e^{-L_p/t_{PD}} \tag{4.25}
\]

\[
g_2 = \frac{1}{2} \left\{ -K_p + \psi_{sc} + g_3 \left[ \cosh \left( \frac{L_c}{2\ell_c} \right) - \left( \frac{\ell_p}{\ell_c} \right) \sinh \left( \frac{L_c}{2\ell_c} \right) \right] + g_4 \left[ \sinh \left( \frac{L_c}{2\ell_c} \right) - \left( \frac{\ell_p}{\ell_c} \right) \cosh \left( \frac{L_c}{2\ell_c} \right) \right] \right\} \tag{4.26}
\]

\[
g_3 = \cosh \left( \frac{L_c}{2\ell_c} \right) \left[ K_p - \psi_{sc} + \gamma_1 (\psi_{N\text{POLY}} + V_{DS} - K_p) e^{-L_p/t_{PD}} + g_6 \gamma_2 \right] \tag{4.27}
\]

\[
g_4 = \sinh \left( \frac{L_c}{2\ell_c} \right) \left[ K_p - \psi_{sc} + \beta_1 (\psi_{N\text{POLY}} + V_{DS} - K_p) e^{-L_p/t_{PD}} + g_6 \beta_2 \right] \tag{4.28}
\]

\[
g_5 = \psi_{N\text{POLY}} + V_{DS} - K_p - g_6 e^{-L_p/t_{PD}} \tag{4.29}
\]

\[
g_6 = \frac{\psi_{N\text{POLY}} - K_p}{\cosh \left( \frac{L_p}{\ell_p} \right) \cosh^2 \left( \frac{L_c}{2\ell_c} \right)} + \frac{K_p - \psi_{sc}}{\cosh^2 \left( \frac{L_c}{2\ell_c} \right)} \tag{4.30}
\]

\[-\alpha_1 \left[ K_p - \psi_{sc} + \gamma_1 (\psi_{N\text{POLY}} + V_{DS} - K_p) e^{-L_p/t_{PD}} \right]
\]
\[-\alpha_2 \tanh^2 \left( \frac{L_c}{2\ell_c} \right) \left[ K_p - \psi_{sc} + \beta_1 (\psi_{NPOLY} + V_{DS} - K_p) e^{-L_p/t_{p0}} \right] \]

\[
/ \left[ \alpha_1 \gamma_2 + \alpha_2 \beta_2 \tanh^2 \left( \frac{L_c}{2\ell_c} \right) \right]
\]

where

\[
\alpha_1 = 1 + \left( \frac{\ell_{pS}}{\ell_c} \right) \tanh \left( \frac{L_c}{2\ell_c} \right) \tanh \left( \frac{L_p}{\ell_{pS}} \right)
\]  

(4.31)

\[
\alpha_2 = 1 + \left( \frac{\ell_{pS}}{\ell_c} \right) \coth \left( \frac{L_c}{2\ell_c} \right) \tanh \left( \frac{L_p}{\ell_{pS}} \right)
\]  

(4.32)

\[
\beta_1 = 1 - \left( \frac{\ell_c}{\ell_{pD}} \right) \coth \left( \frac{L_c}{2\ell_c} \right)
\]  

(4.33)

\[
\beta_2 = 2 - \beta_1 \left( 1 + e^{-2L_p/t_{pD}} \right)
\]  

(4.34)

\[
\gamma_1 = \beta_1 + \frac{\ell_c/\ell_{pD}}{\sinh \left( \frac{L_c}{2\ell_c} \right) \cosh \left( \frac{L_p}{2\ell_c} \right)}
\]  

(4.35)

\[
\gamma_2 = 2 - \gamma_1 \left( 1 + e^{-2L_p/t_{pD}} \right)
\]  

(4.36)

Finally, we can find $\psi_{sc}$ by applying the boundary condition of for $V_{DS} = 0 V$,

$\psi_s(L/2) = \psi_{s0}(L/2)$.

\[
\psi_{sc} = \frac{\cosh \left( \frac{L_p}{2\ell_{c0}} \right)}{1 + \xi_1^0 \cosh \left( \frac{L_p}{2\ell_{c0}} \right) \cosh \left( \frac{L_{p0}}{2\ell_{c0}} \right)} \left\{ \psi_{s0}(L/2) - K_p \right. \nonumber \\
- \gamma_1^0 (\psi_{NPOLY} - K_p) e^{-L_p/t_{p0}} \nonumber \\
- \frac{1}{\xi_1^0} \left[ \psi_{NPOLY} + K_p \left( \cosh \left( \frac{L_{p0}}{2\ell_{c0}} \right) - 1 \right) \right. \nonumber \\
- \alpha_1^0 (K_p + \gamma_1^0 (\psi_{NPOLY} - K_p) e^{-L_p/t_{p0}}) \nonumber \\
\left. \left. - \alpha_2^0 (K_p + \beta_1^0 (\psi_{NPOLY} - K_p) e^{-L_p/t_{p0}}) \tanh^2 \left( \frac{L_c}{2\ell_{c0}} \right) \right\} \right\}
\]  

(4.37)

where

\[
\xi_1^0 = \alpha_1^0 + \frac{1}{\gamma_2^0} \left[ \alpha_2^0 \beta_2^0 \tanh^2 \left( \frac{L_c}{2\ell_{c0}} \right) \right]
\]  

(4.38)
\[ \xi_2^0 = -1 + \frac{1}{\xi_1^0} \left[ \frac{-1}{\cosh^2 \left( \frac{L_c}{2\ell_{c0}} \right)} + \alpha_1^0 + \alpha_2^0 \tanh^2 \left( \frac{L_c}{2\ell_{c0}} \right) \right] \] (4.39)

\[ \alpha_1^0, \alpha_2^0, \beta_1^0, \beta_2^0, \gamma_1^0 \text{ and } \gamma_2^0 \] are directly obtained by replacing \( \ell_{PS}, \ell_{PD} \) and \( \ell_c \) in the expressions for \( \alpha_1, \alpha_2, \beta_1, \beta_2, \gamma_1 \) and \( \gamma_2 \) with \( \ell_{PS0}, \ell_{PO} \) and \( \ell_{CO} \), respectively. Since the boundary conditions of \( \lim_{L_c \to \infty} \psi_{s0}(x = L/2) = \psi_{s(LC)} \) and for \( V_{DS} = 0 \) \( \psi_s(L/2) = \psi_{s0}(L/2) \) are satisfied by the model surface potential equations and DIBL effect at \( x = L/2 \) is negligible in long-channel MOSFETs, the last boundary condition of \( \lim_{x \to \infty} \psi_s(x) = \psi_{s(LC)} \) for a finite \( V_{DS} > 0 \) is automatically satisfied. Equations 4.22 - 4.39 are the model for the subthreshold surface potential of short-channel as well as long-channel pocket devices. Equations 4.7 - 4.9 can be used to calculate the global values of \( \ell_{PS}, \ell_{PD}, \ell_c, \ell_{PO}, \ell_{CO} \text{ and } K_p \) in generating surface potential profiles between source and drain of 0.1-\( \mu m \) pocket n-MOSFETs for \( V_{GS} = 0 \) \( V \) and \( V_{DS} = 1 \) \( V \). However, the model is not in good agreement with the Medici especially for the 0.1-\( \mu m \) pocket n-MOSFETs with short pocket lengths. Therefore, \( \ell_{PS}, \ell_{PD} \) and \( \ell_c \) are also treated as model fitting parameters.

To check the complete model equations and the boundary conditions, the surface potential \( \psi_s(x) \) has been calculated at the position \( x \) between source and drain of pocket MOSFETs with channel length \( L = 0.1 \) and 1 \( \mu m \). using the following simple empirical formulas [24], [60] for calculating \( \ell_{PS}, \ell_{PD}, \ell_c, \ell_{PO} \text{ and } \ell_{CO} \) and assuming \( K_p = 0 \).

\[ \ell_{PS}, \ell_{PD}, \ell_{PO} = 0.16(t_{ox}r_jW_D^2(p))^{1/3} \] (4.40)
where the coefficient 0.16, which is different from the values in the formulas in [24], [60], was found for uniformly doped n-MOSFETs with channel length $L = 0.2\ \mu m$ using an optimization technique to minimize errors between surface potentials of a potential model [24] for deep-submicron MOSFETs and Medici. and $W_{D(p)}$ and $W_{D(c)}$ are the depletion-layer depths (Equation 2.4) under the gate at the pocket and the center regions, respectively. Figure 4.8 shows model results and comparisons of them with Medici surface potentials. The model results in Figure 4.8 show that the surface potential model (Equation 4.22) satisfies all the boundary conditions, including $\lim_{L_c \to \infty} \psi_s(L/2) = \psi_s(L_C)$. However, the model for the deep-submicron pocket MOSFETs neither generates good fits to all the Medici potential profiles with different $L_p$'s nor correctly predicts the highest energy barriers to electron injection, i.e., the minima of the surface potentials between source and drain in a desired error tolerance, e.g., 0.1% in comparing the model potentials to the Medici potentials because the empirical formulas in Equations 4.40 and 4.41 are not accurate. In addition, the empirical formulas do not include the pocket-length dependence of the model parameters. From the general form of z-formulas in Equations 4.14 and 4.15 for the model parameters of deep-submicron pocket MOSFETs described in Section 4.1, the analytical and empirical formulas for $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ in Equation 4.22 of the subthreshold surface potential model for $V_{DS} > 0\ \text{V}$ are proposed and verified against single-device values extracted by fitting model surface
Figure 4.8: Model surface potential $\psi_s$ for $V_{DS} = 1$ V, $V_{GS} = 0$ V checked for pocket n-MOSFETs with (a) $L = 0.1 \mu m$, $L_p = 0.022 \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$, $N_c = 2.375 \times 10^{17} \ cm^{-3}$, (b) $L = 0.1 \mu m$, $L_p = 0.044 \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$, $N_c = 2.138 \times 10^{17} \ cm^{-3}$, (c) $L = 1 \mu m$, $L_p = 0.022 \mu m$, $N_p = 2.135 \times 10^{18} \ cm^{-3}$, $N_c = 2.375 \times 10^{17} \ cm^{-3}$, and (d) $L = 1 \mu m$, $L_p = 0.044 \mu m$, $N_p = 8.557 \times 10^{17} \ cm^{-3}$, $N_c = 2.138 \times 10^{17} \ cm^{-3}$. $t_{ox} = 4 \ nm$ and $r_j = 0.06 \ mu m$ for all the devices. Crude formulas in Equations 4.40 and 4.41 for $\ell_{pS}$, $\ell_{pD}$, $\ell_c$, $\ell_{p0}$ and $\ell_{c0}$, and $K_p = 0$ were used for the model.
potential profiles to Medici surface potential profiles for $V_{GS} = 0 \ V$ and $V_{DS} = 1 \ V$ of the 0.1-$\mu m$ pocket n-MOSFETs in the subsequent subsection.

By minimizing the errors between the surface potentials of the model and Medici at $x_1 = x_m$ at which a minimum of the surface potential $\psi_{s(min)}(x_m)$ exists, $x_2 = L/2 = 0.05 \ \mu m$ for $L_p < 0.038 \ \mu m$ (Type A device), $x_2 = 0.04 \ \mu m$ for $L_p \geq 0.038 \ \mu m$ (Type B device) and $x_3 = L - L_p$, we can find the optimal values of $\ell_{ps}$, $\ell_{pD}$ and $\ell_c$ as the single-device values of all the 0.1-$\mu m$ pocket MOSFETs located at the lower design-space boundary for $I_{off}$. The algorithm described in Appendix C is used for the optimization by slightly modifying the procedure of DYNAMIC-BRACKETS(). Figure 4.9 shows comparisons of the model surface potentials with the Medici surface potentials of the 0.1-$\mu m$ pocket MOSFETs with different doping densities and pocket lengths. The relative errors between the minima $\psi_{s(min)}(x_m)$ of the surface potential distributions of the model and the Medici are within 0.1 % for all the pocket MOSFETs located at the design-space boundary. We observe from the results that the minima of the surface potential profiles of the typical pocket MOSFETs always exist at the pocket region surrounding the source region. The extracted optimal values as the single-device values of the parameters $\ell_{ps}$, $\ell_{pD}$ and $\ell_c$ vs. the parameter $z = LL_p N_p N_c$ of the 0.1-$\mu m$ pocket MOSFETs are shown on log-log scales in Figure 4.10, and the values are listed in Table 4.5.

Rms errors of the potentials of the model surface potential profiles to those of the Medici surface potential profiles for $V_{DS} = 1 \ V$ and $V_{GS} = 0 \ V$ of the twelve
Figure 4.9: Surface potential $\psi_x$ profiles between source and drain of 0.1-μm pocket n-MOSFETs for $V_{GS} = 0 V$ and $V_{DS} = 1 V$. (a) $L_p = 0.022 \mu m$, $N_p = 2.135 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.375 \times 10^{17} \text{ cm}^{-3}$, (b) $L_p = 0.028 \mu m$, $N_p = 1.558 \times 10^{18} \text{ cm}^{-3}$, $N_c = 1.925 \times 10^{17} \text{ cm}^{-3}$, (c) $L_p = 0.034 \mu m$, $N_p = 1.205 \times 10^{18} \text{ cm}^{-3}$, $N_c = 1.825 \times 10^{17} \text{ cm}^{-3}$, and (d) $L_p = 0.044 \mu m$, $N_p = 8.557 \times 10^{17} \text{ cm}^{-3}$, $N_c = 2.138 \times 10^{17} \text{ cm}^{-3}$. $t_{ox} = 4 \text{ nm}$ and $r_j = 0.06 \mu m$ for all the pocket devices. The 0.1-μm pocket MOSFETs chosen here are the same ones used in Figure 4.2. Single-device values of $\ell_{ps}$, $\ell_{pD}$, $\ell_c$, $\ell_{p0}$, $\ell_{c0}$ and $K_p$ were used for the model.
Table 4.5: Structural parameters and $z = L L_p N_p N_c$ of 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary, and single-device values of $\ell_{ps}$, $\ell_{pd}$ and $\ell_c$ of the pocket devices. $V_{DS} = 1 \text{ V}$ and $V_{GS} = 0 \text{ V}$.

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<th>$N_p [cm^{-3}]$</th>
<th>$N_c [cm^{-3}]$</th>
<th>$z [cm^{-3}]$</th>
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<td>$8.048 \times 10^{24}$</td>
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</tbody>
</table>

<table>
<thead>
<tr>
<th>$L_p [\mu m]$</th>
<th>$\ell_{ps} [\mu m]$</th>
<th>$\ell_{pd} [\mu m]$</th>
<th>$\ell_c [\mu m]$</th>
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<td>$1.855 \times 10^{-2}$</td>
<td>$1.529 \times 10^{-2}$</td>
<td>$8.879 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.026</td>
<td>$2.100 \times 10^{-2}$</td>
<td>$1.684 \times 10^{-2}$</td>
<td>$1.969 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.028</td>
<td>$2.268 \times 10^{-2}$</td>
<td>$1.816 \times 10^{-2}$</td>
<td>$1.336 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.030</td>
<td>$2.447 \times 10^{-2}$</td>
<td>$1.935 \times 10^{-2}$</td>
<td>$1.041 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.032</td>
<td>$2.511 \times 10^{-2}$</td>
<td>$2.004 \times 10^{-2}$</td>
<td>$1.477 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.034</td>
<td>$2.668 \times 10^{-2}$</td>
<td>$2.122 \times 10^{-2}$</td>
<td>$9.849 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.036</td>
<td>$2.614 \times 10^{-2}$</td>
<td>$2.113 \times 10^{-2}$</td>
<td>$7.826 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.038</td>
<td>$2.611 \times 10^{-2}$</td>
<td>$2.209 \times 10^{-2}$</td>
<td>$1.003 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.040</td>
<td>$2.670 \times 10^{-2}$</td>
<td>$2.213 \times 10^{-2}$</td>
<td>$1.088 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.042</td>
<td>$2.713 \times 10^{-2}$</td>
<td>$2.229 \times 10^{-2}$</td>
<td>$1.042 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.044</td>
<td>$2.787 \times 10^{-2}$</td>
<td>$2.263 \times 10^{-2}$</td>
<td>$9.567 \times 10^{-3}$</td>
</tr>
</tbody>
</table>
Figure 4.10: The extracted optimal values as single-device values of the model parameters \( \ell_{ps}, \ell_{pd}, \) and \( \ell_c \) vs. \( z = LL_p N_p N_c \) of the 0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary. \( V_{DS} = 1 \) V and \( V_{GS} = 0 \) V.

0.1-\( \mu \)m pocket n-MOSFETs located at the lower design-space boundary are shown in Figure 4.11. The rms error is defined as

\[
\text{rms error} = \sqrt{\frac{1}{N} \sum_{i=1}^{N} \left( \frac{\psi_s(x_i) - \psi_s^*(x_i)}{\psi_s(x_i)} \right)^2}
\]

where \( \psi_s(x_i) \) and \( \psi_s^*(x_i) \) are the Medici and model surface potentials at a location \( x_i \), respectively. \( \psi_s^*(x_i) \) is an approximation to \( \psi_s(x_i) \) and \( N \) is the number of data of \( \psi_s(x_i) \). The rms error in Figure 4.11 ranges up to 7 \%, which is larger than the relative error 0.1 \% of \( \psi_s(\min) \) at \( x_m \). The larger rms error results from the relatively large error of the model surface potential at the drain-depletion region as shown in Figure 4.9.
Figure 4.11: Rms errors of potentials of model surface potential profiles to those of Medici surface potential profiles for $V_{DS} = 1 \, V$ and $V_{GS} = 0 \, V$ of the 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary. The single-device values of $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ in Table 4.5 were used to obtain the model surface potential profiles.

4.2.2 Analytical and Empirical Formulas for $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ for $V_{GS} = 0 \, V$ and $V_{DS} = 1 \, V$

The analytical and empirical formula for the model parameter, $\ell_{pS}$, $\ell_{pD}$ or $\ell_c$, to generate its global parameter values can take one of several forms. In this section two possibilities are considered: the $z$-formulation and the power-law formulation.
4.2.2.1 $z$-Formulation

Equations 4.14 and 4.15 are used to find the empirical formulas of $\ell_{PS}$, $\ell_{PD}$ and $\ell_c$ for $V_{DS} = 1 \text{ V}$ and $V_{GS} = 0 \text{ V}$. The second-degree polynomials in terms of $\log_{10} z$ for $\log_{10} \ell_{PS}$, $\log_{10} \ell_{PD}$ and $\log_{10} \ell_c$ are given by

\[
\log_{10} \ell_{PS} = \beta_{PS}^0 (\log_{10} z)^2 + \beta_{PS}^1 (\log_{10} z) + \log_{10} \alpha_{PS} \quad (4.43)
\]

\[
\log_{10} \ell_{PD} = \beta_{PD}^0 (\log_{10} z)^2 + \beta_{PD}^1 (\log_{10} z) + \log_{10} \alpha_{PD} \quad (4.44)
\]

\[
\log_{10} \ell_c = \beta_{c}^0 (\log_{10} z)^2 + \beta_{c}^1 (\log_{10} z) + \log_{10} \alpha_c \quad (4.45)
\]

The coefficients in Equations 4.43, 4.44 and 4.45 found by polynomial fitting to the single-device values on log-log scales shown in Figure 4.10 are listed in Tables 4.6 - 4.8.

<table>
<thead>
<tr>
<th></th>
<th>Type A $(0.022 \ \mu m \leq L_p \leq 0.032 \ \mu m)$</th>
<th>Type B $(0.032 \ \mu m &lt; L_p \leq 0.044 \ \mu m)$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\log_{10} \alpha_{PS}$</td>
<td>$-1.307558 \times 10^3$</td>
<td>$-6.050021 \times 10^3$</td>
</tr>
<tr>
<td>$\beta_{PS}^0$</td>
<td>$-2.138189$</td>
<td>$-9.730892$</td>
</tr>
<tr>
<td>$\beta_{PS}^1$</td>
<td>$1.055278 \times 10^2$</td>
<td>$4.850488 \times 10^2$</td>
</tr>
</tbody>
</table>

Table 4.6: The coefficients of the second-degree polynomial of $\log_{10} \ell_{PS}$ of two types of 0.1-$\mu m$ pocket n-MOSFETs.

Figures 4.12 - 4.14 show the second-degree polynomial fits to the sets of the single-device values of $\log_{10} \ell_{PS}$, $\log_{10} \ell_{PD}$ and $\log_{10} \ell_c$. By inversely transforming the logarithmic functions $\log_{10} \ell_{PS}$, $\log_{10} \ell_{PD}$ and $\log_{10} \ell_c$ in Equations 4.43 - 4.45 into the corresponding exponential functions $\ell_{PS}$, $\ell_{PD}$ and $\ell_c$, we can directly obtain
Figure 4.12: $\log_{10} \ell_{ps}$ vs. $\log_{10}(L_{p}N_{p}N_{c})$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{ps}$ of (a) Type A ($0.022 \mu m < L_{p} < 0.032 \mu m$) and (b) Type B ($0.032 \mu m < L_{p} < 0.044 \mu m$) of 0.1-\mu m pocket n-MOSFETs.

Figure 4.13: $\log_{10} \ell_{pd}$ vs. $\log_{10}(L_{p}N_{p}N_{c})$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{pd}$ of (a) Type A ($0.022 \mu m < L_{p} < 0.036 \mu m$) and (b) Type B ($0.036 \mu m < L_{p} < 0.044 \mu m$) of 0.1-\mu m pocket n-MOSFETs.
Table 4.7: The coefficients of the second-degree polynomial of \( \log_{10} \ell_{PD} \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs.

<table>
<thead>
<tr>
<th></th>
<th>Type A ((0.022 \ \mu m \leq L_p \leq 0.036 \ \mu m))</th>
<th>Type B ((0.036 \ \mu m &lt; L_p \leq 0.044 \ \mu m))</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \log_{10} \alpha_{PD} )</td>
<td>( 5.991358 \times 10^2 )</td>
<td>( 2.965536 \times 10^3 )</td>
</tr>
<tr>
<td>( \beta_{PD}^0 )</td>
<td>( 9.296407 \times 10^{-1} )</td>
<td>( 4.808374 )</td>
</tr>
<tr>
<td>( \beta_{PD}^1 )</td>
<td>( -4.743914 \times 10^1 )</td>
<td>( -2.390532 \times 10^2 )</td>
</tr>
</tbody>
</table>

Table 4.8: The coefficients of the second-degree polynomial of \( \log_{10} \ell_c \) of two types of 0.1-\( \mu m \) pocket n-MOSFETs.

<table>
<thead>
<tr>
<th></th>
<th>Type A ((0.022 \ \mu m \leq L_p \leq 0.036 \ \mu m))</th>
<th>Type B ((0.036 \ \mu m &lt; L_p \leq 0.044 \ \mu m))</th>
</tr>
</thead>
<tbody>
<tr>
<td>( \log_{10} \alpha_c )</td>
<td>( -2.993293 \times 10^4 )</td>
<td>( -5.523374 \times 10^4 )</td>
</tr>
<tr>
<td>( \beta_c^0 )</td>
<td>( -4.811739 \times 10 )</td>
<td>( -8.920175 \times 10 )</td>
</tr>
<tr>
<td>( \beta_c^1 )</td>
<td>( 2.400053 \times 10^3 )</td>
<td>( 4.439105 \times 10^3 )</td>
</tr>
</tbody>
</table>

the values of the model parameters on a linear scale. Thus,

\[
\ell_{PD} = 10^{\log_{10} \ell_{PD}} = \alpha_{PD} z^{\beta_{PD}^0 (\log_{10} z)} z^{\beta_{PD}^1} \quad (4.46)
\]

\[
\ell_c = 10^{\log_{10} \ell_c} = \alpha_c z^{\beta_c^0 (\log_{10} z)} z^{\beta_c^1} \quad (4.48)
\]

Figure 4.15 shows the model surface potential profiles obtained using the complete model equations that include Equations 4.22 - 4.39 and 4.43 - 4.48, and the comparisons of the model results with the Medici surface potential profiles for \( V_{DS} = 1 \) V and \( V_{GS} = 0 \) V of the 0.1-\( \mu m \) pocket n-MOSFETs (a), (b), (c) and (d) located at the lower design-space boundary in Figure 3.7. The complete surface potential model can generate the surface potential profiles of the 0.1-\( \mu m \) pocket n-MOSFETs
Figure 4.14: \( \log_{10} \ell_c \) vs. \( \log_{10}(LL_pN_pN_c) \) as results of the second-degree polynomial fits to single-device values of \( \log_{10} \ell_c \) of (a) Type A \((0.022 \mu m \leq L_p \leq 0.036 \mu m)\) and (b) Type B \((0.036 \mu m < L_p < 0.044 \mu m)\) of 0.1-\(\mu m\) pocket n-MOSFETs.

fast and correctly. The potential model in [4] is also compared with our model and Medici in Figure 4.15 (a) and (d). Fig. 4.16 shows surface potential profiles for \( V_{GS} = 0 \) \(V\) between source and drain of the 0.1-\(\mu m\) pocket n-MOSFET located at the lower design-space boundary, which were generated by the subthreshold surface potential model for \( \psi_s(x) \). The DIBL effect, which is an increase in the potential minimum due to the increase in the drain voltage, is small in the well-designed 0.1-\(\mu m\) pocket device.

The rms errors of the potentials of the complete model surface potential profiles to those of the Medici surface potential profiles for \( V_{DS} = 1 \) \(V\) and \( V_{GS} = 0 \) \(V\) of the twelve 0.1-\(\mu m\) pocket n-MOSFETs located at the lower design-space boundary
Figure 4.15: Surface potential $\psi_s$ profiles between source and drain for $V_{GS} = 0 \, \text{V}$ and $V_{DS} = 1 \, \text{V}$ of the 0.1-$\mu\text{m}$ pocket n-MOSFETs which are the same ones used in Figure 4.9. The model surface potentials were generated by the complete model equations, including the $z$-formulas for $\ell_{PS}$, $\ell_{PD}$, $\ell_c$, $\ell_{p0}$, $\ell_{c0}$ and $K_p$. (a) $L_p = 0.022 \, \mu\text{m}$, $N_p = 2.135 \times 10^{18} \, \text{cm}^{-3}$, $N_c = 2.375 \times 10^{17} \, \text{cm}^{-3}$, (b) $L_p = 0.028 \, \mu\text{m}$, $N_p = 1.558 \times 10^{18} \, \text{cm}^{-3}$, $N_c = 1.925 \times 10^{17} \, \text{cm}^{-3}$, (c) $L_p = 0.034 \, \mu\text{m}$, $N_p = 1.205 \times 10^{18} \, \text{cm}^{-3}$, $N_c = 1.825 \times 10^{17} \, \text{cm}^{-3}$, and (d) $L_p = 0.044 \, \mu\text{m}$, $N_p = 8.557 \times 10^{17} \, \text{cm}^{-3}$, $N_c = 2.138 \times 10^{17} \, \text{cm}^{-3}$. $t_{ox} = 4 \, \text{nm}$ and $r_j = 0.06 \, \mu\text{m}$. Global values of $\ell_{PS}$, $\ell_{PD}$, $\ell_c$, $\ell_{p0}$, $\ell_{c0}$ and $K_p$ were used for the model.
Figure 4.16: DIBL effect for $V_{GS} = 0$ V of 0.1-$\mu$m pocket n-MOSFET with $L_p = 0.022$ $\mu$m, $N_p = 2.135 \times 10^{18}$ cm$^{-3}$, $N_c = 2.375 \times 10^{17}$ cm$^{-3}$, $t_{ox} = 4$ nm and $r_j = 0.06$ $\mu$m. The subthreshold surface potential model for $\psi_s(x)$ was used to generate the potential profiles.

are shown in Figure 4.17 where the rms error calculated using the global values of the model parameters $\ell_p$, $\ell_{PD}$, $\ell_c$, $\ell_p$, $\ell_c$, and $K_p$ is compared with that calculated using the single-device values of the parameters.

4.2.2.2 Power-Law Formulation

We consider the power-law formulation to formulate the second analytical and empirical formulas for the model parameters, $\ell_{PS}$, $\ell_{PD}$, and $\ell_c$. A typical application of the power-law formulation to an electrical parameter is discussed in [31].
Figure 4.17: Rms errors of potentials of model surface potential profiles to those of Medici surface potential profiles for $V_{DS} = 1 \, V$ and $V_{GS} = 0 \, V$ of the 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary. Either the global or the single-device values of $\ell_{ps}$, $\ell_{pD}$, $\ell_c$, $\ell_{p0}$, $\ell_{c0}$ and $K_p$ were used to obtain the model surface potential profiles.

The following are the power-law formulas for $\ell_{ps}$, $\ell_{pD}$ and $\ell_c$ in $\mu$m found by fitting the formulas to the single-device values of the model parameters in Table 4.5.

\begin{align*}
\ell_{ps} &= 2.506(L_p/L_{p0})^{1.227312}(N_p/N_{p0})^{0.492635}(N_c/N_{c0})^{-0.82164} \\
\ell_{pD} &= 0.178(L_p/L_{p0})^{0.541286}(N_p/N_{p0})^{-0.07049}(N_c/N_{c0})^{-0.506316} \\
\ell_c &= 0.140(L_p/L_{p0})^{-1.588690}(N_p/N_{p0})^{2.033272}(N_c/N_{c0})^{-10.16395}
\end{align*}

where the normalization parameters are $L_{p0} = 10^{-4} \, cm$, $N_{p0} = 10^{18} \, cm^{-3}$ and $N_{c0} = 10^{17} \, cm^{-3}$. The rms errors of the global values of the z- and the power-
Table 4.9: Rms errors of the global values to the single-device values of $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ for the twelve 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.

