Measurement Based Modeling of Power Amplifiers for Reliable Design of Modern Communication Systems

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***Abstract* — The characterization and the modeling of nonlinear memory effects are, nowadays, an integral part of the design process of modern communication systems. Notably, the nonlinear long term memory effects occurring in solid state devices impact considerably system performances. Recently, a new method to characterize and integrate low frequency memory effects in nonlinear behavioral models of SSPAs has been presented [1]. This paper presents a detailed mathematical study and a measurement based extraction principle of the proposed behavioral model. Calibrated time-domain envelope measurements are used for the model extraction and verification procedures. The extraction technique is illustrated by the modeling of a L-Band HFET amplifier.**

I. INTRODUCTION

Long term nonlinear dispersive effects standing in solid state power amplifiers are basically due to biasing circuits embedding active cells and to thermal phenomena.

One of the main challenge of the behavioral modeling is first to integrate an accurate description of high frequency and low frequency memory effects and secondly to enable an accurate prediction of power amplifiers performances in real working conditions (digitally modulated multitones – chirps – …). It appears that, in all modern wireless applications, the distortions caused by power amplifiers result mainly from the combination of nonlinearity and long term memory effects.

Recently, a new modeling approach, based on nonlinear impulse response notion, has been presented [1]. The comparisons between circuit simulations and model predictions have shown the good capabilities of the proposed model to predict long term memory effects.

In the continuity of these works, this paper presents a detailed mathematical study of the approach showing clearly that the adopted impulse response formulation is an effective extension of the Volterra series expansion for band-pass applications modeling. This study is followed by the presentation of a calibrated procedure to extract the model from experimental data. Modeling results of a HFET L-Band mid-power amplifier, based on calibrated time-domain envelope measurement setup, are shown.

II. NONLINEAR IMPULSE RESPONSE MODEL

Consider $x(t)=\Re e\left[\tilde{X}(t).e^{j.\omega_0.t}\right]$ and $y(t)=\Re e\left[\tilde{Y}(t).e^{j.\omega_0.t}\right]$ the band-pass input and output signals of a nonlinear system. We would like to identify an effective mathematical description of the relationship between the input and output envelopes, $\tilde{X}(t)$ and $\tilde{Y}(t)$, of the system. Note that the reference frequency ω_0 in the expressions of the input/output signals, $x(t)$ and $y(t)$, is taken equal to the center frequency of the system operating bandwidth.

The output response at a given time of the system depends on input signal at the same time instant and also input signal at preceding instants in the limit of the memory duration T_m of the system. Therefore, considering a sufficiently small sampling time step, the output response can be written in the following discrete form :

$$\tilde{Y}(t_n) = f_{NL}\left(\tilde{X}(t_n), \tilde{X}^*(t_n), \cdots, \tilde{X}(t_{n-M}), \tilde{X}^*(t_{n-M})\right) \quad (1)$$

where $\tilde{X}(t_{n-i})$ represents the input signal envelope at instant t_{n-i}. The function $f_{NL}(\cdots)$ is a $(M+1)$ dimensional nonlinear characteristic.

From (1), the common approach has been to consider a power series expansion of the characteristic $f_{NL}(\cdots)$ around $\tilde{X}(t_{n-i}) = \tilde{Z}(t_n)$, $i = 0,\cdots,M$ as expressed in (2) below.

If we consider the power series expansion (2) around the time varying envelope DC state $\tilde{X}(t_{n-i}) = \tilde{X}(t_n)\ \forall i$, we obtain the modified Volterra expansion reported in [2-4], which is a powerful formalism. However, due to the complexity of the identification of the power series decomposition terms, we can practically apply it only up to the first order.

