A NEW RAPID ACQUISITION SCHEME FOR BURST MODE DS SPREAD SPECTRUM PACKET RADIO

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ABSTRACT

A new rapid acquisition scheme for DS/SS packet radio systems using PN matched filters is described. The system operates in burst mode, with limited permitted time for synchronization using a fixed length preamble at the beginning of each data burst. Multiple PN codes are available for acquisition at the first portion of the preamble. The second protion of the preamble consists of several PN codes which are modulated with a marker. The receiver may achieve code acquisition anywhere within the first portion of the preamble. Since it is not known a priori in which of the several code periods acquisition will occur, coincidence detection is done using the marker for frame synchronization. The system performance is measured in terms of the probability of packet loss for a given preamble and message length. Appropriate limits are set on the probability of packet blocking caused by false alarms. Analysis shows that the new scheme yields better performance and is more appropriate than some other previous schemes in packet radio system.

1 INTRODUCTION

Direct sequence spread spectrum systems operating in burst or packet mode transmit user data in packets of a few thousand bits where each packet begins with a short synchronization preamble. Thus for receivers which use matched filters like SAW devices spanning one PN code period, only a finite number L PN code periods in the preamble are available at the beginning of a burst for P-N acquisition and to determine the start of user data. Because synchronization must be achieved during the limited time of the preamble, one measure of system performance is the probability of not achieving synchronization within this limited permitted time, i.e. missing the packet. However, packets may be also lost because a false alarm in noise can cause the receiver to be busy, and thus unavailable (blocked) when the packet arrives. Thus the overall system performance is characterized by the probability of packet loss caused by either failed synchronization (missed packet) or receiver blocking.

For maximum information throughput, a tradeoff is made between the probability of successful synchronization and the overhead time needed for synchronization. Thus the length L of the preamble which yields maximum throughput will depend on the number of information symbols in each data burst and the symbol error rate [4].

One previous scheme requires code acquisition at the first PN code period, followed by coincidence detection (verification) on the remaining $L-1$ PN code periods [1]. For any particular choice of acquisition threshold and corresponding false alarm rate, the matched filter may not be long enough for reliable acquisition at low SNR. Therefore most of packet losses may be caused by synchronizations failure.

The two level (TL) code acquisition scheme presented by Rappaport and Wilson in [2, 3] can be applied in the bursty form communication. The reliable sync is guaranteed by the use of several active correlators following one or more fast acquisition matched filters. In the scheme, the detection threshold can be set to be reasonably low for reliable detection in low SNR, because most of the "false-start" signals may be dismissed by a number of active correlators. However, this method needs a relatively long preamble, thus it may not be suitable for packet-switch communication. Furthermore, the hardware complexity may have limit to build many active correlators in one receiver.

The comparisons given at the end of the paper show that the proposed scheme in this paper yields much better performance than the previous scheme in [1]. And for a given short preamble, performances of the new scheme and the TL scheme are almost equivalent. However the new scheme has a much simpler receiver structure.

The paper is organized as follows. In section 2, the acquisition technique is described. The performance analysis is carried out in section 3, followed by numerical results in Section 4 and conclusions in Section 5.

2 ACQUISITION TECHNIQUE

Figure 1a shows the data format used by the new technique for an example preamble length of $L = 8$ PN code periods. We assume one data bit per code period, so that the preamble consists of the data sequence 1111 1001. For comparison, figure 1b shows the preamble data format for the TL method of [2, 3].

Figure 2 shows a simplified CSK receiver structure for the receiver. A bandpass filter eliminates out-of-band noise. The noncoherent matched filters (MF) are designed to match two orthogonal PN codes corresponding to ones and zeros, and are followed by envelope detectors (ED) with a detection threshold b_0. After synchronization is achieved, bit decisions are made by comparing the two ED outputs and selecting the larger output.

The receiver operates by searching the incoming data stream for synchronization in two steps: initial acquisition (correlation detection) and coincidence detection (frame synchronization).

