PAPER

Design of Channel Estimation Filters for Pilot Channel Based DS-CDMA Systems

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SUMMARY The accuracy of channel estimation significantly affects the performance of coherent receiver in a DS-CDMA system. The receiver performance can be improved if an appropriate channel estimation filter is used according to the channel condition. In this paper, we consider the design of channel estimation filters for pilot channel based DS-CDMA systems. When a moving average (MA) FIR filter is used as the channel estimation filter (CEF), the tap size is optimized by minimizing the mean squared error of the estimated channel impulse response. Finally, the analytic design is verified by computer simulation. Numerical results show that the optimum MA FIR CEF provides near optimum performance, i.e., quite similar to that with the use of Wiener filter.
key words: *channel estimation filter, DS-CDMA system, pilot channel*

1. Introduction

It is well known that the use of a rake receiver can take advantages of direct sequence code division multiple access (DS-CDMA) systems in multi-path fading condition [1]. The diversity effect can be obtained in a frequency selective channel by coherently combining the signal components of each finger in the rake receiver. The use of coherent detection in a rake receiver requires the channel impulse response (CIR) including the amplitude and phase of the channel, which can be obtained by employing a channel estimator [2]. In addition, the CIR is required for measurement of the signal to interference power ratio (SIR) for fast power control [3].

In DS-CDMA systems, there have been a number of studies on the channel estimation methods including the pilot-aided, data-aided and hybrid methods [3]. The pilot-aided method extracts the CIR from a known pilot signal. The pilot signal is time-multiplexed with the data signal in the pilot-symbol-aided method, while it is transmitted in parallel with the data signal using a different spreading sequence in the pilot-channel-aided method. In practice, WCDMA and cdma2000 systems adopt the use of a pilot-channel-aided method in both the uplink and downlink [4], [5]. In the data-aided channel estimation method, the CIR can be estimated by removing the data signal component from the received signal, where the data signal is obtained by tentative decision [6]. In the hybrid method, the data message is first tentatively detected using the CIR estimated from the pilot signal and then it is multiplied by the received signal to obtain an additional CIR. Finally, the two CIRs obtained from the pilot and data signals are combined into one [7]. In this paper, we consider the use of a pilot-channel-aided method for channel estimation, which can be applied to the IMT-2000 system. We do not use tentatively decoded data signal considering large additional computational complexity and memory.

In order to improve the accuracy of channel estimation, the obtained CIR is further processed by employing a lowpass filter called channel estimation filter (CEF) [2]. It is known that Wiener filter is optimum as the CEF in a stationary channel in the minimum mean squared error (MMSE) sense [2]. However, it is indispensable for the design of Wiener filter to know the power spectrum of the channel and noise, which may not be obtainable in real time. Moreover, it requires large implementation complexity including a large number of multiplication operations. The Doppler spectrum is usually spread to the maximum Doppler frequency of an experiencing channel [8]. Thus, it may be desirable to employ a brick-wall type lowpass filter as the CEF, whose cut-off frequency is equal to the maximum Doppler frequency of the channel [2], [9], [10]. However, such a CEF may not be practical due to the difficulty of implementation using a small number of filter taps. As a result, the CEF is realized in the form of a conventional lowpass filter such as finite impulse response (FIR) filter, moving average (MA) FIR filter or infinite impulse response (IIR) filter [3], [11], [12]. Such lowpass filters can provide relatively good channel estimation performance if they are appropriately designed according to the channel condition [6], [11], [12].

It is desirable to employ a simple-structured channel estimator, while providing good estimation performance. Although complicated CEF schemes such as Wiener CEF can be employed for channel estimation, their performance improvement is not significant since low-level modulation is used in DS-CDMA systems. In this paper, the use of MA FIR filters is considered as the CEF since it can be implemented simply, while providing relatively good performance [6], [13]. In addition, the characteristics of an MA FIR filter can be simply controlled by adjusting the tap size.

The performance of a channel estimator can significantly vary depending on the channel condition such as the maximum Doppler frequency. Thus, the CEF should be designed considering the channel condition. The CIR cannot be estimated properly if the cut-off frequency of the CEF is smaller than the maximum Doppler frequency. On the other hand, when the CEF has a cut-off frequency much larger than the maximum Doppler frequency of the channel, it pro-

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duces the output with excessive noise, yielding significant performance degradation. It is desirable to design a CEF so as to minimize the estimation error. However, few results have been reported on analytic design of the MA FIR CEF except some numerical results [6], [13], [14]. In this paper, we derive the optimum MA FIR CEF when power control is not employed. The tap size of the MA FIR CEF is analytically determined to minimize the mean squared error (MSE) of the estimated CIR. We also investigate the effect of fast power control on the design of the optimum MA FIR CEF.

