



UNIVERSITÀ
DEGLI STUDI
FIRENZE

FLORE

Repository istituzionale dell'Università degli Studi di Firenze

Physical/electromagnetic pHEMT modeling

Questa è la Versione finale referata (Post print/Accepted manuscript) della seguente pubblicazione:

Original Citation:

Physical/electromagnetic pHEMT modeling / Cidronali, Alessandro; Leuzzi, Giorgio; Manes, Gianfranco; Giannini, Franco. - In: IEEE TRANSACTIONS ON MICROWAVE THEORY AND TECHNIQUES. - ISSN 0018-9480. - STAMPA. - 51:(2003), pp. 830-838. [10.1109/TMTT.2003.808580]

Availability:

This version is available at: 2158/1076069 since: 2017-03-02T16:50:45Z

Publisher:

IEEE / Institute of Electrical and Electronics Engineers Incorporated:445 Hoes Lane:Piscataway, NJ 08854:

Published version:

DOI: 10.1109/TMTT.2003.808580

Terms of use:

Open Access

La pubblicazione è resa disponibile sotto le norme e i termini della licenza di deposito, secondo quanto stabilito dalla Policy per l'accesso aperto dell'Università degli Studi di Firenze (<https://www.sba.unifi.it/upload/policy-oa-2016-1.pdf>)

Publisher copyright claim:

(Article begins on next page)

Physical/Electromagnetic pHEMT Modeling

Alessandro Cidronali, *Member, IEEE*, Giorgio Leuzzi, Gianfranco Manes, *Senior Member, IEEE*, and Franco Giannini

Abstract—An effective technique, which is based only on geometrical and physical data, is presented for the analysis of high-frequency FETs. The intrinsic part of this electron device is described by a quasi-two-dimensional hydrodynamic transport model, coupled to a numerical electromagnetic field time domain solver in three dimensions that analyzes the passive part of the FET. Such an analysis is entirely performed in the time domain, thus allowing linear and nonlinear operations. The obtained data give insights to some parameters affecting the signal distribution through the entire device structure; a comprehensive discussion of these is given for a test device. In order to prove the validity of the approach, the bias-dependent small-signal analysis is compared with the corresponding measurements up to 50 GHz for two 0.3- μm gate-length AlGaAs–InGaAs–GaAs pseudomorphic high electron-mobility transistors, each having two gate fingers of 25- μm and 100- μm width, at bias points ranging from I_{DSS} to the pinchoff regime. The accuracy and the efficiency of the approach make it suitable for device optimization.

Index Terms—Electron device modeling, global modeling, microwave monolithic integrated circuit, pseudomorphic high electron-mobility transistors (pHEMT).

I. INTRODUCTION

THE design of electron devices for microwave and millimeter-wave frequencies is traditionally based upon simple models and basic concepts that are accurate and comprehensive, but use very slow simulators [1]–[4]. These models have become more complex due to the decreasing trend in the device size that has characterized the most recent developments. The actual device geometry (the structure of the electrodes and passive connecting parts) also influences the device characteristics and, hence, may constitute a set of available design parameters. The topology of the passive connecting parts is basically designed by means of cut-and-try procedures; the experience and expertise of the designer is an essential part of this process. For very high frequencies and more advanced devices, this procedure tends to be progressively less effective. Moreover, the growing interest for *a priori* design methods has driven the research effort toward a possibly complete, physical-based approach. The statistical significance and predictive ability are regarded as interesting features for this class of models. It is significant to consider in a self-consistent way the constitutive equations which rule the phenomena appearing in a real electron device. This effort may

be accomplished by solving the models for the electromagnetic (EM) field coupling, the carrier transport, and the thermal propagation.

Recently, the research efforts in this field have been broadly termed as global modeling [5]–[17]. The present paper is intended to be a contribution to this fast growing area, where a 0.3- μm pseudomorphic high electron-mobility transistor (pHEMT) is analyzed for low-noise applications. In the present approach, the thermal processes are neglected.

Recent advances in the construction methods have led to devices having typical dimensions of the order of a particle wavelength. As a direct consequence, the quantum mechanics of the particle in such a structure must be properly considered in the phenomena to be modeled, thus increasing the complexity of the description from both the theoretical and numerical points of view. In the case under investigation, a calculation of the two-dimensional electron gas (2DEG) properties must be accounted for; improper estimation leads to a wrong transconductance value and an inexact pinchoff regime. This consideration is even more relevant in the device optimization stage, where the active layer of the transistor can be regarded as a degree of freedom for the designer.

The goal of the present approach is to consider the above described phenomena which self-consistently contribute to the electron device behavior, in a simple but reliable quasi-three-dimensional time-domain description.

The electron device is assumed to consist of an “*extrinsic passive structure*” connected with a finite, suitable number of elementary “*intrinsic devices*.” The intrinsic device analysis considers two models: the first part is the channel charge control which acts in the vertical direction; the second is the electron channel transport in the horizontal direction. The evaluation of these two effects is obtained on the basis of the voltages at the intrinsic device terminals. As a result, a set of currents characterizes the dynamic response of the intrinsic device that is then coupled to a tri-dimensional EM field solver in order to account for the extrinsic passive structure. In turn, the solution for the EM fields gives the time- and space-dependent control voltages at any intrinsic devices.

Several techniques have already been proposed which are based on different approaches and with varying degrees of accuracy and complexity [6]–[11], [15]–[17]. This paper proposes a comparatively simple code which runs on a PC, thus allowing the designer to predict the device behavior and, potentially, to optimize its performance with a reasonable degree of accuracy and affordable computational time [16].

The organization of this paper is as follows. The description of the quasi-bidimensional (Q2D) pHEMT intrinsic region model [1]–[4], [18]–[23], which is composed of two main parts—the approximate nonstationary charge-transport model

Manuscript received January 28, 2002; revised June 28, 2002.

A. Cidronali and G. Manes are with the Department of Electronics and Telecommunications, University of Florence, 50139 Florence, Italy.

G. Leuzzi is with the Department of Electrical Engineering, University of L’Aquila, Italy, Monteluco di Roio–L’Aquila 67040, Italy.

F. Giannini is with the Department of Electrical Engineering, University of Rome “Tor Vergata,” 00133 Rome, Italy.

Digital Object Identifier 10.1109/TMTT.2003.808580

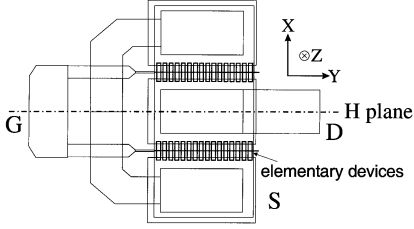


Fig. 1. Structure of the physical/EM FET model.

based on the energy and momentum balance equations and the charge-control model—is presented in Section II. Insights of the most important parameters are given for a test device. The important assumptions on the device operation are introduced in Section III. These enable a strongly simplified link between the Q2D model and the three-dimensional (3-D) EM full-wave model. Finally, small-signal experimental results at microwave frequencies for a $0.3\text{-}\mu\text{m}$ AlGaAs–InGaAs–GaAs pHEMT are presented in Section IV. These confirm the simulation capabilities.

