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# BROADBAND POWER AMPLIFIER DISTORTION CANCELLATION WITH MODEL ESTIMATION IN THE RECEIVER

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## ABSTRACT

This paper proposes a practical receiver cancellation technique for removing nonlinear power amplifier (PA) distortion in OFDM systems. By performing the estimation of the PA model parameters at the receiver, the implementation complexity of the transmitter can be reduced. Furthermore, a simple adaptation rule is provided to enable tracking of the PA model parameters. As a consequence, cancellation of nonlinear distortion can be achieved without assuming that the PA model is known *a priori* at the receiver. Simulation results show that good levels of distortion cancellation are possible for a system with a broadband PA with memory.

## 1. INTRODUCTION

The principal drawback of OFDM system performance is the high Peak-to-Average Power Ratio (PAPR). Real power amplifiers (PAs) have a nonlinear input-output characteristic causing signal compression and clipping that result in signal distortion and adjacent channel interference. Furthermore, broadband PAs introduce memory which gives rise to intersymbol interference (ISI).

Power backoff and PAPR reduction techniques can reduce the nonlinear distortion level (signal compression and clipping) but are incapable of compensating the memory effects of broadband PAs which is necessary for achieving good performance in wireless systems. Furthermore, operating the PA at high back-off levels results in a solution with low power efficiency. Predistortion (PD) techniques [1] show good results with reduced complexity if the PA is memoryless. However, if PA memory effects are accounted for in the PD implementation, the complexity of the associated PA model identification increases considerably. Furthermore, the PA characteristics

may change with time due to temperature or bias point variation [2]. In such case, the PA model needs to be continuously tracked (or re-estimated) [1].

In uplink transmission, it is desirable to move the computationally intensive PA identification and nonlinear distortion suppression to the receiver in base station, where more resources are available, i.e., in terms of power consumption and hardware complexity. For this purpose, iterative nonlinear distortion cancellation techniques applicable at the receiver side have been proposed [3] [4]. These iterative techniques employ a model of the PA to estimate the distortion and cancel it from the received signal. Previous works have assumed a known PA model or that its parameters are estimated at the transmitter and sent to the receiver during the initialization.

In this paper we consider the more practical problem of estimating and tracking the PA model in the receiver. The advantage of our approach is twofold. Firstly the computationally involved PA identification (and tracking) at the transmitter is avoided. Secondly, PA model estimation can be done fully in digital domain. This should be compared to PA identification at the transmitter, where the input-output measurements required to estimate and track the PA model include a feedback loop from analog domain (PA output) to digital domain. On the other hand, the PA model estimation proposed here needs to take into account the estimation and equalization of the time-varying channel. Therefore, we present here the necessary modifications of the nonlinear distortion technique in [3] to function with estimated PA model and channel parameters. In the proposed method, an initial PA model estimation is performed by exploiting an OFDM pilot symbol with uniform amplitude distribution (in time domain). Thereafter, changes in the PA characteristics are tracked by employing an adaptive algorithm. The joint estimation and tracking of the time-varying channel and PA memory effects are performed by employing dedicated pilot carriers in OFDM.

## 2. SYSTEM MODEL

The transmission model used in this paper is illustrated in Figure 2 (a). The OFDM system under consideration has  $N$  sub-

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carriers and the OFDM symbol entering the PA at time  $n$ ,  $\mathbf{x}_{cp}(n) = [x(n, 1) \cdots x(n, N + v)]^T$ , is given by

$$\mathbf{x}_{cp}(n) = \mathbf{G}_{cp}\mathbf{x}(n) = \mathbf{G}_{cp}\mathbf{Q}_N\boldsymbol{\chi}(n) \quad (1)$$

where  $\mathbf{G}_{cp}$  is the  $(N + v) \times N$  cyclic prefix insertion matrix,  $v$  is the length of the cyclic prefix,  $N + v$  is the total length of the OFDM symbol,  $\mathbf{x}(n)$  is the IDFT of the modulated symbols  $\boldsymbol{\chi}(n) \in \mathbb{C}^{N \times 1}$ , and  $\mathbf{Q}_N$  is the  $N \times N$  IDFT matrix.

