Signal Processing and Interference Suppression for Wireless Networks with Any Number of Users

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Abstract— In this paper, we investigate how to send space time codes with full diversity and low decoding complexity for Z channels with any number of users using precoders. First, we assume that we have J transmitters and J receivers. Each transmitter sends code words to respective receiver at the same time. We propose an orthogonal transmission scheme that combines space-time codes and array processing to achieve low-complexity decoding and full diversity for transmitted signals. To our best knowledge, this is the first general scheme which can achieve low-complexity decoding and full diversity for any transmitted code word in Z channel with any number of uses when all the users transmit at the same time. Simulation results validate our theoretical analysis.

Key Words— Space-time codes, array processing, full diversity, precoder design, interference cancellation, Z channels.

I. Introduction

In recent years, wireless multiple-input multiple-output (MIMO) systems with multiple antennas employed at both the transmitter and receiver have gained attention because of their promising improvement in terms of performance and bandwidth efficiency. In the uplink, several single-user techniques have been proposed such as vertical Bell Laboratories layered spaceCtime (V-BLAST), maximum likelihood detection(MLD), and singular value decomposition (SVD)-based techniques. Here we introduce a multiuser MIMO transmit preprocessing technique for the uplink of multiuser MIMO systems. The technique is based on decomposing a multiuser MIMO uplink channel into parallel independent single user MIMO uplink channels. Once the multiuser channels are decomposed, any single-user MIMO technique (such as MLDand BLAST) can be applied in the usual way to each user. Previously, there has been only limited work on multiuser MIMO systems for the uplink. Another issue is that particular linear receiver structures are assumed in both of the systems and these impose certain restrictions on the systems [1]–[6].

Recently, several space-time processing techniques have been used in multiple access channels to reduce the decoding complexity and enhance system performance by canceling the interference from different users [7]–[31].

In this paper, we investigate how to achieve the low complexity decoding and the highest possible diversity to improve the transmission quality for space-time codes in Z channels [11] without losing symbol rate. Our idea to solve this problem is to design proper precoding and decoding schemes based on space-time coding with the assumption of full channel information at the transmitter. The idea of combining space-time coding and precoding in multiuser systems is not new [12], [13]. Note that one can use interference alignment methods to achieve the highest degree of freedom [16]. But under our assumptions, using interference alignment, the diversity will be one. The concentration of this paper is to achieve the highest diversity with low decoding complexity for space-time codes, not achieving the highest degree of freedom.

The outline of the paper follows next. Section II introduces our motivation and the Z channels we discuss in this paper. In Section III, we propose an orthogonal transmission scheme which is necessary to achieve low complexity decoding, high coding gain and full diversity as shown in later sections. In Section IV, our decoding scheme is proposed. We analyze the performance of our scheme in Section V. Simulation results are presented in Section VI and Section VII concludes the paper.

Notation: We use boldface letters to denote matrices and vectors, super-scripts $(\cdot)^T$, $(\cdot)^*$, $(\cdot)^\dagger$ to denote transpose, conjugate and transpose conjugate, respectively. We denote the element in the ith row and the jth column of matrix **X** by $X(i, j)$. We denote the jth column of a matrix **X** by $\mathbf{X}(j)$.

II. Motivation and Channel Model

In a point-to-point MIMO system, i.e., one transmitter with N transmit antennas and one receiver with M receive antennas, one can use space-time codes to achieve symbol-by-symbol decoding and full diversity when the transmitter does not know the channel. The symbol rate is one. Let us consider a channel model as shown in Figure 1. We assume there are J users each with J transmit antennas and J receivers each with J receive antennas. Both users want to send different space-time codes to different receivers on the same frequency band at the same time. As shown in Figure 1, User 1 wants to send codeword C to Receiver 1 without causing interference to other receivers. When channel knowledge is not available at the transmitters, space-time codes combined with TDMA can be used to achieve symbol-by-symbol decoding and full diversity. But the symbol rate reduces to $1/J$. A solution to keep the symbol rate unchanged when space-time codes are used, is to combine space-time coding and array processing. In other words, we allow all transmitters to send space-time codes simultaneously to keep rate one and utilize special array processing techniques to achieve low-complexity decoding and full diversity. In this paper, we achieve the above goals under short-term power constraints, fixed codeword block length and limited delay, when there is outage. We do not claim that our scheme can achieve capacity or full degree of freedom. After

