An Adaptive Channel Estimator in Pilot Channel based DS-CDMA Systems

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Abstract - **The performance of a coherent rake receiver of DS-CDMA systems is significantly affected by the accuracy of channel estimation. The performance of the channel estimator can be improved by employing a channel estimation filter (CEF) whose impulse response is adjusted according to the channel condition. In this paper, we consider the analytic design of an adaptive channel estimator for DS-CDMA systems with a pilot channel. The proposed channel estimator estimates the channel condition including the maximum Doppler frequency, which can be performed without having exact a priori information on the operating condition. The estimated channel condition adjusts the impulse response of the CEF. Numerical results show that the use of the proposed channel estimator can provide BER performance quite robust to a wide range of channel condition.**

I. INTRODUCTION

Coherent rake receivers in the DS-CDMA system require the channel information including the amplitude and phase of the channel to coherently demodulate the received signal [1]. To obtain the channel information, a known signal called the pilot signal is usually transmitted being time-multiplexed or code-multiplexed with the data signal. In practice, the use of a code-multiplexed method is adopted in W-CDMA and Cdma2000 systems proposed for IMT-2000 systems [2,3].

To improve the accuracy of channel estimation, the obtained channel information is usually low-pass filtered to reduce the noise effect by employing a so-called channel estimation filter (CEF). It was shown that the characteristics of the CEF significantly affect the channel estimation performance and thus the receiver performance [4]. Since the channel condition is time-varying and location-dependent, the cut-off frequency of the CEF needs to be adjusted in real-time in response to the channel variation in order to improve the receiver performance.

There have been proposed a number of adaptive schemes applied to channel estimation including the least mean square (LMS) adaptation method, recursive least square (RLS) adaptation method [5]. These methods may not be applicable at low SIR since the reference signal used for calculation of the error metric is severely noise corrupted. Although a modified LMS considers the use of the received signal as the reference [6], the receiver performance can be deteriorated unless the noise power is correctly estimated. In addition, large implementation complexity may inhibit the use of RLS method in practice. Another approach considers the use of multiple CEFs to choose a CEF according to the estimated speed zone [7,8]. However, these methods may have limitation in real application since they assume only Rayleigh fading with a classic spectrum and known SIR of the received signal. If the channel condition does not agree well with the assumption, the maximum Doppler frequency may not be estimated properly. Moreover, even though the maximum Doppler frequency is perfectly estimated, these schemes may not provide optimum channel estimation performance because the CEF is heuristically determined without considering other channel condition parameters.

In this paper, we propose an adaptive channel estimator whose cut-off frequency of the CEF is adaptively controlled in real-time according to the channel condition. The channel condition is estimated by considering various channel condition parameters including the maximum Doppler frequency. It can be applied without exact a priori information on the operating condition such as the fading statistics, the SIR of the received signal and the pilot to data signal power ratio.

Section II describes the structure of the proposed adaptive channel estimator. The performance of the proposed scheme is evaluated in terms of the receiver performance in Section III. Finally, conclusions are summarized in Section IV.

II. PROPOSED CHANNEL ESTIMSTOR

The proposed channel estimator can be applied to both the downlink and uplink where the pilot signal is transmitted with the data signal in code-multiplexed method. In general, since the transmitted power of the dedicated pilot signal in the uplink is much less than that of the common pilot signal in the downlink, the use of an efficient channel estimator becomes much important in the uplink. Thus, we consider the receiver performance of the proposed scheme in the uplink DS-CDMA system.

The block diagram of the l-th branch in a DS-CDMA receiver is depicted in Fig. 1 where the proposed channel estimator scheme is shown in a dotted box. The proposed channel estimator consists of the channel estimation controller (CEC), the channel parameter estimator (CPE) and the channel estimation filter (CEF) module. The CPE estimates the parameters of the channel and transceiver using the received pilot symbols. The CEC determines the design parameters of the CPE and CEF module considering the operating channel characteristics. The output of the CPE is used to control the bandwidth of the CEF. For ease of analytic design and implementation, we consider the use of a simple moving average (MA) FIR as the CEF. The cut-off frequency of the CEF will be controlled by changing the tap size of the MA filter.

