폐루프 다중입출력 시스템을 위한 효율적인 그룹별 공간 다중화 기법 설계

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A New Efficient Group-wise Spatial Multiplexing Design for Closed-Loop MIMO Systems

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요 약

본 논문에서는 폐루프 다중입출력 무선통신 환경을 위한 새로운 공간 다중화 기법을 소개한다. 기존에 제안되었던 직교 공간 다중화 (OSM; orthogonalized spatial multiplexing) 방식을 확장하여, 우리는 임의의 수의 데이터 스트림을 동시에 전송하기 위한 새로운 방식을 제안한다. 이를 위하여 우리는 데이터 스트림을 두 개 이상의 그룹으로 나누고 수신기에서 블록 대각화 과정을 수행한다. 제안하는 기법은 적은 피드백 정보량을 가지며 심볼 단위의 ML (maximum likelihood) 검출을 통해 복잡도를 최소화한다. 실험 결과를 통해 제안하는 기법은 기존의 설계 기법들에 비하여 비트에러율 관점에서 큰 성능 이득을 제공함을 확인한다. 또한 추가적인 피드백을 통해 수신 그룹의 선택을 최적화함으로써 성능을 더욱 향상시킬 수 있음을 관찰한다.

Key Words : Multiple-input Multiple-output(MIMO), Spatial Multiplexing, Orthogonalized Spatial Multiplexing (OSM), Group Detection (GD)

ABSTRACT

This paper introduces a new efficient design scheme for spatial multiplexing (SM) systems over closed loop multiple-input multiple-output (MIMO) wireless channels. Extending the orthogonalized spatial multiplexing (OSM) scheme which was developed recently for transmitting two data streams, we propose a new SM scheme where a larger number of data streams can be supported. To achieve this goal, we partition the data streams into several subblocks and execute the block-diagonalization process at the receiver. The proposed scheme still guarantees single-symbol maximum likelihood (ML) detection with small feedback information. Simulation results verify that the proposed scheme achieves a huge performance gain at a bit error rate (BER) of 10^{-4} over conventional closed-loop schemes based on minimum mean-square error (MSE) or bit error rate (BER) criterion. We also show that an additional 2.5dB gain can be obtained by optimizing the group selection with extra feedback information.

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Ⅰ. Introduction

Ever since the information theoretic analysis identified the great potential of multiple-input multiple-output (MIMO) channels on the system capacity[1,2], there has been intensive study in recent years on the design of practical systems which achieve this huge capacity. One simple approach for taking advantage of the capacity of the MIMO wireless channel is spatial multiplexing (SM) such as Bell Labs layered space-time (BLAST) architecture[3,4]. Based on the perfect channel knowledge at the receiver, a number of receiver structures including maximum likelihood (ML) detector, linear equalizer, and decision- feedback equalizer can be applied to handle the interference and form multiple data streams[5].

When channel state information (CSI) is fully known at the transmitter, it is possible to jointly optimize the transmitter and the receiver under the assumption of linear transformations. There are several designs which aim for the highest performance gain with a variety of meaningful constraints, such as minimum mean-square error (MMSE)[6,7], maximum signal-to-noise ratio (SNR)[8], and minimum bit error rate (BER)[7]. Especially the authors in [9] proposed a unified framework to put all these criteria together by exploiting convex optimization tools. Although these system designs provide good results on the BER performance, their criteria may not be optimum.

Recently, orthogonalized spatial multiplexing (OSM) has been developed in [10] for closed loop MIMO systems, which allows simple maximum likelihood (ML) decoding at the receiver with small feedback information. This scheme achieves orthogonality of the channel by applying a simple transformation at the transmitter and as a result, the subchannels exhibit the same quality. More recently, a new precoder for the OSM has been proposed which can improve the performance substantially[11].

One constraint of the OSM scheme is that the number of supported data streams is limited to two. In this paper, we propose an extended structure of the OSM scheme which can support a larger number of data streams. To achieve this goal, we partition the data streams into several subblocks by applying the block-diagonalization process at the receiver. Such a procedure can be utilized by a low complexity detection technique called group detection (GD)[12,13]. However, they suffer from a considerable performance loss because the noise components are not taken into account. In this paper, we introduce a better GD method which shows a performance gain over the approach used in [13]. The proposed scheme still guarantees simple ML detection and needs small information for feedback. In addition, we consider a group selection issue which can provide an additional performance gain. The simulation results show that the proposed scheme achieves a huge performance gain compared to the conventional power allocation schemes based on the MMSE[6] or the minimum bit error rate (BER)[9] criteria, and that an additional 2.5dB gain can be obtained by optimizing the group selection.

