

Turbo-Coded OFDM Transmission Over a Nonlinear Channel

Hosein Nikopour, Amir K. Khandani, *Member, IEEE*, and S. Hamidreza Jamali

Abstract—In this paper, deliberate level clipping and turbo coding are combined to achieve an orthogonal frequency division multiplexing (OFDM) transmission system with a low peak-to-average power ratio (PAR) and a good performance. Using the linear approximation technique based on the minimum mean square error (MMSE) criterion, we first modify the metric computation for the turbo decoding in order to consider the distortion effects of the nonlinearity, caused by the Cartesian clipper. Also, this paper introduces a modified turbo decoder which simultaneously performs the data estimation and signal reconstruction. In other words, the turbo decoder iteratively recovers the clipped signal by using the estimated data, and then improves the data estimation by using the newly recovered signal. Numerical results are presented showing an improvement in the performance of the OFDM transmission system over the nonlinear channel, an increase in the efficiency of the high power amplifier (HPA), and/or an expansion of the transmitter coverage area.

Index Terms—Cartesian clipper, minimum mean square error (MMSE), nonlinear device, orthogonal frequency division multiplexing (OFDM), peak-to-average power ratio (PAR), turbo code.

I. INTRODUCTION

ORTHOGONAL frequency division multiplexing (OFDM), a popular type of multicarrier transmission [1]–[5], is an effective modulation technique for high-data-rate wireless and wire-line applications, including digital subscriber line (DSL) [6], digital audio broadcasting (DAB) [7], digital video broadcasting (DVB) [8], and wireless local area network (WLAN) [9]–[11]. The main advantage of OFDM is its ability to encounter multipath fading without requiring complex equalizers [12]. Moreover, OFDM is a bandwidth efficient transmission system and can be easily implemented by the fast Fourier transform (FFT) [13].

The superposition of several subcarrier signals leads to a Gaussian-like time domain OFDM waveform [2], [3] with significant envelope variations or a high peak-to-average power ratio (PAR) [14]. This undesirable feature renders the OFDM particularly sensitive to nonlinear distortions [15]–[19].

In practice, transmission devices such as the high power amplifier (HPA) have a limited dynamic range [20]. Therefore, to ensure a distortionless transmission, hardware with a high power back-off is required, but it restricts the OFDM system from utilizing the transmitted power effectively.

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H. Nikopour and A. K. Khandani are with the Department of Electrical and Computer Engineering, University of Waterloo, Waterloo, ON N2L 3G1, Canada (e-mail: hossein@cst.uwaterloo.ca; khandani@cst.uwaterloo.ca).

S. H. Jamali is with the Department of Electrical and Computer Engineering, University of Tehran, Tehran, Iran (e-mail: hjamali@ut.ac.ir).

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There are many approaches to deal with this power control problem [21]–[38]. One class of approaches is based on generating OFDM signals with a low PAR [21]–[33]. Recently, many PAR reduction methods have been proposed in the literature. The most widely known techniques in this category are based on selective mapping [21]–[24], phase shifting [24]–[28], or some form of coding [29]–[33]. Selective mapping and phase shifting offer a considerable reduction of the PAR, but at the price of a significant increase of the overall system complexity. Coding techniques with the capabilities of both PAR reduction, as well as error correction (Reed-Muller codes) [29]–[31], are attractive. However, these codes significantly reduce the overall throughput of the system, especially if there is a relatively large number of subcarriers [29], [30].

Level clipping is a simple technique to reduce the PAR of the OFDM signal [34]–[38]. However, clipping is a nonlinear process which distorts the transmitted signal [39]. OFDM signal can be clipped at the Nyquist rate or at over-sampling rates. Clipping the over-sampled signal produces in-band noise and out-of-band radiation, whereas clipping the Nyquist rate signal generates clipping noise that falls in-band resulting in a more serious degradation of the performance [34], [39]. Conventional coding techniques such as the turbo code [40]–[42] are attractive tools for mitigating such a high performance loss caused by the nonlinear distortion [43], [44]. Consequently, the combination of clipping and turbo coding can be used as an effective approach for both PAR reduction and error correction with a reasonable coding redundancy and system complexity.

To mitigate the nonlinear distortion, a turbo decoder requires knowledge of the characteristic of the clipping noise. This paper shows that a linear approximation of the clipper in the minimum mean square error (MMSE) [45] sense can be used as an accurate model for a nonlinear device. After this model is employed for metric computation, the turbo decoder implicitly incorporates for the effects of the clipping distortion in the decoding procedure.

If the nonlinear characteristic of the transmitter is known, the nonlinear distortion is a deterministic function of the transmitted data. Therefore, by using the estimated data, after some initial turbo decoding iterations, the turbo decoder can partially compensate for the effects of the signal clipping. In other words, the turbo decoder recursively recovers the clipped signal by using the estimated data, and then uses the newly recovered signal to improve its data estimation.

The idea of the decision-aided reconstruction of the clipped signal in an uncoded OFDM system is introduced in [46], and further developed in [47] by deriving a maximum-likelihood detector for an uncoded OFDM system. In this paper, however,

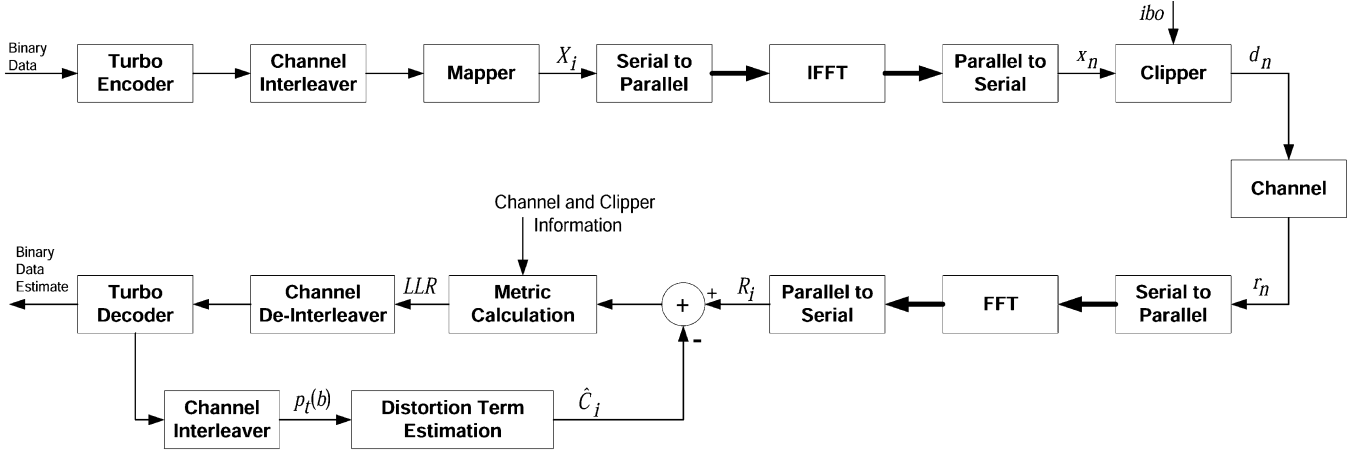


Fig. 1. Block diagram of the underlying OFDM transmission system.

this idea is applied to a turbo coded OFDM system by adopting jointly iterative detection and decoding at the receiver end. In addition, in contrast to these previous works, this paper considers the modified metric in the turbo decoding procedure. This can be effective especially when the OFDM signal is clipped with severe nonlinear distortion.