Initial acquisition is achieved when the MF output $y(t)$ exceeds a threshold b_0. For our acquisition process, this may occur in any one of the N_h code periods in the preamble, not necessarily the first one. When this first step is complete, the receiver sets the threshold to a new value b_1 and samples the MF output once per code period. The coincidence detection is achieved when the current data symbol $\hat{a}\epsilon\{-1, 1, erasure\}$ plus the previous $N_s - 1$ symbols match the N_s symbol marker within a specified symbol error and erasure tolerance. Since it is not known a priori at which point in the N_h preamble symbols acquisition occurred, the marker search (symbol-wise correlation) is performed in successive positions of the marker until a match is found or N_h positions have been tested. If no match is found, then the receiver resumes the search for initial acquisition. If a match is found, then the receiver starts to process the L_D symbols of user data.

The marker of length N_s is chosen to be a sequence with minimum correlation sidelobes when preceded by the N_h '1' symbols in the preamble, while also preserving minimum sidelobes when preceded by random data or noise. The optimum marker sequences for selected values of N_s are determined in [10].

The advantage of this new technique is that successful acquisition can be achieved even if several PN sequences are missed due to noise, thus increasing the reliability at low SNR. The technique in [1] is a special case where $N_h = 1$, so that if the receiver does not achieve acquisition in the first PN code period, then the packet is missed.

3 PERFORMANCE ANALYSIS

In this section, the synchronization performance of the new rapid acquisition technique is analyzed in terms of the probability P_L of packet loss versus SNR. The system parameters which determine

P_L are the PN sequence length $N, N_h, N_s, L_D, b_0, b_1$ and the symbol error tolerance in the marker. An expression for P_L is determined and each component of this expression is evaluated in terms of the system parameters. This is followed by consideration of threshold selection, and the use of two thresholds.

3.1 Preliminaries

A binary code shift keying (CSK) direct-sequence spread-spectrum system is briefly reviewed to establish notation. The received signal

$$\begin{aligned} r(t) &= [\sum_k A a_{i_k}(t-\tau-kT)+n_I]\cos\omega_0 t \\ &\quad -n_Q\sin\omega_0 t \qquad i_k=0,1 \end{aligned} \tag{1}$$

at the output of the bandpass filter, where A is the amplitude of the signal, and $a_0(t)$, $a_1(t)$ are orthogonal spreading sequences time-limited to $[0,T]$ which represent ones and zeros of the binary message. n_I and n_Q are the in-phase and quadrature components of the white Gaussian noise with two sided power spectral density $N_0/2$. The noise power at the output of the bandpass filter is $\sigma^2 = N_0/T_c$. The sampling rate at the output of the matched filter is $1/T_c$. Let N be the total number of chips in a PN code period, we have $T = NT_c$.

Considering the acquisition procedure in which the detector observes $a_1(t)$ over a period $0 < t < N_h T$. The test statistic at the output of the matched filters is

$$y(t) = \{ANc(t)+n'_I\}\cos\omega_0 t - n'_Q\sin\omega_0 t \tag{2}$$

where

$$c(t) \equiv \begin{cases} \frac{1}{T}\int_0^t a_1(\tau)a_1(\tau-t)d\tau & t<T \\ \frac{1}{T}\int_0^T a_1(\tau)a_1(\tau-t)d\tau & t\geq T \end{cases} \tag{3}$$

is the normalized autocorrelation function of the PN sequence $a_1(t)$, and n'_I, n'_Q are filtered noise. The upper equation in (3) corresponds to the partial correlation when the first PN code period of the preamble has not yet come into the matched filter completely, i.e. only a fraction of the first bit (PN period) is in the matched filter. The lower one corresponds to the whole period autocorrelation after the first PN period passed to the matched filter.

Assuming that the receiver has no knowledge of the amplitude A of the received signal, we select the threshold of the acquisition for an acceptable probability of false alarm P_{f0n} caused by the noise. From [6], the output of the envelope detector is according to Rayleigh distribution when no signal is present. Accordingly, the probability of false alarm is

$$P_{f0n} = \exp(-\frac{b_0^2}{2}), \tag{4}$$

where b_0 is the acquisition threshold normalized to σ. Also from [6], the output of the envelope detector is according to Rician distribution when signal plus noise is present. Thus the probability of initial acquisition in the correct position ($c(iT)=1$, $i=1,2,...,N_h$) is given by

$$P_{d0} = Q(\sqrt{2\gamma_o}, b_0), \tag{5}$$

where $\gamma_o = N\gamma_i = N\cdot A^2/2\sigma^2 = N\frac{E_c}{N_0}$ (E_c is chip energy) is the SNR at the output of the MF, and $Q(a,b)$ is the Marcum-Q function.