Following Introduction, Sect. 2 describes the DS-CDMA system where a pilot signal is transmitted in parallel with data signal. The optimum MA FIR CEF is analytically designed by minimizing the MSE of the CIR in Sect. 3. The receiver performance with the use of the optimum MA FIR CEF is evaluated in Sect. 4. Finally, conclusions are summarized in Sect. 5.

2. DS-CDMA System with a Pilot Channel

Consider a DS-CDMA system that employs a common pilot channel in the downlink, where the transmit power of the common pilot signal is usually larger than that of the user signal [4], [5]. On the other hand, we assume that the dedicated pilot signal in the uplink is transmitted with the user signal in parallel [4], [5]. Since the SIR of the dedicated pilot signal in the uplink is usually less than that of common pilot signal in the downlink, the CIR may not be estimated accurately in the uplink as well as in the downlink. Thus, the use of an efficient channel estimation filter is much important in the uplink than in the downlink. In this paper, we focus on the channel estimation problem in the uplink. Note that the design of the channel estimator for the uplink can also be applied for the downlink.

In the transmitter, the user data is first convolutionally encoded with code rate C and constraint length H and then interleaved. Assume that the data and pilot signal are transmitted together through the real and imaginary path, respectively [4], [5]. Letting β^2 be the pilot to data signal power ratio, a baseband equivalent transmit signal $x(t)$ can be represented as

$$x(t) = \sum_{k=-\infty}^{\infty} \sum_{n=0}^{N-1} g[k] \big(x_d[k] w_d[kN+n] + j\beta x_p[k] w_p[kN+n]\big) q[kN+n] p_{T_c}(t-nT_c) \quad (1)$$

where N is the spreading factor, T_c is the chip duration interval, T is the symbol duration time equal to NT_c, $x_d[k]$ is the user message symbol having an amplitude equal to 1 or -1 at time $t = kT$, $x_p[k]$ is the pilot symbol having an amplitude equal to 1, $w_d[n]$ and $w_p[n]$ are bipolar orthogonal spreading sequences allocated to the data and pilot signal, respectively, $g[k]$ is the gain of the k-th symbol employed for fast power control, $p_{T_c}(t)$ is a unit-amplitude rectangular pulse defined in $[0, T_c]$ and $q[n]$ denotes a user specific complex PN scrambling sequence with $|q[n]| = 1$.

Assuming that there are L independent propagation paths with different channel gain and time delay, the CIR of the channel $h(t, \tau)$ at time t can be represented as

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

where $\delta(\tau)$ is Dirac delta function and τ_l is the delay of the l-th path signal. The CIR $h_l(t)$ of the l-th path can be represented by

$$h_l(t) = \alpha_l(t) e^{j\phi_l(t)} \quad (3)$$

where $\alpha_l(t)$ and $\phi_l(t)$ are the amplitude and phase response of the l-th path at time t, respectively. Considering both scattered multipaths and line-of-sight path, we assume a Ricean fading channel. The amplitude $\alpha_l(t)$ is Ricean distributed with Ricean factor K_l and $\phi_l(t)$ is uniformly distributed over $[0, 2\pi]$ [8]. Denoting P_l and θ_l by the power and arrival angle of the specular component and $\sigma^2_{\alpha,l}$ by the average power of the scattered component, K_l is equal to $P_l \sigma^{-2}_{\alpha,l}$. As a special case, the channel becomes a Rayleigh fading channel when K_l is zero.

The received signal $r(t)$ can be expressed by

$$\begin{aligned} r(t) &= \sum_{l=0}^{L-1} h_l(t - d_l T_c) x(t - d_l T_c) + \xi(t) + \nu(t) \\ &= \sum_{l=0}^{L-1} h_l(t - d_l T_c) x(t - d_l T_c) + n(t) \end{aligned} \quad (4)$$

where $n(t)$ denotes the total noise term composed of the background noise $\nu(t)$ and other users' interference $\xi(t)$. Assuming that the number of users is not too small, $n(t)$ can be approximated as additive white Gaussian noise (AWGN) with zero mean and two sided power spectrum $N_0/2$.

