ICI Mitigation by Estimation of Double Carrier Frequency Offsets in High-Speed-Railway Communication Systems for Smart Cities

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Abstract. OFDM is substantial against Inter-symbol-interference due to long symbol duration. However, inter-carrier interference (ICI) caused by high Doppler frequency shift has a severe impact on OFDM in case of high channel variations. In this paper, we propose an ICI mitigation method by utilizing the estimation and pre-compensation of high Doppler shifts in high-speed railway communication systems for smart cities. The estimate of the Doppler shift is based on a preamble frame of data in communication link between EnodeB and user equipment. The simulation results show that the performance of system has been improved using the proposed model.

Keywords: Doppler effect · CFO · OFDM · HSR · ICI · Smart cities

1 Introduction

In smart cities, the means of intelligent transportations need to be connected to a huge communication network to transfer the real-time properties of transportation as well as the control signals, which manages autonomous vehicles in order to reduce urban traffic congestions, to detect accidents and to avoid crashes [1–3]. In high speed railway (HSR), the demands for high-speed data connections of passengers on board as Internet, video calls, online TV are increasing. But the global system for mobile communications-railway (GSM-R) based on GSM technology does not meet the increasing requirements of broadband communications for supervising, controlling the train operation and services for passengers [4]. Currently, there are a lot of multimodal systems supplying broadband communications for HSR. One of them is the global broadband multimodal radio systems (MOWGLY) in Europe, which provides WiFi signal for passengers on board, the satellite signal for communications between train and the ground with a big delay. It is also used on Thalys and TGV East to supply broadcast internet [5]. With the speed of train is 300 kph, the maximum speed of data downlink and uplink are 4 Mbps and 2 Mbps, respectively [6]. Although current technologies temporarily meet the demands for HSR radio communications, they do

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not meet the requirements for the fastest speed trains, high-speed data, real-time and reliable communications. According to recent studies, the required data rate for a train in HSR is around 37.5 Mbps and over 0.5–5 Gbps in the near future [7]. 4G long term evolution (LTE) standard has been widely investigated to satisfy the HSR communication system needs as well as for the train communication and the transmission of train control data [8].

The main core of 4G LTE standard in radio access network is orthogonal frequency division multiplexing (OFDM). OFDM provides high bandwidth efficiency and reliable signal transmission by dividing the available spectrum into group of closely spaced orthogonal sub-carriers, instead of transmitting a high data stream with a single carrier. Nevertheless, a series of difficulties caused by high movement speed such as over-frequent handover and high Doppler shift need to be solved urgently [9]. Particularly, when the OFDM based systems are considered for HSR communications, the Doppler frequency shift compensation techniques are vital due to the fact of that the OFDM systems are vulnerable to the carrier frequency offset (CFO) which may introduce the inter-carrier interference (ICI) to the system. Therefore, CFO problem plays a crucial role in OFDM systems for HSR communications. In the single input single output (SISO) or the co-located multiple input multiple output (MIMO) OFDM systems, there is only one CFO between the transceivers and a lot of researches on CFO approximation have been carried out with both data-aided approach and blind approach [10–16]. On the other side, when distributed transmitters transmit data streams at the same time, multiple CFOs will be obtained at the receiver, making the CFO approximation and compensation more perplexing. A well-known scenario is the so-called orthogonal frequency division multiple access (OFDMA), in which the subcarriers are completely occupied by the many users. Throughout the past few years, several multiuser CFO estimation techniques for OFDMA uplink have been proposed [17, 18], where different CFOs could either be estimated from an iterative methods or be detected from some subspace schemes. There are very few researches on multiple CFO estimation for multiuser OFDM transmission. Besson and Stoica made the first trial in [19] while limited their simulation only for flat fading channels. In [20], a semi-blind scheme was suggested to instantaneously approximate the multiple CFOs and channels in frequency selective fading channels. However, the technique is only effective for zero-padding (ZP) OFDM, as ZP can make the channel estimation and equalization simpler. A joint CFO and channel estimation for multiuser cyclic-prefix (CP) MIMO-OFDM systems based on the maximum likelihood (ML) criterion was introduced [21]. The high-complex multi-dimensional search is reduced from the importance sampling technique. However, the remaining complexity to create sufficient samples for importance sampling may still be large for real application implementation. A suboptimal estimation algorithm was presented in [22]. More recently, the authors of [23] introduced a semi-blind multi-CFO approximation and an independent component analysis based equalization scheme for multiuser coordinated multi-point OFDM systems. In this paper, the problem of double CFOs is investigated. The contributions of this paper is that the estimation of CFOs with only one received antenna rather than many antennas as shown in [19–23] and without the implement of training sequence as in [23]. We also show that our proposed method outperforms the conventional estimation method in [24] at high bit energy to noise ratio.

