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2007

Chiah, S. B. (2007). Unified AC charge and DC current modeling for very-deep-submicron CMOS technology. Doctoral thesis, Nanyang Technological University, Singapore.

<https://hdl.handle.net/10356/3534>

<https://doi.org/10.32657/10356/3534>

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Unified AC Charge and DC Current Modeling for Very-Deep-Submicron CMOS Technology

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School of Electrical and Electronic Engineering

A thesis submitted to the Nanyang Technological University
in fulfilment of the requirement for the degree of
Doctor of Philosophy

2007

Acknowledgments

“Measurement data contains a lot of information that explains physical behavior of MOSFET devices, awaiting us to uncover”. This sentence is really inspiring me from the first day I knew my research supervisor, Dr. Zhou Xing, even until now. He has been a great philosopher. With his enthusiasm in device physical modeling, he guided me through the entire course of my research study in Nanyang Technological University. He is also a friend in sharing thoughts and views on future development in the modeling field, which is enhancing my curiosity to know more and continuously contribute in device modeling.

I would like to express my heartfelt gratitude to the members of SRC in Chartered Semiconductor Manufacturing, Ltd. Singapore, which include Drs. Hsia Liang-Choo, Sanford Chu and Michael Cheng for supporting me in this project and providing process information and data for my modeling work. I am also grateful to Drs. Lap Chan, Alex See and Lim Khee Yong for their guidance in fabrication process in Chartered Semiconductor Manufacturing, Ltd. Singapore. Thanks also go to my friends, especially to Mr. Karthik Chandrasekaran for supporting numerical data in this work, Mr. See Guan Huei and Dr. Shangguan Wangzuo for group discussions.

Finally, I would like to express my deepest gratitude to family - my father, my mother, my brothers, my sisters and especially my wife, Liem Sok Hien for her understanding and mental support during the entire course of my research study. Without their constant encouragement, I would never understand “digging into measurement data” from my advisor.

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Abstract

A novel approach to formulating unified charge and drain current models for MOSFETs is presented. The approach is original and general. The charge modeling methodology is based on three regional surface-potential solutions, which describes three operating regions in MOSFETs. The modeling approach is extended to all regions with an explicit single-piece unified compact charge model. Interpolating functions are used in the compact model. Smoothing parameters are carefully selected so as to maintain model continuity at the transition points. This is also to ensure charge-neutrality across different regions of operation especially at the flat-band condition. The charge modeling approach requires no modification in formulation to include coupled polycrystalline silicon and quantum-mechanical effects for all regions. Both coupled effects are accounted for in regional surface-potential formulations, which are formulated through two-coupled Pao–Sah equations that can be solved regionally and explicitly. This is nontrivial for other modeling methodologies. The proposed model requires 12 fitting parameters, which are extracted in a systematic way following a 10-step extraction procedure with only *one* $C_{gg} - V_g$ measurement data. The approach has been shown to have the potential to be extended to non-conventional bulk-Si MOSFETs structure such as in strain-Si or hetero-structure MOSFETs.

The drain current modeling methodology for nanometer device MOSFETs is presented, which is based on bulk-referenced formulation for symmetric drain current characteristic with proper handling of lateral/transverse-field mobility degradation effects. New length/width-dependent drain current model is formulated with threshold

voltage definition, which requires less effort and is simpler in modeling geometry-dependence effects. Reverse short-channel and short-channel effects are accurately formulated in a unified threshold-voltage expression to capture threshold-voltage roll-up/off effects at decreasing channel length. Similarly, at decreasing channel width, inverse narrow-width and narrow-width effects are excellently modeled. With a geometry-dependent threshold-voltage model, the drain current modeling methodology becomes simpler and efficient for geometry effects due to intrinsic properties such as mobility degradation and channel-length modulation; and extrinsic properties such as series resistance and dog-bone layout dependent effects. The unified threshold-voltage expression requires 16 fitting parameters and 6 bias coefficients, which are extracted in a 4-step calibration approach whereas the unified I_{ds} model requires 10-step to extract 15 fitting parameters plus 7 geometry coefficients. Both unified charge and drain current models are validated and predicted excellently the analytical derivation, (2-D) numerical simulation, and experimental data.

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List of Symbols

Symbol	Remark
V_{gs}	Gate to source voltage
V_{gb}	Gate to bulk voltage
V_{ds}	Drain to source voltage
V_{db}	Drain to bulk voltage
V_{bs}	Bulk to source voltage
V_{cb}	Channel to bulk voltage
V_{deff}	Effective drain to source voltage
$V_{deff,\Delta}$	Effective drain to source voltage with <i>non-pinned</i> ϕ_s -solutions
$V_{deff,\Delta p}$	Effective drain to source voltage with <i>non-pinned</i> ϕ_s/ϕ_p -solutions
V_{geff}	Effective gate to source voltage
$V_{geff,\Delta p}$	Effective gate to source voltage with $V_{tp,\phi}$
$V_{geff,\Delta}$	Effective gate to source voltage with $V_{t,\phi}$
V_{dsat}	Saturation drain to source voltage
V_{dd}	High drain to source voltage ($V_{ds} = 1.32\text{V}$)
V_{d0}	Low drain to source voltage ($V_{ds} = 0.1\text{V}$)
V_{bb}	High bulk voltage ($V_{bs} = -1.2\text{V}$)
V_{b0}	Low bulk voltage ($V_{bs} = 0$)
V_{gg}	High gate to source voltage ($V_{gs} = 1.2\text{V}$)
V_{b0}	Low bulk voltage ($V_{bs} = 0$)
$V_{GBF}, V_{GBA}, V_{gbs}$	Forward unified model
V_{GBR}	Reverse unified model
$V_{tL,i}$	Ideal long-channel threshold voltage
$V_{t,CS}$	Threshold voltage with charge-sharing effect
$V_{t,DIBL}$	Threshold voltage with charge-sharing and barrier-lowering effects
V_t	Unified threshold voltage
$V_{t,\phi}$	“Non-pinned” ϕ_s -based threshold voltage
A_b	Bulk charge factor
$A_{b,\Delta}$	Bulk charge factor with non-pinned ϕ_s -solutions
Q_{BSC}	Short-channel depletion charge
Q_{BL}	Long-channel depletion charge
C_f	Fringing capacitor
n_i	Intrinsic carrier concentration
V_{th}	Thermal voltage
t_{ox}	Gate oxide thickness
ϵ_{ox}	Permittivity of oxide
ϵ_{si}	Permittivity of silicon
χ	Electron affinity
$\hbar$	Plank constant

m_o	Effective mass
m_e	Electron mass
m_h	Hole mass
ϕ_{MS}	Metal to silicon bulk work function difference
W_M	Metal work function
ϕ_{ms}	Poly gate to silicon bulk work function difference
ϕ_m	Poly gate work function
W_S	Work function for silicon
V_{FB}, V_{fb}	Flat-band voltage
N_{ch}	Channel doping concentration
N_{sub}	Uniform substrate doping concentration
N_a	Acceptor
N_d	Donor
N_{pk}	Peak Gaussian doping
N_{tot}	Summation of uniform substrate and peak doping
N_{eff}	Effective channel doping concentration
N_{sd}	Source and drain concentration
κ^{qm}	Quantum-mechanical effect model parameter
W	Effective channel width
W_g	Drawn channel width
L_{eff}	Effective channel length
L_g	Drawn gate length
L_∞	Longest gate length ($L_g=10\mu\text{m}$)
L_{min}	Shortest gate length ($L_g=0.13\mu\text{m}$)
L_{mid}	Mid gate length ($L_g=0.6\mu\text{m}$)
W_∞	Longest gate length ($W_g=10\mu\text{m}$)
W_{min}	Shortest gate length ($W_g=0.13\mu\text{m}$)
X_j	Lightly-doped drain junction depth
X_{dep}	Depletion width
$\psi(x)$	Position dependent surface potential
ψ_{sL}	Channel surface potential at the drain end
ψ_{s0}	Channel surface potential at the source end
ϕ_f	Silicon intrinsic Fermi potential
$\phi_{f,p}$	Poly intrinsic Fermi potential
ϕ_s	Channel surface potential
ϕ_p	Poly surface potential
$\phi_{CC}, \phi_{cc}, \phi_{CC}^{qm}$	Regional accumulation channel surface potential
$\phi_{ACC}, \phi_{acc}, \phi_{acc}^{qm}$	Unified regional accumulation channel surface potential
$\phi_{DD}, \phi_{dd}, \phi_{dd}^{qm}$	Regional depletion channel surface potential
$\phi_{SUB}, \phi_{sub}, \phi_{sub}^{qm}$	Unified regional depletion channel surface potential
$\phi_{SS}, \phi_{ss}, \phi_{ss}^{qm}$	Regional strong inversion channel surface potential

$\phi_{STR}, \phi_{str}, \phi_{str}^{qm}$	Unified regional strong inversion channel surface potential
$\phi_{DS}, \phi_{ds}, \phi_{ds}^{qm}$	Unified regional channel surface potential from depletion to strong inversion regions
$\phi_{SEFF}, \phi_{seff}, \phi_{seff}^{qm}$	Unified full regions channel surface potential
$\phi_{p,cc}, \phi_{p,cc}^{qm}$	Regional accumulation poly surface potential
$\phi_{p,acc}, \phi_{p,acc}^{qm}$	Unified regional accumulation poly surface potential
$\phi_{p,dd}, \phi_{p,dd}^{qm}$	Regional depletion channel surface potential
$\phi_{p,sub}, \phi_{p,sub}^{qm}$	Unified regional depletion channel surface potential
$\phi_{p,ss}, \phi_{p,ss}^{qm}$	Regional strong inversion channel surface potential
$\phi_{p,str}, \phi_{p,str}^{qm}$	Unified regional Strong inversion channel surface potential
$\phi_{peff}, \phi_{peff}^{qm}$	Unified full regions poly surface potential
Δ	Non-pinned surface-potential at the strong inversion region
ΔE_g	Increase of energy gap in van Dort's model
Q_g	Unified gate charge per unit area
Q_d	Unified drain charge per unit area
Q_s	Unified source charge per unit area
Q_b	Unified bulk charge per unit area
Q_i	Inversion charge per unit area
Q_{sc}	Channel induce charge per unit area
$Q_{b,str}$	Bulk charge at strong inversion region per unit area
$Q_{b,acc}$	Bulk charge at accumulation region per unit area
$Q_{b,ds}$	Bulk charge from depletion to strong inversion regions per unit area
$Q_{b,sa}$	Bulk charge from accumulation to strong inversion regions per unit area
Q_{b0}	Unified bulk charge from accumulation to strong inversion regions per unit area for $V_{ds}=0$
$A_{s,\Delta}$	Q_s auxiliary functions
$A_{d,\Delta}$	Q_d auxiliary functions
$D_{b,\Delta}$	Q_b auxiliary functions
C_{gg}	Gate capacitance
C_{dg}	Drain capacitance
C_{sg}	Source capacitance
C_{bg}	Bulk capacitance
σ_a	Smoothing parameter in $\mathcal{G}_r$ function for unified accumulation channel surface-potential formulation
σ_f	Smoothing parameter in $\mathcal{G}_f$ function for unified depletion channel surface-potential formulation
σ_{eff}	Smoothing parameter in $\mathcal{G}_f$ function for unified channel surface-potential increment formulation.

δ_ϕ	Smoothing parameter in $\mathcal{G}_{eff}$ function for unified ϕ_{seff}/ϕ_{ds} formulation
ρ_r	Smoothing parameter in $\mathcal{G}_r$ function for unified accumulation gate surface-potential formulation
ρ_f	Smoothing parameter in $\mathcal{G}_f$ function for unified depletion gate surface-potential formulation
σ^{qm}	Fitting parameter that shifts the model capacitance at the transition from accumulation to depletion to the actual measurement data in QME model
$\mathcal{G}^{qm}$	Smoothing parameter in $\mathcal{G}_f$ function for unified QME model from accumulation to depletion regions
ε^{qm}	Fitting parameter that conserves the charge neutrality at flat-band condition
$I_{ds,lin}$	Linear drain current
$I_{ds,sat}$	Saturation drain current
$I_{ds,str}$	Strong inversion drain current
I_{ds}	Unified drain current
I_{ds}	Drain current
μ_0	Mobility degradation with transverse-field effect
μ_{eff0}	Mobility degradation with lateral/transverse-field effect
μ_{eff}	Effective mobility degradation
$E_\perp$	Transverse-field
E_{eff}	Effective transverse-field
E_{satn}	Effective saturation field
v_0	Velocity field relation
v_{sat}	Saturation velocity
v_h	Velocity overshoot
$E_\parallel$	Lateral-field
δ_0	Length-dependent lateral-field factor
R_{ch}	Channel resistance
R_{sd}	Series source/drain resistance
R_{int}	Intrinsic resistance
R_{ext}	Extrinsic resistance
R_{tot}	Total resistance
V_{Aeff}	Effective early voltage
V_{diff}	Diffusion gate/drain overdrive voltage
V_{drift}	Drift gate/drain overdrive voltage
V_{gg}	Unified drift gate/drain overdrive voltage
V_{gd}	Unified diffusion gate/drain overdrive voltage
V_{geff}	Effective gate to source voltage
γ	Channel Body factor
γ_p	Poly silicon body factor
$V_s(x)$	Position dependent channel potential

$E_s(x)$	Position dependent lateral surface electric field
V_{bi}	Built-in potential between drain-substrate junction
C_d	Depletion capacitance
g_m	Transconductance
g_{ds}	Output drain conductance
I_{on}	Saturation current (I_{ds} @ $V_{ds}=V_{gs}=V_{dd}$)
h	Effective saturation field coefficient
V_{fbb}	Regression parameter that is linked to metal work function
α_L	Long-channel DIBL parameter for characteristic length
λ_0	SCE for charge sharing parameter at V_{b0}
λ_1	The second parameter in parabolic V_{bs} dependence for charge sharing effect
λ_2	The third charge sharing parameter for parabolic V_{bs} dependence
β	Characteristic length parameter for lateral doping profile
κ_0	Ratio between peak halo-doping and the substrate doping at V_{b0}
κ_1	The second parameter in parabolic V_{bs} dependence for the ratio between peak halo-doping and the substrate doping
κ_2	The third parameter in parabolic V_{bs} dependence for the ratio between peak halo-doping and the substrate doping
α	Characteristic length parameter for potential barrier lowering in short-channel
φ_0	Regression parameter that enhances the region of validity for DIBL effect at low V_{bs}
φ_1	The second parameter in parabolic V_{bs} dependence for DIBL effect
φ_2	The third parameter in parabolic V_{bs} dependence for DIBL effect
τ	Process-dependent parameter in the fringing capacitance
α_w	Characteristic length parameter in potential barrier lowering for NWE
β_w	Characteristic length parameter in doping profile along width-dimension for INWE
κ_w	Ratio between peak doping and the substrate doping for INWE
l_w	Pile-up centroid in doping profile along width dimension
σ	Lateral diffusion factor for LDD junction
l_μ	Pile-up centroid in halo-doping profile
ΔW	Optimization for drawn gate width
μ_1	Mobility parameter for Coulombic scattering
μ_2	Mobility parameter for surface roughness scattering
μ_3	Mobility parameter for phonon scattering
ν	Regression parameter that indicates amount of effective field contribution in mobility due to phonon scattering
χ_b	Regression parameter that indicates amount of bulk-charge contribution in effective field
δ_L	Fitting parameter to adjust the amount of mobility degradation due to lateral-field effect

V_{off}	Offset voltage for I_{off} adjustment.
r_1	Extrinsic series resistance parameter
r_2	Intrinsic series resistance parameter
ξ	Process dependent parameter that represents the relationship between carrier temperature and lateral electric field
δ_{s1}	Length-dependent parameter in δ_s for V_{deff} model
δ_{s2}	Length-dependent parameter in δ_s for V_{deff} model
δ_{s3}	Length-dependent parameter in δ_s for V_{deff} model
δ_{s4}	Length-dependent parameter in δ_s for V_{deff} model
μ_{f1}	Width-dependent parameter in coulombic scattering factor for mobility degradation model
μ_{f2}	Width-dependent parameter in coulombic scattering factor for mobility degradation model
μ_{f3}	Width-dependent parameter in coulombic scattering factor for mobility degradation model
μ_{f4}	Width-dependent parameter in coulombic scattering factor for mobility degradation model
θ_s	Dog-bone layout-dependent parameter in the extrinsic series source/drain resistance
θ_d	Dog-bone layout-dependent parameter in the extrinsic series source/drain resistance
μ_{f5}	Geometry-dependent parameter in coulombic scattering factor of mobility degradation model

CHAPTER 1: INTRODUCTION

1.1 Motivation

Tremendous effort has been put into the CMOS technology in order to get high speed, compact and low power consumption electronic products. As a result, the number of transistors on a chip is growing exponentially. Traditional techniques used for circuit analysis and design become inadequate. Circuit simulators offer detailed circuit analysis with graphics capabilities, better user interfaces and more analysis modes (i.e., AC/DC analysis and transient analysis). However, the key for circuit simulators to predict real circuit performance depends mostly on the accurate modeling of real device electrical characteristics through mathematical equations (transistor model) coded in the simulators.

A number of transistor models have been developed, based on various modeling methodologies and parameter calibration approaches. Watts *et al.* [1] summarizes possible approaches, which can be grouped into surface-potential-based, charge-based and threshold-voltage-based approaches. Each modeling approach has its own merits and weaknesses in terms of predicting AC and DC electrical characteristics with physical parameter variation and geometry/bias-dependence, model compactness, model validity in various regions of operation and model parameter extraction procedure. For example, one may offer “binning” approach [2] for model parameter calibration, and another may assume linear or parabolic relationship for geometry-dependent model parameter [1], yet another may not even have physical effect in accumulation region [3].

The fact is that a tractable and reliable model without any approximation and taking into consideration short/narrow-channel effects is not available for circuit

simulation. In addition, solving the governing equations numerically entails enormous computational resources. However, one can still develop an analytical explicit compact model for the circuit simulator based on some approximations with good modeling strategy, the model can match experimental data within acceptable error.

Conventional models are facing great challenges from the trade off between accuracy and simplicity to be used in the circuit simulator [4]. Physical in a way that the models are not only be able to predict electrical characteristics of the classical device structure but the potential of the model to be extended for next generation CMOS. This becomes the challenge of existing models continuously to drive the non-classical device structures, not by re-construction of the model but the model extension from the existing modeling approach. Accuracy in predicting electrical characteristics with minimum measurement data requirement for the model parameter calibration is another concern of the existing modeling approaches and the model parameter calibration methodologies. Simple and easy to understand textbooks' equation that makes circuit designers familiar with, is a plus in model development.

This forms the major motivation of this research work, which is named as "Unified AC Charge and DC Current Modeling for Very-Deep-Submicron CMOS Technology". It is aimed at assisting new technology development and circuit design. The developed model in this work is based on a non-pinned surface-potential-based charge model for AC modeling and a threshold-voltage-based drain-current model for DC modeling. Both the AC and DC modeling methodologies are physics based and is extendable to non-classical device structures [5]. Simple and easy model parameter extraction through a step-by-step calibration approach [6] [7] is developed. Model

predictability for the advanced effects, such as quantum-mechanical and poly-doping effects [8] has also been established. Both models are characterized from experimental and TCAD simulation data truly for advanced nano-scale MOSFETs.

1.2 Objectives

The objectives of the research work are to (1) develop AC MOSFET model by compacting or integrating over the spatial variables on the governing equations such as the Poisson's equation, Gauss' law and charge neutrality equations into a compact form. (2) Enhance and formulate a new unified DC model that is symmetric and properly handles the velocity saturation effect in the lateral-field effective mobility. Geometry effects for short/narrow-channel devices in nano-scale MOSFET technology are investigated and physically modeled. (3) Extract AC and DC model parameters in a *step-by-step* calibration procedure based on full TCAD simulation and experimental data to give a first-order approximation on the device performance.

1.3 Major contribution of the Thesis

The major contributions of this research work are:

- 1) A unified regional approach to consistent and symmetric DC/AC modeling of deep submicron MOSFET is proposed. Charge-based pinned AC model is developed which is fully consistent with DC model without the need for C-V data in bias-dependent calibration, including poly-depletion effect for both DC and AC modeling [9].

- 2) A single-piece unified surface-potential is formulated explicitly from accumulation to strong inversion with smooth transition across operating regions, which consists of three unified regional surface potential expression in the accumulation, depletion and strong inversion regions [10].
- 3) A unified regional charge model with non-pinned surface potential is proposed. The regional models predict the physical behavior correctly in asymptotical regions and are joined seamlessly at the flat-band and threshold voltage with C_{∞} continuity for all device operating regions [11].
- 4) A single piece polycrystalline (silicon) accumulation/depletion/inversion model with implicit/explicit surface-potential solution is demonstrated. For the first time the poly-accumulation/inversion models are formulated physically from two coupled “Pao-Sah” equations [12].
- 5) The physical parameter extraction from decoupled poly-doping/quantum-mechanical effects based on regional surface-potential/gate-charge models is proposed. For the first time the physical parameter extraction procedure from one AC measurement with coupled polycrystalline silicon and quantum-mechanical effects are demonstrated with a *step-by-step* approach [8].
- 6) The unified regional charge-based MOSFET model calibration with the non-pinned surface-potential approach is proposed. Model and smoothing parameter extraction procedure is illustrated in a *step-by-step* procedure [6].
- 7) A full region AC model with quantum-mechanical effect is proposed. Coupled quantum-mechanical and polycrystalline silicon effects in AC modeling are investigated.

- 8) A unified regional MOSFET model with source-drain symmetry for drain current model is suggested. Bulk-referenced effective transverse field is formulated and correct physical effect compared with drain conductance measurement data is validated [9] [13].
- 9) An analytical closed-form “drift + diffusion” model is proposed, based on previously-proposed unified compact model, which approaches the theoretical drift- and diffusion-current asymptotes and guarantees smooth transition in the moderate-inversion region for both current and transconductance-to-current ratio [10] [14].
- 10) Enhance the effective channel doping transformation from vertical non-uniform doping profile and lateral pile-up boron concentration for reverse short-channel effect (RSCE) with lateral pile-up centroid (l_μ) that has great influence on the threshold voltage (V_t) [15].
- 11) Long-channel drain induced barrier lowering effect is investigated with empirical V_{ds} -dependent and finally is derived based on quasi-2D solutions [7], [16], [17].
- 12) The parabolic V_{bs} -dependent for threshold voltage (V_t) model parameter is proposed. *One- and two-iteration* for V_t model parameter extraction with optimization approach for LDD lateral diffusion and halo centroid (σ and l_μ) is suggested [18].
- 13) Narrow-width and inverse narrow-width effects are investigated by physically modeling the fringing capacitance, surface-potential barrier lowering and effective channel doping transformation with two Gaussian doping profiles along the width dimension [19].
- 14) A unified geometry-dependent drain current model for the entire range of drawn length and width without binning is investigated. A unified model is proposed by

taking into consideration all important short-channel/narrow-width effects, including width-dependent vertical-field mobility, bias-dependent source/drain series resistance, velocity saturation/overshoot, as well as layout-dependent effect for narrow-width MOSFET is investigated [20].

15) Full scale extraction methodology for geometry/bias dependence unified drain current model parameter based on technology characterization for threshold voltage and device characterization for drain current is proposed with systematic *step-by-step* parameter extraction procedure [7].

16) The complete analytical AC and DC models for non-pinned surface-potential-based unified regional charge model and compact threshold-voltage-based unified drain current with model parameter extraction procedure and optimization methodology are implemented in MATLAB environment. The device performance fluctuation on physical parameter variation is investigated.

1.4 Organization of the Thesis

The thesis consists of eight chapters. Chapter 1 introduces the motivation and the objective of this work, as well as the main contributions from the research and the organization of the thesis.

Chapter 2 presents the background of the MOSFET modeling, and the approach used to achieve the objective mentioned in Chapter 1. Chapter 3 focuses on surface-potential modeling. An explicit unified regional surface-potential calculation algorithm is developed. Chapter 4 describes AC intrinsic charge modeling with polycrystalline silicon and quantum-mechanical effects valid in all regions of operation, including poly-

gate accumulation, depletion and strong inversion regions. Detailed study on charge conservation at flat-band condition is presented in this chapter.

Chapter 5 explains geometry/bias-dependent threshold voltage equation with detailed modeling for short-channel, reverse short-channel, narrow-width, and inverse narrow-width effects. Chapter 6 describes a symmetric unified bulk-referred drain current equation for short/narrow channel device valid in all regions of operation for the geometry/bias-dependent threshold voltage model,

Chapter 7 is devoted to AC and DC model parameter extraction procedure discussed in the previous chapters. A *step-by-step* parameter extraction for AC and DC models is illustrated. Chapter 8 verifies the model formulation with numerical data from Medici for AC modeling with physical parameter and bias variations and 0.11- μm CMOS technology measurement data for DC modeling with geometry/bias variations. Summaries and conclusions for AC/DC modeling approach, as well as suggestions and recommendations for future investigation are presented in Chapter 9.

CHAPTER 2: BACKGROUND AND METHODOLOGY

2.1 Background

The demand for compact electronic products with multi-functions in daily life is tremendously increased. Integrating hundreds of millions of working transistors into the same chip becomes the challenge for technology developers to make ever shorter and narrower working transistors. In order to meet the demand, not only tremendous efforts and studies have to be carried out to enhance the existing fabrication process, the accuracy of the compact model used in the circuit simulator to simulate electrical characteristics of a real transistor, which is connected together with millions of transistors in circuitry, is critical for predicting the fabricated device performance.

In real life, a physical model that is derived without taking any approximations, yet can still simulate MOSFET electrical characteristics that scale with bias and geometry, and predict the fabrication process parameter variation, is not available. However, process simulation tools, such as Technology CAD (TCAD), which allow technology developers to virtually fabricate a working transistor from hundreds or thousands of steps of a fabrication process by solving the implicit 2-D device equations, are commonly used. Due to the requirement in solving numerical integration, they are computation inefficient and not suitable to be used in any circuit simulator.

Compact model consists of mathematical expression as a function of the four terminal biases of a MOSFET, V_g , V_d , V_s and V_b at the gate, drain, source and bulk, respectively, used in the circuit simulator to simulate electrical characteristics of a transistor. Instead of using 2-D numerical simulator in circuit design, compact model,

which further approximates the governing equations and geometry dependence, can meet the speed required in large circuit simulation.

However, conventional long-channel compact models with ideal textbook equations are not sufficient to simulate MOSFET electrical characteristics at decreasing channel length and width. Geometry effects have to be physically taken into account for deriving compact model, rather than empirically scaling the model parameters with geometry. Enhancement on compact model development from existing ideal long-channel model is necessary such that the electrical characteristics due to physical parameter variation and geometry dependence can be take into consideration for the advance circuit simulators.

In this research work, an AC/DC compact model for circuit designers and technology developers is formulated. The model is physically derived, which allows physical parameter variation such as oxide thickness, channel/poly-gate doping concentration, dielectric constant, and quantum-mechanical parameter with geometry and bias scalability. The model passes symmetry test and conserves charge neutrality with smooth transition across different regions of operation, as well as ease of the model parameter characterization and optimization with minimum measurement data requirement.

2.2 Modeling Methodology

The key to MOSFET operation is the thin layer of charge induced along the channel. The amount of induced charge depends on external terminal voltages. In circuit simulator, four terminal voltages are usually the control parameter in simulating MOSFET electrical characteristics. In MOSFET modeling, external voltages are input

parameters, which change physically the surface-potential, charges and threshold voltage of a transistor.

AC/DC models in the literature are normally grouped into surface-potential (ϕ_s)-based, charge-based, and threshold-voltage (V_t)-based (or the pinned- ϕ_s -based), which use different approximations and methodologies for model calibration to the experimental data. Generally in intrinsic AC modeling, capacitances are related to derivatives of the total integrated terminal charges in the gate, drain, source and bulk terminals, Q_g , Q_d , Q_s and Q_b , respectively. Total charges can be physically derived by integrating the position-dependent charge densities, $Q_g(y)$ and $Q_b(y)$ along the channel, while the drain and source charges, Q_d and Q_s , respectively, can be derived from a “linear partitioning” approach proposed by Ward-Dutton [21]. The inversion charge is defined as the total drain and source charges or similarly can be integrated from position-dependent variable, $Q_i(y)$, which is the sum of $Q_d(y)$ and $Q_s(y)$ along the channel.

The AC modeling methodology is based on unified regional approach to MOSFET compact modeling. The piecewise regional charge model can be formulated from regional surface-potential equations derived analytically from regional Pao-Sah model and unified with smoothing functions. The key to the unified regional approach is to tune the model near transition regions. At the regions of transition, from accumulation to depletion (with smoothing parameters σ_a for channel surface-potential, ϕ_s) and from depletion to strong inversion (with smoothing parameters δ_ϕ and σ_{eff}), smoothing parameters are carefully extracted such that the model will not produce unphysical behaviors, such as discontinuities or kink effects. Beside this, careful tuning on smoothing parameters (σ_f for ϕ_s , ρ_r and ρ_f for poly-gate surface-potential, ϕ_p) is required to satisfy charge conservation for all operating regions, including the flat-band condition.

Coupled polycrystalline silicon (with poly-gate concentration, N_{gate}) and quantum-mechanical (with quantum factor, κ^{qm} , and smoothing parameters, σ^{qm} , δ^{qm} and ε^{qm}) effects in AC modeling are physically derived for all regions through a unified regional formulation. A unified regional AC model is formulated in such a way that the 3 physical parameters (channel doping concentration, N_{ch} , N_{gate} , κ^{qm}) and 9 smoothing parameters (σ_a , σ_f , σ_{eff} , δ_ϕ , ρ_r , ρ_f , σ^{qm} , δ^{qm} and ε^{qm}) follow a unique, *one-iteration*, 10-step (4/6-step for physical/model parameter) extraction methodology. The physical and model parameters are uncorrelated and can be extracted through a *step-by-step* extraction procedure.

Generally in DC modeling, induced current flows along the channel, from drain to source is associated with electrons. Electrons can diffuse and drift along the channel that is physically described by transport, continuity and Poisson's equations [22]-[25]. Total integrated drain current is the sum of integrated drift and diffusion components, which is associated with induced inversion layer along the channel Q_i due to free electrons contribution to the net space charge density. Total integrated drain current can be expressed as a function of threshold voltage or surface potential at the drain and source ends. In this work, threshold-voltage based drain current approach is presented, which is more correlate to technology characterization of MOSFETs and suitable for circuit designers.

For drain current that is derived based on threshold voltage definition, in which the surface potential is “pinned” at $2\phi_f$ in strong inversion, is originated from Sah's equation [26]. Due to the model compactness, physics-based regional approach, flexible and relatively easy to add new features to keep up with technology advancements, the model has been the industrial standard for decades (BSIM model since 1987). However,

increasing the total number of fitting parameters to the model (including “binning”) makes it more and more empirical and unmanageable. Together with the inherent problems in model symmetry and discontinuity, all these inadequacies have made the V_t -based models less attractive, even a “showstopper for the circuit design”.

In fact, asymmetry in the V_t -based model is not inherited by nature but if mobility, effective field and velocity saturation are not handled properly, other modeling approaches can also suffer from model asymmetry, no matter it is formulated by construction or postprocessing (swapping source/drain terminals during model evaluation) [13][27]. As a result, we developed a “non-binnable” compact drain current model based on unified regional V_t -based approach, with proper handling in symmetry and continuity. Bulk-referenced (B-ref) V_t model is formulated from ideal long-channel device and characterized with short/narrow channel effects, in which short-channel, reverse short-channel, narrow-width and inverse narrow-width effects are added in modular sequence with physical description of its own fitting parameters. These effects give physical links to model geometry/bias dependence with V_t model, so that less effort is required to model device electrical characteristics with geometry/bias dependence in the drain current model.

Our unified drain-current (I_{ds}) model has an extremely compact form. It is a single-piece expression and resembles simple textbook long/wide-channel equation with major extrinsic, series source/drain resistance, (R_{sd}) and intrinsic, velocity-overshoot with channel-length modulation, (CLM) for short-channel effects, intrinsic Coulombic mobility correction and extrinsic dog-bone layout for narrow-width effects modeling.

Our unified DC model is formulated in such a way that it requires minimum set of measurement data (using only 6 $I - V$ curves, 6 points (at W_∞) $V_t(L) - L_g$, 2 points (at L_∞

and L_{min}) $V_t(W) - W$ plus points $I_{dsat}(W)$ and $g_{ds}(L)$ measurement data) for complete DC model parameter extraction with a 10/10-step $V_t/I - V$ procedures.

In this thesis, AC model is verified with data from TCAD simulation for all bias dependence and limited experimental data for V_{bs} dependence. On DC model validation, a full set of 0.11- μm CMOS STI technology wafer with drawn gate lengths/widths (L_{drawn}/W_{drawn}) from 10- μm down to 0.13- μm experimental data in this thesis are used for development and validation. Our unified I_{ds} model has also been validated with experimental 0.18- μm CMOS STI technology wafer [28] and NMOS structures from process simulator (Tsuprem4) [9] based on the experimental 90-nm CMOS technology developed in Chartered Semiconductor Manufacturing Ltd. (CSM), Singapore.

CHAPTER 3: SURFACE POTENTIAL MODELING

Surface-potential (ϕ_s) solutions in a MOSFET have been established almost 40 years ago from the well-known Pao-Sah equation [29], but the exact solution is only available from numerical iteration. Although with increased computing speed, implicit or explicit ϕ_s solutions for all regions of operation can be obtained much easier, high accuracy in the ϕ_s solutions is required for ϕ_s -based charge/current formulation, which involves the average and difference of surface potentials at the source and drain ends. It is nontrivial to have a single-piece integrated charge for all regions of operation across flat-band (from accumulation to depletion) but very often, mathematical conditioning [30] (modification of the Pao-Sah model) has to be employed to handle ϕ_s solutions at/near the flat-band condition.

One way to solve Pao-Sah equation numerically for ϕ_s is through the Newton-Raphson [25] method. The closed form approximated explicit ϕ_s expressions have been developed [31], [32] and [33]. It is assumed that there is no voltage drop across poly gate (ϕ_p) in these models. In the real device, however, significant voltage drop across poly gate is observed [34] in the n^+/p^+ dual gate CMOS at low poly gate doping, which cannot be ignored. In a recent study, implicit poly gate doping effect has been formulated [3] within normal operating regions, where two coupled equations (ϕ_s and ϕ_p) have to be solved iteratively. This is computation inefficient. One can derive a regional poly-depletion model [3], [35]-[37] valid from depletion to strong inversion regions by assuming “pinned” $\phi_p = 0$ in accumulation region, resulting *only* one implicit ϕ_s equation that can be solved iteratively or explicitly. These models may have difficulty to predict

poly-doping effect in accumulation region (poly-accumulation effect, PAE) and limits its prediction *only* within normal operating regions. Moreover, it may also result in wrong prediction for “gate capacitance-recovery” effect [38] (poly-inversion effect, PIE) beyond normal operating condition.

In this work, a single-piece full region explicit surface-potential (ϕ_{seff} , ϕ_{peff}) -based on “non-pinned” unified regional surface potential approach is presented in the following sections. The following ϕ_{seff} and ϕ_{peff} expressions give continuous solutions of its functional and higher-order derivatives characteristics, which are tuned to zero at the flat-band condition. Thus, with oxide charge density, Q_{ox} , total charge is conserved in all regions even at the flat-band condition. Before detail discussion on charge modeling methodology, implicit, explicit unified regional ϕ_s and ϕ_p solutions with coupled quantum-mechanical effect (QME) are presented in the following sections.

3.1 Channel Surface Potential Calculation

In this section, implicit/explicit and unified regional ϕ_s solutions are derived. Analytical ϕ_s solutions are regionally approximated (by ignoring ϕ_p as in the conventional approximation), in which three explicit regional ϕ_s models are formulated. Regional ϕ_s models are unified with smoothing functions, which give asymptotic physical ϕ_s solutions from iterative Pao-Sah voltage equation that become the key features in unified regional charge modeling approach. A single-piece explicit ϕ_{seff} model can also be formulated from the unified regional ϕ_s solution and is presented in the following.

3.1.1 Conventional implicit surface potential derivation

Surface potential at the oxide/channel interface is the important parameter to determine the induced channel charge. It is governed by the *potential balance*:

$$V_{gb} - \phi_{MS} - \phi_s - \phi_p - V_{ox} = 0, \quad (3.1.1)$$

where $V_{ox} = Q_g/C_{ox}$ is the voltage drop across the gate oxide thickness t_{ox} associated with $C_{ox} = \epsilon_{ox}/t_{ox}$, and Q_g is the gate charge (per unit area). The work function difference between the poly gate and silicon channel

$$\phi_{MS} = (W_M - W_S)/q = -v_{th} \ln(N_{gate}N_{ch}/n_i^2) \quad (3.1.1a)$$

is determined by the n^+ poly-gate doping, N_{gate} ($= N_d$, donor), and p channel doping, N_{ch} ($= N_a$, Acceptor) assuming that the poly gate and the silicon substrate have the same affinity and bandgap. In Eq. (3.1.1a), v_{th} ($= kT/q$) is the thermal voltage and n_i is the intrinsic carrier concentration. *Charge balance* requires that

$$Q_g + Q_{sc} + Q_{ox} = 0, \quad (3.1.2)$$

where Q_{ox} is the effective fixed oxide charge density. Combining (3.1.1) and (3.1.2), we obtain,

$$V_{gb} - V_{fb} - \phi_s - \phi_p = -Q_{sc}/C_{ox} \quad (3.1.3)$$

where

$$V_{fb} = \phi_{MS} - Q_{ox}/C_{ox} \quad (3.1.3a)$$

is the flat-band voltage. The total induced channel charge (per unit area) in the channel can be determined by applying the *Gauss' law*

$$Q_{sc} = -\epsilon_{si} E_{si} = \epsilon_{si} \left. \frac{d\phi(x)}{dx} \right|_{x=0}, \quad (3.1.4)$$

in which E_{si} is the electric field at the interface. The potential distribution is determined by the *Poisson's equation*:

$$\frac{d^2\phi(x)}{dx^2} = -\frac{\rho}{\epsilon_{si}}, \quad (3.1.5)$$

where

$$\rho = q[p - n - N_a + N_d], \quad (3.1.5a)$$

is the net space charge density (C/cm³) in the space charge region at room temperature,

$$N_d = N_a \exp\left[\frac{-2\phi_f}{v_{th}}\right], \quad (3.1.5b)$$

$$p = N_a \exp\left[\frac{-\phi(x)}{v_{th}}\right], \quad (3.1.5c)$$

and

$$n = N_a \exp\left[\frac{\phi(x) - V_{cb} - 2\phi_f}{v_{th}}\right] \quad (3.1.5d)$$

are the concentrations of ionized donors, free holes and electrons, respectively. N_a is the concentrations of ionized acceptors, V_{cb} is the position-dependent channel-to-bulk voltage and ϕ_f is the bulk Fermi potential. Solving (3.1.5) for potential gradient ($d\phi(x)/dy$) at the oxide/channel interface with boundary conditions ($\phi_s = \phi(x)|_{x=0}$, $d\phi(x)/dx|_{x=X_d} = 0$ and $\phi(x)|_{x=X_d} = 0$) where X_d is the channel depletion width, from (3.1.4), the total induced charge in the channel can be obtained

$$Q_{sc} = -C_{ox}\gamma \operatorname{sgn}(\phi_s)\sqrt{f_\phi}, \quad (3.1.6)$$

where f_ϕ and γ , respectively, are defined by

$$f_\phi = v_{th} \left(\exp \left[\frac{-\phi_s}{v_{th}} \right] - 1 \right) + \phi_s \left(1 - \exp \left[\frac{-2\phi_f}{v_{th}} \right] \right) + v_{th} \exp \left[-\frac{V_{cb} + 2\phi_f}{v_{th}} \right] \left(\exp \left[\frac{\phi_s}{v_{th}} \right] - 1 \right), \quad (3.1.6a)$$

$$\gamma = \sqrt{2q\epsilon_{si}N_{ch}} / C_{ox}, \quad (3.1.6b)$$

where $\text{sgn}(\phi_s) = \pm 1$ represents the sign of ϕ_s . Combining Eq. (3.1.3) with (3.1.6) yields

$$V_{gb} - V_{fb} - \phi_s - \phi_p = \gamma \text{sgn}[\phi_s] \sqrt{f_\phi}. \quad (3.1.7)$$

In the above equation, the potential contribution from holes, acceptors and donors, and electrons are described by the first, second and third terms, respectively. Channel surface potential, ϕ_s can be calculated iteratively with *one* implicit equation by setting $\phi_p = 0$ (conventional approach, Pao-Sah model). To include poly-doping effect, either an explicit ϕ_p solution or another implicit equation for ϕ_p has to be solved together with ϕ_s and this modeling methodology may not be practical for circuit simulation.

