A Measurement Based Distributed Low Frequency Noise HEMT Model : Application to design of Millimeter Wave Automotive Radar Chip sets

A. Laloue*, A. Lyoubi*, M. Camiade**, J.C. Nallatamby*, M. Valenza***, M. Prigent*, J. Obregon*

*IRCOM, CNRS UMR 6615, IUT, 7 rue Jules Vallès, 19100 Brive, France
**UMS, Domaine de Corbeville, 91404 Orsay, France
*** CEM2, CNRS UMR 5507, University of Montpellier II, Place Eugène Bataillon, 34095 Montpellier Cedex 5, France

Abstract **– A fully measurement based extraction procedure of distributed low-frequency nonlinear noise model of PHEMT is proposed. This model describes accurately the distributed nature under the device gate which allows a good noise behavior prediction in non linear circuits. It is used to simulate successfully noise characteristics of MMICs for FMCW automotive radar at 77 GHz. The simulated and experimental results on two different source-chips : a VCO and DRO have been compared and we demonstrate the accuracy of the noise model which results to be independent of the application.**

I. INTRODUCTION

The problem of the global simulation of microwave circuits driven by real signals arises now at the circuit design level. It is particularly the case of monolithic chip-sets for automotive cruise control radars (ACC radar).

In FET based non linear circuits, inter-modulation frequencies are generated along the non linear channel from source to drain, and are directly responsible for phase noise generation. An accurate non linear noise model and suitable simulation methods are needed to calculate the noise characteristics.

In the first part of this paper, the low frequency noise and non linear distributed model of HEMT is presented. The extracted model and the measurements are compared. The second step is the low frequency noise spectral density measurement at the gate and drain accesses of the transistor. Only drain noise spectral density is distributed into the non linear model by specific procedure. In the PHEMT process used, gate noise current is negligible and simulations have shown that its influence on oscillator AM/PM noise is also negligible.

In the second part, the phase noise spectra of a VCO (transmitter) and a DRO (L.O.) MMIC for FM-CW automotive radar are simulated and compared with very good accuracy to the measured ones.

II. NOISY NONLINEAR DISTRIBUTED FET MODEL

Firstly, we present the non linear distributed model, fully extracted from measurements. In this model, which may be applied to all the field effect devices (MESFET, HEMT and MOSFET) [1], the channel is considered from source to drain as a non uniform, non linear, active transmission line [2]-[3]. Fig. 1 shows the intrinsic part of the proposed model.

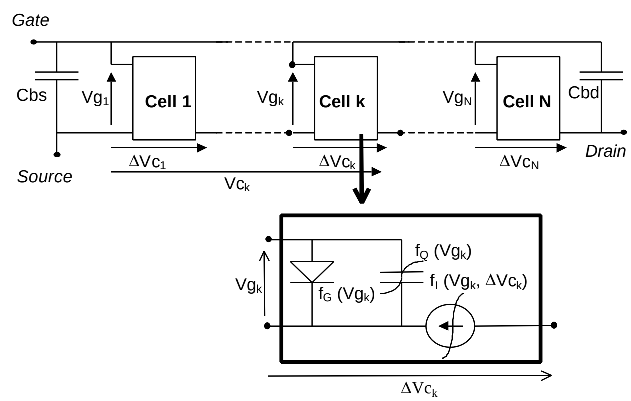

Fig. 1. Topology of the intrinsic distributed model along the gate.

Each cell represents the electrical behavior of a transistor slice. Obviously the number of cells N depends on the gate length L. Practically, we found that for $L \leq 0.3\ \mu m$, N = 10 is a good compromise between accuracy and complexity of this model. Every unit cell includes :

- On one hand : a non linear gate channel capacitance $C = f_C\ (Vg_k)$, in parallel with a schottky diode $I = f_G\ (Vg_k)$. These elements are only function of their own port voltage Vg_k.

