An Underwater Acoustic Telemetry Modem for Eco-Sensing

Ronald A. Iltis, Hua Lee, Ryan Kastner,
Daniel Doonan, Tricia Fu, Rachael Moore and Maurice Chin
Department of Electrical and Computer Engineering
University of California,
Santa Barbara, CA 93106-9560
Tel: (805) 893-4166, E-mail: {iltis,kastner,lee}@ece.ucsb.edu

Abstract— **An underwater acoustic telemetry modem is presented for ecological research (Eco-sensing) applications. The modem is intended for the physical layer (PHY) in an Ad hoc network of "AquaNodes". Each AquaNode will incorporate an acoustic modem, router and host processor with standardized interfaces to oceanographic sensors. The transmitted waveforms use a composite Walsh/m-sequence format in which each bit in an 8-bit long Walsh function is spread by a 7-chip m-sequence. The resulting waveform has a 5 kHz bandwidth for robustness to multipath and achieves a 133 bps data rate. An 11 msec. time-guard band for channel clearing eliminates the need for equalization. The receiver employs a Generalized Multiple Hypothesis Test with a Matching Pursuits (GMHT-MP) inner loop for symbol-by-symbol channel estimation. Symbol-error rate (SER) results demonstrate that the GMHT-MP is superior to the noncoherent RAKE receiver at usable SERs ($< 10^{-2}$) and can accommodate up to 10 msec. multipath spreads with Doppler spreads on the order of 1 Hz.**

I. Introduction

Underwater *Eco-Sensing* deals with telemetry of marine environmental data (conductivity, temperature, depth, currents). There is a great need for undersea wireless networks for Eco-Sensing allowing remote monitoring and adaptive sampling without requiring physical retrieval of instruments. However, underwater wireless networks developed so far for telemetry [1][2] rely on expensive, high power acoustic modems typically based on M-ary FSK [3][4]. The fundamental obstacle to robust UWA communications is multipath which introduces severe intersymbol interference (ISI). M-FSK modems use narrowband tones with duration much greater than the multipath spread, thus eliminating ISI. Alternatively, equalizers [5] and direct-sequence spread-spectrum [6][7][8] modulation have been employed to reduce the effects of ISI and frequency-selective multipath.

An alternative underwater acoustic (UWA) modem using direct-sequence signaling is presented here based on composite Walsh and m-sequence waveforms. The advantage of such Walsh/m-sequence signaling is that the waveform is instantaneously wideband, in contrast to M-FSK, thus providing robustness to frequency-selective multipath. The application considered is underwater telemetry in a short-range ($<$ 500 m) shallow water channel, where multipath spreads can easily extend to 10 msec. [7][9]. The target bit rate is 133 bps, which is adequate for conductivity/temperature/depth (CTD) measurements in Eco-Sensing applications. The UWA modem will be incorporated in "AquaNodes" to form an Ad hoc UWA network. Each AquaNode will contain a host processor, router, and standard sensor interfaces for deployment in a wide range of Long Term Ecological Research (LTER) sites including coral reefs, lakes and coastal areas.

The notation and parameters employed are as follows. $\mathbf{x}$ is a column vector, $\mathbf{M}$ is a matrix, $(\mathbf{x})_i$ is the i-th element of a vector, and $(\mathbf{M})_{i,j}$ is the i,j-th matrix element. Where necessary, Matlab notation is employed with $\mathbf{M}(:(i), i(:))$ denoting the i-th column (row) of a matrix for example. The system parameters are

Chip duration: $T_c = .2$ msec.
Sampling interval: $T_s = T_c/2 = .1$ msec.
m-sequence length: $L_{pn} = 7$ chips.
Walsh sequence length: $N_w = 8$ binary symbols
Symbol duration: $T_{sym} = L_{pn}N_wT_c = 11.2$ msec.
Samples/symbol: $N_s = T_{sym}/T_s = 112$.
Time guard interval: $T_g = T_{sym}$ sec. per symbol.
Bit rate: $R_b = (3\text{bits/symbol})/(T_{sym} + T_g) = 133$ bps.
Binary Walsh sequences: $\mathsf{w_m} \in \{\pm 1\}^{N_W}$ for $\mathsf{m} =$