An explicit regional ϕ_s expression can be formulated analytically by taking similar approach as stated above with only the respective space charge contribution in the respective region of operation, as will be illustrated in the following section. With the explicit regional ϕ_s solution, it will be easier to include poly silicon and quantum-mechanical effects into the unified regional ϕ_s model and, hence, both effects can be decoupled and accurately modeled through our unified regional charge modeling methodology, as described in the following chapter.

3.1.2 Explicit piecewise surface potential derivation

Conventional implicit ϕ_s equation can be formulated from (3.1.7) by dropping the ϕ_p term in the left-hand side of the equation. If the device is biased in the accumulation region ($V_{gb} < V_{fb}$), the net space-charge density in the space charge region will be dominated by free holes, so the charge density can be approximated as $\rho \approx q(p)$ in (3.1.5a) and $f_\phi \approx v_{th} \exp[-\phi_s/v_{th}]$ in (3.1.6a), yielding

$$V_{gb} - V_{fb} - \phi_{CC} = -\gamma \sqrt{v_{th} \exp\left[\frac{-\phi_{CC}}{v_{th}}\right]}. \quad (3.1.8)$$

Setting $\phi_s = \phi_{CC}$ in accumulation, Eq. (3.1.8) can be re-written as

$$\exp(X) + aX + B = 0 \quad (3.1.8a)$$

$$X = \frac{-\phi_{CC}}{2v_{th}}, \quad a = \frac{2\sqrt{v_{th}}}{\gamma}, \quad B = \frac{V_{gb} - V_{fb}}{\gamma\sqrt{v_{th}}}. \quad (3.1.8b)$$

Eq. (3.1.8a) can be solved analytically [39] with the well-known *Lambert W* function, $\mathcal{L}\{W\}$ [32], so that

$$\phi_{CC} = -2v_{th} \left[-\frac{a\mathcal{L}\{W\} + B}{a} \right] = V_{gb} - V_{fb} + 2v_{th}\mathcal{L}\{W\} \quad (3.1.9)$$

where

$$\begin{aligned} W &= \frac{1}{a} \exp\left(-\frac{B}{a}\right) \\ &= \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(-\frac{V_{gb} - V_{fb}}{2v_{th}}\right). \end{aligned} \quad (3.1.9a)$$

If the device is biased in the depletion region, similar approximation results in $\rho \approx q(-N_a)$ in (3.1.5a) and $f_\phi \approx \phi_s$ in (3.1.6a), in which the net space charge region is

dominated by fully ionized acceptors, leading to an approximate (3.1.7) in the depletion region

$$V_{gb} - V_{fb} - \phi_{DD} = \gamma \sqrt{\phi_{DD}}. \quad (3.1.10)$$

The above equation can be readily solved:

$$\phi_{DD} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gb} - V_{fb}} \right]^2. \quad (3.1.11)$$

Similarly in the strong inversion region, space charge region is dominated by both free electrons and fully ionized acceptors so that $\rho \approx -q N_a \{ \exp[(\psi - V_{cb} - 2\phi_f)/v_{th}] + N_a \}$ in (3.1.5a) and $f_\phi \approx \{ v_{th} \exp[-(V_{cb} + 2\phi_f)/v_{th}] [\exp(\phi_s/v_{th}) - 1] + \phi_s \}$ in (3.1.6a). From both approximations, (3.1.7) is re-written

$$V_{gb} - V_{fb} - \phi_{SS} = \gamma \sqrt{\phi_{SS} + v_{th} \exp\left[-\frac{V_{cb} + 2\phi_f}{v_{th}}\right] \left(\exp\left[\frac{\phi_{SS}}{v_{th}}\right] - 1 \right)}. \quad (3.1.12)$$

This equation is only valid in strong inversion region. Approximating $\phi_{SS} \approx \phi_B = \phi_{s0} = 2\phi_f$ inside the radical, solving for ϕ_{SS} , one obtains [33]

$$\phi_{SS} = \phi_{s0} + \Delta_{SS} + V_{cb} \quad (3.1.13)$$

in which $\phi_{s0} = 2\phi_f$ is the channel “pinned” surface potential and the “non-pinned” increment of the surface potential in strong-inversion region is given by [33]

$$\Delta_{SS} = v_{th} \ln \left\{ \frac{1}{v_{th}} \left[\left(\frac{V_{gb} - V_{fb} - \phi_{SS}^*}{\gamma} \right)^2 - (\phi_{s0} + V_{cb}) \right] + 1 \right\} \quad (3.1.13a)$$

$$\phi_{SS}^* = \phi_{s0} + V_{cb} + \frac{\phi_{DD} - (\phi_{s0} + V_{cb})}{\sqrt{1 + \left[\frac{\phi_{DD} - (\phi_{s0} + V_{cb})}{4v_{th}} \right]^2}}. \quad (3.1.13b)$$

Figure 3.1: Explicit regional ϕ_s solutions.

Figure 3.1 shows three explicit regional ϕ_s solutions in accumulation (ϕ_{CC} , solid line), depletion (ϕ_{DD} , dotted line), and strong inversion (ϕ_{SS} , dashed line) regions, which asymptotically approach iterative ϕ_s solutions (symbol). Imaginary ϕ_s solutions (dotted and dashed lines below flat-band and threshold voltages, respectively) and wrong ϕ_s solutions (solid line above flat-band) result when the regional solution is outside its validity region.

3.1.3 Unified regional channel surface potential modeling

In order to extend the explicit regional models beyond their validity boundaries (no imaginary values with correct asymptotic behaviors) without any “if” statement in ϕ_s calculation, the unified regional approach is adopted. In the accumulation region, an explicit regional ϕ_{CC} model is formulated that approaches asymptotic iterative ϕ_s solutions. However, when gate voltage is biased above flat band condition, ϕ_{CC} model over predicts iterative ϕ_s solutions, in which it approaches $V_{gb} - V_{fb}$ (dashed line, figure 3.2). A unified regional model

$$\phi_{ACC} = \phi_{CC} \Big|_{V_{gb}-V_{fb}=V_{GBR}} = V_{GBR} + 2v_{th}\mathcal{L}\{W\}, \quad (3.1.14)$$

$$V_{GBR} = \mathcal{G}_r \{V_{gb} - V_{fb}; \sigma_a\}, \quad (3.1.14a)$$

with the “reverse” smoothing function

$$\mathcal{G}_r \{x; \sigma\} \equiv \left\{ x - \sqrt{x^2 + 4\sigma} \right\} / 2, \quad (3.1.14b)$$

is proposed. ϕ_{ACC} preserves the asymptotic iterative ϕ_s solution in the accumulation region but approaches zero beyond V_{fb} (solid line, figure 3.2). In the depletion region, a modified expression is constructed by

$$\phi_{SUB} = \phi_{DD} \Big|_{V_{gb}-V_{fb}=V_{GBF}} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{GBF}} \right]^2, \quad (3.1.15)$$

$$V_{GBF} = \mathcal{G}_f \{V_{gb} - V_{fb}; \sigma_f\} \quad (3.1.15a)$$

with the “forward” smoothing function

$$\mathcal{G}_f \{x; \sigma\} \equiv \left\{ x + \sqrt{x^2 + 4\sigma} \right\} / 2. \quad (3.1.15b)$$

ϕ_{SUB} preserves the asymptotic ϕ_s solution in the depletion region while avoiding imaginary ϕ_s solutions in the accumulation region, as shown in figure 3.3 (dashed line).

Figure 3.2: Unified regional and regional surface potential solutions at the accumulation region.

Asymptotic iterative ϕ_s solutions can be modeled with ϕ_{ACC} and ϕ_{SUB} in the accumulation and depletion regions, respectively. By adding both regional models, a single-piece solution

$$\phi_{SA} = \phi_{ACC} + \phi_{SUB} \quad (3.1.16)$$

can be obtained which preserves asymptotic behaviors in accumulation and depletion regions with a smooth transition across flat-band (solid line, figure 3.3).

Figure 3.3: Unified regional (ϕ_{SA} , ϕ_{SUB} , ϕ_{ACC}) and regional (ϕ_{DD}) surface potential solutions from accumulation to strong inversion regions.

In the strong inversion region, a “non-pinned” unified surface-potential is formulated as

$$\phi_{STR} = \phi_{s0} + \Delta + V_{cb} \quad (3.1.17)$$

with the unified surface-potential increment

$$\Delta = v_{th} \ln \left\{ \frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(\frac{V_{GA}}{\gamma} \right)^2 - (\phi_{s0} + V_{cb}); \sigma_{eff} \right\} + 1 \right\}, \quad (3.1.17a)$$

in which $V_{gb} - V_{fb}$ in Δ_{ss} is replaced by

$$V_{GA} = \mathcal{G}_f \{ V_{gb} - V_{fb} - \phi_{STR}^*; \sigma_{vga} \}, \quad (3.1.17b)$$

and ϕ_{ss}^* is replaced by

$$\phi_{STR}^* = \phi_{s0} + V_{cb} + \frac{\phi_{SUB} - (\phi_{s0} + V_{cb})}{\sqrt{1 + \left[\frac{\phi_{SUB} - (\phi_{s0} + V_{cb})}{4v_{th}} \right]^2}}. \quad (3.1.17c)$$

σ_{eff} and σ_{vgs} are smoothing parameters. Both the regional ϕ_{ss} and unified ϕ_{STR} expressions give asymptotic iterative ϕ_s solutions in the strong inversion region. However, the regional ϕ_{ss} expression produces imaginary ϕ_s solutions in the depletion and accumulation regions as shown in figure 3.4 (dotted line). While as for the unified ϕ_{STR} expression, imaginary solutions can be avoided with smoothing functions in Δ and V_{GA} to prevent negative argument in the natural log function and the solution from $(V_{gb} - V_{fb} - \phi_{STR}^*)$.

Figure 3.4: Unified regional surface potential solutions in the strong inversion region.

A single-piece unified ϕ_{SEFF} model for all operation regions

$$\phi_{SEFF} = \mathcal{G}_{eff} \{ \phi_{SA}, \phi_{STR}; \delta_\phi \} \quad (3.1.18)$$

is formulated with an interpolating function

$$\mathcal{G}_{eff} \{ x, x_{sat}; \delta \} \equiv x_{sat} - 0.5 \left[x_{sat} - x - \delta + \sqrt{(x_{sat} - x - \delta)^2 + 4\delta x_{sat}} \right] \quad (3.1.18a)$$

to join ϕ_{SA} with ϕ_{STR} smoothly without any discontinuities in full operating regions, in which δ is a smoothing parameter. Similarly, a unified single-piece ϕ_{DD} solution can be formulated

$$\phi_{DS} = \mathcal{G}_{eff} \{ \phi_{SUB}, \phi_{STR}; \delta_\phi \} \quad (3.19)$$

that approaches iterative ϕ_s solutions from depletion to strong inversion regions but approaches at zero surface potential in the accumulation region as shown in figure 3.5 (dotted line).

Figure 3.5: Unified regional ϕ_s solutions across different regions of operation.

A comparison of our single-piece unified ϕ_{SEFF} model (lines) with the iterative ϕ_s solution (symbols) for positive and negative V_{cb} (taking at the source end) is presented in figure 3.6. Good matching of both data with functional behavior across different regions of operation is observed.

Figure 3.6: Bias scalability of the explicit unified regional ϕ_{SEFF} solutions are compared to the iterative ϕ_s solutions in (3.1.7).

3.1.4 Zero surface potential at the flat-band gate voltage

Band bending in the bulk semiconductor occurs when different elements of a MOS system are brought in contact at equilibrium. This band bending is caused by the work function difference, ϕ_{MS} between the metal (poly gate) and the semiconductor given by (3.1.1a). Band bending could also be due to the effective fixed oxide charge density, Q_{ox} which results in a total band bending, V_{fb} in the channel. When external gate terminal is biased at $V_{gb} = V_{fb}$, the energy bands become flat so that $\phi_s = 0$.

In order to meet this condition, a unified regional model, ϕ_{SA} is tuned with smoothing parameter, σ_f in ϕ_{SUB} and σ_a in ϕ_{ACC} , such that $\phi_{SA} = 0$ at $V_{gb} = V_{fb}$ to ensure the smoothness of its functional and higher-order derivatives across V_{fb} . A single-piece unified ϕ_{seff} with tuned σ_f is shown in figure 3.7.

Figure 3.7: Tuning and fitting of σ_f in ϕ_{SUB} with ϕ_{SEFF} solutions to meet zero surface-potential at flat-band condition.

3.1.5 Continuity in higher-order numerical derivatives

Higher-order derivatives of the ϕ_{SEFF} solutions can be numerically computed and compared to those of the implicit ϕ_s solutions, as shown in figure 3.8. Excellent matching

of data is observed with smooth transition across different regions of operating as shown with left-axis for $d\phi_{SEFF}/dV_{gb}$ and right-axis for $d^2\phi_{SEFF}/dV_{gb}^2$.

Figure 3.8: Accurate matching and continuous characteristics for the first/second-order derivatives of surface-potential solutions.

3.2 Calculation of the Poly Gate Surface Potential Solution

In the previous sections of this chapter, one voltage equation (Pao-Sah) to solve ϕ_s iteratively or a single-piece ϕ_{SEFF} to calculate ϕ_s solutions have been demonstrated and validated. Similarly, at the gate/oxide interface, if the device is fabricated with advanced n^+ implanted poly gate n -channel MOSFETs (n^+/p^+ dual-gate CMOS technology), free carriers in the gate space charge region at the gate/oxide interface cannot be neglected. A voltage drop (ϕ_p) across the induced-charge layer at the gate/oxide interface will affect device electrical characteristics, which is dependent on gate doping concentration. The significant amount of voltage drop (ϕ_p) must be modeled by another implicit surface-potential equation at gate/oxide interface. Both the coupled ϕ_s and ϕ_p implicit equations have to be solved iteratively, which is very computationally expensive and might be impractical in circuit simulation.

In the following sections, an implicit ϕ_p equation is derived and the analytical approximation is presented for regional ϕ_p -solutions. A unified regional approach is adopted in this work to unify regional model equations, which is extremely useful in our intrinsic charge formulation with poly-silicon effect in all regions of operation.

3.2.1 Implicit poly-gate surface potential solution

Similar to the channel surface-potential solutions at the oxide/silicon interface, the induced gate charge at the gate/oxide interface with n^+ poly doping can be formulated from the *Gauss' law*, applied to the gate/oxide interface:

$$Q_g = \epsilon_{si} E_{si,p} = \epsilon_{si} \left. \frac{d\phi(x)}{dx} \right|_{x=-t_{ox}}, \quad (3.2.1)$$

in which $E_{si,p}$ is the potential gradient at the gate interface, The *Poisson's equation* is given by

$$\frac{d^2\phi(x)}{dx^2} = -\frac{\rho}{\epsilon_{si}} \quad (3.2.2)$$

where the net space charge density at the gate interface

$$\rho = q[p - n - N_a + N_d] \quad (3.2.2a)$$

$$\text{in which} \quad (3.2.2b)$$

$$p = N_d \exp\left[-\frac{\phi(x) - 2\phi_{f,p}}{v_{th}}\right], \quad (3.2.2c)$$

and

$$n = N_d \exp\left[\frac{\phi(x)}{v_{th}}\right] \quad (3.2.2d)$$

are the concentrations of ionized acceptors, free holes and electrons, respectively, and $\phi_{f,p}$ is the Fermi level in the gate. Integrating the Poisson's equation (3.2.2) with the boundary conditions, $\phi_p = \phi(x)|_{x=-t_{ox}}$, $d\phi(x)/dx|_{x=-(t_{ox}+X_{pd})} = 0$ and $\phi(x)|_{x=-(t_{ox}+X_{pd})} = 0$, where X_{pd} is the gate depletion width, and combining with Eq. (3.2.1), the total induced gate charge can be obtained

$$Q_g = C_{ox}\gamma_p \operatorname{sgn}(\phi_p) \sqrt{f_{\phi,p}}, \quad (3.2.3)$$

$$f_{\phi,p} = v_{th} \left(\exp \left[\frac{-\phi_p}{v_{th}} \right] - 1 \right) + \phi_p \left(1 - \exp \left[\frac{-2\phi_{f,p}}{v_{th}} \right] \right) + v_{th} \exp \left[-\frac{2\phi_{f,p}}{v_{th}} \right] \left(\exp \left[\frac{\phi_p}{v_{th}} \right] - 1 \right), \quad (3.2.3a)$$

$$\gamma_p = \sqrt{2q\epsilon_{si}N_{gate}} / C_{ox}, \quad (3.2.3b)$$

where $\operatorname{sgn}(\phi_p) = \pm 1$ represents the sign of ϕ_p . The gate charge contributions from electrons, donors and acceptors, and holes are denoted by the first, second and third terms in (3.2.3a), respectively. Substituting Eq. (3.1.3) into (3.1.2) and combining with Eq. (3.2.3), one obtains the implicit ϕ_p equation:

$$V_{gb} - V_{fb} - \phi_s - \phi_p = (Q_g + Q_{ox}) / C_{ox}. \quad (3.2.4)$$

This equation has to be solved iteratively for ϕ_p .

3.2.2 Explicit piecewise gate surface potential derivation

Existing models ignore the inversion charge in poly-silicon gate or poly-inversion effect (PIE), and assume $\phi_p = 0$ in accumulation i.e., no poly-accumulation effect (PAE), so that the free carriers in the net space charge region in (3.2.2a), $\rho \approx qN_d$ is assumed solely dominated by ionized donors. This results in the reduced form of (3.2.4),

$$V_{gb} - V_{fb} - \phi_s - \phi_p = \gamma_p \sqrt{\phi_p}, \quad (3.2.5)$$

where the regional poly-depletion effect (PDE) [35], [36]

$$\phi_p = \phi_{p,dd} = \left(-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gb} - V_{fb} - \phi_s} \right)^2, \quad (3.2.6)$$

is formulated from depletion to strong inversion regions, $V_{fb} < V_{gb} < V_{rec}$. V_{rec} is the “gate capacitance recovery” voltage [37], [38]. The gate capacitance “cross-over” [40] behavior cannot be modeled correctly by setting $\phi_p \approx 0$ in the accumulation region. This is because the contribution from accumulated electrons cannot be ignored at the gate interface when the device is operated in the accumulation mode. The free carriers in the space-charge region at the gate interface is solely dominated by the accumulated free electrons while as in the channel oxide/silicon interface, they are dominated by the accumulated free holes, which will lead to

$$f_{\phi,p} \approx v_{th} \left(\exp \left[\frac{-\phi_p}{v_{th}} \right] - 1 \right) \quad (3.2.7)$$

from (3.2.3a) at the gate interface and

$$f_{\phi} \approx v_{th} \left(\exp \left[\frac{-\phi_s}{v_{th}} \right] - 1 \right) \quad (3.2.8)$$

from (3.1.6) at the channel interface. Substituting (3.2.8) into (3.1.6) for the channel charge Q_{sc} and using (3.2.7) in (3.2.3) for the gate charge Q_g , from the charge balance equation (3.1.2); one obtains (by ignoring Q_{ox})

$$\gamma \operatorname{sgn}(\phi) \sqrt{v_{th} \left(\exp \left[\frac{-\phi_s}{v_{th}} \right] - 1 \right)} = \gamma_p \operatorname{sgn}(\phi_p) \sqrt{v_{th} \left(\exp \left[\frac{-\phi_p}{v_{th}} \right] - 1 \right)}, \quad (3.2.9)$$

from which the regional PAE model is obtained

$$\phi_p = \phi_{p,cc} = -v_{th} \ln \left[\frac{\gamma_p^2}{\gamma_p^2} \exp \left(-\frac{\phi_s}{v_{th}} \right) + \left(1 - \frac{\gamma_p^2}{\gamma_p^2} \right) \right] \quad (3.2.10)$$

which gives the potential drop across the induce accumulated layer at the gate interface.

An inverted layer can be induced at the gate/oxide interface if the device is biased beyond the normal operating range ($V_{gb} > V_{rec}$). Simply by extending PDE (3.2.6) to model this region is not sufficient. This is because a hole inversion layer is induced at the gate interface. Apart from donors' contribution to form a fully-depleted layer at the interface, the contribution from the inverted holes in the space charge region has to be considered. This invokes potential balance and Poisson's equation (ignoring Q_{ox}), leading to

$$V_{gb} - V_{fb} - \phi_s - \phi_p = \gamma_p \sqrt{f_{\phi,p}}, \quad (3.2.11)$$

where $f_{\phi,p}$ has been approximated as

$$f_{\phi,p} \approx \phi_p + v_{th} \exp \left[-\frac{2\phi_{f,p}}{v_{th}} \right] \left(\exp \left[\frac{\phi_p}{v_{th}} \right] - 1 \right). \quad (3.2.12)$$

To account for both inverted layers formed at the gate/oxide and oxide/channel interfaces, the “non-pinned” surface potential [11] is adopted so that the poly inversion effect (PIE) can be similarly modeled. A regional poly surface potential solution is proposed as

$$\phi_p = \phi_{p,ss} = \phi_{p0} + \Delta_{p,ss}, \quad (3.2.13)$$

where $\phi_{p0} = 2\phi_{f,p}$ is the “pinned” surface potential,

$$\Delta_{p,ss} = v_{th} \ln \left\{ \frac{1}{v_{th}} \left[\left(\frac{V_{gb} - V_{fb} - \phi_s - \phi_{p,ss}^*}{\gamma_p} \right)^2 - \phi_{p0} \right] + 1 \right\}, \quad (3.2.13a)$$

and

$$\phi_{p,ss}^* = \phi_{p0} + \frac{\phi_{p,dd} - \phi_{p0}}{\sqrt{1 + \left[\frac{\phi_{p,dd} - \phi_{p0}}{4v_{th}} \right]^2}}. \quad (3.2.13b)$$

Figure 3.9: Regional polycrystalline silicon effect (PAE/PDE/PIE) for ϕ_p solutions.

PAE/PDE/PIE models are compared with the extracted numerical data at the gate/oxide interface via solving the coupled implicit gate/channel surface-potential equations using numerical simulator, MEDICI as shown in figure 3.9. However, beyond the valid region, the calculated ϕ_p from the regional models may result in imaginary values or deviate from the implicit ϕ_p solutions. In order to avoid this, we develop a unified regional expression to explicitly model the implicit ϕ_p solution, as discussed in the following.

3.2.3 Unified regional gate surface potential modeling

Piecewise models for ϕ_p in the accumulation (PAE), depletion (PDE) and inversion (PIE) regions has been derived to give accurate regional ϕ_p solutions. However, beyond the respective validity regions, unphysical solutions may result. In

order to overcome this problem, a unified regional approach is adopted so that imaginary solutions in the regional PDE model can be avoided with a single-piece unified PDE model,

$$\phi_{p,sub} = \phi_{p,dd} \Big|_{V_{gb}-V_{fb}-\phi_s=V_{gbs}} = \left(-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gbs}} \right)^2 \quad (3.2.14)$$

where $(V_{gb} - V_{fb} - \phi_s)$ in the regional model is replaced by a forward smoothing function

$$V_{gbs} = \mathcal{G}_f \{ V_{gb} - V_{fb} - \phi_s; \rho_f \}. \quad (3.2.14a)$$

Similarly, a single-piece unified PAE model replaces the regional PAE model

$$\phi_{p,acc} = \phi_{p,cc} \Big|_{\phi_s=\phi_{sr}} = -v_{th} \ln \left[\frac{\gamma_p^2}{\gamma_p^2} \exp \left(-\frac{\phi_{sr}}{v_{th}} \right) + \left(1 - \frac{\gamma_p^2}{\gamma_p^2} \right) \right]^2 \quad (3.2.15)$$

with a reverse smoothing function

$$\phi_{sr} = \mathcal{G}_r \{ \phi_s; \rho_r \} \quad (3.2.15a)$$

so that asymptotic iterative ϕ_p solutions in the accumulation region is preserved, which approaches zero above flat-band voltage. Similarly for the regional PIE model, imaginary ϕ_p solutions may occur if the model is biased to $V_{gb} < V_{rec}$. In order to avoid this, a unified regional PIE model is formulated similar to the ϕ_{STR} modeling approach,

$$\phi_p = \phi_{p,str} = \phi_{p0} + \Delta_p, \quad (3.2.16)$$

$$\Delta_p = v_{th} \ln \left[\frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(V_{gap} / \gamma_p \right)^2 - \phi_{p0}; \rho_{eff} \right\} + 1 \right], \quad (3.2.16a)$$

$$V_{gap} = \mathcal{G}_f \{ V_{gb} - V_{fb} - \phi_s - \phi_{p,str}^*; \rho_{vga} \}, \quad (3.2.16b)$$

$$\phi_{p,str}^* = \phi_{p0} + \frac{\phi_{p,sub} - \phi_{p0}}{\sqrt{1 + \left[\frac{\phi_{p,sub} - \phi_{p0}}{4v_{th}} \right]^2}}. \quad (3.2.16c)$$

Figure 3.10: ϕ_p solutions from iterative, piecewise and unified models.

Single-piece regional and unified PAE/PDE/PIE models are shown in figure 3.10. The unified region models in the accumulation, depletion and inversion regions do avoid the imaginary solutions as calculated from regional models beyond their validity region. By adding both the unified PAE and PDE models, a smooth unified surface potential for ϕ_p can be obtained

$$\phi_{pa} = \phi_{p,acc} + \phi_{p,sub} \quad (3.2.17)$$

It is then joined with $\phi_{p,str}$ to form a single-piece explicit ϕ_{peff} model

$$\phi_{peff} = \mathcal{G}_{eff} \left\{ \phi_{pa}, \phi_{p,str}; \delta_p \right\}, \quad (3.2.18)$$

where δ_p is a smoothing parameter.

Figure 3.11: Single-piece ϕ_{peff} solutions with model components and derivatives (inset).

Figure 3.11 shows the single-piece ϕ_{peff} solutions together with derivatives (inset). Continuous ϕ_{peff} solutions are observed. Accuracy in both the function and derivative are maintained.

3.2.4 Zero gate surface potential at flat-band gate voltage

When V_{gb} is biased exactly at flat-band voltage, similar to the band bending at the channel, bands at gate/oxide interface become flat. At this condition, both ϕ_s and ϕ_p are “pinned” to zero ($\phi_s = 0$, $\phi_p = 0$). In order to meet this condition, smoothing parameters (ρ_f in $\phi_{p,sub}$ and ρ_r in $\phi_{p,acc}$) are fitted at $V_{gb} = V_{fb}$ to maintain $\phi_p = 0$ for various N_{gate} values. With the same set of ρ_f and ρ_r , ϕ_{peff} will always be zero at flat-band as shown in figure 3.12.

Figure 3.12: Zero potential at flat-band condition with explicit ϕ_{peff} for four poly-gate doping concentrations with the same set of ρ_f and ρ_r .

3.3 Channel surface potential with explicit single-piece ϕ_{peff} solutions

Conventionally, ϕ_s can be obtained by solving one implicit equation iteratively with ϕ_p being ignored. The corresponding regional and unified models have been formulated and discussed in the previous section by setting $\phi_p = 0$ for all regions. When the device is fabricated with n^+/p^+ dual gate CMOS technology, its electrical characteristic is highly dependent on the poly gate doping concentration (N_{gate}). In this section, a unified ϕ_{seff} model is re-formulated with non-zero ϕ_p and the influence of ϕ_p on ϕ_{seff} is investigated.

Without ignoring ϕ_p across the induced layer at the gate interface, the equation for ϕ_s in accumulation can be approximated as

$$V_{gb} - V_{fb} - \phi_{cc} - \phi_p = -\gamma \sqrt{v_{th} \exp \left[\frac{-\phi_{cc}}{v_{th}} \right]}, \quad (3.3.1)$$

and can be re-written in the exponential form as

$$\exp(x) + ax + b = 0 \quad (3.3.1a)$$

$$x = \frac{-\phi_{cc}}{2v_{th}}, \quad b = \frac{V_{gb} - V_{fb} - \phi_p}{\gamma\sqrt{v_{th}}}. \quad (3.3.1b)$$

Eq. (3.3.1) can be solved explicitly for ϕ_{cc} with the well-known *Lambert W* function, $\mathcal{L}\{w\}$,

$$\phi_{cc} = -2v_{th} \left[-\frac{a\mathcal{L}\{w\} + b}{a} \right] = V_{gb} - V_{fb} - \phi_p + 2v_{th}\mathcal{L}\{w\} \quad (3.3.2)$$

where

$$\begin{aligned} w &= \frac{1}{a} \exp\left(-\frac{b}{a}\right) \\ &= \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(-\frac{V_{gb} - V_{fb} - \phi_p}{2v_{th}}\right). \end{aligned} \quad (3.3.2a)$$

A unified regional model is proposed in the accumulation region as

$$\phi_{acc} = V_{gbr} + 2v_{th}\mathcal{L}\{w\}, \quad (3.3.3)$$

with a reverse smoothing function

$$V_{gbr} = \mathcal{G}_r\{V_{gb} - V_{fb} - \phi_p; \sigma_a\}. \quad (3.3.3a)$$

The correlation of ϕ_p potential drop with the unified ϕ_{acc} model is investigated and as shown in figure 3.13. No significant change is observed with and without ϕ_p potential drop for the ϕ_{acc} solutions. However, significant changes in functional and derivative of the *Lambert-W* solutions can be observed with the ϕ_p potential drop in the accumulation region. This (PAE) becomes important in formulating unified regional charge model, which has been ignored in the literature.

Figure 3.13: A unified regional channel surface potential, ϕ_{acc} and the Lambert-w function, $\mathcal{L}\{w\}$, with (solid lines) and without (dotted lines) poly-accumulation effect (PAE) on the left-axis and the right-axis for their derivative.

Similarly, in the depletion region, ϕ_{DD} is formulated with non-zero ϕ_p potential drop across the induced layer at the gate/oxide interface, in which the fully ionized acceptors in the space charge region is considered, which gives

$$V_{gb} - V_{fb} - \phi_{dd} - \phi_p = \gamma \sqrt{\phi_{dd}}. \quad (3.3.4)$$

Solving the above equation one obtains

$$\phi_{dd} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gb} - V_{fb} - \phi_p} \right]^2. \quad (3.3.5)$$

A unified regional model in the depletion region can be formulated as

$$\phi_{sub} = \phi_{dd} \Big|_{V_{gb} - V_{fb} - \phi_p = V_{gbf}} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gbf}} \right]^2 \quad (3.3.6)$$

by replacing $(V_{gb} - V_{fb} - \phi_p)$ in the regional ϕ_{dd} model with a forward smoothing function

$$V_{gbf} = \mathcal{G}_f \{ V_{gb} - V_{fb} - \phi_p; \sigma_f \}. \quad (3.3.6a)$$

A single-piece unified regional model that preserves asymptotic iterative ϕ_s solutions from accumulation to depletion regions across flat-band can be formulated from the unified ϕ_{sub} and ϕ_{acc} models

$$\phi_{sa} = \phi_{acc} + \phi_{sub}. \quad (3.3.7)$$

In the strong inversion region, the fully ionized acceptors and free electrons are dominant in the channel space charge region, which results in an approximated implicit ϕ_p expression at the gate interface

$$V_{gb} - V_{fb} - \phi_{ss} - \phi_p = \gamma \sqrt{\phi_{ss} + v_{th} \exp\left[-\frac{V_{cb} + 2\phi_f}{v_{th}}\right] \exp\left[\frac{\phi_{ss}}{v_{th}}\right]}. \quad (3.3.8)$$

Regional ϕ_{ss} model can be formulated

$$\phi_{ss} = \phi_{s0} + \Delta_{ss} + V_{cb}, \quad (3.3.9)$$

in which the regional “non-pinned” surface potential increment is given by

$$\Delta_{ss} = v_{th} \ln \left\{ \frac{1}{v_{th}} \left[\left(\frac{V_{gb} - V_{fb} - \phi_p - \phi_{ss}^*}{\gamma} \right)^2 - (\phi_{s0} + V_{cb}) \right] + 1 \right\}, \quad (3.3.9a)$$

$$\phi_{ss}^* = \phi_{s0} + V_{cb} + \frac{\phi_{dd} - (\phi_{s0} + V_{cb})}{\sqrt{1 + \left[\frac{\phi_{dd} - (\phi_{s0} + V_{cb})}{4v_{th}} \right]^2}}, \quad (3.3.9b)$$

and is unified with forward smoothing function in the argument of the natural log function

$$\Delta = v_{th} \ln \left\{ \frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(\frac{V_{ga}}{\gamma} \right)^2 - (\phi_{s0} + V_{cb}); \sigma_{eff} \right\} + 1 \right\}. \quad (3.3.10)$$

$(V_{gb} - V_{fb} - \phi_p - \phi_{str}^*)$ in Eq. (3.3.9a) is replaced by another forward smoothing function

$$V_{ga} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{str}^* - \phi_p; \sigma_{vga} \right\} \quad (3.3.10a)$$

$$\phi_{str}^* = \phi_{s0} + V_{cb} + \frac{\phi_{sub} - (\phi_{s0} + V_{cb})}{\sqrt{1 + \left[\frac{\phi_{sub} - (\phi_{s0} + V_{cb})}{4v_{th}} \right]^2}} \quad (3.3.10b)$$

which results in a unified “non-pinned” surface potential in the strong inversion region

$$\phi_{str} = \phi_{s0} + \Delta + V_{cb}. \quad (3.3.11)$$

Explicit surface potential valid for all operating regions can be expressed as

$$\phi_{seff} = \mathcal{G}_{eff} \{ \phi_{sa}, \phi_{str}; \delta_\phi \} \quad (3.3.12)$$

or another explicit surface potential valid from depletion to strong inversion but approaching zero in the accumulation region can be expressed similarly as

$$\phi_{ds} = \mathcal{G}_{eff} \{ \phi_{sub}, \phi_{str}; \delta_\phi \}, \quad (3.3.13)$$

which is an important component in our unified charge modeling approach.

Figure 3.14: Unified ϕ_{sub} (solid lines) and ϕ_{SUB} (dotted lines) solutions with and without ϕ_{peff} , respectively. The left axis is for their functional behavior, while the right axis for the derivatives.

Figure 3.15: Unified ϕ_{ds} (solid lines) and ϕ_{DS} (dotted lines) solutions with and without ϕ_{peff} , respectively. The left axis is for their functional behavior, while the right-axis for the derivatives.

The unified ϕ_{sub} and ϕ_{ds} solutions are shown in figure 3.14 and 3.15, respectively. Even with ϕ_{peff} , both ϕ_{sub} and ϕ_{ds} do not have significant effect on their functional and derivative solutions compared with ϕ_{SUB} and ϕ_{DS} solutions. This can be explained with figure 3.11. The amount of ϕ_p solutions has less effect on total solutions from the left-hand side of (3.3.8) within the operating regions. This validates our approximation in formulating the regional and unified ϕ_s models without considering ϕ_p solutions at the beginning but included the minor ϕ_p effect at the later stage for better accuracy. Our unified regional modeling methodology explicitly decouples both ϕ_s and ϕ_p dependency which is the one of the key features in charge modeling, as will be discussed in detail in the following chapter.

3.4 Surface potential with quantum-mechanical effect (QME)

It is well known that the energy of the free electrons in the channel is quantized and free electrons are confined to a well-defined potential well [41]. Previously derived intrinsic charge models treat the MOSFET in a classical way in the sense that the Poisson equation, Gauss's law and potential balance equation are invoked and any effects, which arise from the quantum nature of the 2D electron gas, are completely ignored. A combination of thin gate oxide and high channel doping concentration results in a high vertical field which enhances electron gas quantization in a confined potential well, leading to a decreased capacitance and an increased threshold voltage [42]. The simplest way to model these effects is to adjust the physical oxide thickness and channel doping concentration to be the effective parameters. However, this might not give the best solution to model QME.

In the literature, three models have been proposed to account for quantum mechanical (QM) effect: Hansch [43], van Dort [44] and hybrid [45]. Among those three models, van Dort quantum model has been well adopted in modeling QME via the increase in the energy band gap

$$\Delta E_g = \kappa^{qm} \frac{3}{8} \frac{\hbar^2}{m_e^*} \left[\frac{12 m_e^* q^2}{\epsilon_{si} \hbar^2} \frac{2C_{ox} (V_{gb} - V_{fb} - \phi_s)/q}{3} \right]^{2/3}, \quad (3.4.1)$$

which is obtained under triangular-potential approximation. In Eq. (3.4.1), $\hbar$ is the reduced plank's constant, m_e^* is the effective electron mass, and by default $\kappa^{qm} = 1$. The factor κ^{qm} was introduced in Ref. [46] to account for the effect of quantization level and usually is obtained by fitting to the measurement data. It was suggested by van Dort that QME is accounted for by modification of the intrinsic carrier concentration

$$\begin{aligned}
 n_i^{qm} &= n_i \exp \left[\frac{E_g^{qm} - E_g}{2kT} \right], \\
 &= n_i \exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right]
 \end{aligned} \tag{3.4.2}$$

$$\Delta E_g^{qm} = \frac{\Delta E_g / q}{2}, \tag{3.4.2a}$$

in which the net space-charge density in the quantum well is defined by

$$\rho^{qm} = q \left[p^{qm} - n^{qm} - N_a + N_d^{qm} \right]. \tag{3.4.3}$$

Here the charges in the space-charge region are given by

$$N_d^{qm} = N_a \exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right] \exp \left[\frac{-2\phi_f}{v_{th}} \right], \tag{3.4.3a}$$

$$p^{qm} = N_a \exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right] \exp \left[\frac{-\psi}{v_{th}} \right], \tag{3.4.3b}$$

and

$$n^{qm} = N_a \exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right] \exp \left[\frac{\psi - V_{cb} - 2\phi_f}{v_{th}} \right]. \tag{3.4.3c}$$

Ignoring the spatial dependency of ϕ_s in ΔE_g , combining with Poisson equation and Gauss's law in the channel/oxide interface, one can arrive at the following implicit equation at the channel interface

$$\begin{aligned}
 (V_{gb} - V_{fb} - \phi_s - \phi_p)^2 &= \gamma^2 \left\{ v_{th} \left(\exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right] \exp \left[\frac{-\phi_s}{v_{th}} \right] - 1 \right) + \phi_s \left(1 - \exp \left(\frac{\Delta E_g^{qm}}{v_{th}} \right) \exp \left[\frac{-2\phi_f}{v_{th}} \right] \right) \right. \\
 &\quad \left. + v_{th} \exp \left[-\frac{V_{cb} + 2\phi_f}{v_{th}} \right] \left(\exp \left[-\frac{\Delta E_g^{qm}}{v_{th}} \right] \exp \left[\frac{\phi_s}{v_{th}} \right] - 1 \right) \right\}
 \end{aligned} \tag{3.4.4}$$

in which the equation is the modification of the original Pao-Sah model with QME. It is clear from the implicit equation that no QME is considered in the depletion region.

