BLIND CHANNEL ESTIMATION AND EQUALIZATION IN OFDM SYSTEM WITH CIRCULAR PRECODING

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ABSTRACT

This paper proposes a reliable and efficient blind channel estimation method and a linearly constrained minimum variance (LCMV) receiver with circular precoding for single transmitter and receiver antenna orthogonal frequency division multiplexing (OFDM) system. The proposed blind channel estimation method is based on the interesting cross-correlation property of the cyclic prefix (CP) in an OFDM symbol. Toeplitz structure induced in the cross-correlation matrix allows us to employ a simple and reliable method for extracting channel impulse response effectively. The LCMV optimization criterion is applied for symbol detection of the OFDM symbols. At the same time, a circular precoding scheme for minimizing the maximal power of the dominant interference plus noise in LCMV receiver is considered. In the simulation study, proposed method is compared with other circularly precoded scheme and it demonstrate the superior performance of the proposed method.

1. INTRODUCTION

OFDM has brought about high speed data transmission by effectively handling multipath effect. The OFDM has become a standards in a number of high data transmission such as satellite and terrestrial digital audio broadcasting (DAB), digital terrestrial TV broadcasting (DVB), asymmetric digital subscriber line (ADSL) for high bit-rate digital subscriber services, IEEE 802.11a, .11g, .11n WLAN standards, and IEEE 802.16d, .16e WMAN standards.

In OFDM, since the individual subcarrier signal spectra are affected by frequency-flat rather than frequency-selective fading, equalization is drastically simplified. Pilots-based methods enable channel estimation at the receiver [1], [2]. However, in terms of spectral efficiency, blind channel estimation methods are more preferable than pilots based scheme. Statistical methods for blind channel estimation making use of the cyclostationary induced in the cyclic prefix (CP) are applied on pre-DFT received data [3], [4], [5], [6].

Precoding schemes are employed in order to increase multipath diversity and enable blind channel estimation. Precoding approach at frequency domain is an effective method to create frequency diversity to avoid channel nulls [7], [8]. Also, for blind channel estimation, [9] use the redundant precoding and [10], [11] utilize non-redundant precoding schemes. [12] proposes channel independent precoder for BER minimization. A circular precoding scheme, [13] gives us a simple transmitter structure and blind channel estimation method with singular value decomposition (SVD) at the receiver. We propose efficient blind channel estimation method using CP on pre-DFT received data and a circularly precoded LCMV receiver.

2. SIGNAL MODEL

Let $\boldsymbol{s}_i = [s_i(0) \dots s_i(N-1)]^T$ denote the information symbols of the i-th transmission block. Assume that $s_i(k)$ for $k = 0, \dots, N-1$ are uncorrelated zero mean, i.i.d., and white with ε_s variance. The i-th block, $\boldsymbol{s}_i$ is first transformed by linear precoder and modulated by IDFT matrix, i.e.

$$[u_i(0) \dots u_i(N-1)]^T = \boldsymbol{F}_N^{\boldsymbol{H}} \boldsymbol{C} \boldsymbol{s}_i \qquad (1)$$

where $\boldsymbol{C}$ is an $N \times N$ circular precoding matrix, which satisfies $tr\{E[\boldsymbol{C}\boldsymbol{s}_i\boldsymbol{s}_i^H\boldsymbol{C}^H]\} = \varepsilon_s N$ for maintaining the power at each blocks. $tr\{\cdot\}$ is a trace operator and $\boldsymbol{F}_N$ stands for the size N-DFT matrix with entries of $\boldsymbol{F}_{Nn,k} = e^{j2\pi nk/N}/\sqrt{N}$. For unique decoderbility, $\boldsymbol{C}$ should be full rank. Then L cyclic prefix(CP) samples are added to each i-th block, resulting in a size $P = L + N$ vector $\boldsymbol{u}_i$ where $u_i(-L+n) = u_i(N-L+n)$ for $n = 0, \dots, L-1$. It is assumed that $P = qL$ and $N = (q-1)L$ for $q \geq 2$ where q is positive integer. Then $\boldsymbol{u}_i$ can be split into q sub-blocks of length L

$$\begin{aligned} \boldsymbol{u}_i &= [\boldsymbol{u}_{i,0}^T \dots \boldsymbol{u}_{i,q-1}^T]^T \\ \boldsymbol{u}_{i,b} &= [u_i(bL+0) \dots u_i(bL+L-1)]^T. \end{aligned} \qquad (2)$$

Sub-block of $\boldsymbol{u}_{i,0}$ is corresponding to $\boldsymbol{u}_{i,q-1}$ due to the CP.

