Signal detection through circular convolution reconstruction for OFDM system in fast varying channel

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Abstract: A signal detection algorithm is proposed for the orthogonal frequency division multiplexing (OFDM) system in the presence of fast time-varying channel. The channel is represented by a piece-wise linear variant model with normalized Doppler frequency of less than 0.2. The channel parameters are extracted through ISI/ICI (inter-symbol interference/inter-carrier interference) cancellation and circular convolution reconstruction. Meanwhile, an improved OFDM symbol detection algorithm is also proposed based on circular convolution reconstruction. The channel state information in the OFDM symbol duration can be obtained accurately from the adjacent two block pilots in a linear model. The simulation results show that the proposed method can not only track the channel variation, but also promise better performance gain in the OFDM symbol detection. Furthermore, the bit error ratio (BER) is close to the performance with the perfect channel state information.

Key words: OFDM; inter-symbol interference; inter-carrier interference; circular convolution reconstruction

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1. Introduction

In a terrestrial high-speed mobile environment, the fast time-varying characteristics and bad multipath effects will cause inter-carrier interference (ICI) and inter-symbol interference (ISI) that worsen the system performance in the traditional orthogonal frequency division multiplexing (OFDM) systems [1-2]. Therefore, effective detection of the OFDM signal is one of the core technologies.

Instead of cyclic prefix, block guidance in poor multipath environment is appended as guard and has been applied to the time-domain synchronous OFDM (TDS-OFDM) system, which is a candidate for the China Digital High-Definition Television (HDTV) [3-4]. The guide data can not only prevent the ISI from interfering with the adjacent OFDM symbols, but can also be used for both channel estimation and synchronization purpose [5-6]. In the time-varying environment, when the normalized Doppler frequency is less than 20% in the OFDM data block duration, the channel impulse response (CIR) can be modeled as a linear change process [2].

This model is used directly to describe the channel variation [7], but due to residual interference, the accuracy of the channel estimation is still limited. An algorithm [8] optimizes the method in [7] to eliminate the residual interference to improve the channel estimation. Ref. [9] uses either cycle prefix or two consecutive OFDM blocks to estimate CIR and eliminate ICI. However, the former one is more suitable for the smaller Doppler expansion, while the latter can lead to larger processing time delay.

In high-speed mobile environments, Doppler effects will destroy the orthogonality of the subcarrier and lead to ICI that degrades the performance. The Doppler frequency shift can be estimated by some methods, such as the channel statistic characters or the maximum likelihood (ML) principle based on the training sequence, to compensate and restore the orthogonality of the subcarrier [10-11]. Meanwhile, the autocorrelation of the cyclic prefix can be used to estimate the Doppler shift in OFDM systems [9]. However, the ways of improving synchronization are generally more suitable for slow fading channels [12-13]. An equalizer can be used to compensate for Doppler shift and reduce ICI impact [14]. If the time-varying channel characteristics are known, the ICI cancellation technology through signal processing can be adopted to mitigate ICI, but it requires accurate channel state information [15]. Through poly-

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nomial cancellation coding (PCC) in the frequency domain, a classic ICI interference self-cancellation technique can mitigate the ICI between adjacent sub-carriers to improve system performance [16]. This method is effective and robust, but the spectrum utilization ratio is only 50%. Doppler effects are usually an unfavorable factor and some measures are made to compensate or diminish the Doppler frequency shift (spread) impact in traditional systems [17]. However, Doppler spread can also be considered as a diversity source in the fast fading environment. Collection and use of the energy component in the different Doppler frequency shift direction will improve the performance of the system [18-22]. To date, the existing literature rarely considers joint ICI and ISI suppression for OFDM signal detection in the high-speed mobile environment.

In a fast time-varying environment with high multipath delay, assuming that the CIR satisfies a linear change in the OFDM block period [2], this paper proposes a channel estimation algorithm through iterative ISI and ICI cancellation and circular convolution reconstruction. On this basis, the channel state information over the OFDM symbol duration can be calculated by the adjacent two pilots, and the circular convolution can be reconstructed to complete OFDM symbol detection. Compared to the traditional signal detection method, in which only single pilot is used, the proposed method can improve OFDM symbol detection performance. Simulation results show that the proposed algorithm is able to obtain the CIR and complete OFDM symbol detection effectively. Using the proposed scheme, the bit error ratio (BER) is close to the performance with the ideal channel state information.

2. System model

Fig. 1 depicts the OFDM system, in which the block-type pilot replaces the cyclic prefix as guidance [3]. After phase-shift key (PSK) modulation, the data stream is modulated by N -point inverse fast Fourier transform (IFFT). In the transmitter, the block-type pilot is inserted between the consecutive OFDM data union to construct the transmitted data. For the i th OFDM symbol $\mathbf{S}^i = [S_0^i, S_1^i, \dots, S_{N-1}^i]^T$. After IFFT, the time-domain sequence $\mathbf{s}^i = (s_0^i, s_1^i, s_2^i, \dots, s_{N-1}^i)^T$ is calculated as

$$s_n^i = \frac{1}{\sqrt{N}} \sum_{k=0}^{N-1} S_k^i \exp\left(\frac{j2\pi kn}{N}\right), \quad 0 \leq n < N. \quad (1)$$

The i th time domain pilot is denoted by $\mathbf{p}^i = (p_0^i, p_1^i, p_2^i, \dots, p_{N_p-1}^i)^T$, where N_p is the length of the time domain pilot.