3.5 Regional surface potential with quantum-mechanical effect

An implicit expression to solve ϕ_s iteratively at the channel interface is physically derived (3.4.4) with QME. If only the holes are considered in the channel space charge region in the derivation, it gives

$$\left(V_{gb} - V_{fb} - \phi_s - \phi_p\right)^2 = \gamma^2 \left\{ v_{th} \exp \left[-\frac{\Delta E_g^{qm} \big|_{\phi_s = \phi_{CC}}}{v_{th}} \right] \exp \left[\frac{-\phi_s}{v_{th}} \right] \right\}, \quad (3.5.1)$$

with $\phi_s = \phi_{CC}^{qm}$ in the accumulation region. Ignoring ϕ_p drop at high poly-gate doping, the above equation can be re-written as

$$\exp(X^{qm}) + a^{qm} X^{qm} + B^{qm} = 0 \quad (3.5.1a)$$

$$X^{qm} = \frac{-\phi_{CC}^{qm}}{2v_{th}}, \quad a^{qm} = \frac{2\sqrt{v_{th}}}{\gamma} \exp \left[-\frac{\Delta E_g^{qm} \big|_{\phi_s = \phi_{CC}}}{2v_{th}} \right], \quad B^{qm} = \frac{V_{gb} - V_{fb}}{\gamma\sqrt{v_{th}}} \exp \left[-\frac{\Delta E_g^{qm} \big|_{\phi_s = \phi_{CC}}}{2v_{th}} \right] \quad (3.5.1b)$$

so that ϕ_{CC}^{qm} can be solved explicitly with the well-known *Lambert W* function, $\mathcal{L}\{W^{qm}\}$,

$$\phi_{CC}^{qm} = -2v_{th} \left[-\frac{a^{qm} \mathcal{L}\{W^{qm}\} + B^{qm}}{a^{qm}} \right] = V_{gb} - V_{fb} + 2v_{th} \mathcal{L}\{W^{qm}\} \quad (3.5.2)$$

$$\begin{aligned} W^{qm} &= \frac{1}{a^{qm}} \exp \left(-\frac{B^{qm}}{a^{qm}} \right) \\ &= \frac{\gamma}{2\sqrt{v_{th}}} \exp \left(-\frac{V_{gb} - V_{fb} - \Delta E_g^{qm} \big|_{\phi_s = \phi_{CC}}}{2v_{th}} \right). \end{aligned} \quad (3.5.2a)$$

A regional model can be similarly derived as

$$\left(V_{gb} - V_{fb} - \phi_s\right)^2 = \gamma^2 \left\{ \phi_s + v_{th} \exp \left[-\frac{V_{cb} + 2\phi_f}{v_{th}} \right] \left(\exp \left[-\frac{\Delta E_g^{qm} \big|_{\phi_s = \alpha^{qm} \phi_{SS}}}{v_{th}} \right] \exp \left[\frac{\phi_s}{v_{th}} \right] \right) \right\} \quad (3.5.3)$$

in the strong inversion region when only considering the fully ionized acceptors and the free electrons in the channel space charge region, where $\phi_s = \phi_{SS}^{qm}$ and α^{qm} is the weighting factor with default value 1. A regional ϕ_{SS}^{qm} model can be formulated with QME

$$\phi_{SS}^{qm} = \phi_{s0} + \Delta_{SS}^{qm} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SS}}, \quad (3.5.4)$$

in which the regional “non-pinned” surface potential increment is given as

$$\Delta_{SS}^{qm} = v_{th} \ln \left\{ \frac{1}{v_{th}} \left[\left(\frac{V_{gb} - V_{fb} - \phi_{SS}^{*qm}}{\gamma} \right)^2 - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SS}} \right) \right] + 1 \right\} \quad (3.5.4a)$$

and

$$\phi_{SS}^{*qm} = \phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SS}} + \frac{\phi_{DD} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SS}} \right)}{\sqrt{1 + \frac{\left[\phi_{DD} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SS}} \right) \right]^2}{[4v_{th}]^2}}}. \quad (3.5.4b)$$

Figure 3.16: Regional surface potential solutions in the strong inversion (solid lines) and accumulation (dashed lines) regions are compared with Medici solutions with κ^{qm} variation.

3.6 Unified regional surface potential with quantum-mechanical effect

Both the analytic explicit regional channel surface-potential models, ϕ_{CC}^{qm} and ϕ_{SS}^{qm} give correct asymptotic implicit ϕ_s solutions as shown in figure 3.16. However, beyond their validity region (above flat band and threshold voltage, respectively), both expressions give incorrect solutions. In order to circumvent this problem, a unified regional model is proposed to unify ϕ_{CC}^{qm} model with

$$\phi_{ACC}^{qm} = \phi_{CC}^{qm} \Big|_{V_{gb}-V_{fb}=V_{GBR}} = V_{GBR} + 2v_{th} \mathcal{L}\{W^{qm}\}, \quad (3.6.1)$$

which can be directly added with ϕ_{SUB} model to form

$$\phi_{SA}^{qm} = \phi_{ACC}^{qm} + \phi_{SUB}, \quad (3.6.2)$$

as shown in the figure 3.17 (solid line), with second-order derivative shown in the inset. It gives correct asymptotic behavior from accumulation to depletion regions with only the QME contribution in the accumulation region, in which

$$\Delta E_g \Big|_{\phi_s=\phi_{CC}} \propto [2v_{th} \mathcal{L}\{W\}]^{2/3}, \quad (3.6.3)$$

$$W = \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(-\frac{V_{gb}-V_{fb}}{2v_{th}}\right). \quad (3.6.3a)$$

The well-known discontinuity of the van Dort QM model at flat band is shown by the dotted line in the inset. In order to remove the intrinsic problem in the van Dort QME model, we propose to replace the argument W in the Lambert function with

$$W^{sm} = \frac{\gamma}{2\sqrt{v_{th}}} \exp\left[-\frac{V_{GB,fb} - (V_{fb} - \sigma^{qm})}{2v_{th}}\right] \quad (3.6.4)$$

where V_{gb} in the exponential function of (3.6.3a) is replaced by

$$V_{GB,fb} = \mathcal{G}_{eff}\{V_{gb}, V_{fb} - \sigma^{qm}; \delta^{qm}\}, \quad (3.6.4a)$$

and a QME weighing factor ε^{qm} is added to ΔE_g

$$\Delta E_{g,acc} = \Delta E_g + \varepsilon^{qm}. \quad (3.6.5)$$

Figure 3.17: Unified surface- potential regional models are compared to Medici solutions with (triangle) and without (circle) QME. Inset shows numerical second-order derivatives for the unified models with (dotted line) and without (solid line) QME model.

Three fitting parameters are associated with our accumulation QME model. Gate capacitance and its first-order derivative solutions with 3 fitting model parameter variation are as shown in figure 3.18 (a) and (b), respectively. Glitch in the first-order C_{gg} solutions is shown with W (dashed dot-dot) used in the Lambert- W function, while with fitted model parameter for W^{sm} , smooth transition from accumulation to depletion regions is observed. With the fitted model parameter set, our unified QME model in the accumulation region is able to predict the functional and higher-order derivatives of the capacitance solution without any problem with various κ^{qm} values as defined in the Medici (figure 3.19).

Figure 3.18: (a) C_{gg} and band gap widening solutions and (b) the first-order derivative of C_{gg} solutions are compared with one-fitted model parameter (dotted lines), two-fitted (dashed lines) and unfitted model, W .

Figure 3.19: Predicted QME on (a) capacitances and band gap widening (inset) and (b) surface potentials in the accumulation region with κ^{qm} variations.

In the strong inversion region, the following unified ϕ_{STR}^{qm} solution is formulated with QME:

$$\phi_{STR}^{qm} = \phi_{s0} + \Delta^{qm} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \quad (3.6.6)$$

$$\Delta^{qm} = v_{th} \ln \left\{ \frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(\frac{V_{GA}^{qm}}{\gamma} \right)^2 - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right); \sigma_{eff} \right\} + 1 \right\}, \quad (3.6.6a)$$

$$V_{GA}^{qm} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{STR}^{*qm}; \sigma_{vga} \right\}, \quad (3.6.6b)$$

$$\phi_{STR}^{*qm} = \phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} + \frac{\phi_{SUB} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right)}{\sqrt{1 + \frac{\left[\phi_{SUB} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right) \right]^2}{[4v_{th}]^2}}}. \quad (3.6.6c)$$

A single-piece explicit unified ϕ_{SEFF} model is replaced by

$$\phi_{SEFF}^{qm} = \mathcal{G}_{eff} \left\{ \phi_{SA}^{qm}, \phi_{STR}^{qm}; \delta_\phi \right\} \quad (3.6.7)$$

for full operating region, or

$$\phi_{DS}^{qm} = \mathcal{G}_{eff} \left\{ \phi_{SUB}, \phi_{STR}^{qm}; \delta_\phi \right\} \quad (3.6.8)$$

from depletion and strong inversion regions, as shown in figure 3.20. Figure 3.21 shows the ϕ_{SEFF}^{qm} prediction for various κ^{qm} values compared with Medici data.

Figure 3.20: Models comparison between the unified regional models and Medici solutions with (triangle) and without (circle) QME. Inset shows the second-order numerical derivative of the unified regional model solutions.

Figure 3.21: Unified regional surface-potential, ϕ_{SEFF}^{qm} solutions with QME. Model prediction for various κ^{qm} value used in Medici simulation.

3.7 Regional surface potential with coupled quantum-mechanical and poly-doping effects

In the previous sections, the quantized electron gas confined to a well-defined potential well and the ϕ_p potential drop across an induced charge layer formed at the gate/oxide interface have been discussed independently. The impact of QM and poly-doping effects on the channel surface potential is actually coupled and should be considered simultaneously in the surface-potential formulation. In the following, a new modeling methodology is proposed to couple both effects in our surface-potential formulation.

In the accumulation region, the unified regional gate surface potential has been given by

$$\phi_{p,acc} = \phi_{p,cc} \Big|_{\phi_s = \phi_{sr}} = -v_{th} \ln \left[\frac{\gamma^2}{\gamma_p^2} \exp \left(-\frac{\phi_{sr}}{v_{th}} \right) + \left(1 - \frac{\gamma^2}{\gamma_p^2} \right) \right]^2, \quad (3.7.1)$$

where ϕ_{sr} is the reverse unified model

$$\phi_{sr} = \mathcal{G}_r \{ \phi_s ; \rho_r \} \quad (3.7.1a)$$

which is obtained by equating $\phi_s = \phi_{SEFF}$ without QME. A unified regional gate surface potential with QME is suggested as

$$\phi_{p,acc}^{qm} = \phi_{p,acc} \Big|_{\phi_{sr} = \phi_{sr}^{qm}}, \quad (3.7.2)$$

in which the previous ϕ_{sr} is now replaced by

$$\phi_{sr}^{qm} = \mathcal{G}_r \{ \phi_{SEFF}^{qm,\beta} ; \rho_r \}, \quad (3.7.2a)$$

where

$$\phi_{SEFF}^{qm,\beta} = \mathcal{G}_{eff} \{ \phi_{SA}^{qm,\beta}, \phi_{STR}^{qm} ; \delta_\phi \}, \quad (3.7.2b)$$

$$\phi_{SA}^{qm,\beta} = \phi_{ACC}^{qm,\beta} + \phi_{SUB}, \quad (3.7.2c)$$

$$\phi_{ACC}^{qm,\beta} = \phi_{ACC}^{qm} \Big|_{W^{qm} \rightarrow W^{qm,\beta}}, \quad (3.7.2d)$$

$$W^{qm,\beta} = \frac{\gamma}{2\sqrt{v_{th}}} \exp \left(- \frac{V_{gb} - V_{fb} + \beta^{qm} \Delta E_g^{qm} \Big|_{\phi_s = \phi_{cc}}}{2v_{th}} \right). \quad (3.7.2e)$$

In the above equations, β^{qm} is the weighting factor without considering poly-doping effect in the QME model (ΔE_g^{qm}) at high gate doping concentration. The explicit channel surface-potential with coupled poly-doping and quantum-mechanical effects is formulated as

$$\phi_{acc}^{qm} = V_{gbr}^{qm} + 2v_{th} \mathcal{L} \{ w^{qm} \}, \quad (3.7.3)$$

$$V_{gbr}^{qm} = V_{gbr} \Big|_{\phi_p = \phi_{peff}^{qm}}, \quad (3.7.3a)$$

in which

$$w^{qm} = \frac{\gamma}{2\sqrt{v_{th}}} \exp \left(- \frac{V_{gb} - V_{fb} - \phi_{peff}^{qm} + \Delta E_g^{qm} \Big|_{\phi_s = \phi_{CC}}}{2v_{th}} \right) \quad (3.7.3b)$$

and ϕ_{peff}^{qm} is to be defined in the following.

In our previous discussion, ϕ_{STR}^{qm} has been formulated without poly-doping effect and $\phi_{p, str}$ has been formulated without QME. A unified regional gate surface potential is suggested:

$$\phi_{p, str}^{qm} = \phi_{p, str} \Big|_{\phi_s = \phi_{SEFF}^{qm}}, \quad (3.7.4)$$

where QME is modeled through ϕ_{SEFF}^{qm} in the gate surface-potential model. A single-piece explicit unified gate surface-potential with QME can be formulated as

$$\phi_{peff}^{qm} = \mathcal{G}_{eff} \left\{ \phi_{pa}^{qm}, \phi_{p, str}^{qm}; \delta_p \right\} \quad (3.7.5)$$

$$\phi_{pa}^{qm} = \phi_{p, acc}^{qm} + \phi_{p, sub}. \quad (3.7.5a)$$

Similarly, at the channel/oxide interface, a unified regional surface-potential expression is formulated as

$$\phi_{str}^{qm} = \phi_{s0} + \Delta^{qm} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}^{qm}}, \quad (3.7.6)$$

$$\Delta^{qm} = v_{th} \ln \left\{ \frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(\frac{V_{ga}^{qm}}{\gamma} \right)^2 - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}^{qm}} \right); \sigma_{eff} \right\} + 1 \right\}, \quad (3.7.6a)$$

in which the coupled QM and gate-doping effects are accounted for in the formulation with

$$V_{ga}^{qm} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{str}^{*qm} - \phi_{peff}^{qm}; \sigma_{vga} \right\} \quad (3.7.6b)$$

$$\phi_{str}^{*qm} = \phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} + \frac{\phi_{SUB} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right)}{\sqrt{1 + \frac{\left[\phi_{SUB} - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right) \right]^2}{[4v_{th}]^2}}}}. \quad (3.7.6c)$$

A single-piece explicit unified channel surface-potential is expressed as

$$\phi_{seff}^{qm} = \mathcal{G}_{eff} \left\{ \phi_{sa}^{qm}, \phi_{str}^{qm}; \delta_\phi \right\} \quad (3.7.7)$$

for full operating region, and

$$\phi_{ds}^{qm} = \mathcal{G}_{eff} \left\{ \phi_{SUB}, \phi_{str}^{qm}; \delta_\phi \right\} \quad (3.7.8)$$

from depletion to strong inversion regions. The impact of QME on gate surface potential is shown in figure 3.22, in which significant drop in gate surface potential in accumulation and strong inversion regions are observed. The validation for both coupled effects will be discussed in the next chapter.

Figure 3.22: A unified regional gate surface potential, ϕ_{peff}^{qm} with κ^{qm} variation for QME.

CHAPTER 4: INTRINSIC CHARGE MODELING

The operation principle of a MOSFET is to control the terminal induced charges through the four terminal voltages, namely the gate, source, drain and substrate. Therefore, MOSFET intrinsic charges can be considered as functions of all the external voltages (V_g , V_s , V_d and V_b). For each terminal, the corresponding terminal charge can be derived physically. We use Q_g as the notation to represent the total gate charge density (per unit area) at the gate/oxide. Q_b , Q_d and Q_s as the total source, drain and bulk charge densities at the channel/oxide interface, respectively. In the following sections, the conventional threshold-voltage-based approach with “pinned” surface-potential to model the terminal charges is discussed. Shortcomings of this approach are presented and a way to overcome this with our “non-pinned” surface-potential is formulated.

4.1 Conventional Charge Model

4.1.1 “Pinned” surface-potential threshold-based approach

Charges are distributed across the channel of length L and width W . In order to develop a AC model for circuit simulator with capacitive effect, it requires integrated solutions from the position-dependent induced charge densities,

$$Q_X = W \int_0^L Q_x(y) dy \quad \begin{cases} X = G, B, I \\ x = g, b, i \end{cases} \quad (4.1.1)$$

where the subscripts G , B and I denote total gate, bulk and inversion charges, respectively, and the lower-case subscripts denote the position-dependent charge densities.

$$Q_g(y) = -[Q_i(y) + Q_b(y) + Q_{ox}] \quad (4.1.1a)$$

$$Q_i(y) = Q_{sc}(y) - Q_b(y) \quad (4.1.1b)$$

and

$$Q_b(y) = -C_{ox}\gamma \operatorname{sgn}[\psi_s(y)] \sqrt{v_{th} \left(\exp \left[\frac{-\psi_s(y)}{v_{th}} \right] - 1 \right) + \psi_s(y)} \quad (4.1.1c)$$

are the gate, inversion and bulk position-dependent charge densities, respectively, with the induced channel-charge

$$Q_{sc} = -C_{ox} [V_{gb} - V_{fb} - \psi_s(y) - \psi_p(y)], \quad (4.1.1d)$$

which can be determined by combining the potential balance equation with the charge neutrality equation. The total drain and source charges can be formulated from linear partition of $Q(y)$ based on the Ward-Dutton partitioning scheme [21]:

$$Q_Y = W \int_0^L f_Y Q_i(y) dy \quad \begin{cases} Y = D : f_D = \frac{y}{L} \\ Y = S : f_S = 1 - \frac{y}{L} \end{cases} \quad (4.1.2)$$

The conventional charge model with “pinned” ϕ_s assumes that the channel surface-potential, ϕ_s is pinned at $(\phi_{s0} + V_{cb})$ along the active area in the strong inversion region.

Thus, the position-dependent bulk-charge model is assumed to be $Q_b(y) \approx -\gamma \sqrt{\phi_{s0} + V_{cb}}$ and

the inversion charge can be expressed as

$$Q_i(y) = -\left\{ C_{ox} [V_{gb} - V_{fb} - \phi_{s0} - V_{cb}] - \gamma \sqrt{\phi_{s0} + V_{cb}} \right\}, \quad (4.1.3)$$

where a small voltage drop across the induced layer at the gate interface has been neglected. The above “pinned” ϕ_s -based inversion model is further transformed into voltage-dependent inversion charge in the first-order Meyer model [47]

$$Q_i(V) = -\left\{ C_{ox} [V_{gs} - V_t - V] \right\}, \quad (4.1.4)$$

in which V is defined as the Fermi-potential along the channel and

$$V_t = V_{fb} + \phi_{s0} + \gamma\sqrt{\phi_{s0} + V_{sb}}. \quad (4.1.4a)$$

is the threshold voltage. It has been reported that Meyer's model has its intrinsic problems with charge conservation [48] and capacitances reciprocal [25] behaviors. These problems make it less attractive in circuit simulators [49]. In order to avoid the same intrinsic problems, we perform Taylor's series expansion on $\sqrt{\phi_{s0} + V_{sb} + V}$ in the bulk charge expression and reformulate the potential-dependent inversion charge such that

$$Q_i(V) = -\left\{C_{ox} \left[V_{gs} - V_t - A_b V \right] \right\}, \quad (4.1.5)$$

with the bulk-charge factor,

$$A_b = 1 + \frac{\gamma}{2\sqrt{\phi_{s0} - V_{bs}}}. \quad (4.1.5a)$$

Combining (4.1.5) with the integrated solutions from (4.1.1) and (4.1.2), the following expressions are formulated

$$Q_i = \frac{Q_i}{WL} = \frac{(Q_D + Q_S)}{WL} \quad (4.1.6)$$

$$Q_d = \frac{Q_D}{WL} = -C_{ox} \left(\frac{V_{gs} - V_t}{2} - \frac{A_b}{3} V_{ds} + A_d \right) \quad (4.1.7)$$

$$Q_s = \frac{Q_S}{WL} = -C_{ox} \left(\frac{V_{gs} - V_t}{2} - \frac{A_b}{6} V_{ds} + A_s \right) \quad (4.1.8)$$

$$Q_b = \frac{Q_B}{WL} = -C_{ox} \left(\gamma\sqrt{\phi_{s0} + V_{sb}} + D_b \right) \quad (4.1.9)$$

$$Q_g = \frac{Q_G}{WL} = -(Q_i + Q_b + Q_{ox}) \quad (4.1.10)$$

with auxiliary functions:

$$A_s = \mathcal{A}(1 - \mathcal{B}) \quad (4.1.11)$$

$$A_d = \mathcal{A}\mathcal{B} \quad (4.1.12)$$

$$D_b = (A_b - 1)V_{ds}\mathcal{D} \quad (4.1.13)$$

$$\mathcal{A} = \frac{A_b^2 V_{ds}^2}{12(V_{gs} - V_t - 0.5A_b V_{ds})} \quad (4.1.14)$$

$$\mathcal{B} = \frac{5(V_{gs} - V_t) - 2A_b V_{ds}}{10(V_{gs} - V_t - 0.5A_b V_{ds})} \quad (4.1.15)$$

$$\mathcal{D} = \frac{3(V_{gs} - V_t) - 2A_b V_{ds}}{6(V_{gs} - V_t - 0.5A_b V_{ds})}. \quad (4.1.16)$$

It is obvious that from the above equations, (4.1.6) to (4.1.16) with the “pinned” ϕ_s (ϕ_{so}), the models are formulated in the strong inversion region and they are piecewise regional models. When V_{ds} and $(V_{gs} - V_t)$ are replaced by the following smoothing functions:

$$V_{deff} = \mathcal{G}_{eff} \{V_{dsat}, V_{ds}; \delta_s\}, \quad (4.1.17)$$

and

$$V_{geff} = \frac{2nv_{th} \ln \left(1 + e^{(V_{gs} - V_t)/(2nv_{th})} \right)}{1 + 2n(C_{ox}/C_d) e^{-(V_{gs} - V_t - 2V_{off})/(2nv_{th})}}, \quad (4.1.18)$$

respectively, “unified” regional models result. However, this modeling methodology is based on the “pinned” ϕ_s -based model. Charges and capacitances from this approach are not accurately modeled in the strong inversion region [11], as shown in figure 4.1.

4.1.2 Surface-potential-based model for depletion and strong inversion regions

In the recent studies [50], [51] with the surface-potential-based models, charge model has been developed with position-dependent inversion charge density

$$Q_i(y) = Q_{sc}(y) - Q_b(y). \quad (4.1.19)$$

From the equation (4.1.1d), the channel charge density, $Q_{sc}(y)$ is assumed initially with no poly-doping effect and the bulk charge density is approximated without hole contribution in the accumulation region such that the exponential terms in the square-root in (4.1.1c) can be ignored,

$$Q_b(y) \approx -\gamma \sqrt{\psi_s(y) + v_{th}} \quad (4.1.19b)$$

or

$$Q_b(y) \approx -\gamma \sqrt{\psi_s(y)}. \quad (4.1.19c)$$

The total integrated charges based on this approach may not be as compact as the “pinned” ϕ_s -based. The derived models are only applicable for the regions where the channel is depleted or inverted. Furthermore, the models contain square root of the average/difference between the surface potentials at the source and drain ends that requires precise ϕ_s -solutions.

4.1.3 Charge model in the accumulation region

Although the exact full-region ϕ_s solutions can be calculated implicitly, due to the exponential term in the bulk charge expression, it is nontrivial to formulate an integrated full region single-piece bulk-charge model with V_{ds} variation. The conventional V_t -based approach [52] introduces an effective flat-band voltage, $V_{FB\text{eff}}$, while the ϕ_s -based approach [50] introduces a limiting function with empirical parameters to model total integrated charge in the accumulation region. Both approaches are neither physical nor accurate. In this work, a new modeling methodology is suggested and will be discussed shortly in the following sections.

4.2 Position-dependent inversion charge with “non-pinned” surface potential

In the previous sections, the “pinned” ϕ_s -based intrinsic charge models have been derived with both the surface potentials, $(\phi_{s0} + V_{sb})$ and $(\phi_{s0} + V_{db})$, “pinned” at the source and drain ends, respectively, in the strong inversion region. With the same conditions, drain current can be obtained from integrating the inversion charge density

$$I_{ds} = -\frac{\mu W}{L} \int_{\phi_{s0} + V_{sb}}^{\phi_{s0} + V_{db}} Q_i(y) d\psi \quad (4.2.1)$$

$$= \frac{\mu W C_{ox}}{L} \left\{ (V_{gs} - V_t) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\}$$

where

$$A_b = 1 + \frac{\gamma}{2\sqrt{\phi_{s0}}} \quad (4.2.1b)$$

is defined as the “pinned” ϕ_s -based bulk-charge factor.

A “non-pinned” ϕ_s -based approach is proposed. Two surface potentials, $\phi_{STR,s} = (\phi_{s0} + \Delta_S + V_{sb})$ and $\phi_{STR,d} = (\phi_{s0} + \Delta_D + V_{db})$ are suggested at the source and drain ends, respectively, where $\Delta_S = \Delta|_{V_{cb} = V_{sb}}$ and $\Delta_D = \Delta|_{V_{cb} = V_{db}}$. Integrated drain current can be formulated in a similar way

$$I_{ds} = -\frac{\mu W}{L} \int_{\phi_{STR,s}}^{\phi_{STR,d}} Q_i(y) d\phi = -\frac{\mu W}{L} \int_{\phi_{STR,s}}^{\phi_{STR,d}} Q_{sc}(y) - Q_b(y) d\phi \quad (4.2.2)$$

$$= \frac{\mu W C_{ox}}{L} \left\{ (V_{gs} - V_{fb}) V_{ds,\Delta} - \frac{1}{2} [V_{ds,\Delta}^2 + 2V_{ds,\Delta} (\phi_{STR,s})] - \gamma \sqrt{\phi_{STR,s}} (V_{ds,\Delta}) + \frac{V_{ds,\Delta}^2}{4\sqrt{\phi_{STR,s}}} \right\}$$

$$= \frac{\mu W C_{ox}}{L} \left\{ (V_{gs} - V_{t,\phi}) V_{ds,\Delta} - \frac{1}{2} A_{b,\Delta} V_{ds,\Delta}^2 \right\}$$

in which the “non-pinned” ϕ_s -based approach, threshold voltage, bulk-charge factor and drain-source voltage, respectively, are given by

$$V_{t,\phi} = V_{fb} + \phi_{s0} + \Delta_s + \gamma\sqrt{\phi_{DS,s}}, \quad (4.2.2a)$$

$$A_{b,\Delta} = 1 + \frac{\gamma}{2\sqrt{\phi_{DS,s}}}, \quad (4.2.2b)$$

$$V_{ds,\Delta} = V_{ds} + \Delta_{EFF,ds}, \quad (4.2.2c)$$

with

$$\Delta_{EFF,ds} = \Delta_{EFF,d} - \Delta_{EFF,s} \quad (4.2.2d)$$

$$\Delta_{EFF} = \phi_{SEFF} - (\phi_{s0} + V_{cb}). \quad (4.2.2e)$$

$\phi_{DS,s}$, $\Delta_{EFF,s}$ and $\Delta_{EFF,d}$ solutions approach asymptotic $\phi_{STR,s}$, Δ_s and Δ_d solutions, respectively, in the strong inversion region. They are used in the final integrated drain current instead of $\phi_{STR,s}$, Δ_s and Δ_d . Another approach in deriving drain current is by integrating

$$Q_i(V) = -\left\{C_{ox} \left[V_{gs} - V_{t,\phi} - A_{b,\Delta} V \right] \right\}, \quad (4.2.3)$$

with both the potentials, $V(0) = 0$ and $V(L) = V_{ds,\Delta}$ at the source and drain ends, resulting in

$$\begin{aligned} I_{ds} &= \int_0^L I_{ds}(y) dy = - \int_{V(0)}^{V(L)} [Q_i(V)] dV, \\ &= \mu W C_{ox} \left\{ (V_{gs} - V_{t,\phi}) V_{ds,\Delta} - \frac{1}{2} A_{b,\Delta} V_{ds,\Delta}^2 \right\} \end{aligned} \quad (4.2.4)$$

which gives a similar integrated drain current as (4.2.2). Both derivations explain the transformation of the “pinned” model to the “non-pinned” model with simple replacements of threshold-voltage ($V_t \rightarrow V_{t,\phi}$), bulk-charge factor ($A_b \rightarrow A_{b,\Delta}$) and drain-source voltage ($V_{ds} \rightarrow V_{ds,\Delta}$).

4.2.1 Unified drain, source and bulk charge models

Similar transformation is adopted in our intrinsic charge models (4.1.6) – (4.1.16).

Total integrated drain, source and bulk charges are given by

$$Q_d = \frac{Q_D}{WL} = -C_{ox} \left(\frac{V_{gt,\phi}}{2} - \frac{A_{b,\Delta}}{3} V_{deff,\Delta} + A_{d,\Delta} \right) \quad (4.2.5)$$

$$Q_s = -C_{ox} \left(\frac{V_{gt,\phi}}{2} - \frac{A_{b,\Delta}}{6} V_{deff,\Delta} + A_{s,\Delta} \right) \quad (4.2.6)$$

$$Q_{b,str} = -C_{ox} \left(\gamma \sqrt{\phi_{STR,s}} + D_{b,\Delta} \right) \quad (4.2.7)$$

with the auxiliary functions from equations (4.1.11)-(4.1.16) to be modified as:

$$A_{s,\Delta} = \mathcal{A}_\Delta (1 - \mathcal{B}_\Delta) \quad (4.2.8)$$

$$A_{d,\Delta} = \mathcal{A}_\Delta \mathcal{B}_\Delta \quad (4.2.9)$$

$$D_{b,\Delta} = (A_{b,\Delta} - 1) V_{deff,\Delta} \mathcal{D}_\Delta \quad (4.2.10)$$

$$\mathcal{A}_\Delta = \frac{A_{b,\Delta}^2 V_{deff,\Delta}^2}{12(V_{geff,\Delta_t} - 0.5 A_{b,\Delta} V_{deff,\Delta})} \quad (4.2.11)$$

$$\mathcal{B}_\Delta = \frac{5V_{geff,\Delta} - 2A_{b,\Delta} V_{deff,\Delta}}{10(V_{geff,\Delta} - 0.5 A_{b,\Delta} V_{deff,\Delta})} \quad (4.2.12)$$

$$\mathcal{D}_\Delta = \frac{3V_{geff,\Delta} - 2A_{b,\Delta} V_{deff,\Delta}}{6(V_{geff,\Delta} - 0.5 A_{b,\Delta} V_{deff,\Delta})} \quad (4.2.13)$$

V_{geff} and V_{deff} are modified to

$$V_{geff,\Delta} = \frac{2n\nu_{th} \ln \left(1 + e^{\frac{(V_{gs} - V_{t,\phi})}{2n\nu_{th}}} \right)}{1 + 2n(C_{ox}/C_d) e^{\frac{(V_{gs} - V_{t,\phi} - 2V_{off})}{2n\nu_{th}}}} \quad (4.2.13a)$$

and

$$V_{deff,\Delta} = V_{deff} + \Delta_{EFF,ds}, \quad (4.2.13b)$$

respectively, and

$$V_{gt,\phi} = nV_{th} \ln \left(1 + e^{\frac{(V_{gs} - V_{t,\phi})}{(nV_{th})}} \right). \quad (4.2.13c)$$

Analytic drain and source capacitances data are generated from MEDICI and charges are calculated numerically which are compared with explicit “non-pinned” (solid lines) and “pinned” (dotted lines) ϕ_s solutions, as shown in the figure 4.1 and its inset. Modeled capacitances are numerically computed based on

$$C_{ij} = \delta_{ij} \frac{\partial Q_i}{\partial V_j}, \quad \delta_{ij} = \begin{cases} 1 & (i = j) \\ -1 & (i \neq j) \end{cases}. \quad (4.2.14)$$

Both modeled source/drain charges and capacitances obtained from the “non-pinned” ϕ_s -based model predict accurately the Medici data for the full operating regions while the “pinned” ϕ_s -based model over-predicts the Medici data with sharp transitions near moderate inversion region. These behaviors have been reported [11].

Figure 4.1: “Pinned” (dotted lines) and “non-pinned” (solid lines) ϕ_s -based intrinsic inversion, drain and source capacitances and charges (inset) with the Ward-Dutton partition.

4.2.2 All regions single-piece unified bulk charge model

Depletion approximation is normally adopted in the conventional approach to formulate

an integrated bulk charge model as reported in the literature, [50], [52], [53]. The approximation assumes that the contribution of the holes is negligible in the depletion and strong inversion regions. Thus, the exponential term in (4.1.1c) can be ignored and the equation (4.1.2) can be integrated.

A limiting function with empirical fitting parameters [50] and V_{FBeff} smoothing function [52] have been proposed in the literature to model bulk charge in the accumulation region. These functions are neither physical nor accurate as they are not able to integrate some physical components, such as poly-doping and quantum-mechanical effects into the charge model and cannot give accurate solutions.

Another approach to formulate an integrated bulk charge in the accumulation region is by the subtraction of both the integrated gate and inversion charges. This approach requires precise ϕ_s solutions at the source and drain ends to calculate the difference/average of the surface potentials. Small discrepancies may result in exponentially growth of errors of bulk charge solutions in the accumulation region.

In this section, a unified regional integrated bulk-charge model for all operating regions is presented. In the accumulation region, we ignore the contribution of the inversion charge to the total induced charge due to negligible contribution of electrons in this region. In this operating region, varying V_{ds} will not affect the channel surface-potential, as the potential is spatially constant along the channel. Thus, a unified regional bulk charge can be formulated as

$$Q_{b,acc} \approx Q_{sc} = -C_{ox} [V_{GBR} - \phi_{ACC}], \quad (4.2.15)$$

in which $Q_{b,acc}$ gives the correct asymptotic charge solutions in the accumulation region with zero bulk charge above flat-band, as shown in figure 4.2. V_{GBR} and ϕ_{ACC} are both the reverse unified model and unified accumulation surface-potential solutions. Similarly, in the depletion region, induced channel charge is approximated as the total bulk charge with the unified model

(a)

(b)

Figure 4.2: Capacitances from numerical derivative of: (a) $Q_{b,sa}$ and (b) $Q_{b,ds}$ with their components and charges shown in the inset.

$$Q_{b,sub} \approx Q_{sc} = -C_{ox} [V_{GBA} - \phi_{SUB}]. \quad (4.2.16)$$

$$V_{GBA} = \mathcal{G}_f \{ V_{gb} - V_{fb}; \sigma_a \} \quad (4.2.16a)$$

and

$$\phi_{SUB} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{GBF}} \right]^2 \quad (4.2.16b)$$

are both the forward unified model and unified surface-potential solutions in depletion region,

respectively. However, if we apply the charge-conservation together with the potential balance equations in the depletion region, we find

$$V_{GBA} - \phi_{SUB} \neq \gamma \sqrt{\phi_{SUB}} \quad (4.2.17)$$

but

$$V_{GBF} - \phi_{SUB} = \gamma \sqrt{\phi_{SUB}}, \quad (4.2.18)$$

because in (4.2.18), the same smoothing parameter (σ_f) is used in V_{GBF} and ϕ_{SUB} while in (4.2.17), V_{GBA} uses σ_a . By re-writing (4.2.16) and using (4.2.18):

$$\begin{aligned} Q_{b,sub} &= -C_{ox} [V_{GBA} - V_{GBF} + V_{GBF} - \phi_{SUB}] \\ &= -C_{ox} [V_{GBA} - V_{GBF} + \gamma \sqrt{\phi_{SUB}}] \end{aligned} \quad (4.2.19)$$

The bulk-charge solutions calculated from equations (4.2.16) and (4.2.19) give exactly the same numerical solutions as shown in figure 4.3 with the inset showing the absolute error between both solutions.

Figure 4.3: Unified regional capacitance from numerical derivative solutions of the “non-pinned” ϕ_s -based unified regional bulk-charge model in depletion region. The inset shows absolute error from bulk-charge models from (4.2.16) and (4.2.19).

A unified bulk-charge model with asymptotic numerical Q_b -solutions from the

accumulation to depletion regions can be formulated as

$$Q_{b,sa} = Q_{b,acc} + Q_{b,sub} \quad (4.2.20)$$

A unified bulk-charge model with asymptotic numerical Q_b solutions from depletion to strong inversion regions can also be formulated by replacing ϕ_{SUB} with $\phi_{DS,s}$ in the square-root of (4.2.19)

$$Q_{b,ds} = -C_{ox} \left(V_{GBA} - V_{GBF} + \gamma \sqrt{\phi_{DS,s}} \right). \quad (4.2.21)$$

The single-piece full region ($V_{ds} = 0$ condition) “non-pinned” ϕ_s -based integrated bulk-charge model can be expressed by

$$Q_{b,0} = Q_{b,acc} + Q_{b,ds}, \quad (4.2.22)$$

which is shown in figure 4.4.

Figure 4.4: Capacitances from numerical derivative of $Q_{b,0}$ with their components and charges shown in the inset.

4.2.3 Charge modeling with V_{ds} dependence

A “non-pinned” ϕ_s -based unified single-piece bulk-charge model in the above sections has been formulated without V_{ds} variation. Even with $V_{ds} \neq 0$, the developed

bulk-charge model, Q_{b0} is still valid because the free electrons can be neglected in accumulation and depletion regions, and changes of the quasi-Fermi level due to V_{ds} will not change the free holes and fully ionized acceptors contributions in the space-charge region.

However, in the strong inversion region, the electron contribution in the space charge region highly depends on its quasi-Fermi level. A positive V_{sb} at source relative to substrate lowers the electron quasi-Fermi in the source by qV_{sb} relative to the Fermi level. Similarly, at the drain end with positive V_{db} , the electron quasi-Fermi level is lowered by qV_{db} relative to the Fermi level. If we define $V_{cb}(y)$ as the position-dependent channel potential, surface-potential, $\phi_s(y) = 2\phi_f + \Delta + V_{cb}(y)$ will be defined as the position-dependent surface-potential along the channel, where Δ is the gradual increase of the surface-potential in the strong inversion region.

In our modeling methodology, we define two boundary conditions, $\phi_s(0) = (2\phi_f + \Delta_s + V_{sb})$ and $\phi_s(L) = (2\phi_f + \Delta_d + V_{db})$ at the source and drain ends, respectively. The position-dependent drain, source and bulk charges are integrated in strong inversion region, resulting in the total drain, source and bulk charges as expressed in equations (4.2.5) to (4.2.7), respectively. As shown, the first term in the expressions is always V_{ds} independent. Unlike the ϕ_s -based model [3], [30], [50], it is a complicated charge expression as every term in the expression contains the average/difference of surface-potential at the source and drain ends, which requires high precision of ϕ_s solutions.

Figure 4.5: Bulk-capacitance and bulk-charge (inset) with V_{ds} dependence from accumulation to strong inversion with smooth transition across regions of operation.

Adding both the unified regional $Q_{b,STR}$ in the strong inversion region and single-piece full region bulk-charge Q_{b0} models, one can obtain a unified single-piece bulk-charge model with $D_{b,\Delta}$ for non-zero V_{ds} contribution

$$Q_b = Q_{b0} - C_{ox}D_{b,\Delta} = -C_{ox} \left(V_{GBR} - \phi_{ACC} + V_{GBF} - V_{GBF}^* + \gamma \sqrt{\phi_{DS,s}} + D_{b,\Delta} \right). \quad (4.2.23)$$

A comparison of the model with numerical Medici data is shown in figure 4.5.

4.3 Charge model with poly-doping effect

AC electrical characteristics of an n^+ implanted n -channel MOSFET (n^+/p^+ dual-gate CMOS [34]) vary with poly-gate doping concentration, N_{gate} . In the depletion and strong inversion regions, poly-depletion effect (PDE) has been well known since its experimental observation [37]. In the accumulation region, an accumulation layer is formed at both the gate/oxide and oxide/silicon interfaces, and poly-accumulation effect (PAE) is observed, causing the gate-capacitance “cross-over” near the flat-band [40]. When the gate is biased above “gate capacitance recovery” voltage (V_{rec}) [37], [38],

carrier inversion occurs at both the gate/oxide and channel/oxide interfaces, in which the poly inversion effect (PIE) is observed.

