Approximate Minimum BER Power Allocation for MIMO Spatial Multiplexing Systems

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***Abstract*—Precoding for multiple-input multiple-output (MIMO) spatial multiplexing generally requires high feedback overhead and/or high-complexity processing. Simultaneous reduction in transmitter complexity and feedback overhead is proposed by imposing a diagonal structural constraint to precoding, i.e., power allocation. Minimum bit-error rate (MBER) is employed as the optimization criterion, and an approximate MBER (AMBER) power-allocation algorithm is proposed for a variety of receivers, including zero-forcing (ZF), successive interference cancellation (SIC), and ordered SIC (OSIC). While previously proposed precoding schemes either require ZF equalization for MBER, or use a minimum mean-squared error (MMSE) criterion, we provide a unified MBER solution to power allocation for ZF, SIC, and OSIC receiver structures. Improved error-rate performance is shown both analytically and by simulation. Simulation results also indicate that SIC and OSIC with AMBER power allocation offer superior performance over previously proposed MBER precoding with ZF equalization, as well as over MMSE precoding/decoding. Performance under noisy channels and power feedback is analyzed. A modified AMBER algorithm that mitigates error propagation in interference cancellation is developed. Compared with existing precoding methods, the proposed schemes significantly reduce both transmit processing complexity and feedback overhead, and improve error-rate performance.**

***Index Terms*—Minimum bit-error rate (MBER), multiple-input multiple-output (MIMO), power allocation, precoding, spatial multiplexing, successive interference cancellation (SIC).**

I. Introduction

MULTIPLE-INPUT multiple-output (MIMO) communication offers key advantages over single-input single-output (SISO) communication, such as diversity gain and spatial multiplexing gain [1]. Diversity gain improves link reliability, while spatial multiplexing gain increases the transmission rate. Our goal with this paper is to investigate transmit optimization for MIMO spatial multiplexing, which is receiver-dependent. Signal reception for MIMO spatial multiplexing can employ criteria such as linear zero-forcing (ZF), minimum mean-squared error (MMSE), maximum likelihood (ML), successive interference cancellation (SIC), or ordered SIC (OSIC) as, for example, in the case of the vertical Bell Laboratories layered space–time (V-BLAST) architecture [2].

In order to achieve high MIMO diversity and/or spatial multiplexing gains, appropriate transceiver designs are necessary. Efforts to optimize MIMO transceiver structures have involved joint transmit–receive optimization and linear precoding for specific receivers. Joint precoding/decoding optimization under MMSE criterion is investigated in [3], and [4] studies MMSE precoding for SIC receivers. A unified framework for joint transmit–receive design using convex optimization is proposed in [5]. Minimum bit-error rate (MBER) precoding for ZF equalization of block transmission [6] and block transceivers with MMSE decision-feedback equalization (DFE) [7] are readily applicable to MIMO systems. Precoding for multicarrier MIMO using an ML receiver and pairwise error probability as criterion is proposed in [8]. These designs generally require high-complexity processing at both the transmitter and the receiver, as well as high feedback overhead. Precoded MIMO transmission with reduced feedback has been recently proposed, based on quantized channel state information (CSI) feedback [9] and limited feedback signal design [10]. However, existing precoding schemes with reduced feedback generally also require high processing complexity.

In this paper, we consider simultaneous reduction of transmitter complexity and feedback overhead by constraining precoding to transmit power allocation, i.e., we optimize only the transmitted power of signal streams, but apply a more suitable criterion. Power allocation for multicarrier MIMO systems was considered in [11], where MIMO was operated in a diversity mode and the transmit power was allocated across the frequency dimension (subcarriers). As opposed to MMSE precoding/decoding [3], we consider MBER as the optimization criterion. The block transceiver design for MMSE-DFE [7] provides a closed-form solution to approximate MBER (AMBER) precoding for SIC receivers. Compared with MMSE-DFE [7] and ZF-MBER precoding [6], we provide a unified solution to MBER power allocation for ZF, SIC, and OSIC receivers. General power allocation (with diagonal precoder) by minimizing error rate does *not* have a closed-form solution, and has high computational complexity. An approximate solution can be found instead, which was originally given in [12] to allocate power across channel eigenmodes. In this paper, it is applied to transmit power allocation for ZF, SIC, and OSIC receivers. Recently, it has come to our attention that a similar AMBER power allocation for V-BLAST was proposed independently [13], which is a special case of our proposed solution for well-conditioned channels, as shown in Section III-C. Simulation results show that in general correlated fading channels [14],

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our proposed AMBER power allocation, together with SIC or OSIC (V-BLAST) reception, offers superior performance over existing MBER precoding for ZF equalization, as well as MMSE precoding/decoding.

Transmitter-side power allocation ideally requires CSI or allocated power to be available at the transmitter. In some cases, CSI can be made available at the transmitter, e.g., in time-division duplex (TDD) systems, due to the reciprocity of the uplink and downlink channels. In this case, existing limited feedback schemes do *not* possess any advantages, since feedback overhead is not a concern. However, power allocation is still attractive, due to the significant reduction in transmitter complexity. On the other hand, in channels that lack reciprocity in uplink and downlink, e.g., frequency-division duplex (FDD), complete CSI is not available at the transmitter, and CSI or power information has to be fed back. Regardless of availability, CSI or power feedback is imperfect, in practice, due to channel estimation, quantization, feedback delay, and/or errors introduced by feedback channel [15]. This motivates performance analysis of power allocation under uncertain feedback. While a general analysis is difficult, we analyze the special cases of noisy CSI and power feedback. Based on this analysis, we propose an AMBER power-allocation algorithm that takes statistical knowledge of noisy feedback into account. Furthermore, as a byproduct, a modified algorithm for perfect CSI which takes into account error-propagation effects in SIC and OSIC receivers is devised.

The rest of this paper is organized as follows. MIMO signal reception and performance are introduced in Section II. Section III investigates AMBER power allocation for MIMO with ZF, SIC, and OSIC receivers and their performances. In Section IV, performance degradation and power allocation under imperfect feedback are studied. Section V presents numerical results in general correlated fading channels.

