PERFORMANCE OF TRANSMIT OPTIMIZATION FOR TWO-INPUT MULTIPLE-OUTPUT SPATIAL MULTIPLEXING

Neng Wang

Communications Research Centre
Industry Canada, Ottawa, Ontario, Canada
E-mail: nwang@crc.ca

Steven D. Blostein

Dept. of Electrical and Computer Eng.
Queen's University, Kingston, Ontario, Canada
E-mail: steven.blostein@queensu.ca

ABSTRACT

A two-input multiple-output (TIMO) system represents an important special case of multiple-input multiple-output (MIMO) systems. In this paper, performance of transmit optimization for TIMO spatial multiplexing systems is investigated. Efforts to optimize TIMO transmission have involved minimum mean squared-error (MMSE) precoding and decoding, minimum bit error rate (MBER) precoding for zero-forcing (ZF) receiver, minimum distance (MD) precoding for maximum-likelihood (ML) receiver, approximate MBER transmit power allocation, and approximate MBER transmit beamforming. It is shown both analytically and by simulations that power allocation together with interference cancellation and detection ordering outperforms both linear MMSE and ZF-MBER precoding, and ML-MD precoding and transmit beamforming eliminate error floors in ill-conditioned channels and offer superior performance over both power allocation and linear precoding/decoding methods.

1. INTRODUCTION

Wireless communications using multiple transmit and receive antennas, known as multiple-input multiple-output (MIMO) systems, offers key advantages over single-input single-output (SISO) systems, such as diversity and spatial multiplexing gains [1]. Our goal of this paper is to investigate transmit optimization for a MIMO spatial multiplexing system with two transmit antennas, known also as two-input multiple-output (TIMO). The study of such a system can be motivated in a number of ways: 1) TIMO systems are important in practical scenarios where there are limitations on cost and/or space to install more antennas; 2) a virtual TIMO channel is created when two single-antenna mobiles operate in cooperative communication mode [2]; 3) when transmit antenna selection is employed in MIMO to achieve diversity with reduced cost of transmit radio frequency chains [3], selecting two out of multiple transmit antennas turns MIMO into TIMO; 4) it is easier to analyze TIMO systems than the more general MIMO systems, and these analyses offer insights into MIMO system design and performance.

When channel state information (CSI) is available at the transmitter, system performance can be improved. Transmit optimization is receiver-dependent. Signal reception for spatial multiplexing can employ criteria such as maximum likelihood (ML), zero-forcing (ZF), minimum mean squared-error (MMSE), successive interference cancellation (SIC), or ordered SIC (OSIC) as, for example, in the case of the Vertical Bell Laboratories Layered Space-Time (V-BLAST) [4]. Using MMSE criterion, linear precoding/decoding schemes were investigated in [5]. For ZF equalization, minimum bit error rate (MBER) precoding was proposed in [6]. In [7], a precoder for ML receiver was proposed using minimum Euclidean distance (MD) criterion. These schemes, however, generally require high feedback overhead and/or high complexity processing, e.g., diagonalization of the channel matrix and/or matrix transformations at both the transmitter and the receiver. Efforts to reduce feedback overhead have involved limited feedback precoding [8], as well as approximate MBER (AMBER) transmit power allocation [9–11]. For TIMO systems, AMBER transmit beamforming (BF) schemes were proposed in [12, 13] with the capability of eliminating error floors in ill-conditioned channels. While [12, 13] utilizes instantaneous channel knowledge, the transmit BF method proposed in [14] exploits spatial correlation of transmit antennas.

This paper investigates performance of a variety of transmit optimization schemes for TIMO spatial multiplexing. As in [12, 13], TIMO channels are categorized into well- and ill-conditioned cases. Channel condition is determined by a number of factors, such as Ricean factor and/or spatial correlation. Error rate performances of precoding and, particularly, our proposed transmit power allocation and beamforming [12, 13] are analyzed for both well- and ill-conditioned channels. Performances of these schemes are compared both analytically and by numerical simulations. The rest of this paper is organized as follows: In Section 2, TIMO receiver structures and their performances are reviewed, based on which an ill-conditioned channel model is proposed. Section 3 reviews a variety of transmit optimization schemes. Performance analysis and comparison are presented in Sections 4 and 5 respectively. Section 6 provides numerical results in general correlation fading channels.