Extensive theoretical models have been developed in the literature with the threshold-voltage-based approach [36], [52] or the surface-potential-based approach [3], [35]. However, all existing models are limited by the regional PDE model that is only applicable above flat-band in normal operating regions (i.e., neither PAE nor PIE included). In this section, we present a methodology to model PAE/PDE/PIE for all regions through a unified regional formulation.

4.3.1 Poly-doping effect with “non-pinned” ϕ_s -based approach

The drain current without poly-doping effect has been derived physically in the above sections. A methodology to transform the “pinned” ϕ_s -based drain current formulation to the “non-pinned” ϕ_s -based with simple replacements of threshold voltage, bulk charge factor and drain-source voltage expressions has been discussed. Similar approach is adopted in this section to derive a drain-current equation with poly-doping effect to account for the voltage drop across the induced layer formed at the gate interface, $\phi_p(y)$.

The position-dependent inversion charge has been expressed in (4.1.1b). Integrated drain current can be expressed as

$$I_{ds} = -\frac{\mu W}{L} \int Q_i(y) d\phi, \quad (4.3.1)$$

$$= \frac{\mu W}{L} \left[\int Q_g(y) d\phi + Q_{ox} d\phi + \int Q_b(y) d\phi \right]$$

in which the first term can be formulated with the boundary conditions $\phi_{pL} = \phi_{str,d} + \phi_{p,str,d}$

and $\phi_{p0} = \phi_{str,s} + \phi_{p,str,s}$

$$\begin{aligned}
& \int_{\phi_{p0}}^{\phi_{pL}} [Q_g(y) + Q_{ox}] d\phi \\
& = C_{ox} \int_{\phi_{p0}}^{\phi_{pL}} [V_{gb} - V_{fb} - \phi(y)] d\phi \\
& = C_{ox} \left\{ (V_{gb} - V_{fb}) V_{ds,\Delta p} - \frac{1}{2} [V_{ds,\Delta p}^2 + 2V_{ds,\Delta p} (\phi_{str,s} + \phi_{p, str,s})] \right\}
\end{aligned} \tag{4.3.1a}$$

and the second term can be formulated with the boundary conditions $\phi_{sL} = \phi_{str,d}$ and $\phi_{s0} =$

$$\begin{aligned}
& \phi_{str,s}, \\
& \int_{\phi_{s0}}^{\phi_{sL}} Q_b(y) d\phi \\
& = -C_{ox} \int_{\phi_{s0}}^{\phi_{sL}} \gamma \sqrt{\phi_s(y)} d\phi \\
& = -C_{ox} \left[(V_{ds} + \Delta_{ds}) \gamma \sqrt{\phi_{str,s}} + \frac{1}{2} \left(\frac{\gamma}{2\sqrt{\phi_{str,s}}} \right) (V_{ds} + \Delta_{ds})^2 \right] \\
& = -C_{ox} \left[(V_{ds,\Delta p} - \Delta_{p,ds}) \gamma \sqrt{\phi_{str,s}} + \frac{1}{2} (A_{b,\Delta} - 1) (V_{ds,\Delta p}^2 + \Delta_{p,ds}^2 - 2\Delta_{p,ds} V_{ds,\Delta p}) \right] \\
& = -C_{ox} \left[V_{ds,\Delta p} \gamma \sqrt{\phi_{str,s}} + \frac{1}{2} (A_{b,\Delta} - 1) V_{ds,\Delta p}^2 - (A_{b,\Delta} - 1) \left(1 + \frac{2\sqrt{\phi_{str,s}}}{V_{ds,\Delta p}} - \frac{\Delta_{p,ds}}{2V_{ds,\Delta p}} \right) \Delta_{p,ds} V_{ds,\Delta p} \right] \\
& = -C_{ox} \left[V_{ds,\Delta p} \gamma \sqrt{\phi_{str,s}} + \frac{1}{2} (A_{b,\Delta} - 1) V_{ds,\Delta p}^2 - (A_{b,\Delta} - 1) (1 + \sigma_p) \Delta_{p,ds} V_{ds,\Delta p} \right] \\
& \sigma_p = \frac{2\sqrt{\phi_{str,s}}}{V_{ds,\Delta p}} - \frac{\Delta_{p,ds}}{2V_{ds,\Delta p}}
\end{aligned} \tag{4.3.1b}$$

The integrated drain current can be formulated with the poly-doping effect,

$$\begin{aligned}
I_{ds} & = \frac{\mu W C_{ox}}{L} \left\{ (V_{gb} - V_{fb}) V_{ds,\Delta p} - \frac{1}{2} [V_{ds,\Delta p}^2 + 2V_{ds,\Delta p} (\phi_{str,s} + \phi_{p, str,s})] \right. \\
& \quad \left. - \left[V_{ds,\Delta p} \gamma \sqrt{\phi_{str,s}} + \frac{1}{2} (A_{b,\Delta} - 1) V_{ds,\Delta p}^2 - (A_{b,\Delta} - 1) (1 + \sigma_p) \Delta_{p,ds} V_{ds,\Delta p} \right] \right\} \\
& = \mu W C_{ox} \left\{ (V_{gs} - V_{tp,\phi}) V_{ds,\Delta p} - \frac{1}{2} A_{b,\Delta} V_{ds,\Delta p}^2 \right\}
\end{aligned} \tag{4.3.2}$$

where

$$V_{tp,\phi} = V_{t,\phi} + \phi_{p0} + \Delta_{peff,s} - (A_{b,\Delta} - 1) \Delta_{peff,ds} \tag{4.3.2a}$$

is the “non-pinned” ϕ_s -based threshold voltage with poly-doping effect (assumed $\sigma_p \approx 0$).

$\Delta_{peff,s}$ and $\Delta_{peff,ds}$ solutions approach asymptotic $\Delta_{p,s}$ and $\Delta_{p,ds}$ solutions in the strong inversion region, respectively, which are used in the final integrated drain current instead of $\Delta_{p,s}$ and $\Delta_{p,ds}$. The “non-pinned” ϕ_s -based drain-source voltage is expressed as:

$$V_{ds,\Delta p} = V_{ds,\Delta} + \Delta_{peff,ds}, \quad (4.3.2b)$$

$$\Delta_{peff,ds} = \Delta_{peff,d} - \Delta_{peff,s}, \quad (4.3.2c)$$

$$\Delta_{peff} = \phi_{peff} - \phi_{p0}. \quad (4.3.2d)$$

Similarly, the drain current can be obtained from

$$I_{ds} = - \int_{V(0)}^{V(L)} [Q_{ip}(V)] dV, \quad (4.3.3)$$

$$= \mu W C_{ox} \left\{ (V_{gs} - V_{tp,\phi}) V_{ds,\Delta p} - \frac{1}{2} A_{b,\Delta} V_{ds,\Delta p}^2 \right\}$$

with voltage-dependent inversion charge

$$Q_{ip}(V) = - \left\{ C_{ox} [V_{gs} - V_{tp,\phi} - A_{b,\Delta} V] \right\} \quad (4.3.3a)$$

and the boundary conditions, $V(L) = V_{ds,\Delta p}$ and $V(0) = 0$ at the drain and source ends, respectively.

4.3.2 A unified poly-depletion/inversion effect for intrinsic drain, source and bulk charge models

A “non-pinned” drain-current expression has been derived in the previous section, which gives a similar expression with simple replacements of the threshold-voltage ($V_t \rightarrow V_{t,\phi}$), bulk-charge factor ($A_b \rightarrow A_{b,\Delta}$) and drain-source voltage ($V_{ds} \rightarrow V_{ds,\Delta}$) in the “pinned” drain-current expression. If the potential (ϕ_p) across the induced layer at the gate/oxide interface for the poly-doping effect is considered, similar replacements of the threshold voltage ($V_{t,\phi} \rightarrow V_{tp,\phi}$), and drain-source voltage ($V_{ds,\Delta} \rightarrow V_{ds,\Delta p}$) are observed, yielding the similar

expressions.

Similar replacements of the threshold-voltage and drain-source voltage are suggested for our “non-pinned” ϕ_s -based unified intrinsic charge models and are listed in the following:

$$Q_d = -C_{ox} \left(\frac{V_{gt,p,\phi}}{2} - \frac{A_{b,\Delta}}{3} V_{deff,\Delta p} + A_{d,\Delta p} \right) \quad (4.3.4)$$

$$Q_s = -C_{ox} \left(\frac{V_{gt,p,\phi}}{2} - \frac{A_{b,\Delta}}{6} V_{deff,\Delta p} + A_{s,\Delta p} \right) \quad (4.3.5)$$

$$Q_{b,str} = -C_{ox} \left(\gamma \sqrt{\phi_{ds,s}} + D_{b,\Delta p} \right) \quad (4.3.6)$$

and the auxiliary functions:

$$A_{s,\Delta p} = \mathcal{A}_{\Delta p} (1 - \mathcal{B}_{\Delta p}) \quad (4.3.7)$$

$$A_{d,\Delta p} = \mathcal{A}_{\Delta p} \mathcal{B}_{\Delta p} \quad (4.3.8)$$

$$D_{b,\Delta p} = (A_{b,\Delta} - 1) V_{deff,\Delta p} \mathcal{D}_{\Delta p} \quad (4.3.9)$$

$$\mathcal{A}_{\Delta p} = \frac{A_{b,\Delta}^2 V_{deff,\Delta p}^2}{12 (V_{geff,\Delta p} - 0.5 A_{b,\Delta} V_{deff,\Delta p})} \quad (4.3.10)$$

$$\mathcal{B}_{\Delta p} = \frac{5 V_{geff,\Delta p} - 2 A_{b,\Delta} V_{deff,\Delta p}}{10 (V_{geff,\Delta p} - 0.5 A_{b,\Delta} V_{deff,\Delta p})} \quad (4.3.11)$$

$$\mathcal{D}_{\Delta p} = \frac{3 V_{geff,\Delta p} - 2 A_{b,\Delta} V_{deff,\Delta p}}{6 (V_{geff,\Delta p} - 0.5 A_{b,\Delta} V_{deff,\Delta p})}. \quad (4.3.12)$$

V_{geff} in (4.1.18) is modified to

$$V_{geff,\Delta p} = \frac{2 n v_{th} \ln \left(1 + e^{(V_{gs} - V_{tp,\phi}) / (2 n v_{th})} \right)}{1 + 2 n (C_{ox} / C_d) e^{-(V_{gs} - V_{tp,\phi} - 2 V_{off}) / (2 n v_{th})}} \quad (4.3.13)$$

and $V_{gt,\phi}$ in (4.2.13) and $V_{deff,\Delta}$ in (4.2.4) are modified to be

$$V_{gt,p,\phi} = n v_{th} \ln \left(1 + e^{(V_{gs} - V_{tp,\phi}) / (n v_{th})} \right) \quad (4.3.14)$$

and

$$V_{\text{deff},\Delta p} = V_{\text{deff},\Delta} + \Delta_{\text{peff},ds} \cdot \quad (4.3.15)$$

Figure 4.6: A “non-pinned” ϕ_s -based model with (solid lines) and without (dotted lines) poly-doping effect for unified inversion (circle), drain (triangle) and source (square) capacitances and charges (inset) at low N_{gate} doping concentration.

Figure 4.7: The simulated “non-pinned” ϕ_s -based inversion capacitance, C_{gc} with (solid lines) and without (dashed lines) PIE, and without poly-doping effect (dotted lines) for three different body biases.

Figure 4.6 shows the total inversion, drain and source charges (inset) and their respective capacitances with (solid lines) and without (dotted lines) poly-doping effect for poly gate doping $N_{\text{gate}} = 5 \times 10^{19} \text{ cm}^{-3}$ at $V_{ds} = 0.6\text{V}$, compared with the numerical

Medici data. Our unified “non-pinned” ϕ_s solutions match the Medici solutions with the same set of model parameter and N_{gate} . If the model is extended beyond normal operating range as shown in figure 4.7, the model with only the poly-depletion effect ($\phi_{p,sub}$, dashed lines) clearly cannot predict “capacitance-recovery” effect because the inverted holes at the gate interface can not be modeled with $\phi_{p,sub}$. While with PAE/PDE/PIE (ϕ_{peff}), the total inversion charge can be modeled accurately as shown in figure 4.7, in which the gate capacitance for three difference V_{bs} biases match very well with the numerical solutions. Dotted line in the figure shows the total inversion capacitance behavior with “non-pinned” ϕ_s solutions without poly-doping effect.

4.3.3 A single-piece unified bulk-charge model with poly-doping effect

In the previous section, a single-piece “non-pinned” ϕ_s -based bulk-charge model has been derived without considering poly-doping effect. Strictly speaking, the potential drop across the induced layer at the poly-gate interface cannot be neglected when the poly gate is doped at low doping concentration. A simple replacement of the terms in the expression as discussed is adopted in our explicit single-piece bulk-charge model with poly-doping effect,

$$Q_b = Q_{b0} - C_{ox} D_{b,\Delta p} = -C_{ox} \left(V_{gbr} - \phi_{acc} + V_{gbf} - V_{gbf}^* + \gamma \sqrt{\phi_{ds,s}} + D_{b,\Delta p} \right) \quad (4.3.16)$$

with

$$Q_{b,0} = Q_{b,acc} + Q_{b,ds} \quad (4.3.16a)$$

being the V_{ds} -independent term. A “non-pinned” ϕ_s -based bulk-charge formulation in the accumulation region can be expressed as

$$Q_{b,acc} = -C_{ox} [V_{gbr} - \phi_{acc}] \quad (4.3.16b)$$

and from depletion to strong inversion regions

$$Q_{b,ds} = -C_{ox} \left(V_{gba} - V_{gbf} + \gamma \sqrt{\phi_{ds,s}} \right), \quad (4.3.16c)$$

respectively, with the forward unified model

$$V_{gbf} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{peff,s}; \sigma_f \right\} \quad (4.3.16d)$$

$$V_{gba} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{peff,s}; \sigma_a \right\} \quad (4.3.16e)$$

and the reverse unified model

$$V_{gbr} = \mathcal{G}_r \left\{ V_{gb} - V_{fb} - \phi_{peff,s}; \sigma_a \right\}. \quad (4.3.16f)$$

The second term of (4.3.16), $D_{b,\Delta p}$ describes the V_{ds} -dependent bulk-charge behavior in the strong inversion region.

Figure 4.8: Bulk capacitance solutions with single-piece bulk charge formulation at low and high poly doping concentrations, compared to the numerical solutions. The dotted lines show the bulk capacitance solutions without poly accumulation effect.

Numerical bulk-capacitance data are generated from Medici at two poly-gate doping, $N_{gate} = 10^{22}$ (circle) and $5 \times 10^{19} \text{ cm}^{-3}$ (triangle) and the capacitance “cross-over” is observed from Medici data, as shown in figure 4.8. Our “non-pinned” ϕ_s -based bulk-charge model closely agrees with the Medici data. However, the model without poly-

accumulation effect (dotted line) will overestimate bulk capacitance at low N_{gate} doping concentration in the accumulation region.

4.4 Charge conservation

Q_i can be expressed as the addition of both Q_d and Q_s . Q_g can be expressed as the sum of all charges, Q_i , Q_b and Q_{ox} . Even with these simple additions, care has to be taken of for charge conservation especially at flat-band, which is normally overlooked in the literature. As reported in [25], the channel potential is flatten at flat-band, in which no charge is induced in the channel but at the gate interface, gate charge is offset by the amount of Q_{ox} . This is the important criteria for our unified regional “non-pinned” ϕ_s -based model; ϕ_{sa} model is tuned with σ_f to obtain zero surface-potential at $V_{gb} = V_{fb}$, in which charge neutrality can always be preserved without any further tuning, even with various N_{gate} variation, as shown in figure 4.9.

Figure 4.9: Simulated intrinsic gate, bulk and inversion charge solutions for N_{gate} variation (inset), shows charge conservation for all regions and at flat-band (inset) condition.

4.5 Charge model with quantum-mechanical effect

Quantum-mechanical effect on the MOS structure has been extensively studied. Several QME models have been proposed and the comparison among those models has been summarized [54]. Several approaches to integrate the QME models into the AC compact model have been proposed in the literature. In the ϕ_s -based approach, for example, the modification of the intrinsic carrier density is proposed by [55], [56], in which surface potential can be solved iteratively or explicitly with QME. In the threshold-voltage-based model, a threshold-voltage shift is suggested to account for the QME [57]. Another approach has been suggested to account for capacitance attenuation due to QME through effective oxide thickness [58] or parabolic potential approximation [59].

Our “non-pinned” ϕ_s -based unified charge modeling methodology assumes the QME as an additional physical effect, which can be added in a modular sequence and the poly-doping effect can be coupled with QME in a physical way. So far in the literature, the coupling effect from both the QME and the poly-depletion effect has not been considered. Detailed model formulation is presented in the following.

A single-piece bulk-charge model without QME have been formulated with poly-doping effect as discussed in the previous section. Surface-potential models with both the QME and the poly-depletion effect have been discussed in the previous chapter. A single-piece bulk charge model for all regions is proposed

$$Q_b^{qm} = Q_{b0}^{qm} - C_{ox} D_{b,\Delta p}^{qm} = -C_{ox} \left(V_{gbr}^{qm} - \phi_{acc}^{qm} + V_{gbf}^{qm} - V_{gbf}^{*qm} + \gamma \sqrt{\phi_{ds,s}^{qm}} + D_{b,\Delta p}^{qm} \right) \quad (4.5.1)$$

with simple replacement of the unified regional surface potential models from the one without QME to the one with both coupled effects. The first term of Q_b^{qm} is the V_{ds} -independent term

$$Q_{b,0}^{qm} = Q_{b,acc}^{qm} + Q_{b,ds}^{qm}, \quad (4.5.1a)$$

in which

$$Q_{b,acc}^{qm} = -C_{ox} \left[V_{gbr}^{qm} - \phi_{acc}^{qm} \right] \quad (4.5.1b)$$

gives asymptotic charge solutions in the accumulation region and

$$Q_{b,ds}^{qm} = -C_{ox} \left(V_{gbf}^{qm} - V_{gbf}^{*qm} + \gamma \sqrt{\phi_{ds,s}^{qm}} \right) \quad (4.5.1c)$$

gives another asymptotic charge solutions from depletion to strong inversion. The QME forward unified models are defined as

$$V_{gbf}^{qm} = V_{gbf} \Big|_{\phi_{peff,s} \rightarrow \phi_{peff,s}^{qm}} \quad (4.5.1d)$$

$$V_{gba}^{qm} = V_{gba} \Big|_{\phi_{peff,s} \rightarrow \phi_{peff,s}^{qm}} \quad (4.5.1e)$$

and reverse unified model

$$V_{gbr}^{qm} = V_{gbr} \Big|_{\phi_{peff,s} \rightarrow \phi_{peff,s}^{qm}}. \quad (4.5.1f)$$

The second term of Q_b^{qm} ,

$$D_{b,\Delta p}^{qm} = (A_{b,\Delta}^{qm} - 1) V_{deff,\Delta p}^{qm} \mathcal{D}_{\Delta p}^{qm} \quad (4.5.2)$$

consists of all V_{ds} -dependent charge formulation, in which

$$\mathcal{D}_{\Delta p}^{qm} = \frac{3V_{geff,\Delta p}^{qm} - 2A_{b,\Delta}^{qm} V_{deff,\Delta p}^{qm}}{6(V_{geff,\Delta p}^{qm} - 0.5A_{b,\Delta}^{qm} V_{deff,\Delta p}^{qm})}. \quad (4.5.3)$$

The $V_{geff,\Delta p}$ function is modified to couple both poly-doping and quantum-mechanical effects as

$$V_{geff,\Delta p}^{qm} = \frac{2nV_{th} \ln \left(1 + e^{\left(V_{gs} - V_{tp,\phi}^{qm} \right) / (2nV_{th})} \right)}{1 + 2n \left(C_{ox} / C_d \right) e^{-\left(V_{gs} - V_{tp,\phi}^{qm} - 2V_{off} \right) / (2nV_{th})}} \quad (4.5.4)$$

with the “non-pinned” ϕ_s -based threshold-voltage definition

$$V_{tp,\phi}^{qm} = V_{t,\phi}^{qm} + \phi_{po} + \Delta_{peff,s}^{qm} - \left(A_{b,\Delta}^{qm} - 1 \right) \Delta_{peff,ds}^{qm} \quad (4.5.5)$$

$$V_{t,\phi}^{qm} = V_{fb} + \phi_{s0} + \Delta_s^{qm} + \gamma \sqrt{\phi_{ds,s}^{qm}} \quad (4.5.5a)$$

$$\Delta_{peff,ds}^{qm} = \Delta_{peff,ds} \Big|_{\phi_{peff} \rightarrow \phi_{peff}^{qm}} \quad (4.5.5b)$$

and

$$V_{deff,\Delta p}^{qm} = V_{deff,\Delta p} \Big|_{\phi_{seff}, \phi_{peff} \rightarrow \phi_{seff}^{qm}, \phi_{peff}^{qm}}. \quad (4.5.6)$$

(a)

(b)

Figure 4.10: Bulk Capacitances for: (a) Q_b^{qm} with its unified regional model below and above flat-band voltage at $V_{ds} = 0$ and (b) Q_b^{qm} with unified regional model for Q_{b0}^{qm} and the V_{ds} dependency term.

Similarly, the drain and source charges are modified from previously formulated poly-doping effect coupled with QME such that

$$Q_d^{qm} = -C_{ox} \left(\frac{V_{gtp,\phi}^{qm}}{2} - \frac{A_{b,\Delta}^{qm}}{3} V_{deff,\Delta p}^{qm} + A_{d,\Delta p}^{qm} \right) \quad (4.5.7)$$

and

$$Q_s^{qm} = -C_{ox} \left(\frac{V_{gtp,\phi}^{qm}}{2} - \frac{A_{b,\Delta}^{qm}}{6} V_{deff,\Delta p}^{qm} + A_{s,\Delta p}^{qm} \right), \quad (4.5.8)$$

respectively, with the corresponding auxiliary functions:

$$A_{s,\Delta p}^{qm} = \mathcal{A}_{\Delta p}^{qm} (1 - \mathcal{B}_{\Delta p}^{qm}) \quad (4.5.9)$$

$$A_{d,\Delta p}^{qm} = \mathcal{A}_{\Delta p}^{qm} \mathcal{B}_{\Delta p}^{qm} \quad (4.5.10)$$

$$\mathcal{A}_{\Delta p}^{qm} = \frac{(A_{b,\Delta}^{qm})^2 (V_{deff,\Delta p}^{qm})^2}{12 (V_{geff,\Delta p}^{qm} - 0.5 A_{b,\Delta}^{qm} V_{deff,\Delta p}^{qm})} \quad (4.5.11)$$

$$\mathcal{B}_{\Delta p}^{qm} = \frac{5 V_{geff,\Delta p}^{qm} - 2 A_{b,\Delta}^{qm} V_{deff,\Delta p}^{qm}}{10 (V_{geff,\Delta p}^{qm} - 0.5 A_{b,\Delta}^{qm} V_{deff,\Delta p}^{qm})}. \quad (4.5.12)$$

The gate and inversion charges, respectively, are defined by

$$Q_g^{qm} = -(Q_i^{qm} + Q_b^{qm} + Q_{ox}) \quad (4.5.13)$$

$$Q_i^{qm} = Q_d^{qm} + Q_s^{qm}. \quad (4.5.14)$$

The transcapacitances are obtained from

$$C_{ij} = \delta_{ij} \frac{\partial Q_i}{\partial V_j}, \quad \delta_{ij} = \begin{cases} 1 & (i = j) \\ -1 & (i \neq j) \end{cases}. \quad (4.5.15)$$

Figure 4.11 shows the various capacitances obtained from the terminal charges including QME, and validated with Medici data with the same QME and poly-gate doping.

(a)

Figure 4.11: Capacitances from the numerical derivative of: (a) Q_g^{qm} (solid line), Q_i^{qm} (dotted line) and Q_b^{qm} (dashed line) charges.

CHAPTER 5: THRESHOLD VOLTAGE MODELING

In this and the next chapters, the approach to formulating a DC compact model suitable for circuit designers and technology developers is presented. Our “pinned” ϕ_s -based (or threshold-voltage-based) DC compact model is streamed into two components, firstly the threshold-voltage model, and secondly the unified drain-current model.

A unified compact threshold-voltage expression has been proposed by [60] which has been derived from the source-referenced drain current formulation. Geometry effects have been considered in the threshold-voltage expression at decreasing channel length to account for the short-channel effect (SCE) based on the source-referenced quasi-two-dimensional model [17]. Reverse SCE has been added in a modular sequence into the expression with the peak lateral doping at both the source and the drain ends. This chapter is the continuation of the work developed by [60] on the existing threshold-voltage expression. The threshold-voltage expression is re-formulated to include the geometry effects at decreasing channel length/width, in which two-dimensional lateral doping profiles and surface-potentials are proposed along the width and the length dimensions [15] to account for both the narrow-width effect (NWE) and inverse NWE. Model symmetry is another important aspect which has not been discussed previously and will be addressed in this thesis.

In this work, a bulk-referred threshold-voltage expression is physically formulated. The model is derived in such a way to be characterized from the technology information, such as the technology curves (point data or extracted from curves based on [61], [62]) at decreasing channel length/width with minimum measurement data requirement. Our modeling approach is able to predict electrical characteristics of the

MOSFET at decreasing channel length/width physically with a set of model parameters rather than the empirical approach of joining difference sets of model parameters over geometry variations, as is normally adopted in the literature [2], [63].

Our threshold-voltage model requires less effort in the drain-current formulation as the geometry effect has been considered in threshold-voltage formulation. Threshold voltage is an important parameter, not only used in the drain current formulation but also used frequently by technology developers and industries to gauge the MOSFET electrical performance on the fabricated devices.

5.1 Conventional long-channel threshold-voltage model (source-referenced model)

From theoretical definition, threshold voltage is a special gate voltage at which the channel surface potential is equal to twice of the Fermi potential

$$\phi_{s0} = 2\phi_f = 2v_{th} \ln\left(\frac{N_{sub}}{ni}\right), \quad (5.1.1)$$

where v_{th} , N_{sub} and ni are the thermal voltage, substrate and intrinsic carriers concentration, respectively. This definition is referred to as the *classical criterion for strong inversion*. Another V_t definition, which gives better estimate of the measured threshold voltage, is based on the surface potential reaching $2\phi_f + 6v_{th}$ [64]. However in most of the circuit models, it is preferred to use the *classical criterion* definition, and the conventional threshold voltage based on this definition is discussed.

When the gate terminal is biased at the threshold condition, given that the device is the ideal MOSFET structure, from potential balance equation, the threshold voltage can

be expressed as

$$V_{tL,i} = 2\phi_f + V_{ox} \quad (5.1.2)$$

in which the voltage drop across gate oxide is $V_{ox} = Q_g/C_{ox}$. In real MOSFET devices with the metal gate, the work function between gate and bulk is different and there is oxide charge trapped in the gate oxide, Q_{ox} . The total amount of gate voltage required to offset the potential difference due the different work function between gate and bulk can be calculated from

$$\phi_{ms} = V_{fbb} - \phi_f \quad (5.1.3)$$

$$V_{fbb} = \phi_m - (\chi + E_g/2). \quad (5.1.3a)$$

In the above equation, ϕ_m represents the work function for the gate material, χ and E_g represent the electron affinity and the band-gap of silicon, respectively. From the charge neutrality principle, $Q_g + Q_{sc} + Q_{ox} = 0$, where channel induced charge at threshold condition (when $\phi_s = \phi_{s0} + V_{sb}$) can be expressed as

$$Q_{sc} \approx Q_b \approx -C_{ox}\gamma\sqrt{\phi_{s0} + V_{sb}} \quad (5.1.4)$$

with the body factor $\gamma = \sqrt{2q\epsilon_{si}N_{sub}}/C_{ox}$. The non-ideal threshold voltage equation can be expressed as

$$\begin{aligned} V_{tL} &= \phi_{ms} + \phi_{s0} - \frac{Q_{sc} + Q_{ox}}{C_{ox}} \\ &= \phi_{ms} - Q_{ox} + \phi_{s0} + \gamma\sqrt{\phi_{s0} + V_{sb}} \\ &= V_{FB} + \phi_{s0} + \gamma\sqrt{\phi_{s0} + V_{sb}} \end{aligned} \quad (5.1.5)$$

and V_{FB} is the flat band voltage of the device.

Likewise another approach can be taken to formulate a conventional V_t expression through source-referenced drain current derivation (4.2.1) with the inversion charge in strong inversion region given by

$$Q_i = -C_{ox} \left(V_{gb} - V_{FB} - \phi_s(y) - \gamma \sqrt{\phi_s(y)} \right) \quad (5.1.6)$$

and the integrated current is expressed as

$$\begin{aligned} I_{ds} &= -\frac{\mu W}{L_{eff}} \int_{\phi_{s0} + V_{sb}}^{\phi_{s0} + V_{db}} Q_i(y) d\phi \\ &= \frac{\mu W C_{ox}}{L_{eff}} \left\{ (V_{gs} - V_{tL}) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\} \end{aligned} \quad (5.1.7)$$

$$L_{eff} = L_g - 2\sigma X_j \quad (5.1.7a)$$

$$W = W_g - \Delta W \quad (5.1.7b)$$

with μ and A_b are the mobility coefficient and bulk-charge factor, respectively. σ and ΔW are two optimization parameters.

5.2 Bulk-referenced threshold voltage model

It is well-known that the disadvantage of the above source-referenced model with fitted parameter is the loss of Gummel symmetry [27], [13]. The source of asymmetry in the DC electrical behavior, however, can be avoided since it can be shown (p. 169 in [64]) that (5.1.7) is algebraically equivalent to the symmetric “*bulk-referenced*” (“B-ref”) model in which V_{tL} and A_b are given by

$$V_{tL} = V_{FB} + \phi_{s0} + Q_{BLC} \quad (5.2.1)$$

$$Q_{BLC} = \gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \quad (5.2.1a)$$

$$A_b = 1 + \gamma / (2\sqrt{\phi_{s0}}). \quad (5.2.1b)$$

Our geometry-dependent threshold-voltage model is formulated from analytic B-ref symmetric unified drain current model, which will be discussed in the following

chapter. The starting ground of our geometry-dependent V_t expression is based on the ideal long-channel equation, physical geometry effects are added in a modular sequence.

5.2.1 Vertical doping transformation for non-ideal MOSFET structure

The formulated ideal long-channel V_t expression for B-ref I_{ds} model assumes that the channel doping follows a uniform profile based on ideal MOSFET structure, in which ϕ_{s0} and γ are calculated based on a constant doping. In non-ideal MOSFETs, implantation steps are taken for preventing source/drain punch-through, adjusting the threshold voltage, preventing V_t roll-off with halo implantation, thus doping concentration in the channel is no longer uniform. Many attempts have been reported to perform doping transformation from non-uniform doping profile to an effective channel doping concentration [65] – [69]. In our V_t expression, non-uniform vertical channel doping with Gaussian profile is formulated to be the average vertical doping concentration in the channel [70],

$$N_{ch} = N_{tot} (1 - \eta R_p^2) \quad (5.2.2)$$

where

$$N_{tot} = N_{sub} + N_{pk} \quad (5.2.2a)$$

$$\eta = \frac{I}{3c_2} \left(I - \frac{N_{sub}}{N_{tot}} \right). \quad (5.2.2b)$$

c_2 is a process-dependent fitting parameter and N_{pk} is the peak concentration of the Gaussian's profile. The vertical doping transformation is performed such as N_{sub} is replaced by N_{ch} in

$$\phi_{s0} = 2\phi_f = 2v_{th} \ln \left(\frac{N_{ch}}{n_i} \right) \quad (5.2.3a)$$

and

$$\gamma = \sqrt{2q\epsilon_{si}N_{ch}}/C_{ox} \quad (5.2.3b)$$

5.2.2 Long-channel drain induced barrier lowering (DIBL)

Our B-ref threshold-voltage expression (5.2.1) accurately models the measured threshold-voltage as shown in the figure 5.1 at $V_{ds} = 0.1V$ (solid line). When $V_{ds} = 1.32V$, a minor threshold-voltage drop is observed with the measurement data but the expression (5.2.1) cannot model this behavior without the V_{ds} -dependent term in the expression. One solution to include this effect is based on the quasi-two-dimensional model [17] with the Gaussian box applied in the channel depletion region of height X_{dep} and length Δy :

$$\epsilon_{si} \frac{X_{dep}}{\theta} \frac{dE_s(y)}{dy} + \epsilon_{ox} \frac{V_{gs} - V_{FB} - V_{ch}(y)}{T_{ox}} = qN_s X_{dep} \quad (5.2.4)$$

where $V_{ch}(y)$ is the channel potential, $E_s(y)$ is the lateral surface electric field and θ is an empirical fitting parameter. An “asymmetric” S-ref solution from equation (5.2.4) with boundary conditions $V_{ch}(0) = V_{bi}$ and $V_{ch}(L_{eff}) = V_{bi} + V_{ds}$ [17] at the source and drain ends, respectively, is given

$$V_{ch}(y) = V_{sL} + (V_{bi} + V_{ds} - V_{sL}) \frac{\sinh(y/l_\alpha)}{\sinh(L_{eff}/l_\alpha)} + (V_{bi} - V_{sL}) \frac{\sinh\{(L_{eff} - y)/l_\alpha\}}{\sinh(L_{eff}/l_\alpha)} \quad (5.2.5)$$

where $V_{sL} = V_{gs} - V_{tL} + \phi_{s0}$ represents the long channel surface potential, V_{tL} is the threshold voltage from (5.2.1) without long-channel DIBL effect, V_{bi} is the built-in potential and the characteristic length, l_α is taken as process-dependent parameter. In

order to obtain a symmetric $V_{ch}(y)$, two boundary conditions with $V_{ch}(0) = V_{bi} + V_{sb}$ and $V_{ch}(L_{eff}) = V_{bi} + V_{db}$ are proposed and (5.2.1) is re-written as

$$V_s(y) = V_{sL} + (V_{bi} + V_{db} - V_{sL}) \frac{\sinh(y/l_\alpha)}{\sinh(L_{eff}/l_\alpha)} + (V_{bi} + V_{sb} - V_{sL}) \frac{\sinh\{(L_{eff} - y)/l_\alpha\}}{\sinh(L_{eff}/l_\alpha)}. \quad (5.2.6)$$

For long-channel device, the minimum channel potential is approximately located at $y = L_{max}/2$ even with V_{sb} and V_{db} variations. The minimum channel surface potential is found to be

$$\begin{aligned} V_{ch,min} &= V_{sL} + (V_{bi} + V_{db} - V_{sL}) \frac{\sinh(L_{max}/2l_{\alpha,l})}{\sinh(L_{max}/l_{\alpha,l})} + (V_{bi} + V_{sb} - V_{sL}) \frac{\sinh\{L_{max}/2l_{\alpha,l}\}}{\sinh(L_{max}/l_{\alpha,l})} \\ &= V_{sL} + \left[2(V_{bi} - V_{sL} - V_{bs}) + V_{ds} \right] \frac{\sinh(L_{max}/2l_{\alpha,l})}{\sinh(L_{max}/l_{\alpha,l})} \\ &= V_{sL} + \frac{\left[2(V_{bi} - V_{sL} - V_{bs}) + V_{ds} \right]}{2\cosh(L_{max}/2l_{\alpha,l})} \end{aligned} \quad (5.2.7)$$

where $l_{\alpha,l}$ is the characteristic length for long-channel device. When gate is biased at the threshold condition, we assume $V_{ch,min} = \phi_{s0}$ and $V_{gs} = V_{tL,DIBL}$. Threshold voltage with DIBL effect can be formulated as

$$V_{tL,DIBL} = V_{tL} - \Delta V_{tL,DIBL} \quad (5.2.8)$$

and the threshold voltage increases due to DIBL effect and can be derived from (5.2.7)

$$\Delta V_{tL,DIBL} = \alpha_L \left[2(V_{bi} - V_{sL} - V_{bs}) + V_{ds} \right] \quad (5.2.8a)$$

$$\alpha_L = \frac{1}{2\cosh(L_{max}/2l_{\alpha,l})}. \quad (5.2.8b)$$

It is shown in the figure 5.1 that the threshold-voltage with DIBL effect gives better solutions (dotted line) due to drain voltage variation. The model parameter variation is shown with the dashed line.

Both equations, (5.2.5) and (5.2.6), give the same solutions with $V_{sb} = 0$ at increasing drain potentials as shown in figure 5.2. However, at increasing source potential (inset), the S-ref expression (5.2.5) gives asymmetric solutions (dotted lines) at decreasing drain surface-potential while B-ref (5.2.6) expression gives symmetric solutions at the source end.

Figure 5.1: Bulk-referenced threshold voltage solutions with long-channel DIBL effect and characteristic length, $l_{\alpha,l}$ variation.

Figure 5.2: Channel surface potential along the channel length with V_{db} and V_{sb} (inset) variation at the drain and source ends.

The Gummel's symmetry test is performed for drain current with both the S-ref [17] and B-ref (5.2.6) modeling approaches. A smooth transition across $V_x = 0$ is shown in figure 5.3 that our $\Delta V_{tL,DIBL}$ based on new bulk-referenced $V_{ch}(y)$ model is fully symmetric (solid lines). A glitch (dotted line in figure 5.3) can be observed in I_d'' at $V_x = 0$ for the Gummel's symmetry test with $\Delta V_{tL,DIBL}$ based on S-ref $V_s(x)$ [17].

Figure 5.3: Gummel symmetry test [65] (with $V_{gb0} = 1.2$ V and $V_{b0} = 0$) of the first (right axis) and the second (left axis) derivatives of I_d versus V_x for the bulk-referenced threshold voltage based on bulk-referenced long channel DIBL (solid lines) and source-referenced long channel DIBL [17] (dotted lines).

5.3 Short-channel effects

5.3.1 Charge-sharing effect

The depletion-charge modulation effect has been suggested in (5.2.1a) to model long-channel threshold voltage with source terminal potential variation. However, when the channel length decreases, the model cannot predict the electrical characteristics of a short-channel MOSFET with fixed Q_{BLC} . A trapezoidal approximation in the depletion

region along the channel is proposed (see: p. 262 of [64]). In Ref. [25] (p. 196), the threshold voltage is modified

$$\begin{aligned} V_{t,CS} &= V_{FB} + \phi_{s0} + \frac{Q_{BSC}}{C_{ox}} + \Delta V_{tL,DIBL} \\ &= V_{FB} + \phi_{s0} + Q_{BLC} \frac{Q_{BSC}}{Q_{BLC}} + \Delta V_{tL,DIBL} \end{aligned} \quad (5.3.1)$$

in which the depletion charge ratio of the short to long-channel can be derived based on trapezoidal approximation, as shown in figure 5.4,

$$\frac{Q_{BSC}}{Q_{BL}} = 1 - \frac{\lambda}{L_{eff}} \frac{X_{dep,s} + X_{dep,d}}{2}. \quad (5.3.2)$$

The depletion widths, $X_{dep,s}$ and $X_{dep,d}$, can be calculated at the source and drain ends, respectively,

$$X_{dep,s} = \zeta \sqrt{\phi_{s0} + V_{sb}} \quad (5.3.2a)$$

$$X_{dep,d} = \zeta \sqrt{\phi_{s0} + V_{ds} + V_{sb}} \quad (5.3.2b)$$

$$\zeta = \sqrt{\frac{2\epsilon_{si}}{qN_s}}, \quad (5.3.2c)$$

$$L_{eff} = L_g - 2\sigma X_j \quad (5.3.2d)$$

where λ is a process-dependent parameter with a default value of 1 but can be adjusted to have better fitting to the measurement data. In the S-ref V_t -based model, $X_{dep,d}$ is expanded with Taylor's series expansion, which results in

$$\frac{X_{dep,s} + X_{dep,d}}{2} = \zeta \left(\sqrt{\phi_{s0} + V_{sb}} + \delta \frac{V_{ds}}{\sqrt{\phi_{s0} + V_{sb}}} \right) \quad (5.3.3)$$

with a model parameter δ for best fit to the measurement data [16]. However in the B-ref model, V_t is modified as

$$\begin{aligned}
V_{t,CS} &= V_{FB} + \phi_{s0} + \frac{Q_{BSC}}{C_{ox}} + \Delta V_{tL,DIBL} \\
&= V_{FB} + \phi_{s0} + \left[\gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \right] \left[1 - \frac{\lambda \zeta}{L_{eff}} \frac{\sqrt{\phi_{s0} + V_{sb}} + \sqrt{\phi_{s0} + V_{db}}}{2} \right] + \Delta V_{tL,DIBL}
\end{aligned} \tag{5.3.4}$$

δ -fitting parameter in (5.3.3), which destroys the symmetry of the drain current model is not required in B-ref V_t model.

