Adaptive Channel Estimation in DS-CDMA Systems

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ABSTRACT

The accuracy of channel estimation (CE) significantly affects the performance of the coherent rake receiver in DS-CDMA systems. It is desirable to employ a CE scheme that can provide good estimation performance and fast tracking the channel fading with reduced complexity. Such a CE scheme needs to employ a so-called channel estimation filter (CEF) that can effectively reduce the noise effect according to the channel condition. In this paper, we consider the use of adaptive channel estimators that employ a moving average (MA) FIR filter as the CEF. Numerical results show that the use of adaptive MA FIR CEF for pilot based and hybrid CE schemes can provide a performance similar to the use of Wiener CEF, while significantly reducing the implementation complexity.

Keywords: adaptive channel estimation, channel estimation filter, DS-CDMA system

1. INTRODUCTION

Due to the advantages of spectral efficiency [1], direct sequence code division multiple access (DS-CDMA) system has been widely employed as an air interface for the second and third generations mobile communication systems such as IS-95, IS-2000 and WCDMA systems [2-4]. There have been a great number of studies on the DS-CDMA system including enhancement of IMT-2000 systems, such as high speed downlink packet access (HSDPA), evolution data only (Ev-DO) and evolution data voice (Ev-DV) systems [5,6].

Enhanced DS-CDMA systems employ a coherent detection scheme to increase the receiver gain [7]. For example, coherent detection has been applied to the IMT-2000 system in both the downlink and uplink, while noncoherent detection was employed in the uplink IS-95 system [2,4]. Coherent detection of DS-CDMA signals requires the channel impulse response (CIR) including the amplitude and phase response. The CIR can be estimated using a channel estimation (CE) block employed in each finger of the rake receiver [8]. Since the CIR of each path can be regarded as an independent random process, it can be estimated in each finger without loss of estimation performance.

The CIR is widely estimated using three schemes: reference-based, blind, and hybrid schemes [8]. In the reference-based CE scheme, a predetermined symbol, called pilot signal, is used as the reference for CE, while no pilot symbol is transmitted in a blind scheme. Since there is no reference symbol in a blind scheme, the statistical property of the received signal is usually utilized for CE [11-16]. In a hybrid scheme, the data symbol is first demodulated based on the CIR estimate obtained using the pilot symbol. Then, improved CIR estimate is obtained by combining the CIRs from the pilot and decoded-data symbol since the data and pilot signal experience the same channel condition.

The CE scheme can be designed optimally for a known channel condition. However, the channel condition is location and time dependent. In general, fixed CE schemes are designed to be operable under the worst channel condition. As a result, CE schemes designed for a specific channel condition may suffer from performance degradation when the channel condition is mismatched. It is desirable to employ an adaptive CE scheme whose parameters can be adjusted according to the channel condition. Although complicated schemes have been proposed for adaptive channel estimation, the performance improvement is not significant since low-level modulation is used in the DS-CDMA systems. Thus, we consider the use of moving average (MA) filters as the channel estimation filter (CEF), which is simple to implement and to control the filter characteristics. The tap size of the MA CEF can be adjusted in real time based on the estimated channel condition obtained using pilot signal. We also consider a hybrid CE scheme for further performance enhancement.

Following Introduction, Section II describes the DS-CDMA system that uses pilot signal. Conventional CE schemes are described in Section III. In Section IV, we describe ACE schemes that employ

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an MA FIR filter as the CEF. The tap size of the MA FIR CEF is optimally determined by estimating the channel condition in the pilot based and hybrid schemes. The performance of proposed channel estimators is verified by evaluating the receiver performance in terms of the bit error rate (BER) in Section V. Finally, conclusions are summarized in Section VI.

2. DS-CDMA SYSTEM

We consider a DS-CDMA system that sends a pilot signal in a continuous or discontinuous transmission mode. Unlike other multiple access systems, DS-CDMA system can continuously transmit the pilot signal by allocating orthogonal codes for the pilot signal as well as the data. Depending upon the use of continuous and discontinuous pilot signal, we will refer to pilot-channel based CE (PCCE) and pilot-symbol based CE (PSCE), respectively.

In the transmitter of user-i, the user data is first encoded with code rate C and then interleaved. We will omit the user index i for ease of description. Assuming that the data and pilot signal are transmitted together in the PCCE system [2-4], the transmit signal $x(t)$ can be represented as

$$x(t) = \sum_{k=-\infty}^{\infty} \sum_{N=0}^{N-1} (g_d[k]x_d[k]w_d[kN+n]) \cdot q[kN+n]p_{T_c}(t-nT_c) \quad (1)$$

where β^2 is the pilot to data signal power ratio, N is the spreading factor (SF), T_c is the chip duration interval, $x_d[k]$ is the user message symbol at time $t = kT, T$ is the symbol duration time equal to $NT_c, x_p[k]$ is the pilot symbol having an amplitude equal to 1, $x_d[n]$ and $x_p[n]$ are respectively bipolar orthogonal spreading sequences allocated to the data and pilot signal, $g_d[k]$ and $g_p[k]$ are respectively the gain of the k-th data and pilot signal for fast power control (FPC), $P_{T_c}(t)$ is a unit-amplitude rectangular pulse defined in $[0, T_c)$, and $q[n]$ denotes user and cell specific complex PN scrambling sequence with $|q[n]| = 1$. $g_p[k]$ is usually set to 1 in the downlink system since a common pilot signal is transmitted, while user-dedicated pilot symbol is transmitted with $g_p[k] = g_d[k]$ in the uplink system [3,4]. In the PSCE scheme, $x(t)$ can be written by

