

# A nonlinear empirical I/V model for GaAs and GaN FETs suitable to design power amplifiers

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## Abstract

This article presents an improved nonlinear empirical I/V model suitable for GaAs and GaN FETs. The new drain-to-source current formulation accurately represents the symmetric and the asymmetric bell-shaped transconductance ( $g_m$ ) for all  $V_{DS}$  values. Besides modeling with high accuracy the I/V characteristic, the proposed model can fit the first and second derivatives of the transconductance, from the ohmic to the saturation region, including the pinch-off region. The high correlation between experimental and simulated data of the I/V curves of GaAs and GaN FETs, ACPR, load-pull, and a class-J power amplifier designed in S-band corroborates the usefulness of the proposed model, considering only the static I/V model as the main nonlinear element of the electrical equivalent circuit model of the transistor.

## KEY WORDS

empirical I/V model, MET model, microwave FET model, nonlinear model

## 1 | INTRODUCTION

In the design of microwave circuits based on field-effect transistors (FETs), accurate nonlinear models are required. The nonlinear FET model consists of static or pulsed I/V characteristics and the Q/V charge based model. I/V models are used to predict not only the bias dependent current source ( $I_{DS}$  [ $V_{GS}, V_{DS}$ ]), but also its derivatives: output conductance ( $g_{ds}$ ) and transconductance ( $g_m$ ). Besides, the high-order derivatives of  $g_m$  are related to the harmonics and intermodulation analysis.

In the nonlinear modeling of FETs using electrical equivalent circuits (EECs),  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) is the main nonlinear distortion generator.<sup>1</sup> Hence, it is the most crucial nonlinear element to model. The  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ )

nonlinearity is mostly related to strong nonlinearities in the ohmic to saturation transition, pinch-off, and forward regions, besides the intrinsic nonlinearities like the output conductance and transconductance bias dependence.<sup>1,2</sup>

The basis for modeling the nonlinear current source  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) is through a static model.<sup>1</sup> Then thermal and trapping effects are modeled introducing the temperature dependence in the static model, using sub-circuits that model the traps' capture and emission time. Finally, the capacitance/charge model is used to consider the dynamic behavior of the FET. Nevertheless, a complete nonlinear FET model with all the features mentioned above is less accurate if there is not present an accurate  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) nonlinear model.

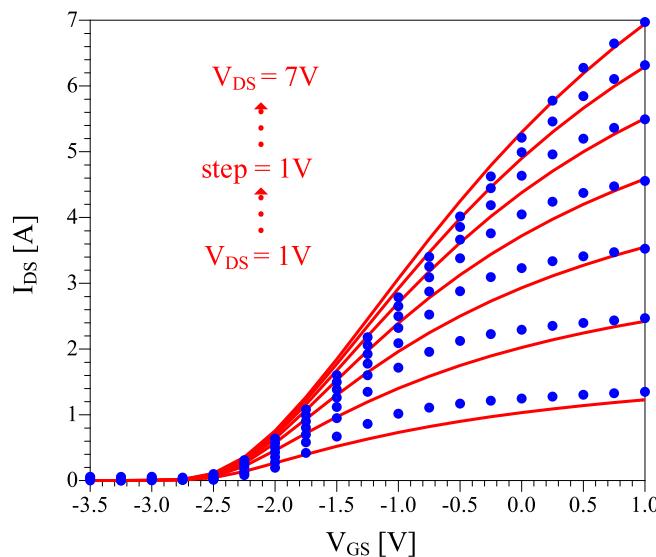
Several semi-empirical analytical  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) modeling techniques for GaN and GaAs HEMT devices have

been developed,<sup>3–18</sup> for example, table-based Artificial Neural Networks (ANNs), or analytical modeling techniques. The table-based model represents each element to be modeled with a lookup table developed from measured data. Although it is easy to implement, this modeling technique cannot predict out-of-range measurements and has convergence problems when dealing with nonlinearities. The ANN modeling technique has higher accuracy even under strong nonlinear operating conditions; however, it has limited prediction capability. The analytical modeling is more efficient in terms of convergence and prediction capability; however, it requires an effort to improve its performance in all regions of the I/V curves.

In that sense, several empirical I/V models for microwave FETs, based on analytical expressions, have been developed. Two of the most accepted analytic formulations are the Angelov-Zirath model<sup>3</sup> and the industry-standard MET<sup>19</sup> model. Both models were modified to make them more suitable for describing transconductance behavior and its dependence on  $V_{DS}$ . In the case of the model from Reference 3, one of its characteristics is the accurate modeling of the peak of the transconductance<sup>4,11,15,16,20</sup> in MESFET and GaN devices. On the other hand, the most notable modification of Reference 19 is the work from Fager<sup>21</sup> to model the  $I_{DS}$  characteristic of LDMOS devices.

With fewer fitting parameters than the other mentioned models, Reference 21 has successfully modeled the  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) characteristic of GaN transistors,<sup>5–8,10</sup> achieving good performance in representing the transconductance when biasing the transistor in the saturation region. Particularly for GaN devices, all the models mentioned above have a common factor: they are suitable for modeling the highly asymmetric bell-shaped transconductance that appears when the FET is in the saturation region ( $V_{DS} > V_{knee}$ ). However, none of them deals with the quasi-symmetrical bell-shaped form, which is present when the device is biased with a  $V_{DS}$  in the ohmic region.<sup>22</sup> As an example of this, Figure 1 shows a comparison of measured and simulated  $I_{DS}$  –  $V_{GS}$  data for a HEMT GaN CGH40045F, where the simulation, based on the I/V model of Reference 21, shows a low correlation with measured data in the ohmic region, being this more evident for values of  $V_{GS}$  close to the threshold voltage.

In this article, using the  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) formulation from Reference 21, we develop an improved nonlinear empirical I/V model capable of modeling the transconductance characteristic and its dependence with  $V_{DS}$  of HEMT GaAs and high power GaN devices. To achieve the above, we have included two functions from Reference 4 and introduced a new function that helps to model the transconductance accurately for low and high values of  $V_{DS}$ . Using pulsed I/V measurements of a packaged



**FIGURE 1** Comparison of measured (circle) and simulated data using the model from Reference 21 (solid line), for a packaged GaN HEMT CGH40045F at a  $V_{GSq} = -2.5$  V and  $V_{DSq} = 28$  V

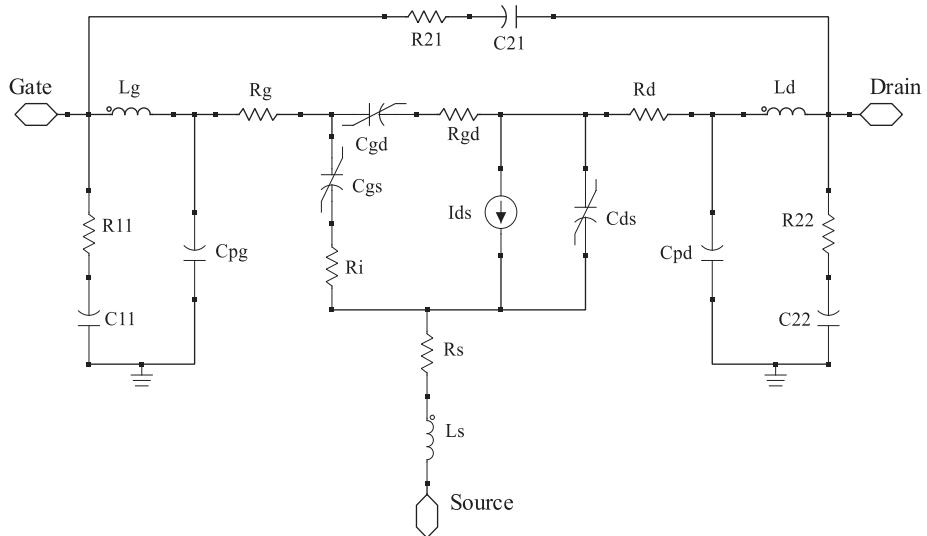
GaN HEMT CGH40045F and a GaAs FET ATF-34143, the proposed model was validated. The experimental results show an excellent correlation between the model and the measurements in all regions of the microwave FETs' I/V characteristic curve. Additionally, the proposed model can accurately determine the  $g_m$  ( $V_{GS}$ ) as well as its first and second derivatives.