To estimate the channel parameters, we propose a new estimation structure that comprises a prefilter, bank of correlators and decision module. Fig. 2 depicts a block diagram of the proposed CPE. The despread pilot symbol $\tilde{h}_l[k]$ is first lowpass filtered to reduce the noise effect. The cut-off frequency of this prefilter needs to be minimized, but it should be larger than the allowable maximum Doppler frequency to be able to preserve the channel information. For simplicity of design, we consider the use of an N_a-tap MA filter as the prefilter. The filtered output is input to a bank of G_l correlators. The i-th correlator of the l-th branch, i=1,2,..., G_l, correlates the prefiltered pilot symbol with the $m_{l,i}$-sample delayed one for an interval of J symbols. Then, the output of the i-th correlator is normalized as

$$w_l(m_{l,i}) \equiv \sum_J \mathrm{Re}\{\bar{h}_l^*[n]\bar{h}_l[n-m_{l,i}]\} \Big/ \sum_J \left|\bar{h}_l[n]\right|^2 \qquad (1)$$

where $m_{l,1} < m_{l,2} < \cdots < m_{l,G_l}$.

Assume that the channel is Ricean faded. As a special case, the channel has Rayleigh fading when the Ricean factor is zero. Assuming that the prefiltered output $\bar{h}_l[n]$ is an ergodic process, $w_l(m_{l,i})$ can be approximated by [4]

$$w_l(m_{l,i}) \approx \frac{P_l \cos(2\pi f_d m_{l,i} T \cos\theta_l) + \sigma_l^2 R_{l,r}(m_{l,i}T)}{P_l + \sigma_l^2 + P_N} \qquad (2)$$

where T is the symbol time, P_l and θ_l are the power and the arrival angle of the direct path ray component of the l-th path signal, respectively, f_d denotes the maximum Doppler frequency. σ_l^2 is the average power of scattered ray component of the l-th path signal whose amplitude and phase are respectively Rayleigh and uniformly distributed, and $R_{l,r}(\tau)$ is the autocorrelaton of the scattered ray components in the received signal given by

$$R_{l,r}(\tau) = \begin{cases} J_0(2\pi f_d \tau), & \text{classic spectrum} \\ \dfrac{\sin(2\pi f_d \tau)}{2\pi f_d \tau}, & \text{flat spectrum.} \end{cases} \qquad (3)$$

Here, $J_0(\tau)$ is the first-kind zero-th order Bessel function. P_N is the prefiltered noise signal power given by

$$P_N = \sigma_n^2 / (\beta^2 N_a) = \sum_{l=0}^{L-1} (P_l + \sigma_l^2) \Big/ (\beta^2 N_a C \gamma_b) \qquad (4)$$

where β^2 is the pilot to data signal power ratio, σ_n^2 is the noise power, C is the code rate, and γ_b is the per-bit signal energy to noise power ratio equal to E_b/N_0. Approximating the noise and interference as an additive white Gaussian noise (AWGN) with zero mean and two-sided power spectrum $N_0/2$, we have $\sigma_n^2 = N_0/T$. Since $w_l(m)$ fast decreases as m increases, the channel environment can be classified by comparing $w_l(m)$ with a given threshold η. If the j-th correlator output of the l-th finger becomes

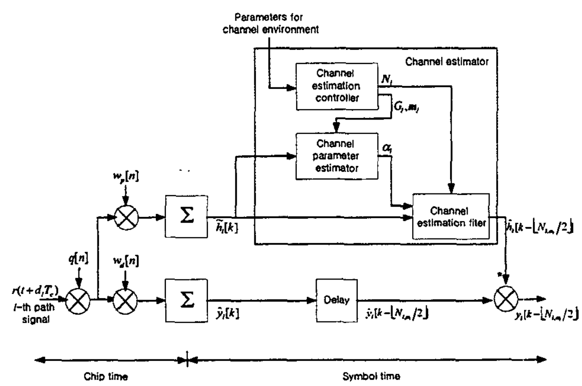


Fig. 1 The l-th finger of the rake receiver employing the proposed channel estimator

