

Novel Time Domain Channel Estimation using a Cyclic Property of Scattered Pilot for OFDM in Large Delay Spread Channel

Taichi Sakaue[†], Chang-Jun Ahn, Tastuya Omori, Ken-ya Hasimoto
 Graduate school of Engineering, Chiba University
 1-33 Yayoi-cho, Inage-ku, Chiba, 263-8522 Japan
 Email:ts-12tm2322@chiba-u.jp[†]

Abstract—Orthogonal frequency division multiplexing (OFDM) based on scattered pilot (SP) can accurately estimate the channel state information (CSI) with small number of pilot signals under the fast fading channel. In fact, the channel response at a particular subcarrier frequency is not supposed to be totally different from its neighboring frequencies, and they must have correlation which depends on the coherence bandwidth of channel. Therefore, the frequency separation of SP should smaller than the coherence bandwidth. However, when the maximum delay spread exceed the length of GI, unacceptable degradation of error performance is caused by inter-symbol-interference (ISI) and inter-carrier-interference (ICI). Moreover, coherence bandwidth is also to be narrow. In this case, channel estimation of SP-OFDM based on the frequency interpolation is poorly operated due to narrow coherence bandwidth. To mitigate this problem, in this paper, we propose the novel time domain channel estimation scheme using a cyclic property of SP for OFDM.

I. INTRODUCTION

Orthogonal frequency division multiplexing (OFDM) is commonly used for broadband wireless transmission like IEEE802.11a and IEEE802.11g because of its ability to combat the multipath channel problem [1], [2]. In order to avoid inter-symbol interference (ISI) caused by multipath, a guard interval (GI) is appended to the OFDM symbol.

In an OFDM, inter-carrier interference (ICI) occurs on a fast fading channel because the orthogonality between subcarriers is collapsed due to a Doppler frequency. To mitigate this problem, OFDM system generally uses the scattered pilot (SP) to identify the channel state information (CSI) and to compensate the ICI [3], [4]. In an scattered pilot OFDM (SP-OFDM), the pilot arrangement is evenly separated at a distance of N_f subcarrier along the frequency domain. In fact, the channel response at a particular subcarrier frequency is not supposed to be totally different from its neighboring frequencies, and they must have correlation which depends on the coherence bandwidth of channel B_c . Therefore, the frequency separation of SP should smaller than the coherence bandwidth. In general, the GI is usually designed to be longer than the delay spread of the channel, and is decided after channel measurements in the desired implementation scenario. However, when the maximum delay spread exceed the length of GI, unacceptable degradation of error performance is caused by ISI and ICI. Moreover, coherence bandwidth is also to

be narrow. In this case, channel estimation of SP-OFDM based on the frequency interpolation is poorly operated due to narrow coherence bandwidth. Until this time, several schemes have been proposed to mitigate the ISI due to the large delay spread channel. Gejoh proposed ISI reduction method by increasing the GI to overcome the above problems [5]. However, transmission efficiency is deteriorated, and transmit power is increased. Lee proposed double window cancellation and combining as ISI canceler scheme [6]. However the complexity is serious problem. Suyama proposed several frequency domain ISI canceling schemes such as a smoothed FFT windows and decision feedback equalizer (DEF) [7]. However it is very sensitive to the severe frequency selective channel. To mitigate this problem, in this paper, we propose the novel time domain channel estimation scheme using a cyclic property of SP. In the proposed channel estimation scheme, we use the time domain processing for the time interpolated channel responses by using a FFT and select the paths to remove the noise term. Moreover, we compensate ISI and ICI by using the replica signals. This paper is organized as follows. The system model and conventional channel estimation scheme are described in Section II. In Section III, we show the proposed channel estimation scheme and the ISI and ICI compensation scheme. In, Section IV, we show the computer simulation results. Finally, the conclusion is given in Section V.

