**ADAPTIVE THRESHOLDING IN CODE ACQUISITION OF**

**DIRECT-SEQUENCE SPREAD SPECTRUM SIGNALS**

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**Abstract*—*Spread spectrum techniques have found extensive applications in communications systems for a variety of reasons, including anti-interference, low detectability and multiple access. In direct sequence spread spectrum systems, it is required that the locally generated pseudo noise (PN) signal in the receiver be synchronized to the received PN signal. This is done in two steps: acquisition (coarse alignment) and tracking (fine alignment). This paper discusses different methods of code acquisition and in particular focuses on methods of adaptive thresholding in code acquisition of direct-sequence spread spectrum signals.**

**I-INTRODUCTION**

Spread spectrum communication signals have been used in military systems for decades because of their ability to reject interference. The interference can be unintentional when another transmitter tries to transmit simultaneously through the channel, or intentional when a hostile transmitter attempts to jam the transmission. By definition, for a communication system to be considered spread spectrum, it must satisfy two conditions. First, the bandwidth of the transmitted data must be much greater than the message bandwidth. Second, the system spreading is accomplished before transmission by some function (e.g., code or a PN sequence) that is independent of the message but known to the receiver. This same code is then used at the receiver to despread the signal so that the original data may be recovered. The two main modulating techniques in spread spectrum communication systems are direct-sequence (DS) or pseudonoise (PN) spread spectrum, and frequency-hop (FH) spread spectrum. A *pseudorandom* or a *pseudonoise sequence*, which is a noiselike spreading code, is used to transform the narrowband data sequence into a wideband sequence. Then, the resulting wideband signal undergoes a second modulation using phase shift keying (PSK) techniques. In frequency-hopping spread spectrum, the information sequence bandwidth is still widened by a pseudonoise sequence but with a changing carrier frequency. Spread spectrum signals appear like random noise, which makes them difficult to demodulate by receivers other than the intended ones, or even difficult to detect in the presence of background noise. Thus, spread spectrum systems are not useful in combating white noise, but have important applications such as antijam capabilities and interference rejection. Interference arises also in multiple access communication, in which a number

of independent users share a common channel. The conventional way to provide multiple access communication uses frequency division multiple access (FDMA) or time division multiple access (TDMA) communication. In FDMA, each user is assigned a particular frequency channel, which presents a fraction of the channel bandwidth until system capacity is reached, when the whole bandwidth is used. In TDMA, the channel time-bandwidth is apportioned into fixed time slots. Each user is assigned a particular time slot until capacity is reached, when all time slots are used. A more efficient way to accomplish multiple access communications is code division multiple access (CDMA). In CDMA, each user is assigned a particular code, which is either a PN sequence or a frequency-hopping pattern, to perform the spread spectrum modulation. Since each user has its own code, the receiver can recover the transmitted signal by knowing the code used by the transmitter. However, each code used must be approximately orthogonalto all other codes; that is, it must have low cross-correlation. CDMA offers secure communication privacy, due to the fact that the messages intended for one user may not be decodable by other users because they may not know the proper codes. In addition, as the number of users increases beyond a certain threshold, a gradual degradation in the performance is tolerated, and thus CDMA can accommodate more users. Because of its low power level, the spread spectrum signal may be hidden in the background noise, and in this case it is called “*covert*.” It has a low probability of being detected and is called a low probability of intercept (LPI) signal. Because of the above advantages, DS-CDMA became in the late 1980s increasingly of interest in cellular type communications for commercial purposes [1].

The receiver for binary DS-CDMA signaling schemes can have one of two equivalently performing structure, a correlator implementation and a matched filter implementation. The correlator receiver performs a correlation operation with all possible signals sampling at the end of each *T*- signaling interval and comparing the outputs of the correlator. In the matched filter receiver, correlators are placed by matched filters. In order to be sure of the successful connection in direct sequence spread spectrum systems, it is necessary to make synchronization between the transmitter and the receiver [2]. The process of synchronizing the local code and the received code is commonly achieved in two stages: initially, the two code signals are aligned in phase to uncertainty less than one chip duration through a process called *code acquisition* or coarse synchronization. In other words, the acquisition is aligning the unknown phase of the received code with the known phase of the local code generated at the receiver. Once the incoming code is acquired, a verification process attests the correct code phase which is continuously maintained by a closed loop tracking system. However, if for some reason the tracking system has gone out of lock, the acquisition system will be re-activated in order to acquire the incoming code and the tracking system takes over again to maintain code synchronization [3].

