Performance Evaluation of Access Channel Slot Acquisition in Cellular DS/CDMA Reverse Link

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CONTENTS

ABSTRACT

In this paper, we consider the acquisition performance of an IS-95 reverse link access channel slot as a function of system design parameters such as postdetection integration length and the number of access channel message block repetitions. The uncertainty region of the reverse link spreading codes compared to that of forward link is very small, since the uncertainty region of the reverse link is determined by a cell radius. Thus, the parallel acquisition technique in the reverse link is more efficient than a serial acquisition technique in terms of implementation and of acquisition time. The parallel acquisition is achieved by a bank of N parallel I/Q noncoherent correlators. The output characteristics of an I/Q noncoherent correlator are analyzed for band-limited noise and the Rayleigh fast fading channel. The detection probability is derived for multiple correct code-phase offsets and multipath fading. The probability of no message error is derived when rake combining, access channel message block combining, and Viterbi decoding are applied. Numerical results provide the acquisition performance for system design parameters such as postdetection integration length and number of access channel message block repetitions in case of a random access on a mobile station.

I. INTRODUCTION

In recent years, direct sequence code division multiple access (DS/CDMA) has become more popular in digital cellular and wireless communication networks. There are several advantages that make DS/CDMA an attractive radio access technique. These advantages include the ability to combat multipath and allow multiple users to simultaneously communicate over a channel. However, the advantages of DS/CDMA can be exploited only if the pseudonoise (PN) code sequence of a receiver and an incoming signal are synchronized within a fraction of a PN chip. Typically, the process of synchronization is done in two steps, acquisition (coarse alignment) and tracking (fine alignment) [1], [2]. This paper is mainly focused on the acquisition of the reverse link in the IS-95 CDMA digital cellular system. In the reverse link, since a pilot channel is not utilized in order to conserve mobile station power, noncoherent demodulation is used for the voice/data traffic channel and access channel [3]. In consideration of hardware complexity, I/Q noncoherent correlators are used for acquisition operations as well.

Typically, cells in digital cellular and wireless communication networks are classified roughly according to cell radius as macrocells and microcells. The maximum cell radiuses of macrocells and microcells are 20 km and 1 km, respectively. In case of macrocells, since the propagation delay is about 3μs/km, the maximum round trip delay corresponds to about 120 μs. When the period of a PN chip, T_c, is 1 μs, the uncertainty region of reverse link is about 120 PN chips. Thus, compared to the uncertainty region of a forward link which is the entire code period (32768 PN chips), that of the reverse link is very small. In case of a small uncertainty region, parallel acquisition is more efficient than serial acquisition in terms of implementation and of acquisition time. Also, a single-dwell detection scheme is known to be more efficient than a multiple-dwell detection scheme over the initial acquisition of a reverse link access channel slot [4].

A parallel acquisition technique that employs a bank of N parallel I/Q noncoherent matched filters (MFs) is discussed in [5], [6]. In this paper, we consider correlators rather than matched filters. The output characteristic of an I/Q noncoherent correlator is analyzed for the band-limited and Rayleigh fast fading channel. The independent fading of all users is assumed. The detection probability is derived for multiple H_1 hypotheses and multipath fading [4]. The probability of no message error is derived when rake combining, access channel message block combining, and Viterbi decoding are applied. Based on these evaluations, suboptimum values of system design parameters such as postdetection integration length (noncoherent summations) and the number of access channel message block repetitions are chosen in terms of mean acquisition time and mean slot acquisition

time. These values are also chosen for cases without antenna diversity and with antenna diversity.

This paper is organized as follows. The procedure of random access and slot acquisition is described in Section II. The output characteristic of an I/Q noncoherent correlator, mean slot acquisition time, and required probability are analyzed in Section III. Numerical results are presented in Section IV, and finally, a concluding discussion is presented in Section V.