2 System Model

The HSR system that we investigate is a two-hoop network structure formed by distributed eNODE-Bs as illustrated in Fig. 1, in which d1 and d2 are distance from e-NODE-B1 and e-NODE-B2 to Mobile Station respectively, d_B is the perpendicular distance of E-NODE-B and railway, d_S is the coverage radius of one E-NODE-B and Δd is the overlap distance of two coverage area. Here we primarily examine the communication of E-NODE-B-MS links. Each E-NODE-B can be designed with one or more antennas. We make the assumption that the coverage region of each E-NODE-B forms a logic cell, and each E-NODE-B has one antenna and the MS also has one antenna. All the antennas have no correlation with each other. The Doppler frequency shifts are our main research objective, so we assume the system has been synchronized in time and oscillator frequency.

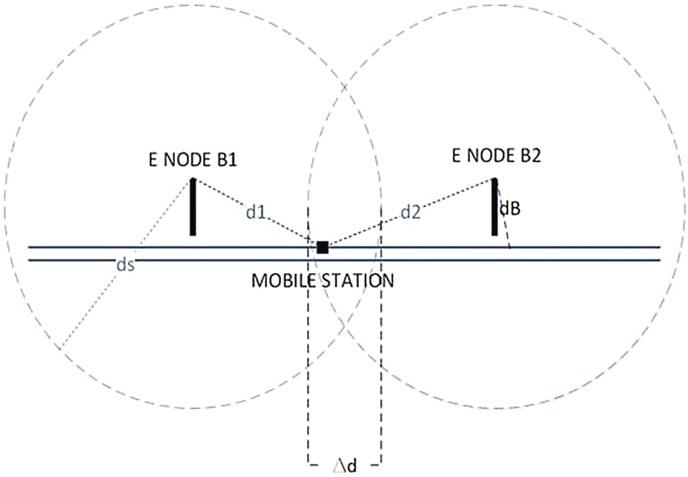

Fig. 1. Communication scenario in HSR

We assign time-frequency resource to the two adjacent E-NODE-Bs just as the two-TX scenario in downlink system in [22]. Desired signals from the two E-NODE-Bs reach the MS almost instantaneously. The two signal streams can not be detached easily because they occupy the same time-frequency slot in the OFDM communication link. Consequently, it is still problematic to compensate Doppler frequency offsets precisely at the receiver. Our proposed solution is given as in Fig. 2. The MS estimates Doppler frequency shifts based on the received data and then feeds the estimated value back to E-NODE-B. When the new data is sent, the frequency offset pre-compensation is made for the data. It is worth to note that the prediction according to variation of Doppler frequency shift is also possible to further enhance the effect of pre-compensation. We also have to point out that error would be presented because of the feedback delay produced by the fast time-varying channel.