II. MIMO Signal Reception and Performance

Consider a MIMO spatial multiplexing communication system with N_t transmit and N_r receive antennas, where $N_r \geq N_t$. The received signal can be modeled as $\mathbf{r} = \mathbf{H}\mathbf{s} + \boldsymbol{\eta}$, where $\mathbf{s}$ is the $N_t \times 1$ transmitted signal vector; $\mathbf{H}$ is the $N_r \times N_t$ channel matrix, which is assumed to be general correlated Ricean fading, as in [14]; and $\boldsymbol{\eta}$ is the $N_r \times 1$ additive Gaussian noise vector. For simplification of analysis, we assume white noise and input, i.e., $\mathbb{E}[\mathbf{s}\mathbf{s}^H] = E_s\mathbf{I}_{N_t}$ and $\mathbb{E}[\boldsymbol{\eta}\boldsymbol{\eta}^H] = N_0\mathbf{I}_{N_r}$, input signal-to-noise ratio (SNR) $\gamma_s \stackrel{\text{def}}{=} E_s/N_0$, and binary phase-shift keying (BPSK) modulation is used without loss of generality (as will be made clear in Section III-A).

1) ZF Receiver: With ZF equalization, the estimate of the transmitted signal $\mathbf{s}$ is given by $\hat{\mathbf{s}} = \mathbf{H}^{\dagger}\mathbf{r} = \mathbf{s} + \mathbf{H}^{\dagger}\boldsymbol{\eta}$, where $(\cdot)^{\dagger}$ denotes the matrix Moore–Penrose pseudoinverse. The decision-point SNR of the kth signal stream, i.e., the signal from the kth transmit antenna, $1 \leq k \leq N_t$, is obtained as

$$\gamma_{\text{Z},k} = \gamma_s[(\mathbf{H}^H\mathbf{H})^{-1}]_{k,k}^{-1} \stackrel{\text{def}}{=} \gamma_s g_{\text{Z},k}^2 \tag{1}$$

where $g_{\text{Z},k}^2 \stackrel{\text{def}}{=} [(\mathbf{H}^H\mathbf{H})^{-1}]_{k,k}^{-1}$ denotes the power gain of the kth stream.

2) SIC Receiver: Without loss of generality, we assume signal stream $k = 1$ is detected first. The interference due to the first stream is then regenerated and subtracted before stream $k = 2$ is detected. This procedure is repeated successively until all streams are detected. We ignore error propagation from early stages, which is a valid assumption at moderate-to-high SNR. Assuming ZF equalization is employed at each stage, the decision-point SNR of the kth stream at the kth stage is given by

$$\gamma_{\text{S},k} = \gamma_s \left[\left(\mathbf{H}_{(k)}^H\mathbf{H}_{(k)}\right)^{\dagger}\right]_{k,k}^{-1} \stackrel{\text{def}}{=} \gamma_s g_{\text{S},k}^2 \tag{2}$$

where $\mathbf{H}_{(k)}$ is generated in a recursive fashion by nulling the $(k-1)$st column of $\mathbf{H}_{(k-1)}$ for $k = 2, \ldots, N_t$, $\mathbf{H}_{(1)} \stackrel{\text{def}}{=} \mathbf{H}$, and $g_{\text{S},k}^2 \stackrel{\text{def}}{=} [(\mathbf{H}_{(k)}^H\mathbf{H}_{(k)})^{\dagger}]_{k,k}^{-1}$.

3) OSIC Receiver: To improve SIC performance, the received signal streams can be reordered based on SNR at each stage. This receiver differs from a SIC receiver only in the detection ordering. An SNR-based ordering scheme that maximizes minimum SNR appears in [2]. The decision-point SNR of the kth stream at the kth stage is given by

$$\gamma_{\text{O},k} = \gamma_s \left[\left(\left(\mathbf{H}_{(k)}^{\text{O}}\right)^H \mathbf{H}_{(k)}^{\text{O}}\right)^{\dagger}\right]_{k,k}^{-1} \stackrel{\text{def}}{=} \gamma_s g_{\text{O},k}^2 \tag{3}$$

where $\mathbf{H}^{\text{O}}$ denotes the reordered channel matrix, and $g_{\text{O},k}^2 \stackrel{\text{def}}{=} [((\mathbf{H}_{(k)}^{\text{O}})^H\mathbf{H}_{(k)}^{\text{O}})^{\dagger}]_{k,k}^{-1}$.

The average BER of the above receivers can be calculated as [16][1]

$$\bar{P}\left(\gamma_s; \{g_k^2\}\right) = \frac{1}{N_t}\sum_{k=1}^{N_t} \mathcal{Q}\left(\sqrt{2\gamma_s g_k^2}\right) \tag{4}$$

where $\mathcal{Q}(x) \stackrel{\text{def}}{=} (1/\sqrt{2\pi})\int_x^{\infty} e^{-y^2/2}dy$; g_k^2 depends on the receiver structure and is given in (1)–(3).

III. AMBER Power Allocation for MIMO With Perfect Feedback

Denote the power allocated to the kth stream as p_k^2. The received signal can be written as $\mathbf{r} = \mathbf{H}\mathbf{P}\mathbf{s} + \boldsymbol{\eta}$, where $\mathbf{P} \stackrel{\text{def}}{=} \text{diag}\{p_1, \cdots, p_{N_t}\}$ denotes the power-allocation matrix. We assume the total transmit power is constrained via

$$\text{tr}(\mathbf{P}\mathbf{P}^T) = \sum_{k=1}^{N_t} p_k^2 = N_t. \tag{5}$$

Compared with general precoding, power allocation constrains the precoder to a diagonal matrix.

[1]This is a lower bound for SIC and OSIC, due to the neglecting of error propagation, which is also an accurate approximate at moderate-to-high SNRs.

