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Abstract A physics based model for the GaAs MESFET is formulated. The physical/geometrical/process parameters, such as device dimensions, material-related parameters, doping profile and channel thickness, etc., are directly applied to the model. The model is embedded into a circuit simulation program based on harmonic balance (HB) method for a computer-aided design (CAD) tool for the analysis and optimization of nonlinear microwave MESFET circuits. It can be used to device and circuit design optimization, statistical modeling, design centering and yield optimization.

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I. INTRODUCTION

Since Shockley provided the original FET model applicable to a long-gate-length device operating in a nonsaturated velocity mode [1], modeling of the FET has been substantially advanced. The models developed can be classified into two main categories.

The first category involves equivalent circuit models which rely on parameter extraction for the equivalent circuit elements, such as resistors, capacitors, inductors and dependent sources, etc., from measured DC and/or RF small-signal S parameters. The models of this type have a high degree of computational efficiency. They can be easily implemented in circuit simulators and are very popular among circuit designers. However, since none of the physical parameters are considered in these models, they are not ideal for studying the effect of process-related FET parameters on the overall performance of the circuits and are not suitable for statistical analysis of the devices.

The second category involves physics based models which attempt to solve the fundamental device equations, such as Poisson's equation and the current density equation, etc., with a minimal number of simplifying assumptions. The model parameters include physical/geometrical/process parameters. Models of this type are quite suitable for design optimization of the device before fabrication, statistical modeling, design centering and yield optimization.

The formulation of a physics based model for GaAs MESFETs presented in this paper is based on Khatibzadeh and Trew's work [2]. Trew's model is a large-signal analytic model for the GaAs MESFET. The model describes the conduction and displacement currents of the FET as a function of instantaneous terminal voltages and their time derivatives. It allows arbitrary doping profiles in the channel, and is thus suitable for the optimization of ion-implanted and buried-channel FETs.

II. FORMULATION OF THE MODEL

A. Basic Model Equations

The active or "intrinsic" region of a FET is shown in Fig. 1. This region consists of the area of the channel directly under the gate electrode. The model is formulated under this region. All other regions of the device are modeled phenomenologically using external or "extrinsic" linear elements. The basic device equations are

$$\nabla^2\psi = -(q/\varepsilon)[N(y) - n(x, y)] \tag{1}$$

$$\mathbf{J} = -qn\mathbf{v} + qD\nabla n \tag{2}$$

$$\nabla\cdot\mathbf{J} = q(\partial n/\partial t) \tag{3}$$

and

$$\mathbf{J}_t = \mathbf{J} + \varepsilon(\partial\mathbf{E}/\partial t) \tag{4}$$

where

$$\mathbf{E} = -\nabla\psi \tag{5}$$

is the electric field, ψ is the electrostatic potential, N is the arbitrary donor concentration in the channel, n is the free-electron density, $\mathbf{v}$ is the electron velocity, D is the diffusion coefficient, $\mathbf{J}$ is the conduction (drift + diffusion) current density, $\mathbf{J}_t$ is the total (conduction + displacement) current density, q is the electronic charge, and ε is the permittivity of GaAs. Equation (1) is the Poisson equation. Equation (2) is the current density equation of which the first term represents the drift current density and the second term accounts for the diffusion current density. Equation (3) is the continuity equation. Equation (4) is the total current density equation of which the second term results from displacement current. It is assumed that $\mathbf{v}$ and $\mathbf{E}$ are codirectional, i.e.,

$$\mathbf{v} = -\mu(E)\mathbf{E} \tag{6}$$

where $\mu(E)$ is the field-dependent mobility.

In our formulation, we allow the velocity-electric field (v-E) curve to vary w.r.t. the parameters in the v-E equations. This is very useful for parameter extraction because the shape

of the v-E curve cannot be exactly measured. For most published methods, the v-E curves are assumed to be piecewise linear or quadratic forms, as shown in Fig. 2. These methods neglect the pick effect near the point of critical electric field E_c, resulting in some error in the simulation. By allowing the change of the v-E curve, we can avoid this disadvantage and get a better fit to the measured DC and S parameter data with parameter extraction. The v-E equations we use in our formulation is based on

$$v_1 = \begin{cases} v_s & 2E_c \le E \\ \mu_0 E - [v_s/(4E_c^2)]E^2 & E < 2E_c \end{cases} \tag{7}$$

where v_s is electronic saturation velocity, μ_0 is the low field mobility and has the relation

$$v_s = \mu_0 E_c \tag{8}$$

We add another term

$$v_2 = \frac{v_s}{C} B\left[\frac{(E/A) - C}{(E/A)^n + B} + \frac{C}{B}\right] \tag{9}$$

into v_1 and form the v-E equation as

$$v = (1 - H)v_1 + Hv_2 \tag{10}$$

where A, B, C, n and H are the parameters to be determined.