Fig. 1 shows the block diagram of the proposed SP-OFDM system. The binary data sequence is QPSK modulated, and scattered pilot signals are discretely inserted. The OFDM symbol with the guard interval (GI) can be express as

$$s(t) = \sum_{i=0}^{N_s-1} g(t-iT) \sqrt{\frac{2S}{N_c}} \sum_{k=0}^{N_c-1} u(k,i) \cdot \exp \{j2\pi(t-iT)k/T_s\}, \quad (1)$$

where S is the average transmit power, N_s is the number of symbols, N_c is the number of subcarriers, T_s is the effective symbol length, T is the OFDM symbol length, $g(t)$ is transmission pulse, respectively. The frequency separation between adjacent orthogonal subcarriers is $1/T_s$ and can be expressed, by using the k th subcarrier of i th modulated symbol $|d(k,i)| = 1$ for $0 \leq i \leq N_s - 1$, as

$$u(k,i) = c_{PN}(k) \cdot d(k,i), \quad (2)$$

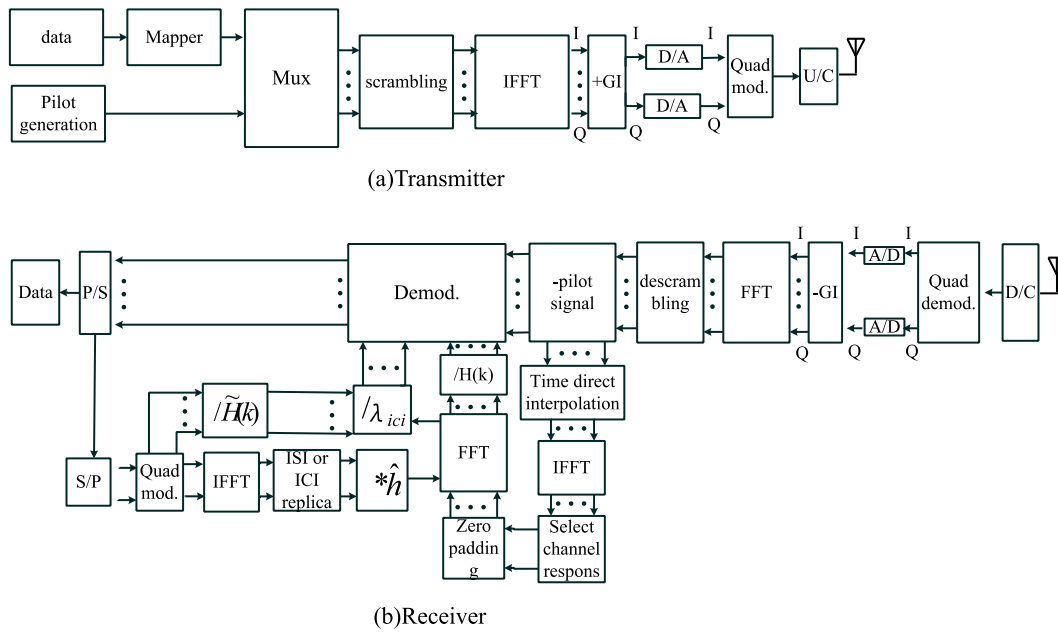


Fig. 1. Proposed system

where c_{PN} is pseudo-noise (PN) code as a scrambling code to reduce the peak to average power ratio (PAPR). The scattered pilot in the frequency domain as shown in Fig. 2 is employed. The scattered pilot is added on every 4 subcarriers ($N_f = 4$) in the OFDM symbol and every 8 OFDM symbols ($N_t = 8$) on the same subcarrier. The scattered pilot arrangement is given by

$$\sum_{\hat{i}=0}^{\lfloor 2N_s/N_t \rfloor - 1} \sum_{\hat{k}=0}^{N_c/N_f - 1} d(N_f \hat{k} + 2\text{mod}(\hat{i}, \frac{N_f}{2}), \frac{N_t}{2} \hat{i}) = 1, \quad (3)$$

where N_f and N_t are SP separations on the frequency and the time domains, “mod” represent modulo, $\lfloor x \rfloor$ denotes the largest integer less than or equal to x . We assume that propagation channel consists of L discrete paths with different time delay. The impulse response is represented as