The signal  $\mathbf{x}_{cp}(n)$  is then passed through a PA here modeled with a Hammerstein structure [1], see Figure 2 (b). The Hammerstein model is formed by a nonlinear static block  $g[\cdot]$  followed by a linear filter  $\mathbf{c}(n)$ , here modeled with an FIR filter (possibly time-varying) with  $L_c$  taps, i.e.,

$$\mathbf{c}(n) = [c_1(n), \cdots, c_{L_c}(n)]^T.$$

The multicarrier signal after the static nonlinearity  $g[\cdot]$  can be represented as (see Figure 2)

$$\mathbf{x}_s(n) = g[\mathbf{x}_L(n)] = K_L \mathbf{x}_{cp}(n) + \mathbf{d}(n) \quad (2)$$

where the first term  $K_L \mathbf{x}_{cp}(n)$  is the distortion-free discrete input-signal vector of Eq. (1) and  $K_L$  is the power amplifier gain. The second term  $\mathbf{d}(n)$  is an  $(N + v) \times 1$  describing the nonlinear distortion which is a function of the modulated symbol vector  $\boldsymbol{\chi}(n)$  and the power amplifier transfer function  $g[\cdot]$ . Under the assumption of low clipping levels [5], the elements of  $\mathbf{d}(n)$  are accurately modeled as AWGN processes with variance  $\sigma_d^2$ .

Finally, the signal passes through the linear PA memory block  $\mathbf{c}(n)$ . Assuming that  $L_c$  is smaller than the cyclic prefix  $v$ , the PA output can be written as (after cyclic prefix removal)

$$\mathbf{x}_g(n) = \mathbf{C}(n)\mathbf{x}_s(n) \quad (3)$$

where  $\mathbf{C}(n)$  is a circular convolution matrix built from vector  $\mathbf{c}(n)$ . The transmitted frequency domain signal is written as

$$\boldsymbol{\chi}_g(n, k) = K_L \mathbf{c}(n, k) [\boldsymbol{\chi}(n, k) + \mathbf{d}(n, k)] \quad (4)$$

where  $\boldsymbol{\chi}(n, k)$  is the transmitted symbol at sub-carrier  $k$ ,  $\mathbf{d}(n, k)$  denotes the nonlinear distortion on sub-carrier  $k$  that is given by the  $k$ th element of vector  $\mathbf{Q}_N \mathbf{d}(n)$ , and  $\mathbf{c}(n, k)$  denotes the  $k$ th component of the  $N \times 1$  frequency response vector of the PA memory, i.e.,

$$\mathbf{c}(n) = [c(n, 1) \cdots c(n, N)]^T = \mathbf{Q}_N^H \begin{bmatrix} \mathbf{c}(n) \\ \mathbf{0}_{N-L} \end{bmatrix}. \quad (5)$$

Finally, the signal is passed through a time-varying channel  $\mathbf{h}(n) \in \mathbb{C}^{L_h \times 1}$ . The cyclic prefix is chosen larger than the length of the effective channel  $\mathbf{h}_{eff}(n) = \mathbf{c}(n) * \mathbf{h}(n)$ , i.e.,  $v > L_{eff} = L_c + L_h$ . Then, the received signal at sub-carrier  $k$ , at time  $n$ , after cyclic prefix removal can be expressed as

$$y(n, k) = \mathbf{h}_{eff}(n, k) [K_L \boldsymbol{\chi}(n, k) + \mathbf{d}(n, k)] + \mathbf{n}(n, k) \quad (6)$$

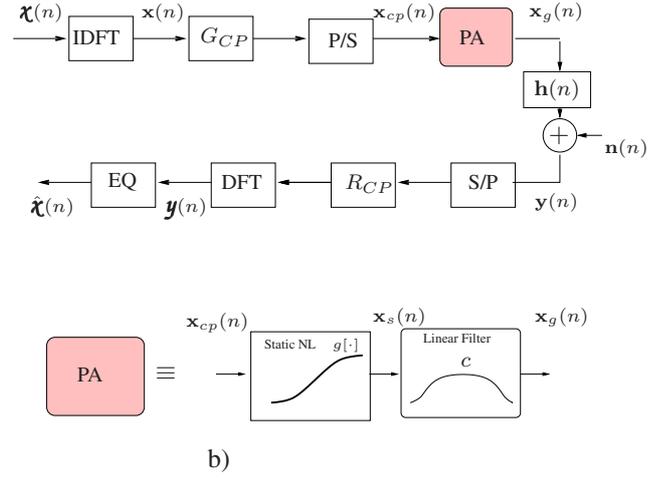


Fig. 1. a) System model b) Hammerstein-type PA model.

