ISI and ICI Compensation for TFI-OFDM in Time-Variant Large Delay Spread Channel

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Abstract: In general, if the maximum delay spread is longer than guard interval (GI), the system performance is significantly degraded. The conventional time-frequency interferometry (TFI) for OFDM does not consider with time-variant large delay spread channel. In this paper, we propose a novel time domain inter-symbol-interference (ISI) cancellation scheme with replica signal based inter-carrier-interference (ICI) compensation for TFI-OFDM.

1. Introduction

Mobile radio communication systems are increasingly demanded to provide a variety of high-quality multimedia services to mobile users. To meet this demand, modern mobile radio transceiver system must be able to support high capacity, variable bit rate information transmission and high bandwidth efficiency. In the mobile radio environment, signals are usually impaired by fading and multipath delay phenomenon. In such channels, severe fading of the signal amplitude and inter-symbol-interference (ISI) due to the frequency selectivity of the channel cause an unacceptable degradation of error performance. Orthogonal frequency division multiplexing (OFDM) is an efficient scheme to mitigate the effect of multipath channel. Since it eliminates ISI by inserting guard interval (GI) longer than the delay spread of the channel [1]. Therefore, OFDM is generally known as an effective technique for high data rate services.

In OFDM systems, the pilot signal assisted channel estimation is generally used to identify the channel state information (CSI) [2]. In this case, large pilot symbols are required to obtain an accurate CSI. As a result, the total transmission rate is degraded due to transmission of large pilot symbols. To reduce this problem, time-frequency interferometry (TFI) for OFDM has been proposed [3]. In general, the GI is usually designed to be longer than the delay spread of the channel, and is decided after channel measurements in the desired implementation scenario. However, if the maximum delay spread is longer than GI, the system performance is significantly degraded. TFI-OFDM does not consider with time-variant large delay spread channel. In this paper, we focus on the large delay spread channel and evaluate the performance on the TFI-OFDM. Until this time, several schemes have proposed to mitigate the ISI due to the large delay spread channel. Lee proposed double window cancellation and combining as ISI canceller scheme [4]. However the complexity is considerable works. Suyama proposed several frequency domain ISI cancellering schemes such as a smoothed FFT windows and decision feedback equalizer (DFE) [5]. However it is very sensitive to the severe frequency selective channel, and also it has a power loss by removing GI. To reduce above-mentioned problems, in this paper, we propose a novel time domain ISI cancellation scheme with replica signal based inter-carrier-interference (ICI) compensation for TFI-OFDM.

2. Proposed System

This section describes the proposed system, which employs time division multiplexing (TDM) transmission for multiple users. The proposed system is illustrated in Fig. 1.

2.1 Channel Model

We assume that a propagation channel consists of L discrete paths with different time delays. The impulse response $h(\tau, t)$ is represented as

$$h(\tau,t)=\sum_{l=0}^{L-1} h_l(t)\delta(\tau-\tau_l), \tag{1}$$

where h_l and τ_l are complex channel gain and the time delay of lth propagation path, respectively, and $\sum_{l=0}^{L-1} E|h_l^2| = 1$, where $E|\cdot|$ denotes the ensemble average operation. The channel transfer function $H(f,t)$ is the Fourier transform of $h(\tau, t)$ and is given by

$$\begin{aligned} H(f,t) &= \int_0^{\infty} h(\tau,t)\exp(-j2\pi f\tau)d\tau \\ &= \sum_{l=0}^{L-1} h_l(t)\exp(-j2\pi f\tau_l). \end{aligned} \tag{2}$$

