OVERVIEW OF RF LARGE-SIGNAL MODELLING TECHNIQUES BASED ON TIME-DOMAIN LARGE-SIGNAL MEASUREMENTS

D. Schreurs

K.U.Leuven, Div. ESAT-TELEMIC, Kasteelpark Arenberg 10, B-3001 Leuven-Heverlee, Belgium
E-mail: Dominique.Schreurs@esat.kuleuven.ac.be

ABSTRACT

Large-signal models for microwave devices are commonly based on DC and S-parameter measurements. The recent availability of full two-port vectorial large-signal measurement systems has initiated the development of new techniques to enhance the ease and accuracy of non-linear device model generation. This paper gives a critical overview of the different modelling techniques that are based on the time-domain representation of those vectorial large-signal measurements. Each of these techniques will be demonstrated by experimental results on microwave transistors.

INTRODUCTION

For decades, the large-signal behaviour of microwave devices has been modelled indirectly using DC and small-signal S-parameter measurements. This linearisation approach was necessary, because vectorial information (i.e., both amplitude and phase) could only be accessed by small-signal ratio measurements, being S-parameters. Since about 15 years, metrology groups focus on developing a measurement set-up that enables absolute vectorial large-signal characterisations [1]-[4]. Once the first prototype systems became mature, also new large-signal modelling techniques using these large-signal data started to be developed. Since this is a novel area, a lot of options are open at the start.
A first way to categorize modelling methods is the representation. The device can either be represented by an equivalent scheme or by a black-box model. The former is often limited to transistors, because the dynamics of microwave circuits easily become too complex to be captured in an a-priori determinable equivalent scheme. Consequently, black-box models are well suited for more complicated components, because they require no physical pre-knowledge.
A second way to subdivide is the modelling technique, namely optimisation or extraction. Optimisation is inherently linked to black-box models, but it is also used to estimate the parameters of empirical functions by which the device's state-functions can be represented. Extraction means that the state-functions are calculated directly from the measurements by evaluating analytical expressions.
The final consideration is the choice of analysis domain. Until now, most methods have been based on time-domain data, but also frequency-domain approaches are under consideration. A reason to prefer the time-domain is the direct compatibility with harmonic-balance analysis, which is standardly available in most microwave circuit simulators.
As demonstration of the above overview, we will present and analyse experimental results obtained from the extraction/optimisation and from the equivalent circuit/black-box model approaches.

EXTRACTION OF EQUIVALENT CIRCUIT MODELS

The first considered modelling technique is closely related to the well-known method of equivalent circuit model extraction using multi-bias S-parameter measurements. To illustrate the principle, we consider the intrinsic quasi-static large-signal model of a FET, as represented by Fig. 1. To be complete, the extrinsic elements, as well as the non-quasi-static effects should normally be added. The simplified scheme consists of the parallel connection of a charge and current source at both port 1 (the gate) and port 2 (the drain), which are the device's state-functions.

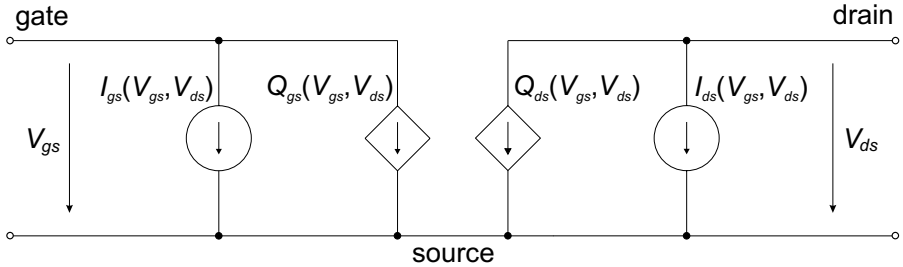

Fig. 1. Intrinsic quasi-static large-signal FET model.

