

Turbo MIMO-OFDM Receiver in Time-Varying Channels

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Abstract

This paper proposes an advanced turbo receiver with joint inter-carrier interference (ICI) self cancellation and channel equalization for multiple-input multiple-output orthogonal frequency division multiplexing (MIMO-OFDM) systems over rapidly time-varying channel environment. The ICI caused by impairment of local oscillators and carrier frequency offset (CFO) is the major problem for MIMO-OFDM communication systems. The existing schemes (conjugate cancellation (CC) and phase rotated conjugate cancellation (PRCC)) that deal with the ICI cancellation and channel equalization can't provide satisfactory performance over time-varying channels. In term of error rate performance and low computational complexity, ICI self cancellation is the best choice. So, this paper proposes a turbo receiver to deal with the problem of joint ICI self cancellation and channel equalization. We employ the adaptive phase rotations in the receiver to effectively track the CFO variations without feeding back the CFO estimate to the transmitter as required in traditional existing scheme. We also give some simulations to verify the proposed scheme. The proposed scheme outperforms the existing schemes.

Keywords: MIMO-OFDM, inter-carrier interference, ICI self-cancellation, channel estimation

1. Introduction

The MIMO transmission system is employed to increase the transmitted data rate and improve the communication link quality compared with single-antenna system [1]-[3]. Furthermore, the Space-Time Block codes provides temporal/spatial multiplexing and have been employed to enhance the reliability of MIMO wireless communication systems and improve the performance of MIMO communications [4]. The OFDM technique can effectively combat the problem of frequency selective fading and can provide the advantages of high-data-rate transmission and high spectral efficiency [5]-[6]. Many wireless standards have employed the OFDM technique [7]-[8]. Therefore, for this reason, the MIMO-OFDM technique exploits the advantages of the MIMO and OFDM and has been a promising technique for future wireless systems. The main problem of the MIMO-OFDM system is that the system is highly sensitive to the CFO.

In order to effectively deal with the problem of the CFO, the estimation the CFO must be derived and the CFO compensation is then performed that is according to the estimated results [9]-[11]. For OFDM communication systems, based on the assumption that the channel variation is with approximate linearity in an OFDM symbol, [12]-[14] proposed the ICI cancellation schemes that use training sequences and pilot symbols to perform frequency-domain channel estimation/equalization over fast fading multipath channels. Also, in the literatures, some techniques without employing pilot symbols or training sequences is available to cancel ICI. Among these schemes, the ICI self-cancellation approaches have attracted much attention. The ICI self-cancellation approaches have the characteristic that the ICI can be cancelled in the reception procedure without performing complicated equalization or inserting extra symbols. The conjugate cancellation (CC) [15] and phase rotated conjugate cancellation (PRCC) [16] are the two most important ICI self-cancellation schemes. The existing schemes (conjugate cancellation (CC) [15] and phase rotated conjugate cancellation (PRCC) [16]) that deal with the ICI cancellation and channel equalization can't provide satisfactory performance over time-varying channels. For the CC scheme, the first path transmits the original OFDM symbol, and the second path transmits the conjugate version of the OFDM symbol. Based on the configuration, we find that the ICI in the first path is restrained by the second path. For the PRCC scheme, by applying a phase rotation to the transmitter in the CC scheme, we find that the ICI caused by both paths can be mutually cancelled more effectively. The existing CC and PRCC schemes can't work well in time-varying channels where the CFOs changes fast with time. Moreover, the existing PRCC scheme must feedback the estimation of CFO to the transmitter to obtain the optimal phase rotation. Therefore, the PRCC scheme needs additional signal overhead.

In this paper, we propose a turbo adaptive receiver that can overcome the above mentioned problem. Furthermore, the existing two-stage IQ imbalance scheme [17] uses a two-stage procedure to estimate and compensate the CFO and the IQ imbalance for MIMO-OFDM communications. Although the two-stage IQ imbalance scheme proposes a new CFO compensation method to cancel the ICI for the MIMO-OFDM communications, the channel equalization and channel estimation that can improve the bit error rate performance have not been considered. The phase error problem is also not considered in the two-stage IQ imbalance scheme. So the two-stage IQ imbalance scheme can't provide satisfactory performance under environments with large CFOs. We propose a turbo receiver for ICI

self-cancellation by taking advantage of conjugate transmission [18] for the OFDM communications. Unlike the existing PRCC method, the proposed scheme uses two phase rotations on both paths at the receiver to cope with the fast fading channel. Furthermore, in order to obtain the formula of between the CFO and the phase rotation, we also derive the optimal phase rotations based in the maximization of the carrier-to-interference ratio (CIR), and also propose an adaptive process employing the normalized block least mean-squared (BLMS) algorithm [19]-[21]. Therefore, the requirement of feeding back the estimation of CFO information can be eliminated. [22] proposed a precoding-based blind channel estimation for MIMO-OFDM systems. This scheme doesn't solve the problem of the CFO and can't be applied to the time-varying channel. [23] proposed a decision-directed channel estimation scheme to deal with the shortage of pilot for the MIMO-OFDM systems [23]. To practically analyze and effectively solve the CFO problem of MIMO-OFDM systems over time-varying channels, this paper joint considers the channel estimation/equalization based on the Kalman algorithm [24] [25], turbo minimum mean square error (MMSE) equalization [26] and the CFO compensation that employs an adaptive modified PRCC receiver [18]. The proposed Kalman algorithm belongs to the pilot-symbol-aided parametric channel estimation method in which the channel responses are characterized as a collection of sparse propagation paths. Based on the signal subspace of the channel samples' correlation matrix, the estimation of channel parameters can be translated into an unconstrained minimization problem. Then, in order to solve this optimization problem, a subspace tracking by Kalman filter is carried out which is characterized in the state equation and the measurement equation. Turbo equalization is joint equalization and decoding and had been proposed for MIMO OFDM systems [27]-[29]. [30] deals with frequency-domain equalization for faster-than-Nyquist (FTN) signaling in doubly selective channels (DSCs). [31] studies the receiver design for a coded faster-than-Nyquist signaling system in time-variant intersymbol interference channels. In this paper, we design a turbo receiver that is joint Kalman channel estimation, CFO compensation and turbo equalization for MIMO OFDM system over time-varying channels. The proposed scheme outperforms the existing schemes. The organization of this paper is as follows. In section 2, the system model, the existing CC scheme and the existing PRCC scheme are described. The proposed scheme is described in section 3, the detailed frequency offset estimation scheme, the Kalman channel estimation algorithm and Turbo equalization are also given in section 3. Finally, some simulation results and conclusions are provided in section 4 and section 5, respectively.

2. System Model

The considered MIMO-OFDM transmitter is described as Fig. 1. The proposed receiver is described as Fig. 2. The time-domain received signals of the two paths can be expressed as follows,

$$\mathbf{y}_{\alpha,l,n}^{(1)} = \sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)} \quad l = 0, 1, \dots, N-1 \quad (1)$$

$$\mathbf{y}_{\alpha,l,n}^{(2)} = \sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n}^* h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} + w_{\alpha,l,n}^{(2)} \quad l = 0, 1, \dots, N-1 \quad (2)$$

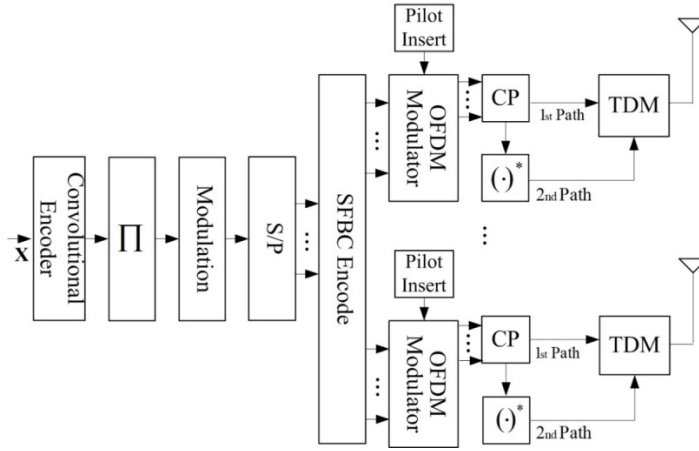


Fig. 1. The transmitter.

where L is the number of channel tap, $\mathbf{y}_{\alpha,l,n}^{(i)}$ and $w_{\alpha,l,n}^{(i)}$ correspond to the received symbol and noise at the α -th receiving antenna, l -th subcarrier in the i -th path within the n -th transmission block before the FFT demodulation, respectively. $h_{\beta,\alpha,p,n}$ corresponds to the channel tap at the β -th transmitting antenna, α -th receiving antenna and p -th path within the n -th transmission block, and then $\beta = 1, \dots, N_t$ and $\alpha = 1, \dots, N_r$, N_t is the total number of transmitting antenna, N_r is the total number of receiving antenna. The channel taps vary per block. If we employ 2x2 STBC encoder that $d_{1,l,n}$ is the 2x2 STBC encoder output at the first antenna in the l -th subcarrier within the n -th transmission block, we obtain

$$\begin{bmatrix} d_{1,l,n} \\ d_{2,l,n} \end{bmatrix} = \begin{bmatrix} x_0 & -x_1^* & \cdots & x_{N-2} & -x_{N-1}^* \\ x_1 & x_0^* & \cdots & x_{N-1} & x_{N-2}^* \end{bmatrix} \quad (3)$$

