PERFORMANCE ANALYSIS OF A MC-CDMA FORWARD LINK OVER NONLINEAR CHANNELS

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ABSTRACT

This paper presents an analytical framework for the performance evaluation of an MC-CDMA forward-link in the presence of nonlinear distortions induced by transmitter's high-power amplifiers. The statistical characterization of the decision variable used for data detection is carried out through the application of the Bussgang theorem. Simulation results validate the accuracy of the proposed method under typical operating conditions[1].

1. INTRODUCTION

Multicarrier Code Division Multiple Access (MC-CDMA) schemes can be considered as very promising candidates for next generation high data rate wireless systems due to their robustness against harsh propagation conditions in multipath fading channels, together with a flexible and efficient multiple access capability [1]-[2]. However, like most multicarrier systems, MC-CDMA exhibits a considerable vulnerability to nonlinear distortions induced by high power amplifiers (HPAs) operating at, or near, the saturation region. Actually, the large peaks in the MC-CDMA signal amplitude (mainly due to its multicarrier structure) give rise to intermodulation products affecting both the MC-CDMA signal itself (as in-band distortion) and the adjacent channels (as out-of-band components) [3], resulting in significant performance degradations that have to be adequately dealt with.

Many works available in the literature, e.g., [4]-[5], have been devoted to the theoretical description of the nonlinear distortion effects. In [4], the authors propose a theoretical framework based on the extension of the Bussgang theorem for bandpass memoryless nonlinearities with complex Gaussian nonzero-mean non-stationary inputs. This approach leads to a fully analytical evaluation of the error probability for Orthogonal Frequency Division Multiplexing (OFDM) systems over nonlinear channels, but appears to be not straightforward applicable in the context of MC-CDMA systems due to the inherent different features of the transmitted signal.

The aim of this contribution is to develop an analytical framework to evaluate the performance degradation of a MC-CDMA forward link due to in-band distortion caused by nonlinear HPA devices. In particular, we consider a MC-CDMA system, originally proposed in [6], wherein the transmitted subcarriers are multiplied by a pseudo-noise scrambling sequence. The above arrangement helps not only in reducing the peak-to-average-power-ratio (PAPR), but also makes easier the application of the extension of the Bussgang theorem to the problem at hand. As detailed in the sequel, this will allow us to achieve a comprehensive statistical description of the decision variable, and accordingly, an accurate enough evaluation of the bit error rate (BER) link performance. A comparison between analytical findings and simulation results is carried out as well, in order to assess the accuracy of the proposed analytical framework.

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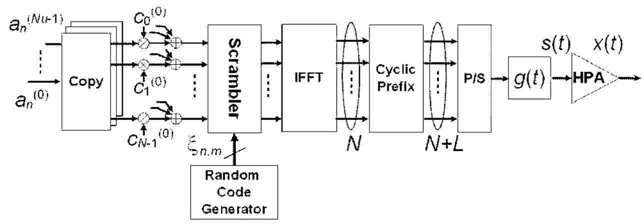

Figure 1: Block diagram of the MC-CDMA transmitter, featuring scrambling and amplification.

The paper organization is as follows. Section II contains a brief discussion on the MC-CDMA system model, while Section III is devoted first to the characterization of the HPA input and output signals, followed by a comprehensive statistical description of the variable at the input of the decision device. The numerical evaluation of BER link performance is illustrated in Section IV together with simulation results to test the accuracy of the proposed method. Finally, in Section V we draw some concluding remarks.

2. SIGNAL MODEL

The transmission scheme considered in this study is the forward link of a MC-CDMA system wherein the base station (BS), whose block diagram is depicted in Fig. 1, serves N_u mobile stations (MSs) through a common nonlinear channel with additive white Gaussian noise (AWGN) having power spectral density $\mathcal{N}_0/2$. Here, the n-th information-bearing symbol of $k-$th user $a_n^{(k)}$, belonging to a $M-$QAM constellation, is first copied into N branches (the same number as the subcarriers), and then, multiplied by the chip spreading sequence $c_m^{(k)} \in \{\pm 1\}$, $0 \leq m \leq N-1$, i.e., the user's channelization code. Next, all of the m-th components of the spread data coming from the N_u active users are summed together, yielding an N-dimensional block. The latter, according to the transmission scheme proposed in [6], is further multiplied by a cell-specific pseudo-noise (PN) scrambling sequence, which greatly helps in reducing the PAPR of the multicarrier transmitted signal. Denoting with $\xi_{n,m} \in \{\pm 1\}$, $0 \le m \le N-1$, the PN scrambling sequence used for the n-th block, we assume that its repetition period is chosen so long that it can be modeled as independent binary random variables with

