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PN Code Acquisition for the N-path RAKE Receiver in the DS/CDMA Systems

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DS/CDMA 이동통신 시스템의 N-경로 RAKE 수신기를 위한 PN 코드동기

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ABSTRACT

This paper presents acquisition algorithm and an improved detection technique for the DS/CDMA (direct sequence code division multiple access) RAKE receiver in a multipath fading channel. The DS/CDMA systems use the RAKE receiver to overcome signal fading due to multipath propagation. To maximize the performance improvement of the RAKE receiver, an accurate code acquisition is required for the RAKE branches. The algorithm is able to find the pseudonoise (PN) code delay estimates for the RAKE branches in a multipath fading channel.

In this paper a numerical method and computer simulation have been developed for the acquisition system. The detection probability and mean acquisition time are investigated as a performance measure of the system using the Monte Carlo method. And also in order to analyze the effect of the acquisition on the RAKE receiver this paper brings out the effect of integration time, doppler frequency, processing gain and the number of users on the acquired code phase.

요 약

본 논문은 주파수 선택적 이동통신 채널에서 RAKE 수신기를 위한 IS-95 DS/CDMA 시스템의 코드동기 및 개선된 검파기법을 제안한다. DS/CDMA 시스템은 다경로 페이딩 성분을 극복하기 위하여 RAKE 수신기를 사용한다. RAKE 수신기의 성능을 극대화하기 위하여 각 RAKE 브랜치에 정확한 PN 코드 오프셋을 제공하여야 한다. 본 논문의 코드동기 알고리듬은 다경로 페이딩 채널에서 RAKE 브랜치를 위한 PN 코드의 오프셋을 추정하

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는 알고리듬이다. Monte Carlo 방법을 적용하여 실험을 행하였고, 동기시스템의 성능을 평가하기 위하여 평가척도로 검파확률 및 평균동기시간을 사용하였다. 그리고 적분시간, 도플러 주파수, 처리이득, 사용자수가 동기시스템에 끼치는 영향을 연구 분석하였다.

I. Introduction

The mobile radio channel is characterized by the time variant multipath propagation and fading. Direct sequence spread spectrum techniques are being strongly considered for use in cellular and microcellular communication systems because of their well-known ability to combat multipath fading, interference, and to allow multiple users to simultaneously communicate, over a mobile radio fading channel, with higher capacity[1].

However, the benefit of direct sequence spread spectrum system is based on the accurate synchronization. The synchronization process generally consists of two parts: acquisition and tracking. Acquisition is the process whereby the received code sequence and the locally generated replica of the code sequence are coarsely aligned, usually to within half a chip duration. Following successful acquisition, a code tracking loop is used to align the two code sequence more accurately, and this is known as tracking [2, 3]. We focus attention on the first part and consider the acquisition process for chip synchronous DS/CDMA systems where the unknown delay of the incomming PN sequence is T_c. Here, is an integer, random variable and Tc is the chip duration.

The acquisition schemes depend on both the search strategy and the detector structure. It is well known that fast acquisition can be realized by employing charge coupled devices(CCD), surface acoustic wave (SAW) convolvers, or other devices to construct PN matched filter(PNMF). The problems of acquisition using the matched filter have attracted considerable interest in recent years. Due to the difficulty of phase-tracking a wideband low spectral- density signal, noncoherent correlation techniques have been widely adopted in the past[2, 3]. In particular, the noncoherent inphase/quadrature-phase(I-Q) matched filter correlation detector, which combines the squares of coherent correlations at both I and Q branches, has received considerable attention because of its fast acquisition capability. But IS-95 DS/CDMA system uses correlator to achieve the code acquisition. In this paper, we employed the PN matched filter to speed up the code acquisition and present the performance analysis. And also, many results were derived assuming a single user in a fading multipath delay mobile environment or multi-users in the additive white gaussian noise channel. But this paper have been done in multipath delay and fading conditions in a multi-user environment.

The search technique is the maximum-likelihood in a serial fashion. Therefore, the input PN signal is serially correlated with all possible code positions of the receiver PNMF using sliding correlation of window and the corresponding detector outputs are stored. At the end of the test, the correct PN alignment is chosen as that local code phase position which produced the maximum output and verification mode is initiated. The acquisition algorithm has two separate modes(stages): the search mode(1st stage), and the verification mode(2nd stage). In the search mode, a tentative decision on the delay of the received signal is made. In the verification mode, a more accurate decision on the delay of the received signal, assumed by the search mode is achieved. This avoids unnecessary false alarm.

In general case of multipath channel model when N-path RAKE receiver, with fixed time delays between adjacent demodulation branches, is used[4] the acquisition procedure is supposed to initially determine delays of the N signal components. The practical approach to the problem depends on the specific values of the delays and dopplers in each of the N paths. Acquisition is based on the received magnitude of the impulse response of the mobile channel and sets the delay times of the RAKE branches. The objective of this paper is to find the accurate code delay offset for the RAKE branches using sliding window and provide an improved detection technique using the postdetection integration. The threshold of the verification mode is determined in the search mode by the proposed threshold control algorithm. The threshold can find a desired threshold value such that the probability of the false alarm is minimized and the probability of the correct detection is maxmized.