II. INTRINSIC pHEMT PHYSICAL MODEL

Fig. 1 shows the schematic representation of a microwave FET where the active part is divided into a convenient number of intrinsic devices. These are analyzed by means of two models: the first is the channel charge control model which operates in the z -direction and the second is the electron channel transport model in the x -direction.

A. Charge-Control Model

The charge-control model of the pHEMT is derived from a self-consistent solution of the Poisson and Schrödinger equations. In this way, it is possible to provide the actual charge-control law as a function of the device layer structure and material composition. This procedure is executed before the calculation of the dynamic behavior, then inserted in the transport calculation by means of a lookup table (LUT) or fitting equation. Hence, the exact description of the 2DEG in the device channel, as a function of the gate-to-channel voltage (V_{GC}), is accounted for in a numerically efficient way.

The steady-state Schrödinger's equation is

$$\frac{-\hbar^2}{2m_o} \frac{d}{dz} \left[\frac{1}{m_z^*} \frac{d\psi_i(z)}{dz} \right] + qV_C(z)\psi_i(z) = E_i\psi_i(z) \quad (1)$$

where m_z^* is the effective mass and E_i and $\psi_i(z)$ are the energy level and the envelope function of the i th subband, respectively. The electrostatic potential is numerically given by the solution of the Poisson's equation

$$\frac{d}{dz} \left[\varepsilon(z) \cdot \frac{dV(z)}{dz} \right] = q [N_D^+(z) - n_g(z)] \quad (2)$$

where $V(z)$ is the electrostatic potential through the semiconductor,¹ $\varepsilon(z)$ is the position-dependent dielectric constant, and $N_D^+(z)$ is the ionized donor density. Since the charge density on the right-hand side of (2) depends on the potential variation,

¹This potential corresponds to the modification of the vacuum level through the semiconductor.

an iterative calculation is required. The electron concentration $n_g(z)$ is described as a sum of two contributions: the 2DEG density localized in the channel $n_{2DEG}(z)$ and the bulk semiconductor electron density localized in the supply layer $n_B(z)$. The first is calculated as described in [24], while the latter is calculated by using the classical model based on 3-D Fermi statistics in which the subband structure is neglected.

In this model, the electrostatic potential $V(z)$ is modified by the conduction band discontinuities ΔE_i at any i th heterointerface. This leads to the definition of the position-dependent conduction band potential

$$V_C(z) = V(z) - \frac{\Delta E_1}{q} u(z - z_1) - \frac{\Delta E_2}{q} u(z - z_2) \quad (3)$$

which is the potential to be used for the numerical solution of (1). The gate bias dependence of $n_g(z)$ is calculated under the following boundary conditions. The conduction band edge at the gate electrode is determined by the sum of the Schottky-barrier height and the gate voltage. The conduction band edge at a point sufficiently far from the channel is fixed to its thermal equilibrium value. The channel is assumed to be grounded (V_{gs} equal to zero) and the Fermi level fixed to zero. The envelope functions at both boundaries are fixed to be zero.

B. The Nonstationary Charge Transport Model

The electron transport in the channel layer which is assumed to take place essentially in the direction from the source to the drain electrodes (x -direction) is modeled by the Energy and Momentum Balance equations [1]

$$\begin{aligned} \frac{\partial n}{\partial t} &= 0 \\ \frac{\partial v}{\partial t} &= \frac{qE_x}{m^*} - \frac{v}{\tau_p} \\ \frac{\partial w}{\partial t} &= qvE_x - \frac{w - w_0}{\tau_w} \end{aligned} \quad (4)$$

where n is the electron density, v is the electron velocity, E_x is the longitudinal electric field, and w is the electron energy, and the relaxation terms τ_p and τ_w as well as w_0 are obtained by steady-state Monte Carlo analysis [25]. The equations are solved in the time domain and account for transient and nonlinear behavior, while the boundary condition is given by time-dependent drain–source voltage. The insertion of the channel control model is obtained from Poisson's equation in the channel

$$\frac{\partial E_x}{\partial x} = \frac{q}{\varepsilon} \left[n - \frac{1}{H} \int_0^h n_g[z, V_{\text{sur}} - V_{\text{ch}}] \cdot dz \right] \quad (5)$$

where H is the device thickness and ε is the dielectric constant. The voltages $V_{\text{sur}}(x, t)$ and $V_{\text{ch}}(x, t)$ represent the air–semiconductor and channel voltages, respectively. The gate- and drain-bias dependences of n and v are calculated under the following boundary conditions. The potentials $V_{\text{sur}}(t)$ and $V_{\text{ch}}(t)$ are fixed to zero at the source terminal and equal to $V_{\text{ds}}(t)$ at the drain terminal. At the gate electrode, $V_{\text{sur}}(t)$ is fixed to the sum of the Schottky-barrier height and the time-dependent gate

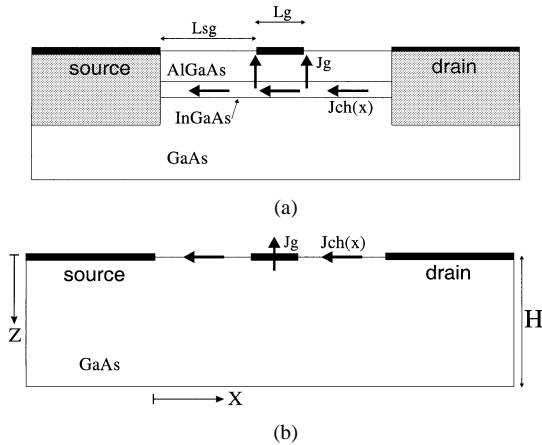


Fig. 2. (a) Schematic representation of the intrinsic device cross section. (b) Model for the EM solver. Design not to scale.

voltage $V_{gs}(t)$. The solution of the system (4) together with (5) allows the channel current density $J_{ch}(t)$ to be calculated. A proper estimation of the conduction dynamic gate current J_g , as well as of the displacement current due to geometric capacitances associated with the electrodes, is required to evaluate the device input impedance and its gain. In the model proposed, this is accounted for by the EM analysis, while the following relation calculates the gate junction current component:

$$J_g(t) = qn(t) \cdot v(t)|_{x=L_{sg}} - qn(t) \cdot v(t)|_{x=L_{sg}+L_g} \quad (6)$$

where L_{sg} is the gate-to-source spacing and L_g is the gate length [see Fig. 2(a)].