$$E\{\xi_{n,m}\xi_{s,l}\} = \begin{cases} 1 & n=s, m=l \\ 0 & \text{otherwise} \end{cases}. \quad (1)$$

The frequency mapping of the N scrambled samples to the N available subcarriers is achieved by using an IFFT (Inverse Fast Fourier Transform) unit. Further, to keep the subcarriers orthogonal with each others and so avoid interference between successive symbols (at least under ideal channel conditions), a conventional cyclic prefix (CP) made of L samples is inserted at the beginning of each IFFT output block. Eventually, pulse shaping is accomplished by a filter whose impulse response $g(t)$ is a root-raised-cosine pulse with rolloff α and whose Fourier transform is $G(f)$. Denoting with $T_s \triangleq (N+L)T$ the transmitted block interval including the CP, and T the chip interval, the complex envelope of the resulting MC-CDMA signal at the HPA input can be expressed as

$$s(t) = \sum_{k=1}^{N_u}\sum_{n} a_n^{(k)} \sum_{l=-L}^{N-1} \gamma_{n,l}^{(k)} g(t-nT_s-lT), \quad (2)$$

where

$$\gamma_{n,l}^{(k)} \triangleq \frac{1}{\sqrt{N}}\sum_{m=0}^{N-1} c_m^{(k)}\xi_{n,m} e^{j2\pi\frac{lm}{N}}, \quad -L \le l \le N-1. \quad (3)$$

The HPA is modelled as a nonlinear memoryless device, with normalized AM/AM and AM/PM responses[2] [2]-[3]

$$\begin{cases} M(\rho) = \frac{\rho}{(1+\rho^q)^{1/q}} \\ \Phi(\rho) = 0 \end{cases}, \quad (4)$$

where $\rho \triangleq |s(t)|$ is the instantaneous amplitude of the signal at the HPA input, while q is an integer-valued parameter which defines the smoothness of the transition between the linear region and the saturation one

At the MS receiver of the intended user (see Fig. 2) the rate $1/T$ samples at the output of the matched filter (MF) are collected into blocks of size $N+L$. After removal of the CP, they are transformed by a FFT unit of size N, despread by multiplying with the intended user's signature, and eventually, after channel equalization, are fed to the decision device. Since, our interest addresses the performance degradation due to the HPA distortions only, in the sequel carrier and both code and chip timing synchronization will be assumed ideal.

3. ANALYSIS OF NONLINEAR DISTORTION

The goal of this section is to derive an analytical framework that can support the theoretical performance evaluation of a forward MC-CDMA link, thereby avoiding time-consuming simulation runs. First in Section 3.1, we will describe the composite MC-CDMA signal at the input of the nonlinear block. Then, in Section 3.2, will follow a characterization of the transmitted signal at the output of the HPA based on the extended version of the Bussgang theorem as proposed in [4]. Finally in Section 3.3, we will pursue the statistical description of the decision variable that can be found at the input of the decision device of the intended user.

[2]The normalization assumption means that the average power of the signal at the HPA input, without any back-off, is coincident with the saturation point of the amplifier.

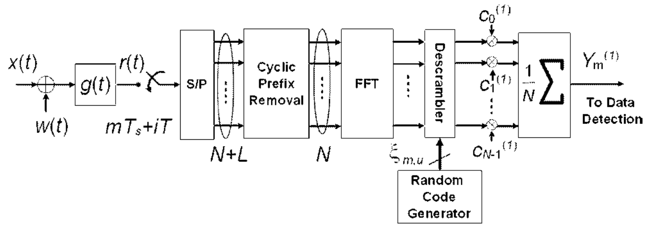

Figure 2: Block diagram of the MC-CDMA receiver, featuring de-scrambling.