(a) The transmitted data

(b) The received data

Fig. 1 Sending and receiving signals in multipath channel

According to Fig. 1(a), the extended OFDM data block $\mathbf{x}^i = (x_0^i, x_1^i, x_2^i, \dots, x_{N_2-1}^i)^T$, where $N_2 = N + N_p$, is constituted with the OFDM data unit and its tail guidance:

$$\mathbf{x}_n^i = \begin{cases} s_n^i, & 0 \leq n < N, \\ p_{n-N}^i, & N \leq n < N_2. \end{cases} \quad (2)$$

The extended OFDM symbol block is defined by $\bar{\mathbf{x}}^i = (\bar{x}_0^i, \bar{x}_1^i, \bar{x}_2^i, \dots, \bar{x}_{N_2-1}^i)^T$, which satisfies

$$\bar{x}_n^i = \begin{cases} p_n^i, & -N_p \leq n < 0, \\ x_n^i, & 0 \leq n < N_2. \end{cases} \quad (3)$$

If the appended pilots are identical, $\mathbf{p}^i$ can be used as the cyclic prefix of $\mathbf{x}^i$. To facilitate the analysis, the identical guidance is assumed in this paper.

3. Channel estimation under time-varying channel

3.1. Channel response decomposition based on received pilot

If the normalized Doppler frequency is less than 20% in an extended OFDM symbol block ($\bar{\mathbf{x}}^i$) duration, the CIR can be approximated to change in a linear model. The channel is modeled as an $(L+1)$ -order FIR filter and the pilot duration is greater than the channel multipath delay.

Let μ_l denote the slope of the l th channel tap in $\bar{\mathbf{x}}^i$ duration, and $h_{n,l}$ denote the tap coefficient at the l th channel tap and time instant n . For convenience, we approximate $h_{N_p/2,l}$ to the estimate of h_l^{ave} , the average channel gain in the pilot $\mathbf{p}^i$ duration. Considering linearization around h_l^{ave} in the $\bar{\mathbf{x}}^i$ duration, $h_{n,l}$ can be approximated as follows:

$$h_{n,l} = h_l^{\text{ave}} + \mu_l \left(n - \frac{N_p}{2} \right), \quad 0 \leq l \leq L, 0 \leq n < N_p. \quad (4)$$

The received time-domain pilot response $\mathbf{r}_p^i = (r_p^i(0), r_p^i(1), r_p^i(2), \dots, r_p^i(N_p - 1))^T$ can be calculated as

$$r_p^i(n) = \begin{cases} \sum_{l=0}^n h_{n,l} p_{\langle n-l \rangle N_p}^i + \sum_{l=n+1}^L h_{n,l} s_{\langle n-l \rangle N}^{i-1} + w_n^i, & 0 \leq n < L, \\ \sum_{l=0}^L h_{n,l} p_{\langle n-l \rangle N_p}^i + w_n^i, & L \leq n < N_p, \end{cases} \quad (5)$$

where w_n^i is Gaussian white noise.

From (5), the first L samples of $\mathbf{r}_p^i$ will be influenced by the adjacent OFDM data. $\mathbf{r}_p^i$ can be decomposed into three parts:

$$\mathbf{r}_p^i = \mathbf{r}_p^{1,i} + \mathbf{r}_p^{2,i} + \mathbf{r}_p^{3,i} + \mathbf{w}^i, \quad (6)$$

in which $\mathbf{w}^i = (w_0^i, w_1^i, w_2^i, \dots, w_{N_p-1}^i)^T$; $\mathbf{r}_p^{1,i}$ represents the expected pilot response, which is the result of circular convolution between the pilot and channel impulse response:

$$\mathbf{r}_p^{1,i} = \mathbf{H}_{\text{CIR}}^i \times \mathbf{p}^i, \quad (7)$$

where $\mathbf{H}_{\text{CIR}}^i$ is an N_p -order square coefficient matrix of CIR: $\mathbf{H}_{\text{CIR}}^i = [H_{\text{CIR}}^i(k, n)]_{N_p \times N_p}$, and

$$H_{\text{CIR}}^i(k, n) = h_{n, \langle n-k \rangle N_p}, \quad 0 \leq k, n < N_p; \quad (8)$$

$\mathbf{r}_p^{2,i}$ represents the ISI from the OFDM data unit $\mathbf{s}^{i-1}$:

$$\mathbf{r}_p^{2,i} = \mathbf{A}_{\text{ISI}}^i \times \mathbf{s}^{i-1}, \quad (9)$$

where $\mathbf{A}_{\text{ISI}}^i = [A_{\text{ISI}}^i(k, n)]_{N_p \times N}$ is a coefficient matrix of ISI, and

$$A_{\text{ISI}}^i(k, n) = h_{k, k-n+N}, \quad 0 \leq k < N_p, 0 \leq n < N; \quad (10)$$