The main feature of (10) is that the overshoot effect of the v-E curve can be represented by changing the parameters A, B, C, n and H. This makes the model more accurate. Fig. 3 shows some typical v-E curves calculated by (10). We can see the overshoot appearing near the critical electric field point.

Based upon the magnitude of the electric field in the channel, specifically at the y = a boundary (E(x, a)), the device can operate in one of three modes:

$$E(0, a) < E(L, a) < E_c \qquad \text{(Mode-A)}$$

$$E(0, a) < E(L_1, a) = E_c < E(L, a) \qquad \text{(Mode-B)}$$

$$E_c < E(0, a) < E(L, a) \qquad \text{(Mode-C)}$$

where E_c is the critical electric field. The plane $x = L_1$ indicated in Fig. 1 separates the

saturation and linear regions of the device. Mode-A has a free channel from source to drain, Mode-B has a free channel only on the source side and Mode-C has no free channel. Fig. 4 shows schematically these three operation modes.

Since (1)-(3) are coupled together, a functional form for n(x, y) is assumed *a priori* in order to solve for ψ analytically. In our formulation the functional form proposed in [2] is adopted. The representation of the function is

$$n(x, y) = [1 + \gamma(x - L_1)]T(d(x), y)N(y) \tag{11}$$

where T(d(x), y) is referred to as the transition function and is defined as

$$T(d(x), y) \equiv 1 - \frac{1}{1 + \exp[(y - d(x))/\lambda]} \tag{12}$$

The term $[1 + \gamma(x - L_1)]$ in (11) is used to allow for charge accumulation or depletion in the channel. For the region $x \leq L_1$ (i.e., the linear region), $\gamma = 0$. The value of the parameter λ in (12) is of the order of the Debye length.

B. Solution for the Potential ψ

The solution for the potential ψ can be represented by a linear superposition of two components, i.e., $\psi = \psi_0 + \psi_1$, where ψ_0 is the Laplacian potential due to the impressed voltages on the electrodes and ψ_1 is due to the space charge in the channel and satisfies Poisson's equation. Mathematically,

$$\nabla^2\psi_0 = 0 \tag{13}$$

with the boundary conditions

$$\psi_0(0, a) = 0 \tag{14a}$$

$$\psi_0(L, a) = V_0 \tag{14b}$$

$$\frac{\partial\psi_0}{\partial y}(x, a) = 0 \tag{14c}$$

$$\psi_0(x, 0) = 0 \tag{14d}$$

and

$$\nabla^2\psi_1 = -(q/\varepsilon)(N - n) \tag{15}$$

with boundary conditions

$$\psi_1(0, a) = 0 \tag{16a}$$

$$\psi_1(L, a) = V_1 \tag{16b}$$

$$\frac{\partial\psi_1}{\partial y}(x, a) = 0 \tag{16c}$$

$$\psi_1(x, 0) = V_{gs} - V_{bi} \tag{16d}$$

where V_{bi} is the built-in voltage of the gate Shottky contact, V_{gs} and $V_{ds} = V_1 + V_0$ are the applied gate-source and drain-source voltages across the intrinsic FET, respectively.

The solution for ψ_0 with the boundary conditions can be approximated by

$$\psi_0(x, y) = \frac{V_0}{\sinh[(\pi L)/(2a)]} \sinh[(\pi x)/(2a)] \sin[(\pi y)/(2a)] \tag{17}$$

Assuming $|\partial^2\psi_1/\partial y^2|$ dominates $|\partial^2\psi_1/\partial x^2|$ and using definition (11) and (12), we can determine ψ_1. The results are: for $0 \le x \le L_1$,

$$\psi_1(x, y) = -\frac{q}{\varepsilon}\int_y^a\int_{y'}^a [1 - T(d(x), y'')N(y'')dy''dy' + \frac{V_1}{L}x \tag{18}$$

and for $L_1 < x \le L$ $(d(x) \equiv d_1)$,

$$\psi_1(x, y) = -\frac{q}{\varepsilon}\int_y^a\int_{y'}^a [1 - T(d_1, y'')N(y'')dy''dy' + \frac{V_1}{L}x$$

$$+ \frac{q}{\varepsilon}\gamma(x - L_1)\int_y^a\int_{y'}^a T(d_1, y'')N(y'')dy''dy' \tag{19}$$

Applying (16d) to (18), we can solve for d(x). The result is an implicit solution of the form

$$-\frac{q}{\varepsilon}F_1(d(x)) + \frac{V_1}{L}x = V_{gs} - V_{bi} \tag{20}$$

where

$$F_1(d) = -\frac{q}{\varepsilon}\int_y^a\int_{y'}^a [1 - T(d, y'')N(y'')dy''dy' \tag{21}$$

For uniform doping ($N(y) \approx N_d$), d(x) can be solved explicitly as

$$d(x) = [(\frac{2\varepsilon}{qN_d})(\frac{V_1}{L}x - V_{gs} + V_{bi})]^{1/2} \tag{22}$$

In general, however, $F_1(d)$ depends on the arbitrary doping profile and cannot be calculated analytically. Instead, for a given doping profile, $F_1(d)$ is tabulated numerically as a function of d for 40 values of d on the interval [0, a]. Then, for any given values of V_{gs} and x, the corresponding value of $F_1(d)$ is calculated from (20), and cubic spline interpolation is used to get d(x).

