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A QUASI-ORTHOGONAL CORRELATOR-BASED BPNN PN CODE ACQUISITION SCHEME FOR UNDERWATER ACOUSTIC DSSS COMMUNICATION

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Key words: PN code acquisition, quasi-orthogonal correlator, modified BPNN, underwater acoustic DSSS communication.

ABSTRACT

This paper proposes a quasi-orthogonal correlator-based back propagation neural network (BPNN) pseudo-noise (PN) code acquisition scheme for underwater acoustic direct sequence spread-spectrum (DSSS) communication. The main idea is to use short-length quasi-orthogonal correlators for coarse phase estimation and a BPNN based discriminator for fine phase calculation. When compared with other neural-net based PN code acquisition schemes, the short quasi-orthogonal correlators effectively locate possible phases of the received signal with high tolerance to interference and require no full-length correlation. These capabilities have been verified by extensive computer simulations and the results clearly show that the proposed scheme has the characteristics that are highly suitable for the implementation of underwater acoustic DSSS communication systems.

I. INTRODUCTION

In recent years, the demand for high data-rate underwater acoustic communication systems has been driving the development of advanced modulation techniques with high spectrum efficiency and multiple access capability [14]. The code division multiple access (CDMA) technique is one of these emerging techniques and is frequently adopted as underlying architecture for underwater acoustic networks [16]. CDMA systems utilize the direct-sequence spread-spectrum (DSSS) technique combining with orthogonal codes to achieve channel separation. The capacity of a CDMA system grows linearly with the increasing number of channels and its performance has been evaluated and reported in several research papers [3, 11, 13, 17]. Although using orthogonal codes for channelization provides good performance in an additive white Gaussian noise (AWGN) channel, the orthogonality will be distorted when communicating in a multipath channel. The multiple access interference (MAI) will also increase enormously as both numbers of channels and paths grow. To suppress the interference from multipath components, a PN code is applied to mask the CDMA signal. The masking PN code is multiplied with all of the CDMA signals to pseudo-randomize the chip sequences. With perfect PN code synchronization, the interference from multiple paths and other channels can be suppressed effectively.

In a DSSS communication system, the state of a channel is unknown when a signal is being transmitted from the sender to the receiver. This phenomenon causes false-alarm during PN code synchronization. Conventional serial PN acquisition uses a matched filter to evaluate the phase of the PN code. The matched filter performs a full-length correlation between the received signal and the PN code at the desired phase and locates the peak of the output correlation. Although serial acquisition is a low-complexity technique, full-length correlation is very time-consuming when the PN code has a long period. For a maximum length sequence (MLS) generated from a linear feedback shift register (LFSR) with $M = 15$, the output of a full correlation can have a peak period of up to 3.2 seconds when the chip rate is at 10 Kchips/s. Additionally, the low velocity of underwater sound wave also increases the variation of the signal's arrival time, which makes the pilot signal used in mobile cellular CDMA systems useless in underwater PN code synchronization.

In order to acquire the phases of the received signal correctly and quickly, several two-stage acquisition systems have been proposed [1, 7, 19]. For example, the coarse/fine (C/F) search scheme is a two-stage acquisition technique consisting of an initial coarse phase estimation followed by a fine phase calculation. The coarse estimation locates the possible phases of the received signal by using specific correlators, and the phase calculation tests all hypotheses within the located range.

