A NOVEL ANALYTICAL APPROACH FOR THE NONLINEAR MICROWAVE CIRCUITS AND EXPERIMENTAL CHARACTERISATION OF THE NONLINEAR BEHAVIOUR OF A NEW MESFET DEVICE STRUCTURE

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ABSTRACT

A novel analytical technique is described based on generalised Volterra series to analyse nonlinear microwave and millimeter-wave circuits and devices. In contrast to previous publications the new method is especially efficient for general-purpose CAD applications and can be easily incorporated into existing CAD programs. The capabilities of the technique has been demonstrated on especially fabricated new MESFET structures, designed to reduce the losses due to the high resistance of the gate electrode. Power and intermodulation distortion measurements has been carried out and good agreement with calculated values has been observed.

INTRODUCTION

Distributed concepts are increasingly utilised for modeling MESFET devices at millimeter-wave frequencies and for the design of wide-band power amplifiers, but they create a severe problem for accurate nonlinear performance characterisation. Present available CAD nonlinear analysis techniques for microwave and millimeter-wave devices and circuits employ the conventional Volterra series analysis /1-3/ or the harmonic-balance method /4-8/. These methods are either analytically complex and topologically inefficient or require a large amount of computer time and are cumbersome to implement into existing CAD programs.

This paper therefore describes results concerning a highly efficient nonlinear analysis of MESFET devices and circuits. Particularly, theoretical and experimental characterisation of the intermodulation distortion (IMD) behaviour of an especially fabricated MESFET device structure is considered. The generalised Volterra series analysis /10,11/ is used and a new set of nonlinear S parameters is defined suitable for nonlinear analysis. Based on the nonlinear S parameters structural modeling of nonlinear microwave circuits with arbitrary topology is possible. Also an arbitrary number of ports can be considered in the analysis, which makes this method most suitable for general-purpose CAD programs. This new analysis technique outperforms conventional approaches because it is easily incorporated into existing microwave CAD packages and is able to deal with arbitrary circuit structures.

NONLINEAR SCATTERING PARAMETERS TECHNIQUE

When analytical results are required, a very promising method for analysing nonlinear microwave circuits is commonly accepted to be that of the Volterra series representation /1-4,9-11/. Although, considerable efforts have been undertaken in the field of Volterra series representation, the studies have been limited to scalar nonlinear systems with fixed topology.

However, it has been shown /9-11/ that the Volterra series representation can be generalised to vector nonlinear systems and hence can overcome the above mentioned tradeoffs. We can write the Volterra series expansion of the response of a nonlinear vector system in the frequency domain having an input vector $\mathbf{U}(f)$ and an output vector $\mathbf{Y}(f)$ as

$$\mathbf{Y}_n(f_1,\ldots,f_n) = \mathbf{H}_n(f_1,\ldots,f_n)\cdot(\mathbf{U}(f_1) \mathbf{x} \ldots \mathbf{x}\, \mathbf{U}(f_n)) \qquad (1)$$

where $\mathbf{U}(*)$ and $\mathbf{Y}_n(*)$ are k and m dimensional vectors respectively and $\mathbf{H}_n(*)$ is a $(m x k^n)$ dimensional transfer-matrix function. The product sign "**x**" denotes the left Kroneker product of matrices /12/. If we set m=k=2 in eq.1 and utilise the harmonic current method/1,9,11/ then we can determine the n-th order multidimensional scattering parameters (MSP)

$$\mathbf{b}_n(f_1,\ldots,f_n) = \mathbf{S}_n(f_1,\ldots,f_n)\cdot(\mathbf{a}(f_1) \mathbf{x} \ldots \mathbf{x}\, \mathbf{a}(f_n)) \qquad (2)$$

where $\mathbf{b}_n(*)=(b_1(*),b_2(*))^T$, $\mathbf{a}(*)=(a_1(*),a_2(*))^T$ and $\mathbf{S}_n(*)$ is a $2x2^n$ dimensional matrix. Specifically, the first order MSP of eq.2 reduce to the known S-parameters. For the second order we can write:

$$\mathbf{b_2}(f_1,f_2) = \mathbf{S_2}(f_1,f_2)\, x \qquad (3)$$
$$x\, [a_1(f_1)a_1(f_2), a_2(f_1)a_1(f_2), a_1(f_1)a_2(f_2), a_2(f_1)a_2(f_2)]^T$$

where $\mathbf{b}(*)$ and $\mathbf{a}(*)$ have the obvious meaning. When dealing with nonlinear microwave circuits we have to devide the circuit under consideration into linear and nonlinear subcircuits. Then with the knowledge of the incident wave variables $\mathbf{a}(*)$ to the nonlinear device we can determine the overall response of the circuit to any desired order utilising the relationships for cascaded, series or paralell connection of nonlinear devices.It has to be emphasized that the choice of the S-parameters above is performed to demonstrate the unique possibilities of the new method for CAD applications. Naturally, any different parameters as Z- or Y-parameters can be chosen for convenience.

The method of nonlinear S-parameters does not only give the correct results for the power, gain and intermodulation distortion (IMD) characteristics of conventional two-port devices, but is also able to handle more complicated phenomena accounted for in distributed MESFET

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devices and circuits with more than two ports. Furthermore, the approach indicated above, allows the correct prediction of the S parameters at the onset of large-signal operation and is therefore of paramount importance in the CAD of microwave oscillators, mixers etc /13/. This can be verified if we consider single frequency excitation. Then in the large-signal case we obtain for the modified S-parameters:

$$(4)\quad \mathbf{b}(f) = \mathbf{S}_1(f)\cdot\mathbf{a}(f) + 3/8\ \mathbf{S}_3(f,f,-f).(\mathbf{a}(f) \times \mathbf{a}(f) \times \mathbf{a}^*(f)) + 3/8\ \mathbf{S}_3(-f,-f,f).(\mathbf{a}^*(f) \times \mathbf{a}^*(f) \times \mathbf{a}(f))$$

Eq.(4) the second and third term give a better approximation to the large-signal S-parameters. These parameters are useful up to the onset of the saturation region of the device.

DEVICE CHARACTERISATION AND RESULTS

To demonstrate the capabilities of the proposed analytical technique we have manufactured a MESFET device incorporating a new device structure. With this structure it is possible to reduce the gate resistance while simultaneously keeping the gate width large and the input impedance to the MESFET and the power-bandwidth product at moderate levels. This new structure is particularly important in millimeter-wave power applications, where the losses on the gate line become the dominating factor in MESFET performance deterioration /14/.

A schematic drawing of the MESFET device and an SEM photograph are shown in fig.1. The improved structure exhibits one gate finger with 2µm gatelength and gatewidth between 400µm and 2mm. The comparatively large gatelength has been only chosen for convenience in measurements. In order to decrease the losses on such large gatewidth we have incorporated a transmission line in parallel to the gate with gate feeders at every 200µm of the gate line. The source pads have been grounded with the aid of via holes, which additionally yield a very low thermal resistance for the device. The gate transmission line parallel to the gate finger as well as the drain line are accessable from both ends. Therefore, a new MESFET test fixture has been constructed allowing for four-ports characterisation of the MESFET device structure.

In order to correctly model the device, a "sliced" transistor model, fig.2, has been chosen as is common practice at millimeter-wave frequencies /16,17/. Similar practice is used when dealing with distributed amplifiers /14,15/. In the calculations we have assumed that each "slice" in the MESFET model is connected by a transmission line of length 200µm representing the parallel gate line and the drain contact. Both ends of the drain contact have been found to be electrically short-circuited at the operating frequency of approximately 3.9GHz. Measurements have been performed on all MESFET structures and those for 400µm and 2mm gatewidth devices are presented here. First, the theoretical and experimental values for the output power at the fundametal and IMD frequencies as a function of input power has been illustrated in fig.3 for the 400µm device. The MESFET with this gatewidth can be still considered as a lumped device at the operating frequency. Good accuracy is obtained well into the