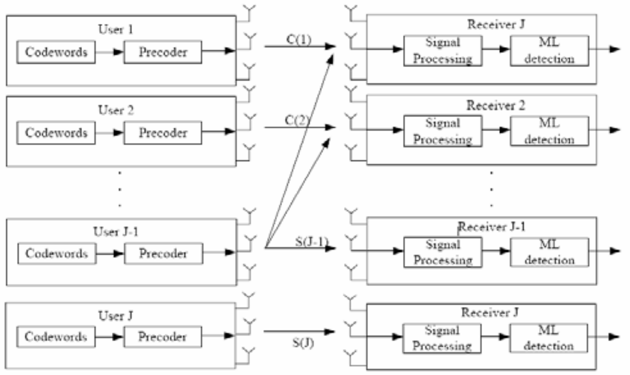

Fig. 1. Z Channel

all, there is a tradeoff between diversity and multiplexing gain, which is outside the scope of this paper.

We assume among J transmitters, J_1 transmitters will cause interference to others while J_2 transmitters will not cause interference. So we have

$$J_1 + J_2 = J \tag{1}$$

Then we introduce the input-output equations. We let user i transmit rate one $J_1 \times J_1$ Codes $\mathbf{C}_i$ [17]. Let $\mathbf{A}_j^t$ be the precoders we need to design for user j. They are combined with the space-time codes sent by User j and this is the first step of our array processing technique. Note that in order to satisfy the short-term power constraint, we need

$$||\mathbf{A}_j^t(J,J)||_F^2 = 1 \tag{2}$$

Where (J,J) the dimension of the matrix is $J \times J$. Similarly, the precoders for User 2 is defined as $\mathbf{B}_j^t$ with the power constraint

$$||\mathbf{B}_j^t(J,J)||_F^2 = 1 \tag{3}$$

The channels are quasi-static flat Rayleigh fading and keep unchanged during two time slots. Then we let $\mathbf{H}_{ij}(J,J)$ denote the channel matrix between User i and Receivers j, respectively. Then the received signals at Receiver j_1 with interference at time slot t can be denoted by

$$\mathbf{y}_{j_1}^t(J,1) = \sum_{i=1}^{J} \mathbf{H}_{ij_1}(J,J)\mathbf{A}_i^t(J,J)\mathbf{C}_i(t) + \mathbf{n}_{j_1}^t(J,1) \tag{4}$$

where $\mathbf{y}_{j_1}^t(J,1)$ and $\mathbf{n}_{j_1}^t(J,1)$ denote the received signals and the noise at Receiver j, respectively, at time slot t. The received signals at Receiver j_2 without interference at time slot t can be denoted by

$$\mathbf{y}_{j_2}^t(J,1) = \mathbf{H}_{j_2 j_2}(J,J)\mathbf{A}_{j_2}^t(J,J)\mathbf{C}_{j_2}(t) + \mathbf{n}_{j_2}^t(J,1) \tag{5}$$

where $\mathbf{y}_{j_2}^t(J,1)$ and $\mathbf{n}_{j_2}^t(J,1)$ denote the received signals and the noise at Receiver j_2, respectively, at time slot t. Equations (4) and (5) are the channel equations on which we will base our design in this paper.

III. Precoder Design and Orthogonal Transmission Structure

In this section, we will build an orthogonal transmission structure by combining the space-time codes and our precoders. This orthogonal transmission structure is necessary because it provides two benefits. The first benefit is that low-complexity decoding can be realized because under this orthogonal transmission structure, different codewords will be sent along different orthogonal vectors. We can easily decode the symbols without the interference at each receiver. The second benefit is that we can achieve full diversity and higher coding gain once we make the proper array processing as shown in later sections. This is the key difference between our array processing method and the interference alignment method. The latter can only achieve the first benefit. Of course, the tradeoff is that we lose the maximum possible degree of freedom in the process.