III. EM ELECTRON TRANSPORT MODEL COUPLING TECHNIQUE

In a microwave FET, the thickness of the active layer (usually less than $0.1 \mu\text{m}$) is much smaller than the total thickness of the device, usually on the order of tens to hundreds of micrometers [4], [see Fig. 2(a)]. An infinitely thin layer can therefore represent the active region when solving for the EM field in the whole structure [16]. This layer has obviously an active behavior and is represented by nonlinear distributed voltage-controlled current sources to be included in the model of the passive structure [see Fig. 2(b)]. It extends in the horizontal plane between the source and drain electrodes and all along the gatewidth. It is worth noting that for the EM problem such an assumption allows a strong reduction in the computational effort through a less severe space discretization in the z -direction, as is otherwise necessary [17], thus making this approach suitable for an efficient analysis of real electron devices. The EM-wave propagation is completely characterized by solving the Maxwell's equations in the time domain by using a 3-D mesh. The equations are solved in the time domain with the conventional iterative scheme [26].

The active region is represented by controlled current sources: $J_{ch}(x, y)$ at the interface between air and semiconductor and $J_g(y)$ placed in the middle of the gate contact in the z -direction. The controlled currents are a function of the gate-source and gate-drain current via the transport equations

and are intrinsically nonlinear; the equations at each time step must be solved iteratively by means of Newton's algorithm.

The insertion of such a contribution is obtained through the Maxwell's curl \vec{H} equations, as discussed in [26]

$$\begin{aligned} E_x|_{i,j,k}^{N+1} &= E_x|_{i,j,k}^N + \frac{\Delta t}{\epsilon} \nabla \times H|_{i,j,k}^{N+1/2} - \frac{\Delta t}{\epsilon} J_{ch}|_{i,j}^{N+1/2} \\ E_z|_{i,j,k}^{N+1} &= E_z|_{i,j,k}^N + \frac{\Delta t}{\epsilon} \nabla \times H|_{i,j,k}^{N+1/2} - \frac{\Delta t}{\epsilon} J_g|_j^{N+1/2}. \end{aligned} \quad (7)$$

The schematic representation for the channel and gate currents for an elementary device is shown in Fig. 2(b).

For comparison with experimental data, an input signal is applied as a sinusoidal electric field source with a $50\text{-}\Omega$ impedance at the input of the device, i.e., at the gate side, while a $50\text{-}\Omega$ load resistance is placed at the output, i.e., at the drain side of the device.

The outline of the actual simulation process is as follows.

- Step 0) The bias voltages V_{gs} and V_{ds} are supplied to any intrinsic region and the static channel and gate currents are evaluated.
- Step 1) The dynamic simulation begins and the electric and magnetic field distributions are calculated in all the space at the time steps $N\Delta t$ and $(N + 1/2)\Delta t$, respectively, as the result of the input signal.
- Step 2) The electric field x component is extrapolated at the time step $(N + 1/2)\Delta t$ on the basis of its value assumed in the previous time steps. The potentials $V_{ds}(y)$ and $V_{gs}(y)$ are then calculated by using

$$V_{is}(y) = - \int_{\text{source}}^{\text{conductor } i} \vec{E}(y) \cdot d\vec{S}. \quad (8)$$

- Step 3) V_{ds} and V_{gs} are added to their relative static components and used to solve the electron transport model.
- Step 4) The electric field, which is effectively defined at $(N + 1)\Delta t$, is calculated by using (7) and the currents obtained at the previous step.
- Step 5) The magnetic field is evaluated at the step $(N + 3/2)\Delta t$.

Steps 1)–5) are repeated until the entire period of interest is exploited.

IV. EXPERIMENTAL RESULTS

The technique proposed has been applied to two pHEMTs manufactured by AMS Alenia-Marconi System. The active layer structure is given in Table I, and the other parameters of interest are: the source-gate and gate-drain distances $L_{sg} = L_{gd} = 1.2 \mu\text{m}$ and gate length $L_g = 0.3 \mu\text{m}$; the devices have two gate fingers with a gatewidth of 25 and $100 \mu\text{m}$, respectively. With reference to Fig. 1, the grid for the definition of the geometry is: $\Delta x = 0.15 \mu\text{m}$, $\Delta y = 5 \mu\text{m}$, $\Delta z = 10 \mu\text{m}$, while the time stepping Δt is 10^{-15} s in order to prevent numerical instability. The analysis involves

TABLE I
SUMMARY OF THE PHEMT LAYER STRUCTURE

Layer Function	Structure	Thickness [\AA]	Dopant [cm^{-3}]	Al or In content
Cap Layer	GaAs: Si	1200	$3.5\text{E}18$	-
1 st supply layer	AlGaAs: Si	300	$1.0\text{E}17$	0.22
2 nd supply layer	AlGaAs: Si	150	$2.0\text{E}18$	0.22
Spacer	AlGaAs	20	-	0.22
Channel	InGaAs	130	-	0.15
Buffer layer	GaAs	5500	-	-
S.I. substrate	GaAs	1E6	-	-

TABLE II
PHYSICAL PARAMETERS USED FOR CHARGE CONTROL CALCULATIONS

Parameter	$\text{Al}_{0.3}\text{Ga}_{0.7}\text{As}$	$\text{In}_{0.15}\text{Ga}_{0.85}\text{As}$	GaAs
band gap [eV]	1.814	1.220	1.439
electron effective mass	0.089	0.063	0.063
light-hole effective mass	0.084	0.076	0.085
heavy-hole effective mass	0.450	0.444	0.410
conduction band density of states at 300K [cm^{-3}]	$6.988\text{E}17$	$4.043\text{E}17$	$4.350\text{E}17$
valence band density of states at 300K [cm^{-3}]	$9.313\text{E}18$	$9.588\text{E}18$	$9.985\text{E}18$
relative dielectric constant	12.1	13.4	13.1

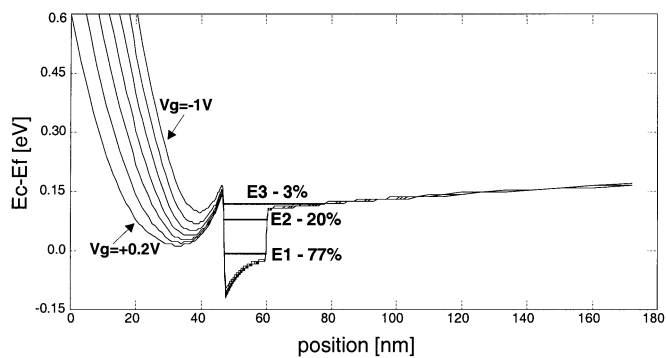


Fig. 3. Band energy diagram calculated for a gate bias voltage V_{gs} ranging from -1 to $+0.2$ V, with 0.2 -V steps and $V_{ds} = 0$ V.

a longitudinal perfect magnetic wall to reduce the simulation domain (see Fig. 1).

A. Intrinsic Physical Results

The analysis begins with the calculation of the charge distribution in the pHEMT active area by using the charge-control model described in Section II-A. The physical parameters used in the present paper are listed in Table II. It is assumed that the conduction band edge discontinuities at the heterointerfaces are 85 percent of the bandgap discontinuity as in [24].