II. SYSTEM DESCRIPTION

The DS/CDMA reverse link consists of the access and reverse traffic channels [7]. The access channel is used for a random access, while the reverse traffic channel is used for communicating user traffic data. When a mobile station attempts to access a network, an access channel slot (pseudorandomly selected transmission time interval) is utilized. Transmissions during an access channel slot consist an access channel preamble unmodulated by data and an access channel message capsule consisting of access messages encoded for error detection with a cyclic redundancy code (CRC). In order to analyze the acquisition performance of access channel slot transmissions, the following assumptions are made :

1) The access channel transmission uses 9600 bps mode.
2) An access channel message block is defined as the result that an access channel message is convolutionally encoded.
3) An access channel message capsule consists of several access channel message blocks.

The random access procedure is briefly described. The transmission timing of an access channel slot is determined pseudorandomly. If the initial acquisition is achieved during a given access channel preamble, the base station proceeds with noncoherent demodulation of M-ary orthogonal signals, Viterbi decoding, and CRC checking over the succeeding access channel message capsule. If no error is detected by CRC checking, the access attempt succeeds tentatively, and the base station sends an acknowledgment to the mobile station. If an acknowledgment is correctly received, an access attempt terminates. But, if errors are detected by CRC checking, the access attempt fails, and the base station does not send an acknowledgment to the mobile station. In case of an access attempt failure, the access channel slot is pseudorandomly repeated at the base of the system time, namely, it is retransmitted a random delay whose average corresponds to several slots.

In the reverse link, since the transmitted signal of a mobile station is always spread by the PN code sequence with zero codephase offset, the uncertainty region is determined by the round-trip delay due to the cell radius. Thus, the uncertainty region, V, can be represented as W/Δ, where

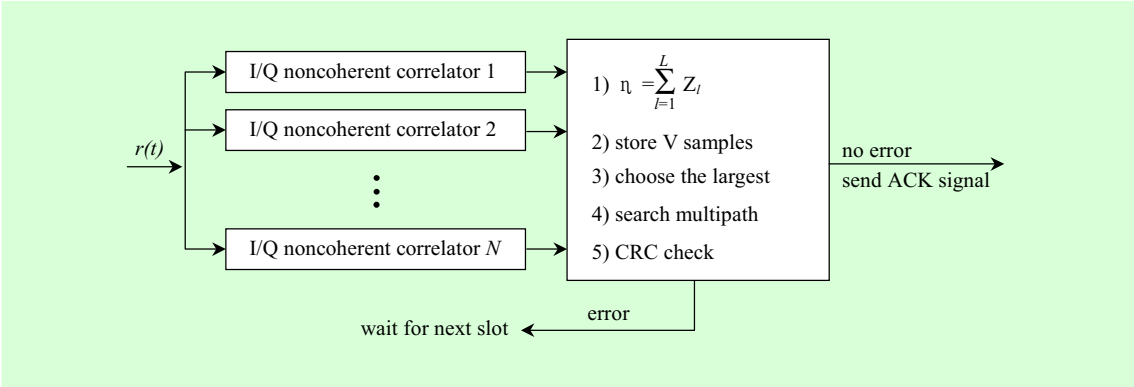

Fig. 10. System structure.

W and Δ are the number of PN chips for round-trip delay and search step size, respectively.

The acquisition system under consideration in this paper is a parallel acquisition using a single-dwell detection scheme. Since the parallel acquisition is done by a bank of N parallel I/Q noncoherent correlators as shown in Fig. 1, the number of total test hypotheses for each I/Q noncoherent correlator is given by V/N. The parallel acquisition algorithm using single-dwell detection is as follows : Defining the sample as the sum η of I/Q noncoherent correlator outputs for the postdetection integration (noncoherent summations) length L as shown in Fig. 1, the samples corresponding to all test hypotheses are compared. Finally, this scheme only chooses the test hypothesis (code-phase offset) that corresponds to the largest one among V samples. After completing this procedure, the multipath acquisition is done over $\pm W_p$ chips range centered on the code-phase that is already selected. W_p is $\lfloor T_m/T_c \rfloor + 1$, where T_m is the delay spread of a channel, T_c is the period of a PN chip, and $\lfloor \cdot \rfloor$ denotes the integer part. This selects the other code-phase offset that corresponds to the largest sample except the code-phase offset that is already selected. Thus, the initial acquisition is terminated by finishing the mulitpath search.