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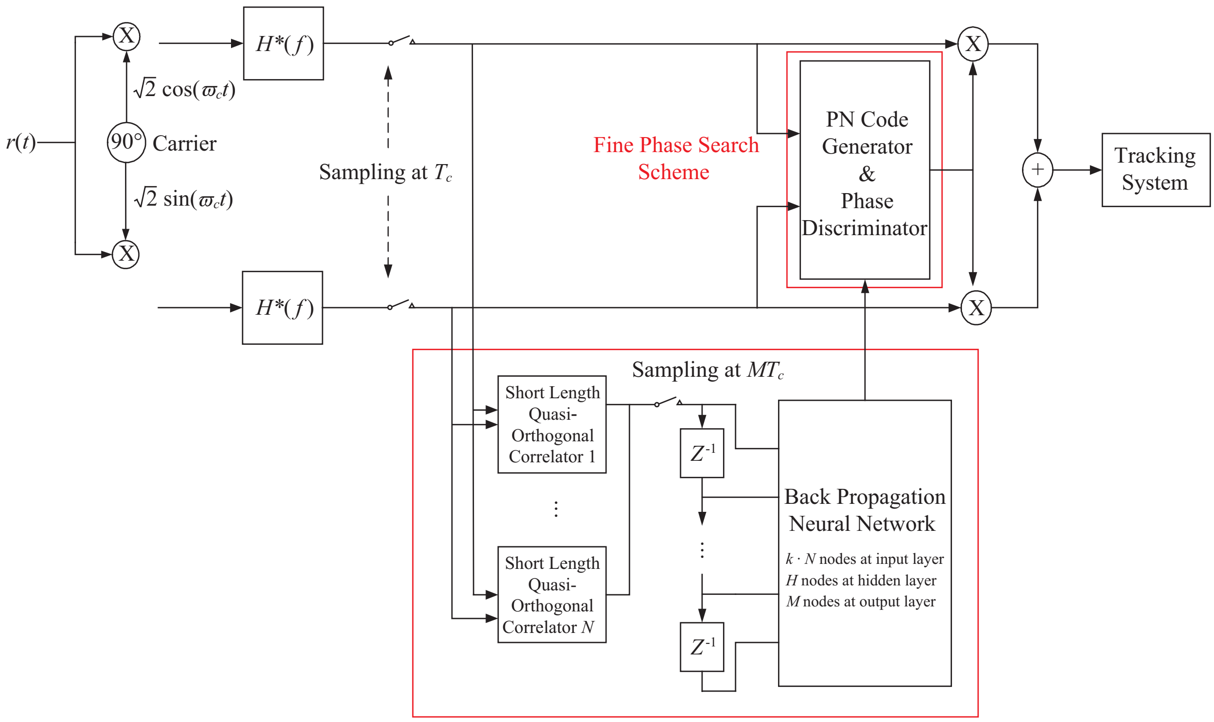

Fig. 1. Architecture of the proposed PN code acquisition system.

However, two-stage acquisition techniques are very time-consuming for full-length correlation. In order to shorten the mean acquisition time, a modified BPNN based PN code acquisition system has been carefully studied in our earlier work [9, 21]. In this paper, we extend our previous work by using short-length quasi-orthogonal correlators for coarse phase estimation and a back-propagation neural network (BPNN) based discriminator for fine phase calculation. The short quasi-orthogonal correlators can locate all the possible phases of the received signal without a full-length correlation. Simulation results show that this proposed scheme can acquire the phase of the received signal effectively.

II. A QUASI-ORTHOGONAL CORRELATOR-BASED BPNN PN CODE ACQUISITION SCHEME

Fig. 1 shows the quasi-orthogonal correlator-based BPNN PN code acquisition scheme proposed in this paper, which includes a coarse phase search scheme and a fine phase search scheme. In the coarse phase search scheme, the received signal passes through N quasi-orthogonal correlators, all of the same length as that of the LFSR within the PN code generator. BPNN estimates the current PN code state according to the output of the correlator group. In the fine phase search scheme, the phase discriminator verifies the estimated PN code state by computing the correlation of the subsequent received signal.

1. Coarse Phase Search Scheme

The coarse phase search scheme is composed of two components, one is the short-length quasi-orthogonal correlator group, and the other is the BPNN-based PN code phase estimator. The quasi-orthogonal correlator group is used to specify the characteristics of the received PN signal. In DSSS communication systems, the PN code is periodically recurring after a certain run-length. In MLS, the period of the PN code is $2^M - 1$, where M is the length of the LFSR. At any given time, the content of the shift register in a MLS PN code generator is one of the $2^M - 1$ distinct nonzero binary sequences of length M. Because PN code is a stationary ergodic sequence, the content of the LFSR after a run-length of n chips should be independent of its current content, when n is larger than the length M of the shift register [20]. The correlation between the received signal and the content of the MLS shift register, i.e. short-length correlation, is maximized if the phases of the received signal and the PN code generator are in synchronization. Therefore, it is possible to estimate the phase of the received signal by computing a short-length correlation instead of a full-length correlation. Each phase of the short-length content is followed by a unique short-length sequence. This property has been used in a fast PN code acquisition algorithm by implementing orthogonal correlators in a parallel search [6], where each correlator can locate a certain range of PN code hypotheses, and the verification is done by calculating the correlation between the following received signal and the PN code generator. Therefore, the mean acquisition time is limited to within one full run-length. Although the orthogonal correlators are capable of doing short-length PN code classification, the inflexibility of the length of the correlators makes them only suitable for PN codes generated from LFSRs with lengths being the power of 2. To overcome this problem, we use short quasi-orthogonal correlators to do the coarse phase search. The orthogonality of the quasi-orthogonal code makes these correlators almost uncorrelated to each other, thus each correlator can locate a certain range of the target short-length phases. In 2000, Cha proposed an algorithm to provide a set of quasi-orthogonal sequence pairs with zero correlation duration [5]. By setting an initial vector matrix G as