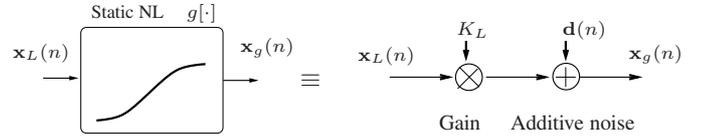


Fig. 2. Equivalent nonlinear model for low clipping levels under Gaussian assumption [5] for the nonlinear distortion

where  $\mathbf{n}(n, k)$  is the additive noise, assumed here to be circular complex Gaussian with variance  $\sigma_n^2$ , and  $\mathbf{h}_{eff}(n, k) = \mathbf{c}(n, k)\mathbf{h}(n, k)$  is the effective channel.

With perfect knowledge of the effective channel, the equalized frequency-domain signal is given by

$$\tilde{y}(n, k) = K_L \boldsymbol{\chi}(n, k) + \mathbf{d}(n, k) + \mathbf{n}(n, k) / \mathbf{h}_{eff}(n, k) \quad (7)$$

We see from Eq. (7) that the nonlinear distortion term  $\mathbf{d}(n, k)$  affects the decision variable in an additive way introducing potential errors in the detection process. In fact, the effective SNR after frequency-domain equalization is

$$SNR = \frac{K_L^2 \sigma_x^2}{\sigma_d^2 + \sigma_n^2 / \mathbf{h}_{eff}^2(n, k)} \quad (8)$$

where  $\sigma_d^2 = E[\mathbf{d}\mathbf{d}^*]$ . From Eq. (8), it is clear that the effective SNR is upper bounded by  $K_L^2 \sigma_x^2 / \sigma_d^2$  which in turn causes an irreducible error floor. Next we consider how to reduce the effect of  $\mathbf{d}(n, k)$ .

### 3. NONLINEAR PA DISTORTION CANCELLATION

The PA nonlinearity cancellation (PANC) technique proposed in [3] considered the case of a memoryless PA whose model parameters were perfectly known. The idea of the PANC

technique is to first use an initial estimate of  $\hat{\chi}(n, k)$  to estimate the distortion  $\hat{\mathbf{d}}(n, k)$ . Then the distortion term is removed from the original received signal and an improved estimate of  $\hat{\chi}_G(n, k)$  is obtained. This second estimation  $\hat{\chi}_G(n, k)$  can be used to re-estimate the distortion vector. The process can be performed iteratively. The described cancellation technique requires the knowledge of the static nonlinearity  $g[\cdot]$  and the effective channel  $\hat{h}_{eff}(n, k)$ . These issues are addressed in Section 4.

Below we summarize the steps of the PANC technique<sup>1</sup>:

At iteration  $m$

- 1) Estimate symbols  $\hat{\chi}^{(m)}(n, k) = \left\langle \frac{y(n, k)}{\hat{h}_{eff}(n, k)} - \hat{\mathbf{d}}(n, k) \right\rangle$
- 2) Switch to time domain  $\hat{\mathbf{x}}^{(m)}(n) = \mathbf{Q}_N \hat{\chi}^{(m)}(n)$
- 3) Estimate distortion term  $\hat{\mathbf{d}}(n) = \hat{g}[\hat{\mathbf{x}}^{(m)}(n)] - \hat{\mathbf{x}}^{(m)}(n)$
- 4) Distortion in frequency domain  $\hat{\mathbf{d}}(n) = \mathbf{Q}_N^H \hat{\mathbf{d}}(n)$
- 5) New iteration

Note that at iteration  $m = 0$ , the distortion term is set to zero. The new distortion terms are taken from vector  $\hat{\mathbf{d}}(n) = [\hat{\mathbf{d}}(n, 1) \cdots \hat{\mathbf{d}}(n, N)]^T$ . For more details on the PANC technique, see [3].