Throughout this paper, the following notation is used. Normal letters represent scalar quantities, bold face letters indicate vectors, and boldface uppercase letters designate matrices. We denote the real and imaginary part of c by $Re[c]$ and $Im[c]$, respectively. The superscripts $(.)^T$ and $(.)^H$ denote transpose and Hermitian operation, respectively. $\boldsymbol{I}_d$ indicates an identity matrix of size d and $\boldsymbol{a} \cdot \boldsymbol{b}$ denotes the inner (dot) product between vectors $\boldsymbol{a}$ and $\boldsymbol{b}$.

The paper is organized as follows: Section II introduces the system model and briefly reviews the OSM scheme and an enhanced version of the OSM scheme. In Section III, we propose an extended structure of the OSM and illustrate the efficacy of the proposed scheme. Also a group selection issue is considered to achieve a performance gain. Section IV provides several simulation results for the proposed schemes. Finally, Section V gives the conclusions of this paper.

Ⅱ. System Description

This section provides a description of the MIMO communication system model with M_t transmit and M_r receive antennas. The cooperation between the transmitter and the receiver required in closed-loop systems is illustrated in Figure 1. The receiver estimates the forward link channel and then feeds back its corresponding information to the transmitter. Using this feedback information, the transmitter applies a precoding operation to the transmitted signals.

Denoting B as the number of data streams ($B \le min(M_t, M_r)$), the sampled baseband signal model for flat-fading channels is given by

$$y = HPx + n$$

where $x = [x_1 \cdots x_B]^T \in C^{B \times 1}$ is the transmitted symbol vector precoded by the linear precoder $P \in C^{M_t \times B}$, and $y \in C^{M_r \times 1}$ denotes the received signal vector. Here the channel response matrix H is modeled as

$$H = \begin{bmatrix} h_{11} & \cdots & h_{1M_t} \\ \vdots & \ddots & \vdots \\ h_{M_r 1} & \cdots & h_{M_r M_t} \end{bmatrix} \quad (1)$$

where h_{ji} represents the channel response between the i-th transmit and the j-th receive antenna. We assume that the transmitted symbols are uncorrelated and have unit-energy, i.e., $E[xx^H] = I_B$, and the noise vector $n \sim N(0, \sigma_n^2 I_{M_r})$ is zero-mean circularly symmetric complex Gaussian. Throughout this paper, we focus on systems supporting full spatial data streams ($B \le min(M_t, M_r)$), which achieves the highest throughput in all range of SNR.

In the following, we review the OSM scheme

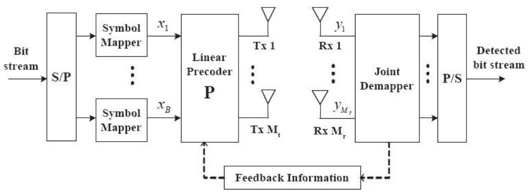

그림 1. M_t개의 송신안테나와 M_r개의 수신안테나를 가진 폐루프 MIMO 시스템의 구조도
Fig. 1. Block diagram of a closed-loop MIMO system with M_t transmit and M_r receive antennas

introduced in [10], then summarize an enhanced version of the OSM proposed in [11]. In the OSM, M_t transmit antennas are considered with two data streams. For illustration purposes, we focus on the $M_t = 2$ case. The OSM scheme encodes two transmitted symbols as

$$F(x, \theta_o) = \begin{bmatrix} 1 & 0 \\ 0 & \exp(j\theta_o) \end{bmatrix} s(x)$$

where θ_o is the rotation phase angle and $s(x)$ is defined by

$$s(x) = \begin{bmatrix} Re[x_1] + jRe[x_2] \\ Im[x_1] + jIm[x_2] \end{bmatrix}.$$

Applying the above transformation, the system model is changed to

$$y = HF(x, \theta_o) + n = H_{\theta_o} s(x) + n \quad (2)$$

where H_{θ_o} accounts for the effective channel matrix denoted as

$$H_{\theta_o} = H \begin{bmatrix} 1 & 0 \\ 0 & \exp(j\theta_o) \end{bmatrix}.$$

Denoting the real-valued representation of H_{θ_o} as $H_{r,\theta_o} = [h_{r,1}^{\theta_o}\ h_{r,2}^{\theta_o}\ h_{r,3}^{\theta_o}\ h_{r,4}^{\theta_o}]$, we can easily check that $h_{r,1}^{\theta_o}$ and $h_{r,2}^{\theta_o}$ are orthogonal to $h_{r,3}^{\theta_o}$ and $h_{r,4}^{\theta_o}$, respectively.

