Rapid Slot Synchronization in the Presence of Large Frequency Offset and Doppler Spread in WCDMA Systems

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Abstract—**Slot synchronization is the most challenging step for rapid cell search in intercell asynchronous code division multiple access systems such as wide-band code division multiple access (WCDMA). For rapid cell search, it is desirable to design the receiver robust to initial frequency offset and Doppler spread. In this letter, we consider combining schemes for rapid slot synchronization of WCDMA signals in such channel impairments. We propose an inner-slot differential combining scheme that exploits partial correlation of slot synchronization code, making the receiver tolerable to a large amount of initial frequency offset and Doppler spread. Unlike conventional differential combining schemes, the proposed scheme is also applicable to the use of transmit antenna diversity. The detection performance of the proposed combining scheme is analyzed with the use of transmit antenna diversity. Finally, the analytic results are verified by computer simulation.**

Index Terms—**Cell search, differential detection, slot synchronization, time-switched transmit diversity (TSTD), wide-band code division multiple access (WCDMA).**

I. Introduction

IN SYNCHRONOUS code division multiple access (CDMA) systems such as IS-95 and cdma2000, the mobile station only needs to search the code timing for initial code acquisition. When each base station uses its own scrambling code in an intercell asynchronous downlink system like wide-band CDMA (WCDMA), the mobile station needs to resolve both the timing and code uncertainty for initial code synchronization [1]. The use of 512 complex Gold codes in the WCDMA downlink makes it impractical to exhaustively search all the possible codes for the code timing. This problem can be alleviated using a three-step cell-search scheme in the WCDMA system [2].

A three-step operation for initial cell search in the WCDMA system can be processed using two synchronization channels (SCHs), i.e., the primary SCH (P-SCH) and secondary SCH (S-SCH), and common pilot channel in the downlink physical channel [3], [4]. The P-SCH provides information on the slot timing (2560-chip timing candidates) and the S-SCH provides information on the code group and frame boundary ($64 \times 15 = 960$ candidates). Finally, the cell searcher selects one candidate among eight scrambling codes in a code group. Among these operations, the first step is the most elaborating process since it should resolve the largest amount of uncertainty.

In the first step of the three-step cell search, the receiver searches for the slot timing by correlating the received signal with the P-SCH code using a matched filter (MF). Since the signal-to-noise power ratio of the transmitted P-SCH code is very low, it is required to combine the correlation values over multiple slots to properly acquire the slot timing. A number of techniques have been proposed for the combining, including coherent combining, noncoherent combining, and differential combining [2], [5], [6].

Coherent combining schemes can provide the optimum performance in an additive white Gaussian noise channel environment, but it may not be applied to a real environment where channel distortion and RF impairment exist. Noncoherent combining schemes have widely been used due to its practical aspects [2]. It was shown that differential coherent combining schemes can outperform noncoherent combining schemes under a mild channel condition [6]. However, they cannot be applied to the use of time-switched transmit diversity (TSTD) [3], since the differential product is extracted from the two consecutive slots. Note that the phase of the received signals between the adjacent slots is usually independent in the TSTD mode. Moreover, it is necessary to employ a slot-timing detector that can operate independently of the use of transmit diversity, which may not be known to the receiver during the initial cell search stage.

When there exists a large amount of initial frequency offset, full correlation of the slot synchronization code may result in a large amount of incoherence loss. This problem can be alleviated using a partial correlation scheme [5]. The partial correlation values are often noncoherently combined for simplicity of implementation. However, differential detection is possible with the use of partial correlation values since the phase variation between the two consecutive partial correlation values heavily depends on the amount of frequency offset, which can be assumed stationary during the combining interval. In this letter, we propose an inner-slot differential combining scheme that exploits partial correlation of the synchronization code, which is also applicable to the transmit antenna diversity mode. Since the proposed scheme uses two consecutive partial correlation values to extract the differential product, it can be more robust to a large amount of Doppler spread than the differential detection scheme proposed in [6].

Following this Introduction, the system model is described in Section II. In Section III, the proposed scheme is described. The performance is analyzed when the transmit antenna diversity is employed in the presence of frequency offset and Doppler spread. The analytic results are verified by computer simulation in Section IV. Finally, conclusions are summarized in Section V.

