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Performance Evaluation of the Joint Detection with Channel Estimation Error

Tongliang Fan, Shengbo Huang, Yandong Sun, Xue Zhao and Dingyi Ji

Electronic Technology Department, China Maritime Police Academy, Ningbo, 315801, China
libufan432@163.com

Abstract

In previous B. Steiner channel estimation algorithm adopted in Time Division-Synchronous Code Division Multiple Access (TD-SCDMA) system, but there are some errors in estimated channel due to noise. However, due to using the short training sequences available, the resulting estimation of the channel and detected data suffer from unavoidable errors. In this paper, we analyze the effect of channel estimation error on joint detection and evaluate the performance of TD-SCDMA systems with channel estimation error. The simulation results are consistent with the theory analysis.

Keywords: *TD-SCDMA, Joint Detection, channel estimation error*

1. Introduction

TD-SCDMA is a time division system that uses an unpaired bandwidth structure; the same bandwidth allocation is used for both downlink and uplink in a time synchronized manner. Joint Detection as key technology depends on the fast and precise estimation of wireless channel impulse responses (CIRs).

B. Steiner [1-2] presented a low-cost channel estimation algorithm based on least square. By taking advantage of the Toeplitz characteristic of the channel estimation matrix, the inverse of matrix can be computed by utilizing fast Fourier transform (FFT) and inverse FFT (IFFT), which increases the operation speed of channel estimation algorithm. It has widely been adopted in current TD-SCDMA systems for its moderate estimation accuracy and low computation complexity. However, it simply treats inter-cell interference as noise, and the results of channel estimation will be greatly affected by the noises. The performance of channel estimation is decreased. Therefore, some joint channel estimation (JCE) algorithms such as minimum mean-square-error (MMSE) are presented and simulated in [5], and JCE can provide more accurate channel impulse responses compared with the conventional Steiner channel estimator. It can eliminate both MAI and ISI and provide the best tradeoff between bias and variance of the estimation results, which suffer from high complexity. By adopting Singular Value Decomposition (SVD), Ali K [6] presents a novel channel estimation method with low complexity. In [7] proposes an iterative parallel interference cancellation channel estimation algorithm. Because the interference has to rebuild for cancelling it, the complexity of this method is also high. Using the channel and active code information from adjacent cells, inter-cell interference cancellation (ICI) [8] has been proposed to combat inter-cell interference.

Joint detection (JD) known as a multi-user detection technique can combat both ISI and MAI. Based on different criterions, typical JD equalizers have three linear joint methods: Whitening Matched Filtering (WMF), ZF-BLE and MMSE-BLE. WMF is based on the

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principle of maximum signal to noise ratio [4]. ZF-BLE and MMSE-BLE are used to solve problems resulted from ISI and MAI [5] and their performance is compared in [6-7].

However, most of previous works did not further present the impact of the accuracy of channel estimation on the JD operation. In this paper, channel estimation error is analyzed. In order to evaluate the performance of TD-SCDMA systems with channel estimation error, we also simulate by computer and compare the results between ideal channel and estimated channel.

The rest of this paper is organized as follows. In Section 2 we briefly go over the TD-SCDMA system model AND B. STEINER algorithm. In Section3, we provide a review of ZF-BLE algorithm. Channel estimation error is analyzed in Section 4. Simulation results of two propagation channel scenarios are presented in Section 5. The final conclusions are given in Section 6.

2. System Model and B. Steiner Algorithm

2.1. Structure of Time Slot

The TD-SCDMA system as defined in the 3GPP standard combines aspects of TDMA with CDMA. It allows for the existence of multiple users in a given uplink or downlink timeslot. The data are transmitted in the form of burst, which is the lasting time of a timeslot. Figure 1 show that each regular time slot composes of two data blocks of 352chips length respectively, a midamble code of 144chips length and a GP with 16chips length. The midamble between the two data bearing blocks is designed to estimate and measure channels.

