Copyright©2007 IEEE. Reprinted from:

F. Gregorio, S. Werner, J. Cousseau and R. Wichman. Split predistortion approach for reduced complexity terminal in OFDM systems, in *Proceedings of the Vehicular Technology Conference, VTC 2007,* pp. 2697-2701, April 2007, Dublin, Ireland.

This material is posted here with permission of the IEEE. Such permission of the IEEE does not in any way imply IEEE endorsement of any of Helsinki University of Technology's products or services. Internal or personal use of this material is permitted. However, permission to reprint/republish this material for advertising or promotional purposes or for creating new collective works for resale or redistribution must be obtained from the IEEE by writing to pubs-permissions@ieee.org.

By choosing to view this document, you agree to all provisions of the copyright laws protecting it.

Split predistortion approach for reduced complexity terminal in OFDM systems

Fernando Gregorio, Stefan Werner and Risto Wichman
Helsinki University of Technology, Signal Processing Laboratory
P.O. Box 3000, FIN-02015 TKK, Finland.
Email: fernando.gregorio@tkk.fi

Juan E. Cousseau
Universidad Nacional del Sur
Dept. of Electrical and Computer Eng.,
Av. Alem 1253, 8000, Baha Blanca, Argentina

***Abstract*— This paper proposes a novel split predistorter structure to remove nonlinear distortion caused by nonlinear power amplifier with memory. Unlike conventional techniques, the new technique does not require an estimation of the memory model at the transmitter leading to a reduced implementation complexity. Simulations have been carried out using the SUI3 channel model developed for IEEE802.16 standard. The results verify the good performance of the proposed predistorter for several relevant power amplifier models.**

I. INTRODUCTION

Multicarrier techniques like OFDM are employed in a vast number of communication systems, e.g., WLAN and WiMAX, due to its multipath mitigation capability and high spectral efficiency. One drawback of OFDM signals is its high Peak-to-Average Power Ratio (PAPR), which requires linear power amplifiers for power-efficient transmission.

Actual power amplifiers (PA) have nonlinear transfer functions that cause in-band (signal waveform) distortion that degrades bit error rate (BER) performance. The nonlinear characteristics of PA also cause adjacent channel interference (out-band distortion). Operating the PA at high back-off levels reduces nonlinear effects but results in a solution with low power efficiency. To this end, predistortion and PAPR reduction can be used to combat this nonlinear distortion and to increase power efficiency.

Predistortion techniques show good results with reduced complexity if PA is memoryless. However, if PA memory effects are accounted for in the predistorter implementation, the complexity increases and the performance may deteriorate. The reason is that the PA memory may need to be continuously tracked (or re-estimated) because it depends on the instantaneous power level [1].

To overcome the high complexity problems related to the conventional predistortion approach, we propose a *split predistorter* implementation. The idea is to use a low complexity memoryless predistorter in the transmitter for nonlinear distortion compensation and a postcompensation filter at the receiver that jointly estimates PA memory and channel effects. The proposed method does not require the knowledge of the PA memory model as is required in the conventional predistorter solution. The price to pay for joint estimation of PA memory and channel effects will be an increase in the minimum number of pilot tones required to obtain a good estimate. This is because the effective length of the channel is increased. On the other hand, employing an OFDM channel estimation approach that exploits dedicated pilot carriers, time-varying channel and PA memory effect can be tracked jointly in digital domain. To put the proposed technique in context, we compare it with the conventional predistorter solution, where the estimation of the input-output measurements required to track the memory effects includes a feedback loop from analog domain (PA output) to digital domain.

The performance of the proposed predistorter solution is evaluated for several relevant PA models. Most PA models found in the literature are only applicable to satellite communication channels where the power amplifier at the satellite transponder is driven near saturation to exploit the maximum power efficiency, see, e.g., [2], [3],[4]. Thus, their applications to systems operating at low power levels, like WLAN and WiMAX, are not easily justifiable. In this paper, we employ a simple model of an actual PA with memory that can be applicable in WLAN and WiMAX systems. The model is based on data sheets provided by a manufacturer. Details on how to build the model are outlined in Section III. More complex models taken from the literature are also considered in our simulations.

The paper is organized as follows. The system model is introduced in Section II. Section III presents modeling of a real power amplifier. The split predistortion approach is developed in Section IV. In Section V, simulations for various PA models models are presented, followed by conclusions in Section VI.

