

Compensation of channel information error using first order extrapolation in eigenbeam space division multiplexing (E-SDM)

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I. Introduction

In multiple-input multiple-output (MIMO) systems, eigenbeam space division multiplexing (E-SDM) [1], [2] has higher performance compared to conventional SDM and an advantage that simple detection algorithms are available at a receiver. However, the transmitter requires MIMO channel information. If it has an error, the performance degradation is inevitable. In this paper, we propose a compensation way against channel information error due to temporal channel transition. Also, the effect is numerically analyzed.

II. E-SDM and channel information

In this paper, we use a MIMO system model with M transmit-antennas (base station side) and N receive-antennas (terminal side) as shown in Fig. 1. For the sake of convenience, we assume a flat Rayleigh fading channel where any channels are uncorrelated. The N -dimensional receive signal vector is represented by the $N \times N$ channel matrix $\mathbf{A}$, the M -dimensional transmit signal vector $\mathbf{x}(t)$, and N -dimensional noise vector $\mathbf{n}(t)$ as

$$\mathbf{r}(t) = \mathbf{A}\mathbf{x}(t) + \mathbf{n}(t), \quad (1)$$

where each element of $\mathbf{n}(t)$ is Gaussian distributed (zero mean and variance σ^2) and mutually independent.

In E-SDM, it is assumed that the channel matrix is known at the transmitter. Let $\text{rank}(\mathbf{A}) = K$ ($K \leq \min(M, N)$). Then, we obtain K eigenvectors by singular value decomposition (SVD) of $\mathbf{A}$. Therefore, we can generate K eigenbeams using the eigenvectors as array weights, and transmit an individual substream in each beam. Thus, the transmit signal vector is given by

$$\mathbf{x}(t) = \sum_{k=1}^K \mathbf{w}_k s_k(t), \quad (2)$$

where $\mathbf{w}_k$ is the k th weight vector (the k th eigenvector of $\mathbf{A}^H \mathbf{A}$), and $s_k(t)$ is the k th substream. Substituting (2) into (1), we obtain

$$\mathbf{r}(t) = \mathbf{A} \sum_{k=1}^K \mathbf{w}_k s_k(t) + \mathbf{n}(t). \quad (3)$$

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To detect the k th substream from spatially multiplexed signals, we apply the receive weight vector $(\mathbf{A}\mathbf{w}_k)^*$ (maximal ratio combining (MRC) weight) to the received signal vector. Then, we have the k th array output

$$\begin{aligned} y_k(t) &= (\mathbf{A}\mathbf{w}_k)^H \left(\mathbf{A} \sum_{m=1}^K \mathbf{w}_m s_m(t) + \mathbf{n}(t) \right) \\ &= \lambda_k s_k(t) + \mathbf{w}_k^H \mathbf{A}^H \mathbf{n}(t), \end{aligned} \quad (4)$$

since the eigenvectors and the channel matrix satisfy

$$\mathbf{w}_i^H \mathbf{A}^H \mathbf{A} \mathbf{w}_j = \begin{cases} \lambda_i & i = j \\ 0 & i \neq j \end{cases}, \quad (5)$$

where λ_i denotes the i th eigenvalue of $\mathbf{A}^H \mathbf{A}$. This fact indicates that the MRC weight can reduce the inter-substream interference perfectly, as a zero-forcing weight, i.e., the channels are orthogonal. Thus, the zero-forcing weight can maximize the output SNR.

The SNR in $y_k(t)$ is given by

$$\gamma_k = \lambda_k E[|s_k(t)|^2] / \sigma^2. \quad (6)$$

This indicates that the quality of each detected substream is different. Therefore, applying the adaptive resource (data rate and transmit power) assignment [2], the channel capacity can be improved.

Assuming time-variant fading channels, we must consider the time-interval τ between the transmit weight determination and transmission. If changes in the channel matrix during the interval are not negligible, the resource adaptation will not be optimum any more. In addition, the MRC weight at the receiver is not applicable since the channel orthogonality has been lost. To reduce effects of the channel information error due to time-variant fading, we propose two countermeasures.