The channel effects are modelled by a FIR filter with $\boldsymbol{h} = [h_0 \dots h_{L-1}]^T$ and noise sample v. Assume that the actual channel length is M and the OFDM system is designed such that the CP is longer than the channel length, i.e. $M \leq L$, which implies that $h_i = 0$ for $i \geq M$. Let $\boldsymbol{H}$ be the $P \times (P+L)$ Toeplitz matrix with the first column $[0 \dots 0]^T$ and the first row $[0 \ h_{L-1} \dots h_0 \ 0 \dots 0]$. Then the received i-th block can be represented as below

$$\boldsymbol{y}_i = \boldsymbol{H}\ [\ \boldsymbol{u}_{i-1,q-1}^T\ \ \boldsymbol{u}_i^T\ ]^T + \boldsymbol{v}_i \tag{3}$$

where $\boldsymbol{u}_{i-1,q-1}$ is $q-1$-th sub-block of $\boldsymbol{u}_{i-1}$ and $\boldsymbol{v}_i = [v_i(0)\ldots v_i(qL-1)]^T$ models the noise. We assume that $v_i(n)$ is complex Gaussian white across subcarriers and across blocks and zero mean and σ^2-variance.

3. BLIND CHANNEL ESTIMATION METHOD

We propose blind channel estimation method utilizing the repeated structure of CP sub-blocks. In order to describe the method efficiently, we define some notations. Let an arbitrary matrix A, square matrix with dimension L, be the circular matrix with the first column is $[h_0 \ldots h_{L-1}]^T$. Then, define lower triangular part of matrix A as $\boldsymbol{H}_1$ and upper off-diagonal part of matrix A as $\boldsymbol{H}_2$ as below,

$$\begin{aligned}
\boldsymbol{H}_1 &= \begin{bmatrix} h_0 & 0 & \cdots & 0 \\ h_1 & h_0 & & 0 \\ \vdots & \vdots & \ddots & 0 \\ h_{L-1} & h_{L-2} & \cdots & h_0 \end{bmatrix}, \\
\boldsymbol{H}_2 &= \begin{bmatrix} 0 & h_{L-1} & \cdots & h_1 \\ 0 & 0 & \ddots & h_2 \\ \vdots & \vdots & \ddots & h_{L-1} \\ 0 & 0 & \cdots & 0 \end{bmatrix}.
\end{aligned} \tag{4}$$

In addition, define length $2L$ vector of $[\boldsymbol{u}_{i-1,q-1}^T\ \boldsymbol{u}_{i,0}^T]^T$ as $\boldsymbol{g}_{i,0}(0)$ and vector delayed from $\boldsymbol{g}_{i,0}(0)$ by $L-1$ as $\boldsymbol{g}_{i,0}(L-1)$, equal to $[u_{i-1,q-1}(L-1)\ \boldsymbol{u}_{i,0}^T\ u_{i,1}(0)\ \cdots\ u_{i,1}(L-2)]^T$. Similarly, define $[\boldsymbol{u}_{i,q-2}^T\ \boldsymbol{u}_{i,q-1}^T]^T$ as $\boldsymbol{g}_{i,q-1}(0)$ and vector delayed from $\boldsymbol{g}_{i,q-1}(0)$ by $L-1$ as $\boldsymbol{g}_{i,q-1}(L-1)$, $[u_{i,q-2}(L-1)\ \boldsymbol{u}_{i,q-1}^T\ u_{i+1,0}(0)\ \cdots\ u_{i+1,0}(L-2)]^T$. The received data vectors corresponding to $\boldsymbol{g}_{i,0}(d)$ and to $\boldsymbol{g}_{i,q-1}(d)$ for $d = 0, L-1$ are denoted by $\boldsymbol{y}_{i,0}(d)$ and $\boldsymbol{y}_{i,q-1}(d)$ for $d = 0, L-1$ respectively : i.e.

