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Iterative Equalization and Decoding Scheme for Underwater Acoustic Coherent Communications

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1. Introduction

Digital communications through underwater acoustic (UWA) channels differ from those in other media, such as radio channels, due to the high temporal and spatial variability of the acoustic channel which make the available bandwidth of the channel limited and dependent on both range and frequency. In order to overcome disadvantage factors and maximize performance to conduct real-time information understanding, underwater acoustic communications require the higher degree of information extraction and development from all kinds of onboard acoustic sensors and processing systems. A higher performance communication technology is needed in order to focus high-performance data processing on the problems and tasks faced by human operators and decision-makers. In order to establiseh reliable data commnication on the severely band-limited underwater acoustic channels, bandwidth-efficient modulation techniques (i.e. coherent communications) should be employed to overcome the inter-symbol interference (ISI) caused by channel multi-path propagation. The effective approach to eliminate the ISI caused by multipath propagation is that adaptive decision feedback equalizer (ADFE) integrates with spatial diversity. That is multi-channel adaptive decision feedback equalizer, which is applied in (Kilfoyle & Baggeroer, 2000; Stojanovic, 1996, 2005; Zhao et al., 2008) represents a more general approach to spatial and temporal signal processing.

However, Single technique, such as equalizer, is difficult to obtain satisfied data transmission because of the complexity of UWA channel, especially in shallow water channel. In recent years, more and more attention has been paid to Turbo codes, including parallel concatenated convolutional code (PCCC) (Berrou et al., 1993) and serially concatenated convolutional code (SCCC) (Benedetto & Montorsi, 1996), because of its near-capacity gains. The range of applications of Turbo codes has expanded to many areas of communications. Trellis-Coded Modulation (TCM) (Ungerboeck, 1982) is a kind of design option combining coding with modulation. It can provide over 3 dB coding gain without bandwidth expansion. Especially, it is interesting to combine PCCC or SCCC with TCM in order to improve the transmission spectral efficiency (i.e. parallel concatenated trellis codes modulation (PCTCM) (Benedetto et al., 1996; Chung & Lou, 2000; Legoff et al., 1994; Yang & Ge, 2005) and serial concatenated trellis codes modulation (SCTCM)) (Benedetto et,al., 1997; Divsalar & Pollara, 1997; Ho, 1997; Shohon et al., 2003) in order to improve the transmission spectral efficiency.

Therefore, iterative equalization and decoding (IED) based on equalizer and deocoding has been developed to obtain higher performance data transmission. Turbo equalizer (Berthet, 2000; Koetter et al., 2004) treats the channel encoder and channel itself as a serial concatenated system that can be decoded in an iterative scheme. A drawback of this iterative receiver is that the complexity of the turbo equalizer is orders of magnitude greater than the decision feedback equalizer (DFE). The turbo equalizer complexity grows exponentially with channel memory length. It isn't suit for the underwater acoustic channel with long delay spreads. In the structure of iterative euqalization and decoding, equalizer can use ADFE (Choi, 2008; Noorbakhsh et al., 2003) or equalizer based on channel estimation (Flanagan & Fagan, 2007; Otnes & Tuchler, 2004; Tuchler et al., 2002). For underwater acoustic channel with severely mutipath propagation and large time delay, adptive channel tracking can get better performance than channel estimation using trainning sequence or pilot symbols.

Comparing PCTCM, SCTCM has the following advantages: (1) It can further reduce the error floor of PCTCM to obtain lower BER (Soleymani & Gao, 2002). (2) It has more flexible coding structure than PCTCM. In this chapter, SCTCM technique is adopted to increase bandwidth efficiency. Furthurmore, a rate R=1 recursive convolutional is adopted as inner encoder of SCTCM encode to get higher performance SCTCM scheme. Therefore, iterative equalization and decoding, based on multi-channel adaptive decision feedback equalizer with variant step tracking factor and decoder of SCTCM, is formed to aid weight update of equalizer utilizing decoding gain provided by decoder of SCTCM such that the performance of equalizer is enhanced. And then, the performance of communication system is improved greatly through iteration calculation between equalizer and decoder.

The structure of this chapter is as follows. Firstly an overview of the channel and system model is provided in Section 2. More specifically, the channel model based on sound speed profile (SSP) measured in the lake and Bellhop method and the system description are discussed in Sections 2.1 and 2.2, respectively. And then, the introduction of the proposed iterative equalization and decoding is presented in Section 3. More specifically, the structures of iterative equalization and decoding are discussed in Section 3.1. The proposed iterative equalization and decoding process is detailed in Section 3.2, commencing with a discussion of the multichannel adaptive equalizer structure in Section 3.2.1, followed by a description of the SCTCM decoding algorithm in Section 3.2.2, the method of the soft symbol estimation in Section 3.2.3. Our simulation results are provided in Section 4, while Section 5 concludes our findings.

2. Channel and system models

We begin with the channel model under consideration and then discuss the description for communication systems.

2.1 Underwater acoustic channel model

We adopt real measured data, sound speed profile (SSP), and a finite-element ray (FER) tracing method (Bellhop) (Porter & Liu, 1994) to model the underwater acoustic multipath propagation. Additionally we model the multipath components as fading due to acoustic propagation loss.

A given multipath arrival l is characterized by its magnitude gain γ_l and delay τ_l. These quantities are dependent on the ray length l_l, which in turn is a function of the given propagation range R. The path magnitude gain is given by

$$\gamma_l = \frac{\Gamma_l}{\sqrt{\beta(l_l)}} \tag{1}$$

where, Γ_l is the amount of loss due to reflection at the bottom and surface. The acoustic propagation loss, represented by $\beta(l_l)$

$$\beta(l_l) = l_l^{\alpha}\left[a(f_c)\right]^{l_l} \tag{2}$$

where, α is constant, f_c is the carrier frequency and absorption coefficient (in db/km) given by Thorp's formula

$$\begin{aligned} &10\log a(f_c) \\ &= \frac{0.11 f_c^2}{1+f_c^2} + \frac{44 f_c^2}{4100+f_c^2} + 2.75\times 10^{-4} f_c^2 + 0.003 \end{aligned} \tag{3}$$

The path delay is given by

$$t_l = \sum_{i=1} \frac{l_{l,i}}{v_{l,i}} \tag{4}$$

where, $v_{l,i}$ is the sound speed of the ith water layer according to SSP. Thus, the overall channel impulse response is given by

$$h(t) = \sum_{l=1}^{L} A_l \delta(t-\tau_l) \tag{5}$$

where, L is the multipath number, A_l and τ_l are the amplitude and relative delay of the lth multipath arrival respectively. $\tau_l = t_l - t_{\min}$, $t_{\min}$ is the minimus delay among the all path delays.