The rotation angle θ_o is decided as to attain the orthogonality between $h_{r,1}^{\theta_o}$ and $h_{r,3}^{\theta_o}$ (or $h_{r,2}^{\theta_o}$ and $h_{r,4}^{\theta_o}$), given as[10]

$$\theta_o = \tan^{-1}\left(\frac{h_{r,1}^{\theta_o} \cdot h_{r,4}^{\theta_o}}{h_{r,1}^{\theta_o} \cdot h_{r,2}^{\theta_o}} \right).$$

Note that this rotation achieves the full orthogonality between two subspaces spanned by the transmitted signals, which is the key concept of the OSM. Exploiting this feature, the optimal ML detection can be carried out with complexity $O(M_c)$, where M_c indicates the size of the signal constellation. Besides, two transmit symbols x_1 and x_2 experience the same channel gain.

More recently, a new precoder design has been proposed for the OSM which can further enhance the performance by maximizing the minimum Euclidean distance, denoted as $d_{\min}$, in each symbol space[11]. We refer to this scheme as *enhanced* OSM (E-OSM). The E-OSM employs a real precoding matrix $\boldsymbol{P} \in R^{2\times 2}$ at the transmitter. The resulting expression for the received signal is rewritten as

$$y = \boldsymbol{H}_{\theta_o}\boldsymbol{P}s(x) + n.$$

Here $\boldsymbol{P}$ consists of two real rotation matrices $\boldsymbol{R}_{\theta 1}$ and $\boldsymbol{R}_{\theta 2}$ and one real diagonal matrix $\boldsymbol{D}$ as[11]

$$\boldsymbol{P} = \boldsymbol{R}_{\theta 1}\boldsymbol{D}\boldsymbol{R}_{\theta 2}, \tag{3}$$

where

$$\boldsymbol{R}_{\theta 1} = \begin{bmatrix}\cos\theta_1 & -\sin\theta_1 \\ \sin\theta_1 & \cos\theta_1\end{bmatrix},\ \boldsymbol{D} = \begin{bmatrix} p & 0 \\ 0 & \sqrt{2-p^2}\end{bmatrix},$$
$$\text{and } \boldsymbol{R}_{\theta 2} = \begin{bmatrix}\cos\theta_2 & -\sin\theta_2 \\ \sin\theta_2 & \cos\theta_2\end{bmatrix}.$$

The E-OSM precoding is performed in two steps. First, the rotation matrix $\boldsymbol{R}_{\theta 1}$ orthogonalizes $h_{r,1}^{\theta_o}$ and $h_{r,2}^{\theta_o}$ as well as $h_{r,3}^{\theta_o}$ and $h_{r,4}^{\theta_o}$. Then, the remaining matrices $\boldsymbol{D}$ and $\boldsymbol{R}_{\theta 2}$ are jointly optimized by the following criteria

$$(\hat{p},\hat{\theta_2}) = \arg\max_{p,\theta} d_{\min}(p,\theta_2).$$

After mathematical derivations, it has been shown in [11] that the optimized components p and θ_2 can be represented by only two and four cases for 4QAM and 16QAM, respectively. The optimal solutions for each constellation are shown in Table I (refer [11] for a parameter k).

Adopting the optimized precoder (3), the E-OSM offers an excellent performance gain compared to other closed-loop schemes with a simple ML detection (called *single-symbol decodable*). Also, the precoder computation is simple in comparison to singular value decomposition (SVD) based schemes. In addition to the feedback information θ_o in the original OSM, the E-OSM needs one more phase value θ_1 and $0.5\log_2 M_c$ bits for feedback.