$\mathbf{r}_p^{3,i}$ is the introduced additional ISI when reconstructing the circular convolution result:

$$\mathbf{r}_p^{3,i} = -\mathbf{B}_{\text{ISI}}^i \times \mathbf{p}^i, \quad (11)$$

where $\mathbf{B}_{\text{ISI}}^i = [B_{\text{ISI}}^i(k, n)]_{N_p \times N_p}$ is an N_p -order square coefficient matrix of ISI, and

$$B_{\text{ISI}}^i(k, n) = h_{k, k-n+N_p}, \quad 0 \leq k, n < N_p. \quad (12)$$

From (10) and (12), $\mathbf{A}_{\text{ISI}}^i$ and $\mathbf{B}_{\text{ISI}}^i$ are upper triangular matrices and sparse matrices, respectively. Substituting (4) into (8), (10), and (12), the corresponding chan-

nel coefficient matrix can be composed of two parts, e.g.,

$$\mathbf{H}_{\text{CIR}}^i = \mathbf{H}_{\text{CIR}}^{\text{ave},i} + \mathbf{M} \times \mathbf{H}_{\text{CIR}}^{\text{var},i}, \quad (13)$$

where $\mathbf{H}_{\text{CIR}}^{\text{ave},i} = [H_{\text{CIR}}^{\text{ave},i}(k, n)]_{N_p \times N_p}$ is dependent on the average channel gain h_l^{ave} , $\mathbf{H}_{\text{CIR}}^{\text{var},i} = [H_{\text{CIR}}^{\text{var},i}(k, n)]_{N_p \times N_p}$ is determined by channel tap slope μ_l , and $\mathbf{M}$ is an N_p -order square coefficient matrix:

$$\left. \begin{aligned} H_{\text{CIR}}^{\text{ave},i}(k, n) &= h_{\langle k-n \rangle N_p}^{\text{ave}}, \quad 0 \leq k, n < N_p, \\ H_{\text{CIR}}^{\text{var},i}(k, n) &= \mu_{\langle k-n \rangle N_p}, \quad 0 \leq k, n < N_p, \\ M(k, k) &= k - N_p/2, \quad 0 \leq k, n < N_p. \end{aligned} \right\} \quad (14)$$

Similarly,

$$\left. \begin{aligned} \mathbf{A}_{\text{ISI}}^i &= \mathbf{A}_{\text{ISI}}^{\text{ave},i} + \mathbf{M} \times \mathbf{A}_{\text{ISI}}^{\text{var},i}, \\ \mathbf{B}_{\text{ISI}}^i &= \mathbf{B}_{\text{ISI}}^{\text{ave},i} + \mathbf{M} \times \mathbf{B}_{\text{ISI}}^{\text{var},i}, \end{aligned} \right\} \quad (15)$$

where

$$\left. \begin{aligned} A_{\text{ISI}}^{\text{ave},i}(k, n) &= h_{k-n+N}^{\text{ave}}, \quad 0 \leq k, n < N_p, \\ A_{\text{ISI}}^{\text{var},i}(k, n) &= \mu_{k-n+N}, \quad 0 \leq k, n < N_p, \end{aligned} \right\} \quad (16)$$

and

$$\left. \begin{aligned} B_{\text{ISI}}^{\text{ave},i}(k, n) &= h_{k-n+N_p}^{\text{ave}}, \quad 0 \leq k, n < N_p, \\ B_{\text{ISI}}^{\text{var},i}(k, n) &= \mu_{k-n+N_p}, \quad 0 \leq k, n < N_p. \end{aligned} \right\} \quad (17)$$

Taking N_p -point FFT to $\mathbf{r}_p^i$, the frequency-domain response $\mathbf{R}_p^i$ can be calculated as

$$\mathbf{R}_p^i = \mathbf{R}_p^{1,i} + \mathbf{R}_p^{2,i} + \mathbf{R}_p^{3,i} + \mathbf{W}^i, \quad (18)$$

where $\mathbf{W}^i$ is the Fourier transform of $\mathbf{w}^i$, and

$$\left. \begin{aligned} \mathbf{R}_p^{1,i} &= \mathbf{F}_{N_p} \times \mathbf{r}_p^{1,i} = \mathbf{G}^{1,i} \times \mathbf{P}^i, \\ \mathbf{R}_p^{2,i} &= \mathbf{F}_{N_p} \times \mathbf{r}_p^{2,i} = \mathbf{G}^{2,i} \times \mathbf{S}^{i-1}, \\ \mathbf{R}_p^{3,i} &= \mathbf{F}_{N_p} \times \mathbf{r}_p^{3,i} = \mathbf{G}^{3,i} \times \mathbf{P}^i, \end{aligned} \right\} \quad (19)$$

where $\mathbf{F}_{N_p}$ represents the FFT matrix of N_p point; $\mathbf{P}^i$ is the frequency-domain response of $\mathbf{p}^i$; $\mathbf{G}^{1,i}$, $\mathbf{G}^{2,i}$, and $\mathbf{G}^{3,i}$ are coefficient matrices.