2. TIMO CHANNEL AND RECEIVER

Consider a TIMO system with $N_r \geq 2$ receive antennas. The received signal can be modelled as

$$\mathbf{r} = \mathbf{H}\mathbf{s} + \boldsymbol{\eta} = s_1\mathbf{h}_1 + s_2\mathbf{h}_2 + \boldsymbol{\eta}, \qquad (1)$$

where $\mathbf{H} = [\mathbf{h}_1 \vdots \mathbf{h}_2]$ is an $N_r \times 2$ channel matrix, which is assumed to be general correlated Ricean fading [15], $\mathbf{s} = [s_1, s_2]^T$ denotes a transmitted signal vector, and $\boldsymbol{\eta}$ is an $N_r \times 1$ additive Gaussian noise vector. For simplification of analysis purposes, we assume white noise and input, i.e., $\mathbb{E}[\mathbf{s}\mathbf{s}^H] = E_s\mathbf{I}_2$ and $\mathbb{E}[\boldsymbol{\eta}\boldsymbol{\eta}^H] = N_0\mathbf{I}_{N_r}$, and define the signal-to-noise ratio (SNR) $\gamma_s \overset{\text{def}}{=} E_s/N_0$. Binary Phase Shift Keying (BPSK) modulation is assumed.

This research was made possible through support from Bell Mobility and Samsung.

2.1. TIMO Signal Reception and Performance

2.1.1. ZF Receiver

With ZF equalization, the decision-point SNR of the k-th signal stream is obtained as

$$\gamma_{Z,k} = \frac{2E_s}{N_0\left[(\mathbf{H}^H\mathbf{H})^{-1}\right]_{k,k}} \stackrel{\text{def}}{=} 2\gamma_s g^2_{Z,k}, \quad (k=1,2),$$

where $g^2_{Z,k}$ denotes the power gain of k-th stream, and can be calculated as

$$g^2_{Z,1} = \frac{\Delta_{\mathbf{H}}}{\|\mathbf{h}_2\|^2}, \qquad g^2_{Z,2} = \frac{\Delta_{\mathbf{H}}}{\|\mathbf{h}_1\|^2}, \tag{2}$$

where $\Delta_{\mathbf{H}} \stackrel{\text{def}}{=} \|\mathbf{h}_1\|^2\|\mathbf{h}_2\|^2 - |\mathbf{h}_2^H\mathbf{h}_1|^2$.

2.1.2. SIC Receiver

Without loss of generality (w.o.l.g.), we assume that stream $k=1$ is detected first. Assuming ZF equalization without error propagation, the power gains can be obtained as [12, 13]

$$g^2_{S,1} = g^2_{Z,1} = \frac{\Delta_{\mathbf{H}}}{\|\mathbf{h}_2\|^2}, \quad g^2_{S,2} = \|\mathbf{h}_2\|^2. \tag{3}$$

2.1.3. OSIC Receiver

To improve SIC performance, the streams can be reordered based on SNR at each stage. The SNR-based ordering scheme [4] detects the stream with largest decision-point SNR first. The power gains of OSIC are given by [12, 13]

$$g^2_{O,1} = \frac{\Delta_{\mathbf{H}}}{\min\{\|\mathbf{h}_1\|^2, \|\mathbf{h}_2\|^2\}}, \quad g^2_{O,2} = \min\{\|\mathbf{h}_1\|^2, \|\mathbf{h}_2\|^2\}. \tag{4}$$

The average BER of the above receivers can be calculated as [16]

$$\bar{P}(\gamma_s; g_1^2, g_2^2) = \frac{1}{2}\mathcal{Q}\left(\sqrt{2\gamma_s g_1^2}\right) + \frac{1}{2}\mathcal{Q}\left(\sqrt{2\gamma_s g_2^2}\right), \tag{5}$$

where the power gains g_1^2 and g_2^2 depend on the receiver structure and are given in (2), (3) and (4); $\mathcal{Q}(x) \stackrel{\text{def}}{=} \frac{1}{\sqrt{2\pi}}\int_x^\infty e^{-y^2/2}dy$. We note that for SIC and OSIC receivers, (5) is only a lower bound due to the neglecting of error propagation. However, at moderate-to-high SNR regimes, this lower bound closely approximates the average BER since error propagation is minimal.