Different users and different codewords may have different diversities. By saying full diversity, we mean the diversity is full for each codeword sent by each user. For example, full diversity for User 1 means at Receiver 1, the diversity for codeword $\mathbf{C}(J,J)$ is full. In this section, we show how to build the orthogonal transmission structure by designing proper precoders. Later, we will show that our proposed orthogonal transmission scheme can achieve low-complexity decoding and full diversity.

Our main idea to build the orthogonal transmission structure is to adjust each signal in the signal space of Z channels by using precoders for each transmitter, such that at the receiver each desired signal is orthogonal to all other signals. In Equation (4), we use

$$\mathbf{H}_{ij_1}^t(J,J) = \mathbf{H}_{ij_1}(J,J)\mathbf{A}_i^t(J,J) \tag{6}$$

to denote the equivalent channel matrices. Then Equation (5) becomes

$$\mathbf{y}_{j_1}^t(J,1) = \sum_{i=1}^{J} \mathbf{H}_{ij_1}^t(J,J)\mathbf{C}_i(t) + \mathbf{n}_{j_1}^t(J,1) \tag{7}$$

Similarly, in Equation (5), if we use

$$\mathbf{H}_{j_2 j_2}^t(J,J) = \mathbf{H}_{j_2 j_2}(J,J)\mathbf{A}_{j_2}^t(J,J) \tag{8}$$

to denote the equivalent channel matrices, we have

$$\mathbf{y}_{j_2}^t(J,1) = \mathbf{H}_{j_2 j_2}^t(J,J)\mathbf{C}_{j_2}(t) + \mathbf{n}_{j_2}^t(J,1) \tag{9}$$

By Equation (7), since the receiver has J_1 receive antennas, each symbol is actually transmitted along a J_1-dimensional vector in a J_1-dimensional space. Because each user sends two symbols at the same time, at the receiver, there are J_1^2 signal vectors in the two-dimensional space.

Since we want to send $\mathbf{C}_i(J,J)$ along orthogonal directions, we let each one of $\mathbf{C}_i(J,J)$ occupy only one dimension. In other words, for any codeword, we should transmit each of the corresponding four symbols in the same direction. In this way, there are only 2 transmit directions. Once we can align the 2 transmit directions of $\mathbf{C}_i(J,J)$ properly, we can separate them completely. This is the main idea to build the orthogonal transmission structure. Note that this is only a general idea and much details are omitted. For example, we will show later that after some array processing and moving the interference at the receiver, each symbol at each receiver will have its own direction. We need to do additional array processing to reduce the decoding complexity and achieve full diversity.

In this section, we only explain the above main idea. By Equation (7), $\mathbf{C}_i(J,J)$ are transmitted along $\mathbf{H}_{ij}^t(J,J)(l)$, respectively, where (J,J) means the dimension of the matrix is $J \times J$ and l means the lth column of the matrix. In order to make $\mathbf{H}_{ij}^t(l)$ along the same direction, by Equation (6), we need

$$\mathbf{A}_{i_1}^t(J,J)(1) = \frac{1}{\alpha_{i_1 l}^t}(J,J)\mathbf{A}_{i_1}^t(l)(J,J) \quad (10)$$

where $\alpha_{i_1 l}^t$ is a constant for the precoder $\mathbf{A}_{i_1}^t(J,J)$ that we will determine later. Here i_1 means the i_1th transmitter that will cause interference. From $||\mathbf{A}_{i_1}^t(J,J)||_F^2 = 1$, we know

$$||\mathbf{A}_{i_1}^t(J,J)(1)||_F^2 = \frac{1}{1+\sum_{l=1}^{J_1}(\alpha_{i_1 l}^t)^2} \quad (11)$$

So when we design precoder $\mathbf{A}_{i_1}^t(J,J)$, Equations (10) and (11) should be satisfied. Similarly, for receivers without interference, we need

$$\mathbf{A}_{i_2}^t(J,J)(1) = \frac{1}{\alpha_{i_2 l}^t}\mathbf{A}_{i_2}^t(J,J)(l) \quad (12)$$

where $\alpha_{i_2 l}^t$ is a constant for the precoder $\mathbf{A}_{i_2}^t(J,J)$ that we will determine later. Here i_2 means the i_2th transmitter that will cause interference. From $||\mathbf{A}_{i_2}^t(J,J)||_F^2 = 1$, we know

$$||\mathbf{A}_{i_2}^t(J,J)(1)||_F^2 = \frac{1}{1+\sum_{l=1}^{J_1}(\alpha_{i_2 l}^t)^2} \quad (13)$$