- On the other hand a non linear channel current controlled source I = $f_I(Vg_k, \Delta Vc_k)$ which depends on its own port voltage ΔVc_k, and on the controlling voltage of the cell : Vg_k.

At last, two linear fringing capacitances are added at the source and drain end of the channel. They take into account the capacitive coupling effects between the metallic electrodes and the edge channel. The other extrinsic linear parasitic elements are described by lumped elements as usual.

It must be pointed out that the non linear functions f_Q, f_G and f_I are identical for all the unit cells. Nevertheless, the unit cell voltages : Vg_k and ΔVc_k vary along the channel from source to drain, following the rank k of the cell under consideration : from 1 to N. It results an accurate non linear, non uniform, distributed model of the channel.

Moreover, while the model is non quasistatic by nature, it does not need the inclusion of any time-delay, nor a questionable charging resistor Ri [4].

A. Non Linear Distributed Model Extraction Procedure

The extraction procedure is based on pulsed measurement of I-V characteristics and S parameters from 2 GHz up to 40 GHz. The pulsed measurement allows to keep constant temperature and trap steady state during the characterization.

The extrinsic linear parasitic elements are obtained by the same extraction methodology as used for the classical lumped FET model [5].

The Fig. 2 shows a good agreement between whole model made of 10 cells and measured I-V characteristics of 4 fingers of 50×0.25 μm^2 PHEMT process (UMS-PH25).

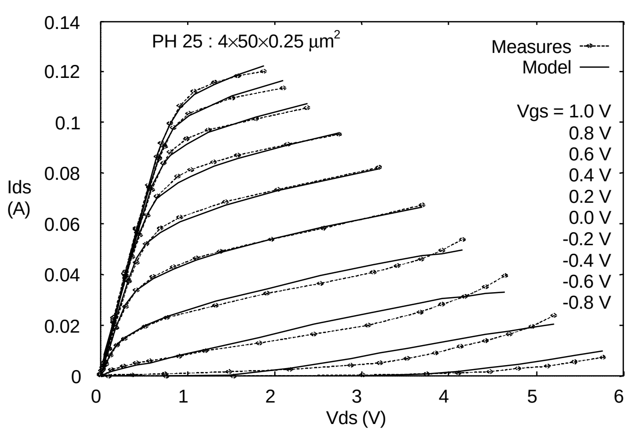

Fig. 2. Modeled and measured I-V characteristics

The non linear function $f_I(Vg_k, \Delta Vc_k)$ of one controlled current source is reported on Fig. 3.

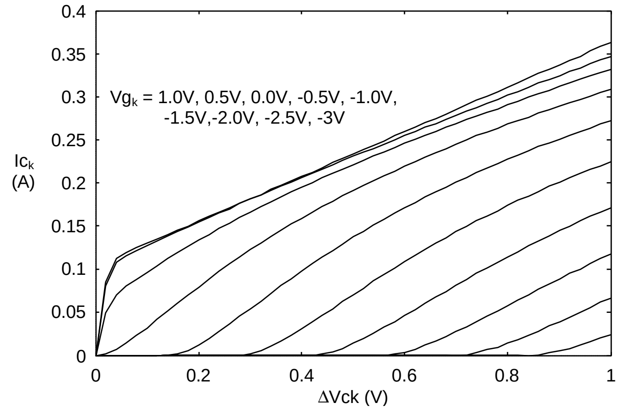

Fig. 3. $Ic_k = f_I(Vg_k, \Delta Vc_k)$

Finally, the N non linear capacitances, C = $f_C(Vg_k)$ and the two edge capacitances Cbs and Cbd are determined from the S parameter measurements. The direct extraction procedure can not be used here, due to the high number of element. Then another software, based on the fast simulated diffusion algorithm is used to obtain the 6 parameter values of the non linear capacitances and the two edge capacitance values. The Fig. 4 presents the modeled and measured S parameters for a pulsed point nearby the bias point :

