ITERATIVE RECEIVER DESIGN FOR SINGLE CARRIER CYCLIC PREFIX WITH CYCLIC DELAY DIVERSITY

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ABSTRACT

In this paper we consider a Single Carrier Cyclic Prefix (SCCP) modulation with Cyclic Delay Diversity (CCD). SCCP is an alternative to the multi-carrier transmission schemes such as Orthogonal Frequency Division Multiplexing (OFDM) and CCD is used to increase the frequency diversity. We propose an iterative receiver which not only makes use of pilot symbols but also data symbols to increase the performance of the system.

The structure of this paper is as follows: the methods presented in section I, section II provides the system model. The iterative receiver design is presented in Section III. The details of the Expectation Maximization (EM)-based channel estimation as well as the Minimum Mean Square Error (MMSE) with Soft Interference Cancelation (SIC) algorithm are given in this Section. Simulation results are given in Section IV. Finally, Section V concludes the paper

TÓM TẮT

Trong bài báo này, chúng tôi xem xét điều chế đơn sóng mang có tiền tố vòng (SCCP) sử dụng phân tập trễ vòng (CCD). SCCP là một thay thế cho truyền dẫn đa sóng mang trực giao (OFDM) và CCD được sử dụng để làm tăng độ phân tập tần số. Chúng tôi đề xuất một máy thu lặp (iterative), để không chỉ sử dụng các ký hiệu pilot mà cả cá kí hiệu data, làm tăng hiệu suất của hệ thống.Cấu trúc của bài báo như sau:Trong phần I giới thiệu về phương pháp. Phần II cung cấp mô hình hệ thống. Thiết kế máy thu lặp (iterative) được trình bày trong phần III. Các chi tiết của phép ước lượng kênh dựa trên thuật toán EM cũng như thuật tóan MMSE-SIC được trình bầy trong phần này. Trong phần IV trình bầy các kết quả mô phỏng. Phần V trình bầy các kết kuận của bài báo.

I. INTRODUCTION

Single Carrier Cyclic Prefix (SCCP) modulation [1] is another choice for multi-carrier transmission schemes in which Orthogonal Frequency Division Multiplexing (OFDM) is the most popular example. SCCP modulation uses a single carrier; hence, it enjoys a low peak to average power ratio (PAPR) as compared to multi-carrier counterparts. Both systems can be categorized in a broader transmission scheme called Cyclic-Prefix (CP) block-based transmission. The insertion of CP into signal blocks is a very effective way to combat intersymbol interference (ISI) arisen in frequency selective environment.

However, in frequency selective fading channels where channel lengths are small, the frequency diversity is poor. To overcome this problem, Cyclic Delay Diversity (CCD) [2] is first introduced to OFDM systems. CCD basically deploys more transmit antennas to transform spatial diversity to frequency diversity. In doing so, CCD does not alter the receiver structure as compared with systems using one transmit antenna. For SCCP systems, we also can deploy CCD to improve the frequency diversity.

To reconstruct the transmitted signals, the receiver needs to know the channel information. In principle, there are two approaches to obtain this information: the first one is to use just the pilot symbols transmitted during the training period and the other is not only uses the pilot symbols but also the detected data symbols to enhance the quality of estimated information. In this paper we adopt the second approach. On the other hand, Expectation Maximization (EM) algorithm [3] provides us an effective iterative algorithm to solve likelihood-based parameter estimation problems when the direct maximization of the likelihood function is extremely complicated or an absence of a nuisance parameter makes the problem hard to solve.

In this paper, we apply the EM algorithm to design the channel estimation algorithm in an iterative receiver for SCCP modulation with CCD. The proposed channel estimation requires the soft information of transmitted symbols. Due to large size of signal blocks, Minimum Mean Square Error (MMSE) with Soft Interference Cancelation (SIC) [4] algorithm is used to provide the soft information. The MMSE-SIC algorithm in [4] is proposed for coded systems; however, it is easy to apply in uncoded systems.

The structure of this paper is as follows: Section II provides the system model. The iterative receiver design is presented in Section III. The details of the EM-based channel estimation as well as the MMSE-SIC algorithm are given in this Section. Simulation results are given in Section IV. Finally, Section V concludes the paper.

