UNDERSTANDING THE BER-PERFORMANCE OF SPACE-TIME BLOCK CODES

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ABSTRACT

In this paper we show that in the case of general Space-Time Block Codes (STBC) not only the Pairwise Error Probabilities (PEPs) between nearest neighboring code words have to be considered to evaluate the total Bit Error Ratio (BER) performance of the STBC. It turns out that code word pairs differing in a rather large large number of positions typically lead to rank deficient Distance Matrices (DMs) causing a severe diversity loss and thus deteriorate the total BER - performance in Rayleigh fading wireless Multiple Input-Multiple Output (MIMO) channels especially at high Signal to Noise Ratio (SNR) levels. In the special case of orthogonal codes from the Generalized Complex Orthogonal Design (GCOD), nearest neighbor code word errors dominate the BER-curve and no diversity loss at high SNR occurs.

1. INTRODUCTION

Recent research in the field of wireless communications heavily focuses on MIMO systems. Such systems promise a huge increase in capacity in case of wireless channels with rich scattering. To exploit this increase in capacity, currently several transmission schemes are under investigation. Space-Time Coding is one method to take advantage of this potential. Several design criteria have already been published to optimize the design of Space-Time Codes (STCs), see for example [1][2].

One of the most famous design criteria are those derived by Tarokh et al [1]. In [1] an upper bound on the PEP of STCs is derived. This upper bound leads to powerful code design criteria and thus to STBCs with good error performance. Evaluating the BER of a code, mostly nearest neighbor error events are considered. For example, if we have a STBC-word as shown in Eqn. (1):

$$\boldsymbol{X} = \begin{pmatrix} x_1 & x_2 & x_3 & x_4 \\ x_4 & x_1 & x_2 & x_3 \\ x_3 & x_4 & x_1 & x_2 \\ x_2 & x_3 & x_4 & x_1 \end{pmatrix}, \qquad (1)$$

which encodes four information symbols (x_1 to x_4) to be transmitted over four transmit antennas within four time slots, then we usually consider only PEPs between code words differing only in a single information symbol. Assuming for example that in Eqn. (1) the code word $\boldsymbol{X}$ is erroneously detected as $\hat{\boldsymbol{X}}$:

$$\hat{\boldsymbol{X}} = \begin{pmatrix} \hat{x}_1 & \hat{x}_2 & \hat{x}_3 & \hat{x}_4 \\ \hat{x}_4 & \hat{x}_1 & \hat{x}_2 & \hat{x}_3 \\ \hat{x}_3 & \hat{x}_4 & \hat{x}_1 & \hat{x}_2 \\ \hat{x}_2 & \hat{x}_3 & \hat{x}_4 & \hat{x}_1 \end{pmatrix}. \qquad (2)$$

If only one of the information symbols is erroneously decoded, for example $\hat{x}_1 \neq x_1$, but $\hat{x}_2 = x_2$, $\hat{x}_3 = x_3$ and $\hat{x}_4 = x_4$, then we speak of a single symbol error. We refer to such code word pairs $\boldsymbol{X}$ and $\hat{\boldsymbol{X}}$ as nearest neighbor code words. If however more than one information symbol differs, then we call this event a multiple symbol error. In this contribution we discuss the following question: *Is the BER dominated by code word errors corresponding to single or to multiple information symbol errors?*

The rest of the paper is organized as follows. In Section 2 we introduce general STBCs and investigate their BER for single and multiple error events. Section 3 is devoted to orthogonal STBCs. The paper closes with some conclusions.

2. NON ORTHOGONAL SPACE TIME BLOCK CODES

2.1. Introduction

In this section we focus on general STBCs and we will answer the question posed in Section 1. In fact, we do not provide a general solution for this problem, but we show interesting results by an illustrative example, which is intended to be generalized in future work. As an example, we consider a cyclic STBC [3] for a MIMO system with four transmit ($n_T = 4$) and four receive antennas ($n_R = 4$) consisting of code words, which are shown in Eqn. (1). The information symbols x_1 to x_4 are assumed to be BPSK modulated symbols taken from the set $\{-1, +1\}$. Then the number of valid code word pairs is 256 and thus not too large to visualize the essential aspects and results. We also want to emphasize that all considerations here are based on the ML-receiver principle. We assume slow, flat Rayleigh fading (block fading) MIMO channels in correspondence to the assumptions in [1]. The entries of the channel matrix $\boldsymbol{H}$ are assumed to be i.i.d. complex Gaussian distributed with zero mean and unit variance. Then Tarokhs well known code design criteria can be stated as follows:

1. The rank r of the code Distance Matrices (DMs) $\boldsymbol{A}_{i,j}$ (defined further ahead in Eqn. (5)) has to be maximized for all code word pairs (i, j):
$$max\left\{rank(\boldsymbol{A}_{i,j})\right\} \ . \tag{3}$$
2. Secondly, the minimum determinant of all code DMs $\boldsymbol{A}_{i,j}$ should be maximized:
$$max\left\{det(\boldsymbol{A}_{i,j})\right\} \ . \tag{4}$$

The code DM $\boldsymbol{A}_{i,j}$ is defined as:

$$\boldsymbol{A}_{i,j} = \boldsymbol{B}_{i,j}^H \boldsymbol{B}_{i,j} = (\boldsymbol{X}_i - \boldsymbol{X}_j)^H (\boldsymbol{X}_i - \boldsymbol{X}_j) \ , \tag{5}$$

where $\boldsymbol{X}_i$ and $\boldsymbol{X}_j$ are distinct code words and $\boldsymbol{B}_{i,j} = (\boldsymbol{X}_i - \boldsymbol{X}_j)$. With the eigenvalues $\lambda_1 \cdots \lambda_r$ of the code DM $\boldsymbol{A}_{i,j}$ an upper bound for the PEP can be calculated:

$$P\left(\boldsymbol{X}_i \to \boldsymbol{X}_j\right) \leq \prod_{k=1}^{r} \left(1 + \lambda_k \frac{\text{SNR}}{4\, n_T}\right)^{-n_R} \ , \tag{6}$$

where the mean SNR is defined as the total received power per antenna divided by the noise power at each antenna σ_n^2, resulting in

$$\text{SNR} = \frac{n_T\, P_s}{\sigma_n^2} \ . \tag{7}$$

The SNR definition implies unit variance channel coefficients. In case of BPSK modulation $P_s = 1$.

2.2. Pairwise Code Word Errors and their Influence on the Upper Bound of the BER

Evaluation of the BER mainly focuses on nearest neighbor code word errors. In contrast, here we take into account all error events. Because of the BPSK modulation and the four information symbols per STBC word, there are only $2^4 = 16$ possible STBC words to be considered, i.e., there exist 16 distinct STBC word matrices $\boldsymbol{X}_i$ and for each code matrix 15 possibilities for decoding erroneous code matrices $\boldsymbol{X}_j$. Therefore, there are $16 \cdot 16 = 256$ combinations of code word pairs $(\boldsymbol{X}_i, \boldsymbol{X}_j)$. For all of these 256 combinations of $(\boldsymbol{X}_i, \boldsymbol{X}_j)$ we can calculate the code word DMs $\boldsymbol{A}_{i,j}$ and their eigenvalues λ_k. For the cyclic STBC given in Eqn. (1) we obtain the following interesting result: There are only eight types of DM, $\boldsymbol{A}^{(i)}$, i=1 .. 8, with eight different eigenvalue distributions. The eigenvalues of these DMs are listed in Table 1. With the knowledge of the DMs and their eigenvalues, we can calculate the upper bound of the BER as a function of the mean SNR. The mean Symbol Error Probability (SEP) results from the various PEPs (PEP1 to PEP8) which can be classified according to the eight different types of DMs listed in Table 1:

$$\begin{aligned} \text{SEP} &= \sum_{j=1}^{16} \sum_{i=1, i\neq j}^{16} P\left(\boldsymbol{X}_i \to \boldsymbol{X}_j\right) P\left(\boldsymbol{X}_i\right) \\ &= \frac{1}{16} \left[64\,\text{PEP1} + 64\,\text{PEP2} + 32\,\text{PEP3} + 32\,\text{PEP4} \right. \\ &\quad \left. + 32\,\text{PEP5} + 8\,\text{PEP6} + 4\,\text{PEP7} + 4\,\text{PEP8}\right] , \end{aligned} \tag{8}$$

type of DM	λ_1	λ_2	λ_3	λ_4	No of code pairs	No of different symbols
	0	0	0	0	16	0
1	4	4	4	4	64	1
2	16	8	8	0	64	2
3	16	16	0	0	32	2
4	36	4	4	4	32	3
5	20	20	4	4	32	3
6	16	16	16	16	8	4
7	32	32	0	0	4	4
8	64	0	0	0	4	4

