Scalable compact modeling for nanometer CMOS technology

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SCALABLE COMPACT MODELING FOR NANOMETER CMOS TECHNOLOGY

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Abstract

This thesis documents the compact model developed for bulk MOSFET and double-gate MOSFET. The unified regional modeling approach is used in the physics-based scalable model development for bulk and double-gate MOSFETs. Surface potential models are developed regionally in accumulation, weak accumulation, depletion, volume and strong inversion regions, which are subsequently combined using interpolation functions to ensure smooth higher order derivatives. New unified regional-based short-channel effects are developed to improve the physical scalability of the charge model. A new concept that defines two separate saturation voltages at source and drain, referenced to bulk (or ground for double-gate), respectively, is introduced to meet the Gummel symmetry requirement and to allow possible extension to asymmetric source/drain devices within the same core model. A novel approach to unifying compact models for different non-classical MOS structures, such as ultra-thin body SOI and symmetric/asymmetric double-gate MOSFETs, is proposed. Explicit surface and zero-field potentials for common-gate asymmetric double-gate MOSFETs are solved regionally and the unified solutions are applied in the explicit drain-current model for double-gate MOSFETs. Explicit surface potentials for double-gate MOSFET with quantum mechanical correction are also developed. The research demonstrates a closer step towards the unification of MOS models for future generation non-classical MOS devices.

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List of Symbols

The following is list of symbols used in the text.

Symbol	Description (unit)
$\overline{A_b}$	Average body factor term (N.A).
$\overline{A_b}'$	Average body factor term with asymmetric series resistance (N.A).
$\overline{A_{b1}}$	Average body charge term at gate 1 (N.A).
$\overline{g_{vo}}$	Average channel length modulation term (N.A).
$\overline{q_i}$	Normalized average inversion charge (C/cm^2).
$\overline{q_i}'$	Average inversion charge term with asymmetric series resistance (C/cm^2).
$\overline{q_{i1}}$	Average normalized inversion charge at gate 1 (C/cm^2).
$\overline{h}$	Planck constant (J.s).
$\overline{\mu_{eff0}}$	Average effective mobility with lateral field degradation ($cm^2/V.s$).
$\overline{\phi_s}$	Average surface potential (V).
$\overline{\phi_s}'$	Average surface potential with asymmetric series resistance (V).
$\overline{\phi_{s1}}$	Average surface potential at gate 1 (V).
$\overline{\phi_{01}}$	Average zero field potential at gate 1(V).
ϑ_r	Reverse interpolation function at V_{FB} (V).
ϑ_f	Forward interpolation function at V_{FB} (V).
$\vartheta_{eff,cc}, \vartheta_{eff,ss},$ $\vartheta_{eff,\delta s}, \vartheta_{eff,ax}$	Smoothing functions (V).
$A_{b,d}$	Drain side, body factor term (N.A).
$A_{b,s}$	Source side, body factor term (N.A).
C_{bg}	Bulk-gate capacitance (F/cm^2).
C_{bottom}	Bottom capacitance (F/cm).
C_{dg}	Drain-gate capacitance (F/cm^2).
C_{ij}	i-j Capacitance, both i and j can be any of the drain, gate, source and bulk terminal (F/cm^2).
C_{ox}	Oxide capacitance (F/cm^2).
C_{ox1}	Oxide capacitance of gate 1 (F/cm^2).
C_{sg}	Source-gate capacitance (F/cm^2).
C_{side}	Side capacitance (F/cm).
E_0	Zero-field in the Si-body, where $\frac{d\phi}{dx} = 0$ (V/cm).
F_s	Electric field in quantum mechanical model (V/cm),

g_m	Transconductance (A/V)
g_d	Output conductance (A/V)
$g_{vo,d}$	Channel length modulation gain factor term at drain side (N.A).
$g_{vo,s}$	Channel length modulation gain factor term at source side (N.A).
H	Intermediate variable for charge model (V).
H'	Intermediate variable for charge model with series resistance (V).
High- K	High relative dielectric constant (NA).
I_{ds}	Drain-source current (V)
I_{ds01}	Long channel drain current at gate 1 (A).
I_{dsat}	Saturation current (A).
L_{eff}	Effective channel length (μm).
L_g	Gate length (μm).
L_{ov}	Overlap length (μm).
L_β	Characteristic length for Gaussian profile (μm).
l_μ	Pile-up centroid (μm).
M^*	Electron/Hole effective mass (g).
N	Free electrons concentration (cm^{-3}).
N^+	Highly concentrated n-type doping region (cm^{-3}).
N_A	Acceptor concentration (cm^{-3}).
N_{ch}	Channel doping concentration (cm^{-3}).
N_D	Donor concentration (cm^{-3}).
N_{sd}	Drain/Source side n^+ doping concentration (cm^{-3}).
$N_{D,d}$	Drain side n^+ doping concentration (cm^{-3}).
$N_{D,s}$	Source side n^+ doping concentration (cm^{-3}).
N_{eff}	Effective doping concentration (cm^{-3}).
N_{gate}	Gate doping concentration (cm^{-3}).
n_i	intrinsic carrier concentration (cm^{-3}).
N_{ov}	Doping concentration of the overlap region (cm^{-3}).
N_{pile}	Gaussian's peak doping concentration (cm^{-3}).
P	Free holes concentration (cm^{-3}).
P^+	Highly concentrated p-type doping region (cm^{-3}).
β^m	Fitting parameter for quantum mechanical effect at low gate doping concentration (N.A).
Q	Magnitude of electronic charge (C).
Q_b	Bulk-charge per unit area (C/cm^2).
Q_B	Bulk charge (C).
$Q_{b,acc}$	Bulk charge accumulation region (C/cm^2).
$Q_{B,Fr}$	Bulk fringing charge (C).
$Q_{B,p}$	Bulk charge with poly gate effect (C).
$Q_{B,p}^{qm}$	Bulk charge with poly gate effect and quantum mechanical effect (C).
$Q_{b,sub}$	Bulk charge depletion region (C/cm^2).
Q_{b1}	Bulk charge at gate 1 (C/cm^2).
Q_{BSC}	Short channel bulk charge, includes poly effect, quantum mechanical effect and extrinsic charges (C).
Q_D	Drain charge (C).
$Q_{D,Fr}$	Drain fringing charge (C).

$Q_{D,OV}$	Drain side overlap charge (C).
$Q_{D,p}$	Drain charge with poly gate effect (C).
$Q_{D,p}^{qm}$	Drain charge with poly gate effect and quantum mechanical effect (C).
Q_{DSC}	Short channel drain charge, includes poly effect, quantum mechanical effect and extrinsic charges (C).
Q_{GSC}	Short channel gate charge, includes poly effect, quantum mechanical effect and extrinsic charges (C).
Q_i	Inversion-charge per unit area (C/cm^2).
q_i'	Inversion charge with series resistance (C/cm^2).
Q_{i1}	Inversion charge at gate 1 (C).
q_{i1}	Normalized inversion charge at gate 1 (C/cm^2).
Q_{ox}	Effective fix oxide charge (C/cm^2).
Q_s	Source charge (C).
$Q_{s,Fr}$	Source fringing charge (C).
$Q_{s,ov}$	Bulk charge in the overlap at source side (C).
$Q_{s,OV}$	Source side overlap charge (C).
$Q_{s,p}$	Source charge with poly gate effect (C).
$Q_{s,p}^{qm}$	Source charge with poly gate effect and quantum mechanical effect (C).
Q_{sc}	Induced channel charge (C/cm^2).
Q_{sc}^{qm}	Quantized channel induce charge (C).
$Q_{Sov,acc}$	Accumulation bulk charge in the overlap at source side (C/cm^2).
$Q_{Sov,sub}$	Depletion bulk charge in the overlap at source side (C/cm^2).
Q_{SSC}	Short channel source charge, includes poly effect, quantum mechanical effect and extrinsic charges (C).
r_1	Fitting parameter for voltage independent series resistance (Ω).
r_2	Fitting parameter for voltage dependent series resistance ($V.s/cm$).
R_d	Series resistance at drain side (Ω).
R_s	Series resistance at source side (Ω).
SiO_2	Silicon oxide (N.A).
t_{ox}	Oxide thickness (cm).
t_p	Poly gate thickness (cm).
t_{si}	Silicon body thickness (cm).
U	Averaged shifted potential (V).
V	Ground referenced channel voltage referenced to ground (V).
V_0	Carrier velocity (cm/s).
$V_{Aeff,d}$	Effective Early voltage at drain side (V).
$V_{Aeff,s}$	Effective Early voltage at source side (V).
v_b	Fitting parameter for bulk fringing charge (NA).
V_{b0}	Low bulk bias, $V_{bs}=0$ (V).
V_{bb}	High bulk bias, $V_{bs}=-1.2$ (V).
V_{bi}	Built in potential (V).
V_{cb}	Channel to bulk voltage (V).
V_d	Ground referenced drain voltage (V).
$V_{d,sat}$	Ground referenced drain saturated voltage (V).
V_{d0}	Low drain bulk bias, $V_{db}=0.05$ (V).

V_{db}	Drain bulk voltage (V).
V_{db}	Drain to bulk voltage (V).
$V_{db,eff}$	Effective drain-bulk voltage (V).
$V_{db,sat}$	Drain-bulk saturation voltage (V).
V_{dd}	High drain bulk bias, $V_{db}=1.2V$.
V_{ds}	Drain-source voltage (V).
$V_{ds,eff}$	Effective drain-source voltage (V).
$V_{ds,sat}$	Drain-source saturation voltage (V).
V_{FB}	Flat-band voltage (V).
$V_{FB,ov}$	Overlap region flat-band voltage (V).
V_{g1FD}	Full depletion voltage (V).
V_{gb}	Gate-bulk voltage (V).
V_{GBA}	Forward interpolated $V_{gb}-V_{FB}$ function with smoothing parameter σ_a , (V).
V_{GBF}	Forward interpolated $V_{gb}-V_{FB}$ function (V).
$V_{gf,V}$	Ground referenced forward interpolated V_g-V_{FB} function (V).
V_{gfl}	Flat-band shifted gate source voltage 1 (V).
V_{gff}	Forward interpolated $V_{gs}-V_{FB}$ function for double gate (V).
V_{gfr}	Reverse interpolated $V_{gs}-V_{FB}$ function for double gate (V).
V_{gg}	High gate to bulk voltage, $V_{gb}=1.2$ (V).
$V_{gr,V}$	Ground referenced reverse interpolated V_g-V_{FB} function (V).
V_{gsf}	Forward interpolated $V_{gs}-V_{FB,ov}$ (V).
V_{gsr}	Reverse interpolated $V_{gs}-V_{FB,ov}$ (V).
$V_{gt,d}$	Drain side, inversion charge normalized to the oxide capacitance (V).
$V_{gt,d}$	Normalized inversion charge for double gate at drain side (V).
$V_{gt,s}$	Normalized inversion charge for double gate at source side (V).
v_i	Fitting parameter for inversion fringing charge (N.A).
V_{ox}	Voltage drop across oxide thickness (V)
V_r	Minimum voltage of source and drain side (V).
V_s	Ground referenced source voltage (V).
$V_{s,sat}$	Ground referenced source saturated voltage (V).
v_{sat}	Carrier saturation velocity (cm/s).
V_{sb}	Source bulk voltage (V).
$V_{sb,eff}$	Effective source-bulk voltage (V).
$V_{sb,sat}$	Source-bulk saturation voltage (V).
$V_{sd,sat}$	Source side saturation voltage (V).
V_t	Threshold-voltage (V).
v_{th}	Thermal voltage (V).
W	Effective width of the transistor (μm).
x_{01}	Zero field location from gate 1 (cm).
x_{d1}	Induced depletion depth from gate 1 (cm).
x_{d2}	Induced depletion depth from gate 2 (cm).
$x_{dmax,d}$	Maximum depletion width at drain side (cm).
$x_{dmax,s}$	Maximum depletion width at source side (cm).
$X_{dov,s}$	Depletion depth of overlap at source side (cm).
$X_{i,d}$	Junction depth at drain side (cm).

$X_{j,s}$	Junction depth at source side (cm).
ΔE_g^{qm}	Energy bandgap difference due to carrier quantization effect (V).
ΔR	Series resistance difference (Ω).
$\Delta\phi_{DS}$	Unified regional potential difference, depletion to strong inversion (V).
$\Delta\phi_{ds}$	Unified regional potential difference, depletion to strong inversion with poly-si gate effect(V).
$\Delta\phi_s$	Surface potential difference (V).
$\Delta\phi_{s1}$	Surface potential difference of gate 1 (V).
E_{eff}	Effective transverse field (V/cm).
α	Fitting parameter for quasi-2D solution (cm^2).
α_{acc}	Fitting parameter for quasi-2D solution at accumulation (cm^2).
α_{dibl}	Fitting parameter for long channel drain induced barrier lowering (NA).
α_{ds}	Fitting parameter for quasi-2D solution at depletion to strong inversion (cm^2).
β	Gain factor for drain current (AV^{-2}).
β^{qm}	Quantum poly-gate gain factor (N.A.).
δ_L	Fitting parameter for lateral field degradation (N.A).
δ_s	Smoothing parameter for effective voltage from linear to saturation (V).
δ_{s1}	Length dependent smoothing parameter coefficients 1 for smoothing parameter for effective voltage from linear to saturation (V).
δ_{s2}	Length dependent smoothing parameter coefficients 2 for smoothing parameter for effective voltage from linear to saturation (V).
δ_{s3}	Length dependent smoothing parameter coefficients 3 for smoothing parameter for effective voltage from linear to saturation (μm).
δ_{s4}	Length dependent smoothing parameter coefficients 4 for smoothing parameter for effective voltage from linear to saturation (N.A).
δ_ϕ	Smoothing parameter for surface potential transition from depletion to strong inversion (V).
$\delta_{\phi,cc}$	Smoothing parameter to join weak accumulation to strong accumulation (V).
$\delta_{\phi,ss}$	Smoothing parameter to join depletion to strong inversion (V).
$\mathcal{E}^{qm}$	Fitting parameter to adjust flat-band shift due to quantum mechanical effect (V).
ϵ_{si}	Dielectric permittivity of silicon (F/cm)
ϕ_{0I}	Zero field potential from gate 1 (V)
ϕ_{0acc1}	Unified regional zero field accumulation potential solution (V).
ϕ_{0eff1}	Full piece unified regional zero field potential solution (V).
ϕ_{0str1}	Unified regional zero field strong inversion potential solution (V).
ϕ_{ACC}	Unified regional accumulation surface potential (V).
ϕ_{acc}	Unified regional surface potential at accumulation with poly gate

	effect (V).
$\phi_{ACC,Q2D}$	Unified regional surface potential at accumulation considering quasi two dimensional effects (V).
ϕ_{ACCI}	Unified regional accumulation surface potential solution of double gate at gate 1 (V).
ϕ_{ADD}	Regional surface potential at weak accumulation (V).
ϕ_{ADD1}	Weak accumulation surface potential solution of gate 1 (V).
ϕ_{ASUB}	Unified regional surface potential at weak accumulation (V).
ϕ_{CC}	Regional surface potential at strong accumulation (V).
ϕ_{CCI}	Regional accumulation surface potential solution of double gate at gate 1 (V).
ϕ_{DD}	Regional surface potential at depletion (V).
ϕ_{dd1}	Depletion surface potential solution of gate 1 (V).
ϕ_{DS}	Unified regional depletion to strong inversion surface potential (V).
$\phi_{DS,D}$	Unified regional depletion to strong inversion surface potential at drain side (V).
ϕ_{ds}	Unified regional surface potential depletion to strong inversion with poly gate effect (V).
$\phi_{ds,Q2D}$	Unified regional surface potential at depletion to strong inversion considering quasi two dimensional effects (V).
$\phi_{DS,S}$	Unified regional depletion to strong inversion surface potential at source side (V).
ϕ_{ds01}	Unified regional depletion to strong inversion surface potential solution of double gate at gate 1 (V).
ϕ_{ds1}	Newton-Raphson improved unified regional depletion to strong inversion surface potential solution of double gate at gate 1 (V).
ϕ_{dv1}	Unified regional depletion to volume inversion surface potential solution of double gate at gate 1 (V).
ϕ_{EFF}	Full-piece unified regional accumulation to strong inversion surface potential (V).
ϕ_{eff}	Unified regional full piece surface potential with poly gate effect (V).
ϕ_{eff1}	Full piece surface potential solution of double gate at gate 1 (V).
ϕ_F	Fermi potential (V).
ϕ_{fd1}	Full depletion surface potential of gate 1 (V).
ϕ_L	Long channel surface potential (V).
ϕ_{MS}	Work function difference between polycrystalline gate and silicon channel (V).
$\phi_{ov,s,acc}$	Accumulation region of surface potential in the overlap (V).
$\phi_{ov,s,sub}$	Depletion region of surface potential in the overlap (V).
ϕ_p	Polycrystalline gate surface potential (V).
ϕ_{acc}	Unified regional gate surface potential at strong accumulation (V).
ϕ_{acc}^{qm}	Unified regional gate surface potential at strong accumulation with quantum mechanical effect (V).
ϕ_{cc}	Regional gate surface potential at strong accumulation (V).
ϕ_{dd}	Regional gate surface potential at depletion (V).

ϕ_{eff}	Unified regional full piece gate surface potential (V).
ϕ_{ss}	Regional gate surface potential at strong inversion (V).
ϕ_{str}^{qm}	Unified regional gate surface potential at strong inversion with quantum mechanical effect (V).
ϕ_{sub}^{qm}	Unified regional gate surface potential at depletion with quantum mechanical effect (V).
ϕ_s	Surface-potential (V).
$\phi_{s,d}$	Surface potential at drain side (V).
$\phi_{s,d}'$	Surface potential with asymmetric series resistance at drain side (V).
$\phi_{s,s}$	Surface potential at source side (V).
$\phi_{s,s}'$	Surface potential with asymmetric series resistance at source side (V).
ϕ_{s1}	Surface potential near gate 1 (V).
$\phi_{s1,d}$	Surface potential at drain side (V).
$\phi_{s1,s}$	Surface potential at source side (V).
ϕ_{s2}	Surface potential near gate 2 (V).
ϕ_{sov}	Overlap region surface potential (V).
ϕ_s^{qm}	Position dependent surface potential considering quantum mechanical effect (V).
ϕ_{ss}	Regional surface potential at strong inversion (V).
ϕ_{ss1}	Regional strong inversion surface potential solution of double gate at gate 1 (V).
$\phi_{str,cc}$	Unified regional strong accumulation surface potential (V).
ϕ_{str}	Unified regional surface potential at strong inversion with poly gate effect (V).
$\phi_{str,ss}$	Unified regional strong inversion surface potential (V).
ϕ_{str1}	Unified regional strong inversion surface potential solution of double gate at gate 1 (V).
ϕ_{SUB}	Unified regional surface potential at depletion (V).
ϕ_{sub}	Unified regional surface potential at depletion with poly gate effect (V).
ϕ_{sub1}	Unified regional depletion surface potential solution of double gate at gate 1 (V).
ϕ_{v1}	Unified regional volume inversion surface potential solution of double gate at gate 1 (V).
ϕ_{vi1}	Volume inversion surface potential of gate 1 (V).
γ	Bulk body factor ($V^{1/2}$).
$\gamma_{eff,d}$	Effective body factor at drain side. ($V^{1/2}$).
$\gamma_{eff,s}$	Effective body factor at source side. ($V^{1/2}$).
γ_{OV}	Body factor term at overlap region ($V^{1/2}$).
γ_{OVBCS}	Body factor term at overlap region with bulk charge sharing effect ($V^{1/2}$).
γ_p	Polycrystalline gate factor ($V^{1/2}$).
γ_{peff}	Polycrystalline gate factor with bulk charge sharing. ($V^{1/2}$).
η	Pre-factor for quasi-2D depletion depth (N.A).
κ	Gain factor of the effective vertical doping concentration (NA).

κ^{qm}	Quantum mechanical effect prefactor (N.A).
λ	Fitting parameter for bulk charge sharing effect at channel (N.A).
λ_{ov}	Fitting parameter for bulk charge sharing effect at overlap region (N.A).
λ_p	Fitting parameter for bulk charge sharing effect at gate (N.A).
μ	Mobility (cm^2/Vs).
μ_1, μ_2, μ_3	Mobility fitting parameters (μ_1 : cm^2/Vs).
V	Mobility fitting surface roughness parameter (N.A).
μ_{co}	Columbic scattering (cm^2/Vs).
μ_{eff0}	Effective mobility with lateral field degradation (cm^2/Vs).
μ_{ph}	Phonon scattering (cm^2/Vs).
μ_{sr}	Surface roughness scattering (cm^2/Vs).
μ_0	Transverse field mobility (cm^2/Vs).
π	Approximately to 3.14159 (N.A).
σ	Junction depth prefactor (N.A).
σ_a	Smoothing parameter for reverse interpolation function to be used in surface potential (V^2).
ρ_p	Smoothing parameter for reverse interpolation function to be used in poly gate surface potential (V^2).
σ_f	Smoothing parameter for forward interpolation function to be used in surface potential (V^2).
ρ_f	Smoothing parameter for forward interpolation function to be used in poly gate surface potential (V^2).
ξ_c	Fitting parameter for channel length modulation (cm^2/V).
ξ_n	Fitting parameter for inversion charge term effective field (N.A).
ξ_b	Fitting parameter for bulk charge term effective field (N.A).

List of Acronyms

The following is list of acronyms used in the text.

Acronyms	Description
AC	Alternative current
a-DG	Asymmetric double-gate
BCS	Bulk charge sharing
BSIM3, BSIM4, BSIM5	Berkeley short-channel insulated-gate field-effect transistor model, version 3, 4 and 5, respectively
Ca-DG	Common (gate) asymmetric double gate
CLM	Channel length modulation
CMC	Compact Model Council
CMOS	Complementary metal oxide semiconductor
DC	Direct current
DIBL	Drain induce barrier lowering
FD	Fully-depleted
FinFET	“Fin”-silicon field-effect-transistor
GST	Gummel symmetry test
HiSIM	Hiroshima-university STARC insulated-gate field-effect transistor model
LHS	Left hand side
MOS9	Philips’ MOS model level 9
MOSFET	Metal-oxide-semiconductor-field-effect transistor
NMOS	N-channel MOSFET
PAE	Poly accumulation effect
PBL	Potential barrier lowering
PD	Partially-depleted
PDE	Poly depletion effect
PIE	Poly inversion effect
Quasi-2D	Quasi-two-dimensional
RF	Radio frequency
RHS	Right hand side
s-DG	Symmetric double-gate
Si	Silicon
SOI	Silicon-on-insulator
TCAD	Technology computer-aided design
URM	Unified regional modeling
URSP	Unified regional surface potential
URZP	Unified regional zero-field potential
UTB	Ultra thin body
VO	Velocity overshoot

w.r.t	With respect to
N.A	Not applicable

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Metal-Oxide-Semiconductor Field Effect Transistor (MOSFET) is one of the key devices in our modern life. We can find it everywhere, such as in mobile phones, televisions, automobiles, computers, etc. The semiconductor experts are always on the go to find ways to shrink the size of MOSFET and to increase the number of transistors per chip. The number of transistors on a chip has doubled about every two years, following the prediction of Moore's Law since 1965. There is more than 1.7 billion transistors in the recent Itanium 2 processor by Intel [1]. Such complexities demand for computer-aided simulations using *compact models* to analyze, design, and optimize an integrated circuit. This work focuses on scalable modeling of an n-channel MOSFET (NMOS) and its extension to next-generation MOSFET, namely, the FinFET.

1.1. Motivations

The industry standard compact model for bulk MOSFETs, namely, BSIM3 [2], has served the industry for more than ten years as the sole standard model [3]. Only until recently, the next-generation standard model, PSP was selected [4]. Both models

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are now supported by the Compact Model Council (CMC) as the industry standard compact models. One of the main reasons the latter model was selected as a standard model is due to a few show-stoppers in the former pertaining to the failure in meeting the Gummel symmetry test (GST) requirement [5] and non-physical behavior in charge/current model [6]. Nevertheless, there is still an inherent problem with the new standard model in meeting the GST requirement at higher-order derivatives, due to the use of smoothing function in the *effective drain-source voltage*, $V_{ds,eff}$ [7]. Another length-dependent fitting parameter was used to correct this problem for higher-order derivatives and for short-channel devices [4]. Therefore, a novel approach of compact modeling that is developed based on the physics without introducing any extra fitting parameters on $V_{ds,eff}$, and maintaining the ability to meet the GST requirement is most desirable.

Most compact models [2, 4] usually ignore the accuracy in the accumulation for the charge/capacitance model. For example, the poly-Si gate accumulation effect is not included in the industry standard core models. This inaccuracy is likely to add to the difficulties in modeling short-channel charge/capacitance in MOSFETs. Therefore, it is desirable to develop a scalable charge/capacitance model that is complete with all the essential physical effects (poly-Si gate effect and quantum-mechanical effect) for both long and short-channel devices. With these secondary effects properly developed, the charge model should be able to fit the long channel capacitance data in all regions and able to scale to short channels physically.

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Conventional bulk MOSFETs are approaching their scaling limit after more than 40 years of continued scaling since 1965. A considerable number of issues have arisen over the years of aggressive scaling, such as gate leakage, quantum effects, poly-Si gate effects, short-channel effects, and mobility degradation due to high channel doping concentration. Various techniques have been explored to overcome these problems. For example, the use of high- K dielectric materials for gate oxide with metal gate electrode [8], mobility enhancement using the strained-silicon (s-Si) channel [9] as well as multiple-gate, fully-depleted (FD) silicon-on-insulator (SOI) MOSFETs [10]. The high- K dielectric material is able to allow the same electrostatic gate control but with a thicker dielectric material compared to silicon oxide (SiO_2), thus reducing the gate leakage. The use of metal electrode is able to reduce the poly-depletion effect as metal is a better conducting material compared to polysilicon. The purpose of s-Si channel is to enhance the electron mobility to achieve higher speed. Fully-depleted body provides a better and more efficient electrostatic control over the channel, and thus improving the short-channel effects. The SOI structure is able to reduce the junction capacitance, hence, improving the electrical isolation (reducing current leakage from other devices) at a relatively lower doping concentration in the channel.

The double-gate (DG) MOSFET is one of the key potential devices that can be used to continue the aggressive scaling of MOSFETs. Before the realization of double-gate MOSFETs as the next-generation device in a large-scale circuit, compact models used in the circuit design need to be developed. Nevertheless, this extension to the next generation MOSFET building blocks may become a test on how much an existing compact model can be extended, but not from the re-construction of model

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formulation [11]. This is due to the fact that the transition from bulk-MOSFET to the variations of double-gate MOSFET has seamless transitions among them as shown in Figure 1.1. A partially-depleted (PD) SOI double-gate MOSFET has a much thicker oxide at its back gate where the induced electric field near the back gate surface can be negligible. If the silicon body is thin enough, then the PD SOI MOSFET may operate in FD mode, as in Figure 1.1(b) for the FD SOI MOSFET. If the thickness of the oxide at the back gate of SOI MOSFET reduced and becomes comparable to that of the front gate, the device is better known in Figure 1.1(c) asymmetric double-gate (a-DG) MOSFET. If both of the gates in (c) are biased at the same voltage and it is also fabricated to have the same oxide material and thickness, it is actually called symmetric double-gate (s-DG) MOSFET as in Figure 1.1(d). In fact, half of the s-DG MOSFET is a bulk-like ultra-thin body (UTB) MOSFET as in Figure 1.1(e). Finally, if the body thickness of the bulk-UTB MOSFET is extended, then it is reverted back to a conventional bulk-MOSFET as in Figure 1.1(f) [11].

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Figure 1.1: Seamless transformation and unification of MOSFETs: (a) partially-depleted (PD) silicon-on-insulator (SOI), (b) fully-depleted (FD) ultra-thin body (UTB) SOI, (c) asymmetric double-gate (a-DG), (d) symmetric double-gate (s-DG), (e) “bulk-UTB”, and (f) bulk [11].