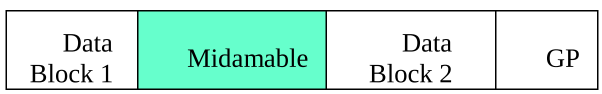

Figure 1. The Structure of a Burst or Timeslot

In TD-SCDMA system, there are 128 basic midambles which are divided into 32 groups. In each cell, the base station chooses one out of the four midamble codes in a given group and each active user in the cell is assigned with a different cyclically shifted version of this basic midamble [3].

The real valued basic midamble code vector m_{basic} is defined as:

$$m_{basic} = \left(m_1 \quad \cdots \quad m_p \quad \cdots \quad m_P \right) \tag{1}$$

where $m_p \in \{1, -1\}$ is the elements of the m_{basic} vector, $P = K \times W$. K is the maximum midamble code which can be used in timeslot, ranging from 2,4,6,8,10,12,14 to 16. W is the window length of the impulse response of wireless channel that depends on the selected number of active users . The different user's midamle sequence in the same cellular and the same timeslot is produced by one midamble code group m_{basic} after cyclic shift.

Modeling the wireless channel of the *k*th user to be the complex impulse response system with window length taken as W , then:

$$\mathbf{h}^{(k)} = (h_1^k, h_2^k, \cdots, h_W^k)^T \qquad k = 1, 2, \cdots, K \tag{2}$$

The midamble sequence of user *k* can be represented as:

$$\mathbf{m}^{(k)} = (m_1^k, m_2^k, \cdots, m_{L_m}^k)^T \qquad k = 1, 2, \cdots, K \tag{3}$$

The midamble code will be sent out along with data information to the receiving terminal via wireless channels. The amplitude and phase distortion happened during the transmitting process, so that the receiving midamble code is:

$$\mathbf{e}^{(k)} = \mathbf{m}^{(k)} * \mathbf{h}^{(k)} + \mathbf{n}^{(k)} \tag{4}$$

Where $\mathbf{n}^{(k)}$ is additive noises. The convolution operation expressed by (4) can be represented by matrix and vector to be:

$$e^{(k)} = \begin{bmatrix} m_1^k & 0 & \cdots & 0 \\ m_2^k & m_1^k & \cdots & \cdots \\ \vdots & \vdots & \vdots & 0 \\ m_W^k & m_{W-1}^k & \cdots & m_1^k \\ \vdots & \vdots & \vdots & \vdots \\ m_{L_m}^k & m_{L_m-1}^k & \cdots & m_{L_m-}^k \\ 0 & m_{L_m}^k & \cdots & m_{L_m-W+1}^k \\ \vdots & \vdots & \vdots & \vdots \\ 0 & 0 & \cdots & m_{L_m}^k \end{bmatrix} \bullet \begin{bmatrix} h_1^k \\ h_2^k \\ \vdots \\ h_W^k \end{bmatrix} + \begin{bmatrix} n_1^k \\ n_2^k \\ \vdots \\ n_W^k \end{bmatrix} \tag{5}$$

After the W is given, the situation that midamble sequence is located between two data fields which cause the interference which former W -1 bits midamble codes of receiver is affected by first data field and last W -1 bits midamble codes affect second data field, so, the only codes decided by midamble sequence is $W \sim W + P$ -1 . Equation (5) is reset to be:

$$\mathbf{e}^{(k)} = \begin{bmatrix} m_W^{(k)} & m_{W-1}^{(k)} & \cdots & m_1^{(k)} \\ m_{W+1}^{(k)} & m_W^{(k)} & \cdots & m_2^{(k)} \\ \cdots & \cdots & \ddots & \vdots \\ m_{L_m-1}^{(k)} & m_{L_m-2}^{(k)} & \cdots & m_{L_m-W}^{(k)} \end{bmatrix} \bullet \begin{bmatrix} h_1^k \\ h_2^k \\ \vdots \\ h_W^k \end{bmatrix} + \begin{bmatrix} n_1^k \\ n_2^k \\ \vdots \\ n_W^k \end{bmatrix} = \mathbf{G}^{(k)}\mathbf{h}^{(k)} + \mathbf{n}^{(k)} \tag{6}$$

