Direct Sequence Spread Spectrum Based Modem for Under Water Acoustic Communication and Channel Measurements

E.M.Sozer[1] J.G.Proakis[2] M.Stojanovic[1] J.A.Rice[3] A.Benson[4]
M.Hatch[3]

[1]Department of Electrical and Computer Engineering, Northeastern University, Boston, MA 02115
[2]Delphi Communication Systems, 2 Clock Tower Place, Maynard, MA 01754
[3]Space and Naval Warfare Systems Center, Acoustic Branch, Code D881, San Diego, CA 92152
[4]Office of Naval Research, 800 North Quincy Street, Arlington, VA 22217-5660

Abstract

An underwater acoustic communication system based on direct-sequence spread-spectrum signaling is presented. The method is intended for use in multiple-access underwater communication networks. In this paper, the transmitter and receiver designs are proposed, discussed, and demonstrated using real data. At the receiver, a rake filter is employed. The same signaling scheme is also used to obtain an adaptive estimate of the time-varying channel coefficients. The channel estimator performance is demonstrated using simulations and experimental data.

I. Introduction

Many applications, such as speech transmission between divers require real-time communication, which is not necessarily in point-to-point links but in network configurations. Hence, acoustic modems that can handle multiple users with efficient utilization of the limited resources of the underwater acoustic (UWA) channel are needed.

Unlike digital communications through radio channels where data is transmitted by means of electromagnetic waves, mostly acoustic waves are used in underwater channels. The propagation characteristics of acoustic waves differ from those of radio waves and create a challenging environment for establishing reliable communication.The available bandwidth of an UWA channel is limited and depends on transmission loss and noise [1], [2]. Sound absorption loss increases with both range and frequency, and limits the available bandwidth. Fig. 1 illustrates the relationship between the available range, bandwidth and signal-to-noise ratio (SNR) at the receiver input [3]. As the range of the system increases, the available bandwidth dramatically decreases.

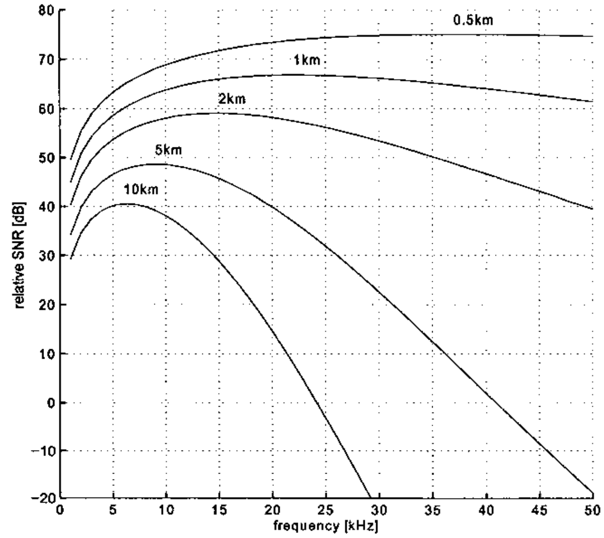

Figure 1: Relationship between range, bandwidth and SNR, which restricts the choice of system parameters.

Within this limited bandwidth the acoustic signals are also subject to time-varying multipath [3]. The multipath characteristics depend on the channel configuration, which can be vertical or horizontal. In the horizontal channel, multipath propagation occurs as a result of surface-bottom reflections. As the depth of the water decreases, the multipath spread of the channel increases and results in severe inter-symbol-interference (ISI). In addition to ISI, the time-varying characteristics of the channel cause Doppler spreading, which can be written in terms of velocity and operating frequency as

$$D = vf/c$$

where v is the speed of the receiver with respect to the transmitter, f is the carrier frequency, and c is the wave propagation speed $3 \times 10^8 m/s$ for free space radio waves [4]. Since acoustic signals propagate at $1500 m/s$, UWA channels produce larger Doppler spread than radio channels and make carrier recovery a challenging task [1].

