HIGH TECHNOLOGY LETTERS | Vol. 20 No. 4 | Dec. 2014 | pp. 415 ~ 420

doi: 10.3772/j.issn.1006-6748.2014.04.012

Error propagation determined iterative channel estimation with ICI mitigation for fast time-varying OFDM channels①

Zhang Jie (张 杰)*, Yang Ruizhe*, Si Pengbo*, Zhang Yanhua②*, Yu Richard* **
(* College of Electronics Information and Control Engineering, Beijing University of Technology, Beijing 100124, P. R. China)
(** Department of Systems and Computer Engineering, Carleton University, Ottawa K1S 5B6, Canada)

Abstract

The intersubcarrier interference (ICI) degrades the performance of the pilot-aided channel estimation in fast time-varying orthogonal frequency division multiplexing (OFDM) systems. To solve the error propagation in joint channel estimation and data detection due to this ICI, a scheme of error propagation determined iterative estimation is proposed, where in the first iteration, Kalman filter based on signal to interference and noise is designed with ICI transformed to be part of the noise, and for the later iterations, a determined iterative estimation algorithm obtains an optimal output from all iterations using the iterative updating strategy. Simulation results present the significant improvement in the performance of the proposed scheme in high-mobility situation in comparison with the existing ones.

Key words: orthogonal frequency division multiplexing (OFDM), channel estimation, intersubcarrier interference (ICI), error propagation, Kalman

0 Introduction

Orthogonal frequency division multiplexing (OFDM)[1] is a promising technique which has been widely applied in wireless communication systems, e.g. mobile WiMAX (IEEE 802.16e), WAVE (IEEE 802.11p), and DVB-T (ETSI EN 300 744), due to the high bandwidth efficiency and robustness against multipath delay. However, the orthogonality between subcarriers can be easily to be broken down in time-varying environment, resulting in the intersubcarrier interference (ICI) and system performance degradation.

In fast time-varying situations, channels may change significantly even within a single OFDM symbol and the ICI effect will be exacerbated. As a consequence, some conventional channel estimation methods are not suitable any more. To mitigate this ICI effect on channel estimation, self-interference cancellation techniques[2] map the information onto a group of subcarriers, leading to a strong cancellation in the self-interference, while with a reduction of the spectral efficiency. An equalization techniques is designed[3] with an operator getting from minimum mean square error (MMSE), but it is high complex and may induce some noise amplification. In Ref. [4], the authors suggest an iterative technique for the equalization of ICI and also the iterative channel estimations are proposed in Refs[5-7], where the detected data are used to enhance the estimation. However, the authors in Ref. [5] do not take the Doppler spread information into account and the the frequency-domain estimation[6,7] affected by ICI evolves lots of iterations to get a good result and may have limited estimation result at low SNR.

In this paper, the fast time-varying OFDM channels having ICI at terminals are considered. To deal with the problem that the accuracy of frequency-domain estimation is affected by ICI, a determined iterative channel estimation with ICI mitigation is proposed based on the analysis of the relationship between ICI and estimation accuracy. Simulation results demonstrate a significant improvement in performance compared with the existing scheme, especially when SNR is low. The main contributions of the present work are as follows:

- A new method is proposed to reduce the interactions between ICI and the estimation accuracy by denoising "SIN—I + N" (sum of ICI and channel noise) with Kalman filter, so that the pilot-aided estimation

① Supported by the National Natural Science Foundation of China (No. 61372089, 61101113, 61072088, 61201198), and the Beijing Natural Science Foundation (No. 4132019).
② To whom correspondence should be addressed. E-mail: zhangyh@bjut.edu.cn
Received on Apr. 3, 2013

can be performed regardless of data symbol when there is no sufficiently accurate data.

- A determined iterative updating strategy is proposed to restrain the error propagation caused by ICI, where the optimal estimation is selected by the criterion designed from all iterations and the accuracy of the channel estimation is improved.

The rest of this paper is organized as follows. Section 1 describes the system model. The proposed scheme is introduced in Section 2. The simulation results are discussed in Section 3. Finally, Section 4 concludes the paper.

1 System model

1.1 OFDM system model

Considering an OFDM system with N subcarriers, and a cyclic prefix (CP) with the length of N_{cp}, the duration of an OFDM symbol is $T = (N_{cp} + N)T_s = N_s T_s$, where T_s is the sampling time. The time-domain received signal can be expressed as

$$v(n) = \sum_{l=0}^{L-1} h(n,l)s(n-l) + w(n) \quad 0 \leqslant n \leqslant N-1 \tag{1}$$

where $s(n)$ and $v(n)$ are the nth sample in the transmitted and received OFDM symbol, respectively, $w(n)$ is white complex Gaussian noise with covariance σ^2, and L is the length of channels. Channel impulse response $h(n,l)$ is wide-sense stationary (WSS) narrow-band complex Gaussian process. For the classical Jakes' model, they are independent for different paths and the normalized correlation function is $R_{h(n,l)}^{(a)} = E[h(n,l)h^*(n+a,l)] = \sigma_{h(n,l)}^2 J_0(2\pi f_d T_s a)$, where $E[\cdot]$ denotes expectation, $J_0(\cdot)$ denotes the zeroth-order Bessel function of the first kind, $f_d = vf_c/c$ is the maximum Doppler frequency with speed v, carrier frequency f_c, speed of light c, and $\sigma_{h(n,l)}^2$ is the average power of the lth path. The average power of the multipath channel is unit, $\sum_{l=0}^{L-1} \sigma_{h(n,l)}^2 = 1$.