2.2. Ill-Conditioned TIMO Channels

Since the $\mathcal{Q}(\cdot)$ function decreases rapidly in its argument, the average BER in (5) is dominated by the term with smaller power gain. In the extreme case with vanishing power gain, the system experiences an error floor. We refer to this as an *ill-conditioned* TIMO channel, which can be modelled as [12, 13]

$$\mathbf{H} \cong \mathbf{h}_1[1 \;\; a], \quad a \in \mathbb{C}. \tag{6}$$

This is also an example of a "pinhole" channel [17]. The least-squares (LS) estimate of a can be found to be

$$\hat{a}_{LS} = \mathbf{h}_1^\dagger \mathbf{h}_2 = \frac{\mathbf{h}_1^H\mathbf{h}_2}{\|\mathbf{h}_1\|^2}. \tag{7}$$

3. TRANSMIT OPTIMIZATION FOR TIMO

We review a variety of transmit optimization schemes. While some schemes apply to general MIMO systems, our treatment in this paper is for TIMO channels. Let $\mathbf{H}^H\mathbf{H} = \mathbf{W}\mathbf{\Lambda}\mathbf{W}^H$ denote eigenvalue decomposition of $\mathbf{H}^H\mathbf{H}$. Denote the eigenvalues as λ_1 and λ_2, and assume $\lambda_1 \geq \lambda_2$. The transmit power is normalized to be $\mathcal{P}_t = 2$.

3.1. Precoding

The MMSE precoding matrix is given by [5]

$$\mathbf{P}_M = \mathbf{W}\mathbf{\Phi},$$

where $\mathbf{\Phi} = \text{diag}\{\phi_1, \phi_2\}$ and ϕ_1 and ϕ_2 satisfy

$$\begin{cases} |\phi_1|^2 = \frac{2}{\gamma_s}, \;\; |\phi_2|^2 = 0, & \text{if } \gamma_s \leq \frac{1}{2}\left(\frac{1}{\sqrt{\lambda_1\lambda_2}} - \frac{1}{\lambda_1}\right) \\ |\phi_k|^2 = \frac{2\gamma_s + \lambda_1^{-1} + \lambda_2^{-1}}{\gamma_s \lambda_k^{1/2}\left(\lambda_1^{-1/2} + \lambda_2^{-1/2}\right)} - \frac{1}{\gamma_s\lambda_k}, \;\; (k=1,2), & \text{o.w.} \end{cases}.$$

For ZF receiver, the MBER precoder is given by [6]

$$\mathbf{P}_Z = \mathbf{W}\mathbf{\Lambda}^{-1/4}\begin{bmatrix} 1 & 1 \\ 1 & -1 \end{bmatrix} \Big/ \left(\text{tr}\left(\mathbf{\Lambda}^{-1/2}\right)\right)^{1/2}.$$

A MD precoder for ML receiver with BPSK modulation is given by [7]

$$\mathbf{P}_D = \mathbf{W}\begin{bmatrix} 1 & j \\ 0 & 0 \end{bmatrix}.$$

3.2. Power Allocation (PA)

Denote $\mathbf{P}_{PA} = \text{diag}(p_1, p_2)$ as the transmit power allocation matrix. Using the approximate BER formula [18]

$$P_b(\gamma) = \mathcal{Q}\left(\sqrt{2\gamma}\right) \cong \frac{1}{5}\exp\{-\gamma\}, \tag{8}$$

an AMBER-PA solution is given by [12, 13]

$$p_k^2 = \gamma_s^{-1} g_k^{-2}\left(\ln g_k^2 + \nu\right)_+, \quad k=1,2, \tag{9}$$

where $(x)_+ \stackrel{\text{def}}{=} \max\{0, x\}$, and ν is chosen to satisfy power constraint.

3.3. Transmit BF for Ill-Conditioned TIMO Channels

Using ZF equalization in ill-conditioned channel (6), two signal streams are coupled and signal detection can be performed in either one-dimensional (1D) or two-dimensional (2D) signal space [12, 13].

An AMBER transmit beamformer for 1D detection is obtained as [12, 13]

$$\mathbf{P}_{1D} = (1+|a|^2)^{-1/2}\begin{bmatrix} 1 \\ a^* \end{bmatrix} \times \sqrt{\frac{2}{5}}\begin{bmatrix} 2 & 1 \end{bmatrix}, \tag{10}$$

which pre-mixes two BPSK streams into one 4-PAM. We refer to this scheme as *4-PAM beamforming*. While the precoder in [14] performs the same pre-mixing, it exploits spatial correlation of transmit antennas. On the other hand, (10) utilizes instantaneous channel knowledge and does not depend on the channel model used.