3.1 Description of the HPA input signal

Let us begin first with a basic assumption concerning the statistical characterization of the MC-CDMA signal given by (2), that feeds the nonlinear HPA. Since the users' contribution have same statistics and are independent with each others, by invoking the central limit theorem, we claim that the latter can be considered as a complex Gaussian random process, provided that N_u is sufficiently large. For the sake of presentation simplicity, and without any loss of generality, we will focus on the link relevant to the user $k=1$. Also, in the sequel we will consider (2) conditioned to the transmission of the generic symbol $a_n^{(1)} = \bar{a}$. Taking into account that the information symbols and the PN scrambling sequence are zero-mean independent random variables, it can be shown that the conditioned mean value of (2) turns out to be

$$\eta_s(t) \triangleq E\left\{ s(t)|\, a_n^{(1)} = \bar{a}\right\} \frac{1}{\sqrt{N}}\bar{a}\sum_{n}\sum_{l=-L}^{N-1}\sum_{m=0}^{N-1} c_m^{(1)} \cdot E\{\xi_{n,m}\} e^{j2\pi\frac{lm}{N}} g(t-nT_s-lT) = 0, \quad (5)$$

whereas its variance is given by

$$\sigma_s^2(t) \triangleq \left\{ |s(t)|^2 \middle| a_n^{(1)} = \bar{a}\right\} \approx A_2(N_u-1)\sum_{n}\sum_{l=-L}^{N-1} g^2(t-nT_s-lT) + |\bar{a}|^2\sum_{n}\sum_{l=-L}^{N-1} g^2(t-nT_s-lT), \quad (6)$$

where $A_2 \triangleq E\{|a_n^{(k)}|^2\}$. Assuming a large number of active users, i.e., $N_u >> 1$, the above expression can then be rearranged as

$$\sigma_s^2(t) \approx A_2 N_u \sum_{n}\sum_{l=-L}^{N-1} g^2(t-nT_s-lT). \quad (7)$$

From (5) and (6), we draw a couple of remarks: *i)* the HPA input signal $s(t)$ is a cyclostationary (CS) Gaussian random process with period T; *ii)* the mean value and the variance conditioned to a given transmitted symbol $a_n^{(1)} = \bar{a}$ of the intended user are independent from the actual value of the symbol itself.

3.2 Characterization of the HPA output signal

Recalling the statistical features of the HPA input signal discussed so far, and applying the extended version of the Bussgang theorem, it can be shown ([4]-[5], [7]-[8]) that signal at the output of the HPA can be written as

$$x(t) = \beta(t)s(t) + \nu(t). \qquad (8)$$

In (8), $\beta(t)$ is a deterministic function given by

$$\beta(t) = \frac{1}{2}\int_0^{\infty}\left[M'(\rho) + \frac{M(\rho)}{\rho}\right]\frac{2\rho}{\sigma^2(t)}e^{-\frac{\rho^2}{\sigma^2(t)}}d\rho, \qquad (9)$$

where $M(\rho)$ is the AM-AM characteristics of the HPA and $M'(\rho)$ its derivative, while $\sigma^2(t)$ is a scaled version of $\sigma_s^2(t)$ depending on the back off (BO) of the HPA[3]. Note that in general $\beta(t)$ is a complex-valued deterministic function, but in the case under study it takes real values since the AM-PM response of the HPA is null. Furthermore, $\nu(t)$ is a non-Gaussian zero-mean additive noise process, labelled in the sequel as nonlinear distortion noise (NLDN), which is uncorrelated with the input process $s(t)$. A close inspection of (7) reveals that $\sigma^2(t)$ is periodic with period T, and therefore, from (9), the function $\beta(t)$ also reveals periodic with the same period. From the considerations above, we can write

$$\beta(t) = \beta_0\left[1+\varphi(t)\right], \qquad (10)$$

where β_0 is the value of $\beta(t)$ corresponding to the case of feeding the HPA by a signal with a constant power equal to the time-average of $\sigma_s^2(t)$, whereas $\varphi(t)$ is a zero-mean T-periodic function that can be expanded as a Fourier series

$$\varphi(t) = \sum_h \Phi_h e^{j2\pi\frac{ht}{T}}, \qquad (11)$$

with

$$\Phi_h = \frac{1}{T}\int_{-T/2}^{T/2}\varphi(t)e^{-j2\pi\frac{ht}{T}}dt. \qquad (12)$$

