A NEW TRAINING SYMBOL STRUCTURE TO ENHANCE THE PERFORMANCE OF CHANNEL ESTIMATION FOR MIMO-OFDM SYSTEMS

Myeongcheol Shin, Hakju Lee, Chungyong Lee and Dae Hee Youn

School of Electrical & Electronic Engineering, Yonsei University, Seoul, Korea

ABSTRACT

In MIMO-OFDM system, a conventional channel estimation technique using comb type training symbol structure has high mean squared error(MSE) at edge subcarriers. To reduce the MSE at those subcarriers, we propose the cyclic comb type training structure and the weighted-averaging method. In the proposed technique, all kinds of comb type training symbols are transmitted cyclically at each transmitter. At the receiver, channel frequency responses estimated using each training symbol are averaged with weight which is obtained from the corresponding MSE. By computer simulations, we have shown that the proposed cyclic comb type training symbols gives higher SNR gain than the conventional training structure.

1. INTRODUCTION

OFDM(Orthogonal Frequency Division Multiplexing) has the ability to combat inter-symbol interference(ISI), a major problem in wideband transmission over multi-path fading channels. This has motivated the adoption of OFDM as a standard for broadband indoor wireless systems[1]. On the other hand, recent information theory has shown that multi-input multi-output (MIMO) systems can support enormous capacities[2], provided that the multi-path scattering of wireless channel is exploited. However, in frequency selective broadband channel, MIMO system requires complicated channel equalization technique to ensure eliminating inter-symbol interference. To reduce the impairments, it is a good solution to use OFDM technique for MIMO system, and recent works show that combining OFDM technique with multiple antenna architecture can provide high performance transmission[3].

In MIMO-OFDM system, a receiver should know the channel frequency responses of the spectral and spatial channels for coherent signal detection. For reliable channel estimation, almost channel estimators make use of training symbols. These training symbols are known data located in fixed subcarriers. Some studies has shown that the performance of channel estimator depends on these training symbol structures and channel estimation techniques[3]-[4]. Therefore, we propose the efficient training symbol structures and the corresponding channel estimation technique for MIMO-OFDM systems.

This research was supported by Samsung Advanced Institute of Technology(SAIT).

2. MIMO-OFDM SYSTEM MODEL

A simplified MIMO-OFDM system is shown in Fig.1. If the time-varying and frequency-selective fading channel is considered, the received signal vector $\mathbf{y}^q(n)$ at the q^{th} antenna can be expressed as a linear combination of channel impulse response and transmitted signals[3] :

$$\mathbf{y}^q(n) = \sum_{p=1}^{N_t} \mathbf{X}^P(n)\mathbf{F}_{[1:L]}\mathbf{h}_n^{p,q}(n) + \mathbf{w}^q(n) \ , \qquad (1)$$

where N_t , N_r mean the number of transmit and receive antennas, respectively. $\mathbf{X}^p(n)$ is diagonalized OFDM symbols at the p^{th} transmit antenna, and can be expressed as

$$\mathbf{X}^p(n) = diag([0 \cdots 0 \ x_1^p \cdots x_{N_\alpha}^p \ 0 \ x_{N_\alpha+1}^p \cdots x_{2N_\alpha}^p \ 0 \cdots 0]), \quad (2)$$

where N_α denotes the number of one sided available subcarriers, the total number of subcarriers for data transmission is $2N_\alpha$, the subscript denotes the data subcarrier index, and $diag(\cdot)$ denotes the diagonalize operator. In (1), $\mathbf{F}$ is the $K \times K$ Fourier transform matrix, and $\mathbf{F}_{[1:L]}$ is the first L columns of $\mathbf{F}$. $\mathbf{h}_n^{p,q}(n) = [h_n^{p,q}(0) \ h_n^{p,q}(1) \cdots h_n^{p,q}(L-1)]$ is the channel impulse response vector between the p^{th} and q^{th} antenna, and L denotes channel disperse length. $\mathbf{w}^q(n)$ is a noise vector at the q^{th} receive antenna with components that have IID(independent identically distributed) wide-sense stationary random processes with variance σ^2. Assume that there are V virtual subcarriers in the system and the subcarrier at DC (center subcarrier) is not used to avoid difficulties in D/A and A/D conversion[1].