2.2 TFI-OFDM

The transmitter block diagram of proposed system is shown in Fig. 1(a). Firstly, the coded binary information data sequence is modulated, and N_p pilot symbols are appended at the beginning of the sequence. The proposed system transmit signal can be expressed in its equivalent baseband representation as

$$\begin{aligned} s(t)=\sum_{i=0}^{N_p+N_d-1} g(t-iT)\cdot\Bigg\{\sqrt{\tfrac{2S}{N_c}}\sum_{k=0}^{N_c-1} u(k,i) \\ \cdot\exp[j2\pi(t-iT)k/T_s]\Bigg\}, \end{aligned} \tag{3}$$

where N_d and N_p are the number of data and pilot symbols, N_c is the number of carriers, T_s is the effective symbol length, S is the average transmitting power, T is the OFDM symbol length, respectively. The frequency separation between adjacent orthogonal subcarriers is $1/T_s$ and can be expressed, by using the kth subcarrier of the ith modulated symbol $d(k,i)$ with $|d(k,i)| = 1$ for $N_p \le i \le N_p + N_d - 1$, as

$$u(k,i)=c_{PN}\cdot d(k,i), \tag{4}$$

where c_{PN} is a long pseudo-noise (PN) sequence as a scrambling code to reduce the peak average power ratio (PAPR). The guard interval T_g is inserted in order to eliminate the ISI due to the multi-path fading, and hence, we have

$$T=T_s+T_g. \tag{5}$$

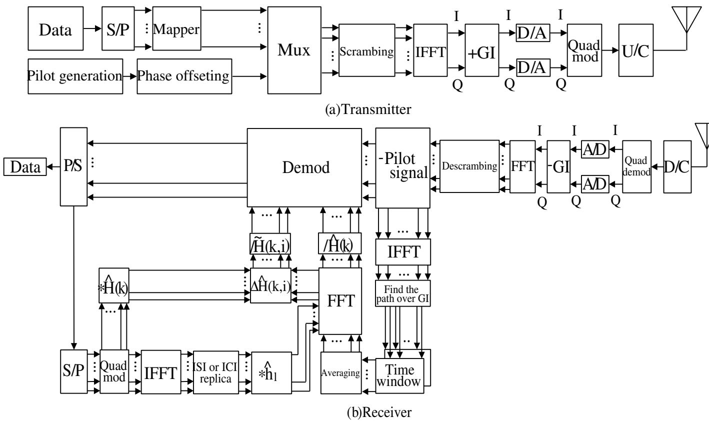

Figure 1. Proposed system.

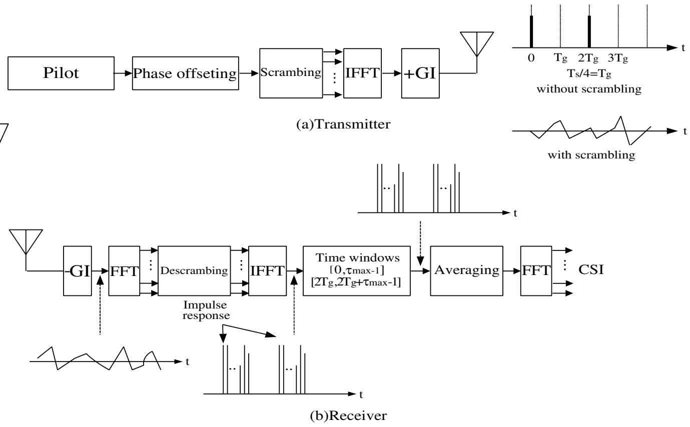

Figure 2. The concept of TFI-OFDM system.

In OFDM systems, T_g is generally considered as $T_s/4$ or $T_s/5$. Thus, we assume $T_g = T_s/4$ in this paper. In Eq. (3), $g(t)$ is the transmission pulse given by

$$g(t)=\begin{cases} 1 & \text{for } -T_g \le t \le T_s \\ 0 & \text{otherwise.} \end{cases} \quad (6)$$

For $0 \le i \le N_p - 1$, the transmitted pilot signal of kth subcarrier is given by

$$d(k,i)=\exp(-j2\pi k/T_s)+\exp(-j4\pi kT_g/T_s), \quad (7)$$

where N_p is the number of pilot symbols. In this case, pilot signal of proposed system can multiplex the same impulse responses in twice on the time domain without overlapping to each other as shown in Fig. 2(a). Moreover, due to the superposition of Eq. (7), the transmission power of pilot signals is $1/2$ for $0 \le i \le N_p - 1$.

