Novel Transmit Precoding Methods for Rayleigh Fading Multiuser TDD-MIMO Systems with CSIT and no CSIR

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***Abstract*—In this paper, we present novel precoding methods for multiuser Rayleigh fading MIMO systems, when channel state information (CSI) is available at the transmitter (CSIT), but not at the receiver (CSIR). Such a scenario is relevant, for example, in time division duplex (TDD) MIMO communications, where, due to channel reciprocity, CSIT can be acquired directly by sending a training sequence from the receiver to the transmitter(s). We propose three transmit precoding schemes that convert the fading MIMO channel into a fixed-gain AWGN channel, while satisfying an average power constraint. We also extend one of the precoding schemes to the multiuser Rayleigh fading Multiple Access Channel (MAC), Broadcast Channel (BC) and Interference Channel (IC). The proposed schemes convert the fading MIMO channel into fixed-gain parallel AWGN channels in all three cases. Hence, they achieve an infinite diversity order, which is in sharp contrast with schemes based on perfect CSIR and no CSIT, which at best achieve a finite diversity order. Further, we show that a polynomial diversity order is retained, even in the presence of channel estimation errors at the transmitter. Monte Carlo simulations illustrate the BER performance obtainable from the proposed precoding scheme compared to existing transmit precoding schemes.**

***Index Terms*—Transmit precoding, diversity order, multiuser MIMO systems, time division duplex communication.**

I. Introduction

Transmit precoding methods utilize the knowledge of channel state information (CSI) at the transmitter to maximize the signal to noise ratio (SNR) at the desired receiver, minimize the interference at unintended receivers, and also equalize or mitigate the effect of fading in the wireless channel. In environments such as in vehicular communications, achieving these goals simultaneously is particularly challenging, since the channel estimation and precoded data transmission phases have to complete within the relatively short coherence time of the channel. Moreover, it is of interest to find precoding methods that are applicable to all types of multiuser channels, such as multiple access channel (MAC), broadcast channel (BC) and interference channel (IC); and this is the focus of this work.

The Diversity-Multiplexing gain Tradeoff (DMT) of the fading Multiple-Input Multiple-Output (MIMO) communication system with Channel State Information (CSI) at the receiver (CSIR) was exactly characterized in the seminal paper by Zheng and Tse [2]. A key finding in this work is that, for Rayleigh fading MIMO channels with perfect CSIR, the maximum diversity order can at most be $N_r N_t$, where N_r and N_t denote the number of antennas at the receiver and transmitter, respectively. Since that early result, DMT has been extended to various cases with full/partial knowledge of CSI at the transmitter (CSIT) and receiver (CSIR), and various multiuser channels. However, transmit diversity schemes, and the corresponding achievable DMT, of a fading MIMO channel with CSI available only at the transmitter and no CSIR has received relatively little attention in the literature. This is perhaps because the acquisition of CSIT has typically been viewed as a two-stage process: CSI is first acquired at the receiver using a known training sequence in the forward-link direction, and then fed back from the receiver to the transmitter over the reverse-link in a quantized or analog fashion. Thus, the existing studies inherently assume an initial estimation of CSI at the receiver. However, when the channel is *reciprocal*, i.e., when the forward and reverse channels are the same, CSI can be directly acquired at the transmitter by sending a known training sequence in the reverse-link direction. The channel can be modeled as being reciprocal, for example, in Time Division Duplex (TDD) communication systems [3]–[7].[1] This reciprocal nature of the channel opens up the possibility of acquiring CSI only at the transmitter without needing to first estimate the channel at the receiver, by sending a known training signal from the receiver to the transmitter. For example, if a user station (STA) wishes to send data to a wireless access point (AP), the STA could send a training signal followed by the data, or the AP could send a training signal to the STA, followed by data from the STA to the AP. In the former case, CSI is available only at the AP (the receiver), while in the latter case, CSI is available only at the STA (the transmitter).

In this context, some important questions that we seek to answer in this work are: If perfect CSI is available only at the transmitter, what is the best diversity order that can be achieved? How does it compare with the diversity order that can be obtained when perfect CSI is available only at the receiver? Can such schemes be used in multiuser fading channels, such as the BC, MAC and IC? We answer these questions by proposing novel transmission schemes based on CSIT and no CSIR. The proposed schemes are fundamentally different from techniques such as Maximum Ratio Transmission (MRT), Zero Forcing (ZF) precoding and channel inver-

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This work will appear in part in [1].

[1]Note that channel reciprocity also requires that the transmit and receive radio-frequency (RF) chains are well-calibrated, which is assumed here [8].

sion based power control in that the MRT and ZF precoding only achieves a finite diversity order, while channel inversion does not satisfy an average power constraint at the transmitter under Rayleigh fading for all antenna configurations [9]. We show that our proposed schemes can convert a Rayleigh fading MIMO channel into fixed-gain parallel AWGN channels, while simultaneously satisfying an average power constraint at the transmitter. Further, we show that the schemes elegantly extend to the fading MAC, BC and IC, and achieve an infinite diversity order in all three cases.

We start with a survey of the relevant literature. As already mentioned, a diversity order of $N_t N_r$ can be achieved with perfect CSIR [2]. This was extended to the MAC channel in [10]. On the other hand, an exponential diversity order[2] can be achieved when perfect CSI is available at *both* transmitter and receiver [11]. It is also known that under partial CSIT and perfect CSIR, a diversity order greater than $N_t N_r$ can be achieved in the single-user case (See [12]–[14], and [7-18] in [15]). Similar results have been obtained in multiuser scenarios with and without partial CSIT (See, e.g., [16]–[20]). In [11], it is shown that polynomial diversity order can be obtained at high SNR region, when imperfect CSIT and CSIR are available. For obtaining the result, it was required to make the assumption that the receiver also have, as side-information, the noisy CSI that is available at the transmitter. In the above papers, the available CSIT is exploited either for inverting dominant modes, or for power control, to improve the diversity order, under the assumption of perfect CSIR. In [9], [21]–[23], channel-inversion based power control and precoding was considered without CSIR. However, channel inversion either fails to satisfy the average power constraint [9], or requires the use of regularization [21] or computationally intensive sphere encoding schemes [22], [23]. Multiuser precoding based on vector perturbation was studied for the broadcast channel in [24].

Our contributions in this paper are as follows. We develop precoding schemes based on CSIT, that work for both single-user and multiuser fading channels such as the MAC, BC, and IC. In the single-user case, we propose three novel and simple-to-implement transmit precoding schemes which require CSI only at the transmitter (see Sec. II). We show that our proposed transmit precoding schemes achieve an infinite diversity order, while satisfying an average transmit power constraint. Added benefits of our proposed approach are that forward-link training is not required, and optimal decoding at the receiver is very simple. In Appendix A, we show that the average power constraint can be satisfied, and a polynomial diversity order can be obtained, even in the presence of channel estimation errors at the transmitter.

We extend the precoding schemes to three kinds of multiuser Rayleigh fading channels: the Multiple Access Channel (MAC), Broadcast Channel (BC) and Interference Channel (IC) (see Sec. III). We show that, in all three cases, the proposed schemes can convert a fading multiuser channel to a fixed-gain multiuser channel, thereby achieving an infinite diversity order, while satisfying an average power constraint.

In Sec. IV, we illustrate the performance of the proposed schemes via Monte Carlo simulations. We show that the probability of error versus SNR curves exhibit the AWGN-like waterfall behavior in both single-user and multiuser scenarios. We demonstrate a significant improvement in performance compared to existing schemes such as space-time block coding, zero forcing, and maximum ratio transmission schemes. We also present simulation results with imperfect CSIT obtained using reverse-link training, as well as with practical peak-to-average power constraints. The results show that, for practical SNRs, the waterfall behavior is still retained. We offer some concluding remarks in Sec. V.

Notation: We use boldface capital letters to denote matrices and boldface small letters to denote vectors. $\mathbf{X}^T, \mathbf{X}^H, \text{tr}(\mathbf{X})$ denote the transpose, hermitian and trace of $\mathbf{X}$, respectively. The ℓ_2 norm of $\mathbf{x}$ is denoted by $\|\mathbf{x}\|$. We denote the real part and absolute value of a complex number c by $\Re\{c\}$ and $|c|$, respectively. We use $\mathcal{N}(0,1)$ (and $\mathcal{CN}(0,1)$) to denote a real (and complex circularly symmetric) Gaussian random variable with zero mean and unit variance. The expectation of $f(\mathbf{x})$ with respect to a random variable $\mathbf{x}$ is denoted by $\mathbb{E}_{\mathbf{x}}[f(\mathbf{x})]$.

In the next section, we describe the proposed precoding schemes in the single-user case. We extend the schemes to the multiuser cases in Sec. III.

