CHAPTER 5: Non-Quasi Static Model

5.1 Background Information

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

5.2 The NQS Model

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

5.3 Model Formulation

The channel of a MOSFET is analogous to a bias dependent RC distributed transmission line (Figure 5-1a). In the Quasi-Static (QS) approach, the gate capacitor node is lumped with both the external source and drain nodes (Figure 5-1b). This ignores the finite time for the channel charge to build-up. One Non-Quasi-Static (NQS) solution is to represent the channel as *n* transistors in series (Figure 5-1c). This model, although accurate, comes at the expense of simulation time. The NQS model in BSIM3v3.2.2 was based on the circuit of Figure 5-1d. This Elmore equivalent circuit models channel charge build-up accurately because it retains the lowest frequency pole of the original RC network (Figure 5-1a). The NQS model has two parameters as follows. The model flag, nqsMod, is now only an element (instance) parameter, no longer a model parameter. To turn on the NQS model, set nqsMod=1 in the instance statement. nqsMod defaults to zero with this model off.

Name	Function	Default	Unit
nqsMod	Instance flag for the NQS model	0	none
elm	Elmore constant	5	none

Table 5-1. NQS model and instance parameters.

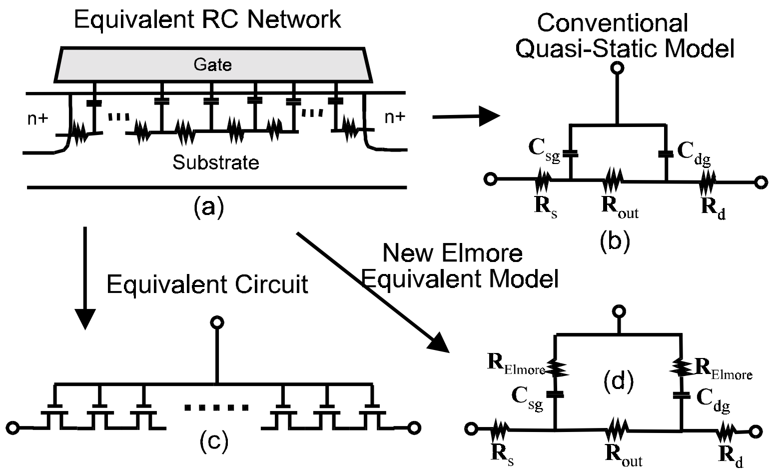

Figure 5-1. Quasi-Static and Non-Quasi-Static models for SPICE analysis.

5.3.1 SPICE sub-circuit for NQS model

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

$$Q_{def}(t) = V_{def} \times (1 \cdot C_{fact}) \tag{5.3.1}$$

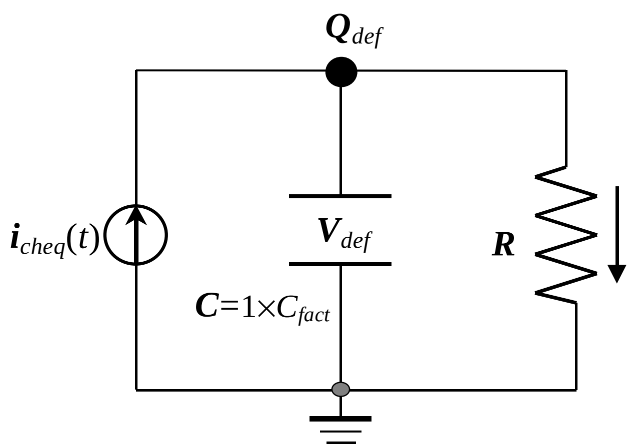

Figure 5-2. NQS sub-circuit for SPICE implementation.

5.3.2 Relaxation time

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

$$\frac{1}{\tau} = \frac{1}{\tau_{diff}} + \frac{1}{\tau_{drift}} \tag{5.3.2}$$

R_{elm} in strong inversoin is calculated from the channel resistance as

$$R_{elm} = \frac{{L_{eff}}^2}{elm \cdot \mu_0 Q_{ch}} \approx \frac{{L_{eff}}^2}{elm \cdot \mu_0 Q_{cheq}} \tag{5.3.3}$$

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

(5.3.4)

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

τ_{diff} has the form of

(5.3.5)

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

5.3.3 Terminal charging current and charge partitioning

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

(5.3.6)

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

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

$$Q_{def}(t) = Q_{cheq}(t) - Q_{ch}(t) \tag{5.3.7}$$

and

$$\frac{\partial Q_{def}(t)}{\partial t} = \frac{\partial Q_{cheq}(t)}{\partial t} - \frac{Q_{def}(t)}{\tau} \tag{5.3.8a}$$

$$\frac{\partial Q_{d,g,s}(t)}{\partial t} = D,G,S_{xpart} \frac{Q_{def}(t)}{\tau} \tag{5.3.8b}$$

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

$$D_{xpart} = \frac{Q_d}{Q_d + Q_s} = \frac{Q_d}{Q_{cheq}} \tag{5.3.9}$$

which is now bias dependent. For example, the derivities of D_{xpart} can be easily obtained based on the quasi-static results:

$$\frac{dD_{xpart}}{dV_i} = \frac{1}{Q_{cheq}} \left(S_{xpart} \cdot C_{di} - D_{xpart} \cdot C_{si} \right) \tag{5.3.10}$$

where i represents the four terminals and C_{di} and C_{si} are the intrinsic capacitances calculated from the quasi-static analysis. The corresponding values for S_{xpart} can be derived from the fact that $D_{xpart} + S_{xpart} = 1$.

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

If $XPART < 0.5$, $D_{xpart} = 0.4$;
Else if $XPART > 0.5$, $D_{xpart} = 0.0$;
Else $D_{xpart} = 0.5$;

5.3.4 Derivation of nodal conductances

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

$$G_{tau} = \frac{C_{fact}}{\tau} \tag{5.3.11}$$

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

$$\frac{dD_{xpart}}{dV_d} \cdot \left(G_{tau} \cdot V_{def}\right) + D_{xpart} \cdot V_{def} \cdot \frac{dG_{tau}}{dV_d} \tag{5.3.12}$$

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

$$D_{xpart} \cdot G_{tau} \tag{5.3.13}$$

Other conductances can also be obtained in a similar way.