A Composite LMMSE Channel Estimator for Spectrum-Efficient OFDM Transmit Diversity

Jeong-Wook Seo[1], ***Non-Member***, **Won-Gi Jeon**[2], ***Non-Member***, **Jong-Ho Paik**[2], ***Member, KSII***, **Minho Jo**[3], ***Member, KSII***, **and Dong-Ku Kim**[1], ***Non-Member***
[1] School of Electrical and Electronic Eng., Yonsei University,
134 Sinchon-dong, Seodaemun-gu, Seoul 120-749, Korea
[e-mail: jwseo@keti.re.kr, dkkim@yonsei.ac.kr]
[2]Advanced Mobile Technology Research Center, Korea Electronics Technology Institute,
1599 Sangam-dong, Mapo-gu, Seoul 121-835, Korea
[e-mail: jeonwg@keti.re.kr, paikjh@keti.re.kr]
[3]School of Information and Communication Eng., Sungkyunkwan University,
300 Cheoncheon-dong, Jangan-gu, Suwon, Gyeonggi-do 440-746, Korea
*Corresponding author: Dong-Ku Kim

Received June 17, 2008; revised August 10, 2008; accepted August 12, 2008; published August 25, 2008

Abstract

In this paper, we propose a subcarrier allocation method and a composite linear minimum mean square error (LMMSE) channel estimator to increase spectrum efficiency in orthogonal frequency division multiplexing (OFDM) transmit diversity. The pilot symbols for OFDM transmit (Alamouti) diversity are exclusively allocated in two OFDM symbols in different antennas, which causes serious degradation of spectrum efficiency. To reduce the number of pilot symbols, our subcarrier allocation method uses repetition-coded data symbols, and the proposed channel estimator maintains good bit error rate (BER) performance.

Keywords: OFDM, transmit diversity, subcarrier allocation, LMMSE channel estimation, maximum ratio combining, normalized effective throughput

This research was supported by the KETI and the MKE (Ministry of Knowledge Economy), South Korea. Parts of this paper were presented at IEEE IWCMC 2008, Crete Island, Greece, August 6, 2008.

DOI: 10.3837/tiis.2008.04.004

1. Introduction

Orthogonal frequency division multiplexing (OFDM) has been applied to various wireless communication systems, since it can mitigate the severe effects of frequency selective fading and provide good spectrum efficiency [1][2]. Recently, there have been many attempts to apply Alamouti's transmit diversity to OFDM systems such as the IEEE 802.16™ WiMAX [3][4]. In addition, many channel estimation methods for OFDM transmit diversity have been studied [5][6]. Although, in terms of practical implementation, pilot symbols should be exclusively allocated in two OFDM symbols in different antennas, this causes serious degradation of spectrum efficiency, due to the number of pilot symbols required to estimate the channel frequency response (CFR) and the number of null symbols required to maintain orthogonality between the pilot symbols. On the other hand, reducing the number of pilot symbols to increase spectrum efficiency, results in severe error flooring, even though an optimal linear minimum mean square error (LMMSE) channel estimator is used. Therefore, it is necessary to design a new subcarrier allocation method and a robust channel estimator, for achieving good spectrum efficiency in OFDM transmit diversity.

The proposed subcarrier allocation method exploits repetition-coded data symbols, instead of corresponding pilot symbols. In order to maintain good bit error rate (BER) performance, the proposed channel estimator consists of a coarse LMMSE estimator, a maximum ratio combining-based decision directed (MRC-DD) estimator, and a fine LMMSE estimator. The coarse LMMSE estimator provides channel frequency response (CFR) estimates at the repetition-coded data locations for the following MRC-DD estimator. Then, the MRC-DD estimator provides more accurate CFR estimates at the same locations. Finally, the find LMMSE estimator provides CFR estimates at the Alamouti-coded data locations by using all CFR estimates from the coarse LMMSE estimator and the MRC-DD estimator. This paper is organized as follows. Section 2 introduces an OFDM system model involving transmit diversity, and our subcarrier allocation method. Section 3 describes the proposed composite LMMSE channel estimator. Section 4 provides simulation results to evaluate the proposed scheme. This paper is concluded in Section 5.

2. Spectrum-Efficient OFDM Transmit Diversity

This section describes OFDM transmit diversity with the proposed subcarrier allocation method. As shown in **Fig. 1**, the original data symbols are encoded and transmitted by Alamouti's method. The pilot and null symbols are allocated, with equal spacing in each OFDM symbol. Especially, extra-data symbols are repeated and interleaved to obtain diversity gain, then inserted between the pilot symbols. In the receiver, both the pilot symbols and the extra-data symbols are used for channel estimation. The detailed explanation is as follows.

