PHYSICAL SIMULATION OF COMPLETE MILLIMETER-WAVE AMPLIFIERS USING FULL-WAVE FDTD TECHNIQUE

S. M. Sohel Imtiaz, *Student Member, IEEE*, Samir M. El-Ghazaly, *Senior Member, IEEE*

Department of Electrical Engineering, Telecommunications Research Center
Arizona State University, Tempe, AZ 85287-7206

ABSTRACT

In this paper, the characterization of high frequency microwave amplifiers using a full-wave analysis coupled with physical modeling of the semiconductor device is presented. The simulation includes the input and output matching networks and the transistor as well. The entire amplifier is simulated with FDTD algorithm which also solves for the electromagnetic fields inside the transistor. The frequency dependence of the scattering parameters for the amplifier is presented.

I. INTRODUCTION

With the advancement of semiconductor devices, the state of the art techniques are developed for their analysis, design, and optimization. After having a stable and accurate device simulator, the characterization of microwave amplifiers using the submicron semiconductor devices, becomes important. The currently available amplifier design and modeling techniques mostly deal with the equivalent circuit parameters or the scattering parameters of the transistors and couple them with the input and the output matching networks. This process of replacing the active device with the equivalent sources or the equivalent circuit parameters, is unable to represent the nonlinearity of the device accurately at microwave and millimeter-wave frequencies. The alternative approach is to simulate the whole amplifier by coupling the physical equations representing the semiconductor device with the electromagnetic fields in the transmission lines.

We have already developed a combined electromagnetic and solid-state (CESS) simulator for the analysis of high frequency MESFETs [1]. The CESS simulator couples a semiconductor model to the 3D time-domain solution of Maxwell's equations using FDTD algorithm. The semiconductor model is based on the moments of the Boltzmann transport equation. The CESS model can predict the nonlinear energy build-up inside the transistor. The other advantage is it's ability to show the dispersive nature of the device, specially at high frequencies. The goal of the present work is to simulate a microwave amplifier with the passive matching networks using the CESS model. The simulation of a complete amplifier using FDTD method requires intensive computer memory and consumes a considerable amount of time. The problem of large computer memory and time can be reduced by breaking the amplifier circuit into active and passive parts, then model the passive parts separately using the FDTD technique and combine them with the full-wave simulation of the transistor. The amplifier network was previously modeled by Kuo et al. [2]. However, the physical transistor was not modeled in this case. An equivalent voltage source approach was used. Breaking a large circuit into sub-circuits was done by using the time-domain diakoptics method [3] and by using the discrete time-domain Green's function [4].

The microwave amplifier consists of input and output matching networks and a transistor. The input and the output matching networks are very large compared to the transistor. But the mesh size and, consequently, the FDTD stability criteria for the amplifier are severely limited by the debye length of the semiconductor. This imposes a constraint on the time step qt of the FDTD algorithm, which becomes on the order of 10^{-17} seconds. If the input and the output matching networks were to be simulated using this criteria, it would take so long time to simulate the amplifier circuit that it becomes almost impossible using the current computers.

In this work, a full-wave analysis is performed to simulate the microwave amplifier with two tuned coplanar wave guides (CPW) as matching networks. The whole amplifier is divided into three regions. The simulation of each region is performed individually and coupled to the next stage properly with all the required informations from the preceding stage. This technique enables one to use large space step, and hence, large time step in matching networks. The computer simulation time is reduced drastically compared to the method incorporating the non-uniform mesh for the whole amplifier. The computer memory requirement is also lowered by approximately 66 % at a certain time. The amplifier analysis, the coupling procedure and the results are summarized in the following sections.

II. ANALYSIS OF THE AMPLIFIER

0-7803-4603-6/97/$5.00 (c) IEEE

To demonstrate the potential of this approach, a millimeter-wave amplifier implemented in a coplanar waveguide configuration is analyzed. The amplifier along with the matching networks is shown in Fig. 1. The input and the output matching networks are designed on the basis of the characteristics of the transistor (Fig. 2). The transistor channel length and aspect ratio are selected to achieve the required transconductance and the cut-off frequency. The cut-off frequency of the device was estimated as 67 GHz. The operating frequency is chosen to be 40 GHz to illustrate the principle of the approach. The scattering parameters are computed for the MESFET. The equivalent circuit admittances of the MESFET are obtained from the simulated s-parameters.