3.3 Definition of the decision variable

Collecting together (8), (10)-(11), defining $p(t) \triangleq g(t)\otimes g(-t)$ as the overall Nyquist impulse response, and denoting with $w(t)$ the channel AWGN component, at the receiver of the intended user the matched filter output signal

$$z(t) = [x(t)+w(t)]\otimes g(-t) \qquad (13)$$

can be put in the alternative form

$$z(t) = z_1(t) + z_2(t) + \vartheta(t) + n(t), \qquad (14)$$

where

$$z_1(t) \triangleq \beta_0 s(t)\otimes g(-t) = \beta_0\sum_{k=1}^{N_u}\sum_n a_n^{(k)}\sum_{l=-L}^{N-1}\gamma_{n,l}^{(k)}p(t-nT_s-lT), \qquad (15)$$

$$z_2(t) \triangleq \beta_0[\varphi(t)s(t)]\otimes g(-t) = \beta_0\sum_h\Phi_h\sum_{k=1}^{N_u}\sum_n a_n^{(k)}\sum_{l=-L}^{N-1}\gamma_{n,l}^{(k)} \cdot\left[e^{j2\pi\frac{ht}{T}}g(t-nT_s-lT)\right]\otimes g(-t), \qquad (16)$$

$$\vartheta(t) \triangleq \nu(t)\otimes g(-t), \qquad (17)$$

and

$$n(t) \triangleq w(t)\otimes g(-t). \qquad (18)$$

Additionally, (16) can be further rearranged as

$$z_2(t) = \beta_0\sum_{h\in\{\pm1\}}\Phi_h\sum_{k=1}^{N_u}\sum_n a_n^{(k)}\sum_{l=-L}^{N-1}\gamma_{n,l}^{(k)}q_h(t-nT_s-lT), \qquad (19)$$

where the pulse

$$q_h(t) \triangleq \left[e^{j2\pi\frac{ht}{T}}g(t)\right]\otimes g(-t) = \int_{-\infty}^{+\infty}G(f-h/T)G^*(f)e^{j2\pi ft}df, \qquad (20)$$

takes non-zero values only for the index $h\in\{\pm1\}$, due the finite support of $G(f)$ in the interval $[-(1+\alpha)/2T,(1+\alpha)/2T]$.

Sampling (14) at the instants $mT_s+iT, 0\le i\le N-1$, yields the m-th block of N samples

$$z_{m,i} \triangleq z(mT_s+iT) = \beta_0\sum_{k=1}^{N_u}a_m^{(k)}\gamma_{m,i}^{(k)} + \beta_0\sum_{h\in\{\pm1\}}\Phi_h\sum_{k=1}^{N_u}a_m^{(k)}\sum_{l=-L}^{N-1}\gamma_{m,l}^{(k)}q_h[(i-l)T] + \vartheta_{m,i} + n_{m,i}, \qquad (21)$$

where $\vartheta_{m,i}$ and $n_{m,i}$ are the samples of NLDN and channel noise, respectively. After removing the CP and assuming negligible the inter symbol interference (ISI), (21) can be reasonably approximated as

$$z_{m,i} \approx \beta_0\sum_{k=1}^{N_u}a_m^{(k)}\gamma_{m,i}^{(k)} + \beta_0\sum_{h\in\{\pm1\}}\Phi_h\sum_{k=1}^{N_u}a_m^{(k)}\sum_{l=0}^{N-1}\gamma_{m,l}^{(k)}q_h[(i-l)T] + \vartheta_{m,i} + n_{m,i}. \qquad (22)$$