Assuming that there are F fingers in the rake receiver, the output of the rake receiver with maximal ratio combining at time $t = kT$ can be represented as

$$\hat{y}[k] = \sum_{l=0}^{F-1} y_l[k] \quad (5)$$

where $y_l[k]$ is the output of the l-th finger given by

$$y_l[k] = \mathrm{Re}\left\{ \int_{kT}^{(k+1)T} r(t + d_l T_c) c_d^*(t) \hat{\alpha}_l(t) e^{-j\hat{\phi}_l(t)} dt \right\}. \quad (6)$$

Here, $\hat{\alpha}_l(t)$ and $\hat{\phi}_l(t)$ are respectively the estimated amplitude and phase of the channel obtained by the channel estimator, the superscript * denotes the complex conjugate and $c_d(t)$ is the data spreading signal represented as

$$c_d(t) = \sum_{n=-\infty}^{\infty} q[n] w_d[n] p_{T_c}(t - nT_c). \quad (7)$$

Assuming that the channel characteristics are unchanged during a symbol time interval, $y_l[k]$ can be rewritten as

$$y_l[k] = \mathrm{Re}\left\{ \int_{kT}^{(k+1)T} r(t + d_l T_c) c_d^*(t) \hat{\alpha}_l[k] e^{-j\hat{\phi}_l[k]} dt \right\} \quad (8)$$

where $\hat{\alpha}_l[k]$ and $\hat{\phi}_l[k]$ are the value of $\hat{\alpha}_l(t)$ and $\hat{\phi}_l(t)$ at time $t = kT$, respectively.

The pilot symbol representing the impulse response of the channel can be obtained by

$$\begin{aligned} \tilde{h}_l[k] &= \frac{-j}{\beta T} \int_{kT}^{(k+1)T} r(t + d_l T_c) c_p^*(t) dt \\ &= \frac{1}{\beta T} \int_{kT}^{(k+1)T} \left(\beta g[k] \alpha_l[k] e^{j\phi_l[k]} - jn(t + d_l T_c) \right) dt \\ &= g[k] \alpha_l[k] e^{j\phi_l[k]} + \frac{1}{\beta} n_l[k] \end{aligned} \quad (9)$$

where $c_p(t)$ is the pilot spreading signal expressed by

$$c_p(t) = \sum_{n=-\infty}^{\infty} q[n] w_p[n] p_{T_c}(t - nT_c) \quad (10)$$

and $n_l[k]$ is AWGN with zero mean and variance $\sigma_{n,l}^2$ equal to N_0/T. Here, we assume that all the fingers suffer from same interference power on the average.

3. Optimum MA FIR CEF

We first consider the design of MA FIR CEF assuming that the large-scale fading is perfectly compensated by slow power control and the small-scale fading effect is only remained in the received signal, i.e., $g[k]$ is constant (=1). When a $(2M_l + 1)$-tap MA filter is employed as the CEF of the l-th finger, the CIR of the l-th path can be estimated by

$$\begin{aligned} \hat{h}_l[k] &= \frac{1}{2M_l + 1} \sum_{i=-M_l}^{M_l} \tilde{h}_l[k + i] \\ &= \frac{1}{2M_l + 1} \sum_{i=-M_l}^{M_l} \left(h_l[k + i] + \frac{1}{\beta} n_l[k + i] \right) \end{aligned} \quad (11)$$

where M_l corresponds to the group delay of the CEF.

The optimum MA FIR CEF is designed so as to minimize the MSE of the estimated CIR. Note that the optimum tap size minimizing the MSE of the estimated CIR also minimizes the BER performance [9]. Define $\varepsilon_l[M_l]$ by the MSE of the l-th path channel estimator

$$\varepsilon_l[M_l] \equiv E\left\{ \left| \hat{h}_l[k] - h_l[k] \right|^2 \right\} \quad (12)$$

where $E\{x\}$ denotes the ensemble average of x. It can easily be shown that the MSE is represented as

$$\begin{aligned} \varepsilon_l[M_l] &= R_{h,l}[0] + \frac{1}{(2M_l + 1)^2} \sum_{i=-M_l}^{M_l} \sum_{j=-M_l}^{M_l} R_{h,l}[i - j] \\ &\quad - \frac{2}{2M_l + 1} \sum_{i=-M_l}^{M_l} R_{h,l}[i] + \frac{1}{2M_l + 1} \frac{\sigma_{n,l}^2}{\beta^2} \end{aligned} \quad (13)$$

where $R_{h,l}[i]$ is the real part of autocorrelation of the l-th sampled impulse response defined as

$$R_{h,l}[i] = E\left\{ \mathrm{Re}\left(h_l^*[n] h_l[n + i] \right) \right\}. \quad (14)$$