II. Transmit Precoding with CSIT and no CSIR: Single-User Case

A. Modified Maximum Ratio Transmission Based Precoding

In this subsection, we present our first precoding scheme, which is based on Maximum Ratio Transmission (MRT) [25]. A key difference between MRT and the scheme proposed below is that our scheme incorporates a different form of power control, that enables one to achieve an infinite diversity order with Rayleigh fading channels, while satisfying an average power constraint at the transmitter. We start with the case of a single receive antenna and $N_t \geq 2$ transmit antennas. Let $\mathbf{h}$ denote the $N_t \times 1$ Rayleigh fading channel vector, with independent and identically distributed (i.i.d.) $\mathcal{CN}(0,1)$ components in the complex baseband representation. In classical MRT, one uses $\mathbf{h}/\|\mathbf{h}\|$ as the beamforming vector at the transmitter. Here, we propose to use $\mathbf{p} \triangleq \mathbf{h}/\|\mathbf{h}\|^2$ to precode the unit-power data symbol x. The received signal, $y \in \mathbb{C}$, can be written as

$$y = \sqrt{\frac{k\rho}{N_t}}\mathbf{h}^H\mathbf{p}x + n = \sqrt{\frac{k\rho}{N_t}}x + n, \tag{1}$$

where $n \in \mathbb{C}$ denotes the receiver noise, distributed as $\mathcal{CN}(0,1)$, and k denotes a normalization constant. The average transmitted power can be written as

$$P_{\text{avg}} = \frac{k\rho}{N_t}\mathbb{E}[x^2]\mathbb{E}[\mathbf{p}^H\mathbf{p}] = \frac{k\rho}{N_t}\mathbb{E}\left[\frac{1}{\|\mathbf{h}\|^2}\right]. \tag{2}$$

It is straightforward to show that $\mathbb{E}\left[1/\|\mathbf{h}\|^2\right] = 1/(N_t-1)$ for $N_t \geq 2$ [26], [27]. Hence, the power normalization constant $k = N_t(N_t-1)$ satisfies the average power constraint

[2] A transmission scheme is said to achieve an exponential diversity order of β if the average probability of error decreases with SNR as $\mathcal{O}(\exp(-\beta\ \text{SNR}))$ [11]. Hence, all schemes that achieve a finite diversity order have an exponential diversity order of zero.

$P_{\text{avg}} = \rho$. This average power constraint is the same as in past work that considers transmission schemes with CSIT, e.g., [11], [12], [14], [15]. The corresponding SNR at the receiver is $\text{SNR} = (N_t - 1)\rho$. Thus, the above modified MRT based precoding scheme fully equalizes the fading channel and achieves an infinite diversity order.

Remark 1. *In Appendix A, we analyze the diversity order of this method, when the precoding vector is computed from imperfect CSI at the transmitter, due to channel estimation errors. We show that a polynomial diversity order of at least $N_t - 1$ is retained, even under imperfect CSIT.*

B. QR-Decomposition Based Precoding Scheme

In this subsection, we consider a Rayleigh fading MIMO channel with N_r receive antennas and $N_t \geq 2N_r$ transmit antennas. The complex baseband signal model for the received signal at the i^{th} receive antenna can be written as

$$y_i = \sqrt{\frac{k\rho}{N_t}}\mathbf{h}_i^H \mathbf{P}\tilde{\mathbf{x}} + n_i, \tag{3}$$

where $\mathbf{h}_i \in \mathbb{C}^{N_t}$ denotes the complex channel coefficients between the N_t transmit antennas and i^{th} receive antenna, $\tilde{\mathbf{x}} = [\tilde{x}_1, \tilde{x}_2, \ldots, \tilde{x}_{N_t}]^T$ denotes an *extended* data vector of dimension N_t, and is derived from a data vector $\mathbf{x} \in \mathbb{C}^{N_r}$ containing the N_r symbols to be transmitted. We assume the normalization $\mathbb{E}[\mathbf{x}^H\mathbf{x}] = 1$. Also, ρ, k and $\mathbf{P}$ denote, respectively, the average transmit power available across the N_t transmit antennas per channel use, a normalization constant, and an $N_t \times N_t$ precoding matrix. The noise is assumed to be i.i.d. across receive antennas with entries from $\mathcal{CN}(0, 1)$.

For ease of presentation, as in the previous subsection, we start with the $N_r = 1$ case. Let $\mathbf{h} \in \mathbb{C}^{N_t}$ denote the channel vector. We set the precoding matrix $\mathbf{P}$ as

$$\mathbf{P} = \mathbf{QU}, \tag{4}$$

where the unitary matrix $\mathbf{Q} \in \mathbb{C}^{N_t \times N_t}$ is obtained from the QR-decomposition of $\mathbf{h}$, i.e., $\mathbf{h} \triangleq \mathbf{Qr}$, with $\mathbf{r} \in \mathbb{C}^{N_t}$ and upper triangular, with first element $r_1 = \|\mathbf{h}\|$ and remaining elements equal to zero. Also, $\mathbf{U} \in \mathbb{C}^{N_t \times N_t}$ is chosen to be an arbitrary, non-diagonal unitary matrix.

Now, given the complex scalar data symbol x, the extended data vector $\tilde{\mathbf{x}}$ is chosen such that the following condition is satisfied:

$$\mathbf{r}^H\mathbf{U}\tilde{\mathbf{x}} = x. \tag{5}$$

It is easy to verify that (5) can be satisfied by choosing $\tilde{x}_1 = x$, $\tilde{x}_2 = x(1 - r_1 u_{1,1})/(r_1 u_{1,2})$, and $\tilde{x}_j = 0$, for $j = 3, \ldots, N_t$, where $u_{i,j}$ is the $(i, j)^{\text{th}}$ element of $\mathbf{U}$, provided $u_{1,2} \neq 0$. Substituting for $\tilde{\mathbf{x}}$ and $\mathbf{P}$, the above precoding scheme leads to the following equivalent channel:

$$y = \sqrt{\frac{k\rho}{N_t}}x + n. \tag{6}$$

In the above, k is a normalization constant independent of the channel instantiation, and its value is specified below. Thus, the proposed precoding scheme inherently equalizes the effect of fading and also cancels the interference caused due to the signal being transmitted from multiple antennas, thereby converting the fading channel into a fixed-gain AWGN channel. Now, we show that, with k appropriately chosen, the above precoding scheme satisfies an average transmit power constraint. The average transmitted power can be written as

$$\begin{aligned} P_{\text{avg}} &= \frac{k\rho}{N_t}\mathbb{E}_{x,\mathbf{h}}\left[\tilde{\mathbf{x}}^H\mathbf{P}^H\mathbf{P}\tilde{\mathbf{x}}\right], \\ &= \frac{k\rho}{N_t}\mathbb{E}_x\left[x^2\right]\left(1 + \frac{|u_{1,1}|^2}{|u_{1,2}|^2} + \frac{1}{|u_{1,2}|^2}\mathbb{E}_{\mathbf{h}}\left[\frac{1}{\|\mathbf{h}\|^2}\right]\right. \\ &\qquad \left. -2\frac{\Re\{u_{1,1}\}}{|u_{1,2}|^2}\mathbb{E}_{\mathbf{h}}\left[\frac{1}{\|\mathbf{h}\|}\right]\right). \end{aligned} \tag{7}$$

For Rayleigh fading channels, it is known that [26], [27]

$$\begin{aligned} \mathbb{E}\left[\frac{1}{\|\mathbf{h}\|^2}\right] &= \frac{1}{N_t - 1}, \\ \mathbb{E}\left[\frac{1}{\|\mathbf{h}\|}\right] &= \frac{\Gamma\left(\frac{2N_t - 1}{2}\right)}{\Gamma\left(N_t\right)}, \end{aligned} \tag{8}$$

where $\Gamma(\cdot)$ denotes the Gamma function [28]. Using (7) and (8), we can satisfy the average transmit power constraint of $P_{\text{avg}} = \rho$ by choosing

$$k = N_t\left(1 + \frac{|u_{1,1}|^2}{|u_{1,2}|^2} + \frac{1}{|u_{1,2}|^2} - 2\frac{\Re\{u_{1,1}\}}{|u_{1,2}|^2}\frac{\Gamma\left(\frac{2N_t - 1}{2}\right)}{\Gamma(N_t)}\right)^{-1}, \tag{9}$$

where $u_{1,2} \neq 0$ is chosen so that $k > 0$ and finite. For example, when $N_t = 2$, one can choose

$$\mathbf{U} = \begin{bmatrix} 0 & 1 \\ 1 & 0 \end{bmatrix}, \tag{10}$$

which results in $k = 1$ and a received SNR of 0.5ρ, i.e., a 3 dB loss compared to the unit-gain AWGN channel. Finding the unitary matrix $\mathbf{U}$ that minimizes the SNR loss is an interesting extension for future work.

In the following, we extend the above QR-based precoding scheme to the case where multiple receive antenna chains are available.

When the receiver is equipped with N_r antennas, with $N_t \geq 2N_r$, our proposed extension leads to N_r parallel, non-interfering AWGN channels. The channel input-output relation is given by

$$\mathbf{y} = \sqrt{\frac{k\rho}{N_t}}\mathbf{H}^H\mathbf{P}\tilde{\mathbf{x}} + \mathbf{n}, \tag{11}$$

where the received vector $\mathbf{y} \in \mathbb{C}^{N_r}$, the channel matrix $\mathbf{H} \in \mathbb{C}^{N_t \times N_r}$, and the noise $\mathbf{n} \in \mathbb{C}^{N_r}$. Denote the QR decomposition of $\mathbf{H}$ by $\mathbf{H} = \mathbf{QR}$, where $\mathbf{Q} \in \mathbb{C}^{N_t \times N_t}$ is unitary and $\mathbf{R} \in \mathbb{C}^{N_t \times N_r}$ is upper triangular. Note that, since $N_t \geq 2N_r$, the rows $N_r + 1$ through N_t of the matrix $\mathbf{R}$ are all zeros.

