

PILOT-ASSISTED ADAPTIVE CHANNEL ESTIMATION USING MULTIPLE SETS OF TAP WEIGHTS FOR COHERENT RAKE RECEPTION OF DS-CDMA SIGNALS

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Abstract

For the coherent reception of DS-CDMA signals, this paper proposes a new adaptive channel estimation scheme using multiple sets of adaptive filter tap weights, each set being used for channel estimation at different symbol position. For updating each set of tap weights using the normalized least mean square (NLMS) algorithm, a block of known pilot symbols is time-multiplexed with the data modulated symbols. The average bit error rate (BER) performance in a frequency-selective Rayleigh fading channel is evaluated by computer simulation. It is confirmed that the proposed channel estimation scheme provides better BER performance than the conventional channel estimation schemes.

1. Introduction

Direct sequence code division multiple access (DS-CDMA) is used in present cellular mobile communications systems [1]. For coherent rake reception of DS-CDMA signals in a multipath fading channel [2], accurate channel estimation is necessary. Among many channel estimation schemes so far proposed [3]-[10], pilot-assisted channel estimation called weighted multi-slot averaging (WMSA) channel estimation uses a finite impulse response (FIR) filter using the time invariant tap weights [9]. However, using the time invariant tap-weights cannot always minimize the bit error rate performance (BER) in a changing multipath propagation environment due to user's movement. In addition to fading, the presence of the frequency offset between a transmitter and a receiver may degrade the BER performance due to constant phase rotation in the received signal. Adaptive channel estimation schemes proposed in [11], [12] adapt the complex-valued tap weights according to a change in the fading conditions and can also track the constant phase rotation due to the frequency offset. The adaptive prediction channel estimation proposed in [12] is an extension of the conventional WMSA channel estimation and uses the forward prediction and backward prediction FIR filters. Their tap weights are updated using the normalized least mean square (NLMS) algorithm [13].

In this paper, a pilot-assisted adaptive channel estimation using multiple sets of tap weights is proposed that simultaneously estimates the channel gains at different symbol positions. Each set of adaptive filter tap weights is updated using the NLMS algorithm, the reference signal for which is obtained by reverse modulating the received data symbol at the symbol position of interest. The remainder of this paper is organized as follows. A transmission system model is presented in Sect. 2. In Sect.

3, the operation principle of the proposed adaptive channel estimation scheme is described. Sect. 4 presents the computer simulation results in a frequency-selective Rayleigh fading channel and compares the simulation results with those using the non-adaptive WMSA channel estimation [9] and the adaptive prediction channel estimation [12]. Also discussed is the BER performance in the presence of frequency offset between a transmitter and a receiver. Sect. 5 concludes the paper.

2. Transmission System Model

The transmission system model and the slot structure are illustrated in Figs. 1 and 2, respectively. At the transmitter, a binary data to be transmitted is transformed into data-modulated symbol sequence. Then, a block of known N_p pilot symbols is time-multiplexed every N_d data symbols as shown in Fig. 2. N_p pilot symbols and succeeding N_d data symbols make a data slot with a length of $N_{slot}=N_p+N_d$. Finally, the pilot-inserted data-symbol sequence is multiplied by a spreading sequence to produce the DS-CDMA signal $s(t)$. $s(t)$ can be expressed in the equivalent baseband representation as

$$\begin{cases} s(t) = \sqrt{2S}d(t)p(t) \\ d(t) = \sum_{k=-\infty}^{\infty} d_k u(t/T - k) \\ p(t) = \sum_{q=-\infty}^{\infty} c_q u(t/T_c - q) \end{cases}, (1)$$

where S denotes the signal power, $d(t)$ represents the data-symbol sequence waveform, $p(t)$ the spreading sequence waveform, and T the data-symbol duration. In Eq. (1), $\{d_k\}$ represents the data-symbol sequence, $\{c_q\}$ represents the spreading sequence with a chip duration of T_c , and $u(t)$ is the rectangular pulse with $u(t)=1$ ($0 \leq t < 1$) and 0 (otherwise). The spreading factor SF is given by $SF=T/T_c$.

