BER Performance of OSTBCs on Correlated Fading Channels With Imperfect Channel Estimation

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Abstract:

In this contribution, we examine the bit error rate (BER) performance of orthogonal space-time block codes (OSTBCs) under flat-fading channels with imperfect channel state information. We present new exact BER expressions for PAM and QAM signal constellations, regardless of the fading distribution. Furthermore, we show how these expressions can be efficiently and accurately evaluated for the case of arbitrarily correlated Nakagami-m fading channels. As the high diversity order resulting from the application of OSTBCs gives rise to small BER values, the numerical evaluation of the presented BER expressions is much faster than straightforward computer simulations.

Keywords: Error analysis, OSTBC, channel estimation, correlated Nakagami fading

1. Introduction

Orthogonal space-time block codes (OSTBCs) are a remarkably beneficial transmit diversity technique, since they provide full diversity gain and result in a very simple maximum-likelihood (ML) detection algorithm, based only on linear processing at the receiver [1, 2]. When imperfect channel estimation (ICE) is taken into account, the performance of OSTBC systems is usually studied considering independent and identically distributed (i.i.d.) Rayleigh fading. In [3], a single-integral BER expression was given for Alamouti's code with pilot-symbol assisted modulation (PSAM). An exact closed-form BER expression for square OSTBCs was derived in [4] for pilot-based minimum mean-square error (MMSE) channel estimation and pulse amplitude modulation (PAM) or quadrature amplitude modulation (QAM) constellations. Using an approach based on characteristic functions, an exact closed-form expression for the pairwise error probability (PEP) of both orthogonal and non-orthogonal space-time codes with least-squares (LS) channel estimation was obtained in [5]. In this paper, we provide an exact analytical BER analysis for OSTBCs with PAM or QAM constellations and ICE on flat-fading channels with an arbitrary joint pdf. Also, we show how these exact BER expressions can be efficiently and accurately evaluated in the particular case of arbitrarily correlated Nakagami-m fading channels.

2. System model

We consider a multiple-input multiple-output (MIMO) wireless communication system with L_t transmit and L_r receive antennas. The transmitted data symbols at each transmit antenna are assumed to be coded according to an $L_t \times K_c$ (K_c denotes the block length) OSTBC matrix $\mathbf{C}$, the entries of which are linear combinations of N_s information symbols $s_i = s_{i,R} + js_{i,I}$, $1 \leq i \leq N_s$, and their complex conjugate s_i^*

$$\mathbf{C} = \sum_{i=1}^{N_s} \left(\mathbf{C}_i s_i + \mathbf{C}_i' s_i^* \right). \tag{1}$$

In (1), the $L_t \times K_c$ matrices $\mathbf{C}_i$ and $\mathbf{C}_i'$ comprise the coefficients of s_i and s_i^*, respectively. Since scaling of $\mathbf{C}$ does not affect its orthogonality, we assume that $\mathbf{C}$ is scaled in such way that

$$\mathbf{C}\mathbf{C}^H = \left(\lambda \sum_{i=1}^{N_s} |s_i|^2 \right) \mathbf{I}_{L_t}, \tag{2}$$

where $\lambda \triangleq K_c/N_s$ and $\mathbf{I}_{L_t}$ is the $L_t \times L_t$ identity matrix. For square OSTBCs, i.e., $L_t = K_c$, it is readily verified that $\mathbf{C}^H\mathbf{C} = \mathbf{C}\mathbf{C}^H$.

We organize the data transmission in frames consisting of K_p known pilot symbols used for channel estimation and K/K_c coded symbol matrices $\mathbf{C}(k)$, with K being a multiple of K_c and k denoting the block index. With $L \triangleq L_t L_r$ denoting the diversity order, the L complex channel coefficients are distributed according to an arbitrary joint pdf that characterizes the fading. The $L_r \times L_t$ complex random matrix $\mathbf{H}$, containing all channel coefficients, is assumed to remain constant during the length of one frame of $K + K_p$ symbols, such that the receiver separately observes the $L_r \times K_c$ matrices

$$\mathbf{R}(k) = \sqrt{E_s}\mathbf{H}\mathbf{C}(k) + \mathbf{W}(k), \tag{3}$$

with $1 \leq k \leq K/K_c$, and the $L_r \times K_p$ matrix

$$\mathbf{R}_p = \sqrt{E_p}\mathbf{H}\mathbf{C}_p + \mathbf{W}_p, \tag{4}$$

where the noise matrices $\mathbf{W}(k)$ and $\mathbf{W}_p$ contain i.i.d. zero-mean (ZM) circularly symmetric complex Gaussian (CSCG) random variables (RVs) with variance N_0. In the remainder of this paper, we will drop the block index k for notational simplicity.

