To appear in the *Proceedings of the 2002 IEEE Military Communications Conference*, (Anaheim, CA), Oct 2002.

PERFORMANCE OF SERIAL, MATCHED-FILTER ACQUISITION WITH ADAPTIVE THRESHOLDS IN DIRECT-SEQUENCE PACKET COMMUNICATIONS

Arvind Swaminathan and Daniel L. Noneaker (corresponding author)
Department of Electrical and Computer Engineering
305 Fluor Daniel Engineering Innovation Building
Clemson University, Clemson, SC 29634

***Abstract*— In this paper we examine a technique to improve the performance of serial, matched-filter acquisition in direct-sequence spread-spectrum packet radio communications. Each packet transmission includes an acquisition preamble, and the preamble sequence is changed at the boundaries of predefined time epochs based on a pseudorandom sequence generator. The acquisition algorithm considered in this paper exploits knowledge of the preamble sequence employed during each time epoch to adaptively choose the acquisition threshold. The performance of the adaptive-threshold acquisition algorithm is evaluated and compared with one previously considered [1] in which the threshold is fixed. It is shown that the adaptive-threshold algorithm results in good average performance, even with a large *a priori* uncertainty in the signal-to-noise ratio. It is also shown that the average performance with the adaptive threshold is superior to that obtained with the fixed-threshold algorithm of [1].**

I. INTRODUCTION

Direct-sequence (DS) spread-spectrum modulation is used widely in both commercial and military communications. It is employed in cellular code-division multiple-access (CDMA) networks, as well as tactical radio networks for the military. Packet radio communications is necessary to achieve good operation in a DS multiple-hop radio network and it has been incorporated into the designs of third-generation and later commercial cellular networks.

In order to successfully receive a DS packet transmission, the receiver must synchronize a locally generated copy of the spreading sequence with the same sequence of the arriving packet. Synchronization is usually performed in two stages: acquisition and tracking. The acquisition stage provides a coarse alignment between the local sequence and the sequence of the arriving signal. This is necessary because of the large *a priori* timing uncertainty of the arriving signal. It has been shown that acquisition can be a significant factor that limits the performance of a DS system employing random-access packet radio communications.

Each transmission in a DS packet radio system includes a preamble that consists of a segment of the spreading sequence of predetermined duration with no data modulation, and the spreading sequence used for each packet transmission is known *a priori* at the receiver. The most effective techniques for acquiring a DS packet transmission are based on the receiver employing a filter matched to the preamble, and that is the approach considered in this paper. In a practical system design, security concerns dictate that the spreading sequence is changed periodically. This can be accomplished by employing a long-period pseudorandom sequence generator and periodically varying the initial sequence phase used for packet transmissions. In this paper, we consider a DS packet radio system in which the preamble sequence is varied periodically and in which the pseudorandom generator determining the preamble in any time epoch is known *a priori* at the receiver.

We examine a technique to improve the performance of serial, matched-filter acquisition in direct-sequence spread-spectrum packet communications. The research is motivated by the results obtained in [1] which account for the effect of the intermediate-frequency (IF) filter and the subsequent automatic gain-control (AGC) subsystem on the performance of a serial acquisition algorithm in a heterodyne receiver. The acquisition algorithm uses the threshold crossing of a matched-filter output to detect the presence of the preamble in each packet. The receiver considered in [1] employs a fixed acquisition threshold that is chosen by optimizing the average performance over all the possible preamble sequences. It is shown in [1] that the interdependence among the IF stage and the AGC subsystem results in a probability of not acquiring that is a non-monotonic function of the signal-to-noise ratio (SNR) of the received signal. In this paper, we consider a modification of the acquisition algorithm in which the receiver exploits knowledge of the preamble sequence employed during each time epoch to adaptively choose the acquisition threshold.

The paper is organized as follows. In Section II, we give a description of the communication system. An analysis of the performance of the serial, matched-filter acquisition technique is presented in Section III. In Section IV we describe the adaptive threshold-selection algorithm, and the improvement in performance that results with the adaptive algorithm is illustrated in Section V. Conclusions are given in Section VI.