The coefficient matrix satisfies

$$\left. \begin{aligned} \mathbf{G}^{1,i} &= \mathbf{F}_{N_p} \mathbf{H}_{\text{CIR}}^{\text{ave},i} \mathbf{F}_{N_p}^H + \mathbf{F}_{N_p} (\mathbf{M} \times \mathbf{H}_{\text{CIR}}^{\text{var},i}) \mathbf{F}_{N_p}^H = \\ &\mathbf{G}_1^{1,i} + \mathbf{G}_2^{1,i}, \end{aligned} \right\} \quad (20)$$

where $\mathbf{G}_1^{1,i} = [G_1^{1,i}(k, k)]_{N_p \times N_p}$ is an N_p -order diagonal matrix, and $\mathbf{G}_2^{1,i} = [G_2^{1,i}(k, m)]_{N_p \times N_p}$ is an N_p -order square

matrix:

$$\left. \begin{aligned} G_1^{1,i}(k,k) &= \sum_{l=0}^L h_l^{\text{ave}} \exp(-j \frac{2\pi}{N_p} kl), \quad 0 \leq k < N_p, \\ G_2^{1,i}(k,m) &= \begin{cases} \sum_{l=0}^L \frac{\mu_l}{N_p} \exp(-j \frac{2\pi}{N_p} ml) \sum_{n=0}^{N_p-1} \left(n - \frac{N_p}{2} \right) \times \\ \exp(j \frac{2\pi}{N_p} (m-k)n), \\ 0 \leq k, m < N_p, k \neq m, \\ 0, \quad k = m. \end{cases} \end{aligned} \right\} \quad (21)$$

Therefore, $\mathbf{R}_p^{1,i}$ can be decomposed into two parts:

$$\mathbf{R}_p^{1,i} = \mathbf{R}_{p,1}^{1,i} + \mathbf{R}_{p,2}^{1,i}, \quad (22)$$

where

$$\left. \begin{aligned} \mathbf{R}_{p,1}^{1,i} &= \mathbf{G}_1^{1,i} \times \mathbf{P}^i, \\ \mathbf{R}_{p,2}^{1,i} &= \mathbf{G}_2^{1,i} \times \mathbf{P}^i. \end{aligned} \right\} \quad (23)$$

Since $\mathbf{G}_1^{1,i}$ is a diagonal matrix, if $\mathbf{R}_{p,1}^{1,i}$ in the first item of (22) is known, the average channel gain $\{h_l^{\text{ave}}\}_{l=0}^L$ can be obtained easily by least square (LS) scheme.

Similarly, $\mathbf{R}_{p,2}^{1,i}$ can be simplified as

$$\mathbf{R}_{p,2}^{1,i}(k) = \frac{1}{N_p} \mathbf{C} \times \boldsymbol{\mu}, \quad (24)$$

where $\boldsymbol{\mu} = [\mu_0, \mu_1, \dots, \mu_L]^T$, and $\mathbf{C}$ is calculated as

$$\mathbf{C} = \mathbf{E} \times \mathbf{F}, \quad (25)$$

where $\mathbf{E} = [E(k,m)]_{N_p \times N_p}$, $\mathbf{F} = [F(m,l)]_{N_p \times L}$, and

$$\left. \begin{aligned} E(k,m) &= \sum_{n=0}^{N_p-1} \left(n - \frac{N_p}{2} \right) \exp\left(j \frac{2\pi}{N_p} (m-k)n\right), \\ 0 \leq k, m < N_p, \\ F(m,l) &= P_m^i \times \exp\left(-j \frac{2\pi}{N_p} ml\right), \\ 0 \leq l \leq L, 0 \leq m < N_p. \end{aligned} \right\} \quad (26)$$

From (26), the channel slope $\boldsymbol{\mu}$ can be obtained:

$$\boldsymbol{\mu} = N_p \mathbf{C}^+ \times \mathbf{R}_{p,2}^{1,i}, \quad (27)$$

where $\mathbf{C}^+$ is the pseudo inverse of $\mathbf{C}$.

Similarly, the coefficient matrices of frequency response of ISI are described as

$$\left. \begin{aligned} \mathbf{G}^{2,i} &= \mathbf{F}_{N_p} (\mathbf{A}_{\text{ISI}}^{\text{ave},i} + \mathbf{M} \times \mathbf{A}_{\text{ISI}}^{\text{var},i}) \mathbf{F}_N^H, \\ \mathbf{G}^{3,i} &= -[\mathbf{F}_{N_p} (\mathbf{B}_{\text{ISI}}^{\text{ave},i} + \mathbf{M} \times \mathbf{B}_{\text{ISI}}^{\text{var},i}) \mathbf{F}_N^H]. \end{aligned} \right\} \quad (28)$$

In conclusion, the received frequency domain pilot response can be shown as

$$\mathbf{R}_p^i = \mathbf{R}_{p,1}^{1,i} + \mathbf{R}_{p,2}^{1,i} + \mathbf{R}_p^{2,i} + \mathbf{R}_p^{3,i} + \mathbf{W}^i, \quad (29)$$

where

$$\left. \begin{aligned} \mathbf{R}_{p,1}^{1,i}(k) &= G_1^{1,i}(k,k) P^i(k), \\ \mathbf{R}_{p,2}^{1,i}(k) &= \sum_{m=0}^{N_p-1} G_2^{1,i}(k,m) P^i(m), \\ \mathbf{R}_p^{2,i}(k) &= \sum_{m=0}^{N-1} G^{2,i}(k,m) S^{i-1}(m), \\ \mathbf{R}_p^{3,i}(k) &= \sum_{m=0}^{N_p-1} G^{3,i}(k,m) P^i(m). \end{aligned} \right\} \quad (30)$$

According to the above analysis, some conclusions can be obtained:

- $\mathbf{R}_p^{1,i}$ is the desired frequency-domain response, which includes $\mathbf{R}_{p,1}^{1,i}$ and $\mathbf{R}_{p,2}^{1,i}$. $\mathbf{R}_{p,1}^{1,i}$, the response of the channel average gains, is defined as the valid part without the ISI and ICI. $\mathbf{R}_{p,2}^{1,i}$, containing the information of tap slopes, is caused by the time variation.