The organization of the article is as follows. Section 2 describes the EEC topology and presents the proposed I/V nonlinear empirical model with the extraction procedure. Section 3 is related to the validation of the proposed model under DC and large-signal conditions. For this purpose, we made comparisons between the proposed model and the commercial model of transistors using measured data from load-pull, ACPR, and a 600 MHz class J power amplifier designed in S-band with a GaN FET CGH40010F, and we found that in all cases, the proposed model matches the measured data much better. These results demonstrate that it is essential to have an accurate  $I_{DS}$  ( $V_{GS}$ ,  $V_{DS}$ ) model to design a power amplifier. Finally, in Section 4 are presented the conclusions.

## 2 | MODEL DESCRIPTION

Figure 2 shows the EEC model of a microwave FET. The EEC has an extrinsic part represented by the parasitic elements  $R_g$ ,  $R_d$ ,  $R_s$ ,  $L_g$ ,  $L_d$ ,  $L_s$ ,  $C_{pg}$ , and  $C_{pd}$ ; these elements are bias independent. Moreover, in the GaN FETs, three R-C series networks (R11-C11, R22-C22, and R21-C21) are considered extrinsic elements.<sup>23</sup> These

**FIGURE 2** Nonlinear electrical equivalent circuit topology used



networks have been successfully adopted in References 5, 24, 25 to account for the substrate's conductivity, which affects the small-signal response.

Regarding the intrinsic part, its elements are nonlinear and bias-dependent. In that sense, the main nonlinear element of the model that contributes significantly to predicting the performance of a transistor or a microwave circuit is the drain-to-source current,  $I_{DS}(V_{GS}, V_{DS})$ , and the remaining intrinsic elements are weak nonlinearities that help to improve the accuracy of the model by modeling the FET dynamic behavior.

## 2.1 | The proposed I/V nonlinear model

The model from Reference 21, initially created for modeling the I/V characteristics of LDMOS and MOSFET devices,<sup>9,19,26</sup> has been used to model the nonlinear HEMT I/V behavior<sup>5-8,10</sup> due to its improved prediction of the nonlinear transconductance shape. Based on References 4 and 2, this work presents an empirical drain-to-source current model for HEMT GaAs and high power GaN devices.

The proposed I/V empirical model is given by

$$I_{DS}(V_{GS}, V_{DS}) = \beta_{eff} \cdot V_{gs3}^2 \cdot (1 + \lambda(V_{GS}) \cdot V_{DS}) \cdot \tanh\left(\frac{\alpha(V_{GS})}{V_{gs3}^{psat}} \cdot V_{DS}\right), \quad (1)$$

where:

$$\beta_{eff} = \frac{\beta}{1 + \mu_{crit} \cdot V_{gs3}^{g_{min}}}, \quad (2)$$

and

$$V_{gs3} = V_{ST} \cdot \ln\left(1 + e^{\left(\frac{V_{gs2}}{V_{ST}}\right)}\right), \quad (3)$$

$$V_{gs2} = V_{gs1} - \frac{1}{2} \left( V_{gs1} + \sqrt{(V_{gs1} - VK)^2 + \Delta^2} - \sqrt{VK^2 + \Delta^2} \right), \quad (4)$$

$$V_{gs1} = V_{GS} - V_{th}. \quad (5)$$

$$V_{th} = V_{th0} + \gamma \cdot V_{DS}, \quad (6)$$

$$g_{mlin} = g_{mlin0} + g_{mlinx} \cdot V_{DS}, \quad (7)$$

$$VK = VK_0 + VK_{sat} \cdot V_{DS}, \quad (8)$$

$$\lambda(V_{GS}) = \lambda_3 - \lambda_0(1 + \lambda_1 V_{GS}) \tanh(\lambda_2 V_{GS}), \quad (9)$$

$$\alpha(V_{GS}) = \alpha_0 + \alpha_1 e^{-\left(\frac{V_{GS} - \alpha_2}{\alpha_3}\right)^2}. \quad (10)$$

where  $V_{th}$ ,  $g_{mlin}$ , and  $VK$  are linear functions that depend on  $V_{DS}$ . Additionally,  $\beta$ ,  $\mu_{crit}$ ,  $V_{ST}$ ,  $\Delta$ ,  $psat$ ,  $\lambda_n$ , and  $\alpha_n$  are fitting parameters;  $V_{th0}$  is the threshold voltage for a  $V_{DS}$  close to 0 V, and  $\gamma$  is the slope of the linear dependence of  $V_{th}$  with  $V_{DS}$ .

It is worth to comment that  $g_{mlin}$  and  $VK$  are used to describe the  $V_{DS}$  dependence of  $g_m$  around its peak value and tune the slope of  $g_m$  in the linear and compression regions, respectively. To illustrate the above, let us analyze Figure 3, where is plotted the DC  $g_m$ , and it is possible to identify four operation regions (Sub-threshold, Quadratic, Linear, and Compression). When the active

device is biased in the saturation region<sup>21</sup> ( $V_{DS} > V_{knee}$ ), the  $g_m$  can be modeled using (2) and (3), and this is because (3) approaches gradually to zero in the sub-threshold region when  $V_{GS} < V_{th}$ , and the turn-on abruptness of  $g_m$ , in the quadratic region, is controlled by the parameter  $VST$ . Notice that  $g_m$  increases linearly when  $V_{GS} \gg V_{th}$ . This behavior is well described because  $V_{gs3}$  has a quasi-linear dependence with  $V_{gs2}$ , and the parameter  $\beta$  in (2) controls the slope. Moreover, the parameter  $\mu_{crit}$  in Equation (2) regulates the transition from the quadratic to the linear region. The compression region of the transconductance is tuned using function  $VK$  and parameter  $\Delta$ . They accurately represent the slow decay and elongation of the transconductance curve after achieving its maximum value (over the threshold voltage), which is the typical highly asymmetrical bell-shaped form of the transconductance when the transistor is biased in the saturation region. This set of equations allows a continuous and smoother transition in the derivative of  $I_{DS}$  as a function of  $V_{GS}$ .