II. SYSTEM DESCRIPTION

A. Transmitted Signal and Received Signal

Each packet transmission consists of a preamble of M chips followed by the packet's data content. The spreading sequence employed for transmission of a packet during a given time epoch is modeled as randomly selected from among all sequences of M chips with equal probability. The spreading se-

This work was supported in part by the DoD Multidisciplinary University Research Initiative (MURI) program administered by the Office of Naval Research under Grant N00014-00-1-0565 and in part by the Army Research Office under grant number DAAD19-00-1-0156 .

quence used during any time epoch is known *a priori* at the receiver. The baseband equivalent of the transmitted signal [1] is given by

$$\tilde{s}(t) = \sqrt{P} c(t) \left[p_{MT_c}(t) + d(t) p_{LT_c}(t - MT_c) \right],$$

where P is the power in the transmitted signal during the preamble, $p_T(t)$ is the rectangular pulse with height 1 over $[0, T)$ and $d(t)$ is the data content of the packet. The quaternary phase-shift-keyed (QPSK) spreading waveform [2] is given by

$$c(t) = \sum_{i=0}^{M+L-1} a_i \psi(t - iT_c)$$

where $\{a_i\}$ represents a complex-valued spreading sequence [2] and each a_i is a random variable uniformly distributed over the values in $\{e^{j(pi/4)}, e^{j(3\pi/4)}, e^{j(5\pi/4)}, e^{j(7\pi/4)}\}$. This *random-sequence* model [3] provides a good approximation to a long-period complex-valued, pseudonoise spreading sequence. The chip waveform $\psi(t)$ is time limited to $[0, T_c)$ and

$$\frac{1}{T_c} \int_{t=0}^{T_c} |\psi(t)|^2 dt = 1. \tag{1}$$

Without loss of generality, the transmitted signal is assumed to arrive unattenuated and undelayed at the receiver over an additive white Gaussian noise channel. Thus the baseband equivalent of the received signal is

$$\tilde{r}(t) = \tilde{s}(t) + \tilde{n}(t) \tag{2}$$

The complex-valued baseband-equivalent white Gaussian-noise voltage $\tilde{n}(t)$ has two-sided power spectral density N_0.

B. Receiver Architecture

The receiver stage is modeled as an AWGN source, a band-pass filter, and an AGC system followed by an acquisition stage. The impulse response of the IF filter is given by

$$h(t) = 2\mathcal{R}e\left\{ \tilde{h}(t) e^{j(2\pi f_c t)} \right\}$$

where $\tilde{h}(t)$ is the baseband-equivalent impulse response of the filter and f_c is the carrier frequency at the transmitter. Without loss of generality, the signal $r(t) = \sqrt{2}\mathcal{R}e\left\{\tilde{r}(t) e^{j2\pi f_c t}\right\}$ is taken as the received signal after demodulation to IF.

A well-designed AGC subsystem achieves minimal variation in its average output over a wide dynamic range for the input power. In this paper, we approximate such a design by considering an AGC subsystem that responds instantly to a step change in the average steady-state input power and maintains a constant steady-state output power. Thus the signal at the AGC output $\hat{r}(t)$ is given by

$$\hat{r}(t) = \sqrt{\alpha(t)} (r * h)(t) \tag{3}$$

where $1/\alpha(t)$ is the average steady-state power at the input of the AGC subsystem at time t. In [1], an expression is developed that characterizes $\alpha(t)$ in terms of the input signal power, the noise power spectral density, and a set of parameters that depend on the chip waveform and IF filter's impulse response. It is shown that

$$\alpha(t) = \begin{cases} \alpha_0 = \left(\gamma_s \beta \frac{N_0}{T_c} \right)^{-1}, & t < 0 \\ \alpha_1 = \gamma_s^{-1} \left(P + \beta \frac{N_0}{T_c} \right)^{-1}, & 0 \le t \le MT_c. \end{cases} \tag{4}$$

The parameters in (4) are given by

$$\gamma_s = \int_{-\infty}^{+\infty} \frac{|\Psi(f)|^2}{T_c} |\tilde{H}(f)|^2 df, \tag{5}$$

$$\gamma_n = T_c \int_{-\infty}^{+\infty} |\tilde{H}(f)|^2 df, \tag{6}$$

$$\text{and } \beta = \frac{\gamma_n}{\gamma_s}, \tag{7}$$

where $\Psi(f)$ is the fourier transform of the chip waveform $\psi(t)$ and $\tilde{H}(f)$ is the baseband equivalent of the IF filter impulse response.

C. Acquisition Algorithm

The acquisition stage employs noncoherent square-law combining of the filter outputs to form the test statistics. The in-phase and quadrature branches of the receiver each contain two filters, denoted $g_1(t)$ and $g_2(t)$. The outputs in each branch at time t are summed to form

$$U(t) = \mathcal{R}e\left\{(y * g)\right\} \tag{8}$$

$$V(t) = \mathcal{I}m\left\{(y * g)\right\} \tag{9}$$

where $y(t) = \sqrt{\alpha(t)}(\tilde{r} * \tilde{h})(t)$ and $g(t) = g_1(t) + jg_2(t)$. The receiver samples the outputs of filters at times $t = iT_c, i$ an integer. The test statistics $X_i = U_i^2 + V_i^2$ are compared with a threshold η. If $X_i > \eta$, a *hit* is declared and a locally generated copy of the spreading waveform is then synchronized to the delay $(i - M)T_c$ and this waveform is used to demodulate the data symbols.

