Three-Dimensional Electromagnetic Approach to the Modeling of Linear Field Effect Transistors

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Abstract **— In this contribution we introduce a 3D electromagnetic approach to the modeling of active devices under the small-signal hypothesis. The proposed technique is validated by comparing measured and calculated results for a pseudomorphic High Electron Mobility Transistor (HEMT) in the millimeter-wave range.**

I. Introduction

In the past decade the design process for high frequency passive components has been greatly enhanced by the commercial availability of powerful, yet easy to use, full-wave tools. In spite of the large number of electromagnetic solvers addressing passive structures, to the best of our knowledge no commercial program is already able to directly model *active* linear devices.

The standard procedure currently used to obtain a reliable model of transistors for high frequencies is much like the one described in [1] or, e.g, in the EMSight user's manual, by AWR. The passive periphery is characterized by a full-wave approach, possibly defining internal ports, where lumped active elements are connected at circuit level. The lumped equivalent circuit of the active element is usually provided by some sort of sample measurement, and by a fitting technique.

Such an approach is very simple, may be applied by using nearly any suitable em solver for *passive* structures and could be even used in the non- linear case. Among these remarkable advantages, the above approach suffers of a major drawback: it cannot predict the effects of possible standing waves along the active finger itself, that, as shown in [2], may have dramatic impact on the performance of an electrically large device. This is particularly relevant when scaling effects have to be evaluated in millimeter wave transistors.

Owing to the need of a full-wave analysis also accounting for the semiconductor equations, a good deal of research work has been devoted to the so-called "Global Modeling" [3]. In this framework in [4] we proposed a 2D Generalized Transverse Resonance-Diffraction Approach (GTRD) that was shown to work well under the small-signal hypothesis. It should be stressed that results in [4] showed the *predictive* ability of the technique, what is out of the scope of anything involving a trim-to-fit procedure.

In particular, in [4] an eigenvalue integral equation was generated by using GTRD: its solution provided the three fundamental quasi-TEM modes of a single FET finger. Using such modal properties, the FET finger was treated as a coupled active multi-conductor transmission line; overall performance was recovered by connecting several fingers at circuit level, along with an em model of the passive periphery obtained by a commercial software package.

The major drawback of the technique in [4] was its modal nature: solving the implicit eigenvalue equation was time-consuming and somehow difficult, due to the existence of spurious modes. Moreover the approach relied onto the quasi-TEM nature of the fundamental modes and the possibility to neglecting high order modes.

In [5] we introduced a deterministic three-dimensional form of GTRD for passive structures. Such an approach was found to have the advantages of standard so-called 2.5D techniques while being 3D in its very nature. It was subsequently implemented in a commercial software package, EM3DS by MEM Research.

The present work is devoted to the extension of the approach [5] to the analysis of active devices, in particular Field-Effect Transistors, under the small signal hypothesis. In this way, multi-mode effects are rigorously accounted for, along with the interactions between periphery and active parts of the device.

As a validation, the technique is applied to a pseudomorphic HEMT, suitable for millimeter-wave applications: the device Lp7512 by Filtronic.

II. Theory

Figure 1 shows a schematic cross-sectional view of a pHEMT featuring a "mushroom" recessed gate. It should be noted that the two-dimensional electron gas (2DEG) is generally only 80 Å thick, while the substrate is 100 µm: the cross-sectional aspect ratio is at least 1250. Moreover the whole substrate appears to be quite complex, including several dielectric layers. These considerations highlight how an approach like GTRD [5], where the layered substrate is modeled by means of a Green's function, is highly suitable.

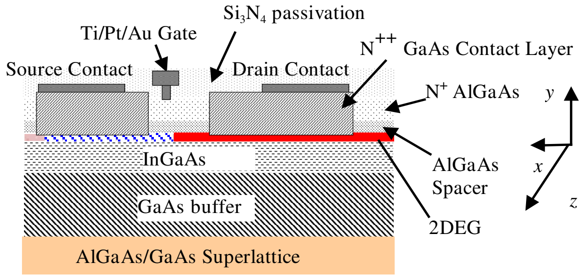

Fig. 1: pHEMT schematic cross-section.

The doped AlGaAs is assumed to be completely depleted under normal operating conditions. The 2DEG is considered to be composed by two separate regions: a first one between source and gate, where carriers are traveling below their saturation velocity and a second one, between gate and drain, where saturation occurs. The small-signal conductivity of such regions being

$$\sigma = q\, n\, \mu \tag{1}$$

where n is the carrier density and μ the carrier mobility, the gate-to-source region appears to the microwave signal as a high conductivity region, owing to the large value of μ. The resulting conductivity in the specific example is $5.2\ 10^5$ S/m. This value may be obtained by Hall's measurements, but in our case it is calculated by using the value of the intrinsic gate-to-source resistance available in the manufacturer's data-sheet, as done in [4] for the 2D case. By the same token the gate-to-drain region, where carriers are travelling at their saturation velocity, features a lower conductivity, around 5400 S/m, partly responsible for the FET output resistance.

