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다중사용자 다중입출력 하향링크 시스템을 위한 최적 수신 결합을 이용한 새로운 빔 형성 기법

(New Beamforming Schemes with Optimum Receive Combining for Multiuser MIMO Downlink Channels)

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요 약

본 논문에서 우리는 다중사용자 다중입출력 하향링크 통신 시스템을 위한 새로운 빔 형성 기법을 제시한다. 최근 block-diagonalization (BD) 알고리즘이 기지국과 각 사용자들이 다중 안테나는 가지는 다중사용자 다중입출력 하향링크를 위해 제안되고 있다. 그러나, BD 알고리즘은 유저당 제공되는 스트림의 개수가 수신기의 개수보다 작은 경우에는 효율적이지 않다. BD 방법이 수신단의 결합을 고려하지 않고 채널 행렬에 기반한 nullspace를 활용하기 때문에, 빔 형성을 위한 자유도는 수신측에서 전부 얻지 못한다. 본 논문에서 우리는 모든 사용자간의 간섭이 0이 되는 zero forcing (ZF) 조건 하에 수신 빔 형성 벡터를 최적화 한다. 우리는 반복적인 과정에 의해 최적 수신 벡터를 찾는 효율적인 알고리즘을 제안한다. 제안된 알고리즘은 수신 결합 벡터를 위해 전방향 정보인 두 phase 값을 요구한다. 또한, 우리는 일반적인 복소 단위 행렬의 분해를 이용하여 단지 한 phase 값만 필요한 또 다른 알고리즘을 제시한다. 시뮬레이션 결과는 에러 확률 관점에서 제안된 빔 형성 기법이 기존 BD 알고리즘보다 성능이 낫고 기지국에서 자유도를 이용함으로써 다이버시티 증가를 획득함을 보여준다.

Abstract

In this paper, we present a new beamforming scheme for a downlink of multiuser multiple-input multipleoutput (MIMO) communication systems. Recently, a block-diagonalization (BD) algorithm has been proposed for the multiuser MIMO downlink where both a base station and each user have multiple antennas. However, the BD algorithm is not efficient when the number of supported streams per user is smaller than that of receive antennas. Since the BD method utilizes the nullspace based on the channel matrix without considering the receive combining, the degree of freedom for beamforming cannot be fully exploited at the transmitter. In this paper, we optimize the receive beamforming vector under a zero forcing (ZF) constraint, where all inter-user interference is driven to zero. We propose an efficient algorithm to find the optimum receive vector by an iterative procedure. The proposed algorithm requires two phase values feedforward information for the receive combining vector. Also, we present another algorithm which needs only one phase value by using a decomposition of the complex general unitary matrix. Simulation results show that the proposed beamforming scheme outperforms the conventional BD algorithm in terms of error probability and obtains the diversity enhancement by utilizing the degree of freedom at the base station.

Keywords : multiple-input multiple-output (MIMO), spatial multiplexing, Beamforming, Multiuser

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I. 서 론

Multiple-input multiple-output (MIMO) channels for future wireless communication systems have attracted considerable attention since the use of multiple antennas at both transmitter and receiver was shown to provide extraordinary high data rates compared to single-input single-output (SISO) systems[1]. More recently, as an interest is being shifted to MIMO Gaussian broadcast channels (BC), the sum capacity of these MIMO Gaussian BC has been extensively studied by several approaches[2~5]. It is well known that dirty paper coding (DPC) introduced in [6] achieves this capacity. However, the encoding process is interference-dependent and largely information-theoretic. Thus, DPC is quite complex and requires techniques which may be incompatible with practical communication systems.

In this paper, we consider a multiuser MIMO downlink where a base station (BS) uses multiple antennas to communicate with several co-channel users. One of the main challenges is to develop a transmission scheme in such multiuser MIMO systems which considers the co-channel interference of other users. In the case where the mobile users are equipped with a single antenna, channel inversion is one of simple beamforming techniques for eliminating all inter-user interference[7~8]. A generalization of channel inversion to the system where each user has multiple antennas, called block diagonalization (BD), was proposed in [9]. The BD algorithm provides a block-diagonal solution, which is less strict than the channel inversion where a complete diagonalization is enforced. The key idea of the BD is to eliminate multiuser interference by placing all the unintended users at nullspace.

In this paper, we investigate a practical beamforming technique which fully utilizes the degree of freedom to improve the performance under a zero forcing (ZF) constraint. We focus on a case where the number of supported streams per user is smaller than the number of receive antennas. This is the case where the performance of the BD degrades. Since in the BD method the nullspace is obtained based on the channel matrix without considering the receive combining, the BD cannot fully utilize the degree of freedom for diversity at the BS. Unlike the BD algorithm, our beamforming method transmits the data along the nullspace of the other users' effective channel matrix while the receive combining process is taken into account. As a result, diversity can be preserved by ensuring that there is no interference on the activated spatial modes.

The problem to optimize the receive combining vector is non-convex since both the intended user channel gain and the interference for other users should be considered simultaneously. In related work[10], an iterative descent algorithm using QR-update is proposed, where the combining gain obtained from cooperation among receive antennas at each user is neglected since a diagonal matrix is employed as a receive beamforming matrix. In contrast, we assume a complex combining receive vector in order to improve the performance by organizing the cooperation among receive antennas. We propose an efficient iterative algorithm using the complex rotation matrix introduced in [11] which achieves the orthogonality between two complex vectors. The optimum solution of the proposed algorithm can be represented by two phase values feedforward information sent from the BS to each user. Also, we present a second algorithm which needs only one phase value by using a decomposition of the general complex unitary matrix. The simulation results show that the proposed algorithm achieves diversity improvement and provides a significant performance gain compared to the BD with comparable complexity. Also, it is confirmed through simulations that two iterations are sufficient to obtain the optimal solution. In addition, we will investigate an effect of the spatial correlation among antennas on the performance in the simulation section.

This paper is organized as follows: In section II, we describe a general system model for the multiuser MIMO downlink and present the correlated channel model. Section III briefly reviews the BD algorithm. In Section IV, we propose new beamforming schemes called Algorithm 1 and 2. Algorithm 2 reduces the feedforward information compared to Algorithm 1. In Section V, the simulation results are presented comparing the proposed method with the BD technique. Finally, the paper is terminated with conclusions in Section VI.