Notations: italic/bold small letters denote vector; meanwhile, italic/bold big letters denote matrix. Transpose, conjugate and, Hermitian conjugate of a vector or a matrix are denoted by $(\bullet)^T$, $(\bullet)^*$ and $(\bullet)^H$, respectively. $\mathrm{diag}\{\boldsymbol{a}\}$ is a diagonal matrix whose diagonal elements belong to $\boldsymbol{a}$. $\boldsymbol{W}$ is the *N*-point discrete Fourier transform matrix. $\boldsymbol{I}_N$ is the identity matrix of size *N*.

II. SYSTEM MODEL

We consider a wireless system using SCCP for transmission. The transmitter has M antennas and the receiver has 1 antenna. The transmitter is depicted in Fig. 1. The channel between the *m*th transmit antenna and the receiver antenna is modeled as a frequency selective fading channel which is characterized by a channel impulse response of $\boldsymbol{g}_m = \begin{bmatrix} g_m(0) & g_m(1) & \cdots & g_m(L-1)\end{bmatrix}^T$ where L is the channel length. We assume that channels $\boldsymbol{g}_m$'s are static over B signal blocks. The *n*th signal block $\boldsymbol{x}(n) = \begin{bmatrix} x(n;0) & x(n;1) & \cdots & x(n;N-1)\end{bmatrix}^T$ of size N is first normalized by a factor of $1/\sqrt{M}$. The normalization is to guarantee that the average transmitted power is independent from the number of antennas in used. The element $x(n;k)$ belongs to a constellation $C = \{c_0, c_1, \cdots, c_{|C|}\}$ of size $|C|$ with average power E_s.

The *n*th signal block after normalization is then cyclically shifted by a cyclic delay δ_m before being transmitted on the *m*th transmit antenna. Here, without loss of generality, δ_0 is set to 0. The resultant signal block after cyclically shifted, $\boldsymbol{x}_m(n) = \begin{bmatrix} x_m(n;0) & \cdots & x_m(n;k) & \cdots & x_m(n;N-1)\end{bmatrix}^T$, is determined as follows

$$x_0(n;k) = \frac{1}{\sqrt{M}} x(n;k), \quad (1)$$

and

$$x_m(n;k) = \frac{1}{\sqrt{M}} x\left(n;(k-\delta_m)_N\right), \; m = 0,1,\cdots M-1 \quad (2)$$

where $n = 0,1,\cdots,B-1$ and $k = 0,1,\cdots,N-1$. Each block belonging to the set $\{\boldsymbol{x}_m(n)\}_{m=0}^{M-1}$ is then CP-added with CP length P. The value of P must satisfy $P \ge L-1$ to avoid interblock interference (IBI).

The received signal vector at the receiver is the superposition of signals from all M transmit antennas. It can be expressed as

$$\boldsymbol{z}(n) = \sum_{m=0}^{M-1} \boldsymbol{G}_m \boldsymbol{x}_m(n) + \boldsymbol{v}(n), \; n = 0,1,\cdots,B-1, \quad (3)$$

where $\boldsymbol{G}_m$ is the circulant matrix having the first column of $\begin{bmatrix} \boldsymbol{g}_m^T & \boldsymbol{0}_{N-L}^T \end{bmatrix}^T$ and $\boldsymbol{v}(n)$ is a realization of a complex Gaussian vector with zero mean and covariance matrix $E\{\boldsymbol{v}(n)\boldsymbol{v}^H(n)\} = N_0 \boldsymbol{I}_N$. Note that matrix $\boldsymbol{G}_m$ can be decomposed as [5]

$$\boldsymbol{G}_m = \boldsymbol{W}^H \boldsymbol{\Lambda}_m \boldsymbol{W}, \; m = 0,1,\cdots,M-1, \quad (4)$$

where $\boldsymbol{\Lambda}_m = \mathrm{diag}\{G_m(0) \;\; G_m(1) \;\; \cdots \;\; G_m(N-1)\}$ and

