"Characterizing Components Under Large Signal Excitation: Defining Sensible 'Large Signal S-Parameters'?!"

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Characterizing Components Under Large Signal Excitation: Defining Sensible "Large Signal S-Parameters"?!

Jan Verspecht, Marc Vanden Bossche and Frans Verbeyst

Hewlett-Packard Network Measurement and Description Group
VUB-ELEC, Pleinlaan 2, 1050 Brussels, Belgium
tel. 32-2-629.2886, fax 32-2-629.2850, email janv@belgium.hp.com

Abstract - A measurement and black-box modeling technique is described enabling the characterization of nonlinear microwave components under periodic large-signal excitation. First, the mathematical model is theoretically described. The model is based on the assumption that the superposition principle holds for the effect of all spectral components, except the fundamental, of the incident travelling voltage waves. This assumption implies that, with fixed fundamental power and biasing conditions, the incident harmonics interact with the component as if it is a linear time-varying circuit. Since the superposition coefficients are a natural extension of the classical scattering parameters they are called "large signal s-parameters". These coefficients are a function of fundamental power and biasing conditions. An automated set-up is described enabling to accurately measure these coefficients. The set-up is based on a "vectorial nonlinear network-analyzer", which accurately measures the phase and amplitude of all spectral components of both incident and reflected travelling voltage waves. The experimental model extraction method is illustrated on a heterojunction bipolar and a field effect transistor, driven hardly nonlinear. The "vectorial nonlinear network-analyzer" is used in order to successfully verify the validity of the extracted model. The black-box model is finally integrated in a commercial harmonic balance simulator.

Introduction

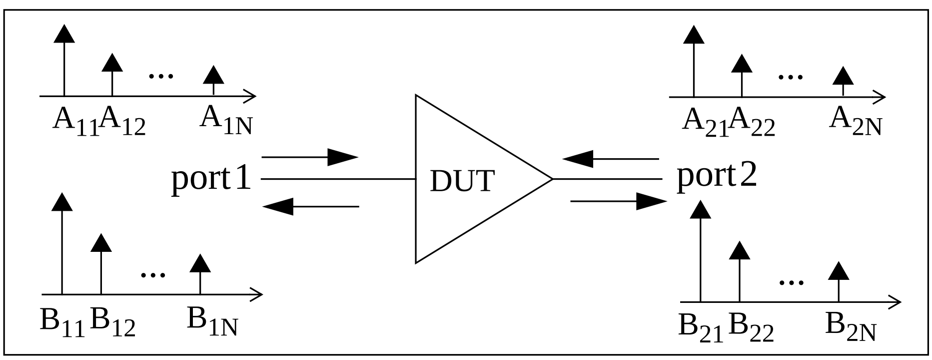
Nowadays, many microwave designers use sophisticated software for their circuit design. Especially when the designer is confronted with hard-nonlinear behavior, e.g. the design of a power amplifier, simulators [1] provide valuable information concerning a circuit behavior before it is even built, thereby saving a lot of time and money. The usefulness of any simulator, however, relies very much on the accuracy of the mathematical device models that are used. For linear devices, accurate models are readily available (e.g. measured scattering parameter matrix, or geometry based models). For nonlinear devices this is not the case, and a lot of effort is put into the accurate modeling of a particular component.

In this paper we show how, under certain explicitly specified conditions, a black-box model for a device can directly be deduced from large-signal measurements. The black-box models, which are completely technology independent, are as accurate as the "nonlinear-network analyzer" measurements [3] on which they rely.

