Decision Direct and Linear Prediction based Fast Fading Compensation for TFI-OFDM

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Abstract: Recently, time-frequency interferometry (TFI)-OFDM has been proposed as a channel identification scheme. TFI-OFDM system can multiplex the same impulse response in twice on the time domain without overlapping to each other. Moreover, by averaging of these impulse responses, the accurated channel impulse responses are obtained. However, under the fast fading channel, the change of channel conditions is too fast to be followed by channel estimator in the receiver. Therefore, too many errors occur in data demodulation processing particularly in the last part of data symbols. To overcome this problem, in this paper, we propose the decision direct and linear prediction based fast fading compensation method for TFI-OFDM.

1. Introduction

High data rate and high quality multimedia services are demanded in a fourth generation mobile communication, since application services are increasing. To meet this demand, orthogonal frequency division multiplexing (OFDM) is attractive and widely studied in recent years [1].Since the signals are transmitted in parallel by using many subcarriers that are mutually orthogonal and the corresponding spectrum is shaped like rectangle, OFDM can achieve high frequency efficiency and high data rate. Moreover, OFDM has been chosen for several broadband WLAN standards like IEEE802.11a, IEEE802.11g and European HIPERLAN/2, and terrestrial digital audio broadcasting (DAB) and digital video broadcasting (DVB) was also proposed for broadband wireless multiple access systems such as IEEE802.16 wireless MAN standard and interactive DVB-T [2]-[4].

In wireless communications, there is the problem that a communication quality deteriorates due to the fading that changes a phase and an amplitude by the influence such as reflection in a channel. In general, the impact of the fading is identified by using the pilot signals [5]. In this case, large pilot symbols are required to identify the accurate channel state information (CSI). Recently, the system called TFI-OFDM has been proposed to reduce this problem [6]. The TFI-OFDM can reduce the pilot symbols with maintaining the channel estimation property. It means that TFI-OFDM can reduce the total transmission power of pilot symbols. However, TFI-OFDM does not consider the channel compensation for fast fading scenario. To overcome this problem, the periodic pilot-symbol-aided (PSA) channel estimation has been proposed [7]. However, this method requires a great deal of the computational effort.

To reduce the above-mentioned problems, in this paper, we propose the decision direct and linear prediction based fast fading compensation for TFI-OFDM. In the proposed system, we make the replica signals by using the demodulated signals for estimating the channel variance between pilot and data symbols. The difference between the replica symbols and received symbols means the channel variance. By using this variance, we can easily compensate the channel variance due to the fast fading. For simplicity, the proposed scheme generates the replica symbols and calculates the channel variance within the adaptation interval. The main purpose of this research is to clarify the optimal adaptation interval for channel variance compensation. Moreover, we consider a linear prediction method to increase the adaptation interval for data packet. This paper is organized as follows. The proposed system is described in section 2. In section 3, we show the computer simulation results. Finally, the conclusion is given in section 4.

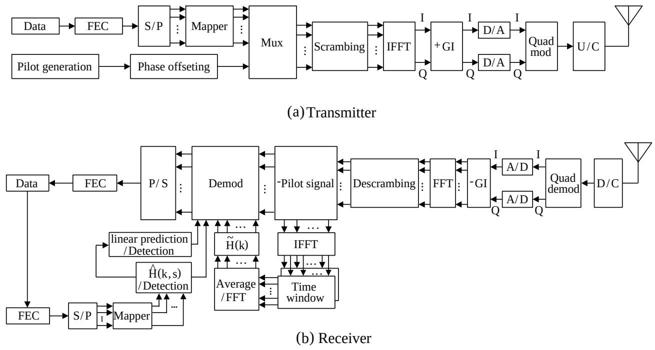

Figure 1. The proposed system.