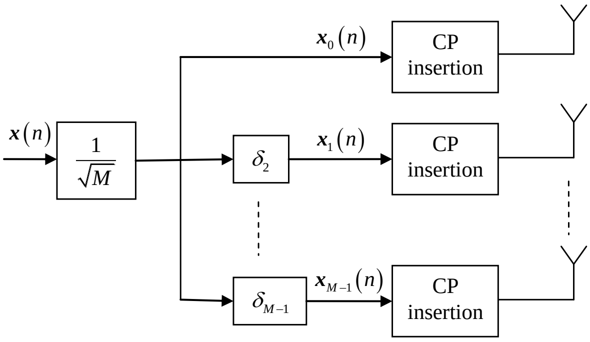

Fig. 1 System model of SCCP modulation with CDD.

$$G_m(k)=\sum_{l=0}^{L-1} g_m(l)e^{-j\frac{2\pi}{N}kl},\ k=0,1,\cdots,N-1. \quad (5)$$

If vector $z(n)$ of (3) is FFT-transformed, we obtain

$$y(n)=Wz(n)=\sum_{m=0}^{M-1}\Lambda_m W x_m(n)+w(n), \quad (6)$$

for $n=0,1,\cdots,B-1$, and $w(n)=Wv(n)$. Note that $w(n)$ is still a realization of a complex Gaussian vector with zero mean and covariance matrix $E\{w(n)w^H(n)\}=N_0 I_N$.

Based on (1), (2) and [5], we have

$$Wx_m(n)=\frac{1}{\sqrt{M}}\Omega_m W x(n),\ n=0,1,\cdots,B-1, \quad (7)$$

where

$$\Omega_m=\mathrm{diag}\left\{1 \quad e^{-j\frac{2\pi}{N}\delta_m} \quad \cdots \quad e^{-j\frac{2\pi}{N}(N-1)\delta_m}\right\}. \quad (8)$$

Therefore, (6) can be written as

$$y(n)=\frac{1}{\sqrt{M}}\sum_{m=0}^{M-1}\Lambda_m\Omega_m W x(n)+w(n), \quad (9)$$

for $n=0,1,\cdots,B-1$. In other words, by deploying CDD, M individual channels from M transmit antennas to the receive antenna become an equivalent channel (in frequency domain) as

$$\Lambda=\frac{1}{\sqrt{M}}\sum_{m=0}^{M-1}\Lambda_m\Omega_m . \quad (10)$$

In time domain, we have the equivalent channel impulse response

$$g=\frac{1}{\sqrt{M}}\sum_{m=0}^{M-1}g_{m,\delta_m}, \quad (11)$$

where g_{m,δ_m} is the cyclically shifted version of g_m with cyclic delay δ_m. It is easy to see that the length g is $(\delta_{M-1}+L)$. Hence, (9) is written as

$$\begin{aligned} y(n)&=\Lambda W x(n)+w(n)\\ &=Hx(n)+w(n),\ n=0,1,\cdots,B-1 \end{aligned}. \quad (12)$$

where $H=\Lambda W$. Eq. (12) can also be written in another form without los of information as below

$$y(n)=X(n)Fg+w(n),\ n=0,1,\cdots,B-1, \quad (13)$$

where F is the matrix consisting of the first $(\delta_{M-1}+L)$ columns of $\sqrt{N}W$ and

$$X(n)=\mathrm{diag}\{Wx(n)\}. \quad (14)$$

Our problem is to estimate g and decode the set of transmitted signal blocks $\{x(n)\}_{n=0}^{B-1}$ based on the set of received signal vectors $\{y(n)\}_{n=0}^{B-1}$. The proposed iterative receiver will be presented in the next Section.

III. PROPOSE ITERATIVE RECEIVER

The proposed iterative receiver consists of two big blocks: channel estimation block and signal detection block. The channel estimation algorithm is based on an application of the EM algorithm. This algorithm requires the soft information of transmitted signals. The information will be provided by the MMSE-SIC algorithm. Details of two blocks are given as follows.