$$x(t) = \sum_{k=-\infty}^{\infty} \sum_{n=0}^{N-1} g[k]x[k]q[kN+n]p_{T_c}(t-nT_c) \quad (2)$$

where $x[k]$ is equal to $x_d[k]$ and $\beta x_p[k]$ during the transmission of data and pilot signal, respectively.

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

$$h(t, \tau) = \sum_{l=0}^{L-1} h_l(t)\delta(\tau - \tau_l) \quad (3)$$

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

$$h_l(t) = \alpha_l(t)e^{j\phi_l(t)} \quad (4)$$

where $\alpha_l(t)$ and $\phi_l(t)$ are the amplitude and phase response of the l-th path at time t, respectively. We assume that the amplitude $\alpha_l(t)$ is Ricean distributed with Ricean factor K_l and $\phi_l(t)$ is uniformly distributed over $[0, 2\pi)$ [10]. 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 one when K_l is zero.

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

$$r(t) = \sum_{l=0}^{L-1} h_l(t)x(t - \tau_l) + n(t) \quad (5)$$

where $n(t)$ denotes the total interference term comprising the background noise, self interference from the desired user and other users interference from the desired and adjacent cells. 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 $I_0/2$.

The data symbol output of an L-finger rake receiver with maximal ratio combining (MRC) can be represented as

$$\hat{y}[k] = \sum_{l=0}^{L-1} r_l[k]\hat{h_l}[k] \quad (6)$$

where $\hat{h_l}[k]$ is the CIR estimate at time $t = kT$, the superscript*denotes the complex conjugate and $r_l[k]$ is the despread symbol of the l-th finger at time $t = kT$ given by

$$r_l[k] = \int_{kT}^{(k+l)T} r(t+\tau_l)(\sum_{n=-\infty}^{\infty} q[n]w_r[n] \cdot p_{T_c}(t-nT_c))^* dt. \quad (7)$$

Here, $w_r[n]$ is equal to $w_d[n]$ and 1 in the PCCE and PSCE, respectively.

The instantaneous CIR of the l-th path can be acquired from the pilot symbol at the l-th finger in the rake receiver by

$$\tilde{h}_l[k] = \frac{1}{\beta T}\int_{kT}^{(k+l)T} r(t+\tau_l)c_p^*(t)dt = g_p[k]h_l[k] + n_l[k]/\beta \tag{8}$$

where $h_l[k]$ is the CIR at time $t = kT$ and $c_p(t)$ is the pilot spreading signal expressed by

$$c_p(t) = \begin{cases} \sum_{n=-x}^{n} q[n]w_p[n]p_{T_c}(t-nT_c); & \text{PCCE} \\ \sum_{n=-x}^{x} q[n]p_{T_c}(t-nT_c); & \text{PSCE} \end{cases} \tag{9}$$

and $h_l[k]$ is zero mean AWGN with variance $\sigma_{\alpha,l}^2$ equal to $I_{0,l}/T(=\xi_l I_0/T)$, where $I_0 = L^{-1}\sum_{l=0}^{L-1} I_{0,l}$. Here, ξ_l varies according to the spreading sequence and channel condition such as the time-delay profile. In this paper, ξ_l is set to one assuming that each finger has the same interference power.

3. FIXED CHANNEL ESTIMATOR

When the pilot symbol is continuously transmitted through a pilot channel, the CIR can be estimated continuously. In the LS scheme, the CIR at time $t = kT$ is estimated by simply dividing the despreaded pilot symbol $r_l[k]$ by the transmitted pilot symbol $x_p[k]$ [17]. Although it is simple to implement, the performance can degrade significantly unless the signal to interference power ratio (SIR) is very high. To mitigate the performance degradation, the CIR can be obtained by filtering the adjacent pilot symbols. Although nonlinear filtering schemes such as neural networking can be employed for CE [19,20], we consider only linear schemes in this paper.

When the channel is stationary, Wiener filter is the optimum linear filter that minimizes the mean square error (MSE) of the estimated CIR [18]. Although both FIR and IIR Wiener filters can be employed, FIR Wiener filter is usually considered for the sake of stability [18]. Denoting $\tilde{\boldsymbol{h}}[k] = \left\{\tilde{h}[k-N]\cdots\tilde{h}[k]\cdots\tilde{h}[k+M]\right\}^T$, the CIR can be estimated using an $(N+M+1)$-tap Wiener filter as

$$\hat{h}[k] = \boldsymbol{w}\tilde{\boldsymbol{h}}[k] \tag{10}$$

where $\boldsymbol{w} = [w_{-N}\cdots w_0\cdots w_M] = \boldsymbol{p}^T\boldsymbol{R}^{-1}$ and

$$\begin{aligned} \boldsymbol{R}(i,j) &= E\left\{\tilde{h}[k+i]\tilde{h}^*[k+j]\right\} \\ \boldsymbol{p}(i) &= E\left\{h[k]\tilde{h}^*[k+i]\right\} \end{aligned} \tag{11}$$

for $i, j \in [-N, M]$.

