ACCURATE RF AND MICROWAVE SYSTEM LEVEL MODELING OF WIDE BAND NONLINEAR CIRCUITS

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ABSTRACT

Accurate system level simulation is indispensable for efficient RF communication system design. However system level models of nonlinear circuits are today very limited by their inability to handle nonlinear memory effects. The paper describes a new approach of system-level modeling which accounts efficiently for nonlinear envelope memory effects of wide band amplifiers, multipliers and mixers.

INTRODUCTION

In the last two decades transistor-level simulation tools for microwave and RF circuits have significantly evolved. Work by various authors on Harmonic Balance (HB) technique [1-3] has eased the calculation of steady state response of nonlinear circuit under single and multi-tone excitation. Initially focused on traditional microwave GaAs circuits, solution algorithms of HB have been recently enhanced by several authors [4-5] to handle more efficiently the new demand on RF silicon ICs. In the same time Envelope transient techniques [6-7] have emerged that handle multi-rate (mixed modulation and carrier) analysis at transistor level. All these techniques form a complete palette of efficient tools for the design of analog communication system components. However despite their today's tremendous power, these tools still need large memory resources and CPU time which make them unsuitable for system analysis and tune. Significant reduction on computer resource needs allowing efficient possibilities for system analysis and tune can only be achieved by a simulation at system level. The main limiting factor of system level simulation today is model accuracy of wide band nonlinear RF circuits (power amplifiers, multipliers and mixers). One of the major limitations is that models in use today do not account for nonlinear memory effects. This paper describes a new system level modeling approach based on "sliding kernels dynamic" Volterra series, which accounts effectively of nonlinear memory effects. Despite it has been widely used in mixer and oscillator design, Volterra series[8] has a bad reputation among RF engineers of being a cumbersome technique; but yet it is the most accurate formalism for representing nonlinear systems with memory, especially when the system is large and distributed. Nevertheless the classical form of Volterra series has poor convergence properties and in practice it is hardly possible to measure its kernels of order more than two to three. These limitations make the classical Volterra series almost useless for most nonlinear IC applications. What we propose to use in this paper is a modified form of Volterra series [9] which resolves the above limitations, providing sufficient accuracy with **only the first order** kernel. We will briefly present the modified Volterra series equation in section II, then show how this fits efficiently to system level modeling in section III. An application example is presented which show the efficiency of the proposed model.

CLASSICAL VS DYNAMIC VOLTERRA SERIES EXPANSION

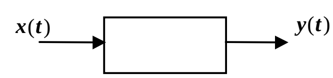

Fig.1 ***Nonlinear system with memory***

The output $y(t_n)$ of a nonlinear system with memory duration $M\Delta t$ can be intuitively expressed as

$$y(t_n) = f(\vec{x})$$
$$\vec{x} = [x(t_n), x(t_n - \Delta t), \dots, x(t_n - M\Delta t)]^T \tag{1}$$

Taylor series expansion of $y(t_n)$ around some arbitrary signal trajectory $\vec{x} = \vec{x}_0$ gives equation (2) below, from which we obtain the well known Volterra series expansion (3) as $\Delta t \to 0$, by choosing $\vec{x}_0 = [0,0,\dots,0]^T$

$$\begin{aligned} y(t_n) = f(\vec{x}_0) \; + \; \Delta f(\vec{x}_0)^T(\vec{x} - \vec{x}_0) \; + \\ \tfrac{1}{2}(\vec{x} - \vec{x}_0)^T[\Delta^2 f(\vec{x}_0)](\vec{x} - \vec{x}_0) \; + \; \dots \end{aligned} \tag{2}$$

$$\begin{aligned} y(t) &= \sum_{n=1}^{\infty} y_n(t) \\ y_n(t) &= \int_0^{\tau}\!..\!\int_0^{\tau} h_n(\lambda_1,..,\lambda_n)\prod_{i=1}^{n} x(t-\lambda_i)\,d\lambda_i \end{aligned} \tag{3}$$

$h_n(\lambda_1,..,\lambda_n)$ is called Volterra kernel of order n. As it can be seen, Volterra kernels are independent of the input signal $x(t)$. They are indeed coefficients of a power series expansion. We thus see that for most nonlinear applications encountered in analog IC systems it is necessary to consider a large number of kernels in order to accurately describe the response. Here comes the two-fold difficulty of identifying high order kernels and computing multiple dimensional integrals, which limits usefulness of this elegant approach.