$$h(\tau, t) = \sum_{l=0}^{L-1} h_l(t) \delta(\tau - \tau_l), \quad (4)$$

where h_l and τ_l are complex channel gain and the time delay of l th propagation path, and $E|\cdot|$ denotes the ensemble average operation. The channel transfer function $H(f, t)$ is the Fourier transform of $h(\tau, t)$ and it is given by

$$\begin{aligned} H(f, t) &= \int_0^\infty h(\tau, t) \exp(-j2\pi f\tau) d\tau \\ &= \sum_{l=0}^{L-1} h_l \exp(-j2\pi f\tau_l). \end{aligned} \quad (5)$$

The maximum path delay τ_L is assumed to exceed the length of GI T_g ($\tau_L > T_g$).

The received signal $r(t)$ in the equivalent baseband representation can be expressed as

$$r(t) = \int_{-\infty}^{\infty} h(\tau, t) s(t - \tau) d\tau + n(t), \quad (6)$$

where $n(t)$ is additive white Gaussian noise (AWGN) with a single sided power spectral density of N_0 . By applying the FFT operation, the received signal $r(t)$ is resolved into N_c subcarriers as

$$\tilde{r}(k, i) = \sqrt{\frac{2S}{N_c}} H(k, i) u(k, i) + \hat{n}(k, i). \quad (7)$$

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

$$\begin{aligned} r(k, i) &= \frac{c_{PN}^*(k)}{|c_{PN}(k)|^2} \tilde{r}(k, i) \\ &= \sqrt{\frac{2S}{N_c}} H(k, i) d(k, i) + \hat{n}(k, i), \end{aligned} \quad (8)$$

where $\frac{c_{PN}^*(k)}{|c_{PN}(k)|^2}$ is descrambling operation, $\hat{n}(k, i)$ is AWGN with variance of $2N_0/T_s$. By using the linear time domain

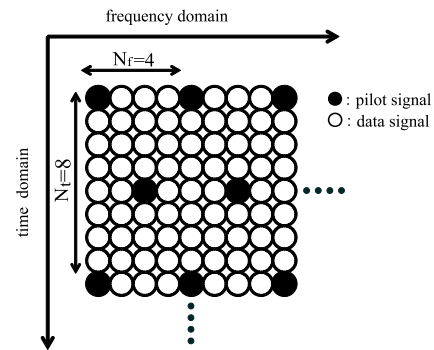


Fig. 2. The assignment of scattered pilot signals

interpolation of scattered pilot, the channel responses are estimated as

$$\hat{H}(k, i) = \begin{cases} \frac{(N_t - \text{mod}(i, N_t))r(k, \lfloor \frac{i}{N_t} \rfloor N_t + \text{mod}(i, N_t)r(k, (\lfloor \frac{i}{N_t} \rfloor + 1)N_t)}{N_t} \\ \text{for } i < \lfloor \frac{N_c - 1}{N_t} \rfloor N_t \text{ \& mod}(k, N_f) = 0 \text{ \& mod}(i, N_t) \neq 0 \\ r(k, i) \text{ for } (\text{mod}(k, N_f) = 0 \text{ \& mod}(i, N_t) = 0) \\ \text{or } (\text{mod}(k, N_f) = \frac{N_f}{2} \text{ \& mod}(i, N_t) = \frac{N_t}{2}) \\ \frac{(N_t - \text{mod}(i - \frac{N_t}{2}, N_t))r(k, \lfloor \frac{i - \frac{N_t}{2}}{N_t} \rfloor N_t + \frac{N_t}{2})}{N_t} \\ + \frac{\text{mod}(i - \frac{N_t}{2}, N_t)r(k, (\lfloor \frac{i - \frac{N_t}{2}}{N_t} \rfloor + 1)N_t + \frac{N_t}{2})}{N_t} \\ \text{for } i > \frac{N_t}{2} \text{ \& mod}(k, N_f) = \frac{N_f}{2} \text{ \& mod}(i - \frac{N_t}{2}, N_t) \neq 0 \\ \frac{(N_t + \text{mod}(i, N_t))r(k, \lfloor \frac{i}{N_t} \rfloor N_t) - \text{mod}(i, N_t)r(k, (\lfloor \frac{i}{N_t} \rfloor - 1)N_t)}{N_t} \\ \text{for } i > \lfloor \frac{N_c - 1}{N_t} \rfloor N_t \text{ \& mod}(k, N_f) = 0 \text{ \& mod}(i, N_t) \neq 0 \\ \frac{(N_t + \frac{N_t}{2} - \text{mod}(i, N_t))r(k, \lfloor \frac{i}{N_t} \rfloor N_t + \frac{N_t}{2})}{N_t} \\ - \frac{(\frac{N_t}{2} - \text{mod}(i, N_t))r(k, (\lfloor \frac{i}{N_t} \rfloor + 1)N_t + \frac{N_t}{2})}{N_t} \\ \text{for } i < \frac{N_t}{2} \text{ \& mod}(k, N_f) = \frac{N_f}{2} \text{ \& mod}(i, N_t) \neq \frac{N_t}{2} \\ 0 \text{ otherwise.} \end{cases} \quad (9)$$

