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Multi-tone, Multi-port, and Dynamic Memory Enhancements to PHD Nonlinear Behavioral Models from Large-signal Measurements and Simulations Measurements

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ABSTRACT — **The PHD nonlinear behavioral model is extended to handle multiple large tones at an arbitrary number of ports, and enhanced for dynamic long-term memory. New capabilities are exemplified by an amplifier model, derived from large-signal network analyzer (LSNA) data, valid for arbitrary impedance environments, and a model of a 50GHz integrated mixer, including leakage terms and IF mismatch dependence. Dynamic memory is demonstrated by an HBT amplifier model identified from up-converted band-limited noise excitations. The models are validated with independent LSNA component data or, for simulation-based models, with the corresponding circuit models.**

Index Terms — **Design automation, distortion, frequency domain analysis, microwave measurements, nonlinear circuits.**

I. INTRODUCTION

The Poly-Harmonic Distortion (PHD) nonlinear model, presented in [1]-[2], is a rigorous, frequency-domain, black-box nonlinear behavioral modeling approach for microwave and rf module and subsystem design. Very accurate models, implemented in commercial simulators have been generated from automated large-signal measurements using large-signal network analyzers (LSNAs), and from simulations starting from detailed circuit-level models of an IC or component. The models can be easily and unambiguously identified from one of several related, simple, multiple-tone experiment designs. Key features of the PHD model include the ability to predict distortion through cascaded chains of nonlinear components under large-signal drive in the presence of small to moderate mismatches at both the fundamental and harmonic frequencies. The model can predict PAE, output match (the proper representation of "Hot S22") including the required terms proportional to the conjugate of the incident waves at the output port, both even and odd harmonics, and the parametric dependence of bias voltages and currents. The resulting models protect intellectual property (IP), thereby promoting its reuse and sharing. The models reduce design breadboard costs, improve simulation speed compared to the detailed circuit model (if one exists), and, in the case of measurement-based models, embody a sound methodology for using large-signal data directly in the design of nonlinear microwave & RF systems. This, and independent research [3], has demonstrated that the rigorous non-analytic spectral mappings of this class of models are required to predict nonlinear figures of merit, such as ACPR, of composite systems of imperfectly matched devices.

II. GENERALIZED PHD FORMULATION

The generalized PHD modeling approach is based on a set of systematic and controlled approximations to the multivariate, nonlinear, complex-valued spectral mappings from input phasors to output phasors that define the multi-port system in the frequency domain. The full nonlinear mapping is approximated by the sum of a simpler nonlinear mapping and a linear mapping. This is indicated in (1).

$$B_{pk}(A_{11}, A_{12}, \ldots A_{21}, A_{22}, \ldots, A_{rm,\ldots}) \approx F_{pk}(LSOP) \cdot P_k^{(F)} + \sum_q \sum_l \left[S_{pq,kl}(LSOP) \cdot P_{kl}^{(S)} \cdot A_{ql} + T_{pq,kl}(LSOP) \cdot P_{kl}^{(T)} \cdot A_{ql}^* \right] \quad (1)$$

Here the indices p, q, and r specify the ports, k, l, and m indicate spectral components (e.g. harmonics or inter-modulation components), A_{ql} is the incident wave at port q at frequency index l, B_{pk} is the transmitted or reflected wave at port p and spectral component k. *LSOP* stands for *large-signal operating point*, which is specified by the subset of A_{rm} terms retained in the nonlinear mappings $F_{pk,}$ (and the S and T terms). The dependence on all other input spectral components is approximated as linear, which is given by the contributions in the summations in (1). That is, the A_{ql} terms appearing in the square brackets are (subsets) of those incident waves that are *not* arguments of the nonlinear functions F, S, and T. P terms are pure phases, which are determined from time-invariance considerations. For example, if we consider the simplest case where the LSOP is established by a single large tone incident on port 1, we obtain the models of [1]-[2], which, can be written, for $P = e^{j\phi(A_{11})}$, according to (2).

$$B_{pk}(A_{11}, A_{12}, \ldots A_{21}, A_{22}, \ldots) \approx F_{pk}(|A_{11}|) \cdot P^k + \sum_q \sum_l \left[S_{pq,kl}(|A_{11}|) \cdot P^{k-l} \cdot A_{ql} + T_{pq,kl}(|A_{11}|) \cdot P^{k+l} \cdot A_{ql}^* \right] \quad (2)$$

We see the nonlinear mappings in this case depend only on a single real variable. All the other incident wave dependences are linear. This is a dramatic simplification, and enables simple characterization and direct identification. F_{pk} is the response of the (matched) system to the stimulus of a single large input tone at the input. The S and T terms can be interpreted as the Jacobian of the input-output mapping evaluated at this simple signal. These S and T terms, however, which are functions of the drive, allow the model to predict the component behavior for small to moderate mismatches at port 2 at the fundamental and harmonic frequencies, since they represent the response, linearized around the simple LSOP.