If the detector fails to declare a *hit* until after the end of a packet's acquisition preamble, a *miss* occurs and the packet is not acquired. Conversely, if the detector declares a hit at least one chip interval before the end of the preamble, a *false alarm* occurs. In this paper the *false-alarm processing interval*, the amount of time required for the detector to detect a false alarm and return to the acquisition mode, is taken to be a constant Q. Hence a packet is not acquired if either a false alarm occurs for $X_i, M - Q \le i \le M - 1$, or if a miss occurs.

The filters for the acquisition stage that are considered in the analysis are those having the form

$$g(t) = \sum_{i=0}^{M-1} a_{M-1-i}^* \varphi(t - iT_c) \tag{10}$$

where $\{a_0, \cdots, a_{M-1}\}$ is the spreading sequence for the acquisition preamble. Furthermore, the impulse response of the baseband-equivalent IF filter $\tilde{h}(t)$ and the function $\varphi(t)$ that are considered satisfy the joint constraint

$$\left(\tilde{h} * \varphi \right)(t) = \psi^*(T_c - t), \qquad \text{for all } t. \tag{11}$$

Under this constraint, the convolution of the IF filter's baseband equivalent impulse response with the complex-valued filter in the acquisition stage is matched to the acquisition preamble in the DS signal.

III. Analysis of Acquisition Performance

The analysis in [1] can be generalized to obtain the statistical characterization of $\{U_i, V_i\}$ for all i. If $SNR = \frac{MPT_c^2}{N_0}$ then it can be shown that

$$E[U_i + jV_i] = \begin{cases} 0, & i \le 0 \\ \frac{T_c C_A(M-i)}{\sqrt{\gamma_s(1+\frac{M\beta}{SNR})}}, & 1 \le i < M \\ \frac{T_c M}{\sqrt{\gamma_s(1+\frac{M\beta}{SNR})}}, & i = M \end{cases} \tag{12}$$

where $C_A(i)$ is the aperiodic autocorrelation function [2] of the sequence $\{a_0, \cdots, a_{(M-1)}\}$. Also

$$Var(U_i) = \begin{cases} \frac{MT_c^2}{2\gamma_s\beta}, & i \le 0 \\ \frac{(M-i)T_c^2}{2\gamma_s\beta} + \frac{MiT_c^2}{2\gamma_s(SNR+M\beta)}, & 1 \le i \le M \end{cases}, \tag{13}$$

$Var(V_i) = Var(U_i)$, and $\{U_{(M-Q)}, V_{(M-Q)}, \cdots, U_M, V_M\}$ are mutually uncorrelated.

Using the same approach as in [1] we can approximate the joint distribution function of the collection of uncorrelated random variables $\{U_{(M-Q)}, V_{(M-Q)}, \cdots, U_M, V_M\}$ by treating them as jointly Gaussian, and therefore independent, random variables with first and second moments given by (12) and (13). Hence, the test statistics $(X_{M-Q}, \ldots, X_0)$ are central chi-square random variables with two degrees of freedom and the test statistics $(X_1, \ldots, X_{M-1})$ are non-central chi-square random variables with two degrees of freedom.

As described in the previous section, acquisition occurs if and only if $X_M > \eta$ and $X_j < \eta, M-Q \le j \le M-1$, where η is the detection threshold. The probability that a false alarm occurs during the false-alarm processing interval is given by

$$\begin{aligned} \mathrm{P}_{fa} &= 1 - \mathrm{P}(X_{M-1} < \eta, \cdots, X_{M-Q} < \eta) \\ &= 1 - \left(\prod_{i=M-Q}^{0} \left[1 - \exp\left(\frac{-\eta}{2\sigma_i^2} \right) \right] \right. \\ &\qquad \left. \times \prod_{i=1}^{M-1} \left[1 - Q\left(\frac{\mu_i}{\sigma_i}, \frac{\sqrt{\eta}}{\sigma_i} \right) \right] \right) \end{aligned} \tag{14}$$