Table 1. Eigenvalue distributions of all types of DMs $\boldsymbol{A}^{(i)}$ and the corresponding No of different symbols.

where PEPk is the PEP corresponding to the DM type k. Note, that we assume that the 16 possible code words $\boldsymbol{X}_i$ are equally likely and therefore $P_i\left(\boldsymbol{X}\right) = \frac{1}{16}$. The double sum over i and j decomposes into the sum over eight different PEPs weighted with the number of code pairs (see Table 1).

The eight different PEPs are characterized by a certain number of different symbols and a distinct set of eigenvalues λ_i. Every symbol error corresponds to a one bit error. The PEP1 corresponds to one bit error, PEP2 to two bit errors, up to PEP8 corresponding to 4 bit errors according to the terms in Table 1. Therefore the overall BER can be calculated as:

$$\begin{aligned} \text{BER} = \frac{1}{16} &\left[64\frac{1}{4}\text{PEP1} + 64\frac{2}{4}\text{PEP2} + 32\frac{2}{4}\text{PEP3} + 32\frac{3}{4}\text{PEP4} \right. \\ &\left. + 32\frac{3}{4}\text{PEP5} + 8\frac{4}{4}\text{PEP6} + 4\frac{4}{4}\text{PEP7} + 4\frac{4}{4}\text{PEP8}\right] . \end{aligned} \tag{9}$$

Substituting the upper bound (Eqn. (6)) for the PEPs as a function of the corresponding eigenvalues and SNR values we get the final result for the upper bound of the BER:

$$\begin{aligned} &\text{BER} \leq \\ &\prod_{i=1}^{r_1} \left(1 + \lambda_i^{(1)} \frac{\text{SNR}}{16}\right)^{-n_R} + 2 \prod_{i=1}^{r_2} \left(1 + \lambda_i^{(2)} \frac{\text{SNR}}{16}\right)^{-n_R} + \\ &\prod_{i=1}^{r_3} \left(1 + \lambda_i^{(3)} \frac{\text{SNR}}{16}\right)^{-n_R} + \frac{3}{2} \prod_{i=1}^{r_4} \left(1 + \lambda_i^{(4)} \frac{\text{SNR}}{16}\right)^{-n_R} + \\ &\frac{3}{2} \prod_{i=1}^{r_5} \left(1 + \lambda_i^{(5)} \frac{\text{SNR}}{16}\right)^{-n_R} + \frac{1}{2} \prod_{i=1}^{r_6} \left(1 + \lambda_i^{(6)} \frac{\text{SNR}}{16}\right)^{-n_R} + \\ &\frac{1}{4} \prod_{i=1}^{r_7} \left(1 + \lambda_i^{(7)} \frac{\text{SNR}}{16}\right)^{-n_R} + \frac{1}{4} \prod_{i=1}^{r_8} \left(1 + \lambda_i^{(8)} \frac{\text{SNR}}{16}\right)^{-n_R} . \end{aligned} \tag{10}$$

Here, the values $\lambda_i^{(k)}$ are the four eigenvalues (i=1,2,3,4) of the code word DM $\boldsymbol{A}^{(k)}$ for the k-th type of pairwise code word error. Their values are given in the Table 1. r_l denotes the rank of the DM $\boldsymbol{A}^{(k)}$. With Eqn. (10), we can draw the eight upper bound contributions for the BER for every type of error event and the total upper bound for the BER as a function of the mean SNR. As we can see from Fig. 1,