A. PHD model for arbitrary impedances from measured data

Using the general formulation (1), we now go beyond (2) by considering two cases in which the LSOP is defined by more complicated signals than a single tone incident on the input. For an amplifier, when a tone injected into the output port becomes very large, due perhaps to a very large mismatch, the approximation (2) of a 1-tone LSOP for the component model begins to degrade. For more accuracy one simply retains the full complex A_{21} dependence of the nonlinear mappings F, S, and T. In this case, the LSOP depends on three real variables, $|A_{11}|$, $|A_{21}|$, and Phase(A_{21}). The extended model is therefore nonlinear in all of the phasors at the fundamental frequency, while remaining linear in all of the harmonics (if they are retained in the model). The sum in (1) is over all ports and all harmonic frequencies; the sum no longer includes A_{21}, as it did in (2), since A_{21} is now an argument of the F, S, and T terms.

A WJ Communications FP2189 GaAs HFET transistor was chosen to demonstrate the capability of the extended PHD amplifier model for arbitrary impedances. For the given example, a fundamental only (no harmonic terms are included) PHD model was identified. The transistor is specified with an output power 1dB compression point of 30dBm and a frequency range from 50MHz to 4GHz. The part comes in a SOT-89 package and was mounted on a PC board, together with carefully designed calibration standards to allow de-embedding to the device planes. A calibration of the LRRM type was performed on the LSNA instrument reported in [4]. To identify a PHD model with two large tones, it is required to define experiments for a very wide range of A_{11} and A_{21} values. Given the device output characteristics, this is equivalent to a synthetic load-pull measurement over the entire Smith chart. New software controlling the LSNA was developed to characterize the device under these conditions. The device was swept in frequency from 0.25 GHz – 4 GHz, over a wide range of input powers. A simplified schematic of the measurement setup is shown in Fig.1. One synthesizer is used to generate A_{11} and a second synthesizer is used to generate the large-signal A_{21} waves, thereby synthesizing reflection coefficients equal to A_{21} divided by the resulting B_{21}.

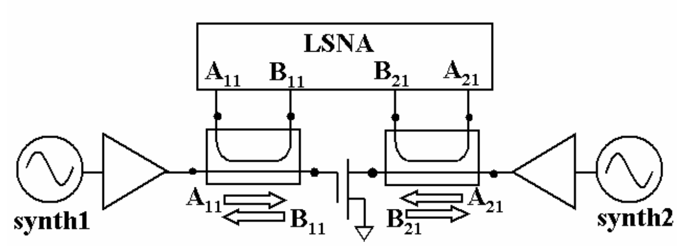

Fig. 1 Simplified schematic of the measurement setup

The model extraction algorithm of [1] was extended to identify the enhanced PHD model. Fig. 2 shows a subset of the measured LSNA data. The operating conditions are: drain bias voltage 8V, drain current 250mA, fundamental frequency 1GHz, incident power (A_{11}) 17dBm. The contours on the Smith chart, generated from the identified model, represent the net output power of the transistor with a step of 100mW. The maximum net output power corresponds to 1.4W. The dots represent the actual synthesized output reflection coefficients that were used to identify the enhanced PHD model, which is in essence an interpolator between the synthesized output reflection coefficients.

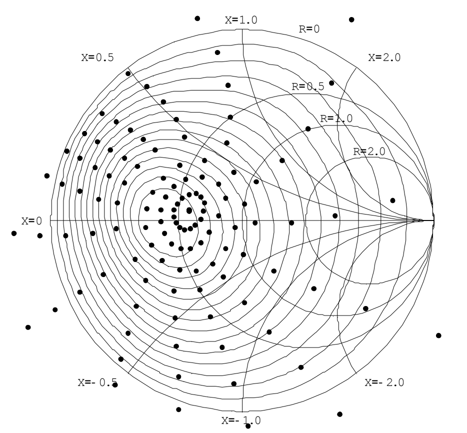

Fig. 2 Synthesized output reflection coefficients (points) and simulated power contours (lines) from enhanced PHD model.

The extended PHD model, when implemented in a simulator, is in perfect agreement with the transistor characteristics at all measured data points, over the entire range of measured characteristics, valid over frequency, input power, and output impedances over the entire Smith chart. Between measured data points, the model interpolates smoothly. This model is an example of applying the black-box behavioral techniques at the transistor level, rather than at the full IC level. It can be a viable substitute for the classic “compact” transistor models for many applications.