For convenience, we modify the definition of a *miss* be the event $\{X_M < \eta\}$ regardless of whether or not a false alarm occurs. Then

$$\begin{aligned} \mathrm{P}_{miss} &= 1 - \mathrm{P}(X_M \ge \eta) \\ &= 1 - Q\left(\frac{\mu_M}{\sigma_M}, \frac{\sqrt{\eta}}{\sigma_M} \right) \end{aligned} \tag{15}$$

where $\mu_i = |E[U_i + jV_i]|$ is calculated using (12), and $\sigma_i = \sqrt{\mathrm{Var}(U_i)}$ from (13), and $Q(.,.)$ is the Marcum Q-function [4]. The probability of not acquiring a packet can be expressed as

$$\begin{aligned} \mathrm{P}_{nacq} &= 1 - \mathrm{P}(X_M \ge \eta, X_{M-1} < \eta, \cdots, X_{M-Q} < \eta) \\ &= 1 - (1 - \mathrm{P}_{miss})(1 - \mathrm{P}_{fa}) \\ &= \mathrm{P}_{miss} + \mathrm{P}_{fa} - \mathrm{P}_{miss}\mathrm{P}_{fa} \end{aligned} \tag{16}$$

IV. Threshold Selection

In this section we describe the algorithm for threshold selection. In [1] a fixed threshold is chosen by optimizing the performance averaged over all preamble sequences, and the performance is evaluated for a receiver that employs the fixed threshold with a time-varying preamble. However, the receiver knows *a priori* the specific preamble sequence that is used during any time epoch. We evaluate a technique in which the receiver exploits this information to adapt the threshold for the specific preamble in use in a given time epoch.

A. Bounds on the Distribution function of the Non-Central Chi-Square Random Variable

In this subsection we develop useful bounds on the distribution function of a non-central chi-square random variable with two degrees of freedom. The random variable can be expressed as $X = U^2 + V^2$, where U and V are independent Gaussian random variables with $\sqrt{(E[U])^2 + (E[V])^2} = \mu$ and equal variance σ^2. We employ three intermediate functions in describing the bounds. Let

$$\begin{aligned} P_w(R, \phi) &= \frac{\phi}{2\pi}(1 - \exp(-\frac{R^2}{2\sigma^2})), \\ R_l(x) &= \frac{|x|\mu}{\sqrt{x^2+1}} + \sqrt{\frac{x^2\mu^2}{x^2+1} + \eta - \mu^2}, \end{aligned}$$

and

$$R_s(x) = \frac{-|x|\mu}{\sqrt{x^2+1}} + \sqrt{\frac{x^2\mu^2}{x^2+1} + \eta - \mu^2}.$$

1) Lower Bound: Let $\Gamma = \{(U, V) | U^2 + V^2 \le \eta\}$. Then it can be shown that

$$\begin{aligned} &P(X \le \eta) \ge P_w(R_s(\cot(0)), 2 \times \pi/8) \\ &\quad + 2P_w(R_s(\cot(\frac{\pi}{8})), \frac{\pi}{4}) + 2P_w(R_s(\cot(\frac{3\pi}{8})), \frac{\pi}{8}) \\ &\quad + 2P_w(R_l(\cot(\frac{\pi}{2})), \frac{\pi}{4}) + P_w(R_s(\cot(\frac{3\pi}{4})), \frac{2\pi}{4}). \end{aligned}$$

Furthermore, the same approach can be extended to obtain an asymptotically accurate sequence of lower bounds.

2) Upper Bounds: A similar approach can be used to obtain a closed-form expression for an upper bound on $P(X \le \eta)$. It can be shown that for $\mu^2 \le \eta$

$$\begin{aligned} &P(X \le \eta) = P((U, V) \in \Gamma) \\ &\quad \le P_w(R_s(\cot(\frac{\pi}{8})), 2 \times \pi/8) + 2P_w(R_s(\cot(\frac{\pi}{4})), \frac{\pi}{8}) \\ &\quad + 2P_w(R_s(\cot(\frac{3\pi}{8})), \frac{\pi}{8}) + 2P_w(R_l(\cot(\frac{\pi}{2})), \frac{\pi}{8}) \\ &\quad + 2P_w(R_l(\cot(\frac{3\pi}{4})), \frac{\pi}{4}) + P_w(R_l(\cot(0), \frac{2\pi}{4}). \end{aligned}$$

The same approach can also be used to obtain an asymptotically accurate sequence of upper bounds. A similar approach can be employed if $\mu^2 > \eta$. It can be shown that

$$\begin{aligned} P(X \le \eta) \le 2 \times \Big[& P_w(R_2^2, \frac{\phi}{2}) - P_w(R_1^2, \frac{\phi}{2}) \\ & + P_w((\mu - \sqrt{\eta})^2, \frac{\phi}{2}) - P_w((\mu + \sqrt{\eta})^2, \frac{\phi}{2}) \Big] \end{aligned}$$

where

$$\begin{aligned}\phi &= \tan^{-1}\left(\sqrt{\frac{\eta}{\mu^2-\eta}}\right)\\ R_1 &= \frac{x\mu}{\sqrt{x^2+1}} - \sqrt{\eta - \frac{\mu^2}{x^2+1}}\\ R_2 &= \frac{x\mu}{\sqrt{x^2+1}} + \sqrt{\eta + \frac{\mu^2}{x^2+1}}\end{aligned}$$

and $x = \cot(\frac{\phi}{2})$.