III. ACQUISITION ANALYSIS

1. Output Characteristic of I/Q Noncoherent Correlator

In order to evaluate the acquisition performance, we consider the discrete multipath Rayleigh fading channel [10]. The discrete multipath intensity profile $\rho(\tau)$ of a fading channel as a function of the time delay τ can be modeled by

$$\rho(\tau) = \sum_{i=1}^{J} E[\alpha_i^2]\delta(\tau - \tau_i), \qquad (1)$$

where α_i and τ_i are the path gain and time delay of the i-th path, respectively. The path gains are independent, identically distributed, Rayleigh random variables with mean square values $E[\alpha_i^2]$. The number of multipath components, J, is $\lfloor T_m/T_c \rfloor + 1$. In IS-95 [7], the transmitted access channel preamble of the k-th user can be represented as

$$S^{(k)}(t) = \sqrt{E_c^{(k)}} \cos\omega_c(t) \sum_n w_{0,n}^{(k)} a_{I,n}^{(k)} h(t - nT_c) + \sqrt{E_c^{(k)}} \sin\omega_c(t) \sum_n w_{0,n}^{(k)} a_{Q,n}^{(k)} h(t - nT_c), \quad (2)$$

where $a_{I,n}^{(k)}$, $a_{Q,n}^{(k)}$ are the I and Q PN sequences, respectively, $w_{0,n}^{(k)}$ is the Walsh index 0 sequence, $E_c^{(k)}$ is the PN chip energy, and $h(t)$ is the impulse response of a band-limited filter.

In the presence of noise with multipath and of other users from the same and other cells in the reverse link, the received signal in cellular DS/CDMA reverse link can be written as

$$\begin{aligned} r(t) = & \sum_{i=1}^{N_u} \sum_{j=1}^{J} \alpha_j^{(i)} \sqrt{E_C^{(i)}} \Big[\cos(\omega_c t - \varphi_j^{(i)}) \\ & \cdot \sum_n w_{s,n}^{(i)} a_{I,n}^{(i)} h(t - nT_c - \tau_j^{(i)}) + \sin(\omega_c t - \varphi_j^{(i)}) \\ & \cdot \sum_n w_{s,n}^{(i)} a_{Q,n}^{(i)} h(t - nT_c - \tau_j^{(i)}) \Big] \\ & + \sum_{m=1}^{N_c} \sum_{i=1}^{N_u} \sum_{j=1}^{J} \alpha_j^{(mi)} \sqrt{E_c^{(mi)}} \Big[\cos(\omega_c t - \varphi_j^{(mi)}) \\ & \cdot \sum_n w_{s,n}^{(mi)} a_{I,n}^{(mi)} h(t - nT_c - \tau_j^{(mi)}) + \sin(\omega_c t - \varphi_j^{(mi)}) \\ & \cdot \sum_n w_{s,n}^{(mi)} a_{Q,n}^{(mi)} h(t - nT_c - \tau_j^{(mi)}) \Big] + n(t), \quad (3) \end{aligned}$$

where N_u and N_c are the number of users and the number of other cells, respectively, $\varphi_j^{(i)}$ is the random phase of the j-th path for the i-th user, $w_{s,n}^{(i)}$ is the Walsh index s sequence of the i-th user, and $n(t)$ is an additive white Gaussian random process with zero mean and two-sided power spectral density $N_0/2$. Also, the subscript m denotes the other cell component.