It may not be practical to directly employ Wiener filter since it requires large implementation complexity and additional information on the channel and power of the desired and interference signals. Complexity-reduced FIR and IIR CEFs can be employed by placing some constraints on the filter coefficients. For example, the filter with real-valued and symmetric coefficients can reduce the complexity without significant performance degradation [21]. The use of MA FIR or one-pole IIR filters can also be considered as the CEF for ease of design and reduced implementation complexity [22-24].

When the channel is non-stationary, it can be described using a state space model as [18]

$$\begin{aligned} \boldsymbol{s}[k+1] &= \boldsymbol{F}\boldsymbol{s}[k] + \boldsymbol{G}\boldsymbol{u}[k] \\ r[k] &= \boldsymbol{X}[k]\boldsymbol{s}[k] + n[k] \end{aligned} \tag{12}$$

where $\boldsymbol{s}[k]$ denotes $N_s \times 1$ state vector representing CIR states at the k-th symbol time, $\boldsymbol{F}$ is $N_s \times N_s$ state transition matrix, $\boldsymbol{u}[k]$ is the zero mean AWGN with covariance matrix of I_{N_s}, $\boldsymbol{G}$ is the gain matrix for $\boldsymbol{u}[k]$, $\boldsymbol{X}[k]$ is the $1 \times N_s$ vector containing pilot symbol and $n[k]$ is the interference and background noise. As the matrix order N_s increases, more accurate channel model can be generated in general [25-27]. Under the model of (12), the optimum CEF can be obtained using a Kalman-type filter [18],

$$\begin{aligned} \boldsymbol{s}[k+1] &= \boldsymbol{F}\boldsymbol{s}[k] + \boldsymbol{K}[k]e^*[k] \\ \boldsymbol{P}[k+1] &= \boldsymbol{F}(\boldsymbol{P}[k] - \frac{\boldsymbol{P}[k]\boldsymbol{X}^H[k]\boldsymbol{X}[k]\boldsymbol{P}[k]}{\boldsymbol{X}[k]\boldsymbol{P}[k]\boldsymbol{X}^H[k]+\sigma_z^2})\boldsymbol{F}^H \\ &\quad + \boldsymbol{G}^H\boldsymbol{G} \\ \boldsymbol{K}[k] &= \frac{\boldsymbol{F}\boldsymbol{P}[k]\boldsymbol{X}^H[k]}{\boldsymbol{X}[k]\boldsymbol{P}[k]\boldsymbol{X}^H[k]+\sigma_z^2} \end{aligned} \tag{13}$$

where $e[k] = r[k] - h[k]x_p[k]$. Since this type of filters requires knowledge on the model parameters, they may not easily be applicable in practice.

The use of the least mean square (LMS) or recursive least square (RLS) method can mitigate this problem [18]. The CIR can be estimated using the LMS method

$$\hat{h}[k+1] = \hat{h}[k] + \mu x_p[k]e^*[k] \tag{14}$$

where μ is the adaptation step size. The LMS scheme can be derived from the Kalman filter assuming a random walk model, i.e., $\boldsymbol{s}[k] = h[k]$. Although the LMS scheme can be implemented without model parameters, it can suffer from slow convergence and approximation error of $h[k]$ in calculating $e[k]$. To accelerate the convergence, the RLS scheme can be employed. The RLS algorithm minimizes the weighted squared MSE as

$$\begin{aligned} \hat{h}[k+1] &= \hat{h}[k] + (\frac{A[k]}{\lambda + x_p^*[k]A[k]x_p[k]})x_p[k]e^*[k] \\ A[k+1] &= \frac{1}{\lambda}(A[k] - \frac{A[k]x_p^*[k]x_p[k]A^H[k]}{\lambda + x_p^*[k]A[k]x_p[k]}). \end{aligned} \tag{15}$$

Although it can also be used without information on the model parameters, it requires a large number of computations. Because of the use of exponential weighting, (15) is referred to exponentially windowed RLS (EW-RLS). Alternatively, a rectangular window RLS (RW-RLS) approach can be used [28], where the window is optimized based on the statistical knowledge of the fading channel and SIR.

When the pilot symbols are discontinuously transmitted, it is required to obtain the CIR estimate during the data transmission from the CIR estimate obtained using the pilot symbols. When the channel condition slowly changes, the estimated CIR using the pilot symbols can also be used for demodulation of the data symbols [31]. However, it may result in severe performance degradation as the maximum Doppler frequency of the channel increases. This problem can be alleviated by using an interpolator to track time-variant channel. Wiener interpolator is optimum in the sense of minimizing the MSE of the estimation error [32,33]. However, it may not be practical to employ Wiener interpolator due to the implementation complexity and need of additional information on the channel condition. A lowpass-type interpolation scheme can be employed for ease of implementation [34]. To reduce the computational complexity, other simple interpolation schemes have been considered including Gaussian, Nyquist, Lagrange and Spline interpolation methods [35-38]. Analytical and computer simulation results show that both the PCCE and PSCE yield the same performance if an efficient CE scheme is employed [39, 49].