$$\begin{aligned}
\boldsymbol{y}_{i,0}(d) &= [\boldsymbol{H}_2\ \boldsymbol{H}_1]\boldsymbol{g}_{i,0}(d) + \boldsymbol{v}_{i,0}(d) \\
\boldsymbol{y}_{i,q-1}(d) &= [\boldsymbol{H}_2\ \boldsymbol{H}_1]\boldsymbol{g}_{i,q-1}(d) + \boldsymbol{v}_{i,q-1}(d)
\end{aligned} \tag{5}$$

, where

$$\begin{aligned}
\boldsymbol{v}_{i,0}(d) &\equiv [v_i(d)\ldots v_i(d+L-1)]^T \\
\boldsymbol{v}_{i,q-1}(d) &\equiv [v_i((q-1)L+d)\ldots v_i(qL-1+d)]^T
\end{aligned} \tag{6}$$

The proposed blind channel estimation method is based on the cross-correlation between $\boldsymbol{y}_{i,0}(d)$ and $\boldsymbol{y}_{i,q-1}(d)$ for $d = 0, L-1$. The cross correlation matrix of the received block pair, $\boldsymbol{y}_{i,0}(d)$ and $\boldsymbol{y}_{i,q-1}(d)$ for $d = 0, L-1$, can be written down as

$$\begin{aligned}
\boldsymbol{R}_0 &\equiv E\{\boldsymbol{y}_{i,0}(0)\boldsymbol{y}_{i,q-1}(0)^H\} \\
&= \boldsymbol{H}_1 E\{\boldsymbol{u}_{i,0}\boldsymbol{u}_{i,q-1}^{\boldsymbol{H}}\}\boldsymbol{H}_1^H \\
&= \boldsymbol{H}_1\boldsymbol{H}_1^H
\end{aligned} \tag{7}$$

$$\begin{aligned}
\boldsymbol{R}_{q-1} &\equiv E\{\boldsymbol{y}_{i,0}(L-1)\boldsymbol{y}_{i,q-1}(L-1)^H\} \\
&= [\boldsymbol{H}_2\boldsymbol{P}\ \ \boldsymbol{H}_1\boldsymbol{e}_1]E\{\boldsymbol{u}_{i,0}\boldsymbol{u}_{i,q-1}^H\}[\boldsymbol{H}_2\boldsymbol{P}\ \ \boldsymbol{H}_1\boldsymbol{e}_1]^H \\
&= [\boldsymbol{H}_2\boldsymbol{P}\ \ \boldsymbol{H}_1\boldsymbol{e}_1][\boldsymbol{H}_2\boldsymbol{P}\ \ \boldsymbol{H}_1\boldsymbol{e}_1]^H
\end{aligned} \tag{8}$$

where $\boldsymbol{P}$ is $L \times L$ lower triangular Toeplitz matrix with the first column is $[0\ 1\ 0 \cdots 0]^T$ and $\boldsymbol{e}_1$ is length L unit column vector with the first element is 1, $[1\ 0\ \cdots\ 0]^T$. It is noteworthy that cross-correlation cancels the noise power.

In practice, cross-correlation matrix $\boldsymbol{R}_0$ and $\boldsymbol{R}_{q-1}$ are approximated by sample averaging on the basis of finite M blocks

$$\begin{aligned}
\boldsymbol{R}_0 &= \tfrac{1}{M}\textstyle\sum_{i=0}^{M-1} \boldsymbol{y}_{i,0}(0)\boldsymbol{y}_{i,q-1}(0)^H \\
\boldsymbol{R}_{q-1} &= \tfrac{1}{M}\textstyle\sum_{i=0}^{M-1} \boldsymbol{y}_{i,0}(L-1)\boldsymbol{y}_{i,q-1}(L-1)^H.
\end{aligned} \tag{9}$$

