Interference Cancellation and Channel Estimation Schemes for Single Carrier Block Transmission with Insufficient Cyclic Prefix

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Abstract

This paper proposes a simple inter-symbol interference (ISI) and inter-block interference (IBI) elimination scheme for a single carrier block transmission with cyclic prefix (SC-CP) system with an insufficient guard interval (GI). In the SC-CP system, only limited number of symbols in a transmitted block cause the interferences, while all the information data contribute to the interferences in the orthogonal frequency division multiplexing (OFDM) system. Therefore, in the SC-CP system, the interferences can be exterminated by only setting transmitted signals to be 0 at certain time slots in a transmitted signal block without changing any parameters or configuration of the system, such as the length of GI. The proposed scheme also can be considered as a simple loading scheme in temporal domain, which tries not to use time slots contaminated by the ISI and the IBI. Moreover, this paper also proposes a pilot signal configuration for the SC-CP system, which enable us to estimate channel impulse response in the discrete frequency domain even when the channel order is longer than the GI length.

1 Introduction

A block transmission with cyclic prefix (CP), including orthogonal frequency division multiplexing (OFDM)[1],[2] and single carrier block transmission with cyclic prefix (SC-CP)[3],[4], has been drawing much attention as a promising candidate for the 4G (4th generation) mobile communications systems. The insertion of the CP as a guard interval (GI) at the transmitter and the removal of the CP at the receiver eliminates inter-block interference (IBI), if all the delayed signals exist within the GI. Moreover, the insertion and the removal of the CP converts the effect of the channel from a linear convolution to a circular convolution, therefore, the inter-symbol interference (ISI) of the received signal can be effectively equalized by a discrete frequency domain equalizer (FDE) using fast Fourier transform (FFT).

The existence of delayed signals beyond the GI causes residual ISI and IBI after the FDE, and hence deteriorates the performance of the block transmission with the CP. In order to overcome the performance degradation due to the insufficient GI, a considerable number of studies have been made on the issue, such as impulse response shortening[5], utilization of an adaptive antenna array[6] and per-tone equalization[7],[8]. All these methods can improve the performance, however, they increase the computational or system complexity, which could spoil the most important feature of the FDE based system, namely, the simplicity.

In this paper, we propose a simple ISI and IBI elimination scheme for the SC-CP system with the insufficient GI. In the SC-CP system, only limited number of symbols in a transmitted block cause the interferences, while all the information data contribute to the interferences in the OFDM system. Therefore, in the SC-CP system, the interferences can be exterminated by only setting transmitted signals to be zero at certain time slots in a transmitted signal block without changing any parameters or configuration of the system. The proposed scheme also can be considered as a simple loading scheme[9] in the temporal domain, which tries not to use time slots contaminated by the ISI and the IBI. Moreover, we also proposes a pilot signal configuration for the SC-CP system, which enable us to estimate channel impulse response in the discrete frequency domain even when the channel order is longer than the GI length. Computer simulations show that the proposed schemes can significantly improve bit error rate (BER) performance and accuracy of the channel estimation when the channel order is larger than the GI length.

2 The SCCP System with Insufficient CP

Fig.1 shows a basic configuration of the SC-CP system. Let $\mathbf{s}(n) = [s_0(n), \ldots, s_{M-1}(n)]^T$, where the superscript $(\cdot)^T$ stands for the transpose, be the nth information signal block of size $M \times 1$. The transmitted signal block $\mathbf{s}'(n)$ of size $(M+K) \times 1$ is generated from $\mathbf{s}(n)$ by inserting the CP of K symbols length as the