A. AMBER Power-Allocation Algorithm

The average BER for power allocation is given by a straightforward generalization of (4), i.e.,

$$\bar{P}\left(\gamma_s; \{p_k^2\}; \{g_k^2\}\right) = \frac{1}{N_t}\sum_{k=1}^{N_t} \mathcal{Q}\left(\sqrt{2\gamma_s g_k^2 p_k^2}\right). \quad (6)$$

To minimize the average BER in (6) under transmit power constraint (5), no closed-form solution exists. However, taking the approach in [17], we approximate the MBER objective function to obtain a closed-form solution. For general constellations, the BER can be approximated as $P_b(\gamma) \approx (1/5)\exp\{-c\gamma\}$, where c is a constellation-specific constant [18]. For BPSK modulation, $c = 1$, and the average BER in (6) can be approximated as

$$\bar{P}\left(\gamma_s; \{p_k^2\}; \{g_k^2\}\right) \approx \frac{1}{5N_t}\sum_{k=1}^{N_t} \exp\left\{-\gamma_s p_k^2 g_k^2\right\} \stackrel{\text{def}}{=} \tilde{P}\left(\gamma_s; \{p_k^2\}; \{g_k^2\}\right). \quad (7)$$

Minimization of (7) under power constraint (5) results in the power allocation [12]

$$p_k^2 = \gamma_s^{-1} g_k^{-2}\left(\ln g_k^2 + \nu\right)_+, \quad 1 \le k \le N_t \quad (8)$$

where $(x)_+ \stackrel{\text{def}}{=} \max\{0, x\}$, and ν is chosen to satisfy the power constraint (5). Note that the total power

$$\sum_k p_k^2(\nu) = \sum_k \gamma_s^{-1} g_k^{-2}\left(\ln g_k^2 + \nu\right)_+$$

is a piecewise-linear increasing function of ν, with breakpoints at $-\ln g_k^2$'s. Therefore, ν is unique and can be readily determined. A recursive algorithm which requires at most N_t recursions can be employed to find the AMBER solution [19]. Extension of the results to other constellations is straightforward.

B. Performance Analysis

1) Asymptotic Performance: By substituting (8) into (6), we obtain the average BER of power allocation as $\bar{P}_{\rm PA}(\gamma_s; \{g_k^2\}) = (1/N_t)\sum_{k=1}^{N_t} \mathcal{Q}(\sqrt{2(\ln g_k^2 + \nu)_+})$, which makes performance comparison difficult, due to the nonlinear operator $(\cdot)_+$. However, at moderate-to-high SNR ($\gamma_s \gg 1$), it is possible to proceed. When the condition

$$\nu \ge \max_k\left\{-\ln g_k^2\right\} \quad (9)$$

is valid, the power allocation in (8) simplifies to

$$p_k^2 = \gamma_s^{-1} g_k^{-2}(\ln g_k^2 + \nu_{\rm a})$$
$$\nu_{\rm a} \stackrel{\text{def}}{=} \frac{\gamma_s - \frac{1}{N_t}\sum_{l=1}^{N_t} g_l^{-2}\ln g_l^2}{\frac{1}{N_t}\sum_{l=1}^{N_t} g_l^{-2}}. \quad (10)$$

From (9) and (10), we observe that the simplified power allocation (10) is valid at moderate-to-high SNR regimes, and/or when the channel is well-conditioned (when the gains g_k's are less spread).[2] Furthermore, if (9) holds, the average BER of the power allocation can be approximated by

$$\bar{P}_{\rm PA}\left(\gamma_s; \{g_k^2\}\right) = \frac{1}{N_t}\sum_{k=1}^{N_t} \mathcal{Q}\left(\sqrt{2\left(\ln g_k^2 + \nu_{\rm a}\right)}\right) \quad (11)$$
$$\approx \frac{1}{5N_t} e^{-\nu_{\rm a}} \sum_{k=1}^{N_t} g_k^{-2} \stackrel{\text{def}}{=} \tilde{P}_{\rm PA}\left(\gamma_s; \{g_k^2\}\right). \quad (12)$$

2) Asymptotic Optimality of AMBER Solution: The exact MBER solution satisfies

$$\sqrt{\frac{\gamma_s g_k^2}{p_k^2}} e^{-\gamma_s g_k^2 p_k^2} = \zeta \quad \forall k \in [1, N_t]$$

where ζ is chosen to satisfy the power constraint (5), [12]. Therefore

$$g_k^2 p_k^2 - g_m^2 p_m^2 = \frac{\ln p_m^2 g_k^2 - \ln p_k^2 g_m^2}{2\gamma_s} \quad \forall k, m \in [1, N_t]$$

which implies

$$g_k^2 p_k^2 \stackrel{(\gamma_s \to \infty)}{=} g_m^2 p_m^2.$$

We obtain

$$\lim_{\gamma_s \to \infty} p_{{\rm MBER},k}^2 = \frac{g_k^{-2}}{\frac{1}{N_t}\sum_{l=1}^{N_t} g_l^{-2}}.$$

On the other hand, from (10), we have

$$\lim_{\gamma_s \to \infty} p_{{\rm AMBER},k}^2 = \frac{g_k^{-2}}{\frac{1}{N_t}\sum_{l=1}^{N_t} g_l^{-2}}$$

i.e., the AMBER solution asymptotically converges to the exact MBER solution.

3) AMBER Versus Uniform Power Allocation: By uniform power allocation, we mean the schemes described in Section II (with the power-allocation matrix $\mathbf{P} = \mathbf{I}_{N_t}$). A comparison between asymptotic performances of schemes with AMBER and uniform power allocation follows [17]. It is shown that power allocation improves error-rate performance at moderate-to-high SNR.

4) ZF Versus SIC: For MIMO with uniform power allocation, it has been shown in [2] that a SIC receiver outperforms a ZF receiver. We now compare the performances of ZF and SIC under power allocation. From (1) and (2), it is easily seen that

[2]We note that power allocation proposed in [13] is equivalent to (10), representing a special case of the general solution (8).

$g_{\mathrm{Z},1}^2 = g_{\mathrm{S},1}^2$. Consider $k = 2, \ldots, N_t$. Denote $\mathbf{h}_k$ as the kth column of matrix $\mathbf{H}$, and $\mathbf{H}_{\bar{k}}$ as the submatrix consisting of all but the kth column of $\mathbf{H}$. We have

$$g_{\mathrm{Z},k}^2 = [(\mathbf{H}^H\mathbf{H})^{-1}]_{k,k}^{-1} = \frac{\det(\mathbf{H}^H\mathbf{H})}{\det(\mathbf{H}_{\bar{k}}^H\mathbf{H}_{\bar{k}})} = \|\boldsymbol{\Upsilon}_{\mathbf{H}_{\bar{k}}}^{\perp}\mathbf{h}_k\|^2$$

where $\boldsymbol{\Upsilon}_{\mathbf{X}}^{\perp} \stackrel{\text{def}}{=} \mathbf{I} - \mathbf{X}(\mathbf{X}^H\mathbf{X})^{-1}\mathbf{X}^H$ is the orthogonal projection matrix. Similarly, $g_{\mathrm{S},k}^2 = \|\boldsymbol{\Upsilon}_{\mathbf{H}_{(k+1)}}^{\perp}\mathbf{h}_k\|^2$. Since $\mathbf{H}_{\bar{k}} = [\mathbf{h}_1 \vdots \cdots \vdots \mathbf{h}_{k-1} \vdots \mathbf{H}_{(k+1)}]$, we have $g_{\mathrm{Z},k}^2 \le g_{\mathrm{S},k}^2$ for $k = 2, \ldots, N_t$. From (12), we observe that $\tilde{P}_{\mathrm{PA}}(\gamma_s; \{g_k^2\})$ is a monotone decreasing function in g_k^2. Therefore, we conclude that $\tilde{P}_{\mathrm{PA}}(\gamma_s; \{g_{\mathrm{S},k}^2\}) \le \tilde{P}_{\mathrm{PA}}(\gamma_s; \{g_{\mathrm{Z},k}^2\})$, i.e., with power allocation, SIC outperforms ZF as well.