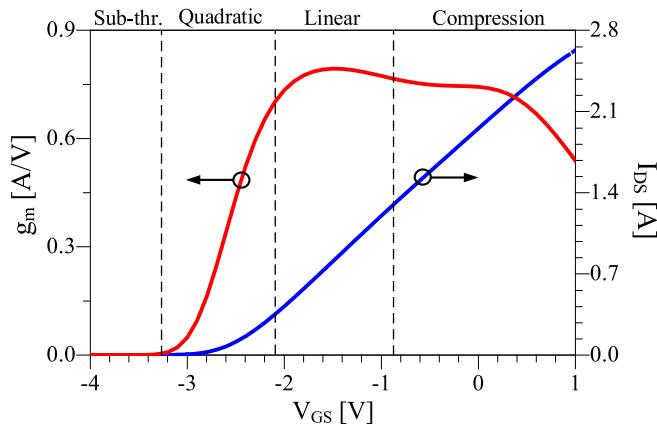
Furthermore, the addition of sub-threshold conduction in Equation (3) is critical to avoid a discontinuity in the drain current expression derivatives at  $V_{GS} = V_{th}$ .<sup>26</sup>

In contrast with the device models presented in References 3 and 16 for GaN FETs, in Reference 21 for LDMOS or Reference 20 for MESFET devices, the empirical formulation for  $I_{DS}(V_{GS}, V_{DS})$  presented in this article accurately represents the nonlinear transconductance shape and its transitions for all values of  $V_{DS}$ . This fact is not only for the highly asymmetrical bell-shaped form of the transconductance (already analyzed and modeled<sup>5,11</sup>), but also for the quasi-symmetrical bell-shaped form obtained when the device is biased with a  $V_{DS}$  in the ohmic region of the  $I_{DS} - V_{DS}$  characteristic<sup>22</sup> as seen in Figure 4, where a comparison between measured and

modeled  $g_m$  using Reference 21, and our proposed model, at low  $V_{DS}$  values obtained from pulsed measurement of the GaN FET CGH40045F, are plotted. The results demonstrate that the model of Reference 21 is unable to follow the transconductance behavior accurately, and hence its correlation with the measurement is low.

Finally, we want to highlight the main features of the proposed model inspired by the models of References 2 and 4:

1. To the best of the author's knowledge, Equation (7) is new. This equation, along with (8), allows us to accurately determine the maximum transconductance value and its shaping around this value, considering the transconductance's nonlinear behavior as  $V_{DS}$  changes.
2. Gate-to-source voltage dependence of the output conductance. This dependency is considered by Equation (9), which allows us to model the drain-to-source current at large  $V_{DS}$  values, especially when the  $V_{GS}$  voltage is close to  $V_{th}$ .
3. Function (10) is useful in reproducing the displacement of the knee voltage ( $V_{knee}$ ) in the  $I_{DS} - V_{DS}$  curves and modeling the subtle complexities in the ohmic region for different active devices. Moreover, it is important to comment that (10) directly influences the improved performance achieved in the transconductance modeling for low values of  $V_{DS}$ , as shown in Figure 4. In fact, in our proposed model, we adopted the term  $V_{gs3}^{psat}$  in Equation (1) instead of the  $\tanh(\alpha(V_{GS}) \cdot V_{DS})$  as proposed in Reference 4. This modification allows the drain saturation current to depend on the gate to source voltage, which causes the transconductance to decrease after the saturation point has been reached, likewise improves the fitting of the data in the ohmic region.



**FIGURE 3** Transconductance regions and drain current for a GaN HEMT CGH40010F at  $V_{GSq} = -2.5$  V and  $V_{DSq} = 28$  V, with a  $V_{DS}$  biased in the saturated region

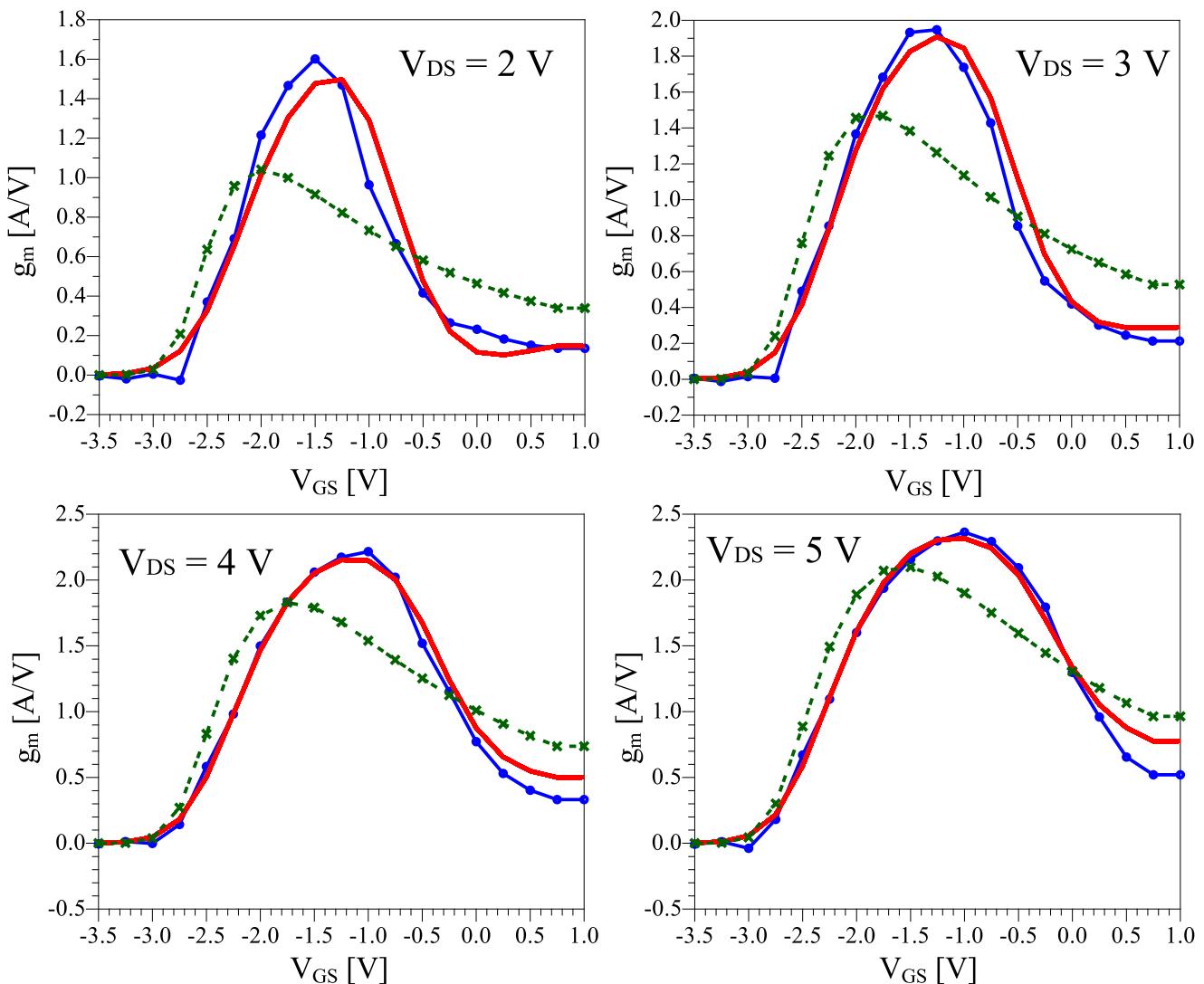
## 2.2 | Extraction procedure of initial values

The proposed model's initial values are determined using the experimental I/V data, such as parameters  $V_{th}$ ,  $V_{th0}$ , and  $\gamma$ . Parameter  $VST$  can be determined as follows<sup>9</sup>:

$$VST = \frac{I_{th}}{\ln(2)}, \quad (11)$$

where  $I_{th}$  is the threshold current at  $V_{GS} = V_{th}$ , extracted at a  $V_{DS}$  value close to 0 V.