Throughout this paper, normal letters represent scalar quantities, boldface letters indicate vectors and boldface uppercase letters designate matrices. For any complex notation c, we denote the real and imaginary part of c by $Re[c]$ and $Im[c]$, respectively. For any matrix $\boldsymbol{A}$, $[\boldsymbol{A}]_{m,n}$ denotes the (m,n)th entry of $\boldsymbol{A}$. $\boldsymbol{A}^H$ and $\boldsymbol{A}^T$ represent the conjugate transpose and the transpose of $\boldsymbol{A}$, respectively. $\mathrm{Tr}(\boldsymbol{A})$ indicates the trace of $\boldsymbol{A}$. Additionally, $\boldsymbol{I}$ and $\boldsymbol{0}$ denote an identity matrix and a zero matrix, respectively.

Ⅱ. System Model

In this section, we consider a multiuser MIMO downlink system where the BS is transmitting to K independent users simultaneously and generates co-channel interference at all users as shown in Fig 1. In this system, the BS is equipped with M transmit antennas and user j has $n_j \geq 2$ receive antennas, referred to in the following as $\{n_1, \cdots, n_K\} \times M$. L_j spatial streams are supported for

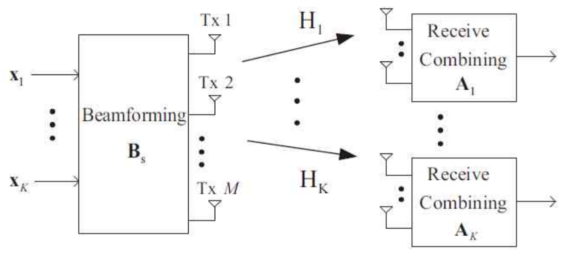

그림 1. 다중사용자 다중입출력 하향 시스템의 구조도.
Fig. 1. Block diagram of multiuser MIMO downlink systems.

the jth user. In the discrete-time complex baseband MIMO case, the channel from the BS to the jth user is modeled as the $n_j \times M$ channel matrix $\boldsymbol{H}_j$, where $[\boldsymbol{H}_j]_{p,q}$ represents the channel gain from antenna q at the BS to antenna p at user j. We assume that $\boldsymbol{H}_j$ is modeled as a block-fading channel, which indicates that the channel is constant during one block. Let $\boldsymbol{x}_j$ represent the L_j transmit data symbol vector for user j. Each user receives a combination of all $L = \sum_{i=1}^{K} L_i$ symbol streams through its own channel. The total number of antennas at all receivers is given as $N_r = \sum_{j=1}^{K} n_j$. We assume that L_j is less than or equal to n_j and M is greater than L.

Denoting $\boldsymbol{B}_j \in \mathbb{C}^{M \times L_j}$ as the transmit beamforming matrix for user j, the received signal at the jth user can be written as

$$\boldsymbol{y}_j = \sum_{i=1}^{K} \boldsymbol{H}_j \boldsymbol{B}_i \boldsymbol{x}_i + \boldsymbol{w}_j = \boldsymbol{H}_j \boldsymbol{B}_j \boldsymbol{x}_j + \sum_{i \neq j} \boldsymbol{H}_j \boldsymbol{B}_i \boldsymbol{x}_i + \boldsymbol{w}_j \quad (1)$$

where $\boldsymbol{y}_j = [y_{j,1}, \cdots, y_{j,n_j}]^T \in \mathbb{C}^{n_j}$ and $\boldsymbol{w}_j = [w_{j,1}, \cdots, w_{j,n_j}]^T \in \mathbb{C}^{n_j}$ are the received signal and noise vectors, respectively. The components $w_{j,i}$ of the noise vector $\boldsymbol{w}_j$ are independently and identically distributed (i.i.d.) with zero mean and variance σ_w^2 for $j = 1, \cdots, K$ and $i = 1, \cdots, n_j$. Note that the term $\sum_{i \neq j} \boldsymbol{H}_j \boldsymbol{B}_i \boldsymbol{x}_i$ indicates the interference from the other users $i \neq j$ to user j.

Applying the receive matrix $\boldsymbol{A}_j \in \mathbb{C}^{L_j \times n_j}$ in equation (1), the filter output $\boldsymbol{z}_j$ for the jth user can be expressed as

$$\boldsymbol{z}_j = \boldsymbol{A}_j \boldsymbol{y}_j = \boldsymbol{A}_j \left(\boldsymbol{H}_j \sum_{i=1}^{K} \boldsymbol{B}_i \boldsymbol{x}_i + \boldsymbol{w}_j \right). \quad (2)$$

For ease of exposition, we define the network channel $\boldsymbol{H}_s$, the receive beamforming matrix $\boldsymbol{A}_s$ and the transmit beamforming matrix $\boldsymbol{B}_s$ as

$$\boldsymbol{H}_s = \begin{bmatrix} \boldsymbol{H}_1 \\ \boldsymbol{H}_2 \\ \vdots \\ \boldsymbol{H}_K \end{bmatrix} \in \mathbb{C}^{N_r \times M},$$

$$\boldsymbol{A}_s = \mathrm{diag}\{\boldsymbol{A}_1, \cdots, \boldsymbol{A}_K\} \in \mathbb{C}^{L \times N_r},$$

$$\boldsymbol{B}_s = [\boldsymbol{B}_1 \boldsymbol{B}_2 \cdots \boldsymbol{B}_K] \in \mathbb{C}^{M \times L},$$

respectively. Then, the corresponding filter output at all the users can be arranged as

$$\boldsymbol{z}_s = \begin{bmatrix} \boldsymbol{z}_1 \\ \boldsymbol{z}_2 \\ \vdots \\ \boldsymbol{z}_K \end{bmatrix} = \boldsymbol{A}_s \boldsymbol{H}_s \boldsymbol{B}_s \boldsymbol{x}_s + \boldsymbol{A}_s \boldsymbol{w}_s = \boldsymbol{H}_e \boldsymbol{B}_s \boldsymbol{x}_s + \boldsymbol{A}_s \boldsymbol{w}_s \quad (3)$$

where

$$\boldsymbol{x}_s = [\boldsymbol{x}_1^T, \boldsymbol{x}_2^T, \cdots, \boldsymbol{x}_K^T]^T \in \mathbb{C}^{N_r \times 1},$$

and

$$\boldsymbol{w}_s = [\boldsymbol{w}_1^T, \boldsymbol{w}_2^T, \cdots, \boldsymbol{w}_K^T]^T \in \mathbb{C}^{L \times 1},$$

and

$$\boldsymbol{H}_e = \boldsymbol{A}_s \boldsymbol{H}_s.$$

In order to satisfy the power constraint, we construct the unnormalized signal $\boldsymbol{x}_s$ and $\boldsymbol{B}_s$ such that

$$E \| \boldsymbol{B}_s \boldsymbol{x}_s \|^2 \le \rho$$

which is a less restrictive power constraint than $E\left(\| \boldsymbol{B}_s \boldsymbol{x}_s \|^2 | \boldsymbol{H}_s \right) \le \rho$ where ρ represents the total transmitted power.

