Hypothesis-Feedback Equalization for Direct-Sequence Spread-Spectrum Underwater Communications

*M.Stojanovic and L.Freitag**

Department of Aeronautics and Astronautics, MIT, Cambridge, MA 02139

AOPE Department, Woods Hole Oceanographic Institution, Woods Hole, MA 02543

Abstract - **Direct-sequence code-division multiple-access is considered for underwater acoustic communication networks. Unlike in the majority of spread-spectrum radio systems, intersymbol interference cannot be neglected, and time variability of the channel requires that receiver adaptation be performed at the chip, rather than the bit rate. Adaptive decision-feedback equalization, which has successfully been used for single-user underwater communications, is not directly applicable to spread-spectrum signals because of the delay in the despreading process and the lack of reliable chip decisions. To overcome this problem, a receiver is proposed which feeds back hypothesized, rather than the actual decisions. Numerical examples demonstrate the receiver's ability to cope with time varying channel distortions and preserve the processing gain when conventional, symbol-rate adaptive methods fail.**

I. INTRODUCTION

In the design of an underwater acoustic communication network, one of the key tasks is the choice of a multiple-access strategy. In this paper, we focus on direct-sequence code division multiple-access (DS CDMA) as a technique for shallow water acoustic networks. This technique is used in conjunction with coherent detection of high-bandwidth signals, which is known to be possible through adaptive equalization.

The network of interest is a shallow water network with moving sources and receivers. The network covers an area of several square kilometers and operates in a 5 kHz band around a center frequency of 15 kHz. The sources are located on underwater crawlers and fast autonomous underwater vehicles. The network must support two modes of operation: single-user access to the base station at high data rate (1 kb/s - 5 kb/s) and multiuser access at low rate (20 b/s - 300 b/s). It is anticipated that up to four users may transmit at the same time. To fully utilize the available bandwidth, spread spectrum signals are transmitted at several kilochips/second (kc/s), while the processing gain L is varied from a few to several hundred chips per bit.

This work was sponsored by the Office of Naval Research under contract N00014-99-1-0287.

Signal processing techniques for DS CDMA communications have been extensively studied, with a large body of literature available on the application of various techniques to radio communications. Comprehensive summaries of these techniques can be found in [1], [2].

In a DS CDMA system, the signals of different users are distinguished by their spreading codes. Two types of receivers exist: single-user, or decentralized receivers, and multiuser, or centralized receivers. Single-user receivers detect only the desired user's signal using the knowledge of its spreading code. Multiuser receivers detect any number of signals using the knowledge of all the users' codes. Their performance is superior to that of single-user receivers, but their complexity is much higher. The problem of computational complexity is emphasized in underwater channels, where in addition to multiple-access interference (MAI), the receiver must overcome the intersymbol interference (ISI). In such channels, single-user receivers may still offer the only practical solution.

A conventional single-user receiver consists of a despreader, i.e., a filter matched to the spreading code of the desired user, followed by a decision device. Ideally, the spreading codes are orthogonal, so that the interfering signals are suppressed by despreading. However, the presence of a multipath channel and the lack of synchronicity among different users' signals destroy the orthogonality of codes in the received signal, leading to the near-far problem and seriously limiting the performance of this receiver. Adaptive single-user receivers have been found to offer a solution to this problem. These receivers are often favored in DS CDMA radio systems because of their additional capability to perform rake filtering, and thus gain multipath diversity. Among the single-user receivers, linear chip-spaced filters have attracted most attention, as they suffice for the majority of radio applications. Fractionally spaced, as well as nonlinear decision-feedback receivers have also been proposed for radio channels [3].

In underwater acoustic channels, multipath propagation causes time spread that may, depending upon L, exceed the duration of one symbol. Nonlinear receivers are better suited to serve the needs of time varying channels which exhibit spectral nulls within the signal bandwidth.

Fractionally spaced decision feedback receivers in both centralized and decentralized configurations have been used in underwater acoustic channels with low spreading gains [4].