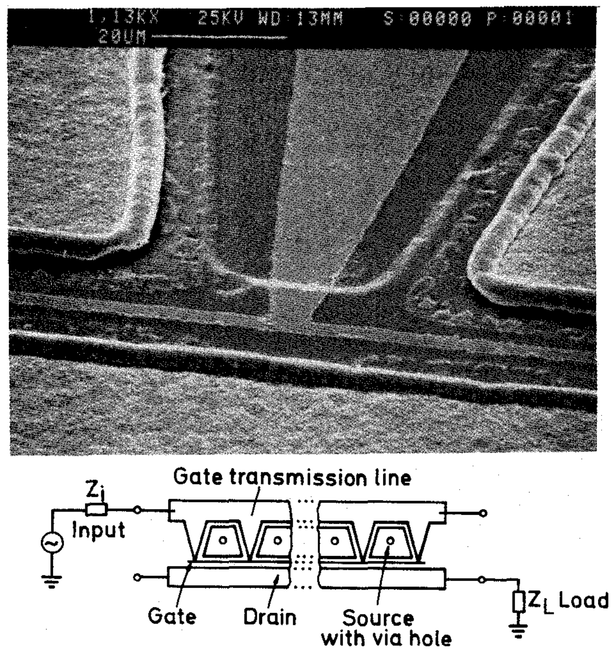

Fig.1: **a**:Schematical drawing of the transistor structure
b: SEM photograph of a detail of the transistor structure

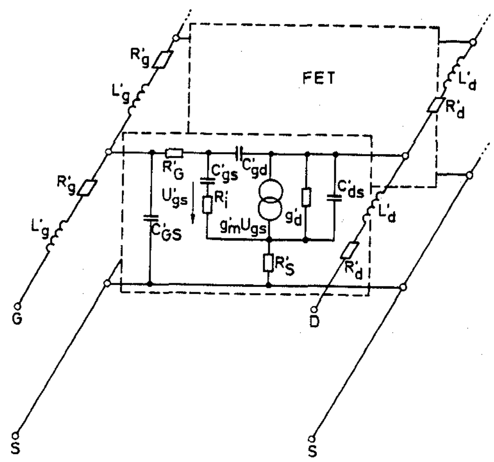

Fig.2: The sliced transistor model taken in calculations

saturation region for the fundamental frequency. However, at very large input powers the calculated values tend to diverge from the measured values. Subsequently, all four ports has been taken into account to demonstrate the capabilities of the analysis method proposed above.

Theoretical and experimental results for the maximum available gain (MAG) and IMD for the MESFET with 2mm gatewidth are indicated in fig.4 and 5. First, the transistor has been matched in the conventional manner and subsequently the tunable load at the otherwise free end of the gate line has been varied without loss of matching. The variation of the tunable load is equivalent to the variation of the

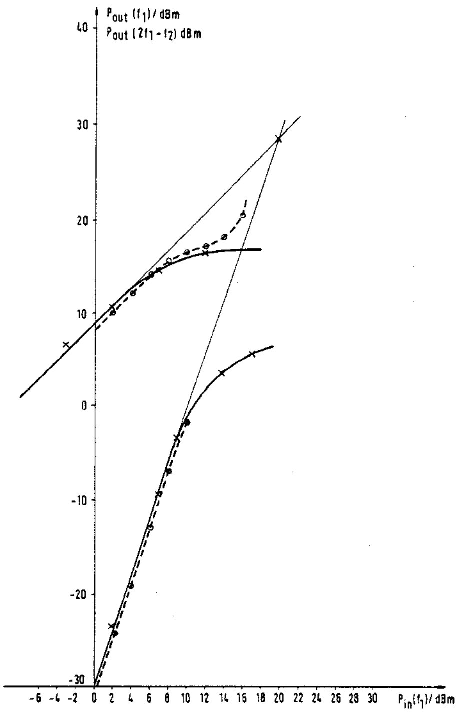

Fig.3: Calculated (broken line) and measured (solid line) of the output power versus input power at the fundamental and inter modulation frequency for the 400µm wide MESFET