In the simulation section (Section 4), the SSP, measured on the lake (shown in Fig. 1), is adopted to model the multipath propagation. The SSP denotes the sound speed is changed with water depth.

Additionally, Doppler frequency f_d is considered in the channel model. It is given by

$$f_d = f_c \frac{v_r}{c} \cos\theta_l \tag{6}$$

Where, c donotes underwater sound speed, v_r is the relative speed between tranmiter and receiver, θ_l is the arrival angle for the lth arrival ray.

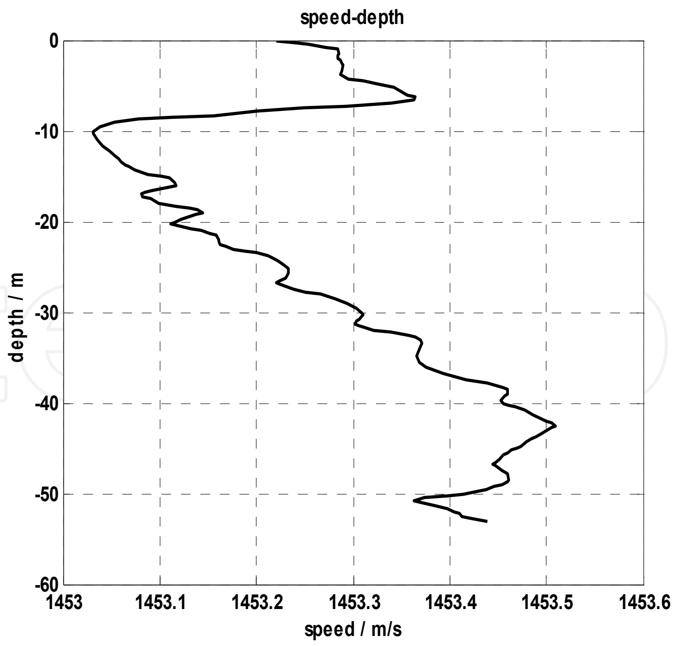


Fig. 1. Sound speed profile (SSP)

2.2 System description

The structure of transmitter is shown in Fig.2.

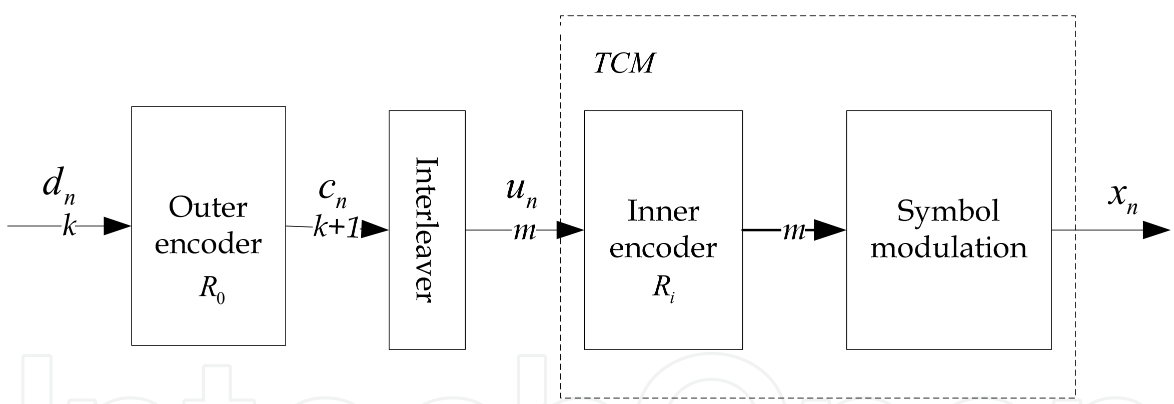


Fig. 2. The structure of transmitter

In comparison with high performance PCTCM scheme (Robertson & Woerz, 1998), the method in (Divsalar et al., 2000), with lower complexity, is adopted to design SCTCM, which can achieves $km/(k+1)\ bit/s/Hz$, using a rate $R_0 = k/k+1$ convolutional encoder with maximum free hamming distance as the outer code. An interleaver permutes the output of the outer code. The interleaved data enters a rate $R_i = m/m = 1$ recursive convolutional inner encoder. The m output bits are then mapped to one symbol belonging to a 2^m level modulation. In our system, the data symbol is QPSK modulated ($m = 2$), i.e. $x_n = \pm\frac{\sqrt{2}}{2} \pm \frac{\sqrt{2}}{2}$ with probability $\frac{1}{4}$. Before data symbol, the pilot symbol is transmitted to probe the channel impulse response (CIR). The LFM signal is used in our system. The frame structure is shown in Fig.3.

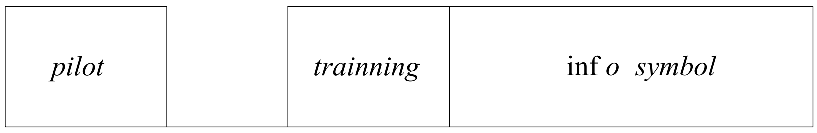


Fig. 3. Data frame structure

In receiver, Spatial diversity is achieved via multiple receiver arrays. The received signal at the kth array is given by

$$r_k(t) = \sum_{l=1}^{L} A_l x(t - \tau_l) e^{j2\pi(1+\Delta_l) f_c t} + n(t) \tag{7}$$

where, $\Delta_l = \frac{v_r}{c}\cos\theta_l$ is the Doppler frequency factor for lth multipah propagation, $n(t)$ is assumed to be a white Gaussian process with zero mean and variance σ_n^2.