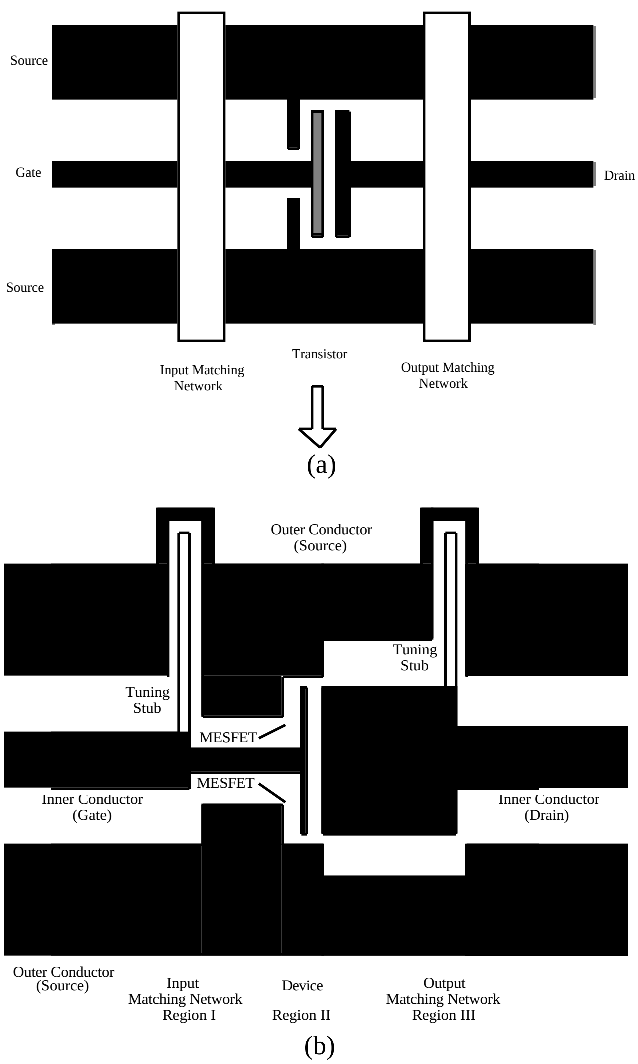

Fig. 1. GaAs transistor amplifier with CPW matching networks (a) Schematic (b) Detailed diagram.

The dimensions of the center conductor and the outer conductor of the matching networks are selected to achieve a characteristic impedance of about 50 ohms and to connect them perfectly at the input and the output sides of MESFET. The characteristic impedances of the input and the output matching networks are shown in Fig. 3 as a function of frequency. The line and the stub lengths of the matching networks are designed from the source and the load admittances of the MESFET.

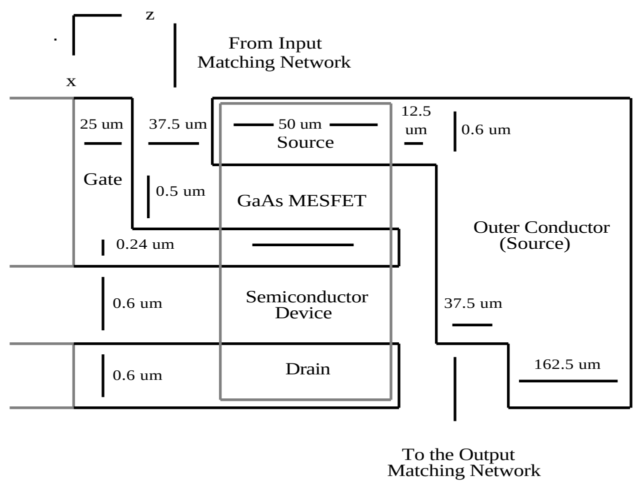

Fig. 2. The top view of one half of the GaAs MESFET structure.

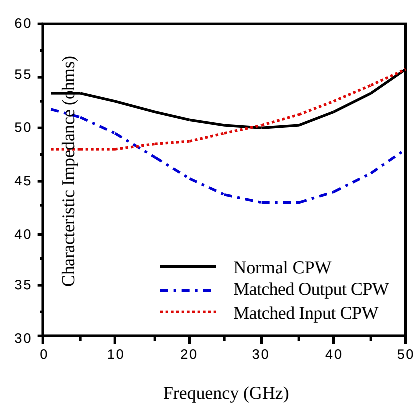

Fig. 3. The characteristic impedance of the input and the output matching networks.

III. COUPLING THE MATCHING NETWORK AND MESFET

The amplifier is divided into three regions. The simulation of each region is performed separately and coupled to the next stage properly with all the required informations of the preceding stage. At the input of Region I, an impulse I_1 is applied. The electromagnetic model is solved for the input matching network using the FDTD algorithm. The response R_2 is collected at point B as shown in Fig. 4. The reflection co-efficient Γ_1 of Region I

0-7803-4603-6/97/$5.00 (c) IEEE

is obtained by applying the same impulse at point B. When the impulse is incident at B, part of it is transmitted and the rest is reflected back. The reflected wave is identified and the time dependent Γ_1 is obtained.