The above N samples are then FFT-transformed, yielding

$$Z_{m,u} = \frac{1}{\sqrt{N}}\sum_{i=0}^{N-1}z_{m,i}e^{-j2\pi\frac{ui}{N}} = Z_{m,u,1} + Z_{m,u,2} + Z_{m,u,3}, \qquad (23)$$

where

$$Z_{m,u,1} \triangleq \beta_0\sum_{k=1}^{N_u}a_m^{(k)}c_u^{(k)}\xi_{m,u}, \qquad (24)$$

[3] Under the normalization assumption of the HPA characteristics described in Sect. 2, the scaling factor of the variance turns out to be coincident with the input back-off.

$$Z_{m,u,2} \triangleq \frac{1}{N}\beta_0 \sum_{h\in\{\pm1\}} \Phi_h \sum_{k=1}^{N_u} a_m^{(k)} \cdot \sum_{l=0}^{N-1}\sum_{i=0}^{N-1} \gamma_{m,l}^{(k)} q_h[(i-l)T]e^{-j2\pi\frac{ui}{N}}, \quad (25)$$

and

$$Z_{m,u,3} \triangleq \frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} (\vartheta_{m,i}+n_{m,i})e^{-j2\pi\frac{ui}{N}}. \quad (26)$$

By descrambling and despreading with the user's signature, we end up to the desired expression of the decision variable for the intended user

$$Y_m^{(1)} = \frac{1}{N}\sum_{u=0}^{N-1} Z_{m,u}\xi_{m,u}c_u^{(1)} = \beta_0 a_m^{(1)} + \Omega_m^{(1)} + \Gamma_m^{(1)}, \quad (27)$$

where

$$\Omega_m^{(1)} \triangleq \frac{1}{N^2}\beta_0 \sum_{h\in\{\pm1\}} \Phi_h \sum_{k=1}^{N_u} a_m^{(k)} \sum_{l=0}^{N-1}\sum_{i=0}^{N-1} \gamma_{m,l}^{(k)} \cdot q_h[(i-l)T] \sum_{u=0}^{N-1} \xi_{m,u}c_u^{(1)} e^{-j2\pi\frac{ui}{N}}, \quad (28)$$

and

$$\Gamma_m^{(1)} \triangleq \frac{1}{N}\sum_{u=0}^{N-1} Z_{m,u,3}\xi_{m,u}c_u^{(1)}, \quad (29)$$

$Z_{m,u,3}$ being defined in (26).

Some remarks on (27)-(29) are now of interest.

1. The information symbol $a_m^{(1)}$ is multiplied by the coefficient $\beta_0 < 1$. This means that the user's constellation undergoes a warping effect according to which each received symbol moves away from its nominal point.
2. The sample $\Omega_m^{(1)}$ is due mainly to the contribution of the $N_u - 1$ interfering users (MAI). In the case of $N_u >> 1$, the central limit theorem enables us to model the MAI contribution as a zero-mean Gaussian random variable.
3. The overall effect of the NLDN and thermal noise is described by $\Gamma_m^{(1)}$, which again can be modeled as a zero-mean Gaussian random variable.
4. After straightforward yet tedious calculations, it can be shown that the variance of the MAI component (28) assumes the form

$$\sigma_\Omega^2 \triangleq E\left\{\left|\Omega_m^{(1)}\right|^2\right\} \quad (30)$$
$$= \frac{A_2}{N^4}\beta_0^2 \sum_{h_1,h_2\in\{\pm1\}} \Phi_{h_1}\Phi_{h_2}^* \sum_{k=1}^{N_u} \sum_{l_1,l_2=0}^{N-1} \sum_{i_1,i_2=0}^{N-1} \cdot q_{h_1}[(i_1-l_1)T]q_{h_2}[(i_2-l_2)T]\left[S_{l_1,l_2,i_1,i_2}^{(k)} - 2T_{l_1,l_2,i_1,i_2}^{(k)}\right],$$

where

$$S_{l_1,l_2,i_1,i_2}^{(k)} \triangleq \sum_{\lambda=0}^{N-1}\sum_{\mu=0}^{N-1}\left[c_\mu^{(k)}c_\mu^{(1)}c_\lambda^{(k)}c_\lambda^{(1)} e^{j2\pi\frac{\mu(2l_1+i_2-i_1)}{N}} \cdot e^{-j2\pi\frac{\lambda(2l_2+i_1-i_2)}{N}} + e^{j2\pi\frac{\mu(l_1-l_2)}{N}} e^{-j2\pi\frac{\lambda(i_1-i_2)}{N}}\right], \quad (31)$$

and

$$T_{l_1,l_2,i_1,i_2}^{(k)} \triangleq \sum_{\mu=0}^{N-1} e^{j2\pi\frac{\mu(l_1-l_2-i_1+i_2)}{N}}. \quad (32)$$