표 1. $\hat{p}$와 $\hat{\theta_2}$ 파라미터 값
Table 1. Closed form expressions of $\hat{p}$ and $\hat{\theta_2}$

modulation	case	$\hat{p}$	$\hat{\theta_2}$
4QAM	$1 \le k < 7$	1	$\pi/4$
	$k \ge 7$	$\sqrt{2}$	0.464
16QAM	$1 \le k < 7.58$	1	$\pi/4$
	$7.58 \le k < 43.11$	1	0.488
	$43.11 \le k < 101.03$	1	0.345
	$k \ge 101.03$	$\sqrt{2}$	0.245

Ⅲ. Proposed Blockwise OSM

As discussed in Section II, the main limitation of the OSM scheme is its constraint on the number of supported data streams. In this section, we propose an extended structure of the OSM scheme which does not have any restriction on the number of data streams. For the ease of our description, we assume $M_t = 2K \le M_r$, where K denotes the number of subblock channels which will be explained later. Also B is assumed to equal M_t.

3.1 Blockwise OSM

We first describe a system structure for the proposed scheme, as shown in Figure 2. One noticeable change is the adoption of the linear receive filter $\boldsymbol{G} \in C^{M_t \times M_r}$. The function of this

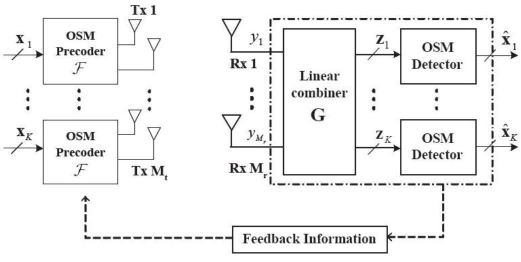

그림 2. B-OSM 시스템의 블록 구조도
Fig. 2. Block diagram of B-OSM systems

combining matrix G is to partition the transmitted data streams into K independent subgroups. After K subgroups are arranged at the transmitter, the OSM is applied into each subgroup to support two data streams. The filtered output signal vector $z \in C^{M_t \times 1}$ at the receiver can be expressed from (2) as

$$z = GHF_B(x,\theta) + Gn = H_G F_B(x,\theta) + v$$

where F_B accounts for the OSM rotation operation at the transmitter, $H_G \in C^{M_t \times M_t}$ is defined as $H_G = GH$, and $v \in C^{M_t \times 1}$ represents the filtered noise vector with the covariance matrix $\sigma_n^2 GG^H$. We call this scheme as *blockwise* OSM (B-OSM). When the same procedure is applied to the E-OSM scheme, we refer to it as *blockwise* E-OSM (BE-OSM). We omit the overlapped description for the BE-OSM for the remaining part of this paper.

Now the question is how to design the receive combining matrix G. The main objective is to suppress the interference coming from other subgroups. This concept was originally suggested for code-division multiple access (CDMA) multi-user detection and was applied to multiple antenna systems, referred to as GD[12,13], where projection matrices were used to obtain G. However, they suffer from a considerable performance loss because they ignore the noise correlation. Moreover, the complexity for filter calculation and detection can be high if the grouping is overlapped or the group size is large.

In the following, we illustrate an improved design method for G, which can overcome the performance degradation by preserving the noise statistics. While the size of each subgroup can be adjusted to an arbitrary value, all the group sizes are fixed to two in the B-OSM system structure. First, we express the original channel matrix H of (1) as

$$H = [H_1 \, H_2 \, \cdots \, H_K] \tag{4}$$

where $H_k \in C^{M_r \times 2}$ $(1 \le k \le K)$ represents the k-th subgroup channel matrix which consists of the $(2k-1)$-th and the $2k$-th column vectors of H. We denote $h_{k,1}$ and $h_{k,2}$ as the first and second column vector of H_k $(H_k = [h_{k,1} \, h_{k,2}])$. Its complementary channel matrix $\widetilde{H}_k \in C^{M_r \times (M_t - 2)}$ is defined as

$$\widetilde{H}_k = [H_1 \cdots \, H_{k-1} \, H_{k+1} \cdots \, H_K].$$

By the same notation as in (4), G can be similarly expressed as

$$G = [G_1^T \, G_2^T \cdots \, G_K^T]^T$$

where $G_k \in C^{2 \times M_r}$ represents the k-th subgroup combining matrix which consists of the $(2k-1)$-th and the $2k$-th row vectors of G. Also $g_{k,1}$ and $g_{k,2}$ denotes the first and second row vector of G_k $(G_k = [g_{k,1}^T \, g_{k,2}^T]^T)$. We want to design G_k such that the k-th subgroup channel H_k is free from the interference channel $\widetilde{H}_k$ as

$$G_k \widetilde{H}_k = 0 \tag{5}$$

However, by relaxing this null constraint, a better solution can be obtained. The solution based on the MMSE criterion is given by