Theoretical Description of the Black-Box Model

The model will typically be valid for a two-port nonlinear device-under-test, called DUT, (typically one power transistor, but the theory can be applied on a whole power amplifier stage containing several transistors) where one spectral component is dominant, and where all other spectral components present are significantly smaller and are harmonically related to the dominant component. The variables used to describe the signals at both ports are the classical incident and scattered voltage waves (cf. definition of the s-parameter matrix [5]), typically defined in a characteristic impedance of 50Ω, together with the dc current and voltage biasing parameters. The input (incident waves) and output (scattered waves) variables are defined in the frequency domain as depicted in Fig. 1: A_{ij} denotes the complex number representing the j^{th} spectral component of the incident voltage wave at port “i” and B_{ij} denotes in an analog manner the scattered voltage waves. The model will be valid provided that A_{11} is the dominant component. This kind of excitation signal is typical for a transistor in a power amplifier under a one-tone excitation. The harmonic (relatively small) input components can be due to harmonics generated in the previous or following stages, or can be harmonics created by the DUT itself and which are reflected on source or load mismatches. Note that the fundamental component is used as a timing reference defining the phases of all other components. This implies that, after phase normalization, the complex number A_{11} has no longer an imaginary part and can be treated as a real number.

Fig. 1 Definition of the signals



The problem is to find a model for the DUT, based upon a limited number of measurements, which allows to describe the scattered components $B_{i1},...,B_{iN}$ and the dependent biasing parameters S_1 and S_2 as a function of $A_{i1},...,A_{iN}$ and the independent biasing conditions R_1 and R_2 [2]. This is illustrated in Eq. 1,

$$B_{ij} = F_{ij}(A^{re}_{11}, A^{re}_{12}, ..., A^{re}_{1N}, A^{im}_{12}, ..., A^{im}_{1N}, A^{re}_{21}..., A^{re}_{2N}, A^{im}_{21}, ..., A^{im}_{2N}, R_1, R_2) \qquad \text{Eq. 1}$$

where F_{ij} stands for a describing function corresponding to the j^{th} harmonic output at signal port “i”, where the superscript “re” and “im” stand for the respective real and imaginary part and where R_1 and R_2 denote the biasing conditions (this can be either current or voltage). Note that, because of the phase definition mentioned earlier, A^{im}_{11} will always equal zero, such that it can be omitted in Eq. 1. To keep the mathematical notations consistent, one defines

$$B_{0j} = S_1 + \Im S_2 , \qquad \text{Eq. 2}$$

with $\Im$ being equal to the complex square root of -1. This way the dependent bias parameters can be treated exactly the same as the scattered travelling voltage waves.

In order to simplify the identification of the describing functions, it is assumed that the superposition principle holds for the relatively small components $A_{i2},...,A_{iN}$. The final mathematical model then becomes:

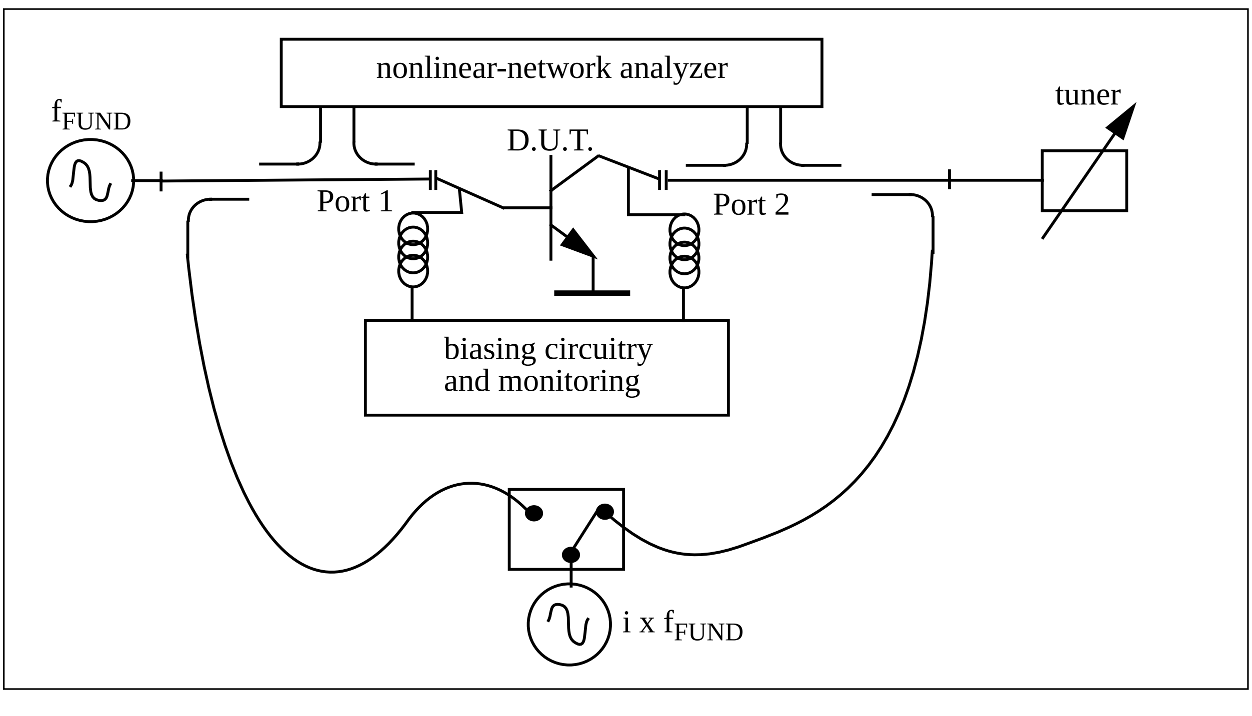
$$B_{ij} = K_{ij} + \sum_{\substack{k = 2 \dots N \\ l = 1, 2}} L_{ijkl} A^{re}_{kl} + \sum_{\substack{k = 2 \dots N \\ l = 1, 2}} M_{ijkl} A^{im}_{kl}, \qquad \text{Eq. 3}$$

where K_{ij} , L_{ijkl} and M_{ijkl} represent linearizing complex coefficients, which are a function of the inputs A^{re}_{11}, A^{re}_{12}, A^{im}_{12} and of the biasing conditions R_1 and R_2 (in what follows these inputs are called the "large signal inputs"). Since the complex coefficients are considered as a natural extension of classical s-parameters valid under large-signal conditions they are called "large-signal s-parameters". Although not exactly the same, the idea of linearization is similar to the "impedance-form conversion matrix" described in [6] and the idea of the "generalized Volterra series" described in [7].

The Measurement Set-up

The measurement set-up to automatically extract the model parameters is illustrated in Fig. 2. A "nonlinear-network analyzer" [3] [4] allows to measure the phase and amplitude of all spectral components A_{ij} and B_{ij}. One synthesizer will excite the D.U.T. at port 1 with a large fundamental (A_{11} with a frequency f_{FUND}), while a tuner allows to create a whole set of A_{21}'s by reflecting B_{21}. The biasing circuitry is computer controlled and allows an accurate setting and monitoring of the biasing conditions. At a particular value of the large signal inputs, relatively small harmonic components are injected one-by-one, at port 1 as well as at port 2, at all frequencies of interest. A second synthesizer and couplers are used for this purpose. If an appropriate number of experiments is performed, the values of K_{ij}, L_{ijkl} and M_{ijkl} (for that particular value of large signal inputs) are estimated by performing a least-squares fit on the measured data. The least-squares fit is straight forward since the model is linear in the parameters K_{ij}, L_{ijkl} and M_{ijkl}. Note that the procedure can be repeated for any value of the large signal inputs.

Fig. 2 The measurement set-up



Experimental Model Extraction and Verification

Next, the technique is illustrated on a heterojunction bipolar transistor in a common emitter configuration. The large signal inputs chosen for the experiment are given in Table 1. The fundamental frequency chosen is 1GHz.