In practice, spatial correlation exists both among transmit antennas and among receive antennas, which indicates that the channel matrix $\boldsymbol{H}_j$ has correlated entries. A spatially correlated MIMO channel can be reasonably described by[12]

$$\boldsymbol{H}_j = \boldsymbol{R}_j^{1/2} \boldsymbol{W}_j \boldsymbol{T}^{1/2} \quad (4)$$

where $\boldsymbol{R}_j \in \mathbb{C}^{n_j \times n_j}$ is the receive spatial correlation matrix among receive antennas for the jth user, $\boldsymbol{T} \in \mathbb{C}^{M \times M}$ represents the transmit spatial correlation matrix among transmit antennas at the BS, and $\boldsymbol{W}_j \in \mathbb{C}^{n_j \times M}$ denotes a random matrix with i.i.d. zero-mean unit-variance circularly symmetric complex Gaussian entries. Although not completely general, this correlation model (4) has been validated through field measurements and regarded as a sufficiently accurate representation of spatially correlated MIMO channels in actual cellular systems.

Ⅲ. Review of Block diagonalization

In this section, we briefly describe the BD algorithm for multiuser MIMO systems presented in [9]. The key idea of the BD algorithm is to find the beamforming (precoding) matrix $\boldsymbol{B}_s$ such that all multiuser interference is zero. To eliminate all the multiuser interference, we impose a constraint as

$$\boldsymbol{H}_i \boldsymbol{B}_j = 0 \text{ for all } i \ne j \text{ and } 1 \le i, j \le K. \quad (5)$$

In order to satisfy the ZF constraint (5), $\boldsymbol{B}_j$ should lie in the nullspace of $\widetilde{\boldsymbol{H}}_j$ where

$$\widetilde{\boldsymbol{H}}_j = [\boldsymbol{H}_1^T \cdots \boldsymbol{H}_{j-1}^T \boldsymbol{H}_{j+1}^T \cdots \boldsymbol{H}_K^T]^T \in \mathbb{C}^{(N_r - n_j) \times M}. \quad (6)$$

Denoting $\widetilde{L}_j$ as $\widetilde{L}_j = \mathrm{rank}(\widetilde{\boldsymbol{H}}_j)$, we define the singular value decomposition (SVD) of $\widetilde{\boldsymbol{H}}_j$ as

$$\widetilde{\boldsymbol{H}}_j = \widetilde{\boldsymbol{U}}_j \widetilde{\boldsymbol{\Lambda}}_j [\widetilde{\boldsymbol{V}}_j^{(1)} \ \widetilde{\boldsymbol{V}}_j^{(0)}]^H$$

where the unitary matrix $\widetilde{\boldsymbol{U}}_j$ contains left singular vectors, the matrix $\widetilde{\boldsymbol{\Lambda}}_j$ consist of ordered singular values of $\widetilde{\boldsymbol{H}}_j$, the matrix $\widetilde{\boldsymbol{V}}_j^{(1)}$ is composed of the first $\widetilde{L}_j$ right singular vectors, and the matrix $\widetilde{\boldsymbol{V}}_j^{(0)}$ holds the last $(M - \widetilde{L}_j)$ right singular vectors. Since $\widetilde{\boldsymbol{V}}_j^{(0)}$ forms an orthogonal basis for the nullspace of $\widetilde{\boldsymbol{H}}_j$, we can construct the beamforming matrix $\boldsymbol{B}_j$ for the jth user using a linear combination of columns of $\widetilde{\boldsymbol{V}}_j^{(0)}$.

Let $\bar{L}_j$ represent the rank of the product $\boldsymbol{H}_j \widetilde{\boldsymbol{V}}_j^{(0)}$. In order for transmission to the jth user to take place under the ZF constraint, $\bar{L}_j$ should be greater than or equal to 1. A sufficient condition for $\bar{L}_j \ge 1$ is that at least one row of $\boldsymbol{H}_j$ is linearly independent of the other rows of $\widetilde{H}_j$. Assuming that all rows of H_j are linearly independent of $\widetilde{H}_j$, this condition would be satisfied.

After applying the nullspace of other users′ channel matrix $\widetilde{V}_j^{(0)}$ to the jth user′s channel H_j, the received signal can be written as

$$y_j = H_{e,j} x_j + w_j$$

where $H_{e,j} = H_j \widetilde{V}_j^{(0)}$. Since the jth user receives its own data stream without interference from other users, any scheme for single user MIMO systems can be applied. We define the SVD of $H_{e,j}$ as

$$H_{e,j} = U_j [\Lambda_j \;\; 0] \left[V_j^{(1)} \;\; V_j^{(0)} \right]^H$$

where Λ_j is a $\overline{L}_j \times \overline{L}_j$ diagonal matrix and $V_j^{(1)}$ represents a matrix with the first $\overline{L}_j$ right singular vectors. Defining the beamforming matrix for the jth user as $B_j = \widetilde{V}_j^{(0)} V_j^{(1)}$, the overall beamforming matrix can be expressed as

$$B_s = \left[\widetilde{V}_1^{(0)} V_1^{(1)} \; \widetilde{V}_2^{(0)} V_2^{(1)} \cdots \widetilde{V}_K^{(0)} V_K^{(1)} \right].$$

Applying the beamforming matrix B_s to the network channel H_s, the effective network channel H_e^{BD} can be obtained as

$$H_e^{BD} = H_s B_s = \begin{bmatrix} U_1 \Lambda_1 & 0 & \cdots & 0 \\ 0 & U_2 \Lambda_2 & \cdots & 0 \\ \vdots & \vdots & \ddots & \vdots \\ 0 & 0 & \cdots & U_K \Lambda_K \end{bmatrix}.$$