In addition to multipath propagation, a major problem that arises in underwater channels is the time variability. This problem is particularly significant in mobile underwater communications where large Doppler shift and spread may arise. To compensate for the resulting signal distortion, adaptive algorithms capable of fast tracking must be employed. The receivers proposed for radio channels use adaptive filtering methods in which the filter coefficients are updated at the symbol rate. This rate of adaptation, however, may not suffice for a mobile underwater channel. In a DS CDMA system, the high-bandwidth received signal offers the possibility to track the channel at the chip rate, rather than the symbol rate. If the channel is time-invariant, or slowly time varying such that it does not change much over the duration of one symbol (L chips), the receiver can extract the processing gain before making a symbol decision. The symbol decisions are thus reliable and can be used to adjust the receiver coefficients for the following detection interval. Symbol-rate adaptation is based on such a principle. However, if the channel changes over one symbol cannot be neglected, the same set of filter coefficients will fail to produce a reliable symbol decision. In systems with fixed bandwidth, such a situation leads to an inverse dependence of performance on the processing gain: with an increase in the processing gain L, the symbol duration becomes longer, thus allowing greater changes. To recover performance, the receiver must adapt its coefficients more often than every symbol interval, e.g., at the chip rate. But to do so, the receiver needs symbol decisions, which are not available *before* despreading. This fact motivates the search for a detection strategy that provides symbol decisions before despreading has taken place. Hypothesis-feedback receiver proposed in this paper offers such a solution. A similar problem is found in adaptive suppression of narrowband interference in DS CDMA systems, and several techniques used to obtain early chip decisions are summarized in [6].

The receiver structure is presented in Sec.II. Its performance is analyzed on a simulated channel in Sec.III. Finally, conclusions are summarized in Sec.IV.

II. HYPOTHESIS-FEEDBACK RECEIVER

In a DS CDMA system, a number of users, I, may transmit simultaneously. On the uplink, their signals travel through different, possibly time varying channels, and arrive at the receiver asynchronously. The signal of each user is a BPSK signal (higher level linear modulations also apply) with a spreading sequence of duration $LT_c = T$, where L is the processing gain of the system, $R_c = 1/T_c$ is the chip rate and $R = 1/T$ is the symbol (bit) rate. The transmitted signal of the i-th user can be represented as

$$u_i(t) = \sum_k d_i(k) g(t - kT_c) \tag{1}$$

where $g(t)$ is the impulse response of the transmitter filter and $d_i(k)$ is the spread information sequence. Denoting the information symbol transmitted at time nT by $D_i(n)$, and the i-th user's code by $p_i(l)$, $l = 0, \ldots, L-1$, the spread sequence is given by

$$d_i(k) = D_i(n) p_i(l),\ k = nL + l,\ l = 0, \ldots L-1 \tag{2}$$

The signal $u_i(t)$ passes through a time varying multipath channel. At the receiver, this signal can be modeled as

$$\bar{v}_i(t) = \sum_p c_{i,p}(t) u_i(t - \tau_{i,p}) e^{j\theta(t)} \tag{3}$$

where $c_{i,p}(t)$ is the time varying complex-valued gain of the p-th propagation path, and $\tau_{i,p}$ is the corresponding path delay, which is assumed to be relatively fixed during a typical data packet so that its time variation is neglected. Additional phase distortion that may arise due to motion or offset of the local carrier is modeled by the term $\theta(t)$. The receiver observes a superposition of all the interfering signals in additive noise $z(t)$:

$$v(t) = \sum_{i=1}^{I} \bar{v}_i(t) + z(t) \tag{4}$$

In a great majority of radio channels, for which DS CDMA receivers have been developed, the delay spread of the channel is such that the ISI can be neglected. The channel is either flat fading (no dispersion at all) or it is frequency selective, but the dispersion is on the order of a few chip intervals only, and because $T >> T_c$, the ISI can be neglected. A rake receiver is employed in such a case to linearly combine the multipath arrivals. The situation in underwater acoustic channels is different. Time dispersion on the order of 10 ms is often found in these channels. At chip rates of several kc/s, dispersion equals tens of chip intervals, which, at moderate processing gains (~ 100) causes non-negligible ISI. In addition, the nature of multipath is such that the channel often exhibits spectral nulls and the rake, or any linear receiver, may not suffice. A decision-feedback equalizer (DFE) has successfully been used on the majority of such channels. However, in order to use a chip-spaced DFE, one must first obtain reliable chip decisions. Because of the delay in the despreading process, a conventional DFE cannot be applied directly. Instead, we propose to use a method of feeding back hypothesized chip decisions.

The receiver structure is shown in Fig.1. The input signal is fractionally sampled, e.g., every $T_s = T_c/2$, and

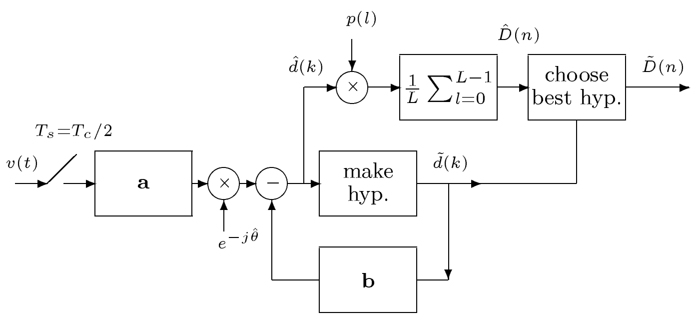

Figure 1. Hypothesis-feedback receiver.

fed into the feedforward filter. The filter output is produced once every chip interval, and a phase correction is performed at this point. The receiver now executes a number of steps.

1. Hypothesize

First, a hypothesis is made on the transmitted information symbol. For BPSK, there are two hypotheses: +1 or -1. If a +1 has been transmitted, the L chips corresponding to the n-th data symbol are equal to the chips of the spreading sequence. If a -1 has been transmitted, the chips are equal to the inverted code. Dropping the user index, these operations are written as:

$$\tilde{D}_{\pm}(n) = \pm 1 \Rightarrow \tilde{d}(nL + l) = \pm p(l), \; l = 0, \ldots L - 1 \quad (5)$$

2. Make chip estimates

For each hypothesized value of $D(n)$, the corresponding chips are fed back, and an estimate of the current chip is made as:

$$\hat{d}_{\pm}(k) = \mathbf{a}'_{\pm}(k)\mathbf{v}(k)e^{-j\hat{\theta}_{\pm}(k)} - \mathbf{b}'_{\pm}(k)\tilde{\mathbf{d}}_{\pm}(k), \quad k = nL, \ldots nL + L - 1 \quad (6)$$

where $\mathbf{a}'$ is the feedforward filter tap weight vector of size $1{\times}N$, $\mathbf{v}(k) = [\cdots v(kT_c{+}T_s)v(kT_c)v(kT_c{-}T_s)\cdots]^T$ is the signal vector stored in the feedforward filter at time kT_c, $\hat{\theta}$ is the phase estimate, $\mathbf{b}'$ is the feedback filter tap weight vector of size $1{\times}M$, and $\tilde{\mathbf{d}}_{\pm}(k) = [\tilde{d}_{\pm}(k{-}1)\tilde{d}_{\pm}(k{-}2)\ldots]^T$ is the vector of chip decisions (corresponding to either a +1 or a -1 hypothesis) stored in the feedback filter.

