Digital Predistortion of Envelope Tracking Amplifiers Driven by Slew-Rate Limited Envelopes

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***Abstract* — This paper presents a new Digital Predistorter (DPD) to compensate for nonlinear distortion that arises in Envelope Tracking (ET) Power Amplifiers (PAs) driven by slew-rate limited versions of the real signal's envelope. The slower version of the transmitted signal's envelope is used to cope with the slew-rate and bandwidth limitations of Envelope Amplifiers (EAs). Unfortunately, the use of slower versions of the real signal's envelope as the drain modulator generates a special kind of nonlinear memory effects. This paper shows experimental results that prove that is possible to compensate for these nonlinear memory effects that appear when exciting the supply of a RF linear PA with a slew-rate limited version of the envelope.**

***Index Terms* — Amplifier distortion, digital predistortion, efficiency, envelope tracking, envelope amplifier.**

I. Introduction

In the field of green communications, several efforts are being made to reduce energy consumption by enhancing the power efficiency of communications equipment. Since the Power Amplifier (PA) is one of the most power hungry devices in radiocommunications, current research is oriented at finding power efficiency structures to cope with the inherent trade-off between linearity and efficiency [1]. The improvement in power efficiency achieved using linear PAs with constant supply combined with linearization techniques is limited in comparison to the efficiency figures obtained considering dynamic supply of the PA.

PA dynamic supply modulation can be carried out by means of well-known structures such as Envelope Elimination and Restoration (EE&R) and Envelope Tracking (ET) architectures in conjunction with DPD. From the implementation point of view, ET is a very attractive technique because it can be applied in conventional transmitters based on linear RF amplification topologies by simply substituting the classical static supply for a dynamic one [2].

As shown in Fig. 1, in ET the dynamic supply is performed by an envelope amplifier (EA). The EA has to efficiently supply the required voltages and currents to the RF transistor drain at the speed imposed by the changes of the RF envelope. In signals with high PAPR the envelope bandwidth is several times (theoretically is infinite) the bandwidth of the baseband complex modulated signal. Therefore, one of the main challenges regarding the envelope drivers consist of supplying the power required by the transistor at the same speed of the signal's envelope. In order to mitigate the high slew-rate requirements of EAs, some solutions have been proposed to reduce the bandwidth (and consequently the slew-rate) of the signal's envelope [3]-[5].

In [5] and [6] a method for generating in real-time suitable envelopes (in terms of speed and bandwidth) to relax the EA slew-rate requirements is presented. Unfortunately, the use of slower versions of the real signal's envelope to supply the PA drain results in nonlinear distortion amplification. A new PA behavioral model suitable to characterize the nonlinear behavior of an ET PA when driven by a slow envelope has been recently submitted for publication in [7]. In this paper, the purpose is to go one step further and use the model proposed in [7] to design a DPD able to compensate the nonlinear memory effects associated to the dynamic supply of an ET PA when it is driven with a slew-rate limited version of the real envelope.

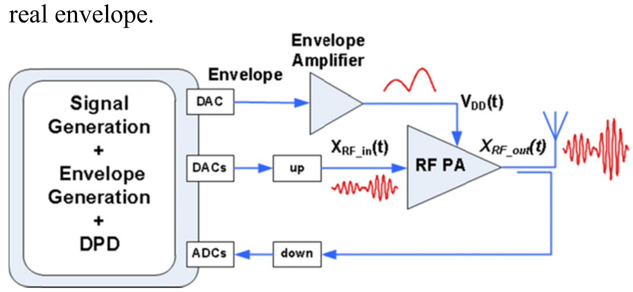

Fig. 1. Block diagram of an envelope tracking PA architecture.

II. The Slew-Rate Limited Envelope: A Compromise Between Efficiency and High Slew-Rate

The slow envelope generation method is described in [5] and applied in [6]. It basically consists of processing the real signal's envelope $E(n)$ for obtaining (in real-time) an slew-rate slower version $E_s(n)$ of the original signal's envelope. This $E_s(n)$ is calculated using the present and future values of the envelope ($E(n+1)$, $E(n+2)$, etc.). Therefore, the resulting slower envelope will suffer a delay with respect to the original modulated signal, and this delay has to be compensated synchronizing the modulated signal and the calculated slower envelope. In fact, in all ET and EE&R amplification structures it is necessary to consider some alignment mechanisms. Fig. 2 shows a RF signal, its envelope and a slew-rate limited version of the envelope considering a reduction of a factor of 100.