After a time domain interpolation, the channel responses are estimated by using a frequency domain interpolation as,

$$\tilde{H}'(k, i) = \begin{cases} \frac{(\frac{N_f}{2} - \text{mod}(k, \frac{N_f}{2}))\hat{H}(\lfloor \frac{k}{N_f/2} \rfloor \frac{N_t}{2}, i)}{\frac{N_f}{2}} \\ + \frac{\text{mod}(k, \frac{N_f}{2})\hat{H}((\lfloor \frac{k}{N_f/2} \rfloor + 1)\frac{N_t}{2}, i)}{\frac{N_f}{2}} \\ \text{for } k < \lfloor \frac{N_c - 1}{N_f/2} \rfloor \frac{N_f}{2} \text{ \& mod}(k, \frac{N_f}{2}) \neq 0 \\ \hat{H}(k, i) \text{ for } \text{mod}(k, \frac{N_f}{2}) = 0 \\ \frac{\frac{N_f}{2} + \text{mod}(k, \frac{N_f}{2})\hat{H}(\lfloor \frac{k}{N_f/2} \rfloor \frac{N_t}{2}, i)}{\frac{N_f}{2}} \\ - \frac{\text{mod}(k, \frac{N_f}{2})\hat{H}((\lfloor \frac{k}{N_f/2} \rfloor - 1)\frac{N_t}{2}, i)}{\frac{N_f}{2}} \\ \text{for } k > \lfloor \frac{N_c - 1}{N_f/2} \rfloor \frac{N_f}{2} \text{ \& mod}(k, \frac{N_f}{2}) \neq 0. \end{cases} \quad (10)$$

The conventional channel estimation is operated by using frequency interpolation. In a frequency interpolation, we should consider the coherence bandwidth. Coherence bandwidth B_c that has more than 50% correlation in frequency is defined as

$$B_c \simeq 1/5\tau_{rms}, \quad (11)$$

where τ_{rms} is the rms delay spread. However, the interval between two adjacent pilot is shorter than coherent bandwidth that defines range of high correlation among adjacent frequencies of channel over the large delay channel. In this case the channel estimation based on frequency interpolation is poorly operated. To mitigate this problem, in this paper, we propose a time domain channel estimation using a cyclic property of scattered pilot.

II. PROPOSED CHANNEL ESTIMATION AND COMPENSATION SCHEMES

A. proposed channel estimation

In the proposed channel estimation, we do not consider the frequency interpolation as Eq. (10). We carry out in the time domain processing. By applying an IFFT operation to Eq.