The receiver structure is illustrated in Fig. 1(b). By applying the FFT operation, the received signal $r(t)$ is resolved into N_c subcarriers. The received signal $r(t)$ in the equivalent baseband representation can be expressed as

$$r(t)=\int_{-\infty}^{\infty} h(\tau,t)s(t-\tau)d\tau+n(t), \quad (8)$$

where $n(t)$ is additive white Gaussian noise (AWGN) with a single sided power spectral density of N_0. The kth subcarrier $\tilde{r}(k,i)$ is given by

$$\begin{aligned}\tilde{r}(k,i) &= \frac{1}{T_s}\int_{iT}^{iT+T_s} r(t)\exp[-j2\pi(t-iT)k/T_s]dt \\ &= \sqrt{\frac{2S}{N_c}}\sum_{e=0}^{N_c-1} u(e,i)\cdot\frac{1}{T_s}\int_0^{T_s}\exp[j2\pi \\ &\quad \cdot(e-k)t/T_s]\cdot\Big\{\int_{-\infty}^{\infty} h(\tau,t+iT)g(t-\tau) \\ &\quad \cdot\exp(-j2\pi e\tau/T_s)d\tau\Big\}dt+\hat{n}(k,i), \end{aligned} \quad (9)$$

where $\hat{n}(k,i)$ is AWGN noise with zero-mean and a variance of $2N_0/T_s$. After abbreviating, Eq. (9) can be rewritten as

$$\begin{aligned}\tilde{r}(k,i) &\approx \frac{1}{T_s}\sqrt{\frac{2S}{N_c}}\sum_{e=0}^{N_c-1} u(e,i)\cdot\int_0^{T_s}\exp[j2\pi \\ &\quad \cdot(e-k)t/T_s]\cdot\Big\{\int_{-\infty}^{\infty} h(\tau,t+iT) \\ &\quad \cdot g(t-\tau)\exp(-j2\pi e\tau/T_s)d\tau\Big\}dt+\hat{n}(k,i) \\ &= \sqrt{\frac{2S}{N_c}}H(k/T_s,iT)u(k,i)+\hat{n}(k,i). \end{aligned} \quad (10)$$

After descrambling, the output signal $r(k,i)$ is given by

$$\begin{aligned} r(k,i) &= \frac{c_{PN}^*(k)}{|c_{PN}(k)|^2}\Big\{\tilde{r}(k,i)\Big\} \\ &= \sqrt{\frac{2S}{N_c}}H(k/T_s,iT)d(k,i)+\hat{n}(k,i), \end{aligned} \quad (11)$$

where $\frac{c_{PN}^*(k)}{|c_{PN}(k)|^2}$ is the descrambling operation. Since $T_g = T_s/4$, TFI-OFDM can multiplex the same impulse responses in twice on the time domain. After the pilot signal separation, the pilot signal is converted to the time domain signal $\hat{r}(t)$ again as

$$\begin{aligned}\hat{r}(t) &= \sum_{i=0}^{N_p-1}\sqrt{\frac{2P}{N_c}}\sum_{k=0}^{N_c-1} r(k,i)\exp[j2\pi(t-iT)k/T_s] \\ &= \sum_{i=0}^{N_p-1}\sqrt{\frac{2P}{N_c}}h(\tau,t+iT)\sum_{k=0}^{N_c-1} d(k,i) \\ &\quad \cdot\exp[j2\pi(t-iT)k/T_s]+\tilde{n}(t) \\ &= \sum_{i=0}^{N_p-1}\sqrt{\frac{2P}{N_c}}\sum_{l=0}^{L-1} h_l(t+iT) \\ &\quad \cdot\frac{1}{\sqrt{2}}\Big\{\delta(\tau-\tau_l)+\delta(\tau-\tau_l-2T_g)\Big\}+\tilde{n}(t), \end{aligned} \quad (12)$$