#### 4. PA MODEL AND CHANNEL ESTIMATION AT THE RECEIVER

In this section, a two-step procedure is proposed for the PA model estimation at the receiver. The transmission starts with two OFDM pilot symbols used for initialization. The first symbol has a low PAPR to enable accurate estimation of the effective channel  $\hat{h}_{eff}(n, k)$  without the influence of the nonlinear static block  $g[\cdot]$ . Thereafter, an OFDM symbol with uniform amplitude distribution is exploited to obtain a good estimate of  $g[\cdot]$ . Tracking of  $\hat{h}_{eff}(n, k)$  and  $g[\cdot]$  is also discussed.

A flow chart of the whole PANC method including PA model and channel tracking is provided in Figure 3 where references to key equations are included.

##### 4.1. Estimation and tracking of effective channel $\hat{h}_{eff}(n, k)$

###### 4.1.1. Initial estimation

It is important that training symbols are not affected by  $g[\cdot]$  to obtain a reliable estimate of  $\hat{h}_{eff}(n, k)$ . It is known that the probability density function (PDF) of an OFDM signal with a large number of carriers is well approximated with a Gaussian distribution  $\mathcal{N}(0, N\sigma^2)$ . Therefore, if only a group  $L < N$  carriers are active, the variance of the OFDM signal is reduced and the PA mostly operates in the linear region. As a consequence there is a tradeoff between the number of pilot carriers, the channel estimation accuracy, and the level of

<sup>1</sup> $\langle \cdot, \cdot \rangle$  denotes the symbol decision process in which the decoder chooses the value that is closest to  $\hat{\chi}(n, k)$  in Euclidean distance

nonlinear distortion. For the frequency domain channel estimation, we will assume comb-type pilot arrangement where a set of  $T$  dedicated pilot carriers are reserved for pilot data. The channel frequency response on these subcarriers without nonlinear distortion  $\hat{\mathbf{d}} = 0$  (i.e., low PAPR symbol), is obtained as

$$\hat{h}_p(n, k) = \frac{y(n, k)}{\chi(n, k)}, \forall k \in \mathcal{T} \quad (9)$$

where  $\mathcal{T} = \{k_1, \dots, k_T\}$  denotes the index set specifying the  $T$  pilot carriers with  $\{k_p\}_{p=1}^T$  taken from the set  $\{1, \dots, N\}$ . The whole frequency domain channel response is obtained through interpolation using truncated DFT matrices [6]

$$\hat{\mathbf{h}}_{eff}(n) = \mathbf{Q}_N [\mathbf{Q}_T^H \mathbf{Q}_T]^{-1} \mathbf{Q}_T^H \hat{\mathbf{h}}_p(n) \quad (10)$$

where matrix  $\mathbf{Q}_T$  is constructed by the  $T$  columns of  $\mathbf{Q}_N$  specified by  $\mathcal{T}$ , and  $\hat{\mathbf{h}}_p(n) = [\hat{h}_{k_1}(n, k) \cdots \hat{h}_{k_T}(n, k)]^T$ .

###### 4.1.2. Tracking

During normal transmission, nonlinear distortion will be present in Eq. (7). With an estimate of the nonlinear distortion at hand (see Section 3), it follows directly from Eq. (7) that Eq. (9) can be modified to (see [7] for details)

$$\hat{h}_p(n, k) = \frac{y(n, k)}{\chi(n, k) + \hat{\mathbf{d}}(n, k)}, \forall p \in \mathcal{T} \quad (11)$$

where  $\hat{\mathbf{d}}(n, k)$  is obtained from the output of the PANC detector. Then, Eq. (10) gives the channel on other sub-carriers.

##### 4.2. Estimation and tracking of static nonlinearity $g[\cdot]$

The static nonlinearity can be modeled using a memoryless polynomial with  $P$  coefficients. We assume that the parameters of the static nonlinearity remain constant over one OFDM symbol (i.e., possible time variation on symbol level  $n$ ). The output signal of the static nonlinearity at the  $j$ th subsymbol at symbol time  $n$ ,  $x_g(n, j)$ , can be modeled as

$$x_g(n, j) = \hat{g}[x(n, j)] = \sum_{k=0}^P a_k(n) |x(n, j)|^k = \mathbf{u}^T(n, j) \mathbf{a}(n) \quad (12)$$

where

$$\mathbf{a}(n) = [a_0(n) \ a_1(n) \ \cdots \ a_P(n)]^T$$

$$\mathbf{u}(n, j) = [1 \ |x_g(n, j)| \ |x_g(n, j)|^2 \ \cdots \ |x_g(n, j)|^P]^T. \quad (13)$$