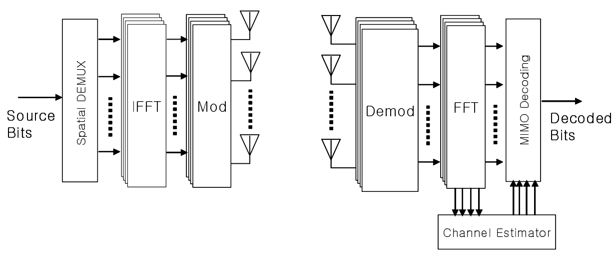

Fig. 1 A simplified MIMO-OFDM system

3. CHANNEL ESTIMATION OF MIMO-OFDM

For MIMO-OFDM system, the comb type training symbol is a well-known structure for the channel estimation[4]. It is assumed that the total number of subcarriers for data transmission is,

$$2N_{\alpha} = N_c N_t \ , \tag{3}$$

where N_c denotes the number of pilot subcarriers reserved for training symbols of each transmit antenna. N_c is assumed to be not smaller than the channel disperse length L . $\mathbf{X}^{p}_{comb}$, the matrix of comb type training symbols of p^{th} transmit antenna, has the elements x^{p}_{i} which satisfies

$$x^{p}_{i} = \begin{cases} c_i & i = (m-1)N_t + p \\ 0 & otherwise \end{cases} , \tag{4}$$

where c_i denotes an arbitrary complex number with magnitude $\sqrt{N_t}$ and m denotes an arbitrary positive integer not greater than N_c . In the comb type training structure, the predetermined training symbol vector for p^{th} antenna, $\mathbf{X}^{p}_{comb}$ is transmitted N_t times repeatedly, and can be expressed as

$$\begin{aligned} \mathbf{X}^{1}_{comb}(1) &= \mathbf{X}^{1}_{comb}(2) = \cdots = \mathbf{X}^{1}_{comb}(N_t) \\ \mathbf{X}^{2}_{comb}(1) &= \mathbf{X}^{2}_{comb}(2) = \cdots = \mathbf{X}^{2}_{comb}(N_t) \\ &\vdots \\ \mathbf{X}^{N_t}_{comb}(1) &= \mathbf{X}^{N_t}_{comb}(2) = \cdots = \mathbf{X}^{N_t}_{comb}(N_t) \ . \end{aligned} \tag{5}$$

At the receiver, each channel frequency response is estimated independently using multiple training symbols with the identical structure, and the channel frequency responses are averaged in time to reduce the noise variance. When using comb type training symbol, we can separate the training symbol transmitted at p^{th} antenna from the received signal vector $\mathbf{y}^{q}(n)$, because the comb type training symbol is orthogonal in the frequency domain. The received training symbol vector of q^{th} receive antenna at n^{th} OFDM block can then be modified from (1) as

$$\mathbf{y}^{p,q}(n) = \mathbf{X}^{p}_{comb}(n)\mathbf{F}_{[1:L]}\mathbf{h}^{p,q}_{n}(n) + \mathbf{w}^{q}(n) \ . \tag{6}$$

In this case, the channel impulse response can be obtained by least square estimation as[5]

$$\hat{\mathbf{h}}^{p,q}_{n}(n) = (\mathbf{X}^{p}_{comb}(n)\mathbf{F}^{p}_{[1:L]})^{\dagger}\mathbf{y}^{p,q}(n) \ . \tag{7}$$

where $(\cdot)^{\dagger}$ denotes the pseudo-inverse operator. The channel frequency response $\mathbf{h}^{p,q}_{f}(n)$ can then be obtained from (6) as