We consider the data vector $\mathbf{x} = [x_1,\ x_2,\ \ldots,\ x_{N_r}]^T$, and choose the extended vector $\tilde{\mathbf{x}}$ such that $\mathbf{R}^H\mathbf{U}\tilde{\mathbf{x}} = \mathbf{x}$, where $\mathbf{U} \in \mathbb{C}^{N_t \times N_t}$ is a fixed non-diagonal unitary matrix. Now, the matrix $\mathbf{R}$ can be partitioned as $\mathbf{R} = [\mathbf{R}_1^H\ \mathbf{0}^H]^H$, where the submatrices $\mathbf{R}_1$ and $\mathbf{0}$ are of dimension $N_r \times N_r$ and $(N_t - N_r) \times N_r$, respectively. We set the first N_r entries of

$\tilde{\mathbf{x}}$ as $\mathbf{x}$. If we partition $\tilde{\mathbf{x}}$ as $\tilde{\mathbf{x}}^H = [\mathbf{x}^H \ \mathbf{x}'^H \ \mathbf{0}^H]$, where $\mathbf{0}$ is a vector of $(N_t - 2N_r)$ zeros, $\mathbf{x}'$ can be written as

$$\mathbf{x}' = \mathbf{R}_{u2}^{-1} \left(\mathbf{I} - \mathbf{R}_{u1}\right) \mathbf{x}, \tag{12}$$

where $\mathbf{R}_{u1} \triangleq \mathbf{R}_1^H \mathbf{U}_{11}$ and $\mathbf{R}_{u2} \triangleq \mathbf{R}_1^H \mathbf{U}_{12}$. The matrix $\mathbf{U}_{11}$ is the $N_r \times N_r$ principal submatrix of $\mathbf{U}$, and the matrix $\mathbf{U}_{12}$ is the $N_r \times N_r$ submatrix of $\mathbf{U}$ obtained by taking the entries from rows 1 through N_r and columns $N_r + 1$ through $2N_r$. Finally, we let $\mathbf{P} = \mathbf{QU}$, as before.

The above described precoding scheme leads to the input-output relation:

$$\mathbf{y} = \sqrt{\frac{k\rho}{N_t}}\mathbf{x} + \mathbf{n}, \tag{13}$$

and hence, we obtain N_r parallel, fixed-gain AWGN channels. By choosing k appropriately, we can satisfy the average power constraint on the data signal, as we show next.

Noting that $\|\tilde{\mathbf{x}}\|^2 = \|\mathbf{x}\|^2 + \|\mathbf{x}'\|^2$, the average transmit power per channel use can be computed from

$$\begin{aligned} P_{\text{avg}} &= \frac{k\rho}{N_t}\mathbb{E}_{\mathbf{x},\mathbf{h}}\left[\mathbf{x}^H\mathbf{x} + \mathbf{x}^H(\mathbf{I} - \mathbf{R}_{u1})^H\mathbf{R}_{u2}^{-H}\mathbf{R}_{u2}^{-1} \right. \\ &\qquad \left. \times(\mathbf{I} - \mathbf{R}_{u1})\mathbf{x}\right], \\ &= \frac{k\rho}{N_t}\left[1 + \frac{1}{N_r}\text{tr}\left(\mathbb{E}_{\mathbf{h}}\left[(\mathbf{I} - \mathbf{R}_{u1})^H\mathbf{R}_{u2}^{-H}\mathbf{R}_{u2}^{-1} \right.\right.\right. \\ &\qquad \left.\left.\left. \times(\mathbf{I} - \mathbf{R}_{u1})\right]\right)\right]. \end{aligned} \tag{14}$$

Since the choice of the unitary matrix $\mathbf{U}$ is arbitrary, we can simply choose $\mathbf{U}_{11} = \mathbf{0}_{N_r}$ and $\mathbf{U}_{12} = \mathbf{I}_{N_r}$. Now, we get $\mathbf{R}_{u1} = \mathbf{0}$ and $\mathbf{R}_{u2} = \mathbf{R}_1^H$. Further, using Lemma 6 in [26], we have

$$\text{tr}\left(\mathbb{E}_{\mathbf{h}}\left[\mathbf{R}_{u2}^{-1}\mathbf{R}_{u2}^{-H}\right]\right) = \frac{N_r}{N_t - N_r}. \tag{15}$$

Hence, we can simplify the average transmit power as

$$P_{\text{avg}} = \frac{k\rho}{N_t}\left[1 + \frac{1}{N_t - N_r}\right] \tag{16}$$

and k can be chosen as

$$k = N_t\left[1 + \frac{1}{N_t - N_r}\right]^{-1} \tag{17}$$

to satisfy the average transmit power constraint of $P_{\text{avg}} = \rho$. The SNR per receive antenna for this scheme is given by

$$\text{SNR} = \frac{\rho(N_t - N_r)}{(1 + N_t - N_r)}. \tag{18}$$

The pseudo-code of the algorithm for the case when $\mathbf{U}_{11} = \mathbf{0}_{N_r}$ and $\mathbf{U}_{12} = \mathbf{I}_{N_r}$ is shown in Algorithm 1.

C. Transmit Precoding Based on Real O-STBC Signaling

For our third proposed scheme, we consider real O-STBC signaling. At the receiver, we consider the real part of the baseband received signal, and, hence, we can consider both the baseband equivalent channel as well as the additive noise as having real-valued components. Mathematically, the received baseband signal at the i^{th} receive antenna can be written as:

$$\mathbf{y}_i = \sqrt{\frac{k\rho}{N_t}}\mathbf{X}\mathbf{h}_i + \mathbf{n}_i, \tag{19}$$

Algorithm 1 QR-based Precoding Algorithm

Inputs: Channel matrix $\mathbf{H} \in \mathbb{C}^{N_t \times N_r}$, data vector $\mathbf{x} \in \mathbb{C}^{N_r}$, unitary $\mathbf{U} \in \mathbb{C}^{N_t \times N_t}$, transmit power ρ.
Outputs: Transmit signal $\mathbf{s} \in \mathbb{C}^{N_t}$
Start

Compute QR decomposition $\mathbf{H} = \mathbf{QR}$
Partition $\mathbf{R} = [\mathbf{R}_1^H \ \mathbf{0}_{(N_t-N_r)\times N_r}^H]^H$,
where $\mathbf{R}_1 \in \mathbb{C}^{N_r \times N_r}$.
Compute $\mathbf{x}' = \mathbf{R}_1^{-H}\mathbf{x}$,
and $\tilde{\mathbf{x}}^H = [\mathbf{x}^H \ \mathbf{x}'^H \ \mathbf{0}_{N_t-2N_r}^H]$
Compute $\mathbf{P} = \mathbf{QU}$
Compute $k = N_t(N_t - N_r)/(N_t - N_r + 1)$
Compute transmit signal $\mathbf{s} = \sqrt{\frac{k\rho}{N_t}}\mathbf{P}\tilde{\mathbf{x}}$

End

where $\mathbf{y}_i \in \mathbb{R}^L$ denotes the received signal vector for $L \geq N_t$ consecutive symbols. The channel vector between the transmit antennas and the i^{th} receive antenna is denoted by $\mathbf{h}_i \in \mathbb{R}^{N_t}$, and is assumed to have Gaussian i.i.d. entries with zero mean and unit variance, denoted by $\mathcal{N}(0,1)$. The real O-STBC codeword is denoted by $\mathbf{X} \in \mathbb{R}^{L \times N_t}$. The noise vector is denoted by $\mathbf{n}_i \in \mathbb{R}^L$, and is assumed to have i.i.d. $\mathcal{N}(0,1)$ entries. Also, ρ is the total transmit power available across the N_t antennas per channel use, and k is a constant used to meet the average transmit power constraint. Using the equivalent representation of the codeword matrix $\mathbf{X}$ in terms of its constituent Hurwitz-Radon matrices [29], it is shown in [30] that (19) can be written as

$$\mathbf{y}_i = \sqrt{\frac{k\rho}{N_t}}\tilde{\mathbf{H}}_i\mathbf{x} + \mathbf{n}_i, \tag{20}$$

where $\tilde{\mathbf{H}}_i \in \mathbb{R}^{L \times L}$ denotes the equivalent channel matrix and the vector $\mathbf{x} \in \mathbb{R}^L$ contains the symbols used to construct $\mathbf{X}$. Note that, $\tilde{\mathbf{H}}_i$ is obtained from $\mathbf{h}_i$ using a simple mapping $\pi : \mathbb{R}^{N_t} \rightarrow \mathbb{R}^{L \times L}$ [30].

For example, consider the 4×4 real O-STBC code designed in [29]. In this case, it can be shown that

$$\begin{aligned} \mathbf{X}^T &= \begin{bmatrix} s_1 & -s_2 & -s_3 & -s_4 \\ s_2 & s_1 & s_4 & -s_3 \\ s_3 & -s_4 & s_1 & s_2 \\ s_4 & s_3 & -s_2 & s_1 \end{bmatrix}, \\ \tilde{\mathbf{H}} &= \begin{bmatrix} h_1 & h_2 & h_3 & h_4 \\ h_2 & -h_1 & h_4 & -h_3 \\ h_3 & -h_4 & -h_1 & h_2 \\ h_4 & h_3 & -h_2 & -h_1 \end{bmatrix}, \end{aligned} \tag{21}$$

where s_j denotes the j^{th} data symbol drawn from a finite size constellation, $\mathbf{x} = [s_1 \ s_2 \ s_3 \ s_4]^T$, and $h_j = h_{ji}, j = 1, 2, \ldots, N_t$ are the channel coefficients between the N_t transmit antennas and i^{th} receive antenna. Here, for simplicity, we have omitted the receive antenna index i in writing the expression for $\tilde{\mathbf{H}}$.

