Evaluation of an Efficient Channel Estimator for the STTD Schemes

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Abstract—This paper evaluates the performance combining space-time transmit diversity (STTD) and an efficient channel estimator (ECE) for wideband code division multiple access (WCDMA) systems in various mobile channels. Using decision variable (DV), we also derive the analytic bit error rate (BER) and mean square error (MSE) for WCDMA applying ECE for STTD schemes. The simulation results show that the ECE performance is superior to the previous works in [1] as because we use additional pilot diversity which is so called secondary common control physical channel (S-CCPCH). The performance in case of the channel estimator using only one-channel or two-channel is worse than that of an ECE as about the maximum 4 dB at BER 1.0E-3 satisfying voice service over Rician fading channel. Our results show that, even with ICE, an ECE algorithm are effective in improving the output SNR and significantly reduce the error floor. In addition, the simulation results investigated in this paper also reveal that WCDMA combining an ECE and the STTD scheme could provide appreciable performance improvements in Rayleigh fading channel.

Index Terms— **STTD, channel estimation, WCDMA**

I. INTRODUCTION

New standards in the framework of so called fourth generation (4G) systems are investigated actively in the field of wireless communications. These systems include networks with multi-mode, multi-band, and multimedia high capacity mobile terminals. Such future systems should be able to fulfill the stringent requirements for quality of service (QoS), mainly in terms of throughput, delay and error rate. This situation necessitated advanced transmission techniques which could not only guarantee QoS but also manage resources efficiently. The space-time codes are considered as one of the most promising techniques to meet the stringent requirements in 4G systems [2]. Motivated by the Alamouti's space time transmit diversity (STTD), many further researches and simulation results were presented [3]-[5]. In most of the previous works on STTD schemes, perfect channel estimation was assumed [2-5]. Therefore most of the related works could get high diversity gain from a large number of multi-paths due to its small coherence bandwidth. It was also assumed that channel gain was always a constant which was defined as quasi-static Rayleigh fading (QSRF) while the transmitter transmits 2 symbols. Accordingly they could achieve the dual diversity obtaining about 3 dB gain being identical performance of maximum ratio combining (MRC) in the downlink of WCDMA system using space-time transmit diversity with perfect channel estimation. In practice, accurate channel estimation is difficult to obtain either due to rapid changes in the channel or due to the overhead needed to estimate a large number of parameters such as in a multi-input multi-output (MIMO) system [6]. In this paper, we evaluate the performance for an ECE and the STTD schemes with imperfect channel estimation in downlink WCDMA systems. The work presented in this paper is motivated by the previous work based on [1]. This channel estimator estimates the channel gain by combining the common pilot symbols in CPICH symbols and the dedicated pilot symbols, which are inserted in DPCH. The additional pilot diversity, secondary common control physical channel (S-CCPCH), is used for our proposed channel estimator. In the future ubiquitous (or pervasive) communication age, all the networks and services may be integrated [7], implying that keeping commonalities of technologies between networks is very important. This commonality requirement calls for evaluation of important 4G technologies in various networks and channels. Therefore, the purpose of this paper is evaluating the performance combining the proposed channel estimation and the STTD scheme in various mobile communication systems. The work presented in this paper is motivated by the observation that for the special case of STTD based on Alamouti's scheme, it is possible to obtain exact closed-form expression for the bit error probability (BEP). Using [8], an exact BEP expression would serve as an attractive alternative to previously derived bounds for evaluation performance applying the proposed channel estimation algorithm to the STTD scheme in downlink WCDMA. WCDMA mobile channel models to simulate the performance of the STTD scheme in various situations. We use Rayleigh multi-path fading and Rician multi-path fading for various terrestrial mobile channels. In case of mobile satellite systems, usually it is very difficult to get an advantage of the STTD scheme in a mobile satellite system due to its inherent large coherence bandwidth. For this reason, authors in [9]

Manuscript received June 9, 2007.

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suggested the space-time minimum mean square error (MMSE) reception with two satellites. In this paper we estimate the performance of the terrestrial STTD scheme having channel estimation error in Rake receiver after we apply the proposed channel estimation algorithm to it, and analyze the results. The rest of this paper is organized as follows. Following this introduction, Section II we describe the algorithm of the proposed channel estimator and compare with the previous work which is in [1]. Section III describes the basic principles of the STTD in downlink WCDMA systems. Section IV introduces bit error probability based on DV approaches in Rayleigh/Rician multi-path fading channel. The simulation results and discussions are given in Section V. Conclusions follow in Section VI. Bold typeface letters (e.g., **X** or **x**) represent vectors.

II. AN EFFICIENT CHANNEL ESTIMATROR

The most demanding task having the most significant effect on the performance of the RAKE receiver is the channel estimation in which the channel complex coefficients are estimated [10]. Channel estimation can be divided in delay estimation and channel amplitude estimation. It has been assumed that propagation delays are fixed and known to the receiver [10]. However, the number of dedicated pilot symbols in a slot is very limited ranging from 1 to 16 symbols depending on the service (slot format). It is clear that reliable channel estimation cannot be accomplished based on dedicated pilots alone [10]. Therefore we propose channel estimator based on common, dedicated and secondary common control channel (S-CCPCH) pilot jointly. In other words, to estimate the complex amplitudes of the propagation paths, the channel estimator combining dedicated pilot symbols time-multiplexed into each slot of the dedicated physical channel (DCH), pilot symbols of the S-CCPCH and the continuous common pilot channel (CPICH) can be used for channel estimation algorithm. The main differences between the method in [1] and the proposed estimator is that pilot symbols of the S-CCPCH additionally used in the method in [1]. In Fig. 1, we show the relative timing relationship for the proposed estimator where α, β, and γ are weighting factor corresponding to the received SNR of DPCCH, S-CCPCH, and CPICH, respectively. To normalize all of the weighting factor we assume $\alpha + \beta + \gamma = 1$.

