Pilot-assisted Adaptive Interpolation Channel Estimation for OFDM Signal Reception

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Abstract- Accurate channel estimation is necessary for coherent detection of the orthogonal frequency division multiplexing (OFDM) signals. This paper studies pilot-assisted adaptive interpolation channel estimation for different pilot arrangements: time-multiplexed pilot, frequency-multiplexed pilot and scattered pilot. Interpolation filter tap weights are adaptively updated according to changes of the propagation channels by using the normalized least mean square (NLMS) algorithm. The average bit error rate (BER) performance in a doubly (frequency and time)-selective Rayleigh fading channel is evaluated by computer simulation. It is confirmed that the scattered pilot with two-dimensional adaptive interpolation channel estimation provides overall a good BER performance, compared to the time (or frequency)-multiplexed pilot case with one-dimensional adaptive interpolation channel estimation.

1. Introduction

Recently, there have been tremendous demands for high-speed data transmission in mobile communications. For achieving high-speed and high-quality data transmissions in a severe frequency-selective fading channel, orthogonal frequency division multiplexing (OFDM) has been considered as a promising wireless transmission technique [1,2]. In a mobile radio, the transmitted signal is reflected and diffracted by many obstacles between a transmitter and a receiver, thus creating the doubly (time and frequency)-selective fading channel [3]. Accurate channel estimation is necessary for coherent detection of OFDM signals in such fading channel.

Pilot-assisted channel estimation using one- and two-dimensional Wiener filtering has been analyzed in [4]-[6]. The pilot-assisted channel estimation using Wiener filtering requires the knowledge of channel statistics, e.g., the delay profile and the Doppler frequency, which are in general unknown to a receiver. Hence, in practical receivers, the Wiener filter is designed assuming the worst-case scenario, e.g., the worst case channel delay spread, the maximum Doppler shift. However, the fading channel condition may change according to user's movement. Hence, in order to achieve a better channel estimation, the conventional fixed Wiener filter can be replaced by an adaptive Wiener filter based on the estimated channel autocorrelation function [7]. Also, a pilot-assisted robust channel estimation using the two-dimensional fast Fourier transform (FFT) and inverse FFT (IFFT) has been proposed for OFDM systems [8].

In this paper, pilot-assisted adaptive interpolation channel estimation is studied for coherent detection of OFDM signals. We consider different pilot arrangements: time-multiplexed pilot, frequency-multiplexed pilot and scattered pilot. The tap weight adaptation presented in [9,10] is applied that uses the normalized least mean square (NLMS) algorithm [11]. The remainder of this paper is organized as follows. In Sect. 2, an OFDM transmission system model is presented. Sect. 3 describes a pilot-assisted adaptive interpolation channel estimation for the time-multiplexed pilot, frequency-multiplexed pilot and the scattered pilot. Then, Sect. 4 presents the computer simulation results on the achievable BER performance in a frequency-selective Rayleigh fading channel. Finally, Sect. 5 offers some conclusions.

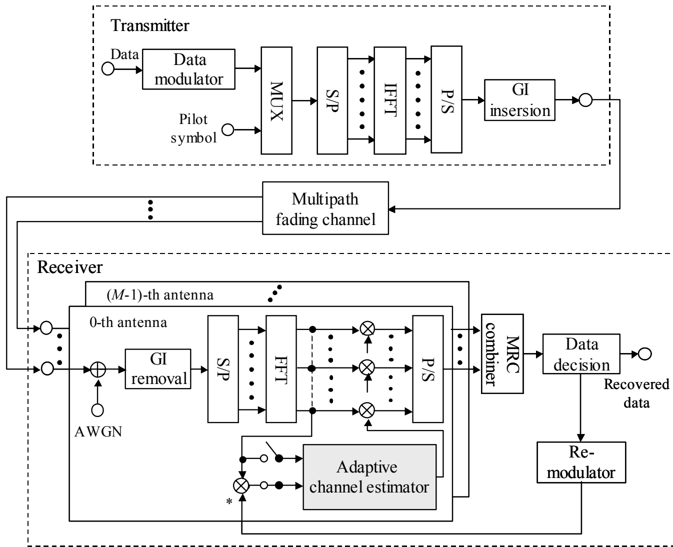

Fig. 1 OFDM transmission system model.