Although throughout this work we assume that the clipping operation is performed in the baseband at the Nyquist sampling rate and the clipper is an ideal soft limiter, the suggested technique can be easily generalized to other forms of nonlinearities. The main objective of this article is to show the effect of nonlinearity on the OFDM system performance. For this purpose, a simple channel model based on additive white Gaussian noise (AWGN) is considered to demonstrate the performance of the proposed method. This is an assumption which is also made in the previous works [46], [47]. The extension of the proposed method to the frequency selective fading channel is straightforward.

The rest of the paper is organized as follows. Section II is a brief description of the turbo coded OFDM transmitter under consideration. After the linear model of the clipper, based on the MMSE criterion, is derived in Section III, the linear model is adopted to modify the metric in Section IV. Section V introduces the corresponding turbo decoder and the process of the signal reconstruction. Section VI is devoted to some numerical examples. The paper finally concludes in Section VII.

Throughout this paper, $E\{\cdot\}$ denotes expectation, $j = \sqrt{-1}$, lowercase letters indicate the time domain samples, capital letters represent the frequency domain samples, and the letters in bold denote the vectors of these time or frequency domain samples.

II. OFDM OVERVIEW

The baseband model of the underlying OFDM transmission system is represented in Fig. 1. Each input block of k bits passes through a systematic turbo encoder [40] of rate $r_c (= k/\ell)$. The block of ℓ bits at the output of the turbo encoder is interleaved, and then mapped to $L (= \ell/m)$ complex samples, selected from

$\mathcal{X}$, an M -ary quadrature amplitude modulation (M -QAM) constellation in which each point represents $m (= \log_2 M)$ bits. This stream of L symbols is partitioned into L/N modulating vectors so that each has the dimension N . The final OFDM signal is the sum of the N subcarriers, each being modulated by the corresponding element of the modulating vector.

If $X_i, i = 0, 1, \dots, N-1$ is defined as the QAM symbol, associated with the i th element of the modulating vector $\mathbf{X}$, then the vector of the Nyquist rate samples of the OFDM baseband signal is expressed as

$$\begin{aligned} \mathbf{x} &= (x_0, x_1, \dots, x_{N-1}) \\ &= \text{IDFT}(\mathbf{X}) \end{aligned} \quad (1)$$

where for $n = 0, 1, \dots, N-1$,

$$x_n = \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} X_i e^{j \frac{2\pi i n}{N}}. \quad (2)$$

We assume that the average power of the QAM symbols is equal to $2\sigma^2$. Due to the orthogonality of the subcarriers, the average power of the OFDM samples is also the same and is equally distributed among the real and imaginary parts.

III. NONLINEARITY MODELING

The clipper is the source of the nonlinearity of the model in Fig. 1. Consequently, this section first describes the characteristics of the nonlinear device, and then models the distortion effects of the nonlinearity on the OFDM signal.

Typically, we can distinguish between two classes of nonlinearity [48]: 1) Cartesian distortion that acts separately on the real part and the imaginary part of the baseband signal (e.g., in D/A conversion), and 2) polar distortion that acts on the envelope and phase of the OFDM signal (e.g., in high power amplification). According to Fig. 2, the Cartesian distortion limits the signal within a square, whereas the polar distortion limits the signal in a circle. In this paper, without the loss of generality, we confine our study to the Nyquist rate Cartesian soft limiter, which belongs to the first class.

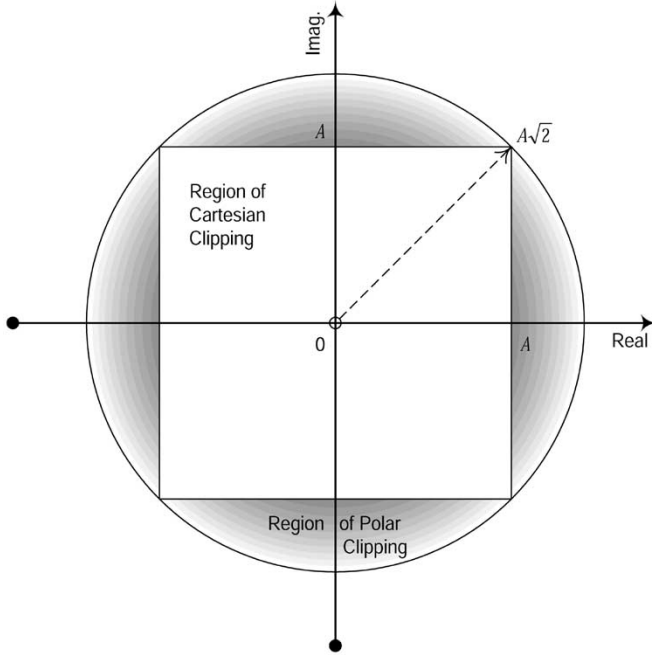


Fig. 2. Comparison of the polar and the Cartesian clipping. Υ_I is the same for the both types and equal to A^2/σ^2 .

Most of the common nonlinear devices in the transmitter can be accurately modeled as memoryless systems [47]. The distorted signal at the output of the memoryless Cartesian clipper is written as

$$d = g(x) = \begin{cases} -A, & x \leq -A \\ x, & -A < x < A \\ A, & x \geq A \end{cases} \quad (3)$$

where A is the saturation level. Since a Cartesian clipper acts on the real and imaginary parts of the complex signal separately, with some misuse of the notation, we use x in (3) to denote either the real or the imaginary part of the complex symbol x_n . If d is rewritten as follows,

$$d = x + c \quad (4)$$

then c represents the distortion term of the clipped signal.

We know that when N is a large number, x can be modeled as a Gaussian random variable [2], [3] with the probability density function (pdf), given by

$$p_x(x) = \frac{1}{\sigma} G(x/\sigma) \quad (5)$$

where $G(\tau) = 1/\sqrt{2\pi}e^{-\tau^2/2}$ is the zero-mean, unit-variance Gaussian function. As a result, the average power of the clipped OFDM sample can be obtained as follows:

$$\begin{aligned} \sigma_d^2 &= \int_{-\infty}^{\infty} g^2(x)p_x(x)dx \\ &= \sigma^2 \left[1 - 2Q(\sqrt{\Upsilon_I}) - \sqrt{2\Upsilon_I/\pi}e^{-\Upsilon_I/2} \right. \\ &\quad \left. + 2\Upsilon_I Q(\sqrt{\Upsilon_I}) \right] \end{aligned} \quad (6)$$

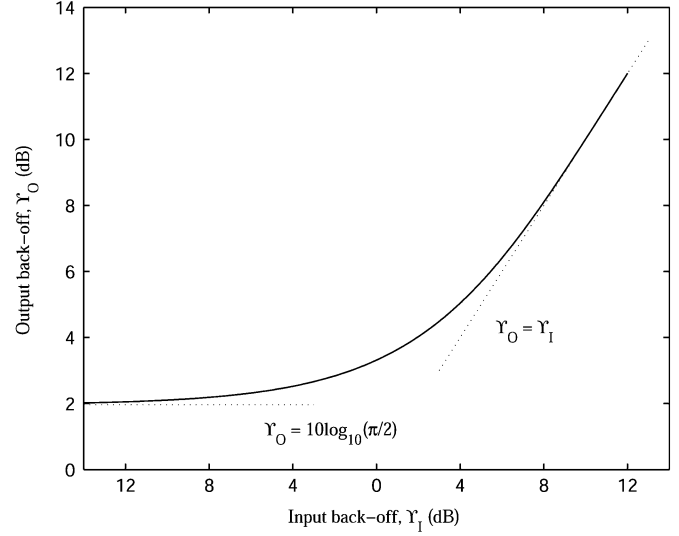


Fig. 3. Υ_O as a function of Υ_I .