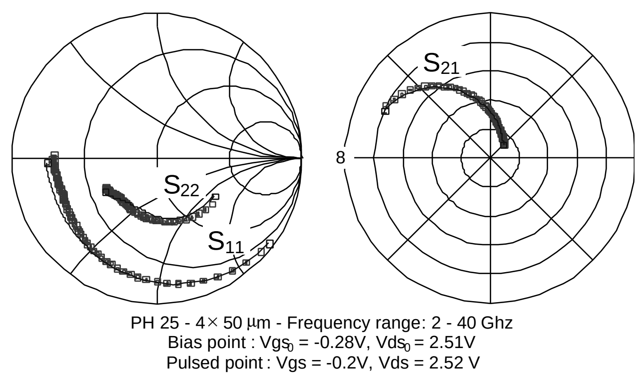

Fig. 4. Modeled and measured S parameters

B. Low Frequency Noise Extraction Procedure

In order to model the low frequency noise sources of the channel, the measurements of the drain noise spectral current density (Sid) for different bias points (Vgs_0, Vds_0) are performed from 1 Hz to 1 MHz frequency range. The setup developed to this purpose is presented Fig. 5 [6].

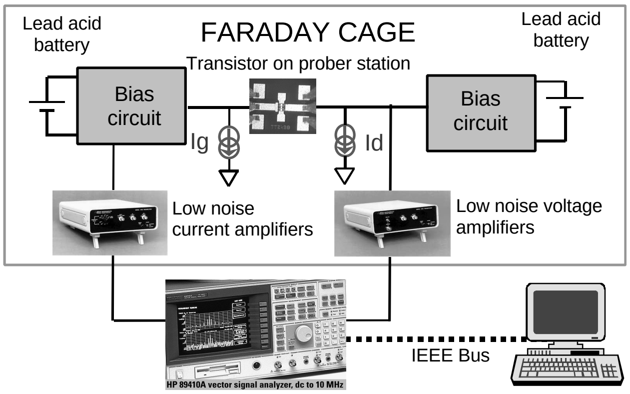

Fig. 5. Low frequency noise measurement setup

By taken into account:

-the parasitic resistances of the FET (Rg, Rd, Rs),

-their noise sources

-the noise generators of the amplifiers (e_i, i_i, e_v, i_v),

we can extract the gate and drain noise current spectra (Fig. 6).

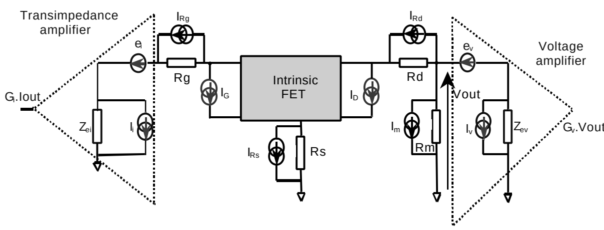

Fig. 6. Low frequency noise measurement setup model

Note that, as explained previously in the introduction, the gate noise source current (i_g) is ten to fifteen orders of magnitude smaller than the drain noise source current (i_d). An example of the extracted Sid spectral density at 100 Hz is shown on Fig. 7.