In CDMA systems, Multiple Access Interference (MAI) is the main problem and has a negative effect on both simple correlator and matched filter method. Therefore the capacity of such system would be limited by the number of users accessing the code synchronization at the same time and not by the number of users accessing the specific BER performance. Therefore improving the acquisition performance in CDMA using advanced acquisition designs can improve the real capacity of the system for access to the capacity based on BER [2].

This paper is organized as follows: Section II provides a brief description of spread spectrum signals in digital communication systems. Different methods of PN code acquisition for DS-SS are presented in section III. Section IV is devoted to adaptive thresholding in code acquisition of DS-SS and section V concludes the paper.

**II-SPREAD SPECTRUM SYSTEMS**

As discussed in previous chapter, the two main modulating techniques in spread spectrum communication systems are direct-sequence (DS) or pseudonoise (PN) spread spectrum and frequency-hop (FH) spread spectrum. In this section, we give a brief description of these modulation techniques [1].

1. ***Direct-Sequence Spread Spectrum Modulation (DS-SS)***

One way of widening the bandwidth of the information-bearing signal is bymodulationof the PN sequence on the spread spectrum carrier, which

can be binary phase-shift keying (BPSK). First, the binary message *m*(*t*) and the PN sequence *p*(*t*) are applied to a product modulator. Since the information sequence *m*(*t*) is narrowband and the PN sequence is wideband, the product signal *s*(*t*) will have a spectrum nearly the same as the PN sequence. That is, the spectrum of the transmitted signal is widened by the PN sequence, which is a *spreading code*. The most widely used PN sequences are the *maximum length sequences*, which are coded sequences of 1s and 0s with certain autocorrelation properties. They have long periods, and are simply generated by a *linear feedback shift register*. An *m*-sequence is periodic with period (length)  bits, and is generated by a shift register of length *m*, which uses *m* flip-flops, as shown in Fig.1.



Fig.1 Maximum-length PN code generator.

The systematic code generated by a shift register of length 3 is shown in Fig.2 as an example.



Fig.2 Maximum-length PN code generator.

In reality, the message is transmitted over a bandpass channel with a carrier frequency**, Thus, for direct-sequence binary phase-shift keying (DS/BPSK) transmission, the transmitted signal is:

 (1)

where is the carrier frequency, and the phase  is given by the truth table in Table 1.

Table 1: Truth Table for Phase θ(t)



The transmitted signal is corrupted by some additive interference *i*(*t*). To recover the original information sequence *m*(*t*), the received signal is applied to a synchronous demodulator. The general model of a direct-sequence spread spectrum phase-shift keying system is shown in Fig.3.



Fig.3 Conceptual model of DS/BPSK system.

1. ***Frequency-Hopped Spread Spectrum Modulation (FH-SS)***

In an FH spread spectrum communications system, the frequency is constant during each time chip but changes from chip to chip. The bandwidth is thus subdivided into a large number of contiguous frequency slots. The modulation of FH systems is commonly binary or *M*-ary frequency shift keying (FH/FSK) or (FH/MFSK). A block diagram of an FH/MFSK transmitter and noncoherent receiver is shown in Fig.4.



Fig.4 Block diagram of an FH/MFSK spread spectrum system.

**III.SYNCHRONIZATION OF SPREAD**

 **SPECTRUM SIGNALS**

As discussed, the practical procedures of spread spectrum signals synchronization are often performed in the form of two successive steps. The first, called acquisition (code cquisition, search), performs a coarse measuring of the necessary parameters and provides preliminary estimates used by the second step, called tracking. To explain the acquisition phase of synchronization let us treat unknown delay and frequency shift of the signal as signal coordinates on the time–frequency plane. Suppose that the initial uncertainty ranges of  and  are ** and , respectively, and that as a result of acquisition those ranges should be reduced to ** and . Then, as Fig.5 shows, signal position is within one of   rectangular cells, where. The acquisition should find out which one of  cells contains the signal.



Fig.5 Search zone and signal position on the delay–frequency plane

Below different methods of PN code acquisition is discussed.