$1, \ldots, N_W$.
m-sequence: $\mathbf{c} \in \{\pm 1\}^{L_{pn}}$.
Transmitted Walsh/m-sequence: $\mathbf{d}_{\mathsf{m}} = \mathbf{w}_{\mathsf{m}} \otimes \mathbf{c} \in \{\pm 1\}^{N_W L_{pn}}$, where $\otimes$ is the Kronecker product.
Complex multipath gains: $\alpha_1(n), \ldots, \alpha_{N_\alpha}(n) \in \mathbb{C}$.
Path delays: $\tau_1(n) < \tau_2(n) \ldots < \tau_{N_\alpha}(n) \in [0, T_{sym}]$.

The Walsh/m-sequence transmitted waveform is shown in Fig. 1. The Walsh sequences are generated via the recursion of Hadamard matrices $\mathbf{H}^{N_W} = \mathbf{H}^{N_W/2} \otimes \mathbf{H}^2$ where $\mathbf{H}_2 = [11; 1 - 1]$ (Matlab notation) and $\mathbf{H}^1 = 1$. Hence the m-th sequence is the m-th row of the Walsh matrix $\mathbf{w}_{\mathsf{m}} = \mathbf{H}^{N_W}(\mathsf{m}, :)$. The specific m-sequence employed is $\mathbf{c} = [1, 1, -1, 1, -1, -1, -1]^T$. The transmitted waveform is then

$$s_T(t) = \sum_{n=0}^{\infty} (\mathbf{d}_{\mathsf{m}(n)})_i g(t - iT_c - 2nT_{sym}), \quad (1)$$

where $g(t)$ is a raised-cosine pulse with bandwidth β/T_c, $.5 < \beta \leq 1$ and $\mathsf{m}(n)$ is the symbol transmitted in epoch $[2nT_{sym}, (2n+1)T_{sym})$. Note that the sampling interval $T_s = T_c/2$ corresponds to an oversampling rate of two for zero percent excess bandwidth.

II. Channel Model and Generalized Multiple Hypothesis Test

The received waveform in $t \in [2nT_{sym}, 2(n+1)T_{sym})$ is given by

$$r(t) = \sum_{p=1}^{N_\alpha} \alpha_p(n) s_{\mathsf{m}(n)}(t - \tau_p(n)) + n(t), \quad (2)$$

where

$$s_{\mathsf{m}(n)}(t) = \sum_{i=0}^{N_w L_{pn}-1} (\mathbf{d}_{\mathsf{m}(n)})_i g(t - iT_c), \quad (3)$$

and $n(t) \in \mathcal{C}$ is additive wideband receiver/ambient noise. The duration of $s_{\mathsf{m}}(t)$ is $T_{sym} = 2N_W L_{pn} T_s$ sec., hence $r(t)$ in (2) includes a $T_g = T_{sym}$ sec. time-guard interval. Under the assumptions that $0 \leq \tau_p(n) \leq T_{sym}$, and coarse symbol acquisition has been achieved, inter-symbol interference is thus eliminated in (2). (Symbol sync is discussed in Section IV.) The noise $n(t)$ is approximated as circular white Gaussian with spectral density N_0. Thus, assuming bandlimiting to $1/T_c = 2/T_s$ Hz., the variance of $n(t)$ is $E\{|n(t)|^2\} = 2N_0/T_s$. The channel parameters are assumed constant in the interval $2T_{sym}$ seconds, but independent from symbol-to-symbol hence the $\alpha_p(n), \tau_p(n)$ vary with epoch n.

