EQUIVALENT-VOLTAGE DESCRIPTION OF LOW-FREQUENCY DISPERSIVE EFFECTS IN LARGE-SIGNAL FET MODELS

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ABSTRACT

The paper introduces a simple and efficient approach for the modelling of low-frequency dispersive phenomena in FETs. It is based on the definition of a virtual, non-dispersive associated device *controlled by* equivalent port voltages *and it is suitable for modelling based on standard nonlinear dynamic approaches, such as lumped-element equivalent circuits. The proposed approach is justified on the basis of a physically-consistent, charge-controlled description of the device, but the results are general and provide a valuable tool for taking into account dispersive effects in FETs by means of an intuitive circuit solution, in the framework of any existing nonlinear dynamic model of the associated non-dispersive device.*

The new equivalent-voltage description, identified on the basis of conventional measurements carried out under static and small-signal dynamic operating conditions, allows for the accurate prediction of dispersive effects above the frequency cut-off, but the formulation is still compatible, without formal modification, for the modelling of the device behaviour under signal excitations having spectral components in the dispersive low-frequency range. Preliminary results are presented which confirm the validity of the proposed approach.

INTRODUCTION

Accurate nonlinear modelling of FETs for microwave circuit design should also account for low-frequency dispersive phenomena of the electrical characteristics due to charge "trapping" and device self-heating. These phenomena cause considerable deviations between static and dynamic (e.g., pulsed) measurements of the I/V characteristics, or, if we think in terms of differential parameters, frequency dependent behaviour of the trans-admittance and output impedance even at low frequencies (e.g., lower than 1MHz). Since microwave large-signal performance prediction involves accurate modelling of both DC and AC components of the drain current, efforts have been made by different research groups to take into account low-frequency dispersion both in mathematical and equivalent circuit models [1-3].

The empirical modelling approach of dispersive phenomena presented in [4] provides very good predictive capabilities of bias-dependent dynamic drain current deviations due to traps and thermal effects in FETs. However, due to the need for non-conventional instrumentation for pulsed measurements in the identification phase, its exploitation could result sometimes impossible. The simplified "backgating"-like approach presented in [5] represents a viable alternative choice based on standard identification data, but a modification of the nonlinear model current source of the intrinsic device is still needed in order to take into account a dependence on both mean and instantaneous port voltages. Moreover, both approaches [4] and [5] neglect possible direct interactions between dispersive phenomena and high-frequency nonlinear device dynamics.

In the following a new approach is presented, which is based on the definition of an intrinsic non-dispersive device controlled by equivalent port voltages and suitable for modelling based on any standard nonlinear dynamic approach, without requiring any modification on model equations.

THE EQUIVALENT-VOLTAGE APPROACH

As well known, after de-embedding from linear extrinsic parasitics by means of available procedures (e.g., see [6]), the resulting intrinsic field effect device still is affected by low-frequency phenomena due to traps and self-heating. The following discussion shows how an intrinsic non-dispersive associated device can be defined when dispersive phenomena are separately taken into account by means of "extrinsic" series voltage sources (as shown in Fig.1a), which are linearly controlled by the voltages at the device ports.

Although dispersive effects could be dealt with in a comprehensive way by taking also device thermal phenomena into account, for the sake of simplicity the discussion presented in this paper is conducted by only considering the presence of trapping phenomena. In any case, the approach described in [5] can be adopted to account for low-frequency dispersion due to self-heating.

Let us consider first an ideal intrinsic field effect transistor, where no low-frequency dispersive phenomena take place, so that a purely algebraic nonlinear relationship can be assumed between charges and voltages. Such a device can be properly described by adopting the following charge-controlled quasi-static model formulation:

$$i(t) = \phi\{q(t)\} + \frac{dq(t)}{dt} \tag{1}$$

$$q(t) = \psi\{v(t)\} \tag{2}$$

where $i = [i_s \; i_d]^T$, $q = [q_{gs} \; q_{gd}]^T$, $v = [v_{gs} \; v_{gd}]^T$ represent the vectors of the source and drain currents, the gate-source and gate-drain equivalent charges, which are dealt with as state-variables, and the intrinsic port voltages respectively. Moreover, $\phi(\cdot) = [\phi_1(\cdot) \; \phi_2(\cdot)]^T$, $\psi(\cdot) = [\psi_1(\cdot) \; \psi_2(\cdot)]^T$ are suitable purely-algebraic nonlinear functions.