$$
\begin{aligned}
\tilde{Y}(t_n) = {} & f_{NL}\left(\tilde{Z}(t_n), \tilde{Z}^*(t_n), \cdots, \tilde{Z}(t_n), \tilde{Z}^*(t_n)\right) \\
& + \sum_{k1=0}^{M} \left(\frac{\partial f_{NL}}{\partial \tilde{X}(t_{n-k1})} \quad \frac{\partial f_{NL}}{\partial \tilde{X}^*(t_{n-k1})} \right) \times \begin{pmatrix} \Delta\tilde{X}(t_{n-k1}) \\ \Delta\tilde{X}^*(t_{n-k1}) \end{pmatrix} \\
& + \frac{1}{2} \sum_{k1=0}^{M} \sum_{k2=0}^{M} \left(\Delta\tilde{X}(t_{n-k2}) \quad \Delta\tilde{X}^*(t_{n-k2}) \right) \times \\
& \begin{pmatrix} \dfrac{\partial^2 f_{NL}}{\partial \tilde{X}(t_{n-k1}) \partial \tilde{X}(t_{n-k2})} & \dfrac{\partial^2 f_{NL}}{\partial \tilde{X}^*(t_{n-k1}) \partial \tilde{X}(t_{n-k2})} \\ \dfrac{\partial^2 f_{NL}}{\partial \tilde{X}(t_{n-k1}) \partial \tilde{X}^*(t_{n-k2})} & \dfrac{\partial^2 f_{NL}}{\partial \tilde{X}^*(t_{n-k1}) \partial \tilde{X}^*(t_{n-k2})} \end{pmatrix} \times \\
& \begin{pmatrix} \Delta\tilde{X}(t_{n-k1}) \\ \Delta\tilde{X}^*(t_{n-k1}) \end{pmatrix} \\
& + \cdots\cdots
\end{aligned} \qquad (2)
$$

with $\Delta\tilde{X}(t_{n-k}) = \tilde{X}(t_{n-k}) - \tilde{Z}(t_n)$

Although as shown in [2-4], a first order truncation of this series offers a fairly accurate modeling even for highly nonlinear systems, its efficiency tends to be limited only to those systems where the nonlinear memory duration T_m is sufficiency small compared to the inverse of the bandwidth of the envelope signal $\tilde{X}(t)$. Unfortunately, it is shown that most of solid state power modules, exhibit an effective nonlinear memory duration much longer than the inverse of their operating bandwidth especially because of bias circuit modulation effects and thermal effects [5].

In order to handle more effectively long term memory effects while considering still a first order truncation of the series, we propose in this paper a slightly modified type of expansion of the discrete characteristic in (1). That is, instead of using a power series decomposition (i.e. polynomial), we propose to use arbitrary nonlinear functions, which are determined a posteriori, i.e. :

$$
\begin{aligned}
\tilde{Y}(t_n) = {} & f_{NL}\left(\tilde{Z}(t_n), \tilde{Z}^*(t_n), \cdots, \tilde{Z}(t_n), \tilde{Z}^*(t_n)\right) \\
& + \sum_{k1=0}^{M} f_{k1}\left(\Delta\tilde{X}(t_{n-k1}), \Delta\tilde{X}^*(t_{n-k1})\right) \\
& + \sum_{k1=0}^{M} \sum_{k2=0}^{M} f_{k1k2}\left(\Delta\tilde{X}(t_{n-k1}), \Delta\tilde{X}^*(t_{n-k1}), \right. \\
& \qquad \left. \Delta\tilde{X}(t_{n-k2}), \Delta\tilde{X}^*(t_{n-k2})\right) \\
& + \cdots\cdots
\end{aligned} \qquad (3)
$$

Thus, we see that (2) is a particular case of (3) where the kernels $f_{k1\cdots kp}\left(\Delta\tilde{X}(t_{n-k1}), \cdots, \Delta\tilde{X}(t_{n-kp})\right)$ are polynomials of order p. Especially, the first term in (2) is a linear function of the signal displacement $\Delta\tilde{X}(t_{n-k1})$ and $\Delta\tilde{X}^*(t_{n-k1})$.