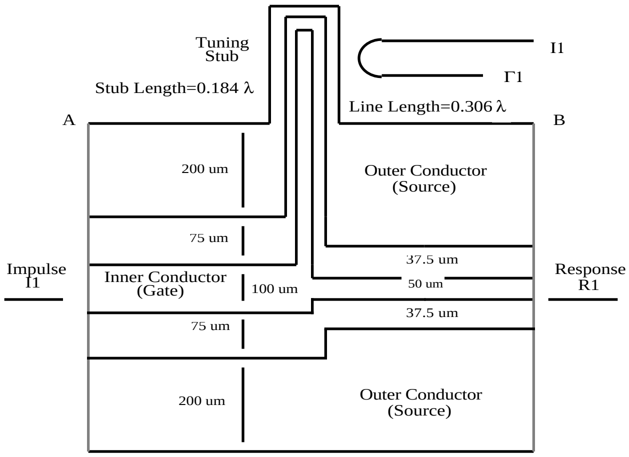

Fig. 4. The input matching network.

A sinusoidal wave is applied to the amplifier input. The response V_2 to the sine wave can be obtained by summing up the impulse responses convoluted in time with proper magnitudes. The signal V_2 is now applied to the transistor as shown in Fig. 5. The combined electromagnetic and solid-state simulator is used to propagate this signal. The reflected wave V_{R1} from the MESFET is identified and convoluted with the reflection co-efficient Γ_1 and the resulting signal is also added to V_2. Thus at point C the applied ac voltage is

$$V_C(t) = V_2(t) + \Sigma\, \Gamma_1(\tau).V_{R1}(t-\tau) \qquad (1)$$

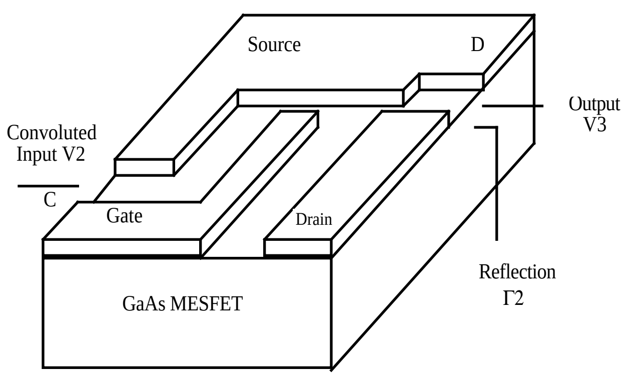

Fig. 5. Coupling in GaAs MESFET.

The reflection co-efficient Γ_2 for an impulse I_2 at the output matching network is obtained (Fig. 6). Γ_2 is convoluted with the output of MESFET and is applied as a secondary source at point D. The output wave V_3 is collected in between the drain and the source at point D.

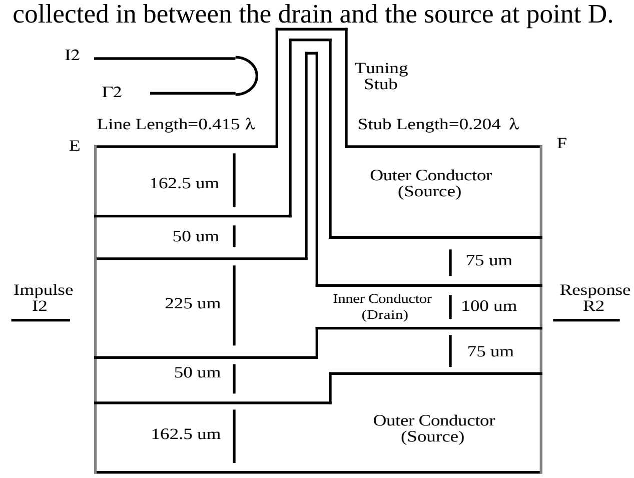

Fig. 6. The output matching network.

In Region III, as stated earlier, an impulse I_2 is applied at input E of the output matching network as shown in Fig. 6. The electromagnetic model is solved in this region. The response R_2 is collected at point F. The wave V_3 can be decomposed to number of narrow pulses. Thus the output wave at point F is calculated by convoluting the impulse response R_2 with proper magnitudes.

In this demonstration, sinusoidal signals are applied to the amplifier input. It should be noted that this is done to illustrate the concept. In general, any arbitrary wave form can be applied to the amplifier without any significant increase in the computational effort.

IV. RESULTS AND DISCUSSIONS

In the transistor portion of the amplifier, only one half of the structure is simulated considering the symmetry. The transfer functions of the input and output matching networks are shown in Fig. 7.