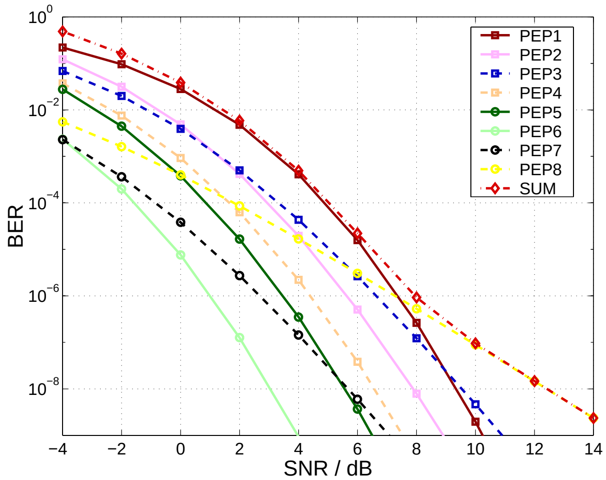

Fig. 1. The upper bound for the BER vs. SNR curves for the eight DMs and the sum BER; 4 receive antennas

PEP1 (solid brown curve, square marker) and PEP8 (dashed yellow curve, circle marker) dominate the BER-sum-curve (dashed-dotted red curve, diamond marker). At low values of SNR the BER-sum-curve is dominated by PEP1, which corresponds to single symbol errors. The BER-sum-curve at high SNR is dominated by PEP8, which corresponds to four symbol errors. The corresponding code DM $\boldsymbol{A}^{(8)}$ is rank deficient and has rank $r_8 = 1$ only, but with a high value of the only non zero eigenvalue $\lambda_1^{(8)}$=64.

Summarizing these observations we conclude: in general, for a STB coded system we have to take into account all possible PEPs and not only single symbol errors, as it is usually done in the case of Single Input - Single Output (SISO) systems. Different PEPs dominate the code performance at low values of SNR and at high values of SNR.

2.3. The Distance Profile of Received Signals

Extremely important parameters of any code are the squared Euclidean distances d^2 between any two different signal matrices obtained at the receiver output corresponding to distinct code words transmitted by the n_T transmit antennas:

$$\begin{aligned} d^2 &= \|\boldsymbol{H}\,\boldsymbol{B}\|^2 = \sum_{i=1}^{n_R} \boldsymbol{h}_i \underbrace{\boldsymbol{B}\boldsymbol{B}^H}_{\boldsymbol{U}\boldsymbol{D}\boldsymbol{U}^H} \boldsymbol{h}_i{}^H = \sum_{i=1}^{n_R} \boldsymbol{h}_i'\,\boldsymbol{D}\,\boldsymbol{h}_i'{}^H \\ &= \sum_{i=1}^{n_R}\sum_{j=1}^{n_T} \lambda_j \left|h_{ij}'\right|^2 = \sum_{j=1}^{n_T} \lambda_j \sum_{i=1}^{n_R} \left|h_{ij}'\right|^2 . \qquad (11) \end{aligned}$$

Here, $\boldsymbol{H}$ is the channel matrix, which is assumed to be i.i.d. complex Gaussian distributed with zero mean and unit variance. $\boldsymbol{h}_i$ is the i-th row of $\boldsymbol{H}$. The unitary transformation of $\boldsymbol{h}_i$ by the matrix $\boldsymbol{U}$ does not change the statistics of $\boldsymbol{h}_i$ and therefore the elements h_{ij}' are still i.i.d. complex Gaussian with zero mean and unit variance. Obviously, the squared distances d^2 are Random Variables (RVs) in fading channels. Because of the eight distinct DMs of our code, there are eight distinct distributions of the eigenvalues λ_i and therefore we can distinguish between eight different squared Euclidean distances $d^2_{\mathrm{DM}_i}$ $(i = 1\ldots 8)$ and their distinct distributions. In order to show why at low SNR values PEP1 and at high SNR values PEP8 dominates the BER, we calculate the statistics of $d^2_{\mathrm{DM}_i}$ for the i.i.d. channel.

