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A Direct Coupling between the Semiconductor Equations describing a GaInP/GaAs HBT in a Circuit Simulator for the Co-design of Microwave Devices and Circuits

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***Abstract* : This paper describes the direct coupling between a physical device simulator and a circuit simulator based on the Harmonic Balance (HB) technique. The semiconductor device equations adopted concern a GaInP/GaAs HBT for power applications. A full computation of the Jacobian matrix for convergence improvement has been implemented. It provides us with a powerful tool for the codesign of devices and circuits which has been successfully tested to simulate the power transfer characteristic of a device operating in class AB.**

1- Introduction.

Circuit and device interaction determination appears to be one of the major challenge of mobile communication engineering in the next few years. The ability of designing simultaneously (co-designing) devices and circuits will be a major feature of CAD tools for the design of MMIC circuits [1]. The coupling of circuit simulators and physical simulators is based either on time-domain methods [2] or HB methods [3]. The time-domain approach suffers from drawbacks relative to time-domain methods for the calculation of steady state regimes for microwave circuits and is frequently constrained by the time steps involved in the discretization scheme of device partial derivative equations. This is in contradiction with a circuit simulation approach where the time constants and subsequently the time steps are determined by the external embedding circuit.

In the case of the coupling with HB methods some approaches have been proposed for direct integration of a physical model in nonlinear CAD [4] for FET device.

The approach presented here deals, for the first time, with the integration of physical HBT model in a general circuit simulator. Thus, the popular HB formulation has been adopted in the proposed approach coupled to a **fully implicit discretization scheme** of device equations. The resulting software allows the optimization of circuit performances in terms of physical and geometrical parameter device as well as in terms of terminating impedances. This result has been achieved by making use of dedicated techniques for convergence improvement including the **exact Jacobian matrix calculation** of the nonlinear system that has to be solved.

2- Description of the coupling method.

Physical HBT models are usually based on Drift Diffusion (DD) [5] or Hydrodynamic approximation [6]. In this work, we have implemented a one dimensional DD model described by equations (1), where the basic unknowns are ψ, n, p which represent the electrostatic potential and the densities of carriers.

$$\begin{aligned}
-\nabla_x\left(\varepsilon\cdot\nabla_x(\psi)\right) &= q\left[p-n+N_D-N_A\right]\\
\frac{\partial n}{\partial t} &= \frac{1}{q}\nabla_x J_n + G - R\\
\frac{\partial p}{\partial t} &= -\frac{1}{q}\nabla_x J_p + G - R \qquad (1)\\
J_n &= qn\mu_n E_n + qD_n\nabla_x n\\
J_p &= qp\mu_p E_p - qD_p\nabla_x p
\end{aligned}$$

Our choice has been motivated by two main reasons. On one hand, two dimensional simulations have proved that the charge transport is mainly done in the vertical direction which first justifies the use of a one dimensional model ; on the other hand, for power application the collector thickness is on the order of 1µm or over 1µm, making velocity overshoot effects - only predicted by Hydrodynamic models - less influent on the overall electrical response of the device [7]. This second reason justifies the choice of a DD model.

The circuit simulator is based on the Harmonic Balance Method [8] which is a nonlinear frequency domain analysis technique widely used to simulate harmonic/intermodulation distortion in nonlinear high frequency circuits. This technique allows a rapid calculation of the steady state of the circuit which presents a great interest for mobile communication circuits. Moreover this method constitutes the basis of new algorithms dedicated to the computation of steady state regime for complex modulated signals at the input of the nonlinear circuit [9].

Considering the whole circuit represented in figure 1, the problem is expressed as two set of equations. The first set relates currents and voltages at the **linear** circuit ports in the frequency domain :

$$\mathbf{V}(\omega) - \mathbf{A}_I(\omega)\cdot\mathbf{I}(\omega) - \mathbf{A}_G\cdot\mathbf{E}(\omega) = 0 \qquad (2)$$

where V, I, E are vectors respectively of the nonlinear controlling variables, the nonlinear subcircuit port currents and the linear subcircuit Thevenin voltage source for each study frequency and A_I and A_G are matrix related to the linear subcircuit.