Similarly, for a compact model, the transitions from one type of device to another should ideally be seamless. This is, however, a very challenging task, if not impossible, for model developers to formulate a compact model infrastructure that is simple yet scalable, in terms of terminal voltages, geometry scaling and physical transitions. On top of that, the compact model also needs to be carefully developed with Gummel symmetry in mind, with the lessons learned from BSIM model over the past many years.

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The core model in this work is developed based on the unified regional modeling (URM) approach to include short-channel effects, poly-Si gate effects and quantum-mechanical effects. It will also be extended to double-gate MOSFETs. Ultimately, it will provide a circuit model that is applicable for current and future-generation MOSFET circuit design and simulation.

1.2. Objectives

The objectives of this research are to:

- i. Develop and improve on the direct current/alternating current (DC/AC) bulk-MOSFET compact model so that it is simple, scalable and fulfilling most of the industry standards, especially the GST requirement.
- ii. Extend the AC bulk-MOSFET model to include short-channel, quantum-mechanical and poly-Si gate effects based on the URM approach.
- iii. Demonstrate the scalability of URM approach through compact modeling of FinFETs (double-gate MOSFETs), including unified regional surface-potential solutions and meeting the GST requirement.

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1.3. Major contributions of the thesis

The major contributions of this thesis are listed below:

- i. Proposed a new unified regional surface potential (URSP) solution for strong accumulation, weak accumulation and strong inversion. The new solutions are physical yet simpler compared to the Lambert- W function in accumulation with reduced total number of smoothing parameter in strong inversion [12].
- ii. Changed the DC/AC model to be based on URSP solutions instead of threshold-voltage based non-pinned surface potential modeling approach [7].
- iii. Proposed new formulations for short-channel effects with URM approach, namely, potential-barrier lowering and bulk-charge sharing [7].
- iv. Developed a new approach in defining two saturation voltages, at the source and drain ends, respectively, referring to the bulk-terminal to improve the Gummel symmetry behaviour in the effective drain-source voltage, $V_{ds,eff}$. The new model allows the evaluation of negative drain-source voltage, V_{ds} without swapping the drain-source terminal when $V_{ds} < 0$ as is done conventionally. It is now possible to handle asymmetric source/drain structure in the same core MOSFET model [7].
- v. Proposed a new asymmetric source/drain series resistance drain-current model that is based on the URSP approach.
- vi. Extended the AC charge model to include extrinsic charge model such as the overlap and fringing capacitances that are formulated in the charge model [13].

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- vii. Derived full-depletion voltage for common-gate asymmetric double-gate structure [12], which allows seamless modeling of asymmetric/symmetric double-gate MOSFETs.
- viii. Introduced the URSP solution that includes volume inversion for double-gate, which subsequently can be unified with the strong inversion to form a single-piece surface-potential solution [14]. The surface-potential solutions are tested for different doping, body thickness and oxide thickness.
- ix. Proposed regional solutions for zero-field potential [12] in symmetric-DG, which is an explicit function of the surface potential.
- x. Extended the URSP model for common-asymmetry double-gate in drain-current models. Accurate surface-potential solutions are not required in the model evaluation [12].
- xi. Proposed a new explicit model for quantum-mechanical effect in the common a-DG surface-potential model [15].

1.4. Organization of the thesis

This thesis is written in five chapters. Chapter one introduced the motivations, objectives and major contributions of this work.

Chapter two discusses the background of compact modeling methodologies that currently exist in DC/AC bulk-MOSFETs compact models. The URM approach

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for bulk-MOSFET is discussed in details. The compact model is physically derived based on the regional Poisson's equations, which are then unified with interpolation functions for smooth higher order derivatives. The drain-current and charge models are then derived based on bulk-terminal referenced equations. This leads to a core model without requiring terminal-voltage swapping for negative drain-source voltage; thus, allowing asymmetric source/drain modeling within the same core model. The model is able to exhibit perfect Gummel symmetry for all higher order derivatives in an ideal source/drain symmetry condition. The model is also extended to include the short-channel, poly-Si gate and quantum-mechanical effects explicitly based on the URM approach.

Chapter three focuses on the extension of the URM approach to double-gate MOSFETs modeling, namely, FinFETs. The regional surface-potential solutions are derived based on similar ideas to bulk-MOSFET modeling, namely, accumulation, weak accumulation, depletion, volume and strong inversion. Based on full-depletion assumption, a novel approach for seamless modeling of asymmetric/symmetric DG has been proposed. The zero-field potential solutions are derived based on explicit functions of the URSP from the regional Poisson equations. Both URSP and unified regional zero-field potential (URZP) are able to scale physically for most of the practical operating conditions.

Chapter four demonstrates the DC/AC model parameter extraction procedure for bulk MOSFETs. A *step-by-step* model parameter extraction is illustrated. The compact model is verified using Medici numerical simulations from a 90-nm CMOS technology. The measurement data of long-channel capacitance of 0.13- μm CMOS

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technology has also been used to verify the extraction procedure. Similar procedure is applicable to FinFET compact model as both models are developed based on the same URM approach.

Finally, chapter five summarizes and concludes the AC/DC MOSFET modeling methodology with suggestions and recommendations for future investigations.

CHAPTER 2: Bulk MOSFET Model With URSP

CHAPTER 2: Bulk MOSFET Model with Unified Regional Surface Potential

Device modeling has become tremendously complex nowadays. A check on one of the industry standard models for bulk MOSFETs, BSIM4.6.1's user manual [16], reveals that it has 78 basic model parameters (not counting other parameters for secondary effects such as asymmetric source/drain junction diode, asymmetric bias-dependent series resistance, gate-induced drain leakage, gate dielectric tunnelling, charge and capacitance, high-speed/RF model, noise models, etc). The list is also likely to increase as the technology continues to advance. Device model development requires physical understanding on the device layout, fabrication, and operation to properly develop a compact model that is able to match real MOSFET electrical behaviours in the most advanced technology. This chapter will discuss the development of a scalable model based on unified regional modeling approach for bulk MOSFETs with the help of numerical simulations using Medici for model validation.

CHAPTER 2: Bulk MOSFET Model With URSP

2.1. Background of bulk MOSFET compact modeling

The MOSFET has been the key electronic device for the advancement of the semiconductor industry. The number of transistors on a single chip has grown tremendously over the years according to Moore's law. With more than 1.7 billion transistors in the recent Itanium 2 processor by Intel [1], simple analytical (mathematical) descriptions of the electrical behaviour of the MOSFET in the form of compact models are needed for circuit simulation and design.

A MOSFET compact model is built based on the physics that governs the carrier transport in a three-dimensional (3D) structure [17]. The MOSFET transport differential equations are well described and can be solved by technology computer-aided design (TCAD) simulators [18]. These simulators, however, are not suitable for circuit design because it is too computational demanding to run simulations of many transistors in a circuit by solving the differential equations numerically. Nevertheless, numerical simulation tools provide very important insights for verification in the development of compact models. Therefore, one of the tools, Medici [18], will be used extensively in this work for model validation.

The bulk MOSFET compact model that is derived based on the Poisson equation and Gauss' law has been established by Pao and Sah in the mid 1960s [19]. The drain current is described as a double integral, in which the solution has to be obtained by numerical integration. Again, it is not suitable for use in circuit simulation.

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Therefore, it has to be “compacted” (i.e., integrated out by hand) for the closed-form analytical solution. With charge-sheet approximation, the double-integral equation for the drain current can be expressed as an explicit function of the surface potential [20] without losing much accuracy (1~2% error [21]). The charge-sheet drain current model, however, still needs the surface potential solutions, which can be solved iteratively at the drain and source ends of the channel using the input voltage equation according to Pao and Sah [19].

Recently, it was pointed out that the earlier surface-potential solution had caused a numerical error near the flat-band voltage, V_{FB} [22]. In response to the error, Sah has corrected the solution in [23], which used the correct remote boundary condition for the minority carrier concentration. It becomes the reference for comparison of surface-potential solutions for all compact models. Based on this (Pao and Sah) voltage equation, three different approaches to compact solutions have been developed over the years [23], namely, the threshold-voltage (V_t) based, surface-potential (ϕ_s) based, and inversion-charge (Q_i) based. The classification of these different approaches is based on the intermediate variable (i.e., V_t , ϕ_s , or Q_i) used to derive the charge and current models.

The first generation bulk MOSFET compact models were all V_t -based. V_t -based compact models have been popular due to simplicity of equations relating the input gate voltage and surface potential. The surface potential is assumed a linear function of input gate-bulk voltage, V_{gb} when it is below V_t and constant above V_t . This is also largely referred to as the “pinned” surface potential model, where the

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surface potential is pinned to twice the Fermi potential, $2\phi_F$ in strong inversion ($V_{gb} > V_t$). The drain current model is obtained with a Taylor expansion at the source-end from the charge-sheet drain current model, resulting in a very simple form with all the current and charges to be expressed as a function of V_t . This model has been implemented in the BSIM family of models and MOS9. BSIM3v3 was subsequently chosen as the industry standard compact model in 1995 [3]. Nevertheless, this standard model suffers from the Gummel symmetry problems [5] due to its source-referenced simplification as well as discontinuities and accuracy problems in moderate inversion. These non-physical effects have led to the need for the Compact Model Council to search for a next generation MOSFET model, which must be inherently symmetric, continuous and physically correct for charges and currents [6].

The ϕ_s solution can either be obtained iteratively (such as in HiSIM) [24] or explicitly (such as in SP) [17], which later was known as PSP [23] after it merged with Philip's MOS 11 ϕ_s -based model [25]. The terminal currents and charges are formulated based on the solution of the surface potential. The ϕ_s -based approach was not well accepted in the early days. The iterative approach was relatively slow at that time and non-iterative approach cannot extend to accumulation region. However, with the newly developed iterative algorithms, the execution time for ϕ_s -based models has improved tremendously [24]. It is no longer considered as impractical and becoming more popular instead due to its strong physical contents [23]. Nevertheless, high accuracy of the surface-potential solution is required in the ϕ_s -based model so that it is able to produce proper subthreshold (or weak inversion region) behaviour [26], correct trans-capacitance and, subsequently, to achieve stable circuit simulation [23].

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The third approach in compact modeling is Q_i -based models. From the voltage equation, it is to first approximate the surface potential with a linear function of inversion charge density [27] in the inversion region. The terminal charges and currents are then formulated based on the solution of the charges. Examples of such models are the EKV model [27] and BSIM5 model [23]. The key feature of the Q_i -based model is that it is able to correctly handle the weak inversion region [23], which is important for analog applications. This type of model is well accepted by analog designers.

All of these different approaches to compact modeling have their strengths and shortcomings in terms of electrical characteristics, predictability, physical parameter variations, geometry and bias dependence, and model parameter extraction. Some adopt the option of “binning” [28] in model calibration to model different segments (gate length/width) of transistors in a given technology. Others employ empirical geometry dependence [4]. Nevertheless, the ϕ_s -based solution is favoured by the industry due to its strong physical contents.

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2.2. Introduction to the unified regional modeling approach

The core model of this work takes the unified regional modeling (URM) approach. It is based on the works of two previous PhD students [29, 30] since 2001. Initially, it was developed based on a V_t -based core model but has evolved and improved over the years. Nevertheless, it retained some important features such as the parameter extraction procedure.

The extraction procedure for this core model is “technology based” rather than “transistor based” [31]. In technology-based characterization, a set of transistors of different lengths/widths in a given technology node are modeled and calibrated, as compared to “golden die” (or “best-transistor”) approach in the transistor-based characterization. It provides better physical geometry scaling for a given technology.

The URM builds in more physics [32] with the regional surface potential. It is derived from the input voltage equation in different regions. The operation of a transistor, which can be classified into three main regions of operation, namely, accumulation, depletion, and strong inversion, depending on the gate-bulk bias (and channel doping concentration). The boundaries of each region of operation are well defined, in which the flat-band voltage, V_{FB} is the boundary between accumulation and depletion regions, and the threshold voltage, V_t is the boundary between weak inversion and strong inversion regions [26]. In each region of operation, an explicit surface potential solution can be derived. Unlike the explicit ϕ_s approach [4] which

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“glues” the pieces of surface potentials together [6], the unified regional surface potentials are “stitched” using interpolation/smoothing functions [33]. This ensures the model continuity for all derivatives.

The drain current is carefully re-derived to ensure the fulfilment of the Gummel symmetry test (GST) [34] and physical behaviour for all geometries without the need for accurate surface potentials. It shows a good GST behaviour up to sixth order derivatives while maintaining physical bias scalability as well as possible extension to asymmetry source/drain within the same core model [7, 35]. In addition, the model has been extended to include poly-Si gate doping effect from accumulation to depletion [36], quantum effect [33], and short-channel effect [13] in AC charge/capacitance model. Furthermore, a complete short-channel/narrow-width drain current model has also been demonstrated in [37]. The model has also been used for technology development and technology prediction through a process correlation modeling approach [17].

The URM approach has also been extended to model future devices, such as strained-Si [38] and DG MOSFETs [11, 39, 40]. In this work, common-gate asymmetric double-gate model will be discussed in detail. One unique feature of the URM approach is that the model does not require accurate surface-potential solution for its correct model behaviour. Finally, the recent model extension to three-terminal MOSFETs without a body contact is also included. More details of the URM approach for bulk-MOSFET will be discussed in the following sections, starting from surface potential, then the charge, and finally the current model.

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2.3. Unified regional surface potential bulk-MOSFET model

Unified regional surface potential (URSP) is formulated according to the regional solution of the input voltage equation and joined with interpolation and smoothing functions [33, 36, 41]. The essential physics is contained in the regional surface potential without the need to solve surface potentials accurately. It is a reasonable compromise between model complexity and simplicity, and sustainable for extension to future devices. In this chapter, the focus is on the model for bulk-MOSFET. The basic bulk-MOSFET model formulation is governed by the potential balance equation

$$V_{gb} = \phi_{MS} + \phi_s + \phi_p + V_{ox}, \quad (2.1)$$

where ϕ_s and ϕ_p are the potentials at the channel surface (SiO₂/Si interface) and poly-Si gate surface (poly-Si/SiO₂ interface), respectively, $V_{ox} = Q_g/C_{ox}$ is the potential drop across the gate oxide of thickness t_{ox} associated with oxide capacitance (per unit area) $C_{ox} = \epsilon_{ox}/t_{ox}$ and Q_g is the gate charge (per unit area). The work function difference between poly-Si gate and silicon channel for the n^+ gate CMOS technology

$$\phi_{MS} = \frac{(W_M - W_S)}{q} \approx -v_{th} \ln \left(\frac{N_{gate} N_{ch}}{n_i^2} \right) \quad (2.2)$$

is determined by the n^+ poly-Si gate doping concentration, N_{gate} and p^+ channel doping concentration, N_{ch} and n_i is the intrinsic carrier concentration. The charge balance equation demands:

$$Q_g + Q_{sc} + Q_{ox} = 0, \quad (2.3)$$

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in which Q_{ox} is the effective fixed oxide charge and Q_{sc} is the induced channel charge.

Combining both balance equations (2.1) and (2.3), we have

$$V_{gb} - V_{FB} - \phi_s - \phi_p = -\frac{Q_{sc}}{C_{ox}} \quad (2.4)$$

in which the flat-band voltage is given by

$$V_{FB} = \phi_{MS} - \frac{Q_{ox}}{C_{ox}}. \quad (2.5)$$

For very high gate doping concentration, the surface potential drop at the poly-Si gate, ϕ_p can be ignored. The right hand side of (2.4) can be determined by applying the Gauss' Law at the SiO₂/Si interface in the channel,

$$Q_{sc} = -\epsilon_{si} E_{si} = \epsilon_{si} \left. \frac{d\phi}{dx} \right|_{x=0} \quad (2.6)$$

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

$$\frac{d^2\psi}{dx^2} = -\frac{\rho}{\epsilon_{si}} \quad (2.7)$$

where

$$\rho = q[p - n - N_A + N_D] \quad (2.8)$$

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

$$p = n_i \exp\left(\frac{-\phi + \phi_F}{v_{th}}\right) \quad (2.9)$$

$$n = n_i \exp\left(\frac{\phi - V_{cb} - \phi_F}{v_{th}}\right), \quad (2.10)$$

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p and n are the concentrations of holes and electrons, respectively. The concentrations of the acceptors, N_A and donors, N_D are related to the concentrations of the remote holes and electrons, respectively [42],

$$N_A = p_{\infty} \Big|_{\phi \rightarrow 0} = n_i \exp\left(\frac{\phi_F}{v_{th}}\right) \quad (2.11)$$

$$N_D = n_{\infty} \Big|_{\phi \rightarrow 0} = n_i \exp\left(\frac{-V_{cb} - \phi_F}{v_{th}}\right) \quad (2.12)$$

V_{cb} is the position-dependent channel-to-bulk voltage and

$$\phi_F = v_{th} \operatorname{arcsinh}\left(\frac{N_A - N_D}{2n_i}\right) \quad (2.13)$$

is the bulk Fermi level [43]. Integrating (2.7) once with the boundary conditions at surface and deep in the bulk, $\phi|_{x=0} = \phi_s$, $d\phi/dx|_{x \rightarrow \infty} = 0$ and $\phi|_{x \rightarrow \infty} = 0$, from (2.6) the total induced charge in the channel can be obtained:

$$Q_{sc} = -C_{ox} \gamma \operatorname{sgn}(\phi_s) \sqrt{f_{\phi}} \quad (2.14)$$

with

$$f_{\phi} = v_{th} \left(e^{-\frac{\phi_s}{v_{th}}} - 1 \right) + \phi_s + e^{-\frac{V_{cb} + 2\phi_F}{v_{th}}} \left(v_{th} e^{\frac{\phi_s}{v_{th}}} - v_{th} - \phi_s \right). \quad (2.15)$$

$$\gamma = \sqrt{2q\epsilon_{si} N_{ch}} / C_{ox} \quad (2.16)$$

is the body factor, and

$$N_{ch} = n_i e^{\frac{\phi_F}{v_{th}}} \quad (2.17)$$

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is effective channel doping concentration, where $N_{ch} \approx N_A$ when $N_A \gg N_D$. The $\text{sgn}(\phi_s)$ $= \pm 1$ represents the sign of ϕ_s . Therefore, the solution is now an implicit solution in the form of [42]:

$$V_{gb} - V_{FB} - \phi_s = \gamma \text{sgn}(\phi_s) \sqrt{ \underbrace{v_{th} \left(e^{\frac{-\phi_s}{v_{th}}} - 1 \right)}_{\text{Holes}} + \underbrace{\phi_s \left(1 - e^{\frac{-V_{cb} + 2\phi_F}{v_{th}}} \right)}_{\text{Ionized Acceptor \& Donor}} + \underbrace{v_{th} e^{\frac{-V_{cb} + 2\phi_F}{v_{th}}} \left(e^{\frac{\phi_s}{v_{th}}} - 1 \right)}_{\text{Electrons}} }, \quad (2.18)$$

where the contributions due to holes, ionized acceptors and donors, and electrons, respectively, are shown in (2.18). This implicit solution can either be solved iteratively with the Newton-Raphson (NR) algorithm [26] or approximately [44, 45], which uses explicit and linearized-iteration approach to improve the accuracy of surface potential. Explicit regional solutions exist when only considering the contributions of the dominant terms in accumulation, depletion, and strong inversion, respectively. This will be detailed in the ensuing sections.

2.3.1. Piecewise explicit regional surface potential solution

In this section, we derive piecewise unified regional surface potentials, ignoring poly-Si gate doping effect (i.e., assuming $\phi_p=0$ when N_{gate} is very high). When the NMOS is biased in strong accumulation ($V_{gb} < V_{FB}$) or strong inversion ($V_{gb} > V_t$), the net charge is dominated by the holes or electrons, respectively. In strong accumulation when only holes are considered, we can approximate the implicit solution of equation (2.18) by replacing $f_\phi \approx v_{th} \exp[-\phi_s/v_{th}]$ and rearranging it as:

$$\ln \left(\frac{(V_{gb} - V_{FB})^2}{\gamma^2 v_{th}} \right) + \ln(1 - a\phi_s + b\phi_s^2) = -\phi_s/v_{th} \quad (2.19)$$

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and $f_\phi \approx v_{th} \exp[(\phi_s - 2\phi_F - V_{cb})/v_{th}]$ as:

$$\ln \left(\frac{(V_{gb} - V_{FB})^2}{\gamma^2 v_{th}} \right) + \ln(1 - a\phi_s + b\phi_s^2) = (\phi_s - 2\phi_F - V_{cb})/v_{th} \cdot \quad (2.20)$$

with

$$a = \frac{2}{V_{gb} - V_{FB}} \quad (2.21)$$

and

$$b = \frac{1}{(V_{gb} - V_{FB})^2} \cdot \quad (2.22)$$

The second natural-log term is approximated with a Taylor series expansion

$$\begin{aligned} \ln(1 - a\phi_s + b\phi_s^2) &= -a\phi_s + b\phi_s^2 - \frac{1}{2}(-a\phi_s + b\phi_s^2)^2 + \frac{1}{3}(-a\phi_s + b\phi_s^2)^3 + \dots \\ &\approx -a\phi_s + \left(b - \frac{a^2}{2}\right)\phi_s^2 + \left(ab - \frac{a^3}{3}\right)\phi_s^3 \end{aligned} \quad (2.23)$$

and rearranged to express as a general cubic function so that it can be solved analytically [46],

$$\phi_s^3 + p\phi_s^2 + q\phi_s + r = 0 \quad (2.24)$$

where

$$p = s \left(b - \frac{a^2}{2} \right) \quad (2.25)$$

$$q = \begin{cases} s \left(\frac{1}{v_{th}} - a \right), & \text{strong accumulation} \\ -s \left(\frac{1}{v_{th}} + a \right), & \text{strong inversion} \end{cases} \quad (2.26)$$

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$$r = \begin{cases} s \ln \left(1 + \frac{(V_{gb} - V_{FB})^2}{v_{th} \gamma^2} \right), & \text{strong accumulation} \\ s \left(\ln \frac{(V_{gb} - V_{FB})^2}{v_{th} \gamma^2} + \frac{2\phi_F + V_{cb}}{v_{th}} \right), & \text{strong inversion} \end{cases} \quad (2.27)$$

in which

$$s = \left(ab - \frac{a^3}{3} \right)^{-1}. \quad (2.28)$$

In the above derivations, an extra ‘1’ term in r is added to adjust the strong-accumulation solution at high channel doping concentration so that it approaches zero in the region beyond its validity, as shown in Figure 2.1.

Figure 2.1: Comparison of unified regional surface potential using the intermediate solution r as of (2.28) with Newton-Raphson iterative solution. The extra ‘1’ term in r for accumulation region has improved the unified regional accumulation at high doping concentration for the region beyond its validity.

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The following intermediate variables are introduced:

$$A = \frac{1}{3}(3q - p^2) \quad (2.29)$$

$$B = \frac{1}{27}(2p^3 - 9pq + 27r). \quad (2.30)$$

The cubic solution is dependent on the sign of the discriminant of

$$D = 4\frac{A^3}{27} + B^2 \quad (2.31)$$

where D is always positive in strong accumulation or strong inversion. Therefore, the analytical solution of the cubic surface-potential equation (2.24) has a unique solution:

$$\phi_s = \left(\frac{\sqrt{D} - B}{2} \right)^{1/3} - \frac{A}{3} \left(\frac{2}{\sqrt{D} - B} \right)^{1/3} - \frac{p}{3} \quad (2.32)$$

and we use the following notations for the piecewise regional solutions

$$\phi_s = \begin{cases} \phi_{CC}, V_{gb} < V_{FB} \\ \phi_{SS}, V_{gb} > V_t \end{cases} \quad (2.33)$$

in strong-accumulation (ϕ_{CC}) and strong-inversion (ϕ_{SS}), respectively.

In depletion region where $f_\phi \approx \phi_s$, the regional surface potential solution is given by:

$$\phi_{DD} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gb} - V_{FB}} \right]^2 \quad (2.34)$$

This equation models the constant slope for different value of N_A in the depletion region, as shown in Figure 2.2.

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Figure 2.2: The depletion slope is approaching the ideal subthreshold swing (i.e. 60mV/decade) as the channel doping decreases. The slope changes are already captured by the depletion region solution.

In “weak-accumulation” near and below the V_{FB} , although there is no equivalent “depletion” (for NMOS), from Figure 2.2 it can be seen that ϕ_s slope changes with the channel doping concentration. This can be modeled by a “fictitious” depletion solution in weak accumulation, which is similar to (2.34) but with a sign change $\phi_s = -\phi_{ADD}$:

$$V_{gb} - V_{FB} + \phi_{ADD} = \gamma \sqrt{-\phi_{ADD}} \quad (2.35)$$

and its solution is given by

$$\phi_{ADD} = - \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} - V_{gb} + V_{FB}} \right]^2. \quad (2.36)$$

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2.3.2. Unified regional surface potential solution

The piecewise regional solutions derived so far are only physically correct for the intended regions. To produce a surface potential solution that is usable across every region, the unified regional approach is adopted. In the unification of the regional solutions at flat-band voltage, the complementary reverse and forward interpolation functions are used, which are in the form:

$$\vartheta_r \{x; \sigma\} \equiv \left\{ x - \sqrt{x^2 + 4\sigma} \right\} / 2 \quad (2.37)$$

and

$$\vartheta_f \{x; \sigma\} \equiv \left\{ x + \sqrt{x^2 + 4\sigma} \right\} / 2, \quad (2.38)$$

respectively. When a single value of σ is used with x equal to $V_{gb} - V_{FB}$, the sum of these two functions is exactly equal to $V_{gb} - V_{FB}$ [30]. The interpolated, flat-band shifted gate-bulk voltage ($V_{gb} - V_{FB}$) for below and above V_{FB} are given as:

$$V_{GBR} = \vartheta_r \{V_{gb} - V_{FB}; \sigma_a\} \quad (2.39)$$

and

$$V_{GBF} = \vartheta_f \{V_{gb} - V_{FB}; \sigma_f\} \quad (2.40)$$

where σ_a and σ_f are the smoothing parameters, respectively, which are used to ensure charge neutrality at flat-band ($\phi_s=0$ at $V_{gb}=V_{FB}$) and smoothness near V_{FB} . Therefore, for voltages below V_{FB} , the interpolated strong accumulation and weak accumulation solutions are

$$\phi_{STR,CC} = \phi_{CC} \Big|_{V_{gb}-V_{FB}=V_{GBR}}, \quad (2.41)$$

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$$\phi_{ASUB} = - \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} - V_{GBR}} \right]^2, \quad (2.42)$$

respectively. For voltages above V_{FB} , the interpolated depletion (or weak inversion) and strong inversion solutions are

$$\phi_{SUB} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{GBF}} \right]^2 \quad (2.43)$$

and

$$\phi_{STR,SS} = \phi_{SS} \Big|_{V_{gb}-V_{FB}=V_{GBF}}, \quad (2.44)$$

respectively. In order to obtain a smooth transition from strong accumulation to weak accumulation and depletion to strong inversion, respectively, another two smoothing functions are used, namely

$$\vartheta_{eff,cc} \{x, x_{sat}; \delta\} \equiv x_{sat} - 0.5 \left[x_{sat} - x + \delta - \sqrt{(x_{sat} - x + \delta)^2 - 4\delta x_{sat}} \right] \quad (2.45)$$

and

$$\vartheta_{eff,ss} \{x, x_{sat}; \delta\} \equiv x_{sat} - 0.5 \left[x_{sat} - x - \delta + \sqrt{(x_{sat} - x - \delta)^2 + 4\delta x_{sat}} \right]. \quad (2.46)$$

These two smoothing functions ensure that strong accumulation will smoothly join weak accumulation, and depletion will smoothly join strong inversion, respectively. Therefore, the unified regional surface potential solution from strong accumulation to weak accumulation and depletion to strong inversion are

$$\phi_{ACC} = \vartheta_{eff,cc} \{ \phi_{ASUB}, \phi_{STR,CC}; \delta_{cc} \} \quad (2.47)$$

$$\phi_{DS} = \vartheta_{eff,ss} \{ \phi_{SUB}, \phi_{STR,SS}; \delta_{ss} \} \quad (2.48)$$

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where δ_{cc} and δ_{ss} are two smoothing parameters that control the abruptness of the respective transitions. An optional correction in the URSP similar to the explicit NR correction in [45] or [47], can be introduced by using V_{GBF}/V_{GBR} in (2.39)/(2.40) and ϕ_{ACC}/ϕ_{DS} in (2.47)/(2.48) in place of $V_{gb}-V_{FB}$ and ϕ_s , respectively, to reduce dependency of the smoothing parameters, δ_{cc}/δ_{ss} on the drain current model in all doping ranges. Finally, a single-piece URSP expression valid in all operating regions is given by

$$\phi_{SEFF} = \phi_{ACC} + \phi_{DS}, \quad (2.49)$$

where the final solution is a one piece solution that follows the iterative surface potential. Numerical simulations using Medici are generated for verification of the URSP model.