II. SYSTEM MODEL

The transmission model is illustrated in Figure 1 (a). The OFDM system under consideration has N sub-carriers; the signal $\mathbf{x}_{cp}(n) \in \mathbb{C}^{(N+v)\times 1}$ at symbol time n, before the PA, is given by

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

where $\mathbf{G}_{cp}$ is the $(N+v)\times N$ *cyclic prefix insertion matrix* [5], 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 with nonlinear transfer function. We assume here that the PA is modeled with a Wiener structure [6], see Figure 1 (b). As depicted in

Figure 1 (a), the Wiener model is formed by a linear filter followed by a nonlinear static block. The Wiener model is frequently used to model nonlinear PAs with memory. Other models, like Hammerstein and Wiener-Hammerstein can also be considered. As discussed in Section VI, the performance of the proposed split predistorter structure has good performance even in the case of model mismatch.

The multicarrier signal $\mathbf{x}_L(n)$, at the output of the linear block $\mathbf{h}_a$ of the PA, assuming that the length of the memory filter can be compensated by the cyclic prefix, is given by

$$\mathbf{x}_L(n) = \mathbf{H}_a \mathbf{x}_{cp}(n) \tag{2}$$

where $\mathbf{H}_a$ is a circular convolution matrix built from vector $\mathbf{h}_a = [h_{a1}, \cdots, h_{aL}]$.

Finally, after the nonlinear static block, under the assumption of low clipping levels [7], the transmitted signal becomes

$$\mathbf{x}_g(n) = g[\mathbf{x}_L(n)] = K_L\, \mathbf{x}_L(n) + \mathbf{d}(n) \tag{3}$$

where the first term $K_L\, \mathbf{x}_L(n)$ is the distortion-free discrete input signal vector (time domain) of Eq. (1) and K_L is the power amplifier gain. The second term $\mathbf{d}(n)$ is the nonlinear distortion which is a function of the modulated symbol vector $\boldsymbol{\mathcal{X}}(n)$ and the power amplifier transfer function $g[\cdot]$.

The transmitted signal in frequency domain can be written as

$$\mathcal{X}_g(n,k) = K_L \mathcal{H}_a(n,k) \mathcal{X}(n,k) + \mathcal{D}(n,k) \tag{4}$$

where $\mathcal{X}(n,k)$ is the transmitted symbol at subcarrier k, $\mathcal{D}(n,k) = \mathbf{Q}_N \mathbf{d}(n)$ is the value of the nonlinear distortion on sub-carrier k, and $\mathcal{H}_a(n,k)$ is the kth component of the $N \times 1$ frequency response vector of the PA memory, i.e.,

$$\boldsymbol{\mathcal{H}}_a = \boldsymbol{Q}_N \bar{\mathbf{h}}_a; \tag{5}$$

where $\bar{\mathbf{h}}_a = [\mathbf{h}_a, \mathbf{0}_{N-L}]$ is a $N \times 1$ vector.

Let us assume that the cyclic prefix is chosen larger than the effective channel length $L_{eff} = L_a + L_c$, where L_a is the length of the FIR filter $\boldsymbol{h}_a$ which model the time-variant part and L_c is the channel length. Then, the received signal at subcarrier k, at time n, can be expressed as

$$\begin{aligned} \mathcal{Y}(n,k) &= \mathcal{H}_{eff}(n,k) \mathcal{X}_g(n,k) + \mathcal{N}(n,k) \\ &= K_L \mathcal{H}_{eff}(n,k) \mathcal{X}(n,k) + \mathcal{H}_c(n,k) \mathcal{D}(n,k) \\ &\quad + \mathcal{N}(n,k) \end{aligned} \tag{6}$$

where $\mathcal{N}(n,k)$ is the additive noise, assumed here to be circular complex Gaussian with variance σ_n^2, and $\mathcal{H}_{eff}(n,k) = \mathcal{H}_a(n,k) \mathcal{H}_c(n,k)$ is the effective channel, where $\mathcal{H}_c(n,k)$ is the channel response at sub-carrier k.