In the following, we roughly show the way of calculating the Probability Density Functions (PDFs) of the squared distances $d^2_{\mathrm{DM}_i}$. In principle, in our case of the cyclic STBC given in Eqn. (1) there are two different types of PDFs. If the non-zero eigenvalues of the code word DMs $\boldsymbol{A}^{(i)}$ are all equal (as they are for DM1, DM3, DM6, DM7, DM8), then we get PDFs of the first type, namely χ^2-distributed RVs with a certain degree of freedom and a certain variance. In Table 1, we find two different eigenvalues for DM2, DM4 and DM5. In the following these two different eigenvalues are denoted by $\lambda^{(1)}$ and $\lambda^{(2)}$ and the multiplicity of these eigenvalues are denoted by $n^{(1)}$ and $n^{(2)}$, respectively. With this definition we can rewrite Eqn. (11):

$$d^2 = \underbrace{\sum_{i=1}^{n_R\, n^{(1)}} \lambda^{(1)}\,|h_i|^2}_{\alpha} + \underbrace{\sum_{j=1}^{n_R\, n^{(2)}} \lambda^{(2)}\,|h_j|^2}_{\beta} . \qquad (12)$$

The sum of squared magnitudes of independent complex Gaussian RVs with equal variance and zero mean is a χ^2-RV with a certain degree of freedom and variance. Additionally, we know that the PDF of the sum of two independent RVs $(\alpha+\beta)$ is the convolution of the PDFs of α and β. Equivalently, the Characteristic Function (ChF) of the sum of independent RVs is the product of the ChFs (The ChF essentially is the Fourier transform of the PDF.). From page 42 of [4] we know the ChF of a χ^2-distributed RV with n degrees of freedom and variance σ^2:

$$\Psi(j\,\omega) = \frac{1}{(1 - j\,2\,\omega\,\sigma^2)^{\frac{n}{2}}} . \qquad (13)$$

Then, the ChF of the squared distance $\Psi_{d^2}(j\,\omega)$ is:

$$\begin{aligned} \Psi_{d^2}(j\,\omega) &= \Psi_\alpha(j\,\omega)\,\Psi_\beta(j\,\omega) \qquad (14) \\ &= \frac{1}{(1 - j\,\omega\,\lambda^{(1)})^{n_R\, n^{(1)}}\,(1 - j\,\omega\,\lambda^{(2)})^{n_R\, n^{(2)}}} . \end{aligned}$$

Knowing the ChF we can calculate the PDFs of the squared distance d^2 for the DM2, DM4 and DM5 by applying the inverse Fourier transform to $\Psi_{d^2}(j\,\omega)$. The weighted PDFs of $d^2_{\mathrm{DM}_i}$ are depicted in Fig. 2 for all DMs. The weighting is done according to the number of code pairs shown in Table 1.

Assuming ML-detection, an error occurs if the noise amplitude n is larger than half the distance d between two competing code words. For low SNR values the noise variance σ_n^2 is large, i.e., there are a lot of large noise amplitudes

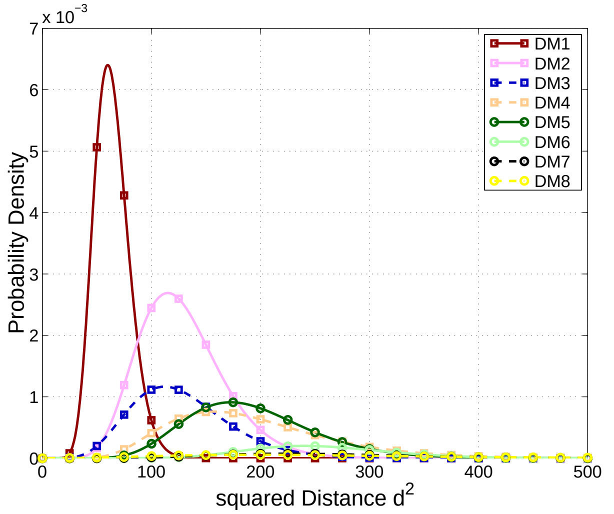

Fig. 2. The weighted PDFs of $d^2_{\mathrm{DM}_i}$ (Weighted according to the No of code pairs).

n. If we look at the PDF of the squared Euclidean distance d^2 for DM1 which is shown by the solid brown curve in Fig. 2, then we can see that the probability of having a small value of d^2 in case of DM1 is relatively high. This means that for low SNR-values the minimum Euclidean distance of the code word pairs is dominated by DM1. Note, that there are distances which are smaller than the distances of the DM1, but they are less frequent and therefore they can be neglected. In Fig. 3 we show the zoomed PDFs of $d^2_{\mathrm{DM}_i}$