Figure 2.3 shows the components of the regional surface potential solutions in strong accumulation, weak accumulation, weak inversion (depletion), and strong inversion at undoped body, in which $N_D=N_A=0$ and ϕ_s vs. $V_{gb}-V_{FB}$ is expected to be symmetric, i.e., an odd function. To demonstrate the symmetry behaviour, the ϕ_{ACC} is “flipped” at $V_{gb}=V_{FB}$ and compared with the ϕ_{DS} solution. The sum is shown in the inset of Figure 2.3 with negligible difference. This capability is not available in other existing models.

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Figure 2.3: Components of unified regional surface potential at undoped: strong accumulation, weak accumulation, weak inversion and strong inversion. Inset: Addition of the left-right flipped ϕ_{acc} and ϕ_{is} , only negligible noise observed signify both unify solution is symmetrical at $\phi=0$ (generated without the extra “1” in r in (2.27)), as predicted from the equation.

As the channel doping concentration increases, the weak accumulation region is disappearing and becomes negligible. The strong accumulation is dominating the whole $V_{gb} < V_{FB}$ region. This is readily being taken care of by the smoothing function (2.45), as shown in Figure 2.4 when the channel doping is at $N_{ch}=10^{17} \text{ cm}^{-3}$.

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Figure 2.4: Components of unified regional surface potential at $N_{ch}=10^{17}\text{ cm}^{-3}$: strong accumulation, depletion and strong inversion. The smoothing function (2.45) ensures that the dominant strong accumulation is used when the doping is higher.

Figure 2.5 shows the comparison of single-piece URSP based on equation (2.49) and iterative surface potential solutions for different body doping. The channel doping starts from the undoped body to highly doped body of 10^{18} cm^{-3} . The single-piece URSP solutions are able to follow the iterative solutions very well. The inset shows the errors (in the mV range) for each doping concentration. Figure 2.6 shows the first-order derivatives of URSP for the corresponding body doping, with the second-order derivatives shown in the inset. All the numerical derivatives shown in this work are computed based on central difference formula. To ensure fair and consistent comparison with the numerical data from TCAD simulations, all model evaluations are using the same voltage step sizes as those in TCAD simulations and maximum number of significant digits are used in all TCAD data for better accuracy in higher order numerical derivatives. The URSP solutions are smooth, continuous and

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matching all NR solutions. Finally, the direct play-back of the single-piece URSP at different body bias (or channel voltage) is shown in Figure 2.7 and compared with the NR solutions with the first-order derivatives shown in the inset.

Figure 2.5: Comparison of unified regional surface potential with improved accumulation solution with respect to Newton-Raphson iterative solution of surface potential for different channel doping concentration. Inset: The error relative to Newton-Raphson solution.

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Figure 2.6: Comparison of derivative of surface potential for unified regional surface potential with improved accumulation solution derivative with Newton-Raphson solution for different channel doping concentration. Inset: 2nd derivative of corresponding surface potentials.

Figure 2.7: Direct playback of single-piece explicit surface potential, ϕ_{seff} and first order derivative (inset) for fermi level ($V_{cb} = V_{sb}$) variations, including positive bias condition, is compared with iterative solutions from Pao and Sah voltage equation (symbols).

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2.3.3. Poly-Si gate doping effect

Lower doping concentration in the poly-Si gate has a larger influence on the channel potential. Poly-depletion effect (PDE) is well known and has been modeled in most of the compact models [2, 4]. However, poly-accumulation effect (PAE) that is observed in the accumulation region with a “crossing-over” in the gate capacitance near the flat-band voltage [30, 48] is not included in most models. Under normal operating region, it is reasonable to ignore poly-inversion effect (PIE) [48] as it is unusual to bias at such a high gate voltage. Nevertheless, PAE is still within the normal operating region in accumulation that should be included in the compact model. This is especially critical for MOS varactors, which may operate in accumulation region [49].

The poly-Si gate surface potential, ϕ_p can be calculated with similar approach to that for ϕ_s . The voltage and charge balanced equations including n^+ poly-gate doping effect give,

$$V_{gb} - \phi_{ms} - \phi_s - \phi_p = (Q_g + Q_{ox}) / C_{ox} . \quad (2.50)$$

The “induced” gate charge (per unit area) in the poly-Si gate can be similarly obtained from Poisson’s equation in poly-Si gate with Gauss’ law applied to the poly-Si/SiO₂ interface:

$$Q_g = +C_{ox} \gamma_p \text{sgn}(\phi_p) \sqrt{f_{\phi,p}} \quad (2.51)$$

where

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$$f_{\phi,p} = v_{th} \left[\exp\left(-\frac{\phi_p}{v_{th}}\right) - 1 \right] + \phi_p \left[1 - \exp\left(\frac{-2\phi_{F,p}}{v_{th}}\right) \right] \\ + v_{th} \exp\left(-\frac{2\phi_{F,p}}{v_{th}}\right) \left[\exp\left(\frac{\phi_p}{v_{th}}\right) - 1 \right] \quad (2.52)$$

and

$$\gamma_p = \sqrt{2q\epsilon_{si}N_{gate}} / C_{ox} . \quad (2.53)$$

Note that there is no imref-splitting in the gate surface potential in (2.52) due to isolation of the gate electrode from the source/drain. The ϕ_s is coupled with ϕ_p in the balanced equation, which requires coupled iterative solutions together with (2.4), (2.14), and (2.15). This can be too difficult for compact modeling.

In the URM approach [30], explicit unified regional ϕ_p and ϕ_s are decoupled by first deriving the regional ϕ_s solutions, ignoring ϕ_p effect on ϕ_s at high gate doping. Then, regional ϕ_p solutions are solved with $V_{gb}-V_{FB}-\phi_s$ as the argument in the ϕ_p -related smoothing function. Finally, ϕ_p is included in ϕ_s with $V_{gb}-V_{FB}-\phi_p$ in ϕ_s -related smoothing function.

For PAE, only accumulated electrons and holes in the gate and channel, respectively, are considered. Equating the two types of carriers

$$\gamma \operatorname{sgn}(\phi_s) \sqrt{v_{th} (e^{-\phi_s/v_{th}} - 1)} = \gamma_p \operatorname{sgn}(\phi_p) \sqrt{v_{th} (e^{-\phi_p/v_{th}} - 1)} \quad (2.54)$$

one can solve for the regional accumulation ϕ_p as [48],

$$\phi_{p,cc} = -v_{th} \ln \left[\frac{\gamma^2}{\gamma_p^2} \exp\left(-\frac{\phi_s}{v_{th}}\right) + \left(1 - \frac{\gamma^2}{\gamma_p^2}\right) \right]. \quad (2.55)$$

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For PDE, only donor in poly-gate is considered and, therefore, poly-depletion regional solution is given by

$$\phi_{p,dd} = \left[-\frac{\gamma_p}{2} + \sqrt{\frac{\gamma_p^2}{4} + V_{gb} - V_{FB} - \phi_s} \right]^2. \quad (2.56)$$

For PIE [48], similar procedure as in section 2.3.1 for ϕ_{ss} derivation can be followed to obtain poly-gate surface potential in strong inversion, $\phi_{p,ss}$ with $V_{gb} - V_{FB}$ to be replaced by $V_{gb} - V_{FB} - \phi_s$ and γ replaced by γ_p . The unified regional poly-gate surface potentials for PAE, PDE, and PIE are obtained by replacing $V_{gb} - V_{FB} - \phi_s$ with the smoothing function V_{gbs}

$$\phi_{p,acc} = \phi_{p,cc} \Big|_{\phi_s = \phi_{sr}} \quad (2.57)$$

$$\phi_{p,sub} = \phi_{p,dd} \Big|_{V_{gb} - V_{FB} - \phi_{ds} = V_{gbs}} \quad (2.58)$$

$$\phi_{p,str} = \phi_{p,ss} \Big|_{V_{gb} - V_{fb} - \phi_{ds} = V_{GBSF}} \quad (2.59)$$

where

$$\phi_{sr} = \vartheta_r \{ \phi_{acc}; \rho_r \} \quad (2.60)$$

$$V_{gbs} = \vartheta_f \{ V_{gb} - V_{FB} - \phi_{ds}; \rho_f \}. \quad (2.61)$$

Then, the combined unified regional poly-Si gate surface potential from depletion to strong inversion is given by

$$\phi_{p,ds} = \vartheta_{eff,ss} \{ \phi_{p,sub}, \phi_{p,str}; \delta_\rho \}. \quad (2.62)$$

The final single-piece poly-Si gate surface potential is given by

$$\phi_{peff} = \phi_{p,acc} + \phi_{p,ds}. \quad (2.63)$$

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Next, we derive the channel surface potential ϕ_s with the influence of ϕ_p , which can be carried out similar to section 2.3.1, replacing $V_{gb}-V_{FB}$ with $V_{gb}-V_{FB}-\phi_p$, in which $\phi_p=\phi_{p,eff}$ is from the unified regional ϕ_p solutions derived above. We use lower-case subscripts to denote inclusion of poly-Si doping effects. In strong accumulation

$$\phi_{str,cc} = \phi_{cc} \Big|_{V_{gb}-V_{FB}-\phi_{p,acc}=V_{gbr}}, \quad (2.64)$$

in which $\phi_{str,cc}$ is from (2.33), replacing $V_{gb}-V_{FB}$ with $V_{gb}-V_{FB}-\phi_{p,acc}$, using the smoothing function, $V_{gbr}=\vartheta_r$.

In weak accumulation,

$$\phi_{asub} = \phi_{add} \Big|_{V_{gb}-V_{FB}-\phi_{p,acc}=V_{gbr}} = - \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} - V_{gbr}} \right]^2 \quad (2.65)$$

in depletion,

$$\phi_{sub} = \phi_{dd} \Big|_{V_{gb}-V_{FB}-\phi_{p,ds}=V_{gbf}} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gbf}} \right]^2 \quad (2.66)$$

and in strong inversion,

$$\phi_{str,ss} = \phi_{ss} \Big|_{V_{gb}-V_{FB}-\phi_{p,ds}=V_{gbf}}. \quad (2.67)$$

The combined solution for strong accumulation to weak accumulation and depletion-to-strong inversion, respectively

$$\phi_{acc} = \vartheta_{eff,cc} \{ \phi_{asub}, \phi_{str,cc}; \delta_{cc} \} \quad (2.68)$$

$$\phi_{ds} = \vartheta_{eff,ss} \{ \phi_{sub}, \phi_{str,ss}; \delta_{ss} \} \quad (2.69)$$

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and full-piece URSP,

$$\phi_{seff} = \phi_{acc} + \phi_{ds} . \quad (2.70)$$

In the above equations, lower-case ϕ_{seff} related expressions have included the ϕ_p -related expressions, in which ϕ_s is also involved. The decoupled ϕ_s and ϕ_p derivation starts with ϕ_{peff} including ϕ_{SEFF} (upper-case, assuming $\phi_p=0$ at high gate doping), followed by ϕ_{seff} including ϕ_{peff} at a given N_{gate} . An optional second iteration can improve the accuracy of the initial ϕ_{peff} using the calculated ϕ_{seff} .

Figure 2.8 shows the comparison of poly-Si gate surface potential from numerical simulation and the URSP solutions. Three regions of operation, namely, PAE, PDE, and PIE, are clearly observable. The inset shows the first-order derivative of the poly-Si gate surface potential. The model is able to match the first-order derivative smoothly and accurately.

Figure 2.8: Comparison of unified regional poly-Si gate surface potential, accumulation, depletion and strong inversion. Inset: The first-order derivatives showing PAE, PDE and PIE, respectively.

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2.3.4. Quantum-mechanical correction

The gate oxide thickness (t_{ox}) in advanced n^+ implanted poly-Si gate MOSFET (n^+/p^+ dual-gate CMOS [50]) has been scaled into the regime in which quantum-mechanical effect (QME) cannot be ignored. MOSFET electrical characteristics are significantly affected by QME as t_{ox} continues to reduce. The basic compact model equation must capture these physical behaviours in order to maintain acceptable accuracy in circuit simulator over the required bias conditions and physical variations.

It was suggested by van Dort *et. al.* that the quantization of the space charge density can be accounted for by a simple modification of the intrinsic carrier density for free carriers [51],

$$n_i^{qm} = n_i f^{qm}, \quad (2.71)$$

$$f^{qm} = \exp\left[-\Delta E_g^{qm} / v_{th}\right] \quad (2.72)$$

in which n_i is the classical intrinsic carrier density and $\Delta E_g^{qm} = \Delta E_g / 2q$. The increase of silicon energy bandgap (ΔE_g) due to the effect of carrier energy quantization in the surface potential well [52, 53]

$$\Delta E_g = \kappa^{qm} \frac{3\hbar^2}{8qm^*} \left[\frac{12m^* q^2}{\epsilon_{si} \hbar^2} F_q \right]^{2/3}, \quad (2.73)$$

$$F_q = \frac{2C_{ox}(V_{gb} - V_{FB} - \phi_s)/q}{3} \quad (2.74)$$

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is derived through triangular-shaped potential approximation with self-consistent solutions of the coupled Schrödinger and Poisson's equations, $\hbar$ is the reduced Planck's constant, m^* is the effective electron/hole mass and the factor κ^{qm} , with a default value of one, was introduced in [54]. Following the URSP approach, the electric field in QME model, F_q in (2.74) in accumulation is redefined as:

$$F_{qa} = \frac{2C_{ox}(V_{gbr} - \phi_{acc})/q}{3}, \quad (2.75)$$

and in strong inversion at source end:

$$F_{qs} = \frac{2C_{ox}(V_{gbf} - \phi_{ds,s})/q}{3} \quad (2.76)$$

and at the drain end,

$$F_{qd} = \frac{2C_{ox}(V_{gbf} - \phi_{ds,d})/q}{3}. \quad (2.77)$$

Similar to bulk MOSFET derivation, the voltage equation becomes

$$V_{gb} - V_{FB} - \phi_s^{qm} = -Q_{sc}^{qm}/C_{ox}, \quad (2.78)$$

in which the superscript “qm” denotes inclusion of QME. With the van Dort model for QME using (2.71) for n_i^{qm} , ignoring the spatial dependence of ϕ_s in ΔE_g , and following the same approach to solving Poisson's equation with Gauss' law applied to the SiO₂/Si interface, the induced channel charge density can be derived [30]:

$$Q_{sc}^{qm} = -C_{ox} \gamma \text{sgn}(\phi_s^{qm}) \sqrt{f_\phi^{qm}} \quad (2.79)$$

where

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$$f_{\phi}^{qm} = v_{th} \left[f^{qm} \exp\left(-\frac{\phi_s^{qm}}{v_{th}}\right) - 1 \right] + \phi_s^{qm} \left[1 - \exp\left(-\frac{V_{cb} + 2\phi_F}{v_{th}}\right) \right] \\ + v_{th} \exp\left(-\frac{V_{cb} + 2\phi_F}{v_{th}}\right) \left[f^{qm} \exp\left(\frac{\phi_s^{qm}}{v_{th}}\right) - 1 \right] \quad (2.80)$$

Following the URSP approach in section 2.3.1, all the unified regional pieces can be derived as follows,

$$\phi_{str,cc}^{qm} = \phi_{CC}^{qm} \Big|_{V_{gb}-V_{FB}-\phi_{p,acc}^{qm}=V_{gbr}} \quad , \quad (2.81)$$

with

$$f_{acc}^{qm} = \exp\left(-\frac{\Delta E_s^{qm}}{q v_{th}}\right) - \varepsilon^{qm} \quad (2.82)$$

where ε^{qm} is a fitting parameter to adjust the flat-band shift due to QME for the accumulation piece, with (2.28) modified to

$$r^{qm} = s \ln \left(1 + \frac{(V_{gb} - V_{FB})^2}{v_{th} \gamma^2 f_{acc}^{qm}} \right) \quad (2.83)$$

for the ϕ_{cc}^{qm} expression, where $s = 1/(ab - a^3/3)$ is the same as (2.28).

In the depletion region, there is no QME, so

$$\phi_{sub}^{qm} = \phi_{sub} \quad . \quad (2.84)$$

In strong inversion region,

$$\phi_{str}^{qm} = \phi_{SS}^{qm} \Big|_{V_{gb}-V_{FB}-\phi_{p,ds}^{qm}=V_{gbf}} \quad , \quad (2.85)$$

with (2.28) modified to

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$$r^{qm} = s \left(\ln \frac{V_{gbf}^2}{v_{th} \gamma^2 f^{qm}} + \frac{2\phi_F + V_{cb}}{v_{th}} \right), V_{gb} > V_{FB} \quad (2.86)$$

in the ϕ_{ss}^{qm} expression.

Numerical data are generated with Medici for verification of the URSP with QME. Figure 2.9 shows the comparison with Medici data for surface potential with and without QME. The inset shows the second derivative of the corresponding surface potentials. The URSP model matches the numerical data correctly and smoothly.

Figure 2.9 : Comparison of unified regional surface potential with Medici for with and without quantization effect. Inset: The second-order derivatives of full-piece unified regional surface potential.

For poly-Si gate surface potential incorporating with the QME, similar expressions can be derived.

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$$\phi_{p,acc}^{qm} = \phi_{p,acc}^{qm} \Big|_{\phi_s = \phi_r^{qm}}, \quad (2.87)$$

where

$$\phi_{sr}^{qm} = \vartheta_r \left\{ \phi_{acc}^{qm} + \beta^{qm} \frac{\Delta E_g}{q} \Big|_{\phi_s = \phi_{acc}^{qm}} ; \rho_r \right\} \quad (2.88)$$

and β^{qm} is introduced to adjust the quantum effect at high poly-Si gate doping. In depletion,

$$\phi_{p,sub}^{qm} = \phi_{p,sub} \Big|_{\phi_s = \phi_{sub}^{qm}}. \quad (2.89)$$

In strong inversion,

$$\phi_{p,str}^{qm} = \phi_{p,str}^{qm} \Big|_{V_{gb} - V_{FB} - \phi_{s,ds}^{qm} = V_{gbs}}. \quad (2.90)$$

The URSP model with and without QME is shown in Figure 2.10 at high and low poly-Si gate doping concentrations. The threshold shift of the transistor is mainly due to the PDE, while the QME has strong effect in strong inversion. Similar plot of the poly-Si gate surface potential is also shown in Figure 2.11.

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Figure 2.10 : Surface potential with and without both quantization and poly-Si gate effects.

Figure 2.11 Poly-Si gate surface potential with and without both quantization effect and poly-Si gate effects.

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2.3.5. Short channel effects and reverse short channel effects

As the channel length reduces, the effectiveness of the gate-voltage control over the channel charge is degraded due to lateral-field effect. Effectively, this causes the threshold voltage reduction at decreasing gate length or increasing drain voltage. Two short channel models, namely, the bulk-charge sharing (BCS) and quasi-two-dimensional (quasi-2D) drain-induced barrier lowering (DIBL) have been used in threshold-voltage based modeling [55]. These models are extended to include the short-channel effect on the surface potentials. The BCS assumes the depletion width in the channel takes a trapezoidal shape as shown in Figure 2.12 due to overlapping of electrical field from source/drain and the channel.

Figure 2.12. Charge sharing model for calculating the charge reduction.

Therefore, it can be rewritten as a modification of the bulk-charge term, γ from equation (2.18) as,

$$\gamma_{eff,s} = \gamma \left(1 - \frac{\lambda}{L_{eff}} x_{d\max,s} \right) \quad (2.91)$$

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$$\gamma_{eff,d} = \gamma \left(1 - \frac{\lambda}{L_{eff}} x_{d\max,d} \right) \quad (2.92)$$

at the source and drain ends, respectively, with a maximum depletion depth at source end

$$x_{d\max,s} = \sqrt{\frac{2\epsilon_{si}(2\phi_F + V_{sb})}{qN_{ch}}} \quad (2.93)$$

and a maximum depletion depth at drain end

$$x_{d\max,d} = \sqrt{\frac{2\epsilon_{si}(2\phi_F + V_{db})}{qN_{ch}}} \quad (2.94)$$

where λ is a fitting parameter for BCS and L_{eff} is the effective channel length. Various poly-gate length (L_g) devices are created in Medici to generate the TCAD simulation data for verification of the compact model. The model is, however, using a slightly different concept when considering the effective channel length, defined as $L_{eff} = L_g - 2\sigma x_j$ where x_j is the LDD junction depth and σ is a fitting parameter, which is extracted together with the halo parameters (peak, spread, and centroid) for all gate lengths. To include the BCS in the surface potential, γ_{eff} replaces the γ in URSP solutions in depletion and strong-inversion regions. Note that this is different from the previous work, which is taking an average of maximum depletion depth of the source and drain ends [30], with separate effective bulk-charge terms at the drain and source sides, respectively, which will ensure fulfilment of Gummel symmetry at short channel.

Similar BCS model can be introduced for short-channel devices together with the PDE by changing γ to γ_{eff} , with

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$$\gamma_{peff} = \gamma_p \left(1 - \frac{\lambda_p}{L_{eff}} \sqrt{\frac{2\epsilon_{si} (2\phi_F)}{qN_{gate}}} \right). \quad (2.95)$$

The second short-channel model is the quasi-2D potential-barrier lowering (PBL) effect. The Gauss law is applied to a fixed rectangular box (Gaussian box) of depth, at X_{dmax} and effective channel length, L_{eff} as shown in Figure 2.13.

Figure 2.13. Gaussian box and boundary condition for quasi-two dimensional solution

Under the given boundary conditions [56]:

$$\begin{aligned} \phi(0, y) &= \phi_s(y) \\ \phi(x_{dmax}, y) &= -V_{sb} \\ \left. \frac{d\phi(x, y)}{dy} \right|_{x=0} &= -\frac{C_{ox}}{\epsilon_{si}} (V_{gb} - V_{FB} - \phi_s(y)) \\ \left. \frac{d\phi(x, y)}{dy} \right|_{x=x_{dmax}} &= 0 \end{aligned} \quad (2.96)$$

Then, quasi-2D Gauss Law can be written as

$$C_{ox} \left[V_{gb} - V_{FB} - \phi(y) + \frac{\epsilon_{si}}{C_{ox}} \frac{x_{dmax}}{\eta} \frac{d\phi^2(y)}{dy^2} \right] = -\frac{\rho}{\epsilon_{si}} \quad (2.97)$$

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where, $E_s(y)$ is the lateral surface electric field, $\phi(y)$ is the channel potential at Si/SiO₂ interface, and η is the prefactor for quasi-2D solution [57]. Assuming the total gate charge from the long channel is Q_g (y-independent term); therefore, it can be rewritten as,

$$\left[V_{gb} - V_{FB} - \phi(y) + \alpha \frac{d\phi^2(y)}{dy^2} \right] = \frac{Q_g}{C_{ox}} \quad (2.98)$$

where

$$\alpha = \frac{\epsilon_{si}}{\eta C_{ox}} x_{d\max}. \quad (2.99)$$

The depth boundary can also be replaced with varying depletion depth. However, the maximum depletion width is used as the second boundary in the bulk so that the same form of expression for α can be used in all regions. Physically, all the depletion charges are assumed to be confined within the maximum depletion depth. Nevertheless, two different fitting parameters, namely α_{acc} and α_{ds} , are to be used for accumulation and depletion-to-strong inversion, respectively, due to different amount of charges that are present in the Gaussian box in each region. The partial equation can be rewritten as follows

$$\left[\alpha \frac{d\phi^2(y)}{dy^2} - \phi(y) \right] = -\phi_L \quad (2.100)$$

where ϕ_L is considered as the long-channel surface potential resulting from the solution by replacing $Q_g = C_{ox}(V_{gb} - V_{FB} - \phi_s)$

$$\phi_L = V_{gb} - V_{FB} - \frac{Q_g}{C_{ox}}. \quad (2.101)$$

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When the boundary conditions $\phi(0)=V_{bi}+V_{sb}$ and $\phi(L)=V_{bi}+V_{db}$ are applied at source and drain ends, respectively, the solution to (2.100) is given as [56, 58]

$$\begin{aligned}\phi_s(y) &= \phi_L + \text{sgn}(\phi_L) \delta\phi_s \\ \delta\phi_s &= (V_{bi} + V_{db} - \phi_L) \frac{\sinh(y/\alpha)}{\sinh\left(\frac{L_{eff}}{\alpha}\right)} + (V_{bi} + V_{sb} - \phi_L) \frac{\sinh((L_{eff} - y)/\alpha)}{\sinh\left(\frac{L_{eff}}{\alpha}\right)}\end{aligned}\quad (2.102)$$

which is considered as the y -dependent short-channel surface potential from the quasi-2D solution.

At long channel, the minimum potential is approximately located at $L_{max}/2$, even with V_{db} and V_{sb} variations. Therefore, the changes in potential is approximately given by

$$\begin{aligned}\delta\phi_L &= (V_{bi} + V_{db} - \phi_L) \frac{\sinh(L_{max}/2\alpha)}{\sinh(L_{max}/\alpha)} + (V_{bi} + V_{sb} - \phi_L) \frac{\sinh(L_{max}/2\alpha)}{\sinh(L_{max}/\alpha)} \\ &= \left[2(V_{bi} - \phi_L - V_{bs}) + V_{ds} \right] \frac{\sinh(L_{max}/2\alpha)}{\sinh(L_{max}/\alpha)} \\ &= \frac{\left[2(V_{bi} - \phi_L - V_{bs}) + V_{ds} \right]}{2\cosh(L_{max}/2\alpha)}.\end{aligned}\quad (2.103)$$

It is assumed that this change can be formulated as a change in the Fermi potential, equivalent to the long-channel drain-induced barrier lowering (DIBL). Therefore, (2.102) is rewritten as $\delta\phi_F$ and the Fermi potential with DIBL is given by

$$\phi_{F,dibl} = \phi_F + \delta\phi_F \quad (2.104)$$

$$\delta\phi_F = \alpha_{dibl} \left[2(V_{bi} - 2\phi_F - V_{bs}) + V_{ds} \right] \quad (2.105)$$

$$\alpha_{dibl} = \frac{1}{2\cosh(L_{max}/2\alpha)}. \quad (2.106)$$

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As the channel length continues to decrease and when $V_{ds} \neq 0$, $\phi(y)$ is approximated as [58]

$$\phi(y) = \phi_L + (V_{bi} + V_{db} - \phi_L) e^{(y-L_{eff})/\alpha} + (V_{bi} + V_{sb} - \phi_L) e^{-y/\alpha} + (V_{bi} + V_{db} - \phi_L) e^{-L_{eff}/\alpha}. \quad (2.107)$$

The minimum potential occurs at $y=y_{\min}$. It can be calculated by equating the derivative of (2.107) to zero:

$$\frac{d\phi(y)}{dy} = \frac{(V_{bi} + V_{db} - \phi_L)}{\alpha} e^{(y_{\min}-L)/\alpha} - \frac{(V_{bi} + V_{sb} - \phi_L)}{\alpha} e^{-y_{\min}/\alpha} = 0, \quad (2.108)$$

and it is found to be [13]

$$y_{\min} = \frac{L_{eff}}{2} - \frac{\alpha}{2} z, \quad (2.109)$$

$$z = \begin{cases} z_{acc} = \ln \frac{V_{bi} + V_{db} - \phi_{acc}}{V_{bi} + V_{sb} - \phi_{acc}}; & V_{gb} < V_{FB} \\ z_{ds} = \ln \frac{V_{bi} + V_{db} - 2\phi_F}{V_{bi} + V_{sb} - 2\phi_F}; & V_{gb} > V_{FB} \end{cases}. \quad (2.110)$$

The final URSP solutions with quasi-2D PBL are expressed as

$$\phi_{acc,Q2D} = \phi_{acc} - \delta\phi_s [y_{\min}(z_{acc})] \quad (2.111)$$

and

$$\phi_{ds,Q2D} = \phi_{ds} + \delta\phi_s [y_{\min}(z_{ds})] \quad (2.112)$$

evaluated using (2.102), (2.109), and (2.110) with ϕ_L being replaced by ϕ_{acc} and ϕ_{ds} .

