Space-Time Differentially Coded Orthogonal Matrix Modulation using QAM

Anil Mengi
Institute for Experimental Mathematics
University Duisburg-Essen
Ellernstr. 29, D-45326
Essen, Germany
Email: mengi@exp-math.uni-essen.de

Gerhard Bauch
DoCoMo Euro-Labs
Landsberger Strasse 308-312, D-80687
Munich, Germany
Email: bauch@docomolab-euro.com

A. J. Han Vinck
Institute for Experimental Mathematics
University Duisburg-Essen
Ellernstr. 29, D-45326
Essen, Germany
Email: vinck@exp-math.uni-essen.de

Abstract

This work addresses digital communication in a wireless system with multiple antennas at transmitter and/or receiver when the channel state information is unknown at the transmitter as well as at the receiver. We propose a generalized bandwidth-efficient space-time differentially coded orthogonal non-unitary matrix modulation scheme with non-coherent soft-output detector. Most proposals which are available in literature are based on unitary matrices which limit the achievable data rate and performance. In contrast to schemes which use unitary matrices, in non-unitary matrix modulation, symbols are required to be normalized at the transmitter. This normalization factor guarantees that the differential scheme meets a power constraint but at the same time makes it hard to design a soft-output detector. Compared to the existing differential orthogonal matrix modulation schemes, our main contribution is a new soft-output detector which does not require knowledge of channel coefficients, channel statistics or noise variance.

I. Introduction

Space-time differentially coded unitary matrix modulation schemes [1], [3], [4] are attractive alternatives since they don't require channel estimation at the receiver. However, for those techniques which require the constellation elements to be unitary matrices, the achievable data rate and performance are limited due to the distance properties of phase shift-keying (PSK). In order to increase the bandwidth efficiency, a differential transmit diversity approach based on multiple amplitude levels was proposed in [5]. A more generalized extension of unitary matrix modulation by an additional matrix norm modulation was proposed in [6]. This proposal still includes unitary matrices.

In this paper, we construct a communication system with an idea of increasing the bandwidth efficiency by sacrificing the unitary matrix design. This leads to the concept of space-time differential non-unitary matrix modulation. Our proposed scheme is based on orthogonal designs with quadrature amplitude modulation (QAM) modulation. Earlier proposals for non-unitary differential matrix modulation based on orthogonal designs with QAM modulation were presented in [8]–[11]. However, we provide a non-coherent soft-output detector and an alternative way to the calculation of the normalization factor which does not lead to an error propagation.

II. MIMO Channel Model

We consider a flat fading multiple-input multiple-output (MIMO) channel for a system with n_T transmit and n_R receive antennas. The path gain from transmit antenna i to receive antenna j at matrix index k is described as $h_k^{(ij)}$. The channel coefficients are collected in the matrix $\mathbf{H}_k$.

The antennas are separated far enough to ensure independently fading channels from each transmit to each receive antenna. Therefore, independent fading may be modelled by selecting the elements of $\mathbf{H}_k$ as an equal variance complex Gaussian random variables with i.i.d. real and imaginary parts.

At the receiver, we observe

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k, \tag{1}$$

where $\mathbf{X}_k$ and $\mathbf{Y}_k$ denote the transmitted and received symbols, respectively, and $\mathbf{N}_k$ contains the noise samples which are assumed to be independent and Gaussian with variance

$$\sigma^2 = \frac{N_o}{2} \tag{2}$$

per real dimension.

III. REVIEW ON SPACE-TIME DIFFERENTIALLY CODED UNITARY MATRIX MODULATION

Space-time differential unitary modulation was introduced in [12] and simultaneously in [3] and [4]. In this work, we consider a differential detection scheme where encoding is done on a matrix basis and the reference matrix and the info matrices are constructed according to the orthogonal designs. The incoming info bits are mapped on an $L \times L$ dimensional info matrix $\mathbf{C}_k$. In order to allow simple non-coherent detection, $\mathbf{C}_k$ must be unitary, i.e.

