LS Codes Aided Channel Estimation for MIMO-OFDM Systems in Multipath Environment

Wonsop Kim, Hyun Kyu Chung, Kwang Chun Lee
Mobile Telecommunication Research Laboratory
Electronics and Telecommunications Research Institute
161, Gajeong-dong, Yuseong-gu, Daejeon, 305-700, Korea
E-mail: {topsop, hkchung, kclee}@etri.re.kr

Hyoungsuk Jeon, Jongsub Cha, Hyuckjae Lee
School of Engineering
Information and Communications University
103-6, Munji-dong, Yuseong-gu, Daejeon, 305-732, Korea
E-mail: {hschun, s1009, hjlee}@icu.ac.kr

Abstract—**MIMO-OFDM system is one of the promising schemes for achieving high data rate in communication systems. In this paper, channel estimation techniques based on Loosely Synchronous (LS) codes are proposed for MIMO-OFDM systems. Since LS codes have perfect auto-correlation and cross-correlation functions within certain vicinity of the zero shifts, multiple antenna and multipath signal interference can be reduced substantially. Through simulations, it is shown that the proposed methods outperform both of least square and linear minimum mean square error (LMMSE) estimators.**

Keywords-LS codes; channel estimation; MIMO-OFDM systems; multipath channels

I. INTRODUCTION

The requirement of high-data rate services in mobile communication systems becomes more and more prevalent in recent years. Strong candidates for these needs are MIMO systems, OFDM systems and combined MIMO-OFDM systems. MIMO systems have shown their potential by improving the diversity gain with scheme as Space-Time Block Codes (STBC) [1] and increasing the capacity gain significantly with scheme as Bell Lab Layered Space-Time Architecture (BLAST).

As an attractive solution for high data rate transmission in multipath environment, MIMO systems have been applied to the OFDM systems. However, since channel fading characteristics are not perfectly known at the receiver, reliable channel estimations are needed to take the advantages of the MIMO-OFDM systems. For the SISO-OFDM systems, various channel estimation techniques such as least square [4] and linear minimum mean square error (LMMSE) [4] estimators have been studied. In MIMO-OFDM systems, these techniques are inadequate for channel estimation, because signals from other transmission antennas act as interference.

Smart codes [7] have been studied previously by many researchers for interference-free communications. LS codes [6], one of the smart codes based on Golay complementary pairs (GCP) [5], have the zero correlation zone (ZCZ) or Interference Free Window (IFW). Utilizing the ZCZ or IFW, it is proposed that the multipath and multiple antennas interference can be significantly reduced.

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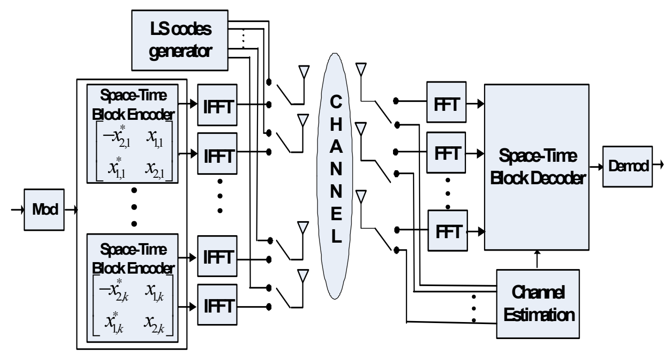


Figure 1. A simplified block diagram for a MIMO-OFDM system based on LS codes channel estimation

In this paper, we propose two channel estimation techniques for MIMO-OFDM systems by using LS codes as training sequences. Since LS codes have perfect auto-correlation and cross-correlation functions within certain vicinity of the zero shifts, multiple antenna and multipath signal interference can be reduced substantially.

The outline of this paper is as follow. In section II, a system model of MIMO-OFDM systems is described. In section III, properties of LS codes are described. In section IV, the proposed channel estimation techniques are described. Simulation results demonstrating the performance improvement are provided in section V. Finally, conclusions are given in section VI.

