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RATE 1 QUASI ORTHOGONAL UNIVERSAL TRANSMISSION AND COMBINING FOR MIMO SYSTEMS ACHIEVING FULL DIVERSITY

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ABSTRACT

This work addresses general multiple-input multiple-output systems and develops combined diversity transmission and combining schemes that achieve rate one and full diversity with reduced decoding complexity, while being universal in the sense that the operations performed at both transmission ends are channel independent. Such schemes may be useful in a scenario where a multiple-antenna source node communicates with the cloud via a multiple-antenna "dumb relay" that forwards the received vector over a rate-constrained digital front-haul link or serves as relay performing an amplify-forward operation over the air. The proposed schemes are derived by establishing an operational equivalence relation between the true channel and an associated multiple-input single-output channel.

***Index Terms*—** MIMO systems, diversity methods, space-time codes, relays, fronthaul

1. INTRODUCTION

A well-known challenge for achieving the goal of ultra-reliable communication over a wireless link is overcoming channel fading, especially so when small payloads are to be transmitted and/or when operating over a narrow frequency band. One of the fundamental approaches to reduce the impact of fading is to use multiple transmit and/or receive antennas. Specifically, the maximal achievable diversity of a $N_t \times N_r$ multiple-input multiple-output (MIMO) channel with i.i.d. Rayleigh fading, is $N_r N_t$ [1].[1]

Simple methods can be used to take advantage of the additional antennas. For instance, full diversity can be achieved via repetition (sending each symbol over the different antennas in sequence). However, repetition results in a drastic reduction in the effective symbol rate. Therefore, schemes that attain maximal diversity while retaining a high symbol rate are of interest.

The difficulty of meeting this goal greatly depends on two factors: the availability of channel state information (CSI) and the MIMO configuration. Clearly, the task is simple if CSI is available at both transmission ends. Another simple scenario is when only the receiver is equipped with multiple antennas where it suffices for the receiver to have CSI in order to attain maximal diversity while supporting a single stream (symbol rate of 1) by employing maximal ratio combining (MRC). Moreover, the latter effectively converts the channel to an equivalent single-input single output (SISO) channel, thus leading also to reduced detection complexity. Conversely, the same is true in the multiple-input single-output (MISO) scenario by employing transmit beamforming assuming the transmitter has access to CSI.

The MISO channel with CSI available only at the receiver, a scenario that is very common in practice, is more challenging. A well known approach to attain full diversity is to employ space-time codes, including space-time trellis codes and space-time block codes (STBC); see, e.g., [2,3].

Since the decoding complexity of maximum likelihood (ML) detection of "generic" full-diversity space-time codes scales exponentially with the number of transmit antennas, considerable work has been devoted to deriving space-time codes that allow for reduced-complexity ML detection, as well as to developing non-ML detection techniques that can attain full diversity (when full-diversity codes are used). An example of the latter approach is lattice-reduction aided detection [4–7] and its ramifications [8–10].

A well-known family of space-time block code (STBC) allowing for reduced-complexity ML detection, is that of orthogonal space-time block-codes (OSTBC), for which ML detection can be done symbol-by-symbol (for uncoded symbols). As is well known, for the case of two transmit antennas the Alamouti OSTBC [11] supports a symbol rate of 1 with minimal possible blocklength, i.e., two channel uses, and is thus "ideal". When the number of antennas grows however, the symbol rate of orthogonal OSTBCs decreases, ultimately down to $1/2$, see, e.g. [12]. Moreover, the blocklength of orthogonal STBCs grows rapidly with the number of antennas.

When the number of transmit antennas is large, quasi-orthogonal space-time block code (QOSTBC) [13–15] with symbol rotation [16–18] are an attractive transmission scheme as they achieve full diversity while maintaining a symbol rate of 1 as well as offering reduced-complexity ML detection. Specifically, it was shown in [17] that the generalization of the code suggested in [13] achieves rate one, full diversity and a ML detection complexity that amounts to separate ML detection of each half of the transmitted symbols. Another approach is the design of STBCs that allow to achieve full diversity with the aid of *linear equalization* in conjunction with symbol-by-symbol detection as developed in [19, 20]; however, the symbol rate approaches 1 only as the block length tends to infinity.

[1]The diversity order is defined as the slope of the error probability curve at (asymptotically) high signal-to-noise ratio (SNR).