In time division duplex (TDD) mode, each frame is partitioned into consecutive time slots, where each slot contains one data vector $\mathbf{S}_l$, $l = 0,1$ expressed in the frequency domain as

$$\mathbf{S}_l = \left[S_l(0), S_l(1), \cdots, S_l(N_d - 1) \right]^T \tag{1}$$

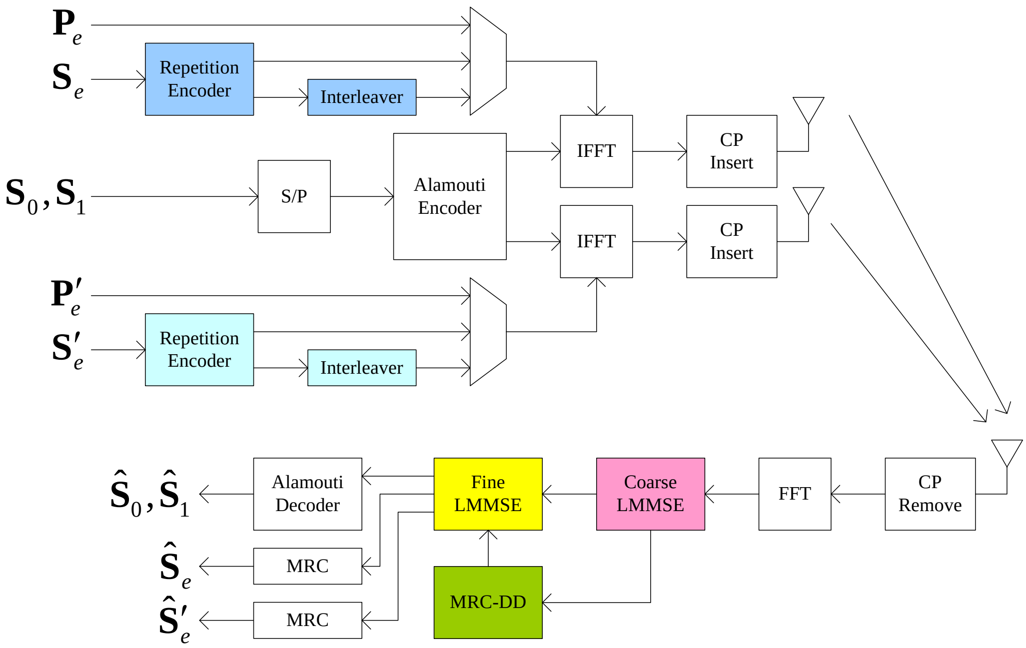

Fig. 1. OFDM transmit diversity with proposed subcarrier allocation and channel estimation

where N_d is the number of data symbols or subcarriers, and $S_l(n)$ is the *n*th modulated data symbol with the *M*-ary QAM constellation. Using Alamouti's encoding method, two consecutive data vectors $\left[\mathbf{S}_0, -\mathbf{S}_1^*\right]$ and $\left[\mathbf{S}_1, \mathbf{S}_0^*\right]$ are generated, and mapped to transmitted vector $\mathbf{X}_l^1$ and $\mathbf{X}_l^2$ for the 1st and 2nd antennas, respectively. The transmitted vector $\mathbf{X}_l^\alpha$, $\alpha = 1, 2$ is expressed as

$$\mathbf{X}_l^\alpha = \left[X_l^\alpha(0), X_l^\alpha(1), \cdots, X_l^\alpha(N-1)\right]^T \tag{2}$$

where N is the total number of subcarriers. The mapping rules of the 1st antenna are expressed as

$$\left[\mathbf{X}_0^1\right]_{j_n} = S_0(n) \text{ and } \left[\mathbf{X}_1^1\right]_{j_n} = -S_1^*(n) \tag{3}$$

where $\{j_n \,|\, 0 \le n \le N_d - 1\}$ denotes data locations. For the 2nd antenna, the mapping rules are expressed as

$$\left[\mathbf{X}_0^2\right]_{j_n} = S_1(n) \text{ and } \left[\mathbf{X}_1^2\right]_{j_n} = S_0^*(n). \tag{4}$$

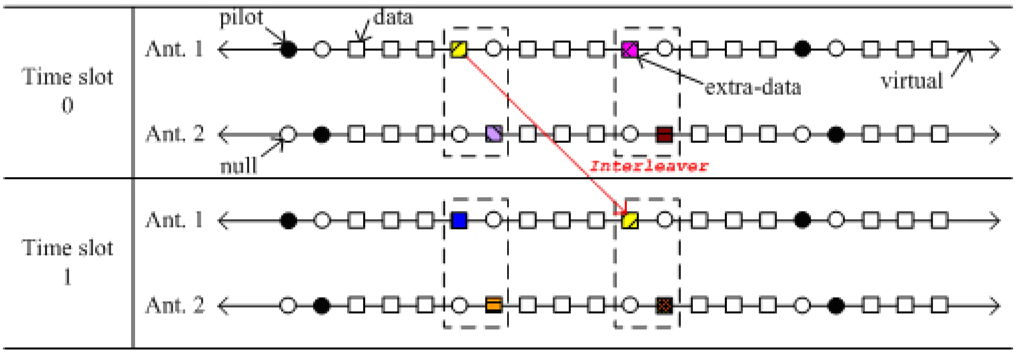

Fig. 2. Proposed subcarrier allocation method: pilot symbols and repetition coded extra-data symbols