where P is the power of pilot signals. The converted time domain signal $\hat{r}(t)$ is shown in Fig. 2(b). From Eq. (7), $\sum_{k=0}^{N_c-1} d(k,i)\exp[j2\pi(t-iT)k/T_s]$ shows two impulses with time shift as $\delta(\tau-2T_g)$, and the output signals are equivalent to time domain multiplexed impulse responses. By averaging of these impulse responses, we can reduce the noise power. Therefore, the impulse response of kth subcarrier $\hat{H}(k)$ is obtained by

$$\begin{aligned}\hat{H}(k) &= \frac{1}{\sqrt{P/N_c}}\sum_{e=0}^{N_c-1}\frac{1}{T_s}\int_0^{T_s}\Big\{\sum_{l=0}^{L-1} h_l(t+iT) \\ &\quad \cdot\Big\{\delta(\tau-\tau_l)+\delta(\tau-\tau_l-2T_g)\Big\} \\ &\quad \cdot\exp(-j2\pi e\tau/T_s)d\tau\Big\}dt+\eta(k), \end{aligned} \quad (13)$$

where $\eta(k)$ is AWGN component with $E[|\eta(k)|]^2 = E[|\frac{\hat{n}(k,i)}{2}|]^2 = \frac{\sigma^2}{2}$. For $L < T_g$, the faded channel is easily compensated by using Eq. (13). Alternately, Eq. (12) for $N_p \le i \le N_p + N_d - 1$ can be rewritten in matrix form as

$$\mathbf{R_i}=\mathbf{H_i}\mathrm{F}\mathbf{D_i}+\mathbf{N_i}, \quad (14)$$

Where $\mathbf{H_i}$ is the $N_c \times N_c$ time-domain matrix for the ith symbol, $\mathbf{N_i}$ is the $N_c \times 1$ noise matrix, F is the IFFT operation, respectively. For $L > T_g$, Eq. (14) can be written as

$$\mathbf{R_i}=\mathbf{H_{i-1,isi}}\mathrm{F}\mathbf{D_{i-1}}+\mathbf{H_{i,ici}}\mathrm{F}\mathbf{D_i}+\mathbf{N_i}, \quad (15)$$

where $\mathbf{H_{i-1,isi}}$ and $\mathbf{H_{i,ici}}$ denote ISI and ICI channel matrices for the $(i-1)$th and ith symbols.

2.3 Conventional ISI and ICI Equalization

In general, the first symbol of the received signals has no ISI for $L > T_g$ [4], [5]. In this case, Eq. (15) can be written as

$$\mathbf{R_i}=\mathbf{H_{i,ici}}\mathrm{F}\mathbf{D_i}+\mathbf{N_i}. \quad (16)$$

From Eq. (16), the equalization coefficient can be easily calculated by the Hermitian of $\mathbf{H_{i,ici}}$F. However, since the symbols behind the first symbol include the ISI, ISI equalization should be performed with the previously detected symbol $\hat{\mathbf{D}}_{i-1}$ and ISI channel matrix $\mathbf{H_{i-1,isi}}$. In this case, the

ISI equalized signal $\tilde{\mathbf{R}}_i$ is given by

$$\begin{aligned}\tilde{\mathbf{R}}_i &= \mathbf{R_i} - \mathbf{H_{i-1,isi}}\mathrm{F}\mathbf{D_{i-1}} + \tilde{\mathbf{N}}_\mathbf{i} \\ &= \lambda_{\mathbf{i,ici}}\mathrm{F}\mathbf{D_i} + \tilde{\mathbf{N}}_\mathbf{i},\end{aligned} \tag{17}$$

where $\tilde{\mathbf{N}}_i$ is the noise term with the residual ISI, $\lambda_{\mathbf{i,ici}}$ is the remained ICI channel matrix, respectively. Observing Eq. (17), since the residual ISI and ICI terms exist, a further equalization processing is necessary. Zero-forcing (ZF) uses a combining weight that is inversely proportional to the estimated channel matrix $\lambda_{\mathbf{i,ici}}$, in order to mitigate the residual ISI and ICI. The ZF weight is given by

$$\omega_{\mathbf{i,ZF}} = \frac{1}{\lambda_{\mathbf{i,ici}} \cdot \mathrm{F}}. \tag{18}$$