If signal detection is performed in 2D space, an AMBER precoder is found to be

$$\mathbf{P}_{2D} = (1+|a|^2)^{-1/2}\begin{bmatrix} 1 \\ a^* \end{bmatrix} \times \begin{bmatrix} 1 & j \end{bmatrix}, \tag{11}$$

which pre-mixes two BPSK streams into one QPSK. We refer to this scheme as *QPSK beamforming*. Compared to ML-MD precoding [7], which performs similar pre-mixing, (11) applies to ZF receiver for ill-conditioned channel (6).

4. PERFORMANCE ANALYSIS

Performances of transmit optimization schemes reviewed in Section 3 are analyzed for both well- and ill-conditioned channels.

4.1. Precoding

The average BER performance of the MMSE precoding/decoding is given by [5]

$$\bar{P}_{\text{M}}\left(\gamma_s;\mathbf{h}_1,\mathbf{h}_2\right)=\frac{1}{2}\sum_{k=1}^{2}\mathcal{Q}\left(\sqrt{2\gamma_s\lambda_k\left|\phi_k\right|^2}\right).$$

At moderate-to-high SNR ($\gamma_s \gg 1$), the above BER can be approximated by

$$\tilde{P}_{\text{M}}\left(\gamma_s;\mathbf{h}_1,\mathbf{h}_2\right)\cong\frac{1}{2}\sum_{k=1}^{2}\mathcal{Q}\left(\sqrt{\frac{4\gamma_s\lambda_k^{1/2}}{\lambda_1^{-1/2}+\lambda_2^{-1/2}}}\right). \quad (12)$$

By calculating the decision-point SNR, the average BER of ZF-MBER precoding can be found to be

$$\bar{P}_{\text{Z}}\left(\gamma_s;\mathbf{h}_1,\mathbf{h}_2\right)=\mathcal{Q}\left(\sqrt{\frac{8\gamma_s}{\left(\lambda_1^{-1/2}+\lambda_2^{-1/2}\right)^2}}\right). \quad (13)$$

Similarly, the average BER of ML-MD precoding can be obtained as [7]

$$\bar{P}_{\text{MD}}^{\text{ML}}\left(\gamma_s;\mathbf{h}_1,\mathbf{h}_2\right)=\mathcal{Q}\left(\sqrt{2\gamma_s\lambda_1}\right). \quad (14)$$

In ill-conditioned channels, the smaller eigenvalue,

$$\lambda_2=\frac{1}{2}\left(\|\mathbf{h}_1\|^2+\|\mathbf{h}_2\|^2-\sqrt{(\|\mathbf{h}_1\|^2+\|\mathbf{h}_2\|^2)^2-4\Delta_{\mathbf{H}}}\right),$$

is vanishing. From (12) and (13), we see that both linear MMSE and ZF-MBER precoding experience error floors. On the other hand, ML-MD precoding eliminates error floors since signal streams are pre-mixed and transmitted over the dominant eigen-mode.

4.2. Power Allocation

We consider well-conditioned channels first. As shown in [10], at moderate-to-high SNR ($\gamma_s \gg 1$), the BER of power allocation (9) can be approximated as

$$\tilde{P}_{\text{PA}}\left(\gamma_s;g_1^2,g_2^2\right)\cong\frac{1}{10}\left(g_1^{-2}+g_2^{-2}\right)\exp\left\{-\frac{2\gamma_s}{g_1^{-2}+g_2^{-2}}\right\}. \quad (15)$$

For ill-conditioned channels, we assume w.o.l.g., $|a|\le 1$ in (6). Consider a OSIC receiver since it has been shown to outperform both ZF and SIC [9, 10]. The power gains of OSIC can be obtained as $g_{O,1}^2\cong 0$ and $g_{O,2}^2=|a|^2\|\mathbf{h}_1\|^2$. Applying power allocation in (9), we obtain $p_1^2=0$, and $p_2^2=2$. The average BER of power allocation for ill-conditioned channels can be approximated as

$$\bar{P}(\gamma_s;\mathbf{h}_1,a)\cong\frac{1}{10}+\frac{1}{10}\exp\left\{-4\gamma_s|a|^2\|\mathbf{h}_1\|^2\right\}, \quad (16)$$

which experiences an obvious error floor.