To try resolving these limitations, in [9] it was suggested that carrying Taylor series expansion in eq (2) around a well behaved trajectory $\vec{x}_0$ given by prior knowledge of the system response, rather than the null vector can give excellent convergence properties to the resulting series. One simple and efficient trajectory has been found by authors in [10] to be $\vec{x}_0 = [x(t_n), x(t_n),..., x(t_n)]^T$. In this case the resulting modified Volterra series takes the form

$$y(t) = y_{dc}(x(t)) + \int_0^\tau .. \int_0^\tau \hat{h}_n(x(t),\lambda_1,..,\lambda_n) \prod_{i=1}^{n} [x(t-\lambda_i) - x(t)] d\lambda_i \qquad (4)$$

In the above expression $y_{dc}(x(t))$ is the static characteristic (DC) of the system, $\hat{h}_n(x(t),\lambda_1,..,\lambda_n)$ are the dynamic Volterra kernels, called so to imply their dependence on the input signal. This modified series has the important property to separate purely static effects from memory effects, which are intimately mixed in the classical series. Hence if the system is purely static, the series converges only with the static term, irrespective of the input power. In fact for most components where memory effects are parasitic effects rather than desired ones, the nonlinearity lies mainly in the static term and memory effects are just mildly nonlinear. Hence series (4) can be limited to only first order, with good accuracy.

$$y(t) = y_{dc}(x(t)) + \int_0^\tau h(x(t),\lambda)[(x(t-\lambda) - x(t)] d\lambda \qquad (5)$$

A more convenient form of eq (5) is obtained by considering a Fourier integral in place of convolution to find.

$$y(t) = y_{dc}(x(t)) + \frac{1}{2\pi} \int_{-\infty}^{\infty} \hat{H}(x(t),\omega) X(\omega) e^{j\omega t} d\omega \qquad (6)$$

$$\hat{H}(x(t),\omega) = H(x(t),\omega) - H(x(t),0)$$

In the above, $H(x(t),\omega)$ is indeed the small signal transfer function of the system computed around the pump $x(t)$. So this model does not need complicate simulation and extraction procedures.

Eq (6) has been used with some success by several authors [10-12] to model GaAs transistors in harmonic balance simulator. In the following we are extending this principle to system level modeling.

SYSTEM LEVEL MODELING

At system level, we need to model nonlinear components like power amplifier, multiplier, mixers and VCO's.

The basic assumption on system level modeling is that signal $x(t)$ at any component port can be expressed as band limited modulation around a carrier frequency:

$$x(t) = \Re e[\hat{X}(t) e^{j\omega_{0x} t}] \qquad (7)$$

where $\hat{X}(t)$ and ω_{0x} are respectively the complex envelope (modulation) and reference carrier (mid-channel frequency for a multiplex) of the signal $x(t)$. All idle frequencies are supposed to be sufficiently filtered-out within the subsystem, or constitute otherwise distinct ports.

For sake of notational simplicity let us consider a component block with only one input and output, as depicted in Fig.2, the concept can be generalized to multiple input and output blocs.