In TD-SCDMA system, midamble sequences of all K users are transmitted simultaneously. Received signals are commonly determined by midamle sequences of all K users and additive noises, as the following equation:

$$\begin{aligned} \mathbf{e} &= \sum_{k=1}^{K} \mathbf{e}^{(k)} = \sum_{k=1}^{K} \left(\mathbf{G}^{(k)}\mathbf{h}^{(k)} + \mathbf{n}^{(k)} \right) \\ &= \left[\mathbf{G}^{(1)} \quad \mathbf{G}^{(2)} \quad \cdots \quad \mathbf{G}^{(K)} \right] \begin{bmatrix} \mathbf{h}^{(1)} \\ \mathbf{h}^{(2)} \\ \vdots \\ \mathbf{h}^{(K)} \end{bmatrix} + \mathbf{n} = \mathbf{G}\,\mathbf{h} + \mathbf{n} \end{aligned} \tag{7}$$

Where $\mathbf{G}$ is P order square cyclic matrix that collects $\mathbf{G}^{(k)}$ for all the K users in the same serving cell. At the same time, $\mathbf{h}$ is the $P \times 1$ channel impulse vector for the K users.

The CIR of vector $\mathbf{h}$ can be obtained from (7):

$$\hat{\mathbf{h}} = \mathbf{G}^{-1}\mathbf{e} = \mathbf{h} + \mathbf{G}^{-1}\mathbf{n} \tag{8}$$

The key to solve $\mathbf{h}$ is to find the converse of $\mathbf{G}$ inverse a circular Toeplitz matrix which contains the converse of nonsingular circular matrix. Therefore it only needs to solve the first column and the first row of the circular converse matrix. By using the property of a cyclic matrix, Discrete Fourier Transform (DFT) and Inverse Discrete Fourier Transform (IDFT) can be employed to simplify the calculation greatly, and therefore the channel of K users can be estimated as:

$$\hat{\mathbf{h}} = IFFT\left(\frac{FFT(\mathbf{e})}{FFT(\mathbf{m}_{basic})}\right) \tag{9}$$

Since different columns of matrix $\mathbf{G}$ generated by the same basic midamble in the serving cell are quasi-orthogonal vectors, the B. Steiner algorithm, applied as a low-computational method, can achieve satisfying performance with little amplification of noise power.

2.2. Signal Model

Assume there are K users in TD-SCDMA system, the transmitted data symbol of the *k*th user can be represented as:

$$\mathbf{d}^{(k)} = \begin{bmatrix} d_1^k & d_2^k & \cdots & d_N^k \end{bmatrix}^T \quad k = 1, \cdots, K \tag{10}$$

where N is the number of data symbols in one data block. Each data symbol is spread by a user specific $\mathbf{c}^{(k)}$ code of length Q . $\mathbf{c}^{(k)}$ is:

$$\mathbf{c}^{(k)} = \begin{bmatrix} c_1^k & c_2^k & \cdots & c_N^k \end{bmatrix}^T \quad k = 1, \cdots, K \tag{11}$$

Define the CIR of the signal received at the *k*th user as:

$$\mathbf{h}^{(k)} = \begin{bmatrix} h_1^k & h_2^k & \cdots & h_N^k \end{bmatrix}^T \quad k = 1, \cdots, K \tag{12}$$