In what follows, we analyze the dimension of the nullspace utilized for beamforming considering a receive combining matrix. We define $\hat{H}_j$ as

$$\hat{H}_j = \begin{bmatrix} A_1 H_1 \\ \vdots \\ A_{j-1} H_{j-1} \\ A_{j+1} H_{j+1} \\ \vdots \\ A_K H_K \end{bmatrix} \in \mathbb{C}^{(L-L_j) \times M}. \qquad (7)$$

Then, the dimension of the nullspace of $\hat{H}_j$ is $M-(L-L_j)$. In other words, we can employ the $M-L+L_j$ orthonormal basis vectors for transmit beamforming corresponding to the jth user. In contrast, as observed in (6), the BD algorithm does not consider the number of actual supported streams for each user in the process of obtaining the nullspace of $\widetilde{H}_j$. Since the receive combining process is not included in the BD method, the BD cannot fully utilize the degree of freedom for beamforming. In the BD, the number of orthonormal basis vectors for the jth user is $M-(N_r-n_j)$. Since $L-L_j$ should be less than or equal to N_r-n_j, the BD has a smaller dimension for determining the beamforming matrix. For instance, for the case of $2,2\times 4$ where one stream is assumed to be supported per user, each user with a receive combining matrix is assigned $3(=4-2+1)$ dimensions in determining the beamforming matrix, while the BD algorithm only utilizes $2(=4-4+2)$ dimensions. This added degree of freedom in computing the receive combining matrix allows us to design a system with a better performance, and it will be confirmed in the simulation section that the increased dimension for beamforming improves the diversity order.

Ⅳ. New Beamforming scheme

In this section, we present a procedure for finding the optimal receive matrix A_s which minimizes the power used for eliminating all multiuser interference. We propose an efficient iterative algorithm based on the result in [10] using the rotation matrices in [11]. For simplicity, we consider the case where each user is equipped with two receive antennas $(n_j=2)$ and assume that the channel matrix H_s is known perfectly at the BS. Note that the proposed algorithm can be generalized to the case with $n_j > 2$ receive antennas. Also, we assume that one stream is supported for each user $(L_j=1)$. Since the receive matrix A_j reduces to a row vector in this case, we will use a_j to denote the receive combining vector from now on.

In order to satisfy the ZF constraint, we can select a pseudo-inverse of the effective channel matrix considering the receive combining vector at all users. Thus, the overall transmit beamforming matrix $\boldsymbol{B}_s$ for (3) can be obtained as

$$\boldsymbol{B}_s = \boldsymbol{H}_e^H\left(\boldsymbol{H}_e\boldsymbol{H}_e^H\right)^{-1}.$$

Applying the transmit beamforming matrix $\boldsymbol{B}_s$ in equation (3), the filter output vector $\boldsymbol{z}_s$ can be rewritten as

$$\boldsymbol{z}_s = \frac{1}{\sqrt{\gamma}}\boldsymbol{x}_s + \boldsymbol{A}_s\boldsymbol{w}_s$$

where γ is referred to as the power loss factor. From the power constraint, γ is computed as

$$\gamma = \frac{1}{\rho}\mathrm{Tr}\left(\left(\boldsymbol{H}_e\boldsymbol{H}_e^H\right)^{-1}\right).$$

Our objective is to obtain the optimum receive vector for minimization of the power loss factor γ, which is significantly large when $\boldsymbol{H}_s$ is poorly conditioned. This problem is formulated as

$$\left(\boldsymbol{a},\cdots,\boldsymbol{a}_K\right)_{opt} = \arg\min_{(\boldsymbol{a},\cdots,\boldsymbol{a}_K)}\mathrm{Tr}\left(\left(\boldsymbol{H}_e\boldsymbol{H}_e^H\right)^{-1}\right). \tag{8}$$

As observed in (8), this joint optimization problem is a complicated non-convex problem. In general, this can be solved by a gradient-descent algorithm which is computationally expensive.

In what follows, we present an iterative algorithm, which will be referred to as Algorithm 1, to solve the optimization problem. First, we define $\boldsymbol{P}_j$ as the permutation matrix which satisfies the following condition

$$\boldsymbol{P}_j\boldsymbol{H}_{e,perm} = \boldsymbol{H}_e$$

where the permuted effective channel matrix $\boldsymbol{H}_{e,perm}$ is given by

$$\boldsymbol{H}_{e,perm} = \begin{bmatrix}\hat{\boldsymbol{H}}_j \\ \boldsymbol{a}_j\boldsymbol{H}_j\end{bmatrix}.$$

Note that the permutation of the effective channel matrix does not change the optimization metric as

$$\mathrm{Tr}\left(\left(\boldsymbol{H}_e\boldsymbol{H}_e^H\right)^{-1}\right) = \mathrm{Tr}\left(\left(\boldsymbol{H}_{e,perm}\boldsymbol{H}_{e,perm}^H\right)^{-1}\right). \tag{9}$$

Next, we obtain the optimum solution of (8) by utilizing the result in [10]. Due to non-convexity of the optimum problem (8), it is difficult to solve directly the overall receive beamforming matrix $\boldsymbol{A}_s$. To overcome this difficulty, we calculate iteratively the optimum combining vector $\boldsymbol{a}_j$ for the jth user by fixing the receive combining vectors of the other users. Instead of performing joint optimization of (8), a solution for $\boldsymbol{a}_j$ with other $\boldsymbol{a}_j$'s for $i \neq j$ fixed can be computed as (see Theorem 1 in [10] for details)

$$\begin{aligned}&\arg\min_{\boldsymbol{a}_j}\mathrm{Tr}\left(\left(\boldsymbol{H}_{e,perm}\boldsymbol{H}_{e,perm}^H\right)^{-1}\right)\\ &= \boldsymbol{e}_1\left(\left(\boldsymbol{I}+\boldsymbol{H}_j\boldsymbol{Q}\boldsymbol{R}^{-1}\boldsymbol{R}^{-H}\boldsymbol{Q}^H\boldsymbol{H}_j^H\right)^{-1}\left(\boldsymbol{H}_j\boldsymbol{H}_j^H - \boldsymbol{H}_j\boldsymbol{Q}\boldsymbol{Q}^H\boldsymbol{H}_j^H\right)\right)\end{aligned} \tag{10}$$

where the QR factorization of $\hat{\boldsymbol{H}}_j^H$ is defined as $\hat{\boldsymbol{H}}_j^H = \boldsymbol{Q}\boldsymbol{R}$, and $\boldsymbol{e}_1(\boldsymbol{\Gamma})$ refers to the eigenvector corresponding to the maximum eigenvalue of $\boldsymbol{\Gamma}$. After obtaining $\boldsymbol{a}_j$ for $j = 1,\cdots,K$ from (10) and repeating this procedure iteratively, we can determine the optimum vector set which minimizes the power loss factor.