2. OFDM Transmission System Model

2.1 Signal representation

The OFDM transmission system model considered in this paper is illustrated in Fig. 1. OFDM signal transmission using N_c subcarriers is assumed. In the transmitter, the binary date sequence is transformed into the quadrature phase shift keying (QPSK) modulated symbol sequence. Then, the known pilot symbols are inserted into the data symbol sequence. After the pilot-inserted symbol sequence $\{d_n\}$ is serial-parallel

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(S/P) converted into N_c parallel data sequences each with a lower rate, OFDM signal waveform with N_c subcarriers is generated by applying N_c-point inverse fast Fourier transform (IFFT). Finally, the cyclic prefix of N_g samples is inserted into the guard interval (GI) in order to eliminate the inter-symbol interference (ISI) resulting from the frequency-selective fading.

The baseband equivalent representation of the OFDM signal for the i-th OFDM signaling period is given by

$$s(t)=\sqrt{\frac{2E_s}{(N_c+N_g)T_c}}\sum_{n=0}^{N_c-1}d(i,n)\exp\left(-j2\pi\frac{n}{N_c}(t'\bmod N_c)\right) \quad (1)$$

for $t'=t-i(N_c+N_g)= -N_g \sim N_c-1$, where E_s represents the transmitted symbol energy, $d(i,n)=d_{iN_c+n}$, and T_c represents the IFFT sampling interval given by $T/(N_c+ N_g)$ for a data symbol rate per subcarrier of $1/T$. The resultant OFDM signal is transmitted via a frequency-selective fading channel and received by M antennas for diversity reception.

A T_c-spaced delay-time model of the propagation channel is assumed. Assuming that the channel has L independent propagation paths with T_c-spaced time delays of $0 \sim (L-1)T_c$, the discrete-time impulse response $h_m(t)$ of the multipath channel experienced by the m-th antenna, m=0~M-1, may be expressed as

$$h_m(t)=\sum_{l=0}^{L-1}h_{m,l}\delta(t-l) \quad (2)$$

with $\sum_{l=0}^{L-1}\mathrm{E}[|h_{m,l}|^2]=1$, where $\delta(t)$ is the delta function and E[.] denotes ensemble average operation. Assuming slow fading so that the channel impulse response remains constant over one OFDM signaling interval, time dependency of the channel has been dropped for simplicity. It is assumed that the maximum time delay of the channel is shorter than GI.

The received signal is sampled at a rate of $1/T_c$. The signal sample sequence on the m-th antenna is

$$r_m(t)=\sum_{l=0}^{L-1}h_{m,l}s(t-l)+\mu_m(t)\,, \quad (3)$$

where $\mu_m(t)$ is the zero mean noise sample having variance $2N_0/T_c$ due to the additive white Gaussian noise (AWGN). After removal of GI, the received signal sample sequence is resolved into N_c-subcarrier components by applying N_c-point FFT to obtain

$$R_m(i,n)=\sqrt{\frac{2E_s}{(N_c+N_g)T_c}}d(i,n)H_m(i,n)+\Pi_m(i,n) \quad (4)$$

for the n-th subcarrier, where $H_m(i,n)$ and $\Pi_m(i,n)$ are the channel gain and the zero-mean noise sample having variance $2(N_0/T_c)N_c$, respectively, at the n-th subcarrier. Then, using the channel gain estimates $\{\widetilde{H}_m(i,n)\}$, coherent detection and antenna diversity reception using maximal ratio combining (MRC) [3] are carried out as

$$\eta(i,n)=\sum_{m=0}^{M-1}R_m(i,n)\widetilde{H}_m^*(i,n)\Big/\sum_{m=0}^{M-1}\left|\widetilde{H}_m(i,n)\right|^2\,, \quad (5)$$

where * denotes the complex conjugate operation. Finally data demodulation for symbol $d(i,n)$ is carried out as

$$\bar{d}(i,n)=\arg\min_{d}|\eta(i,n)-d|^2\,. \quad (6)$$

For coherent detection, estimation of $\{H_m(i,n)\}$ is necessary. In Sect. 3, the pilot-assisted adaptive interpolation channel estimation is described.