Until recently, due to the severe Doppler spread, non-coherent modulation techniques were exclusively used for UWA communications. Since the channel varies rapidly, it is almost impossible to make an estimate that would be valid for the next symbol for low symbol rates. When the symbol rates increased, inter-symbol-interference also increases which makes reliable communication a very difficult task. In [3] it is shown that high-speed data transmission is possible using bandwidth efficient phase-coherent communications. In the present work, we use bit rates that assure the channel does not to vary significantly during two bit intervals. Thus, differentially coherent modulation can be employed.

For systems that operate in a network environment, in addition to channel impairments, multiple access interference is an important issue. In order to achieve high throughput, the network users must share the available frequency band in an efficient way. Frequency division multiple access (FDMA) is the simplest method, but since it requires separate frequency bands for each user, it cannot utilize UWA channels efficiently. Time division multiple access (TDMA) is an effective method used in radio channels, such as GSM. However, TDMA requires framing and exact synchronization, which is a difficult task in an UWA channel where propagation delays are extensive. Different from both methods, code division multiple access (CDMA) uses pseudo-noise codes to share the available bandwidth and time which results in an efficient multiuser environment [5]. The existing multiuser under water applications use FDMA. We employ direct-sequence spread-spectrum (DSSS) modulation that creates a basis for a CDMA based multi user environment.

In this paper, we present a DSSS based acoustic modem. The system operates in a very shallow water horizontal channel within a network environment where multiple users can access the channel randomly. We tested the proposed modem using experimental data. Section II. explains the DSSS modem. Section III. gives an explanation of the channel measurement device and illustrates the performance of the receiver used for channel measurements. Sections IV. and refsec-con contains results and conclusions, respectively.

II. Direct Sequence Spread Spectrum Modem

The transmitter employs direct sequence spread spectrum (DSSS) modulation. Two Gold codes of length 2047 are used for spreading the differentially coded data bits that are generated by a binary source. The same data bit is spread in both quadrature and in-phase branches of the acoustic channel. The bit rate of the system is 100 bps with a processing gain of 25 chips per bit. The block diagram of the transmitter is given in Fig. 2.

After spectral shaping the signals are carrier modulated. The carrier frequency is set to 12kHz. This frequency selection together with 40kHz sampling rate requires only twelve distinct values to modulate the signal. Thus instead of implementing a sinusoidal generator in the modem, we can store these six values in the memory. After carrier modulation the signals from in-phase and quadrature branches are combined and fed to the transducer.

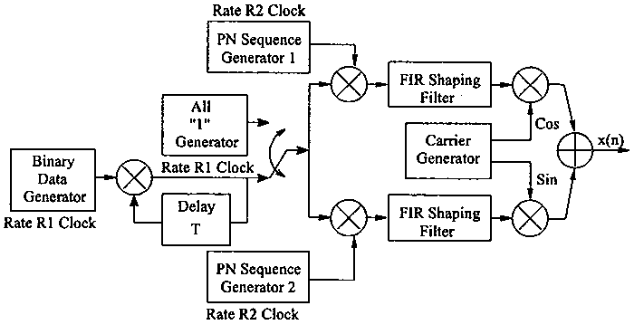

Figure 2: DSSS transmitter block diagram for underwater acoustic communication.

The first step in demodulating a spread spectrum signal is to despread the received pseudonoise (PN) signal. Despreading is accomplished by generating a replica of the PN signal (in our case two different codes for in-phase and quadrature components) and correlating this local PN signal with the incoming signal. To maximize the spreading gain, the local PN signal should be synchronized to the one superimposed on the received signal. Synchronization is performed in two stages:

- Acquisition, where the two PN signals are coarsely aligned within a small fraction of time;
- Tracking, where the fine tuning is performed using a closed-loop system.

