Channel Estimation for WLAN System Employing CCK Modulation in Multipath Fading Channels

Jin-Woong Cho · Cheol-Ho Kang

Abstract

This paper considers a channel estimation technique in a wireless local area network (WLAN) system with complementary code keying (CCK) signaling over multipath fading channels. This scheme uses the maximum correlator output of RAKE receiver for the extraction of the channel parameters. The performance of the proposed detection technique is compared with that of a noncoherent detection technique, and a significant improvement of performance is observed in terms of the bit error probability.

Ⅰ. INTRODUCTION

Recently, there is a demand for higher data-rate wireless personal area network (WPAN) such as Bluetooth, shared wireless access protocol (SWAP), and wireless local area network [1]~[3]. In this effort, both the Bluetooth SIG and IEEE 802.15 committee are working for the high-rate version, in which the maximum data rate will be upgraded up to 12 Mbps from the current 1Mbps, along with a reduction in interference with other wireless communication systems, such as IEEE 802.11b. On the other hand, Harris and Lucent have proposed complementary code keying (CCK) as an appropriate high rate waveform for use in the IEEE 802.11 standard for WLAN's [3]. These CCK codes perform well when used with a RAKE receiver in an indoor multipath environment. However, it can be efficiently demodulated with an accurate knowledge of the multipath channel. Recently, several approaches of the channel estimation have been investigated for the various transmission systems [4]~[6].

This paper is concerned with a simple method of channel estimation in WPAN systems with an emphasis on high rate WLANs. The proposed estimation method is based on the idea that channel parameters can be extracted from the maximum correlator output. The proposedestimation technique has a capability of estimating the channel distortion with little more additional hardware and a significant improvement of performance is obtained in comparison with a concoherent RAKE receiver. Furthermore, the estimation scheme can be applied to the next high-rate version of the Bluetooth system.

The outline of the paper is organized as follows. Section Ⅱ describes the transceiver and channel model. In Section Ⅲ and Ⅳ, the proposed channel estimator is presented in detail, and some numerical results are discussed, respectively. Finally, the concluding remarks are given in Section Ⅴ.

Ⅱ. TRANSCEIVER AND CHANNEL MODEL

IEEE 802.11 standard adopts the CCK modulation method for high-rate transmission. The IEEE 802.11 complementary spreading codes have a code length 8

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Table 1. Phase parameter encoding method.

Data bit	Phase parameter
(d_1, d_0)	ϕ_1
(d_3, d_2)	ϕ_2
(d_5, d_4)	ϕ_3
(d_7, d_6)	ϕ_4

Table 2. Differential QPSK modulation of phase parameters.

Data bit (d_{i+1}, d_i)	Phase
00	0
01	π
10	$\pi/2$
11	$-\pi/2$

and a clipping rate of 11Mcps. In the IEEE 802.11 standard, the 8-bit CCK code word, $C^m(t) = C_I^m(t) + jC_Q^m(t)$, are derived from the following formula:

$$C = \{e^{j(\phi_1+\phi_2+\phi_3+\phi_4)}, e^{j(\phi_1+\phi_3+\phi_4)}, e^{j(\phi_1+\phi_2+\phi_4)}, -e^{j(\phi_1+\phi_4)}, e^{j(\phi_1+\phi_2+\phi_3)}, e^{j(\phi_1+\phi_3)}, -e^{j(\phi_1+\phi_2)}, e^{j\phi_1}\} \quad (1)$$

where where $\phi_1 \in \left\{0, \frac{\pi}{2}, \pi, -\frac{\pi}{2}\right\}$ according to the transmitted information bits and the phase parameter ϕ_1 contained in all 8 chips of the code word rotates the whole vector. The parameters ϕ_1-ϕ_4 determins the phase values of the complex code set and are defined in the IEEE 802.11 high rate standard [3]. For the case of 11 Mbps mode, the data bit stream is partitioned into bytes as $(d_7, d_6, \cdots, d_0)$ where d_0 is the LSB and is first in time. The 8 bits are used to encode the phase parameters ϕ_1- ϕ_4 according to method described in Table 1. The encoding scheme is based on differential QPSK modulation as specified in Table 2. Therefore, the CCK code words are expressed in terms of 64 complex codes $\Psi^m(t) = \Psi_I^m(t) + j\Psi_Q^m(t)$ $(m = 1, 2, \cdots, 64)$ as follows

$$C_I^m(t) = \cos\phi_1 \Psi_I^m(t) - \sin\phi_1 \Psi_Q^m(t) \quad (2)$$

and

$$C_Q^m(t) = \sin\phi_1 \Psi_I^m(t) + \cos\phi_1 \Psi_Q^m(t) \quad (3)$$

Using eqns. (2) and (3), the transmitted signal can be expressed as

$$s(t) = Re[\sqrt{P_w}\{C_I^m(t) + jC_Q^m(t)\}e^{-jw_c t}], \quad 0 \le t \le T_w \quad (4)$$

where P_w is the signal power, w_c is the carrier angular frequency, and T_w is the symbol duration.