$$G = \begin{bmatrix} 1 & 1 & 1 & -1 \\ 1 & 1 & -1 & 1 \\ 1 & -1 & 1 & 1 \\ -1 & 1 & 1 & 1 \end{bmatrix} \tag{1}$$

each row of G can be used as an initial vector $S_4^{(a)}$ = [$s_0^{(a)}$ $s_1^{(a)}$ $s_2^{(a)}$ $s_3^{(a)}$]. Then, we can determine another initial vector $S_4^{(b)}$ orthogonal to $S_4^{(a)}$, where $s_i^{(b)} = (-1)^i s_i^{(a)}$. To increase the number of usable quasi-orthogonal sequences, the initial vector can be expanded to form matrix D, shown in Eq. (2), where the number of sequences is quadrupled and the length of the expanded sequences is doubled.

$$D = \begin{bmatrix} (s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2}^{(a)},...,s_{m-1}^{(a)}) & (s_0^{(a)},...,s_{m/2-1}^{(a)}) & -(s_{m/2-1}^{(a)},...,s_{m-1}^{(a)}) \\ (s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2}^{(a)},...,s_{m-1}^{(a)}) & -(s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2-1}^{(a)},...,s_{m-1}^{(a)}) \\ (s_0^{(a)},...,s_{m/2-1}^{(a)}) & -(s_{m/2}^{(a)},...,s_{m-1}^{(a)}) & (s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2-1}^{(a)},...,s_{m-1}^{(a)}) \\ -(s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2}^{(a)},...,s_{m-1}^{(a)}) & (s_0^{(a)},...,s_{m/2-1}^{(a)}) & (s_{m/2-1}^{(a)},...,s_{m-1}^{(a)}) \end{bmatrix} \tag{2}$$

However, because the length of the sequences can only expand by multiples of 2, these sequences can only be applied to special cases where the length of LSFR is power of 2. To overcome this limitation, we have developed a method to generate quasi-orthogonal sequences of any length. Our method works as follows.

Given a sequence length M, factorization can be performed such that $M = M_0 \cdot M_1 \cdot \ldots \cdot M_N$. Let A_i be identity matrices of order i, where $i = M_0, M_1, \ldots, M_N$. By expanding all the identity matrices, we are able to generate the set of quasi-orthogonal sequences of desired length. For example, consider a set of quasi-orthogonal correlators of length $M = 15$. By factorizing M, we form two identity matrices of orders 5 and 3 as follows:

$$A_0 = \begin{bmatrix} 1 & -1 & -1 & -1 & -1 \\ -1 & 1 & -1 & -1 & -1 \\ -1 & -1 & 1 & -1 & -1 \\ -1 & -1 & -1 & 1 & -1 \\ -1 & -1 & -1 & -1 & 1 \end{bmatrix} \tag{3}$$

The expanded matrix A_{exp} can then be derived as follows, where each row is quasi-orthogonal to the others.

$$A_{exp} = \begin{bmatrix} A_0 & -A_0 & -A_0 \\ -A_0 & A_0 & -A_0 \\ -A_0 & -A_0 & A_0 \end{bmatrix} \tag{4}$$

Although the short-length quasi-orthogonal correlators can classify the range of received DSSS signal phases, we still need a PN code phase estimator to identify the exact phase. In order to acquire the phase of the received signal correctly, many neural network based acquisition systems have been proposed [8, 9, 15, 21]. Neural network systems store information of the PN codes as network weights by training the system on all possible phases of the received signal. The trained network can then acquire the phase of the received signal correctly. Yoon, *et al*. [21] proposed the first BPNN based acquisition system, where the received spread-spectrum signal passed through a shift register and was classified by the neural network. The threshold of operation is determined by the probability of false-alarm. When the output of the neural network exceeds the threshold, the system will determine the phase of the received signal and a local PN code will be generated to despread the received signal. The neural network proposed by Yoon was trained on a certain phase of PN code with additive Gaussian noise. Their simulation result showed that its performance was similar to the systems using the serial search technique with same threshold.