In Fig.1, n, k and l denote n-th slot, k-th symbol, and l-th resoluble multi-path, respectively. In addition $P'_{CPICH}(n,k,l)$, $P'_{S\text{-}CCPCH}(n,k,l)$, and $P'_{DPCCH}(n,k,l)$ mean estimates of CPICH, S-CCPCH, and DPCCH, respectively. A problem of this approach is that dedicated and common pilot symbols are received with different power levels which makes their optimal combining difficult. However in order to reduce system complexity, combining methods used each of the channels is given by Table I. Since the Cramer-Rao lower bound (CRLB) is useful to compare the

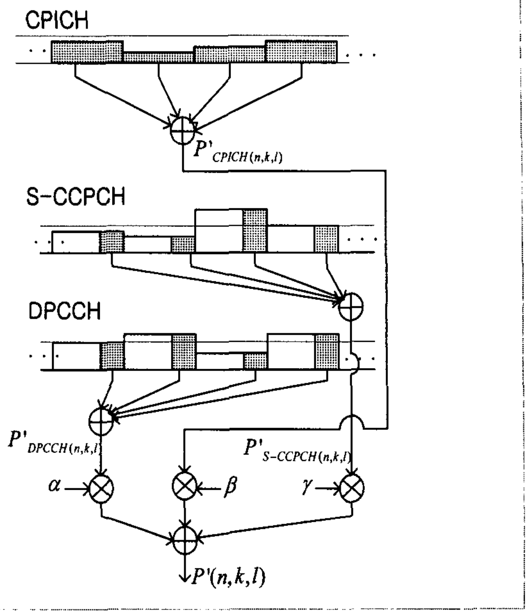

Fig. 1 Example of a timing diagram for an efficient channel estimator.

Table 1 Combining methods for channel estimation.

	In case of [1]	*An efficient channel estimator*
FIRST	1/3 < 2/3	1/7 < 2/7 < 4/7
SECOND	SNR1 < SNR2	SNR1 < SNR2 <SNR3
THIRD [11]	0.5 < 0.6	0.5 < 0.6 < 0.7

efficiencies of two different estimators, in the following we prove that combining the CPICH symbols, pilot symbols of the DPCCH, and pilot symbols of the S-CCPCH can lower the CRLB if we know the power ratio γ(i) perfectly. Using [1] in conclusion we have

$$\frac{1}{\frac{N}{\sigma_n^2}+\frac{M}{\mu\sigma_m^2}+\frac{K}{\mu\mu_1\sigma_k^2}} \le \frac{1}{\frac{N}{\sigma_n^2}+\frac{M\lambda^2}{\sigma_m^2}} \le \frac{\sigma_n^2}{N} \tag{1}$$

where N, M, and K are the number pilot symbol within one slot CPICH, DPCCH, and S-CCPCH, respectively. In addition, $\lambda(j)=(1/\mu(j))^{1/2}$ means power of pilot symbols of the DPCCH to power of the CPICH ratio. Also $\lambda_1(k)=(1/\mu_1(k))1/2$ stands for power of pilot symbols of the S-CCPCH to power of the CPICH ratio. σ_n^2, σ_m^2, and σ_k^2 are variance of AWGN in case of only CPICH is used for channel estimation, combining DPCCH and CPICH is used for channel estimation, and combining DPCCH, CPICH, and S-CCPCH, respectively. Therefore, CRLB of channel estimator combining CPICH, DPCCH, and S-CCPCH has a lower CRLB as compared with the CRLB of the estimator that uses only the CPICH symbols or combining CPICH and DPCCH.

III. STTD IN DOWNLINK WCDMA SYSTEMS

A. Encoder Structure

We present complex orthogonal design architecture used by Alamouti which is the simplest in the design of the space-time codes. S_1 and S_2 are transmitted symbols. From time t to $t+T$, antenna 1 and antenna 2 transmit S_1 and $-S_2^*$ respectively, where T is symbol duration. From time $t+T$ to $t+2T$, they transmit S_2 and S_1^*, respectively, where * is complex conjugate operation. The Hurwitz-Radon theorem showed that square complex linear processing orthogonal designs cannot achieve full diversity and full rate simultaneously, except in the two transmit antenna case [12]. Therefore the space-time codes design architecture can be used only for two transmitters.

B. Transmitter Structure

Figure 2 shows the block diagram of a transmitter in the downlink WCDMA system using space-time codes.

From time t to $t+T$, the transmitted complex signal of each antenna is expressed as

$$\begin{aligned} S_{A1,t}(t) &= h_1(t)C_cS_c(b_0\cos w_ct - b_1\sin w_ct) + n_1 \\ S_{A2,t}(t) &= h_2(t)C_cS_c(-b_2\cos w_ct - b_3\sin w_ct) + n_2 \end{aligned} \quad (2)$$

where $S_{A1,t}$ and $S_{A2,t}$ represent the transmitted complex signal of antenna 1 and antenna 2 at time t, respectively. In addition, $h_i(t)=a_ie^{j\theta}$, C_c, S_c, b_i, n_i and omegac represent channel gain, orthogonal variable spreading factor (OVSF) codes, scrambling codes, random number of {-1,1}, zero-mean complex Gaussian random variable with variance of $N_0/2$ per dimension and carrier frequency, respectively.