Hybrid CE schemes can be applied to both the PCCE and PSCE schemes. In the PSCE scheme, the data symbol is demodulated using the estimated CIR obtained by interpolating the CIR estimated from the pilot symbols. If no decision error occurs, the data symbol can be used as the pilot symbol. This enables the use of PCCE scheme. In the PCCE scheme, the data symbol is demodulated using the CIR estimated from the pilot channel. Then, the CIR is additionally estimated using the data. When there is no decision error, the two CIR estimates can be combined optimally in an MRC manner. To reduce the decision error probability, the CIR can be estimated using the data channel decoding [41,42]. However, this requires a large amount of memory due to the delay of channel decoding. Thus, the data is often decoded based on the combined output $\hat{y}[k]$ of the rake receiver. It is desirable in practice to consider the effect of erroneous data decision to alleviate the performance loss.

4. ADAPTIVE CHANNEL ESTIMATOR WITH MA FIR CEF

Fixed CE schemes are designed assuming that the statistical parameters of the channel such as the correlation of the CIR, SIR and channel model parameters are known and time-invariant. In practice, these parameters can vary as the location of the mobile station and time change. Thus, it is desirable to employ adaptive channel estimator whose parameters are adjusted according to channel conditions.

The LMS, Kalman and RLS CE schemes can be applied to time-variant channel. However, the step size μ for LMS CE and the forgetting factor λ for the RLS CE schemes need to be adjusted according to the channel statistics and SIR [28,50]. The step size can be adjusted according to the average path power and CIR correlation [48]. Similarly, the parameters of Kalman filter are changed according to the channel statistics [26,52]. When these values are different from the optimum ones, the CE error increases significantly [51].

An $(N+M+1)$-tap Wiener filter can be realized as

$$\hat{y}[k] = \boldsymbol{w}[k]\tilde{\boldsymbol{h}}[k] \tag{16}$$

where $\boldsymbol{w}[k] = \{w_{-N}[k]\cdots w_0[k]\cdots w_M[k]\}^T$. Here, the tap coefficient vector is updated using the LMS tracking scheme as [53,54]

$$\boldsymbol{w}[k+1] = \boldsymbol{w}[k] + \mu\hat{\boldsymbol{h}}[k]e^*[k]. \tag{17}$$

Since accurate error signal $e[k](=h[k]-\hat{h}[k])$ cannot be obtained in practice, an approximate one can be obtained by $w_0[k]=0$ and $e[k]=\tilde{h}[k]-\hat{h}[k]$ [54]. This can result in performance degradation since the main tap is not included for filtering. Another approximation was considered by updating the main tap coefficient in a heuristic manner [53]. An RLS tracking scheme can also be employed for adaptation of the filter coefficients [53,54].

To reduce the implementation complexity and the risk of instability, the CEF can be selected from a filter bank according to the estimated channel conditions [24,29,30]. In the PSCE scheme, some ACE schemes have been proposed where the interpolator structure is changed according to the channel conditions [12,16]. Although various types of adaptive channel estimators (ACE) have been proposed, they may not be applicable to real situation because they consider a certain channel condition or they require high implementation complexity. We consider the use of MA FIR filters as the CEF since it is simple to implement and easy to control because only a single parameter (i.e., the number of tap size) is required for adaptation to the channel condition. It was shown that the MA FIR CEF can provide relatively good performance if the tap size is properly determined according to the channel condition [42]. Although we consider WCDMA downlink system where a common pilot symbol is continuously transmitted through a common pilot channel (CPICH) [3], it can also be applied other DS-CDMA system employing a pilot channel.

