Transmit Power Allocation for Successive Interference Cancellation in Multicode MIMO Systems

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Abstract—Multiple-input multiple-output (MIMO) system with multicode transmission can provide high speed data services by transmitting independent parallel substreams from multiple antennas and through multicode channelization. In this paper, we first introduce an iterative two-stage successive interference cancellation (SIC) detection scheme for a multicode MIMO system. Next, we derive various transmit power allocation schemes over different data substreams for the proposed detection process to improve error rate performance. The power allocation is performed to make the signal-to-interference-plus-noise ratio (SINR) become balanced for all data substreams. Numerical results show that the proposed transmit power allocation schemes for the two-stage SIC significantly outperform the equal power allocation scheme.

I. INTRODUCTION

Multiple-input multiple-output (MIMO) systems, which employ multiple antennas at both the transmitter and receiver, have drawn considerable attention in recent years due to its enormous capacity enhancement [1], [2]. On the other hand, multicode channelization scheme has been proposed for a high rate transmission in a code division multiple access (CDMA) system [3]. Thus, a MIMO system combined with a multicode transmission, referred to as multicode MIMO system in this paper, may be promising for providing high speed data services [4].

In a frequency-selective fading channel, the orthogonality between two different spreading codes cannot be preserved for the received signals with different path delays. This interference between multicode signals is called multicode interference (MCI) and results in the performance degradation. In [4], space-time detector based on a group detection technique has been introduced for a cellular MIMO CDMA system with multicode transmission, and mitigated the MCI. However, the detection scheme in [4] may cause huge computational complexity when the number of data substreams $D$ is large, since the detection needs a $D \times D$ matrix inversion operation.

In this paper, we first introduce an iterative two-stage successive interference cancellation (SIC) detection scheme for a multicode MIMO system. This two-stage approach makes the matrix dimension for inverse operation become limited to the number of data substreams in the space domain for a specific spreading code, not to that of all the parallel data substreams in both space and code domains as in [4]. Next, we derive various transmit power allocation schemes over different data substreams for the proposed two-stage SIC detection process. The power allocation is performed to make the signal-to-interference-plus-noise ratio (SINR) become balanced for all data substreams, based on the notion that the substream with the smallest SINR may dominate the overall error performance of the system.

The performance of the proposed power allocation schemes for the two-stage SIC detection is evaluated, and compared with that of the equal power allocation scheme. Numerical results show that the use of the proposed power allocation schemes significantly improves the bit error rate (BER) performance of a multicode MIMO system.\(^1\)

II. SYSTEM AND CHANNEL MODELS

A multicode MIMO communication system considered in this paper is depicted in Fig. 1. We assume that the transmitter and the receiver are equipped with $M$ and $N$ antennas respectively, and $K$ spreading codes are used for multicode channelization under single user environment. An input data stream is divided into $KM$ parallel substreams through a serial-to-parallel converter. The substreams are partitioned into $M$ groups. Each group consists of $K$ data substreams, which are spread using different spreading waveforms $c_k(t)$ $(k = 1, 2, \ldots, K)$ and are transmitted from the same transmit antenna. The spreading waveforms are reused for the substreams at all the other transmit antennas. Hence, the complex baseband transmitted signal of the $m$th transmit antenna for one data symbol duration $T$, may be expressed as

$$s_m(t) = \sum_{i=1}^{K} \sqrt{P_{m,i}} d_{i,m} c_{i,m}^*(t), \quad m = 1, 2, \ldots, M, \quad 0 \leq t \leq T$$

where $P_{m,i}$ is the transmit power for the $k$th spreading code at the $m$th transmit antenna. A quadrature phase-shift keying (QPSK) is assumed to be employed for a modulation scheme. Hence, the encoded data symbol $d_{i,m}$ takes one of four complex values in $\{(1 \pm j)/\sqrt{2}\}$ with equal probability. $c_k(t)$ is the spreading waveform for the $k$th code given as

$$c_k(t) = \sum_{j=0}^{G-1} c_{k,j} e^{j\pi/2}, \quad k = 1, 2, \ldots, K$$

where $G = T/T_c$ is the spreading gain, $T_c$ is the chip duration, $c_{k,j}$ is the $j$th chip for the $k$th code, and $\psi(t)$ is the chip pulse shape which is assumed to be rectangular, i.e., one for $0 \leq t \leq T_c$ and zero otherwise. The chip sequence $\{c_{k,j}\}$ is assumed to be a complex spreading sequence, and is given by $c_{k,j} = c_{k,j}^{(0)} + j c_{k,j}^{(1)}$, where $\{c_{k,j}^{(0)}\}$ and $\{c_{k,j}^{(1)}\}$ take on the random values of $+1/\sqrt{2}$ and $-1/\sqrt{2}$ with equal probability. Moreover, $\{c_{k,j}\}$ are mutually orthogonal for all $k$, i.e., $\sum_{j=0}^{G-1} c_{k,j}^* c_{k',j} = 0$ for $k \neq k'$, where the superscript $*$ denotes the complex conjugate. The total power constraint is given as

$$\sum_{k=1}^{K} \sum_{i=1}^{M} P_{i,m} = P_T$$

where $P_T = \sum_{m=1}^{M} \sum_{i=1}^{N} P_{i,m}$.