$$g_{k,i} = [(\hat{H}_{k,i}^H \hat{H}_{k,i} + \sigma_n^2 I_{M_t - 1})^{-1} \hat{H}_{k,i}^H]_{(1,:)}, \quad i = 1,2 \tag{6}$$

where $[A]_{(1,:)}$ denotes the first row of A, and $\widehat{H_{k,i}}$ consists of the desired channel vector $h_{k,i}$ and the interference channel matrix $\widetilde{H}_k$ as

$$\widehat{H_{k,i}} = [h_{k,i} \; \widetilde{H_k}].$$

By taking the first row of the MMSE based filter matrix (6) of $\widehat{H_{k,i}}$, the interference channel $\widetilde{H}_k$ can be nearly eliminated in the k-th subgroup channel. We note that although each $g_{k,i}$ is optimal, the solution G_k obtained from (6) may not be optimum in the MMSE sense. Nevertheless, the proposed method provides a good performance in low SNRs by taking the noise components into account, as will be shown in the simulation section.

Then, we perform the Gram-Schmidt process in the row space of G_k in (6) to preserve the statistics of the subblock noise vectors as follows:

- $g_{k,1} \leftarrow g_{k,1} / \parallel g_{k,1} \parallel$
- $g_{k,2} \leftarrow g_{k,2}\left(I_{M_t} - g_{k,1}^H g_{k,1} / \parallel g_{k,1} \parallel^2\right)$, then
  $g_{k,2} \leftarrow g_{k,2} / \parallel g_{k,2} \parallel$.

This simple projection process finds the orthogonal basis without a noise whitening filter using eigen-value decomposition (EVD).

Now we return to our system model. By applying the receive combining filter G obtained by the above method, the resulting effective channel model H_G is represented as

$$H_G = GH = \begin{bmatrix} H_{G,1} & G_1H_2 & \cdots & G_1H_K \\ G_2H_1 & H_{G,2} & \cdots & G_2H_K \\ \vdots & & \ddots & \vdots \\ G_KH_1 & G_KH_2 & \cdots & H_{G,K} \end{bmatrix}.$$

where $H_{G,k} = G_kH_k$ is the k-th subblock channel matrix of size 2×2, and G_kH_j accounts for residual interference from other subblock $j \neq k$ to the k-th subblock.

The output signal vector z_k at the k-th subblock channel can be written as

$$z_k = H_{G,k}F(x_k, \theta_o^{(k)}) + \sum_{j \neq k} G_kH_jF(x_k, \theta_o^{(j)}) + v_k \quad (7)$$

where x_k, z_k, and v_k ($\in C^{2\times 1}$) denote the transmit signal vector, the received signal vector, and the noise vector for the k-th subblock, respectively. Also $\theta_o^{(k)}$ stands for the rotation angle for the k-th subblock. Note that thanks to the previous Gram-Schmidt process, the noise vector v_k remains white, i.e. $E[v_kv_k^H] = \sigma_n^2 G_kG_k^H = \sigma_n^2 I_2$. As will be shown in the following section, this property results in good performance.

Applying the Gaussian approximation to the interference term in (7), it is clear that the resulting effective channel model can be considered as being equivalent to the original channel model (2). Thus, the OSM can be directly applied to each transmission subblock. The receiver calculates the rotation angle $\theta_o^{(k)}$ independently and feeds back to the transmitter. Then at the transmitter, the k-th input symbol vector x_k is precoded by the function $F(x_k, \theta_o^{(k)})$ in (7), and is transmitted over its corresponding transmit antennas. For the case of BE-OSM, the additional precoding P_k is applied in each subblock before transmission for enhanced performance. Note that the B-OSM scheme still maintains the single-symbol decodability for each subblock channel with complexity $O(M_c)$, regardless of the number of data streams.

We comment on the feedback overhead of the proposed scheme. The amount of feedback information for the B-OSM and the BE-OSM is presented in Table II. In general, the optimal SVD based transmissions require the full CSI to determine the precoders at the transmitter[9]. When the number of data streams B grows, the feedback overhead of our scheme increases only linearly with K ($K \propto B$), while the overhead of the SVD based schemes is proportional to K^2. Thus compared to SVD based designs, our proposed schemes have an advantage for systems with a limited feedback channel.