The paper is organized as follows. The total system model is described in section Ⅱ. Section Ⅲ provides the description of the acquisition scheme. Section Ⅳ contains the acquisition performance results. Finally, conclusions drawn from this investigation are enumrerated.

Ⅱ. System model

The service area is assumed to be partitioned into cells, each cell having a base station and numerous mobile units. Each cell has both the forward and reverse links operating in different frequency bands. The analysis is relevant to reverse link, assuming that perfect power control is present.

A. Multipath Propagation Channel

The spread-spectrum signal at the output of the mobile transmitter is asynchronously transmitted (uplink) through the channel. This signal propagates along various paths (multipath) so that at the receiver a constructive or destructive superposition of the attenuated and time-delayed replicas of the transmitted signal (depending on the their relative phases) is produced. The multipath channel modelled in this investigation is based on the the fact that the complex lowpass impulse response of the multipath channel can be expressed as:

$$h(t;\tau)=\sum_{l}\alpha_l\delta(\tau-\tau_l(t))\exp(j\phi_l(t)) \qquad (1)$$

where α_l and ϕ_l are respectively the magnitude and phase of the l-th resolvable path, τ_l is its propagation time delay and $\delta()$ is the Dirac delta function specifying a unit impulse. In the fading channel considered here, the fading process is regarded as a constant over a single chip period. The block diagram of the Rayleigh fading channel is shown in Fig. 1. The system includes K users and each user transmits through L multiple paths and each path has a propagation delay τ and attenuation factor. The arrow denotes complex signal flow.

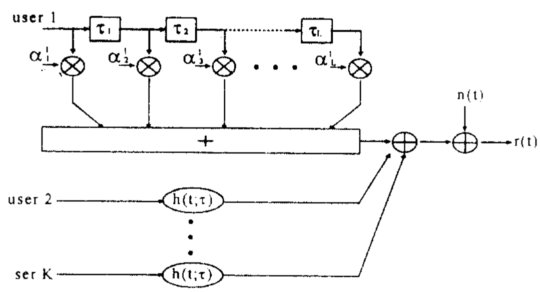


Fig 1. The multipath delay and multi-user channel model

B. Receiver Structure

We consider quaternary direct sequence systems where the same data is modulated onto the in-phase and quadrature channels using different PN codes. The channel is a multipath fading with L resolvable paths. The determination of the maximum number of the resolvable paths between the transmitter and the receiver depends on the time resolution of the direct sequence signal which is given by the chip time T_c. This implies that two versions of a spread spectrum signal must be separated in time by T_c seconds if their correlation peaks are to be unambiguously distinguished[6]. Therefore, if the transmission chip rate

$1/T_c$ exceeds the reciprocal of the delay spread T_m of the channel, the mutipath components can be resolved into a number of discrete fading paths L, given by

$$L \leq \lfloor T_m/T_c \rfloor + 1 = \lfloor T_m W \rfloor + 1 \quad (2)$$

where $\lfloor x \rfloor$ refers to the integer part of x and W is the bandwidth of the spread spectrum signal. Eqn. (2) shows that a maximum of L statistically independent paths can be resolved directly from the channel and combined so as to maximize the total wanted signal upon which to make a decision. But the RAKE receiver uses only N paths among L paths for the demodulation. The receiver has two parallel arms: the PN code synchronization arm and the data demodulating arm. The receiver structure is shown in Figure 2.

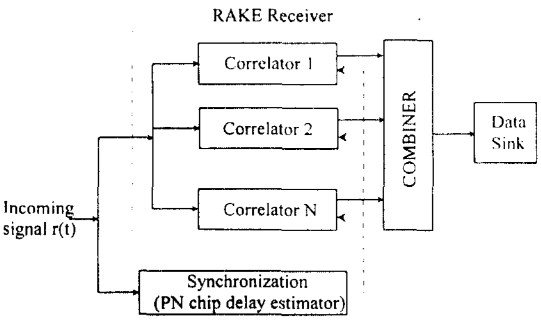


Fig. 2. Simplified Block diagram of the RAKE receiver with synchronization

The synchronization device scans the received power of the impulse responses on all possible paths in the delay of the received signal. After this process is completed, it sets the delay times of the branches corresponding to the N paths.

Ⅲ. Acquisition System Description

In CDMA system, the acquisition problem of the downlink is trivial, since there is pilot signal which is transmitted at the base station. The uplink, however, is much more complicated, because multiusers simultaneously transmit to a base station with a different propagation delay and random delay computed by a hash function. Therefore, we focus on the design of fast acquisition algorithms for the uplink. The random delay plus the propagation delay forms the delays of a message received at the base station.

In order to make the signals synchronous at the antenna of the base station, it is necessary to know two variables, that are generally different for all users: First, the clock offsets between users and base station and second, the propagation delay or user-base distance. Once these variables are known, they can be corrected by applying a delay to the transmitted codes that is opposite to the propagation delay plus the clock offset between base station and user.