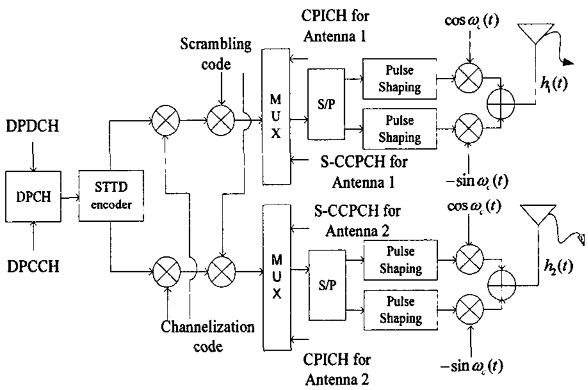

Fig. 2 transmitter in downlink WCDMA systems combining the ECE and STTD schemes.

Similarly, from time $t+T$ to $t+2T$, the transmitted complex signal of each antenna is given by

$$\begin{aligned} S_{A1,t+T}(t) &= h_1(t)C_cS_c(b_2\cos w_ct - b_3\sin w_ct) + n_3 \\ S_{A2,t+T}(t) &= h_2(t)C_cS_c(b_0\cos w_ct + b_1\sin w_ct) + n_4 \end{aligned} \quad (3)$$

where $S_{A1,t+T}$ and $S_{A2,t+T}$ represent the transmitted complex signal of antenna 1 and antenna 2 at time $t+T$, respectively.

C. Receiver Structure

Figure 3 shows the block diagram of a receiver in the downlink WCDMA system using STTD.

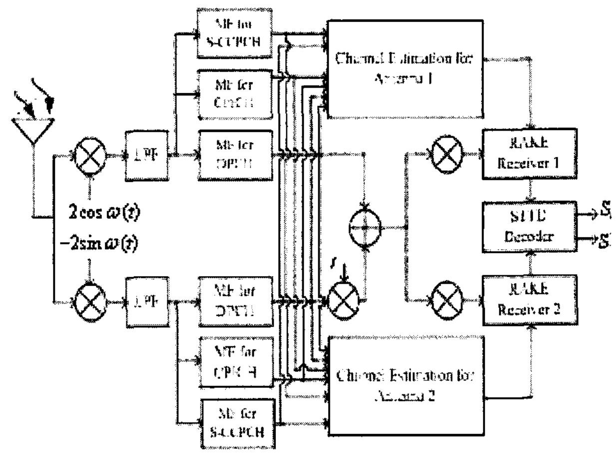

Fig. 3 Receiver in downlink WCDMA systems combining the ECE and STTD schemes.

The receiver is composed of two RAKE receivers, one of which includes M fingers, receiving complex signals. The received s signal from the low pass filter (LPF) is multiplied by a de-scrambling code and in succession by OVSF code. Then we can detect the desired signal using a simple ML decoder, where we use several channel estimation algorithms [13]. After the received signal is down converted to base-band, the incoming signal of the first RAKE receiver is described by

$$\begin{aligned} S_{R1,t}(t) &= h_1(t)C_cS_cb_0 + jh_1(t)C_cS_cb_1 \\ &\quad + 2(t)n_1 - j2(t)n_2 \\ S_{R1,t+T_s}(t) &= h_1(t)C_cS_cb_2 + jh_1(t)C_cS_cb_3 \\ &\quad + 2(t)n_2 - j2(t)n_2 \end{aligned} \quad (4)$$

where $S_{R1,t}$ and $S_{R1,t+T}$ represent incoming signal of the first RAKE receiver at time t and $t+T$, respectively. Similarly, incoming signal of the second RAKE receiver is expressed as

$$\begin{aligned} S_{R2,t}(t) &= -h_2(t)C_cS_cb_2 + jh_2(t)C_cS_cb_3 + 2n_1 - j2n_1 \\ S_{R2,t+T}(t) &= h_2(t)C_cS_cb_0 - jh_2(t)C_cS_cb_1 + 2n_2 - j2n_2 \end{aligned} \quad (5)$$

where $S_{R2,t}$ and $S_{R2,t+T}$ represent in coming signal of the second RAKE receiver at time t and $t+T$, respectively. The output values of the first and the second RAKE receiver having four fingers at time t are given by

$$\begin{aligned} Z_1(t) &= \sum_{j=1}^{4} W_{1,t}^j\{h_1(t)(b_0 + jb_1) + N_1\}, \\ Z_2(t) &= \sum_{j=1}^{4} W_{2,t}^j\{h_2(t)(-b_2 + jb_3) + N_3\} \end{aligned} \quad (6)$$

$W_{i,t}^j$ and $W_{i,t+T}^j$ represent weighting factors used in the channel estimation algorithm applying to the j-th finger of the i-th RAKE receiver, T is the bit duration. Similarly, the output values of the received signal through each finger of the first and the second RAKE receiver at time $t+T$ are given by

$$\begin{aligned} Z_1(t+T) &= \sum_{j=1}^{4} W_{1,t+T}^{j}\left\{h_1(t)(b_2+jb_3)+N_2\right\}, \\ Z_2(t+T) &= \sum_{j=1}^{4} W_{2,t+T}^{j}\left\{h_2(t)(b_0-jb_1)+N_4\right\} \end{aligned} \tag{7}$$