By changing the summation form in (13) into an integral form, the MSE can be approximated by

$$\begin{aligned} \varepsilon_l(\kappa_l) &\approx R_{h,l}(0) + \frac{1}{(2\kappa_l)^2} \int_{-\kappa_l}^{\kappa_l} \int_{-\kappa_l}^{\kappa_l} R_{h,l}(t - u) dt du \\ &\quad - \frac{1}{\kappa_l} \int_{-\kappa_l}^{\kappa_l} R_{h,l}(t) dt + \frac{T}{2\kappa_l} \frac{\sigma_{n,l}^2}{\beta^2} \\ &= R_{h,l}(0) + \frac{1}{2\kappa_l^2} \int_0^{2\kappa_l} R_{h,l}(t)(2\kappa_l - t) dt \\ &\quad - \frac{2}{\kappa_l} \int_0^{\kappa_l} R_{h,l}(t) dt + \frac{T}{2\kappa_l} \frac{\sigma_{n,l}^2}{\beta^2} \end{aligned} \quad (15)$$

where κ_l is equal to $(M_l + 1/2)T$ and $R_{h,l}(\tau)$ is the autocorrelation of the impulse response of the l-th path channel given by

$$\begin{aligned} R_{h,l}(\tau) &= E\left\{ \mathrm{Re}\left(h_l^*(t) h_l(t + \tau) \right) \right\} \\ &= P_l \cos(2\pi f_d \tau \cos\theta_l) + \sigma_{\alpha,l}^2 R_{\alpha,l}(\tau). \end{aligned} \quad (16)$$

Here, f_d denotes the maximum Doppler frequency and $R_{\alpha,l}(\tau)$ is the autocorrelaton of scattered components in the l-th path signal given by [8]

$$R_{\alpha,l}(\tau) = \begin{cases} J_0(2\pi f_d \tau), & \textit{Classic spectrum} \\ \frac{\sin(2\pi f_d \tau)}{2\pi f_d \tau}, & \textit{Flat spectrum} \end{cases} \quad (17)$$

where $J_0(\cdot)$ is the first kind, zero-th order Bessel function.

The optimum κ_l minimizing $\varepsilon_l(\kappa_l)$ can be obtained from

$$\begin{aligned} \left.\frac{\partial \varepsilon_l(\kappa_l)}{\partial \kappa_l}\right|_{\kappa_l = \hat{\kappa}_l} &= -\frac{1}{\hat{\kappa}_l^2} \int_0^{2\hat{\kappa}_l} R_{h,l}(t) dt + \frac{1}{\hat{\kappa}_l^3} \int_0^{2\hat{\kappa}_l} t R_{h,l}(t) dt \\ &\quad + \frac{2}{\hat{\kappa}_l^2} \int_0^{\hat{\kappa}_l} R_{h,l}(t) dt - \frac{2}{\hat{\kappa}_l} R_{h,l}(\hat{\kappa}_l) - \frac{T}{2\hat{\kappa}_l^2} \frac{\sigma_{n,l}^2}{\beta^2} \\ &= 0. \end{aligned} \quad (18)$$

Substituting the following Taylor approximations for $0 \le t \le 2\pi/3$,

$$\cos(t) = \sum_{k=0}^{\infty} \frac{(-1)^k t^{2k}}{(2k)!} \approx 1 - \frac{t^2}{2!} + \frac{t^4}{4!}$$

$$J_0(t) = \sum_{k=0}^{\infty} \frac{(-1)^k t^{2k}}{2^{2k}(k!)^2} \approx 1 - \frac{t^2}{4} + \frac{t^4}{64}$$

$$\frac{\sin(t)}{t} = \sum_{k=0}^{\infty} \frac{(-1)^k t^{2k}}{(2k+1)!} \approx 1 - \frac{t^2}{3!} + \frac{t^4}{5!} \tag{19}$$

(16) can be approximated as

$$R_{h,l}(\tau) = \begin{cases} P_l\left(1 - \frac{(\omega_\tau \cos\theta_l)^2}{2} + \frac{(\omega_\tau \cos\theta_l)^4}{24}\right) \\ \quad + \sigma_{\alpha,l}^2\left(1 - \frac{\omega_\tau^2}{4} + \frac{\omega_\tau^4}{64}\right), \ Classic \\ P_l\left(1 - \frac{(\omega_\tau \cos\theta_l)^2}{2} + \frac{(\omega_\tau \cos\theta_l)^4}{24}\right) \\ \quad + \sigma_{\alpha,l}^2\left(1 - \frac{\omega_\tau^2}{6} + \frac{\omega_\tau^4}{120}\right), \ Flat \end{cases} \tag{20}$$

where $\omega_\tau = 2\pi f_d \tau$. Note that t satisfying (18) lies in $0 \le t \le 2\pi/3$. Substituting (20) into (18), $\hat{k}_l$ can be determined by

$$\hat{k}_l = (2\pi f_d)^{-4/5}\left(\frac{T}{2}\frac{\sigma_{n,l}^2}{\beta^2}\right)^{1/5}\left(\frac{P_l\cos^4\theta_l}{9} + \frac{\sigma_{\alpha,l}^2}{\chi_l}\right)^{-1/5} \tag{21}$$

where χ_l is a constant equal to 24 and 45 in the case of classic and flat spectrum, respectively.