where $Q(x) = \int_x^{\infty} G(\tau)d\tau$. The input back-off Υ_I is defined as the ratio of the maximum allowable input power to the average input power (e.g., refer to [10]), which is described as

$$\Upsilon_I = A^2/\sigma^2. \quad (7)$$

The output back-off Υ_O is another parameter of the clipper which indicates the relative level of the clipping, and is defined as the ratio of the maximum output power to the average power of the clipped signal, i.e. [10],

$$\Upsilon_O = A^2/\sigma_d^2. \quad (8)$$

Indeed, the parameter Υ_O is equal to the maximum PAR of the clipped signal. If (6) is substituted into (8), Υ_O is described as a function of Υ_I ,

$$\begin{aligned} \Upsilon_O &= \Upsilon_I / \left[1 - 2Q(\sqrt{\Upsilon_I}) - \sqrt{2\Upsilon_I/\pi}e^{-\Upsilon_I/2} \right. \\ &\quad \left. + 2\Upsilon_I Q(\sqrt{\Upsilon_I}) \right] \end{aligned} \quad (9)$$

which is illustrated in Fig. 3. We can see that for the high values of back-off, Υ_O and Υ_I are approximately the same.

Let us approximate the nonlinear function $g(x)$ by the linear function αx . The linear coefficient α can be set so that the error of the approximation

$$\epsilon = g(x) - \alpha x \quad (10)$$

is minimized in the mean square sense, i.e.,

$$\alpha = \arg \min_{\alpha} E\{\epsilon^2\}. \quad (11)$$

As a result,

$$\alpha = 1 - 2Q(\sqrt{\Upsilon_I}) \quad (12)$$

and

$$\begin{aligned} \sigma_{\epsilon}^2 &= E\{\epsilon^2\} \\ &= \sigma_d^2 - \alpha^2 \sigma^2 \end{aligned}$$

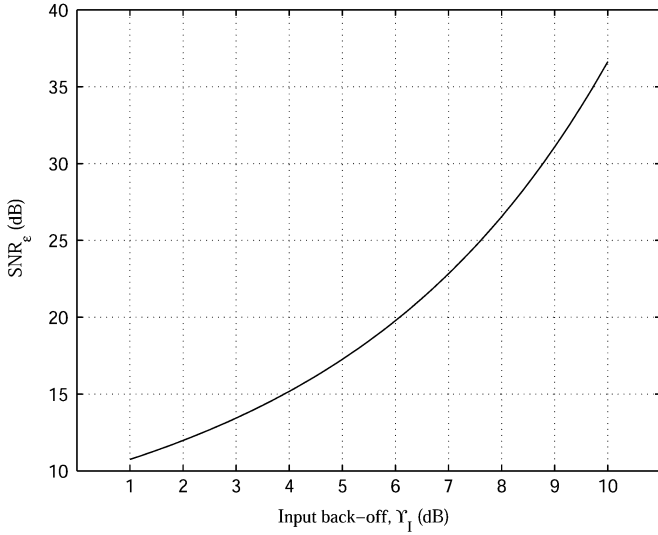


Fig. 4. Signal to nonlinear noise power ratio as a function of Υ_I for the clipper.

$$= \sigma^2 \left[2Q(\sqrt{\Upsilon_I})(1 - 2Q(\sqrt{\Upsilon_I})) - \sqrt{2\Upsilon_I/\pi} e^{-\Upsilon_I/2} + 2\Upsilon_I Q(\sqrt{\Upsilon_I}) \right]. \quad (13)$$

Consequently, according to the following equation,

$$d = \alpha x + \epsilon \quad (14)$$

the effect of the nonlinear distortion is modeled as an attenuation α and an additive “nonlinear” noise ϵ . In addition, due to the property of the MMSE approximation, the nonlinear noise is uncorrelated with the input signal x .

According to (14), the signal to nonlinear noise power ratio is defined as

$$\text{SNR}_{\epsilon} = \frac{\alpha^2 \sigma^2}{\sigma_{\epsilon}^2}. \quad (15)$$

If (12) and (13) are substituted into (15), SNR_{ϵ} is expressed as a function of Υ_I as follows:

$$\text{SNR}_{\epsilon} = (1 - 2Q(\sqrt{\Upsilon_I}))^2 / \left[2Q(\sqrt{\Upsilon_I})(1 - 2Q(\sqrt{\Upsilon_I})) - \sqrt{2\Upsilon_I/\pi} e^{-\Upsilon_I/2} + 2\Upsilon_I Q(\sqrt{\Upsilon_I}) \right]. \quad (16)$$

Fig. 4 exhibits that SNR_{ϵ} is obviously an ascending function of the Υ_I , because increasing Υ_I reduces the distortion effects of the clipper.

Although these equations are derived for the clipper, their generalization to other forms of $g(x)$ is straightforward.

IV. OFDM RECEIVER

The transmitted signal d_n is corrupted by an AWGN w_n with an average power equal to $2\sigma_w^2$. Therefore, the transmit signal to the AWGN power ratio is equal to

$$\text{SNR}_w = \frac{\sigma_d^2}{\sigma_w^2}. \quad (17)$$

The received signal

$$r_n = d_n + w_n \quad (18)$$

passes through the OFDM FFT demodulator where the output is expressed as

$$R_i = D_i + W_i. \quad (19)$$

The maximum a posteriori probability (MAP) [45], [49] receiver requires the knowledge of the conditional probability

$$P(X_i = X | R_0, R_1, \dots, R_{N-1}) \quad (20)$$

to compute the metrics that are used in the turbo decoding. The presence of the nonlinear device results in a dependency among the subcarriers. An ideal turbo decoder uses (20) to compute the turbo decoding metric. However, due to the enormous computational complexity that is associated with (20), we employ the following conditional probability:

$$P(X_i = X | R_i). \quad (21)$$

In fact, for the case of no clipping, X_i is correlated only with R_i and hence (21) is exact. However, when the OFDM signal is clipped, X_i and $\mathbf{R}$ become correlated and (21) is only an approximation of (20). The accuracy of this approximation depends on the level of clipping. The key point is that for the practical levels of clipping, X_i is mostly correlated with R_i and its correlation with other $R_k, k \neq i$ is negligible. Therefore, the iterative algorithm starts with this approximated metric and then tries to reconstruct the clipped signal in an iterative fashion.

If it is assumed that all constellation points are acquired with equal probability, the following conditional distribution probability is used to compute the turbo decoding metric:

$$p(R_i | X_i = X). \quad (22)$$

Using the linear model, we rewrite the received signal as

$$R_i = \alpha X_i + E_i + W_i \quad (23)$$

where $E_i, i = 0, 1, \dots, N-1$ is a zero-mean, nonlinear noise in the frequency domain that is described as

$$E_i = \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \epsilon_n e^{-j \frac{2\pi i n}{N}} \quad (24)$$

or

$$\mathbf{E} = \mathbf{DFT}(\epsilon). \quad (25)$$

Since N is large, by invoking the central limit theorem, E_i can be modeled as a Gaussian variable [15], [18], [48]. This is accurate, especially in a practical situation, where the clipping level A is not very large, causing most of the ϵ_n samples to have nonzero values. Although Gaussian approximation of the clipping noise is not valid for high back-offs (large clipping levels), numerical results will show there is no noticeable degradation for such high back-offs. The proposed algorithm would be used in low back-offs where the Gaussian approximation of the nonlinear noise is valid. Consequently, E_i is, approximately, a Gaussian variable, independent of X_i and with a power that

is equal to $2\sigma_\epsilon^2$. Considering this fact and according to (23), we represent the turbo decoder metric as

$$p(R_i | X_i = X) \propto \exp(-|R_i - \alpha X|^2 / 2\sigma_t^2) \quad (26)$$

where

$$2\sigma_t^2 = 2\sigma_\epsilon^2 + 2\sigma_w^2 \quad (27)$$

is the total noise power.