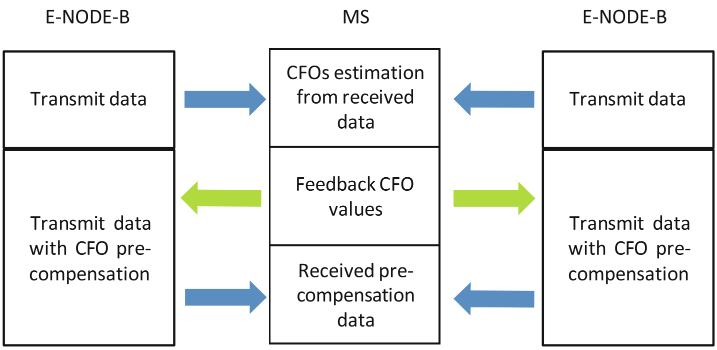

Fig. 2. Scheme for pre-compensation CFOs communication

As in Figs. 1 and 2, we can see that the received signal is the mixture of two signals from two separated E-NODE-Bs. As the MS is moving away from E-NODE-B1 and toward E-NODE-B2, the Doppler frequency shift of these two signals are undoubtedly different. In this paper, through mathematical models of channel and communication scenarios, we find out the two frequency shifts based on a preamble frame of data in communication link between E-NODE-Bs and MS.

3 Theoretical Analysis

3.1 Mathematical Model

In OFDM, multiple sinusoidal with frequency separation $1/T_s$ is used where T_s is the symbol duration.

Denote:

N: Number of subcarriers
P: Number of slots for data in N subcarriers (P < N)
s(k): the k^{th}OFDM symbol.

We have:

$$s(k) = [s_1(k)\, s_2(k) \ldots s_P(k)]^T \tag{1}$$

The OFDM modulation is performed by applying an inverse discrete fourier transform (IDFT) operator to the data stream. Using matrix representation, the resulting N-point time domain signal is given by:

$$x(k) = [x_1(k)\, x_2(k) \ldots x_N(k)]^T = \boldsymbol{W}_P s(k) \tag{2}$$

where $\boldsymbol{W_P}$ is a first P column of the $N \times N$ IDFT matrix $\boldsymbol{W}$:

$$\boldsymbol{W} = IDFT(N) = [w_l \ldots w_N] \tag{3}$$

$$\boldsymbol{W}_P = [w_1 \ldots w_P] \tag{4}$$

In practice, to prevent transmit filtering in OFDM system, some subcarriers are not modulated. In other words, the number of sub-channels carrying the information is generally smaller than the size of the discrete fourier transform (DFT) block. Without adjacent-channel-interference, the outputs from these virtual carriers are zero. Without loss of generality, we assume that carriers from 1 to P are used for data transmission so that we have Eq. (4). In DFT-based OFDM, a cyclic prefix (CP) is added to the multiplexed output of the IDFT before it is transmitted through a fading channel [23]. After that, the CP is detached at the receiver. Here, we do not put it into equation.

Denote:

$\boldsymbol{h}$: *Channel Impulse Response.*
$[H(1)\ldots H(N)] = FFT(\boldsymbol{h})(N - point\, FFT\, of\, \boldsymbol{h})$.

$$\boldsymbol{H} = diag[H(1)\ldots H(N)] \tag{5}$$

$$\boldsymbol{H}_P = diag[H(1)\ldots H(P)] \tag{6}$$

The receiver input for the k-th block from one E-NODE-B without noise and CFO is given by:

$$y(k) = \boldsymbol{h}\boldsymbol{W}_P s(k) = \boldsymbol{W}_P \boldsymbol{H}_{\boldsymbol{P}} s(k) \tag{7}$$

It is clear that, each subchannel, with a scalar ambiguity, can be recovered by applying a DFT to *y(k)*:

$$\boldsymbol{W}_P^H y(k) = \boldsymbol{H}_{\boldsymbol{P}} s(k) \tag{8}$$

Denote φ the normalized carrier frequency offset and Ng the guard interval duration of OFDM symbol. In the presence of a carrier offset, $e^{j\varphi}$, the receiver input *y(k)* without noise is modulated by:

$$\boldsymbol{E}(\varphi) = diag[1, e^{j\varphi}, e^{j2\varphi}, \ldots, e^{j(N-1)\varphi}] \tag{9}$$

and becomes:

$$y(k) = \boldsymbol{E}(\varphi)\boldsymbol{W}_P \boldsymbol{H}_{\boldsymbol{P}} s(k) e^{j(k-1)(N+Ng)\varphi} \tag{10}$$

