Short Papers

Fast and Accurate Extraction of Capacitance Parameters for the Statz MESFET Model

Sven Van den Bosch and L. Martens

***Abstract*— A modified approach to S-parameter fitting of measured MESFET data to the Statz's model equations is presented. The technique only requires one sweep of measured S-parameters versus frequency at a single bias point. In addition, there is a reduced need for optimization during the extraction procedure. The additional information in the data-fitting code reduces problems with measurement noise and calibration error. The presented method was developed and tested in HP-ICCAP, a software tool for parameter extraction. All measurements are controlled by this software.**

***Index Terms*—Capacitance, MESFET's, modeling.**

I. Introduction

In order to design, develop, and produce high-yield low-cost microwave circuits, adequate models of all components are required. One of the first models used for nonlinear GaAs MESFET's was proposed by Curtice *et al.* [1]. The capacitance–voltage relations were, to a first degree of approximation, assumed to be independent of the current–voltage (*C–V*) relation, and were modeled by a simple diffusion capacitance. Later, Statz *et al.* [2] proposed modified equations resulting in a more accurate prediction of measured S-parameters, valid in a broader area of MESFET operation. These equations also conserve charge at the device terminals [3], [4]. Even now, a lot of work is still done on nonlinear capacitance MESFET modeling [5]. Performing accurate circuit simulations not only requires accurate device models, but also accurate extraction of all model parameters.

The generally used method to determine capacitance parameters is described in Section II-A. It is similar to the approach of [6] but uses S-parameters instead of *C–V* measurements. The method in [6] requires S-parameters measured over a range of bias points and yields large and hard-to-manage data sets.

The new method eliminates one of three capacitance parameters by applying a well-selected bias voltage and reduces the mutual influence in the extraction of the remaining two parameters. It is presented in Section II-B. The method fits measured data to a predefined functional form based on information available in the Statz model, instead of fitting it as a function of bias voltage. This reduces the problems associated with measurement noise and calibration errors. Section III applies both the new and the old approach on a 6 × 100 μm GaAs MESFET. The conclusion repeats the most important results.

II. Theory

Fig. 1 shows the nonlinear Statz MESFET model [2], with added parasitic resistances and inductances, as it is available in most commercial circuit simulators. The model has three dominant nonlinearities: the transconductance $I_{\rm ds}$, the gate–source capacitance $C_{\rm gs}$ and the gate–drain capacitance $C_{\rm gd}$. These are expressed in terms of the intrinsic gate–source and gate–drain voltages, $V_{\rm gs}$ and $V_{\rm ds}$. This paper focuses on $C_{\rm gs}$ and $C_{\rm gd}$. The S-parameters versus the intrinsic voltages are obtained by de-embedding the measured S-parameters from knowledge of device parasitics (R_i and L_i with $i = g, d, s$). In the Statz MESFET model, the dependence of $C_{\rm gs}$ and $C_{\rm gd}$ on the intrinsic terminal voltages [2] is given by (1)

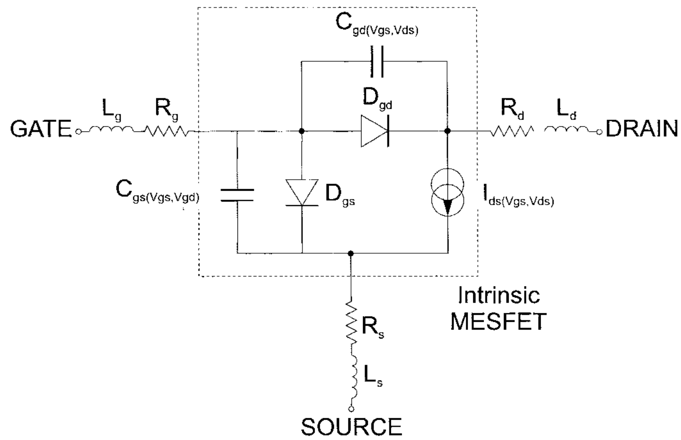

Fig. 1. The nonlinear Statz MESFET model.

