LS Codes Assisted Multipath Channel Interference Canceller for MIMO-OFDM Systems

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***Abstract*—MIMO-OFDM system is a promising way to achieve high data-rate wireless transmission. In this paper, Loosely Synchronous (LS) codes assisted Multipath Channel Interference Canceller (MCIC) is proposed for the channel estimation of MIMO-OFDM systems. Since LS codes have perfect auto-correlation and cross-correlation functions within certain vicinity of the zero shifts, LS codes assisted MCIC can substantially reduce multipath channel interference. Simulation results show that the MCIC estimator outperforms least square estimator and has slightly inferior performance as compared with linear minimum mean square error (LMMSE) estimator. Moreover the MCIC estimator has the advantage of not requiring any information about the channel statistics like auto-correlation of channels.**

Keywords- channel estimation, LS code, MCIC, MIMO-OFDM systems, multipath channel.

I. INTRODUCTION

High data-rate transmission techniques in mobile communication systems have gained considerable interest in recent years. For instance, there are Multiple-Input Multiple-Output (MIMO) systems, Orthogonal Frequency Division Multiplexing (OFDM) systems, and MIMO-OFDM systems. MIMO systems have shown their potential by providing high data-rate and diversity gain obtained through spatial multiplexing system commonly referred to as Bell Lab Layered Space-Time (BLAST) and Space-Time Block Coding (STBC) [1] respectively.

Since OFDM is commonly used for high data-rate wireless communication due to its inherent error susceptibility, the combination of OFDM with MIMO systems is regarded as one of the most promising schemes for the next generation high speed wireless communication. However, it is impossible to know perfect channel characteristics at the receiver in terms of realistic aspect. Thus, reliable channel estimations are needed to take the advantages of the MIMO-OFDM systems. For the OFDM systems with Single-Input Single Output (SISO), various channel estimation techniques such as least square [3] and linear minimum mean square error (LMMSE) [3] estimators was introduced to estimate channel. In MIMO-OFDM systems, on the other hand, these techniques can not be applied directly for channel estimation because signals from other transmission antennas act as interference.

LS codes [5], one of the smart codes based on Golay complementary pairs (GCP) [4], have the zero correlation zone (ZCZ) or Interference Free Window (IFW). It is proposed that the multiple antennas and multipath interference can be significantly reduced by utilizing the ZCZ or IFW.

In this paper, we propose the MCIC for the channel estimation of 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, the MCIC can substantially reduce multipath channel interference by using these correlation properties of LS codes.

This paper is organized as follows. 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 MCIC is described. Finally, Section V and VI give simulation results and conclusions respectively.

II. SYSTEM MODEL

MIMO-OFDM systems with k synchronous Alamouti system [2] blocks at the transmitters are shown in Fig. 1. The output over two consecutive periods at the j^{th} receive antenna, $Y_{1,j}(k)$ and $Y_{2,j}(k)$, for $k=1,2,..,N$, can be written as;

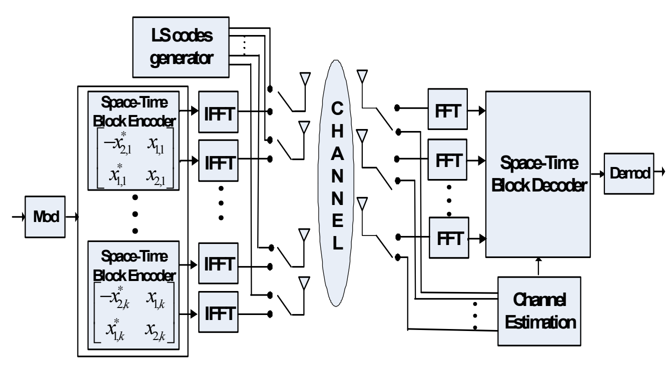

Figure 1. A simplified block diagram for a MIMO-OFDM system with LS codes generator

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$$Y_{1,j}(k)=\sum_{i=1}^{k}\left\{H^{i}_{1,j}(k)X_{1,i}(k)+H^{i}_{2,j}(k)X_{2,i}(k)\right\}+W_{1,j}(k),$$
$$Y_{2,j}(k)=\sum_{i=1}^{k}\left\{H^{i}_{2,j}(k)X^{*}_{1,i}(k)-H^{i}_{1,j}(k)X^{*}_{2,i}(k)\right\}+W_{2,j}(k), \tag{1}$$

where $\left[X_{1,i}(k),\ X_{2,i}(k)\right]$ are transmitted from the i^{th} block simultaneously during the first symbol period, $\left[-X^{*}_{2,i}(k),\ X^{*}_{1,i}(k)\right]$ are transmitted from the i^{th} block during the second symbol period, $H^{i}_{1,j}$ and $H^{i}_{2,j}$ denote the channel frequency responses of the i^{th} transmission block and the j^{th} receive antenna respectively, $W_{1,j}(k)$ and $W_{2,j}(k)$ represent the frequency response of Additive White Gaussian Noise (AWGN). We also define following equations.