After demodulation by multiplying the local carrier frequency and Doppler frequency compensation, the received baseband signal at the kth array is given by

$$y_k(t) = \sum_{l=1}^{L} A_l x(t - \tau_l) e^{j\theta_k} + n(t) \tag{8}$$

where, θ_k is the remain phase distoration.

And then, the iterative equalization and decoding (IED) is performed on the received multichannel baseband signals. The concrete IED algorithm will be presented in the next section.

3. Iterative equalization and decoding (IED)

In this section, we first present the structures of iterative equalization and decoding and analyze the merits and drawbacks of different structures in Section 3.1. And then, the proposed iterative receiver is detailed in Section 3.2.

3.1 The structures of iterative equalization and decoding

Since the underwater acoustic ISI channel can be treated as a convolutional encoder with rate 1, it is possible to treat the channel encoder and channel itself as a serial concatenated system that can be decoded in an iterative scheme such as the turbo equalizer structure that is illustrated in Fig.5. The motivation for the study of this receiver algorithm is to improve equalizer performance beyond that attainable by the optimum parameters decision feedback equalizer which also employs the all-training sequence.

The turbo equalizer consists of two soft input soft output (SISO) modules for the channel equalizer and the decoder that are arranged in a serial fashion. A drawback of this receiver algorithm is that the complexity of the turbo equalizer is orders of magnitude greater than the DFE. The turbo equalizer complexity grows exponentially with channel memory length, modulation level, and spatial diversity combining. It should be noted that traditionally the turbo equalizer has been used for known channels with reasonable ISI. It still needs to be demonstrated that this type of receiver can be used with modification to track the time-varying underwater channel and provide performance that exceeds the performance of the DFE using known training sequence throughout the entire data packet.

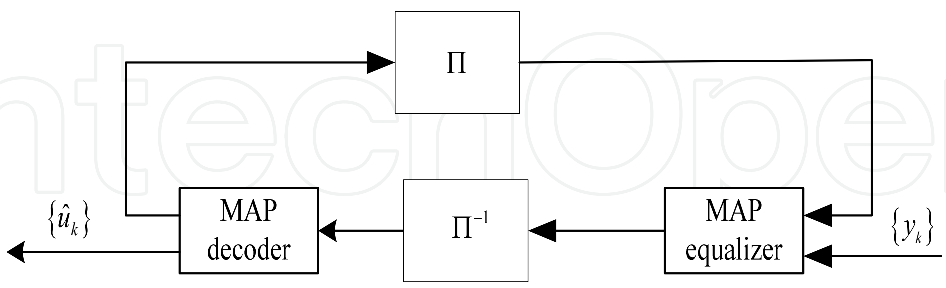


Fig. 4. Turbo equalizer

The second structure is hard iterative (shown in Fig.5.)

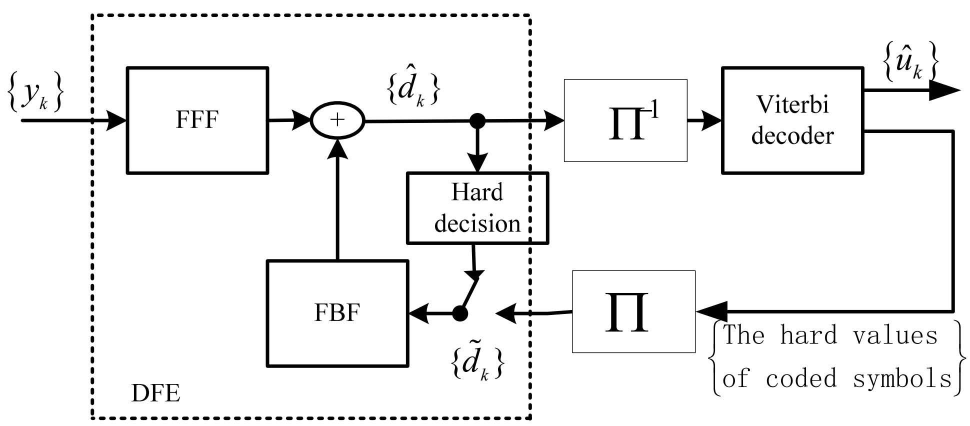


Fig. 5. Hard iteration

In this structure, the decoded symbols or hard decisions from the decoder after a first pass through the receiver system are then re-encoded to be used as the new training sequence to be used over the entire received data packet. Therefore, in the second pass or iteration, decision directed equalization after the short initial training sequence is not employed. Subsequent passes or iterations through the data in this fashion can be made. However diminishing improvements are obtained due to the hard decision nature of this algorithm. Feedback error propagation can still occur in this algorithm due to uncorrected errors at the output of the soft input hard output decoder. The desire for performance improvement by using SISO decoders as well as added information that the equalizer can provide to correct decoder errors provides the motivation for the soft iterative approach.

An improved receiver algorithm as compared to the hard iterative approach would be to employ all the information regarding the received symbols to generate the new training sequence by combining soft values of the coded symbols out of the decoder and the soft information about the detected symbols provided by the decision directed mode of the equalizer. This is intent of the soft iterative manner.

In the soft iteration, the decision feedback equalizer is modified so that it can use soft a priori information from the decoder from previous iterations. In order to obtain soft a priori information as opposed to hard decisions, the decoder structure must now take the form of a SISO device. One such device is the Maximum A posteriori (MAP) decoder. The method in which the information streams from the decoder and equalizer are combined is crucial because log likelihood ratio (LLR) values are produced from the decoder feedback path. In this chapter, the soft iteration structure is adopted.

3.2 Soft iterative equalization and decoding (IED)

The structure of the proposed IED with phase compensation is shown in Fig.6.