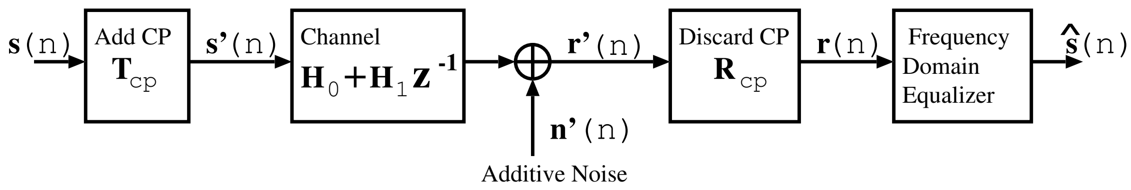

Fig. 1: Basic Configuration of SC-CP system

GI, namely,

$$\mathbf{s}'(n) = \mathbf{T}_{cp}\mathbf{s}(n), \tag{1}$$

where $\mathbf{T}_{cp}$ denotes the CP insertion matrix of size $(M+K) \times M$ defined as

$$\mathbf{T}_{cp} = \begin{bmatrix} \mathbf{I}_{cp} \\ \mathbf{I}_{M\times M} \end{bmatrix}, \tag{2}$$

$$\mathbf{I}_{cp} = \begin{bmatrix} \mathbf{0}_{K\times(M-K)} & \mathbf{I}_{K\times K} \end{bmatrix}.$$

$\mathbf{0}_{K\times(M-K)}$ is a zero matrix of size $K \times (M-K)$, and $\mathbf{I}_{M\times M}$ is an identity matrix of size $M \times M$.

The received signal block $\mathbf{r}'(n)$ is written as

$$\mathbf{r}'(n) = \mathbf{H}_0\mathbf{s}'(n) + \mathbf{H}_1\mathbf{s}'(n-1) + \mathbf{n}'(n), \tag{3}$$

where $\mathbf{n}'(n)$ is a channel noise vector of size $(M+K)\times 1$. $\mathbf{H}_0$ and $\mathbf{H}_1$ denote $(M+K)\times(M+K)$ channel matrices defined as

$$\mathbf{H}_0 = \begin{bmatrix} h_0 & 0 & \cdots & \cdots & \cdots & 0 \\ \vdots & \ddots & \ddots & & & \vdots \\ h_L & & \ddots & \ddots & & \vdots \\ 0 & \ddots & & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & & \ddots & 0 \\ 0 & \cdots & 0 & h_L & \cdots & h_0 \end{bmatrix}, \tag{4}$$

$$\mathbf{H}_1 = \left[\begin{array}{c|ccc} & h_L & \cdots & h_1 \\ & 0 & \ddots & \vdots \\ \mathbf{0}_{(M+K)\times(M+K-L)} & \vdots & \ddots & h_L \\ & \vdots & & 0 \\ & \vdots & & \vdots \\ & 0 & \cdots & 0 \end{array}\right], \tag{5}$$

where $\{h_0, \ldots, h_L\}$ denotes the channel impulse response.

After discarding the CP portion of the received signal block $\mathbf{r}'(n)$, the received signal block $\mathbf{r}(n)$ of size $M \times 1$ can be written as

$$\begin{aligned} \mathbf{r}(n) &= \mathbf{R}_{cp}\mathbf{r}'(n) \\ &= \mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}\mathbf{s}(n) + \mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp}\mathbf{s}(n-1) + \mathbf{n}(n), \end{aligned} \tag{6}$$

where $\mathbf{R}_{cp}$ denotes the CP discarding matrix of size $M \times (M+K)$ defined as

$$\mathbf{R}_{cp} = \begin{bmatrix} \mathbf{0}_{M\times K} & \mathbf{I}_{M\times M} \end{bmatrix}, \tag{7}$$

and $\mathbf{n}(n) = \mathbf{R}_{cp}\mathbf{n}'(n)$.

If the length of the GI is sufficiently long, namely, $K > L-1$, it can be easily verified that $\mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp}$ becomes a zero matrix, and the nth received signal block has no IBI component from the $(n-1)$th transmitted signal block. Moreover, if $K > L-1$, $\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$ becomes a circulant matrix of size $M \times M$, which means the one-tap FDE can equalize the ISI effectively.