2.2 Pilot arrangement

The different pilot symbol arrangements (time-multiplexed pilot, frequency-multiplexed pilot and scattered pilot case) are considered as shown in Fig. 2. For the time (or frequency)-multiplexed pilot case, one pilot symbol (or subcarrier) is inserted every N_p symbols (or N_p subcarriers), while one pilot symbol is inserted every $\sqrt{N_p}$ OFDM symbols and $\sqrt{N_p}$ subcarriers for the scattered pilot case (the transmission rate is kept the same for the three pilot cases).

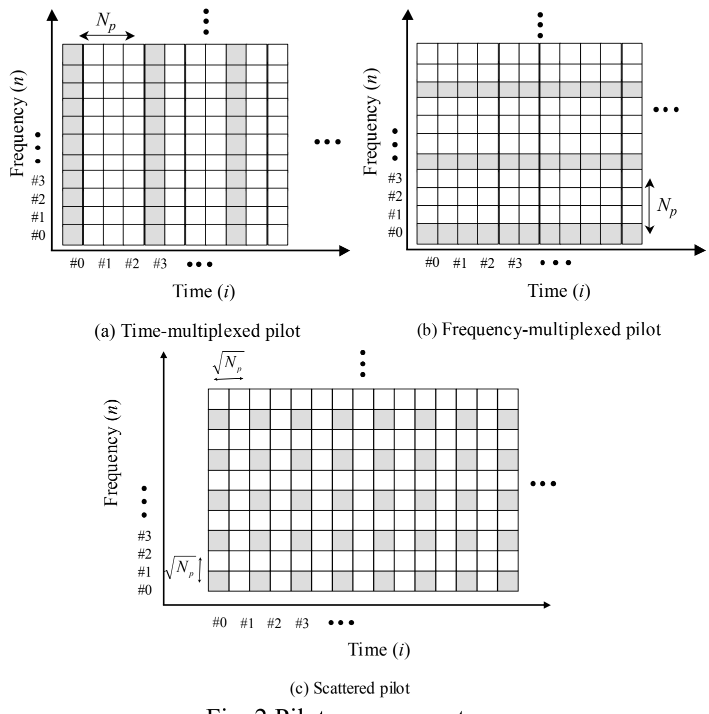

Fig. 2 Pilot arrangement.

3. Adaptive Interpolation Channel Estimation

3.1 Time-multiplexed and frequency-multiplexed pilot cases

One-dimensional adaptive interpolation filter in time (or frequency) is applied to the time (or frequency)-multiplexed pilot. The filter structure is shown in Fig. 3. Pilot symbol and succeeding N_p-1 data symbols in time (or frequency) constitutes the slot. The adaptive interpolation filter estimates the channel gains at N_p-1 different data symbol positions in a slot. Hence, a total of N_p-1 sets of 2K tap weights are used. The same set of 2K weights is reused for channel estimation at the same symbol position in each slot.

First, the instantaneous channel gains $\{\hat{H}_m(i,n)\}$ at the 2K pilot positions in time (or frequency) are estimated using the received pilot for time-multiplexed case (or frequency-multiplexed case). Assuming that the pilot symbol is 1+j0, $\{\hat{H}_m(i,n)\}$ are obtained using a smoothing filter as follows.

(a) time-multiplexed pilot (i mod N_p=0)

$$\hat{H}_m(i,n)=\begin{cases} R_m(i,n) \text{ for } n=0, N_c-1 \\ \sum_{\substack{k=-\alpha(n)\\ k\neq 0}}^{k=\alpha(n)} \beta_{m,j}(k) R_m(i,n+k) \text{ otherwise} \end{cases}, (7)$$

where $\{\beta_{m,j}(k);\ k=-K+1\sim K\}$ are the complex tap weights of a smoothing filter in frequency and $\alpha(n)$ is given by [9]

$$\alpha(n)=\begin{cases} n & \text{for } 1\leq n\leq K-1 \\ K & \text{for } K\leq n\leq N_c-K-1 \\ N_c-1-n & \text{for } N_c-K\leq n\leq N_c-2 \end{cases}. (8)$$

(b) frequency-multiplexed pilot (n mod N_p=0)

$$\hat{H}_m(i,n)=\sum_{\substack{k=-K\\ k\neq 0}}^{k=K} \beta_{m,j}(k) R_m(i+k,n), (9)$$