$$C_{\rm gs} = \frac{C_{\rm gs0}}{\sqrt{1-\dfrac{V_{\rm new}}{V_{\rm bi}}}}\,\frac{1}{4}\,(1+K_1)(1+K_2) + \frac{1}{2}\,C_{\rm gd0}(1-K_2)$$
$$C_{\rm gd} = \frac{C_{\rm gs0}}{\sqrt{1-\dfrac{V_{\rm new}}{V_{\rm bi}}}}\,\frac{1}{4}\,(1+K_1)(1-K_2) + \frac{1}{2}\,C_{\rm gd0}(1+K_2) \tag{1}$$

where

$$V_{\rm new} = \tfrac{1}{2}\cdot\left[V_{\rm eff1} + V_T + \sqrt{(V_{\rm eff1}-V_T)^2+\delta^2}\right]$$
$$V_{\rm eff1} = \tfrac{1}{2}\cdot\left[V_{\rm gs} + V_{\rm gd} + \sqrt{(V_{\rm gs}-V_{\rm gd})^2+\alpha^{-2}}\right] \tag{2}$$

and

$$K_1 = \frac{V_{\rm eff1}-V_T}{\sqrt{(V_{\rm eff1}-V_T)^2+\delta^2}}$$
$$K_2 = \frac{V_{\rm gs}-V_{\rm gd}}{\sqrt{(V_{\rm gs}-V_{\rm gd})^2+\alpha^{-2}}}. \tag{3}$$

V_T is the threshold voltage, $C_{\rm gs0}$ and $C_{\rm gd0}$ are the zero-bias gate–source and gate–drain capacitances and α the saturation voltage. The δ-parameter models the behavior of $C_{\rm gs}$ and $C_{\rm gd}$ around and below pinchoff. Its default value is 0.2 [2].

A. Generally Used Method

The generally used extraction routine uses the de-embedded Y-parameters at one bias point to extract capacitance values at this bias point [7]

$$C_{\rm gs} = {\rm slope}[{\rm im}\,(Y_{11})] - C_{\rm gd}$$
$$C_{\rm gd} = -\,{\rm slope}[{\rm im}\,(Y_{12})] \tag{4}$$

Manuscript received March 14, 1996; revised April 25, 1997. The work of S. Van den Bosch was supported by the Vlaams Instituut voor de Bevordering van het Wetenschappelijk-Technologisch onderzoek in de industrie.

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Publisher Item Identifier S 0018-9480(97)05381-7.

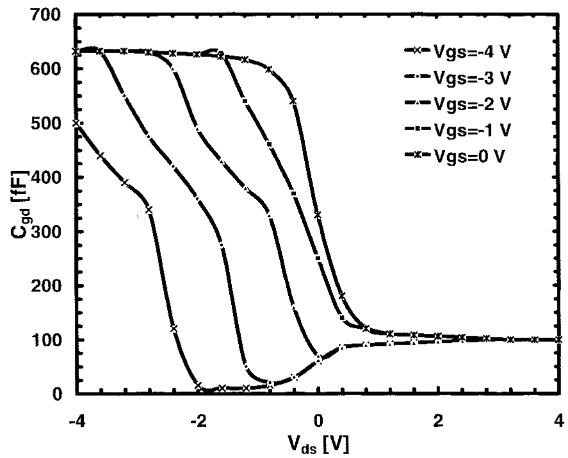

Fig. 2. Simulations of $C_{\rm gs}$ versus $V_{\rm ds}$ for different values of $V_{\rm gs}$; $C_{\rm gs0} = 500$ fF and $C_{\rm gd0} = 100$ fF.

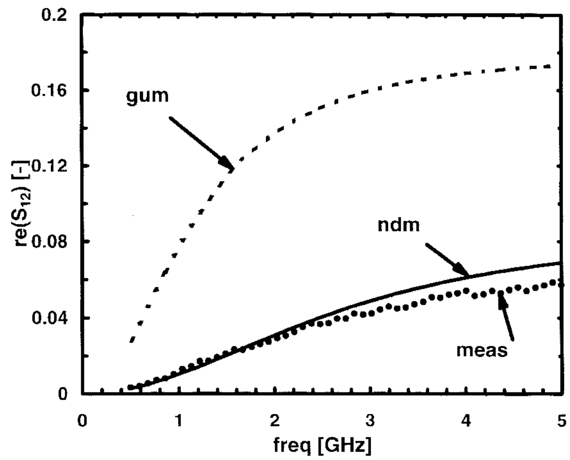

Fig. 4. Measured (meas) $\mathrm{re}(S_{12})$ versus simulations using the generally used method (gum) and the newly developed method (ndm).