Fig. 3 shows the calculated conduction band profile for a gate bias voltage ranging from -1 to $+0.2$ V. The induction of a 2DEG system in the InGaAs channel is confirmed; the calculation allows the subband levels to be evaluated and their fractional occupation. For the case under investigation, $E1 = -1.84$ meV, $E2 = 75$ meV, and $E3 = 91.2$ meV with corresponding probabilities of occupation of 77%, 20%, and 3%, respectively. From this calculation, it is possible to observe that only the first two subband energy levels give a contribution to the channel electron density, while $E3$ and higher levels are empty of charge.

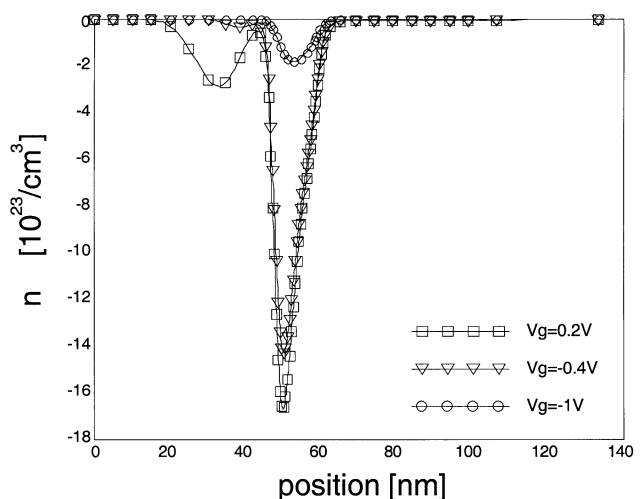


Fig. 4. Total charge density calculated for a gate bias voltage V_{gs} ranging from -1 to $+0.2$ V, with $V_{ds} = 0$ V.

Although not reported here, subbands are also formed in the supply and spacer layers, their energy distribution being very sensitive to the bias voltage and responsible for the well-known MESFET-like parasitic channel. The energy diagram shows a flat-band behavior in the buffer layer and the semi-insulating substrate; hence, the component of the electric field E_z is equal to zero.

Fig. 4 reports the corresponding total charge density distribution profile in the semiconductor structure including the ionized donor atoms and the electrons in the conduction band, for the gate potential ranging from -1 to $+0.2$ V, with the drain and source potentials kept to zero. The graph shows that the charge distribution vanishes for a distance from the gate electrode larger than 70 nm. This means that any electron flow through the substrate during the dynamic operation is only due to displacement currents, while conduction currents are

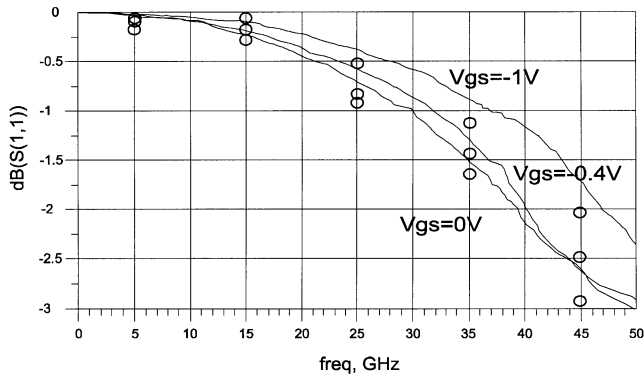


Fig. 5. Comparison between simulated (symbols) and measured (solid lines) S_{11} for $V_{ds} = 3$ V and different gate bias for the $2 \times 25 \mu\text{m}$ gatewidth pHEMT.

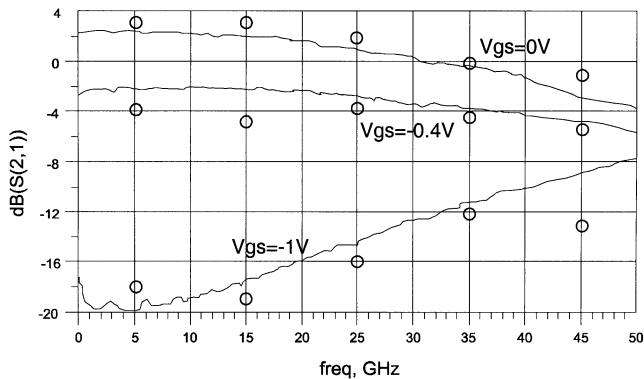


Fig. 6. Comparison between simulated (symbols) and measured (solid lines) S_{21} for $V_{ds} = 3$ V and different gate bias for the $2 \times 25 \mu\text{m}$ gatewidth HEMT.

not allowed. Consequently, the assumption made in Section III is consistent with the device physics.

The charge control model presented above has been implemented by using a finite difference one-dimensional (1-D) spatial mesh and solved by the Gummel method; the results are then made available in terms of an LUT then inserted into the dynamic calculation by using (5).

B. Microwave Results

In order to validate this approach, S -parameters measured and modeled up to 50 GHz are shown in Figs. 5–8 for both the HEMT structures, three gate voltages from open channel to pinchoff, and 3-V drain bias voltage. In order to compare the results of the analysis with the device performance, the actual device layout is considered which includes the in–out transmission lines and the ground–signal–ground (G–S–G) contact pads. The analysis is performed entirely in the time domain, and linear and nonlinear operations of this device are allowed. Postprocessing of the data provides a deep insight into the device dynamics and the small-signal results at microwave and millimeter-wave frequencies.

The analysis of the $2 \times 25 \mu\text{m}$ transistor has been performed in two separated steps, first by feeding from the gate terminal, then from the drain terminal and sensing the reflected and the transmitted signals. The incoming signal (a_1 - and a_2 -parameters) have been supplied through a $50\text{-}\Omega$ impedance by using the technique described in [27], while the calculation of the

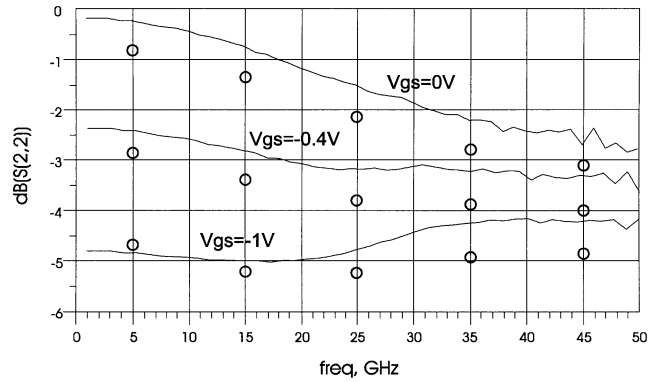


Fig. 7. Comparison between simulated (symbols) and measured (solid lines) S_{22} for $V_{ds} = 3$ V and different gate bias for the $2 \times 25 \mu\text{m}$ gatewidth pHEMT.

