USIM Design Considerations

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ABSTRACT

Compact models need to be carefully designed to predict circuit behavior efficiently, accurately, and robustly. A variety of effects must work together and be consistently integrated into the model equations without redundancy.

USIM is a compact MOSFET model that has complete symmetry and handles all regions of operation seamlessly. A simple, implicit relation for inversion charge is used without the need for calculating the surface potential. No limiting for V_{ds} or unification functions for drift and diffusion current are needed. In USIM, the threshold condition is defined to occur when the normalized inversion charge is unity.

Keywords: MOSFET transistors, inversion change density, bulk charge, saturation voltage, channel voltage.

1 MODEL DESIGN GOALS

Compact model design involves a careful tradeoff between accuracy and efficiency. The model should be based on analog simulation requirements, which are usually more demanding than for digital simulations. Enough detail must be captured for circuit behavior to be well approximated, both quantitatively and qualitatively. Minimizing errors is important, but the model should also be transparent enough that designers trust the results. Typically MOSFETs are constructed with source/drain symmetry, and for such structures the model should reflect that symmetry. Models that give different results depending on which node is the source and drain (e.g. C_{gs} vs. C_{gd}) are much less credible.

On the other hand, the model implementation must be efficient and compact enough to allow large circuits to be simulated. The model equations should go as far as possible with clean physics, and then use well-behaved empirical relations. Including as much physics as appropriate is likely to improve the ease of extraction, reduce the number of fitting parameters, and make the model better at extrapolation or scaling. Simplifying assumptions should be tested against data and/or detailed numerical simulations. The model needs to be fast, smooth and robust. Depending on the simulator in which the model is to be used, overshoot avoidance or other implementation features may be needed to guarantee proper convergence in the simulator.

2 INVERSION CHARGE DENSITY

The first step in modeling MOSFET behavior is to determine the local charge. For constant substrate doping, the 1-D Poisson equation can be integrated to obtain the total silicon charge density[1]

$$q_{tot} = \gamma\sqrt{\psi_s + e^{\psi_s - 2\phi_F - v_{ch}}} \tag{1}$$

where v_{gb} is the (normalized) gate-to-bulk voltage, v_{ch} is the local channel voltage, and the other symbols have their usual meaning. This applies to a vertical slice in the interior of the channel, assuming that the gradual channel approximation holds. From charge balance

$$v_{gb} = v_{fb} + q_{tot} + \psi_s \tag{2}$$

For a given gate-to-bulk voltage v_{gb} and channel voltage v_{ch}, we can solve (1) and (2) for ψ_s and q_{tot}. The inversion charge density can then be calculated by

$$q_m = q_{tot} - \gamma\sqrt{\psi_s} \tag{3}$$

In this way we obtain q_m as a function of v_{gb} and v_{ch}. Such curves are shown by the symbols in Fig. 1.

Now, instead of plotting q_m, let us plot $(q_m-1)+\ln(q_m)$ vs. v_{ch} for constant v_{gb} as shown by the symbols in Fig. 2. It is seen that we obtain nearly straight lines of about unity slope. Positive values represent inversion and negative values subthreshold. As Fig. 2 suggests, we define the threshold condition to exist when the ordinate is zero, i.e. when q_m is unity.

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[1] To simplify equations, lower case voltages have been normalized by the thermal voltage kT/q and charge densities have been normalized by $-C_{ox}kT/q$.

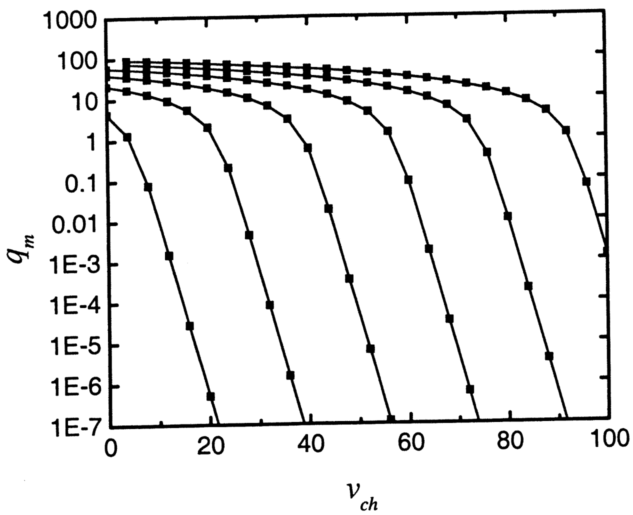

Figure 1: Inversion charge vs. channel voltage for V_{gb}= 0.5,1,...3V. Symbols from eq.(1-3); curves from eq.(5-6)