However, if the length of the GI is insufficient ($K \leq L-1$), $\mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp}$ is no longer a zero matrix. Instead, $\mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp}$ can be written as

$$\mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp} = \left[\begin{array}{c|ccc} & h_L & \cdots & h_{K+1} \\ & 0 & \ddots & \vdots \\ \mathbf{0}_{M\times(M-L+K)} & \vdots & \ddots & h_L \\ & \vdots & & 0 \\ & \vdots & & \vdots \\ & 0 & \cdots & 0 \end{array}\right]. \tag{8}$$

This means that the IBI from the $(n-1)$th transmitted signal block remains even after the CP removal operation at the receiver.

Also, $\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$ can be written as

$$\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp} = \begin{bmatrix} h_0 & 0 & \cdots & \cdots & \cdots & 0 & h_K & \cdots & h_1 \\ \vdots & \ddots & \ddots & & & \vdots & \vdots & & \vdots \\ \vdots & & \ddots & \ddots & & 0 & h_L & & \vdots \\ \vdots & & & \ddots & \ddots & & \ddots & \ddots & \vdots \\ h_L & & & & \ddots & \ddots & & \ddots & h_L \\ 0 & \ddots & & & & \ddots & \ddots & & 0 \\ \vdots & \ddots & \ddots & & & & \ddots & \ddots & \vdots \\ \vdots & & \ddots & \ddots & & & & \ddots & 0 \\ 0 & \cdots & \cdots & 0 & h_L & \cdots & \cdots & \cdots & h_0 \end{bmatrix}. \tag{9}$$

Note that $\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$ is no longer a circulant matrix. Therefore, it is difficult for the one-tap FDE to equalize the received signal block distorted by the ISI.

3 Proposed Interference Elimination Scheme

In order to present the proposed interference elimination scheme, we firstly separate the matrix $\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$ into the two matrices as

$$\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp} = \mathbf{C} - \mathbf{C}_{ISI}, \tag{10}$$

where $\mathbf{C}$ is a circulant matrix, whose first column is the same as that of the matrix $\mathbf{R}_{cp}\mathbf{H}_0\mathbf{T}_{cp}$, namely,

$$\mathbf{C} = \begin{bmatrix} h_0 & 0 & \dots & 0 & h_L & \dots & h_1 \\ \vdots & \ddots & \ddots & & \ddots & \ddots & \vdots \\ \vdots & & \ddots & \ddots & & \ddots & h_L \\ h_L & & & \ddots & \ddots & & 0 \\ 0 & \ddots & & & \ddots & \ddots & \vdots \\ \vdots & \ddots & \ddots & & & \ddots & 0 \\ 0 & \dots & 0 & h_L & \dots & \dots & h_0 \end{bmatrix}, \tag{11}$$

and $\mathbf{C}_{ISI}$ is the compensation term given by

$$\mathbf{C}_{ISI} = \left[\begin{array}{cccc} & h_L & \dots & h_{K+1} & \\ & 0 & \ddots & \vdots & \\ & \vdots & \ddots & h_L & \\ \mathbf{0}_{M\times(M-L)} & \vdots & & 0 & \mathbf{0}_{M\times K} \\ & \vdots & & \vdots & \\ & 0 & \dots & 0 & \end{array} \right]. \tag{12}$$

Using $\mathbf{C}$ and $\mathbf{C}_{ISI}$, the nth received signal block $\mathbf{r}(n)$ after the CP removal can be rewritten as

$$\mathbf{r}(n) = \mathbf{C}\mathbf{s}(n) - \mathbf{C}_{ISI}\mathbf{s}(n) + \mathbf{C}_{IBI}\mathbf{s}(n-1) + \mathbf{n}(n), \tag{13}$$

where $\mathbf{C}_{IBI}$ is defined as

$$\mathbf{C}_{IBI} = \mathbf{R}_{cp}\mathbf{H}_1\mathbf{T}_{cp}. \tag{14}$$

From (13), we can see that the first term of the right hand, $\mathbf{C}\mathbf{s}(n)$, can be equalized using the FDE, since $\mathbf{C}$ is a circulant matrix, while the second and the third terms could result in the ISI and the IBI component respectively at the FDE output.