In a real device, implantation steps are used to prevent the source/drain punch-through and threshold voltage roll-off. The doping concentration in the channel is no

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longer uniform. The vertical doping profile can be modeled with transformation of an equivalent effective doping concentration [59-61]. Nevertheless, the dopant pile-up at the source and drain ends of the channel due to halo implantation results in increased effective doping concentration as the gate length decreases. In the literature, lateral non-uniformly doped channel is handled empirically with linear approximation [62], step function [2], and Gaussian profile [63]. It can be used to model the reverse short-channel effect (RSCE).

Two Gaussian profiles are proposed [63] at the source and drain ends with pile-up centroid l_μ , Gaussian peak (N_{pile}), and the characteristic length (l_β), as shown in Figure 2.14.

Figure 2.14: Modeling of lateral doping concentration with two Gaussian profiles.

The effective doping concentration including the lateral Gaussian doping profile is given by

$$N_{eff} = \frac{\sqrt{\pi} N_{pile}}{L_{eff} / l_\beta} \left[\text{erf} \left(\frac{L_{eff} - l_\mu}{l_\beta} \right) + \text{erf} \left(\frac{l_\mu}{l_\beta} \right) \right] + N_{ch} \quad (2.113)$$

where $\text{erf}(\cdot)$ is the error function and the Gaussian peak N_{pile} is defined as

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$$N_{pile} = \kappa N_{ch} \quad (2.114)$$

with κ as a fitting parameter.

2.4. Unified regional drain current model

Most of the existing compact models [2, 4] are developed based on the charge-sheet approximation [20]. It is assumed that the *depletion region under the gate is free of mobile carriers so that the depletion approximation is valid* [26]. The bulk-charge density is therefore approximated as

$$Q_b = -\gamma C_{ox} \sqrt{\phi_s(y)}, \quad (2.115)$$

where $\phi_s(y)$ is position dependent surface potential along the intrinsic channel between source and drain. Therefore, the inversion charge density can be calculated with the charge conservation relating to gate charge, Q_g by

$$\begin{aligned} Q_i &= -Q_g - Q_b \\ &= -C_{ox} (V_{gb} - V_{FB} - \phi_s(y) - \gamma \sqrt{\phi_s(y)}). \end{aligned} \quad (2.116)$$

From current continuity, $I_{ds} = -\mu W Q_i(y) dV/dy$, the drain current can be found by integrating the inversion charge density from source $y=0$ to drain $y=L_{eff}$:

$$\int_0^{L_{eff}} I_{ds} dy = -\mu W \int_0^{V_{ds}} Q_i dV. \quad (2.117)$$

If one performs a change of variable of the surface potential and quasi-Fermi potential while assuming charge-sheet approximation, (2.117) becomes

$$\int_0^{L_{eff}} I_{ds} dy = -\mu W \int (Q_i(y) d\phi(y) - v_{th} dQ_i(y)). \quad (2.118)$$

The integrated drain current is therefore given as [26]

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$$I_{ds} = \beta \left\{ (V_{gb} - V_{FB}) (\phi_{s,d} - \phi_{s,s}) - \frac{1}{2} (\phi_{s,d}^2 - \phi_{s,s}^2) - \frac{2\gamma}{3} (\phi_{s,d}^{3/2} - \phi_{s,s}^{3/2}) \right. \\ \left. + v_{th} \left[\gamma (\phi_{s,d}^{1/2} - \phi_{s,s}^{1/2}) + (\phi_{s,d} - \phi_{s,s}) \right] \right\} \quad (2.119)$$

where $\beta = W\mu C_{ox}/L_{eff}$ is the gain factor, $\phi_{s,d}$ and $\phi_{s,s}$ are surface potentials at drain and source ends, respectively. The first line in (2.119) is usually referred to as the drift-component corresponding to the results of integration for the first (drift) term in (2.118) while the second line is the diffusion-component corresponding to the second (diffusion) term in (2.118). This drain current model is accurate [21], but it also results in a complicated charge model [22]. Therefore, a compromise between complexity and physics is possible by taking a Taylor expansion of (2.119). This has always been done with the V_t -based model, where the Taylor expansion is done on the $\phi_s^{3/2}$ terms for drift current and the $\phi_s^{1/2}$ terms for diffusion current, referenced to source. Nevertheless, this will cause a violation of the Gummel symmetry test (GST) [5]. However, if the Taylor expansion is done on the inversion charge density (2.116) with symmetric bulk-charge linearization [64],

$$Q_i(x) \approx \overline{Q_i} \Big|_{\phi_s = \overline{\phi_s}} + \frac{d\overline{Q_i}}{d\phi_s} \Big|_{\phi_s = \overline{\phi_s}} (\phi_s - \overline{\phi_s}) \\ = -C_{ox} \left(\overline{q_i} - \overline{A_b} (\phi_s - \overline{\phi_s}) \right) \quad (2.120)$$

where

$$\overline{q_i} = V_{gb} - V_{FB} - \overline{\phi_s} - \gamma \sqrt{\overline{\phi_s}}, \quad (2.121)$$

$$\overline{A_b} = 1 + \frac{\gamma}{2\sqrt{\overline{\phi_s}}}, \quad (2.122)$$

in which

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$$\overline{\phi_s} = \frac{\phi_{s,d} + \phi_{s,s}}{2} \quad (2.123)$$

is the average surface potential, the drain current expression will pass the GST. This also results in a simpler drift-diffusion current expression, given by

$$I_{ds} = \beta \left(\overline{q_i} + v_{th} \overline{A_b} \right) \Delta \phi_s \quad (2.124)$$

where

$$\Delta \phi_s = \phi_{s,d} - \phi_{s,s}. \quad (2.125)$$

Both models given by equation (2.119) and equation (2.124) are classified under the surface-potential based model. Both require accurate solution of the surface potential, in particular, in the subthreshold region, due to the fact that I_{ds} is an exponential function of the surface potential difference, $\phi_{s,d} - \phi_{s,s}$. However, $\phi_{s,s}$ in the subthreshold region is almost equal to $\phi_{s,d}$. A small error in the surface potential solution can result in a large error in I_{ds} .

In the URM approach, the drain-current model (2.124) is further approximated with a “pinned surface potential” concept where $\phi_s \approx \phi_{s,s} + V_{cb}$, therefore, it can be rewritten as

$$I_{ds} = \beta \left(\overline{q_i} + v_{th} \overline{A_b} \right) V_{ds}, \quad (2.126)$$

with

$$\Delta \phi_s \approx \left(\phi_{s,s} + V_{db} \right) - \left(\phi_{s,s} + V_{sb} \right) = V_{ds} = V_{ds,eff}, \quad (2.127)$$

in which V_{ds} is to be replaced by the effective drain-source voltage, $V_{ds,eff}$, in current evaluation for correct linear (and strong inversion) to saturation (and subthreshold)

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behaviour. This form of drain current equation does not require accurate surface potential solution, as long as it has the correct subthreshold slope in $V_{ds,eff}$, as shown in Figure 2.15.

Figure 2.15 The drain current as a function of V_{gs} based on URSP drain current model, drift current (I_{drift}), diffusion current (I_{diff}) and combined total current I_{ds} (solid line)

2.4.1. Gummel symmetry test requirement

Source-drain symmetry is a fundamental feature of an ideal MOSFET. Gummel symmetry test (GST) [5] was introduced as a benchmark test to qualify a MOSFET compact model, which requires the drain-source current $I_{ds}(V_{ds})$ to be strictly an odd function of V_{ds} . In the GST, the source and drain terminals are to be

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biased at certain potential plus/minus a varying voltage V_x , and it is required that at least the second-order derivatives of I_{ds} with respect to (w.r.t.) V_x be smoothly passing through $V_x = 0$. Besides the “odd-function” requirement, it is also necessary that $d^{(n)}I_{ds}(0) \equiv d^n I_{ds}(V_x)/dV_x^n|_{V_x \rightarrow 0}$ exist (i.e., no singularity at $V_x = 0$) [34] in order to pass the GST with n th-order derivative. It has been recognized [5] that Gummel symmetry was one of the main showstoppers in the industry *de facto* standard model BSIM3 [2]. Special care has been taken in passing the GST in the next-generation standard model, PSP [4]. However, Gummel symmetry at higher-order derivatives, which is often required for distortion analyses [34, 65], may still be a problem due to use of the mathematical smoothing function for the “effective drain–source voltage” ($V_{ds,eff}$), which has to be traded off for the appropriate range of the smoothing parameter or introduce extra parameter for Gummel symmetry [4] when modeling the geometry-dependent output (drain current and drain conductance) characteristics at short channel. The drain current model has been developed to fulfil the GST requirement without needing any extra parameter and, as an added advantage, it is readily applicable to asymmetry source/drain MOS structure with the bulk-referenced core model.

2.4.2. Effective mobility and transverse field degradation

One of the main sources that cause a model to fail the GST is from the effective mobility expression, which includes transverse and lateral field degradation, as well as the effective transverse field. The transverse field-dependent mobility is

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usually expressed with the well-known Matthiessen's rule for three scattering mechanisms [66],

$$\begin{aligned}\mu_0 &= \left(\frac{1}{\mu_{co}} + \frac{1}{\mu_{ph}} + \frac{1}{\mu_{sr}} \right)^{-1} \\ &= \frac{\mu_1}{1 + \left(\frac{\mu_1}{\mu_2} \right) E_{eff}^{\frac{1}{3}} + \left(\frac{\mu_1}{\mu_3} \right) E_{eff}^{\nu}},\end{aligned}\quad (2.128)$$

in which semi-empirical models for phonon scattering, surface roughness scattering, and Coulombic scattering are related to the effective transverse field, E_{eff} [26] by

$$\mu_{ph} = \frac{\mu_2}{E_{eff}^{1/3}}, \quad (2.129)$$

$$\mu_{sr} = \frac{\mu_3}{E_{eff}^{\nu}}, \quad (2.130)$$

and

$$\mu_{co} = \mu_1, \quad (2.131)$$

respectively. μ_1 is assumed as a constant with respect to the effective transverse field, μ_2 and μ_3 are related to acceptor density and absolute temperature, and ν is (by default) 2 for NMOS and 1 for PMOS. Nevertheless, μ_1 , μ_2 , μ_3 and ν are usually treated as fitting parameters to match the measurement data. The effective transverse field is related to the inversion and bulk charges as [26]

$$E_{eff}(y) = [\zeta_b Q_b(y) + \zeta_n Q_i(y)] / \epsilon_{si}, \quad (2.132)$$

where ζ_n and ζ_b are 0.5 and 1, respectively, for electrons in <100> silicon crystal orientation. Both ζ_n and ζ_b are fitting parameters related to inversion charge and bulk

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charge, respectively. The symmetric bulk-charge linearization approach is used to calculate E_{eff} to ensure fulfillment of the GST, as follows:

$$\begin{aligned}
 E_{eff} &= \frac{1}{\phi_{s,d} - \phi_{s,s}} \int_{\phi_{s,s}}^{\phi_{s,d}} E_{eff}(y) d\phi_s \\
 &= \frac{1}{\Delta\phi\epsilon_{si}} \left[\int_{\phi_{s,s}}^{\phi_{s,d}} \zeta_n \overline{Q_i} d\phi_s + \zeta_b \int_{\phi_{s,s}}^{\phi_{s,d}} \overline{Q_b} d\phi_s \right] \\
 &= \frac{C_{ox}}{\Delta\phi\epsilon_{si}} \left[\int_{\phi_{s,s}}^{\phi_{s,d}} \zeta_n \left(\overline{q_i} - \overline{A_b} (\phi_s - \overline{\phi_s}) \right) d\phi_s + \zeta_b \gamma \int_{\phi_{s,s}}^{\phi_{s,d}} \left(\sqrt{\phi_s} + \frac{1}{2\sqrt{\phi_s}} (\phi_s - \overline{\phi_s}) \right) d\phi_s \right] \\
 &= \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left(\overline{q_i} + \frac{\zeta_b}{\zeta_n} \gamma \sqrt{\overline{\phi_s}} \right) \quad (2.133)
 \end{aligned}$$

The inversion charge density can either be obtained with left-hand-side (LHS) of the input-voltage equation (2.18), or the right-hand-side (RHS) of (2.18), thus yielding

$$Q_i(y) = \gamma C_{ox} V_{gt}(y), \quad (2.134)$$

with

$$V_{gt}(y) = V_{gb} - V_{FB} - \phi_s(y) - \gamma \sqrt{\phi_s(y)} \quad (\text{LHS}) \quad (2.135)$$

$$V_{gt}(y) = \sqrt{\phi_s(y)} + v_{th} e^{\left[\phi_s(y) - 2\phi_F - V_{cb} \right] / v_{th}} - \sqrt{\phi_s(y)} \quad (\text{RHS}). \quad (2.136)$$

It is shown in Figure 2.16 that the RHS of the voltage equation shows the correct subthreshold-slope trend as compared to that using LHS of the equation, which becomes negative below threshold voltage that cannot be plotted on logarithm scale. Using RHS of the voltage equation is the key for the URSP approach without requiring very accurate surface-potential solutions.

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Figure 2.16. Comparison of normalized inversion charge using the Left-hand-side of equation (2.18), as given by (2.135) and Right-hand-side of equation (2.18,) as given by (2.136)

2.4.3. Series resistance

Conventional compact models treat bias-independent series resistances as external to the core model and being solved by the circuit simulator. We adopt the previous work [67] in which the series resistance is incorporated into the core model, with extension for symmetry and inclusion of diffusion current. There are two parts of the series resistance in a MOSFET [67]. The first part is the intrinsic bias-dependent term, R_{int} to model the intrinsic bias-dependent resistance [30],

$$R_{int} = \frac{r}{C_{ox} W \left(V_{gs} - V_t - \frac{A_b V_{ds}}{2} \right)} \quad (2.137)$$

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where r is a fitting parameter. The second part is the extrinsic bias-independent term, R_{ext} , which is mainly caused by the sheet resistance in the source/drain extension,

$$R_{ext} = \frac{2\rho S}{x_j W}, \quad (2.138)$$

where ρ/x_j is the sheet resistance under the LDD spacer of thickness S . R_{int} in [30] is previously derived according to V_t -based approach, which ignored the contribution of diffusion current. The diffusion contribution is included in this work with symmetric bulk-charge linearization approach as

$$R_{int} = \frac{r}{C_{ox} W \left(\overline{V_{gt}} + \overline{A_b v_{th}} \right)}. \quad (2.139)$$

The final source resistance is formulated as

$$R_s = R_{ext} + R_{int} = r_1 + \frac{r_2}{C_{ox} W \left(\overline{V_{gt}} + \overline{A_b v_{th}} \right)} \quad (2.140)$$

where $r_1 = \frac{2\rho S}{x_j W}$ and r_2 are treated as fitting parameters. Drain resistance is assumed

identical to source side in conventional MOSFET structure. Thus, yielding

$$R_s = R_d \quad (2.141)$$

and

$$R_{sd} = R_s + R_d \quad (2.142)$$

where R_{sd} is the sum of the source (R_s) and drain (R_d) series resistance.

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2.4.4. Lateral field mobility degradation

The mobility degradation due to lateral field, $E_y = -d\phi_s/dy$ begins to have more significant role over transverse-field degradation when V_{ds} increases beyond the gradual-channel approximation. At low lateral field (i.e., low V_{ds}), the velocity of carriers in the channel is proportional to E_y and begins to saturate when lateral field is larger than the saturation field, E_{sat} . The two-region piecewise velocity–field relation is [68]

$$v_0 = \frac{\mu_0 E_y}{1 + E_y/E_{sat}} \quad \text{if } E_y < E_{sat} \quad (2.143)$$

and

$$v_0 = v_{sat} \quad \text{if } E_y > E_{sat}, \quad (2.144)$$

with

$$E_{sat} = \frac{v_{sat}}{2\mu_0} \quad (2.145)$$

for electrons. E_y in the denominator is approximately a linear function of V_{ds} and, therefore, the velocity is given as

$$v_0 = \frac{\mu_0 E_y}{1 + \frac{\delta_L V_{ds}}{L_{eff} E_{sat}}} \quad (2.146)$$

and the effective mobility with lateral field degradation is

$$\mu_{eff0} = \frac{\mu_0}{1 + \frac{\delta_L V_{ds}}{L_{eff} E_{sat}}} \quad (2.147)$$

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where δ_L is a fitting parameter for lateral-field mobility degradation. Nevertheless, this function is known to fail the GST due to non-odd function of V_{ds} [69]. We have proposed the average lateral-field mobility following the similar idea to the symmetric bulk-charge linearization [7]

$$\overline{\mu_{eff0}} = \frac{\mu_0}{2} \left(\frac{1}{1 + \frac{\delta_L V_{db}}{L_{eff} E_{sat}}} + \frac{1}{1 + \frac{\delta_L V_{sb}}{L_{eff} E_{sat}}} \right). \quad (2.148)$$

The GST is performed on I_{ds} with the mobility models using (2.147) and (2.148), as shown in Figure 2.17. The average mobility shows a smooth transition even for the seventh-order derivative of V_x , while the previous approach using (2.147) fails the GST.

Figure 2.17: Comparison of Gummel symmetry test on I_{ds} -model with existing lateral mobility degradation with (2.147) and the average lateral mobility degradation with (2.148). The existing approach has a discontinuity at $V_x=0$ at the 7th-order numerical derivatives of I_{ds} while the average mobility is smooth across $V_x=0$.

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2.4.5. Saturation voltage

The saturation voltage is defined as the voltage at which the current reaches its saturated velocity. Unlike conventional compact models in which source and drain are defined by MOS convention, where drain terminal voltage is always higher than source terminal voltage for NMOS, in this work, the source and drain terminals are treated as node labels (designated by layout), in which V_{ds} can be negative and the source/drain terminals are not swapped in model evaluation in a circuit simulator.

The saturation voltage can be determined from the intercept point of the saturation currents according to the V_t -based approach [55]. The saturation current with a Taylor expansion at the source end is given as

$$I_{dsat} = v_{sat} W C_{ox} (V_{gt,s} - A_{b,s} \Delta\phi), \quad (2.149)$$

where $V_{gt,s} = Q_{i,s}/C_{ox}$ is the inversion charge normalized to the oxide capacitance C_{ox} , evaluated at the source end,

$$V_{gt,s} = \gamma \sqrt{\phi_{s,s} + v_{th} e^{\left[\phi_{s,s} - 2\phi_F - V_{sb}\right]/v_{th}}} - \gamma \sqrt{\phi_{s,s}} \quad (2.150)$$

$$A_{b,s} = 1 + \gamma / (2\sqrt{\phi_{s,s}}), \quad (2.151)$$

and drift-diffusion current (Taylor expanded at source end) can be written as

$$I_{dd,s} = \mu_{eff0,s} C_{ox} \frac{W}{L_{eff}} \left(V_{gt,s} - \frac{A_{b,s} \Delta\phi}{2} + A_{b,s} v_{th} \right) \Delta\phi. \quad (2.152)$$

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where

$$\mu_{eff0,s} = \frac{\mu_0}{1 + \frac{\delta_L V_{sb}}{L_{eff} E_{sat}}} \quad (2.153)$$

is (2.147) evaluated at the source end. Solving (2.149) and (2.152) with (2.127), drain saturation voltage is given by

$$V_{ds,sat} = \frac{V_{gt,s} L_{eff} E_{sat}}{V_{gt,s} + A_{b,s} L_{eff} E_{sat} + 2A_{b,s} v_{th}}. \quad (2.154)$$

The onset of saturation actually happens at the drain side (“drain” by label in the layout) in “forward” operation when $V_{db} > V_{sb}$; therefore, from the convention $V_{ds} = V_{db} - V_{sb}$, it is possible to define the “saturation *drain* voltage” w.r.t. bulk as $V_{ds,sat} = V_{db,sat} - V_{sb}$, from which

$$V_{db,sat} = V_{ds,sat} + V_{sb}. \quad (2.155)$$

Similar analysis can be applied at the drain end when $V_{sb} > V_{db}$, in which the surface potential in (2.150) and (2.151) are replaced with the drain-side surface potential:

$$V_{gt,d} = \gamma \sqrt{\phi_{s,d} + v_{th}} e^{\left[\phi_{s,d} - 2\phi_F - V_{db}\right]/v_{th}} - \gamma \sqrt{\phi_{s,d}} \quad (2.156)$$

and

$$A_{b,d} = 1 + \gamma / \left(2\sqrt{\phi_{s,d}} \right). \quad (2.157)$$

The saturation voltage evaluated at the drain end is therefore given as:

$$V_{sd,sat} = \frac{V_{gt,d} L_{eff} E_{sat}}{V_{gt,d} + A_{b,d} L_{eff} E_{sat} + 2A_{b,d} v_{th}}. \quad (2.158)$$

Therefore, saturation source–bulk voltage is

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$$V_{sb,sat} = V_{sd,sat} + V_{db}. \quad (2.159)$$

For short-channel devices, the series-resistance effect is included in the saturation voltage. Similar to [55], the saturation current considering series-resistance can be derived as

$$I_{dsat} = \frac{v_{sat} WC_{ox} (V_{gt,s} - A_{b,s} V_{ds,sat})}{1 - v_{sat} WC_{ox} (A_{b,s} R_{sd} - R_s)} \quad (2.160)$$

where R_s and R_{sd} are given in (2.140) and (2.142), respectively. The linear current in equation (2.152) including R_{sd} and evaluated at $V_{ds} = V_{ds,sat}$, is given by

$$I_{dsat} = \frac{\mu_{eff,0,s} C_{ox} \frac{W}{L_{eff}} \left(V_{gt,s} - \frac{A_{b,s} V_{ds,sat}}{2} + A_{b,s} v_{th} \right) V_{ds,sat}}{1 + \mu_{eff,0,s} C_{ox} \frac{W}{L_{eff}} \left(V_{gt,s} - \frac{A_{b,s} V_{ds,sat}}{2} + A_{b,s} v_{th} \right) R_{sd}}. \quad (2.161)$$

Solving the above two equations, we obtain

$$V_{ds,sat} = \frac{-b_s - \sqrt{b_s^2 - 4a_s c_s}}{2a_s}, \quad (2.162)$$

in which

$$a_s = v_{sat} WC_{ox} A_{b,s} R_s \quad (2.163)$$

$$b_s = - \left[V_{gt,s} + v_{sat} WC_{ox} V_{gt,s} (2R_s + A_{b,s} R_{sd}) + A_{b,s} E_{sat} L_{eff} + 2A_{b,s} v_{th} (1 + v_{sat} WC_{ox} R_s) \right] \quad (2.164)$$

$$c_s = V_{gt,s} E_{sat} L_{eff} + 2v_{sat} WC_{ox} R_{sd} V_{gt,s} (V_{gt,s} + A_{b,s} v_{th}), \quad (2.165)$$

which includes both drift and diffusion currents. Following similar discussions for the drain side, $V_{sd,sat}$ including the R_{sd} effect and drift-diffusion current can be derived and, subsequently, used in (2.162) accordingly. The effective drain–bulk voltage, $V_{db,eff}$ and

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effective source–bulk voltage, $V_{sb,eff}$ are smoothed using the following smoothing function,

$$\vartheta_{eff,\delta_s}\{V_1, V_2, V_{sat}; \delta_s\} \equiv V_{sat} - \frac{1}{2} \left(V_{sat} - V_1 + \sqrt{(V_{sat} - V_1)^2 + 4\delta_s(V_{sat} - V_2 + \delta_s)} \right) \quad (2.166)$$

where

$$V_{db,eff} = \vartheta_{eff,\delta_s}\{V_{db}, V_{sb}, V_{db,sat}; \delta_s\} \quad (2.167)$$

and

$$V_{sb,eff} = \vartheta_{eff,\delta_s}\{V_{sb}, V_{db}, V_{sb,sat}; \delta_s\}. \quad (2.168)$$

The effective drain–source voltage is then given by

$$V_{ds,eff} = V_{db,eff} - V_{sb,eff}. \quad (2.169)$$

Figure 2.18 shows the behaviour of (2.167) to (2.169) at $V_{sb}=0$. In this case, $V_{db,eff}$ is equal to $V_{ds,eff}$ due to zero source bias.

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Figure 2.18: Components of effective drain-source voltage sweeping gate-bulk voltage. $V_{ds,eff}$ is the subtraction of $V_{db,eff}$ and $V_{sb,eff}$. The $V_{sb,eff}$ is however negligible when $V_s=0$. Therefore, $V_{ds,eff}$ is essentially same as $V_{db,eff}$.

Another popular smoothing function used among the compact models is

$$\vartheta_{eff,ax}\{V_{ds}, V_{ds,sat}; a_x\} \equiv \frac{V_{ds}}{\left[1 + (V_{ds}/V_{ds,sat})^{a_x}\right]^{1/a_x}} \quad (2.170)$$

where a_x is the smoothing parameter. When Gummel symmetry at higher-order derivatives becomes important, (2.170) has shown a major limitation since to avoid singularity in the n th-order derivative at $V_x = 0$, a_x has to be larger than n due to the $(V_{ds}/V_{ds,sat})^{a_x-n}$ term in the n th-order derivative, but short-channel devices often require a small value of a_x . Another length-dependent smoothing parameter has to be introduced to tackle this problem [4]. Figure 2.19 shows the GST on the second-, fourth-, and sixth-order derivatives of $V_{ds,eff}$ using (2.169) and (2.170). It can be seen that the model (2.170) fails the GST at n th-order derivative when a_x is smaller than n , while our model (2.169) has passed the GST for all the eventh-order derivatives.

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Figure 2.19: Comparison of the $V_{ds,eff}$ using (2.169) and using (2.170) for the (a) second-, (b) fourth-, and (c) sixth-order derivatives of GST for two smoothing parameters as indicated from the respective models. Inset: GST circuit.

If the similar idea is used for the $\vartheta_{eff,ax}(x)$ function in (2.170) as in (2.167) and (2.168),

$$\begin{aligned} V_{ds,eff} &= V_{db,eff} - V_{sb,eff} \\ &= \vartheta_{eff,ax} \{V_{db}, V_{db,sat}; a_x\} - \vartheta_{eff,ax} \{V_{sb}, V_{sb,sat}; a_x\}, \end{aligned} \quad (2.171)$$

it will solve the singularity problem for this smoothing function. The singularity at the fourth-order derivative has been removed, as shown in Figure 2.20, after applying the bulk-referencing concept. Nevertheless, the smoothing function with square-root function in (2.169) is much more computationally effective than the power function in (2.171).

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Figure 2.20: GST for the fourth-order derivative based on (2.171) applying the concept of bulk-referenced to original(old) approach as in (2.170). The singularity has been removed.

The drain current is rewritten in a unified expression with average lateral field mobility in (2.148) and effective drain-source voltage in (2.169):

$$I_{ds0} = \frac{WC_{ox}}{L_{eff}} \overline{\mu_{eff0}} \left(\overline{q_i} + v_{th} \overline{A_b} \right) V_{ds,eff}. \quad (2.172)$$

2.4.6. Channel length modulation

Channel-length modulation (CLM) and velocity overshoot (VO) have been modeled based on energy-balance formulation [70] through an “effective Early voltage,” which has been modified with the same idea (“bulk-referencing”) in the new

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saturation-voltage formulations to maintain model symmetry [7]. The final short-channel current equation including series resistance and CLM/VO effects is given as

$$I_{ds} = \frac{\overline{g_{vo}} I_{ds0}}{1 + R_{sd} I_{ds0} / V_{ds,eff}} \quad (2.173)$$

where I_{ds0} is the drain current without R_{sd} effect from (2.172).

$$\overline{g_{vo}} = (g_{vo,s} + g_{vo,d}) / 2 \quad (2.174)$$

is the averaged factor for the CLM/VO effect evaluated at the source and drain sides.