By using Eq. (18), the detected signal $\hat{\mathbf{D}}_i$ can be written as

$$\begin{aligned}\hat{\mathbf{D}}_\mathbf{i} &= \omega_{\mathbf{i,ZF}} \cdot \lambda_{\mathbf{i,ici}}\mathrm{F}\mathbf{D_i} + \omega_{\mathbf{i,ZF}} \cdot \tilde{\mathbf{N}}_\mathbf{i} \\ &= \mathbf{D_i} + \frac{\tilde{\mathbf{N}}_\mathbf{i}}{\lambda_{\mathbf{i,ici}} \cdot \mathrm{F}}.\end{aligned} \tag{19}$$

From Eq. (19), we can observe that the first term is the desired signal, the second term is a noise term. From the second term, ZF scheme can equalize the residual ISI and ICI, but it enhances the noise term for a deep faded subcarrier.

2.4 Replica Signal Insertion Based ICI Equalization

To avoid the noise enhancement due to the residual ISI and ICI, we consider the replica signal insertion based ICI equalization. ICI equalization is performed with the detected symbol $\tilde{\mathbf{D}}_i$ and ICI channel matrix $\mathbf{H_{i,ici}}$ of Eq. (19). From Eq. (17), ISI is deducted from the received signal. However, the orthogonality is destroyed by the deducted signal due to the ISI compensation. In this case, we can easily compensate the problem with inserting the deducted part of the signal. The ICI equalized signal $\check{\mathbf{R}}_i$ with deducted the part of the signal is given by

$$\begin{aligned}\check{\mathbf{R}}_i &= \tilde{\mathbf{R}}_i + \mathbf{H_{i,ici}}\mathrm{F}\hat{\mathbf{D}}_\mathrm{i} + \check{\mathbf{N}}_\mathrm{i} \\ &= \lambda_\mathbf{i}\mathrm{F}\hat{\mathbf{D}}_\mathrm{i} + \check{\mathbf{N}}_\mathrm{i}.\end{aligned} \tag{20}$$

where $\check{\mathbf{N}}_i$ is the noise term with the residual ICI, $\lambda_\mathbf{i}$ is the ICI deducted channel matrix, respectively. By using Eq. (20), the detected signal $\tilde{\mathbf{D}}_i$ can be written as

$$\begin{aligned}\tilde{\mathbf{D}}_i &= \tilde{\omega}_{i,ZF} \cdot \lambda_\mathbf{i}\mathrm{F}\hat{\mathbf{D}}_\mathrm{i} + \tilde{\omega}_{\mathrm{i,ZF}} \cdot \check{\mathbf{N}}_\mathrm{i} \\ &= \hat{\mathbf{D}}_i + \frac{\check{\mathbf{N}}_i}{\lambda_\mathbf{i} \cdot \mathrm{F}},\end{aligned} \tag{21}$$

where $\tilde{\omega}_{i,ZF} = \frac{1}{\lambda_\mathbf{i} \cdot \mathrm{F}}$. Observing Eqs. (19), (21), the replica signal based ICI equalization can increase the detection property. However, the detection property is depended on the channel identification. $\lambda_\mathbf{i}$ is calculated by $\mathbf{H_{i-1,isi}}$ and $\mathbf{H_{i,ici}}$, thus, the estimation errors are included in the detected signal. To eliminate the estimation errors, we calculate the estimation errors as

