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Wiener-based ICI Cancellation Schemes for OFDM Systems over Fading Channels

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Abstract. The subcarriers of orthogonal frequency division multiplexing (OFDM) systems may fail to keep orthogonal to each other under time-varying channels. The loss of orthogonality among the subcarriers will degrade the system performace, and this effect is named intercarrier interference (ICI). In this paper, a Wiener-based successive interference cancellation (SIC) scheme is proposed to detect the OFDM signals. It provides good ICI cancellation performance; however, it suffers large computation complexity. Therefore, a modified Wiener-based SIC scheme is further proposed to reduce the computation complexity. Simulation results show the performance of the Wiener-based SIC scheme is better than those of zero forcing, zero forcing plus SIC and original Wiener-based schemes. Furthermore, with the modified Wiener-based SIC scheme, the performance is still better than the others. Although the performace of the modified Wiener-based SIC scheme suffers little degradation compared to Wiener-based SIC scheme, the computation complexity can be dramatically reduced.

Keywords: Orthogonal frequency division multiplexing (OFDM), fading channels, intercarrier interference (ICI), Wiener-based, successive interference cancellation (SIC).

1. Introduction

Orthogonal frequency division multiplexing (OFDM) has been applied in many digital transmission systems, such as digital audio broadcasting (DAB) system, digital video broadcasting terrestrial TV (DVB-T) system, asymmetric digital subscriber line (ADSL), IEEE 802.11a/g wireless local area network (WLAN), IEEE 802.16 worldwide interoperability for microwave access (WiMax) systems, and ultra-wideband (UWB) systems [1-6]. It can also be applied to cooperative communication systems [7]. OFDM systems split a high-rate data stream into numbers of low-rate data stream. Since the available channel is divided into several narrowband subchannels, OFDM systems have such advantages: immunity to delay spread, resistance to frequency selective fading, simple equalization, and efficient bandwidth usage. However, OFDM systems have several disadvantages: the problem of synchronization; hardware complexity of FFT units at transmitter and receiver; the problem of high peak to average power ratio (PAPR); intercarrier interference (ICI) effect. The performance degrades significantly for intercarrier interference, and several methods have been proposed to mitigate the ICI effect with different efficiency and complexity.

The remainder of the paper is organized as follows. Related work is given in Section 2. In section 3, channel model of OFDM system is introduced. In section 4, signal detection and interference cancellation schemes are introduced. The simulation results are shown in section 5. Finally, the conclusion is given in section 6.

2. Related Work

Carrier frequency offset, caused by Doppler shift, and time-varying channel bring the intercarrier interference. Several ICI cancellation schemes have been proposed, and ZF (zero forcing) detection scheme is one of them. Although conventional ZF detection scheme is widely used in noise free environment, the noise enhancement occurs while suppressing the ICI effect. Wiener solution has been proved to be able to detect signals without noise enhancement [8]. On the other hand, successive interference cancellation scheme has been successfully used in MC-CDMA and OFDM systems to mitigate multiple access interference and intercarrier interference respectively [9-10]. In this paper, we first study the performance of Wiener-based SIC for OFDM systems over fading channels. Although the Wiener-based SIC scheme can provide good ICI cancellation performance, its computation complexity increases as number of subcarriers increases [11]. This is a trade-off between bit error rate (BER) performance and computation complexity. Therefore, we further study a modified Wiener-based SIC ICI cancellation scheme to reduce computation complexity without reducing BER performance or with minor BER performance degradation.

3. Channel Model

The block diagram of OFDM system shown in Fig. 1 has several propagation paths between transmitter and receiver. The schematic of multipath communication environment is shown in Fig. 2. Each path introduces different phase, amplitude attenuation, delay and Doppler shift to the signal.

Since the transmission environment is time-varying; therefore, the phase, attenuation, delay and Doppler shift of the signal are random variables.

For a time-varying multipath channel, the impulse response could be expressed as:

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

where the amplitude of $h_l(t)$ is modeled as Rayleigh distribution with the maximum Doppler shift f_d, and it denotes the channel impulse response as l-th delay path at the time t. According to (1) the time delay and the attenuation are function of time. The Fig. 3 shows the time-varying channel. In the mobile radio channels, the Rayleigh distribution is commonly used to describe the statistical time-varying channel. It is well known the envelope of sum of two quadrature Gaussian noise signals obeying a Rayleigh distribution. Fig. 4 shows a Rayleigh distributed signal envelope as a function of time.

