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A bandwidth-efficient method for cancellation of ICI in OFDM systems

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Abstract

Orthogonal frequency division multiplexing (OFDM) is a very important modulation technique in wideband wireless communication and multimedia communication systems. While, it can effectively deal with multipath delay spread produced by frequency fading channels, its main drawback is the effect of frequency offset (FO) produced by the receiver local oscillator or by motion-induced Doppler. The FO breaks the orthogonality among the subcarriers and hence causes intercarrier interference (ICI). In this paper, ICI caused by frequency drift is eliminated by equalizing the complex weighting coefficients of interference. In most of the commonly used ICI cancellation schemes, bandwidth efficiency suffers because of the requirement of redundancy in the transmission. In the proposed scheme, repetition of data symbols or transmission of training sequence is not required. Thus, the bandwidth efficiency of normal OFDM system is maintained. The improved performance of the present scheme is confirmed through extensive simulations.

Keywords: OFDM; Intercarrier interference; Self-ICI cancellation; Interference coefficients equalization

1. Introduction

Orthogonal frequency division multiplexing (OFDM) is becoming the chosen multi-carrier modulation technique for wireless and multimedia communication systems. Multimedia wireless services require high-bit-rate transmission over mobile radio channels [1,2]. OFDM can provide large data rates with sufficient robustness to radio channel impairments. OFDM is a method that allows to transmit high data rates over extremely hostile channels at a comparable low complexity [3]. The OFDM has been used in wireless LAN standards such as American IEEE802.11a and European equivalent HYPERLAN/2 and in multimedia wireless services such as Japanese Multimedia Mobile Access Communications.

In OFDM, the entire bandwidth is divided into N small parts, and a block of N data symbols is modulated on N corresponding subcarriers, which are orthogonal to each other. These sub-channels are transmitted in parallel, thereby increasing the symbol duration and reducing the intersymbol interference (ISI). In OFDM systems the effect of ISI can be significantly reduced by cyclically extending the OFDM symbol.

Therefore, OFDM is a very attractive technique for the transmission of the high-bit-rate data. The major limitation of the OFDM is that it is very sensitive to frequency errors caused by frequency differences between the local oscillators in the transmitter and receiver. The FO may also arise due to motion-induced Doppler. Since in an OFDM system a block of N data symbols is modulated into N corresponding subcarriers, which are orthogonal to each other, the FO will destroy the orthogonality of the subcarriers and in turn induce the intercarrier interference and degrade the system performance significantly [4–7].

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A number of methods, namely self-ICI cancellation schemes [8–11], discrete Fourier transform (DFT)-based ICI cancellation scheme [12], and channel estimation and frequency domain equalization [13,14], have been developed to reduce this sensitivity to FO. The self-ICI cancellation scheme requires repetition of input data which reduces the bandwidth efficiency of the normal OFDM system. The DFT-based scheme performs well only for a small FO. The channel estimation methods require pilot signals, which also limits the throughput.

In the present work, a new bandwidth-efficient ICI cancellation scheme is presented. The proposed scheme cancels interference in OFDM systems by equalization of interference coefficients without requiring any training data.

The paper is organized as follows. First we introduce the structure of the OFDM system in Section 2. In Section 3, the self-ICI cancellation schemes of [8,9] and the DFT-based ICI cancellation scheme of [12] are briefly discussed and analyzed. Then the proposed scheme is introduced in Section 4. Section 5 presents the analysis of the performance of these schemes through simulation. The efficacy of all the schemes considered in this study is judged on the basis of symbol error rate (SER) at different values of signal to noise ratio (SNR) and FO. Finally, the overall findings of the study are summarized in Section 6.