Figure 5.4: Trapezoidal approximation with $V_{db} \neq V_{sb}$ at the drain and source ends, respectively.

5.3.2 Potential barrier lowering

The potential along the channel in y -direction from source to drain is governed by the quasi 2-D solutions [17], in which the potential is not fixed to a constant ϕ_{s0} as in long-channel but varies with channel length, L_{eff} . Based on quasi 2-D model, when $V_{db} = V_{sb}$ with $l_\alpha = \alpha(\phi_{s0})^{0.25}$ as the characteristic length where α is a process-dependent proportionality factor, the minimum surface potential for any channel length is located at $L_{eff}/2$ and thus, quasi 2-D solutions can be simplified as

$$\phi_s(L_{eff}) = \phi_{s0} + \frac{[2(V_{bi} - \phi_{s0} - V_{bs}) + V_{ds}]}{2\cosh(L_{eff}/2l_\alpha)} \quad (5.3.5)$$

and

$$\Delta\phi_s = \frac{[2(V_{bi} - \phi_{s0} - V_{bs}) + V_{ds}]}{2\cosh(L_{eff}/2l_\alpha)} \quad (5.3.6)$$

is defined as the surface-potential increment, which varies with the channel length.

When $V_{db} \neq V_{sb}$, the minimum surface potential is not located at $L_{eff}/2$ in the channel anymore but is shifted. The amount of shift can be calculated by considering that the effective channel length of a device is always much greater than the characteristic length,

l_α such that the potential along the channel from (5.2.6) can be approximated as

$$V_{ch}(y) = \phi_{s0} + (V_{bi} + V_{db} - \phi_{s0})e^{(y-L)/l_\alpha} + (V_{bi} + V_{sb} - \phi_{s0})e^{-y/l_\alpha} + (V_{bi} + V_{db} - \phi_{s0})e^{-L/l_\alpha} \quad (5.3.7)$$

and the minimum point of equation (5.3.7) can be found by taking $dV_{ch}(y)/dy = 0$

$$\frac{dV_{ch}(y)}{dy} = \frac{(V_{bi} + V_{db} - \phi_{s0})}{l_\alpha} e^{(y_{min}-L)/l_\alpha} - \frac{(V_{bi} + V_{sb} - \phi_{s0})}{l_\alpha} e^{-y_{min}/l_\alpha} = 0 \quad (5.3.8)$$

where the minimum potential is found to be

$$y_{min} = \frac{L_{eff}}{2} - \frac{l_\alpha}{2} z \quad (5.3.9)$$

$$z = \ln \frac{V_{bi} + V_{ds} + V_{sb} - \phi_{s0}}{V_{bi} + V_{sb} - \phi_{s0}}. \quad (5.3.9a)$$

The minimum potential can be calculated by substituting y_{min} into equation (5.2.6) and

the surface potential increment from (5.3.6) is modified to

$$\Delta\phi_s = (V_{bi} + V_{ds} + V_{sb} - V_{sL}) \frac{\sinh\left(\frac{L_{eff}}{2l_\alpha} - \frac{z}{2}\right)}{\sinh\left(\frac{L_{eff}}{l_\alpha}\right)} + (V_{bi} + V_{sb} - V_{sL}) \frac{\sinh\left(\frac{L_{eff}}{2l_\alpha} + \frac{z}{2}\right)}{\sinh\left(\frac{L_{eff}}{l_\alpha}\right)}. \quad (5.3.10)$$

Since $\sinh(A \pm B) = \sinh(A)\cosh(B) \pm \cosh(A)\sinh(B)$, $\Delta\phi_s$ is reduced to

$$\begin{aligned}\Delta\phi_s &= \left\{ (V_{bi} + V_{ds} + V_{sb} - V_{sL}) \left(\sinh\left[\frac{L_{eff}}{2l_\alpha}\right] \cosh\left[\frac{z}{2}\right] - \cosh\left[\frac{L_{eff}}{2l_\alpha}\right] \sinh\left[\frac{z}{2}\right] \right) + \right. \\ &\quad \left. (V_{bi} + V_{sb} - V_{sL}) \left(\sinh\left[\frac{L_{eff}}{2l_\alpha}\right] \cosh\left[\frac{z}{2}\right] + \cosh\left[\frac{L_{eff}}{2l_\alpha}\right] \sinh\left[\frac{z}{2}\right] \right) \right\} / \sinh\left[\frac{L_{eff}}{l_\alpha}\right] \quad (5.3.11) \\ &= \left[2(V_{bi} + V_{sb} - V_{sL}) \sinh\left[\frac{L_{eff}}{2l_\alpha}\right] \cosh\left[\frac{z}{2}\right] + V_{ds} \sinh\left[\frac{L_{eff}}{2l_\alpha} - \frac{z}{2}\right] \right] / \sinh\left[\frac{L_{eff}}{l_\alpha}\right]\end{aligned}$$

Since $\frac{2 \sinh(L_{eff}/2l_\alpha)}{\sinh(L_{eff}/l_\alpha)} = \frac{1}{\cosh(L_{eff}/2l_\alpha)}$, we have

$$\Delta\phi_s = \frac{1}{\cosh\left(\frac{L_{eff}}{2l_\alpha}\right)} \left[(V_{bi} + V_{sb} - \phi_{s0}) \cosh\left(\frac{z}{2}\right) + \frac{V_{ds}}{2} \frac{\sinh\left(\frac{L_{eff}}{2l_\alpha} - \frac{z}{2}\right)}{\sinh\left(\frac{L_{eff}}{2l_\alpha}\right)} \right] \quad (5.3.12)$$

In order to match the measurement data for various V_{ds} variation, fitting parameter, φ is proposed and $\Delta\phi_s$ is re-written as

$$\Delta\phi_s = \frac{1}{\cosh\left(\frac{L_{eff}}{2l_\alpha}\right)} \left[(V_{bi} + V_{sb} - \phi_{s0}) \cosh\left(\frac{z}{2}\right) + \frac{\varphi V_{ds}}{2} \frac{\sinh\left(\frac{L_{eff}}{2l_\alpha} - \frac{z}{2}\right)}{\sinh\left(\frac{L_{eff}}{2l_\alpha}\right)} \right] \quad (5.3.13)$$

The V_t model is further modified as

$$\begin{aligned}V_{t,DIBL} &= V_{FB} + \phi_s + \frac{Q_{BSC}}{C_{ox}} + \Delta V_{tL,DIBL} \\ &= V_{FB} + \phi_s + \left[\gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \right] \left[1 - \frac{\lambda \zeta}{L} \frac{\sqrt{\phi_s + V_{sb}} + \sqrt{\phi_s + V_{db}}}{2} \right] + \Delta V_{tL,DIBL} \quad (5.3.14)\end{aligned}$$

in which the length-dependent surface potential is expressed as

$$\phi_s = \phi_{s0} + \Delta\phi_s \quad (5.3.14a)$$

and the bulk-charge factor is re-written as

$$A_b = 1 + \gamma / (2\sqrt{\phi_s}). \quad (5.3.14b)$$

5.3.3 Reverse short-channel effect

V_t roll-off effect at decreasing channel length is commonly known as short-channel effect (SCE), which has been discussed in detail with charge sharing and potential barrier lowering effects. In order to prevent significant V_t roll-off for short-channel devices, the CMOS technology is enhanced with inserting halo-implant step into the process. An additional dopant pile-up at the source and drain ends after the post-LDD annealing changes the effective doping concentration in the channel at decreasing channel length. The effect of the pile-up doping concentration at the two sides of the channel will eventually get closer to each other when channel length decreases. This increase in the effective channel doping concentration causes V_t to increase. Many models have been suggested in the literature with empirical modeling of the lateral doping profile [71], with two step functions [2] as shown in figure 5.5(a), two Gaussian profiles [60] as shown in figure 5.5 (b) at two ends of the channel, or linear approximation of the impurity profile along the lateral dimension [72].

(a) (b)

Figure 5.5: Lateral doping profile modeled as (a) two step functions, (b) two Gaussian profiles, at the two ends of the channel.

Figure 5.6: Two Gaussian profiles with pile-up centroid at two sides of the channels.

In this work, two Gaussian profiles are proposed at the source and drain ends with pile-up centroid, l_μ for the location of the Gaussian's peak, N_{pile} and l_β are the peak and characteristic length for the Gaussian profiles along the channel, as shown in figure 5.6. The effective channel doping concentration is defined as the average channel doping that is contributed from the lateral and vertical doping profiles through the effective doping transformation [15]

$$\begin{aligned}
N_{eff} &= \frac{1}{L_{eff}} \int_0^{L_{eff}} \left\{ N_{pile} \exp \left[- \left(\frac{y - l_\mu}{l_\beta} \right)^2 \right] + N_{pile} \exp \left[- \left(\frac{L_{eff} - (y + l_\mu)}{l_\beta} \right)^2 \right] + N_{ch} \right\} dy \\
&= \frac{N_{pile}}{L_{eff}/l_\beta} \left\{ \left[\int_{-(l_\mu/l_\beta)}^{(L_{eff}-l_\mu)/l_\beta} \exp(-t^2) dt \right] \right. \\
&\quad \left. - \left[\int_{(L_{eff}-l_\mu)/l_\beta}^{-(l_\mu/l_\beta)} \exp(-t^2) dt \right] \right\} + N_{ch} \\
&= \frac{\sqrt{\pi} N_{pile}}{L_{eff}/l_\beta} \left[\operatorname{erf} \left(\frac{L_{eff} - l_\mu}{l_\beta} \right) + \operatorname{erf} \left(\frac{l_\mu}{l_\beta} \right) \right] + N_s
\end{aligned} \tag{5.3.15}$$

where erf is the error function and the Gaussian's peak, N_{pile} and lateral spread, l_β are defined by

$$N_{pile} = \kappa N_{ch} \tag{5.3.15a}$$

$$l_\beta = \beta \phi_{s0}^{0.25}. \tag{5.3.15b}$$

Here, κ is the concentration ratio of the peak Gaussian profile to the average vertical doping concentration and l_β takes the same form as the characteristic length in DIBL without the V_{bs} dependence. Surface-potential and bulk-factor expressions is re-written by replacing N_{ch} with N_{eff} as

$$\phi_{s0} = 2\phi_f = 2v_{th} \ln \left(\frac{N_{eff}}{n_i} \right) \tag{5.3.16a}$$

and

$$\gamma = \sqrt{2q\epsilon_{si} N_{eff}} / C_{ox}, \tag{5.3.16b}$$

respectively.

V_t roll-off/up effect cannot be simulated accurately to match the measurement data without SCE and reverse SCE models as described in the previous sections, and as shown with the dotted line in figure 5.7. V_t model with roll-off effect (SCE) is clearly shown with the dashed lines, roll-up effect (reverse SCE) with the dash-dotted lines, and

the coupled effects (both SCE and reverse SCE) in the final model (solid lines), compared with the measurement data (symbol) at decreasing channel length.

Figure 5.7: Short-channel and reverse short-channel effects at decreasing channel length. Solid lines: coupled effect, dashed lines: SCE, dash-dotted lines: Reverse SCE and dotted line: no SCE and reverse SCE.

Figure 5.8: Simulated channel-doping concentration along the normalized channel length for the device of gate length 10, 1 and 0.13- μm with pile-up centroid variation for 0.04, 0 and 0.04- μm .

Pile-up centroid, l_μ is another physical parameter that indicates the Gaussian's peak location in the channel. Physical interpretation of its effect on channel doping concentration along the channel is clearly shown in figure 5.8. Its effect becomes

significant for short-channel devices with the positive/negative l_μ . It defines the Gaussian's peak location inside/outside the channel at the source or drain ends of the channel, which is related to the amount of lateral Gaussian profiles contributed to the effective doping concentration. The maximum amount of lateral Gaussian profile contribution to the effective doping concentration occurs when l_μ is equal to half of effective channel length, in which the two Gaussian profiles at the source and the drain ends overlap. The significance of l_μ variation on V_t behavior is shown in figure 5.9.

Figure 5.9: Threshold voltage with pile-up centroid variation.

5.4 Narrow-width effect

Experimentally observed V_t roll-off and roll-up effect at decreasing channel length, known as the SCE and reverse SCE, have been discussed and modeled with length-dependent V_t expression. Similar behaviors in V_t roll-off and roll-up at decreasing channel width for trench-isolated CMOS devices have been observed and reported in [73]-[75], which are known as the narrow-width effect (NWE) and inverse NWE. Our previously formulated length-dependent V_t expression is modified to model NWE and inverse NWE at decreasing channel-width, as discussed in the following.

5.4.1 Narrow-width effect and inverse narrow-width effect

The observed long-channel threshold-voltage roll-off at decreasing channel width has been attributed to the fringing field from sidewalls, which contributes to the fringing capacitance that resulted in an increase in the total gate capacitance [29],

$$C_g = C_{ox} + 2C_f \quad (5.4.1)$$

where the fringing capacitance is formulated as

$$C_f = \left(\frac{2\epsilon_{ox}}{W\pi} \right) \ln(2\tau), \quad (5.4.1a)$$

$$\tau = \frac{\tau_p t_d}{t_{ox}}. \quad (5.4.1b)$$

Here, t_d is the depth of recessed oxide and τ_p is a process-dependent fitting parameter. In our V_t formulation, the oxide capacitance C_{ox} is modified with extra fringing capacitance, $2C_f$, in which the first-order V_t model with NWE is clearly shown in figure 5.10 at decreasing channel width with only *one* fitting parameter, τ_p for various V_{ds} and V_{bs} dependence.

Figure 5.10: Long-channel threshold voltage at decreasing channel width with (solid lines) and without (dotted lines) fringing capacitance for various V_{ds} and V_{bs} .

The first-order NWE V_t model simulates reasonably well the measured data for long-channel but not the short-channel devices at decreasing channel width, as shown in figure 5.11 (dotted line). Instead of V_t roll-off at decreasing channel width for long-channel devices, the measurement data show V_t roll-up (inverse NWE) before rolling-off (NWE) for short-channel devices. V_t roll-up effect for short-channel devices has been attributed to the increase in the effective channel doping concentration at decreasing channel width [22].

Figure 5.11: Threshold voltage at short-channel for width variation with (solid lines) and without (dotted lines) fringing capacitance at various V_{ds} and V_{bs} .

Figure 5.12: Device structure with two Gaussian profiles along the channel length and two Gaussian profiles along the channel width dimensions.

In our inverse NWE modeling, two Gaussian doping profiles are proposed along the width dimension with peak doping concentration, $N_{pile,w}$ pile-up centroid, l_w and characteristic length, $l_{\beta,w}$ at $z = 0$ and $z = W_{eff}$, respectively, as shown schematically in figure 5.12. The inverse NWE is modeled in a similar way as our reverse SCE V_t model. Effective channel doping transformation is performed along the width dimension and channel length/width-dependent model is formulated as

$$N_{eff} = \frac{\sqrt{\pi}\kappa N_{ch}}{L_{eff}/l_{\beta}} \left[\operatorname{erf}\left(\frac{L_{eff} - l_{\mu}}{l_{\beta}}\right) + \operatorname{erf}\left(\frac{l_{\mu}}{l_{\beta}}\right) \right] \quad (5.4.2)$$

$$+ \frac{\sqrt{\pi}\kappa_w \left(\frac{L_{min}}{L}\right)^{\omega_2} N_{ch}}{W/l_{\beta,w}} \left[\operatorname{erf}\left(\frac{W_{eff} - l_w}{l_{\beta,w}}\right) + \operatorname{erf}\left(\frac{l_w}{l_{\beta,w}}\right) \right] + N_{ch}$$

$$l_{\beta,w} = \beta_w (\phi_{s0} - V_{bs})^{0.25} \quad (5.4.2a)$$

with four error functions to characterize the pile-up profiles at the corners of the channel length and width dimensions. κ_w is defined as the ratio of the Gaussian's peak concentration along the width dimension to the average vertical doping concentration modeled by an empirical length-dependent function, $\kappa_w = N_{pile,w} / [(L_{min}/L)^{\omega_2} N_{ch}]$, β_w and ω_2 (with default 1) are fitting parameters. In all the above equations, effective channel width is W .

The threshold-voltage roll-up effect is shown by the dashed line in figure 5.11 for short-channel device at decreasing channel width. However, the model is not able to simulate the measurement data accurately due to V_t roll-off behavior. V_t roll-off for short-channel device at decreasing channel width has been attributed to the potential barrier lowering effect in small geometry devices [21]. This potential barrier lowering

effect for small geometry device is modeled in a similar way as for SCE with quasi 2-D formulation

$$\varepsilon_{si} X_{dep} \frac{dE_w(z)}{dz} + \varepsilon_{ox} \frac{V_{gs} - V_{FB} - \phi_s(z)}{t_{ox}} = q N_{ch} X_{dep}, \quad (5.4.4)$$

where $E_w(z)$ is the vertical field along the width dimension (z), $\phi_s(z)$ is the channel potential at the Si/SiO₂ interface. The solution to (5.4.4) under the boundary conditions $V(0) = V_{bi}$ and $V(W_{eff}) = V_{bi}$ is given by

$$\begin{aligned} \phi_s(W) &= \phi_{so} + (V_{bi} - \phi_{so}) \frac{\sinh[w/l_{\alpha,w}]}{\sinh[W_{eff}/l_{\alpha,w}]} + (V_{bi} - \phi_{so}) \frac{\sinh[(W_{eff} - w)/l_{\alpha,w}]}{\sinh[W_{eff}/l_{\alpha,w}]} \\ &= \phi_{so} + [(V_{bi} - \phi_{so})] \frac{1}{\cosh[W/2l_{\alpha,w}]} \end{aligned} \quad (5.4.5)$$

To account for the narrow width potential-lowering effect, surface potential is again modified to be

$$\phi_s = \phi_{s0} - \Delta\phi - \Delta\phi_{s,w} \quad (5.4.6)$$

where

$$\Delta\phi_{s,w} = \frac{V_{bi} - \phi_{s0}}{\cosh(W/2l_{\alpha,w})} \quad (5.4.6a)$$

and $l_{\alpha,w}$ is defined as

$$l_{\alpha,w} = \alpha_w \left(\frac{L_{min}}{L} \right)^{\omega_l} (\phi_{s0} - V_{bs})^{0.25} \quad (5.4.6b)$$

α_w and ω_l (default 1) are fitting parameters. V_t expression with $\Delta\phi_{s,w}$ alone (without inverse NWE) cannot simulate measured data accurately at decreasing channel width, as shown by the dashed-dot-dot line in figure 5.11. However, with the coupled effect from NWE and inverse NWE models, the complete geometry-dependent V_t model is expressed as

$$V_t = V_{FB} + \phi_s + \left[\gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \right] \left[1 - \frac{\lambda \zeta}{L_{eff}} \frac{\sqrt{\phi_s + V_{sb}} + \sqrt{\phi_s + V_{db}}}{2} \right] + \Delta V_{tL, DIBL} \quad (5.4.7)$$

in which the geometry-dependent surface potential (5.4.6) and effective channel doping concentration (5.4.2) are the key features in our V_t modeling approach.

5.5 Effect of body bias

Width of the depletion layer in the channel can be changed at different body biases. Our geometry-dependent V_t model discussed in the previous section may not have strong V_{bs} dependency in simulating the measurement data. The V_{bs} -dependent fitting parameters are suggested to have a better fit to the measurement data. A linear or parabolic relationships are proposed for the charge-sharing fitting parameter (λ), ratio of the peak Gaussian lateral doping profile (κ) to the average vertical doping, and enhancement factor ϕ for the potential barrier lowering effect. Fitting parameters with parabolic relationship are expressed as

$$\begin{aligned} \lambda &= \lambda_0 + b_{\lambda 1} V_{bs} + b_{\lambda 2} V_{bs}^2 \\ \kappa &= \kappa_0 + b_{\kappa 1} V_{bs} + b_{\kappa 2} V_{bs}^2, \\ \phi &= \phi_0 + b_{\phi 1} V_{bs} + b_{\phi 2} V_{bs}^2 \end{aligned} \quad (5.5.1)$$

which are found to have better prediction to the measurement data than the linear approximation, as shown in figure 5.13. However, in compact modeling, the selection of linear or parabolic relationship highly depends on the available measurement data.

Figure 5.13: Threshold voltage with parabolic and linear V_{bs} approximation for λ , κ and ϕ , and compared with the measurement data (symbols).

CHAPTER 6: DRAIN CURRENT MODELING

The source-referenced drain-current modeling has gone through many generations of development, starting from regional/piecewise (MOS1 and MOS2) and empirical (MOS3) models [76] in the late 60s and mid 80s. Now, the model has been the industry's *de facto* standard for decades (BSIM family [77]) and widely used in the chip design communities. Recently, the trend in drain-current model development is towards inversion charge Q_i -based models (EKV [78] and the recent SPP [59]) and surface-potential ϕ_s -based model, either explicit (SP [79]) or iterative (HiSIM [80]), however, industries are still using the threshold-based drain current model in their circuit simulators. Threshold voltage is still the key monitoring parameter in fabricating the working MOSFETs on the wafers.

It is well known that model parameters used in industry standard model (V_T -based model) increase tremendously [4] in order to fit all non-ideal measurement data. “Binning” approach is allowed in the model for technology characterization, in which geometry/bias dependence can be empirically modeled. With more data being fitted, the model description becomes less physical. On the other hand, keep adding nonphysical model parameters into the “core” model will make it more and more empirical and unmanageable. Together with the “intrinsic” problems in model symmetry, continuity and physical behaviors in higher-order derivatives of the drain current solutions make it less attractive. In this work, the “intrinsic” problems for the existing drain-current model are addressed to overcome the “showstopper” and make the model attractive for circuit designers. The model symmetry, continuity and narrow-width effects that have not been presented by [60] are investigated with the bulk-referenced drain current formulation.

The first-order derivative of the output-conductance has not been well modeled due to the effective-transverse field [11], [77], [63] used in the existing modeling approach, which is discussed in this chapter. A physical effective-transverse field modeling approach is proposed. The derivation of our model preserves the minimum measurement data requirement with a step-by-step model calibration to technology for our threshold-voltage in the drain current modeling methodology, which will be discussed in the next chapter.

6.1 Long/wide-channel drain current model

A common feature of drain current equation in the literature is the drift-diffusion formulation from the first moment of Boltzmann Transport Equation (BTE) [81]. In the strong inversion condition, the drift current

$$\begin{aligned} I_{ds} &= WvQ_i(y) \\ &= -\mu WQ_i(y) \frac{d\phi_s(y)}{dy} \end{aligned} \quad (6.1.1)$$

is the dominant component, where v is the drift velocity, μ is the carrier mobility and position-dependent inversion charge $Q_i(y)$ is defined by (4.1.1b) based on the depletion approximation

$$Q_i = -C_{ox} \left(V_{gb} - V_{FB} - \phi_s(y) - \gamma \sqrt{\phi_s(y)} \right). \quad (6.1.1a)$$

Under the strong inversion condition, the integrated drain current across the channel from 0 to L_{eff} can be expressed as

$$\int_0^{L_{eff}} I_{ds} = \mu W \int_{\phi_0}^{\phi_L} Q_i(y) d\phi_s(y). \quad (6.1.2)$$

The integrated drain current can be formulated by substituting Q_i from (6.1.1a) into (6.12), resulting in

$$I_{ds} = \frac{W}{L_{eff}} \mu C_{ox} \left[(V_{gb} - V_{FB})(\phi_L - \phi_0) - \frac{1}{2}(\phi_L^2 - \phi_0^2) - \frac{2}{3}(\phi_L^{3/2} - \phi_0^{3/2}) \right] \quad (6.1.3)$$

where the surface potentials at the drain and source ends, ϕ_L and ϕ_0 , respectively, are the intermediate variables linking output drain-current with the input terminal voltages. Output drain current can also be related to the terminal voltages with the boundary conditions such that surface potentials are “pinned” at the drain and source

$$\phi_L \approx \phi_{s0} + V_{db} \quad (6.1.3a)$$

and

$$\phi_0 \approx \phi_{s0} + V_{sb}, \quad (6.1.3b)$$

respectively. Drain current can also be expressed as

$$I_{ds} = \frac{W}{L_{eff}} \mu C_{ox} \left\{ (V_{gb} - V_{FB})(V_{ds}) - \frac{1}{2}[(\phi_{s0} + V_{ds} + V_{sb})^2 - (\phi_{s0} + V_{sb})^2] - \frac{2}{3}[(\phi_{s0} + V_{ds} + V_{sb})^{3/2} - (\phi_{s0} + V_{sb})^{3/2}] \right\}, \quad (6.1.4)$$

with drain-source voltage, V_{ds} or

$$I_{ds} = \frac{W}{L_{eff}} \mu C_{ox} \left\{ (V_{gb} - V_{FB})(V_{db} - V_{sb}) - \frac{1}{2}[(\phi_{s0} + V_{db})^2 - (\phi_{s0} + V_{sb})^2] - \frac{2}{3}[(\phi_{s0} + V_{db})^{3/2} - (\phi_{s0} + V_{sb})^{3/2}] \right\} \quad (6.1.5)$$

with drain-bulk voltage, V_{db} . In compact modeling, the above drain-current equation is compacted by Taylor's series expansion on the 3/2 power terms such that it can be expressed as

$$\begin{aligned}
& (\phi_{s0} + V_{ds} + V_{sb})^{3/2} - (\phi_{s0} + V_{sb})^{3/2} \\
&= (\phi_{s0} + V_{ds} + V_{sb})(\phi_{s0} + V_{ds} + V_{sb})^{1/2} - (\phi_{s0} + V_{sb})^{3/2} \\
&= (\phi_{s0} + V_{ds} + V_{sb}) \left(\sqrt{\phi_{s0} + V_{sb}} + \frac{V_{ds}}{2\sqrt{\phi_{s0} + V_{sb}}} - \frac{V_{ds}^2}{8(\phi_{s0} + V_{sb})^{3/2}} + \dots \right) - (\phi_{s0} + V_{sb})^{3/2} \\
&= \left[(\phi_{s0} + V_{sb})^{3/2} + \frac{3V_{ds}}{2} \sqrt{\phi_{s0} + V_{sb}} + \frac{3V_{ds}^2}{8\sqrt{\phi_{s0} + V_{sb}}} - \frac{V_{ds}^3}{8\sqrt{\phi_{s0} + V_{sb}}} + \dots \right] - (\phi_{s0} + V_{sb})^{3/2} \quad (6.1.6) \\
&= \frac{3V_{ds}}{2} \sqrt{\phi_{s0} + V_{sb}} + \frac{3V_{ds}^2}{8\sqrt{\phi_{s0} + V_{sb}}} \left(1 - \frac{V_{ds}}{3} + \dots \right) \\
&\approx \frac{3V_{ds}}{2} \sqrt{\phi_{s0} + V_{sb}} + \frac{3n_v V_{ds}^2}{8\sqrt{\phi_{s0} + V_{sb}}}
\end{aligned}$$

where the bulk-charge coefficient, n_v represents a constant for V_{ds} dependence.

Alternatively

$$\begin{aligned}
& \left[(\phi_{s0} + V_{ds})^{3/2} - (\phi_{s0} + V_{sb})^{3/2} \right] \\
&= \left[(\phi_{s0} + V_{ds} + V_{sb})^{3/2} - (\phi_{s0} + V_{sb})^{3/2} \right] \\
&\approx \phi_{s0}^{3/2} + \frac{3}{2} \phi_{s0}^{1/2} (V_{ds} + V_{sb}) + \frac{3}{2} \frac{1}{2} \frac{(V_{ds} + V_{sb})^2}{\sqrt{\phi_{s0}}} \quad (6.1.7) \\
&\quad - \left[\phi_{s0}^{3/2} + \frac{3}{2} \phi_{s0}^{1/2} (V_{sb}) + \frac{3}{2} \frac{1}{2} \frac{(V_{sb})^2}{\sqrt{\phi_{s0}}} \right] \\
&\approx \frac{3}{2} \phi_{s0}^{1/2} V_{ds} + \frac{3}{4} \frac{V_{ds}^2 + 2V_{ds}V_{sb}}{2\sqrt{\phi_{s0}}}
\end{aligned}$$

with the expansion up to third term without n_v . A compact drain-current model is formulated as

$$I_{ds} = \frac{W}{L_{eff}} \mu C_{ox} \left\{ (V_{gs} - V_{tL})(V_{ds}) - \frac{1}{2} A_b V_{ds}^2 \right\}. \quad (6.1.8)$$

Substituting (6.1.6) into (6.1.4) results in the well-known S-ref drain-current model with the threshold voltage,

$$V_{tL} = V_{FB} + \phi_{s0} + \gamma \sqrt{\phi_{s0} - V_{bs}} \quad (6.1.9a)$$

and the bulk-charge factor

$$A_b = 1 + \frac{n_v \gamma}{2 \sqrt{\phi_{s0} - V_{bs}}}, \quad (6.1.9b)$$

respectively. On the other hand, substituting Taylor's series from (6.1.7) into (6.1.5) gives the B-ref drain-current model with

$$V_{tL} = V_{FB} + \phi_{s0} + \gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \quad (6.1.10a)$$

and

$$A_b = 1 + \frac{\gamma}{2 \sqrt{\phi_{s0}}}. \quad (6.1.10b)$$

The above B-ref drain-current expression is derived based on ideal long-channel MOSFET's device physics. In real devices, non-ideal device behaviors in the output $I - V$ characteristics at decreasing channel length/width cannot be modeled with the above drain-current equation. This is because geometry effects, such as intrinsic mobility degradation effect due to transverse/lateral field, extrinsic series source/drain resistances, channel-length modulation with velocity-overshoot and extrinsic lay-out effects, have to be considered in the drain-current formulation. Geometry dependence for non-ideal device physics effects in drain-current formulation will be discussed in the following.

6.1.1 Transverse-field-dependent mobility degradation with bulk-referenced model

Main sources of model symmetry arise from the effective mobility expression. The expression includes transverse and lateral field dependencies, as well as the effective transverse-field. The transverse field-dependent mobility can be expressed based on the well-known Matthiessen's summation of three scattering mechanisms [82],

$$\begin{aligned}\mu_0 &= \left(\frac{1}{\mu_{co}} + \frac{1}{\mu_{ph}} + \frac{1}{\mu_{sr}} \right)^{-1} \\ &= \frac{\mu_1}{1 + \left(\frac{\mu_1}{\mu_2} \right) E_{eff}^{\frac{1}{3}} + \left(\frac{\mu_1}{\mu_3} \right) E_{eff}^{\nu}}\end{aligned}\quad (6.1.11)$$

where μ_{co} , μ_{ph} and μ_{sr} , respectively, are the semi-empirical model for Coulombic, 2-D phonon scattering and surface roughness scattering can be obtained from [83], [84]

$$\mu_{co} = \mu_1 \quad (6.1.11a)$$

$$\mu_{ph} = \frac{\mu_2}{E_{eff}^{1/3}} \quad (6.1.11b)$$

$$\mu_{sr} = \frac{\mu_3}{E_{eff}^{\nu}}. \quad (6.1.11c)$$

μ_1 is assumed a constant with respect to the effective transverse field. μ_2 and μ_3 are related to acceptor density and absolute temperature, and ν defaults to 2. μ_1 , μ_2 , μ_3 and ν are usually treated as fitting parameters to match the measurement data [82]. The effective transverse field is related to the inversion and bulk charges as [85]

$$E_{\perp} = (Q_b + \zeta_n Q_i) / \epsilon_{si}, \quad (6.1.12)$$

where ζ_n is 0.5 for <100> electrons. Normally in S-ref drain-current model, the inversion charge at low drain bias condition is assumed to be

$$Q_i = C_{ox} (V_{gs} - V_{tL}), \quad (6.1.13)$$

and the bulk charge is approximated to

$$Q_b \approx C_{ox} V_{tL}. \quad (6.1.14)$$

Substituting Q_b and Q_i into (6.1.12) gives the effective transverse field

$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[V_{gs} + \left(\frac{1}{\zeta_n} - 1 \right) V_{tL} \right] \approx \frac{V_{gs} + V_{tL}}{6t_{ox}}. \quad (6.1.15)$$

Another approach can be taken to formulate effective transverse field with our B-ref modeling methodology. Transverse field expression is formulated from the position-dependent inversion charge

$$Q_i(y) = C_{ox} \left[V_{gb} - V_{FB} - \phi_s(y) - \frac{Q_b(y)}{C_{ox}} \right], \quad (6.1.16a)$$

and bulk charge is defined by

$$Q_b(y) = C_{ox} \gamma \sqrt{\phi_s(y)}, \quad (6.1.16b)$$

respectively. Based on the depletion approximation from the source end along the channel

$$\begin{aligned} E_{eff} &= \frac{1}{\phi_L - \phi_0} \int_{\phi_0}^{\phi_L} E_{eff}(y) d\phi(y) \\ &= \frac{1}{V_{ds} \epsilon_{si}} \left[\int_{\phi_0}^{\phi_L} \zeta_n Q_i(y) d\psi_s + \int_{\phi_0}^{\phi_L} Q_b(y) d\phi \right], \quad (6.1.17) \\ &= \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\left(V_{gs} - V_{tL} - \frac{A_b V_{ds}}{2} \right) + \frac{\gamma \sqrt{\phi_{s0}}}{\zeta_n} \left(1 + \frac{V_{ds} - 2V_{bs}}{4\phi_{s0}} \right) \right] \end{aligned}$$

where V_{tL} and A_b are expressed in B-ref drain-current model given by (6.1.10). A process-dependent parameter, χ_b is proposed to indicate the amount of bulk-charge contribution to the integrated transverse field, and (6.1.17) is modified to

$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\left(V_{gs} - V_{tL} - \frac{A_b V_{ds}}{2} \right) + \frac{\chi_b \gamma \sqrt{\phi_{s0}}}{\zeta_n} \left(1 + \frac{V_{ds} - 2V_{bs}}{4\phi_{s0}} \right) \right]. \quad (6.1.18)$$

Figure 6.1: First-order derivative solutions of g_{ds} versus V_{ds} for (solid line) B-ref V_{tL} and E_{eff} , compared to (dotted line) B-ref V_t and S-ref E_{eff} , (dashed line) S-ref V_{tL} and S-ref E_{eff} , as well as (circle) numerical data.

Figure 6.1 shows the first-order derivative of output drain conductance, $g'_{ds} = dg_{ds}/dV_{ds}$ compared with the Medici data (symbols) and measurement data in the appendix, which is physically modeled in our new effective transverse field (6.1.18) formulation. Inaccurate g'_{ds} behavior is observed when effective field is formulated at low drain voltage condition (dotted and dashed lines [86]), which is well-adopted in the literature [11], [67], [77]. Our regional strong inversion drain-current expression is modified with mobility degradation (6.1.11) effect due to the transverse field (6.1.18) as

$$I_{ds} = \frac{W}{L_{eff}} \mu_0 C_{ox} \left\{ (V_{gs} - V_{th}) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\}. \quad (6.1.19)$$

6.1.2 Lateral-field-dependent mobility degradation modeling

When the device is operating in the linear region (low V_{ds} condition), mobility degradation effect due to lateral field is relatively small. However, when V_{ds} is gradually increased, carrier drift velocity in the channel will eventually saturate to a fixed value, v_{sat} [87]. The output drain-current electrical characteristic (6.1.1), which is a function of the drift velocity, will change accordingly. In the literature, a two-region piecewise velocity-field relation [88] is proposed and has been well-adopted in compact modeling,

$$v_0 = \frac{\mu_{eff0} E_{||}}{1 + \delta_0 E_{||} / E_{sat}} \quad \text{if } E_{||} < E_{sat} \quad (6.1.20a)$$

with a “ δ_0 ” factor suggested [53] in the linear region and

$$v_0 = v_{sat} \quad \text{if } E_{||} > E_{sat}, \quad (6.1.20b)$$

when the lateral field is greater than the saturation field

$$E_{sat} = (1 + \delta_0) \frac{v_{sat}}{\mu_{eff0}} \quad (6.1.20c)$$

with the saturation drift velocity, v_{sat} . $\delta_0 = 1$ by default and the lateral-field is defined as the potential gradient,

$$E_{||} = -\frac{d\psi_s}{dy}. \quad (6.1.20d)$$

Substituting the drift velocity in (6.1.20) into the drain-current equation (6.1.1), the following integrated drain current (taking the similar derivation step as in section 4.1) can be formulated,

$$I_{ds} = \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_{th}) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\} \quad (6.1.21)$$

in which the mobility degradation due to the transverse field (μ_o) is formulated in section 6.1.11, and in this section, the coupled lateral and transverse field effects is derived as

$$\mu_{eff0} = \frac{\mu_o}{1 + \frac{\delta_0 V_{ds}}{E_{sat} L_{eff}}} \quad (6.1.22)$$

Many empirical δ_0 models have been proposed and they are summarized in [53]. In our drain-current modeling approach, the following empirical length-dependent δ_0 is proposed as

$$\delta_0 = \left[\left(\frac{L_g - L_{min}}{L_{max} - L_{min}} \right) (\delta_L - 1) + 1 \right] \left(\frac{V_{ds}}{V_{dsat}} \right) \quad (6.1.23)$$

where δ_L is the fitting parameter to adjust the amount of mobility degradation due to the lateral-field effect and V_{dsat} is the saturation drain voltage. The first-order derivative on the mobility degradation effect due to the transverse field (μ_o , left axis) and coupled transverse and lateral field (μ_{eff0} , right axis) is shown in figure 6.2 [13] with B-ref (solid lines) and S-ref (dashed and dotted line) transverse field, and B-ref (solid and dotted lines) and S-ref (dashed) threshold voltage. The removal of singularity for the Gummel symmetric test circuit [27] at $V_x = 0$ for coupled lateral and transverse field is clearly shown in the inset

Figure 6.2: Derivatives of μ_{eff0} [right axis] and μ_0 [μ_{eff} with $\delta_0 = 0$, left axis] w.r.t. V_{ds} for three models in figure 6.1. The inset shows the Gummel symmetry plots for the B-ref V_{tL} and E_{eff} models when μ_{eff0} (right axis) and μ_0 (left axis) is used as the effective mobility in drain current, I_{ds} , respectively.

Drain-current symmetry behavior is investigated with Gummel symmetry test [27], as shown in figure 6.3 for the three models mentioned in the above section. It demonstrates that with the correct lateral and transverse field effects in the mobility model, and correct V_t modeling methodology in drain-current derivation, symmetric drain-current model can be formulated. This shows that asymmetry in drain-current model may not be inherited by nature to the V_t -based model.

Figure 6.3: Gummel symmetry test [27] (with $V_{gb0} = 1.2$ V and $V_{b0} = 0.2$ V) of the first (right axis) and second order (left axis) derivatives of I_d versus V_x for long/wide-channel device.

