

# An Adaptive Inter-Carrier Interference Mitigation Method for MIMO-OFDM Systems in HST Environment

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Abstract—In this paper, we propose an adaptive intercarrier interference (ICI) mitigation algorithm based on phase rotated conjugate cancellation (PRCC) scheme. For the MIMO-OFDM systems in high-speed train (HST) environment, the ICI is caused by the non-stationary time-varying channels. By updating the optimal phases every two OFDM symbols, our method achieves additional optimal rotated phases applicable for HST wireless channels. Simulation results demonstrate that the carrier-to-interference ratio (CIR) increase and the bit error rate (BER) decreases enormously in our method. Therefore, the proposed adaptive algorithm is effective for mitigating the ICI in HST environment comparing with the traditional ICI selfcancellation algorithm.

Keywords—MIMO-OFDM; high-speed train environment; inter-carrier interference; adaptive phase rotated conjugate cancellation

# I. INTRODUCTION

The characteristics of HST communications are different from territorial cellular communications. Due to the ultra-fast movement of terminals, the HST channel are both timevarying and non-stationary [1]. The time-varying HST channel during an OFDM symbol destroys the orthogonality and causes ICI. As a result, the BER of OFDM systems may be degraded significantly. It is obvious that the ICI mitigation for OFDM systems in HST environment is urgent.

So far, a large amount of ICI mitigation methods have been put forward. An extended mixed algorithm of minimum mean square error (MMSE) and successive interference cancellation (SIC) was proposed in [2]. In [3], an iterative algorithm of successive interference suppression (SIS) was proposed. In [4], a combination algorithm of PIC and semi-definite relaxation (SDR) was studied. Progressive parallel interference canceller (PPIC) technology was developed in [5]. The self-cancellation schemes were developed in [6,7]. The main idea of selfcancellation is to process both transmitted signal and received signal to suppress ICI. Recently, ICI self-cancellation scheme through two-path conjugate transmission was studied. As extensions of conjugate cancellation (CC) algorithm [8], phase rotated conjugate cancellation (PRCC) [9] and adaptive PRCC [10] were developed. The adaptive PRCC scheme adjusts the optimal rotated phases to adapt to the channel variations, thus

works well in time-variant channels. In this paper, the ICI mitigation problem for non-stationary time-varying HST channels is explored. Based on the simulated channel model of HST communications in [11], we extend the adaptive PRCC scheme. A general ICI mitigation scheme implemented by updating the optimal rotated phases every two OFDM symbols is established for MIMO-OFDM systems in HST environment.

The rest of the paper is organized as follows. The adaptive ICI mitigation scheme is introduced in Section II. In Section III, simulation results on different methods are discussed, and conclusion is given at last.

#### II. ADAPTIVE ICI MITIGATION SCHEME

## A. System Model

There are strong line-of-sight (LOS) and several non-lineof-sight (NLOS) between base stations and HST. Rice channel is generally used for HST channel model. Based on the nonstationary geometry-based stochastic model (GBSM) in [11], the discrete-time channel impulse response at the p th transmitting antenna and the q th receiving antenna in HST environment can be described as

$$h_{p,q}(n,l) = \begin{cases} h_{p,q}^{LOS}(n,0) + h_{p,q}^{NLOS}(n,0), & (l=0) \\ h_{p,q}^{NLOS}(n,l), & (1 \le l \le L-1) \end{cases}$$
(1)

$$h_{p,q}^{LOS}(n,0) = \sqrt{\frac{K_{pq}}{K_{pq}+1}} e^{-j2\pi f_c \tau_{pq}(n)} \cdot e^{j[2\pi f_{max}nt_s \cos(\phi_p^{LOS}(n) - r_R)]}$$
(2)

$$h_{p,q}^{NLOS}(n,0) = \sqrt{\frac{1}{K_{pq}+1}} \sum_{n_{1}=1}^{S} \frac{1}{\sqrt{S}} e^{j[\psi_{n_{1}}-2\pi f_{c}\tau_{pq,n_{1}}(n)]} \cdot e^{j[2\pi f_{max}nt_{s}\cos(\phi_{R}^{(m)}(n)-r_{R})]}$$
(3)

