SMALL-SIGNAL TIME-DOMAIN PHYSICAL/ELECTRICAL FET MODELING APPROACH

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ABSTRACT

In this paper, a reliable small-signal time-domain FET modeling approach is proposed. Based on physical/electrical parameters, the proposed model can efficiently characterize high-frequency transistors.

Index Terms— FDTD, MESFET, time domain, wave effects

Nomenclature

a	Channel thickness (m)
E	Electric field (V/m)
ε_s	Semiconductor permittivity (F/m) with $\varepsilon_s = \varepsilon_o \varepsilon_r$
ε_o	Permittivity of free space (8.85*10^{-12} F/m)
ε_r	Relative permittivity (13.1 for GaAs)
h	Depletion width (m)
H	Height of the gate strip (m)
i_{ds}	Drain to Source current (A)
i_{sub}	Substrate current (A)
i_{ch}	Channel current (A) with $i_{ch} = i_{ds} - i_{sub}$
I_p	Saturation current (A)
K_B	Boltzmann's constant (1.38*10^{-16} /K)
L_g	Gate length (m)
L_{gd}	Gate to drain length (m)
L_{sg}	Source to gate length (m)
m	Effective electron mass (kg)
n	Free electron density (m^{-3})
N_A^-	Ionized acceptor doping density (m^{-3})
N_D^+	Ionized donor doping density (m^{-3})
N_D	Donor doping density with $n = N_D$(m^{-3})
p	Free hole density (m^{-3})
p_s	Number of parallel strips
φ	Electrostatic potential (V)
$\emptyset_b$	Schottky barrier height (V)
q	Electron charge (1.60*10^{-19} C)
T_e	Electron temperature (K)
τ_p	Momentum relaxation time (s)
τ_w	Energy relaxation time (s)
μ	Mobility of the free carrier (electron) (cm^2/V.s)
μ_p	Free space permeability (12.56*10^{-6} H/m)
v_s	Saturation velocity (m/s)
v	Carrier velocity $v = \mu E$ (m/s)
w	Total average electron energy (eV)
Z	Device width (m)
l_g	Gate inductance (H)
R_s	Source resistance (Ω)
R_d	Drain resistance (Ω)
R_g	Gate resistance (Ω)
R_c	Ohmic contact resistance (Ω)
R_i	Channel resistance (Ω)
C_{gs}	Gate to source capacitance (F)
C_{gd}	Gate to drain capacitance (F)
V_T	Threshold voltage (V)
V_{gs}	Gate to source voltage (V)
V_{ds}	Drain to source voltage (V)
g_m	Transconductance (S)
g_{ds}	Output conductance (S)

1. INTRODUCTION

The ever increasing interest in high-gain low-noise high-frequency communication systems has significantly emphasized the use of GaAs Field Effect Transistors (FETs) like Metal-Semiconductor FETs (MESFETs). The flexibility in their technology emphasized their extensive use in several high-frequency applications advertising low-noise figures and high-gains [1]-[8].

A physical model can efficiently capture the effects of channel parameters on the transistor behavior while an equivalent electrical circuit is more flexible and suitable for circuit design and optimization. In fact, existing 2D/3D physical models are accurate but still require intensive CPU time and memory as well as certain user expertise to manage data while electrical models are faster and easier to use, but their accuracy is mainly dependent on how precisely the circuit elements are related to the device technology and physical behavior.

In this paper, we merged both approaches for efficient FET modeling. They started from physical relationships to accurately derive the component values and thus built a reliable electrical equivalent circuit.

To demonstrate the proposed approach, a low-noise Ku-K band GaAs MESFET was modeled and the obtained simulated values successfully compared with measured data. To further improve the transistor model, wave-propagation effects have been included to take into account the miniaturization effects associated to high frequency operation.