The frequency-domain received signals of these two paths can be written as follows,

$$\begin{aligned} \mathbf{Y}_{\alpha,m,n}^{(1)} &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} (\mathbf{y}_{\alpha,l,n}^{(1)}) e^{-j\frac{2\pi}{N}lm} \\ &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \left(\sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\ &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \sum_{\beta=1}^{N_t} \left(\frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} \left[D_{\beta,i,n} e^{j\frac{2\pi}{N}(l-p)i} \left(\sum_{p=0}^{L-1} h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} \right] \right. \\ &\quad \left. + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\ &= \sum_{\beta=1}^{N_t} \sum_{i=0}^{N-1} D_{\beta,i,n} \left[\frac{1}{N} \sum_{l=0}^{N-1} \left(\sum_{p=0}^{L-1} h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{-j\frac{2\pi}{N}(i+\epsilon_{\alpha,n}-lm)} \right] \\ &\quad + \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} w_{\alpha,l,n}^{(1)} e^{-j\frac{2\pi}{N}lm} \\ &= D_{\beta,m,n} H_{\alpha,\beta,p,n} C(-\epsilon_{\alpha,n}) + \sum_{\substack{i=0 \\ i \neq m}}^{N-1} D_{N_t,i,n} H_{N_t,N_r,i,n} C(m-i-\epsilon_{\alpha,n}) + W_{\alpha,m,n}^{(1)} \end{aligned} \quad (4)$$

$$m = 0, 1, \dots, N-1$$

$$\begin{aligned}
\mathbf{Y}_{\alpha,m,n}^{(2)} &= \frac{\mathbf{1}}{\sqrt{N}} \sum_{l=0}^{N-1} (\mathbf{y}_{\alpha,l,n}^{(2)})^* e^{-j\frac{2\pi}{N}lm} \\
&= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \left(\sum_{p=0}^{L-1} d_{\beta,(l-p),n}^* h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}l(\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n})} + w_{\alpha,l,n}^{(2)} \right)^* e^{-j\frac{2\pi}{N}lm} \\
&= D_{\beta,m,n} H_{\alpha,\beta,p,n} C(\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n}) \\
&\quad + \sum_{\substack{i=0 \\ i \neq m}}^{N-1} D_{\beta,i,n} H_{\alpha,\beta,i,n} C(m - i + (\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n})) + W_{\alpha,m,n}^{(2)}, \\
m &= 0, 1, \dots, N - 1
\end{aligned} \tag{5}$$

where $D_{\beta,m,n}$ is the FFT of 2x2 STBC encoder output. By combining and averaging the signal processed in (4) and (5), and define $C(v) = \frac{1}{N} \sum_{n=0}^{N-1} e^{-j\frac{2\pi}{N}nv}$, we obtain

$$\begin{aligned}
\mathbf{R}_{\alpha,m,n} &= \frac{1}{2} (\mathbf{Y}_{\alpha,m,n}^{(1)} + \mathbf{Y}_{\alpha,m,n}^{(2)}) \\
&= \frac{1}{2} [e^{j\varphi_{\alpha,n}} C(-\varepsilon_{\alpha,n}) + e^{-j\varphi_{\alpha,n}} C(+\varepsilon_{\alpha,n})] \sum_{\beta=1}^{N_t} D_{\beta,m,n} H_{\beta,\alpha,m,n} \\
&\quad + \frac{1}{2} \sum_{\substack{i=0 \\ i \neq m}}^{N-1} [e^{j\varphi_{\alpha,n}} C(m - i - \varepsilon_{\alpha,n}) + e^{-j\varphi_{\alpha,n}} C(m - i + \varepsilon_{\alpha,n})] D_{\beta,m,n} H_{\beta,\alpha,m,n} \\
&\quad + \frac{1}{2} [W_{\alpha,m,n}^{(1)} + W_{\alpha,m,n}^{(2)}], \quad m = 0, 1, \dots, N - 1
\end{aligned} \tag{6}$$

The CIR of the CC scheme in 2x2 MIMO-OFDM can be defined as

$$CIR_{\alpha,cc} = \frac{|C(-\varepsilon_{\alpha,n}) + C(\varepsilon_{\alpha,n})|^2}{\sum_{i=0}^{N-1} |C(i - \varepsilon_{\alpha,n}) + C(i + \varepsilon_{\alpha,n})|^2} \tag{7}$$

For the PRCC scheme, the first path transmits the original MIMO-OFDM signals with phase rotation $\varphi_{\alpha,n}$, while the second path transmits the conjugate version of the original MIMO-OFDM signals with phase rotation $-\varphi_{\alpha,n}$, *i.e.*, the $d_{\beta,l,n} e^{j\varphi_{\alpha,n}}$ signal is transmitted in the first path and the $\{d_{\beta,l,n} e^{-j\varphi_{\alpha,n}}\}^*$ is transmitted in the second path, respectively. Therefore, the time-domain received signals of the two paths can be expressed as follows,

$$\mathbf{y}_{\alpha,l,n}^{(1)} = \sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\beta,\alpha,p,n} e^{j\varphi_{\alpha,n}} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)}, \quad l = 0, 1, \dots, N - 1 \tag{8}$$

$$\mathbf{y}_{\alpha,l,n}^{(2)} = \sum_{p=0}^{L-1} d_{\beta,(l-p),n}^* h_{\beta,\alpha,p,n} e^{j\varphi_{\alpha,n}} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} + w_{\alpha,l,n}^{(2)}, \quad l = 0, 1, \dots, N - 1 \tag{9}$$

And the frequency-domain received signals of the two paths can be written as follows,

$$\mathbf{Y}_{\alpha,m,n}^{(1)} = \frac{\mathbf{1}}{\sqrt{N}} \sum_{l=0}^{N-1} (\mathbf{y}_{\alpha,l,n}^{(1)}) e^{-j\frac{2\pi}{N}lm} \tag{10}$$

$$\begin{aligned}
&= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \left(\sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\alpha,\beta,p,n} e^{j\phi_{\alpha,n}} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\
&= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \sum_{\beta=1}^{N_t} \left(\frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} \left[D_{\beta,i,n} e^{j\frac{2\pi}{N}li} \left(\sum_{p=0}^{L-1} h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{j\phi_{\alpha,n}} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} \right] + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\
&= \sum_{\beta=1}^{N_t} \sum_{i=0}^{N-1} D_{\beta,i,n} \left[\frac{1}{N} \sum_{l=0}^{N-1} \left(\sum_{p=0}^{L-1} h_{\alpha,\beta,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{j\phi_{\alpha,n}} e^{-j\frac{2\pi}{N}(m-i-\varepsilon_{\alpha,n})} \right] + \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} w_{\alpha,l,n}^{(1)} e^{-j\frac{2\pi}{N}lm} \\
&= D_{\beta,m,n} H_{\alpha,\beta,p,n} e^{j\phi_{\alpha,n}} C(-\varepsilon_{\alpha,n}) + \sum_{\substack{i=0 \\ i \neq m}}^{N-1} D_{N_t,i,n} H_{\alpha,\beta,i,n} e^{j\phi_{\alpha,n}} C(m-i-\varepsilon_{\alpha,n}) + W_{\alpha,m,n}^{(1)}, \\
&\quad m = 0, 1, \dots, N-1 \\
\mathbf{Y}_{\alpha,m,n}^{(2)} &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} (\mathbf{y}_{\alpha,l,n}^{(2)})^* e^{-j\phi_{\alpha,n}} e^{-j\frac{2\pi}{N}lm} \\
&= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \left(\sum_{p=0}^{L-1} d_{\beta,(l-p),n}^* h_{\alpha,\beta,p,n} \varepsilon_{\alpha,n} e^{j\frac{2\pi}{N}l\varepsilon_{\alpha,n}} + w_{\alpha,l,n}^{(2)} \right)^* e^{-j\frac{2\pi}{N}lm} \\
&= D_{\beta,m,n} H_{\alpha,\beta,p,n} e^{-j\phi_{\alpha,n}} C(\varepsilon_{\alpha,n}) + \sum_{\substack{i=0 \\ i \neq m}}^{N-1} D_{\beta,i,n} H_{\alpha,\beta,i,n} C(m-i+\varepsilon_{\alpha,n}) + W_{\alpha,m,n}^{(2)}, \\
&\quad m = 0, 1, \dots, N-1
\end{aligned} \tag{11}$$