We investigated how these 4 state-functions can directly be determined from full two-port vectorial large-signal measurements [5]. The terminal current I_{m2} at port 2 can be expressed in the time domain by:

$$I_{m2}(t) = I_{ds}\left(V_{gs}(t), V_{ds}(t)\right) + \frac{dQ_{ds}\left(V_{gs}(t), V_{ds}(t)\right)}{dt} \tag{1}$$

By defining

$$C_{21}\left(V_{gs}(t), V_{ds}(t)\right) = \frac{\partial Q_{ds}\left(V_{gs}(t), V_{ds}(t)\right)}{\partial V_{gs}} \qquad C_{22}\left(V_{gs}(t), V_{ds}(t)\right) = \frac{\partial Q_{ds}\left(V_{gs}(t), V_{ds}(t)\right)}{\partial V_{ds}} \tag{2}$$

(1) can be rewritten as:

$$I_{m2}(t) = I_{ds}\left(V_{gs}(t), V_{ds}(t)\right) + C_{21}\left(V_{gs}(t), V_{ds}(t)\right)\frac{dV_{gs}(t)}{dt} + C_{22}\left(V_{gs}(t), V_{ds}(t)\right)\frac{dV_{ds}(t)}{dt} \tag{3}$$

A similar equation can be written down for the terminal current I_{m1}. This results into a set of 2 equations with 6 unknowns, which can be solved if we have 3 distinct measurements by which the instantaneous terminal voltages remain unchanged. This imposes special experimental conditions on the large-signal measurements, such as e.g., the presence of a loadpull system [6], which is a drawback of this extraction method.

The extraction technique is illustrated on a HEMT. Fig. 2 presents the extracted C_{11} and I_{ds}, which are the two most important non-linear elements. The extractions agree well with reference results, obtained by standard DC and S-parameter measurements on the same device.

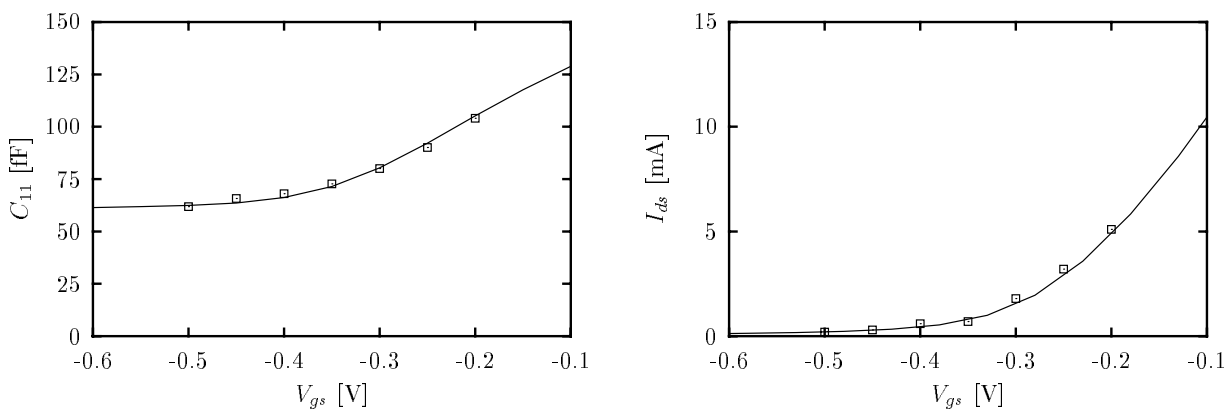

Fig. 2. Comparison of the DC/VNA measured (—) and large-signal measurement based extracted (□) C_{11} and I_{ds} of a HEMT at V_{dsDC}= 1.2 V.

PARAMETER ESTIMATION OF EMPIRICAL MODELS

The device's state-functions are often represented by particular analytical expressions. Such models are typically denoted as "compact" or "empirical". The aim of the modelling method is here to estimate the parameters of these expressions through optimisation. The classical procedure is to optimise the functions towards the DC measured state-functions, e.g., I_{ds}, and/or the S-parameter measurement based state-functions, e.g., C_{gs} or the corresponding Q_{gs}.