3.Despread

The chip estimates are now used for despreading:

$$\hat{D}_{\pm}(n) = \frac{1}{L}\sum_{l=0}^{L-1} p(l)\hat{d}_{\pm}(nL + l) \quad (7)$$

4.Choose best hypothesis

The squared error associated with each hypothesized data symbol is evaluated as

$$Q_{\pm}(n) = |\pm 1 - \hat{D}_{\pm}(n)|^2 \quad (8)$$

The final decision rule is to choose that data symbol which results in the lowest squared error:

$$\tilde{D}(n) = \arg \min_{+1,-1} Q_{\pm}(n) \quad (9)$$

5.Update receiver coefficients

An adaptive receiver may choose to update its parameters at the symbol rate or at the chip rate. The filter coefficients are updated using an algorithm such as LMS or RLS, while the phase estimate is updated using a decision-directed phase-locked loop (PLL). The adaptation towards the MMSE solution is driven by an error signal which is different under each hypothesis. Thus, under each hypothesis a set of receiver parameters is computed.

If chip rate updating is used, the error signal that drives the adaptation is the chip estimation error:

$$e_{\pm}(k) = \pm p(l) - \hat{d}_{\pm}(k), \; k = nL{+}l, \; l = 0, \ldots L{-}1 \quad (10)$$

The receiver coefficients are updated throughout the L chip intervals allocated for the detection of the n-th data symbol. Upon making the final symbol decision, filter coefficients and phase estimates corresponding to the winning hypothesis are retained for the next iteration.

If symbol rate updating is used, adaptation is driven by the symbol error:

$$E_{\pm}(n) = \pm 1 - \hat{D}_{\pm}(n) \quad (11)$$

Two sets of receiver parameters are evaluated for the two BPSK hypotheses. This is done in a single step, i.e., no chip-rate updating is required. The values of the winning set are used in the next iteration.

The five algorithm steps given above can also be cast in the multichannel receiver framework. This is a very important aspect for applications to underwater acoustic communications where spatial diversity often provides the margin necessary for reliable detection. In fact, it was shown in [4] that given array processing, a decentralized multiuser receiver matches the performance of its centralized counterpart.

Computational complexity is another important consideration in the implementation of underwater acoustic systems. The computational complexity of the hypothesis-feedback algorithm is reduced if symbol rate, rather than chip rate updating is used, but this causes performance degradation on a rapidly varying channel. In the underwater acoustic channels, it is the signal phase that varies most rapidly. Thus, it is advantageous to consider strategies in which the phase estimate $\hat{\theta}$ is updated every chip interval, while the filters are updated only every symbol interval. The benefits of various updating strategies will depend on the actual channel, as well as the system parameters. The principles of hypothesis-feedback are applicable to systems with arbitrary processing gain, as well as to non-spread systems which use a conventional DFE. Because hypothesized decisions, rather than (possibly unreliable) actual decisions are fed back, this receiver must outperform any of its linear or decision-feedback counterparts.

In implementing decision-feedback in a spread-spectrum system, one cannot count on sufficient SNR per chip to have reliable chip decisions. One solution is to feed back symbol decisions, after front-end processing of the signal by a chip-spaced (or a fractionally chip-spaced) feedforward filter which extracts the processing gain and any multipath diversity. Such an approach was proposed in [3]. In this reference, two methods of front-end filtering were investigated. In the first, a sequence-matched filter is used prior to adaptive fractionally spaced feedforward filtering. In the second, no explicit sequence-matched filtering is performed, and this task is left entirely to the feedforward filter. Because a training sequence is employed, the two approaches were shown to have comparable performances.

We use a slightly modified version of the receiver from [3]. As shown in Fig.2, despreading is performed after fractionally chip-spaced feedforward filtering and a PLL is added. Decision-directed phase correction can only be accomplished at the symbol rate because chip decisions are not available before despreading. The equalizer coefficients are updated using an RLS algorithm. Note that fewer feedback taps are needed in this implementation. The number of feedback taps is reduced by L, the processing gain of the system. However, because symbol-rate adaptation is used, L must be kept small in applications to rapidly varying channels.