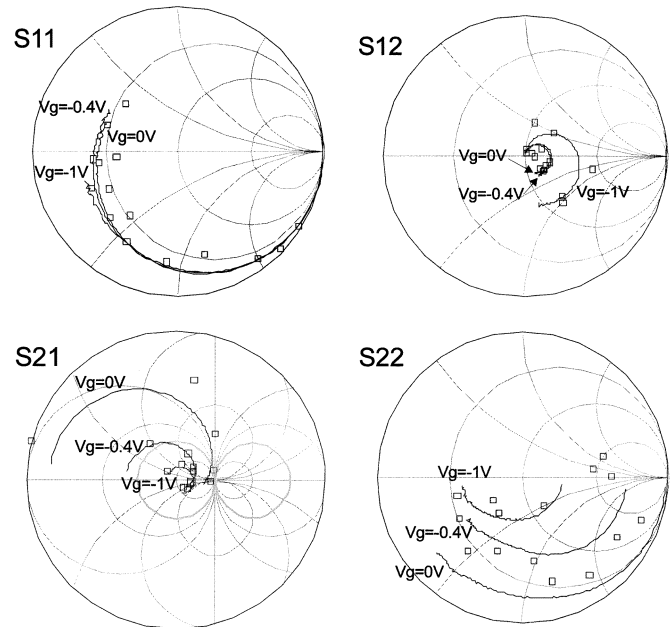


Fig. 8. Comparison between simulated (symbols) and measured (solid lines) S_{22} for $V_{ds} = 3$ V and different gate bias for the $100\text{-}\mu\text{m}$ gatewidth HEMT; the simulation points are 5–45 GHz in 10-GHz steps.

b_1 - and b_2 -parameters is obtained in the same way as that of measuring the voltage for a lumped resistor. This has allowed the calculation of the magnitude for S_{11} , S_{21} , and S_{22} reported in Figs. 5–7, respectively, while the corresponding phase responses are reported in terms of percent relative errors in Table III. Finally the same comparison is reported in Table IV for the S_{12} -parameter.

In order to test the actual model capability, the $2 \times 100 \mu\text{m}$ device has been modeled and the results compared with corresponding measurements. The transistor has exactly the same semiconductor structure and was conceived by applying the same design rules; the only change with respect to the previous device is the gatewidth. In that case, the EM effects would be more significant. The comparisons between measurement and simulation, up to 50 GHz, are reported in Fig. 8. From the graphs reporting S_{21} for both the pHEMTs, Figs. 6 and 8(c), it is possible to note that the gate voltage of the data measured is actually shifted by 0.2 V with respect to the simulated data; this data is easily estimated by the $I_d \setminus V_g$ graph reported in

TABLE III
S11, S21, AND S22 PHASE PERCENT ERROR AT DIFFERENT GATE BIASES AND FREQUENCIES FOR THE 2×25 GATEWIDTH pHEMT

Freq	Phase (S11)			Phase (S21)			Phase (S22)		
	Vg=0V	Vg=-.4V	Vg=-1V	Vg=0V	Vg=-.4V	Vg=-1V	Vg=0V	Vg=-.4V	Vg=-1V
5GHz	-1.0E1	2.4E1	-8.6E1	7.4	-1.4E1	-2.8E1	-2.3E1	-1E1	-0.2
25GHz	-1.3E1	1.3E1	-2.0E1	-3.7E1	3.8	-3.4E2	-1.7E1	-1.0E1	1.2E1
45GHz	1.3E1	2.0E2	-1.3E1	-6.8E1	2.0E2	3.6E1	1.2E2	1.1E2	1.0E2

TABLE IV
S12 PERCENT ERROR AT DIFFERENT GATE BIASES AND FREQUENCIES FOR THE 2×25 GATEWIDTH pHEMT

Freq	Magnitude (S12)			Phase (S12)		
	Vg=0V	Vg=-.4V	Vg=-1V	Vg=0V	Vg=-.4V	Vg=-1V
5GHz	4.4E1	6.8E1	-1.2E2	-34.1	-16.0	-7.2
25GHz	6.4E1	1.1E1	-2.6E1	97.2	-75.2	-63.4
45GHz	2.1E1	-1.2E1	-1.2E1	20.9	20.3	16.5

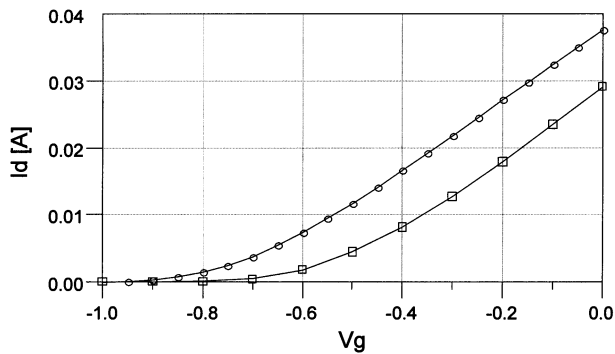


Fig. 9. Comparison between simulated (circle) and measured (square) drain current for $V_{ds} = 3$ V and different gate bias for the $2 \times 100 \mu\text{m}$ gatewidth pHEMT.

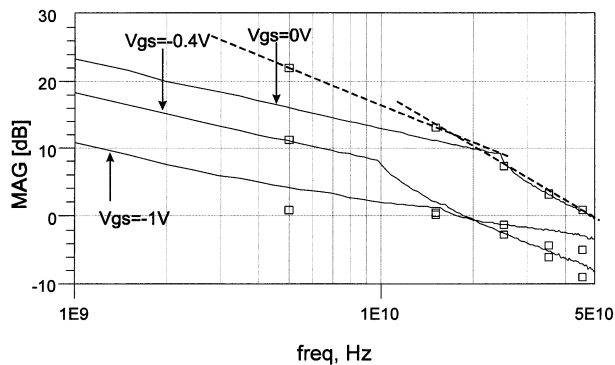


Fig. 10. Comparison between simulated (symbols) and measured (solid) MAG for $V_{ds} = 3$ V and different gate bias for the 2×100 gatewidth HEMT.

Fig. 9 for the $2 \times 100 \mu\text{m}$ device. This is probably due to the uncertainty of the technological process. The shift in the current parameter is also responsible for a fictitious increase in drain current and consequently in a gain increase. The results are otherwise satisfactory and promising for an accurate prediction of the device performance.