Since $\boldsymbol{W}_P^H \boldsymbol{E}(\varphi)\boldsymbol{W}_P \neq \boldsymbol{I}$, the $\boldsymbol{E}(\varphi)$ matrix breaks the orthogonality among the sub-channels and, as a result, introduces ICI. To recover *s(k)*, the carrier offset, φ requires be estimating and compensating before applying the DFT.

In the situation that we investigate, the received signal is the combination of signals from two sources. From Eq. (10), we denote the received signal from one transmitter without noise is:

$$y1(k) = \boldsymbol{E}(\varphi_l)\boldsymbol{W}_P\boldsymbol{H}_{\boldsymbol{1P}}s(k)e^{j(k-1)(N+Ng)\varphi} \tag{11}$$

where φ_1 is the normalized carrier frequency shift from the first E-NODE B and $\boldsymbol{H_{1P}}$ is the channel frequency response matrix similar to $\boldsymbol{H_P}$. With K is the number of OFDM symbols in one frame, we have one frame of the received signal from one E-NODE B is:

$$\boldsymbol{y1} = \boldsymbol{E}(\varphi_l)\boldsymbol{W}_P\boldsymbol{H}_{1P}\boldsymbol{s}^{\boldsymbol{T}}\boldsymbol{B}(\varphi_l) \tag{12}$$

in which $\boldsymbol{s} = [s(1), s(2), \ldots s(K)]^T$

and $\boldsymbol{B}(\varphi) = diag[1, e^{j(N+Ng)\varphi}, e^{j2(N+Ng)\varphi}, \ldots, e^{j(K-1)(N+Ng)\varphi}]$ are the sending data in one frame and the carrier frequency offset matrix along OFDM symbols.

When there are two E-NODE-Bs, the received data frame without noise is:

$$\boldsymbol{y} = \boldsymbol{E}(\varphi_l)\boldsymbol{W}_P\boldsymbol{H}_{\boldsymbol{1P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_l) + \boldsymbol{E}(\varphi_2)\boldsymbol{W}_P\boldsymbol{W}_{\boldsymbol{2P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_2) \tag{13}$$

In order to recover $s(k)$, the carrier offsets, φ_1 and φ_2, need to be estimated and compensated before performing the DFT.

3.2 Estimation Algorithm

Since $\boldsymbol{W}_P$ contains of a subset of the columns of the IDFT matrix, $\boldsymbol{W}$, its orthogonal complement, $\boldsymbol{W}^{\perp} = [w_{P+1} \ldots w_N]$, is known. We define:

$$\boldsymbol{Z}(\omega) = diag[1, e^{j\omega}, e^{j2\omega}, \ldots, e^{j(N-1)\omega}] \tag{14}$$

It is obvious that when $\omega = \varphi, \boldsymbol{Z}(\omega) = E(\varphi)$.

Considering a preamble data frame in which $\boldsymbol{s}$ is known in prior. Let the number of symbols per frame K greater than the number of data slots P, we construct $\dot{\boldsymbol{s}}$ from $\boldsymbol{s}$. Firstly, we make singular value decomposing $\boldsymbol{s}^{\boldsymbol{T}}$ into $\boldsymbol{U}, \boldsymbol{D}, \boldsymbol{V}(\boldsymbol{s}^{\boldsymbol{T}} = \boldsymbol{UDV})$ in which $\boldsymbol{U}$ and $\boldsymbol{V}$ are $P \times P$ and $K \times K$ unitary matrix respectively. $\boldsymbol{D}$ is a rectangular diagonal matrix with non-negative real numbers on the diagonal. Then, we set $\dot{\boldsymbol{s}}$ to be the last K −P column of the $K \times K$ matrix $\boldsymbol{V} : \dot{\boldsymbol{s}} = [v_{K-P+1}, \ldots, v_K] = [\dot{s}_1, \dot{s}_2, \ldots, \dot{s}_{K-P}]$. Consequently, we have $\boldsymbol{s}^{\boldsymbol{T}}\dot{s}_k$ is a null vector.