Consider now the case of a SIMO channel. Recently, [21] and [22] showed that for the 1×2 channel, channel-independent combining at the receiver can achieve full diversity by introducing the dual of Alamouti modulation.

Such universal combining is beneficial in a cloud radio access networks (C-RAN) architecture in which the transmitted symbols are received by a relay (remote radio head) and are then forwarded to the cloud for detection; see, e.g. [23]. It is desirable to transfer as many functionalities as possible (without significantly scarifying performance) to the cloud, i.e., making the relay as “dumb” as possible. Thus, channel oblivious combining, allowing channel estimation to be performed only at the cloud, where detection takes place, is of interest.

In this work, we extend the approach to more general MIMO channels. It is shown that rate 1 full-diversity universal transmission and combining can be achieved for any MIMO channel for which the number of transmit and receive antennas are integer powers of two, while supporting detection complexity that amounts to separate detection for each half of the transmitted information symbols.

2. SYSTEM MODEL

We consider a topology in which the receiver is composed of two blocks: a combining unit and a detector where these two blocks are not necessarily collocated. We begin by describing the channel between the transmitter and the combining unit.

We further assume that the transmitter uses space-time block codes as defined in [24]. The transmitter wishes to transmit K symbols $\tilde{x}_1, \dots, \tilde{x}_K$ over T channels uses that may be described by a *space-time transmission matrix* denoted by $\mathbf{C}^{K,T} \in \mathbb{C}^{T \times N_t}$, where the entry $c_{t,i}$ is the transmitted symbol at time t from antenna i , $t \in \{1, \dots, T\}$ and $i \in \{1, \dots, N_t\}$. Recall that the minimum value of T required to achieve full diversity is $T = N_t$ [24]. Codes having the minimum value of T are called “delay optimal”. The induced symbol rate is denoted by $R_t = \frac{K}{T}$.

We assume that the entries of the transmission matrix may be one of the indeterminates $\pm x_1, \dots, \pm x_K$ and their conjugates $\pm x_1^*, \dots, \pm x_K^*$ (up to a power normalization factor). Those indeterminates are derived from the information symbols by (possibly) applying a certain transformation on them (e.g., symbols rotations as in [17]).

We set $\mathbb{E}\left[|x_l|^2\right] = P$ for $l \in \{1, \dots, K\}$, so the power constraint is met for a normalization factor of N_t, i.e. $\mathbb{E}\left[|c_{t,i}|^2\right] = \frac{P}{N_t}$ for any i, t. It is assumed that the channel remains fixed during the transmission of the codeword $\mathbf{C}^{K,T}$. Thus, we have

$$\mathbf{R} = \mathbf{C}^{K,T}\mathbf{H} + \mathbf{Z}, \tag{1}$$

where $\mathbf{R} \in \mathbb{C}^{T \times N_r}$ represents the received matrix (over receive antennas and over time). Similarly, $\mathbf{Z} \in \mathbb{C}^{T \times N_r}$ is the space-time noise matrix. We denote by $r_{t,i}$ and $z_{t,i}$ the received symbols and the noise at receive antenna i at time t (see Figure 1).

We turn now to describe the link between the combining unit and the detector. The combining unit takes the matrix $\mathbf{R}$ as its input and returns as output a vector $\mathbf{s} \in \mathbb{C}^{\mathbf{T}}$

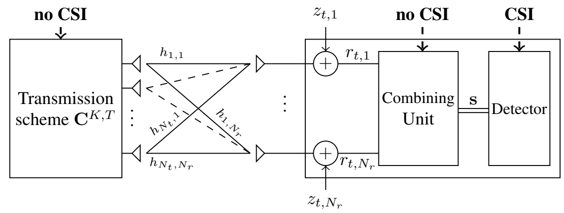

Fig. 1: Combining scheme for $N_t \times N_r$ system where CSI is available only at the detector.

which represents the stream of symbols forwarded to the detector (and used to recover the K information symbols). Hence, the combining unit performs dimension reduction. Of course, in practice, the symbol stream after dimension reduction would also need to be quantized before transmission over the fronthaul link.