The DS-CDMA signal $s(t)$ is transmitted over a multipath propagation channel. It is assumed that the propagation channel is frequency-selective and has L discrete paths having different time delays of integer multiple of T_c , and experiencing independent fading. The receiver has a total of M spatially separated antennas. The channel impulse response $h_m(t, \tau)$ at time t , seen on the m th antenna, $m=0 \sim M-1$, can be expressed as [14]

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

where $\xi_{m,l}(t)$ and τ_l denote the complex path gain and time delay of the l th path, respectively, with $E[\sum_{l=0}^{L-1} |\xi_{m,l}(t)|^2] = 1$; $E[.]$ is the ensemble average operation.

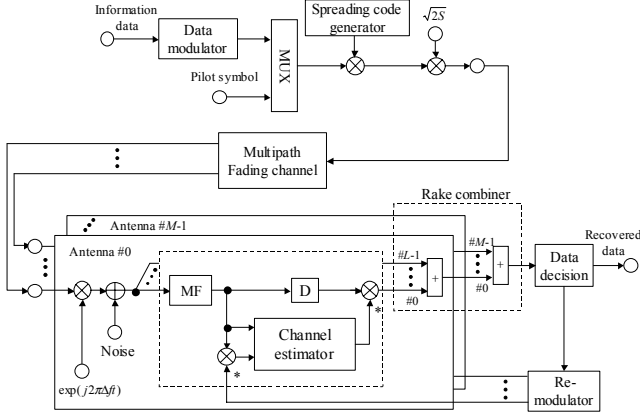


Fig. 1 Transmission system model.

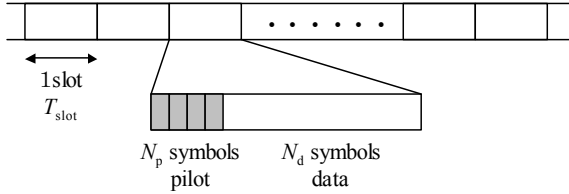


Fig. 2 Slot structure.

The received signal $r_m(t)$ on the m th antenna in the presence of the frequency offset Δf between the transmitter and the receiver can be expressed as

$$r_m(t) = \sum_{l=0}^{L-1} \xi_{m,l}(t) \exp(j2\pi\Delta f t) s(t - \tau_l) + v_m(t), \quad (3)$$

where $v_m(t)$ represents the additive white Gaussian noise (AWGN) having the single-sided power spectrum density of N_0 . We have assumed that the receiver filter is wide enough not to distort the received signal at all due to the frequency offset. The received DS-CDMA signal is despread and resolved into L matched copies of transmitted data-symbol sequence by a matched filter (MF). The MF output $r_{m,l}(g, n)$ at the n th symbol time epoch of the g th data slot, associated with the l th path, is represented as

$$\begin{aligned} r_{m,l}(g, n) &= \frac{1}{T} \int_{gT_{slot} + nT + \tau_l}^{gT_{slot} + (n+1)T + \tau_l} r_m(t) p(t - \tau_l) dt \\ &= \sqrt{2S} \xi_{m,l}(g, n) d(g, n) \\ &\quad \times \exp[j2\pi\Delta f (gT_{slot} + nT)] + v_{m,l}(g, n) \end{aligned}, \quad (4)$$

where $\xi_{m,l}(g, n) = \xi_{m,l}(gT_{slot} + nT)$ with $T_{slot} = N_{slot}T$,

$d(g, n) = d_{g(N_p + N_d) + n}$, and $v_{m,l}(g, n)$ represents the noise plus inter-path interference (IPI) component. In the rake combiner, a total of $M \times L$ MF outputs are coherently summed up based on maximal ratio combining (MRC) [2]. Representing the channel gain estimate as $\hat{\xi}_{m,l}(g, n)$, the rake combiner output $\eta(g, n)$, which is the decision variable, can be expressed as

$$\eta(g, n) = \sum_{m=0}^{M-1} \sum_{l=0}^{L-1} r_{m,l}(g, n) \hat{\xi}_{m,l}^*(g, n), \quad (5)$$

where $*$ denotes the complex conjugation.