Since $c(t)$ itself is a periodic function after the first period with period T, the probabilities of false initial acquisition at an incorrect position where $c(t) < 1$ can be written

$$\begin{aligned} P_{f0d,j} &= Q(\sqrt{2\gamma_o c^2(jT_c)}, b_0), \\ &\quad t<T, j=1,...,N-1. \end{aligned} \tag{6}$$

$$\begin{aligned} P_{f1d,j} &= Q(\sqrt{2\gamma_o c^2((kN+j)T_c)}, b_0), \\ &\quad t>T, j=1,...,N-1, k=1,2,...,N_h-1. \end{aligned} \tag{7}$$

[5] proved that all the tests at incorrect and correct positions can be taken as the statistically independent events if the PN sequence is long enough.

3.2 Calculation of performance

There are two cases for a packet to be captured. The first one is when no blocking occurs while a packet arrives. And the sync is obtained successfully under the above condition. The second one happens when blocking interval ends before the last PN sequence for acquisition comes into the matched filter. Up to $N_h - 1$ acquisition bits may be lost due to the blocking, but synchronization is still possible. If it is achieved, the packet is captured. For simplicity, we only concern the worst case which assumes the probability of successful sync when the second case occurs to be zero. From [1], the probability of packet loss P_L can be written by

$$P_L = 1-(1-P_B)P_{syn} \tag{8}$$

where P_B is the probability of receiver blocking due to false alarms, and P_{syn} is the probability of successful synchronization provided that the receiver is not blocked. P_{syn} can be written

$$P_{syn} = \sum_{i=1}^{N_h} P_{acq}(i)\cdot P_c(i) \tag{9}$$

where $P_{acq}(i)$ is the probability of acquisition in the ith PN period. $P_c(i)$ is the probability of coincidence when the acquisition occurs in the ith PN period.

To determine P_L, it remains to evaluate $P_{acq}(i), P_c(i)$ and P_B.

3.2.1 Probabilities of initial acquisition $P_{acq}(i)$

As we mentioned early that the decision tests for acquisition are statisticly independent. Thus the probability of initial acquisition $P_{acq}(i)$ at the ith PN sequence can be simply given by

$$\begin{aligned} P_{acq}(i) &= \prod_{j=1}^{N-1}(1-P_{f0d,j})[\prod_{j=1}^{N-1}(1-P_{f1d,j})]^{i-1} \\ &\quad \cdot(1-P_{d0})^{i-1}P_{d0} \end{aligned} \tag{10}$$

The first term in this equation means no false acquisition before the first main correlation peak. The second term means no false acquisition before any of the other $i-1$ main correlation peaks. The third one means that the first $i-1$ main peaks are missed. And the last one implies that acquisition occurs in the ith peak.

3.2.2 Probability of coincidence detection

To calculate the second term $P_c(i)$ in (9), we note that the coincidence test is carried out in up to N_h positions of the marker. The probability of successful coincidence detection $P_c(i)$ will depend on the probability of a false marker detection when testing at an incorrect position before the correct position is reached, as well as on the probability of a successful marker test when testing in the correct position. The number of incorrect positions tested, and thus the probability of false marker detection, will depend on the particular PN code period i in the preamble at which the search for the marker begins.

The probability $P(h, h_x, n)$ of marker detection in a particular position depends on the specific choice of marker, the amount of overlap n, the error tolerance h and the erasure tolerance h_x between the observed marker bits and the known or expected marker bits. The overlap n is defined to be N_s when there is complete overlap (i.e. correct position) and may be 0 or negative if there is no overlap . $P(h, h_x, n)$ is evaluated in [7] and given by

$$
\begin{aligned}
& P(h, h_x, n) \\
= \; & Prob(h \text{ or fewer disagreements and } h_x \text{ or fewer} \\
& \text{erasures with overlap } n) \\
= \; & \sum_{j=0}^{h} \sum_{k=0}^{\min(h_x, N_s - j)} \sum_{m=\max(0, k-h_0(n))}^{\min(k, N_s - h_0(n))} \sum_{\ell=\max(0, j-h_0(n)+k-m)}^{\min(j, N_s - h_0(n) - m)} \\
& \binom{N_s - h_0(n)}{\ell, m} q^{N_s - h_0(n) - \ell - m} \, p^{\ell} \, s^{m} \\
& \times \binom{h_0(n)}{j-\ell, \; k-m} q^{j-\ell} \, p^{h_0(n)-j+\ell-k+m} \, s^{k-m}, \qquad (11)
\end{aligned}
$$

$$
\text{where } \binom{h_0(n)}{\ell, \; m} = \frac{h_0(n)!}{\ell! m! (h_0(n) - \ell - m)!}. \qquad (12)
$$

where symbol error probability p and erasure probability s for noncoherent CSK receiver are derived in the Appendix.