Since $\boldsymbol{H}_1$ is lower triangular Toeplitz matrix, it would be possible to extract channel impulse response, $\boldsymbol{h}$, from $\boldsymbol{R}_0$ using Cholesky factorization. [6] introduces similar approach with Cholesky factorization. However, since Cholesky factorization which, in practice, breaks down frequently due to non-positive definiteness resulted from noise in correlation matrix, it is quietly restricted to be applied. Therefore, Cholesky based approach is not feasible for blind channel estimation. The method in [6] relies on an autocorrelation matrix, which computes correlation for the entire $\boldsymbol{y}_i$ vector whose size is $P \times 1$. And in order to extract valid channel impulse response, it must be assumed that the first channel coefficient is $h_0 \neq 0$. Differently, proposed scheme generates two different cross-correlation matrix, $\boldsymbol{R}_0$ and $\boldsymbol{R}_{q-1}$ whose size is $L = P/q$ and does not assume $h_0 \neq 0$. In order to extract channel information, we compute

$$\begin{aligned}
\boldsymbol{R}_{0,h} &= \boldsymbol{R}_0 - \boldsymbol{P}\boldsymbol{R}_0\boldsymbol{P}^T \\
\boldsymbol{R}_{q-1,h} &= \boldsymbol{R}_{q-1} - \boldsymbol{P}^T\boldsymbol{R}_{q-1}\boldsymbol{P}.
\end{aligned} \tag{10}$$

Then, equal gain combining these two matrix extracts outer product of estimated channel impulse response of $\hat{\boldsymbol{h}}$

$$\hat{\boldsymbol{h}}\hat{\boldsymbol{h}}^H = \frac{\boldsymbol{R}_{0,h} + \boldsymbol{R}_{q-1,h}}{2}. \tag{11}$$

Since the dimension of the above eq(11) is L, which is several fractions of P, and it has only one dominant singular value, if we apply iterative SVD algorithm, it will converge quickly to one dominant singular value with its corresponding singular vector. The phase ambiguity can be easily dealt with using a single training symbol or making it independent by using differential modulation.

4. LCMV BASED EQUALIZATION

The general idea of blind LCMV equalization is to minimize the output power of the receiver subject to given constraint. We apply LCMV optimization procedure to the OFDM receiver before DFT. It can be formulated as below

$$\boldsymbol{w}_k = \boldsymbol{Arg\ min}\ \boldsymbol{w}_k^H\boldsymbol{R}_x\boldsymbol{w}_k\ \ subject\ to\ \ c_{dk}^*\hat{\boldsymbol{h}}^H\boldsymbol{w}_k = 1 \tag{12}$$

where $\boldsymbol{R}_x = E\{\boldsymbol{x}_i\boldsymbol{x}_i^H\}$ is the covariance matrix of $\boldsymbol{x}_i$, which is the i-th received block after removal of the CP from $\boldsymbol{y}_i$ resulting in $\boldsymbol{x}_i = \boldsymbol{H}_c[\boldsymbol{u}_{i,1}^T \dots u_{i,q-1}^T]^T + \boldsymbol{v}_i$, where $\boldsymbol{H}_c$ is $N \times N$ circular matrix with its first column is $[\boldsymbol{h}^T\ 0 \dots 0]^T$, c_{dk} is the k-th diagonal element of diagonal matrix of $\boldsymbol{C}_D = \boldsymbol{F}_N^{\boldsymbol{H}}\boldsymbol{C}\boldsymbol{F}_N$, and $\boldsymbol{w}_k$ indicates equalizing weight vector for recovering the k-th information symbol of $\boldsymbol{s}_i(k)$. It is well known that optimal solution for eq(12) is

$$\begin{aligned} \boldsymbol{w}_k &= \boldsymbol{R}_x^{-1}\hat{\boldsymbol{h}}c_{dk}(c_{dk}^*\hat{\boldsymbol{h}}^H R_x^{-1}\hat{\boldsymbol{h}}c_{dk})^{-1} \\ &= \boldsymbol{R}_x^{-1}\hat{\boldsymbol{h}}(c_{dk}^*\hat{\boldsymbol{h}}^H R_x^{-1}\hat{\boldsymbol{h}})^{-1} \qquad (13) \end{aligned}$$

Once we get vector $\boldsymbol{w}_0$, equalizing weights for the rest part of $\boldsymbol{s}_i$ are easily obtained by altering c_{d0} in eq(13) as c_{dk} for $k = 1, \dots, N-1$ and shifting obtained vector $\boldsymbol{w}_k$ circularly by index k because of the circularity of channel matrix H_c [14]. Let $\boldsymbol{W} \equiv [\boldsymbol{w}_0\ \dots\ \boldsymbol{w}_{N-1}]$ be the equalizing matrix.