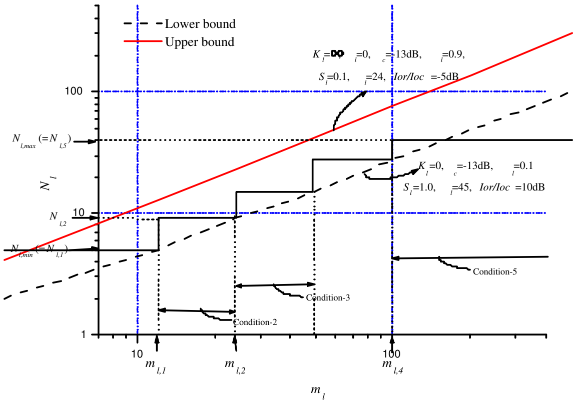


Fig.1: *The tap size of the MA FIR CEF*

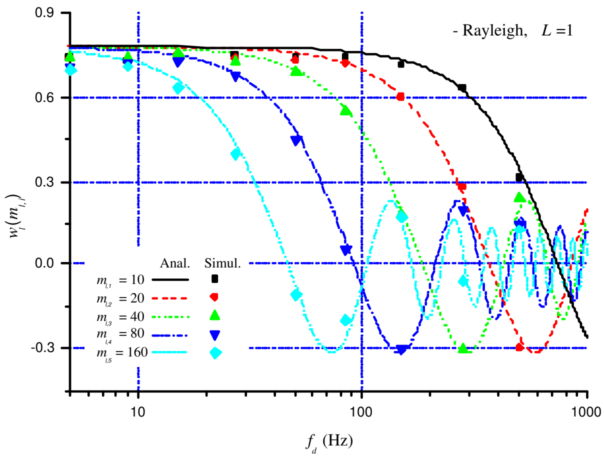


Fig.2: *The correlator output as a function of f_d*

Let $h_l^{CP}[k]$ and $h_l^{DP}[k]$ be the complex-valued channel gain multiplied by the transmit gain of the CPICH and dedicated physical data channel (DPDCH), respectively (i.e., $h_l^{CP}[k] = \beta h_l[k]$ and $h_l^{DP}[k] = h_l[k]$). Here, $h_l[k]$ is the normalized complex-valued channel gain of the l-th path at the k-th symbol time (i.e., $\sum_{l=0}^{L-1} |h_l[k]|^2 = 1$).

The despread pilot symbol is low-pass filtered to reduce the interference effect using an N_l-tap MA FIR CEF as

$$\hat{h}_l^{CP}[k] = \frac{1}{N_l} \sum_{i=-(N_l-1)/2}^{(N_l-1)/2} \hat{h}_l^{CP}[k+i] \tag{18}$$

where $(N_l-1)/2$-symbol delay is introduced by the CE block. The optimum tap size $\hat{N}_l$ of the MA FIR CEF can be determined by [43]

$$\hat{N}_l = \left(\frac{\upsilon_l + I_{oc}I_{or}}{(\pi f_d T)^4 \psi_p \gamma_c S_l} \cdot \frac{1+K_l}{K_l \cos^4\theta_l/9 + 1/\chi_l}\right)^{1/5} \tag{19}$$

where f_d is the maximum Doppler frequency, S_l is the ratio of the l-th path signal to total signal power, ψ_p is the spreading factor of the CPICH, γ_c denotes the CPICH E_c/I_{or}, E_c is the chip energy of the desired signal, χ_l is equal to 24 and 45 in the case of classic and flat spectrum, respectively [45], I_{or} is the total transmitted power from the desired base station, I_{oc} is the interference power from other cells and υ_l is the orthogonality factor. For example, υ_l is equal to zero in flat fading and it increases to one as the number of multipaths increases.

Since f_d is the most influencing factor determining the optimum filter tap size, it can be estimated in terms of the speed of CIR variation by exploiting the correlation property of the received pilot symbol [44]. We employ a correlator bank whose normalized correlation output is given by

$$w_l(m_l) \equiv \frac{\sum_{k=0}^{J-1} \mathrm{Re}\left\{\bar{h}^{CP}[k]\bar{h}_l^{CP^*}[k+m_l]\right\}}{\sum_{k=0}^{J-1} \left|\bar{h}^{CP}[k]\right|^2} \tag{20}$$

where $\bar{h}^{CP}[k]$ is the pre-filtered output of $\tilde{h}^{CP}[k]$ to suppress the noise term whose frequency is higher than the maximally allowable Doppler frequency. By finding $\hat{m}_l$ satisfying $w_l(\hat{m}_l) = \eta$, we can estimate the channel condition, where η is a given threshold level.

For a given $\hat{m}_l$, the optimum tap size of the MA FIR CEF can be determined considering the variation of other channel parameters as

$$\hat{N}_l = \left(\frac{16\hat{m}_l^4(\upsilon_l + I_{oc}I_{or})}{\psi_p \gamma_c S_l \hat{f}_l^4} \cdot \frac{1+K_l}{K_l \cos^4\theta_l/9 + 1/\chi_l}\right)^{1/5} \tag{21}$$

where $\hat{f}_l$ is the normalized maximum Doppler frequency equal to $2\pi f_d \hat{m}_l T$. Note that $\hat{f}_l$ is not a function of f_d but a function of the other channel condition parameters.

As an example, Fig.1 depicts an upper and lower bound of the tap size as a function of $\hat{m}_l$ in a typical channel condition, where the channel parameters are given as $0 \le K_l < \infty, -90° \le \theta_l \le 90°, \gamma_c = -13\text{dB}, -5\text{dB} \le I_{or}/I_{oc} \le 10\text{dB}, 0.1 \le S_l \le 1, 0.1 \le \upsilon_l \le 0.9$ and $24 \le \chi_l \le 45$. The tap size is between the upper and lower bound for $\hat{m}_l$. However, it may not be practical to employ too many correlators. For practical application, the channel condition can be approximated into a small number of cases using a staircase approximation as shown in Fig.1, where $m_{l,i}$ is $\hat{m}_l$ specified in the staircase. As shown in Fig.2, $w_l(m_{l,i})$ decreases faster than $w_l(m_{l,j})$ for $m_{l,i} > m_{l,j}$. Thus, the channel condition can be classified by comparing $w_l(m_l)$ with the threshold η. If the j-th correlator output of the l-th branch becomes less than η for the first time, i.e.,

$$w_l(m_{l,j}) < \eta \text{ and } w_l(m_{l,j-1}) \ge \eta,\ j = 1, 2, \ldots, G_l \tag{22}$$

we infer that the channel condition belongs to the j-th case. In this case, the tap size of the corresponding MA FIR filter is set to $N_{l,j}$, $1 \leq j \leq G_l$. Note that the tap size N_{l,G_l+1} corresponds to the case when no correlator output is less than the threshold, which can occur when the channel response varies very slowly.

