Cooperative Multi-User MIMO Wireless Systems Employing Precoding and Beamforming

Wibowo Hardjawana, Branka Vucetic, Yonghui Li, and Zhendong Zhou
Telecommunications Lab, School of Electrical and Information Engineering
The University of Sydney, Australia
{whar5698*, branka, lyh, kyle}@ee.usyd.edu.au

ABSTRACT

Interference among multiple base stations that co-exist in the same location limits the capacity of wireless networks. In this paper, we propose a method to design a spectrally efficient cooperative downlink transmission scheme employing precoding and beamforming for multi-user multiple-input-multiple-output (MIMO) systems. The algorithm eliminates the interference and achieves symbol error rate (SER) fairness among different users. To eliminate the interference, Tomlinson Harashima precoding (THP) is used to cancel part of the interference while the transmit-receive antenna weights are chosen to cancel the remaining interference. A novel iterative method is applied to generate the transmit-receive antenna weights. To achieve SER fairness among different users and further improve the performance of multi-user MIMO systems, we develop algorithms that provide equal signal-to-interference-plus-noise-ratio (SINR) across all users. The users are also ordered so that the minimum SINR for each user is maximized. The simulation results show that the proposed scheme considerably outperforms existing cooperative transmission schemes in terms of the SER performance and complexity and approaches an interference free performance under the same configuration.

I. INTRODUCTION

The spectral efficiency in existing cellular mobile [1] and WLANs [2] networks is limited by interference. In cellular mobile networks, the dominant interference comes from adjacent cells [1], while in co-working WLANs [2], the interference from other networks, operating in the same area, is a major limiting factor [2]. In the cooperative transmission scheme proposed here, multiple base stations (BSs) share information about the transmitted messages to their respective users and wireless channels via a backbone network. Individual BSs are equipped with multiple transmit antennas. Each BS transmitter uses the information of the transmitted signals from other BSs and wireless channel condition to precode its own signal. The precoded signal for each BS is broadcast through all BS transmit antennas in the same frequency band at a given time slot. The precoding operation and transmit-receive antenna coefficients are chosen in such a way as to minimize the interference coming from other BS transmissions. The calculated receive antenna coefficients are then sent from the transmitter to the receiver through the wireless channel prior to the data transmission. As MIMO is a widely accepted technology for all future wireless standards, due to its high spectral efficiency, we consider multiple antennas at both the transmitter and receivers.

A multi-user MIMO systems with multiple transmit-receive antennas has been considered by several researchers. In [3], the authors showed how a ZF method can be used to exploit the availability of multiple receive antennas. Here transmit-receive antenna weights are first jointly optimized by a ZF diagonalization technique and then a water-filling power allocation method is applied to allocate power to each user. The scheme in [3] is further improved in [4] by using an iterative method. Nonlinear methods, utilizing a combination of a ZF method with DPC and a combination of a ZF method with THP [5], [6] for a multi-user MIMO system were considered in [7], [8], respectively. The authors use the ZF method to eliminate part of the inter-link interference. DPC or THP are then used to cancel the remaining interference. These schemes, however, are not practical for cooperative MIMO systems, since their symbol-error-rate (SER) performance varies from user to user. In particular, this SER variation is not desirable since the MIMO systems can be deployed by different operators and they expect the systems to have similar performance.