5. CIRCULAR PRECODING CRITERION

We design the circular precoding matrix $\boldsymbol{C}$ in a sense of minimizing the maximal power of interference plus noise in LCMV output vector $\tilde{\boldsymbol{x}}_i = \boldsymbol{W}^{\boldsymbol{H}}\boldsymbol{x}_i$. For this purpose, we conduct the interference plus noise analysis in LCMV receiver and then obtain asymptotic expansion for $\tilde{\boldsymbol{x}}_i$. The equalizing matrix, $\boldsymbol{W}^{\boldsymbol{H}}$ can be rewritten as below

$$\boldsymbol{W}^{\boldsymbol{H}} = [\boldsymbol{diag}(\boldsymbol{C}_{\boldsymbol{D}}^{\boldsymbol{H}}\boldsymbol{H}_{\boldsymbol{c}}^{\boldsymbol{H}}\boldsymbol{R}_{\boldsymbol{x}}^{\boldsymbol{-1}}\boldsymbol{H}_{\boldsymbol{c}}\boldsymbol{C}_{\boldsymbol{D}})]^{-\boldsymbol{H}}\boldsymbol{C}_{\boldsymbol{D}}^{\boldsymbol{H}}\boldsymbol{H}_{\boldsymbol{c}}^{\boldsymbol{H}}\boldsymbol{R}_{\boldsymbol{x}}^{\boldsymbol{-1}} \quad (14)$$

, where $\boldsymbol{diag}(\boldsymbol{A})$ means a diagonal matrix consisted of diagonal elements of $\boldsymbol{A}$. Since $\boldsymbol{H}_c$ is circular matrix, $\boldsymbol{H}_D = \boldsymbol{F}_N\boldsymbol{H}_c\boldsymbol{F}_N^H$ holds, where $\boldsymbol{H}_{\boldsymbol{D}}$ is a diagonal matrix. Let SVD of matrix $\boldsymbol{H}_{\boldsymbol{D}}\boldsymbol{C}$ as $\boldsymbol{U}\sum(\sqrt{\lambda_i})\boldsymbol{V}^{\boldsymbol{H}}$. Then, $\boldsymbol{H}_{\boldsymbol{D}}\boldsymbol{C}\boldsymbol{C}^{\boldsymbol{H}}\boldsymbol{H}_{\boldsymbol{D}}^{\boldsymbol{H}}$ can be represented as $\boldsymbol{U}\sum(\lambda_i)\boldsymbol{U}^{\boldsymbol{H}}$ where $\sum(\lambda_i)$ is diagonal matrix with its diagonals are λ_i for $i = 0, \dots, N-1$, singular values of $\boldsymbol{H}_{\boldsymbol{D}}\boldsymbol{C}\boldsymbol{C}^{\boldsymbol{H}}\boldsymbol{H}_{\boldsymbol{D}}^{\boldsymbol{H}}$. Then, it is straightforward that

$$\boldsymbol{R}_x = \varepsilon_s\boldsymbol{F}_N^H\boldsymbol{U}\boldsymbol{diag}(\lambda_i + \frac{\sigma^2}{\varepsilon_s})\boldsymbol{U}^H\boldsymbol{F}_N$$

$$\boldsymbol{C}_{\boldsymbol{D}}^{\boldsymbol{H}}\boldsymbol{H}_{\boldsymbol{c}}^{\boldsymbol{H}}\boldsymbol{R}_{\boldsymbol{x}}^{\boldsymbol{-1}}\boldsymbol{H}_{\boldsymbol{c}}\boldsymbol{C}_{\boldsymbol{D}} = \boldsymbol{F}_{\boldsymbol{N}}^{\boldsymbol{H}}\boldsymbol{V}\boldsymbol{diag}(\frac{\frac{\lambda_i}{\varepsilon_s}}{\lambda_i + \frac{\sigma_2}{\varepsilon_s}})\boldsymbol{V}^{\boldsymbol{H}}\boldsymbol{F}_{\boldsymbol{N}} \quad (15)$$

$$where\ \tfrac{1}{\varepsilon_s}\tfrac{\lambda_i}{\lambda_i + \frac{\sigma^2}{\varepsilon_s}} = \tfrac{1}{\varepsilon_s}\big(1 - \tfrac{\sigma^2}{\lambda_i\varepsilon_s} + O(\sigma^4)\big).$$