3. Adaptive Channel Estimation Using Multiple Sets of Tap Weights

Fig. 3 illustrates the block diagram of adaptive FIR filter. Q sets of the $2K$ tap weights are provided for simultaneous channel estimation, where $Q \leq N_d$ (with N_d/Q being an integer). This adaptive channel estimation filter is necessary for each resolved path. First, the instantaneous path gain is estimated by coherent addition of N_p received pilot symbols as in the conventional WMSA channel estimation [9]. Without loss of generality, the pilot symbol $d=1+j0$ is assumed. The instantaneous channel estimate, $\hat{\xi}_{m,l}(g)$, at the beginning of the g th slot is given by

$$\hat{\xi}_{m,l}(g) = \frac{1}{N_p} \sum_{n=0}^{N_p-1} r_{m,l}(g, n). \quad (6)$$

Then, using the K past and K future instantaneous channel gains, $\{\hat{\xi}_{m,l}(g+k); k = -K+1 \sim +K\}$, the channel gains at N_d different data symbol positions in the g th slot of interest are estimated by the $2K$ -tap adaptive filter. Q sets of the $2K$ tap weights are used. When $Q=N_d$, channel estimation is carried out at each data symbol position in a slot. On the other hand, when $Q=1$, channel estimation using one set of the $2K$ tap weights is carried out at each data symbol position in a slot. The channel gain estimate at the n th data symbol position of the g th slot, associated with the l th path, is given by

$$\begin{aligned} \tilde{\xi}_{m,l}(g, n) &= \sum_{k=-K+1}^K w_{m,l,q}(g, k) \hat{\xi}_{m,l}(g+k) \\ &= \mathbf{W}_{m,l,q}^T(g) \mathbf{X}_{m,l}(g) \end{aligned} \quad (7)$$

for $n=q(N_d/Q) \sim (q+1)(N_d/Q)-1$, with

$$\begin{cases} \mathbf{W}_{m,l,q}(g) = [w_{m,l,q}(g, K), \dots, w_{m,l,q}(g, -K+1)]^T \\ \mathbf{X}_{m,l}(g) = [\hat{\xi}_{m,l}(g+K), \dots, \hat{\xi}_{m,l}(g-K+1)]^T \end{cases}, \quad (8)$$

where $\mathbf{W}_{m,l,q}(g)$ and $\mathbf{X}_{m,l}(g)$ are the complex tap weight vector of the q th weight set ($q=0 \sim Q-1$) and the instantaneous channel gain estimate vector used for the g th

slot, respectively, and $[\cdot]^T$ denotes transposition.

For the adaptation of q th set of $2K$ tap weights, the NLMS algorithm [13] is applied. After channel estimation and data decision for the g th slot, the $2K$ -tap weight vector is updated. The reference signal $\bar{\xi}_{m,l,q}(g)$ for NLMS algorithm is obtained using reverse modulation and N_d/Q -symbol averaging as

$$\begin{cases} \bar{\xi}_{m,l,q}(g) = \frac{Q}{N_d} \sum_{k=0}^{N_d/Q-1} \hat{\xi}_{m,l,q}(g, N_p + qN_d/Q + k), & (9) \\ \hat{\xi}_{m,l,q}(g, n) = r_{m,l}(g, n) \hat{d}^*(g, n) \end{cases}$$

where $\hat{d}(g, n)$ is the decision result on the n th symbol in the g th slot. The recursive relation for updating the tap weight vector is expressed as