Now Equations (7) and (9) become

$$\mathbf{y}_{j_1}^t(J,1) = \sum_{i_1=1}^{J_1}[\mathbf{H}_{i_1 j_1}^t(J,J)(1), \mathbf{H}_{i_1 j_1}^t(J,J)(1)]\cdot \mathbf{C}_{i_1}(t) + \mathbf{n}_{j_1}^t(J,1) \quad (14)$$

and

$$\mathbf{y}_{j_2}^t(J,1) = [\mathbf{H}_{i_2 j_2}^t(J,J)(1), \mathbf{H}_{i_2 j_2}^t(J,J)(1)]\cdot \mathbf{C}_{i_2}(t) + \mathbf{n}_{j_2}^t(J,1) \quad (15)$$

where $\mathbf{H}_{i_2 j_2}^t(J,J)(1)$ denote the first column of matrix $\mathbf{H}_{i_2 j_2}^t(J,J)$, respectively. At receiver j_1, after we combine the channel equations in J_1 time slots, we have

$$\mathbf{y}_{j_1}(J,1) = \sum_{i_1=1}^{J_1}\begin{pmatrix} \alpha_{i_1 1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(1) & \cdots & \alpha_{i_1 J_1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(J_1) \\ \vdots & \ddots & \vdots \\ \alpha_{i_1 J_1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(1) & \cdots & \alpha_{i_1 1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(J_1) \end{pmatrix} \times \mathbf{C}_{i_1}(1) + \mathbf{n}_{j_1}(J,1) \quad (16)$$

where

$$\mathbf{y}_{j_1}(J,1) = \begin{pmatrix} \mathbf{y}_{j_1}^1(J,1) \\ \vdots \\ (\mathbf{y}_{j_1}^{J_1}(J,1))^* \end{pmatrix} \quad (17)$$

and

$$\mathbf{n}_{j_1}(J,1) = \begin{pmatrix} \mathbf{n}_{j_1}^1(J,1) \\ \vdots \\ (\mathbf{n}_{j_1}^{J_1}(J,1))^* \end{pmatrix} \quad (18)$$

Similarly, at receiver j_2 without interference, after we combine the channel equations in J_1 time slots, we have

$$\mathbf{y}_{j_2}(J,1) = \begin{pmatrix} \alpha_{i_2 1}^t \mathbf{H}_{i_2 j_2}^1(J,J)(1) & \cdots & \alpha_{i_2 J_2}^t \mathbf{H}_{i_2 j_2}^1(J,J)(J_2) \\ \vdots & \ddots & \vdots \\ \alpha_{i_2 J_1}^t \mathbf{H}_{i_2 j_2}^1(J,J)(1) & \cdots & \alpha_{i_2 1}^t \mathbf{H}_{i_2 j_2}^1(J,J)(J_2) \end{pmatrix} \times \mathbf{C}_{i_2}(J,J)(1) + \mathbf{n}_{j_2}(J,1) \quad (19)$$

where

$$\mathbf{y}_{j_2}(J,1) = \begin{pmatrix} \mathbf{y}_{j_2}^1(J,1) \\ \vdots \\ (\mathbf{y}_{j_2}^{J_2}(J,1))^* \end{pmatrix} \quad (20)$$

and

$$\mathbf{n}_{j_2}(J,1) = \begin{pmatrix} \mathbf{n}_{j_2}^1(J,1) \\ \vdots \\ (\mathbf{n}_{j_2}^{J_2}(J,1))^* \end{pmatrix} \quad (21)$$

By Equation (16), we can see that once we make vector $\mathbf{H}_{i_1 j_1}^t(J,J)(1)$ orthogonal to each other with different i_1 at time slot t, signal vectors for different elements of $\mathbf{C}_{i_1}(J,J)(1)$ will lie in a subspace which is orthogonal to the subspace created by other signal vectors. Because of this orthogonality, at the receiver one, we can easily separate the desired signals from the interference signals. At Receiver j_2, since there is no interference, by Equation (19), we can easily decode the desired signals $\mathbf{C}_{i_2}(J,J)(1)$. This is our main idea to achieve interference-free transmission in this interference channel.