3.1 MMSE-SIC algorithm

MMSE-SIC algorithm will provide us the soft information of individual elements of $x(n)$. Suppose that after i_{EM} iteration of EM-based channel estimation algorithm we have the estimate value $\hat{g}^{[i_{\mathrm{EM}}]}$. This value will be used in the MMSE-SIC algorithm to find an estimate of $\hat{H}^{[i_{\mathrm{EM}}]}$ (based on (12)). After that, model in (12) is used as in the following algorithm

• At the zero iteration of MMSE-SIC, a conventional MMSE equalizer is applied for each element of $\boldsymbol{x}(n)$, i.e., for $x(n;k)$, we need to determine

$$\hat{\boldsymbol{a}}^{[i_{\text{EM}},0]}(n;k)=\left(E_s\hat{\boldsymbol{H}}^{[i_{\text{EM}}]}\left(\hat{\boldsymbol{H}}^{[i_{\text{EM}}]}\right)^H+N_0\boldsymbol{I}_N\right)^{-1}E_s\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}, \tag{15}$$

where $\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}$ is the kth column of $\hat{\boldsymbol{H}}^{[i_{\text{EM}}]}$. The output of MMSE equalizer after bias removal is given by

$$\begin{aligned}\bar{x}^{[i_{\text{EM}},0]}(n;k)&=\frac{\left(\boldsymbol{a}^{[i_{\text{EM}},0]}(n;k)\right)^H\boldsymbol{y}(n)}{\hat{\phi}^{[i_{\text{EM}},0]}(n;k)}.\\&=x(n;k)+\hat{e}^{[i_{\text{EM}},0]}(n;k)\end{aligned} \tag{16}$$

where $\hat{\phi}^{[i_{\text{EM}},0]}(n;k)=\left(\boldsymbol{a}^{[i_{\text{EM}},0]}(n;k)\right)^H\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}$ and $\hat{e}^{[i_{\text{EM}},0]}(n;k)$ comprises of the residual interference plus noise. The variance of $e^{[i_{\text{EM}},0]}(n;k)$ is

$$\begin{aligned}\hat{\rho}^{[i_{\text{EM}},0]}(n;k)&=\frac{1}{\left|\hat{\phi}^{[i_{\text{EM}},0]}(n;k)\right|^2}\left(\hat{\boldsymbol{a}}^{[i_{\text{EM}},0]}(n;k)\right)^H\\&\left(\sum_{l=0,\,l\neq k}^{N-1}E_s\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\left(\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\right)^H+N_0\boldsymbol{I}_N\right)^{-1}\hat{\boldsymbol{a}}^{[i_{\text{EM}},0]}(n;k)\end{aligned}. \tag{17}$$

Based on (16), the soft information of $x(n;k)$ is determined by

$$\begin{aligned}\tilde{x}^{[i_{\text{EM}},0]}(n;k)&=E\left\{x(n;k)\middle|\bar{x}^{[i_{\text{EM}},0]}(n;k)\right\}\\&=\frac{\sum_{i=0}^{|C|-1}c_i\,f\left(\bar{x}^{[i_{\text{EM}},0]}(n;k)\middle|x(n;k)=c_i\right)}{\sum_{i=0}^{|C|-1}f\left(\bar{x}^{[i_{\text{EM}},0]}(n;k)\middle|x(n;k)=c_i\right)}\end{aligned} \tag{18}$$

where $f\left(\bar{x}^{[i_{\text{EM}},0]}(n;k)\middle|x(n;k)=c_i\right)$ is the conditional probability density function (p.d.f.) of $\bar{x}^{[i_{\text{EM}},0]}(n;k)$ given that $x(n;k)$ is the constellation point c_i. If we assume that $\hat{e}^{[i_{\text{EM}},0]}(n;k)$ is a Gaussian random variable, the conditional p.d.f. is easily calculated.

• We assume that we have obtained the soft information of $x(n;k)$ at $(i_{\text{MMSE}}-1)$ iteration, $\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;k)$. During the i_{MMSE} iteration, the following soft interference cancelation is formed for $x(n;k)$

$$\begin{aligned}&\hat{\boldsymbol{y}}^{[i_{\text{EM}},i_{\text{MMSE}}]}(n;k)\\&=\boldsymbol{y}(n)-\sum_{l=0,\,l\neq k}^{N-1}\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)=\\&\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}x(n;k)+\sum_{l=0,\,l\neq k}^{N-1}\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\left(x(n;l)-\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)\right)\\&+\boldsymbol{w}(n)\end{aligned}. \tag{19}$$