The DPCH data signal can also be used for improved channel estimation, similar to the use of the CPICH. The only difference is that the decoded data symbol is used for channel estimation instead of known pilot symbol. The DPCH despread signal $r_l^{DP}[k]$ can be obtained by dispreading over ψ_d chips as

$$r_l^{DP}[k] = h_l^{DP}[k]x_d[k] + n_l^{'}[k] \tag{23}$$

where $n_l^{'}[k]$ represents the interference term. Denoting the decision data by $\hat{x}_d[k]$, the CIR can be estimated as

$$\hat{h}_l^{DP}[k] = \frac{1}{M_l} \sum_{j=-(M_l-1)/2}^{(M_l-1)/2} r_l^{DP}[k+j]\hat{x}_d^*[k+j] \tag{24}$$

where M_l is the tap size of the MA FIR CEF for the DPCH. Fig.3 depicts the structure of the proposed hybrid ACE [47]. In the proposed hybrid scheme, the channel is first estimated by the pilot-channel based ACE as (18). This CIR estimate is used by the rake symbol combiner to obtain tentative data decision. The CIR is re-estimated using this tentatively decoded data as (24). Finally, the channel estimate $\hat{h}_l[k]$ is obtained by combining these two channel estimates as

$$\hat{h}_l[k] = w_{cp}\hat{h}_l^{CP}[k] + w_{dp}\hat{h}_l^{DP}[k] \tag{25}$$

where w_{cp} and w_{dp} are the combining weights to be optimized. The tap size M_l of the MA FIR CEF of the DPCH can also be adjusted similarly as that of the CPICH. However, this procedure requires an additional processing block for estimation of the channel condition parameters. Since the DPCH and CPICH experience the same fading, the tap size M_l can be determined using a pre-determined tap size N_l. Thus, M_l can be obtained by making the two CEFs have the same averaging time interval, i.e.,

$$M_l = \frac{\psi_p}{\psi_d} N_l. \tag{26}$$

Since the decision errors are unavoidable in practice, they should be taken into account for combining $\hat{h}_l^{DP}[k]$ and $\hat{h}_l^{CP}[k]$. Since the data symbol $x_d[k]$ is QPSK modulated with unit amplitude [3], it can be shown that

$$\begin{aligned} E\left\{r_l^{DP}[k]\hat{x}_d^*[k]\right\} &= P_e(1-P_e)(h_l^{DP}[k]e^{j\frac{\pi}{2}}) \\ &\quad + P_e(1-P_e)(h_l^{DP}[k]e^{-j\frac{\pi}{2}}) \\ &\quad + P_e^2(-h_l^{DP}[k]) + (1-P_e)^2(h_l^{DP}[k]) \quad (27) \\ &= (1-2P_e)h_l^{DP}[k] \end{aligned}$$

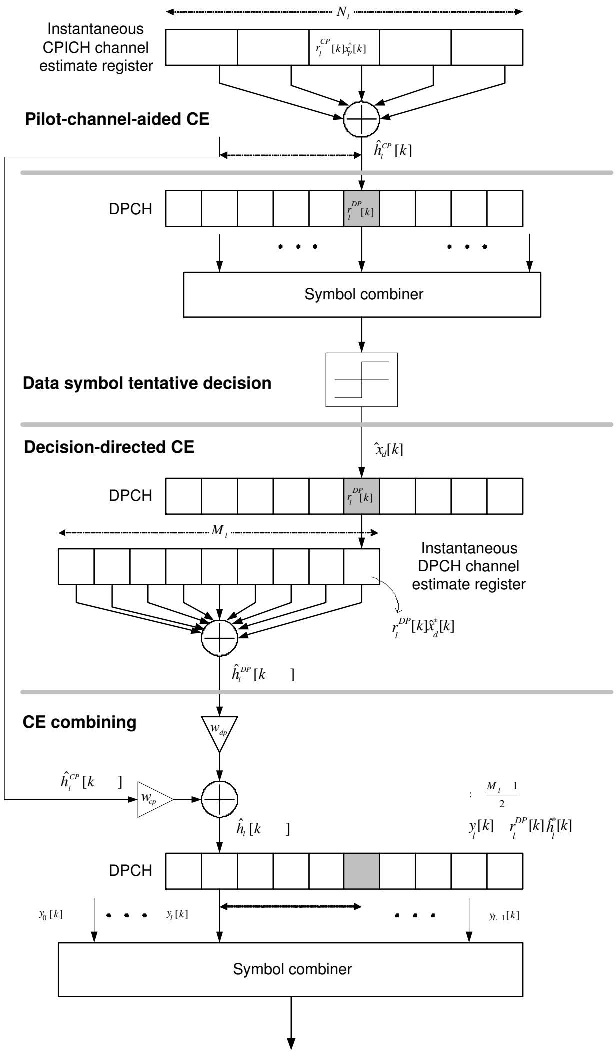


__Fig.3:__ Structure of the proposed hybrid channel estimator

where P_e is the BER due to tentative decision. Therefore, the term $r_l^{DP}[k]\hat{x}_d^*[k]$ can be rewritten as