III. Modeling and Digital Predistortion Linearization

In [6] the slow envelope method described in [5] was used for exciting the drain of a GaN transistor, and the power efficiency obtained when exciting with the slower version of the envelope was compared with the one obtained when exciting with the real envelope. The advantage of using the slower version of the envelope is the relaxation of the EA bandwidth and slew-rate requirements, and thus current wideband modulation signals can meet the up to date technology of efficient DC-DC converters. However, in [6] we found out that when exciting with the slow envelope, nonlinear memory effects were present in the ET PA behavior. Therefore, some DPD techniques taking into account memory effects compensation, such as in [8], were needed. The characterization and modeling of these unwanted memory effects has been recently submitted to [7], where a new behavioral model was presented and validated using experimental data. As a consequence, in this paper we are using this behavioral model to estimate its corresponding DPD linearizer.

If we have a look at Fig. 3 the input-output relationship of the PA can be described as

$$y_A = g_A(|x_A|) \cdot x_A \tag{1}$$

with x_A and y_A being complex base-band modulated signals representing the PA input and output, respectively. The function $g_A(\cdot)$ is a complex gain, only dependent of the modulus (that is, the envelope) of the input signal, that characterizes the PA nonlinear behavior. The PA nonlinear complex gain function can be defined by the well-known Amplitude to Amplitude (AM-AM) and Amplitude to Phase (AM-PM) characteristics, as

$$g_A(|x_A|) = g_{A\_AM}(|x_A|) \cdot e^{j g_{A\_PM}(|x_A|)} \tag{2}$$

The static nonlinear amplitude $g_{A\_AM}(\cdot)$ and phase $g_{A\_PM}(\cdot)$ terms of the gain function can be expressed as polynomials

$$g_{A\_P}(|x_A|) = \sum_{i=0}^{N_1} c_i^{A-P} \cdot |x_A|^i \tag{3}$$

where $g_{A\_P}(\cdot)$ can be either $g_{A\_AM}(\cdot)$ or $g_{A\_PM}(\cdot)$, and where the constant coefficients c_i^{A-P} can be either c_i^{A-AM} or c_i^{A-PM}. Moreover, in our case, since the PA behavior presents an envelope-variant nonlinear gain, the coefficients c_i^{A-P} are not constant but dependent on the slower version of the envelope E_s, as it is described in [7]. Therefore, each one of these coefficients can be replaced by a polynomial ($p_i^{A-P}(E_s)$) that depends on E_s

$$\begin{aligned} g_{A\_P}(|x_A|, E_s) &= \sum_{i=0}^{N_1} p_i^{A-P}(E_s) \cdot |x_A|^i \\ &= \sum_{j=0}^{N_2} \sum_{i=0}^{N_1} \gamma_{ij}^{A-P} \cdot |x_A|^i \cdot E_s^j \end{aligned} \tag{4}$$

The same polynomial structure can be used to design the DPD. The predistortion function is obtained using the indirect learning approach shown in Fig. 3. First, a postdistortion function is extracted by means of a Least Squares fitting using the input (x_A) and output (y_A) measured data

$$\hat{x}_A = g_{DPD}(|y_A|, E_s) \cdot y_A \tag{5}$$

where $g_{DPD}(\cdot)$ is the estimated postdistorter function that will be copied and used as the predistortion function (see Fig. 3), and where $\hat{x}_A$ is the estimated PA input.

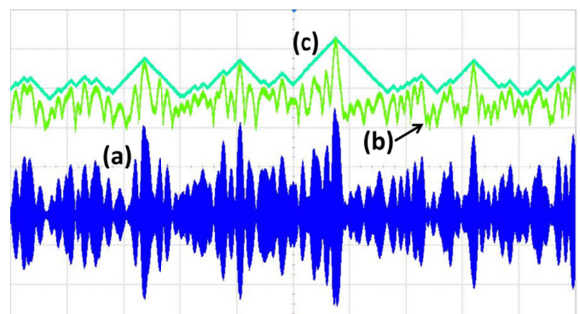

Fig. 2. Oscilloscope capture of: a) OFDM 16-QAM RF signal @2 GHz, b) real envelope and (c) slew-rate limited envelope.