- $\mathbf{R}_p^{2,i}$ and $\mathbf{R}_p^{3,i}$ are the frequency response and are caused by multipath impact.

3.2. Channel estimation

In the time-varying channel, the received pilot will be influenced by ISI and ICI in the OFDM systems from

Fig. 2 Channel estimation model diagram

the above analysis. To extract the channel parameters: $\{h_l^{\text{ave}}\}_{l=0}^L$ and $\{\mu_l\}_{l=0}^L$ over the pilot duration, a channel estimation algorithm is described in Fig. 2.

3.3. Signal detection based on circular convolution reconstruction

An OFDM symbol detection algorithm is studied in this subsection. According to the linear model [2], with the known channel average gains $\{h_l^{\text{ave},i}\}_{l=0}^L$ and $\{h_l^{\text{ave},i+1}\}_{l=0}^L$ by the pilots $\mathbf{p}^i$ and $\mathbf{p}^{i+1}$, the following channel parameters during the area 1 in Fig. 3 can be obtained.

The channel tap slopes:

$$\lambda_{\text{OFDM},l} = \frac{h_l^{\text{ave},i+1} - h_l^{\text{ave},i}}{N + N_p}, \quad 0 \leq l \leq L. \quad (31)$$

The channel average gain:

$$h_{\text{OFDM},l}^{\text{ave}} = \frac{h_l^{\text{ave},i+1} + h_l^{\text{ave},i}}{2}, \quad 0 \leq l \leq L. \quad (32)$$

The channel impulse response over the OFDM data unit $\mathbf{x}^i$ can be calculated as

$$h_{n,l}^{\text{OFDM}} = h_{\text{OFDM},l}^{\text{ave}} + \lambda_{\text{OFDM},l} \left(n - \frac{N}{2} \right), \quad 0 \leq l \leq L, \quad 0 \leq n < N. \quad (33)$$

where $h_{n,l}^{\text{OFDM}}$ represents the tap coefficient at the l th channel tap and time instant n . The i th received OFDM data is

$$\mathbf{r}_{\text{OFDM}}^i = (r_{\text{OFDM}}^i(0), r_{\text{OFDM}}^i(1), r_{\text{OFDM}}^i(2), \dots, r_{\text{OFDM}}^i(N-1))^T,$$

where

$$r_{\text{OFDM}}^i(n) = \begin{cases} \sum_{l=0}^n h_{n,l}^{\text{OFDM}} s_{\langle n-l \rangle N}^i + \sum_{l=n+1}^L h_{n,l}^{\text{OFDM}} p_{\langle n-l \rangle N_p}^i + w_n^i, & 0 \leq n < L, \\ \sum_{l=0}^L h_{n,l}^{\text{OFDM}} s_{\langle n-l \rangle N}^i + w_n^i, & L \leq n < N_p, \end{cases} \quad (34)$$

The first L samples in $\mathbf{r}_{\text{OFDM}}^i$ may be disturbed by ISI for

Fig. 3 The linear model for CIR

the lack of cyclic prefix. To demodulate the received time-domain OFDM signal $\mathbf{r}_{\text{OFDM}}^i$, the ISI must be mitigated and circular convolution $\tilde{\mathbf{r}}_{\text{OFDM}}^i$ should be reconstructed:

$$\tilde{\mathbf{r}}_{\text{OFDM}}^i = (\tilde{r}_{\text{OFDM}}^i(0), \tilde{r}_{\text{OFDM}}^i(1), \tilde{r}_{\text{OFDM}}^i(2), \dots, \tilde{r}_{\text{OFDM}}^i(N-1))^T,$$

where

$$\tilde{r}_{\text{OFDM}}^i(n) = \begin{cases} r_{\text{OFDM}}^i(n) - r_{\text{Re_ISI}}^i(n) + r_{\text{Add_ISI}}^i(n), & 0 \leq n < L, \\ r_{\text{OFDM}}^i(n), & L \leq n < N, \end{cases} \quad (35)$$

and

$$\begin{cases} r_{\text{Re_ISI}}^i(n) = \sum_{l=n+1}^L h_{n,l}^{\text{OFDM}} p_{\langle n-l \rangle N_p}^i, & 0 \leq n < L, \\ r_{\text{Add_ISI}}^i(n) = \sum_{l=n+1}^L h_{n,l}^{\text{OFDM}} s_{\langle n-l \rangle N}^i, & 0 \leq n < L. \end{cases} \quad (36)$$

Two steps could be taken to complete ISI cancellation and reconstruct circular convolution:

- Firstly, $\mathbf{r}_{\text{Re_ISI}}^i$ can be calculated from $\mathbf{p}^i$ and the estimated CIR $\mathbf{h}_{n,l}^{\text{OFDM}}$ in (36).