C. Solution for the Electric Field E

The electric field is defined by (5) and has two components: one arising from ψ_0 and one due to ψ_1. Differentiating (17), (18) and (19) w.r.t. y, we have for the linear region ($0 \le x \le L_1$)

$$E_y(x, y) = -\frac{q}{\varepsilon}\int_y^a [1 - T(d(x), y')N(y')dy' - \xi\sinh(\frac{\pi x}{2a})\cos(\frac{\pi y}{2a})\frac{V_0}{L} \tag{23}$$

where

$$\xi = \frac{(\pi L)/(2a)}{\sinh[(\pi L)/(2a)]}$$

and for the saturation region ($L_1 < x \le L$)

$$E_y(x, y) = -\frac{q}{\varepsilon}\int_y^a [1 - T(d_1, y')N(y')dy' - \xi\sinh(\frac{\pi x}{2a})\cos(\frac{\pi y}{2a})\frac{V_0}{L}$$

$$+ \frac{q}{\varepsilon}\gamma(x - L_1)\int_y^a T(d_1, y')N(y')dy' \tag{24}$$

The x component of **E** in both linear and saturation regions ($0 \le x \le L$) can be

approximated by [4]

$$E_x(x, y) = -\frac{V_1}{L}\left[1 - \left(\frac{y}{a} - 1\right)^2\right] - \xi\cosh\left(\frac{\pi x}{2a}\right)\sin\left(\frac{\pi y}{2a}\right)\frac{V_0}{L} \tag{25}$$

At the interface y = a, the magnitude of **E** becomes

$$E(x, a) = \left| E_x(x, a) \right| = \frac{V_1}{L} + \xi\cosh\left(\frac{\pi x}{2a}\right)\frac{V_0}{L} \tag{26}$$

From the definition of the plane $x = L_1$, we have $E(L_1, a) = E_c$, combining this with (26) and solving for L_1, we get

$$L_1 = \left(\frac{2a}{\pi}\right)\ln[\eta + (\eta^2 + 1)^{1/2}] \tag{27}$$

where

$$\eta \equiv \left(\frac{2a}{\pi}\right)\left(\frac{E_cL - V_1}{V_0}\right)\sinh\left(\frac{\pi L}{2a}\right) \tag{28}$$

Thus, given V_{ds} and V_1, we can explicitly solve for L_1, and determine the operation modes of the device.

D. Source and Drain Conduction Currents

The source and drain conduction currents are calculated by integrating the conduction current density as given by (2) over the planes x = 0 and x = L, respectively. Here, we derive the conduction current expressions by considering the Mode-B operation. The conduction currents for the operations of Mode-A and Mode-C can be obtained by taking the limits $L_1 \rightarrow 0$ and $L_1 \rightarrow L$, respectively.

For Mode-B operation, the carrier concentration does not depend strongly on x in the linear region, therefore only the drift component of **J** is considered. While in the saturation region both drift and diffusion terms are included, since the carrier concentration varies with x.

The x component of the conduction current density at the source (x = 0) is given by

$$J_x(0, y) = q\mu(E(0, y))n(0, y)E_x(0, y) \tag{29}$$

Thus, the magnitude of the source conduction current becomes

$$I_s = - qW \int_0^a \mu(E(0, y))n(0, y)E_x(0, y)dy \tag{30}$$

where W is the total gate width of the FET. Since $E_x(0, y)$ and $\mu(E(0, y))$ are functions of V_1, I_s depends on V_1 and thus on the bias voltages. For more computational efficiency we can define an average mobility and rewrite I_s as

$$I_s = - qW\overline{\mu}_s \int_0^a n(0, y)E_x(0, y)dy \tag{31}$$

where $\overline{\mu}_s$ is a weighted mobility at the source (x = 0) and is defined as

$$\overline{\mu}_s = \frac{\sum_{i=1}^{M} \omega_i \mu_{si}}{\sum_{i=1}^{M} \omega_i} \tag{32}$$

where

$$\mu_{si} = \mu(E(0, y_i))$$

$$y_i = \left(\frac{i - 1}{M - 1}\right)a \qquad i = 1, 2, \cdots, M$$

The weights ω_i are defined by

$$\omega_i = \left(\frac{i - 1}{M - 1}\right)^{\alpha} \qquad i = 1, 2, \cdots, M \tag{33}$$

where α is an empirical input parameter of the model.