Without using the linear model, it is directly shown in the Appendix that for a large N

$$E\{R_i | X_i = X\} \approx \alpha X. \quad (28)$$

This is a match to (23) and (26), and therefore justifies the validity of the linear model.

According to (23), the signal to the total noise power ratio is

$$\text{SNR}_t = \frac{\alpha^2 \sigma^2}{\sigma_t^2} \quad (29)$$

which, by using (15), (17), and (27), is

$$\text{SNR}_t = \frac{\text{SNR}_\epsilon \text{SNR}_w}{1 + \text{SNR}_\epsilon + \text{SNR}_w}. \quad (30)$$

V. MODIFICATION OF THE TURBO-DECODER

The conventional turbo decoder cannot mitigate the effects of the nonlinear distortion. Therefore, to achieve a reasonable performance through a nonlinear transmission, the turbo decoder must be modified to correspond to the behavior of the channel. The block of the metric calculation in Fig. 1 uses $p(R_i | X_i = X)$ to compute the probability of the ρ th bit $\rho = 0, 1, \dots, m-1$, associated to the received signal R_i as follows:

$$P(b_\rho = b; R_i, i) \propto \sum_{X \in \mathcal{X}(\rho, b)} p(R_i | X_i = X). \quad (31)$$

In (31), $b \in \{0, 1\}$ and $\mathcal{X}(\rho, b) \subset \mathcal{X}$ is the set of the QAM points for which the ρ th bit is equal to b .

The logarithmic likelihood ratio (LLR) is related to the bit probabilities as follows [40]:

$$\text{LLR}_\rho(R_i, i) = \log \frac{P(b_\rho = 1; R_i, i)}{P(b_\rho = 0; R_i, i)}. \quad (32)$$

The modified turbo decoder uses (26) to incorporate the distortion caused by the nonlinearity. Substituting (26) in (32), we can write

$$\text{LLR}_\rho(R_i, i) = \log \frac{\sum_{X \in \mathcal{X}(\rho, 1)} \exp\left(\frac{-|R_i - \alpha X|^2}{2\sigma_t^2}\right)}{\sum_{X \in \mathcal{X}(\rho, 0)} \exp\left(\frac{-|R_i - \alpha X|^2}{2\sigma_t^2}\right)}. \quad (33)$$

Therefore, the turbo decoder requires the knowledge of the clipping parameters to calculate the modified metric.

As described in the standards [6]–[11], the pilots and data subcarriers are transmitted simultaneously in the same OFDM symbol. Therefore, the pilots like the data subcarriers are affected by the nonlinear device. Fortunately, in this situation, the clipper parameters can be found by using the pilots during the

channel and noise power estimation operations [50] without an additional estimator block. In this way, a conventional turbo decoder implicitly calculates the modified metric.

The sequence of the modified LLRs, corresponding to all ℓ coded bits after channel deinterleaving, passes through the turbo decoder to estimate k information bits.

This procedure assumes that the receiver has no knowledge of the distortion term and can only model the nonlinearity effects as signal attenuation and additive noise. The performance of the turbo decoder can be further enhanced by estimating the distortion term toward compensating the effect of the nonlinear device. According to (4), the received signal over the i th subcarrier can be written as

$$R_i = X_i + C_i + W_i. \quad (34)$$

Obviously, $\mathbf{C} = (C_0, C_1, \dots, C_{N-1})$ is a function of the transmitted QAM vector, $\mathbf{X}$

$$\mathbf{C} = \text{DFT}(\mathbf{g}(\text{IDFT}(\mathbf{X}))) - \mathbf{X} \quad (35)$$

where $\mathbf{g}(\mathbf{Y}) = (g(y_0), g(y_1), \dots, g(y_{N-1}))$.

Since $\mathbf{X}$ is unknown, the receiver must use an estimate of $\mathbf{C}$, denoted as $\hat{\mathbf{C}}$, by knowing $g(\cdot)$ and having an estimate of $\mathbf{X}$ from the previous iterations of the turbo decoding. This process can be continued in an iterative manner.

Let us assume $P_t(b), t = 0, 1, \dots, \ell - 1, b \in \{0, 1\}$ is the interleaved output of the turbo decoder which denotes the probability of the t th transmitted bit to be equal to b . Since the interleaved bits are approximately independent, the i th transmitted QAM symbol $i = 0, 1, \dots, L - 1$ is $X = \{b_0, b_1, \dots, b_{m-1} | b_\rho = 0 \text{ or } 1, 0 \leq \rho \leq m - 1\} \in \mathcal{X}$ with the probability $P_i(X)$ being expressed as

$$P_i(X) = P_i(\{b_0, b_1, \dots, b_{m-1}\}) \approx \prod_{\rho=0}^{m-1} P_{im+\rho}(b_\rho). \quad (36)$$

The estimate of X_i is defined as the average of the constellation points, i.e.,

$$\hat{X}_i = \sum_{X \in \mathcal{X}} P_i(X) X. \quad (37)$$

Let us define $\nabla_I(\cdot)$ as the previous procedure that estimates X_i s of the all transmitted OFDM symbols from the sequence of $\text{LLR}_\rho(R_i, i), \rho = 0, 1, \dots, m - 1, i = 0, 1, \dots, L - 1$ denoted as $\mathbf{LLR}$, i.e.,

$$(\hat{X}_0, \hat{X}_1, \hat{X}_2, \dots) = \nabla_I(\mathbf{LLR}) \quad (38)$$

where I is the number of turbo decoding iterations. The modified turbo decoding procedure can be summarized in the following steps, starting with $u = 1$. In the following discussion, the subscript within the parentheses denotes the stage number.

- 1) The LLR of the first stage $\mathbf{LLR}^{(0)}$ is calculated as a function of the received signal by utilizing (33).
- 2) The u th estimate of X_i s, after the $(u - 1)$ th reconstruction of the received signal, is [see (38)]

$$(\hat{X}_0^{(u)}, \hat{X}_1^{(u)}, \hat{X}_2^{(u)}, \dots) = \nabla_I(\mathbf{LLR}^{(u-1)}). \quad (39)$$

- 3) According to (35), the estimate of the distortion term is

$$\hat{\mathbf{C}}^{(u)} = \mathbf{DFT}(\mathbf{g}(\mathbf{IDFT}(\hat{\mathbf{X}}^{(u)}))) - \hat{\mathbf{X}}^{(u)}. \quad (40)$$

This procedure is performed on all the OFDM symbols.

- 4) The received signal is reconstructed as follows:

$$R_i^{(u)} = R_i - \hat{C}_i^{(u)}. \quad (41)$$

- 5) If we neglect the remaining distortion term after the signal reconstruction, then, according to (34) and (41), the new LLR can be written as

$$\text{LLR}_{\rho}^{(u)}(R_i^{(u)}, i) = \log \frac{\sum_{X \in \mathcal{X}(\rho, 1)} \exp\left(\frac{-|R_i^{(u)} - X|^2}{2\sigma_w^2}\right)}{\sum_{X \in \mathcal{X}(\rho, 0)} \exp\left(\frac{-|R_i^{(u)} - X|^2}{2\sigma_w^2}\right)}. \quad (42)$$

- 6) The next stage of the signal reconstruction begins with $u \leftarrow u + 1$. Go to step 2.