Coefficient  $\mu_{crit}$  can be assigned as  $0.5 \text{ V}^{-1}$ , whereas coefficient  $psat$  can take an initial value close to one. Fitting parameters  $g_{mlin0}$ ,  $g_{mlinx}$ ,  $VK_0$ ,  $VK_{sat}$ , and  $\Delta$  can be



**FIGURE 4** Comparison of measured (circles) and modeled transconductance using the proposed model (solid lines), and the model from Reference 21 ( $\times$ ), for the GaN CGH40045F at  $V_{G Sq} = -2.5$  V and  $V_{D Sq} = 28$  V, illustrating the quasi-symmetrical bell-shaped characteristic of the transconductance at  $V_{DS}$  values inside the ohmic region of the  $I_{DS} - V_{DS}$  characteristic

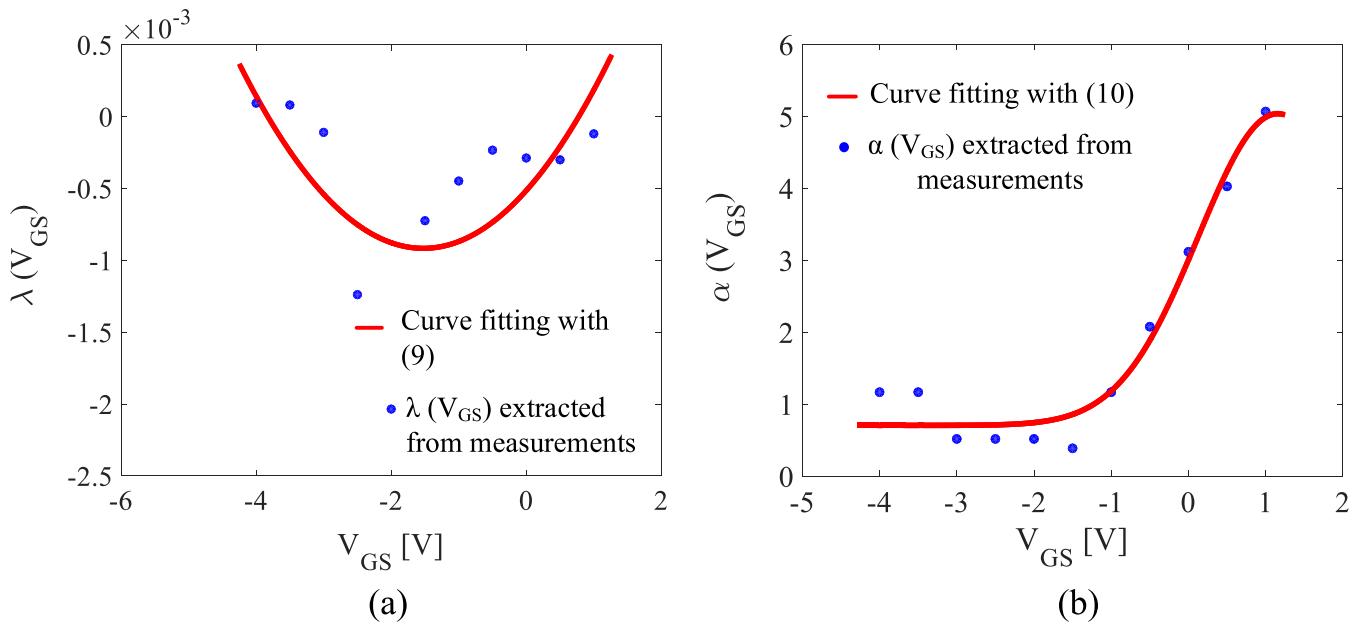
determined by numerical optimization. On the other hand, coefficient  $\beta$  can be fitted with reasonable accuracy by numerical experiments since this parameter has a well-defined meaning in terms of its influence on the shape of  $g_m$ .

Regarding the extraction of parameters in Equations (9) and (10), the output conductance ( $\lambda$ ) is determined from the slope of the  $I_{DS} - V_{DS}$  characteristic at a saturated channel condition, and the saturation voltage ( $\alpha$ ) is 25% of the knee voltage.<sup>13</sup> Thus,  $\lambda$  and  $\alpha$ , in the proposed model, are determined for all  $V_{GS}$  values greater than the pinch-off voltage following the above criteria. The result of these operations is a set of points of  $\lambda(V_{GS})$  and  $\alpha(V_{GS})$ , one for each  $V_{GS}$  value, as shown in Figure 5. Once  $\lambda(V_{GS})$  and  $\alpha(V_{GS})$  are known, the parameters  $\lambda_n$  ( $n = 0, 1, 2, 3$ ) and  $\alpha_n$  ( $n = 0, 1, 2, 3$ )

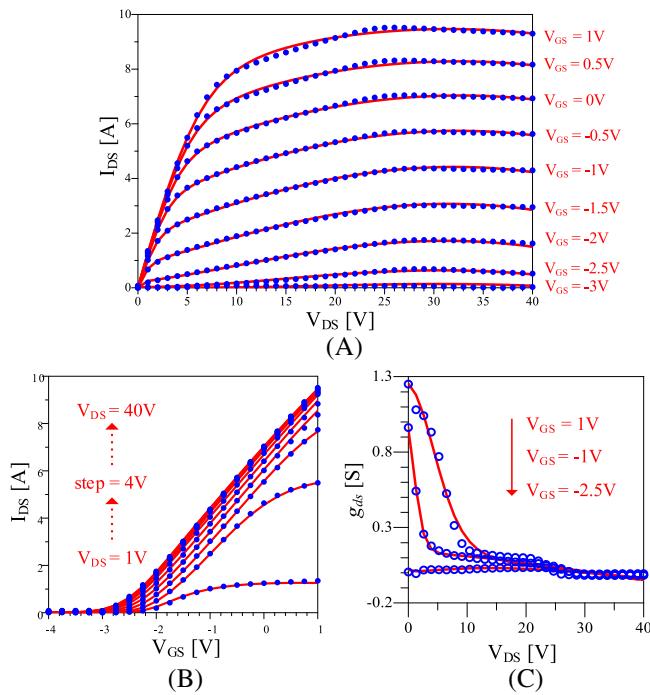
from Equations (9) and (10), respectively, are determined using the curve fitting toolbox from MATLAB and employing a nonlinear least-square optimization method.

### 3 | EXPERIMENTAL RESULTS

Three packaged active devices (two GaN HEMTs from Wolfspeed, CGH40010F and CGH40045F, and a GaAs ATF-34143) were measured and characterized under different operation conditions to validate the proposed model. The three packaged transistors' experimental I/V data were measured using the pulsed I/V system Auriga AU4950 configured with a pulse width of 4, 0.5, and 5  $\mu$ s for the packaged GaN HEMT CGH40045F, CGH40010F,



**FIGURE 5** Curve fitting analysis of functions  $\lambda(V_{GS})$  and  $\alpha(V_{GS})$  for the GaN CGH40045F employing the curve fitting toolbox of MATLAB with Equations (9) and (10)