For conventional PN code trained neural network based systems, the effect of multipath fading has an adverse effect on the mean acquisition time. To cope with this difficulty, the BPNN proposed in this paper is trained on the output of a group of short-length correlators, rather than the PN code itself. The multipath components of the received signal are independent of the main-path signal; therefore, the distribution of the output of the short-length correlators is Gaussian. Consequently, the short-length correlators can classify the received signal into certain groups, given a proper detection rule. The input to the BPNN is the output from the group of k correlators with a sampling rate of $1/(M \cdot T_c)$. The number of nodes in the input layer, the hidden layer, and the output layer are $k \cdot N$, H, and M, respectively. The transfer function for the hidden layer and the output layer are a linear function and a Tan-Sigmoid function, respectively. The training procedure for the BPNN is shown in Fig. 2. How many samples should

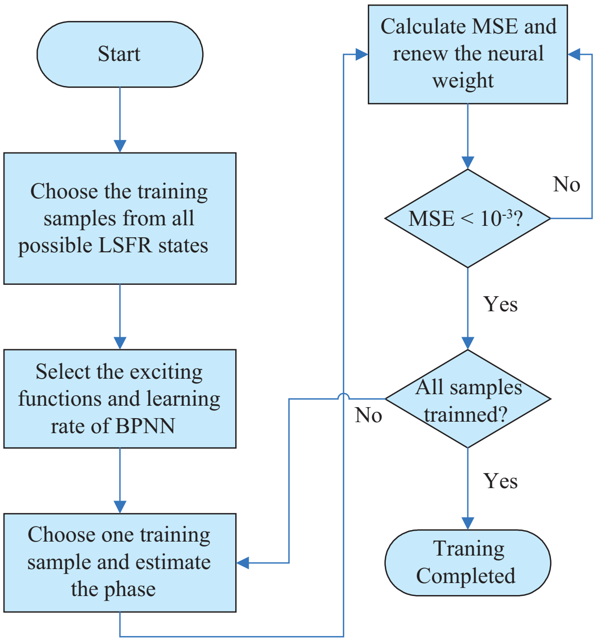

Fig. 2. Flow chart for the proposed BPNN training procedure.

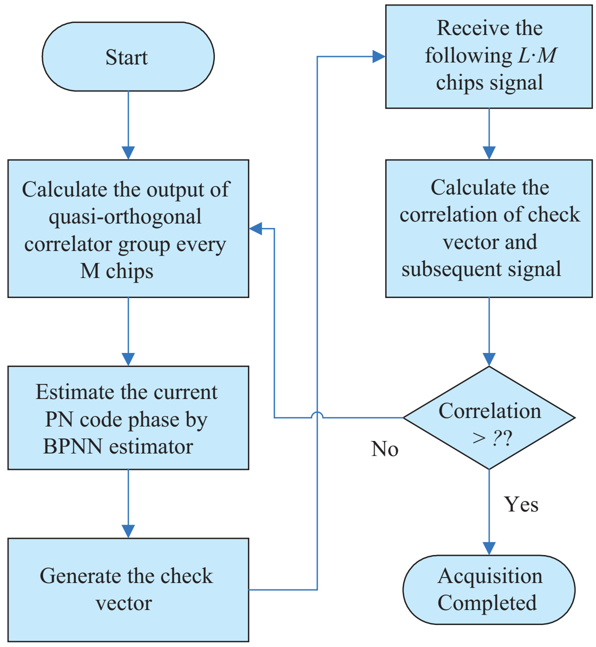

Fig. 3. PN code acquisition procedure.

be trained is a trade-off between the training time required and the mean acquisition time desired. The more samples are trained, the shorter the acquisition time is needed.