In order to simplify the receiver and subsequent presentation, an approximate tapped-delay line model for the multipath is assumed [10]. Thus

$$r(t) \approx \sum_{l=0}^{N_s-1} f_l s_{\mathsf{m}(n)}(t - lT_s) + n(t). \quad (4)$$

where $f_l \in \mathbb{C}$ are the coefficients from an interpolation formula approximation to the true channel. Since the channel is typically sparse, the f_l will be estimated under the constraint that only $N_f << N_s$ coefficients are nonzero. The received signal is sampled at rate $1/T_s$ over an interval $2T_{sym}$ to include the time-guard interval. The resulting received vector model is then

$$\mathbf{r}(n) \approx \sum_{l=0}^{N_s-1} f_l \mathbf{s}_{\mathsf{m}(n)}(l) + \mathbf{n}(n), \quad (5)$$

where $\mathbf{r}(n), \mathbf{s}_{\mathsf{m}(n)}(l) \in \mathbb{C}^{2N_s}$ are

$$\mathbf{r}(n) = \quad (6)$$
$$[r((2(n+1)N_s - 1)T_s) r((2(n+1)N_s - 2)T_s) \ldots r(2nN_sT_s)]^T$$
$$\mathbf{s}_{\mathsf{m}(n)}(l) = [s_{\mathsf{m}(n)}((2N_s - 1 - l)T_s) s_{\mathsf{m}}((2N_s - 2 - l)T_s) \ldots, s_{\mathsf{m}(n)}(-lT_s)]^T.$$

The channel coefficients $f_l(n)$ are unknown and may be rapidly varying (e.g. Doppler spreads of .5 Hz are reported in [11].) Channel estimation techniques such as Kalman filtering are too complex for implementation in the envisioned low-cost telemetry modem. Here, a Generalized Multiple Hypothesis Test receiver is proposed, using the Matching Pursuits algorithm [12][13] for instantaneous channel estimation. The GMHT is compactly expressed as the decision rule (dropping the time index)

$$\hat{\mathsf{m}} = \arg\min_{\mathsf{m}} \left\{ \min_{\mathbf{f}:|\mathbf{f}|=N_f} ||\mathbf{r} - \mathbf{S}_{\mathsf{m}}\mathbf{f}||^2 \right\}, \quad (7)$$

where $\mathbf{f} = [f_0 f_1 \ldots f_{N_s-1}]^T$. The constraint $|\mathbf{f}| = N_f$ indicates that at most $N_f << N_s$ elements of $\mathbf{f}$ are nonzero. The signal matrix $\mathbf{S}_n \in \mathbb{C}^{2N_s \times N_s}$ is

$$\mathbf{S}_{\mathsf{m}} = [\mathbf{s}_{\mathsf{m}}(0)\mathbf{s}_{\mathsf{m}}(1) \ldots \mathbf{s}_{\mathsf{m}}(N_s - 1)]. \quad (8)$$

Unfortunately, the exact minimization over $\mathbf{f}$ in the GMHT (7) has complexity growing exponentially in N_s. Hence, the GMHT is replaced by

$$\hat{\mathsf{m}} = \arg\min_{\mathsf{m}} \left\{ ||\mathbf{r} - \mathbf{S}_{\mathsf{m}}\hat{\mathbf{f}}_{\mathsf{m}}^{MP}||^2 \right\}, \quad (9)$$

where $\hat{\mathbf{f}}_{\mathsf{m}}^{MP}$ is the Matching Pursuits estimate of the channel conditioned on the m-th Walsh sequence being transmitted. In the sequel, $\hat{\mathbf{f}}_{\mathsf{m}}$ will always refer to the sequence conditional MP estimate.

III. GMHT-MP Receiver

The Matching Pursuits algorithm is described next. It is computationally efficient to employ a sufficient statistics representation of the cost function in (9) as follows:

$$-||\mathbf{r} - \mathbf{S}_{\mathsf{m}}\hat{\mathbf{f}}||^2 \propto 2Re\{\mathbf{v}_{\mathsf{m}}^H\hat{\mathbf{f}}\} - \hat{\mathbf{f}}^H\mathbf{A}_{\mathsf{m}}\hat{\mathbf{f}}, \qquad (10)$$

where $\mathbf{v}_{\mathsf{m}} = \mathbf{S}_{\mathsf{m}}^H\mathbf{r}$ and $\mathbf{A}_{\mathsf{m}} = \mathbf{S}_{\mathsf{m}}^H\mathbf{S}_{\mathsf{m}}$.