Recently, radio propagation measurements in the indoor channel at 2.4~GHz ISM band were reported by several researchers [7]~[9]. From these propagation measurements, we assume that the WLAN system operates in the indoor environment with a root mean square (RMS) delay spread τ_{rms} of 20 ~ 50ns, a maximum excess delay spread $\tau_{\max}$ of 30 ~ 200ns, and a maximum Doppler spread B_d of 10Hz based on the mobile speed of pedestrians. Using these parameters, the coherence bandwidth Δf_c and the coherence time Δt_c can be approximated as follows

$$\Delta f_c \ll \frac{1}{\tau_{rms}} = 50\,MHz$$
$$\text{and} \quad \Delta t_c \ll \frac{1}{B_d} = 100\,ms \quad (5)$$

From these assumptions, the channel of 11 Mbps WLAN systems operated at 2.4 GHz ISM band is modeled as a slowly varying frequency-selective fading as follows

$$h(t) = \sum_{n=1}^{N} \alpha_n e^{-j\theta_n} \delta(t - \tau_n) \quad (6)$$

where N is the number of multipath components. α_n, θ_n, and τ_n are the amplitude, phase, and delay of the n-th multipath, respectively.

Assuming the j-th complex code word $C^j(t)$ is transmitted, the complex baseband received signals can be expressed as

$$d(t) = \frac{1}{2}\sum_{n=1}^{N}\sqrt{P_w}\alpha_n C^j(t-\tau_n)e^{j\theta_n} + \frac{1}{2}n(t) \quad (7)$$

where $n(t) = n_c(t) + jn_s(t)$ is a zero mean Gaussian noise process with spectral density $N_0/2$.

Ⅲ. CHANNEL ESTIMATION SCHEME FOR WLAN SYSTEMS

Fig. 1 shows the block diagram of the baseband receiver with the channel estimation. For the l-th tracked branch, multiplying the complex code word $\Psi^m(t)$ to $d(t)$ and correlating with the m-th code word, the decision variable for the l-th branch can be determined as follows

$$\begin{aligned} R(m,l) &= \frac{1}{\sqrt{P_w}}\int_{\tau_{n_l}}^{T_w+\tau_{n_l}} d(t)\,\Psi^m(t-\tau_{n_l})^* dt \\ &= R_I(m,l) + jR_Q(m,l) \end{aligned} \quad (8)$$

where $R_I(m,l) = R_{II}(m,l) + R_{QQ}(m,l)$ and $R_Q(m,l) = R_{QI}(m,l) - R_{IQ}(m,l)$. In eqn. (8), the decision variable $R_{II}(m,l)$ is given by

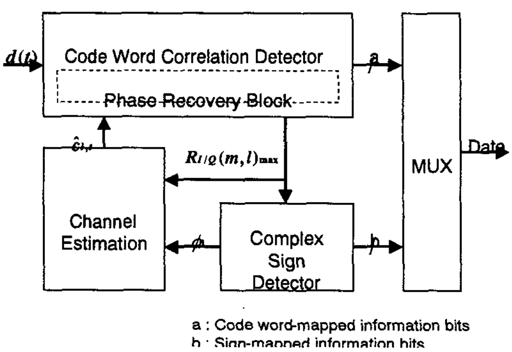

Fig. 1. Functional block diagram of the channel estimation.