B. PHD Mixer model with leakage and IF mismatch

A second, and distinct generalization of (2) is embodied by a three-port device, this time with the two large signals determining the LSOP, incident at different ports, at distinct frequencies. This is the case of a nonlinear mixer (e.g. where both the LO and RF signals are assumed large and produce their own harmonics and inter-modulation products). Assuming the LO and RF frequencies are incommensurate, it can be proved that the LSOP depends only on the magnitude of the large LO and RF signal amplitudes.

The extended PHD methodology was applied in simulation to a 50GHz integrated mixer manufactured by Agilent Technologies' High Frequency Technology Center. The IC contains over 40 III-V hetero-junction bipolar transistors, each modeled by the Agilent HBT nonlinear transistor model [5]. Linear terms at the IF port at intermodulation frequencies were retained to model mismatch effects. Fig. 3 compares simulations of conversion gain with the PHD model to the underlying circuit model from which it was generated. The simulation were performed using Agilent ADS. In this comparison, the IF load is 10 ohms, far from the 50 ohm match in which the model was identified. This is a direct validation of the accuracy of the linear approximations in (1) for mismatch terms at the IF frequencies. The conversion gain is a sensitive function of the IF load, yet the PHD model is remarkably accurate. The extended PHD model predicts LO and RF leakage terms, as well as scores of additional port-to-port couplings in comparable agreement. Finally, the PHD model achieves a factor of about 6 times speed improvement over the detailed circuit model, providing a significant advantage for system simulation.

C. Long-term Dynamic Memory Extension

The formulation (1) describes instantaneous mappings from the incident A-waves to the output B-waves. The above results demonstrate that (1) is sufficient to produce accurate models for CW applications. There are important cases, however, where the static mapping is no longer accurate for many components. The reason has to do with the fact that complex stimuli (e.g. broad-band modulation, pulsed RF, etc.) may excite additional slow dynamical variables of the component, such as temperature or bias supply voltage, that in turn affect the nonlinear mapping of the rf signals. When the output phasors depend not only on the instantaneous input RF phasors, but also on other slowly varying state variables of the system, we get long-term dynamic memory effects.

It is possible to extend the PHD model to incorporate additional dynamic "hidden" variables. The example presented here is the DC-20 GHz HBT amplifier model considered in [1] but this time excited by band-limited, normally distributed noise. It is possible to extend the PHD model to incorporate additional dynamic "hidden" variables.

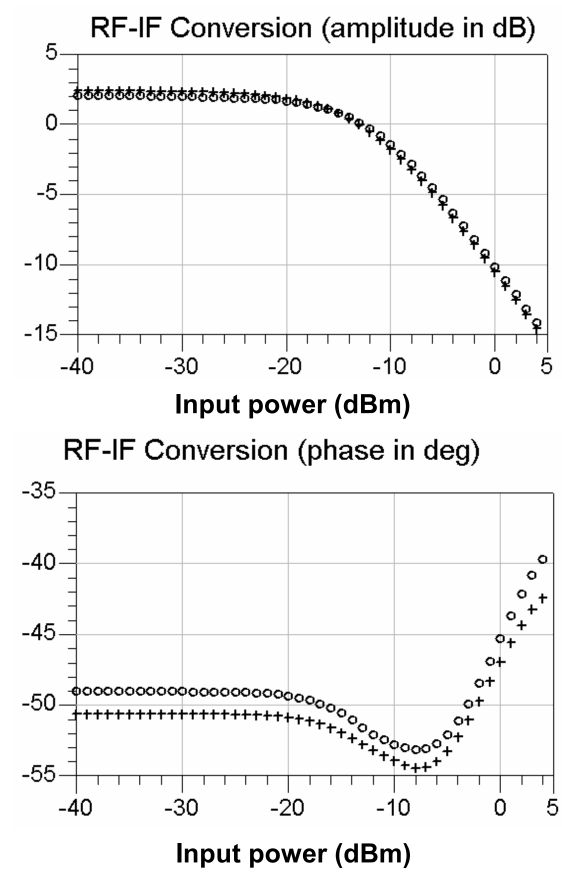

Fig.3: Extended PHD mixer model (o); Circuit model (+) (a) Conversion gain (up-conversion) in dB versus RF power (dBm) and (b) conversion phase in degrees versus RF power (dBm) for an LO power of 0 dBm and an IF load of 10 Ohm.

The example presented here is the DC-20 GHz HBT amplifier model considered in [1], but this time excited by band-limited, normally distributed noise. The modeled amplifier includes a non-ideal bias circuit through which can pass low-frequency signals, generated by the nonlinearity of the amplifier. We assume neither the bias pin nor the bias circuit is accessible; both are considered part of the black box. The only access we have is to the RF ports.

The model was simplified by assuming a perfect output match, and only the dependence of B_{21} on A_{11} and the hidden state variable was considered. For this example, 20MHz BW noise, centered at 2.6 GHz, is used to excite the system.