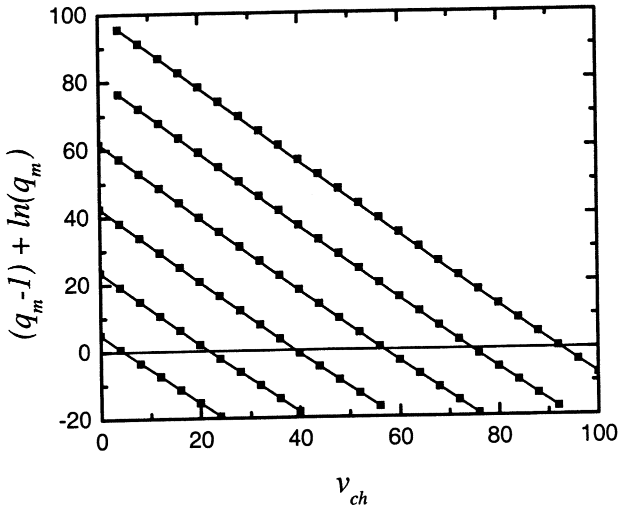

Figure 2: $(q_m-1)+ln(q_m)$ vs. v_{ch} for same conditions as Fig 1.

We define the saturation voltage v_{sat} as that value of v_{ch} at which the line for a given v_{gb} crosses zero. v_{sat} is a function of v_{gb}. This leads to the approximation

$$(q_m-1)+\ln(q_m) = v_{sat}-v_{ch} \tag{4}$$

Relations similar to this (in different notation) were obtained by several authors[1,2,3]. While Fig. 2 appears to show straight lines of unity slope, that is only approximately the case. A better approximation is given by the inversion charge relation[4,5]

$$h_x(q_m-1)+\ln(q_m) = v_{sat}-v_{ch} \tag{5}$$

where h_x is a correction factor depending only on v_{gb}. To represent the channel average of the inversion charge for short channel transistors, a factor t_e (of order unity) is needed in front of the $ln(q_m)$ term to model Gate Induced Barrier Lowering. GIBL reduces the subthreshold slope in short channel transistors. The saturation voltage is given by

$$v_{sat}=v_{gb}-v_{fb}-1-q_b-\ln(2q_b+1)+\ln(\gamma^2)-2\phi_F \tag{6}$$

In Fig. 1, the symbols represent an exact solution to the 1-D Poisson equation and the curves are calculated using the inversion charge relation. That relation provides an excellent fit to the numerical and measured data and seamlessly captures both the subthreshold and inversion regimes.

The above discussion has been for the case of a constant doping profile. When the substrate doping is non-uniform, the bulk charge will be a different function of $(v_{gb}-q_m)$. To make (6) more general, the hole Fermi potential ϕ_F may be incorporated into the gamma term and into v_{fb}, using the gate-substrate work function difference ϕ_{ms} :

$$v_{sat}=v_{gb}-\phi_{ms}-1-q_b-\ln(2q_b+1)+\ln(\gamma_z^2) \tag{7}$$

where

$$\gamma_z^2 = \frac{2q\varepsilon_{si}n_i}{C_{ox}}\frac{q}{C_{ox}kT} \tag{8}$$

Eq. (5) still applies, although the correction factor h_x would be different and, depending on the doping profile, may be somewhat larger than for the constant doping case.