The first term in the expansion (3), $f_{k1}(\cdots)$, is not constrained to be a linear function of the input signal displacement, as it is the case in the power series expansion. For this reason, we can intuitively think that such a development even truncated at the first order will be more capable of predicting nonlinear effects and particularly long term memory effects as compared to a first order power series truncation. We may either consider a decomposition around the point $\tilde{Z}(t_n) = 0$ or around the point $\tilde{Z}(t_n) = \tilde{X}(t_n)$, in a similar way to respectively the classical Volterra series expansion and the dynamic Volterra series expansion. In this paper, for a sake of simplicity, we will consider a decomposition around $\tilde{Z}(t_n) = 0$. In this case, a first order truncation of the series results to :

$$
\tilde{Y}(t_n) = \sum_{k=0}^{M} f_k\left(\tilde{X}(t_{n-k}), \tilde{X}^*(t_{n-k})\right) \qquad (4)
$$

Equation (4) is a statement that the output signal at a given time instant t_n is a sum of the system responses to the individual elementary input pulses $\tilde{X}(t_{n-k})$ along the history $k = 0, \cdots, M$. It assumes that each elementary input pulse propagates nonlinearly through the system but sums up linearly.

Normalizing (4) by a constant sampling time step $\Delta t = t_n - t_{n-1}$ and taking the limit as $\Delta t \to 0$, we obtain the following integral form.

$$
\tilde{Y}(t) = \int_0^{Tm} \tilde{f}\left(\tilde{X}(t-\tau), \tilde{X}^*(t-\tau), \tau\right).d\tau \qquad (5)
$$

Now it is useful to express the envelope signal $\tilde{X}(t)$ as :

$$
\begin{aligned}
\tilde{X}(t) &\stackrel{\Delta}{=} \left|\tilde{X}(t)\right|.e^{j.\varphi_{\tilde{X}}(t)} \stackrel{\Delta}{=} \left|\tilde{X}(t)\right|.e^{j.\int_0^t \Omega_{\tilde{X}}(\tau).d\tau} \\
\tilde{X}^*(t) &\stackrel{\Delta}{=} \left|\tilde{X}(t)\right|.e^{-j.\varphi_{\tilde{X}}(t)} \stackrel{\Delta}{=} \left|\tilde{X}(t)\right|.e^{-j.\int_0^t \Omega_{\tilde{X}}(\tau).d\tau}
\end{aligned} \qquad (6)
$$

where $\left|\tilde{X}(t)\right|$ and $\Omega_{\tilde{X}}(t)$ are respectively the magnitude and the instantaneous frequency of the envelope $\tilde{X}(t)$. In (1), we expressed the output envelope $\tilde{Y}(t)$ as being a function of $\tilde{X}(t)$ and its conjugate $\tilde{X}^*(t)$. From (6), we see that we may equally express $\tilde{Y}(t)$ as a function of the magnitude $\left|\tilde{X}(t)\right|$ and instantaneous frequency $\Omega_{\tilde{X}}(t)$. So equation (6) can be rewritten, as below :

$$\tilde{Y}(t)=\int_{0}^{Tm}\tilde{f}\left(\left|\tilde{X}(t-\tau)\right|,\Omega_{\tilde{X}}(t-\tau),\tau\right).d\tau \tag{7}$$

It is interesting to rewrite (7) in the classical form of a convolution integral, by defining :

$$\tilde{h}\left(\left|\tilde{X}(t-\tau)\right|,\Omega_{\tilde{X}}(t-\tau),\tau\right)\overset{\Delta}{=}\frac{\tilde{f}\left(\left|\tilde{X}(t-\tau)\right|,\Omega_{\tilde{X}}(t-\tau),\tau\right)}{\tilde{X}(t-\tau)} \tag{8}$$

The kernel $\tilde{h}\left(\left|\tilde{X}(t-\tau)\right|,\Omega_{\tilde{X}}(t-\tau),\tau\right)$ will be termed "**nonlinear impulse response**" of the system and depends on the magnitude and the instantaneous frequency of the input envelope signal.