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

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

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

Equation (3) can be reduced to the following form;

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

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 to avoid overlapping between the two subsequences. 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 a simple example of LS codes whose components are equal to ± 1 or 0. As a result of inserted zeros, LS codes have features that aperiodic auto-correlation sidelobes and cross-correlations are zero within IFW W_0. Fig. 3 shows auto-correlation and cross-correlation properties of a mate with length $N = 64$ and IFW $W_0 = 32$.

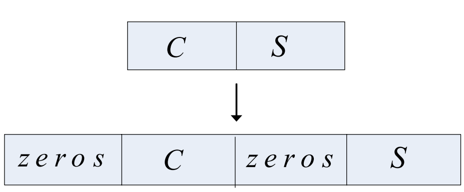
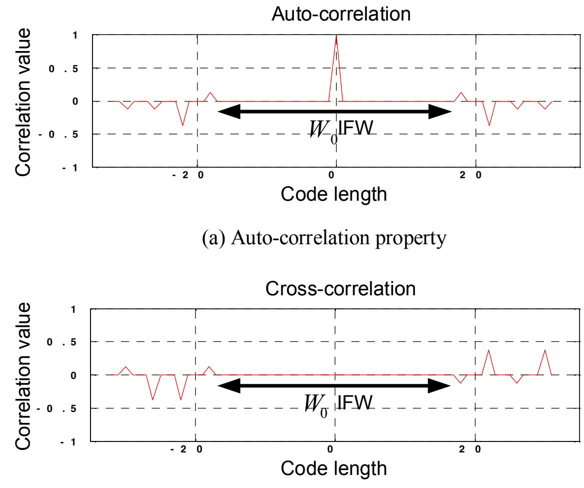


Figure 2. Formation of LS codes



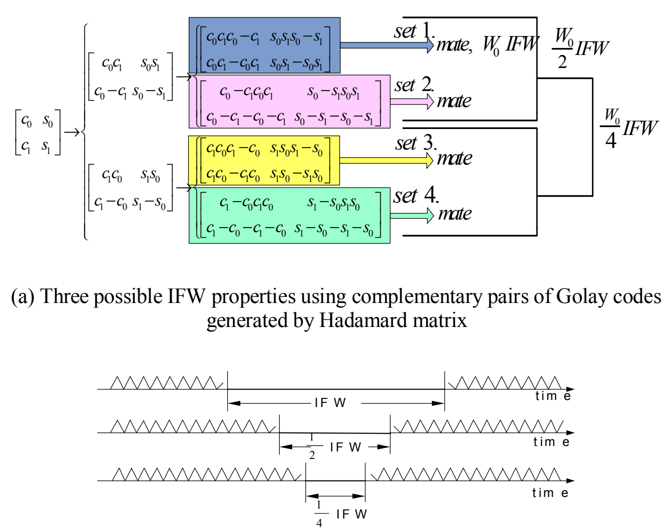
(a) Auto-correlation property

(b) Cross-correlation property

Figure 3. Correlation properties of a LS code mate

B. *Properties of correlated LS codes in interference free windows*

The main purpose of zeros insertion of the LS codes is to avoid the sequences C_0 and C_1 overlapping with the sequences S_0 and S_1. Note that it is also necessary to insert enough guard intervals between sequences with length longer than the maximum delay of the multipath channel. Under the assumption that $W_0/2$ is the maximum delay of the multipath channel without decreasing the energy efficiency suggested in [5], 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=16 which are generated by using Hadamard matrix [5], the properties of possible IFWs are as follows;



(a) Three possible IFW properties using complementary pairs of Golay codes generated by Hadamard matrix

(b) Three possible IFWs

Figure 4. Three possible cross-correlation properties

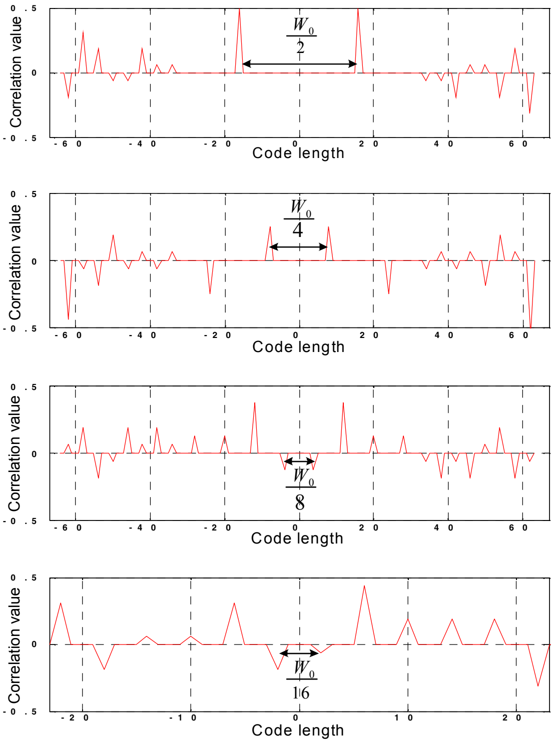
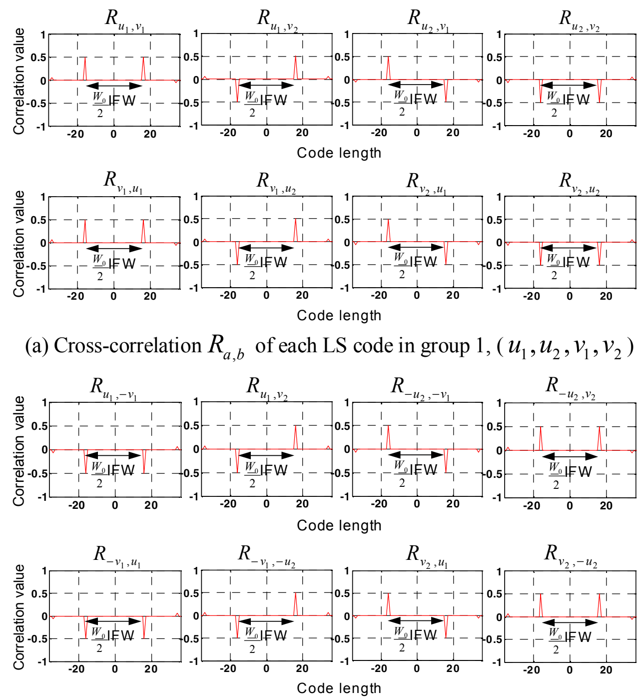