6.2 Short/wide-channel drain-current model

At decreasing channel length, geometry effects in threshold-voltage expression has been well-formulated and re-written here for convenience,

$$V_t = V_{FB} + \phi_s + \left[\gamma \sqrt{\phi_{s0}} - (A_b - 1) V_{bs} \right] \left[1 - \frac{\lambda \zeta}{L} \frac{\sqrt{\phi_s + V_{sb}} + \sqrt{\phi_s + V_{db}}}{2} \right] + \Delta V_{tL, DIBL}. \quad (6.2.1)$$

The expression contains DIBL, charge-sharing effect and long-channel DIBL effects in the second, third and fourth terms, respectively. The reverse SCE is accounted for by forming the effective doping concentration through the doping transformation from lateral and vertical doping profiles, which is used in the ϕ_s calculation. The geometry-dependent threshold voltage model (6.2.1) converges to long-channel threshold-voltage expression (6.1.10a) where geometry effects in (6.2.1) are negligibly small for long-channel device. However, replacing V_{tL} (6.1.10a) by V_t (6.2.1) in drain-current expression (6.1.20)

$$I_{ds} = \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_t) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\} \quad (6.2.2)$$

to model drain-current electrical characteristics for short/narrow devices is not accurate because not only the threshold voltage changes, significant voltage drop across series resistances cannot be ignored in the geometry-dependent drain-current modeling.

6.2.1 Source and drain series resistance effect in drain current model

Drain current in (6.2.2) has been formulated with mobility degradation due to lateral and transverse field, μ_{eff0} , with the replacement of long-channel V_{tL} (6.1.10a) by short-channel V_t (6.2.1). When a large current is flowing through the short-channel MOSFET, it causes a significant voltage drop across the series resistance in the drain and source regions. From Ohm's law, the drain current is the ratio of the terminal drain-source voltage over the total resistance of the channel, R_{ch} and the series resistances at the drain and source, R_d and R_s , respectively,

$$\begin{aligned} I_{ds} &= \frac{V_{ds}}{R_{ch} + R_{sd}} \\ &= \frac{V_{ds} / R_{ch}}{1 + R_{sd} / R_{ch}}, \\ &= \frac{I_{ds0}}{1 + R_{sd} / R_{ch}} \end{aligned} \quad (6.2.3)$$

where I_{ds0} is defined by (6.2.2) as the drain-current without R_{sd} effect. Without considering velocity-overshoot and channel-length modulation effects, (6.2.3) can be rewritten in a very compact form, similar to (6.2.2)

$$I_{ds} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} \left\{ (V_{gs} - V_t) V_{ds} - \frac{1}{2} A_b V_{ds}^2 \right\} \quad (6.2.4)$$

where effective mobility with R_{sd} effect is expressed as

$$\mu_{eff} = \frac{\mu_{eff0}}{1 + R_{sd} \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_t) - \frac{1}{2} A_b V_{ds} \right\}}. \quad (6.2.4a)$$

The source/drain series resistance is decomposed into two parts [89]. The first part is the intrinsic bias-dependent term, R_{int} to model the accumulation resistance, R_{ac} and spreading resistance, R_{sp} [90],

$$R_{int} = \frac{E\Delta y/2}{I_{ds}} = \frac{\nu}{C_{ox} W (V_{gs} - V_t)} \quad (6.2.5)$$

where E is the electric field, I_{ds} is the channel current and ν is a fitting parameter. The second part is the extrinsic bias-independent term, R_{ext} to model sheet resistance, R_{sh} and contact resistance, R_{co} . R_{ext} is assumed constant, mainly from sheet resistance,

$$R_{ext} = \frac{2\rho S}{x_j W}, \quad (6.2.6)$$

where ρ/x_j is the sheet resistance under the LDD spacer of thickness S . R_{int} in [89] was the previously derived intrinsic resistance, assuming no bulk-charge contribution in the drain current [i.e., ignoring the second term in (6.2.2)]. This assumption is commonly used in the existing model, which is not valid as the drain current consists of both the inversion and bulk-charge contributions. In this work, R_{int} in (6.2.5) is proposed as

$$R_{int} = 2 \frac{\nu}{C_{ox} W \left(V_{gs} - V_t - \frac{A_b}{2} V_{ds} \right)}. \quad (6.2.7)$$

The final source/drain resistance is formulated as

$$R_{sd} = R_s + R_d \quad (6.2.8)$$

$$R_s = R_d \quad (6.2.8a)$$

$$R_s = \frac{R_{ext} + R_{int}}{2} = r_1 + \frac{r_2}{C_{ox}W \left(V_{gs} - V_t - \frac{A_b}{2} V_{ds} \right)} \quad (6.2.8b)$$

where

$$r_1 = \frac{2\rho S}{x_j W}, \quad (6.2.8c)$$

$$r_2 = \nu. \quad (6.2.8d)$$

6.2.2 Saturation drain-current model

The derived drain-current model (6.2.4) describes length-dependent drain-current electrical characteristics that are only valid in the linear operating region. If the terminal voltage is biased such that the device is operated in the saturation region, the previously derived I_{ds} expression is not longer valid, as shown by the dashed line (linear current model, $I_{d,lin}$) in figure 6.4. However, from the measurement data, it is observed that the drain current is saturated when $V_{ds} > V_{dsat}$, where the saturation voltage V_{dsat} is located at the peak of the dashed line which can be determined by taking the derivative of drain current (6.1.8): $V_{dsat} = (V_{gs} - V_{tL})/A_b$ [64]. This is a crude approximation to determine V_{dsat} where the lateral-field effect is ignored. Another way to formulate V_{dsat} is by equating the saturation drain current, which can be determined from the gradual-channel approximation

$$I_{dsat} = W v_{sat} C_{ox} (V_{gs} - V_t - A_b V_{dsat}) \quad (6.2.9)$$

to the drain-current defined by (6.2.4)

$$I_{dsat} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} \left\{ (V_{gs} - V_t) V_{dsat} - \frac{1}{2} A_b V_{dsat}^2 \right\}. \quad (6.2.10)$$

Replacing V_{ds} with V_{dsat} , ignoring R_{sd} effect ($\mu_{eff} \approx \mu_{eff0}$) for long-channel, and setting $\delta_L = 1$, V_{dsat} can be derived as

$$V_{dsat} = \frac{E_{sat} L_{eff} (V_{gs} - V_t)}{V_{gs} - V_t + A_b E_{sat} L_{eff}} \quad (6.2.11)$$

for long-channel devices with lateral-field effect. However, for the short-channel devices, R_{sd} effect cannot be ignored and the drain-current in (6.2.9) is modified to

$$\begin{aligned} I_{dsat} &= W v_{sat} C_{ox} \left\{ \left[V_{gs} - I_{dsat} R_s \right] - V_t - A_b \left[V_{dsat} - I_{dsat} (R_s + R_d) \right] \right\} \\ &= \frac{W v_{sat} C_{ox} (V_{gs} - V_t - A_b V_{dsat})}{1 - W C_{ox} v_{sat} \left[(A_b - 1) R_s + A_b R_d \right]} \end{aligned} \quad (6.2.12)$$

Equating (6.2.12) with (6.2.10), length-dependent V_{dsat} can be formulated as [91]

$$V_{dsat} = \frac{-b - \sqrt{b^2 - 4ac}}{2a} \quad (6.2.13)$$

$$a = (A_b W C_{ox} v_{sat} R_s) \quad (6.2.13a)$$

$$b = -\left\{ V_{gs} - V_t + (V_{gs} - V_t) W C_{ox} v_{sat} \left[2R_s + A_b (R_s + R_d) \right] + A_b E_{sat} L_{eff} \right\} \quad (6.2.13b)$$

$$c = E_{sat} L_{eff} (V_{gs} - V_t) + 2 W C_{ox} v_{sat} (R_s + R_d) (V_{gs} - V_t)^2. \quad (6.2.13c)$$

Saturation drain-current model, I_{dsat} (6.2.10) with the derived V_{dsat} (6.2.13) can accurately predict the measured drain current, as shown in figure 6.4(a) (dotted line).

(a)

(b)

Figure 6.4: Drain current solutions based on strong inversion model with unified (V_{deff} , solid line), linear (V_{ds} , dashed line) and saturation (V_{dsat} , dotted) current equations for a) long-channel, b) short-channel.

The difference between linear and saturation drain current models in (6.2.4) and (6.2.10) is only the drain-source voltage used in the drain-current expression, where V_{ds} and V_{dsat} are both referred to the one used in drain current expression in linear and saturation region, respectively. An effective drain-source voltage is introduced,

$$V_{deff} = V_{dsat} - \frac{1}{2} \left[V_{dsat} - V_{ds} - \delta_s + \sqrt{(V_{dsat} - V_{ds} - \delta_s)^2 + 4\delta_s V_{dsat}} \right], \quad (6.2.14)$$

with a smoothing parameter, δ_s . A unified drain-current model that smoothly joins both the linear and saturation drain-current models can be expressed with V_{ds} in (6.2.4) replaced by V_{deff} , yielding

$$I_{ds, str} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} \left\{ (V_{gs} - V_t) V_{deff} - \frac{1}{2} A_b V_{deff}^2 \right\} \quad (6.2.15)$$

as shown by the solid line in figure 6.4. One can observe that the simulated short-channel drain current with piecewise linear and saturation drain current models can smoothly joint with unified drain-current model (6.2.15). However, in saturation region, the unified (solid line) and the regional (dashed line) drain-current models cannot simulate the measured drain current accurately. The following section gives detailed study on the electrical characteristics in this region. A theoretical model is suggested based on velocity-overshoot and lateral-field effects.

6.2.3 Channel-length modulation (CLM) effect for unified drain current model

The drain-current model in (6.2.15) has been derived based on two-region piecewise velocity-field relation model, as shown in figure 6.4. Velocity-field relationship assumes that when the lateral-field reaches at saturation condition, drift velocity will be saturated at v_{sat} . This is clearly shown in figure 6.4a for long-channel device but not the case for short-channel (figure 6.4b) as the measurement data shows the increase of the drain-current at increasing V_{ds} . Channel-length modulation (CLM) and

velocity-overshoot effects become important when short-channel devices are operated in this region.

High-field electron effective velocity, v_h is suggested to account for the velocity-overshoot effect. From Boltzmann transport equation (BTE), in the velocity saturation region (VSR)

$$v_h = v_0 \left(1 + \frac{k}{qE_{\parallel}} \frac{\partial T_e}{\partial y} \right), \quad (6.2.16)$$

which includes nonlocal electron temperature in the second term but excluding particle diffusion. In the above equation, k is the Boltzmann constant and T_e the electron temperature. Based on quasi-2D solution [25] of the electric field in VSR, compact high-field electron effective velocity can be formulated [92]

$$v_h = \frac{\mu_0 E_{\parallel}}{1 + \frac{\delta_0 E_{\parallel}}{E_{satn}}}, \quad (6.2.17)$$

in which the effective saturation field is given by

$$E_{satn} = E_{sat} \left[1 + h(V_{ds} - V_{deff}) \right], \quad (6.2.17a)$$

$$h = \frac{\frac{\xi c}{l^2} \left[1 + \sqrt{1 + \left(\frac{V_{ds} - V_{deff}}{l E_{sat}} \right)^2} \right]}{\sqrt{1 + \left(\frac{V_{ds} - V_{deff}}{l E_{sat}} \right)^2} - \frac{\xi c}{l^2} (V_{ds} - V_{deff})}. \quad (6.2.17b)$$

In (6.2.17b), c is related to the thermal voltage, electronic charge and mobility, and l is the characteristic length, and ξ is a fitting parameter.

Replacing v with v_h in position-dependent drain-current expression (6.1.1) and following the similar drain-current integration step as in section 4.1 (ignoring series source/drain resistance), a new drain-current with CLM effect can be formulated as

$$I_{deff} = \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_t) V_{deff} - \frac{1}{2} A_b V_{deff}^2 \right\}. \quad (6.2.18)$$

The previously formulated mobility expression μ_{eff0} (6.1.22) is modified with the proposed velocity-overshoot effect as

$$\begin{aligned} \mu_{eff0} &= \frac{\mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{satn}}} \\ &= \frac{\mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})]}} \\ &= \frac{\mu_0}{\frac{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})] + \delta_0 V_{deff}}{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})]}} \\ &= \frac{\mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{sat}}} \frac{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})] + \delta_0 V_{deff} + \delta_0 V_{deff} h(V_{ds} - V_{deff})}{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})] + \delta_0 V_{deff}} \\ &= \frac{\mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{sat}}} \left(1 + \frac{\delta_0 V_{deff} h(V_{ds} - V_{deff})}{L_{eff} E_{sat} [1 + h(V_{ds} - V_{deff})] + \delta_0 V_{deff}} \right) \\ &= \frac{\mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{sat}}} \left(1 + \frac{(V_{ds} - V_{deff})}{V_{A_{eff}}} \right) \\ &= \frac{g_{v0} \mu_0}{1 + \frac{\delta_0 V_{deff}}{L_{eff} E_{sat}}} \end{aligned} \quad (6.2.18a)$$

where

$$g_{vo} = 1 + \frac{V_{ds} - V_{dsat}}{V_{Aeff}} \quad (6.2.18b)$$

and V_{Aeff} is an effective early voltage defined as

$$V_{Aeff} = \frac{L_{eff} E_{sat} \left[1 + h(V_{ds} - V_{deff}) \right] + \delta_0 V_{deff}}{\delta_0 h V_{deff}}. \quad (6.2.18c)$$

Single-piece strong-inversion drain current model from linear to saturation with series resistance and CLM effects can be derived similarly as in section 4.2.1

$$I_{ds, str} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} \left\{ (V_{gs} - V_t) V_{deff} - \frac{1}{2} A_b V_{deff}^2 \right\}, \quad (6.2.19)$$

in which the previously derived effective mobility with series resistance effect (6.2.4a) is modified to account for the velocity-overshoot effect

$$\mu_{eff} = \frac{\mu_{eff0}}{1 + R_{sd} \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_t) - \frac{1}{2} A_b V_{deff} \right\}}, \quad (6.2.19a)$$

in which μ_{eff0} is defined by (6.2.18a).

Figure 6.5: Simulated drain-current (lines) based on velocity-overshoot (solid and dotted) and two-region piecewise velocity-field (dashed) models, compared to the measured drain current (symbol).

Figure 6.5 shows the short-channel saturation drain current with (dotted line) and without (dashed line) CLM effect. It is clearly shown that with the carrier velocity-overshoot effect and effective saturation field, our CLM effect in the saturation region is modeled accurately compared to the measurement data. Unified drain current is formulated by replacing V_{ds} with V_{deff} , which gives a smooth transition from linear to saturation (solid line) regions. Gummel symmetry test [27] is performed and the simulated results are shown in figure 6.6. It indicates that our unified drain-current model smoothly across $V_x = 0$ for long-channel (figure 6.3) as well as for short-channel (figure 6.6) devices.

Figure 6.6: Gummel symmetry test (with $V_{gb0} = 1.2$ V and $V_{b0} = 0$) of the first- (right axis) and second- (left axis) derivatives of I_d versus V_x for short-channel device.

6.3 Narrow-channel drain-current model

In the previous sections of this chapter, we have extensively discussed the length-dependent unified drain-current model in strong inversion region (linear to saturation regions). As channel width decreases, the previously derived model (dotted lines, figure 6.7) will not predict accurately the measurement data (symbols) with narrow-width effect

related to the stress in STI structure. The NEW degrades the mobility in the narrow-width device [93] and reduces turn-on current I_{on} , output drain conductance g_{dsat} and transconductance g_{msat} at decreasing channel width for the same channel length, as shown in figure 6.7 (dotted lines).

(a)

(b)

Figure 6.7: Long-channel turn on (a) current, (b) trans/output conductance at decreasing channel width with (solid lines) and without (dotted lines) width-dependent factor, μ_f in the coulombic mobility scattering model are compared to the measured data (symbols).

Based on experimental data and the observed increase in mobility degradation for long channel at decreasing channel width [19], we found that the drain current with width-dependent electrical characteristic is more correlated to the Coulombic scattering mobility. We propose the modification of

$$\mu_{co} = \mu_f \mu_1, \quad (6.3.1)$$

where μ_f is the width-dependent factor in Coloumbic mobility, and is empirically modeled as

$$\mu_f = \mu_{f1} + \frac{(\mu_{f2} - \mu_{f1})}{1 + \left(\frac{W}{\mu_{f3}} \right)^{\mu_{f4}}}. \quad (6.3.1a)$$

Four fitting parameters are used to determine the width-dependent Coloumbic factor μ_f , which are independent of bias variation and can be extracted from technology data ($I_{on} - W$), as shown in figure 6.8 with the solid line from the model. With this model (from fixed to width-dependent μ_f), our first-order unified width-dependent drain-current model is formulated and predicts accurately not only the functional but also the higher-order derivatives of terminal characteristics.

Figure 6.8: Width-dependent coulumbic scattering factor with extracted data (symbol) from technology data (I_{on}), and compared with the simulated data from the model.

Another geometry effect due to layout-dependent behavior at decreasing channel width arises from the “dog-bone” layout of the S/D resistance in narrow-width devices [19], [94], which can be explained schematically as shown in figure 6.9 [20]. The channel width, W is usually smaller than the contacted active area, W_{dog} , which is layout dependent. Besides the channel and contact width, the distance between the contact holes to the poly gate, S , S_1 and S_2 (each may have its own resistivity, ρ and junction depth, x_j) is another factor that affects source/drain series resistance. Although transistor layout is outside the intrinsic MOSFET model, the effect on measured drain current is extremely significant due to the voltage across the external resistance, which affects the model parameter extraction if it is not included in the intrinsic drain-current model. Our formulated series resistance in the previous section, R_{sd} model, includes a V_{gs} -dependent intrinsic part (R_{int}) and a constant extrinsic part (R_{ext}), and is repeated here for convenience,

$$R_{sd} = R_s + R_d = R_{int} + R_{ext}, \quad (R_s = R_d) \quad (6.3.2)$$

$$R_{\text{int}} = 2 \frac{\nu}{C_{\text{ox}} W (V_{\text{gs}} - V_t)}, R_{\text{ext}} = \frac{2\rho S}{x_j W} \quad (6.3.2a)$$

Although the “dog-bone” layout effect is considered outside the core intrinsic drain current model, in R_{sd} model, the total external resistance is modified to

$$\begin{aligned} R_{\text{ext}} &= \frac{2\rho S}{x_j W} + \frac{2\rho_1 S_1}{x_{j1} W} + \frac{2\rho_2 S_2}{x_{j2} W} \\ &= \frac{2\rho}{x_j W} \left[(S + \theta_r S_1) + \frac{(\theta_r S_2) W}{W_{\text{dog}}} \right] \\ &= \frac{2\rho S'}{x_j W} \left[1 - \theta_s \left(1 - \frac{W}{\theta_d W_{\text{dog}}} \right) \right] \end{aligned} \quad (6.3.3)$$

where $S' = S + \theta_r(S_1 + S_2)$, and θ_r is defined as

$$\theta_r = \frac{\rho_1 / x_{j1}}{\rho / x_j} \quad (6.3.3a)$$

which is taken as a fitting parameter if the layout parameters (S_1 and S_2) are given; or alternatively,

$$\theta_s = \frac{\theta_r S_2}{S + \theta_r S_1 + \theta_r S_2} \quad (6.3.3b)$$

is taken as a fitting parameter (hence, θ_r can be calculated given S_1 , S_2 , and S), together with the added fitting parameter θ_d , as well as the original parameter ρ .

For short-/narrow-channel devices, conventional scaling rule $R_{sd} \propto W^{-1}$ by (6.3.2) underestimates the measured drain current for short/narrow channel devices (dotted line), as shown in figure 6.10. With the dog-bone layout-dependent model, it allows external series resistance, R_{ext} to be interpreted physically at decreasing channel width (solid line). However, the dog-bone layout model does not have any effect on long-channel devices as

the voltage drop across the series resistance is negligibly small due to small amount of drain current flowing through the device.

Figure 6.9: Schematic layout for dog-bone transistor (Co: contact, Poly: Polysilicon, LDD: lightly-doped drain).

Figure 6.10: Drain current for short- and long-channel devices showing the effect of dog-bone geometry. Solid lines indicate width-dependent external source/drain resistance model and dotted line shows no dog-bone model in R_{ext} for narrow-width devices.

Based on experimental data, even with R_{sd} modification for dog-bone layout effect, the drain current model underestimates the measurement data for very small

geometry devices, as shown in figure 6.11. We found out from measurement data that this is more related to coulombic-scattering term in the mobility expression for small geometry devices. The width-dependent μ_f factor is proposed to be modified with simple linear width and length dependence as

$$\mu_f = \left[1 + \mu_{f5} \frac{L_{\min}}{L_g} \frac{W_{\min}}{W} \right] \left[\mu_{f1} + (\mu_{f2} - \mu_{f1}) / \left(1 + \left(\frac{W}{\mu_{f3}} \right)^{\mu_{f4}} \right) \right], \quad (6.3.4)$$

where μ_{f5} is a fitting parameter. Figure 6.11 shows a comparison between the model with and without dog-bone effect and linear width-/length-dependent μ_f for coulombic-scattering term in mobility expression at decreasing channel length/width, with the unified drain current (6.2.19).

Figure 6.11: Turn on current for narrow-width at decreasing channel length (left/bottom axis) and short-channel at decreasing channel width (right/top axis). Unified drain current model is simulated and compared with the measurement data (symbols) for dog-bone effect, DB with and without μ_{f5} (solid and dotted lines), and both without DB and μ_{f5} (dashed lines).

6.4 Subthreshold drain-current modeling

The total intrinsic drain current flowing through the channel consists of two components, which have been physically derived through BTE and continuity equation, the drift, I_{drift} and diffusion current, I_{diff} . The previously derived unified geometry/bias-dependent drain current (6.2.19) is only valid in strong inversion region. Beyond its validity region, the model is inaccurate. Drain current in the subthreshold region, where the channel is weakly inverted ($V_{gs} < V_t$), is equally important in circuit simulation. In this section, we will formulate a unified drain-current model valid in all operating regions for all geometry/bias conditions.

From BTE [81], position-dependent diffusion current can be expressed as

$$I_{ds} = -\mu W v_{th} \frac{dQ_i(y)}{dy} \quad (6.4.1)$$

where v_{th} is the thermal voltage. Integrating the above equation from $y = 0$ to $y = L_{eff}$ gives the diffusion current

$$\begin{aligned} I_{ds} &= -\mu v_{th} \frac{W}{L_{eff}} \int_{Q_{i,s}}^{Q_{i,d}} dQ_i \\ &= -\mu v_{th} \frac{W}{L_{eff}} (Q_{i,d} - Q_{i,s}) \end{aligned} \quad (6.4.2)$$

where

$$Q_{i,d} = -\frac{\gamma C_{ox}}{2\sqrt{\phi_s}} v_{th} e^{(\phi_s - 2\phi_f - V_{db})/v_{th}} \quad (6.4.2a)$$

and

$$Q_{i,s} = -\frac{\gamma C_{ox}}{2\sqrt{\phi_s}} v_{th} e^{(\phi_s - 2\phi_f - V_{sb})/v_{th}} \quad (6.4.2b)$$

are the inversion charge densities at the drain and source ends, respectively. Hence, the diffusion current becomes

$$I_{ds} = \mu \frac{W}{L_{eff}} C_d v_{th}^2 e^{(\phi_s - 2\phi_f - V_{sb})/v_{th}} (1 - e^{-V_{ds}/v_{th}}) \quad (6.4.3)$$

and $C_d = \gamma C_{ox} / 2\sqrt{\phi_s}$ is the depletion capacitance. Bulk-charge linearization with depletion approximation (p. 141 in [64]) gives

$$Q_b(\phi_s) \cong Q_b(\phi_{s0}) + (\phi_s - \phi_{s0}) \left. \frac{\partial Q_b}{\partial \phi_s} \right|_{\phi_s = \phi_{s0}}, \quad (6.4.4)$$

$$\left. \frac{\partial Q_b}{\partial \phi_s} \right|_{\phi_s = \phi_{s0}} = C_{d0} = C_d \big|_{\phi_s = \phi_{s0}} = \frac{\gamma C_{ox}}{2\sqrt{\phi_{s0}}}, \quad (6.4.4a)$$

and from charge neutrality linking to the potential-balance equation in the subthreshold region,

$$V_{gb} - V_{FB} - \phi_s \approx Q_b(\phi_s), \quad (6.4.5)$$

Substituting (6.4.4) into (6.4.5) yields

$$\phi_s - \phi_{s0} = \frac{V_{gb} - V_t}{n_0}, \quad (6.4.6)$$

where $n_o = 1 + C_{d0}/C_{ox}$ and diffusion current from (6.4.3) can also be expressed as

$$I_{ds} = \mu \frac{W}{L_{eff}} C_d v_{th}^2 e^{(V_{gs} - V_t)/(n_0 v_{th})} (1 - e^{-V_{ds}/v_{th}}). \quad (6.4.7)$$

The above diffusion current is modified by adding an offset voltage, V_{off} to characterize the leakage current at zero gate bias, which determines the turn off current from real measurement data,

$$I_{ds} = \mu \frac{W}{L_{eff}} V_{diff} \quad (6.4.8)$$

$$V_{diff} = C_d v_{th}^2 e^{(V_{gs} - V_t - V_{off}) / (n_0 v_{th})} (1 - e^{-V_{ds} / v_{th}}). \quad (6.4.8a)$$

Regional drift and diffusion currents in the strong inversion and subthreshold regions are shown in figure 6.12. Direct addition of the two currents not only gives wrong result compared with measured long/wide channel device, but also introduces glitches at the transition. In order to obtain the asymptotic drain current behavior from subthreshold to strong inversion regions with continuous transition across regions, in the following section, we will formulate a unified continuous drain current model, which will take into account of the influence from physical SCEs, RSCE and NWEs in the subthreshold region.

Figure 6.12: Piecewise drain-current model based on drift (dotted line) and diffusion (solid line) current equations.

6.4.1 Continuous unified drain-current model

In this section, a single-piece continuous unified drain-current model across all operating regions is formulated. The regional drift current in strong inversion region (6.2.19) is re-written as

$$I_{ds, str} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} V_{drift}, \quad (6.4.9)$$

in which the regional gate/drain voltage

$$V_{drift} = (V_{gs} - V_t) V_{deff} - \frac{A_b}{2} V_{deff}^2 \quad (6.4.9a)$$

is replaced with the unified gate/drain voltage [95] as

$$V_{gg} = \frac{2n_0 v_{th} \ln \left[1 + \exp \left(\frac{V_{gs} - V_t}{2n_0 v_{th}} \right) \right] V_{de}}{2n_0 V_{de} C_{ox} \exp \left(-\frac{V_{gs} - V_t - 2V_{off}}{2n_0 v_{th}} \right) + 1 + \frac{v_{th} C_d \left[1 - \exp \left(-\frac{V_{ds}}{v_{th}} \right) \right]}{1}} \quad \begin{cases} V_{gs} \geq V_t : & V_{gg} = V_{drift} \\ V_{gs} < V_t : & V_{gg} = V_{diff} / W_{ge} \end{cases} \quad (6.4.10)$$

where V_{diff} is the diffusion gate/drain overdrive,

$$V_{de} = V_{deff} \left(1 - \frac{A_b}{2} \frac{V_{deff}}{V_{geff}} \right), \quad (6.4.10a)$$

$$W_{ge} = \frac{A_b}{n_0} \frac{1 - \exp \left(-\frac{V_{ds}}{v_{th}} \right)}{\exp \left(\frac{V_{gs} - V_t}{2n_0 v_{th}} \right)} \quad (6.4.10b)$$

and the effective gate overdrive expression is given [96] by

$$V_{geff} = \frac{2n_0 v_{tm} \ln \left[1 + \exp \left(\frac{V_{gs} - V_t}{2n_0 v_{tm}} \right) \right]}{2n_0 C_{ox} \exp \left(-\frac{V_{gs} - V_t - 2V_{off}}{2n_0 v_{tm}} \right) + \frac{C_d \left[1 - \exp \left(-\frac{V_{ds}}{v_{tm}} \right) \right]}{1}}. \quad (6.4.11)$$

The regional effective mobility is expressed as

$$\mu_{eff} = \frac{\mu_{eff0}}{1 + R_{sd} \frac{W}{L_{eff}} \mu_{eff0} C_{ox} \left\{ (V_{gs} - V_t) - \frac{1}{2} A_b V_{deff} \right\}}, \quad (6.4.12)$$

in which the contribution from the regional source/drain series resistance

$$R_{sd} = R_{ext} + R_{int} \quad (6.4.13)$$

is formulated as the sum of the extrinsic part

$$R_{ext} = \frac{2\rho S'}{x_j W} \left[1 - \theta_s \left(1 - \frac{W}{\theta_d W_{dog}} \right) \right] \quad (6.4.13a)$$

and the intrinsic part, given by

$$R_{int} = \frac{r_2}{C_{ox} W \left(V_{gs} - V_t - \frac{A_b}{2} V_{ds} \right)}. \quad (6.4.13b)$$

The denominator of the intrinsic series resistance can be replaced with

$$V_{ageff} = \frac{2n_0 v_{th} \ln \left[1 + \exp \left(\frac{V_{gs} - V_t}{2n_0 v_{th}} \right) \right] V_{ade}}{2n_0 V_{ade} C_{ox} \exp \left(-\frac{V_{gs} - V_t - 2V_{off}}{2n_0 v_{th}} \right) + \frac{v_{th} C_d \left[1 - \exp \left(-\frac{V_{ds}}{v_{th}} \right) \right]}{1}} \quad (6.4.14)$$

and

$$V_{ade} = \left(1 - \frac{A_b}{2} \frac{V_{deff}}{V_{geff}} \right), \quad (6.4.14a)$$

leading to a unified R_{int} formulation as

$$R_{int} = \frac{r_2}{C_{ox} W V_{ageff}}. \quad (6.4.15)$$

This gives asymptotic regional R_{int} (6.4.13b) solutions in the strong inversion region.

In the subthreshold region, the intrinsic series resistance is derived in a similar way as for strong inversion, but diffusion current (6.4.3) is considered as described in section 4.2.1, which gives,

$$R_{int} = \frac{r_2}{C_{ox} W v_{th} A_b \left[1 - \exp\left(-\frac{V_{ds}}{v_{th}}\right) \right]}. \quad (6.4.15)$$

A unified intrinsic series resistance valid from subthreshold to strong inversion regions is proposed as

$$R_{int} = \frac{r_2}{C_{ox} W \left\{ V_{ageff} + v_{th} A_b \left[1 - \exp\left(-\frac{V_{ds}}{v_{th}}\right) \right] \right\}} \quad (6.4.16)$$

which gives both asymptotic solutions in the strong inversion (6.4.13b) and subthreshold (6.4.15) regions. In order to represent mobility expression valid from subthreshold to strong inversion regions, gate overdrive expression in the denominator of mobility expression (6.4.12) is replaced with V_{ageff} , so that

$$\mu_{eff} = \frac{\mu_{eff0}}{1 + R_{sd} \frac{W}{L_{eff}} \mu_{eff0} C_{ox} V_{ageff}}. \quad (6.4.17)$$

Mobility degradation due to transverse and lateral fields and velocity-overshoot (g_{vo}) effects is represented by μ_{eff0} (6.2.18). Mobility (6.1.11) degradation (μ_o) due to transverse field is highly dependent on the regional effective transverse field (6.1.17) in the strong inversion region, re-written for convenience

$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\left(V_{gs} - V_t - \frac{A_b V_{ds}}{2} \right) + \frac{\chi_b V_{tsat}}{\zeta_n} \right], \quad (6.4.18)$$

$$V_{tsat} = \frac{\gamma \sqrt{\phi_s}}{\zeta_n} \left(1 + \frac{V_{deff} - 2V_{bs}}{4\phi_s} \right), \quad (6.4.18a)$$

with the first and second terms in the bracket being the contribution from inversion and bulk charge, respectively. In the subthreshold region, effective transverse field is assumed to be dependent on the bulk-charge term

$$\begin{aligned} E_{eff} &= \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\frac{\chi_b \gamma \sqrt{\phi_{sub}}}{\zeta_n} \right], \\ &= \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\frac{\chi_b V_{tsub}}{\zeta_n} \right], \end{aligned} \quad (6.4.19)$$

$$V_{tsub} = -\frac{\gamma^2}{2} + \gamma \sqrt{\frac{\gamma^2}{4} + V_{gs} - V_{bs} - V_{FB}}. \quad (6.4.19a)$$

A unified transverse field from subthreshold to strong inversion region is proposed as

$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\left(V_{geff} - \frac{A_b V_{deff}}{2} \right) + \frac{\chi_b V_{teff}}{\zeta_n} \right], \quad (6.4.20)$$

$$V_{teff} = \mathcal{G}_{eff} \{ V_{tsub}, V_{tsat}; \delta_b \} \quad (6.4.20a)$$

that smoothly joins V_{tsub} in the subthreshold to V_{tsat} in the strong inversion regions with the interpolating function,

$$\mathcal{G}_{eff} \{ x, x_{sat}; \delta \} \equiv x_{sat} - 0.5 \left[x_{sat} - x - \delta + \sqrt{(x_{sat} - x - \delta)^2 + 4\delta x_{sat}} \right]. \quad (6.4.21)$$

A unified effective mobility (6.4.17) expression, which gives asymptotic effective mobility (6.4.12) in subthreshold and strong inversion regions, can be formulated from the effective transverse field (6.4.20) and series resistance (6.4.16), with the replacement of denominator $(V_{gs} - V_t - 0.5 A_b V_{deff})$ in (6.4.12) and (6.4.13b) by V_{ageff} . A unified drift current in the strong inversion region is formulated as

$$I_{drift} = \frac{W}{L_{eff}} C_{ox} \mu_{eff} V_{gg} . \quad (6.4.22)$$

The unified diffusion current in the subthreshold region is formulated as

$$I_{diff} = \frac{W}{L_{eff}} \mu_{eff} C_{ox} V_{gd} \quad (6.4.23)$$

with the newly proposed effective gate/drain voltage in subthreshold region:

$$V_{gd} = V_{gg} W_{ge} \quad \begin{cases} V_{gs} \geq V_t : & V_{gd} = V_{drift} W_{ge} \\ V_{gs} < V_t : & V_{gd} = V_{diff} \end{cases} , \quad (6.4.24)$$

in which $C_{ox} V_{gd}$ approaches V_{diff} in subthreshold but decreases exponentially in strong inversion region due to the exponential function in W_{ge} . The total unified drain current from subthreshold to strong inversion regions is formulated as the summation of the unified drift and diffusion currents:

$$I_{ds} = I_{drift} + I_{diff} . \quad (6.4.25)$$

CHAPTER 7: MODEL PARAMETER EXTRACTION

A methodology is presented to extract device physical parameters, such as channel doping concentration (N_{ch}), poly-gate doping (N_{gate}) and QME (κ^{qm}), as well as smoothing parameters in our unified charge (AC), drain-current and threshold voltage (DC) models. Our unified regional charge model has less parameter dependency, as the physical parameter is not correlated with the smoothing parameter. Our unified drain current model is formulated from long/wide-channel, with SCE, reverse SCE, NWE and inverse NWE being modularly added to short/narrow-channel devices. This chapter illustrates the calibration sequence and data requirement for our unified charge and drain-current modeling.

7.1 Unified AC (charge) model parameter extraction

Implicit channel surface-potential from Pao-Sah model is repeated here for convenience,

$$V_{gb} - V_{fb} - \phi_s - \phi_p = \gamma \operatorname{sgn}[\phi_s] \sqrt{v_{th} \left(\exp\left[\frac{-\phi_s}{v_{th}}\right] - 1 \right) + \phi_s \left(1 - \exp\left[\frac{-2\phi_f}{v_{th}}\right] \right) + v_{th} \exp\left[-\frac{V_{cb} + 2\phi_f}{v_{th}}\right] \left(\exp\left[\frac{\phi_s}{v_{th}}\right] - 1 \right)} \quad (7.1.1)$$

It is nontrivial to extract physical parameters from the above implicit equation because the channel/gate surface-potentials are not measurable. Physical parameters, such as N_{gate} , N_{ch} and t_{ox} are closely correlated in V_{fb} , ϕ_f and C_{ox} expressions, which can have multiple solutions from the above implicit equation.

In our model parameter extraction methodology, the unified charge model requires *one* C_{gg} versus V_{gs} measurement data (measured at $V_{ds} = V_{bs} = 0$) with no further

calibration needed for bias dependence, including source/drain charge partitioning and bulk-charge models for non-zero V_{ds} dependence. The model requires 10 steps to extract its 12 fitting parameters: 3 physical parameters (N_{ch} , N_{gate} , κ^{qm}), 4 smoothing parameters for channel surface potential ϕ_s (σ_f , σ_a , σ_{eff} and δ_ϕ), 2 smoothing parameters for poly surface potential ϕ_s (ρ_f and ρ_r), and 2 model (σ^{qm} , ε^{qm}) and 1 smoothing (δ^{qm}) parameters for quantum-mechanical effect in the surface potential equation.

7.1.1 Physical parameter extraction

Gate capacitance can be expressed from the first-order derivative of gate charge from the left-hand or right-hand side of the above implicit equation as a function of ϕ_s and ϕ_p . If the parameter extraction procedure requires surface potentials to be calculated implicitly from (7.1.1), this makes all physical parameters correlated. In order to illustrate the correct physical content in our unified regional charge-based model formulation, C_{gg} data are generated numerically with known device parameters ($t_{ox} = 2\text{nm}$, $N_{ch} = 7.5 \times 10^{17} \text{ cm}^{-3}$) at four N_{gate} values (10^{22} , 5×10^{20} , 10^{20} and $5 \times 10^{19} \text{ cm}^{-3}$) with and without QME ($\kappa^{qm} = 0, 1$). Gate charge can be formulated regionally (ignoring ϕ_p at high poly doping) from the left-hand side of the implicit equation in the accumulation region

$$Q_g = C_{ox} (V_{gb} - \phi_{MS} - \phi_{CC}) \quad (7.1.2)$$

with the surface potential

$$\phi_{CC} = -2v_{th} \left[-\frac{a\mathcal{L}\{W\} + B}{a} \right] = V_{gb} - V_{fb} + 2v_{th}\mathcal{L}\{W\}. \quad (7.1.2a)$$

The regional gate capacitance is given by

$$C_{gg} = -\frac{dQ_g}{dV_{gb}} = C_{ox} \frac{\mathcal{L}\{W\}}{1 + \mathcal{L}\{W\}} \quad (7.1.3)$$

which is a strong function of oxide thickness, as can be observed from the simulated data shown in figure 7.1. The effect from channel doping in this region is not significant as can be observed in figure 7.1b.

(a)

(b)

Figure 7.1: Regional gate capacitance for (a) oxide thickness and (b) channel doping variations in the strong accumulation region, compared with Medici data. The corresponding gate charges are shown in the insets.

Similarly, in the depletion region, the gate charge can be expressed by the right-hand side of (7.1.1)

$$\begin{aligned} Q_g &= \gamma C_{ox} \sqrt{\phi_{dd}} \\ &= \sqrt{2q\epsilon_{si} N_{ch} \phi_{dd}} \end{aligned} \quad (7.1.4)$$

with the explicit regional surface potential in the depletion region

$$\phi_{dd} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gb} - V_{fb}} \right]^2. \quad (7.1.4a)$$

The regional gate capacitance in the depletion region can be expressed by taking the first order derivative of gate charge in the depletion region with respect to V_{gb}

$$\begin{aligned} C_{gg} &= \frac{dQ_g}{dV_{gb}} \\ &= C_{ox} \frac{\gamma}{2\sqrt{\frac{\gamma^2}{4} + V_{gb} - V_{fb}}} \\ &= \sqrt{\frac{2q\epsilon_{si} N_{ch}}{\gamma^2 + 4(V_{gb} - V_{fb})}} \\ &\approx \sqrt{\frac{q\epsilon_{si} N_{ch}}{2(V_{gb} - V_{fb})}} \quad \gamma^2 = 2 \frac{q\epsilon_{si} N_{ch} t_{ox}^2}{\epsilon_{ox}^2} \ll 4(V_{gb} - V_{fb}) \end{aligned} \quad (7.1.5)$$

The above analytical depletion gate capacitance expression shows that the physical parameters are uncorrelated because t_{ox} is not present in the above expression, while it is a strong function of the average channel doping concentration, N_{ch} as shown in figure 7.2 for various N_{ch} .