1. ***Serial Search method***

In a serial search only one cell at a time is tested, i.e. only a single correlation is calculated of the observation and a local signal replica, having some specific time–frequency shift. The correlation magnitude is then analysed in order to decide whether the cell is true or false. Various criteria may serve to take the decision. For example, the search may continue until all the cells inside the uncertainty region (see Fig5) are tested, all the time storing in memory the maximal correlation observed up to now along with the values of ,  corresponding to it. Then, after the last cell is analysed, the cell believed to be true is known automatically by its coordinates kept in memory, and all to be done is just reading them out. This strategy is equivalent to implementing the ML estimation rule, but calculating the necessary correlations not simultaneously but sequentially in time for successively arriving signal segments.

Still more typical of practical receivers is another version of a serial search, where the currently found correlation magnitude is just compared with a threshold. If the correlation is larger than the threshold, the decision is made that the current cell is true and the search finishes. Otherwise the search system examines the next cell and so forth.

From the point of view of performance analysis, it does not matter how many parameters are unknown and to be estimated in the course of searching: both time and frequency (or whatever else) or some one of them. The only material thing is the overall number of cells to be checked. Yet to make further deliberations more transparent we will treat them as though an acquisition consists in only measuring the time delay of a received signal, the frequency being known a priori with sufficient precision [3].

In order to test synchronism at each time instant, the cross-correlation is performed over fixed intervals of , called *search dwell time*. The correlator output signal is compared to a preset threshold, as shown in Fig.6. If the output is below the threshold, the phase of the locally generated reference code signal is advanced in time by a fraction (usually one-half) of a chip and the correlation process is repeated. These operations are performed until a signal is detected; that is, when the threshold is exceeded. In this case, the PN code is assumed to have been acquired, the phase incrementing process of the local reference code is inhibited, and the tracking phase is initiated.



Fig.6 A sliding correlator for DS serial search acquisition.

 If chips are examined during each correlation, the maximum time required for a fully serial DS search , assuming increments of, is:

 (2)

where  chips is the time uncertainty between the local reference code and the receiver code (searched region). The mean acquisition time can be shown, for , to be:

 (3)

where is the probability of detection, is the probability of false alarm, andthe time interval needed to verify a detection.

A similar process may also be used for frequency-hopping signals. In this case, the problem is to search for the correct hopping pattern of the FH signal [1].

1. ***Serial-parallel search method***

An evident resource of search acceleration involves several parallel correlators, each operating autonomously and scanning a separate part of the uncertainty region. In this case an initial uncertainty region just breaks into sub-regions each covering cells, where  is the number of parallel channels, and acquisition time accordingly reduces  times. In the uttermost case when  the search becomes fully parallel and does not require serial steps [3]. In this case, as illustrated in Fig.7, we observe that the incoming signal is correlated with the locally generated code and its delayed versions with one-half chip  apart. If the time uncertainty between the local code and the received code is chips, then we need 2 correlators to make a complete parallel search in a single search time. The locally generated code corresponding to the correlator with the largest output is chosen. As the number of chips  increases, the probability of choosing the incorrect code alignment (synchronization error) decreases, and the maximum acquisition time given by

 (4)

increases. Thus, *N* is chosen as a compromise between the acquisition time and the error probability of synchronization. The mean acquisition time is :

 (5)



Fig.7 Correlator for DS parallel search acquisition.

The number of correlators can be large, which makes this parallel acquisition less attractive [1].

1. ***Sequential detection acquisition method***

The search algorithm discussed, acquires the correct code phase using single fixed integration for a given threshold level. Such an algorithm is incapable of quickly dismissing a false phase cell or extending the integration time during phase search in a given cell. Indeed, the algorithm does not make use of the additional information that could be available, such as: whether the threshold statistic is close to, greater or smaller than the threshold level. Sequential detection suggests the decision in a given code phase cell by using two or more sequences in a successive order. Consider a detection process with two thresholds A and B such that A*>*B. If the decision statistic*>*A, the signal is declared present and the test ends; if the decision statistic *<*B, the signal is declared absent and the test also ends. However, if B*<*decision statistic *<*A, no decision is made about the presence or absence of the signal and the test continues by extending the integration time.