$$\begin{aligned} \hat{\mathbf{h}}^{p,q}_{f}(n) &= \dot{\mathbf{F}}_{[1:L]}\hat{\mathbf{h}}^{p,q}_{n}(n) \\ &= \mathbf{h}^{p,q}_{f}(n) + \dot{\mathbf{F}}_{[1:L]}(\mathbf{X}^{p}_{comb}(n)\mathbf{F}_{[1:L]})^{\dagger}\mathbf{w}^{q}(n) \\ &= \mathbf{h}^{p,q}_{f}(n) + \mathbf{e}^{p,q}(n) \ , \end{aligned} \tag{8}$$

where $\dot{\mathbf{F}}_{[1:L]}$ is the reduced Fourier transform matrix with $2N_{\alpha} \times L$ dimension, which has rows corresponding to data transmission carriers, $\mathbf{e}^{p,q}(n)$ denotes channel estimation error at n^{th} time block. The corresponding MSE of each data carrier can be expressed as

$$\boldsymbol{\sigma}^{2}_{p,q}(n) = \boldsymbol{\Psi}\left(\mathbf{E}\{\mathbf{e}^{p,q}(n)\mathbf{e}^{p,q}(n)^{H}\}\right), \tag{9}$$

where $\boldsymbol{\Psi}(\cdot)$ indicates a de-diagonalize operator which makes a vector from a matrix by taking the diagonal term of the matrix, and $\mathbf{E}\{\cdot\}$ denotes an expectation operator.

The estimated channel frequency responses are averaged in time domain under the assumption that channel is invariant during training phase. The averaged channel frequency response can be expressed as

$$\bar{\mathbf{h}}^{p,q}_{f} = \frac{1}{N_t}\sum_{n=1}^{N_t}\hat{\mathbf{h}}^{p,q}_{f}(n). \tag{10}$$

After averaging, the MSE is reduced as

$$\boldsymbol{\sigma}^{2}_{p,q} = \frac{1}{N_t}\boldsymbol{\sigma}^{2}_{p,q}(n) \ . \tag{11}$$

The overall normalized MSE can then be obtained as

$$NMSE = \frac{1}{2N_{\alpha}N_t N_r}\sum_{p=1}^{N_t}\sum_{q=1}^{N_r} trace(diag(\boldsymbol{\sigma}^{2}_{p,q})) \ , \tag{12}$$

where $trace(\cdot)$ indicates a trace operator.

Fig.2 shows the MSE of each data carriers using the conventional training strucure. In this figure, 4 transmit antennas are used in MIMO-OFDM system which has 64 subcarriers including 52 data subcarriers, 11 virtual subcarriers and DC subcarrier. Fig.2(a) shows the MSE of channel estimation using the comb type training symbols transmitted at 1st antenna $\mathbf{X}^{1}_{comb}$, and Fig.2(b),(c),and (d) give the MSEs using the conventional training symbols at second, third, and fourth antenna, respectively. Fig.2 clearly shows that the MSEs of the edge subcarriers are relatively large. The origin of this is a lack of training symbols to interpolate the edge subcarriers, and thus the channel estimation error is increased[5].