4.3. Transmit Beamforming

The 4-PAM and QPSK beamforming schemes are developed for ill-conditioned channels. In ill-conditioned channels, their average BER can be approximated as, respectively, [12, 13]

$$\tilde{P}_{\text{4-PAM}}^{\text{BF}}(\gamma_s;\mathbf{h}_1,a)\cong\frac{3}{20}\exp\left\{-\frac{2}{5}\gamma_s\|\mathbf{h}_1\|^2(1+|a|^2)\right\} \quad (17)$$

$$\tilde{P}_{\text{QPSK}}^{\text{BF}}(\gamma_s;\mathbf{h}_1,a)\cong\frac{1}{5}\exp\left\{-\gamma_s\|\mathbf{h}_1\|^2(1+|a|^2)\right\}. \quad (18)$$

Therefore, both 4-PAM and QPSK beamforming do not experience error floors in ill-conditioned channels. Comparing (17) with (18), we observe an approximate SNR gain of 2.5 ($\cong$ 4 dB), at the expense of increased detection complexity. This will be verified by simulations in Section 6.

It is of interest to study performances of beamforming in well-conditioned channels. The performances of 4-PAM and QPSK beamforming can be obtained by substituting $\hat{a}_{LS}$ in (7) into (17) and (18), respectively,

$$\tilde{P}_{\text{4-PAM}}^{\text{BF}}(\gamma_s;\mathbf{h}_1,\mathbf{h}_2)\cong\frac{3}{20}\exp\left\{-\frac{2\gamma_s}{5}\frac{\|\mathbf{h}_1\|^4+|\mathbf{h}_1^H\mathbf{h}_2|^2}{\|\mathbf{h}_1\|^2}\right\} \quad (19)$$

$$\tilde{P}_{\text{QPSK}}^{\text{BF}}(\gamma_s;\mathbf{h}_1,\mathbf{h}_2)\cong\frac{1}{5}\exp\left\{-\gamma_s\frac{\|\mathbf{h}_1\|^4+|\mathbf{h}_1^H\mathbf{h}_2|^2}{\|\mathbf{h}_1\|^2}\right\}. \quad (20)$$

From (19) and (20), we observe that 1D detection has a 4 dB power penalty relative to 2D detection in well-conditioned channels as well.

5. PERFORMANCE COMPARISON

For convenience of description, we define the SNR gain of a scheme with error rate $P_e(\gamma_s)$:

$$\Gamma\overset{\text{def}}{=}\lim_{\gamma_s\to\infty}-\frac{1}{\gamma_s}\ln P_e(\gamma_s),$$

which is a measure of asymptotic error rate performance. We now compare a variety of transmit optimization schemes. For beamforming, we consider QPSK beamforming first and address 4-PAM briefly later. Their SNR gains are given by, respectively,

$$\Gamma_{\text{QPSK}}^{\text{BF}}=\frac{\|\mathbf{h}_1\|^4+|\mathbf{h}_1^H\mathbf{h}_2|^2}{\|\mathbf{h}_1\|^2},\quad\Gamma_{\text{4-PAM}}^{\text{BF}}=\frac{2}{5}\Gamma_{\text{QPSK}}^{\text{BF}}.$$

5.1. QPSK Beamforming versus Power Allocation

Consider the asymptotic BER of power allocation in (15). The SNR gain is given by $\Gamma_{\text{PA}}=2\left(g_1^{-2}+g_2^{-2}\right)^{-1}$, where the power gains g_1^2 and g_2^2 depend on the receiver structure and are given in Section 2.1. We consider OSIC with power allocation, since it has been shown to outperform both ZF and SIC. We have

$$\Gamma_{\text{PA}}^{\text{O}}=\frac{2}{g_{\text{O},1}^{-2}+g_{\text{O},2}^{-2}}=\frac{2\Delta_{\mathbf{H}}}{g_{\text{O},1}^2+g_{\text{O},2}^2}\le\Delta_{\mathbf{H}}^{1/2}\le\|\mathbf{h}_1\|^2\le\Gamma_{\text{QPSK}}^{\text{BF}}.$$

Therefore, QPSK beamforming outperforms OSIC power allocation, at least asymptotically.

5.2. QPSK Beamforming versus Precoding

- *ZF-MBER precoding:* From (13) and (8), the SNR gain for ZF-MBER precoding can be found to be

$$\Gamma_{\text{MBER}}^{\text{ZF}} = 4\left(\lambda_1^{-1/2} + \lambda_2^{-1/2}\right)^{-2}.$$

We have

$$\Gamma_{\text{MBER}}^{\text{ZF}} \leq (\lambda_1\lambda_2)^{1/2} = \Delta_{\mathbf{H}}^{1/2} \leq \|\mathbf{h}_1\|^2 \leq \Gamma_{\text{QPSK}}^{\text{BF}}. \quad (21)$$

Therefore, we have

$$\bar{P}_{\text{BF}}^{\text{QPSK}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2) \leq \bar{P}_{\text{Z}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2).$$

We note that in (21), the equality holds only when $\mathbf{h}_1^H \mathbf{h}_2 = 0$ and $\|\mathbf{h}_1\| = \|\mathbf{h}_2\|$. Such a channel has condition number 1, and is referred to as a *perfect channel*.