The propagation delay is a random variable since the geometric position of mobiles in a cell site are randomly distributed. The delay value of the received signal is assumed to be uniformly distributed over D chips. The random variable D is determined by a chip rate and cell radius. Therefore, acquisition process tests all possible delay positions(also called "code phases") and look for the maximum received energy at delay position $\tau = \delta T_c$, where the index δ is in [0, D]. The code acquisition in the reverse link is accomplished by the preamble of the access channel. An access channel slot is the maximum message capsule size plus the preamble size rounded up to an integer number of access channel frames in length (20ms)[7]. The preamble frames are chosen to ensure that the receiving base station has enough time to perform correlation with all possible offsets.

The transmitted signal of the i-th user, $i = 1, 2, \ldots, K$, is given by

$$s_i(t) = \sqrt{P}\,[a_I^i(t)\cos(w_c t) + \phi^i) + a_Q^i(t)\sin(w_c t + \phi^i)] \quad (3)$$

where P is the signal power, ϕ^i is the phase of the i-th user, c is the carrier frequency, a_I^i and a_Q^i are the in-phase and quadrature spreading PN code of i-th user, respectively and K denotes the number of users. Noncoherent acquisition detector structure of the QPSK spreading and the matched filter correlator are shown in Figure 3 and 4 respectively. The received signal arriving from the multipath fading channel can be represented as

$$r(t)=\sum_{i=1}^{K}\sum_{l=1}^{L}\alpha_l^i\, s_i(t-\tau_l^i)+n(t) \tag{4}$$

where L is the number of multipaths, τ_l^i is the propagation delay of the l-th path of the i-th user, assumed to be uniformly distributed in the uncertainty region limited by the round trip delay of the cell, $n(t)$ is an additive white Gaussian noise with zero mean and two sided power spectral density $N_o/2$, α_l^i ($l=1, 2, \ldots, L$) is an attenuation factor for the signal received from l-th path of the i-th user and is a Rayleigh distribution. The recevied waveform, after frequency down conversion, is sampled at the chip rate.

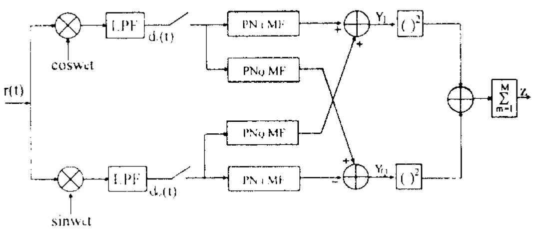


Fig. 3. Noncoherent MF acquisition detector structure of the QPSK spreading.

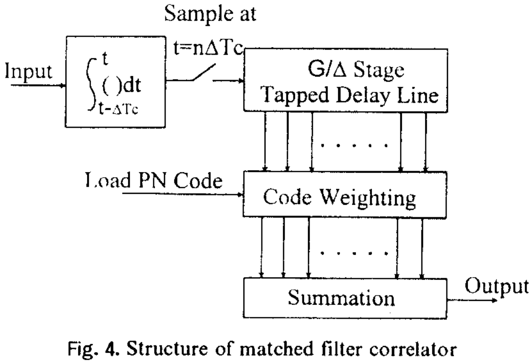


Fig. 4. Structure of matched filter correlator

The sampled baseband I and Q signal is then fed into an I/Q PN MF, summed, squared in each arm, and finally added to yield value Z. In the figures 3, M is the postdetection integration period. G and Δ are the number of chips summed in the PN matched filter and the step size respectively in the figure 4. We consider the chip synchronous CDMA system. Therefore Δ is equal to 1. Postdetection integration period M has effect on the acquisition performance under fading channels. The effect will be analyzed in section Ⅳ. The outputs of the low pass filter, designated as $d_I(t)$ and $d_Q(t)$ for I and Q channel, are described as

$$\begin{aligned} d_I(t) &= LPF[r(t)\cos w_c t] \\ &= \sum_{i=1}^{K}\sum_{l=1}^{L}\alpha_l^i\sqrt{P}\,[\alpha_I^i(t-\tau_l^i)\frac{\cos\theta_l^i}{2} \\ &\quad + a_Q^i(t-\tau_l^i)\frac{\sin\theta_l^i}{2}]+\frac{n_c(t)}{2} \end{aligned} \tag{5}$$

$$\begin{aligned} d_Q(t) &= LPF[r(t)\sin w_c t] \\ &= \sum_{i=1}^{K}\sum_{l=1}^{L}\alpha_l^i\sqrt{P}\,[\alpha_Q^i(t-\tau_l^i)\frac{\cos\theta_l^i}{2} \\ &\quad - a_I^i(t-\tau_l^i)\frac{\sin\theta_l^i}{2}]+\frac{n_s(t)}{2} \end{aligned} \tag{6}$$

where $\theta_l^i = w_c\,\tau_l^i+\phi_l^i$. ϕ_l^i is the phase of i-th user of the l-th path and $n_c(t)$ and $n_s(t)$ are the in-phase and quadrature component of the white Gaussian noise with two sided power spectral density $N_o/2$. Let the j-th user be the user of interest. The outputs of the matched filter Y_I, Y_Q in the I and Q channels are as follows respectively.