<table>
<thead>
<tr>
<th>Formulation</th>
<th>$\ell_{pS}$</th>
<th>$\ell_{pD}$</th>
<th>$\ell_c$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Z-formula</td>
<td>0.66 %</td>
<td>0.81 %</td>
<td>15.58 %</td>
</tr>
<tr>
<td>Power-law formula</td>
<td>2.97 %</td>
<td>1.21 %</td>
<td>115.6 %</td>
</tr>
</tbody>
</table>

law formulas for $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ to the single-device values for the twelve 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary are tabulated in Table 4.9. The z-formulation has better accuracy than the power-law formulation. However, the power-law formulation covers all the values of a model parameter for both Types A and B with one model formula. The accuracy of the power-law formulation might be improved at the expense of the increase in the number of parameters of the formulas when the single-device values are split into two sets.

Figure 4.18 shows comparisons of the global values of the z- and the power-law formulas for $\ell_{pS}$ with the single-device values for the 0.1-$\mu$m pocket n-MOSFETs. Figure 4.19 shows the global values of the z- and the power-law formulas for $\ell_{pD}$ and $\ell_c$, and the corresponding single-device values for the 0.1-$\mu$m pocket devices.

4.3 Modeling of Off-State Current and Model Results

Expressions for the subthreshold current and the location at which the minimum of the surface potential distribution of pocket n-MOSFETs exists are derived to predict the off-state current in 0.1-$\mu$m pocket devices. The off-state current
Figure 4.18: Comparisons of global values with single-device values of the parameter $\ell_{ps}$ of 0.1-$\mu m$ pocket n-MOSFETs. (a) $\ell_{ps}$ vs. $L_p$, (b) $\ell_{ps}$ vs. $N_p/N_{p0}$, (c) $\ell_{ps}$ vs. $N_c/N_{c0}$ and (d) $\ell_{ps}$ vs. $z/z_0$. The global parameter values were calculated using power-law and $z$-formulas. $N_{p0} = 10^{18} \text{ cm}^{-3}$, $N_{c0} = 10^{17} \text{ cm}^{-3}$ and $z_0 = 10^{24} \text{ cm}^{-4}$. 
model includes the subthreshold surface potential models described in the preceding sections, and will be used to generate the curves of $N_p$ vs. $L_p$ and $N_c$ vs. $L_p$ of the lower design-space boundary for the 0.1-µm pocket n-MOSFETs to be compared with the two-dimensional simulation results shown in Figure 3.7. The model for the off-state current will also be used to predict design-space boundaries for 0.1-µm pocket devices satisfying the specification of $I_{off} \leq 1 \text{nA/µm}$ regardless of the specification of the sensitivity of off-state current to channel length.

4.3.1 Off-State Current Model

The diffusion current term of the current transport equation in Equation 2.31 can be used to derive an expression for the subthreshold current $I_{D(sub)}$ per unit
width because the diffusion process is a dominant part in the subthreshold current flow within semiconductor. We intend to evaluate the current at the potential minimum where the lateral field is zero, so only diffusion is possible. Thus, at the Si surface \((y = 0)\)

\[ I_{D_{\text{(sub)}}} = qD_n \frac{dN_f}{dx} \]  

(4.52)

where \(D_n(x) = \mu_{\text{eff}}(x)V_{\text{th}}\) (the Einstein relation) is the diffusion constant for electrons, \(\mu_{\text{eff}}(x)\) is the effective mobility of electrons, and \(N_f(x)\) is the electron density per unit area in weak inversion layer. Using the coordinate system in Figure 2.9, we let \(x_m\) be a starting point of the effective channel length \(L_{\text{eff}}\) at the Si surface. For the purpose of the modeling of the subthreshold current in deep-submicron pocket n-MOSFETs, let \(x_m\) be a location at which a minimum of a surface potential profile between source and drain exists. Then, Equation 4.52 can be approximated to:

\[
I_{D_{\text{(sub)}}} \simeq qD_n \frac{N_f(x_m + L_{\text{eff}}) - N_f(x_m)}{L_{\text{eff}}} \\
= qD_n \frac{N_f(x_m)e^{-\Delta \phi_s/V_{\text{th}}} - N_f(x_m)}{L_{\text{eff}}} \\
= -\frac{q\mu_{\text{eff}}V_{\text{th}}}{L_{\text{eff}}} N_f(x_m)(1 - e^{-\Delta \phi_s/V_{\text{th}}})
\]  

(4.53)

where

\[
\Delta \phi_s = \phi_s(x_m + L_{\text{eff}}) - \phi_s(x_m)
\]  

(4.54)
Note that $I_{D_{(sub)}} [A/\mu m]$ is constant at the Si surface along the effective channel, and the effective mobility $\mu_{\text{eff}}$ was replaced with the effective transverse-field-dependent mobility $\mu_{\text{eff}0}$ at a low lateral field $E_x$ in deriving the expression for the subthreshold current because $E_x$ vanishes at the point where the minimum of the surface potential distribution exists.

To derive an expression for the electron density $N_i(x_m)$ in inversion layer, some equations described in Chapter 2 are used. From Equation 2.18, we have

$$N_i(x_m) = n_i W L = \left( \frac{n_i^2}{N_p} e^{\phi_{s(min)}} \right) \left( \frac{V_{th}}{E_{\text{eff}}} \right)$$

(4.55)

Note that according to model results and Medici simulations, the minimum surface potential of typical pocket MOSFETs, e.g., the pocket devices located within viable design space always exists at the pocket region near source. Since the electron charge density is negligible compared to the depletion-layer charge density at weak inversion, the effective transverse electric field $E_{\text{eff}}$ (Equation 2.1) is approximated to

$$E_{\text{eff}} = \frac{q N_i(x_m)}{2 \epsilon_s} + \frac{q N_p W_D}{\epsilon_s} \approx \frac{q N_p W_D}{\epsilon_s} = E_D$$

(4.56)

Therefore,

$$N_i(x_m) \approx \left( \frac{n_i^2}{N_p} e^{\phi_{s(min)}} \right) \left( \frac{V_{th}}{E_D} \right) = \left( \frac{n_i^2}{N_p} e^{\phi_{s(min)}} \right) \left( \frac{L_D^2}{W_D} \right)$$

(4.57)

where $L_D$ is the Debye length at the pocket region, and $W_D$ is the depletion-layer depth under gate at the pocket region. The expression for $N_i(x_m)$ has exactly
the same form as Equation 2.21 derived by using the binomial series expansion.

Applying the result (Equation 2.11) of the continuity of electric displacement at
the interface between Si and SiO₂ or Gauss's law, Equation 2.12 and $\mathcal{E}_D \approx \mathcal{E}_s$ to
Equation 4.57, we obtain

$$N_l(x_m) = \frac{q(n_i L_D)^2 e^{\beta \phi_{s(min)}}}{C_{ox}(\phi_G - \phi_{s(min)})}$$  (4.58)

From Equations 4.53 and 4.58, we derive a final expression for $I_{D(sub)}$.

$$I_{D(sub)} = \frac{-\mu_{eff0} V_{th} (q n_i L_D)^2}{L_{eff}} \frac{e^{\beta \phi_{s(min)}}}{C_{ox}(\phi_G - \phi_{s(min)})} (1 - e^{-\Delta \phi_s})$$  (4.59)

In Equation 4.59 when $\Delta \phi_s \gg V_{th}$, $I_{D(sub)}$ is a function of $\phi_{s(min)}$ and $L_{eff}$ alone.

The surface potential minimum $\phi_{s(min)}$ of deep-submicron pocket n-MOSFETs will
be modeled below using the solutions of the quasi-two-dimensional Poisson's equa­
tion in Sections 4.1 and 4.2, and will include the DIBL effect. The exponential
factor in Equation 4.59 is expected for any potential barrier model of current flow,
and $L_{eff}$ also absorbs two-dimensional effects on the barrier shape when used as a
fitting parameter.

Assuming $\Delta \phi_s \gg V_{th}$ for $V_{DS} = 1 \text{ V}$, Equation 4.59 is directly used to model
$I_{off}$ for $V_{GS} = 0 \text{ V}$ and $V_{DS} = 1 \text{ V}$, which can be represented by the following
functional form.

$$I_{off} = \frac{1}{L_{eff}} f_{off}[\psi_{s(min)}(x_m), \mu_{eff0}]$$  (4.60)
where

\[ \psi_{s(min)}(x_m) = \phi_{s(min)}(x_m) - V_{th} \ln \left( \frac{N_p}{n_i} \right) \]  \hspace{1cm} (4.61)

Differentiation of Equation 4.22 with respect to \( x \) at the pocket region near source yields the location \( x_m \), which is given by

\[ x_m = 0.5[L_p + \epsilon_{ps} \ln(g_1/g_2)], \quad 0 < x_m < L_p \] \hspace{1cm} (4.62)

The minimum of the surface potential \( \psi_{s(min)}(x_m) \) is obtained by combining Equations 4.22 and 4.62.

The effective channel length \( L_{eff} \) can be expressed as a function of \( z = LL_pN_pN_c \).

Thus,

\[ L_{eff} = 10^{\log_{10} L_{eff}} = \alpha L_e \ln \beta_{L_e} \ln(z) \ln L_e \] \hspace{1cm} (4.63)

Taking the logarithmic function for both sides of Equation 4.63, we obtain a second-degree polynomial in terms of \( \log_{10} z \)

\[ \log_{10} L_{eff} = \beta^0_{L_e} (\log_{10} z)^2 + \beta^1_{L_e} (\log_{10} z) + \log_{10} \alpha L_e \] \hspace{1cm} (4.64)

The coefficients of the second-degree polynomial can be found by polynomial fitting to single-device values of \( L_{eff} \) in a log-log plot of \( L_{eff} \) vs. \( z = LL_pN_pN_c \). The single-device values of \( L_{eff} \) are obtained by using Equation 4.59, the model equations for \( \phi_{s(min)}(x_m) \) and \( \mu_{eff0} \), and Medici values of \( I_{off} \simeq 0.714 \, nA/\mu m \). The values are tabulated in Table 4.10 together with the structural parameters of the 0.1-\( \mu m \)
Table 4.10: Structural parameters of 0.1-μm pocket n-MOSFETs located at the lower design-space boundary and single-device values of $L_{eff}$ of the pocket devices. $V_{DS} = 1 \text{ V}$ and $V_{GS} = 0 \text{ V}$.

<table>
<thead>
<tr>
<th>$L_p , [\mu m]$</th>
<th>$N_p , [\text{cm}^{-3}]$</th>
<th>$N_c , [\text{cm}^{-3}]$</th>
<th>$L_{eff} , [\mu m]$</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.022</td>
<td>$2.135 \times 10^{18}$</td>
<td>$2.375 \times 10^{17}$</td>
<td>$1.037 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.024</td>
<td>$1.906 \times 10^{18}$</td>
<td>$2.175 \times 10^{17}$</td>
<td>$1.268 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.026</td>
<td>$1.715 \times 10^{18}$</td>
<td>$2.050 \times 10^{17}$</td>
<td>$9.538 \times 10^{-3}$</td>
</tr>
<tr>
<td>0.028</td>
<td>$1.558 \times 10^{18}$</td>
<td>$1.925 \times 10^{17}$</td>
<td>$1.208 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.030</td>
<td>$1.424 \times 10^{18}$</td>
<td>$1.850 \times 10^{17}$</td>
<td>$1.558 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.032</td>
<td>$1.306 \times 10^{18}$</td>
<td>$1.825 \times 10^{17}$</td>
<td>$1.292 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.034</td>
<td>$1.205 \times 10^{18}$</td>
<td>$1.825 \times 10^{17}$</td>
<td>$1.325 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.036</td>
<td>$1.116 \times 10^{18}$</td>
<td>$1.838 \times 10^{17}$</td>
<td>$1.919 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.038</td>
<td>$1.039 \times 10^{18}$</td>
<td>$1.875 \times 10^{17}$</td>
<td>$1.888 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.040</td>
<td>$9.707 \times 10^{17}$</td>
<td>$1.925 \times 10^{17}$</td>
<td>$1.621 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.042</td>
<td>$9.095 \times 10^{17}$</td>
<td>$2.025 \times 10^{17}$</td>
<td>$2.156 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.044</td>
<td>$8.557 \times 10^{17}$</td>
<td>$2.138 \times 10^{17}$</td>
<td>$2.055 \times 10^{-2}$</td>
</tr>
</tbody>
</table>

Table 4.11: The coefficients of the second-degree polynomial of $\log_{10} L_{eff}$ of two types of 0.1-μm pocket n-MOSFETs.

<table>
<thead>
<tr>
<th>Type A (0.022 μm ≤ $L_p$ ≤ 0.034 μm)</th>
<th>Type B (0.034 μm &lt; $L_p$ ≤ 0.044 μm)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$\log_{10} \alpha_{L_e}$</td>
<td>$2.192966 \times 10^3$</td>
</tr>
<tr>
<td>$\beta_{L_e}^0$</td>
<td>$3.502999$</td>
</tr>
<tr>
<td>$\beta_{L_e}^1$</td>
<td>$-1.755321 \times 10^2$</td>
</tr>
<tr>
<td>$\beta_{L_e}^2$</td>
<td>$-2.444054 \times 10^4$</td>
</tr>
<tr>
<td>$\beta_{L_e}^3$</td>
<td>$-3.938411 \times 10^3$</td>
</tr>
<tr>
<td>$\beta_{L_e}^4$</td>
<td>$1.961984 \times 10^6$</td>
</tr>
</tbody>
</table>

Figure 4.20 shows the second-degree polynomial fits of the analytical and empirical formula (Equation 4.64) to the single-device values of $\log_{10} L_{eff}$. Note that $L_{eff}$ of the off-state current is evaluated using Equation 4.64 and the first equation in Equation 4.63.
Figure 4.20: $\log_{10} L_{\text{eff}}$ vs. $\log_{10}(L_{\text{p}}N_pN_c)$ as results of the second-degree polynomial fits to single-device values of $\log_{10} L_{\text{eff}}$ of (a) Type A ($0.022\mu m \leq L_p \leq 0.034\mu m$) and (b) Type B ($0.034\mu m < L_p \leq 0.044\mu m$) of 0.1-µm pocket n-MOSFETs.

4.3.2 Results of $I_{\text{off}}$ Model and Prediction of Lower Design-Space Boundary for $I_{\text{off}}$

The closed-form model equations described in the preceding sections are used to generate the off-state current in deep-submicron ($L = 0.1 \mu m$) pocket n-MOSFETs, and the lower design-space boundary for $I_{\text{off}}$ fast and correctly.

Figure 4.21 shows how to calculate $I_{\text{off}}$ schematically when using the model equations rather than Medici. Model results of $I_{\text{off}}$ are listed in Table 4.12 and compared to the Medici data for the 0.1-µm pocket MOSFETs located at the lower design-space boundary for $I_{\text{off}}$. In order to generate the curves $N_p$ vs. $L_p$ and $N_c$
Input $L, L_p, N_p, N_c, t_{ox}$, and physical constants.

- Calculate $\xi_p, \xi_c$ and $K_p$ for $V_{DS} = 0$ V and $V_{GS} = 0$ V.
- Calculate $\psi_{sc}$ for $x = L/2$, $V_{DS} = 0$ V and $V_{GS} = 0$ V.
- Calculate $\xi_p, \xi_c$ and $\xi_c$ for $V_{DS} = 1$ V and $V_{GS} = 0$ V.
- Calculate $x_n$ and $\psi_{(min)}(x_n)$ for $V_{DS} = 1$ V and $V_{GS} = 0$ V.

Calculate $\mu_{eff}$

Calculate $L_{eff}$

Calculate $I_{off} = f_{off}(\mu_{eff}, \psi_{(min)}(x_n))/L_{eff}$

Return $I_{off}$.

Figure 4.21: Calculation of $I_{off}$.

vs. $L_p$ in Figure 3.7 of the lower design-space boundary for $I_{off} = 0.714 \, \text{nA/\mu m}$ and $\frac{\Delta I_{off}}{I_{off}} = 40\%/10\%$ of the 0.1-\mu m pocket MOSFETs by using only the model equations, we have to solve the model equations, i.e., $I_{off}(\text{Model}) - I_{off}(\text{Medici})(\simeq 0.714 \, \text{nA/\mu m}) = 0$ for $N_p$ at given values of $N_c$ and $L_p$, or for $N_c$ at given values of $N_p$ and $L_p$, using a numerical method, e.g., the secant method [61]. Figure 4.22 shows the comparisons of the model with the Medici results. The model can generate the design-space boundary within 1.2% error tolerance with reference to the Medici values. The model equations are checked further, and $I_{off}(\text{Model})$
Figure 4.22: Model results of (a) \(N_p\) vs. \(L_p\) and (b) \(N_c\) vs. \(L_p\) at the lower design-space boundary for \(I_{off} = 0.714\ \text{nA/\mu m}\) and \(\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40%/10\%\) of the 0.1-\(\mu\)m pocket n-MOSFETs are compared with the two-dimensional simulation results shown in Figure 3.7. The lines in the figures are the model results.
Table 4.12: Model values of $I_{off}$ compared to Medici data for the 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.

<table>
<thead>
<tr>
<th>$L_p$ [$\mu$m]</th>
<th>$I_{off}$ [nA/$\mu$m] (Medici)</th>
<th>$I_{off}$ [nA/$\mu$m] (Model)</th>
<th>Percent error</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.022</td>
<td>0.713</td>
<td>0.694</td>
<td>2.7 %</td>
</tr>
<tr>
<td>0.024</td>
<td>0.713</td>
<td>0.827</td>
<td>16.0 %</td>
</tr>
<tr>
<td>0.026</td>
<td>0.715</td>
<td>0.597</td>
<td>16.6 %</td>
</tr>
<tr>
<td>0.028</td>
<td>0.714</td>
<td>0.707</td>
<td>1.1 %</td>
</tr>
<tr>
<td>0.030</td>
<td>0.713</td>
<td>0.855</td>
<td>20.0 %</td>
</tr>
<tr>
<td>0.032</td>
<td>0.718</td>
<td>0.685</td>
<td>4.5 %</td>
</tr>
<tr>
<td>0.034</td>
<td>0.713</td>
<td>0.682</td>
<td>4.5 %</td>
</tr>
<tr>
<td>0.036</td>
<td>0.717</td>
<td>0.757</td>
<td>5.6 %</td>
</tr>
<tr>
<td>0.038</td>
<td>0.716</td>
<td>0.739</td>
<td>3.2 %</td>
</tr>
<tr>
<td>0.040</td>
<td>0.716</td>
<td>0.616</td>
<td>13.9 %</td>
</tr>
<tr>
<td>0.042</td>
<td>0.715</td>
<td>0.761</td>
<td>6.4 %</td>
</tr>
<tr>
<td>0.044</td>
<td>0.712</td>
<td>0.697</td>
<td>2.1 %</td>
</tr>
</tbody>
</table>

$-1 \text{ nA/}$$\mu$m = 0 is solved for $N_p$ at given values of $N_c$ and $L_p$ to predict design-space boundaries for the worst-case specification of $I_{off} = 1 \text{ nA/}$$\mu$m in 0.1-$\mu$m pocket n-MOSFETs regardless of the specification of the sensitivity of off-state current to channel length. The model predictions are shown in Figure 4.23. The plot in Figure 4.23 (b) shows three different contours of equal $N_p$, $N_c$ and $L_p$. The results show that in order to satisfy the specification of $I_{off} = 1 \text{ nA/}$$\mu$m, $N_p$ has to be decreased as $L_p$ is increased at a fixed $N_c$ or as $N_c$ is increased at a fixed $L_p$. Finally, Figure 4.24 shows the design-space boundary $N_p$ vs. $L_p$ with $N_c$ and $I_{off}$ as parameters, predicted by the models for 0.1-$\mu$m pocket n-MOSFETs with $L_p = 0.022 - 0.033 \mu$m satisfying the specification of $I_{off} \leq 1 \text{ nA/}$$\mu$m regardless of the specification of the channel-length sensitivity of off-state current.
Figure 4.23: Model predictions of design-space boundaries for the worst-case specification of $I_{off} = 1 \, nA/\mu m$ in 0.1-\mu m pocket n-MOSFETs with $L_p = 0.022 - 0.033 \, \mu m$. (a) Two-dimensional design-space boundary $N_p$ vs. $L_p$ with $N_c$ as a parameter, and (b) three-dimensional design-space boundary.
Figure 4.24: Model predictions of design-space boundary of $N_p$ vs. $L_p$ with $N_c$ and $I_{off} \leq 1 \text{nA/\mu m}$ as parameters of 0.1-\mu m pocket n-MOSFETs.

4.4 Simple Model for Off-State Current in 0.1-\mu m Pocket n-MOSFETs and Design-Space Boundary for Off-State Current

This section is devoted to simplification of model equations for the off-state current in the 0.1-\mu m pocket n-MOSFETs and generation of $N_p$ vs. $L_p$ and $N_c$ vs. $L_p$ of the lower design-space boundary for the off-state current using simple model equations. $N_p$ vs. $L_p$ is generated, using the off-state current model in Section 4.3, and then $N_c$ vs. $L_p$ of the design-space boundary is generated, using the surface potential model for $V_{DS} = 0 \text{ V}$ and $V_{GS} = 0 \text{ V}$ in Section 4.1 and the model result
of \( N_p \) vs. \( L_p \). All analytical formulas for model parameters of the simple model are
simply expressed by polynomials in terms of the geometrical parameter \( L_p \).

4.4.1 Off-State Current Model and \( N_p \) vs. \( L_p \) of Lower Design-Space Boundary

The expression for the off-state current for \( V_{QS} = 0 \text{V} \) and \( Y_{DS} = 1 \text{V} \) is given in
Equations 4.59 - 4.61. Since the difference of the Medici surface potential minimum
\( \psi_{s(min)} \) between the 0.1-\( \mu \text{m} \) pocket n-MOSFETs located at the lower design-space
boundary is small (Figure 3.8 (a) in Chapter 3), the value of \( \psi_{s(min)} = 0.266 \text{V} \) of the
pocket device with \( L_p = 0.034 \mu \text{m} \) is used to extract the single-device values of \( L_{eff} \)
of all the 0.1-\( \mu \text{m} \) pocket n-MOSFETs. The single-device values of \( L_{eff} \), which

\[
\begin{array}{|c|c|c|c|}
\hline
L_p [\mu \text{m}] & N_p [\text{cm}^{-3}] & N_c [\text{cm}^{-3}] & L_{eff} [\mu \text{m}] \\
\hline
0.022 & 2.135 \times 10^{18} & 2.375 \times 10^{17} & 1.120 \times 10^{-2} \\
0.024 & 1.906 \times 10^{18} & 2.175 \times 10^{17} & 1.206 \times 10^{-2} \\
0.026 & 1.715 \times 10^{18} & 2.050 \times 10^{17} & 1.288 \times 10^{-2} \\
0.028 & 1.558 \times 10^{18} & 1.925 \times 10^{17} & 1.371 \times 10^{-2} \\
0.030 & 1.424 \times 10^{18} & 1.850 \times 10^{17} & 1.455 \times 10^{-2} \\
0.032 & 1.306 \times 10^{18} & 1.825 \times 10^{17} & 1.526 \times 10^{-2} \\
0.034 & 1.205 \times 10^{18} & 1.825 \times 10^{17} & 1.614 \times 10^{-2} \\
0.036 & 1.116 \times 10^{18} & 1.838 \times 10^{17} & 1.685 \times 10^{-2} \\
0.038 & 1.039 \times 10^{18} & 1.875 \times 10^{17} & 1.762 \times 10^{-2} \\
0.040 & 9.707 \times 10^{17} & 1.925 \times 10^{17} & 1.838 \times 10^{-2} \\
0.042 & 9.095 \times 10^{17} & 2.025 \times 10^{17} & 1.913 \times 10^{-2} \\
0.044 & 8.557 \times 10^{17} & 2.138 \times 10^{17} & 1.992 \times 10^{-2} \\
\hline
\end{array}
\]

Table 4.13: Structural parameters of 0.1-\( \mu \text{m} \) pocket n-MOSFETs located at the
lower design-space boundary and single-device values of \( L_{eff} \) of the pocket devices
to be used for developing an analytical formula for \( L_{eff} \) of the simple model. \( V_{DS} = 1 \text{V} \) and \( V_{GS} = 0 \text{V} \).
are tabulated in Table 4.13, is obtained by using Equation 4.59 for $\Delta \phi_s \gg V_{th}$. $\psi_{s(min)} = 0.266 \, V$, the modified Arora mobility model of the unified mobility model in Chapter 2 and Medici values of $I_{off} \simeq 0.714 \, nA/\mu m$. Figure 4.25 shows the single-device values of the parameter $L_{eff}$ which are compared with its global parameter values. The global parameter values of $L_{eff}$ were calculated from the following analytical model formula.

$$L_{eff} = 0.394L_p + 2.636 \times 10^{-3}$$  \hspace{1cm} (4.65)

where both $L_{eff}$ and $L_p$ have the dimension of $\mu m$. Equation 4.65 and Figure 4.25 together with Figure 3.10 show that the effective channel length $L_{eff}$ is a fraction of the length $L_p$ of the source-end pocket region. In addition, as the pocket length is increased, the channel length modulation decreases.

Using the model equations, i.e., Equations 4.59 – 4.61 and 4.65 and solving the equation $I_{off}(N_p) - 0.714 \, nA/\mu m = 0$ for $N_p$ numerically with a given value of $L_p$, we can find a model $N_p$ vs. $L_p$ of the lower design-space boundary for $I_{off}$ in the 0.1-$\mu m$ pocket n-MOSFETs. Figure 4.26 shows a model result generated by the model equations and compared with the Medici lower design-space boundary. The errors of the model $N_p$ with reference to the Medici $N_p$ are within 1.5% for all the twelve 0.1-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary.
Figure 4.25: Comparison of global parameter values with single-device values of $L_{eff}$ of 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.

Figure 4.26: Comparison of the model lower design-space boundary for $I_{off}$ in 0.1-$\mu$m pocket n-MOSFETs with the Medici lower design-space boundary.
4.4.2 Surface Potential Model and $N_c$ vs. $L_p$ of Lower Design-Space Boundary

We can use the surface potential model for $V'_{DS} = 0$ V and $V_{GS} = 0$ V, i.e., Equations 4.6 and 4.10 – 4.12, in Section 4.1 to generate $N_c$ vs. $L_p$ of the lower design-space boundary for $0.1$-$\mu m$ pocket n-MOSFETs. In the model equations, $\psi_{s(LC)}$ is the long-channel surface potential at the center region calculated from Equations 3.4 and 3.7 in Chapter 3. $K_p$ is a potential parameter at the pocket region and directly related to the surface potential at the pocket region. The characteristic length $\ell_{p0}$ at the pocket region increases as the doping density $N_p$ at the pocket region decreases. The characteristic length $\ell_{c0}$ at the center region increases as the doping density $N_c$ at the center region decreases, too.

The single-device values of $\ell_{p0}$, $\ell_{c0}$ and $K_p$ are shown in Figure 4.3 and tabulated in Table 4.1. In order to include dependence of the model surface potential $\psi_{s0}(x)$ on the pocket doping density $N_p$, we introduce another model parameter. $\zeta \equiv \psi_{s0(pock)} - K_p$, in which $\psi_{s0(pock)}$ is the long-channel surface potential which is a function of $N_p$, and is evaluated using Equations 3.4 and 3.7. The single-device values of $\zeta$ of the $0.1$-$\mu m$ pocket n-MOSFETs located at the lower design-space boundary are tabulated in Table 4.14. Polynomial fits to the single-device values of $\zeta$, $\ell_{p0}$ and $\ell_{c0}$ can provide simple analytical formulas for $\zeta$, $\ell_{p0}$ and $\ell_{c0}$. Thus,

$$\zeta = \psi_{s0(pock)} - K_p$$

$$= -1.411 \times 10^6 L_p^4 + 2.849 \times 10^5 L_p^3 - 2.008 \times 10^4 L_p^2$$  (4.66)
Table 4.14: Structural parameters of 0.1-μm pocket n-MOSFETs located at the lower design-space boundary and single-device values of ζ of the pocket devices to be used for developing an analytical formula for ζ. $V_{DS} = 0$ V and $V_{GS} = 0$ V.