- *MMSE precoding/decoding:* Define

$$f(x) \stackrel{\text{def}}{=} \mathcal{Q}\left(\sqrt{\frac{4\gamma_s}{\left(\lambda_1^{-1/2} + \lambda_2^{-1/2}\right)x}}\right).$$

It is readily shown that $f(x)$ convex if $x < \frac{4\gamma_s}{3\left(\lambda_1^{-1/2} + \lambda_2^{-1/2}\right)}$. Therefore, $f(x)$ is asymptotically convex at $\gamma_s \gg 1$. By using Jensen's inequality, we have, asymptotically,

$$\begin{aligned} \bar{P}_{\text{M}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2) &= \frac{1}{2}f(\lambda_1^{-1/2}) + \frac{1}{2}f(\lambda_2^{-1/2}) \\ &\geq f\left(\frac{1}{2}\left(\lambda_1^{-1/2} + \lambda_2^{-1/2}\right)\right) \\ &= \bar{P}_{\text{Z}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2), \end{aligned}$$

where the equality holds only for perfect channels.

- *ML-MD precoding:* The average BER of ML-MD precoding can be calculated as

$$\bar{P}_{\text{MD}}^{\text{ML}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2) = \mathcal{Q}\left(\sqrt{2\gamma_s\lambda_1}\right) \cong \frac{1}{5}e^{-\gamma_s\lambda_1}.$$

Its SNR gain is therefore given by $\Gamma_{\text{MD}}^{\text{ML}} = \lambda_1$. It can be shown that

$$\lambda_1 \geq \|\mathbf{h}_1\|^2 + \frac{|\mathbf{h}_1^H\mathbf{h}_2|^2}{\|\mathbf{h}_1\|^2} = \Gamma_{\text{QPSK}}^{\text{BF}},$$

with equality only if $\|\mathbf{h}_1\|^2\|\mathbf{h}_2\|^2 = |\mathbf{h}_1^H\mathbf{h}_2|^2$, i.e., the channel has rank one. Hence, in general

$$\bar{P}_{\text{MD}}^{\text{ML}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2) \leq \bar{P}_{\text{QPSK}}^{\text{BF}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2);$$

and in ill-conditioned channels,

$$\bar{P}_{\text{MD}}^{\text{ML}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2) \cong \bar{P}_{\text{QPSK}}^{\text{BF}}(\gamma_s; \mathbf{h}_1, \mathbf{h}_2).$$

5.3. 4-PAM Beamforming

For 4-PAM beamforming, due to the 4-dB power penalty, the SNR gain $\Gamma_{\text{4-PAM}}^{\text{BF}}$ may or may not be larger than those of other power allocation and precoding schemes. This makes adaptation of transmission schemes based on channel conditions possible. When 1D detection is employed, transmit optimization can be adapted relative to channel characteristics under a MBER criterion. As an example, we consider adaptive transmission of 4-PAM beamforming and power allocation for OSIC. A simple approximate MBER adaptation rule is

$$\begin{cases} \text{power allocation} & \text{if } \Gamma_{\text{PA}}^{\text{O}} \geq \Gamma_{\text{BF}}^{\text{4-PAM}}; \\ \text{4-PAM beamforming} & \text{otherwise.} \end{cases} \quad (22)$$

This can viewed as *adaptive modulation*. Application to other receiver structures is similar.

6. NUMERICAL RESULTS AND DISCUSSIONS

We now compare our analytical asymptotic performance results derived in Sections 4 and 5 with simulations. In our simulations, we adopt the spatial fading correlation model for general non-isotropic scattering given in [15]. The following parameters are chosen: $N_r = 4$ receive antennas; transmit and receive antenna spacings expressed in wavelengths are 0.5 and 10, respectively; angles of arrival and departure of the deterministic component are $\pi/6$ and 0, respectively; angle spread 10°; and BPSK modulation is used.

Fig. 1 is a plot of the average BER for a variety of transceivers in an uncorrelated Rayleigh fading channel. To clarify the plot, performances of ZF with power allocation and SIC without power allocation are not shown since they are nearly identical to that of MMSE precoding/decoding; OSIC without power allocation (also not shown) has performance close to that of ZF-MBER precoding. We observe that both SIC and OSIC with power allocation outperform precoding schemes, e.g., at a BER of 10^{-3}, OSIC with power allocation offers 0.9 and 1.9 dB SNR gains over ZF-MBER and MMSE precoding/decoding, respectively. In Fig. 1, it is also observed that QPSK beamforming offers superior performance to all other simulated schemes except ML-MD precoding, e.g., at a BER of 10^{-3}, its SNR gain over OSIC with power allocation is 3.3 dB.

