

Compensation of Nonlinear Distortions with Memory Effects in OFDM Transmitters

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Abstract—The Orthogonal Frequency Division Multiplexing (OFDM) modulation scheme with its many advantages, such as robustness to multipath fading, is characterized by the Gaussian-like signal behavior with a relatively high peak-to-average power ratio (PAPR). As a result, the performance of the OFDM transceivers is very sensitive to nonlinear distortions, which arise mainly from the *high power amplifier* (HPA). In addition, the wideband characteristics of the transmitted signals cause the nonlinear distortions to be frequency-dependent. This paper proposes an algebraic solution to compensate at the transmitter for nonlinearity of the HPA with memory effects, where the HPA behavioral model is represented by the Hammerstein structure, a cascade of a memoryless nonlinear block followed by a linear filter. In particular, a frequency domain parameter identification methodology is developed that first estimates the parameters of the unknown nonlinearity, which is modelled through a polynomial expansion. The frequency response of the unknown filter is then calculated, in order to capture the memory effects in the system. Using the identified nonlinear system parameters, an inverse Wiener structure, consisting of a cascade of the inverse filter and the inverse memoryless nonlinearity, is constructed preceding the HPA, in order to predistort the input signals so as to achieve overall linear transmitter characteristics.

The performance of the proposed scheme is examined through computer simulations for OFDM, though the scheme is more general and applicable to any modulation scheme with high PAPR. Specifically, improvements in the reduction of the *bit error rate* (BER) and out-of-band spectrum regrowth are demonstrated for the *travelling wave tube* (TWT) HPA model. The results show that the proposed method is effective in compensating for *amplitude-to-amplitude* (AM/AM) distortions with memory using a relatively small number of data points in the identification stage.

I. INTRODUCTION

An important objective in the design of the digital communication systems is often to minimize equipment cost, complexity and power consumption, while also reducing the bandwidth occupied by the transmitted signal and/or the transmission time. The conventional approach to increasing bandwidth utilization is to employ high-level linear modulation schemes such as M-ary *quadrature amplitude modulation* (QAM) and its multi-carrier variant OFDM which can support bit rates of $\log_2(M)$ bits/sec in 1Hz of bandwidth. However, the use of such high-level modulation schemes necessitates the use of highly linear *power amplifier* (PA), which is problematic in wireless systems where PAs operate close to the saturation point [1]. The nonlinear distortions in the *high*

power amplifier (HPA) result in such negative factors as: (i) constellation warping which increases the *bit error rate* (BER) in the system and (ii) spectrum outgrowth resulting in *adjacent channel interference* (ACI). The new communication systems also more frequently use wideband channels. This causes the HPA nonlinearity to become frequency-dependent, which means that the distortion effects differ when they affect different frequency signal components. Therefore, reliable and realizable transmitter linearization techniques are very critical to the success and acceptance of the next generation of wireless systems, which will be based on high-level modulation schemes.

This paper investigates signal processing solutions which compensate for HPA nonlinearity by introducing a *predistortion* (PD) algorithm at the input to the RF PAs represented at the transmitter side by the diagram in Fig. 1, which includes all the necessary working RF components. In the design of the

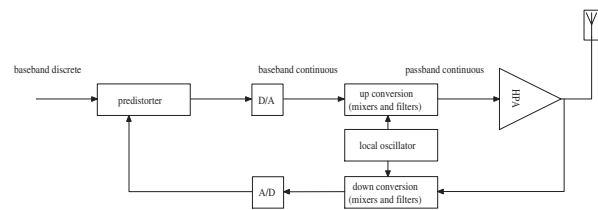


Fig. 1. A predistorter in a RF communication system.

predistorter which is the main topic of this paper, the processed signals to achieve HPA linearization and compensation are all at the baseband, although the nonlinear effects actually occur at the IF and the RF stages of the transmitter. This is a standard practice when the sampling rate of D/A and A/D converters cannot match the carrier frequency or when the DSP device processing speed is a limiting factor. Thus, in the analysis and discussions that follow in this paper, all the discrete time waveforms are either baseband or baseband equivalent (e.g., a complex envelopes of the bandpass signals). Correspondingly, all frequency responses are either lowpass functions or lowpass equivalents of bandpass functions. In order to make the presentation of the material more concise, the baseband and lowpass equivalency property of the signals and systems encountered in the following sections will not be

discussed explicitly.