5) SIC Versus OSIC: Comparison between SIC and OSIC involves examining whether the SNR-based ordering is still optimal under power allocation. While exact analysis is difficult, we provide the following observations. At moderate-to-high SNR ($\gamma_s \gg 1$), ν_{a} in (10) can be approximated as $\nu_{\mathrm{a}} \approx N_t\gamma_s / (\sum_{k=1}^{N_t} g_k^{-2})$. We can further approximate (12) as

$$\begin{aligned}\tilde{P}_{\mathrm{PA}}\left(\gamma_s; \{g_k^2\}\right) &\approx \frac{1}{5}\left(\frac{1}{N_t}\sum_{k=1}^{N_t} g_k^{-2}\right) \\ &\quad \times \exp\left\{-\gamma_s\left(\frac{1}{N_t}\sum_{k=1}^{N_t} g_k^{-2}\right)^{-1}\right\} \\ &\stackrel{\text{def}}{=} \tilde{P}'_{\mathrm{PA}}\left(\gamma_s; \frac{1}{N_t}\sum_{k=1}^{N_t} g_k^{-2}\right). \end{aligned} \tag{13}$$

From (13), $\tilde{P}'_{\mathrm{PA}}(\gamma_s; (1/N_t)\sum_{k=1}^{N_t} g_k^{-2})$ is a monotone increasing function in $(1/N_t)\sum_{k=1}^{N_t} g_k^{-2}$, and is dominated by small g_k^2's. SNR-based ordering maximizes $\min_k\{g_k^2\}$. Therefore, heuristically, SNR-based ordering is expected to offer improved performance in power-allocation schemes. We verify this by simulation in Section V.

C. Remarks

1) AMBER Power Allocation Versus General Precoding: While other precoding schemes either apply an MBER criterion to ZF equalization [6], or use an MMSE criterion [3], [7], the proposed scheme provides a unified AMBER solution to power allocation for ZF, SIC, and OSIC receivers.

2) Feedback Overhead: In channels that lack reciprocity between uplink and downlink, precoding requires either CSI (N_tN_r complex coefficients) or precoding matrix (N_t^2 complex coefficients) feedback. On the other hand, using power allocation only N_t real coefficients are required at the transmitter, a factor of $(1/2N_r)$ (for CSI feedback) or $(1/2N_t)$ (for power feedback) savings.

3) Transceiver Complexity: Precoding schemes require diagonalization of a channel matrix [3], [6], [7]. The transmitter of precoding schemes performs matrix-vector multiplication. Using power allocation, operations performed at the transmitter are trivial. At the receiver, due to channel diagonalization, precoding requires lower complexity compared with SIC and OSIC with power allocation.

4) Quantization of Power Allocation: If power allocation is performed at the receiver and fed back to the transmitter, quantization of power is necessary, which results in quantization noise. An analysis of performance under imperfect (noisy) power feedback is given in Section IV-B.

5) Simplified Scenarios: Some aspects of power allocation for general MIMO spatial multiplexing are open problems, as described above. However, for the special case of two-input multiple-output (TIMO) systems, closed-form analytical results can usually be obtained [20].

IV. AMBER Power Allocation With Imperfect Feedback

Here we assume that perfect CSI is available at the receiver, while noisy feedback of CSI or allocated power is available at the transmitter. Feedback noise is modeled as a zero-mean Gaussian random variable. Such a noisy CSI model arises in, e.g., ML channel estimation [11]. A SIC receiver is considered. For an OSIC receiver, the analysis applies directly after ordering. Extension to ZF receivers is also straightforward.

A. Power Allocation With Noisy CSI Feedback

Since only the CSI feedback $\hat{\mathbf{H}}$ is available at the transmitter, the allocated power is a function of $\hat{\mathbf{H}}$, i.e., $\{p_k^2(\hat{\mathbf{H}})\}_{k=1}^{N_t}$, while the power gains are functions of perfect CSI at the receiver, i.e., $\{g_k^2(\mathbf{H})\}_{k=1}^{N_t}$. From (7), the approximate average BER is obtained as

$$\tilde{P}\left(\gamma_s; \mathbf{H}; \hat{\mathbf{H}}\right) = \frac{1}{5N_t}\sum_{k=1}^{N_t} \exp\{-\gamma_s p_k^2(\hat{\mathbf{H}}) g_k^2(\mathbf{H})\}. \tag{14}$$

By averaging both sides of (14) over the Gaussian conditional distribution $f_{\mathbf{H}|\hat{\mathbf{H}}}(\mathbf{H}|\hat{\mathbf{H}})$ as in [21], we obtain the approximate BER

$$\begin{aligned}\tilde{P}(\gamma_s; \hat{\mathbf{H}}) = \frac{1}{5N_t}\sum_{k=1}^{N_t}\int \exp\left\{-\gamma_s p_k^2(\hat{\mathbf{H}}) g_k^2(\mathbf{H})\right\} \\ \times f_{\mathbf{H}|\hat{\mathbf{H}}}(\mathbf{H}|\hat{\mathbf{H}}) d\mathbf{H}. \end{aligned} \tag{15}$$

Generally, it is difficult to find a closed-form expression for (15), due to nonlinearity of $g_k^2(\mathbf{H})$. In what follows, we study a special case when CSI at the transmitter is noisy, and obtain a closed-form error rate by using approximation techniques.