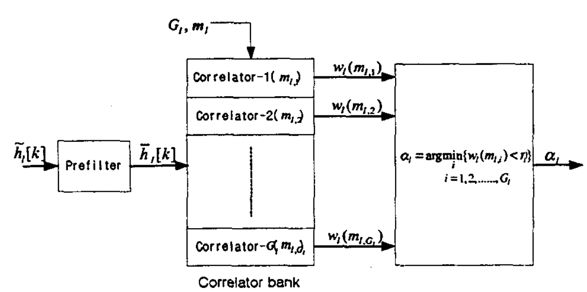


Fig. 2 The proposed channel parameter estimator

smaller than η for the first time, *i.e.*, for i = 1, 2, ... , j-1, $j=1,2,...,G_l$,

$$w_l(m_{l,j}) < \eta \text{ and } w_l(m_{l,i}) \geq \eta \tag{5}$$

the CPE infers that the channel environment belongs to channel condition-j. In this case, tap size of the corresponding MA filter is set to $N_{l,j}$, $1 \leq j \leq G_l + 1$, where $N_{l,1} < N_{l,2} < \cdots < N_{l,G_l+1}$. Note that N_{l,G_l+1} corresponds to the case when no correlators output is smaller than the threshold, which happens when the channel response too slowly varies. In this case, it suffices to set the filter tap size to a maximum value since the optimum tap size is not much changed due to the effect of fast power control [4].

To illustrate the operation of the proposed CPE, the values of $w_l(m_{l,i})$ obtained from analysis and simulation are plotted in Fig. 3 in a single path channel in terms of f_d when C = 1/2, γ_b = 5dB, T = 1/19200 sec, N_a = 9 and J = 1920. The simulation results are obtained by averaging 50 runs. It can be seen that the analysis agrees very well with the simulation results. For example, when the channel has Rayleigh fading with f_d = 100 Hz and η = 0.3, the output of the correlator-4 becomes less than the threshold for the first time as can be seen in Fig. 3. The channel condition can be effectively estimated with a size of J corresponding to 100msec interval. This enables the channel estimator to perform properly even if the channel environment varies fast.

Now, it is required to determine the number of correlators, G_l, the values of $m_{l,i}$ and $N_{l,i}$, $i=1,2,...,G_l$. It is performed in the CEC. It is reported that the optimum filter tap size $\hat{N}_l$ of the MA CEF can be approximately determined in a minimum mean squared error (MMSE) sense by [4]

$$\hat{N}_l = \left(\frac{2^4 m^4 A_l}{\beta^2 C \gamma_b \varsigma_l^2} \cdot \frac{1+K_l}{K_l \cos^4 \theta_l / 9 + 1/\chi_l} \right)^{1/5} \tag{6}$$

where A_l is the ratio of the total signal to the l-th path signal power, K_l is the Ricean factor equal to P_l/σ_l^2 and χ_l is equal to 24 and 45 in the case of the classic spectrum and the flat spectrum, respectively. Here, ς_l is equal to $(2\pi f_d mT)^2$, where m satisfies $w_l(m) = \eta$. Using (2) and forth-order polynomial approximations of the cosine, Bessel and sinc functions, ς_l can be written by

$$\varsigma_l = \left(b_l - \sqrt{b_l^2 - 4a_l c_l} \right) \Big/ 2a_l \,. \tag{7}$$

where

$$a_l = \begin{cases} P_l \cos^4 \theta_l / 24 + \sigma_l^2/64 \,; \text{ classic} \\ P_l \cos^4 \theta_l / 24 + \sigma_l^2/120 \,; \text{ flat} \end{cases} \quad b_l = \begin{cases} P_l \cos^2 \theta_l / 2 + \sigma_l^2/4 \,; \text{ classic} \\ P_l \cos^2 \theta_l / 2 + \sigma_l^2/6 ; \text{ flat} \end{cases}$$

$$c_l = (1-\eta)(P_l + \sigma_l^2) - \eta P_N \,.$$

Taking the logarithm on (6), we have

$$\log \hat{N}_l = 0.8 \log m + \xi_l \tag{8}$$

where ξ_l is related to the channel parameters represented as

$$\xi_l = \frac{1}{5}\left[\log(2^4 A_l) - \log(C\beta^2 \gamma_b \varsigma_l^2) + \log\left(\frac{1+K_l}{K_l \cos^4 \theta_l / 9 + 1/\chi_l} \right) \right]. \tag{9}$$