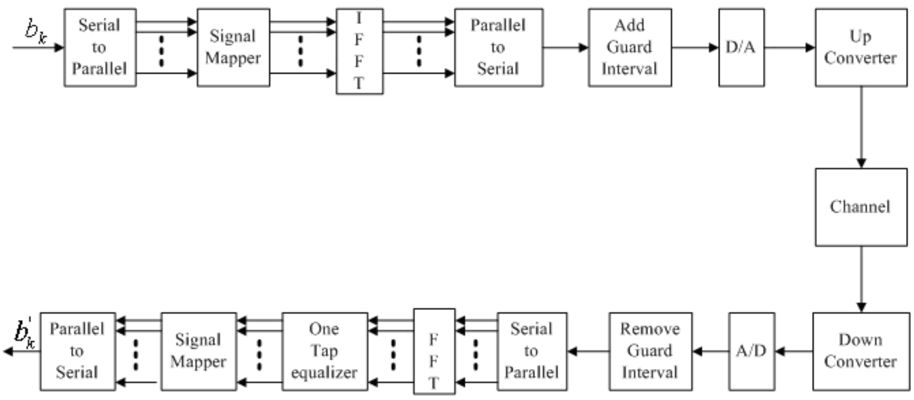

Fig. 1. Block diagram of OFDM systems

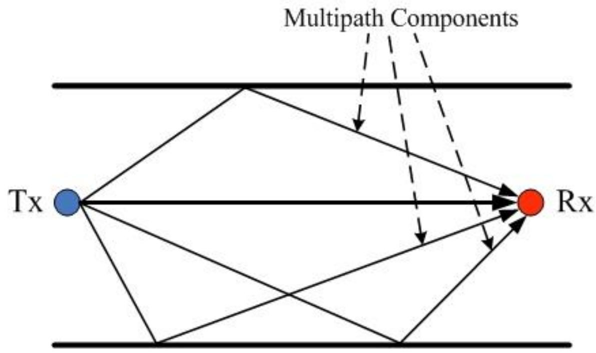

Fig. 2. The multipath environment

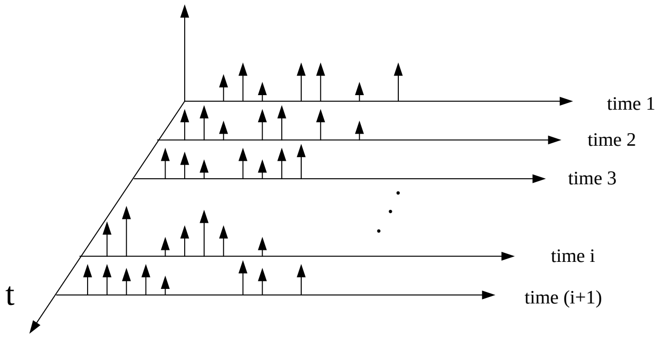

Fig.3. The impulse response of the time-varying channel

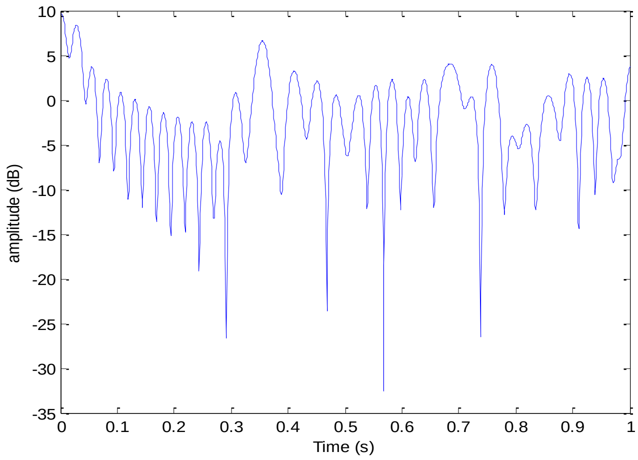

Fig. 4. Rayleigh distributed signal envelope

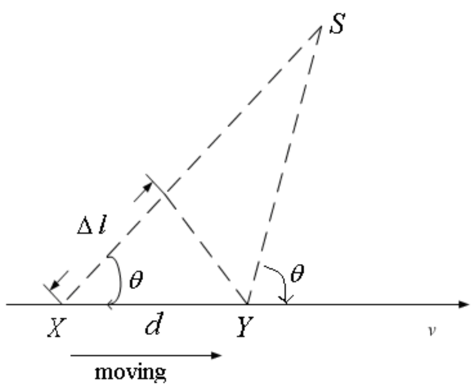

Fig. 5. Illustration of the Doppler effect

As shown in Fig. 5, consider a mobile moving at a constant velocity v, along a path with length d between point X and Y, it receives signal from a remote source S. The difference in path lengths traveled by the signal wave from source S to the mobile at point X and Y is $\Delta l = d\cos\theta = v\Delta t\cos\theta$, where Δt is the time required for the mobile to travel from X and Y, and the angles θ are assumed to be the same at points X and Y since the source is assumed to be very far away from the mobile. Therefore, the phase change in the received signal due to the difference in path lengths is

$$\Delta\phi = \frac{2\pi\Delta l}{\lambda} = \frac{2\pi v\Delta t}{\lambda}\cos\theta, \tag{2}$$

and the apparent change in frequency, or Doppler shift is given by f_m, where

$$f_m = \frac{v}{\lambda}\cos\theta = \frac{v}{c} f_c \cos\theta, \tag{3}$$

where c is velocity of light and f_c is carrier frequency. The Doppler shift f_d could be maximized when $\cos\theta$ is equal to 1.

The time-varying channel is expressed in (1), and the time-varying path gain $h_l(t)$ is generally represented by a Rayleigh random process. For the classical Doppler spectrum [12], the spectral density of $h_l(t)$ is

$$p(f) = \begin{cases} \dfrac{1}{\pi f_d \sqrt{1-\left(\dfrac{f}{f_d}\right)^2}} & \text{, if } |f| < f_d \\ 0 & \text{, else} \end{cases}, \tag{4}$$

where f_d is the maximum Doppler frequency. Therefore, if the channel is time-varying, the ICI would be occurred.

In the orthogonal frequency division multiplexing (OFDM) system, the transmission bandwidth is divided into many narrow subchannels, and they are transmitted in parallel. Because the bandwidth of subchannel is very narrow, channel response could be seen constant. In contrast to time domain, the symbol duration increases, such that the intersymbol interference (ISI) would be happened. If the guard interval is greater than the maximum delay path, the ISI will be removed. This is the reason why OFDM could against the frequency selective fading. Increase in the symbol duration makes it much more vulnerable to time selective fading due to the Doppler spread effect.

The output of IFFT (inverse fast Fourier transform) in OFDM system could be expressed as:

$$x(n) = \frac{1}{N}\sum_{k=0}^{N-1} X(k) e^{j\frac{2\pi}{N}nk} \qquad , n = 0,1,2,\ldots,N-1 \tag{5}$$

The transmitted signal could be represented as:

$$x(t)=\frac{1}{N}\sum_{k=0}^{N-1}X(k)e^{j\frac{2\pi}{T}kt} \qquad , -T_{g}\le t<T \tag{6}$$

where the interval $-T_g \leqq t < 0$ is the guard interval for opposing to the intersymbol interference (ISI).

Then the received signal $r(t)$ could be obtained as:

$$\begin{aligned} r(t) &= (x(t)e^{j2\pi f_c t}) * h(t,\tau) + w'(t) \\ &= \int_{-\infty}^{\infty} h(t,\tau')x(t-\tau')e^{j2\pi f_c(t-\tau')}d\tau' + w'(t) \\ &= \int_{-\infty}^{\infty}\sum_{l=0}^{L-1} h_l(t)\delta(\tau'-\tau_l)x(t-\tau')e^{j2\pi f_c(t-\tau')}d\tau' + w'(t) \\ &= \sum_{l=0}^{L-1} h_l(t)x(t-\tau_l)e^{j2\pi f_c(t-\tau_l)} + w'(t). \end{aligned} \tag{7}$$

where f_c is the carrier frequency.