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 the complex channel gain and the time delay of the lth propagation path, respectively, and $\sum_{L-1}^{l=0} E|h_l^2| = 1$, 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 Transmitter and Receiver Structure

The transmitter block diagram of the proposed system is shown in Fig. 1(a). Firstly, the coded binary information data sequence is modulated, and N_p pilot symbols are appended

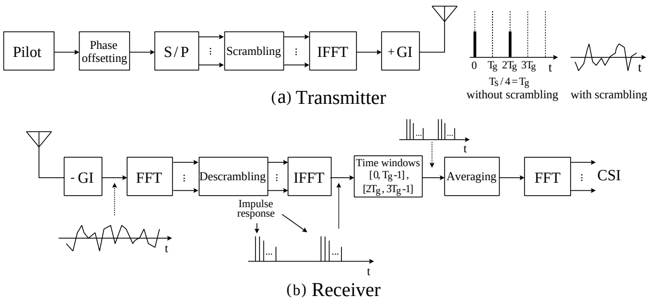

Figure 2. The concept of TFI-OFDM system.

at the beginning of the sequence. The transmit signal of the proposed system can be expressed in its equivalent baseband representation as

$$s(t) = \sum_{i=0}^{N_p+N_d-1} g(t-iT) \cdot\left\{\sqrt{\frac{2S}{N_c}}\sum_{k=0}^{N_c-1} u(k,i)\cdot\exp[j2\pi(t-iT)k/T_s]\right\}, \quad (3)$$

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

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

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 inter-symbol interference (ISI) due to the multi-path fading, and hence, we have

$$T=T_s+T_g. \quad (5)$$

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 the 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 fast fourier transform (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

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

where $\hat{n}(k,i)$ is AWGN noise with zero-mean and a variance of $2N_0/T_s$. After abbreviating [1], Eq. (9) can be rewritten as

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

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

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

where $\frac{c^*_{PN}(k)}{|c_{PN}(k)|^2}$ is the descrambling operation. Since $T_g = T_s/4$, the proposed system 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

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

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 $\tilde{H}(k)$ is obtained by

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

where $\eta(k)$ is AWGN component. In low Doppler frequency, the CSI is almost static. The received signal can be compensated by using Eq. (13) as

$$R(k,i)=r(k,i)/\tilde{H}(k), \quad (14)$$

where $R(k,i)$ is the compensated signal.

2.3 Fast Fading Compensation

Under the fast fading channel, the change of channel conditions is too fast to be followed by channel estimator in the

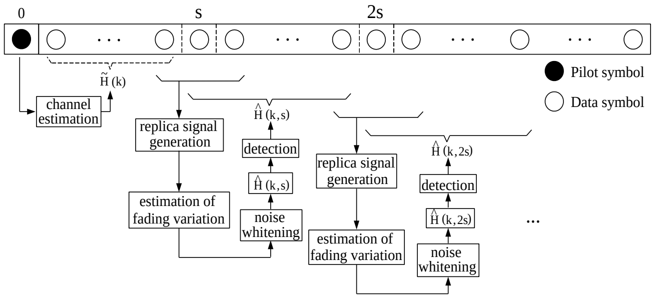

Figure 3. The proposed fading compensation method.

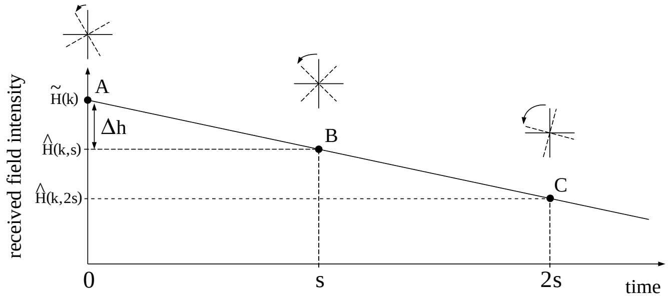

Figure 4. The linear approximated fading characteristics.

receiver. Therefore, too many errors occur in data demodulation processing particularly in the last part of data symbols. To overcome this problem, we proposed the decision direct and linear prediction based fast fading compensation method. Under the fast fading, the phase and amplitude are changed due to the Doppler shift. In this case, the change of channel condition is too fast to be followed by channel estimator. Therefore, the channel estimation and fading compensation is required for data symbols behind sth symbol that the error is caused by an increase of Δh.

