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 portion 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 marker 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 systems.

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 PN 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 alarms" 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 one and zero, and are followed by envelope detectors (ED) with a detection threshold $b_i$. 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 $p(i)$ exceeds a threshold $b_i$. For our acquisition process, this may occur in any one of the $N_s$ 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 $b_i$ and samples the MF output once per code period. The coincidence detection is achieved when the current data symbol $d_i (i=1, 2, \ldots)$ plus the previous $N_s-1$ symbols match the $N_s$ symbols marker within a specified symbol error and erasure tolerance. Since it is not known a priori at which position in the $N_s$ preamble symbol acquisition occurred, the marker search (symbol-wise correlation) is performed in successive positions of the marker until a match is found or $N_s$ 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 $L_p$ symbols of user data.

The marker of length $N_s$ is chosen to be a sequence with minimum correlation sidelobes when preceded by the $N_s$ '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 [2] is a special case where $N_s=1$, 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 $N$, $N_k$, $L_p$, $b_0$, $b_1$, 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

$$r(t) = \sum_{k} A_{0k}(t - \tau - kT) + n_0 \cos \omega_{\text{ut}} - n_{0T} \sin \omega_{\text{ut}}$$

$$k = 0, 1$$

(1)

at the output of the bandpass filter, where $A$ is the amplitude of the signal, and $a(t)$, $a(t)$ are orthogonal spreading and time-limited to $[0, T]$ which represent ones and zeros of the binary message. $n_0$ and $n_{0T}$ 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 $\sigma^2 = N_0/T$. The sampling rate at the output of the matched filter is $1/T$. Let $N$ be the total number of chips in a PN code period, we have $T = NT$.

Considering the acquisition procedure in which the detector observes $a(t)$ over a period $0 < t < N_0T$. The test statistic at the output of the matched filters is

$$y(t) = (ANc(t) + n_0)^2 - n_{0T}^2 \sin \omega_{\text{ut}}$$

(2)

where

$$c(t) \equiv \left\{ \begin{array}{ll} \frac{1}{T} \int_0^T a(t)a(t - \tau) \, d\tau & t < T \\ \frac{1}{T} \int_0^T a(t)a(t + \tau) \, d\tau & t \geq T \end{array} \right.$$ (3)

is the normalized autocorrelation function of the PN sequence $a(t)$, and $n_0, n_{0T}$ 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 $P_{\text{fa}}$ caused by the noise. From [9], the output of the envelope detector is according to Rayleigh distribution when no signal is present. Accordingly, the probability of false alarm is

$$P_{\text{fa}} = \exp(-\frac{y_0^2}{2\sigma^2})$$

(4)

where $y_0$ is the detection threshold normalized to $\sigma$. Also from [9], the output of the envelope detector is according to Rayleigh distribution when signal plus noise is present. Thus the probability of initial acquisition in the correct position ($c(t/T) = 1$, $i = 1, 2, ..., N_0$) is given by

$$P_{\text{fa}} = Q\left(\sqrt{2\sigma^2} \cdot \frac{y_0}{\sigma}\right)$$

(5)

where $y_0 = \sigma_0^2 = N \cdot A^2/2\sigma^2 = \sum_{i=0}^{x_{\text{SN}}}(E_i + E_{\text{as}})$ 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

$$P_{\text{fa}k} = Q\left(\sqrt{2\sigma^2} \cdot \frac{y_0}{\sigma}\right)$$

(6)

where $k > 0$, $j = 1, ..., N_0 - 1$.

### 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 $N_k - 1$ acquisition bits may be lost due to the blocking, but synchronisation 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

$$P_L = 1 - (1 - P_0)P_{\text{fa}}$$

(8)

where $P_0$ is the probability of receiver blocking due to false alarms, and $P_{\text{fa}}$ is the probability of successful synchronization provided that the receiver is not blocked. $P_{\text{fa}}$ can be written

$$P_{\text{fa}} = \sum_{i=1}^{N_0} P_{\text{fa}}(i)$$

(9)

where $P_{\text{fa}}(i)$ is the probability of acquisition in the $i$th PN period.

$$P_{\text{fa}}(i)$$

(10)

as we mentioned, $P_{\text{fa}}(i)$ is the probability of coincidence when the acquisition occurs in the $i$th PN period.