Figure 7.2: Regional gate capacitance for channel doping variation in the depletion region compared with Medici data.

In the strong inversion region, the regional gate charge can be expressed with the left-hand side of the Pao-Sah model,

$$Q_g = C_{ox} (V_{gb} - \phi_{MS} - \phi_{SS}) \quad (7.1.6)$$

with regional surface potential in the strong inversion region,

$$\phi_{SS} = \phi_{s0} + \Delta_{SS} + V_{cb}. \quad (7.1.6a)$$

Strong inversion gate capacitance can be formulated similarly from the first order derivative of gate charge

$$C_{gg} = \frac{dQ_g}{dV_{gb}} = C_{ox} \left(1 - \frac{d\phi_{SS}}{dV_{gb}} \right) \quad (7.1.7)$$

$$\begin{aligned} \frac{d\phi_{SS}}{dV_{gb}} &= \frac{d\Delta_{SS}}{dV_{gb}} \\ &= \frac{2v_{th} [V_{gb} - \phi_{MS} - (\phi_{s0} + V_{cb} + 4v_{th})]}{[V_{gb} - \phi_{MS} - (\phi_{s0} + V_{cb} + 4v_{th})]^2 - (\phi_{s0} + V_{cb} - v_{th})^2 \gamma^2}, \\ &\approx \frac{2v_{th}}{[V_{gb} - \phi_{MS} - (\phi_{s0} + V_{cb} + 4v_{th})]} \end{aligned} \quad (7.1.7a)$$

with

$$\phi_{SS}^* \approx \phi_{s0} + V_{cb} + 4v_{th} \quad (7.1.7b)$$

and

$$(\phi_{s0} + V_{cb} - v_{th})\gamma^2 \ll [V_{gb} - \phi_{MS} - (\phi_{s0} + V_{cb} + 4v_{th})]. \quad (7.1.7c)$$

The analytical expression shows that the gate capacitance is a strong function of oxide thickness and channel doping concentration in strong inversion region; as shown in figure 7.3. From figure 7.3b, one can observe that the gate capacitance has a very weak channel-doping dependence. This can be explained physically from the analytical expression (7.1.7a) in the strong inversion region, which shows the effect of channel doping variation is negligibly small in the denominator of (7.1.7a) that only has minor influence on ϕ_{s0} .

(a)

(b)

Figure 7.3: Regional gate capacitance for (a) oxide thickness and (b) channel doping variations in the strong inversion region, compared to the Medici data.

The above regional capacitance formulations assume a negligible voltage drop across the induced layer at the gate interface, which is true for highly doped poly-gate MOSFET. However, with low N_{gate} , the above regional strong-inversion gate charge model has to be modified to include ϕ_p

$$Q_g = C_{ox} (V_{gb} - \phi_{MS} - \phi_{ss} - \phi_{p,dd}) \quad (7.1.8)$$

with the regional (depletion) model ϕ_p

$$\phi_{p,dd} = \left(-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gb} - \phi_{fb} - \phi_s} \right)^2 \quad (7.1.8a)$$

at the gate interface due to poly-depletion effect (PDE). The corresponding regional gate capacitance with PDE can be analytically derived in a similar way

$$\begin{aligned} C_{gg} &= \frac{dQ_g}{dV_{gb}} \\ &= C_{ox} \left(1 - \frac{d\phi_{ss}}{dV_{gb}} - \frac{d\phi_{p,dd}}{dV_{gb}} \right) \end{aligned} \quad (7.1.9)$$

The extra PDE term results in the decrement in gate capacitance, which is given by

$$\frac{d\phi_{p,dd}}{dV_{gb}} = 1 + \frac{\gamma_p \left(-1 + \frac{d\phi_{ss}}{dV_{gb}} \right)}{\sqrt{\gamma_p^2 + 4(V_{gb} - V_{fb} - \phi_{ss})}} - \frac{d\phi_{ss}}{dV_{gb}}. \quad (7.1.9a)$$

Regional gate capacitance in the strong inversion region can be re-written as

$$C_{gg} = C_{ox} C_p \left(1 - \frac{d\phi_{ss}}{dV_{gb}} \right) \quad (7.1.10)$$

where the poly degradation factor

$$C_p = \frac{\gamma_p}{\sqrt{\gamma_p^2 + 4(V_{gb} - V_{fb} - \phi_s)}} \quad (7.1.10a)$$

is strongly N_{gate} dependent. The predicted result is shown in figure 7.4 from high to low N_{gate} concentrations.

Figure 7.4: Gate capacitance with N_{gate} variation in strong inversion region.

Another important phenomenon in MOSFET AC modeling is the quantum-mechanical effect. In this work, QME has been accounted for by the modification of the (gate and the channel) surface potentials with the band-gap widening, ΔE_g^{qm} from the van

Dort's model (3.4.2). ΔE_g^{qm} model has significant impact on the accumulation and strong inversion region charge models as it physically modifies the conventional intrinsic carrier concentration, making the space-charge density in the accumulation (holes) and strong inversion (electrons) regions strongly correlated. With poly gate terminal doped with n^+ doping, the poly-doping and QM effects are coupled. Then unified regional gate charge in strong inversion region, with coupled QM and poly-doping effects is expressed as

$$Q_g = C_{ox} (V_{gb} - \phi_{MS} - \phi_{str}^{qm} - \phi_{p,sub}^{qm}), \quad (7.1.11)$$

in which the unified regional channel surface potential (ϕ_{str}^{qm}) and the gate surface potential ($\phi_{p,sub}^{qm}$) are both coupled with QM and poly-doping effects. The regional gate capacitance is expressed as

$$\begin{aligned} C_{gg} &= \frac{dQ_g}{dV_{gb}} \\ &= C_{ox} \left(1 - \frac{d\phi_{ss}^{qm}}{dV_{gb}} - \frac{d\phi_{p,dd}^{qm}}{dV_{gb}} \right), \end{aligned} \quad (7.1.12)$$

$$\frac{d\phi_{p,sub}^{qm}}{dV_{gb}} = 1 + \frac{\gamma_p (-1 + \phi_{str}'^{qm})}{\sqrt{\gamma_p^2 + 4(V_{gb} - V_{fb} - \phi_{str}^{qm})}} - \phi_{str}'^{qm} \quad (7.1.12a)$$

in the strong inversion region. Figure 7.5 shows the regional gate capacitance with κ^{qm} variations with the QM model at low poly-gate doping.

Figure 7.5: Gate capacitance with coupled QM and poly-doping effects at low poly doping for κ^{qm} variations, compared to the Medici data.

The detailed physical parameter extraction procedure is illustrated in the following. We assume that only one C_{gg} data is given from measurement, while numerical data is generated for illustration with known device physical parameters ($t_{ox} = 20\text{nm}$, $N_{ch} = 7.5 \times 10^{17} \text{ cm}^{-3}$, $N_{gate} = 5 \times 10^{19} \text{ cm}^{-3}$, and $\kappa^{qm} = 1$). Following a step-by-step physical parameter extraction below with our unified regional charge models, the corresponding physical parameters can be extracted.

Step 1: Physical oxide thickness normally can be obtained from process. However, in our extraction procedure, initially we will assume an effective oxide thickness in our model before N_{gate} and QME extraction. This is an important step to physically extract N_{gate} as will be illustrated in Step 2. Effective oxide thickness is extracted from regional capacitance model,

$$C_{gg} = \frac{\epsilon_{ox}}{t_{ox}} \left(1 - \frac{d\phi_{ss}}{dV_{gb}} \right) \quad (7.1.13)$$

by setting other physical parameters at the default condition, $N_{gate} = 10^{22} \text{ cm}^{-3}$, $\kappa^{qm} = 0$ and $N_{ch} = 7.5 \times 10^{17} \text{ cm}^{-3}$. Automated extraction procedure can be implemented by searching the minimum error between the last two simulated and measured (point) data in strong accumulation region. As t_{ox} extraction is correlated with QME, the extracted $t_{ox} = 23.5 \text{ \AA}$ is higher than the actual physical $t_{ox} = 20 \text{ \AA}$ [97]. This will be the key to physically decouple the N_{gate} extraction from QM effect in the strong inversion region, which will be described in the following steps.

Step 2: Effective channel concentration, N_{ch} is weakly correlated to t_{ox} , κ^{qm} and N_{gate} , as illustrated in depletion gate capacitance derivation and in Chapter 3. Accurate N_{ch} can be extracted from

$$C_{gg} = \sqrt{\frac{q\epsilon_{si}N_{ch}}{2(V_{gb} - V_{fb})}} \quad (7.1.14)$$

by setting other physical parameters at the default condition, $N_{gate} = 10^{22} \text{ cm}^{-3}$ and $\kappa^{qm} = 0$ with the extracted electrical t_{ox} in step 1. Automated extraction procedure can be implemented by searching the minimum error between the simulated and measured (point) data in depletion region, as illustrated in figure 7.6.

Step 3: Poly gate-doping concentration, N_{gate} is related to the gate capacitance model

$$C_{gg} = \frac{\epsilon_{ox}}{t_{ox}} \left(1 - \frac{d\phi_{ss}}{dV_{gb}} - \frac{d\phi_{p,dd}}{dV_{gb}} \right), \quad (7.1.15)$$

which is derived without QME in the strong inversion region. However, with the extracted electrical oxide thickness in step 1, (which contains QM effect) the gradient of strong-inversion gate capacitance is found to be a strong function of N_{gate} , which is used

to extract N_{gate} . The extracted N_{gate} from this step has the same value as that used in the Medici as illustrated in the inset of figure 7.6.

Figure 7.6: Physical parameter extraction procedure for t_{ox} , N_{ch} and N_{gate} based on regional gate capacitance and first-order derivative of gate capacitance (inset).

Step 4: QME parameter, κ^{qm} is eventually coupled with N_{gate} in the strong inversion region. However N_{gate} has been extracted in the previous step and in this step, instead of using the effective oxide thickness, physical oxide thickness ($t_{ox} = 2\text{nm}$), which is available from the process, is used in κ^{qm} extraction. Without QME, it is clearly shown in figure 7.7 that the gate capacitance changes only in the strong inversion region, the capacitance gradient also changes as is shown in the inset. As N_{gate} has been extracted from the previous step, QME with κ^{qm} extraction is related to the derived regional gate capacitance model

$$C_{gg} = C_{ox} \left(1 - \frac{d\phi_{ss}^{qm}}{dV_{gb}} - \frac{d\phi_{p,dd}^{qm}}{dV_{gb}} \right) \quad (7.1.16)$$

$$\phi_{p,dd}^{qm} = \left(-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gb} - \phi_{fb} - \phi_{ss}^{qm}} \right)^2, \quad (7.1.16a)$$

in which κ^{qm} can be tuned to match the level and the gradient of the measured gate capacitance with the physical t_{ox} . This is shown in figure 7.7 with all the extracted parameters matching those used in Medici C_{gg} data.

Figure 7.7: Physical parameter extraction for κ^{qm} based on regional gate capacitance and the first-order derivative of gate capacitance (inset).

7.1.2 Model parameter extraction

A one-iteration physical parameter extraction has been illustrated in the previous section based on analytical regional capacitance models, which consist of 3 physical parameters with known physical oxide thickness in a 4-step extraction procedure. Our “non-pinned” ϕ_s -based charge model attempts to build in all physical effects, such as PAE/PDE/PIE and QME from the extracted physical parameters, into the unified regional model in all operating regions and bias conditions. Our unified regional charge model consists of 9 fitting parameters: σ_a , σ_f , σ_{eff} , δ_ϕ , ρ_r , ρ_f , σ^{qm} , δ^{qm} and ε^{qm} . The extraction is

based on *one-iteration* approach following 6-step calibration procedure, as illustrated in the following.

Step 1: From accumulation to depletion region: σ_a is considered in this calibration step to ensure smoothness. In this step, set $N_{gate} = 10^{22} \text{ cm}^{-3}$, $\kappa^{qm} = 0$ and extract N_{ch} and t_{ox} from the process. The unified regional gate charge model from accumulation to depletion can be expressed as

$$Q_g = C_{ox} (V_{gb} - \phi_{MS} - \phi_{SA}) \quad (7.1.17)$$

and the channel induced charge

$$Q_{sc} = -C_{ox} (V_{gb} - V_{fb} - \phi_{SA}), \quad (7.1.18)$$

with the unified regional surface potential

$$\phi_{SA} = \phi_{ACC} + \phi_{SUB}. \quad (7.1.18a)$$

In the above equations,

$$\phi_{ACC} = V_{GBR} + 2v_{th} \mathcal{L}\{W\} \quad (7.1.18b)$$

and

$$\phi_{SUB} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{GBF}} \right]^2 \quad (7.1.18c)$$

are the unified regional surface potentials in the accumulation and depletion regions, respectively, and

$$V_{GBR} = \mathcal{G}_r \{V_{gb} - V_{fb}; \sigma_a\} \quad (7.1.18d)$$

$$V_{GBF} = \mathcal{G}_f \{V_{gb} - V_{fb}; \sigma_f\} \quad (7.1.18e)$$

are unified reversed and forward smoothing functions. σ_a can be extracted from the model by varying its value from the default 0.001 until numerical dC_{gg}/dV_{gb} shows a smooth transition across the flat-band voltage, as shown in figure 7.8 (inset).

Figure 7.8: Unified numerical first-order and second-order derivatives of gate charge, C_{gg} and dC_{gg}/dV_{gs} (inset), respectively, in σ_a extraction. Dotted line represents the default setting while solid line for the extracted σ_a .

Step 2: Charge neutrality condition at flat-band: σ_f is a fitting parameter in the V_{GBF} function for ϕ_{SUB} . In order to satisfy charge neutrality at flat-band condition, σ_f is tuned such that

$$Q_{sc} \big|_{V_{gb}=V_{fb}} = 0. \quad (7.1.19)$$

Initially by setting σ_f to σ_a as the default value (dotted line, figure 7.9), smooth transition at the flat-band for gate capacitance gradient is observed but charge neutrality at flat-band cannot be satisfied (dotted line, inset). By fitting σ_f to satisfy charge neutrality condition at flat-band when $V_{gs} = V_{fb}$, charge neutrality is satisfied but not for the smoothness as shown by the dashed line in figure 7.9. With the fitted σ_f in step 2, σ_a is re-fitted in step 1

until both conditions for the smoothness and charge-neutrality can be satisfied (solid line).

Figure 7.9: First-order derivative of gate capacitance, dC_{gg}/dV_{gb} and the channel induced charge (inset) for σ_f and σ_a extraction in one-iteration (dashed lines), iterated loops (solid lines) and default σ_f value.

Step 3: From depletion to strong inversion region: σ_{eff} is a smoothing parameter in the unified surface potential,

$$\phi_{str}^{qm} = \phi_{s0} + \Delta^{qm} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \quad (7.1.20)$$

in which the “non-pinned” surface-potential increment is expressed as

$$\Delta^{qm} = v_{th} \ln \left\{ \frac{1}{v_{th}} \mathcal{G}_f \left\{ \left(\frac{V_{ga}^{qm}}{\gamma} \right)^2 - \left(\phi_{s0} + V_{cb} + \Delta E_g^{qm} \Big|_{\phi_s = \alpha^{qm} \phi_{SEFF}} \right); \sigma_{eff} \right\} + 1 \right\} \quad (7.1.20a)$$

$$V_{ga}^{qm} = \mathcal{G}_f \left\{ V_{gb} - V_{fb} - \phi_{str}^{qm} - \phi_{peff}^{qm}; \sigma_{vga} \right\}. \quad (7.1.20b)$$

The smoothing functions prevent the argument in the natural log from becoming negative beyond its validity regions. In this extraction procedure, σ_{vga} is set to be a constant (0.001) and σ_{eff} is extracted in the region of interest, when V_{gb} is greater than the peak

dC_{gg}/dV_{gb} location as shown by the inset of figure 7.9. Unified regional gate charge for all operating regions with poly-doping effect and QME is given by

$$Q_g^{qm} = -(Q_i^{qm} + Q_b^{qm} + Q_{ox}). \quad (7.1.21)$$

By varying σ_{eff} in the Q_g^{qm} expression, the gradient of the gate capacitance (inset) can be numerically calculated with the *same step size* as in the measurement data and σ_{eff} can be extracted based on minimum root-mean-square error between the simulated and measured data in the region of interest.

Figure 7.10: Gate capacitance and its first-order derivative (inset) showing the effect on σ_{eff} variation in the region of interest (box) with fitted (solid lines), default (dotted lines) and overestimated (dashed lines) σ_{eff} values.

Step 4: From depletion to strong inversion region: δ_ϕ is a smoothing parameter in the unified surface potential,

$$\phi_{ds}^{qm} = \mathcal{G}_{eff} \{ \phi_{SUB}, \phi_{str}^{qm}, \delta_\phi \}, \quad (7.1.22)$$

which joints smoothly ϕ_{SUB} (for the depletion region) and ϕ_{str}^{qm} (for the strong inversion region). From the extracted σ_{eff} value in the previous step, gradient of gate capacitance is

numerically calculated from the unified gate charge expression (7.1.22) with default $\delta_\phi (= 0.1\mu)$, in which glitches can be seen in the inset of figure 7.11. However, correct δ_ϕ can be extracted from the first-order derivative of C_{gg} (inset) by tuning δ_ϕ to a value until a smooth transition from depletion to strong inversion regions is obtained (solid line). δ_ϕ is extracted without using any measurement data but with fine step size. It is also observed from figure 7.11 that σ_f is uncorrelated to σ_{eff} at difference tuning region (V_{gb} below the peak of dC_{gg}/dV_{gb})

Figure 7.11: Gate capacitance and the first-order derivative (inset), showing the fitted (solid lines) and default (dotted lines) δ_ϕ values in the region of interest (box).

Step 5: N_{gate} dependence for charge-neutrality conservation: ρ_r and ρ_f are the smoothing parameters in the unified gate surface potential

$$\phi_{p,sub} = \phi_{p,dd} \Big|_{V_{gb}-V_{fb}-\phi_s=V_{gbs}} = \left(-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gbs}} \right)^2 \quad (7.1.23)$$

$$V_{gbs} = \mathcal{G}_f \{ V_{gb} - V_{fb} - \phi_s; \rho_f \} \quad (7.1.23a)$$

when the gate is depleted, and

$$\phi_{p,acc} = \phi_{p,cc} \Big|_{\phi_s = \phi_{sr}} = -v_{th} \ln \left[\frac{\gamma^2}{\gamma_p^2} \exp \left(-\frac{\phi_{sr}}{v_{th}} \right) + \left(1 - \frac{\gamma^2}{\gamma_p^2} \right) \right]^2 \quad (7.1.24)$$

$$\phi_{sr} = \mathcal{G}_r \{ \phi_s; \rho_r \} \quad (7.1.24a)$$

when the gate accumulates with electrons. A unified poly-gate surface potential that gives asymptotic $\phi_{p,acc}$ and $\phi_{p,sub}$ solutions in the accumulation and depletion regions, respectively, can be expressed as

$$\phi_{p,sa} = \phi_{p,acc} + \phi_{p,sub} . \quad (7.1.25)$$

Channel induced charge in (7.1.18) is modified with the surface-potential drop across the induce layer at the gate interface as

$$Q_{sc} = -C_{ox} (V_{gb} - V_{fb} - \phi_{sa} - \phi_{p,sa}) . \quad (7.1.26)$$

Both ρ_r and ρ_f are extracted with a range of N_{gate} values at flat-band condition such that $Q_{sc} = 0$ is always satisfied as shown in figure 7.12. The inset shows the $(V_{gb} - V_{fb})$ sweep, channel induced charge always passes through $Q_{sc} = 0$ for various N_{gate} values with the fitted pair of ρ_r and ρ_f .

Figure 7.12: ρ_r and ρ_f extraction at flat-band for charge conservation condition with a range of N_{gate} . Induced channel charge smoothly across zero charge at flat-band (inset) with a range of N_{gate} as indicated.

Step 6: QME model parameter extraction: δ^{qm} , σ^{qm} and ε^{qm} are the three parameters in the QME model in the accumulation region (3.6.5), which are re-written here for convenience,

$$\Delta E_{g,acc} = \Delta E_g + \varepsilon^{qm}, \quad (7.1.23)$$

$$\Delta E_g = \kappa^{qm} \frac{3}{8} \frac{\hbar^2}{m_e^*} \left[\frac{12 m_e^* q^2}{\varepsilon_{si} \hbar^2} \frac{2 C_{ox} (-2 v_{th} L \{W^{qm}\}) / q}{3} \right]^{2/3}, \quad (7.1.23a)$$

$$W^{qm} = \frac{\gamma}{2 \sqrt{v_{th}}} \exp \left[- \frac{V_{GB,fb} - (V_{fb} - \sigma^{qm})}{2 v_{th}} \right], \quad (7.1.23b)$$

$$V_{GB,fb} = \mathcal{G}_{eff} \{ V_{gb}, V_{fb} - \sigma^{qm}; \delta^{qm} \}. \quad (7.1.23c)$$

If these parameters are all set to the default values (zero), the consequence is well known from [45] that singularity at flat-band occurs as shown in figure 7.13 (dotted lines). In order to avoid this, a smoothing function, $V_{GB,fb}$ with smoothing parameter δ^{qm} , is

proposed such that the singularity in dC_{gg}/dV_{gb} can be removed, thereby δ^{qm} is extracted (solid lines).

Figure 7.13: δ^{qm} extraction with coupled poly-doping effect and QME at default values for σ^{qm} and ϵ^{qm} .

Although extracted δ^{qm} gives smooth transition from accumulation to depletion region, the simulated gate capacitance deviates from the measurement data across the transition region. Without altering the smoothness in the gradient of the C_{gg} across transition region, σ^{qm} is fitted to match the measurement data, as shown in figure 7.4.

Figure 7.14: σ^{qm} extraction with coupled poly-doping effect and QME at default value for ϵ^{qm} but fitted value for σ^{qm} . By fitting δ^{qm} to the measurement data, deviation of the simulated gate capacitance data is corrected (solid line) without scarifying the smoothness in the gradient of the gate capacitance.

However, even the gate capacitance is accurately modeled (compared to the measured data) with smooth transition from accumulation to depletion regions with the fitted σ^{qm} and δ^{qm} , charge-neutrality is still violated at the flat-band condition (see figure 7.15), as the induced channel charge is deviated from zero at flat-band condition. ϵ^{qm} is proposed in the model to tune back the charge neutrality condition at flat-band, as shown in figure 7.15 (solid line, inset), which shows smooth crossing zero charge at flat-band voltage, but minor deviation from the measured C_{gg} is observed in figure 7.15 (dotted line). σ^{qm} is re-fitted (solid line) to satisfy the charge-neutrality.

Figure 7.15: σ^{qm} re-calibration to the measurement data (symbol) with fitted ε^{qm} to satisfy charge neutrality at flat-band condition (inset).

Our unified regional charge model consists of 2 calibration sequences: firstly, physical parameters are extracted through a 4-step extraction procedure; and secondly, model (smoothing) parameters are extracted through a 6-step extraction procedure. The model involves 3 physical parameters with a given physical oxide thickness and 9 model parameters. With a set of extracted parameter, no further calibration is needed to predict bias-dependent bulk, source/drain (based on Ward-Dutton partitioning) and gate (based on charge neutrality) charges, as will be verified in the next chapter.

7.2 Unified DC (V_t , I_{ds}) model parameter extraction

This section presents the calibration methodology for our unified length/width-dependent MOSFET drain-current (I_{ds}) model with length/width-dependent threshold-voltage (V_t) model in the entire geometry/bias range based on 0.11 CMOS technology. The model has been formulated with built-in physical geometry effects to account for

electrical characteristics of the fabricated short-channel/narrow-width devices while maintaining Gummel symmetry.

The unified V_t model requires 4 steps to extract its 16 fitting parameters and 6 bias coefficients using point data ($V_t - L/W$ at corner bias), whereas the unified I_{ds} model requires 10 steps to extract its 15 fitting parameters and 7 geometry coefficients. The experimental V_t point data used in this research work are based on 6 x 6 matrix devices of $L_{drawn} = 10, 1, 0.5, 0.3, 0.15$ and $0.13 \mu\text{m}$ and $W = 10, 3, 1.2, 0.6, 0.3$ and $0.13 \mu\text{m}$. The following “extreme” or corner conditions are defined: the widest/longest-gate, $W_{\infty}/L_{\infty} = 10/10$, the narrowest/shortest-gate, $W_{min}/L_{min} = 0.13/0.13$, the medium gate width, $W_m = 0.6 \mu\text{m}$, low $V_{ds} = V_{d0} = 0.1 \text{ V}$, high $V_{ds} = V_{dd} = 1.32 \text{ V}$, low $V_{bs} = V_{b0} = 0$, high $V_{bs} = V_{bb} = -1.2 \text{ V}$ and medium $V_{bs} = V_{bl} = -0.6 \text{ V}$. Out of 6 x 6 matrixes for various geometry/bias sets of measured $I-V$ data from 0.11- μm technology wafer, the I_{ds} fitting procedure only required 6 $I-V$ data (at the corner geometry) plus the point $I_{dsat}(W)$ at L_{∞}/L_{min} and $g_{ds}(L)$ at W_{∞} data. The rest of the $I-V$ data are used for model verification. Even the point V_t data is not given from the measurement, it can be calculated based on the critical current at linear threshold (I_{crit} at V_{t0}) method [62] from the $I-V$ data.

7.2.1 Threshold-voltage parameter extraction

Our V_t model calibration starts from the simple long/wide-channel equation [17], ignoring short/narrow-channel effects that have not been calibrated initially, which have extremely weak influence on long/wide-channel devices. Parameters associated with short/narrow-channel effects are calibrated to long/wide-channel equation by adding SCEs, reverse SCEs, NWE, and inverse NWE in a *step-by-step* sequence at its extreme

bias and/or geometry conditions where the effect is most pronounced. V_t model consists of 3 long/wide-channel parameters (N_{ch} , V_{fbb} , α_L), 3 SCEs parameters to account for charge sharing (λ), barrier lowering and DIBL (α , ϕ), together with 2 reverse SCE parameters (β , κ), 2 NWE parameters for fringing capacitance (τ) and barrier lowering (α_w), and 3 inverse NWE parameters (β_w, κ_w, l_w) for lateral nonuniform profiles. Another 3 optimization parameters (σ , ΔW , l_u) for L_{eff} , W_{eff} and the pile-up centroid [15] for lateral doping. The V_t parameter extraction follows a 10-step procedure as illustrated in the following based on *technology-characterization* approach.

Step 1: at W_∞/L_∞ , measure V_t for a few values of V_{bs} at $V_{ds} = V_{d0}$. Fit the long/wide-channel $V_{tL} = f(V_{bs})$ equation to the measured $V_t - V_{bs}$ data to extract the channel doping parameter (N_{ch}) and metal to substrate work function ($\phi_{ms} = V_{fbb} - \phi_f$), which are then fixed in the following steps.

Step 2: at W_∞/L_∞ , measure V_t for a few values of V_{bs} at $V_{ds} = V_{dd}$. Fit long/wide-channel $V_{tL} = f(V_{bs}; N_{ch}, V_{fbb})$ equation to the measured $V_t - V_{bs}$ data to extract long-channel DIBL parameter (α_L), which is then fixed in the following steps.

Step 3: at W_∞ for each L_{drawn} , measure V_t for each L_{drawn} at $V_{ds} = V_{d0}$ and $V_{bs} = V_{bb}$. Fit $V_t = f(L_{drawn}; N_{ch}, V_{fbb}, \alpha_L)$ equation to the measured $V_t - L_{drawn}$ data to extract the charge sharing parameter (λ'), reverse SCE parameters (β , κ') and surface-potential barrier lowering parameter (α) with ϕ being set to zero. Other NWE and inverse NWE parameters in the V_t equation are irrelevant and set to negligibly small values. Extracted β and α are then fixed for the following steps but λ' and κ' will be re-calibrated for V_{bs} dependence.

Step 4: at W_∞ for each L_{drawn} , measure V_t for each L_{drawn} at $V_{ds} = V_{dd}$ and $V_{bs} = V_{bb}$.

Fit $V_t = f(L_{drawn}; N_{ch}, V_{fbb}, \alpha_L, \alpha, \beta, \lambda'$ and $\kappa')$ equation to the measured $V_t - L_{drawn}$ data to extract DIBL parameter (ϕ') at high drain bias condition. ϕ' is fixed at this bias condition and will be re-calibrated at different V_{bs} conditions for V_{bs} dependence. *Two-iteration* approach can be adopted by re-fitting model parameters in step 3 with the extracted ϕ' in step 4.

Step 5: At W_∞ for each L_{drawn} : Measure V_t for each L_{drawn} at $V_{ds} = V_{d0}$ and $V_{bs} = V_{b0}$. Fit $V_t = f(L_{drawn}; N_{ch}, V_{fbb}, \alpha_L, \alpha$, and $\beta)$ equation to the measured $V_t - L_{drawn}$ data to extract charge sharing parameter (λ'') and reverse SCE parameters (κ'') with ϕ being set to zero at this V_{bs} bias condition. If linear approximation is assumed,

$$\kappa = \kappa_0 + \kappa_1 V_{bs} \quad (7.2.1)$$

$$\kappa_0 = \kappa'', \quad \kappa_1 = (\kappa' - \kappa'')/V_{bb} \quad (7.2.1a)$$

is formulated for reverse SCE and

$$\lambda = \lambda_0 + \lambda_1 V_{bs} \quad (7.2.2)$$

$$\lambda_0 = \lambda'', \quad \lambda_1 = (\lambda' - \lambda'')/V_{bb} . \quad (7.2.2a)$$

for charge-sharing effect.

Step 6: at W_∞ for each L_{drawn} : measure V_t for each L_{drawn} at $V_{ds} = V_{dd}$ and $V_{bs} = V_{bb}$. Fit $V_t = f(L_{drawn}; N_{ch}, V_{fbb}, \alpha_L, \alpha, \beta, \lambda''$ and $\kappa'')$ equation to the measured $V_t - L_{drawn}$ data to extract DIBL parameter (ϕ'') at high drain bias condition. *Two-iteration* approach can be adopted by re-fitting model parameters in step 5 with the extracted ϕ'' in step 6. ϕ' is fixed at this bias and a linear approximation is assumed for V_{bs} dependence

$$\varphi = \varphi_0 + \varphi_1 V_{bs} \quad (7.2.3)$$

$$\varphi_0 = \varphi'', \quad \varphi_1 = (\varphi' - \varphi'')/V_{bb}. \quad (7.2.3a)$$

σ and l_μ optimization loop is performed. For each σ (outer loop) and l_μ value [18], the best extracted parameter set (N_{ch} , V_{fbb} , α_L , α , β , λ , κ and φ) from step 3–7 is selected with minimum error in all $V_t(L_{drawn})$ values. If parabolic relationship is assumed, step 7–8 may proceed for more accurate $V_t(V_{bs})$ prediction.

Step 7–8: at W_∞ for each L_{drawn} , repeat step 5 and 6 for $V_{bs} = V_{b1}$. λ'' and κ'' are extracted at $V_{ds} = V_{d0}$ condition, and φ'' is extracted at $V_{ds} = V_{dd}$. λ , κ and φ are formulated as

$$\kappa = \kappa_0 + \kappa_1 V_{bs} + \kappa_2 V_{bs}^2 \quad (7.2.4a)$$

$$\lambda = \lambda_0 + \lambda_1 V_{bs} + \lambda_2 V_{bs}^2 \quad (7.2.4b)$$

$$\varphi = \varphi_0 + \varphi_1 V_{bs} + \varphi_2 V_{bs}^2 \quad (7.2.4c)$$

Figure 7.16: Measured (symbols) and modeled (lines) $V_t - L_{drawn}$ curves at the indicated bias conditions showing a 8-step V_t parameter extraction procedure for SCEs and reverse SCE at W_∞ condition.

Following the above 8-step extraction procedure, 10 V_t model parameters plus 6 bias coefficients, which are associated with SCE and reverse SCE parameters, are extracted. V_t equation with width dependence is characterized in the NWE and inverse NWE, following a 2-step calibration procedure as illustrated in the following.

Step 9: at L_∞ for each W_{drawn} , measure V_t for each W_{drawn} at $V_{ds} = V_{d0}$ and $V_{bs} = V_{bb}$. With a few trial values of ΔW (ranging, e.g. $0 \sim W_{min}$), fit $V_t = f(W_{drawn}; N_{ch}, V_{fbb}, \alpha_L, \alpha, \beta, \lambda, \kappa \text{ and } \varphi)$ equation to the measured $V_t - W_{drawn}$ data to extract fringing capacitance parameter (τ). τ is fixed in the following step.

Figure 7.17: Measured (symbols) and modeled (line) $V_t - W_{drawn}$ curves at the indicated bias condition showing V_t extraction for fringing capacitance at L_∞ condition.

Step 10: at L_{min} for each W_{drawn} , measure V_t for each W_{drawn} at $V_{ds} = V_{d0}$ and $V_{bs} = V_{bb}$. Fit $V_t = f(W_{drawn}; N_{ch}, V_{fbb}, \alpha_L, \alpha, \beta, \lambda, \kappa, \varphi \text{ and } \tau)$ equation to the measured $V_t - W_{drawn}$ data to extract NWE for barrier lowering parameter (α_w) and 3 inverse NWE (β_w, κ_w, l_w) for nonuniform profiles along the width dimension. Hence, this completes the V_t model parameter extraction procedure and it is ready for geometry/bias prediction.

Figure 7.18: Measured (symbols) and modeled (line) $V_t - W_{drawn}$ curves at the indicated bias conditions showing step 10 V_t extraction for NWE and inverse NWE at L_∞ condition.

7.2.2 Drain-current parameter extraction

Our unified I_{ds} model has an extremely compact form, which has been derived physically to model device geometry effects, i.e., SCEs and NWEs and to predict bias-dependent electrical characteristics of real MOSFET electrical behavior.

In our unified I_{ds} calibration procedure, starting from simple long/wide-channel equations, geometry effects are added into the model in a *step-by-step* sequence at the corner geometry/bias conditions where each effect is most pronounced. Our unified I_{ds} model is constructed from:

- (1) Vertical-field mobility (μ_o) with 4 parameters (μ_1 , μ_2 , μ_3 , ν) as a function of the effective field (E_{eff}), which has 1 parameter for bulk-bias dependence (χ_b); and lateral-field mobility (μ_{eff0}), which includes velocity saturation (v_{sat}) and overshoot (g_{vo}) derived from energy-balance formulation together with quasi-2D solution and expressed in the form of an effective Early voltage (V_{Aeff}), with 1 parameter (ξ) for all geometry and bias, as well as the factor $\delta_0 = \delta_L V_{deff}/V_{dsat}$ for Gummel symmetry, with 1 parameter (δ_L);

- (2) The effective drain voltage (V_{deff}), which is a smoothing function for the transition from liner to saturation region, with 1 parameter (δ_s); and the gate-drain overdrive voltages (V_{gg} , V_{gd}), which are interpolating functions for the transition from subthreshold to strong inversion region, with 1 parameter (V_{off}) for calibrating the subthreshold current;
- (3) Source/drain (S/D) series resistance (R_{sd}), which has 2 parameters (ρ , ν), included in the “short-channel” effective mobility (μ_{eff}) that approaches μ_{eff0} for long-channel drain current when R_{sd} can be ignored; and
- (4) Width dependence in I_{ds} model, including 1 parameter (μ_f) for Coulombic mobility correction and 2 parameters for the extrinsic resistance due to dog-bone layout (θ_s , θ_d).

The I_{ds} parameter extraction follows a 10-step procedure as illustrated in the following based on *transistor-characterization* approach. Initially before starting the parameter extraction procedure, all the parameters are set to the default values ($v_{sat} = 8 \times 10^6$ cm/s, $\chi_b = 1$, $\delta_o = 0$, $\delta_s = 0.005$, $g_{ov} = 1$, $V_{off} = 0$ and $R_{sd} = 0$, $\mu_f = 1$) such that the parameters that have not been calibrated initially do not have appreciable influence in the current calibration step. The detailed extraction steps are as follows.

Step 1: at W_∞/L_∞ , measure I_{ds} - V_{gs} data at $V_{ds} = V_{d0}$ and $V_{bs} = V_{b0}$. Fit the drain current equation in the strong inversion region,

$$I_{ds} = f(V_{gs}, V_t) \quad (7.2.5)$$

with the vertical field mobility

$$\mu_0 = \frac{\mu_1}{1 + \left(\frac{\mu_1}{\mu_2}\right) E_{eff}^{\frac{1}{3}} + \left(\frac{\mu_1}{\mu_3}\right) E_{eff}^\nu} \quad (7.2.5a)$$

to the measurement data to extract mobility parameters (μ_1 , μ_2 , μ_3 , ν), which are then fixed in the following steps.

Figure 7.19: Drain-current calibration with long/wide-channel device for mobility parameter extraction as shown by the fitting with vertical-field effect (solid lines) and constant coulombic-scattering parameter, μ_1 (dashed lines).

Step 2: at W_∞/L_∞ , measure $I_{ds} - V_{gs}$ data at $V_{ds} = V_{dd}$ and $V_{bs} = V_{b0}$. Fit the drain current equation in the saturation region,

$$I_{ds} = f(V_{gs}, V_t, \mu_0), \quad (7.2.6)$$

with the lateral field mobility

$$\mu_{eff0} = \frac{g_{ov}\mu_0}{1 + \frac{\delta_0 V_{dsat}}{E_{sat} L_{eff}}}; \quad g_{ov} = 1, \quad (7.2.6a)$$

and $\delta_0 = f(\delta_L)$ to the measurement data to extract δ_L , which is then fixed in the following step.

Figure 7.20: Drain-current calibration with long/wide-channel device for effective mobility with lateral-field effect with fitted (solid lines) and (dashed lines) default values.

Step 3: at W_{∞}/L_{∞} , measure $I_{ds} - V_{gs}$ data at $V_{ds} = V_{d0}$ and $V_{bs} = V_{bb}$. Fit the drain current equation in the linear region,

$$I_{ds} = f(V_{gs}, V_t, \mu_{eff0}) \quad (7.2.7)$$

with the effective transverse field

$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left[\left(V_{gs} - V_{tL} - \frac{A_b V_{ds}}{2} \right) + \frac{\chi_b \gamma \sqrt{\phi_{s0}}}{\zeta_n} \left(1 + \frac{V_{ds} - 2V_{bs}}{4\phi_{s0}} \right) \right] \quad (7.2.7a)$$

to the measurement data to extract χ_b , which is then fixed in the following step. However, the extracted χ_b may not be the same as the default value due to different contribution of bulk-charge in the effective field for different processes. If the extracted χ_b value is different from the default value, iterate steps 1 to 3 with the extracted χ_b as the new initial guess.

Figure 7.21: Drain-current calibration with long/wide-channel device for effective transverse field bulk-bias dependence effect with fitted (solid lines) and (dashed lines) default value.

Step 4: at W_∞/L_∞ , measure $I_{ds} - V_{gs}$ data at $V_{ds} = V_{d0}$ and $V_{bs} = V_{b0}$. Fit the drain current equation in the subthreshold region,

$$I_{diff} = f(V_{gs} : V_t, \mu_{eff0}, \chi_b) \quad (7.2.8)$$

to the measurement data to extract V_{off} , which is then fixed in the following steps.

Figure 7.22: Subthreshold-current calibration by fitting the regional diffusion current $I_{diff}(V_{off})$ (dashed line) after the regional I_{drift} calibration (dotted line), compared to the one before fitting ($V_{off} = 0$). The final unified I_{ds} is shown by the solid line.