It is easy to see that the matrix $\tilde{\mathbf{H}}$ is orthogonal. In fact, this property is true for all real O-STBCs. By the equivalence of the two representations, we have $\mathbf{Xh} = \tilde{\mathbf{H}}\mathbf{x}$. Multiplying by $\mathbf{X}^T$ on both sides, we get

$$\beta\mathbf{h} = \mathbf{X}^T\tilde{\mathbf{H}}\mathbf{x}$$

where $\mathbf{X}^T\mathbf{X} = \beta\mathbf{I}_{N_t}$, and $\beta = \sum_{i=1}^{N_t} s_i^2 > 0$, since $\mathbf{X}$ is a real O-STBC codeword. Also, $\mathbf{I}_{N_t}$ represents the $N_t \times N_t$ identity matrix. Now, suppose $\mathbf{h} \neq \mathbf{0}$, but the columns of $\tilde{\mathbf{H}}$ are linearly dependent. Then, there exists a nonzero $\mathbf{x}$ that lies in the null space of $\tilde{\mathbf{H}}$, and substituting such an $\mathbf{x}$ in the above leads to $\mathbf{h} = \mathbf{0}$, i.e., a contradiction. Hence, any nonzero channel vector $\mathbf{h}$ leads to an $\tilde{\mathbf{H}}$ with full column rank. Next, we show that $\tilde{\mathbf{H}}$ is orthogonal.

Let $\mathbf{x}_1$ and $\mathbf{x}_2$ denote two data vectors and $\mathbf{X}_1$ and $\mathbf{X}_2$ denote their corresponding O-STBC matrices. Further, let $\mathbf{x}_{k,j}$ denote the j^{th} column of $\mathbf{X}_k$, for $k = 1, 2$. Due to the structure of O-STBC codes, $\mathbf{x}_{2,j}^T\mathbf{x}_{1,i} = -\mathbf{x}_{2,i}^T\mathbf{x}_{1,j}$ for $j \neq i$, and $\mathbf{x}_{1,i}^T\mathbf{x}_{2,i} = \mathbf{x}_{1,j}^T\mathbf{x}_{2,j} = \mathbf{x}_1^T\mathbf{x}_2$ [31]. Hence,

$$\begin{aligned}\mathbf{h}^T\mathbf{X}_1^T\mathbf{X}_2\mathbf{h} &= \sum_i\sum_j h_ih_j\mathbf{x}_{1,i}^T\mathbf{x}_{2,j} \\ &= \sum_i h_i^2\mathbf{x}_{1,i}^T\mathbf{x}_{2,i} = \mathbf{x}_1^T\mathbf{x}_2\sum_i h_i^2. \quad (22)\end{aligned}$$

Using the fact that $\mathbf{X}\mathbf{h} = \tilde{\mathbf{H}}\mathbf{x}$, we get

$$\mathbf{x}_1^T\left(\tilde{\mathbf{H}}^T\tilde{\mathbf{H}}\right)\mathbf{x}_2 = \mathbf{x}_1^T\mathbf{x}_2\sum_i h_i^2. \quad (23)$$

The above equation holds for any pair of vectors $\mathbf{x}_1$ and $\mathbf{x}_2$, if and only if $\tilde{\mathbf{H}}$ is orthogonal and $\tilde{\mathbf{H}}^T\tilde{\mathbf{H}} = \tilde{\mathbf{H}}\tilde{\mathbf{H}}^T = (\sum_{i=1}^{N_t} h_i^2)\mathbf{I}_L$. Thus, the equivalent channel matrix $\tilde{\mathbf{H}}$ is an orthogonal matrix.

1) Proposed Transmit Precoding Scheme: As before, we first consider the single receive antenna case. We premultiply the data vector $\mathbf{x}$ with the matrix $\mathbf{P} \triangleq \tilde{\mathbf{H}}^T/\alpha$ where $\alpha = h_1^2 + h_2^2 + \ldots + h_{N_t}^2$ is a scalar. Then, we use the vector $\mathbf{P}\mathbf{x}$ to generate the real O-STBC codeword $\mathbf{X}$. Since the channel matrix $\tilde{\mathbf{H}}$ is orthogonal, such a precoding equalizes the effective channel, i.e.,

$$\mathbf{y} = \sqrt{\frac{k\rho}{N_t}}\tilde{\mathbf{H}}\mathbf{P}\mathbf{x} + \mathbf{n} = \sqrt{\frac{k\rho}{N_t}}\mathbf{x} + \mathbf{n}. \quad (24)$$

In the above, the constant k is used to satisfy the transmit power constraint; we derive its value below. Note that, with the aforementioned precoding, optimal data decoding at the receiver is very simple, as the equivalent channel consists of L parallel Single-Input Single-Output (SISO) AWGN channels with their gain independent of the channel instantiation. Since the effect of fading has been perfectly equalized at the transmitter, the proposed scheme achieves an infinite diversity order. Moreover, as the equivalent channel has a fixed gain, channel estimation is not required at the receiver.

In the proposed scheme, the columns of actual O-STBC matrix transmitted, $\mathbf{X}$, are constructed using permutations and sign-inversions of the entries of the precoded vector $\mathbf{P}\mathbf{x}$. Hence, the average transmit power over L channel uses, which is given by $\text{tr}(\mathbf{X}^T\mathbf{X})$, can be written as

$$P_{\text{avg}} = \frac{k\rho L}{N_t}\left(\mathbb{E}_{\mathbf{h},\mathbf{x}}\left[\mathbf{x}^T\mathbf{P}^T\mathbf{P}\mathbf{x}\right]\right) \quad (25)$$

where $\mathbb{E}_{\mathbf{h},\mathbf{x}}$ refers to the expectation over the distributions of $\mathbf{h}$ and $\mathbf{x}$. Since ρ is the total power available for transmission per channel use, to satisfy the average transmit power constraint of $P_{\text{avg}} = L\rho$, we need

$$\rho L = \frac{k\rho L}{N_t}\mathbb{E}_{\mathbf{h}}\left[\frac{1}{\alpha}\right]\mathbb{E}_{\mathbf{x}}[\|\mathbf{x}\|^2], \quad (26)$$

where the orthogonality property of $\tilde{\mathbf{H}}$ is used. Now, α is a $\chi^2_{N_t}$ random variable when the channel is Rayleigh fading with i.i.d. $\mathcal{N}(0,1)$ entries, and it can be shown that

$$\mathbb{E}\left[\frac{1}{\alpha}\right] = \frac{1}{N_t - 2}, \text{ for } N_t > 2. \quad (27)$$

Hence, assuming that each entry of $\mathbf{x}$ is normalized to have unit energy, $\mathbb{E}_{\mathbf{x}}[\|\mathbf{x}\|^2] = L$, and the average transmit power constraint can be satisfied by choosing

$$k = \frac{N_t(N_t - 2)}{L}. \quad (28)$$

Moreover, the SNR at the receiver can be computed as $\rho(N_t - 2)/L$.

Remark 2. *Note that the O-STBC based precoding scheme converts a Rayleigh fading channel into L parallel SISO AWGN channels with a fixed gain over L channel uses. Hence, we obtain an infinite diversity order with a single receive antenna. Having additional receive antennas can improve the received SNR, but does not increase the diversity order. One of the ways to handle multiple receive antennas is to do antenna selection. For this, the transmitter chooses the receive antenna for which the transmit power required is minimum. The selection index represented in $\log_2 N_r$ bits is conveyed to the receiver through other channels such as control channels. The average transmit power required in this case in naturally lower than that is required for single receive antenna. Applying* (47) *in Appendix B, for the case of $N_t = 4, L = 4$ with real $\mathbf{h}_i$, we get $k \approx N_t$ in the 2 receive antenna case with antenna selection, in contrast with $k = (N_t - 2)$ for the single antenna case. Thus, when $N_t = 4$, the above precoding scheme offers a nearly 3 dB improvement in the performance with 2 receive antennas and antenna selection, compared to the single receive antenna case.*

Remark 3. *The equivalent channel representation in* (20) *and the orthogonality property in* (23) *also hold for the complex 2×2 Alamouti code (see Exercise 9.4 of [32]). Hence, the above scheme also works with complex signaling when $N_t = 2$, by using the complex 2×2 Alamouti code as the underlying O-STBC. In this case, with one receive antenna, it can be shown that $\mathbb{E}[1/\alpha] = 1$. Thus, the average power constraint is satisfied in this case as well. However, the orthogonality property does not necessarily hold for other complex O-STBCs. Due to this, the real O-STBC based transmit precoding scheme does not directly extend to complex O-STBC signaling, except when $N_t = 2$.*

Remark 4. *Note that the instantaneous transmit power required in the above three schemes can be very large, especially when the channel coefficients are small. However, under the practical constraints on the peak power in practical power* *amplifiers, one has to limit the peak power. Under this constraint, it is straightforward to obtain k to meet the average transmit power constraint, in all the three precoding schemes proposed in this work.*

Next, we present CSIT-based precoding schemes for the fading multiuser MAC, BC and IC.

III. Precoding Schemes for Multiuser Channels

In this section, we extend the above transmit precoding schemes to the multiuser MAC, BC and IC. We assume that the wireless channels between transmit and receive antenna pairs are i.i.d. and Rayleigh distributed. An interesting feature of the proposed precoding schemes is that they require each transmitter to have knowledge only of the channel between itself and the receiver(s), and not the other users' channels. Hence, the proposed schemes do not require the exchange of CSI between transmitters. We start with the multiuser MAC with CSIT.