Table 1   Large signal inputs

base-emitter voltage	0.35V
collector-emitter voltage	5.00V
fundamental input power (I_{11})	12dBm

Unfortunately, no tuner was available at the time of the measurements. Instead, a matched load was chosen and the behavior was also linearized in the parameter A_{21}. This is done by performing the summation in Eq. 3 for k=1...N in stead of k=2...N. The consequence is that the model will only be valid for relatively small values of A_{21} (corresponding in practice to a fundamental reflection coefficient at the output with an amplitude significantly smaller than 1). Next, several experiments are performed in order to extract the “large-signal s-parameters”. Note that in what follows “one measurement” corresponds to measuring all components B_{ij} and A_{ij}, for j going from 1 to 18. First a measurement is done without injecting harmonics. Next harmonics 2 and 3 (frequencies 2GHz and 3GHz) are injected towards the transistor base, and harmonics 1, 2 and 3 (frequencies 1GHz, 2GHz and 3GHz) are injected towards the collector and each time 6 independent measurements are done where the phase relationship between the fundamental synthesizer and the harmonic synthesizer is randomized. This results in a set of 31 measurements. The amplitude of the injected A_{12}, A_{13}, A_{21}, A_{22} and A_{23} is about -9dBm. In order to have an idea about the degree of nonlinearity, two periods of the collector current waveform, without harmonics injected, are depicted in Fig. 3. A hard clipping is clearly visible. The level of the harmonics produced at the output is given in Table 2.

Table 2   Level of the harmonics at the output

Power	B_{21}	B_{22}	B_{23}	B_{24}	B_{25}	B_{26}
dBm	3.5	-0.6	-9.8	-26.6	-17.5	-25.9
mVolt-peak	476	294	102	14.8	41.9	16.0

Next the “large signal s-parameters”, valid at the specified large signal input, are calculated and a verification of the model is performed.

The verification is done as follows. The harmonic synthesizer injects power into the base of the transistor, with an output frequency of 3GHz. Next 75 measurements are performed while randomizing the phase of the harmonic synthesizer relative to the fundamental and while keeping the amplitude of I_{13} equal to -10dBm during the first 50 measurements and equal to -16dBm during the last 25 measurements. The corresponding measured complex values of A_{31} are depicted in Fig. 4. The corresponding modeled and measured values of B_{22} are depicted in Fig. 5. As one can see the correspondence between the measured and modeled values is very good.

The effect of injecting a small third harmonic into the base of the transistor can also be interpreted in the time domain. The difference between the collector current with and without the presence of the harmonic, and this for an amplitude of -10 dBm and -16 dBm, is depicted in Fig. 6. As one can theoretically expect, the difference between the two is zero when the collector current is zero. While the transistor is conducting however, the harmonic will contribute to the output current. Since superposition holds, this difference current waveform is proportional to the amplitude of the injected harmonic (assuming the phase of the harmonic remains constant). This proportionality can clearly be noted in Fig. 6, thereby confirming the superposition assumption.