For acquisition purposes, three short pulses, which are modulated in the same way as the data symbols, are added at the beginning of each transmission. These pulses are passed through filters that are matched to the PN signals used for modulation. At the output of the filters, three peaks are obtained that are used to coarsely estimate the PN signal timing. Fig. 3 shows an example of the received acquisition pulses and the output of the matched filters.

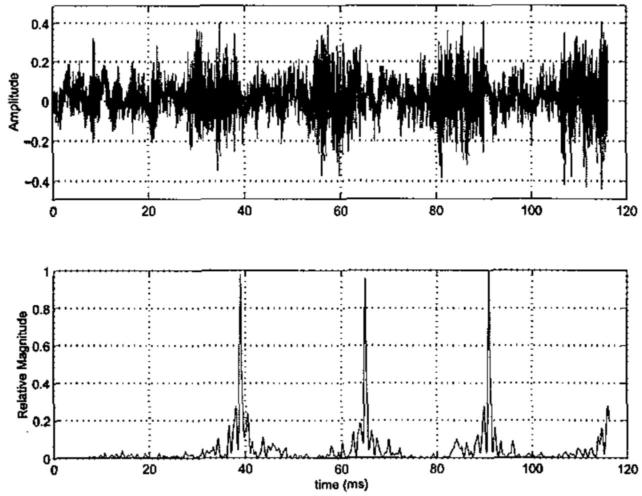

Figure 3: Received acquisitions pulses and the output of the matched filters.

A delay-locked loop (DLL), which is a type of early-late gate, is implemented for tracking the PN signals. The block diagram of the DLL is given in Fig. 4. After obtaining synchronization, the receiver begins demodulation of the data symbols.

DSSS systems are resistant to multipath propagation. These systems can reject the effects of multipath by utilizing the correlation properties of the PN sequences. However, rake receivers can make use of the energy present in multiple propagation paths. This energy extraction process is also known as time diversity.

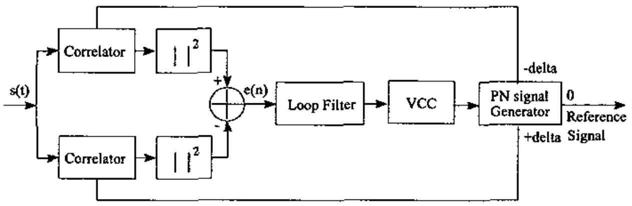

Figure 4: Block diagram of the DLL used in Type X receiver.

A rake receiver is a matched filter. If we represent the channel as a tap delay line, the rake taps are chosen as the conjugates of the channel taps and an optimum receiver is built. The same structure can also be realized with correlators. At each tap the time delayed replica of the received signal is correlated with the replica of the PN signal used to spread the data symbols. Because PN codes ideally have maximum correlation at zero delay and zero correlation at non-zero delays, we will get signal from taps that correspond to the multiple propagation paths. The rest of the taps won't give any output. Fig. 5 shows the rake filter where correlators are shown by simple multiplication.

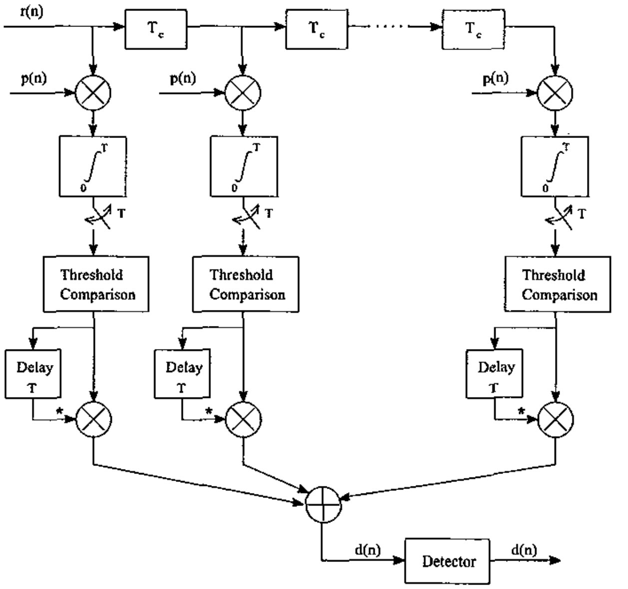

Figure 5: Rake receiver used for data demodulation.