$$\Delta\hat{H}(k,i) = \frac{\check{d}(k,i)}{\tilde{d}(k,i) \cdot \hat{H}(k)}, \tag{22}$$

where $\check{d}(k,i)$ the frequency domain components of $\check{\mathbf{R}}_i$, $\tilde{d}(k,i)$ is the frequency domain components of $\tilde{\mathbf{D}}_i$. From Eq. (22), the estimation errors include the noise. In order to control the influence of noise, we average the previous symbols. As a results, the averaged estimation errors $\varepsilon(k,i)$ is given by

$$\varepsilon(k,i) = \frac{\Delta\hat{H}(k,i) + \Delta\hat{H}(k,i-1)}{2}. \tag{23}$$

Therefore, the improved CSI is given by

$$\tilde{H}(k,i) = \hat{H}(k) \cdot \varepsilon(k,i). \tag{24}$$

By using $\tilde{H}(k,i)$, we compensate the channel estimation error. The detected data signal $\acute{d}(k,i)$ is given by

$$\acute{d}(k,i) = \frac{\check{d}(k,i)}{\tilde{H}(k,i)}. \tag{25}$$

3. Computer Simulated Results

In this section, we show the performance of the proposed method. Figure 1 shows a simulation model of the proposed system. On the transmitter, the pilot signals are assigned for each transmitter using Eq. (7). In this case, the proposed system can multiplex the same impulse responses in the receiver antenna in twice on the time domain without overlapping to each other as shown in Fig. 2(a). The coded bits are QPSK modulated, and then the pilot signal and data signal are multiplexed with scrambling using PN code to reduce the PAPR. The OFDM time signals are generated by an IFFT and transmitted to frequency selective and time variant radio channel after cyclic extensions have been inserted. The transmitted signals are subject to broadband channel propagation. In the simulation, we assume that OFDM symbol period is 4 μs, guard interval is 1 μs, and channel model is $L = 2$ paths Rayleigh fadings. The maximum Doppler frequency is assumed to be 10 Hz. In the receiver, the guard interval is erased from the received signals and S/P converted. The parallel sequences are passed to an FFT operator, which converts the signal back to the frequency domain. After descrambling and IFFT, each impulse response can be estimated by extracting and averaging twice impulse responses using the time windows with Eq. (15) as shown in Fig. 2(b). After FFT operation, the frequency domain data signal is demodulated using the estimated channel impulse response. Since the detected data signals include the ISI and ICI, it is necessary to remove them. ISI equalization should be performed with the previous detected symbol and ISI channel matrix as Eq. (17). By using ZF, the ISI and ICI can be equalized, but it enhances the noise term for a deep faded subcarrier. To overcome this problem, replica signal insertion based ICI equalization is considered. From Eq. (17), ISI is deducted from the received signal. However, the orthogonality is destroyed by the deducted signal due to the ISI compensation. By using the replica signal insertion, ICI is deducted as Eq. (21). From Eq. (21), the estimation errors are included in the detected signal. To eliminate the estimation errors, the proposed method calculate the estimation errors as Eqs. (23) and (24). Finally, by using Eq. (24), the data signal is detected. The packet consists of $N_p = 1$ pilot symbol and $N_d = 20$ data symbols.

Figure 3 shows the BER versus the total power ratio β for E_b/N_0=10, 20, and 30 dBs. In order to compare the performance of different total power ratio between in side and outside of GI, the electric power ratio β is defined as

$$\beta = 10log_{10}\left|\frac{P_{in\_GI}}{P_{out\_GI}}\right|, \tag{26}$$

where P_{in_GI} and P_{out_GI} are the total power of paths of inside and outside as GI. From the simulation results, the BER performances are saturated when $\beta \geq 22$. It means that ISI

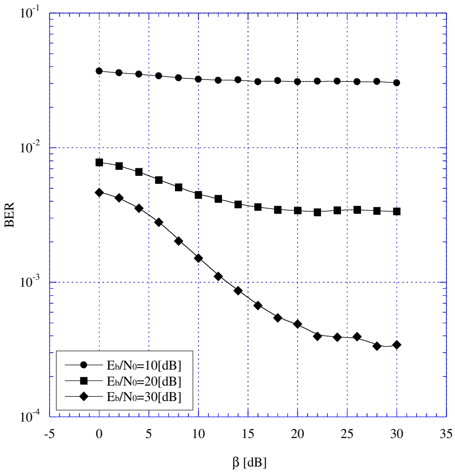

Figure 3. BER versus the total power ratio β for E_b/N_0=10, 20, and 30 dBs.