Recall that the problem is to estimate $\mathbf{f} \in \mathbb{C}^{N_s}$ under the constraint that only $|\mathbf{f}| = N_f << N_s$ coefficients are nonzero. In order to obtain such a sparse solution, MP iteratively updates the assumed channel order and estimates the next coefficient using a canceled signal. Specifically, at iteration $k = 1$ of MP, it is assumed that $N_f = 1$ and the index of the detected path is [12]

$$q_1 = \arg\max_i \left(2Re\{(\mathbf{v}_{\mathsf{m}})_i^*\hat{\mathbf{f}}_i\} - (\mathbf{A}_{\mathsf{m}})_{i,i}|\hat{f}_i|^2\right). \qquad (11)$$

The estimate $\hat{\mathbf{f}}_i$ is

$$\hat{\mathbf{f}}_i = \arg\max_f 2Re\{(\mathbf{v}_{\mathsf{m}})_i^* f\} - (\mathbf{A}_{\mathsf{m}})_{i,i} f = \frac{(\mathbf{v}_{\mathsf{m}})_i}{(\mathbf{A}_{\mathsf{m}})_{i,i}}. \qquad (12)$$

Thus, the detected index on MP iteration $k = 1$ reduces to $q_1 = \arg\max_i |(\mathbf{v}_{\mathsf{m}})_i|^2/(\mathbf{A}_{\mathsf{m}})_{i,i}$.

At stage k of MP, a canceled signal is formed as

$$\mathbf{r}^k = \mathbf{r} - \sum_{i=1}^{k-1}\mathbf{S}_{\mathsf{m}}(:, q_i)\hat{f}_{q_i}. \qquad (13)$$

The signal $\mathbf{r}^k$ is then used to detect the next path index q_k [12][13]. The sufficient statistic becomes

$$\mathbf{v}_{\mathsf{m}}^k = \mathbf{S}_{\mathsf{m}}^H\mathbf{r}^k = \mathbf{v}_{\mathsf{m}} - \sum_{i=1}^{k-1}\mathbf{A}_{\mathsf{m}}(:, q_i)\hat{f}_{q_i}. \qquad (14)$$

Using (14), the k-th detected coefficient index is

$$q_k = \arg\max_{i\neq q_1,\ldots,q_{k-1}} \frac{|(\mathbf{v}_{\mathsf{m}}^k)_i|^2}{(\mathbf{A}_{\mathsf{m}})_{i,i}}. \qquad (15)$$

Note that $(\mathbf{A}_{\mathsf{m}})_{i,i}$ is not required in eq. (15) for the equal-energy signal set employed here. Table I summarizes the combination of Matching Pursuits with the GMHT.

```
For m = 1, 2, ... N_W
    Get current received vector r
    v_m^1 = S_m^H r
    Initialize channel estimate f̂_m = 0
    For k = 1, 2, ..., N_f
        q_k = arg max_{i≠q_1,...,q_{k-1}} |(v_m^k)_i|^2 / (A_m)_{i,i}
        (f̂_m)_{q_k} = (v_m^k)_{q_k} / (A_m)_{q_k,q_k}
        v_m^{k+1} = v_m^k − A_m(:, q_k)(f̂_m)_{q_k}
    Next k
Next m
Make transmitted symbol decision using GMHT
m̂ = arg max_m 2Re{v_m^H f̂_m} − f̂_m^H A_m f̂_m
```

TABLE I

GMHT-MP Algorithm for the UWA modem.

IV. Symbol Synchronization Algorithm

The GMHT-MP algorithm in Table I assumes coarse symbol synchronization, such that the beginning of the vector $\mathbf{r}(n)$ for symbol n is aligned with the direct path. In order to obtain symbol sync, a training sequence comprising $2MN_s$ samples is first transmitted using M known Walsh symbols with $\mathsf{m}(n) = 1$ for $n = 1, \ldots, M$. Recall that $2N_s = 2T/T_s$ is the length in samples of the symbol plus time-guard interval. The synchronization algorithm first estimates the time-of-arrival of the direct path, and then advances the symbol clock appropriately.