The presence of low-frequency dispersive effects due to traps in inter-electrode surface regions and in channel-substrate interface deep layers causes modifications in the charge-based state variables, introducing a dependence of charges on past voltages with relatively-long memory duration. In such conditions the stored charge $q(t)$ can not be correctly predicted by purely algebraic non-dispersive Eq.(2) since a charge perturbation $\Delta q(t)$ due to the slow dynamics of dispersive phenomena must also be taken into account. Thus the dispersive charge/voltage model equation becomes:

$$q(t) = \psi\{v(t)\} + \Delta q(t). \tag{3}$$

However, an equivalent result can be obtained by still using the non-dispersive Eq.(2) (i.e. the associated non-dispersive device model) provided that the actual port voltages $v(t)$ are replaced by equivalent port voltages $\widetilde{v}(t)$. These clearly must satisfy the equivalence condition:

$$\widetilde{v}(t) = \psi^{-1}\{q(t)\} = \psi^{-1}\left\{\psi\{v(t)\} + \Delta q(t)\right\} \doteq v(t) + \Delta v(t) \tag{4}$$

where dispersive effects are taken into account by the voltage perturbations $\Delta v(t) = \widetilde{v}(t) - v(t)$. It can be shown that the above defined non-dispersive associated device is not necessarily a device where traps are not present, but rather a device where trapped charges are "frozen" in a particular state.

This shows that any intrinsic field effect transistor affected by dispersive trapping phenomena and excited by port voltages v, can be described in terms of a virtual non-dispersive associated device excited by *equivalent port voltages* $\widetilde{v} = v + \Delta v$. When a suitable identification procedure exists for the Δv terms, the nonlinear modelling problem of a dispersive device is transformed into the modelling of the associated non-dispersive device (e.g., any nonlinear dynamic approach can be adopted such as, for example, lumped-element equivalent circuits). By adopting an equivalent voltage description for the associated non-dispersive device, we have:

$$i(t) = F\{\widetilde{v}(t)\} + C\{\widetilde{v}(t)\} \cdot \frac{d\widetilde{v}(t)}{dt} \tag{5}$$

where $F\{\widetilde{v}\} = \phi\{\psi\{\widetilde{v}\}\}$ and $C\{\widetilde{v}\} = \frac{d\psi(\widetilde{v})}{d\widetilde{v}}$ are purely-algebraic functions. Eq.(4) clearly corresponds to the intrinsic device circuit schematic shown in Fig.1a, where the Δv terms have been interpreted as series voltage-controlled voltage sources, yet to be identified. Once the two sources are known, the associated non-dispersive device can be obtained from a "deembedding-like" operation on the intrinsic device.

Since the dispersive phenomena due to traps in FETs, although by no means negligible, are usually not so strong to involve highly nonlinear effects, a linear dependence of the Δv terms on device port voltages is assumed in this work. The validity of such an assumption is also empirically confirmed by comparisons between model prediction and experimental data. The following frequency-domain vector relationship is therefore assumed:

$$\Delta V = A(\omega) \cdot V(\omega) \tag{6}$$

where ΔV, V are the Fourier-transforms of Δv, v respectively and A is a suitable matrix of transfer functions, whose frequency-dependence corresponds to the low-pass behaviour of dispersive phenomena. In particular, for RF operation at microwave frequencies, Eq.(6) becomes [1]:

$$\Delta V = A_0 \cdot V_0 \tag{7}$$

where $A_0 = A(0)$ and $V_0 = V(0)$ are the DC components of $A(\omega)$ and $V(\omega)$ respectively.