표 2. $M_t = 2K$일 때 B-OSM과 BE-OSM의 피드백 정보량
Table 2. Feedback information amount for B-OSM and BE-OSM with $M_t = 2K$

B-OSM	$\{\theta_o^{(k)}\}_{k=1}^K$
BE-OSM	$\{\theta_o^{(k)}, \theta_1^{(k)}\}_{k=1}^K$, $(K/2 \cdot \log_2 M_c)\ bits$

3.2 Group Selection for B-OSM and BE-OSM

Although the BE-OSM achieves the optimal performance for a given subblock, the BE-OSM may not guarantee the optimality in the overall sense. This corresponds to a group selection issue in the GD technique. By choosing the groups of the channel columns properly, the performance of the GD can be further increased. In [13], an adaptive grouping method was proposed which considers the channel correlation. However, this algorithm suffers from high complexity cost, as it allows overlapped grouping. Also we note that the correlation alone does not account for the performance of the ML detection much.

In this subsection, we present a group selection criteria based on the minimum distance $d_{\min}$ and illustrate how this group selection can be efficiently combined with the B-OSM. Our design goal is to minimize the error performance with the optimal ML detection. In this case, for uncoded transmission, $d_{\min}$ is directly related to the symbol error rate (SER) from the union bound approach[14]. Thus, the following criteria is used:

$$\hat{n} = \arg\max_{1 \le n \le N}\left[\min_{1 \le k \le K} d_{\min}(n,k)\right] \quad (8)$$

where $d_{\min}(n,k)$ represents the minimum Euclidean distance for the k-th subblock of the rearranged channel with the n-th group pairing. Here N denotes the total number of all possible combinations of pairs among M_t channel columns.

Then we have $N = \binom{M_t}{2}\binom{M_t-2}{2}\cdots\binom{2}{2}/K!$. For example, $N=3$ and $N=15$ combinations exist for $M_t=4$ and $M_t=6$, respectively. In the criteria (8), we identify the combination of pairs n whose worst case minimum distance is the highest among all of N pair combinations.

In the original GD, this criteria is somewhat useless because of its complexity for the overall candidate search. However, in the proposed scheme, $d_{\min}(n,k)$ can be efficiently computed by utilizing reduced candidate search in one dimension from the OSM property. Actually it was shown in [10] that the required number of candidate vectors in computing $d_{\min}(n,k)$ for the OSM is only 2, 5, and 19 for 4QAM, 16QAM, and 64QAM, respectively. Therefore, the complexity related to the group selection is low in the proposed scheme. We will verify in the next section that an additional gain is achieved if the grouping is utilized.

Based on this criteria (8), the receiver determines the optimal pairing set and feeds back its index to the transmitter. This requires $\lceil \log_2 N \rceil$ additional bits, where $\lceil a \rceil$ indicates the smallest integer greater than or equal to a. For example, $\lceil \log_2 3 \rceil = 2$ bits and $\lceil \log_2 15 \rceil = 4$ bits are needed for $M_t=4$ and $M_t=6$, respectively.

Ⅳ. Simulation Results

Through Monte-Carlo simulations, this section demonstrates the efficacy of the proposed scheme in flat Rayleigh fading channels. We consider a MIMO channel with $M_t=4$ and $M_r=4$ ($K=2$) over which we transmit $B=4$ independent data streams.

First, we compare the BER performance of the B-OSM and the BE-OSM scheme with two optimal power allocation schemes based on MMSE[6] and ARITH-BER criteria[9], respectively,

in Figure 3. The group selection is not considered here and thus a fixed grouping is used instead of optimized one. The simulation employs 4QAM for all substreams and the receive combining filter in (6). The proposed B-OSM provides about 10dB and 2dB gains at a BER of 10^{-4} over two above-mentioned schemes, respectively, with only two phase values $\theta_o^{(1)}$ and $\theta_o^{(2)}$ feedback. Allowing two bits and two phase values $\theta_1^{(1)}$ and $\theta_1^{(2)}$ more for feedback, the BE-OSM obtains an additional 6dB gain and even largely outperforms the ARITH-BER

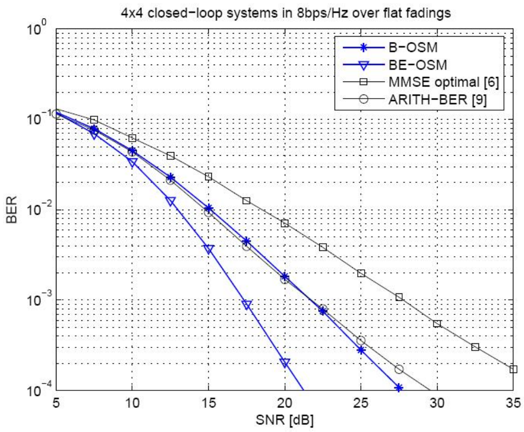

그림 3. $M_t=4$ 및 $M_r=4$의 안테나 환경에서 제안하는 기법의 비트에러율 성능
Fig. 3. BER performance of the proposed schemes with $M_t=4$ and $M_r=4$

method at a BER of 10^{-4}. Note that the proposed schemes do not require the additional complexity for power allocation.

