A Novel OFDM Channel Estimation Algorithm with ICI Mitigation over Fast Fading Channels

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Abstract. *Orthogonal frequency-division multiplexing (OFDM) is well-known as a high-bit-rate transmission technique, but the Doppler frequency offset due to the high speed movement destroys the orthogonality of the subcarriers resulting in the intercarrier interference (ICI), and degrades the performance of the system at the same time. In this paper a novel OFDM channel estimation algorithm with ICI mitigation based on the ICI self-cancellation scheme is proposed. With this method, a more accurate channel estimation is obtained by comb-type double pilots and then ICI coefficients can be obtained to mitigate the ICI on each subcarrier under the assumption that the channel impulse response (CIR) varies in a linear fashion. The theoretical analysis and simulation results show that the bit error rate (BER) and spectral efficiency performances are improved significantly under high-speed mobility conditions (350 km/h – 500 km/h) in comparison to ZHAO's ICI self-cancellation scheme.*

Keywords

OFDM, channel estimation, high-speed mobility, ICI mitigation.

1. Introduction

Broadband wireless access under high speed-mobility conditions has received much attention, and the high speed railway broadband wireless access is one of the typical scenarios. OFDM, known as an attractive technique for the transmission of the high-bit-rate data, has been investigated as a candidate for the next generation wireless communication [1] for combating the frequency selective fading caused by the multipath channel. But OFDM is very sensitive to the ICI, which may be caused by the carrier frequency offset (CFO), phase noise, timing offset, and the Doppler spread under high-mobility condition [2]. For the ICI induced by the first three impairments, OFDM system can completely compensate or correct it. However, in high-mobility scenarios (such as high speed railway whose velocity reaches 350 km/h – 500 km/h), the channel fluctuates during communication due to the Doppler spread induced by the mobility. Since the Doppler spread or shift is random, we can only mitigate its impact but not cancel it completely. To deal with this problem, several different ICI mitigation techniques have been developed currently including time-domain windowing [3], frequency equalization [4], ICI self-cancellation [5] and Doppler diversity [6]. Furthermore, GE and SUN [7] make use of all phase spectrum analysis technique to reduce the impact of the side lobe on each subcarrier in frequency domain but the spectral efficiency is lower. LIU et al. [8] propose a scheme based on fractional basis expansion model (BEM) for the estimation of doubly selective channel parameters, which can reduce the complexity of channel estimation.

A classical ICI self-cancellation scheme has been proposed by ZHAO [5]. The main idea of the scheme is based on the principle that the difference of ICI between adjacent subcarriers is small, so that ICI can be “self-cancelled” with each other by modulating one data symbol onto the next subcarrier with the weighting coefficient “-1”. This scheme is simple and has sufficient robustness to frequency offset; however, it only has a spectral efficiency of 50 %, which can not satisfy the requirement for high speed broadband wireless access in modern communication.

In this paper, a novel OFDM channel estimation algorithm with ICI mitigation based on the comb-type double pilots is proposed. The method first estimates the accurate channel state information (CSI) with less ICI and builds the time-varying CIR matrix through linear interpolation with two consecutive OFDM symbols. Meanwhile, the received symbols are detected less reliably. Then after the transformation of matrix we can get the ICI coefficients matrix. Finally ICI can be mitigated from the received signal combining the ICI coefficient matrix with the detected signals. The theoretical analysis and simulation results show that the new scheme can reduce the effects of Doppler spread and increase the spectral efficiency significantly at the same time.

This paper is organized as follows. In section 2, we first introduce the structure of the OFDM system and the time-varying channel models. Section 3 discusses and analyzes the principle of ZHAO's self-cancellation scheme [5]. Then we introduce our proposed scheme in section 4. Section 5 depicts the performances of these schemes through simulations on the basis of BER and the spectral efficiency. Finally, we draw the conclusion in section 6.

2. OFDM System and Channel Models

2.1 OFDM System Model

Assuming that there are N subcarriers in an OFDM symbol, $X[m]$ is the complex-valued transmitted data on the mth subcarrier. N_p comb-pilots are inserted in the OFDM symbol for channel estimation, and the spacing between two adjacent pilots is $\Delta P = N/N_P$. For simplicity, the virtual subcarriers and DC tone are ignored. After N-point IFFT, the discrete-time transmitted signal $x(n)$ can be expressed as

$$x(n) = \frac{1}{N}\sum_{m=0}^{N-1} X[m]exp\left(j2\pi\frac{mn}{N}\right), \quad 0 \le n \le N-1 \qquad (1)$$

where $x(n)$ denotes the nth time sample in the OFDM symbol. The cyclic prefix (CP) is added as a guard interval at the beginning of each OFDM symbol to eliminate ISI, and its length is longer than the maximum delay of the channel. Then the OFDM symbol is transmitted through a time-varying multipath fading channel.