Figure 5. Four weakly correlated cross-correlations



(a) Cross-correlation $R_{a,b}$ of each LS code in group 1, (u_1, u_2, v_1, v_2)

(b) Cross-correlation $R_{a,b}$ of each LS code in group 2, ($u_1, -u_2, -v_1, v_2$)

Figure 6. Cross-correlation of each LS code in groups

III.B.1) Set 1, 2, 3 and 4 are mates, respectively.

III.B.2) If set 1 is initial mate, IFW between set 1 and 2 is $W_0/2$.

III.B.3) If set 1 is initial mate, IFW between set 1 and 3 is $W_0/4$.

III.B.4) If set 1 is initial mate, IFW between set 1 and 4 is $W_0/4$.

III.B.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, the initial mate.

For the mates having length $N = 128$ and IFW $W_0 = 64$, Fig. 5 shows four weakly correlated cross-correlations

IV. LS Codes Assisted Multipath Channel Interference Canceller

A. The MCIC using weakly correlated LS codes

For the mates having length $N = 128$ and IFW $W_0 = 64$, Fig. 6 shows each cross-correlation $R_{a,b}$ of four weakly correlated LS codes in group 1, (u_1, u_2, v_1, v_2), and group 2, $(u_1, -u_2, -v_1, v_2)$, respectively. Based on the following seven properties, we propose the MCIC for MIMO-OFDM systems using four LS codes as training sequences.

IV.A.1) The length of IFW W_0 is equal to $N_p - 1$, where N_p is length of a complementary pair.

IV.A.2) A mate of LS codes has maximum IFW W_0.

IV.A.3) (p_1, p_2) and (z_1, z_2) are mates of LS codes respectively, and they are weakly correlated with IFW $W_0/2$.

IV.A.4) $W_0/2$ is assumed to be the maximum delay of the channel.

IV.A.5) Guard interval of OFDM systems is about a quarter of an OFDM symbol duration.

IV.A.6) Since the locations and values of correlations among four LS codes are known in advance, interference cancellation method can be used.

IV.A.7) To estimate the channel, transmission antennas must transmit at least two different groups, group 1 and group 2.

Here we consider the MIMO-OFDM system as shown in Fig. 1. Suppose that four transmission antennas simultaneously transmit group 1 at a given symbol period. After transmitting group 1 from the first four antennas, another four antennas also transmit group 1, and repeat transmission of the group 1 for the rest of transmission antennas. After transmitting group 1 for all antennas, transmission antennas also transmit group 2 like the method of group 1 transmission for all antennas. At the receiver, we can estimate the channel with Double Space Time Transmit Diversity (DSTTD) [8] blocks. If we restrict ourselves to a DSTTD-OFDM system with four transmission antennas and two receive antennas, the time domain received signals at the first and second receive antenna can be represented respectively as;

$$\begin{aligned} y_{1,1}(t) &= h^1_{1,1}(\tau,t) * p_1(t) + h^1_{2,1}(\tau,t) * p_2(t) + h^2_{1,1}(\tau,t) * z_1(t) + h^2_{2,1}(\tau,t) * z_2(t) + w_{1,1}(t) \\ y_{1,2}(t) &= h^1_{1,2}(\tau,t) * p_1(t) + h^1_{2,2}(\tau,t) * p_2(t) + h^2_{1,2}(\tau,t) * z_1(t) + h^2_{2,2}(\tau,t) * z_2(t) + w_{1,2}(t) \end{aligned} \quad (5)$$

where " $*$ " denotes the convolution, $(p_1(t), p_2(t))$ and $(z_1(t), z_2(t))$ are mates at time t respectively, each has length N, $(p_1(t), p_2(t))$ and $(z_1(t), z_2(t))$ are weakly correlated with IFW $W_0/2$, both $h^i_{1,j}(\tau,t)$ and $h^i_{2,j}(\tau,t)$ 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 taking respective cross correlation between received signals and LS codes. The channel parameters can be written as follow;