$x(t) = \Re e[\hat{X}(t) e^{j\omega_{0x} t}]$ $\qquad$ $y(t) = \Re e[\hat{Y}(t) e^{j\omega_{0y} t}]$

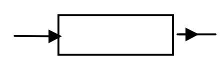

Fig.2 *component bloc diagram*

Reference carrier frequencies ω_{0x} and ω_{0y} of excitation and response being known a priori, they don't bear any information. For modeling the bloc we need only to identify the relationship between the two complex envelopes $\hat{X}(t)$ and $\hat{Y}(t)$. This problem of nonlinear system envelope modeling has deserved a large amount of literature [13-18]. The models proposed have been aiming on traveling wave amplifier (TWTA) non-linearity. Hence their efficiency is limited in handling nonlinear memory effects, as those due to solid state power matching and biasing circuits [17].

Basic dynamic volterra series model equation

The envelopes $\hat{X}(t)$ and $\hat{Y}(t)$ being analytic signals, $\hat{Y}(t)$ is actually a function of $\hat{X}(t)$ and its conjugate. Hence reconsidering equation (5), with $\hat{X}(t)$ and $\hat{X}^*(t)$ as input signals, we readily find

$$\hat{Y}(t)=\hat{Y}_{dc}(\hat{X}(t),\hat{X}^*(t))+$$
$$\int_0^\tau h_1(\hat{X}(t),\hat{X}^*(t),\lambda)[(\hat{X}(t-\lambda)-\hat{X}(t)]d\lambda+ \qquad (8)$$
$$\int_0^\tau h_2(\hat{X}(t),\hat{X}^*(t),\lambda)[(\hat{X}^*(t-\lambda)-\hat{X}^*(t)]d\lambda$$

Replacing convolution by Fourier integral in (8), we find equivalently

$$\hat{Y}(t)=\hat{Y}_{dc}(\hat{X}(t),\hat{X}^*(t))+$$
$$\frac{1}{2\pi}\int_{-BW/2}^{BW/2}\hat{H}_1(\hat{X}(t),\hat{X}^*(t),\Omega)\times\hat{X}(\Omega)e^{\mathrm{j}\Omega t}d\Omega+ \qquad (9)$$
$$\frac{1}{2\pi}\int_{-BW/2}^{BW/2}\hat{H}_2(\hat{X}(t),\hat{X}^*(t),-\Omega)\times\hat{X}^*(\Omega)e^{-\mathrm{j}\Omega t}d\Omega$$

where

BW is the modulation bandwidth,

$\hat{X}(\Omega)$ the input modulation spectrum.

$\hat{Y}_{dc}(\hat{X}(t),\hat{X}^*(t))$ is the static characteristic of the subsystem, i.e. the response of the subsystem under a non modulated carrier excitation $x(t)=\Re e[\hat{X}_0 e^{j\omega_{0x}t}]$. If we consider the case of an amplifier, this corresponds to the well known AM/AM and AM/PM curves.

$\hat{H}_1(\hat{X}(t),\hat{X}^*(t),\Omega)$ and $\hat{H}_2(\hat{X}(t),\hat{X}^*(t),\Omega)$ are the *synchronous* and *image* Volterra transfer functions of the subsystem. These can be easily computed by applying a two-tone signal of the form $x(t)=\Re e[\hat{X}_0 e^{j\omega_{0x}t}]+\Re e[\delta\hat{X}e^{j(\omega_{0x}+\Omega)t}]$, $\left|\delta\hat{X}\right|<<1$ to the component, as depicted in Fig.4.