D. Decoder Structure

We can detect desired symbols by a decoder of the simple STTD using ECE. Each of the estimated symbols S_1' and S_2' can be expressed as

$$\begin{aligned} S_1' &= Z_1(t)+Z_2^{*}(t+T) \\ &= \left\{\sum_{j=1}^{4} W_{1,t}^{j}h_1^{j}+\sum_{j=1}^{4} W_{2,t+T}^{j}h_2^{j}\right\}\times(b_0+jb_1)+N_{1,total} \\ S_2' &= -Z_2^{*}(t)+Z_1(t+T) \\ &= \left\{\sum_{j=1}^{4} W_{2,t}^{j}h_2^{j}+\sum_{j=1}^{4} W_{1,t+T}^{j}h_1^{j}\right\}\times(b_2+jb_3)+N_{2,total} \end{aligned} \tag{8}$$

E. Imperfect channel estimation (ICE)

In most of the previous works on STTD schemes, perfect channel estimation was assumed [2-5]. It was also assumed that channel gain was always a constant which was defined as quasi-static Rayleigh fading (QSRF) while the transmitter transmits 2 symbols. Accordingly they could achieve the dual diversity obtaining about 3 dB gain being identical performance of maximum ratio combining (MRC) in the downlink of satellite/terrestrial WCDMA system using space-time codes with perfect channel estimation. However, in practice we can neither get ideal channel estimation in downlink nor QSRF.

In this paper, we evaluate the performance of the STTD schemes with imperfectly channel estimation. This is because, in practice, we cannot implement an ideal channel estimator in downlink WCDMA systems. We use a channel estimation error of (9).

$$W_i = H_i + E_i, \quad i = 0,1 \tag{9}$$

where W_i is the estimate for channel gain H_i which it is equivalent to $P'(n,k,l)$. Ei is the channel estimation error according to an independent zero-mean complex Gaussian random variable with variance of σ_z^2. The correlation coefficients of H_i and W_i are approximately σ_H/σ_W [6]. It is also assumed that W_i is a zero-mean complex Gaussian random variable only dependent on H_i with correlation coefficients ρ'. In general, correlation coefficients ρ' is given by

$$\rho' = \frac{1}{\sqrt{1+\sigma_z^2}} \tag{10}$$

The cross-correlation of H_i and W_i is easily seen to be $E\{H_i W_i^{*}\}= \rho'$ [14], where * is the complex conjugate of W_i.

IV. BIT ERROR PROBABILITY (BEP)

To evaluate the bit error probability, we use DV (Decision variable)-based approach in [15].

A. BER in Rayleigh multipath fading channel

Detection is carried out codeword by codeword, for known channel matrix **H**, the maximum likelihood (ML) detector is given by

$$\left[S_1'\ S_2'\right] = \arg\max \mathrm{Re}\left(\left(\mathbf{S}_k^{*}\right)^{T}\left(\mathbf{W}^{*}\right)^{T}\mathbf{A}_k\right) \tag{11}$$

where T means transpose operation and $\mathbf{A}_k$ denotes equation derived from (11). Note that the ML detector performs a maximum ratio combining operation on the received signal vector. Unlike the conventional MRC, here there are two weight vectors resulting in a 2×1 output. Let

$$\mathbf{D} = \left[Z_1'\ Z_2'\right]^{T} = \left(\mathbf{W}^{*}\right)^{T}\mathbf{A}_k \tag{12}$$

Note that D_1 is dependent only on the symbol S_1. Indeed, from (10)

$$\begin{aligned} Z_1' &= W_{1,t}^{*}\left\{h_1(b_0+jb_1)+N_1\right\} \\ &+ W_2\left\{h_2^{*}(-b_2+jb_3)+N_3\right\} \end{aligned} \tag{13}$$

The BEP of S_1' can be obtained from the probability density function (PDF) of Z_1'. Define the random variables: $J_1=W_1^{*}$, $J_2=W_2$, $K_1=h_1(b_0+jb_1)+N_1$ and $K_2=h_2^{*}(-b_2+jb_3)+N_3$ then Z_1' can be expressed

$$Z_1' = \sum_{i=1}^{2} J_i K_i^{*} \tag{14}$$

Condition on the symbol b_0+jb_1, the sets (X_i, Y_i), i=1,2, are two pairs of correlated, complex-valued, zero-mean, Gaussian random variables. The two pairs are however, mutually statistically independent and identically distributed. Define $Z_r=Re(Z_1')$ and $Z_i=Im(Z_1')$. The joint characteristic function $\phi(jv_1,jv_2)$ of the random variable Z_r and Z_i can be obtained using [7]. An alternative interpretation to (11) is that the phase of Z_1' is the decision variable for the detection of (b_0+jb_1). Define $R= (Z_r^2 +Zi^2)^{1/2}$, and $\phi=\tan^{-1}(Z_i/Z_r)$. Our goal is to obtain the PDF $p(\phi)$, where lower case notations for realization of the corresponding upper case denoted random variables. This is achieved as follows: compute the joint PDF of Z_r and Z_i, $p(z_r,z_i)$, from the Fourier transform of the joint characteristic function $\phi(jv_1,jv_2)$. From $p(z_r,z_i)$ obtain $p(r,\theta)$, the joint PDF of the envelope R and the phase θ . By integrating $p(r,\theta)$ over the variable r, we can obtain the PDF $p(\theta)$. The result can be found in [8]. In order to obtain normalized covariance in case of Rayleigh multi-path fading, we perform as follows.