2.2 Time-Varying Channel Model in High-Mobility Scenarios

The Doppler frequency offset under high-mobility conditions will make the channel vary fast with time. While the high speed train is moving at a constant velocity making an angle with the direction of wave motion, the Doppler frequency offset of the direct wave is given by

$$f_d = \frac{\nu}{\lambda}cos\alpha = \frac{\nu}{c}\cdot f_c \cdot cos\alpha = f_{max}\cdot cos\alpha \qquad (2)$$

where λ denotes the wavelength of the carrier signal, c is the speed of light, f_c is the transmission center carrier frequency and $f_{max} = \nu/\lambda$ means the maximum Doppler frequency.

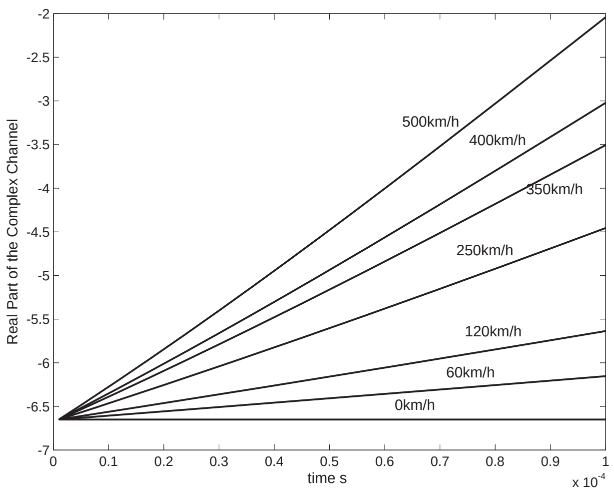

Fig. 1. The channel time-varying characteristics with different speeds.

Fig. 1 shows the time-varying characteristics of the channel with different velocities, from which we can see that the tap of the channel varies more severely when the speed is 500 km/h. Since the high speed railway is usually built in the open, the radio channel has its own characteristics different from the urban channel model in which the scatters are dense and uniformly distributed. The transmitted signal travels along a dominant line-of-sight or direct path called Rician path, and the other muiltpath power spectrum can be described by classical Jakes model [9].

Here, we define the normalized Doppler frequency offset to describe the time-varying characteristics of the channel in OFDM system, which can be denoted as $f_N = T_{sys}\cdot f_d$, where T_{sys} is the duration of the OFDM symbol defined by $T_{sys} = NT_s$ and T_s is the sampling interval. If $f_N \le 0.1$, it is considered that each tap of the channel varies in a linear fashion with time during a block period [8].

In the time-varying multipath channel model, when the transmitted signal $x(n)$ passes through the channel $h(n,l)$, the received signal can be represented as

$$\begin{aligned} y(n) &= h(n,l) * x(n) + w(n) \\ &= \sum_{l=0}^{L-1} h(n,l)x(n-l) + w(n) \end{aligned} \qquad (3)$$

where $*$ denotes the convolution, L is the number of discrete multipaths, $h(n,l)$ represents the time-varying complex gain of the l^{th} path at the n^{th} sample instant, and $w(n)$ is the additive white Gaussian noise (AWGN) with variance σ^2. At the receiver side, perfect synchronization is assumed. After removing the CP and taking N-point FFT, the demodulated signal on the mth subcarrier in the frequency domain is [8]

$$\begin{aligned} Y[m] &= \sum_{k=0}^{N-1}\sum_{l=0}^{L-1} X[k]H_l^{m-k}e^{-j2\pi lk/N} + W[m] \\ &= \left(\sum_{l=0}^{L-1} H_l^0 e^{-j2\pi lk/N}\right) X[m] \\ &+ \sum_{\substack{k=0\\k\neq m}}^{N-1}\sum_{l=0}^{L-1} X[k]H_l^{m-k}e^{-j2\pi lk/N} + W[m] \end{aligned} \qquad (4)$$

where $W[m]$ is the FFT of $w(n)$, H_l^{m-k} represents the FFT of the time-varying multipath channel tap l, which indicates the time-varying characteristics and can be expressed as

$$H_l^{m-k} = \frac{1}{N}\sum_{k=0}^{N-1}\sum_{l=0}^{L-1} h(n,l)e^{-j2\pi lk/N}. \qquad (5)$$

The first term in the right-hand side of (4) contains the desired signal and the fading coefficient resulting from the multipath without interference of other subcarriers. The second term is the ICI component on the m^{th} subcarrier.