Firstly, we decide the sth symbol that has the big difference of the channel frequency response between the pilot and data symbol, where s is $N_p \leq s \leq N_p+N_d-1$. As shown in Fig. 1, the replica signals $rep(k, s{-}1), rep(k, s), rep(k, s{+}1)$ of (s–1, s, $s+1$)th symbols are generated by $R(k, s)$, where $R(k, s)$ is the demodulated signal for the kth subcarrier and the sth symbol. By using $\tilde{H}(k)$ as Eq. (13), the fading variations $\Delta h(k, s-1)$, $\Delta h(k, s)$, $\Delta h(k, s+1)$ of $(s-1,\ s,\ s+1)$th symbols can be calculated by

$$\begin{aligned}\Delta h(k,s{-}1) &= r(k,s{-}1)/\{rep(k,s{-}1)\cdot\tilde{H}(k)\}\\ \Delta h(k,s) &= r(k,s)/\{rep(k,s)\cdot\tilde{H}(k)\}\\ \Delta h(k,s{+}1) &= r(k,s{+}1)/\{rep(k,s{+}1)\cdot\tilde{H}(k)\},\end{aligned} \quad (15)$$

where $r(k, s)$ is the output signal for the kth subcarrier and the sth symbol in Eq.(11). The adjusted CSI $H(k, s-1)$, $H(k, s)$, $H(k, s{+}1)$ of (s–1, s, s+1)th symbols are generated by

$$\begin{aligned}H(k,s{-}1) &= \Delta h(k,s{-}1)\cdot\tilde{H}(k)\\ H(k,s) &= \Delta h(k,s)\cdot\tilde{H}(k)\\ H(k,s{+}1) &= \Delta h(k,s{+}1)\cdot\tilde{H}(k).\end{aligned} \quad (16)$$

Observing Eq. (16), each CSI includes the noise. For improving the accuracy of the CSI, we used the averaging for each component of Eq. (16) as

$$\hat{H}(k,s)=\{H(k,s{-}1)+H(k,s)+H(k,s{+}1)\}/3. \quad (17)$$

To prevent the performance deterioration due to rapid channel variance, we consider the following condition,

$$\hat{H}(k,s)=\begin{cases}\hat{H}(k,s) & \text{for } \frac{|\hat{H}(k,s)|^2}{|\tilde{H}(k)|^2}\geq|\delta|^2\\ \tilde{H}(k) & \text{otherwise},\end{cases} \quad (18)$$

where δ is a threshold, and $\frac{|\hat{H}(k,s)|^2}{|\tilde{H}(k)|^2}\geq|\delta|^2$ means an acceptable channel variance. As shown in Fig. 3, data symbols behind sth symbol are compensated by the adjusted $\hat{H}(k, s)$. By iteration with the above processing, we can increase the packet length to maintain the system performance.

When the packet is long, the proposed decision direct must repeat the estimation of compensation value at s symbol intervals as shown in Fig. 3. The computational effort necessary to estimate of the CSI can be reduced by the following linear prediction. Although the fading characteristic nonlinearly varies in fast fading channel, this variance can be linearly approximated in packets as shown in Fig. 4. Therefore, if the CSIs in two points have been already known, the CSIs of the symbols behind these points can be linearly predicted. For example, when we know the pilot channel frequency response $\tilde{H}(k)$ and the adjusted CSI $\hat{H}(k, s)$ for the sth symbol, $\hat{H}(k, 2s)$ can be calculated as

$$\hat{H}(k,2s)=2\cdot\hat{H}(k,s)-\tilde{H}(k). \quad (19)$$

For all symbols behind these symbols, by using Eq. (19), the next CSI can be calculated with the linear prediction method. In this case, we can reduce the computational effort.

3. Computer Simulated Results

In this section, the performance of the proposed fading compensation method is compared with the conventional method. Fig. 1 shows a simulation model of the proposed system. In 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 receive antenna in twice on the time domain without overlapping to each other as shown in Fig. 2(a). The data stream is encoded. Here, convolutional codes (rate $R = 1/2$, constraint length $\mathcal{K}=7$) are used. The coded bits are QPSK modulated, and then the pilot signal and the data signal are multiplexed with scrambling using PN code to reduce the PAPR. The OFDM time signals are generated by an inverse fast fourier tansform (IFFT) and transmitted to the frequency selective and time variant radio channel after cyclic extensions have been inserted. In this simulation, we assume that OFDM symbol period is 4 μs, guard interval is 1 μs, and $L=15$ path Rayleigh fadings have exponential shapes and a path separation $T_{path} = 62.5$ $n$sec. The maximum Doppler frequencies (f_d) are assumed to be 10 Hz and 600 Hz. In the receiver, the guard interval is erased from the received signals and the received signals are serial-to-parallel (S/P) converted. The parallel sequences are passed to an FFT operator, which converts the signal back to the frequency domain. After descrambling and an IFFT, each impulse responses can be estimated by extracting and averaging twice impulse responses using the time windows with Eq. (13) as shown in Fig. 2(b). The frequency domain data