In this paper, we propose a cooperative transmission scheme employing precoding and beamforming for the downlink of multi-user MIMO systems. In this algorithm, the THP cancels part of the interference while the transmit-receive antennas weights cancel the remaining interference. A new novel iterative method is used to generate the transmit-receive antennas weights. This transmit-receive antenna weights are optimized based on the iterative optimization method in [9]. The receive and transmit weights are optimized iteratively until the SINR for each user converges to a fixed value. The convergence behaviour of the proposed method is investigated numerically. In addition to the iterative joint transmit-receive antenna weights optimization and THP above, we also employ SINR equalization, and an adaptive precoding ordering (APO) in the algorithm. The SINR equalization process [10] is used to allocate power to users in such a way that all users have the same SINRs, which ensures SER fairness among all users. The APO is then used to further improve the performance of MIMO systems by maximizing the minimum SINR for each user [11]. Simulation results show that the proposed scheme is significantly superior to the existing methods and is only 0.25 dB away from an interference free channel, under the
The proposed method offers a significant improvement over a nonlinear cooperative precoding algorithm presented in [3], [4], [7], [8]. The first improvement is the enhancement of the SER performance compared to [3], [4], [7], [8]. The second improvement is the relaxation of the zero forcing constraints. Unlike [7], [8], here we allow transmit signals intended for different users to interfere with each other. This interference is cancelled at the receiver where the signal is multiplied by the receive antenna weights. The third improvement comes from the complexity reduction. The proposed scheme has a much lower computational complexity than the methods in [3], [4], [7], [8]. The fourth improvement is that it can be written as

\[ y = \text{RHTPu + RN} = (D + F + B)Pu + RN \]

where \( R = \text{Diag}(r_1, \ldots, r_K) \), \( D = DiT(\text{RHT}) \), \( B = UpT(\text{RHT}) \), \( F = LoT(\text{RHT}) \) and \( y = [y_1, \ldots, y_K]^T \), \( y_j \) is the received signal at the input of the THP decoder for link \( j \), \( r_j \) is the receive antenna weights vector for link \( j \). \( \text{DPu} \) is a vector of scaled replicas of the transmitted symbols for \( K \) links. \( \text{FP} \) is defined as the front-channel interference matrix, since the rows \( j = 1, \ldots, K \) of \( \text{FP} \) represent the interference caused by \( \text{front} \ links 1, \ldots, j - 1 \). Similarly, \( \text{BP} \) is defined as the rear-channel interference matrix, since the rows \( j = 1, \ldots, K \) of \( \text{BP} \) represent the interference caused by \( \text{rear} \ links j + 1, \ldots, K \).
precodes \( \mathbf{u} \) into \( \mathbf{v} = [v_1 \cdots v_j \cdots v_K]^T \in \mathbb{C}^{K \times 1} \). The front-channel interference is then subtracted from \( \mathbf{u} \) as in [12],
\[
\mathbf{v} = \mathbf{u} + \mathbf{d} - (\mathbf{D}\mathbf{P})^{-1}\mathbf{FPv} \tag{2}
\]
where \( (\mathbf{D}\mathbf{P})^{-1} \) is used to normalize the front-channel interference with respect to \( \mathbf{u} \). \( \mathbf{d} = [d_1 \cdots d_K]^T \), \( d_j = 2\sqrt{M}\Delta \) and \( \Delta \) is a complex number whose real and imaginary parts are suitable integers selected to ensure the real and imaginary parts of \( \Delta \) are constrained into \((-\sqrt{M}, \sqrt{M})\). Here, we use the vector \( \mathbf{d} \) as an offset to ensure the energy of \( \mathbf{v} \) to be between \((-\sqrt{M}, \sqrt{M})\), since the value of \( \mathbf{v} \) after pre-subtraction of the front-channel interference can be very large and exceed \((-\sqrt{M}, \sqrt{M})\). By using a modulo operation, defined as [13]
\[
\text{mod}_M(\mathbf{u}_j) = \mathbf{u}_j - \sqrt{M}\left[\frac{\mathbf{u}_j + \sqrt{M}}{2}\right]/\sqrt{M} \tag{3}
\]
for \( j = 1, \ldots, K \), the output of the THP precoder in (2) can be re-written as
\[
\mathbf{v}_j = \text{mod}_M(\mathbf{u}_j - \sum_{i=1}^{j-1}[\mathbf{M}_{THP}]_{j,i}\mathbf{v}_i), \quad j = 1, \ldots, K \tag{4}
\]
where \([\mathbf{M}_{THP}]_{j,i}\) denotes the \((j, i)^{th}\) component of \( \mathbf{M}_{THP} \).