Noisy CSI is modeled by $[\mathbf{H}]_{m,n} = [\hat{\mathbf{H}}]_{m,n} + \epsilon_{m,n}^h$, for $1 \le m \le N_r, 1 \le n \le N_t$, where $\epsilon_{m,n}^h$ denotes a zero-mean complex Gaussian CSI feedback noise with variance σ_h^2. From Section III, power gains of SIC can be expressed as

$$g_k^2 = \left[\left(\mathbf{H}_{(k)}^H\mathbf{H}_{(k)}\right)^{\dagger}\right]_{k,k}^{-1} = \left\|\boldsymbol{\Upsilon}_{\mathbf{H}_{(k+1)}}^{\perp}\mathbf{h}_k\right\|^2 \tag{16}$$

for $1 \le k \le N_t$. Conditioned on $\hat{\mathbf{H}}$, $\mathbf{H}_{(k)}^H\mathbf{H}_{(k)}$ is distributed as complex noncentral Wishart [22]. While it is difficult to obtain a closed-form density function of g_k^2, which is the Schur complement of the (k,k)th entry of a noncentral Wishart matrix, we approximate the density function. In (16), we note that conditioned on $\hat{\mathbf{H}}$, both $\boldsymbol{\Upsilon}^{\perp}_{\mathbf{H}_{(k+1)}}$ and $\mathbf{h}_k$ are random, which makes analysis difficult. We therefore approximate

$$g_k^2 = \|\boldsymbol{\Upsilon}^{\perp}_{\mathbf{H}_{(k+1)}}\mathbf{h}_k\|^2 \approx \|\boldsymbol{\Upsilon}^{\perp}_{\hat{\mathbf{H}}_{(k+1)}}\mathbf{h}_k\|^2 \stackrel{\text{def}}{=} \tilde{g}_k^2$$

i.e., we use $\hat{\mathbf{H}}_{(k+1)}$ to approximate $\mathbf{H}_{(k+1)}$ at stage k.

Claim 1: The approximate power gain $\tilde{g}_k^2$, conditioned on $\hat{\mathbf{H}}$ has a noncentral chi-square density function with $2(N_r - N_t + k)$ degrees of freedom and noncentrality parameter $\hat{g}_k^2 \stackrel{\text{def}}{=} \hat{\mathbf{h}}_k^H \boldsymbol{\Upsilon}^{\perp}_{\hat{\mathbf{H}}_{(k+1)}}\hat{\mathbf{h}}_k$.

We now use $\tilde{g}_k^2$ to approximate g_k^2 in (15), and obtain an approximate BER in closed form. From the distribution of $\tilde{g}_k^2$ given by *Claim 1*, we obtain its characteristic function as [16]

$$\psi_{\tilde{g}_k^2}(j\omega) \stackrel{\text{def}}{=} \mathbb{E}\left\{e^{j\omega\tilde{g}_k^2}\right\} = \frac{\exp\left(\frac{j\omega\hat{g}_k^2}{1-j\omega\sigma_h^2}\right)}{(1-j\omega\sigma_h^2)^{N_r-N_t+k}}. \quad (17)$$

Using the characteristic function in (17), we can approximate the average BER (15) as

$$\begin{aligned}\tilde{P}\left(\gamma_s;\hat{\mathbf{H}};\sigma_h^2\right) &\approx \frac{1}{5N_t}\sum_{k=1}^{N_t}\int e^{-\gamma_s p_k^2\tilde{g}_k^2} f\left(\tilde{g}_k^2\middle|\hat{\mathbf{H}}\right)d\tilde{g}_k^2 \\ &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\frac{\exp\left(-\frac{\gamma_s p_k^2\hat{g}_k^2}{1+\gamma_s p_k^2\sigma_h^2}\right)}{(1+\gamma_s p_k^2\sigma_h^2)^{N_r-N_t+k}} \\ &\stackrel{\text{def}}{=} \tilde{\tilde{P}}\left(\gamma_s;\hat{\mathbf{H}};\sigma_h^2\right).\end{aligned} \quad (18)$$

1) Perfect CSI: For perfect CSI at the transmitter, $\hat{\mathbf{H}} = \mathbf{H}$ and $\sigma_h^2 = 0$, and (18) reduces to

$$\begin{aligned}\tilde{\tilde{P}}(\gamma_s;\hat{\mathbf{H}};0) &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\exp\left(-\gamma_s p_k^2\mathbf{h}_k^H\boldsymbol{\Upsilon}^{\perp}_{\mathbf{H}_{(k+1)}}\mathbf{h}_k\right) \\ &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\exp\left(-\gamma_s p_k^2 g_k^2\right)\end{aligned}$$

which is the approximate BER for power allocation with perfect feedback.

2) Asymptotic Performance: Since

$$\lim_{\gamma_s\to\infty}\frac{\tilde{\tilde{P}}\left(\gamma_s;\hat{\mathbf{H}};\sigma_h^2\right)}{\gamma_s^{-(N_r-N_t+1)}} = \frac{1}{5N_t}\frac{\exp\left(-\hat{g}_1^2/\sigma_h^2\right)}{(p_1^2\sigma_h^2)^{(N_r-N_t+1)}} \stackrel{\text{def}}{=} \kappa_h$$

we have, for $\gamma_s \gg 1, \tilde{\tilde{P}}(\gamma_s;\hat{\mathbf{H}};\sigma_h^2) \approx \kappa_h \cdot \gamma_s^{-(N_r-N_t+1)}$, decreasing exponentially as $(N_r - N_t + 1)$th power of γ_s. Therefore, the performance improvement increases with the inherent diversity order. On the other hand, from (4), the BER of MIMO with uniform power allocation can be approximated as $\bar{P}(\gamma_s;\{g_k^2\}) \approx (1/5N_t)\sum_{k=1}^{N_t} e^{-\gamma_s g_k^2}$, decreasing exponentially in γ_s. Thus, for sufficiently large γ_s, power-allocation schemes with noisy CSI at the transmitter are inferior to MIMO with uniform power allocation, which is not affected by noisy CSI feedback.

B. Power Allocation With Noisy Power Feedback

Denote $\mathbf{p} = [p_1, p_2, \ldots, p_{N_t}]^T$ and $\hat{\mathbf{p}} = [\hat{p}_1, \hat{p}_2, \ldots, \hat{p}_{N_t}]^T$, where $\hat{p}_k$ denotes a noisy feedback of power. Noisy power feedback is modeled as $\hat{\mathbf{p}} = \mathbf{p} + \boldsymbol{\epsilon}^p$, where $\boldsymbol{\epsilon}^p$ is a noise vector with distribution $\mathcal{N}(\mathbf{0},\mathbf{R})$. With noisy power, the conditional approximate average BER can be written as shown in (19) at the bottom of the page.