To determine $P_L$, it remains to evaluate $P_{\text{fa}}(i), P_{\text{fa}}(i)$ and $P_0$.

### 3.2.1 Probabilities of initial acquisition $P_{\text{fa}}(i)$

As we mentioned earlier that the decision tests for acquisition are statistically independent. Thus the probability of initial acquisition $P_{\text{fa}}(i)$ at the $i$th PN period can be simply given by

$$P_{\text{fa}}(i) = \prod_{j=1}^{N_0 - 1} (1 - P_{\text{fa}j})$$

(11)

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 i-th peak.

### 3.2.2 Probability of coincidence detection

To calculate the second term $P_{\text{fa}}(i)$ in (11), we note that the coincidence test is carried out in up to $N_k$ positions of the marker.

The probability of successful coincidence detection $P_{\text{fa}}(i)$ 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 $P(h, h_0, n)$ 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 $h_0$ between the observed marker bits and the known or expected marker bits. The overlap $n$ is defined to be $N_k$ when there is complete overlap (i.e. correct position) and may be 0 or negative if there is no overlap. $P(h, h_0, n)$ is evaluated in [7] and given by
\[ P(h, h_a, n) = \text{Prob}(\text{at least one false marker} \text{ in} N_a \text{ marker tests}) \]

\[ = \sum_{j=0}^{N_a-1} \sum_{k=0}^{n} \sum_{m=0}^{n} \text{Pr}(h, h_a, n) \frac{\binom{n}{k} \cdot \binom{n}{m}}{\binom{N_a}{n}} \cdot \left( \frac{h_a(n)}{h_a(n) - m} \right) \]

\[ \cdot \left( \frac{h_a(n)}{h_a(n) - j - (j + m) + k} \right) \cdot \left( \frac{h_a(n)}{h_a(n) - m} \right)^{-1} \cdot \left( \frac{h_a(n)}{h_a(n) - j - (j + m) + k} \right)^{-1} \cdot \left( \frac{h_a(n)}{h_a(n) - m} \right)^{-2} \cdot \left( \frac{h_a(n)}{h_a(n) - j - (j + m) + k} \right)^{-2} \]

\[ \text{where} \quad h_a(n) = \frac{\binom{n}{m}}{\binom{N_a}{n}} \left( \frac{h_a(n) - m}{h_a(n) - m} \right) \cdot \left( \frac{h_a(n) - j - (j + m) + k}{h_a(n) - m} \right)^{-1} \cdot \left( \frac{h_a(n) - m}{h_a(n) - m} \right)^{-2} \cdot \left( \frac{h_a(n) - j - (j + m) + k}{h_a(n) - m} \right)^{-2} \]

\[ P_b \leq \frac{\mu_a}{1 + \rho} \quad (19) \]

Define \( \{a, s, p_a, p_b\} \) as the probabilities of decisions \( \{+, \), erasure, \(-\)\)], then

\[ P_{\text{Fm}} = \text{Pr}(\text{at least one false marker detection occurs in noise during} N_a \text{ marker tests}) \]

\[ \leq N_a P_{\text{Fm}} \cdot (\text{one test succeeds}) \]

\[ = N_a P_{\text{Fm}} \quad (20) \]

where

\[ P_{\text{Fm}} = \sum_{j=0}^{N_a-1} \sum_{k=0}^{n} \binom{n}{k} \cdot \binom{n}{m} \cdot \left( \frac{N_a}{N_a - j} \right)^{-1} \cdot \left( \frac{N_a}{N_a - j} \right)^{-2} \cdot \left( \frac{N_a}{N_a - j} \right)^{-3} \cdot \left( \frac{N_a}{N_a - j} \right)^{-4} \]

The calculations of \( s, p_a, \) and \( p_b \) are also given in the Appendix.