$$Y_I = \begin{cases} \sum_{k=1}^{G}(\alpha_{l_n}^j\sqrt{P}\cos\theta_{l_n}^j+\sum_{\substack{l=1\\ l\neq l_n}}^{L} I_{l,l}^{jj}(kT_c)+\sum_{\substack{i=1\\ i\neq j}}^{K}\sum_{l=1}^{L} I_{l,l}^{ij}(kT_c)+N_I(kT_c)); \\ \qquad \text{if } \tau_l^j=\tau^j \text{ for } l=l_n \text{ (in sync.)} \\ \sum_{k=1}^{G}\sum_{l=1}^{L} I_{l,l}^{jj}(kT_c)+\sum_{\substack{i=1\\ i\neq j}}^{K}\sum_{l=1}^{L} I_{l,l}^{ij}(kT_c)+N_I(kT_c)); \text{else(not in sync.)} \end{cases} \tag{7}$$

$$Y_Q = \begin{cases} \sum_{k=1}^{G} (\alpha_{l_n}^{j}\sqrt{P}\sin\theta_{l_n}^{j} + \sum_{\substack{l=1 \\ l\neq l_n}}^{L} I_{Q,l}^{jj}(kT_c) + \sum_{\substack{i=1 \\ i\neq j}}^{K}\sum_{l=1}^{L} I_{Q,l}^{ij}(kT_c) + N_Q(kT_c)); \\ \qquad \text{if } \tau_l^j = \tau^j \text{ for } l = l_n \text{ (in synch.)} \\ \sum_{k=1}^{G}\sum_{l=1}^{L} I_{Q,l}^{jj}(kT_c) + \sum_{\substack{i=1 \\ i\neq j}}^{K}\sum_{l=1}^{L} I_{Q,l}^{ij}(kT_c) + N_Q(kT_c)); \text{else(not in synch.)} \end{cases} \tag{8}$$

where $I_{I,l}^{jj}(kT_c)$, $I_{I,l}^{ij}(kT_c)$, and $N_I^j(kT_c)$ are, respectively, the self-interference caused by the l-th multipath component, the multiple access interference of the i-th user and Gaussian noise and defined as follows:

$$I_{I,l}^{jj}(kT_c) = \alpha_l^j \frac{\sqrt{P}}{2}[(\alpha_I^j(kT_c-\tau_l^j)a_I^j(kT_c-\tau^j) + a_Q^j(kT_c-\tau_l^j)a_Q^j(kT_c-\tau^j))\cos\theta_l^j + (a_Q^j(kT_c-\tau_l^j)a_I^j(kT_c-\tau^j) - a_I^j(kT_c-\tau_l^j)a_Q^j(kT_c-\tau^j))\sin\theta_l^j] \tag{9}$$

$$I_{I,l}^{ij}(kT_c) = \frac{\sqrt{P}}{2}\alpha_l^i[(\alpha_I^i(kT_c-\tau_l^i)a_I^j(kT_c-\tau^j) + a_Q^i(kT_c-\tau_l^i)a_Q^j(kT_c-\tau^j))\cos\theta_l^i + (a_Q^i(kT_c-\tau_l^i)a_I^j(kT_c-\tau^j) - a_I^i(kT_c-\tau_l^i)a_Q^j(kT_c-\tau^j))\sin\theta_l^i] \tag{10}$$

$$N_I^j(kT_c) = \frac{1}{2}[n_c(kT_c)a_I^j(kT_c-\tau^j) + n_s(kT_c)a_Q^j(kT_c-\tau^j)] \tag{11}$$

We can also express $I_{Q,l}^{jj}(kT_c)$, $I_{Q,l}^{ij}(kT_c)$ and $N_Q^j(kT_c)$ in a form similar to (9)~(11). After correlation over G chips, with assumptions that the amplitude α_l^i, phase θ_l^i, and delay τ_l^i of the multipath remain constant over the G chips accumulated, we have

$$Y_I = G\,\alpha_{l_n}^j\sqrt{P}\cos\theta_{l_n}^j + \sum_{\substack{l=1 \\ l\neq l_n}}^{L}\alpha_l^j\frac{\sqrt{P}}{2}[(R_{II}^{jj}(\tau_l^{jj}) + R_{QQ}^{jj}(\tau_l^{jj}))\cos\theta_l^j + (R_{QI}^{jj}(\tau_l^{jj}) - R_{IQ}^{jj}(\tau_l^{jj}))\sin\theta_l^j] + \sum_{\substack{i=1 \\ i\neq j}}^{K}\sum_{l=1}^{L}\frac{\sqrt{P}}{2}\alpha_l^i[(R_{II}^{ij}(\tau_l^{ij}) + R_{QQ}^{ij}(\tau_l^{ij}))\cos\theta_l^i + (R_{QI}^{ij}(\tau_l^{ij}) - R_{IQ}^{ij}(\tau_l^{ij}))\sin\theta_l^i] + \sum_{k=1}^{G} N_I(kT_c) : \text{if } \tau_l^j = \tau^j \text{ for } l = l_n \text{ (in sync.)} \tag{12}$$