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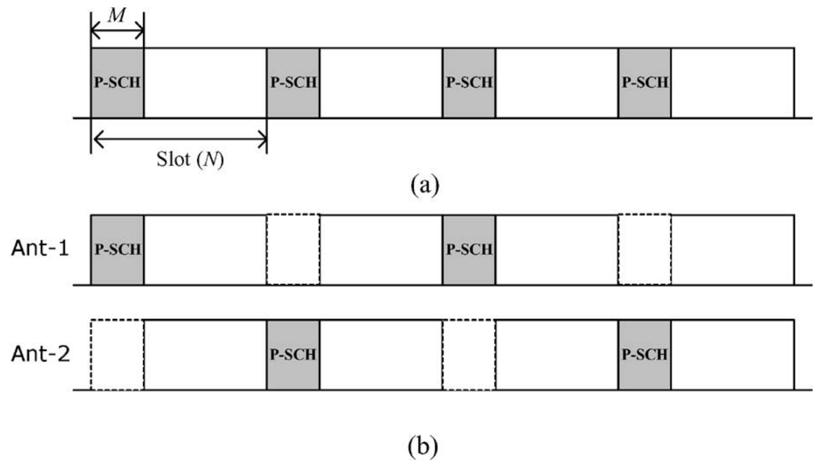

Fig. 1. Transmission of synchronization code. (a) No transmit antenna diversity. (b) Two transmit antenna diversity.

II. System Model

The slot timing is acquired using a P-SCH code in the first step of the three-step cell search operation. The base station repeatedly transmits the P-SCH code of length M during the first symbol of each slot comprising N chips. As shown in Fig. 1, one P-SCH code is alternately transmitted using a single antenna for each slot in the transmit diversity mode.

A complex baseband-equivalent received signal at the kth slot can be represented as

$$r_k[i] = \begin{cases} \sqrt{E_\mathrm{c}}\alpha_k \mathrm{e}^{j2\pi f_\mathrm{o} i T_\mathrm{c}+\theta} c_p[i] + n_k[i], & i \in [0, M-1] \\ n_k[i], & i \in [M, N-1] \end{cases} \tag{1}$$

where E_c is the chip energy, f_o is initial frequency offset due to the oscillator instability, T_c is the chip duration, θ is the phase offset, which can be assumed zero without the loss of generality, $c_p[i]$ denotes the P-SCH code of length M with $c_p[i]c_p^*[i] = 1$, $n_k[i]$ denotes all the interference terms, having one-sided power spectral density N_0, and α_k denotes the Rayleigh fading channel gain. It is assumed that the channel fading is static during M-chip interval and the signal undergoes flat fading. The power of the received signal is constant during each slot interval since the SCH signal is time multiplexed with the primary common control channel [3]. Thus, we can assume that the power of $n_k[i]$ is constant because the power of the P-SCH code is sufficiently low compared to that of the received signal.

In the initial code acquisition stage, there can be a large amount of frequency offset due to inherent oscillator instability. It was shown that the use of a partial correlation scheme could provide detection performance robust to the initial frequency offset [5]. The correlation interval is divided into L subintervals and L M_s-tap MFs are employed to obtain the partial correlation values, where $M = LM_s$. The partial correlation value of the lth subinterval at the kth slot can be represented as

$$a_{kl}[i] = \frac{1}{M_s} \sum_{m=(l-1)M_s}^{lM_s-1} r_k[i+m] c_p^*[m], \qquad k \in [1, N_s]; \quad l \in [1, L]. \tag{2}$$

The timing detection can be modeled as a simple hypothesis problem. Let H_1 and H_0 be the hypothesis that the two codes are aligned and not, respectively. Then the partial correlation value $a_{kl}[i]$ can be modeled as

$$a_{kl}[i] = \begin{cases} s_{kl} + \eta_{kl}, & H_1 \\ \eta_{kl}, & H_0 \end{cases} \tag{3}$$

where $s_{kl} = \alpha_k \sqrt{E_\mathrm{c}}[\sin(M_s \varepsilon \pi)/M_s \sin(\varepsilon\pi)]\exp(j2\pi\varepsilon M_s l)$ $\exp[-j\pi\varepsilon(M_s+1)]$, $\varepsilon = f_\mathrm{o} T_\mathrm{c}$, and η_{kl} denotes the interference term approximated as a zero-mean complex Gaussian random variable (RV) with $E\{|\eta_{kl}|^2\} = N_0/M_s$. The self-noise term can be ignored when M_s is not too small since E_c/N_0 is very low in a nominal operating environment.