Fig. 2 illustrates average BER's in correlated Ricean fading channels. Performance of SIC without power allocation (not shown) is nearly identical to that of MMSE precoding/decoding. Again, SIC and OSIC with power allocation outperform ZF-MBER and MMSE precoding/decoding. We also observe that the proposed 4-PAM and QPSK beamforming offer significant gain over the power allocation schemes shown: at a BER of 10^{-3}, 7.5 and 11.5 dB SNR gain over OSIC with power allocation are observed. This is as excepted since in Ricean fading, due to the existence of a line-of-sight (LOS) component, the channel matrix is likely to be ill-conditioned.

General Observations: From Figs. 1-2, we observe the SNR losses of QPSK beamforming relative to ML-MD precoding are less than 0.5 dB, and decrease as channel correlation increases.

7. CONCLUSIONS

Performance of transmit optimization for TIMO spatial multiplexing is investigated in this paper. A number of transmit optimization schemes are considered, including linear MMSE precoding and decoding and MBER precoding for ZF receiver, MD precoding for ML receiver, AMBER transmit power allocation, and AMBER transmit beamforming. It is shown that SIC and OSIC with AMBER power allocation outperform linear ZF-MBER and MMSE precoding/decoding schemes in Rayleigh fading channels, e.g., at a BER of 10^{-3}, OSIC with power allocation offers 0.9 dB and 1.9 dB SNR gains over ZF-MBER and MMSE precoding/decoding respectively. The 4-PAM and QPSK transmit beamforming schemes eliminate error floors and offers superior performance over both power allocation and ZF-MBER and MMSE precoding/decoding in Ricean fading channels, e.g., at a BER of 10^{-3}, their SNR gains over OSIC

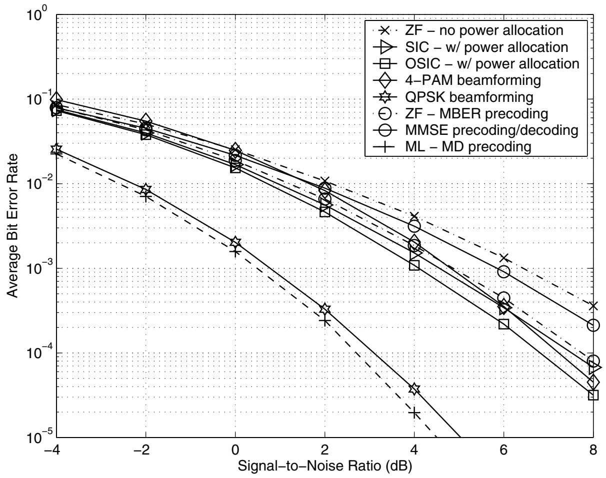

Fig. 1. Average BER in uncorrelated Rayleigh fading channel.

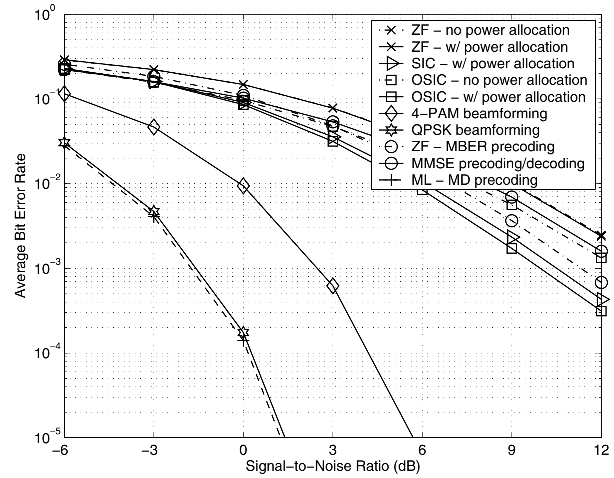

Fig. 2. Average BER in correlated Ricean fading channel ($K = 8$ dB).

with power allocation are, respectively, 7.5 and 11.5 dB; and QPSK beamforming offers performance close to ML-MD precoding.