$$Y_Q = G\,\alpha_{l_n}^j\sqrt{P}\sin\theta_{l_n}^j + \sum_{\substack{l=1 \\ l\neq l_n}}^{L}\alpha_l^j\frac{\sqrt{P}}{2}[(R_{II}^{jj}(\tau_l^{jj}) + R_{QQ}^{jj}(\tau_l^{jj}))\sin\theta_l^j + (R_{IQ}^{jj}(\tau_l^{jj}) - R_{QI}^{jj}(\tau_l^{jj}))\cos\theta_l^j] + \sum_{\substack{i=1 \\ i\neq j}}^{K}\sum_{l=1}^{L}\frac{\sqrt{P}}{2}\alpha_l^i[(R_{II}^{ij}(\tau_l^{ij}) + R_{QQ}^{ij}(\tau_l^{ij}))\sin\theta_l^i + (R_{IQ}^{ij}(\tau_l^{ij}) - R_{QI}^{ij}(\tau_l^{ij}))\cos\theta_l^i] + \sum_{k=1}^{G} N_Q(kT_c) : \text{if } \tau_l^j = \tau^j \text{ for } l = l_n \text{ (in synch.)} \tag{13}$$

where $R_{II}^{ij}(\tau_l^{ij})$, the partial correlation of the chi sequence, is given by

$$R_{II}^{ij}(\tau_l^{ij}) = \sum_{k=1}^{G} a_I^i(kT_c - a_l^i)a_I^j(kT_c-\tau^j) \tag{14}$$

Similarly, we can get Y_I and Y_Q for the other case. Note that, in (14), the sequences a_I^i and a_I^j are the phase-shifted sequences of the same PN generator, and hence $R_{II}^{ij}(\tau_l^{ij})$ is a partial autocorrelation function. The output of the noncoherent square law combining is obtained as follows,

$$Z = \sum_{m=1}^{M}(Y_I^2 + Y_Q^2) \tag{15}$$

Assume that the received signal consists of L path signal components and the receiver has N despreading correlator paths with a delay of one code chip between them. Therefore, the acquisition scheme searchs for the N paths by sliding correlation. The matched filter output Z is shifted into the memory until acquisition process tests all possible delay positions.

After sliding over the uncertainty region, peak value at some position will be found. The receiver window is then centered around this position by shifting the local code, and the verification mode is initiated. The threshold for verification mode(Th_v) is determined at the search mode by averaging over the uncerataInty cell D and the threshold used in verification mode has the relation of $Th_v = R \cdot Th$. R is a threshold multiplier. Th is defined as follows:

$$Th=\frac{1}{D}\sum_{k=1}^{D} Z(kT) \qquad (16)$$

where T is equal to GMTc. The block diagram of the acquisition algorithm is shown in Figure 5.

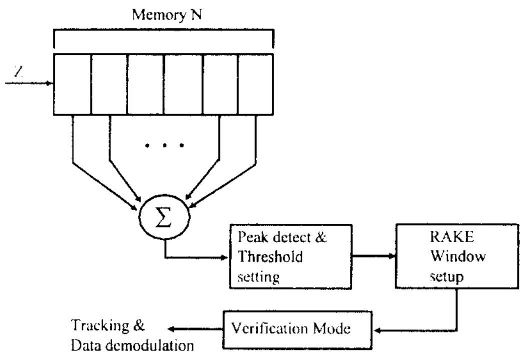


Fig. 5. Block diagram of the acquisition algorithm

Ⅳ. Performance Results

In this section we present the result of performance. In order to illustrate the performance of the acquisition algorithm described in the previous section, the Monte Carlo method is used to simulated the acquistion system. Unlike other multiple access techniques, CDMA is interference linited, not by noise. Thus, the effect of mutiuser interference on the acquisition performance was analyzed and AWGN was not considered in the simulation. The detection probability and mean acquistion time($E[T_{acq}]$) are used as a meausre of performance. The detection probability is the probability to correctly acquire 3-paths in the search mode. And threshold in the verification mode is determined in the search mode. In our simulation studies, four directions are studied: 1) Doppler frequency ($f_d T_c$) 2) number of users(K) 3) Postdetection integration period(M) 4) processing gain (G). As for the parameters used in this scheme: the PN code length is 32767 chips, in the verification mode the parameters A and B are given as 4 and 2 respectively[3], code rate is 1.2288MHz and the random delay D is uniformly distrubuted over 500 chips. The processing gain G is an integer multiple of 256 chips and false alarm penalty is 1000 times 256chips. We assume that a delay spread is 4μs and system bandwidth is 1.23MHz. Hence, from the equation (2) the number of the multipath components(L) is 6. The 3-path RAKE receiver is used. Multipath power delay profiles for typical urban, hilly terrain, and rural areas are derived from those specified for the GSM (Global System for Mobile communications) system.

In this paper, multipath power delay profile for typical urban area has been used to assess the acquisition performance of the CDMA system against a frequency selective fading channel in multiuser envirnoment. The multipath channel profile for typical urban is shown at the table 1.

Table 1. Multipath Channel Profile

Ray Number	Number of chip delays	Delay (μsec)	Relative power(dB)
1	0	0	0
2	1	0.813	−3.484
3	2	1.626	−6.968
4	3	2.439	−10.453
5	4	3.252	−13.937
6	5	4.065	−17.421

Various values of fdTc were used to investigate the effect of Doppler frequency on the acquisition performance. The probabilities of detection are plotted in Fig. 6~10 as a function of the number of user. Inspection of Figs 6~10 shows that the detection probability is increased and the amount of performance improvement is decreased gradually as the Doppler frequency increases. However, the proposed postdetection integration method provides an enhanced performance especially in the case of high fdTc under the condition of the same tap length (GM = 512) in the matched filter.