III. Proposed Combining Scheme

A. Combining Scheme

A conventional detection scheme noncoherently combines the partial correlation value over N_s slots for reliable detection as [5]

$$z_n[i] = \sum_{k=1}^{N_s} \sum_{l=1}^{L} |a_{kl}[i]|^2 . \tag{4}$$

The slot timing can be obtained using a maximum likelihood method, i.e., by finding i such that

$$i_n = \underset{i \in [0, N-1]}{\arg\max} \left(z_n[i] \right). \tag{5}$$

In the presence of initial frequency offset, the phase variation between the two consecutive partial correlation values significantly depends on the amount of initial frequency offset. Assuming that the amount of frequency offset does not change during the combining interval (i.e., N_s slots), the inner-slot partial correlation values can be combined differentially for

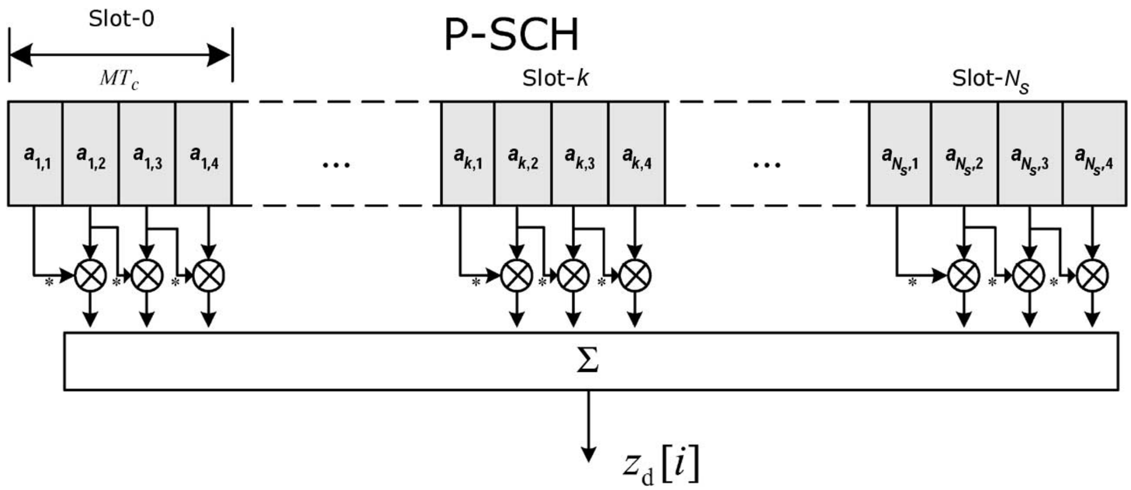

Fig. 2. Proposed combining scheme.

peak detection as shown in Fig. 2. Using the differentially combined output $\tilde{z}_{\mathrm{d}}[i]$ over N_s slots

$$\tilde{z}_{\mathrm{d}}[i] = \sum_{k=1}^{N_s}\sum_{l=1}^{L-1} a_{kl}^*[i]a_{k(l+1)}[i] \tag{6}$$

the slot timing can be obtained by finding the chip index i_{d} maximizing $z_{\mathrm{d}}[i] = |\tilde{z}_{\mathrm{d}}[i]|^2$.

The initial frequency offset can also be estimated using the phase information of $\tilde{z}_{\mathrm{d}}[i]$ since the amount of phase variation between the two consecutive partial correlation values is closely related to the initial frequency offset. The initial frequency offset can be estimated as $\hat{f}_{\mathrm{o}} = (1/2\pi M_s T_{\mathrm{c}})\arg\{\tilde{z}_{\mathrm{d}}[i_{\mathrm{d}}]\}$. This frequency offset estimator can facilitate the initial operation of automatic frequency control, providing rapid initial stabilization of the receiver.

B. Performance Analysis

The proposed scheme is evaluated in terms of the detection probability as a performance measure. The detection probability of the first step can be obtained by

$$P_{\mathrm{d}} = \int_0^\infty f_{H_1}(z)F_{H_0}^{N-1}(z)\mathrm{d}z \tag{7}$$

where $f_{H_1}(z)$ denotes the probability density function (pdf) of z under hypothesis H_1 and $F_{H_0}(z)$ denotes the cumulative distribution function (cdf) of z under hypothesis H_0. For simplicity of mathematical description, we will omit the chip index i because there are only two simple hypotheses.