A. The Multiple Access Channel

1) Real O-STBC Signaling Scheme: Consider the M user MAC with N_t antennas at each transmitter (user) and a single antenna at the receiver. The received signal $\mathbf{y} \in \mathbb{R}^L$ can be written as

$$\mathbf{y} = \sum_{i=1}^{M} \sqrt{\frac{k\rho_i}{N_t}} \tilde{\mathbf{H}}^{(i)} \mathbf{P}^{(i)} \mathbf{x}_i + \mathbf{n}, \tag{29}$$

where $\mathbf{n} \in \mathbb{R}^L$ is the additive noise at the receiver, distributed as $\mathcal{N}(0,1)$; $\mathbf{x}_i \in \mathbb{R}_i^L$ is the O-STBC data vector; and ρ_i denotes the average transmit power from the i^{th} user. Also, $\mathbf{P}^{(i)}$ denotes the precoding matrix employed by the i^{th} transmitter corresponding to its channel to the receiver, $\tilde{\mathbf{H}}^{(i)}$ is the equivalent channel matrix as defined in Sec. II-C, and k denotes the power normalization constant. Now, we choose $\mathbf{P}^{(i)} \triangleq \tilde{\mathbf{H}}^{(i)\ T}/\alpha_i$ where $\alpha_i = \|\mathbf{h}_i\|^2$, and $\mathbf{h}_i$ is the channel from the i^{th} transmitter to the receiver, with i.i.d. $\mathcal{N}(0,1)$ entries. Then, as in Sec. II-C, the precoding scheme equalizes the channel, and we obtain L parallel Gaussian MACs with transmit powers $\rho_i, i = 1, 2, \ldots, K$. That is, the received signal can be written as

$$\mathbf{y} = \sum_{i=1}^{M} \sqrt{\frac{k\rho_i}{N_t}} \mathbf{x}_i + \mathbf{n}, \tag{30}$$

where $k = N_t(N_t - 2)/L$, for $N_t > 2$. Hence, the precoding scheme converts a Rayleigh flat-fading MISO MAC channel into a fixed-gain Gaussian MAC channel. Moreover, the scheme only requires each transmitter to have knowledge of its own channel to the receiver, and not the other users' channels.

Remark 5. *It is straightforward to extend the modified MRT based precoding scheme to the case of MAC. As in the case of the O-STBC based scheme, this results in a fixed-gain Gaussian MAC albeit with complex signaling. We omit the details here for the sake of brevity.*

2) QR-Based Precoding Scheme: Consider the M user Rayleigh fading MAC with N_r antennas at the receiver and $N_t \geq 2N_r$ antennas at each transmitter (user). Using precoding scheme described in the previous section, the received signal $\mathbf{y} \in \mathbb{C}^{N_r}$ can be written as

$$\mathbf{y} = \sum_{i=1}^{M} \sqrt{\frac{k\rho_i}{N_t}} \mathbf{H}_i^H \mathbf{P}_i \tilde{\mathbf{x}}_i + \mathbf{n}, \tag{31}$$

where $\mathbf{H}_i \in \mathbb{C}^{N_t \times N_r}$ denotes the channel between the i^{th} user and the receiver, distributed as i.i.d. $\mathcal{CN}(0,1)$, $\tilde{\mathbf{x}}_i \in \mathbb{C}^{N_t}$ denotes an extended data vector, and is derived from the complex data vector $\mathbf{x}_i$ as explained earlier in the single-user case. Also, ρ_i and $\mathbf{P}_i \in \mathbb{C}^{N_t \times N_t}$ denote the average transmit power available and the precoding matrix, respectively, corresponding to the i^{th} user, and k is a normalization constant. The components of the AWGN vector $\mathbf{n}$ are assumed to be i.i.d. $\mathcal{CN}(0,1)$. At the i^{th} transmitter, we choose the matrix $\mathbf{P}_i$ as in Sec. II-B. With this precoding scheme, the received data vector becomes

$$\mathbf{y} = \sum_{i=1}^{M} \sqrt{\frac{k\rho_i}{N_t}} \mathbf{x}_i + \mathbf{n}. \tag{32}$$

Thus, the precoding scheme converts the $N_t \times N_r$ MIMO Rayleigh fading MAC channel into N_r parallel Gaussian MAC channels with a fixed gain, when CSI is available at the transmitters.

B. The Broadcast Channel

We now present an adaptation of the proposed QR-decomposition based precoding scheme to the M user BC with N_r antennas at each user terminal and $N_t \geq 2MN_r$ antennas at the transmitter. Here, the combined channel matrix $\mathbf{H} \in \mathbb{C}^{N_t \times MN_r}$ between N_t transmit antennas and M user terminals can be considered as a virtual MIMO channel, but with MN_r individual messages. Let $\mathbf{x} = [\sqrt{\rho_1}\mathbf{s}_1, \sqrt{\rho_2}\mathbf{s}_2, \ldots, \sqrt{\rho_M}\mathbf{s}_M]^T$ denote the vector containing the messages intended to the M users, where ρ_i denotes the transmit power used by user i such that $\sum_i \rho_i = \rho$, the total available transmit power, and the transmitted symbols $\mathbf{s}_i \in \mathbb{C}^{N_r}$ are drawn from a constellation satisfying $\mathbb{E}[\mathbf{s}_i^H \mathbf{s}_i] = 1$. Let $\tilde{\mathbf{x}} \in \mathbb{C}^{N_t}$ denote an extended message vector, derived from $\mathbf{x} \in \mathbb{C}^{MN_r}$ as described in the previous section. Hence, one can write the signal model as

$$\mathbf{y} = \sqrt{\frac{k}{N_t}} \mathbf{H}^H \mathbf{P} \tilde{\mathbf{x}} + \mathbf{n}, \tag{33}$$

where $\mathbf{P} \in \mathbb{C}^{N_t \times N_t}$ is now a common precoding matrix for all users, k is a normalization constant and $\mathbf{n} \in \mathbb{C}^{MN_r}$ denotes the complex Gaussian noise vector at all the M receivers.

Now, the scheme proposed in Sec. II-B in the single-user case is directly applicable to the multiuser BC. Note that, due to the possibly unequal power allocation across the users, we have $\mathbf{C}_{\mathbf{x}} = \mathbb{E}\left[\mathbf{x}\mathbf{x}^H\right] = \text{diag}(\rho_1 \mathbf{I}_{N_r}, \rho_2 \mathbf{I}_{N_r}, \ldots, \rho_M \mathbf{I}_{N_r})$. Hence, the average power equation (14) is modified to:

$$P_{\text{avg}} = \frac{k\rho}{N_t} \operatorname{tr}\left(\mathbf{C}_{\mathbf{x}} \left\{\mathbf{I}_{MN_r} + \mathbb{E}_{\mathbf{h}}\left[(\mathbf{I} - \mathbf{R}_{u1})^H \mathbf{R}_{u2}^{-H} \mathbf{R}_{u2}^{-1} (\mathbf{I} - \mathbf{R}_{u1})\right]\right\}\right). \tag{34}$$

Correspondingly, the transmit power normalization constant k is given by

$$k = \frac{N_t}{\text{tr}\left(\mathbf{C_x}\left\{\mathbf{I} + \mathbb{E}_{\mathbf{h}}\left[\mathbf{R}_1^{-H}\mathbf{R}_1^{-1}\right]\right\}\right)}, \tag{35}$$

where we have used $\mathbf{U}_{11} = \mathbf{0}_{MN_r}$ and $\mathbf{U}_{12} = \mathbf{I}_{MN_r}$. Thus, the average power constraint can be satisfied, and the MIMO channel $\mathbf{H}^H \in \mathbb{C}^{N_t \times MN_r}$ is converted into MN_r parallel AWGN channels. Due to this, data received at the other users are not required for symbol detection and decoding at a given receiver.

C. The Interference Channel

In this subsection, we extend the transmit precoding proposed in the previous subsection to an M user IC. For ease of presentation, we consider the $M = 2$ user IC, with $N_t \geq 2MN_r$ antennas at each transmitter and N_r antennas at each receiver. In contrast with the BC, we now have $M-1$ interfering transmitters. The received signal at i^{th} receiver can be modeled as

$$\mathbf{y}_i = \sqrt{\frac{k}{N_t}} \sum_{j=1}^{2} \mathbf{H}_{i,j}^H \mathbf{P}_j \tilde{\mathbf{x}}_j + \mathbf{n}_i, \tag{36}$$

where $\mathbf{H}_{i,j} \in \mathbb{C}^{N_t \times N_r}$ denotes the channel matrix between the i^{th} transmitter and j^{th} receiver, having i.i.d. $\mathcal{CN}(0,1)$ entries, and $\mathbf{n}_i \in \mathbb{C}^{N_r}$ denotes the Gaussian noise at the i^{th} receiver, having i.i.d. $\mathcal{CN}(0,1)$ entries.