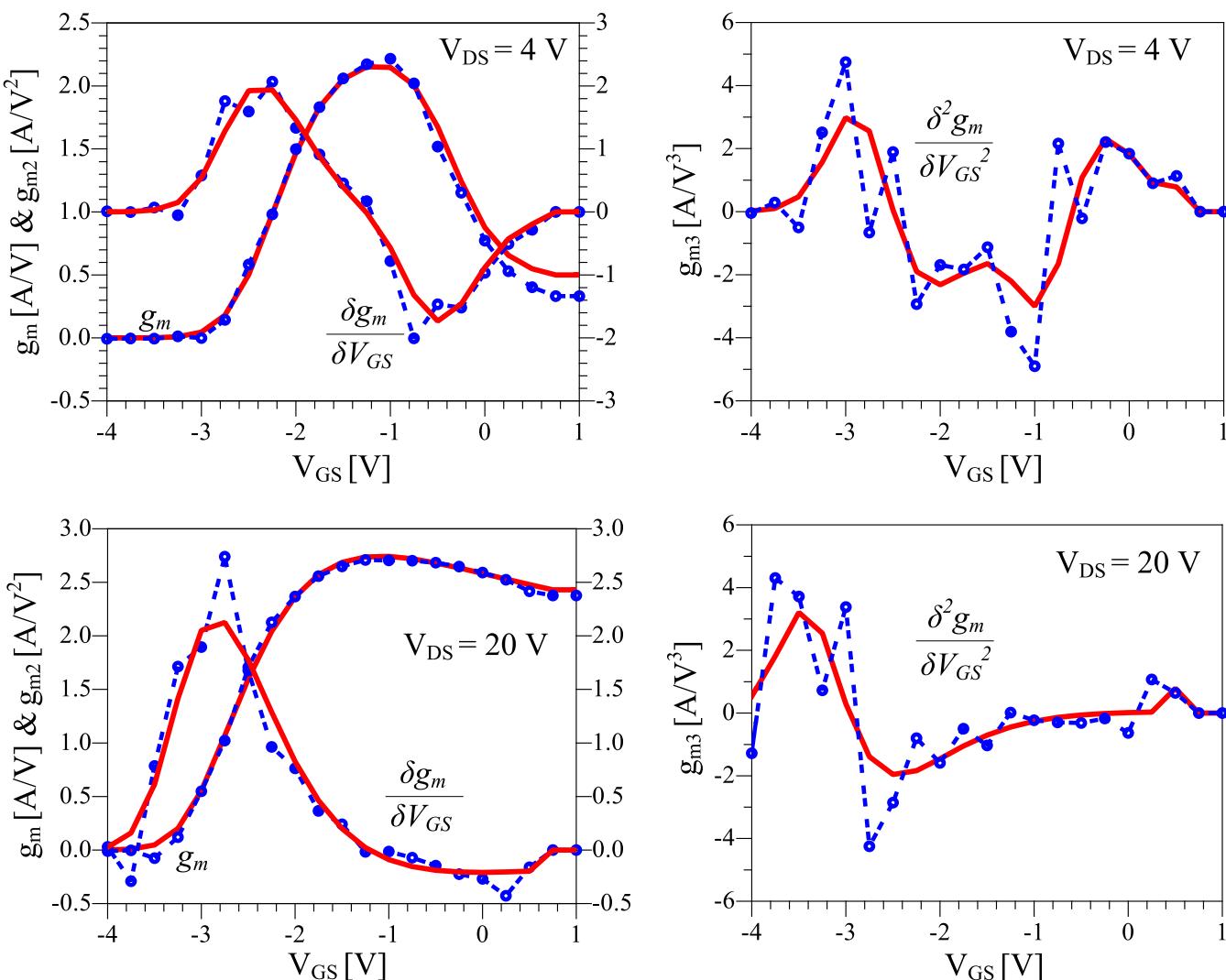
**FIGURE 6** Comparison of measured (circle) and simulated data using the proposed model (solid line), for a packaged GaN HEMT CGH40045F. (A)  $I_{DS}$  –  $V_{DS}$ , (B)  $I_{DS}$  –  $V_{GS}$ , and (C) output conductance, at a  $V_{GSq} = -2.5$  V and  $V_{DSq} = 28$  V. Nonquiescent drain voltage was swept from  $V_{DS} = 0$  V to  $V_{DS} = 40$  V with steps of 1 V, and the nonquiescent gate voltage was swept from  $V_{GS} = -4$  V to  $V_{GS} = 1$  V with steps of 0.5 V

**TABLE 1** Model parameter values for the CGH40045F

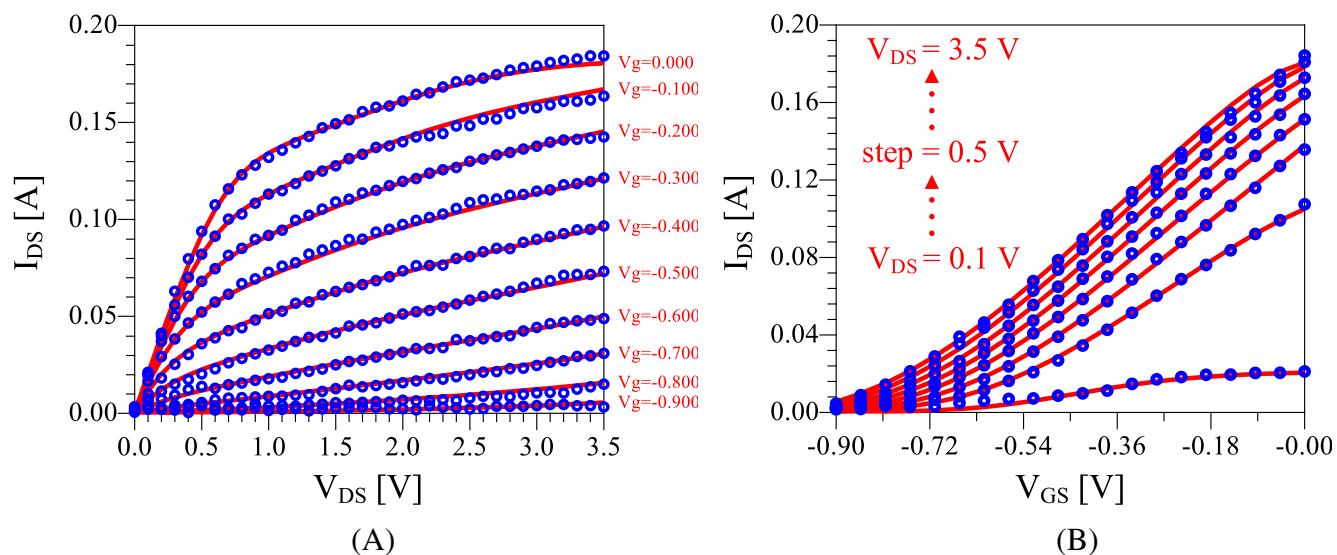
VST (V)	$\mu_{crit}$ (V <sup>-1</sup> )	g <sub>mlin0</sub>	V <sub>K0</sub> (V)	V <sub>tho</sub> (V)
0.25	0.26	1.692	0.04681	-2.377
$\gamma$	$\beta$ (1/Ω)	g <sub>mlin0</sub> (V <sup>-1</sup> )	V <sub>Ksat</sub>	psat
-0.04027	1.78555	0.01463	5.306	0.62249
$\lambda_0$ (1/V)	$\lambda_1$ (1/V)	$\lambda_2$ (1/V)	$\lambda_3$ (1/V)	$\Delta$ (V)
-1.98237	-0.006269	0.0072705	0.0174368	13.2896
$\alpha_0$ (1/V)	$\alpha_1$ (1/V)	$\alpha_2$ (V)	$\alpha_3$ (V)	
0.657506	3.75704	-3.06779	-1.82375	

and the pseudomorphic HEMT ATF-34143, respectively. Each active device was biased as follows; for the CGH40010F a  $V_{GSq} = -2.8$  V and  $V_{DSq} = 28$  V, for the CGH40045F a  $V_{GSq} = -2.5$  V and  $V_{DSq} = 28$  V, and finally for the ATF34143 a  $V_{GSq} = -0.54$  V and  $V_{DSq} = 2$  V.