Let $\boldsymbol{x}^{(r)} = [x_0^{(r)}, x_1^{(r)}, \cdots, x_{N-1}^{(r)}]^{\mathrm{T}}$, $\boldsymbol{y}^{(r)} = [y_0^{(r)}, y_1^{(r)}, \cdots, y_{N-1}^{(r)}]^{\mathrm{T}}$ be the rth transmitted and received OFDM symbol in frequency-domain, from Eq. (1) there is

$$\boldsymbol{y}^{(r)} = \boldsymbol{H}^{(r)} \boldsymbol{x}^{(r)} + \boldsymbol{w}^{(r)} \tag{2}$$

where $\boldsymbol{w}^{(r)}$ is frequency-domain white noise, $\boldsymbol{H}^{(r)}$ is a $N \times N$ channel matrix with elements given by

$$\boldsymbol{H}^{(r)}(m,m') = \sum_{l=0}^{L-1} \boldsymbol{G}_l^{(r)}(m, m') e^{-j2\pi(\frac{m'-1}{N} - \frac{1}{2})\tau_l} \tag{3}$$

$$\boldsymbol{G}_l^{(r)}(m,m') = \frac{1}{N} \sum_{n=0}^{N-1} h^{(r)}(n,l) e^{-j2\pi(m-m')n/N} \tag{4}$$

where m, m' denotes the $[m,m']$th element of the matrix, $\tau_l \times T_s$ is the delay of the lth tap. $\boldsymbol{H}^{(r)}(m,m')$, $m \neq m'$ is the ICI coefficient from subcarrier m' to subcarrier m. For time-invariant channels, $\boldsymbol{H}^{(r)}$ can be regarded as diagonal. However, as f_d increases, the ICI terms on the off-diagonals of $\boldsymbol{H}^{(r)}$ can not be neglected, and $\boldsymbol{H}^{(r)}$ should be treated as banded or even a full matrix.

Assuming that each OFDM symbol contains N_p pilots and $N_d = N - N_p$ data. In this paper, Comb-type pilot is used, where the N_p pilot subcarriers are fixed during the transmission and inserted into N subcarriers with equal intervals. The interval N/N_p can be selected without respecting the sampling theorem, but N_p must fulfill the requirement that $N_p \geqslant L$. Let $\boldsymbol{p}_s$, $\boldsymbol{d}_s$ denote the pilot and data indices respectively in frequency direction: $\boldsymbol{p}_s = [p_1, p_2, \cdots, p_{N_p}]$, $\boldsymbol{d}_s = [d_1, d_2, \cdots, d_{N_d}]$, where $p_i = (i-1)N/N_p + 1$, $i = 1, 2, \cdots, N_p$. Therefore, signals $\boldsymbol{x}_p^{(r)} = \boldsymbol{x}^{(r)}(\boldsymbol{p}_s) = [x_{p_1}^{(r)}, x_{p_2}^{(r)}, \cdots, x_{p_{N_p}}^{(r)}]^{\mathrm{T}}$ are pilots, while others are data $\boldsymbol{x}_d^{(r)} = \boldsymbol{x}^{(r)}(\boldsymbol{d}_s)$.

1.2 BEM channel model

Basis expansion model (BEM) is utilized to approximate the doubly selective fading channels. In BEM, the multipath channel response is modeled as a FIR filter and each filter tap is represented as a superposition of some basis functions. The channel impulse response is given by

$$h^{(r)}(n,l) = \boldsymbol{Q}(n,:)\boldsymbol{c}_l^{(r)} + \xi_l^{(r)}(n), \quad -N_{cp} \leqslant n \leqslant N-1 \tag{5}$$

where $\xi_l^{(r)}(n)$ represents the corresponding modeling error, $\boldsymbol{Q}$ is a $N_s \times Q$ orthonormal basis function matrix, vector $\boldsymbol{c}_l^{(r)} = [c_{1,l}^{(r)}, \cdots, c_{Q,l}^{(r)}]^{\mathrm{T}}$ is the BEM coefficients of size Q for the lth complex gain of the rth OFDM symbol, generally, $Q \geqslant 2\lceil f_d N_s T_s \rceil$. Based on different basic functions, various traditional BEM designs have been reported to model the channel time-variations. Without loss of generality, polynomial BEM (P-BEM) $\boldsymbol{Q}(m,q) = (m - N_{cp} - 1)^{q-1}$ is used[8], but the proposed algorithm can be applied with arbitrary basis. Using BEM, the channel variety is tracked with LQ parameters $c_{q,l}^{(r)}$ instead of the LN_s samples of the complex gains, and the whole channel response can be retrieved by the estimated $\hat{c}_{q,l}^{(r)}$ using Eq. (5). Using this polynomial approximation, the received signal in Eq. (2) can be rewritten as

$$\boldsymbol{y}^{(r)} = \boldsymbol{B}^{(r)} \boldsymbol{g}^{(r)} + \boldsymbol{w}^{(r)} \tag{6}$$