Denote:

$$\boldsymbol{T}(\omega) = diag[1, e^{j(N+Ng)\omega}, e^{j2(N+Ng)\omega}, \ldots, e^{j(K-1)(N+Ng)\omega}] \tag{15}$$

In the absence of noise, if $\omega_l = \varphi_l$ and $\omega_2 = \varphi_2$, we have:

$$
\begin{aligned}
& w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\boldsymbol{y}\boldsymbol{T}(\omega_2)^{-\mathbf{1}}\dot{s}_k \\
& = w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\{\boldsymbol{E}(\varphi_1)\boldsymbol{W}_P\boldsymbol{H}_{\mathbf{1P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_1)+\boldsymbol{E}(\varphi_2)\boldsymbol{W}_P\boldsymbol{H}_{\mathbf{2P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_2)\}\boldsymbol{T}(\omega_2)^{-\mathbf{1}}\dot{s}_k \\
& = w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\boldsymbol{E}(\varphi_1)\boldsymbol{W}_P\boldsymbol{H}_{\mathbf{1P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_1)\boldsymbol{T}(\omega_2)^{-\mathbf{1}}\dot{s}_k \\
& + w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\boldsymbol{E}(\varphi_1)\boldsymbol{W}_P\boldsymbol{H}_{\mathbf{2P}}\boldsymbol{s}^T\boldsymbol{B}(\varphi_2)\boldsymbol{T}(\omega_2)^{-\mathbf{1}}\dot{s}_k \\
& = 0,\ (i=1,\ldots,N-P; k=1,\ldots,K-P).
\end{aligned}
\tag{16}
$$

This observation suggests that we form a cost function given a finite number of data vectors as follows:

$$
\begin{aligned}
P(\omega_1,\omega_2) &= \sum_{k=1}^{K-P}\sum_{i=1}^{N-P}\left||w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\boldsymbol{y}\,\boldsymbol{T}(\omega_2)^{-\mathbf{1}}\dot{s}_k|\right|^2 \\
&= \sum_{k=1}^{K-P}\sum_{i=1}^{N-P} w_{p+i}^{H}\boldsymbol{Z}(\omega_1)^{-\mathbf{1}}\boldsymbol{y}\,\boldsymbol{T}(\omega_2)^{-1}\dot{s}_k\dot{s}_k^{H}\boldsymbol{T}(\omega_2)\boldsymbol{y}^H\boldsymbol{Z}(\omega_1)w_{p+i}
\end{aligned}
\tag{17}
$$

Clearly, $P(\omega_1,\omega_2)$ is zero when $\omega_1=\varphi_1$ and $\omega_2=\varphi_2$. Therefore, one can find the carrier offsets by evaluating $P(\omega_1,\omega_2)$ along the two independent unit circles. However, we can see that the cost function also has root $\omega_1=\varphi_2$ and $\omega_2=\varphi_1$; hence, the cost function has two minimas.