Based on (19), a new MMSE equalizer for $x(n;k)$ is derived as

$$\begin{aligned}\hat{\boldsymbol{a}}^{[i_{\text{EM}},i_{\text{MMSE}}]}(n;k)&=\left(E_s\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}\left(\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}\right)^H+\right.\\&\left.\sum_{l=0,\,l\neq k}^{N-1}\breve{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\left(\hat{\boldsymbol{h}}_l^{[i_{\text{EM}}]}\right)^H+N_0\boldsymbol{I}_N\right)^{-1}\\&\times E_s\hat{\boldsymbol{h}}_k^{[i_{\text{EM}}]}\end{aligned} \tag{20}$$

where

$$\breve{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)=E\left\{\left|x(n;l)-\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)\right|^2\middle|\bar{x}^{[i_{\text{EM}},i_{\text{MMSE}}-1]}(n;l)\right\}.$$

After obtaining the new MMSE equalizer, a similar steps can be derived to find $\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}]}(n;k)$. Therefore, when i_{MMSE} iterations of MMSE-SIC is completed, we obtain the soft information of vector $\boldsymbol{x}(n)$ as $\tilde{\boldsymbol{x}}^{[i_{\text{EM}},i_{\text{MMSE}}]}=\left[\tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}]}(n;0)\ \cdots\ \tilde{x}^{[i_{\text{EM}},i_{\text{MMSE}}]}(n;N-1)\right]^T$. This soft information vector is transferred to the EM-based channel estimation algorithm to find a new updated value of channel vector $\boldsymbol{g}$.

3.2 EM-based channel estimation algorithm

The set $\{\boldsymbol{y}(n)\}_{n=0}^{B-1}$ is the *incomplete data space* for the parameter we want to estimate, $\boldsymbol{g}$. The signal set $\{\boldsymbol{x}(n)\}_{n=0}^{B-1}$ is considered as the *missing data space*. We define $\left\{\{\boldsymbol{y}(n)\}_{n=0}^{B-1},\{\boldsymbol{x}(n)\}_{n=0}^{B-1}\right\}$ as the *complete data space* for $\boldsymbol{g}$. The EM-based channel estimation algorithm consists of the following two steps.

1. *E-step:* In this step, we determine

$$Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right)=E\left\{f\left(\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\left\{\boldsymbol{x}(n)\right\}_{n=0}^{B-1}|\boldsymbol{g}\right)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}. \quad (21)$$

Due to the independence between $\left\{\boldsymbol{x}(n)\right\}_{n=0}^{B-1}$ and $\boldsymbol{g}$, we have

$$Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right)=E\left\{f\left(\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1}\left|\left\{\boldsymbol{x}(n)\right\}_{n=0}^{B-1},\boldsymbol{g}\right.\right)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}. \quad (22)$$

where

$$f\left(\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1}\left|\left\{\boldsymbol{x}(n)\right\}_{n=0}^{B-1},\boldsymbol{g}\right.\right)=\frac{1}{(\pi N_0)^N}\exp\left\{-\frac{1}{N_0}\sum_{n=0}^{B-1}\left\|\boldsymbol{y}(n)-\boldsymbol{X}(n)\boldsymbol{F}\boldsymbol{g}\right\|^2\right\}. \quad (23)$$

Substituting (23) into (22) and dropping some terms that do not relate to $\boldsymbol{g}$, we obtain

$$\begin{aligned}Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right)=&\ \boldsymbol{g}^H\boldsymbol{F}^H\left(\sum_{n=0}^{B-1}E\left\{\boldsymbol{X}^H(n)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}\boldsymbol{y}(n)\right)\\&+\left(\sum_{n=0}^{B-1}\boldsymbol{y}^H(n)E\left\{\boldsymbol{X}(n)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}\right)\boldsymbol{F}\boldsymbol{g}\\&-\boldsymbol{g}^H\boldsymbol{F}^H E\left\{\sum_{n=0}^{B-1}\boldsymbol{X}^H(n)\boldsymbol{X}(n)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}\boldsymbol{F}\boldsymbol{g}.\end{aligned} \quad (24)$$

Eq. (24) can be determined by the following quantities:

- We define $\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}=\sum_{n=0}^{B-1}E\left\{\boldsymbol{X}^H(n)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}\boldsymbol{y}(n)$.