III. Power amplifier modeling

It is important to employ a realistic power amplifier model when evaluating the system performance. In this section we describe a simple method to extract the parameters related to the Wiener model. The method is based on standard curves (i.e., S-parameters versus frequency, and gain compression (AM/AM) curves) that can be obtained from PA manufacturers. In particular, the method is used to extract the Wiener model parameters of a HMC409LP4 power amplifier that is targeted for WLAN and WiMAX applications.[1] This PA model is used in the simulations provided in Section VI.

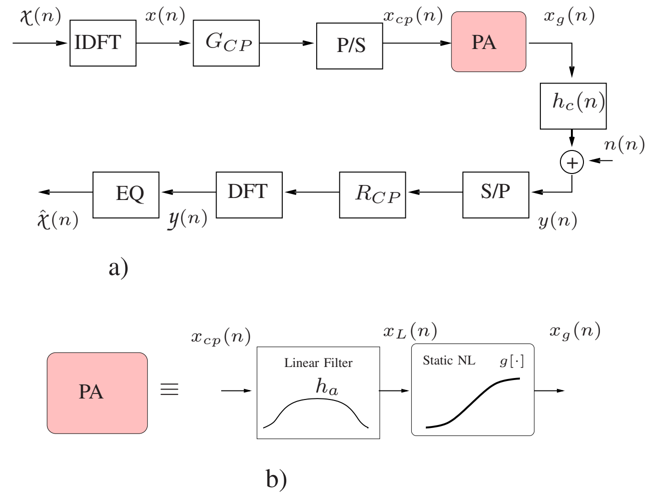


Fig. 1. a) System model b) Wiener model for the power amplifier

A. Static nonlinearity modeling

The gain compression curve defines the static nonlinear block of the Wiener model in Figure 1. Figure 2(c) shows the power compression (AM/AM) curve of the HMC409LP4 power amplifier operating at 3.5 GHz. In the following, we only consider the effect of AM/AM conversion. The AM/PM effect is assumed negligible.

The static nonlinearity can be modeled using a memoryless polynomial as

$$y(t) = \hat{g}[x(t)] = \sum_{j=0}^{P} a_j |x(t)|^j \tag{7}$$

where $x(t)$ is the input signal, $y(t)$ is the output signal of the power amplifier, and $\{a_j\}_{j=0}^{P}$ are the polynomial coefficients, P being the polynomial order. By exciting the input with a power-swept single-tone signal, the coefficients of the polynomial can be estimated using a least-squares (LS) approach, see, e.g., [9]. Since memory effects are mostly evident with broadband signals, they will not be important using a power-swept single tone signal. In this form, the static block will be suitably modeled.

For illustration purposes, Figure 2(c) depicts the compression curve of the HMC409LP4 PA and the estimated curve using a 5th-order polynomial.

B. Small-signal frequency response modeling

The linear dynamic part of the Wiener model, h_a, can be obtained from the small-signal frequency response of the power amplifier. Let us assume that an Lth-order FIR filter is used to model the frequency response. The filter coefficients

[1] Data sheets were obtained from Hittite Microwave Corporation [8]

can be obtained by minimizing the following weighted LS function [9]

$$E = \sum_{k=0}^{K-1} W(w_k) \left| H_d(e^{jw_k}) - \sum_{l=0}^{L} h_a(l) e^{jw_k(l+D)} \right|^2 \quad (8)$$

where $H_d(e^{jw_k})$ is the complex-valued power amplifier frequency response (S_{21} parameter), w_k is the angular frequency, D is a delay placed before the filter, and $W(w_k)$ is a real-valued nonnegative weighting function. The linear block modeled using an FIR filter is designed using a weighted LS approximation as proposed in [9].

Figures 2(a) and 2(b) illustrate the magnitude and phase frequency response approximation of the HMC409LP4 PA. The order of the FIR filter, L, was chosen equal 8. The weighting function was optimized to provide an accurate approximation in the frequency band of interest, $3.2-4$ GHz. We can see that modeling and measurements results agree well.

IV. Split Predistortion structure

This section proposes a split predistortion structure to reduce the complexity at the transmitter when compared with a conventional predistorter. The idea is to distribute the predistorter tasks such that the transmitter compensates for the static nonlinearity, and the receiver equalizes the PA memory. The advantages of such a solution are numerous. Firstly, the out-of-band distortion is reduced as with the conventional predistorter. Furthermore, the identification of the PA memory can now be easily performed at the receiver jointly with the wireless channel. This feature is particularly attractive because the computationally involved identification and tracking at the transmitter is avoided. The price to pay is an increase in the effective length of the combined PA and wireless channel.