We constrain the dimension-reduction operation to be any linear operation over the reals.[2] The combining operation applied to the matrix $\mathbf{R}$ is denoted by $\mathfrak{C}$, such that

$$\begin{aligned} \mathbf{s} = \mathfrak{C}(\mathbf{R}) &= \mathfrak{C}(\mathbf{C}^{K,T}\mathbf{H} + \mathbf{Z}) \\ &\overset{(a)}{=} \mathfrak{C}(\mathbf{C}^{K,T}\mathbf{H}) + \mathfrak{C}(\mathbf{Z}), \end{aligned} \tag{2}$$

where (a) follows since the combining operator is assumed to be widely linear.

We note that the vector $\mathbf{s}$ is of length T so that one information symbol per channel use is conveyed over the frauthaul, i.e. there is no bandwidth (BW) expansion.[3]

We refer to the space-time code along with the combining operation as a *universal transmission-combining scheme*. For a $N_t \times N_r$ channel, such a scheme is characterized by the pair $\{\mathbf{C}^{K,T}, \mathfrak{C}\}_{N_t,N_r}$, were $\mathbf{C}^{K,T}$ is the the STBC applied at the transmitter and where $\mathfrak{C}$ is the combining operation applied at the receiver.

The vector $\mathbf{s}$ is used by the detector to recover the K information symbols where perfect CSI is assumed. For simplicity, in the sequel, we will consider the bit-error rate (BER) of different uncoded transmission schemes.

3. SCHEME EQUIVALENCY

We introduce the concept of *scheme equivalency* which is the main tool we use in this paper. Consider a $N_t \cdot N_r \times 1$ MISO transmission scheme with the STBC transmission of $\hat{\mathbf{C}}^{K,\hat{T}} \in \mathbb{C}^{\hat{T} \times N_{\text{eff}}}$ where we denote $N_{\text{eff}} = N_t \cdot N_r$. The transmission scheme $\hat{\mathbf{C}}^{K,\hat{T}}$ is also comprised of the same indeterminants $\pm x_1, \dots, \pm x_K$ and their conjugates $\pm x_1^*, \dots, \pm x_K^*$ (up to a power normalization factor) so its rate is $\hat{R}_t = \frac{K}{\hat{T}}$. Noting that in the case of a MISO channel, no combining is needed (the combining unit forwards the received symbols as is), we can denote the above scheme by $\{\hat{\mathbf{C}}, \mathfrak{I}\}_{N_{\text{eff}},1}$ where $\mathfrak{I}(\mathbf{R}) = \mathbf{R}$, i.e., forwarding the received symbols without applying any operation on it.

[2] Alternatively, the operations are assumed to be widely linear over the complex field.

[3] We shall refer to this property as “no BW expansion” constraint in the sequel.

For a MISO channel with channel vector $\hat{\mathbf{h}} \in \mathbb{C}^{N_{\text{eff}}}$, the received vector $\hat{\mathbf{r}} \in \mathbb{C}^{N_{\text{eff}}}$ is given by

$$\hat{\mathbf{r}} = \hat{\mathbf{C}}^{K,\hat{T}}\hat{\mathbf{h}} + \hat{\mathbf{z}}, \tag{3}$$

where the noise $\hat{\mathbf{z}} \in \mathbb{C}^{N_{\text{eff}}}$ is i.i.d circularly-symmetric complex Gaussian with unit variance. We denote $\hat{\mathbf{s}} = \hat{\mathbf{r}}$ as the combined vector here is simply the received vector.

Consider now a $N_t \times N_r$ MIMO channel. We define a *conjugate-symmetric reordering transformation* $\mathcal{M}$ as a mapping from $\mathbf{H}$ to $\hat{\mathbf{h}}$, such that each entry of the matrix $\mathbf{H}$ is mapped to a distinct location in the vector $\hat{\mathbf{h}}$, up to conjugation and/or negation. Such a mapping $\mathcal{M}$ is associated with a one-to-one mapping m that maps each pair of indices $(i,j) : 1 \le i \le N_t, 1 \le j \le N_r$ to a distinct index $k : 1 \le k \le N_t \cdot N_r$, i.e we write $m(i,j) = k$ and hence $m^{-1}(k) = (i,j)$. Thus, for any k, we have

$$h_k \overset{\pm *}{=} H_{m^{-1}(k)} \tag{4}$$

where the notation $\overset{\pm *}{=}$ stands for equality with possibly negation and/or conjugation. We denote $\hat{\mathbf{h}} = \mathcal{M}(\mathbf{H})$.