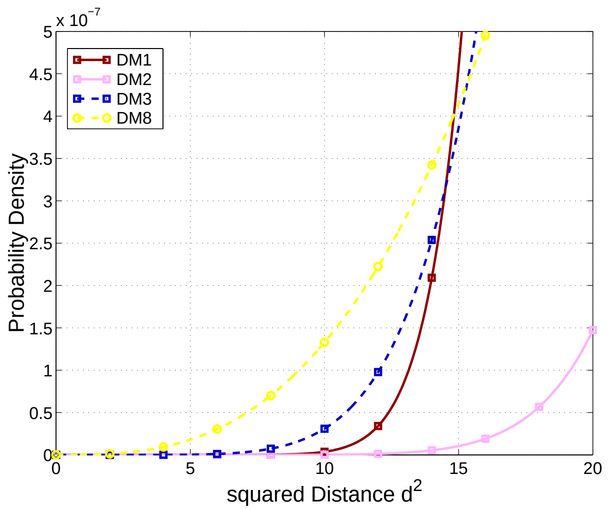

Fig. 3. The zoomed part of the PDFs of $d^2_{\mathrm{DM}_i}$ (Weighted according to the No of code pairs).

for low values of d^2. In this figure the PDF values of some DMs are so small that they do not show up. In particular this is true for DM4, DM5, DM6 and DM7. For high values of SNR, where the noise variance σ_n^2 and thus most amplitudes of n are small, things are quite different from the case of low SNR values. Fig. 3 shows that for very small values of d the dashed yellow curve which corresponds to DM8 dominates the Euclidean distances. Therefore, for high SNR values PEP8 dominates the total BER-curve.

Summarizing these results in Fig. 2 and 3 showing different distributions on d^2, we observe, that different PEPs dominate the BER-curve at low SNR values and at high SNR values.

As a consequence, it turns out that STBCs with rank deficiencies do not necessarily lead to poor performance, if they are operated at low and medium values of SNR, where single symbol errors dominate the BER curve as can be seen in Fig. 1. Rank deficiency deteriorates the BER curve shown in Fig. 1 only at high values of SNR, where the PEP8 dominates the BER curve resulting in a low degree of diversity. However, in the example of our cyclic STBC this deficiency is only relevant for BER values below 10^{-6}.

3. ORTHOGONAL SPACE TIME BLOCK CODES

In this section the influence of multiple symbol errors on the BER-curve in the case of orthogonal codes is discussed. The first results about orthogonal codes have been presented in [5]. The codes presented in [5] are designed according to the GCOD. Such codes are defined in Theorem 5.5.1 in [5].

3.1. BER Performance of GCOD Codes

We start with the following Lemma:

Lemma: *For GCOD-Codes, all eigenvalues of any DM are identical.*

Proof: As stated in [6], the code word DM of all distinct code pairs of codes of the GCOD is a weighted identity matrix:

$$\begin{aligned} \boldsymbol{A}_{i,j} &= (\boldsymbol{X}_i - \boldsymbol{X}_j)^H (\boldsymbol{X}_i - \boldsymbol{X}_j) \\ &= \underbrace{\left(|\Delta x_1|^2 + |\Delta x_2|^2 + \cdots + |\Delta x_N|^2\right)}_{c} \boldsymbol{I}\,, \end{aligned} \tag{15}$$

which is valid for all pairs of (i, j). In case of such an GCOD, we also have different DMs, but since $\boldsymbol{A}_{i,j} = c\boldsymbol{I}$ all eigenvalues of any DM are identical due to the specific structure of the DM in Eqn. (15).

Although all eigenvalues corresponding to one DM are identical, the eigenvalues for different DMs are distinct (see Table 2). The trace of a distinct code word DM $\boldsymbol{A}^{(k)}$ with identical eigenvalues λ_k is: $trace(\boldsymbol{A}^{(k)}) = n_T\,\lambda_k$. The more symbol errors occur, the larger is the factor c in Eqn. (15). If the diagonal elements of $\boldsymbol{A}^{(k)}$ are large, then the corresponding eigenvalues λ_k are high. Therefore, we conclude that single symbol errors, which correspond to small diagonal elements of $\boldsymbol{A}$ according to Eqn. (15) and thus to small eigenvalues λ, dominate the BER-curve of GCODs.