II. SYSTEM MODEL

A MIMO-OFDM system having k synchronous co-channel terminals at the transmitters is shown in Fig. 1. Each terminal uses Alamouti coded OFDM system [3]. The receiver is equipped with m receive antennas. Let, $H^i_{1,j}$ and $H^i_{2,j}$ denote the channel frequency responses of the k^{th} tone of the OFDM block, corresponding to the channel between the 1^{st} and 2^{nd} transmission antennas of the i^{th} terminal and the j^{th} receive antenna respectively, where $i = 1, \ldots k;\ j = 1, \ldots, m$. Let $\left[X_{1,i}(k),\ X_{2,i}(k)\right]$ are transmitted from the i^{th} terminal simultaneously during the first symbol period. During the second symbol period, $\left[-X^*_{2,i}(k),\ X^*_{1,i}(k)\right]$ are transmitted from the i^{th} terminal. Half of the total power is transmitted from each antenna to maintain the transmitted power equal to a single antenna system.

Assuming that the channel responses are constant over two consecutive block periods, the output of the OFDM demodulator over two consecutive OFDM block periods at the j^{th} receive antenna, $Y_{1,j}(k)$ and $Y_{2,j}(k)$, for $k=1,2,..,N$ could be written as

$$\begin{aligned} Y_{1,j}(k) &= \sum_{i=1}^{k}\left\{H_{1,j}^{i}(k)X_{1,i}(k)+H_{2,j}^{i}(k)X_{2,i}(k)\right\}+W_{1,j}(k), \\ Y_{2,j}(k) &= \sum_{i=1}^{k}\left\{H_{2,j}^{i}(k)X_{1,i}^{*}(k)-H_{1,j}^{i}(k)X_{2,i}^{*}(k)\right\}+W_{2,j}(k), \end{aligned} \quad (1)$$

where $W_{1,j}(k)$ and $W_{2,j}(k)$ represent the frequency response of AWGN. We also define following equations.

$$\begin{aligned} &\mathbf{Y}_j(k)=\left[Y_{1,j}(k),\, Y_{2,j}^{*}(k)\right]^T,\ \mathbf{X}_i(k)=\left[X_{1,i}(k),\, X_{2,i}(k)\right]^T, \\ &\mathbf{W}_j(k)=\left[W_{1,j}(k),\, W_{2,j}^{*}(k)\right]^T,\ \mathbf{H}_j^i(k)=\begin{bmatrix} H_{1,j}^{i}(k) & H_{2,j}^{i}(k) \\ H_{2,j}^{i\,*}(k) & -H_{1,j}^{i\,*}(k)\end{bmatrix}^T. \end{aligned} \quad (2)$$

The received signals from all receive antennas at the k^{th} tone over two consecutive OFDM block periods can be written in a matrix form as

$$\begin{bmatrix} \mathbf{Y}_1(k) \\ \mathbf{Y}_2(k) \\ \vdots \\ \mathbf{Y}_m(k) \end{bmatrix} = \begin{bmatrix} \mathbf{H}_1^1(k) & \mathbf{H}_1^2(k) & \cdots & \mathbf{H}_1^k(k) \\ \mathbf{H}_2^1(k) & \mathbf{H}_2^2(k) & \cdots & \mathbf{H}_2^k(k) \\ \vdots & \vdots & \vdots & \vdots \\ \mathbf{H}_m^1(k) & \mathbf{H}_m^2(k) & \cdots & \mathbf{H}_m^k(k) \end{bmatrix} \begin{bmatrix} \mathbf{X}_1(k) \\ \mathbf{X}_2(k) \\ \vdots \\ \mathbf{X}_k(k) \end{bmatrix} + \begin{bmatrix} \mathbf{W}_1(k) \\ \mathbf{W}_2(k) \\ \vdots \\ \mathbf{W}_m(k) \end{bmatrix}. \quad (3)$$

The representation of (3) is rewritten by

$$\mathbf{Y}(k)=\mathbf{H}(k)\mathbf{X}(k)+\mathbf{W}(k). \quad (4)$$

III. LOOSELY SYNCHRONOUS CODES

A. Properties of uncorrelated LS codes in interference free window

LS codes are defined as the combination of C and S subsequences, a Golay complementary pair, with zeros inserted between them. If (C_0, S_0) and (C_1, S_1) are both Golay pairs of LS codes, we say that two LS codes are a mate. Fig. 2 shows formation of LS codes whose components are equal to ± 1 or 0. The reason for the zeros insertion is to avoid overlapping between the subsequences so as to form the desired aperiodic orthogonal zone.