To obtain $f_{H_1}(z)$ in the presence of Doppler spread, the fading channel gain should be characterized. According to Jake's model, α_k is a complex Gaussian random process whose autocorrelation is given by [10]

$$E\left\{\alpha_k\alpha_{k+m}^*\right\} = J_0(2\pi f_{\mathrm{d}} m N T_{\mathrm{c}}) \tag{8}$$

where $J_0(\bullet)$ is the zeroth order Bessel function of the first kind, $E\{X\}$ denotes the expectation of X, and f_{d} denotes the maximum Doppler frequency. Let $\boldsymbol{\alpha} = [\,\alpha_1 \quad \alpha_2 \quad \cdots \quad \alpha_{N_s}\,]^{\mathrm{T}}$ be the channel vector representing the channel gain during the combining interval. Taking the transmit antenna diversity into consideration, the (k,l)th element R_{kl} of the channel covariance matrix $\mathbf{R}$ can be obtained by

$$\begin{aligned} R_{kl} &= E\left\{\alpha_k\alpha_l^*\right\} \\ &= \begin{cases} J_0(2\pi f_{\mathrm{d}} N T_{\mathrm{c}}|k-l|)\,, & \mathrm{mod}(|k-l|, K) = 0 \\ 0, & \text{otherwise} \end{cases} \end{aligned} \tag{9}$$

where K is the number of transmit antennas and $\mathrm{mod}(a,b)$ denotes the remainder of a/b. It is assumed that there is no correlation between the antennas.

To characterize the combined correlation value during multiple slots, let us define γ by

$$\gamma = \boldsymbol{\alpha}^{\mathbf{H}}\boldsymbol{\alpha} = \sum_{i=1}^{N_s}|\alpha_i|^2. \tag{10}$$

The characteristic function of γ can be obtain as [6]

$$\Phi_\gamma(w) = E\{jw\boldsymbol{\alpha}^{\mathbf{H}}\boldsymbol{\alpha}\} = \prod_{k=1}^{N_s}\frac{1}{1 - jw\lambda_k} \tag{11}$$

where λ_k is the eigenvalue of $\mathbf{R}$. Note that all the eigenvalues are nonzero since $\mathbf{R}$ is a Toeplitz matrix. When all the eigenvalues are distinct, $\Phi_\gamma(w)$ can be expressed in a summation form of $\Phi_\gamma(w) = \sum_{k=1}^{N_s} g_k/(1 - jw\lambda_k)$ where $g_k = \Pi_{l=1,l\neq k}^{N_s}(\lambda_k/\lambda_k - \lambda_l)$. Then, the pdf of γ can be obtained by

$$f_\gamma(\gamma) = \sum_{k=1}^{N_s}\frac{g_k}{\lambda_k}\mathrm{e}^{-\frac{\gamma}{\lambda_k}}. \tag{12}$$

When all the eigenvalues are not distinct, $f_\gamma(\gamma)$ can easily be derived using a partial fraction decomposition method.

1) Conventional Combining: Under hypothesis H_0, z_n in (4) can be modeled as a central χ^2-distributed RV with $2LN_s$ degrees of freedom. The cdf of z_n under hypothesis H_0 can be expressed as [11]

$$F_{H_0}(z) = 1 - \mathrm{e}^{-\frac{z}{2\sigma_a^2}} \sum_{m=0}^{LN_s-1} \frac{1}{m!}\left(\frac{z}{2\sigma_a^2}\right)^m \tag{13}$$

where $\sigma_a^2 = N_0/2M_s$.