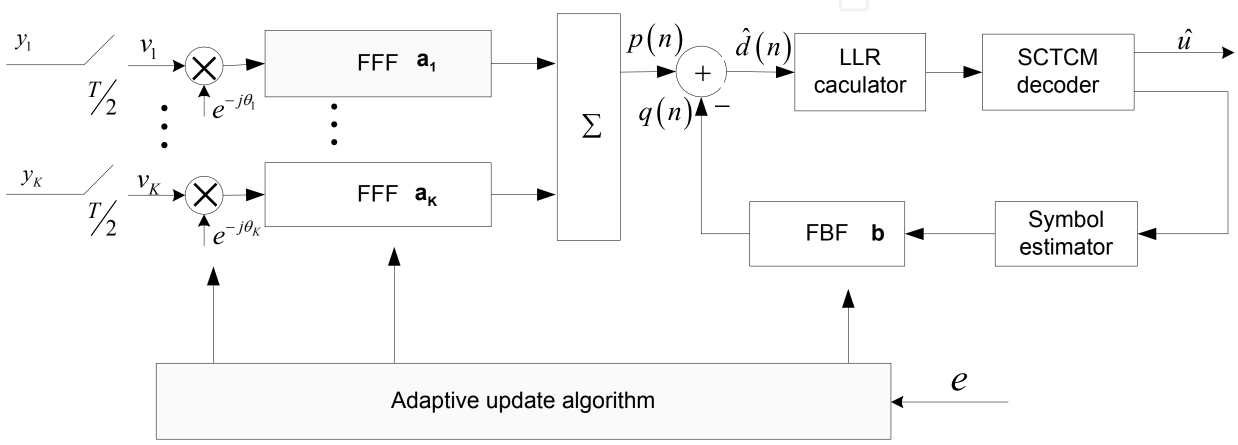


Fig. 6. The proposed IED

As shown in Fig.6, the received signal at each array elements is T/2 fractional sampled. The digital phase lock loop (DPLL) is adopted to correct phase distortion. Then, the feedforward and feedback filters are applied to obtain the estimation of transmitted symbol. In the IED scheme, the multichannel adaptive decision feedback qualizaer with phase compensation and decoder of SCTCM exchange soft information in an iterative manner. Specifically, at the output of the equalizer, the likelihood ratio (LLR) calculator computes soft information of coded bits based on the symbol estimation $\hat{d}(n)$. This soft information is delivered to the maximum a posteriori (MAP) decoder of SCTCM. In addition to providing the decoded output, the decoder also computes soft information on the coded bits, which is converted to soft estimates of the symbols. These soft symbol estimates are used to aid the operation of the equalizer and its adaptive weight update algorithm.

3.2.1 Mutichannel adaptive equalization

The received signal will carry out equalization proccessing (including carrier phase compensation) after demodulation. The main task of multichannel adaptive decision feedback qualizaer is eliminate inter-symbol interference (ISI) caused by multipath propagation.

According to minimum mean square error (MMSE) scheme, an error signal is used to update receiver parameters. The error signal $e(n)$ of adaptive update algorithm as follows

$$e(n) = d(n) - \hat{d}(n) \tag{9}$$

$$\begin{aligned}\hat{d}(n) &= p(n) - b^H \tilde{\mathrm{d}}(n) \\ &= \begin{bmatrix} a_1^H & \cdots & a_L^H & -b^H \end{bmatrix} \begin{bmatrix} v_1(n)\mathrm{e}^{-\mathrm{j}\theta_1} \\ \vdots \\ v_L(n)\mathrm{e}^{-\mathrm{j}\theta_L} \\ \tilde{\mathrm{d}}(n) \end{bmatrix} = \mathrm{w}^H \mathrm{u}(n)\end{aligned} \tag{10}$$

Where, $\mathrm{w}(n)$=$[\mathrm{a}(n)\ \text{-}\mathrm{b}(n)]^{\mathrm{T}}$ denotes overall equalizer coefficient vector, $\mathrm{u}(n) = [\mathrm{x}(n)\ \tilde{\mathrm{d}}(n)]^T$ denotes composite input vector, $\mathrm{a}_k^H = [a_0^k \cdots a_{N-1}^k]^*$ denotes the coefficients vector of feedforward filter, $\mathrm{b}^H = [b_1 \cdots b_M]^*$ denotes the coefficients vector of feedback filter, *N* and *M* denote the feedforward and feedback filter taps respectively, * denotes complex conjugate, $^{\mathrm{H}}$ denotes transpose conjugate, $^{\mathrm{T}}$ denotes transpose. $\tilde{\mathrm{d}}(n) = \left[\tilde{d}(n-1) \cdots \tilde{d}(n-M)\right]^T$ denotes the vector of M previously detected symbols stored in the feedback filter, $p(n)$ represents the output of the linear part of the equalizer, it can be written as

$$p(n) = \sum_{k=1}^{K} a_k^H v_k(n) e^{-\mathrm{j}\theta_k} = \sum_{k=1}^{K} p_k(n) \tag{11}$$

As shown in Fig.6., the baseband signals firstly perform carrier phase compensation. Using DPLL technology (Proakis, 2003), we can obtain carrier phase compensation as follows

$$\hat{\theta}_{k+1}(n) = \theta_k(n) + K_{f_1}\phi_k(n) + K_{f_2}\sum_{m=1}^{n}\phi_k(m) \tag{12}$$

Where, $\Phi_k(n+1) = \mathrm{Im}\left\{p_k(n)\left[p_k(n) + e(n)\right]^*\right\}$, K_{f_1} and K_{f_2} are constants, $K_{f_2} \le K_{f_1}$.