First, the zero-forcing weight is applied to the receiver. This solves the inter-substream interference problem. However, SNR degradation due to the zero-forcing weight becomes considerable when the channel orthogonality is severely damaged. Secondly, therefore, we apply first order extrapolation of channel information. The concept is shown in Fig. 2. In the paper, it is assumed that the channel matrix is measured at the base station using ACK packet in the TDD reverse link. Then, the transmit weight determination and resource adaptation are carried out based on the predicted channel $\hat{a}(i_2 + \tau)$ from the last two successive channels $a(i_1)$ and $a(i_2)$ where

$$\hat{a}(i_2 + \tau) = a(i_2) + \tau(a(i_2) - a(i_1)) / (i_2 - i_1).$$

It is expected that the extrapolation reduces degradation in the channel orthogonality and the resource adaptation. Hereafter, we call the method using $\hat{a}(i_2 + \tau)$ as “extrapolated” scheme and one using $a(i_2)$ (the last channel) as “fixed” scheme.

III. Numerical analysis

Uncoded BER performance of E-SDM in the time-variant fading channel is numerically analyzed assuming a 4-tx 4-rx MIMO system where the total carried bit per symbol duration is 8. Thus, we prepared QPSK, 16QAM, 64QAM, and 256QAM, as modulation schemes. As the number of substreams, the appropriate modulation is selected. The frame duration ($i_2 - i_1$) is 2 ms (similar to HIPERLAN/2) and time interval τ is 1.6 ms. The packet size is 48 symbols (384 bits). For the performance comparison, conventional SDM and weighted SDM (or called as per antenna rate control [3] with antenna selection) are also evaluated. In E-SDM, joint detection (maximum likelihood detection) is also applied in addition to zero-forcing. In SDM and W-SDM, joint detection, zero-forcing, and V-BLAST using MMSE are applied.

Fig. 3 shows the BER performance of E-SDM, W-SDM, and SDM in a time-variant fading condition where the maximum Doppler shift (f_D) is 40 Hz ($f_D\tau = 0.064$). The abscissa axis represents transmit (TX) power normalized by the power which achieves E_s/N_0 of 0 dB when a single omni antenna is used. Therefore, it is equal to E_s/N_0 in SDM since no transmit beams are used. The BER performance of E-SDM with the zero-forcing (ZF) weight degrades about 1 dB at the BER of 10^{-4} from one with joint detection (JD). This means that loss of orthogonality due to channel changes yields SNR degradation in zero-forcing. In the following discussions, only zero-forcing is considered for E-SDM and W-SDM detectors.

Fig. 4 shows the BER performance with applying first order extrapolation to E-SDM and W-SDM where the maximum Doppler shift is 40 Hz. It should be noted that SDM needs no channel information at the transmitter. It is observed that about 1 dB and 2 dB improvements in E-SDM are obtained at the BER of 10^{-4} and 10^{-5} , respectively. From this result, it can be said that the first order extrapolation is very effective to compensate the channel state transition. Thus, the E-SDM with zero-forcing outperforms SDM with joint detection by about 3.5 dB at the BER of 10^{-4} . This improvement is still observable in a much fast fading case as shown in Fig. 5 where the maximum Doppler shift is 100 Hz ($f_D\tau = 0.16$). The performance of E-SDM with zero-forcing is almost the same as that of SDM with joint detection which requires very complicated processing compared to zero-forcing algorithm. Finally, let us evaluate required TX power for the BER of 10^{-4} versus the maximum Doppler frequency. A constant improvement by the first order extrapolation is kept until $f_D \simeq 50$ Hz.

IV. Conclusions

We have proposed the channel prediction based on the simple extrapolation, and indicated that the obtained improvement in E-SDM is valid in indoor WLAN environments with pedestrians even with very simple zero-forcing algorithm.

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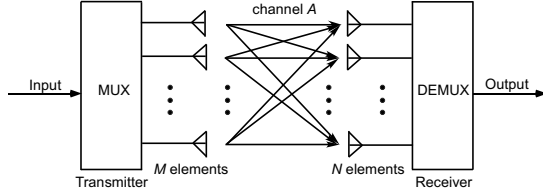


Fig. 1. A MIMO system model.