By averaging both sides of (19) over the distribution of $\boldsymbol{\epsilon}^p$, the approximate BER is obtained in (20), shown at the bottom of the page, where $\mathbf{e}_k$ denotes the kth column of $\mathbf{I}_{N_t}$. The derivation follows [21]. When the noise terms ϵ_k^p's are independent and identically distributed as $\mathcal{N}(0,\sigma_p^2)$, we have

$$\begin{aligned}\tilde{P}(\gamma_s;\{g_k^2\}) &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\frac{\exp\left\{-\frac{\gamma_s p_k^2 g_k^2}{1+2\gamma_s g_k^2\sigma_p^2}\right\}}{(1+2\gamma_s g_k^2\sigma_p^2)^{1/2}} \\ &\stackrel{\text{def}}{=} \tilde{\tilde{P}}\left(\gamma_s;\left\{g_k^2\right\};\sigma_p^2\right).\end{aligned} \quad (21)$$

1) Perfect Power Feedback: For perfect power feedback, $\hat{\mathbf{p}} = \mathbf{p}$ and $\sigma_p^2 = 0$, and the above analysis does not apply because $\mathbf{R} = \mathbf{0}$. However, from (21), we have the limiting case of high-quality feedback

$$\lim_{\sigma_p^2\to 0}\tilde{\tilde{P}}\left(\gamma_s;\left\{g_k^2\right\};\sigma_p^2\right) = \frac{1}{5N_t}\sum_{k=1}^{N_t}\exp\left(-\gamma_s p_k^2 g_k^2\right)$$

which reduces to the approximate BER for power allocation with perfect feedback. Therefore, (21) includes perfect power feedback as a special case.

$$\tilde{P}(\gamma_s;\{g_k^2\};\{\hat{p}_k^2\}) = \frac{1}{5N_t}\sum_{k=1}^{N_t}\exp\left\{-\gamma_s\hat{p}_k^2 g_k^2\right\} \quad (19)$$

$$\begin{aligned}\tilde{P}(\gamma_s;\{g_k^2\};\{p_k^2\}) &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\int \exp\{-\gamma_s(p_k+\epsilon_k^p)^2 g_k^2\} f_{\boldsymbol{\epsilon}^p}(\boldsymbol{\epsilon}^p)d\boldsymbol{\epsilon}^p \\ &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\int\frac{\exp\left\{-\gamma_s g_k^2(\mathbf{p}+\boldsymbol{\epsilon}^p)^T\mathbf{e}_k\mathbf{e}_k^T(\mathbf{p}+\boldsymbol{\epsilon}^p) - \frac{1}{2}(\boldsymbol{\epsilon}^p)^T\mathbf{R}^{-1}\boldsymbol{\epsilon}^p\right\}}{(2\pi)^{N_t/2}[\det(\mathbf{R})]^{1/2}}d\boldsymbol{\epsilon}^p \\ &= \frac{1}{5N_t}\sum_{k=1}^{N_t}\left(1+2\gamma_s g_k^2[\mathbf{R}]_{k,k}\right)^{-1/2}\exp\left\{-\frac{\gamma_s p_k^2 g_k^2}{1+2\gamma_s g_k^2[\mathbf{R}]_{k,k}}\right\}\end{aligned} \quad (20)$$

2) Asymptotic Performance: From (21), we have

$$\lim_{\gamma_s \to \infty} \frac{\tilde{\tilde{P}}(\gamma_s; \{g_k^2\}; \sigma_p^2)}{\gamma_s^{-1/2}} = \frac{1}{5N_t} \sum_{k=1}^{N_t} \frac{\exp\left\{-\frac{p_k^2}{2\sigma_p^2}\right\}}{\sqrt{2} g_k \sigma_p} \stackrel{\text{def}}{=} \kappa_p$$

i.e., when $\gamma_s \gg 1, \tilde{\tilde{P}}(\gamma_s; \{g_k^2\}; \sigma_p^2) \approx \kappa_p \cdot \gamma_s^{-1/2}$, decreasing in $\gamma_s^{-1/2}$. Again, in view of (4), for sufficiently large γ_s, power-allocation schemes with noisy power feedback are inferior to MIMO with uniform power allocation.

C. Power Allocation Using Feedback Noise Variance

When knowledge of the variance of noisy power feedback σ_p^2, or noisy CSI σ_h^2, is available at the transmitter, power allocation can be modified to take feedback noise variance into account as in [8]. To this end, a constrained optimization problem, referred to as *modified AMBER power allocation*, can be formulated as

$$\begin{cases} \min \tilde{\tilde{P}} \\ \text{subject to} \quad \sum_{k=1}^{N_t} p_k^2 = N_t \end{cases} \tag{22}$$

where $\tilde{\tilde{P}}$ is the objective function from (18) or (21). It can be verified that $(d^2\tilde{\tilde{P}}/d(p_k^2)^2) > 0$, i.e., $\tilde{\tilde{P}}$ is convex in p_k^2's. A solution to the convex optimization problem (22) is given by [12]

$$p_k^2 = (\phi_k)_+ \tag{23}$$

where ϕ_k is the root of the equation $(d\tilde{\tilde{P}}/d(p_k^2)) = \mu$, and μ is chosen to satisfy the transmit power constraint. By noting the normalized transmit power constraint $\sum_{k=1}^{N_t} p_k^2 = N_t$, we have $\min_k p_k^2 \le 1 \le \max_k p_k^2$, and the parameter μ can be bounded as

$$\min_k \left[d\tilde{\tilde{P}}/d\left(p_k^2\right)\right]_{p_k^2=1} \le \mu \le \max_k \left[d\tilde{\tilde{P}}/d\left(p_k^2\right)\right]_{p_k^2=1}.$$

An iterative algorithm can be used to solve this problem numerically.