The OFDM is effective in high data rate applications due to its high spectral efficiency and immunity to multipath fading and impulse noise. As a result it has been standardized in several widely accepted systems, such as DVB, DAB and WLAN. However, the modulated OFDM signal envelope assumes a Rayleigh distribution, resulting in a very large PAPR value. Consequently, it is well known that OFDM systems are far more sensitive to the nonlinear distortion usually caused by the transmitter HPA than are single-carrier systems.

While the general concept of PD techniques has been investigated in the digital communication literature quite extensively in the context of memoryless nonlinearity [2], [3], [4], this paper addresses the compensation of a cascade of a memoryless nonlinear block followed by a linear filter, i.e., a Hammerstein system, which may be applicable in the case of *digital audio broadcasting* (DAB) system in the AM band. Once the parameters of the HPA model are obtained, the inverse Wiener structure is constructed preceding the HPA, so that the overall effect after passing the signal through non-ideal amplifiers will result in acceptable system performance in terms of reduced BER and limited out-of-band spectrum spreading in the channel. It is expected that the investigations in this paper will be applicable to any radio system with multilevel modulation schemes, where there is a need to use nonlinear amplifiers with memory.

II. BACKGROUND AND PROBLEM FORMULATION

A. Zero-Memory Nonlinear System Models

If the wireless communication system is narrow-band or the PA frequency response exhibits flat characteristics over its entire working frequency range, the nonlinearity is said to be frequency-independent or memoryless. Generally, in this situation, at any instant in time PAs exhibit nonlinear distortions in both amplitude (AM/AM conversion) and phase (AM/PM conversion) which are dependent only on the current signal value. The Saleh *travelling wave tube* (TWT) PA amplitude-phase nonlinear model has the advantage of exhibiting greater simplicity and accuracy than other models and is the most widely referenced memoryless HPA model in the literature. Its AM/AM and AM/PM functions, $A(\cdot)$ and $\Phi(\cdot)$, take the following forms [5]:

$$\begin{aligned} A(r) &= \alpha_a r / (1 + \beta_a r^2) \\ \Phi(r) &= \alpha_\phi r^2 / (1 + \beta_\phi r^2) \end{aligned} \quad (1)$$

In this paper, the above TWT model is chosen as the submodel

Function	α	β
$A(r)$	2	1
$\Phi(r)$	60^0	1

TABLE I

SIMPLIFIED PARAMETERS OF THE SALEH TWT HPA AMPLITUDE-PHASE MODEL.

for the HPA nonlinear behavior with memory, and is used in

the system simulations in later sections. For this purpose, a simplified version of the parameters in (1) according to [6] will be used, as shown in Table I.

B. Nonlinear System Models With Memory

1) *Volterra system*: The Volterra series is a generalization of the Taylor series and has been used frequently for modelling nonlinear functions with memory [7]. In fact, the p th-order inverse, an inverse method based on the general Volterra series model, has already been applied to predistorter design for nonlinear telecommunication channels [8]. Basically, this method connects another Volterra system in series with the nonlinear system to be compensated in such a way that the overall system Volterra kernels with an order greater than one and less than or equal to p are zero. However, the design of a p th-order inverse system is very complex and must be based on a known Volterra series model of the nonlinear channel. Moreover, as the order of the nonlinearity increases, so does the complexity of identifying the original Volterra system and designing the p th-order inverse system. In order to make the PD design practical and implementable, the methods used in this paper are based on simplified block-oriented models, which does not require a large number of nonlinear system parameters to be identified. This approach is based on the assumption that the nonlinear system consists of relatively simple subsystems with structures which are known a priori. In particular, there are two commonly used systems, referred to as the Wiener and Hammerstein nonlinear systems, which are discussed in Sections 2.2.2 and 2.2.3. .