The development of single dwell detection into sequential algorithm is similar in concept to the evolution of a hard decision into a soft decision decoding in digital signalling. The similarity is clear when one considers the fixed integration time with a single threshold level used in the search algorithm, with the single threshold level in the hard decision decoding on one hand and the extended integration time with the multiple threshold levels in the sequential detection with multiple quantization levels in the soft decision decoding on the other. The comparison of data decoding and search algorithms can be extended to system performance. While soft decision decoding improves bit error rate in data transmission, the sequential detection improvement is evident in shorter mean acquisition times. In sequential detection, the integration time is increased in discrete steps until the test fails and the false phase position is dismissed in a short time [4].

1. ***Matched filter acquisition method***

In serial search, the received spreading code sequence plus noise is multiplied by continuously running local reference spreading code sequence and, after the removal of the possible modulation using square envelope detection, the output is integrated to make an acquisition decision. The process leading to the acquisition test is known as active correlation. Consequently, a new set of Ti /Tc chips from the reference code is used in each acquisition test. This means that, if the test fails, the code phase is updated only every Ti-second intervals. The search rate can be significantly increased by using a matched filter. As we know, matched filtering is basically passive correlation which maximizes the signal-to-noise ratio at its output when the input signal is embedded in additive white Gaussian noise. The received signal continuously slides the stationary (stored) spreading code until the two code sequences are in synchronism. The output of the matched filter is applied to the input of the square law envelope detector and tested against a threshold. The maximum output occurs when the system acquires the correct code phase. The matched filter acquisition system is shown in Fig.8 when a perfect coherent system is used [4].



Fig.8 Baseband matched filter acquisition system.

**IV. ADAPTIVE THRESHOLD IN CODE**

 **ACQUISTION OF DS-SS**

Different methods of PN code acquisition was presented in previous section. As discussed, the complexity of parallel implementation increases as the number of correlators increase. The most popular acquisition approach is the conventional serial search which uses a fixed threshold. The received signal power in mobile communications is inversely proportional to some power *n* of the distance between the base station and the mobile station, where *n* is normally between 3 and 4. Furthermore, the received signals are also subject to the rapid Rayleigh fading about a slowly varying mean signal strength. Thus, since the received signal levels are unknown and location-varying, we cannot achieve good acquisition performance of PN sequences if we employ the fixed threshold acquisition scheme. This fixed threshold acquisition scheme may cause too many false alarms or result in low detection probability according to the threshold value selected. Thus, Threshold setting plays an important role in the performance of the system, since it is the base for the decision of synchronization. Several methods for setting the threshold have been published in the literature. In the last seven years, the concept of adaptive CFAR thresholding has been introduced. Consider a single dwell serial search scheme with a noncoherent detection. This system consists of a single adaptive detector with a correlation tap size N. The adaptive detector consists of two blocks, as shown in Fig.9. The first block is the conventional noncoherent matched filter (MF) detector. The second block illustrates the adaptive CFAR operation for the decision process.



Fig.9 Adaptive serial search acquisition scheme.

 The received PN signal plus noise and any interference are arriving at the input of the adaptive detector. If the adaptive detector declares that the present cell is the correct one, the tracking loop is activated, and the relative time delay of the local PN signal is retarded by , where  is the chip time, to examine the next cell. The whole testing procedure is repeated. Usually, the value of  is 0.25, 0.5, or 1. On the other hand, if the adaptive detector declares , the phases of the two codes (incoming and local) are automatically adjusted to the next offset position, and the test is repeated. As shown in Fig.10, for the adaptive operation of the decision processor, the threshold value of the comparator in the adaptive detector is adapted in accordance with the magnitude of the incoming signals.



Fig.10 Block diagram of adaptive detector.

 Accordingly, the outputs of the correlator are serially fed into a shift register of length . The first register, denoted as *Y*, stores the output of the multiplication of the power of the incoming signal with the value of the partial correlation between the local and incoming PN sequences. The following *M* registers, denoted by , *j* = 1, 2, …, *M*, and called reference windows, store the output of the previous *M* phases. A selection logic is then used to set the threshold based on a fixed probability of false alarm. Linatti [5], while studying threshold principles in code acquisition of direct sequence spread spectrum signals, showed that better performances may be obtained using CFAR criterion under certain conditions [1]. Different CFAR algorithms have been suggested in the literature. Below we present some of these methods.