$$\begin{aligned} R_{II}(m,l) &= \frac{1}{\sqrt{P_w}}\int_{\tau_{n_l}}^{T_w+\tau_{n_l}} \sqrt{P_w}\alpha_n \\ &\left\{C_I^j(t-\tau_{n_l})\frac{\cos\theta_n}{2} - C_Q^j(t-\tau_{n_l})\frac{\sin\theta_n}{2}\right\} \\ &\times \Psi_I^m(t-\tau_{n_l})dt + J_{II}(l) + W_{II}(l) \end{aligned} \quad (9)$$

where $J_{II}(l)$ is the self-interference to the l-th branch of the RAKE receiver due to multipath and $W_{II}(l)$ is due to the AWGN. Similarly, we can obtain $R_{QQ}(m,l)$, $R_{IQ}(m,l)$, and $R_{QI}(m,l)$ as follows

$$\begin{aligned} R_{QQ}(m,l) &= \frac{1}{\sqrt{P_w}}\int_{\tau_{n_l}}^{T_w+\tau_{n_l}} \sqrt{P_w}\alpha_n \\ &\left\{C_I^j(t-\tau_{n_l})\frac{\sin\theta_n}{2} + C_Q^j(t-\tau_{n_l})\frac{\cos\theta_n}{2}\right\} \\ &\times \Psi_Q^m(t-\tau_{n_l})dt + J_{QQ}(l) + W_{QQ}(l) \end{aligned} \quad (10)$$

$$\begin{aligned} R_{IQ}(m,l) &= \frac{1}{\sqrt{P_w}}\int_{\tau_{n_l}}^{T_w+\tau_{n_l}} \sqrt{P_w}\alpha_n \\ &\left\{C_I^j(t-\tau_{n_l})\frac{\cos\theta_n}{2} - C_Q^j(t-\tau_{n_l})\frac{\sin\theta_n}{2}\right\} \\ &\times \Psi_Q^m(t-\tau_{n_l})dt + J_{IQ}(l) + W_{IQ}(l) \end{aligned} \quad (11)$$

$$\begin{aligned} R_{QI}(m,l) &= \frac{1}{\sqrt{P_w}}\int_{\tau_{n_l}}^{T_w+\tau_{n_l}} \sqrt{P_w}\alpha_n \\ &\left\{C_I^j(t-\tau_{n_l})\frac{\sin\theta_n}{2} + C_Q^j(t-\tau_{n_l})\frac{\cos\theta_n}{2}\right\} \\ &\times \Psi_I^m(t-\tau_{n_l})dt + J_{QI}(l) + W_{QI}(l) \end{aligned} \quad (12)$$

The decision variable $R_I(m,l)$ and $R_Q(m,l)$ have two different forms according to the transmitted value of ϕ_1. For $\cos\phi_1 \neq 0$ (i.e., $\phi_1 = 0$ or π), first, the inphase and quadrature components of $R(m,l)$ are given by

$$R_I(m,l) = \begin{cases} \alpha_{n_l}\sqrt{E_w}\cos\phi_1\frac{\cos\theta_{n_l}}{2} + J_{II}(l) \\ + J_{QQ}(l) + W_{II}(l) + W_{QQ}(l), \\ \quad (m=j) \\ J_{II}(l) + J_{QQ}(l) + W_{II}(l) + W_{QQ}(l), \\ \quad (m \neq j) \end{cases} \quad (13)$$

and

$$R_Q(m,l)=\begin{cases}\alpha_{n_l}\sqrt{E_w}\cos\phi_1\dfrac{\sin\theta_{n_l}}{2}+J_{IQ}(l)\\ +J_{QI}(l)+W_{IQ}(l)+W_{QI}(l),\\ \quad (m=j)\\ J_{IQ}(l)+J_{QI}(l)+W_{IQ}(l)+W_{QI}(l),\\ \quad (m\neq j)\end{cases} \tag{14}$$

For $\sin\phi_1\neq 0$ (i.e., $\phi_1=\frac{\pi}{2}$ or $-\frac{\pi}{2}$), on the other hand, $R_I(m,l)$ and $R_Q(m,l)$ can be written by

$$R_I(m,l)=\begin{cases}-\alpha_{n_l}\sqrt{E_w}\sin\phi_1\dfrac{\sin\theta_{n_l}}{2}\\ +J_{II}(l)+J_{QQ}(l)+W_{II}(l)+W_{QQ}(l),\\ \quad (m=j)\\ J_{II}(l)+J_{QQ}(l)+W_{II}(l)+W_{QQ}(l),\\ \quad (m\neq j)\end{cases} \tag{15}$$

and

$$R_Q(m,l)=\begin{cases}\alpha_{n_l}\sqrt{E_w}\sin\phi_1\dfrac{\cos\theta_{n_l}}{2}+J_{IQ}(l)\\ +J_{QI}(l)+W_{IQ}(l)+W_{QI}(l),\\ \quad (m=j)\\ J_{IQ}(l)+J_{QI}(l)+W_{IQ}(l)+W_{QI}(l),\\ \quad (m\neq j)\end{cases} \tag{16}$$