$$h_{p,q}^{NLOS}(n,l) = \sum_{n_l=1}^{S} \frac{1}{\sqrt{S}} e^{j[\psi_{n_l} - 2\pi f_c \tau_{pq,n_l}(n)]} \cdot e^{j[2\pi f_{max}nf_s \cos(\phi_R^{(m_l)}(n) - r_R)]}$$
(4)

where  $K_{pq}$ ,  $f_c$ ,  $f_{max}$ ,  $t_s$ ,  $r_R$ , S are listed as the Ricean factor, the carrier frequency, the maximum Doppler frequency offset, sample interval, direction of the train, and the number of scatterers, respectively.  $\tau_{pq,n_l}$  is the time for a signal to travel from p th transmitting antenna to q th receiving antenna through  $n_l$  th scatterer.  $\phi_p^{LOS}$  is the angle of arrival (AOA),  $\phi_R^{(n_l)}$  is AOA from  $n_l$  th scatterer to the antenna on the train, and  $\psi_{n_l}$  is the random phase with dependent uniform distribution over  $[-\pi, \pi)$ .

Consider an MIMO-OFDM system with  $M_T$  transmitting antennas and  $M_R$  receiving antennas. An OFDM symbol contains N subcarriers.  $X_p(k)$  denotes the transmitted data at the subcarrier k ( $0 \le k \le N-1$ ). The transmitted signal in the time-domain through IFFT is expressed as

$$x_{p}(n) = \frac{1}{N} \sum_{k=0}^{N-1} X_{p}(k) e^{j\frac{2\pi nk}{N}}, (0 \le n \le N-1).$$
 (5)

The OFDM symbol is transmitted twice continuously at the transmitter. Two paths are received as  $y_q^{(1)}$  and  $y_q^{(2)}$  through the GBSM channel as follows

$$y_q^{(1)}(n) = \sum_{p=1}^{M_T} \sum_{l=0}^{L-1} h_{p,q}^{(1)}(n,l) \, \mathbf{x}_p(n-l) + w_q^{(1)}(n) \tag{6}$$

$$y_{q}^{(2)}(n) = \sum_{p=1}^{M_{T}} \sum_{l=0}^{L-1} h_{p,q}^{(2)}(n,l) \mathbf{x}_{p}^{*}(n-l) + w_{q}^{(2)}(n)$$
(7)

where  $\mathbf{x}_{p}^{*}$  is the conjugate of  $\mathbf{x}_{p}$ , and  $w_{q}(n)$  is the additive white Gaussian noise.

The received signals are added with independent rotated phase  $\phi_1$  and  $\phi_2$  separately, then transformed into frequency domain by FFT. We can obtain the first paths as follows

$$Y_{q}^{(1)}(k) = \sum_{\substack{p=1\\m\neq k}}^{M_{T}} \sum_{l=0}^{N-1} \sum_{l=0}^{L-1} \left[ H_{p,q}^{(1)}(l,k-m)\overline{\varpi}_{l,m} \right] \cdot X_{p}(m) \cdot e^{j\phi_{l}} + \underbrace{\sum_{\substack{m\neq k\\m\neq k}}^{M_{T}} \sum_{l=0}^{L-1} \left[ H_{p,q}^{(1)}(l,0)\overline{\varpi}_{l,k} \right] \cdot X_{p}(k) \cdot e^{j\phi_{l}} + W_{q}^{(1)}(k) \cdot e^{j\phi_{l}}}_{desired signal}$$
(8)

Supposing  $G_{p,q}^{(1)}(k,k) = \sum_{l=0}^{L-1} H_{p,q}^{(1)}(l,0) \overline{\sigma}_{l,k}$ , and

$$G_{p,q}^{(1)}(m,k) = \sum_{l=0}^{L-1} H_{p,q}^{(1)}(l,k-m)\overline{\omega}_{l,m}$$
, we get

$$Y_{q}^{(1)}(k) = \underbrace{\sum_{p=1}^{M_{T}} G_{p,q}^{(1)}(k,k) \cdot X_{p}(k) \cdot e^{j\phi_{1}}}_{desired signal} . (9)$$

$$+ \underbrace{\sum_{p=1}^{M_{T}} \sum_{m=0}^{N-1} G_{p,q}^{(1)}(m,k) \cdot X_{p}(m) \cdot e^{j\phi_{1}}}_{m \neq k} + W_{q}^{(1)}(k) \cdot e^{j\phi_{1}}}_{ICI \ components}$$