By combining and averaging the signal processed in (10) and (11), we obtain

$$\begin{aligned}
\mathbf{R}_{\alpha,m,n} &= \frac{1}{2} (\mathbf{Y}_{\alpha,m,n}^{(1)} + \mathbf{Y}_{\alpha,m,n}^{(2)}) \\
&= \frac{1}{2} [e^{j\phi_{\alpha,n}} C(-\varepsilon_{\alpha,n}) + e^{-j\phi_{\alpha,n}} C(+\varepsilon_{\alpha,n})] \sum_{\beta=1}^{N_t} D_{\beta,m,n} H_{\beta,\alpha,m,n} \\
&\quad + \frac{1}{2} \sum_{\substack{i=0 \\ i \neq m}}^{N-1} [e^{j\phi_{\alpha,n}} C(m-i-\varepsilon_{\alpha,n}) + e^{-j\phi_{\alpha,n}} C(m-i+\varepsilon_{\alpha,n})] D_{\beta,m,n} H_{\beta,\alpha,m,n} \\
&\quad + \frac{1}{2} [W_{\alpha,m,n}^{(1)} + W_{\alpha,m,n}^{(2)}], \quad m = 0, 1, \dots, N-1
\end{aligned} \tag{12}$$

From (12), the CIR of the PRCC scheme in 2x2 MIMO-OFDM can be defined as

$$CIR_{\alpha,prcc} = \frac{|e^{j\phi_{\alpha,n}} C(-\varepsilon_{\alpha,n}) + e^{-j\phi_{\alpha,n}} C(\varepsilon_{\alpha,n})|^2}{\sum_{i=0}^{N-1} |e^{j\phi_{\alpha,n}} C(i-\varepsilon_{\alpha,n}) + e^{-j\phi_{\alpha,n}} C(i+\varepsilon_{\alpha,n})|^2} \tag{13}$$

We can optimize $\phi_{\alpha,n}$ in terms of maximizing the CIR in (13),

$$\phi_{\alpha,n}^{(opt)} = \operatorname{argmax} CIR_{PRCC}(\varepsilon_{\alpha,n}, \phi_{\alpha,n}) \tag{14}$$

The optimal phase rotation can be express as follows

$$\phi_{\alpha,n}^{(opt)} = -\pi \varepsilon_{\alpha,n} \frac{N-1}{N} \tag{15}$$

3. The Proposed Turbo Receiver

The proposed turbo receiver for MIMO-OFDM system is shown as **Fig. 2**. With the presence of frequency offset $\varepsilon_{\alpha,n}$, the time-domain received signals from the two transmission paths can be described as follows,

$$\mathbf{y}_{\alpha,l,n}^{(1)} = \sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)} \quad l = 0, 1, \dots, N-1 \quad (16)$$

$$\mathbf{y}_{\alpha,l,n}^{(2)} = \sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n}^* h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l(\epsilon_{\alpha,n} + \Delta\epsilon_{\alpha,n})} + w_{\alpha,l,n}^{(2)} \quad l = 0, 1, \dots, N-1 \quad (17)$$

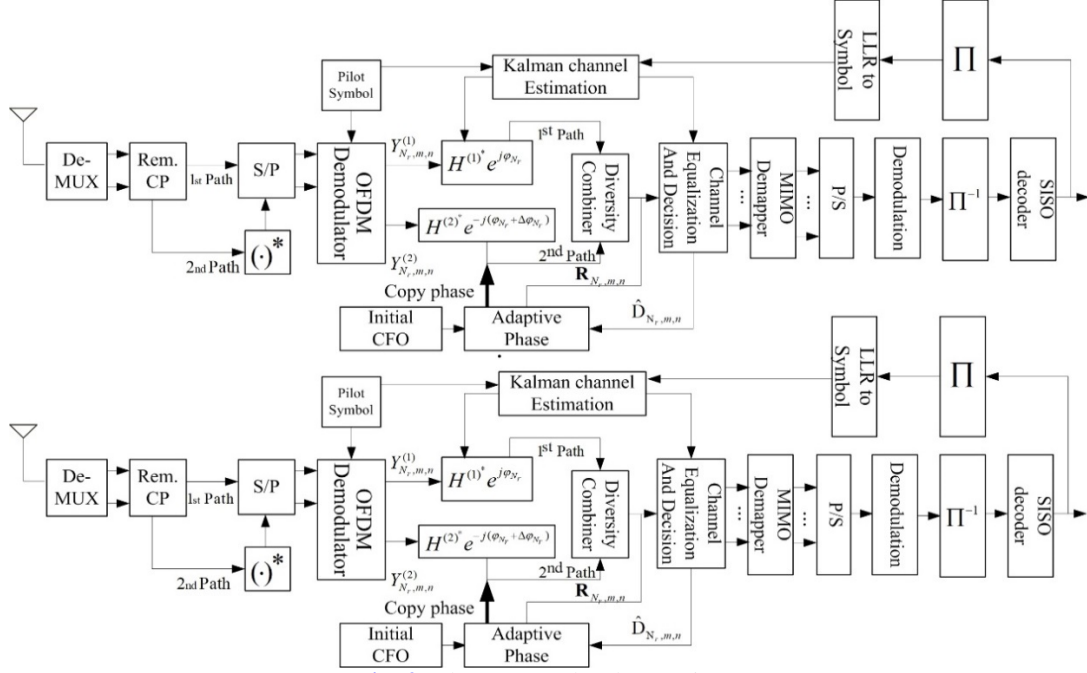


Fig. 2. The proposed turbo receiver.

After performing FFT operation, we obtain the following frequency-domain signals as follows,

$$\begin{aligned} \mathbf{Y}_{\alpha,m,n}^{(1)} &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} (\mathbf{y}_{\alpha,l,n}^{(1)}) e^{-j\frac{2\pi}{N}lm} \\ &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \left(\sum_{\beta=1}^{N_t} \sum_{p=0}^{L-1} d_{\beta,(l-p),n} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\ &= \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} \sum_{\beta=1}^{N_t} \left(\frac{1}{\sqrt{N}} \sum_{i=0}^{N-1} \left[D_{\beta,i,n} e^{j\frac{2\pi}{N}li} \left(\sum_{p=0}^{L-1} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{j\frac{2\pi}{N}l\epsilon_{\alpha,n}} \right] + w_{\alpha,l,n}^{(1)} \right) e^{-j\frac{2\pi}{N}lm} \\ &= \sum_{\beta=1}^{N_t} \sum_{i=0}^{N-1} D_{\beta,i,n} \left[\frac{1}{N} \sum_{l=0}^{N-1} \left(\sum_{p=0}^{L-1} h_{\beta,\alpha,p,n} e^{j\frac{2\pi}{N}ip} \right) e^{-j\frac{2\pi}{N}(m-i-\epsilon_{\alpha,n})l} \right] + \frac{1}{\sqrt{N}} \sum_{l=0}^{N-1} w_{\alpha,l,n}^{(1)} e^{-j\frac{2\pi}{N}lm} \\ &= D_{\beta,m,n} H_{\alpha,\beta,p,n} e^{j\theta_{\alpha,n}} C(-\epsilon_{\alpha,n}) + \sum_{\beta=1}^{N_t} \sum_{i=0}^{N-1} D_{N_t,i,n} H_{\alpha,\beta,i,n} e^{j\theta_{\alpha,n}} C(m-i-\epsilon_{\alpha,n}) + W_{\alpha,m,n}^{(1)} \end{aligned} \quad (18)$$

$$\begin{aligned} \mathbf{Y}_{\alpha,m,n}^{(2)} &= D_{\beta,m,n} H_{\alpha,\beta,p,n} C(\epsilon_{\alpha,n} + \Delta\epsilon_{\alpha,n}) + \sum_{\beta=1}^{N_t} \sum_{i=0}^{N-1} D_{\beta,i,n} H_{\alpha,\beta,i,n} C(m-i+\epsilon_{\alpha,n} + \Delta\epsilon_{\alpha,n}) + W_{\alpha,m,n}^{(2)} \quad m = 0, 1, \dots, N-1 \end{aligned} \quad (19)$$

where $\mathbf{Y}_{\alpha,m,n}^{(i)}$, $H_{\alpha,\beta,p,n}$ and $W_{\alpha,m,n}^{(i)}$ correspond to the received symbol, channel response and noise at the β -th transmitting antenna, the α -th receiving antenna, the m -th subcarrier in the i -th path within the n -th transmission after FFT modulation, respectively. $\varepsilon_{\alpha,n}$ and $\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n}$ denote the CFOs in the two paths at the α -th receiving antenna within the n -th transmission, respectively. After performing the individual phase rotations and the weighting coefficients, the signals in both paths can be written as follows,

$$\mathbf{R}_{\alpha,m,n}^{(1)} = e^{j\varphi_{\alpha,n}} \mathbf{H}_{\beta,\alpha,m,n}^{(1)*} \mathbf{Y}_{\alpha,m,n}^{(1)} \tag{20}$$