Similarly, the magnitude of the drain conduction current can be derived as

$$I_d = - qW\overline{\mu}_d \int_0^a n(L, y)E_x(L, y)dy + qWD\gamma \int_0^a T(d_1, y)N(y)dy \tag{34}$$

The last term in (34) accounts for diffusion current. The average mobility at the drain (x = L), $\overline{\mu_d}$ is defined as

$$\overline{\mu_d} = \frac{\sum_{i=1}^{M} \omega_i \mu_{di}}{\sum_{i=1}^{M} \omega_i} \tag{35}$$

where

$$\mu_{di} = \mu(E(L, y_i))$$

The coefficient γ is calculated by

$$\gamma = - \frac{\varepsilon V_1}{qLF_2(d_1)} \tag{36}$$

where

$$F_2(d) = - (\varepsilon/q)(V_{po} - V_{bi}) - F_1(d) \tag{37}$$

Here V_{po} is the pinch-off voltage defined by

$$V_{po} = V_{bi} - \frac{q}{\varepsilon} \int_y^a \int_{y'}^a N(y'')dy''dy' \tag{38}$$

E. Displacement Currents

The displacement current density is defined as

$$J_d = \varepsilon \, (\partial E/\partial t) \tag{39}$$

Thus the drain, source and gate displacement currents can be evaluated by

$$i_d = W \int_0^a \varepsilon \frac{\partial E_x}{\partial t} \Big|_{(L, y)} dy \tag{40a}$$

$$i_s = W \int_0^a \varepsilon \frac{\partial E_x}{\partial t} \Big|_{(0, y)} dy \tag{40b}$$

$$i_g = W \int_0^L \varepsilon \frac{\partial E_y}{\partial t}\Big|_{(x,\,0)} dy \tag{40c}$$

The final results are

$$i_d = -\frac{\varepsilon}{L} W\left[\frac{4}{3} a - \xi \cosh\left(\frac{\pi L}{2a}\right)\frac{2a}{\pi}\right]\frac{\partial V_1}{\partial t}$$

$$+ \frac{\varepsilon}{L} W \xi \cosh\left(\frac{\pi L}{2a}\right)\frac{2a}{\pi}\frac{\partial V_{ds}}{\partial t} \tag{41a}$$

$$i_s = -\frac{\varepsilon}{L} W\left(\frac{4}{3} a - \xi\frac{2a}{\pi}\right)\frac{\partial V_1}{\partial t} - \frac{2a\varepsilon W\xi}{\pi L}\frac{\partial V_{ds}}{\partial t} \tag{41b}$$

$$i_g = \left\{\frac{\varepsilon}{L} W \int_0^{L_1} \frac{F_3(d)}{\partial F_1/\partial d} x\,dx - \frac{\varepsilon}{L}\frac{2a}{\pi}\xi W\left[\cosh\left(\frac{\pi L}{2a}\right) - 1\right]\right\}\frac{\partial V_1}{\partial t}$$

$$- \varepsilon W \int_0^{L_1} \frac{F_3(d)}{\partial F_1/\partial d}\,dx \frac{\partial V_{gs}}{\partial t} + \frac{2a\varepsilon\xi W}{\pi L}\left[\cosh\left(\frac{\pi L}{2a}\right) - 1\right]\frac{\partial V_{ds}}{\partial t} \tag{41c}$$

where

$$F_3(d) = \int_0^a [1 - T(d(x),\, y')]\frac{\partial T}{\partial d}\,dy'$$

In the HB simulation, the displacement currents are equivalently replaced by the accumulation charges in the drain, source and gate contacts, i.e., Q_d, Q_s and Q_g. The equations for calculating Q_d, Q_s and Q_g are

$$Q_d = \varepsilon W \int_0^a E_x(L,\, y)dy\ , \tag{42a}$$

$$Q_s = \varepsilon W \int_0^a E_x(0,\, y)dy \tag{42b}$$

and

$$Q_g = \varepsilon W \int_0^L E_y(x,\, 0)dx\ . \tag{42c}$$

F. The Total Currents

The total currents are the summation of conduction currents and displacement currents. From (31), (34) and (41), we can, in general, write the total gate, drain and source currents as

$$I_{gt} = f_{Igt}(\phi, V_1, V_{gs}, V_{ds}, \partial V_1/\partial t, \partial V_{gs}/\partial t, \partial V_{ds}/\partial t) \tag{43a}$$

$$I_{dt} = f_{Idt}(\phi, V_1, V_{gs}, V_{ds}, \partial V_1/\partial t, \partial V_{gs}/\partial t, \partial V_{ds}/\partial t) \tag{43b}$$

$$I_{st} = f_{Ist}(\phi, V_1, V_{gs}, V_{ds}, \partial V_1/\partial t, \partial V_{gs}/\partial t, \partial V_{ds}/\partial t) \tag{43c}$$

where f_{Igt}, f_{Idt} and f_{Ist} are nonlinear functions, ϕ is a parameter vector including gate length, gate width, channel thickness, doping density, etc.