The second set of equations is obtained thanks to equations (1). Indeed the spatial discretization of (1) provides a set of nonlinear time domains equations of the form

$$\mathbf{f}(\psi,\mathbf{n},\mathbf{p},\dot{\mathbf{n}},\dot{\mathbf{p}}) = 0 \qquad (3)$$

where $\psi = \{\psi(1)...\psi(ml)\}$ is the electrostatic potential

$\mathbf{n} = \{n(1)...n(ml)\}$ the electron density

$\mathbf{p} = \{p(1)...p(ml)\}$ the hole density

ml is the index of the last node of the mesh.

$\dot{\mathbf{n}}$, $\dot{\mathbf{p}}$ are the time derivatives of n and p.

The applied terminal bias voltages Ve, Vb, Vc correspond to the boundary conditions on the mesh of the heterostructure

$Ve = \psi(1)$, $Vb = \psi(mb)$, $Vc = \psi(ml)$

However, the knowledge of (Ve, Vb, Vc) or (Jb, Ve, Vc)... and their time derivative fixes in a **bijective** manner ψ, n, p along the heterostructure.

So the HB computation may be done on the mesh variables ψ, n, p as well as on the others "extern" commands Ve, Vb, Vc as chosen in this work and the complementary variables such as (Je, Jb, Jc) or (Vb, Je, Jc) or ...may be determined.

Thus solving equation (3) in the spatial domain leads to an implicit equation relating the voltages and the currents such as

$$\mathbf{f}\left(\mathbf{v(t)},\frac{\mathbf{dv(t)}}{\mathrm{dt}},\mathbf{i(t)},\frac{\mathbf{di(t)}}{\mathrm{dt}}\right) = \mathbf{0} \qquad (4)$$

Equation (3) is then transposed in the frequency domain by means of Fourier transform and the whole system of equations (2) and (3) is solved for the unknowns V(ω) with a Newton Raphson algorithm using an **exact computation** of the Jacobian matrix.

The Jacobian matrix may be expressed as

$$\mathbf{J_{II}} = \mathbf{1} - \mathbf{A_I}(\omega_k)\left[\frac{\partial I_k}{\partial V_l}\right] \qquad (5)$$

where I_k is the k^{th} harmonic of the nonlinearity vector and where V_l is the l^{th} harmonic of the command vector. The computation of the term $\left[\frac{\partial I_k}{\partial V_l}\right]$ is based on a double Fourier Transform (5) of the partial derivative of the nonlineary samples versus command samples :

$u_m(p) = \dfrac{\partial i(p)}{\partial v(p-m)}$ $\quad 0 \le m < N$, $0 \le p < N$ where N is the total number of time samples on a period.We define then

$$U_m(k\text{-}l) = \frac{1}{N}\sum_{p=0}^{N-1} u_m(p)e^{-j\omega_0(k-l)p\Delta t} \qquad (6)$$

which represents the $(k\text{-}l)^{th}$ term of the Discrete Fourrier Transform (DFT) of $u_m(p)$

$$\hat{U}(l) = \frac{1}{N}\sum_{m=0}^{N-1} NU_m(k-l)e^{-jl\omega_0 m\Delta t} \qquad (7)$$

which represents the l^{th} term of the DFT of $NU_m(k - l)$

According to (6) and (7), the nonlinear term of the jacobian matrix expresses as :

$$\frac{\partial I_k}{\partial V_l} = \hat{U}(l) \qquad (8)$$

The exact computation of the jacobian matrix is the key point of our technique as it ensures convergence of the system in very few iterations.

Once the steady state regime is obtained, all the electrical quantities relevant to circuit performances such as waveforms, output power, power gain, power added efficiency can be calculated. We are also able to determine the electrostatic potential, the electric field or the carrier densities inside the transistor.