A fast self-optimized LMS (FOLMS) algorithm (Bragard & Jourdain, 1990) is used to update the equalizer vector $\mathrm{w}(n)$. But in (Bragard & Jourdain, 1990), the formulations are conducted based on the single channel line equalizer (LE). In this chapter, we extent it to multi-channel decision feedback equalizer and consider the effect of carrier phase compensation. It can be deducted by the composite input data $\mathrm{u}(n)$ and the error signal $e(n)$. So, we can rewrite the FOLMS algorithm as follows

$$\mathbf{w}(n+1) = \mathbf{w}(n) + \mu(n)\mathbf{x}(n)e^*(n) \tag{13}$$

$$\mu(n+1) = \mu(n) + \beta \mathrm{Re}\left[\mathbf{G}^{\mathrm{H}}(n)\mathbf{x}(n)e^*(n)\right] \tag{14}$$

$$g(n) = \mathbf{x}^{\mathrm{H}}(n)\mathbf{G}(n) \tag{15}$$

$$x^{'}(n) = e^*(n) / \mu(n) \tag{16}$$

$$\xi(n) = x'(n) - g(n) \tag{17}$$

$$\mathbf{G}(n+1) = \mathbf{G}(n) + \mu(n)\mathbf{x}(n)\xi(n) \tag{18}$$

Where, $\mu(n)$ is the step-size factor for controlling the convergence ratio of the equalizer, which can adaptively update, $g(n)$ is temporary variant for updating $\mu(n)$, α is constant. $\mathrm{Re}(\cdot)$ denotes the real part of data.

3.2.2 Decoding of SCTCM

In order to simplify the decoding algorithm of SCTCM, the symbol decoding of SCTCM is transformed into bit decoding through calculating the LLR of coded bits. The position of LLR calculator is shown in Fig.6. The calculation is detailed as follows.

For MPSK, the corresponding $m = \log_2 M$ coded bits are mapped to an M-ary signal. The probability $p(b_i = 1 \mid y_k)$ of ith coded bit of kth received symbol can be calculated as

$$\begin{aligned} p(b_i = 1 \mid y_k) &= \frac{p(y_k \mid b_i = 1) \cdot p(b_i = 1)}{p(y_k)} \\ &= \frac{1}{p(y_k)} \left\{ \sum_{b_1} \sum_{b_2} \cdots \sum_{b_m} p(y_k \mid b_i = 1, b_1, ..., b_m) \cdot p(b_i = 1, b_1, ..., b_m) \right\} \end{aligned} \tag{19}$$

Let the probabilities $p(b_i = 1)$ and $p(b_i = 0)$ of coded bit b_i are the same. Therefore:

$$p(b_i = 1, b_1, ..., b_m) = p(b_1 = 1) \cdot p(b_2) \cdots p(b_m) = \frac{1}{2^m} \tag{20}$$

So, (19) can be simplified as

$$p(b_i = 1 \mid y_k) = \frac{1}{2^m \cdot p(y_k)} \sum_{b_1} \sum_{b_2} \cdots \sum_{b_m} p(y_k \mid b_i = 1, b_1, ..., b_m) \tag{21}$$

The probability $p(b_i = 0 \mid y_k)$ of ith coded bit of kth received symbol can be calculated as

$$p(b_i = 0 \mid y_k) = \frac{1}{2^m \cdot p(y_k)} \sum_{b_1} \sum_{b_2} \cdots \sum_{b_m} p(y_k \mid b_i = 0, b_1, ..., b_m) \tag{22}$$

The LLR value of ith coded bit of kth received symbol is

$$\Lambda_i = \ln \frac{p(b_i = 1 \mid y_k)}{p(b_i = 0 \mid y_k)} = \ln \frac{\sum_{b_1} \sum_{b_2} \cdots \sum_{b_m} p(y_k \mid b_i = 1, b_1, ..., b_m)}{\sum_{b_1} \sum_{b_2} \cdots \sum_{b_m} p(y_k \mid b_i = 0, b_1, ..., b_m)} \tag{23}$$

From the received signal $y_k = d_k + n_k$,and from the noise distribution it follows that $p(y_k \mid d_k)$ is given by

$$p\left(y_k \mid d_k\right) = \exp\left(-\left(y_k - d_k\right)^2 / \left(2\sigma^2\right)\right) \Big/ \sqrt{2\pi\sigma^2} \tag{24}$$

So, the (23) can be calculated as

$$\Lambda_{k,i} = \ln \frac{\sum_{d \in \mathrm{B}_{i=1}} \exp\left(-\frac{\left(\hat{d}_k - d\right)^2}{2\sigma^2}\right)}{\sum_{d \in \mathrm{B}_{i=0}} \exp\left(-\frac{\left(\hat{d}_k - d\right)^2}{2\sigma^2}\right)} \tag{25}$$

where, $\mathrm{B} = \{d_1, d_2, \cdots, d_M\}$ denotes the finite alphabet used for MPSK signals, $\mathrm{B}_{i=1}$ and $B_{i=0}$ denote the sets of all possible symbol values, in which the ith coded bit is 1 and 0 respectively.

We can simplify symbol decoding into bit decoding using Eq.(25). The probability distributions of the output sequences $\tilde{P}_k^O(u)$ and $\tilde{P}_k^O(c)$ can be calculated as follows:

$$\tilde{P}_k^O(u) = \tilde{B}_u \sum_{e:u(e)=u} \alpha_{k-1}\left[s^S(e)\right] P_k^I\left[u(e)\right] P_k^I\left[c(e)\right] \beta_k\left[s^E(e)\right] \tag{26}$$

$$\tilde{P}_k^O(c) = \tilde{B}_c \sum_{e:c(e)=c} \alpha_{k-1}\left[s^S(e)\right] P_k^I\left[u(e)\right] P_k^I\left[c(e)\right] \beta_k\left[s^E(e)\right] \tag{27}$$

where, $\tilde{B}_u$ and $\tilde{B}_c$ are normalization constants as

$$\tilde{B}_u \to \sum_u P_k^O(u) = 1 \tag{28}$$

$$\tilde{B}_c \to \sum_c P_k^O(c) = 1 \tag{29}$$

The forward recursion $\alpha_k[\cdot]$ and backward recursion $\beta_k[\cdot]$ are given by

$$\alpha_k(s) = \sum_{e:s^E(e)=s} \alpha_{k-1}\left[s^S(e)\right] P_k^I\left[u(e)\right] P_k^I\left[c(e)\right] \ , \ k = 1,2,\ldots,n \tag{30}$$

$$\beta_k(s) = \sum_{e:s^S(e)=s} \beta_{k+1}\left[s^S(e)\right] P_{k+1}^I\left[u(e)\right] P_{k+1}^I\left[c(e)\right] \ , \ k = n-1,\ldots,0 \tag{31}$$

The $\alpha_k[\cdot]$ computation will be initialized as

$$\alpha_0(s) = \begin{cases} 1, \, s = S_0 \\ 0, \, other \end{cases} \tag{32}$$

If the trellis is terminated to a known state S_N, then the $\beta_k[\cdot]$ computation will be initialized as

$$\beta_n(s) = \begin{cases} 1, \ s = S_N \\ 0, \ other \end{cases} \tag{33}$$

Otherwise

$$\beta_n(s) = 1 / M_s \ , \ \forall s \tag{34}$$

In this chapter, log-map algorithm (Soleymani & Gao, 2002) is used to simplify calculation through transforming multiplication into addition.