$$\begin{aligned} m_{xx} &= E\left(|J_1|^2\right) = E\left(|W_1^*|^2\right) = E\left(|h_1+\varepsilon_1|^2\right) \\ &= 1+\sigma_z^2+2\left(\rho'-1\right) \end{aligned} \tag{15}$$

$$\begin{aligned} m_{yy} &= E\left(|K_1|^2\right) \\ &= E\left(|h_1(b_0+jb_1)|^2\right) = 1/2+N_0 \end{aligned} \tag{16}$$

where, $E(|h_1|^2)=1$ and $E(|(b_0+jb_1)|^2)=1/2$ because normalized power is used for two transmit antennas on WCDMA using STTD.

$$\begin{aligned} m_{xy} &= E\left(J_i K_i^*\right) = E\left(\left(W_1^*\right)\left(h_1(b_0+jb_1)\right)^*\right) \\ &= E\left(W_1^* h_1\right)\sqrt{1/2} = \sqrt{1/2}\rho' \end{aligned} \tag{17}$$

Therefore we can archive the normalized cross-correlation μ as follows.

$$\begin{aligned} \mu &= m_{xy}/\sqrt{m_{xx}m_{yy}} \\ &= \sqrt{1/2}\rho'/\left(\sqrt{1+\sigma_z^2+2\left(\rho'-1\right)}\sqrt{1/2+N_0}\right) \end{aligned} \tag{18}$$

The error probability of the received signal can be obtained by integration $p(\theta)$ over the angle interval complementary to the correct decision. According to [8], we can obtain BEP for WCDMA using STTD over Rayleigh fading channel in QPSK modulation while integrates $p(\theta)$ from $\pi/4$ to $5\pi/4$. The symbol error probability is expressed as

$$P_{4b}' = 2\int_{\pi/4}^{3\pi/4} p(\theta)d\theta + 4\int_{3\pi/4}^{\pi} p(\theta)d\theta \tag{19}$$

The bit error probability therefore is given by

$$P_{4b} = \frac{1}{2}\left(1-\rho-\frac{1}{2}\rho(1-\rho^2)\right) \tag{20}$$

where $\rho=\mu/((2-\mu^2))^{1/2}$.The analytic BEP for WCDMA using STTD according to ICE is shown by Fig. 4.

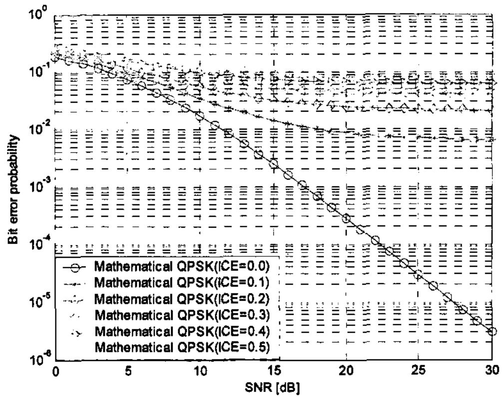

Fig. 4 BER for WCDMA using STTD according to ICE.

B. BER in Rician multipath fading channel

To evaluate the bit error probability, in similar manner we use equation (11)-(14) in BER in Rayleigh multi-path fading channel.

In order to obtain normalized covariance μ in case of Rician multi-path fading, we perform as follows.

$$\begin{aligned} m_{xx} &= E\left(|J_k-J_k'|^2\right) \\ &= E\left(\left(h_1^*+\varepsilon_1^*-K\right)\left(h_1+\varepsilon_1-K\right)\right) \\ &= 2\rho'-K^2-1/2+\sigma_z^2 \end{aligned} \tag{21}$$

where K_k' denotes mean of the random variable K_k and K is line-of-sight (LOS) which is so called Rice factor.

$$\begin{aligned} m_{yy} &= E\left(|K_k-K_k'|^2\right) \\ &= E\begin{pmatrix} \left(h_1(b_0+jb_1)+N_1-\sqrt{1/2}K\right)\times \\ \left(h_1^*(b_0+jb_1)^*+N_1^*-\sqrt{1/2}K\right) \end{pmatrix} \\ &= 1/4+N_0 \end{aligned} \tag{22}$$

where K_k' means expectation of the random variable K_k. Therefore we assume mean of random variable of the received signal is $(1/2K)^{1/2}$.

$$\begin{aligned} m_{xy} &= E\left(\left(J_k-J_k'\right)\left(K_k^*-K_k'^*\right)\right) \\ &= \sqrt{1/2}\left(\rho'-K^2\right) \end{aligned} \tag{23}$$

Therefore we can archive the normalized cross-correlation μ in Rician multi-path fading channel as follows.

$$\mu = \frac{\sqrt{1/2}\left(\rho'-K^2\right)}{\sqrt{\left(2\rho'-K^2-1/2+\sigma_\varepsilon^2\right)\left(1/4+N_0\right)}} \tag{24}$$

Using (20), the BEP for Rician fading channel is shown by Fig. 11 and Fig. 12. From Fig. 11 and Fig. 12, that the amount of variance of the ICE affects the performance of WCDMA systems is much larger than that the amount of K factor affects it.