- Secondly, estimate $\mathbf{r}_{\text{Add_ISI}}^i$ in (35). Since $\mathbf{s}^i$ is unknown, it is not easy to calculate $\mathbf{r}_{\text{Add_ISI}}^i$ directly. The initial L samples of the time-domain response $\mathbf{r}_p^{i+1}$ of $\mathbf{p}^{i+1}$ contain the value information to reconstruct $\mathbf{r}_{\text{Add_ISI}}^i$. Then, $\mathbf{r}_p^{i+1} = (r_p^{i+1}(0), r_p^{i+1}(1), r_p^{i+1}(2), \dots, r_p^{i+1}(N_p-1))^T$ are calculated as

$$r_p^{i+1}(n) = r_{p,1}^{i+1}(n) + r_{p,2}^{i+1}(n), \quad 0 \leq n < L, \quad (37)$$

where

$$\begin{cases} r_{p,1}^{i+1}(n) = \sum_{l=0}^n h_{n,l}^{i+1} p_{\langle n-l \rangle N_p}^{i+1} + w_n^{i+1}, & 0 \leq n < L, \\ r_{p,2}^{i+1}(n) = \sum_{l=n+1}^L h_{n,l}^{i+1} s_{\langle n-l \rangle N}^{i+1} + w_n^{i+1}, & 0 \leq n < L, \end{cases} \quad (38)$$

$h_{n,l}^{i+1}$ represents the channel tap coefficient over the pilot $\mathbf{p}^{i+1}$ duration, and $\mathbf{r}_{p,2}^{i+1}$ can be obtained by subtracting $\mathbf{r}_{p,1}^{i+1}$ in (37). Since the CIR is mainly decided by the channel average gain when the normalized Doppler frequency is less than 20%, it is reasonable that

$$r_{\text{Add_ISI}}^i(n) = r_{p,2}^{i+1}(n), \quad 0 \leq n < L. \quad (39)$$

Combining (35), (36), and (37), we obtain the corrected data $\tilde{\mathbf{r}}_{\text{OFDM}}^i$:

$$\tilde{\mathbf{r}}_{\text{OFDM}}^i = (\mathbf{H}^{\text{ave}} + \mathbf{M} \times \mathbf{A}) \times \mathbf{s}^i + \mathbf{w}^i, \quad (40)$$

where $\mathbf{H}^{\text{ave}}$, $\mathbf{M}$, and $\mathbf{A}$ are all N -order square matrices, and the specific expressions for their elements are as follows:

$$\left. \begin{aligned} H^{\text{ave}}(n, m) &= h_{\text{OFDM}, \langle n-m \rangle N}, \quad 0 \leq n, m < N, \\ A(n, m) &= \lambda_{\text{OFDM}, \langle n-m \rangle N}, \quad 0 \leq n, m < N, \\ M(n, m) &= n - \frac{N}{2}, \quad 0 \leq n < N. \end{aligned} \right\} \quad (41)$$

$$\left. \begin{aligned} D_1(k, k) &= \sum_{l=0}^L h_{\text{OFDM}, l} \exp\left(-j \frac{2\pi}{N} kl\right), \quad 0 \leq k < N, \\ D_2(k, m) &= \begin{cases} \sum_{l=0}^L \frac{\lambda_{\text{OFDM}, l}}{N} \exp\left(-j \frac{2\pi}{N} ml\right) \sum_{n=0}^{N-1} \left(n - \frac{N}{2}\right) \exp\left(j \frac{2\pi}{N} (m-k)n\right), \\ 0 \leq k, m < N, k \neq m, \\ 0, \quad k = m. \end{cases} \end{aligned} \right\} \quad (44)$$

The frequency domain equalizer can be used by normalized mean square error (NMSE) in (42) to obtain $\{\hat{S}_k^i\}_{k=0}^{N-1}$. After demodulation, the transmitted signal can be detected.

3.4. Traditional OFDM detection

In Fig. 1, the linear model remains in the OFDM data block $\bar{\mathbf{x}}^i$, and the identified pilot is used in the system. In order to make a comparison between the proposed detection method and the traditional method, we studied a traditional OFDM detection algorithm as follows.

For the estimated $\{h_l^{\text{ave}}\}_{l=0}^L$ and $\{\mu_l\}_{l=0}^L$ over the p^i duration, the channel coefficient over the OFDM symbol block ($\bar{\mathbf{x}}^i$) duration can be analyzed according to Fig. 4:

$$h_{n,l}^i = h_l^{\text{ave}} + \mu_l \left(n - \frac{N_p}{2}\right), \quad 0 \leq n < N_3, \quad 0 \leq l \leq L, \quad (45)$$

where $h_{n,l}^i$ represents the channel coefficient at the l th

Fig. 4 The linear model for CIR

The frequency-domain response $\tilde{\mathbf{R}}_{\text{OFDM}}^i$ of $\tilde{\mathbf{r}}_{\text{OFDM}}^i$ is

$$\tilde{\mathbf{R}}_{\text{OFDM}}^i = \mathbf{D} \times \mathbf{S}^i + \mathbf{W}^i, \quad (42)$$

where $\mathbf{D}$ consists of two parts:

$$\mathbf{D} = (\mathbf{D}_1 + \mathbf{D}_2), \quad (43)$$

where $\mathbf{D}_1 = [D_1(k, k)]_{N \times N}$ is an N -order diagonal matrix, and $\mathbf{D}_2 = [D_2(k, m)]_{N \times N}$ is an N -order square matrix:

channel tap and time instant n .