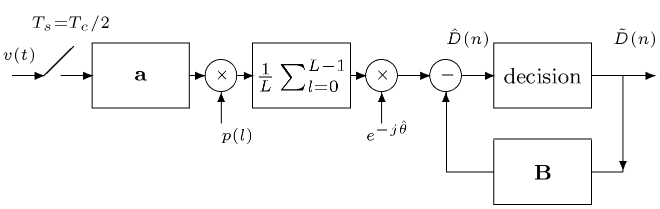

Figure 2. Decision-feedback receiver.

III. RECEIVER PERFORMANCE

To demonstrate the receiver performance, a simulated channel is used. The shallow water channel is modeled using a direct and several surface-reflected paths. The delays and average path gains can roughly be calculated from the system geometry. For example, assuming a range of 5 km and a depth of 75 m, and taking into account only the path loss due to practical spreading and energy absorption at the carrier frequency of 15 kHz, the first paths are found at relative delays of 0, 1.5 ms, 6 ms, 13.5 ms, etc. The relative average path gains are 1, 0.3, 0.05, and quickly decay thereafter. Thus, we choose to model the channel as having three paths with a multipath spread of T_m=6 ms. However, the geometric model is a simplified one, and it is known from practical experience that stronger multipath is often present in shallow water. To allow for such a possibility, the three paths are taken to be of equal energy.

Each path gain can now be modeled as a random process in time. For lack of a well-accepted fading model, we adopt a Gauss-Markov modeling approach. In our model, each path gain is represented as a second-order auto-regressive (AR2) process driven by complex-valued zero-mean white Gaussian noise. The 3 dB bandwidth B_d of the Doppler power spectrum defines the Doppler spread, which may be different for each path. In addition to the path gain variation, there is in practical channels a time varying phase deviation, which causes a frequency offset from the nominal carrier f_c. At a relative velocity v, the frequency offset, or the Doppler shift is given by $f_d = f_c v/c$, where c=1500 m/s is the speed of sound.

The BPSK signals are shaped by a transmitter filter that has a raised cosine spectrum with roll-off factor α=0.25. The transmitter bandwidth B=5 kHz is fully utilized at the chip rate $R_c = B/(1+\alpha)$=4 kc/s, while the spreading gain is varied. The spreading codes are chosen from the Kasami sequences [5] (we are not concerned with the design of optimal sequences). A set of Kasami sequences generated using a shift register of length m contains $2^{m/2}$ sequences, each of length $2^m - 1$. Sequences of length 15, 63 and 255 are used to achieve the data rates of 266 b/s, 64 b/s and 16 b/s, respectively.

The characteristics of a simulated channel for the desired user are shown in Fig.3. Listed in the figure are the various simulation parameters. The SNR per bit is defined as E_b/N_0, the ratio of the bit energy to the power spectral density of the AWGN. The path gain magnitudes are normalized such that $\sum_p E\{|c_{ip}^2(t)|\} = 1$. The chip energy is $E_c = E_b/L$, so that the SNR per chip is SNR$_c$=SNR$/L$. The Doppler spread is set to $B_d = 1.2$ Hz on all three paths. This value results in a normalized Doppler spread $\pi B_d T_c = 10^{-3}$, which is close to the practical limit for the performance of coherent detection. The path gain variation, shown over 500 symbol intervals, is indeed considerable (in fact, a practical channel is likely to exhibit a much slower variation). Assuming a velocity of v=2 knots, the frequency offset is 10 Hz, which causes significant carrier phase degradation. At the same time, motion induces pulse compression/dilation by a factor of $(1 \pm v/c)$. With the chosen system parameters, it can be assumed that a $T_c/2$ fractionally spaced adaptive receiver is capable of recovering correct timing.