###### 4.2.1. Initial estimation

In order to get a reliable initial estimate of  $g[\cdot]$ , a properly designed training symbol is needed. During the transmission of useful data, the amplitude of the OFDM sub-symbols will be Rayleigh distributed. In other words, the number of time

instants with signal levels near the clipping region are few. Thus, accurate modeling with conventional OFDM symbols requires a high number of training symbols which is not practical in a real system. To overcome this problem, a particular set of frequency domain symbols has been designed such that the OFDM symbol (in time domain) has a uniformly distributed amplitude, for details, see [8]. By exploiting such a training symbol, designed to have adequate dynamic range, the model in (12) can be estimated at the receiver as

$$\hat{\mathbf{a}}(n) = [\mathbf{U}^H(n)\mathbf{U}(n)]^{-1} \mathbf{U}^H(n)\tilde{\mathbf{y}}(n) \quad (14)$$

where  $\tilde{\mathbf{y}}(n)$  is the  $N \times 1$  time-domain received signal after equalization and  $\tilde{\mathbf{U}}(n)$  is an  $N \times P$  *Vandermonde* matrix constructed with the  $N$  input vectors  $\{\mathbf{u}(n, j)\}_{j=1}^N$ , i.e.,

$$\mathbf{U}(n) = \begin{bmatrix} \mathbf{u}^T(n, 1) \\ \mathbf{u}^T(n, 2) \\ \vdots \\ \mathbf{u}^T(n, N) \end{bmatrix} \quad (15)$$

$$\begin{aligned} \tilde{\mathbf{y}}(n) &= [\tilde{y}(n, 1) \tilde{y}(n, 2) \cdots \tilde{y}(n, N)]^T \\ &= \mathbf{Q}_N [\tilde{\mathbf{y}}(n, 1) \tilde{\mathbf{y}}(n, 2) \cdots \tilde{\mathbf{y}}(n, N)]^T. \end{aligned}$$

#### 4.2.2. Model tracking

To track variations in PA between consecutive OFDM symbols, we employ a normalized least mean-square (NLMS) algorithm operating on the OFDM subsymbols, i.e., sample-by-sample processing. The update recursion for  $\hat{\mathbf{a}}(n)$  are summarized below:

- 1) Initialize for each  $n$ :  $\tilde{\mathbf{a}}(n, 0) = \hat{\mathbf{a}}(n - 1)$ .
- 2) For each  $j$ , update  $\tilde{\mathbf{a}}(n, j)$  according to:

$$\begin{aligned} \tilde{\mathbf{a}}(n, j + 1) &= \tilde{\mathbf{a}}(n, j) + \mu \frac{\mathbf{u}(n, j)e^*(n, j)}{\mathbf{u}^H(n, j)\mathbf{u}(n, j)} \\ e(n, j) &= y(n, j) - \tilde{\mathbf{a}}^H(n, j)\mathbf{u}(n, j) \end{aligned} \quad (16)$$

- 3) Finally, update estimate:  $\hat{\mathbf{a}}(n) = \tilde{\mathbf{a}}(n, N)$ .

In the recursion above,  $0 < \mu < 1$  is a step size that controls the stability and convergence speed. A small value for the step size  $\mu$  is recommended due to the slow variation of the PA parameters and the amplitude distribution of the OFDM signal which only rarely results in levels in the saturation region.

## 5. SIMULATIONS

An OFDM system with 16-QAM on  $N = 512$  sub-carriers is considered. The length of the cyclic prefix  $v$  is equal to 16, and  $T = 32$  pilot carriers are used for channel estimation. The channel is Rayleigh fading with four independent propagation paths each generated according to a Jakes' Doppler spectrum. The subcarrier frequency is 5 GHz, and a bandwidth of 20 MHz is used. The mobile speed is set to 2 km/h.