$$\mathbf{C}_k \mathbf{C}_k^H = \mathbf{I}_L, \qquad k = 1, 2, ... \tag{3}$$

where $\mathbf{C}_k^H$ represents the conjugate transpose of $\mathbf{C}_k$ and $\mathbf{I}_L$ is the $L \times L$ dimensional identity matrix. A suitable construction of unitary matrices is using orthogonal designs [12] with PSK symbols $c_{k,l}$. In the case of $n_T = 2$ transmit antennas, we have the Alamouti scheme where $K = 2$ symbols are mapped to the matrix

$$\mathbf{C}_k = \begin{pmatrix} c_{k,1} & c_{k,2} \\ -c_{k,2}^* & c_{k,1}^*. \end{pmatrix} \tag{4}$$

The $n_T \times L$ dimensional transmit matrix $\mathbf{X}_k$ is obtained by $\mathbf{C}_k$ and the previously transmitted matrix $\mathbf{X}_{k-1}$ according to the differential encoding rule given by

$$\mathbf{X}_k = \mathbf{X}_{k-1} \mathbf{C}_k. \tag{5}$$

The major drawback with space-time differentially coded unitary matrix modulation is since the unitary condition is achieved by symbols based on PSK mapping, the performance will be poor for more bandwidth-efficient transmission, i.e. $M \geq 8$.

IV. REVIEW ON SPACE-TIME DIFFERENTIALLY CODED AMPLITUDE AND UNITARY MATRIX MODULATION

Extensions of differential unitary matrix modulation with an additional differential amplitude modulation for higher bandwidth-efficiency with improved performance have been introduced in [5]–[7]. We consider the general approach according to [6].

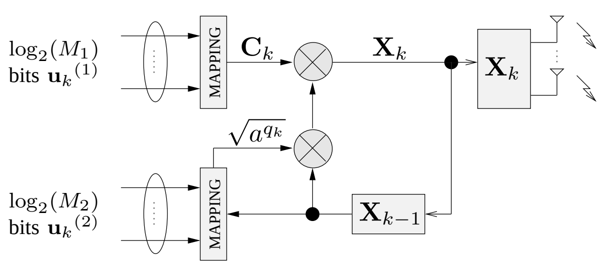


Fig. 1. Differentially coded amplitude and unitary matrix modulation, transmitter.

Figure 1 depicts the transmitter structure. The encoding is characterized by two separate parts. First part applies the conventional differential unitary modulation where the first $b_1 = \log_2(M_1)$ bits are mapped on a unitary matrix $\mathbf{C}_k$ where M_1 denotes the number of different info matrices. Second part encodes $b_2 = \log_2(M_2)$ bits into the amplitude differences of the transmit matrix $\mathbf{X}_k$ compared to the previously transmitted matrix $\mathbf{X}_{k-1}$ where M_2 shows the number of possible amplitude values. Due to the additional amplitude modulation, transmit matrix $\mathbf{X}_k$ is not unitary anymore but the underlying structure is still an orthogonal design.

At the output of the encoder, the transmit matrix can be written as

$$\mathbf{X}_k = \sqrt{a^{q_k}} \mathbf{X}_{k-1} \mathbf{C}_k. \tag{6}$$

In Equation (6), a is just a constant and evaluated in [6] according to the BER analysis. Amplitude exponent coming from the differential amplitude modulation is described as $q_k \in \{-M_2+1, -M_2+2, \ldots, -1, 0, 1, \ldots, M_2-1\}$.

A drawback of this scheme is that the amplitude modulation has to be done per matrix rather than per symbol which reduces the bandwidth-efficiency with increasing matrix size, i.e. increasing number of transmit antennas.

V. SPACE-TIME DIFFERENTIALLY CODED ORTHOGONAL NON-UNITARY SPACE-TIME MODULATION

In the previous sections, we gave a brief description of different space-time matrix modulation schemes where the mapping of the info bits onto info matrix $\mathbf{C}_k$ is chosen from the PSK constellation. However, as we remarked earlier, as higher modulations are employed, minimum distance properties of PSK causes poor performance. The new proposal increases the bandwidth-efficiency by using differential non-unitary matrix modulation. The idea of differentially coded non-unitary modulation has been also proposed in [5], [8], [9]. However, we provide a non-coherent soft-output detector and an alternative way to the calculation of the normalization factor which does not lead to an error propagation.