$\mathbf{r}_d^i$ can be calculated as

$$\mathbf{r}_d^i = \mathbf{H}_T^{\text{ave}} \times \mathbf{x}^i + \mathbf{M}_T \times \mathbf{H}_T^{\text{var}} \times \mathbf{x}^i + \mathbf{w}^i, \quad (46)$$

where $\mathbf{H}_T^{\text{ave}}$ and $\mathbf{H}_T^{\text{var}}$ are all N -order square matrices, $\mathbf{M}_T$ is an N -order diagonal matrix, and the specific expressions for their elements are as follows:

$$\left. \begin{aligned} H_T^{\text{ave}}(n, k) &= h_{\langle n-k \rangle N_2}^{\text{ave}}, \quad 0 \leq n, k < N_2, \\ H_T^{\text{var}}(n, k) &= \mu_{\langle n-k \rangle N_2}, \quad 0 \leq n, k < N_2, \\ M_T(n, k) &= k + \frac{N_p}{2}, \quad 0 \leq k < N_2, \end{aligned} \right\} \quad (47)$$

Similarly, after FFT in (46), the frequency-domain equalization is completed to obtain the time-domain transmitted signal $\hat{\mathbf{x}}^i$. Then, the transmitted OFDM data block $\hat{\mathbf{s}}^i$ can be calculated as

$$\hat{\mathbf{s}}^i = \Phi_{N \times N_2} \hat{\mathbf{x}}^i, \quad (48)$$

where $\Phi_{N \times N_2} = [\mathbf{I}_N, \mathbf{0}_{N \times N_p}]$, $\mathbf{I}_N$ is an N -dimensional unit matrix, and $\mathbf{0}_{N \times N_p}$ is a zero matrix. After FFT and demodulation, we could obtain the transmitted signal.

In the traditional detection method, a single pilot is used to estimate the channel coefficient over the OFDM symbol block and complete detection. It requires that the linear variation relationship should maintain over $(N + 2N_p) \times T_s$ duration, where T_s is the sampling period. However, the assumption is not always appropriate and may degrade the performance in the fast varying

environment. In the proposed improved signal detection, however, two block pilots are used to estimate the channel coefficient and complete signal detection with interference cancellation and circular convolution reconstruction. It only requires that the CSI satisfies a linear model in the $(N + N_p) \times T_s$ duration. The proposed method may cause some slight processing delay, but brings better performance.

4. Simulation results

The system parameters in a fast time-varying environment are shown in Table 1. A two-delay Rayleigh channel, as described in Table 2, is used to analyze the performance. The minimum mean square error (MMSE) of channel estimation is calculated as

$$\text{MMSE} = \frac{1}{M} \sum_{i=0}^M \left\{ \frac{\sum_{n=0}^{N_p-1} \sum_{l=0}^L |h(n,l) - \hat{h}(n,l)|^2}{\sum_{n=0}^{N_p-1} \sum_{l=0}^L |h(n,l)|^2} \right\}, \quad (49)$$

where $\hat{h}(n,l)$ represents the estimated channel coefficient of $h(n,l)$, and M the number of pilot blocks. The ISI ratio is defined as the ratio of the normalized maximum multipath delay to the block pilot duration.

To analyze the performance in the two delayed Rayleigh channel with normalized Doppler frequency of 0.1 and different high multipath delay, the channel estimation performance is depicted in Fig. 5 and Fig. 6 with signal band of 40 KHz. It shows that the proposed algorithm can effectively track the channel state information. Even if the ISI proportion is 15%, the MMSE can be reduced to 4×10^{-3} at SNR of 24 dB after three iterations in Fig. 6. The BER is shown in Fig. 7 and Fig. 8.

Compared to the signal detection performance with the estimated channel information obtained without ISI cancellation (ISIC), the improved performance can be provided with channel state information obtained by ISIC and circular convolution reconstruction in the traditional OFDM detection, labeled by method 1, or the

Table 1 System parameters

Modulation	FFT piont	Length of pilot	Number of iterations
QPSK	512	400	$I=3$

Table 2 Two delayed Rayleigh channel

Tap	Delay (μ s)	Power (dB)
1	0	-0.01
2	40/60	-0.01

proposed improved method, labeled by method 2.

Fig. 9 shows the BER performance with the standard TDS-OFDM system parameter in [3] in the two delayed Rayleigh channel. It implies that the proposed method, method 2, can effectively improve the BER gain. Furthermore, in contrast to method 1, the BER obtained by the proposed way is nearly close to that with perfect channel state information.

Fig. 5 Performance of MMSE with ISI of 10%

Fig. 6 Performance of MMSE with ISI of 15%

Fig. 7 BER vs. SNR with ISI of 10%

Fig. 8 BER vs. SNR with ISI of 15%

Fig. 9 BER vs. SNR with ISI 15% in TDS-OFDM

5. Conclusion

In the fast time-varying environment, the channel impulse response can be approximated to vary in a linear fashion when the normalized Doppler spread is less than 20%. This paper proposes an effective detection algorithm to complete the channel estimation and OFDM symbol demodulation through ISI cancellation and circular convolution reconstruction. The simulation results show that the proposed algorithm can effectively track the channel variation even under a large Doppler frequency and high multipath delay. Although some slight processing delay may appear, the proposed method provides better gain in symbol detection than the traditional way and the BER is close to the performance with perfect channel state information.