Due to the fact that the we have to solve in the form (φ_1,φ_2) and (φ_2,φ_1); we propose finding maxima of the following symmetric cost function to further enhance the performance:

$$
\boldsymbol{Q}(\omega_1,\omega_2)=P(\omega_1,\omega_2)+P(\omega_2,\omega_1) \tag{18}
$$

The proposed algorithm is summarized as follows:

(1) Constructing $\dot{\boldsymbol{s}}$ at the receiver based on the preamble data $\boldsymbol{s}$.
(2) Forming the cost function $\boldsymbol{Q}(\omega_1,\omega_2)$ as in (17) and (18) using the known receiver outputs $\boldsymbol{y}$ and the constructed $\dot{\boldsymbol{s}}$.
(3) Estimating the carrier offset as the minima of $\boldsymbol{Q}(\omega_1,\omega_2)$:

$$
(\hat{\varphi}_1,\hat{\varphi}_2)=(\omega_{0_1},\omega_{0_2}):\boldsymbol{Q}(\omega_{0_1},\omega_{0_2})=min\,\boldsymbol{Q}(\omega_1,\omega_1)
$$

(4) Compensating the results $(\hat{\varphi}_1,\hat{\varphi}_2)$ into the sending data for several next rounds of communication.
(5) Redoing the preamble data estimation after data transmission with another preamble data.

4 Simulation Results

We use the parameters shown in Table 1 to establish the communication scenario. In the simulation, we focus on the two communication links between E-NODE-Bs and MS. We assume that the system has already achieve the signal time synchronization at the receiver.

Table 1. Simulation parameters for double CFO estimation

Parameters	Values
Cell radius (ds)	500 m
Overlap coverage region (Δd)	400 m
Initial distance between train and E-NODE-B1 (d1)	100 m
Initial distance between train and E-NODE-B2 (d2)	400 m
Distance between railway and E-NODE-Bs (d_B)	50 m
Velocity of train (v)	360 km/h
Total number of subcarriers	128
Modulation	BPSK
Subcarrier spacing (Δf)	120 kHz
Bandwidth (BW)	10 MHz
Sampling frequency (fs)	15.36 MHz
Frequency	2.6 GHz

Figures 3 and 4 show the CFO of communication links when the MS is going far away from E-NODE-B1 and toward E-NODE-B2. Similar with Single CFO estimation, the proposed algorithm was not accurate at low E_b/N_0. However, it achieves high accuracy at higher E_b/N_0.

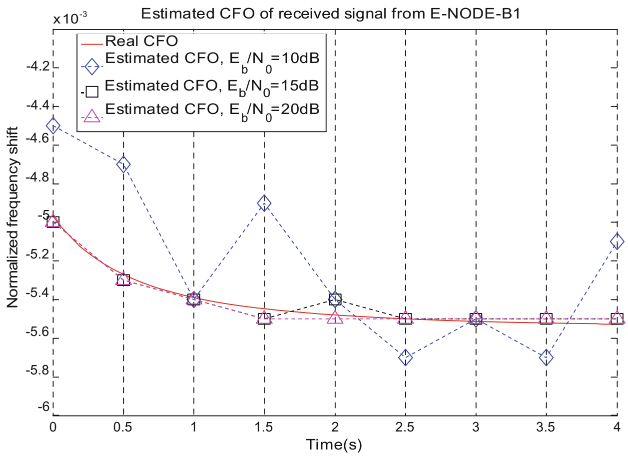

Fig. 3. The estimated CFO and the actual CFO at E-NODE-B1

The mean square error of our proposed method is described in Fig. 5. In the simulation, we compare our proposed method to the conventional solution as shown in [18]. From Fig. 5, it can be seen clearly that at higher *16 dB* E_b/N_0, our algorithm has shown its dominance.