The signal in the output of down converter is

$$\begin{aligned} y'(t) &= r(t)e^{-j2\pi(f_c-\Delta f)t} \\ &= \left[\sum_{l=0}^{L-1} h_l(t)x(t-\tau_l)e^{j2\pi f_c(t-\tau_l)} + w'(t)\right]e^{-j2\pi(f_c-\Delta f)t} \\ &= \sum_{l=0}^{L-1} h_l(t)x(t-\tau_l)e^{-j2\pi f_c\tau_l}e^{j2\pi\Delta f t} + w'(t)e^{-j2\pi(f_c-\Delta f)t} \end{aligned} \tag{8}$$

where Δf is the carrier frequency offset. After passing through the lowpass filter, the signal could be obtained as:

$$y(t)=\sum_{l=0}^{L-1} h_l(t)x(t-\tau_l)e^{j2\pi\Delta f t} + w(t) \tag{9}$$

Sampling the received signal $y(t)$ with the rate N/T and removing the portion of cyclic prefix, the received signal could be obtained as $y(n)$=$y(T/N)$, n =0, 1, 2,…, N-1, within one symbol interval. The received signal could be rewritten as:

$$\begin{aligned} & y(n\frac{T}{N}) = \sum_{l=0}^{L-1} h_l(n\frac{T}{N})x(n\frac{T}{N}-l\frac{T}{N})e^{j2\pi\Delta f(n\frac{T}{N})} + w(n\frac{T}{N}) \\ & \Rightarrow y(n) = \sum_{l}^{L-1} h_l(n)x(n-l)e^{j\frac{2\pi}{N}\varepsilon n} + w(n), \quad n=0,1,2,\ldots,N-1 \end{aligned} \tag{10}$$

where $|\tau_{l+1}-\tau_l|$ is assumed equal to T/N, and the normalized frequency offset is represented as ε=$\Delta f\times T$=$\Delta f / f$ in which T is the symbol duration, and f is the subcarrier spacing.

The FFT of $y(n)$ could be expressed as:

$$
\begin{aligned}
Y(m) &= \sum_{n=0}^{N-1} y(n) e^{-j\frac{2\pi}{N}nm} \\
&= \sum_{n=0}^{N-1}\left(\sum_{l=0}^{L-1} h_l(n) x(n-l) e^{-j\frac{2\pi}{N}\varepsilon n} + w(n)\right) e^{-j\frac{2\pi}{N}nm} \\
&= \sum_{n=0}^{N-1}\left(\sum_{l=0}^{L-1} h_l(n)\left(\sum_{k=0}^{N-1}\frac{1}{N} X(k) e^{j\frac{2\pi}{N}k(n-l)}\right) e^{-j\frac{2\pi}{N}\varepsilon n} + w(n)\right) e^{-j\frac{2\pi}{N}nm} \\
&= \sum_{n=0}^{N-1}\left(\sum_{l=0}^{L-1} h_l(n)\left(\sum_{k=0}^{N-1}\frac{1}{N} X(k) e^{j\frac{2\pi}{N}k(n-l)}\right) e^{-j\frac{2\pi}{N}\varepsilon n}\right) e^{-j\frac{2\pi}{N}nm} + W(m) \\
&= \sum_{n=0}^{N-1}\left(\sum_{l=0}^{L-1}\left(\frac{1}{N}\sum_{k=0}^{N-1} h_l(n) e^{j\frac{2\pi}{N}(k+\varepsilon-m)n}\right) e^{-j\frac{2\pi}{N}kl}\right) X(k) + W(m)
\end{aligned}
\tag{11}
$$

where the ICI term is defined as:

$$
\begin{aligned}
\mathrm{ICI} &= \sum_{l=0}^{L-1}\left(\frac{1}{N}\sum_{k=0}^{N-1} h_l(n) e^{j\frac{2\pi}{N}(k+\varepsilon-m)n}\right) e^{-j\frac{2\pi}{N}kl} \\
&= \left(\sum_{l=0}^{L-1} h_l(n) e^{-j\frac{2\pi}{N}kl}\right)\sum_{n=0}^{N-1}\frac{1}{N} e^{j\frac{2\pi}{N}(k+\varepsilon-m)n}
\end{aligned}
\tag{12}
$$

According to the (12), it is clear that if there is no frequency offset, $\varepsilon=0$, and the channel is stationary, $h_l(n)=h_l$, then the ICI= $h(k)\delta(m\text{-}k)$, there will be no intercarrier interference. If there is no frequency offset, $\varepsilon=0$, due to the time-varying channel fading characteristic of the mobile channel, ICI would exist in OFDM systems for the mobile application. In contrarily, the channel is stationary but the frequency offset is not equal to zero, ICI would still exist in OFDM systems.

In this paper we focus on the time-varying channel fading characteristic of the mobile channel, so we set frequency offset ε equal to zero. In time-varying channels, the ICI term is defined as:

$$
\begin{aligned}
\mathrm{ICI} &= \sum_{l=0}^{L-1}\left(\frac{1}{N}\sum_{k=0}^{N-1} h_l(n) e^{j\frac{2\pi}{N}(k-m)n}\right) e^{-j\frac{2\pi}{N}kl} \\
&= \sum_{l=0}^{L-1}\big(H_l(m-k)\big) e^{-j\frac{2\pi}{N}kl}.
\end{aligned}
\tag{13}
$$

where $H_l(m\text{-}k)$ is the ICI effect of the k-th subcarrier to the m-th subcarrier. Then output of FFT (fast Fourier transform) is also written as:

$$
\begin{aligned}
Y(m) &= \sum_{k=0}^{N-1}\left(\sum_{l=0}^{L-1}\big(H_l(m-k)\big) e^{-j\frac{2\pi}{N}kl}\right) X(k) + W(m) \\
&= \sum_{l=0}^{L-1}\big(H_l(0)\big) e^{-j\frac{2\pi}{N}kl} X(m) + \sum_{\substack{k=0\\k\neq m}}^{N-1}\left(\sum_{l=0}^{L-1}\big(H_l(m-k)\big) e^{-j\frac{2\pi}{N}kl}\right) X(k) + W(m)
\end{aligned}
\tag{14}
$$

where the first term is the desired signal and the second term is the ICI component.

For a time-varying fading channel, the channel variations would lead to the loss of orthogonality between subcarriers, hence the ICI effect could be occurred. The ICI effect in OFDM systems would result in an error floor. In next section, we would propose a scheme to suppress ICI.