where $\boldsymbol{g}^{(r)} = [(\boldsymbol{c}_1^{(r)})^{\mathrm{T}}, \cdots, (\boldsymbol{c}_L^{(r)})^{\mathrm{T}}]^{\mathrm{T}}$, $\boldsymbol{B}^{(r)} = \frac{1}{N}[\boldsymbol{Z}_1^{(r)}, \cdots, \boldsymbol{Z}_L^{(r)}]$, $\boldsymbol{Z}_l^{(r)} = \frac{1}{N}[\boldsymbol{M}_1 \mathrm{diag}(\boldsymbol{x}^{(r)})\boldsymbol{f}_l, \cdots, \boldsymbol{M}_Q \mathrm{diag}(\boldsymbol{x}^{(r)})\boldsymbol{f}_l]$. $\boldsymbol{f}_l$ is the lth column of the Fourier matrix $\boldsymbol{F}(m,k) = e^{-j2\pi(\frac{m-1}{N}-\frac{1}{2})\tau_k}$ and $\boldsymbol{M}_q(m,k) = \sum_{n=0}^{N-1} n^{q-1} e^{-j2\pi(m-k)n/N}$, $q = 1, \cdots, Q$. It shows that $\boldsymbol{B}$ is related to all the transmitted symbols $\boldsymbol{x}^{(r)}$, therefore if estimation is done on the pilot subcarriers in frequency-domain, its accuracy will be affected by the ICI from unknown data.

2 Channel estimation and ICI mitigation

2.1 SIN channel estimation

Since $\boldsymbol{C}_l^{(r)}$ is correlated complex Gaussian variable with zero- means and its correlation matrix can be expressed as: $\boldsymbol{R}_{\mathbf{C}_l}^{(a)} = (\boldsymbol{Q}^H\boldsymbol{Q})^{-1}\boldsymbol{Q}^H\boldsymbol{R}_{\mathbf{h}(:,l)}^{(a)}\ \boldsymbol{Q}(\boldsymbol{Q}^H\boldsymbol{Q})^{-1}$, the AR model for channel BEM parameters $\boldsymbol{g}_{(r)}$ is[6]:

$$\boldsymbol{g}^{(r)} = \boldsymbol{A}_1\boldsymbol{g}^{(r-1)} + \boldsymbol{u}^{(r)} \tag{7}$$

where $\boldsymbol{A}_1 = \mathrm{diag}(\boldsymbol{\alpha}_1, \boldsymbol{\alpha}_2, \cdots, \boldsymbol{\alpha}_L)$, $\boldsymbol{V}[\boldsymbol{u}_{(r)}] = \mathrm{diag}(\boldsymbol{\beta}_1, \boldsymbol{\beta}_2, \cdots, \boldsymbol{\beta}_L)$ and $\boldsymbol{\alpha}_l = \boldsymbol{R}_{\boldsymbol{C}_l}^{(1)}(\mathbf{R}_{\mathbf{C}_l}^{(0)})^{-1}$, $\boldsymbol{\beta}_l = \boldsymbol{R}_{\boldsymbol{C}_l}^{(0)} + \boldsymbol{\alpha}_l\boldsymbol{R}_{\boldsymbol{C}_l}^{(-1)}$ can be obtained from Yule-Walker equations. $\boldsymbol{V}[\cdot]$ stands for the covariance matrix.

Therefore, using the state Eq. (7) and the measurement Eq. (6) in the form of Kalman, coefficient $\boldsymbol{g}^{(r)}$ can be tracked. However, as mentioned previously, transmitted symbol $\boldsymbol{x}^{(r)}$ is required for parameter $\boldsymbol{B}$ that in fact, only the pilot is exactly known at the receiver and the data is to be detected or maybe in accurately detected. In Ref. [7], for the first iteration, the authors have the data in $\boldsymbol{B}$ detected with the predicted channel matrix from the previous time using $\boldsymbol{g}^{(r)} = \boldsymbol{A}_1\boldsymbol{g}^{(r-1)}$. This initial inaccuracy data degrades the whole Kalman filtering and causes a performance floor which may not be retrieved, even though the iterative joint channel estimation and data detection are adopted. It likely happened in the case of low SNR or high speed where prediction error in AR model is non-negligible, seen simulations.

To overcome this performance floor, a new method is proposed, where the sum of the statistics ICI and the additive channel noise (SIN—I + N) are considered as the equivalent noise and therefore the Kalman filter estimates $\boldsymbol{g}^{(r)}$ using exact pilots without the data bothering in the first iteration. That is, the received signals on the pilots subcarriers in Eq. (6) are divided into three parts, related with the pilots, data and noise respectively,

$$\boldsymbol{y}_p^{(r)} = \boldsymbol{y}^{(r)}(\boldsymbol{p}_s) = \underbrace{\boldsymbol{H}_{(\mathbf{p}_s,\mathbf{p}_s)}^{(r)}\boldsymbol{x}_p^{(r)}}_{\text{pilots component}} + \underbrace{\boldsymbol{H}_{(\mathbf{p}_s,\mathbf{d}_s)}^{(r)}\boldsymbol{x}_d^{(r)}}_{\text{data's ICI}} + \boldsymbol{w}^{(r)}(\boldsymbol{p}_s)$$