The number of pilot symbols is related to the minimum pilot spacing D_f bounded by the frequency-domain sampling theorem which is expressed as

$$D_f \leq NT / 2\tau_{\max} \tag{5}$$

where $\tau_{\max}$ is the maximum excess channel delay, and T is the sampling time. As D_f was decreased, the number of pilot symbols increased, according to Eq. (6)

$$N_p = \left\lfloor N / D_f \right\rfloor \tag{6}$$

where $\lfloor \cdot \rfloor$ denotes the floor function. In addition, due to the increased number of pilot symbols, the mean square error (MSE) performance of channel estimation was improved, while spectrum efficiency was degraded. In the proposed subcarrier allocation method, we selected D_f such that it was as close to the bound in Eq. (3) as possible, and inserted extra-data symbols between pilot symbols as shown in **Fig. 2**. Moreover, repetition coding and interleaving were applied to the extra-data symbols in order to enhance the reliability of the regenerated symbols for the MRC-DD estimator.

The pilot and null symbols in the 1st antenna were mapped as

$$\left[\mathbf{X}_l^1\right]_{i_n} = P_e(n) \text{ for } \{i_n \mid 0 \leq n \leq N_p - 1\} \tag{7}$$

$$\left[\mathbf{X}_l^1\right]_{k_n} = 0 \text{ for } \{k_n \mid 0 \leq n \leq N_z - 1\} \tag{8}$$

where i_n and k_n denote pilot locations and null locations, respectively. N_p and N_z denote the number of pilot subcarriers and the number of null subcarriers, respectively. The extra-data symbols in the 0th time slot and the 1st time slot were mapped as

$$\left[\mathbf{X}_0^1\right]_{u_n} = S_e\left(n\right) \text{ for } \left\{u_n \mid 0 \le n \le N_e - 1\right\} \tag{9}$$

$$\left[\mathbf{X}_1^1\right]_{\delta_n} = S_e\left(n\right) \text{ for } \left\{\delta_n \mid 0 \le n \le N_e - 1\right\} \tag{10}$$

where N_e denotes the number of extra-data subcarriers, and δ_n denotes the interleaved extra-data locations. In other words, $\delta_n = \Pi\left(u_n\right)$ and $u_n = \Pi^{-1}\left(\delta_n\right)$ where $\Pi\left(\cdot\right)$ denotes the interleaving function. The mapping rules of the pilot symbols, the null symbols, and the extra-data symbols in the 2nd antenna differ from the previous ones, such as

$$\left[\mathbf{X}_l^2\right]_{i'_n} = P'_e\left(n\right) \text{ for } \left\{i'_n \mid 0 \le n \le N'_p - 1\right\} \tag{11}$$

$$\left[\mathbf{X}_l^2\right]_{k'_n} = 0 \text{ for } \left\{k'_n \mid 0 \le n \le N'_z - 1\right\} \tag{12}$$

$$\left[\mathbf{X}_0^2\right]_{u'_n} = S'_e\left(n\right) \text{ for } \left\{u'_n \mid 0 \le n \le N'_e - 1\right\} \tag{13}$$

$$\left[\mathbf{X}_1^2\right]_{\delta'_n} = S'_e\left(n\right) \text{ for } \left\{\delta'_n \mid 0 \le n \le N'_e - 1\right\} \tag{14}$$

where N'_p denotes the number of pilot subcarriers, N'_z denots the number of null subcarriers, N'_e denotes the number of extra-data subcarriers, and δ'_n denotes the interleaved extra-data locations where $\delta'_n = \Pi\left(u'_n\right)$ and $u'_n = \Pi^{-1}\left(\delta'_n\right)$.

The transmitters in the two antennas perform an inverse discrete Fourier transform (IDFT) operation $\mathbf{D}^H$ on the transmitted vectors. Then, a cyclic prefix (CP) of length N_{cp} is added in order to mitigate inter-symbol interference (ISI) due to the multipath fading channels. H denotes the Hermitian tranpose, and $\mathbf{D}$ denotes an $N \times N$ DFT matrix with entries which are expressed as

$$\left[\mathbf{D}\right]_{m,n} = \frac{1}{\sqrt{N}} e^{-j2\pi mn/N} . \tag{15}$$

Assuming that there is perfect synchronization, quasi-stationary multipath fading during consecutive two OFDM symbols, and the CP length is longer than the maximum excess delays of both channels, the received vector, after removing CP and completing the DFT operation, is expressed as

$$\mathbf{Y}_l = \mathbf{H}_l^1 \mathbf{X}_l^1 + \mathbf{H}_l^2 \mathbf{X}_l^2 + \mathbf{W}_l \tag{16}$$

where $\mathbf{H}_l^\alpha$, $\alpha = 0,1$ denote the channel matrices from the 1st and 2nd antennas, and $\mathbf{W}_l$ denotes the additive zero mean circularly symmetric complex Gaussian noise vector with covariance matrix $N_0\mathbf{I}_N$. Here, $\mathbf{I}_N$ denotes an $N \times N$ identity matrix. $\mathbf{H}_l^\alpha$ can be expressed as

$$\mathbf{H}_l^{\alpha} = diag\left(H_l^{\alpha}(0), H_l^{\alpha}(1), \cdots, H_l^{\alpha}(N-1)\right) \tag{17}$$

where $H_l^{\alpha}(k) = \sum_{n=0}^{L_\alpha - 1} h_l^{\alpha}(n) e^{-j2\pi kn/N}$, L_α is the number of channel paths, the nth non-zero complex channel gain $h_l^{\alpha}(n)$ is wide-sense stationary uncorrelated scattering (WSSUS) process with the Jakes' power spectrum, and $diag(\mathbf{x})$ is a diagonal matrix with $\mathbf{x}$ on its diagonal.