In order to estimate the channel, we assume orthogonal pilot sequences, i.e., $\mathbf{C}_p\mathbf{C}_p^H = K_p\mathbf{I}_{L_t}$, and LS channel estimation [6]

$$\hat{\mathbf{H}} = \frac{1}{K_p\sqrt{E_p}}\mathbf{R}_p\mathbf{C}_p^H, \tag{5}$$

such that the estimated channel $\hat{\mathbf{H}}$ can be decomposed into the sum of two statistically independent contributions, i.e., $\hat{\mathbf{H}} = \mathbf{H} + \mathbf{N}$. Note that the $L_r \times L_t$ estimation noise matrix $\mathbf{N} \triangleq \left[1/(K_p\sqrt{E_p}) \right] \mathbf{W}_p\mathbf{C}_p^H$ consists of ZM CSCG RVs, the real and imaginary parts of which have variance $\sigma_{\mathbf{N}}^2 = N_0/(2K_pE_p)$. Hence, when conditioned on $\mathbf{H}$, the estimated channel coefficients $\hat{\mathbf{H}}_{\ell,k}$ ($1 \leq \ell \leq L_r, 1 \leq k \leq L_t$) are CSCG RVs with mean $\mathbf{H}_{\ell,k}$ and variance $2\sigma_{\mathbf{N}}^2$. Increasing the total energy K_pE_p allocated to pilot symbols

improves the channel estimate, but also reduces the symbol energy E_s available for data transmission. Indeed, with E_b, $\gamma = E_p/E_s$, M, and $\rho = N_s/(L_t K_c)$ denoting the energy per information bit, the ratio of the pilot energy E_p to the symbol energy E_s, the number of constellation points and the code rate, respectively, we have $E_s = \frac{K}{K+\gamma K_p}\rho \log_2(M)E_b$, which is a decreasing function of K_p.

We consider a mismatched ML receiver that uses the estimated channel $\hat{\mathbf{H}}$ in the same way as an ML receiver would apply $\mathbf{H}$. In this way, the detection algorithm for the information symbols s_i can be shown to reduce to symbol-by-symbol detection

$$\hat{s}_i = \arg\min_{\tilde{s}} |u_i - \tilde{s}|, \quad 1 \le i \le N_s, \tag{6}$$

where the minimization is over the symbols $\tilde{s}$ belonging to the considered signaling constellation, and the decision variables $u_i = u_{i,R} + j u_{i,I}$ are given by

$$u_i = \frac{\mathrm{tr}\left(\mathbf{C}_i^H \hat{\mathbf{H}}^H \mathbf{R} + \mathbf{R}^H \hat{\mathbf{H}} \mathbf{C}_i'\right)}{\lambda \sqrt{E_s} \left\|\hat{\mathbf{H}}\right\|_F^2}, \tag{7}$$

with $\mathrm{tr}(\cdot)$ and $\|\cdot\|_F$ denoting the trace and the Frobenius norm of a matrix, respectively.