The drain-side VO factor is derived as

$$g_{vo,d} = 1 + \left(\frac{V_{db} - V_{db,eff}}{V_{Aeff,d}} \right) \quad (2.175)$$

with the drain side effective Early voltage, $V_{Aeff,d}$ defined by

$$V_{Aeff,d} = \frac{\left(L_{eff} E_{sat} \left[1 + h_d (V_{db} - V_{db,eff}) \right] \right) + \delta_L V_{db,eff}}{\delta_L h_d V_{db,eff}}, \quad (2.176)$$

$$h_d = \frac{\frac{\xi c}{l^2} \left[1 + \sqrt{1 + \left(\frac{V_{db} - V_{db,eff}}{l E_{sat}} \right)^2} \right]}{\sqrt{1 + \left(\frac{V_{db} - V_{db,eff}}{l E_{sat}} \right)^2} - \frac{\xi c}{l^2} (V_{db} - V_{db,eff})} \quad (2.177)$$

where ξ is a fitting parameter. The effective Early voltage is the model including electron-temperature gradient from energy-balance equation [70] expressed in the (effective) form of the Early voltage. The source side of the effective Early voltage can be derived by replacing $V_{db,eff}$ with $V_{sb,eff}$ and V_{db} with V_{sb} accordingly. Numerical data are generated with Medici simulation for verification of the URM I_{ds} model.

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Figure 2.21 shows the second derivative of I_{ds} for long- and short-channel devices, comparing the GST from Medici data and the URSP I_{ds} model (2.173). The model passes the GST smoothly at $V_x=0$. The inset shows that the model is still passing the GST at sixth-order derivative, which indicates passing of the GST for all derivatives. Figure 2.22a and 2.22b show the comparison of the model with Medici long-channel device with changing V_d , V_g , V_s , V_b at long channel. The model confirms correct terminal-bias dependencies.

Figure 2.21: GST ($V_g = 1.2 \text{ V}$, $V_d = V_g + V_x$, $V_s = V_g - V_x$, $V_b = 0$, step size: 0.02 V) for long- and short-channel devices (lines) and compared with Medici numerical data (symbols). The inset shows the sixth-order derivatives of I_{ds} for the long (solid line) and short (dotted line) channel devices.

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Figure 2.22: Comparison of modeled drain current (lines) with Medici numerical data (symbols), transfer characteristics with (a) V_d and V_s variations (at fixed $V_{ds} = 1.2$ V) referenced to bulk, (b) V_b variations (at fixed $V_{ds} = 0.05$ V)

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2.4.7. Extension to asymmetric source/drain modeling

Conventional MOSFET model always assumes source/drain symmetry. But in real devices, this may never really happen due to different angles of ion implantation at source/drain, misalignment of poly-Si gate resulting in different source/drain extensions with poly-Si gate, and other process/layout variations, such as source/drain junctions. There are also devices intentionally designed to have asymmetry source/drain for performance enhancement, for example, in “true-LDD” devices where lightly-doped-drain is indeed only formed at the drain side [71, 72] instead of at both source and drain. Other examples of asymmetric MOSFETs include asymmetric-halo [73] and dual/hetero-material gate [74, 75] MOSFETs.

With separate drain and source saturation voltages identified by label, the model can be evaluated with positive or negative V_{ds} . Internally, the model source/drain terminals are not swapped when $V_{ds} < 0$. Therefore, it has the advantage to be extended to include asymmetric source/drain structures. There are model structural changes if one intends to model source/drain asymmetry, e.g., due to different LDD S/D junction depth and doping. Physically, saturation velocity and series resistance will be different at the source and drain side, resulting in different I_{ds} for positive/negative V_{ds} sweeping.

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In this section, we derive asymmetric I_{ds} equation due to $R_s \neq R_d$, given by two different sets of (r_1, r_2) parameters:

$$R_s = r_{1s} + \frac{r_{2s}}{C_{ox} W \left(\overline{V_{gt}} + \overline{A_b} v_{th} \right)} \quad (2.178)$$

$$R_d = r_{1d} + \frac{r_{2d}}{C_{ox} W \left(\overline{V_{gt}} + \overline{A_b} v_{th} \right)}. \quad (2.179)$$

Using the notation (“prime”) $\phi'_s(y)$ to denote surface potential with R_{sd} effect, we have

$$\Delta \phi'_s = \phi'_{s,d}(V'_d) - \phi'_{s,s}(V'_s) = \Delta \phi_s - I_{ds}(R_s + R_d), \text{ where } \Delta \phi_s = \phi_{s,d}(V_d) - \phi_{s,s}(V_s) \text{ when } R_s = R_d.$$

The effective surface potentials at drain and source, after considering the voltage drops on R_d and R_s , respectively, are given by

$$\phi'_{s,d} = \phi_{s,d} - I_{ds} R_d \quad (2.180)$$

$$\phi'_{s,s} = \phi_{s,s} + I_{ds} R_s \quad (2.181)$$

Therefore, all the surface potential terms are modified to include R_{sd} (i.e., due to intrinsic voltage $V'_{ds} = V'_d - V'_s$):

$$q'_i = V_{gb} - V_{FB} - \phi'_s - \gamma \sqrt{\phi'_s} \quad (2.182)$$

$$q'_i \approx \overline{q'_i} - \overline{A_b}' (\phi'_s - \overline{\phi'_s}) \quad (2.183)$$

where

$$\overline{q'_i} = V_{gb} - V_{FB} - \overline{\phi'_s} - \gamma \sqrt{\overline{\phi'_s}} \quad (2.184)$$

$$\overline{A_b}' = 1 + \frac{\gamma}{2\sqrt{\overline{\phi'_s}}} \quad (2.185)$$

in which

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$$\begin{aligned}\overline{\phi_s}' &= \frac{\phi_{s,d}' + \phi_{s,s}'}{2} \\ &= \frac{\phi_{s,d} - I_{ds}R_d + \phi_{s,s} + I_{ds}R_s}{2} \\ &= \overline{\phi_s} - I_{ds} \frac{\Delta R}{2}\end{aligned}\quad (2.186)$$

and

$$\Delta R = R_d - R_s. \quad (2.187)$$

Substituting (2.180) and (2.181) into (2.184),

$$\overline{q_i}' = V_{gb} - V_{FB} - \overline{\phi_s} + I_{ds} \frac{\Delta R}{2} - \gamma \sqrt{\overline{\phi_s} - I_{ds} \frac{\Delta R}{2}}. \quad (2.188)$$

It is then Taylor-expanded around the average surface potential up to the second order:

$$\overline{q_i}' \approx V_{gb} - V_{FB} - \overline{\phi_s} + I_{ds} \frac{\Delta R}{2} - \gamma \sqrt{\overline{\phi_s}} \left(1 - \frac{I_{ds} \Delta R}{4 \overline{\phi_s}} \right) \quad (2.189)$$

The final form of the average normalized inversion charge with series resistance is given by

$$\overline{q_i}' = \overline{q_i} + \overline{A_b} \frac{I_{ds} \Delta R}{2}, \quad (2.190)$$

and the body factor after similar Taylor expansion is

$$\overline{A_b}' \approx \overline{A_b} + \frac{\gamma}{4 \overline{\phi_s}^{3/2}} \frac{I_{ds} \Delta R}{2}. \quad (2.191)$$

The integrated drain current is

$$I_{ds} = \overline{\beta} (\overline{q_i}' + \overline{A_b}' v_{th}) \Delta \phi' \quad (2.192)$$

where

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$$\bar{\beta} = \frac{WC_{ox}\overline{\mu_{eff0}}}{L_{eff}}, \quad (2.193)$$

and after replacing (2.190) and (2.192),

$$I_{ds} \approx \bar{\beta} \left(\bar{q}_i + \bar{A}_b \frac{I_{ds} \Delta R}{2} + v_{th} \bar{A}_b + v_{th} \frac{\gamma}{4\phi_s^{3/2}} \frac{I_{ds} \Delta R}{2} \right) (\Delta\phi_s - I_{ds} R_{sd}) \quad (2.194)$$

which is a quadratic equation with the following solution:

$$I_{ds} = \frac{-b + \sqrt{b^2 - 4ac}}{2a} \quad (2.195)$$

where

$$\begin{aligned} a &= \bar{\beta} \left(\bar{A}_b + v_{th} \frac{\gamma}{4\phi_s^{3/2}} \right) \frac{\Delta R}{2} R_{sd} \\ b &= \left(1 + \bar{\beta} (\bar{q}_i + \bar{A}_b v_{th}) R_{sd} - \bar{\beta} \left(\bar{A}_b + v_{th} \frac{\gamma}{4\phi_s^{3/2}} \right) \frac{\Delta R}{2} \Delta\phi \right) \\ c &= -\bar{\beta} (\bar{q}_i + \bar{A}_b v_{th}) \Delta\phi \end{aligned} \quad (2.196)$$

If $\Delta R=0$, then it is reduced to the conventional symmetric source/drain resistance

($R_s=R_d$) MOSFET equation with

$$I_{ds} = -\frac{c}{b} = \frac{\bar{\beta} (\bar{q}_i + \bar{A}_b v_{th}) \Delta\phi}{1 + \bar{\beta} (\bar{q}_i + \bar{A}_b v_{th}) R_{sd}}, \quad (2.197)$$

where $\Delta\phi$ is to be replaced with $V_{ds,eff}$.

Figure 2.23 shows an example of model matching with Medici data of an asymmetric device, using the drain current ($|I_{ds}|$) with asymmetric source/drain junction depths ($X_{j,s} = 0.8$ nm, $X_{j,d} = 0.3$ nm) and doping concentrations ($N_{D,s} = 1 \times 10^{19}$ cm⁻³, $N_{D,d} = 1 \times 10^{18}$ cm⁻³), which are two physical parameters in the model. The output

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resistance is also shown in the inset, together with the schematic of the asymmetric device being modeled. This model extension is not feasible with a model that depends on terminal voltage swapping for negative V_{ds} , for which two complete sets for model parameters have to be extracted for forward/reverse operations.

Figure 2.23: Modeled drain current (lines) of an asymmetric MOSFET with different source/drain junction depth and doping with positive and negative drain-source voltage sweeping, and compared with the same numerical device (symbols). The inset shows the corresponding output resistance and the schematic of the asymmetric nMOS.

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2.5. Unified regional charge model

The dynamic behaviour of a MOSFET is due to the small-signal capacitive effects, which depends on the physical modeling of the charges in intrinsic (channel) and extrinsic (source/drain overlaps).

2.5.1. Intrinsic long channel charge model

When the gate length is very long (channel length $\gg$ source/drain overlap length), the extrinsic charges are small and negligible compared to the charges in the intrinsic channel. In accumulation, the dominant charge is hole, therefore, according to the unified regional approach, the induced charge can be approximated with the LHS of the voltage equation as,

$$Q_{b,ACC} \approx Q_{sc} = -C_{ox} [V_{GBR} - \phi_{ACC}]. \quad (2.198)$$

in which $Q_{b,ACC}$ gives the correct asymptotic charge solution in accumulation and zero bulk charge in other regions. Similarly, the depletion charge can be formulated using the depletion approximation,

$$Q_{b,SUB} \approx Q_{sc} = -C_{ox} [V_{GBA} - \phi_{SUB}] \quad (2.199)$$

where

$$V_{GBA} = \vartheta_f \{V_{gb} - V_{FB}; \sigma_a\} \quad (2.200)$$

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is the same forward smoothing function as V_{GBF} in (2.40) but with the same parameter σ_a as in V_{GBR} . With the charge balance equation in the depletion region,

$$V_{GBF} - \phi_{SUB} = \gamma \sqrt{\phi_{SUB}}, \quad (2.201)$$

bulk charge in depletion is as [30]:

$$\begin{aligned} Q_{b,SUB} &= -C_{ox} [V_{GBA} - V_{GBF} + V_{GBF} - \phi_{SUB}] \\ &= -C_{ox} [V_{GBA} - V_{GBF} + \gamma \sqrt{\phi_{SUB}}] \end{aligned} \quad (2.202)$$

When the device charge dependence on V_{ds} variations is to be taken into account, the total charge needs to be integrated across the channel. Following the symmetric bulk-charge linearization [64] for the inversion charge, Q_i in (2.120) and integrating (2.118) from source to any y in the channel, one obtains

$$y = \frac{W\mu_0 C_{ox}}{I_{ds}} \left(\bar{q}_i (\phi_s - \phi_{s,s}) - \bar{A}_b \left(\frac{\phi_s^2 - \phi_{s,s}^2}{2} - \bar{\phi}_s (\phi_s - \phi_{s,s}) \right) + v_{th} \bar{A}_b (\phi_s - \phi_{s,s}) \right), \quad (2.203)$$

which can be linearized as

$$y \approx y(\bar{\phi}_s) + \frac{dy}{d\bar{\phi}_s} \bigg|_{\bar{\phi}_s = \bar{\phi}_s} (\phi_s - \bar{\phi}_s) + \frac{d^2 y}{d\bar{\phi}_s^2} \bigg|_{\bar{\phi}_s = \bar{\phi}_s} \frac{(\phi_s - \bar{\phi}_s)^2}{2}, \quad (2.204)$$

using (2.124) for I_{ds} , $y(\bar{\phi}_s)$ is given by

$$\begin{aligned} y(\bar{\phi}_s) &= \frac{L_{eff}}{(\bar{q}_i + v_{th} \bar{A}_b) \Delta \phi} \left((\bar{q}_i + v_{th} \bar{A}_b) (\bar{\phi}_s - \phi_{s,s}) - \bar{A}_b \left(\frac{\bar{\phi}_s^2 - \phi_{s,s}^2}{2} - \bar{\phi}_s (\bar{\phi}_s - \phi_{s,s}) \right) \right) \\ &= \frac{L_{eff}}{2} \left(1 + \frac{\Delta \phi}{4H} \right), \end{aligned} \quad (2.205)$$

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$$H = \frac{\bar{q}_i}{A_b} + v_{th}. \quad (2.206)$$

The first derivative is

$$\frac{dy}{d\phi_s} = \frac{L_{eff}}{(\bar{q}_i + v_{th} \bar{A}_b) \Delta\phi} (\bar{q}_i + v_{th} \bar{A}_b - \bar{A}_b (\phi_s - \bar{\phi}_s))$$

$$\left. \frac{dy}{d\phi_s} \right|_{\phi_s = \bar{\phi}_s} = \frac{L_{eff}}{\Delta\phi}, \quad (2.207)$$

and the second derivative is

$$\frac{d^2 y}{d\phi_s^2} = -\bar{A}_b \frac{L_{eff}}{(\bar{q}_i + v_{th} \bar{A}_b) \Delta\phi}$$

$$\left. \frac{d^2 y}{d\phi_s^2} \right|_{\phi_s = \bar{\phi}_s} = -\frac{L_{eff}}{\Delta\phi H}. \quad (2.208)$$

Introducing an intermediate variable $u = \phi_s - \bar{\phi}_s$, (2.204) becomes

$$y \approx \frac{L_{eff}}{2} \left(1 + \frac{\Delta\phi}{4H} \right) + \frac{L_{eff}}{\Delta\phi} \left(u - \frac{u^2}{2H} \right). \quad (2.209)$$

Taking its derivative, it gives

$$\frac{dy}{du} = \frac{L_{eff}}{\Delta\phi} \left(1 - \frac{u}{H} \right). \quad (2.210)$$

The integrated terminal bulk charge in strong inversion can be calculated,

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$$\begin{aligned}
 Q_B &= W \int_0^{L_{eff}} Q_b(y) dy \\
 &= -WC_{ox} \int_{u_s}^{u_D} \left(\gamma \sqrt{\phi_s} + (\bar{A}_b - 1)u \right) \frac{dy}{du} du \\
 &= -\frac{WC_{ox} L_{eff}}{\Delta\phi} \int_{u_s}^{u_D} \left(\gamma \sqrt{\phi_s} + (\bar{A}_b - 1)u \right) \left(1 - \frac{u}{H} \right) du \\
 &= -WC_{ox} L_{eff} \left(\gamma \sqrt{\phi_s} - \frac{(\bar{A}_b - 1) \Delta\phi_s^2}{12H} \right). \tag{2.211}
 \end{aligned}$$

The above derivations are based on [64], which is valid above V_{FB} but cannot be extended to below V_{FB} . In the URM approach, we extend (2.211) to below V_{FB} by replacing ϕ_s with ϕ_{DS} (from depletion to strong inversion), which approaches (2.202) in depletion:

$$Q_{B,DS} = -WC_{ox} L_{eff} \left(V_{GBA} - V_{GBF} + \gamma \sqrt{\phi_{DS}} - \frac{(\bar{A}_b - 1) \Delta\phi_{DS}^2}{12H} \right). \tag{2.212}$$

Then, the unified bulk-charge model for all regions can be written as

$$\begin{aligned}
 Q_B &= Q_{B,ACC} + Q_{B,DS} \\
 &= -WC_{ox} L_{eff} \left(V_{GBR} - \phi_{ACC} + V_{GBA} - V_{GBF} + \gamma \sqrt{\phi_{DS}} - \frac{(\bar{A}_b - 1) \Delta\phi_{DS}^2}{12H} \right) \tag{2.213}
 \end{aligned}$$

where

$$\overline{\phi_{DS}} = \frac{\phi_{DS,s} + \phi_{DS,d}}{2} \tag{2.214}$$

is the average depletion-to-strong inversion unified surface potential solution and

$$\Delta\phi_{DS} = \phi_{DS,d} - \phi_{DS,s}. \tag{2.215}$$

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The inversion charge can be partitioned into source and drain sides with the Ward and Dutton partition scheme [76, 77]:

$$Q_Y = W \int_0^L f_Y Q_i(y) dy \quad \begin{cases} Y = D : f_D = \frac{y}{L} \\ Y = S : f_S = 1 - \frac{y}{L} \end{cases} \quad (2.216)$$

Following similar steps to the integrated bulk-charge formulations and [64], the integrated unified regional drain charge in strong inversion is

$$Q_D = -\frac{WC_{ox}L_{eff}}{2} \left(\frac{-}{q_i} - \frac{\bar{A}_b \Delta \phi_{DS}}{6} \left(1 - \frac{\Delta \phi_{DS}}{2H} - \frac{\Delta \phi_{DS}^2}{20H^2} \right) \right) \quad (2.217)$$

and the integrated unified regional source charge is

$$Q_S = -\frac{WC_{ox}L_{eff}}{2} \left(\frac{-}{q_i} + \frac{\bar{A}_b \Delta \phi_{DS}}{6} \left(1 + \frac{\Delta \phi_{DS}}{2H} - \frac{\Delta \phi_{DS}^2}{20H^2} \right) \right). \quad (2.218)$$

From the transcapacitance definition:

$$C_{ij} = \delta_{ij} \frac{\partial Q_i}{\partial V_j}, \quad \delta_{ij} = \begin{cases} 1 & (i = j) \\ -1 & (i \neq j) \end{cases} \quad (2.219)$$

four capacitances are calculated from the terminal charge equations, and validated with Medici data, as shown in Figure 2.24.

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Figure 2.24: Direct playback of normalized transcapacitances at $V_{ds} = 0.6$ V (left axis) and compared with numerical data, with the smoothness in second order derivative of gate charge (right axis).

Poly-Si gate doping effects, namely, the PAE and PDE, are to be included in the charge model in a decoupled way as described in section 2.3.3. The surface potentials without poly-Si gate doping effect are replaced with the surface potentials with poly-Si gate doping effect. The terminal charges given in (2.213), (2.217), and (2.218) are to be modified to include the poly-Si gate doping effect with the inclusion of ϕ_p by replacing $V_{gb}-V_{FB}$ with $V_{gb}-V_{FB}-\phi_p$ in ϕ_s -related smoothing functions. Besides, the equation given in (2.215) needs to be modified to

$$\Delta\phi_{ds} = \phi_{DS,d} - \phi_{DS,s} + \Delta\phi_{p,ds}, \quad (2.220)$$

where $\Delta\phi_{p,ds}$ is given by

$$\Delta\phi_{p,ds} = \phi_{p,ds} \Big|_{\phi_s=\phi_{ds,d}} - \phi_{p,ds} \Big|_{\phi_s=\phi_{ds,s}}, \quad (2.221)$$

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which includes the effect in ϕ_p for non-zero V_{ds} .

The model play-back of gate capacitance with different poly-Si gate doping concentrations is shown in Figure 2.25.

Figure 2.25: Model play-back of gate capacitance with different poly-Si gate doping concentration.

Quantum mechanical correction in charge is included in the surface potential as shown in section 2.3.4. Similar approach to adding the poly-Si gate doping effect in all the terminal charges, the surface potentials are replaced with the surface potentials with quantum effects, thus yielding

$$\phi_{acc} \rightarrow \phi_{acc}^{qm} \quad (2.222)$$

$$\phi_{ds} \rightarrow \phi_{ds}^{qm} \quad (2.223)$$

$$\phi_{p,acc} \rightarrow \phi_{p,acc}^{qm} \quad (2.224)$$

$$\phi_{p,ds} \rightarrow \phi_{p,ds}^{qm} \quad (2.225)$$

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Therefore, applying (2.222) to (2.225) into the bulk, drain and source charges given in (2.213) to (2.218), the terminal charge models are now included with both poly-Si gate doping and quantum mechanical effects. The total gate charge is solved with charge conservation

$$Q_g^{qm} = -(Q_b^{qm} + Q_d^{qm} + Q_s^{qm} + Q_{OX}) \quad (2.226)$$

where Q_{OX} is the fixed oxide charge. Medici data are generated for model verification.

Figure 2.26 shows the comparison of gate capacitance, C_{gg} with different quantum-mechanical constant, κ^{qm} . The coupled poly-Si gate doping effect and QME are shown in Figure 2.27, in which the poly-Si gate doping concentration is low, $N_{gate}=5 \times 10^{19} \text{ cm}^{-3}$. Figure 2.28 shows the intrinsic channel gate capacitance including poly-Si gate effect, QME, drain and bulk bias variations.

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Figure 2.26: Model play-back with different QME parameters, κ^{qm} .

Figure 2.27 Coupled QME and poly-gate effect.

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Figure 2.28 Charge model including both quantum mechanical correction and poly gate effect for different bulk and drain-source biases.

2.5.2. Intrinsic short channel charge model

The key to charge-model scalability is the correct modeling of intrinsic charges. Intrinsic-charge model needs to be scalable with decreasing length; to do this it must include the channel dependent effects, i.e., short-channel effects and/or lateral- field effects.

At short channel, when $V_{ds}=0$, two short-channel effects, namely, potential-barrier lowering (PBL) and bulk-charge sharing (BCS), have been included as a

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modification to the surface potential, as described in section 2.3.5. Both BCS and quasi-2D PBL will automatically modify the charges when the channel is short. Nevertheless, at $V_{ds} \neq 0$, the drain current has included the lateral mobility degradation and series resistance, which has modified the form of current expression. In this section, the short-channel charges neglecting poly-Si gate and quantum-mechanical effects are discussed.

The drain current without CLM/VO is given as

$$I_{ds} = \frac{WC_{ox}\mu_0(\bar{q}_i + v_{th}\bar{A}_b)\Delta\phi}{L_{eff} + \frac{\delta_L\Delta\phi}{E_{sat}} + R_{sd}WC_{ox}\mu_0(\bar{q}_i + v_{th}\bar{A}_b)} . \quad (2.227)$$

The y-dependency from drain current in which series resistance is also expressed as a ratio of $\Delta\phi$, without which it can end up with the Ward and Dutton [76] partition for the V_{ds} -independent subthreshold inversion charge, is given by

$$y = \frac{L_{eff}\left(1 + \frac{\delta_L\Delta\phi}{L_{eff}E_{sat}} + \frac{R}{L_{eff}}\right)}{\left(\bar{q}_i + v_{th}\bar{A}_b\right)\Delta\phi} \left(\left(\bar{q}_i + v_{th}\bar{A}_b\right)(\phi - \phi_{s,s}) - \bar{A}_b\left(\frac{\phi_s^2 - \phi_{s,s}^2}{2} - \bar{\phi}_s(\phi_s - \phi_{s,s})\right) \right) - \frac{\delta_L(\phi_s - \phi_{s,s})}{E_{sat}} - \frac{R(\phi_s - \phi_{s,s})}{\Delta\phi} \quad (2.228)$$

where

$$R = R_{sd}WC_{ox}\mu_0(\bar{q}_i + v_{th}\bar{A}_b). \quad (2.229)$$

After linearization, it becomes

$$y \approx \frac{L_{eff}}{2} \left(1 + \frac{\Delta\phi}{4H'}\right) + \frac{L_{eff}}{\Delta\phi} \left(u - \frac{u^2}{2H'}\right) \quad (2.230)$$

where

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$$H' = \frac{(\bar{q}_i + v_{th} \bar{A}_b)}{\bar{A}_b \left(1 + \frac{\delta_L \Delta \phi}{L_{eff} E_{sat}} + \frac{R_{sd} W C_{ox} \mu_0 (\bar{q}_i + v_{th} \bar{A}_b)}{L_{eff}} \right)} \quad (2.231)$$

which automatically reduces to (2.206) at long channel. It can be further transformed to

$$H' = \frac{(\bar{q}_i + v_{th} \bar{A}_b)}{\bar{A}_b \left(1 + \frac{\delta_L (V_{db,eff} + V_{sb,eff})}{2L_{eff} E_{sat}} + \frac{R_{sd} W C_{ox} \mu_0 (\bar{q}_i + v_{th} \bar{A}_b)}{L_{eff}} \right)} \quad (2.232)$$

to meet the Gummel symmetry requirement.

The final forms of short-channel charges with lateral-field mobility reduction and series resistance are similar to long-channel charges, but with the replacement of $H \rightarrow H'$.

2.5.3. Extrinsic short channel charge model

At short channel, the extrinsic capacitance is on the same order of magnitude as the intrinsic capacitance. It must be included in the MOSFET charge model for the correct short-channel dynamic model behaviour [78, 79].

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The bias-dependent overlap charge can be derived with two parallel N-type MOS. At the source side, integration of the Poisson equation neglecting minority electrons yields the following equation for the surface potential ϕ_{sov} ,

$$V_{gs} - V_{FB,ov} - \phi_{sov} = \gamma_{ov} \sqrt{v_{th} \left(\exp \frac{\phi_{sov}}{v_{th}} - 1 \right) - \phi_{sov}} \quad (2.233)$$

where $V_{FB,ov}$ is the flat-band voltage in the overlap region, $\gamma_{ov} = (2q\epsilon_{Si}N_{ov})^{1/2}/C_{ox}$ is the body factor of the overlap region. The overlap regional surface potential can be solved with unified regional approach described in section 2.3.2.

The source-side overlap region bulk charge in accumulation and depletion are modeled through the unified regional surface potentials in accumulation ($\phi_{ov,s,ACC}$) and depletion ($\phi_{ov,s,SUB}$):

$$\begin{aligned} Q_{Sov,acc} &= -WL_{ov}C_{ox} (V_{GSF} - \phi_{ov,s,ACC}) \\ Q_{Sov,sub} &= -WL_{ov}C_{ox} (V_{GSR} - \phi_{ov,s,SUB}) \end{aligned} \quad (2.234)$$

where $Q_{S,OV} = Q_{Sov,acc} + Q_{Sov,sub}$ is the total bulk charge at the source side of the overlap region of length L_{ov} , V_{gsf} and V_{gsr} are the forward and reverse interpolation functions, respectively. Figure 2.29 shows the comparison of iterative overlap surface potential and overlap unified regional surface potential. The first derivative is included to show its smooth transition at overlap flat-band voltage.