2) *Wiener System*: A nonlinear system with memory is considered to be a Wiener system if it can be represented by a linear dynamic subsystem (*linear time invariant* (LTI) filter) followed by a zero-memory nonlinear subsystem (see Fig. 2). The impulse response of the LTI filter is $h_{eq}(n)$, while the memoryless nonlinearity is $f(\cdot)$.

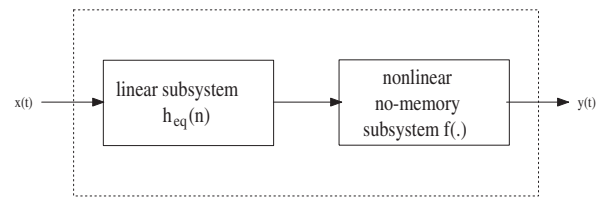


Fig. 2. Structure of a Wiener nonlinear system.

3) *Hammerstein System*: The Hammerstein system consists of the same subsystems as the Wiener system, however the subsystems are connected in the reverse order [9]. The impulse response of the LTI filter is $h_{HPA}(n)$, while the memoryless nonlinearity is $g(\cdot)$. After identifying the parameters of the Hammerstein system, the idea of the proposed cascaded compensation is to

- 1) find function $f(\cdot)$ so that

$$f((g(x))) = x \quad (2)$$

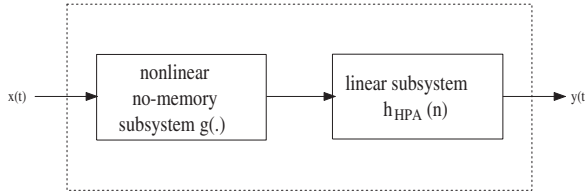


Fig. 3. Structure of a Hammerstein nonlinear system.

Equivalently, the function $f(\cdot)$ is the inverse of $g(\cdot)$, i.e., $f(\cdot) = g^{-1}(\cdot)$.

- 2) design the equalizer filter $h_{eq}(n)$ so that in the time domain

$$h_{eq}(n) \star h_{HPA}(n) = \delta(n) \quad (3)$$

where $\delta(n)$ is the unit impulse response and $\star$ represents the convolution.

In the frequency domain, the condition in (3) is equivalently represented as:

$$H_{eq}(f) \cdot H_{HPA}(f) = \text{const} \quad (4)$$

With the conditions in both (2) and (3) satisfied, the cascaded connection of the Wiener (predistorter) and Hammerstein (HPA) systems will achieve overall linear characteristics with respect to both amplitude and time.

III. PREDISTORTION DESIGN

A. Identification Algorithm

As to develop an identification method for a nonlinear system with memory, a general power series is used to represent the nonlinearity in the memoryless block of the Hammerstein system. No assumptions are made concerning the filter that represents the following linear dynamic subsystem, other than that the filter is adequately represented by its sampled frequency response. Both of these subblock representations are very general and can be adjusted by choosing a suitable order of polynomial and number of samples, so as to best approximate the true characteristics of the real wideband HPA. The proposed model is illustrated in Fig. 4, where

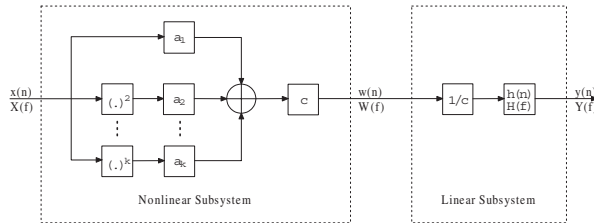


Fig. 4. System structure of a nonlinear HPA with memory effects.