Plotting the output B_{21}-wave as a function of the input A_{11} wave proves the presence of memory. One input amplitude A_{11} results in a whole range of output amplitudes, especially at the high input levels. This is shown in Fig. 4. At an input amplitude of 7dBm the output amplitude ranges from 13dBm to 16dBm. This is a clear manifestation of a memory effect.

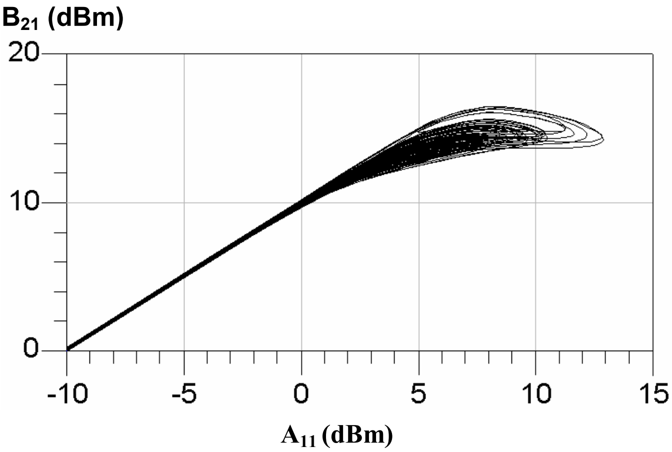

Fig. 4: Amplitude of B_{21} (dBm) vs. amplitude of A_{11} (dBm)

System identification techniques were used to "discover" the hidden variable. The state variable was estimated by an exponential moving average filter applied to the squared magnitude of the input signal. The correlation between B_{21} and this estimate, as a function of the filter delay parameter, was used to identify the timescale. In this case, an optimal time delay parameter, noted τ, was determined to be 110 ns. With the new variable identified, the two-dimensional functional mapping from A_{11} and the new state variable is easily determined. The final relationship between $B_{21}(t)$ and $A_{11}(t)$ is described by the following equation, whereby F(.,.) is implemented as a two dimensional lookup table.

$$B_{21}(t)=F\left(\left|A_{11}(t)\right|,\int_{-\infty}^{t}\left|A_{11}(u)\right|^2 e^{\frac{u-t}{\tau}}\frac{du}{\tau}\right)e^{j\varphi(A_{11}(t))} \qquad (3)$$

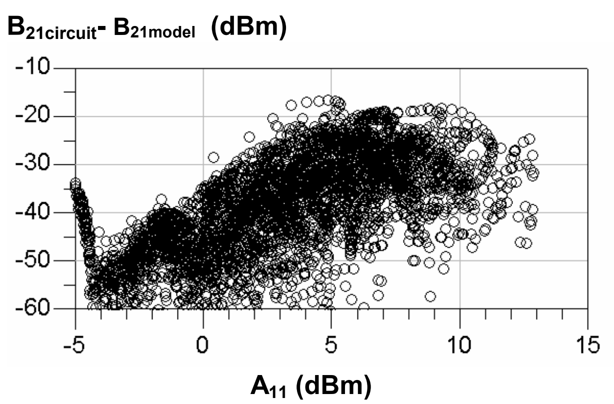

Fig. 5 Difference (dBm) between the PHD model with dynamic memory and the circuit-level model as a function of the amplitude of A_{11} (dBm)

The dynamic memory extended PHD model is implemented as a first-order differential equation in the envelope domain, using the frequency domain device (FDD) of Agilent ADS, according to [6]. The model is compared to the detailed circuit model of the amplifier in Fig. 5. The results demonstrate that the extended PHD model with dynamic memory tracks the result of the circuit model very well, even for very large values of A(t). The maximum deviation between the model and the circuit is at a level of about -16dBm, 32dB lower than the maximum output amplitude. Note that the static (memoryless) PHD models of [1]-[2] would result in a single line characteristic for figure 4, completely unable to capture the multi-valued set of B-waves for a given A-wave.

IV. Conclusion

The PHD nonlinear behavioral model has been generalized for an arbitrary number of RF and DC ports, fundamental frequencies, and large signal tones. It has been demonstrated to be accurate over power, frequency, and bias, at matched as well as strongly mismatched conditions. The modeling methodology has been applied, successfully, at the transistor level and the complete IC level, for both two and three-port nonlinear components. The present infrastructure allows the user to trade speed for accuracy by including more signals in the nonlinear mapping and/or by keeping more linear terms representing mismatch at harmonic or inter-modulation frequencies. The model has been extended for long-term, nonlinear dynamic memory, such as required to model self-heating and bias-line interactions under complex stimulus-response conditions. These model capabilities enable accurate simulations of cascaded chains of multi-port nonlinear devices for RF system design from measurement and simulation.

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