MODEL IDENTIFICATION

The circuit schematic in Fig.1b is considered, where the intrinsic device is shown in common-source configuration instead of common-gate as previously assumed. Although unneeded from a theoretical point of view, this description change is found more convenient for model identification. Simple linear algebraic relations allow

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[1] A common case in microwave circuit analysis and design is that every spectral component of the involved signals is above the cut-off frequency of trapping effects (a part from the DC value). On the other hand, the behaviour of the $A(\omega)$ functions could become important for signals with spectral components in the dispersive low-frequency range.

for the transformation of the voltage perturbations, expressed for the common-gate device by Eq.(7), into the corresponding terms valid for the common-source device. Moreover, it is assumed from now on that the vectors $F\{\cdot\}$ and $C\{\cdot\}$ of purely-algebraic functions describing the associated non-dispersive device are defined in the v_{GS}, v_{DS} domain. In other words, for the sake of notation simplicity, the two functions are not assigned a new symbol, even if they are, in the new voltage domain, formally different from those introduced for the common-gate device by Eq.(5).

Since all the dynamic drain current characteristics give i_D=0 for any v_{GS} when v_{DS}=0, the constraint: $A^S_{21_0}$=$A^S_{22_0}$=0 must be satisfied, $A^S_{ij_0}$ $(i, j = 1, 2)$ being the elements of the A_0 matrix in the common-source description. The circuit schematic in Fig.1b can be therefore simplified since no voltage perturbation has to be considered at the drain port when a common-source device description is adopted. Thus, circuit equations become: $\widetilde{v}_{GS}(t) = v_{GS}(t) + A^S_{11_0} \cdot V_{G0} + A^S_{12_0} \cdot V_{D0}$ and $\widetilde{v}_{DS}(t) = v_{DS}(t)$, where V_{G0}, V_{D0} are the mean values of $v_{GS}(t)$, $v_{DS}(t)$ and $A^S_{11_0}$, $A^S_{12_0}$ are the only two model parameters to be identified.

To this aim, let us now consider static device operation, such that: $v_{GS}(t)$=V_{G0} and $v_{DS}(t)$=V_{D0}. Thus:

$$i(t) = I_0 = F\{\widetilde{V}_{G0}, V_{D0}\} = F^{DC}\{V_{G0}, V_{D0}\} = \; F^{DC}\left\{\frac{\widetilde{V}_{G0} - A^S_{12_0} \cdot V_{D0}}{1 + A^S_{11_0}}, V_{D0}\right\} \tag{8}$$

where $F^{DC}\{\cdot\}$ are the DC current characteristics of the intrinsic device and $i = [i_G \; i_D]^T$. By differentiation of the drain current around a generic $\widehat{V}_{G0}$, $\widehat{V}_{D0}$ voltage pair, we obtain:

$$\left\{ \begin{array}{l} \widehat{g}_m^{DC} = \widehat{g}_m^{AC} \cdot (1 + A^S_{11_0}) \\ \widehat{g}_d^{DC} = \widehat{g}_m^{AC} \cdot A^S_{12_0} + \widehat{g}_d^{AC} \end{array} \right. \tag{9}$$

where $\widehat{g}_m^{DC}$, $\widehat{g}_d^{DC}$, $\widehat{g}_m^{AC}$, $\widehat{g}_d^{AC}$ are the static and low-frequency-dynamic device trans- and output-conductances respectively. Note that derivatives around a static condition of the drain current function in $F\{\cdot\}$, involved in Eqs.(9), coincide with the corresponding quantities under low-frequency dynamic regime, since the associated device is non-dispersive. Moreover, the dynamic conductances of the virtual non-dispersive associated device coincide with those of the real device since the voltage perturbations do not modify differential parameters.

Identification of the two model coefficients $A^S_{11_0}$, $A^S_{12_0}$ can be easily carried out by means of Eqs.(9). In fact, by considering a set of different bias conditions for the measurement of the differential parameters [2] in Eqs.(9), an overdetermined linear system is obtained to be solved for $A^S_{11_0}$ and $A^S_{12_0}$. Closed-form, analytical least-squares solution allows for robust model identification.