$x(n)$, $w(n)$ and $y(n)$ are the sampled input, intermediate and output signal of the nonlinear HPA system with memory, respectively. A k degree complex polynomial with constant coefficients $a_1, a_2, \dots, a_k$ for the corresponding power terms is used to describe the nonlinear subsystem to include both of the

AM/AM and AM/PM nonlinearity. The constant c following the summation operation is an ambiguity scalar factor which is cancelled by another constant, $1/c$, at the front-end of the following linear dynamic subsystem. The discrete time impulse response of that linear subsystem is $h(n)$. The Fourier transforms $X(f)$, $W(f)$ and $Y(f)$ are the representations of the input, intermediate and output signals in the frequency domain. In practice, since the intermediate signal components $w(n)$ or $W(f)$ are in fact not accessible, they cannot be used in the identification algorithms. They are included here merely for the sake of analytic convenience.

From the block diagram in Fig. 4, it can be seen that to describe a specific HPA system, it is necessary to identify the polynomial coefficients $a_1, a_2, \dots, a_k$ and the impulse response $h(n)$, all being complex numbers, where it is assumed initially that the polynomial and the filter orders are known a priori.

The analytic functional relationship between the system input and output can be described either in the time domain or the frequency domain. In the time domain, the following formulae relating the system parameters are obtained:

$$\begin{aligned} w(n) &= a_1 x^1(n) + a_2 x^2(n) + \dots + a_k x^k(n) \\ y(n) &= w(n) \star h(n) \end{aligned} \quad (5)$$

where $x^1(n)$, $x^2(n)$ and $x^k(n)$ are used to represent the original input signal and its 2nd and k th power values. Similarly, in the frequency domain, the corresponding relationships are:

$$\begin{aligned} W(f) &= [a_1 X_1(f) + a_2 X_2(f) + \dots + a_k X_k(f)] \cdot c \\ Y(f) &= W(f) \cdot \left[\frac{1}{c} \cdot H(f) \right] \end{aligned} \quad (6)$$

The Fourier transforms $X_1(f)$, $X_2(f)$ and $X_k(f)$ are the corresponding counterparts in the frequency domain to $x^1(n)$, $x^2(n)$ and $x^k(n)$ which can be easily calculated based on the available $x(n)$.

To reuse the calculations in the identification and predistorter design stages so that the overall algorithm complexity is minimized, a frequency domain identification approach is proposed as follows. For the HPA Hammerstein system with a memoryless nonlinear subsystem represented by a polynomial series up to the k th degree, for identification purposes, $k+1$ batches of samples of the input and output signals are taken, with each batch having N samples in the frequency domain, which is the length of the *discrete Fourier transform* (DFT) operation on the discrete time signals. That is, the j th batch of input data is represented in the frequency domain at the f th frequency point resulting from the DFT of the input $x^m(n)$ ($m = 1, \dots, k$) as $X_{j,m}(f)$, while the j th batch of output data resulting from the DFT of the output $y(n)$ as $Y_j(f)$.

By using the relations between these DFTs as in (6) and normalizing them through the relation in the last $(k+1)$ batch to remove the effect of the dynamic subsystem frequency response $H(f)$, the following formulae are obtained:

$$\begin{aligned} b_1 &= \frac{Y_1(f)}{Y_{k+1}(f)} = \frac{[a_1 X_{1,1}(f) + a_2 X_{1,2}(f) + \dots + a_k X_{1,k}(f)] \cdot c}{[a_1 X_{k+1,1}(f) + a_2 X_{k+1,2}(f) + \dots + a_k X_{k+1,k}(f)] \cdot c} \\ &\vdots \\ b_k &= \frac{Y_k(f)}{Y_{k+1}(f)} = \frac{[a_1 X_{k,1}(f) + a_2 X_{k,2}(f) + \dots + a_k X_{k,k}(f)] \cdot c}{[a_1 X_{k+1,1}(f) + a_2 X_{k+1,2}(f) + \dots + a_k X_{k+1,k}(f)] \cdot c} \end{aligned} \quad (7)$$

