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Beam Selection and Antenna Selection: A Hybrid Transmission Scheme over MIMO Systems Operating with Vary Antenna Arrays

Feng Wang, Marek E. Bialkowski
School of Information Technology and Electrical Engineering, University of Queensland Brisbane, Brisbane, Australia
E-mail: {*fwang*, *meb*}*@itee.uq.edu.au*
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Abstract

In this paper, we have proposed a hybrid transmission scheme which involves beam selection and antenna selection techniques over a MIMO system operating with vary antenna array. Optimal subset of transmit antennas are selected via fast successive selection scheme designed to optimize the target eigenbeam. Optimal eigenbeams corresponding to the largest singular values of the new MIMO channel formed by the selected antennas are exploited for data transmission. To evaluate the performance of the proposed scheme, different array structures, including uniform linear array (ULA) and including uniform circular array (UCA) are employed in the simulations. The transmitter is assumed to be surrounded by scattering objects while the receiver is postulated to be free from scattering objects. The Laplacian distribution of angle of arrival (AoA) of a signal reaching the receiver is postulated. The results presented by this paper indicate that the proposed scheme can significantly improve the performance of data transmission in term of symbol error rate (SER).

Keywords: MIMO, Beam Selection, Antenna Selection, ULA, UCA

1. Introduction

In recent years, wireless multiple-input multiple-output (MIMO) systems featured by employing multiple antennas at the transmitter and receiver have attracted constant attention. Research has revealed that MIMO is capable of improving system capacity and link robustness significantly in a rich scattering environment without costing extra transmit power and bandwidth [1-3]. The data transmission schemes over a MIMO channel can be categorized into three types. The first one is multiplexing gain maximizing schemes, which are exploited to transmit independent data streams in parallel through multiple spatial channels. Bell Labs Layered space-time (BLAST) techniques [4,5] are typical multiplexing schemes to increase the data rate. The second one is diversity gain maximizing schemes, which are exploited to combat the fading effect of wireless channel and improve the bit-error-rate (BER) performance. To obtain the diversity gain, the signals carrying the same information are transmitted through redundant independent fading channels. As a result, the data transmission reliability is assured. Space-time coding techniques [6,7] are typical schemes of exploiting spatial diversity. However, maximizing multiplexing gain may not necessarily maximize diversity gain and vice versa. Consequently, scholars also tried to achieve trade-off between the multiplexing and diversity gain [8].

With the perfect channel state information (CSI), the optimal design, in terms of maximum data rate is multichannel beamforming (MB) [9]. This strategy transmits data on the eigenmodes (or eigenbeams) of the MIMO channel using linear transmit-receive processing [10,12]. It has been well noted that the eigenmodes corresponding to the small singular values are showing a poor performance. Targeting at this problems, power allocating and bit-loading schemes were proposed in [11]. By allocating different power and constellations among a subset of available eigenmodes, a significantly improved performance can be achieved. The complexity of such a scheme is significant since each channel eigenmode requires a different combination of signal constellation and codes, depending on the allocated power [13]. For this reason, equal power and bit allocation scheme is still a popular transmission scheme. With equal power and bit allocation, an improved performance can also be obtained by selecting the optimal eigenmodes (which are corresponding to the largest singulars) [9].

Antenna selection (AS) has been proposed for enhances performance in correlated fading [14,15]. By selecting a small number of optimal antennas from a large set of antennas, AS is capable of capturing a large portion of the channel capacity of MIMO [19]. AS also results in a reduced hardware cost and computational complexity [17]. AS have received a plenty of attentions. A couple of antenna selection criteria have been studied in [14]. To avoid the theory-optimal exhaustive search scheme, fast antennas selection schemes have been proposed [17-19].

In this paper, we proposed a hybrid transmission scheme over a MIMO system, in where, the optimal eigenmodes for the optimal antenna subset are employed to transmit data. To achieve this goal, a fast antenna selection scheme designed for the eigenbeam selection is proposed. The performance of the proposed scheme is evaluated in a spatial correlation fading environment. Previous researches show that spatial correlation always has a negative impact on the MIMO performance [21,22]. It can be expected that difference configurations of antenna arrays will result in different spatial correlations of transmitted and received signals. As a result, the channel properties between the transmitter and receiver could be different. In this paper, we select uniform linear array (ULA) and uniform circular array (UCA) as the objects. However, it is worthwhile noticing that we can reach similar conclusions by reasoning the similarity between UCA and other alike structures such as triangular, square, pentagonal or hexagonal arrays.