The proposed interference elimination scheme is based on the following facts;

- The ISI and the IBI due to the insufficient GI can be eliminated if $\mathbf{C}_{ISI}\mathbf{s}(n) = \mathbf{0}_{M\times1}$ and $\mathbf{C}_{IBI}\mathbf{s}(n) = \mathbf{0}_{M\times1}$.

- Since only the limited number of columns of the matrices $\mathbf{C}_{ISI}$ and $\mathbf{C}_{IBI}$ have nonzero elements, only the corresponding symbols in the transmitted signal block $\mathbf{s}(n)$ cause the ISI and the IBI.

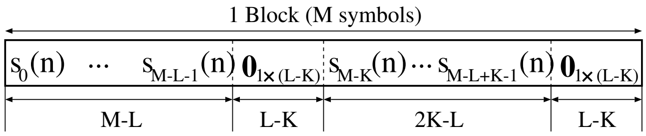

Fig. 2: Proposed Transmission Scheme for ISI and IBI elimination

From the facts, we can see that the interferences can be eliminated by only setting the information data to be zero at certain time slots (or symbols) in the information block $\mathbf{s}(n)$. The time slots are selected so that $\mathbf{C}_{ISI}\mathbf{s}(n) = \mathbf{0}_{M\times1}$ and $\mathbf{C}_{IBI}\mathbf{s}(n) = \mathbf{0}_{M\times1}$. To be more precise, the information block in the proposed scheme can be written as

$$\mathbf{s}(n) = \left[s_0(n), \dots, s_{M-L-1}(n), \mathbf{0}_{1\times(L-K)}, s_{M-K}(n), \dots, s_{M-L+K-1}(n), \mathbf{0}_{1\times(L-K)}\right]^T. \tag{15}$$

Also, Fig.2 show the proposed transmitted signal block configuration.

Note that the proposed transmission scheme does not require the transmitter to know the detailed channel state information (CSI), such as an instantaneous channel impulse response. The transmitter only has to know the channel order L. It also should be noted that the transmission rate of the proposed scheme is $(M-(L-K)\times2)/M$ times the transmission rate of the original SC-CP system, since the $(L-K)\times2$ time slots (symbols) in a block are not used for the signal transmission. Therefore, it can be said that the proposed scheme eliminates the interferences due to the insufficient GI at the cost of small reduction of the transmission rate.

With the proposed transmission scheme (15), the received signal block $\mathbf{r}(n)$ can be written as

$$\mathbf{r}(n) = \mathbf{C}\mathbf{s}(n) + \mathbf{n}(n). \tag{16}$$

Since $\mathbf{C}$ is a circulant matrix, it can be diagonalized by the discrete Fourier transform (DFT) matrix $\mathbf{F}$ of size $M\times M$ as[10]

$$\mathbf{C} = \mathbf{F}^H\mathbf{\Lambda}\mathbf{F}, \tag{17}$$

where the superscript H denotes the Hermitian transpose, and $\mathbf{\Lambda}$ is a diagonal matrix, whose diagonal elements are $\{\lambda_0, \dots, \lambda_{M-1}\}$.

Also, $\mathbf{\Lambda}$ can be calculated as

$$\mathbf{\Lambda} = \mathrm{diag}\left\{\mathbf{F}\begin{bmatrix} h_0 \\ \vdots \\ h_L \\ \mathbf{0}_{(M-L-1)\times1}\end{bmatrix}\right\}, \tag{18}$$

where diag$\{\mathbf{a}\}$ denotes a diagonal matrix, whose diagonal elements are the same as the elements of vector $\mathbf{a}$.