The formulae in (7) hold for N frequency points, generically represented by f , resulting from the N -point DFT. For every frequency point in (7), there is always a constant c that makes the denominators equal to 1. This yields the following relations:

$$\begin{bmatrix} b_1 \\ b_2 \\ \vdots \\ b_k \end{bmatrix} = \begin{bmatrix} X_{1,1}(f) & X_{1,2}(f) & \cdots & X_{1,k}(f) \\ X_{2,1}(f) & X_{2,2}(f) & \cdots & X_{2,k}(f) \\ \vdots & \vdots & \ddots & \vdots \\ X_{k,1}(f) & X_{k,2}(f) & \cdots & X_{k,k}(f) \end{bmatrix} \times \begin{bmatrix} a_1 \\ a_2 \\ \vdots \\ a_k \end{bmatrix} \cdot c \quad (8)$$

With

$$B = [b_1 \ b_2 \ \cdots \ b_k]^T \quad (9)$$

and

$$A = [a_1 \ a_2 \ \cdots \ a_k]^T \quad (10)$$

and $^{-1}$ representing the matrix inverse, according to the matrix manipulation rules [10]:

$$A \cdot c = \begin{bmatrix} X_{1,1}(f) & X_{1,2}(f) & \cdots & X_{1,k}(f) \\ X_{2,1}(f) & X_{2,2}(f) & \cdots & X_{2,k}(f) \\ \vdots & \vdots & \ddots & \vdots \\ X_{k,1}(f) & X_{k,2}(f) & \cdots & X_{k,k}(f) \end{bmatrix}^{-1} \times B \quad (11)$$

From (11), the polynomial coefficients can be calculated using the available solutions to the system of linear equations based on the DFTs of the $k + 1$ batches of input and output signals of the HPA Hammerstein system. In order to minimize the measurement noise and numerical calculation errors, as well as to optimize the identification accuracy of A , an average of the N groups of coefficients is taken to obtain a more accurate estimate of $a_1, a_2, \dots, a_k$.

After obtaining the coefficients of the power series in the nonlinear subsystem, to obtain the frequency response of the linear dynamic subsystem at the frequency point f , (6) is manipulated further to obtain:

$$\begin{aligned} H_1(f) &= \frac{Y_1(f)}{[a_1 X_{1,1}(f) + a_2 X_{1,2}(f) + \cdots + a_k X_{1,k}(f)]} \\ &\vdots \\ H_k(f) &= \frac{Y_k(f)}{[a_1 X_{k,1}(f) + a_2 X_{k,2}(f) + \cdots + a_k X_{k,k}(f)]} \end{aligned} \quad (12)$$

Based on (12), there are k groups of frequency response functions, each having N frequency sampling points. Again, to obtain a reliable estimate of $H(f)$, an average of k frequency responses derived from k different batches is taken.

The total number of input/output signal pairs required in order to identify the HPA system with memory is: $(k+1) \times N$. If the order of the polynomial series is low and the length of the FIR filter capturing memory effects is short, this represents a very small number of signal samples, which means that the proposed identification method could be very effective.

B. Compensation Algorithm

As explained in Section 2, the most suitable compensator for the nonlinear distortions of a Hammerstein system is a Wiener system which has the inverse characteristics. The proposed compensation scheme including the identification stage is shown in a complete PD design system block diagram in Fig. 5.

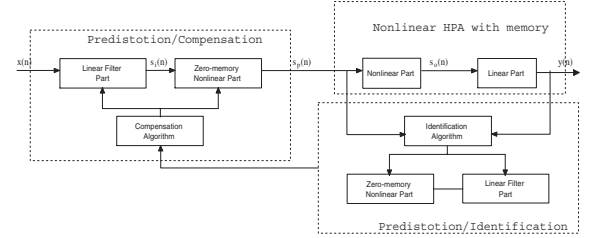


Fig. 5. System structure of the proposed PD.