B. Adaptive Threshold-Selection Algorithm

Since the signal-to-noise ratio of the incoming transmission is not known *a priori*, the threshold chosen should provide good performance over the range values of SNR over which the system is designed to operate. In any practical application, there is some smallest value of preamble signal-to-noise ratio (denoted SNR_{min}) for which it is necessary to achieve a low probability of not acquiring. (For example, SNR_{min} might be determined from the smallest signal-to-noise ratio in the data symbols that results in acceptable data-detection performance after acquisition.) Hence, the objective we consider for the threshold-selection algorithm is to obtain a small worst-case probability of not acquiring for $\text{SNR} \geq \text{SNR}_{min}$.

The algorithm is designed to select the threshold that results in approximately equal probabilities of not acquiring for $SNR = SNR_{min}$ and as $SNR \to \infty$ for each preamble sequence that is used. Note that P_{nacq} at SNR_{min} is approximately equal to P_{miss}, P_{nacq} as $SNR \to \infty$ is approximately equal to P_{fa}, and both of these quantities can be expressed in terms of the distribution function of non-central chi-square random variables using (14) and (15). The iterative algorithm employs a predetermined initial range of values for the search region for the threshold. Let $P_{miss}(\eta, SNR)$ denote the probability of miss with a threshold η at a signal-to-noise ratio SNR, and let $P_{fa}(\eta, SNR)$ denote the corresponding probability of false alarm. The difference $\delta(\eta) = P_{miss}(\eta, SNR_{min}) - P_{fa}(\eta, \infty)$ is evaluated for the threshold at the middle of the range. The search region is then divided in half by choosing either the upper or lower half of the current region depending on the sign of $\delta(\eta)$. The quantities $P_{miss}(\eta, SNR_{min})$ and $P_{fa}(\eta, \infty)$ are approximated in closed form by upper bounds based on the bounds for the distribution function presented in the previous subsection. Since, the upper bound on P_{miss} is an increasing function of the threshold and the upper bound on P_{fa} is a decreasing function of the threshold, for a given signal-to-noise ratio, $\delta(\eta)$ is a monotonic function of η and hence has a single zero crossing. This is sufficient to guarantee that the algorithm converges provided that the initial search region is chosen to include the zero crossing of $\delta(\eta)$.

V. Results

In this section, we compare the performance that results with an optimal fixed threshold and the performance that results with the adaptive-threshold algorithm introduced in the previous section. All numerical results in the paper are for a system with a preamble of 400 chips and a chip waveform and receiver filters that result in $\beta = 2.08$. The false alarm processing interval is 1000 chips, and thresholds are selected to yield good performance for $SNR \geq 16$dB, unless otherwise specified.

Consider first a receiver that employs a fixed acquisition threshold (i.e., a threshold that is not varied as the preamble sequence is changed), for which the performance for a given SNR is characterized by the probability of not acquiring averaged over all possible preamble sequences. If the objective is to minimize the worst-case average probability of not acquiring for $SNR \geq SNR_{min}$, it is shown in [1] that the optimal fixed threshold results in the same average probability of not acquiring for $SNR = SNR_{min}$ and $SNR = \infty$. Quite varied performance is observed with different preamble sequences if the optimal fixed threshold is used, however. If the preamble sequence has good autocorrelation properties, the probability of not acquiring is much higher for $SNR = SNR_{min}$ than that for $SNR = \infty$, as illustrated in Fig. 1. In contrast, if the sequence has poor autocorrelation properties, the probability of not acquiring is higher for $SNR = \infty$ than for $SNR = SNR_{min}$, as illustrated in Fig. 2 for $SNR_{min} = 16dB$.