$$r_l^{DP}[k]\hat{x}_d^*[k] = (1-2P_e)h_l^{DP}[k] + n_l^{'}[k]\hat{x}_d^*[k] + \epsilon[k] \tag{28}$$

where

$$\epsilon[k] = (x_d[k]\hat{x}_d^*[k] - 1 + 2P_e)h_l^{DP}[k] \tag{29}$$

is an additional interference term due to tentative decision error. It can be shown that $\epsilon[k]$ has zero mean and variance of $4P_e(1-P_e)\left|h_l^{DP}[k]\right|^2$ Since the variance σ_n^2 of $n_l^{'}[k]\hat{x}_d^*[k]$ is much larger than $4P_e(1-P_e)\left|h_l^{DP}[k]\right|^2$ in nominal channel condition, the SIR ρ_{dp} of $h_l^{DP}[k]$ can be approximated as

$$\rho_{dp} = \frac{(1-2P_e)^2\left|h_l^{DP}[k]\right|^2}{\sigma_n^2} M_l \tag{30}$$

Table 1: *Simulation condition*

	Parameter	Value
DPCH		144 Kbps (ψ_d=16)
CPICH	Ec/Ior	-13dB, -10dB
Ior/Ioc		-3dB, 9dB
Channel	Rayleigh (Classic spectrum)	3GPP model (Case-1)
CEF		MA FIR
Power control		Step size: 1dB, Control period: 1 slot (0.667ms)

while the SIR ρ_{cp} of $h_l^{CP}[k]$ is

$$\rho_{cp} = \frac{\left|h_l^{CP}[k]\right|^2}{\sigma_n^2} N_l = \frac{\left|h_l^{CP}[k]\right|^2}{\sigma_n^2} M_l \tag{31}$$

using the Cauchy-Schwartz inequality, the maximum SIR ρ_{hyb} of $\hat{h_l[k]}$ is given by

$$\rho_{hyb,max} = \left[1 + \frac{1-2P_e}{\beta}^2\right] \tag{32}$$

when

$$\hat{w_{dp}} = \frac{1}{\beta}(1-2P_e)\hat{w_{cp}} \tag{33}$$

Note that the second term in the bracket of (32) represents the SIR gain by the proposed hybrid scheme and it increases as P_e decreases. In particular, when P_e is fixed, the achievable SIR increases as the power ratio of the DPCH and CPICH increases. It can also be seen that, if there is no decision error, the ratio of optimum weights $\hat{w_{cp}}$ and $\hat{w_{dp}}$ is equal to the ratio of the CPICH and DPCH transmitted power. That is, the two-channel estimates are combined in an MRC manner. If is equal to 0.5, $\hat{w_{dp}}$ becomes zero and thus the decision-directed channel estimation should not be used. Note that (33) can be obtained in practice by approximating the ensemble average into the time average because $E\{\hat{h}_l^{DP}[k]\}/E\{\hat{h}_l^{CP}[k]\} = (1-2P_e)/\beta$. As shown in Fig. 3, the MA CEF requires symbol delays due to CE block. This problem can be alleviated using a CEF with zero group delay with little performance degradation [44].

5. PERFORMANCE EVALUATION

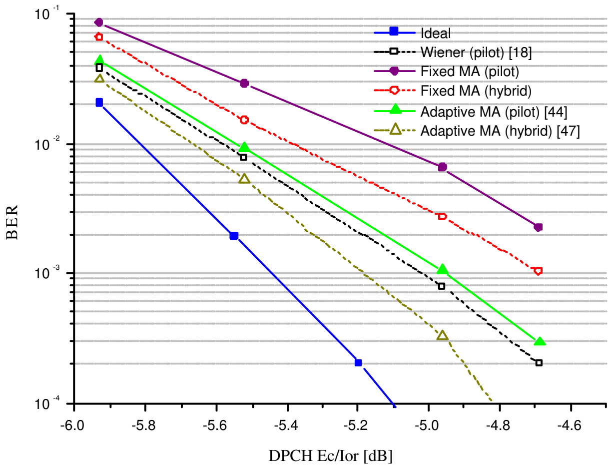
To investigate the performance of the CE scheme, the receiver performance is examined under two extreme channel conditions: one is when CPICH E_c/I_{or} =-13dB and I_{or}/I_{oc} =-3dB, and the other is when CPICH E_c/I_{or} =-10dB and I_{or}/I_{oc} =9dB. The former condition corresponds to a poor channel condition that the user resides near the cell boundary and the CPICH is transmitted with relatively low power. The latter one corresponds to a good channel condition that the user is near the base station and the CPICH is transmitted with relatively high power. Table 1 summarizes the simulation condition.