$$\mathbf{R}_{\alpha,m,n}^{(2)} = e^{j(\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n})} \mathbf{H}_{\beta,\alpha,m,n}^{(2)*} \mathbf{Y}_{\alpha,m,n}^{(2)} \tag{21}$$

where $\varphi_{\alpha,n}$ and $\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n}$ denote the designed phase rotations in the two paths at the α -th receiving antenna within the n -th transmission, respectively. Finally, the output signal can be expressed as follows,

$$\begin{aligned} \mathbf{R}_{\alpha,m,n} &= \frac{1}{2} (\mathbf{Y}_{\alpha,m,n}^{(1)} + \mathbf{Y}_{\alpha,m,n}^{(2)}) \\ &= \frac{1}{2} \sum_{\beta=1}^{N_t} \left(D_{\beta,m,n} |H_{\beta,\alpha,m,n}^{(1)}|^2 e^{j\varphi_{\alpha,n}} \mathcal{C}(-\varepsilon_{\alpha,n}) \right. \\ &\quad \left. + D_{\beta,m,n} |H_{\beta,\alpha,m,n}^{(2)}|^2 e^{-j\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n}} \mathcal{C}(-\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n}) \right) \\ &\quad + \sum_{\substack{l=0 \\ l \neq m}}^{N-1} \left[D_{\beta,l,n} H_{\beta,\alpha,m,n}^{(1)*} H_{\beta,\alpha,m,n}^{(1)} e^{j\varphi_{\alpha,n}} \mathcal{C}(m-l-\varepsilon_{\alpha,n}) \right. \\ &\quad \left. + D_{\beta,l,n} H_{\beta,\alpha,m,n}^{(2)*} H_{\beta,\alpha,m,n}^{(2)} e^{-j\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n}} \mathcal{C}(m-l + (\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n})) \right] \\ &\quad + (e^{j\varphi_{\alpha,n}} H_{\beta,\alpha,m,n}^{(1)} W_{\alpha,m,n}^{(1)} + e^{-j(\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n})} H_{\beta,\alpha,m,n}^{(2)*} W_{\alpha,m,n}^{(2)}) \end{aligned} \tag{22}$$

From (22), the CIR of the proposed adaptive modified PRCC scheme can be defined as follows,

$$\begin{aligned} CIR_{\alpha,proposed}(\varepsilon_{\alpha,n}, \Delta\varepsilon_{\alpha,n}, \varphi_{\alpha,n}, \Delta\varphi_{\alpha,n}) \\ = \frac{|e^{j\varphi_{\alpha,n}} \mathcal{C}(-\varepsilon_{\alpha,n}) + e^{-j(\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n})} \mathcal{C}(-\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n})|^2}{\sum_{l=1}^{N-1} |e^{j\varphi_{\alpha,n}} \mathcal{C}(l - \varepsilon_{\alpha,n}) + e^{-j(\varphi_{\alpha,n} + \Delta\varphi_{\alpha,n})} \mathcal{C}(l + (\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n}))|^2} \end{aligned} \tag{23}$$

Given $\varepsilon_{\alpha,n}$ and $\Delta\varepsilon_{\alpha,n}$, the optimal phase rotations can be determined by maximizing the CIR,

$$(\varphi_{\alpha,n}^{(opt)}, \Delta\varphi_{\alpha,n}^{(opt)}) = \operatorname{argmax} CIR_{\alpha,PRCC}(\varepsilon_{\alpha,n}, \Delta\varepsilon_{\alpha,n}, \varphi_{\alpha,n}, \Delta\varphi_{\alpha,n}) \tag{24}$$

By solving (24), we can derive the following sufficient condition for the optimal phase rotations,

$$2\varphi_{\alpha,n}^{(opt)} + \Delta\varphi_{\alpha,n}^{(opt)} = -\pi \frac{(2\varepsilon_{\alpha,n} + \Delta\varepsilon_{\alpha,n})(N-1)}{N} \tag{25}$$

So the corresponding solution is as follows,

$$\varphi_{\alpha,n}^{(opt)} = -\pi \frac{\varepsilon_{\alpha,n}(N-1)}{N} \tag{26}$$

$$\Delta\varphi_{\alpha,n}^{(opt)} = -\pi \frac{\Delta\varepsilon_{\alpha,n}(N-1)}{N} \tag{27}$$

Assuming that the phase rotations $\varphi_{\alpha,n}$ and $\Delta\varphi_{\alpha,n}$ have approached the optimal values after sufficient adaptations, the CFOs can be estimated as follows,

$$\hat{\varepsilon}_{\alpha,n} = -\varphi_{\alpha,n} \frac{N}{\pi(N-1)} \quad (28)$$

$$\Delta\hat{\varepsilon}_{\alpha,n} = -\Delta\varphi_{\alpha,n} \frac{N}{\pi(N-1)} \quad (29)$$

3.1 The Cost Function for the Phase Update

The equalization output based on the MMSE criteria can be expressed as follows,

$$\hat{\mathbf{D}}_{\alpha,m,n} = ((\mathbf{H}_{\beta,\alpha,m,n} \mathbf{H}_{\beta,\alpha,m,n}^H + \mathbf{I}\sigma^2)^{-1} \mathbf{H}_{\beta,\alpha,m,n}^H) \mathbf{R}_{\alpha,m,n} \quad (30)$$

Then, the desired signal can be written as

$$\mathbf{d}\mathbf{r}_{\alpha,m,n} = e^{j\varphi_{\alpha,n}} \hat{\mathbf{D}}_{\alpha,m,n} e^{-j(\hat{\varepsilon}_{\alpha,n})n/N} \quad (31)$$

Therefore, we can define the error signal as follows,

$$\mathbf{e}_{\alpha,m,n} = \mathbf{d}\mathbf{r}_{\alpha,m,n} - \mathbf{R}_{\alpha,m,n} \quad (32)$$

If we average the squared error signals of all the subcarriers, the cost function used in the receiver can be defined as follows,

$$J_{\alpha,n}(\varphi_{\alpha,n}, \Delta\varphi_{\alpha,n}, \hat{\varepsilon}_{\alpha,n}, \Delta\hat{\varepsilon}_{\alpha,n}) \equiv E \left\{ |\mathbf{e}_{\alpha,m,n}|^2 \right\} \cong \frac{1}{N} \sum_{m=0}^{N-1} |\mathbf{e}_{\alpha,m,n}|^2 \quad (33)$$

The phase rotation update equation employing normalized block least mean squared (BLMS) algorithm can be described as follows,

$$\begin{bmatrix} \varphi_{\alpha,n+1} \\ \Delta\varphi_{\alpha,n+1} \end{bmatrix} = \begin{bmatrix} \varphi_{\alpha,n} \\ \Delta\varphi_{\alpha,n} \end{bmatrix} + \begin{bmatrix} \mu_1 & 0 \\ 0 & \mu_2 \end{bmatrix} \frac{\nabla J(\varphi_{\alpha,n}, \Delta\varphi_{\alpha,n}, \hat{\varepsilon}_{\alpha,n}, \Delta\hat{\varepsilon}_{\alpha,n})}{\xi + E \left\{ \mathbf{Y}_{\alpha,n}^{(1)H} \mathbf{Y}_{\alpha,n}^{(1)} + \mathbf{Y}_{\alpha,n}^{(2)H} \mathbf{Y}_{\alpha,n}^{(2)} \right\}} \quad (34)$$

where $\mathbf{Y}_{\alpha,n}^{(i)} = [Y_{\alpha,1,n}^{(i)} \ Y_{\alpha,2,n}^{(i)} \ \dots \ Y_{\alpha,N,n}^{(i)}]^T$, μ_i is the step size employed in the i -th paths, ξ is a small positive constant.

3.2 Kalman Channel Estimation

Fig. 3 shows the block diagram of Kalman channel estimation in the proposed receiver. We first serve the mean of the transmitted symbol $\bar{\mathbf{X}} = \mathbf{E}[\mathbf{X}]$ as the measurement matrix \mathbf{M}_n . In fact, we should set the received signal $\mathbf{Y} = \bar{\mathbf{X}}\mathbf{H}_{\beta,\alpha,n} + \mathbf{N}$ as the measurement equation, and the transmitted signal \mathbf{X} is the measurement matrix. But because we do not know about the transmitted signal, we may just employ the mean of the transmitted signal to be the measurement matrix of the proposed Kalman algorithm. We substitute $\bar{\mathbf{X}} = \mathbf{E}[\mathbf{X}]$ into $\mathbf{Y} = \bar{\mathbf{X}}\mathbf{H}_{\beta,\alpha,n} + \mathbf{N}$, then we serve the \mathbf{N} as the measurement noise \mathbf{v}_n . Finally we can obtain a new measurement equation $\mathbf{Y} = \bar{\mathbf{X}}\mathbf{H}_{\beta,\alpha,n} + \mathbf{v}_n$. The Kalman algorithm has two equations, the process equation and the measurement equation shown as follows,

$$\mathbf{H}_{\beta,\alpha,n+1} = \mathbf{F}_{n+1} \mathbf{H}_{\beta,\alpha,n} + \mathbf{p}_n \quad (35)$$

$$\mathbf{Y} = \mathbf{M}_n \mathbf{H}_{\beta,\alpha,n} + \mathbf{v}_n \quad (36)$$