Then, the channel gain estimates for all N_p-1 data symbol positions in a slot are estimated by the adaptive interpolation filter using $\{\hat{H}_m(i,n)\}$ as

$$\tilde{H}_m(i,n)=\mathbf{W}_{m,j}{}^{\mathrm{T}}(q)\mathbf{X}_m(i,n), (10)$$

where $\mathbf{W}_{m,j}(q)=\{w_{m,j}(q,k);k=-K+1\sim K\}^{\mathrm{T}}$ is the 2K-by-1 complex tap weight vector of adaptive interpolation filter, obtained after jth updating, of the q-th symbol position in a slot, with q=i mod N_p and q=n mod N_p for the frequency- and time-multiplexed pilot cases, respectively. $\mathbf{X}_m(i,n)=\{x_m(i,n,k);k=-K+1\sim K\}^T$ is the 2K-by-1 instantaneous channel gain estimate vector with

$$x_m(i,n,k) = \begin{cases} \hat{H}_m(\lfloor i/N_p \rfloor N_p+kN_p, n) \text{ for time-mux pilot case} \\ \hat{H}_m(i,\lfloor n/N_p \rfloor N_p+kN_p) \text{ for freq.-mux pilot case} \end{cases}. (11)$$

In channel estimation for subcarriers close to the edge for the frequency-multiplexed pilot case, the instantaneous channel gains outside the transmission bandwidth are assumed to be zero, i.e, $\{\hat{H}_m(i,n)=0;\ n\leq 0 \text{ and } n\geq N_c\}$.

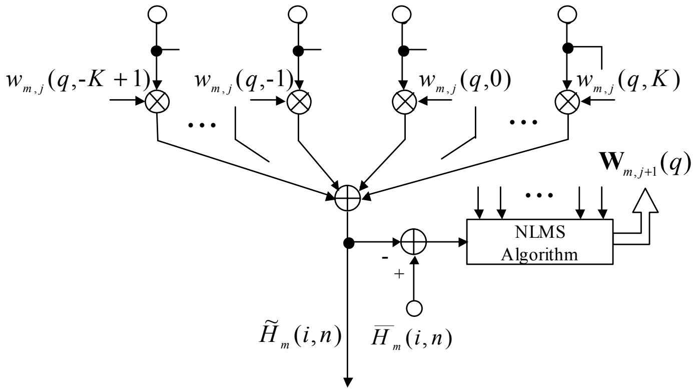

Fig. 3 Adaptive interpolation filter structure.

3.2 Scattered pilot case

We take a heuristic approach to implement a two-dimensional adaptive interpolation filter as in [5,6]; the one-dimensional interpolation filter in time is followed by one-dimensional interpolation filter in frequency. When the channel gain estimates provided by first one-dimensional filter are used as the input to the second one-dimensional filter, the optimal tap weights of the second filter are a function of the first filter's tap weighs. However, assuming that the input to the second filter does not depend on the first filter's tap weighs, the simplified approach [12] is taken to reduce the number of tap weight sets.

For the scattered pilot case, using $\{\hat{R}_m(i,n)\}$ at subcarriers of i mod N_p=0, one-dimensional adaptive interpolation filtering in time is carried out to obtain the channel gain estimates at a subcarrier of n mod N_p=0. The channel gain estimate $\breve{H}_m(i,n)$ is expressed as

$$\breve{H}_m(i,n)=\mathbf{W}_{m,j}{}^{\mathrm{T}}(q)\mathbf{X}_m(i,n), (12)$$

where $\mathbf{X}_m(i,n)$ is the 2K-by-1 instantaneous channel gain estimate vector in time at a subcarrier of n mod N_p=0. Then, one-dimensional adaptive interpolation filtering in frequency is carried out to obtain the channel estimate of all subcarrier and time positions. The final channel gain estimate is obtained using $\{\breve{H}_m(i,n)\}$ as

$$\tilde{H}_m(i,n)=\mathbf{W}_{m,j}{}^{\mathrm{T}}(q)\breve{\mathbf{X}}_m(i,n), (13)$$

where $\breve{\mathbf{X}}_m(i,n)=\{\breve{H}_m(i,\lfloor n/N_p \rfloor N_p+kN_p;k=-K+1\sim K\}^{\mathrm{T}}$ is the 2K-by-1 channel gain estimate vector obtained from Eq. (12).