C. MSE in Rayleigh/Rician multipath fading channel

We also consider mean square error to evaluate MSE in Rayleigh/Rician multi-path fading channel as follows.

$$MSE = E\left(\left|W_i - H_i\right|^2\right) \tag{25}$$

V. SIMULATION RESULTS AND DISCUSSION

A. BEP in Rayleigh multi-path fading channel

A comparison between methods based on [1] and ECE is given in Fig. 5-7, where we assume transmit bits is 50,000 bits and the received bit is synchronized by the WCDMA receiver.

We also compare fixed implementation to variable implementation. From Fig. 5 to Fig. 10, both MSE and BER of channel estimator combining CPICH, DPCCH, and S-CCPCH has a lower BER/MSE as compared with the BER/MSE of the estimator that uses only the CPICH symbols or combining CPICH and DPCCH in spite of imperfectly channel estimation. We also turn out performance of the fixed weighted channel coefficients operation is better than that of variable weighted channel coefficients.

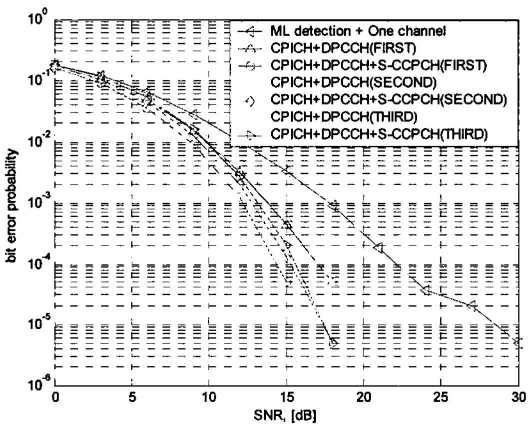

Fig. 5 BER for WCDMA combining ECE and STTD [ICE=0.0 dB].

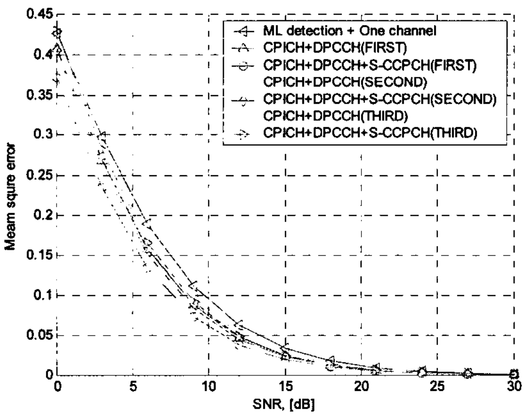

Fig. 6 MSE for WCDMA combining ECE and STTD [ICE=0.0dB].

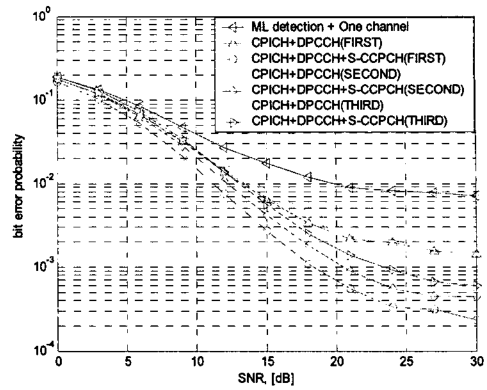

Fig. 7 BER for WCDMA combining ECE and STTD [ICE=0.3 dB].

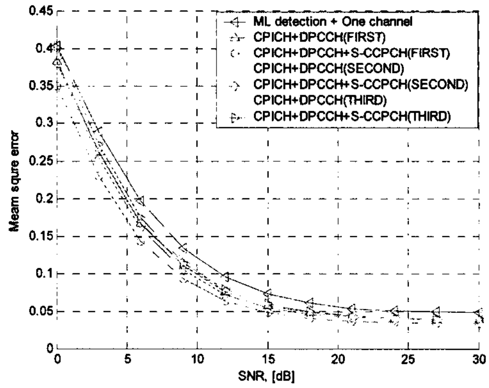

Fig. 8 MSE for WCDMA combining ECE and STTD [ICE=0.3 dB].

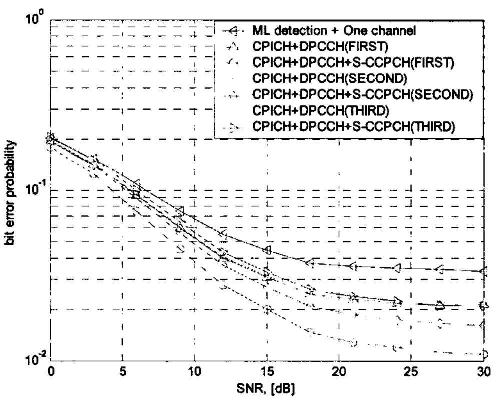

Fig. 9 BER for WCDMA combining ECE and STTD [ICE=0.5 dB].

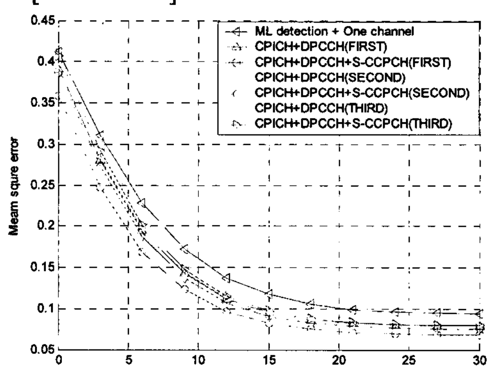

Fig. 10 MSE for WCDMA combining ECE and STTD [ICE=0.5 dB].