In narrow-band applications, which is common case in most modern communication systems, the dependence of $\tilde{h}(\cdots)$ to the instantaneous frequency is small and thus can be neglected. In such conditions, we can simplify the previous equation by neglecting the instantaneous frequency $\Omega_{\tilde{X}}(t)$, which reduces the model expression to:

$$\boxed{\tilde{Y}(t)=\int_{0}^{Tm}\tilde{h}\left(\left|\tilde{X}(t-\tau)\right|,\tau\right).\tilde{X}(t-\tau).d\tau} \tag{9}$$

Thus, the system characterization requires the extraction of the nonlinear impulse response $\tilde{h}(\cdots)$. The important thing with the new model is that its single kernel $\tilde{h}(\cdots)$ can be readily and easily obtained by using existing simulation tools or time-domain measurement setups [6], as will be outlined in the next section. We will limit ourselves below to the model equation (9) for sake of conciseness.

The observation of equation (9) shows that the nonlinear impulse response can be easily obtained by driving the system with a unit step function $x(t)=\Re e\left[X_0.U(t).e^{j.\omega_0.t}\right]$, where ω_0 (reference frequency) is set to the center frequency of the system bandwidth. The nonlinear impulse response $\tilde{h}(\cdots)$ can thus be readily obtained considering the time derivative of the envelope response $\tilde{Y}_U(\cdots)$ of the device to the step function excitation as expressed below.

$$\boxed{\tilde{h}(X_0\ ,t)=\frac{1}{X_0}\frac{\partial\left\{\tilde{Y}_U(X_0\ ,t)\right\}}{\partial t}} \tag{10}$$

III. Application to the Modeling of a HFET Amplifier

A calibrated time-domain measurement setup [7] has been built in order to characterize an HFET mid-power amplifier. This is a two stage hybrid HFET mid-power amplifier with an output power of 350 mW operating at 1.6 GHz. Time-domain envelope measurements are used to carry the kernel extraction process outlined above.

The setup is based on frequency translations of complex envelope with I/Q modulator and demodulator. The base-band signal is synthesized using an arbitrary waveform generator (AWG), as depicted in Fig. 1. Then, this signal is up-converted to the operating RF frequency of the device by the modulator. The device output signal is down-converted back to the base-band by using the demodulator, and is stored in a digital oscilloscope (TDS). Input and output envelope acquisitions are carried out simultaneously to avoid the possible drifts of phase of the local oscillator.

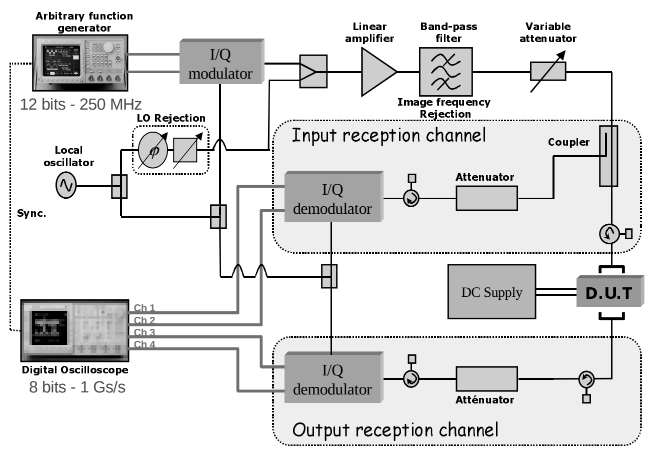
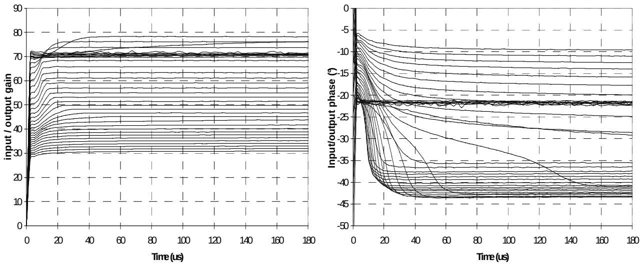


Fig. 1. Time-Domain measurement setup used for model extraction

The reception channels are calibrated to correctly de-embed envelopes at the DUT reference plans. The equipments used (Tektronix TDS 754D) provide a 40 dBc measurement dynamic, which results in non negligible errors on extracted impulse responses, that are minimized using smoothing functions.