V. Numerical Results and Discussions

Performance of transmission methods discussed earlier are simulated and compared with MBER precoding with ZF equalization [6], as well as optimal MMSE precoding/decoding using trace criterion [3] in fading channels. The realization for a Ricean MIMO channel is modeled as

$$\mathbf{H} = \sqrt{\frac{K}{1+K}}\mathbf{H}_d + \sqrt{\frac{1}{1+K}}\mathbf{H}_s$$

where Ricean K-factor is defined as the ratio of deterministic-to-scattered power, and $\mathbf{H}_d$ and $\mathbf{H}_s$ denote the deterministic and scattered components, respectively [23]. The deterministic component $\mathbf{H}_d$ is modeled as $\mathbf{H}_d = \mathbf{a}_r(\theta_r)\mathbf{a}_t(\theta_t)^T$, where $\mathbf{a}_r(\theta_r) \stackrel{\text{def}}{=} [1\ e^{-j2\pi d_r \sin(\theta_r)}\ \cdots\ e^{-j2\pi d_r \sin((N_r-1)\theta_r)}]^T$ and $\mathbf{a}_t(\theta_t) \stackrel{\text{def}}{=} [1\, e^{-j2\pi d_t \sin(\theta_t)}\ \cdots\ e^{-j2\pi d_t \sin((N_t-1)\theta_t)}]^T$ are array response vectors for uniform linear receiver and transmitter antenna arrays, θ_r and θ_t are the angles of arrival and departure of the deterministic component, and d_r and d_t are the receiver and transmitter antenna spacing expressed in wavelength. The entries of the scattered component are modeled as zero-mean complex Gaussian random variables, with cross-correlations determined by antenna array geometry, angle spread, mean direction of signal arrival/departure, etc. In our simulations, we adopt the spatial fading correlation model for general non-isotropic scattering given in [14]. The following parameters are chosen: $N_t = 4$ transmit and $N_r = 8$ receive antennas; antenna spacings are $d_t = 0.5$ and $d_r = 10$; angles of arrival/departure of deterministic component are $\pi/6, 0$, respectively; angle spread $10°$; $K = 8$ dB for Ricean fading channels; and BPSK modulation is used for the purposes of comparison with [6].

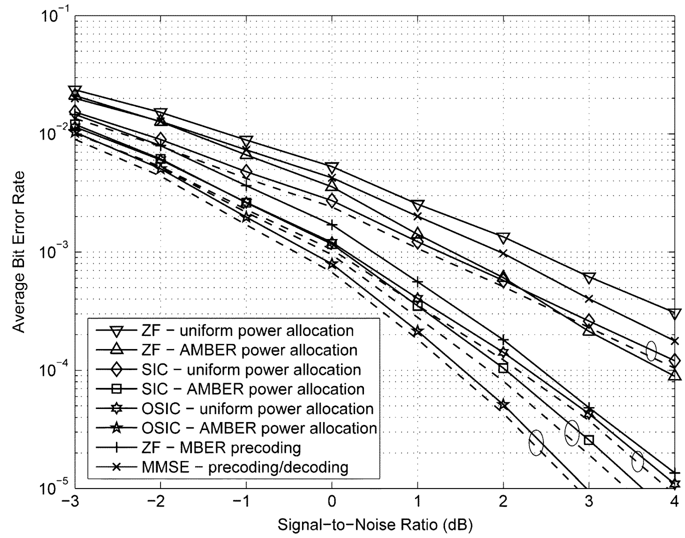

Fig. 1. Average BER performance in uncorrelated Rayleigh fading MIMO channel ($N_t = 4, N_r = 8$). Dashed curves associated with SIC and OSIC curves stand for BER without error propagation.

A. AMBER Power Allocation With Perfect Feedback

1) Rayleigh Fading: Fig. 1 is a plot of the average uncoded BER of different transceivers in an uncorrelated Rayleigh fading channel. We observe that at a BER of 10^{-3}, AMBER power allocation offers 1.0, 1.1, and 0.4 dB SNR gains over ZF, SIC, and OSIC receivers, respectively. At all SNRs shown, MMSE precoding/decoding offers performance between that of ZF with uniform and AMBER power allocation, while MBER precoding with ZF equalization has performance between that of SIC with uniform and AMBER power allocation.

2) Ricean Fading: In Fig. 2, we illustrate average BERs in correlated Ricean fading channels. It is observed that at a BER of 10^{-3}, SNR gains offered by AMBER power allocation for SIC and OSIC are 2.7 and 1.6 dB, respectively. We also observe that MMSE precoding/decoding has performance similar to that of SIC with uniform power allocation, while performances of MBER precoding with ZF equalization and OSIC with uniform power allocation are nearly identical.

Remarks:

- *Effects of Error Propagation*: The results for SIC and OSIC discussed above take error propagation into account, i.e., actual decisions are used for interference cancellation. Performances of SIC and OSIC with and without error propagation are shown in solid and dashed

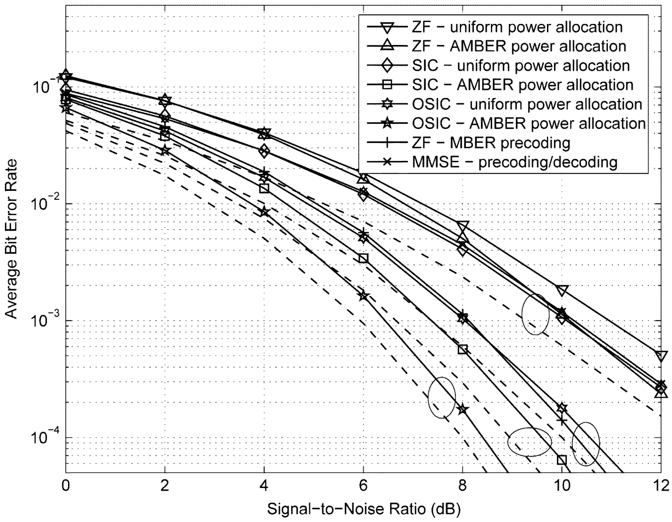

Fig. 2. Average BER performance in correlated Ricean fading MIMO channel ($N_t = 4, N_r = 8, K = 8$ dB). Dashed curves associated with SIC and OSIC curves stand for BER without error propagation.