3. BER analysis

For square M-QAM transmission with Gray mapping, the BER can be obtained by averaging the BERs related to the in-phase and quadrature-phase bits of all information symbols s_i in the code matrix

$$\mathrm{BER} = \frac{1}{2N_s}\sum_{i=1}^{N_s} \mathbb{E}\left[\mathrm{BER}_{i,R}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}}) + \mathrm{BER}_{i,I}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}})\right], \tag{8}$$

where $\mathrm{BER}_{i,R}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}})$ and $\mathrm{BER}_{i,I}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}})$ denote the conditional BERs related to the in-phase and quadrature-phase information bits of s_i, respectively, conditioned on $\mathbf{s}$, $\mathbf{H}$, and $\hat{\mathbf{H}}$. Denoting by Ψ the M-QAM constellation, $\mathbf{s}$ is uniformly distributed over Ψ^{N_s}. Furthermore, Ψ_R and Ψ_I are the sets containing the real and imaginary parts of the constellation points, respectively. Referring to the projections of the decision area of a QAM symbol $b = b_R + j b_I$ on the real and imaginary axis as the decision regions of b_R and b_I, respectively, we can write $\mathrm{BER}_{i,q}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}})$, with q = R or q = I, as follows

$$\mathrm{BER}_{i,q}(\mathbf{s},\mathbf{H},\hat{\mathbf{H}}) = 2\sum_{b_q \in \Psi_q} \frac{N(s_{i,q}, b_q)}{\log_2 M} P_{i,q}(\mathbf{s}, b_q, \mathbf{H}, \hat{\mathbf{H}}), \tag{9}$$

where $N(s_{i,q}, b_q)$ represents the Hamming distance between the bits allocated to $s_{i,q}$ and b_q, and $P_{i,q}(\mathbf{s}, b_q, \mathbf{H}, \hat{\mathbf{H}})$ is the probability that $u_{i,q}$ is located inside the decision area of b_q, when $\mathbf{s}$, $\mathbf{H}$, and $\hat{\mathbf{H}}$ are known. Taking (1) and (3) into account, expanding the decision variable (7) yields $u_i = u_i' + n_i$, where u_i' is a function of $\mathbf{s}$, $\mathbf{H}$, and $\hat{\mathbf{H}}$, and $n_{i,R}$ represents ZM CSCG noise with variance $N_0/(\lambda E_s \|\hat{\mathbf{H}}\|_F^2)$. When $d_1(b_q)$ and $d_2(b_q)$ denote the boundaries of the decision area of b_q, with $d_1(b_q) < d_2(b_q)$, $P_{i,q}(\mathbf{s}, b_q, \mathbf{H}, \hat{\mathbf{H}})$

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is given by

$$P_{i,q}(\mathbf{s}, b_q, \mathbf{H}, \hat{\mathbf{H}}) = Q\left\{\sqrt{2\lambda\frac{E_s}{N_0}}\|\hat{\mathbf{H}}\|_F \left[d_1(b_q) - u'_{i,q}\right]\right\}$$

$$- Q\left\{\sqrt{2\lambda\frac{E_s}{N_0}}\|\hat{\mathbf{H}}\|_F \left[d_2(b_q) - u'_{i,q}\right]\right\}, \quad (10)$$

where $Q(.)$ denotes the Gaussian Q-function [7, Eq. (4.1)]. Note from (8)-(10) that the evaluation of the BER requires averaging over $4L$ real-valued continuous RVs, i.e., the real and imaginary parts of the elements of $\mathbf{H}$ and $\hat{\mathbf{H}}$, and over N_s discrete RVs, i.e., the N_s symbols in $\mathbf{s}$. By using a carefully selected coordinate transformation, however, we will decrease the number of RVs involved in the expectation (8) and, accordingly, the computational complexity related to its numerical evaluation. Let us introduce the $2L$-dimensional real-valued column vectors $\hat{\mathbf{h}}$ and $\mathbf{h}$, consisting of the real and imaginary parts of all elements of $\hat{\mathbf{H}}$ and $\mathbf{H}$, respectively:

$$\begin{cases} \hat{\mathbf{h}} = \left[\hat{\mathbf{h}}_{1,R}^T, \hat{\mathbf{h}}_{1,I}^T, \cdots, \hat{\mathbf{h}}_{N_s,R}^T, \hat{\mathbf{h}}_{N_s,I}^T\right]^T \\ \mathbf{h} = \left[\mathbf{h}_{1,R}^T, \mathbf{h}_{1,I}^T, \cdots, \mathbf{h}_{N_s,R}^T, \mathbf{h}_{N_s,I}^T\right]^T \end{cases}, \quad (11)$$