Designing a filter inverse to that identified in the Hammerstein system, zero-forcing is used to design the inverse filter. Specifically, since the frequency response function of the linear subsystem in the HPA has already been found in the identification process by (12), the frequency-dependent distortion effect can also be eliminated in the frequency domain simply by dividing the Fourier transform of the input signal entering the predistorter on a sample-to-sample basis by the function which corresponds to the use of the FIR filter with an impulse response of length N . To design a polynomial series inverse to that identified in the Hammerstein system could require the use of very high order polynomials and thus be impractical in DSP applications. Therefore, in order to implement the inverse of the polynomial series function, a more practical DSP solution, constructing an inverse look-up table (LUT) is proposed. Since HPA AM/AM distortion is the main factor responsible for communication system performance degradation, a computationally efficient compensation scheme is proposed here only for the AM/AM nonlinearity resulting from the identified complex polynomial memoryless nonlinear function included in the complete HPA with memory modelling. If the input signal is represented in the polar form $x = r \cdot e^{j\psi}$, where $r = |x|$ and $e^{j\psi} = \frac{x}{|x|}$, after passing it through the Saleh memoryless amplitude-phase nonlinear model, represented as function $g(\cdot)$, the following output signal y is obtained:

$$y = g(x) = A(r) \cdot e^{j(\psi + \Phi(r))} \quad (13)$$

where $A(r)$ and $\Phi(r)$ are the normalized AM/AM and AM/PM nonlinear functions for the amplitude-phase HPA model. The absolute value of the output signal y can then be represented as

$$|y| = |A(r)| = A(r) = b_1 r + b_2 r^2 + b_3 r^3 + \cdots \quad (14)$$

where the normalized signal envelope r assumes values from 0 to 1. In the above formula, a real polynomial is used to approximate the memoryless HPA AM/AM nonlinear function

$A(r)$ with real coefficients $b_1, b_2, b_3, \dots$. In fact, by using the developed identification algorithm, another memoryless nonlinear complex polynomial has already been obtained:

$$\hat{g}(x) = a_1x + a_2x^2 + a_3x^3 + \dots \quad (15)$$

which is an approximation to $g(x)$, where the coefficients $a_1, a_2, a_3, \dots$, assume complex values. A set of input signals x is constructed with phase value equal to 0, or $x = r \cdot e^{j \cdot 0} = |x|$. These are then substituted into (15) and the absolute value is taken, yielding

$$\begin{aligned} |\hat{g}(x)|_{x=r \cdot e^{j \cdot 0}} &\approx |g(x)| = A(r) = b_1r + b_2r^2 + b_3r^3 + \dots \\ &= |a_1xr + a_2r^2 + a_3r^3 + \dots| \end{aligned} \quad (16)$$

With complex coefficients $a_1, a_2, a_3, \dots$ being calculated using the identification algorithm developed at the beginning of the section, a set of coefficients $b_1, b_2, b_3, \dots$, capturing AM/AM distortions, is obtained by using the MMSE polynomial fitting method. The number of the calculated coefficients b_k is adjusted to achieve the required accuracy which will satisfy the particular application performance requirements in terms of BER.

In order to obtain more accurate approximations of the real coefficients $b_1, b_2, b_3, \dots$, additional input signals are substituted with different phase values of $\frac{1}{n} \cdot 2\pi, \frac{2}{n} \cdot 2\pi, \dots, \frac{n-1}{n} \cdot 2\pi$, and the MMSE polynomial fitting method is used to obtain a total of n sets of approximated coefficients $b_1, b_2, b_3, \dots$. In the proposed method, for the sake of efficiency, 8 slices of input signals are taken in the complex domain, or $x = r \cdot e^{j \cdot \frac{1}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{2}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{3}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{4}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{5}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{6}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{7}{4} \cdot \pi}, r \cdot e^{j \cdot \frac{8}{4} \cdot \pi}$, for the purpose of approximating the AM/AM function real polynomial. The average of the 8 sets of coefficients is then taken to minimize the measurement noise and numerical calculation errors.

Hence, an inverse LUT can be constructed to eliminate the lower order AM/AM nonlinear distortions which are included in the real polynomial approximation in (16).