Using eqns. (13)~(16), therefore, we select the decision variables to give the maximum value of $|R_I(m,l)|$ and $|R_Q(m,l)|$ for $m=1,2,\cdots,64$, which are denoted as $R_I(m,l)_{\max}$ and $R_Q(m,l)_{\max}$, and generally it is obtained when $m=j$. From this generality, we may deduce that $R_I(m,l)_{\max}$ and $R_Q(m,l)_{\max}$ are a real and imaginary terms of the channel coefficient plus AWGN, respectively. In eqns. (13)~(16), the self-interferences due to multipa.th and interference due to AWGN can be modelled as zero-mean AWGNs [10]. Therefore, we can design an unbiased estimator based on $R_I(m,l)_{\max}$ and $R_Q(m,l)_{\max}$. Fading channel parameters may be represented as a first-order Gauss-Markov process[11], i.e., the channel parameters to be estimated are correlated with the value of previous channel parameter. From eqns. (13)~(16), therefore, to track the fading channel, the update rule for a multipath fading channel can be designed as follows

$$\hat{c}_{l,t}=\begin{cases}\beta\,\hat{c}_{l,t-1}+(1-\beta)\dfrac{2\cos\phi_1}{\sqrt{E_w}}\\ \{R_I(m,l)_{\max}+jR_Q(m,l)_{\max}\},\\ \quad \text{for }\cos\phi_1\neq 0\\ \beta\,\hat{c}_{l,t-1}+(1-\beta)\dfrac{2\sin\phi_1}{\sqrt{E_w}}\\ \{R_Q(m,l)_{\max}-jR_I(m,l)_{\max}\},\\ \quad \text{for }\sin\phi_1\neq 0\end{cases} \tag{17}$$

where $\hat{c}_{l,t}$ is the estimated complex value of the channel of the l-th path at time t, and β is the update (or forgetting) factor in the estimation of the channel. It is easily expected that the choice of the forgetting factor β involves a trade-off between tracking speed and accuracy in the coefficient estimate, and the selection of the optimum β should be required according to the Doppler frequency of a multipath fading channel, which will be validated by simulation results in Section Ⅳ.

If the estimation of the channel parameter is assumed to be perfect, the output of the phase recovery block, $d'(t)$, can be written as

$$d'(t)=\frac{1}{2}\sum_{n=1}^{N}\sqrt{P_w}(\alpha_n)^2C^j(t-\tau_n)+\frac{1}{2}n(t) \tag{18}$$

Using eqn. (9), we can obtain the phase-recovered decision variables $R_I'(m,l)$ and $R_Q'(m,l)$, the decision variable obtained from the l-th multipath is $R'(m,l)=R_I'(m,l)+R_Q'(m,l)$. Combining all N multipaths, the decision variable can be obtained as $R_m=\sum_{l=1}^{N}R'(m,l)$, which may be seen as a Gaussian random variable with mean $\sum_{l=1}^{N}2(\alpha_{n,l}/2)^2\sqrt{P_w}$ when $m=j$ or 0 when $m\neq j$ since the phase-recovered decision variables $R_I'(m,l)$ and $R_Q'(m,l)$ are Gaussian random variables as described in [10]. For the CCK demodulation in the 11 Mbps mode, as shown in Fig. 1, a bank of 64 complex correlators determines which code was 6 bits of the data word. The other 2 bits of the 8-bit data word are determined from the complex sign detector.

IV. SIMULATION RESULTS AND DISCUSSIONS

To validate the effectiveness of the proposed channel estimation technique, extensive series of Monte-Carlo simulations have been conducted in Rayleigh-fading channel using the parameters specified in IEEE 802.11 standard with (1) Number of RAKE branches : $N=4$ and (2) Doppler frequency : $B_d=10\text{Hz}$.

Fig. 2 shows the probability density function (PDF) of the normalized estimation error for various values of β. 2. It is shown from the figure that the

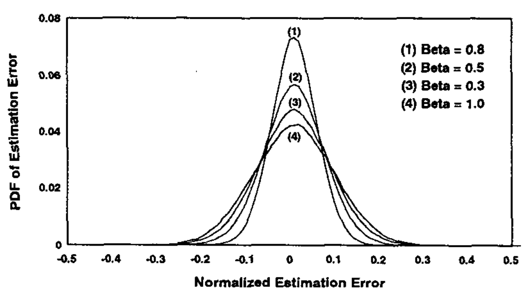

Fig. 2. The PDF of estimation error for various values of β.