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2. 2D HYDRODYNAMIC FET MODEL

Fundamental semiconductor equations can be divided into two main groups namely, electromagnetic and charge transport. At low frequencies, the Poisson's equation correlates the electrostatic potential to a given charge distribution as [9]:

$$\nabla(\varepsilon_0\varepsilon_r\nabla\varphi) + q(N_D^+ - N_A^+ + p - n) = 0 \qquad (1)$$

where all variables used in this paper are defined in the nomenclature list. Thus, the electric field can be deduced from:

$$\vec{E} = -\nabla\vec{\varphi} \qquad (2)$$

However, in the submicron regime, the hydrodynamic approach is more reliable to simulate the carrier transport, where the carrier, momentum and energy balance equations are solved simultaneously with the Poisson's equation. Since the GaAs FET is dominated by the single-electron gas approximation so, by omitting the carrier heat and kinetic energy terms, the Boltzman's transport equations are

$$\frac{\partial \mathrm{n}}{\partial \mathrm{t}} + \nabla(\mathrm{n}\vec{\mathrm{v}}) = 0 \qquad (3)$$

$$\frac{\partial \mathrm{v}}{\partial \mathrm{t}} + \vec{\mathrm{v}}\nabla\vec{\mathrm{v}} + \frac{\mathrm{q}\vec{\mathrm{E}}}{\mathrm{m}} + \frac{2}{3}\nabla(\mathrm{nw}) = -\frac{\vec{\mathrm{v}}}{\tau_\mathrm{p}} \qquad (4)$$

$$\frac{\partial \mathrm{w}}{\partial \mathrm{t}} + \mathrm{q}\vec{\mathrm{v}}\vec{\mathrm{E}} + \frac{2}{3\mathrm{n}}\nabla(\mathrm{n}\vec{\mathrm{v}}\mathrm{w}) = -\frac{\mathrm{w}-\mathrm{w}_0}{\tau_\mathrm{w}} \qquad (5)$$

where τ_p, τ_w, and $\vec{v}$ (function of the applied electrical field, doping and lattice temperature) are derived from three-valley ensemble Monte-Carlo simulations. The energy of the carriers can be expressed as [9]

$$\mathrm{w} = \frac{3}{2}\mathrm{K_B T_e} + \frac{1}{2}\mathrm{mv}^2, \qquad \mathrm{w}_0 = \frac{3}{2}\mathrm{K_B T_0} \qquad (6)$$

In the microwave range, the displacement current cannot be neglected. Therefore, the total conduction current is the sum of the drift and displacement currents given in [9] by

$$\vec{\mathrm{J}}(\mathrm{t}) = -\mathrm{qn}\vec{\mathrm{v}}(\mathrm{t}) + \varepsilon_0\varepsilon_\mathrm{r}\frac{\partial\vec{\mathrm{E}}}{\partial \mathrm{t}} \qquad (7)$$

$$\rightarrow \quad \vec{\mathrm{J}}(\mathrm{t}) = -\mathrm{qn}\mu_\mathrm{n}\vec{\mathrm{E}}(\mathrm{t}) + \varepsilon_0\varepsilon_\mathrm{r}\frac{\partial\vec{\mathrm{E}}}{\partial \mathrm{t}} \qquad (8)$$

3. TIME-DOMAIN WAVE EFFECTS

For long channel FETs and under boundary conditions, the Schottky barrier potential can be derived from the Poisson's equation as [10]

$$\emptyset_b = \frac{qN_Dh^2}{2\varepsilon_s} \qquad (9)$$

According to [11], the threshold voltage can be expressed as

$$V_T = \emptyset_b - V_{gs} + V_{ds} \qquad (10)$$

Thus, from the drain current square law given in [12], the transconductance $g_m$and the g_{ds}output conductance can be obtained after the saturation onset as

$$g_m = \left.\frac{\partial i_{ds}}{\partial V_{gs}}\right|_{Vds=cst} = \frac{I_p}{V_T}\sqrt{\frac{\emptyset_b - V_{gs}}{V_T}} \qquad (11)$$

$$g_{ds} = \left.\frac{\partial i_{ds}}{\partial V_{ds}}\right|_{Vgs=cst} = \frac{I_p}{V_T} \qquad (12)$$

with

$$I_p = \frac{q^2N_d^2\mu Zh^3}{2\varepsilon_s L_g} \qquad (13)$$

The intrinsic capacities are given by [13] as:

$$C_{gs} = \left(\frac{\pi}{2} + \frac{L_g}{h}\right)\varepsilon_s Z, \quad C_{gd} = \frac{\pi}{2}\varepsilon_s Z \qquad (14)$$