<table>
<thead>
<tr>
<th>$L_p$ [μm]</th>
<th>$N_p$ [cm$^{-3}$]</th>
<th>$N_c$ [cm$^{-3}$]</th>
<th>$ζ$ [V]</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.022</td>
<td>$2.135 \times 10^{18}$</td>
<td>$2.375 \times 10^{17}$</td>
<td>$2.816 \times 10^{-2}$</td>
</tr>
<tr>
<td>0.024</td>
<td>$1.906 \times 10^{18}$</td>
<td>$2.175 \times 10^{17}$</td>
<td>$1.303 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.026</td>
<td>$1.715 \times 10^{18}$</td>
<td>$2.050 \times 10^{17}$</td>
<td>$2.051 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.028</td>
<td>$1.558 \times 10^{18}$</td>
<td>$1.925 \times 10^{17}$</td>
<td>$2.371 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.030</td>
<td>$1.424 \times 10^{18}$</td>
<td>$1.850 \times 10^{17}$</td>
<td>$2.654 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.032</td>
<td>$1.306 \times 10^{18}$</td>
<td>$1.825 \times 10^{17}$</td>
<td>$2.505 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.034</td>
<td>$1.205 \times 10^{18}$</td>
<td>$1.825 \times 10^{17}$</td>
<td>$2.743 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.036</td>
<td>$1.116 \times 10^{18}$</td>
<td>$1.838 \times 10^{17}$</td>
<td>$2.164 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.038</td>
<td>$1.039 \times 10^{18}$</td>
<td>$1.875 \times 10^{17}$</td>
<td>$2.200 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.040</td>
<td>$9.707 \times 10^{17}$</td>
<td>$1.925 \times 10^{17}$</td>
<td>$2.036 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.042</td>
<td>$9.095 \times 10^{17}$</td>
<td>$2.025 \times 10^{17}$</td>
<td>$1.871 \times 10^{-1}$</td>
</tr>
<tr>
<td>0.044</td>
<td>$8.557 \times 10^{17}$</td>
<td>$2.138 \times 10^{17}$</td>
<td>$1.810 \times 10^{-1}$</td>
</tr>
</tbody>
</table>

\[+5.926 \times 10^2 L_p - 5.993\]

\[
\ell_{p0} = 8.375 \times 10^3 L_p^4 + 5.882 \times 10^2 L_p^3 - 1.411 \times 10^2 L_p^2 \quad (4.67)
\]

\[+6.574 L_p - 6.968 \times 10^{-2}\]

\[
\ell_{c0} = -1.210 \times 10^5 L_p^4 + 1.704 \times 10^4 L_p^3 - 9.122 \times 10^2 L_p^2 \quad (4.68)
\]

\[+22.02 L_p - 0.1838\]

where $\ell_{p0}$, $\ell_{c0}$ and $L_p$ have the dimension of μm while $ζ$ has the dimension of V.

A dependence of $ζ$, $\ell_{p0}$ or $\ell_{c0}$ on $L_p$ alone is expected because $N_p$ and $N_c$ both are functions of $L_p$ at the lower design-space boundary. The global parameter values of $K_p$ of the surface potential model is directly calculated from Equation 4.66 for given values of $N_p$ and $L_p$. Figures 4.27 and 4.28 show comparisons of the
Figure 4.27: Comparison of global parameter values with single-device values of $\zeta$ of 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.

global parameter values with the single-device values of $\zeta$, $\ell_p$ and $\ell_c$ of the 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.

Since $V_{DS} = 0$ V and $I_D = 0$ nA/$\mu$m for the surface potential model, we have to use the potential profile of $\psi_{so}(x)$ in order to generate $N_c$ vs. $L_p$ of the lower design-space boundary. Figure 4.2 shows that all the pocket n-MOSFETs have their potential minima and maxima. Comparing the Medici surface potential maximum, $\psi_{so(max)}(L/2)$, or minimum, $\psi_{so(min)}(x_m)$, with the model surface potential generated by Equations 4.6, 4.10 - 4.12 and 4.66 - 4.68 at the point where the potential maximum or minimum exists, the model equations can generate $N_c$ vs. $L_p$ of the lower design-space boundary for the given model $N_p$ vs. $L_p$ in Figure 4.26.
Figure 4.28: Comparisons of global parameter values with single-device values of (a) $\ell_{po}$ and (b) $\ell_{co}$ of 0.1-\mu m pocket n-MOSFETs located at the lower design-space boundary.
The surface potential maximum $\psi_{s0}(\text{max})(L/2)$ is chosen to generate the model $N_c$ vs. $L_p$.

Solving $\psi_{s0}(\text{max})(L/2)(\text{Model}) - \psi_{s0}(\text{max})(L/2)(\text{Medici}) = 0$ for $N_c$ numerically, we find a model $N_c$ vs. $L_p$ shown in Figure 4.29. The Medici values of $\psi_{s0}(\text{max})(\text{Medici})$ of the twelve 0.1-μm pocket n-MOSFETs are shown in Figure 4.30. The errors of the model $N_c$ with reference to the Medici $N_c$ are within 1.1% for all the twelve 0.1-μm pocket n-MOSFETs located at the lower design-space boundary. Figure 4.31 shows a model prediction of a variation of the surface potential profile with the doping density at the center region of the 0.1-μm pocket n-MOSFET with the fixed values of $N_p$ and $L_p$. As discussed in Chapter 2, the surface potential of a MOS system decreases as the doping density is increased for $N_A > 1 \times 10^{17} \text{ cm}^{-3}$ in Figure 2.1 (a). Likewise, as $N_c$ of the 0.1-μm pocket n-MOSFET is increased in Figure 4.31, the surface potential at the center region as well as the surface potential minimum at the source-end pocket region decreases. The variation of the surface potential profile in Figure 4.31 can also be generated by the surface potential model, including the z-formulas for $K_p$, $\ell_{p0}$ and $\ell_{c0}$, in Section 4.1.

4.4.3 Model Prediction of Barrier Lowering Effect of Pocket n-MOSFETs

Polynomial fits to the single-device values of $\ell_{pS}$, $\ell_{pD}$ and $\ell_c$ for $V_{GS} = 0 \text{ V}$ and $V_{DS} = 1 \text{ V}$ can also provide the following analytical formulas for the model
Figure 4.29: Comparison of the model with the Medici $N_c$ vs. $L_p$ of the lower design-space boundary for $I_{off}$ in 0.1-$\mu$m pocket n-MOSFETs. Twelve Medici values of $\psi_{s0(\text{max})}(L/2)(\text{Medici})$ were used to find the model $N_c$ vs. $L_p$.

Figure 4.30: Medici values of $\psi_{s0(\text{max})}(L/2)(\text{Medici})$ of the twelve 0.1-$\mu$m pocket n-MOSFETs located at the lower design-space boundary.
Figure 4.31: Surface potential $\psi_{s0}$ profiles between source and drain for $V_{DS} = 0 \, V$. $V_{GS} = 0 \, V$ and three different values of $N_c$ of 0.1-$\mu m$ pocket n-MOSFET with $N_p = 2.104 \times 10^{18} \, cm^{-3}$ and $L_p = 0.022 \, \mu m$.

\[ \psi_{s0} \]

\[ \ell_{pS} = 1.261 \times 10^{5} L_p^4 - 1.461 \times 10^4 L_p^3 + 5.748 \times 10^2 L_p^2 - 8.025 L_p + 4.017 \times 10^{-2} \]  

\[ \ell_{pD} = 4.143 \times 10^4 L_p^4 - 5.146 \times 10^3 L_p^3 + 2.137 \times 10^2 L_p^2 - 2.883 L_p + 1.884 \times 10^{-2} \]  

\[ \ell_c = 5.105 \times 10^6 L_p^4 - 5.906 \times 10^5 L_p^3 + 2.383 \times 10^4 L_p^2 - 3.888 \times 10^2 L_p + 2.158 \]  

where $\ell_{pS}$, $\ell_{pD}$, $\ell_c$ and $L_p$ are in micrometers. Figures 4.32 - 4.33 shows the comparisons of the global values of the model parameters calculated from
Figure 4.32: Comparisons of global values with single-device values of (a) $\ell_{PS}$ and (b) $\ell_{PD}$ of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0$ V and $V_{DS} = 1$ V. Fourth-degree polynomials in terms of $L_p$ were used to formulate the model parameters.
Figure 4.33: Comparison of global values with single-device values of $\ell_c$ of 0.1-$\mu$m pocket n-MOSFETs for $V_{GS} = 0$ V and $V_{DS} = 1$ V. Fourth-degree polynomial in terms of $L_p$ was used to formulate the model parameter.

Equations 4.69 - 4.71 with their single-device values of 0.1-$\mu$m pocket devices for $V_{GS} = 0$ V and $V_{DS} = 1$ V. Figure 4.34 shows a model prediction of the barrier lowering effect due to the scaling down of the channel length of one of the well-designed pocket devices located at the lower design-space boundary. The barrier lowering is also small in the 0.1-$\mu$m pocket n-MOSFET for low-voltage applications. The subthreshold surface potential model for $\psi_s(x)$ in Section 4.2 together with Equations 4.66 - 4.68 and 4.69 - 4.71 was used to generate the potential profiles.
Figure 4.34: Variation of surface potential profile between source and drain of pocket n-MOSFET with channel length \( L \) for \( V_{GS} = 0 \) V and \( V_{DS} = 1 \) V. \( L_p = 0.024 \, \mu m \), \( N_p = 1.906 \times 10^{18} \, cm^{-3} \), \( N_c = 2.175 \times 10^{17} \, cm^{-3} \), \( t_{ox} = 4 \, nm \) and \( r_j = 0.06 \, \mu m \). The subthreshold surface potential model for \( \psi_s(x) \) was used to generate the potential profiles.

4.5 Summary

The subthreshold surface potential model, including the solutions of the quasi-two-dimensional Poisson's equation and the analytical and empirical formulas for the model parameters, was developed to generate surface potential profiles between source and drain of 0.1-\( \mu m \) pocket n-MOSFETs fast and accurately. The potential model proposed in this dissertation satisfies rigorously the boundary conditions for pocket n-MOSFETs, and predicts the DIBL effect in the well-designed 0.1-\( \mu m \) pocket n-MOSFETs, which are effective in suppressing short-channel effects.
The barrier lowering due to the scaling down of the channel length of the pocket devices is also predicted by the potential model. The empirical formulas were completed by comparing the quasi-two-dimensional model with two-dimensional device simulation results. The model for the off-state current together with the surface potential model for the pocket devices was used to generate the lower design-space boundary for $I_{off}$ in the deep-submicron ($L = 0.1 \, \mu m$) pocket n-MOSFETs fast and correctly; the closed-form model equations for the deep-submicron pocket n-MOSFETs reduce time and cost of two-dimensional simulation. Reduced time means more complete exploration of design space is possible within time constraints of design. The models also provide explicit relations between process, device and model parameters of the deep-submicron pocket n-MOSFETs.

In Chapter 5, the design-space boundary for $I_{on}$ in 0.1-$\mu m$ pocket n-MOSFETs is constructed using an algorithm and the two-dimensional device simulator. Complete model equations for $I_{on}$ developed to generate $I_D - V_{DS}$ curves and $I_{on}$ are presented. The model equations are used to generate the upper design-space boundary for the pocket devices.
CHAPTER 5

MODELING OF ON-STATE CURRENT

One of advantages of the scaling of MOSFETs down to deep-submicron dimensions is enhancement of the on-state current in the devices biased to strong inversion, which contributes to improvement in circuit performance. However, an increased doping level necessary to suppress short-channel effects of the deep-submicron MOSFETs can degrade the performance. This chapter discusses the modeling and design of deep-submicron \( L = 0.1 \, \mu m \) pocket n-MOSFETs biased to strong inversion at \( V_{GS} = 1 \, V \) and satisfying the boundary design specification of \( I_{on} = 0.2 \, mA/\mu m \) as well as the specifications of \( I_{off} \leq 1 \, nA/\mu m \) and \( \frac{\Delta I_{on}/I_{off}}{\Delta L/L} \leq 40\%/10\% \). In Section 5.1, the 0.1-\( \mu m \) pocket n-MOSFETs meeting the design specifications are explored using two-dimensional device simulation, and the values of the design variables \( N_p, N_c \) and \( L_p \) of the devices make up the unique upper design-space boundary for \( I_{on} = 0.2 \, mA/\mu m \). The characteristics of the 0.1-\( \mu m \) pocket n-MOSFETs located at the upper design-space boundary for \( I_{on} \) are presented. The MOSFET transconductance or the effective velocity averaging out velocities of all electrons participating in current flow over entire channel length and depth is evaluated to measure switching performance of the pocket n-MOSFETs. Section 5.2 provides model equations for the above-threshold and on-state current
in the deep-submicron pocket n-MOSFETs. The drift-diffusion current transport and Poisson's equations based on the charge-sheet approximation [50] are solved numerically within a gradual channel region to predict \( I_D - V_{DS} \) characteristics and \( I_{on} \) of the deep-submicron pocket n-MOSFETs. The transverse- and lateral-field-dependent mobility models described in Chapter 2 are essentially used for the modeling of the above-threshold current in the deep-submicron (0.1-\( \mu m \)) pocket n-MOSFETs. When the velocity saturation [3], [65], [66] of mobile electrons occurs along the channel between source and drain, a quasi-two-dimensional velocity saturation model [3], [65] is used to calculate a voltage drop across the velocity saturation region. Section 5.3 presents an algorithm developed on the basis of the model equations to generate \( I_D - V_{DS} \) curves and design-space boundaries for \( I_{on} \) in the deep-submicron pocket n-MOSFETs. Model results as well as model parameter values obtained through an optimization process are shown in the section.

Section 5.4 describes the analytical formulas for the model parameters, which are functions of structural parameters of the pocket n-MOSFET, developed to complete the model of the above-threshold current in the 0.1-\( \mu m \) pocket n-MOSFETs. Section 5.5 shows model results and predictions of design-space boundaries for \( I_{on} \).

Finally, we summarize this chapter and draw conclusions in Section 5.6.
5.1 Characteristics of Deep-Submicron Pocket n-MOSFETs Biased to Strong Inversion

As the channel length of a strongly inverted MOSFET is reduced (holding bias voltages fixed), longitudinal electric fields in the MOSFET increase, thereby increasing the electron velocity $v$ and the current density ($J = -qnv$). The electron velocity in Si increases monotonically as the electric field increases and saturates at $v_{sat} \approx 9.2 \times 10^6 \text{ cm/s}$ at a high field [68], [69]. However, when electrons enter a high-field region from a low-field region in a very short-channel n-MOSFET of the Si material, they are accelerated by the high electric field and then the electron velocity depending on the longitudinal-electric-field gradient in space, temperature and low-field mobility [73]-[77] can exceed the limited value of $v_{sat}$. The phenomenon is called velocity overshoot [69]-[83]. Velocity overshoot in an ultrashort-channel pocket n-MOSFET is discussed further with Medici simulation results in Appendix A. Note that this dissertation deals with n-MOSFETs of the Si material rather than devices of the direct-bandgap materials.

Assuming parasitic source and drain resistances of an n-MOSFET are negligible, the intrinsic transconductance is the same as the extrinsic transconductance. The intrinsic transconductance per unit width of the n-MOSFET defined in Equation 5.1 is directly proportional to the electron effective velocity [75]. Thus,

$$g_m \equiv \frac{\partial I_D}{\partial V_{GS}} \quad (5.1)$$
\[ g_m = C_{ox} \langle v(x, y) \rangle = C_{ox} v_{eff} \quad (5.2) \]

where \( I_D \) is the drain current per unit width, \( C_{ox} \) is the oxide capacitance per unit area, \( v(x, y) \) is the velocity [39], [40] of electrons at the position \((x, y)\) in channel region, and \( v_{eff} \) that is an average over entire channel length and depth is the effective velocity averaging out velocities of all electrons participating in current flow in the channel region of the n-MOSFET. The upper bound of \( g_m \) of long-channel n-MOSFETs is \( C_{ox} v_{sat} \). If the velocity becomes higher than \( v_{sat} \) over a large portion of the channel region of a very short-channel n-MOSFET, the velocity overshoot effect is measured by the increased transconductance.

The characteristics of 0.1-\(\mu\)m pocket n-MOSFETs biased to strong inversion are studied using some models available in Medici. Note that all Medici simulations concerning above-threshold currents in Chapters 5 and 6 were performed, using a new version of Medici (Version 1999.2) [40].

- Model C:

Drift-diffusion transport model,

longitudinal-field-dependent mobility model [41]–[43],

and transverse-field-dependent mobility model [44], [45]

for a low-longitudinal-field mobility.

Key Medici statements

\[ \text{MOBILITY SILICON FLDMOB=1 BETAN=2. VSATN=9.2E6} \]
Although measurements of \( v_{\text{sat}} \) have been reported to be changeable and the saturation velocity depends on the effective transverse electric field [66], [67], the value of the saturation velocity \( v_{\text{sat}} = 9.2 \times 10^6 \text{ cm/s} \) [45] is fixed for all the Medici simulations because the dependence of the saturation velocity on the effective transverse field is very weak [66] and Medici does not provide any model for the dependence of the saturation velocity on a local transverse electric field.

Based on the fact from the device physics that both the on-state current and the electron mobility decrease as \( N_p \) is increased at fixed \( N_c, L_p, L, t_{ox} \) and \( r_j \), a computer algorithm that is similar to the procedure POCKET_DOPING in Chapter 3 for the Medici simulation has been developed to automatically and accurately find the unique upper design-space boundary for any \( I_{\text{on}} \) within a desired error tolerance of \( I_{\text{on}} \). In order to construct a unique viable design space to locate the deep-submicron pocket n-MOSFETs meeting all the specifications of \( I_{\text{on}}, I_{\text{off}} \) and \( \frac{\Delta I_{\text{off}}}{I_{\text{off}}} \), the choice of the specification of \( I_{\text{on}} \) is important and depends on the specification of \( I_{\text{off}} \). If the boundary specification of \( I_{\text{on}} \) is too large, the viable
design space between the upper and the lower design-space boundaries for the deep-submicron pocket devices cannot be formed. However, the goal of this chapter is development of the computer algorithm for finding the upper design-space boundary for any choice of \( I_{on} \) in deep-submicron pocket n-MOSFETs from Medici simulations, and the model and algorithm for the above-threshold current to be used to generate \( I_D - V_{DS} \) curves and the upper design-space boundary. Figure 5.1 shows the upper design-space boundary \( N_p \) vs. \( L_p \) for \( I_{on} = 0.2 \, mA/\mu m \) in 0.1-\( \mu m \) pocket n-MOSFETs found for the same \( N_c \) values in Figure 5.1 (b) as those in Figure 3.7 (b). \( t_{ox} = 4 \, nm \) and \( r_j = 0.06 \, \mu m \). Note that the upper design-space boundary is unique because \( N_p \) is uniquely determined for given values of \( L, L_p, N_c, t_{ox} \) and \( r_j \). As described in Chapter 1, \( I_{on} = 0.2 \, mA/\mu m \) is the specification for low-power applications. The off-state current at the upper design-space boundary in Figure 5.1 is smaller than \( I_{off} = 0.714 \, nA/\mu m \) at the lower design-space boundary in Figure 3.7 because \( N_p \) at the upper design-space boundary is larger than that at the lower design-space boundary. Also, the sensitivity of off-state current to channel length at the upper design-space boundary is lower than the sensitivity \( \frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40\% / 10\% \) at the lower design-space boundary because \( N_c \) vs. \( L_p \) in Figure 5.1 (b) is based on the boundary specification of the sensitivity \( 40\% / 10\% \) and the off-state current at the upper design-space boundary is smaller than that at the lower design-space boundary. Thus, the increased doping density of the pocket region at the upper design-space boundary is more effective in suppressing the short-channel effects than that at
Figure 5.1: (a) The unique upper design-space boundary $N_p$ vs. $L_p$ for $I_{on} = 0.2 \text{ mA/\mu m}$ in 0.1-\textmu m pocket n-MOSFETs, and (b) $N_c$ vs. $L_p$ of all the devices located at the design-space boundary. This figure (b) is a repeat of Figure 3.7 (b) for the pocket devices satisfying the channel-length sensitivity and the off-current specifications.
the lower design-space boundary. Comparisons of the 0.1-μm pocket n-MOSFETs
located at the upper design-space boundary with the pocket devices located at the
lower design-space boundary will be made in detail in Chapter 6.

In order to examine profiles of potential, electric field and electron concentration
at the Si-SiO₂ interface, and \( I_D - V_{DS} \) characteristic of the deep-submicron pocket
n-MOSFET, a 0.1-μm pocket n-MOSFET with \( N_p = 2.713 \times 10^{18} \text{ cm}^{-3}, N_c =
2.175 \times 10^{17} \text{ cm}^{-3} \) and \( L_p = 0.024 \mu\text{m} \) is selected from the devices located at the
upper design-space boundary in Figure 5.1. Figure 5.2 shows Medici simulation
results of the potential, the longitudinal electric field, the electron density at the
Si-SiO₂ interface and the \( I_D - V_{DS} \) characteristic of the pocket device. The surface
potential increases monotonically in the channel from the source end \((x = 0)\) toward
the drain end \((x = L)\). The potential difference at the ends, \( \psi_s(L) - \psi_s(0) \), is
the potential drop in the channel region given by the drain voltage \( V_{DS} = 1 \text{ V} \).
Figure 5.2 (c) shows that the electron density in the vicinity of the source peaks,
while in the vicinity of the drain has a minimum. The maximum of the electron
density results from an increase in the electron density due to two-dimensional
effects at the source-end pocket and supply of electrons from the source, while the
minimum of the electron density results from a decrease in the oxide field due to
the high surface potential at the drain-end pocket.

Figure 5.3 shows \( I_D - V_{DS} \) characteristics of the 0.1-μm pocket n-MOSFET. The
drain current in the saturation mode on the \( I - V \) curves increases noticeably as
Figure 5.2: (a) Potential (b) electric field and (c) electron density profiles between source and drain at the Si-SiO₂ interface \((x,0)\) for \(V_{DS} = 1\) V and \(V_{GS} = 1\) V, and (d) \(I_D - V_{DS}\) characteristic at \(V_{GS} = 1\) V of the 0.1-\(\mu\)m pocket n-MOSFET with \(N_p = 2.713 \times 10^{18} \text{ cm}^{-3}\), \(N_c = 2.175 \times 10^{17} \text{ cm}^{-3}\) and \(L_p = 0.024 \mu\)m.
Figure 5.3: Medici $I_D - V_{DS}$ characteristics of the 0.1-$\mu m$ pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$ and $L_p = 0.024 \mu m$. The model C was used for the two-dimensional device simulations.

$V_{DS}$ is increased at a fixed $V_{GS}$ because the length $L_{sat}$ of the velocity saturation region increases; the channel-length modulation [3], [87] effect becomes larger, and then the ratio of $L_{sat}$ to the short-channel length $L$ of the pocket device is not negligible. Without bulk punchthrough, the higher the gate bias voltage, the larger the channel-length modulation in the deep-submicron pocket n-MOSFET. Figure 5.4 shows $I_D - V_{DS}$ characteristics at $V_{GS} = 1 \text{ V}$ of four different pocket n-MOSFETs located at the design-space boundary in Figure 5.1.
Figure 5.4: Medici $I_D - V_{DS}$ characteristics at $V_{GS} = 1 \text{ V}$ of 0.1-$\mu$m pocket n-MOSFETs with (a) $L_p = 0.022 \mu m$, $N_p = 2.995 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.375 \times 10^{17} \text{ cm}^{-3}$, (b) $L_p = 0.028 \mu m$, $N_p = 2.276 \times 10^{18} \text{ cm}^{-3}$, $N_c = 1.925 \times 10^{17} \text{ cm}^{-3}$, (c) $L_p = 0.038 \mu m$, $N_p = 1.580 \times 10^{18} \text{ cm}^{-3}$, $N_c = 1.875 \times 10^{17} \text{ cm}^{-3}$, (d) $L_p = 0.044 \mu m$, $N_p = 1.307 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.138 \times 10^{17} \text{ cm}^{-3}$, and of 0.1-$\mu$m conventional bulk n-MOSFET with $N_A = 1.098 \times 10^{18} \text{ cm}^{-3}$. All the devices are located at the design-space boundary in Figure 5.1. $t_{ox} = 4 \text{ nm}$ and $r_j = 0.06 \mu m$. The model C was used for the two-dimensional device simulations.
and of a 0.1-μm conventional bulk n-MOSFET, i.e., \( L_p = 0.05 \mu m \) case, with
\[ N_A = 1.098 \times 10^{18} \text{ cm}^{-3} \] and \( I_{on} = 0.2 \text{ mA/μm} \).

The MOSFET transconductance \( g_m \) at \( V_{GS} = 1 \text{ V} \) and \( V_{DS} = 1 \text{ V} \) are evaluated using the definition of \( g_m \) in Equation 5.1, and the drain currents from the
\( I_D - V_{DS} \) characteristics for \( V_{GS} = 1 \text{ V} \) and \( V_{GS} = 0.09 \text{ V} \). Then, the electron effective velocity of the n-MOSFET with a fixed \( t_{ox} = 4 \text{ nm} \) is directly calculated using Equation 5.2. Table 5.1 lists the transconductance, the electron effective velocity and the on-state current of the 0.1- and 1-μm pocket n-MOSFETs. The results of the model C in the table show that although the electron effective velocity of the 0.1-μm pocket n-MOSFET is higher than that of the 1-μm pocket n-MOSFET, neither of the effective velocities exceeds the electron saturation velocity (\( = 9.2 \times 10^6 \text{ cm/s} \)). Of course, this result is expected in a drift-diffusion transport model because velocity overshoot cannot happen. Comparison of the model C with the model B in the Appendix A, which does include the hydrodynamic model, also predicts that the effective velocity \( v_{eff} \) does not exceed \( v_{sat} \). Therefore, local velocity overshoot, when it occurs, does so only in a limited region of the channel.

<table>
<thead>
<tr>
<th>( L = 0.1 \mu m )</th>
<th>( L = 1 \mu m )</th>
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<tbody>
<tr>
<td>( g_m )</td>
<td>( v_{eff} )</td>
</tr>
<tr>
<td>Model C</td>
<td>582</td>
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Table 5.1: Transconductance \([\mu S/\mu m]\), effective electron velocity \([\text{cm/s}]\) and on-state current \([\text{A/μm}]\) at \( V_{GS} = 1 \text{ V} \) and \( V_{DS} = 1 \text{ V} \) of the 0.1- and 1-μm pocket n-MOSFETs with \( t_{ox} = 4 \text{ nm} \), \( L_p = 0.024 \mu m \), \( N_p = 2.713 \times 10^{18} \text{ cm}^{-3} \), \( N_e = 2.175 \times 10^{17} \text{ cm}^{-3} \) and \( r_f = 0.06 \mu m \).
and does not have a large effect on the effective velocity and the on-state current. Therefore, we can use the model C, which does not include velocity overshoot, in modeling the on-state current in the 0.1-\(\mu\)m pocket MOSFET and in determining the upper design-space boundary for \(I_{on}\).

The MOSFET transconductance is one of the most important parameters to evaluate the switching performance of MOSFETs. To know whether or not the deep-submicron pocket n-MOSFETs have advantages in performance over the deep-submicron conventional bulk n-MOSFETs, the 0.1-\(\mu\)m pocket n-MOSFETs existing at the upper design-space boundary for \(I_{on} = 0.2 \text{ mA/\(\mu\)m}\) in Figure 5.1 are compared to a 0.1-\(\mu\)m conventional bulk n-MOSFET with \(N_A = 1.098 \times 10^{18} \text{ cm}^{-3}\), which is the doping density found using Medici for the same \(I_{on} = 0.2 \text{ mA/\(\mu\)m}\), \(t_{ox}\) and \(r_j\) as those of the 0.1-\(\mu\)m pocket n-MOSFETs, regarding the transconductance \(g_m\) and the effective velocity \(v_{eff}\). Figure 5.5 shows that all the 0.1-\(\mu\)m pocket n-MOSFETs have higher switching performance than the 0.1-\(\mu\)m conventional bulk n-MOSFET for the same \(I_{on} = 0.2 \text{ mA/\(\mu\)m}\). At \(L_p = 0.05 \text{ \(\mu\)m}\), the 0.1-\(\mu\)m pocket n-MOSFET becomes the conventional bulk n-MOSFET. The conventional device has \(N_A = 1.098 \times 10^{18} \text{ cm}^{-3}\), \(I_{on} = 0.2 \text{ mA/\(\mu\)m}\), \(I_{off} = 3.66 \text{ pA/\(\mu\)m}\) and \(S = 486\%/10\%\) (Figure 6.12). Thus, the device satisfies the specifications of the on- and off-state currents, but not the specification of the sensitivity of off-state current to channel length. Figure 5.6 shows comparisons between the 0.1-\(\mu\)m
Figure 5.5: Comparisons between 0.1-μm pocket n-MOSFETs located at the upper design-space boundary for $I_{on}$ in Figure 5.1 and a 0.1-μm conventional bulk n-MOSFET with $N_A = 1.098 \times 10^{18} \text{ cm}^{-3}$ and the same $I_{on} = 0.2 \text{ mA/μm}$, $t_{ox} = 4 \text{ nm}$ and $r_j = 0.06 \text{ μm}$ as those of the pocket n-MOSFETs regarding (a) $g_m$ and (b) $v_{eff}$. $v_{eff}$ was calculated using Equations 5.1 and 5.2. The model C was used for all the Medici simulations. $V_{DS} = 1 \text{ V}$ and $V_{GS} = 1 \text{ V}$. 
Figure 5.6: (a) Electric potential and (b) longitudinal electric field profiles at the Si surface for the 0.1-\(\mu\)m pocket n-MOSFET with \(N_p = 2.713 \times 10^{18} \text{ cm}^{-3}\), \(N_c = 2.175 \times 10^{17} \text{ cm}^{-3}\) and \(L_p = 0.024 \text{ \(\mu\)m}\), and the 0.1-\(\mu\)m conventional bulk n-MOSFET with \(N_A = 1.098 \times 10^{18} \text{ cm}^{-3}\). The devices have the same \(I_{on} = 0.2 \text{ mA/\(\mu\)m}\). \(V_{DS} = 1 \text{ V}\) and \(V_{GS} = 1 \text{ V}\).
pocket n-MOSFET with $L_p = 0.024 \, \mu m$ and the 0.1-$\mu m$ conventional bulk n-MOSFET with the same $I_{on} = 0.2 \, mA/\mu m$ regarding $\psi_s$ and $E_x$. The high electric field around the boundary between the source-end pocket and center regions can contribute to improvement in the switching performance of the deep-submicron pocket n-MOSFET. As $L_p$ shrinks, $N_p$ and $N_c$ both increase in Figure 5.1, and the threshold voltages of the 0.1-$\mu m$ pocket n-MOSFETs increase. On a long-channel model basis, $g_m$ would decrease as $L_p$ reduces on this account. However, $g_m$ is larger for smaller $L_p$ in Figure 5.5 (a) because as $L_p$ is decreased, the electric field around the boundary between the source-end pocket and center regions becomes higher. Although the doping in the pocket region is $N_p = 2.7 \times 10^{18} \, cm^{-3}$ compared to $N_A = 1.1 \times 10^{18} \, cm^{-3}$ for the conventional device, Figure 5.6 shows that the potential and the field in the pocket and conventional structures coincide near the source because of large electron density near the source in both the pocket and conventional structures. Thus, threshold near the source does not matter because of a two-dimensional effect. The 0.1-$\mu m$ pocket n-MOSFETs located within the viable design space will be compared with 0.1-$\mu m$ conventional bulk n-MOSFETs selected to meet the same specifications in more detail in Chapter 6.