$$
\begin{aligned}
\tilde{h}_{1,1}^{1} &= R_{y_{1,1},p_1}(m) = h_{1,1}^{1} + R_{h_{2,1}^{1}*p_2,p_1}(m) + R_{h_{1,1}^{2}*z_1,p_1}(m) + R_{h_{2,1}^{2}*z_2,p_1}(m) + R_{w_{1,1},p_1}(m) \\
\tilde{h}_{1,2}^{1} &= R_{y_{1,2},p_1}(m) = h_{1,2}^{1} + R_{h_{2,2}^{1}*p_2,p_1}(m) + R_{h_{1,2}^{2}*z_1,p_1}(m) + R_{h_{2,2}^{2}*z_2,p_1}(m) + R_{w_{1,2},p_1}(m) \\
\tilde{h}_{2,1}^{1} &= R_{y_{1,1},p_2}(m) = h_{2,1}^{1} + R_{h_{1,1}^{1}*p_1,p_2}(m) + R_{h_{2,1}^{2}*z_2,p_2}(m) + R_{h_{1,1}^{2}*z_1,p_2}(m) + R_{w_{1,1},p_2}(m) \\
\tilde{h}_{2,2}^{1} &= R_{y_{1,2},p_2}(m) = h_{2,2}^{1} + R_{h_{1,2}^{1}*p_1,p_2}(m) + R_{h_{2,2}^{2}*z_2,p_2}(m) + R_{h_{1,2}^{2}*z_1,p_2}(m) + R_{w_{1,2},p_2}(m) \\
\tilde{h}_{1,1}^{2} &= R_{y_{1,1},z_1}(m) = h_{1,1}^{2} + R_{h_{1,1}^{1}*p_1,z_1}(m) + R_{h_{2,1}^{2}*z_2,z_1}(m) + R_{h_{2,1}^{1}*p_2,z_1}(m) + R_{w_{1,1},z_1}(m) \\
\tilde{h}_{1,2}^{2} &= R_{y_{1,2},z_1}(m) = h_{1,2}^{2} + R_{h_{1,2}^{1}*p_1,z_1}(m) + R_{h_{2,2}^{2}*z_2,z_1}(m) + R_{h_{2,2}^{1}*p_2,z_1}(m) + R_{w_{1,2},z_1}(m) \\
\tilde{h}_{2,1}^{2} &= R_{y_{1,1},z_2}(m) = h_{2,1}^{2} + R_{h_{1,1}^{1}*p_1,z_2}(m) + R_{h_{2,1}^{1}*p_2,z_2}(m) + R_{h_{1,1}^{2}*z_1,z_2}(m) + R_{w_{1,1},z_2}(m) \\
\tilde{h}_{2,2}^{2} &= R_{y_{1,2},z_2}(m) = h_{2,2}^{2} + R_{h_{1,2}^{1}*p_1,z_2}(m) + R_{h_{2,2}^{1}*p_2,z_2}(m) + R_{h_{1,2}^{2}*z_1,z_2}(m) + R_{w_{1,2},z_2}(m)
\end{aligned} \tag{6}
$$

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 of the two properties, IV.A.3) and IV.A.4), we can not clearly eliminate the multipath channel interference. When four antennas transmit (u_1,u_2,v_1,v_2), group 1, (6) can be represented by following methods;

$$
\begin{aligned}
&\tilde{h}_{1,1}^{1} = h_{1,1}^{1} + \tfrac{1}{2}h_{1,1}'^{2} + \tfrac{1}{2}h_{2,1}'^{2} + \tfrac{1}{2}h_{1,1}''^{2} - \tfrac{1}{2}h_{2,1}''^{2} + A_1,\ \tilde{h}_{1,2}^{1} = h_{1,2}^{1} + \tfrac{1}{2}h_{1,2}'^{2} + \tfrac{1}{2}h_{2,2}'^{2} + \tfrac{1}{2}h_{1,2}''^{2} - \tfrac{1}{2}h_{2,2}''^{2} + A_2 \\
&\tilde{h}_{2,1}^{1} = h_{2,1}^{1} - \tfrac{1}{2}h_{1,1}'^{2} - \tfrac{1}{2}h_{2,1}'^{2} + \tfrac{1}{2}h_{1,1}''^{2} - \tfrac{1}{2}h_{2,1}''^{2} + A_3,\ \tilde{h}_{2,2}^{1} = h_{2,2}^{1} - \tfrac{1}{2}h_{1,2}'^{2} - \tfrac{1}{2}h_{2,2}'^{2} + \tfrac{1}{2}h_{1,2}''^{2} - \tfrac{1}{2}h_{2,2}''^{2} + A_4 \\
&\tilde{h}_{1,1}^{2} = h_{1,1}^{2} + \tfrac{1}{2}h_{1,1}'^{1} + \tfrac{1}{2}h_{2,1}'^{1} + \tfrac{1}{2}h_{1,1}''^{1} - \tfrac{1}{2}h_{2,1}''^{1} + A_5,\ \tilde{h}_{1,2}^{2} = h_{1,2}^{2} + \tfrac{1}{2}h_{1,2}'^{1} + \tfrac{1}{2}h_{2,2}'^{1} + \tfrac{1}{2}h_{1,2}''^{1} - \tfrac{1}{2}h_{2,2}''^{1} + A_6 \\
&\tilde{h}_{2,1}^{2} = h_{2,1}^{2} - \tfrac{1}{2}h_{1,1}'^{1} - \tfrac{1}{2}h_{2,1}'^{1} + \tfrac{1}{2}h_{1,1}''^{1} - \tfrac{1}{2}h_{2,1}''^{1} + A_7,\ \tilde{h}_{2,2}^{2} = h_{2,2}^{2} - \tfrac{1}{2}h_{1,2}'^{1} - \tfrac{1}{2}h_{2,2}'^{1} + \tfrac{1}{2}h_{1,2}''^{1} - \tfrac{1}{2}h_{2,2}''^{1} + A_8
\end{aligned} \tag{7}
$$