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Figure 2.29: Comparison of overlap region surface potential using unified regional surface potential approach and iterative surface potential. Inset: The first-order derivative showing the smooth transition after unification.

Similar to the intrinsic channel region, the overlap region also has charge sharing effect. Therefore, the BCS model is extended to the overlap region charge modeling. The source-side body factor is modified to

$$\gamma_{ovscs} = \gamma_{ov} \left(1 - \frac{\lambda_{ov}}{L_{ov}} \left(\frac{X_{dov,s}}{2} \right) \right) \quad (2.235)$$

where $X_{dov,s}$ is the source-side overlap region depletion depth. For the drain-side overlap charge, similar analysis is applied by replacing V_{sb} by V_{db} .

The extrinsic fringing capacitance models [78, 80] are formulated as a function of the inversion charge,

$$Q_{I,FR} = -v_i \frac{Q_I}{L_g C_{ox}} (C_{side} + C_{bottom}) \quad (2.236)$$

and bulk charge

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$$Q_{B,FR} = -v_b \frac{Q_B}{L_g C_{ox}} (C_{side} + C_{bottom}) \quad (2.237)$$

where

Q_I is the long-channel inversion charge ($=Q_S+Q_D$), Q_B is the long-channel bulk charge,

$$C_{side} = \frac{\epsilon_{ox}}{\pi} \ln \left(2 \left(1 + \frac{t_p}{t_{ox}} \right) \sqrt{\left(\left(1 + \frac{t_p}{t_{ox}} \right)^2 - 1 \right)} + 2 \left(1 + \frac{t_p}{t_{ox}} \right) - 1 \right), \quad (2.238)$$

$$C_{bottom} = \frac{\epsilon_{ox}}{\pi} \ln \left(2 - \ln 4 - \ln \left[1 - 2 \exp \left(-2 \left(1 + \frac{\pi L_g}{2 t_{ox}} \right) \right) \right] \right)$$

t_p is the gate thickness, L_g is the gate length, v_i and v_b are fitting parameters for inversion fringing charge and bulk fringing charge, respectively.

Complete final charges with all the short-channel effects are written as

$$\begin{aligned} Q_{DSC} &= Q_D \{ \gamma_{eff}, \phi_{ds}^{qm} + \delta\phi_{ds}^{qm}, H^{qm} \} + Q_{D,OV} + Q_{D,Fr} \\ Q_{SSC} &= Q_S \{ \gamma_{eff}, \phi_{ds}^{qm} + \delta\phi_{ds}^{qm}, H^{qm} \} + Q_{S,OV} + Q_{S,Fr} \\ Q_{BSC} &= Q_B \{ \gamma_{eff}, \phi_s^{qm} + \text{sgn}(\phi_s^{qm}) \delta\phi_s^{qm}, H^{qm} \} + Q_{B,Fr} \\ Q_{GSC} &= -(Q_{DSC} + Q_{SSC} + Q_{BSC} + Q_{OX}) \end{aligned} \quad (2.239)$$

where γ_{eff} represents the bulk-charge sharing effects, small case potentials related terms represent the poly-gate effect, superscript 'qm' represents the quantum-mechanical correction, $\delta\phi_{ds}$ represents the potential-barrier lowering effect, H represents the lateral-field mobility degradation and series resistance effects, $Q_{D,OV}$ and $Q_{S,OV}$ are the overlap charges, and $Q_{D,Fr}$, $Q_{S,Fr}$, and $Q_{B,Fr}$ are the fringing charges for drain, source, and bulk terminals, respectively. The final terminal capacitances are numerically calculated from (2.219).

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The model behaviour of intrinsic short-channel charge is included in Figure 2.30 and extrinsic short-channel charge is included in Figure 2.31. From these two plots, each individual effect of the model can be observed, by turning off each of the effect one at a time, thus providing some information on short-channel charge model calibration and extraction.

Figure 2.30: Comparison of normalized Medici numerical data $|C_{bg}|$ and $|C_{sg}+C_{dg}|$ with complete model model-Xsim (solid) and turning off the intrinsic effects individually. Xsim model without PBL at accumulation (long dashed), without PBL at depletion-to-strong inversion (medium dashed), without BCS (short dashed).

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Figure 2.31: Comparison of normalized Medici numerical data $|C_{bg}|$ and $|C_{sg}+C_{dg}|$ with complete model-Xsim (solid) and turning off the extrinsic effects individually. Xsim model without overlap capacitances (long dashed), and without BCS in overlap capacitances (medium dashed), without bulk charge extrinsic fringing (short dashed) and without inversion charge extrinsic fringing (dotted).

2.6. Chapter 2 conclusion

In this chapter, the background and the development of the compact model for bulk MOSFET has been discussed. The current trend is moving toward a more physically-based compact model, i.e., surface-potential based model. Therefore, unified regional modeling approach is to introduce a modeling approach that is scalable, simple yet good enough to capture most of the essential physics of MOSFETs. Unified regional surface potential solutions are physically derived in each region and combined with smoothing functions. It is a symmetric model, fulfilling

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most of the qualities of an industry standard model and easily extendable. Other secondary effects, such as poly-Si gate effect, quantum-mechanical effect, short-channel effect, and asymmetric source/drain effect, have been included to demonstrate the physical effects and ease of extension. The next chapter will further explore the unified regional modeling approach for a common asymmetry double-gate MOSFET.

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CHAPTER 3: Double-gate MOSFET Model with Unified Regional Model

Conventional bulk-MOSFET is fast approaching its scaling limit [81]. In order to continue technology scaling according to Moore's law, non-classical alternative device structures with multiple gates like FinFETs and gate-all-around (GAA) Si nanowires have been studied [82-85].

Figure 3.1: Multiple gates MOSFET: a) FinFET and (b) Gate all around with twin silicon nanowire [84].

The fundamental building block of these devices is actually the double-gate. These devices exhibit better electrostatic control as compared to bulk-MOSFET. DG MOSFETs have better subthreshold slope, improved short-channel effects and higher drive current. Nevertheless, there are still quite a number of challenges regarding the multiple-gate technology [85]. Similarly for compact modeling, the coupling effects of two gates that need to be considered in the presence of holes, donors, acceptors and electrons are very challenging tasks for compact modellers.

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An asymmetric-DG MOSFET structure is depicted in Figure 3.2, in which gate asymmetry can arise due to differences in gate oxide thickness and dielectric material, gate material (workfunction) and oxide/interface charges, and the applied biases on the two gates.

Figure 3.2 : Schematic of an asymmetric DG NMOS

The electrostatic potential that obeys the Poisson equation, considering holes, body dopings and electrons, with the applied boundary conditions at two oxide/silicon surfaces cannot be solved easily as it has two unknown surface potentials (ϕ_{s1} and ϕ_{s2}). It is, however, possible to solve the DG-MOSFET regionally [86, 87]. In this work, compact model of a special case of a-DG will be considered, namely, common gate a-DG MOSFET.

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3.1. Common gate asymmetric double-gate MOSFET model

Common gate a-DG (ca-DG) is a DG MOSFET, in which both gates are tied together and, thus, sharing the same gate voltage bias. The example of such a structure is the FinFET [83]. Due to its common gate configuration, it is likely to have a zero-field, E_0 location inside the body [88, 89]. E_0 in a ca-DG, however, may not locate at the centre of the Si film due to asymmetry, which may arise from differences in gate material, oxide thickness, and oxide dielectric in the two gates.

3.2. Ca-DG unified regional surface potential

Applying the Gauss' law at one of the gates and at the zero-field location, the boundary conditions are therefore

$$\begin{aligned}
 \phi(0) &= \phi_{s1} \\
 \phi'(0) &= -\frac{C_{ox1}}{\epsilon_{si}}(V_{gs1} - V_{FB1} - \phi_{s1}) \\
 \phi(x_{01}) &= \phi_{01} \\
 \phi'(x_{01}) &= 0
 \end{aligned} \tag{3.1}$$

where x_{01} is the zero-field location and ϕ_{01} is the potential at that location corresponding to gate 1. Similar set of equations can be written by changing the subscript from 1 to 2 for the second gate. Therefore, without loss of generality, the subscript 1 is dropped for simplicity in the subsequent potential derivations.

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Integrating the Poisson equation once and applying the boundary conditions, it is possible to express the input voltage equation as

$$V_{gs} - V_{FB} - \phi_s = \gamma \sqrt{f_\phi}$$

$$f_\phi = v_{th} \left(e^{-\phi_s/v_{th}} - e^{-\phi_0/v_{th}} \right) + (\phi_s - \phi_0) + e^{-(2\phi_F + V)/v_{th}} \left(v_{th} \left(e^{\phi_s/v_{th}} - e^{\phi_0/v_{th}} \right) - (\phi_s - \phi_0) \right). \quad (3.2)$$

There is no simple solution to this implicit equation [87]. But, regional solutions are possible [14].

If only hole is considered in accumulation region, the starting Poisson's equation is given by

$$\frac{d^2 \phi}{dx^2} = -\frac{qN_{ch}}{\epsilon_{si}} \left(e^{-\phi/v_{th}} \right). \quad (3.3)$$

The analytical solution of the differential equation is similar to the one considering only the electrons [90], with appropriate change of sign and applying the aforementioned boundary conditions, the Poisson solution for holes is given by

$$\phi(x) = 2v_{th} \ln \left(\cos \left(\arccos \left(\sqrt{\frac{B_{cc}}{A_{cc}}} e^{\phi_{cc}/2v_t} \right) \pm \frac{x \sqrt{B_{cc}}}{2v_t} \right) \right) - v_{th} \ln \frac{B_{cc}}{A_{cc}} \quad (3.4)$$

where

$$B_{cc} = A_{cc} e^{\frac{-\phi_{cc}}{v_{th}}} - (\phi'(0))^2 \quad (3.5)$$

and

$$A_{cc} = \frac{2qN_{ch}v_{th}}{\epsilon_{si}}. \quad (3.6)$$

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Differentiating the integrated Poisson's solution (3.4) results in the implicit equation [90],

$$z_{cc} = \cos\left(\frac{x_0 \sqrt{A_{cc}}}{2V_{th}} e^{-\phi_{cc}/2V_{th}} z_{cc}\right) \quad (3.7)$$

where

$$z_{cc} = \sqrt{\left(1 - \frac{1}{A_{cc}} \left(\frac{C_{ox}}{\epsilon_{si}} (\phi_{cc} - V_{gs} + V_{FB})\right)^2 e^{\phi_{cc}/V_{th}}\right)} \quad (3.8)$$

ϕ_{cc} is the regional solution in accumulation. This is an implicit equation which can be solved iteratively using Newton-Raphson algorithm. Nevertheless, an explicit solution is possible if assuming the surface potential dominates and the zero-field potential, which is very small and approaching zero in strong accumulation, $\phi_{cc} \approx 0$. This results in the solution that is identical to the regional solution in the bulk-MOSFET.

Beyond the flat-band voltage, the space-charge region is now dominated by the ionized dopants (either acceptors or donors). If only acceptors term (for NMOS) is considered, then

$$\frac{d^2\phi}{dx^2} = \frac{qN_{ch}}{\epsilon_{si}}. \quad (3.9)$$

Ca-DG device may be partial-depleted, fully-depleted or both in the subthreshold region depending on the channel dopant concentration, N_{ch} and body thickness, t_{si} . The threshold that separates the two regimes is defined as the full-depletion voltage, V_{gs1FD} . The channel is fully-depleted when the sum of the depletion regions of both gates reaches t_{si} .

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$$x_{d1} + x_{d2} = t_{si} \quad (3.10)$$

where x_{d1} and x_{d2} are the depletion widths due to gate 1 and gate 2 alone, respectively.

Since the channel is in depletion, it is assumed that the full-depletion approximation is valid. Therefore, the depletion widths are given as

$$x_{d1} = \sqrt{\frac{2\epsilon_{Si}}{qN_{ch}}} \left[-\frac{\gamma_1}{2} + \sqrt{\frac{\gamma_1^2}{4} + V_{gs1} - V_{FB1}} \right], \quad (3.11)$$

$$x_{d2} = \sqrt{\frac{2\epsilon_{Si}}{qN_{ch}}} \left[-\frac{\gamma_2}{2} + \sqrt{\frac{\gamma_2^2}{4} + V_{gs2} - V_{FB2}} \right]. \quad (3.12)$$

Another equation is needed to solve for the $V_{gs1}=V_{gs1FD}$. In common-gate configuration, these two terminals are tied together; therefore,

$$V_{gs2} = V_{gs1}. \quad (3.13)$$

Replacing (3.11), (3.12) and (3.13) into (3.10), the solution for V_{gs1FD} is given by

$$V_{gs1FD} = \frac{k^2 - 4G_1G_2}{4(k + G_1 + G_2)} \quad (3.14)$$

where

$$k = \left(\frac{qN_{ch}t_{si}^2}{2\epsilon_{Si}} + (\gamma_1 + \gamma_2) \sqrt{\frac{qN_{ch}t_{si}^2}{2\epsilon_{Si}}} + V_{FB1} + V_{FB2} + \frac{\gamma_1\gamma_2}{2} \right) \quad (3.15)$$

$$G_1 = \frac{\gamma_1^2}{4} - V_{FB1} \quad (3.16)$$

$$G_2 = \frac{\gamma_2^2}{4} - V_{FB2}. \quad (3.17)$$

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The zero-field location from gate 1 and gate 2 can be determined by applying V_{gs1FD} and V_{gs2FD} into (3.11) and (3.12), respectively. The prediction of zero-field location automatically goes back to $t_{si}/2$ when the device is symmetrical.

The surface begins to be depleted of the charges beyond flat-band voltage, similar to bulk MOSFET, the surface potential in depletion for DG MOSFET is given by

$$\phi_{DD} = \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} + V_{gs} - V_{FB}} \right]^2, \quad (3.18)$$

in which the reference voltage is now changed from bulk to source terminal, i.e., V_{gb} to V_{gs} .

In the weak accumulation, the prevailing term due to acceptor can be extended for $V_{gs} < V_{FB}$ by changing the sign of the surface potential and, thus, yielding

$$\phi_{ADD} = - \left[-\frac{\gamma}{2} + \sqrt{\frac{\gamma^2}{4} - V_{gs} + V_{FB}} \right]^2. \quad (3.19)$$

The surface potential of DG MOSFETs is similar to that for bulk MOSFETs except for the volume inversion region. The volume inversion is given by the regional solution of (3.2), which is

$$\phi_{VI} = V_{gs} - V_{FB} - \gamma \sqrt{\frac{qN_{ch}x_d^2}{2\epsilon_{si}}}, \quad (3.20)$$

where x_d is the depletion depth that will eventually reach a maximum when volume inversion begins.

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In strong inversion, both the acceptors and electrons should be considered. Nevertheless, second-integration of Poisson's equation is not possible when the acceptors-term is considered together with electrons. Therefore, dropping the acceptors-term and consider only the electrons-term, (3.2) becomes

$$\frac{d^2\phi}{dx^2} = \frac{qN_{ch}}{\epsilon_{si}} \left(e^{(\phi - 2\phi_F - V)/v_{th}} \right). \quad (3.21)$$

With the aforementioned boundary conditions, the Poisson's solution is given as:

$$\phi(x) = -2v_{th} \ln \left(\cos \left(\arccos \left(\sqrt{\frac{B_{ss}}{A_{ss}}} e^{\phi_{ss}/2v_{th}} \right) \pm \frac{x\sqrt{B_{ss}}}{2v_{th}} \right) \right) + v_{th} \ln \frac{B_{ss}}{A_{ss}}. \quad (3.22)$$

This results in an implicit solution expressed as in [90]

$$z_{ss} = \cos \left(\frac{z_{ss} x_d \sqrt{A_{ss}}}{2v_{th}} e^{\phi_{ss}/2v_{th}} \right), \quad (3.23)$$

which can be solved with Newton-Raphson algorithm, with

$$z_{ss} = \sqrt{1 - \frac{1}{A_{ss} e^{\phi_{ss}/2v_{th}}} \left(\frac{C_{ox}}{\epsilon_{si}} (V_{gs} - V_{FB} - \phi_{ss}) \right)^2} \quad (3.24)$$

and

$$A_{ss} = \frac{2qN_{ch}v_{th}e^{-2\phi_F - V/v_{th}}}{\epsilon_{si}}. \quad (3.25)$$

Similarly, if the zero-field potential is assumed weakly coupled to the surface potential, we have a strong-inversion solution identical to that for bulk MOSFET.

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Unified regional solutions are obtained through the two complementary interpolation functions as shown in section 2.2.2 and repeated here for convenience,

$$V_{GSF} = \vartheta_f \{x; \sigma_f\} \equiv 0.5 \left(x + \sqrt{x^2 + 4\sigma_f} \right) \quad (3.26)$$

$$V_{GSR} = \vartheta_r \{x; \sigma_a\} \equiv 0.5 \left(x - \sqrt{x^2 + 4\sigma_a} \right) \quad (3.27)$$

by replacing $x=V_{gs}-V_{FB}$ in the regional piece-wise solutions. The regional solutions need to be unified and combined to form a single-piece solution. The unified strong accumulation solution is similar to (2.41) but replacing V_{gb} with V_{gs} due to change in the reference voltage

$$\phi_{STR,CC} = \phi_{CC} \Big|_{V_{gs}-V_{FB}=V_{GSR}}, \quad (3.28)$$

and the weak accumulation piece is given by

$$\phi_{ASUB} = \phi_{ADD} \Big|_{V_{gs}-V_{FB}=V_{GSR}}. \quad (3.29)$$

These two pieces will form the unified accumulation solution similar to (2.47) and repeated here

$$\phi_{ACC} = \vartheta_{eff,cc} \{ \phi_{ASUB}, \phi_{STR,CC}, \delta_{cc} \}. \quad (3.30)$$

The unified depletion solution is similar to (2.43) with

$$\phi_{SUB} = \phi_{DD} \Big|_{V_{gs}-V_{FB}=V_{GSR}}. \quad (3.31)$$

For DG MOSFET, the depletion layer will continue to increase starting from flat-band voltage until the fully-depleted voltage, V_{gsFD} at which depletion stops. The corresponding full-depletion potential can be found by applying V_{gsFD} given in (3.14) into (3.18):

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$$\phi_{FD} = \phi_{DD} \Big|_{V_{gs}=V_{gsFD}}. \quad (3.32)$$

Therefore, a unified solution from depletion to full-depletion potential, ϕ_{DF} can be found using

$$\phi_{DF} = \vartheta_{eff,ss} \{ \phi_{DD}, \phi_{FD}; \delta_{dd} \}. \quad (3.33)$$

Then, applying the ϕ_{DF} solution in the volume-inversion equation (3.20), it gives the depletion to volume-inversion solution, ϕ_{DV} :

$$\phi_{DV} = V_{GSF} - \gamma \sqrt{\phi_{DF}}. \quad (3.34)$$

The strong inversion solution is similar to (2.44), given by

$$\phi_{STR,SS} = \phi_{SS} \Big|_{V_{gs}-V_{FB}=V_{GSF}}. \quad (3.35)$$

The depletion to strong inversion solution, ϕ_{DS} is obtained similarly to (2.48), given by

$$\phi_{DS} = \vartheta_{eff,ss} \{ \phi_{STR,SS}, \phi_{DV}; \delta_{ss} \}. \quad (3.36)$$

An alternative improvement, similar to NR iteration as done in bulk MOSFET, can be achieved to reduce dependency of the smoothing parameter δ_s . The final single-piece surface potential is therefore given by

$$\phi_{SEFF} = \phi_{ACC} + \phi_{DS}. \quad (3.37)$$

3.3. Ca-DG unified regional zero-field potential

The zero-field potential can be obtained as an explicit function of the surface potential with the second-integration of the Poisson solution considering only one type of carriers as shown in equations (3.4) and (3.22), for hole and electron, respectively.

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The unified regional zero-field potential (URZP) in accumulation, ϕ_{ACC} given in (3.30) and interpolated function V_{GSR} given in (3.26) are used in replacement of ϕ_s and $V_{gs}-V_{FB}$ respectively. The URZP in accumulation can be obtained from

$$\phi_{0ACC} = 2v_{th} \ln \left(\cos \left(\arccos \left(\sqrt{\frac{B_{acc}}{A_{cc}}} e^{\phi_{ACC}/2v_{th}} \right) + \frac{\sqrt{B_{acc} x_d}}{2v_{th}} \right) \right) - v_{th} \ln \frac{B_{acc}}{A_{cc}} \quad (3.38)$$

where

$$B_{acc} = A_{cc} e^{-\phi_{ACC}/2v_{th}} - \left(\frac{C_{ox}}{\epsilon_{si}} (V_{GSR} - \phi_{ACC}) \right)^2 \quad (3.39)$$

and A_{cc} is given in (3.6). The zero-field potential remains almost constant when the surface is depleted. It can therefore be neglected. The zero-field potential starts to increase when the gate voltage is larger than V_{gs1FD} . Therefore, considering only the electron term and applying the unified regional surface potentials, ϕ_{DF} and ϕ_{DS} given in (3.33) and (3.36), respectively, and V_{GSF} given in (3.27), the URZP for strong inversion is given as

$$\phi_{0DS} = v_{th} \ln \left[\frac{B_{ss}}{A_{ss}} \right] - 2v_{th} \ln \left(\cos \left(\arccos \left(\sqrt{\frac{B_{ss}}{A_{ss}}} e^{-(\phi_{DS}-\phi_{DF})/v_{th}} \right) + \frac{\sqrt{B_{ss} x_d}}{2v_{th}} \right) \right) \quad (3.40)$$

where

$$B_{ss} = \left(A_{ss} e^{(\phi_{DS}-\phi_{DF})/v_{th}} - \left(\frac{C_{ox}}{\epsilon_{si}} [V_{GSF} - (\phi_{DS} - \phi_{DF})] \right)^2 \right). \quad (3.41)$$

The $\phi_{DS} - \phi_{DF}$ term allows the model to have the correct “turn-on” of the zero-field potential. Therefore, the final single-piece URZP is

$$\phi_{0EFF} = \phi_{0ACC} + \phi_{0DS}. \quad (3.42)$$

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The derived unified regional solutions are compared with Medici data for verification. Figure 3.3(a) shows the unified regional surface potential solutions for “pure” body MOSFET. It can be separated into strong accumulation, weak accumulation, weak inversion and strong inversion regions. The zero-field potential seems to saturate in strong accumulation and inversion. At higher body doping concentration, the surface potential has accumulation, depletion, weak inversion and strong inversion as shown in Figure 3.3(b). The zero-field potential in strong inversion still increases compared to Medici data; however, it will not have much influence on both the current and charge, as the surface potential will play the dominant role. Comparison with Medici data is also done for different body doping concentration, from pure body to high-doping concentration. The surface potentials and zero-field potentials are able to capture the changes in body doping concentration correctly and smoothly, as shown in Figure 3.4 and its inset. Finally, ca-DG surface potential solutions are shown in Figure 3.5, where the oxide thickness from the first gate is varied. The first gate surface potential is changing with different oxide thickness, and the inset shows the smooth derivatives of the corresponding surface potentials.

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Figure 3.3: Surface and zero-field potentials for (a) undoped (pure) body ($N_A=0 \text{ cm}^{-3}$) (b) highly doped body, $N_A=10^{18} \text{ cm}^{-3}$ s-DG and their regional components.

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(a)

(b)

Figure 3.4: (a) Surface potential and (b) Zero-field potential model validation from undoped to doped and fully-depleted to partially-depleted for changing channel doping concentration.

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Figure 3.5: First gate surface potential validation when second gate oxide thickness changes from s-DG to a-DG with common gate. Inset: the first-order derivative of surface potential showing smooth transitions.

3.4. Quantum-mechanical correction for ca-DG unified regional surface potential

The quantum-mechanical correction for DG is included similar to bulk-MOS according to van Dort *et. al.* approach [51]. The modification of the intrinsic carrier density is repeated here,

$$n_i^{qm} = n_i f^{qm} \quad (3.43)$$

in which f^{qm} is given in (2.72). Similar to QME derivation in bulk MOSFETs, the Poisson solution for s-DG is given as

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$$V_{gs} - V_{FB} - \phi_s^{qm} = -Q_{sc}^{qm} / C_{ox} \quad (3.44)$$

if the spatial dependence in (3.43) is ignored. Note that the induced charge modifies only the free carrier at the surface,

$$Q_{sc}^{qm} = -C_{ox} \gamma \text{sgn}(\phi_s^{qm}) \sqrt{f_\phi^{qm}} \quad (3.45)$$

where

$$f_\phi^{qm} = v_{th} \left[f_{qm} \exp\left(-\frac{\phi_s^{qm}}{v_{th}}\right) - \exp\left(-\frac{\phi_0}{v_{th}}\right) \right] + (\phi_s^{qm} - \phi_0) \left[1 - \exp\left(-\frac{V + 2\phi_F}{v_{th}}\right) \right] \\ + v_{th} \exp\left(-\frac{V + 2\phi_F}{v_{th}}\right) \left[f_{qm} \exp\left(\frac{\phi_s^{qm}}{v_{th}}\right) - \exp\left(\frac{\phi_0}{v_{th}}\right) \right] \quad (3.46)$$

(3.44) with (3.45) and (3.46) is solved using explicit regional approximation. In either strong accumulation or strong inversion, the contribution of zero-field potential can be ignored. Following the derivation shown in section 2.3.4 and section 3.2, the DG MOSFETs unified regional potential solutions with QME can be derived. The model is compared with the simulated Medici data.

The comparison of surface potential in classical model and with quantum mechanical correction for different body doping concentration is shown in Figure 3.6 (with default value $\kappa=1$, $V_{ds}=0$ V unless specified explicitly). The model follows well for the whole range of practical body doping concentration, beginning from ideal pure semiconductor (i.e., $N_A = 0$) to highly doped body, $N_A=10^{18} \text{ cm}^{-3}$. Figure 3.7(a) shows the comparison of unified regional surface and zero-field potential with quantum mechanical correction for different κ values, which is taken directly from the Medici simulation for undoped body. Figure 3.7(b) shows the corresponding derivatives of the surface and zero-field potentials. Similar plot is included in Figure 3.8 but at high

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doping concentration, $N_A = 10^{18} \text{ cm}^{-3}$ and the derivative of the surface potential is included in the inset. The surface potential is following the data physically and smoothly (from the derivatives). Though, the zero-field potential has over-predicted the data for highly-doped body, it has negligible effect in the current model, as most current only conduct in the depletion layer at the surface. Figure 3.9 and Figure 3.10 show the prediction for surface potential (and inset shows the zero-field potential) for different body thickness and asymmetric oxide thickness, respectively. The surface potential is able to predict the changes physically.

Figure 3.6: Comparison of unified regional surface potential with Medici data for classical and quantum mechanical corrected model.

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Figure 3.7: Comparison of a) unified regional surface potential and zero-field potential with Medici data with quantum mechanical correction model for different κ parameter at undoped body. b) The corresponding derivative of the potentials.

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Figure 3.8: Comparison of unified regional surface potential and zero-field potential with Medici data with quantum mechanical correction model for different κ parameter at highly doped body, $N_A=10^{18}$ cm $^{-3}$. Inset: The corresponding derivatives of surface potentials.

Figure 3.9: Comparison of unified regional surface potential with quantum mechanical correction for different body thickness. Inset: The corresponding zero-field potential.

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Figure 3.10: Comparison of unified regional surface potential with quantum mechanical correction for different asymmetric oxide thickness. Inset: The corresponding zero-field potential.