We use the tapped delay line multipath channel model [5]. Thus, the complex channel impulse response from the $m$th transmit antenna to the $p$th receive antenna may be expressed as

$$h_{p,m}(t) = \sum_{l=0}^{L-1} h_{p,m} \delta(t - lT)$$

where $L$ is the number of resolvable multipath components. $h_{p,m}$

\(^1\) Recently, the authors have proposed an effective scheme for multicode MIMO systems in the 3rd generation partnership project (3GPP) standardization (included in 3GPP TR 25.876 v1.4.0).
denotes the fading coefficient from the mth transmit antenna to the pth receive antenna for the kth multipath, and they are assumed to be circularly symmetric complex Gaussian random variables with zero mean. It is assumed that all \( h_{p,m,l} \)'s are constant over several symbol duration, and are independent for all \( p, m, \) and \( l \). The second moment of channel amplitude \( E[|h_{p,m,l}|^2] = \Omega_p \) is assumed to have the exponential multipath intensity profile (MIP) as

\[
\Omega_p = E[|h_{p,m,l}|^2] = \Omega_p e^{\delta l}, \quad l = 0, 1, \cdots, L - 1
\]

where \( E[\cdot] \) denotes the expectation and the parameter \( \delta \) represents the rate of the exponential decay of the average path power. The small delay spread relative to the symbol interval is assumed, so that the effect of the intersymbol interference (ISI) on the error performance is negligible compared to that of the intrasymbol interference. Under this assumption, the ISI is ignored in this paper.

Correspondingly, the received complex baseband equivalent signal at the pth \( (p = 1, 2, \cdots, N) \) receive antenna may be expressed as

\[
r_p(t) = \sum_{m=1}^{M} \sum_{l=0}^{L-1} h_{p,m,l} s_m(t-lT_c) + w_p(t)
\]

(6)

where \( w_p(t) \) is the additive white Gaussian noise (AWGN) at the pth receive antenna with one-sided power spectral density \( \sigma^2 \).

Throughout this paper, the superscripts \([\cdot]^T\) and \([\cdot]^H\) denote the transpose and conjugate transpose, respectively. Moreover, \([\cdot]_{p,q}\) denotes the element in the pth row and qth column of a matrix, and \([\cdot]_{p,q} \) represents the Euclidean norm of a vector.

The channel state information (CSI) is estimated and assumed to be perfectly known at the receiver. As depicted in Fig. 1, the transmit power values are calculated at the receiver using the CSI, and fed back to the transmitter through the feedback channel.

III. DETECTION PROCESS

A. Output Signal of a Correlator Bank Without SIC

Prior to the SIC processing, we describe the overall output signal of the correlator bank for the kth spreading code, which is assumed to be the first code for signal detection. The correlator output of the lth rake finger at the pth receive antenna for the kth spreading code \((k = 1, \cdots, K, \quad p = 1, \cdots, N, \quad l = 0, \cdots, L - 1)\) may be written as

\[
z_{b_k,p,l} = C_{b_k,l}^T (t-lT_c) dt.
\]

Then, the NL-dimensional correlator output vector for the kth spreading code, \( z_{b_k} \), may be expressed as

\[
z_{b_k} = [z_{b_k,1,0}, z_{b_k,1,1}, \cdots, z_{b_k,1,L-1}, z_{b_k,2,0}, \cdots, z_{b_k,N,L-1}]^T
\]

(8)

where

\[
Y_{b_k} = [Y_{b_k,1}^T, Y_{b_k,2}^T, \cdots, Y_{b_k,N}^T]^T,
\]

\[
Y_{b_k,1} = \sum_{l=1}^{L-1} h_{p,m,l} R_{b_k}(l) \cdots h_{p,m,l} R_{b_k}(l - (i - 1)),
\]