Since now we are using the THP scheme, we are transmitting THP precoded symbol streams, \( \mathbf{v} \) instead of \( \mathbf{u} \). The received signal \( \mathbf{y} \) is now given by (1) with \( \mathbf{u} \) is replaced by \( \mathbf{v} \). By using (1) and (2), the received signal is now given as
\[
\mathbf{y} = (\mathbf{D} + \mathbf{F})\mathbf{P}\mathbf{v} + \mathbf{B}\mathbf{P}\mathbf{v} + \mathbf{RN} = \mathbf{DPv} + \mathbf{BPv} + \mathbf{RN} \tag{5}
\]
The estimates of the transmitted symbols for link \( j \), denoted by \( \hat{\mathbf{u}}_j \), can be recovered from \( \mathbf{y}_j \) by applying an element-wise modulo operator in (3) to each \( \mathbf{y}_{\hat{u}_j} \), as
\[
\hat{\mathbf{u}}_j = \text{mod}_M(\mathbf{y}_{\hat{u}_j}), \quad j = 1, \ldots, K \tag{6}
\]
Here, the effect of offset vector \( \mathbf{d} \) on the desired transmitted signal is removed at each MS receiver by applying the modulo operation in (3) to each \( \mathbf{y}_j \) in (5). This is shown in Fig. 1(b). In the proposed scheme, THP cancels the interference caused by the front-channel interference, while the interference caused by the rear-channel interference is eliminated by the transmit-receive antenna weights optimization process.

III. ITERATIVE ANTENNA WEIGHTS OPTIMIZATION

In this section, we propose a joint iterative transmit-receive antenna weights optimization method based on ZF to cancel the rear-channel interference, while maximizing the SINR for each link and maintaining the same SER for all links. To do this, we use the fact that 1) we can set \( \mathbb{E}[\mathbf{vv}^H] = \mathbf{I} \) and 2) the effect of vector \( \mathbf{d} \) on the received signals is completely removed by the THP decoder modulo operation. By using (5), the received downlink SINR for each link \( j \) can then be written as
\[
\text{SINR}_j = \frac{p_j\mathbf{r}_j^H\mathbf{H}_j\mathbf{t}_j(\mathbf{H}_j\mathbf{t}_j)^H\mathbf{r}_j}{\mathbf{r}_j^H(\sum_{i=j+1}^{K} p_i\mathbf{H}_i\mathbf{t}_i(\mathbf{H}_i\mathbf{t}_i)^H + \sigma^2\mathbf{I})\mathbf{r}_j} \tag{7}
\]
where \( \sigma^2 \) is MS receiver noise power. Maximizing the minimum SINR for each link, while maintaining it equal for all links, can be formulated as follows
\[
\max_{\mathbf{R}, \mathbf{T}, \mathbf{P}} \min_{1 \leq i \leq K} \text{SINR}_i \quad \text{subject to} \quad (1) \quad \mathbf{T}^H\mathbf{T} = \mathbf{I}, \quad (2) \quad \mathbf{r}_j^H\mathbf{r}_j = 1 \tag{8}
\]
\[
(3) \quad \mathbf{1}^T\mathbf{p} = P_{\text{max}}, \quad (4) \quad \mathbf{r}_j^H\mathbf{H}_j\mathbf{t}_i = 0
\]
for \( j = 1, \ldots, K, i = j + 1, \ldots, K \) where \( P_{\text{max}} \) and \( \mathbf{p} = [p_1 \cdots p_K]^T = \mathbf{P}^2\mathbf{1} \) are the power constraint at the cooperative transmitter and the set of the powers assigned to each link, respectively. Here the objective of (8) is to maximize the minimum SINR for each link. The first, second and third constraints in (8) are to ensure that the transmit-receive weight vectors are unitary vectors and the sum of the power allocated to each link does not exceed the maximum power available at the transmitter. These constraints will bound the possible solution for \( \mathbf{R}, \mathbf{T}, \mathbf{P} \), and \( \text{SINR}_i \) ensuring the convergence of (8)
to a solution. Finally, the fourth one is the ZF constraint which ensures the interference from links \( j + 1, \ldots, K \) to link \( j \) are fully cancelled. Here, to maximize the minimum SINR in (8), we reduce the SINR of the best link until the SINR of all links are equal. Thus, the optimal solution is reached when all links attain an equal SINR [10], [14]. This optimization problem, however, is difficult to solve as it is not jointly convex in variables \( \mathbf{R}, \mathbf{T} \) and \( \mathbf{p} \). To solve (8), we propose a sub-optimal solution that splits the problem into a 2-step optimization. The first step is to solve \( \mathbf{R} \) and \( \mathbf{T} \) iteratively, when \( \mathbf{p} \) is fixed. Hence in this step we simply ignore the equalization of SINRs among all links. The second step is to solve \( \mathbf{p} \) in a way that equalizes SINR for all links under fixed \( \mathbf{R} \) and \( \mathbf{T} \). The process is described in Fig. 2, where \( i, f_1(\cdot) \) and \( g_j(\cdot) \) are the iteration number, a function to generate transmit antenna weights for \( K \) links and a function to generate receive antenna weights vector for link \( j \), respectively.