If we let $\boldsymbol{F}_{\boldsymbol{N}}^{\boldsymbol{H}}\boldsymbol{s}_{\boldsymbol{i}}$ as $\tilde{\boldsymbol{s}}_{\boldsymbol{i}}$, apply eq(15) into eq(14), and expand resulting equation using long division, the asymptotic expansion for $\tilde{\boldsymbol{x}}_{\boldsymbol{i}}$ is

$$\tilde{\boldsymbol{x}}_{\boldsymbol{i}} = \tilde{\boldsymbol{s}}_{\boldsymbol{i}} + \boldsymbol{F}_{\boldsymbol{N}}^{\boldsymbol{H}}\boldsymbol{C}^{\boldsymbol{-1}}\boldsymbol{H}_{\boldsymbol{D}}^{\boldsymbol{-1}}\boldsymbol{F}_{\boldsymbol{N}}\boldsymbol{v}_{\boldsymbol{i}} + O(\sigma^2). \qquad (16)$$

The k-th receiver output $\tilde{x}_i(k)$ can be represented as $\tilde{x}_i(k) = \tilde{s}_i(k) + \tau_i(k)$ where $\tau_i(k) = \sum_{j=0}^{N-1} \boldsymbol{f}_{\boldsymbol{k}}^{\boldsymbol{H}}\boldsymbol{C}^{\boldsymbol{-1}}\boldsymbol{H}_{\boldsymbol{D}}^{\boldsymbol{-1}}\boldsymbol{f}_{\boldsymbol{j}}\, v_i(j) + O(\sigma^2)$ which represents interference plus noise where $\boldsymbol{f}_j$ is the j-th column of $\boldsymbol{F}_N$. The power of $\tau_i(k)$, $E\{|\tau_i(k)|^2\}$, varies depending on the values in $\boldsymbol{H}_{\boldsymbol{D}}$. We design the circular precoding matrix $\boldsymbol{C}$ to be independent of $\boldsymbol{H}_{\boldsymbol{D}}$. The criterion to design precoding matrix $\boldsymbol{C}$, where $\boldsymbol{C}$ is an $N \times N$ circular matrix with its first column is $[\frac{1}{\sqrt{\rho+1}}\ \ -\sqrt{\frac{\rho}{\rho+1}}\sqrt{\frac{1}{N-1}}\ \dots\ -\sqrt{\frac{\rho}{\rho+1}}\sqrt{\frac{1}{N-1}}]^T$ for $0 < \rho < 1$, is to minimize the possible maximum power of the first dominant term in $\tau_i(k)$. The $E\{|\tau_i(k)|^2\}$ can be represented as below

$$\begin{aligned} E\{|\tau_i(k)|^2\} &= \sigma^2 \textstyle\sum_{j=0}^{N-1} |\boldsymbol{f}_{\boldsymbol{k}}^{\boldsymbol{H}}\boldsymbol{C}^{\boldsymbol{-1}}\boldsymbol{H}_{\boldsymbol{D}}^{\boldsymbol{-1}}\boldsymbol{f}_{\boldsymbol{j}}|^2 + O(\sigma^4) \\ &= \sigma^2\boldsymbol{f}_k^H\boldsymbol{C}^{\boldsymbol{-1}}\boldsymbol{H}_D^{-1}\boldsymbol{H}_{\boldsymbol{D}}^{-\boldsymbol{H}}\boldsymbol{C}^{-\boldsymbol{H}}\boldsymbol{f}_k + O(\sigma^4) \\ &= \sigma^2\boldsymbol{\alpha}^{\boldsymbol{H}}\boldsymbol{\alpha} + O(\sigma^4) \qquad (17) \end{aligned}$$

, where $\boldsymbol{\alpha}$ equals $\boldsymbol{H}_{\boldsymbol{D}}^{-\boldsymbol{H}}\boldsymbol{C}^{-\boldsymbol{H}}\boldsymbol{f}_k$.