Now we show how to derive the above orthogonality by designing precoders for Users j_1 and Users j_2 simultaneously. Assume the Singular Value Decomposition of channel matrices $\mathbf{H}_{i_1 j_1}(J,J)$, $\mathbf{H}_{i_2 j_2}(J,J)$ as follows

$$\mathbf{H}_{i_1 j_1}(J,J) = \mathbf{V}_{H_{i_1 j_1}(J,J)}\mathbf{\Lambda}_{H_{i_1 j_1}(J,J)}\mathbf{U}_{H_{i_1 j_1}(J,J)}^{\dagger} \quad (22)$$

$$\mathbf{H}_{i_2 j_2}(J,J) = \mathbf{V}_{H_{i_2 j_2}(J,J)}\mathbf{\Lambda}_{H_{i_2 j_2}(J,J)}\mathbf{U}_{H_{i_2 j_2}(J,J)}^{\dagger} \quad (23)$$

At time slot j_1, we let User j_1 transmit along its best direction. In this case, we can choose the precoder

$$\mathbf{A}^{i_1 j_1}(J,J) = \frac{1}{\sqrt{\sum_{l=1}^{J_1}(\alpha_{i_1 l}^t)^2}}[\alpha_{i_1 1}^t \mathbf{U}_{H_{i_1 j_1}(J,J)}(1), \ldots, \alpha_{i_1 J_1}^t \mathbf{U}_{H_{i_1 j_1}(J,J)}(1)] \quad (24)$$

Then we design precoders for User $j_1^{'} \in \{1,\ldots,J_1\}$, but $j_1^{'} \neq j_1$ such that at receiver j_1, the signal vectors from other users are orthogonal to the signal vectors of User j_1. Note that at receiver j_1, the signal from other users is interference, while at other receivers, signals from user j_1 is interference. We need to consider the signals at all receivers when we design precoders. So, at time slot j_1, the precoder $\mathbf{A}_{i_1^{'} j_1^{'}}^1$ needs to satisfy the following J_1 equations. At receiver j_1,we need

$$(\mathbf{H}_{i_1^{'} j_1^{'}}^1(J,J)(1))^{\dagger}\cdot \mathbf{H}_{i_1 j_1}^1(J,J)(1) = 0 \quad (25)$$

where $j_1^{'} = 1,\ldots,J_1$ but $j_1^{'} \neq j_1$ and the power constraint

$$||\mathbf{A}^{i_1^{'} j_1^{'}}(J,J)(1)||^2 = \frac{1}{\sum_{l=1}^{J_1}(\alpha_{i_1 l}^t)^2} \quad (26)$$

By solving the above J_1 equations, we can get the precoder $\mathbf{A}^{i_1' j_1'}(J, J)$ for User j_1' at time slot j_1. At time slot j_2, we let user j_2 transmit along its best direction. In this case, we can choose the precoder

$$\mathbf{A}^{j_2}(J, J) = \frac{1}{\sqrt{\sum_{l=1}^{J_2} (\alpha_{i_1 l}^t)^2}} [\mathbf{U}_{H_{j_2}}(J, J)(1), \mathbf{U}_{H_{j_2}}(J, J)(1)] \tag{27}$$

and the power constraint

$$||\mathbf{A}^{j_2}(J, J)(1)||^2 = \frac{1}{\sum_{l=1}^{J_2} (\alpha_{i_1 l}^t)^2} \tag{28}$$

By solving the above equations, we can get the precoder $\mathbf{A}^{j_1}(J, J)$ for User j_1 that will cause interference at time slot j_1. With our precoders $\mathbf{A}^{j_1}$ at time slot j_1 and $\mathbf{A}^{j_2}(J, J)$ at time slot j_2, we can show that we can achieve interference-free transmission with low decoding complexity and full diversity simultaneously as shown in the next two sections.