Since the energy of the received signal per bit can be represented as

$$\begin{aligned} E_b &= \frac{T}{C}\sum_{l=0}^{L-1}(P_l + \sigma_{\alpha,l}^2) \\ &= \frac{N_0}{C\sigma_{n,l}^2}\sum_{l=0}^{L-1}(P_l + \sigma_{\alpha,l}^2) \end{aligned} \tag{22}$$

$\hat{k}_l$ can be rewritten by

$$\begin{aligned} \hat{k}_l &= (2\pi f_d)^{-4/5}\beta^{-2/5}\left(\frac{T}{2}\right)^{1/5} \\ &\times \left[\frac{CE_b}{N_0\sum_{l=0}^{L-1}(P_l + \sigma_{\alpha,l}^2)}\left(\frac{P_l\cos^4\theta_l}{9} + \frac{\sigma_{\alpha,l}^2}{\chi_l}\right)\right]^{-1/5} \end{aligned} \tag{23}$$

Letting S_l be the average power ratio of the l-th path signal to the total signal,

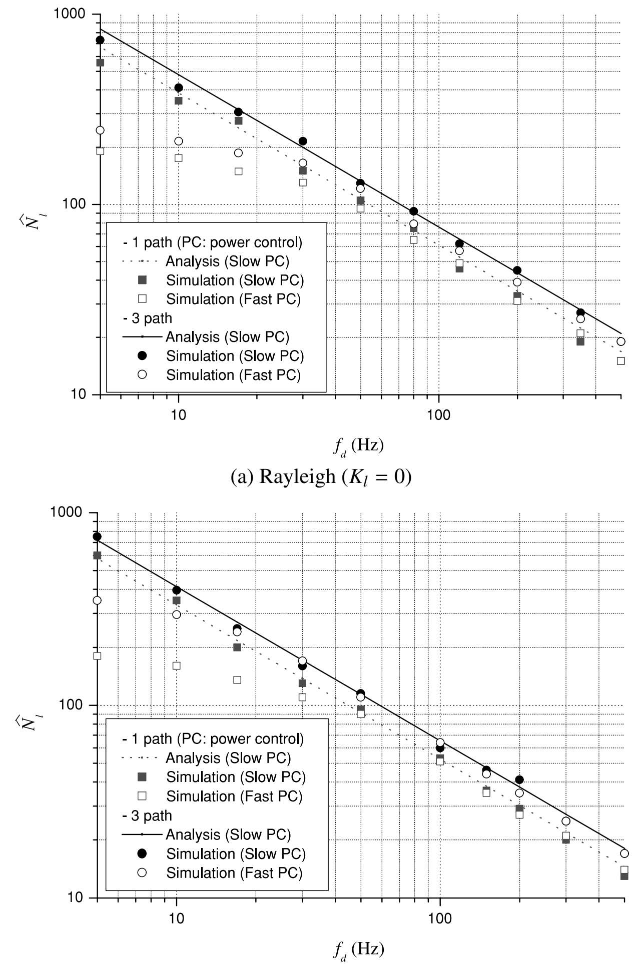
$$S_l = \left(P_l + \sigma_{\alpha,l}^2\right)\Big/\sum_{i=0}^{L-1}(P_i + \sigma_{\alpha,i}^2) \tag{24}$$

the optimum tap size $\hat{N}_l$ ($=2\hat{M}_l+1$) of the MA FIR CEF for the l-th path can approximately be determined by

$$\begin{aligned} \hat{N}_l &= \lfloor 2\hat{k}_l/T \rfloor \\ &= \left\lfloor\left(\frac{(\pi f_d T)^{-4}\beta^{-2}}{(E_b/N_o)CS_l}\frac{1+K_l}{K_l\cos^4\theta_l/9 + 1/\chi_l}\right)^{1/5}\right\rfloor \end{aligned} \tag{25}$$

where $\lfloor x \rfloor$ denotes an integer less than or equal to x.

It can be seen from (25) that $\hat{N}_l$ decreases as f_d, E_b/N_0, β^2 and/or S_l increase and is most influenced by f_d. The difference between the optimum tap sizes corresponding to the classic and flat spectrum becomes maximum when K_l is equal to zero. Since the maximum ratio of $\hat{N}_l$ of the two spectra is $(45/24)^{1/5} \approx 1.13$, the shape of the channel spectrum has small influence on the optimum tap size of the MA FIR CEF. To verify the analytic design, Fig. 1 compares the analytic design with the simulation result in Rayleigh ($K_l = 0$) and Ricean ($K_l = 2$, $\theta_l = 0°$) fading channels with a classic spectrum, where $T = 1/19200$ sec, $\beta^2 = 1/4$, $C = 1/2$, $\chi_l = 24$ and $E_b/N_0 = 5$ dB. Here, we assume that the all the multipaths have the same gain on the average. It can be seen that the analytic design is quite accurate.