\( A. \) First Step

In the first step, we assume an equal power allocation for each link by setting \( \mathbf{P} = \mathbf{I} \). (8) can then be simplified as

\[
\max_{\mathbf{R}, \mathbf{T}} \text{SINR}_j
\]

subject to

\[
\begin{align}
(1) & \quad \mathbf{T}^H \mathbf{T} = \mathbf{I}, \\
(2) & \quad \mathbf{r}_j^H \mathbf{H}_j^H \mathbf{t}_i = 0, \\
(3) & \quad \mathbf{r}_j^H \mathbf{H}_j \mathbf{t}_i = 0
\end{align}
\]

for \( j = 1, \ldots, K, i = j + 1, \ldots, K \). To solve (9), we propose to alternately optimize \( \mathbf{R} \) and \( \mathbf{T} \) until they converge, under the ZF constraint in (8). We first assign the initial value of the receive antenna weights for \( K \) links. The initial receive weights of \( K \) links are given as \( \mathbf{r}_j^{(0)} = \mathbf{u}_{svd}(\mathbf{H}_j^H), j = 1, \ldots, K \) where \( \mathbf{u}_{svd}(\cdot) \) is the Singular Value Decomposition operation (SVD) [15] to select the left singular vector of \( \mathbf{H}_j \) corresponding to the largest singular value. We then transform the system into a downlink multi-link MISO system by fixing

\[
\mathbf{R} = \text{Diag}(\mathbf{T}_1^{(0)H}, \ldots, \mathbf{r}_K^{(0)H}).
\]

(1) can then be written as

\[
\mathbf{y} = \mathbf{R} \mathbf{H} \mathbf{T} \mathbf{v} + \mathbf{R} \mathbf{N} = \mathbf{H}_c \mathbf{H} \mathbf{T} \mathbf{v} + \mathbf{N}.
\]

Here, we know from (5) that the interference from links \( 1, \ldots, j - 1 \) to links \( j = 1, \ldots, K \) does not exist at the receiver, after performing decoding, since this front-channel interference is totally cancelled by the THP described in Section II-C. The remaining interference is the rear-channel interference, coming from links \( j + 1, \ldots, K \) to links \( j = 1, \ldots, K \) which needs to be cancelled. At each iteration, we apply a QR decomposition [15] to \( \mathbf{H}_c^H \) to find \( \mathbf{T} \) that forces this interference to zero.