We have elaborated a parameter estimation procedure based on vectorial large-signal measurements only [7]. The advantage is that only one type of measurements, i.e., vectorial large-signal measurements, and only one type of simulation, i.e., harmonic-balance analysis, are needed. The procedure starts by performing a number of measurements, where it is possible to sweep any degree of freedom, such as input power, excitation frequency, DC bias, load impedance, …, but typically one focuses on particular experimental conditions depending on the application. Subsequently, the model parameters are estimated during one global optimisation process in which all the measurements are combined. The optimisation goals are expressed in terms of minimising the difference between the measured and simulated spectral components where we consider all significant harmonics and, if present, intermodulation products.

To demonstrate the method, we have taken the Chalmers model as empirical HEMT model [8]. All model parameters are simultaneously optimised towards a set of power-swept measurements at a fixed DC bias (class B operation). Fig. 3 compares the measured and simulated $I_{m1}(t)$ as function of $V_{gs}(t)$, and $I_{m2}(t)$ as function of $V_{gs}(t)$ at a high input power. This Figure clearly indicates a very good agreement and hence high model accuracy.

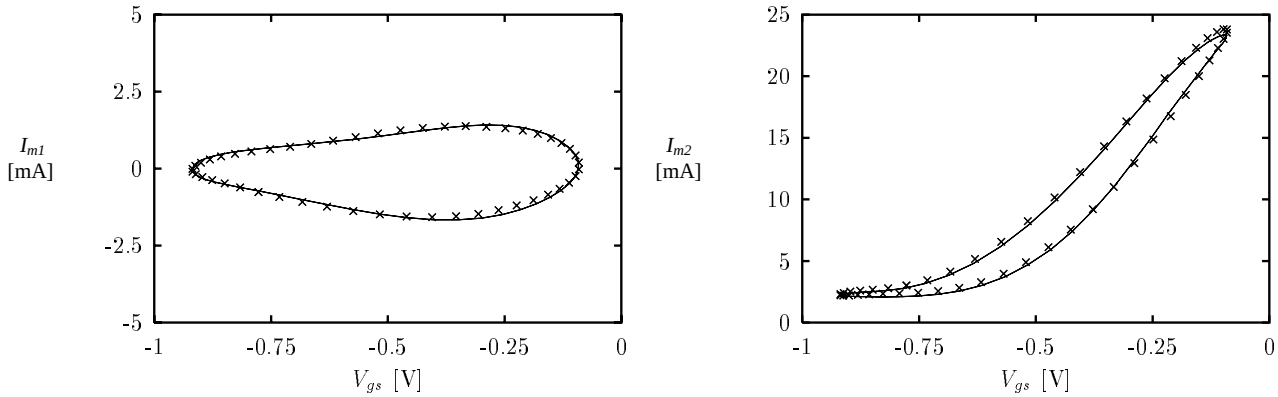

Fig. 3. Comparison of the measured (x) and simulated (—) I_{m1}(t) versus V_{gs}(t) and I_{m2}(t) versus V_{gs}(t) of a HEMT (V_{gsDC}= -0.5 V, V_{dsDC}= 1.5 V, f_0= 3.6 GHz, P_{in}= -3.4 dBm).

OPTIMISATION OF BLACK-BOX MODELS

A third possible model is the so-called black-box model, which is completely based on measurements and does not require any physical pre-knowledge of the device. In other words, the main difference with the model types discussed above is that the equivalent scheme does not need to be known. In fact, the number of state variables is treated as an unknown, which needs to be determined by an additional step in the black-box modelling procedure.