This paper is organised as follows. In Section 2, we briefly introduce the signal model and channel model which are used in the rest of the paper. In Section 3, the hybrid scheme is presented by introducing the eigenbeam selection scheme with LMMSE receiver and the fast antenna selection scheme designed for the eigenbeam selection scheme. In Section 4, we provide error rate results to show the performance of the proposed scheme in spatial correlated fading channels. Conclusions are given in Section 5.

2. Signal Model

2.1. System Model

Consider a wireless MIMO system employing N transmit and M receiver antennas, The impulse response for the MIMO channel can be modeled by a $M \times N$ matrix $\mathbf{H}$ with the $(m,n)^{\text{th}}$ element containing the complex fading parameter between the n^{th} transmit and m^{th} receive antenna. The baseband equivalent signal model can be represented by

$$\mathbf{r} = \mathbf{HX} + \mathbf{n} \tag{1}$$

$\mathbf{r} \in \mathbb{C}^{M\times 1}$ is the received signal vector, $\mathbf{X} \in \mathbb{C}^{N\times 1}$ represents the transmitted signal vector, $\mathbf{n} \in \mathbb{C}^{M\times 1}$ is the additive white Gaussian noise vector with covariance matrix given by

$$E\left(\mathbf{nn}^{\dagger}\right) = \sigma_n^2\, \mathbf{I}_M \tag{2}$$

We assume that perfect Channel State Information (CSI) is available for both the receiver and transmitter sides. The transmitted signal vector can be written as

$$\mathbf{X} = \mathbf{ws} \tag{3}$$

where $\mathbf{w} \in \mathbb{C}^{N\times L}$, $L \le \min(N,M)$ is the transmit adaptive switching matrix which maps the L modulated data symbols s_i ($1 \le i \le L$) onto the N transmit antennas. The symbol s_i is the elements of $\mathbf{s}$, with E $\{\mathbf{ss}^{\dagger}\}$ = $\mathbf{I}_L$, and chosen from all possibly signal constellations. $\mathbf{X}$ is subject to the power constraint, which is given by

$$E\left\{\left\|\mathbf{X}\right\|_2^2\right\} = tr\left(\mathbf{ww}^{\dagger}\right) = E_s \tag{4}$$

2.2. Channel Model

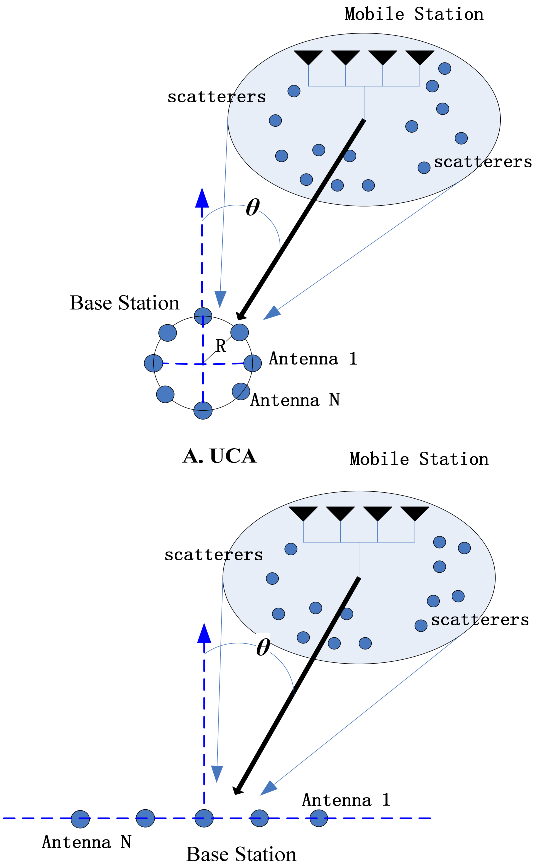
We assume that the wireless MIMO link between the transmitter and receiver is undergoing a flat-fading narrow-band fading. The Kronecker model [21] is utilized, in where the spatial correlations at the transmitter and receiver are independent and separable. As a result, the complex channel matrix $\mathbf{H}$ can be given as

$$\mathbf{H} = \mathbf{R}_R^{1/2}\mathbf{G}\mathbf{R}_T^{1/2} \tag{5}$$

where $\mathbf{R}_R$ is the spatial correlation matrix at the receiver and $\mathbf{R}_T$ represents the spatial correlation at the transmitter. In a scattering rich signal propagation environment, the antenna array is surrounded by scattering objects. Based on the assumption that these scattering objects are uniformly distributed in a circle, the correlation experienced by a pair of antennas with large inter-antennas spacing in an array can be written as

$$R\,(m,n) = J_0\,[2\pi(m-n)/\lambda] \tag{6}$$

On the other hand, if the antenna array is free from any surrounding objects, the correlations matrices at the transmitter and receiver sides are subjective to the array structure. **Figure 1** demonstrates the model under considered, where the receiver are surrounded by uniformly distributed scatterers and transmitter equipped with ULA or UCA is located high above ground where there are no scattering objects. All the antenna elements at receiver and transmitter have an omni-directional radiation pattern in the azimuth plane. The θ is the central Angle of Departure (AoD) or Angle of Arrival (AoA), we assume