Under the above assumptions, the received signal during the training period is approximated via a tapped-delay line multipath channel as

$$r(t) \approx \sum_{l=0}^{2N_s-1} f_l s_1(t - lT_s) + n(t). \qquad (16)$$

The sparse channel with coefficients f_l corresponds to unambiguous delays ranging from 0 to $(2N_s - 1)T_s$ s. The acquisition problem is then

1) Estimate the $\{f_l\}$ under a numerosity constraint $|\mathbf{f}| = N_f << 2N_s$. Here, the channel vector is twice the length as in GMHT-MP, specifically $\mathbf{f} = [f_0 f_2 \ldots f_{2N_s-1}]^T$ and $|\mathbf{f}|$ is the number of nonzero elements in $\mathbf{f}$.
2) Find the index of the direct path l_d in $\mathbf{f}$ such that the multipath channel is defined by $[f_{l_d}, f_{l_d+1\bmod 2N_s} \cdots f_{l_d+N_s-1\bmod 2N_s}]$. That is, if the direct-path delay is $2N_s - 1$ for example, and the next multipath coefficient is nonzero, the next coefficient will have associated delay 0, due to the $2N_s$ delay ambiguity.

3) Collect vectors $\mathbf{r}(n)$ for subsequent demodulation with a timing advance of l_d samples. This aligns the onset of each direct-path waveform $s_1(t)$ with the beginning of the $2N_s$ sample window defined by $\mathbf{r}(n)$.

V. Packet-Based Matching Pursuits

The packet-based MP algorithm is defined somewhat differently from the GMHT-MP algorithm. First, since a training sequence is transmitted only one MP algorithm for $\mathsf{m} = 1$ needs to be implemented. Second, a vector of length $2MN_s$, rather than N_s samples with $M > 1$ is collected. The packet length M is selected to achieve accurate initial channel estimates, but short enough so that the Doppler variation is negligible in $M2T_{sym}$ sec. Define the n-th received sample vector $\mathbf{r}(n)$ as

$$\mathbf{r}(n) = \sum_{l=0}^{2N_s-1} f_l \mathbf{s}_1(l) + \mathbf{n}(n), \tag{17}$$

where with $r(n) = r_c(nT_s)$ and $r_c(t)$ the continuous-time received waveform, we have

$$\mathbf{r}(n) = [r(2(n+1)N_s-1) r(2(n+1)N_s-2) \dots r(2nN_s)]^T. \tag{18}$$

For training sequence transmission, the signal vector corresponds to

$$\mathbf{s}_1(l) = [s_1(2N_s - 1 - l) s_1(2N_s - 2 - l) \dots s_1(-l)]^T. \tag{19}$$

However, unlike the GMHT-MP receiver, continual training sequence transmission results in a circular form for the vector $\mathbf{s}_1(l)$. Specifically, the k-th component $s_1(k)$ is given by

$$s_1(k) = \sum_{q=0}^{1} s_1^c(kT_s + q2N_sT_s), \tag{20}$$

where

$$s_1^c(t) = \sum_{i=0}^{N_w L_{pn}-1} (\mathbf{d}_1)_i g(t - iT_c), \tag{21}$$

where $\mathbf{d}_1 = \mathbf{b}_1 \otimes \mathbf{c}$ is the $N_w = 8, L_{pn} = 7$ or 56 chip long Walsh/m-sequence corresponding to training symbol $\mathsf{m} = 1$. Thus, the sequence $s_1(k)$ is a *periodic* version of the $2T_{sym}$ long symbol plus guard interval with period $2N_s$. Hence $\mathbf{s}_1(l)$ in (19) is the l-th circular shift of the $2N_s$ sample long Walsh/m-sequence waveform.