Under hypothesis H_1 and given γ, z_n can be modeled as a noncentral χ^2-distributed RV with $2LN_s$ degrees of freedom [11]

$$f_{H_1}(z|\gamma) = \frac{1}{2\sigma_a^2}\left(\frac{z}{s^2\gamma}\right)^{\frac{LN_s-1}{2}} \mathrm{e}^{-\frac{(s^2\gamma+z)}{2\sigma_a^2}} I_{LN_s-1}\left(\frac{s}{\sigma_a^2}\sqrt{\gamma z}\right) \tag{14}$$

where $I_n(\cdot)$ is the nth order modified Bessel function of the first kind and $s = \sqrt{E_\mathrm{c} L}\sin(M_s\varepsilon\pi)/(M_s \sin\varepsilon\pi)$. Averaging $f_{H_1}(z|\gamma)$ with respect to γ, using (12) and (14), it can be shown that

$$\begin{aligned} f_{H_1}(z) &= \int_0^\infty f_{H_1}(z|\gamma) f_\gamma(\gamma)\mathrm{d}\gamma \\ &= \sum_{k=1}^{N_s} \beta_k \left[1 - \frac{\Gamma\left(LN_s-1, \frac{s^2\lambda_k z}{2s^2\lambda_k\sigma_a^2+4\sigma_a^4}\right)}{\Gamma(LN_s-1)}\right] \mathrm{e}^{-\frac{z}{(s^2\lambda_k+2\sigma_a^2)}} \end{aligned} \tag{15}$$

where $\Gamma(x)$ is the gamma function, $\Gamma(x,y)$ denotes the incomplete gamma function defined by $\Gamma(x,y) = \int_y^\infty t^{x-1}\mathrm{e}^{-t}\mathrm{d}t$, and β_k is a constant term expressed as

$$\beta_k = \frac{g_k\lambda_k\left(\frac{2\sigma_a^2}{s^2\lambda_k}+1\right)^{LN_s} s^2}{(s^2\lambda_k+2\sigma_a^2)^2}. \tag{16}$$

2) The Proposed Combining: Under hypothesis H_0, $\tilde{z}_\mathrm{d}$ can be expressed as

$$\tilde{z}_\mathrm{d} = \sum_{k=1}^{N_s}\sum_{l=1}^{L-1} \eta_{kl}^*\eta_{k(l+1)}. \tag{17}$$

Since η_{kl} is a zero-mean complex Gaussian RV, the pdf of the multiplication of two η_{kl}s can be expressed as a form that includes the modified Hankel function [8]. Since $\tilde{z}_\mathrm{d}$ is the sum of $(L-1)N_s$ RVs, it can be approximated as a zero-mean complex Gaussian RV with $E\{|\tilde{z}_\mathrm{d}|^2\} = (L-1)N_s N_0^2/M_s^2$ from the central limit theorem. Thus, we can model z_d as a central χ^2-distributed RV with two degrees of freedom. The cdf of z_d is represented as

$$F_{H_0}(z) = 1 - \mathrm{e}^{-\frac{M_s^2 z}{(L-1)N_s N_0^2}}. \tag{18}$$

Under hypothesis H_1, we have

$$\begin{aligned} z_\mathrm{d}^{(k)} &\triangleq \sum_{l=1}^{L-1} a_{kl}^* a_{k(l+1)} \\ &= \sum_{l=1}^{L-1} \left(s_{kl}^* + \eta_{kl}^*\right)\left(s_{k(l+1)} + \eta_{k(l+1)}\right) \\ &= \sum_{l=1}^{L-1} s_{kl}^* s_{k(l+1)} + \sum_{l=1}^{L-2}\left(s_{kl}^*\eta_{k(l+1)} + s_{k(l+2)}\eta_{k(l+1)}^*\right) \\ &\quad + \left(s_{k2}\eta_{k1}^* + s_{k(L-1)}^*\eta_{kL}\right) + \sum_{l=1}^{L-1}\eta_{kl}^*\eta_{k(l+1)}. \end{aligned} \tag{19}$$

Here, the first term is the signal component expressed as

$$\sum_{l=1}^{L-1} s_{kl}^* s_{k(l+1)} = (L-1)|\alpha_k|^2 \mathrm{e}^{j2\pi M_s\varepsilon} \frac{\sin^2(M_s\varepsilon\pi)}{M_s^2\sin^2(\varepsilon\pi)} E_\mathrm{c} \tag{20}$$

and the second term is the noise term due to successive differential summation, represented as