If U represents the number of times that the turbo decoder with I iterations is used, then the total number of the turbo decoding iterations is $H = IU$. Therefore, this modified decoding algorithm is comparable with the classical turbo decoder with a total of H iterations. However, this algorithm requires two additional FFTs and one additional LLR computation in the decoder for each stage of the signal reconstruction.

VI. NUMERICAL RESULTS

In all of the following, the OFDM system has a turbo code with a rate of $1/2$, a memory size of 4, and an interleaver length of 4096, and therefore the channel interleaver has the length of 8192 bits.

Fig. 5 displays the performance of the 256-QAM OFDM system for the different subcarrier numbers. The vertical axis is the bit-error rate (BER), and the horizontal axis is the signal to noise ratio per information bit that is defined as

$$\frac{E_b}{N_o} = \frac{\text{SNR}_w}{r_c m}. \quad (43)$$

The turbo decoder has 12 iterations without any signal reconstruction, and the turbo decoder uses only the modified metric. From this point in the paper, this type of turbo decoder is called the “classical decoder.” In Fig. 5, Υ_I is set to 5 dB. The accuracy of the modified metric depends on the number of subcarriers. As plotted in Fig. 5, the performance of the 1024-subcarrier system is better than the system with $N = 64$ and $N = 256$. Although the metric modification significantly improves the performance, there is still a substantial performance degradation compared to the performance of the system in an ideal AWGN channel in which the signal is not clipped. The suggested iterative signal reconstruction method tries to compensate for this performance loss.

The results of exploiting the proposed turbo decoder (the modified decoder) are illustrated in Fig. 6. The example of curves for $\Upsilon_I = 5$ dB given in Fig. 6 provides an indication of the performance of the proposed method in a case in which

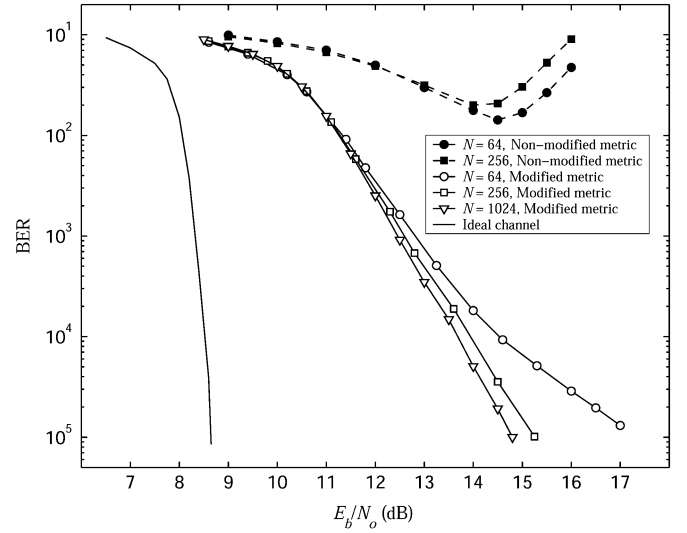


Fig. 5. Effect of the modified metric on 256-QAM OFDM systems with different subcarrier numbers. $\Upsilon_I = 5$ dB.

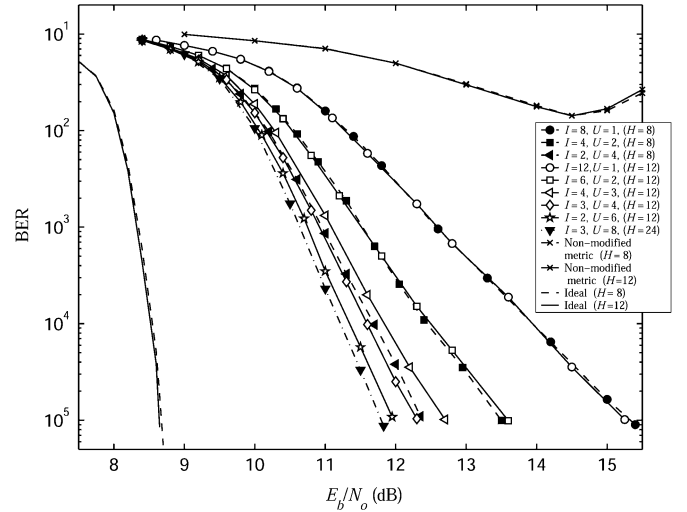


Fig. 6. Performance of the 256-subcarrier, 256-QAM OFDM system with the modified turbo decoder. $\Upsilon_I = 5$ dB.

the signal is distorted severely. As this figure shows, the system achieves a better performance if the number of the signal reconstruction iterations is increased. It seems that U is a dominant factor in comparison to I . For example, a system with $I = 2$ and $U = 4$ ($H = 8$) has approximately the same performance as the system with $I = 3$ and $U = 4$ ($H = 12$). However, for $H = 12$, it is possible to increase U to 6, which leads to a further improvement. As Fig. 6 confirms, there is a little improvement for $U > 6$; consequently, the performance of the system with $I = 3$ and $U = 8$ ($H = 24$) appears to be the best that can be achieved by this modified turbo decoder. In this case, the degradation in performance in comparison with the ideal AWGN channel is approximately 2.4 dB at a BER of 10^{-4} . This gap cannot be recovered by the proposed method because of the severe signal distortion for $\Upsilon_I = 5$ dB. In other words, the power of the remaining nonlinear noise after some iterations of the signal

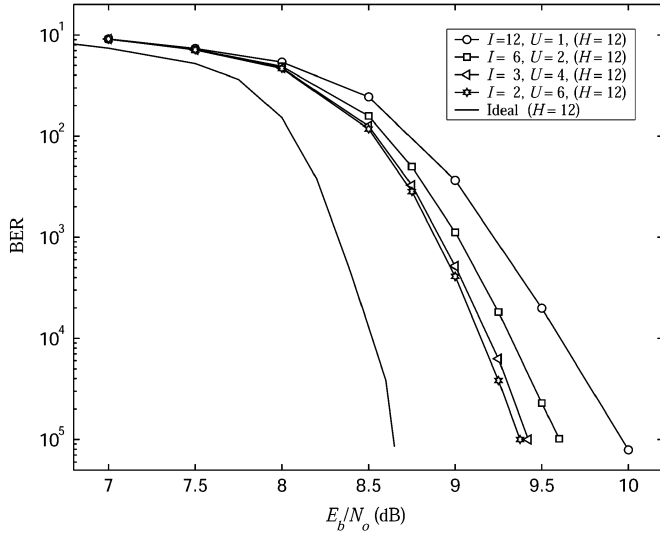


Fig. 7. Performance of the 256-subcarrier, 256-QAM OFDM system with the modified turbo decoder. $\Upsilon_I = 7$ dB.