with $\hat{\mathbf{h}}_{i,R} + j\hat{\mathbf{h}}_{i,I}$ and $\mathbf{h}_{i,R} + j\mathbf{h}_{i,I}$ denoting the i-th column of $\hat{\mathbf{H}}$ and $\mathbf{H}$, respectively. It follows from (11) that $|\hat{\mathbf{h}}| = \|\hat{\mathbf{H}}\|_F$ and $|\mathbf{h}| = \|\mathbf{H}\|_F$, with $|\cdot|$ denoting the norm of a vector. Using (11), it can be shown that $u'_{i,q}$ in (10) can be written as

$$u'_{i,q} = \frac{1}{\lambda(x_1^2 + x_2^2 + z^2)} \left(\lambda x_1 s_{i,q}|\mathbf{h}| + x_2|\mathbf{M}_{i,q}\mathbf{h}|\right), \quad (12)$$

where the $2L \times 2L$ matrix $\mathbf{M}_{i,q}$ is a function of the transmitted symbol vector $\mathbf{s}$ and the coefficient matrices $\mathbf{C}_j$ and $\mathbf{C}'_j$, $1 \leq j \leq N_s$. Because of space limitations, $\mathbf{M}_{i,q}$ is not given in explicit form. Also, it can be shown that x_1, x_2, and z, when conditioned on $\mathbf{h}$, are independent RVs satisfying the following properties: x_1 and x_2 are Gaussian RVs with mean $|\mathbf{h}|$ and zero, respectively, and variance σ_N^2; z/σ_n follows the chi-distribution with $2L - 2$ degrees of freedom. Hence, considering that $|\hat{\mathbf{h}}|^2 = x_1^2 + x_2^2 + z^2$, it is easily seen that the conditional BERs given by (9) can be rewritten as functions that depend on $\mathbf{H}$ through the random vector $\mathbf{h}$, and on $\hat{\mathbf{H}}$ through only 3 RVs: x_1, x_2, and z; we denote these functions by $B_{i,q,1}(\mathbf{s}, \mathbf{h}, x_1, x_2, z)$. Note that the dependence on $\mathbf{h}$ is only through $|\mathbf{h}|$ and $|\mathbf{M}_{i,q}\mathbf{h}|$, with $\mathbf{M}_{i,q}$ depending on $\mathbf{s}$. Because of this substitution, the BER reduces to an expectation over $2L + 3$ continuous RVs, i.e., the $2L$ components of $\mathbf{h}$, x_1, x_2, and z, and N_s discrete RVs, i.e., the components of $\mathbf{s}$:

$$\text{BER} = \frac{1}{2N_s} \sum_{i=1}^{N_s} \mathbb{E}\left[B_{i,R,1}(\mathbf{s}, \mathbf{h}, x_1, x_2, z) + B_{i,I,1}(\mathbf{s}, \mathbf{h}, x_1, x_2, z)\right]. \quad (13)$$

In the case of square OSTBCs, the BER expression (13) can be considerably simplified. For these OSTBCs, it can be shown that the magnitude of $\mathbf{M}_{i,q}\mathbf{h}$ used in (12) is given by $|\mathbf{M}_{i,q}\mathbf{h}| = \lambda|\mathbf{h}|\sqrt{|\mathbf{s}|^2 - s_{i,q}^2}$. Hence, $B_{i,q,1}(\mathbf{s}, \mathbf{h}, x_1, x_2, z)$ in (13), with q = R or q = I can be rewritten as a function that depends on $\mathbf{h}$ through only the norm $|\mathbf{h}|$

of the channel vector; we denote this function by $B_{i,q,2}(\mathbf{s}, |\mathbf{h}|, x_1, x_2, z)$. Moreover, since the statistical properties of $\mathbf{s}_{i,q}$ are independent of i and q, it is easily seen that the BER for square OSTBCs reduces to

$$\text{BER} = \mathbb{E}[B_{i,q,2}(\mathbf{s}, |\mathbf{h}|, x_1, x_2, z)], \tag{14}$$

which involves the expectation over only 4 continuous RVs, i.e, the norm $|\mathbf{h}|$ of the channel vector $\mathbf{h}$, x_1, x_2 and z, and N_s discrete RVs, i.e., the components of $\mathbf{s}$.