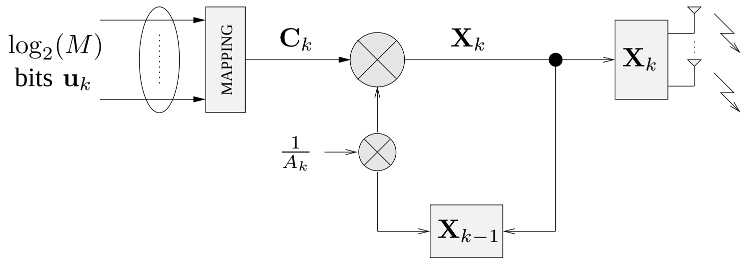


Fig. 2. Differentially coded non-unitary matrix modulation, transmitter.

Differential non-unitary space-time transmitter is shown in Figure 2. Let M denote the number of different info matrices at matrix index k; a group of data bits $\mathbf{u}_k = u_{k,1}, ..., u_{k,b}$ where $u_{k,t} \in \{+1, -1\}$ and $b = \log_2(M)$ is mapped onto $L \times L$ dimensional info matrix $\mathbf{C}_k$. Once again, the distribution of the symbols with its conjugates in the data matrix are designed to achieve orthogonality. However, different from the previous schemes, mapping is based on QAM modulation. In the sense of orthogonality, the design rule of the info matrix $\mathbf{C}_k$ can be written as

$$\begin{aligned} \mathbf{C}_k \mathbf{C}_k^H &= a_k \mathbf{I}_L \\ &= \left(\sum_{i=1}^{K} |c_{k,i}|^2 \right) \mathbf{I}_L, \end{aligned} \tag{7}$$

where $\mathbf{I}_L$ is $L \times L$ identity matrix and a_k is defined as the amplitude of the info matrix $\mathbf{C}_k$ with variables $c_{k,1}, c_{k,2}, \ldots, c_{k,K}$.

At the beginning of each frame, the differential encoder requires an $L \times L$ identity matrix in order to start the transmission. We note that, as a result of non-unitary info matrices, the transmit power does not stay constant after differential encoding. Thus, a normalization factor $1/A_k$ should be introduced at the feedback loop. In that case, the transmit matrix can be written as

$$\mathbf{X}_k = \frac{1}{A_k} \mathbf{X}_{k-1} \mathbf{C}_k, \tag{8}$$

where the normalization factor is added in order to satisfy the energy constraint given

$$\varepsilon_{k,l} \left\{ \sum_{n=1}^{n_T} |x_{k,l}^{(n)}|^2 \right\} = E_s. \tag{9}$$

In Equation (9), ε denotes the expectation with respect to the matrix index k and time slot l. At the right side of the equality, E_s indicates the total average transmit energy per time slot. As long as the transmit matrix is kept unitary, the result of differential encoding does not have any effect on the average transmit energy. In the case of non-unitary space-time modulation, restriction in Equation (9) may be achieved by normalizing the transmit matrix with an appropriate normalization factor. A propose of our work is to calculate the normalization factor by the energy of the previously transmitted matrix, i.e.

$$A_k = \sqrt{\text{trace}\{\mathbf{X}_{k-1} \mathbf{X}_{k-1}^H\}/L} \tag{10}$$

in order to fulfill the energy constraint in Equation (9). Since $\mathbf{C}_k$ is an orthogonal design, Equation (10) is equivalent to choosing the normalization factor as the amplitude of the previous matrix, i.e.

$$A_k = \sqrt{a_{k-1}}. \tag{11}$$

A. Soft Output Detector

One of the concern of this work is to design a soft-output receiver structure that produces reliable log-likelihood ratios at the output.