Referring to (3) and I/Q noncoherent correlator of Fig. 2, Y_I and Y_Q for the h-th path of the k-th user in the home cell can be represented as

$$Y_I = N_w \sqrt{E_c^{(k)}} \sum_{j=1}^{J} \alpha_j^{(k)} R(\lambda_h^{(j)}) \cos\varphi_j^{(k)} + \sum_{n=1}^{N_w} n_{I,n} \quad (4)$$

and

$$Y_Q = N_w \sqrt{E_c^{(k)}} \sum_{j=1}^{J} \alpha_j^{(k)} R(\lambda_h^{(j)}) \sin\varphi_j^{(k)} + \sum_{n=1}^{N_w} n_{Q,n}, \quad (5)$$

where $R(\lambda_h^{(j)}) = \int |H(f)|^2 \cos(2\pi f \lambda_h^{(j)}) df$, $\lambda_h^{(j)}$ is the timing error between the j-th path and PN code sequence of local PN code generator corresponding to the h-th path, N_w is the number of PN chips for a Walsh symbol duration.

The variance consists of the contribution from background noise, multipath interference, multiple access interference, and other cell interference. From (4) and (5), the variance of Y_I and Y_Q is given by

$$Var[Y_I] = Var[Y_Q] = \sigma_1^2, \quad (6)$$

where σ_1^2 is the interference variance for the h-path of the k-th user.

To evaluate the interference variance on the reverse link, we assume that the number of users is the same in all cells and users

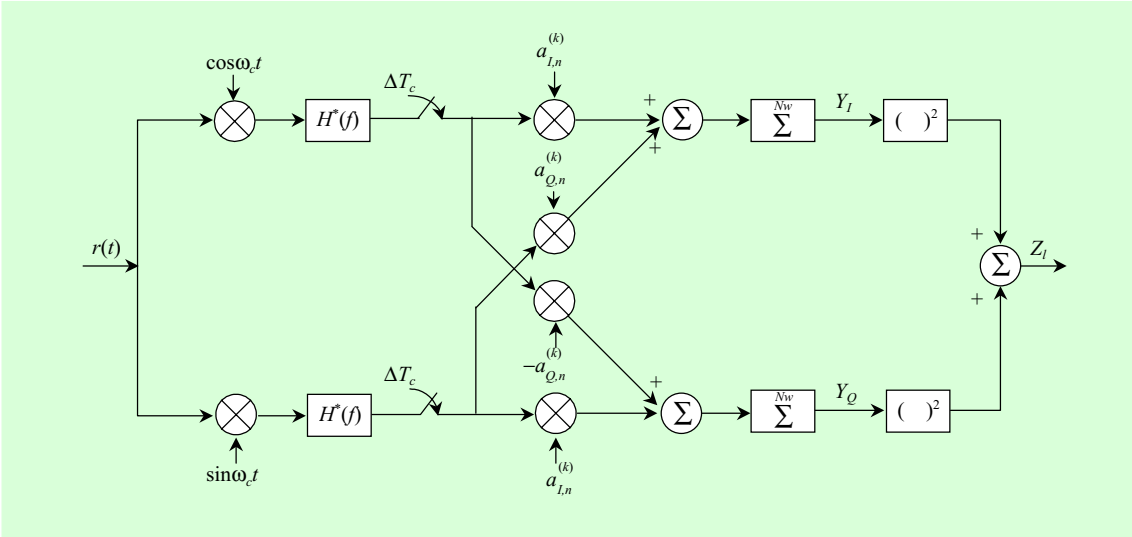

Fig. 11. I/Q noncoherent correlator.

are uniformly distributed. Also, the fading of all users in all cells is assumed to be independent. In general, the interference described above is treated as an additive Gaussian noise. In this case, under hypothesis H_1 corresponding to the h-th path of the k-th user, the interference variance σ_1^2 can be represented as