In the subsequent sections and chapters, the model C described in this section is used for all Medici simulations on above-threshold currents.
5.2 Model Equations for Above-Threshold Current

The model for the above-threshold current in deep-submicron pocket n-MOSFETs consists of model equations for the gradual channel and the velocity saturation regions. In the gradual channel region, magnitudes of transverse fields in the direction perpendicular to the Si surface are much greater than magnitudes of longitudinal fields in the direction of current flow, and the velocity of electrons in the channel region is lower than \( v_{sat} \). On the other hand, the velocity of electrons saturates at \( v_{sat} \) at high longitudinal fields in the velocity saturation region. In Figure 5.7, \( L_{ch} \)

Figure 5.7: Pocket n-MOSFET structure illustrating the gradual channel and velocity saturation regions.

and \( L_{sat} \) are the lengths of the gradual channel and the velocity saturation regions.
respectively. The channel length $L$ as a metallurgical source-to-drain separation or an effective physical-channel length is given by

$$L = L_{ch} + L_{sat} = 2L_p + L_c$$  \hspace{1cm} (5.3)

First, we derive the model equations for the above-threshold current flowing in the gradual channel region of the n-MOSFET with a non-uniform doping profile $N_A(x)$ between source and drain. Starting from the drift-diffusion current transport equation (Equation 2.31) and the Einstein relation, we have

$$I_D = qN_I(x)\mu_{eff}(x)E_z(x, y = 0) + qD_n \frac{dN_I}{dx}$$ \hspace{1cm} (5.4)

$$= -qN_I(x)\mu_{eff}(x)\frac{d\phi_s}{dx} + q\mu_{eff}(x)\nu_{ih} \frac{dN_I}{dx}$$

where the surface potential $\phi_s(x)$ is defined in Equation 2.7. Note that $I_D$ has a negative value because the drain current flows in the negative $x$-direction from the drain to the source. Assuming that there are no fixed and interfacial charges in the SiO$_2$, we can derive the following equation using the charge-sheet approximation [50], a solution of Poisson’s equation, the continuity of electric displacement at the Si-SiO$_2$ interface (Equation 2.11) and Equation 2.12.

$$\epsilon_{ox}E_{ox} = \epsilon_{ox} \frac{\phi_G(x) - \phi_s(x)}{\ell_{ox}} = C_{ox}(\phi_G(x) - \phi_s(x))$$

$$= \epsilon_s E_s = qN_I(x) + qN_A(x)W_D(x)$$

$$= qN_I(x) + qN_A(x)L_D(x)\sqrt{2\sqrt{\beta \phi_s(x)}} - 1$$  \hspace{1cm} (5.5)
where $\beta = 1/V_{th}$. From Equation 5.5 the electron charge density in the inversion layer is given by

$$qN_i(x) = \frac{C_{ox}(\phi_G(x) - \phi_s(x)) - \alpha \sqrt{N_A(x)} \sqrt{\beta \phi_s(x)} - 1}{\alpha}$$  (5.6)

where

$$\phi_G(x) = V_{GS} - V_{FB(x)} = V_{GS} + \psi_{NPOLY} + V_{th} \ln \left( \frac{N_A(x)}{n_i} \right)$$  (5.7)

$$\alpha \equiv \sqrt{2q\epsilon_s V_{th}}$$  (5.8)

Combining Equation 5.4 and Equation 5.6, we obtain

$$-\beta \frac{d\phi_s}{dx} \left[ 1 + \frac{qN_i(x)\beta}{C_{ox}} + \frac{\alpha'}{2} \sqrt{\frac{N_A(x)}{N_A(0)}} \sqrt{\beta \phi_s(x)} - 1 \right]$$

$$= \frac{\beta^2 I_D}{\mu_{eff}(x)C_{ox}} + \frac{1}{N_A(x)} \frac{dN_A(x)}{dx} \left[ \frac{\alpha'}{2} \sqrt{\frac{N_A(x)}{N_A(0)}} \sqrt{\beta \phi_s(x)} - 1 - 1 \right]$$  (5.9)

where

$$\alpha' \equiv \frac{\alpha \beta \sqrt{N_A(0)}}{C_{ox}}$$  (5.10)

We use the effective mobility model of Equation 2.22 in Chapter 2 for electrons in the gradual channel region. Thus,

$$\mu_{eff}(x) = \frac{\mu_{eff0}(x)}{\sqrt{1 + \left( \frac{\mu_{eff0}(x)\varepsilon_{sat}}{K_{sat}v_{sat}} \right)^2}}$$

$$= \frac{\mu_{eff0}(x)}{\sqrt{1 + \left( \frac{\varepsilon_{sat}}{\varepsilon_{sat}} \right)^2}}$$  (5.11)
where

\[ |E_x| = \frac{d\phi_s}{dx} > 0 \]

\[ E_{sat}(x) = K_{sat} \frac{v_{sat}}{\mu_{eff}(x)} \]

\[ v_{sat} = 9.2 \times 10^6 \text{ cm/s} \]

(5.12) (5.13) (5.14)

\( \mu_{eff}(x) \) is the effective transverse-field-dependent mobility \([44]\) of electrons in inversion layer at a low \( E_x \) given by Equation 2.26. \( K_{sat} \) is a dimensionless model parameter necessary to adjust model values of the saturation electric field \( E_{sat} \) by fitting model \( I_D - V_{DS} \) curves to two-dimensional simulation results. Evaluation of the parameter \( K_{sat} \) for deep-submicron pocket n-MOSFETs is discussed in Section 5.3. Combining Equation 5.9 and Equation 5.11, and solving a resulting quadratic equation for \( \frac{d(\beta \phi_s)}{dx} \), we obtain

\[ \frac{d(\beta \phi_s)}{dx} = \frac{-\mu_{eff}^{2}ac \pm \sqrt{(\mu_{eff}^{2}ac)^2 - \left\{ (\mu_{eff}^{2}a) - \left( \frac{b}{\beta E_{sat}} \right)^2 \right\} [(\mu_{eff}^{2}c)^2 - b^2]}}{(\mu_{eff}^{2}a)^2 - \left( \frac{b}{\beta E_{sat}} \right)^2} \]

(5.15)

where

\[ a \equiv 1 + \frac{qN_f(x)\beta}{Cox} + \frac{\alpha'}{2} \sqrt{\frac{N_A(x)}{N_A(0)}} \frac{1}{\sqrt{\beta \phi_s(x)} - 1} \]

(5.16)

\[ b \equiv \frac{\beta I_D}{Cox} \leq 0 \]

(5.17)

\[ c \equiv \frac{1}{N_A(x)} \frac{dN_A(x)}{dx} \left[ \frac{\alpha'}{2} \sqrt{\frac{N_A(x)}{N_A(0)}} \sqrt{\beta \phi_s(x)} - 1\right] \]

(5.18)

Since \( |E_x| = \frac{d\phi_s}{dx} \) has positive values, we choose the positive solution in Equation 5.15, and then obtain an expression for the longitudinal electric field and the
relation between the incremental channel length $\Delta x$ and the normalized incremen
tal surface potential $\Delta(\beta \phi_s)$ at a position $x$ in the channel.

$$|E_\varepsilon(x)| = V_{th} \frac{-\mu_{eff}ac + \sqrt{(\mu_{eff}ac)^2 - \left[(\mu_{eff}ac)^2 - \left(\frac{b}{\beta \varepsilon_{sat}}\right)^2\right]^2}}{(\mu_{eff}ac)^2 - \left(\frac{b}{\beta \varepsilon_{sat}}\right)^2} (5.19)$$

$$\Delta x = \frac{\Delta(\beta \phi_s) \left[(\mu_{eff}ac)^2 - \left(\frac{b}{\beta \varepsilon_{sat}}\right)^2\right]}{-\mu_{eff}ac + \sqrt{(\mu_{eff}ac)^2 - \left[(\mu_{eff}ac)^2 - \left(\frac{b}{\beta \varepsilon_{sat}}\right)^2\right]^2}} (5.20)$$

Note that $\Delta x > 0$ since $\Delta \phi_s > 0$. Assuming that the electron quasi-Fermi
level [50], [56] at the source end ($x = 0$) of the channel layer at non-equilibrium,
i.e., $I_D \neq 0$ is almost the same as the corresponding electron Fermi level at thermal
equilibrium, we can find the surface potential at $x = 0$ by solving the following
equation, i.e., Equation 2.13 for $\phi_s(0)$, using a numerical method.

$$C_{ox}(\phi_C(0) - \phi_s(0)) = qN_A(0)L_D(0)\sqrt{2} \sqrt{\beta \phi_s(0) - 1 + \left(\frac{n_i}{N_A(0)}\right)^2} e^\beta \phi_s(0) (5.21)$$

The incremental channel length $\Delta x$ at $x$ and $\phi_s(x)$ for given $I_D$ and $\Delta \phi_s$ is found
using Equation 5.20. Expanding the surface potential $\phi_s(x + \Delta x)$ for a non-zero
$I_D$ in a Taylor series, we have a relation between the surface potentials at $x$ and
$x + \Delta x$. Thus,

$$\phi_s(x + \Delta x) = \phi_s(x) + \frac{\partial \phi_s}{\partial x} \Delta x + \cdots (5.22)$$

$$\simeq \phi_s(x) + \Delta \phi_s (5.23)$$
Increasing both \( x \) and \( \phi_s(x) \) gradually from \( x = 0 \) up to \( x = L_{ch} \), we can find the voltage drop across the gradual channel region. Thus,

\[
V_{ch} = \phi_s(L_{ch}) - \phi_s(0)
\]

The position of \( x = L_{ch} \) is determined when \( |\mathcal{E}_x| \) of Equation 5.19 is equal to \( \mathcal{E}_{sat} \) of Equation 5.13.

We can reduce the generalized model equations described above to the case of the pocket n-MOSFET with a lateral-step doping profile, and apply the reduced model equations to each of the pocket and the center regions in the pocket n-MOSFET, subject to the following boundary conditions of the continuity of the potential at each of the boundaries between the pocket and the center regions.

\[
\phi_s(x^-_1) = \phi_s(x^+_1)
\]

\[
\phi_s(x^-_2) = \phi_s(x^+_2)
\]

The electric field at the boundaries between the pocket and the center regions is continuous, too. Thus,

\[
\frac{\Delta \phi_s}{\Delta x}igg|_{x=x^-_1} = \frac{\Delta \phi_s}{\Delta x}igg|_{x=x^+_1}
\]

\[
\frac{\Delta \phi_s}{\Delta x}igg|_{x=x^-_2} = \frac{\Delta \phi_s}{\Delta x}igg|_{x=x^+_2}
\]

However, the difference between the results, i.e., \( I_D - V_{DS} \) curves obtained with and without using the boundary conditions of Equations 5.27 and 5.28 is very small. For the pocket n-MOSFETs, \( c = 0 \) in Equation 5.18 and \( N_A(x) = N_A(0) \) at each of
the pocket and center regions. Then, we have the reduced forms of Equations 5.19 and 5.20 as follows.

\[ |E_x(x)| = V_{th} \frac{\sqrt{(\mu_{eff} a)^2 - \left(\frac{b}{\beta E_{sat}}\right)^2}}{(\mu_{eff} a)^2 - \left(\frac{b}{\beta E_{sat}}\right)^2} \]  

\[ \Delta x = \frac{\Delta(\beta \phi_s) \left[ (\mu_{eff} a)^2 - \left(\frac{b}{\beta E_{sat}}\right)^2 \right]}{|b|\sqrt{(\mu_{eff} a)^2 - \left(\frac{b}{\beta E_{sat}}\right)^2}} \]  

where

\[ a \equiv 1 + \frac{qN_i(x)\beta}{C_{ox}} + \frac{\alpha\beta\sqrt{N_x}}{2C_{ox}} \frac{1}{\sqrt{\beta \phi_s(x)} - 1} \]  

\[ b \equiv \frac{\beta^2 I_D}{C_{ox}} \]  

\[ N_x = N'_p \text{ or } N_c \]  

\[ N'_p = \kappa_p N_p \]

\( \kappa_p \) in Equation 5.34 is the second dimensionless model parameter to be found by fitting the model \( I_D - V_{DS} \) curves to two-dimensional simulation results. The model parameter is required for the model of the deep-submicron pocket n-MOSFETs to include an increase in the electron density due to two-dimensional effects at the heavily doped and short pocket region near source. For example, two-dimensional simulation results for the 0.1-\( \mu m \) pocket n-MOSFET with \( L_p = 0.024 \mu m \), \( N_p = 2.713 \times 10^{18} \text{ cm}^{-3} \) and \( N_c = 2.175 \times 10^{17} \text{ cm}^{-3} \), existing at the upper design-space boundary in Figure 5.1, show that a minimum of the electron density at the pocket region near source at non-equilibrium is \( 1.39 \times 10^{12} \text{ cm}^{-2} \) at \( x = 0.024 \mu m \) for
\( V_{GS} = 1 \, \text{V} \) and \( V_{DS} = 1 \, \text{V} \), and a minimum value at equilibrium is \( 1.86 \times 10^{12} \, \text{cm}^{-2} \) at \( x = 0.024 \, \mu\text{m} \) for \( V_{GS} = 1 \, \text{V} \) and \( V_{DS} = 0 \, \text{V} \), while the electron density at equilibrium evaluated by using Equations 5.6 and 5.21 is just \( 2.51 \times 10^{10} \, \text{cm}^{-2} \) for \( V_{GS} = 1 \, \text{V} \) and \( V_{DS} = 0 \, \text{V} \), which is far smaller than the minimum values of the two-dimensional simulations. In addition, the long-channel threshold voltage \( V_T \) of the source-end pocket predicted by the model is 1.05 \( \text{V} \), which is higher than the applied gate bias voltage 1 \( \text{V} \). This model result indicates that the source pocket region is not in strong inversion. The small electron density and the high threshold voltage of the model lead to a large difference of the above-threshold current between the model and the Medici. With \( \kappa_p = 0.3875 \), the electron density at the pocket region evaluated using the model equations is \( 1.55 \times 10^{12} \, \text{cm}^{-2} \), which is a correct value comparable to the values of the two-dimensional simulation, and the threshold voltage is lowered to 0.59 \( \text{V} \). Therefore, without transformation from the heavily doped pocket to a less heavily doped pocket, the model equations can correctly predict neither the \( I_D - V_{DS} \) characteristic at \( V_{GS} = 1 \, \text{V} \) nor an increase in \( I_{on} \) of the 0.1-\( \mu\)m pocket n-MOSFET. The surface potential \( \phi_s(0) \) at the source end of the gradual channel region of the deep-submicron pocket n-MOSFET is found by solving the following equation numerically. Note that the beginning of the gradual channel region is assumed to start at \( x = 0 \) in Figure 5.7.

\[
C_{ox}(\phi_G(0) - \phi_s(0)) = qN'_pL_{D(p)}\sqrt{2} \sqrt{\beta \phi_s(0)} - 1 + \left( \frac{n_i}{N'_p} \right)^2 e^{\beta \phi_s(0)} \tag{5.35}
\]
where $N_p'$ is defined in Equation 5.34 and

$$\phi_G(0) = V_{GS} + \psi_{NPOLY} + V_{th} \ln \left( \frac{N_p'}{n_i} \right)$$  \hspace{1cm} (5.36)

$$L_{D(p)} = \sqrt{\frac{\varepsilon_s V_{th}}{q N_p'}}$$  \hspace{1cm} (5.37)

In the region called the velocity saturation region expanding from the end of the gradual channel region ($x = L_{ch}$) to the end of the physical channel region ($x = L$), the electron velocity saturates at $v_{sat}$ due to high longitudinal electric fields, which violate the gradual channel approximation, and electrons spread out widely toward the drain region, which also violates the charge sheet approximation. Therefore, the charge sheet model and the gradual channel approximation are no longer valid within the velocity saturation region. The quasi-two-dimensional velocity saturation model proposed in [65] is used to calculate the voltage drop across the velocity saturation region. The underlying assumptions or approximations made for developing the model are as follows:

1. The drain junction is abrupt, and the resistance at the drain region is null.

2. The current flow in the velocity saturation region is confined within the junction depth.

3. Electric flux or field lines passing through the bottom edge of the velocity saturation region as a Gaussian box of height $r_j$ are much smaller than those crossing top, right and left edges, and thus are ignored.
4. The longitudinal electric field inside the velocity saturation region is independent of the y-coordinate.

5. The mobile electron density within the velocity saturation region is independent of the position \((x, y)\).

6. The gradient of the longitudinal electric field at \(x = L_{ch}\) is zero.

Assumption 1 agrees with the assumptions of the pocket MOSFET structure and the zero drain resistance made in Chapter 1. Assumption 2 is acceptable for \(r_j = 0.06 \mu m\), \(V_{GS} = 1 V\) and \(V_{DS} = 1 V\). Two-dimensional simulation results also support it. Assumption 3 is acceptable because the transverse electric field decreases from the maximum at the Si-SiO\(_2\) interface to zero in the Si bulk. Assumption 4 is acceptable because the gradual channel approximation is no longer valid within the velocity saturation region, the drain-end pocket region is uniformly doped and the drain resistance was assumed to be zero. Assumptions 5 and 6 can cause some errors that should not be serious. However, the two assumptions are necessary to make the quasi-two-dimensional velocity saturation model as simple as possible.

Consider a Gaussian box of height \(r_j\) and of length \(\Delta x = x - L_{ch}\) in the velocity saturation region in Figure 5.7, containing mobile electron and depletion charges. Applying Gauss's law to the Gaussian box, we obtain

\[
\varepsilon_sE_{sat}r_j - \varepsilon_s|E_z(x)|r_j - \varepsilon_{oz}\int_{L_{ch}}^{x} \frac{\phi_G - \phi_z(x')}{t_{oz}} dx' = (-qN'_p r_j - qnr_j)(x - L_{ch}) \quad (5.38)
\]
where

\[ |\mathcal{E}_x(x)| = \frac{dV(x)}{dx} \quad (5.39) \]

\[ \phi_G = V_{GS} - V_{FB} \quad (5.40) \]

\[ \phi_s(x) = 2|\psi_B| + V(x) \quad (5.41) \]

and \( n \) is the mobile electron density that is independent of the position \((x, y)\) within the velocity saturation region. Note that the electric field at the Si-SiO₂ interface of the velocity saturation region decreases from the maximum at \( x = L_{ch} \) to the minimum at \( x = L \). Differentiating Equation 5.38 with respect to \( x \) yields

\[ \epsilon_s r_j \frac{d|\mathcal{E}_x(x)|}{dx} + C_{oz}(\phi_G - 2|\psi_B| - V(x)) = qr_j(N'_p + n) \quad (5.42) \]

Assuming \( \frac{d|\mathcal{E}_x(x)|}{dx} = 0 \) at \( x = L_{ch} \), then we obtain from Equation 5.42

\[ V_{ch} \approx \phi_G - 2|\psi_B| - \frac{qr_j(N'_p + n)}{C_{oz}} \quad (5.43) \]

Combining Equations 5.42 and 5.43 we have

\[ \epsilon_s r_j \frac{d|\mathcal{E}_x(x)|}{dx} + C_{oz}(-V(x) + V_{ch}) = 0 \quad (5.44) \]

Solving the differential equation of Equation 5.44 for \( V(x) \) subject to the following boundary conditions,

\[ V(L_{ch}) = V_{ch} \quad (5.45) \]

\[ |\mathcal{E}_x(L_{ch})| = \mathcal{E}_{sat} \quad (5.46) \]
we obtain the solution $V(x)$. Thus,

$$V(x) = V_{ch} + \ell_{sat}E_{sat} \sinh \left( \frac{x - L_{ch}}{\ell_{sat}} \right), \text{ for } L_{ch} \leq x \leq L \quad (5.47)$$

where $\ell_{sat}$ is a characteristic length at the velocity saturation region given by

$$\ell_{sat} = \sqrt{\frac{\epsilon_s}{\epsilon_{ox}T_j}} \quad (5.48)$$

The expression for the electric field $E_x(x)$ is derived from Equations 5.39 and 5.47.

$$|E_x(x)| = E_{sat} \cosh \left( \frac{x - L_{ch}}{\ell_{sat}} \right), \text{ for } L_{ch} \leq x \leq L \quad (5.49)$$

The voltage drop, $V_{DS} = V(L)$, across the whole channel region between source and drain, which is the drain voltage, is given by

$$V_{DS} = V_{ch} + \ell_{sat}E_{sat} \sinh \left( \frac{L - L_{ch}}{\ell_{sat}} \right)$$

$$= V_{ch} + \ell_{sat}E_{sat} \sinh \left( \frac{L_{sat}}{\ell_{sat}} \right) \quad (5.50)$$

Without finding accurate values of $\ell_{sat}$, $\kappa_p$ and $K_{sat}$, the quasi-two-dimensional and the charge-sheet models cannot generate correctly $I_D - V_{DS}$ curves and $I_{on}$ values of Medici two-dimensional simulation results for the deep-submicron pocket n-MOSFETs. Therefore, $\ell_{sat}$, $\kappa_p$ and $K_{sat}$ as the model parameters will be found by fitting the model $I_D - V_{DS}$ curves to the Medici simulation results.
5.3 Model Results and Parameter Extraction

The $I_D - V_{DS}$ curve and $I_{on}$ of a deep-submicron pocket n-MOSFET are found for given optimal values of the model parameters $K_{sat}$, $\kappa_p$ and $\ell_{sat}$, using the procedure described below.

1. Set $V_{GS} = 1 \text{ V}$. Let $I_D = 0 \text{ A/cm}$ and $V_{DS} = 0 \text{ V}$ be the initial drain current and voltage, respectively. Let $\Delta I_D = -0.01 \text{ A/cm}$ be an incremental drain current.

2. $I_D \leftarrow I_D + \Delta I_D$, and $x \leftarrow 0$.

3. Find the surface potentials at the starting points $x = 0$, $x_1$ and $x_2$ at the pocket region near source, the center region, and the pocket region near drain, respectively.

- If $x = 0$ at the pocket region near source, then compute the surface potential $\phi_s(0)$ numerically using Equation 5.35. $\beta \phi_s(x) \leftarrow \beta \phi_s(0)$.

- If $x = x_1$ at the center region, then $\phi_s(x_1^+) = \phi_s(x_1^-)$ (boundary condition 5.25). $\beta \phi_s(x) \leftarrow \beta \phi_s(x_1^+)$.

- If $x = x_2$ at the pocket region near drain, then $\phi_s(x_2^+) = \phi_s(x_2^-)$ (boundary condition 5.26). $\beta \phi_s(x) \leftarrow \beta \phi_s(x_2^+)$.  

Set $\Delta(\beta \phi_s) = 10^{-3}$ (a normalized incremental surface potential). If necessary, normalize the distance $x$ in each region by $L_p$ or $L_c$. 
4. Calculate $\mu_{e f f 0}(x), N_i(x), E_{sat}(x), |E_z(x)|$ and $\Delta x$ at $x$, using Equations 2.26, 5.6, 5.13, 5.29 and 5.30, respectively.

- If $x > x_1$, $x > x_2$ or $x > L$, and $|E_z(x)| < E_{sat}(x)$, decrease $\Delta(\beta_0)$, e.g., $\Delta(\beta_0) \leftarrow \Delta(\beta_0)/2$ to accurately find all the parameter values at the boundary $x_1$, $x_2$ or $L$. $x \leftarrow x - \Delta x$ and $\beta_0(x) \leftarrow \beta_0(x - \Delta x)$. Go to the step 5, and repeat the steps 4 and 5 until $x = x_1$, $x = x_2$ or $x = L$ within an error tolerance, e.g., $10^{-7}$ under the condition of $|E_z(x)| < E_{sat}(x)$.

- If $|E_z(x)| > E_{sat}(x)$ and $x < L$, decrease $\Delta(\beta_0)$, e.g., $\Delta(\beta_0) \leftarrow \Delta(\beta_0)/2$ to accurately find all the parameter values at $x = L_{ch}$. $x \leftarrow x - \Delta x$ and $\beta_0(x) \leftarrow \beta_0(x - \Delta x)$. Go to the step 5, and repeat the steps 4 and 5 until $|E_z(L_{ch})| = E_{sat}(L_{ch})$ within an error tolerance, e.g., $10^{-7}$.

5. $\beta_0(x + \Delta x) = \beta_0(x) + \Delta(\beta_0)$, using Equation 5.23. $x \leftarrow x + \Delta x$ and $\beta_0(x) \leftarrow \beta_0(x + \Delta x)$.

6. - If $x = L$ and $|E_z(L)| < E_{sat}(L)$, calculate $V_{DS} = \phi_s(L) - \phi_s(0)$ for the given $I_D$. Return $V_{DS}$.

- If $L_{ch} \leq L$, calculate $V_{DS} = \phi_s(L_{ch}) - \phi_s(0) + \ell_{sat}E_{sat} \sinh \left( \frac{L - L_{ch}}{\ell_{sat}} \right)$ (Equations 5.24, 5.50) for the given $I_D$. If $V_{DS} = 1 V$, $I_{on} = I_D$. Return $V_{DS}$.

7. Repeat the steps 3 - 6.
8. Repeat the steps 2 – 7 until \( V_{DS} = 1 \) V.

Figures 5.8 shows comparisons of model \( I_D - V_{DS} \) characteristics to Medici’s of four different 0.1-\( \mu m \) pocket n-MOSFETs located at the upper design-space boundary for \( I_{on} = 0.2 \) mA/\( \mu m \). The model results were obtained by using the algorithm described above and an optimization technique for minimizing errors between the model and Medici \( I_D - V_{DS} \) curves. Note that all the pocket n-MOSFETs have the fixed \( t_{ox} = 4 \) nm and \( r_j = 0.06 \) \( \mu m \). Optimal values of the model parameters \( K_{sat} \), \( \kappa_p \) and \( \ell_{sat} \) for the 0.1-\( \mu m \) pocket n-MOSFETs have been found through the optimization process. Since optimized results showed that \( K_{sat} \) depends only weakly on a new parameter \( z = LL_pN_pN_{ctox}r_j [cm^{-2}] \) or the structural parameters of the pocket n-MOSFET for 0.022 \( \mu m \) \( \leq L_p \leq 0.044 \) \( \mu m \), one value for \( K_{sat} = 2.3 \) selected from the optimal values of \( K_{sat} \) has been used to generate excellent fits to Medici \( I_D - V_{DS} \) curves. Note that the parameter \( z \) as a composite variable consists of all the design variables, including \( t_{ox} \) and \( r_j \), of the pocket n-MOSFET for further extensions of the dependence of the model parameters on the structural parameters of the pocket n-MOSFET. However, since no test has been made of \( t_{ox} \) and \( r_j \) dependence, we do not know that \( z = LL_pN_pN_{ctox}r_j \) is a useful variable as far as \( L, t_{ox} \) and \( r_j \) are concerned.

A single-device value of a model parameter in this chapter is an optimal value, which is extracted by fitting a model \( I_D - V_{DS} \) curve to a Medici \( I_D - V_{DS} \) curve. The single-device values of the parameters \( \kappa_p \) and \( \ell_{sat} \) vs. the parameter \( z \) for the
Figure 5.8: Model $I_D - V_{DS}$ characteristics at $V_{GS} = 1\ V$ compared to Medici data for the 0.1-$\mu$m pocket n-MOSFETs with (a) $L_p = 0.022\ \mu m, N_p = 2.995 \times 10^{18}\ cm^{-3}, N_c = 2.375 \times 10^{17}\ cm^{-3}$, (b) $L_p = 0.028\ \mu m, N_p = 2.276 \times 10^{18}\ cm^{-3}, N_c = 1.925 \times 10^{17}\ cm^{-3}$, (c) $L_p = 0.038\ \mu m, N_p = 1.580 \times 10^{18}\ cm^{-3}, N_c = 1.875 \times 10^{17}\ cm^{-3}$, and (d) $L_p = 0.044\ \mu m, N_p = 1.307 \times 10^{18}\ cm^{-3}, N_c = 2.138 \times 10^{17}\ cm^{-3}$. Single-device values of $\kappa_p, \ell_{sat}$ and $K_{sat} = 2.3$ were used for the model.
Figure 5.9: The extracted optimal values as single-device values of the model parameters (a) $\kappa_p$ and (b) $\ell_{sat}$ vs. $z = LL_pN_pN_{ctox}r_j$ of the 0.1-$\mu m$ pocket n-MOSFETs located at the upper design-space boundary for $I_{on} = 0.2 \text{ mA}/\mu \text{m}$. $K_{sat} = 2.3$ and $V_{GS} = 1 \text{ V}$. 
Table 5.2: Structural parameters of the 0.1-\(\mu\)m pocket n-MOSFETs located at the upper design-space boundary in Figure 5.1, and the best single-device values of \(\kappa_p\) and \(\ell_{sat}\) of the pocket devices. \(K_{sat} = 2.3\) and \(V_{GS} = 1\) V.