(i.e., \(Y_{b_k,1,p,l} = \sum_{j=0}^{L-1} h_{p,j,l} R_{b_k}(l - (j - i))\))

\[
R_{b_k}(l) \triangleq \int_{T_c}^{T_c+T_c} c_{b_k}^*(t) c_{b_k}^*(t-lT_c) dt.
\]

\[
\begin{align*}
\mathbf{P}_{b_k} &= \text{diag}\left(\sqrt{P_{b_k,1}}, \sqrt{P_{b_k,2}}, \cdots, \sqrt{P_{b_k,N}}\right), \\
\mathbf{d}_{b_k} &= [d_{b_k,1}, d_{b_k,2}, \cdots, d_{b_k,M}]^T, \\
\mathbf{n}_{b_k} &= [n_{b_k,1,0}, n_{b_k,1,1}, \cdots, n_{b_k,1,L-1}, n_{b_k,2,0}, \cdots, n_{b_k,N,L-1}]^T,
\end{align*}
\]

and

\[
\begin{align*}
n_{b_k,p,l} &= \int_{T_c}^{T_c+T_c} w_{p,l}(t) c_{b_k}^*(t-lT_c) dt. \\
\sum_{k=1}^{K} \sum_{m=1}^{M} Y_{b_k,m} \mathbf{P}_{b_k} + \mathbf{n}_{b_k} \text{ in (8) is the MCI-plus-noise vector for the } k \text{th code whose covariance matrix may be expressed as}
\end{align*}
\]

\[
E \left[ \begin{bmatrix} \sum_{k=1}^{K} \sum_{m=1}^{M} Y_{b_k,m} \mathbf{P}_{b_k} + \mathbf{n}_{b_k} \end{bmatrix} \begin{bmatrix} \sum_{k=1}^{K} \sum_{m=1}^{M} Y_{b_k,m} \mathbf{P}_{b_k} + \mathbf{n}_{b_k} \end{bmatrix}^H \right] = \sum_{k=1}^{K} \sum_{m=1}^{M} \mathbf{P}_{b_k} + \mathbf{R}_{b_k}^{\text{MCI}} + \mathbf{R}_{b_k}^{\text{noise}}
\]

where \( \mathbf{R}_{b_k}^{\text{MCI}} = E[\mathbf{Y}_{b_k,m} \mathbf{P}_{b_k} (\mathbf{Y}_{b_k,m} \mathbf{P}_{b_k})^H] \) and \( \mathbf{R}_{b_k}^{\text{noise}} = E[\mathbf{n}_{b_k} \mathbf{n}_{b_k}^H] \).

Although the crosscorrelations \( R_{b_k,l}(k) \) between the different spreading waveforms can be computed at the receiver, we average these values over random orthogonal code sequences to find \( \mathbf{R}_{b_k}^{\text{MCI}} \) in (9) for computational efficiency. The expectation of the product of \( R_{b_k,l}(k) \) can be found as

\[
E \left[ R_{b_k,l}(k) R_{b_k,l}(j) \right] = \begin{cases} G^2 T_c^2, & k = j = 0 \\
(G - |l|) T_c^2, & i = j = 0, |l| \leq G - 1 \\
0, & \text{else}
\end{cases}
\]

(10)

Using (10), we can find the element in the \( i^\prime = (L(p-i)+1) \) th row and \( j^\prime = (L(q-j)+1) \) th column of \( \mathbf{R}_{b_k,l}^{\text{MCI}} \) as

\[
[R_{b_k,l}]_{i^\prime,j^\prime} = \left\{ \begin{array}{ll} \sum_{m=1}^{M} \sum_{l=0}^{L-1} h_{p,m,l} R_{b_k}(l) (G - |l-i|) T_c^2, & i = j, j = 0 \leq j \leq G - 1 \leq \frac{L}{2} \\
\sum_{m=1}^{M} \sum_{l=0}^{L-1} h_{p,m,l} R_{b_k}(l) (G - |l-j|) T_c^2, & j = i, j = 0 \leq j \leq G - 1 \leq \frac{L}{2} \end{array} \right.
\]

where \( k_0 \neq k, \quad p, q = 1, 2, \cdots, N, \quad i, j = 1, 2, \cdots, L. \)

(11)

Covariance matrix of noise vector, \( \mathbf{R}_{b_k}^{\text{noise}} \), can be found as

\[
\mathbf{R}_{b_k}^{\text{noise}} = \sigma^2 \text{diag}(\hat{R}_{b_k}, \hat{R}_{b_k}, \cdots, \hat{R}_{b_k})
\]

(12)

where \( \hat{R}_{b_k} \) is the \( L \times L \) correlation matrix for the spreading waveforms of the \( k \)th code, with elements

\[
[R_{bb}]_{i,j} = \int_{T_c}^{T_c+T_c} c_{b_k}^*(t-iT_c) c_{b_k}^*(t-jT_c) dt = R_{b_k,b_k}(j-i).
\]