3. ICI Self-Cancellation Scheme [5]

The main idea of ZHAO's ICI self-cancellation scheme is to modulate one data symbol onto the adjacent subcarrier with the weighting coefficient "-1". At the receiver side, the received signal is linearly combined on the adjacent subcarrier with the corresponding coefficient, so that ICI contained in the received signals can then be further reduced. The time-varying channel can be modeled based on the Doppler frequency offset as follows

$$h(n,l) = \sum_{l=0}^{L-1} a_l exp\Big(j\frac{2\pi}{N} f_{d-l} T_{sys} n\Big)\delta(\tau-\tau_l) \qquad (6)$$

where a_l, f_{d-l}, *and* τ_l denote the time-varying attenuation coefficient, the Doppler frequency offset and the relative transmission delay of the l^{th} discrete path respectively. For the sake of simplicity, only one path is considered, i.e. $L=1$ (the time-varying attenuation coefficient is α), and f_d is the Doppler frequency offset. Substituting (6) into (4), the signal on the mth subcarrier in frequency domain can be written as

$$\begin{aligned} Y[m] &= a\sum_{k=0}^{L-1} X[k] exp\left(j\pi(1-1/N)\left(k+f_d T_{sys}-m\right)\right) \\ &\quad \cdot \frac{sin\left(\pi\left(k+f_d T_{sys}-m\right)\right)}{Nsin\left(\pi\left(k+f_d T_{sys}-m\right)/N\right)} + W'[m] \\ &= aX[m] exp\left(j\pi(1-1/N) f_d T_{sys}\right) \\ &\quad \cdot \frac{sin\left(\pi\left(f_d T_{sys}\right)\right)}{Nsin\left(\pi\left(f_d T_{sys}\right)/N\right)} + a\sum_{\substack{k=0\\k\neq m}}^{N-1} X[k] \\ &\quad \cdot exp\left(j\pi(1-1/N)\left(k+f_d T_{sys}-m\right)\right) \\ &\quad \cdot \frac{sin\left(\pi\left(k+f_d T_{sys}-m\right)\right)}{Nsin\left(\pi\left(k+f_d T_{sys}-m\right)/N\right)} + W'[m] \end{aligned} \qquad (7)$$

where $W[m]$ is also AWGN. The ICI coefficient is

$$\begin{aligned} S[k-m] &= exp\Big(j\pi(1-\frac{1}{N})(k+f_N-m)\Big) \\ &\quad \cdot \frac{sin\Big(\pi(k+f_N-m)\Big)}{Nsin\Big(\pi(k+f_N-m)/N\Big)}. \end{aligned} \qquad (8)$$

Then (7) can be rewritten as

$$Y[m] = X[m]S(0) + \underbrace{\sum_{\substack{k=0\\k\neq m}}^{N-1} X[k]S(k-m) + W'[m]}_{\substack{ICI\\m=0,1,\ldots,N-1}} \qquad (9)$$

In (9), $S(k-m)$ can be seen as the ICI coefficient that the k^{th} subcarrier works on subcarrier m. Fig. 2 shows the amplitude of the ICI coefficient $S(k-m)$ for $m=0, N=64$ and the normalized Doppler frequency offset values $f_N = 0.1$, $f_N = 0.2$, $f_N = 0.3$. It is evident that the ICI coefficient increases with the increasing Doppler frequency offset, but the ICI coefficient values on the adjacent subcarriers can be approximated equivalently[5]. This is the main idea of ZHAO's ICI self-cancellation technique.

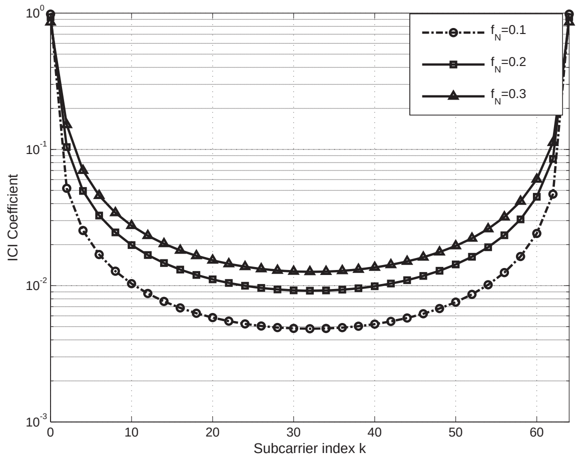

Fig. 2. The amplitude of $S(k-m)$.

ZHAO's self-cancellation scheme in [5] is based on data allocation of $X[1] = -X[0]$, $X[3] = -X[2]$, $\ldots$, $X[N-1] = -X[N-2]$, so the m^{th} received subcarrier can be represented as

$$Y'[m] = \sum_{\substack{k=0\\k=even}}^{N-2} X[k][S(k-m) - S(k+1-m)] + W[m]. \qquad (10)$$

The adjacent $(m+1)^{\text{th}}$ subcarrier can be expressed as

$$\begin{aligned} Y'[m+1] = &\sum_{\substack{k=0\\k=even}}^{N-2} X[k][S(k-m-1) - S(k-m)] \\ &+ W[m+1]. \end{aligned} \qquad (11)$$