2. System model

Fig. 1 shows the structure of the OFDM communication system. The N bits of high-speed input data signal $X_0, X_1, \ldots, X_{N-1}$ modulate the N orthogonal subcarriers to produce the output of the OFDM transmitter for the ith-transmitted symbol as

$$x(t) = \mathrm{e}^{(j2\pi f_c t)} \sum_{k=0}^{N-1} x_{k,i}\, p(t - kT/N) \tag{1}$$

where f_c is the carrier frequency and $p(t)$ is the impulse response of the low-pass filter in the transmitter and the inverse discrete Fourier transform (IDFT) $x_{k,i}$ of $X_{k,i}$, in the ith symbol, is given by

$$x_{k,i} = \frac{1}{N} \sum_{l=0}^{N-1} X_{l,i} \exp(j2\pi kl/N) \quad \text{for } k = 0, 1, \ldots, N-1 \tag{2}$$

If t_s is the bit duration of the input, the OFDM symbol duration T will be equal to Nt_s. Thus the duration of the OFDM symbol is increased by N times the duration of the input data bit. This greatly reduces the effect of ISI [1]. To reduce the effect of ISI completely, the cyclic prefixing of the OFDM symbol is done. At the receiver, the noisy signal received is mixed with a local oscillator signal having frequency Δf above the correct frequency f_c. Ignoring the effect of noise as considered in [8], the demodulated signal is given by

$$y(t) = \mathrm{e}^{(j2\pi\Delta f t + \theta_0)} \sum_{k=0}^{N-1} x_{k,i}\, q(t - kT/N) \tag{3}$$

where $q(t)$ is the combined impulse response of the channel, transmitter, and receiver filters. θ_0 is the difference between the phase of the receiver local oscillator and the carrier phase at the start of the received symbol. As in [8], we assume that the $q(t)$ satisfies the Nyquist criterion for samples taken at intervals T/N. If $y(t)$ is sampled at the optimum instant, then the samples of $y(t)$ are given by

$$y_{k,i} = \mathrm{e}^{(j\theta_0)} x_{k,i} \mathrm{e}^{(j2\pi k\varepsilon/N)} \tag{4}$$

where $\varepsilon = \Delta f T$ is the normalized FO.

These samples are applied to the receiver DFT, after removing the cyclic prefix. The output of receiver DFT is again converted into serial form. The output of the DFT is given by

$$Y_{m,i} = \sum_{k=0}^{N-1} y_{k,i} \mathrm{e}^{(-j2\pi km/N)} \tag{5}$$

Then from (4) and (5) it can be shown that

$$Y_{m,i} = \frac{1}{N} \mathrm{e}^{(j\theta_0)} \sum_{l=0}^{N-1} X_{l,i} \times \sum_{k=0}^{N-1} \exp(j2\pi k(l - m + \varepsilon)/N) \tag{6}$$

After some simplification this can also be expressed as

$$Y_{m,i} = \mathrm{e}^{(j\theta_0)} \sum_{l=0}^{N-1} c_{l-m} X_{l,i} \tag{7}$$

where

$$c_{l-m} = \frac{1}{N} \sum_{k=0}^{N-1} \exp(j2\pi k(l - m + \varepsilon)/N) \tag{8}$$

From (7), it can be seen that if $\varepsilon = 0$ then $Y_{m,i} = \mathrm{e}^{(j\theta_0)} X_{m,i}$ and each decoded value is simply the phase-rotated version of the transmitted value. If $\varepsilon \neq 0$, the ICI will occur and each output-decoded value will depend on all the input values. The decoded output $Y_{m,i}$, therefore, consists of a wanted component, which is due to $X_{m,i}$ but is subject to a change in the phase and amplitude given by

$$c_0 = \frac{1}{N} \sum_{k=0}^{N-1} \exp(j2\pi\varepsilon/N) \tag{9}$$

Thus, the ICI depends on the N complex weighting coefficients $c_0, c_1, \ldots, c_{N-1}$. A plot of real and imaginary coefficients of these weighting coefficients, for $N = 16$, is shown