3. Proposed Channel Estimation

If the number of pilot symbols is greatly reduced, a conventional LMMSE channel estimator may cause severe error flooring and fail to maintain the desired service quality. In order to solve this problem, a composite LMMSE channel estimator is proposed, which consists of a coarse LMMSE estimator, an MRC-DD estimator, and a fine LMMSE estimator. Although a pilot-symbol-aided (PSA) channel estimator with decision directed estimation was previously proposed in [7][8], it does not provide satisfactory performance, since severe decision errors due to deep fading results in the use of unreliable data symbols in least square (LS) estimation. In our channel estimator, the coarse LMMSE estimator was utilized first to estimate CFRs at repetition-coded data locations. Subsequently, the MRC-DD estimator was utilized to obtain more accurate CFRs at the same locations. Finally, the fine LMMSE estimator was utilized to estimate the CFRs needed for Alamouti decoding.

In order to estimate the $\mathbf{H}_l^1$ channel matrix, the coarse LMMSE estimator and the MRC-DD estimator were utilized as follows. Firstly, the coarse LMMSE estimator was utilized and expessed as

$$\tilde{H}_l^1(u_m) = \sum_{n=0}^{N_p-1} Q_{lmmse}^1(u_m, i_n)\hat{H}_l^1(i_n) \tag{18}$$

where $\hat{H}_l^1(i_n)$ are the LS estimates at pilot locations, and $Q_{lmmse}^1(m,n)$ are the LMMSE coefficients, which are expressed as

$$\hat{H}_l^1(i_n) = \frac{Y_l(i_n)}{P_e(n)} \tag{19}$$

$$Q_{lmmse}^1(m,n) = \left[\mathbf{R}_{\mathrm{H}_l^1\hat{\mathrm{H}}_l^1}\mathbf{R}_{\hat{\mathrm{H}}_l^1\hat{\mathrm{H}}_l^1}^{-1}\right]_{m,n} \tag{20}$$

where $\mathbf{R}_{\mathrm{H}_l^1\hat{\mathrm{H}}_l^1} = E\left(\mathrm{H}_l^1\hat{\mathrm{H}}_l^{1H}\right)$ is the cross-correlation matrix of the desired channel vector H_l^1 and the estimated channel vector $\hat{\mathrm{H}}_l^1$, and $\mathbf{R}_{\hat{\mathrm{H}}_l^1\hat{\mathrm{H}}_l^1} = E\left(\hat{\mathrm{H}}_l^1\hat{\mathrm{H}}_l^{1H}\right)$ is the auto-correlation matrix of the estimated channel vector. Here, $E(\cdot)$ denotes the expectation function. Note that the cross- and auto-correlation matrices can be pre-computed and stored, if we assume a uniform channel power delay profile in [9]. In other words, if the indices of the pilot and the extra-data subcarriers are given, any LMMSE coefficients can be obtained automatically. Secondly, the MRC-DD channel estimator was utilized as follows. The MRC operation is expressed as

$$\hat{X}_l^1(u_m) = \sum_{l=0}^{1} G_l^1(u_m) Y_l(u_m) \tag{21}$$

where the weight coefficients are given by

$$G_l^1(u_m) = \frac{\tilde{H}_l^{1*}(u_m)}{\sqrt{\sum_{l=0}^{1} \left| \tilde{H}_l^1(u_m) \right|^2}}. \tag{22}$$

The hard-decision function $f_D(\cdot)$ generated data symbols as follows

$$\bar{X}_l^1(u_m) = f_D\left(\hat{X}_l^1(u_m)\right). \tag{23}$$

The LS estimation provided the CFR as follows

$$\bar{H}_l^1(u_m) = \frac{Y_l(u_m)}{\bar{X}_l^1(u_m)}. \tag{24}$$

Finally, the fine LMMSE estimator was utilized as follows

$$\hat{H}_l^1(j_n) = \sum_{m=0}^{N_p-1} Q_{lmmse}^1(j_n, i_m) \hat{H}_l^1(i_m) + \sum_{m=0}^{N_e-1} Q_{lmmse}^1(j_n, u_m) \bar{H}_l^1(u_m). \tag{25}$$

Since the $\mathbf{H}_l^2$ channel matrix can be estimated by the aforementioned method, we do not describe the estimation procedure.