IV. SIMULATION RESULTS

In this section, simulations are used in order to demonstrate the validity of the proposed PD technique in compensating for the HPA nonlinear and memory effects in the OFDM system. A total of 256 subcarriers are used to modulate/demodulate the 16-ary QAM complex signals in the subcarriers. The IBO values of 8dB and 6dB are selected to verify the effectiveness of the proposed PD algorithm.

In each independent simulation trial, 1024 bits are fed into a 16-ary QAM digital baseband modulator, and from these 256 16-ary QAM symbols are created and passed through the 256-point IFFT. These complex signals are then upsampled 8 times ($F_s/F_d = 8$) and filtered as they pass through the transmitter root raised cosine (RRC) filter, with a rolloff factor of $\alpha = 0.5$ for pulse shaping or band limiting purposes, where F_s and F_d are the OFDM symbol and sampling rates, respectively. In order to facilitate a close examination of the effectiveness of the predistorter in compensating for the nonlinear distortion,

it was assumed that the multipath fading effect is completely compensated, resulting in an AWGN channel. At the receiver side, the corresponding inverse operations are performed. A nonlinear system with memory effect, constructed with a Saleh amplitude-phase TWT HPA model followed by a 3-tap FIR filter with the coefficients [0.7692, 0.1538, 0.0769] [11], represents the original distortion source to be compensated. The inverse LUT constructed to cancel the HPA AM/AM nonlinear distortion is designed with a step size of 0.01 and an amplitude ranging from 0 to 1, which corresponds to the HPA input saturation voltage value. By examining the results of simulated signal samples under the conditions of a linear HPA and a nonlinear HPA with and without compensation, different aspects of the PD performance are analyzed and compared in terms of BER and spectrum outgrowth in the tail of power spectral density (PSD). Based on extensive simulations and complexity concerns, 5-order complex polynomial is selected in the identification stage to model the structure of the memoryless nonlinear subsystem to be identified.

A. BER Results

To obtain the BER performance curves, for each of the three systems: linear, nonlinear without compensation, and nonlinear with compensation, the simulations were repeated 1000 times, and a total number of $256 \times 4 \times 1000 = 1024000$ bits was generated and collected. The linear HPA system is simulated here for the purpose of verifying the *Matlab*[®] code. Figures 6 and 7, compare the BER performance of different systems for IBO of 8 and 6 dB, respectively. From both figures, it is clear that the nonlinear HPA distortion with memory effects severely degrades the system performance. With the aid of the proposed PD method, however, the system performance can be greatly enhanced. At an IBO of 8dB, the BER approaches 10^{-5} for a SNR value of 18dB, which constitutes a great improvement. When the IBO value is decreased to 6dB, which correspond to working with more non-linear system, the BER increases to a little below 10^{-4} for the same SNR value.

B. Spectrum Spreading Results

In Figs 8 and 9, the PSDs of the output signals of the linear HPA and the nonlinear HPA with and without compensation are shown for the OFDM system for IBO of 8 and 6 dB, respectively. In both figures, it can be seen that when a 5-order complex polynomial is used for the predistorter design, decreases of nearly 18dB and 15dB in the PSD value are obtained as compared with the uncompensated situation, for IBO values of 8dB and 6dB, respectively. This can dramatically reduce the ACI distortion. As compared with the ideal situation, even with compensation, there is still an increase in the PSD value of about 10dB. However, in accordance with the trend indicated in these two figures, a further increase in the IBO value can further reduce the spectrum spreading.

V. CONCLUSIONS

In this paper, we have proposed and analyzed a baseband digital compensation method for nonlinear with memory distortions in digital transmitters. The novelty of this work stems

from the fact that the proposed scheme is non-parametric with respect to the HPA memoryless nonlinearity and the cascaded filter, and can therefore cope with any invertible Hammerstein structure. In addition, the identification methodology which has been developed is independent of the specific modulation format [12].