Fig. 4 depicts the BER performance with the use of the proposed ACE scheme and fixed CE scheme that uses an MA FIR CEF whose tap size is fixed to one slot time interval (= 0.667ms) [57]. The performance of the ideal channel estimator (i.e., known CIR) and 40-tap Wiener CEF is also presented for reference. It can be seen that the proposed ACE [44] can provide performance similar to Wiener CEF. This implies that the use of an MA FIR filter is a good practical choice as the CEF considering the implementation complexity and performance. Moreover, the performance of the hybrid ACE [47] is better than that of Wiener CEF due to the gain with the use of data signal. It can also be seen that adaptive schemes outperform fixed schemes in poor channel environment because the bandwidth adaptation of the CEF becomes more important. Fig. 5 depicts the required DPCH power to obtain a BER of 10^-3 in terms of the maximum Doppler frequency. It can be seen that the proposed adaptive schemes work well particularly for low maximum Doppler frequency. Note that the numerical results are obtained under the two-path Case-1 channel condition [45], but similar results are obtained even when the channel has more paths (e.g., the four-path Case-3 channel).

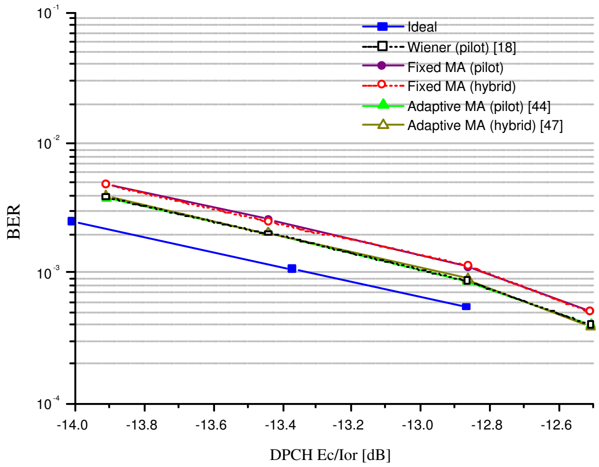
The proposed ACE scheme can also be applied to DS-CDMA systems employing multiple transmitted antennas where the pilot signal of each transmitted antenna needs to be separated from the compound signal using an appropriate orthogonal pilot pattern. In this case, the use of an efficient CE scheme becomes more important since CE error significantly affects on the receiver performance compared to the use of single transmit antenna [46]. Moreover, the estimated CIR can be more inaccurate due to the split pilot power between the transmitted antennas.

6. CONCLUSIONS

In this paper, we have reviewed various conventional CE schemes for the rake receiver of DS-CDMA systems. It is desirable to employ CE schemes that can provide good estimation performance and fast track the channel fading with a simple structure. To this end, we have proposed low-complexity adaptive channel estimators for DS-CDMA systems employing continuous pilot signal, where the tap size of MA FIR CEF is adjusted based on the estimated chan-

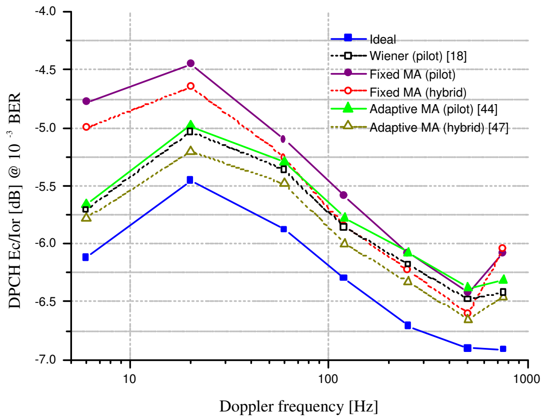


(a)Ior/Ioc = -3dB, CPICH Ec/Ior = -13dB

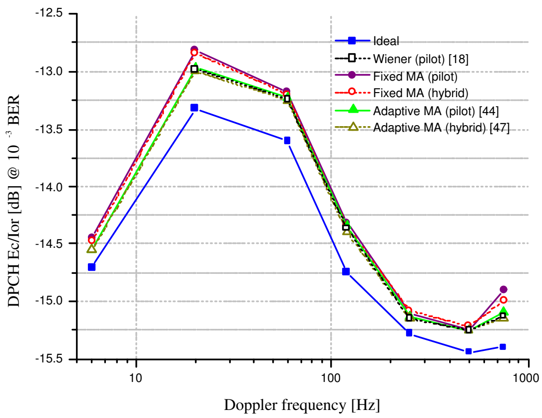


(b)Ior/Ioc = 9dB, CPICH Ec/Ior = -10dB

Fig.4: *BER performance when* f_d*=20Hz*



(a)Ior/Ioc = -3dB, CPICH Ec/Ior = -13dB



(b)Ior/Ioc = 9dB, CPICH Ec/Ior = -10dB

Fig.5: *Required Ec/Ior at* 10^{-3} *BER in the Case-1 channel*

nel condition parameters. The channel condition is estimated by exploiting the correlation property of the pilot signal. Numerical results show that the use of adaptive MA FIR CEF can provide the receiver performance similar to that of Wiener CEF.

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