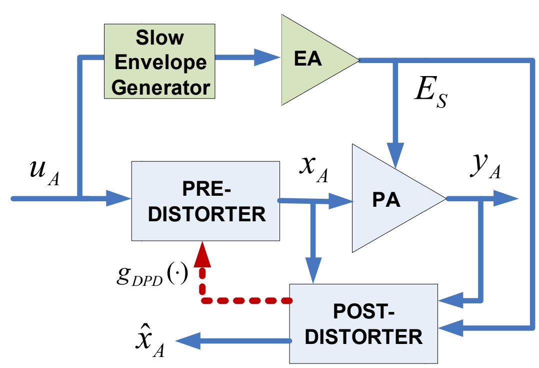

Fig. 3. Block diagram of the DPD indirect learning approach.

The postdistortion functions to compensate for both AM-AM and AM-PM time-variant nonlinear distortion can be describe as

$$g_{DPD}(|y_A|, E_s) = g_{DPD\_AM}(|y_A|, E_s) \cdot e^{j\, g_{DPD\_PM}(|y_A|, E_s)} \quad (6)$$

with the general postdistortion function defined as a polynomial that depends on the slow envelope of the signal

$$g_{DPD\_P}(|y_A|, E_s) = \sum_{i=0}^{N_1} p_i^{DPD\_P}(E_s) \cdot |y_A|^i \quad (7)$$

where $g_{DPD\_P}(\cdot)$ can be either $g_{DPD\_AM}(\cdot)$ or $g_{DPD\_PM}(\cdot)$, and each one of these terms will have the same structure as in (7). Therefore, expanding $p_i^{DPD\_P}(E_s)$ we obtain

$$g_{DPD\_P}(|y_A|, E_s) = \sum_{j=0}^{N_2} \sum_{i=0}^{N_1} \gamma_{ij}^{DPD\_P} \cdot |y_A|^i \cdot E_s^j \quad (8)$$

This expression is similar to the one used in [3] for estimating DPD values. Now, in order to reduce the error due to the finite polynomial expansion (of orders N_1 and N_2), it is possible to expand the DPD coefficients around the most probable E_s value. This value can be obtained calculating the statistical mode of the histogram of the E_s values. Therefore, the DPD gain is obtained as incremental around the nominal DPD gain (this corresponds to the nominal or most probable value, E_{s_Nom}).

The input-output relationship in the DPD, using a copy of the postdistortion function (see Fig. 3), is described as

$$x_A = g_{DPD}(|u_A|, E_s) \cdot u_A \quad (9)$$

with the generic predistortion polynomial function defined as

$$g_{DPD\_P}\left(|u_A|, E_s - E_{s\_Nom}\right) = \sum_{i=0}^{N_1} \gamma_{i0}^{DPD\_P} \cdot |u_A|^i + \sum_{j=1}^{N_2} \sum_{i=0}^{N_1} \gamma_{ij}^{DPD\_P} \cdot |u_A|^i \cdot (E_s - E_{s\_Nom})^j \quad (10)$$

The predistortion function in (10) is expressed as the combination of an static nonlinear part (coefficients $\gamma_{i0}^{DPD\_P}$) that is only dependent of the input signal; and a nonlinear memory (time-variant gain) compensation part (coefficients $\gamma_{ij}^{DPD\_P}$, $j \neq 0$) that depends on both the slow version of the envelope, E_s, and the original signal to be transmitted, u_A.

IV. Experimental Digital Predistortion Results

For testing purposes we used a Cree Inc. Evaluation Board CGH40006P-TB (GaN transistor) at 2 GHz. The signal generation and measurement equipment consist of: an Agilent MXG N5182A RF vector generator, a Tabor WW2572A arbitrary wave generator, an Agilent Infinium DSO90404A oscilloscope for capturing the signals and an Agilent N2783A probe for measuring the transistor drain current. The overall system is controlled by a PC running Matlab. We used as the EA the high-speed (35 MHz bandwidth and 900V/µs slew-rate at Av=2 and 10 Ω load) high-current (1.1 A) Linear Technology IC LT1210. For the sake of simplicity we have considered the slightly efficient IC LT1210 as the envelope driver because the scope of this work is to prove the performance of the proposed envelope-dependent DPD. Fig. 4 shows both the measured and the modeled AM-AM characteristic of the ET amplifier when it is excited with the slew-rate limited envelope. Due to the fact that the dynamic supply is performed with the slow envelope, the gain of the transistor is nonlinear and time-variant, and thus it has to be linearized with a DPD that takes into account these effects. The predistortion function proposed in (10) is able to compensate for these nonlinear memory effects as it can be observed in the linearized AM-AM characteristic shown in Fig. 5.