When the channel condition where the base station handles is given, the upper and the lower boundary of the tap size can be determined as in Fig. 4. That is, the lower boundary of the filter tap ize is obtained by substituting the parameters $A_{l,\min}$, $\chi_{l,\max}$, $K_{l,\min}$, $|\theta_l|_{\max}$, $\beta_{\max}$ and $\gamma_{b\max}$ into (6). Similarly, the upper

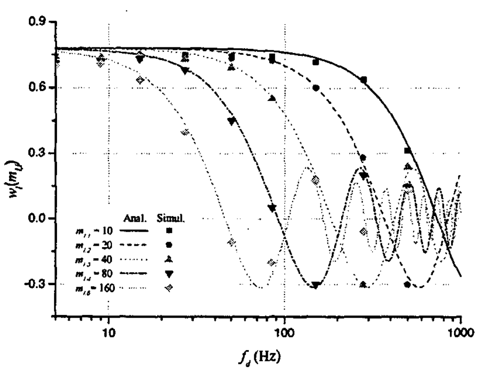


Fig. 3 The correlator output as a function of f_d

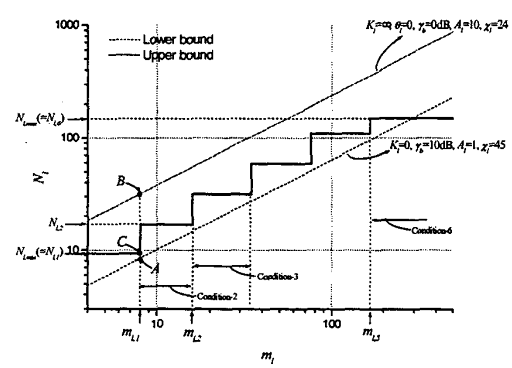


Fig. 4 An example of filter tap size selection

boundary of the filter tap size is obtained under $A_{l,\max}$, $\chi_{l,\min}$, $K_{l,\max}$, $|\theta_l|_{\min}$, $\beta_{\min}$ and $\gamma_{b\min}$. Here, the subscripts max and min respectively denote the maximum and minimum values of the parameter. Note that the range of the channel parameters can significantly vary depending on the characteristics of the service condition such as the geometry and channel environment. The smaller the difference between the lower and the upper boundary, the better the accuracy of the channel information can be obtained. To be able to apply the proposed scheme without the a priori information on the channel environment, the channel parameters can be set for a presumed operating condition. Given these boundaries, the magnitudes $\boldsymbol{m}_l$ of the delay parameters and the tap size N_l of the CEFs can be determined sequentially in a recursive manner as follows.

Let's introduce the lower boundary margin factor λ_L and upper boundary margin factor λ_U $(0 \le \lambda_L, \lambda_U < 1)$ to have a freedom in the design of the CPE. The lower bound margin λ_L can be determined such that the point A in Fig. 4 is represented as $(m_{l,1}, (1+\lambda_L)^{-1} N_{l,1})$. It can be shown that, for $1 \le i \le G_l$,

$$m_{l,i} = 0.5\left[\frac{N_{l,i}{}^5 C\gamma_{b\max}\varsigma_{l,\max}{}^2\beta_{\max}{}^2(K_{l,\min}\cos^4|\theta_l|_{\max}/9+1/\chi_{l,\max})}{(1+\lambda_L)^5 A_{l,\min}(1+K_{l,\min})}\right]^{1/4} \quad (10)$$

where

$$\begin{aligned}\varsigma_{l,\max} &= \frac{y_{\max}-\sqrt{y_{\max}{}^2-4x_{\max}z_{\max}}}{2x_{\max}} \\ x_{\max} &= K_{l,\min}\cos^4|\theta_l|_{\max}/24+1/120 \\ y_{\max} &= K_{l,\min}\cos^2|\theta_l|_{\max}/2+1/6 \\ z_{\max} &= (1-\eta)(1+K_{l,\min})-\frac{\eta A_{l,\min}(1+K_{l,\min})}{C\beta_{\max}{}^2 N_a\gamma_{b\max}}.\end{aligned} \quad (11)$$