The measured step responses of the amplifier are presented on Fig. 2 as time varying gain amplitude and phase for different input powers.

Then, comparison between amplifier measurements and the model predictions has been performed for two types of significant input stimuli. Fig. 3 shows the third order intermodulation (IM3) curves obtained as a function of the two tones frequency spacing ranging from –0.5 MHz to 0.5 MHz and an input power ranging from 0 to 3 dB gain compression.



(a) (b)

Fig. 2. Amplifier step responses : amplitude (a) and phase (b) of input/output gain

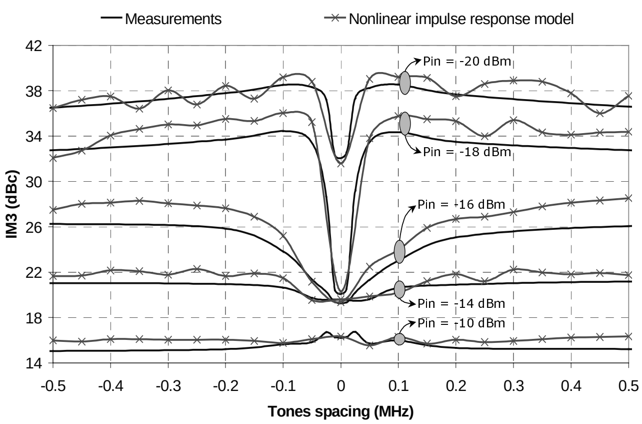


Fig. 3. IM3 measured and simulated vs tones spacing and input power

We can observe that the proposed model provides a good prediction of the nonlinearity and memory effects. The deep resonance of IM3 curves for small tones spacing shows the presence of important low frequency memory effects. The oscillations observed on the model output are a result of the small dynamic of the measurement setup.

We have also tested the model when it is driven by a QPSK modulated signal with 1 MB/s bit rate. Fig. 4 shows a short time windows of the output envelope waveform obtained by measurement and the model prediction.

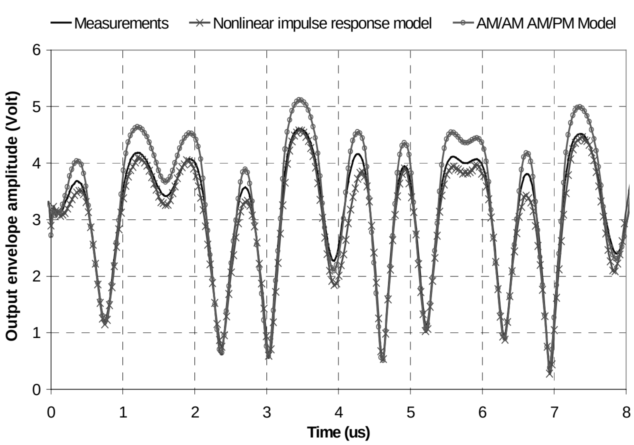


Fig. 4. QPSK Output waveform measured and simulated

A large improvement of the prediction with respect to the classical memory less model is obtained, showing the good versatility of the proposed model.

IV. Conclusion

An improved behavioral modeling technique has been presented. The proposed model provides an accurate prediction of the low frequency memory effects, which will improve reliability in system design and verification. Its extraction principle is simple and has been proved by extracting the model of a space qualified HFET amplifier from a time-domain measurement setup. The good prediction of the amplifier output obtained for two types of stimuli different from the stimuli used for model extraction have shown the effectiveness of the proposed approach.

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