3.3 Tap weight adaptation

For tap weight adaptation of one-dimensional adaptive interpolation channel estimation, the NLMS algorithm is applied that uses the noisy instantaneous channel gains obtained by reverse modulation as the reference signals [10]. Tap weight adaptation of two-dimensional adaptive interpolation channel estimation is the same as that for one-dimensional case. Each set of tap weights is adaptively updated N_p times every slot according to the NLMS algorithm:

$$\begin{cases} \mathbf{W}_{m,j+1}(q) = \mathbf{W}_{m,j}(q) + \mu \dfrac{e_{m,j}(q)\mathbf{X}_m^*(i,n)}{\left\|\mathbf{X}_m(i,n)\right\|^2} & \text{for all } i, n \\ e_{m,j}(q) = R_m(i,n)\bar{d}^*(i,n) - \widetilde{H}_m(i,n) \end{cases} \quad (14)$$

where μ is the step size, $q=i$ (n) mod N_p for time (frequency)-multiplexed pilot. The updating method of tap weights $\beta_{m,j}(k)$ for the smoothing filter in Eqs. (7) and (9) is described in [9].

4. Computer Simulation

The average bit error rate (BER) performance in a frequency-selective Rayleigh fading channel is evaluated by computer simulation. The simulation condition is summarized in Table 1. Quadrature phase shift keying (QPSK) data modulation and 2-branch antenna diversity reception are considered and the number of subcarriers is N_c=1024. The multipath fading channel is assumed to have an L=8-exponentially decaying power delay profile with time delay separation of $32T_c$. Each path is subject to an independent Rayleigh fading. It is assumed that the maximum time delay is shorter than the guard interval of N_g samples and the complex channel gains remain almost constant over the one OFDM signaling period $T=(N_c+N_g)T_c$. At the receiver, M=2-branch MRC antenna diversity reception is assumed. The number of adaptive interpolation filter tap weights $2K$ is 6 for all pilot cases.

Table 1 Simulation condition.

OFDM	Data modulation	QPSK
	Number of subcarriers	N_c=1024
	Guard interval samples	$N_g=N_c/4$
Propagation channel model		8-path Rayleigh having an exponential power delay profile
Antenna diversity		2-branch MRC
No. of tap weights		$2K$=6

We compare the average BER performances with time-multiplexed pilot, frequency-multiplexed pilot and scattered pilot with N_p=9. Figure 4 plots the simulated average BER performance as a function of the average received signal energy per bit-to-AWGN power spectrum density ratio E_b/N_0. The normalized maximum Doppler frequency f_DT is 0.02 and the normalized delay spread τ_{rms}/T_s is 0.0074. For comparison, the BER performance of ideal channel estimation is also plotted. We can see from Fig. 4 that the BER performances of scattered pilot and time-multiplexed pilot are superior to that of frequency-multiplexed pilot, since the channel estimation accuracy at the edge frequency is significantly degraded for the frequency-multiplexed pilot case. The scattered pilot case using two-dimensional adaptive interpolation channel estimation is found to provide a slightly better BER performance than the time-multiplexed pilot case. The degradation in the required average E_b/N_0 for BER=10^{-3} from the ideal channel estimation case is 1.1 dB, 1.3 and 2.1dB for scattered, time- and frequency-multiplexed pilot cases, respectively.

Figures 5 and 6 plot the simulated BERs at the average received E_b/N_0 per antenna=20dB as a function of f_DT and τ_{rms}/T, respectively. The time (or frequency)-multiplexed pilot using one-dimensional adaptive interpolation in time (or frequency) is very robust against frequency (or time)-selective fading; however, its robustness is significantly lost as the channel time (or frequency)-selectivity tends to become stronger. On the other hand, the scattered pilot using two-dimensional adaptive interpolation channel estimation has overall a good robustness against both time- and frequency-selectivity of the fading channel.

5. Conclusion

In this paper, a pilot-assisted adaptive interpolation channel estimation for OFDM signal reception was studied for for different pilot arrangements (time-multiplexed pilot, frequency-multiplexed pilot and scattered pilot). One-dimensional and two-dimensional adaptive interpolation filtering using the normalized least mean square (NLMS) algorithm were presented. The average BER performance in a doubly (time and frequency)-selective Rayleigh fading channel was evaluated by computer simulation. It was found that the scattered pilot using two-dimensional adaptive interpolation channel estimation provides overall a good BER performance.