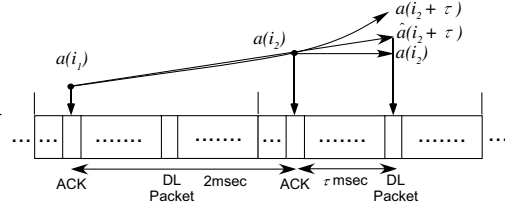


Fig. 2. A concept of 1st-order channel extrapolation.

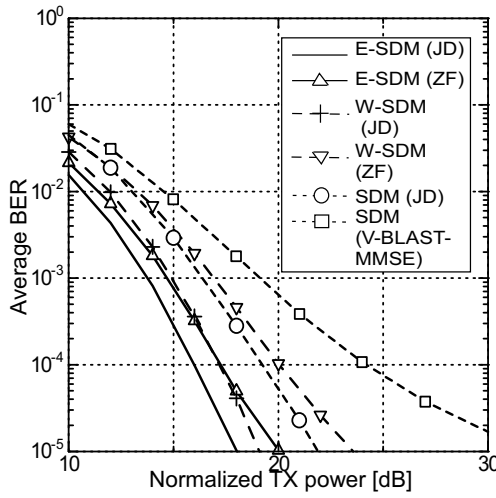


Fig. 3. BER performance (40 Hz).

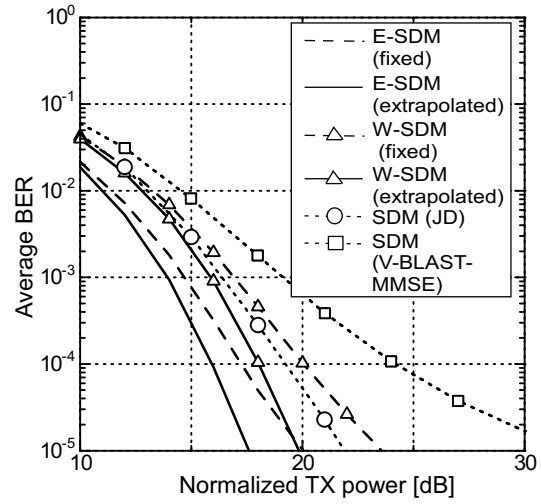


Fig. 4. BER performance with channel extrapolation (40 Hz).

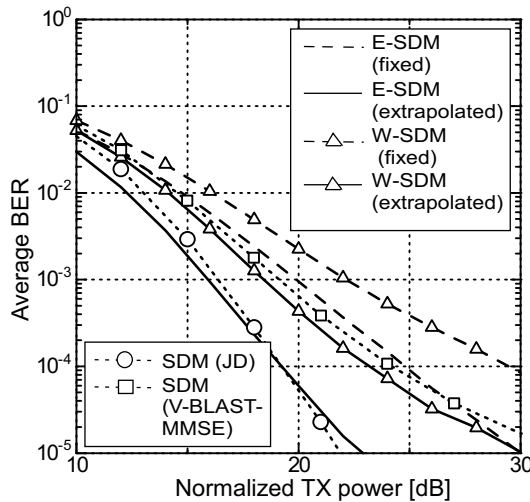


Fig. 5. BER performance with channel extrapolation (100 Hz).

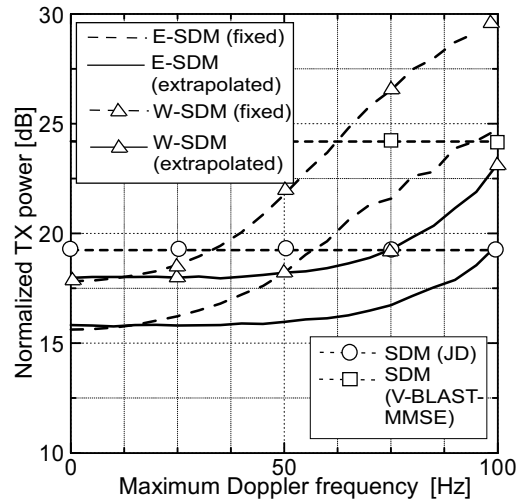


Fig. 6. Required TX power for BER of 10^{-4} .