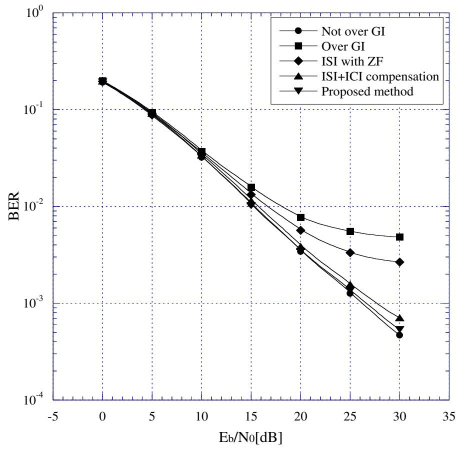

Figure 4. BER of not over the GI, over the GI, the ISI with ORC, the ISI+ICI compensation, and the proposed method for β=0 dB.

and ICI are not affect the system performance. Therefore, we can abbreviate the compensation processing for ISI and ICI.

Fig. 4 shows the BER of the proposed method and the conventional methods for β=0 and Doppler frequency of 10 Hz. For a delay spread that is longer than the GI, the BER rises rapidly due to ISI and ICI. From the simulation results, the BER with a delay spread that is longer than GI, show about 10 times compared with no ISI and ICI case. In the ISI compensation with ZF as a ICI compensation method, the BER performance shows the error floor. This is because the ICI is still remained and the ICI and noise components are amplified by ZF. The proposed method shows the approximately same BER performance like the case with no ISI and ICI. This is because the proposed method can mitigate the ISI and ICI by using the accurate CSI that estimated from the replica signal and the received signal.

Fig. 5 shows the BER versus various β for E_b/N_0=30 and Doppler frequency of 10 Hz. The proposed method shows the best BER when $\beta < 8$. For $8 \leq \beta < 22$, the proposed method and the conventional ISI+ICI method show the approximately same BER performance. This is because the amount of the interference is degraded with increasing β. Therefore, the estimation error is also degraded. As a result, we can adaptively abbreviate the compensation complexity with considering β.

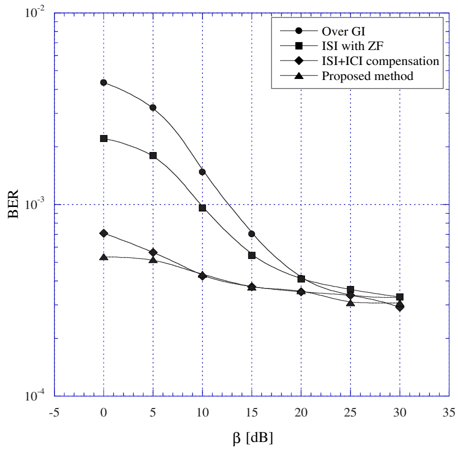

Figure 5. BER of over the GI, the ISI with ORC, the ISI+ICI compensation, and the proposed method for E_b/N_0=30 dB.

4. Conclusion

In this paper, we focus on the large delay spread channel and proposed the ISI and ICI compensation method for TFI-OFDM. From the simulation results, it is shown that the proposed method achieves the approximately same BER performance like the case with no ISI and ICI. For $8 \leq \beta < 22$, the proposed method and the conventional ISI+ICI method show the approximately same BER. Therefore, we can adaptively abbreviate the compensation complexity with considering β.

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