The one-tap FDE can be formulated as $\mathbf{F}^H\mathbf{\Gamma}\mathbf{F}$, where $\mathbf{\Gamma}$ is a diagonal matrix with the diagonal elements of $\{\gamma_0, \ldots, \gamma_{M-1}\}$. For zero-forcing (ZF) equalization, the ith equalizer weights $\gamma_{i,ZF}$ of the one-tap FDE is given by

$$\gamma_{i,ZF} = \frac{1}{\lambda_i}, \quad i = 0, \ldots, M-1, \tag{19}$$

and for minimum mean-square-error (MMSE) equalization, the equalizer weights $\gamma_{i,MMSE}$ is given by

$$\gamma_{i,MMSE} = \frac{\lambda_i^*}{|\lambda_i|^2 + \sigma_n^2/\sigma_s^2}, \quad i = 0, \ldots, M-1, \tag{20}$$

where the superscript $*$ denotes the complex conjugate, σ_n^2 is the variance of the additive channel noise, and σ_s^2 is the variance of the transmitted data symbols.

In this way, the same equalization methods as the conventional SC-CP system can be applied to the proposed scheme. The fundamental difference between the equalization in the proposed transmission scheme and in the conventional SC-CP system is that the channel order is longer than the length of the GI in the proposed scheme.

4 Channel Estimation Scheme for SCCP system with Insufficient CP

The proposed transmission scheme can effectively eliminate the ISI and the IBI component with the one-tap FDE, however, in order to equalize the received signal, it requires the channel impulse response, whose order may be longer than the length of the GI. In this section, we propose a pilot signal configuration for the computationally efficient channel estimation in the discrete frequency domain for the SC-CP system with the insufficient GI.

Let $\mathbf{p}(n) = [p_0(n), \ldots, p_{M-1}(n)]^T$ denotes the nth pilot signal block of length M. After the CP removal, the corresponding received pilot signal block, $\mathbf{r}_p(n)$, can be written as

$$\mathbf{r}_p(n) = \mathbf{C}\mathbf{p}(n) - \mathbf{C}_{ISI}\mathbf{p}(n) + \mathbf{C}_{IBI}\mathbf{p}(n-1) + \mathbf{n}(n), \tag{21}$$

where $\mathbf{C}$, $\mathbf{C}_{ISI}$ and $\mathbf{C}_{IBI}$ are the same as the matrices defined in (11), (12) and (14) respectively.

When the length of the GI is insufficient, namely, $K \leq L-1$, it becomes difficult to estimate the channel response in the DFT domain due to the ISI and the IBI components. However, from (21), we can see that if

$$\mathbf{C}_{ISI}\mathbf{p}(n) = \mathbf{C}_{IBI}\mathbf{p}(n-1) \tag{22}$$

holds, the received pilot signal block $\mathbf{r}_p(n)$ can be written as

$$\mathbf{r}_p(n) = \mathbf{C}\mathbf{p}(n) + \mathbf{n}(n), \tag{23}$$

which is the same form as the received pilot signal block with the sufficient GI. Therefore, conventional channel estimation schemes are available.

Inspection of (9) and (12) reveals that the two matrices, $\mathbf{C}_{ISI}$ and $\mathbf{C}_{IBI}$, share the same elements with the same arrangement, although they are not the same matrices. It is easily verified that $\mathbf{C}_{ISI}$ and $\mathbf{C}_{IBI}$ can be related as

$$\mathbf{C}_{ISI}\mathbf{S}^K = \mathbf{C}_{IBI}, \tag{24}$$

where the $M \times M$ shifting matrix $\mathbf{S}$ is defined as

$$\mathbf{S} = \begin{bmatrix} 0 & 1 & 0 & \ldots & 0 \\ \vdots & \ddots & \ddots & \ddots & \vdots \\ \vdots & & \ddots & \ddots & 0 \\ \vdots & & & \ddots & 1 \\ 0 & \ldots & \ldots & \ldots & 0 \end{bmatrix}. \tag{25}$$

Then we derive a condition imposed on the transmitted pilot signals for the equality in (22) to be true. Using (24), (22) can be modified as

$$\mathbf{C}_{ISI}\mathbf{p}(n) = \mathbf{C}_{ISI}\mathbf{S}^K\mathbf{p}(n-1). \tag{26}$$