3.5. Ca-DG unified regional full-depletion dominated drain current

At low body doping and thinner silicon body, the device is operating in full-depletion mode in subthreshold region. This is very much like undoped DG behaviour that has no visible depletion region. Therefore, the normalized inversion charge, $V_{gt,s}$ is derived by rearranging (3.23) to obtain an input voltage equation as

$$V_{gf} - \phi_s = \gamma \sqrt{v_{th} e^{\frac{\phi_s - 2\phi_F - V}{v_{th}}}} \left(\sin \left(\frac{\gamma C_{ox}}{\epsilon_{si}} \frac{x_0}{2v_{th}} \sqrt{v_{th} e^{\frac{\phi_s - 2\phi_F - V}{v_{th}}}} - \frac{1}{\gamma^2} (V_{gf} - \phi_s)^2 \right) \right). \quad (3.47)$$

Nevertheless, in the URM approach, ϕ_s in (3.47) is derived explicitly and is not accurate enough for direct evaluation in the drain current. Therefore, we rewrite $V_{gt,s}$

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by approximating the argument in the square-root with the first integration of the Poisson equation (taking only the electrons term in (3.2)) as

$$V_{gt,s} = \gamma \sqrt{v_{th} e^{\frac{\phi_s - 2\phi_F - V}{v_{th}}}} \left(\sin \left(\frac{\gamma C_{ox}}{\epsilon_{si}} \frac{x_0}{2v_{th}} \sqrt{v_{th} e^{\frac{\phi_0 - 2\phi_F - V}{v_{th}}}} \right) \right). \quad (3.48)$$

The current expression used for the undoped-DG can be modified for use in full-depletion dominant drain current model, including channel doping and considering only one gate,

$$I_{ds} = \beta \left(\left(V_{gf} - \overline{\phi_s} + 2v_{th} \right) \Delta\phi_s + \frac{t_{si} q v_{th} N_{ch}}{2C_{ox}} \left(e^{\frac{(\phi_{0,d} - 2\phi_F - V_{ds})}{v_{th}}} - e^{\frac{\phi_{0,s} - 2\phi_F}{v_{th}}} \right) \right) \quad (3.49)$$

where

$$\overline{\phi_s} = 0.5(\phi_{s,d} + \phi_{s,s}) \quad (3.50)$$

$$\Delta\phi_s = \phi_{s,d} - \phi_{s,s}. \quad (3.51)$$

The third-term in (3.49) is much smaller than the first two terms, especially in strong inversion, and is negligible [91]. Therefore, the drain current model is simplified and the “non-pinned” approximation, $\phi(L) \approx \phi(0) + V_{ds}$ is used to derive the effective drain-source voltage, $V_{ds,eff}$ as done in bulk, shown in section 2.4.5. The drain current with piece-wise lateral-field degradation is given as

$$I_{ds} = \frac{W \mu_0 C_{ox} \left(V_{gt,s} + 2v_{th} - \frac{V_{ds}}{2} \right) V_{ds}}{L_{eff} (1 + V_{ds} / L_{eff} E_{sat})} \quad (3.52)$$

while the saturation current is approximately

$$I_{ds,sat} = W v_{sat} C_{ox} (V_{gt,s} - V_{ds,sat}). \quad (3.53)$$

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$V_{ds,sat}$ can be solved by equating (3.52) and (3.53) at $V_{ds}=V_{ds,sat}$

$$V_{ds,sat} = \frac{V_{gts} L_{eff} E_{sat}}{V_{gts} + 4v_{th} + L_{eff} E_{sat}} . \quad (3.54)$$

For short-channel devices, (3.52) with series resistance is given by

$$I_{ds1} = \frac{\mu_0 W C_{ox} \left(V_{gts} + 2v_{th} - \frac{V_{ds}}{2} \right) V_{ds}}{L_{eff} \left(1 + V_{ds} / L_{eff} E_{sat} \right) + \mu_0 W C_{ox} R_{sd} \left(V_{gts} + 2v_{th} - \frac{V_{ds}}{2} \right)} \quad (3.55)$$

and saturation current with series resistance

$$V_{ds,sat} = \frac{-b - \sqrt{b^2 - 4ac}}{2a} \quad (3.56)$$

where

$$a = v_{sat} W C_{ox} R_s \quad (3.57)$$

$$b = - \left(V_{gt,s} + v_{sat} W C_{ox} V_{gt,s} (2R_s + R_{sd}) + L_{eff} E_{sat} + 4v_{th} (1 + v_{sat} W C_{ox} R_s) \right) \quad (3.58)$$

$$c = V_{gt,s} L_{eff} E_{sat} + 2v_{sat} W C_{ox} R_{sd} (V_{gt,s} + 2v_{th}) V_{gt,s} . \quad (3.59)$$

$V_{ds,eff}$ is a result of the smoothing of V_{ds} and $V_{ds,sat}$, using the same smoothing function as surface potential given in (2.46)

$$V_{ds,eff} = \vartheta_{eff,ss} \{ V_{ds}, V_{ds,sat}; \delta_s \} . \quad (3.60)$$

The drain current without series resistance in the current expression is

$$I_{ds0} = \frac{W \mu_0 C_{ox}}{L_{eff} (1 + V_{ds,eff} / L_{eff} E_{sat})} \left((V_{gf} - \overline{\phi_{s,eff}} + 2v_{th}) V_{ds,eff} \right) . \quad (3.61)$$

The CLM/VO is included similarly to bulk-MOSFET, the final drain current expression is therefore

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$$I_{ds} = \frac{g_{vo} I_{ds0}}{1 + R_{sd} I_{ds0} / V_{ds,eff}} \quad (3.62)$$

where

$$g_{vo} = 1 + \left(\frac{V_{ds} - V_{ds,eff}}{V_{Aeff}} \right) \quad (3.63)$$

$$V_{Aeff} = \frac{\left(L_{eff} E_{sat} \left[1 + h(V_{ds} - V_{ds,eff}) \right] \right) + \delta_L V_{ds,eff}}{\delta_L h V_{ds,eff}} \quad (3.64)$$

$$h = \frac{\frac{\xi c}{l^2} \left[1 + \sqrt{1 + \left(\frac{V_{ds} - V_{ds,eff}}{l E_{sat}} \right)^2} \right]}{\sqrt{1 + \left(\frac{V_{ds} - V_{ds,eff}}{l E_{sat}} \right)^2} - \frac{\xi c}{l^2} (V_{ds} - V_{ds,eff})} \quad (3.65)$$

3.6. Ca-DG unified regional partial-depletion dominated drain current

If the body is highly doped and/or with a thicker body, the subthreshold region may have both the depletion and volume-inversion regions. In this case, the depletion term due to N_A cannot be ignored. However, at the same time, the zero-field potential is close to zero in depletion region. Therefore, the drain current expression is derived similarly to bulk-MOSFET, while ignoring the zero-field potential. The inversion charge is given by

$$Q_i = -C_{ox} \left(V_{gf} - \phi_s - \gamma \sqrt{\phi_s - \phi_0} \right) \quad (3.66)$$

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where the bulk charge is

$$Q_b = -\gamma C_{ox} \sqrt{\phi_s - \phi_0} . \quad (3.67)$$

The normalized inversion charge is further simplified according to symmetric bulk-charge linearization,

$$q_i \approx \bar{q}_i - \bar{A}_b (\phi_s - \bar{\phi}_s) \quad (3.68)$$

where

$$\bar{q}_i = V_{gf} - \bar{\phi}_s - \gamma \sqrt{\bar{\phi}_s - \bar{\phi}_0} \quad (3.69)$$

and

$$\bar{A}_b = 1 + \frac{\gamma}{2\sqrt{\bar{\phi}_s - \bar{\phi}_0}} . \quad (3.70)$$

$$\bar{\phi}_s = \frac{\phi_{s,d} + \phi_{s,s}}{2} \quad (3.71)$$

$$\bar{\phi}_0 = \frac{\phi_{0,d} + \phi_{0,s}}{2} . \quad (3.72)$$

The drain current can be derived with charge-sheet approximation. The internal functions that have the bulk-charge term are changed accordingly with (3.67), for example the effective transverse field becomes

$$E_{eff} = \frac{C_{ox}}{\epsilon_{si}} \left(\zeta_n \bar{V}_{gt,s} + \gamma \sqrt{\bar{\phi}_s - \bar{\phi}_0} \right) \quad (3.73)$$

and the normalized inversion charge for computing the $V_{ds,eff}$ is

$$V_{gt,s} = \gamma \sqrt{\phi_{s,s} - \phi_{0,s}} + v_{th} \left(e^{(\phi_{s,s} - 2\phi_F)/v_{th}} - e^{(\phi_{0,s} - 2\phi_F)/v_{th}} \right) - \gamma \sqrt{\phi_{s,s} - \phi_{0,s}} . \quad (3.74)$$

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The drain saturation voltage considering doping has only slightly different form in the quadratic pre-factors (3.57)–(3.59),

$$a = A_{b,s} v_{sat} W C_{ox} R_s \quad (3.75)$$

$$b = -\left(V_{gt,s} + v_{sat} W C_{ox} V_{gt,s} (2R_s + R_{sd}) + L_{eff} E_{sat} + 2A_{b,s} v_{th} (1 + v_{sat} W C_{ox} R_s)\right) \quad (3.76)$$

$$c = V_{gt,s} L_{eff} E_{sat} + 2v_{sat} W C_{ox} R_{sd} (V_{gt,s} + A_{b,s} v_{th}) V_{gt,s} . \quad (3.77)$$

The drain saturation voltage has negligible difference (<2.5%) even for undoped body with these expressions, as seen in Figure 3.11.

Figure 3.11: Comparison of drain saturation voltage using undoped pre-factor (3.57)–(3.59) and doped pre-factor (3.75)–(3.77).

The final DG drain current is given as (3.62), where

$$I_{ds01} = \frac{W \mu_0 C_{ox}}{L_{eff} (1 + V_{ds,eff} / L_{eff} E_{sat})} (\overline{q_{t1}} + \overline{A_{b1}} v_{th}) V_{ds,eff} , \quad (3.78)$$

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which is similar to bulk-MOSFET. The model is compared with Medici data for verification.

The drain current data are generated for channel doping concentration variations, as shown in Figure 3.12. The changes in the subthreshold slope are clearly visible with the increase in body doping concentration. The combined change due to body thickness and doping concentration is shown in Figure 3.13. The variation of back-gate oxide thickness for ca-DG is shown in Figure 3.14. The long-channel drain current at different gate voltages, $V_{gs}=0.4, 0.8,$ and 1.2 V, is given in Figure 3.15. The inset shows the output resistance of the corresponding current. Figure 3.16 and Figure 3.17 show the comparison with the numerical data at short channel, $L_g=90$ nm. The insets for these figures show the corresponding first-order derivative, transconductance, g_m and output conductance, g_d , respectively.

Figure 3.12: S-DG drain current validation for channel doping variation, from undoped to $N_{ch} = 10^{17}$ and 10^{18} cm^{-3} .

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Figure 3.15: S-DG drain current validation for long channel at $V_{gs} = 0.4, 0.8$ and 1.2 V. Inset: The output resistance of the corresponding output drain current.

Figure 3.16: S-DG drain current validation for short channel at $V_{ds} = 0.05$ and 1.2 V. Inset: The transconductance of the corresponding drain current.

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Figure 3.17: S-DG drain current validation for short channel at $V_{gs} = 0.4, 0.8$ and 1.2 V. Inset: The output conductance of the corresponding drain current.

3.7. Ca-DG Ground-referenced unified regional surface potential

The potential solutions and the drain current model described in the sections above are derived based on the source-referenced convention. All input voltages and internal variables are described relative to source terminal, e.g., V_{gs} and V_{ds} . Source-referenced convention had created problems in bulk-MOSFET models and caused failure to meet the Gummel symmetry requirement. For example, the problems were found in intermediate variables used in the transverse electric field, E_{eff} , effective drain-source voltage, $V_{ds,eff}$ and threshold-voltage based drain current expression [5, 7, 34, 69]. Similarly, for DG-MOSFET model, all models developed based on source-

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referenced convention are bound to fail the GST due to similar reasons, i.e., failing to keep the drain current strictly an odd function of V_{ds} .

Therefore, the input voltage equation needs to be developed based on external ground reference, where (3.2) has to be redefined. If we consider only the undoped body and ignoring the hole-term, the input voltage equation is given by

$$V_{gf} - \phi_s = \gamma \sqrt{v_{th} e^{-(2\phi_F + V)/v_{th}} \left(e^{\phi_s/v_{th}} - e^{\phi_b/v_{th}} \right)} \quad (3.79)$$

where ground referenced flat-band shifted gate voltage, $V_{gf} = V_g - V_{FB}$ and V_r is the minimum voltage in the silicon body (either V_d or V_s , whichever is lower)

$$V_r = \min(V_d, V_s) \quad (3.80)$$

and V is channel voltage, it is V_d at drain side and V_s at source side. Following the URM approach, the surface potential can be decoupled by first ignoring the contribution of the zero-field potential, thus further simplifying (3.79) to

$$V_{gf} - \phi_s = \gamma \sqrt{v_{th} e^{(\phi_s - 2\phi_F - V)/v_{th}}}, \quad (3.81)$$

which can be solved with Lambert- W solution [87] as

$$\phi_{ss} = V_{gf} - 2v_{th} L\{W\} \quad (3.82)$$

where

$$W = \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(\frac{V_g - V_{FB} - V}{2v_{th}}\right). \quad (3.83)$$

The smoothing functions need to be separated between source- and drain-side using

$$V_{GFY} = \vartheta_f \{V_{gf}, V; \sigma_f\} \equiv V + 0.5 \left(V_{gf} - V + \sqrt{(V_{gf} - V)^2 + 4\sigma_f} \right) \quad (3.84)$$

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where Y is either D or S depending on V_d or V_s being used. Therefore, the unified regional surface potential at source side is given by applying V_s into (3.81) and (3.84)

$$\phi_{DS,s} = V_{GFS} - 2v_{th}L \left\{ \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(\frac{V_g - V_{FB} - V_s}{2v_{th}}\right) \right\}. \quad (3.85)$$

Similarly, for the drain end, the unified regional surface potential is by replacing V_d with V_s as

$$\phi_{DS,d} = V_{GFD} - 2v_{th}L \left\{ \frac{\gamma}{2\sqrt{v_{th}}} \exp\left(\frac{V_g - V_{FB} - V_d}{2v_{th}}\right) \right\}. \quad (3.86)$$

In order to meet the GST, two separate saturation voltages from drain side and source side, referenced to external ground are derived similarly to bulk-referenced concept. In “forward” mode operation, $V_d > V_s$, where the channel pinch-off happens at drain side. Therefore, the long-channel ground-referenced drain saturation voltage is given as (by rearranging $V_{ds} = V_d - V_s$),

$$V_{d,sat} = V_{ds,sat} + V_s = \frac{V_{gt,s}LE_{sat}}{V_{gt,s} + 4v_{th} + LE_{sat}} + V_s. \quad (3.87)$$

Similarly, at source side for “reverse” mode, $V_s > V_d$

$$V_{s,sat} = V_{sd,sat} + V_d = \frac{V_{gt,d}LE_{sat}}{V_{gt,d} + 4v_{th} + LE_{sat}} + V_d, \quad (3.88)$$

both of which will then be combined using the following smoothing function to form a single-piece solution from linear to saturation region,

$$\vartheta_{eff,\delta} \{x, x_{sat}, x_{xsat}; \delta\} \equiv x_{sat} - 0.5 \left[x_{sat} - x - \sqrt{(x_{sat} - x)^2 + 4\delta x_{xsat} + \delta} \right]. \quad (3.89)$$

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This smoothing function will ensure the consistency of drain current model regardless of any reference voltage when it is applied to all input terminals. The effective drain-source voltage is then given as

$$V_{ds,eff} = V_{d,eff} - V_{s,eff} \quad (3.90)$$

with

$$V_{d,eff} = \vartheta_{eff,\delta} \{V_d, V_{d,sat}, V_{ds,sat}; \delta\} \quad (3.91)$$

$$V_{s,eff} = \vartheta_{eff,\delta} \{V_s, V_{s,sat}, V_{sd,sat}; \delta\}. \quad (3.92)$$

The final long-channel drain current equation is given as

$$I_{ds} = \frac{W \mu_0 C_{ox}}{L \sqrt{1 + (V_{ds,eff} / LE_{sat})^2}} (V_G - V_{FB} - \overline{\phi_{s,eff}} + 2v_{th}) V_{ds,eff} \quad (3.93)$$

with

$$\overline{\phi_{s,eff}} = \frac{\phi(V_{d,eff}) + \phi(V_{s,eff})}{2}. \quad (3.94)$$

The drain current model is checked for consistency with an arbitrary reference voltage, $V_0 = 10$ V is added to all the terminals and compared to the drain current with $V_0 = 0$ V. It shows no difference with the $V_0 = 0$ V case, as can be seen in Figure 3.18. The drain currents with V_{gs} and V_{ds} sweeping are shown in Figure 3.19 and Figure 3.20, respectively. Finally, the GST is done on the ground-referenced s-DG I_{ds} model, which is able to match the Medici data correctly without any glitch, as shown in Figure 3.21. The sixth-order derivative is also included in the upper inset, which shows that it meets the GST requirement for all its lower-order derivatives. The lower inset shows the GST schematic circuit used to test the model.

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Figure 3.18: Model consistency test when arbitrary voltage, V_0 is added to all terminal voltages. No difference between $V_0=0$ and $V_0=10$ V is observed.

Figure 3.19: Drain current comparison with Medici data from subthreshold to strong inversion at $V_{ds}=0.05$, 0.6 and 1.2 V.

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Figure 3.20: Drain current comparison with Medici data from linear to saturation at $V_{gs}=0.3, 0.6, 0.9$ and 1.2 V.

Figure 3.21 : Comparison of GST with Medici data at second-order derivative. Upper inset: sixth order derivative of the drain current. Lower inset: Schematic circuit for GST.

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3.8. Chapter 3 conclusion

The unified regional modeling approach has been extended to model the doped ca-DG MOSFETs, which can be technically challenging for other modeling approaches. The unified regional surface potential and unified regional zero-field potential models are able to scale physically with different body doping concentration, body thickness, and oxide thickness seamlessly. Quantum-mechanical correction has been included in the surface potential solutions to show the extendibility of the URM approach to DG-MOSFET models. It is able to match the process variations correctly. The new ground-referenced unified regional potential solution for undoped s-DG MOSFETs is able to ensure that the model is always an odd function of V_{ds} (in order to pass the Gummel symmetry requirement) and still consistent for any arbitrary reference voltage.

CHAPTER 4: Extraction of Physical and Model Parameters**CHAPTER 4: Extraction of Physical and Model Parameters**

A good model is not only scalable and physical but it should be also easy to understand and to extract its parameters. This chapter will illustrate the procedures to extract the physical parameters (i.e., t_{ox} , N_{ch} , N_{gate}) and model parameters (interpolation/smoothing parameters: σ_a , σ_f , δ_ϕ , etc. and fitting parameters: μ_1 , μ_2 , μ_3 , etc.). The URM-based charge model has less parameter dependency as compared to other compact models, as the physical parameters are uncorrelated to the smoothing parameters. In this chapter, the model extraction procedure is described based on a bulk MOSFET with the help of Medici data and measurement data. Similar procedure can also be extended to double-gate model extraction as both shared a similar core model equation.

4.1. URSP bulk-MOSFET charge model extraction

To demonstrate the correct behaviour of the URSP charge model, C_{gg} data are generated numerically using known physical parameters ($t_{ox} = 2$ nm, $N_{ch} = 5 \times 10^{17}$ cm⁻³), with and without poly-Si gate effect ($N_{gate} = 10^{22}$, 5×10^{20} , and 5×10^{19} cm⁻³), and

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with and without QME ($\kappa^m = 0, 0.5, 1$) for long channel ($L_g = 10 \mu\text{m}$) and short channel ($L_g = 90 \text{ nm}$) MOSFETs.

It can be seen in Figure 4.1 that the regional capacitance model is able to follow the accumulation and strong inversion capacitance of the MEDICI data correctly with the matching value of the physical oxide thickness ($t_{ox} = 2\text{nm}$, dotted line). Other oxide thicknesses have either over- or under-estimated the capacitance relative to the data. Similarly, the matching value of channel doping concentration will give the matching regional depletion capacitance, while other doping concentrations have failed to match the data, as shown in Figure 4.2. The effect of poly-Si gate concentration variations on each individual unified regional charge/capacitance model can be seen in Figure 4.3. It is observed that the poly-Si gate effect is stronger at lower doping concentration, especially for the strong-inversion capacitance, where the poly-depletion effect is most pronounced. Besides, the poly-accumulation has caused the accumulation capacitance to have a “cross over” in the accumulation region. Other models that have not included the PAE will not have this behaviour. For very deep-submicron process, the oxide thickness is so thin that the quantum mechanical effect can no longer be ignored. QME can be either modelled with effective oxide thickness or modification of intrinsic carrier concentration. In this work, the latter was chosen as it had better physical meaning. Effectively, QME has caused the strong accumulation and strong inversion to vary, while depletion region remains unchanged, as seen in Figure 4.4.

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(a)

(b)

Figure 4.1: The comparison of unified regional model's effective oxide thickness variations with Medici data for a) accumulation capacitance and b) strong inversion

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Figure 4.2: The comparison of unified regional model's effective channel doping concentration variations with Medici data for depletion capacitance.

Figure 4.3: The comparison of unified regional model's effective poly-gate doping concentration variations with Medici data.

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Figure 4.4: Comparison of model's unified regional capacitance with Medici data for quantum mechanical fitting parameter.

In a model, all of the device parameters are essentially effective values, which need to be extracted from the measurement data. Therefore in this validation, a set of long-channel C_{gg} is numerically generated from a device with known parameters ($t_{ox} = 2 \text{ nm}$, $N_{ch} = 5 \times 10^{17} \text{ cm}^{-3}$, $N_{gate} = 5 \times 10^{19} \text{ cm}^{-3}$, $\kappa^{qm} = 1$), which is used to illustrate the calibration steps for a given measurement data of a long-channel device.

Step 1: The model assumes the oxide thickness, t_{ox} can be initially estimated from the accumulation C_{gg} data, without PAE and QME, using the unified regional accumulation piece while taking $N_{ch} = n_i$. (Extracted $t_{ox} = 2.24 \text{ nm}$)

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Step 2: The initial value can be extracted by following the level and shape of the given C_{gg} data in depletion using the unified regional solution. A simple rule-of-thumb is by examining the minimum value of the model and C_{gg} data. At this stage, the correct turn-on of the threshold voltage is ignored. (Extracted $N_{ch} = 4.8 \times 10^{17} \text{ cm}^{-3}$; close to known value.)

Step 3: The poly-gate doping concentration is estimated from the strong inversion capacitance. Once the poly-gate doping is extracted, the threshold voltage is likely to be close to the given data. (Extracted $N_{gate} = 5.51 \times 10^{19} \text{ cm}^{-3}$; close to known value.)

Step 4: The QME effect can be included by matching the turning on of the threshold voltage. (Extracted $\kappa^{qm} = 0.77$.)

Step 5: The t_{ox} is re-tuned to match the data at strong inversion. The procedure from step 2 to step 5 can be repeated until the model is matched to data with satisfactory accuracy. (Extracted $t_{ox} = 2.0 \text{ nm}$; same as known value.)

Step 6: The remaining two parameters are fitting parameter that are introduced due to coupled QME and poly-gate effects. β^{qm} is extracted to match the level in accumulation. (Extracted $\beta^{qm} = 1.35$)

Step 7: ϵ^{qm} is extracted to match the transition at the V_{FB} . (Extracted $\epsilon^{qm} = 3.45 \times 10^{-2}$)

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The physical effects of each tuning step are shown in Figure 4.5.

Figure 4.5: Device parameters extraction procedure: (a) Step 1 until step 6, (b) Step 7.

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The model smoothing parameters are tuned to match the charge neutrality condition at the flat-band voltage, V_{FB} after the device parameters are extracted ($t_{ox} = 2.0$ nm, $N_{ch} = 4.8 \times 10^{17} \text{ cm}^{-3}$, $N_{gate} = 5.51 \times 10^{19} \text{ cm}^{-3}$, $\kappa^{qm} = 0.77$, $\beta^{qm} = 1.35$ and $\epsilon^{qm} = 3.45 \times 10^{-2}$), the first two smoothing parameters, σ_a and σ_f are to be calibrated at high poly-Si gate doping concentration, $N_{gate} = 10^{22} \text{ cm}^{-3}$.

Step 1: Assuming the smoothing parameters are identical $\sigma_f = \sigma_a$. Then select a σ value that will not produce any visible glitch in the second derivative of gate charge (i.e., dC_{gg}/dV_{gb}) compared with the default value of 2×10^{-3} . For example, after tuning, $\sigma = \sigma_f = \sigma_a = 5e^{-3}$, as shown in Figure 4.6.

Figure 4.6: Select an initial value of the model smoothing parameters that give smooth transition near V_{FB} . Inset: the second order derivative of gate charge of the fitted value.

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Step 2: Adjust σ_f so that charge neutrality is fulfilled at flat-band voltage. After tuning,

$$\sigma_f = 7.25e^{-2}, \sigma_a = 5e^{-3}.$$

Step 3: Fine-tune σ_a so that no visible glitch in the second derivative of gate charge.

Repeat steps 2 and 3 until both charge neutrality and C_{gg} smoothness are achieved.

After tuning, $\sigma_f = 9.29e^{-3}$, $\sigma_a = 6.0e^{-3}$.

The calibration from step 1 to step 3 is shown in Figure 4.7.

Figure 4.7: Smoothing parameter calibration at V_{FB} to achieve charge neutrality and smooth transition.

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Step 4: Similar to the above 3-step procedure, the smoothing parameters (ρ_a, ρ_f) of the poly-Si gate at the extracted N_{gate} are extracted. After calibration, $\rho_a = 9.34e^{-7}$, $\rho_f = 3.84e^{-2}$ as illustrated in Figure 4.8.

Figure 4.8: Calibration of smoothing parameter (ρ_a, ρ_f) at extracted N_{gate} .

Step 5: Fine-tune ϵ^{qm} to match the data near V_{FB} . Repeat the 3-step procedure for charge neutrality condition for both high and low poly-Si gate doping. ($\epsilon^{qm} = 6.93 \times 10^{-3}$, $\sigma_f = 1.15e^{-2}$, $\sigma_a = 5.76e^{-3}$, $\rho_a = 4.31e^{-5}$, $\rho_f = 1.07e^{-2}$)

The smoothing parameter, δ_ϕ is no longer needed to be adjusted, as the default value (0.002) is quite close to fit the measurement data, as shown in Figure 4.9 due to the improvement in the model equation with explicit Newton-Raphson iteration.

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Figure 4.9: No re-tuning of δ_ϕ due to good matching with the given data.

Figure 4.10 shows model play-back using known Medici physical parameters and the extracted parameters, which demonstrates both the physical meaning of parameters and the validity of the extraction procedure.

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Figure 4.10: Comparison of model using known device parameters and extracted parameters.

Both short-channel bulk charge (capacitance) and inversion charge (capacitance) are needed in short-channel charge calibration. Intrinsic short-channel effects have the major impact on the bulk-charge while the extrinsic capacitances are mainly seen in the inversion charge. Using the same technology with the long channel, a set of short-channel gate-bulk capacitance and gate-inversion capacitance are generated to illustrate the extraction.

From gate-inversion capacitance, there are also five fitting parameters to be extracted:

Step 1: Extract the fitting parameter for junction lateral spread, σ ($L_{ov} = \sigma x_j$ is the length of the overlap region) to match the minimum capacitance in strong accumulation. ($\sigma = 0.41$)

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Step 2: Extract the fitting parameter for potential-barrier lowering from depletion-to-strong inversion in depletion region. ($\alpha_{ds} = 0.13$).

Step 3: Extract the fitting parameter for poly-Si gate bulk-charge sharing model in strong inversion. ($\lambda_p = 4.31$).

Step 4: Extract the source-drain doping concentration to match the transition near the threshold voltage. ($N_{sd} = 5.6 \times 10^{20} \text{ cm}^{-3}$)

Step 5: Extract the overlap region bulk-charge sharing in strong accumulation. ($\lambda_{ov} = 0.29$).

The above extraction procedure is illustrated in Figure 4.11.

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Figure 4.11: Short channel parameters extraction from inversion capacitance.