where $h_{1,j}'^{i}$ and $h_{2,j}'^{i}$ are part of the each channel from $W_0/4$ to $W_0/2$ delay and have bad effects on part of the $h_{1,j}^{i}$ and $h_{2,j}^{i}$ from zero to $W_0/4$ delay for $i,j=1,2$, $h_{1,j}''^{i}$ and $h_{2,j}''^{i}$ are part of the each channel from zero to $W_0/4$ delay and have bad effects on part of the $h_{1,j}^{i}$ and $h_{2,j}^{i}$ from $W_0/4$ to $W_0/2$ delay for $i,j=1,2$, $A_k = R_{w_{1,i}^{1},u_j}(m)$ for $i,j=1,2,\ k=1,\ldots,4,$ $A_k = R_{w_{1,i}^{1},v_j}(m)$ for $i,j=1,2,\ k=5,\ldots,8$. Fig. 7 shows the meanings of h' and h''. We also define following equations;

$$
\begin{aligned}
&\tilde{h}_{1}^{1} = \tilde{h}_{1,1}^{1} - \tilde{h}_{2,1}^{1} = h_{1,1}^{1} - h_{2,1}^{1} + h_{1,1}'^{2} + h_{2,1}'^{2} + B_1,\ \tilde{h}_{2}^{1} = \tilde{h}_{1,2}^{1} - \tilde{h}_{2,2}^{1} = h_{1,2}^{1} - h_{2,2}^{1} + h_{1,2}'^{2} + h_{2,2}'^{2} + B_2 \\
&\tilde{h}_{3}^{1} = \tilde{h}_{1,1}^{1} + \tilde{h}_{2,1}^{1} = h_{1,1}^{1} + h_{2,1}^{1} + h_{1,1}''^{2} - h_{2,1}''^{2} + B_3,\ \tilde{h}_{4}^{1} = \tilde{h}_{1,2}^{1} + \tilde{h}_{2,2}^{1} = h_{1,2}^{1} + h_{2,2}^{1} + h_{1,2}''^{2} - h_{2,2}''^{2} + B_4 \\
&\tilde{h}_{1}^{2} = \tilde{h}_{1,1}^{2} - \tilde{h}_{2,1}^{2} = h_{1,1}^{2} - h_{2,1}^{2} + h_{1,1}'^{1} + h_{2,1}'^{1} + B_5,\ \tilde{h}_{2}^{2} = \tilde{h}_{1,2}^{2} - \tilde{h}_{2,2}^{2} = h_{1,2}^{2} - h_{2,2}^{2} + h_{1,2}'^{1} + h_{2,2}'^{1} + B_6 \\
&\tilde{h}_{3}^{2} = \tilde{h}_{1,1}^{2} + \tilde{h}_{2,1}^{2} = h_{1,1}^{2} + h_{2,1}^{2} + h_{1,1}''^{1} - h_{2,1}''^{1} + B_7,\ \tilde{h}_{4}^{2} = \tilde{h}_{1,2}^{2} + \tilde{h}_{2,2}^{2} = h_{1,2}^{2} + h_{2,2}^{2} + h_{1,2}''^{1} - h_{2,2}''^{1} + B_8
\end{aligned} \tag{8}
$$

where $B_k = R_{w_{1,i}^{1},u_1}(m) \pm R_{w_{1,i}^{1},u_2}(m)$ for $i=1,2,\ k=1,\ldots,4,$ $B_k = R_{w_{1,i}^{1},v_1}(m) \pm R_{w_{1,i}^{1},v_2}(m)$ for $i=1,2,\ k=5,\ldots,8.$ When four antennas transmit $(u_1,-u_2,-v_1,v_2)$, group 2, (6) can be represented by following methods;