At this time, the ICI coefficient of the mth subcarrier is

$$S'(k-m) = S(k-m) - S(k+1-m). \qquad (12)$$

In order to further mitigate ICI, at the receiver side the adjacent subcarriers are combined with the weighting coefficient "-1", which can be derived as

$$\begin{aligned} Y''[m] &= Y'[m] - Y'[m+1] \\ &= \sum_{\substack{k=0\\k=even}}^{N-2} X[m][-S(k-m-1) + 2S(k-m) \\ &\quad - S(k-m+1)] + W[m] - W[m+1]. \end{aligned} \qquad (13)$$

The corresponding ICI coefficient then becomes

$$S''(k-m) = -S(k-m-1) + 2S(k-m) - S(k-m+1) \qquad (14)$$

For a constant subcarrier m and most $k-m$,

$$|S''(k-m)| \ll |S'(k-m)| \ll |S(k-m)| \qquad (15)$$

so the ICI is minimized after procedure (12).

Through "self-cancelling" ICI on the adjacent subcarriers, the performance of the system can be improved greatly with less complexity. However, the spectral efficiency of the scheme is reduced by half due to the repetition symbols in frequency domain. Furthermore, although there is no need for channel estimation due to the differential modulation, high-order modulation such as QAM modulation can not be employed. So this scheme can not realize the effective transmission in modern communication and thus restricts its application in reality.

4. The Novel OFDM Channel Estimation Algorithm with ICI Mitigation

When the OFDM symbol is passing through the time-varying channel, (4) can be simply written as

$$\begin{aligned} Y[m] &= H[m]X[m] + ICI_m + W[m] \\ & m = 0, 1, \ldots, N-1 \end{aligned} \tag{16}$$

where ICI_m and $H[m]$ represent the ICI and the channel frequency response (CFR) resulting from multipath on the m^{th} subcarrier respectively. From (13), if $X[m]$ is to be detected more accurately, both $H[m]$ and ICI_m are to be estimated.

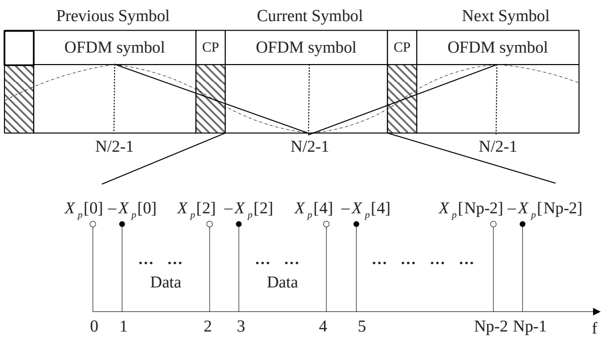

Fig. 3. The structure of the OFDM symbol in the proposed scheme.

Fig. 3 shows the structure of the OFDM symbol in our proposed method. The pilots clusters are inserted in the OFDM symbol like the comb-pilots, and each cluster contains two pilots satisfying $X_p[m_p+1] = -X_p[m_p]$, m_p=0, 2,...,N_p-2, where N_p is the number of the pilots by which the $H[m]$ with less ICI in (13) can be obtained at the receiver side. In Fig. 3, the dashed line in each OFDM symbol is the variation of the channel tap and the solid line is corresponding to the approximation of the tap in linear fashion.

Fig. 4 shows the system model at the receiver side. The detection procedure is described in detail as follows.

Step 1: The double pilots channel estimation. The received pilots are used to estimate the channel information exploiting LS algorithm [12] with the ICI self-cancellation, which is expressed as

$$\begin{aligned} \tilde{H}_p[m_p] &= \frac{Y_p[m_p] - Y_p[m_p+1]}{2X_p[m_p]} \\ &= \frac{H_p[m_p]X_p[m_p] + H_p[m_p+1]X_p[m_p]}{2X_p[m_p]} \\ &+ \frac{ICI_{m_p} - ICI_{m_p+1} + W[m_p] - W[m_p+1]}{2X_p[m_p]} \end{aligned} \tag{17}$$

where $Y_p[m_p]$ and $Y_p[m_p+1]$ represent the two adjacent pilots in one pilots cluster, and $X_p[m_p]$ is the pilot at the transmit side. Owing to $ICI_{m_p} \approx ICI_{m_p+1}$, $\tilde{H}_p[m_p]$ is affected by less ICI. Then we can get the CIR for the time-varying multipath channel, as well as the channel estimation value $\tilde{H}[m]$ after the transform domain channel estimation [11], [12].

On one hand, $\tilde{H}[m]$ is used in the coarse zero forcing (ZF) equalization for the received signals. The equalized signal can be expressed as

$$\tilde{X}[m] = Y[m]\tilde{H}[m] / (\tilde{H}[m]\tilde{H}^*[m]) \tag{18}$$

where $(\quad)^*$ denotes conjugate operation. Then the less reliable binary bits from the decision of $\tilde{X}[m]$ are re-modulated, which is represented as $\tilde{X}'[m]$

On the other hand, the CIR is used for the multi-symbols channel estimation, which will be described in detail in Step 2.