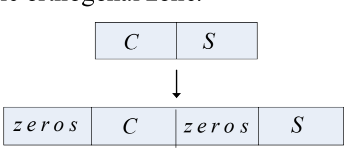
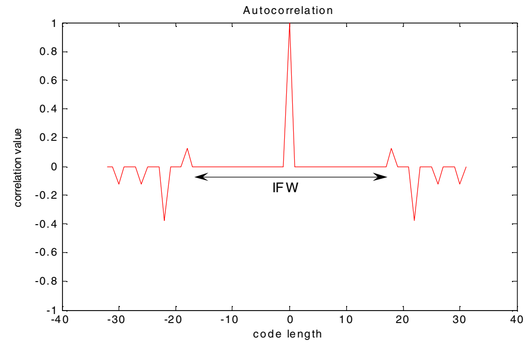
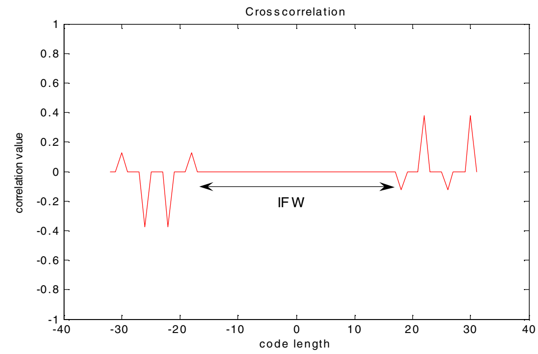


Figure 2. Formation of LS codes

As a result of zeros insertion, LS codes have shown that aperiodic auto-correlation sidelobes and cross-correlations are zero within IFW W_0. Fig. 3 shows auto-correlation and cross-correlation property of a mate with length $N=64$ and IFW $W_0=32$.



(a) Auto-correlation property



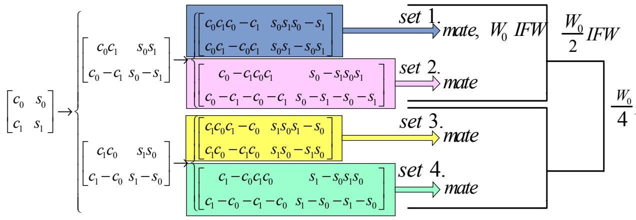
(b) Cross-correlation property

Figure 3. Correlation properties of a mate

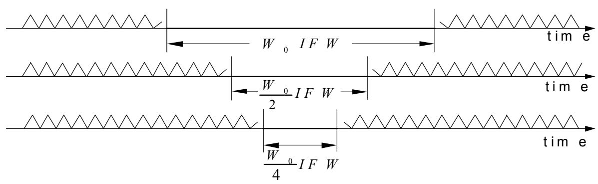
B. Properties of correlated LS codes in interference free windows

The basic idea of the LS codes is the insertion of zeros to avoid the sequences C_0 and C_1 overlapping with the sequences S_0 and S_1. It is also necessary to insert enough "guard" intervals between sequences of length longer than the maximum dispersion delay of the channel. If we assume that $W_0/2$ is the maximum delay dispersion of the channel, in order not to decrease the energy efficiency which are mentioned in [6], we define that the length of IFW W_0 is equal to N_p-1, where N_p is length of a complementary pair. As shown in Fig. 4, using Golay codes of $N_p=8$ which are generated by using systematic method [6], the properties of possible IFWs are as follows.