From (26), it can be said that if the two consecutive pilot signal blocks, $\mathbf{p}(n-1)$ and $\mathbf{p}(n)$, have the relation of

$$\mathbf{p}(n) = \mathbf{S}^K\mathbf{p}(n-1), \tag{27}$$

the equality of (22) is always true. However, the condition of (27) can be much more relaxed because only the limited number of columns of $\mathbf{C}_{ISI}$ and $\mathbf{C}_{IBI}$ have nonzero elements. To be more precise, if the elements of $\mathbf{p}(n-1)$ and $\mathbf{p}(n)$ have the relation of

$$p_m(n) = p_{m+K}(n-1), \quad m = M-L, \ldots, M-K-1, \tag{28}$$

(22) is always true. Fig. 3 also shows the proposed pilot signal configuration for the channel estimation.

Now that we have the received pilot signal block given by (23), with the relation of the transmitted pilot signal blocks of (28), conventional channel estimation schemes can be used for the channel estimation. For example, since the received pilot signal block can be modified as

$$\begin{aligned} \mathbf{r}_p(n) &= \mathbf{C}\mathbf{p}(n) + \mathbf{n}(n) \\ &= \mathbf{Q}(n)\mathbf{h} + \mathbf{n}(n), \end{aligned} \tag{29}$$

where $\mathbf{Q}(n)$ is a $M \times (L+1)$ circulant matrix defined as

$$\mathbf{Q}(n) = \begin{bmatrix} p_0(n) & p_{M-1}(n) & \ldots & p_{M-L+1}(n) \\ p_1(n) & p_0(n) & \ddots & \vdots \\ \vdots & & \ddots & p_{M-1}(n) \\ \vdots & \vdots & & p_0(n) \\ \vdots & & & \vdots \\ p_{M-1}(n) & p_{M-2}(n) & \ldots & p_{M-L}(n) \end{bmatrix}, \tag{30}$$

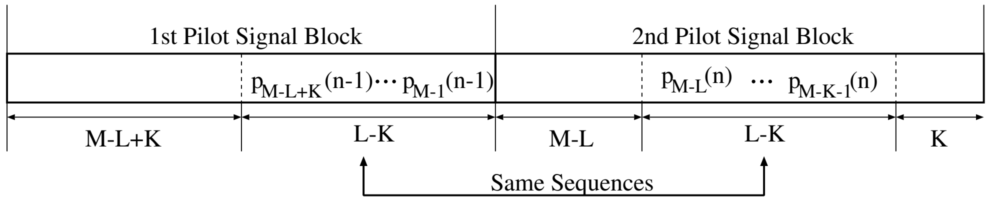

Fig. 3: Proposed Pilot Signal Configuration for Insufficient GI

and $\mathbf{h}$ is the channel impulse response vector defined as $\mathbf{h} = [h_0, \ldots, h_L]^T$, the maximum likelihood (ML) estimate of the channel impulse response is given by[11]

$$\hat{\mathbf{h}} = (\mathbf{Q}(n)^H\mathbf{Q}(n))^{-1}\mathbf{Q}(n)^H\mathbf{r}_p. \tag{31}$$

Although the ML method can provide the channel estimation of high precision, it requires high computational complexity. Since the proposed transmission scheme in Sec.2 aims at simplicity, such a channel estimation method may not be suited for the proposed transmission scheme. More computationally efficient channel estimation could be achieved in the DFT domain signal processing.

After the DFT of the received pilot signal $\mathbf{r}_p(n)$, we have

$$\begin{aligned} \mathbf{F}\mathbf{r}_p(n) &= \mathbf{F}\mathbf{C}\mathbf{p}(n) + \mathbf{F}\mathbf{n}(n) \\ &= \mathbf{F}\mathbf{F}^H\mathbf{\Lambda}\mathbf{F}\mathbf{p}(n) + \mathbf{N}(n) \\ &= \mathbf{\Lambda}\mathbf{P}(n) + \mathbf{N}(n) \\ &= \mathrm{diag}\{\mathbf{P}(n)\}\mathbf{H} + \mathbf{N}(n), \end{aligned} \tag{32}$$

where $\mathbf{P}(n) = \mathbf{F}\mathbf{p}(n)$, $\mathbf{N}(n) = \mathbf{F}\mathbf{n}(n)$ and $\mathbf{H} = F\left[\mathbf{h}^T \ \mathbf{0}_{1\times(M-L-1)}\right]^T$. From (32), the frequency response of the channel, $\mathbf{H}$, may be estimated as

$$\bar{\mathbf{H}} = (\mathrm{diag}\{\mathbf{P}(n)\})^{-1}\mathbf{F}\mathbf{r}_p(n). \tag{33}$$