Based on (14), we have

$$\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}=\sum_{n=0}^{B-1}\left(\mathrm{diag}\left\{\boldsymbol{W}\tilde{\boldsymbol{x}}^{[i_{\mathrm{EM}},i_{\mathrm{MMSE}}]}(n)\right\}\right)^H\boldsymbol{y}(n). \quad (25)$$

- Let $\hat{\boldsymbol{B}}^{[i_{\mathrm{EM}}]}=E\left\{\sum_{n=0}^{B-1}\boldsymbol{X}^H(n)\boldsymbol{X}(n)\left|\left\{\boldsymbol{y}(n)\right\}_{n=0}^{B-1},\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right\}$. We have $\hat{\boldsymbol{B}}^{[i_{\mathrm{EM}}]}=\left|\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}\right|^2$.

In other words, (24) is rewritten as

$$Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right)=\boldsymbol{g}^H\boldsymbol{F}^H\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}+\left(\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}\right)^H\boldsymbol{F}\boldsymbol{g}-\boldsymbol{g}^H\boldsymbol{F}^H\hat{\boldsymbol{B}}^{[i_{\mathrm{EM}}]}\boldsymbol{F}\boldsymbol{g}. \quad (26)$$

2. *M-step:* In this step, the new value of $\boldsymbol{g}$, $\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}+1]}$, is calculated by

$$\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}+1]}=\arg\max_{\boldsymbol{g}}Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{EM}]}\right.\right). \quad (27)$$

Differentiating (26) with respect to $\boldsymbol{g}$ [6], we have

$$\frac{\partial Q\left(\boldsymbol{g}\left|\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}]}\right.\right)}{\partial\boldsymbol{g}}=\left(\boldsymbol{F}^H\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}\right)^*-\left(\boldsymbol{F}^H\hat{\boldsymbol{B}}^{[i_{\mathrm{EM}}]}\boldsymbol{F}\boldsymbol{g}\right)^*. \quad (28)$$

Equating (28) to $\boldsymbol{0}$, we obtain

$$\hat{\boldsymbol{g}}^{[i_{\mathrm{EM}}+1]}=\left(\boldsymbol{F}^H\hat{\boldsymbol{B}}^{[i_{\mathrm{EM}}]}\boldsymbol{F}\right)^{-1}\boldsymbol{F}^H\hat{\boldsymbol{A}}^{[i_{\mathrm{EM}}]}. \quad (29)$$

IV. SIMULATION RESULTS

In this Section we present some simulation results of a system with the following parameters: signal block size $N=64$, the number of antennas (in case CDD is deployed) $M=4$, the cyclic delay $\delta_m=8m$ for $m=0,1,\cdots,M-1$. The channel between the *m*th transmit antenna to the receive antenna is a frequency selective fading channel with channel impulse response of length $L=4$. Each element of the impulse response is modeled as a complex Gaussian random variable with zero mean and variance $1/(LM)$, i.e., uniform power delay profile is used. This assumption guarantees that for any number of transmit antennas in used, the signal-to-noise ratios at the receiver are the same. We assume that M channels are static over $B=10$ signal blocks. QPSK modulation is adopted with $E_s=2$. SNR is defined as $\mathrm{SNR}=E_s/N_0$. Of B signal blocks, the first block is used to carry pilot symbols. Those pilot symbols help us to determine the initial value of $\boldsymbol{g}$, $\hat{\boldsymbol{g}}^{[0]}$, to start our iterative receiver. We first look at the effect of CDD and then performance of our proposed receiver is presented.

4.1 Effect of CDD

Fig. 2 and Fig. 3 provide the bit-error-rate (BER) of the system without CDD $(M=1)$ and with CDD $(M=4)$. In this case, we assume that the receiver has the perfect knowledge of $\boldsymbol{g}$. The number of iteration of MMSE-SIC is 5. Due to high signal block size, we cannot provide the Maximum Likelihood (ML) detection performance bound; instead, we present the Single User – Matched Filter Bound (SU-MFB) [7] as a benchmark. We observe from the two figures that using CDD improves the BER performance a lot. With CDD, the gap between SU-MFB and MMSE-SIC becomes small.