The combined CIR of the *k*th user is introduced and defined as the convolution of its spreading code and CIR, expressed as:

$$\mathbf{b}^{(k)} = \mathbf{h}^{(k)} * \mathbf{c}^{(k)} = \begin{bmatrix} b_1^k & b_2^k & \cdots & b_{Q+W-1}^k \end{bmatrix}^T \tag{13}$$

where $*$ means the convolution operation. The received signal is the addition of users' data and noise:

$$\mathbf{r} = \sum_{k=1}^{K} \mathbf{h}^{(k)} * \mathbf{c}^{(k)} \otimes \mathbf{d}^{(k)} + \mathbf{n} \tag{14}$$

where $\otimes$ is the kronecker operator symbol. The combined CIR matrix $\mathbf{A}^{(k)}$ for the *k*th user can be constructed by arranging $\mathbf{b}^{(k)}$ in its columns, with each column corresponding to one symbol. Thus, the received signal is rewritten as:

$$\mathbf{r} = \sum_{k=1}^{K} \mathbf{A}^{(k)}\mathbf{d}^{(k)} + \mathbf{n} = \mathbf{A}\mathbf{d} + \mathbf{n} \tag{15}$$

3. Joint Detection Algorithms

If a zero forcing-block linear equalizer (ZF-BLE) [4] is used at the receiver of note B or UE, the transmitted data symbol vector is estimated as:

$$\mathbf{d} = (\mathbf{A}^H\mathbf{A})^{-1}\mathbf{A}^H\mathbf{r} \tag{16}$$

ZF-BLE algorithm is the optimal weighted least squares estimation based on Gauss-Markov theorem. The solution of ZF-BLE algorithm is obtained by the following equation:

$$\mathbf{q} = \arg\min \left\| \mathbf{r} - \mathbf{A}\hat{\mathbf{d}} \right\|^2 \tag{17}$$

It can obtain a continuous unbiased estimates.

The derivation of ZF-BLE algorithm is given in the following. The Lagrangian equation of problem (17) is presented by:

$$\frac{\partial \mathbf{q}}{\partial (\hat{\mathbf{d}})} = \mathbf{0} \tag{18}$$

According to the derivation rule of the matrix function:

$$\begin{aligned} \frac{\partial \mathbf{q}}{\partial (\hat{\mathbf{d}})} &= \frac{\partial [(\mathbf{r} - \mathbf{A}\hat{\mathbf{d}})^H (\mathbf{r} - \mathbf{A}\hat{\mathbf{d}})]}{\partial (\hat{\mathbf{d}})} \\ &= \frac{\partial [\mathbf{r}\mathbf{r}^H - \mathbf{r}^H \mathbf{A}\hat{\mathbf{d}} - \hat{\mathbf{d}}\mathbf{A}^H \mathbf{r} - \hat{\mathbf{d}}^H \mathbf{A}^H \mathbf{A}\hat{\mathbf{d}}]}{\partial (\hat{\mathbf{d}})} \\ &= \mathbf{0} - \mathbf{A}^H \mathbf{r} - \mathbf{A}^H \mathbf{r} - [\mathbf{A}^H \mathbf{A} + (\mathbf{A}^H \mathbf{A})^H]\hat{\mathbf{d}} \\ &= -2\mathbf{A}^H (\mathbf{r} - \mathbf{A}\hat{\mathbf{d}}) \end{aligned} \tag{19}$$

The superscript H is complex-conjugate (Hermitian) transpose. By substituting (19) into (18):

$$\frac{\partial \mathbf{q}}{\partial (\hat{\mathbf{d}})} = -2\mathbf{A}^H (\mathbf{r} - \mathbf{A}\hat{\mathbf{d}}) = \mathbf{0} \tag{20}$$

Therefore, the detection data of ZF-BLE is:

$$\hat{\mathbf{d}}_{ZF-BLE} = (\mathbf{A}^H \mathbf{A})^{-1} \mathbf{A}^H \mathbf{r} = \mathbf{d} + (\mathbf{A}^H \mathbf{A})^{-1} \mathbf{n} \tag{21}$$

From the Eq. (20), the zero-forcing (ZF) equalizer applies the inverse of the system matrix to separate the user signals and eliminate MAI and ISI, but the noise has some residual. This scheme is very popular and is considered as the early TD-SCDMA demos [4]. A Minimum-Mean Square Error (MMSE) detector minimizes the error between the weighted received signal and the desired bits and can result in lower BER at high SNR levels. The computational costs of the ZF equalizer are smaller than that for the MMSE detector because the latter requires an estimate of the noise covariance matrix; however, the implementation issues are very similar. This paper only addresses the zero-forcing detector.