Since $\mathbf{P}(n)$ is known to the receiver a priori, the calculation of the channel response $\bar{\mathbf{H}}$ is efficiently conducted using FFT.

5 Computer Simulation

In order to confirm the validity of the proposed transmission and the channel estimation schemes, we have conducted computer simulations. System parameters used in the computer simulations are summarized in table 1.

As a modulation/demodulation scheme, QPSK modulation with a coherent detection is employed. The FFT length (or the information block size), the length of GI and the channel order are set to be $M = 64$, $K = 16$ and $L = 20$ respectively. 9-path frequency selective Rayleigh fading channel is used for the channel model. Also, additive white Gaussian noise (AWGN) is assumed as the channel noise.

Table 1: System parameters

Mod./Demod. Scheme	QPSK/Coherent Detection
FFT Length	M=64
Guard Interval	K=16
Channel Order	L=20
Channel Model	9-path Rayleigh Fading Channel
Channel Noise	Additive White Gaussian Noise

Fig.4 shows the BER performance versus the ratio of the energy per bit to the noise power density (E_b/N_0) of the proposed transmission scheme with the MMSE based FDE or the ZF based FDE. The BER performances of the SC-CP system without the proposed transmission scheme are also plotted in the same figure. In this computer simulation, perfect channel estimation is assumed in order to evaluate the validity of only the proposed transmission scheme. From the figure, we can see that the proposed scheme can improve the BER performance significantly, while the performance of the SC-CP system without the proposed scheme is degraded due to the ISI and the IBI caused by the insufficient GI.

Fig.5 shows the mean-square-errors (MSEs) of the ML channel estimation and the simple discrete frequency domain channel estimation versus the E_b/N_0 with and without the proposed pilot signal configuration. The MSE is defined as

$$MSE = \frac{1}{N_{trial}} \sum_{i=1}^{N_{trial}} \frac{||\mathbf{h} - \bar{\mathbf{h}}||^2}{||\mathbf{h}||^2}, \tag{34}$$

where N_{trial} denotes the number of channel realizations and was set to be $1,000$ in the simulations. From the figure, we can see that the proposed pilot signal configuration can significantly improve the accuracy of the channel estimation both for the ML channel estimation and the simple discrete frequency domain channel estimation.

6 Conclusion

We have proposed a simple ISI and IBI elimination scheme for the SC-CP system with the insufficient GI. The proposed transmission scheme can exterminate the