Alamouti decoding was performed with the following parameters:

$$X_0^1(j_n) = S_0(n) \text{ and } X_0^2(j_n) = S_1(n) \tag{26}$$

$$X_1^1(j_n) = -S_1^*(n) \text{ and } X_1^2(j_n) = S_0^*(n) \tag{27}$$

$$H_1(j_n) = H_0^1(j_n) = H_1^1(j_n) \tag{28}$$

$$H_2(j_n) = H_0^2(j_n) = H_1^2(j_n). \tag{29}$$

The received data vector $\mathbf{Y}_d = \left[Y_0(j_n), Y_1^*(j_n)\right]^T$ can be expressed as

$$\mathbf{Y}_d = \mathbf{H}_{eff}\mathbf{S} + \mathbf{W}_d = \begin{bmatrix} H_1(j_n) & H_2(j_n) \\ H_2^*(j_n) & -H_1^*(j_n) \end{bmatrix} \begin{bmatrix} S_0(n) \\ S_1(n) \end{bmatrix} + \begin{bmatrix} W_0(j_n) \\ W_1^*(j_n) \end{bmatrix}. \tag{30}$$

Thus, if we assume perfect channel estimation, the decoded data vector $\hat{\mathbf{S}} = \left[\hat{S}_0(n), \hat{S}_1(n)\right]^T$ is expressed as

$$\hat{\mathbf{S}} = \mathbf{H}_{eff}^H \mathbf{Y}_d = \mathbf{H}_{eff}^H \mathbf{H}_{eff}\mathbf{S} + \mathbf{H}_{eff}^H \mathbf{W}_d = \|\mathbf{h}\|^2 \mathbf{I}_2\mathbf{S} + \hat{\mathbf{W}}_d \tag{31}$$

where $\mathbf{h} = \left[H_1(j_n), H_2(j_n)\right]$, and $\hat{\mathbf{W}}_d$ is a 2×1 zero mean noise vector with covariance matrix $N_0\|\mathbf{h}\|^2 \mathbf{I}_2$. The extra-data symbols $\hat{S}_e(n)$ and $\hat{S}'_e(n)$ can be decoded by an MRC operation, given by Eq. (21) and Eq. (22).

The normalized effective throughput of the proposed OFDM transmit diversity, which is derived from [10], can be expressed as

$$T_h = \frac{2N_d \log_2 M \times (1-p_b)^{2N_d} + N_e \log_2 M_e \times (1-p_e)^{N_e} + N'_e \log_2 M'_e \times (1-p'_e)^{N'_e}}{2N} \tag{32}$$

where M and p_b are the modulation order and the bit error rate (BER) of the original data symbols, respectively. M_e and p_e denote those of the extra-data symbols in the 1st antenna. In addition, M'_e and p'_e denote those of the extra-data symbols in the 2nd antenna.

4. Simulation Results

To verify the proposed channel estimator, we considered an OFDM system with transmit diversity in the 10 MHz band at 2.3 GHz, with the following parameters: $N = 1024$, $Nv = 127$, $Ncp = 128$, and there was QPSK modulation for the original data symbols. For the multipath fading channel, the "Vehicular A" channel model was used, as defined by ETSI for the evaluation of UMTS radio interface proposals. For the subcarrier allocation, we set the following parameters:

$$\{i_n = nD_f\},\ \{k_n = nD_f + 1\},\ \text{and}\ \{u_n\} \subset \{i_n\} \tag{33}$$

$$\{i'_n = k_n\},\ \{k'_n = i_n\},\ \text{and}\ \{u'_n\} \subset \{i'_n\}. \tag{34}$$

In addition, the random interleavers were utilized for $\{\delta_n\}$ and $\{\delta'_n\}$.

Fig. 3 and **Fig. 4** compare mean square error (MSE) performances in the 1st and 2nd antennas, respectively. *Conv*(D_{f1},D_{f2}) represents a conventional LMMSE channel estimator with the minimum pilot spacing: D_{f1} in the 1st antenna and D_{f2} in the 2nd antenna. *Prop*(D_{f1}/λ_{f1}, D_{f2}/λ_{f2}) corresponds to the proposed LMMSE channel estimator, where λ_{f1} and λ_{f2} denote the extra-data spacings in D_{f1} and D_{f2}, respectively. As shown in **Fig. 3** and **Fig. 4**, a smaller (D_{f1},D_{f2}) resulted in better MSE performance. The MSE performance of *Prop*(16/8, 16/8) was superior to that of *Prop*(32/8, 32/8) across the entire signal-to-noise ratio (SNR) range. The MSE performance of *Prop*(16/8, 16/8) was inferior to *Conv*(8, 8) at low SNRs, but the performance was identical at high SNRs.

Fig. 5 shows the BER performances of the conventional and proposed channel estimators. Here, we considered the modulation types for the extra-data symbols as follows: *Prop*(D_{f1}/λ_{f1}, D_{f2}/λ_{f2}, *ModT ype*). *Prop*(16/8, 16/8, *QPSK*) resulted in performance identical to *Conv*(8, 8). For a BER of 10^{-3}, there was a greater than 10 dB SNR gain over *Conv*(32, 32). If *Prop*(32/8, 32/8, *QPSK*) was considered, there was error flooring at high SNRs exceeding 20 dB. Moreover, the modulation order change of the extra-data symbols resulted in slight performance degradation at low SNRs.