For a given sequence the performance in the low signal-to-noise ratio region is improved if the threshold is decreased, while in the high signal-to-noise ratio region the performance is improved if the threshold is increased. Indeed, optimal performance is obtained with most preamble sequences if the threshold is selected to ensure that the probability of not acquiring with that preamble sequence is the same for $SNR = SNR_{min}$ and $SNR = \infty$. (A modification of the adaptive algorithm can be used to account for the small fraction of sequences for which this is not true, but it is not addressed in this paper.) The adaptive threshold-selection algorithm achieves this by adapting the threshold to the current preamble sequence

A. Performance Comparison for Individual Sequences

The improvement in performance that results with the adaptive-threshold algorithm is illustrated in Figs. 1-3. The probability of not acquiring with the optimal fixed threshold is compared the probability of not acquiring with the adaptive threshold-selection algorithm. The sequence considered in Fig. 1 has good autocorrelation properties, and the threshold selected by the adaptive threshold-selection algorithm is smaller than the fixed threshold. This results in improved performance at low signal-to-noise ratios as compared to the fixed threshold. For this sequence the worst-case probability of not acquiring with a fixed threshold is 0.0089 at $SNR = SNR_{min}$. The worst-case probability of not acquiring with the adaptive threshold is 0.0015, which occurs at both $SNR = SNR_{min}$ and $SNR = \infty$. In Fig. 2 and Fig. 3 the sequences have poor autocorrelation properties, and in each case the adaptive threshold-selection algorithm selects a threshold that is higher than the fixed threshold. As a result, the adaptive threshold results in reasonable performance at high signal-to-noise ratios while the fixed threshold results in poor performance. In particular, in Fig. 3 the autocorrelation properties of the sequence are sufficiently poor that the the probability of not acquiring approaches one at high signal-to-noise ratios with the optimal

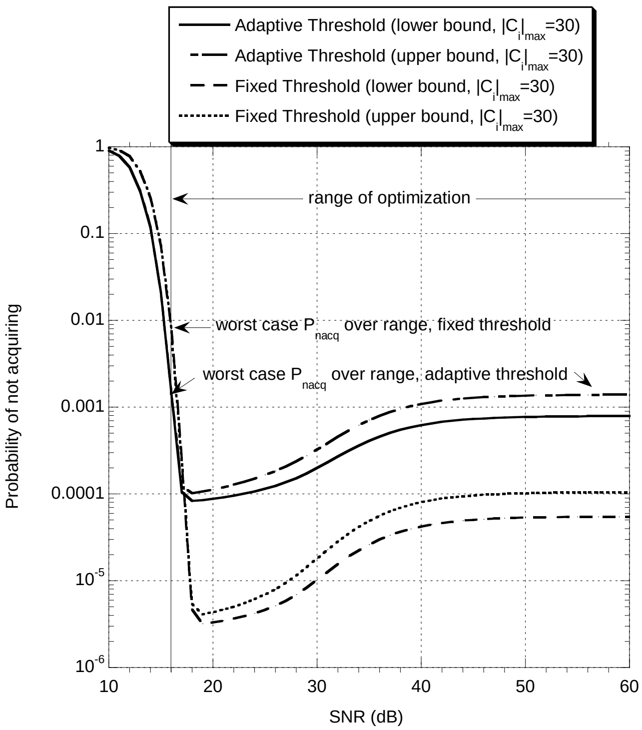

Fig. 1. Acquisition performance for a sequence with maximum autocorrelation sidelobe magnitude of 30 and $SNR_{min} = 16$ dB.

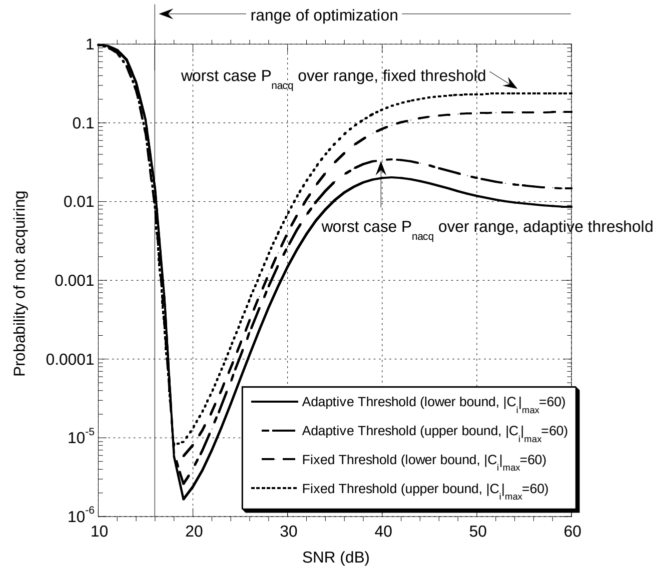

Fig. 2. Acquisition performance for a sequence with maximum autocorrelation sidelobe magnitude of 60 and $SNR_{min} = 16$ dB.

fixed threshold. Note that the adaptive-threshold algorithm results in a lower worst-case probability of not acquiring over $SNR \geq 16$ dB for each of the three sequences. The worst case probability of not acquiring is 0.0014, 0.035 and 0.1 with the adaptive threshold for sequences with $|C_i|_{max} = 30, 60$, and 70, respectively, while the corresponding values with the optimal fixed threshold are 0.0089, 0.25 and 1.0, respectively. Hence the worst-case probability of not acquiring is almost an order of magnitude greater with the optimal fixed threshold.

