OFDM RECEIVERS WITH ITERATIVE NONLINEAR DISTORTION CANCELLATION

Leonardo G. Baltar[1], *Stefan Dierks*[*1], *Fernando H. Gregorio*[2], *Juan E. Cousseau*[2], *Josef A. Nossek*[1]

[1]Technische Universität München
Institute for Circuit Theory and Signal Processing
Arcisstr. 21, 80290 München, Germany
{leo.baltar, josef.a.nossek}@tum.de,
stefan.dierks@mytum.de

[2]Universidad Nacional del Sur
Dept. of Electrical and Computer Eng.-
IIIE (CONICET)
Av. Alem 1253, 8000 Bahía Blanca, Argentina
{fernando.gregorio, jcousseau}@uns.ar

ABSTRACT

We propose two practical schemes for the cancellation of nonlinear distortions caused by the power amplifiers (PA) in Orthogonal Frequency Division Multiplexing (OFDM) systems. The algorithms are iterative and located at the receiver. They utilize a soft channel decoder to improve the capabilities of the distortion estimation and relay all the computational complexity to the receiver. Simulation results show that the algorithm can improve the coded bit error rate (BER) to levels close to the linear system case.

1. INTRODUCTION

Cyclic prefix based OFDM system is the current choice as the multicarrier modulation scheme for the 4-th generation of mobile communications. Its advantages are well known, like for example its robustness to multipath propagation and its efficient implementation. However, one of the main drawbacks of OFDM is that the transmitted signal has a near Gaussian amplitude distribution and consequently presents a high Peak-to-average Power Ratio (PAPR).

Practical (low cost) PA used in wireless communications systems present a nonlinear input-output characteristic. This means that, if the input signals have a high PAPR and the operation at the linear region is required, the energy efficiency will drop dramatically. This is particularly critical at the mobile station (MS) during an uplink transmission, because the MS works on batteries and its PA should be operated in an efficient way.

There are many techniques to reduce the PAPR at the transmitter [1]. The problem of those techniques is that they either reduce the data throughput or reduce the useful energy and indirectly the SNR, provoking a higher BER at the receiver. They also increase the computational complexity at the transmitter, what is also undesired in an uplink scenario. Similarly, some predistortion techniques can be employed, having, however, similar drawbacks.

*This work was partially supported by the German Academic Exchange Service (DAAD).

If the PA is to be efficiently used, some nonlinear distortion can be accepted. There has to be a compromise between the efficient use of the PA and the level of distortion. This compromise is achieved by means of adjusting the input back-off (IBO) or equivalently the clipping ratio. The main drawbacks of clipping the transmitted signal are an increase in the out-of-band radiation and the introduction of an in-band distortion. Our focus here is the cancellation of the in-band distortion.

The authors of [2] consider iterative scheme that is not maximum likelihood (ML) and due to its highly nonlinear nature, it is difficult to analyse and compare. In [3] the authors propose an iterative algorithm that approximates the performance of the ML distortion estimation scheme. In this work we propose two algorithms that are based on that same scheme, which we call power amplifier nonlinearity cancellation (PANC) [4]. We first consider the introduction of a channel decoder inside the iteration loop to improve the distortion estimation, a variation we called coded PANC (CPANC) algorithm and that has similarities to the scheme presented in [5]. Then we considered the use of the soft information from the channel decoder to calculate a reliability coefficient that scales the estimated distortion before the subtraction from the received channel. We called this last algorithm scaled CPANC (sCPANC).

The outline of the work is the following. In Section 2 a brief description of the system model and notation is included. The novel efficient techniques for nonlinear distortion cancellation are introduced in Section 3. Simulation results are presented in Section 4. Finally, our conclusions are presented in Section 5.