III. DYNAMIC INTEGRATION OF THE PHYSICAL MODEL WITH HARMONIC BALANCE SIMULATION

In the circuit simulation with physics based model described in Section II, the key variable to be solved for is V_1. In the original approach of [2, 3], the condition $I_d = I_s$ was used to solve for V_1, neglecting the current through the gate contact. However, this condition is strictly valid for DC, and not for instantaneous currents under AC excitation. Also, a double loop is required in HB simulation. The first loop is devoted to solving V_1 iteratively, then the value of V_1 is substituted into the second loop for solving the HB equations.

In our formulation, V_1 is integrated directly into the HB equations. For example, the HB equation for a nonlinear circuit with one FET can be written as

$$\overline{\mathbf{F}}(\overline{\mathbf{V}}) = \overline{\mathbf{Y}} \begin{bmatrix} \overline{\mathbf{V}}_{gs} \\ \overline{\mathbf{V}}_{ds} \\ \overline{\mathbf{V}}_1 \end{bmatrix} + \overline{\mathbf{A}} \begin{bmatrix} \overline{\mathbf{I}}_{gt} \\ \overline{\mathbf{I}}_{dt} \\ \overline{\mathbf{I}}_{st} \end{bmatrix} + \overline{\mathbf{I}}_{ss} = \mathbf{0} \ , \tag{44}$$

where a bar is used to denote the split real and imaginary parts of a complex quantity at DC, fundamental frequency and all harmonics. For example, $\overline{\mathbf{V}}_{gs}$ is a vector containing the real and imaginary parts of intrinsic gate voltage at all harmonics. $\overline{\mathbf{I}}_{gt}$, $\overline{\mathbf{I}}_{st}$ and $\overline{\mathbf{I}}_{dt}$ represents the sum of conducting and displacement currents through the gate, source and drain, respectively. $\overline{\mathbf{Y}}$ is

the admittance matrix of linear part. Matrix $\overline{A}$ is a simple incidence matrix containing 1's and 0's. $\overline{I}_{ss}$ contains excitations. Equation (43) automatically ensures the current continuity at all harmonics, i.e.,

$$\overline{I}_g + \overline{I}_s + \overline{I}_d = 0 \tag{45}$$

Therefore, our condition is valid not only for DC but also for small- or large-signal RF operation. Only a single iterative loop, i.e., the loop for solving the HB equation is required. This is generally much more efficient than the double loop approach. The algorithm for the physics based model is shown in Fig. 5.

IV. ALGORITHM FOR THE PHYSICAL MODEL

The following algorithm for the physical model can be used to calculate the drain and source conduction currents, I_d and I_s, and accumulative charges in the gate, drain and source electrodes, Q_g, Q_d and Q_s. The calculation is carried out in the time domain. The values of the model variables V_1, V_{gs} and V_{ds} are assumed coming from the frequency domain analysis with the HB method. The outputs of the model, i.e., I_d, I_s, Q_g, Q_d and Q_s, are sent to the frequency domain analysis by Fourier transform to solve the HB equation (44).

Step 1 Input the physical/geometrical/process parameters such as gate length L, gate width W, channel thickness and doping profile, etc., model parameters such as α and λ, and parasitic parameters such as R_g, R_d, R_s, L_g, L_d and L_s, etc.

Step 2 Pre-calculations.

Step 2.1 Given N values of d_i, i = 1, 2, •••, N, calculate the corresponding values of $F_1(d_i)$ using (21) to obtain a set of data $(d_i, F_1(d_i))$.

Step 2.2 Create the cubic spline coefficients for interpolation using the data $(d_i, F_1(d_i))$ obtained in *Step 2.1*.

Step 2.3 Calculate the pinch-off voltage V_{po} using (38).

Step 3 From the values of V_1, V_{gs} and V_{ds} given by the frequency domain analysis determine the operation mode of the MESFET. Compute L_1 using (27) and (28). If $L_1 \geq L$, the operation mode is Mode-A and L_1 is set to be L. If $L_1 \leq 0$, the operation mode is Mode-C and L_1 is set to be 0. Otherwise the operation mode is Mode-B.

Step 4 Calculate the source and drain conduction currents I_s and I_d.