In Figure 4, we present the simulation results to confirm the benefits of the optimized pairing for the proposed schemes. In this simulation, the group pairing is performed with a 2-bit feedback signal based on the selection criteria (8) with 4QAM. This criteria maximizes the overall minimum distance such that any ill-conditioned subblock can be taken care of. The plot shows that optimizing the grouping pairs achieves about a 2.5dB gain at a BER of 10^{-4}

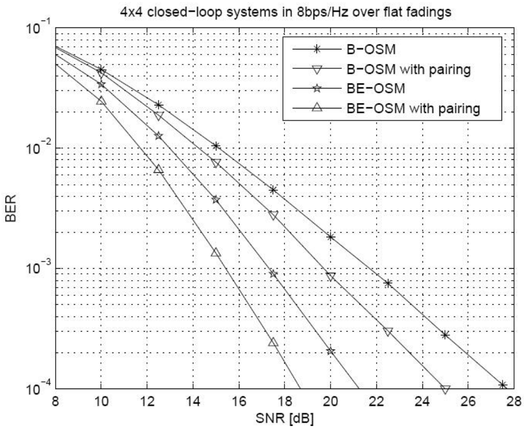

그림 4. $M_t = 4$ 및 $M_r = 4$의 안테나 환경에서 그룹 선택을 결합한 제안하는 기법의 비트에러율 성능
Fig. 4. BER performance of the proposed schemes with group selection with $M_t = 4$ and $M_r = 4$

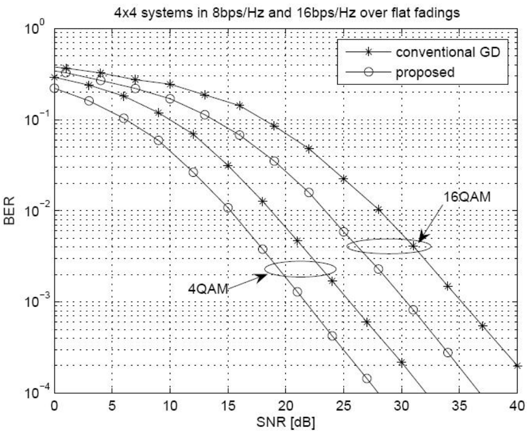

그림 5. $M_t = 4$ 및 $M_r = 4$의 안테나 환경에서 제안하는 기법과 기존의 GD와의 비트에러율 성능 비교
Fig. 5. BER comparison of the proposed grouping method with conventional GD with $M_t = 4$ and $M_r = 4$

for both the B-OSM and the BE-OSM schemes.

In the last simulations, we compare the BER performance of the proposed grouping method for the B-OSM with the conventional GD method in [13]. In our method, a simple ML detection is carried out in one dimension. Figure 5 shows that the proposed method performs 4-5dB better than the conventional GD at all SNRs with both 4QAM and 16QAM. An improved performance gain is achieved by managing the noise whitening issue properly in the high SNR regime. Also, the performance gain at low SNRs comes from our MMSE based approach by relaxing the null constraint (5). As a result, by taking the noise components into account, the proposed block-wise transmission schemes accomplish highest performance gain.

V. Conclusion

In this paper, we have presented a new efficient design scheme for closed loop MIMO systems. The main objective of this work is the extension of the OSM scheme to multiple data stream transmission. To achieve this goal, we have partitioned data streams into several subblocks by executing the block-diagonalization process at the receiver. The simulation results confirm the efficiency of the B-OSM and the BE-OSM scheme in practical situations. The proposed schemes allow a simple single-symbol detection process, and also provide the flexibility for the feedback information amount compared to conventional SVD-based schemes. In addition, the pairing issue has been studied to optimize the performance. We have shown that the combination with the pairing offers an additional gain for the BER performance.

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