2. SYSTEM MODEL

The OFDM system model we considered, including the PA model is depicted in Fig. 1. The data bit stream is first grouped in blocks of length $N_\mathrm{u}\log_2(M)R$, where N_u is the number of data filled subcarriers, M is size of the M-QAM constellation and R is the code rate. Those blocks of bits are then en-

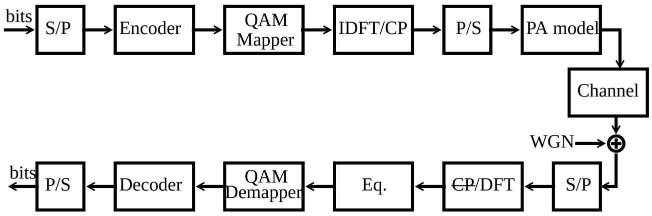

Fig. 1. System Model

coded, generating a vector of length $N_{\mathrm{u}} \log_2(M)$, in which each group of $\log_2(M)$ bits is mapped into one M-QAM symbol. $\mathbf{x}[n] \in \mathbb{C}^N$ contains complex symbols in N_{u} of its elements and zeros elsewhere. Next the OFDM multicarrier modulation is performed by the transform $\mathbf{Q}^{-1}\mathbf{x}[n]$ and by the addition of the cyclic prefix (CP) generating the vector $\tilde{\mathbf{x}}[n] \in \mathbb{C}^{N+L_{\mathrm{CP}}}$, where $\mathbf{Q}$ is the DFT matrix with elements given by $[\mathbf{Q}]_{i,j} = \frac{1}{\sqrt{N}} \mathrm{e}^{\mathrm{j}\frac{2\pi(i-1)(j-1)}{N}}$, $\mathbf{Q}^{-1}$ is the IDFT matrix and L_{CP} is the length of the cyclic prefix.

In a real transmission system the vectors $\tilde{\mathbf{x}}[n]$ would be serialized, upsampled, transformed to the analog domain and filtered before being passed to the radio frequency (RF) front end. In the RF front-end the analog signal is then upconverted (possibly in steps) to the RF frequency, filtered again and amplified by the PA before entering the antenna.

In our system model we consider the PA as the only processing after the insertion of the CP. We assume a memoryless PA, because the memory of the PA and all previous analog filters can be considered inside the channel model, as far as the total length of the resulting channel is not longer than the CP into consideration. In our simulations we consider the model of a Solid State Power Amplifier (SSPA) with input-output characteristic given by [6]

$$\mathrm{F_a}(\tilde{\mathbf{x}}[n]) = \frac{|\tilde{\mathbf{x}}[n]|}{\left[1 + \left(\frac{|\tilde{\mathbf{x}}[n]|}{A}\right)^{2p}\right]^{\frac{1}{2p}}}, \tag{1}$$

$$\mathrm{F_p}(\tilde{\mathbf{x}}[n]) = 0, \tag{2}$$

where $F_a(\bullet)$ is the magnitude transfer and $F_p(\bullet)$ is the phase transfer. p describes the smoothness of the transition from the linear to the saturation region. A is the clipping threshold and the clipping ratio is defined as

$$\gamma = \frac{A}{\sqrt{\mathrm{E}[|\tilde{\mathbf{x}}[n]|^2]}}. \tag{3}$$

The output of the PA is then given by

$$\tilde{\mathbf{x}}_g[n] = \mathrm{F_a}(\tilde{\mathbf{x}}[n]) \exp(j[\arg(\tilde{\mathbf{x}}[n]) + \mathrm{F_p}(\tilde{\mathbf{x}}[n])]). \tag{4}$$

After the PA the transmitted signal goes through a memory channel that can be regarded as a finite impulse response (FIR) filter. If the length of the channel is shorter than the CP, the received signal can be represented in vector form and the convolution can be described by a circulant matrix as it is shown in Section 3.

2.1. Linear PA model

For low clipping ratios, the nonlinear distortion can be approximated by a Gaussian process. In this case the nonlinear PA can be represented by a linear model [7], where the input signal is weighted by a real coefficient and a Gaussian noise is added as shown in Fig. 2. The output of the PA is given by $\tilde{\mathbf{x}}_g[n] = K_L\tilde{\mathbf{x}}[n] + \tilde{\mathbf{d}}[n]$.

