A High Rate Open-Loop MIMO Multi-User Downlink Transmission System

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Abstract—In this paper, we first propose a new class of full rate full diversity Quasi-Orthogonal Space Time Block Code (QO-STBC), namely QO-STBC with Blind Precoding (BP-QO-STBC). Based on this code, an open-loop MIMO multi-user downlink transmission scheme is derived. Presuming $M$ blocks of BP-QO-STBC are transmitted from the access point simultaneously, thus achieving a multiplexing gain of $M$, we show that the broadcasted signal can be separated by any user with only $m_R \geq M$ receive antennas, which is much smaller than that required by conventional linear schemes. By exploiting the special structure of BP-QO-STBC, we derived two decoding algorithms with different complexities, namely an optimal Maximum Likelihood Decoder (MLD) and a suboptimal linear decoder (BP-QO-STBC) respectively. We also propose in [5] for the uplink. However our scheme, differs in the following aspects: Firstly, our scheme is based on the blind linear precoding techniques of BP-QO-STBC, whereas the scheme proposed in [5] is based on the differential decoding of QO-STBC [6]. Secondly, by utilizing the special structure of the code our scheme achieves the lowest decoding complexity (which we denote as LDC1), for all known full rate full diversity QO-STBC, which is only half the complexity in [5]. Thirdly, our scheme is more flexible in the tradeoff between complexity and performance. Specifically, we can use a very simple suboptimal linear decoder to achieve a transmit diversity of 2 which is not available for the scheme in [5]. The simulation results also show that without any feedback from the user terminals to the access point, our proposed scheme performs better than RBF in terms of average BER.

Throughout this paper, we use bold upper case letters to denote matrices and bold lower case letters to denote vectors. Superscripts $(\cdot)^*$, $(\cdot)^T$, $(\cdot)^H$, $(\cdot)^R$, $(\cdot)^I$ denote conjugation, transpose, Hermitian transpose, real and imaginary part respectively. $I_n$ denotes a $n \times n$ identity matrix, $\theta_n$ denotes a $n \times n$ all zero matrix and $\theta_n'$ denotes a $n \times 1$ all zero vector. We also let $\mathcal{C}$ and $\mathbb{R}^+$ denote the field of complex numbers and positive real numbers respectively. $A \otimes B$ denotes the Kronecker product of matrix $A$ and $B$.

I. INTRODUCTION

Closed-loop MIMO multi-user schemes such as Dirty Paper Coding (DPC) [1] and Random Beamforming (RBF) [2] [3] generally have good performance. But their requirements for perfect channel state information (CSI) feedback [1], and sufficiently large user pool [2] [3] respectively are usually unrealistic for practical implementation.

To solve the practical problems for the two schemes above, we propose an open-loop MIMO multi-user system based on full rate full diversity QO-STBC, which we named as BP-QO-STBC(QO-STBC with Blind Precoding). We assume $M$ BP-QO-STBC codewords with size $n_T^{(m)} \times n_T^{(m')}$, $m = 1, 2, \cdots, M$ are transmitted from the access point simultaneously, thus achieving a spatial multiplexing gain of $M$. By exploiting the special structure of BP-QO-STBC, we show that the multiplexed BP-QO-STBC codewords can be easily separated by users with only $m_R \geq M$ receive antennas, instead of $m_R \geq M \times n_T$ [4]. Transmit diversity order of $n_T^{(m)}$ is achieved, and the receive diversity order of $r$ can be achieved when the number of receive antennas $m_R = M + r - 1$, where $r$ is an integer and $r \geq 1$.