$$= \underbrace{\begin{bmatrix} \boldsymbol{H}^{(r)}(p_1,p_1) & \boldsymbol{H}^{(r)}(p_1,p_2) & \cdots & \boldsymbol{H}^{(r)}(p_1,p_{N_p}) \\ \boldsymbol{H}^{(r)}(p_2,p_1) & \boldsymbol{H}^{(r)}(p_2,p_2) & \cdots & \boldsymbol{H}^{(r)}(p_2,p_{N_p}) \\ \vdots & \vdots & \ddots & \vdots \\ \boldsymbol{H}^{(r)}(p_{N_p},p_1) & \boldsymbol{H}^{(r)}(p_{N_p},p_2) & \cdots & \boldsymbol{H}^{(r)}(p_{N_p},p_{N_p}) \end{bmatrix} \begin{bmatrix} \boldsymbol{x}^{(r)}(p_1) \\ \boldsymbol{x}^{(r)}(p_2) \\ \vdots \\ \boldsymbol{x}^{(r)}(p_{N_p}) \end{bmatrix}}_{\text{pilots component}}$$

$$+ \underbrace{\begin{bmatrix} \boldsymbol{H}^{(r)}(p_1,d_1) & \boldsymbol{H}^{(r)}(p_1,d_2) & \cdots & \boldsymbol{H}^{(r)}(p_1,d_{N_d}) \\ \boldsymbol{H}^{(r)}(p_2,d_1) & \boldsymbol{H}^{(r)}(p_2,d_2) & \cdots & \boldsymbol{H}^{(r)}(p_2,d_{N_d}) \\ \vdots & \vdots & \ddots & \vdots \\ \boldsymbol{H}^{(r)}(p_{N_p},d_1) & \boldsymbol{H}^{(r)}(p_{N_p},d_2) & \cdots & \boldsymbol{H}^{(r)}(p_{N_p},d_{N_d}) \end{bmatrix} \begin{bmatrix} \boldsymbol{x}^{(r)}(d_1) \\ \boldsymbol{x}^{(r)}(d_2) \\ \vdots \\ \boldsymbol{x}^{(r)}(d_{N_d}) \end{bmatrix}}_{\text{non-pilot data ICI}}$$

$$+ \begin{bmatrix} \boldsymbol{W}^{(r)}(p_1) \\ \boldsymbol{W}^{(r)}(p_1) \\ \vdots \\ \mathbf{W}^{(r)}(p_{N_p}) \end{bmatrix} \tag{8}$$

where $\boldsymbol{V}[\boldsymbol{w}^{(r)}] = \sigma^2\boldsymbol{I}_{N_p}$, $\boldsymbol{I}_{N_p}$ is an $N_p \times N_p$ identity matrix. Part 1 is the desired pilot term without ICI of unknown data and part 2 is the ICI term, which represents the ICI coefficient from data subcarriers. If part 2 is nonexistent in the observation, Kalman estimation can be performed only with observation of the transmitted pilots. To this end, the data's ICI term is transformed into part of the equivalent channel noise $\boldsymbol{w}^{SIN(r)} = \boldsymbol{H}_{[\mathbf{p}_s,\mathbf{d}_s]}^{(r)}\boldsymbol{x}_d^{(r)} + \boldsymbol{w}^{(r)}(\boldsymbol{p}_s)$. Then, the measurement equation with SIN is rewritten as

$$\boldsymbol{y}_p^{(r)} = \boldsymbol{B}^{SIN(r)}\boldsymbol{g}^{(r)} + \boldsymbol{w}^{SIN(r)} \tag{9}$$

where $\boldsymbol{B}^{SIN(r)}$ can be calculated without the unknown data:

$$\boldsymbol{B}^{SIN(r)} = \frac{1}{N}[\boldsymbol{Z}_1^{SIN(r)}, \boldsymbol{Z}_2^{SIN(r)}, \cdots, \boldsymbol{Z}_L^{SIN(r)}]$$

$$\boldsymbol{Z}_l^{SIN(r)} = \frac{1}{N}[\boldsymbol{M}_1^{SIN}\mathrm{diag}(\boldsymbol{x}_p^{(r)})\boldsymbol{f}_l^{SIN}, \cdots, \boldsymbol{M}_Q^{SIN}\mathrm{diag}(\boldsymbol{x}_p^{(r)})\boldsymbol{f}_l^{SIN}]$$

$$\boldsymbol{f}_l^{SIN} = \begin{bmatrix} e^{-j2\pi(\frac{p_1-1}{N}-\frac{1}{2})\tau_l} & e^{-j2\pi(\frac{p_2-1}{N}-\frac{1}{2})\tau_l} & \cdots & e^{-j2\pi(\frac{p_{N_p}}{N}-\frac{1}{2})\tau_l} \end{bmatrix}^{\mathrm{T}}$$

$$\boldsymbol{M}_q^{SIN} = \begin{bmatrix} \sum_{n=0}^{N-1} n^{q-1} & \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_1-p_2)n}{N}} & \cdots & \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_1-p_{N_p})n}{N}} \\ \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_2-p_1)n}{N}} & \sum_{n=0}^{N-1} n^{q-1} & \cdots & \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_2-p_{N_p})n}{N}} \\ \vdots & \vdots & \ddots & \vdots \\ \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_{N_p}-p_1)n}{N}} & \sum_{n=0}^{N-1} n^{q-1} e^{\frac{-j2\pi(p_{N_p}-p_2)n}{N}} & \cdots & \sum_{n=0}^{N-1} n^{q-1} \end{bmatrix}$$