B. Average Performance

The average performance obtained with the optimal fixed threshold and with the adaptive threshold-selection algorithm is characterized in two ways. In Fig. 4, the probability of not acquiring averaged over all preamble sequences is shown for the optimal fixed threshold and the adaptive-threshold algorithm if $SNR_{min} = 16$ dB. If the optimal fixed threshold is used, the worst-case probability of not acquiring for $SNR \geq 16$ dB is 0.0051. But if the adaptive-threshold algorithm is used, the worst-case probability of not acquiring for $SNR \geq 16$ dB is only 0.00215. Thus the adaptive-threshold algorithm results in almost a two-and-one-half-fold decrease in the worst-case probability of not acquiring over the range of SNR of interest.

Suppose instead that the greatest acceptable probability of not acquiring is 0.0051. From the result discussed above, the performance objective is achieved with a fixed threshold for $SNR \geq 16$ dB if the the threshold is optimized over that range. If instead, the adaptive-threshold algorithm is used with $SNR_{min} = 15.35$ dB in the algorithm, the same performance objective is achieved for $SNR \geq 15.35$ dB. Thus the use of the adaptive-threshold algorithm allows the range of SNR over which the performance objective is achieved to be expanded by 0.65 dB. This is illustrated in Figs. 5 and 6.

VI. Conclusions

In this paper, we present an adaptive threshold-selection algorithm for use with serial, matched-filter acquisition in DS spread-spectrum packet radio communications. The algorithm adapts the threshold based on the autocorrelation function of the preamble spreading sequence and the range of signal-to-noise ratios over which acquisition performance is to be optimized. It is shown that the acquisition performance that results with adaptive threshold selection is better than the performance that results if the optimal fixed threshold is used.

References

[1] Daniel L. Noneaker, Arvind R. Raghavan, and Carl W. Baum, “The effect of automatic gain control on serial, matched-filter acquisition in direct-sequence packet radio communications,” *IEEE Trans. Vehicular Technol.*, vol. 50, no. 4, pp. 1140-1150, July 2001.

[2] D.V. Sarwate and M.B. Pursley, “Crosscorrelation properties of pseudo-random and related sequences, ” *Proc. IEEE*, vol. 68, no. 5, pp. 593-619, May 1980.

[3] K.K. Chawla and D.V. Sarwate, “ Acquisition of PN sequences in chip synchronous DS/SS systems using a random sequence model and the SPRT, ” *IEEE Trans. Commun.*, vol. 42, no. 6, pp.2325-2334, June 1994.

[4] M. Schwartz, W.R. Bennett, and S. Stein, *Communication Systems and Techniques*. New York: McGraw-Hill, 1966.

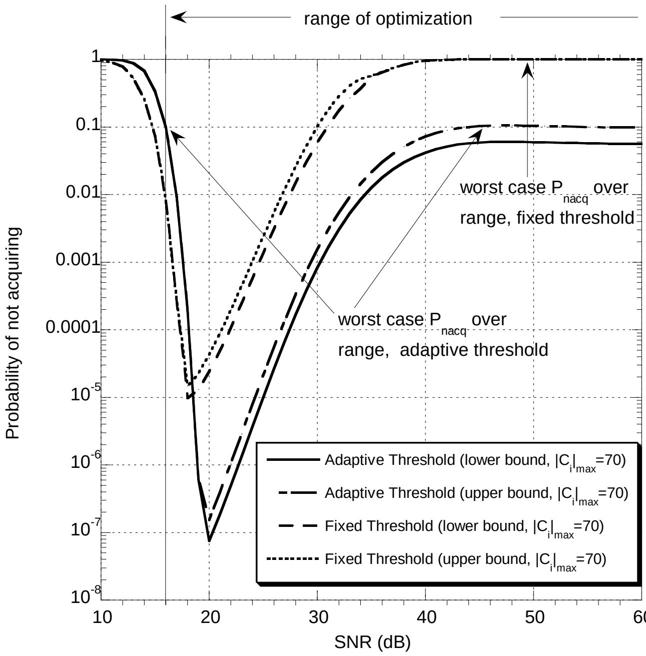

Fig. 3. Acquisition performance for a sequence with maximum autocorrelation sidelobe magnitude of 70 and $SNR_{min} = 16$ dB.