A. UCA

B. ULA

Figure 1. Signal model of a downlink MIMO multiuser system.

that θ follows the Laplacian distribution. For the case of ULA, the real and imaginary components of the spatial correlation parameter between the m^{th} and n^{th} elements are given as [23]

$$\mathrm{Re}\left(R_{m,n}\right)=J_0\left(Z_l\right)+2\sum_{k=1}^{\infty}\frac{a^2\left(a-\mathrm{e}^{-a\pi}\right)}{a^2+4k^2}J_{2k}\left(Z_l\right)\cos(2k\theta) \tag{7}$$

$$\mathrm{Im}\left(R_{m,n}\right)= 4C_l\sum_{k=1}^{\infty}\frac{a\left(a+\mathrm{e}^{-a\pi}\right)}{a^2+(2k+1)^2}\times J_{2k+1}\left(Z_l\right)\sin\left[(2k+1)\theta\right] \tag{8}$$

Similarly, for the case of UCA, the real and imaginary components of spatial correlation parameter between the m^{th} and n^{th} elements are given as [23]

$$\mathrm{Re}\left(R_{m,n}\right)= J_0\left(Z_c\right)+2\sum_{k=1}^{\infty}\frac{a^2\left(a-\mathrm{e}^{-a\pi}\right)}{a^2+4k^2}J_{2k}\left(Z_c\right)\cos\left[2k\left(\theta+\alpha\right)\right] \tag{9}$$

$$\mathrm{Im}\left(R_{m,n}\right)= 4C_l\sum_{k=1}^{\infty}\frac{a\left(a+\mathrm{e}^{-a\pi}\right)}{a^2+(2k+1)^2}\times J_{2k+1}\left(Z_c\right)\sin\left[(2k+1)(\theta+\alpha)\right] \tag{10}$$

In the expressions (7)-(10), a is the decay factor related to the angle spread, specifically, as a increases, the angle spread decreases. It also decides the normalizing constant [23]

$$C_l=\frac{a}{2\left(1-\mathrm{e}^{-a\pi}\right)} \tag{11}$$

$J_n(*)$ represents the n^{th} order Bessel function of the first kind. The parameter α is the relative angle between the m^{th} and n^{th} elements in a UCA. Let φ_m and φ_n represent the angle of the m^{th} and n^{th} elements in an azimuth plane, then we have

$$\sin\alpha=\frac{\cos\varphi_m-\cos\varphi_n}{\sqrt{2-2\left(\cos\left(\varphi_m-\varphi_n\right)\right)}} \qquad \cos\alpha=\frac{\sin\varphi_m-\sin\varphi_n}{\sqrt{2-2\left(\cos\left(\varphi_m-\varphi_n\right)\right)}} \tag{12}$$

Z_l and Z_c are related to the antenna spacing in ULA and UCA, they can be expressed as

$$Z_l=\frac{2\pi d}{\lambda}\left|m-n\right| \tag{13}$$

$$Z_c=\frac{2\pi R}{\lambda}\sqrt{2-2\left(\cos\left(\varphi_m-\varphi_n\right)\right)} \tag{14}$$

The formulas (7)-(10) show that the spatial correlations between two antenna elements in an ULA and UCA are characterized by a couple of parameters, including inter-element spacing (d/λ or d/lambda), AoA and decay factor. **Figure 2** shows the spatial correlation between the antenna 1 and 2 in regards to inter-element spacing with different decay factors. We can see from **Figure 2** that for both ULA and UCA, large decay factors result in an increased spatial correlation. **Figure 3** and **Figure 4** are the correlation surface in regards to the inter-element spacing and AoA. One can see from **Figure 3** and **Figure 4** that for a given inter-element spacing, the spatial correlation in a ULA increase significantly with AoA (it becomes more apparent when inter-element spacing is larger than 0.5), while in a UCA the spatial correlation remains almost constant in regards to AoA.