1) Set 1, 2, 3 and 4 are mates, respectively.
2) Considering set 1 is initial mate, IFW between set 1 and 2 is $W_0/2$.
3) Considering set 1 is initial mate, IFW between set 1 and 3 is $W_0/4$.
4) Considering set 1 is initial mate, IFW between set 1 and 4 is $W_0/4$.
5) If set 2, 3 or 4 are initial mates respectively, properties of IFWs among three other mates are the same as the set 1 is initial mate.



(a) Three possible IFW properties using complementary pairs of Golay codes generated by systematic method



(b) Three possible IFWs

Figure 4. Three possible cross-correlation properties

IV. LS CODES AIDED CHANNEL ESTIMATIONS

A. Proposed channel estimation using uncorrelated LS codes in interference free window (proposed estimation 1)

Using the following properties, we propose the first channel estimation method for MIMO-OFDM systems using a mate of LS codes as training sequences in multipath channels.

1) The length of IFW W_0 is equal to $N_p - 1$, where N_p is length of a complementary pair.
2) A mate of LS codes has maximum IFW.
3) $W_0 / 2$ is assumed to be the maximum delay dispersion of the channel
4) Guard interval of OFDM systems is about a quarter of an OFDM symbol duration.

Let's consider a MIMO-OFDM system as shown in Fig. 1. During the training period, two transmission antennas simultaneously transmit a mate of LS codes at a given symbol period by DPSK. After transmitting a mate from the first two antennas, another two transmission antennas also transmit a mate and so on. LS codes are transmitted from each transmission antenna without IFFT process. The receivers also do not need FFT process and estimate the channel with Alamouti coded blocks. Once the initial channel parameters are estimated using LS codes, these estimated channel parameters are used to decode the data blocks.

B. Proposed channel estimation using weakly correlated LS codes in interference free window with interference cancellation (Proposed estimation 2)

In the proposed estimation 1, it is limited to use only two transmit antennas at given time. In order to extend the number of transmit antenna, it is necessary pursuit more on the properties of LS codes. For the mates having length $N = 128$ and IFW $W_0 = 64$, we can have four weakly correlated cross-correlations. Fig. 5 shows four weakly correlated cross-correlations. If we use the properties of weakly correlated LS codes having $W_0 / 2$ IFW, we may reduce the computation

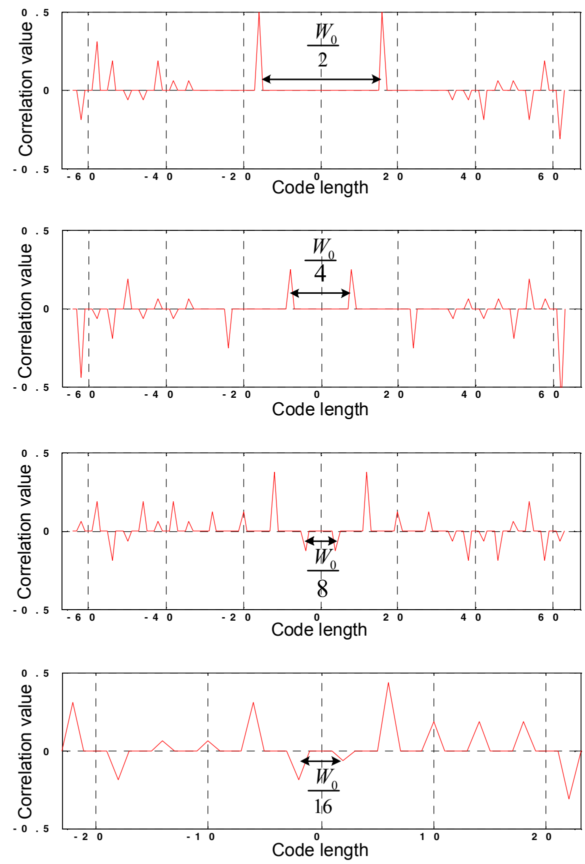


Figure 5. Four weakly correlated cross-correlations

complexity and, then achieve higher data rate communication than the proposed estimation 1 at the cost of a little performance degradation. Using the following properties, we propose the second channel estimation method for MIMO-OFDM systems using four LS codes as training sequences with interference cancellation.