2. Fine Phase Search Scheme

Since the BPNN has estimated the possible phase of the PN code, we can verify the hypothesis by subsequent received signals. The predictable content of the LFSR can be used as the check vector to verify our hypothesis. For each hypothesis, there exists a unique check vector of length L. By computing the correlation between the check vectors and the subsequent received signal within the period from $t + \tau$ to $t + \tau + L \cdot M - 1$, where τ is a random time delay, we can determine the exact phase of the received signal. To minimize the mean acquisition time, we set τ to zero. The length $L \cdot M$ and the threshold for phase discriminator in the fine phase search scheme will affect the probability of false-alarm P_F. Fig. 3 shows the procedure for PN code acquisition. For each acquisition test, $M \cdot (k + L)$ chips of the received signal are needed.

III. PERFORMANCE ANALYSIS

Direct-sequence spread-spectrum communication uses PN codes to combat the interference of noise by spreading the spectrum of the original signal [20]. In AWGN channel, the received signal can be expressed as

$$r(t) = \sqrt{\frac{2E_C}{T_C}} C(t - t_d)\cos(\omega_C(t - t_d) + \theta) + n(t) \tag{5}$$

where E_c is the energy of per chip, $C(t)$ is the PN code signal, n is Gaussian noise, t_d is the time delay, θ is the phase angle, and T_c is the chip duration. When the spread spectrum signal is received, the receiver will generate a local PN code to despread the received signal. In order to synchronize the phase of the local PN code and the received signal, the signal is fed through a correlation detector to acquire the phase. When the received signal is in synchronization with the correlation detector, the output of the correlation detector can be expressed as

$$\begin{aligned} y(t) &= \int_0^{l \cdot T_c} r(t) \cdot \sqrt{\frac{2}{T_C}} C(t - \hat{t}_d)\cos(\omega_C(t - \hat{t}_d)) \\ &= l\sqrt{E_C} + \beta \end{aligned} \tag{6}$$

where $\hat{t}_d$ is the estimation of time delay, l is the length of PN code, and β is Gaussian noise energy. The distribution of β is $N(0, l\sigma^2)$, where σ^2 is the noise power. If the received signal and the correlation detector are not in synchronization, the output of the correlation detector reduces to β only. During acquisition, the output of the correlation detector indicates the phase of the received signal. When the output of the correlation detector exceeds the threshold, the phase of the received PN code is acquired and the local PN code will be generated to despread the received signal. The probability of error is

$$P_e = Q\sqrt{\frac{lE_C}{\sigma^2}} \tag{7}$$

The value of the threshold affects the probability of false-alarm. False-alarm occurs when the output of the correlation detector exceeds the threshold and the phases are different between the received signal and the correlation detector. When false-alarm occurs, a wrong PN code is generated and the receiver gets wrong information. The system will then try to recover the correct phase, thus resulting in lower data rate. For a given value of threshold γ, the probability of false-alarm is

$$P_F = Q(\frac{\gamma}{\sqrt{l\sigma^2}}) \tag{8}$$

When the correlation detector and the received signal are in synchronization, the distribution of the correlation detector output is $N(l\sqrt{E_C}, l\sigma^2)$. The probability of detection can be expressed as

$$P_D = Q(\frac{\gamma}{\sqrt{l\sigma^2}} - \sqrt{\frac{lE_C}{\sigma^2}}) \tag{9}$$

The mean acquisition time of this system depends on the distribution of the trained hypotheses in the coarse phase search scheme. In MLS, the distribution of the hypotheses is uniform because of the stationary ergodic property. Hence the average separation duration T_d between any two nearby hypotheses is

$$T_d = \frac{M \cdot (2^M - 1)}{N_{hyp}} \tag{10}$$

Fig. 4 show the state diagram of phase search, when the probability of detection is P_D and the probability of false-alarm is P_F [13]. The states at the outer ring are the phases of the received signals which are not within the acquisition system's hypotheses. The states at the inner ring indicate the phases of false-alarm. The parameter Z is the time delay for hypothesis testing. Because the short-length quasi-orthogonal correlator group can be seen as a circuit doing parallel processing, the testing time is a duration of M chips. If we assume the first phase of the received signal is on the i^{th} node of the outer ring counterclockwise from the node where a successful detection occurs, the transfer function from the initial node to the node of successful detection is