$$\begin{aligned} &\sum_{l=1}^{L-2}\left(s_{kl}^*\eta_{k(l+1)} + s_{k(l+2)}\eta_{k(l+1)}^*\right) \\ &\quad = \sum_{l=1}^{L-2}\left(s_{kl}^*\eta_{k(l+1)} + \mathrm{e}^{j4\pi M_s\varepsilon} s_{kl}\eta_{k(l+1)}^*\right) \\ &\quad = \mathrm{e}^{j2\pi M_s\varepsilon}\sum_{l=1}^{L-2} 2\mathrm{Re}\left[\mathrm{e}^{j2\pi M_s\varepsilon} s_{kl}\eta_{k(l+1)}^*\right] \\ &\quad = \mathrm{e}^{j2\pi M_s\varepsilon}|\alpha_k|\varsigma_k \end{aligned} \tag{21}$$

where ς_k is a zero-mean real Gaussian RV with variance $\sigma_\varsigma^2 = 2(L-2)N_0E_\mathrm{c}[\sin^2(M_s\varepsilon\pi)/M_s^3\sin^2(\varepsilon\pi)]$. The third term can be modeled as $\alpha_k\kappa_k$, where κ_k is a zero-mean complex Gaussian RV with real and imaginary parts having the same variance $\sigma_\kappa^2 = N_0E_\mathrm{c}[\sin^2(M_s\varepsilon\pi)/M_s^3\sin^2(\varepsilon\pi)]$. Ignoring the last noise term of (19), the inner-slot combined correlation value $z_\mathrm{d}^{(k)}$ can be approximated as

$$z_\mathrm{d}^{(k)} \cong \mathrm{e}^{j2\pi M_s\varepsilon}\left[(L-1)|\alpha_k|^2\frac{\sin^2(M_s\varepsilon\pi)}{M_s^2\sin^2(\varepsilon\pi)}E_\mathrm{c} + |\alpha_k|\varsigma_k\right] + \alpha_k\kappa_k. \tag{22}$$

Thus, we have

$$\begin{aligned} z_\mathrm{d} &= \left|\sum_{k=1}^{N_s} z_\mathrm{d}^{(k)}\right|^2 = \left|\mathrm{e}^{-j2\pi M_s\varepsilon}\sum_{k=1}^{N_s} z_\mathrm{d}^{(k)}\right|^2 \\ &= \left|\left[(L-1)\gamma\frac{\sin^2(M_s\varepsilon\pi)}{M_s^2\sin^2(\varepsilon\pi)}E_\mathrm{c} + \sum_{k=1}^{N_s}|\alpha_k|\varsigma_k\right]\right. \\ &\quad \left. + \sum_{k=1}^{N_s}\mathrm{e}^{-j2\pi M_s\varepsilon}\alpha_k\kappa_k\right|^2 \end{aligned} \tag{23}$$

TABLE I
SYSTEM PARAMETERS

Item	Value
Carrier frequency	2 GHz
Chip rate	3.84 Mcps
Slot duration (N)	2560 chips
Code length (M)	256 chips
Number of tx. antenna (K)	1 or 2

where the last summation term is a complex RV representing the correlation between the signal and noise. Since the power of this term is negligible compared to the others, z_{d} can further be approximated as

$$z_{\mathrm{d}} \cong \left| \left[(L-1)\gamma \frac{\sin^2(M_s \varepsilon \pi)}{M_s^2 \sin^2(\varepsilon \pi)} E_{\mathrm{c}} + \sum_{k=1}^{N_s} |\alpha_k| \varsigma_k \right] \right.$$
$$\left. + \operatorname{Re}\left[\sum_{k=1}^{N_s} \mathrm{e}^{-j2\pi M_s \varepsilon} \alpha_k \kappa_k \right] \right|^2$$
$$= (\mu\gamma + \nu)^2 \tag{24}$$

where $\mu = (L-1)(\sin^2(M_s \varepsilon \pi)/M_s^2 \sin^2(\varepsilon \pi)) E_{\mathrm{c}}$ and ν denotes a zero-mean Gaussian RV with variance $\sigma_\nu^2 = \gamma(\sigma_\varsigma^2 + \sigma_\kappa^2)$. The conditional pdf of z_{d} can be expressed as a noncentral χ^2 distributed pdf with one degree of freedom

$$f_{H_1}(z|\gamma) = \frac{1}{\sqrt{2\pi z \gamma (\sigma_\varsigma^2 + \sigma_\kappa^2)}} \mathrm{e}^{-\frac{(z+\mu^2\gamma^2)}{2\gamma(\sigma_\varsigma^2+\sigma_\kappa^2)}} \cosh\left(\frac{\sqrt{z}\mu}{\sigma_\varsigma^2 + \sigma_\kappa^2} \right). \tag{25}$$