3.2.3 Soft symbol estimation

As shown in Fig.6, the LLR values of coded bits, output from decoder of SCTCM, are used to implement symbol estimation. And then, these symbols are fed back to feedback filter of MC-ADFE to perform joint iterative scheme. So, the symbol estimation is key module to perform soft IED.

There are two methods to estimate data symbol: hard estimation and soft estimation. Compare with hard estimation, soft estimation can void error symbols spread during the course of iterations. What's more, soft estimation can more sufficiently utilize decoding gain to update system performance. In this chapter, soft method is adopted to estimate data symbols.

The soft symbol estimation can be obtained as follows:

$$\hat{d}_k = \sum_{d \in \mathrm{B}} d \prod_{i=1}^{m} p(b_i) \tag{35}$$

where, $\mathrm{B} = \{d_1, d_2, \cdots, d_M\}$ denotes the finite alphabet used for MPSK, b_i denotes the ith coded bit, $i = 1,2,\cdots m$.

The probability distributions $P(b_i)$ of coded bits can be obtained from the corresponding LLR values $L(b_i)$. Therefore:

$$P(b_i) = \frac{e^{b_i \cdot L(b_i)}}{1 + e^{L(b_i)}} \tag{36}$$

4. Simulation results

In this section, we use simulation experiments to verify the performance of the proposed soft IED algorithm. The system parameters of computer simulation are shown in Table 1. The outer decoders of SCTCM adopt convolutional codes encoder with 4-state 1/2 code rate for QPSK modulation. In this paper, we integrate the multi-path fading and additive white Gaussian noise to simulate underwater acoustic channel.

Based on the sound speed profile (SSP) measured in the lake and finite-element ray (FER) tracing method (Bellhop) (Porter & Liu, 1994), the channel impulse response is shown in

Fig.7. From the SSP (shown in Fig.1.), we know that the water with mixed gradient SSP is about 53m deep. The transmission distance is 2000m.

Carrier frequency	10 KHz
Symbol rate	5 Kbps
Doppler	10 Hz
Array elements	4
Training symbols	200 sys

Table 1. Simulation parameters

4.1 Soft iterative performance of iterative equalization and decoding

The system parameters of simulation are shown in Table 1. The channel impulse responses are shown in Fig.7. As shown in Fig.7, the multi-path propagation is very seriously.

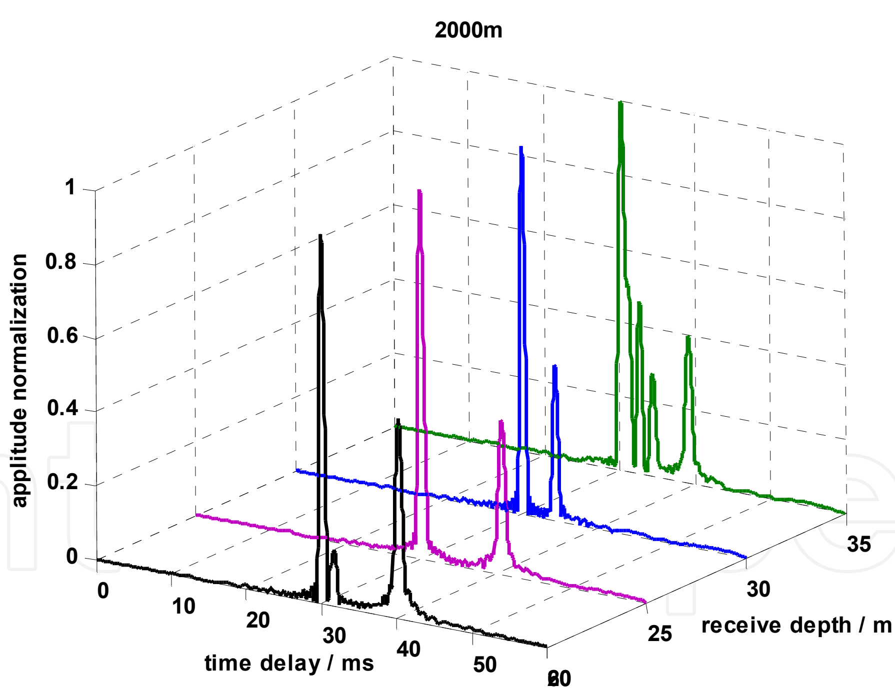


Fig. 7. Channel impulse response (2000m)

Fig.8 shows the BER curves of IED algorithm with soft iteration for QPSK modulation. As show in Fig.8, the iteration algorithm can sufficient utilize the decoding gain provided by decoder of SCTCM to enhance the equalizer performance such that the system performance is increased and the data transmission with lower BER can be obtained.