The distortion $\tilde{\mathbf{d}}[n]$ is actually dependent on the input signal $\tilde{\mathbf{x}}[n]$. The factor K_L is chosen as to minimize the average power of the distortion $\tilde{\mathbf{d}}[n] = g(\tilde{\mathbf{x}}[n]) - K_L\tilde{\mathbf{x}}[n]$. In this case the input signal and the distortion have correlation close to 0. K_L is can be calculated analytically for simple PA models [7], but in our simulation results we have calculated it numerically.

It is worth noting that the approximation of the nonlinear distortion by a Gaussian process cannot be applied for very high clipping ratios, where an impulsive noise model would be more adequate. In such a case the PA has to operate almost only in the linear region, causing an inefficient energy consumption and a very low level of in-band and out-of-band distortion. As a consequence an iterative receiver would not be necessary and the channel decoder should be able to compensate it. In [4] the authors showed that even for high clipping level (to meet the spectral requirements fixed in WLAN specifications), the nonlinear distortion can be modeled as a Gaussian uncorrelated noise.

3. RECEIVER

The received signal after the removal of the CP is given by

$$\tilde{\mathbf{y}}[n] = K_L\mathcal{H}\mathbf{Q}^{-1}\mathbf{x}[n] + \mathcal{H}\tilde{\mathbf{d}}[n] + \boldsymbol{\eta}, \tag{5}$$

where $\mathcal{H}$ is the channel circulant matrix. After the DFT the received signal is given by

$$\mathbf{y}[n] = K_L\mathbf{H}\mathbf{x}[n] + \mathbf{H}\mathbf{d}[n] + \mathbf{Q}\boldsymbol{\eta}, \tag{6}$$

where $\mathbf{H} \in \mathbb{C}^{N\times N}$ is a diagonal matrix with the DFT of the channel impulse response in its main diagonal and $\mathbf{d}[n]$ is the distortion in the frequency domain or $\mathbf{d}[n] = \mathbf{Q}\tilde{\mathbf{d}}[n]$. Provided that the receiver has perfect channel state information (CSI) and that the CP is long enough, then, since $\mathbf{H}$ is diagonal, the equalizer is just the inverse of $\mathbf{H}$.

The iterative receiver we propose comes immediately after the equalization. For that matter we introduce the subscript m to represent the iteration index. Moreover, we drop the block index n for simplicity reasons. The receiver is based on the PANC [3, 4] with the main difference that the channel decoder is included inside the iteration cycle. In this work we

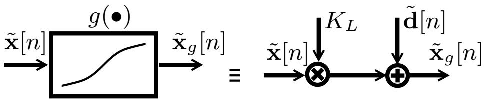

Fig. 2. Equivalent linear PA model

consider a QAM demapper that provides soft bits by means of log-likelihood ratios to a soft decoder [8]. Besides having better performance than hard decoder, the soft decoder also provides us means of checking the reliability of the decoded bits. That is why we also included a reliability factor in the case of sCPANC.

The idea of the PANC algorithm is to use an estimation of the transmitted symbols together with a memoryless model of the PA in order to calculate an estimation of the distortion. Then, this estimation is subtracted from the received signal and a new estimation of the distortion can be calculated. This is done iteratively until a reliable estimation of the distortion is obtained.