IV. Decoding with Low Complexity

In the last section, we have shown how to build the orthogonal transmission structure. Once the orthogonal structure is built, it is easy to realize low-complexity decoding. In this section, we will show how to decode and analyze the decoding complexity. We first consider the decoding at receiver j_1. In Equation (16), if we let

$$\overline{\mathbf{H}}_{i_1 j_1}(J, J) = \begin{pmatrix} \alpha_{i_1 1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(1) & \cdots & \alpha_{i_1 J_1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(J_1) \\ \vdots & \ddots & \vdots \\ \alpha_{i_1 J_1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(1) & \cdots & \alpha_{i_1 1}^t \mathbf{H}_{i_1 j_1}^1(J,J)(J_1) \end{pmatrix} \tag{29}$$

then Equation (16) becomes

$$\mathbf{y}_{j_1}(J, 1) = \sum_{i_1=1}^{J_1} \overline{\mathbf{H}}_{i_1 j_1}(J, J) \cdot \mathbf{C}_{i_1}(1) + \mathbf{n}_{j_1}(J, 1) \tag{30}$$

Note that at receiver j_1, $\mathbf{C}_{i_1}(J, J)(1)$ are the desired signal and all others are the interference. We can cancel the interference by multiplying both sides of Equation (30) by matrix $\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger$. Then we get

$$\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \cdot \mathbf{y}_{j_1}(J, 1) = \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J) \mathbf{C}_{i_1}(1) + \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \mathbf{n}_{j_1}(J, 1) \tag{31}$$

Here we have canceled the interference because $\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1' j_1'}(J, J) = 0$. In order to decode the symbols, we first multiply both sides of Equations (31) by matrix $(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}}$ to whiten the noise, i.e.,

$$(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \cdot \mathbf{y}_1(J, 1) = (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{\frac{1}{2}} \mathbf{C}_{i_1}(J, J)(1) + (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \mathbf{n}_{j_1}(J, 1) \tag{32}$$

Then we can detect $\mathbf{C}_{i_1}(1)$ by

$$\widehat{\mathbf{C}_{i_1}}(1) = \arg \min_{\mathbf{C}_{i_1}(1)} \left\| (\overline{\mathbf{H}}_{i_1 j_1}^\dagger \overline{\mathbf{H}}_{i_1 j_1})^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}^\dagger \cdot \mathbf{y}_{j_1}(J, 1) - (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1})(J, J)^{\frac{1}{2}} \mathbf{C}_{i_1}(1) \right\|_F^2 \tag{33}$$

Further, note that $\overline{\mathbf{H}}_{i_1 j_1}^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J)$ is real. So, when QAM is used, Equation (32) is equivalent to the following two equations.

$$\begin{aligned} &\text{Real}\{(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \times \mathbf{y}_{j_1}(J, 1)\} \\ &= (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{\frac{1}{2}} \text{Real}\{\mathbf{C}_{i_1}(1)\} + \\ &(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \mathbf{n}_{j_1}(J, 1) \end{aligned} \tag{34}$$

$$\begin{aligned} &\text{Imag}\{(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \times \mathbf{y}_{j_1}(J, 1)\} \\ &= (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{\frac{1}{2}} \text{Imag}\{\mathbf{C}_{i_1}(1)\} + \\ &(\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \mathbf{n}_{j_1}(J, 1) \end{aligned} \tag{35}$$

So we can detect the real part and the imaginary part of c_1, c_2 separately as follows:

$$\begin{aligned} &\text{Real}\{\widehat{\mathbf{C}_{i_1}}(1)\} = \arg \min_{\text{Real}\{\mathbf{C}_{i_1}(1)\}} \\ &\Big\| \text{Real}\{(\overline{\mathbf{H}}_{i_1 j_1}^\dagger \overline{\mathbf{H}}_{i_1 j_1} \leq)^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \cdot \mathbf{y}_{j_1}(J, 1)\} \\ &- (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{\frac{1}{2}} \text{Real}\{\mathbf{C}_{i_1}(1)\} \Big\|_F^2 \end{aligned} \tag{36}$$