<table>
<thead>
<tr>
<th>(L_p [\mu m])</th>
<th>(N_p [cm^{-3}])</th>
<th>(N_c [cm^{-3}])</th>
<th>(\kappa_p)</th>
<th>(\ell_{sat} [cm])</th>
</tr>
</thead>
<tbody>
<tr>
<td>0.022</td>
<td>(2.995 \times 10^{18})</td>
<td>(2.375 \times 10^{17})</td>
<td>(3.554 \times 10^{-1})</td>
<td>(6.990 \times 10^{-7})</td>
</tr>
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<td>0.024</td>
<td>(2.713 \times 10^{18})</td>
<td>(2.175 \times 10^{17})</td>
<td>(3.875 \times 10^{-1})</td>
<td>(7.764 \times 10^{-7})</td>
</tr>
<tr>
<td>0.026</td>
<td>(2.479 \times 10^{18})</td>
<td>(2.050 \times 10^{17})</td>
<td>(4.175 \times 10^{-1})</td>
<td>(8.386 \times 10^{-7})</td>
</tr>
<tr>
<td>0.028</td>
<td>(2.276 \times 10^{18})</td>
<td>(1.925 \times 10^{17})</td>
<td>(4.472 \times 10^{-1})</td>
<td>(9.044 \times 10^{-7})</td>
</tr>
<tr>
<td>0.030</td>
<td>(2.104 \times 10^{18})</td>
<td>(1.850 \times 10^{17})</td>
<td>(4.740 \times 10^{-1})</td>
<td>(9.503 \times 10^{-7})</td>
</tr>
<tr>
<td>0.032</td>
<td>(1.948 \times 10^{18})</td>
<td>(1.825 \times 10^{17})</td>
<td>(5.008 \times 10^{-1})</td>
<td>(1.004 \times 10^{-6})</td>
</tr>
<tr>
<td>0.034</td>
<td>(1.815 \times 10^{18})</td>
<td>(1.825 \times 10^{17})</td>
<td>(5.277 \times 10^{-1})</td>
<td>(1.057 \times 10^{-6})</td>
</tr>
<tr>
<td>0.036</td>
<td>(1.690 \times 10^{18})</td>
<td>(1.838 \times 10^{17})</td>
<td>(5.531 \times 10^{-1})</td>
<td>(1.091 \times 10^{-6})</td>
</tr>
<tr>
<td>0.038</td>
<td>(1.580 \times 10^{18})</td>
<td>(1.875 \times 10^{17})</td>
<td>(5.799 \times 10^{-1})</td>
<td>(1.139 \times 10^{-6})</td>
</tr>
<tr>
<td>0.040</td>
<td>(1.487 \times 10^{18})</td>
<td>(1.925 \times 10^{17})</td>
<td>(6.054 \times 10^{-1})</td>
<td>(1.170 \times 10^{-6})</td>
</tr>
<tr>
<td>0.042</td>
<td>(1.393 \times 10^{18})</td>
<td>(2.025 \times 10^{17})</td>
<td>(6.322 \times 10^{-1})</td>
<td>(1.200 \times 10^{-6})</td>
</tr>
<tr>
<td>0.044</td>
<td>(1.307 \times 10^{18})</td>
<td>(2.138 \times 10^{17})</td>
<td>(6.590 \times 10^{-1})</td>
<td>(1.221 \times 10^{-6})</td>
</tr>
</tbody>
</table>

The optimal value \(K_{sat} = 2.3\) are shown on logarithmic scales in Figure 5.9, and the values are also listed in Table 5.2. In Figure 5.9 the lower branches of the curves are for the 0.1-\(\mu\)m pocket n-MOSFETs with \(N_p\) and \(N_c\) for \(L_p \leq 0.034\ \mu\)m (Type A devices) in Figure 5.1 while the upper ones are for the pocket devices with \(N_p\) and \(N_c\) for \(L_p > 0.034\ \mu\)m (Type B devices) in Figure 5.1. As discussed in Chapter 3, the double-branch curves in Figure 5.9 result solely from the characteristic of \(N_c\) vs. \(L_p\) shown in Figure 5.1 (b).

Rms errors of the drain voltages of the model \(I_D - V_{DS}\) curves to those of the Medici \(I_D - V_{DS}\) curves in Figure 5.8 for the twelve 0.1-\(\mu\)m pocket n-MOSFETs located at the upper design-space boundary are shown in Figure 5.10. The rms
Figure 5.10: Rms errors of drain voltages of model $I_D - V_{DS}$ curves to those of Medici $I_D - V_{DS}$ curves at $V_{GS} = 1\, V$ of the 0.1-$\mu m$ pocket n-MOSFETs located at the upper design-space boundary. The single-device values of $\kappa_p$ and $\ell_{sat}$ in Table 5.2, and $K_{sat} = 2.3$ were used to obtain the model $I_D - V_{DS}$ curves.

Error is defined as

\[
\text{rms error} = \sqrt{\frac{1}{N} \sum_{i=1}^{N} \left( \frac{V_{DS}(I_{Di}) - V_{DS}^*(I_{Di})}{V_{DS}(I_{Di})} \right)^2}
\]  

(5.51)

where $V_{DS}(I_{Di})$ and $V_{DS}^*(I_{Di})$ are Medici and model drain voltages, respectively, for a drain current $I_{Di}$ on a Medici $I_D - V_{DS}$ curve at $V_{GS} = 1\, V$. $V_{DS}^*(I_{Di})$ is an approximation to $V_{DS}(I_{Di})$ and $N$ is the number of data of $V_{DS}(I_{Di})$. 
5.4 Formulation of Model Parameters of $\kappa_p$ and $\ell_{sat}$

A global parameter value is calculated from an analytical and empirical formula for a model parameter. This formula is fitted to the single-device values available for the model parameter. The fitted formula can take one of several forms. In this section two possibilities are considered: the $z$-formulation and the power-law formulation.

5.4.1 $z$-Formulation

The formulas of Equations 4.14 and 4.15 proposed in Section 4.1 are used to formulate the model parameters of $\kappa_p$ and $\ell_{sat}$ in terms of the composite variable $z = LL_p N_p V_c t_{ox} r_j$. Taking the second-degree polynomials of $\log_{10} z$ for the model formulas, we have

$$\log_{10} \kappa_p = \beta_{\kappa_p}^0 (\log_{10} z)^2 + \beta_{\kappa_p}^1 (\log_{10} z) + \log_{10} \alpha_{\kappa_p} \quad (5.52)$$

$$\log_{10} \ell_{sat} = \beta_{\ell_{sat}}^0 (\log_{10} z)^2 + \beta_{\ell_{sat}}^1 (\log_{10} z) + \log_{10} \alpha_{\ell_{sat}} \quad (5.53)$$

$\kappa_p$ and $\ell_{sat}$ are directly obtained by transforming inversely the logarithmic functions $\log_{10} \kappa_p$ and $\log_{10} \ell_{sat}$ in Equations 5.52 and 5.53 into their corresponding exponential functions. Thus,

$$\kappa_p = 10^{\log_{10} \kappa_p} = \alpha_{\kappa_p} z^{\beta_{\kappa_p}^0} (\log_{10} z)^{\beta_{\kappa_p}^1} \quad (5.54)$$

$$\ell_{sat} = 10^{\log_{10} \ell_{sat}} = \alpha_{\ell_{sat}} z^{\beta_{\ell_{sat}}^0} (\log_{10} z)^{\beta_{\ell_{sat}}^1} \quad (5.55)$$
Note that the first equations in Equations 5.54 and 5.55 are recommended in evaluating the model parameters in order to avoid possible overflow or underflow in using the second equations in Equations 5.54 and 5.55. Fitting the polynomials in Equations 5.52 and 5.53 to the single-device values of $\kappa_p$ and $\ell_{sat}$ in Table 5.2 on logarithmic scales, we can find the coefficients of the second-degree polynomials for the two different types of the 0.1-$\mu$m pocket n-MOSFETs, shown in Tables 5.3 and 5.4.

Figures 5.11 and 5.12 show comparisons of the second-degree polynomials of $\log_{10} \kappa_p$ and $\log_{10} \ell_{sat}$ to the corresponding single-device values of the 0.1-$\mu$m pocket n-MOSFETs.

Table 5.3: The coefficients of the second-degree polynomial of $\log_{10} \kappa_p$ of the two types of the 0.1-$\mu$m pocket n-MOSFETs.

<table>
<thead>
<tr>
<th></th>
<th>Type A</th>
<th>Type B</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>(0.022 $\mu$m $\leq L_p \leq$ 0.034 $\mu$m)</td>
<td>(0.034 $\mu$m $&lt; L_p &lt; 0.044$ $\mu$m)</td>
</tr>
<tr>
<td>$\log_{10} \alpha_{\kappa_p}$</td>
<td>$6.733224 \times 10^2$</td>
<td>$-6.734498 \times 10^3$</td>
</tr>
<tr>
<td>$\beta_0^{\kappa_p}$</td>
<td>$3.613345$</td>
<td>$-3.710655 \times 10^2$</td>
</tr>
<tr>
<td>$\beta_1^{\kappa_p}$</td>
<td>$-9.868428 \times 10^1$</td>
<td>$9.997744 \times 10^2$</td>
</tr>
</tbody>
</table>

Table 5.4: The coefficients of the second-degree polynomial of $\log_{10} \ell_{sat}$ of the two types of the 0.1-$\mu$m pocket n-MOSFETs.

<table>
<thead>
<tr>
<th></th>
<th>Type A</th>
<th>Type B</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>(0.022 $\mu$m $\leq L_p \leq$ 0.034 $\mu$m)</td>
<td>(0.034 $\mu$m $&lt; L_p &lt; 0.044$ $\mu$m)</td>
</tr>
<tr>
<td>$\log_{10} \alpha_s$</td>
<td>$4.463457 \times 10^2$</td>
<td>$-6.359378 \times 10^3$</td>
</tr>
<tr>
<td>$\beta_0^s$</td>
<td>$2.395769$</td>
<td>$-3.504952 \times 10^2$</td>
</tr>
<tr>
<td>$\beta_1^s$</td>
<td>$-6.585583 \times 10^1$</td>
<td>$9.437922 \times 10^2$</td>
</tr>
</tbody>
</table>
Figure 5.11: $\log_{10} \kappa_p$ vs. $\log_{10}(L_{Lp}N_{p}N_{ctoxr_j})$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \kappa_p$ of (a) Type A ($0.022\mu m < L_p \leq 0.034\mu m$) and (b) Type B ($0.034\mu m < L_p \leq 0.044\mu m$) of the 0.1-μm pocket n-MOSFETs.

Figure 5.12: $\log_{10} \ell_{sat}$ vs. $\log_{10}(L_{Lp}N_{p}N_{ctoxr_j})$ as results of the second-degree polynomial fits to single-device values of $\log_{10} \ell_{sat}$ of (a) Type A ($0.022\mu m \leq L_p \leq 0.034\mu m$) and (b) Type B ($0.034\mu m < L_p \leq 0.044\mu m$) of the 0.1-μm pocket n-MOSFETs.
The complete model equations, including the z-formulas for $\kappa_p$ and $\ell_{sat}$ given in Equations 5.52 - 5.55, and the algorithm described in Section 5.3 can generate $I_D - V_{DS}$ curves and $I_{on}$ of the two-dimensional simulations for the 0.1-$\mu$m pocket n-MOSFETs fast and correctly within an acceptable error tolerance. Note that the global values of the model parameters $\kappa_p$ and $\ell_{sat}$ are calculated using the first equations in Equations 5.54 and 5.55 after Equations 5.52 and 5.53 are evaluated. Figure 5.13 shows the $I_D - V_{DS}$ curves and $I_{on}$ as the complete model results of the 0.1-$\mu$m pocket n-MOSFETs located at the upper design-space boundary.

The rms errors of the drain voltages of the complete model $I_D - V_{DS}$ curves to those of the Medici $I_D - V_{DS}$ curves at $V_{GS} = 1$ V of the twelve 0.1-$\mu$m pocket n-MOSFETs located at the upper design-space boundary are shown in Figure 5.14 where the rms error calculated using the global values of the model parameters $\kappa_p$ and $\ell_{sat}$ is compared with that calculated using the single-device values of the parameters.

5.4.2 Power-Law Formulation

We consider the power-law formulation to formulate the second analytical and empirical formulas for $\kappa_p$ and $\ell_{sat}$. The following are the power-law formulas for $\kappa_p$ and $\ell_{sat}$ found by fitting the formulas to the single-device values of the model parameters in Table 5.2.

$$\kappa_p = 0.432(L_p/L_{p0})^{-0.241075}(N_p/N_{p0})^{-0.924385}(N_c/N_{c0})^{-0.110454} \quad (5.56)$$
Figure 5.13: $I_D - V_{DS}$ characteristics at $V_{GS} = 1\, V$ generated by the complete model equations, including the $z$-formulas for $\kappa_p$, $\ell_{sat}$ and $K_{sat} = 2.3$, and compared to Medici $I_D - V_{DS}$ for the 0.1-$\mu m$ pocket n-MOSFETs with (a) $L_p = 0.022\, \mu m$, $N_p = 2.995 \times 10^{18}\,\text{cm}^{-3}$, $N_c = 2.375 \times 10^{17}\,\text{cm}^{-3}$, (b) $L_p = 0.028\,\mu m$, $N_p = 2.276 \times 10^{18}\,\text{cm}^{-3}$, $N_c = 1.925 \times 10^{17}\,\text{cm}^{-3}$, (c) $L_p = 0.038\,\mu m$, $N_p = 1.580 \times 10^{18}\,\text{cm}^{-3}$, $N_c = 1.875 \times 10^{17}\,\text{cm}^{-3}$, and (d) $L_p = 0.044\,\mu m$, $N_p = 1.307 \times 10^{18}\,\text{cm}^{-3}$, $N_c = 2.138 \times 10^{17}\,\text{cm}^{-3}$. The 0.1-$\mu m$ pocket n-MOSFETs chosen here are the same ones used in Figure 5.8. Global values of $\kappa_p$ and $\ell_{sat}$ were used for the model.
Figure 5.14: Rms errors of drain voltages of model $I_D - V_{DS}$ curves to those of Medici $I_D - V_{DS}$ curves at $V_{GS} = 1\,V$ of the 0.1-$\mu m$ pocket n-MOSFETs located at the upper design-space boundary. Either the global or the single-device values of $\kappa_p$ and $\ell_{sat}$ were used to obtain the model $I_D - V_{DS}$ curves. The $z$- and power-law formulas were used to calculate the global parameter values. The power-law formulas for $\kappa_p$ and $\ell_{sat}$ are provided in the subsequent subsection. $K_{sat} = 2.3$.

$$\ell_{sat} = 3.09 \times 10^{-3}(L_p/L_{p0})(N_p/N_{p0})^{0.2}(N_c/N_{c0})^{-0.2}$$

where the normalization parameters are $L_{p0} = 10^{-4}\, cm$, $N_{p0} = 10^{18}\, cm^{-3}$ and $N_{c0} = 10^{17}\, cm^{-3}$. The rms errors of the global values of the $z$- and power-law formulas for $\kappa_p$ and $\ell_{sat}$ to the single-device values for the twelve 0.1-$\mu m$ pocket n-MOSFETs located at the upper design-space boundary are tabulated in Table 5.5. The power-law formulation has better accuracy than the $z$-formulation in these cases. However, the accuracy of the $z$-formulation might be improved by increasing
Table 5.5: Rms errors of the global values to the single-device values of $\kappa_p$ and $\ell_{sat}$ for the twelve 0.1-µm pocket n-MOSFETs located at the upper design-space boundary.

<table>
<thead>
<tr>
<th>Formula</th>
<th>$\kappa_p$</th>
<th>$\ell_{sat}$</th>
</tr>
</thead>
<tbody>
<tr>
<td>Z-formula</td>
<td>1.37 %</td>
<td>1.38 %</td>
</tr>
<tr>
<td>Power-law formula</td>
<td>0.29 %</td>
<td>0.76 %</td>
</tr>
</tbody>
</table>

the degree of the polynomial of $\log_{10} z$ from 2 to a higher number. Covering all the values of a model parameter for both Types A and B with one model formula, the power-law formulation has an advantage over the $z$-formulation. Figure 5.15 shows comparisons of the global values of the $z$- and the power-law formulas for $\ell_{sat}$ with the single-device values for the 0.1-µm pocket n-MOSFETs.

5.5 Model Result and Prediction of Upper Design-Space Boundary for $I_{on}$

Solving the above-threshold current model equations, i.e., $V_{OS}^{\text{(Model)}} - 1 \ V = 0$ numerically for $N_p$ for the given values of $I_{on} = 0.2 \ mA/\mu m$ and $N_c$ vs. $L_p$ shown in Figure 5.1 (b), we can find the upper design-space boundary for $I_{on} = 0.2 \ mA/\mu m$ in the 0.1-µm pocket n-MOSFETs. Figure 5.16 shows the upper design-space boundary as a model result compared to Medici data. Note that if there exist more than one solution, i.e., $N_p$ for the given $N_c$, $L_p$ and $I_{on}$ in solving the model equations, a solution as the closest to the Medici datum as possible is chosen.

Figure 5.17 (a) shows the two-dimensional design-space boundary $N_p$ vs. $L_p$ with $N_c$ as a parameter for $I_{on} = 0.2 \ mA/\mu m$ in 0.1-µm pocket n-MOSFETs, predicted
Figure 5.15: Comparisons of global values with single-device values of the parameter \( \ell_{\text{sat}} \) of 0.1-\( \mu \)m pocket n-MOSFETs. (a) \( \ell_{\text{sat}} \) vs. \( L_p \), (b) \( \ell_{\text{sat}} \) vs. \( N_p/N_{p0} \), (c) \( \ell_{\text{sat}} \) vs. \( N_c/N_{c0} \) and (d) \( \ell_{\text{sat}} \) vs. \( z/z_0 \). The global parameter values were calculated using power-law and \( Z \)-formulas. \( N_{p0} = 10^{18}\) cm\(^{-3}\), \( N_{c0} = 10^{17}\) cm\(^{-3}\) and \( z_0 = 10^{13}\) cm\(^{-2}\).
Figure 5.16: Model result of $N_p$ vs. $L_p$ at the upper design-space boundary for $I_{on} = 0.2 \, mA/\mu m$ in the 0.1-\mu m pocket n-MOSFETs. The values of $N_c$ are shown in Figure 5.1 (b).

using the model equations for the above-threshold current, including the analytical formulas for $\kappa_p$, $\ell_{sat}$ and $K_{sat} = 2.3$ for the Type B devices. The upper design-space boundary in Figure 5.16 is also shown in Figure 5.17 (a). The upper design-space boundary for $I_{on} = 0.2 \, mA/\mu m$ in Figure 5.16 varies with both $L_p$ and $N_c$, while each of the design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ in Figure 5.17 (a) varies only with $L_p$ for a fixed $N_c$. Figure 5.17 (b) shows the three-dimensional design-space boundary for $I_{on} = 0.2 \, mA/\mu m$ in the 0.1-\mu m pocket devices in a space whose axes are the design variables $N_p$, $L_p$ and $N_c$ of the pocket devices, and the
Figure 5.17: Model predictions of design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ in 0.1-\(\mu m\) pocket n-MOSFETs. (a) Two-dimensional design-space boundary $N_p$ vs. $L_p$ with $N_c$ as a parameter, and (b) three-dimensional design-space boundary. The solid line in Figure (a) is the upper design-space boundary for $I_{on} = 0.2 \, mA/\mu m$ in Figure 5.16.
plot shows three different contours of equal $N_p$, $N_c$ and $L_p$. From the predictions, $N_p$ has to be decreased as $L_p$ is increased at a fixed $N_c$ or as $N_c$ is increased at a fixed $L_p$ in order to satisfy the given specification of $I_{on} = 0.2 \text{ mA/um}$. Figure 5.18 shows design-space boundaries for the different boundary specifications of $I_{on} = 0.2 - 0.6 \text{ mA/um}$ in the 0.1-μm pocket n-MOSFET with a fixed value of $N_c = 2 \times 10^{17} \text{ cm}^{-3}$. predicted using the model equations for the above-threshold current.

![Figure 5.18: Model predictions of design-space boundaries for different specifications of $I_{on}$ in the 0.1-μm pocket n-MOSFET with $N_c = 2 \times 10^{17} \text{ cm}^{-3}$.](image-url)
5.6 Summary

The characteristics of 0.1-μm pocket n-MOSFETs biased to strong inversion were studied using two-dimensional device simulation. The high electric field at the boundary between the pocket and center regions near source can improve switching performance of the deep-submicron pocket n-MOSFETs. The electron effective velocity $v_{ef}$ at $V_{GS} = 1\,V$ and $V_{DS} = 1\,V$ averaging out velocities of all electrons participating in current flow in the deep-submicron pocket n-MOSFET was evaluated using Medici $I_D - V_{DS}$ characteristics and simple formulas. The upper design-space boundary for $I_{on} = 0.2\,mA/\mu m$ in the 0.1-μm pocket n-MOSFETs was uniquely determined using the model C that consists of the drift-diffusion current transport equation, Texas transverse-field-dependent and Caughey-Thomas longitudinal-field-dependent mobility models. Using the model C as well as the model B in Appendix A, all the values of $v_{ef}$ of the 0.1-μm pocket devices at $V_{GS} = 1\,V$ and $V_{DS} = 1\,V$ are lower than $v_{sat}$. Therefore, the model for the on-state current in the 0.1-μm pocket n-MOSFETs based on the model C does not have to include any model for velocity overshoot. A computer algorithm for finding automatically and accurately the unique upper design-space boundary for $I_{on}$ was developed and implemented in the programming language of the two-dimensional device simulator Medici.
The model for the above-threshold current in the 0.1-μm pocket n-MOSFETs based on solutions of the 1-D Poisson's and the drift-diffusion current transport equations, the quasi-two-dimensional velocity saturation model, the lateral-field-dependent, the transverse-field-dependent mobility models and analytical formulas for the model parameters $\kappa_p$, $\ell_{sat}$ and $K_{sat} = 2.3$, and the algorithm based on the model were developed to generate $I_D - V_{DS}$ curves, $I_{on}$ and the upper design-space boundaries for $I_{on}$ in the pocket n-MOSFETs. The models can provide explicit relations between process, device and model parameters of the deep-submicron pocket n-MOSFETs, and a solution to the problem of time-consuming tasks of two-dimensional simulation. The model for the above-threshold current developed generates $I_D - V_{DS}$ curves of the 0.1-μm pocket n-MOSFETs efficiently and correctly.

In Chapter 6, unique viable design spaces locating deep-submicron pocket n-MOSFETs that meet device design specifications of $I_{off}$, $\frac{\Delta I_{off}/I_{off}}{\Delta L/L}$ and $I_{on}$ are constructed for 0.1, 0.15 and 0.2 μm channel-length pocket n-MOSFETs, using the algorithms developed for finding the unique upper and lower design-space boundaries. The 0.1-μm pocket n-MOSFETs located within the viable design space are compared with 0.1-μm conventional bulk n-MOSFETs selected to meet the same specifications.
CHAPTER 6

VIABLE DESIGN SPACE AND ADVANTAGES OF
0.1-\(\mu\)m POCKET N-MOSFETS

Well-designed deep-submicron pocket n-MOSFETs satisfying the device design specifications of \(I\text{off} \leq 1 \text{nA/}\mu\text{m} \), \(\frac{\Delta I_{\text{on}}/I_{\text{off}}}{\Delta L/L} \leq 40\%/10\%\), \(I_{\text{on}} \geq 0.2 \text{mA/}\mu\text{m}\) and \(V_{\text{DD}} = 1 \text{V}\) are located within a unique viable design space. The short-channel immunity, performance and design trade-off of the 0.1-\(\mu\)m pocket n-MOSFETs located within the viable design space are discussed and evaluated in this chapter. The 0.1-\(\mu\)m pocket n-MOSFETs selected from the viable design space are compared to 0.1-\(\mu\)m conventional bulk n-MOSFETs selected to meet the same specifications and prove to be effective in controlling short-channel effects as well as in improving switching performance. Section 6.1 presents the unique viable design spaces constructed for 0.1, 0.15 and 0.2 \(\mu\)m channel-length pocket n-MOSFETs, and dependencies of the off-state current \((I_{\text{off}})\), the sensitivity of off-state current to channel length \((\frac{\Delta I_{\text{off}}/I_{\text{off}}}{\Delta L/L})\), the threshold voltage \((V_{\text{T}})\), the on-state current \((I_{\text{on}})\), the transconductance \((g_m)\) and the electron effective velocity \((v_{\text{eff}})\) on the pocket length \((L_p)\) of the 0.1-\(\mu\)m pocket n-MOSFETs located at the upper and the lower design-space boundaries. Section 6.2 provides comparisons between the 0.1-\(\mu\)m
pocket n-MOSFETs located within the viable design space and the 0.1-\( \mu m \) conventional bulk n-MOSFETs regarding \( I_{\text{off}} \), \( \frac{\Delta I_{\text{off}}/I_{\text{off}}}{\Delta L/L} \), \( I_{\text{on}} \), \( g_m \) and \( v_{\text{eff}} \), and advantages of the pocket n-MOSFETs over the conventional bulk n-MOSFETs. Finally, this chapter is summarized and conclusions are drawn in Section 6.3.

6.1 Viable Design Space for Deep-Submicron Pocket n-MOSFETs

A design space for the pocket n-MOSFETs is a space whose axes are the design variables for the device: \( N_p \), \( L_p \) and \( N_c \) for fixed values of \( L \), \( t_{\text{ox}} \) and \( r_j \). The viable design space or viable window is a portion of the design space locating pocket n-MOSFETs that meet the specifications of \( I_{\text{off}} \leq 1 \) nA/\( \mu m \), \( S = \frac{\Delta I_{\text{off}}/I_{\text{off}}}{\Delta L/L} \leq 40%/10\% \), \( I_{\text{on}} \geq 0.2 \) mA/\( \mu m \) and 1 V power-supply voltage. The specifications of \( I_{\text{off}} \) and \( I_{\text{on}} \) are requirements of the digital memory and logic technology. The specification of the channel-length sensitivity of \( I_{\text{off}} \) enables uniformity of manufactured deep-submicron pocket MOSFETs. The viable design space for 0.1-\( \mu m \) pocket n-MOSFETs with fixed \( t_{\text{ox}} = 4 \) nm and \( r_j = 0.06 \) \( \mu m \) is constructed using the upper design-space boundary for \( I_{\text{on}} = 0.2 \) mA/\( \mu m \), and the lower design-space boundary for \( \frac{\Delta I_{\text{off}}/I_{\text{off}}}{\Delta L/L} = 40%/10\% \) and \( I_{\text{off}} = 0.714 \) nA/\( \mu m \), which were already found in Chapters 3 and 5. Figure 6.1 (a) shows the viable design space for the devices. The values of the doping density of the center region for all the pocket n-MOSFETs in the viable design space are shown in Figure 6.1 (b), which
Figure 6.1: (a) The zone pointed by an arrow is the viable design space $N_p$ vs. $L_p$ for 0.1 $\mu$m channel-length pocket n-MOSFETs. The devices located within the unique viable design space meet all the design specifications of $I_{on} \geq 0.2 \, mA/\mu m$, $I_{off} \leq 1 \, nA/\mu m$ and $S = \frac{\Delta I_{off}/I_{off}}{\Delta L/L} \leq 40%/10\%$. (b) $N_c$ vs. $L_p$ for all the devices located within the viable design space. This figure (b) is a repeat of Figure 3.7 (b) for the pocket devices satisfying both the sensitivity and the off-current specifications.
is the same as Figure 3.7 (b). The viable design space for the 0.1-\(\mu m\) pocket n-MOSFETs is unique because the lower and the upper design-space boundaries were determined uniquely. Both the off-state current and the sensitivity of off-state current to channel length decrease as \(N_p\) is increased in the viable design space from the lower design-space boundary for \(I_{off} = 0.714 \, nA/\mu m\) and \(\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40\%/10\%\) towards the upper design-space boundary for \(I_{on} = 0.2 \, mA/\mu m\). The on-state current also decreases as \(N_p\) is increased from the lower design-space boundary to the upper one.