Biasing the FET has two first-order effects: the first one is that a static carrier flow, with some velocity gradient, is established. The microwave (small-) signal sees this flow like a conductive region with a conductivity gradient, that may be approximated as discussed above.

The second effect is that the microwave signal impinging on the gate Schottky junction modulates the 2DEG density. Under the small-signal hypothesis, such a modulation appears as a controlled current source for the microwave signal, flowing across the channel.

Hence a controlled current (density) source is sufficient to describe the power exchange between the static biasing and the dynamic microwave signal. The above general considerations are valid regardless the kind of em approach that will be used in the subsequent analysis, provided that it has the ability to account for controlled sources.

In this paper the em technique is the GTRD approach, in its deterministic 3D formulation given in [5], but improved ion order to account for controlled sources.

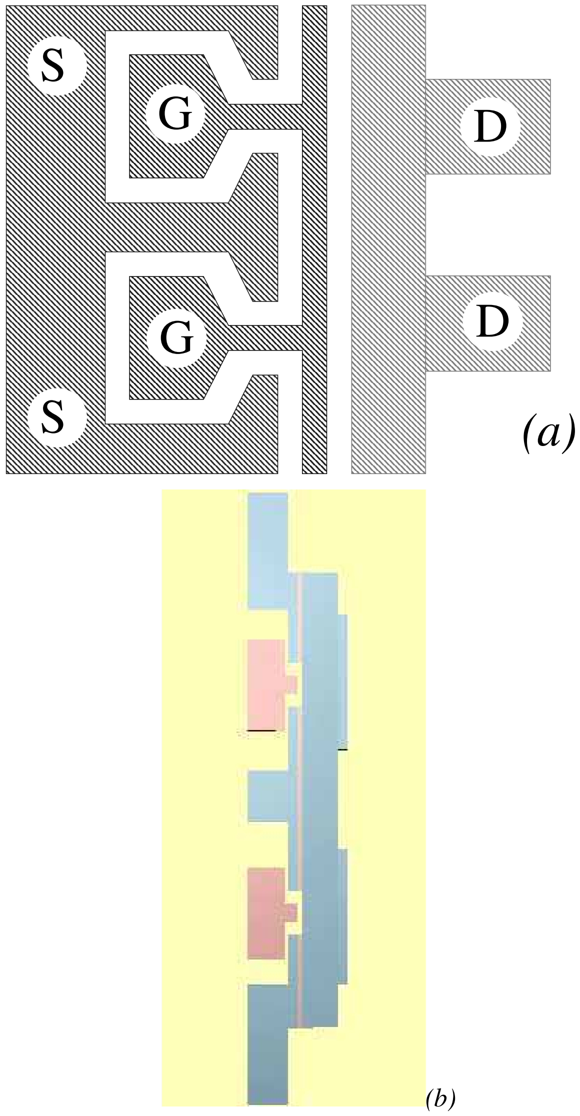

Fig. 2: Schematic topology of Lp7512 (*a*) and its model (with actual aspect ratio) (*b*). Part of the source metal is not modeled as unnecessary.

Given the Green's function of a dielectric stack **Z**, it relates current sources to electric field according to

$$\mathbf{E}(\mathbf{r}) = -\iiint_V d\mathbf{r'}\, \mathbf{Z}(\mathbf{r},\mathbf{r'}) \cdot \mathbf{J}(\mathbf{r'}) \tag{2}$$

Expression (2) just makes explicit the relationship between field and currents appearing in the Maxwell's equations. Moreover the Green's function accounts for every boundary condition pertaining the horizontally (namely along the xz plane, fig.1) uniform dielectric stack. Current **J** may be either a displacement current, an ohmic current or also an impressed source, and it is used to describe any entity not being horizontally uniform, like conductors, dielectric discontinuities and controlled sources. By defining a complex dielectric permittivity, both ohmic and displacement currents are related to the field by [5]

$$\mathbf{E}(\mathbf{r}) = \frac{\mathbf{J}(\mathbf{r})}{j\omega\varepsilon_0\left(\tilde{\varepsilon}(\mathbf{r}) - \tilde{\varepsilon}'(\mathbf{r})\right)} \tag{3}$$

where $\tilde{\varepsilon}$ is the complex permittivity of the additional dielectric or conductor, while $\tilde{\varepsilon}'$ is the one of the embedding medium.