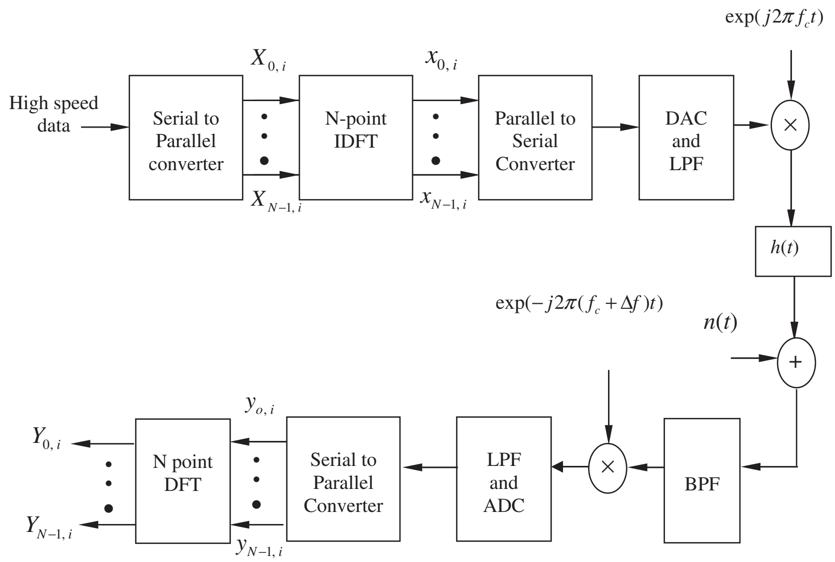

Fig. 1. OFDM system model.

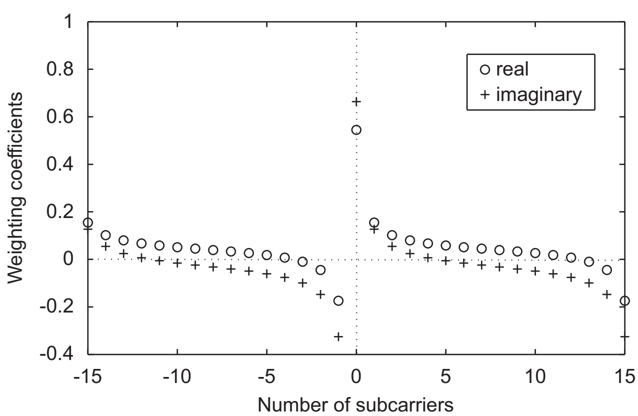

Fig. 2. Weighting coefficients for $N = 16$ and $\Delta f T = 0.2$.

in Fig. 2. It can be observed from this figure that the variation in the weighting coefficients is smooth as the distance moves from -15 to $+15$, except between -1 and 0 and between 0 and 1. The main component of the symbol corresponds to the weighting coefficient at zero.

The received signal can also be written in the matrix form, using (7) as

$$\mathbf{Y} = \mathbf{CX} \tag{10}$$

where $\mathbf{C}$ is given by

$$\mathbf{C} = \begin{bmatrix} c_0 & c_1 & \cdot & \cdot & c_{N-1} \\ c_{-1} & c_0 & \cdot & \cdot & c_{N-2} \\ c_{-2} & c_{-1} & c_0 & \cdot & c_{N-3} \\ \cdot & \cdot & \cdot & \cdot & \cdot \\ c_{-N+1} & c_{-N+2} & \cdot & \cdot & c_0 \end{bmatrix} \tag{11}$$

where $\mathbf{X} = [X_{0,i}, \ldots, X_{N-1,i}]^{\mathrm{T}}$ is the input data and $\mathbf{Y} = [Y_{0,i}, \ldots, Y_{N-1,i}]$ is the received data.

3. ICI cancellation schemes

The ICI cancellation schemes, which do not require channel estimation, are briefly revised here.

3.1. Self-ICI cancellation scheme

This scheme describes a method of reducing sensitivity to frequency errors. In this scheme the input data are mapped onto adjacent pair of subcarriers rather than onto a single subcarrier, such that $X_{o,i} = -X_{1,i}$, $X_{2,i} = -X_{3,i}, \ldots, X_{N-2} = -X_{N-1,i}$ as proposed in [8]. The decoded value of the ith carrier is expressed as a function of weighting coefficients as

$$Y_{0,i} = \exp(j\theta_0)\{(c_0 - c_1)X_{0,i} + (c_2 - c_3)X_{2,i} + \cdots + (c_{N-2} - c_{N-1})X_{N-2,i}\} \tag{12}$$

The ICI now depends on the difference between the adjacent weighting coefficients rather than on the coefficients themselves. As the difference between adjacent coefficients is small, this results in a substantial reduction in ICI. In this process the component of ICI, which is constant between an adjacent pair of subcarriers, is cancelled out. This scheme, termed as constant ICI cancellation, improves the performance at any FO, because the ICI cancellation depends only on the coefficients being a slowly varying function of offset.