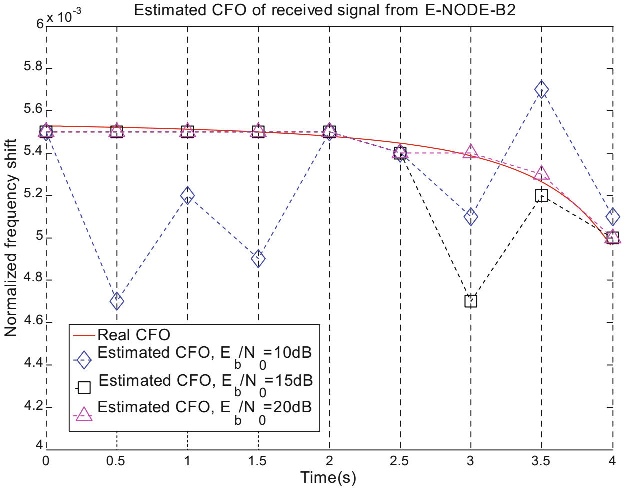

Fig. 4. The estimated CFO and the actual CFO at E-NODE-B2

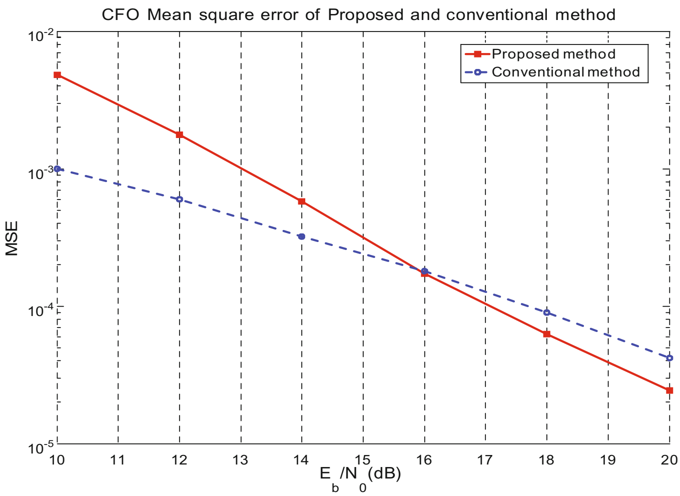

Fig. 5. MSE of the proposed algorithm and conventional method

Figure 6 is the BER of our communication scheme compare to communication link with no CFO and communication link with CFO but no CFO compensation. As E_b/N_0 got higher, our algorithm results approached the real CFO value; hence, the performance got closer to ideal case.

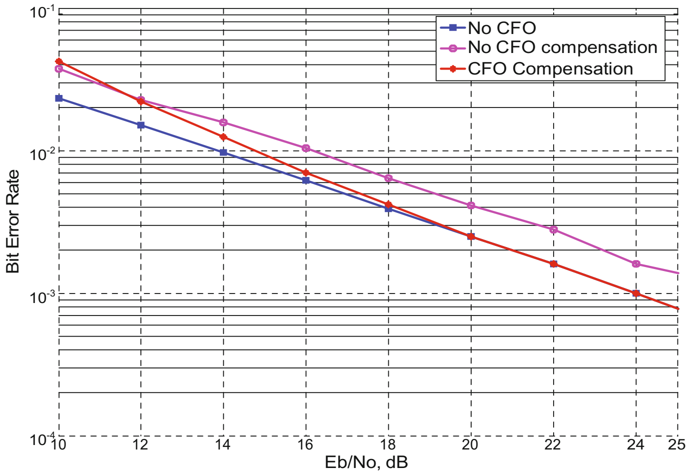

Fig. 6. BER with and without CFO compensation

5 Conclusions

We propose an inter-symbol-interference mitigation method by estimation of double carrier frequency offsets and pre-compensation of high Doppler shifts in High-Speed-Railway Communication Systems with distributed E-NODE-B. Our proposed method is based on a preamble frame of data and received signal without the implement of multiple received antennas. In our proposed solution, the MS estimates Doppler frequency shifts based on the received data and then feeds the estimated value back to E-NODE-B. When the new data is sent, the frequency offset pre-compensation is made for the data. The system is simulated and the double CFOs are estimated using the proposed algorithm. The mean square error of estimation is compared to conventional solution in order to show the performance. The proposed method gets a better accuracy and the CFOs are eliminated effectively and at high E_b/N_0. By the proposed method, the passengers's demand and transportation management networks require high-speed data services can be provided with reliable connections. This can be considered as a promising solution to solve problem of transportation infrastructures and High Speed Railway (HSR) in digital or smart cities.

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