$$\sigma_1^2 = \frac{N_w}{2}\Bigg[N_0 + E_c^{(k)} E\{(\alpha_h^{(k)})^2\} \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} [R(nT_c + \lambda_h^{(h)})]^2$$
$$+E_c^{(k)} \sum_{\substack{j=1 \\ j\neq h}}^{J} E\{(\alpha_j^{(k)})^2\} \sum_{n=-\infty}^{\infty} [R(nT_c + \lambda_h^{(j)})]^2$$
$$+\sum_{\substack{i=1 \\ i\neq k}}^{N_u} E_c^{(i)} \sum_{j=1}^{J} E\{(\alpha_j^{(i)})^2\} \sum_{n=-\infty}^{\infty} [R(nT_c + \lambda_h^{(ij)})]^2$$
$$+R_f \sum_{i=1}^{N_u} E_c^{(i)} \sum_{j=1}^{J} E\{(\alpha_j^{(i)})^2\} \sum_{n=-\infty}^{\infty} [R(nT_c + \lambda_h^{(ij)})]^2\Bigg] \quad (7)$$

where $\lambda_h^{(j)}$ is the timing difference between the h-th path and the j-th path of the k-th user, $\lambda_h^{(ij)}$ is the timing difference between the h-th path of the the k-th user and the j-th path of the i-th user, and R_f is the ratio of total other cell user interference to home cell other user interference. Typical value for R_f is 0.55 [8]. In (7), the first term is background noise component, the second term is the interchip interference component, the third term is the multipath interference component, the fourth and fifth terms are the other user interference components of home cell and of other cells, respectively.

Then, if we assume that the number of users is the same in all cells and users are uniformly distributed, the interference variance σ_1^2 for the h-th path of the k-th user can be rewritten as

$$\sigma_1^2 = \frac{N_w}{2}\Bigg[N_0 + E_c E\{\alpha_h^2\} \sum_{\substack{n=-\infty \\ n\neq 0}}^{\infty} [R(nT_c + \lambda_h^{(h)})]^2$$

$$+E_c\sum_{\substack{j=1\\ j\neq h}}^{J}E\{\alpha_j^2\}\sum_{n=-\infty}^{\infty}[R(nT_c+\lambda_h^{(j)})]^2$$

$$+\{(N_u-1)+R_fN_u\}E_c\sum_{j=1}^{J}E\{\alpha_j^2\}$$

$$\cdot\sum_{n=-\infty}^{\infty}[R(nT_c)]^2\Bigg], \tag{8}$$

where σ_1^2 is also based on an implicit assumption that the phase $\varphi_j^{(k)}$ in (4) and (5) remains constant over the N_w PN chips accumulated.

Also, the variance σ_0^2 of Y_I and Y_Q under hypothesis H_0 (incorrect acquisition case) can be represented as

$$\sigma_0^2=\frac{N_w}{2}\Bigg[N_0+N_u(1+R_f)E_c\sum_{j=1}^{J}E\{\alpha_j^2\}$$

$$\cdot\sum_{n=-\infty}^{\infty}[R(nT_c)]^2\Bigg]. \tag{9}$$

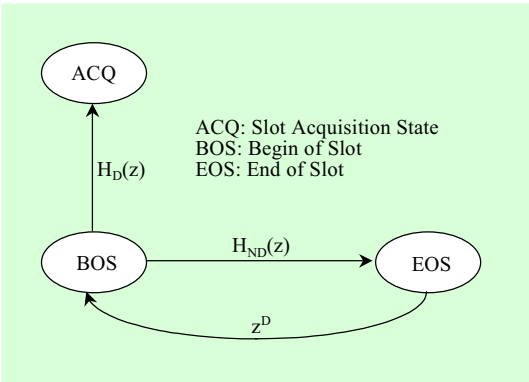

Fig. 12. State diagram of the access channel slot acquisition.

2. Mean Slot Acquisition Time

State diagram is used to derive the probability generating function of acquisition time owing to Markovian nature of the acquisition process. The state diagram of access channel slot acquisition described in Section II is shown in Fig. 3. $H_D(z)$ and $H_{ND}(z)$ of Fig. 3 are given by

$$H_D(z)=P_DP_Cz^S \tag{10}$$

and

$$H_{HD}(z)=(1-P_DP_C)z^S, \tag{11}$$

where P_D and P_C are the detection probability and probability of no message error, respectively. S is the size of access channel slot comprised of access channel preamble and access channel message capsule. From Fig. 3, the generating function is

$$U(z)=\frac{H_D(z)}{1-z^DH_{ND}(z)}, \tag{12}$$

where D is the average random-delay. Thus, the mean slot acquisition time, $E[T_{ACQ}]$, is given by

$$E[T_{ACQ}]=\frac{d}{dz}\ln U(z)\Big|_{z=1}$$
$$=\frac{S+D(1-P_DP_C)}{P_DP_C}. \tag{13}$$