Step5: at W_{∞}/L_{min} , measure $I_{ds} - V_{gs}$ data at $V_{ds} = V_{d0}$ and $V_{bs} = V_{b0}$. Fit the drain-current equation in the linear region,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff0}, \chi_b, V_{off}) \quad (7.2.9)$$

with source/drain series resistance

$$R_s = \frac{R_{ext} + R_{int}}{2} = r_1 + \frac{r_2}{C_{ox}W \left(V_{gs} - V_t - \frac{A_b}{2} V_{ds} \right)}. \quad (7.2.9a)$$

to the measurement data to extract ρ and ν (6.2.8), which is then fixed in the following steps.

Figure 7.23: Drain-current calibration with short/wide-channel device for source/drain series resistance extraction after fitting (solid lines) and before fitting (dashed lines).

Channel-length modulation has no effect in this operating region for R_{sd} parameter extraction and it is safe to set $g_{ov} = 1$. Even R_{sd} is well calibrated to the measured $I_{ds} - V_{gs}$ data with the default $\delta_s = 0.005$ in the smoothing function

$$V_{deff} = V_{dsat} - \frac{1}{2} \left[V_{dsat} - V_{ds} - \delta_s + \sqrt{(V_{dsat} - V_{ds} - \delta_s)^2 + 4\delta_s V_{dsat}} \right], \quad (7.2.10)$$

g'_d from the model as shown in the inset of figure 7.24 clearly cannot match the measured $g'_d - V_{ds}$ data for short-channel device. Before calibrating channel-length modulation, δ_s is finely tuned by increasing the value together with re-calibration of R_{sd} until matching the measured $g'_d - V_{ds}$ data as shown in the inset. The drain current is improved in the transition region with the tuned δ_s and fitted R_{sd} .

Figure 7.24: dg_d/dV_{ds} calibration with short/wide-channel device for δ_s and source/drain series resistance after fitting (solid lines) and before fitting (dashed lines).

Step 6: at W_∞/L_{min} , measure $I_{ds} - V_{ds}$ data at $V_{gs} = V_{gg}$ and $V_{bs} = V_{b0}$. g_d can be calculated numerically from the measurement or simulated $I_{ds} - V_{ds}$ data. Fit the numerical g_d from drain current equation,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s, R_{sd}) \quad (7.2.11)$$

with velocity overshoot in the channel-length modulation effect,

$$g_{vo} = 1 + (V_{ds} - V_{deff}) / V_{Aeff} \quad (7.2.11a)$$

with $V_{Aeff}(\xi)$ to the measured g_d data to extract ξ . Figure 7.23 shows the drain current and g'_d (inset) for the value of ξ before (dashed lines) and after (solid lines) extraction. The

calibrated ξ is fixed for the following steps. However, if simulated I_{on} ($V_{gs} = V_{ds} = V_{dd}$) cannot match the measured data, default v_{sat} value can be finely tuned and the extraction needs to re-start from step 5 to match the modeled I_{on} to the measured data.

Figure 7.25: g_d calibration with short/wide-channel device for $V_{aef}(\xi)$ extraction after fitting (solid lines) and before fitting (dashed lines).

This completes the parameter extraction at two corners (L_{∞} , L_{min}) for W_{∞} for the drain-current equation,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s, R_{sd}, \xi) \quad (7.2.12)$$

which consists of geometry/bias-dependent V_t , effective vertical/lateral-field mobility with overshoot, subthreshold current, effective drain voltage, series resistance and channel-length modulation. The predicted g_d with drain-current equation (7.2.12) and fixed δ_s from step 5 is shown in figure 7.26 (dotted lines). One can see that g_d is overestimated in the saturation region as channel length increases. However, it is clearly shown that with length-dependent δ_s (solid lines), g_d in the saturation region is greatly improved. The length-dependent δ_s extraction procedure is discussed in the following steps.

Figure 7.26: Predicted $g_d - V_{ds}$ data with length-dependent δ_s (solid lines) and constant δ_s extracted at short-channel (dashed lines), compared to the measurement data (symbols) for various channel-length devices.

Step 7: at W_∞ , measure $g_d - L_g$ (point) data at $V_{gs} = V_{gg}$, $V_{bs} = V_{b0}$ and $V_{ds} = 0$. For each gate length, g_d is calculated numerically from the drain-current equation,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, R_{sd}, \xi) \quad (7.2.13)$$

and δ_s is extracted (re-tuning δ_s as shown in figure 7.27 (a) and (b)) when the simulated g_d matches the measured data. A length-dependent δ_s model is proposed,

$$\delta_s = \delta_{s1} + \frac{(\delta_{s2} - \delta_{s1})}{1 + \left(\frac{L_g}{\delta_{s3}}\right)^{\delta_{s4}}} \quad (7.2.13a)$$

and is fitted to the extracted $\delta_s - L_g$ data as shown in figure 7.28. Thus, the length-dependent I_{ds} model

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s[L], R_{sd}, \xi) \quad (7.2.14)$$

parameter extraction procedure is complete.

(a)

(b)

Figure 7.27: Drain conductance at $V_{ds} = 0$, $V_{bs} = 0$ and $V_{gs} = V_{gg}$ for (a) $L_g = 0.13\text{-}\mu\text{m}$ and (b) $L_g = 0.5\text{-}\mu\text{m}$ devices.

Figure 7.28: $\delta_s - L_g$ behavior with the length-dependent model (line) fitted to the extracted δ_s .

The length-dependent I_{ds} model is extended to width dimension, by following the extraction procedure for NWEs (μ_f) and dog-bone layout parameters (θ_s, θ_d).

Step 8: at L_∞ : measure $I_{dsat} - W$ (point) data at $V_{gs} = V_{ds} = V_{gg}$, $V_{bs} = V_{b0}$ and $V_{ds} = V_{dd}$. For each W , fit

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s[L], R_{sd}, \xi) \quad (7.2.15)$$

to each measured I_{dsat} value to extract the NWE parameter (μ_f). The extracted $\mu_f - W$ data are plotted and fitted with the Coulombic model parameters ($\mu_{f1}, \mu_{f2}, \mu_{f3}, \mu_{f4}$) in the μ_f expression,

$$\mu_f = \mu_{f1} + \frac{(\mu_{f2} - \mu_{f1})}{1 + \left(\frac{W}{\mu_{f3}}\right)^{\mu_{f4}}}, \quad (7.2.15a)$$

which are then fixed in the I_{ds} model.

Step 9: at W_m/L_{min} , measure $I_{ds} - V_{gs}$ data at $V_{ds} = V_{d0}$ and $V_{bs} = V_{b0}$. Fit the I_{ds} equation,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s [L], R_{sd}, \xi, \mu_f [W]) \quad (7.2.16)$$

to the measured I_{ds} data to extract dog-bone layout parameters (θ_s, θ_d), which are fixed in the following steps.

Figure 7.29: Drain-current calibration with short/median-width device for dog-bone layout effect.

Step 10: at W_{min}/L_{min} : measure I_{dsat} (point) data at $V_{gs} = V_{ds} = V_{gg}$ and $V_{bs} = V_{b0}$.

Fit the I_{ds} equation,

$$I_{ds} = f(V_{gs} : V_t, \mu_{eff}, \chi_b, V_{off}, \delta_s [L], R_{sd}, \xi, \theta_s, \theta_d, \mu_f [W]) \quad (7.2.16)$$

with

$$\mu_f = \left[1 + \mu_{f5} \frac{L_{min}}{L_g} \frac{W_{min}}{W} \right] \left[\mu_{f1} + (\mu_{f2} - \mu_{f1}) / \left(1 + \left(\frac{W_{eff}}{\mu_{f3}} \right)^{\mu_{f4}} \right) \right], \quad (7.2.16a)$$

to measured I_{dsat} data to extract the Coulombic model parameter (μ_{f5}) in the μ_f expression.

The predicted result with and without dog-bone effect (DB) parameters and μ_{f5} at decreasing channel width for short-channel device is illustrated in figure 7.30.

Figure 7.30: Predicted turn-on current at decreasing channel-width for short-channel devices with fitted μ_{f5} and dog-bone (DB) effect, (solid line) and without μ_{f5} for fitted DB (dotted line) and no DB (dashed line).

In summary, to extract the 15 parameters plus 7 geometry coefficients in the I_{ds} model, only 6 $I - V$ data (at corner geometry) plus the $I_{ds}(W) - L_{\infty}/L_{min}$ and $g_{ds}(L) - W_{\infty}$ (point) data are needed. The calibrated I_{ds} model is capable of predicting the whole range of geometry/bias variations, which will be verified with a 0.11- μm CMOS technology data in the following chapter.

CHAPTER 8: MODEL VERIFICATION

This chapter is devoted to demonstration and verification of our AC/DC modeling methodology with a set of experimental data based on 0.11- μm CMOS technology wafer. As we have limited AC measurement data (1 C_{gg} and 3 C_{gc}), most of the AC model verification is based on 2-D Medici numerical simulation for various physical parameter and bias variations. However, one measured C_{gg} is used for calibration and another 3 C_{gc} for prediction. We have a complete set (6 x 6 matrix, lengths x widths devices) DC measurement data for various geometry/bias variations. In our DC calibration approach, only point data ($V_t - L/W$ at corner bias) for V_t calibration and a few $I - V$ measurement data 6 $I - V$ data at corner geometries plus the $I_{ds}(W) - L_{\infty}/L_{min}$ and $g_{ds}(L) - W_{\infty}$ point data) for I_{ds} calibration are needed. Other available DC measurement data have not been used in parameter extraction and are used for model prediction.

8.1 Unified AC (charge) model verification and discussion

As discussed in chapter four, our AC model consists of terminal intrinsic charges, Q_d , Q_s and Q_b at the drain, source and bulk, respectively. Inversion charge is defined as the sum of the drain and source charges. Gate charge is formulated from charge neutrality, which is the sum of the inversion, bulk and oxide trapped charges. Intrinsic capacitance can be numerically calculated from terminal charge with respect to voltage from (4.2.14). Measurement data is extracted from 0.11- μm -technology wafer and numerical data are generated based on a 90-nm process developed at Chartered Semiconductor Manufacturing Ltd. (CSM). Measurement $C - V$ data can be extracted from wafer using split $C - V$ method and numerical $C - V$ can be generated from 2-D

simulator for various physical parameters (i.e., poly-gate doping concentration, N_{gate} , QME factor, κ^{qm} , permittivity, ϵ_r and oxide thickness, t_{ox}) with various bias dependencies. Based on the AC extraction procedure outlined in section 7.1.1 of chapter seven, only one C_{gg} data from measurement or numerical simulation is used for calibration, the set of model parameters σ_f , σ_a , σ_{eff} , δ_ϕ , ρ_f , ρ_r , σ^{qm} , ϵ^{qm} and δ^{qm} and physical parameter t_{ox} , N_{gate} , N_{ch} and κ^{qm} values are extracted and tabulated in Table 8.1 and 8.2, respectively. Table 8.3 lists physical constants and model parameters used in the 90-nm numerical process/device simulation.

Table 8.1: AC model smoothing parameters extracted for CSM's 90nm nMOSFET 2-D numerical simulator and 0.11- μm technology experimental data.

Symbol	Appear in Eq. #	Extracted value	Unit	Remark
σ_a	(3.1.14a)	0.007		Smoothing parameter in $\mathcal{G}_r$ function for unified accumulation channel surface- potential formulation
σ_f	(3.1.15a)	0.0155		Smoothing parameter in $\mathcal{G}_f$ function for unified depletion channel surface-potential formulation
σ_{eff}	(3.1.17a)	0.1		Smoothing parameter in $\mathcal{G}_f$

				function for unified channel surface-potential increment formulation
δ_ϕ	(3.1.18)	0.66×10^{-3}		Smoothing parameter in $\mathcal{G}_{eff}$ function for unified ϕ_{seff}/ϕ_{ds} formulation
ρ_r	(3.2.15a)	0.011		Smoothing parameter in $\mathcal{G}_r$ function for unified accumulation gate surface-potential formulation
ρ_f	(3.2.14a)	0.188×10^{-3}		Smoothing parameter in $\mathcal{G}_f$ function for unified depletion gate surface-potential formulation
σ^{qm}	(3.6.4)	0.1	V	Fitting parameter that shifts the model capacitance at the transition from accumulation to depletion to the actual measurement data in QME model

δ^{gm}	(3.6.4a)	0.01		Smoothing parameter in $\mathcal{G}_f$ function for unified QME model from accumulation to depletion regions
ε^{gm}	(3.6.5)	-0.04	EV	Fitting parameter to conserve charge neutrality at flat-band condition

Table 8.2: AC model physical parameters used for CSM's 90nm nMOSFET 2-D numerical simulator and (0.11- μm technology experimental) data.

Quantity	Symbol	Value	Unit
Gate Oxide Thickness	t_{ox}	20 (20.05)	$\text{\AA}$
Channel doping concentration	N_{ch}	$7.5 (6.5) \times 10^{17}$	cm^{-3}
Poly gate doping concentration	N_{gate}	$5 \times 10^{19} (3 \times 10^{20})$	cm^{-3}
QME model parameter	κ^{gm}	1 (1.01)	
Dielectric constant	K_{ox}	3.9	

Table 8.3: Physical constants and model parameters used in CSM's 90nm nMOSFET Medici and 0.11- μm technology experimental data.

Quantity	Symbol	Value	Unit
Oxide trapped charge	Q_{ox}	3×10^{10}	C/cm^2
Source/drain doping conc.	N_{sd}	1×10^{21}	cm^{-3}

Source/drain junction depth	X_j	0.075	μm
Intrinsic concentration of Si	n_i	1.5×10^{10}	cm^{-3}
Electron charge	q	1.6×10^{-19}	C
Permittivity of free space	ϵ_0	8.854×10^{-14}	F/cm
Relative permittivity of Si	ϵ_{si}	11.8	
Thermal voltage	V_{th}	0.0259	V
Planck's constant	$\hbar$	$6.6260755 \times 10^{-34}$	Js
Effective mass	m_o	9.10956×10^{-31}	kg
Electron mass	m_e	0.31	m_o
Hole mass	m_h	0.16	m_o

Figure 8.1 shows the comparison of the model results (lines) to the measurement and numerical data (symbols). Solid line is the fitted solution at $V_{ds} = V_{bs} = 0$ bias condition while the dashed lines show the predicted model results on gate-to-channel capacitance, C_{gc} at various $V_{bs} = 0, -0.6\text{V}$ and -1.2V . Since we have limited set of measurement data as shown in figure 8.1a, numerical data are generated for various V_{ds}/V_{bs} bias conditions ($V_{ds}=0, 0.6\text{V}$ and 1.2V , and $V_{bs}=0, -0.6\text{V}$ and -1.2V) as shown in figure 8.1b. Solid line is the calibrated model solution and dashed lines are the predicted results. Our AC model predicts both sets of measurement and numerical data excellently.

(a)

(b)

Figure 8.1: Unified AC model with fitted (solid line) and predicted (dashed lines) results and compared to the (a) measurement, (b) numerical data for various V_{bs} and V_{ds} bias conditions.

In order to verify our AC model, sets of numerical data are generated for various gate oxide thicknesses, poly doping concentrations and quantum factors, κ^{qm} as tabulated in the table 8.4. With the numerical simulator, the generated data is very useful to show our unified charge model prediction for the particular physical effect on MOSFET

electrical characteristics, which might be difficult with measurement data as all effects are coupled.

Table 8.4: Medici generated gate capacitances with a list of physical parameter variation in figure 8.2 (a), (b) and (c).

Figure 8.2	$N_{gate} (\text{cm}^{-3})$	κ^{qm}	$t_{ox}(\text{\AA})$	K_{ox}
a	10^{22} , 5×10^{20} and 5×10^{19}	0	20	3.9
b	10^{22}	0, 0.5, 1, 1.5	20	3.9
c	5×10^{19} and 10^{19}	0	20, 40	3.9, 7.5

Without re-fitting to the numerical data, our unified charge model is used to predict numerical solutions as shown in figure 8.2. Unique features in the AC characteristics are well captured with our modeling approach, as shown in the insets of the figure, which will be illustrated in the following.

(a)

(b)

(c)

Figure 8.2: Gate capacitances with physical parameter variation listed in table 8.4. Our unified AC model is simulated accordingly and model data (lines) is compared to the generated Medici data (symbols).

Poly-accumulation/depletion effect is shown in figure 8.2 (a), QME with κ^{qm} on output AC electrical characteristics is shown in figure 8.2(b), and the effect from gate oxide thicknesses and dielectric constants (7.5 for nitrided oxide) is shown in figure 8.2 (c).

Note that in figure 8.2(a), the capacitance “cross-over” behavior is modeled excellently by our model with the smooth transition across operating regions. In figure 8.2(b), the model solutions agree to the charge conservation at flat-band condition (inset). In figure 8.2(c), our model solutions with PIE well capture AC characteristics beyond recovery voltage (V_{rec}) as defined in section 3.2.2. Without re-fitting to the numerical data, our unified AC model (lines) matches Medici data (symbols) excellently for all physical parameter variations.

Our proposed intrinsic gate charge has been validated with measurement and numerical $C - V$ data as shown in figure 8.1 and 8.2 for various terminal biases and physical parameter variations. In the following, our unified intrinsic drain, source and bulk charges are validated with numerical data for various bias conditions. Numerical $C - V$ data are generated for bulk, drain/source-to-gate capacitances, $C_{bg}(Q_b)$, $C_{dg}(Q_d)$ and $C_{sg}(Q_d)$, as shown in figures 8.3 – 8.5 at $V_{ds}=0, 0.6$ and $1.2V$, and $V_{bs}=0, -0.6$ and $-1.2V$. Our unified charge models are used to predict AC characteristics with various terminal bias combinations and compared with the numerical data. It is shown that our unified AC model (lines) predicts Medici data (symbols) excellently based on “non-pinned” surface potential for our unified regional charge formulation.

Figure 8.3: A comparison for our bulk capacitance, C_{bg} to Medici data at various bias conditions. The inset shows model prediction on the first-order derivative of the bulk capacitance.

Figure 8.4: A comparison for our unified drain charge model (lines) to the Medici data (symbols) on (a) Drain capacitance, C_{dg} and (b) drain charge, Q_d at various bias conditions.

Figure 8.5: A comparison for our unified source charge model (lines) to the Medici data (symbols) on (a) Source capacitance, C_{sg} and (b) source charge, Q_s at various bias conditions.

Figure 8.6: Sixteen capacitances (self/trans-capacitances) vs. drain voltage at high gate terminal bias with the Medici data (symbol) and AC model (lines).

Since MOSFET consists of four terminals, there are 16 possible capacitances, including four self-capacitances (C_{gg} , C_{dd} , C_{ss} and C_{bb}) and twelve trans-capacitances (C_{gd} , C_{gs} , C_{gb} , C_{dg} , C_{ds} , C_{db} , C_{sg} , C_{sd} , C_{bs} , C_{bg} , C_{bd} and C_{bs}). These capacitances are

generated from Medici (symbols) as shown in figure 8.6. Our AC model (lines) predicts excellently the Medici data. Our AC model shows a symmetric AC behavior at $V_{ds} = 0$ condition with the smooth transition across regions of operation.

Our unified AC model has been calibrated to both the measurement and Medici data with one C_{gg} data as illustrated in the previous chapter. With the same set of model parameters, our model predicts excellently the coupled poly-doping and quantum-mechanical effects at various bias conditions. Our modeling methodology allows decoupling of both poly-doping and quantum-mechanical effects in parameter extraction sequence as well as model playback by setting N_{gate} to high value (no poly-doping effect) or setting $\kappa^{qm} = 0$ (no QME). Since our unified AC model consists of drain, source and bulk charges, gate and inversion charges can be formulated from charge neutrality. Lastly, all the 16 trans-capacitances are accurately modeled without any discontinuity and asymmetry problems.

8.2 Threshold-voltage model verification

The threshold-voltage (V_t) model outlined in chapter five is verified with a set of measurement data with drawn length $L_g = 10, 1, 0.5, 0.3, 0.15$ and $0.13 \mu\text{m}$, and drawn width $W = 10, 3, 1.2, 0.6, 0.3$ and $0.13 \mu\text{m}$ (6 X 6 matrix). The data consist of a set of $I - V$ measurement, in which the substrate bias varies from 0 to -1.2V with -0.6V step and drain biases are 0.1V (low), 0.6V (medium) and 1.2V (high). Point V_t data can be extracted from measured $I - V$ with *Constant Current* in “ $I_{crit} @ V_{t0}$ ” definition [62].

Based on a 10-step threshold-voltage calibration procedure in section 7.2.1 with 6 $V_t(L_g)$ at W_∞ and 2 $V_t(W)$ at L_{min}/L_∞ point measured data, a set of model parameters (N_{ch} ,

ϕ_{ms} , α_L , λ_0 , λ_1 , λ_2 , β , κ_0 , κ_1 , κ_2 , α , φ_0 , φ_1 , φ_2 , τ , α_w , β_w , κ_w and l_w) and optimization parameters (σ , l_μ and ΔW) are extracted and tabulated in Table 8.5 with the same physical constant as tabulated in Table 8.3.

Table 8.5: Threshold-voltage model parameters extracted for 0.11- μm technology experimental data.

Symbol	Appear in Eq. #	Extracted value	Unit	Physical interpretation
N_{ch}	(5.2.3a)	10.3031×10^{17}	cm^{-3}	The average vertical doping concentration in the channel
V_{fbb}	(5.1.3)	-0.5293	V	Regression parameter linked to metal-gate work function
α_L	(5.2.8a)	0.001		Long-channel DIBL parameter for characteristic length
λ_0	(7.2.4b)	0.1401		SCE for charge sharing parameter at V_{b0}
λ_1	(7.2.4b)	-0.2245	V^{-1}	The second parameter in parabolic V_{bs} dependence for charge sharing effect
λ_2	(7.2.4b)	-0.2497	V^{-2}	The third charge sharing parameter for parabolic V_{bs}

				dependence
β	(5.3.15b)	0.4055	$\mu\text{m}/V^{0.25}$	Characteristic length parameter for lateral doping profile
κ_0	(7.2.4a)	0.2012		Ratio of peak halo-doping to the substrate doping at V_{b0}
κ_1	(7.2.4a)	-0.0202	V^{-1}	The second parameter in parabolic V_{bs} dependence for the ratio of peak halo-doping to the substrate doping
κ_2	(7.2.4a)	-0.1034	V^{-2}	The third parameter in parabolic V_{bs} dependence for the ratio of peak halo-doping to the substrate doping
α	(5.3.5)	0.01	$\mu\text{m}/V^{0.25}$	Characteristic length parameter for potential barrier lowering in short-channel
φ_0	(7.2.4c)	7.5829		Regression parameter that enhances the region of

				validity for DIBL effect at low V_{bs}
φ_1	(7.2.4c)	6.8906	V^{-1}	The second parameter in parabolic V_{bs} dependence for DIBL effect
φ_2	(7.2.4c)	4.7880	V^{-2}	The third parameter in parabolic V_{bs} dependence for DIBL effect
τ_p	(5.4.1b)	25000		Process-dependent parameter in the fringing capacitance
α_w	(5.4.6b)	0.05	$\mu\text{m}/V^{0.25}$	Characteristic length parameter in potential barrier lowering for NWE
β_w	(5.4.2a)	0.07	$\mu\text{m}/V^{0.25}$	Characteristic length parameter in doping profile along width-dimension for INWE
κ_w	(5.4.2)	0.7		Ratio of peak doping to the substrate doping for INWE
l_w	(5.4.2)	0	μm	Pile-up centroid in doping profile along width dimension

σ	(5.1.7a)	0.4		Lateral diffusion factor for LDD junction
l_μ	(5.3.15)	-0.04	μm	Pile-up centroid in halo-doping profile
ΔW	(5.1.7b)	0	μm	Optimization for drawn gate width

Figure 8.7 shows a comparison for our model (lines) with the fitted (solid lines) and predicted (dotted lines) solutions to the measurement data (symbols) at decreasing channel length for wide-channel at various terminal bias conditions, $V_{bs} = 0$ to -1.2V with -0.3V step for $V_{ds} = 0.1\text{V}$ and 1.32V .

Figure 8.7: Threshold voltage at decreasing channel length for wide-channel model results lines (solid and dashed lines for fitted and predicted, respectively) are compared to the measurement data (symbols).

(a)

(b)

Figure 8.8: Threshold voltage at decreasing channel width for (a) long-channel, (b) short-channel model results (solid and dashed or dotted lines for fitted and predicted, respectively) are compared to the measurement data (symbols).

Lines in figure 8.7 and 8.8 are generated from our unified V_t model as discussed in chapter five. It shows our model solutions matches the measurement data. Once the correct model parameters are extracted following a 10-step V_t extraction procedure, V_t

model not only can fit to the measurement data well but also predict the measured V_t data accurately as shown by the dotted/dashed lines.

8.3 Drain-current model verification

After extracting correct model parameter for our geometry/bias-dependent V_t model, subsequently our unified drain current model parameter (μ_1 , μ_2 , μ_3 , ν , χ_b , δ_L , r_1 , r_2 , ξ , δ_{s1} , δ_{s2} , δ_{s3} , δ_{s4} , μ_{f1} , μ_{f2} , μ_{f3} , μ_{f4} , θ_s , θ_d , and μ_{f5}) as tabulated in Table 8.6, can be extracted from a 10-step extraction procedure as presented in section 7.2.2 with 6 $I - V$ measurement data (at corner geometry) plus the $I_{ds}(W) - L_{\infty}/L_{min}$ and $g_{ds}(L) - W_{\infty}$ (point) data.

Table 8.6: Drain-current model parameters extracted for 0.11- μm technology experimental data.

Symbol	Appear in Eq. #	Extracted value	Unit	Physical interpretation
μ_1	(6.1.11)	2.7812×10^8	$\text{cm}^2/(\text{MV.s})$	Mobility parameter for Coulombic scattering
μ_2	(6.1.11)	1.7571×10^{10}	$\text{cm}^{5/3}/(\text{MV}^{2/3}.\text{s})$	Mobility parameter for surface roughness scattering
μ_3	(6.1.11)	9.1612×10^8	$\text{cm}^{2-[v]}/(\text{MV}^{1-[v]}.s)$	Mobility parameter for phonon scattering

ν	(6.1.11)	4.4762		Regression parameter for the amount of effective field contribution in mobility due to phonon scattering
χ_b	(6.1.18)	0.3060		Regression parameter for the amount of bulk-charge contribution in effective field
δ_L	(6.1.23)	5.3244		Fitting parameter to adjust the amount of mobility degradation due to lateral-field effect
V_{off}	(6.4.10)	-0.12	V	Offset voltage for I_{off} adjustment
r_1	(6.2.8c)	0.6404×10^{-4}	$\Omega\text{-m}$	Extrinsic series resistance parameter
r_2	(6.2.8d)	0.0044×10^{-4}	$\Omega\text{V-m}$	Intrinsic series resistance parameter
ξ	(6.2.17b)	0.24	V^{-1}	Effective Early voltage parameter
δ_{s1}	(7.2.13a)	0.0045		Length-dependent parameter in δ_s for V_{deff}
δ_{s2}	(7.2.13a)	0.055		Length-dependent parameter in δ_s for V_{deff}

δ_{s3}	(7.2.13a)	0.4734	μm	Length-dependent parameter in δ_s for V_{deff}
δ_{s4}	(7.2.13a)	2.5021		Length-dependent parameter in δ_s for V_{deff}
μ_{f1}	(7.2.15a)	1.0003		Width-dependent parameter in Coulombic mobility degradation
μ_{f2}	(7.2.15a)	0.7634		Width-dependent parameter in Coulombic mobility degradation
μ_{f3}	(7.2.15a)	1.6203	μm	Width-dependent parameter Coulombic mobility degradation
μ_{f4}	(7.2.15a)	2.3613		Width-dependent parameter in Coulombic mobility degradation
θ_s	(6.3.3)	0.5		Dog-bone layout-dependent parameter in the extrinsic series source/drain resistance
θ_d	(6.3.3)	10	μm	Dog-bone layout-dependent parameter in the extrinsic series source/drain resistance

μ_{f5}	(6.3.4)	0.5		Geometry-dependent parameter in Coulombic mobility degradation
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Figures 8.9 – 8.15 show the combined plots of $I_{ds}/\log(I_{ds})$ and $g_m/ (g_m/I_{ds})$ vs. V_{gs} at low substrate bias for the longest /widest-channel (10/10 μm), shortest/narrowest-channel (0.13/0.13 μm) and medium W/L -channel (0.3/0.3 μm) devices with (a) $V_{ds} = 0.1\text{V}$ and (b) $V_{ds} = 1.2\text{V}$. Higher-order derivatives of the drain current, g_m are calculated numerically. Symbols represent measurement data and lines indicate the simulated results from the I_{ds} model.

(a)

(b)

Figure 8.9: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} from the I_{ds} model (lines), compared to the measurement data (symbols) (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 10/10$, $V_{bs} = 0$.

(a)

(b)

Figure 8.10: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} from the I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 10/0.13$, $V_{bs} = 0$.

(a)

(b)

Figure 8.11: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} from the I_{ds} model (lines), compared to the measurement data (symbols) at (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 0.13/0.13$, $V_{bs} = 0$.

(a)

(b)

Figure 8.12: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 10/0.3$, $V_{bs} = 0$.

(a)

(b)

Figure 8.13: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 0.3/10$, $V_{bs} = 0$.

(a)

(b)

Figure 8.14: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{ds} = 0.1$ V and (b) $V_{ds} = 1.32$ V, for $W/L = 0.3/0.13$, $V_{bs} = 0$.

(a)

(b)

Figure 8.15: $I_{ds}/\log(I_{ds})$ on the left axis and $g_m/(g_m/I_{ds})$ on the right axis vs. V_{gs} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{ds} = 0.1V$ and (b) $V_{ds} = 1.32V$, for $W/L = 0.3/0.3$, $V_{bs} = 0$.

In the following plots, figures 8.16 – 8.21 show the combined plot of $I_{ds}/\log(g_{ds})$ at the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} at high gate ($V_{gs} = 1.2V$) bias for the longest /widest-channel (10/10 μm), shortest/narrowest-channel (0.13/0.13 μm) and medium W/L -channel (0.6/0.5 μm) with (a) $V_{bs} = 0$ and (b) $V_{bs} = -1.2V$. Higher-order derivatives of the drain current, g_{ds} and dg_{ds}/V_{ds} are calculated numerically

(a)

(b)

Figure 8.16: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results

from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{bs} = 0$ and (b)

$V_{bs} = -1.2$ V, for $W/L = 10/10$, $V_{gs} = 1.2$ V.

(a)

(b)

Figure 8.17: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results from I_{ds} model (lines) compared to the measurement data (symbols) at $V_{bs} = 0$ and (b) $V_{bs} = -1.2$ V, for $W/L = 10/0.13$, $V_{gs} = 1.2$ V.

(a)

(b)

Figure 8.18: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{bs} = 0$ and (b)

$V_{bs} = -1.2$ V, for $W/L = 0.13/10$, $V_{gs} = 1.2$ V.

(a)

(b)

Figure 8.19: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{bs} = 0$ and (b)

$V_{bs} = -1.2V$, for $W/L = 0.13/0.13$, $V_{gs} = 1.2V$.

(a)

(b)

Figure 8.20: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{bs} = 0$ and (b)

$V_{bs} = -1.2V$, for $W/L = 10/0.3$, $V_{gs} = 1.2V$.

(a)

(b)

Figure 8.21: $I_{ds}/\log(g_{ds})$ on the left axis and $(dg_{ds}/dV_{ds})/r_{out}$ on the right axis vs. V_{ds} results from I_{ds} model (lines) compared to the measurement data (symbols) (a) $V_{bs} = 0$ and (b)

$V_{bs} = -1.2V$, for $W/L = 0.3/0.3$, $V_{gs} = 1.2V$.

Our unified I_{ds} model has been validated with 0.11- μm technology measurement data as shown in the above figures. Our proposed I_{ds} model can predict the drain current and conductance very well for geometry/bias dependence with a set of model parameter extracted from a 10-step calibration procedure with 6 $I - V$ measurement and points data. Data for other intermediate channel length/wide are shown in Appendices for various

condition as presented in Chapter 3, the existing modeling approach cannot model “ C_{gg} recovery” due to poly-inversion effect and poly-accumulation effect [12]. This is because the complexity in the coupled implicit equations at gate/oxide and oxide/channel interfaces, which cannot be easily solved with the iterative approach. Our modeling approach solves both effects regionally and explicitly, which is simpler and easier to account for both effects in our AC model. As have been shown in [5] that our modeling approach can be easily extended to Strain-Si MOSFET, which might be difficult for existing modeling approaches without re-construction of the model.

The key features of our DC model are on the symmetry and continuity for “pinned” surface-potential-based (threshold-voltage-based) approach. The asymmetric S-ref threshold-voltage-based model has been addressed and a method is proposed with B-ref drain-current formulation. One of the keys to model symmetry is illustrated [13] in chapter 6 with B-ref mobility formulation. Assumption made in the existing mobility expression, which has been widely used in the existing modeling approach is discussed. With our proposed mobility expression, our DC model not only gives symmetric solutions but also shows correct output conductance behavior, which has been neglected in the exiting modeling approach [86].

Our model parameters are characterized to a given technology rather than “binning” to all the measurement data, which is very different compared to the existing modeling approaches. Expansion in empirical geometry-dependent parameters (such as “binning”), as normally adopted in existing modeling approaches, is not suitable for our modeling methodology as it makes the model more and more unmanageable. Our

proposed calibration approach is simpler with a step-by-step procedure and the measurement data requirement is minimum as discussed in the previous chapter.

CHAPTER 9: CONCLUSION, ENHANCEMENT AND RECOMMENDATION

This chapter presents conclusions on the AC/DC modeling methodology in this work. The enhancement of this work and recommendation for future development are discussed.

9.1 Conclusions and model enhancement

A unified single-piece intrinsic terminal charge model (AC model) based on “non-pinned” surface-potential modeling methodology has been developed. Analytic regional AC model has been formulated from explicit regional “non-pinned” ϕ_s -solutions by linking the potential balance to charge-neutrality equations and solving Poisson’s equation. Analytic regional AC model has been extended beyond its validity region through smoothing functions in regional surface-potential expressions, leading to a unified AC model valid in all regions.

Poly-doping effect has been studied with the potential drop, ϕ_p , across the induced space-charge layer at the gate interface. ϕ_p potential can be accurately formulated by following the same approach as for the channel potential, ϕ_s in all regions of operation. Poly-accumulation/inversion effect, which has been neglected in the literature, is presented in this work. Our unified charge model is extended to accumulation region and even beyond normal operating region (when the device is biased beyond recovery voltage) with significant impact from low gate doping concentration results significant ϕ_p potential across an induced layer at gate/oxide interface.

Quantum-mechanical effect has been studied with the van Dort's model, which is formulated via the increase of the energy bandgap that is derived through triangular potential approximation with the self-consistent solution to the coupled Schrödinger and Poisson equations. Our unified AC model expression with van Dort's QME model has been proposed with significant change of the surface-potentials, ϕ_s/ϕ_p due to the modification of intrinsic carrier concentration to account for the effect of free carriers' quantization level. QME has been incorporated without any modification in our AC model derivation, and extended with smooth transition across all regions.

Both the poly-doping and quantum-mechanical effects are coupled in our AC model. They are calculated explicitly through ϕ_s/ϕ_p solutions. Model parameters for our AC model are extracted in a *step-by-step* sequence. Polycrystalline silicon and quantum mechanical effects have been demonstrated, and can be extracted separately in our model parameter extraction methodology.

A set of model parameters has been extracted for our AC model to validate the model with a set of numerical and limited set of measurement data. Our AC model fits well to the numerical and measurement data, and predicts excellently the poly-doping and quantum-mechanical effects.

Unified single-piece drain-current model (DC model) based on bulk-referenced "pinned" ϕ_s -based (threshold-voltage based) modeling methodology has been developed. The model has an extremely compact form. It resembles simple textbook ideal long/wide-channel equations with major geometry effects included, such as short-channel and narrow-width effects. Threshold voltage is a sensitive parameter in the drain-current expression. It is physically formulated with geometry/bias effects that take into account

for the threshold voltage roll-up (reverse SCE and inverse NWE) and roll-off (SCEs and NWEs) at decreasing channel length/width as well as long/short-channel DIBL and non-uniform channel doping.

A unified single-piece drain-current model has been formulated by proper handling on the mobility degradation effect due to transverse/lateral effective field. The model preserves model symmetry and removes singularity in Gummel symmetric test. Geometry effects in drain-current model, which includes series source/drain resistance, channel-length modulation effect, dog-bone layout effect and Coulombic scattering effect for narrow-width devices, have been presented.

Model parameter extraction methodology for both the threshold-voltage and drain-current expressions has been illustrated with minimum measurement data requirement. Firstly, our threshold voltage parameters are extracted based on technology characterization (“binning” is not an option) at decreasing channel length and width. Geometry effects are added in a *step-by-step* sequence from a 10-step extraction procedure. Drain-current model parameters are extracted in modular sequence following a 10-step extraction procedure at corner geometry and bias conditions in which the geometry effects are most pronounced. The model has been verified and predicts excellently to the measurement data based on the requirement of 6 $I - V$ measurement and points V_t and I_{on} data for our model extraction algorithm. Such minimum measurement data requirement for model calibration can be a challenge for other modeling approaches.

In summary, the thesis has presented a physics-based unified “non-pinned” ϕ_s terminal charge model and a “pinned” ϕ_s -based drain-current model. Our unified

regional approach (combining threshold-voltage/surface-potential/charge-based approaches) is an attempt to build the core model for achieving the ultimate goal with potential for easy extension to model non-idealities and future generation of device structures.

9.2 Recommendations for future research

Based on this work in modeling methodology, the following recommendations are suggested for future development:

- 1.) Temperature is an important parameter but has not been studied in detail. Temperature variation will affect the AC/DC electrical characteristics due to the impact of temperature changes on the intrinsic carrier concentration, mobility and surface potential. With a temperature scalable compact model, circuit simulator will be able to predict the performance of a working MOSFET at difference environment.
- 2.) The developed AC model is applicable only for long-channel device with coupled poly-doping and quantum-mechanical effects. Geometry effects should be considered in future models. Extrinsic capacitances such as overlap and fringing capacitances are also important for short/narrow devices.
- 3.) Our DC model is based on “pinned” ϕ_s -based (threshold voltage based), in which the threshold voltage is derived based on “pinned” ϕ_s -solutions that physically links the input geometry/bias dependence to the output drain-current electrical characteristics. Less effort is required for geometry/bias dependent drain-current modeling. A “non-pinned” ϕ_s -based threshold voltage can be another modeling

- methodology to model DC characteristics. Surface potential above threshold condition is not “pinned” to ϕ_{s0} but with $\Delta\phi_s$ increment as defined in our AC modeling. All major physical effects formulated in our AC model can be consistently used in “non-pinned” ϕ_s -based DC model.
- 4.) Conventional bulk-Si based MOSFETs are approaching their limits as the technology scales into the nanometer regime and the demand for new materials and technologies is increasing. We recommend for extending our modeling methodology presented in this thesis to non-ideal and non-bulk-Si based structures such as strained silicon, SiGe, double-gate FET, and ultra-thin body SOI, which may have many challenges and opportunities. The proposed AC model has been derived as the bulk-Si based core model with poly-doping and quantum-mechanical effects, which has the potential for extension to the non-ideal, non-classical CMOS devices.

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Appendix A

$I_{ds}-V_{gs}$ at $V_{ds} = 0.1V$ curves (Symbols: expt.; Lines: model)

Appendix A

$I_{ds}-V_{gs}$ at $V_{ds} = 1.32$ V curves (Symbols: expt.; Lines: model)

Appendix A

$I_{ds}-V_{ds}$ at $V_{bs} = 0$ and -1.2V curves (Symbols: expt.; Lines: model)

Appendix A

Appendix A

g_m - V_{gs} at $V_{ds} = 0.1$ V and 1.32 V curves (Symbols: expt.; Lines: model)

Appendix A

Appendix A

g_{ds} – V_{gs} at $V_{bs} = 0$ and -1.2 V curves (Symbols: expt.; Lines: model)

Appendix A

Appendix A