Now, we exploit the fact that a 2 user IC can be viewed as a combination of two interfering BCs. We employ the power allocation scheme described for the BC, and choose $\rho_1 = \rho$ and $\rho_2 = 0$ at transmitter 1, and $\rho_1 = 0$ and $\rho_2 = \rho$ at transmitter 2, with ρ denoting the per-user transmit power constraint, assumed to be the same for both users. We apply the precoding scheme presented for the BC in the previous subsection. Due to the zero power allocation to the signal component from each transmitter to the unintended receiver, the transmitters do not need to know the data symbols being transmitted by the other transmitter. Also, the receivers see only their intended messages, and hence do not need joint decoding or multiuser detection, and the Rayleigh fading IC is converted into MN_r parallel AWGN channels. Further, it is interesting to note that, when $M = 2$, the number of parallel AWGN channels corresponds precisely to the degrees of freedom of the two user $N_t \times N_r$ MIMO IC with perfect CSIT and CSIR [33].

IV. Simulation Results

In this section, we demonstrate the performance of the proposed precoding schemes using Monte Carlo simulations. We consider a Rayleigh flat-fading MIMO system with $N_t = 2$ or 4 antennas, and $N_r = 1$ or 2 antennas. We consider uncoded QPSK or 4-PAM constellations and compute the BER by averaging over 10^6 noise and 10^4 channel instantiations. We compare the BER performance of the proposed scheme with other existing schemes in the literature that assume perfect CSIR and/or perfect CSIT, such as MRT [25], space-time coding [29], [34], and ZF precoding [35], and vector perturbation based precoding for the BC [24].

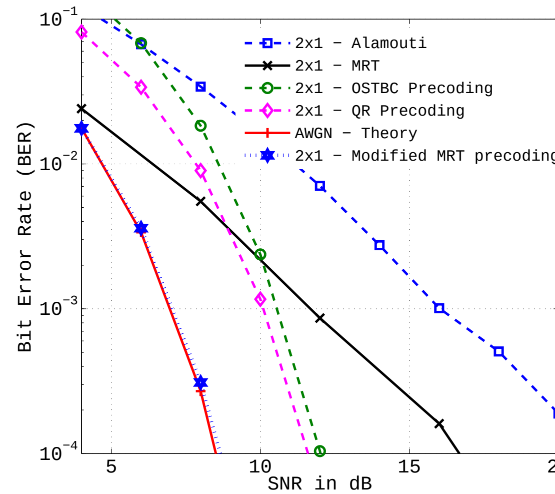

Fig. 1. BER comparison of the proposed precoding schemes along with Alamouti code with perfect CSIR and MRT precoding with perfect CSIR and CSIT for a 2×1 system using the QPSK constellation.

A. Single-User Channels

Figures 1 and 2 show the BER performance corresponding to the $N_t \times N_r = 2 \times 1$ and 2×2 MIMO systems, respectively. We compare the performance of the Alamouti scheme [34] under perfect CSIR, and MRT [25] under perfect CSIR and CSIT, with that of the proposed modified MRT, O-STBC and QR based precoding schemes under perfect CSIT. It can be observed that O-STBC based precoding needs about 0.5 dB higher transmit power to achieve the same BER, compared to the QR-based precoding scheme, while the performance of the modified MRT-based precoding scheme matches with the AWGN performance. Moreover, all three proposed precoding schemes exhibit the waterfall-like behavior, as in the AWGN channel, which is a significant improvement over the finite diversity order offered by existing schemes. Note that the O-STBC based scheme is simpler to implement compared to the QR-based scheme. When $N_r = 2$, both Alamouti and OSTBC based schemes use antenna selection at the receiver.[3] For the ZF precoding, we consider the scheme presented in [35], with two users and $N_t = 2$ transmit antennas at each user. We plot the BER performance of one of the users, when the users employ equal transmit power. Also, the performance of the O-STBC precoding scheme without antenna selection is about 3 dB worse than the unit-gain SISO AWGN channel, as predicted by the theory. Employing the antenna selection between two receive antennas fills most of this gap. Thus, the proposed scheme converts a MIMO fading channel into an equivalent SISO fixed gain AWGN channel.

To demonstrate the O-STBC based scheme with a higher number of transmit antennas, we show the performance of

[3] With receive antenna selection, we use the receive antenna for which the average transmit power required is the minimum, for finding the precoding matrix at the transmitter. This corresponds to choosing the antenna for which the ℓ_2 norm of the channel vector is the highest among all the receive antennas. Note that data decoding requires limited CSI at the receiver, since the receiver requires knowledge of the antenna selected by the precoding scheme.

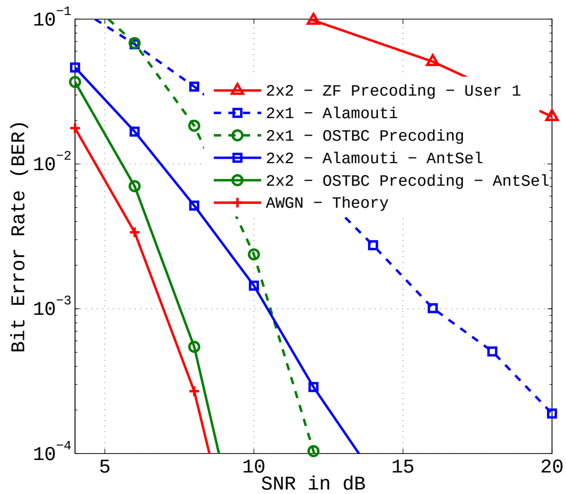

Fig. 2. BER comparison of the proposed precoding schemes along with Alamouti code with perfect CSIR with antenna selection, and ZF precoding for a 2×2 system using the QPSK constellation.

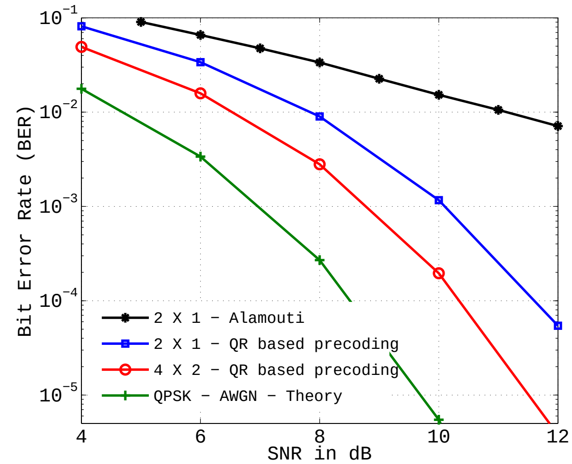

Fig. 4. BER comparison of the Alamouti code under perfect CSIR and proposed QR-based scheme under perfect CSIT, for the 2×1 and 4×2 systems, with uncoded QPSK signaling.

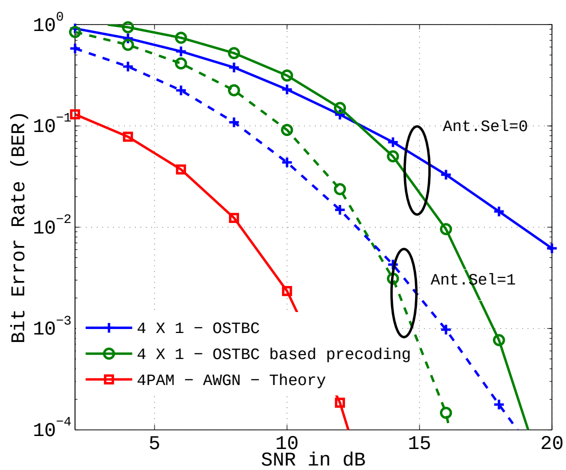

Fig. 3. BER comparison of the real O-STBC transmission scheme in [29] with perfect CSIR and proposed O-STBC based precoding scheme with perfect CSIT for a 4×1 system with 4-PAM constellation. The dashed curves correspond to the scheme with $N_r = 2$ and antenna selection at the receiver.

a 4×1 system employing the full-rate 4×4 real O-STBC code in (21) with 4-PAM constellation symbols in Fig. 3. Also shown is the performance of the 4×2 system with antenna selection at the receiver. In both cases, we see that the proposed precoding scheme renders the effective channel to be a fixed-gain AWGN channel at all SNRs, as expected. Also, the antenna selection between two antennas results in about 3 dB gain in the BER performance for the proposed precoding scheme, while it results in a diversity order improvement from 4 to 8 for the CSIR-based O-STBC transmission scheme.

Figure 4 shows the BER performance the QR based precoding scheme for the 2×1 and 4×2 systems. We also show the performance of the complex Alamouti code with uncoded QPSK transmission and perfect CSIR. It can be seen the BER of the proposed scheme is parallel to that of the unit-gain SISO AWGN channel. The gap between the two is about 3 dB and 1.7 dB for the 2×1 and 4×2 systems, respectively, which corroborates well with the theory in (18). Further, the proposed scheme far outperforms the perfect CSIR-based Alamouti coding scheme.

1) Precoding with CSI Estimated at the Transmitter: Now, we present simulation results when the channel is estimated at the transmitter using a reverse-link training sequence consisting of 10 known symbols transmitted with 10 dB power boosting compared to the forward-link data SNR. The MMSE channel estimator is used for estimating the CSIT. Simulation results are provided for the O-STBC based precoder in Fig. 5; the behavior of the modified MRT and QR based schemes is similar, but it is not shown here to avoid repetition. It can be seen that the BER performance is close to that obtained with perfect CSIT, and that the waterfall-type behavior of the curves is retained.

2) Transmit Precoding with a Peak Power Constraint: Here, we present the simulation results when the peak power used by the transmitter is restricted to a practical limit (say, to 20 dB higher than the average power). Limiting the peak power does not invert the channel perfectly for those channel realizations where the peak power required is more than 20 dB above the average power constraint, but the transmit power constraint is still satisfied with the normalization factor k derived earlier. The BER performance is plotted as a function of the SNR for the O-STBC scheme in Fig. 6; the behavior of the other two precoding schemes is similar. It can be observed that the BER performance is very close to the one with no peak power limit, and the peak power constraint does not significantly alter the behavior of the curves at practical SNRs.