The correlator at each tap of the rake receiver is shown in detail in Fig. 6. Signals at the taps are first multiplied by the corresponding PN sequences and the carrier. Next, the two pairs of signals are added and subtracted as shown in the figure. The final forms of the in-phase and quadrature signals at the output of the correlator are given in Equations (1) and (2), respectively:

$$y_c(n) = c_1 b_n (Cos\phi)(A_c^2(n) + A_s^2(n)) \tag{1}$$

$$y_s(n) = c_1 b_n (Sin\phi)(A_c^2(n) + A_s^2(n)) \tag{2}$$

where it is assumed that the channel has only one path with constant path gain c_1 and there is a ϕ degree of phase offset between receiver and transmitter local oscillators. $A_c(n)$ is the PN signal used in the in-phase branch and $A_s(n)$ is the PN signal used in the quadrature branch.

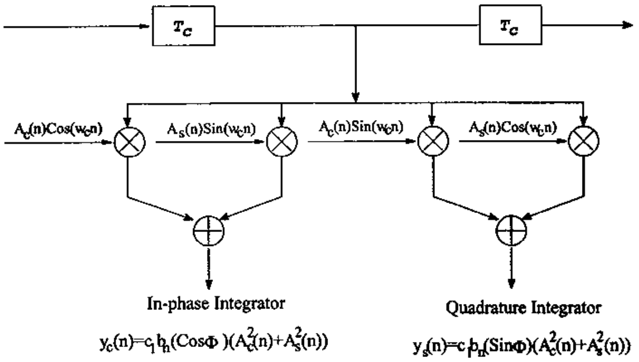

Figure 6: Correlator used at each tap of the rake receiver.

At the correlator output, we obtain the despread complex signal y(n), which is defined as

$$y(n) = y_c(n) + jy_s(n). \tag{3}$$

y(n) is than fed to the differential decoding block.

Since Gold codes have finite correlation values for other than zero delay, if there is not a multipath component at the corresponding tap, the output will be a signal-dependent self-noise, which degrades the performance of the receiver [4]. This problem is handled by including threshold devices at each tap. The output signal level of each tap is compared with a threshold. The ones that stay below the threshold are discarded. In this way the amount of noise at the combining stage is reduced. Finally, the signals that are above the threshold are summed and the polarity of these signals is used to make a decision on the incoming bit.

III. Channel Measurements

An important task in designing a communication system is to characterize the channel. Understanding the random time-varying behavior of the underwater acoustic channel will let us design transmitter-receiver pairs that can maximally exploit the limited capabilities of this environment. Therefore, channel measurements are required for the characterization of the subject environment.

Unmodulated spreading sequences are used for channel measurements; that is, the data sequence is all ones and differential coding is not employed. Thus only the unmodulated PN sequence is sent. The received signal is passed through a tap-delay line. At each tap, filters matched to the PN signals that are used in modulation are employed. Fig. 7 shows the channel estimator configuration.

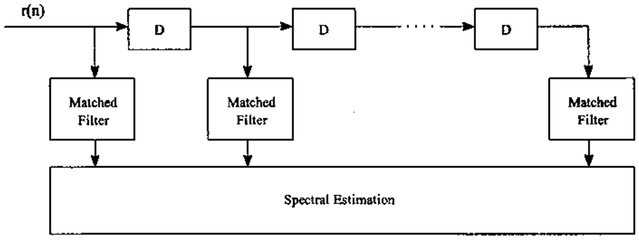

Figure 7: The system used for channel measurements. The filters are matched to the PN signals. Bartlett's method is used for spectral density estimation.