Here, the covariance of $\boldsymbol{w}^{SIN(r)}$ is no longer just the white complex Gaussian noise, but the sum of ICI and channel noise. In Ref. [9], it shows that the residual ICI of OFDM has high normalized autocorrelation and this normalized autocorrelation is insensitive to the multipath channel profile as well as a variety of other system and channel conditions. Assuming transmitted OFDM symbol to be white, due to the independence between ICI and channel noise, the approximate mathematical expression for covariance matrix of SIN can be derived as

$$\begin{aligned} \boldsymbol{V}[\boldsymbol{w}^{SIN(r)}] &= \boldsymbol{V}[\boldsymbol{H}^{(r)}_{[D_s,D_s]}\boldsymbol{x}^{(r)}_d] + \boldsymbol{V}[\boldsymbol{W}^{(r)}(\boldsymbol{P}_s)] \\ &= \boldsymbol{V}[\boldsymbol{H}^{(r)}_{[D_s,D_s]}\boldsymbol{x}^{(r)}_d] + \sigma^2\boldsymbol{I}_{N_p} \end{aligned} \quad (10)$$

where $\boldsymbol{V}_{ICI} = \boldsymbol{V}[\boldsymbol{H}^{(r)}_{[p_s,d_s]}\boldsymbol{x}^{(r)}_d]$ can be calculate by:

$$\boldsymbol{V}_{ICI}(m,m') = R^{(m-m')}_{ICI}$$

$$\begin{aligned} R^{(a)}_{ICI} &= \frac{E_s}{N^2}\sum_{l=0}^{L-1}\sum_{n=0}^{N-1}\sum_{n'=0}^{N-1}\sum_{k\neq 0,k\neq -a}\sigma^2_{h(n,l)}R^{(n-n')}_{h(n,l)}e^{j2\pi[n'(k+a)-nk]/N} \\ &\approx 4\pi^2T_s^2E_s(\sum_{l=0}^{L-1}\sigma^2_{h(n,l)}\sigma_{D_l})\rho(a,N) \end{aligned} \quad (11)$$

where E_s is the average transmitted symbol energy, $\sigma_{D_l} = \int_{f_d}^{fl} P_l(f)f^2\mathrm{d}f$, $P_l(f) = P_v/2\pi\sqrt{f_d^2 - f^2}$ is Doppler power spectral density with average power P_v and:

$$\rho(m,N) \approx \begin{cases} N^2/12, & m = 0 \\ N^2/(2\pi^2m^2), & m \neq 0 \end{cases} \quad (12)$$

Thus, using the SIN measurement Eq. (9) together with the state Eq. (7), the stages in the first SIN iteration for each OFDM symbols in Kalman filtering are

1) Time update equation:

$$\begin{aligned} \hat{\boldsymbol{g}}^{(r|r-1)}_1 &= \boldsymbol{A}_1\hat{\boldsymbol{g}}^{(r-1)}_o \\ \boldsymbol{P}^{(r|r-1)}_1 &= \boldsymbol{A}_1\boldsymbol{P}^{(r-1)}_o\boldsymbol{A}^H_1 + \boldsymbol{V}[\boldsymbol{W}^{SIN(r)}] \end{aligned} \quad (13)$$

2) Measurement update equations:

$$\begin{aligned} \boldsymbol{K}^{(r)}_1 &= \boldsymbol{P}^{(r|r-1)}_1(\boldsymbol{B}^{SIN(r)})^H(\boldsymbol{B}^{SIN(r)})\boldsymbol{P}^{(r|r-1)}_1(\boldsymbol{B}^{SIN(r)})^H \\ &\quad \boldsymbol{V}[\boldsymbol{w}^{SIN(r)}])^{-1} \\ \hat{\boldsymbol{g}}^{(r)}_1 &= \hat{\boldsymbol{g}}^{(r|r-1)}_1 + \boldsymbol{K}^{(r)}_1(\boldsymbol{y}^{(r)}_p - \boldsymbol{B}^{SIN(r)}\hat{\boldsymbol{g}}^{(r|r-1)}_1) \\ \boldsymbol{P}^{(r)}_1 &= \boldsymbol{P}^{(r|r-1)}_1 - \boldsymbol{K}^{(r)}_1\boldsymbol{B}^{SIN(r)}\boldsymbol{P}^{(r|r-1)}_1 \\ \hat{\boldsymbol{g}}^{(r)}_o &= \hat{\boldsymbol{g}}^{(r)}_1,\ \boldsymbol{P}^{(r)}_o = \boldsymbol{P}^{(r)}_1 \end{aligned} \quad (14)$$

where $\hat{\boldsymbol{g}}^{(r)}_i$, $\boldsymbol{K}^{(r)}_i$ and $\boldsymbol{P}^{(r)}_i$ are the estimated result, Kalman gain and covariance matrix for the rth OFDM symbol in ith iteration, respectively, $\hat{\boldsymbol{g}}^{(r-1)}_o$, $\boldsymbol{P}^{(r)}_o$ are the final optimal output for the rth OFDM symbol which will be updated in the following. $\boldsymbol{P}^{(0)}_o = \mathrm{diag}(\boldsymbol{R}^{(0)}_{C_1}, \boldsymbol{R}^{(0)}_{C_2}, \cdots, \boldsymbol{R}^{(0)}_{C_L})$ and $\hat{\boldsymbol{g}}^{(0)}_o = 0_{LQ,1}$ are initialized and r above starts from 1.