In **Fig. 6**, the advantages of the proposed subcarrier allocation and channel estimator are significant. The proposed method improved the normalized effective throughput. Especially, *Prop*(32/8, 32/8, *QPSK*) was better than *Conv*(8, 8), in spite of its BER saturation. *Prop*(16/8, 16/8, 16*QAM*) outperformed *Prop*(16/8, 16/8, *QPSK*) at SNRs exceeding 21 dB.

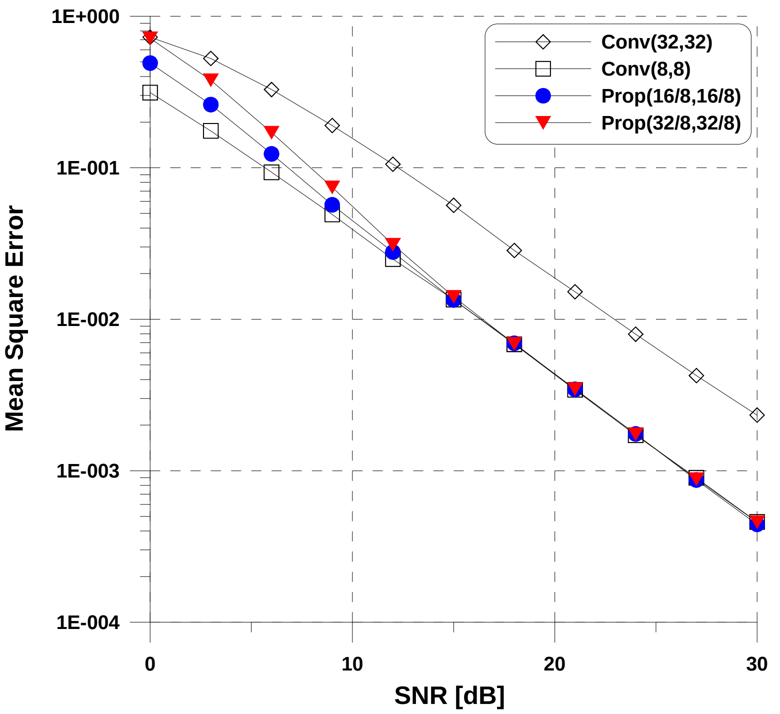

Fig. 3. MSE performance comparison between the conventional and proposed channel estimators, from the 1st antenna to the receiver

5. Conclusions

In this paper, we proposed a subcarrier allocation method, and a composite LMMSE channel estimator, which resulted in an efficient OFDM transmit diversity spectrum. By reducing the number of pilot symbols required, and inserting repetition-coded data symbols, the spectrum efficiency was improved. Moreover, the proposed channel estimator guarantees that the BER performance is as good as a conventional LMMSE channel estimator using full pilot symbols.

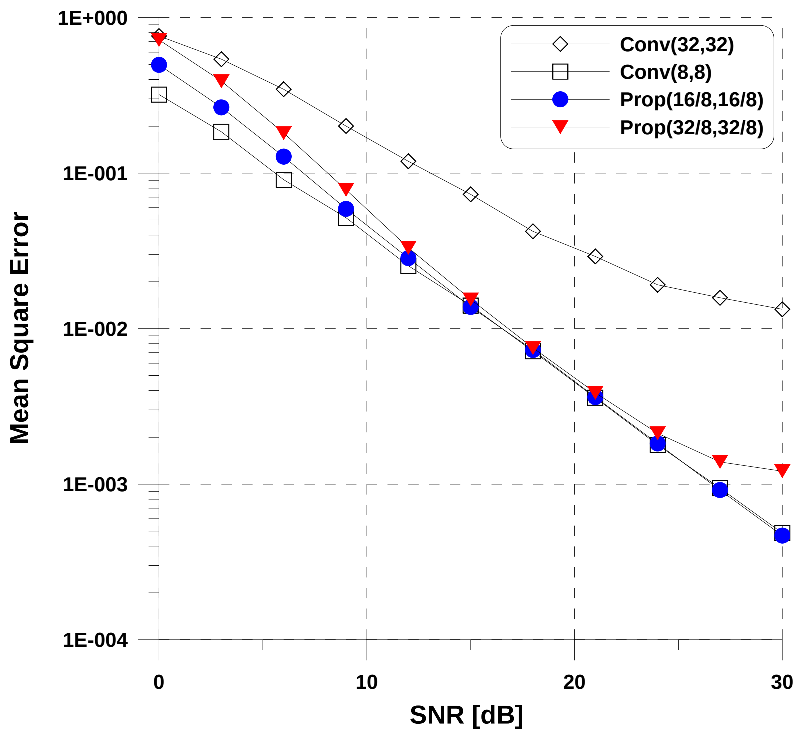

Fig. 4. MSE performance comparison between the conventional and proposed channel estimators, from the 2nd antenna to the receiver