while the channel resistance can be stated as [14], [15]

$$R_i = \frac{v_s L_g}{\mu\, i_{ch}} \qquad (15)$$

Finally, the parasitic elements of the FET terminals are given by [11], [13]

$$R_s = R_c + \frac{L_{sd}}{q\mu_p N_D\, a\, Z}, \quad R_d = R_c + \frac{L_{gd}}{q\mu_p N_D\, a\, Z}$$

$$R_g = \frac{\rho Z}{3.p_s^2\, HL_g}, \quad l_g = \frac{h\,\mu_p}{p_s^2 L_g}Z \qquad (16)$$

Using the well-known de-embedding procedure presented in [16], a small-signal FET electrical equivalent circuit was obtained. However associated to ever-increasing miniaturization, transistor models operating at high frequencies should include electromagnetic effects such as wave propagation effects [17]-[19].

In the proposed modeling approach, such effects have been considered. Therefore, our model has been divided into two parts: passive and active (Fig. 1). Note that the passive part represents the device electrodes modeled as a set of three transmission lines. Note also that the voltages and currents are time-dependent and function of the position *x* along the device width; the proposed approach is indeed a time-domain approach to capture the real device behavior.

By biasing the device with Gaussian pulse, the full wave analysis can be attributed and the transmission line elements provided along the wide frequencies band [17]-[19]. To obtain the intrinsic elements of the circuit, a 2-D hydrodynamic simulation was performed for various bias conditions.

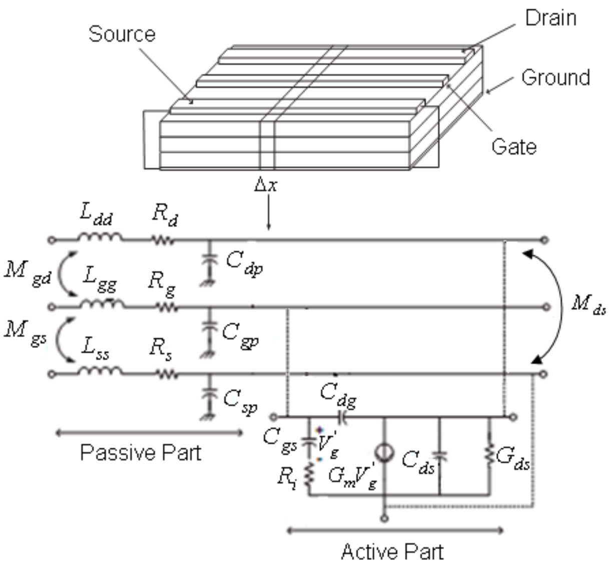

Fig. 1. Distributed representation of FET terminals.

The time-space finite-difference method proposed by Roden *et al.* [20] was chosen to discretize the model equations and thus to obtain the voltages along the electrodes of the transistor as well as the voltage across the gate–source capacitance [19]. Then the active part can be replaced by the intrinsic equivalent circuit obtained above.

4. RESULTS

To demonstrate the proposed approach, the NE71000, a low-noise Ku-K band GaAs MESFET, was characterized (Table I). As shown in Fig. 2 (for the I-V curves) and Fig. 3 (Small-signal S-parameters), the obtained simulated values were successfully compared with simulated data obtained from a commercial high-frequency simulator (ADS [21]) and/or measured data. To further improve the proposed model, wave-propagation effects have been included to take into account the miniaturization effects associated to high frequency operation (Fig. 4).

Table I. MESFET physical parameters (uniform distribution).

Physical Parameters	Value (unit)
a	0.1μm
p-buffer layer	0.15μm
Substrate thickness	0.15μm
Z(μm)*p_s	70*4
L_g	0.3μm
Recessed gate depth	0.028μm
Donor doping density	$2.5*10^{17}cm^{-3}$