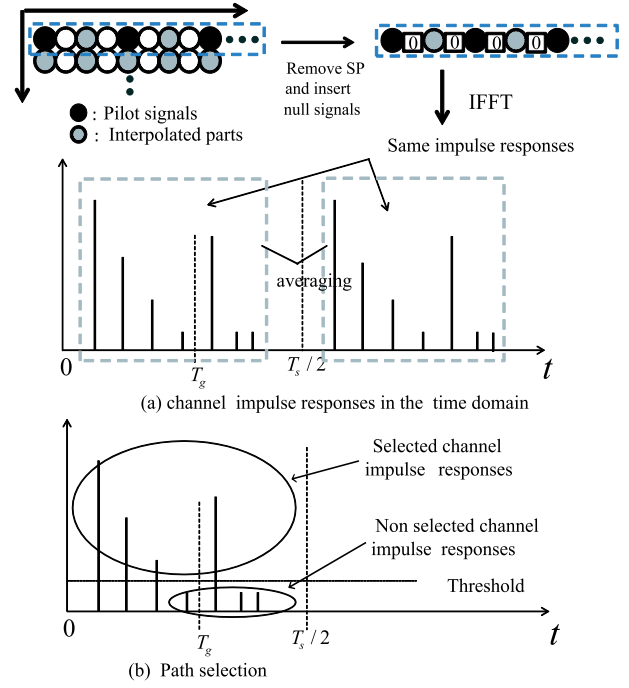


Fig. 3. The channel responses and path selection in the time domain

(9), we can obtain the two same impulse responses on the time domain as shown in Fig.3(a) [8]. In this case, Eq. (9) is converted to the time domain response $\hat{h}(t)$ by using an IFFT as

$$\begin{aligned} \hat{h}(t) &= \sum_{i=0}^{N_s-1} \sqrt{\frac{2P}{N_c}} \sum_{k=0}^{N_c-1} \tilde{H}'(k, i) \exp \{j2\pi(t - iT)k/T_s\} \\ &= \sum_{i=0}^{N_s-1} \sum_{l=0}^{L-1} \sqrt{\frac{2P}{N_c}} h_l(t + iT) \frac{1}{\sqrt{2}} [\delta(\tau - \tau_l) \\ &\quad + \delta(\tau - \tau_l - \frac{T_s}{2})] + \tilde{n}(t), \end{aligned} \quad (12)$$

where P is the power of frequency responses. From Eq.(12), we can see two impulses with time shift as $\delta(\tau - \frac{T_s}{2})$, and the output signals are equivalent to time domain multiplexed impulse responses. By averaging of these impulse responses, we can reduce the noise power. However, if the total channel paths are reduced, the selected time signals include the noise terms. By an FFT operation, these noise terms are spread in the frequency domain. Therefore, the channel identification is poorly operated. Here we consider a path selection method by using the threshold ξ_0 . The concept of path selection is shown in Fig. 3(b). We estimate the noise term as

$$\begin{aligned} \sigma_N &= \frac{1}{N_c/N_f - 1} \sum_{\hat{k}=0}^{N_c/N_f - 1} \{r(N_f \hat{k}, 0) \\ &\quad - \{r(N_f \hat{k}, 0) + r(N_f \hat{k}, N_t)\}/2\}. \end{aligned} \quad (13)$$

From Eq. (13), we choose the threshold ξ as

$$|\xi|^2 = |\sigma_N|^2. \quad (14)$$

Therefore, the selected paths are represented as

$$\hat{h}'(t+iT) = \begin{cases} \hat{h}(t+iT) & |\hat{h}_l(t+iT)|^2 > |\xi|^2 \\ 0 & \text{otherwise} \end{cases} \quad (15)$$