1) The length of IFW W_0 is equal to $N_p - 1$, where N_p is length of a complementary pair.
2) (c_1, c_2) and (s_1, s_2) are mates of LS codes respectively, and they are weakly correlated with IFW $W_0 / 2$.
3) Since the locations and values of correlations among four LS codes are known in advance, interference cancellation method can be used.
4) $W_0 / 2$ is assumed to be the maximum delay dispersion of the channel
5) Guard interval of OFDM systems is usually about a quarter of an OFDM symbol duration.

Let's consider a MIMO-OFDM system as shown in Fig. 1. Suppose that four transmission antennas simultaneously transmit four weakly correlated LS codes at a given symbol period. After transmitting four codes from the first four antennas, another four transmission antennas also transmit four codes and so on. At the receiver, we can estimate the channel with Double Space Time Transmit Diversity (DSTTD) [2] blocks. The time domain received signals at the first and second receive antenna can be represented respectively as

$$
\begin{aligned}
y_{1,1}(t) &= h_{1,1}^1(\tau,t)*c_1(t)+h_{2,1}^1(\tau,t)*c_2(t)+h_{1,1}^2(\tau,t)*s_1(t)+h_{2,1}^2(\tau,t)*s_2(t)+w_{1,1}(t), \\
y_{1,2}(t) &= h_{1,2}^1(\tau,t)*c_1(t)+h_{2,2}^1(\tau,t)*c_2(t)+h_{1,2}^2(\tau,t)*s_1(t)+h_{2,2}^2(\tau,t)*s_2(t)+w_{1,2}(t),
\end{aligned}
\tag{5}
$$

where " $*$ " denotes the convolution, ($c_1(t)$, $c_2(t)$) and ($s_1(t)$, $s_2(t)$) are mates respectively at time t , each has length N , $(c_1(t), c_2(t))$ and $(s_1(t), s_2(t))$ are weakly correlated with IFW $W_0/2$, both $h_{1,j}^i(\tau,t)$ and $h_{2,j}^i(\tau,t)$ are time domain response of channel for $i, j = 1,2$, $w_{1,1}$ and $w_{1,2}$ are AWGN. At the receiver, we can get the channel parameters by following methods.

$$
\begin{aligned}
\tilde{h}_{1,1}^1 &= R_{y_{1,1},c_1}(m) = h_{1,1}^1 + R_{h_{2,1}^1*c_2,c_1}(m) + R_{h_{1,1}^2*s_1,c_1}(m) + R_{h_{2,1}^2*s_2,c_1}(m) + R_{w_{1,1},c_1}(m), \\
\tilde{h}_{1,2}^1 &= R_{y_{1,2},c_1}(m) = h_{1,2}^1 + R_{h_{2,2}^1*c_2,c_1}(m) + R_{h_{1,2}^2*s_1,c_1}(m) + R_{h_{2,2}^2*s_2,c_1}(m) + R_{w_{1,2},c_1}(m), \\
\tilde{h}_{2,1}^1 &= R_{y_{1,1},c_2}(m) = h_{2,1}^1 + R_{h_{1,1}^1*c_1,c_2}(m) + R_{h_{1,1}^2*s_1,c_2}(m) + R_{h_{2,1}^2*s_2,c_2}(m) + R_{w_{1,1},c_2}(m), \\
\tilde{h}_{2,2}^1 &= R_{y_{1,2},c_2}(m) = h_{2,2}^1 + R_{h_{1,2}^1*c_1,c_2}(m) + R_{h_{1,2}^2*s_1,c_2}(m) + R_{h_{2,2}^2*s_2,c_2}(m) + R_{w_{1,2},c_2}(m), \\
\tilde{h}_{1,1}^2 &= R_{y_{1,1},s_1}(m) = h_{1,1}^2 + R_{h_{1,1}^1*c_1,s_1}(m) + R_{h_{2,1}^1*c_2,s_1}(m) + R_{h_{2,1}^2*s_2,s_1}(m) + R_{w_{1,1},s_1}(m), \\
\tilde{h}_{1,2}^2 &= R_{y_{1,2},s_1}(m) = h_{1,2}^2 + R_{h_{1,2}^1*c_1,s_1}(m) + R_{h_{2,2}^1*c_2,s_1}(m) + R_{h_{2,2}^2*s_2,s_1}(m) + R_{w_{1,2},s_1}(m), \\
\tilde{h}_{2,1}^2 &= R_{y_{1,1},s_2}(m) = h_{2,1}^2 + R_{h_{1,1}^1*c_1,s_2}(m) + R_{h_{2,1}^1*c_2,s_2}(m) + R_{h_{1,1}^2*s_1,s_2}(m) + R_{w_{1,1},s_2}(m), \\
\tilde{h}_{2,2}^2 &= R_{y_{1,2},s_2}(m) = h_{2,2}^2 + R_{h_{1,2}^1*c_1,s_2}(m) + R_{h_{2,2}^1*c_2,s_2}(m) + R_{h_{1,2}^2*s_1,s_2}(m) + R_{w_{1,2},s_2}(m),
\end{aligned}
\tag{6}
$$