Our proposed receiver structure combined with transmitter is shown in Figure 3. Non-coherent soft-output receiver considers two successively received matrices

$$\mathbf{Y}_{k-1} = \mathbf{H}_{k-1} \mathbf{X}_{k-1} + \mathbf{N}_{k-1}, \tag{12a}$$

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{N}_k = \frac{1}{A_k} \mathbf{H}_k \mathbf{X}_{k-1} \mathbf{C}_k + \mathbf{N}_k. \tag{12b}$$

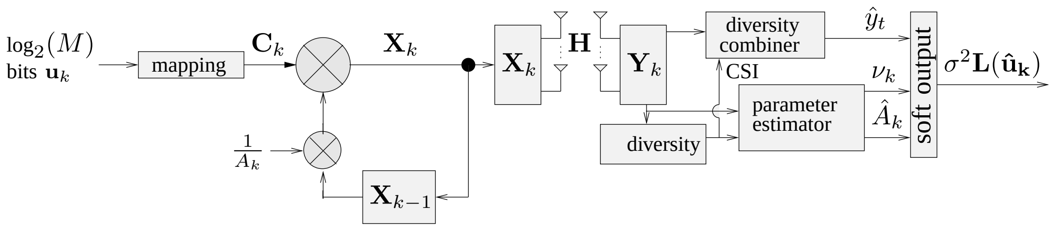


Fig. 3. Differential non-unitary space-time system.

where the channel is assumed to be constant for at least two successively transmitted matrix, i.e. $\mathbf{H}_{k-1} = \mathbf{H}_k$.

Plugging (12a) into (12b) yields

$$\begin{aligned} \mathbf{Y}_k &= \frac{1}{A_k}\mathbf{Y}_{k-1}\mathbf{C}_k - \frac{1}{A_k}\mathbf{N}_{k-1}\mathbf{C}_k + \mathbf{N}_k \\ &= \frac{1}{A_k}\mathbf{Y}_{k-1}\mathbf{C}_k + \tilde{\mathbf{N}}_k, \end{aligned} \tag{13}$$

Accordingly, Equation (13) models a non-coherent system with L transmit and n_R receive antennas where the info matrix $\mathbf{C}_k$ is transmitted over a flat fading channel whose coefficients can be represented in terms of the previously received matrix and the normalization factor, i.e. $\tilde{\mathbf{H}}_k = \frac{1}{A_k}\mathbf{Y}_{k-1}$. It is also observed that additive white Gaussian noise $\tilde{\mathbf{N}}_k$ may be expressed in terms of two parts which are both additive white Gaussian noise with the power σ^2. In the design of the non-coherent soft-output detector, two parameters must be calculated. These are the variance of the additive white Gaussian noise $\tilde{\mathbf{N}}_k$ which is obtained in Equation (13) and the normalization factor A_k.

The variance of the additive Gaussian noise $\tilde{\mathbf{N}}_k$ per real dimension at each virtual receive antenna can be calculated as

$$\tilde{\sigma}^2 = \sigma^2\left(1 + \frac{a_k}{A_k^2}\right), \tag{14}$$

where σ^2 is the noise variance per real dimension at each physical receive antenna. We denote the required estimate as

$$\nu_k = \frac{a_k}{A_k^2}. \tag{15}$$

Our proposed possibility to obtain estimates for those parameters is to find an approximation of $\frac{a_k}{A_k^2}$ is to compute

$$\nu_k = \frac{\text{trace}\{\mathbf{Y}_k\mathbf{Y}_k^H\}}{\text{trace}\{\mathbf{Y}_{k-1}\mathbf{Y}_{k-1}^H\}} \tag{16}$$

in the absence of noise.

Due to the orthogonality of the space-time block code matrix, a significant simplification in the implementation of the detector occurs by using a diversity combiner [12], [13]. In such a case, the detection rule simplifies since we can obtain decoupled expressions for the M-ary symbols $c_{k,l},\ l = 1,\dots,K$ where we replace the channel coefficient matrix $\mathbf{H}_k$ by $\frac{1}{A_k}\mathbf{Y}_{k-1}$ as a channel state information. This yields an equivalent system

$$\hat{y}_{k,l} = \frac{1}{A_k}\text{trace}\{\mathbf{Y}_{k-1}^H\mathbf{Y}_{k-1}\}c_{k,l} + \hat{n}_{k,l},\ l = 1,\dots,K, \tag{17}$$

of K single-input single-output (SISO) channels with noise variance

$$\hat{\sigma}^2 = \tilde{\sigma}^2\text{trace}\{\mathbf{Y}_{k-1}^H\mathbf{Y}_{k-1}\}. \tag{18}$$

Because of the dependence of the normalization factor in the transmitted data , we also need an estimate on A_k itself. In this paper, we propose three methods of estimating normalization factor at the receiver.