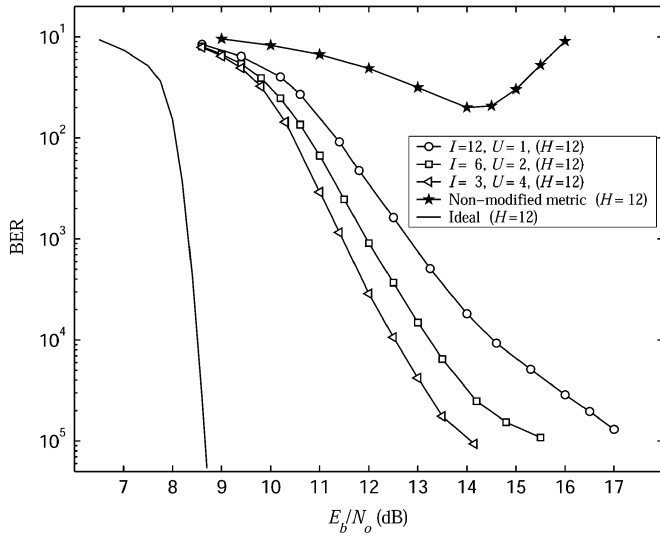


Fig. 8. Performance of the 64-subcarrier, 256-QAM OFDM system with the modified turbo decoder. $\Upsilon_I = 5$ dB.

reconstruction is still high, and this remaining noise degrades the system performance. Obviously, as shown in the next examples, by increasing Υ_I , this gap reduces. However, in some practical situations, for example when the channel is in a deep fade, the transmitter needs to increase its power and would work near the saturation level of the HPA. In Fig. 7, we increase the clipping level by increasing Υ_I to 7 dB. Obviously, due to less clipping distortion, the remaining degradation, with respect to the ideal channel, is less than 0.8 dB for $I = 2$ and $U = 6$ ($H = 12$).

In Fig. 8, we decrease the number of the subcarriers from 256 to 64. The proposed decoder still converges so that the performance is improved but at a slower rate in comparison to Fig. 6.

Fig. 9 depicts the BER of the 1024-subcarrier, 256-QAM OFDM system for practical conditions as a function of the amplifier clipping level Υ_O . According to (8), Υ_O is a clipping

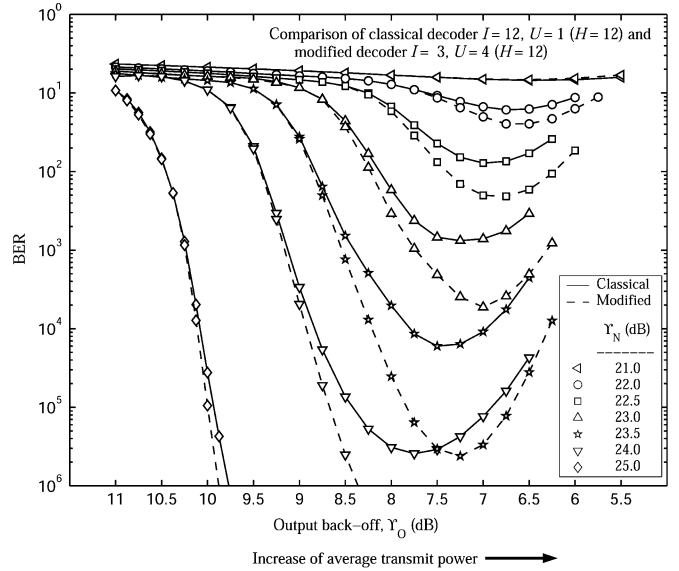


Fig. 9. Comparison of the modified, $I = 3, U = 4$ ($H = 12$), and the classical, $I = 12, U = 1$ ($H = 12$), turbo decoder in a practical situation. The results are for a 1024-subcarrier, 256-QAM OFDM system.

parameter which indicates the average power of the clipped signal. Since in practice the transmit power over the channel is of interest, instead of Υ_I , the parameter Υ_O is selected to show BER as a function of the average transmit power. As shown in Fig. 3, for practical cases ($\Upsilon_I > 6$ dB), Υ_I and Υ_O are approximately equal. Therefore, if one is interested in the Υ_I -BER relationship, this is approximately the same as the Υ_O -BER relationship. In general, the Υ_I -BER relationship can be obtained by combining Figs. 3 and 9.

The results of Fig. 9 are for several values of the noise back-off Υ_N , which is defined as follows:

$$\Upsilon_N = \frac{A^2}{\sigma_w^2}. \quad (44)$$

In fact, Υ_N is the maximum achievable signal to noise power ratio in a transmission channel for which the maximum transmit power is $2A^2$, and the noise power of the channel is $2\sigma_w^2$. Consequently, Υ_N indicates the quality of the transmission channel. When the channel is in a deep fade (or equivalently the relative noise power is high), Υ_N is low; in contrast, when the received power is high, the channel is in a good condition, and Υ_N has a relatively high value.

Fig. 9 exhibits that for a fixed Υ_N , when Υ_O is high, and as a result the HPA works in the linear region, by decreasing Υ_O (or equivalently by increasing the average transmit power), the system performance improves. However, this situation changes when the HPA approaches the saturation level, and the nonlinear noise becomes dominant. Therefore, after a specific Υ_O , which is the optimum operating point of the HPA, despite an increase in the transmit power, the nonlinear noise distorts the transmit signal and prevents the system from achieving a better performance. Consequently, the HPA is not able to effectively utilize all of its dynamic range due to the nonlinearity. This causes the system to fail, especially when the receiver is far from the

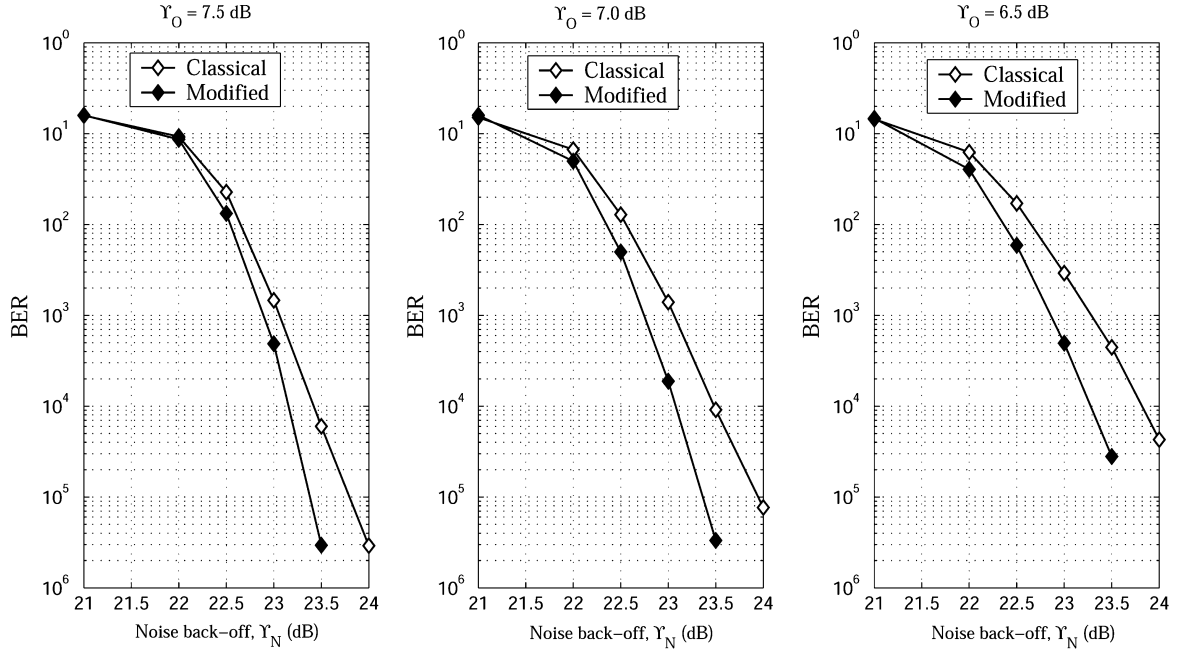


Fig. 10. The BER as a function of Υ_N for $\Upsilon_O = 6.5, 7.0, 7.5$ dB. A comparison of the modified turbo decoder, $I = 3, U = 4$ ($H = 12$), and the classical turbo decoder, $I = 12, U = 1$ ($H = 12$). The results are for a 1024-subcarrier, 256-QAM OFDM system.

transmitter, or when the channel is in deep fade, and so Υ_N is relatively low. For example, for $\Upsilon_N = 22.5$ dB, the classical decoder is not able to achieve a BER of less than 1.27×10^{-2} .

