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Novel downlink beamforming method using selective STBC with common eigenvectors for SDM-OFDM systems

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1. Introduction

Due to the growing popularity of broadband wireless LANs, high-speed transmission with a limited frequency band is required for future wireless communication systems. Orthogonal Frequency Division Multiplexing (OFDM) systems are well known for combating the effects of multi-path fading in broadband wireless LAN systems. To further improve the transmission rate, Space Division Multiplexing (SDM) [1] has been used in OFDM systems, e.g., in [2].

If the transmitter in the SDM scheme knows the channel status information (CSI), the channel capacity can be maximized by employing the Eigenbeam-Space Division Multiplexing (E-SDM) scheme [3]. The E-SDM scheme requires singular value decomposition (SVD) to obtain the array weight vector for transmission. Thus, transmitters are required to calculate the SVD for each sub-carrier to maximize the channel capacity in OFDM systems. This method, however, proportionally increases the calculation load according to the number of sub-carriers. Moreover, the transmission quality is considerably degraded when the CSI error increases.

To address these problems, this paper proposes a new downlink beamforming method that uses the common transmission weight vector derived for all sub-carriers and employs the transmission diversity. The transmission weight vector is obtained as an eigenvector of the matrix derived from the CSI of all sub-carriers and Space Time Block Cording (STBC) schemes such as in [4] is used in the eigenvector beams corresponding to second or subsequent eigenvalues to compensate for the transmission quality. Compared to the conventional E-SDM, the proposed method is robust against CSI error while it reduces the calculation complexity of the SVD.

2. Proposed Downlink Beamforming Method

Figure 1 shows a block diagram of the proposed beamforming system. The number of transmission antennas, reception antennas, sub-carriers, and beams are N, M, F, and L ($N \geqq L$, $N \geqq M$), respectively. Initially, the access point (AP) receives signals that are transmitted at the mobile station (MS) and perceived by both the AP and MS. The AP estimates the uplink channel matrix, $\mathbf{H}'_i$, where i is the sub-carrier index. Matrix $\mathbf{H}'_i$ is calculated using the following equation.

$$\widetilde{\mathbf{H}}'_i = \mathbf{X}_i \mathbf{S}_0^{-1}, \qquad \mathbf{X}_i = \mathbf{H}'_i \mathbf{S}_0 + \mathbf{N} \tag{1}$$

where $\widetilde{\mathbf{H}}'_i$ denotes the estimated channel matrix in the uplink, $\mathbf{N}$ denotes the $N\times 1$ vector comprising additive complex Gaussian noise with $\mathrm{CN}(0,\sigma)$, $\mathbf{X}_i$ of the $N\times M$ matrix denotes the received signals at the AP, and $\mathbf{S}_0$ of the $M\times M$ matrix represents the signals transmitted at the MS. The estimated channel matrix $\widetilde{\mathbf{H}}_i$ in the downlink is obtained as $\widetilde{\mathbf{H}}_i'^{T}$. The weight vector for each sub-carrier can be obtained as follows. The averaged correlation matrix, $\mathbf{R}_{\mathrm{all}}$, is calculated using the following equation.

$$\mathbf{R}_{all} = \frac{1}{F}\sum_{i=1}^{F}\mathbf{R}_i, \quad \mathbf{R}_i = \widetilde{\mathbf{H}}_i^H \widetilde{\mathbf{H}}_i \tag{2}$$

Subsequently, N eigenvalues and eigenvectors are obtained from $\mathbf{R}_{\mathrm{all}}$. Note that $\mathbf{R}_{\mathrm{all}}$ is averaged for all sub-carriers, it is not affected by the error of the estimated channel matrix, and extends the number of eigenvalues from M to N in the proposed method compared to using $\mathbf{R}_i$. Moreover, the calculation load significantly decreases. Next, we sequentially select L vectors from a large eigenvalue as transmission weight vectors ($\mathrm{W}_1,\cdots,\mathrm{W}_L$). Because the eigenvalue of $\mathbf{W}_1$ is the largest of all eigenvalues, the propagation channel of Beam #1, which uses transmission weight $\mathbf{W}_1$, has a relatively stable and high capacity. On the other hand, the channel capacities of the residual beams are unstable. Therefore, we apply STBC to such beams to compensate for the unstable and weak transmission quality. This scheme is called selective STBC in this paper.