1) A first method is to ignore the normalization factor $1/A_k$ at the receiver by assuming $1/A_k = 1$. This works surprisingly well particularly in time varying channels, where the channel is not exactly constant during transmission of two successive matrices.
2) Second solution is, since A_k is determined by the previously transmitted matrix $\mathbf{X}_{k-1}$, the estimates $\hat{A}_k$ on A_k can be obtained from the hard decisions $\hat{C}_t,\ t < k$. More precisely: Knowing $\mathbf{X}_{k-1}$ yields $\hat{A}_k$ for detection of $\mathbf{C}_k$. From the hard decision $\hat{\mathbf{C}}_k$, we obtain an estimate for $\mathbf{X}_k$ which yields $\hat{A}_{k+1}$ etc.. The problem with this approach is that it imposes error propagation.

3) Another possibility of estimating $1/A_k$ at the receiver is to neglect noise and estimate by using Equation 16, i.e.

$$\nu_k = \frac{\text{trace}\{\mathbf{Y}_k \mathbf{Y}_k^H\}}{\text{trace}\{\mathbf{Y}_{k-1} \mathbf{Y}_{k-1}^H\}} \approx \frac{a_k}{A_k^2}, \tag{19}$$

where a_k is the amplitude of the orthogonal designed info matrix $\mathbf{C}_k$. From (11), we know that

$$A_{k+1}^2 = a_k. \tag{20}$$

Since the first transmitted matrix $\mathbf{X}_0$ is a known reference matrix, A_1 is known at the receiver. Typically, the reference matrix will be the identity matrix and, hence $A_1 = 1$. For $k > 1$, an estimate $\hat{A}_k$ is obtained using (19) and (11) from

$$\hat{A}_{k+1}^2 = \nu_k \hat{A}_k^2. \tag{21}$$

Method 3 turns out to be advantageous in quasistatic channels, whereas method 2 performs better in time varying channels, where the condition $\mathbf{H}_{k-1} = \mathbf{H}_k$ is violated.

B. Computation of Log-Likelihood Ratios

With estimates on $\tilde{\sigma}^2$ and $\hat{A}_k$, soft-output receiver after a standard diversity combiner computes the logarithmic probability as

$$\begin{aligned}\log p(c_{k,l} \mid \hat{y}_{k,l}, \mathbf{Y}_{k-1}) &= \text{const} + \frac{1}{2\hat{\sigma}^2}\left(\frac{2}{\hat{A}_k}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}\text{real}\{y_{k,l}^* c_{k,l}\}\right.\\ &\left. -\frac{1}{\hat{A}_k^2}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}^2 |c_{k,l}|^2\right) + \frac{1}{2}\mathbf{u}_{k,l}^T \mathbf{L}_a(\mathbf{u}_{k,l}),\end{aligned} \tag{22}$$

where $u_{k,l}$ contains the $\log_2 M$ bits which determine the M-QAM symbol $c_{k,l}$. Using (17), (18) and (15), the APP loglikelihood ratios yield