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References

- [1] X. Cai, G.B. Giannakis, Low-complexity ICI suppression for OFDM over time- and frequency-selective Rayleigh fading channels, In: *Proc. Asilomar Conf. Signals, Systems and Computers*, Minnesota Univ., USA, Nov. 2002: 822-826.
- [2] W.G. Jeon, K.H. Chuang, Y.S. Cho, An equalization technique for orthogonal frequency division multiplexing systems in time-variant multipath channels, *IEEE Trans. Commun.*, 1999, **49**(1): 1185-1191.
- [3] W. Zhang, Y. Guan, W. Liang, et al., An introduction of the Chinese DTTB standard and analysis of the PN595 working modes, *IEEE Trans. Broadcast.*, 2007, **53**(1): 8-13.
- [4] H. Steendam, Different guard interval techniques for OFDM: performance comparison, In: *Proceedings from International Workshop on Multicarrier Spread Spectrum*, Herrsching, Germany, May. 2007, Vol. 1: 11-24.
- [5] T. Hwang, C. Yang, G. Wu, et al., OFDM and its wireless applications: a survey, *IEEE Trans. On Vehicular Technology*, 2009, **58**(4): 1673-1694.
- [6] S. Raza, S. Poonpat, Iterative channel estimation using joint zero padding and nonzero padding for TDS-OFDM systems, In: *2010 International Conference on Electrical Engineering/Electronics Computer Telecommunications and Information Technology (ECTI-CON)*, Chaing Mai, May 19-21, 2010: 983-987.
- [7] S. Chen, T. Yao, Inter-carrier interference suppression and channel estimation for OFDM systems in time-varying frequency-selective fading channels, *IEEE Transactions on Consumer Electronics*, 2004, **50**(2): 429-435.
- [8] Z. Li, X. Lei, W. Tang, et al., Channel estimation for OFDM time-variant multi-path environment, In: *IEEE Vehicular Technology Conference, VTC*, Nat. Key Lab. of Commun., Univ. of Electron., May 11-14, 2008: 356-360.
- [9] Y. Mostofi, D.C. Cox, ICI mitigation for pilot-aided OFDM mobile systems, *IEEE Transactions on Wireless Communications*, 2005, **4**(2): 65-74.
- [10] J.M. Holtzman, A. Sampath, Adaptive averaging methodology for handoffs in cellular systems, *IEEE Transactions on Vehicular Technology*, 1995, **44**(1): 59-66.
- [11] J.J. Beek, M. Sandell, P.O. Bourjesson, ML estimation of time and frequency offset in OFDM system, *IEEE Transactions on Signal Processing*, 1997, **45**(7): 1800-1805.
- [12] C. Jueping, S. Wentao, L. Zan, Doppler spread estimation for mobile OFDM systems in Rayleigh fading channels, consumer electronics, *IEEE Transactions on Consumer Electronics*, 2003, **49**(4): 973-977.
- [13] S. Alamouti, Doppler frequency determination for mobile wireless devices, Patent US 2007070554, 2007-11-12.
- [14] A. Seyedi, G.J. Saulnier, General ICI self-cancellation scheme for OFDM systems, *IEEE Transactions on Vehicular Technology*, 2005, **54**(1): 198-210.

- [15] J.H. Ni, Z.M. Liu, A Joint ICI estimation and mitigation scheme for OFDM systems over fast fading channels, In: *Global Mobile Congress 2009*, Shanghai, Oct. 12-14, 2009: 1-6.
- [16] Y. Zhao, S.G. Haggman, Inter-carrier interference self-cancellation scheme for OFDM mobile communication systems, *IEEE Transactions on Commun.*, 2001, **49**(7): 1185-1191.
- [17] R. Merched, N. Yousef, Efficient Doppler compensation method and receiver for orthogonal-frequency-division-multiplexing (OFDM) systems, Patent US 20087424062, 2008-09-09.
- [18] A.M. Sayeed, B. Aazhang, Joint multipath-Doppler diversity in mobile wireless communications, *IEEE Trans on Commun.*, 1999, **47**(1): 123-132.
- [19] A. Dammann, On the influence of cyclic delay diversity and Doppler diversity on the channel characteristics in OFDM systems, In: *IEEE International Conference on Communications (ICC'07)*, Glasgow, June 24-28, 2007: 4179-4184.
- [20] B.C. Kim, I.T. Lu, Doppler diversity for OFDM wireless mobile communications. part I: frequency domain approach, In: *IEEE Vehicular Technology Conference, VTC*, South Korea, April 22-25, 2003, Vol. 4: 2677-2681.
- [21] X. Ma, G.B. Giannakis, Maximum-diversity transmissions over time-selective wireless channels, In: *Proc. of Wireless Comm. and Networking Conf*, March 17-21, 2002: 1497-501.
- [22] J. Wu, Y.R. Zheng, K.B. Letaief, et al., On the error performance of wireless systems with frequency selective fading and receiver timing phase offset, *IEEE Transactions on Wireless Communications*, 2007, **6**(2): 720-729.

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