A similar multi-user interference cancellation technique is also proposed in [5] for the uplink. However our scheme,
The symbols\(^2\) for the \(K\) users are denoted by \(s_k, k = 1, 2, \ldots, K\) which are encoded by \(M\) BP-QO-STBC blocks where the \(m\)th BP-QO-STBC block can be expressed as

\[
G_{mT}^H = \sum_{k=(m-1)\times T+1}^{m\times T} (s_k^R A_k + j s_k^I B_k) \tag{1}
\]

where \(s_k = [s_k^R + j s_k^I, \ldots, s_k^{R^{K-1}} + j s_k^{I^{K-1}}]^T\), subjecting to the power constraint \(E[s_k]\) and \(\rho\) is the total transmit power at the access point. Matrices \(A_k \in C^{T\times n_T}\) and \(B_k \in C^{T\times n_T}\) are the linear dispersion matrices and \(T\) is the codeword length in time domain. For our full rate BP-QO-STBC, \(T = n_T\). The received matrix for user \(k\) is given as

\[
Y_k = \sum_{m=1}^{M} G_{mT}^H H_k(m) + N_k \tag{2}
\]

where \(Y_k = [y_{k,1}, y_{k,2}, \ldots, y_{k,m\times T}]\) is the received signal matrix of user \(k\) over \(T\) time slots and \(y_{k,j} = [y_{k,j,1}, y_{k,j,2}, \ldots, y_{k,j,T}]^T\). The matrix \(H_k(m) = [h_{k,j,1}, h_{k,j,2}, \ldots, h_{k,j,m\times T}]\) is the channel coefficient matrix between user \(k\) and the \(n_T\) transmit antennas for the \(m\)th BP-QO-STBC codeword, where \(h_{k,j}(m) = [h_{k,j,1}(m), h_{k,j,2}(m), \ldots, h_{k,j,m\times T}(m)]\) is the channel coefficient vector from the \(n_T\) transmit antennas for the \(m\)th BP-QO-STBC codeword to the \(j\)th receive antenna of user \(k\). For i.i.d. Rayleigh fading channel, we have \(h_{k,j}(m) \sim CN(\tilde{\Theta}_{n_T}, I_{n_T})\). The noise matrix is denoted by \(N_k = [n_{k,1}, n_{k,2}, \ldots, n_{k,m\times T}]\) where \(n_{k,j} = [n_{k,j,1}, n_{k,j,2}, \ldots, n_{k,j,T}]\) is the noise vector for the \(j\)th receive antenna of user \(k\) and \(n_{k,j} \sim CN(\tilde{\Theta}_{n_T}, I_{n_T})\). We also assume a block fading channel such that the channel is static in an interval of \(T\) time slots. It is obvious that the objective of the receiver at user \(k\) is to retrieve his own raw data symbol from \(Y_k\).

### III. BP-QO-STBC Interference Cancellation

In this section, we illustrate the proposed interference cancellation method by considering a simple scenario, where \(M = 2\) BP-QO-STBC codewords with \(n_T = 4\) are transmitted simultaneously. Since the derivation is applicable for all users, we drop the subscript \(k\) and assume \(m_{(k)} = m_{(i)} = 2 \forall k, i = 1, 2, \ldots, K\). The results derived here is generalized in [8]. With the above assumptions, (2) can be simplified as

\[
Y = G_1^H H(1) + G_2^H H(2) + N \tag{3}
\]

Let \(y_1\) and \(y_2\) denote the first and second column of \(Y\) respectively, and we obtain the following equation which is equivalent to (3),

\[
y_j = G_j^H h_j(1) + G_j^H h_j(2) + n_j \tag{4}
\]

where \(y_j = [y_{j,1}, y_{j,2}, y_{j,3}, y_{j,4}]^T, j = 1, 2, h_j(m) = [h_{j,m,1}, h_{j,m,2}, h_{j,m,3}, h_{j,m,4}]^T, m = 1, 2\)

\(^2\)As explained in the next section, do note that \(s_k\) does not correspond to the raw data symbol for the \(k\)th user.

The \(j\)th column of \(H(m)\) and \(n_j = [n_{j,1}, n_{j,2}, n_{j,3}, n_{j,4}]^T\) denotes the \(j\)th column of \(N\).