To maximize the overall SNR, the values of output symbols can be subtracted in pairs, resulting in further reduction of ICI:

$$Y_{0,i} - Y_{1,i} = \exp(j\theta_0)\{(-c_{-1}+2c_0-c_1)X_{0,i} + \cdots + (-c_{N-3}+2c_{N-2}-c_{N-1})X_{N-2,i}\} \tag{13}$$

In this process the remaining ICI depends on the factors of the form $\{-c_1 + 2c_2 - c_3\}$. If the three weighting coefficients in each term were linear functions of offset, the ICI would completely be zero. This is termed as the linear ICI cancellation scheme [8].

3.2. DFT-based ICI cancellation scheme

In this scheme, the input data symbols, after serial to parallel conversion, are divided into two groups, each of size $N/2$. These groups are separately discrete Fourier transformed; after this, the two groups are again combined and then N point IDFT is taken. At the receiver side, after demodulation, the N point DFT of the received data symbols is taken and after this the symbols are divided into two groups again and each inverse-discrete Fourier transformed, respectively, to rebuild the transmitted symbols. It has been shown in [12] that by doing this ICI is greatly reduced.

For the DFT-based scheme the transmitted data can be described in matrix form as

$$\mathbf{x} = \frac{1}{\sqrt{N}}\mathbf{W}_N^{\mathrm{H}}\mathbf{B}\mathbf{X} \tag{14}$$

where $\mathbf{W}_N^{\mathrm{H}}$ is the $N \times N$ IDFT matrix, $\mathbf{B}$ is the $N \times N$ matrix obtained after combining the two $N/2$ DFT blocks, $\mathbf{X} = [X_{0,i}, X_{1,i}, \ldots, X_{N-1,i}]^{\mathrm{T}}$, and $^{\mathrm{H}}$ denotes the Hermitian transpose. The data received are described in matrix form as

$$\mathbf{Y} = \mathbf{C}^{(1)}\mathbf{X} \tag{15}$$

where $\mathbf{C}^{(1)} = 1/N\mathbf{B}^{\mathrm{H}}\mathbf{W}_N\mathbf{\Lambda}\mathbf{W}_N^{\mathrm{H}}\mathbf{B}$ and $\mathbf{\Lambda} = \mathrm{diag}(1, \mathrm{e}^{j2\pi\varepsilon/N}, \mathrm{e}^{j4\pi\varepsilon/N}, \ldots, \mathrm{e}^{j2(N-1)\pi\varepsilon/N})$.

A plot of interfering coefficients of matrix $\mathbf{C}^{(1)}$ is shown in Figs. 3 and 4 for $N=16$ and at $\varepsilon=0.1\&0.2$ respectively. Fig. 3 shows that the value of the coefficient corresponding to the main symbol is significant compared to other interfering coefficients at small FOs, for $\varepsilon \leqslant 0.1$. It can also be noted that the coefficient corresponding to the main symbol decreases and the interfering coefficients increase at $\varepsilon = 0.2$. Thus the scheme is effective only at small FOs.

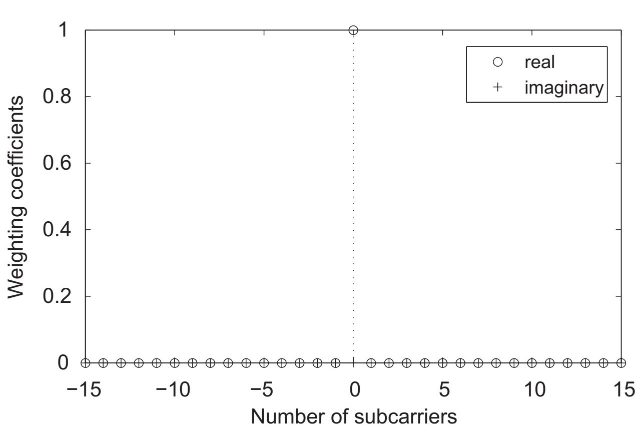

Fig. 3. Weighting coefficients of matrix $\mathbf{C}^{(1)}$ for $N = 16$ and $\Delta fT = 0.1$.