4. Signal Detection and Interference Cancellation Schemes

4.1. Zero-forcing Detection Scheme

The received signals $y(n)$ in the time-varying channel could be obtained as:

$$y(n)=\sum_{l=0}^{L-1} h_l(n)x(n-l)+w(n) \qquad , n=0,1...,N-1 \tag{15}$$

The received signals can be represented as a matrix form, and it is represented as:

$$\mathbf{y}=\mathbf{hx}+\mathbf{w}. \tag{16}$$

Each element of the received signal $\mathbf{y}$, the channel matrix $\mathbf{h}$, the transmitted signal $\mathbf{x}$, and the AWGN (additive white Gaussian noise) $\mathbf{w}$ can be expressed as:

$$\begin{bmatrix} y(0) \\ y(1) \\ \vdots \\ y(N-1) \end{bmatrix} = \begin{bmatrix} h_0(0) & 0 & \cdots & 0 & h_{L-1}(0) & h_{L-2}(0) & \cdots & h_1(0) \\ h_1(1) & h_0(1) & & \cdots & 0 & h_{L-1}(1) & & h_2(1) \\ \vdots & \ddots & & \ddots & & & & \vdots \\ 0 & \cdots & 0 & h_{L-1}(N-1) & h_{L-2}(N-1) & \cdots & \cdots & h_0(N-1) \end{bmatrix} \begin{bmatrix} x(0) \\ x(1) \\ \vdots \\ x(N-1) \end{bmatrix} + \begin{bmatrix} w(0) \\ w(1) \\ \vdots \\ w(N-1) \end{bmatrix} \tag{17}$$

The element $h_l(n)$ in $\mathbf{h}$ denotes the channel response of the l-th path at the n-th sample time. If N is the number of subcarriers, then $\mathbf{x}$, $\mathbf{y}$, and $\mathbf{w}$ can be expressed as an N-by-1 vector, and $\mathbf{h}$ is an N-by-N matrix.

The Fourier Transform of the received signal, $\mathbf{y}$, can be multiplied by $\mathbf{F}$ on both sides of (16). Hence, the received signal vector $\mathbf{Y}$ in the frequency domain will be expressed as:

$$\mathbf{Y}=\mathbf{HX}+\mathbf{W}, \tag{18}$$

where $\mathbf{X}$, $\mathbf{Y}$, and $\mathbf{W}$ denote the Fourier series of $\mathbf{x}$, $\mathbf{y}$, and $\mathbf{w}$, respectively. $\mathbf{H}$ is defined as the frequency domain channel matrix, and it can be expressed in terms of matrix $\mathbf{h}$. The frequency domain channel matrix $\mathbf{H}$ can be obtained as $\mathbf{FhF}^H$, where $\mathbf{F}$ denotes the N-by-N Fourier Transform matrix, which can be seen in (19), and $\mathbf{F}^H$ denotes applying the Hermitian operation on $\mathbf{F}$.

$$\mathbf{F}=\begin{bmatrix}1 & 1 & 1 & \cdots & 1\\ 1 & e^{-j\frac{2\pi}{N}} & e^{-j\frac{2\pi}{N}*2} & \cdots & e^{-j\frac{2\pi}{N}*(N-1)}\\ 1 & e^{-j\frac{2\pi}{N}*2} & e^{-j\frac{2\pi}{N}*4} & \cdots & e^{-j\frac{2\pi}{N}*2(N-1)}\\ \vdots & \vdots & \vdots & \ddots & \vdots\\ 1 & e^{-j\frac{2\pi}{N}*(N-1)} & e^{-j\frac{2\pi}{N}*2(N-1)} & \cdots & e^{-j\frac{2\pi}{N}*(N-1)(N-1)}\end{bmatrix}. \tag{19}$$

In order to detect the signals in (18), the zero forcing (ZF) detection scheme can be used by inversing the channel matrix $\mathbf{H}$. If the matrix $\mathbf{H}$ is not a square matrix, the inverse of the matrix will be replaced with the pseudo-inverse operation. Hence, the detected signals can be obtained as $\hat{\mathbf{X}}=\mathbf{H}^{+}\mathbf{Y}$, where the matrix $\mathbf{H}^{+}=(\mathbf{H}^H\mathbf{H})^{-1}\mathbf{H}^H$ is the pseudo-inverse of $\mathbf{H}$. Noise enhancement will occur when the ZF method is used, because the detected signals will be obtained as $\hat{\mathbf{X}}=\mathbf{H}^{+}\mathbf{Y}$.

$$\hat{\mathbf{X}}=\mathbf{H}^{+}\mathbf{Y}=\mathbf{H}^{+}(\mathbf{HX}+\mathbf{W})=\mathbf{IX}+\mathbf{H}^{+}\mathbf{W} \tag{20}$$

In (20), the first term becomes an identical matrix multiplied with the signal $\mathbf{X}$, and the second term cannot become a zero vector. Then, the second term $\mathbf{H}^{+}\mathbf{W}$ may enhance the noise term if some components in $\mathbf{H}^{+}$ become large due to the operation of $(\mathbf{H}^H\mathbf{H})^{-1}$. In the noise free environment, ZF detection will be widely used. However, noise enhancement occurs when the ZF detection is used.

4.2. Wiener-based Detection Scheme

In the adaptive theory [8], the Wiener filter is useful for communication systems. The Wiener filter theory is formulated for the general case of a complex valued stochastic process with the filter specified in terms of its impulse response.

In frequency domain, the received signals are $\mathbf{Y}=\mathbf{HX}$ where $\mathbf{X}$ is the transmitted signal. In order to estimate signals at the receiver, the estimated signal $\hat{\mathbf{X}}$ can be obtained by the Wiener solution. The Wiener solution $\mathbf{K}$ can be obtained by the following algorithm.

In order to find the Wiener solution, we must minimize the cost function. We define the cost function as the mean square error between the estimated signal $\hat{\mathbf{X}}$ and the transmitted signal $\mathbf{X}$. Then, the cost function can be expressed as

$$C = E\left(\left(\mathbf{X} - \hat{\mathbf{X}}\right)^2\right) \tag{21}$$

The Wiener solution is shown as follow:

$$\mathbf{K} = \arg\min_{\mathbf{Q}} C = \arg\min_{\mathbf{Q}} E\left(\left\|\mathbf{X} - \mathbf{QY}\right\|^2\right), \tag{22}$$

The Wiener solution can be achieved by the "orthogonal principle," and the geometric interpretation is presented in Fig. 6.