3. Beam Selection and Antenna Selection: A Hybrid Scheme

In this section, we present the ideal of the hybrid trans-

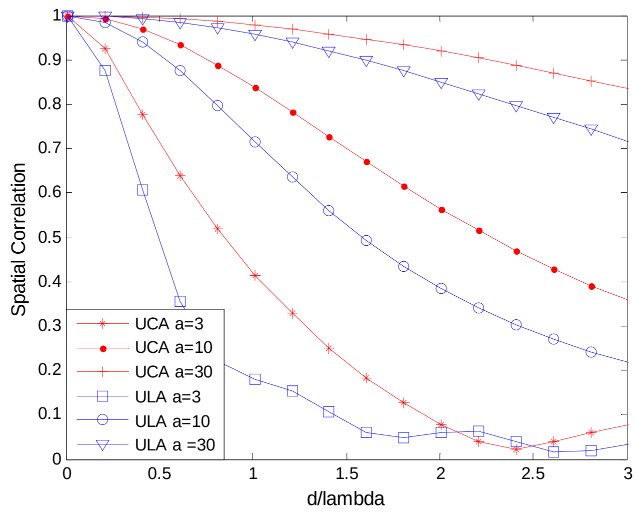


Figure 2. Spatial correlation between antenna 1 and 2 (four-antenna ULA and UCA, AoA = 0°).

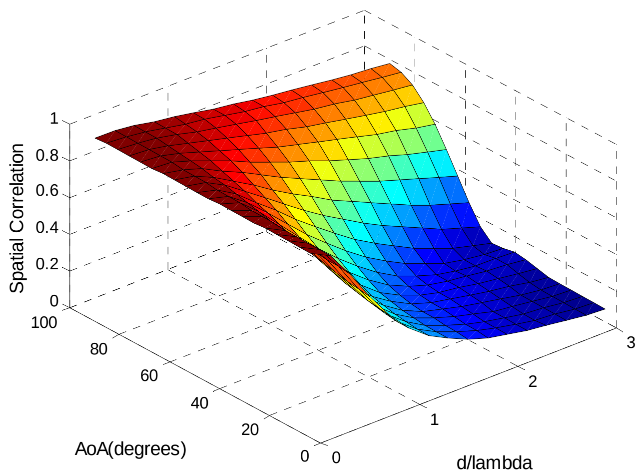


Figure 3. Spatial correlation between antenna 1 and 2 (four-antenna ULA, decay factor = 5).

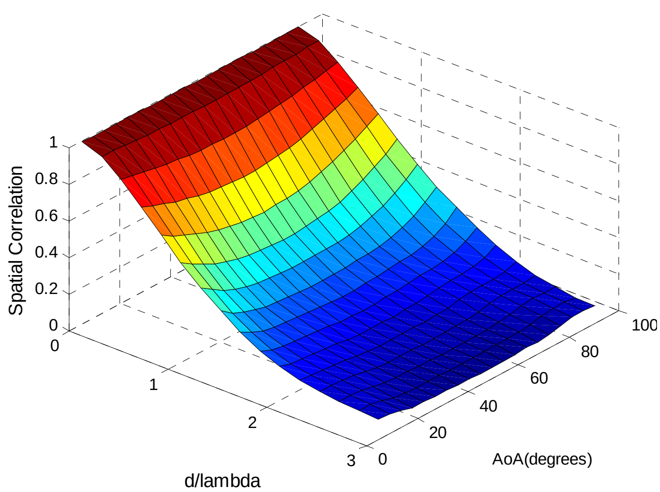


Figure 4. Spatial correlation between antenna 1 and 2 (four-antenna UCA, decay factor = 5).

mission scheme by introducing beam selection scheme and antenna selection scheme respectively. Logically, the design of antenna selection scheme and performance metrics is based on beamforming and beam selection scheme. The beamforming and beam selection scheme are based on the current channel station information, and the antenna selection scheme is going change the MIMO channel matrix. Consequently, beamforming and beam selection is carried out after antenna selection. In our scheme, beamforming and beam selection is performed at transmitter side, and antenna selection can be implemented at either transmitter or receiver side.

3.1. Beam Selection Scheme

We assume that antenna selection scheme has finished. The output of the antenna selection scheme is a new channel matrix between the transmitter and receiver with smaller dimensions (smaller number of columns for transmits antenna selection and smaller number of rows for receives antenna selection). We represent the new complex matrix as can be given as $\hat{\mathbf{H}}_{P\times Q}(P \le M, Q \le N)$. Without loss of generality, we assume that $P \le Q$. As a result, by applying SVD technique, the complex channel can be given as

$$\hat{\mathbf{H}}_{P\times Q} = \mathbf{U\Sigma V}^H = \underbrace{[u_1 \cdots u_P]}_{\mathbf{U}} \begin{bmatrix} d_1 & & & & & \\ & \ddots & & & & \\ & & d_P & 0 & \cdots & 0 \end{bmatrix} \underbrace{[v_1 \cdots v_Q]}_{\mathbf{V}}^H \tag{15}$$

where$(\cdot)^H$ denotes the hermitian operation and d_p is the p^{th} non-negative singular values with $d_1 \ge d_2 \ge \cdots \ge d_P$ and **U** and **V** are the left and right unitary matrices, respectively. We have

$$\mathbf{U}^H\mathbf{U} = \mathbf{I} \in \mathbb{C}^{P\times P}$$
$$\mathbf{V}\mathbf{V}^H = \mathbf{I} \in \mathbb{C}^{Q\times Q} \tag{16}$$