2.2 Error propagation determined iterative estimation

With the BEM parameters $\hat{\boldsymbol{g}}^{(r)}_1$ from the SIN Kalman estimation, channel matrix is computed by

$$\hat{\boldsymbol{H}}^{(r)}_i = \frac{1}{N}\sum_{q=1}^{Q}\boldsymbol{M}_q\mathrm{diag}(\boldsymbol{F}\hat{\boldsymbol{g}}^{(r)}_{i,q}) \quad (15)$$

where $\hat{\boldsymbol{g}}^{(r)}_{i,q} = [\hat{c}^{(r)}_{q,1}, \hat{c}^{(r)}_{q,2}, \cdots, \hat{c}^{(r)}_{q,L}]^{\mathrm{T}}$ can be obtain from $\hat{\boldsymbol{g}}^{(r)}_i$ in Eq. (14). Then using the QR detection[7] with the estimated $\hat{\boldsymbol{H}}^{(r)}_1$, the data symbol $\hat{\boldsymbol{x}}^{(r)}_{d,1}$ can be detected. Hereafter, the iterative standard Kalman estimation can be executed to update the estimated channels $\hat{\boldsymbol{H}}^{(r)}_i$ using the detected $\hat{\boldsymbol{x}}^{(r)}_{d,i-1}(\boldsymbol{B}^{(r)}_{i-1})$ and detect the data $\hat{\boldsymbol{x}}^{(r)}_{d,i}$ with the estimated channels $\hat{\boldsymbol{H}}^{(r)}_i$. However, it is worth noting that not every estimated subcarrier channel coefficient is improved along the estimation iterations, sometimes errors in the previous result will affect the follow-up iterations by ICI, even update some correct ones into wrong. The essence behind this phenomenon is that the promotion between channel estimation and data detection is limited by ICI between them. The accuracy of channel estimation limits data detection and inaccurate data cannot improve the channel estimation. Unknown data or data errors will not only lead to estimation error on the corresponding subcarriers but also influence other subcarriers through ICI, this is the **error propagation.** In order to restrain the error propagation, an iterative updating strategy with a criteria for judging the accuracy of the Kalman is proposed to determine whether the estimation result is updated or not. Since the measurement residual $\boldsymbol{y}^{(r)} - \hat{\boldsymbol{y}}^{(r)}_i$ denotes the error between the real observation and the estimated received symbol[10] which can determine the Kalman estimation's accuracy roughly, the determined criteria to analyze the degree of error propagation are defined:

$$e^i = \frac{1}{N}\sum_{m=1}^{N}(|[\boldsymbol{y}^{(r)}]_m - [\boldsymbol{B}^{(r)}_i\hat{\boldsymbol{g}}^{(r)}_i]_m|^2) \quad (16)$$

where $[g]_m$ denotes the mth element of g, i represents the iteration number. It decides whether to update the estimated channel BEM parameters $\hat{\boldsymbol{g}}^{(r)}_o$ and corresponding covariance matrix $\boldsymbol{P}_{(r)}$ and find the optimal estimated result. Therefore, the determined iterative estimation is

1) Time update equation:

$$\begin{aligned} \hat{\boldsymbol{g}}^{(r|r-1)}_i &= \boldsymbol{A}_1\hat{\boldsymbol{g}}^{(r-1)}_o \\ \boldsymbol{P}^{(r|r-1)}_i &= \boldsymbol{A}_1\boldsymbol{P}^{(r-1)}_o\boldsymbol{A}^H_1 + \boldsymbol{V}[\boldsymbol{u}^{(r)}] \end{aligned} \quad (17)$$

2) Measurement update equations:

$$\begin{aligned} \boldsymbol{K}^{(r)}_i &= \boldsymbol{P}^{(r|r-1)}_i(\boldsymbol{B}^{(r)}_i)^H(\boldsymbol{B}^{(r)}_i\boldsymbol{P}^{(r|r-1)}_i(\boldsymbol{B}^{(r)}_i)^H \\ &\quad + \boldsymbol{V}[\boldsymbol{w}^{(r)}])^{-1} \\ \hat{\boldsymbol{g}}^{(r)}_i &= \hat{\boldsymbol{g}}^{(r|r-1)}_i + \boldsymbol{K}^{(r)}_i(\boldsymbol{y}^{(r)} - \boldsymbol{B}^{(r)}_i\hat{\boldsymbol{g}}^{(r|r-1)}_i) \\ \boldsymbol{P}^{(r)}_i &= \boldsymbol{P}^{(r|r-1)}_i - \boldsymbol{K}^{(r)}_i\boldsymbol{B}^{(r)}_i\boldsymbol{P}^{(r|r-1)}_i \end{aligned} \quad (18)$$

3) Determined estimation:

if $e^i < e^o$

$$\hat{\boldsymbol{g}}^{(r)}_o = \hat{\boldsymbol{g}}^{(r)}_i,\ \boldsymbol{P}^{(r)}_o = \boldsymbol{P}^{(r)}_i \quad (19)$$

The scheme flow is shown in Fig. 1.

3 Simulation results and analysis

In this section, the performance of the algorithm proposed by Matlab simulations is presented, compared with the method in Ref. [7]. The normalized channel model is Rayleigh as recommended by GSM recommendations 05.05[11]. The delays are assumed invariant (over several OFDM symbols) and perfectly estimated[12]. Assume an OFDM system using QPSK modulation with the sampling time $T_s = 0.5\mu s$, carrier frequency $f_c = 5$GHz. Let the number of subcarriers $N = 128$, the number of pilots $N_p = 32$, the length of cyclic prefix $N_{cp} = N/8$ and the number of BEM coefficients $Q = 3$. The performance is evaluated by the mean square error (MSE).