Fig. 9 compares the performance of the classical turbo decoder $I = 12, U = 1$ ($H = 12$), and the modified turbo decoder $I = 3, U = 4$ ($H = 12$). When the channel has a good quality, for example $\Upsilon_N = 25$ dB, both decoders behave approximately the same, because the system can achieve a good performance with a relatively low transmit power in which Υ_O is sufficiently high, and the HPA behaves like a linear system. However, if the channel quality degrades, for example Υ_N drops to 23 dB, then the transmitter has to increase the transmit power to keep the system performance within a reasonable range at an acceptable level. In this situation, the classical decoder cannot offer a performance that is better than 1.3×10^{-3} , whereas the best BER of the modified decoder is approximately 1.88×10^{-4} . Therefore, the ability of the modified decoder to improve the system performance is especially pronounced in a degraded channel condition (when the transmit power is required to be high, and the HPA is forced to operate in the nonlinear region with a comparatively low Υ_O).

Fig. 10 illustrates the BER of the classical and the modified decoder as a function of Υ_N and for $\Upsilon_O = 6.5, 7.0$, and 7.5 dB. These results are extracted from Fig. 9. Fig. 10 depicts that the modified decoder can handle the worse channel quality while holding the BER at the same level as the BER of the classical decoder operating in a better channel condition. For example, the gain of Υ_N is approximately 0.6 dB for $\text{BER} = 10^{-5}$ and $\Upsilon_O = 7.0$ dB, indicating that for the modified decoder, the signal attenuation can be 0.6 dB greater, in comparison to that of the classical decoder, while maintaining a BER that is equal to 10^{-5} . Therefore, by exploiting the modified decoder, the transmitter can either expand its coverage area without requiring

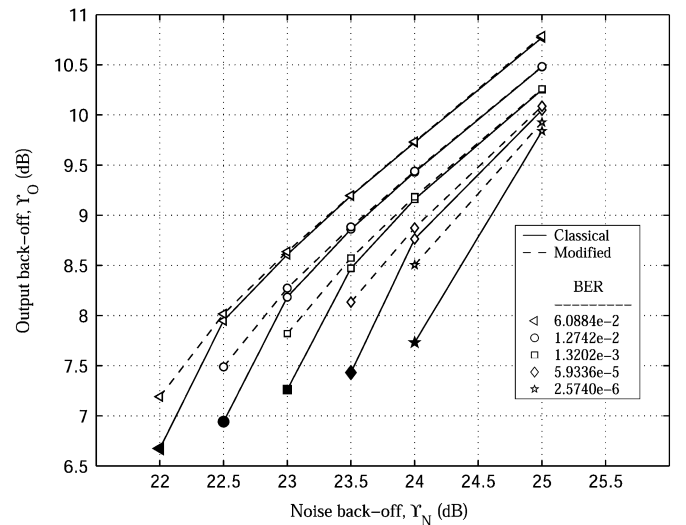


Fig. 11. Υ_O as a function of Υ_N for the several BERs. A comparison of the modified turbo decoder, $I = 3, U = 4$ ($H = 12$), and the classical turbo decoder, $I = 12, U = 1$ ($H = 12$). The results are for a 1024-subcarrier, 256-QAM OFDM system.

additional power, or retain its coverage area, and the average output power is decreased by 0.6 dB.

The average transmit power gain, with the value of the BER as a parameter, is offered in Fig. 11, in which the classical and modified decoder are compared. These BERs are selected at the optimum performance of the classical decoder so that the dark points correspond to the optimum operating point of the HPA for a given Υ_N . According to this figure, the modified decoder can achieve the same performance as the classical one but with a lower transmit power (higher Υ_O). This is the case for Υ_O at approximately the optimum point of operation, where Υ_O

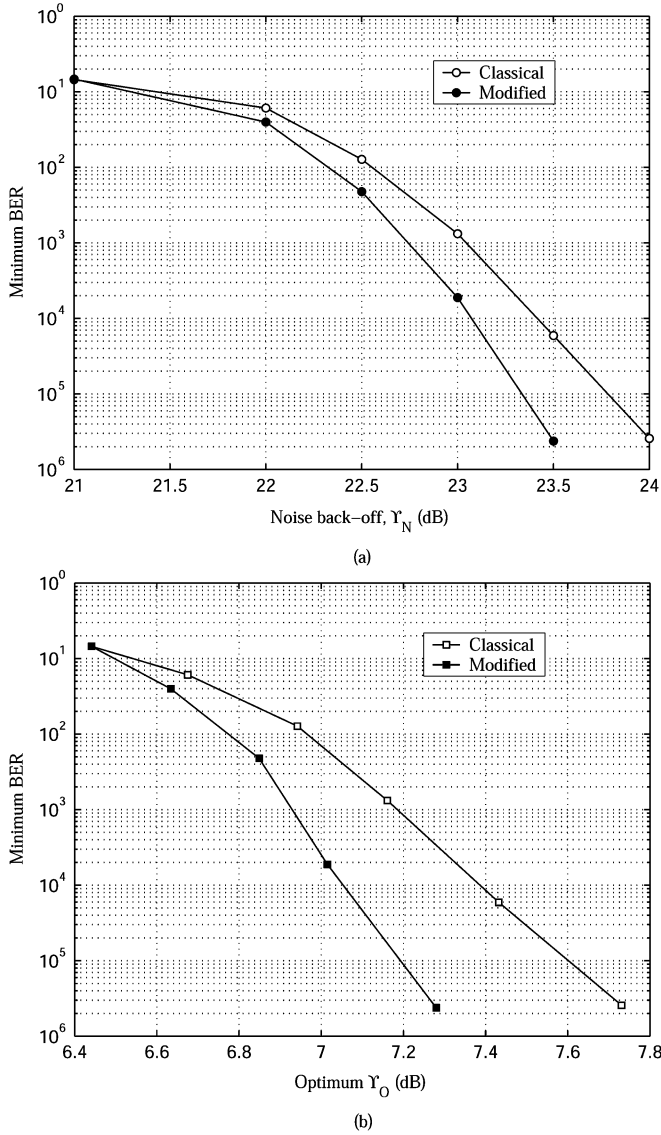


Fig. 12. Comparison of the modified turbo decoder, $I = 3, U = 4$ ($H = 12$), and the classical turbo decoder, $I = 12, U = 1$ ($H = 12$) at the optimum operating points. The results are for a 1024-subcarrier, 256-QAM OFDM system. (a) The minimum BER as a function of Υ_N . (b) The minimum BER as a function of the optimum Υ_O .

is relatively low, and the HPA operates near the nonlinear region. Therefore, by using the modified decoder, the transmission system can save the power through the higher Υ_O selection.

The behavior of the classical and the modified decoder at the optimum operating points are compared in Fig. 12(a) and (b). The minimum BER is shown as a function of Υ_N in Fig. 12(a). Similar to Fig. 10, this figure signifies that the functionality threshold of the modified decoder is approximately 0.5 dB higher than the functionality threshold of the classical one. Fig. 12(b) illustrates the minimum BER as a function of the optimum Υ_O . Each point in each curve indicates the optimum operation point of the amplifier. As shown in Fig. 12(b), the optimum Υ_O of the modified decoder is approximately 0.4 dB less than the optimum point of the classical decoder. This demonstrates that the modified decoder utilizes the dynamic range of

the HPA more effectively; i.e., the HPA efficiency, in converting the DC to AC power, is better when the system performs at the optimum operating point. It is clear that with the same BER, the channel quality of the modified receiver is worse than that of the classical one.