Fig. 11 shows the sensitivity of the acquisition time with respect to the threshold multiplier(R) when the number of user is 2. It is shown that the acquisition time is more sensitive in the large Doppler frequency than in small Doppler frequency. And the optimum value is around 3. All the simulations are performed in this value. If the acquisition threshold factor R is smaller than 2, the false acquisition probability will be high and false penalty is given. Therefore, the mean acquisition time is increased. On the other hand, as the factor is increased more than 3 the detection probability is decreased. And also $E[T_{acq}]$ is increased gradually. At the optimal threshold setting the mean acquisition time reaches a minimum value that is decided by the acquisition system parameters. This minimum mean acquisition time is $(D+AGM)$ Tc seconds when detection probability approaches unity.

Fig. 12 to Fig. 16 show the combined effect of G and fdTc on the acquisition time. At low value of fdTc, G = 512 perform better than with G = 256. But at fdTc = 5e-3, the detector with G = 256 performs better. Hence, it shows that up to about 1e-3, G can be increased to reduce the mean acquisition time but after that it begins to deteriorate as G increases. Therefore, we can say that in a fading channel, increasing the number of taps in the matched filter may not necessarily result in any performance improvement, but it may even cause performance degration if the Doppler frequency is large. If further performance improvement is required in a severe fading environment parallel implementation similar to those discussed in [10] and [11] may be the option. But the proposed postdetection integration method can also provide performance improvement. Fig. 7 to Fig. 9 show the combined effect of Doppler frequency and postdetection integration period and that for low fdTc, the effect of the postintegration is not clear, but the postdetection integration shows a clear performance improvement for a high fdTc under the same integration time(MG = 512).

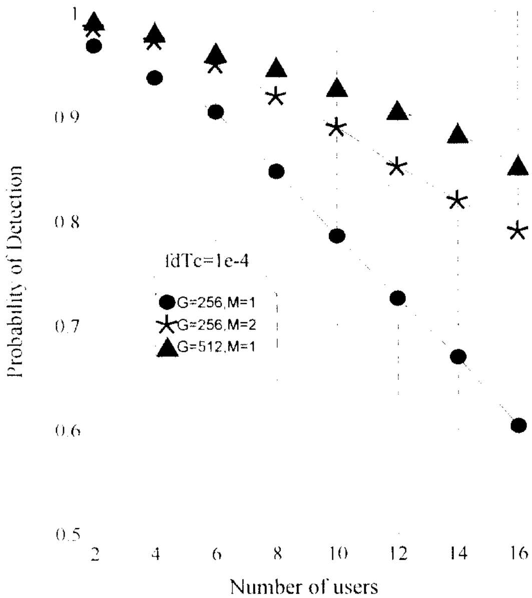


Fig. 6. Probability of detection versus number of users for fdTc = 1e-4 and different G, M

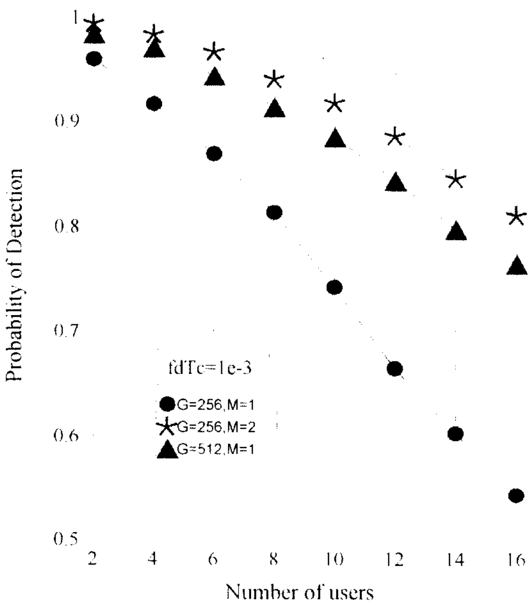
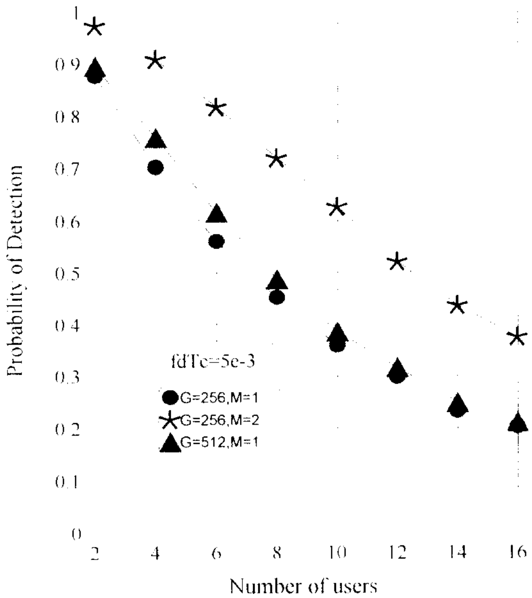