8. REFERENCES

[1] A. J. Paulraj, D. A. Gore, R. U. Nabar, and H. Bölcskei, "An Overview of MIMO Communications—A Key to Gigabit Wireless," *Proc. of IEEE,* vol. 92, no. 2, pp. 198-218, Feb. 2004.

[2] A. Sendonaris, E. Erkip, and B. Aazhang, "User Cooperation Diversity–Part I: System Description," *IEEE Trans. Commun.,* vol. 51, no. 11, pp. 1927-1938, Nov. 2003.

[3] D. A. Gore, R. W. Heath, Jr., and A. J. Paulraj, "Transmit Selection in Spatial Multiplexing Systems," *IEEE Commun. Lett.,* vol. 6, no. 11, pp. 491-493, Nov. 2002.

[4] G. J. Foschini, G. D. Golden, R. A. Valenzuela, P. W. Wolniansky, "Simplified Processing for High Spectral Efficiency Wireless Communication Employing Multi-Element Arrays," *IEEE J. Select. Areas Commun.,* vol. 17, no. 11, pp. 1841-1852, Nov. 1999.

[5] A. Scaglione, P. Stoica, S. Barbarossa, G. B. Giannakis and H. Sampath, "Optimal Designs for Space-Time Linear Precoders and Equalizers," *IEEE Trans. Signal Processing,* vol. 50, no. 5, pp. 1051-164, May 2002.

[6] Y. Ding, T. N. Davidson, Z.-Q. Luo, and K. M. Wong, "Minimum BER Block Precoders for Zero-Forcing Equalization," *IEEE Trans. Signal Processing,* vol. 51, no. 9, pp. 2410-2423, Sept. 2003.

[7] L. Collin, O. Berder, P. Rostaing, and G. Burel, "Optimal minimum distance-based precoder for MIMO spatial multiplexing," *IEEE Trans. Signal Processing,* vol. 52, no. 3, pp. 617-627, Mar. 2004.

[8] D. J. Love and R. W. Heath Jr., "Limited Feedback Precoding for Spatial Multiplexing Systems," *Proc. IEEE Globecom 2003,* vol.4, pp. 1857-1861, San Francisco, CA, Dec. 2003.

[9] N. Wang and S. D. Blostein, "Minimum BER Power Allocation for MIMO Spatial Multiplexing Systems," *Proc. IEEE ICC 2005*, vol. 4, pp. 2282-2286, Seoul, Korea, May 2005.

[10] N. Wang and S. D. Blostein, "Approximate Minimum BER Power Allocation for MIMO Spatial Multiplexing Systems," *IEEE Trans. Commun.*, accepted for publication, July 2005.

[11] S. H. Nam, O.-S. Shin, and K. B. Lee, "Transmit Power Allocation for a Modified V-BLAST System," *IEEE Trans. Commun.,* vol. 52, no. 7, pp. 1074-1079, July 2004.

[12] N. Wang and S. D. Blostein, "Minimum BER Transmit Power Allocation and Beamforming for Two-Input Multiple-Output Spatial Multiplexing System," submitted to *IEEE Trans. Veh. Technol.*, Dec. 2004, revised Jan. 2006.

[13] N. Wang and S. D. Blostein, "Minimum BER Transmit Optimization for Two-Input Multiple-Output Spatial Multiplexing," *Proc. IEEE Globecom 2005*, vol. 6, pp. 3774-3778, St. Louis, MO, Nov.-Dec. 2005.

[14] J. Akhtar and D. Gesbert, "A Closed-Form Precoder for Spatial Multiplexing over Correlated MIMO Channels," *Proc. IEEE Globecom 2003,* vol.4, pp. 1847-1851, San Francisco, CA, Dec. 2003.

[15] A. Abdi and M. Kaveh, "A Space-Time Correlation Model for Multielement Antenna Systems in Mobile Fading Channels," *IEEE J. Select. Areas Commun.,* vol. 20, no. 3, pp. 550-560, Apr. 2002.

[16] J. G. Proakis, *Digital Communications,* 4th ed., New York: McGraw-Hill, 2001.

[17] D. Gesbert, H. Bölcskei, D. Gore, and A. J. Paulraj, "Outdoor MIMO Wireless Channels: Models and Performance Prediction," *IEEE Trans. Commun.,* vol. 50, no. 12, pp. 1926-1934, Dec. 2002.

[18] S. Zhou and G. B. Giannakis, "Adaptive Modulation for Multi-Antenna Transmissions with Channel Mean Feedback," *Proc. of IEEE ICC 2003*, vol. 4, pp. 2281-2285, Anchorage, AK, May 2003.