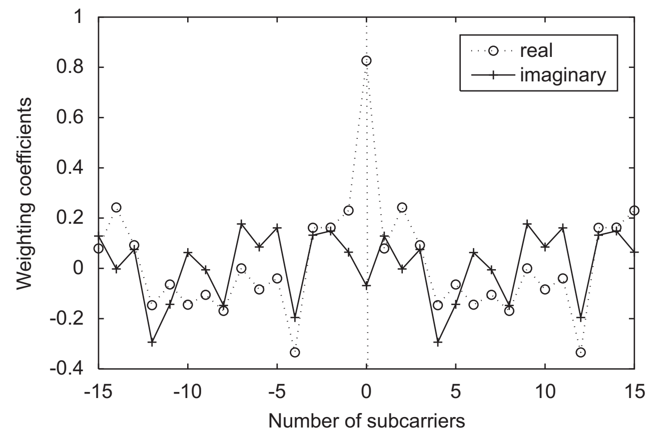

Fig. 4. Weighting coefficients of matrix $\mathbf{C}^{(1)}$ for $N = 16$ and $\Delta fT = 0.2$.

4. Proposed scheme

As discussed in the previous section, the ICI depends on the weighting coefficients $c_o, c_1, \ldots, c_{N-1}$. In order to design an equalizer for equalizing the weighting coefficients that cause ICI, the matrix of these coefficients is obtained from (11). The received signal at the receiver in the matrix form is given by

$$\mathbf{Y} = \mathbf{W}_N\mathbf{\Lambda}\mathbf{W}_N^{-1}\mathbf{X}$$

or $\mathbf{Y} = \mathbf{C}\mathbf{X}$ where $\mathbf{W}_N$ is an $N \times N$ DFT matrix and $\mathbf{W}_N^{-1}$ is an $N \times N$ IFDT matrix.

When an equalizer is used, the output is given by

$$\mathbf{Z} = \mathbf{D}\mathbf{Y} \tag{16}$$

where $\mathbf{D} = \mathbf{C}^{-1}$ is a circulant matrix, which denotes the equalizer filter matrix.

Commonly, the equalizer is designed by using training symbols. In the present scheme, however, an appropriate filter is chosen from a small set of predefined filters. This is based on the observation that an equalizer designed for a given value of ε is effective for a variation $\pm\Delta\varepsilon$ around ε. If the maximum expected value of normalized FO is $\varepsilon_{\max}$, then the given range of ε can be divided into p segments, where $p = \varepsilon_{\max}/2\Delta\varepsilon$. Let the midpoint of these segments be denoted by $\varepsilon^{(i)}$ $i = 1, 2, \ldots, p$. Now the equalizers $\mathbf{D}^{(\mathbf{i})}$ for each $\varepsilon^{(i)}$ $i = 1, 2, \ldots, p$ can be obtained from (16). The appropriate equalizer from this set of predefined equalizers $\mathbf{D}^{(\mathbf{i})}$, $i=1, 2, \ldots, p$ is selected according to the the following criterion:

$$\underset{i}{\text{minimize}}\{J(\mathbf{D}^{(i)}) = E[|\mathbf{Z}^{(\mathbf{i})} - \mathbf{g}(\mathbf{Z}^{(\mathbf{i})})|^2]\} \tag{17}$$

with $\mathbf{Z}^{(\mathbf{i})} = \mathbf{D}^{(\mathbf{i})}\mathbf{Y}$ where $\mathbf{g}(\cdot)$ is a nonlinear estimator of transmitted symbols and $E\{\cdot\}$ denotes the expectation.

Theorem. *If the objective function of* (17) *is to be used for selection of the equalizer*, *then the nonlinear estimator*

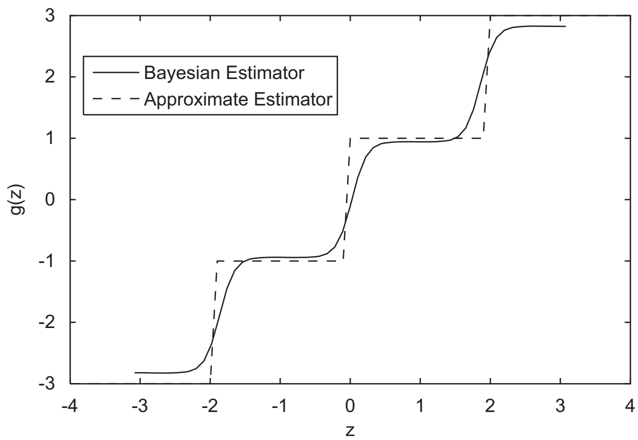

Fig. 5. Plot of nonlinear Bayes estimator $g(Z)$ and its approximation.

g(·) *will depend upon the probability density functions of transmitted symbols and convolution noise*.