From the ABBA code given by [9], we have

\[
G_1^4 = G_2^4 = \begin{bmatrix}
\begin{array}{cccc}
s_1 & s_2 & s_3 & s_4 \\
-\bar{s}_2^* & \bar{s}_1^* & -\bar{s}_4 & \bar{s}_3^* \\
s_3 & s_4 & s_1 & s_2 \\
-\bar{s}_4^* & -\bar{s}_3^* & \bar{s}_1^* & \bar{s}_2^*
\end{array}
\end{bmatrix} \tag{5}
\]

Substituting them into (4), and after some manipulations we arrive at the more intuitive expression given by

\[
y_j' = H_j^4(1) s_1 + H_j^4(2) s_2 + n_j' \tag{7}
\]

for \(j = 1, 2\), where

\[
H_j^4(m) = H_4^4 \{h_j(m, t)\}_{t=1}^{T-4} = \begin{bmatrix}
h_j(m, 1) & h_j(m, 2) & h_j(m, 3) & h_j(m, 4) \\
h_j(m, 2)^* & -h_j(m, 1)^* & h_j(m, 4)^* & -h_j(m, 3)^* \\
h_j(m, 3) & h_j(m, 4) & h_j(m, 1) & h_j(m, 2)^* \\
h_j(m, 4)^* & -h_j(m, 3)^* & h_j(m, 2)^* & -h_j(m, 1)^*
\end{bmatrix} \tag{8}
\]

is the equivalent channel matrix from transmit antennas for \(m\)th BP-QO-STBC block to the \(j\)th receive antenna of the user. \(y_j' = [y_{j,1}, y_{j,2}, y_{j,3}, y_{j,4}]^T, s_1 = [s_1, s_2, s_3, s_4]^T, s_2 = [s_5, s_6, s_7, s_8]^T\), and \(n_j' = [n_{j,1}, n_{j,2}, n_{j,3}, n_{j,4}]^T\).

### A. Interference Cancellation

Suppose the desired symbol of this specific user is in \(s_2\), then \(s_1\) is interference and has to be eliminated. At both of the two receive antennas, matched filtering is applied,

\[
H_j^4(1)^H y_j' = H_j^4(1, 1) s_1 + H_j^4(1, 2) s_2 + H_j^4(1)^H n_j' \tag{9}
\]

where

\[
H_j^4(1, 1) = H_j^4(1)^H \begin{bmatrix}
\alpha_j \\
\beta_j
\end{bmatrix} \odot I_2
\]

\[
H_j^4(1, 2) = H_j^4(1)^H \begin{bmatrix}
\alpha_j \\
\beta_j
\end{bmatrix} \odot I_2 \tag{10}
\]

and \(\alpha_j = |h_j(1, 1)|^2 + |h_j(1, 2)|^2 + |h_j(1, 3)|^2 + |h_j(1, 4)|^2, \beta_j = 2(h_j(1,1) h_j(1, 3)^* + h_j(1, 2) h_j(1, 4)^*), j = 1, 2\).

Since \(H_j^4(1, 1)\) are circular matrices, they can be diagonalized easily as

\[
H_j^4(1, 1) = U_4 A_j^4 U_4^H \tag{11}
\]

where \(A_j^4 = \text{diag}[d_{j,1}, d_{j,2}, d_{j,3}, d_{j,4}], d_{j,1}, d_{j,2} \in \mathbb{R}^+\) are eigenvalues of \(H_j^4(1, 1)\). Unitary matrix \(U_4\) is the eigenvector matrix and can be expressed as

\[
U_4 = \frac{1}{\sqrt{2}} \begin{bmatrix}
1 & 1 \\
1 & -1
\end{bmatrix} \odot I_2 \tag{12}
\]