By numerically evaluating (13) and (14), the BER for OSTBCs can be efficiently and accurately obtained. In order to minimize the associated computation time, a proper combination of numerical integration techniques, such as the *quadrature rule* [8, Sec. 4.1], in which case a *J*-fold integral is approximated by a *J*-fold sum, and *Monte-Carlo integration* [8, Sec. 7.7] with *importance sampling* [8, Sec. 7.9.1] can be used. For instance, the expectation over x_1, x_2, and z can be evaluated by means of the quadrature rule, while the expectation over $\mathbf{s}$ and $\mathbf{h}$ can be evaluated through Monte-Carlo integration.

4. Numerical results

In this section, BER results are presented for correlated Nakagami-m fading channels with uniformly distributed phases, assuming that $E_p = E_s$. The Nakagami-m distribution is considered as a versatile statistical distribution that accurately models a variety of fading environments [7, 9–11]. Denoting by α the magnitude of a complex-valued channel coefficient, its pdf in the case of Nakagami-m fading is given by [7, Eq. (2.20)]

$$p_\alpha(r) = \frac{2}{\Gamma(m)}\left(\frac{m}{\Omega}\right)^m r^{(2m-1)} \exp\left(-\frac{m}{\Omega}r^2\right), \qquad r \geq 0, \tag{15}$$

with $\Gamma(\cdot)$ being the Gamma function [12, Eq. (8.310/1)], $\Omega = \mathbb{E}[\alpha^2]$ being the average fading power, and $m \geq 1/2$ being the fading parameter. In case of a correlated Nakagami fading MIMO channel, the marginal pdf of the magnitude $\alpha_{i,j}$ of the channel coefficient $\mathbf{H}_{i,j}$ follows (15), with parameters $m_{i,j}$ and $\Omega_{i,j}$ possibly depending on i and j. Stacking the $L = L_t L_r$ fading magnitudes $\alpha_{i,j}$ into an L-dimensional column vector $\boldsymbol{\alpha}$, such that $\boldsymbol{\alpha}_{i+(j-1)L_r} = \alpha_{i,j}$, the elements of the $L \times L$ power correlation matrix $\boldsymbol{\Sigma}$ are defined as [7, Eq. (9.195)]. Also, we use the MIMO channel model proposed in [13], such that $\boldsymbol{\Sigma} = \boldsymbol{\Sigma}_t \otimes \boldsymbol{\Sigma}_r$, where $\boldsymbol{\Sigma}_t$ and $\boldsymbol{\Sigma}_r$ are the $L_t \times L_t$ transmit and $L_r \times L_r$ receive power correlation matrices, respectively, and $\otimes$ denotes the Kronecker product.