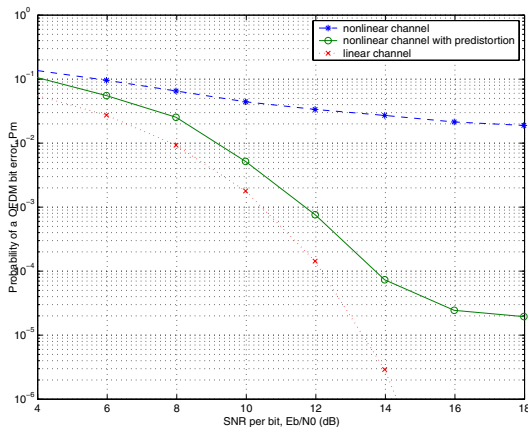


Fig. 6. OFDM BER curves for a linear HPA, an uncompensated nonlinear HPA, and a nonlinear HPA compensated with the proposed 5-order predistorter, at an IBO value of 8 dB.

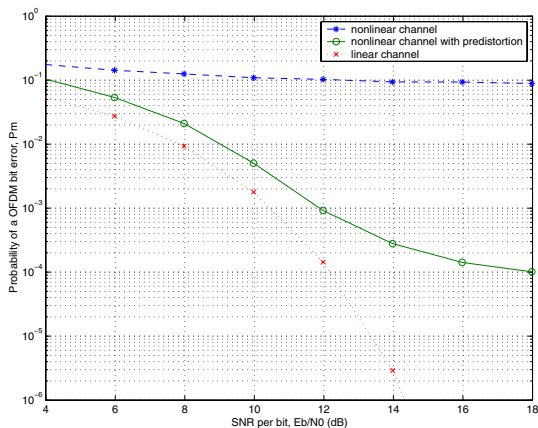


Fig. 7. OFDM BER curves for a linear HPA, an uncompensated nonlinear HPA, and a nonlinear HPA compensated with the proposed 5-order predistorter, at an IBO value of 6 dB.

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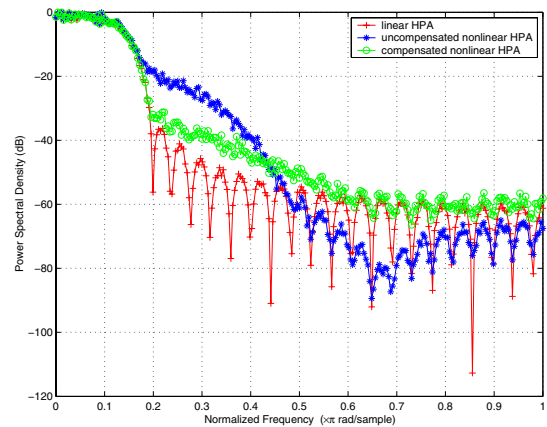


Fig. 8. OFDM PSD curves for a linear HPA, an uncompensated nonlinear HPA, and a nonlinear HPA compensated with the proposed 5-order predistorter, at an IBO value of 8 dB.

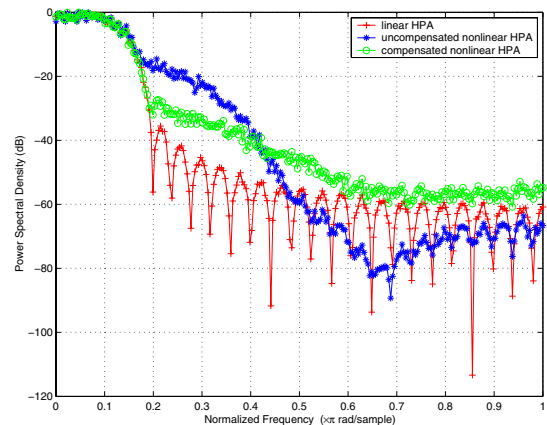


Fig. 9. OFDM PSD curves for a linear HPA, an uncompensated nonlinear HPA, and a nonlinear HPA compensated with the proposed 5-order predistorter, at an IBO value of 6 dB.