3- Results.

The previous method has been implemented in our laboratory HB simulator for GalnP/GaAs HBTs manufactured at the Thomson LCR foundry. Those devices have been optimized for power X band applications [10] although they have been used for radiomobile communication. More details about physical and geometrical characteristics will be given in the final paper.

The devices have been used for the optimization of a class AB power amplifier for mobile communication at 1.8 GHz. The HBT device is represented in figure 2. The intrinsic part of the device is represented by the physical simulator whereas the extrinsic one is modeled by lumped elements determined by electrical measurements. The whole transistor is biased through a λ/4 transmission line which allows to short circuit the second harmonic of the collector voltage. Taking into account transmission lines in the time domain would be a cumbersome task. With the actual coupling method, optimization of the amplifier becomes feasible in terms of device characteristics or embedding impedances for a particular device as the total CPU time required for the simulation of the steady state regime takes less than 20 min per power point using an HP712/60 workstation.

Figures 3 and 4 present first a comparison between simulated and pulsed measurements characteristics for a one finger Thomson 2μm * 30μm HBT. We can see a good agreement between measurements and models which first validate our one dimensional DD approach. Then we present in figure 5 the results of the optimization of an 1.8 GHz amplifier in terms of embedding impedances for the fabricated device. Collector current waveform as well as the load line in the Ic-Vce plane of the device are shown in figure 6 and 7. Figure 7 exhibits also the current excursion in the Ic-Vce plane. Figure 8 shows the output power versus the input one. Three points have been labeled in figure 7 corresponding to three different functioning areas. We have also represented in figures 9 to 11 for these points respectively the electron, the hole densities and the electric field along the heterostructure. This work has been dynamically realized. It means that all these quantities have been computed dynamically as a function of all the command samples all over the period. Figures 9 to 11 show that in the A point, the HBT begins to present some Kirk effect. The electric field slope inverts in figure 11. This corresponds to the increase of the electron density over the collector doping in figure 9 and a base push out in the collector in figure 10.

4- Conclusion.

For the first time an efficient coupling technique between HBT device and HB circuit equations has been proposed for the codesign of microwave devices and circuits. The use of device equations instead of lumped or analytical models for the active device make this approach as a predictive one. This technique has proved to be very efficient for calculation of power amplifiers for mobile communication. Moreover it has been shown to require only reasonable amounts of CPU time. It will now be extensively used for optimization of devices in their circuit environment and the extension to two dimensional and hydrodynamic models is currently under progress.