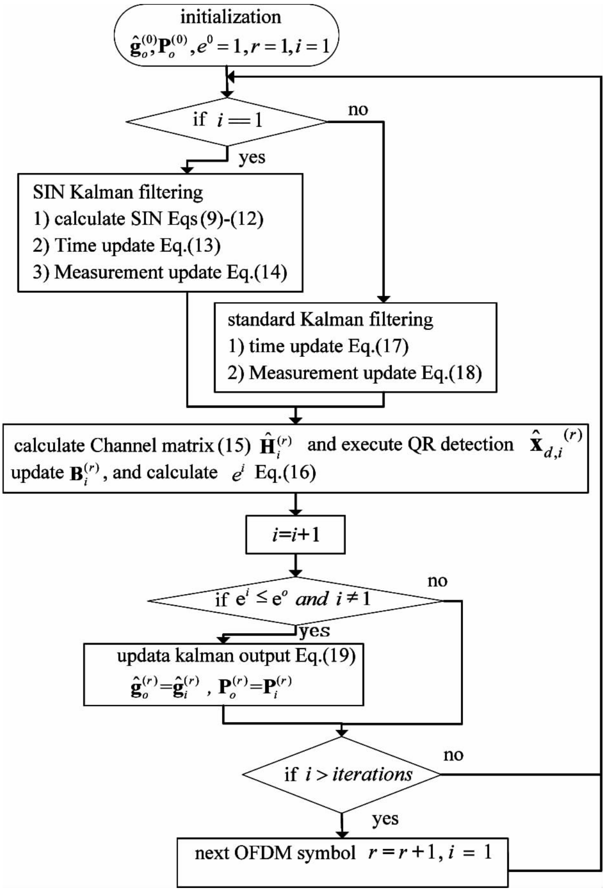

Fig. 1 The flow figure of the scheme

Fig. 2 shows the MSE performance versus SNR of the proposed algorithm for $f_d T = 0.3$, compared to the algorithms in Ref. [7]. The RCRB is the upper limit in theory of Kalman estimation algorithm under the assumption that all data are known at the receiver, given in Ref. [13]. It is observed that, the SIN method improves the performance in the first iteration at low SNR. This is because the ICI is more severe when SNR is low and SIN works well in confronting ICI and error propagation. After several iterations, the determined algorithm provides further improvements by restraining the error propagation, which can be seen from the comparison between the proposed and the SIN without being determined at iteration 10. Due to the SIN filtering and the determined iteration, it can be seen that the proposed algorithm only needs 3 iterations to achieve the equivalent performance of 10 iterations in Ref. [7] and the MSE is close to the RCRB after 10 iterations.

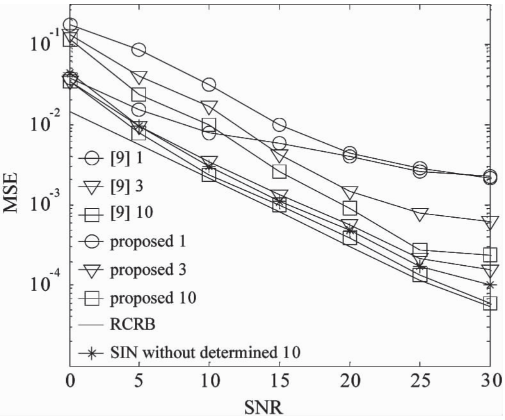

Fig. 2 Comparation of MSE versus SNR with 1, 3 and 10 iterations

Fig. 3 shows the determined iteration of the proposed algorithm which is the last time the optimal results updated in the 10 iterations for different SNR situations. It can be seen that for the low SNR, the iterations needed factually are more than the ones for high SNR, which indicates that due to the high SNR, the error correction by the iterative algorithm is fast but easy to have an exactly opposite effect and therefore the determined step plays an important role to keep the optimal results.

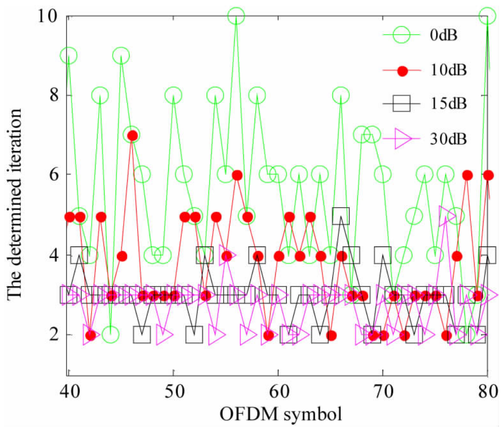

Fig. 3 The determined iteration of the proposed scheme from the 40 th OFDM symbol to the 80th at different SNR

Fig. 4 compares the MSE versus speed ($f_d T$ Doppler spread) between the proposed and the algorithm in Ref. [7] when $SNR = 20$dB. It shows that in the first iteration the proposed algorithm performs better than Ref. [7] only at high speed due to the ICI is not serious at low speed environment and the probability of occurring error propagation is small when $SNR = 20$dB. Actually, SIN is more suitable for large ICI situations which always happen at low SNR or high speed. High speed enlarges ICI terms on $\boldsymbol{H}^{(r)}$ while low SNR leads to error of data estimation. Compared with Ref. [7], the significant improvement occurs after 2 and 5 iterations.