We are interested in such transformations as they do not change the distribution of the elements of $\mathbf{H}$ assuming they are conjugate-symmetrically distributed (e.g., circularly-symmetric complex Gaussian).

Definition 1 (Scheme equivalency)**.** *A universal MIMO transmission-combining scheme* $\{\mathbf{C}^{K,T}, \mathfrak{C}\}_{N_t,N_r}$ *is equivalent to the MISO transmission scheme* $\{\hat{\mathbf{C}}^{K,\hat{T}}, \mathfrak{I}\}_{N_t \cdot N_r,1}$ *if there exists a conjugate-symmetric reordering transformation* $\hat{\mathbf{h}} = \mathcal{M}(\mathbf{H})$ *such that for every realization of* $\mathbf{H}$ *we have* $\mathfrak{C}(\mathbf{C}^{K,T}\mathbf{H}) = \hat{\mathbf{C}}^{K,\hat{T}}\hat{\mathbf{h}}$*, and the resulting noise* $\mathfrak{C}(\mathbf{Z})$ *has the same distribution as* $\hat{\mathbf{z}}$*.*

We denote such equivalency as

$$\{\mathbf{C}^{K,T}, \mathfrak{C}\}_{N_t,N_r} \longleftrightarrow \{\hat{\mathbf{C}}^{K,\hat{T}}, \mathfrak{I}\}_{N_{\text{eff}},1}.$$

Remark 1. *We note that scheme equivalency does not necessarily mean that both transmission schemes have the same symbol rate. Since our goal is to obtain equal-rate scheme equivalency, we will require that* $T = \hat{T}$*.*

Corollary 1. *For any quasi-static channel, two equivalent schemes have the same BER performance (and hence the same diversity order) when the same decoder is used.*

4. A UNIVERSAL TRANSMISSION-COMBINING SCHEME FOR 2×2 MIMO CHANNELS

4.1. Trivial schemes for 2×2 Channels

The simplest scheme for achieving full diversity is to perform repetition: Each information symbol is transmitted four times (twice from each antenna). The receiver forwards to the detector the output of the different antennas in every time slot. While resulting in full diversity and meeting the "no BW expansion" constraint, the effective rate of this scheme is $1/4$.

Applying Alamouti modulation (either at the transmitter while using a single receive antenna or at the receiver while using a single transmit antenna) satisfies the no BW expansion constraint and achieves rate 1. However, the diversity order is 2 (rather than 4).

Conversely, one could achieve full diversity (with no BW expansion) by applying Alamouti coding (either at the transmitter or at the receiver) in conjunction with repetition. But in this case the symbol rate is reduced to $1/2$.

4.2. Universal Combining Scheme for 2×2 Channels

As we have seen, trivial schemes sacrifice either diversity or symbol rate. Our goal now is to overcome this drawback.

To that end, we first introduce the EA-QOSTBC matrix for the 4×1 MISO channel, defined as [13]:

$$\breve{\mathbf{X}}^{(4)} = \frac{1}{\sqrt{4}} \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix}.$$

The scheme is attractive as it is rate 1 (whereas no rate 1 OSTBC exists for four transmit antennas [12]). The structure of the transmission matrix is comprised of two sets of orthogonal columns. Although it does not enjoy the low complexity symbol-wise detection of orthogonal design, it does allow for pairwise detection. Furthermore, if one employs proper symbol rotation precoding (designed for the chosen constellation) to the information symbols vector as proposed by [17], full diversity is achieved.

Building on the EA code, We suggest the following transmission scheme for the 2×2 channel:

$$\mathbf{C}^T = \frac{1}{\sqrt{2}} \begin{bmatrix} x_1 & -x_2^* & x_3^* & -x_4 \\ x_2 & x_1^* & x_4^* & x_3 \end{bmatrix}$$

The combining unit applies the following operations to the received matrix

$$\mathfrak{C}(\mathbf{R}) = \frac{1}{\sqrt{2}} \left[r_{1,1} + r_{3,2}^* \quad r_{2,1} + r_{4,2}^* \quad -r_{3,1} + r_{1,2}^* \quad -r_{4,1} + r_{2,2}^* \right]^T.$$

Note that the combining scheme reduces the 8 symbols of $\mathbf{R}$ to a vector of size 4, the latter being the number of channel uses per transmission. Hence, the no BW expansion constraint is met. Further, it supports a symbol rate of 1.

