A Globally-Continuous, Charge-Conservative, Non-linear Equivalent Circuit Model For RF MOSFETs

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***Abstract*— A non-linear equivalent circuit model for MOSFETs valid for DC, small and large-signal simulations of high frequency circuit design is presented. The model is valid for a wide range of bias conditions and is globally continuous. Capacitances are derived from a single charge model and charge conservation is taken into account. Simulations of the model, following parameter extraction, are validated by comparisons with experimental data.**

I. INTRODUCTION

MOSFET gate lengths and device sizes have continuously decreased presenting us with more and more interesting high frequency performances. There continues to be an urgent need for accurate, compact large signal models for MOSFETs which can be used for computer aided design of monolithic integrated non-linear circuits operating in the millimeter-wave ranges.

Several non-linear equivalent circuit models for MESFETs and HEMTs have been developed for high frequency application [1][2], but not so much for MOSFETs. For large signal models it is important to include charge conservation to improve the model consistency and the convergence properties of nonlinear simulation [3], [4], [5]. Existing industrial RF MOSFET models include the BSIM series [6], [7]. These models require a very large number of parameters however, and to improve accuracy, they rely on smoothing functions and may require parameter values that are not often available from chip manufacturers.

The aim of this paper is to present a non-linear equivalent cicuit model for MOSFETs which is charge conservative, easy to extract and contains simple equations to represent the voltage dependent elements of the MOSFET model very accurately. This model describes accurately the electrical characteristics and is fully consistent between DC, small signal simulations and large signal nonlinear equations. It includes a non-linear drain current expression that is continuous and infinitely derivable, leading to a continuous transconductance and drain conductance. The capacitances are derived from a new gate charge expression that ensures charge conservation. The model parameters were extracted for a MOSFET ($L_g = 0.18\,\mu m$, $W_g = 200\,\mu m$) and verified with both small signal and large signal measurements at different bias conditions.

II. THE NONLINEAR DRAIN CURRENT MODEL

In Fig. 1 the equivalent circuit of the large signal model for an RF MOSFET is depicted. The source and substrate have been connected together as is common [8], [7].The current generator I_{ds} and the bias dependent capacitances C_{gs} and C_{gd} are considered the nonlinear elements in the model.

Bias dependent S-parameter measurements and dc-measurements from a MOSFET ($L_g = 0.18\,\mu m$, $W_g = 200\,\mu m$) were used for the extraction of the model parameters. The intrinsic elements were calculated after de-embedding of the parasitics from measured S-parameters.

Equation (1) gives the drain current equation implemented in the model for RF MOSFET devices. Equation (1) is based on the COBRA model [1] developed for MESFET and PHEMT devices.

$$
\begin{gathered}
I_{ds,DC}(V_{gs}, V_{ds}) = \\
\beta \cdot V_{eff}^{\frac{\lambda}{1+\mu \cdot V_{ds}^2 + \xi \cdot V_{eff}}} \cdot \tanh\left[\alpha \cdot V_{ds}\left(1 + \zeta \cdot V_{eff}\right)\right] \\
V_{eff} = \frac{1}{2}\left(V_{gst} + \sqrt{V_{gst}^2 + \delta^2}\right) \\
V_{gst} = V_{gs} - (1 - \beta_r^2) \cdot V_{TO} + \gamma \cdot V_{ds}
\end{gathered}
\tag{1}
$$