From Eq. (15), the channel response is represented as

$$\tilde{h}'(t) = \sum_{i=0}^{N_s-1} \sqrt{\frac{2P}{N_c}} \sum_{l=0}^{L'-1} \hat{h}'(t+iT) \cdot \frac{1}{\sqrt{2}} \{ \delta(\tau - \tau_l) + \delta(\tau - \tau_l - 2T_g) \} + n'(t), \quad (16)$$

where L' is the number of the selected paths. After a FFT operation, the channel response of the k th subcarrier is given by

$$\tilde{H}'(k) = \frac{1}{\sqrt{P/N_c}} \sum_{e=0}^{N_c-1} \frac{1}{T_s} \int_0^{T_s} \sum_{l=0}^{L'-1} h_l(t+iT) \cdot \int_{-\infty}^{\infty} \{ \delta(\tau - \tau_l) + \delta(\tau - \tau_l - 2T_g) \} \cdot \exp(-j2\pi e\tau/T_s) d\tau dt + \bar{\mu}'(k), \quad (17)$$

where $\bar{\mu}'(k)$ is AWGN after proposed channel estimation. Observing Eq. (17), we can reduce the total noise power due to the path selection and null assignment [8]. In this case, we achieve an accurate CSI compared to the conventional frequency interpolation method.

B. ISI and ICI compensations

When $\tau_L \leq T_g$, we represent (6) for matrix in frequency domain as

$$\mathbf{R}_i = \mathbf{H}_i \mathbf{F} \mathbf{D}_i + \mathbf{N}_i, \quad (18)$$

where, $\mathbf{H}_i$ is the $N_c \times N_c$ channel matrix for i th symbol, $\mathbf{N}_i$ is the $N_c \times 1$ noise matrix, $\mathbf{F}$ is the IFFT operation and, τ_l is maximum delay time, respectively. For $\tau_L > T_g$, (18) can be written as

$$\mathbf{R}_i = \mathbf{H}_{i,ici} \mathbf{F} \mathbf{D}_i + \mathbf{H}_{i-1,isi} \mathbf{F} \mathbf{D}_{i-1} + \mathbf{N}_i \quad (19)$$

where, $\mathbf{H}_{i-1,isi}$ is the ISI channel matrix for $(i-1)$ th symbol and the ICI channel matrix $\mathbf{H}_{i,ici}$ for i th symbol. In general, the first symbol of received signals has no ISI. In this case, (19) can be written as

$$\mathbf{R}_0 = \mathbf{H}_{0,ici} \mathbf{F} \mathbf{D}_0 + \mathbf{N}_0 \quad (20)$$

From (20), the equalization coefficient can be easily calculated by the Hermitian of $\mathbf{H}_{i,ici} \mathbf{F}$. However, since the symbols behind the first symbol include the ISI, ISI equalization should be performed with the previously detected symbol $\hat{\mathbf{D}}_{i-1}$ and ISI channel matrix $\mathbf{H}_{i-1,isi}$. In this case, the ISI equalized signal $\tilde{\mathbf{R}}_i$ is given by

$$\tilde{\mathbf{R}}_i = \mathbf{R}_i - \tilde{\mathbf{H}}_{i-1,isi} \mathbf{F} \mathbf{D}'_{i-1} = \lambda_{i,ici} \mathbf{F} \mathbf{D}_i + \tilde{\mathbf{N}}_i, \quad (21)$$

where $\tilde{\mathbf{N}}_i$ is the noise term with the residual ISI, $\lambda_{i,ici}$ is the channel matrix with residual ICI. $\tilde{\mathbf{H}}_{i-1,isi}$ is constructed by the estimated channel response. $\tilde{\mathbf{H}}_{i-1,isi} \mathbf{F} \mathbf{D}'_{i-1}$ represent the replica signals and are deducted from received signals to

cancel ISI. After ISI compensation, we perform FFT operation and equalize the received signals by Zero-forcing(ZF) in frequency domain as

$$\hat{\mathbf{D}}_i = \frac{\mathbf{F}^{-1} \tilde{\mathbf{R}}_i}{\hat{\mathbf{H}}_i} = \mathbf{D}_i + \frac{\tilde{\mathbf{N}}_i}{\hat{\mathbf{H}}_i} \quad (22)$$