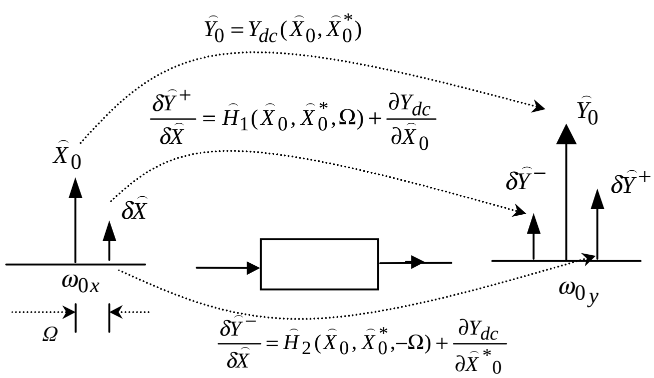

Fig.4 ***Dynamic Volterra kernels extraction set-up***

Hence all characteristics of the model can be computed by a simple two-tone (small signal mixer) HB simulation of the subsystem. The static characteristic $\hat{Y}_{dc}()$ accounts for the purely static nonlinear effects and the two nonlinear transfer functions $\hat{H}_1()$ and $\hat{H}_2()$ for the memory effects. The separation Ω between the two tones is to be swept throughout the bandwidth BW, it plays the role of modulation frequency.

Sliding kernels model

Nonlinear functions with complex conjugate variables $\hat{X}(t)$ and $\hat{X}^*(t)$ are convenient for mathematical development, but not straightforward in practice, where it is preferable to use their counter parts which are amplitude and phase: $\left|\hat{X}(t)\right|$ and $\phi_{\hat{X}(t)}$. In the case of an amplifier, taking account of the system causality, we find that:

$$\hat{Y}_{dc}(\hat{X}(t),\hat{X}^*(t))=Y_{dc}(\left|\hat{X}(t)\right|)e^{\mathrm{j}\phi_{\hat{X}(t)}} \qquad (10)$$
$$\hat{H}_1(\hat{X}(t),\hat{X}^*(t),\Omega)=H_1(\left|\hat{X}(t)\right|,\Omega)$$
$$\hat{H}_2(\hat{X}(t),\hat{X}^*(t),\Omega)=H_2(\left|\hat{X}(t)\right|,\Omega)e^{\mathrm{j}2\phi_{\hat{X}(t)}}$$

Experimenting the above presented model we have found that excellent accuracy over wider bandwidths can be achieved if the term $\hat{Y}_{dc}(..)$ is not considered purely static, but rather quasi-static, as outlined below. Reconsidering eq (2), we thus obtain a model extension termed "*sliding kernels dynamic*" Volterra series having the expression below:

$$\hat{Y}(t)=\tilde{Y}_{dc}(\left|\hat{X}(t)\right|,\frac{d\phi_{\hat{X}(t)}}{dt})e^{\mathrm{j}\phi_{\hat{X}(t)}}+ \qquad (11)$$
$$\int\tilde{H}_1(\left|\hat{X}(t)\right|,\Omega,\frac{d\phi_{\hat{X}(t)}}{dt})\times\hat{X}(\Omega)e^{\mathrm{j}\Omega t}d\Omega/2\pi \quad +$$
$$\int\tilde{H}_2(\left|\hat{X}(t)\right|,-\Omega,\frac{d\phi_{\hat{X}(t)}}{dt})e^{j2\phi_{\hat{X}(t)}}\times\hat{X}^*(\Omega)e^{-\mathrm{j}\Omega t}d\Omega/2\pi$$

where the sliding kernels $\tilde{Y}_{dc}()$, $\tilde{H}_1()$ and $\tilde{H}_2()$ are computed by sliding also the reference frequency ω_{0x} throughout the amplifier bandwidth, in the extraction procedure depicted in Fig.4. Note that $d\phi_{\hat{X}(t)}/dt$ plays the role of reference frequency shift as to the amplifier mid-band frequency ω_{mid}. For each value of ω_{0x}, the procedure of Fig.4 is carried, so that referring to (10) we have:

$$\tilde{Y}_{dc}(\left|\hat{X}(t)\right|,\frac{d\phi_{\hat{X}(t)}}{dt})=Y_{dc}(\left|\hat{X}(t)\right|)\Big|_{\omega_{0x}=\omega_{\mathrm{mid}}+\frac{d\phi_{\hat{X}(t)}}{dt}} \qquad (12)$$
$$\tilde{H}_i(..)=[H_i(\left|\hat{X}(t)\right|,\Omega)-H_i(\left|\hat{X}(t)\right|,\frac{d\phi_{\hat{X}(t)}}{dt})]\Big|_{\omega_{0x}=\omega_{\mathrm{mid}}+\frac{d\phi_{\hat{X}(t)}}{dt}}$$