$$\begin{cases} \mathbf{W}_{m,l,q}(g+1) = \mathbf{W}_{m,l,q}(g) + \mu \frac{e_{m,l,q}(g) \mathbf{X}_{m,l}^*(g)}{\sum_{k=-K+1}^K |\hat{\xi}_{m,l,q}(g+k)|^2}, & (10) \\ e_{m,l,q}(g) = \bar{\xi}_{m,l,q}(g) - \tilde{\xi}_{m,l,q}(g, n) \end{cases}$$

where $e_{m,l,q}(g)$ is the estimation error and μ is the step size.

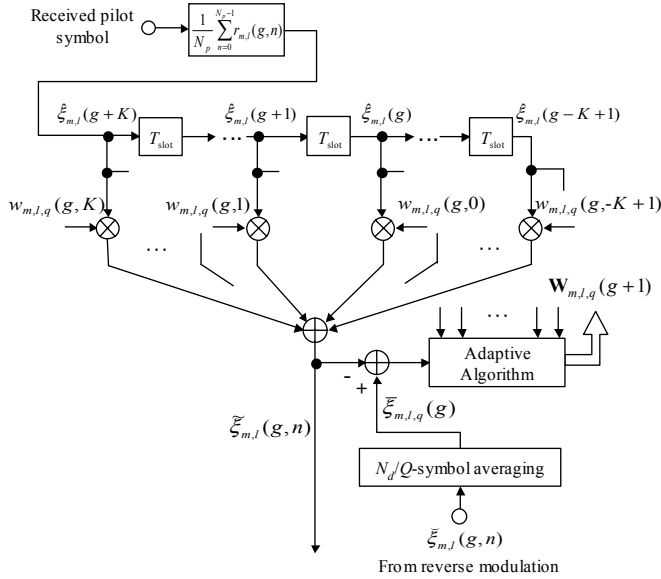


Fig. 3 Adaptive filter structure.

4. Computer Simulation

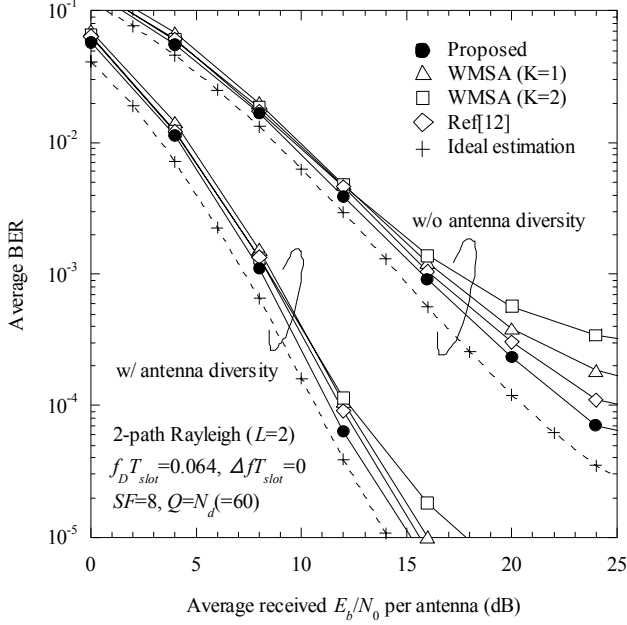
We assume a frequency-selective Rayleigh fading channel with a 2-path uniform power delay profile. Data modulation and spreading modulation are quadrature phase shift keying (QPSK) and binary PSK (BPSK), respectively. Since channel estimation accuracy depends on the value of N_p , the selection of N_p is an important issue. As N_p increases, the signal-to-noise ratio (SNR) of the instantaneous channel estimate $\hat{\xi}_{m,l,q}(g)$ of Eq. (6) improves due to coherent addition of N_p received pilot symbols. Accordingly, the BER performance improves.