By applying matrix norm inequalities with the assumption of normalized channel power, we can obtain the upper bound of the first dominant term in eq(17) as below

$$\begin{aligned} \boldsymbol{\alpha}^{\boldsymbol{H}}\boldsymbol{\alpha} &= \|\boldsymbol{\alpha}\|^2 \\ &\leq \|\boldsymbol{H}_D^{-H}\boldsymbol{C}^{-H}\|_F^2\|\boldsymbol{f}_k\|_2^2 \leq \|\boldsymbol{C}^{-H}\|_F^2. \qquad (18) \end{aligned}$$

When the absolute value of each element in a matrix $\boldsymbol{C}^{-H}$ has the same magnitude, $\|\boldsymbol{C}^{-H}\|_F$ is mnimized, where $\|\cdot\|_2$ and $\|\cdot\|_F$ are 2-norm and Frobenious-norm respectively. Moreover, when the channel nulls are located on some subcarriers in frequency domain, the matrix $\boldsymbol{C}$ improves symbol detectability due to frequency diversity.

6. SIMULATION RESULTS

Simulations of the QPSK OFDM system are carried out with 16 subcarriers and $\rho = 0.08$ for the proposed precoding and blind channel estimation. The channel model is a 3-tap FIR filter coefficients selected from complex white Gaussian process and the length of the cyclic prefix is 3. Correlation matrix for channel estimation and LCMV equalizing weights are computed based on 150 blocks. The performance of channel estimation is evaluated according to NMSE,

$$NMSE = \frac{1}{J}\sum_{i=1}^{J}(\sum_{m=0}^{N-1}|H_{Dm}^i - \hat{H}_{Dm}^i|^2)/(\sum_{m=0}^{N-1}|H_{Dm}^i|^2) \quad (19)$$

where J is the number of independent channel realization, and H_{Dm}^i, $\hat{H}_{Dm}^i$ are the i-th realization of the channel and its estimation, respectively. The detection is performed after DFT of $\tilde{\boldsymbol{x}}_{\boldsymbol{i}}$, $\boldsymbol{F}_{\boldsymbol{N}}\tilde{\boldsymbol{x}}_{\boldsymbol{i}}$. The phase ambiguity of the received signal can be compensated via the use of a pilot symbol.

The results are shown in Fig.1 and Fig.2. In Fig.1, The performance of the proposed blind channel estimation method is compared with other schemes of [6] and [13] where the former utilizes autocorrelation for the entire symbol block and extracts $\boldsymbol{R}_{\boldsymbol{o}}$ only and the later uses circular precoding for blind channel estimation with SVD. The proposed method reveals more faster convergence than methods of [6] and [13]. Fig.2 shows bit-error-rate (BER) using the estimated channel for receiver [13] with MMSE equalization and proposed method. Though the proposed scheme reveals similar performance with the method of [13] at low SNR region, the performance at high SNR is improved drastically.