Following [7, 8], $P_c(i)$ can be calculated from

$$
\begin{aligned}
P_c(i) \; = \; & Pr\{\text{No detection occurred at } n \neq N_s\} \\
& \cdot Pr\{\text{Detection at } n = N_s \mid \text{no detection} \\
& \text{occurred at } n \neq N_s\} \\
= \; & [1 - Pr\{\text{At least one detection occurred} \\
& \text{at } n \neq N_s\}] \cdot Pr\{\text{Detection at } n = N_s \mid \text{no} \\
& \text{detection occurred at } n \neq N_s\} \qquad (13)
\end{aligned}
$$

A lower bound on $P_c(i)$ may be written

$$
P_c(i) \geq [1 - \sum_{n=N_s+i-N_h}^{N_s - 1} P(h, h_x, n)] \cdot P(h, h_x, N_s) \qquad (14)
$$

3.2.3 Probability of blocking P_B

P_B can be approximated by $\bar{F}/(\bar{A} + \bar{F})$, where $\bar{F}$ and $\bar{A}$ are the mean blocking time and mean silence time respectively within any sufficiently long time interval. $\bar{F}$ and $\bar{A}$ may be determined according to figure 3, in which time axis is characterized by an alternating sequence $\{A_j, F_j\}$, where A_j and F_j denote the jth silence interval and jth blocking interval of the receiver respectively.

The false acquisition in noise is modeled as a Poisson process with arrival rate $\lambda = P_{fan}/T_c$. Since the inter-arrival times of a Poisson process are i.i.d. exponential, we have

$$
\bar{A} = E\{A_j\} = 1/\lambda \qquad (15)
$$

where E means the statistical expectation. The blocking or busy time once a false acquisition occurs is a random variable which depends on the probability P_{fcn} of false coincidence in noise only. If no false coincidence occurs, the receiver will be busy for $(N_h + N_s - 1)$ PN periods. Otherwise it will be unavailable for about a time equal to the packet length. Thus $\bar{F}$ can be determined as

$$
\begin{aligned}
\bar{F} \; = \; & E\{F_j\} = (1 - P_{fcn})(N_h + N_s - 1)T \\
& + P_{fcn}(T_R + T_D) \qquad (16)
\end{aligned}
$$

where $T_D = L_D T$ is the data length. T_R is a random variable which depends on which of the $(N_h - 1)$ coincidence positions is detected. Obviously, $T_R \leq (N_h + N_s - 1)T$. Thus

$$
\bar{F} \leq (N_h + N_s - 1)T + P_{fcn} T_D \qquad (17)
$$

If we define

$$
\begin{aligned}
\rho \; = \; & \frac{(N_h + N_s - 1)T + P_{fcn} T_D}{\bar{A}} \\
= \; & P_{fan}(N_h + N_s - 1 + P_{fcn} L_D) N \qquad (18)
\end{aligned}
$$

we obtain the upper bound of P_B as

$$
P_B \leq \frac{\rho}{1 + \rho} \qquad (19)
$$

Define $\{q_n, s_n, p_n\}$ as the probabilities of decisions $\{+, \text{erasure}, -\}$, then

$$
\begin{aligned}
P_{fcn} \; = \; & Pr\{\text{at least one false marker detection occurs} \\
& \text{in noise during } N_h \text{ marker tests}\} \\
\leq \; & N_h Pr\{\text{one test succeeds}\} \\
= \; & N_h P_{fn} \qquad (20)
\end{aligned}
$$

where

$$
P_{fn} = \sum_{j=0}^{h} \sum_{k=0}^{min(h_x, N_s - j)} \binom{N_s}{k, \; j} p_n^j \, s_n^k \, q_n^{N_s - j - k}. \qquad (21)
$$

The calculations of s_n, p_n and q_n are also given in the Appendix.