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Figures

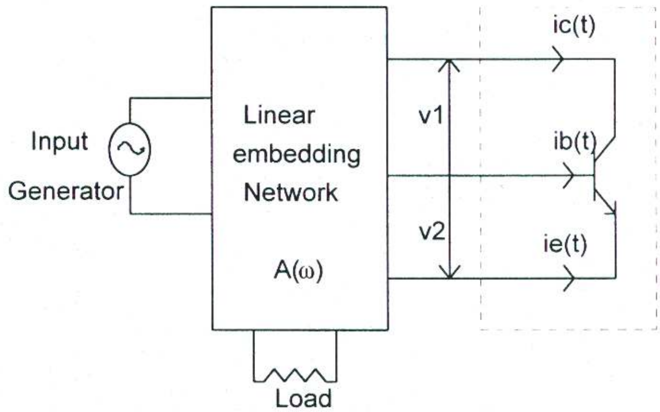

Figure 1 : Topology of Simulated Circuit

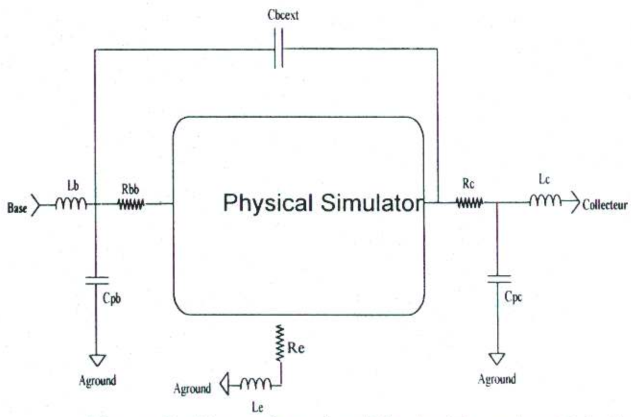

Figure 2 : Heterojunction Bipolar Transistor Model

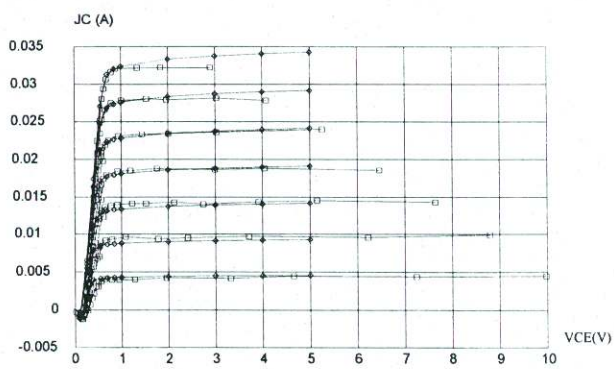

Figure 3 : Simulated and measured output characteristics Ic(Vce)

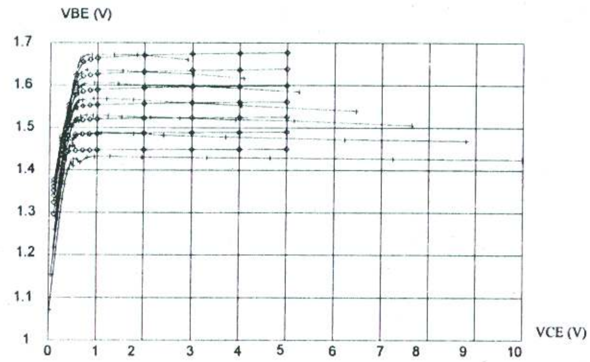

Figure 4 : Simulated and measured input characteristics Vbe(Vce)