Averaging $f_{H_1}(z|\gamma)$ with respect to γ, we have

$$f_{H_1}(z) = \sum_{k=1}^{N_s} \frac{g_k \exp\left[-\left\{ \left(\frac{2}{\lambda_k} + \frac{\mu^2}{\sigma_\varsigma^2+\sigma_\kappa^2} \right) \frac{z}{\sigma_\varsigma^2+\sigma_\kappa^2} \right\}^{\frac{1}{2}} \right]}{\sqrt{\left\{ 2\lambda_k \left(\sigma_\varsigma^2 + \sigma_\kappa^2 \right) + \lambda_k^2 \mu^2 \right\} z}}$$
$$\times \cosh\left(\frac{\sqrt{z}\mu}{\sigma_\varsigma^2 + \sigma_\kappa^2} \right). \tag{26}$$

Substituting (18) and (26) into (7), we can obtain the detection probability of the proposed scheme. The performance analysis of the proposed scheme can be extended into a multipath environment using a method similar to [9].

IV. NUMERICAL RESULTS

To verify the synchronization performance of the proposed scheme, the detection probability is compared as a function of E_{c}/N_0 with the use of single and two antennas. The parameters for performance evaluation are summarized in Table I [7]. Fig. 3 depicts the detection probability in the flat Rayleigh fading channel with $f_{\mathrm{d}} = 10$ Hz when f_{o} is 10 and 20 kHz, and $N_s = 15$. It can be seen that the analytic results agree well with the simulation results and that the proposed scheme provides better performance than the conventional one particularly when E_{c}/N_0 is low. Note that the use of two antennas may not work better than the use of a single antenna when E_{c}/N_0 is low because the use of antenna diversity may not be appropriate in a noise-dominant environment. It can also be seen that the proposed scheme is quite effective in low E_{c}/N_0 even with the use of two antennas.

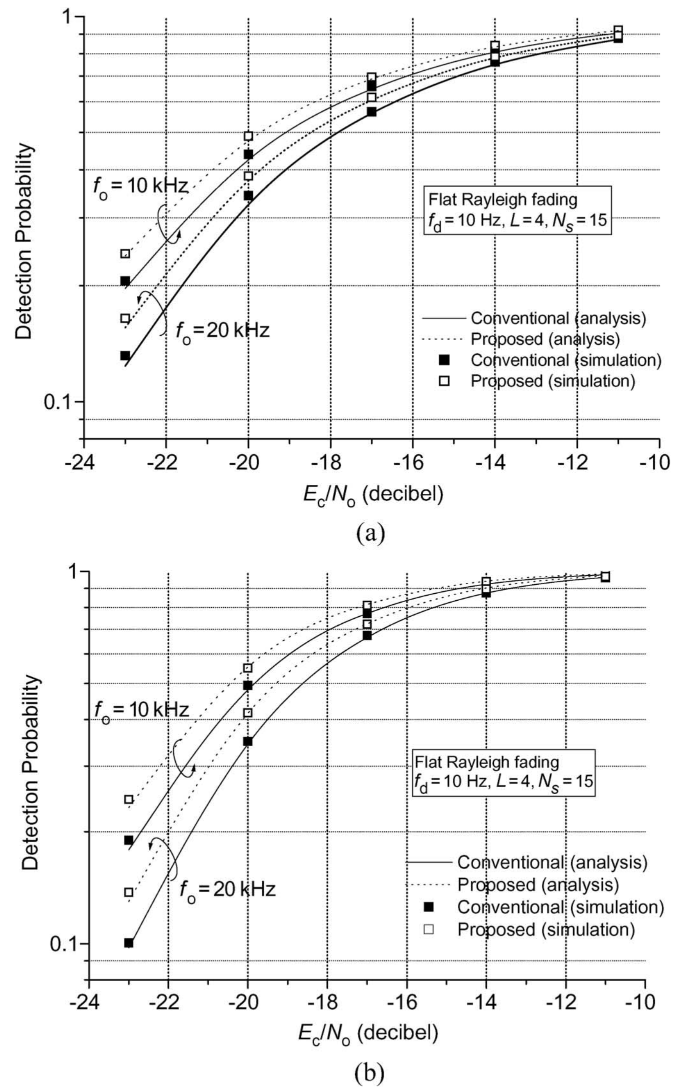

Fig. 3. Detection probability of the proposed scheme. (a) $K = 1$. (b) $K = 2$.