In (1) V_{TO} is the pinchoff voltage and α, β, β_r, γ,

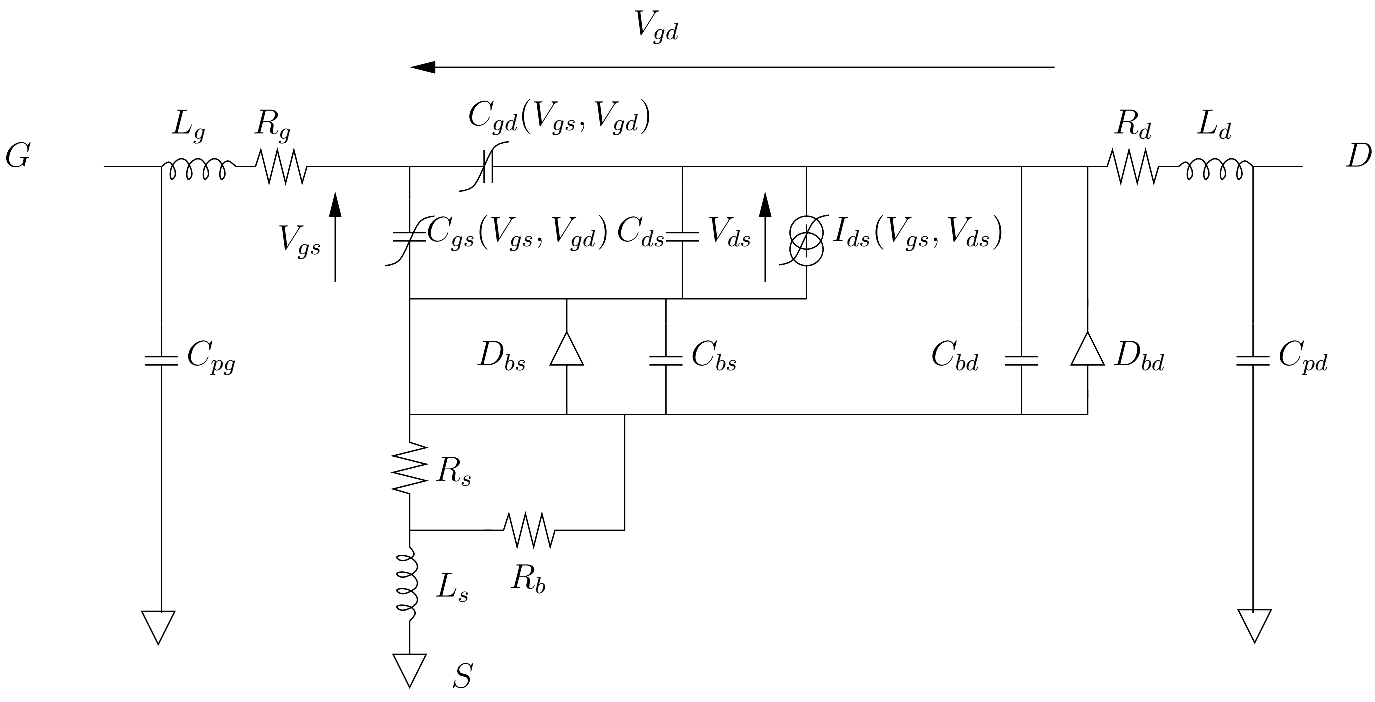

Fig. 1. Large Signal equivalent circuit model of the RF MOSFET

δ, λ, μ, ξ, ζ are model parameters and β_r is a dimensionless parameter, numerically equal to β (when I_{ds} is expressed in Amperes). The model is continuous over the entire bias plane and its derivatives are continuous, which is important for a good representation of the intermodulation characteristics.

The model transconductance's and conductance's equations are given by (2).

$$g_{mi}(V_{gsi}, V_{dsi}) = \frac{dI_{ds}}{dV_{gsi}} \qquad g_{di}(V_{gsi}, V_{dsi}) = \frac{dI_{ds}}{dV_{dsi}}_{|V_{gsi}=cte} \tag{2}$$

III. The nonlinear capacitance model

The best way to model capacitance and to maintain charge conservation is to consider the total charge at a given physical point of the device [9]. The total gate charge depends on the gate to source bias and the gate to drain bias : $Q(V_{gs}, V_{gd})$. The capacitances $C_{gs}$and C_{gd} will be derived using equation (3) [10].

$$C_{gs} = \left.\frac{\partial Q}{\partial V_{gs}}\right|_{V_{gd}} \qquad C_{gd} = \left.\frac{\partial Q}{\partial V_{gd}}\right|_{V_{gs}} \tag{3}$$

The capacitances C_{gs} and C_{gd} are derived from the same charge expression and are interdependent. Using a single charge function from which the expressions for C_{gs} and C_{gd} are calculated, the model accounts for charge conservation if [5]:

$$\frac{\partial C_{gd}}{\partial V_{gs}} = \frac{\partial^2 Q}{\partial V_{gs}\partial V_{gd}} = \frac{\partial^2 Q}{\partial V_{gd}\partial V_{gs}} = \frac{\partial C_{gs}}{\partial V_{gd}}$$

Following these principles the gate charge for the MOSFET transistor was modelled by (4).

In (4), c_p, c_{g1}, c_{g2}, c_{g3}, m, κ, ψ, V_{T1} are the model parameters. (3) and (4) give the explicit functions for C_{gs} and C_{gd}.

IV. Simulation Results

We present here the results obtained for a MOSFET with $W_g = 200\,\mu m$ and $L_g = 0.18\,\mu m$. The parameters were extracted using an optimisation procedure by fitting the model equations on the measured DC and AC data. The intrinsic and extrinsic elements were deduced from bias dependent S-parameter measurements [11].

The model was implemented in the Agilent Technologies ADS circuit simulator. Small signal and large signal simulations were carried out in order to perform some comparisons with experimental data and validate the model.