Step 2: The consecutive symbols linear channel estimation in time domain. When the normalized Doppler frequency offset $f_N \le 0.1$, the time variations of the tap coefficients, for all L paths, are approximated by straight lines with low slops during a block period. As a result, we can approximate the CIR in an OFDM symbol combining adjacent symbols, the previous and next symbols. As shown in Fig. 3, the time-varying CIR of the present symbol is interpolated in time domain by combination of neighboring symbols. First, the $\tilde{H}_p[m_p]$ is converted to the time domain by taking the IFFT:

$$\begin{aligned} \tilde{h}(n)^{ave} &= \frac{1}{N_p} \sum_{m_p=0}^{N_p-1} \tilde{H}_p[m_p] e^{j\frac{2\pi n m_p}{N_p}}, \\ & 0 \le n \le (N_p - 1). \end{aligned} \tag{19}$$

For the reason that $|\tilde{h}(n)^{ave} - h(n)^{(\frac{N}{2}-1)}|^2$ is minimized for the n^{th} path, which means that the variance between $\tilde{h}(n)^{ave}$ and the CIR at time $(N/2-1)T_s$ is the smallest [13], [14], it is reasonable to approximately represent the multipath tap coefficients at time $t = N/2-1$ by use of $\tilde{h}(n)^{ave}$, i.e.

$$\tilde{h}(n)^{(\frac{N}{2}-1)} \approx \tilde{h}(n)^{ave} \tag{20}$$

Then it is easy to interpolate the CIRs at other time instants in the current OFDM symbol if the consecutive three symbols' CIRs at time $(N/2-1)T_s$ ($\tilde{h}(n-1)^{(\frac{N}{2}-1)}$,$\tilde{h}(n)^{(\frac{N}{2}-1)}$ and $\tilde{h}(n+1)^{(\frac{N}{2}-1)}$) are known. Then we can get the "cyclic" convolution matrix $\mathbf{h}_c$ of the current OFDM symbol, which can be expressed as (19), where $\tilde{h}(n,l)$ denotes the l^{th} estimated tap coefficient at time n.

Step 3: Calculate the channel estimation matrix and the interference coefficient matrix. If the channel is assumed to be time-invariant during a block period, each row of $\mathbf{h}_c$ referred in Step 2 is the right circular shift of a constant row vector $\mathbf{h}_n = [\tilde{h}(0), \tilde{h}(1),...,\tilde{h}(L-1), 0, ...,0]$, hence, the equivalent channel information in frequency domain is

$$\Lambda = \mathbf{F}\,\mathbf{h}_c\,\mathbf{F}^H \tag{21}$$

where $\mathbf{F}$ and $\mathbf{F}^H$ represent the N-point FFT and IFFT matrix with the elements $e^{j2\pi ik/N}$ and $e^{-j2\pi ik/N}$ respectively, all of which are unitary matrix. According to the characteristics of the matrix, Λ will be a diagonal matrix if $\mathbf{h}_c$ is rotate right Toeplitz matrix. Under the condition of $f_N \leq 0.1$, $\mathbf{h}_c$ is no longer a rotate right Toeplitz matrix when each element of the row vector $\mathbf{h}_c$ varies linearly. Then (18) can be expressed as (20).

$$\mathbf{h}_c = \begin{bmatrix} \tilde{h}(0,0) & 0 & \dots & 0 & \tilde{h}(0,L-1) & \dots & \tilde{h}(0,1) \\ \tilde{h}(1,1) & \tilde{h}(1,0) & 0 & \dots & 0 & \dots & \tilde{h}(1,2) \\ \vdots & \ddots & \ddots & \ddots & \ddots & 0 & \vdots \\ \tilde{h}(L-1,L-1) & \ddots & \ddots & \tilde{h}(L-1,0) & \ddots & \ddots & 0 \\ 0 & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ 0 & \dots & 0 & \tilde{h}(N-1,L-1) & \dots & \tilde{h}(N-1,1) & \tilde{h}(N-1,0) \end{bmatrix}_{N\times N} \tag{22}$$

$$\mathbf{A} = \begin{bmatrix} a(0,0) & a(0,1) & \dots & a(0,q) & 0 & 0 & \dots & a(0,N-2) & a(0,N-1) \\ a(1,0) & a(1,1) & \dots & a(1,q) & a(1,q+1) & \dots & \dots & \dots & a(1,N-1) \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ a(q,0) & a(q,1) & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & 0 \\ 0 & a(q+1,1) & \ddots & \ddots & \ddots & \ddots & \ddots & a(N-q,N-2) & a(N-q,N-1) \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \ddots & \vdots \\ a(N-2,0) & \dots & \dots & a(N-2,q) & a(N-2,N-q-1) & a(N-2,N-q) & \dots & a(N-2,N-2) & a(N-2,N-1) \\ a(N-1,0) & a(N-1,1) & \dots & a(N-1,q) & 0 & a(N-1,N-q) & \dots & a(N-1,N-2) & a(N-1,N-1) \end{bmatrix}_{N\times N} \tag{23}$$