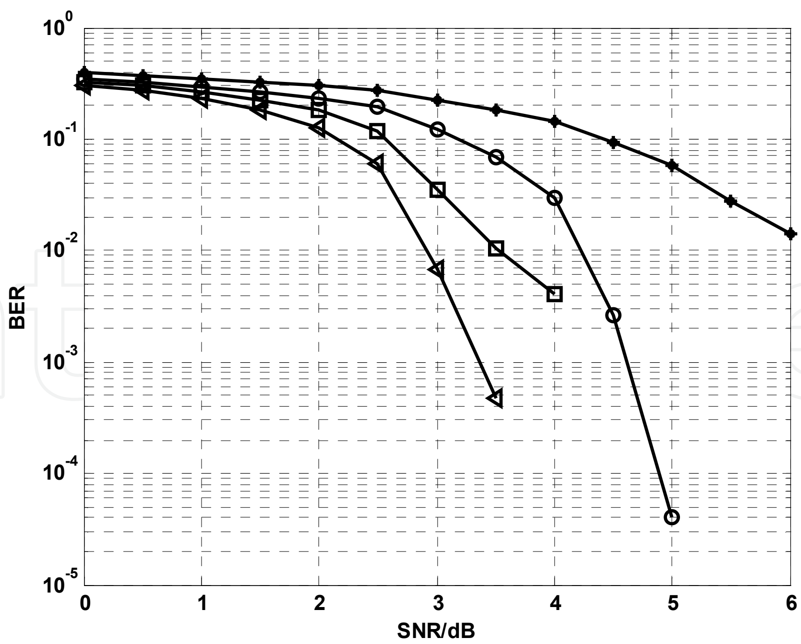


Fig. 8. Performance of the soft IED

4.2 Dual model of iterative equalization and decoding

The dual model IED is shown in Fig.9. There are two parts: (1) iterative equalization and decoding; (2) iterative decoding. As mentioned in Section 3, the mutichannel ADFE and decoder of SCTCM exchange soft information in an iterative manner in the IED scheme. So, we can perform decoding iteration before IED. And thus, the accuracy of symbol estimation is further improved such that the equalizer performance is improved greatly.

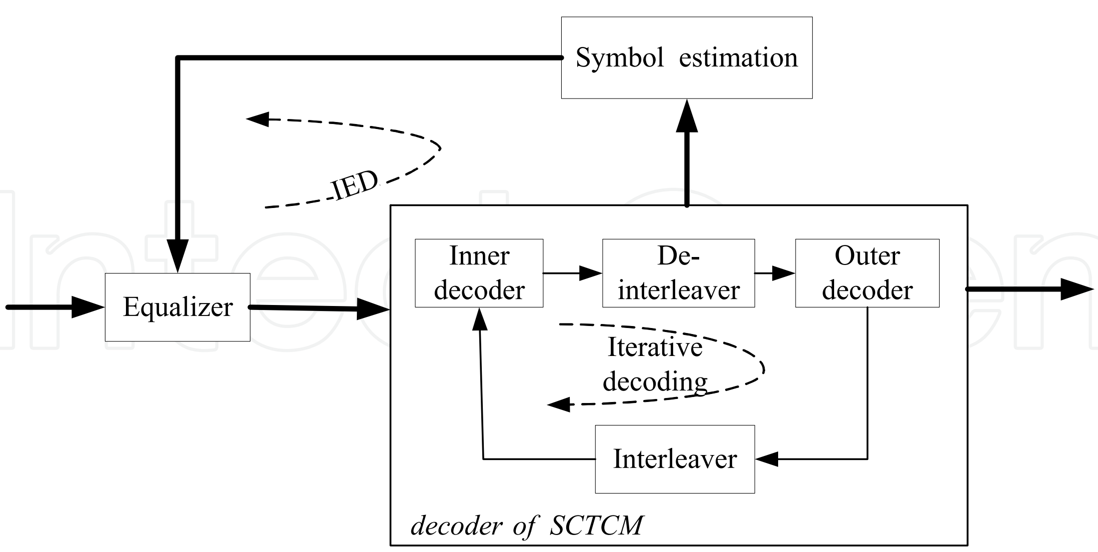


Fig. 9. Dual model IED

As shown in Table 2, in dual model IED, the soft symbols estimation, based on the coded bits output from decoder of SCTCM, are more accurated. And thus, the propagation errors can be futherly reduced.

IED iteration	0		1		2	
Decoder iteration	0	1	0	1	0	1
3 dB	0.2217	0.2072	0.1205	0.0824	0.0351	0.0185
3.5 dB	0.1856	0.1582	0.0682	0.0326	0.0105	0.0042
4 dB	0.1451	0.0987	0.0295	0.0123	0.0041	0.0024
4.5 dB	0.0941	0.0358	0.0026	1e-4	0	0
5 dB	0.0589	0.0097	4e-5	0	0	0
5.5 dB	0.0282	0.0016	0	0	0	0
6 dB	0.0141	3e-4	0	0	0	0

Table 2. Performance of dual model IED

5. Conclusions

In this chapter, according to the characteristics of underwater acoustic channel, SCTCM technology with rate-1 inner code, is adopted to improve the bandwidth efficiency of underwater acoustic channel. Simultaneously, LLR calculation is introduced to simplify symbol decoder into bits decoder. The soft IED scheme with soft symbol estimation is proposed to overcome the multi-path fading of underwater acoustic channel and enhance the performance of equalizer through utilizing decoding gain provided by decoder and the information symbols with soft symbol estimation fed back to equalizer so that the performance of communication system is improved greatly. What's more, the dual model IED scheme is proposed to obtain lower BER in order to meet the demands of higher system performance. The simulation results verify the proposed algorithm can obtain satisfied data transmission with the small iterations, especially the dual model IED.

6. References

Bragard P., Jourdain G.. A fast self-optimized algorithm for non-stationary identification: application to underwater equalization, *Proceedings of IEEE ICASSP*, pp. 1425-1428, 1990.

Berrou C., Glavieux A., and Thitimajshima P.. Near Shannon Limit Error-Correcting Coding:Turbo Codes, *Proceeding of IEEE Int. Conf on Communications*, pp. 1064-1070, Geneva, Switzerland, May 1993.

Benedetto S., Montorsi G.. Serial concatenation interleaved codes: performance analysis, design and iterative decoding. TDA Progress Report 42-126, 1996.

Benedetto S., Divsalar D., Montorsi G., Pollara F.. Parrallel concatenated trellis coded modulation, *Proceeding of IEEE ICC'96*, Vol.2, pp. 974-978.

Benedetto S., Divsalar D., Montorsi G., Pollara F.. Serial Concatenated Trellis Coded Modulation with Iterative Decoding: Design and Performance, *IEEE Global Telecommunications Conference*, (CTMC), Nov. 1997.