In fact, **V** is the matrix with all the columns are the eigenvectors of $\hat{\mathbf{H}}^H\hat{\mathbf{H}}$, which is related to the eigen-modes of the MIMO communication channel. The equation (15) can be written in a different way, which is given as

$$\left(\hat{\mathbf{H}}^H\hat{\mathbf{H}}\right)\mathbf{V} = \mathbf{V}\left(\mathbf{\Sigma}^H\mathbf{\Sigma}\right) \tag{17}$$

For transmit beamforming, the optimal beam directions are along the eigenvectors of the $\hat{\mathbf{H}}^H\hat{\mathbf{H}}$ [12]. In other words, eigen-beamforming [9,12] utilized the eigen-modes of the auto-correlated $\hat{\mathbf{H}}^H\hat{\mathbf{H}}$ to transmit signal symbols. By assuming that there are L ($L \le \min$ (P, Q)) data streams are being transmitted to the receiver, the L columns of $\boldsymbol{V}$ corresponding to the L largest eigen-values are selected as the transmitting beamforming matrix. Consequently, the received signal vector is given as

$$\mathbf{r} = \sqrt{\frac{E_s}{N}}\hat{\mathbf{H}}\mathbf{V}_L\mathbf{s} + \mathbf{n} \tag{18}$$

By weighting the received signal vector, the estimated data at the receiver can be defined as

$$\tilde{\mathbf{s}} = \mathbf{F}^H\mathbf{r} \tag{19}$$

The error vector then can be given as

$$\mathbf{e} = \tilde{\mathbf{s}} - \mathbf{s} \tag{20}$$

The mean squared error (MSE) matrix will be then defined as the covariance matrix of the error vector, which can be presented as

$$\mathbf{E} = E\left[\mathbf{e}\mathbf{e}^H\right] = \mathbf{F}^H\mathbf{R}_r\mathbf{F} + \mathbf{I} - \mathbf{F}^H\hat{\mathbf{H}}\mathbf{V}_L - \mathbf{V}_L^H\hat{\mathbf{H}}^H\mathbf{F} \tag{21}$$

where

$$\mathbf{R}_r = \hat{\mathbf{H}}\mathbf{V}_L\mathbf{V}_L^H\hat{\mathbf{H}}^H + \mathbf{R}_n \tag{22}$$

The MSE of the l^{th} symbol transmitted to the receiver is the l^{th} diagonal element of **E**. Given the transmit beamforming matrix $\mathbf{V}_L$, the optimal receive matrix $\mathbf{F}_{opt}$ is obtained such that diagonal elements of **E** are minimized. This is equivalent to the solve

$$\min_{\mathbf{F}^H} \mathbf{c}^H\mathbf{E}\mathbf{c} = \min_{\mathbf{F}^H} Tr(\mathbf{E}\mathbf{c}\mathbf{c}^H), \ \forall \mathbf{c} \tag{23}$$

By setting the gradient of (23) to zero, and particularizing **c** for all the vectors of the canonical base, it follows that

$$\mathbf{F}_{opt} = \left(\hat{\mathbf{H}}\mathbf{V}_L\mathbf{V}_L^H\hat{\mathbf{H}}^H + \mathbf{R}_n\right)^{-1}\hat{\mathbf{H}}\mathbf{V}_L \tag{24}$$

which is the linear minimum MSE (LMMSE) receiver (Wiener solution). With this choice of linear receive and transmit filtering, the MIMO channel is decomposed into L parallel eigenmode sub-channels, with each can be expressed as

$$\hat{s}_l = \kappa_l\left(d_l\sqrt{p_l}s_l + n_l\right), \ l = 1, \cdots, L \tag{25}$$

where κ_l is a constant, which does not affect the received sub-channel SNR, p_l is the transmit power allocated to the l^{th} subchannel and d_l is the l^{th} largest singular value of $\hat{\mathbf{H}}_{P\times Q}(P \le M, Q \le N)$. The instantaneous received SNR via the l^{th} suchannel is given by