B. Two-Stage SIC Scheme

Based on the results in the previous subsection, an iterative two-stage SIC scheme is described in this subsection. The receiver structure for detection process is depicted in Fig. 2. In the first stage, space domain SIC (SD-SIC) successively cancels the intracode interference caused by the data substreams from the different transmit antennas for a specific spreading code. Here, the multipath diversity reception by rake receiver is regarded as the virtual receive antenna diversity reception. In the second stage, code domain SIC (CD-SIC) cancels the MCI. In this processing, the contributions of pre-detected code domain signals are subtracted from the received signal after the space domain signals for each code channel are all detected through SD-SIC in the previous stage. If we assume that the detection order of code domain signals is from 1 to \( K \), i.e., any particular code ordering criterion is not used, the MCI cancelled correlator output of the lth rake finger at the pth receive antenna for
the $k$th spreading code, $\tilde{z}_{k,p,j}$, may be written as

$$\tilde{z}_{k,p,j} = \int_{T_c}^{T_c+T} \left( r_i(t) - \sum_{k=1}^{L-1} \sum_{l=0}^{K} P_{k,l} h_{p,m} \hat{d}_{k,m} c_j(t-iT_c) \right) \cdot e^{j\omega_n t} \, dt \quad (13)$$

where $\hat{d}_{k,m}$ is the hard estimate for the data symbol $d_{k,m}$ in the previous SD-SIC detection stage. Then, the MCI cancelled NL-dimensional correlator output vector for the $k$th spreading code, $\tilde{z}_{k,i}$, may be expressed as

$$\tilde{z}_{k,i} = [\tilde{z}_{k,i-1}, \tilde{z}_{k,i-2}, \ldots, \tilde{z}_{k,i-L+1}]^T = \sum_{k=1}^{K} Y_{k,i} P_{k,i} \hat{d}_{k,i} + n_{k,i} \quad (14)$$

where $\hat{d}_{k,i}$ is the quantization operation $d_{k,i}$ for $k_0 \neq k$, as indicated in (13).

This two-stage SIC described above is recursively performed until all data symbols are detected. The overall detection procedure can be presented as follows. For simplicity, zero-forcing (ZF) criterion is used to calculate the linear weighting vector of SD-SIC [6].

Step 1) Initialization for CD-SIC:

$k_0 = 1$

$$\tilde{z}_1 = z_1 = Y_{1,1} P_{1,1} \hat{d}_{1,1} + \sum_{k=2}^{K} Y_{1,k} P_{1,k} \hat{d}_{k,1} + n_1 \quad (15a)$$

Step 2) Initialization for SD-SIC:

$i = 1$

$$\tilde{z}_i, (1) = \tilde{z}_{k_0}, \quad G_{k_0}(1) = \left( \tilde{z}_{k_0} \right)^*$$

$$g(1) = \arg\min_j \left[ G_{k_0}(1) R_{\text{order}}^{\text{MC},j} \left( G_{k_0}(1) \right)^H \right]_{j,j} \quad (15b)$$

Step 3) SD-SIC for the $(g)$th substream of the $k_0$th code:

$$w_{k_0,g}(g) = [G_{k_0}(g)]^H$$

$$\tilde{z}_{k_0}(i+1) = \tilde{z}_{k_0}(i) - \frac{P_{k_0,g}(g)}{P_{k_0,g}(g)} \hat{d}_{k_0,g}(g) \left( Y_{k_0,k_0} \right)_{i,M} \quad (15d)$$

$$G_{k_0}(i+1) = \left( \left( Y_{k_0,k_0} \right)_{i,M} \right)^* \quad (15e)$$

$$g(i+1) = \arg\min_{j\neq g} \left[ G_{k_0}(i+1) R_{\text{order}}^{\text{MC}}(j) \left( G_{k_0}(i+1) \right)^H \right]_{j,j} \quad (15f)$$

Step 4) Repetition or termination for SD-SIC: If $i < M$, increase $i$ by one and go to Step 3. Otherwise, go to Step 5.

Step 5) CD-SIC for the $k_0$th code:

$$\tilde{z}_{k_0+1} = \sum_{k=1}^{K} Y_{k_0,k_0+1} \hat{d}_{k_0+1} + \sum_{k=1}^{K} Y_{k_0,k} P_{k,k} \hat{d}_{k} + n_{k_0+1} \quad (15g)$$

Step 6) Repetition or termination for CD-SIC: If $k_0 < K$, increase $k_0$ by one and go to Step 2. Otherwise, terminate the whole detection procedure.