3.3 Selection of thresholds

From (8), since $P_{syn} \leq 1, P_L$ can never be less than

$$
P_L(min) = P_B, \qquad (22)
$$

which is bounded by $\rho/(1 + \rho)$. Thus the system designer can select a desired $P_L(min) \ll 1$ by an appropriate choice of b_0 and b_1. Typical values of $P_L(min)$ may be in the range 10^{-3} to 10^{-6}.

3.4 Two thresholds for initial acquisition

At high SNR, correlation sidelobes of the PN sequences may cause false acquisition at an incorrect position prior to the correct positions, because the threshold b_0 is determined independent of the received signal power. Thus P_L will have a minimum value at some SNR, and will increase for higher as well as lower values of SNR. This effect has been observed for $N_h = 1$ in [1].

One way to eliminate this effect is to use hard limiter receiver given in [9] with the cost of some degradation in performance in low SNR for AWGN channel. Another way is to construct a maximum likelihood receiver which can compare all the samples in a whole period of PN correlations and pick up the largest one as the main correlation peak. The simplest and practical way is to apply the two-level threshold method of [1] to the present case. The priority in the sync decision is given to the higher level. Whenever an acquisition detection by a decision device with the lower threshold level b_0 is followed by the crossing of a higher threshold $b_{0h} = b_0 + \Delta$, within one PN period, the latter is taken for the correct sync detection [2]. By this method, enough dynamic range can be obtained.

4 NUMERICAL RESULTS

In this section, the performance of the receiver is illustrated for $L = 8$ and $L_D = 1050$. b_0 and b_1 are selected to get the required $P_L(min)$. Performance comparison with the TL acquisition scheme at low SNR is made for $L = 9$.

For $L = 8$, figure 4 shows P_L versus SNR for different N_h and N_s respectively with b_0 and b_1 chosen so that $P_L(min) = 10^{-3}$. The double threshold method of [1] is used with $\Delta = 4$ dB. For each value of N_h, h and h_x were selected for the widest dynamic range. The best overall performance is obtained with $(N_h, N_s) = (4, 4)$ or $(5,3)$. From these curves, we can see that almost 3 dB improvement can be achieved at low SNR compared to the method of [1] where $(N_h, N_s) = (1, 7)$, with only a slight reduction at high SNR.

Figure 5 shows P_L versus SNR with thresholds chosen to obtain $P_L(min) = 10^{-3}$ to 10^{-6} and parameters $(N_h, N_s) = (4, 4)$ and $(1,7)$. The amount of improvement at low SNR obtained with the new sync technique is essentially independent of the choice of $P_L(min)$.

A comparison between the proposed acquisition scheme and the TL code acquisition scheme [2, 3] is made in case of low SNR where the influence of correlation sidelobes can be neglected. For this case the average rate of false code start signals when a packet comes is the same as that in noise. The analysis for high SNR will be difficult because the false code start rate when signal is present will be different from that when signal is absent. To use the result from [2, 3] for probability of packet loss (miss probability defined in [2, 3]), we neglect the effect of false lock on blocking probability. Based on the above case and assumption, P_L for the single copy message leader of TL acquisition can be calculated using the same expression as (8) with P_b replaced by Erlang B formula B(c,a), where c is the number of correlators and $a = P_{f0n}M$ is the offered load. M is the number of chips in the long code period. The probability P_{syn} of successful synchronization is written

$$\begin{aligned} P_{syn} &= P_{d0}P_{d1} \\ &= Q(\sqrt{2N\gamma_{in}}, b_0)Q(\sqrt{2M\gamma_{in}}, b_1) \qquad (23) \end{aligned}$$

The results are shown in figure 6 for the same $P_L(min) = 10^{-3}$ and preamble length $L = 9$. This figure shows that, for a short preamble, the proposed method yields better performance than that of single copy TL method unless there is a large number of active correlators for TL method, for example more than 16 correlators in this example. On the other hand, the new method has relatively simpler hardware construction because it does not use any active correlators. We also calculate the sync performance of multiple prefix preamble of TL scheme. We have found that, for $L = 9$, 3-prefix preamble will give the minimum P_L. The comparison of the new scheme with this case in figure 7 shows that the two methods are essentially equivalent in performance unless there is a large number of active correlators.

5 CONCLUSIONS

For burst mode DS spread spectrum communications, where a fixed length preamble is used at the beginning of each data packet for synchronization, and noncoherent matched filters are used for detection, the new acquisition technique shows the advantages over the previous schemes in either sync performance or receiver complexity.