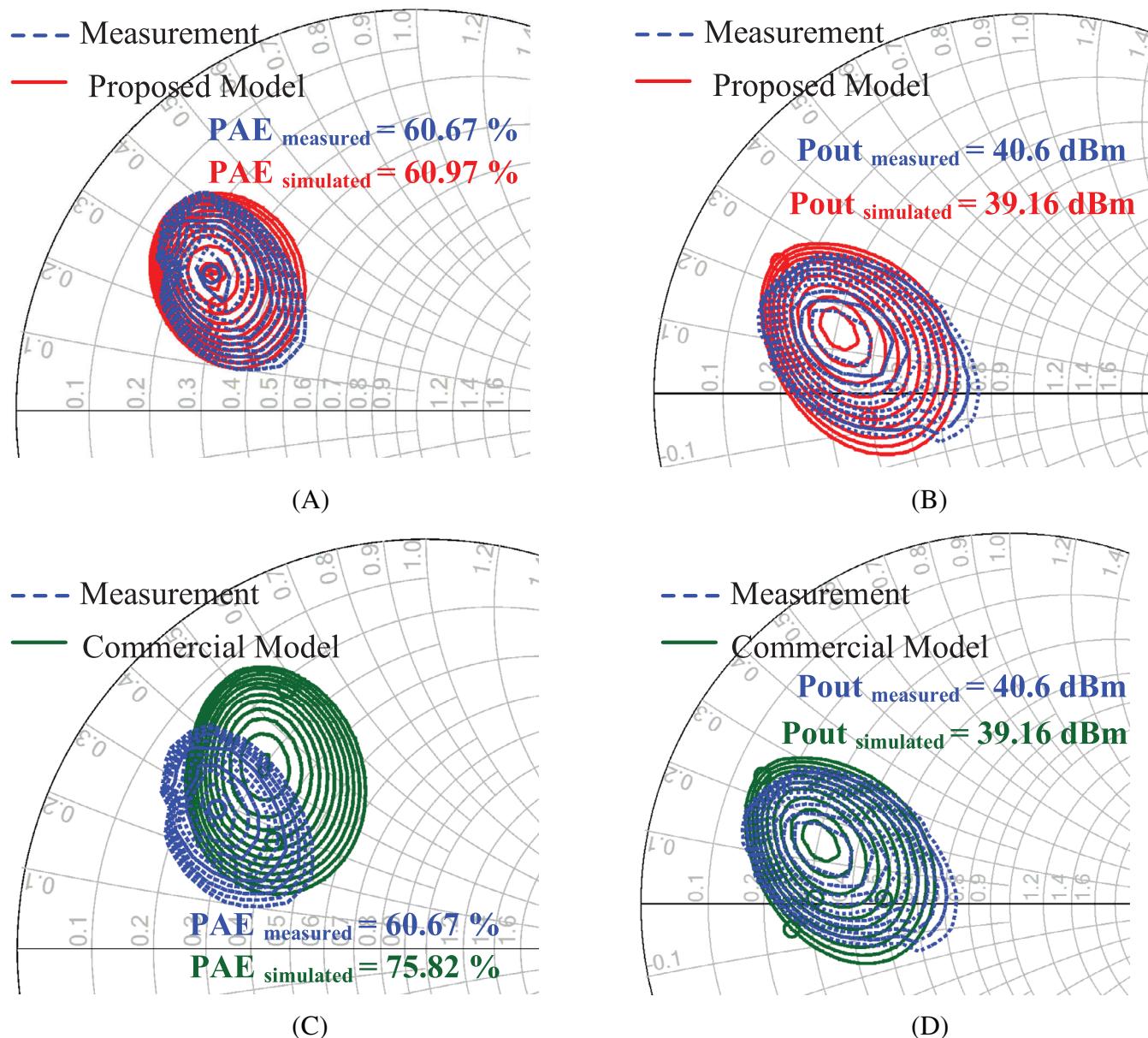
On the other hand, S-parameters measured with a calibrated<sup>27</sup> PNA-X N5242A from 0.4 to 4.0 GHz were used to extract the devices' extrinsic and intrinsic elements. In that sense, the first step was to obtain the R-C series networks (R11-C11, R22-C22, and R21-C21) from the admittance parameters of an open dummy structure.<sup>25,28</sup> Considering that the number of parasitic elements surpasses the number of equations representing the real and imaginary parts of the admittance



**FIGURE 7** Comparison of measured (circles) and modeled transconductance and its first and second derivatives using the proposed model (solid lines) for the GaN CGH40045F. At a  $V_{GSq} = -2.5\text{ V}$  and  $V_{DSq} = 28\text{ V}$ , illustrating the transconductance accurate representation at  $V_{DS} = 4\text{ V}$  (top) and  $V_{DS} = 20\text{ V}$  (below)



**FIGURE 8** Comparison of measured (circles) and modeled drain-to-source current using the proposed model (solid lines) with  $V_{GSq} = -0.54\text{ V}$  and  $V_{DSq} = 2\text{ V}$ , for the GaAs ATF34143. (A)  $I_{DS} - V_{DS}$  and (B)  $I_{DS} - V_{GS}$



**FIGURE 9** Comparison of load-pull PAE and  $P_{\text{out}}$  contours at 2 GHz with an input power of 31 dBm for the HEMT CGH40010F, at a  $V_{\text{GSq}} = -2.8$  V and  $V_{\text{DSq}} = 28$  V. Measured data (dot) and simulated with the proposed model (solid line) (A,B), and with the commercial model (solid line), (C,D)

parameters, an optimization was done using the commercial CAD tool Advance Design System (ADS) to obtain these six elements.<sup>24,28</sup> Afterward, biasing the transistors under forward and reverse Cold-FET conditions, as suggested in References 29–31, the extrinsic elements were extracted. After de-embedding the extrinsic elements from hot S-parameters measurements, using a similar procedure described by References 32, 33, the intrinsic elements were determined using the method reported in Reference 34.

The EEC (see Figure 2) considering the proposed I/V model is easy to implement in the microwave circuit

simulator ADS using the SDD component. It is worth commenting that thermal effects and traps are not considered in this work. Besides, the I/V pulsed data are not isodynamic I/V<sup>35, 36</sup> because we currently do not have a test bench for making these measurements. However, our main interest is to highlight the importance of an accurate I/V model for designing microwave circuits like power amplifiers, even when it does not consider the thermal or traps effects. In that sense, we will compare simulation results using our proposed model with the transistors' commercial models under different large-signal tests.

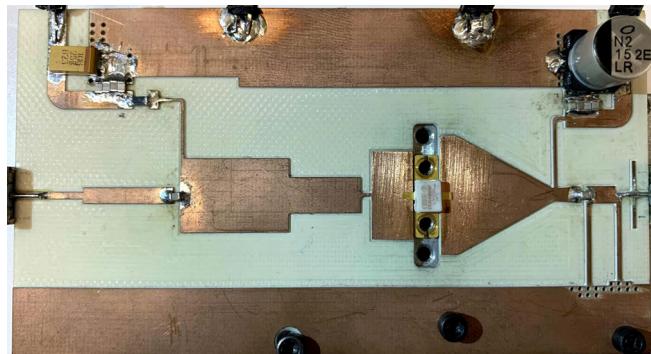
### 3.1 | DC validation of the proposed model

The I/V models developed for the packaged GaN FETs CGH40010F, CGH40045F, and the GaAs FET ATF34143 use pulsed I/V measured data for mitigating self-heating and trapping conditions. In the CGH40010F and ATF34143, the quiescent gate voltage was set to less than the pinch-off voltage. Therefore, the devices dissipate no power and, consequently, are free of any static self-heating conditions.<sup>11</sup> However, in the GaN CGH40045F, the quiescent gate voltage was a bias point where the transistor is in conduction, as a class AB, and therefore it presents a prominent kink effect when compared with the other two transistors.