There are three fitting parameters to be extracted from the gate-bulk capacitance:

Step 1: Extract the accumulation region C_{bg} using the potential-barrier lowering fitting parameter, α_{acc} from the data near the flat-band voltage. ($\alpha_{acc} = 0.35$)

Step 2: Extract the depletion region C_{bg} using the bulk-charge sharing fitting parameter, λ from the data in depletion region. ($\lambda = 0.95$)

Step 3: Extract the strong accumulation C_{bg} with external fringing pre-factor. ($v_b = 0.26$)

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The above extraction procedure is illustrated in Figure 4.12.

Figure 4.12: Short channel extraction fitting parameters extraction for gate-bulk capacitance.

Figure 4.13 shows the final extracted short-channel C_{gg} model compared with Medici data.

Figure 4.13: Model playback for extracted short channel capacitance, including QME and Poly-Si gate effect.

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4.2. URSP bulk-MOSFET charge model verification

The measurement data of a long/wide channel device, from a 0.11- μm technology node is used to verify the extraction procedures. The extraction is done using only one C_{gg} data, and play-back to predict the C_{gc} data, with the extracted physical parameters listed in the figure. It can be seen that the model is able to follow the data very well.

Figure 4.14: Physical parameter extraction using C_{gg} data and model prediction for different body biases using C_{gc} data.

For short-channel charge model verification, short-channel Medici structures are used. The model is first validated with only short-channel model included (i.e., ignoring QME and poly-Si gate effects) and then the final full model. The following model parameters are extracted from the simulated data, as listed in

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Table 1.

Table 1: Extracted value of model parameters from capacitance

MODEL	MODEL	EXTRACTED
<i>Oxide thickness</i>	$t_{ox} (cm)$	2×10^7
<i>Effective channel doping concentration</i>	$N_{ch} (cm^{-3})$	4.8×10^{17}
<i>Effective Poly-gate doping</i>	$N_{gate} (cm^{-3})$	5.5×10^{19}
<i>Effective source/drain doping concentration</i>	$N_{sd} (cm^{-3})$	5.6×10^{20}
<i>Junction lateral diffusion spread parameter</i>	σ	0.41
<i>Poly-gate charge sharing parameter</i>	λ_p	4.31
<i>Quantum mechanical correction</i>	κ^{qm}	0.68
<i>Quantum poly-gate gain factor</i>	β^{qm}	1.35
<i>Quantum adjustment at accumulation</i>	ϵ^{qm}	6.93×10^{-3}
<i>Overlap region doping concentration</i>	$N_{ov} (cm^{-3})$	8.0×10^{19}
<i>Overlap charge sharing parameter</i>	λ_{ov}	0.29
<i>Extrinsic inversion charge gain factor</i>	v_i	9.16×10^{-1}
<i>Extrinsic bulk charge gain factor</i>	v_b	0.26
<i>Complimentary function parameter for surface potential</i>	σ_a	7×10^{-3}
	σ_f	1.15×10^{-2}
<i>Complimentary function parameter for gate surface potential</i>	ρ_a	4.31×10^{-5} ,
	ρ_f	1.07×10^{-2}
<i>Complimentary function parameter for overlap surface potential</i>	σ_{aov}	2.15×10^{-1}
	σ_{fov}	3.39×10^{-1}

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<i>Smoothing function parameter from depletion to strong inversion for surface potential</i>	δ_ϕ	6.62×10^{-4}
<i>Potential barrier lowering (charge)</i>	α_{acc}	0.35
	α_{ds}	1.31×10^{-1}
<i>Bulk-charge sharing (charge)</i>	λ	9.5×10^{-1}

Figure 4.15 shows the model playback of C_{bg} and C_{gg} of a short-channel data for different imref-split, namely $V_{ds}=0, 0.6$, and 1.2 V. The geometry prediction and body-bias variation of the model is shown in Figure 4.16. The model is able to predict the intermediate gate lengths and body-bias variations quite well.

Figure 4.15: Charge model validation for different imref split, $V_{ds}=0, 0.6$ and 1.2 V (ignoring QME and Poly-Si gate effects).

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Figure 4.16: Charge model validation with different lengths for different body biases (ignoring QME and Poly-Si gate effects).

Finally, the full-model (i.e., with QME and poly-Si gate effects) is validated by comparing the C_{gg} for a few gate lengths, $L_g = 10, 0.5$, and $0.25 \mu\text{m}$, as shown in Figure 4.17. The model is able to predict the scaling of the capacitance with reasonable accuracy.

Figure 4.17: Model scalability to intermediate effective lengths, $L_g = 0.25$ and $0.5 \mu\text{m}$ and model play-back to $L_g = 10 \mu\text{m}$, with both QME and poly gate effect.

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Other model quality tests, such as the reciprocity test is also shown in Figure 4.18. The model is passing the reciprocity test at $V_{ds}=0$.

a

b

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Figure 4.18: Reciprocity test on capacitances at $V_{ds} = 0V$, sweeping a.) V_{ds} and sweeping b.) V_{gb} at $L_g = 90$ nm

4.3. URSP bulk-MOSFET current model extraction

The bulk-MOSFET drain current model has been formulated with the essential physical effects built-in. SCE, poly-Si gate effects, and QME are included in the surface potential formulation, which account for the respective effects in the charge and drain current models. Besides, the lateral-field effects such as lateral-field mobility degradation and channel-length modulation are modeled through velocity saturation and effective Early voltage, respectively. All these effects are included in the modular form and, thus, can be turned off individually. Special care is given to ensure the fulfilment of the Gummel symmetry requirement through bulk-referenced core formulation and bulk-charge symmetric linearization. The internal model evaluation includes the condition for $V_{ds} < 0$ and thus has the possibility to include asymmetry in source/drain in the same core model. This section discusses the model calibration for conventional bulk-MOSFET for both long- and short-channel effects.

The physical parameters, such as N_{ch} , N_{gate} , t_{ox} , QME parameters, and unified regional smoothing parameters, are properly calibrated in the long-channel charge model. Similarly, for the drain current calibration procedure, it begins with the extreme geometry conditions, long/wide device. Some of the parameters are first set to the default values: $\zeta_n = 0.5$ and $\zeta_b = 1$ (in E_{eff}), $v_{sat} = 1.15 \times 10^7$ cm/s, and $\delta_s = 0.0005$

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(in $V_{ds,eff}$), $\delta_L = 0.01$ (lateral mobility degradation), $\alpha_{dibl} = 2.4 \times 10^{-3}$ (long channel DIBL).

Step 1: At W_{∞}/L_{∞} : Measure I_{ds} vs. V_{gs} data at $V_{ds} = V_{d0}$ (linear region, $V_{d0} = 0.05$ V) and $V_{bs} = 0$ V. Fit the drain current equation in the strong inversion region, where transverse field mobility parameters (μ_1, μ_2, μ_3, ν) are tuned to match the data. The transverse-field mobility model is repeated here for convenience

$$\mu_0 = \frac{\mu_1}{1 + \left(\frac{\mu_1}{\mu_2} \right) E_{eff}^{\frac{1}{3}} + \left(\frac{\mu_1}{\mu_3} \right) E_{eff}^{\nu}}. \quad (4.1)$$

The effect of the step-1 extraction is illustrated in Figure 4.19.

Figure 4.19: Model calibration for transverse field mobility parameters.

Step 2: At W_{∞}/L_{∞} : Measure I_{ds} vs. V_{gs} data at $V_{ds} = V_{d0}$ (linear region, $V_{d0} = 0.05$ V) and $V_{bs} = V_{bb} = -V_{dd}$ V. Fit the drain current equation in the strong inversion region, where ζ_b is tuned to match the data. The effective electric transverse field model is repeated here for convenience

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$$E_{eff} = \frac{\zeta_n C_{ox}}{\epsilon_{si}} \left(\bar{q}_i + \frac{\zeta_b}{\zeta_n} \gamma \sqrt{\phi_s} \right). \quad (4.2)$$

When ζ_b is adjusted, it is likely it will also affect the zero V_{bs} condition. Therefore, ζ_n is re-tuned to fit the data at $V_{bs} = V_{b0}$ condition. Step 1 and step 2 can be repeated until both I_{ds} vs. V_{gs} at $V_{ds} = V_{d0}$, $V_{bs} = V_{b0}$ and V_{bb} match the data. This two-step iterations procedure is illustrated in Figure 4.20.

Figure 4.20: Model calibration for (a) ζ_b that also cause the modification at $V_{bs} = 0$ V, (b) repeated step 1 and step 2 until both $V_{bs} = 0$ and -1.2 V matched the data.

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Step 3: At W_{∞}/L_{min} , measure I_{ds} vs. V_{gs} data, $V_{bs} = V_{b0}$ at $V_{ds} = V_{d0}$. Fit the drain current equation in the linear region for source/drain series resistance extraction. The series resistance model is repeated here for convenience

$$R_s = R_{ext} + R_{int} = r_1 + \frac{r_2}{C_{ox} W \left(\overline{V}_{gt} + \overline{A}_b v_{th} \right)} . \quad (4.3)$$

The extracted value of the fitting parameters are $r_1 = 1.01 \, \Omega$ and $r_2 = 0.22 \, \Omega$, respectively. Figure 4.21 shows the effect with the fitted series resistance.

Figure 4.21: Series resistance extraction at shortest channel length.

Step 4: At W_{∞}/L_{min} , measure I_{ds} vs. V_{ds} data at $V_{gs} = V_{gg}$ and $V_{bs} = V_{b0}$. Fit the drain current equation in saturation to extract v_{sat} and ξ . Adjust the δ_s at the transition from linear to saturation to match the data smoothly. Figure 4.22 shows the extraction of these three parameters from I_{ds} vs. V_{ds} at $V_{gs} = V_{dd} = 1.2 \, \text{V}$.

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Figure 4.22: Extraction of δ_s , v_{sat} and ξ from I_{ds} - V_{ds} at $V_{gs}=1.2$ V, $V_{bs}=0$ V.

Step 5: At W_∞/L_{min} , measure I_{ds} vs. V_{ds} data at $V_{gs} = V_{gg}$ and $V_{bs} = V_{bb}$. Fit the drain current equation at saturation with δ_L . Figure 4.23 shows the extraction of the δ_L .

Figure 4.23: Extraction of δ_L at low body bias.

Step 6: At W_∞/L_{min} , measure I_{ds} vs. V_{gs} data, $V_{bs} = V_{b0}$ at $V_{ds} = V_{dd}$. Fit the threshold voltage with I_α .

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Step 7: At W_{∞}/L_{min} , measure I_{ds} vs. V_{gs} data, $V_{bs} = V_{bb}$ at $V_{ds} = V_{dd}$. Fit the threshold voltage with λ .

Steps 6-7 sometimes may need iteration to achieve a reasonable match with the data, as shown in Figure 4.24.

Figure 4.24: Extraction of the short channel coefficients at low V_{bs} , high V_{ds} .

Step 8: At W_{∞} : Measure r_{out} vs. V_{ds} data at $V_{gs} = V_{gg}$, $V_{bs} = V_{b0}$. For each gate length, δ_s is tuned to fit the data to match the level of output resistance. A length-dependent δ_s model is proposed as

$$\delta_s = \delta_{s1} + \frac{(\delta_{s2} - \delta_{s1})}{1 + \left(\frac{L_g}{\delta_{s3}}\right)^{\delta_{s4}}}. \quad (4.4)$$

Figure 4.25 show the comparison of length-depended δ_s and single δ_s in r_{out} of a single technology. Figure 4.26 shows the model match using (4.4) and extracted δ_s of the corresponding gate lengths.

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Figure 4.25: Extraction of length dependent δ_s from the output resistance.

Figure 4.26: Model fitting using the geometry dependent, δ_s

With this, the length dependent drain current model calibration is completed.

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4.4. URSP bulk-MOSFET current model verification

Table 2 shows the extracted values used in the drain-current model for bulk-MOSFET to generate the data for validation used in this work. It is shown that the total number of parameters needed for the drain current model is only 22 for the whole technology. Other parameters are calibrated with the charge model. The model is played back and compared with Medici data for the whole technology with gate lengths $L_g=10, 1, 0.5, 0.25, 0.18, 0.09 \mu\text{m}$, for the currents and first derivatives, as shown in Figure 4.27 to Figure 4.62.

Table 2: Extracted value of model parameters from drain current

MODEL	MODEL	EXTRACTED
<i>Mobility</i> μ_2 dimension: ($\text{cm}^{5/3}/\mu^{-1/3}\text{V}^{2/3}.\text{s}$) μ_3 dimension: ($\text{cm}^{2-\nu}/\mu^{-\nu}\text{V}^{\nu-1}.\text{s}$) (where $\mu = 10^{-6}$)	μ_1 ($\text{cm}^2/\text{V}.\text{s}$)	365
	μ_2	684
	μ_3	389
	ν	1.73
<i>Lateral field mobility degradation</i>	δ_L	8.15×10^{-4}
<i>Long channel drain induced barrier lowering</i>	α_{dibl}	1.25×10^{-4}
<i>Effective transverse field</i>	ζ_n	0.43
	ζ_b	3.24
<i>Velocity saturation</i>	v_{sat} (cm/s)	6.68×10^6
<i>Lengths dependent smoothing function for effective drain-source voltage</i>	δ_{s1}	2.65×10^{-3}
	δ_{s2}	8.10×10^{-2}
	δ_{s3}	0.29
	δ_{s4}	1.28

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<i>Bulk-charge sharing (current)</i>	λ	0.56
<i>Potential barrier lowering (current)</i>	α_{acc}	1.12×10^{-2}
	α_{ds}	2.70×10^{-3}
<i>Gain factor for Gaussian peak</i>	κ	1.29
<i>Characteristic lengths</i>	l_{β}	2.90×10^{-2}
<i>Gaussian's peak location</i>	l_{μ} (cm)	-1.10×10^{-2}
<i>Series resistance</i>	r_1 (Ω/cm)	1.05
	r_2 ($\Omega \cdot \text{V}/\text{cm}$)	0.14
<i>Channel length modulation</i>	ξ	1×10^{-3}

The model data are generated and compared with Medici simulations and arranged in the following manner:

- I_{ds} vs. V_{gs}
 1. The first figure at longest channel, i.e., $L_g = L_{max} = 10 \mu\text{m}$, Figure 4.27 depicts the I_{ds} vs. V_{gs} in linear mode, i.e., $V_{ds} = V_{d0} = 0.05 \text{ V}$ for different body biases, namely, $V_{bs} = 0, -0.4, -0.8$, and -1.2 V , shown for both linear- and log-scale.
 2. The second figure (i.e., Figure 4.28) shows the corresponding g_m vs. V_{gs} of the first figure in linear mode.
 3. Similar condition to the first figure but in saturation mode, i.e., $V_{ds} = V_{dd} = 1.2 \text{ V}$ (e.g., Figure 4.29).
 4. The corresponding g_m vs. V_{gs} of third figure in saturation mode (e.g., Figure 4.30).
- I_{ds} vs. V_{ds}

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5. The fifth figure (e.g., Figure 4.31) (a) depicts the output drain-current for $V_{gs} = 0.4, 0.8,$ and 1.2 at zero body bias, $V_{bs} = 0$. Fifth figure (b) shows the corresponding g_{ds} .
6. Similar to the fifth figure (e.g., Figure 4.32) but at low body bias, i.e., $V_{bs} = -1.2$ V.

These six figures are then repeated for $L_g = 1, 0.5, 0.25, 0.18,$ and $0.09 \mu\text{m}$.

Figure 4.27: Validation by model play-back and predict for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=10 \mu\text{m}$.

Figure 4.28: Validation by model play-back and predict for g_{ms} vs. V_{gs} at various body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=10 \mu\text{m}$.

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Figure 4.29: Validation by model play-back and prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=10$ μm .

Figure 4.30: Validation by model play-back and predict for g_{ms} vs. V_{gs} at various body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=10$ μm .

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(a)

(b)

Figure 4.31: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs}=0$ for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g=10 \mu m$.

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(a)

(b)

Figure 4.32: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4$, 0.8 and 1.2 V. (b) The corresponding drain conductance, g_{ds} at $L_g = 10$ μm .

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Figure 4.33: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=1$ μ m.

Figure 4.34: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=1$ μ m.

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Figure 4.35: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds} = 1.2 \text{ V}$, $L_g = 1 \mu\text{m}$.

Figure 4.36: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds} = 1.2 \text{ V}$, $L_g = 1 \mu\text{m}$.

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(a)

(b)

Figure 4.37: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs}=0$ for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g=1 \mu\text{m}$.

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(a)

(b)

Figure 4.38: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4$, 0.8 and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g = 1 \mu\text{m}$.

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Figure 4.39: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.5$ μm .

Figure 4.40: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.5$ μm .

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Figure 4.41: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.5$ μm .

Figure 4.42: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.5$ μm .

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(a)

(b)

Figure 4.43: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs}=0$ for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g=0.5 \mu\text{m}$.

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(a)

(b)

Figure 4.44: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4$, 0.8 and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g = 0.5$ μm .

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Figure 4.45: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.25$ μm .

Figure 4.46: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.25$ μm .

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Figure 4.47: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.25$ μm .

Figure 4.48: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.25$ μm .

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(a)

(b)

Figure 4.49: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs}=0$ V for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g=0.25$ μm .

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(a)

(b)

Figure 4.50: Validation by model prediction for I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4, 0.8$ and 1.2 V. Inset: The corresponding drain conductance, g_{ds} , at $L_g = 0.25$ μm.

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Figure 4.51: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.18$ μm .

Figure 4.52: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05$ V, $L_g=0.18$ μm .

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Figure 4.53: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.18$ μm .

Figure 4.54: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.18$ μm .

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(a)

(b)

Figure 4.55: Validation by model prediction for I_{ds} vs. V_{ds} at $V_{bs}=0$ V for various $V_{gs} = 0.4, 0.8$ and 1.2 V. Inset: The corresponding drain conductance, g_{ds} , at $L_g=0.18$ μ m.

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(a)

(b)

Figure 4.56: Validation by model prediction for I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4, 0.8$ and 1.2 V. Inset: The corresponding drain conductance, g_{ds} , at $L_g = 0.18$ μm.

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Figure 4.57: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05\text{V}$, $L_g=0.09 \mu\text{m}$.

Figure 4.58: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at linear region, $V_{ds}=0.05\text{V}$, $L_g=0.09 \mu\text{m}$.

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Figure 4.59: Validation by model prediction for I_{ds} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.09$ μm .

Figure 4.60: Validation by model prediction for g_{ms} vs. V_{gs} at body biases, $V_{bs} = 0, -0.4, -0.8$ and -1.2 V at saturation region, $V_{ds}=1.2$ V, $L_g=0.09$ μm .

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(a)

(b)

Figure 4.61: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs}=0$ for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g=0.09$ μm.

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(a)

(b)

Figure 4.62: Validation by model prediction for: (a) I_{ds} vs. V_{ds} at $V_{bs} = -1.2$ V for various $V_{gs} = 0.4, 0.8$ and 1.2 V. (b) The corresponding drain conductance, g_{ds} , at $L_g = 0.09$ μm .

It is worth noting that not all data for every gate-length device in a technology are used for parameter extraction. Only a limited set of data are used, which is a set of six I_{ds} vs. V_{gs} (at longest length, i.e., $L_g = L_{max}$, both $V_{bs} = V_{b0}$ and V_{bb} ; at shortest length, i.e., $L_g = L_{min}$, $V_{bs} = V_{b0}$ and V_{bb} for $V_{ds} = V_{d0}$ and V_{dd}), two I_{ds} vs. V_{ds} (at $L_g = L_{min}$, $V_{bs} = V_{b0}$ and V_{bb}) and r_{out} vs. V_{ds} for all lengths.

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For a model to be able to effectively fit the data, there are some basic sequence of calibration that needs to be done. For example, the long channel parameters (e.g., t_{ox} , N_{ch} , N_{gate} , mobility parameters etc.) need to be calibrated for a particular technology before the model can fit the short channel data. In this work, the main purpose of using Medici data is to demonstrate the idea and the feasibility of the URM approach. In addition, Medici and compact model start with the same physical equations, so model fit to Medici data can validate (or invalidate) compact model approximations and effectiveness of the fitting parameters. A good and physical model should be applicable to measured devices in the same or similar way as fitting to numerical data. The developed model has been applied to measured 90nm technology data. To demonstrate that this model should be feasible to fit the larger saturated slope, the data for 0.097 μm MOS provided by the reviewer were digitized and fitted, based on $V_{gs}=1.2\text{ V}$, as shown below.

Figure 4.63: The model fit of a larger saturation slope for $V_{gs}=1.2\text{V}$ from a digitized figure of a 0.097 μm MOS provided by reviewer.

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In this work, all the rest of the data are predictions generated from the model. The modeled data match very well with the all Medici data. Therefore, the model is very physical and able to scale for the different geometries and voltages.

4.5. Chapter 4 conclusions

The model extraction procedure is done systematically due to uncorrelated model parameters. The physical parameters, such as the t_{ox} , N_{ch} , N_{gate} , k^{qm} , β^{qm} , and ϵ^{qm} , are able to be extracted from accumulation, depletion, and strong inversion of the C_{gg} capacitance data at long channel. The model smoothing parameters (σ_a , σ_f , δ_{s1} , and δ_{s2}) are being optimized from the derivatives of C_{gg} and V_{FB} condition. The mobility models (μ_1 , μ_2 , μ_3 , and ν) are extracted from long-channel I_{ds} vs. V_{gs} at linear condition. The long-channel body effect can be included with the optimization of the transverse field, E_{eff} (ζ_n , ζ_b) at low body bias. If short-channel charge model is available, then short-channel effect and reverse short-channel effect (α , l_β , κ , λ , λ_p) can be effectively optimized through the short-channel capacitance data. Alternatively, technology characterization approach can be used to extract these model parameters using “pinned” V_t model to match the V_t vs. L_g curves [37]. The series resistances (r_1 , r_2) are to be extracted from the short-channel I_{ds} vs. V_{gs} in linear region. v_{sat} and ξ can be optimized to match the short-channel I_{ds} vs. V_{ds} . At each gate length in a single technology, δ_s parameter can be extracted from the r_{out} vs. V_{ds} at $V_{gs}=V_{gg}$, after which

CHAPTER 4: Extraction of Physical and Model Parameters

these data can be fitted with empirical function (4.4). For short-channel body effect, δ_L can be used to optimize the model for low body bias output current.

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CHAPTER 5: Conclusions and Recommendations for Future Work

The scalable modeling approach developed in this work is not only applicable for n-channel MOSFET (NMOS), it is also possible to be extended to p-channel MOSFET (PMOS). One can easily follow the sequence of development of the compact model as shown in this work, beginning from the surface potential, then to current, and finally to charge model. Alternatively, one may reuse the existing NMOS equation but with additional sign change options that can switch for PMOS. For example, the input voltage equation with an appropriate sign change option can be written as follows:

$$V_{gs} - V_{FB} - \phi_s = \text{sgn}(\phi_s) \gamma \sqrt{v_{th} \left(e^{\frac{-(TYPE)\phi_s}{v_{th}}} - 1 \right) + (TYPE)\phi_s \left(1 - e^{\frac{-(TYPE)(V_{gs} + 2\phi_F)}{v_{th}}} \right) + v_{th} e^{\frac{-(TYPE)(V_{gs} + 2\phi_F)}{v_{th}}} \left(e^{(TYPE)\left(\frac{\phi_s}{v_{th}}\right)} - 1 \right)} \quad (5.1)$$

where

TYPE = 1 for NMOS,

TYPE = -1 for PMOS.

Other areas that may need further adjustments for PMOS model are the semi-empirical mobility model and quantum mechanical correction. The work presented in this thesis is a summary of an approach that allows one to develop a physics-based scalable compact model. This chapter summarizes the conclusion, possible enhancements and recommendations for future work in the related study.

CHAPTER 5: Conclusions and Recommendations

5.1. Conclusions

In this work, a simpler and more computational efficient unified regional surface potential solution that is based on Taylor-expansion up to the third-order has been derived for strong accumulation/inversion in bulk-MOSFET. Similar to depletion region, a weak accumulation model can be derived before the surface potential saturates. Unified regional strong accumulation and inversion solutions are subsequently combined with the weak accumulation and depletion model, respectively, to form the unified regional accumulation, depletion-to-strong inversion surface potential that is valid for all doping concentration.

Unlike other surface potential based drain-current models that require accurate surface potential solution for correct subthreshold behaviour, this work managed to do away the accuracy requirement by taking the induced charge terms in the transverse electric field and combined with “pinned”-like effective drain-source voltage for the drain-current model.

In addition, to achieve perfect Gummel symmetry, bulk-terminal referenced formula is introduced for the evaluation of saturation voltages, which separates the saturation voltages at the drain and source sides, respectively. According to this method, the drain and source terminals are not swapped for negative V_{ds} in the model evaluation. The drain and source terminals are treated as node labels rather than by convention, the latter always assumes higher potential at the drain than the source

CHAPTER 5: Conclusions and Recommendations

terminal (for NMOS). The same idea has been extended for all short-channel models with bulk-referenced formulism, namely, bulk-charge sharing, quasi-2D potential-barrier lowering, channel-length modulation, and other effects such as the poly-Si gate and quantum-mechanical effects.

AC charge model has been extended to include the short-channel effects, namely, the intrinsic and extrinsic short-channel effects. The intrinsic short-channel effects are mainly due to bulk-charge sharing and potential-barrier lowering, and extrinsic short-channel effects consist of the overlap charges and fringing charges. Bulk-charge sharing effects have also been applied to both overlap charges and poly-Si gate effect. Quantum-mechanical and poly-Si gate effects that are coupled with short-channel effects have also been formulated.

Unified regional surface potential approach has been extended to common asymmetric double-gate MOSFET model. The regional surface potentials for common gate configuration have been derived, namely, the strong accumulation, weak accumulation (also known as “conjugated” depletion), depletion (for high channel doping concentration and thick body), volume inversion, and strong inversion. The zero-field potential is derived with the explicit function of unified regional surface potential. The induced charge terms are used for the evaluation of the effective drain-source voltage. With this, the model does not require a highly accurate surface potential solution for the drain-current model.

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Finally, the calibrations for both charge and drain current are explained. It has been demonstrated that both the charge and drain current models are able to fit and predict for different geometries and biases physically. Though the demonstration is only done on the bulk-MOSFET, similar extraction procedure can be readily applied to double-gate MOSFET model.

5.2. Recommendations

The following works are recommended for future study.

- i. Both the charge and drain current models developed in this work are limited to the length dimension. The narrow width effect has been ignored. This work can be continued to include the narrow width effect according to the URM approach that has been used.
- ii. Channel position dependent mobility and non-quasi-static modeling are not considered, which require integration across the channel. One of the best approaches is through channel segmentation, and the core model in this thesis provide an easy way for extension since the saturation effect is added on the intrinsic linear current (I_{ds0}) that can be called repeatedly through sub-circuit expansion approach.

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- iii. Other areas of interest to the semiconductor industry are the device lifetime, temperature effect, and high frequency effects. In this work, all of the secondary effects like short channel effects, poly-Si gate effects and quantum mechanical correction are included modularly and verified independently. As such, the compact model calibration can be done in an uncorrelated manner. Similarly, additional effects may be included regionally in the compact model modularly. The physics of additional effects like the device lifetime, temperature and high frequency may not yield any compact solution if considering the whole region of device operation. However, a simpler solution may be possible regionally and, therefore, can be unified accordingly as demonstrated in the URM approach.
- iv. The source/drain asymmetry capability of the model is not fully calibrated or understood in the current work. This work can be extended to study the impact of asymmetric source/drain, namely, the series resistance, saturation voltages on the transistor performance.
- v. The double-gate model can be extended to include the poly-gate effects. The model can be easily extended to include these effects following the bulk-MOSFET model.
- vi. Independent asymmetric double-gate MOSFET modeling is the most challenging to handle due to the two unknown implicit variables in the Poisson solution. The common asymmetric double-gate approach in solving the full-depletion voltage can be extended to cover all the possible cases in the independent double-gate MOSFET operation.

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- vii. Another emerging device is the strained-silicon MOSFET. The mobility modeling for strained-Si MOSFET should be studied and understood. With the unified regional approach, strained-Si surface potential model can be handled easily.

Author Publications

Journal

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