When considering (3), we notice that the terminal currents are a function of the terminal voltages and the first derivatives of the terminal voltages. This can be generalised for an electrical two-port by equations of the form [9]

$$\begin{aligned} I_{m1}(t) &= f_1\left(V_1(t),V_2(t),\dot{V}_1(t),\dot{V}_2(t),\ddot{V}_1(t),\ddot{V}_2(t),\cdots,\dot{I}_{m1}(t),\dot{I}_{m2}(t),\cdots\right) \\ I_{m2}(t) &= f_2\left(V_1(t),V_2(t),\dot{V}_1(t),\dot{V}_2(t),\ddot{V}_1(t),\ddot{V}_2(t),\cdots,\dot{I}_{m1}(t),\dot{I}_{m2}(t),\cdots\right) \end{aligned} \tag{4}$$

with $I_{m1}(t)$ and $I_{m2}(t)$ the terminal currents, and $V_1(t)$ and $V_2(t)$ the terminal voltages.

The objective of the modelling technique is to find the functional relationships $f_1(.)$ and $f_2(.)$ by fitting the measured terminal currents to the measured independent variables or state variables.

Since the black-box modelling approach supposes that no physical background information is available, we have to determine the independent (or state) variables. This can be accomplished by the so-called “embedding” technique, based on the “false nearest neighbour” principle [10]. The method consists in unfolding the characteristics of the dependent variables I_{m1} and I_{m2} in an increasing dimensional space by gradually increasing the number of independent variables $\left(V_1(t),V_2(t),\dot{V}_1(t),\dot{V}_2(t),\ddot{V}_1(t),\ddot{V}_2(t),\cdots\right)$ until a single-valued function for each current is obtained. By applying this technique e.g., to the measured data of a HEMT, we found we must include state variables up to the 2nd derivative of $V_1(t)$ and $V_2(t)$, for both I_{m1} and I_{m2}.

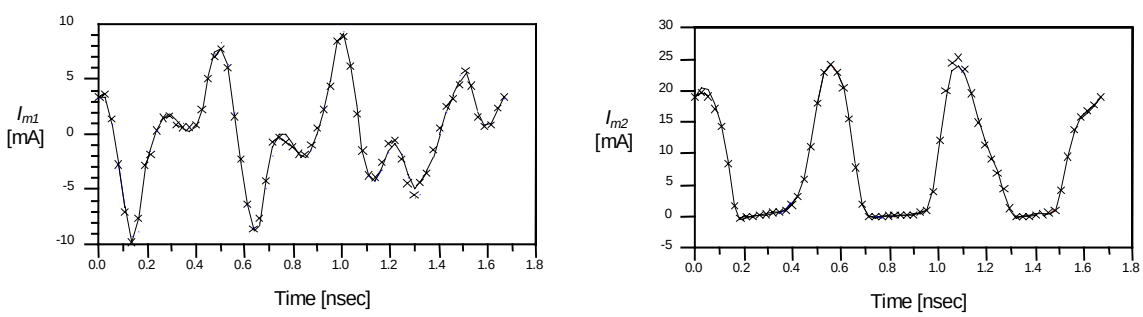

Fig. 4. Comparison of the measured (x) and ANN-based simulated (solid line) waveforms of the terminal currents for a HEMT (V_{qsDC} = -0.4 V, V_{dsDC} = 0.7 V, f_0=1.8 GHz, a_1= -3.9 dBm (with a secondary excitation at port 1 of -15.4 dBm at 4.2 GHz), a_2= -0.7 dBm, $\phi(a_2)$-$\phi(a_1)$=22°).

Next, the functional relationships $f_1(.)$ and $f_2(.)$ are determined by fitting the measured time domain terminal currents to the measured independent variables, determined in the preceding step. As fitting function, we can use several representations, such as polynomials and artificial neural networks (ANNs). Finally, we implement the obtained model in a commercial microwave circuit simulator by means of a Symbolically Defined Device (SDD). The SDD can determine the time-derivatives of the terminal voltages at each time-step in the simulation, enabling the fitting functions for the currents to be evaluated.