Fig. 7. Probability of detection versus number of users for fdTc = 1e-3 and different G, M



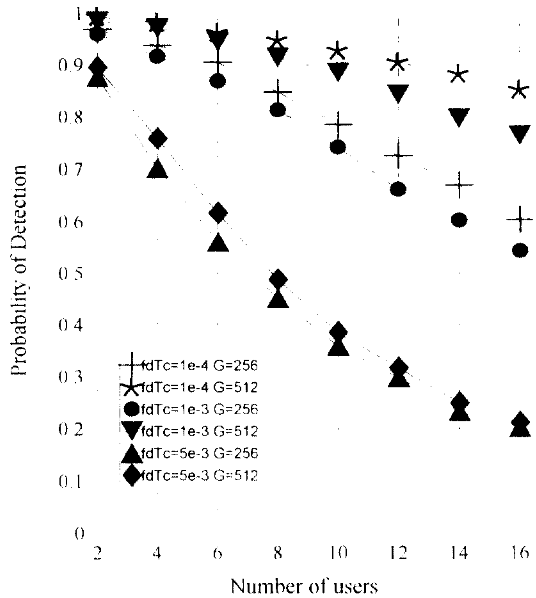


Fig. 9. Probability of detection versus number of users for fdTc=5e-3 and different G, M and G

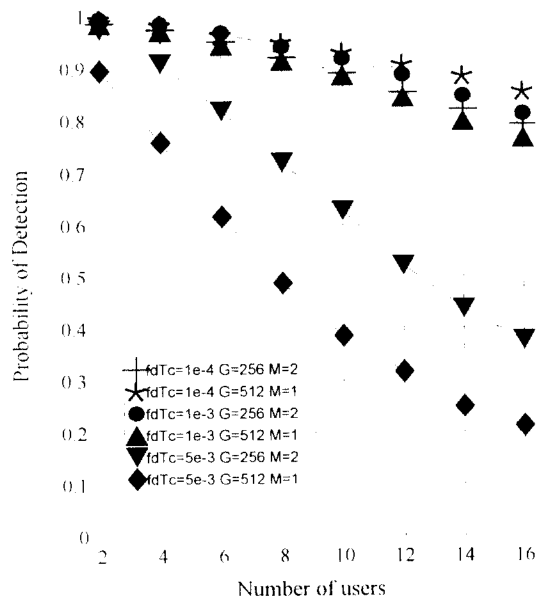


Fig. 10. Probability of detection versus number of users for different G, M

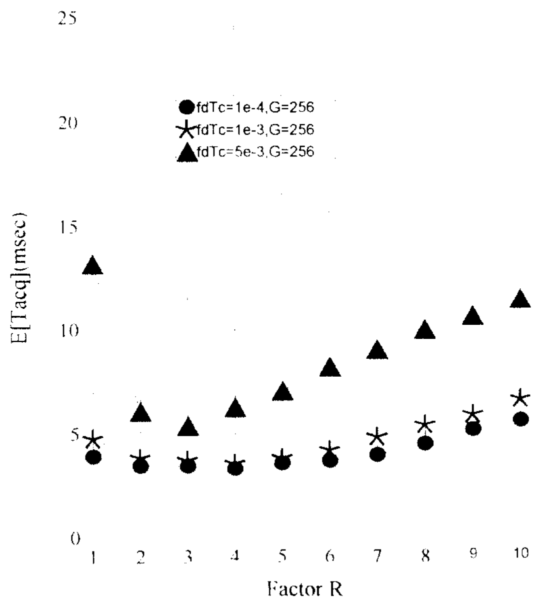


Fig. 11. Mean acquisition time for different threshold factor R and fdTc

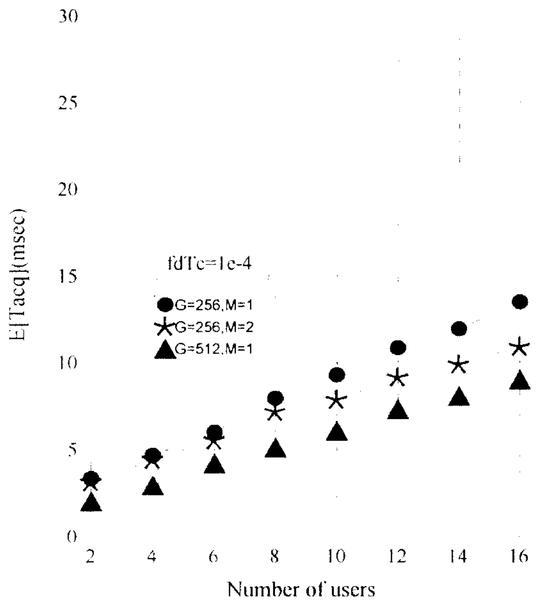


Fig. 12. Mean acquisition time for different G and M, fdTc = 1e-4.