3. Detection Probability and Probability of No Message Error

To facilitate the derivation of detection probability, we made the following assumptions : 1) all test statistics are independent, 2) there are IH_1 cells in the uncertainty region, 3) all test cells are a priori equally likely, and 4) the output characteristic of an I/Q noncoherent correlator is analyzed for the fast fading model

with Rayleigh-distributed envelope statistics. If the decision variable η is the sum of I/Q noncoherent correlator outputs Z_l, $l = 1, 2, \cdots, L$, which are mutually independent, η under the hypothesis H_0 is central chi-square distributed with $2L$ degrees of freedom. Hence, the probability density function of η under the hypothesis H_0 is given by

$$f_\eta(\eta|H_0) = \frac{1}{(L-1)!V_N^L}\eta^{L-1}e^{\frac{-\eta}{V_N}}, \qquad (14)$$

where $V_N = 2\sigma_0^2 = N_w I_0$, and I_0 is the interference density under hypothesis H_0.

From (4) and (5), the I/Q noncoherent correlator output means V_{F_i}, $i = 1, 2, \cdots, I$ under the hypothesis H_1 are given by

$$V_{F_i} = N_w^2 E_c \sum_{j=1}^{J} E\{\alpha_j^2\}[R(\lambda_i^{(j)}]^2 + 2\sigma_1^2,$$
$$i = 1, 2, \cdots, I, \qquad (15)$$

where $R(\lambda_i^{(j)}) = \int |H(f)|^2 \cos(2\pi f \lambda_i^{(j)})df$, α_j is the path gain of the j-th path, $\lambda_i^{(j)}$ is the timing error between the j-th path and code sequence of local PN code generator corresponding to the H_{1i} cell, and $N_w I_0 = 2\sigma_1^2$. Using (15), the probability density function of η under the hypothesis H_1 is given by

$$f_\eta(\eta|H_{1i}) = \frac{1}{(L-1)!V_{F_i}^L}\eta^{L-1}e^{\frac{-\eta}{V_{F_i}}},$$
$$i = 1, 2, \cdots, I. \qquad (16)$$

Using the probability density functions of (14) and (16) above, the detection probability can be derived. Since the detection probability $P_{D_{1i}}$ corresponding to the H_{1i} cell is the probability that the H_{1i} is larger than (V-1) cells, the detection probability $P_{D_{1i}}$ is represented as

$$P_{D_{1i}} = \int_0^\infty f_\eta(\eta|H_{1i}) \left[\int_0^\eta f_x(x|H_0)dx\right]^{(V-1)}$$
$$\cdot \prod_{\substack{l=1 \\ l\neq i}}^{I} \left[\int_0^\eta f_y(y|H_{1l})dy\right] d\eta. \qquad (17)$$

Since there are IH_1 cells in the uncertainty region, the detection probability P_D is given by

$$P_D = \sum_{i=1}^{I} P_{D_{1i}}. \qquad (18)$$

The probability of no message error P_C is the probability that no error is detected within an access channel message block, B_L. After CRC checking, it is given by [11]

$$P_C = (1 - P_B)^{B_L}, \qquad (19)$$

where P_B is the bit error probability after the hard-decision Viterbi decoding. Then, P_B is upper bounded by

$$P_B \leq \frac{1}{2}\frac{dT(D,N)}{dN}\Big|_{N=1, D=\sqrt{4\varepsilon(1-\varepsilon)}}, \qquad (20)$$

where ε is the bit error probability of M-ary orthogonal signals for the multipath Rayleigh fading channel. We assume that a received access channel message block experiences independent fading. Under these fading environments, the symbol error probability of M-ary orthogonal signals for $T(\leq J)$ path combining and N_r access channel message block combining is given by [10]