(a) Rayleigh ($K_l = 0$)

(b) Ricean ($K_l = 2$, $\theta_l = 0°$)

Fig. 1 Optimum tap of the MA FIR CEF.

To further reduce the implementation complexity, the analytic results obtained with the use of an MA FIR CEF can be applied to the use of an one-pole IIR CEF. The frequency response of an MA FIR filter can be approximated by that of an one-pole IIR filter if the forgetting factor μ_l of the one-pole IIR filter is determined as [15]

$$\hat{h}_l[k] = \mu_l\hat{h}_l[k-1] + (1-\mu_l)\tilde{h}_l[k] \tag{26}$$

where $\mu_l = (\hat{N}_l - 3)\big/(\hat{N}_l - 1)$. Here, μ_l is determined to have the same group delay at zero frequency as the corresponding MA FIR filter.

In practice, the power of the transmit signal is fast controlled to compensate the fast fading as well as the near-far effect [3]. However, most of studies on the channel estimation have assumed only the use of slow power control and few analytical results have been reported with the use of fast power control. Since fast power control includes time-variant nonlinear feedback process, it is not easy to analytically describe the effect of fast power control on the received signal. It would be useful for the design of a channel estimator if the effect of fast power control can be approximated by a simple model.

To see the effect of fast power control, Fig. 2 illustrates the power spectral density of the received pilot signal with the use of slow and fast power control. We assume that the fast power control is performed with a step size of 1 dB at a rate of 800 Hz. We also assume that the SIR is perfectly estimated and that the power control bit is transmitted with no error. Since the fast power control adjusts real-valued gain $g[k]$, it affects only the power of the received signal, but not the phase of the received signal. Thus, as illustrated in Fig. 2, it can be seen that the power spectral density is little changed by fast power control. The use of fast power control makes the transmit signal have fluctuation in magnitude at a rate higher than f_d, resulting in increase of high frequency term in the power spectral density. From Figs. 2(a) and (b), it can be seen that the increase of high frequency components in the power spectral density decreases as f_d increases since the difference between f_d and the power control rate decreases. Thus, it can be inferred that the use of fast power control little affects the shape of the Doppler spectrum of the channel except for low f_d. This implies that the gain can be set to a constant, say $g[k]$=1, for design of the CEF except for low f_d.

Figure 1 also depicts the optimum tap size with fast power control obtained by simulation. It can be seen that the optimum tap size is almost the same irrespective of the use of power control except for low f_d. When f_d is low, the designed tap size can be larger than the actual one due to the effect of fast power control as described before. In this case, it may be practical to fix the tap size of the MA FIR CEF when f_d is less than a threshold, say 30 Hz in Fig. 1. Although the threshold frequency may need to be optimized according to the channel condition, the use of a fixed threshold little affects the performance since the SIR of the estimated CIR is already much higher than that of the data signal using an MA FIR CEF with a large tap size.