As λ_L increases, the difference between the break points of the staircase (e.g., the point C in Fig. 4) and the lower boundary becomes large. With λ_U and $m_{l,1}$, the tap size of the second CEF in the l-th finger, $N_{l,2}$, can be determined such that $(m_{l,1}, (1+\lambda_U)N_{l,2})$ is on the upper boundary (e.g., the point B in Fig. 4). It can be shown that, for $1 \le i \le G_l$,

$$N_{l,i+1} = \frac{1}{1+\lambda_U}\left(\frac{164A_{l,\max}m_{l,i}{}^4(1+K_{l,\max})}{C\gamma_{b\min}\beta_{\min}{}^2\varsigma_{l,\min}{}^2(K_{l,\max}\cos^4|\theta_l|_{\min}/9+1/\chi_{l,\min})}\right)^{1/5} \quad (12)$$

where

$$\begin{aligned}\varsigma_{l,\min} &= \frac{y_{\min}-\sqrt{y_{\min}{}^2-4x_{\min}z_{\min}}}{2x_{\min}} \\ x_{\min} &= K_{l,\max}\cos^4|\theta_l|_{\min}/24+1/64 \\ y_{\min} &= K_{l,\max}\cos^2|\theta_l|_{\min}/2+1/4 \\ z_{\min} &= (1-\eta)(1+K_{l,\max})-\frac{\eta A_{l,\max}(1+K_{l,\max})}{C\beta_{\min}{}^2 N_a\gamma_{b\min}}.\end{aligned} \quad (13)$$

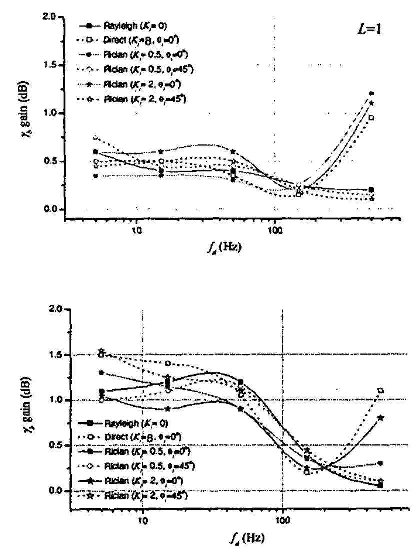
Similarly, $m_{l,2}$ can be calculated from $N_{l,2}$ and (10) with i=2. Thus, $N_{l,3}$ can be calculated from $m_{l,2}$ and (12) with i=2. In this way, $m_{l,i}$, i=1,2,..., G_l, and $N_{l,i}$, i=1,2,..., G_l+1, can be calculated iteratively until the tap size becomes larger than a predetermined maximum value $N_{l,\max}$. When $N_{l,i+1} \ge N_{l,\max}$, $N_{l,i+1}$ is set to a value of $N_{l,\max}$.

To illustrate a design example, Fig. 4 depicts the upper and the lower bound of $\hat{N}_l$ satisfying $w_l(m)=\eta$ as a function of m when $\lambda_L = 0.15$, $\lambda_U = 0.75$, 0 (*i.e.*, Rayleigh) $\le K_l < \infty$ (*i.e.*, direct path only), $-90° \le \theta_l \le 90°$, 0dB $\le \gamma_b \le$ 10dB, β^2 =0.25 (-6dB), $1 \le A_l \le 10$ and $24 \le \chi_l \le 45$. Here, the minimum and maximum values of the channel parameters are determined considering the normal operating condition. It can be seen that the difference between the maximum and minimum values of $\hat{N}_l$ are not large for a given m. For simplicity of implementation, we can approximate the channel condition into a small number of channel types using a staircase as illustrated in Fig. 4, where G_l = 5, $\boldsymbol{m}_l$ = {8, 16, 35, 76, 166} and $\boldsymbol{N}_l$ = {9, 17, 31, 57, 105, 151}.