$$
\begin{aligned}
&\hat{h}_{1,1}^{1} = h_{1,1}^{1} - \tfrac{1}{2}h_{1,1}'^{2} + \tfrac{1}{2}h_{2,1}'^{2} - \tfrac{1}{2}h_{1,1}''^{2} - \tfrac{1}{2}h_{2,1}''^{2} + D_1,\ \hat{h}_{1,2}^{1} = h_{1,2}^{1} - \tfrac{1}{2}h_{1,2}'^{2} + \tfrac{1}{2}h_{2,2}'^{2} - \tfrac{1}{2}h_{1,2}''^{2} - \tfrac{1}{2}h_{2,2}''^{2} + D_2 \\
&\hat{h}_{2,1}^{1} = h_{2,1}^{1} - \tfrac{1}{2}h_{1,1}'^{2} + \tfrac{1}{2}h_{2,1}'^{2} + \tfrac{1}{2}h_{1,1}''^{2} + \tfrac{1}{2}h_{2,1}''^{2} + D_3,\ \hat{h}_{2,2}^{1} = h_{2,2}^{1} - \tfrac{1}{2}h_{1,2}'^{2} + \tfrac{1}{2}h_{2,2}'^{2} + \tfrac{1}{2}h_{1,2}''^{2} + \tfrac{1}{2}h_{2,2}''^{2} + D_4 \\
&\hat{h}_{1,1}^{2} = h_{1,1}^{2} - \tfrac{1}{2}h_{1,1}'^{1} + \tfrac{1}{2}h_{2,1}'^{1} - \tfrac{1}{2}h_{1,1}''^{1} - \tfrac{1}{2}h_{2,1}''^{1} + D_5,\ \hat{h}_{1,2}^{2} = h_{1,2}^{2} - \tfrac{1}{2}h_{1,2}'^{1} + \tfrac{1}{2}h_{2,2}'^{1} - \tfrac{1}{2}h_{1,2}''^{1} - \tfrac{1}{2}h_{2,2}''^{1} + D_6 \\
&\hat{h}_{2,1}^{2} = h_{2,1}^{2} - \tfrac{1}{2}h_{1,1}'^{1} + \tfrac{1}{2}h_{2,1}'^{1} + \tfrac{1}{2}h_{1,1}''^{1} + \tfrac{1}{2}h_{2,1}''^{1} + D_7,\ \hat{h}_{2,2}^{2} = h_{2,2}^{2} - \tfrac{1}{2}h_{1,2}'^{1} + \tfrac{1}{2}h_{2,2}'^{1} + \tfrac{1}{2}h_{1,2}''^{1} + \tfrac{1}{2}h_{2,2}''^{1} + D_8
\end{aligned} \tag{9}
$$

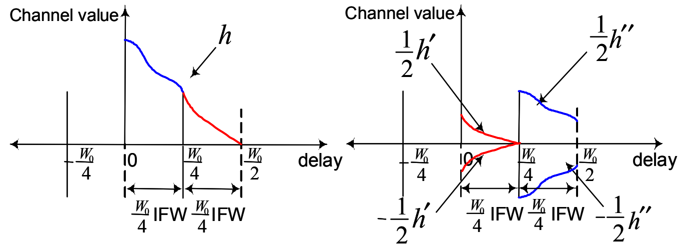


Figure 7. The meanings of h' and h''

where $D_k = R_{w_{1,i}^{2},u_1}(m)$ for $i=1,2,\ k=1,2,$ $D_k = R_{w_{1,i}^{2},-u_2}(m)$ for $i=1,2,\ k=3,4,$ $D_k = R_{w_{1,i}^{2},-v_1}(m)$ for $i=1,2,\ k=5,6,$ $D_k = R_{w_{1,i}^{2},v_2}(m)$ for $i=1,2,\ k=7,8$. We also define following equations;

$$
\begin{aligned}
&\hat{h}_{1}^{1} = \hat{h}_{1,1}^{1} - \hat{h}_{2,1}^{1} = h_{1,1}^{1} - h_{2,1}^{1} - h_{1,1}''^{2} - h_{2,1}''^{2} + E_1,\ \hat{h}_{2}^{1} = \hat{h}_{1,2}^{1} - \hat{h}_{2,2}^{1} = h_{1,2}^{1} - h_{2,2}^{1} - h_{1,2}''^{2} - h_{2,2}''^{2} + E_2 \\
&\hat{h}_{3}^{1} = \hat{h}_{1,1}^{1} + \hat{h}_{2,1}^{1} = h_{1,1}^{1} + h_{2,1}^{1} - h_{1,1}'^{2} + h_{2,1}'^{2} + E_3,\ \hat{h}_{4}^{1} = \hat{h}_{1,2}^{1} + \hat{h}_{2,2}^{1} = h_{1,2}^{1} + h_{2,2}^{1} - h_{1,2}'^{2} + h_{2,2}'^{2} + E_4 \\
&\hat{h}_{1}^{2} = \hat{h}_{1,1}^{2} - \hat{h}_{2,1}^{2} = h_{1,1}^{2} - h_{2,1}^{2} - h_{1,1}''^{1} - h_{2,1}''^{1} + E_5,\ \hat{h}_{2}^{2} = \hat{h}_{1,2}^{2} - \hat{h}_{2,2}^{2} = h_{1,2}^{2} - h_{2,2}^{2} - h_{1,2}''^{1} - h_{2,2}''^{1} + E_6 \\
&\hat{h}_{3}^{2} = \hat{h}_{1,1}^{2} + \hat{h}_{2,1}^{2} = h_{1,1}^{2} + h_{2,1}^{2} - h_{1,1}'^{1} + h_{2,1}'^{1} + E_7,\ \hat{h}_{4}^{2} = \hat{h}_{1,2}^{2} + \hat{h}_{2,2}^{2} = h_{1,2}^{2} + h_{2,2}^{2} - h_{1,2}'^{1} + h_{2,2}'^{1} + E_8
\end{aligned} \tag{10}
$$