$$U_i(Z) = \frac{((1-P_F)Z + P_F Z^{1+L})^i P_D Z^{k+L}}{1-(1-P_D)Z((1-P_F)Z + P_F Z^{1+L})^{v-1}} \tag{11}$$

where v is the total number of starting nodes. Since all nodes at the outer-ring are a priori equally likely to be the starting node and

$$H_o(Z) = (1-P_F)Z + P_F Z^{1+L} \tag{12}$$

the averaged total transfer function is

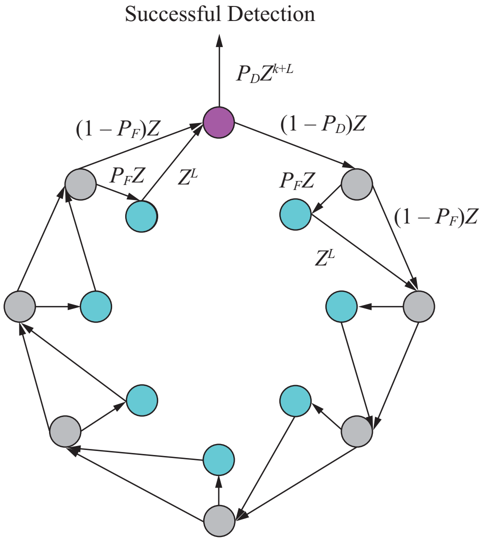

Fig. 4. Phase search state diagram.

$$\begin{aligned} U(Z) &= \frac{1}{v}\sum_{i=0}^{v-1} U_i(Z) \\ &= \frac{1}{v}\frac{P_D Z^{k+L}[1 - H_o^v(Z)]}{[1-(1-P_D)Z \cdot H_o^{v-1}(Z)][1-H_o(Z)]} \\ &= \sum_{n=1}^{\infty} U_n Z^n \end{aligned} \tag{13}$$

In an ideal channel, where $P_D = 1$ and $P_F = 0$, the mean acquisition time is

$$\begin{aligned} T_{ideal} &= M \cdot \sum_{n=1}^{\infty} nU_n = M \cdot \frac{dU(Z)}{dZ}\bigg|_{Z=1} \\ &= \frac{M}{v} \cdot \frac{d}{dZ} \frac{Z^{k+L}[1-Z^v]}{[1-Z]}\bigg|_{Z=1} \\ &= M \cdot (k + L + \frac{v-1}{2}) \end{aligned} \tag{14}$$

From Eq. (14), the mean acquisition time is a linear function of k, L, and v. The parameter v is the ratio of the number of all possible samples to the number of trained ones. When the full run-length of MLS PN code is large, the full-sample training will be time-consuming. Fortunately, the training process can be sped up easily by reducing the number of training samples with only a slight increase of the mean acquisition time.

IV. SIMULATION

In underwater acoustic communication, the main limitation arises from the time-varying multipath channel. The long time spread caused by multipath propagation restricts the bandwidth of the communication system. The multipath conditions

Table 1. Parameters of short range shallow water channel model.

Range	100 m
Water depth	14.5 m
Source depth	3 m
Receiver depth	2 m
Carrier frequency	40 KHz
Chip Rate	2 KHz
Doppler spread	10 Hz
Water Sound Speed	1539 m/s
Fading in direct path	Ricean fading, $K = 1.7$
Fading in reflected path	Rayleigh fading
Ambient noise	Symmetric α-stable distribution, $\alpha = 1.7$

heavily depend on sea state, boundary conditions, ocean depth, transmitter-receiver position configuration, and the range of communication [2, 4, 18]. In a short-range shallow water (SRSW) acoustic communication environment, the dominant components of multipath interference arise mainly from sound waves bounced off the boundaries. To estimate the propagation delay and power decay scope of each multipath signal, these reflected components can be described using ray-tracing technique [12]. The experimental results are in agreement with the predicted multipath scope. In this paper, we use a SRSW underwater acoustic model [10] to evaluate the performance of the proposed PN code acquisition system. Table 1 summarizes the parameters of the channel model depicting the time-varying channel impulse response. In this model, the multipath delay can extend to 3 ms, thus causes the inter-chip interference.