Proof. Using the Bussgang principle of blind deconvolution, the equalizer must minimize an objective function of the type: $J(n) = E[e^2(n)]$, as shown in [15], where the error is given by $e(n) = (g(Z_k^{(i)}) - Z_k^{(i)})$; $k = 0, 1, \ldots, N-1$, and g(·) is an estimator of the transmitted symbol. The Bayes estimate of transmitted symbol X_k is given by

$$\hat{X}_k = E[X_k/Z_k] = \int_{-\infty}^{\infty} X_k f_X(X_k/Z_k)\,\mathrm{d}X_k \tag{18}$$

for $k = 0, 1, \ldots, N-1$, where $f_X(X_k/Z_k)$ denotes the conditional probability density function (pdf) of X, given Z. The observation Z_k is expressed as $Z_k = \alpha_0 X_k + v_k$, where v_k is the convolutional noise. In accordance with the statistical model for the convolutional noise v, X_k, v_k are statistically independent. With v modeled to have zero mean and variance σ^2, the scaling factor α_0 is given by $\alpha_0 = \sqrt{1-\sigma^2}$. Using the Bayes' rule, (18) becomes

$$\hat{X}_k = \frac{1}{f_Z(Z_k)} \int_{-\infty}^{\infty} X_k f_Z(Z_k/X_k) f_X(X_k)\,\mathrm{d}X_k \tag{19}$$

Furthermore, from $Z_k = \alpha_0 X_k + v_k$, it follows that

$$f_Z(Z_k/X_k) = f_v(Z_k - \alpha_0 X_k) \tag{20}$$

where $f_v(Z_k - \alpha_0 X_k)$ is the pdf of convolutional noise.

By using (19) and (20) we get the estimate of the transmitted symbol as

$$\hat{X}_k = g(Z_k) = \frac{1}{f_Z(Z_k)} \times \int_{-\infty}^{\infty} X_k f_v(Z_k - \alpha_0 X_k) f_X(X_k)\,\mathrm{d}X_k \tag{21}$$

For the uniformly distributed symbols $\mathbf{X}$, the plot of estimator g(·) and its approximation used in the simulation is shown in Fig. 5. With this approximation the criterion for selection of the equalizer, now becomes

$$\underset{i}{\text{minimize}}\{J(\mathbf{D}^{(i)}) = E[|\mathbf{Z}^{(\mathbf{i})} - \mathbf{Z}_{\mathbf{s}}^{(\mathbf{i})}|^2]\} \tag{22}$$

where $\mathbf{Z}_{\mathbf{s}}^{(\mathbf{i})}$ denotes the decision on $\mathbf{Z}^{(\mathbf{i})}$.

4.1. Complexity comparison

The complexity of the proposed scheme and the conventional equalizer with a training sequence using the LMS algorithm is shown in the following Table 1.

Alternatively, in place of the cost function defined in (17), another cost function such as the CMA cost function [15] can also be used. However, our simulation has shown that the cost function of (17) requires fewer samples to select the optimum equalizer in comparison to the CMA cost function. The plot of the proposed cost function and CMA cost function is shown in Fig. 6.

In the proposed scheme the cost function is merely used for the selection of the appropriate equalizer filter and is not minimized iteratively. It can be noted that the cost function is required to be differentiable when it is used for designing the equalizer using some higher-order statistics (HOS)-based blind equalization scheme such as CMA. These

Table 1. Comparison of complexity

Operations	Proposed scheme	Equalizer designed with a training sequence using the LMS algorithm
Number of complex additions	$(N^2+N-1)p$	(N^2+N-1) per iteration
Number of complex multiplications	$(N^2+N+1)p$	(N^2+N+1) per iteration
Number of nonlinear function	N	–