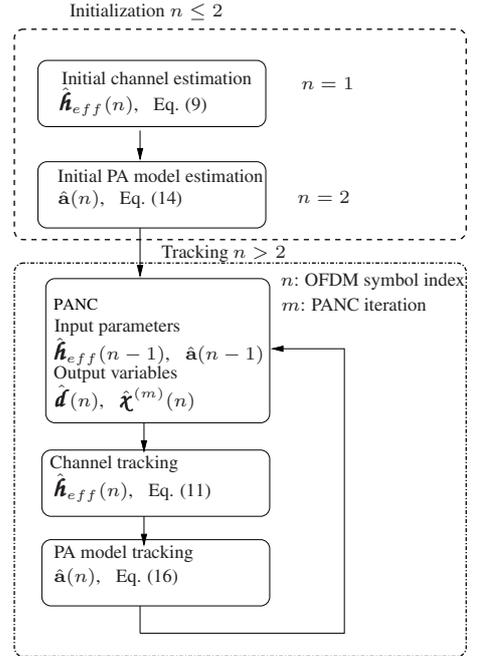


Fig. 3. PANC with PA model estimation

The PA memory in the Hammerstein model (see Figure 1) was modeled by  $C(z) = \frac{1}{1.5} \frac{1+0.25z^{-2}}{1+0.4z^{-1}}$ . As the static nonlinearity  $g[\cdot]$ , a third-order polynomial specified by vector  $\mathbf{a} = [0.024, 0.9634, 0.0847, 0.1170]^T$  was used to model a Solid State Power Amplifier (SSPA) with smoothness factor  $p = 2$  and clipping level  $\nu = 4$  dB [7].

To assess the PA model tracking capability of the proposed detection technique, a random-walk model [9] was adopted for the variation of the third-order coefficient in the PA polynomial, i.e.,

$$a_3(n) = a_3(n - 1) + q(n) \quad (17)$$

where  $q(n)$  denotes a random perturbation with  $\sigma_q^2 = 0.001$ . We note that more accurate models for the time variation is beyond the scope of the paper. The variation in (17) will modify the output third-order intercept point, the 1dB-compression point, and saturation level. Other coefficients were assumed constant throughout the simulations.

Figure 4 shows the BER versus SNR when using PANC with PA model estimation at the receiver for the case when the third-order coefficient of  $g[\cdot]$  is varying according to (17). We considered the case when the estimated model parameters are kept fixed, and the case when they are recursively updated using the algorithm in Section 4.2.2. Curves for linear PA with channel estimation (CE) and channel state information (CSI), and nonlinear PA without PANC are included for reference.

Finally, we evaluate the performance for the case of model mismatch. For this purpose we employ a Wiener-type PA model, i.e., a linear filter followed by a static nonlinearity [1]. Figure 5 shows the BER curves when employing a Wiener

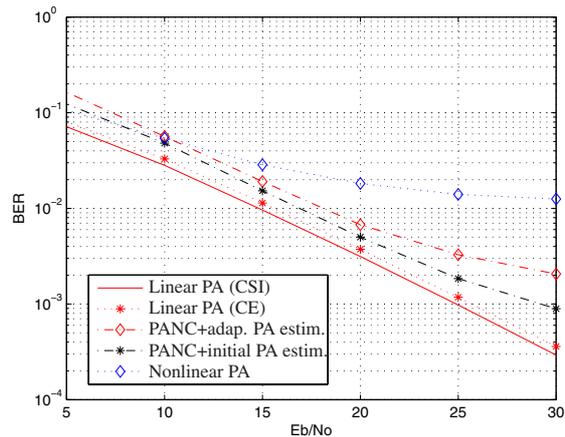
model of the HMC409LP4 PA which is suitable for WLAN and WiMAX implementations. The taps of the linear part are specified by  $h = [1 \ -0.011 \ -0.101 \ -0.022 \ -0.035 \ -0.038 \ 0.002]^T$ . The time-variation in  $g[\cdot]$  was the same as for the Hammerstein model. We see that a significant reduction of the nonlinear distortion is still obtained. However, a BER floor appears at high SNRs because the nonlinear distortion estimation disregards the PA memory. Techniques to overcome this problem are under investigation.