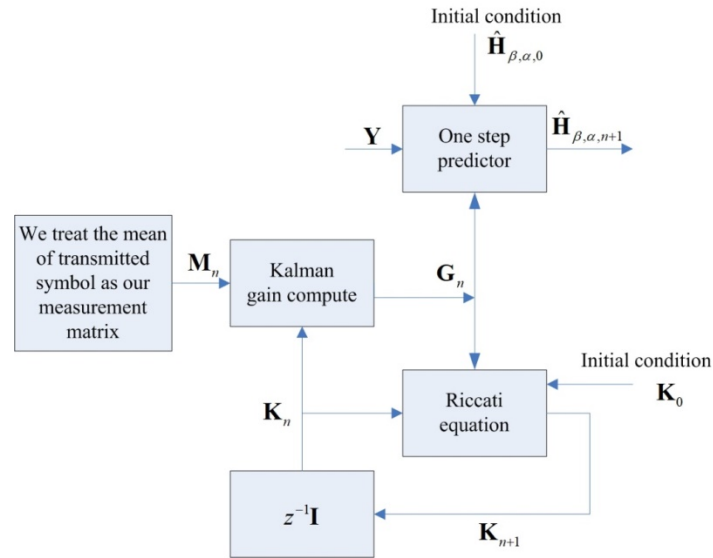


Fig. 3. Block diagram of Kalman channel estimation

The above measurement equation can be rewritten as follows,

$$\hat{\mathbf{Y}} = \mathbf{M}_n \hat{\mathbf{H}}_{\beta,\alpha,n} + \mathbf{v}_n \tag{37}$$

Based on the following assumptions

$$E[\mathbf{p}_n \mathbf{p}_m^H] = \begin{cases} \mathbf{p}_n & n = m \\ \mathbf{0} & n \neq m \end{cases} \quad E[\mathbf{Y}_m \mathbf{v}_n^H] = \mathbf{0}, \quad 0 \leq m \leq n-1$$

$$E[\mathbf{v}_n \mathbf{v}_m^H] = \begin{cases} \mathbf{Q}_n & n = m \\ \mathbf{0} & n \neq m \end{cases} \quad E[\mathbf{Y}_m \mathbf{p}_m^H] = \mathbf{0}, \quad 0 \leq m \leq n$$

$$E[\mathbf{p}_n \mathbf{v}_m^H] = \mathbf{0}$$

The following estimate can be obtained

$$\hat{\mathbf{Y}} = \mathbf{M}_n \hat{\mathbf{H}}_{\beta,\alpha,n} \tag{38}$$

We further compute the innovation vector that can be described as follows,

$$\alpha_n = \mathbf{Y} - \mathbf{M}_n \hat{\mathbf{H}}_{\beta,\alpha,n} \tag{39}$$

So we obtain

$$\alpha_n = \mathbf{M}_n \mathbf{e}_n + \mathbf{v}_n \tag{40}$$

where $\mathbf{e}_n = \mathbf{H}_{\beta,\alpha,n} - \hat{\mathbf{H}}_{\beta,\alpha,n}$ is the estimation error. This estimate can be expressed as a linear combination of the sequence of the innovation process. That is to say, we obtain

$$\hat{\mathbf{H}}_{\beta,\alpha,i} = \sum_{r=1}^n \mathbf{B}_{i,r} \alpha_r \tag{41}$$

Using the assumption that $E[\mathbf{e}_{i,n} \alpha_j^H] = \mathbf{0}, j = 1, \dots, n$, and using $E[\mathbf{p}_n \mathbf{p}_m^H] = \begin{cases} \mathbf{R}_n & n = m \\ \mathbf{0} & n \neq m \end{cases}$, We can derive

$$E[\mathbf{H}_{\beta,\alpha,i} \alpha_j^H] = \mathbf{B}_{i,j} \mathbf{R}_j \tag{42}$$

Then, we multiply the inverse of \mathbf{R}_j for the both side of (42), and then we obtain

$$\mathbf{B}_{i,j} = E[\mathbf{H}_{\beta,\alpha,i}\alpha_j^H][\mathbf{R}_j]^{-1} \quad (43)$$

Substituting (43) into (41), we can derive

$$\hat{\mathbf{H}}_{\beta,\alpha,i} = \sum_{r=1}^{n-1} E[\mathbf{H}_{\beta,\alpha,i}\alpha_r^H][\mathbf{R}_r]^{-1}\alpha_r + E[\mathbf{H}_{\beta,\alpha,i}\alpha_n^H][\mathbf{R}_n]^{-1}\alpha_n \quad (44)$$

Setting $i = n+1$, the following equation can be derived

$$\hat{\mathbf{H}}_{\beta,\alpha,n+1} = \sum_{r=1}^{n-1} E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_r^H][\mathbf{R}_r^{(u)}]^{-1}\alpha_r + E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_n^H][\mathbf{R}_n]^{-1}\alpha_n \quad (45)$$

The above equation indicate the relation between $\mathbf{H}_{\beta,\alpha,n+1}$ and $\mathbf{H}_{\beta,\alpha,n}$, and we can derive the following equation

$$E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_r^H] = \mathbf{F}_{n+1}E[\mathbf{H}_{\beta,\alpha,n}\alpha_r^H] \quad (46)$$

The summation term of (45) can be rewritten as follows,

$$\sum_{r=1}^{n-1} E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_r^H][\mathbf{R}_r]^{-1}\alpha_r^{(u)} = \mathbf{F}_{n+1}\hat{\mathbf{H}}_{\beta,\alpha,n} \quad (47)$$

Define the gain as follows,

$$\mathbf{G}_n = E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_n^H][\mathbf{R}_n]^{-1} \quad (48)$$

The following equations can be derived,

$$E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_n^H] = \mathbf{F}_{n+1}E[\mathbf{H}_{\beta,\alpha,n}\alpha_n^H] = \mathbf{F}_{n+1}E[\mathbf{H}_{\beta,\alpha,n}\mathbf{e}_n^H]\mathbf{M}_n^H \quad (49)$$

$$E[\mathbf{H}_{\beta,\alpha,n+1}\alpha_n^H] = \mathbf{F}_{n+1}E[\mathbf{e}_n\mathbf{e}_n^H]\mathbf{M}_n^H = \mathbf{F}_{n+1}\mathbf{K}_n\mathbf{M}_n^H \quad (50)$$

Finally, we obtain the gain as follows,

$$\mathbf{G}_n = \mathbf{F}_{n+1}\mathbf{K}_n\mathbf{M}_n^H[\mathbf{R}_n]^{-1} \quad (51)$$

where $\mathbf{K}_n = E[\mathbf{e}_n\mathbf{e}_n^H]$ is error correlation matrix, $\mathbf{R}_n = E[\alpha_n\alpha_n^H]$ is correlation matrix of innovation vector. After some manipulations, \mathbf{R}_n can be derived as

$$\mathbf{R}_n = \mathbf{M}_n\mathbf{K}_n\mathbf{M}_n^H + \mathbf{Q}_n \quad (52)$$

Then, using the definition of the gain \mathbf{G}_n , we can rewrite (45) as our estimation equation and can be expressed as follows,

$$\hat{\mathbf{H}}_{\beta,\alpha,n+1} = \mathbf{F}_{n+1}\hat{\mathbf{H}}_{\beta,\alpha,n} + \mathbf{G}_n\alpha_n \quad (53)$$

The following procedure is to derive the equation of the error correlation matrix. We know that

$$\mathbf{e}_{n+1} = \mathbf{H}_{\beta,\alpha,n+1} - \hat{\mathbf{H}}_{\beta,\alpha,n+1} \quad (54)$$

So we can obtain

$$\mathbf{e}_{n+1} = \mathbf{F}_{n+1}[\mathbf{H}_{\beta,\alpha,n} - \hat{\mathbf{H}}_{\beta,\alpha,n}] - \mathbf{G}_n[\mathbf{Y} - \mathbf{M}_n\hat{\mathbf{H}}_{\beta,\alpha,n}] + \mathbf{p}_n \quad (55)$$

Next, based on (36), we get

$$\mathbf{e}_n = [\mathbf{F}_{n+1} - \mathbf{G}_n\mathbf{M}_n]\mathbf{e}_n + \mathbf{p}_n - \mathbf{G}_n\mathbf{v}_n \quad (56)$$

Substituting (56) into $\mathbf{K}_n = \mathbf{E}[\mathbf{e}_n \mathbf{e}_n^H]$, we obtain

$$\mathbf{K}_{n+1} = [\mathbf{F}_{n+1} - \mathbf{G}_n \mathbf{M}_n] \mathbf{K}_n [\mathbf{F}_{n+1} - \mathbf{G}_n \mathbf{M}_n]^H + \mathbf{P}_n + \mathbf{G}_n \mathbf{Q}_n \mathbf{G}_n^H \quad (57)$$