In computer simulation, the chosen PN code is MLS with a period of 2^{10}-1 chips. The length of the LFSR is $M = 10$ and its characteristic polynomial is $p(D) = D^{10} + D^9 + D^8 + D^5 + 1$. The number of short-length correlators is equal to M. The number of input sets for the BPNN is k; the threshold is γ; and the length of the check vector in the phase discriminator is either M or $2M$. For a stationary ergodic pseudo noise sequence, the state of the LSFR is uniformly distributed. In a noise free environment, at any given time, the probability of false alarms caused by signal itself depends on the threshold γ of the correlator. In order to suppress false alarms, multiple samples are fed to the BPNN as its input. As a result, the probability of false alarm reduces to $(1 - \gamma)^k$. By choosing $k = 4$ with threshold γ set to 0.8, the probability of false alarm caused by signal itself has an upper bound of 1.6×10^{-3}.

The BPNN consists of a 40-node input layer, a 50-node hidden layer, and a 10-node output layer. All the training samples for the BPNN are taken from the possible contents of the LFSR. With an ideal channel, the mean acquisition time of the simulated system fluctuates between 50 and 60 chips, depending on the length of the check vector.

To evaluate the proposed system, a full run-length MLS PN code with a random initial phase is used and the simulation results are shown in Figs. 5-8. Fig. 5 shows that the mean

Table 2. Comparison of the mean acquisition time of different PN code acquisition methods in symbol duration (T_S) of 1023 chips.

Proposed method	$(M + L)\log_2 T_s$
Ancken's C/F search method [1]	$3\ T_s$
Full Matched Filter [16]	$1.5\ T_s$
K Parallel search method [1]	$(1 + T_s/2K)\ T_s$
Yoon's BPNN search method [21]	$1.5\ T_s$

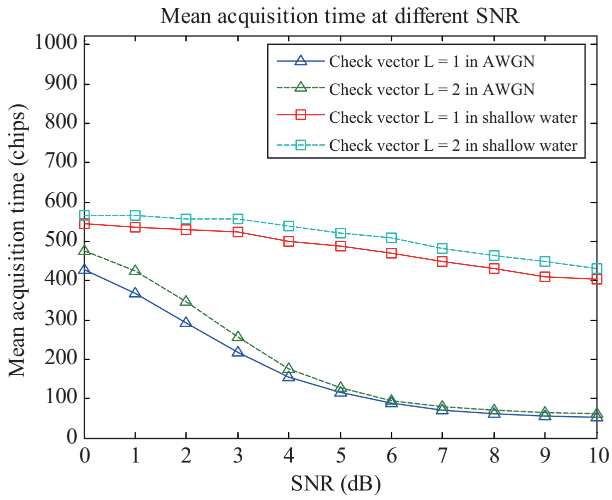

Fig. 5. Mean acquisition time of the proposed BPNN-based coarse/fine search scheme in underwater PN code acquisition.

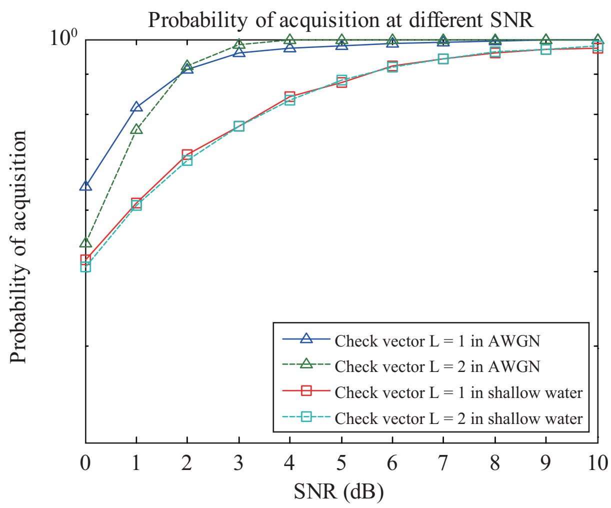

Fig. 6. Probability of acquisition.

acquisition time over an AWGN channel with high SNR is on par with that over an ideal channel. Also, with an SRSW channel, the mean acquisition time is longer than that in an AWGN channel due to the interference caused by multipath propagation.