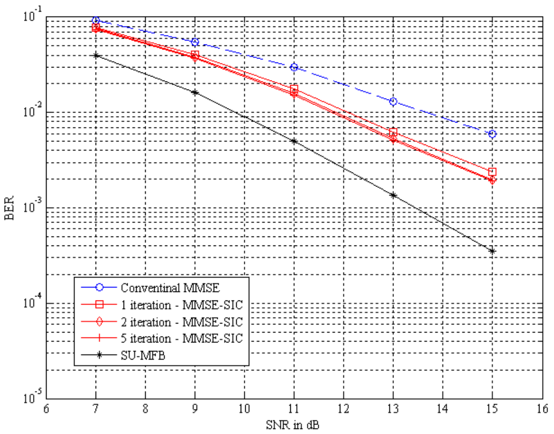

Fig. 2 BER performance without CDD.

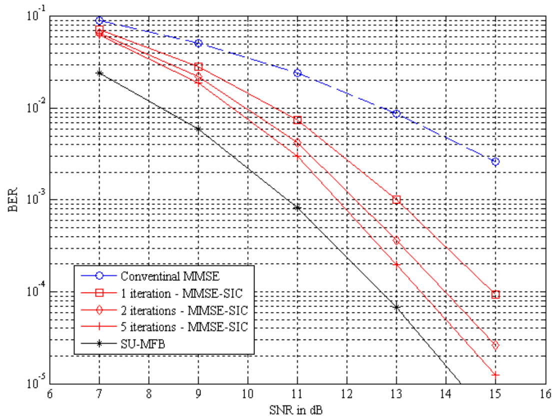

Fig. 3 BER performance with CDD.

4.2 Performance of proposed iterative receiver

BER performance of the proposed receiver with 5 iterations of MMSE-SIC and 3 iterations of the EM-based algorithm is given in Fig. 4. Least-square principle [6] is applied on the received signal vector associated with the first signal block (pilot block) to obtain $\hat{\boldsymbol{g}}^{[0]}$. We see that after 1 iteration of EM-based channel estimation, the BER improves greatly. The curve with legend "Initial" is the performance of MMSE-SIC with $\hat{\boldsymbol{g}}^{[0]}$. Between 2 and 3 iterations of EM-based algorithm, the improvement is marginal. Note that after 3 iterations of EM-based algorithm, our receiver approaches the performance of MMSE-SIC with perfect channel information at high SNR region.

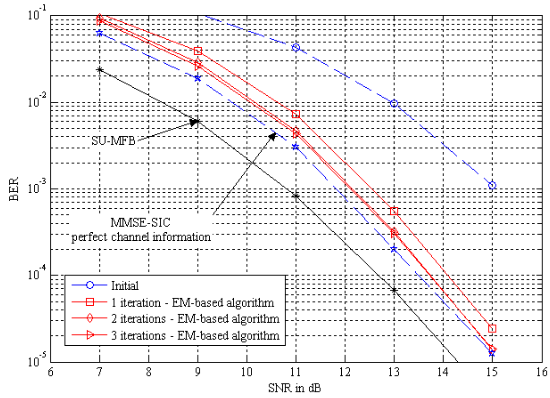

Fig. 4 BER performance of proposed iterative receiver.

Fig. 5 provides the Mean Square Error (MSE) performance of $\boldsymbol{g}$. We observe that by using EM-based algorithm, the MSE dramatically reduces as compared with only using pilot block.

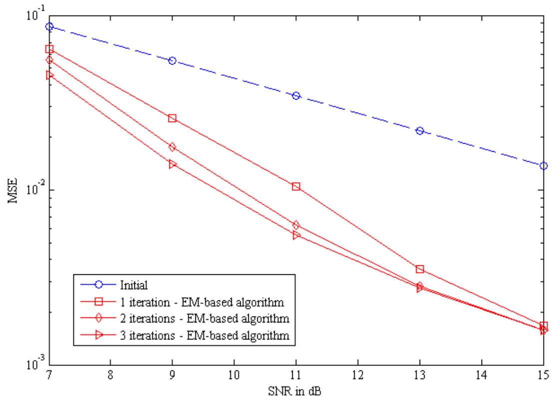

Fig. 5 MSE performance of proposed iterative receiver.

V. CONCLUSIONS

In this paper we proposed an iterative receiver for a wireless system using SCCP with CDD. The receiver consists of two blocks: MMSE-SIC algorithm to provide soft information of transmitted signals which is used at the second block, EM-based channel estimation algorithm. Simulation results show that our receiver has the performance very close to that of perfect channel information.

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