The multipath intensity profile (MIP) of the channel is the energy present in the channel for different time delays when it is excited with a unit pulse. Due to the correlation properties of the PN signals, the matched filter output averaged in power over independent trials represents the MIP of the channel. The signals obtained at different taps correspond to the MIP of the channel at different time instants. The scattering plot of the channel is determined by employing Bartlett's spectral density estimation method.

IV. Experimental Results

We tested the proposed modem using the experimental data provided by Datasonics, Inc in an experiment performed in the Baltic sea in March 99. The receiver was placed at about 30 m, while the transmitter was at 6 m. The approximate range is about 3000 m. We present the results of two events, where the second event is performed with 12 dB less output power with respect to the first event. Blocks of 200 bits are transmitted during both events. In addition to the data modulated signals, we used unmodulated signals to probe the channel.

Fig. 8 shows an ensemble of estimated channel responses. Each response shown along the delay axis represents an estimate of the channel at a given time instant. The separation between consecutive estimates is 60 ms; that is, the delay between receiver taps is set to 60 ms. Using this figure, we estimated the multipath spread of the channel as about 2.5 ms, which is equivalent to 6 rake taps. The multipath spread of the channel determines the number of taps that should be used in the rake receiver. If we use more taps than the multipath spread of the channel, the receiver performance degrades due to self-noise generated at the taps that do not contain signal components. On the other hand, if the span of the rake receiver is kept too small, we cannot make use of the time-diversity that is present in the channel.

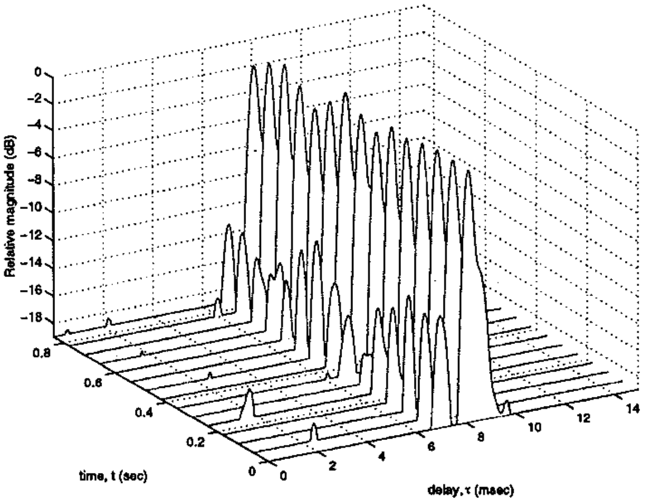

Figure 8: Variation of the channel. The multipath spread is about 2.5 ms.

The scattering function of the channel is given in Fig. 9. Each function along the frequency axis represents the Doppler power spectrum of the channel at the corresponding delay value. While Fig. 8 describes the signal coherence in time, the Doppler power spectrum describes signal spreading in frequency. From the results presented in Fig. 9, we found that there is a Doppler spread of about 0.5 Hz, which occurs as a shift in the frequency axis. In addition to the Doppler shift, variations in the channel creates code Doppler in DSSS systems that results in a change in the period of the received PN signals. This makes the tracking and demodulation a challenging task. If we know the Doppler shift a-priori, we can remove it from the received signal in the process of demodulation.

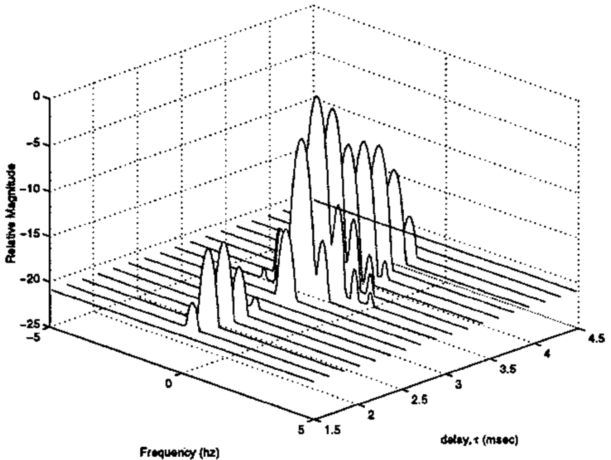

Figure 9: Scattering function of the channel. Doppler spread is about 0.5 Hz.