Step 4.1 Determine the depletion layer thickness at x = 0 and x = L_1, i.e., d_0 and d_1. First obtain $F_1(d_0)$ and $F_1(d_1)$ using (20) and then get d_0 and d_1 by cubic spline interpolation.

Step 4.2 Calculate model parameter γ using (36).

Step 4.3 Compute the weighted average mobility $\overline{\mu_s}$ and $\overline{\mu_d}$ using (32), (33) and (35).

Step 4.4 Evaluate the source and drain conduction currents I_s and I_d using (31) and (34).

Step 5 Calculate the accumulative charges in the drain, source and gate electrodes Q_d, Q_s and Q_g using (42a), (42b) and (42c).

V. CONCLUSIONS

The formulation of the physics based model for the GaAs MESFET has been discussed in this paper. The model accepts, as input data, physical/geometrical/process parameters of the device, such as gate length, channel thickness and doping profile, etc. It represents a breakthrough over the conventional equivalent circuit approach. With this model the sensitivity analysis can be carried out directly w.r.t. the physical parameters of the device. This is very useful in both device and circuit design optimization and is quite suitable for statistic modeling, design centering and yield optimization of FET circuits and MMICs.

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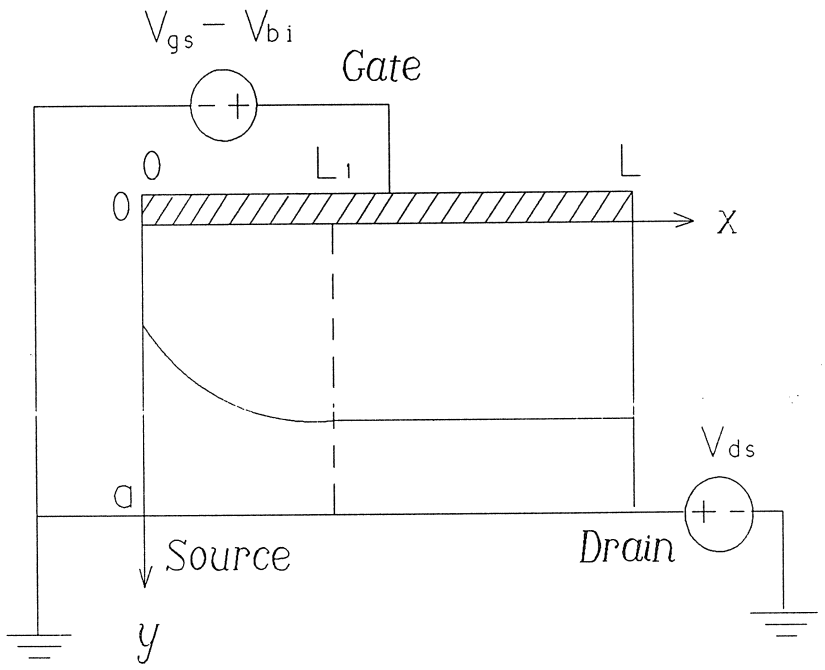


Fig. 1 Active region of the FET used in the model.

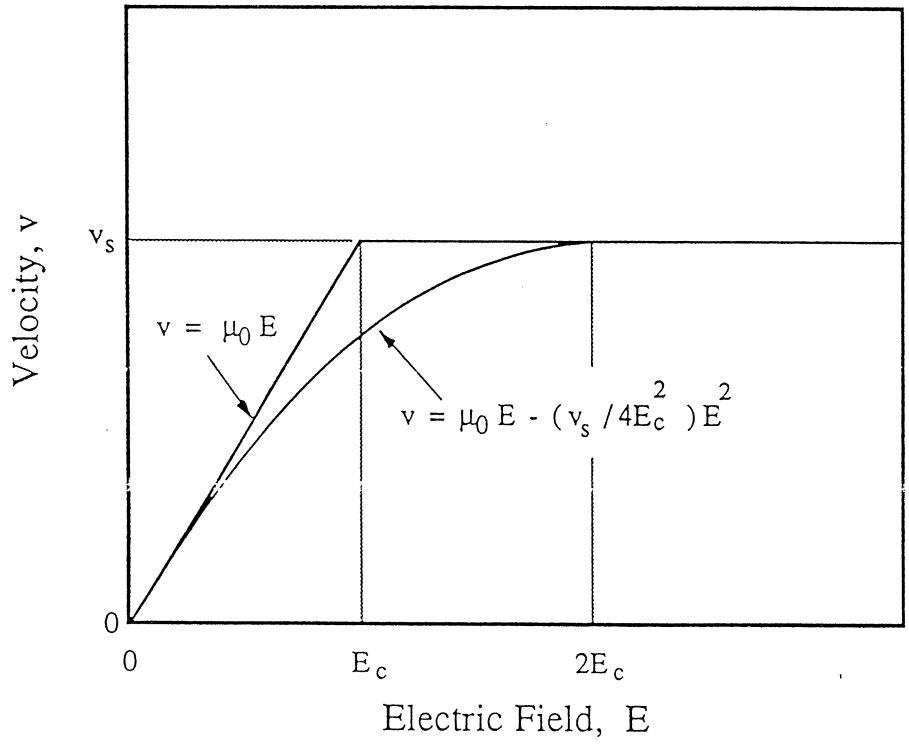


Fig. 2 Two analytic formulations for the velocity as a function of the electric field.