We observe an excellent agreement in Fig. 2 of the static simulation for the drain current versus V_{ds} for different values of V_{gs}.

In Fig. 3, the small signal simulations are compared to the measured ones. The four S-parameters are plotted for frequencies varying from $0.5\,GHz$ to $10\,GHz$ and the agreement is quite good.

$$
\begin{aligned}
Q(V_{gs},V_{gd}) \;=\;& c_p\cdot(V_{gs}+V_{gd})+\frac{c_{g1}\cdot V_{bi}}{1-m}\left(\left(1-\left(1-\frac{V_{gs}}{V_{bi}}\right)^{1-m}\right)+\left(1-\left(1-\frac{V_{gd}}{V_{bi}}\right)^{1-m}\right)\right)\\
&+c_{g2}\cdot\left(V_{gs}+V_{gd}+\frac{1}{\kappa}\cdot\ln\left(\frac{\cosh(\kappa\cdot(V_{gs}-V_{T1}))}{\cosh(\kappa\cdot V_{T1})}\right)+\frac{1}{\kappa}\cdot\ln\left(\frac{\cosh(\kappa\cdot(V_{gd}-V_{T1}))}{\cosh(\kappa\cdot V_{T1})}\right)\right)\\
&+\frac{1}{\psi}\cdot c_{g3}\cdot V_{gs}\cdot\ln\left(\cosh\left(\psi\cdot(V_{gs}-V_{gd})\right)\right)
\end{aligned}
\tag{4}
$$

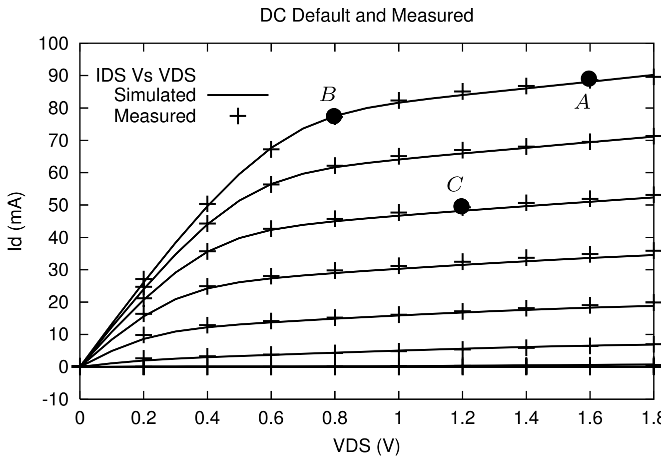

Fig. 2. DC measurements and simulation results ($VGS = 1.6\,V$ down to $0\,V$ in steps of $0.2\,V$)

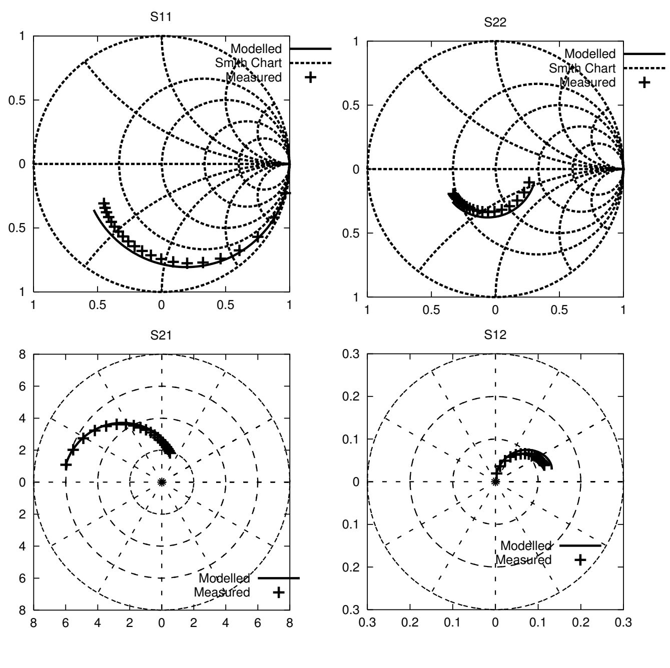

Fig. 3. S-parameter measurements and simulation results ($VGS = 1.6\,V$ and $VGS = 1.6\,V$, point A in fig. 2)