$$P_M = 1 - \int_0^\infty \frac{z^{R-1}e^{-z/(1+\overline{r}_c)}}{(R-1)!(1+\overline{r}_c)^R}$$
$$\cdot \left[1 - e^{-z/V_N}\sum_{k=0}^{R-1}\frac{(z/V_N)^k}{k!}\right]^{M-1} dz, \qquad (21)$$

where $R = TN_r$, $\overline{r}_c = \frac{1}{R}\sum_{j=1}^{R} E[\alpha_j^2]N_w E_c$.
Thus, the bit error probability of M-ary orthogonal signals is given by

$$\varepsilon = \frac{M/2}{M-1} P_M. \tag{22}$$

IV. NUMERICAL RESULTS

To evaluate the acquisition performance, we consider the following parameters : the period of a PN chip T_c=813.8 ns, number of PN chips per Walsh symbol duration N_w=256, number of I/Q noncoherent correlators N=6, number of rake combining paths T=4, total search window length W=148 PN chips (cell radius = 20 km), multipath search region W_p=6 PN chips, search step size Δ=1/2, reset time T_r=1.25 ms, average random-delay D=760 ms, number of Walsh functions M=64, access channel message block B_L=552 bits.

The average path powers $E[\alpha_i^2]$, $i = 1,2,\cdots,6$ of the discrete multipath intensity profile considered in this work are given by 0.5562, 0.2493, 0.1118, 0.0501, and 0.0101, respectively. In order to calculate the bit error probability P_B given in (18), $\frac{dT(D,N)}{dN}\Big|_{N=1}$ of the best (3,1,9) convolutional code in (20) is given by [9]

$$\begin{aligned}\frac{dT(D,N)}{dN}\Big|_{N=1} = {} & 11D^{18} + 32D^{20} + 195D^{22} \\ & + 564D^{24} + 1473D^{26} + 5129D^{28} \\ & + 17434D^{30} + 54092D^{32} \\ & + 17117D^{34} + \cdots \end{aligned} \tag{23}$$

In this paper, we intend to choose the suboptimum system design parameters of postdetection integration length L and the number of access channel message block repetitions N_r in terms of acquisition time and its sensitivity.

From (13), we can calculate the mean acquisition time when $P_C = 1$. Figures 4 and 5 show the mean acquisition times in terms of E_s/I_0 in various values of L for both cases without antenna diversity and with antenna diversity, respectively. Here, E_s is $N_w E_c$ and I_0 is the interference spectral density that is represented as $2\sigma_1^2/N_w$ under hypothesis H_1 and $2\sigma_0^2/N_w$ under hypothesis H_0, respectively. In view of the mean acquisition time and its sensitivity to changes in E_S/I_0, better performances for both cases are achieved when L=3 and L=2, respectively.

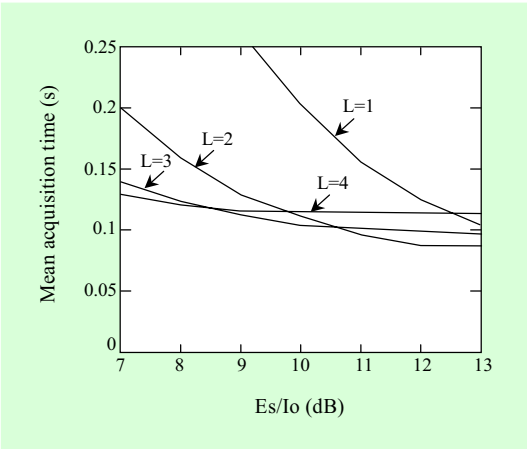

Fig. 13. Mean acquisition time for no antenna diversity.