\[
\mathbf{T} = f_1(\mathbf{R}), \quad f_1(\mathbf{R}) = [\mathbf{Q} | \mathbf{R} \mathbf{H} \mathbf{H}_c^H].
\]

We choose the unitary matrix \( \mathbf{Q} \) obtained from the QR decomposition of \( \mathbf{H}_c^H \) in (11) as \( \mathbf{T} \). We now need to compute \( \mathbf{R} \). Here, we choose \( \mathbf{r}_j \) by aligning it in the direction of \( \mathbf{h}_j \) and by ensuring the total power of receive weight vector \( \mathbf{r}_j \) is normalized to 1,

\[
r_j(i) = \frac{\mathbf{H}_c \mathbf{t}_i(i)}{||\mathbf{H}_c \mathbf{t}_i(i)||}.
\]

This ensures the second constraint in (8) is satisfied and we refer to this receiver structure as a Matched Filter (MF) design.

\( B. \) Second Step

In the second step, we use \( \mathbf{R} \) and \( \mathbf{T} \) obtained in the first step to find \( \mathbf{p} \). Using the fact that at the optimal solution all links will attain equal SINR and letting \( a_{i,j} = \mathbf{r}_i \mathbf{H}_j \mathbf{t}_j, \) (7) can be written as

\[
\sum_{i=1}^{j-1} |a_{j,i}|^2 p_i + \sigma^2 = \frac{p_j |a_{j,j}|^2}{\text{SINR}}.
\]

(13) can be further represented in a matrix format as

\[
\mathbf{A}^{-1} \mathbf{Bp} + \sigma \mathbf{A}^{-1} \mathbf{1} = \frac{\mathbf{p}}{\text{SINR}}
\]

where \( \mathbf{A} = \text{Di} \mathbf{T}(\mathbf{M}), \mathbf{B} = \mathbf{U} p \mathbf{T}(\mathbf{M}) \) and \( \mathbf{M} \) is a \( K \) by \( K \) matrix with entries \( |a_{i,j}|^2 \) in row \( i \) and column \( j \). By multiplying both sides of (14) with \( \mathbf{1}^T \), we obtain [10]

\[
\frac{1}{P_{\text{max}}} (\mathbf{1}^T \mathbf{A}^{-1} \mathbf{Bp} + \sigma \mathbf{1}^T \mathbf{A}^{-1} \mathbf{1}) = \frac{1}{\text{SINR}}.
\]

By defining the extended power vector \( \mathbf{p}_e = [\mathbf{p}^T \mathbf{1}]^T \), we can then combine (14) and (15) to obtain an equations matrix given as [10]

\[
\begin{bmatrix}
\mathbf{A}^{-1} \mathbf{B} & \sigma \mathbf{A}^{-1} \mathbf{1} \\
\mathbf{1}^T \mathbf{A}^{-1} \mathbf{B} / P_{\text{max}} & \sigma \mathbf{1}^T \mathbf{A}^{-1} \mathbf{1} / P_{\text{max}}
\end{bmatrix}
\mathbf{p}_e = \frac{\mathbf{p}_e}{\text{SINR}}.
\]

Hence the optimum \( \mathbf{p} \) can be obtained by selecting \( \mathbf{p}_e \) that corresponds to the maximum eigenvalue of \( \mathbf{\Psi} \). This is the only possible solution for (16) satisfying \( p_j \geq 0 \) for \( j = 1, \ldots, K \) and \( \text{SINR} \geq 0 \). The proof is described in detail in [16].