VII. CONCLUSION

The combination of clipping and turbo coding is used to reduce the PAR and, simultaneously, achieve an acceptable performance with a reasonable redundancy. First, we demonstrate that the linear MMSE approximation of the memoryless, nonlinear device leads to an almost optimal symbolwise metric. This linear model is more accurate for a large number of subcarriers. If the turbo decoder metric is computed with this model, the turbo decoder can reduce the BER by several orders of magnitude. In the next step, the turbo decoder is modified to recursively reconstruct the distorted signal during the decoding iterations. The numerical examples confirm that for a fixed number of decoding iterations, the modified turbo decoder results in a noticeable improvement in the BER with a slight increase in the complexity. Finally, the modified decoder is tested in a practical situation. The simulation results prove that a system in which the modified decoder is exploited can develop its coverage area and use the dynamic range of the HPA more effectively. Although this work focuses on the AWGN channel, the proposed approach can be easily extended to the frequency selective fading channel. The only effect of the frequency selective channel is that the SNR varies through the subcarriers due to the different channel gain. Therefore, the metric in (26) must be modified to consider the different fading channel gain over each subcarrier. In addition, in the frequency selective fading channel case, the joint detection-decoding-reconstruction can be considered as an iterative tool to combat both the clipping and fading destructive effects simultaneously.

APPENDIX

PROOF OF (28)

Let us start with $E\{d_n | X_i = X\}$, $n, i = 0, 1, \dots, N-1$. According to (2), $x_n, n = 0, 1, \dots, N-1$ can be rewritten as follows:

$$x_n = \bar{x}_n^{(i)} + \frac{1}{\sqrt{N}} \sum_{q=0}^{N-1} X_q e^{j \frac{2\pi q n}{N}} \quad (45)$$

where

$$\bar{x}_n^{(i)} = \frac{1}{\sqrt{N}} X e^{j \frac{2\pi i n}{N}} \quad (46)$$

is the mean of $x_n | X_i = X$. When N is large, the summation term in (45) is, approximately, a complex Gaussian process with a variance that is expressed as follows:

$$2\sigma^2 = \frac{N-1}{N} 2\sigma^2. \quad (47)$$

Therefore, if x and $\bar{x}$ respectively denote either the real part or the imaginary part of x_n and $\bar{x}_n^{(i)}$, then the pdf of $x | X_i = X$

can be expressed as follows:

$$p_{x|X}(x|X_i = X) = \frac{1}{\bar{\sigma}} G\left(\frac{x - \bar{x}}{\bar{\sigma}}\right). \quad (48)$$

According to (3), the clipped signal is a function of x , and then we can easily show that

$$E\{d|X_i = X\} = \int_{-\infty}^{\infty} g(x)p_{x|X}(x|X_i = X)dx. \quad (49)$$

If (3) and (48) are substituted into (49), after some manipulations, we can write

$$E\{d|X_i = X\} = \sigma\{[G(\xi^+) - G(\xi^-)] - [\xi^+Q(\xi^+) - \xi^-Q(\xi^-)]\} + \bar{x} \quad (50)$$

where

$$\xi^{\pm} = \frac{A \pm \bar{x}}{\bar{\sigma}}. \quad (51)$$

If $N \rightarrow \infty$, it is clear from (46), (47), (51), and (7) that

$$\begin{cases} \bar{x} \rightarrow 0 \\ \bar{\sigma}^2 \rightarrow \sigma^2 \end{cases} \Rightarrow \begin{cases} \xi^{\pm} \rightarrow \sqrt{\Upsilon_1} \\ \xi^+ - \xi^- \rightarrow \frac{2\bar{x}}{\sigma} \end{cases} \quad (52)$$

and then

$$\begin{cases} G(\xi^+) - G(\xi^-) \rightarrow 0 \\ \xi^+Q(\xi^+) - \xi^-Q(\xi^-) \rightarrow \frac{2\bar{x}}{\sigma}Q(\sqrt{\Upsilon_1}) \end{cases}. \quad (53)$$

Therefore, according to (53) and (12), when N is large, (50) reduces to

$$E\{d_n|X_i = X\} \approx \alpha \bar{x}_n^{(i)}. \quad (54)$$

Transforming from the time to the frequency domain, (54) is changed for $h = 0, 1, \dots, N-1$ as follows:

$$E\{D_h|X_i = X\} \approx \frac{1}{\sqrt{N}} \sum_{n=0}^{N-1} \alpha \bar{x}_n^{(i)} e^{-j\frac{2\pi hn}{N}} \quad (55)$$

which, by using (46), is reduced to

$$E\{D_h|X_i = X\} \approx \alpha X \delta_{h-i} \quad (56)$$

where δ_{ρ} is a Dirac function. Therefore

$$E\{D_i|X_i = X\} \approx \alpha X. \quad (57)$$

Since W_i is a zero mean noise, (28) can be inferred from (19) and (57).

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Hosein Nikopour was born in Shiraz, Iran, in 1973. He received the B.Sc. degree from Shiraz University, Shiraz, in 1995, the M.Sc. degree from Sharif University of Technology, Tehran, Iran, in 1997, and the Ph.D. degree from the University of Tehran, Tehran, Iran, in 2004, all in electrical engineering. During the Ph.D. program, he was with the University of Waterloo, Waterloo, ON, Canada, for one year as a Visiting Scholar.

Dr. Nikopour is currently working as a Postdoctoral Fellow at the University of Waterloo. His research interests include wireless communications and coding theory.



Amir K. Khandani (M'02) received the M.A.Sc. degree from the University of Tehran, Tehran, Iran, and the Ph.D. degree from McGill University, Montreal, QC, Canada, in 1985 and 1992, respectively.

Following graduation, he worked for one year as a Research Associate at INRS-Telecommunications, Montreal. In 1993, he joined the Department of Electrical and Computer Engineering, University of Waterloo, Waterloo, ON, Canada, where he is currently working as a Professor. He currently holds an NSERC Industrial Research Chair funded by the Natural Sciences and Engineering Research Council of Canada (NSERC) and Nortel Networks on "Advanced Telecommunications Technologies" and a Canada Research Chair (Tier I) on "Wireless Systems."

Dr. Khandani is currently serving as an Associate Editor for the *IEEE Transactions on Communications* in the area of Coding and Communication Theory.



S. Hamidreza Jamali was born in Zanjan, Iran, in March 1955. He received the B.Sc. and M.Sc. degrees from University of Tehran, Tehran, Iran, in 1978 and 1980, respectively, and the Ph.D. degree from Concordia University, Montreal, QC, Canada, in 1991.

From 1982 to 1983, he worked at the Electronics Research and Production Center, a subsidiary of the Iran Broadcast Organization. In 1983, he joined the Department of Electrical and Computer Engineering, University of Tehran, as a Lecturer. While at Concordia University, he was a Research Assistant from

1987 to 1991 and Postdoctoral Fellow from January 1992 to September 1992 in the Department of Electrical and Computer Engineering. In September 1992, he rejoined the Department of Electrical and Computer Engineering, University of Tehran, Iran, where he is currently an Associate Professor. He also has been working as a Visiting Professor at the Department of Electrical and Computer Engineering, University of Waterloo, Waterloo, Canada, since November 2004. He is the coauthor of a book entitled *Coded Modulation Techniques for Fading Channel* (New York: Kluwer, 1994). His current research interests include applications of coding and diversity techniques to wireless communications.