Figure 6 compares the measured and modeled data of the intrinsic  $I_{DS} - V_{DS}$ ,  $I_{DS} - V_{GS}$ , and the output conductance for the packaged HEMT CGH40045F, whereas in Table 1 are reported the values of the fitting parameters. Notice the excellent fitting of the proposed model with the measured data compared with the simulation results using the model from Reference 21 (see Figure 1). These results demonstrate the proposed model's versatility for modeling atypical behaviors on the FET's I/V curves. Moreover, Figure 7 shows measured and modeled  $g_m$  data and its first and second derivatives ( $g_{m2}$  and  $g_{m3}$ ), where the relative error is less than 10% with our proposed model.

Besides, Figure 8 shows the comparison results between measured and modeled data of the intrinsic  $I_{DS} - V_{DS}$  and  $I_{DS} - V_{GS}$  for the GaAs ATF34143; also, a high degree of precision is achieved resulting from using the proposed  $I_{DS}(V_{GS}, V_{DS})$  model.

In practice, the measured pulsed I/V curves depend on the quiescent point, particularly with  $V_{GS}$ . In some cases, depending on each particular FET, the slope of the I/V curves in the ohmic and saturation regions can change significantly. In that sense, it is important to

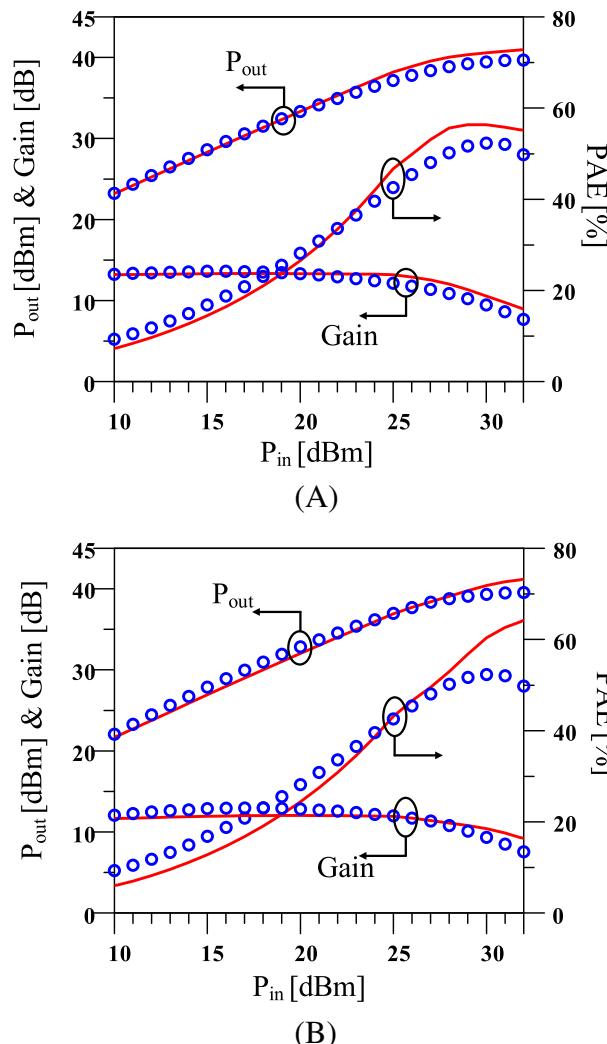


**FIGURE 10** Photograph of the implemented class-J PA using a HEMT CGH40010F

comment that in the proposed model, these changes can be considered by tuning the variables  $\alpha_n$ ,  $VST$ ,  $V_{th0}$ ,  $psat$ , and  $\gamma$  for the ohmic region; and  $\lambda_n$  for the saturation region. This feature helps to quickly model the pulsed I/V characteristics at a different quiescent point of the same FET.

### 3.2 | GaN HEMT CGH40010F: load-pull measurements and broadband high efficiency Class-J PA design

Now the proposed model is validated under large-signal conditions. For this purpose, the output power ( $P_{out}$ ) and power added efficiency (PAE) contours of the GaN

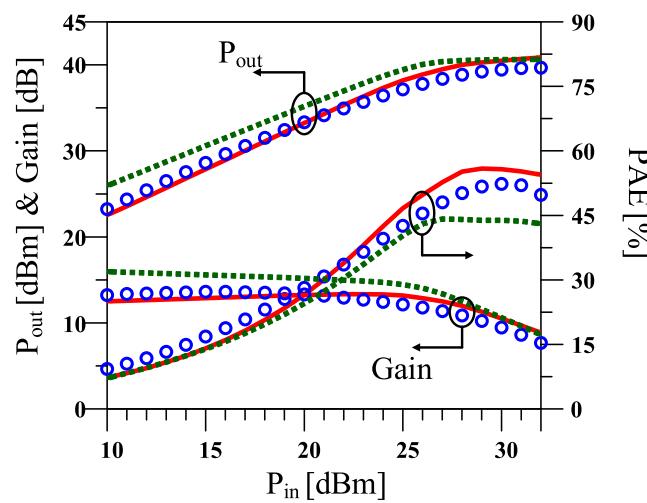


**FIGURE 11** Comparison of measured (circles) and simulated (solid line)  $P_{out}$ , PAE and Gain vs input power for the class-J PA designed using the proposed model under CW operation, for (A) 3 GHz and (B) 3.6 GHz.  $V_{GSq} = -2.8$  V and  $V_{DSq} = 28$  V

HEMT CGH40010F were measured using an impedance tuner i1818 from Focus Microwaves.

Figure 9 shows the load-pull contours; these contours correspond to 2 GHz with an input power of 31 dBm and biased with quiescent voltages of  $V_{DSq} = 28$  V and  $V_{GSq} = -2.8$  V (class-B operation). Notice that the simulated contours with the proposed model, Figure 9A,B, can predict the behavior of the GaN CGH40010F very well, where the maximum simulated PAE has an absolute error of 0.3%, and the absolute error for the maximum power prediction is 1.4 dB. On the other hand, Figure 9C,D shows the simulated load-pull contours results using the nonlinear commercial model, where the user cannot access or modify the model's equations. Notice that although the commercial model contains parasitic elements, nonlinear intrinsic elements, and thermal/charge trapping models, the PAE contours prediction is highly inaccurate. It is worth to comment that we also compared the simulated Load-Pull data using the model of Reference 21, and the results are slightly different from our model. This fact was expected as both models fit very well in the transistor's I/V curves' saturation region. However, for applications other than designing an amplifier such as designing a low distortion mixer,<sup>37</sup> our proposed model is likely more suitable since in the latter case the I/V model has to fit very well in the ohmic region.