\( C. \) Adaptive Precoding Order

In the THP and the 2-step optimization process described in the previous sections, we fix the order of \( u_j \), resulting in a fixed permutation matrix \( \mathbf{M}_{\text{perm}} \). The performance of the system, however, differs when a different \( \mathbf{M}_{\text{perm}} \) is used. In addition, the performance of the system also depends on the weakest link. In this section we propose an APO scheme. APO arranges the order of \( x \) by selecting \( \mathbf{M}_{\text{perm}} \) that maximizes the minimum SINR for each user. We formulate the optimization process to find a permutation matrix \( \mathbf{M}_{\text{perm}} \) in \( \mathbf{M}_{\text{perm}} \) that gives the maximum SINR as \( \mathbf{M}_{\text{perm}} = \arg \max_{\mathbf{M}_{\text{perm}}} \max \{ \text{SINR}_1(\mathbf{M}_{\text{perm}}), \ldots, \text{SINR}_K(\mathbf{M}_{\text{perm}}) \} \) where \( \text{SINR}_j(\mathbf{M}_{\text{perm}}) \) is the SINR of link \( j \), given that the permutation matrix \( \mathbf{M}_{\text{perm}} \) is used. To find the \( \mathbf{M}_{\text{perm}} \) without searching from \( K! \) possible orderings, we use the idea of the Myopic Optimization method proposed in [11], which is
proven to be optimal. Using this idea we now only need to search \( \sum_{i=0,i\neq 1}^{K-1} K - i \) possible orderings.

IV. NUMERICAL RESULTS AND DISCUSSION

Monte Carlo simulations have been carried out to assess the performance of the proposed method in a MIMO-OFDM environment. We investigate its performance and compare it with [3], [4], [7], [8] and with an interference free performance. Here, an interference free performance is defined as the performance of any random single link \( i \) transmitted assuming there is no interference from other links at all. The transmitted power of this single user is set to 1. In this case, the received signal of cooperative transmission system is given as \( y_i = r_i^H (H_i t_i x_i + n_i) \) where \( r_i \) and \( t_i \) are the left and right eigenvectors associated with the maximum eigenvalue of \( H_i H_i^H \) using SVD. The comparison of the schemes is performed at \( \text{SER}=10^{-4} \). We use a fixed permutation matrix that orders MSs \( 1, \ldots, K \) as links \( K, \ldots, 1 \), when we are not using APO, for all the simulation results except stated otherwise. For convenience, we will use the notations \( (N_{BS}, N_{MS}, K) \) in all figures to denote the number of transmit antennas per BS, the number of receive antennas per MS and the number of BS in the network, respectively. Perfect CSI is assumed to be available at both ends. Rectangular 64-QAM (\( M=64 \)) modulation is used for all transmissions. The wireless channel model we used is a Rayleigh fading channel. This channel model is commonly used for cellular networks or WLANs, since in most cases there is no line-of-sight path between the transmitter and receiver in these networks. In order to simulate the wireless channel, we set each entry of \( H_j \) channel matrix as an i.i.d. complex Gaussian variable with a zero mean and unit variance. In all simulations, we fix the the Signal-to-Noise-Ratio of each TPH precoded symbol to be \( \text{SNR} = \frac{E[x_i^2]}{2\sigma^2} \), where \( E[x_i^2] \) is normalized to 1 and \( P_{\text{max}} = K \). In all simulation figures, the proposed method refers to the proposed algorithm with TPH, joint iterative transmit-receive weights optimization, SINR equalization (SINRE) under a total power constraint unless stated otherwise and APO.

A. Convergence Study

Figs. 3 and 4 show the convergence characteristics of the proposed method with a total power constraint for (2, 2, 3) and (1, 2, 4) systems. Note that here it does not matter whether total power or per BS constraints are used. This is because, as shown in Fig. 2, the SINR equalization is not an iterative process. We plot the number of iterations versus the average error and scaled output SINR (after SINR equalization), while fixing the SNR at 21 dB. The average error is defined as the average of the maximum entries of the front-channel interference \( \text{BP} \) over all channel realizations. The output SINRs for (1, 2, 4) and (2, 2, 3) systems are scaled up by 4 dB and 0 dB to fit in one figure. The scaling does not matter here since we only want to observe the convergence rate.