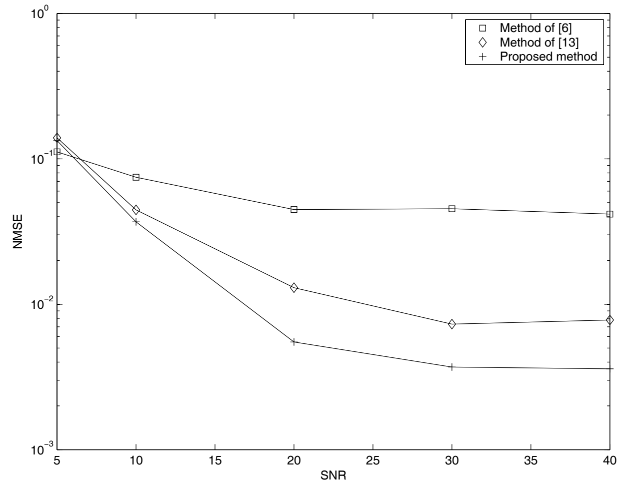

Fig. 1. NMSE performance for methods of [6], [13],and the proposed method

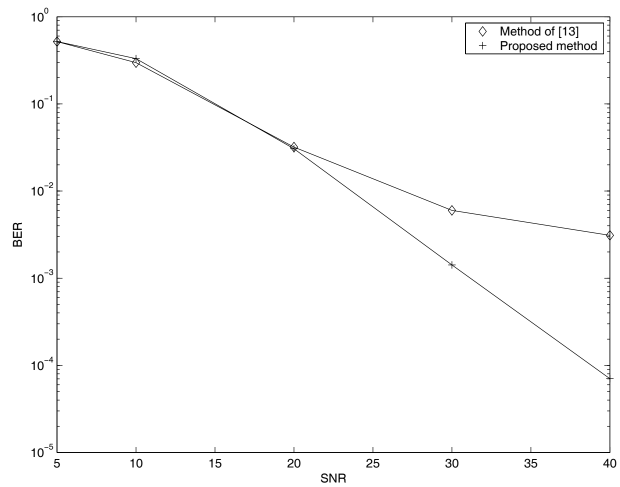

Fig. 2. BER performance for method of [13] and the proposed method

7. REFERENCES

[1] Y. Li, L. J. Cimini Jr., and N. R. Sollengerger, "Robust channel estimation for OFDM systems with rapid dispersive fading channels," *IEEE Trans. Commun.*, vol. 46, pp. 902.915, Jul. 1998.

[2] M. Morelli and U. Mengali, "A comparison of pilot-aided channel estimation methods for OFDM systems," *IEEE Trans. Signal Process.*, vol.49, pp. 3065.3073, Dec. 2001.

[3] G. B. Giannakis, "Filterbanks for blind channel identification and equalization," IEEE Signal Process. Lett., vol. 4, pp. 184.187, Jun. 1997.

[4] R. W. Heath and G. B. Giannakis, "Exploiting input cyclostationarity for blind channel identification in OFDM systems," *IEEE Trans. Signal Process.*, vol. 47, pp. 848.856, Mar. 1999.

[5] B. Muquet, M. de Courville, and P. Duhamel, "Subspace-based blind and semi-blind channel estimation for OFDM systems," *IEEE Trans. Signal Process.*, vol. 50, pp. 1699.1712, Jul. 2002.

[6] B. Muquet and M. de Courville, "Blind and semi-blind channel identification methods using second order statistics for OFDM systems" *IEEE ICASSP Proc.*, Mar. 1999.

[7] Z. Wang and G. B. Giannakis, "Linearly Precoded or coded OFDM against wireless fades?," *SPAWC'01 Proc.*, Taoyuan, Taiwan, March 2001.

[8] Z. Liu, Y. Xin, and G. B. Giannakis, "Linear constellation precoding for OFDM with maximum multipath diversity and coding gains," *IEEE Trans. Commun.*, vol. 51, pp. 416.427, Mar. 2003.

[9] R. Zhang, "Blind OFDM Channel Estimation through Linear Precoding: A Subspace Approach," *Asilomar's 02 Proc.*, Pacific Grove, CA, Nov. 2002.

[10] A.Petropulu, R. Zhang, and R. Lin, "Blind OFDM Channel Estimation through Simple Linear Precoding," *IEEE Trans. Wireless Commun.*, vol.3, no.2, pp.647-655, March, 2004.

[11] Rui Lin and Athina P. Petropulu, "Linear Precoding Assisted Blind Channel Estimation for OFDM System," *IEEE Trans. Vehicular Technology.*, vol.54, no.3, pp.983-995, May, 2005.

[12] Y. Lin and S. Phoong, "BER minimized OFDM systems with channel independent precoders," *IEEE Trans. Signal Process.*, vol. 51, pp. 2369.2380, Sept. 2003.

[13] R. Zhang, "Blind Channel Estimation for Precoded OFDM System," *IEEE ICASSP Proc.*, Mar. 2005.

[14] Taewoo Han and Xiaohua Li, "Minimum-output-energy Method for Blind Equalization of OFDM and Systems with Sufficietn or Insufficietnt Cyclic Prefix," *IEEE ICASSP Proc.*, Mar. 2003.