APPLICATION

The new method has been used to model a 4 stage 6Watts, 3GHz MMIC MESFET amplifier. We have computed the output of the amplifier with the new model for a single and a two-tone stimulus in the 300MHz bandwidth of the amplifier. Fig.5-6 compare the output of the model with the complete circuit simulation by harmonic balance. The good accuracy and tremendous improvements made over the classical memoryless AM/AM/PM model is clearly visible, as the latter is invariably independent of frequency. As expected, the new model is well describing the circuit over a large bandwidth. Variation of intermodulation figure with mixing frequency distance is well reproduced quite deep into saturation.

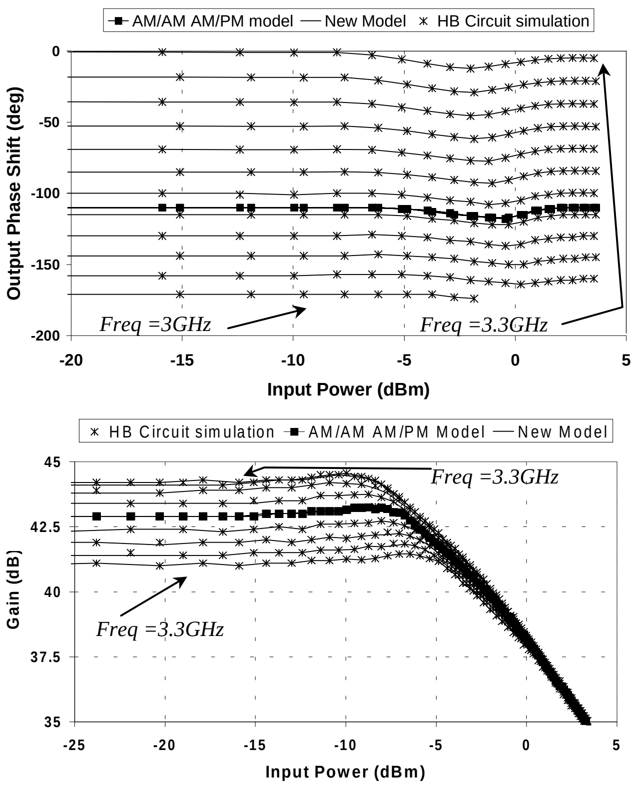

Fig.5 ***Amplifier phase shift and gain over 300MHz bandwidth, from 3GHz to 3.3GHZ and 30MHz step***

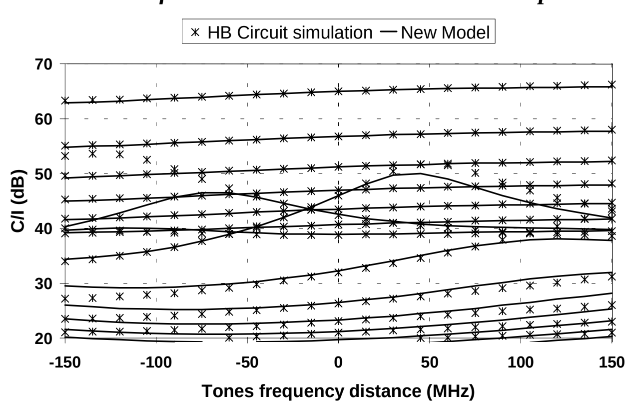

Fig.6 ***Amplifier intermodultation C/I3 figure, from -20dBm to –3dBm input (2dB gain compression)***

CONCLUSION

The paper has presented the basics of a new and powerful behavioral modeling mechanism based on modified Volterra series, so-called sliding kernels dynamic Volterra series model. The model is simple to derive from circuit simulation and also with some accommodations from load pull measurements setup. It can be applied to various functions like amplifiers, multipliers and mixers.

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