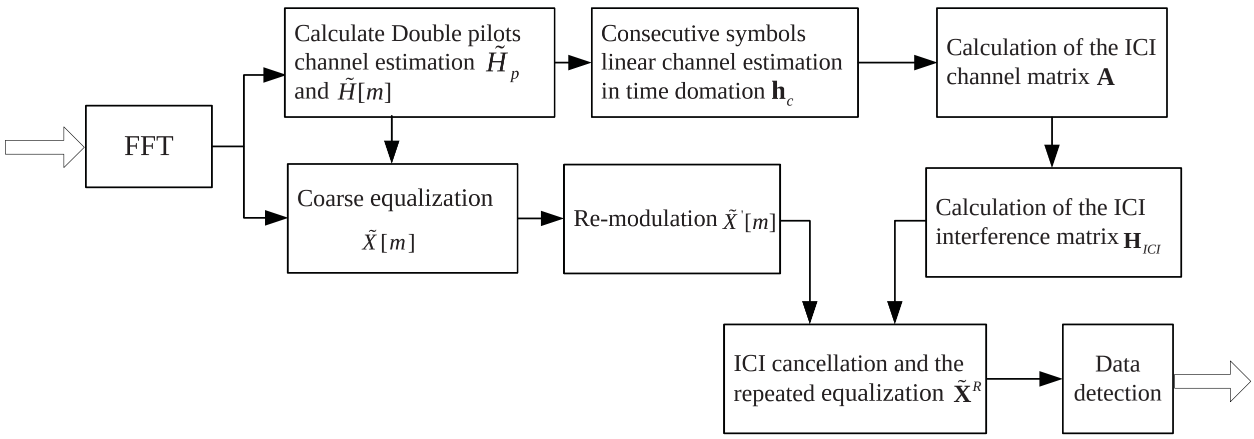

Fig. 4. Block diagram of the novel OFDM channel estimation algorithm with ICI mitigation.

It is evident that $\mathbf{A}$ is the ICI channel matrix, and it is a non-diagonal matrix with cross-terms between subcarriers. $a(i,j)$ denotes the ICI coefficient between adjacent subcarriers. When the channel is time-invariant, $a(i,j)$ is

$$a(i,j) = \begin{cases} H[i], & i=j \\ 0, & otherwise. \end{cases} \tag{24}$$

At this time $\mathbf{A}$ is a diagonal matrix, i.e. $\mathbf{A} = \Lambda$

Then (13) can be expressed in matrix form as follows

$$\mathbf{Y} = \mathbf{AX} + \mathbf{W}. \tag{25}$$

The diagonal elements of $\mathbf{A}$, $H_{diag} = diag\mathbf{A}$ = $[a(0,0)$, $a(1,1)$,...,$a(N-1,N-1)]$ are the fading factors on corresponding subcarriers, while $\mathbf{H}_{ICI} = \mathbf{A} - \mathbf{H}_{diag}$ is the ICI interference coefficient matrix.

Step 4: ICI mitigation and re-equalization. From the analysis above, we know that under the assumption of linear time-varying channel, the received signal detection is converted to the solution of equation $\mathbf{Y} = \mathbf{AX}$. Under ordinary conditions, we do not know the value of q in the frequency response matrix $\mathbf{A}$, so it is impossible to solve the equation directly. Since ICI on a subcarrier suffers from the adjacent subcarriers more and it gradually decreases apart from this subcarrier [10], most energy is concentrated in the neighborhood of the diagonal line in (20). [1] and [10] have employed reduced channel models to solve the equation, so they can only coarsely calculate the ICI on each subcarrier.

Here, we combine the re-modulation $\tilde{\mathbf{X}}'$ in Step 1 with the ICI interference coefficient matrix $\mathbf{H}_{ICI}$, and the result is the ICI interference matrix. Then the interference can be mitigated from the received signals, the procedure can be represented in matrix form as

$$\mathbf{Y}_{offICI} = \mathbf{Y} - \mathbf{H}_{ICI}\tilde{\mathbf{X}}' \tag{26}$$

where $\tilde{\mathbf{X}}'$ is the re-modulation signal vector after the first coarse decision and $\mathbf{Y}_{offICI}$ is the received signal without ICI interference.