$$\gamma_l = \frac{d_l^2 p_l}{\sigma_n^2}, \quad l = 1, \cdots, L \tag{26}$$

Clearly, the overall received SNR can be lower bounded by

$$SNR_{overall} \ge \frac{d_L^2 p_L}{\sigma_n^2} \tag{27}$$

3.2. Antenna Selection Scheme

Equation (27) shows that the received SNR for the MIMO system with the choice of transmits and receives matrices is lower bounded by a monotonically increasing function of the L^{th} largest singular value of $\hat{\mathbf{H}}$. In the case that all the singulars are exploited for data transmission, the received SNR then will be lower bounded by a monotonically increasing function of the smallest singular value of $\hat{\mathbf{H}}$, which is confirmed in [14,16]. As a result, to match the aforementioned beam selection scheme, the antenna selection scheme in this paper is to seek the subset of transmit antennas and receive antennas with the largest L^{th} largest singular value. An optimal selection can be achieved by exhaustively searching over all possible combinations transmit and receive antennas. However, such an exhaustive search is hardly suitable for real-time implementation because of prohibitively long computational time. A number of complexity reduced antenna selection schemes [17-19] have been proposed, however these antenna selection schemes are targeting at maximizing the channel Shannon capacity.

In this paper, we utilize the successive selection approach which was firstly described in [18] in the context of channel capacity. However, in this paper, the target of the antenna selection has changed to maximizing the L^{th} largest singular value of $\hat{\mathbf{H}}$. Without loss of generality, we assume that the subset of antenna selection is performed at the receiver side. The receive antenna selection approach begins with full set of receive antennas available and then removes one of receive antenna per step. In each step, the antenna with the smallest contribution to the L^{th} largest singular value of the updated channel matrix is removed. This process is repeated until the required number of antennas remained. This approach can also be straightforwardly applied to the transmitter antenna selection.

Assume that we select O ($M \geq O \geq L$) out of M available receive antennas. The selection algorithm can be presented as follows

$$
\begin{aligned}
&Set\ \mathbf{G} = \mathbf{H}_{M\times N},\ Pt = 0\\
&\quad for\ i = 1\ to\ (M-O),\\
&\qquad for\ j = to\ (M-Pt)\\
&\qquad\quad Update\ \mathbf{G} := [\mathbf{g}_1^T, \cdots, \mathbf{g}_{j-1}^T, \mathbf{g}_{j+1}^T, \cdots, \mathbf{g}_{M-Pt}^T];\\
&\qquad\quad Calculate\ SVD\ over\ updated\ \mathbf{G}\\
&\qquad end\\
&\qquad find\ \hat{p} := \underset{j\leq \hat{p}\leq (M-Pt)}{\arg\min} \left(L^{th}\ Largest\ singular\ of\ G\right)\\
&\qquad Update\ \mathbf{G} := [\mathbf{g}_1^T, \cdots, \mathbf{g}_{\hat{p}-1}^T, \mathbf{g}_{\hat{p}+1}^T, \cdots, \mathbf{g}_{M-Pt}^T];\\
&\qquad Update\ Pt := Pt + 1\\
&\quad end\\
&let\ \hat{\mathbf{H}} := \mathbf{G}
\end{aligned}
\tag{28}
$$

The strategy requires *M*-*O* iterations, and the i^{th} iteration requires *M*-*I* + 1 space searches. As a result, the size of the search space is given as

$$S = \sum_{i=1}^{M-O} M - i + 1 \tag{29}$$

which is far less than the one obtained from the exhaustive search scheme. For instance, the total search space for the exhaustive search scheme amounts to 1820 when $O = 4$ and $M = 16$, while for the fast search employed in this paper only 126 iterations are used.

4. Numerical Results

Monto-carol simulations are performed to evaluate the performance of the proposed scheme in term of symbol-error-rate (SER). QPSK modulation with Gray coding is used for the data streams. By assuming that a Gray encoding is employed to map the bits into the constellation point, the bit-error-rate (BER) can be approximately obtained from SER by

$$BER \approx SER/R \tag{30}$$

where $R = \log_2 K$ is the number of bits per symbol and K is the constellation size.

Figure 5 shows the SER performance of transmit antenna selection without beam selection. We assume that LMMSE receiver is employed. For simplicity reason, the complex Rayleigh channel is utilized in the simulations. The receiver is equipped with 2 antennas. The number of RF chains at transmitter is 2, which means 2 optimal antennas out of all the available antennas will be activated during data transmission. We can see from **Figure 5** that for a give SNR, the SER performance improved significantly with the transmit antenna pool. When SNR = 10 dB, the system without antenna selection (2 × 2/2) achieves SER approximately at 6.5×10^{-2}, however when we increase the size of transmit antenna pool to 4 and 8, the SER then are decreased to 6.5×10^{-3} and 6.5×10^{-4}, respectively.