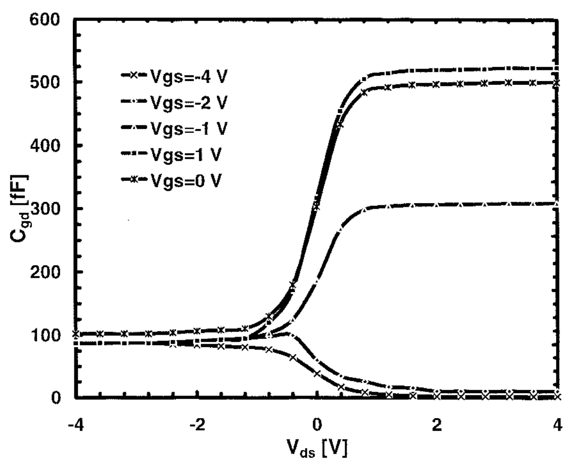

Fig. 3. Simulations of $C_{\rm gd}$ versus $V_{\rm ds}$ for different values of $V_{\rm gs}$; $C_{\rm gs0} = 500$ fF and $C_{\rm gd0} = 100$ fF.

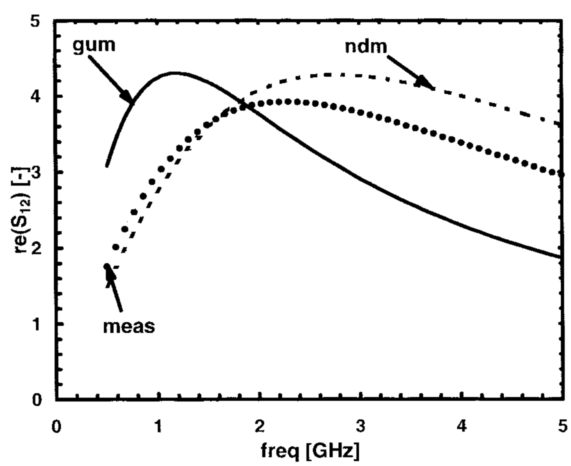

Fig. 5. Measured (meas) $\mathrm{im}(S_{21})$ versus simulations using the generally used method (gum) and the newly developed method (ndm).

where *slope* indicates the slope of the parameter versus frequency. Setting $V_{\rm gs}$ equal to $V_{\rm gd}(K_2 = 0)$ in (2) and neglecting $\delta^2(V_{\rm new} = V_{\rm eff1}$ and $K_1 = 1)$ in (2) and (3) simplifies (1) to

$$C_{\rm gs} = \frac{C_{\rm gs0}}{\sqrt{1 - \dfrac{V_{\rm gs} + \dfrac{1}{2 \cdot \alpha}}{V_{bi}}}} \cdot \frac{1}{2} + \frac{1}{2} \cdot C_{\rm gd0}. \tag{5}$$

By performing the Y-parameter measurements at several bias points, (5) yields a number of values for $C_{\rm gs}$. When the built-in voltage $V_{\rm bi}$ is known, plotting these values of $C_{\rm gs}$ versus $\left[\sqrt{1 - (V_{\rm gs} + \frac{1}{2} \cdot \alpha)/V_{\rm bi}}\right]^{-1}$ allows for linear regression, yielding $C_{\rm gs0}$ as slope and $C_{\rm gd0}$ as intercept. When linear regression is performed while sweeping $V_{\rm bi}$, the solution with the highest regression coefficient yields the best values for $C_{\rm gs0}$ and $C_{\rm gd0}$, since a regression coefficient close to one indicates highly linear data. In order to be sufficiently accurate, the generally used method requires that a significant number of bias points are considered, thus yielding a slow and error-prone routine. The authors, therefore, developed and implemented in HP-ICCAP,[1] a new method, which considerably simplifies and speeds up the extraction procedure.

[1] HP85190 Series High-Frequency IC-CAP Software User's Manual, Hewlett-Packard, 1995.

B. Newly Developed Method

The key features of the newly developed method include a reduction of the number of parameters to be simultaneously extracted and a further simplification of the capacitance equations. The most important simplification of the new procedure is achieved by eliminating $V_{\rm bi}$ from the equations. From (1), it is clear that this can only be done by making $V_{\rm new}$ equal to zero. Substituting $V_{\rm new} = 0$ in (2), one obtains (6) for $V_{\rm eff1}$

$$\begin{aligned} V_{\rm new} &= 0 \\ \Leftrightarrow (V_{\rm eff1} + V_T)^2 &= (V_{\rm eff1} - V_T)^2 + \delta^2 \\ \Leftrightarrow V_{\rm eff1} &= \frac{\delta^2}{4 \cdot V_T} \\ &= \frac{0.01}{V_T}, \qquad \delta = 0.2. \end{aligned} \tag{6}$$