5. The variance of the overall noise component (29) is

$$\sigma_\Gamma^2 \triangleq E\left\{\left|\Gamma_m^{(1)}\right|^2\right\} = \sigma_\vartheta^2 + \frac{2\mathcal{N}_0}{N}, \quad (33)$$

where $\sigma_\vartheta^2 \triangleq E\{|\vartheta_{m,i}|^2\}/N$ is the variance of the samples of the NLDN at the matched filter output, defined in (17).

In order to reach a comprehensive statistical description of the decision variable (27), we are required now to evaluate σ_ϑ^2 that is included in (33). Let us assume for the time being that $\sigma_\Gamma^2 >> \sigma_\Omega^2$. This assumption will be verified in the next section when discussing the receiver performance results. Accordingly, the MAI component can be considered negligible as far as the approximation $\beta(t) \approx \beta_0$ holds true, and so, (8) turns out to be

$$x(t) \approx \beta_0 s(t) + \nu(t). \quad (34)$$

Using (34) and recalling that the NLDN $\nu(t)$ is uncorrelated with the HPA input signal $s(t)$, we get

$$E\left\{|\nu(t)|^2\right\} = E\left\{|x(t)|^2\right\} - \beta_0^2 E\left\{|s(t)|^2\right\}, \quad (35)$$

where $E\{|x(t)|^2\}$ can be computed as $E\{|M(\rho)|^2\}$. Hence, letting $\sigma_\vartheta^2 \approx E\{|\nu(mT_s+iT)|^2\}/N$, we would end up to an overestimation of the required variance, in that the power spectral density of the noise process $\nu(t)$ has a support wider than that of the matched filter due to sidelobe regrowth induced by the HPA. Indeed, a more accurate approach shows that the NLDN can be modelled as a white Gaussian process (at least within the signal bandwidth) having PSD $S_\vartheta(f) \approx S_\vartheta(0)$, as detailed in [9]. Hence, its variance can be accurately approximated by multiplying $S_\vartheta(0)$ with twice the MF's equivalent noise bandwidth.

4. BER PERFORMANCE EVALUATION

The results obtained in the following refer to a MC-CDMA system using 16-QAM as modulation format, root-raised cosine (RRC) pulse shaping with roll-off $\alpha = 0.125$, $N = 256$ subcarriers, a cyclic prefix of $L = 64$ samples. The sampling period is $T = 50$ ns. The channelization sequences belong to an N-size Walsh-Hadamard (WH) set yielding synchronous orthogonal spreading, and the scrambling code is a sequence of random binary-valued chips. The simulation scenario includes $N_u = 32$ active users, a solid state power amplifier (SSPA) at the transmitter, modelled according to eqn. (4) with $q = 10$, and an AWGN transmission channel.

Figure 3 shows the attenuation factor β_0 of the useful symbol vs. HPA's output back-off (OBO), obtained in the absence of AWGN, both analytically (solid line) according to (9) and by simulation (circular marks). Figure 4 presents the variance of the NLDN σ_ϑ^2 vs. OBO, in the absence of AWGN. The analytical curve (solid line) was derived by neglecting the MAI contribution, while marks represent simulation results. We remark that, in the interval of OBO of practical interest, a satisfactory agreement between analytical and simulation results is obtained both for the attenuation factor β_0 and variance of NLDN σ_ϑ^2.