where $\hat{d}_{k,i}$ denotes the Moore-Penrose pseudo-inverse, $\left[ \cdot \right]_{g,g}(g)$ is the $(g)$th row of a matrix, and $O(\cdot)$ is the quantization operation appropriate to the modulation scheme. $\left[ \cdot \right]^*$ denotes the $(g)$th column of a matrix, and $\left[ \cdot \right]_{g,g}^{g}$ is the deflated version of a matrix, in which columns $g(1), g(2), \ldots, g(i)$ have been zeroed. $R_{\text{order}}^{\text{MC}}$ is a matrix used for SD-SIC detection ordering, and is defined as

$$R_{\text{order}}^{\text{MC},k} = \sum_{k=1}^{K} R_{\text{MC},k} + R_{\text{noise}}^{\text{MC}} \quad (16)$$

Note that the detection ordering processes using this $R_{\text{order}}^{\text{MC},k}$ in (15f) and (15k) are performed to select the substream with the best SINR at each iteration, when interfering signal powers $P_{k,m}$’s for $k \geq k_0 + 1$ are given and $P_{k,m}$’s are assumed to be equal for all $m$. Our simulation results show that the performance degradation is negligible, even when the matrix $R_{\text{order}}^{\text{MC}}$ is set to “identity matrix” for computational simplicity.

IV. TRANSMIT POWER ALLOCATION

In this section, we describe several effective transmit power allocation schemes. The joint power allocation scheme, which makes the post-detection SINR become the same for all substreams in both space and code domains, is developed in Section IV-A. The two-stage power allocation scheme is derived as a simplified version of the joint power allocation scheme in Section IV-B. The variable and constant power ratio (PR) schemes are developed to reduce the feedback overhead in Section IV-C.

A. Joint Power Allocation

We first define the post-detection SINR of the $g(i)$th substream for the $k_0$th code, $\Gamma_{k_0,g(i)}$, as the SINR of the decision statistic $w_{k_0,g(i)} \tilde{z}_{k_0}$ in (15h) when the error propagation is ignored, i.e.,

$$\Gamma_{k_0,g(i)} = \frac{P_{k_0,g(i)}}{\Lambda_{k_0,g(i)}} \quad (17)$$

where $\Lambda_{k_0,g(i)} = \sum_{k=1}^{K} P_{k,M}^{\text{MC}} + R_{\text{noise}}^{\text{MC}}$ denotes the variance of MCI-plus-noise component. For large $K$, it may be difficult to find a closed-form solution for the post-detection SINRs to be equal for all substreams in both space and code domains. In this case, an adaptive method may be employed to find a solution in an iterative manner as follows.

Step 1) Initialization for Loop 1: Set an iteration number $l = 1$ and an arbitrary initial $K$th code power $P_l(1)$, $0 < P_l(1) < P_l$.

Step 2) Initialization for Loop 2:

$k_0 = K$

Step 3) Initialization for Loop 3:

$i = 1$

$$G_{k_0}(1) = \left( Y_{k_0,k_0} \right)^*$$

$$g(1) = \arg\min_{j\neq g} \left[ G_{k_0}(1) R_{\text{order}}^{\text{MC}}(j) \left( G_{k_0}(1) \right)^H \right]_{j,j} \quad (18a)$$

Step 4) Calculation of MCI-plus-noise variance $\Lambda_{k_0,g(i)}(l)$ for the $g(i)$th substream of the $k_0$th code:

$$w_{k_0,g(i)} = [G_{k_0}(i)]^H$$

$$\Lambda_{k_0,g(i)}(l) = \sum_{k=1}^{K} P_{k,M}^{\text{MC}} + R_{\text{noise}}^{\text{MC}}$$

$$G_{k_0}(i+1) = \left( \left( Y_{k_0,k_0} \right)^* \right)_{i,M} \quad (18d)$$

$$g(i+1) = \arg\min_{j\neq g} \left[ G_{k_0}(i+1) R_{\text{order}}^{\text{MC}}(j) \left( G_{k_0}(i+1) \right)^H \right]_{j,j} \quad (18e)$$

Step 5) Repetition or termination for Loop 3: If $i < M$, increase $i$ by one and go to Step 4. If $i = M$ and $k_0 = K$, go to Step 6. Otherwise, go to Step 7.

Step 6) Adjustment of the reference SINR $\Gamma_{\text{ref}}(l)$ for all data substreams:
\[ \Gamma_{\text{ref}}(l) = \frac{P_k(l)}{\sum_{m=1}^{M} \Lambda_{k,m}(l)} = \frac{\sum_{m=1}^{M} P_{k,m}(l)}{\sum_{m=1}^{M} \Lambda_{k,m}(l)} \]  

(18i)

Step 7) Calculation of the transmit power set for the \( k \)th code, \( \{P_{k,m}(l) : m=1, 2, \ldots, M\} \):

\[ P_k(l) = \Gamma_{\text{ref}}(l) \sum_{m=1}^{M} \Lambda_{k,m}(l) \]  

(18j)

\[ P_{k,m}(l) = \Lambda_{k,m}(l) \frac{P_k(l)}{\sum_{m=1}^{M} \Lambda_{k,m}(l)} = \Lambda_{k,m}(l) \Gamma_{\text{ref}}(l) \]  

(18k)

Step 8) Repetition or termination for Loop 2: If \( k_0 > 1 \), decrease \( k_0 \) by one and go to Step 3. Otherwise, go to Step 9.