B. Multiuser Channels

In Fig. 7, we demonstrate the performance of the O-STBC precoding scheme for the MAC channel with $M = 2$ users,

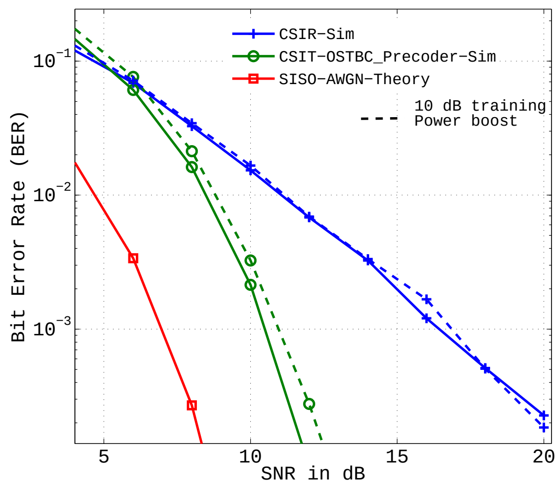

Fig. 5. BER performance the $N_t = 2,\ N_r = 1$ system with O-STBC based precoding and estimated CSIT. Here, estimated channel coefficients are used for both CSIR based CSIT based schemes.

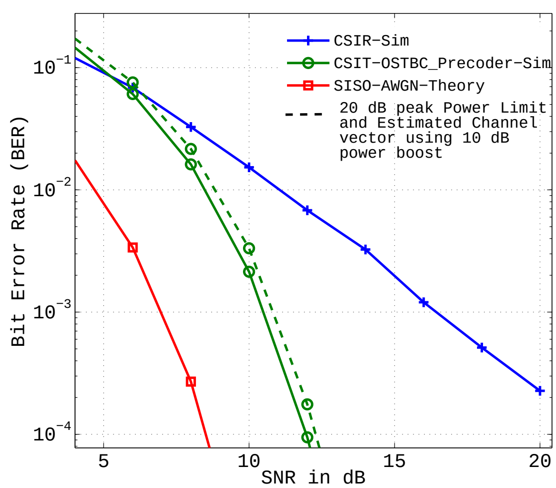

Fig. 6. BER performance of the $N_t = 2,\ N_r = 1$ system with O-STBC based precoding and a peak power constraint. The peak power was limited to be 15 dB higher than the average transmit power. As another example, peak power limit of 20 dB is used along with estimated channel vector using training sequence with 10 dB additional power than the data transmission.

$N_t = 2, N_r = 1$ and $L = 2$. We compare the performance of the complex Alamouti code constructed using QPSK symbols with that of the proposed O-STBC based and QR based precoding schemes. Here, users 1 and 2 are allocated $9/10$ and $1/10$ of the total transmit power, respectively. For decoding symbols from the two users, a joint Maximum Likelihood (ML) decoder is used at the receiver. We see, again, that the proposed precoding schemes are able to convert the fading MAC into a fixed-gain Gaussian MAC, with the QR based precoding scheme marginally outperforming the O-STBC based precoding scheme.

We next illustrate the BER performance of the proposed precoding scheme for the two-user BC, in Fig. 8. We consider

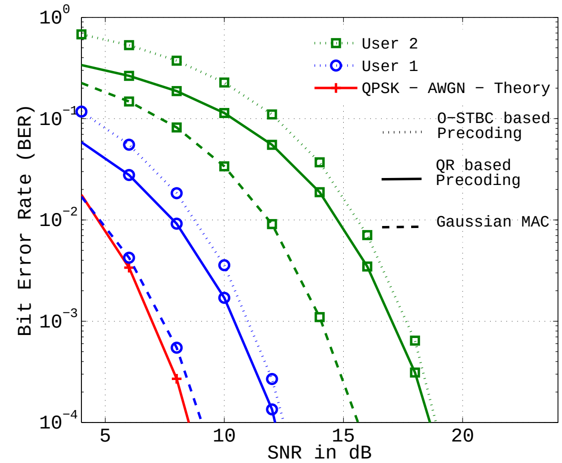

Fig. 7. BER performance of users 1 and 2, with QPSK signaling in a 2×1 MAC. Here, the transmit powers at the users are set using $\rho_1 = 9/10$SNR and $\rho_2 = 1/10$SNR, and joint ML decoding is employed at the receiver.

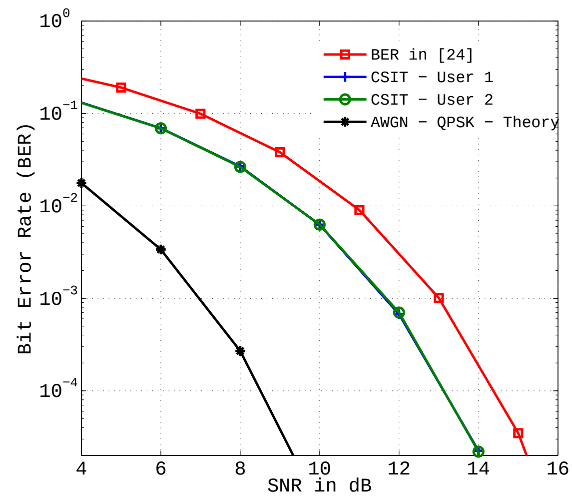

Fig. 8. BER performance of users 1 and 2 with uncoded QPSK signaling in a 4×1 BC with $\rho_1 = \rho_2 = 1/2$SNR.

a 4×1 system with uncoded QPSK signaling. Equal power is allocated to both users, and, hence, the power normalization constant k with the QR-based precoding scheme is given by (17). We see that the performance of the QR-based precoding scheme is parallel to the that of uncoded QPSK symbols in a unit-gain AWGN channel. Thus, the fading MIMO BC is converted into 2 parallel fixed-gain AWGN channels. In the plot, we also show the performance of the vector perturbation method for multiuser BC in [24] for the same antenna configuration, which also requires CSIT. The proposed scheme is not only simpler from an implementation point of view at both the transmitter and receiver, but also outperforms the vector perturbation approach by about 1 dB.

Note that, since the precoding scheme for the IC follows from that of the BC, it results in exactly the same performance as in the BC at the two receivers. Hence, we do not explicitly

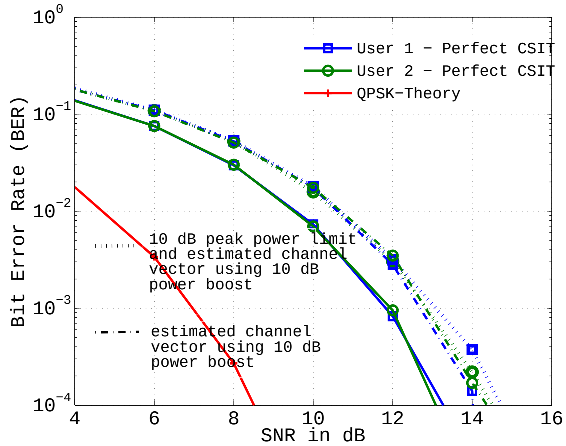

Fig. 9. BER performance of users 1 and 2 with uncoded QPSK signaling in a 4×1 BC channel with $M = 2$, $\rho_1 = \text{SNR}/2$ and $\rho_2 = \text{SNR}/2$. The precoder used estimated channel matrix and peak power limitation.

illustrate the performance of the proposed scheme for the 2-user IC.

Finally, in Fig. 9, we show the BER performance obtained for the 2 user BC with $N_t = 4$, when the CSI is estimated at the transmitter using a training signal with 10 dB power boosting compared to the data SNR. Also, the peak power is limited to be 10 dB higher than the average power. We see that imperfect CSIT and practical peak-to-average power constraints have little impact on the BER performance at practical SNRs.

V. Conclusions

In this paper, we proposed three novel, simple-to-implement precoding schemes which utilize CSIT to convert a Rayleigh fading MIMO channel into a fixed-gain AWGN channel, thereby achieving an infinite diversity order, while satisfying an average power constraint. Thus, if perfect CSI could be made available either at the transmitter, or at the receiver, but not both, the perfect CSIT option provides significantly better resilience to fading. The proposed schemes not only offer an improvement over CSIR-based techniques in terms of the diversity order, but also admit single symbol ML decoding at the receiver. We extended the precoding schemes to the fading multiuser MIMO multiple access, broadcast and interference channels. In all three cases, we showed that the fading MIMO channel is converted into parallel fixed-gain AWGN channels. Numerical simulations illustrated the significant performance advantage of the proposed scheme compared to CSIR-based transmit diversity schemes. Thus, the proposed precoding schemes are promising for use in reciprocal MIMO systems, where it is practically feasible to directly acquire CSI at the transmitter.