Acknowledgement

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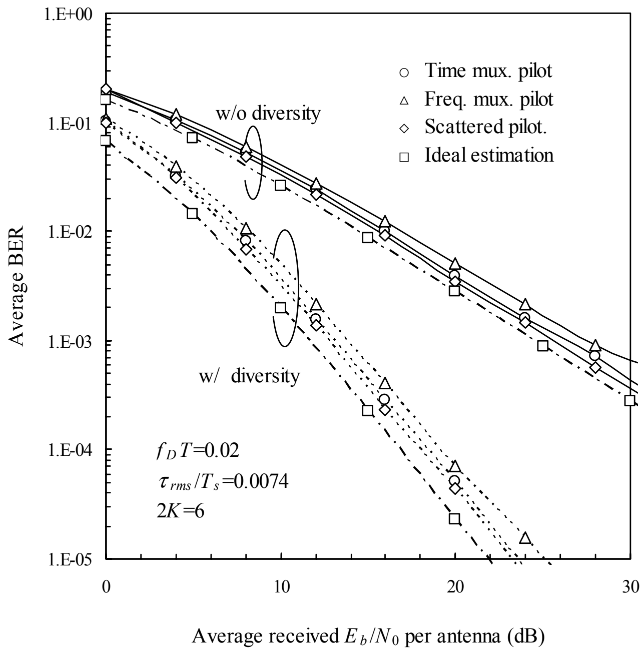

Fig. 4 Average BER performance.

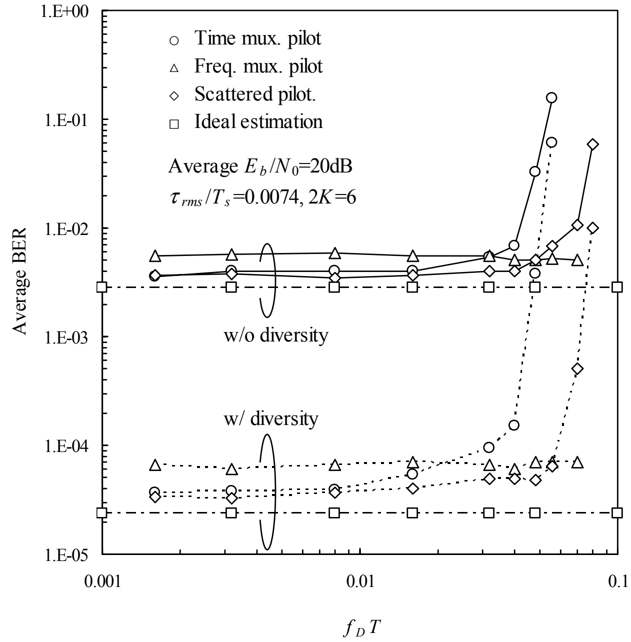

Fig. 5 Impact of time-selectivity of fading channel.

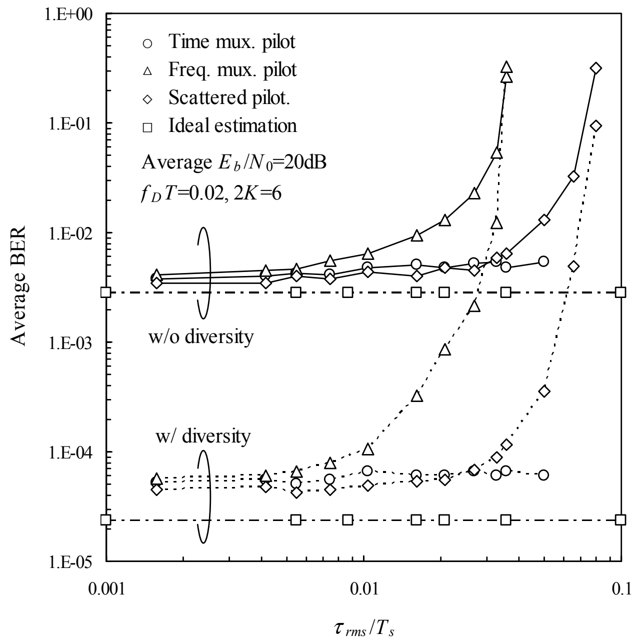

Fig. 6 Impact of frequency-selectivity of fading channel.