The probability of packet loss is lower bounded by the probability of receiver blocking caused by false alarms. The system designer can determine this lower bound by selecting appropriate decision thresholds. For a preamble length of 9 PN code periods, the proposed scheme is equivalent to multiple-prefix TL scheme in performance and is much simpler in receiver construction.

Appendix Determination of q, p, s, q_n, p_n, s_n

In this appendix, we determine the probabilities q, p, s, q_r, p_r, s_r for the non-coherent receiver of Figure 2. Two matched filters are used for the marker test. We assume that the two input signals are orthogonal, where u_1 and u_0 corresponding to "1" and "0" respectively are the outputs of the envelopes normalized to σ.

The distribution at each MF output is written [5]

$$p(u_1) = u_1 \exp(-\frac{u_1^2 + 2\gamma_0}{2}) I_0(\sqrt{2\gamma_0} u_1) \qquad (24)$$

$$p(u_0) = u_0 \exp(-\frac{u_0^2}{2}) \qquad (25)$$

provided that "1" was sent.

For reliable coincidence detection ($P(h, h_x, N_s) \simeq 1$) without excessive false alarms ($P_{f2n} \simeq 0$), it is necessary to use a normalized threshold $b_1 \neq 0$ with the comparison device, thus erasures may occur if both MF outputs are less than b_1. If we define

$$Q(\alpha, \beta) = \int_\beta^\infty x exp(-\frac{x^2 + \alpha^2}{2}) I_0(\alpha x) dx \qquad (26)$$

then q, p, s, q_n, p_n, s_n are calculated as follows:

$$\begin{aligned} s &= \int_0^{b_1} \int_0^{b_1} p(u_1, u_0) du_1 du_0 \\ &= [1 - Q(\sqrt{2\gamma_0}, b_1)][1 - \exp(-\frac{{b_1}^2}{2})] \qquad (27) \end{aligned}$$

$$\begin{aligned} q &= \int_{b_1}^\infty P_r[U_0 < u_1 | U_1 = u_1] p(u_1) du_1 \\ &= \int_{u_1=b_1}^\infty \int_{u_0=0}^{u_1} p(u_0) du_0 p(u_1) du_1 \\ &= \int_{b_1}^\infty [1 - exp(-u_1^2/2)] p(u_1) du_1 \\ &= Q(\sqrt{2\gamma_0}, b_1) - \frac{1}{2}\exp(-\frac{\gamma_0}{2}) Q(\sqrt{\gamma_0}, \sqrt{2} b_1) \qquad (28) \end{aligned}$$

$$p = 1 - q - s \qquad (29)$$

$$\begin{aligned} q_n &= p_n = \int_{u_1=b_1}^\infty \int_{u_0=0}^{u_1} u_0 exp(-\frac{u_0^2}{2}) du_0 u_1 exp(-\frac{u_1^2}{2}) du_1 \\ &= \exp\{-\frac{b_1^2}{2}\} - \frac{1}{2}\exp\{-b_1^2\} \qquad (30) \end{aligned}$$

$$s_n = 1 - 2q_n \qquad (31)$$

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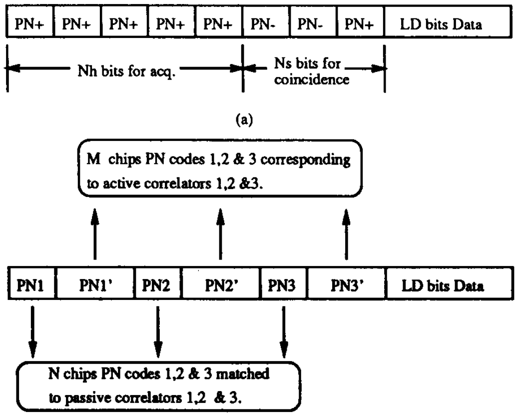

(a)

(b)

Fig . 1 Packet data format for the new scheme (a) and the 3 prefix preamble TL scheme (b)

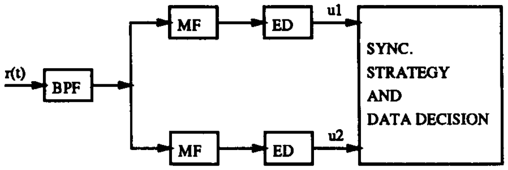

Fig. 2 Simplified CSK receiver

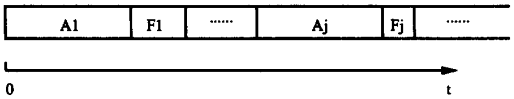

Fig. 3 State of receiver in noise. A : idle; F : busy.