Let 
$$Y_{p,q}^{(1)}(k) = G_{p,q}^{(1)}(k,k) \cdot X_p(k) + \sum_{\substack{m=0 \ m \neq k}}^{N-1} G_{p,q}^{(1)}(m,k) \cdot X_p(m)$$
,

then (9) can be written simply as

$$Y_q^{(1)}(k) = \sum_{p=1}^{M_T} \left[ Y_{p,q}^{(1)}(k) + W_q^{(1)}(k) \right] \cdot e^{j\phi_1}, \ (0 \le k \le N - 1) \ . \ (10)$$

Equally, the second path can be written as

$$Y_q^{(2)}(k) = \sum_{p=1}^{M_T} \left[ Y_{p,q}^{(2)}(k) + W_q^{(2)}(k) \right] \cdot e^{-j\phi_2}, \ (0 \le k \le N - 1). \ (11)$$

Furthermore, we merge two paths by maximum ratio combing (MRC). Two paths signals on the k th subcarrier of the q th receiving antenna can be rewritten as

$$R_{q}^{(1)}(k) = \sum_{p=1}^{M_{T}} \left\{ \left[ G_{p,q}^{(1)}(k,k) \right]^{*} \cdot \left[ Y_{p,q}^{(1)}(k) + W_{q}^{(1)}(k) \right] \right\} \cdot e^{j\phi_{1}} \quad (12)$$

$$R_q^{(2)}(k) = \sum_{p=1}^{M_T} \left\{ \left[ G_{p,q}^{(2)}(k,k) \right]^* \cdot \left[ Y_{p,q}^{(2)}(k) + W_q^{(2)}(k) \right] \right\} \cdot e^{-j\phi_2} .$$
(13)

Eventually, the merged output is the actual received signal.

#### B. Calculation of Optimal Rotated Phases

It is noted that the independent rotated phases  $\phi_1, \phi_2$  are crucial to the accuracy of the received signal. Therefore, we



calculate the optimal rotated phases  $\phi_{1opt}, \phi_{2opt}$  by adjusting two rotated phases to maximize the CIR.

CIR is used to evaluate the ICI interfere level to the desired signal. It is defined as the ratio of the desired signal power to the ICI power.

$$CIR = \frac{\left|\sum_{p=1}^{M_{T}} \left[ \left| G_{p,q}^{(1)}(k,k) \right|^{2} \cdot e^{j\phi_{1}} + \left| G_{p,q}^{(2)}(k,k) \right|^{2} \cdot e^{-j\phi_{2}} \right] \right|^{2}}{\left| \sum_{p=1}^{M_{T}} \sum_{\substack{m=0\\m\neq k}}^{N-1} \left\{ \left[ G_{p,q}^{(1)}(k,k) \right]^{*} G_{p,q}^{(1)}(m,k) \cdot e^{j\phi_{1}} \right\} \right|^{2}} .$$
(15)

By calculating the partial derivative of  $\phi_1, \phi_2$  in (16), and making it equivalent to 0, we can obtain  $\phi_{1opt}, \phi_{2opt}$  as follows

$$\begin{cases} \partial CIR/\partial \phi_1 = 0 \Longrightarrow \phi_{lopt} \\ \partial CIR/\partial \phi_2 = 0 \Longrightarrow \phi_{2opt} \end{cases}.$$
(16)