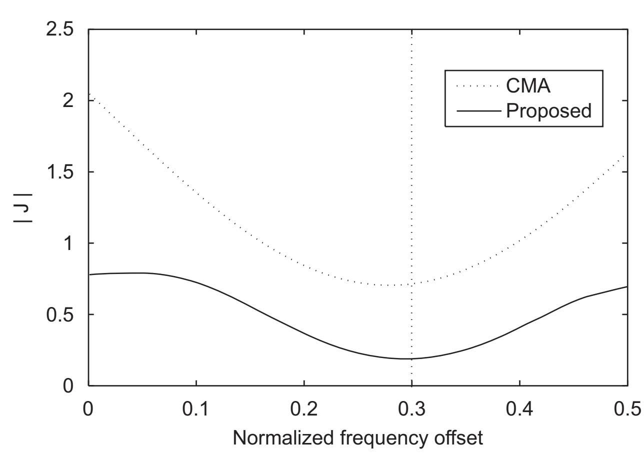

Fig. 6. CMA and proposed cost function.

schemes usually suffer from slow convergence and are not suitable for the present case.

5. Simulation and results

To evaluate the performance of various ICI cancellation schemes, considered in this paper, an OFDM system is simulated to implement these schemes. In the proposed scheme the appropriate equalizer is selected from a set of four filters designed at $\varepsilon^{(1)} = 0.05$, $\varepsilon^{(2)} = 0.1$, $\varepsilon^{(2)} = 0.15$ and $\varepsilon^{(3)} = 0.2$. For better accuracy a lager set of filters may be considered. The performance of various schemes of ICI cancellation is judged on the basis of the SER. In our simulation of the OFDM system, QPSK modulation with 64 subcarriers is considered. We assume the ideal reception of OFDM signals in the self-ICI cancellation scheme, and DFT-based ICI cancellation scheme as used in [8] and [12], respectively. For the computation of SER curves, the transmission of 10 000 symbols is considered.

The SER for various ICI cancellation schemes and standard OFDM is observed. The SER of standard OFDM, DFT-based OFDM, self-ICI cancellation scheme, and the proposed scheme is plotted in Fig. 7. This figure shows that the SER of standard OFDM, self-ICI cancellation, and DFT-based schemes is comparable at ε=0.05, while it is lower for the proposed scheme. As ε increases the SER of equalized OFDM, DFT-based, and the self-ICI cancellation scheme improves, as shown in Fig. 8. The improvement in SER of the proposed scheme is better than that of the standard OFDM, self-ICI cancellation scheme, and DFT-based ICI cancellation scheme. Figs. 9 and 10 show the SER for all these schemes at a higher value of ε. From Figs. 9 and 10, it is clear that the SER of the self-ICI cancellation scheme is better than the DFT-based ICI cancellation scheme. The main drawback of the self-ICI cancellation scheme is its throughput, which is just half as compared to standard and DFT-based ICI cancellation schemes. It can also be observed that the SER of the DFT-based ICI cancellation scheme is

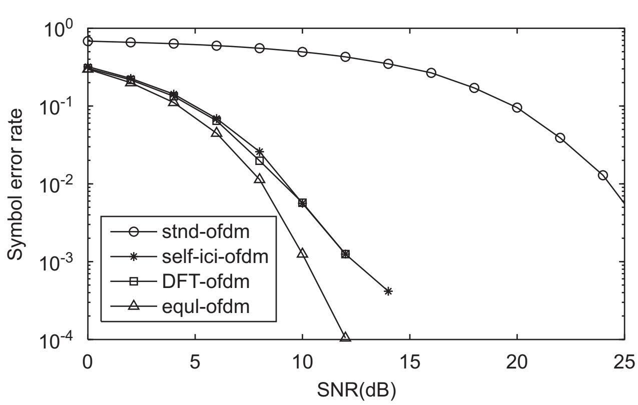

Fig. 7. SER of standard, self-ICI cancellation, DFT-based ICI cancellation, and equalized OFDM $\Delta f T = 0.05$.