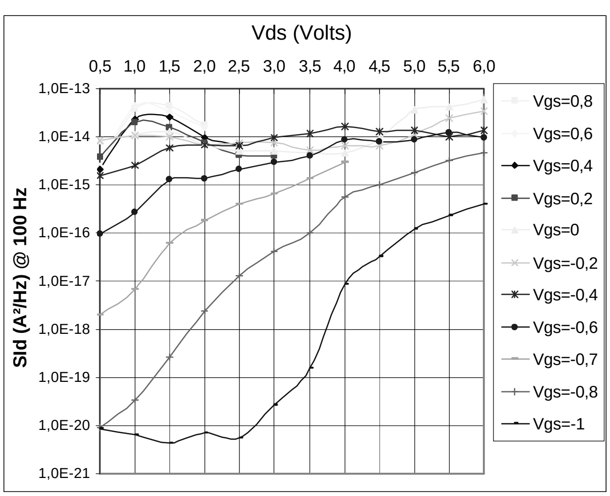

Fig. 7. Drain current spectral density measurement

C. Low Frequency Noise modeling

In each unit cell of the distributed model, an elementary noise source is added, calculated from previous spectral density Sid. These elementary Norton current sources are presumed uncorrelated [8]. At a given noise frequency and bias point, the measured noise spectral density Si_{drain} (Vgs_0 ,Vds_0) can be related to the local noise sources $\langle ic_k^2 \rangle$ as:

$$Si_{drain}\left(Vgs_0, Vds_0\right) = \sum_{k=1}^{N} \left\langle ic_k^2 \right\rangle . h_k^2\left(Vg_k, \Delta Vc_k\right) \quad (1)$$

Where $h_k(Vg_k,\Delta Vc_k)$ are the transfer functions, computed from the linearized model, and k is the number of the cells. To extract the value of every local power spectral density $\langle ic_k^2 \rangle$, a physically based expression, in function of the applied local voltages Vg_k and ΔVc_k, must be defined. We assume that the local low frequency noise sources are proportional to the local electron velocity v_k and the drain current I_{ds0}.

$$\left\langle ic_k^2 \right\rangle = \alpha . v_k\left(\Delta Vc_k\right) . I_{ds_0} \quad (2)$$

where α is a constant which depends on physical parameters of the transistor. The drain side zone of the channel is well-known for its high electric field. Thus the local noise sources have high levels above all in the last cells of the model.

As an example, the Fig. 8 gives the modeled low frequency noise source spectral density of a cell number 9, versus the local voltages Vg_9 and ΔVc_9.

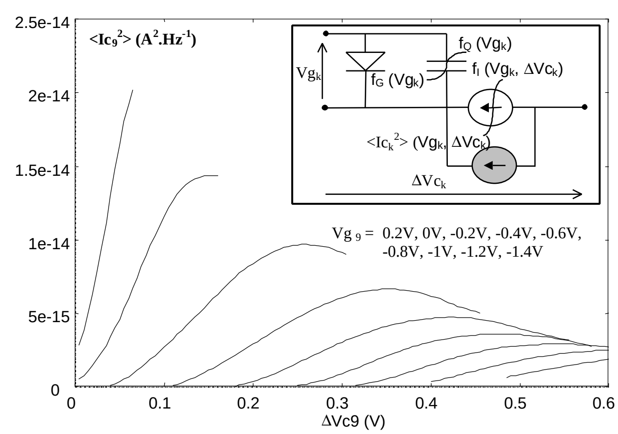

Fig. 8. Local noise source $\langle Ic_9^2 \rangle = f(Vg_9, \Delta Vc_9)$ at 10 kHz of the cell number 9

Note that this representation does not need any explicit correlation coefficient between gate and drain noise currents because it arises in a natural way by means of the distributed along the channel.

III. MMIC Non Linear Noise Simulations

Now we have used the previous distributed model, to compute PM noise spectrum of two MMIC sources of an automotive radar microwave system [7].

A. VCO chip

The oscillator circuit (Fig. 9) is composed of a VCO at 12.75 GHz followed by a frequency tripler. An external medium Q passive resonator gives the center frequency and optimizes phase noise performance. It is based on a micro-strip filter.