$$\begin{aligned} &\text{Imag}\{\widehat{\mathbf{C}_{i_1}}(1)\} = \arg \min_{\text{Imag}\{\mathbf{C}_{i_1}(1)\}} \\ &\Big\| \text{Imag}\{(\overline{\mathbf{H}}_{i_1 j_1}^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{-\frac{1}{2}} \overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \cdot \mathbf{y}_{j_1}(J, 1)\} \\ &- (\overline{\mathbf{H}}_{i_1 j_1}(J, J)^\dagger \overline{\mathbf{H}}_{i_1 j_1}(J, J))^{\frac{1}{2}} \text{Imag}\{\mathbf{C}_{i_1}(1)\} \Big\|_F^2 \end{aligned} \tag{37}$$

The decoding complexity is symbol-by-symbol. Similarly, we can detect s_1, s_2 with symbol-by-symbol complexity at receiver two.

V. Diversity Analysis

In this section, we show that our proposed scheme can achieve full diversity for each user. We only prove that at receiver j_1, the diversity for $\mathbf{C}_{i_1}(1)$ from user j_1 is full. The proof for other signals will be similar. First, the diversity is defined as

$$d = -\lim_{\rho \to \infty} \frac{\log P_e}{\log \rho} \tag{38}$$

where ρ denotes the SNR and P_e represents the probability of error. We let $\mathbf{e} = \begin{pmatrix} e_1 \\ \vdots \\ e_{J_1} \end{pmatrix} = \mathbf{C}_{i_1}(1) - \widehat{\mathbf{C}}_{i_1}(1)$ denote the error vector. Here we add a rotation matrix $\mathbf{R}(J, J)$ on the transmitted codewords to improve the system performance. Based on Equation (32), the pairwise error probability (PEP) for c_1, c_2 can be written as [18]

$$\begin{aligned}
&P(\mathbf{c} \to \overline{\mathbf{c}} | \overline{\mathbf{H}}_{j_1 j_1}(J,J)) = \\
&Q\left(\sqrt{\frac{\rho ||(\overline{\mathbf{H}}_{j_1 j_1}(J,J)^\dagger \overline{\mathbf{H}}_{j_1 j_1}(J,J))^{\frac{1}{2}} \mathbf{R}(J,J)\mathbf{e}||_F^2}{4}}\right) \\
&= Q\left(\sqrt{\frac{\rho \mathbf{e}^\dagger \mathbf{R}(J,J)^\dagger \overline{\mathbf{H}}_{j_1 j_1}^\dagger \overline{\mathbf{H}}_{j_1 j_1}(J,J)\mathbf{R}(J,J)\mathbf{e}}{4}}\right) \\
&\leq \exp\left(\frac{\rho \mathbf{e}^\dagger \mathbf{R}(J,J)^\dagger \overline{\mathbf{H}}_{j_1 j_1}(J,J)^\dagger \overline{\mathbf{H}}_{j_1 j_1}(J,J)\mathbf{R}\mathbf{e}}{-4}\right) \\
&= \exp\left(-\frac{\rho\lambda}{4}\right) \qquad (39)
\end{aligned}$$

where

$$\lambda = ||\mathbf{H}_{j_1 j_1}^1(J,J)(1)||_F^2 |\widehat{c}_1 + \widehat{c}_2|^2 + ||\mathbf{H}_{j_1 j_1}^2(J,J)(1)||_F^2 |\widehat{c}_1 - \widehat{c}_2|^2 \qquad (40)$$

and

$$\widehat{\mathbf{c}} = \begin{pmatrix} \widehat{c}_1 \\ \vdots \\ \widehat{c}_{J_1} \end{pmatrix} = \mathbf{R}\mathbf{e} \qquad (41)$$

Since

$$||\mathbf{H}_{j_1 j_1}^1(J,J)(1)||_F^2 \geq \frac{||\mathbf{H}_{j_1 j_1}(J,J)||_F^2}{2} \cdot \frac{1}{2} \qquad (42)$$