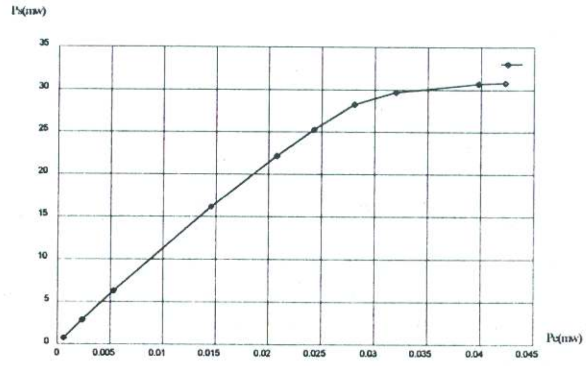

Figure 8 : Output power versus input power

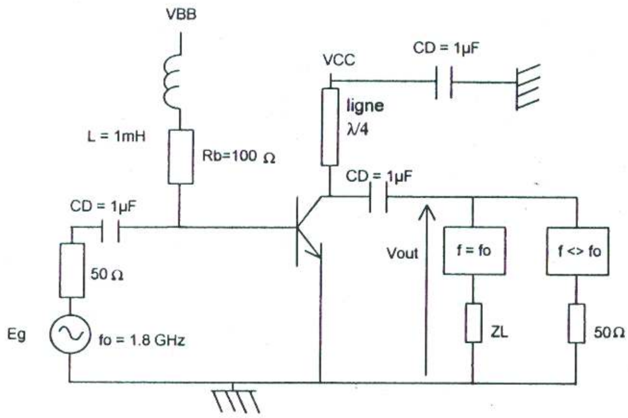

Figure 5 : Simulated Circuit

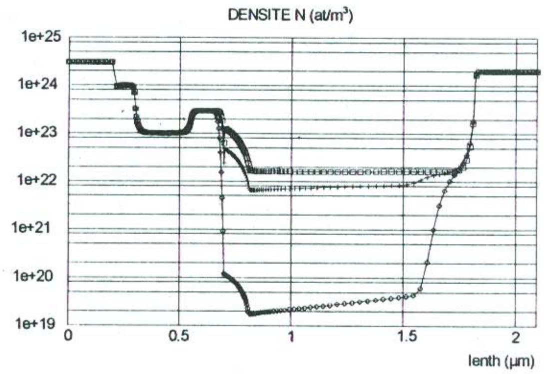

Figure 9 : Electron density

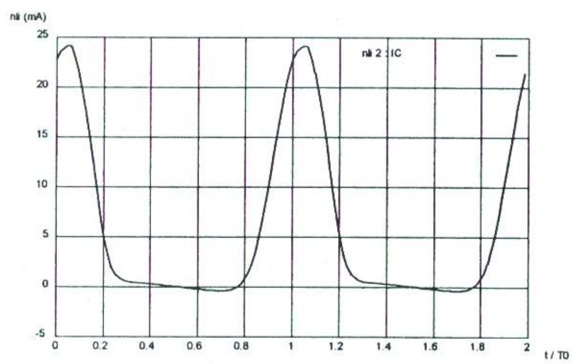

Figure 6 : Collector current waveform Ic(t)

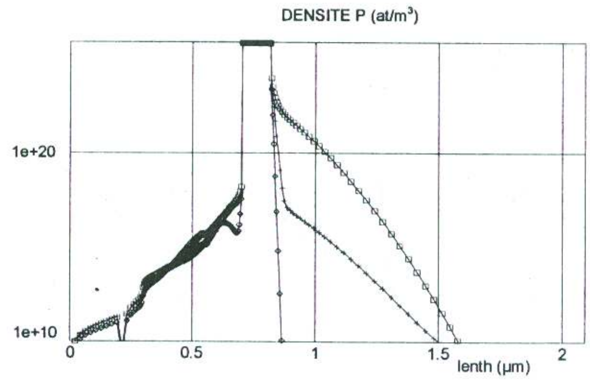

Figure 10 : Hole density

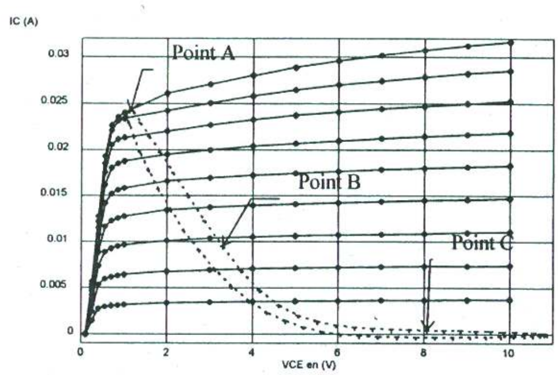

Figure 7 : Load line Ic = f(Vce)

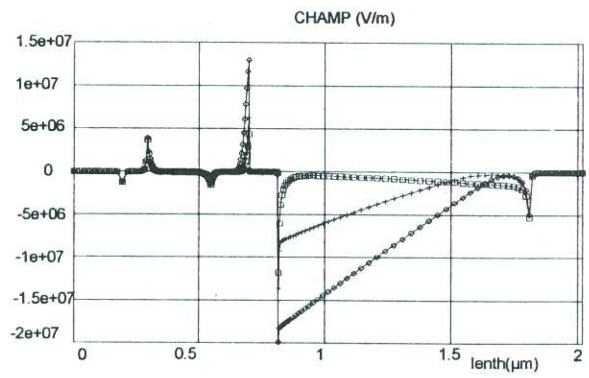

Figure 11 : Electric field