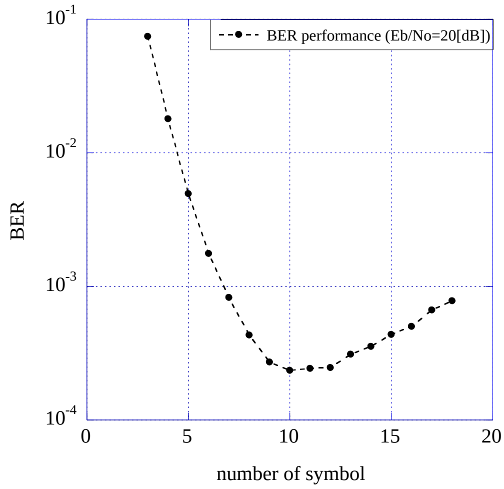

Figure 5. BER vs. adaptation interval s.

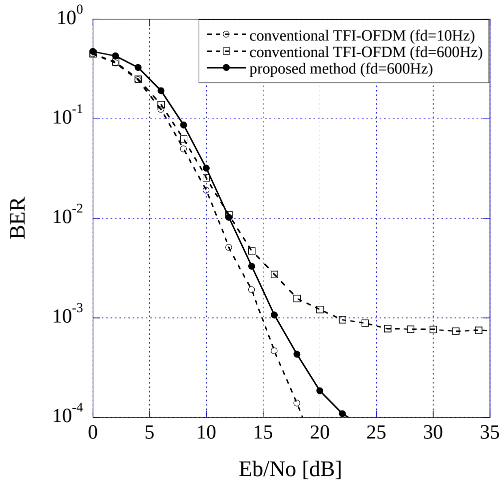

Figure 6. BER vs. Eb/No (s=10, δ=0.1).

signal is detected and demodulated using the estimated channel impulse response. After detection, bits are decoded by the Viterbi soft decoding algorithm. From the decoded bits, the replica signals are generated. Using the replica signals, the channel time variance is estimated by decision direct, and then the received signals are compensated by using the adjusted CSI. This processing is repeated with the same adaptation interval until the end of packet. Finally, the data signals are demodulated and decoded by the Viterbi soft decoding algorithm. Table 1 shows the simulation parameters.

Fig. 5 shows the BER of the proposed fading compensation method versus adaptation interval s, where s is the value of symbol intervals for fading compensation using the decision direct and linear prediction at $Eb/No = 20$ dB. From the simulation result, the proposed method shows the best BER with $s = 10$. In this evaluation, we used the adaptation interval as $s = 10$.

Fig. 6 shows the BER of the conventional and proposed compensation methods at Doppler frequencies (f_d) of 10 Hz and 600 Hz. By increasing the Doppler frequency, the conventional method shows the error floor. On the other hand, the proposed method can compensate the channel time variance by using the decision direct and linear prediction. Therefore, the proposed method can mitigate the error floor. As a result, the proposed method with $f_d = 600$ Hz shows the approximately same BER performance like that of the conventional method with f_d=10 Hz.

Table 1. Simulation parameters.

Data Modulation	QPSK
Data detection	Coherent
Symbol duration	4.0 μs
Frame size	21 symbols ($N_p = 1, N_d = 20$)
FFT size	64
Number of carriers	64
Guard interval	16 sample times
Fading	15 path Rayleigh fading
Doppler frequency	10, 600Hz
FEC	Convolutional code (R=1/2, $\mathcal{K}$=7)

4. Conclusion

In this paper, we have proposed the decision direct and linear prediction based fast fading compensation method. The proposed scheme repeatedly estimates the CSI by using the decision direct method with the adaptation interval s. From the simulation results, the conventional method shows the error floor. On the other hand, the proposed method can compensate the channel time variance by using the decision direct and linear prediction. Therefore, the proposed method can mitigate the error floor. As a result, the proposed method with f_d=600 Hz shows the approximately same BER performance like that of the conventional method with f_d=10 Hz.

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