Step 9) Repetition or termination for Loop 1:

If \( \sum_{k=1}^{K} P_k(l) - P_l > \varepsilon \), update \( P_k(l) \) as

\[ P_k(l+1) = \begin{cases} P_k(l) - \Delta, & \text{if } \sum_{k=1}^{K} P_k(l) > P_l \\ P_k(l) + \Delta, & \text{otherwise} \end{cases} \]  

(18l)

where \( \varepsilon (\varepsilon > 0) \) is a maximum allowable error of the sum of all updated power components, and \( \Delta (\Delta > 0) \) is a step size for \( P_k(l) \). Then, increase \( l \) by one, and go to Step 2. Otherwise, terminate the whole joint power allocation algorithm.

Note that the iterative procedure described above obtains the power components for each code in the reverse order of code index, since the values of \( \Lambda_{k,g}(l) \) for the \( k \)th code require the knowledge of interfering signal powers \( P_{k,m}(l) \)’s for \( k \geq k_0 + 1 \). It should also be noted that all final values for the nulling vector and detection ordering, calculated in the power allocation procedure, can be used in the detection procedure described in Section III without repeating calculations.

B. Two-Stage Power Allocation

In this subsection, we derive two-stage power allocation scheme as a simplified version of the joint power allocation scheme. Proposed two-stage scheme performs the code domain power allocation, followed by the space domain power allocation. In the first stage, an approximate closed-form solution for the code power set \( \{P_k(l) : k=1, 2, \ldots, K\} \) is derived. In the second stage, based on the values calculated in the first stage, the antenna power components for each code, \( \{P_{k,m}(l) : m=1, 2, \ldots, M\} \), are found in an iterative manner.

1) Code Domain Power Allocation (First Stage): To find the code domain power allocation, we first define the code domain SINR for the \( k \)th code, \( \Gamma_{\text{CD}}^{k} \), as

\[ \Gamma_{\text{CD}}^{k} \triangleq \frac{E \left[ \left\| \mathbf{Y}_{k,l} - \mathbf{P}_k \mathbf{d}_k \right\|^2 \right]}{E \left[ \sum_{k=1}^{K} \left\| \mathbf{Y}_{k,l} - \mathbf{P}_k \mathbf{d}_k + n_k \right\|^2 \right]}, \quad k_0 = 1, \ldots, K \]  

(19)

where the expectations are taken with respect to the spreading sequences as well as the data and noise vectors. Note that the elements of the vectors in (19) are the signals posterior to the CD-SIC detection process and prior to the SD-SIC detection process. Based on the equal code domain SINR design, we allocate the code power so that \( \Gamma_{\text{CD}}^{k_0} \) becomes equal for all codes as

\[ \Gamma_{\text{CD}}^{k} = \Gamma_{\text{CD}}^{k_0}, \quad k_0 = 1, 2, \ldots, K. \]  

(20)

To find the code power set satisfying (20), we assume that \( P_{k,m} = P_k/M \) for all \( k \) and \( m \). Using (10), it can be shown that the equation (20) is reduced to

\[ \frac{\alpha P_k}{\beta \sum_{k=k_0+1}^K P_k + \eta} = \Gamma_{\text{CD}}, \quad k_0 = 1, 2, \ldots, K \]  

(21)

where

\[ \alpha = \frac{T_c^2}{M} \sum_{p=1}^N \sum_{l=1}^L \left[ \left| h_{p,m(l)} \right|^2 \right]^2 \frac{1}{G} \sum_{i=0}^{L-1} \left[ \left| G - |f - (i-1)| \right| \right], \]  

(22)

\[ \beta = \frac{1}{M} \sum_{p=1}^N \sum_{l=1}^L \sum_{i=0}^{L-1} \sum_{j=0}^{L-1} \left| h_{p,m(l)} \right|^2 \left[ (G - |f - (i-1)|) \right] \]  

(23)

\[ \eta = N L \sigma^2 T. \]  

(24)

Solving \( K+1 \) simultaneous equations in (3) and (21), we can find the code power values as

\[ P_k = \left[ \frac{1 - \gamma}{1 - \gamma^k} \right] P_l, \quad k = 1, 2, \ldots, K \]  