The controlled source, flowing across an ohmic region, may be described as

$$\mathbf{J}(\mathbf{r}) = \mathbf{G}(\mathbf{r},\mathbf{r}')\cdot\mathbf{E}(\mathbf{r}') + \sigma(\mathbf{r})\mathbf{E}(\mathbf{r}) \tag{4}$$

where σ is the channel conductivity, and **G** a transconductive operator that relates the impressed controlled source to the impinging microwave signal. A suitable expression is

$$\mathbf{G}(\mathbf{r},\mathbf{r}') = \frac{g_m}{wd} e^{-j\omega\tau}\,\delta(u'-ctrl)\,rect(y',sh)\,\mathbf{uy} \tag{5}$$

where

$\mathbf{r}' \in rect(u',sh)\,rect(v',sh)\,rect(y',sh)$

$\mathbf{r} \in rect(u,ch)\,rect(v,ch)\,rect(y,ch)$

In (5) g_m is the *static* device transconductance, w its total width, d the 2-DEG thickness, τ the transit-time. u and v are either x and z respectively if the controlled source is oriented along x, or z and x, if the controlled source is oriented along z. δ is the Dirac impulsive function, while *rect* are functions assuming value 1 in a given region and 0 elsewhere. In particular *ctrl* identifies the region assumed to be the control plane, namely where the amplitude of the impinging signal is considered to control the current source (usually the center of the gate), *sh* is the Shottky contact while *ch* is the 2DEG. Operator (5), when applied to a field, relates the impressed current into the channel to the voltage calculated across the Shottky contact.

The deterministic integral equation is obtained by combining expressions (2-5) and by defining some kind of excitation: in this paper we use standard delta-gap field excitation [5], and a port p is defined by applying the unitary impressed field. Combination of equations (2-5) and of the impulsive excitation provides the integral equation

$$\iiint d\mathbf{r}''\Big\{\frac{1}{j\omega\varepsilon_0\left(\tilde{\varepsilon}(\mathbf{r}) - \tilde{\varepsilon}'(\mathbf{r})\right)}\mathbf{G}(\mathbf{r},\mathbf{r}')\cdot\mathbf{Z}(\mathbf{r}',\mathbf{r}'') + \mathbf{Z}(\mathbf{r},\mathbf{r}'') + \frac{\mathbf{I}(\mathbf{r},\mathbf{r}'')}{j\omega\varepsilon_0\left(\tilde{\varepsilon}(\mathbf{r}) - \tilde{\varepsilon}'(\mathbf{r})\right)}\Big\}\cdot\mathbf{J}(\mathbf{r}'') = \delta(v-v_0)\,rect(p)\,\mathbf{u} \tag{7}$$

$v = x$, y or z, and **u** is the direction normal to the port plane. Currents across the whole structure, and network parameters, are obtained by solving (7).

III. Results

The method has been validated by modeling the pseudomorphic HEMT by Filtronic Lp7512, for which sufficient information was available. It is a .25 µm gate "mushroom" gate HEMT, having 200 µm of total length. The EM3DS was used to edit the 3D model, as shown in figure 2. Lumped inductances have been added at circuit level to gate, drain and source ports in order to account for bonding wires. Also an additional gate-to-drain capacitor was included, as the residual differential capacitor C_{gd} was not included into the model. Such a capacitor is usually very small when operating in the saturation mode, but is quite important to correctly predict S_{12} and stability characteristics. On the other hand the most important differential capacitor, C_{gs}, is basically accounted for by the passive gate-to-channel capacitance, as discussed for the 2D case in [4]. Comparison between measured and preliminary calculated results for S-parameters is reported in figure 3. The agreement is quite good, also considering uncertainties about structural parameters. In fact data were obtained partly from the manufacturer' data sheet and partly from on-line application notes. Anyway further investigation about sensitivity to the foundry parameters is in progress. The technique is also being applied to a number of additional structures (including MESFETs).

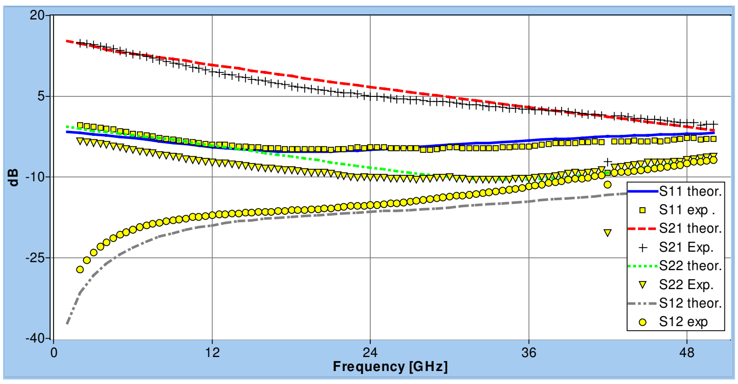

Fig. 3: Comparison between calculated and measured data for pHEMT LP7512.

IV. Conclusion

In this contribution we introduce a 3D full-wave approach for the modeling of circuits including active linear effects. The proposed technique accounts for the controlled source flowing in the channel and provides a self-consistent em modeling of microwave FETs. The approach is validated by comparing theoretical results to experimental data for the commercial low-noise pseudomorphic HEMT LP7512. Results are shown to be in good agreement in the whole 1-50GHz band.

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