Benedetto S., Divsalar D., Montorsi G., Pollara F.. Serial concatenation of interleaved codes: performanceanalysis, design, and iterative decoding. *IEEE Trans. on Information Theory*, Vol. 44, May 1998, pp. 909-926.

Berthet, A.O.; Visoz, R.; Unal, B.; Tortelier, P.. A comparison of several strategies for iteratively decoding serially concatenated convolutional codes in multipath Rayleigh fading environment, *Proceeding of Global Telecommunications Conference*, Vol.2, pp. 783-789, 2000.

Chung S.Y., Lou H.. TCM and turbo TCM for audio broadcasting, *Proceeding of Vehicular Technology Conference*, Vol.3, pp. 2222-2226, Tokyo, Japan, May 15-18, 2000.

Choi J.W.. Iterative Multi-Channel Equalization and Decoding for High Frequency Underwater Acoustic Communications, *IEEE SAM*, 2008.

Divsalar D., Pollara F.. Hybrid Concatenated Codes and Iterative Decoding, JPL TMO Prog. Report, Aug. 15, 1997.

Divsalar D., Dolinar S., Pollara F.. Serial concatenated trellis coded modulation with rate-1 inner code, *Proceedings of IEEE Global Telecommunications Conference*, Vol.2, pp. 777-782, Nov.27-Dec.1, 2000.

Flanagan M.F., Fagan A.D.. Iterative Channel Estimation, Equalization, and Decoding for Pilot-Symbol Assisted Modulation Over Frequency Selective Fast Fading Channels. *IEEE Transactions on Vehicular Technology*, Vol. 56, No. 4, July 2007, pp. 1661-1670.

Ho M. "Performance bounds for serially-concatenated trellis-coded modulation", Signals, Systems & Computers, 1997. Conference Record of the Thirty-First Asilomar Conference on Volume 2, 2-5 Nov. 1997:1364 – 1368.

Kilfoyle D.B. and Baggeroer A.B.. The state of the art in the underwater acoustic communications. *IEEE Journal of Oceanic Engineering*, Vol. 25, No. 1, 2000, pp. 4-27.

Koetter R., Singer A.C., Tuchler M.. Turbo equalization. *IEEE Signal Processing Magazine*, Vol. 21, No. 1, Jan. 2004, pp. 67-80.

Legoff S., Glavieux A., and Berrou C.. Turbo codes and high spectral efficiency modulation, *Proceeding of IEEE ICC'94,* New Orleans, LA, May 1-5 1994.

Li X., Song W.T., Luo H.W.. Joint turbo equalization and turbo TCM for mobile communication systems, *Proceeding of 12th IEEE International Symposium on Personal, Indoor and Mobile Radio Communications*, Vol.1, pp. A184-A188, Sep. 30.-Oct. 3, 2001.

Noorbakhsh M., Mohamed-Pour K.. Combined turbo equalisation and block turbo coded modulation, *Proceeding of IEE Communications*, Vol. 150, No. 3, pp. 149-152, June 2003.

Otnes, R., Tuchler M.. Iterative channel estimation for turbo equalization of time-varying frequency-selective channels. *IEEE Transactions on Wireless Communications*, No.6, Nov. 2004, pp. 1918-1923.

Porter M.B., Liu Y.C.. Finite-Element Ray Tracking. *Theoretical and computational Acoustics*, Vol. 2, 1994, pp. 947-956.

Proakis J.G.. Digital communication (4th Edition) . Beijing: Publishing House of Electronics Induxtry, 2003.

Robertson P., Woerz T.. Bandwidth efficient turbo trellis coded modulation using punctured component codes. *IEEE Journal on Selected Areas in Communications*, Vol.16, No.2, Feb.1998.

Stojanovic M., Catipovic J., Proakis J.. Recent advances in high-speed underwater acoustic communications. *IEEE J .Oceanic Eng.*, Vol. 21, No. 2, April 1996, pp. 125-136.

Soleymani M.R., Gao Y.Z., Vilaipornsawai U.. Turbo Coding for Satellite and Wireless CommunicationsBoston. MA: Kluwer Academic Publishers, 2002.

Shohon T., Jinbo K., Ogiwara H. Performance improvement with increasing interleaver length for serially concatenated trellis coded modulation. *Proceeding of IEEE International Symposium on Information Theory*, June 29-July 4, 2003.

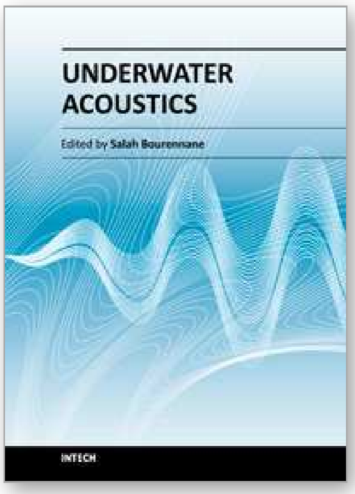
Stojanovic M., Frieitag L.. Wideband Underwater CDMA: Adaptive Multichannel Receiver design, *Proceeding of MTS Oceans 2005*, pp. 1-6, USA 2005.

Tuchler M., Singer A.C., Koetter R.. Minimum mean squared error equalization using a priori information. *IEEE Transactions on Signal Processing*, Vol.50, No.3, March 2002, pp. 673-683.

Ungerboeck G.. Channel coding with multilevel phase signaling. *IEEE Trans. Inf. Th.*, Vol.IT-25, Jan. 1982, pp. 55-67.

Yang P., Ge J.H.. Combination of turbo equalization and turbo TCM for mobile communication system, *Proceeding of IEEE International Symposium on Communications and Information Technology*, Vol.1, pp. 383-386, Oct.12-14, 2005.

Zhao L., Zhu W.Q., Zhu M.. Adaptive Equalization Algorithms for Underwater Acoustic Coherent Communication System. *Journal of Electronics & Information Technology*, Vol. 30, No. 3, 2008, pp. 648-651.



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