(25)

where

\[ \gamma = \frac{1}{\left( 1 + P_l \frac{\beta}{\eta} \right)^{1/k}}. \]  

(26)

Using (23) and (24), \( \beta/\eta \) in (26) is calculated as

\[ \frac{\beta}{\eta} = \frac{T_c^2}{MN L \sigma^2 T} \sum_{p=1}^N \sum_{l=1}^L \sum_{i=0}^{L-1} \left| h_{p,m(l)} \right|^2 \left[ (G - |f - (i-1)|) \right]. \]  

(27)

If we assume that \( G \gg L \), (27) may be approximated as

\[ \frac{\beta}{\eta} = \frac{T_c^2 G L (1 - \frac{1}{L})}{MN L \sigma^2 T} \sum_{p=1}^N \sum_{l=1}^L \sum_{i=0}^{L-1} \left| h_{p,m(l)} \right|^2 \approx \rho. \]  

(28)

Note that \( \gamma \) in (26) is the power ratio between the two adjacent code powers, i.e., \( \gamma = P_{k+1}/P_k \) for \( k = 1, 2, \ldots, K-1 \).

2) Space Domain Power Allocation (Second Stage): Based on the code power values calculated in the first stage, the space domain power components for each code, \( \{P_{k,m}(l) : m=1, 2, \ldots, M\} \), are computed in an iterative manner. Essentially, this iterative procedure follows the joint power allocation algorithm described in Section IV-A, except for the reference SINR adjustment in (18i) and some procedures in Loop 1 for adaptive code power calculation.

Note that the two-stage power allocation in this subsection makes the post-detection SINRs become equal for all the substreams in the same code domain, however, the post-detection SINRs between the data substreams in the different code domains may be different.

C. Variable and Constant Power Ratio (PR) Schemes

The joint and two-stage schemes described in Section IV-A and IV-B respectively, require the feedback information for all the transmit power components \( P_{k,m} \)'s from the receiver to the transmitter. As the number of transmit antennas and spreading codes increases, the amount of feedback information increases. As shown in (25), for given some channel states, the code domain power allocation needs only “single” power ratio value \( \gamma \) for the feedback information, since the given total transmit power \( P_T \) is known at the transmitter. Thus, it may be desirable to allocate the different transmit powers in the code domain only as in (25), with the space domain powers for a specific code distributed equally, in order to reduce the amount of feedback information. In this case, the power allocation may be written as

\[ P_{k,m} = \frac{1}{M} \left[ 1 - \frac{\gamma}{1 - \gamma^k} \right] P_{l}, \quad k = 1, \ldots, K, \quad m = 1, \ldots, M \]  

(29)

where \( \gamma = (1 + P_l \rho)^{-1/k} \). We refer to this scheme as the variable PR scheme, in which the power ratio \( \gamma \) varies with the instantaneous channel gains as shown in (28).
The ratio factor $\rho$ in (28) is an important parameter determining the power ratio $\gamma$ for the variable PR scheme. The expectation of $\rho$ in (28) with respect to the fading channel amplitudes $\{h_{p,m}\}$, may be calculated as
\[
E[\rho] = \frac{T(L-1)}{GL\sigma^2} \Omega_{\text{sum}} = \rho_C \tag{30}
\]
where $\Omega_{\text{sum}} = \sum_{i=0}^{L-1} \Omega_i \Omega_{\text{sum}}^{-1} \sum_{i=0}^{L-1} E[|h_{p,m,i}|^2]$. From the law of large numbers [7], it can be seen that as $MN$ goes to infinity, $\rho$ in (28) converges to a finite value $\rho_C$ in (30). From this relationship between $\rho$ and $\rho_C$, we propose the power allocation written as
\[
P_{k,m} = \frac{1}{M} \left(\frac{1 - \gamma_{C}}{1 - \gamma_{K}}\right)^{k-1}, \quad k=1, \ldots, K, \quad m=1, \ldots, M \tag{31}
\]
where $\gamma_{C} = \left(1 + P_{T} \rho_C \right)^{-1/K}$, for the case of large number of antennas. This power allocation in (31) is the same as that in (29), except that a parameter $\rho$ in (28) is replaced by $\rho_C$ in (30). We refer to this scheme as the constant PR scheme. Note that the constant PR scheme uses the power ratio determined by the long-term statistical properties of the fading channel amplitudes, i.e., this scheme is independent on the instantaneous channel gains. As a result, in addition to the advantage of the “small amount of feedback information” for given channel states, the constant PR scheme provides the significantly “reduced feedback rate” over the time-varying fading channels.