To summarize, the proposed algorithm 1 can be described as follows:

1) Initialize $\boldsymbol{a}_j$ as $\boldsymbol{a}_j = [1\ \ 0]$ for $j = 1,\cdots,K.$.

2) Set $iter\_count = 0$ for outer loop.

3) Set $j=0$ for inner loop.

4) Obtain $\hat{\boldsymbol{H}}_j$ and $\boldsymbol{H}_j$ from $\boldsymbol{H}_e$.

5) Compute $\boldsymbol{Q}$ and $\boldsymbol{R}$ by performing the QR factorization of $\hat{\boldsymbol{H}}_j^H$.

6) Determine $\boldsymbol{a}_j$ as the unit-norm eigenvector corresponding to the maximum eigenvalue of the matrix

$$\left(\boldsymbol{I}+\boldsymbol{H}_j\boldsymbol{Q}\boldsymbol{R}^{-1}\boldsymbol{R}^{-H}\boldsymbol{Q}^H\boldsymbol{H}_j^H\right)^{-1}\left(\boldsymbol{H}_j\boldsymbol{H}_j^H - \boldsymbol{H}_j\boldsymbol{Q}\boldsymbol{Q}^H\boldsymbol{H}_j^H\right).$$

7) Go back to step 4 with $j \leftarrow j+1$ unless $j > K$.

8) Go back to step 3 with $iter\_count \leftarrow iter\_count + 1$ if $iter\_count < max\_iteration$.

As stated in [10], this iterative algorithm may converge to a local optimum due to nature of a non-convex problem. Complexity can be reduced by employing simpler QR-updates instead of the QR-factorization at step 5 in the above algorithm. Note that unlike[10], our algorithm works directly on complex-valued matrices.

Once the receive beamforming vector $\boldsymbol{a}_j$ is computed, the given $\boldsymbol{a}_j$ can be represented using a general complex unitary matrix expressed by[11]

$$\boldsymbol{C}(\theta_1,\theta_2)=\begin{bmatrix}\cos(\theta_1,\theta_2) & \sin(\theta_1,\theta_2)\\ -\sin^*(\theta_1,\theta_2) & \cos^*(\theta_1,\theta_2)\end{bmatrix} \tag{11}$$

where we define $\cos(\alpha,\beta)$ and $\sin(\alpha,\beta)$ as

$$\begin{aligned}\cos(\alpha,\beta)&=\cos(\alpha)\cos(\beta)+i\sin(\alpha)\sin(\beta),\\ \sin(\alpha,\beta)&=\sin(\alpha)\cos(\beta)+i\cos(\alpha)\sin(\beta),\end{aligned}$$

and $i=\sqrt{-1}$. In [11], the above complex rotation matrix is used for the complex vector orthogonalization which provides a constructive basis for beamforming in single user MIMO channels. In our formulation, we employ the row vectors of $\boldsymbol{C}(\theta_1,\theta_2)$ as a general expression for complex-valued unit-norm vectors. In this paper, we use the first row of this matrix for receive beamforming, although the second row of the matrix can be employed as well. Then, we define the receive beamforming vector $\boldsymbol{a}_j$ as

$$\boldsymbol{a}_j=[\cos(\theta_{j,1},\theta_{j,2})\quad \sin(\theta_{j,1},\theta_{j,2})]. \tag{12}$$

To compute two phase values $\theta_{j,1}$ and $\theta_{j,2}$ which identify $\boldsymbol{a}_j$, we utilize the relation between the components of the above equation. Denoting a_1 and a_2 as the first and the second element of $\boldsymbol{a}_j$, respectively, we can easily compute $\theta_{j,1}$ and $\theta_{j,2}$ as

$$\begin{aligned}\theta_{j,1}&=\arctan(Re[a_2]/Re[a_1]),\\ \theta_{j,2}&=\arctan(Im[a_2]/Re[a_1])).\end{aligned} \tag{13}$$

Thus, after the maximum number of iterations is reached in Algorithm 1, $\theta_{j,1}$ and $\theta_{j,2}$ can be obtained from (13). Then, these values are transmitted to the j th user to identify $\boldsymbol{a}_j$ using (12).

Now we will explain why a performance gain is expected with the proposed $\boldsymbol{a}_j$ in (12) in comparison to the solution in [10]. Unlike our scheme, the receive combining filter in [10] is extended to a matrix form in the case with two receive antennas, which is given as

$$\boldsymbol{A}_j=\begin{bmatrix}e^{j\theta_{j,1}} & 0\\ 0 & e^{j\theta_{j,2}}\end{bmatrix}. \tag{14}$$

By using this combining matrix, two data streams are allocated for each user. However, since the modulation is restricted to one-dimensional PAM for each symbol in [10], the spectral efficiency is the same as our scheme using two-dimensional QAM.

Applying the above $\boldsymbol{A}_j$ to equation (2) and taking the real part of the result, the filter output $\boldsymbol{z}_j$ for the scheme in [10] is obtained as

$$\boldsymbol{z}_j=Re[\boldsymbol{A}_j\boldsymbol{y}_j]=\begin{bmatrix}Re[e^{j\theta_{j,1}}y_{j,1}]\\ Re[e^{j\theta_{j,2}}y_{j,2}]\end{bmatrix}$$

where $\boldsymbol{z}_j=[z_{j,1}\ z_{j,2}]^T$ and $\boldsymbol{y}_j=[y_{j,1}\ \ y_{j,2}]^T$. By utilizing only the real part of complex-valued symbols for transmission, the scheme in [10] eliminates the inter-user interference which is confined to the imaginary part of the filter output. However, as off-diagonal elements in (14) are zero, this receive combining matrix cannot obtain a performance gain from cooperation among receive antennas. In the simulation section, we will show that the proposed scheme outperforms the scheme in [10] by allowing non-zero off-diagonal elements in $\boldsymbol{A}_j$. By adopting a more general form of the receive combining matrix compared to [10], the proposed scheme exhibits a 2-3dB gain over the method in [10] with the same computational complexity, as will be demonstrated in the simulation section.