However, the transmission of too many non-information bearing pilot symbols degrades the BER performance due to the increased power loss. In the following computer simulations, we set $N_p=4$ and $N_d=60$ as in [12]. For comparison, we assume the pilot-aided adaptive prediction channel estimation using linear interpolation [12] (hereafter, simply referred to as adaptive prediction channel estimation) and the WMSA channel estimation [9] using $K=1$ and 2 (the tap weight vectors for $K=1$ - and $K=2$ -WMSA channel estimation are $\{1.0, 1.0\}$ and $\{0.6, 1.0, 1.0, 0.6\}$, respectively [9]).

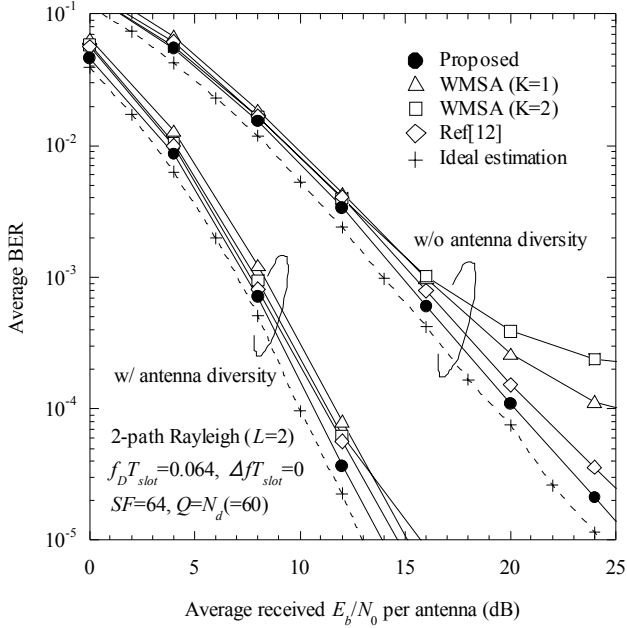
Table 1. Simulation condition

Transmitter	Spreading factor	$SF=8, 64$	
	Spreading sequence	Long PN sequence	
	Modulation	Data	QPSK
		Spreading	BPSK
Slot structure	Pilot	$N_p=4$ symbols	
	Data	$N_d=60$ symbols	
Propagation channel model	Rayleigh fading with 2-path uniform power delay profile		
Receiver	Diversity	$M=2$ -branch antenna diversity + 2-finger rake	
	No. of tap weights	$2K=8$	

The BER performance achievable with the proposed channel estimation scheme is compared with the WMSA channel estimation and the adaptive prediction channel estimation. First, we consider the case of no frequency offset Δf between the transmitter and receiver. Fig. 4 plots the average BER performance as a function of the average received E_b/N_0 per antenna for $f_D T_{\text{slot}}=0.064$. Fig. 4(a) and 4(b) are for $SF=8$ and 64, respectively. $Q=N_d$ ($=60$ symbols) sets of $2K$ tap weights are used (i.e., each symbol position in a slot is provided a different set of $2K$ tap weights for channel estimation). The WMSA channel estimation is designed to emphasize on reducing the noise effect at the cost of slightly losing the tracking ability against fading. In a low (high) E_b/N_0 region, $K=2(1)$ -WMSA channel estimation provides better BER performance than $K=1(2)$ -WMSA channel estimation. In a high E_b/N_0 region, however, the WMSA channel estimation does not perform well and produces the BER floors. However, the proposed adaptive channel estimation always provides the best BER performance. When $SF=64$, the proposed adaptive channel estimation with 2-antenna diversity reception reduces the required average E_b/N_0 per antenna for $\text{BER}=10^{-4}$ by 0.6(0.4) dB compared with $K=1(2)$ -WMSA channel estimation and by 0.3 dB compared with adaptive prediction channel estimation. The BER performance using the proposed adaptive channel estimation approaches that using ideal channel estimation and the degradation in E_b/N_0 for achieving $\text{BER}=10^{-4}$ is as small as 0.7 dB (0.28dB of which is due to power loss owing to pilot insertion).