It is obvious now if we perform precoding to the raw data vector \(x_1 = [x_1, x_2, x_3, x_4]^T\), with \(U_4\), i.e. \(s_1 = U_4 x_1\), then \(x_1\) is decoupled at the receiver by postcoding with \(U_4^H\). It can also be easily verified that with the same precoding
and postcoding to $x_2 = [x_5, x_6, x_7, x_8]^T$, $H^2_j(1, 2)$ can be block diagonalized and each block in the diagonal is an Alamouti’s STBC block [10]. The equations with precoding and postcoding at the receive antennas can be expressed as

$$r_j = \Lambda_{j}^2 x_1 + \left[ D_{j,1}^2 \theta_2 \begin{array}{c} 1 \\ \end{array} D_{j,2}^2 \right] x_2 + U^H_j H^2_j(1, 1)^H n_j$$

(13)

where $r_j = [r_{j,1}, r_{j,2}, r_{j,3}, r_{j,4}]^T$, $j = 1, 2$ and $D^2_{j,i}, l = 1, 2$ is an Alamouti’s STBC block$^3$ whose elements are combinations of channel coefficients. It is obvious that $x_1$ can be simply cancelled by using $r_1$ and $r_2$ such that

$$\hat{r}_1 = \begin{bmatrix} \frac{r_{1,1}}{d_{1,1}} \\ \frac{r_{1,2}}{d_{1,2}} \end{bmatrix} - \frac{1}{d_{2,1}} \begin{bmatrix} r_{2,1} \\ r_{2,2} \end{bmatrix} = \hat{H}^A_1 x_5 + \hat{n}_1$$

(14)

$$\hat{r}_2 = \begin{bmatrix} \frac{r_{1,3}}{d_{1,2}} \\ \frac{r_{1,4}}{d_{1,2}} \end{bmatrix} - \frac{1}{d_{2,2}} \begin{bmatrix} r_{2,3} \\ r_{2,4} \end{bmatrix} = \hat{H}^A_2 x_7 + \hat{n}_2$$

(15)

where $\hat{r}_1 = [\hat{r}_{1,1}, \hat{r}_{1,2}]^T$, $\hat{r}_2 = [\hat{r}_{1,3}, \hat{r}_{1,4}]^T$, $\hat{n}_1 = [\hat{n}_{r,1}, \hat{n}_{r,2}]^T$, $\hat{n}_2 = [\hat{n}_{r,3}, \hat{n}_{r,4}]^T$, $\hat{H}^A_1 = \frac{D_{1,1}^A}{d_{1,1}}$ and $\hat{H}^A_2 = \frac{D_{1,2}^A}{d_{1,2}}$. Due to the closure properties of Alamouti’s STBC blocks [11], $\hat{H}^A_1$ and $\hat{H}^A_2$ are also Alamouti’s STBC blocks. We let

$$\hat{H}^A_1 = \begin{bmatrix} \hat{n}_{1,1} \\ -\hat{n}_{1,2} \end{bmatrix}, \quad \hat{H}^A_2 = \begin{bmatrix} \hat{n}_{2,3} \\ -\hat{n}_{2,4} \end{bmatrix}$$

(16)

where $\hat{n}_1 \in C, t = 1, 2, 3, 4$. We also note that $\hat{n}_1$ and $\hat{n}_2$ are still i.i.d Gaussian noise vectors.

It is clear now that the transmission of $x_2$ is equivalent to two parallel independent Alamouti’s STBC transmissions, both of which are free from interference from $x_1$ and each other. The receiver can then apply the standard decoding method [10] for an Alamouti code to retrieve symbols of interest.