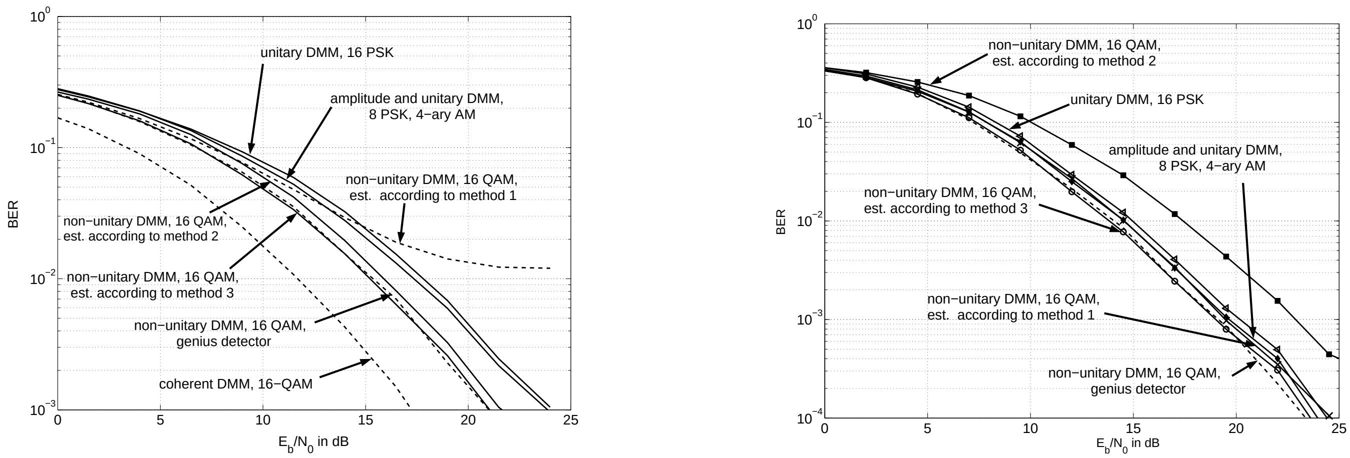
$$\begin{aligned}\sigma^2 L(\hat{u}_{k,t}) \approx\ & \max_{\substack{C_k \\ u_{k,t=+1}}} \left\{ \frac{1}{2\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}(1+\nu_k)} \left(\frac{2}{\hat{A}_k}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}\text{real}\{y_{k,l}^* c_{k,l}\}\right.\right.\\ &- \left.\left.\frac{1}{\hat{A}_k^2}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}^2 |c_{k,l}|^2\right) + \frac{\sigma^2}{2}\mathbf{u}_k^T \mathbf{L}_a(\mathbf{u}_k)\right\}\\ &- \max_{\substack{C_k \\ u_{k,t=-1}}} \left\{ \frac{1}{2\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}(1+\nu_k)} \left(\frac{2}{\hat{A}_k}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}\text{real}\{y_{k,l}^* c_{k,l}\}\right.\right.\\ &- \left.\left.\frac{1}{\hat{A}_k^2}\text{trace}\{\mathbf{Y}_{k-1}^H \mathbf{Y}_{k-1}\}^2 |c_{k,l}|^2\right) + \frac{\sigma^2}{2}\mathbf{u}_k^T \mathbf{L}_a(\mathbf{u}_k)\right\},\end{aligned} \tag{23}$$

where $t = 1, \ldots, \log_2(M)$ and $l = 1, \ldots, K$. We compute simply the APP log-likelihood ratios multiplied by σ^2 in order to make the right hand side of (23) independent of the unknown noise variance σ^2. If the noise variance is constant over a frame, which is a reasonable assumption, all log-likelihood ratios are scaled by a constant factor σ^2. This has no effect on the hard output of an outer Viterbi or Max-Log-type APP decoder. However, the APP loglikelihood ratios of the outer decoder will also be scaled by the same factor.

VI. SIMULATION RESULTS

In our simulation system, for forward error control (FEC) coding, we use convolutional encoder which generates a recursive and systematic convolutional code with a rate $1/2$ and with a memory 4. The generators for this code are given in octal form as $(1, 35/23)$. We compare the performance of the schemes mentioned in the last sections for $n_T = 2$ transmit antennas and $n_R = 1$ receive antenna in a spatially uncorrelated channel. We consider transmission of 8 bits per matrix. This can be achieved with space-time differentially coded unitary matrix modulation (unitary DMM) based on 16-PSK or with non-unitary matrix modulation (non-unitary DMM) based on 16-QAM. Alternatively, we can use the space-time differentially coded amplitude and unitary matrix modulation described in Section IV based on 8-PSK modulation plus 4-ary amplitude modulation. ($M_1 = 8, M_2 = 4$).

We consider a quasistatic channel, i.e. the channel is constant during a coded block and changes independently from one block to the next. BER results are given in Fig 4.