The resulting combined vector is:

$$\begin{bmatrix} s_1 \\ s_2 \\ s_3 \\ s_4 \end{bmatrix} = \frac{1}{\sqrt{4}} \begin{bmatrix} x_1 & x_2 & x_3 & x_4 \\ -x_2^* & x_1^* & -x_4^* & x_3^* \\ -x_3^* & -x_4^* & x_1^* & x_2^* \\ x_4 & -x_3 & -x_2 & x_1 \end{bmatrix} \cdot \begin{bmatrix} h_{11} \\ h_{21} \\ h_{12}^* \\ h_{22}^* \end{bmatrix} + \frac{1}{\sqrt{2}} \left[z_{1,1} + z_{3,2}^* \quad z_{2,1} + z_{4,2}^* \quad -z_{3,1} + z_{1,2}^* \quad -z_{4,1} + z_{2,2}^* \right]^T.$$

Now, defining the transformation

$$\mathcal{M}\left(\begin{bmatrix} h_{11} & h_{12} \\ h_{21} & h_{22} \end{bmatrix} \right) = \begin{bmatrix} h_{11} & h_{21} & h_{12}^* & h_{22}^* \end{bmatrix}^T = \begin{bmatrix} \hat{h}_1 & \hat{h}_2 & \hat{h}_3 & \hat{h}_4 \end{bmatrix}^T, \tag{5}$$

we obtain that

$$\{\mathbf{C}, \mathfrak{C}\}_{2,2} \longleftrightarrow \{\breve{\mathbf{X}}^{(4)}, \mathfrak{I}\}_{4,1}.$$

Therefore, by Corollary 1, we conclude that this scheme has full diversity and rate 1, while meeting the no BW expansion constraint on the fronthaul link.

5. UNIVERSAL TRANSMISSION-COMBINING FOR $N_T \times N_R$ CHANNELS

We generalize the universal transmission-combining scheme described in Section 5 to $N_t \times N_r$ channels satisfying that $N_t = 2^p$ and $N_r = 2^q$ are integer powers of 2. We denote by $\breve{\mathbf{X}}^{(N)}$ the generalized EA-QOSTBC for N transmit antenna as defined recursively in [17]. For convenience, we denote $\mathbf{X}^{(p)} \triangleq \breve{\mathbf{X}}^{(2^p)}$. The proposed scheme, denoted by $\{\mathbf{C}^{(p,q)}, \mathfrak{C}\}_{N_t,N_r}$, is as follows. The information symbols are split into two groups, denoted by $\mathbf{x} = \{x_1, \cdots, x_{\frac{N}{2}}\}$ and $\mathbf{y} = \{x_{\frac{N}{2}+1}, \cdots, x_N\}$.

The transmission scheme is defined by the following recursion:

$$\begin{aligned} \mathbf{C}^{(p,q)} &= \mathbf{X}^{(p)} && \text{for } q = 0,\ p \geq 0 \\ \mathbf{C}^{(p,q)} &= \left[\begin{array}{c} \mathbf{C}^{(p,q-1)}_{\{\mathbf{x}\}} \\ \hline \mathbf{C}^{(p,q-1)*}_{\{\mathbf{y}\}} \end{array}\right] && \text{for } q > 0,\ p \geq 0, \end{aligned} \quad (6)$$

where $\mathbf{C}^{(p,q-1)}_{\{\mathbf{x}\}}$ and $\mathbf{C}^{(p,q-1)}_{\{\mathbf{y}\}}$ are the matrices obtained by placing in the transmission matrix of the $2^p \times 2^{q-1}$ MIMO channel the sets $\mathbf{x}$ and $\mathbf{y}$ correspondingly. The induced transmission rate is 1 for any $p, q > 0$.