Several other compact models obtain the surface potential iteratively and then compute the inversion charge. However deep subthreshold q_m values depend very sensitively on ψ_s; thus very high accuracy is required for ψ_s. Instead, USIM obtains q_m directly and robustly by use of (5). As shown in the appendix of [4], a very efficient procedure has been developed to calculate q_m using one pass through about 20 lines of simple code. In this case, a clean physical approach is both fast and accurate.

3 DRAIN CURRENT

The drain current is proportional to $q_m\, dv_{ch}/dy$. By using the gradient of channel voltage, i.e. the gradient of the electron quasi-Fermi potential, the total current is obtained; there is no need to handle diffusion current and drift current separately. If q_m is expressed as a function of ψ_s, as is done in many approaches, then obtaining dv_{ch}/dy as a function of ψ_s becomes complicated. However in USIM, q_m varies along the channel as a function of only v_{ch}, and the derivative of q_m with respect to v_{ch} is h_x+t_e/q_m. Thus

$$q_m\frac{dv_{ch}}{dy} = \left(h_xq_m+t_e\right)\frac{dq_m}{dy} \tag{9}$$

which can be directly integrated along the channel. The result is the channel current relation[4]

$$I = \beta\left(h_x \frac{q_{ms}^2 - q_{md}^2}{2} + t_e\left(q_{ms} - q_{md}\right)\right) \tag{10}$$

where

$$\beta = \mu_{eff} C_{ox}\left(\frac{kT}{q}\right)^2 \frac{W}{L} \tag{11}$$

μ_{eff} represents an effective mobility, i.e. averaged along the channel length. In order to approximate the dependence of μ_{eff} on vertical electric field, we construct a (symmetrical) function using q_{ms} and q_{md}. Modeling horizontal mobility degradation is discussed in Section 5.

This channel current relation (10) covers all regions of operation smoothly, satisfies the MOSFET symmetry test [8, p.540], and is consistent with published work [2,6,7].

The channel current relation has various simple limits. In subthreshold, q_{ms} and q_{md} are much smaller than unity so it becomes

$$I = \beta t_e \left(q_{ms} - q_{md}\right) \tag{12}$$

and for $q_m << 1$, the (q_m-1) term of (5) can be dropped. Thus

$$I = \beta t_e e^{v_{sat}/t_e}\left(e^{-v_{sb}/t_e} - e^{-v_{db}/t_e}\right) \tag{13}$$

which has the proper symmetry.

Another consequence of (9) is that, because the current is constant, q_m can be calculated as a function of channel position. Integrating, this leads to analytical expressions for the quasi-static source and drain terminal charge and thus to clean capacitance formulas.[5]

4 THRESHOLD

In conventional MOSFET modeling, much emphasis has been placed on the threshold voltage. There is a large variety of definitions of threshold voltage and a multitude of extraction methods. However, what is important for a model in a circuit simulator is faithful representation of the whole current-voltage and capacitance characteristics. If the model does that, then the threshold voltage, however defined, is automatically represented properly.

There are cases where the threshold voltage is not representative of the whole *I-V* characteristic. Consider, for example, a set of pocket implanted devices of different lengths. The surface potential at small v_{ds} might look as shown in [9, see Fig. 3]. In deep subthreshold, the current is dominated by the end regions and tends to be independent of channel length, as pointed out in [10, see Fig. 1], while in the strong inversion regime the normal channel length dependence takes over.

In USIM, we fit the output *I-V* characteristics for transistors of a given channel length by a few key parameters, such as γ, ϕ_F, and v_{fb}, which determine v_{sat} directly, and also by β. By adjusting these parameters for low v_{ds} characteristics, good fits are typically obtained. If, in addition, conventional modeling of mobility is performed, then excellent fits are obtained. For channel length dependence, some or all of the key parameters can now be modeled empirically or by physically based reasoning as a function of *L*. While the fitting described above covers some short channel effects, additional modeling for DIBL and velocity saturation has to be performed.

5 SYMMETRY

For symmetrically constructed MOSFETs, symmetry is a fundamental property and should be designed into the model equations for smoothness, accuracy, and credibility.

There are two complementary approaches to developing compact model equations. One approach is to start from more basic physical equations and make appropriate simplifications to reduce computation time. For example, to simplify the integration along the channel leading to Eq. (10), an average value of the mobility is used.