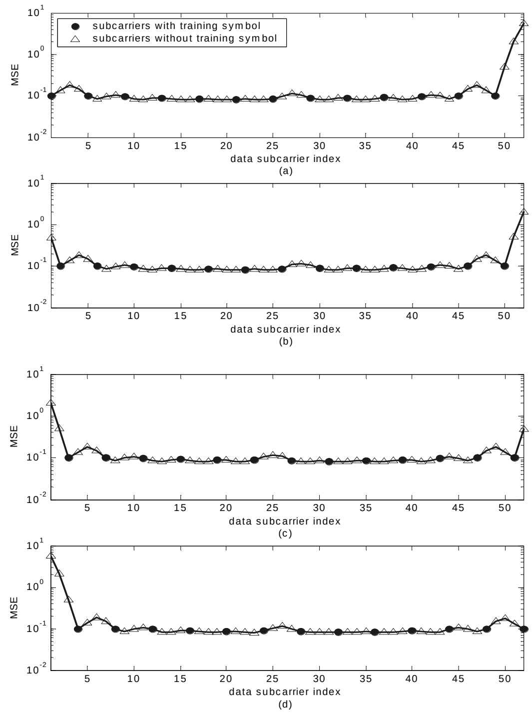

Fig. 2 The MSEs of the conventional channel estimation

4. THE PROPOSED TRAINING SYMBOL STRUCTURE & CORRESPONDING CHANNEL ESTIMATION TECHNIQUE

In conventional training structure, the MSEs of estiamated channel frequency reponses using the identical training symbols which are transmitted N_t times are equal, so that the MSEs of the averaged channel frequency response shows the same pattern as Fig.2. As a result, it is not a good approach that each transmitter sends the same training symbol repeatedly and the estimated channel frequency responses are averaged in time.

Therefore, we propose a new training symbol structure, so called "the cyclic comb type training structure". In the cyclic comb type training structure, a comb type training symbol which was transmitted through the 1st antenna is transmitted through the 2nd antenna, and will be transmitted through the 3rd antenna the next time, and so on. Using this cyclical scheme, all types of comb type training symbols are transmitted at each antenna. The cyclic comb type training structure can be expressed as

$$\begin{aligned} &\mathbf{X}^1_{comb}(1) = \mathbf{X}^2_{comb}(2) = \cdots = \mathbf{X}^{N_t}_{comb}(N_t) \\ &\mathbf{X}^2_{comb}(1) = \mathbf{X}^3_{comb}(2) = \cdots = \mathbf{X}^1_{comb}(N_t) \\ &\qquad\vdots \\ &\mathbf{X}^{N_t}_{comb}(1) = \mathbf{X}^1_{comb}(2) = \cdots = \mathbf{X}^{N_t-1}_{comb}(N_t)\ . \end{aligned} \tag{13}$$

In this case, the estimated channel frequency response $\hat{\mathbf{h}}^{p,q}_f(n)$ can be also obtained as (8). However, the estimated channel frequency responses should not be averaged in the same manner as (10), because the channel estimation error variance $\boldsymbol{\sigma}^2_{p,q}(n)$ is not the same with respect to time index n. The channel frequency responses should be weighted according to the estimation error variance before averaging to ensure that the averaged channel frequency response has minimum error variance, which will be called hereinafter "weighted-averaging". The weight averaged channel frequency response can be expressed as

$$\begin{aligned} \hat{\mathbf{h}}^{p,q}_f &= (\sum_{n=1}^{N_t} \mathbf{C}^{p,q}(n))^{-1} \sum_{n=1}^{N_t} \mathbf{C}^{p,q}(n) \hat{\mathbf{h}}^{p,q}_f(n) \\ &= \mathbf{h}^{p,q}_f + \left\{ \sum_{n=1}^{N_t} \mathbf{C}^{p,q}(n) \right\}^{-1} \sum_{n=1}^{N_t} \mathbf{C}^{p,q}(n) \mathbf{e}^{p,q}(n) \\ &= \mathbf{h}^{p,q}_f + \boldsymbol{\Omega}^{p,q}\ , \end{aligned} \tag{14}$$

where $\mathbf{C}^{p,q}(n)$ is the diagonal weight matrix to minimize the averaged estimation error variance, and $\boldsymbol{\Omega}^{p,q}$ denotes channel estimation error between p^{th} and q^{th} antennas after weighted-averaging. The diagonal weight matrix $\mathbf{C}^{p,q}(n)$ can be obtained by solving the minimization proble :

$$\min_{\mathbf{C}^{p,q}} \left\{ \mathbf{E}\{ \boldsymbol{\Omega}^{p,q^H} \boldsymbol{\Omega}^{p,q} \} \right\}\ . \tag{15}$$

The diagonal weight matrix to minimize (15) can be obtained as

$$\mathbf{C}^{p,q}(n) = (diag(\boldsymbol{\sigma}^2_{p,q}(n)))^{-1}\ . \tag{16}$$