In Algorithm 1, two phase values should be transmitted for each user. To reduce this overhead, we present a simpler algorithm which requires only one phase value at the receiver, which will be refered to as Algorithm 2. The simplification is carried out based on the fact that the complex rotation matrix in (11) can be decomposed into two matrices, i.e.,

$$C(\theta_1,\theta_2)=\begin{bmatrix}\cos(\theta_1) & \sin(\theta_1)\\ -\sin(\theta_1) & \cos(\theta_1)\end{bmatrix}\begin{bmatrix}\cos(\theta_2) & i\sin(\theta_2)\\ i\sin(\theta_2) & \cos(\theta_2)\end{bmatrix}.$$

By choosing one of two matrices above for the receive combining, we can reduce the feedforward information to only one phase value. Since any choice between two matrices does not affect the overall performance, we use the first row of the first matrix as the receive combining vector. Then, the receive combining vector for the jth user for Algorithm 2 can be expressed as

$$\boldsymbol{a}_j=[\cos(\theta_j)\ \ \sin(\theta_j)]. \tag{15}$$

Compared to the complex solution (12), $\boldsymbol{a}_j$ in (15) has only real-valued elements.

Now we will discuss how to determine the real-valued optimum combining vector for Algorithm 2. In [10], the result (10) was derived using the following well-known result

$$\mathrm{argmin}_{\boldsymbol{a}_j}\left\{\frac{\boldsymbol{a}_j\boldsymbol{M}_1\boldsymbol{a}_j^H}{\boldsymbol{a}_j\boldsymbol{M}_2\boldsymbol{a}_j^H}\mid \|\boldsymbol{a}_j\|=1\right\}=\boldsymbol{e}_1\left(\boldsymbol{M}_1^{-1}\boldsymbol{M}_2\right).$$

In the real-valued representation, the above equation is equivalent to

$$\mathrm{argmin}_{\bar{\boldsymbol{a}}_j}\left\{\frac{\bar{\boldsymbol{a}}_j\overline{\boldsymbol{M}}_1\bar{\boldsymbol{a}}_j^H}{\bar{\boldsymbol{a}}_j\overline{\boldsymbol{M}}_2\bar{\boldsymbol{a}}_j^H}\mid \|\bar{\boldsymbol{a}}_j\|=1\right\}=\boldsymbol{e}_1\left(\overline{\boldsymbol{M}}_1^{-1}\overline{\boldsymbol{M}}_2\right) \tag{16}$$

where

$$\overline{\boldsymbol{a}}_j=[Re[\boldsymbol{a}_j]\ \ Im[\boldsymbol{a}_j]],\ \overline{\boldsymbol{M}}_j=\begin{bmatrix}Re[\boldsymbol{M}_i] & -Im[\boldsymbol{M}_i]\\ Im[\boldsymbol{M}_i] & Re[\boldsymbol{M}_i]\end{bmatrix}$$

for $i=1,2$.

Since $\boldsymbol{a}_j$ is a real-valued vector ($Im[\boldsymbol{a}_j]=0$) for Algorithm 2, the solution of (16) reduces to

$$\boldsymbol{e}_1\left(Re[\boldsymbol{M}_1]^{-1}Re[\boldsymbol{M}_2]\right).$$

Employing the above result, we can obtain Algorithm 2 by making a slight change in step 6 of Algorithm 1 as

6) Determine $\boldsymbol{a}_j$ as the unit-norm eigenvector corresponding to the maximum eigenvalue of the matrix

$$Re\left[\left(I+H_jQR^{-1}R^{-H}Q^HH_j^H\right)\right]^{-1}Re\left[\left(H_jH_j^H-H_jQQ^HH_j^H\right)\right].$$

Similarly, after $\boldsymbol{a}_j$ is determined from the iterative algorithm, θ_j for the receive combining vector $\boldsymbol{a}_j$ is easily computed from (15). In the following simulation section, we will see that the performance gap between Algorithm 1 and 2 is small.

In our proposed algorithms, for decoding the signal at the receiver, the transmitter needs to send information about the power loss factor γ and phase values to each user. Note that this amount of information is inevitable in any beamforming scheme for multiuser systems. Since the optimization and the training can be done once per transmission block, this may incur no additional penalty in this regard.

Ⅴ. Simulation Results

In this section, we present the simulation results for the proposed beamforming schemes to compare the performance with conventional algorithms. Throughout the simulation, the number of iterations is set to 5 for Algorithm 1 and 2.

Figures 2 and 3 show the simulation results for $2,2\times 4$ systems in terms of bit error rate (BER) with respect to signal-to-noise ratio (SNR) in dB at

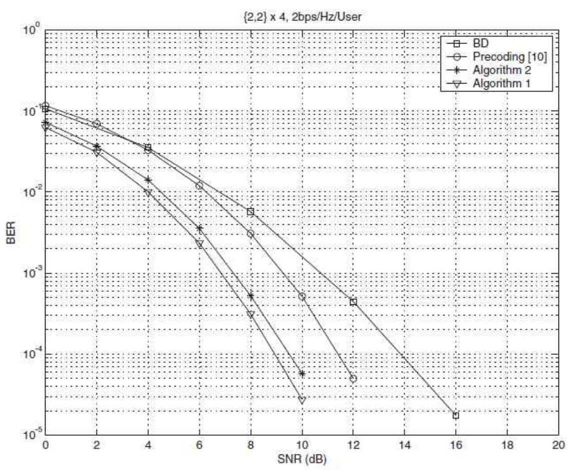

그림 2. 2 bps/Hz/User에서의 여러 가지 기법의 비트 오율 성능

Fig. 2. Bit error probability of different methods at 2bps/Hz/user.

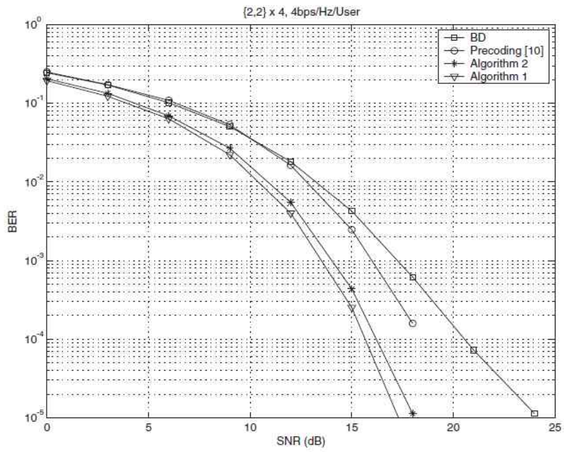

그림 3. 2 bps/Hz/User에서의 여러 가지 기법의 비트 오율 성능

Fig. 3. Bit error probability of different methods at 4bps/Hz/user.