From Eq. (22), we can observe that the first term is the desired signal, the second term is a noise term. Next, we modify CSI by using received signal $\tilde{\mathbf{R}}_i$ and detected signal $\hat{\mathbf{D}}_i$ as (23). As a result, the modified channel matrix $\tilde{\mathbf{H}}_i$ is

$$\tilde{\mathbf{H}}_i = \tilde{\mathbf{R}}_i \hat{\mathbf{D}}_i^{-1}. \quad (23)$$

where $\tilde{H}(k, i)$ is the modified channel response. We compensated received signal's ISI as (21). However, the orthogonality of signals is destroyed by the subtracted signal due to the ISI compensation. To reconstruct the orthogonality, we consider the ICI compensation scheme with inserting the replica signals using the subtracted signal due to ISI compensation. The ICI equalized signal $\tilde{\mathbf{R}}_i$ with inserting the part of the signal previously detected $\hat{\mathbf{D}}_i$ into the signal lack part is given by

$$\begin{aligned} \tilde{\mathbf{R}}_i &= \mathbf{R}_i - \mathbf{H}_{i,isi} \mathbf{F} \hat{\mathbf{D}}_{i-1} + \tilde{\mathbf{H}}_i \mathbf{F} \hat{\mathbf{D}}_i + \tilde{\mathbf{N}}_i \\ &= \lambda_{i,pro} \mathbf{F} \hat{\mathbf{D}}_i + \tilde{\mathbf{N}}_i \end{aligned} \quad (24)$$

where $\tilde{\mathbf{N}}_i$ is the noise term with residual ISI and the residual ICI, $\lambda_{i,p}$ is the ISI and ICI subtracted channel matrix, respectively. By using the replica signal $\tilde{\mathbf{R}}'_i$ and the detected signal $\hat{\mathbf{D}}_i$, we get the channel response $\lambda_{i,pro}$ as

$$\lambda_{i,pro} = \mathbf{F}^{-1} \tilde{\mathbf{R}}'_i / \hat{\mathbf{D}}_i, \quad (25)$$

where $\tilde{\mathbf{R}}'_i = \tilde{\mathbf{H}}_{i,p} \mathbf{F} \hat{\mathbf{D}}_i + \tilde{\mathbf{H}}_{i,q} \mathbf{F} \hat{\mathbf{D}}_i$. By using $\lambda_{i,pro}$, received signal $\tilde{\mathbf{R}}_i$ is equalized, and the detected signal $\hat{\mathbf{D}}_i$ can be written as

$$\tilde{\mathbf{D}}_i = \frac{\mathbf{F}^{-1} \tilde{\mathbf{R}}_i}{\lambda_{i,pro}} = \mathbf{D}_i + \frac{\mathbf{F}^{-1} \tilde{\mathbf{N}}_i}{\lambda_{i,pro}} \quad (26)$$

Therefore, we can get the accurate signal $\tilde{\mathbf{D}}_i$.

III. COMPUTER SIMULATION RESULTS

In this section, we show the performance of the proposed method. Fig. 1 shows a simulation model of the proposed system. On the transmitter, the scattered pilot (SP) signals and QPSK modulated signals are multiplexing, and parallel to serial (P/S) converted. The arrangement of SP signals is shown in Fig. 2. The OFDM time signals are generated by an IFFT and transmitted to frequency selective and time variant radio channel after cyclic extensions have been inserted. The transmitted signals are subject to broadband channel propagation. In the simulation, we assume that OFDM symbol period is $5\mu s$, guard interval is $1\mu s$, channel model is L=2 path Rayleigh fading and the maximum delay spread is $1.06\mu s$. The maximum Doppler frequencies are assumed to be 10 and 600 Hz. The number of symbols N_s is 21, the number of subcarriers N_c is 64, the number of pilot signals N_p is 96, and the FFT size is 64. In the receiver, the guard interval is removed from the received signals, and S/P converted.