From Fig. 4, we could see that the average error decreases rapidly after 5 iterations (for (1, 2, 4)) and 7 iterations (for (2, 2, 3)). That means the front-channel interference, \( \text{BP} \) goes to \( \approx 0 \) as the number of iterations increases. This satisfies the 3\(^{rd} \) constraint (e.g. ZF constraint) in (9). We also observe that MF converges a specific SINR value. This is shown in Fig. 3. MF's SINR reaches a plateau after 8 iterations and not much performance improvement can be obtained by increasing the number of iterations further. In all simulations for SER comparison, we set the maximum iteration number for the proposed scheme to 10.

B. Performance of Overall SER and Complexity

Here, we studied the performance of the overall SER. Overall SER is defined as the average SER of all links. The overall SER performance for the proposed method with or without APO, and [3], [4], [7], [8] for a (2,2,3) system is shown in Fig. 5. The proposed method without APO outperforms the methods in [7], [8] and [3] by 5 dB and 3 dB, respectively, and is only 1 dB away from an interference free performance when \( \text{SER}=10^{-4} \). The large performance
improvement in the proposed scheme with respect to [7], [8] comes from an increase in the degree of freedom and the iteration process used in determining the transmit-receive antenna weights. Fig. 5 also shows the performance of the proposed method. APO moves the SER performance of the proposed method within 0.25 dB from an interference free transmission when SER=10\(^{-4}\). Thus, APO gives about 1 dB gain over the proposed method without APO.

To show that the proposed method performs better than the iterative scheme in [4], we set the iteration number for the scheme in [4] to be 5 for a (2,2,3) system. The performance of [4] is shown in Fig. 5. Here, we could see clearly that [4] is worse than the proposed method with or without APO. It is not possible to get much performance improvement in [4] by raising SNR above 21 dB. We refer to the SNR value, above which there is no further SER decrease, a saturation point. Here, we must stress that the performance of [4] can be further improved by increasing the number of iterations. This is shown in Fig. 5 when we increase the number of iterations to 11. Due to the space limitation, we do not show the overall SER result for a (1,2,4) system. Our simulations show that the SER performance of the proposed method for a (1,2,4) system is also higher than for the schemes in [3], [4], [7], [8].

We calculate the computational complexities for the proposed schemes and comparison schemes in terms of the number of floating point operations (flops) defined in [15]. The complexity of our proposed scheme and [3], [4], [7], [8] are approximately \(O(i \times \max(K,N_{BS},N_{MS})^3)\), \(O(9K^4N_{BS}^2N_{MS})\), and \(O(9K^4N_{BS}^2N_{MS})\), respectively. Here, \(i\) is the number of the iterations. Thus, the computational complexity of the proposed method for a (2,2,3) system with 10 iterations is on average about 75\% less than the complexity of methods in [3], [7], [8] and [4] with 11 iterations. Thus, the proposed method utilizes THP, transmit antennas and receive antennas in a more optimal way with less complexity to create non interfering spatial channels.

V. CONCLUSION

In this paper, we propose a method to design a spectrally efficient cooperative downlink transmission scheme employing precoding and beamforming. THP and iterative transmit-receive weights optimization are used to cancel the interference. A new method to generate transmit-receive antenna weights is proposed. SINR equalization and APO are used to achieve symbol error rate (SER) fairness among different users and further improve the system performance, respectively. For a (2,2,3) cooperative system, the proposed method outperforms the existing schemes by at least 3 dB and is only 0.25 dB away from the interference free performance when SER=10\(^{-4}\). In addition, the proposed method eliminates the dependency between the numbers of transmit and receive antennas. The complexity of the proposed method is also shown to be much lower than for the existing schemes. The complexities of the proposed method for a (2,2,3) system is 75\% less than the complexities of the existing schemes with the same configurations. The proposed method can improve the performance and capacity of co-working WLANs and mobile networks.

REFERENCES