4. Analyze of Channel Estimation Error

In TD-SCDMA system, various key technologies are built on the basis of precise estimation of CIRs, so the existence of errors can seriously affect the system performance. It is clearly found from (9) that the estimated CIRs have error compared with the real ones. The error is caused by the noise in the channel, which makes the noise power from the output is greater than in the real channel. What's more, it has a great impact on joint detection according to the law of error propagation.

Here, we use $\mathbf{\Delta}^{(k)} = (\Delta_1^k, \Delta_2^k \cdots \Delta_W^k)^T$ to build a bridge between the accurate channel response $\mathbf{h}^{(k)} = (h_1^k, h_2^k \cdots h_W^k)^T$ and the estimated ones $\hat{\mathbf{h}}^{(k)} = (h_1^k, h_2^k \cdots h_W^k)^T$, that is

$$\mathbf{h}^{(k)} = \hat{\mathbf{h}}^{(k)} + \mathbf{\Delta}^{(k)} \tag{22}$$

The combined CIR of the *k*th user can be rewritten as:

$$\begin{aligned}\mathbf{b}^{(k)} &= \mathbf{h}^{(k)} * \mathbf{c}^{(k)} \\ &= (\hat{\mathbf{h}}^{(k)} + \boldsymbol{\Delta}^{(k)}) * \mathbf{c}^{(k)} \\ &= \hat{\mathbf{h}}^{(k)} * \mathbf{c}^{(k)} + \boldsymbol{\Delta}^{(k)} * \mathbf{c}^{(k)} \\ &= \hat{\mathbf{b}}^{(k)} + \xi^{(k)}\end{aligned} \tag{23}$$

where $\xi^{(k)}$ is resulted errors and has the same structure as $\hat{\mathbf{b}}^{(k)}$. Due to constructing according to $\hat{\mathbf{b}}^{(k)}$, the system matrix $\mathbf{A}$ has the form resemble equation.

$$\mathbf{A} = \hat{\mathbf{A}} + \xi \tag{24}$$

Let $\mathbf{A}^{\dagger}$ is the pseudo-inverse of $\mathbf{A}$ and $\hat{\mathbf{r}} = \mathbf{r} - \mathbf{n}$. The solution $\hat{\mathbf{d}}_{ZF} = \mathbf{A}^{\dagger}\mathbf{r}$ is then perturbed to and the residual to $\hat{\mathbf{d}}_{ZF} + \delta\hat{\mathbf{d}}_{ZF} = \mathbf{A}^{\dagger}\hat{\mathbf{r}}$ and the residual $\boldsymbol{\zeta} = \mathbf{r} - \mathbf{A}\mathbf{d}_{ZF}$ to $\boldsymbol{\varepsilon} + \delta\boldsymbol{\varepsilon} = \hat{\mathbf{r}} - \mathbf{A}\mathbf{d}$. Let κ denote the spectral condition number $\kappa = \left\|\hat{\mathbf{A}}\right\|^2 \left\|\hat{\mathbf{A}}^{\dagger}\right\|^2$ and take $\varepsilon = \left\|\xi\right\|^2 \left\|\hat{\mathbf{A}}\right\|^2$ and $\mathbf{y} = \hat{\mathbf{A}}^{\dagger H}\mathbf{d}_{ZF} = (\hat{\mathbf{A}}\hat{\mathbf{A}}^{H})^{\dagger}\mathbf{r}$.