Step 5: The second time data detection. The received signal $\mathbf{Y}_{offICI}$ without ICI interference is detected again, and then the more reliable decision signal $\tilde{\mathbf{X}}^R$ can be expressed as

$$\tilde{\mathbf{X}}^R = \mathbf{Y}_{offICI}\mathbf{H}_{diag}^H(\mathbf{H}_{diag}\mathbf{H}_{diag}^H)^{-1}. \tag{27}$$

5. Simulation Results

In this section, we compare the performance of the proposed system with the conventional OFDM and ZHAO's ICI self-cancellation scheme by the Monte Carlo simulation. The parameters used in the simulation are shown in Tab. 1. The virtual subcarriers and the DC tone are ignored and the center carrier frequency is set to 2.4 GHz. For the reason that our application scenario is the high speed railway access, the COST 207 rural area channel model [15] is exploited as the high speed railway wireless channel model. In the model, the first path is a line-of-sigh path, i.e. the strong Rician path. And the spectrum of the other paths is still the classical Doppler spectrum. The tapped delay line (TDL) model is used in the simulation. The path gains and the path delays are shown in Tab. 2.

The simulation mainly focuses on the performance analysis of the high speed railway broadband access with different velocities. Tab. 3 shows the normalized Doppler frequency offset in terms of velocity, from which we can see that the normalized Doppler frequency offset is less than 0.1 even when the speed reaches 500 km/h, so the channel still varies in a linear fashion with time.

Next we will show the performance analysis of the proposed channel estimation algorithm, the channel estimation based on the transform domain and ZHAO's ICI self-cancellation scheme. The analysis is mainly about the performance of combating the Doppler frequency and the spectral efficiency. The modulation scheme is DBPSK for ZHAO's scheme without channel estimation, and the spectral efficiency is only 50 %, which is the same as case **I** in Tab. 1.

Parameters	Values
System bandwidth	5MHz
FFT/IFFT points	256
CP Length	32
OFDM symbol period	51.2μs
Pilot/data Modulation	BPSK
Pilots number N_p	**I**: 64× 2=128 **II**: 32× 2=64 **III**:16× 2=32
Pilots interval ΔP	**I**: 256/64 =4 **II**: 256/32 = 8 **III**: 256/16 =16
Spectral efficiency	**I**: 50% **II**: 75% **III**: 87.5%

Tab. 1. OFDM system simulation parameters.

Path	Path delay (μs)	Path gain (dB)	Doppler spectrum
1	0	0	Rice
2	0.4	-2	class
3	0.4	-10	class
4	0.6	-20	class

Tab. 2. COST 207 rural area channel model.

Velocity(km/h)	120	250	350	400	500
$f_N = T_{sys} \cdot f_{max}$	0.014	0.028	0.04	0.046	0.057

Tab. 3. The normalized Doppler frequency offsets.

Fig. 5 and Fig. 6 show the BER performances in terms of the average signal-to-noise ratio (SNR) and velocity for the three different schemes. Fig. 5 is mainly about the medium and low speed (120 km/h and 250 km/h) while Fig. 6 is about the high speed. We can see that the proposed scheme outperforms the other two schemes whether at medium and low speed or high speed. In Fig. 5, the three schemes have the similar performance when the SNR is less than 20 dB, but after the SNR exceeds 25 dB, ZHAO's scheme meets an "error floor" rapidly. Furthermore, when the speed is 120 km/h, the conventional DFT channel estimation with spectral efficiency 50 % performs even better than ZHAO's scheme, while the proposed algorithm performances similarly to the static channel (the speed is 0 km/h). When the speed is 250 km/h, our proposed algorithm also performs superior over the other two schemes. In Fig. 6, the conventional DFT channel estimation, ZHAO's scheme and our proposed algorithm reach error floor when SNR is 20 dB, 25 dB and 35 dB respectively. The performance of our algorithm is less than 10^{-4} when the speed is 350 km/h and much better than the other two schemes when the speed is 500 km/h.

Fig. 7 shows the BER performances in terms of SNR and spectral efficiency for the three different schemes, from which we can see that the performances of these three schemes decrease as the spectral efficiency increases. Here, ZHAO's scheme has a steady performance at medium and low speeds, and the "error floor" is around $10^{-3.7}$; while the proposed algorithm with the spectral efficiency of 75 % and 87.5 % at high speed has a better performance than ZHAO's scheme with the spectral efficiency of 50 % at medium and low speeds, though the performance of the former is a little worse than the latter when the SNR is less than 20 dB. Moreover, the proposed algorithm with the spectral efficiency of 87.5 % at the speed of 350 km/h or with the spectral efficiency of 75 % at the speed of 500 km/h has the similar performance with the conventional DFT channel estimation scheme with spectral efficiency of 50 % at the speed of 120 km/h. Then, we can conclude that the performance of the proposed algorithm is superior to that of other two methods in respect to mobility and spectral efficiency.

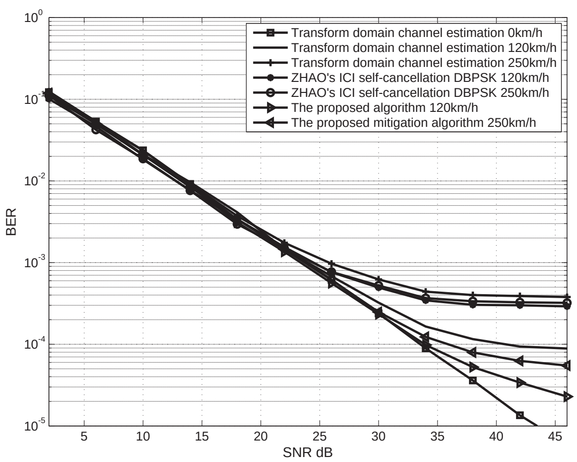

Fig. 5. BER performance comparison of three schemes with medium and low speeds.