V. NUMERICAL RESULTS

In this section, the performance of the transmit power allocation schemes described in Section IV are evaluated and compared with one another. The feedback channel is assumed to be an ideal error-free channel without feedback delay. The average BER is calculated over sufficient number of randomly generated channel coefficients $\{h_{p,m}\}$, spreading sequences $\{c_{i}\}$, and data symbols $\{d_{k,m}\}$. The spreading gain $G = 32$, the number of multipaths $L = 3$, the exponential decay rate of the MIP $\delta = 0.5$, and $\Omega_{\text{sum}}$ is set to 1. The average SNR is defined to be $P_{T}T/K\sigma^2$. For spreading codes, we use the orthogonal Walsh-Hadamard codes multiplied by a common random complex scrambling sequences [8].

Fig. 3 compares the performance of the transmit power allocation schemes, when $M = N = 4$ and $K = 8$. The scheme denoted by “SD-SIC w/o CD-SIC” in Figs. of this section allocates equal power (denoted by “EP”) to all substreams and uses the SIC detection process in space domain, not in code domain. “Two-stage SIC” in Figs. is also assumed to allocate the equal power to all data substreams. It can be seen that the performance of these two equal power schemes suffer from irreducible error floors at high SNR range. The joint and two-stage power allocation schemes improve the BER performance significantly over these equal power schemes. It is noticeable that the performance of the joint and two-stage power schemes are almost indistinguishable at all SNR range. Therefore, the joint power scheme can be replaced by the two-stage power scheme without any performance degradation but with significant savings in computational complexity. Fig. 3 also shows that the variable and constant PR schemes are considerably superior to the equal power schemes. It is also noticeable that the performance of the variable and constant PR schemes are almost indistinguishable. This indicates that, in this MIMO case of Fig. 3, the constant PR scheme can be an alternative to the variable PR scheme, and hence the feedback rate for power allocation may be reduced significantly without any performance degradation.

In Fig. 4, the BER performance of the power allocation scheme according to the equation (31) is illustrated with the same parameters as in Fig. 3, when the fixed code power ratio $\gamma_C$ in (31) is set to arbitrary values as well as the value by the constant PR scheme. The code power ratio by the constant PR scheme is shown to provide almost the best performance for all given fixed power ratios. Note that $\gamma_C$ of the constant PR scheme decreases as the average SNR increases. The reason for this is that, as the SNR increases, the MCI effect on the BER performance is larger than the Gaussian noise effect, and hence the more power is allocated to the data substreams corresponding to the earlier detected code index.

Fig. 5 shows the performance comparison of the transmit power allocation schemes, when $M = N = 1$ and $K = 8$. As opposed to the indistinguishable performance difference between the variable and constant PR schemes in the MIMO case of Fig. 3, the constant PR scheme is seen to be significantly inferior to the variable PR scheme in this SISO case. This indicates that the SISO case of $M = N = 1$ does not provide high probability of $\gamma$ approaching $\gamma_C$, and the fixed power ratio $\gamma_C$ regardless of the instantaneous channel states may result in the significant performance degradation.

The effects of the number of antennas in the case of $M = N$ on the BER performance are shown in Fig. 6, when $K = 8$ and the average SNR = 20 dB. It can be seen that the performance of constant PR scheme is almost the same as that of variable PR scheme for $M(N) \geq 2$. As the number of antennas increases, the performance of variable and constant PR schemes is found to approach that of the two-stage power scheme. This indicates that the effects of space domain power allocation on the BER performance may be reduced when the number of antennas becomes larger.

VI. CONCLUSIONS

In this paper, we have presented a simple two-stage SIC detection scheme for a multicode MIMO system, and developed the transmit power allocation schemes to improve error rate performance for this detection process. The joint power allocation makes the post-detection SNR become the same for all substreams in both space and code domains. A computationally efficient two-stage power allocation scheme has also been derived, and found to provide almost the same BER performance as the joint power allocation scheme. The variable and constant PR schemes have been developed to reduce the feedback overhead. In particular, the constant PR scheme has been shown to achieve significantly reduced feedback rate to the transmitter. It has also been found that the performance of the variable and constant PR schemes are almost indistinguishable in MIMO systems, and those approach the performance of the two-stage power allocation scheme, as the number of transmit and receive antennas increases.

REFERENCES


Figures

Fig. 1. Multicode MIMO communication system.

Fig. 2. Two-stage SIC receiver structure.

Fig. 3. BER performance comparison of power allocation schemes for $M = N = 4$.

Fig. 4. BER performance versus the fixed code power ratio for various SNR values, when $M = N = 4$.

Fig. 5. BER performance comparison of power allocation schemes for $M = N = 1$ (SISO case).

Fig. 6. Effects of the number of antennas on the BER performance of power allocation schemes, when $M = N \geq 2$. 