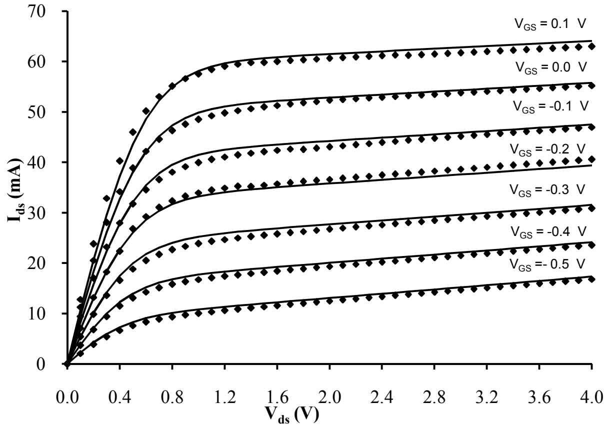

Fig. 2: NE71000. I-V curves. Comparison between measurements (♦) and our simulations (—).

5. CONCLUSION

A robust physical/electrical distributed time-domain field effect transistor model was proposed. By including wave-propagation effects, this model can be successfully used in high-frequency FET-based circuit design and optimization.

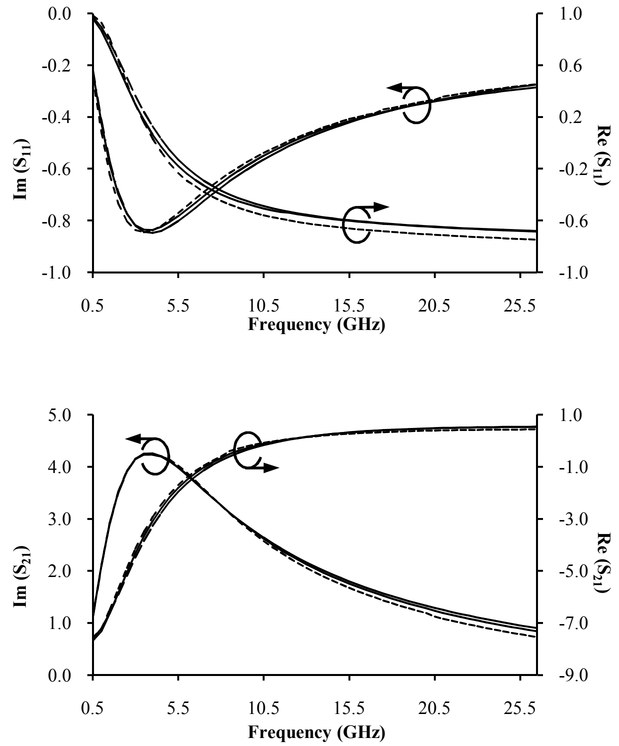

Fig.3: NE 71000. Comparison between simulated results from ADS (—·—), measurements (- - -) and our simulated results (—)

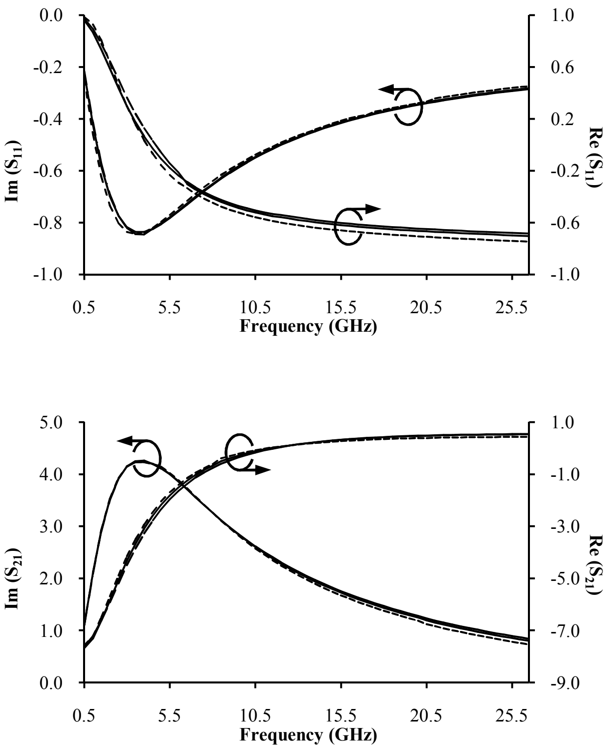

Fig.4: Comparison between simulated results with wave propagation effects (–·–), measurements (- - -) and our simulated results without wave propagation effects (—)

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