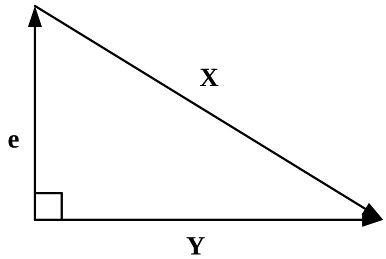

Fig. 6. Geometric interpretation of the relationship between the desired signal $\mathbf{X}$, the output of FFT $\mathbf{Y}$, and the mean square error $\mathbf{e}$

To achieve the orthogonal principle, the inner product between $\mathbf{Y}$ and $\mathbf{e}$ is held to zero

$$E\left(\left(\mathbf{X} - \hat{\mathbf{X}}\right)^H \mathbf{Y}\right) = 0 \Rightarrow E\left(\mathbf{X}^H \mathbf{Y}\right) = E\left(\mathbf{Y}^H \mathbf{Y}\right)\mathbf{K} \tag{23}$$

The Wiener solution can be determined by (23). Then, this solution is expressed as follows:

$$\mathbf{K} = \left(E\left(\mathbf{Y}^H \mathbf{Y}\right)\right)^{-1} E\left(\mathbf{X}^H \mathbf{Y}\right) = \left(\mathbf{H}^H \mathbf{H} + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{H}^H \tag{24}$$

Therefore, the detected signal $\hat{\mathbf{X}}$ can be obtained by the Wiener solution as $\hat{\mathbf{X}} = \mathbf{KY}$.

4.3. Wiener-based ICI Cancellation Scheme

Since the Wiener solution can detect a reliable signal, the results will be applicable to signal detection in the successive interference cancellation scheme.

In the successive interference cancellation scheme, in order to utilize ICI as a source of diversity, both reliable signal detection and an efficient ICI cancellation are needed. First, in order to achieve reliable signal detection, we utilize the method of ordering received signals based on signal-to-interference and noise ratio (SINR).

In order to fully utilize the time diversity while suppressing the residual interference and the noise enhancement, the signal detection is successive, but not detecting all the signals simultaneously.

The detection orders with subcarriers in the SIC scheme are decided by the SINR. The SINR can be obtained in the following manner. The vector of the received signal $\mathbf{Y}$ can be expressed as $\mathbf{Y}$=$\mathbf{HX}$+$\mathbf{W}$. The received signal can also be represented as:

$$\mathbf{Y}=\begin{bmatrix}\vec{h}_0 & \vec{h}_1 & \cdots & \vec{h}_{N-2} & \vec{h}_{N-1}\end{bmatrix}\begin{bmatrix}X(0)\\X(1)\\\vdots\\X(N-2)\\X(N-1)\end{bmatrix}+\mathbf{W}, \tag{25}$$

where $\vec{h}_k$ is the k-th column vector of the channel matrix $\mathbf{H}$, and the $X(k)$ is the k-th subcarrier signal at the input of IFFT.

Then, the vector of the received signal can be rewritten as the following form:

$$\mathbf{Y}=\vec{h}_0X(0)+\vec{h}_1X(1)+\cdots+\vec{h}_{N-1}X(N-1)+\mathbf{W}, \tag{26}$$

Then, we can use the Wiener solution to detect the signal $\hat{\mathbf{X}}$. Therefore, the k-th signal of the k-th subcarrier can be detected by $\hat{X}(k)=\vec{k}_k\mathbf{Y}$, where $\vec{k}_k$ is the k-th row vector of the Wiener solution $\mathbf{K}$, and the k-th signal of the k-th subcarrier can be obtained as:

$$\vec{k}_k\mathbf{Y}=\vec{k}_k(\vec{h}_0X(0)+\vec{h}_1X(1)+\cdots+\vec{h}_kX(k)+\cdots+\vec{h}_{N-1}X(N-1)+\vec{h}_NX(N))+\vec{k}_k\mathbf{W} \tag{27}$$

In (27) the desired signal is denoted as $\vec{k}_k\vec{h}_kX(k)$, and the others are the ICI and noise component. Hence, for the particular subcarrier k, the $SINR_k$ is defined as:

$$SINR_k=\frac{E\left[\left\|\vec{k}_k\vec{h}_kX(k)\right\|^2\right]}{E\left[\sum_{\substack{q=0\\q\neq k}}^{N-1}\left\|\vec{k}_k\vec{h}_qX(q)\right\|^2\right]+E\left[\left\|\vec{k}_k\mathbf{W}\right\|^2\right]}. \tag{28}$$

Each subcarrier's SINR can be obtained by (28). Hence, the subcarrier with the highest SINR is decided, following which we first detect the signal $\hat{X}(k)$. Equivalently, we choose the k-th row vector of $\mathbf{K}$ to detect the signal of the k-th subcarrier. After making a hard decision, the detected signal of the k-th subcarrier $\hat{X}(k)$ is reconstructed as the ICI component of the k-th subcarrier. Then, ICI effect for the k-th subcarrier will be cancelled in the received signal.

Following this, after canceling the ICI for the k-th subcarrier, the received signal $\mathbf{Y}$ will be obtained to

$$\mathbf{Y}^{(j+1)} = \mathbf{Y}^{(j)} - \vec{h}_k \hat{X}(k), \tag{29}$$

where $\hat{X}(k)$ is the hard decision signal of the k-th subcarrier. $\vec{h}_k \hat{X}(k)$ is the ICI term corresponding to k-th subcarrier. As long as this hard decision data is correct, the new vector $\mathrm{Y}^{(j+1)}$ has less interference. After this operation, the ICI

term of the detected signal will be removed and the channel matrix $\mathbf{H}$ should be reconstructed by removing the k-th column vector and k-th row vector. The column number and row number of the new channel matrix is then reduced. Therefore, the proposed Wiener-based SIC scheme repeats these steps until all the subcarriers are detected completely. According to this process, we will be able to detect all signals completely. The simulation results will be shown in section 5.