The combining scheme is defined over the matrix of received symbols $\mathbf{R}$, where $\mathbf{R}$ has dimensions of $T \times 2^q$. Matrix $\mathbf{R}$ can be written as follows

$$\mathbf{R} = \left[\begin{array}{c|c} \mathbf{A} & \mathbf{B} \\ \hline \mathbf{C} & \mathbf{D} \end{array}\right]$$

where each sub-matrix $(\mathbf{A},\ \mathbf{B},\ \mathbf{C},\ \mathbf{D})$ is a $T/2 \times 2^{q-1}$ matrix. The combining scheme is defined recursively as

$$\begin{aligned} \mathfrak{C}(\mathbf{R}) &= \mathbf{R} && \text{for } q = 0,\ p > 0 \\ \mathfrak{C}(\mathbf{R}) &= \frac{1}{\sqrt{2}} \left[\begin{array}{c} \mathfrak{C}(\mathbf{A}) + \mathfrak{C}(\mathbf{D})^* \\ \hline -\mathfrak{C}(\mathbf{C}) + \mathfrak{C}(\mathbf{B})^* \end{array}\right] && \text{for } q > 0,\ p > 0. \end{aligned} \quad (7)$$

The recursion ends when the interim input is a column vector with dimensions $2^p \times 1$. A code that generates the universal transmission and combining schemes can be found in [25].

Proposition 1. *The following scheme equivalency*

$$\{\mathbf{C}^{(p,q)}, \mathfrak{C}\}_{N_t,N_r} \longleftrightarrow \{\breve{\mathbf{X}}^{(N_{\text{eff}})}, \mathfrak{I}\}_{N_t \cdot N_r \times 1}, \quad (8)$$

holds where $\mathbf{C}^{(p,q)}$ is defined in (6) and $\mathfrak{C}$ in (7).

Proof. Due to space limitations the proof is omitted. It Can be found in [26]. □

Thus, the proposed transmission-combining scheme has the same rate (1), same diversity (full diversity) and same decoding complexity as that of the QOSTBC.

6. SIMULATIONS

The simulations assume i.i.d. Rayleigh fading. Figure 2 depicts the results for a 2×2 channel. As performance benchmarks, we consider the following schemes: "standard Alamouti" (applied either at the transmitter or receiver, making use only of one antenna at the other side) and a "double BW" scheme, in which the combining unit forwards the symbols received at the two antennas (violating the no BW expansion constraint).

The proposed scheme from Section 4.2 is simulated under three scenarios: QPSK constellation and zero forcing (ZF) detection, QPSK constellation and ML detection, and rotated QPSK and ML detection. The first enjoys a simple low complexity symbol-wise detector and outperforms the "standard Alamouti" scheme by approximately 1 dB. The second is slightly more complicated but yields an improvement of approximately 2 dB w.r.t to the "transmit Alamouti", but yet is not a full-diversity scheme. Using rotated QPSK leads to a full-diversity scheme and its bit error ratio (BER) curve slope matches that of the "double BW" scheme, albeit, with a degradation of approximately 3 dB.

Figure 3 depicts the performance of the proposed scheme for more general MIMO systems. The symbol constellation is taken with suitable rotation such that full diversity is achieved.

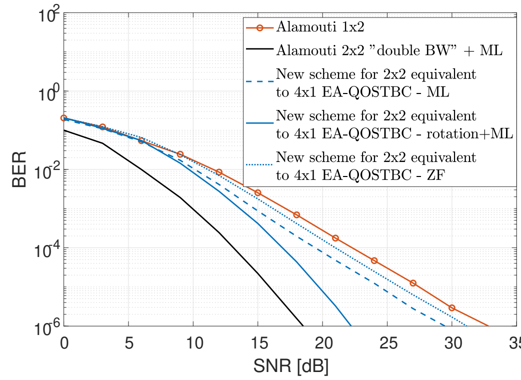

Fig. 2: BER performance of different schemes and detection methods using QPSK constellation

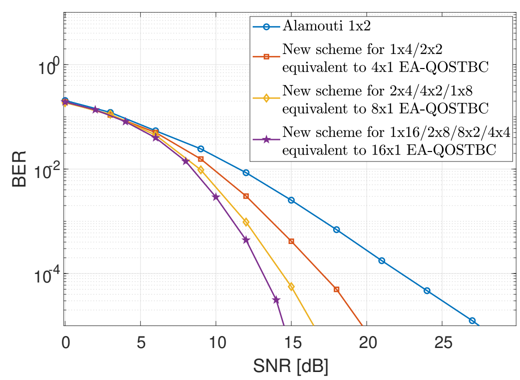

Fig. 3: BER performance of different universal transmission-combining schemes for QPSK with ML detection.

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