### 3.3 Selection of thresholds

From (8), since \( P_{\text{Fm}} \leq 1, P_b \) can never be less than

\[ P_b(\text{min}) = P_b \]

which is bounded by \( \rho/(1 + \rho) \). Thus the system designer can select a desired \( P_b(\text{min}) \) < 1 by an appropriate choice of \( \rho_b \) and \( b_b \). Typical values of \( P_b(\text{min}) \) may be in the range \( 10^{-3} \) to \( 10^{-4} \).

### 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 \( \rho_b \) is determined independent of the received signal power. Thus \( P_b \) 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 \( N_a = 1 \) 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 \( \rho_b \) is followed by the crossing of a higher threshold \( \rho_a = \rho_b + \Delta \), 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 \( L_D = 100 \). \( \rho_b \) and \( \rho_a \) are selected to get the required \( P_b(\text{min}) \). Performance comparison with the TL acquisition scheme at low SNR is made for \( L = 9 \). For \( L = 8 \), figure 4 shows \( P_b \) versus SNR for different \( N_a \) and \( N_b \) respectively with \( \rho_b \) and \( \rho_a \) chosen so that \( P_b(\text{min}) = 10^{-3} \). The double threshold method of [1] in used with \( \Delta = 4 \) dB. For each value of \( N_a, \rho_b \) and \( \rho_a \) were selected for the widest dynamic range. The best overall performance is obtained with \( (N_a, N_b) = (4, 4) \) 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 \( (N_a, N_b) = (1, 7) \), with only a slight reduction at high SNR.

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

A comparison between the proposed acquisition scheme and the TL code acquisition scheme [2, 3] is made in case of low SNR.

34.6.3.
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, $P_f$ for the single copy message leader of TL acquisition can be calculated using the same expression, as (8) with $P_f$ replaced by Erlang B formula $B(c,a)$, where $c$ is the number of correlators and $a = P_{th}M$ is the offered load. $M$ is the number of chips in the long code period. The probability $P_{syn}$ of successful synchronization is written

$$P_{syn} = \frac{P_fP_{at}}{Q(\sqrt{2}N_{th},h_1)Q(\sqrt{2}M_{th},h_1)} \quad (23)$$

The results are shown in figure 6 for the same $P_f(\min) = 10^{-3}$ and preamble length $L = 5$. 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 $P_f$. 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 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.

References


(a) PN+, PN+, PN+, PN+, PN+, PN+, PN+ & LD bias Data

Nh bias for acquision

Nc bias for coincidence

M chips PN codes 1, 2, & 3 corresponding to active correlators 1, 2 & 3.

(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.

$P_L$ versus SNR for different $N_A$.
a: $N_A = 3, N_c = 5, h_0 = 5.304, h_1 = 2.6, h = 0, h_p = 2$
b: $N_A = 4, N_c = 4, h_0 = 5.239, h_1 = 2.6, h = 0, h_p = 1$
c: $N_A = 5, N_c = 3, h_0 = 5.190, h_1 = 2.5, h = 0, h_p = 2$
d: $N_A = 6, N_c = 2, h_0 = 5.060, h_1 = 2.1, h = 0, h_p = 0$
e: $N_A = 1, N_c = 2, h_0 = 5.004, h_1 = 2.2, h = 1, h_p = 2$

$P_L$ versus SNR for different $N_A$.

$c$ : number of active correlators.

Fig. 4. $P_L$ versus SNR for different $N_A$.

$c$ : number of active correlators.

Fig. 5. $P_L$ versus SNR for different $P_L$(min), comparison of new and previous techniques with $N_A = 1$.

$N_A = 4, N_c = 4, h_1 = 2.6, h = 0, h_p = 1$

$P_L$(min) $= 10^{-3}, h_0 = 5.239$

$P_L$(min) $= 10^{-4}, h_0 = 5.662$

$P_L$(min) $= 10^{-5}, h_0 = 6.005$

$P_L$(min) $= 10^{-6}, h_0 = 6.424$

$N_A = 1, N_c = 2, h_1 = 2.2, h = 1, h_p = 2$

$P_L$(min) $= 10^{-3}, h_0 = 5.349$

$P_L$(min) $= 10^{-4}, h_0 = 5.793$

$P_L$(min) $= 10^{-5}, h_0 = 6.150$

$P_L$(min) $= 10^{-6}, h_0 = 6.514$

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.

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