Fig. 10 represents the eye patterns obtained at the input of the decision device without using threshold devices. The receiver output SNR is 9.3 dB for the first event while it drops to 8 dB in the second event. In both events, there are no errors out of 200 bits. Fig. 11 shows the eye patterns with threshold set to 0.1. Since threshold devices prevent self-noise from entering the decision devices, we observe that the spread of the plots is reduces and the output SNR values increase to 9.33 dB and 8.5 dB, respectively. As the amount of noise in the system increases, thresholding becomes more effective.

V. Conclusions

We presented a DSSS based system for very shallow water communications. We employed rake filter at the receiver which makes use of the time-diversity that is present in the channel. The self-noise generated at the rake taps that do not contain useful energy is eliminated by employing threshold devices. The same signaling scheme is also used for probing the channel and obtained the channel characteristics that play a key role in determining the system parameters. Using experimental data, we demonstrated that the proposed system is capable of reliable communication in a very shallow water channel.

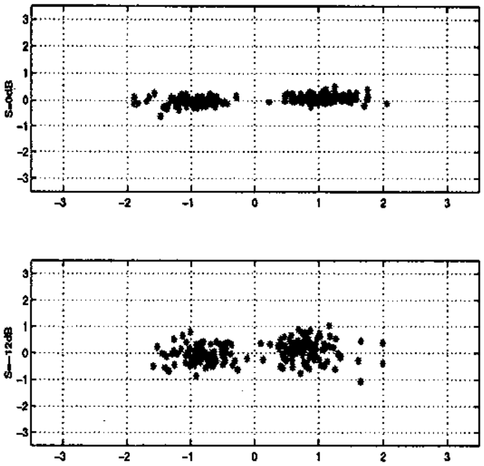

Figure 10: The eye plots obtained at the output of the receiver. Lower plot is obtained with 12 dB less output power with respect the the upper one. Threshold devices are not used. There are no detection errors in the block of 200 data bits.

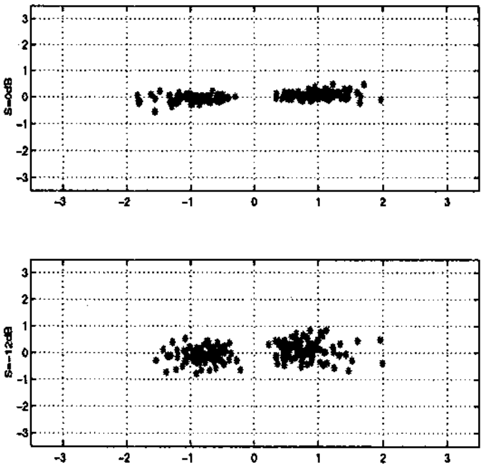

Figure 11: The eye plots obtained at the output of the receiver. Lower plot is obtained with 12 dB less output power with respect the the upper one. Threshold is set to 0.1. There are no detection errors in the block of 200 data bits.

References

[1] L. Berkhovskikh and Y. Lysanov, *Fundamentals of Ocean Acoustics*, New York:Springer, 1982.

[2] A. Quazi and W. Konrad, "Underwater acoustic communications," *IEEE Comm. Magazine*, pp. 24-29, Mar. 1982.

[3] M. Stojanovic,"Recent advances in high-speed underwater acoustic communications," *IEEE J. Oceanic Eng.*, vol. 21, pp. 125-136, Apr. 1996.

[4] J.G. Proakis, *Digital communications*, New York: McGraw-Hill, 1995.

[5] T.S. Rappaport, *Wireless communications*, NJ.: Prentice Hall, 1996.