## 6. CONCLUSIONS

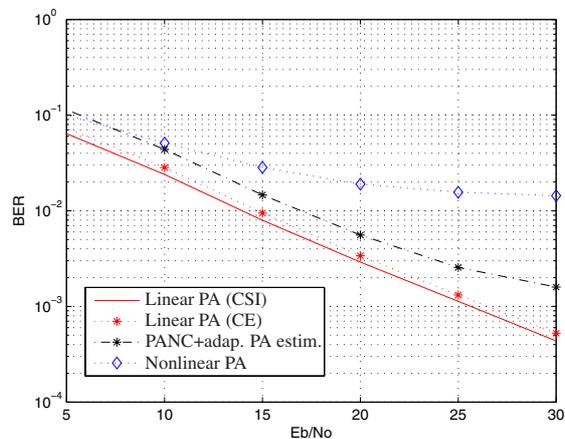
In this paper, we addressed the problem of identifying and tracking the model parameters of a broadband power amplifier (PA) at the receiver. The PA parameters are then used in the receiver to cancel the nonlinear PA distortion caused by high peak-to-average power ratio OFDM signals. The approach greatly simplifies implementation complexity as the need for a computationally complex PA identification step at the transmitter is avoided. This is an attractive scheme for uplink processing where the computational burden is concentrated at the base station. One important feature of the approach is that the PA memory can be estimated jointly with the wireless channel. Simulations illustrated the good performance of the nonlinear distortion cancellation for the case when the PA model estimation was carried out in the receiver. It was also shown that time-varying PA parameters could be tracked by employing a simple stochastic gradient algorithm.

## 7. REFERENCES

- [1] D. Morgan, Z. Ma, J. Kim, M. Zierdt, and J. Pastalan, "A generalized memory polynomial model for digital predistortion of RF power amplifiers," *IEEE Trans. Signal Proc.*, vol. 54, pp. 3852–3860, October 2006.
- [2] S. Boumaiza and F. M. Ghannouchi, "Thermal memory effects modeling and compensation in rf power amplifiers and predistortion linearizers," *IEEE Microw. Theory Tech.*, vol. 51, pp. 2427–2433, Dec. 2003.
- [3] F. Gregorio, T. Laakso, and J. Cousseau, "Receiver cancellation of nonlinear power amplifier distortion in SDMA-OFDM systems," in *IEEE, ICASSP 2006, France*, May 2006.
- [4] J. Tellado, L. Hoo, and J. Cioffi, "Maximum-likelihood detection of nonlinearly distorted multicarrier symbols by iterative decoding," *IEEE Trans. Commun.*, vol. 51, pp. 218–228, Feb. 2003.
- [5] D. Dardari, V. Tralli, and A. Vaccari, "A theoretical characterization of nonlinear distortion effects in OFDM systems," *IEEE Trans. Commun.*, vol. 48, pp. 1755–1764, Oct. 2000.
- [6] L. Deneire, P. Vandenameele, L. van der Perre, B. Gyselinckx, and M. Engels, "A low-complexity ML channel estimator for OFDM," *IEEE Trans. Commun.*, vol. 51, pp. 135–140, Feb. 2003.
- [7] F. Gregorio, S. Werner, J. Cousseau, and T. Laakso, "Channel estimation for multiuser OFDM systems in the presence



**Fig. 4.** BER versus  $E_b/N_o$  for PANC with adaptive PA estimation and using initial PA model estimate (no tracking), FD channel estimator ( $T = 32$  pilot subcarriers) and time-varying Hammerstein-type PA model. Linear PA with channel estimation (CE) and channel state information (CSI), and nonlinear PA without PANC are included for reference.



**Fig. 5.** BER versus  $E_b/N_o$  for PANC with adaptive PA estimation, FD channel estimator ( $T = 32$  pilot subcarriers) and time-varying Wiener-type PA model. Linear PA with channel estimation (CE) and channel state information (CSI), and nonlinear PA without PANC are included for reference.

of power amplifier nonlinearities," *IEEE PIMRC 2006*, Sept. 2006.

- [8] W. G. Jeon, K. H. Chang, and Y. S. Cho, "An adaptive data predistorter for compensation of nonlinear distortion in OFDM systems," *IEEE Trans. Comm.*, vol. 45, pp. 1167–1172, Oct. 1997.
- [9] A. Papoulis and S. Pillai, *Probability, Random Variables, and Stochastic Processes*, McGraw-Hill, New York, 4th Ed., 2002.