The following theorem gives bounds on $\left\|\delta\hat{\mathbf{d}}_{ZF}\right\|$ [13-14].

Theorem*:* Assume that $rank(\mathbf{A}) = rank(\hat{\mathbf{A}})$, and $\kappa\varepsilon < 1$. Then:

$$\delta x_{LS} \le \frac{\kappa}{(1-\kappa\varepsilon)\left\|\hat{\mathbf{A}}\right\|^2}\left(\varepsilon\left\|\hat{\mathbf{d}}_{ZF}\right\|^2\left\|\hat{\mathbf{A}}\right\|^2 + \left\|\mathbf{n}\right\|^2 + \kappa\varepsilon\left\|\zeta\right\|^2\right) + \varepsilon\left\|\mathbf{y}\right\|^2\left\|\hat{\mathbf{A}}\right\|^2 \tag{25}$$

NOTE. The last term of (24) vanishes if $rank(\hat{\mathbf{A}}) = rank(\mathbf{A})$, and $\left\|\hat{\mathbf{A}}^{\dagger}\right\|^2\left\|\xi\right\|^2 < 1$. then :

$$\delta\hat{\mathbf{d}}_{ZF} \le \frac{\kappa}{(1-\kappa\varepsilon)\left\|\hat{\mathbf{A}}\right\|^2}\left(\varepsilon\left\|\hat{\mathbf{d}}_{ZF}\right\|^2\left\|\hat{\mathbf{A}}\right\|^2 + \left\|\mathbf{n}\right\|^2 + \kappa\varepsilon\left\|\hat{\mathbf{A}}\right\|^2\right) \tag{26}$$

PROOF. From the decomposition theorem, $\hat{\mathbf{d}}_{ZF} = \hat{\mathbf{A}}^{\dagger}\mathbf{r}$ and $\hat{\mathbf{d}}_{ZF} + \delta\hat{\mathbf{d}}_{ZF} = \mathbf{A}^{\dagger}\hat{\mathbf{r}}$, it is seen that:

$$\delta\hat{\mathbf{d}}_{ZF} = \mathbf{A}^{\dagger}\hat{\mathbf{r}} - \hat{\mathbf{A}}^{\dagger}\mathbf{r} = (\mathbf{A}^{\dagger} - \hat{\mathbf{A}}^{\dagger})\mathbf{r} + \mathbf{A}^{\dagger}\mathbf{n} = (-\mathbf{A}^{\dagger}\xi\hat{\mathbf{d}}_{ZF} + \mathbf{A}^{\dagger}\zeta + \mathbf{A}^{\dagger}\mathbf{n}) - P_{N(\mathbf{A})}\hat{\mathbf{d}}_{ZF} \tag{27}$$

where the first part belongs to $R(\mathbf{A})$ and the second belongs to $N(\mathbf{A})$.

$$\left\|\mathbf{A}^{\dagger}\xi\hat{\mathbf{d}}_{ZF}\right\| \le \left\|\mathbf{A}^{\dagger}\right\|^2\left\|\xi\right\|^2\left\|\hat{\mathbf{d}}_{ZF}\right\|^2 \le \frac{\left\|\mathbf{A}^{\dagger}\right\|^2}{1-\kappa\varepsilon}\left\|\xi\right\|^2\left\|\hat{\mathbf{d}}_{ZF}\right\|^2 = \frac{\kappa\varepsilon}{1-\kappa\varepsilon}\left\|\hat{\mathbf{d}}_{ZF}\right\|^2 \tag{28}$$

$$\left\|\mathbf{A}^{\dagger}\zeta\right\| \le \left\|\mathbf{A}^{\dagger}\right\|^2\left\|\xi\right\|^2\left\|\zeta\right\|^2 \le \frac{\left\|\mathbf{A}^{\dagger}\right\|^2}{1-\kappa\varepsilon}\left\|\xi\right\|^2\left\|\zeta\right\|^2 \le \frac{\kappa^2\varepsilon}{1-\kappa\varepsilon}\frac{\left\|\zeta\right\|^2}{\left\|\hat{\mathbf{A}}\right\|^2} \tag{29}$$