Figure 6 shows the SER performance of beam selection without antenna selection. We assume that LMMSE receiver is employed. Both the receiver and transmitter are equipped with 4 elements ULA array with inter-element spacing equals to 0.5 λ. Under Rayleigh channel, there are 4 eigenbeams at most can be exploited for data transmission. We can see from Figure 6 that for a give SNR, the SER performance improved significantly by beam selection. When SNR = 5 dB, the system without beam selection achieves SER approximately at 1.7×10^{-1}. However when we block the worst beams and select the best beams for data transmission, the SER performance is significantly improved. The SER is decreased to 5.5×10^{-2} when the worst eigenbeam is blocked

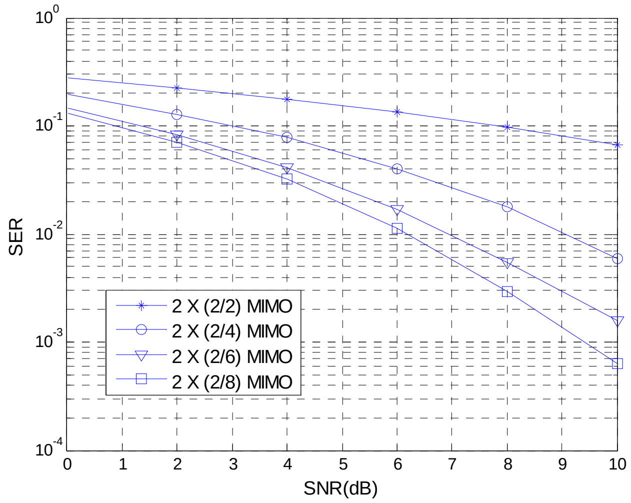


Figure 5. SER performance of Transmit antenna selection with LMMSE receiver under Gaussian Channel.

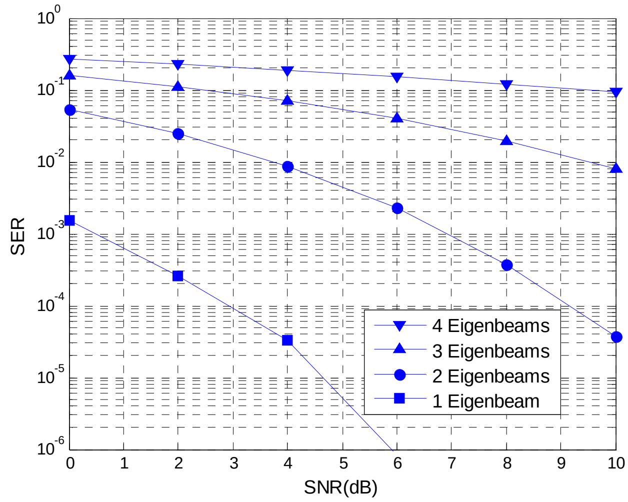


Figure 6 . SER performance of beam selection with LMMSE receiver under Rayleigh Channel.

from data transmission, to 5.5×10^{-3} when the worst 2 eigenbeams are blocked and to 1.7×10^{-5} when the worst 3 eigenbeams are blocked.

Figure 7 shows a compare of the SER performances of proposed scheme with beam selection. The receiver is equipped with 4 antennas. The transmitter has 4 RF branches. The results confirm the observations from Figure 6 that the performance is significantly improved when the worst beams are deactivated from data transmission. The MIMO system employs all the 4 eigenbeams show a poor SER performance (stared and dotted curves). When the 2 optimal eigenbeams out of 4 available eigenbeams are exploited, the SER is improved dramatically (circled and up-triangled curves). The improved performance is achieved by sacrificing data rate. When the number of available antennas (in **Figure 7**, the number is 8) is larger than the number of RF branches, the proposed fast antenna selection scheme can be utilized together with beam selection, thus rendering a hybrid scheme. It can be seen from Figure 7 that the proposed scheme is capable of achieving a better BER without sacrificing data rate. When SNR = 10 dB, the SER is further increased by the proposed scheme from 2.4×10^{-2} to 1.7×10^{-2} for UCA and 1.6×10^{-3} to 1.3×10^{-3} for ULA.

Figure 8 and **Figure 9** show the SER performance of hybrid scheme in where beam selection and antenna selection are both employed. We assume that there are 2 data streams to be transmitted. As a result, only 2 eigenbeams are exploited for data transmission. The receiver is equipped with a 4-element ULA or UCA. The transmitter has 4 RF chains and is equipped with a ULA with