In the following we briefly review the memoryless predistorter of [10] that we will employ in the transmitter. Note that it is possible to use other memoryless predistorters with the proposed structure. Thereafter, we present the details on how to estimate the PA memory and the wireless channel jointly.

A. Memoryless predistorter design [10]

The predistorter function $P(x)$ is designed to linearize the static nonlinearity. The function $P(x)$ can be expressed as a Q-order polynomial

$$\hat{P}[x(t)] = \sum_{i=0}^{Q} b_i |x(t)|^i \quad (9)$$

With a polynomial model of the static nonlinearity (see previous section), the output of the predistorter-amplifier combination can be written as

$$y(t) = \sum_{j=0}^{P} a_j \left(\sum_{i=0}^{Q} b_i |x(t)|^i \right)^j \quad (10)$$

The inverse of the power amplifier function can be fitted by another polynomial from training data

$$x_t(t) = g^{-1}(y_t(t)) = \sum_{j=0}^{P} \beta_j y_t^j(t) \quad (11)$$

where $x_t(t)$ is an input training sequence and $y_t(t)$ is the system output.

The coefficients β_i can be obtained using an LS estimate as

$$\hat{\boldsymbol{\beta}} = \left[\bar{\mathbf{Y}}_t^{\mathrm{H}} \bar{\mathbf{Y}}_t \right]^{-1} \bar{\mathbf{Y}}_t^{\mathrm{H}} \mathbf{X}_t \quad (12)$$

where $\bar{\mathbf{Y}}_t$ is the *Vandermonde* matrix formed by the output signal vector, and $\hat{\boldsymbol{\beta}} = [\beta_0, \beta_1, \ldots, \beta_Q]$ are the estimated coefficients that describe the inverse nonlinearity polynomial $g^{-1}[\cdot]$.

Finally, the predistortion function $P(x)$ is given by

$$\hat{P}[x(t)] \quad = \quad g^{-1}(y_t(t)) = \sum_{j=0}^{Q} \beta_j a_j x(t)^j = \sum_{j=0}^{Q} b_j x(t)^j$$

with $b_j = a_j \beta_j$.

B. Postcompensation of memory effects

Taking advantage of the OFDM system characteristics, we now propose a joint estimation of the linear part $\mathcal{h}_a$ and transmit channel $\mathcal{h}_c$ in frequency domain. The solution eliminates the model estimation process at the transmitter. In addition, time variations due to PA memory and wireless channel can be tracked because the estimation of the effective channel is carried out for each OFDM symbol.

Let us assume a comb-type pilot arrangement where a set of T carriers are dedicated to channel estimation [11]. Furthermore, let $\mathcal{T}\,(k_1, \cdots, k_T)$ denote the set specifying the T pilot carriers. The channel frequency response on these sub-carriers ($k \in \mathcal{T}$) may be estimated as

$$\begin{aligned} \hat{\mathcal{h}}_{eff}(n,k) &= \mathcal{y}(n,k)/\chi(n,k) \quad (13) \\ &= \mathcal{h}_{eff}(n,k) K_L + \mathcal{h}_c(n,k) \frac{\mathcal{d}'(n,k)}{\chi(n,k)} + \frac{\mathcal{n}(n,k)}{\chi(n,k)} \end{aligned}$$

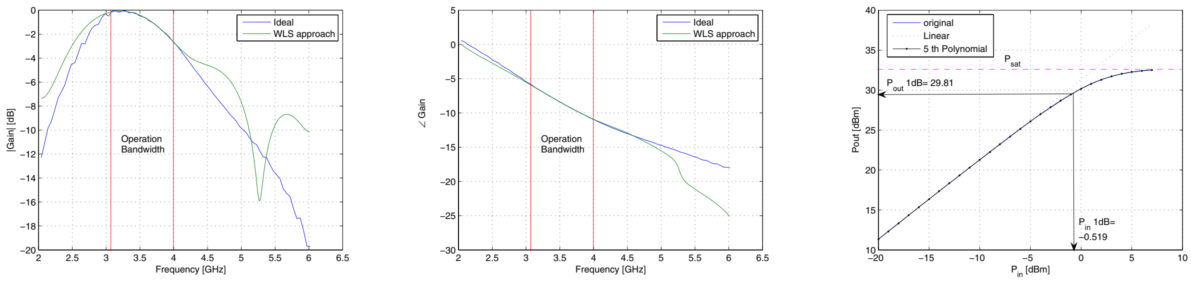
where $\mathcal{y}(n,k)$ is the received signal at sub-carrier $k \in \mathcal{T}_L$, $\mathcal{d}'(n,k)$ is the residual nonlinear distortion that cannot be removed with the predistorter, and $\chi(n,k)$ is the training symbol transmitted at the corresponding sub-carrier.