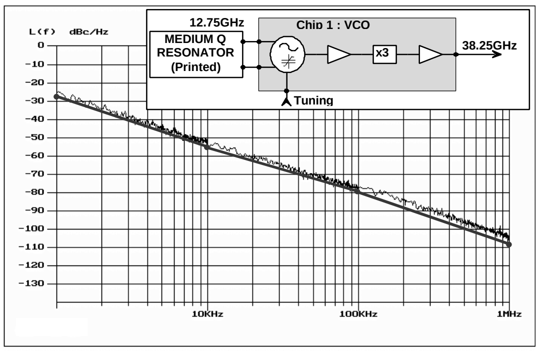

Fig. 9. Measured and computed phase noise at the VCO output at 38.25 GHz for Vtune=0.6V

The distributed models of transistors used in the chip are extracted, following the procedures described in section II.A and II.B. Thanks to conversion matrix analysis [6], the PM noise spectrum is computed and then compared to the measurement results. (Fig. 9)

B. DRO chip

This oscillator chip (Fig. 10) is composed of a high Q resonator centered at 19GHz. With the same transistor models and simulation tools used for the VCO MMIC, the PM noise spectrum is performed. Fig. 10 shows the comparison between the measured and simulated ones. The comparison shows an excellent accuracy between simulations and measurements for the VCO and also for the DRO.

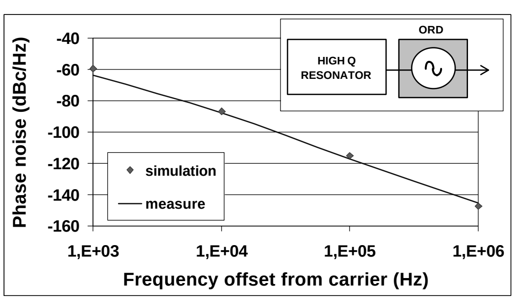

Fig. 10. Measured and computed phase noise of the DRO

IV. Conclusion

A non linear, non uniform transmission line model of HEMT, including low frequency noise sources is presented. Thanks to I-V and S parameter pulsed measurements and a dedicated low frequency noise measurement setup, this low frequency nonlinear noise model is fully extracted from measurement. Due to the robustness of the extraction procedure, it may be applied without any significant modification to other field effect devices. This model efficiency is due to the accurate representation of the distributed nature of the channel from source to drain, which reproduces naturally the non-quasistatic behavior of the transistor.

We showed that noise simulation with the distributed model allows an accurate prediction of the PM and AM noise in microwave and millimeter wave fixed frequency and voltage tuned oscillators. These results validate the proposed modeling.

Acknowledgment

This work was partly supported by the Centre National d'Etude Spatiale (Convention n°714/CNES/98/7305/00) and the European Community Program Esprit 38311 Locomotive.

References

[1] J. Portilla, R. Quéré, J. Obregon, “An improved CAD oriented FET model for large-signal and noise applications”, *San Diego 1994 IEEE MTT-S, pp 849-852.*

[2] D. Sodini, D. Rigaud , “A new approach to model d.c. and a.c. characteristics of junction gate field effect transistors”, *Solid State Electronics, Vol. 24, pp. 635-647, 1981.*

[3] D. Huang, H. C. Lin, “DC and transmission line models for a high electron mobility transistor”, *IEEE Trans. MTT, Vol. 37, No 9, pp 1361-1370, 1989.*

[4] R. Anholt, “Electrical and Thermal Characterization of MESFETs, HEMTs and HBTs”, *Artech House Editor.*

[5] J.P. Teyssier, J.P. Viaud, R. Quéré, “A new Nonlinear I(V) model for FET including Breakdown Effects”, *IEEE MGWL, Vol 2, No 4, pp 104–106, April 1994.*

[6] J. Obregon, et al “Modeling and design of low noise microwave oscillators.”, *IMS 2000 Boston, IEEE MTT-S,Workshop : Microwave Oscillators.*

[7] M. Camiade, D. Domnesque, Z. Ouarch, A. Sion, “Fully MMIC-based front end for FMCW automotive radar at 77 GHz”, *EUMC2000 Symposium, Paris, pp 9-12*

[8] S. Perez, T. Gonzalez, S.L. Delage, J. Obregon, “Microscopic analysis of generation-recombination noise in semiconductors under dc and time-varying electric fields”, *Journal of Applied Physics, vol. 88, No 2, July. 2000, pp 800-807*