Next define a packet-length vector $\mathbf{r} \in \mathbb{C}^{2MN_s}$ as

$$\mathbf{r} = [\mathbf{r}(M-1)^T \mathbf{r}(M-2)^T \dots \mathbf{r}(0)]^T, \tag{22}$$

where $\mathbf{r}(n) \in \mathbb{C}^{2N_s}$ is given by (18). Also define $\mathbf{S} \in \mathbb{C}^{2MN_s \times 2N_s}$ with i-th column $\mathbf{S}_i = \mathbf{1}_M \otimes \mathbf{s}_1(i-1)$, where $\mathbf{1}_M$ is the length M all-ones vector and $\otimes$ is the Kronecker product. Then the i-th column of $\mathbf{S}$ is the i-th circular shift of the transmitted waveform over $2MN_s$ samples. The received packet is then

$$\mathbf{r} = \mathbf{S}\mathbf{f} + \mathbf{n}. \tag{23}$$

Using $\mathbf{r}$, the Matching Pursuits algorithm yields an estimate $\hat{\mathbf{f}}$ of the channel coefficients. Given $\hat{\mathbf{f}}$, we have to find the best-fit multipath channel such that the N_f nonzero coefficients occupy at most N_s contiguous samples, modulo $2N_s$ starting at delay l_d. The following test for l_d is proposed, under the assumption that the maximum multipath spread is N_s samples, and hence at least N_s contiguous coefficients (modulo N_s) $\hat{f}_l$ must be zero.

$$l_d = \arg\max_l \frac{\sum_{i=l}^{N_s+l-1} |\hat{f}_{i \bmod 2N_s}|^2}{\sum_{i=N_s+l}^{2N_s+l-1} |\hat{f}_{i \bmod 2N_s}|^2}. \tag{24}$$

That is, we try to find the direct path delay which corresponds to the maximum of the ratio of the energy in the hypothesized multipath profile to the energy in the hypothesized set of N_s zero coefficients $\hat{f}_l$.

The overall synchronization/channel estimation algorithm is summarized in Table II.

Precompute $\mathbf{A} = \mathbf{S}^H\mathbf{S} \in \mathbb{C}^{2N_s \times 2N_s}$
Collect received packet $\mathbf{r} \in \mathbb{C}^{2MN_s}$ over $2MN_s$ samples
$\quad \mathbf{r} = [\mathbf{r}(M-1)^T \mathbf{r}(M-2) \dots \mathbf{r}(0)]^T$
Initialize sufficient statistic vector
$\quad \mathbf{v}^1 = \mathbf{S}^H\mathbf{r} \in \mathbb{C}^{2N_s}$
For $k = 1, 2, \dots, N_f$
$\quad q_k = \arg\max_{i \neq q_1,\dots,q_{k-1}} |(\mathbf{v}^k)_i|^2/(\mathbf{A})_{i,i}$
$\quad \hat{f}_{q_k} = (\mathbf{v})_{q_k}/(\mathbf{A})_{q_k,q_k}$
$\quad \mathbf{v}^{k+1} = \mathbf{v}^k - \mathbf{A}(:, q_k)\hat{f}_{q_k}$
Next k
Find direct-path index
$\quad l_d = \arg\max_l \frac{\sum_{i=l}^{N_s+l-1} |\hat{f}_{i \bmod 2N_s}|^2}{\sum_{i=N_s+l}^{2N_s+l-1} |\hat{f}_{i \bmod 2N_s}|^2}$
Advance subsequent timing for demodulation by l_d samples

TABLE II

MP Algorithm for Initial Timing/Channel Estimation

VI. Results

The GMHT-MP algorithm was simulated using independent Rayleigh fading channels $\alpha_p(n)$ from symbol-to-symbol. The delays $\tau_p(n)$ were generated as independent random variables uniform on $[0, T_{sym}]$ on each epoch n.

An example of instantaneous channel estimation using one symbol epoch of data, $\mathbf{r}(n)$ is shown in Fig. 2. The true channel multipath intensity profile (MIP) is plotted along with the estimated channel MIP using $\hat{\mathbf{f}}$. In this case, there are $N_\alpha = 6$ actual paths, and Matching Pursuits assumes $N_f = 8$ paths are present. The Matching Pursuits channel estimator clearly identifies the major paths using just one symbol epoch of data.