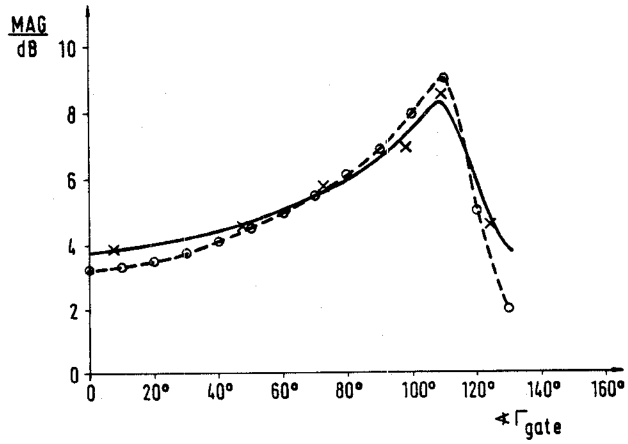

Fig.4: Calculated (broken line) and measured (solid line) of the maximum avalilable gain versus the phase of the reflec tion coefficient at the end of the gate of the 2mm wide MESFET

phase of the reflection coefficient at the according port. Fig.4 determines the increase of MAG due to the variation of the phase of the reflection coefficient at the end of the gate line (Γ_{gate}). We can infer from fig.4 that leaving the end of the gate open gives not optimal performance. We found theoretically and experimentally that an inductive load at the end of the gate can improve the gain performance of the device considerably. The calculated values overestimate the improvement in MAG due to variations of the phase of Γ_{gate}. This is probabely due to the parasitic capacitive and inductive coupling which has not been taken into account in the calculations and due to higher losses in the real transistor structure than those assumed in the calculations. In fig.5 we find the curves for the output power at the fundamental and IMD frequency versus the input power with Γ_{gate} as a parameter. The theoretical values obtained for optimum phase of Γ_{gate} are also higher than the measured ones in fig.5 but the variation of the output power at the fundamental and IMD frequency is still well approximated.

Finally it has to be emphasized that the increase in MAG due to the appropriate termination at the free end of the gate line is achievable only at low to moderate input power levels At this value we could obtain an improvement of 100% in MAG and 200% in power-added efficiency as compared to the open end case with an output power of 1W at 4GHz.

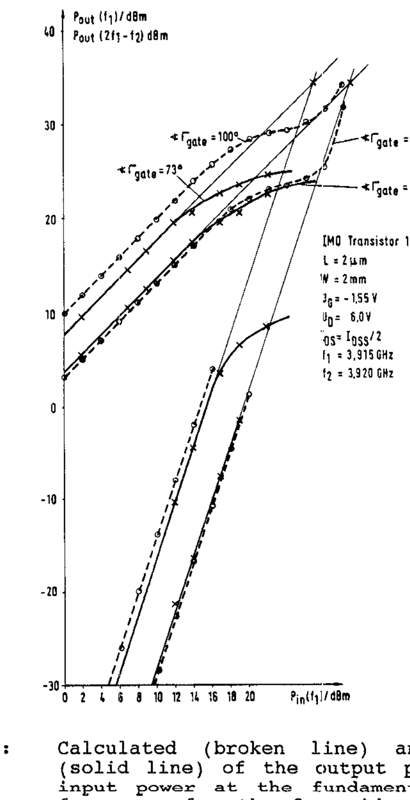

Fig.5: Calculated (broken line) and measured (solid line) of the output power versus input power at the fundamental and IMD frequency for the 2mm wide MESFET with the phase of the reflection coefficient at the end of the gate as a parameter

CONCLUSIONS

A novel nonlinear analysis technique has been proposed which is especially efficient for CAD applications. The new method relies on the generalised Volterra series representation. Nonlinear S- parameters can be defined using this technique in order to accurately predict circuit performance or device S parameters at the onset of large-signal excitation. This method is perfectly suited for distributed circuits and devices operating at millimeter-wave frequencies. Calculated results have been compared with experimental data taken from an especially fabricated MESFET. The MESFET employed in the characterisation and measurements has a unique design to reduce the attenuation losses on the gateline without decreasing the gatewidth and the input impedance of the device. When the MESFET is operated as a four-port an additional improvement in MAG and power-added efficiency can be obtained when the otherwise free end of the gate line is terminated in an inductive load. With the appropriate termination it is also possible to increase the power-bandwidth product.

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