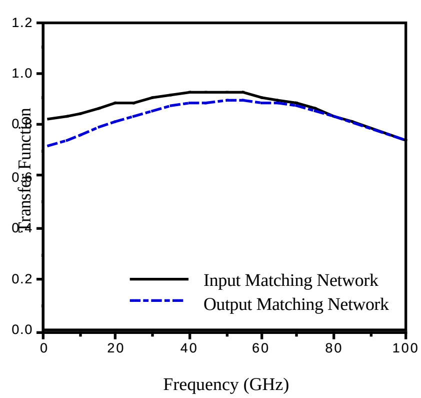

Fig. 7. The transfer functions of the input and output matching networks.

The scattering parameters are obtained at different frequencies. In Fig. 8, S_{21} is shown as a function of frequency. The magnitude of S_{21} remains within a certain range for the frequency band of 20 - 60 GHz. S_{21} has its peak value at the design frequency which is expected. The value of S_{21} for this amplifier is low because the transistor width is very small compared to any commercial amplifier. However, increasing the aspect ratio would increase the value of S_{21}. In Fig. 9, the variation of S_{12} is shown with frequency. The magnitude of S_{12} decreases with frequency. The value of S_{12} is small which is also anticipated. The corresponding variation of S_{11} and S_{22} are presented in Fig. 10. The return loss is less than 25 dB at 40 GHz.

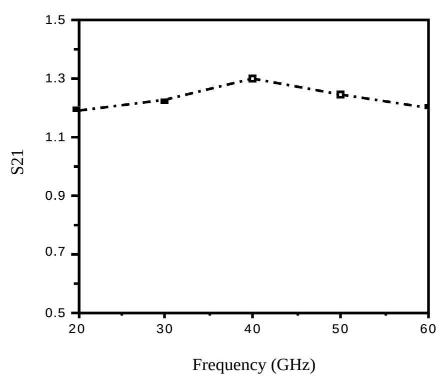

Fig. 8. The dependence of S_{21} on frequency.

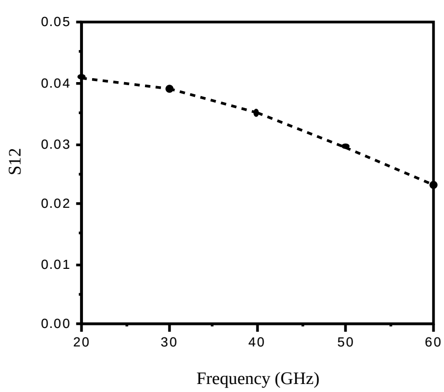

Fig. 9. The variation of S_{12} with frequency.

It should be noted that no attempts were made to optimize the amplifier performance in this paper, since the aim is to report on the technique rather than on the amplifier itself.

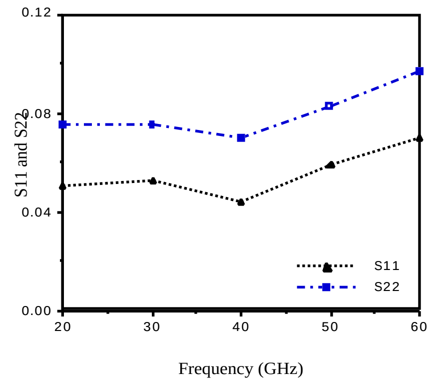

Fig. 10. The dependence of S_{11} and S_{22} on frequency.

V. CONCLUSION

The characterization of microwave amplifier is performed at high frequencies using a full-wave approach. The amplifier consists of matching networks and a transistor. The electromagnetic characteristics of the passive parts are simulated using the FDTD technique. The characteristics of the active part are modeled by coupling the FDTD solution to a physical model for the transistor. The transfer functions of the input and the output matching networks are presented. The scattering parameters are extracted for the entire amplifier for a frequency band of 20 - 60 GHz.

ACKNOWLEDGMENT

This work is supported by the Army Research Office under contract # DAAH04-95-1-0252.

REFERENCES

[1] M. A. Alsunaidi, S. M. Sohel Imtiaz, and S. M. El-Ghazaly, "Electromagnetic wave effects on microwave transistors using a full-wave time domain model," *IEEE Trans. Microwave Theory and Tech.*, vol. 44, pp. 799-808, June 1996.

[2] C. -N. Kuo, R. -B. Wu, B. Houshmand, and T. Itoh, “Modeling of microwave active devices using the FDTD analysis based on the voltage-source approach,” *IEEE Microwave and Guided Wave Lett.*, vol. 6, pp. 199-201, May 1996.

[3] P. B. Johns and K. Akhtarzad, “The use of time domain diakoptics in time discrete models of fields,” *Int. J. Num. Meth. Engg.*, vol. 17, pp. 1-14, 1981.

[4] W. J. R. Hoefer, “The discrete time-domain Green’s function or Johns matrix - A new powerful concept in transmission line modeling (TLM),” *Int. J. Num. Meth. Engg.*, vol. 2, pp. 215-225, 1989.