The corresponding minimized MSE of each subcarrier is expressed as

$$\boldsymbol{\sigma}^2_{p,q} = \boldsymbol{\Psi}\left(\left(\sum_{n=1}^{N_t} (diag(\boldsymbol{\sigma}^2_{p,q}(n)))^{-1} \right)^{-1} \right)\ , \tag{17}$$

and the normalized MSE using the proposed cyclic training structure can be expressed as

$$NMSE = \frac{1}{2N_\alpha N_t N_r} \sum_{p=1}^{N_t} \sum_{q=1}^{N_r} trace(diag(\boldsymbol{\sigma}^2_{p,q})). \tag{18}$$

Fig.3 shows the MSE of channel estimation using the conventional and cyclic comb type training symbols transmitted at 1st transmit antenna for example. The conventional training structure shows a similar pattern as the MSE of Fig.2: the channel estimation error at both edges are large and these dominant channel estimation errors lead to an increase in the overall NMSE. On the contrary, in the proposed cyclic training structure, the channel estimation errors of each data carrier are relatively small, because the channel frequency responses are obtained using the cyclic training structure and weighted-averaging. As a result, the proposed cyclic training structure performs better for channel estimation than the conventional training structure.

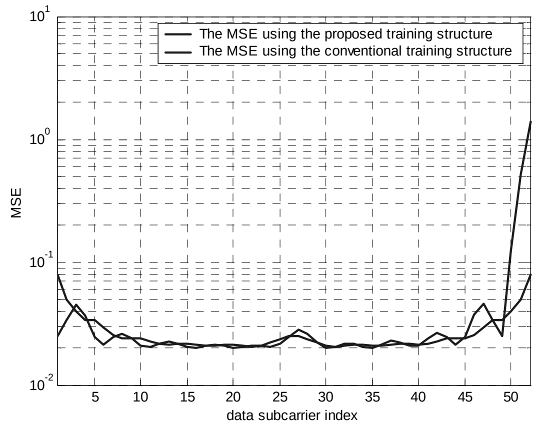

Fig. 3 The comparison of the MSEs of the conventional and the proposed training symbol structure

5. SIMULATION RESULTS

The target MIMO-OFDM system was based on IEEE 802.11a system which has 64 subcarriers including 52 data subcarriers, 11 virtual subcarriers at edge band and DC subcarrier. Only 12 Mbps rate which uses QPSK was considered. As a MIMO configuration, 2×2 and 4×4 antenna systems were assumed. For the channel estimation, training symbols are transmitted 2 times in preamble field for a 2×2 system and 4 times for a 4×4 system.

Fig.4 shows the NMSE of channel estimation for the 2×2 system. It indicates that the cyclic training symbol structure has 0.2dB SNR gain in NMSE over the conventional training structure. Fig.5 shows the NMSE under the 4×4 system. We can observe that the cyclic training structure has 2.5dB SNR gain over the conventional training structure. In the conventional training structure case, the number of the subcarriers for training symbols in the 4×4 system is less than that in the 2×2 system, so that the interpolation error in the 4×4 system becomes larger than the 2×2 system. However, in the proposed cyclic training structure, it is possible to compensate this interpolation error effectively, so that the SNR gain of the cyclic training structure in the 4×4 antenna system is much higher than that in the 2×2 antenna system. According to these results, it would be predicted that the larger the number of transmit antennas is, the more SNR gain is available.

Fig.6 shows bit error rate(BER) performance in the 4×4 antenna system for both structures. This figure indicates that the cyclic training structure has 1.3dB SNR gain at BER=10^{-3} over the conventional training structure. In the conventional training structure, it is the large channel estimation errors at the edge subcarriers that affects the overall BER to be increased. Thus, the proposed cyclic training structure designed to minimize estimation error at these subcarriers are effective in improving the detection performance.

6. CONCLUSIONS

In this paper, we proposed the cyclic comb type training structure and the weighted-averaging technique to enhance the channel estimation performance of MIMO-OFDM system. By computer simulations, we have shown that the proposed cyclic comb type training symbols gives higher SNR gain than the conventional training structure. Consequently, the proposed cyclic training structure and the weighted-averaging method can enhance the channel estimation performance of MIMO-OFDM systems.