3.2. Example

To clarify the before mentioned Lemma, we analyze a typical example of a code of the GCOD [6] with rate 3/4 in more detail:

$$\boldsymbol{X} = \begin{pmatrix} x_1 & x_2 & x_3 & 0 \\ -x_2^* & x_1^* & 0 & x_3 \\ -x_3^* & 0 & x_1^* & -x_2 \\ 0 & -x_3^* & x_2^* & x_1 \end{pmatrix} . \tag{16}$$

For this code, the code word DM $\boldsymbol{A}_{i,j}$ can be written as:

$$\begin{aligned} \boldsymbol{A}_{i,j} &= (\boldsymbol{X}_i - \boldsymbol{X}_j)^H (\boldsymbol{X}_i - \boldsymbol{X}_j) \qquad (17) \\ &= \left(\left| x_1^i - x_1^j \right|^2 + \left| x_2^i - x_2^j \right|^2 + \left| x_3^i - x_3^j \right|^2 \right) \boldsymbol{I} , \end{aligned}$$

which is valid for all pairs (i, j). For BPSK modulation ($x_i \in \{+1, -1\}$) and because of Eqn. (17), we obviously get only three different types of DMs. The eigenvalues of these DMs are shown in Table 2. Obviously, all DMs

type of DM	λ_1	λ_2	λ_3	λ_4	No of code pairs	No of different symbols
	0	0	0	0	8	0
1	4	4	4	4	24	1
2	8	8	8	8	24	2
3	12	12	12	12	8	3

Table 2. Different eigenvalue constellations for all DMs.

have four identical eigenvalues ($\lambda_1 = \lambda_2 = \lambda_3 = \lambda_4$). For these three DMs, the upper bounds for the BER are shown in Fig. 4. As expected and already theoretically explained, the

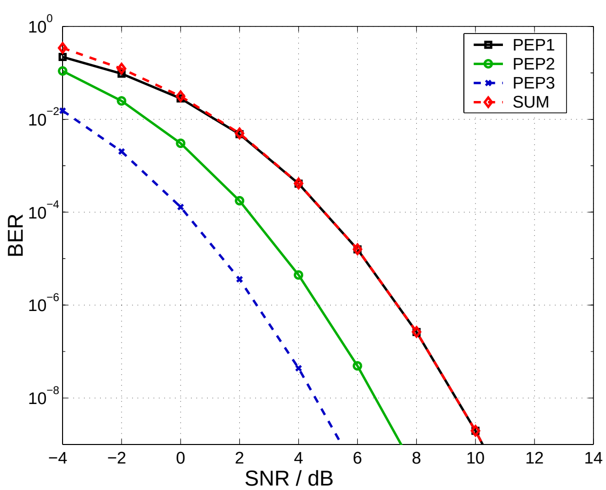

Fig. 4. The upper bound for the BER vs. SNR curves for the 3 PEPs and the sum BER; 4 receive antennas

sum BER-curve exclusively depends on PEP1 with single symbol errors which dominate the BER in the whole SNR range.

Interestingly enough, for orthogonal codes of the GCOD only single symbol errors and no other PEPs dominate the BER performance in the whole range of SNR values.

4. CONCLUSION

For STBCs in general we have to consider all PEPs and not only the nearest neighbor error events in order to adequately describe the BER performance of these codes. The upper bound of the BER-curve shows that for low SNR the BER-curve is dominated by the nearest neighbor errors (PEP1) whereas for high SNR the PEPs corresponding to multiple symbol errors dominate the error performance. This fact is explained by means of the Euclidean distance profile of pairwise error events.

Additionally, we showed that for reasonable low SNR values it is not that important to use full rank STBCs and to consider all possible PEPs. If rank deficiencies occur quite rarely, then they do not influence the total BER in the range of low to medium SNR that is of major interest in practical applications.

In contrast to general STBC, the BER-performance of GCOD-codes is dominated by single symbol errors in the whole range of SNR. For such codes it is not necessary to consider all distinct PEPs to characterize the BER performance.

5. ACKNOWLEDGMENT

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