The unique design-space boundaries for 0.15 and 0.2 \(\mu m\) channel-length pocket n-MOSFETs with fixed \(t_{ox} = 4 \, nm\) and \(r_j = 0.06 \, \mu m\) have been found automatically and accurately using the computer algorithms for Medici simulations with adjustments of the initial guesses of doping density values, described in Chapters 3 and 5. Figures 6.2 and 6.3 show the viable design spaces constructed from the design-space boundaries for the 0.15- and 0.2-\(\mu m\) pocket n-MOSFETs, respectively. These viable design spaces are wider than that for the 0.1-\(\mu m\) pocket n-MOSFET in Figure 6.1. The plots of \(N_c\) vs. \(L_p\) in Figures 6.1 – 6.3 show that the doping density at the center region becomes less sensitive to the pocket length as the channel length is increased, and would be constant at a long channel.

Figures 6.4 and 6.5 show potential contours and depletion-region edges for \(V_{GS} = 0 \, V\) and \(V_{DS} = 1 \, V\) inside a few 0.1-\(\mu m\) pocket n-MOSFETs located
Figure 6.2: (a) The viable design space for 0.15 $\mu$m channel-length pocket n-MOSFET. The devices within the viable design space meet all the design specifications of $I_{on} \geq 0.2$ mA/$\mu$m, $I_{off} \leq 1$ nA/$\mu$m and $S = \frac{\Delta I_{off}/I_{off}}{\Delta L/L} \leq 40\%/10\%$. (b) $N_c$ vs. $L_p$ for all the devices within the viable design space.

Figure 6.3: (a) The viable design space for 0.2 $\mu$m channel-length pocket n-MOSFET. The devices within the viable design space meet all the design specifications of $I_{on} \geq 0.2$ mA/$\mu$m, $I_{off} \leq 1$ nA/$\mu$m and $S = \frac{\Delta I_{off}/I_{off}}{\Delta L/L} \leq 40\%/10\%$. (b) $N_c$ vs. $L_p$ for all the devices within the viable design space.
Figure 6.4: Potential contours for $V_{GS} = 0\, \text{V}$ and $V_{DS} = 1\, \text{V}$ inside 0.1-$\mu\text{m}$ pocket n-MOSFETs located at (a), (f), (c) and (d) in Figure 6.1. Dashed lines are depletion-region edges.
Figure 6.5: Potential contours for $V_{DS} = 0$ V and $V_{DS} = 1$ V inside 0.1-$\mu$m pocket n-MOSFETs located at (b) and (g) in Figure 6.1. Dashed lines are depletion-region edges.

within the viable design space in Figure 6.1. The depletion-region edges and potential contours are close to gate channel, source and drain regions, and thus a saddle point [28], i.e., a potential barrier to carrier injection does not exist in the bulk. Therefore, the devices are not in bulk punchthrough. The gates of the 0.1-$\mu$m pocket n-MOSFETs are controlling depletion charges or potentials within the channel region even though the gate has nearly lost control of the depletion edge at the center region of the device (a) or (f). Short-channel effects such as bulk punchthrough or sub-surface drain-induced barrier lowering (DIBL) are controlled by preventing strong-field lines at the drain junction from penetrating into the source. As explained in Chapter 3, the pocket n-MOSFETs (a) and (f) are Type
A devices while the pocket n-MOSFETs (c) and (d) are Type B devices. Comparing the devices located at the upper design-space boundary to those located at the lower one for fixed $L_p$ and $N_c$, the pocket n-MOSFETs located at the upper area of the viable design space have better short-channel immunity than those located at the lower area because, for example, the pocket device (a) has smaller depletion-region depth than the pocket device (f). Similarly, the device (c) has smaller depletion-region depth than the device (d).

Figures 6.6 and 6.7 show comparisons of the 0.1-μm pocket n-MOSFETs located at the upper and lower design-space boundaries shown in Figure 6.1 regarding the off-state current, the percent change of off-state current due to 10% reduction of channel length from $L = 0.1 \mu m$, i.e., the sensitivity of off-state current to channel length, and the threshold voltage. The results show that the 0.1-μm pocket devices located at the upper area within the viable design space have better short-channel immunity and lower standby power consumption than the pocket devices located at the lower area within the viable design space because the increased pocket doping at the upper area is more effective in preventing high-field lines at the drain junction of the pocket n-MOSFET from penetrating into source, and making depletion-region depths smaller. Therefore, the pocket n-MOSFETs for low-power applications are located in the upper area of the viable design space. The results in Figure 6.6 show that the 0.1-μm pocket n-MOSFET with about $L_p = 0.034 \mu m$ can be the best candidate for the low-power applications. At the
Figure 6.6: (a) Off-state current and (b) channel-length sensitivity of off-state current variations with pocket length for $V_{GS} = 0$ V and $V_{DS} = 1$ V of 0.1-$\mu$m pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2$ mA/$\mu$m and $S = 40\%/10\%$ ($I_{off} = 0.714$ nA/$\mu$m), respectively.
Figure 6.7: Threshold voltage variation with pocket length for $V_{DS} = 1 \text{ V}$ of 0.1-$\mu$m pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \text{ mA/\mu m}$ and $S = 40\%/10\%$ ($I_{off} = 0.714 \text{ nA/\mu m}$), respectively.

minimum in the sensitivity of the pocket n-MOSFETs located at the upper design-space boundary in Figure 6.6 (b), the device is on the borderline of the two types of the pocket devices, i.e., Type A and Type B. Since for shorter $L_p$, $N_c$ rises and for longer $L_p$, $N_c$ rises in Figure 6.1 (b), to first order there is no change in $N_c$ with $L_p$ at the borderline, leading to the smallest sensitivity of off-state current to channel length. $N_c$ has a direct effect on the sensitivity of off-state current to channel length because the sensitivity was evaluated by reducing the length of the center region. Note that the 0.1-$\mu$m pocket devices located at the lower design-space boundary have the same $I_{off} = 0.714 \text{ nA/\mu m}$ and $S = \frac{\Delta I_{off}/L}{\Delta L/L} = 40\%/10\%$ because the values of $N_c$ and $N_p$ for each of the given $L_p$ were found in Chapter 3.
for the pocket devices meeting the boundary specification of the channel-length sensitivity of off-state current, i.e., $\Delta I_{off}/I_{off}(L) = 40\%$, $I_{off}(L) = 0.714 \text{ nA}/\mu\text{m}$ and $I_{off}(0.9L) = 1 \text{ nA}/\mu\text{m}$ (Figures 3.5 and 3.7). The low sensitivity of $I_{off}$ to $L$ of the 0.1-$\mu$m pocket n-MOSFETs enables the channel length to be scaled down to the sub-100 nm regime. (a), (b), (c), (d), (f) and (g) in Figure 6.6 (b) mean the values of the sensitivity of $I_{off}$ to $L$ of the 0.1-$\mu$m pocket devices (a), (b), (c), (d), (f) and (g) located within the viable design space in Figure 6.1 (a), respectively. Note that all the pocket devices located within the viable design space have their own identifications ($L_p, N_p, N_c$) and are directly mapped into the plot in Figure 6.6 (b) to assign the sensitivity to each of the pocket devices. In Figure 6.7 the strong and short pocket implant (Type A device) shows a higher threshold voltage than the weak and long pocket implant (Type B device). The observation from the simulation result agrees with the experimental one in [5]. The fact implies that the strong and short pocket implant can be more effective in suppressing $V_T$ roll-off of sub-100 nm pocket n-MOSFETs than the weak and long one. Note that $V_T$ is defined as the intercept with the $V_{GS}$ axis from the point of a maximum slope on the $I_D - V_{GS}$ curve. Figure 6.8 shows the minimum $\psi_{s(min)}$ of the surface potential distribution between source and drain, and the location $x_m$, measured from the source, where the surface potential minimum exists in the 0.1-$\mu$m pocket n-MOSFETs located within the viable design space. As discussed in Chapter 4, $\psi_{s(min)}$ directly determines the off-state current in the 0.1-$\mu$m pocket n-MOSFETs.
Figure 6.8: (a) Minima of surface potential distributions, i.e., the highest potential barriers to electron injection of the 0.1-μm pocket n-MOSFETs located at lower and upper design-space boundaries for $S = 40%/10\%$ ($I_{off} = 0.714 \text{ nA/μm}$) and $I_{on} = 0.2 \text{ mA/μm}$, respectively, and (b) locations at which the highest potential barriers exist. $V_{GS} = 0 \text{ V}$ and $V_{DS} = 1 \text{ V}$. 
The results shown in Figures 6.6 (a) and 6.8 (a) confirm that $I_{off}$ in the pocket devices located within the viable design space depends strongly on $\psi_{s(min)}$. Since $\psi_{s(min)}$ of all the pocket n-MOSFETs is located in the source-end pocket region (Figure 6.8 (b)), short-channel effects are controlled at the pocket region.

Figures 6.9 and 6.10 show the comparisons of the 0.1-\(\mu\)m pocket n-MOSFETs located at the upper and lower design-space boundaries regarding the on-state current, the MOSFET transconductance and the electron effective velocity, which directly affect switching performance. All the parameter values of $I_{on}$, $g_m$ and $v_{eff}$ for a given $L_p$ of the 0.1-\(\mu\)m pocket n-MOSFETs located at the upper and lower design-space boundaries in Figure 6.1 are plotted in Figure 6.9.

![Graph of on-state current variation with pocket length](image)

Figure 6.9: On-state current variation with pocket length for $V_{GS} = 1\ V$ and $V_{DS} = 1\ V$ of 0.1-\(\mu\)m pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2\ mA/\mu m$ and $S = 40%/10\%$ ($I_{off} = 0.714\ nA/\mu m$), respectively.
Figure 6.10: (a) Transconductance and (b) electron effective velocity variations with pocket length for $V_{GS} = 1 \, V$ and $V_{DS} = 1 \, V$ of 0.1-$\mu m$ pocket n-MOSFETs located at the upper and lower design-space boundaries for $I_{on} = 0.2 \, mA/\mu m$ and $S = 40\%/10\% \ (I_{off} = 0.714 \, nA/\mu m)$, respectively.
lower design-space boundary are greater than those of the devices located at the upper one because the electron mobility increases as $N_p$ is decreased. Therefore, the pocket n-MOSFETs for high-performance applications are located in the lower area of the viable design space. Figure 6.10 (b) shows that neither of the 0.1-$\mu$m pocket n-MOSFETs located within the viable design space is in effective velocity overshoot. Thus, $v_{eff} < v_{sat} = 9.2 \times 10^6$ cm/s.

Considering the 0.1-$\mu$m pocket n-MOSFETs with identical $L_p$ and $N_c$ located within the viable design space in Figure 6.1 (a), Figures 6.6 - 6.10 show that the increasing of $N_p$ for the fixed $L_p$ and $N_c$ from the lower design-space boundary to the upper one contributes to improvement in short-channel effects, but leads to performance degradation. This is a trade-off of deep-submicron pocket MOSFET design. In order to improve performance of the deep-submicron pocket n-MOSFETs, the LDD (Lightly Doped Drain) structure or an oxide thinner yet limited to about 2.3 nm by gate-to-channel tunneling leakage needs to be used.

6.2 Advantages of 0.1-$\mu$m Pocket n-MOSFETs over 0.1-$\mu$m Conventional Bulk n-MOSFETs

One of the most important roles of the pocket implant of the deep-submicron pocket MOSFET is to prevent large DIBL as well as bulk punchthrough, which gives rise to a significant increase in off-state current and standby power dissipation. Figure 6.11 shows the variations of the off-state current and the threshold
Figure 6.11: (a) $I_{off}$ vs. $L$ for $V_{GS} = 0$ V and $V_{DS} = 1$ V and (b) $V_T$ vs. $L$ for $V_{DS} = 1$ V of typical pocket n-MOSFETs with $N_c = 2.175 \times 10^{17}$ cm$^{-3}$, $N_p = 1.906 \times 10^{18}$ cm$^{-3}$ and $L_p = 0.024$ $\mu$m, and of conventional bulk n-MOSFETs with $N_A = 2.175 \times 10^{17}$ cm$^{-3}$. 
voltage with channel length of typical pocket n-MOSFETs with fixed values of the pocket length, the pocket and the center doping densities, and of conventional bulk n-MOSFETs with the doping density equal to the center doping density of the pocket devices. The pocket n-MOSFET with \( L = 0.1 \mu m \) in Figure 6.11 is exactly the same as that located at \( L_p = 0.024 \mu m \) of the lower design-space boundary for \( S = 40\%/10\% \) and \( I_{off} = 0.714 \, nA/\mu m \) in Figure 6.1 (a). The results in Figure 6.11 show that heavily doped pocket implants around drain and source of the pocket n-MOSFET scaled down into the channel length sub-100 nm range are very effective in suppressing short-channel effects particularly in preventing bulk punchthrough, which causes the significant increases in the off-state leakage and the threshold voltage roll-off of the sub-100 nm channel-length conventional bulk devices. The highest potential barrier inside the source-end pocket n-MOSFET formed by the heavily doped pocket implant is so effective as to prevent majority carriers, i.e., electrons in source from being injected into the bulk of the pocket device. Humps like the one in Figure 6.11 (a) have been seen in pocket devices experimentally [4], [9], [88], [89] and called the "reverse short-channel effect." However the mechanisms behind these observed effects, which is usually explained in terms of doping distribution, do not occur in the devices we have simulated. This dissertation does not include more detailed investigation of the reverse short-channel effect of the deep-submicron pocket n-MOSFETs.
The remaining part of this section will be devoted to comparisons of 0.1-\(\mu m\) pocket and conventional bulk n-MOSFETs in short-channel effects and performance. There are three design variables, \(N_p\), \(L_p\) and \(N_c\), for pocket devices with fixed \(L = 0.1 \mu m\), \(t_{ox} = 4 \ nm\) and \(r_j = 0.06 \mu m\) while there is one design variable, \(N_A\), for conventional bulk devices with fixed \(L = 0.1 \mu m\), \(t_{ox} = 4 \ nm\) and \(r_j = 0.06 \mu m\).

The advantage of the 0.1-\(\mu m\) pocket n-MOSFETs over the 0.1-\(\mu m\) conventional bulk n-MOSFETs in performance was already explored in Section 5.1 of Chapter 5. The 0.1-\(\mu m\) pocket devices located at the upper design-space boundary for \(I_{on} = 0.2 \ mA/\mu m\) and the 0.1-\(\mu m\) conventional bulk device with \(N_A = 1.098 \times 10^{18} \ cm^{-3}\) have the same \(I_{on} = 0.2 \ mA/\mu m\), but the MOSFET transconductances and the electron effective velocities of the pocket n-MOSFETs are higher than those of the conventional n-MOSFET as shown in Figure 5.5. The transconductance is one of the most important MOSFET parameters used to evaluate switching performance. The time constant necessary to charge a load capacitor is inversely proportional to the MOSFET transconductance. Comparing the 0.1-\(\mu m\) pocket n-MOSFETs located at the upper design-space boundary to the 0.1-\(\mu m\) conventional bulk n-MOSFET with the same \(I_{on} = 0.2 \ mA/\mu m\) as those of the pocket devices in short-channel immunity, Figure 6.12 (a) shows that the channel-length sensitivities of off-state current of the pocket devices are much lower than that of the conventional device while Figure 6.12 (b) shows that the off-state currents of
Figure 6.12: Comparisons of the 0.1-μm pocket n-MOSFETs and a 0.1-μm conventional bulk n-MOSFET with \( N_A = 1.098 \times 10^{18} \text{ cm}^{-3} \) regarding (a) sensitivity of off-state current to channel length and (b) off-state current for \( V_{GS} = 0 \text{ V} \) and \( V_{DS} = 1 \text{ V} \). The devices have the same \( I_{on} = 0.2 \text{ mA/μm} \).
the pocket n-MOSFETs are larger than or comparable to that of the conventional n-MOSFET. Therefore, the 0.1-\(\mu\)m pocket n-MOSFETs, particularly with medium or long pocket lengths, located at the upper design-space boundary have an advantage over the 0.1-\(\mu\)m conventional bulk n-MOSFET in short-channel immunity, in addition to the advantage in switching performance.

The 0.1-\(\mu\)m pocket n-MOSFETs located at the lower design-space boundary for \(I_{\text{off}} = 0.714 \ \text{nA}/\mu\text{m}\) and \(\frac{\Delta I_{\text{off}}}{\Delta L/L} = 40%/10\%\), and a 0.1-\(\mu\)m conventional bulk n-MOSFET with \(N_A = 7.297 \times 10^{17} \ \text{cm}^{-3}\) have the same \(I_{\text{off}} = 0.714 \ \text{nA}/\mu\text{m}\). Figure 6.13 shows that the MOSFET transconductances and electron effective velocities of the 0.1-\(\mu\)m pocket devices are higher than those of the 0.1-\(\mu\)m conventional device. Figure 6.14 (a) shows that the channel-length sensitivities (\(=40%/10\%\)) of off-state current of the 0.1-\(\mu\)m pocket n-MOSFETs located at the lower design-space boundary are far lower than that of the 0.1-\(\mu\)m conventional bulk n-MOSFET with the same \(I_{\text{off}} = 0.714 \ \text{nA}/\mu\text{m}\) as the pocket n-MOSFETs’ \(I_{\text{off}}\) values. Thus, the 0.1-\(\mu\)m pocket n-MOSFETs located at the lower design-space boundary have much better short-channel immunity than the 0.1-\(\mu\)m conventional bulk n-MOSFET with the same \(I_{\text{off}}\) as those of the pocket devices. Figure 6.14 (b) shows that the on-state currents of the pocket devices are smaller or larger than that of the conventional device. Thus, performance of the 0.1-\(\mu\)m pocket devices is comparable to that of the conventional device. Increasing the doping of the conventional device to bring the high sensitivity into those of the pocket n-MOSFETs...
Figure 6.13: Comparisons of the 0.1-μm pocket n-MOSFETs and a 0.1-μm conven­tional bulk n-MOSFET with $N_A = 7.297 \times 10^{17}$ cm$^{-3}$ regarding (a) transcon­ductance and (b) electron effective velocity for $V_{GS} = 1$ V and $V_{DS} = 1$ V. The devices have the same $I_{off} = 0.714 \, nA/\mu m$. 
Figure 6.14: Comparisons of the 0.1-μm pocket n-MOSFETs and a 0.1-μm conventional bulk n-MOSFET with $N_A = 7.297 \times 10^{17} \text{ cm}^{-3}$ regarding (a) sensitivity of off-state current to channel length for $V_{GS} = 0 \text{ V}$ and $V_{DS} = 1 \text{ V}$, and (b) on-state current for $V_{GS} = 1 \text{ V}$ and $V_{DS} = 1 \text{ V}$. The devices have the same $I_{off} = 0.714 \text{ nA/μm}$. 
in Figure 6.14 (a) will make the conventional device worse than the pocket devices in terms of \( I_{on} \) for the entire range of values of \( L_p \), but will make the conventional device better than the pocket devices in terms of \( I_{off} \).

Now, a 0.1-\( \mu \)m pocket n-MOSFET with \( N_p = 1.627 \times 10^{18} \text{ cm}^{-3} \), \( L_p = 0.032 \mu \text{m} \) and \( N_c = 1.825 \times 10^{17} \text{ cm}^{-3} \) located in the middle of the viable design space is compared to a 0.1-\( \mu \)m conventional bulk n-MOSFET with \( N_A = 9.1 \times 10^{17} \text{ cm}^{-3} \) for the same \( I_{on} \), and to another 0.1-\( \mu \)m conventional bulk n-MOSFET with \( N_A = 8.925 \times 10^{17} \text{ cm}^{-3} \) for the same \( I_{off} \).

Table 6.1 shows the comparisons of

<table>
<thead>
<tr>
<th></th>
<th>( I_{on} )[mA/( \mu )m]</th>
<th>( g_m )[( \mu )S/( \mu )m]</th>
<th>( I_{off} )[pA/( \mu )m]</th>
<th>( \Delta I_{off} / I_{off} )[%]</th>
<th>( V_T )[V]</th>
</tr>
</thead>
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<td>Pocket</td>
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<td>54.5</td>
<td>19</td>
<td>0.6137</td>
</tr>
<tr>
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<td>568</td>
<td>42.3</td>
<td>611</td>
<td>0.6016</td>
</tr>
<tr>
<td>Convent 2</td>
<td>0.26</td>
<td>570</td>
<td>54.6</td>
<td>619</td>
<td>0.5940</td>
</tr>
</tbody>
</table>

Table 6.1: Comparisons of a 0.1-\( \mu \)m pocket n-MOSFET located in the middle of the viable design space in Figure 6.1 with 0.1-\( \mu \)m conventional bulk n-MOSFETs. "Pocket" is the 0.1-\( \mu \)m pocket device with \( N_p = 1.627 \times 10^{18} \text{ cm}^{-3} \), \( L_p = 0.032 \mu \text{m} \) and \( N_c = 1.825 \times 10^{17} \text{ cm}^{-3} \). "Convent 1" is the 0.1-\( \mu \)m conventional device with \( N_A = 9.1 \times 10^{17} \text{ cm}^{-3} \) and the same \( I_{on} \) as that of the pocket device. "Convent 2" is the 0.1-\( \mu \)m conventional device with \( N_A = 8.925 \times 10^{17} \text{ cm}^{-3} \) and the same \( I_{off} \) as that of the pocket device.

The 0.1-\( \mu \)m pocket n-MOSFET located in the middle of the viable design space has much better short-channel immunity and slightly better switching performance than the 0.1-\( \mu \)m conventional bulk n-MOSFET (Convent 2) with both the same
$I_{on}$ and $I_{off}$ as those of the pocket device. Figure 6.12 and Table 6.1 imply that a group of 0.1-$\mu$m pocket n-MOSFETs located in an area of the viable design space have the same $I_{on}$ and $I_{off}$ as those of 0.1-$\mu$m conventional bulk devices, but they have much lower channel-length sensitivities of $I_{off}$ than those of the conventional bulk devices.

6.3 Summary

Viable design spaces locating deep-submicron devices that meet device design specifications of $I_{off}$, $\Delta I_{off}/\Delta L/L$, $I_{on}$ and $V_{DD}$ have been constructed for 0.1, 0.15 and 0.2 $\mu$m channel-length pocket n-MOSFETs, using the computer algorithms developed for finding design-space boundaries. Dependencies of device parameters of the 0.1-$\mu$m pocket n-MOSFETs on the pocket length show that the deep-submicron pocket devices for low-power applications are located at the upper area of the viable design space, whereas the pocket devices for high-performance applications are located at the lower area of the viable design space.

Comparisons of the 0.1-$\mu$m pocket n-MOSFETs located within the viable design space with the 0.1-$\mu$m conventional bulk n-MOSFETs selected to meet the same specifications show that the deep-submicron pocket devices are effective in suppressing short-channel effects and improving switching performance.

Conclusions of this dissertation is drawn in the next chapter.
CHAPTER 7

CONCLUSIONS

Laterally non-uniformly doped deep-submicron pocket n-MOSFETs that meet the device specifications of $I_{off} \leq 1 \, nA/\mu m$, $\frac{\Delta I_{off}/I_{off}}{\Delta L/L} \leq 40%/10\%$, $I_{on} \geq 0.2 \, mA/\mu m$ and 1 V power-supply voltage have been designed for low-voltage low-power applications. Unique viable design spaces locating the deep-submicron devices that meet the specifications have been constructed for 0.1, 0.15 and 0.2 $\mu m$ channel-length pocket n-MOSFETs, using the computer algorithms developed on the basis of device physics and two-dimensional device simulation results for finding automatically and accurately the design-space boundaries for $\frac{\Delta I_{off}/I_{off}}{\Delta L/L} = 40%/10\%$ ($I_{off} = 0.714 \, nA/\mu m$) and $I_{on} = 0.2 \, mA/\mu m$. The pocket n-MOSFETs for low-power applications are located in an upper area of the viable design space while the devices for high-performance applications are located in a lower area of the viable design space. Well-designed pocket n-MOSFETs are so effective in suppressing short-channel effects as to enable MOSFET dimensions to be scaled down to deep-submicron or sub-100 nm regime, and also improve switching performance. The 0.1-$\mu m$ pocket n-MOSFETs located within the viable design space have much lower sensitivity of off-state current to channel length and higher transconductance than those of 0.1-$\mu m$ conventional bulk n-MOSFETs with either the same on-state
or off-state current as that of the 0.1-\(\mu m\) pocket devices, or with both the same on-state and off-state currents as those of the 0.1-\(\mu m\) pocket devices. Additionally, a 0.1-\(\mu m\) pocket n-MOSFET with weak and long pocket implants located within the viable design space has slightly larger on-state current, far lower channel-length sensitivity of off-state current and higher transconductance than those of a 0.1-\(\mu m\) conventional bulk n-MOSFET with the same off-state current as that of the 0.1-\(\mu m\) pocket device.

A novel and simple technique for extracting the effective transverse-field-dependent mobility from Medici using the drift-diffusion current transport equation and some Medici data has been developed for comparing the extracted mobility from two-dimensional device simulations with the universal mobility curve. Analytical models for the subthreshold and the above-threshold currents in deep-submicron pocket n-MOSFETs have been developed for the first time to predict the off-state and the on-state currents, and the design-space boundaries for \(I_{off}\) and \(I_{on}\) in the 0.1-\(\mu m\) pocket devices. The models are based on solutions of the Poisson’s and the drift-diffusion current transport equations, the quasi-two-dimensional models and the analytical formulas for model parameters as functions of structural parameters of the pocket n-MOSFETs. The model equations reduce time and cost of two-dimensional simulation and provide explicit relations between process, device and model parameters of the deep-submicron pocket n-MOSFETs.
APPENDIX A

VELOCITY OVERSHEEOT IN
ULTRASHORT-CHANNEL POCKET N-MOSFET

The velocity of electrons accelerated by high electric fields in an ultrashort-channel Si n-MOSFET can exceed the limited value of $v_{sat}$. The ultrashort-channel n-MOSFET in effective velocity overshoot [69]–[83] has local velocity overshoot over a wide portion of its channel region. Few scattering events within the n-MOSFET occur where velocity overshoot occurs [69], which is observed for a period shorter than the energy relaxation time [40], [71], [72], [86] of the order of $10^{-12}$ [23] or less seconds. Most important parameters to affect velocity overshoot are known to be longitudinal-electric-field gradient in MOSFET channel, temperature and low-field mobility [73]–[77]. Also, the electron-density gradient can cause the electron velocity in a low-field region to increase [72]. The effects of electron velocity overshoot in very short-channel n-MOSFETs are increases in transconductance and on-state current, which bring improvement in switching circuit performance by reducing the time constant for charging a load capacitor. The increase in the MOSFET transconductance due to velocity overshoot has been observed experimentally [75], [80], [81], and can be predicted by Monte Carlo [71], [73], [78], [82], [86] or hydrodynamic [70]–[72], [84], [85] simulations. The Monte Carlo simulator can be used to predict the
phenomenon of velocity overshoot the most accurately among simulators available today at the expense of computation time. The full solution of the Boltzmann equation [71] is used to accurately describe electron dynamics. However, the use of a coupled system of simplified momentum- and energy-balance equations, and the relaxation time approximation of the hydrodynamic model [72], [84] could provide satisfactory descriptions of a basic physical effect underlying velocity overshoot and electron dynamics. The hydrodynamic model consists of basic equations of Poisson’s equation, the carrier-continuity equation, the energy-balance equation and the momentum-balance equation. The following three models for electrons are proposed for Medici simulations, and their results are compared.

1. Model A:

Hydrodynamic model [40], [70]–[72], [84], [85], including energy-balance equation, and carrier temperature-based mobility model [40], [76], [79].

**Key Medici statements**

```
MATERIAL SILICON WTN0=1.6852e-13(sec) WTN1=1.0299e-13(sec)
WTN2=-5.1845e-15(sec) WTN3=0(sec) WTN4=0(sec) WTN5=0(sec)
WTNL=6.8e-13(sec) TNL=2979.8(Kelvin)
MOBILITY SILICON FLDMOB=0 BETAN=2. VSATN=9.2E6
MODELS TEMPERAT=300. FLDMOB=F TFLDMOB=F
BOLTZMAN=T E.EFFECT=F ET.MODEL=T TMPMOB=T
TMPDIFF=T EF.TMP=T TMPTAUW=T EBLT.HT=T
```
II.TEMP=F EFI.TMP=F

SYMBOLIC NEWTON CARRIERS=1 ELECTRON ELE.TEMP=T

COUP.ELE=T

METHOD ETX.TOLER=1.5e-2 CX.TOLER=1.e-2 PX.TOLER=1.e-3

N.DAMP=T (for V_{DS} > 0)

2. Model B:

Hydrodynamic model, including energy-balance equation, longitudinal-field-dependent mobility model [41]-[43] and transverse-field-dependent mobility model [44], [45] for a low-longitudinal-field mobility.

Key Medici statements

| MATERIAL SILICON WTN0=1.6852e-13(sec) WTN1=1.0299e-13(sec) |
| WTN2=-5.1845e-15(sec) WTN3=0(sec) WTN4=0(sec) WTN5=0(sec) |
| WTNL=6.8e-13(sec) TNL=2979.8(Kelvin) |

| MOBILITY SILICON FLDMOB=1 BETAN=2 VSATN=9.2E6 |
| TEMP.N.UT=1150.0 PHONN.UT=3.2E-9 SURFN.UT=6.0E14 |
| COULN.UT=1.1E21 INV.N.UT=1.0 ETAN=1 ETAP=1 |

| MODELS TEMPERAT=300. FLDMOB=T TFLDMOB=T |
| BOLTZMAN=T E.EFFECT=T ET.MODEL=T TMPMOB=F |
| TMPDIFF=T EF.TMP=F TMPTAUW=T EBLT.HT=T |

II.TEMP=F EFI.TMP=F

SYMBOLIC NEWTON CARRIERS=1 ELECTRON ELE.TEMP=T
COUP.ELE=T

METHOD ETX.TOLER=1.5e-2 CX.TOLER=1.e-2 PX.TOLER=1.e-3

N.DAMP=T (for $V_{DS} > 0$)

3. Model C:
   Drift-diffusion transport model, longitudinal-field-dependent and transverse-field-dependent mobility models.