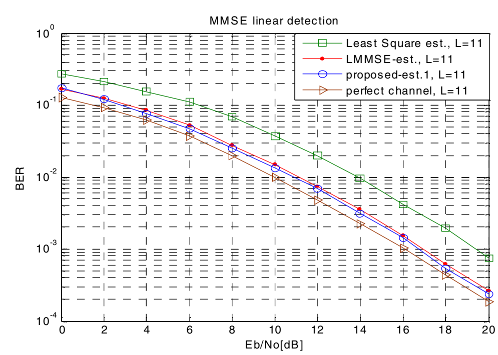
where $R_{a,a}(m)$ is the aperiodic auto-correlation of a and a , $R_{a,b}(m)$ is the aperiodic cross-correlation of a and b . Because four LS codes are correlated in IFW, we can not clearly eliminate the interference by taking respective cross-correlation. So, they are affected to each other. However, since we already know how the cross-correlations are occurring among four codes, it is possible to eliminate the interference. For example, we can assume that $R_{h_{1,1}^2*s_1,c_1}(m) \approx R_{\tilde{h}_{1,1}^2*s_1,c_1}(m)$, which is given by calculating the $\tilde{h}_{1,1}^2$ in (6). Using above facts, we can describe an interference cancellation method as follow.

$$
\begin{aligned}
\hat{h}_{1,1}^1 &= \tilde{h}_{1,1}^1 - R_{\tilde{h}_{2,1}^1*c_2,c_1}(m) - R_{\tilde{h}_{1,1}^2*s_1,c_1}(m) - R_{\tilde{h}_{2,1}^2*s_2,c_1}(m), \\
\hat{h}_{1,2}^1 &= \tilde{h}_{1,2}^1 - R_{\tilde{h}_{2,2}^1*c_2,c_1}(m) - R_{\tilde{h}_{1,2}^2*s_1,c_1}(m) - R_{\tilde{h}_{2,2}^2*s_2,c_1}(m), \\
\hat{h}_{2,1}^1 &= \tilde{h}_{2,1}^1 - R_{\tilde{h}_{1,1}^1*c_1,c_2}(m) - R_{\tilde{h}_{1,1}^2*s_1,c_2}(m) - R_{\tilde{h}_{2,1}^2*s_2,c_2}(m), \\
\hat{h}_{2,2}^1 &= \tilde{h}_{2,2}^1 - R_{\tilde{h}_{1,2}^1*c_1,c_2}(m) - R_{\tilde{h}_{1,2}^2*s_1,c_2}(m) - R_{\tilde{h}_{2,2}^2*s_2,c_2}(m), \\
\hat{h}_{1,1}^2 &= \tilde{h}_{1,1}^2 - R_{\tilde{h}_{1,1}^1*c_1,s_1}(m) - R_{\tilde{h}_{2,1}^1*c_2,s_1}(m) - R_{\tilde{h}_{2,1}^2*s_2,s_1}(m), \\
\hat{h}_{1,2}^2 &= \tilde{h}_{1,2}^2 - R_{\tilde{h}_{1,2}^1*c_1,s_1}(m) - R_{\tilde{h}_{2,2}^1*c_2,s_1}(m) - R_{\tilde{h}_{2,2}^2*s_2,s_1}(m), \\
\hat{h}_{2,1}^2 &= \tilde{h}_{2,1}^2 - R_{\tilde{h}_{1,1}^1*c_1,s_2}(m) - R_{\tilde{h}_{2,1}^1*c_2,s_2}(m) - R_{\tilde{h}_{1,1}^2*s_1,s_2}(m), \\
\hat{h}_{2,2}^2 &= \tilde{h}_{2,2}^2 - R_{\tilde{h}_{1,2}^1*c_1,s_2}(m) - R_{\tilde{h}_{2,2}^1*c_2,s_2}(m) - R_{\tilde{h}_{1,2}^2*s_1,s_2}(m).
\end{aligned}
\tag{7}
$$