2bps/Hz/user and 4bps/Hz/user, respectively. The proposed beamforming scheme using Algorithm 1 provides about a 5 dB power gain at a BER of 10^{-4} over the BD algorithm.

Note that the diversity improvement is achieved by transmitting the data along the nullspace of the other users′ effective channel matrix while optimizing the receive combining vector. That is, the proposed scheme can fully utilize the remaining nullspace dimension at the BS for the diversity improvement. As can be shown in these plots, the performance loss of Algorithm 2 is less than 0.5dB compared to Algorithm 1. In comparison to the scheme proposed in [10], we can see that the utilization of cooperation among receive antennas provides a performance gain up to 3dB. It should be emphasized that with reduced feedforward overhead compared to [10], the proposed algorithm 2 achieves a 2dB gain over the scheme in [10].

Figures 4 and 5 present the simulation results for $2,2,2,2\times 8$ systems at 2bps/Hz/user and 4bps/Hz/user, respectively. As observed in these plots, the proposed Algorithm 1 outperforms the BD algorithm by about 8 dB at a BER of 10^{-4}. We can see that the gain of the proposed scheme over the BD method increases compared to the case of $2,2\times 4$. This improvement results from the increased degree of freedom in orthogonal basis utilized for beamforming. In the case of $2,2,2,2\times 8$, we can use $5(=8-3)$ dimension for beamforming, while $3(=4-1)$ dimension is utilized in the case of $2,2\times 4$.

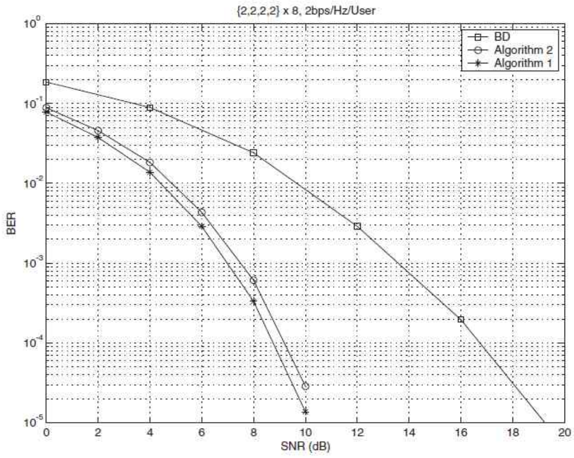

그림 4. 2 bps/Hz/User에서의 여러 가지 기법의 비트 오율 성능

Fig. 4. Bit error probability of different methods at 2bps/Hz/user.

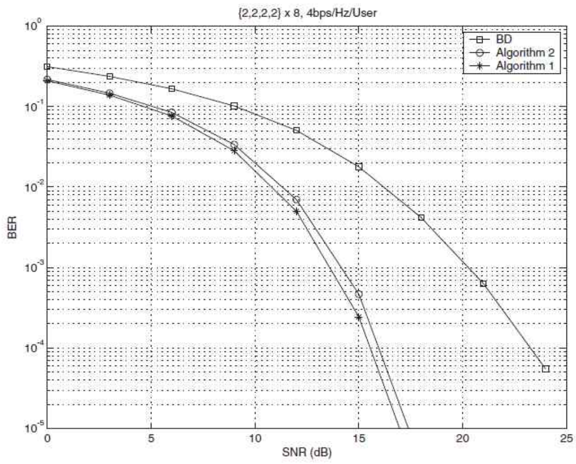

그림 5. 4 bps/Hz/User에서의 여러 가지 기법의 비트 오율 성능

Fig. 5. Bit error probability of different methods at 4bps/Hz/user.

Figure 6 depicts the performance of Algorithm 1 in terms of iteration numbers. It is clear from the plot that only two iterations are sufficient to arrive at near optimal solutions.

Next, we present the evaluation of the BER performance of the proposed beamforming scheme over spatially correlated MIMO Rayleigh-fading channels. For generating the spatial correlation, we

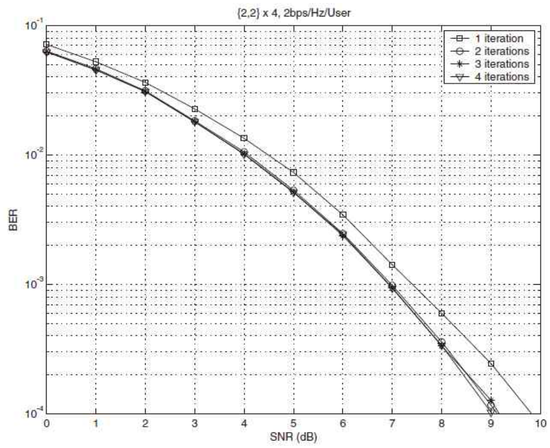

그림 6. 반복횟수에 따른 성능 비교
Fig. 6. Comparison of performance in terms of the number of iterations.

employ the correlation matrix for (4) with entries from [13]

$$[\boldsymbol{T}]_{m,n}(\text{or}\,[\boldsymbol{R}_k]_{m,n}) = \int_{-180}^{180} \frac{d\phi}{\sqrt{2\pi\delta_k^2}} e^{j\pi(m-n)\sin(\pi\phi/180)-(\phi-\theta_k)^2/2\delta_k^2}$$

where m, n denote the indices of antennas and $\theta_{k,}\ \delta_k$ represent the mean angle and the angular spread, respectively.

Figure 7 shows the performance comparison between the proposed beamforming schemes and the BD method for $2,2\times4$, assuming the receiver correlation only. In this plot, the solid line indicates the correlated case, and the dashed line represents the uncorrelated case. The adopted strong receiver correlation matrix is obtained with $\theta_k = 40^\circ$ and $\delta_k = 3^\circ$ as

$$\boldsymbol{R} = \begin{bmatrix} 1 & 0.4278 - i0.8951 \\ 0.4278 + i0.8951 & 1 \end{bmatrix}$$

Figure 7 shows that for the correlated case the proposed beamforming schemes outperform the BD method by 7 dB at a BER of 10^{-4}. In the presence of the spatial correlation at the receiver, the performance gap between the BD and the proposed scheme increases compared to the uncorrelated case. This results from diversity degradation due to the spatial correlation.