Once the $A^S_{11_0}$, $A^S_{12_0}$ parameters are known, Eq.(8) acts like identification formula for the virtual non-dispersive device current characteristics $F\{\cdot\}$ on the basis of the static measurements $F^{DC}\{\cdot\}$ carried out on the actual device [3]. Note that the proposed approach for the modelling of dispersive phenomena is fully compatible with any existing high-frequency dynamic nonlinear model for the virtual associated non-dispersive device, simply involving proper preliminary "deembedding-like" operation on intrinsic device characterisation data.

EXPERIMENTAL VALIDATION

The proposed equivalent-voltage approach has been applied for the modelling of low-frequency dispersive phenomena of different devices. In Fig.2 the comparison is presented for a $300\mu m$ MESFET between predicted dynamic drain current characteristics and on-wafer measurements obtained by applying short, simultaneous voltage pulses at the gate/drain electrodes starting from different quiescent conditions. Similar results are shown in Fig.3 for a packaged medium-power MESFET. Since quite important self-heating takes place into this device, as can be seen from DC characteristics, slightly better prediction accuracy has been obtained (not shown) by also including thermal effects modelling in a similar way as in [5].

Finally, the equivalent voltage approach has been tested in conjunction with a high-frequency nonlinear model for the virtual non-dispersive associated device, namely the Finite Memory Model (FMM) [3]. In particular, a $600\mu m$ PHEMT has been characterised on-wafer under static and small-signal dynamic conditions up to 50 GHz for FMM identification. Fig.4 shows the agreement between model predictions and measurements of the S-parameters for three different bias conditions as well as the harmonic distortion prediction at 5 GHz, confirming the accuracy of the proposed approach also under nonlinear high-frequency operation.

CONCLUSION

A new approach for the modelling of low-frequency dispersive phenomena in FETs has been presented. The model is based on the definition of a non-dispersive associated device, which is controlled by equivalent port

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[2] Low-frequency dynamic conductances can be obtained from measured S-parameters after conversion to Y-parameters, so that g_m^{AC}=$Real\{Y_{21}\}$ and g_d^{AC}=$Real\{Y_{22}\}$.

[3] After parasitic network resistances de-embedding.

voltages and can be identified on the basis of conventional DC and small-signal S-parameter measurements, providing accurate predictions of bias-dependent, dynamic current characteristics. The associated non-dispersive device is suitable for modelling based on conventional nonlinear dynamic approaches in order to take into account also high-frequency junction charge-storage phenomena. Experimental validation under pulsed and RF large-signal operation confirms the accuracy of the proposed approach.

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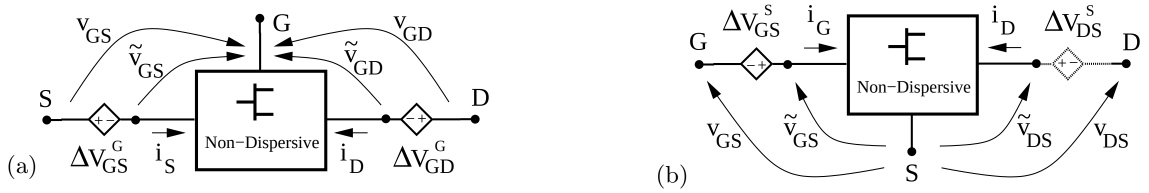

Fig.1 - Intrinsic device circuit schematics describing the equivalent-voltage approach for dispersive effects modelling. The "G" or "S" labels into the ΔV sources denote the common-gate (a) or common-source (b) configuration. The ΔV_{DS}^S voltage-controlled voltage source (shaded in figure) can be omitted as described in the paper.