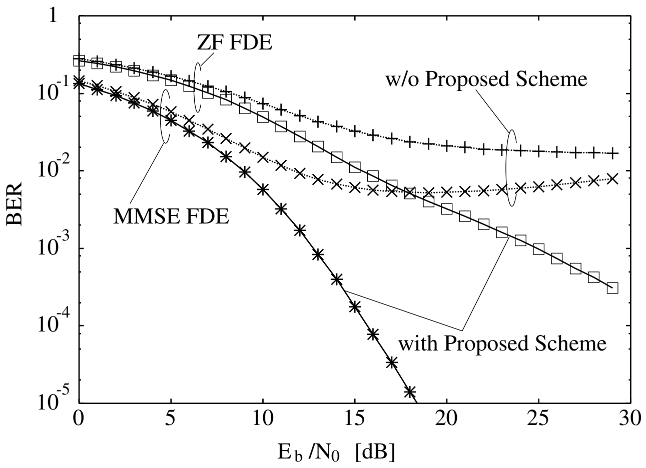

Fig. 4: Computer Simulation Results: BER Performance

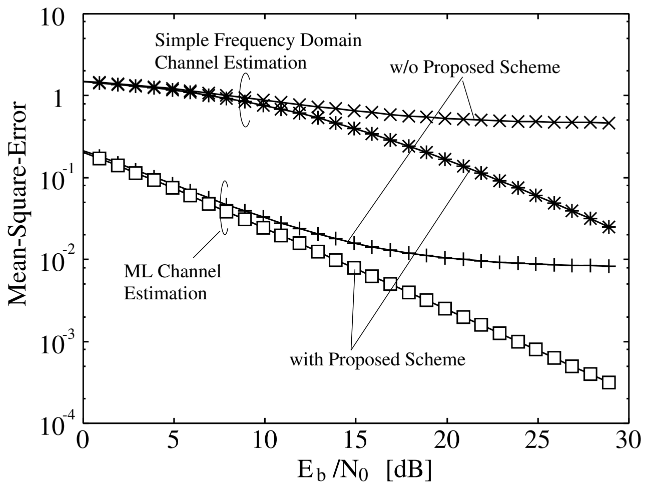

Fig. 5: Computer Simulation Results: Channel Estimation Error

interferences without changing any system parameters or the configuration from the basic SC-CP system. Moreover, we have proposed a pilot signal configuration for the channel estimation in the discrete frequency domain for the SC-CP system with the insufficient GI. Furthermore, the validity of the proposed transmission and the channel estimation schemes are confirmed via computer simulations. From all the results it can be concluded that the proposed schemes could be a simple but powerful solution for the SC-CP system with the insufficient GI.

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References

[1] S. Hara and P. Prasad, "Overview of multicarrier CDMA," *IEEE Commun. Mag.*, vol.35, pp.126-133, Dec. 2002.

[2] H. Sari, G. Karam and I. Jeanclaude, "Transmission Techniques for Digital Terrestrial TV Broadcasting," *IEEE Commun. Mag.*, vol.33, pp.100-109, Feb. 1995.

[3] D. Falconer et al., "Frequency Domain Equalization for Single-Carrier Broadband Wireless System," *IEEE Commun. Mag.*, vol.40, pp.58-66, Apr. 2002.

[4] Z. Wang and G. B. Giannakis, "Wireless Multicarrier Communications," *IEEE Signal Processing Mag.*, vol.47, pp.126-144, Dec. 2001.

[5] P. J. W. Melsa, R. C. Younce, and C. E. Rohrs, "Impulse Response Shortening for Discrete Multitone Transceivers," *IEEE Trans. Commun.,* vol. 44, pp. 1662-1672, Dec. 1996.

[6] K. Hayashi, T. Kojima and H. Sakai, "A New Beamforming Method for Single Carrier Modulation System with Cyclic Prefix," *Proceedings of the 6th Baiona Workshop on Communications*, pp. 75-80, Baiona, SPAIN, Sept. 2003.

[7] K. Van Acker, G. Leus, M. Moonen, O. van de Wiel, and T. Pollet, "Per Tone Equalization for DMT-Based Systems," *IEEE Trans. Commun.*, vol. 49, pp. 109-119, Jan. 2001.

[8] G. Leus and M. Moonen, "Per-Tone Equalization for MIMO OFDM Systems," *IEEE Trans. Signal Processing*, vol. 51, pp. 2965-2975, Nov. 2003.

[9] P. S. Chow, J. M. Cioffi, and J. A. C. Bingham, "A Practical Discrete Multitone Transceiver Loading Algorithm for Data Transmission over Spectrally Shaped Channels," *IEEE Trans. Commun.*, vol. 43, pp. 773-775, Feb. 1995.

[10] P. J. Davis, *Circulant Matrices*, John Willey & Sons,1979.

[11] J. Choi, "A Joint Channel Estimation and Detection for Frequency-Domain Equalization Using an Approximate EM Algorithm," *Signal Processing*, vol. 84, pp. 865-880, 2004.