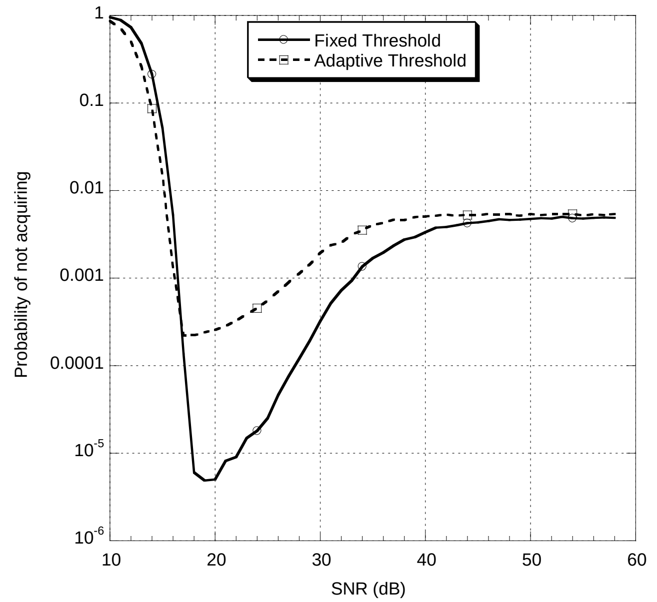

Fig. 5. Acquisition performance averaged over all sequences with worst-case P_{nacq} = 0.0051.

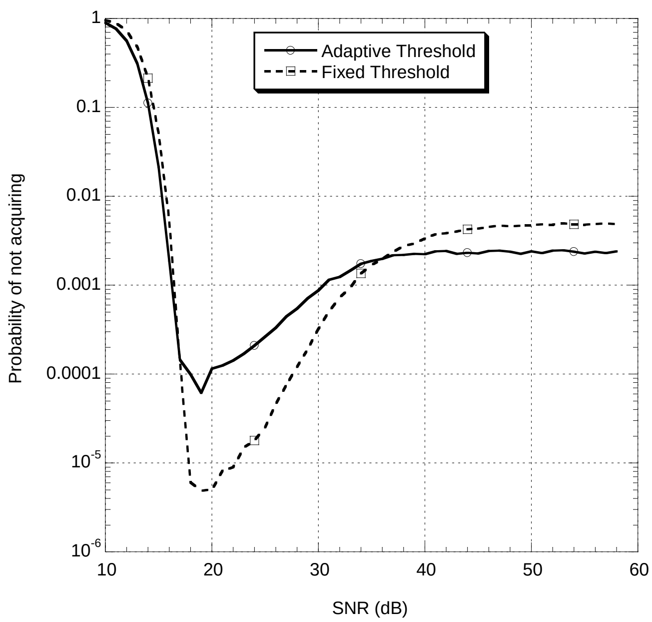

Fig. 4. Acquisition performance averaged over all sequences with $SNR_{min} = 16$ dB.

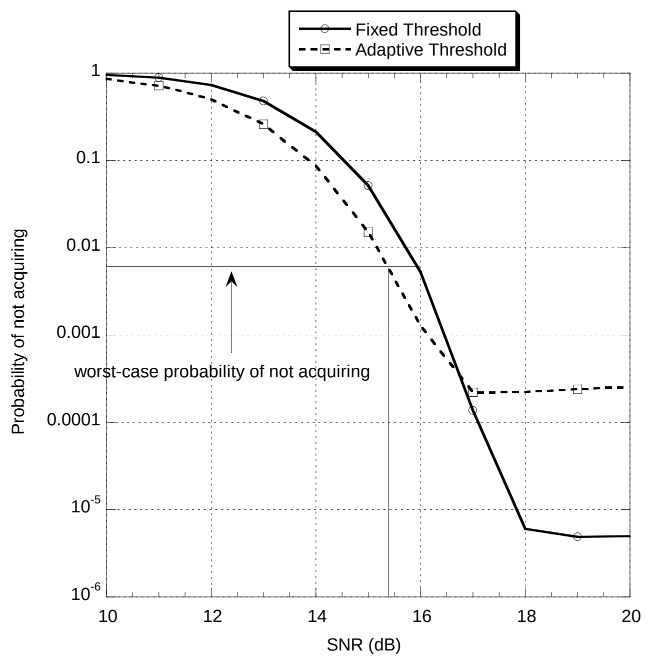

Fig. 6. Acquisition performance averaged over all sequences with worst-case P_{nacq} = 0.0051. (enlarged x-axis)