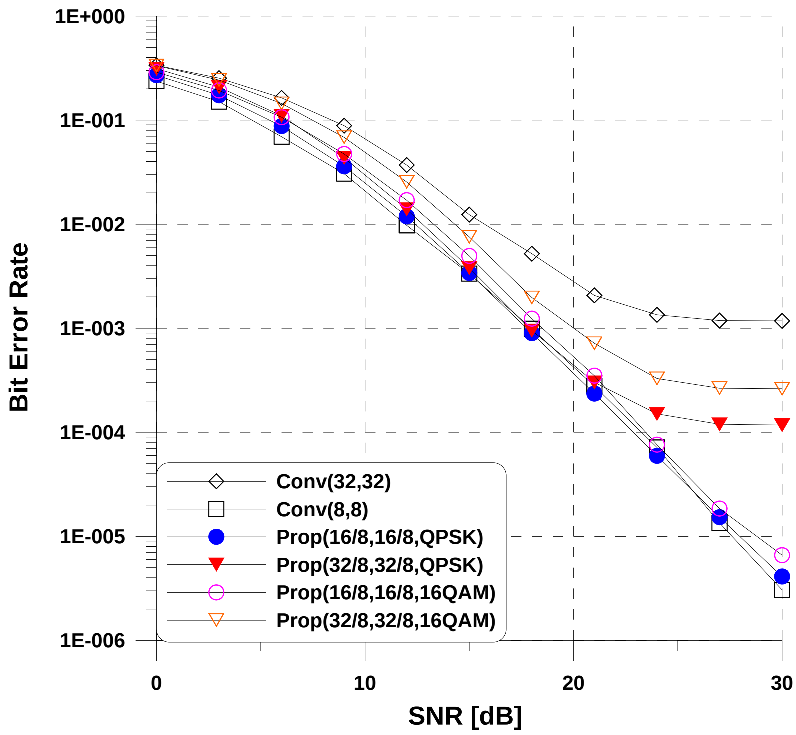

Fig. 5. BER performance comparison between the conventional and proposed channel estimators,

according to SNR [dB]

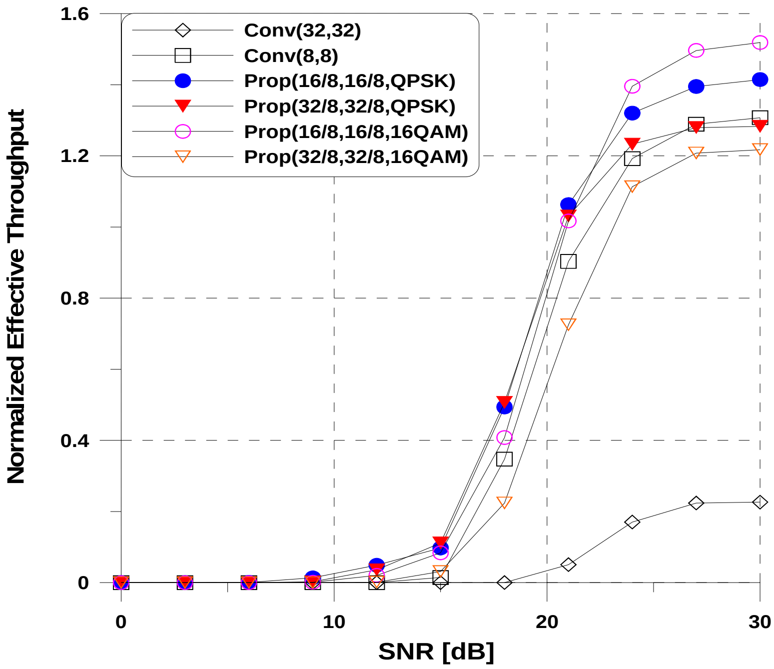

Fig. 6. Normalized effective throughput comparison between the conventional and proposed channel estimators, according to SNR [dB]

References

[1] Richard van Nee and Ramjee Prasad, "OFDM for Wireless Multimedia Communications," Artech House, 2000.

[2] David Tse and Pramod Viswanath, "Fundamentals of Wireless Communication," Cambridge Univ. press, 2005.

[3] IEEE Std. 802.16^{TM}-2004, "IEEE standard for local and metropolitan area networks- Part 16: Air interface for fixed broadband wireless access systems," Oct. 2004.

[4] Yue Li, Jianhua He, and Weixi Xing, "Bandwidth management of WiMAX systems and performance modeling," *KSII Transactions on Internet and Information Systems*, vol. 2, no. 2, pp. 63-81, April 2008.

[5] B. W. Song, Y. F. Guan, and W. J. Zhang, "An efficient training sequences strategy for channel estimation in OFDM systems with transmit diversity," *Journal of Zhejiang Univ. SCI*, pp. 613-618, 2005.

[6] D. Shen, Z. Diao, K. K. Wong, and V. O. K. Li, "Analysis of pilot-assisted channel estimators for OFDM systems with transmit diversity," *IEEE Transactions on Broadcasting*, vol. 52, no. 2, June 2006.