This is pregnant results in practical systems because it is impossible that variable weighting channel coefficients is used for the channel estimation. For example in Fig. 7 the required SNR of the channel estimator using the fixed weighting factor conditioned on BER at 1.0E-03 is approximately 17 dB. However the required SNR is about 20 dB. In addition, channel estimator based on [1] is inferior to ECE as all that. We similarly known the above same result is shown in sense of MSE.

B. BEP in Rician multi-path fading channel

The higher K factors result in lower bit error probability, as expected. For the high SNR, error floor exist for almost all channel estimation methods in similar with simulation results of Rayleigh fading channel in presence of ICE. However we make out the performance of WCDMA system using STTD is highly dependent on K factor even imperfect channel estimation in a comparison between Fig. 14 and Fig. 18. Furthermore, it is noted performance in case of the channel estimator using only one-channel or two-channel is worse than that of the ECE as about the maximum 4 dB at BER 1.0E-3 satisfying voice service in Fig. 18.

VI. CONCLUSIONS

In this paper, we analyzed the performance of the downlink WCDMA system combining the STTD and the ECE over various mobile communications channels. From the simulation results, it is noted that performance in case of the channel estimator using only one-channel or two-channel is worse than that of both [1]-based approach and ECE as about the maximum 4 dB at BER 1.0E-3 satisfying voice service. Our results show that, even with imperfect channel estimation (ICE), the ECE algorithm are effective in improving the output SNR and significantly reduce the error floor. Furthermore, the simulation results investigated in this paper also reveal that WCDMA combining the ECE and the STTD scheme could provide appreciable performance improvements in the presence of ICE over Rayleigh multi-path fading channel. The performance improvement is confirmed by simulation results.

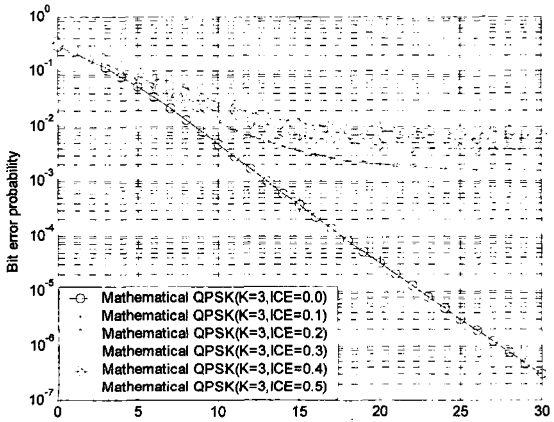

Fig. 11 BER for WCDMA using STTD according to [K=3 dB].

For further improvement in performance and efficiency, additional advanced techniques are required. These include powerful error correction coding, smart antennas, advanced channel estimation, and optimum combination of these.

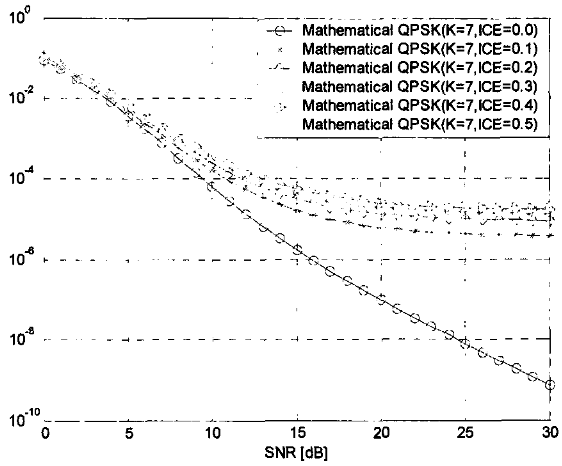

Fig. 12 BER for WCDMA using STTD according to [K=7 dB].

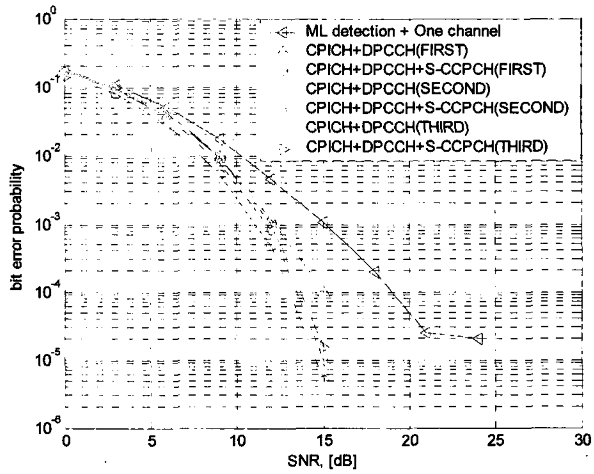

Fig. 13 BER for WCDMA combining ECE and STTD [K=3 dB, ICE=0.0 dB].