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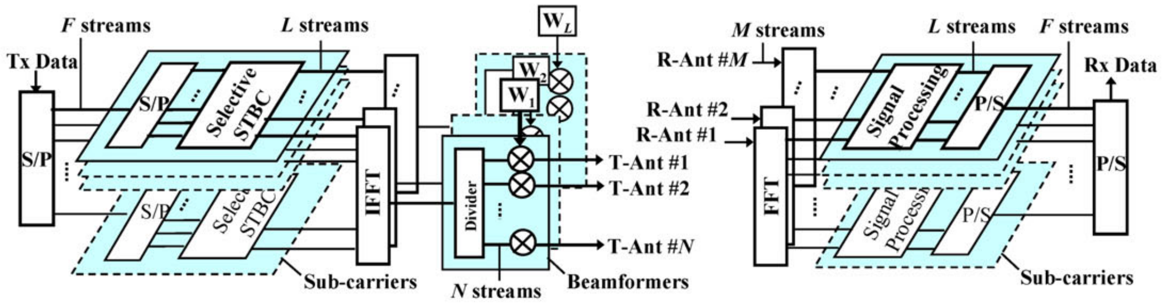

Figure 1. Block diagram of proposed downlink beamforming method

After the AP determines the weight using the above scheme, signals are transmitted using the following method. Input data are divided into F sub-streams and selective STBC is used for each sub-carrier. The Inverse Fast Fourier Transform (IFFT) is applied to the L output signals and these signals are sent to each transmission antenna after multiplying them by the transmission weight vector at each beamformer. The transmit antennas (#1 to #N) sum the L signals from the beamformers and up-convert them to Radio Frequency (RF) signals. At the MS, the received signals are down-converted to baseband signals and transformed to sub-carrier signals using the Fast Fourier Transform (FFT). The MMSE algorithm is used as the decoding algorithm.

To explain the principle using the equations of the proposed scheme (in detail), the proposed scheme is described in the following for the case of $N = 4$, $M = 2$, and $L = 3$. The second and third beams employ STBC, and four data symbols are transmitted in each block, which has the length of 2. Table 1 shows the transmission block. We then set the received signals, $\mathbf{X}^{(t)} = (X_1^{(t)}, X_2^{(t)})^{\mathrm{T}}$, where the subscript denotes receive antenna number (#1 or #2), at time t (1 or 2) for a certain sub-carrier as

$$\mathbf{X}^{(t)} = \mathbf{G}\mathbf{S}^{(t)} + \mathbf{N}^{(t)} \tag{3}$$

$$\mathbf{G} = \begin{pmatrix} G_{11} & G_{12} & G_{13} \\ G_{21} & G_{22} & G_{23} \end{pmatrix} = \begin{pmatrix} H_{11} & H_{12} & H_{13} & H_{14} \\ H_{21} & H_{22} & H_{23} & H_{24} \end{pmatrix} \begin{pmatrix} \mathbf{W}_1 & \mathbf{W}_2 & \mathbf{W}_3 \end{pmatrix} \tag{4}$$

Table 1. Transmission Block

	$t = 1$	$t = 2$
Beam #1	S_1	S_2*
Beam #2	S_3	-S_4*
Beam #3	S_4	S_3*

where $\mathbf{S}^{(1)}$ and $\mathbf{S}^{(2)}$ represent transmitted symbols as $(S_1, S_3, S_4)^{\mathrm{T}}$ and $(S_2{}^*, -S_4{}^*, S_3{}^*)^{\mathrm{T}}$, respectively, $\mathbf{W}_1$, $\mathbf{W}_2$, and $\mathbf{W}_3$ of the 4×1 vector are the transmission weight vectors in the proposed method, H_{ij} is the propagation coefficient between receive antenna #i and transmit antenna #j, and G_{ij} is the propagation coefficient between receive antenna #i and beam #j, and $\mathbf{N}^{(1)}$ and $\mathbf{N}^{(2)}$ denote the 2×1 vector comprising additive complex Gaussian noise. Equation (3) can be rewritten as

$$\mathbf{X} = \mathbf{G}'\mathbf{S} + \mathbf{N}, \tag{5}$$

where

$$\mathbf{X} = \begin{pmatrix} X_1^{(1)} \\ X_2^{(1)} \\ X_1^{(2)*} \\ X_2^{(2)*} \end{pmatrix}, \quad \mathbf{G}' = \begin{pmatrix} G_{11} & 0 & G_{12} & G_{13} \\ G_{21} & 0 & G_{22} & G_{23} \\ 0 & G_{11}^* & G_{13}^* & -G_{12}^* \\ 0 & G_{21}^* & G_{23}^* & -G_{22}^* \end{pmatrix}, \quad \mathbf{S} = \begin{pmatrix} S_1 \\ S_2 \\ S_3 \\ S_4 \end{pmatrix} \tag{6}$$

In order to obtain the transmitted symbols at the MS, $\mathbf{G}'$ must be estimated. We can acquire $\mathbf{G}'$ from $\mathbf{G}$, which is obtained using the same procedure described in Eq. (1). The estimated transmitted symbols, $\widetilde{\mathbf{S}}$, are given by

$$\widetilde{\mathbf{S}} = \mathbf{W}_R \mathbf{X}, \tag{7}$$

$$\mathbf{W}_R = \left(\mathrm{E}\left[\mathbf{X}\mathbf{X}^H\right]\right)^{-1} \mathbf{G}'^H, \tag{8}$$

where $\widetilde{\mathbf{S}}$ is the 4×1 estimated transmitted vector and E[] is the expectation operator. Through transmission diversity, the transmission quality levels of S_3 and S_4 are improved. Consequently, the transmission quality levels of S_1 and S_2 are also improved because S_3 and S_4 are removed more accurately from the received signals.