Finally, we employ the gain and the correlation matrix of the innovation vector to obtain the Riccati equation. So the estimated error correlation matrix can be computed as follows,

$$\mathbf{K}_{n+1} = \mathbf{F}_{n+1} \mathbf{K}_n \mathbf{F}_{n+1}^H + \mathbf{P}_n \quad (58)$$

$$\mathbf{K}_n = \mathbf{K}_{n-1} - \mathbf{F}_n \mathbf{G}_n \mathbf{M}_n \mathbf{K}_{n-1} \quad (59)$$

3.3 Turbo equalization

The prior LLR $L(\mathbf{D}^{(n_r)})$ is used for computing the mean and variance of the transmitted symbol described as follows, [32]-[35]

$$\begin{aligned} \bar{\mathbf{D}}^{(n_r)} &= \sum_{X^{(u)} \in \beta} \mathbf{D}^{(n_r)} P(\mathbf{D}^{(n_r)} = D) = P(\mathbf{D}^{(n_r)} = +1) - P(\mathbf{D}^{(n_r)} = -1) \\ &= \frac{e^{L(\mathbf{D}^{(n_r)})}}{1 + e^{L(\mathbf{D}^{(n_r)})}} - \frac{1}{1 + e^{L(\mathbf{D}^{(n_r)})}} = \tanh(L(\mathbf{D}^{(n_r)})/2) \end{aligned} \quad (60)$$

$$v = \sum_{X^{(u)} \in \beta} |\mathbf{D}^{(n_r)} \bar{\mathbf{D}}^{(n_r)}|^2 p(D = D) = 1 - |\bar{\mathbf{D}}|^2 \quad (61)$$

The mean and variance of the transmitted symbol and the output signal of the multiuser detector $\tilde{\mathbf{Y}}_u$ are the input signal of the MMSE equalizer. The turbo equalizer then final output the estimate of the transmitted symbol and extrinsic LLR. As we know, the extrinsic LLR is equal to a posteriori LLR minus a prior LLR as follows,

$$L_E(D) = \ln \frac{p(D = +1|\hat{D})}{p(D = -1|\hat{D})} - \ln \frac{p(D = +1)}{p(D = -1)} = \ln \frac{p(\hat{D}|D = +1)}{p(\hat{D}|D = -1)} \quad (62)$$

Here for reducing the computational complexity, we assume that the probability density function (PDF) $p(\hat{X}^{(u)}|X^{(u)} = X)$ is Gaussian distribution with parameters

$$p(\hat{D}|D = D) \approx \phi\left(\frac{(\hat{D} - \mu_D)}{\sigma_D}\right) / \sigma_D \quad (63)$$

where $\mu_D = E(\hat{D}|D = D)$ is the mean of the Gaussian distribution, $\sigma_D^2 = Cov(\hat{D}, \hat{D}|D = D)$ is the variance of the Gaussian distribution. In order to perform the exact implementation of the MMSE equalizer, we need to compute the extrinsic LLRs, the estimates of the transmitted symbols, and the coefficients of the equalizer. Then given the separated signal of each receiver antenna $\mathbf{Y} = [\tilde{\mathbf{Y}}_1 \quad \tilde{\mathbf{Y}}_2 \quad \dots \quad \tilde{\mathbf{Y}}_{N_t}]^T$, the channel estimation matrix

$$\hat{\mathbf{H}}_{n_r, n_t} = \begin{bmatrix} \hat{\mathbf{H}}_{1,1} & \hat{\mathbf{H}}_{1,2} & \dots & \hat{\mathbf{H}}_{1,N_t} \\ \hat{\mathbf{H}}_{2,1} & \hat{\mathbf{H}}_{2,2} & \ddots & \hat{\mathbf{H}}_{2,N_t} \\ \vdots & \vdots & \ddots & \vdots \\ \hat{\mathbf{H}}_{N_r,1} & \hat{\mathbf{H}}_{N_r,2} & \dots & \hat{\mathbf{H}}_{N_r,N_t} \end{bmatrix},$$

the mean matrix of the transmitted symbol $\bar{\mathbf{D}} = [\bar{\mathbf{D}}_1 \quad \bar{\mathbf{D}}_2 \quad \dots \quad \bar{\mathbf{D}}_{N_t}]^T$ and the variance matrix of the transmitted symbol $\mathbf{V} = \text{Diag}(v_1, v_2, \dots, v_{N_t})$, the following equation

$$\hat{\mathbf{D}} = E(\mathbf{D}) + \text{Cov}(\mathbf{D}, \mathbf{Y}) \text{Cov}(\mathbf{Y}, \mathbf{Y})^{-1} (\mathbf{Y} - E(\mathbf{Y})) \quad (64)$$

It can be rewritten as follows,

$$\hat{\mathbf{D}} = \bar{\mathbf{D}} + \mathbf{V} \tilde{\mathbf{H}}^H (\sigma_w^2 \mathbf{I} + \tilde{\mathbf{H}} \mathbf{V} \tilde{\mathbf{H}}^H)^{-1} (\mathbf{Y} - \tilde{\mathbf{H}} \bar{\mathbf{D}}) \quad (65)$$

Substituting the above equation into the MMSE equalization $\hat{\mathbf{D}} = \sum \mathbf{C}(\mathbf{Y} - E(\mathbf{Y}))$, and then do partial deviation to the coefficients of the equalizer \mathbf{C} , we can get the equalizer as follows,

$$\mathbf{C} = \left(\sigma_w^2 \mathbf{I} + \tilde{\mathbf{H}}^{(u)} \mathbf{V}^{(u)} \tilde{\mathbf{H}}^H + (\mathbf{I}_{diag} - \mathbf{V}) \right)^{-1} \tilde{\mathbf{H}} \quad (66)$$

The above equation can also be used to compute the mean and variance of the transmitted symbol as follows,

$$\mu_D = E(\hat{\mathbf{D}} | D = D) = D \mathbf{C}^H \quad (67)$$

$$\sigma_D^2 = \text{Cov}(\hat{\mathbf{D}}, \hat{\mathbf{D}} | D = D) = \mathbf{C}^H (\mathbf{I} - \mathbf{C}) \quad (68)$$

We then find the following equation to get the Gaussian distribution described above,

$$L_E(\mathbf{D}) = \ln \frac{\phi \left(\frac{(\hat{\mathbf{D}} - \mu_{+1}) / \sigma_{+1}}{\sigma_{+1}} \right) / \sigma_{+1}}{\phi \left(\frac{(\hat{\mathbf{D}} - \mu_{-1}) / \sigma_{-1}}{\sigma_{-1}} \right) / \sigma_{-1}} = \frac{2 \hat{\mathbf{D}} \mu_{+1}}{\sigma_{+1}^2} = 2 \mathbf{C}^H (\mathbf{Y} - \tilde{\mathbf{H}} \bar{\mathbf{D}}) / (\mathbf{I} - \mathbf{C}) \quad (69)$$

The extrinsic LLR is used for the decoder. For initial step, the a priori information is zero. We substitute this condition into the above equations and the initial values can be described as follows,

$$\hat{\mathbf{D}} = \tilde{\mathbf{H}}^H (\sigma_w^2 \mathbf{I} + \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H)^{-1} \mathbf{Y} \quad (70)$$

$$\mathbf{C}_{NA} = \left(\sigma_w^2 \mathbf{I} + \tilde{\mathbf{H}} \mathbf{V} \tilde{\mathbf{H}}^H + (\mathbf{I}_{diag} - \mathbf{V}) \right)^{-1} \tilde{\mathbf{H}} \Big|_{L(\mathbf{D})=0} = (\sigma_w^2 \mathbf{I} + \tilde{\mathbf{H}} \tilde{\mathbf{H}}^H)^{-1} \tilde{\mathbf{H}} \quad (71)$$

$$L_E(\mathbf{D}) = 2 \mathbf{C}_{NA}^H \mathbf{Y} / (\mathbf{I} - \mathbf{C}_{NA}) \quad (72)$$

4. Simulation Results

In this section, we compare the performances of existing schemes (CC, PRCC, Two stage IQ-Imbalance schemes), optimal solution (known CFO and known channel state information) and the proposed turbo receiver. The adopted channel models are multipath Rayleigh fading and ITU channel models defined in the IEEE 802.11 Working Group [33]. We simulate those schemes in the multipath Rayleigh fading channel model and ITU channel model for vehicular test environments with 6 path taps. In these systems, we use QPSK modulation and 16QAM modulation for MIMO-OFDM systems, and then we apply two antennas at transmitter, and two antennas at receiver. The simulated MIMO-OFDM systems assume there are 256 subcarriers (*i.e.*, FFT size=256), 20 cyclic prefixes (*i.e.*, CP=20), difference of speed (*i.e.*, $V=5, 60, 100, 200, 300$ km/hr), and different frequency offset (*i.e.*, $\varepsilon = 0, 0.15, 0.25$). The turbo encoder employs convolution code, and the code rate is 1/2. The turbo decoder employs MAP decoder.