Using (2), with α large, and (6), one finally obtains the desired gate–bias voltage $V_{\rm gs}$

$$\begin{aligned} V_{\rm eff1} &\cong V_{\rm gs} \qquad (\alpha \text{ large}) \\ \Rightarrow V_{\rm gs} &\cong \frac{0.01}{V_T}. \end{aligned} \tag{7}$$

On the other hand, one finds from (2) that for small δ, $V_{\rm new} = V_{\rm eff1}$. Because $V_{\rm new}$ was assumed to be zero, $V_{\rm eff1}$, and thus, $V_{\rm gs}$, must be small. With the value of $V_{\rm gs}$ found in (7), this condition is

TABLE I
CAPACITANCE PARAMETERS EXTRACTED WITH BOTH APPROACHES UNDER CONSIDERATION

Parameter	Generally used method	Newly developped method
C_{gs0}	682.6 fF	621.4 fF
C_{gd0}	33.4 fF	32.1 fF
V_{bi}	1.92 V	0.921 V

satisfied. Setting a large gate–drain voltage guarantees that the term $(V_{\rm gs} - V_{\rm gd})^2$ in (2) is large compared to α^{-2}, and the assumption of a large α is valid. Under these conditions, $V_{\rm new}$ is negligible in (1) and $K_1 = 1$ and $K_2 = 1$. The capacitance equations then reduce to (8)

$$C_{\rm gs} = C_{\rm gs0}$$
$$C_{\rm gd} = C_{\rm gd0}. \tag{8}$$

Evidence for (8) can also be found on simulated characteristics (Figs. 2 and 3, since $V_{\rm gs}$ is nearly zero, large $(V_{\rm gs} - V_{\rm gd})^2$ is equivalent to large $V_{\rm ds}$). It has been demonstrated that the parameters wanted to be extracted can be calculated easily from measured S-parameters for one particular bias point. All necessary optimizations need only be performed for a single bias point. In addition, the influence of the (earlier extracted) α-parameter and $V_{\rm bi}$ were eliminated. The value of $V_{\rm bi}$ can be easily determined from measurements at one additional bias point.

III. MEASUREMENTS

The modeled parameter extraction was first tested on modeled data. In that case, both approaches under consideration work almost equally well. However, when the extraction was performed on a real device, an important discrepancy between measured S-parameters and those simulated with parameters extracted using the generally used routine (Figs. 4 and 5) was observed. The device used was a 6 $\times$ 100 μm GaAs MESFET with a 0.7-μm gate length. The DC-parameters [7] and parasitics [8], [9] were extracted with established extraction procedures. The extracted capacitance parameters are listed in Table I. To accentuate the discrepancy, a comparison is shown (in Figs. 4 and 5) between measurement and simulation of the real part of S_{12} where the new method performs best, and the imaginary part of S_{11}, where it performs worst (but still better than the generally used method). Figs. 4 and 5 also indicate two striking features of simulation results obtained with the new method. First, a significant reduction in the error magnitude is achieved and also a better correspondence between the shapes of measured and simulated curves is obtained. The reason for these large differences for the measured results, in comparison with the extraction on modeled data, lies in the contribution of errors introduced by the influence of $V_{\rm bi}$ on the extraction procedure. It is clear that this new method is far less sensitive to these errors.

IV. CONCLUSION

A new, fast, and reliable method for the extraction of the nonlinear Statz-MESFET-model capacitance parameters was developed. The method gives good simulation results, both on real and modeled data. This method is far more robust and insensitive to measurement errors than the formerly used one because the number of parameters to be simultaneously extracted was reduced from three to two, which eliminated the influence of earlier extracted parameters such as α. The mutual influence of $C_{\rm gs0}$ and $C_{\rm gd0}$ was also eliminated.

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A Novel Ultrawideband Microwave Differential Phase Shifter

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***Abstract*—Wideband microwave phase shifters are usually constructed using cascaded coupled-line sections connected together at their far ends. By utilizing the unique frequency-independent quadrature property of symmetric couplers, a new class of phase shifter is proposed. The use of ultrawideband couplers in its realization results in a device which offers greater freedom with respect to bandwidth and ripple. The tendency of symmetrical networks to cancel small errors due to manufacturing tolerances is also exploited in order to improve its high-frequency performance.**

***Index Terms*— Microwave phase shifter, stripline components, wideband.**

I. INTRODUCTION

Differential phase shifters find wide application in microwave systems, for example, in hybrid circuits and wideband phased-array antennas. They are three-port or four-port networks providing a

Manuscript received August 23, 1996; revised April 25, 1997.
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Publisher Item Identifier S 0018-9480(97)05382-9.