3. Simulation Conditions

We assume that the number of transmit antennas is four ($N = 4$), the number of receive antennas is two ($M = 2$), a linear array antenna is used for the transmit and receive antennas, and the element spacing is 0.5

wavelengths. We consider the SDM-OFDM system such that RF is 5.0 GHz, the bandwidth is 20 MHz and the number of sub-carriers is 51 ($F = 51$).

In our simulations, the CSI error in the uplink was considered. In general, the CSI error is caused by thermal noises, the change in CSI with time, and the calibration error. Although, the calibration error can be corrected by the perfect calibration method [5] and the change in the CSI during the period of a few frames can be neglected, the influence of thermal noise should be taken into account. This is because, in OFDM systems, the symbol length of each sub-carrier becomes long and long training symbols cannot be used to suppress the influence of the thermal noise in the channel estimation scheme. Thus, we assume that the CSI error level in the uplink is expressed as the variance, σ, in Eq. (1).

4. Performance of Proposed Method

We show that the proposed method suppresses the deterioration in the communication quality due to CSI error. It is assumed that the propagation environment has four incoming waves. The Angles of Departure (AoD) are 30˚, 120˚, and 210˚, respectively, and the power drops exponentially. The delay spread is 100 *nsec*.

The transmit directivity using conventional E-SDM is plotted in Fig. 2. This figure shows that the radiation pattern averaged over all sub-carriers where the variances of the CSI error, σ, are $-\infty$ dB (no error), -20 dB, and -10 dB, respectively. In the method, the AP can form only two beams based on the restriction that the number of receive antennas is two. When σ is $-\infty$ dB (no error), the directions of the main beams of Beam #1 and Beam #2 are 30˚ and 120˚, respectively. However, when σ is -20 dB, the power of Beam #2 in the direction of 120˚ decreases by 0.6 dB. Moreover, when σ is -10 dB, the power of beam #2 in the direction of 120˚ decreases by 3.0 dB, and the power of Beam #1 is also reduced by 0.3 dB. Figure 3 shows the radiation pattern when similarly applying the proposed method. The proposed method can form three beams, because there are three transmit antennas. Even for -10 dB error, the decrease in the power of Beam #2 in the direction of 120˚ is within 0.3 dB. In addition, the direction of the main Beam #3 approximately 210˚, although the direction shifts from 210˚ as the CSI error, σ, increases. This contributes to the transmit diversity.