A summary of the sCPANC Algorithm after dropping the block index n is given as follows:

1. Estimate $\hat{\boldsymbol{x}}_m$
   (a) Subtract distortion: $K_L\mathbf{x} + \mathbf{d} - \hat{\mathbf{d}}_m + \mathbf{H}^{-1}\boldsymbol{\eta}$
   (b) Scale signal: $\mathbf{x} + \frac{\mathbf{d}-\hat{\mathbf{d}}_m+\mathbf{H}^{-1}\boldsymbol{\eta}}{K_L}$
   (c) Calculate approximate LLRs for each k-th subcarrier of encoded bits:
   $\hat{\lambda}_{i,m,k} = -\frac{|h_k|^2}{\sigma_n^2}\left(\min_{s\in\mathbb{A}_0}\|x_k - s\| - \min_{x\in\mathbb{A}_1}\|x_k - s\|\right)$,
   where $\mathbb{A}_0$ and $\mathbb{A}_1$ are the sets of constellation points in which i-th bit is 0 and 1, respectively.
   (d) Decode input LLRs and obtain decoded and encoded bit sequences: $\hat{\mathbf{b}}_m$, $\tilde{\mathbf{b}}_m$
   (e) Modulate encoded bit sequence: $\hat{\mathbf{x}}_m$
2. Transform estimated symbols: $\mathbf{Q}^{-1}\hat{\mathbf{x}}_m$
3. Estimate distortion: $\mathrm{g}(\mathbf{Q}^{-1}\hat{\mathbf{x}}_m) - K_L\mathbf{Q}^{-1}\hat{\mathbf{x}}_m$
4. Transform distortion:
   $\hat{\mathbf{d}}_m = \mathbf{Q}\mathrm{g}(\mathbf{Q}^{-1}\hat{\mathbf{x}}_m) - K_L\hat{\mathbf{x}}_m$
5. Calculate Reliability Coefficient α based on the LLRs of the decoded and corrected bits.
6. Scale distortion with the Reliability Coefficient: $\alpha\hat{\mathbf{d}}_m$
7. Go to step 1.

For the initialization ($m = 0$), the distortion is set to zero.

A block diagram representation of the algorithm is shown if Fig. 3 The block called Scaling and the arrow coming from and to it are drawn in dashed lines because this block is only included in the sCPANC and not in the CPANC algorithm.

3.1. Calculation of the reliability coefficient

One scaling coefficient has to be calculated for each received OFDM block. If the bits transmitted are considered independent, the most intuitive way of calculating the scaling would be the product of the probabilities of correct bit decoding. Because the number of bits in one OFDM block is very high this product will tend to zero even for medium to high SNR. As an alternative we considered the minimum probability of correct decoding for all bits in one block as given by

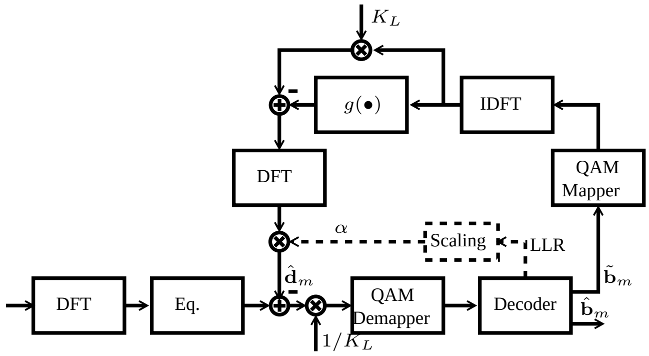

Fig. 3. Coded PANC (CPANC) and scaled Coded PANC (sCPANC). The block called Scaling and the arrow coming from and to it are drawn in dashed lines because this block is only included in the sCPANC.

$$\alpha = \min_{0\le i\le B} \mathrm{P}(b_i = \hat{b}_i|\boldsymbol{y}) = \min_{0\le i\le B} \frac{\exp|\mathrm{LLR}(b_i)|}{1+\exp|\mathrm{LLR}(b_i)|}. \quad (7)$$

where B is the number of uncoded bits in one OFDM block.

4. SIMULATIONS

In our simulations, we have considered an oversampling of the transmitted signal before the PA model. The oversampling factor was 4 and a root raised cosine (RRC) with roll-off factor $\rho = 0.5$ was used for the interpolation. After the PA model the signal was downsampled again with the same filter in order to reduce the simulation complexity involved in an oversampled channel modeling.