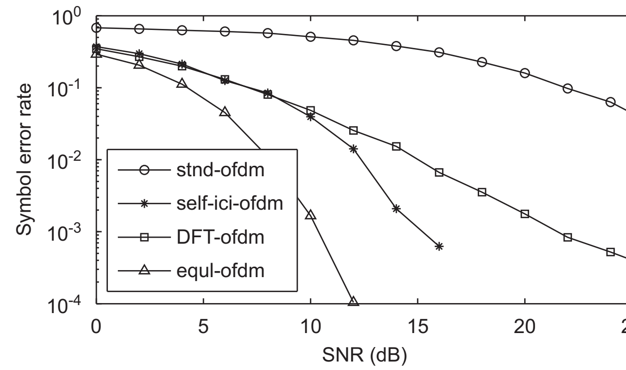

Fig. 8. SER of standard, self-ICI cancellation, DFT-based ICI cancellation, and equalized OFDM at $\Delta f T = 0.1$.

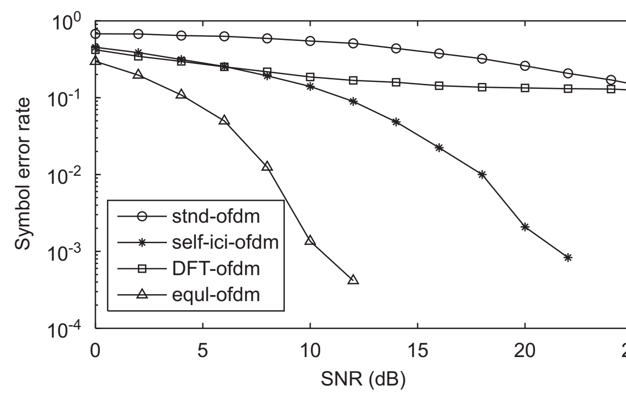

Fig. 9. SER of standard, self-ICI cancellation, DFT-based ICI cancellation, and equalized OFDM at $\Delta f T = 0.15$.

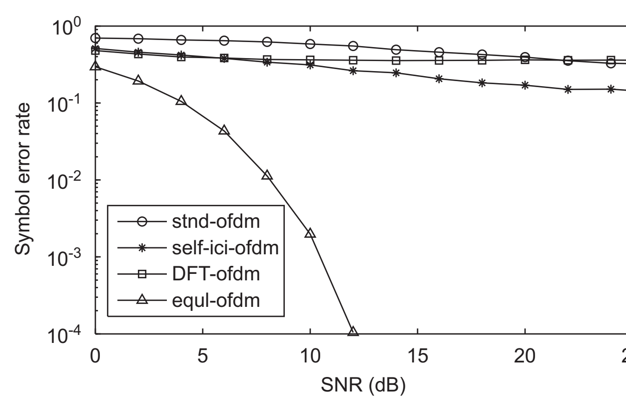

Fig. 10. SER of standard, self-ICI cancellation, DFT-based ICI cancellation, and equalized OFDM at $\Delta f T = 0.2$.

better than the standard OFDM only for low values of ε. As ε increases the SER of the DFT-based ICI cancellation scheme increases and becomes comparable to the SER of the standard OFDM.

It is revealed by Figs. 7–10 that the present scheme has better SER performance as compared to the standard

OFDM, self-ICI cancellation and DFT-based ICI cancellation schemes at all FOs. Also, the spectral efficiency of present scheme is maintained as that of the standard or the normal OFDM system.

6. Conclusions

A new scheme based on equalization of interference coefficients for cancellation of ICI in an OFDM system is presented. The proposed scheme does not need any pilot symbols for equalization, maintaining the high spectral efficiency of a standard OFDM system. A comparison with other popular methods of cancellation of ICI shows that the proposed scheme outperforms the other methods at all FOs. The self-ICI cancellation scheme of [8] results in a low SER as compared to the standard OFDM and DFT-based ICI cancellation scheme, but its throughput is just half as compared to the DFT-based scheme of [12] and the proposed scheme. On the other hand, the DFT-based scheme yields a low SER only at low values of FO, and is therefore better than the standard OFDM at low values of FO. However, at a high value of FO the performance of the DFT-based ICI cancellation scheme deteriorates over the standard OFDM. The SER resulting from the proposed method is very low as compared to the standard OFDM, self-ICI cancellation scheme, and DFT-based ICI cancellation scheme at any value of FO. Also, in the present scheme repetition of data is not required, and hence its spectral efficiency is maintained as that of the standard OFDM.

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