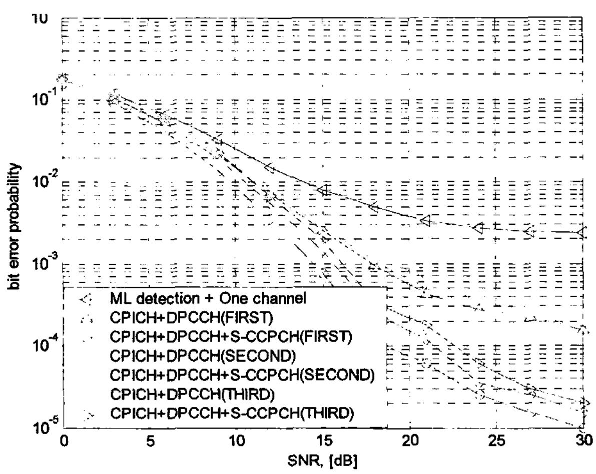

Fig. 14 BER for WCDMA combining ECE and STTD [K=3 dB, ICE=0.3 dB].

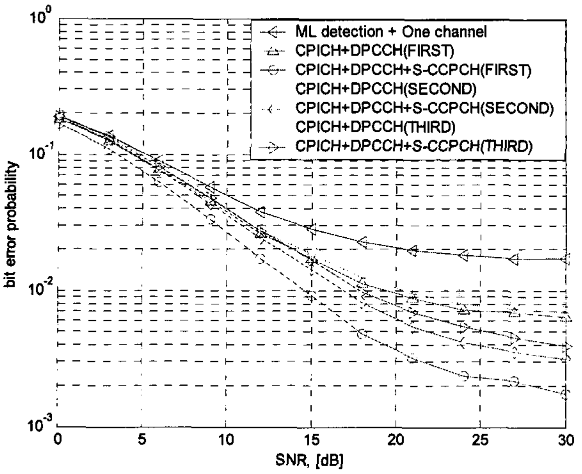

Fig. 15 BER for WCDMA combining ECE and STTD [K=3 dB, ICE=0.5 dB].

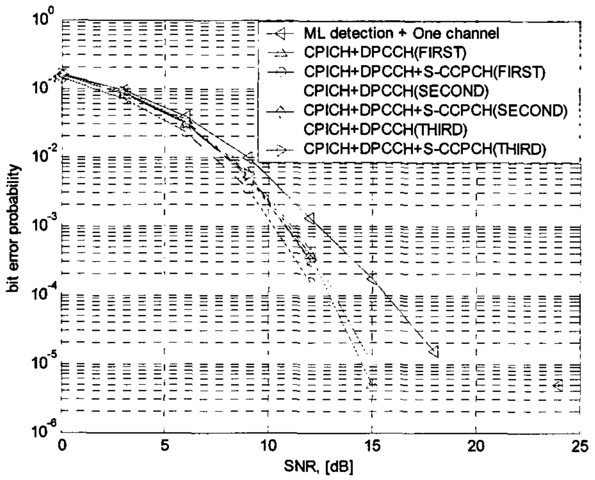

Fig. 16 BER for WCDMA combining ECE and STTD [K=7 dB, ICE=0.0 dB].

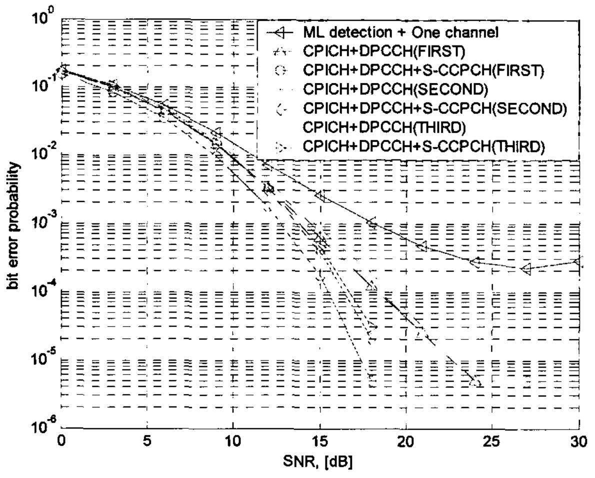

Fig. 17 BER for WCDMA combining ECE and STTD [K=7 dB, ICE=0.3 dB].

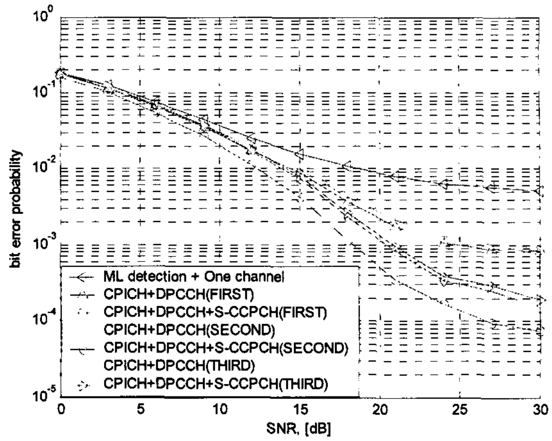

Fig. 18 BER for WCDMA combining ECE and STTD [K=7 dB, ICE=0.5 dB].

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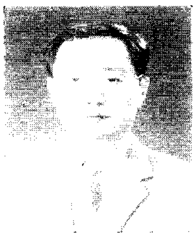

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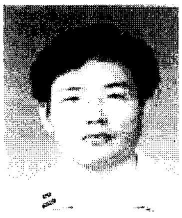

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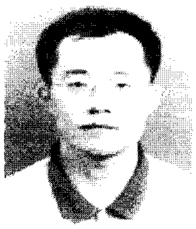

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