1. ***Adaptive Thresholding using CML-CFAR***

***Processor***

An adaptive acquisition scheme of PN sequences which estimate the background power level using censored mean level processor(CML), multiply it with the threshold coefficient to keep the false alarm rates constant and use it as a threshold is presented in [6]. In this section, we will focus on this method.

**Model description**:

We assume that the spreading sequence timing is determined before any phase measurement is attempted. Hence, the noncoherent hypothesis testing device should be employed, as shown in Fig.11. We also assume that there is no frequency error.



Fig.11 Kth user’s non-coherent detector

When noncoherent reception of Rayleigh-faded signals in AWGN is considered, the normalised likelihood functions become :

 (6)

 (7)

where ,  is the average signal-to-noise ratio. For the adaptive operation of the decision processor, the detection threshold should be updated in accordance with the local situation. As discussed, The outputs of detection variable ** are sent serially into a shift register of length *M* + 1. The first register, denoted as Y, stores the output of the test phase. The following *M* registers, denoted by*,* and generally called a window, store the outputs of previous *M* phases. The statistic ** is formed by processing the contents of the cells in the window. Similar to CML-CFAR method in radar literature the power level is estimated by summing the smallest values of the window:

 (8)

where denotes the smallest value after sorting the data in the window, is the largest value and  is the number of excised data in the window. The threshold is set as the power level estimate *Z,* times threshold coefficient*.* Here,  is a scaling factor used to achieve the desired false alarm probability for a given window of size *M.* We call this adaptive acquisition processor*.*

**Analysis of AAP**:

Acquisition of PN sequences is declared if *Y* exceeds the threshold  as follows:

 (9)

Since *X* is a random variable, the performance is evaluated by average detection and false alarm probabilities. The false alarm and detection probabilities are:

 (10)

 (11)

respectively. In [7] it has be shown that the detection and false alarm probabilities are expressed as :

 (12)

 (13)

where

 (14)

**Results and discussion:**

Table 2 lists the value of threshold coefficient T for the adaptive scheme to achieve the false alarm rate of *Pfa* =10-4 for *M=* 8.

Fig.12 shows the detection probability of *AAP* against the signal-to-noise ratio for different numbers of excised cells. From the Figure, we can see that *AAP(0)* has the best detection performance under the assumptions that *Y* has the probability density function (PDF) of (7) and all the *X,* have a PDF of (6).

Table 2 : Threshold coefficient *T* of *AAP* for *Pfa* =10-4

 and *M* = 8

.



Fig.12 Detection probability against SNR for different number

 of excision, Pfa = 0.0001, M = 8

Since the transmitted signal in terrestrial communications is usually reflected by a variety of terrains or buildings, it is replicated at the receiver with several time delays. We consider the situation in which the three multipath signals are received in the window and their distances are within eight chips as shown in Fig.13. In this case, the detection probability of *AAP(0)* degrades seriously with the multipath components included in the window. The threshold of *AAP(0)* is raised as the previously reflected signal enters the window, so that acquisition of the second and third multipath components in Fig.13 has failed.



Fig.13 Thresholds of AAP(0) and AAP(2) under multipath

 situations

 (i) AAP(0)

 (ii) AAP(2)

 (iii) input data

Conversely, *AAP(2)* can detect these multipath signals without raising the threshold because the threshold of *AAP(2)* is determined while the 7th and 8th largest samples in the window are removed. Thus, the *Ne* parameter has a great influence on false alarm rate and detection probability. Generally, we can select the  parameter by predicting the maximum number of multipath components likely to be present in the given window.

1. ***Adaptive thresholding using Excision-CFAR processor***

As we know, multipath interferences usually exist in the spread spectrum communication systems and have disastrous effects on the performance of conventional mean-level detectors. In this section, we present the excision CFAR (E-CFAR) method which is introduced into the adaptive acquisition in [8]. The E-CFAR detector excises the strong samples that exceed the excision threshold from the sample set prior to the cell-averaging operation. If the excision threshold is properly set, the effect of multipath fading under the excision threshold should be tolerable.

**Model description:**

The structure of the proposed E-CFAR detector is shown in Fig. 14b. In the Figure, BE presents the excision threshold. We assume that the spreading sequence timing is determined before any phase measurement is attempted. Hence the noncoherent correlator is employed and illustrated in Fig. 14a.