The results are also reported in terms of MSG-MAG for this device (see Fig. 10). This latter parameter assumes relevance because it may be used as a device figure-of-merit during the

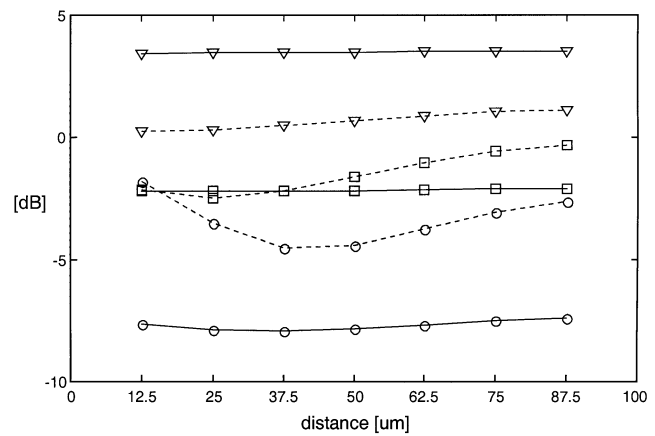


Fig. 11. Signal power distribution through the drain (solid lines) and gate (dashed lines) electrode at 25 GHz (triangle), 50 GHz (square), and 75 GHz (circle). $V_{ds} = 3$ V and $V_{gs} = 0$ V.

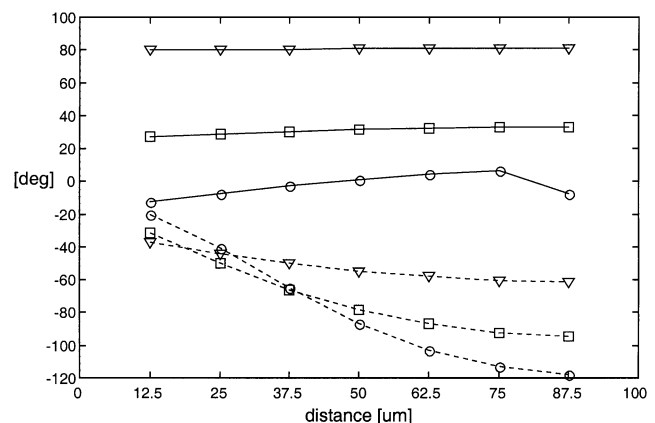


Fig. 12. Signal phase distribution through the drain (solid lines) and gate (dashed lines) electrode at 25 GHz (triangle), 50 GHz (square), and 75 GHz (circle). $V_{ds} = 3$ V and $V_{gs} = 0$ V.

optimization of the device structure itself. From measurements, we see that the frequency limits, below which the transistor becomes potentially unstable, change with the gate potential, as expected. In the case of $V_{gs} = 0$ V, we estimate a frequency limit around 20 GHz for both the measured and simulated data.

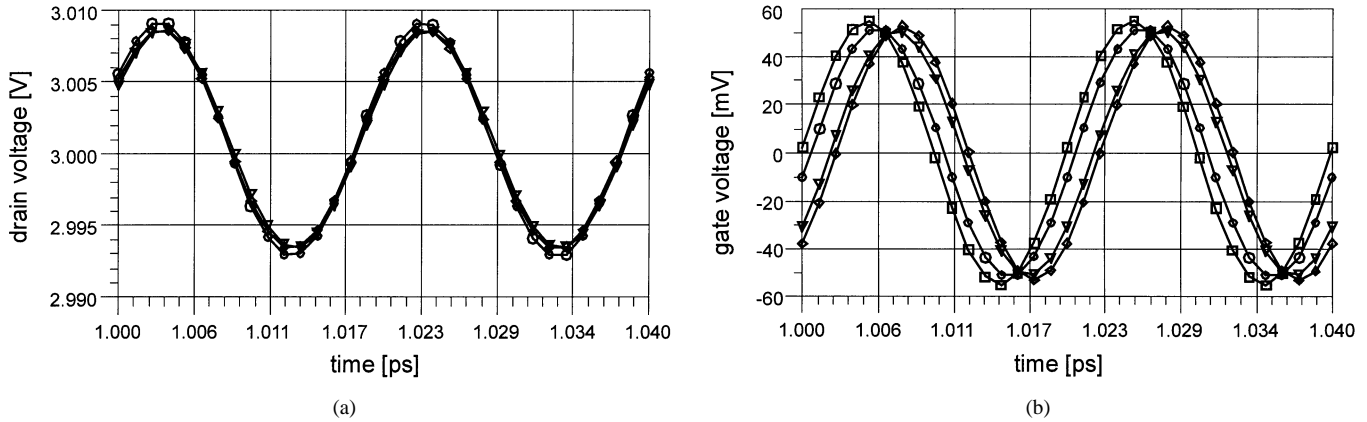


Fig. 13. Signal distribution through the: (a) drain and (b) gate electrodes sampled at the following points. Circles: $y = 12.5 \mu\text{m}$. Triangles: $y = 50 \mu\text{m}$. Diamond lines: $y = 87.5 \mu\text{m}$. In (b), the input 50-GHz sinusoidal signal is also reported (squares).

Consequently, the design procedure can be driven by the maximization of this parameter around the operation frequency. Its accuracy depends on the accuracy with which the scattering parameters are predicted; as a result, it may be slightly poorer, although the comparison in Fig. 10 shows a fairly constant prediction over the gate bias voltages.

C. Distributed Effects

The approach described herein allows an insight to the intrinsic device behavior that becomes important during the device design. Besides the electron charge accumulation reported in Fig. 4, it is also interesting to observe the signal distribution through the gate and drain electrodes for the $2 \times 100 \mu\text{m}$ device. This calculation is made by considering the integration of the electric field in the x -direction by keeping $z = 0$ (see Figs. 1 and 2). The potential is calculated with respect to the source electrode for y which spans over the whole intrinsic active region. In order to show how the distribution of the input and output signals depend upon the frequency, the calculation is repeated for 25, 50, and 75 GHz for a test input signal of 0 dBm of available power. Fig. 11 reports the difference between the inner gate-source voltages at any intrinsic devices and the input voltage in decibels as a function of the distance from the input, while Fig. 12 depicts the same difference in phase.

At this point, some remarks can be conveniently made. The most important phenomenon consists in the signal attenuation between the supply generator and the inner gates; this can be avoided with a matching network. Once the signal is fed through the gate, it is appreciably delayed (dashed line in Fig. 12), but only slightly attenuated when propagating through the electrode (dashed lines in Fig. 11). The phase delay depends upon the frequency and is more noticeable for the gate than for the drain. The attenuation is not very sensitive to the distance from the access point; in particular, the drain voltage is constant through the electrode even though it is strongly dependent on the frequency. This is due to the size of the drain electrode that acts as a reference plane for the drain of the intrinsic devices; a comparable result is not obtained for the gate due to its very narrow width. The nonlinear dependence on the distance of the gate signal magnitude, noticeable at 50 and 75 GHz, is due to the combination of attenuation and feedback which are associated with