(a) $SF=8$



(b) $SF=64$

Fig. 4 BER performance comparison.

Fig. 5 shows how the fading rate impacts the channel estimation accuracy when $SF=64$ and $L=2$. The fading rate is represented by the normalized maximum Doppler frequency $f_D T_{slot}$. The average BERs with diversity reception at average received $E_b/N_0=12\text{dB}$ are plotted in the figure. With the proposed adaptive channel estimation scheme, almost the same BER can be achieved if $f_D T_{slot} < 0.2$. However, with the other channel estimation schemes, as $f_D T_{slot}$ increases, the tracking ability of channel estimation against fading tends to be lost and hence, the achievable BER increases.

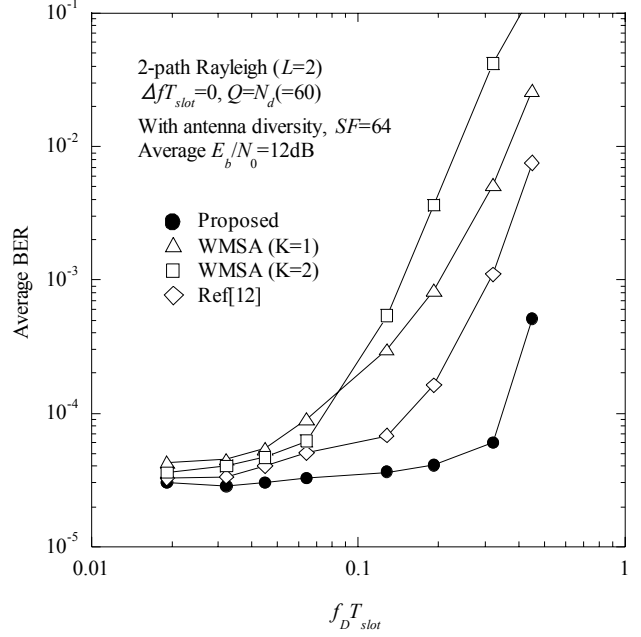


Fig. 5 Impact of the normalized maximum Doppler frequency $f_D T_{slot}$ when $SF=64$.

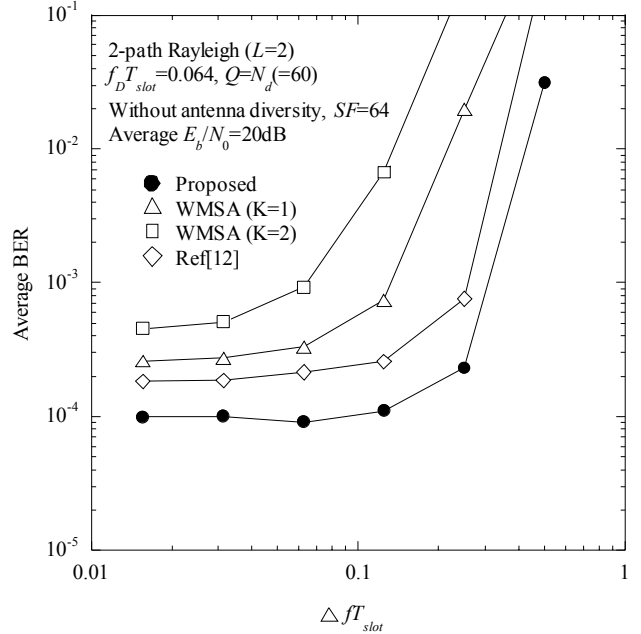


Fig. 6 Impact of the normalized frequency offset $\Delta f T_{slot}$ when $f_D T_{slot}=0.064$.