Appendix

A. Diversity Analysis with Imperfect CSIT in the Modified MRT based Precoding

In this section, we show that a diversity order of $N_t - 1$ can be obtained with the MRT based precoding scheme presented in Sec. II-A. For simplicity, consider $N_r = 1$ and M-PSK modulated data. Let $\mathbf{h} \in \mathbb{C}^{N_t}$ denote the channel vector. Let $\hat{\mathbf{h}} = \mathbf{h} + \Delta\mathbf{h}$ denote the channel vector estimated using the uplink training sequence, which is used for precoding. We assume that MMSE channel estimation is employed. Because of this, $\hat{\mathbf{h}}$ and $\Delta\mathbf{h}$ are independent random variables [36]. Moreover $\Delta\mathbf{h}$ is i.i.d. Gaussian with zero mean and variance $\sigma_d^2 = \sigma_n^2/\rho_{\text{tr}}$, where ρ_{tr} is the transmit power used during the training phase, which is assumed to be equal to the data transmit power, i.e., $\rho_{\text{tr}} = \rho$. Note that, the average power constraint can be satisfied even when $\hat{\mathbf{h}}$ is used for precoding instead of $\mathbf{h}$. The norm of $\hat{\mathbf{h}}$ is a scaled χ^2 random variable with the scaling factor $\left(1 + \sigma_n^2/\rho\right)$. Define $\tilde{\rho} = \rho/\sigma_n^2$. Hence, using $k = N_t(N_t - 1)\left(1 + \tilde{\rho}^{-1}\right)$ ensures that the average power constraint is satisfied, for complex signaling.

Now, the received data signal, in (1), can be re-written as

$$\begin{aligned} y &= \sqrt{\frac{k\rho}{N_t}}\mathbf{h}^H \frac{\hat{\mathbf{h}}}{\|\hat{\mathbf{h}}\|^2} x + n, \\ &= \sqrt{\frac{k\rho}{N_t}} x - \sqrt{\frac{k\rho}{N_t}} \frac{\hat{\mathbf{h}}^H \Delta\mathbf{h}}{\|\hat{\mathbf{h}}\|^2} x + n. \end{aligned} \tag{37}$$

Hence, the probability of error at the decoder is given by

$$\begin{aligned} P_e &\triangleq \mathbb{E}_{\hat{\mathbf{h}},x_i}\left\{ \Pr\left[\left| y - \sqrt{\frac{k\rho}{N_t}} x_i \right|^2 > \left| y - \sqrt{\frac{k\rho}{N_t}} x_j \right|^2, \right.\right. \\ &\qquad \left.\left. \forall j \neq i \right] \right\} \\ &\leq \mathbb{E}_{\hat{\mathbf{h}},x_i}\left\{ (M-1)\Pr\left[4\Re\left\{ \sqrt{\frac{k\rho}{N_t}} \frac{\hat{\mathbf{h}}^H \Delta\mathbf{h}}{\|\hat{\mathbf{h}}\|^2} x_i + n \right\} \right.\right. \\ &\qquad \left.\left. > \sqrt{\frac{k\rho}{N_t}} d_{\text{min}}^2, \right] \right\}, \end{aligned} \tag{38}$$

where we have taken the union bound, and used the fact that, for M-PSK data, $|x_i|^2 = 1$ and $d_{\text{min}}^2 \leq |x_i - x_j|^2 \leq 2$. Note that, with MMSE channel estimation, conditioned on $\hat{\mathbf{h}}$, $\hat{\mathbf{h}}$ and $\Delta\mathbf{h}$ are uncorrelated. Due to this, the term $\hat{\mathbf{h}}^H \Delta\mathbf{h}/\|\hat{\mathbf{h}}\|^2$ becomes a scaled Gaussian random variable with zero mean and variance $1/(\tilde{\rho}\|\hat{\mathbf{h}}\|^2)$. Hence, we can write the probability of error expression as

$$\begin{aligned} P_e &\leq \mathbb{E}_{\hat{\mathbf{h}}}\left\{ (M-1) Q\left(\frac{d_{\text{min}}}{2} \sqrt{\frac{k\rho}{N_t\left(\sigma_n^2 + \frac{2k\rho}{\|\hat{\mathbf{h}}\|^2 \tilde{\rho} N_t}\right)}} \right) \right\} \\ &\leq \frac{M-1}{2} \mathbb{E}_{\hat{\mathbf{h}}}\left\{ e^{-\left(\frac{d_{\text{min}}^2 k\rho}{4N_t\left(\sigma_n^2 + \frac{2k\rho}{\|\hat{\mathbf{h}}\|^2 \tilde{\rho} N_t}\right)} \right)} \right\} \\ &\leq \frac{M-1}{2} \int_0^\infty e^{-\left[\frac{\gamma\tilde{\rho} d_{\text{min}}^2 N_t(N_t-1)(1+\tilde{\rho}^{-1})}{4\left(\gamma N_t + 2N_t(N_t-1)(1+\tilde{\rho}^{-1})\right)} \right]} f_\gamma(\gamma) \mathrm{d}\gamma, \end{aligned} \tag{39}$$

where $\gamma = \|\hat{\mathbf{h}}\|^2$. We split the above integral into two parts, between the ranges $(0, c]$ and (c, ∞), where $c \triangleq 2(N_t-1)(1+\tilde{\rho}^{-1})$. Then, we can further upper bound the probability of error as

$$\begin{aligned}P_e &\leq \frac{M-1}{2}\left[\int_0^c e^{-\left[\frac{\gamma\tilde{\rho}d_{\min}^2 N_t(N_t-1)(1+\tilde{\rho}^{-1})}{16N_t(N_t-1)(1+\tilde{\rho}^{-1})}\right]}\right.\\&\quad\left.+\int_c^\infty e^{-\left[\frac{\gamma\tilde{\rho}d_{\min}^2 N_t(N_t-1)(1+\tilde{\rho}^{-1})}{8\gamma N_t}\right]}\right] f_\gamma(\gamma)\mathrm{d}\gamma\\&\leq \frac{M-1}{2}\left[\left(1-F_\gamma\left(2(N_t-1)(1+\tilde{\rho}^{-1})\right)\right)\right. \qquad (40)\\&\quad\left.\times e^{-\tilde{\rho}\left[\frac{d_{\min}^2(N_t-1)(1+\tilde{\rho}^{-1})}{8}\right]}\right]\\&\quad+\frac{M-1}{2^{N_t+1}(N_t-1)!(1+\tilde{\rho}^{-1})^{N_t-1}}\int_0^c e^{-\gamma\left[\frac{\tilde{\rho}d_{\min}^2}{16}\right]}\\&\quad\times\gamma^{N_t-1}e^{-\frac{\gamma}{(1+\tilde{\rho}^{-1})}}\mathrm{d}\gamma. \qquad (41)\end{aligned}$$

It is now easy to see that, as $\tilde{\rho}$ gets large, the first term goes to zero exponentially in $\tilde{\rho}$, and the diversity order behavior is determined by the second term. Now, the second term in the above equation is given by

$$\int_0^c \gamma^{N_t-1}e^{-\alpha\gamma}\mathrm{d}\gamma = \frac{1}{\alpha^{N_t-1}}\gamma\left(N_t-1,\frac{c}{\alpha}\right). \qquad (42)$$

where $\alpha = \left[\tilde{\rho}d_{\min}^2/16 + 1/(1+\tilde{\rho}^{-1})\right]$, and $\gamma(s,x)$ is the lower incomplete Gamma function [28]. Using an expansion of $\gamma(s,x)$, we can write

$$\gamma(s,x) = \Gamma(s)\left[1-e^{-x}\left(\frac{x^{s-1}}{(s-1)!}+\frac{x^{s-2}}{(s-2)!}+\cdots+1\right)\right].$$

Now, substituting for $s = (N_t-1)$ and $x = c/\alpha = 32(N_t-1)(1+\tilde{\rho}^{-1})^2/\left[\tilde{\rho}d_{\min}^2(1+\tilde{\rho}^{-1})+16\right]$, and taking $\tilde{\rho}$ to infinity, we get

$$\frac{1}{\alpha^{N_t-1}}\gamma\left(N_t-1,\frac{c}{\alpha}\right) \leq \mathcal{O}\left(\tilde{\rho}^{-(N_t-1)}\right). \qquad (43)$$

Hence, we obtain a polynomial diversity order of N_t-1 when using the proposed modified MRT based precoding scheme with channel estimation errors.

B. Mean of the Inverse of the Maximum of Two χ_K^2 Distributed Random Variables

The CDF of the random variable $X \triangleq \max(X_1,X_2)$, where X_i's are χ^2-distributed with K degrees of freedom, can be written as

$$F_X(x) = \left(\frac{\gamma\left(\frac{K}{2},\frac{x}{2}\right)}{\Gamma\left(\frac{K}{2}\right)}\right)^2. \qquad (44)$$

The PDF of X can be obtained by differentiating the above with respect to x, as

$$f_X(x) = \frac{1}{\Gamma^2\left(\frac{K}{2}\right)}\gamma\left(\frac{K}{2},\frac{x}{2}\right)\left(\frac{x}{2}\right)^{\frac{K}{2}-1}e^{-\frac{x}{2}}, \quad x\geq 0. \quad (45)$$

We expand $\gamma(s,x)$ into an infinite series as

$$\gamma(s,x) = x^s\Gamma(s)e^{-x}\sum_{i=0}^{\infty}\frac{x^i}{\Gamma(s+i-1)}. \qquad (46)$$

Substituting in (45) and taking expectation of $1/X$, it is easy to show that

$$\mathbb{E}\left[\frac{1}{X}\right] = \frac{2^{1-K}}{\Gamma\left(\frac{K}{2}\right)}\sum_{i=0}^{\infty}\frac{\Gamma(K-1+i)}{2^i\Gamma\left(\frac{K}{2}+1+i\right)}, \text{ for } K>2. \quad (47)$$

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