Collecting the channel estimates obtained with pilot carriers into vector $\hat{\boldsymbol{\mathcal{h}}}_p(n) = [\hat{\mathcal{h}}_{eff}(n,k_1), \cdots, \hat{\mathcal{h}}_{eff}(n,k_T)]^T$, the whole channel frequency domain response can be obtained through interpolation using truncated DFT matrices [12], as given in the following expression

$$\hat{\boldsymbol{\mathcal{h}}}_{eff}(n) = \mathbf{Q}_N [\mathbf{Q}_{\mathcal{T}}^{\mathrm{H}} \mathbf{Q}_{\mathcal{T}}]^{-1} \mathbf{Q}_{\mathcal{T}}^{\mathrm{H}} \hat{\boldsymbol{\mathcal{h}}}_p(n) \quad (14)$$

where $\mathbf{Q}_N$ is the $N \times N$ IDFT matrix, $\mathbf{Q}_{\mathcal{T}}$ is formed by the T columns of $\mathbf{Q}_N$ associated with the initially estimated sub-carriers and the $L_a + L_h$ rows associated with the non zero time-domain channel taps.

By design, the proposed technique obtains the best performance when the PA can be accurately modeled using a Hammerstein model (the reverse of the Wiener model). However,



(a) (b) (c)

Fig. 2. HMC409LP4 PA modelling. a) Magnitude response using 8th-order FIR approximation, b) Phase response using 8th-order FIR approximation, c) Power compression (AM/AM) curve using a 5th-order least squares polynomial fit.

considering small levels of residual nonlinear distortion, the technique shows excellent performance for Wiener structures as verified in the next section.

V. Simulations

In this section, the performance of the novel split predistorter approach is evaluated in a realistic scenario. The results are compared with a system using conventional predistortion, where the linear model of the PA memory is assumed to be known.

An OFDM system with 16-QAM on $N = 512$ sub-carriers is considered. Frequency-domain channel estimation at the receiver is performed, see Eq. (14). The cyclic prefix length L is equal to 16, and $T = 32$ pilot carriers are used. This results in 6.25% of overhead for channel estimation.

The frequency-selective fading channel is generated according to the Stanford University Interim model, SUI 3, which is used in the evaluation of IEEE 802.16 WiMAX (2004) standard [13].

The power amplifier operates with a back-off level of $IBO = -4$ dB.

The performance of the system is evaluated with the following power amplifier models where a 5-th order polynomial is used for the static nonlinearity:

1) Model 1: WLAN low-power amplifier (see Fig 2). The HMC409LP4 power amplifier, suitable for WLAN and WiMAX implementations modeled with a Wiener filter is considered. An FIR filter of 7 coefficients is used to model the linear part.
$h = [1 - 0.011, -0.101, -0.022, -0.035, -0.038, 0.002]$
2) Model 2: Wiener system. The model is extracted from a Class AB power amplifier [3]. The linear part is modeled using an IIR filter with transfer function given by
$$h(z) = \frac{1 + 0.3z^{-2}}{1 - 0.2z^{-1}} \qquad (15)$$
3) Model 3: Wiener-Hammerstein Model. In this case the memory effects to be compensated obey a Wiener-Hammerstein model. The block transfer functions are given by [3]
$$h(z) = \frac{1}{1.5}\frac{1 + 0.25z^{-2}}{1 + 0.4z^{-1}} \qquad (16)$$
$$s(z) = \frac{1}{0.52}\frac{1 - 0.1z^{-2}}{1 - 0.2z^{-1}} \qquad (17)$$

Performance is evaluated for the following cases: the proposed split technique, a perfect linear amplifier (to have a BER lower bound), a nonlinear amplifier with memoryless predistorter without compensation (in this case the number of pilot tones is $T = 8$ that can not estimate the effective channel length with good accuracy), and a nonlinear amplifier with PD and precompensation in the transmitter as described in [1]. For this last precompensation method, the FIR coefficients modeling memory effects are assumed known without errors.