4.4. Modified Wiener-Based SIC ICI Cancellation Scheme

From above, we clearly know that the computation complexity of the Wiener-based successive interference cancellation scheme is very high. The size of the channel matrix $\mathbf{H}$ will be increased with the number of subcarriers. Hence, the computation complexity will increase. Therefore, in this section, we will propose an algorithm to reduce the computation complexity for the Wiener-based SIC scheme.

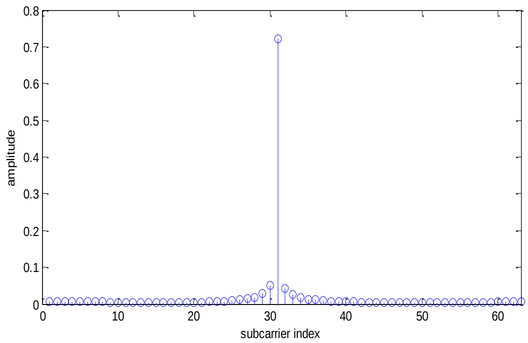

Fig. 7. The ICI amplitude for the desired signal at subcarrier 30

The traditional Wiener filter or a ZF equalizer is too complicated to be implemented, since it involves an N-by-N matrix inverse and matrix multiplication. N is usually fairly large, for example, N = 64 for the IEEE 802.11a standard, and N = 1024 for the IEEE 802.16 standard. As a matter of the fact, the ICI power arises from the neighboring subcarriers around the subject subcarrier. If we only focus on the neighboring subcarriers around the subject subcarrier, the computation complexity can be reduced significantly. Consequently, a simple scheme is investigated in Fig. 7 and Fig. 8, which show the ICI amplitude and ICI power for the desired signal.

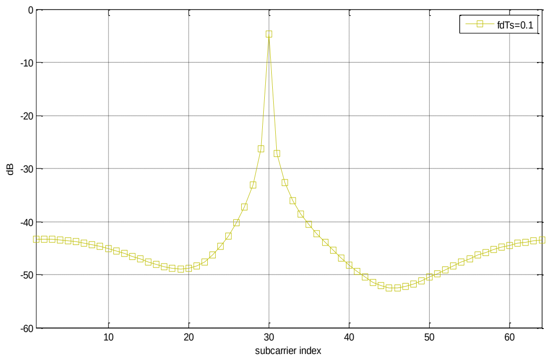

Fig. 8. The ICI power for the desired signal at subcarrier 30

Fig. 7 and Fig. 8 show the effective subcarriers that contribute the ICI to specific subcarrier are actually much smaller than the number of subcarriers in one OFDM symbol. Using this fundamental observation, we are going to focus on few subcarriers around the desired subcarrier. Now, if we only focus on q subcarriers around the desired subcarrier, we can rewrite the frequency-domain channel matrix $\mathbf{H}$ as $\bar{\mathbf{H}}$, and $\bar{\mathbf{H}}$ can be expressed as follows:

$$\bar{\mathbf{H}} = \begin{bmatrix} H_{0,0} & H_{0,1} & \cdots & H_{0,q-1} & 0 & \cdots & H_{0,N-q} & H_{0,N-q+1} & \cdots & H_{0,N-1} \\ H_{1,0} & H_{1,1} & & & & \ddots & & & & H_{1,N-1} \\ \vdots & & H_{2,2} & & & & & & & \vdots \\ H_{q-1,0} & & & \ddots & & & & & & H_{q-1,N-1} \\ 0 & & & & \ddots & & & & & 0 \\ \vdots & & & & & \ddots & & & & \vdots \\ H_{N-q,0} & \ddots & & & & & \ddots & & & H_{N-q,N-1} \\ H_{N-q+1,0} & & \ddots & & & & & & & H_{N-q+1,N-1} \\ \vdots & & & \ddots & & & & & & \vdots \\ H_{N-1,0} & H_{N-1,1} & \cdots & H_{N-1,q-1} & 0 & \cdots & H_{N-1,N-q} & H_{N-1,N-q+1} & \cdots & H_{N-1,N-1} \end{bmatrix} \quad (30)$$

The channel matrix $\bar{\mathbf{H}}$ is shown as (30). Hence, each signal on the subcarrier in the output of FFT can be rewritten as:

$$Y(m) = \sum_{k=m-q}^{m+q} H_{m,k} X(k), \quad (31)$$

where

$$H_{m,k} = \frac{1}{N} \sum_{l=0}^{L-1} e^{-j\frac{2\pi}{N}kl} \sum_{n=0}^{N-1} h_l(n) e^{-j\frac{2\pi}{N}(m-k)n}$$

Hence, for detecting the k-th subcarrier, we only used the partial element of $\mathbf{Y}$ to detect the signal. For example, if the value q is set as 2, two neighboring subcarriers (two on each side) are employed in the simplified equalizer; if the desired subcarrier is $X(3)$, then received signal signals $Y(1)$, $Y(2)$, $Y(4)$, and $Y(5)$ are all used. Therefore, if we want to detect the k-th subcarrier, the vector of the received signal and the vector of the channel matrix in the frequency-domain can be reduced as:

$$\mathbf{Y}_k = \mathbf{H}_k \mathbf{X}_k + \mathbf{W}_k, \quad k = 0,1,2,...,N-1 \tag{32}$$

$$\mathbf{Y}_k = \mathbf{Y}(k-q:k+q) \tag{33}$$

$$\mathbf{H}_k = \bar{\mathbf{H}}(k-q:k+q, k-2q:k+2q) \tag{34}$$

$$\mathbf{X}_k = \mathbf{X}(k-q:k+q) \tag{35}$$

where H_k indicates the partial matrix of $\bar{\mathbf{H}}$ that is the consecutive row vector from the $(k-q)$-th vector to the $(k+q)$-th vector and column vector from the $(k-2q)$-th vector to the $(k+2q)$-th vector. $\mathbf{Y}_k$ means the partial vector whose elements are consecutive from $Y(k\text{-}q)$ to $Y(k+q)$. Here, if k-q<0, the related (k-q)-th vector and (k-q)-th element is redefined as ((k-q) mod N)-th vector and ((k-q) mod N)-th element.

In the modified algorithm, the channel matrix H is reduced to H_k and the size is also reduced from N-by-N to $(2q+1)$-by-$(4q+1)$. Then, the Wiener solution will be obtained as:

$$\mathbf{G}_k = \left(\mathbf{H}_k{}^H \mathbf{H}_k + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{H}_k{}^H = [\vec{g}_0\ \vec{g}_1 \cdots \vec{g}_{2q}]^T, \tag{36}$$

where $\vec{g}_n$ is the n-th row vector of $\mathbf{G}_k$.