Fig. 3 Collector Current Waveform

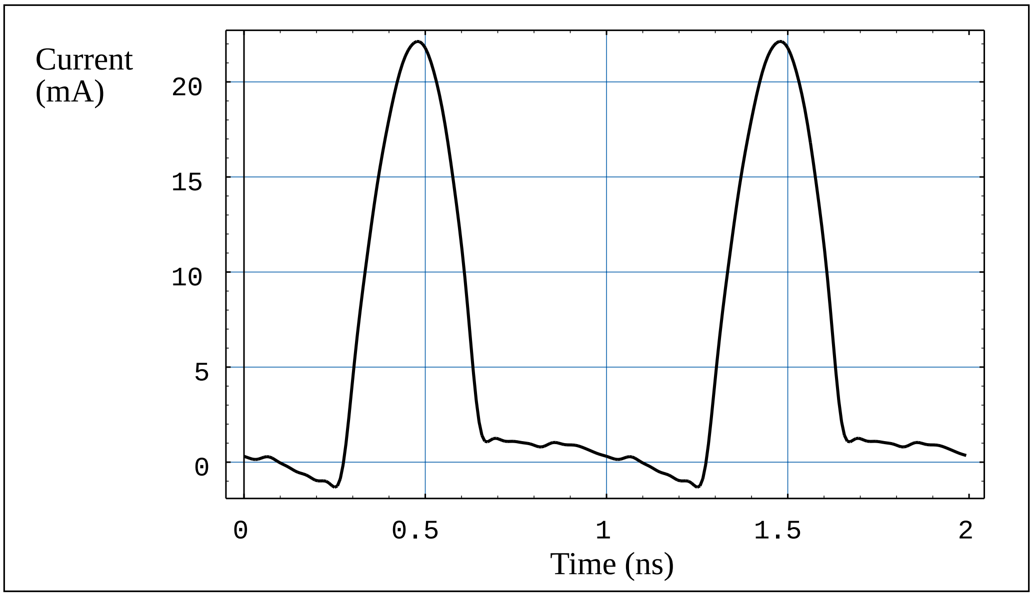


Fig. 4 Measured Complex Values of A_{31}

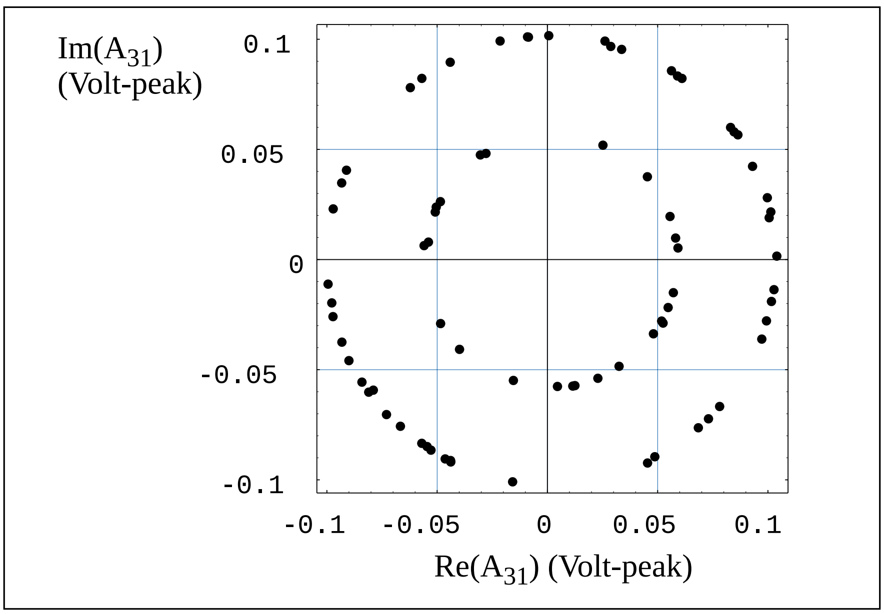


Fig. 5 Measured and Modeled Complex Values of B_{22}

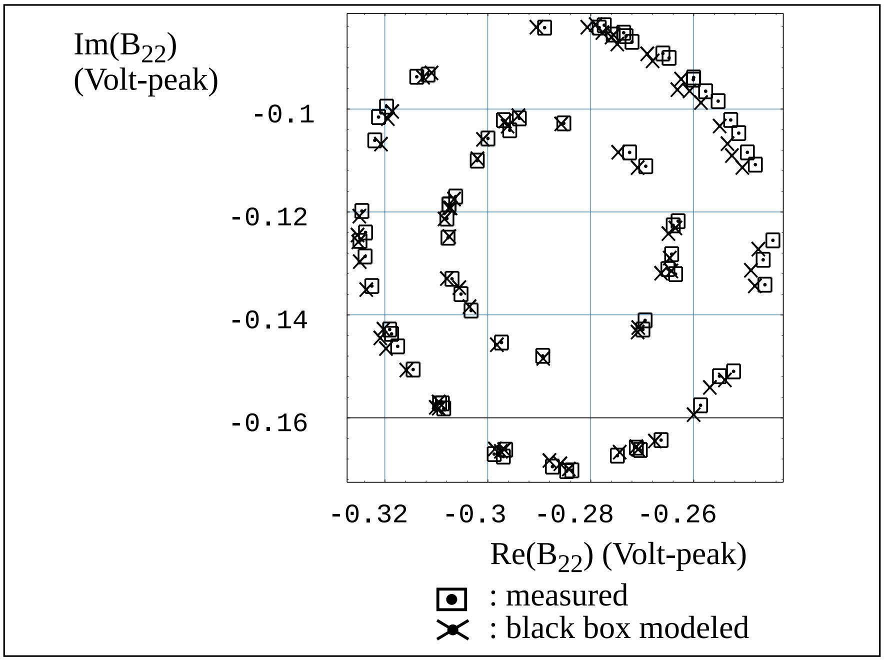
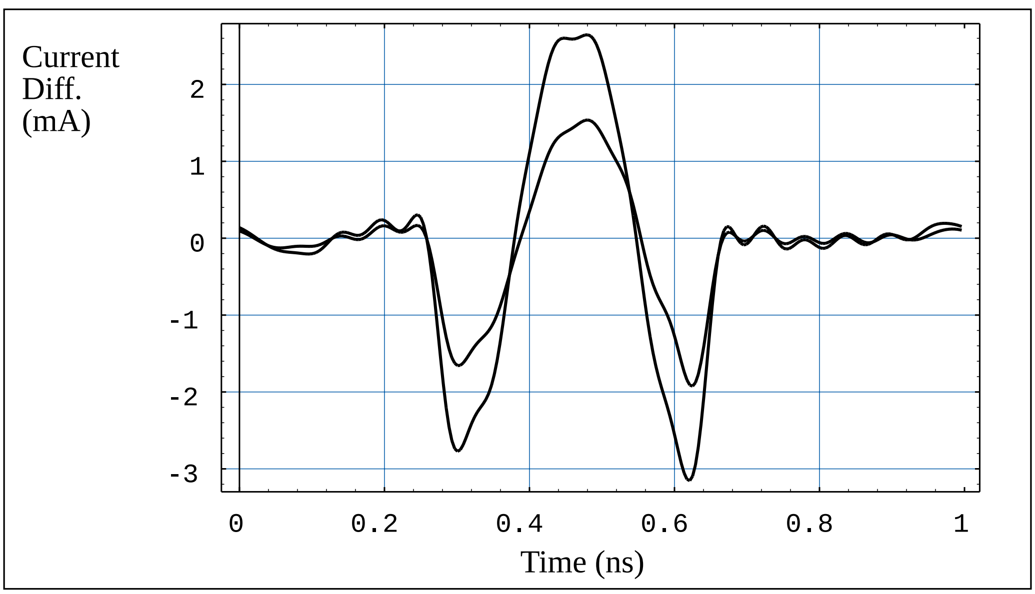


Fig. 6 Difference Between Collector Current Waveforms



Integration of the Model in an Harmonic Balance Simulator

In order to be useful, it is very important that the black-box model can be integrated in existing CAD packages. In what follows, a black-box model of a packaged FET transistor, extracted by means of the set-up described earlier, is integrated in a commercial harmonic-balance simulator, this is illustrated in Fig. 7. The circuit build around the model allows to simulate the behavior of the component with a whole range of biasing condition and RF excitation signals (provided that the excitation and biasing are within the limits of the black-box model). Several “nonlinear network-analyzer” measurements can then be done, and the measured behavior can be compared with the behavior resulting from the harmonic balance simulation. For one such a measurement, the result is illustrated in Fig. 8. The four graphs illustrate the modeled and measured large signal time domain waveforms for the current and voltage at the gate (port 1) and drain (port 2) of the packaged transistor. The fundamental frequency used for the experiment is 1 GHz. One notes that there is a very good correspondence between the measured and modeled values. A third harmonic is injected towards the gate (this can clearly be seen on the gate voltage waveform), the gate voltage equals -1.78 V, the drain voltage equals 2.71 V. The power of the measured spectral components of incident and scattered voltage waves is given in Table 3 (up to the fifth harmonic).