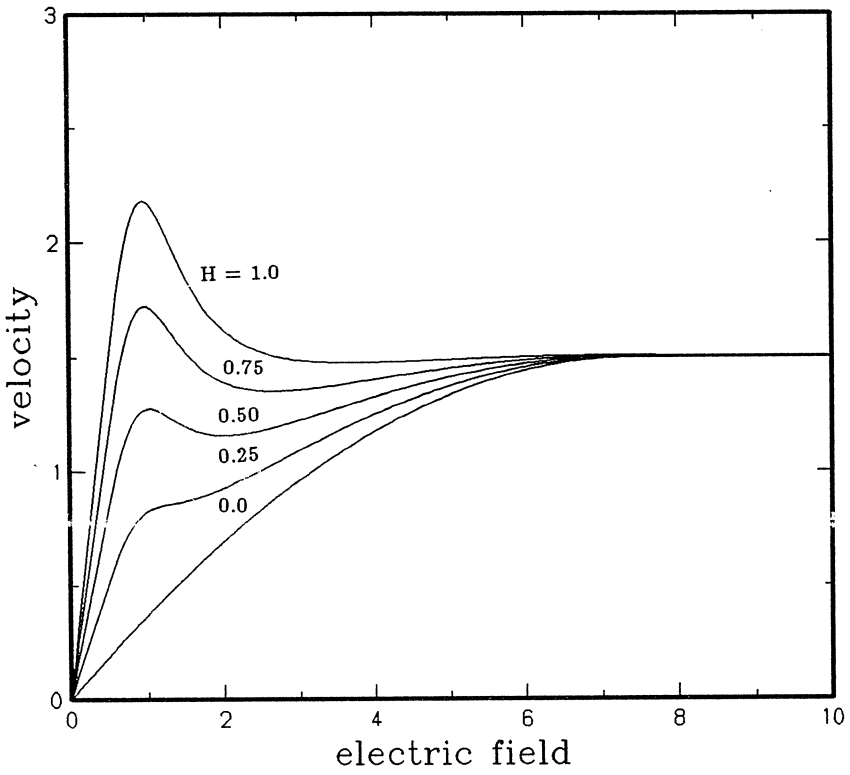
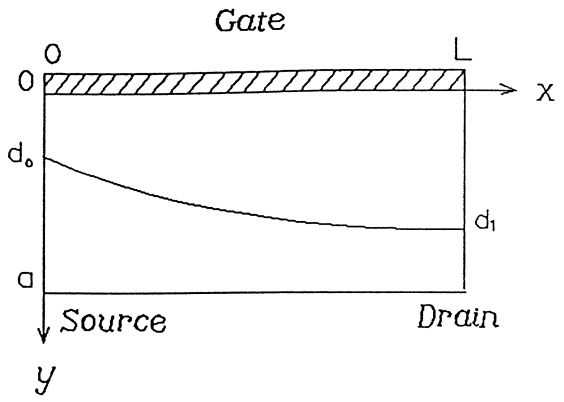
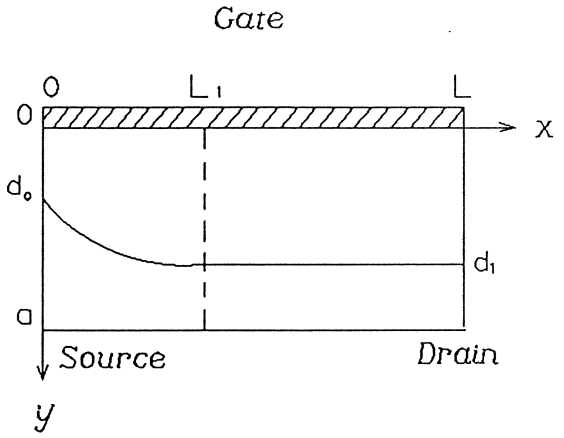


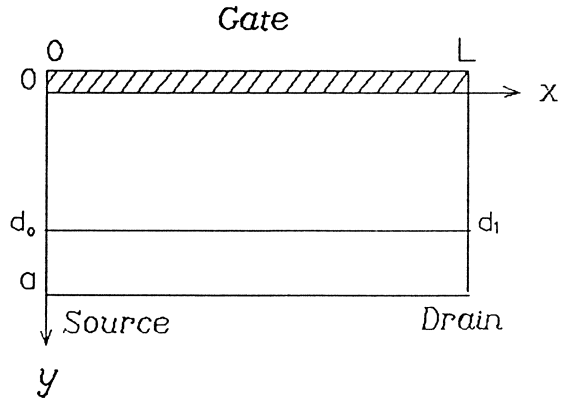
Fig. 3 Velocity-electric field curves determined by (10). E_c = 3.75kV/cm, v_s = 1.5×10^7cm/s, μ_0 = 4000cm^2/V.s, A = 1, B = 1, C = 0.4 and n = 4.