Fading produces random phase rotations in the received DS-CDMA signal. If the frequency offset is present between the transmitter and receiver, the received signal experiences the constant phase rotation in addition to the random phase rotation. Below, we will discuss how the frequency offset affects the channel estimation accuracy and then, the achievable BER performance. Fig. 6 plots the average BER at average received $E_b/N_0=20\text{dB}$ as a function of the normalized frequency offset $\Delta f T_{slot}$ for the case of no diversity and $SF=64$. $Q=60$ sets of $2K$ tap weights are used. For very small values of $\Delta f T_{slot}$, the predominant cause of

decision errors is the AWGN and hence, the BER is almost constant. As $\Delta f T_{slot}$ increases, the BER starts to increase due to the presence of frequency offset. However, the proposed adaptive channel estimation scheme is very robust against the constant phase rotation due to frequency offset as well as random phase rotation due to fading, as seen in Fig. 5.

So far, we have assumed that each symbol position is provided the different set of $2K$ tap weights for channel estimation ($Q=N_d=60$). The same weight set is shared for channel estimation at N_d/Q consecutive symbol positions. However, reducing the number of tap weight sets may degrade the achievable BER performance. This is because different symbol positions may experience different channel gains in a fast fading environment. Fig. 7 plots the average BER performance with diversity as a function of Q for various values of $f_D T_{slot}$ when $SF=64$. Although the achievable BER increases as Q becomes smaller, as small as $Q=6$ sets of tap weights can be used with only a slight performance degradation.

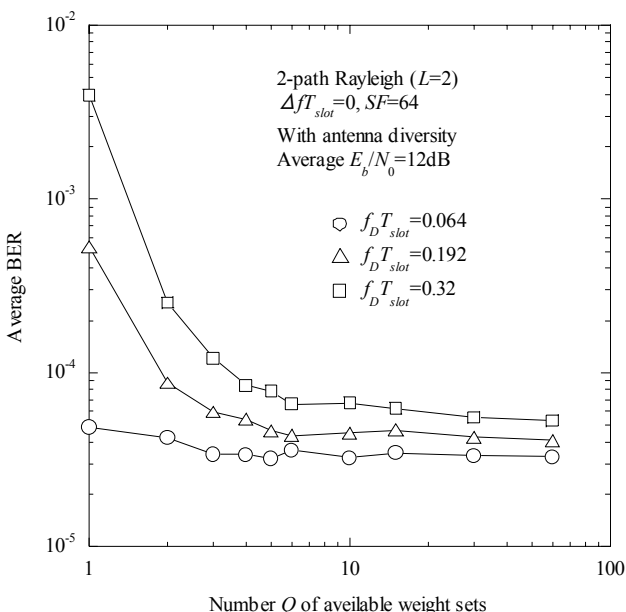


Fig. 7 Impact of the number Q of tap weight sets.

5. Conclusion

In this paper, for the coherent rake reception of DS-CDMA signals, a pilot-assisted adaptive channel estimation using multiple sets of tap weights was proposed that simultaneously estimates the channel gains at different symbol positions. In the proposed adaptive channel estimation scheme, each set of tap weights is updated using the NLMS algorithm. The reference signal for NLMS algorithm is obtained by reverse modulating the received data symbol. The achievable BER performance in a frequency-selective Rayleigh fading channel was evaluated by computer simulation and was compared with those of the non-adaptive WMSA channel estimation [9] and the adaptive prediction channel estimation using linear interpolation [12]. It was confirmed that the proposed adaptive channel estimation always provides the best BER performance. The proposed channel estimation was found

to be almost insensitive to the fading rate; almost the same BER can be achieved until $f_D T_{slot}$ approaches 0.2. The E_b/N_0 degradation at $BER=10^{-4}$ from ideal channel estimation is as small as 0.7 dB when $f_D T_{slot}=0.064$. It was also confirmed that the proposed adaptive channel estimation is very robust against the frequency offset and that the number Q of tap weight sets can be reduced to as small as $Q=6$ with only a slight performance degradation.

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