Suppose

$$\begin{split} A_{d}^{(1)} &= \sum_{p=1}^{M_{T}} \left| G_{p,q}^{(1)}(k,k) \right|^{2}, \ A_{lcl}^{(1)} e^{j\phi_{lcl}^{(1)}} = \sum_{p=1}^{M_{T}} \sum_{\substack{m=0\\m\neq k}}^{N-1} \left[ G_{p,q}^{(1)}(k,k) \right]^{*} G_{p,q}^{(1)}(m,k) \quad , \\ A_{d}^{(2)} &= \sum_{p=1}^{M_{T}} \left| G_{p,q}^{(2)}(k,k) \right|^{2}, \ A_{lcl}^{(2)} e^{j\phi_{lcl}^{(2)}} = \sum_{p=1}^{M_{T}} \sum_{\substack{m=0\\m\neq k}}^{N-1} \left[ G_{p,q}^{(2)}(k,k) \right]^{*} G_{p,q}^{(2)}(m,k) \quad , \end{split}$$

then (16) can be rewritten as

$$CIR = \frac{\left|A_{d}^{(1)}e^{j\phi_{1}} + A_{d}^{(2)}e^{-j\phi_{2}}\right|^{2}}{\left|A_{ici}^{(1)}e^{j\phi_{ci}^{(1)}}e^{j\phi_{1}} + A_{ici}^{(2)}e^{j\phi_{ci}^{(2)}}e^{-j\phi_{2}}\right|^{2}} = \frac{\left(A_{d}^{(1)}\right)^{2} + \left(A_{d}^{(2)}\right)^{2} + 2A_{d}^{(1)}A_{d}^{(2)}\cos\left(\phi_{1} + \phi_{2}\right)}{\left(A_{ici}^{(1)}\right)^{2} + \left(A_{ici}^{(2)}\right)^{2} + 2A_{ici}^{(1)}A_{ici}^{(2)}\cos\left(\phi_{ici}^{(1)} - \phi_{ici}^{(2)} + \phi_{1} + \phi_{2}\right)}.$$
(17)

Let 
$$A_{d,1} = (A_d^{(1)})^2 + (A_d^{(2)})^2$$
,  $A_{d,2} = A_d^{(1)} A_d^{(2)}$ ,  
 $A_{ici,1} = (A_{ici}^{(1)})^2 + (A_{ici}^{(2)})^2$ ,  $A_{ici,2} = A_{ici}^{(1)} A_{ici}^{(2)}$ ,  $\phi_{ici} = \phi_{ici}^{(1)} - \phi_{ici}^{(2)}$ ,

CIR can be calculated shortly as

$$CIR = \frac{A_{d,1} + 2A_{d,2}\cos(\phi_1 + \phi_2)}{A_{ici,1} + 2A_{ici,2}\cos(\phi_{ici} + \phi_1 + \phi_2)} = \frac{a(\phi_1 + \phi_2)}{b(\phi_1 + \phi_2)}.$$
 (18)

Supposing the partial derivative of  $\phi_1$  in (19) is 0, we get

$$\begin{bmatrix} A_{d,1}A_{ici,2}\cos(\phi_{ici}) - A_{d,2}A_{ici,1} \end{bmatrix} \cdot \sin(\phi_1 + \phi_2) + A_{d,1}A_{ici,2}\sin(\phi_{ici}) \cdot \cos(\phi_1 + \phi_2) + 2A_{d,2}A_{ici,2}\sin(\phi_{ici}) = 0$$
(19)

Let  $A' = A_{d,1}A_{ici,2}\cos(\phi_{ici}) - A_{d,2}A_{ici,1}$ ,  $A'' = A_{d,1}A_{ici,2}\sin(\phi_{ici})$ ,  $A'' = 2A_{d,2}A_{ici,2}\sin(\phi_{ici})$ , then (20) is calculated as

$$\sqrt{(A')^{2} + (A'')^{2}} \sin(\phi_{1} + \phi_{2} + \phi_{A}) + A'' = 0, \text{ where } \phi_{A} = \arctan\frac{A'}{A'}.$$
  
Let  $\phi_{0} = \arcsin\left(-\frac{A''}{\sqrt{(A')^{2} + (A'')^{2}}}\right) - \phi_{A}$ , we have  
 $\phi_{1} + \phi_{2} = \phi_{0}.$  (20)