Table 2 shows that our method has a clear edge over several other PN code acquisition methods, regarding the mean ac-

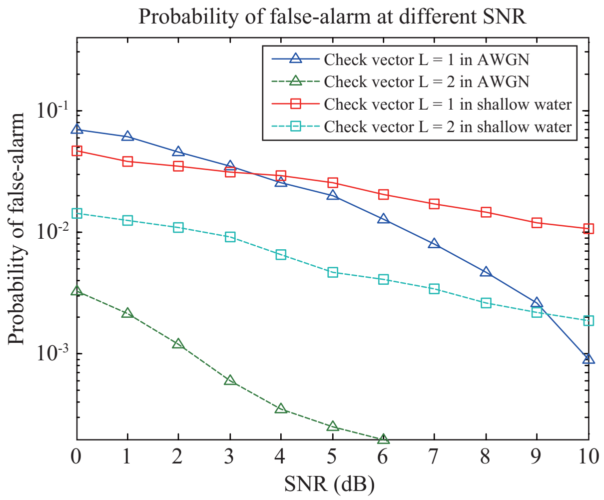

Fig. 7. Probability of false-alarm.

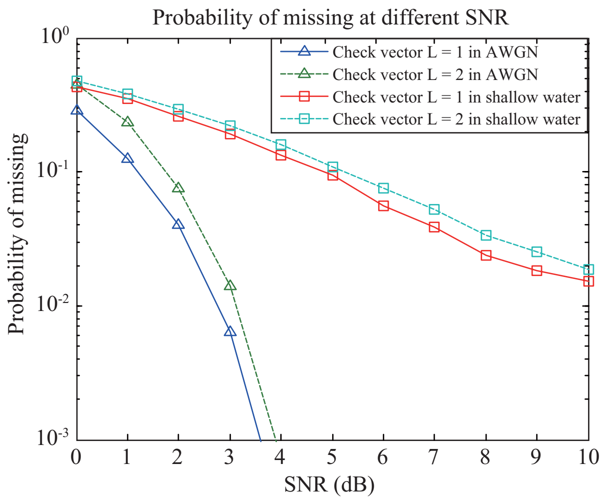

Fig. 8. Probability of missing.

quisition time. Most PN code acquisition methods need a full-length symbol to gain enough information to determine the correct phase. In doing so, time is inevitably wasted on waiting for the completion of code template initialization, even when SNR is high. To make things worse, underwater acoustic DSSS communication systems often suffer from PN code re-acquiring problem because of the time-varying environment and limited bandwidth [10]. Our method solves the problem by adopting continuous short-length sequences, thus needs not to wait for code template initialization. Shorter mean acquisition time also works in our advantage when re-acquiring occurs.

The probabilities of acquisition, false-alarm, and missing signals are shown in Figs. 6-8, respectively. The length of the check vector in the fine search scheme is the major factor that affects these probabilities. Longer check vectors result in lower probability of false-alarm and higher probability of missing at high SNR for the proposed acquisition system. With the same probability of false-alarm, the check vector with $L = 2$ shows a gain of 9 dB in SNR over $L = 1$. However, longer check vectors cause the mean acquisition time to become longer. It's a trade-off between lower probability of false-alarm and shorter mean acquisition time. The overall simulation results indicate that the proposed PN code acquisition system is highly efficient in PN code acquisition for both AWGN and SRSW channels. With sufficient SNR, the phase of the PN code can be acquired correctly within one full length running duration.

V. CONCLUSION

We have proposed in this paper a quasi-orthogonal correlator-based BPNN PN code acquisition scheme for underwater DSSS modem. The quasi-orthogonal correlators quickly classify the received signal and the BPNN identifies the phase of the received signal accurately. The proposed system has two major advantages, namely, the mean acquisition time is relatively short and the amount of training samples required is small. The small amount of training samples is due to the fact that our system is trained on the output of the correlation detectors, which results in a small input space for the neural network and reduces the time of training. Simulation results show that the system can acquire the correct phase of the PN code efficiently in a short range shallow water channel. When compared with other neural-net based systems, the system proposed in this paper achieves both higher probability of acquisition and shorter mean acquisition time which is always within one full run-length. The property of rapid acquisition makes our PN code acquisition system more suitable for underwater acoustic communication.

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