   Key Medici statements

MOBILITY SILICON FLDMOB=1 BETAN=2 VSATN=9.2E6
   TEMPN.UT=1150.0 PHONN.UT=3.2E-9 SURFN.UT=6.0E14
   COULN.UT=1.1E21 INV.N.UT=1.0 ETAN=1 ETAP=1

MODELS TEMPERAT=300 TFLDMOB=T FLDMOB=T
   BOLTZMAN=T E.EFFECT=T TMDIFF=F

SYMBOLIC NEWTON CARRIERS=1 ELECTRON

The model A is the the hydrodynamic model with an unrealistic mobility because the mobility does not include the degradation due to the field component vertical to the Si-SiO₂ interface, while the model B is the hydrodynamic model with a realistic mobility.

Note that the default parameter values of the electron energy relaxation time formula [40] for the hydrodynamic model on the MATERIAL statement are used for Medici simulations. T and F on the Medici statements stand for logical True
and False of a logical parameter, respectively. TEMPERAT on the MODELS statement is the initial lattice temperature in Kelvins for the device structure. FLD-MOB on the MODELS statement specifies that the longitudinal-field-dependent mobility model is used. TFLDMOB specifies that the transverse-field-dependent mobility model for a low-longitudinal-field mobility is used. TMPMOB specifies that the carrier temperature-based mobility model is used. BOLTZMAN specifies that the Maxwell-Boltzmann statistics are used. E.EFFECT specifies that an effective electric field is calculated at the Si-SiO₂ interface for the use in the transverse-field-dependent mobility model. ET.MODEL specifies that a current density expression [40] is used when solving the hydrodynamic set of equations. TMPDIFF specifies that a simplified thermal diffusion term in the current density expression is used when the hydrodynamic model is used. EF.TMP specifies that the hydrodynamic set of equations is solved locally in Si material to determine the effective electric field used in the carrier temperature-based mobility model. TMP- TAUW enables an electron temperature-dependent energy relaxation time model used for the hydrodynamic model. EBLT.HT (default: True) is effective only when a lattice heating model is used. II.TEMP and EFI.TMP are effective only when an impact ionization model is used. The last three parameters are not necessary for the models and Medici simulations for this dissertation. ELE.TEMP on the SYMBOLIC statement specifies that the solution of the hydrodynamic model is initiated. COUP.ELE specifies that the fully-coupled hydrodynamic system is solved.
Thus, the electron temperature equation is fully coupled with the carrier-continuity equation and Poisson's equation.

The models A and B that include the hydrodynamic model have a problem of much longer simulation time than the model C, and produce serious convergence problems in simulation. The convergence problems can be solved by specifying larger values of the parameters relevant to error tolerance than the default values on the METHOD statement. Thus, ETX.TOLER = 1.5 \times 10^{-2} (the carrier temperature update tolerance, 1 \times 10^{-2} (default)), CX.TOLER = 1 \times 10^{-2} (the carrier concentration update tolerance, 1 \times 10^{-5} (default)), and PX.TOLER = 1 \times 10^{-3} (the potential update tolerance, 1 \times 10^{-5} (default)).

Figures A.1 and A.2 show Medici simulation results of $I_D - V_{DS}$ characteristics at $V_{GS} = 1$ V, potential, longitudinal electric field, electron density and electron velocity at the Si-SiO$_2$ interface of a 0.1-$\mu$m pocket n-MOSFET, which were obtained using the models A, B and C. The drain current, the electron velocity and the slope of the electron density of the results of the model A in comparison with the models B and C are noticeably increased. The markedly increased electron velocity around the boundary of the center and the source-end pocket regions of the model A in Figure A.2 (d) results from the rapidly changing high electric field and the steeper gradient of the electron density at the source pocket region. The carrier temperature-based mobility model, which is a function of the low-longitudinal-field mobility, the carrier and lattice temperatures and the position $(x, y)$, also plays
Figure A.1: Comparison of the three different models A, B and C with respect to $I_D - V_{DS}$ characteristics at $V_{GS} = 1$ V of the 0.1-$\mu$m pocket n-MOSFET with $N_p = 2.713 \times 10^{18}$ $cm^{-3}$, $N_c = 2.175 \times 10^{17}$ $cm^{-3}$ and $L_p = 0.024$ $\mu$m.

so important role to markedly increase the electron velocity at the channel region.

There are no great differences of the model B from the model C in the results in Figures A.1 and A.2, particularly in the $I_D - V_{DS}$ characteristics.

For the purpose of examining the dependence of the electron velocity on physical parameters for the model C, we consider the following basic equations for the electron current density at a position in the channel region.

$$ J_x(x, y) = -qnv $$

(A.1)
Figure A.2: Comparisons of the three different models A, B and C with respect to (a) potential (b) electric field (c) electron density and (d) electron velocity profiles between source and drain at the Si-SiO₂ interface for $V_{DS} = 1 \text{ V}$ and $V_{GS} = 1 \text{ V}$ of the 0.1-μm pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \text{ cm}^{-3}$, $N_c = 2.175 \times 10^{17} \text{ cm}^{-3}$, and $L_p = 0.024 \mu m$. $v_{sat} = 9.2 \times 10^6 \text{ cm/s}$. 
\[ J_x(x, y) = qn\mu E_x + qD_n \frac{\partial n}{\partial x} = -qn\mu \frac{\partial \phi}{\partial x} + q\mu \nu_{\text{in}} \frac{\partial n}{\partial x} \] (A.2)

where \( \mu(x, y) \) is the electron local-field-dependent microscopic mobility for a volume electron density \( n(x, y) \) at the position \((x, y)\) in the channel region, which is computed by the two-dimensional device simulator using both the longitudinal- and the transverse-field-dependent mobility models, and \( D_n(x, y) = \nu_{\text{in}} \mu(x, y) \) is the Einstein relation, which holds for the Maxwell-Boltzmann distribution. From Equations A.1 and A.2 we can directly derive an expression for the electron velocity \( v(x, y) \).

\[ v(x, y) = \mu \frac{\partial \phi}{\partial x} - \mu \nu_{\text{in}} \frac{1}{n} \frac{\partial n}{\partial x} = \mu \frac{\partial \phi}{\partial x} - \mu \nu_{\text{in}} \frac{\partial (\ln n)}{\partial x} \equiv v_{\text{drift}} + v_{\text{diff}} \] (A.3)

The electron velocity consists of the drift velocity that is directly dependent on the longitudinal electric field and the electron local-field-dependent microscopic mobility, and the diffusion velocity that directly depends on the gradient of electron density, the local-field-dependent microscopic mobility and the electron density. The profile of the electron velocity \( v(x, 0) \) of the model C in Figure A.2 (d) can be obtained using Equation A.3 and some Medici data. If velocity overshoot is taken account of in a one-dimensional model, an additional velocity term proposed in [76] needs to be included in Equation A.3.
Figure A.3: Comparisons of the three different models A, B and C for the identical low-field mobility model, i.e., the generalized mobility curve mobility model with respect to (a) electron velocity profile at the Si-SiO₂ interface for $V_{DS} = 1 \ V$ and $V_{GS} = 1 \ V$, and (b) $I - V$ characteristics at $V_{GS} = 1 \ V$ of the 0.1-μm pocket n-MOSFET with $N_p = 2.713 \times 10^{18} \ cm^{-3}$, $N_c = 2.175 \times 10^{17} \ cm^{-3}$ and $L_p = 0.024 \ \mu m$.

Figure A.3 shows another comparison of the models A, B and C for the same low-field mobility model, i.e., the generalized mobility curve mobility model [55], [58] that can be specified together with the carrier temperature-based mobility model for the model A. Comparing the results in Figure A.3 to those in Figures A.1 and A.2, the electron velocity and the drain current especially for the model A depend strongly on the low-field mobility. The high field and velocity around the boundary between the source-end pocket and center regions can contribute to improvement in performance of the deep-submicron pocket n-MOSFET. The model A with the generalized mobility curve mobility model can be the most accurate
one among the models, and the use of the model can make the viable design space of the ultrashort-channel pocket n-MOSFETs wider by shifting the upper design-space boundary upward. However, the hydrodynamic model for the model A is complicated, and takes much longer simulation time than the drift-diffusion transport model for the model C does. The model C was used to construct the upper design-space boundaries for pocket n-MOSFETs in Chapters 5 and 6 for the practical reasons.

We accept the following definitions of velocity overshoot.

1. \( v_{\text{eff}} > v_{\text{sat}} \) means a MOSFET is in effective velocity overshoot.
2. \( u(x,y) > v_{\text{sat}} \) means a MOSFET has local velocity overshoot at the position \((x,y)\).

Then, \( v_{\text{eff}} > v_{\text{sat}} \) is a sufficient condition for \( u(x,y) > v_{\text{sat}} \), but \( v_{\text{eff}} > v_{\text{sat}} \) is not a necessary condition for \( u(x,y) > v_{\text{sat}} \) because the effective velocity overshoot does not occur if the local velocity overshoot occurs in a small portion of the channel. Also, \( u(x,y) > v_{\text{sat}} \) is a necessary condition for \( v_{\text{eff}} > v_{\text{sat}} \). To confirm that effective velocity overshoot is a phenomenon in an ultrashort-channel n-MOSFET, the 0.1-\( \mu \)m pocket n-MOSFET in effective velocity overshoot is compared to a 1-\( \mu \)m pocket n-MOSFET with the same parameter values except the channel length, using the model A. Figure A.4 shows comparisons between the 0.1-\( \mu \)m and 1-\( \mu \)m pocket n-MOSFETs regarding the \( I_D - V_{DS} \) characteristic, the longitudinal electric
Figure A.4: (a) Surface longitudinal electric field and (b) surface electron velocity vs. normalized distance from source for $V_{DS} = 1 \, V$ and $V_{GS} = 1 \, V$, and (c) $I - V$ characteristics at $V_{GS} = 1 \, V$ of the pocket n-MOSFETs with $L = 0.1, 1 \, \mu m$, $N_p = 2.713 \times 10^{18} \, cm^{-3}$, $N_c = 2.175 \times 10^{17} \, cm^{-3}$ and $L_p = 0.024 \, \mu m$. 
field and the electron velocity. The magnitudes of the longitudinal electric field, the electron velocity at the Si-SiO₂ interface and the on-state current of the 0.1-μm pocket n-MOSFET are much greater than those of the 1-μm pocket n-MOSFET.

The MOSFET transconductance or the effective velocity averaging out the velocities of all electrons participating in current flow over entire channel length and depth is evaluated to judge whether or not the 0.1-μm pocket n-MOSFET are in effective velocity overshoot. Table A.1 lists the transconductance and the electron effective velocity of the 0.1- and 1-μm pocket n-MOSFETs, calculated from the Medici-simulated $I_D - V_{DS}$ characteristics for $V_{GS} = 1 \, V$ and $V_{GS} = 0.09 \, V$ obtained using the model A, Equations 5.1 and 5.2. The effective velocity, higher than the saturation velocity $v_{sat} = 9.2 \times 10^6 \, cm/s$ [45], of the 0.1-μm pocket n-MOSFET indicates that the pocket MOSFET is in effective velocity overshoot, which means that the velocity $v(x,y)$ is higher than $v_{sat}$ over a wide portion of the channel region. However, the long-channel 1-μm pocket n-MOSFET is not in effective velocity overshoot even using the model A because $v_{eff} < v_{sat}$. The results of the model B or C in Table A.2 show that although the electron effective

<table>
<thead>
<tr>
<th></th>
<th>$L = 0.1 , \mu m$</th>
<th></th>
<th>$L = 1 , \mu m$</th>
</tr>
</thead>
<tbody>
<tr>
<td>$g_m , [\mu S/\mu m]$</td>
<td>$v_{eff} , [cm/s]$</td>
<td>$g_m , [\mu S/\mu m]$</td>
<td>$v_{eff} , [cm/s]$</td>
</tr>
<tr>
<td>Model A</td>
<td>1781</td>
<td>$2.06 \times 10^7$</td>
<td>385</td>
</tr>
</tbody>
</table>
Table A.2: Transconductance $[\mu S/\mu m]$, electron effective velocity $[cm/s]$ and on-state current $[A/\mu m]$ at $V_{GS} = 1 V$ and $V_{DS} = 1 V$ of the 0.1- and 1-$\mu$m pocket n-MOSFETs with $t_{ox} = 4 nm$, $L_p = 0.024 \mu m$, $N_p = 2.713 \times 10^{18} cm^{-3}$, $N_c = 2.175 \times 10^{17} cm^{-3}$ and $r_j = 0.06 \mu m$.

<table>
<thead>
<tr>
<th></th>
<th>$L = 0.1 \mu m$</th>
<th>$L = 1 \mu m$</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>$g_m$  $v_{eff}$  $I_{on}$</td>
<td>$g_m$  $v_{eff}$  $I_{on}$</td>
</tr>
<tr>
<td>Model B</td>
<td>586  $6.79 \times 10^6$  $2.048 \times 10^{-4}$</td>
<td>165  $1.91 \times 10^6$  $6.854 \times 10^{-5}$</td>
</tr>
<tr>
<td>Model C</td>
<td>582  $6.75 \times 10^6$  $2.005 \times 10^{-4}$</td>
<td>171  $1.98 \times 10^6$  $6.806 \times 10^{-5}$</td>
</tr>
</tbody>
</table>

velocity of the 0.1-$\mu$m pocket n-MOSFET is higher than that of the 1-$\mu$m pocket n-MOSFET, neither of the effective velocities exceeds the electron saturation velocity. The percent differences in $v_{eff}$ and $I_{on}$ between the two models for the 0.1-$\mu$m pocket n-MOSFET are $[v_{eff}(\text{model B}) - v_{eff}(\text{model C})]/v_{eff}(\text{model B}) = 0.59 \%$ and $[I_{on}(\text{model B}) - I_{on}(\text{model C})]/I_{on}(\text{model B}) = 2.10 \%$, respectively. Thus, the simulation results of the models B and C are almost the same for the 0.1-$\mu$m pocket n-MOSFET. Here is a physical view to explain why the pocket n-MOSFET is not in effective velocity overshoot even though the model B is used. As $v_{eff}$ is the electron effective velocity averaging out the velocities of all electrons participating in current flow over the entire channel length and depth of the pocket n-MOSFET, let $t_{eff}$ and $\tau_{eff}$ be the electron effective transit time from the source to the drain and the electron effective energy relaxation time, respectively. When the n-MOSFET is in effective velocity overshoot, very few or a small amount of scattering events occur theoretically, and then $t_{eff}$ is shorter than or comparable to $\tau_{eff}$. Using the
effective quantities, we have a simple relation between $v_{\text{eff}}$ and $t_{\text{eff}}$. Thus,

$$v_{\text{eff}} = \frac{L}{t_{\text{eff}}} \quad (A.4)$$

where $L$ is the effective physical channel length between the source and the drain. $t_{\text{eff}}$ increases as $v_{\text{eff}}$ decreases, the scattering events increase, the electron mobility decreases or the electric field decreases. Therefore, $t_{\text{eff}}$ depends strongly on the electron mobility and the electric field within the semiconductor device.

The transverse-field-dependent mobility model includes the phonon, surface roughness and screened Coulomb scattering mechanisms. Also, the longitudinal-field-dependent mobility model has the inherent property of velocity saturation at uniform high fields. Therefore, $t_{\text{eff}}$ of the model B using both the field-dependent mobility models must be longer than $T_{\text{eff}}$ of the hydrodynamic model, and thereby the results of the model B using the hydrodynamic model are almost the same as those of the model C using the drift-diffusion model and the field-dependent mobility models. For example, $t_{\text{eff}} = \frac{0.1 \mu \text{m}}{6.79 \times 10^6 \text{ cm/s}} = 1.47 \times 10^{-12} \text{ sec}$ of the 0.1-$\mu$m pocket n-MOSFET using the model B, while $\tau_{\text{eff(max)}} = 6.8 \times 10^{-13} \text{ sec}$ [40] that is the maximum of the electron energy relaxation time for $T_n$ (electron temperature) > 2979.8°K. Thus, $t_{\text{eff}} > \tau_{\text{eff(max)}}$. Therefore, the 0.1-$\mu$m pocket n-MOSFET is not in effective velocity overshoot. However, the field-dependent mobility models can be inadequate for the modeling of currents in ultrashort-channel pocket n-MOSFETs in effective velocity overshoot.
In order to confirm further that velocity overshoot is negligible in the 0.1-\(\mu m\) pocket n-MOSFET, we consider the hydrodynamic model, the carrier temperature-based mobility model and the generalized mobility curve mobility model that is almost the same as the transverse-field-dependent mobility model at strong inversion used for the models B and C. Table A.3 shows the two-dimensional device simulation results of transconductance, electron effective velocity and on-state current of the 0.1-\(\mu m\) pocket n-MOSFET, obtained using the model A and the generalized mobility curve mobility model. The results show that \(v_{\text{eff}} < v_{\text{sat}}\), and the percent difference in \(I_{\text{on}}\) between the model A with the generalized mobility curve mobility model and the model C is \([I_{\text{on}}(\text{model A}) - I_{\text{on}}(\text{model C})]/I_{\text{on}}(\text{model A}) = 5.42\%\). Therefore, we do not have to take into account velocity overshoot effects for the modeling of the on-state current in the 0.1-\(\mu m\) pocket n-MOSFET.

<table>
<thead>
<tr>
<th>(L = 0.1\ \mu m)</th>
<th>(g_m[\mu S/\mu m])</th>
<th>(v_{\text{eff}}[\text{cm/s}])</th>
<th>(I_{\text{on}}[\text{A/\mu m}])</th>
</tr>
</thead>
<tbody>
<tr>
<td></td>
<td>786</td>
<td>(9.1 \times 10^6)</td>
<td>(2.120 \times 10^{-4})</td>
</tr>
</tbody>
</table>

Table A.3: Transconductance, electron effective velocity and on-state current at \(V_{\text{GS}} = 1\ V\) and \(V_{\text{DS}} = 1\ V\) of the 0.1-\(\mu m\) pocket n-MOSFET with \(t_{ox} = 4\ nm\), \(L_p = 0.024\ \mu m\), \(N_p = 2.713 \times 10^{18}\ cm^{-3}\), \(N_c = 2.175 \times 10^{17}\ cm^{-3}\) and \(r_j = 0.06\ \mu m\).
APPENDIX B

IMPLEMENTATION OF ALGORITHMS FOR FINDING

DESIGN-SPACE BOUNDARY FOR $I_{off}$ AND $\frac{\Delta I_{off}}{\Delta L/L}$

TITLE Find $N_c$ and $N_p$ of 0.1-$\mu$m pocket MOSFETs satisfying the sensitivity
+ specification of 40% change of $I_{off}$ ($I_{off}(0.9L)=1[nA/\mu m]$)
+ with 10% change (decrease) of $L$ (Sensitivity of $I_{off}$ to $L$) of pocket
+ MOSFETs for $L=0.1[\mu m], L_P=0.02-0.045[\mu m], V_{gs}=0[V], V_{ds}=1[V]$.
+ Find $I_{off}(L)$. $L_0=L$, $L_2=0.9*L$.
+ Error tolerances of sensitivity of $I_{off}$ to $L$, and $I_{off}(0.9L)$ from given specifications: within 0.5%.
+ Modified Arora mobility model, transverse-field-dependent mobility
+ model. lateral-field-dependent mobility model. $VSAT_N=9.2E6[c7n/5]$.
+ $L_G=L+L_D+L_S[\mu m]$, $L_D=L_S=0.1[\mu m]$, $tox=40[\AA]$, and $r_j=0.06[\mu m]$.
+ Find potential distribution between source and drain.
+ Check short-channel effects such as bulk punchthrough of the designed
+ pocket MOSFETs from the subthreshold slope in $I_D-V_{GS}$ characteristics.

COMMENT MEDICI Version 2.3.1 or Version 1999.2.0 was used for this program.

COMMENT channel length and initial guesses of $N_c$ and $N_p$
ASSIGN NAME=chan.le0 N.VALUE=0.1
ASSIGN NAME=nc.guess N.VALUE=2.E-17 print
ASSIGN NAME=np.guess N.VALUE=2.E-18 print

loop steps=14 print
assign name=number n.value=1 delta=1 print
ASSIGN NAME=Lp N.VALUE=0.020 DELTA=0.002 print
if cond=(@number=14)
ASSIGN NAME=Lp N.VALUE=0.045 print
else
if.end

$\text{length of center region for } L=L_0.$
EXTRACT NAME=Lc0 EXPRESS="@chan.le0-2.*Lp" NOW clear print
ASSIGN NAME=junc.dep N.VALUE=0.06
ASSIGN NAME=del.x N.VALUE=1.e-4
ASSIGN NAME=ls N.VALUE=0.1
ASSIGN NAME=Ld N.VALUE=0.1
ASSIGN NAME=interfa N.VALUE=@Ls+@Lp
ASSIGN NAME=Lc.dec N.VALUE=@Lc0-@chan.Ie0*0.1

ASSIGN NAME=FILE1 C.VALUE=nc_np_sens.dat.00 DELTA=1
ASSIGN NAME=FILE2 C.VALUE=phis_vds1.0.ps.00 DELTA=1
ASSIGN NAME=FILE3 C.VALUE=phis_vds1.0.dat.00 DELTA=1
ASSIGN NAME=FILE4 C.VALUE=phis_vds0.0.ps.00 DELTA=1
ASSIGN NAME=FILE5 C.VALUE=phis_vds0.0.dat.00 DELTA=1
ASSIGN NAME=FILE6 C.VALUE=ptnl_cont.vg0.vd1.ps.00 DELTA=1
ASSIGN NAME=FILE7 C.VALUE=vds0.1.log.00 DELTA=1
ASSIGN NAME=FILE8 C.VALUE=vds1.0.log.00 DELTA=1
ASSIGN NAME=FILE9 C.VALUE=ids_vgs.ps.00 DELTA=1
ASSIGN NAME=FILE10 C.VALUE=ids_vgs_log scale.ps.00 DELTA=1
ASSIGN NAME=FILE11 C.VALUE=ids_vgs_vds1.0.dat.00 DELTA=1
ASSIGN NAME=FILE12 C.VALUE=ids_vgs_vds0.1.dat.00 DELTA=1

$Assign nc.guess and np.guess to Nc and Np, respectively
+for 0.02um<Lp<0.045um.
ASSIGN NAME=nc.den N.VALUE=@nc.guess print
ASSIGN NAME=np.den N.VALUE=@np.guess print

$Initial delta Nc and delta Np
ASSIGN NAME=del.nc N.VALUE=1.e16 print
ASSIGN NAME=del.np N.VALUE=0.5e18 print

ASSIGN NAME=Ivr.pre N.VALUE=0. print
$spec of [Ioff(0.9*L)-Ioff(L)]/Ioff(L)=40%
ASSIGN NAME=Ivr.spec N.VALUE=0.4 print

ASSIGN NAME=ids.pre N.VALUE=0. print
$spec of Ioff(0.9*L)=1[nA/um]
ASSIGN NAME=ids.spec N.VALUE=1.e-9 print

ASSIGN NAME=iter N.VALUE=2
LOOP STEPS=100 print
if cond=(@iter=1 | @iter=3)
EXTRACT NAME=Lc EXPRESS="@Lc0" now clear print
$else cond=(@iter=2)
else
EXTRACT NAME=Lc EXPRESS="@Lc.dec" now clear print
if.end
ASSIGN NAME=chan.len N.VALUE=@Lc+2.*@Lp print
EXTRACT NAME=space.lp EXPRESS="20.*@Lp/(@chan.len/4.)+1." now clear
EXTRACT NAME=space.lc EXPRESS="40.*@Lc/(@chan.len/2.)+1." now clear
ASSIGN NAME=interfa2 N.VALUE=@interfa1+@Lc
ASSIGN NAME=interfa3 N.VALUE=@interfa2+@Lp
ASSIGN NAME=length N.VALUE=@interfa3+@Ld
ASSIGN NAME=x.pos0 N.VALUE=@Ls-@del.x
ASSIGN NAME=x.pos1 N.VALUE=@Ls+2.*@del.x
ASSIGN NAME=x.pos2 N.VALUE=@interfa1-2.*@del.x
ASSIGN NAME=x.pos3 N.VALUE=@interfa1+2.*@del.x
ASSIGN NAME=x.pos4 N.VALUE=@interfa2-2.*@del.x
ASSIGN NAME=x.pos5 N.VALUE=@interfa2+2.*@del.x
ASSIGN NAME=x.pos6 N.VALUE=@interfa3-2.*@del.x
ASSIGN NAME=x.pos7 N.VALUE=@interfa3+@del.x

COMMENT Specify a rectangular mesh
MESH SMOOTH.K=1
$grid section 1: source electrode, and pocket_left
X.MESH X.MIN=0.00 X.MAX=@x.pos0 N.SPACES=11 H2=1.E-3
$grid section 2: interface, source electrode, and pocket_left
+ (DEL=1e-4[μm], DEL+2*DEL)
X.MESH X.MIN=@x.pos0 X.MAX=@x.pos1 N.SPACES=3 min.spac=1.e-6

$grid section 3: pocket_left
X.MESH X.MIN=@x.pos1 X.MAX=@x.pos2 N.SPACES=@space.lp
+ H1=1.E-4 H2=1.E-4 min.spac=1.e-6

if cond=(@Lc < 1.e-8)
X.MESH X.MIN=@x.pos2 X.MAX=@interfa1 N.SPACES=2 min.spac=1.e-6
X.MESH X.MIN=@interfa1 X.MAX=@x.pos6 N.SPACES=@space.lp
+ H1=1.E-4 H2=1.E-4 min.spac=1.e-6
else
$grid section 4: interface, pocket_left, and center
+ (DEL=1e-4[μm], 2*DEL+2*DEL)
X.MESH X.MIN=@x.pos2 X.MAX=@x.pos3 N.SPACES=4 min.spac=1.e-6

$grid section 5: center
X.MESH X.MIN=@x.pos3 X.MAX=@x.pos4 N.SPACES=@space lc H1=1.E-4 + H2=1.E-4 min.spac=1.e-6
$grid section 6: interface, center, and pocket_right
+ (DEL=1e-4[\mu m], 2*DEL+2*DEL)
X.MESH X.MIN=@x.pos4 X.MAX=@x.pos5 N.SPACES=4 min.spac=1.e-6

$grid section 7: pocket_right
X.MESH X.MIN=@x.pos5 X.MAX=@x.pos6 N.SPACES=@space lp H1=1.E-4 + H2=1.E-4 min.spac=1.e-6
if.end

$grid section 8: interface, pocket_right, and drain electrode
+ (DEL=1e-4[\mu m], 2*DEL+DEL)
X.MESH X.MIN=@x.pos6 X.MAX=@x.pos7 N.SPACES=3 min.spac=1.e-6

$grid section 9: drain electrode, and pocket_right
X.MESH X.MIN=@x.pos7 X.MAX=@length N.SPACES=10 H1=1.E-3

$-0.004[\mu m] \leq y \leq 0.24[\mu m]
COMMENT Gate oxide thickness=4[nm]=40[Å]
Y.MESH N=1 L=-40.0E-4
Y.MESH N=2 L=-25.0E-4
Y.MESH N=3 L=-6.5E-4
Y.MESH N=4 L=-1.5E-4
Y.MESH N=5 L=-0.5E-4 min.spac=1.e-6
Y.MESH N=6 L=0.0 min.spac=1.e-6
$for channel region
Y.MESH N=7 L=0.5E-4 min.spac=1.e-6
Y.MESH N=8 L=1.0E-4 min.spac=1.e-6
Y.MESH N=9 L=1.5E-4 min.spac=1.e-6
Y.MESH Y.MIN=1.50E-4 Y.MAX=2.00E-2 N.SPACES=38 H1=1.E-4
$y \leq r_j
Y.MESH Y.MIN=2.00E-2 Y.MAX=6.00E-2 N.SPACES=16 H1=1.E-3 H2=2.E-3
Y.MESH Y.MIN=6.00E-2 Y.MAX=0.22 N.SPACES=13 H1=2.E-3 H2=5.E-3
Y.MESH Y.MIN=0.22 Y.MAX=0.24 N.SPACES=3 H1=5.E-3 H2=1.E-3

COMMENT Specify oxide and silicon regions
if cond=(@Lc>1.e-8)
REGION NAME=CENTER SILICON X.MIN=@interfa1 X.MAX=@interfa2
+ Y.MIN=0.0 Y.MAX=0.24
else
if.end
REGION NAME=POCKET SILICON X.MIN=@Ls X.MAX=@interfa1
+ Y.MIN=0.0 Y.MAX=@junc.dep
REGION NAME=POCKET SILICON X.MIN=0.0 X.MAX=@interfa1
+ Y.MIN=@junc.dep Y.MAX=0.24
REGION NAME=POCKET SILICON X.MIN=@interfa2 X.MAX=@interfa3
+ Y.MIN=0.0 Y.MAX=@junc.dep
REGION NAME=POCKET SILICON X.MIN=@interfa2 X.MAX=@length
+ Y.MIN=@junc.dep Y.MAX=0.24
REGION NAME=OXIDE OXIDE IY.MAX=6
REGION NAME=NPOLY_S POLYSILI X.MIN=0.0 X.MAX=@Ls Y.MIN=0.0
+ Y.MAX=@junc.dep
REGION NAME=NPOLY_D POLYSILI X.MIN=@interfa3 X.MAX=@length
+ Y.MIN=0.0 Y.MAX=@junc.dep