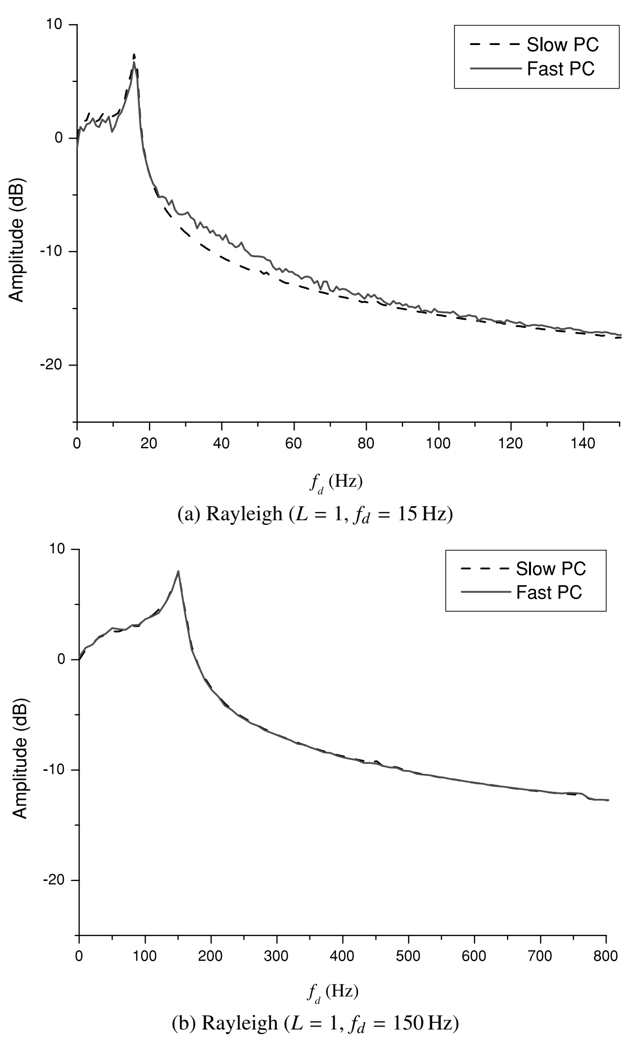


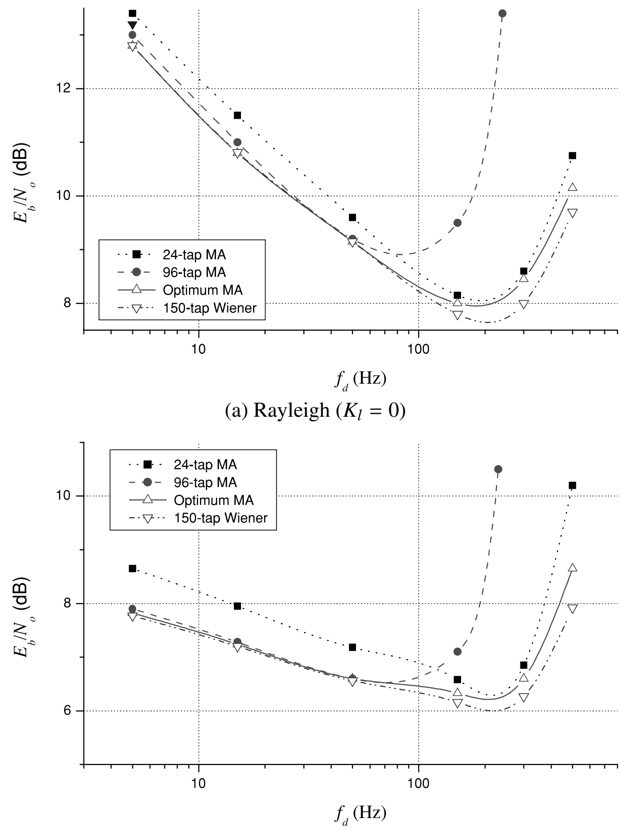
Fig. 2 Power spectral density of the received signal.

4. Performance Evaluation

To investigate the effect of channel estimation on the receiver performance, the required E_b/N_0 is evaluated by computer simulation to provide 10^{-3} BER in fading channels with a classic spectrum when the transmit signal power is fast controlled at a rate of 800 Hz. The pilot to data signal power ratio β^2 is set to 1/4 (i.e., −6 dB) by considering the trade-off between the pilot redundancy and accuracy of channel estimation. The simulation condition is summarized in Table 1. For comparison, we consider the use of 24-tap or 96-tap MA FIR filter as the fixed CEF, corresponding to an average time of 1.25 msec or 5 msec, respectively. Note that the use of 24-tap MA FIR filter is effective for f_d of nearly up to 500 Hz, corresponding to the case of high-speed train at a carrier frequency of 2 GHz and that the use of 96-tap MA FIR CEF is considered as a typical one for low f_d. We also consider the use of Wiener FIR CEF with 150-tap [9]. Numerical results show that the performance improvement with the use of Wiener CEF with a tap size

Table 1 Simulation parameters.