We have considered a single channel realization of the ITU channel model Vehicular B. The channel realization we have chosen has low frequency selectivity. This selection is made, because in the case of very frequency selective channel realizations, the loss in SNR in the subcarriers where the channel has a high attenuation will cause a BER loss that dominates over the effect of the nonlinear distortion. In this case, the iterative algorithm will not contribute much to improve the BER.

The parameters of the OFDM system were taken from the Mobile WiMAX OFDM standards and are shown in Table 1.

We have assumed that the PA model of (4) and its parameters are perfect known at the receiver.

As channel encoder we have employed a convolutional code with the parameters presented in Table 2.

In Fig. 4 the coded BER results after 2 iterations and for a clip ratio of $\gamma = 1.4$ are shown. For comparison purposes

Parameter	Value
OFDM symbol length, N	512
Number of data filled subcarriers, N_{u}	420
Number of null/guardband subcarriers	92
Cyclic prefix, L_{CP}	128
Channel bandwidth	5 MHz
OFDM symbol duration with CP	102.9 μs
Subcarrier frequency spacing	10.94 kHz
Modulation	16-QAM

Table 1. Parameters of the OFDM system

Parameter	Value
Code rate, R	$1/2$
Code polynomials	$1 + D^1 + D^2 + D^3 + D^6$
	$1 + D^2 + D^3 + D^5 + D^6$
Octal description	$[171; 133]$
Type of decoder	Max-log-MAP algorithm
Appended zero bits	6

Table 2. Parameters of the convolutional code

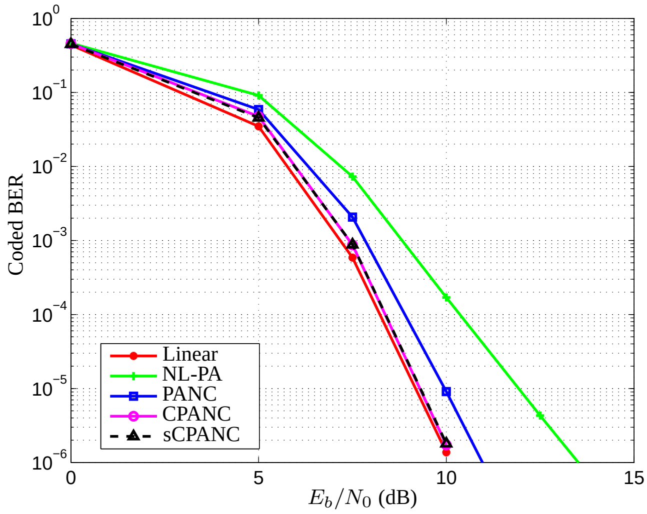

Fig. 4. BER after 2 iterations for $\gamma = 1.4$

we have plotted the BER for the linear and nonlinear cases, where no PAPR reduction, predistortion or nonlinear distortion cancellation schemes are included. We have also plotted the basic PANC with the channel decoder localized after the algorithm. It can be seen that the CPANC algorithm presents a significant performance improvement when compared to the simple PANC. But for the scenario into consideration the sCPANC shows almost no improvement when compared to the CPANC.

In Fig. 5 the coded BER results after 3 iterations and for a clip ratio of $\gamma = 1$ are shown. Despite of this critical clip level case the CPANC and sCPANC algorithms show a performance very close to the linear case and much better than the basic PANC.

5. CONCLUSIONS

In this work we have proposed some extensions of the PANC algorithm. Namely the CPANC, where the channel decoder is included in the iteration loop and the sCPANC, where besides the inclusion of the channel decoder, a reliability coefficient is used to weight the estimated distortion.

The simulation results showed that the algorithms provide a great improvement in comparison to the simple PANC. Of course that this comes at the cost of higher complexity, but, following our proposal, at the receiver side.