**B. Full Transmit Diversity**

However, with the above precoding, it is easy to verify [8] that the minimum determinant of the codeword distance matrix will be zero, i.e. $\text{det}_{\text{min}}(C - E)(C - E)^H = 0$, where $C \neq E$, $C, E \in \Theta$ and $\Theta$ is the codebook. This means that if we apply (14) and (15) directly, this decoding method will not achieve full transmit diversity [12]. In fact, we can see this more intuitively from (14) and (15), where $x_5$ and $x_6$ are only transmitted through the sub-equivalent channel $\hat{H}^A_1$ but not through sub-equivalent channel $\hat{H}^A_2$, whereas $x_7$ and $x_8$ are only transmitted through $\hat{H}^2_2$ but not $\hat{H}^A_1$. Since the transmission of each symbol does not utilize both sub-equivalent channels, full transmit diversity cannot be achieved.

In order to achieve full diversity, we introduce a unitary linear transform matrix, namely $R$ to “disperse” symbols evenly throughout all the sub-equivalent channels and hence make it possible for $\text{det}_{\text{min}}(C - E)(C - E)^H > 0$ with any $C \neq E$ by applying proper constellation rotation [14] to some of the transmitted symbols. $R$ for $n_T = 4$ is given as

$$R_4 = \frac{1}{\sqrt{2}} \begin{bmatrix} 1 \\ j \\ -1 \\ -j \end{bmatrix} \otimes I_2$$

(17)

$^3$Matrix with superscript $A$ denotes a $2 \times 2$ matrix having the form of an Alamouti’s STBC block [10].

**C. Lowest Decoding Complexity**

In order to achieve lowest decoding complexity (LDC) [7], we apply signal mapping to ensure that at the receiver the number of symbols to be jointly decoded is minimized,

$$\pi_i = (x_i) + j\theta_4(x_i)^4$$

(18)

where $i = 1, 2$, and the rotation matrix

$$\theta_4 = \begin{bmatrix} \theta_0 I_2 \\ I_2 \theta_2 \end{bmatrix}$$

(19)

Finally, the two transmit signal matrices of BP-QO-STBC for $n_T = 4$ are $G_1 = \frac{1}{\sqrt{2}} (s_{k,1})^\ast$ and $G_2 = \frac{1}{\sqrt{2}} (s_{k,2})^\ast$ as given in (5) and (6), where $s_1 = U_4 R_1 \pi_1$ and $s_2 = U_4 R_4 \pi_2$.

Now, after matched filtering in (14) and (15) we arrive at

$$\hat{r}_1 = (\hat{H}^A_1)^H \hat{n}_1 = \frac{1}{\sqrt{2}} \begin{bmatrix} \gamma_1 \\ 0 \end{bmatrix} \begin{bmatrix} x_5^R - x_3^R + j(x_5^R + x_3^R) \\ 0 \end{bmatrix} + \hat{n}_1$$

(20)

$$\hat{r}_2 = (\hat{H}^A_2)^H \hat{n}_2 = \frac{1}{\sqrt{2}} \begin{bmatrix} 0 \\ \gamma_2 \end{bmatrix} \begin{bmatrix} -x_7^R + x_5^R - j(x_7^R + x_5^R) \\ 0 \end{bmatrix} + \hat{n}_2$$

(21)

where $\gamma_1, \gamma_2 \in \mathbb{T}^+$. From (20) and (21) we can see that each symbol is “dispersed” across both sub-equivalent channels, thus achieving full transmit diversity. Note that since $\gamma_1, \gamma_2 \in \mathbb{T}^+$, we can freely separate $x_k, k = 5, 6, 7, 8$ by interchanging some of the elements in $\hat{r}_j^R$ and $\hat{r}_j^I$, $j = 1, 2$ to obtain

$$\hat{r}_1 = \frac{1}{\sqrt{2}} \begin{bmatrix} \gamma_1 \gamma_1 \\ 0 \gamma_2 \end{bmatrix} \begin{bmatrix} x_5^R - x_3^R \\ -x_7^R - x_5^R \end{bmatrix} + \hat{n}_1$$

(22)

$$\hat{r}_2 = \frac{1}{\sqrt{2}} \begin{bmatrix} \gamma_1 \gamma_2 \gamma_1 \\ 0 \gamma_2 \end{bmatrix} \begin{bmatrix} x_5^R - x_6^R \\ -x_7^R - x_6^R \end{bmatrix} + \hat{n}_2$$