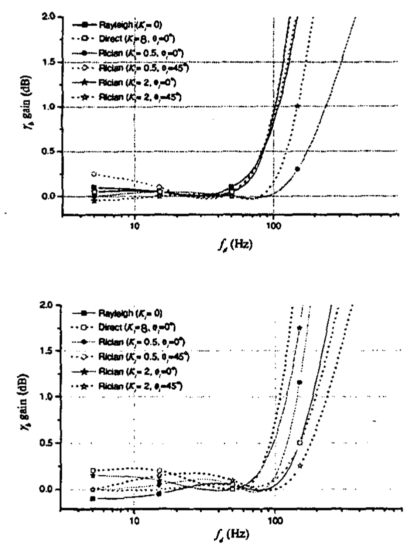
III. PERFORMANCE EVALUATION

To verify the performance improvement due to the use of the proposed channel estimator, the BER performance is evaluated when the power control is employed. The channel is assumed to have one or three paths with the same power gain in average. It is assumed that fast power control is performed with a step size of 1dB at a rate of 800Hz. We consider data transmission at a rate of 9600bps when β^2 = -6dB with the use of a convolutional coder with code rate 1/2 and constraint length 9, interleaving with 20msec (24x16) depth, two Walsh spreading sequences with a length of 64 for the data signal and the pilot signal, and m-sequences with a period of $(2^{15}-1)$. We also consider a maximum Doppler frequency of up to 500Hz.

Fig. 5 depicts the effective γ_b gain with the use of the proposed channel estimator over the use of a conventional one optimized for a specific channel condition. Compared to the use of a 24-tap MA FIR filter (*i.e.*, 1.25 msec averaging time), it can be seen that the use of the proposed scheme provides significant gain



(a) γ_b gain over the 1.25ms fixed MA CEF



(b) γ_b gain over the 5ms fixed MA CEF

Fig. 5 γ_b gain of the proposed channel estimator

except at a Doppler frequency of around 200Hz. This is due to that the use of a 24-tap MA filter is optimum for f_d of about 200Hz. Compared to the use of a 96-tap MA filter (*i.e.*, 5 msec averaging time), it can be seen that the use of the proposed scheme provides a γ_b gain of a fractional dB when f_d is low, but it can provide a large γ_b gain when f_d is higher than 80Hz. As the number of multipaths increases, the use of an efficient channel estimator becomes more important since the SIR of each path signal is decreased. Thus, the gain γ_b with the use of the proposed channel estimator increases as the number of multipaths increases.

IV. CONCLUSIONS

In this paper, we have proposed an adaptive channel estimator that employs a simple but efficient channel parameter estimator which classifies the channel condition using a bank of correlators. The impulse response of the channel estimation filter is adaptively adjusted according to the estimated channel condition. The proposed channel estimator can be applied without requiring exact information on the operating condition such as the fading statistics, the received SIR and the pilot to data signal power ratio. Numerical results show that the rake receiver with the use of the proposed channel estimator can provide relatively good BER performance over a wide range of the channel condition.

RERFERENCE

[1] F. Ling, "Coherent detection with reference symbol based channel estimation for direct sequence CDMA uplink communications," *Proc. VTC '93*, pp. 400-403, May 1993.

[2] 3GPP, *3G TS 25.211 – Physical channels and mapping of transport channels onto physical channels (FDD)*, June 2001.

[3] TIA, *The cdma2000 ITU-R RTT candidate submission*, June 1998.

[4] J.-W. Choi and Y.-H. Lee, "Design of channel estimation filter for coherent demodulation of DS-CDMA signals," *Proc. ISPACS'2000*, pp. 921-926, Nov. 2000.

[5] B. Widrow and A. D. Stearns, *Adaptive signal processing*, Prentice-Hall, 1985.

[6] S. K. Jo, J. H. Choi, Y. S. Woo, J. W. Soh and Y. H. Lee, "Modified leaky LMS algorithm for channel estimation in DS-CDMA systems," *in Proc. VTC'2001*, pp. 623-627, May 2001

[7] M. Sakamoto, J. Huoponen and I. Niva, "Adaptive channel estimator with velocity estimator for W-CDMA receiver," *Proc. VTC'2000*, pp. 2024-2028, May. 2000.

[8] H. J. Oh and J. M. Cioffi, "An adaptive channel estimation scheme for DS-CDMA systems," *Proc. VTC'2000*, pp. 2839-2843, Sept. 2000.