(a)



(b)



(c)

Fig. 4 Operation modes of the FET. (a) Mode-A, (b) Mode-B and (c) Mode-C.

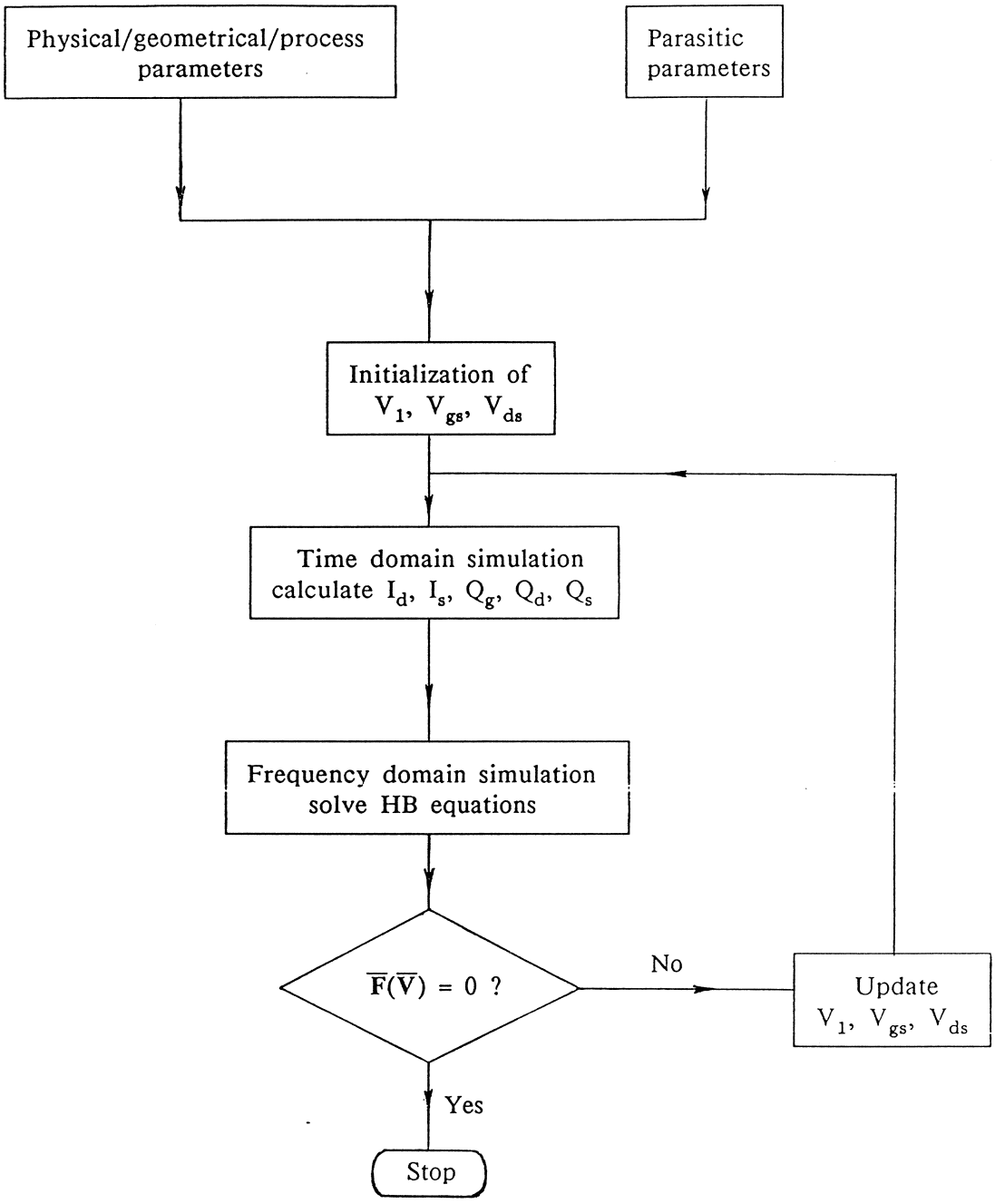


Fig. 5 Algorithm for the physics based model.