(23)

$$\hat{r}_3 = \frac{1}{\sqrt{2}} \begin{bmatrix} \gamma_1 \gamma_2 \gamma_1 \\ 0 \gamma_2 \end{bmatrix} \begin{bmatrix} x_5^R + x_3^R \\ -x_7^R + x_3^R \end{bmatrix} + \hat{n}_3$$

(24)

$$\hat{r}_4 = \frac{1}{\sqrt{2}} \begin{bmatrix} \gamma_1 \gamma_2 \gamma_1 \\ 0 \gamma_2 \end{bmatrix} \begin{bmatrix} x_5^R + x_6^R \\ -x_7^R + x_6^R \end{bmatrix} + \hat{n}_4$$

(25)

We can see that each of these 4 vectors contains only one of the 4 symbols in $x_2$, and the user need to solve only one of the above equations to decode the intended symbol just by linear decoding and without performing any joint detection, which is required for all the other known QO-STBC in [5]–[7], [9], [13]. For example, the decoding for $x_5$ is given as

$$\hat{r}_1 = \frac{1}{\sqrt{2}} \begin{bmatrix} x_5^R \\ x_5^I \end{bmatrix} + \frac{1}{\gamma_1} \begin{bmatrix} -1 \\ -\gamma_2 \end{bmatrix} \hat{r}_1$$

(26)

However, since the linear decoding basically counteracts the effect of matrix $R$, and thus it cannot achieve full transmit diversity as stated earlier. Alternatively, by performing MLD to (14) and (15), much better performance in terms of coding gain and diversity gain can be achieved at the expense of higher complexity. The MLD can be performed by minimizing the metrics given in (27). It is obvious that joint detection of only two real symbols (=1 complex symbol) is required, thus the single symbol decoding [13] (which is equivalent to LDC when $n_T = 4$) is achieved.
Generally, when \( M \) BP-QO-STBC blocks are transmitted simultaneously, with \( m_R = M \) receive antennas we will have \( M \) equations with the form of (7), each of which contains symbols from \( M \) BP-QO-STBC codewords. After cancelling the effect of the first BP-QO-STBC codeword, we have \( M - 1 \) sets of independent equations having the form of (14) and (15), each of which contains symbols from the \( M - 1 \) remaining BP-QO-STBC codewords. Due to the closure property under multiplication and addition for Alamouti’s STBC blocks, we can easily continue to perform matched filtering and then eliminate the second BP-QO-STBC codeword and so on. After eliminating \( k \) BP-QO-STBC codewords, we have \( M - k \) independent sets of equations in the form of (14) and (15) left. In the end, we are left with only a single set of equations to decode the desired signal without interference from other BP-QO-STBC codewords. When the number of receive antennas \( m_R = M + r - 1 \), where \( r \geq 1 \), we have \( r \) sets of equations left, thus providing a receive diversity order of \( r \). Since the extension is straightforward, we omit the detailed proof due to the space limit.

It has been shown in [12] that full rate full diversity QO-STBC is able to support an arbitrary number of transmit antennas by eliminating columns from a QO-STBC with codeword length of power of 2. Since this eliminating operation will not affect \( \tilde{H}_1 \) and \( \tilde{H}_2 \) in (14) and (15) to have a structure of Alamouti’s STBC block, our interference cancellation scheme discussed above can be applied directly to support BP-QO-STBC with \( n_T < T \). Furthermore, due to the iterative construction method of BP-QO-STBC, multiplexed codewords with different codeword size can be separated by the same method without any extra effort. Thus our proposed scheme is able to support any number of transmit antennas with a tradeoff between spatial multiplexing gain and transmit diversity. The detailed proofs are provided in [8].