The SERs for the GMHT-MP and RAKE receivers are plotted in Fig. 3 for randomly generated channels on each symbol. Note that the RAKE detector corresponds to GMHT-MP in Table I with the cancellation step removed, i.e. with $\mathbf{v}_\mathsf{m}^k = \mathbf{S}_\mathsf{m}^H \mathbf{r}(n)$ at each iteration. Again, the α_p were chosen as i.i.d. zero-mean circular Gaussian r.v.s (Rayleigh fading), such that the channel energy gain was normalized to unity. The noncoherent combining loss and fading impacts the SER so that an E_b/N_0 of 20 dB is required to obtain a 10^{-3} SER. It is seen that the GMHT-MP receiver significantly outperforms the RAKE at usable SERs below 10^{-2}. This result is not surprising, since as SNR increases, the GMHT-MP channel estimates become increasingly accurate, and the receiver approaches the coherent ML detector. In contrast, the RAKE performance is limited to that of a noncoherent detector. Finally, observe that the GMHT-MP performance is not adversely affected by channel overparameterization (e.g. $N_\alpha = 32$ with $N_f = 48$ assumed.

An example run from the synchronization algorithm in Table II is shown in Fig. 5. The estimated MIP closely follows the true MIP, and the algorithm correctly detects the time-of-arrival of the direct path with delay l_d samples.

VII. Hardware Implementation

The GMHT-MP modem is currently being implemented at UCSB using the TI F2812 fixed-point DSP device with custom amplifiers, matching networks and transducers. The GMHT-MP real-time software has been successfully tested in the lab and verified using Code Composer. The F2812 board with amplifier/matching module is shown in Fig. 4.

In the receive mode, the signal from the 25 kHz center frequency transducer is amplified and filtered, with a large gain-adjust range in the amplifiers (84 dB) allowing adaptation to a wide range of conditions. The filtered signal is applied to the 12-bit ADC which is a module

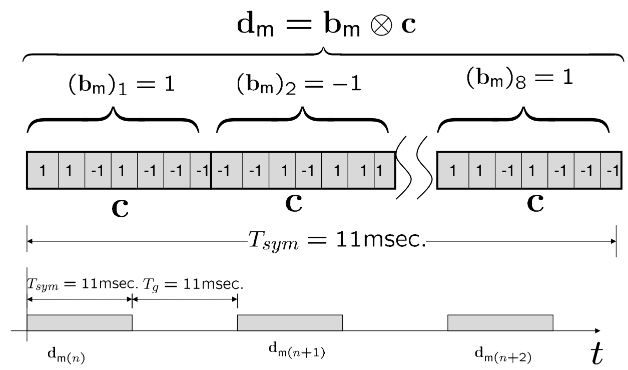

Fig. 1. Walsh/m-sequence signals.

on the DSP chip. In the transmit mode, the digital signal generated by the DSP is applied to a pair of DACs, one of which is dedicated to scaling the transmitted signal (for power control). The DAC output passes through a filter network identical to the receiver filter; this filter removes DAC harmonic artifacts. A Maxim MAX9703 class D power amplifier amplifies the signal to a maximum of 12 watts RMS, and a network consisting of a transformer and inductor cancel most of the transducer capacitance and approximately match the transducer impedance to the amplifier's design load. The preamp input remains coupled to the transducer during transmit but the power flow into the preamp is limited to prevent damage.

VIII. Conclusions

A design for an UWA modem was presented based on Walsh/m-sequence signaling. The proposed GMHT-MP receiver significantly outperforms a RAKE detector at usable SNRs. The Matching Pursuits algorithm in particular is effective in estimating long, sparse underwater acoustic multipath channels, and is fairly robust to overparamterization.

Hardware implementation of the modem is underway in programmable DSP, and sea testing is anticipated by Fall 2005. Future research on the modem will include optimization of the Walsh signal dimension and spreading to increase data rates and obtain further robustness to frequency-selective multipath. Implementation in reconfigurable hardware will also be investigated as an alternative to DSP for lower power and accommodation of larger acoustic bandwidths and data rates.