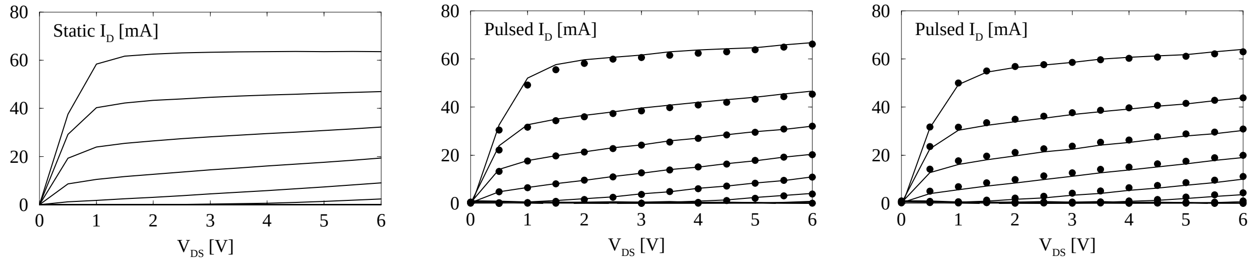

Fig.2 - Drain current characteristics of an "on-wafer" 300 μm MESFET. From left to right: DC, Pulsed ($V_{G_0}=-0.8V$; $V_{D_0}=4V$), Pulsed ($V_{G_0}=0V$; $V_{D_0}=6V$). Measurements (•) versus predictions (—).

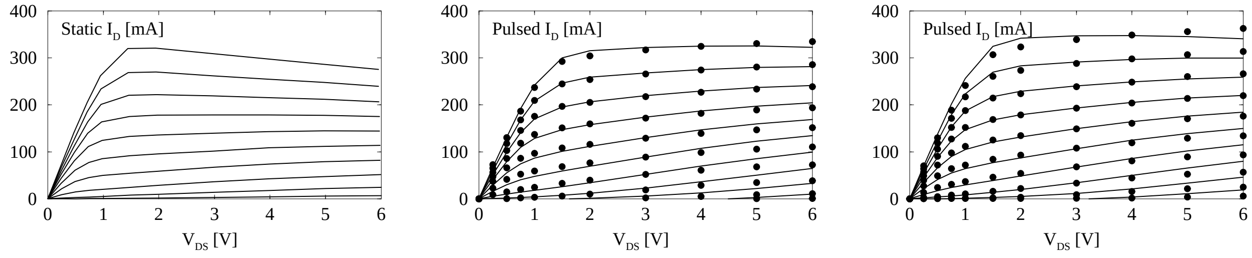

Fig.3 - Drain current characteristics of a packaged medium-power MESFET. From left to right: DC, Pulsed ($V_{G_0}=-1V$; $V_{D_0}=3V$), Pulsed ($V_{G_0}=-3V$; $V_{D_0}=3V$). Measurements (•) versus predictions (—).

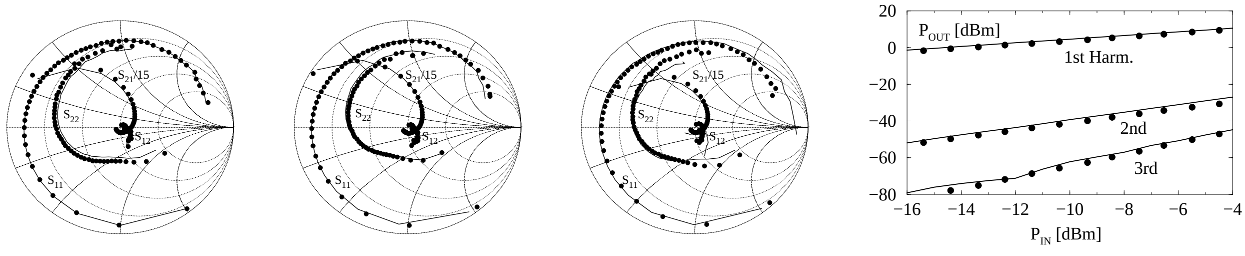

Fig.4 - From left to right: S-parameters (1-50GHz) of an "on-wafer" 600μm PHEMT at three (V_{G_0},V_{D_0}) biases: 1:(-0.55V,6V), 2:(-0.25V,3V), 3:(-0.7V,4V). Harmonic distortion in a 50Ω-loaded power amplifier (Bias:1,f_0=5GHz). Measurements (•) versus predictions (—) by means of a nonlinear dynamic model [3] and the proposed approach.