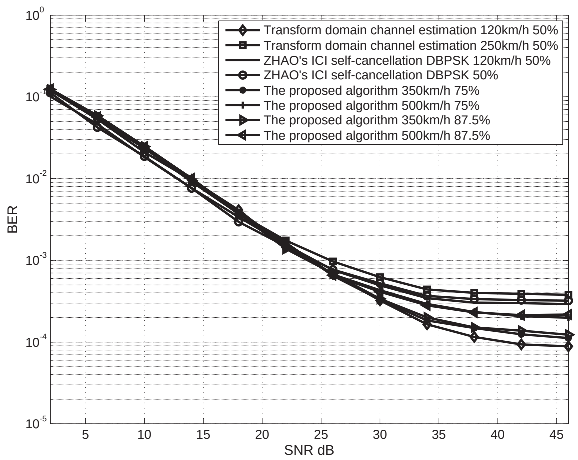

Fig. 7. BER performance comparison of three schemes with different spectral efficiencies.

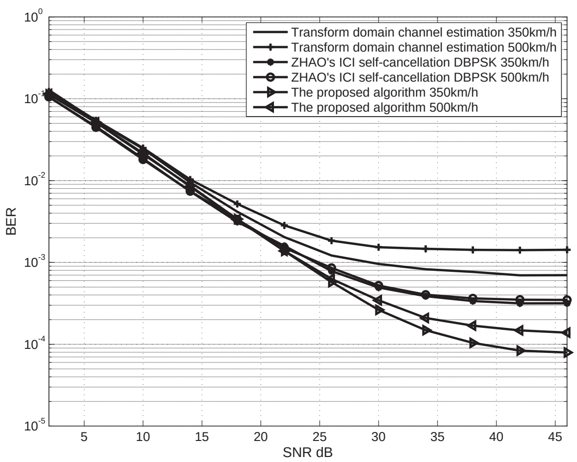

Fig. 6. BER performance comparison of three schemes with high speeds.

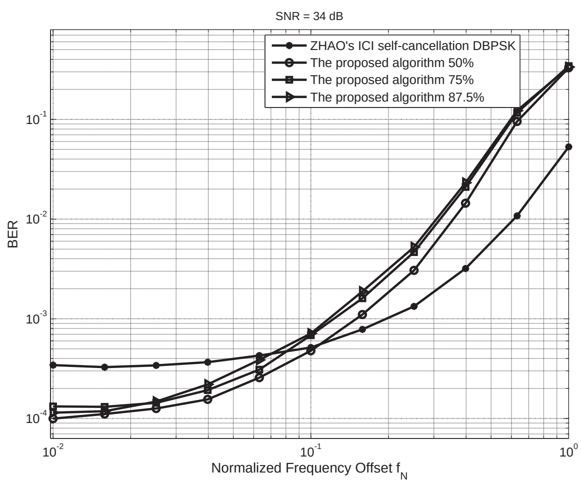

Fig. 8. BER performance comparison of three schemes with different normalized Doppler frequency offsets.

Fig. 8 shows the BER performance in terms of the normalized Doppler frequency offset and spectral efficiency for these three different schemes when the SNR is 34 dB. We can see that when the normalized Doppler frequency offset $f_N \approx 0.1$ corresponding to a vehicle speed of 781 km/h, which is the upper bound that the multipath tap varies linearly in a block period. Under this condition, the proposed algorithm outperforms ZHAO's scheme greatly. When $f_N > 0.1$, the assumption that the CIR varies in a linear fashion during a block period no longer holds, so the proposed algorithm which utilizes the consecutive symbols linear interpolation to estimate the channel can not track the actual channel variance and the ICI matrix is not accurate any more. This results in the performance degradation. Nevertheless, since the speed corresponding to $f_N \approx 0.1$ has reached the ceiling speed for the land high speed movement, our scheme can be employed in practice more easily.

6. Conclusion

Under the condition of high speed movement, ICI caused by the Doppler frequency offset degrades the performance of OFDM system significantly. This paper analyzes ZHAO's ICI self-cancellation scheme and proposes a novel OFDM channel estimation algorithm with ICI mitigation. By the assumption that the channel varies in a linear fashion in a block period, the comb-type double pilots are used to estimate the CFR with less ICI and the ICI matrix is obtained by exploiting the adjacent OFDM symbols. After iterative interference cancellation, the system performance in high-mobility scenarios can be improved and the spectral efficiency is increased at the same time.

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