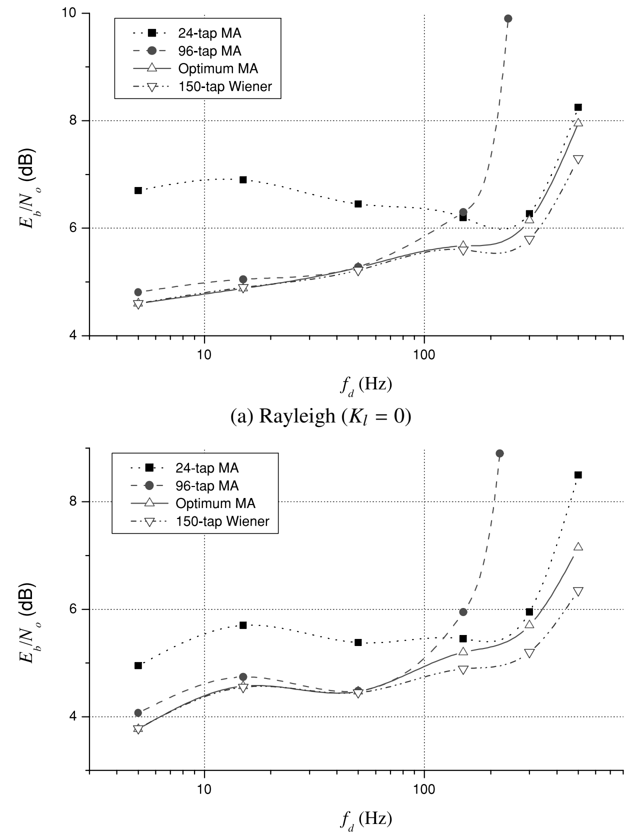
Parameters	Values
Data rate	9.6 kbps
Chip rate	1.2288 Mcps
Spreading factor	64
Spreading sequence	m-sequence with a period of $2^{15} - 1$
β^2	0.25
Channel coding	Convolutional (C=1/2 and H=9)
Interleaver	24 × 16 (20 ms)
Channel	Rayleigh and Ricean (Classic)
Multipaths	L=1 and 3 (equal average gain)
f_d	5–500 Hz
Power control	Step size=1 dB, control rate=800 Hz
Number of fingers	≤ 3



(a) Rayleigh (K_l = 0)

(b) Ricean (K_l = 2, θ_l = 0°)

Fig. 3 Required E_b/N_0 in a single-path channel.



(a) Rayleigh (K_l = 0)

(b) Ricean (K_l = 2, θ_l = 0°)

Fig. 4 Required E_b/N_0 in a three-path channel.

larger than 150 is negligible.

Figures 3 and 4 depict the required transmit E_b/N_0 to provide 10^{-3} BER under single-path and three-path channel condition, respectively, when the optimum MA and Wiener FIR filters are employed as the CEF. The tap size of the optimum MA FIR CEF is determined by (25). For comparison, we assume that the maximum Doppler frequency is perfectly estimated. The proposed scheme requires the information on the maximum Doppler frequency. Although the estimation of the maximum Doppler frequency is beyond the scope of this paper, it can be achieved using conventional schemes [16]–[19].

In slow fading channel, deep fading can result in burst errors due to the lack of time diversity, increasing the required E_b/N_0. This problem can occur in a single-path fading channel as in Fig. 3 (L=1). However, this problem becomes negligible as the number of multipaths increases as in Fig. 4 (L=3) due to the effect of multipath diversity. On the other hand, as f_d increases, the required E_b/N_0 increases since the accuracy of channel estimation becomes degraded. As a result, the effect of time diversity becomes less than the effect of inaccurate channel estimation as f_d increases. Thus, the required E_b/N_0 in a single-path channel has an 'U' shape as in Fig. 3, unlike a monotonically increasing shape in multi-path channel as in Fig. 4. It can be seen that the optimum MA FIR CEF provides receiver performance similar to or slightly poorer than the use of Wiener CEF except for very high maximum Doppler frequency. Thus, it can be suggested that the use of an MA FIR filter is a good practical choice as the CEF considering the implementation complexity and performance.

Figures 3 and 4 also depict the required E_b/N_0 when a fixed MA filter is used as the CEF. The use of 96-tap MA FIR CEF can provide BER performance similar to that of the optimum MA FIR CEF when f_d is low. As f_d increases, however, the averaging interval becomes larger than the optimum value. As a result, the BER performance with the use of 96-tap MA FIR CEF rapidly deteriorates when f_d is higher than 100 Hz. On the other hand, the performance with the use of 24-tap MA FIR CEF is optimum for around 250 Hz, but it is substantially degraded for f_d other than 250 Hz. This verifies that the tap size of the MA FIR filter significantly affects the receiver performance.

When there exist multipaths in the channel, the signal power will be split, reducing the signal power in each finger of the rake receiver. In this case, the use of an efficient CEF becomes much important issue. Thus, the E_b/N_0 gain with the use of the optimum MA FIR CEF increases as the number of multipaths increases.

5. Conclusions

In this paper, we have derived the optimum tap size of MA FIR CEF for pilot channel based DS-CDMA systems by minimizing the MSE of the channel estimate. The optimum tap size is a function of the channel condition parameters including the maximum Doppler frequency, SIR of the pilot symbol, Ricean factor. Numerical results show that the optimum MA FIR CEF can provide nearly optimum receiver performance in most of channel condition, i.e., quite similar to Wiener CEF. Thus, it can be a good practical choice to use an MA FIR filter as the CEF considering the implementation complexity and performance.

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