Let us consider Alamouti's code ($L_t = K_c = N_s = 2$), which is given by [1, Eq. (32)]. Fig. 1 displays the BER curves for Alamouti's code along with 4-QAM signaling, operating over correlated identically distributed (i.d.) Nakagami-m channels with $\Omega = 1$ and $L_r = 3$. In order to obtain the BER, we evaluate the expectation over $|\mathbf{h}|$, x_1, x_2, and z in (14) by means of the quadrature rule, while the expectation over $\mathbf{s}$ is exactly obtained by a finite summation over all possible symbol vectors. The pdf of $|\mathbf{h}|$ can be easily obtained from [9] in the case of identical and integer m. The BER results are shown for both a receiver with perfect channel knowledge (PCK) and a mismatched receiver using (5) with $K = 200$, and $K_p = 20$. Further, we assume that $m = \{1, 3\}$, that no antenna correlation occurs at the transmitter side, and that the receive antennas are either uncorrelated (unc) or constructed as a linear array (lin), the configuration of which is depicted in [10, Fig. 4(b)], with power correlation matrix $\boldsymbol{\Sigma}_r$ given by [10, Eq. (38)]. From Fig. 1, we can see how the fading parameter m, ICE,
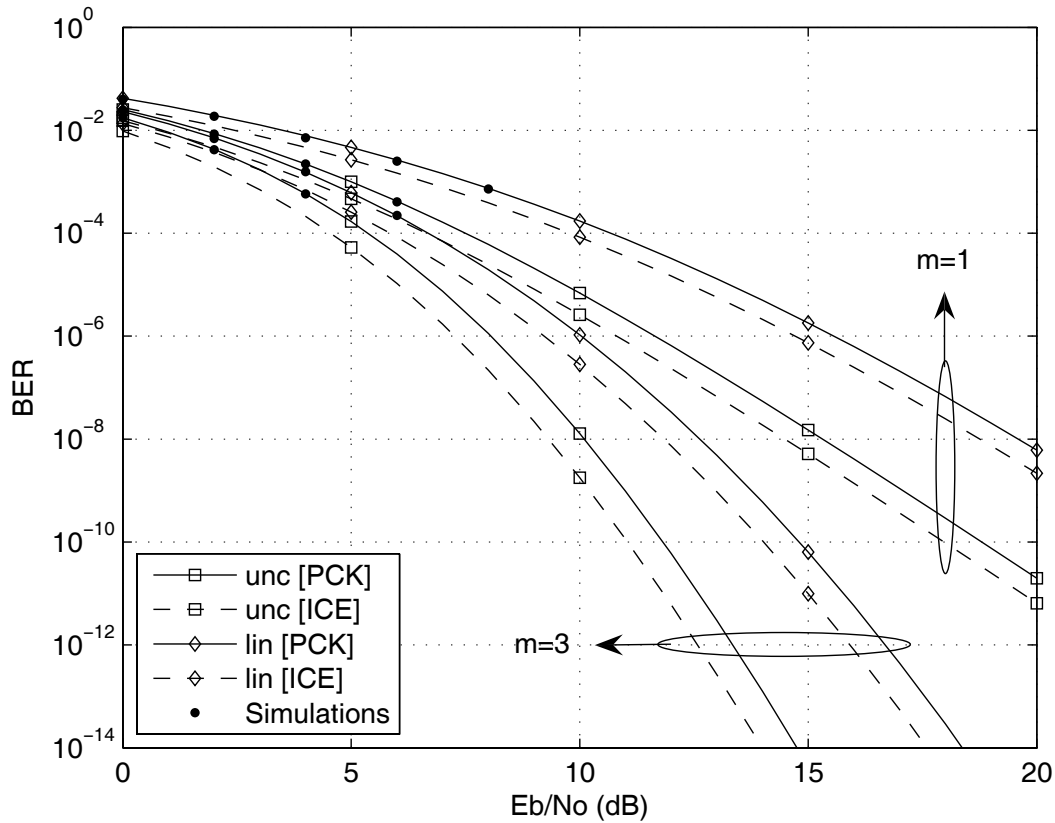
Figure 1: BER for Alamouti's code with 4-QAM signaling operating over correlated i.d. Nakagami-m channels, with $m = \{1, 3\}$ and $\Omega = 1$. Computer simulations confirm the analytical results obtained from numerically evaluating (14).

and antenna correlation affect the BER of maximal-ratio diversity (MRD) reception. A larger m indicates less severe fading, such that the BER performance is improved. Both imperfect channel estimation and antenna correlation degrade the BER through a horizontal shift of the BER curve for large E_b/N_0. Interestingly, the BER degradation due to antenna correlation is much larger than the degradation due to ICE.

Fig. 2 shows the BER versus the number of pilot symbols K_p for the 3×4 OSTBC given by [1, Eq. (39)], operating over correlated i.d. Nakagami-m fading channels with $m = 2$ and $\Omega = 1$, under the assumption that $E_b/N_0 = 10$ dB. To obtain the BER, the expectation over $\mathbf{s}$, $\mathbf{h}$, x_1, x_2, and z in (13) has been evaluated by means of Monte-Carlo integration with importance sampling. Clearly, the expectation over $\mathbf{h}$ requires the generation of L correlated Nakagami-m RVs; in [11], it is shown how these RVs can be obtained from $2m$ i.i.d. L-dimensional ZM Gaussian random vectors, in case of identical and integer m. The power correlation matrix $\mathbf{\Sigma}_r$ of the mismatched dual-antenna receiver ($L_r = 2$) is given by