COMMENT Electrodes:
+ Gate, Source Drain contacts
ELECTR NAME=Gate TOP
ELECTR NAME=Source X.MIN=0.0 X.MAX=@Ls Y.MIN=0.0
+ Y.MAX=@junc.dep
ELECTR NAME=Drain X.MIN=@interfa3 X.MAX=@length Y.MIN=0.0
+ Y.MAX=@junc.dep
ELECTR NAME=Sub BOTTOM

COMMENT Specify impurity profiles
if cond=(@Lc>1.e-8)
PROFILE P-TYPE REGION=CENTR n.peak=@nc.den uniform
+ OUT.FILE=DOPING
PROFILE P-TYPE REGION=POCKET n.peak=@np.den uniform
else
PROFILE P-TYPE REGION=POCKET n.peak=@np.den uniform
+ OUT.FILE=DOPING
if.end

COMMENT Interface fixed charge density[cm⁻²]
INTERFACE QF=0.0

COMMENT Regrid on doping
REGRID DOPING LOG IGNORE=OXIDE RATIO=2 SMOOTH.K=1
+ IN.FILE=DOPING

COMMENT Specify contact parameters
CONTACT NAME=Gate N.POLY workfunc=4.05
CONTACT NAME=Source N.POLY workfunc=4.05
CONTACT NAME=Drain N.POLY workfunc=4.05
CONTACT NAME=Sub NEUTRAL
$CONTACT NAME=Sub P.POLY workfunc=5.174 print

COMMENT Material parameters
MATERIAL SILICON EG300=1.124 EGALPH=0 V0.BGN=0 NC300=2.8E19
+ NV300=1.04E19 EG.MODEL=1 AFFINITY=4.05 PERMITTI=11.7
MATERIAL OXIDE PERMITTI=3.9

COMMENT Specify mobility, and physical models to be used.
MOBILITY SILICON FLDMOB=1 BETAN=2. VSATN=9.2E6
+ TEMPN.UT=1150.0 PHONN.UT=3.2E-9 SURFN.UT=6.0E14
+ COULN.UT=1.1E21 INV.N.UT=1.0 ETAN=1 ETAP=1
+ MUN1.ARO=0.000000e+00 MUN2.ARO=1.01620e+03 AN.ARO=7.29358e-01
+ CN.ARO=1.08319e+17
MODELS TEMPERAT=300. TFDMOB=T FLDMOB=T BOLTZMAN=T
+ E.EFFECT=T TMPDIFF=F ARORA=T

COMMENT Symbolic factorization, solve, regrid on potential
SYMBOLIC CARRIERS=0
METHOD ICCG DAMPED
SOLVE
REGRID POTENTIA IGNORE=OXIDE RATIO=0.2 MAX.LEVE=1
+ SMOOTH.K=1 IN.FILE=DOPING

COMMENT Use Newton's solution method.
SYMBOLIC NEWTON CARRIERS=1 ELECTRONS

COMMENT Vgs=0[V],Vds=1.0[V]
if cond=(@iter=3)
SOLVE V(Gate)=0.0 V(Drain)=0.0
else
SOLVE V(Gate)=0.0 V(Drain)=0.0 ELEC=Drain VSTEP=0.1 NSTEP=10
$SAVE SOLUTION OUT.FILE=VGSO.VDS1.O VV.MODELS
print solution X.MIN=@interfa3 X.MAX=@interfa3 Y.MIN=0.0 Y.MAX=0.0
if.end
if cond=(@iter=1)
Serror of Ioff(0.9*L) is within tolerance
EXTRACT NAME=Ioff.L0 EXPRESS="@I(Drain)" now units=[A/\mu m]
  + clear print
EXTRACT NAME=IovarL2 EXPRESS="ABS((@Ioff.L0-@Ids.L2)/@Ioff.L0)"
  + now clear print
EXTRACT NAME=Ivar.acc EXPRESS="ABS((@IvarL2-@Ivr.spec)/@Ivr.spec)"
  + now clear print

COMMENT The procedure CENTER_POCKET_DOPING(L,Lp)
IF COND=(@Ivar.acc < 5.0E-3)
Serror of sensitivity is within 0.5% tolerance

ASSIGN NAME=nc.guess N.VALUE=@nc.den print
ASSIGN NAME=np.guess N.VALUE=@np.den print
Ssave to file
EXTRACT NAME=Nc EXPRESS="@nc.den" now out.file=@FILE1 twb
  + units=[cm^{-3}] clear print
EXTRACT NAME=Np EXPRESS="@np.den" now out.file=@FILE1 twb
  + units=[cm^{-3}] clear print
EXTRACT NAME=Lpocket EXPRESS="@Lp" now out.file=@FILE1 twb
  + units=[\mu m] clear print
EXTRACT NAME=Lcenter EXPRESS="@Lc" now out.file=@FILE1 twb
  + units=[\mu m] clear print
EXTRACT NAME=Lchan0 EXPRESS="@chan.len" now out.file=@FILE1 twb
  + units=[\mu m] clear print
EXTRACT NAME=Ioff.L0 EXPRESS="@I(Drain)" now out.file=@FILE1 twb
  + units=[A/\mu m] clear print
EXTRACT NAME=Lchan2 EXPRESS="@L2" now out.file=@FILE1 twb
  + units=[\mu m] clear print
EXTRACT NAME=Ioff.L2 EXPRESS="@Ids.L2" now out.file=@FILE1 twb
  + units=[A/\mu m] clear print
EXTRACT NAME=IovarL2
  + EXPRESS="ABS((@Ioff.L0-@Ioff.L2)/@Ioff.L0)*100." now out.file=@FILE1
twb units=[\%] clear print
EXTRACT NAME=Acc.foff EXPRESS="@IOFF.ACC*100."
  + NOW out.file=@FILE1 twb units=[\%] clear PRINT
EXTRACT NAME=Acc.Ivar EXPRESS="@Ivar.acc*100."
  + NOW out.file=@FILE1 twb units=[\%] clear PRINT
\$Plot surface potential at \( y = 0 \).

COMMENT \( V_{gs}=0[V], V_{ds}=1.0[V] \)

PLOT.1D POTENTIAL X.START=@Ls X.END=@interfa3 Y.START=0.0 + Y.END=0.0 + TITLE="surface ptnl(\( NC = \@nc.den \), \( \alpha Ls \) < \( x < \) \( @interfa3 \) \( \mu m \), \( y = 0 \mu m \))" + DEVICE="POSTSCRIPT" PLOT.OUT=@FILE2 out.file=@FILE3 LABEL LABEL="VGS=0.0V" X=0.02 LABEL LABEL="VDS=1.0V" LABEL LABEL="Lp="@Lp[\mu m], Lc="@Lc[\mu m]" LABEL LABEL="Ids="@Ioff.L0[A/\mu m], NP="@np.den[cm^{-3}]" COMMENT POTENTIAL CONTOURS PLOT.2D BOUND JUNC DEPL + TITLE="POTENTIAL CONTOURS(\( NC = \@nc.den[cm^{-3}] \))" + DEVICE="POSTSCRIPT" PLOT.OUT=@FILE6 FILL SCALE CONTOUR POTENTIAL MIN=-1. MAX=2. DEL=0.1 COLOR=6 LABEL LABEL="VGS=0.0V" X=0.02 Y=0.15 LABEL LABEL="VDS=1.0V" LABEL LABEL="Lp="@Lp[\mu m], Lc="@Lc[\mu m]" LABEL LABEL="Ids="@Ioff.L0[A/\mu m], NP="@np.den[cm^{-3}]" \$Vgs=1[V], Vds=1.0[V]

METHOD N.DAMP=T CX.TOLER=1.e-2

SOLVE V(Gate)=0.1 V(Drain)=1.0 ELEC=Gate VSTEP=0.1 NSTEP=9 print solution X.MIN=@interfa3 X.MAX=@interfa3 Y.MIN=0.0 Y.MAX=0.0 EXTRACT NAME=Ion.L0 EXPRESS="@I(Drain)" now out.file=@FILE1 twb + units=[A/\mu m] clear print

ASSIGN NAME=iter N.VALUE=3 print

ELSE COND=(@IovarL2<@Ivr.spec)

if cond=(@Ivr.pre>@Ivr.spec)

ASSIGN NAME=nc.den N.VALUE=@nc.den+@del.nc/2. print

ASSIGN NAME=del.nc N.VALUE=@del.nc/2. print

ASSIGN NAME=Ivr.pre N.VALUE=@IovarL2 print

else

ASSIGN NAME=nc.den N.VALUE=@nc.den+@del.nc print

ASSIGN NAME=Ivr.pre N.VALUE=@IovarL2 print

if.end

ASSIGN NAME=iter N.VALUE=2 print

ASSIGN NAME=ids.pre N.VALUE=0. print

ELSE
if cond=((@IovarL2>@Ivr.spec)(@Ivr.pre<@Ivr.spec)(@Ivr.pre>0.))
ASSIGN NAME=nc.den N.VALUE=@nc.den-@del.nc/2. print
ASSIGN NAME=del.nc N.VALUE=@del.nc/2. print
ASSIGN NAME=Ivr.pre N.VALUE=@IovarL2 print
else
ASSIGN NAME=nc.den N.VALUE=@nc.den-@del.nc print
ASSIGN NAME=Ivr.pre N.VALUE=@IovarL2 print
if end
ASSIGN NAME=iter N.VALUE=2 print
ASSIGN NAME=ids.pre N.VALUE=0. print
IF.END
else cond=(@iter=3)
$Find potential plot for Vgs=Vds=0V, and Ids vs. Vgs for Vds=0.1V and 1.0V
COMMENT Vgs=0[V],Vds=0[V]
PLOT.1D POTENTIAL X.START=@Ls X.END=@interfa3 Y.START=0.0
+ Y.END=0.0
+ TITLE="surface ptnl(NC="@nc.den","@Ls"< X <"@interfa3"μm, y = 0μm")"
+ DEVICE="POSTSCRIPT" PLOT.OUT=@FILE4 out.file=@FILE5
LABEL LABEL="VGS=0.0V" X=0.02
LABEL LABEL="VDS=0.0V"
LABEL LABEL="Lp="@Lp[μm],Lc="@Lc[μm]"
LABEL LABEL="NP="@np.den[cm^-3]"
COMMENT Vgs=0[V],Vds=0.1[V]
SOLVE V(Gate)=0.0 V(Drain)=0.1
SAVE SOLUTION OUT.FILE=VGS0_VDS0.1 W.MODELS
print solution X.MIN=@interfa3 X.MAX=@interfa3 Y.MIN=0.0 Y.MAX=0.0
COMMENT Vgs=0[V],Vds=0.1[V]
SOLVE V(Gate)=0.0 V(Drain)=0.2 ELEC=Drain VSTEP=0.1 NSTEP=8
SAVE SOLUTION OUT.FILE=VGS0_VDS1.0 W.MODELS
print solution X.MIN=@interfa3 X.MAX=@interfa3 Y.MIN=0.0 Y.MAX=0.0

COMMENT Set up log file for Ids-Vgs(Vgs=0 1[V]) at Vds=0.1[V]
LOG OUT.FILE=@FILE7
load in.file=VGS0_VDS0.1
METHOD N.DAMP=T CX.TOLER=1.e-2
SOLVE V(Gate)=0.0 V(Drain)=0.1 ELEC=Gate VSTEP=0.05 NSTEP=20
COMMENT Set up log file for Ids-Vgs(Vgs=0 1[V]) at Vds=1.0[V]
LOG OUT.FILE=@FILE8
load in.file=VGS0_VDS1.0
METHOD N.DAMP=T CX.TOLER=1.e-2
SOLVE V(Gate)=0.0 V(Drain)=1.0 ELEC=Gate VSTEP=0.05 NSTEP=20
PLOT.1D IN.FILE=@FILE8 Y.AXIS=I(Drain) X.AXIS=V(Gate) LINE=1
+ TITLE="Ids-Vgs CHARACTERISTICS"
+ DEVICE="POSTSCRIPT" PLOT.OUT=@FILE9 OUT.FILE=@FILE11
LABEL LABEL="Vds=1.0[V](line=1)," X=0.2
LABEL LABEL="Vds=0.1[V](line=2)"
LABEL LABEL="Lp="@Lp"[\mu m],Lc="@Lc"[\mu m]"
LABEL LABEL="NC="@nc.den"[cm^{-3}],NP="@np.den"[cm^{-3}]"
PLOT.1D IN.FILE=@FILE7 Y.AXIS=I(Drain) X.AXIS=V(Gate) LINE=2
+ UNCHANGE DEVICE="POSTSCRIPT" PLOT.OUT=@FILE9
+ OUT.FILE=@FILE12
PLOT.1D IN.FILE=@FILE8 Y.AXIS=I(Drain) X.AXIS=V(Gate)
+ Y.LOG LINE=1
+ TITLE="Ids-Vgs CHARACTERISTICS"
+ DEVICE="POSTSCRIPT" PLOT.OUT=@FILE10
LABEL LABEL="Vds=1.0[V](line=1)," X=0.2
LABEL LABEL="Vds=0.1[V](line=2)"
LABEL LABEL="Lp="@Lp"[\mu m],Lc="@Lc"[\mu m]"
LABEL LABEL="L="@chan.len"[nm]"
LABEL LABEL="NC="@nc.den"[cm^{-3}],NP="@np.den"[cm^{-3}]"
PLOT.1D IN.FILE=@FILE7 Y.AXIS=I(Drain) X.AXIS=V(Gate)
+ Y.LOG LINE=2
+ UNCHANGE DEVICE="POSTSCRIPT" PLOT.OUT=@FILE10

L.MODIFY break PRINT
else
$iter=2 case
EXTRACT NAME=Ids.L2 EXPRESS="@I(Drain)" now units=[.4/\mu m] clear
+ print
EXTRACT NAME=IOFF.ACC
+ EXPRESS="ABS((@Ids.L2-@ids.spec)/@ids.spec)" now clear print

COMMENT The procedure POCKET_DOPING(Nc,L,Lp)
IF COND=(@IOFF.ACC<5.0E-3)
$ioff(0.9*L) is within 0.5% error tolerance with reference to ids.spec=1nA/\mu m
ASSIGN NAME=iter N.VALUE=1 print
EXTRACT NAME=L2 EXPRESS="@chan.len" now units=[\mu m] clear print
ELSE COND=(@Ids.L2<@ids.spec)
if cond=(@ids.pre>@ids.spec)
ASSIGN NAME=np.den N.VALUE=@np.den-@del.np/2. print
ASSIGN NAME=del.np N.VALUE=@del.np/2. print
ASSIGN NAME=ids.pre N.VALUE=@Ids.L2 print
else
ASSIGN NAME=np.den N.VALUE=@np.den-@del.np print
ASSIGN NAME=ids.pre N.VALUE=@Ids.L2 print
if.end
ELSE
if cond=((@Ids.L2>@ids.spec)(@ids.pre<@ids.spec)(@ids.pre>0.))
ASSIGN NAME=np.den N.VALUE=@np.den+@del.np/2. print
ASSIGN NAME=del.np N.VALUE=@del.np/2. print
ASSIGN NAME=ids.pre N.VALUE=@Ids.L2 print
else
ASSIGN NAME=np.den N.VALUE=@np.den+@del.np print
ASSIGN NAME=ids.pre N.VALUE=@Ids.L2 print
if.end
IF.END
if.end
L.END
l.end
APPENDIX C

ALGORITHM TO FIND SINGLE-DEVICE VALUES OF

\[ \ell_{p0}, \ell_{c0} \text{ AND } K_p \]

This appendix describes the algorithm to find single-device values, i.e., optimal values of the characteristic lengths \( \ell_{p0}, \ell_{c0} \) and the fitting potential parameter \( K_p \) in the surface potential model equation (Equation 4.6) for the deep-submicron pocket MOSFETs by fitting model surface potentials to Medici surface potentials. A simple empirical formula proposed in [24], [60] is used to ease finding brackets for the solutions of the characteristic lengths \( \ell_{p0} \) and \( \ell_{c0} \) at the pocket and center regions. Thus,

\[
\ell_{p0} = m_p(t_{ox}r_jW_{D(p)}^2)^{1/3} \quad (C.1)
\]

\[
\ell_{c0} = m_c(t_{ox}r_jW_{D(c)}^2)^{1/3} \quad (C.2)
\]

where \( W_{D(p)} \) and \( W_{D(c)} \) are the depletion-layer depths (Equation 2.4) under gate at the pocket and center regions, respectively.

The algorithm is implemented in three procedures (functions): MULTIPLIER(), CENTER(), SOLUTIONS() and DYNAMIC_BRACKETS(), which are to find the single-device values of \( \ell_{p0}, \ell_{c0} \) and \( K_p \) with relative errors within 0.1 % of the
model surface potentials \( \psi_0(x_1 = L/6), \psi_0(\text{min})(x_2) \) and \( \psi_0(\text{max})(x_3) \) with reference to the corresponding Medici potentials \( \psi_0(\text{MED})(x_1 = L/6), \psi_0(\text{MED})(x_2) \) and \( \psi_0(\text{MED})(x_3) \).

function MULTIPLIER_CENTER(\( \ell_{p0}, K_p \))

Let \( f(m_c) = [\psi_0(\text{MED})(x_1) - \psi_0(x_1)] + [\psi_0(\text{MED})(x_2) - \psi_0(\text{min})(x_2)] + [\psi_0(\text{MED})(x_3) - \psi_0(\text{max})(x_3)] \) be a function of \( m_c \). The positions \( x_1, x_2 \) and \( x_3 \) are found from the Medici potential profile. Given \( \ell_{p0} \) and \( K_p \), this procedure solves \( f(m_c) = 0 \) for \( m_c \) using the secant method [61].

begin

\( \text{left\_bracket} \leftarrow 0.5; \, \text{right\_bracket} \leftarrow 0.5; \)

\( \Delta \text{right\_bracket} \leftarrow \text{right\_bracket}/100; \, j \leftarrow 0; \)

while (\( j \leq 100 \)) {

/*Use a procedure implementing the secant method from here.*/

Calculate \( \psi^i_0(x_1), \psi^i_0(\text{min})(x_2) \) and \( \psi^i_0(\text{max})(x_3) \) at the \( i \)th iteration of the secant method for the given \( \ell_{p0} \) and \( K_p \) and \( m_c^i \), i.e., \( \ell_{c0}^i \), using Equation 4.6.;

if \( \left| \left( \frac{f_{n-1} - f_{n-2}}{f_{n} - f_{n-1}} \right) (m_c^i - m_c^{i-1}) \right| < \epsilon_{\text{secant}} (= 10^{-7}) \) \&\& \( m_c^i > 0 \) \&\& \( m_c^i < 1 \) {

/*convergence*/

\( \text{flag\_conv} \leftarrow 1; \)

return \( m_c^i \) and \( \ell_{c0}^i; \)
}

else{

}
263

/*adjust right bracket*/

\begin{align*}
right\_bracket^i+1 & \leftarrow right\_bracket^i - \Delta right\_bracket;
\end{align*}

} /*end of while(j)*/

\end /*end of MULTIPLIER\_CENTER()*/

\begin{function}{SOLUTIONS}(K_p1, K_p2, n_{K_p}, m_{p1}, m_{p2}, n_{m_p})

This procedure finds the solutions $K_p$, $m_p$ and $m_c$ satisfying \(\text{error}_\psi_0(x_1) = \frac{|\psi_0(\text{MED})(x_1) - \psi_0(x_1)|}{\psi_0(\text{MED})(x_1)} < \epsilon_{\psi_0}(=0.1 \%)\), \(\text{error}_\psi_0(x_2) = \frac{|\psi_0(\text{MED})(x_2) - \psi_0(m_{\text{min}})(x_2)|}{\psi_0(\text{MED})(x_2)} < \epsilon_{\psi_0}\), and \(\text{error}_\psi_0(x_3) = \frac{|\psi_0(\text{MED})(x_3) - \psi_0(m_{\text{max}})(x_3)|}{\psi_0(\text{MED})(x_3)} < \epsilon_{\psi_0}\). If all the error criteria are not satisfied, SOLUTIONS() finds $K_p$, $m_p$ and $m_c$ which make the sum of $\text{error}_\psi_0(x_1)$, $\text{error}_\psi_0(x_2)$ and $\text{error}_\psi_0(x_3)$ minimal.

\begin{begin}

/*$K_p1$ and $K_p2$ are lower and upper brackets of the solution $K_p$, respectively. $m_{p1}$ and $m_{p2}$ are lower and upper brackets of the solution $m_p$, respectively. $n_{K_p}$ and $n_{m_p}$ determine the number of iterations of loops.*/

\[\Delta K_p \leftarrow \frac{K_p2 - K_p1}{n_{K_p}}; \quad \Delta m_p \leftarrow \frac{m_{p2} - m_{p1}}{n_{m_p}};\]

flag\_visit = 0 (to be used to find a minimum of the sum of total errors);

for \(i = 0, K_p = K_p1; i \leq n_{K_p}; i++\), $K_p+ = \Delta K_p$) {

\begin{align*}
\text{for}(j = 0, m_p = m_{p1}; j \leq n_{m_p}; j++, m_p+ = \Delta m_p) & \{
\end{align*}

/*Calculate $\ell_{p0}$ at the pocket region for given values of $m_p$ and
geometrical parameters. */

$$\ell_{p0} \leftarrow m_p (t_{ox} r_j W_{D(p)}^2)^{1/3};$$

/* Find $$m_c$$ using the procedure MULTIPLIER_CENTER */

$$m_c \leftarrow \text{MULTIPLIER\_CENTER}(\ell_{p0}, K_p);$$

/* Calculate $$\ell_{c0}$$ at the center region. */

$$\ell_{c0} \leftarrow m_c (t_{ox} r_j W_{D(c)}^2)^{1/3};$$

/* Calculate the model surface potentials at $$x_1$$, $$x_2$$ and $$x_3$$ using Equation 4.6, and the the sum of errors of the model surface potentials with reference to the corresponding Medici surface potentials. */

$$\text{sum\_error} \leftarrow \text{error\_}\psi_0(x_1) + \text{error\_}\psi_0(x_2) + \text{error\_}\psi_0(x_3);$$

if(flag\_conv == 1 && error\_}\psi_0(x_1) < \epsilon_\psi, && error\_}\psi_0(x_2) < \epsilon_\psi, && error\_}\psi_0(x_3) < \epsilon_\psi) 

return $$m_p$$, $$m_c$$ and $$K_p;$$

/* Find a minimum of the sum of error\_}\psi_0(x_1), error\_}\psi_0(x_2) and error\_}\psi_0(x_3). */

else if(flag\_conv == 1 && flag\_visit == 0) {

$$\text{flag\_visit} \leftarrow 1; \text{min\_sum\_error} \leftarrow \text{sum\_error}; K'_p \leftarrow K_p; m'_p \leftarrow m_p;$$

$$m'_c \leftarrow m_c;$$

}

else if(flag\_conv == 1 && flag\_visit == 1 

&& min\_sum\_error > sum\_error) {

}
$\text{min} \cdot \text{sum} \cdot \text{error} \leftarrow \text{sum} \cdot \text{error}; \ K_p' \leftarrow K_p; \ m_p' \leftarrow m_p; \ m_c' \leftarrow m_c;$

\}

else

;

if(m_p1 == m_p2)

break loop(j);

\} /*end of loop(j)*/

if(K_p1 == K_p2)

break loop(i);

\} /*end of loop(i)*/

return $K_p'$, $m_p'$ and $m_c'$;

\end /*end of SOLUTIONS()*/

\begin function \text{DYNAMIC\_BRACKETS()}

This procedure finds $m_p^0$ with $K_p = 0$, and $K_p^0$ with $m_p^0$ found, as initial guesses of solutions $m_p$ and $K_p$. Whenever the solutions $m_p$ and $K_p$ are out of brackets, the brackets are adjusted automatically to find final solutions satisfying the given error tolerance $\epsilon_{\text{es}} (=0.1 \%)$ for all the three surface potentials $\psi_{s0}(x_1)$, $\psi_{s0}(x_2)$ and $\psi_{s0}(x_3)$.

\begin begin

/*Set initial lower and upper brackets. and increments of $K_p^0$ and $m_p^0.*/$
\[ n_{m_p}^0 \leftarrow 10 - 40; \quad n_{K_p}^0 \leftarrow 10 - 40; \quad m_{p1}^0 \leftarrow 0; \quad m_{p2}^0 \leftarrow 0.25; \quad K_{pl}^0 \leftarrow -0.25; \]

\[ K_{p2}^0 \leftarrow -0.1; \quad \Delta m_p^0 \leftarrow \frac{m_{p2}^0 - m_{p1}^0}{n_{mp}^0}; \quad \Delta K_p^0 \leftarrow \frac{K_{p2}^0 - K_{p1}^0}{n_{K_p}^0}; \]

/*initial guess of \( m_p \) with \( K_p = 0^* /\)

\[ \text{flag.conv} \leftarrow 0; /* used to indicate convergence in MULTIPLIER.CENTER()*/ \]

\[ K_p \leftarrow 0; \quad m_{p2} \leftarrow m_{p2}^0; \]

\[ \text{while}(\text{flag.conv} == 0) \{ \]

\[ \text{call SOLUTIONS}(K_p, K_p, n_{K_p}^0, m_{p1}, m_{p2}, n_{mp}^0); \]

\[ m_{p2} \leftarrow 2 \times m_{p2}; \]

\[ \} \]

/*initial guess of \( K_p \) with initial solution \( m_p \) found*/

\[ \text{call SOLUTIONS}(K_{p1}, K_{p2}, n_{K_p}^0, m_p, m_p, n_{mp}^0); \]

\[ m_{p1} \leftarrow m_p - \Delta m_{p1}^0; \quad m_{p2} \leftarrow m_p + \Delta m_{p1}^0; \]

\[ K_{p1} \leftarrow K_p - \Delta K_{p1}^0; \quad K_{p2} \leftarrow K_p + \Delta K_{p1}^0; \]

/*Call SOLUTIONS(). If solutions \( m_p \) and \( K_p \) are out of brackets, adjust the brackets dynamically.*/

\[ n_{mp} \leftarrow n_{mp}^0; \quad n_{K_p} \leftarrow n_{K_p}^0; \quad i \leftarrow 1; \]

\[ \text{MAX.ITER} \leftarrow 10 - 40; \]

\[ \text{while}(i \leq \text{MAX.ITER}) \{ \]

\[ \Delta K_p \leftarrow \frac{K_{p2} - K_{p1}}{n_{K_p}^0}; \quad \Delta m_p \leftarrow \frac{m_{p2} - m_{p1}}{n_{mp}^0}; \]

\[ \text{call SOLUTIONS}(K_{p1}, K_{p2}, n_{K_p}, m_{p1}, m_{p2}, n_{mp}); \]

\[ \text{if}(\text{flag.conv} == 1 \&\& \text{error} \cdot \psi_0(x_1) < e_\psi \&\& \text{error} \cdot \psi_0(x_2) < e_\psi, \]

\[ \]
\&\& \text{error}\_\psi_0(x_3) < \epsilon_\psi,

return final solutions \(m_p, m_c\) and \(K_p\).

if\((K_p > K_{p1} \&\& K_p < K_{p2} \&\& m_p > m_{p1} \&\& m_p < m_{p2})\) \{ 

\(K_{p1} \leftarrow K_p\); \(K_{p2} \leftarrow K_p + \Delta K_p^0\);

\(m_{p1} \leftarrow m_p\); \(m_{p2} \leftarrow m_p + \Delta m_p^0\);

\}

else\{ 

/*adjust \(K_p\) brackets*/

if\((K_p == K_{p1} \&\& K_p < K_{p2})\) \{ 

\(K_{p1} \leftarrow K_p - \Delta K_p^0\); \(K_{p2} \leftarrow K_p\);

\}

else if\((K_p > K_{p1} \&\& K_p == K_{p2})\) \{ 

\(K_{p1} \leftarrow K_p\); \(K_{p2} \leftarrow K_p + \Delta K_p^0\);

\}

else : 

/*adjust \(m_p\) brackets*/

if\((m_p == m_{p1} \&\& m_p < m_{p2})\) \{ 

\(m_{p1} \leftarrow m_p - \Delta m_p^0\); \(m_{p2} \leftarrow m_p\);

\}

else if\((m_p > m_{p1} \&\& m_p == m_{p2})\) \{ 

\(m_{p1} \leftarrow m_p\); \(m_{p2} \leftarrow m_p + \Delta m_p^0\);
else:
}/*end of else*/

i++;

if(i > MAX_ITER){
    MAX_ITER ← 2 * MAX_ITER; \( n_{K_p} \leftarrow 2 \times n_{K_p} \); \( n_{m_p} \leftarrow 2 \times n_{m_p} \);
}

}/*end of while(i)*/

}/*end of DYNAMIC_BRACKETS()*/
REFERENCES