The proposed techniques, sCPANC and CPANC, reduce the in-band distortion allowing the operation of the PA in a high power efficient region. However, the out-of-band distortion is not mitigated employing receiver-side compensation techniques. Therefore, the clipping level (PA operation point) needs to be set in a region where the interference over neighboring users satisfies the spectral mask imposed by the wireless standards. We also have not considered here the problem of estimating the PA parameters [9] and PA models with memory, like for example the Wiener-Hammerstein. This gives some room for future extensions of our schemes.

Acknowledgments

We would like to thank Amine Mezghani for his contributions to this work and for the fruitful discussions.

6. REFERENCES

[1] S. H. Han and J. H. Lee, "An overview of peak-to-average power ratio reduction techniques for multicarrier transmission," *IEEE Wireless Commun. Mag.*, vol. 12, no. 2, pp. 56–65, April 2005.

[2] D. Kim and G.L. Stuber, "Clipping noise mitigation for OFDM by decision-aided reconstruction," *Communications Letters, IEEE*, vol. 3, no. 1, pp. 4 –6, jan 1999.

[3] J. Tellado, L. M. C. Hoo, and J. M. Cioffi, "Maximum-likelihood detection of nonlinearly distorted multicarrier symbols by iterative decoding," *IEEE Trans. Commun.*, vol. 51, no. 2, pp. 218–228, Feb. 2003.

[4] F. Gregorio, S. Werner, T. I. Laakso, and J. Cousseau, "Receiver cancellation technique for nonlinear power amplifier distortion in SDMA-OFDM systems," *IEEE Trans. Veh. Technol.*, vol. 56, no. 5, pp. 2499–2516, Sept. 2007.

[5] H. Chen and A. M. Haimovich, "Iterative estimation and cancellation of clipping noise for OFDM signals," *Communications Letters, IEEE*, vol. 7, no. 7, pp. 305 – 307, july 2003.

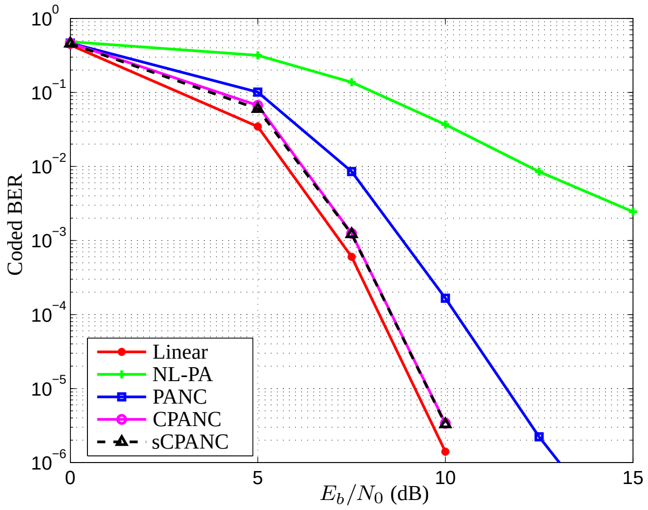

Fig. 5. BER after 3 iterations for $\gamma = 1$

[6] H. E. Rowe, "Memoryless nonlinearities with gaussian inputs: elementary results," *Bell Syst. Tech. J.*, vol. 61, pp. 1519–1525, September 1982.

[7] M. Friese, "On the degradation of OFDM-signals due to peak-clipping in optimally predistorted power amplifiers," in *Proc. IEEE Global Telecommun. Conf. GLOBECOM 98*, 8–12 Nov. 1998, vol. 2, pp. 939–944.

[8] A. Goldsmith, *Wireless Communications*, Cambridge University Press, 2005.

[9] F. H. Gregorio, S. Werner, J. E. Cousseau, and R. Wichman, "Broadband power amplifier distortion cancellation with model estimation in the receiver," in *Proc. IEEE 8th Workshop on Signal Processing Advances in Wireless Commun. SPAWC 2007*, 17–20 June 2007, pp. 1–5.