Fig. 4 shows the BER performance with different iteration for the proposed turbo receiver. The signal format is QPSK modulation, CFO $\varepsilon=0.15$ and the channel is multipath channel. In this figure, we can see that the performances with third and fourth iterations are almost the same, so the third iteration can be chosen in the proposed turbo receiver. **Fig. 5** shows the BER performance with different iteration in the proposed turbo receiver. The signal format is QPSK modulation, CFO $\varepsilon=0.25$ and the channel is multipath channel model. We can see that the performances with third and fourth iterations are almost the same, so the third iteration can be chosen in the proposed turbo receiver. **Fig. 6** shows the BER comparison of the no CFO Solution, CC, PRCC, IQ Imbalance, the optimal scheme, the adaptive modified PRCC schemes and the proposed turbo receiver with QPSK modulation and $\varepsilon=0.15$ under ITU Channel A of 100km/hr. It demonstrates that the performance of the proposed turbo modified receiver is better than those of CC, PRCC and IQ imbalance. The turbo receiver with modified PRCC improve the performance for about 5dB. **Fig. 7** shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme, the adaptive modified PRCC scheme and the proposed turbo receiver with QPSK modulation, $\varepsilon=0.25$ under ITU Channel A with speed 100km/hr. It demonstrates that the performance of the proposed turbo receiver is better than those of CC, PRCC and IQ imbalance. This is because that the developed turbo receiver can track CFOs and update the phase rotations based on the estimated CFOs from the two paths in fast fading channel.

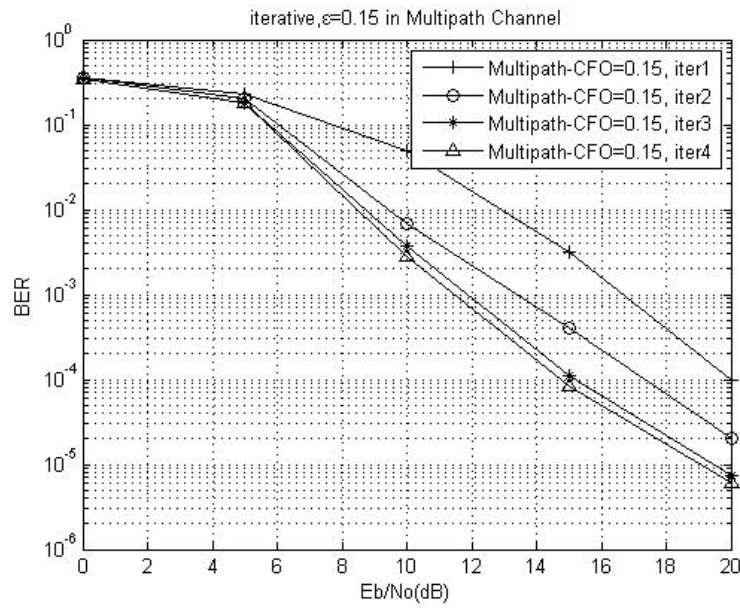


Fig. 4. The BER performance of different iteration for the proposed receiver with QPSK in multipath channel model, CFO=0.15.

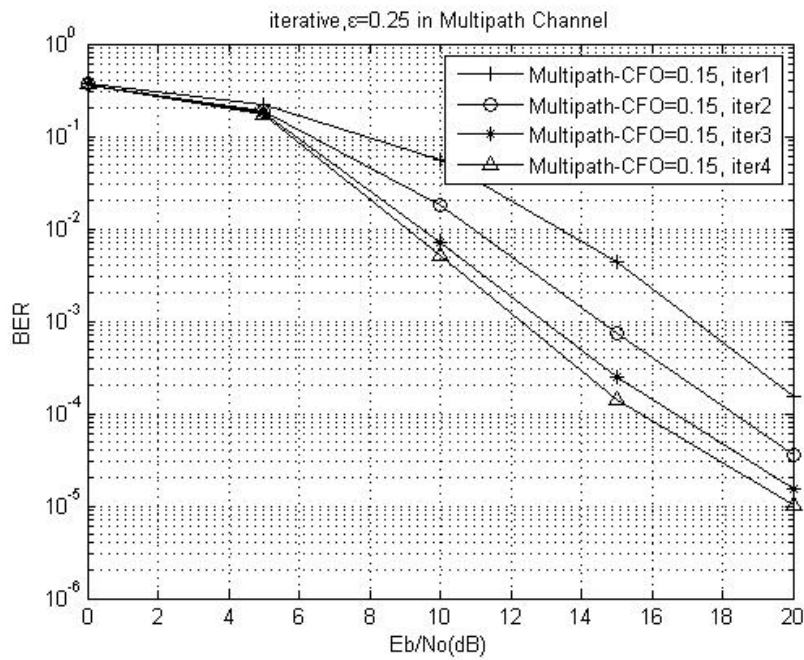


Fig. 5. The BER performance of different iteration for the proposed receiver with QPSK in multipath channel model, CFO=0.25.

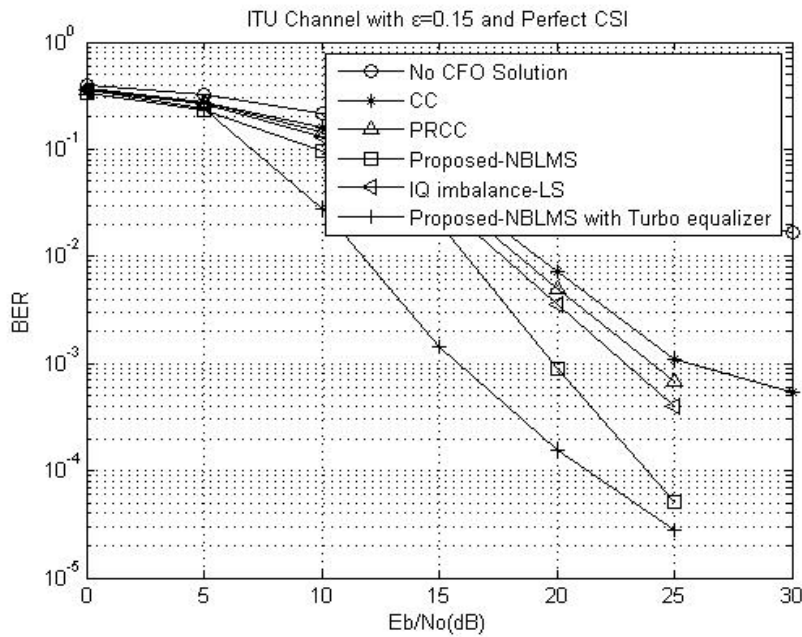


Fig. 6. The BER comparisons of existing methods and the proposed turbo receiver, $\epsilon=0.15$ under ITU Channel A of 100km/hr.

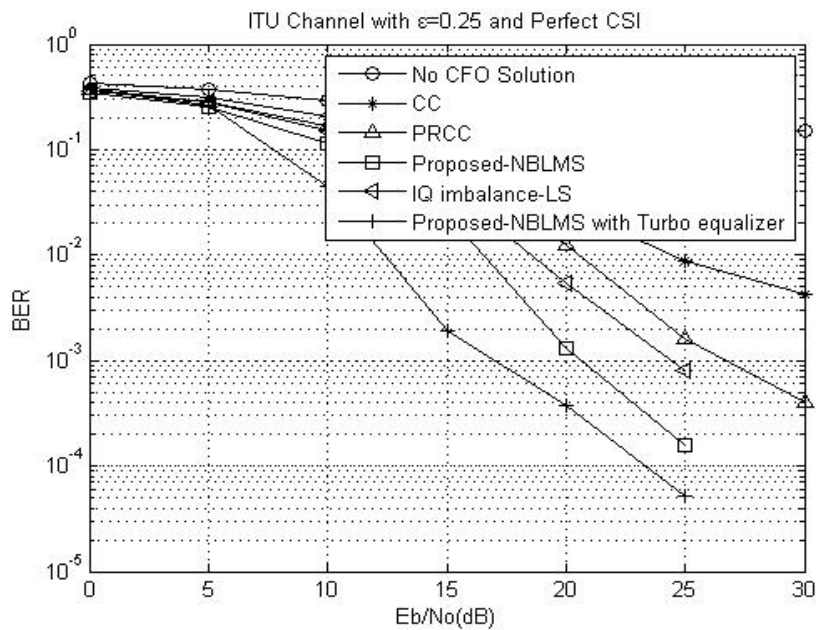


Fig. 7. The BER comparisons of existing methods and the proposed turbo receiver, $\epsilon=0.25$ under ITU Channel A of 100km/hr.