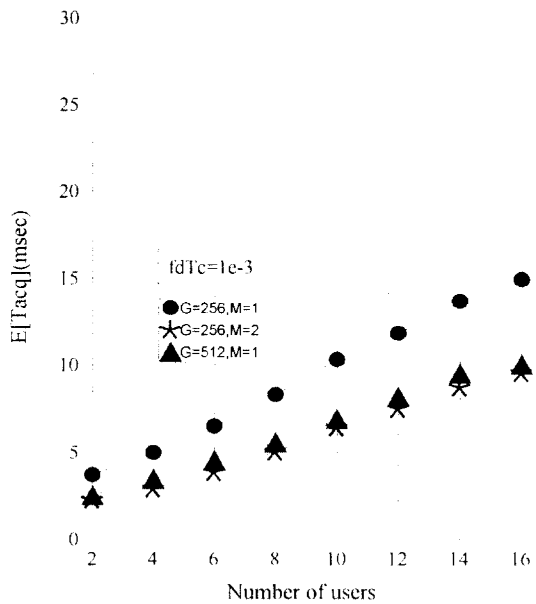


Fig. 13. Mean acquisition time for different G and M, fdTc = 1e-3.

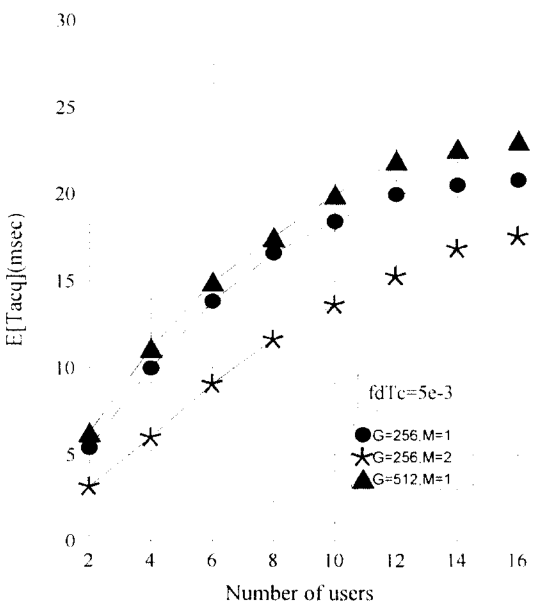


Fig. 14. Mean acquisition time for different G and M, fdTc = 5e-3.

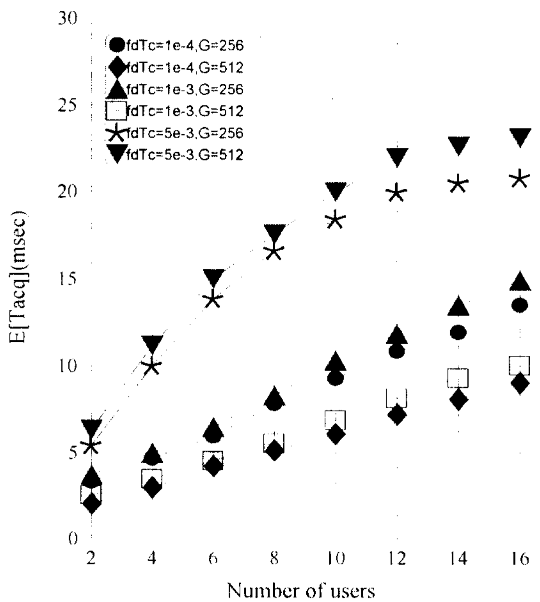


Fig. 15. Mean acquisition time for different fdTc and G

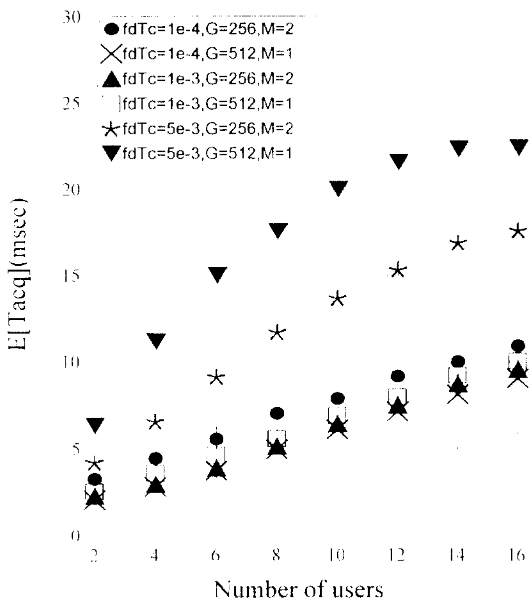


Fig. 16. Mean acquisition time for different fdTc, G and M.

V. Conclusions

In this paper acquisition algorithm has been studied for the DS/CDMA RAKE receiver. The performance of the acquisition scheme is investigated in the frequency selective fading channel and multiuser environment. The detection probability and the mean acquisition time have been obtained by simulation. And also the postintegration detection is studied to overcome the effect of doppler frequency on the acquisition system. It has been found that the postdetection integration technique improves the detection probability and then reduces acquisition time in the case of high fdTc. But as M increases more than 3 the performance of the system is not expected to be improved. Therefore $M=2$ is best choice in a high fdTc. And for fast and reliable acquisition in a frequency selective channel, suboptimal threshold set method is provided.

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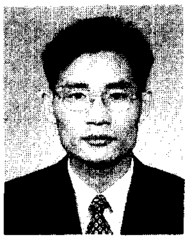

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