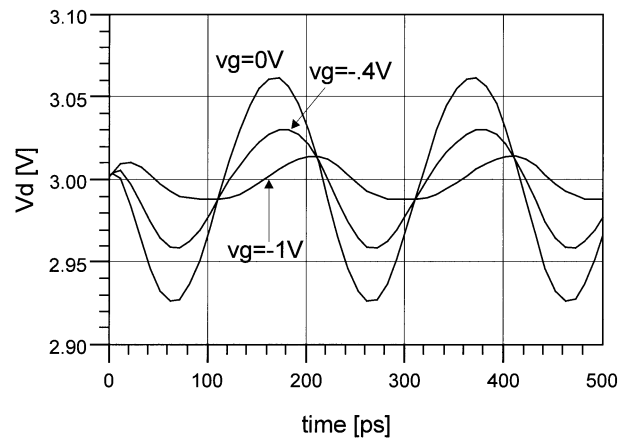


Fig. 14. Internal drain voltages at different gate biases for $V_{ds} = 3 \text{ V}$ bias voltage.

the drain-gate capacitance. For the particular case of 50 GHz, it is also apparent that the magnitude of the drain signal is in excess of the gate signal up to a distance of $37.5 \mu\text{m}$; this means that at this frequency the complete device is not efficiently exploited. This result is also consistent with the cut-off frequency calculated from the MAG shown in Fig. 10. It can be concluded that the device should be shortened to a gatewidth of less than $37.5 \mu\text{m}$, if intended for an operation frequency of 50 GHz.

The same results in the time domain are reported in Fig. 13 for an input signal of 50 GHz. This approach allows the large-signal model capability [18]–[28]. In Fig. 14, the drain voltage is reported which is sampled in the middle of the gate electrode ($y = 50 \mu\text{m}$), as the response to an input signal at 5 GHz 1 V peak to peak. The distortion of the waveform confirms the capability of this technique to deal with a nonlinear regime although a complete discussion of this topic is beyond the aim of this paper.

V. CONCLUSION

The approach presented herein deals with the problem of global modeling for state-of-the-art pHEMTs operating at frequencies where the EM-field interaction significantly affects the device performance. This technique is based only on geometrical and material data and includes the transport phenomena in the active device, charge control due to quantum

mechanical properties, and the EM interaction between the intrinsic device and the electrodi-electric structure of the access elements. The approach proposed in this paper takes advantage of the physically expected assumption that the charge distribution within the intrinsic device could be considered as thin and located near the semiconductor-air interface; the resulting intrinsic device description is more easily inserted into the EM code. This leads to a comparatively very simple code which runs on a PC in a few hours per bias point depending on the platform, thus allowing the designer to predict the behavior of the device and potentially optimizing its performance with a reasonable degree of accuracy and limited computer-time consumption. Although the EM problem has been presently solved by performing a 3-D analysis, this approach allows simplified schemes to be considered like the planar 3-D technique implemented in many commercial tools.² In order to validate this technique, the comparison between experimental and simulated small-signal results up to 50 GHz is presented for two gate fingers devices manufactured by Alenia-Marconi with a 25- μm and 100- μm gatewidth, respectively, and a 0.3- μm gate length, at bias points ranging from I_{dss} to the pinchoff regime.

REFERENCES

- [1] P. A. Sandborn, A. Rau, and P. A. Blakey, "An assessment of approximate nonstationary charge transport models used for GaAs device modeling," *IEEE Trans. Electron Devices*, vol. 36, pp. 1244–1253, July 1989.
- [2] P. A. Sandborn, J. R. East, and G. I. Haddad, "Quasi two-dimensional modeling of GaAs MESFET's," *IEEE Trans. Electron Devices*, vol. ED-34, pp. 981–985, May 1987.
- [3] M. Snowden and R. R. Pantoja, "Quasitwo-dimensional MESFET simulation for CAD," *IEEE Trans. Electron Devices*, vol. 36, pp. 1564–1574, Sept. 1989.
- [4] T. Shawki, G. Salmer, and O. El Sayed, "MODFET 2-D hydrodynamic energy modeling: Optimization of subquarter-micron-gate structures," *IEEE Trans. Electron Devices*, vol. 37, pp. 21–29, Jan. 1990.
- [5] R. O. Grondin, S. El-Ghazaly, and S. Goodnick, "A review of global modeling of charge transport in semiconductors and full-wave electromagnetics," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 817–829, June 1999.
- [6] K. P. Ma, M. Chen, B. Houshmand, Y. Qian, and T. Itho, "Global time-domain full-wave analysis of microwave circuits involving highly nonlinear phenomena and EMC effects," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 859–866, June 1999.
- [7] M. Farina and T. Rozzi, "Full-wave modeling of linear FET's for millimeter waves," *IEEE Trans. Microwave Theory Tech.*, vol. 49, pp. 1443–2450, Aug. 2001.
- [8] S. Goasguen and S. M. El-Ghazaly, "A practical large-signal global modeling simulation of a microwave amplifier using artificial neural network," *IEEE Microwave Guided Wave Lett.*, vol. 10, pp. 273–275, July 2000.
- [9] P. Ciampolini, L. Roselli, G. Stopponi, and R. Sorrentino, "Global modeling strategies for the analysis of high-frequency integrated circuits," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 950–955, June 1999.
- [10] I. Witzig, C. Schuster, P. Regli, and W. Fichtner, "Global modeling of microwave applications by combining the FDTD method and a general semiconductor device and circuit simulator," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 919–928, June 1999.
- [11] S. Beausart, O. Perrin, M. R. Friscourt, and C. Dalle, "Millimeter-wave pulsed oscillator global modeling by means of electromagnetic, thermal, electrical, and carrier transport physical coupled models," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 929–934, June 1999.
- [12] M. B. Steer *et al.*, "Global modeling of spatially distributed microwave and millimeter-wave systems," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 830–839, June 1999.
- [13] M. A. Alsunaidi, S. M. Sohel Intiaz, and S. M. El-Ghazaly, "Electromagnetic wave effects on microwave transistors using a full-wave time domain model," *IEEE Trans. Microwave Theory Tech.*, vol. 44, pp. 799–807, June 1996.
- [14] A. Cidronali, G. Collodi, G. Vannini, A. Santarelli, and G. Manes, "A new approach to FET model scaling and MMIC design based on electromagnetic analysis," *IEEE Trans. Microwave Theory Tech.*, vol. 47, pp. 900–907, June 1999.
- [15] A. Cidronali, G. Collodi, G. Leuzzi, and G. Manes, "Numerical analysis of a 0.2- μm AlGaAs/GaAs HEMT including electromagnetic effects," in *Proc. IEEE Int. Symp. Compound Semiconductor*, San Diego, CA, Sept. 7–11, 1997, pp. 635–638.
- [16] A. Cidronali, "Modelli fisici e comportamentali di FET ad onde millimetriche," Ph.D. dissertation, Univ. of Florence, Florence, Italy, 1998.
- [17] A. Monorchio, A. Cidronali, G. Manara, and G. Pelosi, "A hybrid finite element/finite difference time domain electromagnetic approach to the analysis of high frequency FET devices," *Int. J. Numer. Modeling*, vol. 13, no. 1, pp. 289–300, Jan.–Feb. 2000.
- [18] P. J. Rudge, R. E. Miles, M. B. Steer, and C. M. Snowden, "Investigation into intermodulation distortion in HEMT's using a quasi-2-D physical model," *IEEE Trans. Microwave Theory Tech.*, vol. 49, pp. 2315–2321, Dec. 2001.
- [19] R. Singh and C. M. Snowden, "A quasitwo-dimensional HEMT model for DC and microwave simulation," *IEEE Trans. Electron Devices*, vol. 45, pp. 1165–1169, June 1998.
- [20] C. G. Morton, C. M. Snowden, and M. J. Howes, "HEMT physical model for MMMIC CAD," in *Proc. 25th Eur. Microwave Conf.*, Sept. 1995, pp. 199–204.
- [21] C. M. Snowden and R. R. Pantoja, "GaAs MESFET physical models for process-oriented design," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 1401–1409, July 1992.
- [22] R. Drury and C. M. Snowden, "A quasitwo-dimensional HEMT model for microwave CAD applications," *IEEE Trans. Electron Devices*, vol. 42, pp. 1026–1032, June 1995.
- [23] C. G. Morton and C. M. Snowden, "Comparison of quasi-2D and ensemble Monte Carlo simulations for deep submicron HEMT's," in *IEEE MTT-S Int. Microwave Symp. Dig.*, vol. 1, 1998, pp. 153–156.
- [24] J. Yoshida, "Classical versus quantum mechanical calculation of the electron distribution at the n-AlGaAs/GaAs heterointerface," *IEEE Trans. Electron Devices*, vol. 33, pp. 154–156, Jan. 1986.
- [25] B. Carnez, A. Cappy, A. Kaszynski, E. Constant, and G. Salmer, "Modeling of a submicrometer gate field-effect transistor including the effect of nonstationary electron dynamics," *J. Appl. Phys.*, vol. 51, pp. 784–790, Jan. 1980.
- [26] W. Sui, D. A. Christensen, and C. Durney, "Extending the two-dimensional FDTD method to hybrid electromagnetic systems with active and passive lumped elements," *IEEE Trans. Microwave Theory Tech.*, vol. 40, pp. 724–730, Apr. 1992.
- [27] M. Picket-May, A. Taflove, and J. Baron, "FD-TD modeling of digital signal propagation in 3-D circuits with passive and active loads," *IEEE Trans. Microwave Theory Tech.*, vol. 42, pp. 1514–1524, Aug. 1994.
- [28] F. Giannini, G. Leuzzi, M. Kopanski, and G. Salmer, "Large-signal analysis of quasi-2D physical model of MESFETs," *Electron. Lett.*, vol. 29, no. 21, pp. 1891–1893, Oct. 1993.