Fig. 8 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed turbo receiver with QPSK modulation, $\epsilon=0.15$ in multipath fading channel with speed 100km/hr. It demonstrates that the performance of the proposed turbo receiver is better than those of CC, PRCC and IQ imbalance. We can see that the excess

delay of ITU channel A is larger than the excess delay of multipath fading channel, and then the coherence bandwidth of multipath fading channel is larger than the coherence bandwidth of ITU channel A, so the performance affected by the multipath fading channel is lower than that by ITU channel A. Fig. 9 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed turbo receiver with QPSK modulation, $\epsilon=0.25$ in multipath fading channel with speed 100 km/hr. It demonstrates that the performance of the proposed turbo receiver better than those of CC, PRCC and IQ imbalance.

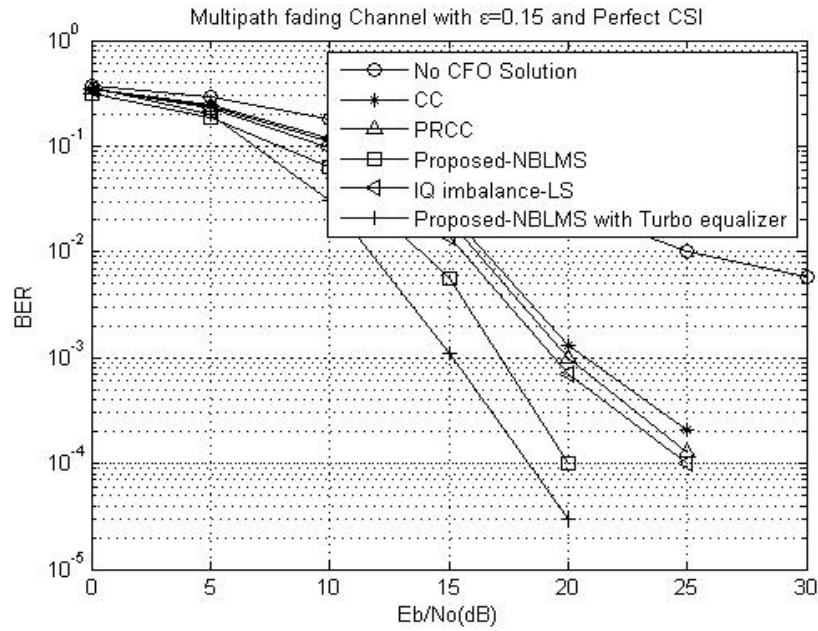


Fig. 8. The BER comparisons of existing methods and the proposed turbo receiver, $\epsilon=0.15$ under multipath fading channel of 100km/hr.

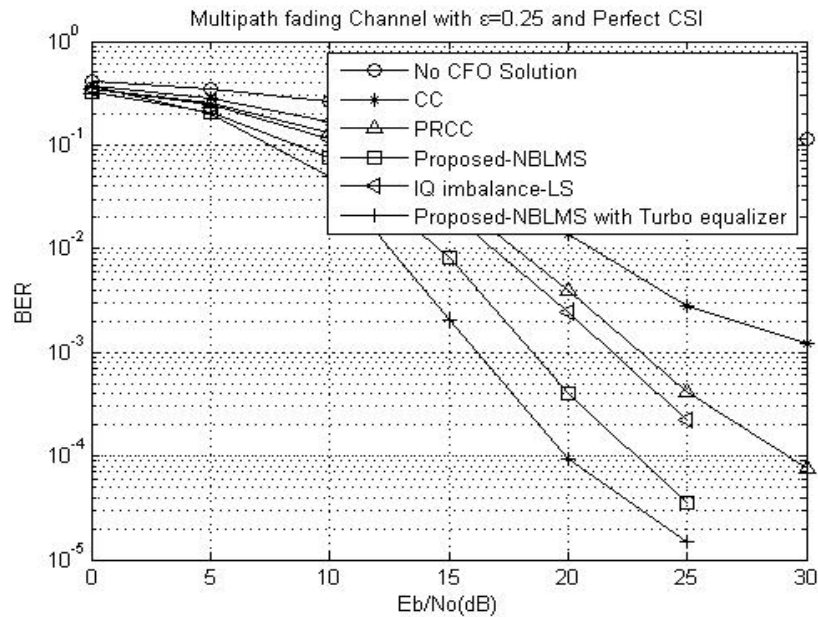


Fig. 9. The BER comparisons of existing methods and the proposed turbo receiver, $\epsilon=0.25$ under multipath fading channel of 100km/hr.

Fig. 10 shows the BER comparison of the no CFO solution, CC, PRCC, IQ Imbalance, the optimal scheme and the proposed turbo receiver with QPSK modulation, $\epsilon=0.25$ in multipath fading channel with speed 100km/hr. The Kalman channel estimation is applied. It demonstrates that the performance of the proposed turbo receiver is better than those of CC, PRCC and IQ imbalance schemes. This is because the developed turbo receiver can track CFOs and update the phase rotations based on the estimated CFOs from the two paths in fast fading channel. **Fig. 11** shows the BER comparison of the existing PRCC scheme and the proposed scheme at different speed with OPSK modulation and $\epsilon = 0.25$ at SNR=25dB. We can see that the proposed scheme outperforms the existing PRCC scheme at different speed. This difference is due to the proposed scheme can track CFOs and update the phase rotations, and the existing PRCC scheme can't bear the large CFO in fast speed.

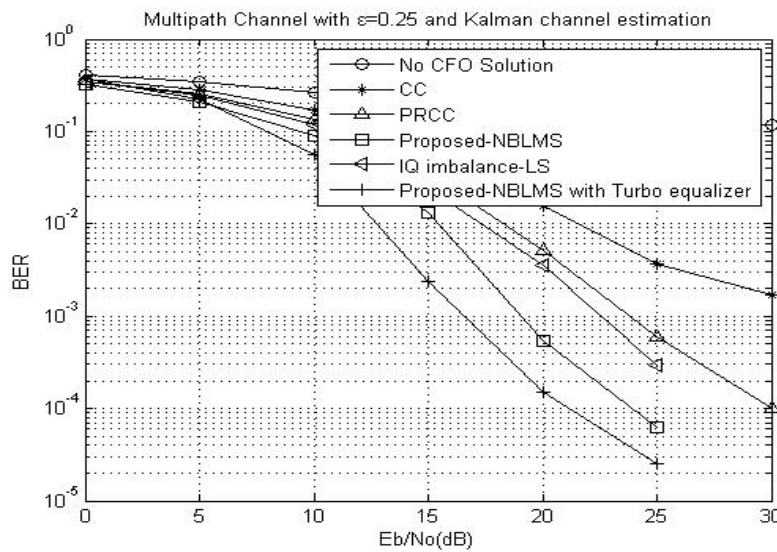


Fig. 10. The BER comparisons of existing methods and the proposed turbo receiver, $\epsilon=0.25$ under Multipath Fading Channel of 100km/hr with Kalman channel estimation.

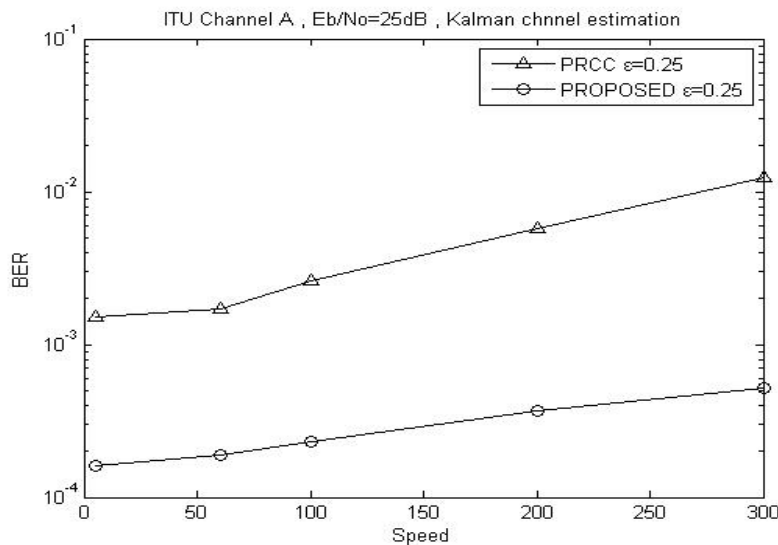


Fig. 11. The BER comparisons the proposed scheme and the existing PRCC scheme under different speed for $\epsilon=0.25$ with QPSK modulation and kalman channel estimation at SNR =25dB.

5. Conclusion

This paper proposes an advanced turbo receiver with ICI self cancellation, Kalman channel estimation and turbo equalization for MIMO-OFDM systems in time-varying channels. We employ an adaptive receiver and construct two-path conjugate transmission that is based on the PRCC concept. Additionally, the Kalman filter can efficiently track the time-varying channels. We also use iterative property of turbo equalization to update the Kalman filter, equalizer and decoder to improve the receiver performance. With such adaptive phase rotations in the receiver, the CFO variations due to the channel effect and the mismatch between oscillators at the transmitter and receiver can effectively be tracked without feeding back the CFO estimate to the transmitter as required in conventional PRCC scheme.

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