The detection scheme presented above is modified in the Wiener-based SIC scheme. If the size of the matrix is reduced, the computation complexity of matrix multiplication and matrix inverse is also reduced. In the modified scheme, first, we find the SINR for each subcarrier and sort each. The detection order is from the maximum SINR to the minimum SINR. Equivalently, we apply the modified detection scheme in the Wiener-based scheme. The algorithm of the modified scheme for reducing the computation complexity is shown in the following steps.

Step 1. Find the Wiener solution and compute the SINR for each subcarrier.

Step 2. Find the maximum SINR_k with the k-th subcarrier of undetected subcarriers, and find the Wiener solution.

$\mathbf{G}_k = \left(\mathbf{H}_k{}^H \mathbf{H}_k + \sigma^2 \mathbf{I}\right)^{-1} \mathbf{H}_k{}^H.$

Step 3. Detect the k-th subcarrier signal that has the maximum SINR.

$\hat{X}(k) = decision(\vec{g}_q \mathbf{Y}_k)$ where $\vec{g}_q$ is the $q-$th row of $\mathbf{G}_k$

Step 4. Cancel the intercarrier interference for $\hat{X}(k)$.

$\mathbf{Y}^{(j+1)} = \mathbf{Y}^{(j)} - \vec{h}_k \hat{X}(k)$ where $\vec{h}_k$ is the $k-$th column of $\mathbf{H}$

Step 5. Let the k-th column vector of $\mathbf{H}$ equal zero. ($\vec{h}_k = 0$)

Step 6. If $\mathbf{H}$ becomes a zero matrix, stop the scheme; if not, return to Step 2.

According to the modified algorithm for reducing the computation complexity, we analyze the complexity for different methods, and compare the order of computation complexity in section 4.5. In section 5, we show the BER performance of the modified SIC scheme for a different value q.

4.5. Complexity Analysis

The evaluation of the computation complexity for matrix operations follows the rules in [13–14]. For an N-by-N matrix multiplication or inversion, its order of the computation complexity is equivalent to $O(N^{2.376})$. For an M-by-N matrix multiplication with an N-by-M matrix and an M-by-N matrix, it is equivalent to $O(N^{1.376+r})$ of computation complexity where $r = \log_M N$.

In the Wiener-based SIC scheme, it is necessary to undertake matrix multiplication and inverse operation for each iteration. The Wiener solution can be obtained as $\mathbf{K}=(\mathbf{H}^H\mathbf{H}+\sigma^2\mathbf{I})^{-1}\mathbf{H}^H$; hence, the computation complexity of the Wiener solution for each OFDM symbol is obtained as:

$$k^{(r+1.376)} + k^{2.376} + k^{(r+1.376)}, \tag{37}$$

where k is the size of matrix $\mathbf{H}$. The first term denotes the computation complexity of the matrix multiplication, $\mathbf{H}^H\mathbf{H}$, the second term denotes the computation complexity of the inverse operation, and third term denotes the computation complexity of the results that are caused by the inverse of $\mathbf{H}^H\mathbf{H}+\sigma^2\mathbf{I}$ multiplication with $\mathbf{H}^H$. According to the Wiener-based SIC scheme in section 4.3 , the complexity for an OFDM symbol can be computed as:

$$\sum_{k=2}^{N}[2(k^{(r+1.376)}) + k^{2.376}] = \sum_{k=2}^{N}[2(k^{2.376}) + k^{2.376}] = \sum_{k=2}^{N}3(k^{2.376}). \tag{38}$$

In the modified Wiener-based SIC scheme, the Wiener solution is rewritten as $\mathbf{G}_k=(\mathbf{H}_k^H\mathbf{H}_k+\sigma^2\mathbf{I})^{-1}\mathbf{H}_k^H$; hence, the computation complexity for an OFDM symbol is also computed as:

$$3N^{2.376} + N((2q+1)^{r+1.376} + (2q+1)^{2.376} + (2q+1)^{r+.376})$$
$$= 3N^{2.376} + N(2(2q+1)^{2.376} + (2q+1)^{r+1.376}), \quad (39)$$

where r = $\log_{(4q+1)}(2q+1)$. Table 1. shows the complexity for each method.

Table 1. Order of computation complexity for the presented methods

Method	Multiplication	Results (N=64)
ZF	$3N^{2.376}$	58,696
Wiener solution	$3N^{2.376}$	58,696
ZF+SIC	$\sum_{k=2}^{N} 3(k^{2.376})$	1,142,237
Wiener+SIC	$\sum_{k=2}^{N} 3(k^{2.376})$	1,142,237
Reduced complexity scheme	$3N^{2.376} + 2N(2q+1)^{r+1.376} + N(2q+1)^{2.376}$	175,075 (for q=8) 85,002 (for q=4)

According to Table 1., we can discover that the computation complexity of the modified Wiener-based SIC scheme is less than that of the original Wiener-based SIC scheme. If the number of subcarriers is very large, the gap of computation complexity between the original Wiener-based SIC scheme and the modified Wiener-based SIC scheme will increase. In the next section, we show the performance for the different value q.

5. Simulation Results

In this section, we demonstrate the BER performance of our proposed scheme. We investigate the performance of the proposed scheme over Rayleigh fading channels. The environment of simulation is shown in Table 2.

Table 2. The environment of the simulation

Modulation scheme	QPSK
Number of subcarriers	64
Channel	Rayleigh fading multipath channel
Normalized Doppler frequency $f_d T_s$	0.1, 0.05
Path number	6

In Fig. 9, we simulated the BER performance of the OFDM system with different schemes. The simulation results show that the BER performance would be error floor when ICI occurs. The performance with the Wiener solution scheme is better than that with the ZF scheme, because the latter does not consider the noise term. Therefore, using the ZF method involves noise enhancement. The use of the Wiener solution to detect the signal in the Wiener-based SIC scheme ensures that the noise enhancement will be avoided. Hence, the Wiener-based SIC scheme's performance is also better than that of ZF-SIC scheme.