In order to guarantee the coefficient of desired signal  $R_q(k)$  is real, no additional phase is added. That means  $A_d^{(1)} \cdot e^{j\phi_1} + A_d^{(2)} \cdot e^{-j\phi_2}$  equals to real value, which is  $A_d^{(1)} \cos \phi_1 + A_d^{(2)} \cos \phi_2 + j \left[ A_d^{(1)} \sin \phi_1 - A_d^{(2)} \sin \phi_2 \right]$  equals to real value. Thus, we have

$$\begin{cases} A_d^{(1)} \cos \phi_1 + A_d^{(2)} \cos \phi_2 \neq 0\\ A_d^{(1)} \sin \phi_1 - A_d^{(2)} \sin \phi_2 = 0 \end{cases}.$$
 (21)

Combining (17), (22) and (23), we can obtain the optimal rotated phases as follows

$$\begin{cases} \phi_{1opt} = \phi_B \\ \phi_{2opt} = \phi_0 - \phi_B \end{cases}$$
(22)

where 
$$\phi_B = \arctan\left(\frac{A_d^{(2)}\sin\phi_0}{A_d^{(1)} + A_d^{(2)}\cos\phi_0}\right).$$

In order to reduce the computational complexity,  $\phi_{1opt}, \phi_{2opt}$  are updated every 2N data samples to fit the twopath transmission period. The rotated phases are updated accordingly during the whole period of the continuous receiving of OFDM symbols.

### **III.** SIMULATIONS

In this section, we present CIR and BER performance comparisons of the traditional ICI self-cancellation scheme in [6], denoted by Zhao's scheme, the standard MIMO-OFDM system and our proposed PRCC scheme with the HST channel model. The system parameters are listed in Table I.

TABLE I. SIMULATION SYSTEMS PARAMETERS

Modulation	Carrier	Subcarrier	CP	Sampling	Ricean	Multipath	Train	Simulation
method	frequency	number	length	interval	factor	number	speed	times
QPSK	2.4GHz	512	16	$0.8 \ \mu s$	10dB	6	50~500km/h	50

Comparisons of the desired signal power and ICI power, and related CIRs at different speed of train for Zhao's scheme, the standard MIMO-OFDM system and the proposed PRCC scheme are shown in Figure I and Figure II. In Figure I, the faster the train goes, the lower the received power of desired signal is, but the higher the power of ICI is. The desired signal power of proposed scheme is higher than that of standard MIMO-OFDM system when the train speed is lower than 400km/h. When the train is faster than 400km/h, the desired signal power of proposed PRCC scheme is lower than that of standard MIMO-OFDM systems, but ICI power reduces in our proposed scheme. Therefore, CIR can remain matter-of-factly in Figure II by proposed scheme. It is proved that the CIR is greatly improved by our proposed PRCC scheme, while Zhao's scheme is not a good choice for HST environment.



FIGURE I. THE CHANGE OF DESIRED SIGNAL POWER AND ICI POWER WITH DIFFERENT TRAIN SPEED



FIGURE II. THE CHANGE OF CIR WITH DIFFERENT TRAIN SPEED

The comparison of BER performances for proposed PRCC scheme and standard MIMO-OFDM system are

demonstrated in Figure III and Figure IV. In Figure III, SNR is set to 30dB, and the proposed scheme can effectively reduce BER comparing with standard MIMO-OFDM system. In Figure IV, BER performance is analysed with different SNR and train speed. The BER decrease at the rate of 95.4%, 90.0% and 66.7% at the speed of 300km/h, 400km/h and 500km/h, respectively. It is verified that the proposed PRCC scheme is suitable for mitigating ICI in HST environment.



FIGURE III. THE CHANGE OF BER WITH DIFFERENT TRAIN SPEED



FIGURE IV. THE CHANGE OF BER WITH DIFFERENT SNR

#### IV. CONCLUSION

In this paper, an adaptive ICI mitigation method based on the PRCC scheme for HST channels in MIMO-OFDM systems was proposed. The algorithm was implemented by updating optimal rotated phases every two OFDM symbols.



Simulation results show that the proposed method outperforms the traditional method. The CIR increases, and the BER decreases enormously comparing to those in the traditional ICI self-cancellation algorithm and standard MIMO-OFDM systems at different train speed. It is verified that the proposed method is suitable for solving the ICI problem in HST environment.

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