$$\left\|\mathbf{A}^{\dagger}\delta b\right\| \le \left\|\mathbf{A}^{\dagger}\right\|^2\left\|\mathbf{n}\right\|^2 \le \frac{\kappa}{1-\kappa\varepsilon}\frac{\left\|\mathbf{n}\right\|^2}{\left\|\hat{\mathbf{A}}\right\|^2} \tag{30}$$

$$\begin{aligned}\left\|P_{N(\mathbf{A})}\hat{\mathbf{d}}_{ZF}\right\| &= \left\|(\mathbf{I} - \mathbf{A}^{H}\mathbf{A}^{H\dagger})\hat{\mathbf{A}}^{\dagger}\hat{\mathbf{A}}\hat{\mathbf{A}}^{\dagger}\mathbf{r}\right\| \\ &= \left\|(\mathbf{I} - \mathbf{A}^{H}\mathbf{A}^{H\dagger})\hat{\mathbf{A}}^{H}\hat{\mathbf{A}}^{\dagger H}\mathbf{r}\right\| \le \left\|\xi\right\|^2\left\|\hat{\mathbf{A}}^{\dagger H}\hat{\mathbf{d}}_{ZF}\right\| = \left\|\xi\right\|^2\left\|\mathbf{y}\right\|\end{aligned} \tag{31}$$

And arrive at (24) from the expression for $\delta\hat{\mathbf{d}}_{ZF}$. Where $R(A)$ is Subspace A matrix, $N(A)$ is Null space matrix A , $\|\bullet\|$ is Frobenius norm

When $\boldsymbol{\xi} = \mathbf{0}$ and $\mathbf{n}$ is a zero mean white Gaussian noise vector, $\hat{\mathbf{d}}_{ZF}$ has zero bias. It also is the maximum-likelihood estimator. When $\boldsymbol{\xi} \neq \mathbf{0}$, $\hat{\mathbf{d}}_{ZF}$ will in general be biased, and will exhibit a pool performance. It suffers from perturbation of noise errors and increase covariance due to the accumulation of interference in $\mathbf{A}^H\mathbf{A}$.

5. Simulation Result and Analysis

In this section, simulations are carried out to evaluate the performance of the ZF-BLE under the CIR with error. The TD-SCDMA parameters used in our simulations are listed in Table 1 and all the datum have been conducted according to 3GPP specifications. Wireless fading channel according to the multipath fading channel environment 3GPP TR25.945 required to set up multi-path delay and power. Two scenarios of propagation channels Vehicular A and Indoor B are considered, which are listed in Table 2.

Table 1. Simulation Parameters

Chip Rate	1.28 Mcps
Modulation	QPSK
User data rate	12.2kbps
Channel length (*W*)	16 chips
Active users per slot (K)	8users
Interfering users	4users
Carrier frequency	2.0GHz
Effective training sequence length (P)	128 chips
Spreading factor (Q_k)	16 for each user
RRC roll-off factor	0.22

Table 2. Parameters of Propagation Channel Environment

Number of paths	Case 1, speed 3km/h		Case 3, speed 120km/h	
	Relative delay [ns]	Relative mean power [dB]	Relative delay [ns]	Relative mean power [dB]
1	0	0	0	0
2	2928	-10	781	-3
3			1563	-6
4			2344	-9

In the following, we compare the performance of ZF-BLE under estimated channel and ideal channel. Figure 2-3 present the results for both scenarios, assuming channel state information is known at the receiver. We can clearly see that the error of channel has a great impact on the performance of joint detection.