7. REFERENCES

[1] IEEE 802.11a, “Wireless LAN Medium Access control (MAC) and Physical Layer (PHY) specifications : High speed physical layer in the 5GHz band,” *IEEE Standard*, 1999

[2] G. Foschini, “Layered space-time architecture for wireless communication in a fading environments using multi-element antennas,” *Bell Labs Technical Journal*, vol.1, pp.41-59, Aut. 1996.

[3] I. Barhumi, G. Leus and M. Moonen, “Optimal training sequences for channel estimation in MIMO OFDM systems in mobile wireless channels,” in *Int. Zurich Symp. on Broadband Communications*, pp. 44-1~ 44-6. Feb. 2002.

[4] W. G. Jeon, K. H. Paik and Y.S. Cho, “An efficient channel estimation techniques for OFDM system with transmitter diversity,” in *IEEE Int. Symp. on Personal, Indoor and Mobile Radio Communications*, 2000, vol.2, pp. 1246-1250.

[5] Michele Morelli and Umberto Mengali, “A comparison of pilot-aided channel estimation methods for OFDM systems,” *IEEE Trans. on Signal Processing*, vol.49, no.12, pp.3065-3073, Dec. 2001.

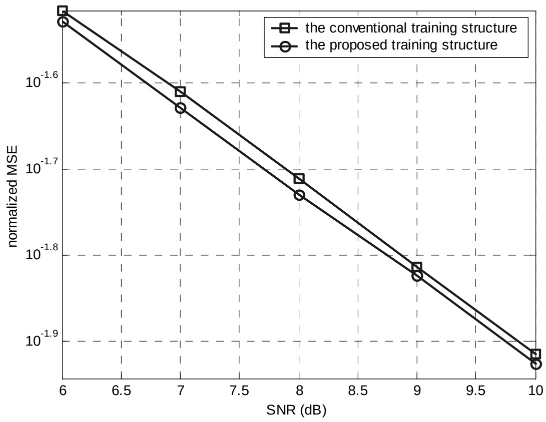

Fig. 4 The comparison of NMSEs in the 2×2 system

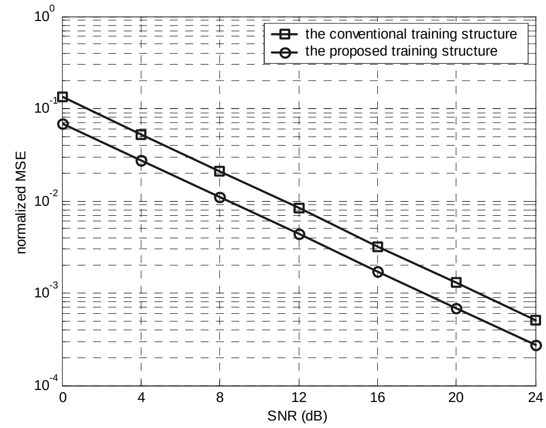

Fig. 5 The comparison of NMSEs in the 4×4 system

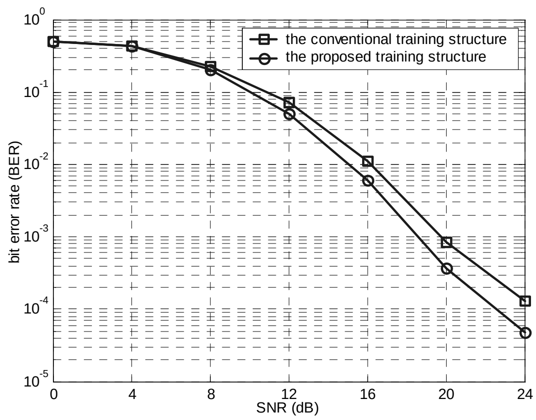

Fig. 6 The comparison of bit error rate in the 4×4 system