Inequality (39) can be written as

$$\begin{aligned}
&P(\mathbf{c} \to \overline{\mathbf{c}} | \overline{\mathbf{H}}_{j_1 j_1}(J,J)) \leq \exp\left(-\frac{\rho\lambda}{4}\right) \\
&= \exp\left(-\frac{\rho ||\mathbf{H}_{j_1 j_1}(J,J)||_F^2 |\widehat{c}_1 + \widehat{c}_2|^2}{16}\right) \qquad (43)
\end{aligned}$$

Therefore, we have

$$\begin{aligned}
&P(\mathbf{c} \to \overline{\mathbf{c}}) = E[P(\mathbf{c} \to \overline{\mathbf{c}} | \overline{\mathbf{H}}_{j_1 j_1}(J,J))] \\
&= E\left[\exp\left(-\frac{\rho ||\mathbf{H}_{j_1 j_1}(J,J)||_F^2 |\widehat{c}_1 + \widehat{c}_2|^2}{16}\right)\right] \\
&= \frac{1}{\prod_{j=1}^{J_1^2}(1 + \frac{\rho\tau}{16})} \qquad (44)
\end{aligned}$$

where

$$\tau = |\widehat{c}_1 + \widehat{c}_2|^2 \qquad (45)$$

At high SNR region, (44) can be written as

$$P(\mathbf{c} \to \overline{\mathbf{c}}) \leq \left(\frac{\rho\tau}{16}\right)^{-J_1^2} \qquad (46)$$

So the diversity is J_1^2, full diversity, as long as $\tau \neq 0$. Also the coding gain is affected by τ and we can choose rotation matrix $\mathbf{R}(J,J)$ properly to maximize τ. The best choice for rotation matrix depends on the adopted constellation. Such an optimization is a straightforward optimization that has been discussed in many existing literature [19]. Similarly, we can prove that the diversity for other codewords is also full.

VI. Simulation Results

In this section, we provide simulation results to evaluate the performance of the proposed scheme. First, we assume there are 3 transmitters each with 3 transmit antennas and 3 receivers each with 3 antennas. Each user uses our proposed scheme to transmit Alamouti codes to its receiver. Figure 2 presents simulation results using 8-PSK. We compare the performance of our scheme with that of two other scenarios that can achieve interference cancellation. In the first scenario, we use TDMA and beamforming. That is, at each time slot, only one transmitter sends signals to one receiver using beamforming. 64-QAM is used to have the same bit-rate. In the second scenario, each user uses the multi-user detection(MUD) method to send its codewords. The results show that our proposed scheme can achieve full diversity and symbol rate one. Note that we combine the array processing and space-time coding to avoid symbol rate loss. This does not mean that we cannot change the bit rate. We can always adapt the bit rate by changing the constellation according to the channel condition. In comparison, the TDMA and beamforming method can achieve full diversity but the rate is one half. The MUD method can achieve full rate, but it cannot achieve full diversity. As shown in the figure, our scheme provides the best performance due to its high diversity and increased coding gain without any rate loss.

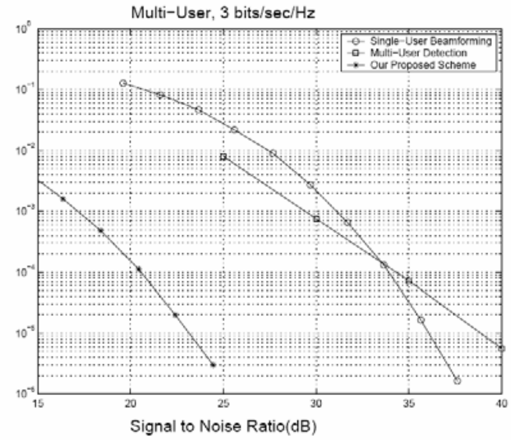

Fig. 2. Simulation results for 3 users each with 3 transmit antennas and 3 receivers each with 3 receive antennas. The constellation is 8-PSK.

VII. Conclusions

In this paper, we propose an orthogonal transmission scheme for Z channels by combining array processing and space-time coding to achieve full diversity and low decoding complexity. To the best of our knowledge, this is the first scheme to achieve the highest possible diversity and low-complexity decoding simultaneously for Z channels when all users transmit simultaneously. We analytically prove that our scheme can achieve low-complexity decoding and full diversity. Simulation results validate our theoretical analysis.

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