Fig. 14 Block diagram of proposed E-CFAR detector

a Time hypothesis testing scheme for BPSK spreading

b Structure of E-CFAR detector

In an AWGN environment, each noise sample at the output of the square-law device is a random variable with an exponential probability density function (pdf) and each signal sample is a noncentral chi-square pdf. In the analysis of a communication system, the amplitude of the received signal with a Rayleigh pdf is widely accepted in a fading channel. So each signal sample also has an exponential pdf. The pdfs of the noise sample and the signal sample can be written as:

 (15)

where µ is the noise variance and is the signal-to-noise ratio (SNR). All sample outputs are assumed to be independent and identically distributed (IID).

**Analysis of E-CFAR:**

The function of the excisor is to excise the samples that exceed the excision threshold. The output *Y* of the excisor is a random variable with a pdf given by

 (16)

When the output of the excisor is a signal sample, the variable  is replaced by . When the output of the excisor is a noise sample,

the variable is replaced by . The probability that a noise sample can survive the excisor is given by:

 (17)

The outputs after the excisor are averaged to obtain the power level estimate. The detection threshold S is determined by the estimate  multiplied by the threshold coefficient as follows

 (18)

Here, *K* represents the number that survived the excisor. T is a scaling factor used to achieve the desired false alarm probability. The characteristic function of the random variable  can be calculated from (17). We denote this function as:

 (19)

Where and *K* are the parameters of the characteristic function and *V* is the variable. When all the outputs of the excisor are noise samples, the parameter  is replaced by . The detection probability conditioned on the number of surviving samples after the excisor is the probability that exceeds the detection threshold . Thus, the probability in the benign environment is:

 (20)

Having such an expression for the conditional detection probability, it is obvious that an expression for the conditional false alarm probability can be derived by setting in (20):

 (21)

To remove the condition *K*, the probability that *K* samples remain after the excisor must be obtained. It is a binomial distribution function as follows:

 (22)

The unconditional detection probability and unconditional false alarm probability can be obtained by averaging (21) and (22) over the distribution:

 (23)

(24)

**Results and discussion:**

We define the excision coefficient as α=BE/µ. When , the E-CFAR detector becomes the CA-CFAR detector. Fig.15 shows the detection probability of the E-CFAR detector against SNR for different a in signal path situations. The detection coefficient T is derived by solving (24) for the required values of PFA and . Under the homogeneous noise environment without multipath signals, the detectability losses with respect to the conventional CA-CFAR detector can be evaluated by comparing the curves with the curve corresponding to . From Fig.15 we can see that the detection probability rapidly converges to the maximal value as the excision coefficient increases. Fig.16 shows the detection probability of E-CFAR and CA-CFAR detectors against SNR in multipath situations. In Fig.16,  present the average SNR of the detection signal sample and  is the multipath signal sample. The left-hand curve corresponds to the CA-CFAR detector without multipath. The lower solid curve presents the performance deterioration of the CA-CFAR detector with two multipath signals. The other curves present the performance of the E-CFAR detector with two multipath signals. For small , since the excision probability of the signals and noise samples increases, the performance deteriorates seriously. For large , as the excision probability decreases,  decreases gradually to the value corresponding to the CA-CFAR detector with two multipath signals. If proper a is selected, we can see that the degradation is quite small under the multipath fading, in comparison with a conventional cell-averaging method. When  increases, the excision probability of the multipath signals increases and the performance improves accordingly.



Fig. 15 Detection probability against SNR for no multipath

 signals

 Upper——– Lower——– 

 Upper------- Lower------- 

 Upper…….. Lower…..... 

 

Fig. 16 Detection probability against SNR for mutipath signals

 

Upper——– CA-CFAR without multipath signal

Lower——– CA-CFAR with two multipath signals

-------- α=.5 with two multipath signals

……... α=4 with two multipath signals

-.-.-.-.-. α=20 with two multipath signals

**V. CONCLUSION**

A very challenging problem arising in DS-SS systems is to achieve reliable PN code acquisition in the presence of fading and interference. In this paper, we have considered this Problem. Different methods of PN code acquisition has been discussed and it has been shown that CFAR processors, primarily developed for radar signal detection, can also be effective in PN code acquisition. The effectiveness of this adaptive technique has been assessed using computer simulation results.

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