Fig. 10 shows that the BER performance with different normalized Doppler frequencies. The system has better BER performance when the normalized Doppler frequency is large. Meanwhile, as the f_dT_s gets large, the Wiener-based SIC scheme achieves more diversity gain. This shows that the Wiener-based SIC scheme can utilize the ICI as a source of diversity.

According to the Fig. 11, the modified Wiener-based SIC scheme's performance is not better than that of the original Wiener-based SIC scheme. Comparing the modified Wiener-based SIC scheme with the original Wiener-based SIC scheme, the performance loss of the modified Wiener-based SIC scheme is 2 dB for q = 8 at BER = 10^{-3}. However, according to the Table 2, the computation complexity of the modified Wiener-based SIC scheme is much lower than that of the original Wiener-based SIC scheme. Therefore, a suitable value, q, is a trade-off problem between performance and computation complexity.

6. Conclusions

A refined SIC detection for OFDM systems under Rayleigh fading channels has been presented. The performance for a low SINR subcarrier can be significantly improved due to ICI reduction scheme. According to the simulation results, SIC detection is very suitable for high fading rate mobile communications, such as the high-speed rail communication systems. The algorithm and the BER performance for the Wiener-based SIC scheme have been presented. According to the simulation results, we could clearly realize the performance of the Wiener-based SIC scheme is better than the ZF-SIC scheme's. Because the detection scheme may have noise enhancement in ZF-SIC scheme, the performance would be degraded. Although the Wiener-based SIC scheme has better performance, it has high computation complexity. In order to reduce the computation complexity for the Wiener-based SIC scheme, the modified Wiener-based SIC scheme is proposed.

According to the analysis, the computation complexity of the modified Wiener-based SIC scheme with q = 8 is 15% of the original Wiener-based SIC scheme. According to complexity analysis and simulation results, the performance with a large q is better than the performance with a small q, but the computation complexity is higher. Hence, this is a trade-off problem between the system performance and the computation complexity.

The schemes studied in this paper require perfect channel state information. To obtain a precision channel state information becomes an important issue worthy of further studying. Besides, due to the population of MIMO-OFDM technology, applying the proposed scheme to MIMO-OFDM systems is also worthy to investigate.

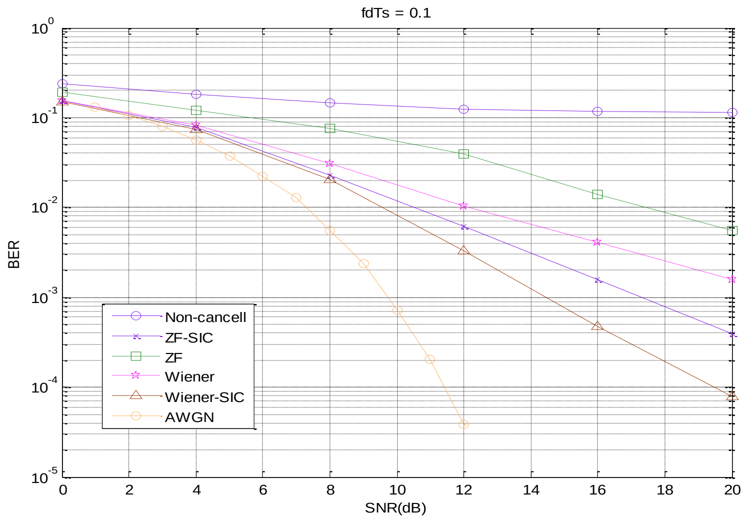

Fig. 9. BER versus SNR of four methods for f_dT_s = 0.1

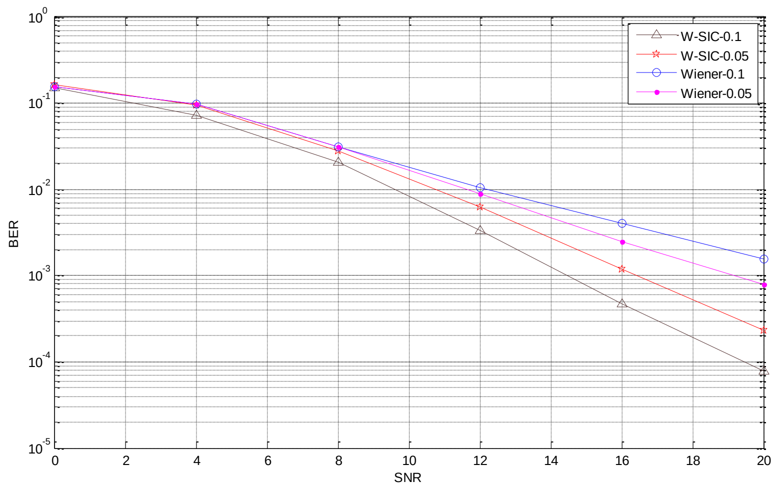

Fig. 10. BER versus SNR for different f_dT_s

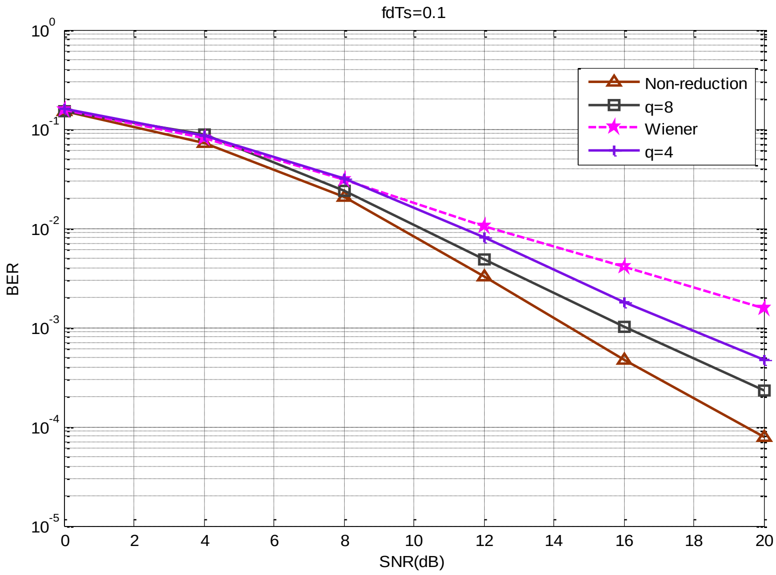

Fig. 11. BER versus SNR for different *q*

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