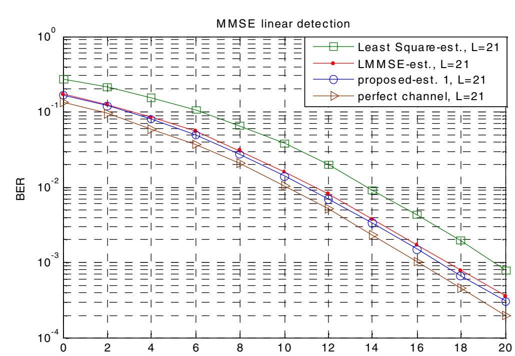
V. SIMULATION RESULTS

We consider a DSTTD-OFDM system with four transmit antennas and two receive antennas. The parameters of the simulation are as follow. Assume that the channel bandwidth W is 20MHz, which is divided into 128 subchannels N_s . A BPSK signal constellation is used. The OFDM symbol duration T_s is 8 μs . An additional 1.6 μs guard interval T_g is used to provide protection from intersymbol interference (ISI). Two channel models are used, namely, Channel A with $\tau_{rms} = 50$ ns, and Channel B with $\tau_{rms} = 100$ ns. Both channels assume an exponentially decaying power delay profile with 11 multipaths and 21 multipaths respectively, which are independently generated using Jake's model. We have used Monte-Carlo simulations to generate channel auto-correlation matrix for LMMSE estimation. To suppress the error propagation, training symbols are periodically inserted in the data stream after every 20 OFDM blocks. Table I summarizes both training symbols per transmission antenna and total training periods of the channel estimators for the DSTTD-OFDM system.

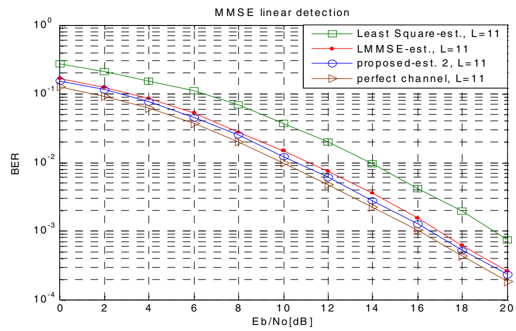
The performance of the system is measured in terms of BER versus E_b/N_0 for a minimum mean square error equalizer based on the estimated channel. Fig. 6 (a) and (b) show the BER performance versus E_b/N_0 among Least Square estimation, LMMSE estimation and the proposed estimation 1. We can see a about 3 dB gain in E_b/N_0 for the proposed estimation 1 over Least Square estimation at the BER of 10^{-3} . The proposed estimation 1 also has a little better performance by about 0.2 dB as compared with LMMSE estimation at the BER of 10^{-3} . This proposed estimation 1 method performs within 1.0 dB limit of the system with known channel. Fig. 6 (c) and (d) show the BER performance versus E_b/N_0 among Least Square estimation, LMMSE estimation and the proposed estimation 2. The performance results of the proposed estimation 2 are almost the same for proposed estimation 1. Fig. 7 (a) and (b) show that in the channel with relatively small delay spread like channel A or with exponentially decaying power delay profile like both channel A and B, the proposed estimation 2 saves more computation time and achieves higher data rate than the proposed estimation 1 at the cost of a little performance degradation.