BER results for the amplifier models 1,2,3 are presented in Figures 3, 4 and 5, respectively. It is possible to conclude that the split predistorter structure allows to obtain similar BER performance than the conventional PD. Simulation results shows also that our technique is robust to different power amplifier models presenting a good performance for the 3 different models evaluated. Even when the performance in terms of BER is similar, the advantage of using the split structure is the reduced complexity of a memoryless PD required when compared with the implementation of PD with memory.

VI. Conclusions

A new split predistorter approach for OFDM systems was presented. Unlike conventional predistortion techniques, the new method does not require an estimation of the memory model at the transmitter. As a consequence, a reduced implementation complexity can be expected. The new method estimates the effective channel response given by the convolution of the linear part of the power amplifier model and the channel. Simulations showed that the performance of the novel split predistorter technique reaches similar performance to conventional predistorters. The new technique can be implemented in systems where reduced implementation complexity is required in the transmitter. Thus, the split predistorter is a good option for the uplink transmission, where memory compensation is performed in the access point simplifying the hardware complexity of the subscriber units.

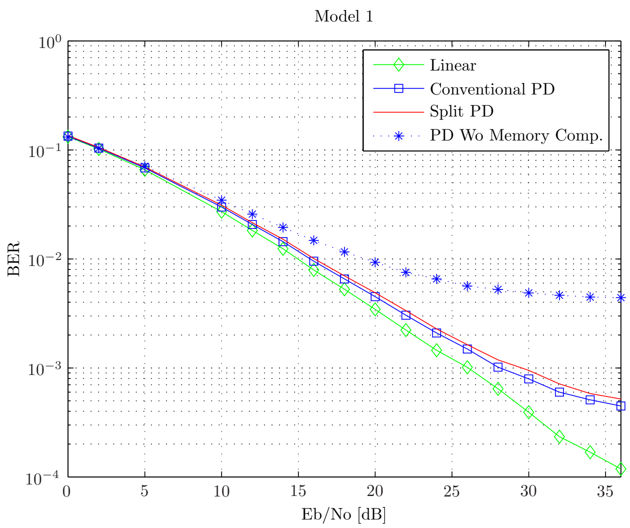


Fig. 3. BER versus E_b/N_o of a OFDM system with 16-QAM modulation, for Split predistortion (Split PD) and conventional PD ($T = 32$ pilot subcarriers) with nonlinear PA model 1 (HMC409LP4). Results obtained with linear PA and memoryless PD without memory compensation are included for reference.

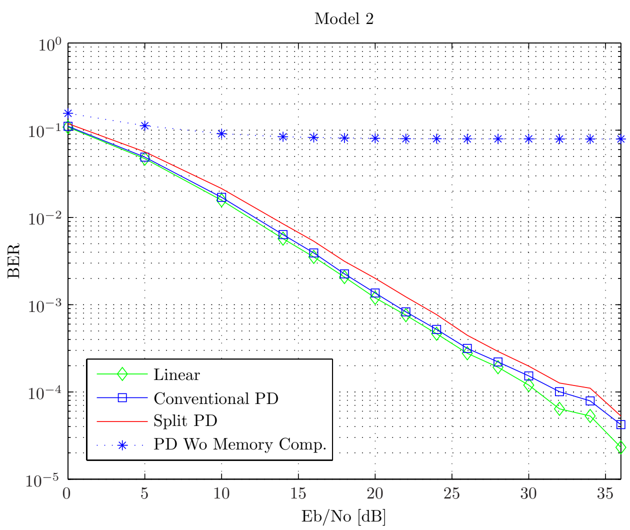


Fig. 4. BER versus E_b/N_o of a OFDM system with 16-QAM modulation, for Split predistortion (Split PD) and conventional PD ($T = 32$ pilot subcarriers) with nonlinear PA model 2 (Class AB). Results obtained with linear PA and memoryless PD without memory compensation are included for reference.