The performance of the receiver on this channel without multiple-access interference is illustrated in Fig.4. This case is of interest to the present application where it is expected that most of the time a single user will be transmitting to the receiver. The eye pattern of the chip

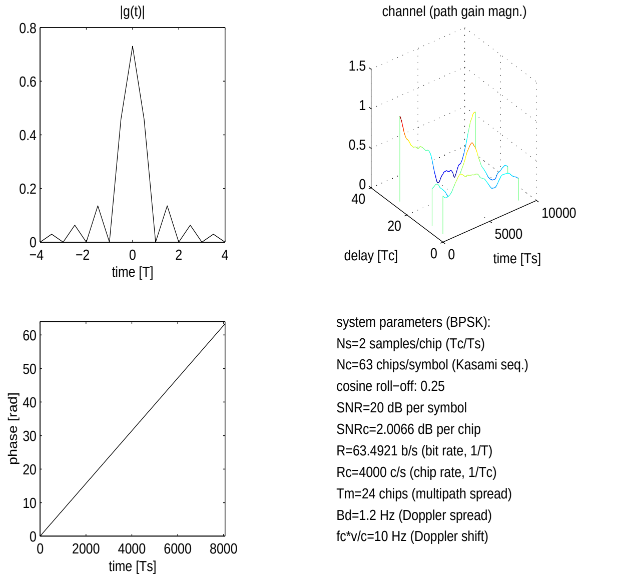

Figure 3. Example of a simulated channel: transmitter filter impulse response, channel path gain magnitudes, and carrier phase variation.

estimates $\hat{d}(k)$ corresponding to the winning hypotheses is completely closed due to the low SNR per chip. Nevertheless, excellent symbol detection is achieved, as illustrated by the eye pattern of the estimated data symbols $\hat{D}(n)$. The receiver parameters are listed in the figure. The length of RLS training is determined approximately as $2{\cdot}(N{+}M)$ chips. If training is accomplished within the code length L, the receiver can as well operate blindly, as indicated in the figure by $N_t = 0$ training symbols.

If the same signal is processed by the symbol rate adaptive DFE of Fig.2, convergence cannot be established even if perfect feedback is used. Failure of the receiver is explained by the fact that symbol rate adaptation at the processing gain of 63 chips per symbol is too slow to track the channel. In fact, phase variation alone has a similar effect on the receiver performance.

Fig.5 illustrates the performance of the hypothesis-feedback receiver on the three-path channel where path gain variation is negligible, but the carrier phase varies due to a velocity of 2 knots. There are four users present in the system now. Because the interfering users transmit from different locations, their channels are likely to be different. Accordingly, each user is independently assigned an equal-energy three-path channels, with path delays distributed in a similar, but not identical manner. Different users also experience different Doppler shifts. The interfering signals are thus received asynchronously. In particular, the path delays are 0.5, 3, 23 chips for the first interferer; 1, 4, 11.5 for the second, and 1.5, 2.5, 13 chips for the third interferer. Doppler shifts are calculated for the velocities of 0.2, 0.5 and 3 knots. The interfering users' powers are equal, and the signal-to-

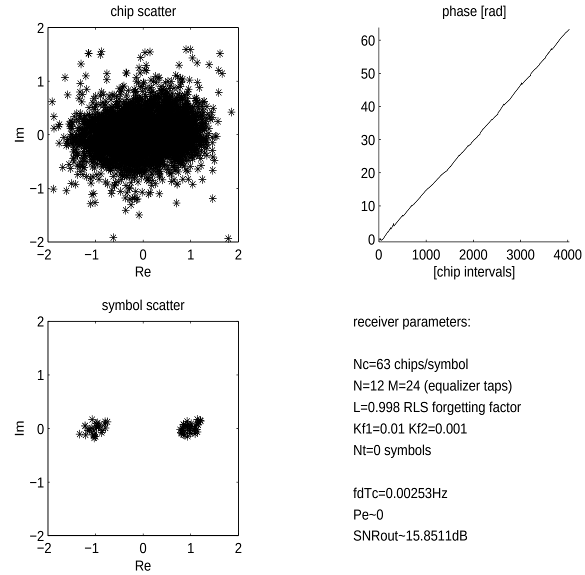

Figure 4. Performance of the hypothesis-feedback receiver. Single user, SNR=20 dB.

interference ratio is SIR=0 dB. In the presence of strong interference, the receiver needs a longer training period, but the performance is excellent.