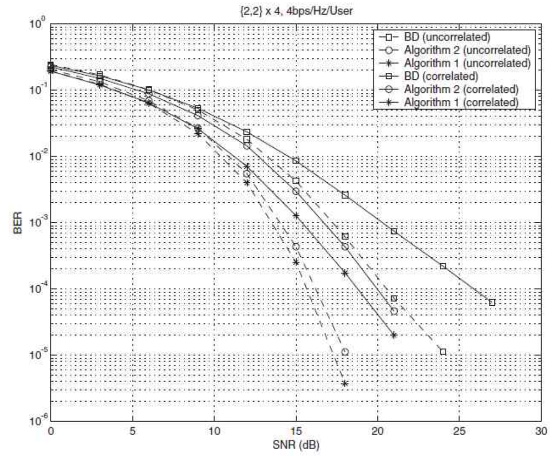

그림 7. 수신 측에 spatial correlation 존재 하에 비트 오율 성능
Fig. 7. Bit error probability in the presence of spatial correlation at the receiver.

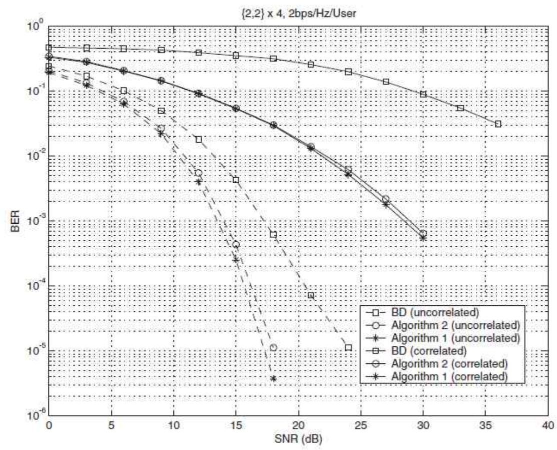

그림 8. 송신 측에 spatial correlation 존재 하에 비트 오율 성능
Fig. 8. Bit error probability in the presence of spatial correlation at the transmitter.

Fig. 8 presents the performance comparison between the proposed beamforming schemes and the BD method for $\{2,2\}\times4$ for the case of the transmitter correlation. The adopted strong transmitter correlation matrix is obtained with $\theta_k = 20^\circ$ and $\delta_k = 3^\circ$ as

$$\boldsymbol{T} = \begin{bmatrix} 1 & 0.4718 - i0.8683 & -0.5189 - i0.7999 & -0.8955 + i0.0704 \\ 0.4718 + i0.8683 & 1 & 0.4718 - i0.8683 & -0.5189 - i0.7999 \\ -0.5189 + i0.7999 & 0.4718 + i0.8683 & 1 & 0.4718 - i0.8683 \\ -0.8955 - i0.0704 & -0.5189 + i0.7999 & 0.4718 + i0.8683 & 1 \end{bmatrix}$$

As can be seen in Fig. 8, in a correlated channel at the BS the performance degradation of the proposed beamforming scheme is much smaller compared to the BD. This indicates that the proposed beamforming scheme is robust to spatially correlated channels.

Now we explain why the performance of beamforming schemes is more sensitive to the spatial correlation at the BS than at the mobile user. In the presence of the spatial correlation at the transmitter, the channel matrix $\boldsymbol{H}_s$ from (4) can be written as $\boldsymbol{H}_s = \boldsymbol{W}\boldsymbol{T}^{1/2}$ where $\boldsymbol{W} \in \mathbb{C}^{N_r \times M}$ is a spatially white matrix. Thus, the rank of $\boldsymbol{H}_s$ is obtained as

$$\begin{aligned}\mathrm{rank}(\boldsymbol{H}_s) &= \mathrm{rank}(\boldsymbol{W}\boldsymbol{T}^{1/2}) \\ &= \min(\mathrm{rank}(\boldsymbol{W}), \mathrm{rank}(\boldsymbol{T}^{1/2})) \\ &= \min(\min(N_r, M), \mathrm{rank}(\boldsymbol{T}^{1/2})).\end{aligned}$$

Since we have $1 \le \mathrm{rank}(\boldsymbol{T}^{1/2}) \le M$ due to the correlation, the $\mathrm{rank}(\boldsymbol{H}_s)$ is in the range of $1 \le \mathrm{rank}(\boldsymbol{H}_s) \le \min(N_r, M)$. We can see that the rank of the channel matrix approaches one as a MIMO channel is highly correlated at the BS.

In contrast, the channel matrix $\boldsymbol{H}_s$ for the case of the receiver correlation can be given as $\boldsymbol{H}_s = \boldsymbol{R}^{1/2}\boldsymbol{W}$ where $\boldsymbol{R}^{1/2} = \mathrm{diag}\{\boldsymbol{R}_1^{1/2}, \cdots, \boldsymbol{R}_K^{1/2}\} \in \mathbb{C}^{N_r \times N_r}$. Similarly, the rank of $\boldsymbol{H}_s$ is computed as

$$\mathrm{rank}(\boldsymbol{H}_s) = \min(\mathrm{rank}(\boldsymbol{R}^{1/2}), \min(N_r, M)).$$

As we assume that no correlation exists among users, $\boldsymbol{R}^{1/2}$ has a block diagonal matrix form. As a result, the minimum of $\mathrm{rank}(\boldsymbol{R}^{1/2})$ is K. Thus the $\mathrm{rank}(\boldsymbol{H}_s)$ is in the range of $\min(N_r, M, K) \le \mathrm{rank}(\boldsymbol{H}_s) \le \min(N_r, M)$. Unlike the case of the correlation at the BS, the rank of $\boldsymbol{H}_s$ is guaranteed to be $\min(N_r, M, K)$ for the fully correlated case at the receiver. For example, in our simulation configuration of $\{2,2\} \times 4$, the system with the full receiver correlation has rank 2. Thus the diversity degradation due to the receiver correlation is smaller than the correlation at the BS.

Ⅵ. Conclusion

In this paper, we have proposed a new beamforming scheme with the optimum receive combining vector for multiuser MIMO downlink systems where each user has more than one antenna. Considering the receive combining process, the proposed scheme can fully utilize the remaining degrees of freedom for diversity. Also, unlike conventional schemes, we can achieve a performance improvement from the cooperation among receive antennas. An efficient iterative algorithm for computation of the receive combining vector has been proposed. Also, we present a simplified algorithm which requires only one phase value per user at the expense of a small performance loss. The simulation results confirm that the proposed schemes outperform the conventional BD method.

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