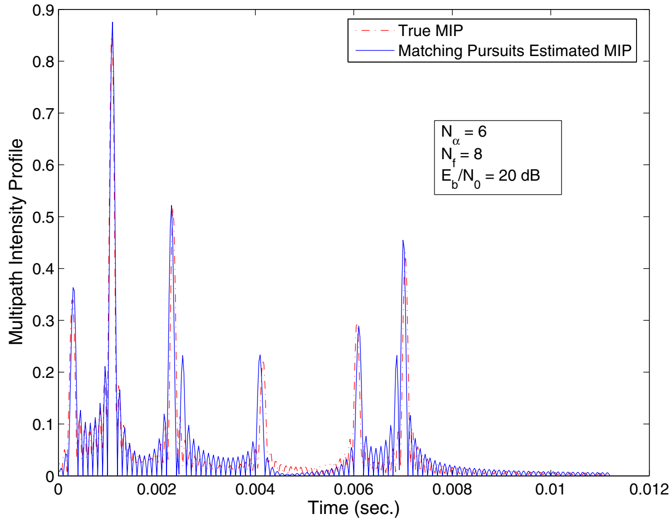

Fig. 2. Channel estimates at $E_b/N_0 = 20$ dB, $N_\alpha = 6, N_f = 8$

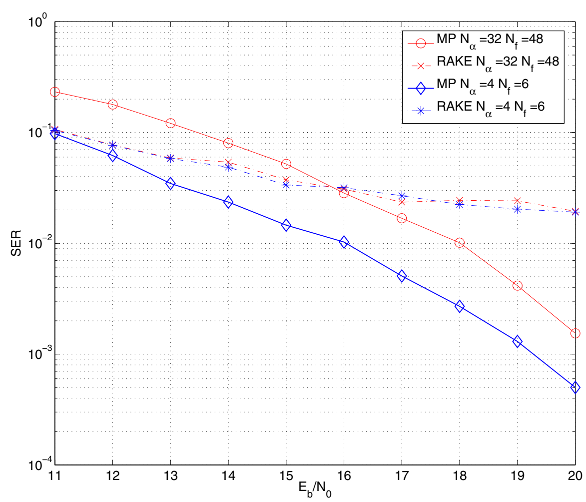

Fig. 3. SERs for $N_\alpha = 6, N_f = 8$ and $N_\alpha = 32, N_f = 48$.

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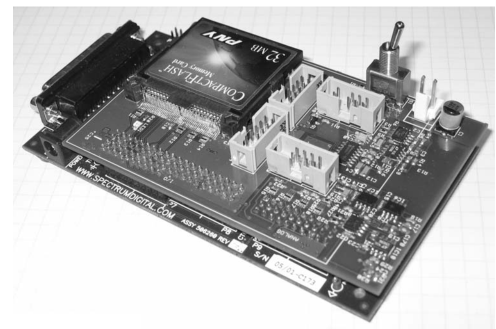
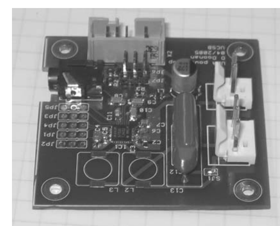


(a)



(b)

Fig. 4. (a) F2812 DSP with ADC/DAC. (b) Power amplifier

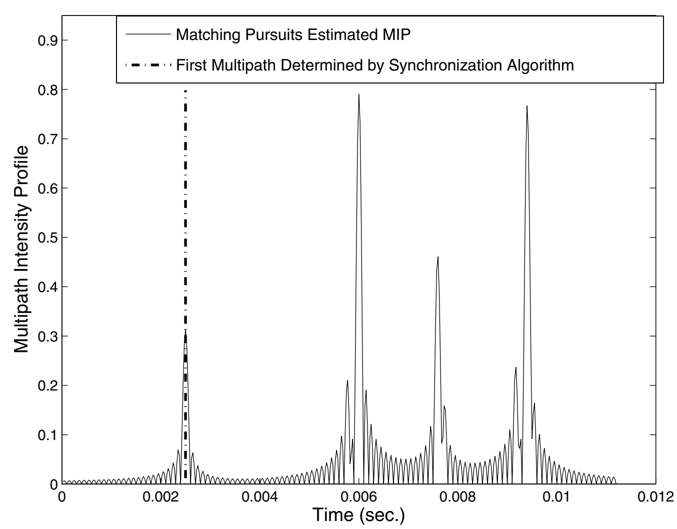


Fig. 5. Estimate of l_d direct-path delay from synchronization algorithm

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