where $E_k = R_{w_{1,i}^{2},u_1}(m) \pm R_{w_{1,i}^{2},-u_2}(m)$ for $i=1,2,k=1,\ldots,4,$ $E_k = R_{w_{1,i}^{2},-v_1}(m) \pm R_{w_{1,i}^{2},v_2}(m)$ for $i=1,2,k=5,\ldots,8$. Applying (10) to (7), we can get the first reestimated channels. We can describe the first part of the MCIC as follow;

$$
\begin{aligned}
&\widehat{h}_{1,1}^{1} = \tilde{h}_{1,1}^{1} - \tfrac{1}{2}\hat{h}_{3}'^{2} - \tfrac{1}{2}\hat{h}_{1}''^{2},\ \widehat{h}_{1,2}^{1} = \tilde{h}_{1,2}^{1} - \tfrac{1}{2}\hat{h}_{4}'^{2} - \tfrac{1}{2}\hat{h}_{2}''^{2},\ \widehat{h}_{2,1}^{1} = \tilde{h}_{2,1}^{1} + \tfrac{1}{2}\hat{h}_{3}'^{2} - \tfrac{1}{2}\hat{h}_{1}''^{2} \\
&\widehat{h}_{2,2}^{1} = \tilde{h}_{2,2}^{1} + \tfrac{1}{2}\hat{h}_{4}'^{2} - \tfrac{1}{2}\hat{h}_{2}''^{2},\ \widehat{h}_{1,1}^{2} = \tilde{h}_{1,1}^{2} - \tfrac{1}{2}\hat{h}_{3}'^{1} - \tfrac{1}{2}\hat{h}_{1}''^{1},\ \widehat{h}_{1,2}^{2} = \tilde{h}_{1,2}^{2} - \tfrac{1}{2}\hat{h}_{4}'^{1} - \tfrac{1}{2}\hat{h}_{2}''^{1} \\
&\widehat{h}_{2,1}^{2} = \tilde{h}_{2,1}^{2} + \tfrac{1}{2}\hat{h}_{3}'^{1} - \tfrac{1}{2}\hat{h}_{1}''^{1},\ \widehat{h}_{2,2}^{2} = \tilde{h}_{2,2}^{2} + \tfrac{1}{2}\hat{h}_{4}'^{1} - \tfrac{1}{2}\hat{h}_{2}''^{1}
\end{aligned} \tag{11}
$$

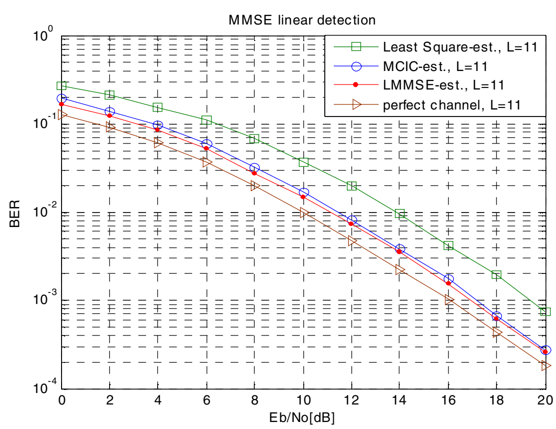
Applying (8) to (9), we can get the second reestimated channels. We can also describe the second part of the MCIC as follow;

$$
\begin{aligned}
&\widehat{\widehat{h}}_{1,1}^{1} = \hat{h}_{1,1}^{1} + \tfrac{1}{2}\tilde{h}_{1}'^{2} + \tfrac{1}{2}\tilde{h}_{3}''^{2},\ \widehat{\widehat{h}}_{1,2}^{1} = \hat{h}_{1,2}^{1} + \tfrac{1}{2}\tilde{h}_{2}'^{2} + \tfrac{1}{2}\tilde{h}_{4}''^{2},\ \widehat{\widehat{h}}_{2,1}^{1} = \hat{h}_{2,1}^{1} + \tfrac{1}{2}\tilde{h}_{1}'^{2} - \tfrac{1}{2}\tilde{h}_{3}''^{2} \\
&\widehat{\widehat{h}}_{2,2}^{1} = \hat{h}_{2,2}^{1} + \tfrac{1}{2}\tilde{h}_{2}'^{2} - \tfrac{1}{2}\tilde{h}_{4}''^{2},\ \widehat{\widehat{h}}_{1,1}^{2} = \hat{h}_{1,1}^{2} + \tfrac{1}{2}\tilde{h}_{1}'^{1} + \tfrac{1}{2}\tilde{h}_{3}''^{1},\ \widehat{\widehat{h}}_{1,2}^{2} = \hat{h}_{1,2}^{2} + \tfrac{1}{2}\tilde{h}_{2}'^{1} + \tfrac{1}{2}\tilde{h}_{4}''^{1} \\
&\widehat{\widehat{h}}_{2,1}^{2} = \hat{h}_{2,1}^{2} + \tfrac{1}{2}\tilde{h}_{1}'^{1} - \tfrac{1}{2}\tilde{h}_{3}''^{1},\ \widehat{\widehat{h}}_{2,2}^{2} = \hat{h}_{2,2}^{2} + \tfrac{1}{2}\tilde{h}_{2}'^{1} - \tfrac{1}{2}\tilde{h}_{4}''^{1}
\end{aligned} \tag{12}
$$