Alessandro Cidronali (M'89) was born in Florence, Italy, in 1965. He received the Laurea and Ph.D. degrees in electronic engineering from the University of Florence, Florence, Italy, in 1992 and 1998, respectively.

In 1993, he joined the Department of Electronics Engineering, University of Florence, where he became an Assistant Professor in 1999. During his academic career, he has been a Lecturer in courses on "Applied Electronics and Solid State Electronics" and currently teaches "Microwave Electronics."

His research activities cover the study of active and passive compact structures for monolithic microwave integrated circuits (MMICs), the design of multifunction MMICs for low-power wireless applications, computer-aided design (CAD), and numerical modeling of HBTs and HEMTs for microwave and high-speed applications. He is currently involved with basic research on quantum functional devices and their applications to microwave circuits.

²em," Sonnet Software Inc., Liverpool, NY.



Giorgio Leuzzi received the degree (*cum laude*) from the University of Roma "La Sapienza," Rome, Italy, in 1982. His thesis concerned transmission lines for microwave integrated circuits.

Following military service, he became a Teaching and Research Assistant with the University of Rome "Tor Vergata," Rome, Italy, where, since 1991, he has taught "Microwave Electronics." In 1990, he spent several months with the Centre Hyperfréquences et Semiconducteurs, University of Lille, Lille, France. In 1998, he became an

Associate Professor and, in 2001, a Full Professor of electronic devices with the University of L'Aquila, Monteluco di Roio-L'Aquila, Italy. His research activities involve the simulation of semiconductor devices for microwaves and millimeter waves, the linear and nonlinear characterization of microwave and millimeter-wave active devices, and the development of methods and algorithms for microwave nonlinear circuit design. He took part in several research projects, both Italian (CNR MADESS, MURST Microelettronica, and ASI) and international (ESPRIT Projects COSMIC, MANPOWER, FUSE, EDGE, and ESA). He regularly teaches international summer courses of the ITSS and has regular collaborations with corporations such as Alenia Marconi Systems and Agilent Technologies.



Gianfranco Manes (SM'02) was born in Florence, Italy, on November 16, 1944.

He became an Associate Professor in 1980 and a Full Professor in 1985 with the University of Florence, Florence, Italy. He has contributed, since the early stages, to the field of surface-acoustic-wave (SAW) technology for radar signal processing and electromagnetic compatibility (ECM) applications. His major contributions have been the introduction of novel FIR synthesis techniques, fast analog spectrum analysis configurations, and FH waveform synthesis. Since the early 1980s, he has been active in the field of microwave modeling and design. In the early 1990s, he founded and currently leads the Microelectronics Laboratory, University of Florence, where he is committed to research in the field of microwave devices. In 1982, he was committed to beginning a facility for the design and production of SAW and MIC/MMIC devices, as a subsidiary of the Florence radar company SMA Spa. In 1984, this facility became a standalone privately owned microwave company Micrel SpA, which operates in the field of defence electronics and space communications. During 1996–2000, he was involved in IV FW projects in the field of information technology applied to the cultural heritage and was invited to orientation meetings and advisory panels for the Commission. His current research interest is in the field of RITD devices for microwave applications in a scientific collaboration with the Group at the Physical Science Research Laboratories, Motorola Corporation, Tempe, AZ. He was the founder and is currently President of MIDRA, a research consortium between the University of Florence and the Motorola Corporation. He is Director of the Italian Ph.D. School in Electronics. In November 2000, he was appointed Deputy Rector for the Information System of the University of Florence. He has authored over 100 scientific papers in learned society journals and presentations at International Conferences.

Dr. Manes is member of the Board of Italian Electronics Society.



Franco Giannini was born in Galatina (LE), Italy, on November 9, 1944. HE received the Electronics Engineering degree (*summa cum laude*) in 1968.

Since 1980, he has been a Professor of applied electronics. He is currently with the University of Roma "Tor Vergata," Rome, Italy. He has been involved with problems concerning modeling, characterization, and design methodologies of passive and active microwave components, circuits, and subsystems, including GaAs microwave and millimeter-wave monolithic integrated circuits. He is a consultant for various national and international industrial and governmental organizations, including the International Telecommunication Union and the European Union.

Prof. Giannini has been a member of many committees for international scientific conferences on GaAs and on microwave and millimeter-wave techniques and applications.