It is also possible to implement our BP-QO-STBC based scheme in a round-robin fashion, where at each \( T \) time slots, \( M \) BP-QO-STBC blocks containing the symbols for one user are transmitted, and all the \( K \) users are served one after another. However, the implementation discussed here is superior in the following two points: Firstly, symbols of each user are transmitted through \( K \times T \) time slots instead of only \( T \) time slots for the TDMA implementation, and thus extra diversity gain in time domain can be obtained with the help of channel coding. Secondly, time slot resolution in our scheme is higher than that for the TDMA implementation, thus allowing for a more efficient resource allocation among users to meet their different QoS requirements.

IV. SIMULATION RESULTS

In this section, we compare the performance of our proposed scheme with the multi-user interference cancellation scheme proposed in [5], as well as a simple RBF scheme [2] [3], which requires the feedback of SINR from all the active users.

The comparison results of our proposed scheme and the scheme in [5] are shown in Fig. 1. For our proposed scheme, two kinds of receiver are applied: the suboptimal linear decoder (26) achieves transmit diversity order of 2 without joint detection, whereas full transmit diversity order is achieved by performing the MLD with LDC as discussed in the previous section. Similar performance is also achieved by the scheme in [5]. However, for the case of \( n_T = 4 \), joint detection of 2 complex symbols is required for [5]. Generally, the decoding for scheme proposed in [5] needs joint detection of \( n_T/2 \) complex symbols, whereas joint detection of only 2 complex symbols is required for our BP-QO-STBC scheme.

In Fig. 2, we compare the average BER performance of all users for our proposed scheme and a simple RBF scheme. For RBF, we follow the scheme proposed in [3], where \( n_T \) spatially orthogonal beams are formed by the transmitter randomly, and then transmissions to the users with the highest SINR are performed. Here we assume that there are \( W \geq n_T \) active users to choose from and each terminal has \( m_R = 2 \) receive antennas. We omit the proportional fair user scheduling and water-filling power allocation algorithms here for simplicity. We consider RBF with two configurations: one is with \( n_T = 8 \), \( W = 500 \), the other one is with \( n_T = 4 \), \( W = 100 \), and both of them apply BPSK modulation. For our proposed scheme, BP-QO-STBC with \( n_T^{(m)} = 4 \) \( \forall m \) is applied and two cases where \( M = 1 \) and \( M = 2 \) are considered. In order to achieve the same data rate, we apply 16QAM for both cases of \( n_T = 4 \) and \( n_T = 8 \).

It is clear to observe the significant performance gap at high SNR region between RBF and our proposed scheme, and this result coincide with the conclusion derived in [2].

4 RBF can serve at most \( n_T \) out of the \( W \) active users.
eliminated and the performance is only dominated by the mismatch which will dominate the performance of RBF.

However for our proposed scheme, all the interference is mitigated and this “mismatch” will dominate the performance of RBF.

It is more difficult to find a set of spatially orthogonal users whose channels “match” the randomly generated beamformer, and this “mismatch” will dominate the performance of RBF. However for our proposed scheme, all the interference is eliminated and the performance is only dominated by the AWGN.

V. CONCLUSION

We proposed a high rate multi-user downlink transmission scheme based on BP-QO-STBC, by which spatial multiplexing gain and transmit diversity gain is achieved at the same time, without the need for CSI at the transmitter. Due to the inherent structure of BP-QO-STBC, the minimum number of receive antennas needed for each user to separate signal from $M$ multiplexed BP-QO-STBC codewords is only $M$, instead of $M \times n_T$ which is required for conventional linear method, and the receive diversity order is readily achieved by using extra receive antennas.

A suboptimal linear receiver is possible for our scheme to achieve a low decoding complexity at the expense of some loss in diversity gain. However, full transmit diversity order is achieved by using our proposed MLD which achieves LDC. It is also shown that our proposed scheme has a better average BER performance than RBF which requires feedback of SINR from all the active user terminals.

REFERENCES