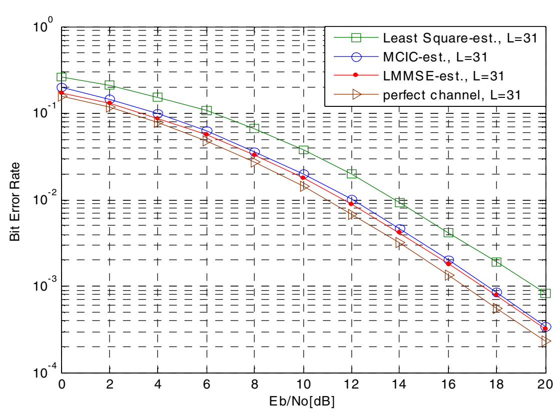
V. SIMULATION RESULTS

We consider a DSTTD-OFDM system with four transmission and two receive antennas. The parameters of the simulation are as follow. The entire channel bandwidth W, 20MHz, 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. Two channel models are used, namely, Channel A with $\tau_{rms}=50$ ns, and Channel B with $\tau_{rms}=150$ ns. Both channels assume an exponentially decaying power delay profile with 11multipaths and 31 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 1 summarizes both training symbols per transmission antenna and total training periods of the channel estimators.

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. 8 (a) and (b) show the comparative BER performance for Least Square estimation, LMMSE estimation, and the MCIC estimation. From Fig. 8 (a) at the target BER of 10^{-3}, we can see a about 2.3dB gain and 0.2dB loss in E_b / N_0 for the MCIC estimation over Least Square and LMMSE estimations respectively. From Fig. 8 (b) at the target BER of 10^{-3}, we can also see a about 2dB gain and 0.2dB loss in E_b / N_0 for the MCIC estimation over Least Square and LMMSE estimations respectively. From Fig. 8 (a) and (b), The MCIC estimator performs within 1 dB limit of the system with known channel. Fig. 8 (a) and (b) show that the MCIC estimator substantially reduces multipath channel interference in the channel with relatively small or large delay spread.



(a) The MCIC, channel A



(b) The MCIC, channel B

Figure 8. BER performance of Proposed MCIC for 4×2 DSTTD-OFDM system

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

Estimators	Least square	LMMSE	MCIC
Training symbols per transmission antenna	$1\ N_s$	$1\ N_s$	$4\ N_s$
Total training periods	$4\ T_s$	$4\ T_s$	$4\ T_s$

VI. CONCLUSIONS

In this paper, LS codes assisted MCIC is proposed for MIMO-OFDM systems. The proposed MCIC uses both weakly correlated LS codes and interference cancellation method in the presence of channel with relatively small or large delay spread. The simulation results show that the MCIC estimator outperforms least square estimator, although it has slightly inferior performance as compared with LMMSE estimator. Moreover the MCIC estimator has the advantage of not requiring any information about the channel statistics like auto-correlation of channels. As a result, the proposed MCIC estimator is more efficient for reducing multipath channel interference than least square and LMMSE estimators in the channel with relatively small or large delay spread.

REFERENCES

[1] S. M. Alamouti, "A Simple Transmit Diversity Technique for Wireless Communications," *IEEE Journal on Selected Areas in Communications,* vol. 16, no. 8, pp. 1451-1458, Oct. 1998.

[2] J. K. Kim, R. W. Heath, and E. J. powers, "Receiver Designs for Alamouti Coded OFDM Systems in Fast Fading Channels," *IEEE Transaction on wireless Communications*, vol. 4, no. 2, pp. 550-559, Mar. 2005.

[3] O. Edfors, M. Sandell, J. J van de Beek, S. K. Wilson, and P. O. Borjesson, "OFDM channel estimation by singular value decomposition," *IEEE Transaction on Communications*, vol. 46, pp. 931-939, July 1998.

[4] M. J. E. Golay, "Complementary series," *IRE Transactions on Information Theory,* IT-7, pp. 82-87, Apr. 1961.

[5] S. Stanczak, H. Boche, and M. Haardt, "Are LAS-codes a miracle**?,**" Global Telecommunications Conference, 2001. *GLOBECOM '01. IEEE* vol. 1, pp. 589-593, 25-29, Nov. 2001.

[6] B. Choi and L. Hanzo, "On the design of LAS spreading codes," Vehicular Technology Conference, 2002. Proceedings. *VTC 2002-Fall 2002 IEEE 56th ,* vol. 4, pp. 2172-2176, 24-28, Sep. 2002.

[7] D. Li, "A High Spectrum Efficient Multiple Access Code," Fifth Asia-Pacific Conference on Communications and Fourth Optoelectronics and Communications Conference, vol. 1, pp. 598-605, 18-22 Oct. 1999.

[8] A. F. Naguib, N. Seshadri, and A. R. Calderbank, "Applications of space-time block codes and interference suppression for high capacity and high data rate wireless systems," *Signals, Systems & Computers, 1998. Conference Record of the Thirty-Second Asilomar Conference on,* vol. 2, pp. 1803-1810, Nov. 1998.

[9] T. S. Rapparport, *Wireless Communication: Principles and Practice*, Prentice Hall PTR, 1996.