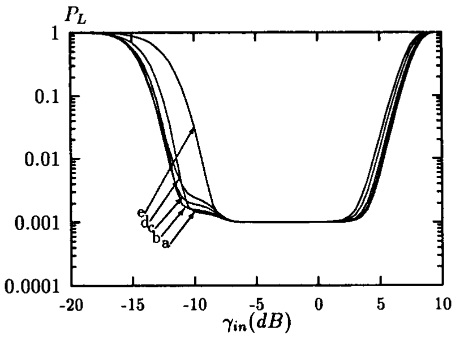

Fig.4. P_L versus SNR for different N_h.
a:$N_h = 3, N_s = 5, b_0 = 5.304, b_1 = 2.6, h = 0, h_x = 2$
b:$N_h = 4, N_s = 4, b_0 = 5.239, b_1 = 2.6, h = 0, h_x = 1$
c:$N_h = 5, N_s = 3, b_0 = 5.130, b_1 = 2.8, h = 0, h_x = 0$
d:$N_h = 6, N_s = 2, b_0 = 5.028, b_1 = 3.2, h = 0, h_x = 0$
e:$N_h = 1, N_s = 7, b_0 = 5.004, b_1 = 2.2, h = 1, h_x = 2$

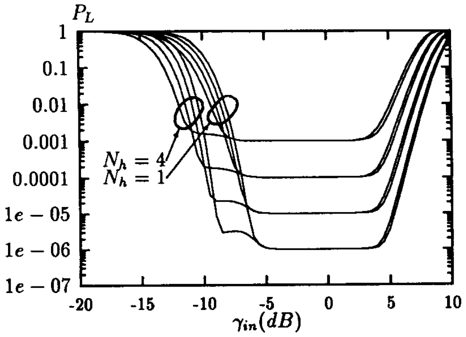

Fig.5. P_L versus SNR for different $P_L(min)$, comparison of new and previous techniques with $N_h = 1$.
$N_h = 4, N_s = 4, b_1 = 2.6, h = 0, h_x = 1$
$P_L(min) = 10^{-3}, b_0 = 5.239$
$P_L(min) = 10^{-4}, b_0 = 5.662$
$P_L(min) = 10^{-5}, b_0 = 6.005$
$P_L(min) = 10^{-6}, b_0 = 6.424$
$N_h = 1, N_s = 7, b_1 = 2.2, h = 1, h_x = 2$
$P_L(min) = 10^{-3}, b_0 = 5.349$
$P_L(min) = 10^{-4}, b_0 = 5.763$
$P_L(min) = 10^{-5}, b_0 = 6.150$
$P_L(min) = 10^{-6}, b_0 = 6.514$

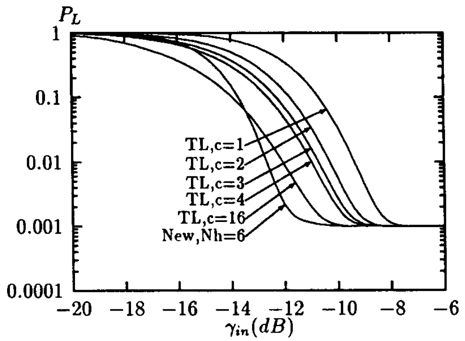

Fig.6. Comparison between the new method and the single copy TL method for a given preamble length $L = 9$ and $P_L(min) = 10^{-3}$. c : number of active correlators.

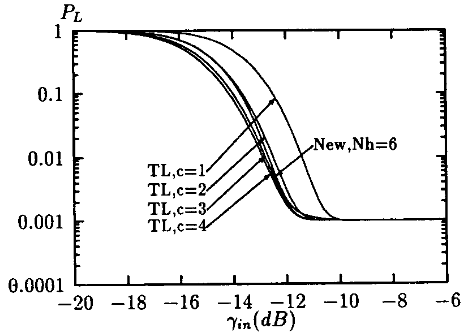

Fig.7. Comparison between the new method and the 3-prefix preamble TL method for a given preamble length $L = 9$ and $P_L(min) = 10^{-3}$. c : number of active correlators.