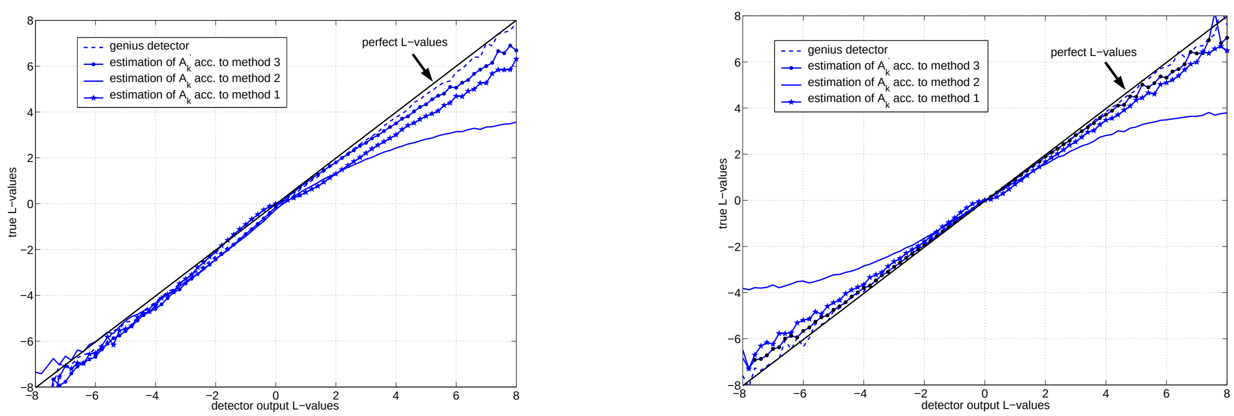
(uncoded transmission) **(coded transmission)**

Fig. 4. Uncoded and coded BER for transmission over quasistatic channel. 8 bits per matrix, $n_T = 2$, $n_R = 1$

The performance of the non-unitary scheme depends on the estimation method for the effective channel parameters as discussed in Section V-A where we introduced 3 methods. For comparison, we include the performance of the genius detector which perfectly knows the otherwise estimated effective channel parameters.

From Fig. 4, in the case of uncoded transmission, it can be observed that for 16-QAM, the performance of the detector with the method 3 (estimation of A_k based on ν) is almost the same as that of genius detector while the performance of the method 2 (estimation of A_k based on hard decision) is about 0.5 dB worse than that with proposed scheme at BER of 10^{-3}. This degradation is due to the fact that estimates $\hat{A}_k$ are calculated from the previous decoded info matrix $\hat{C}_{k-1}$, which results in error propagation. It is noticeable that with the optimum detector, non-unitary DMM performs 2.2 dB better than amplitude and unitary DMM and 2.5 dB better than unitary DMM. We also included the performance of a receiver which neglects the normalization factor according to the method 1 and sets $A_k = 1$. This method results in an error floor.

The respective BER results for FEC coded transmission are also depicted in Fig. 4. Here, the performance advantage of non-unitary DMM over amplitude and unitary DMM reduces to 0.5 dB. This is mainly due to the reason that the coding gain varies according to the computation of log-likelihood ratios especially in the case of non-unitary DMM where the correct estimation of the normalization factor is very crucial for the correct weighting of the output log-likelihood ratios. The quality of the soft-output log-likelihood ratios of the non-unitary scheme is worse than for amplitude and unitary DMM resulting in a performance degradation of the FEC decoder.



(Gray mapping) **(non-Gray mapping)**

Fig. 5. Quality of log-likelihood ratios of Non-unitary DMM at SNR=15 dB with 16-QAM over quasistatic channel. 8 bits per matrix, $n_T = 2$, $n_R = 1$

Figure 5 illustrates the quality comparison of the log-likelihood ratios at the output of the differential space-time soft-output detector at SNR$=15$ dB. X-axis shows the output log-likelihood ratios of the differential detector while y-axis shows the true log-likelihood ratios. For perfect soft-outputs we would obtain a diagonal. Notice that in the case of method 2 of estimating normalization factor, the effect of error propagation on the soft-output values can be clearly seen in Fig. 5.

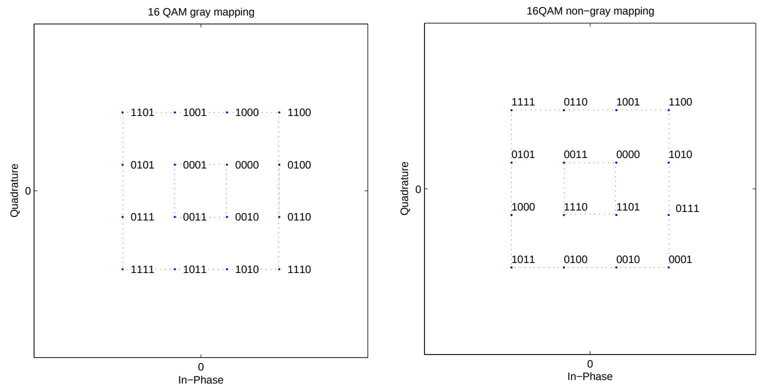


Fig. 6. Symbol constellation of 16 QAM with gray and non-gray mapping.