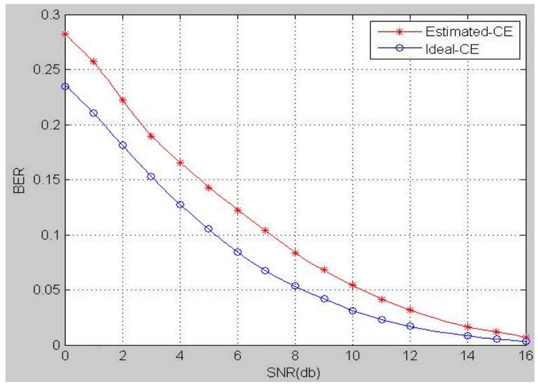

Figure 2. BER Comparison between Estimated Channel and Ideal Channel for Case 1

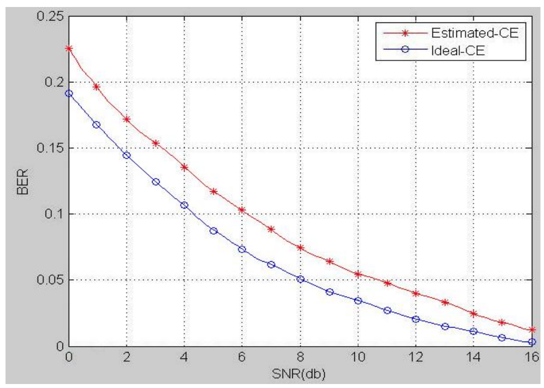

Figure 3. BER Comparison between Estimated Channel and Ideal Channel for Case 3

6. Conclusion

In this paper, we analyze the performance of joint detection with channel estimation error. The result is verified by computer simulation. So, the detection algorithm should be taken in account the channel estimation error and noise.

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Authors

Tongliang Fan, is currently a Lecturer at Electronic Technology Department, China Maritime Police Academy since June 2012. He obtained his Ph. D. degree in the Center of Communication and Tracking Telemetry Command, Chongqing University, China, in June 2012. He obtained his M.S. in Chongqing University of Posts and Telecommunications and studied network management and network flow measurement from 2003 to 2006. He worked as an algorithm engineer with Chongqing CYIT Communication Technologies CL., LTD from 2006 until 2008. His research interest and activities include PHY layer design for terrestrial (4G) communications based on OFDM, antenna diversity techniques for wireless communications broadcast systems and positioning in wireless communications, shortwave and Ultrashort-wave in military communications.

Shengbo Huang, is currently a Professor at Electronic Technology Department, China Maritime Police Academy. He enjoy the Ministry of public security science and technology grants, engaged in police communication teaching and research work for more than 25 years, focusing on the information integration technology, military radio communication and emergency communication guarantee standard operation etc.. Has presided over the "public security marine police equipment and information integration construction plan", "border control forces of public security police information training operation mechanism and evaluation system research" and other ministry project, The results of the promotion of national border troop.

Yandong Ma, received the B.S. degree from Zhengzhou University in 2000, his M.S. from Zhejiang University in 2008. He has been working on wireless communication technologies including TD-SCDMA, Antenna, GSM since 2000.His research interests include communication simulation and multi-carrier based communication systems, in particular channel estimation and synchronization techniques. He has published 2 works as well as more than 10 research papers. Now, he works as an associate professor of electronics technology department in China maritime police academy.

Xue Zhao, is currently a Lecturer at Electronic Technology Department, China Maritime Police Academy. She received her B.E. and M.E. degree in Communication Engineering in 2003 and 2006, respectively, from Hehai University, China. She currently works in China Maritime Police Academy. Her research interests include signal processing, communication command and learning mechanism in the field of wireless communication.

Dingyi Ji, is currently a Lecturer at Electronic Technology Department, China Maritime Police Academy. He received his B.E. and M.E. degree in Automation Engineering in 2001 and 2007, respectively, from Nanjing University of Aeronautics and Astronautics, China. He currently works in China Maritime Police Academy. His research interests include signal processing, navigation, wireless communication, shortwave and Ultrashort-wave in military communications.