ACKNOWLEDGMENT

Fernando Gregorio is supported by the Programme ALβAN, European Union Programme of High Level Scholarships for Latin America. Identification Number: **E03D19254AR** and by Nokia Foundation. Fernando Gregorio, Stefan Werner and Risto Wichman are with the Smart and Novel Radios Research Group (SMARAD) Center of Excellence. This work was partially supported by Universidad Nacional del Sur, Argentina, Project # 24/K023.

REFERENCES

[1] A. Chaker, M. Atiaudo, I. Fijalkow, and J. Gautier, "Pre-compensation of the frequency-dependence of a non-linear amplifier in a multi-carrier transmission," in *IEEE International Conference on Communications, ICC 2004*, June 2004, vol. 4, pp. 2464–2467.

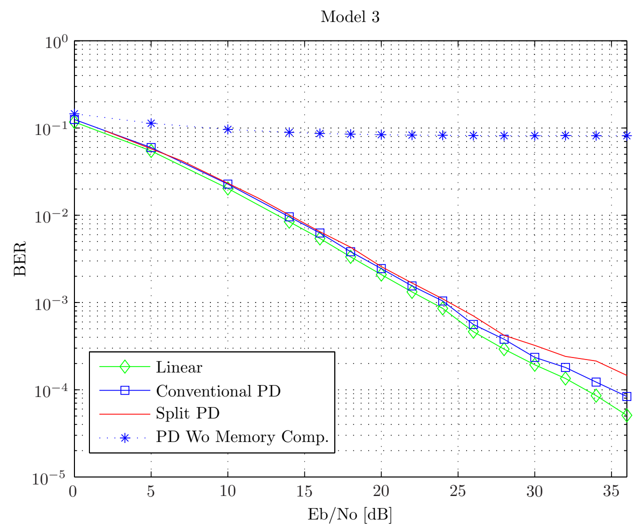


Fig. 5. BER versus E_b/N_o of a OFDM system with 16-QAM modulation, for Split predistortion (Split PD) and conventional PD ($T = 32$ pilot subcarriers) with nonlinear PA model 3 (Wiener-Hammerstein structure). Results obtained with linear PA and memoryless PD without memory compensation are included for reference.

[2] A. A. M. Saleh, "Frequency-independent and frequency-dependent nonlinear models of TWT amplifiers," *IEEE Trans. Commun.*, vol. 29, pp. 1715–172096, Nov. 1981.

[3] R. Raich, H. Qian, and G. T. Zhou, "Orthogonal polynomials for power amplifier modeling and predistorter design," *IEEE Trans. Veh. Technol.*, vol. 53, pp. 1468–1479, Sept. 2004.

[4] C. Eun and E. J. Powers, "A predistorter design for a memory-less nonlinearity preceded by a dynamic linear system," in *IEEE GLOBECOM 95*, Sept. 1995, vol. 1, pp. 152 – 156.

[5] T. Roman, M. Enescu, and V. Koivunen, "Joint time-domain tracking of channel and frequency offsets for MIMO OFDM systems," *Kluwer, Wireless Personal Communications*, vol. 31, pp. 181–200, 2004.

[6] 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.

[7] 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.

[8] Hittite Microwave Corporation, *http://www.hittite.com*.

[9] P. Jantunen, G. Gamez, and T. Laakso, "Measurements and modelling of nonlinear power amplifiers," in *Signal Processing Symposium, 2004. NORSIG 2004*, June 2004, vol. 1, pp. 328–331.

[10] Y. Guo and J. Cavallaro, "Enhanced power efficiency of mobile OFDM radio using predistortion and post-compensation," in *IEEE, Vehicular Technology Conference, VTC 2002-Fall.*, Sept. 2002, vol. 1, pp. 214 – 218.

[11] S. Werner, M. Enescu, and V. Koivunen, "Combined frequency and time domain channel estimation in mobile MIMO-OFDM systems," in *IEEE, International Conference on Acoustics, Speech, and Signal Processing , ICASSP 2006*, May 2006.

[12] 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.

[13] V. Erceg et .al,, "Channel models for fixed wireless applications," *tech. rep., IEEE 802.16 Broadband Wireless Access Working Group*, January 2001.