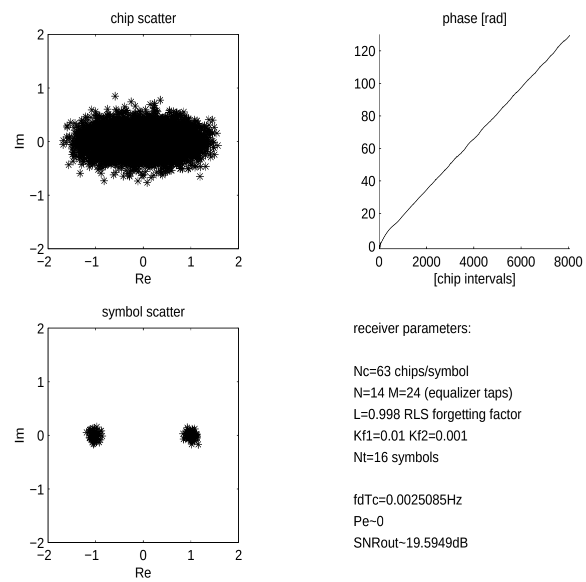

Figure 5. Performance of the hypothesis-feedback receiver. Four users, SIR=0 dB, SNR=20 dB.

Fig.6 summarizes the receiver performance in the four-user multiple-access configuration. System parameters are unchanged from the previous example. The performance, measured by the output SNR per bit (inverse of the average squared error in bit estimates), is shown as a function of the processing gain. The performance of hypothesis-feedback receiver improves with an increase in processing gain (even though the SNR per chip decreases). For the symbol rate adaptive DFE, the situation is reversed. At the processing gain of 15, it provides performance close to that of the hypothesis-feedback receiver; however, at processing gains of 63 and 255, it fails despite the fact that thermal noise power is negligible. While reduction of the processing gain does enable this receiver to operate, it increses sensitivity to MAI. The chip rate adaptive hypothesis-feedback receiver does not suffer from the effect of inverse performance dependence, and is thus free to use a higher processing gain.

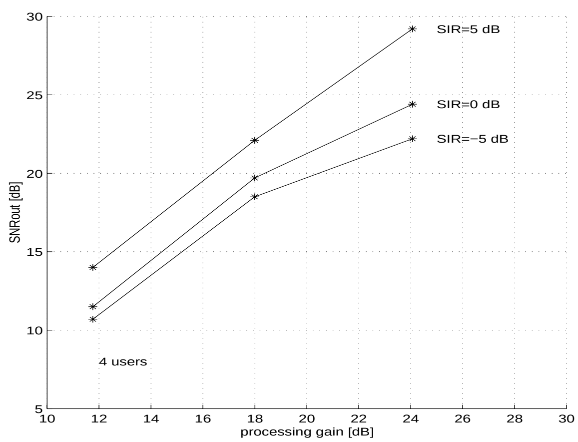

Figure 6. Output SNR as a function of processing gain. SNR=35 dB.

The above observations should not discourage the use of the symbol rate adaptive feedback receiver. Its good feature is computational efficiency, which results from both the fact that equalizer is updated only every symbol interval, and from the fact that only a short feedback filter is needed. The number of feedback taps is equal to the number of *symbols* that span the multipath spread. This number varies with the processing gain. For the three-path channel in our example, 2, 1 and 0 taps would be needed for processing gain of 15, 63 and 255. At the same time, the chip-spaced feedback always uses 24 taps, or the number of *chips* needed to span the multipath spread. (The same number of of feedforward taps is used in both receivers.) Thus, to reduce the computational complexity, but retain the adaptation speed, a receiver structure that combines chip rate phase tracking and symbol rate equalizer updating should be considered.