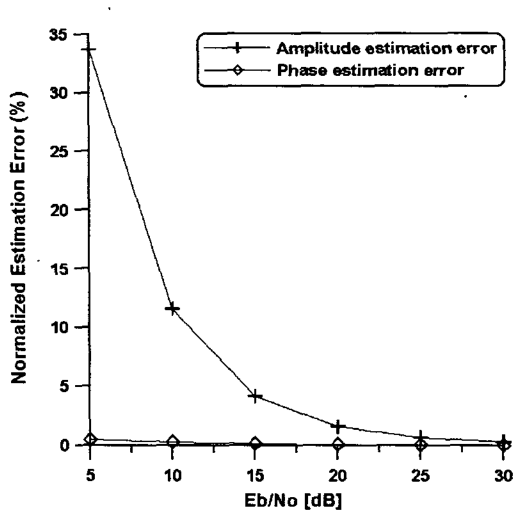

Fig. 3. The normalized estimation error ratio of the proposed channel estimator versus E_b/N_0 with $\beta=0.8$.

PDF of the estimation error has a minimum variance when $\beta=0.8$. Since β is sensitive to the channel characteristics, in a slow fading channel the large value of β is good, whereas in a fast fading channel small value is appropriate. Futhermore, the variance of the estimation error has a maximum when $\beta=1$, which is due to the fact that the channel coefficient should be updated without the aid of the present decision variables. As described earlier, however, the WLAN system operates in the indoor channel with a maximum Doppler spread of 10Hz based on the mobile speed of pedestrians. For a fixed value of β, therefore, the proposed channel estimation scheme performs efficiently.

The channel amplitude and phase estimation errors in a normalized logarithmic scale is shown in Fig. 3. We see from Fig. 3 that the proposed channel estimator corrects more efficiently the phase distortion of the channel than the amplitude distortion of the channel. As described above, since the phase error of the channel is a dominating term compared with the amplitude error of the channel, the compensation of the phase distortion affects

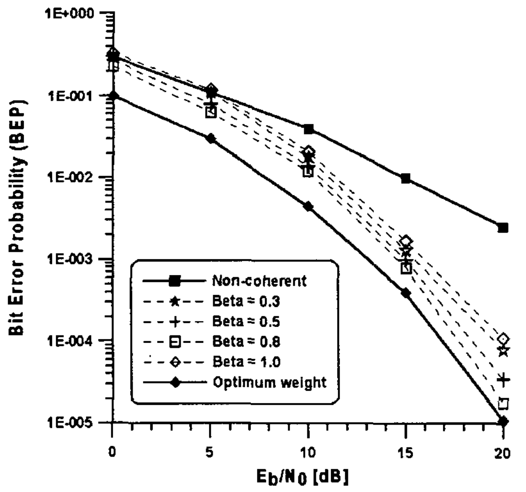

Fig. 4. The bit error probability of the channel estimator versus E_b/N_0 for various values of β.

dramatically bit error probability (BEP) performance.

The BEP versus E_b/N_0 for various values of β is shown in Fig. 4. It is shown from the Fig. 4 that the proposed detection method with phase recovery outperforms a conventional noncoherent method at high SNR. As expected, the channel estimator with $\beta=0.8$ gives the best BEP performance. For both methods, however, the performance degradation is observed at low SNR due to the large estimation error. Such a problem results from the self-interference due to the multipath, which is circumvented by combining the proposed channel estimator with the interference canceller.

V. CONCLUDING REMARKS

In this paper, a channel estimation method in a WLAN system with CCK modulation over a multipath fading channel. The proposed estimation method is based on the idea that channel parameters can be extracted from the maximum correlator output in a receiver block. The estimation scheme has an accurate ability of the estimation with little more extra hardware. The performance improvement results from the coherent demodulation with the help of the channel estimation. Further performance improvement can be obtained by combing the proposed channel estimation scheme with the self-interference canceller.

Especially, the investigated estimation scheme can be applicable to the high-rate version of Bluetooth system, which will be released after Bluetooth Ver. 1.1, and is very effective for solving the problem of the channel distortion for the implementation of the high-rate Bluetooth system without degradation of frame and spectral efficiency.

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