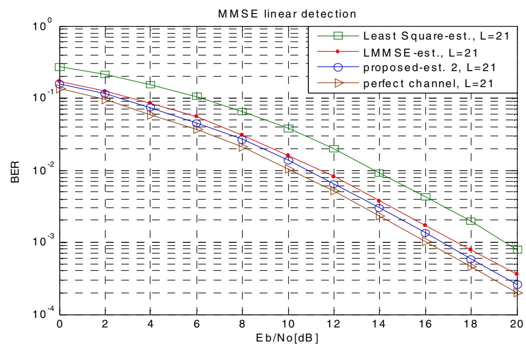
(a) Proposed estimation 1, channel A



(b) Proposed estimation 1, channel B

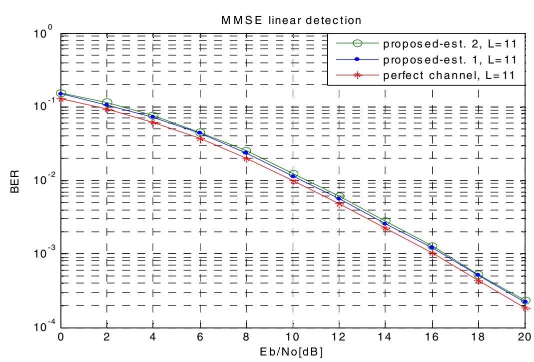


(c) Proposed estimation 2, channel A

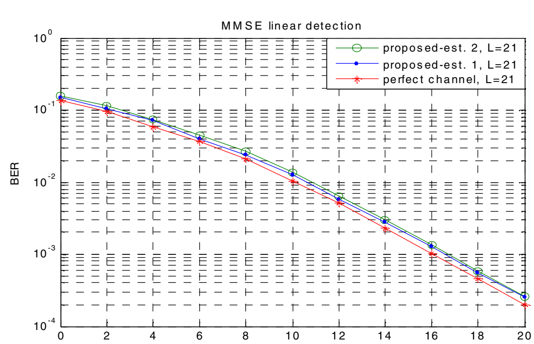


(d) Proposed estimation 2, channel B

Figure 6. BER performance of channel estimators for 4×2 DSTTD-OFDM system (Type-I)



(a) Proposed estimation 1 and 2, Channel A



(a) Proposed estimation 1 and 2, Channel B

Figure 7. BER performance comparison of channel estimators for 4×2 DSTTD-OFDM system (Type-II)

TABLE I. TRAINING SYMBOLS PER TRANSMISSION ANTENNA AND TOTAL TRAINING PERIODS

Estimators	**Type-I**		**Type-II**	
	Training symbols	***Training periods***	***Training symbols***	***Training periods***
Least Square	1 N_s	4 T_s	-	
LMMSE	1 N_s	4 T_s	-	
Proposed 1	2 N_s	4 T_s	4 N_s	8 T_s
Proposed 2	4 N_s	4 T_s	4 N_s	4 T_s

VI. CONCLUSIONS

In this paper, two channel estimation techniques based on the LS codes are proposed for MIMO-OFDM systems. The first one uses uncorrelated LS codes in IFW. The second one uses weakly correlated LS codes in IFW and interference cancellation method. The simulation results show that the BER performance of proposed estimation techniques are considerably improved as compared with that of Least Square estimator and slightly improved as compared with that of LMMSE estimator. Moreover the proposed techniques have the advantage of not requiring any information about the channel statistics like auto-correlation of channels. We also show that in the channel with relatively small delay spread or with exponentially decaying power delay profile, the second method achieves higher data rate than the first one at the cost of a little performance degradation.

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