Symmetry arises naturally when starting from more basic equations, but it also needs to be incorporated into the empirical equations. For example, maintaining symmetry may be difficult in conventional modeling approaches for the horizontal mobility and output conductance degradation. While the underlying physics is extremely complex, the observed behavior can be very simple, as seen in Fig. 3, which shows measured drain current at V_{gb}=3.6V normalized to the slope at the origin for various transistor lengths.

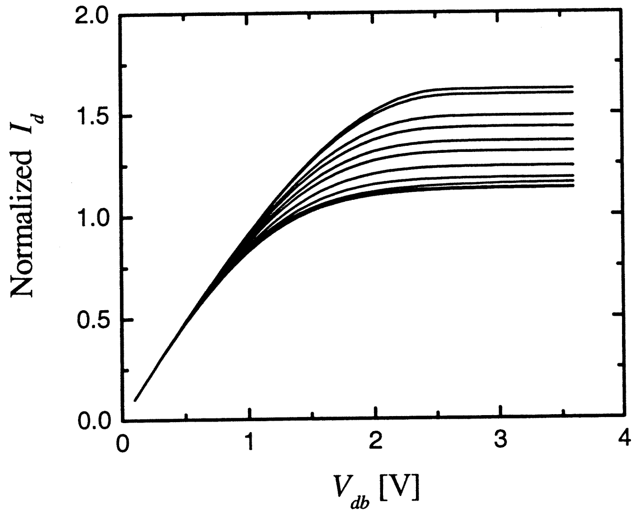

Figure 3: Output I_d-V_{db} data for several *L* from 0.28 to 15μm normalized to slope at $V_{db}=0$

To model this behavior, the quadratic term in the channel current relation (10) is replaced by

$$H = \frac{2}{f+4} \frac{q_{ms}^{f+4} - q_{md}^{f+4}}{q_{ms}^{f+2} + q_{md}^{f+2}} \tag{14}$$

where f is a parameter that depends upon bias and channel length. This function is shown in Fig. 4 for various values of f. Although empirical, this expression is simple, maintains full symmetry, and, for $f=0$, includes the original case without horizontal mobility degradation. It provides just enough flexibility to fit the data while keeping the number of parameters small.

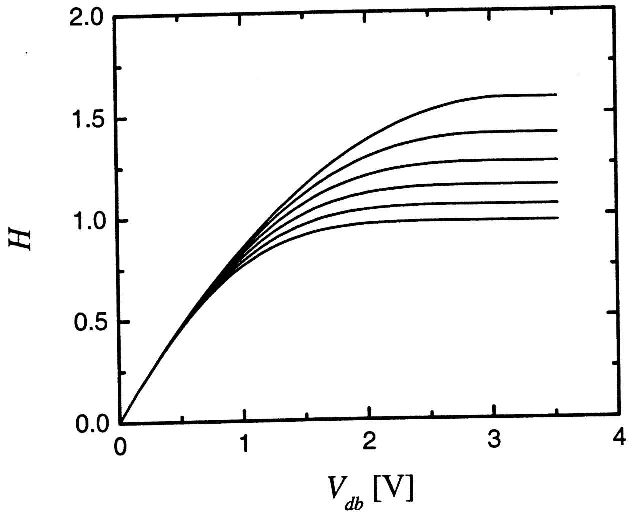

Figure 4: Eq. (12) results for f=0,0.5,...2.5

6 SUMMARY

The balance between accuracy and efficiency of a compact model is best achieved by using bulk-based symmetrical equations and staying with clean physics as long as possible. When short channel effects and channel engineering effects have to be modeled by empirical expressions, these too should represent the inherent source-drain symmetry. Seamless formulations that cover all operating regimes are preferable to regional approaches that are complicated by the need for artificial limiting functions and unification schemes.

In USIM, the inversion charge is obtained as a simple, through implicit, function of gate and channel voltage. As a result, both drain current and capacitive stored charges are expressed in terms of the inversion charge densities at the source and drain, q_{ms} and q_{md}, leading to a highly efficient, "understandable" model.

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