IV. CONCLUSIONS

Multiple-access communications based on DS CDMA are being considered for use in underwater acoustic networks. The focus is on the design of an adaptive decentralized receiver that can operate in the presence of time varying multipath in addition to multiple-access interference. Similar receivers developed for radio channels are based on symbol rate adaptation, which may be too slow for applications to shallow water acoustic channels.

Better channel tracking can be accomplished by a chip rate adaptive receiver. However, to form the error signals necessary for the system adaptation, chip decisions are needed *before* despreading, but the SNR per chip is not sufficient for these decisions to be reliable. To overcome this problem, the principle of hypothesized decisions is proposed. The receiver maintains two sets of receiver coefficients, one corresponding to each hypothesized BPSK symbol. Once the despreading has been accomplished, the hypothesis with a lower squared error is chosen. In this manner, hypothesis-feedback receiver enables chip rate adaptation. This, in turn, eliminates performance dependence on the symbol rate. In systems with fixed chip rate it thus allows the use of higher processing gains, which are needed for effective suppression of strong multiple-access interference.

The improved performance is obtained at the price of increased computational complexity. Better efficiency can be achieved if chip rate phase tracking is combined with symbol rate adaptive equalization. Also, a symbol-spaced hypothesis-feedback filter can be considered for reducing the receiver size. The benefits of such an approach are ultimately determined by the channel.

In a practical implementation, a multichannel receiver configuration should be used if possible. The array processing gain effectively increases SIR by separating the interferers spatially. At high SIR, hypothesis-feedback receiver can also operate blindly, which makes it an appealing choice for code acquisition in DS CDMA systems.

REFERENCES

[1] M.Majmundar, N.Sandhu and J.Reed, "Adaptive Single-User Receivers for Direct-Sequence Spread-Spectrum CDMA Systems," *IEEE Trans. Vehic. Technology*, pp.379-389, March 2000.

[2] G.Woodward and B.Vucetic, "Adaptive Detection for DS-CDMA," *Proceedings of the IEEE*, pp.1413-1434, July 1998.

[3] M.Abdulrahman, A.Sheikh and D.Falconer, "Decision feedback equalization for CDMA indoor wireless communications," *IEEE J. Select. Areas Commun.*, pp.698-706, May 1994.

[4] M.Stojanovic and Z.Zvonar, "Multichannel processing of broadband multiuser communication signals in shallow water acoustic channels," *IEEE J. Oceanic Eng.*, pp. 156-166, Apr. 1996.

[5] J.G.Proakis, *Digital Communications*, New York: Mc-Graw Hill, 1995.

[6] L.Rush and H.Poor, " Narrowband Interference Suppression in CDMA Spread Spectrum Communications," *IEEE Trans. Commun.*, pp.1969-1979, Apr. 1994.

HYPOTHESIS-FEEDBACK EQUALIZATION FOR DIRECT-SEQUENCE SPREAD-SPECTRUM UNDERWATER COMMUNICATIONS

*M.Stojanovic and L.Freitag**

Department of Aeronautics and Astronautics, MIT, Cambridge, MA 02139

AOPE Department, Woods Hole Oceanographic Institution, Woods Hole, MA 02543

Direct-sequence code-division multiple-access is considered for underwater acoustic communication networks. Unlike in the majority of spread-spectrum radio systems, intersymbol interference cannot be neglected, and time variability of the channel requires that receiver adaptation be performed at the chip, rather than the bit rate. Adaptive decision-feedback equalization, which has successfully been used for single-user underwater communications, is not directly applicable to spread-spectrum signals because of the delay in the despreading process and the lack of reliable chip decisions. To overcome this problem, a receiver is proposed which feeds back hypothesized, rather than the actual decisions. Numerical examples demonstrate the receiver's ability to cope with time varying channel distortions and preserve the processing gain when conventional, symbol-rate adaptive methods fail.