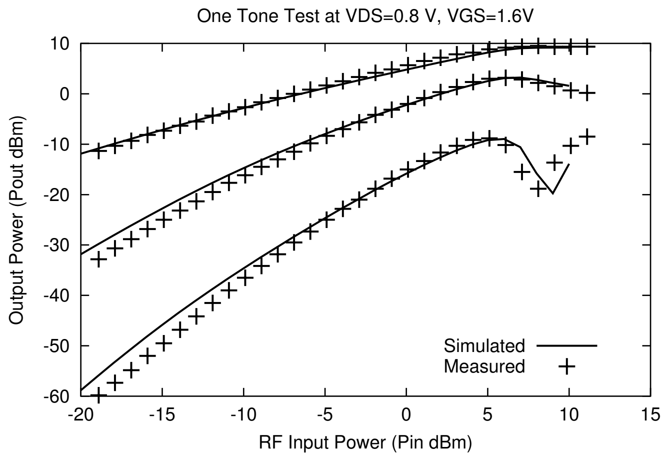

Fig. 4. Large Signal measurements and simulation results. $1st$, $2nd$ and $3rd$ harmonics of a one-tone test at $VGS = 0.8\,V$ and $VDS = 1.6\,V$ (point B in Fig. 2)

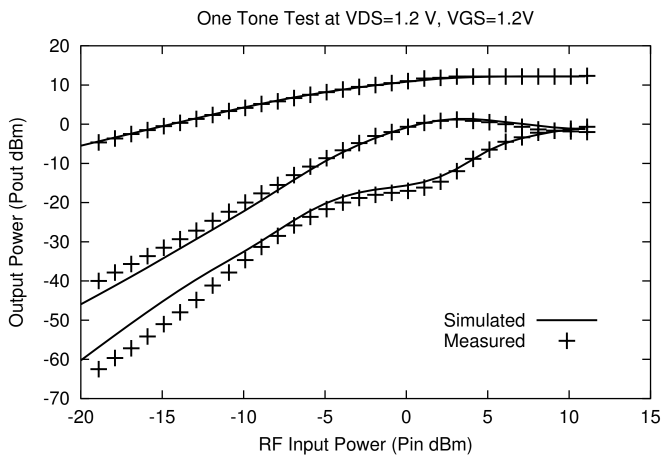

Fig. 5. Large Signal (one-tone test) measurements and simulation results. $1st$, $2nd$ and $3rd$ harmonics of a one-tone test at $VGS = 1.2\,V$ and $VDS = 1.2\,V$ (point C in Fig. 2)

In order to validate the model, single tone large signal measurements and two tone intermodulation tests were carried out. The single tone large signal measurements were conducted with a microwave power P_{in} applied at the input of the device using a $50\,\Omega$ generator. The 1^{st} to 3^{rd} harmonics of the output power P_{out} dissipated by the $50\,\Omega$ output load were measured. The results are compared to simulations at two different bias points in Figs. 4 and 5.

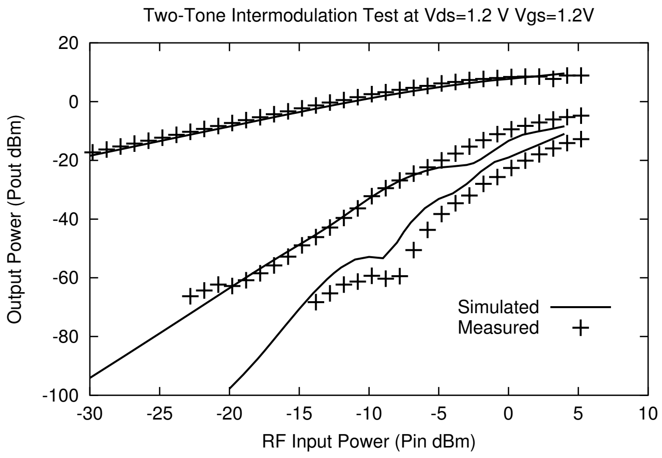

Fig. 6. 1^{st}, 3^{rd} and 5^{th} intermodulation distortion products of a two-tone test at VGS=1.2 V and VDS=1.2 V (point C in Fig. 2)

For the two tone intermodulation measurements, we have applied at the input of the device a two tone signal of microwave power P_{in} centred at $2\,GHz$ with a tone spacing of $4\,MHz$ using a $50\,\Omega$ generator and measured the intermodulation distortion products at the output load of $50\,\Omega$. The results are shown in Fig. 6. A good agreement has been observed with both the single tone and two-tone simulations carried out using ADS.

V. Conclusions

In this paper a compact non-linear equivalent circuit model for RF MOSFETs has been presented. This model is based both upon closed form expressions of the non-linear drain current and a non-linear gate charge. The parameters of the model can be easily extracted using optimisation routines and the comparisons in terms of static and dynamic performances are in good agreement with the experimental ones. The non-linear drain current expression is globally continuous and infinitely derivable. The model is charge conservative with the capacitances derived from a single gate charge function. Finally, single tone large signal and two-tone intermodulation measurements have been made and compared with simulations.

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