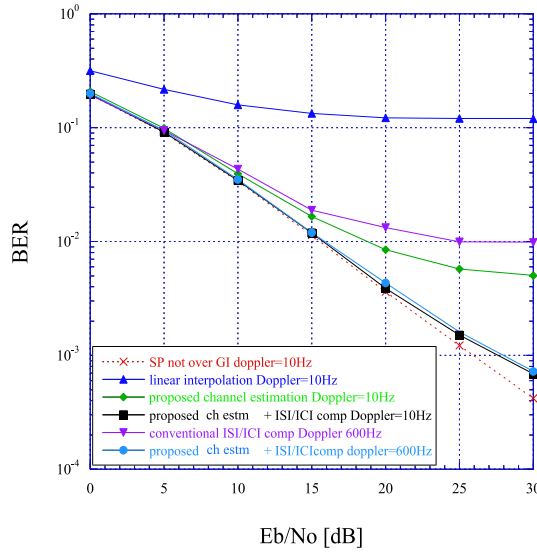


Fig. 4. BER vs. E_b/N_o

The parallel sequences are passed to an FFT operator, which converts the signal back to the frequency domain. After descrambling, SP signals are extracted. We estimate the CSI with the proposed channel estimation scheme, and demodulate the detected signals. Since the detected data signals include the ISI and ICI, it is necessary to remove them. At first, ISI compensation should be performed with the previous detected symbol and ISI channel matrix as Eq. (21). By using ZF, the ISI can be equalized. However, since it is destroyed the orthogonality of subcarriers due to the residual ICI, we compensate the ICI to reconstruct the orthogonality by inserting the replica signals made from the detected signals again. Fig.4 shows the BER of the proposed method and conventional method for large delay spread channel. The BER performance of the proposed method shows about 24 times smaller than that of the conventional frequency interpolation method at $E_b/N_o = 30\text{dB}$. In addition, ISI and ICI compensation methods improve the BER performance about more 7 times and show the approximately same BER performance compared with no ISI and ICI cases. Moreover it is shown that the proposed method is effective for Doppler frequency of 600Hz. Fig.5 shows the throughput performance for large delay spread channel. Throughput is calculated by

$$T_{tp} = \frac{N_{D_{bit}} R}{(N_d + N_p) T} (1 - P_{per}), \quad (27)$$

where $N_{D_{bit}}$ is the number of total data bit, R is the code rate, N_d is the number of data symbols, N_p is the number of pilot symbols, and P_{per} is the packet error rate, respectively. Fig.5 shows that the throughput performance of proposed system can achieve 5.7 % enhancement compared with conventional OFDM system.

IV. CONCLUSION

In this paper, we have focused on the channel estimation in the large delay spread channels. We have proposed the

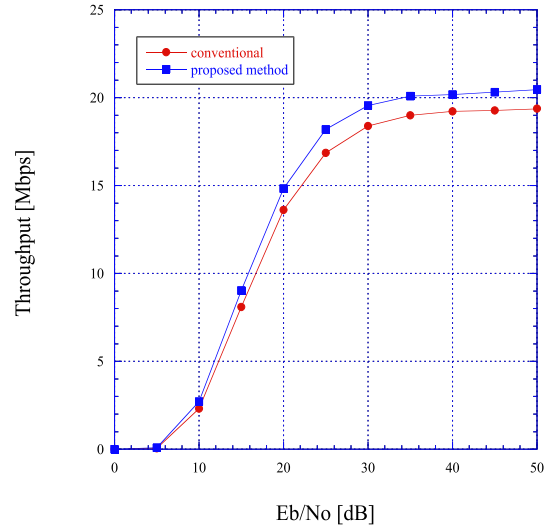


Fig. 5. Throughput vs. E_b/N_o

channel estimation scheme and ISI and ICI compensation scheme for SP-OFDM. In the proposed channel estimation scheme, we have performed time domain processing for the channel responses by using the IFFT and selected path to remove the noise term and the interference term by using the threshold. In addition, we can compensate ISI and ICI by using the replica signals. From the simulation results, it is shown that the proposed schemes achieve the approximately same BER performance like the case with no ISI and ICI. In addition, the proposed method is effective for the case with fast fading and achieves 5.7 % throughput enhancement compared with the conventional system.

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