$$\mathbf{\Sigma}_r = \begin{pmatrix} 1 & 0.2 \\ 0.2 & 1 \end{pmatrix}, \tag{16}$$

and the transmitted information symbols belong to a 64-QAM constellation. The results are shown for $K = \{48, 200, 800\}$ information symbols and for the power correlation
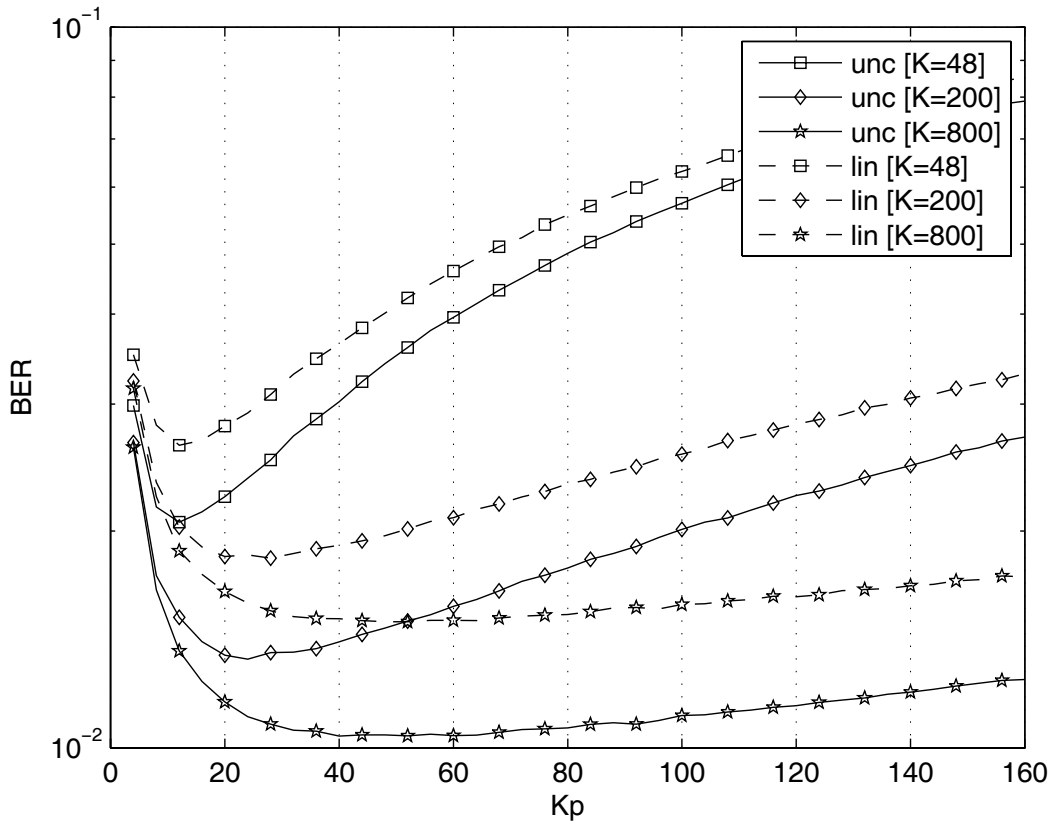
Figure 2: BER versus K_p for the 3×4 OSTBC given by [1, Eq. (39)] with 64-QAM.

matrix $\mathbf{\Sigma}_\mathrm{t}$ at the transmitter side being given by either [10, Eq. (38)] (lin) or the identity matrix (unc). From Fig. 2, we notice that the optimal number of pilot symbols grows with the number of information symbols K, and that antenna correlation does not affect this optimal number. For large K, obtaining the optimal number of pilot symbols is not very critical, as the BER grows only slowly when more pilot symbols are added.

5.   Conclusions and remarks

In this contribution, we investigated the effect of ICE on the BER performance of OS-TBCs operating over correlated flat-fading channels. For both non-square and square OSTBCs, we proposed new exact BER expressions which can be efficiently and accurately evaluated by a combination of numerical integration techniques. Our numerical results illustrate how the proposed analysis can be applied to investigate the effect of channel estimation and fading correlation on the BER performance of OSTBC systems in the case of Nakagami-m fading.

Acknowledgments

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