[7] M. Rim, "Optimal combining decision-directed and pilot-symbol-aided channel estimators," *Electronics Letters*, vol. 39, no. 6, pp. 558–560, Mar. 2003.

[8] Y. Ma, Y. Huang, N. Yi, and F. Liu, "Combined channel estimator for pilot-assisted wireless OFDM communication systems," In *Proceedings of IEEE VTC 2004-Spring*, vol. 4, pp. 1888–1891, May 2004.

[9] O. Edfors, M. Sandel, J.-J. van de Beek, S. K. Wilson, and P. O. Börjesson, "OFDM channel estimation by singular value decomposition," *IEEE Transactions on Communications*, vol. 46, no. 7, pp. 931-939, July 1998.

[10] D. Shen, Z. Pan, K. K. Wong, and V. O. K. Li, "Effective throughput: A unified benchmark for

pilot-aided OFDM/SDMA wireless communication systems," In *Proceedings of IEEE INFOCOM 2003*, vol. 3, pp. 1603–1613, April 2003.

Jeong-Wook Seo received the B.S. and M.S. degrees in the Department of Telecommunication & Information Engineering from Korea Aerospace University, Gyeonggi, Korea, in 1999 and 2001, respectively. He is currently working toward the Ph.D. degree in the School of Electrical & Electronic Engineering from Yonsei University, Seoul, Korea. Since 2001, he has been with Advanced Mobile Technology Research Center in Korea Electronics Technology Institute (KETI), Gyeonggi, Korea. His research interests include statistical signal processing, digital communications, and OFDM-based wireless systems.

Won-Gi Jeon received the B.S., M.S., and Ph.D degrees in electronic engineering from Chung-Ang University, Seoul, Korea, in 1994, 1996, and 1999, respectively. From 1999 to 2001, he was a BK contract professor of School of Electrical and Electronic Engineering in Chung-Ang University. In 2001, he joined Korea Electronics Technology Institute (KETI), Seongnam-si, Gyeonggi-do, Korea, where he is presently a Team Leader. His research interests are in the area of signal processing for high-speed digital communication and multicarrier transmission system.

Richard (Jong-Ho) Paik received the B.S., M.S., and Ph. D. degrees in the school of Electrical and Electronic Engineering from Chung-Ang University, Seoul, Korea, in 1994, 1997, and 2007, respectively. Since 1997, he has been working at Korea Electronics Technology Institute (KETI), Seongnam-si, Gyeonggi-do, Korea, where he is presently the Director of DxB·Communication Convergence Research Center. His research interests are in the areas of wireless/wired communications system design, video communications system design, and system architecture for realizing advanced digital communications system, for next generation digital any x broadcasting (DxB) networks.

Minho Jo received B.S. degree in Industrial Engineering, Chosun Univ., South Korea and his Ph.D. in Computer Networks in Department of Industrial and Systems Engineering, Lehigh University, Bethlehem, PA, USA in 1994. He worked as a staff researcher with Samsung Electronics, South Korea and was professor of School of Ubiquitous Computing and Systems, Sejong Cyber University, Seoul, South Korea. He is now a Research Associate Professor in School of Information and Communication Engineering, Sungkyunkwan Univ., Suwon, South Korea. Prof. Minho Jo is Executive Director of Korean Society for Internet Information (KSII) and Board of Trustees of Institute of Electronics Engineers of Korea (IEEK), respectively. He is Editor-in-Chief and Chair of the Steering Committee of KSII Transactions on Internet and Information Systems, General Chair of International Ubiquitous Conference, and Co-chair of International Conference on Ubiquitous Convergence Technology. Dr. Jo is Editor of Journal of Wireless Communications and Mobile Computing, and Associate Editor of Journal of Security and Communication Networks published by Wiley and Sons, respectively. He serves on an Associate Editor of Journal of Computer Systems, Networks, and Communications published by Hindawi. He serves as Chairman of IEEE/ACM WiMax/WiBro Services and QoS Management Symposium, IWCMC 2008. He is Technical Program Committee of IEEE ICC 2008, 2009 and IEEE GLOBECOM

2008, 2009. His interesting area lies in wireless ad-hoc sensor networks & wireless mesh networks, RFID, security in communication networks, machine intelligence in communications, and ubiquitous and mobile computing.

Dong Ku Kim received the B. Eng. Degree from Korea Aerospace University, Korea, in 1983 and the M.Eng and Ph.D. degrees from the University of Southern California, Los Angeles, in 1985 and 1992, respectively. He was a Research Engineer with the Cellular Infrastructure Group, Motorola by 1994, and he has been a professor in the school of Electrical and Electronic Engineering, Yonsei University, Seoul, since 1994. He is a director of Radio Communication Research Center at Yonsei University. He has been also a director of Qualcomm Yonsei CDMA Joint Research Lab since 1999. His main research interests are multi-user mulit-mode MIMO, OFDM, scheduling schemes, mobile multihop relay, and location finding algorithms.