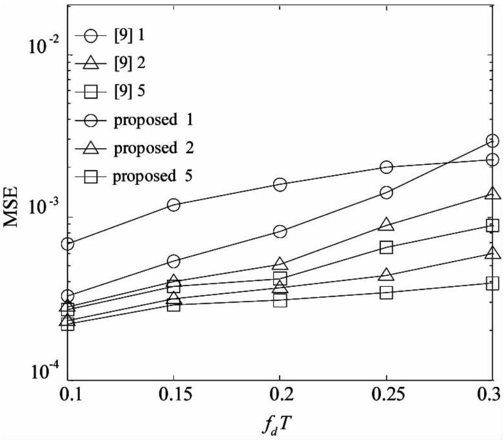

Fig. 4 Comparation of MSE versus Doppler spread after different iterations

Fig. 5 gives the BER performance for $f_dT = 0.2$. Since the error propagation is decreasing in terms of SNR, the algorithms in Ref. [7] catch up and exceed the proposed estimation gradually when SNR increases in the first iteration, which is similar to the conclusion in Fig. 4. However, it can be observed that the proposed algorithm has better performance than the algorithms proposed in Ref. [7] after several iterations.

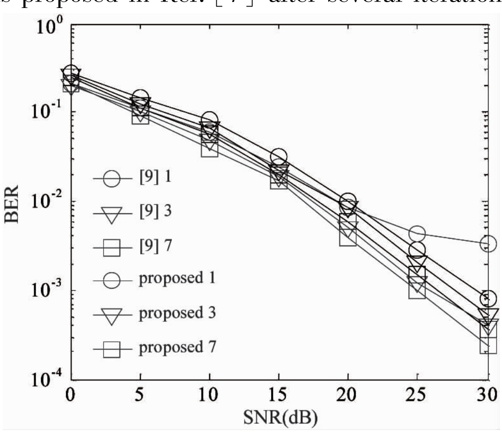

Fig. 5 Comparation of BER versus SNR after different iterations

4 Conclusions

In this paper, a new determined iterative channel estimation scheme for high- mobility OFDM systems is proposed. The frequency-domain Kalman channel estimation is performed with ICI transformed to be part of the noise in the first iteration, which mitigated ICI from the unknown data. In addition, the iterative updating strategy is designed to restrain the error propagation of ICI in the later iterations and therefore the accuracy of the estimation results is improved. The simulation results indicate that the proposed scheme achieves a good result with less iteration in fast time-varying situation especially when SNR is low.

References

[1] Cimini Jr L. Analysis and simulation of a digital mobile channel using orthogonal frequency division multiplexing, *IEEE Trans on Communication*, 1985, COM-33: 665-675

[2] Zhao Y, Haggman S G. Intercarrier interference self-cancellation scheme for OFDM mobile communication systems, *IEEE Trans on Communication*, 2001, 49(7): 1185-1191

[3] Jeon W G, Chang K H, Cho Y S. An equalization technique for orthogonal frequency-division multiplexing systems in time-variant multipath channels, *IEEE Trans on Communication*, 1999, 47(1): 27-32

[4] Molisch A F, Toeltsch M, Vermani S. Iterative methods for cancellation of intercarrier interference in OFDM systems. *IEEE Trans Vehicular Technologh*, 2007, 56(4): 2158-2167

[5] Zhao M, Shi Z, Reed M C. Iterative turbo channel estimation for OFDM systems over rapid dispersive fading channel. *IEEE Trans on Communication*, 2008, 7(7): 1-11

[6] Hijazi H, Ros L. Polynomial estimation of time-varying multipath gains with intercarrier interference mitigation in OFDM systems. *IEEE Trans Vehicular Technologh*, 2008, 57(6): 140-151

[7] Hijazi H, Ros L. Joint data QR-detection and Kalman estimation for OFDM time-varying Rayleigh channel complex gains. *IEEE Trans on Communication*, 2010, 58: 170-178

[8] Tomasin S, Gorokhov A, Yang H, et al. Iterative interference cancella-tion and channel estimation for mobile OFDM, *IEEE Trans on Wireless Communication*, 2005, (1): 238-245

[9] Wang H W, Lin D W, Sang T H. OFDM signal detection in doubly selective channels with blockwise whitening of residual intercarrier interference and noise. *IEEE Vehicular Technology Conference*, 2012, 30(4): 684-694

[10] Song Z K, Yang R Z, Zang L, et al. Multi-model based channel estimation for doubly selective wireless fading channels. In: Proceedings of the IEEE International Conference on Computational Photography. 2012. 515-525

[11] European Telecommunications Standards Institute, European Digital Cellular Telecom-munication System (Phase 2); Radio Transmission and Reception, GSM 05. 05, vers. 4.6.0, Sophia Antipolis, France, July 1993

[12] Hijazi H, Ros L. Polynomial estimation of time-varying multi-path gains with intercarrier interference mitigation in OFDM systems, *IEEE Trans Vehicular Technologh*, 2009, 58(1): 140-151

[13] Hijazi H, Ros L. Analytical analysis of Bayesian Cramer-Rao bound for dynamical Rayleigh channel complex gains estimation in OFDM system. *IEEE Trans on Signal Processing*, 2009, 57(5): 1889-1900

Zhang Jie, born in 1988. He received the M. S. degree in Information and Communication Engineering, Beijing University of Technology in 2013. He received his Bachelor degree in Electronic Information Engineering from Beijing University of Technology in 2010.