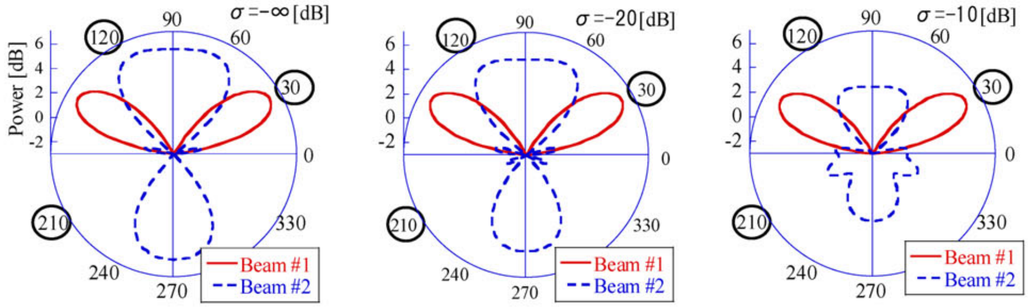

Figure 2. Radiation pattern with E-SDM in one sub-carrier

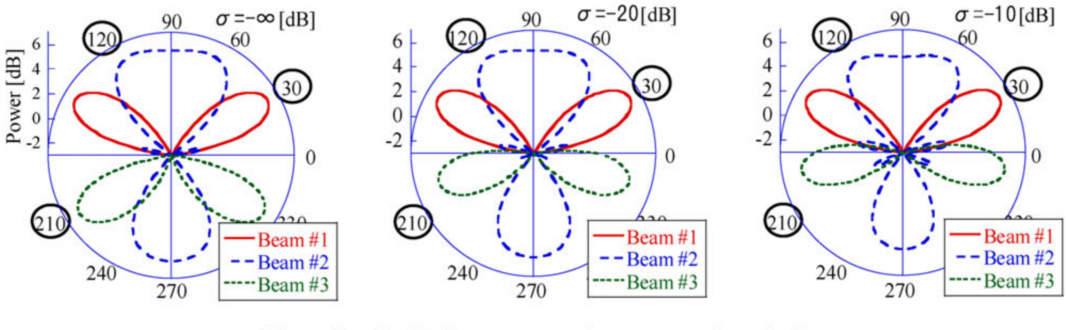

Figure 3. Radiation pattern using proposed method

5. Effect of Proposed Method

We evaluate the proposed method by computer simulation. We consider the indoor environment scattering model shown in Fig. 4 and adopt Ricean fading (the Ricean K-factor is 3 dB, the delay spread is 33 *nsec*) which is characterized by 100 reflections from five clusters and one line-of-sight path. Each cluster comprises scatterers and has the same angular spread of 25˚ with a Laplacian distribution. Beam #1 employs 64QAM modulation and Beam #2 and Beam #3 employ QPSK modulation with STBC. The power distribution for each beam is 2:1:1 to allocate equal power to each symbol.

We compare the proposed method to the conventional E-SDM method that uses the correlation matrix eigenvector of the channel matrix as the transmission weight vector for each sub-carrier. The conventional method can only form two beams due to the restriction of the number of receive antennas. We apply 64QAM modulation to Beam #1 and QPSK to Beam #2, and allocate equal power to each beam. To make a fair comparison, we introduce normalized transmit power where the average Carrier-to-Noise-power-ratio (CNR) of 30 dB is observed at the MS.

We simulated the Bit Error Rate (BER) that is average of all symbols in SDM scheme one thousand times. Each time, we changed the direction of the cluster and cluster's scatterers, and the phase of each path at random. Therefore, we obtained the cumulative probability of the BER. Fig. 5 shows the BER at the cumulative probability of 90% versus the CSI error. The figure shows that the performance of the conventional method decreases as the CSI error increases. The performance of the proposed method is better if the CSI error is greater than -18 dB.

Figure 6 shows the cumulative probability versus the BER when the CSI error is -10 dB. This figure indicates that the proposed method improves the cumulative probability for the BER of 10^{-3} from 35% to 75%.

Obviously, the modulation and the power allocation are not optimal; however, in the conventional method it is difficult to obtain the optimal parameter without knowing of the amount of error because the performance changes according to the error. In the proposed method it is easier to determine the optimal parameters.

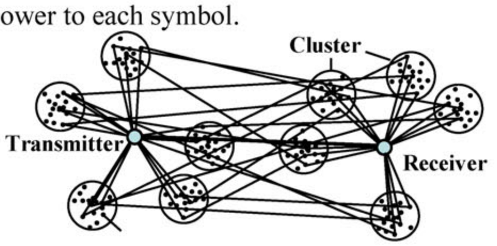

Figure 4. Model of propagation environment

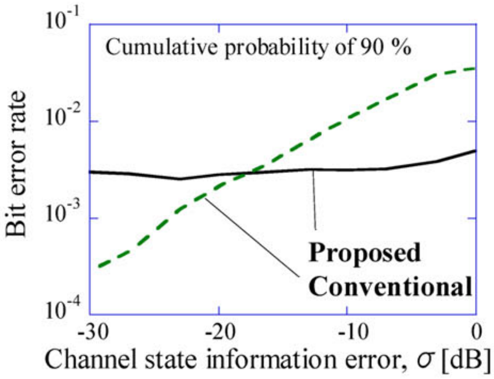

Figure 5. BER versus CSI error

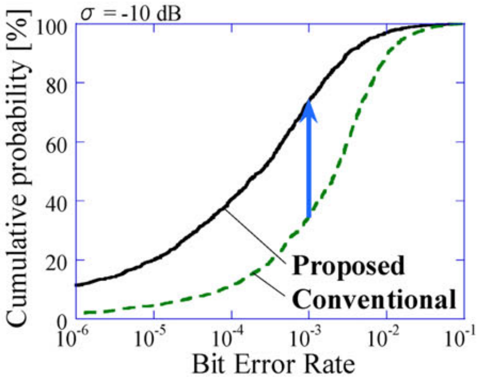

Figure 6. Cumlative probability

6. Conclusions

We proposed a new downlink beamforming method that uses the common transmission weight vector derived for all sub-carriers and that employs the selective STBC. Numerical simulations demonstrated that the transmission quality of the proposed method is robust against CSI error in the uplink. Despite the significant decrease in the calculation load, the performance of the proposed method is better than the conventional E-SDM method when the CSI error is greater than -18 dB. We clarified that when the CSI error is -10 dB, the proposed method improves the cumulative probability for the BER of 10^{-3} by 40%.

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