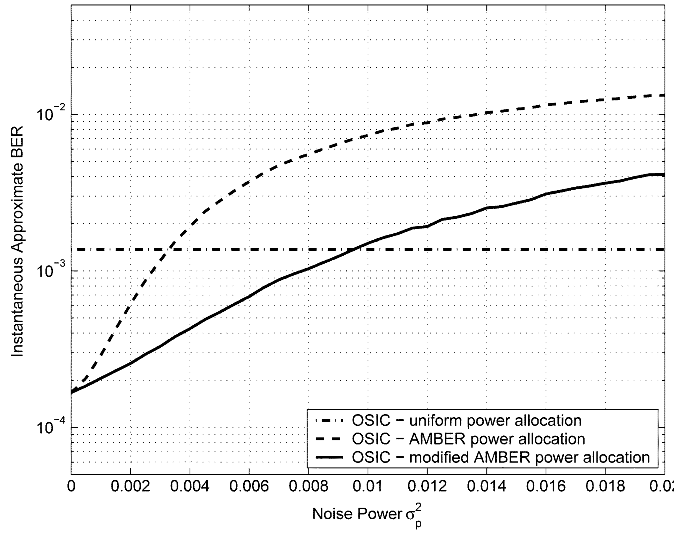

Fig. 3. Example of approximate instantaneous BER versus noise variance of power feedback ($N_t = 4, N_r = 8, K = 8$ dB, $\gamma_s = 8$ dB).

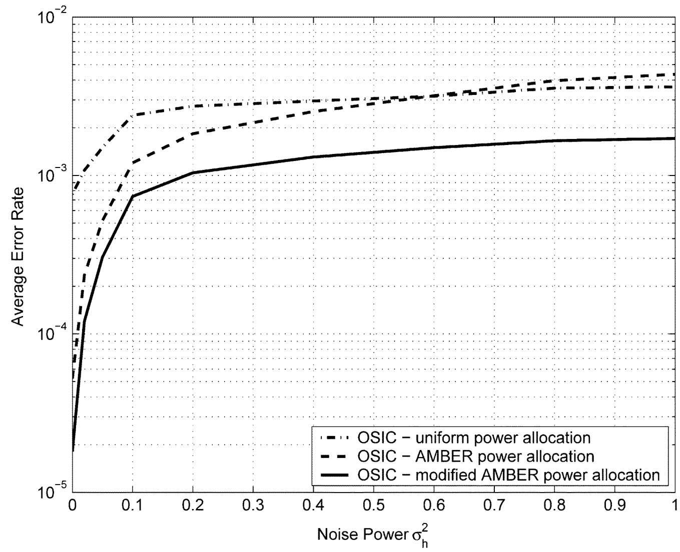

Fig. 4. Average BER performance versus noisy CSI variance in correlated Ricean fading MIMO channel ($N_t = 4, N_r = 8, K = 8$ dB, $\gamma_s = 10$ dB).

curves, respectively, in Figs. 1 and 2. We can clearly see the advantage of the proposed AMBER power allocation and the relatively modest effects of error propagation.

- *Comparison Among ZF, SIC, and OSIC with AMBER Power Allocation*: For all channels simulated and employing AMBER power allocation, SIC outperforms ZF and OSIC outperforms SIC, which agrees with the heuristic results in Section III-C.
- *AMBER Power Allocation versus Ordering for SIC*: From all simulations, it is also observed that SIC with AMBER power allocation outperforms OSIC with uniform power allocation, i.e., AMBER power allocation outperforms SNR-based ordering for SIC receivers.

B. AMBER Power Allocation With Imperfect Feedback

1) Noisy Power Feedback: Fig. 3 shows an example of instantaneous approximate BER of OSIC with uniform power allocation, with AMBER power allocation (8), and with modified AMBER power allocation (23) as a function of feedback power noise variance. The channel is randomly generated with Ricean distribution. From Fig. 3, when power feedback noise variance σ_p^2 is larger than 0.003, AMBER power allocation (8) is inferior to uniform power allocation. Using modified AMBER power allocation (23), tolerance to feedback noise power increases to $\sigma_p^2 = 0.009$.

2) Noisy CSI: Fig. 4 depicts average BER performances of OSIC transceivers with uniform power allocation, with AMBER power allocation (8), and with modified AMBER power allocation (23), respectively, in a correlated Ricean fading channel. Perfect knowledge of noise variance σ_h^2 is assumed. Ordering is carried out at the transmitter based on noisy CSI. We observe that performance of OSIC with uniform power allocation also degrades with an increase of CSI noise power. Also, when the CSI noise variance is larger than 0.6, AMBER power allocation (8) has performance inferior to that of uniform power allocation. At all CSI noise variances shown, modified power allocation (23) outperforms the other OSIC methods.

Remark: From Fig. 4, we also observe that when $\sigma_h^2 = 0$, i.e., perfect CSI, using (23) outperforms (8). This can be explained as follows: the modified power-allocation solution for perfect CSI is given by $(p'_k)^2 = (\phi_k)_+$, where ϕ_k is the solution to

$$\left.\frac{d\tilde{\tilde{P}}(\gamma_s; \hat{\mathbf{H}}; \sigma_h^2)}{d\left({p'_k}^2\right)}\right|_{\sigma_h^2=0} = \mu$$

which is equivalent to

$$(p'_k)^2 = \left(\frac{\ln\left[g_k^2 - \gamma_s^{-1}(N_r - N_t + k)\right] + \mu'}{\gamma_s g_k^2}\right)_+ \quad (24)$$

where μ and μ' are chosen to satisfy the transmit power constraint. Comparing (24) with AMBER power allocation using (8), it is obvious that in the modified scheme, more power is allocated to earlier SIC stages. This change has the benefit of reducing error propagation from earlier stages to later ones, which improves the error-rate performance. Note that the recursive algorithm for (8) can be easily adapted to solve (24) without an increase in complexity, resulting in a modified power-allocation algorithm that takes error propagation of interference cancellation into account.

VI. Conclusion

Power allocation using an AMBER criterion for MIMO spatial multiplexing is studied in this paper. AMBER power-allocation schemes for a variety of receiver structures have been proposed. Compared with existing precoding schemes, the proposed schemes reduce both transmitter complexity and feedback overhead significantly. This method is motivated by an approximate BER analysis, which is also used to develop an AMBER power-allocation scheme that uses the variance of the feedback noise. Simulation results show that the proposed power-allocation method improves performance of ZF, SIC, and OSIC receivers. In particular, SIC and OSIC employing AMBER power allocation have the potential to offer superior performances over some existing precoding schemes, e.g., in a correlated Ricean fading channel, at a BER of 10^{-3}, AMBER power allocation for OSIC offers, respectively, 1.7 and 3.8 dB SNR gains over MBER precoding with ZF equalization and MMSE precoding/decoding.

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