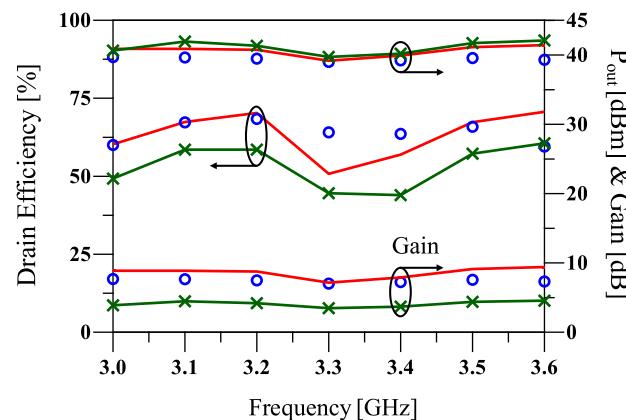
Furthermore, we designed a broadband high-efficiency class-J PA with the aid of the proposed model. The designed class-J PA, biased with  $V_{DS} = 28$  V and  $V_{GS} = -2.8$  V, operates in the S-band (3 GHz). The input



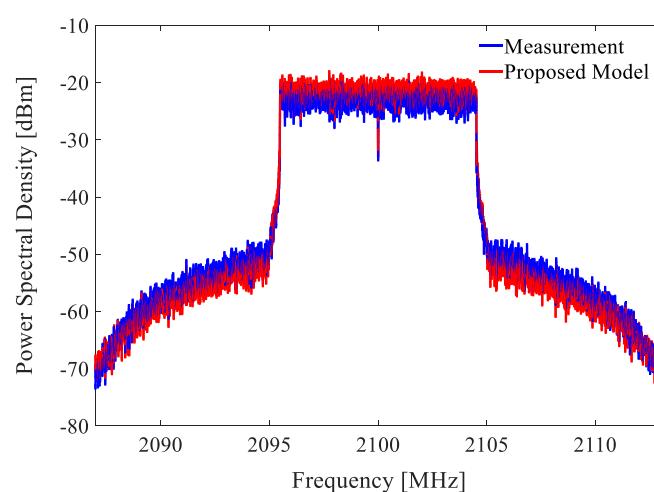
**FIGURE 12** Comparison of measured (circle), simulated with the commercial model (discontinuous line) and simulated with the proposed model (solid line)  $P_{out}$ , PAE and Gain vs input power for the designed class-J PA under 3 GHz CW operation, at a  $V_{GSq} = -2.8$  V and  $V_{DSq} = 28$  V

and output matching networks were designed based on the characteristics of the Rogers high-frequency laminate RO4003C with  $\epsilon_r = 3.55$ . Figure 10 shows a photograph of the fabricated power amplifier.

Figure 11 compares the class-J PA simulation results using the proposed model with measured data of  $P_{out}$ , PAE, and gain under CW operation at two different frequencies, 3 and 3.6 GHz. The experimental results show a class-J PA with a maximum  $P_{out}$  and PAE of 10 W and 52%, respectively, with an associated gain of approximately 8 dB for the two frequencies. Notice the high correlation between the simulated and measured data of the  $P_{out}$  and gain, using the proposed model, where errors less than 1 dB are observed. Regarding the PAE



**FIGURE 13** Comparison of measured (circle), simulated with the commercial model (x) and simulated with the proposed model (solid line)  $P_{out}$ , Gain and Drain Efficiency vs frequency for the designed class-J PA with an input power of 32 dBm, at a  $V_{GSq} = -2.8$  V and  $V_{DSq} = 28$  V



**FIGURE 14** Comparison of measured and simulated, with the proposed model, output power spectra using a 10 MHz bandwidth LTE signal with  $V_{GSq} = -0.54$  V,  $V_{DSq} = 2$  V and  $P_{in} = -2$  dBm

**TABLE 2** Measured and simulated results for the GaAs HEMT ATF-34143 under digital modulated stimulus

$V_{GS}$ (V)	$V_{DS}$ (V)	$P_{in}$ (dBm)	$P_{out}$ (dBm)			ACPR-L (dBc)			ACPR-U (dBc)		
			Meas.	Sim.	Comm.	Meas.	Sim.	Comm.	Meas.	Sim.	Comm.
-0.54	2	-4	8.06	9.28	11.03	-38.1	-41.7	-30.7	-37.87	-41.8	-30.64
-0.54	2	-2	9.81	11.1	12.87	-32.57	-34.8	-28.48	-32.48	-35.01	-28.54
-0.8	2	-2	7.3	10.1	12.94	-32.92	-35.55	-28.37	-33.58	-36.36	-28.44

Note: Sim. and Comm. correspond to simulated results using the proposed model and the commercial model, respectively.

simulation, it deviates slightly from the measurements. In the worst case, there is a difference of 10%. Nevertheless, compared with simulation results using the nonlinear commercial model at 3 GHz, the proposed model predicts the  $P_{out}$ , PAE, and gain accurately, as shown in Figure 12.

Moreover, Figure 13 shows the measured and simulated  $P_{out}$ , drain efficiency, and gain results in the frequency band of 3 to 3.6 GHz. Good agreement between measurement and simulation with the proposed model was achieved, not being the case when using the commercial model.

### 3.3 | GaAs HEMT ATF34143: intermodulation distortion

Finally, we compared the simulated and measured ACPR-generated from a pseudomorphic GaAs HEMT ATF-34143, using an LTE signal with 10 MHz of bandwidth, 7 dB of PAPR, and centered at a frequency of 2.1 GHz. The ACPR was measured using a vector signal generator (MXG N5182A) and a digital signal analyzer (DSA90404A), both controlled by the VSA software, and with the aid of ADS was obtained the output power spectrum. Figure 14 shows the measured and simulated output spectrum, where it is remarkable the excellent correlation between the simulation and the measurement. Table 2 reports the measured and simulated data, using the proposed and commercial models, of the output power and ACPR at different input power levels and bias points. An error of less than 1.3 dB is observed in  $P_{out}$ , and in the case of the ACPR, the maximum error is 3.93 dB when using the proposed model, which is not the case when using the commercial model where the error predicting the  $P_{out}$  ascend to 3 dB and the ACPR error to 7.4 dB. These results demonstrate that an accurate I/V model is the most critical nonlinear element to consider in the EEC of microwave FETs; this allows us to get a good approximation of the principal figures of merit of the power amplifier.

## 4 | CONCLUSIONS

An improved nonlinear empirical I/V model suitable for modeling the drain-to-source current of GaAs and GaN FETs has been presented. The proposed model introduces modifications to an LDMOS behavioral equations, allowing us to consider the output conductance and saturation voltage dependence with  $V_{GS}$ . Moreover, the proposed model is able to describe the highly nonlinear shape of the transconductance for all  $V_{DS}$  values. That is possible, thanks to a new function that considers the dependence of  $V_{DS}$  in the linear region of the transconductance. Furthermore, with the proposed model, GaN, and GaAs HEMTs may be modeled from the ohmic to the saturation region, including the pinch-off region and the first and second derivatives of the transconductance. A microwave FET model that consists of package parasitic, linear extrinsic, and bias-invariant intrinsic elements, along with the proposed I/V model, has demonstrated the capacity to represent the behavior under large-signal excitation of GaAs and GaN transistors. Load-Pull and ACPR measurements data at different frequencies validate the proposed model. Furthermore, data measurements of a class-J PA designed at S-band frequencies with 600 MHz of bandwidth were used, demonstrating the proposed model's usefulness to simulate microwave power amplifiers.

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## DATA AVAILABILITY STATEMENT

The data that support the findings of this study are available from the corresponding author upon reasonable request.

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