An example using ANNs is shown in Fig. 4. The ANN has first been trained using the back-propagation algorithm, and has consequently been fine-tuned using the quasi-Newton method, as implemented in the software developed by Prof. Zhang and his team [11]. Fig. 4 shows the good agreement between measured and simulated time-domain waveforms of the terminal currents.

CONCLUSIONS

We have shown that several interesting non-linear modelling approaches based on vectorial large-signal measurements are being developed. If the underlying equivalent scheme represents well the physical behaviour of the device, the extraction method yields the best validity range, but the measurements require special attention. The drawback of methods relying on optimisation is that extrapolation beyond the operating conditions of which measurement data were included in the optimisation process is limited and depends on the formulation of the model expressions. Black-box models are widely applicable, but also here the measurements require caution, because the state variables are unknowns and have in fact to be deduced from the available measurements. As general conclusion, it is clear that there is not such thing as a “golden” modelling technique, but that the best trade-off has to be found for each device type and application.

ACKNOWLEDGEMENTS

D. Schreurs is supported by the Fund for Scientific Research-Flanders as a post-doctoral fellow. K.U.Leuven acknowledges Agilent Technologies for the donation of the NNMS.

REFERENCES

[1] F. van Raay and G. Kompa, “A new on-wafer large-signal waveform measurement system with 40 GHz harmonic bandwidth,” *IEEE MTT-S Int. Microwave Symp.*, pp. 1435-1438, 1992.

[2] M. Demmler, P. Tasker, and M. Schlechtweg, “A vector corrected high power on-wafer measurement system with a frequency range for the higher harmonics up to 40 GHz,” *European Microwave Conf.*, pp. 1367-1372, 1994.

[3] C. Wei, Y. Lan, J. Hwang, W. Ho, and J. Higgins, “Wave-form characterization of microwave power heterojunction bipolar transistors,” *IEEE MTT-S Int. Microwave Symp.*, pp. 1239-1242, 1995.

[4] J. Verspecht, P. Debie, A. Barel, and L. Martens, “Accurate on wafer measurement of phase and amplitude of the spectral components of incident and scattered voltage waves at the signal ports of a nonlinear microwave device,” *IEEE MTT-S Int. Microwave Symp.*, pp. 1029-1032, 1995.

[5] D. Schreurs, J. Verspecht, B. Nauwelaers, A. Van de Capelle, and M. Van Rossum, “Direct extraction of the non-linear model for two-port devices from vectorial non-linear network analyzer measurements,” *European Microwave Conf.*, pp. 921-926, 1997.

[6] M. Curras-Francos, P. Tasker, M. Fernandez-Barciela, Y. Campos-Roca, and E. Sanchez, “Direct extraction of nonlinear FET Q-V functions from time domain large-signal measurements,” *IEEE Microwave and Guided Wave Letters*, pp. 531-533, 2000.

[7] D. Schreurs, J. Verspecht, S. Vandenberghe, G. Carchon, K. van der Zanden, and B. Nauwelaers, “Easy and accurate empirical transistor model parameter estimation from vectorial large-signal measurements,” *IEEE Int. Microwave Symp.*, pp. 753-756, 1999.

[8] I. Angelov, H. Zirath, and N. Rorsman, “Validation of a nonlinear transistor model by power spectrum characteristics of HEMT’s and MESFET’s,” *IEEE Trans. Microwave Theory Techn.*, pp. 1046-1052, 1995.

[9] D. Schreurs, J. Wood, N. Tufillaro, D. Usikov, L. Barford, and D.E. Root, “The construction and evaluation of behavioral models for microwave devices based on time-domain large-signal measurements,” *IEEE International Electron Devices Meeting (IEDM)*, pp. 819-822, 2000.

[10] M. Kennel, R. Brown, and H. Abarbanel, “Determining embedding dimension for phase-space reconstruction using a geometrical construction,” *Phys. Rev. A*, pp. 3403-3411, 1992.

[11] Q.J. Zhang and K.C. Gupta, *Neural networks for RF and microwave design*, Artech House, 2000.