Fig. 4 depicts the detection performance for different values of L as a function of the initial frequency offset. It can be seen that the use of full correlation $(L = 1)$ is quite impractical when the initial frequency offset is large. The detection performance becomes robust to the frequency offset as L increases since incoherent loss can be mitigated by reducing M_s. Moreover, the performance gain with the use of the proposed scheme over the conventional one increases as L increases. This is mainly due to the fact that coherent combining with large LN_s makes the decision variable of the proposed scheme more reliable.

Fig. 5 depicts the effect of the Doppler frequency on the detection performance, when $L = 4$, $N_s = 15$, $f_{\mathrm{o}} = 10$ kHz, $E_{\mathrm{c}}/N_0 = -17$, -20 dB, and $K = 1$ and 2. It can be seen

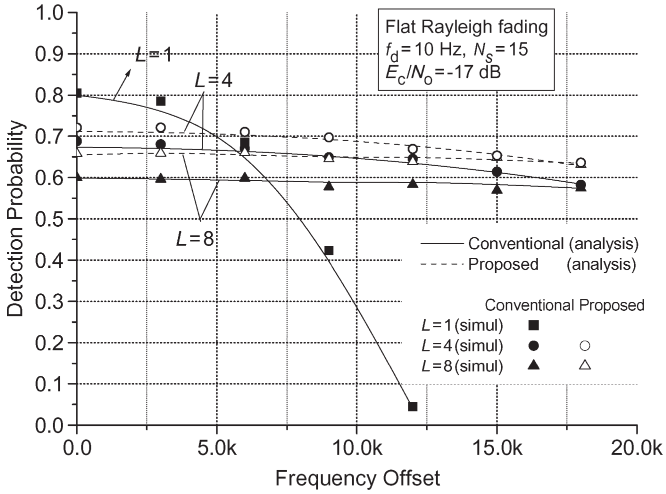

Fig. 4. Detection probability of the single-antenna scheme in the presence of initial frequency offset.

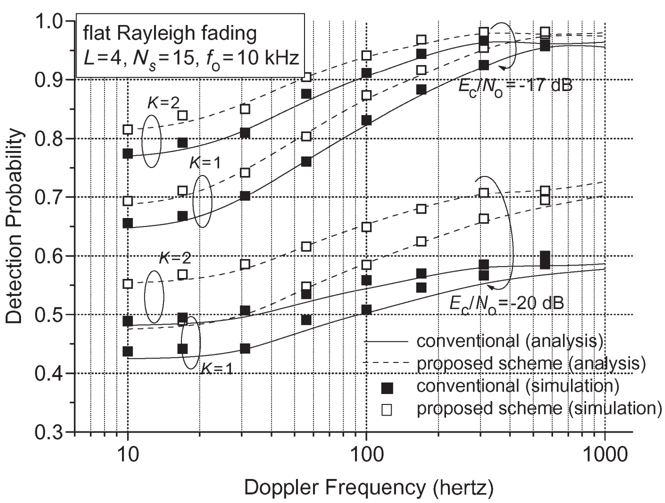

Fig. 5. Detection probability as a function of the maximum Doppler frequency.

that the detection performance is improved as the Doppler frequency increases due to the time diversity effect obtained from the combining process. It can also be seen that the proposed scheme outperforms the conventional one and that additional antenna diversity gain can also be obtained with the use of a TSTD scheme ($K = 2$). However, the performance difference between $K = 1$ and $K = 2$ becomes marginal as the Doppler frequency increases since the diversity gain is sufficiently obtained in the time domain. The combining process can obtain full time diversity gain when the slot interval is larger than the coherence time of the fading channel. As a result, the improvement of the detection performance becomes marginal for $f_{\mathrm{d}} > 750$ Hz.

V. Conclusion

To improve the cell search performance in the WCDMA system, we have proposed an inner-slot differential combining scheme that can be combined with the use of transmit antenna diversity. The performance of the proposed scheme has been analyzed and verified by computer simulation when multiple antennas are used in the presence of initial frequency offset and Doppler spread. Numerical results show that the proposed scheme can provide performance robust to a large amount of initial frequency offset and Doppler spread, outperforming conventional noncoherent combining schemes.

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