Table 3 Measured Power Levels (dBm)

j	1	2	3	4	5
A_{1j}	9.15	-23.2	-6.30	-34.1	-55.7
A_{2j}	-8.80	-22.18	-27.2	-25.8	-34.5
B_{1j}	8.73	-11.5	-7.97	-17.2	-49.8
B_{2j}	15.7	6.99	-5.61	-3.96	-10.1

One notes that the two main incident spectral components are A_{11} (the injected fundamental), with a value of 9.15 dBm, and A_{13} (the injected third harmonic) with a value of -6 dBm. As expected, the power of the other incident components is significantly lower. The power of the complex difference between the model and the measurements is given in Table 4.

Table 4 Complex Difference between Model and Measurements (dBm)

j	1	2	3	4	5
D_{1j}	-47	-40	-44	-31	-39
D_{2j}	-14	-20	-23	-19	-24

D_{ij} is a complex residual defined as: $D_{ij} = \text{modeled}(B_{ij}) - \text{measured}(B_{ij})$.
One can conclude that the residual powers are small, which indicates that the model is good (this was already apparent from the time domain waveforms).

Fig. 7 The Black-Box Model in an Harmonic-Balance Simulator

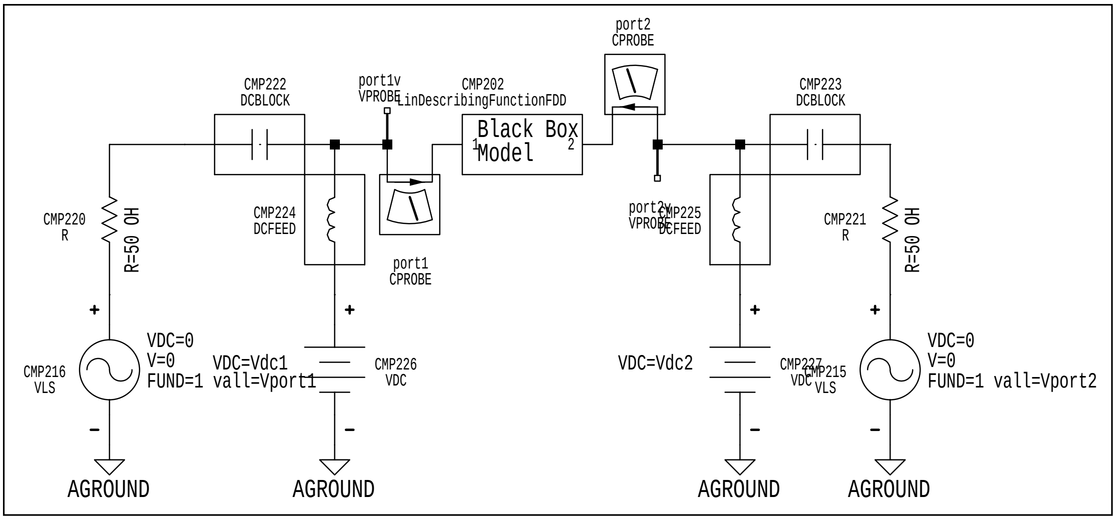
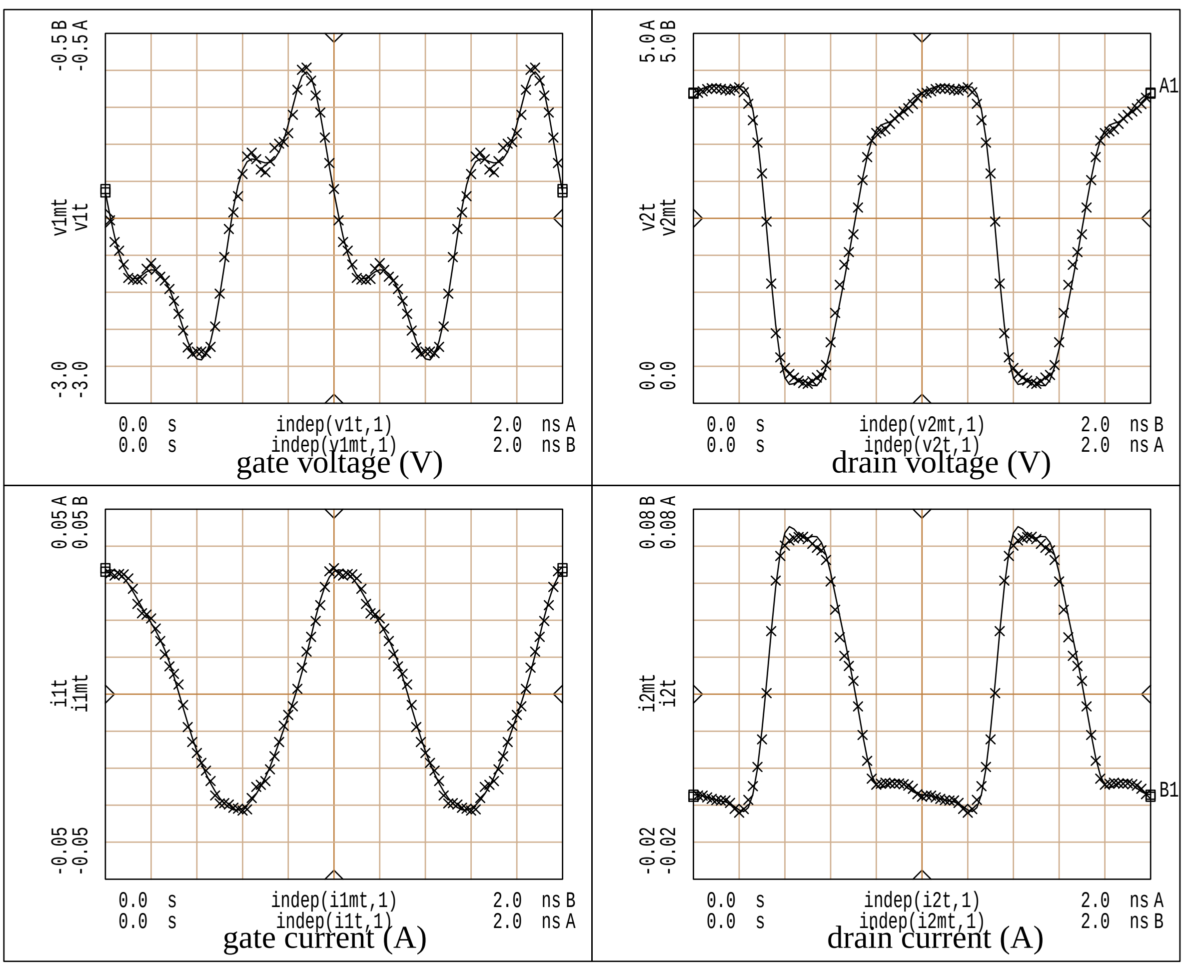


Fig. 8 Modeled and Measured Current and Voltage Waveforms



× : measured waveform

—— : harmonic balance simulation

Conclusions

“Large signal s-parameters” can accurately be measured with an automated set-up containing a “vectorial nonlinear-network analyzer”. The corresponding model accurately describes the behavior of the hard-nonlinear microwave component under a large-signal one-tone excitation (relatively small harmonics may be present), and can be integrated in a commercial harmonic balance simulator.

Acknowledgments

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