Figures 6 and 7 show the mean slot acquisition time versus E_s/I_0 for the case

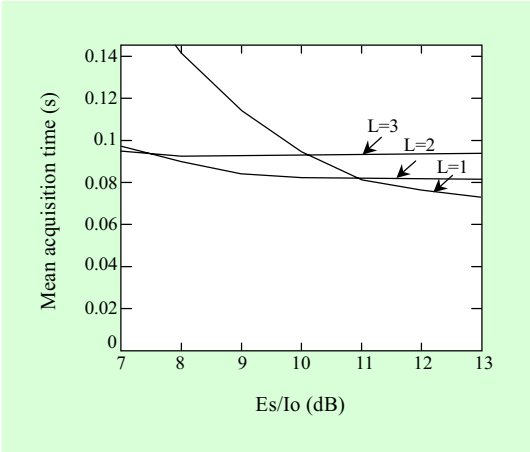

Fig. 14. Mean acquisition time for antenna diversity.

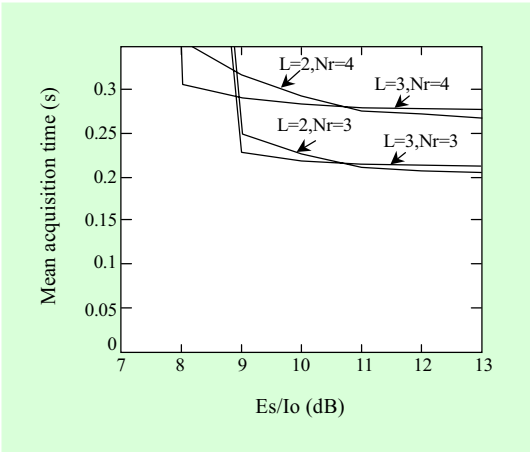

Fig. 15. Mean slot acquisition time for no antenna diversity.

without antenna diversity and the case with antenna diversity. From the results of two figures, the suboptimum choice of the case without antenna diversity is L=3 and N_r=4, while in the case with antenna diversity, the suboptimum system design parameters are L=2 and N_r=2. From these suboptimum system design parameters, the

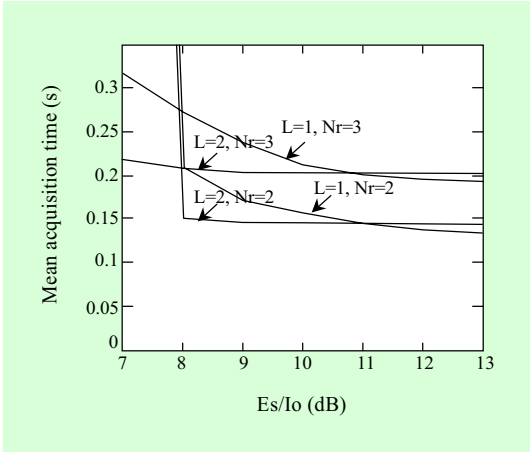

Fig. 16. Mean slot acquisition time for antenna diversity.

mean acquisition times of both cases are about 0.1 s and 0.08 s, respectively, while the mean slot acquisition times are about 0.3 s and 0.15 s, respectively. Hence, the case with antenna diversity provides two times shorter mean slot acquisition time than the case without antenna diversity. If we also see the sensitivity of mean slot acquisition time with respect to L and N_r, its sensitivity to changes in L is small more or less, but that to changes in N_r is very large. Hence, the mean slot acquisition time is influenced by the probability of no message error more than the detection probability.

V. CONCLUSION

The performance of an access channel slot acquisition scheme has been evaluated in terms of mean acquisition time and mean

slot acquisition time, where the parallel acquisition using the single-dwell detection scheme is applied. Comparing the mean slot acquisition time of both cases without antenna diversity and with antenna diversity, the case with antenna diversity can achieve two times shorter acquisition time than the case without antenna diversity. The suboptimum system design parameters such as L and N_r have also been suggested for both cases without antenna diversity and with antenna diversity, respectively. In view of the mean slot acquisition time and its sensitivity to changes in E_s/I_0, the case with antenna diversity performs better when L=3 and N_r=4, while the case with antenna diversity does better when L=2 and N_r=2.

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