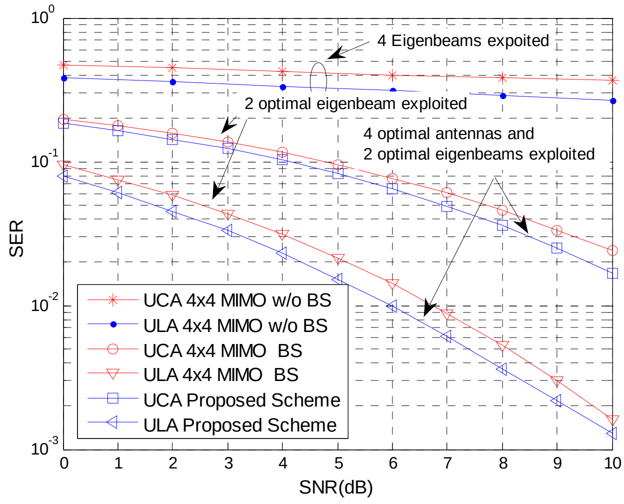


Figure 7. SER performance of the hybrid scheme with LMMSE receiver (AoA = π/6, *a* = 5).

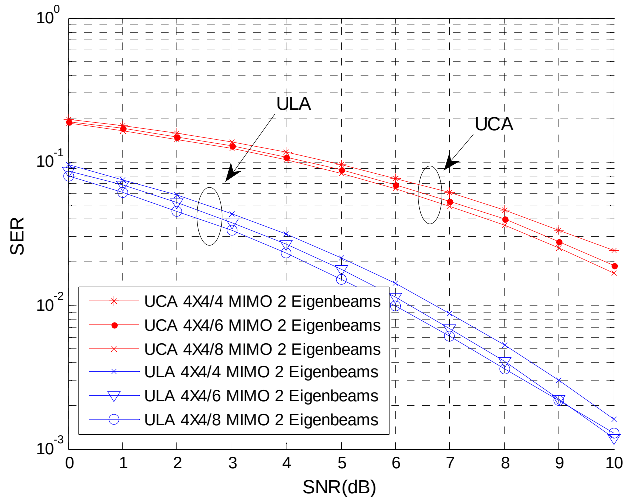


Figure 8. SER performance of the hybrid scheme with LMMSE receiver with vary array structures (AoA = π/6, *a* = 5).

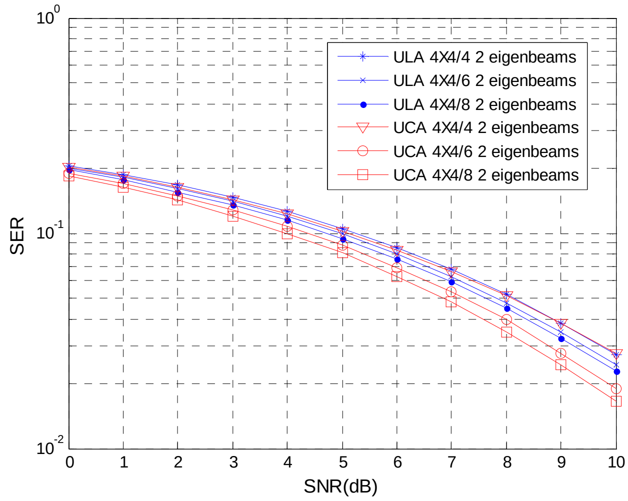


Figure 9. SER performance of the hybrid scheme with LMMSE receiver with vary array structures (AoA = π/3, *a* = 5).

inter-element spacing equals to 0.5 λ. Transmit antenna selection and beam selections are performed at transmitter side. We can see from these figures that with extra antennas the hybrid scheme is capable of improving the SER performance further. One can see from Figure 8, the system employing ULA shows a much better performance when AoA = $\pi/6$. However, when AoA increased to $\pi/3$, the system employing UCA takes dominant position (as shown in Figure 9). By comparing Figure 8 and Figure 9, it is easy to be noted that the increased AoA leads to a significant decrease in the SER performance for ULA. This observation confirms the results shown in Figures 3 and 4 that a UCA receiver is robust to AoA. The change of AoA results in little change of SER performance for UCA receievr. On the contrary, the receiver equipped with ULA is sensitive to AoA. An increased AoA could result in a significant decrease in the SER performance.

5. Conclusions

In this paper, we have proposed a hybrid transmission scheme which involves beam selection and antenna selection techniques over a MIMO system operating with vary antenna array. Optimal subset of transmit antennas are selected via fast successive selection scheme designed to optimize the target eigenbeam. Optimal eigenbeams corresponding to the largest singular values of the new MIMO channel formed by the selected antennas are exploited for data transmission. We also evaluated the performance of the proposed scheme with different array structures. In our simulations, the transmitter is assumed to be surrounded by scattering objects while the receiver is postulated to be free from scattering objects. The Laplacian distribution of angle of arrival (AoA) of a signal reaching the receiver is postulated. The results show that the proposed scheme is capable of achieving an improved SER performance. In regards to the array structure, we can conclude that ULA is preferred when AoA is small the constant while UCA is favoured when AoA is varying significantly.

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