The in-band and out-of-band compensation achieved by the DPD can be observed in the demodulated constellations of Fig. 6 and the output power spectra of Fig. 7, respectively. The ACPR improvement was of 10 dB. In addition, Table I shows the Normalized Mean Square Error (NMSE) and Adjacent Channel Error Power Ratio (ACEPR) values for different DPD configurations (static DPD and envelope-dependent DPD with 9 and 16 coefficients). The error signal is defined as the difference between the PA output after DPD and the original input signal to be transmitted.

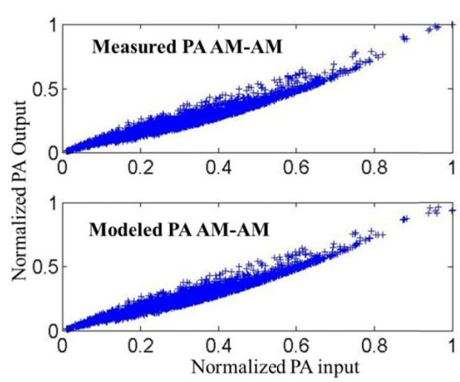

Fig. 4. AM-AM characteristics (OFDM 16-QAM modulation).

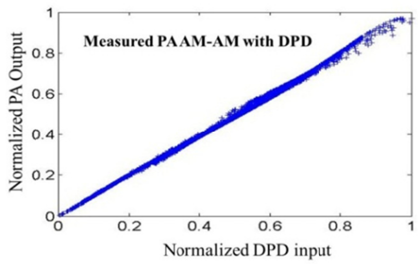

Fig. 5. AM-AM characteristic (OFDM 16-QAM modulation).

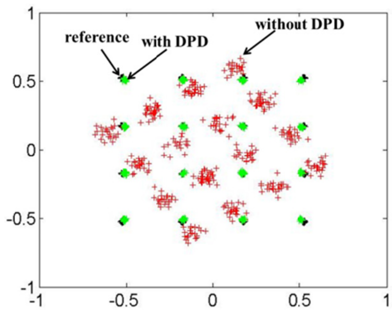

Fig. 6. OFDM 16-QAM constellations.

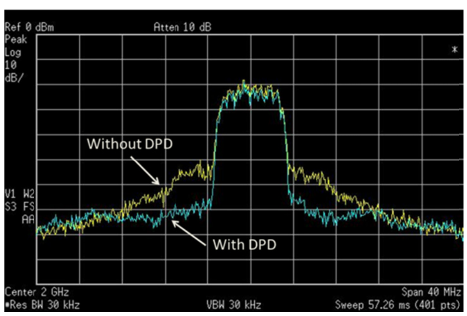

Fig. 7. Output power spectra of a 64-QAM signal (BW=5 MHz).

TABLE I
COMPARISON BETWEEN STATIC AND ENVELOPE-DEPENDENT DPD

	# of coeff.	NMSE	ACEPR
Static DPD (only γ_{i0} coeff.)	9	-28 dB	-29 dB
Envelope-dependent DPD	9	-34 dB	-37 dB
Envelope-dependent DPD	16	-35 dB	-38 dB

V. CONCLUSION

This work has shown a method for linearizing envelope tracking PAs when using as drain signal a slew-rate limited version of the real envelope of the signal. As reported in [6], the PA power efficiency using ET with the slower version of the envelope is slightly worse than using ET with the real envelope (27% versus 34%), but better than considering a fixed supply (13%). Relaxing the slew-rate requirements of efficient EAs is of crucial importance because this way is possible to extend the applicability bandwidth of some commercial efficient EAs [9].

Experimental results showed that the nonlinear memory effects arising due to the envelope-variant nonlinear gain of the ET PA when it is driven with the slow envelope are compensated with the proposed envelope-dependent DPD.

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