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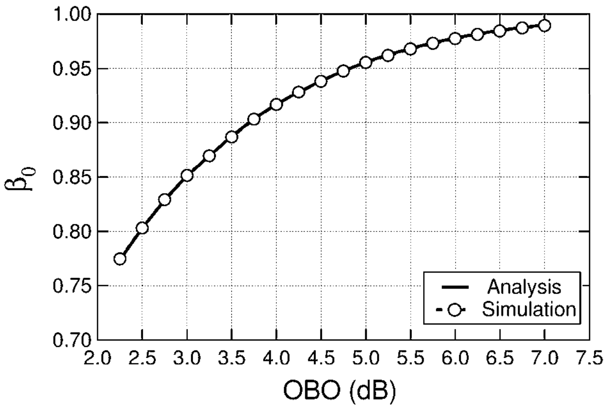

Figure 3: Average signal attenuation vs. OBO.

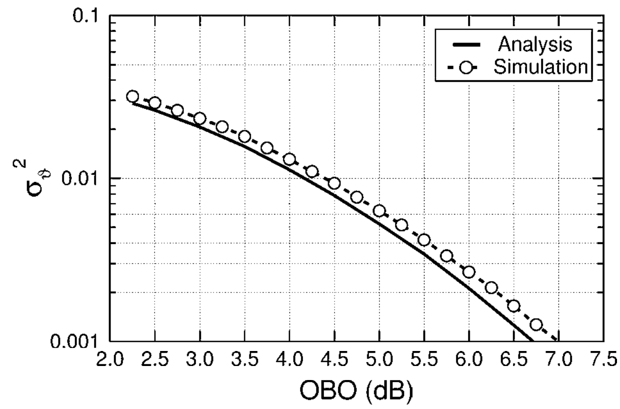

Figure 4: Nonlinear distortion noise variance vs. OBO.

These findings validate the hypothesis, made in the previous section, that the MAI contribution to the overall noise is negligible. The proposed framework can be then applied to derive a simple yet accurate analytical expression for the BER when a M-QAM modulation is employed. Focusing on the case of 16-QAM, we get

$$P_b = \frac{1}{4}\left\{\mathrm{Q}(\sqrt{\mu}) + \mathrm{Q}\left(\frac{2-\beta_0}{\beta_0}\sqrt{\mu}\right) + \mathrm{Q}\left(\frac{3\beta_0-2}{\beta_0}\sqrt{\mu}\right)\right\} \quad (36)$$

where

$$\mu \triangleq \frac{4}{5}\frac{E_b}{\mathcal{N}_0}\frac{1}{1+\frac{\sigma_\Gamma^2}{\beta_0^2\mathcal{N}_0}}, \quad (37)$$

is the equivalent signal-to-noise ratio, E_b is the average received energy per bit, and $\mathrm{Q}(x)$ is the Gauss integral function. The BER curves, both analytical (solid lines) and simulated (dashed lines), are plotted vs. $E_b/\mathcal{N}_0$ in Fig. 5, for several values of OBO. Again, we attain a satisfactory match between theoretical and simulation result.

5. CONCLUDING REMARKS

In this paper, we have proposed a detailed analytical framework to be employed for the bit-error performance evaluation of a MC-CDMA forward-link over nonlinear channels.

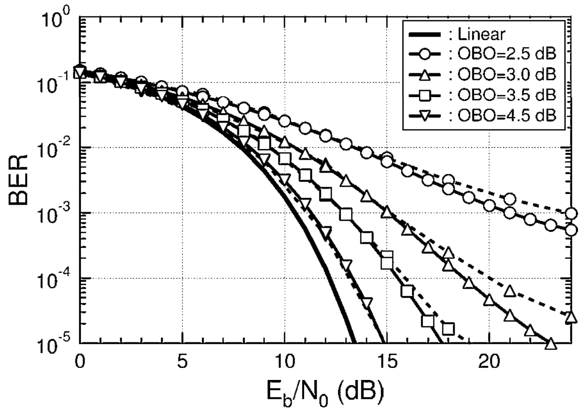

Figure 5: BER performance.

The key point is based on the proper application of the extension of the Bussgang theorem which enables an accurate statistical characterization of the decision variable used for data detection. The analytical approach is corroborated by simulation results obtained under typical multiple-access operating conditions.

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