Another observation from Figure 5 is the asymmetric behaviour of the soft-output detector for the positive log-likelihood ratios compared to the negative log-likelihood ratios. We also observed a shift at the origin compared to the perfect L-values. These effects can be explained with the unbalanced distribution of the $1's$ and $0's$ at the inner square and at the outer square in the case of Gray mapping (see Fig. 6). Figure 5 shows also results according to the non-Gray mapping which has a equal number of $1's$ and $0's$ at the inner square and at the outer square. It can be also seen that the behavior of the soft-output detector is not asymmetric anymore and no shift at the origin is observed . In our simulations Gray mapping is used due to its more robust characteristic against error.

VII. Conclusion

We presented a generalized bandwidth-efficient space-time differentially coded orthogonal non-unitary matrix modulation scheme where neither the receiver nor the transmitter has access to channel state information. In quasistatic fading channels, as higher modulations are employed, our proposed scheme with optimal receiver outperforms space-time differential unitary matrix modulation by 2.5 dB and space-time differential amplitude and unitary modulation by 2.2 dB in uncoded BER. Due to the difficulties in computing soft output values which are passed to the FEC decoder, this advantage is decreased to 1.2 dB and 0.5 dB, respectively, in coded transmission.

References

[1] V. Tarokh and H. Jafarkhani. A differential detection scheme for transmit diversity. *IEEE Journal on Selected Areas in Communications*, 18(7):1169-1174, July 2000.

[2] H. Jafarkhani and V. Tarokh. Multiple transmit antenna differential detection from generalized orthogonal designs. *IEEE Transactions on Information Theory*, 47(6):2626-2631, September 2001.

[3] B. L. Hughes. Differential space-time modulation. *IEEE Transactions on Information Theory*, 46(7):2567-2578, November 2000.

[4] B. Hochwald and W. Swelden. Differential unitary space-time modulation. *IEEE Transactions on Information Theory*, 48(12):2041-2052, December 2000.

[5] X.-G. Xia. Differentially en/decoded orthogonal space-time block codes with APSK signals. *IEEE Communications Letters*, 6(4):150-152, April 2002.

[6] G. Bauch. Differential amplitude and unitary space-time modulation. In *International ITG Conference on Source and Channel Coding*, pages 135-142, January 2004.

[7] G. Bauch. A bandwidth-efficient scheme for non-coherent transmit diversity. In *IEEE Globecom*, December 2003.

[8] M. Tao and R.S. Cheng. Differential space-time block codes. In *IEEE Globecom Conference*, pages 1098-1102, November 2001.

[9] Z. Chen, G. Zhu, J. Shen, and Y. Liu. Differential space-time block codes from amicable orthogonal designs. In *Wireless Communications and Networking Conference (WCNC)*, pages 768-772, March 2003.

[10] C.-S. Hwang, S. H. Nam, J. Chung, and V. Tarokh. Differential space-time block codes using nonconstant modulus constellations. *IEEE Transactions on signal Processing*, 51(11):2955-2964, November 2003.

[11] C.-S. Hwang, S. H. Nam, J. Chung, and V. Tarokh. Differential space-time block codes using QAM constellations. In *International Sympososium on Personal, Inddor and Mobile Radio Communications (PIMRC)*, September 2003.

[12] V. Tarokh, H. Jafarkhani, and A. R. Calderbank. Space-time block codes from orthogonal designs. *IEEE Transactions on Information Theory*, 45(5):1456-1467, June 1999.

[13] G. Bauch, J. Hagenauer, and N. Seshadri. Turbo processing in transmit antenna diversity. *Annals of Telecommunications, Special Issue: Turbo codes - a widespreading technique*, 56(7-8):455-471, August 2001.