# QPSK System Design for Underwater Acoustic Communication

(수중음향통신용의 QPSK 시스템 설계)

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A thesis submitted in partial fulfillment of the requirements
for the degree of
Master of Engineering

in the Department of Telematics Engineering, Graduate School,
Pukyong National University

February 2003

# QPSK System Design for Underwater Acoustic Communication

A Dissertation

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December 26th, 2002

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## Abstract

Underwater acoustic(UWA) communications have been developed rapidly in recent years. Their applications are beginning to shift from military towards commercial. UWA communications has mostly relied on the use of noncoherent modulation techniques, such as FSK or DPSK, due to the characteristic of underwater channels such as multipath.

The UWA channel is characterized as a time-dispersive rapidly fading channel, which exhibits Doppler instabilities. While vertical channels exhibit little multipath propagation, horizontal channels suffer from extended multipath propagation which usually increases with range and, depending on the signaling rate, causes the intersymbol interference(ISI). In some applications, unpredictable motion of the receiver and transmitter, as well as changes in the transmission medium, cause severe phase fluctuations. This is the main reason why coherent communications are often not considered feasible.

The main limitation of an UWA channel is therefore the combination of time-varying multipath and phase instabilities. However, in order to achieve the bandwidth and power efficiency on

the bandlimited UWA channels, phase-coherent modulation techniques should be employed, such as M-ary PSK.

In this work the design of UWA communication systems focuses on the QPSK modulation, and the ways to remove ISI caused by multipath, therefore coherent underwater acoustic communications (QPSK Modem) with adaptive equalization is proposed, then the systems characteristic and performance will be examined by numerical simulation in the cases of horizontal and vertical UWA channels transmission. Finally the simulation results illustrate the necessity to employ the equalizer for UWA channel transmission. And it's seen that the rate of convergence of RLS is faster than that of LMS, and for the channel with spectral null, the nonlinear equalizer DFE is very suitable to be employed to cancel ISI. In all adaptive equalization makes the phase coherent modulation possible and improves the transmission data rate in UWA channel.

# Ⅰ. Introduction

The first underwater source that transmitted sound was developed in 1915, and the extensive use of sonar (sound navigation and ranging) for submarine detection was in the World War Ⅱ. In 1919 soon after the end of the World War Ⅰ, the first scientific paper on underwater sound was published. From a scientific standpoint, the most notable accomplishment of the years between the World War I and the World War Ⅱ was the obtaining of an understanding of the vagaries of sound propagation.[8]

With the development of electronics, the acoustic sensors became more effective, greater ranges could be achieved and smaller targets could be identified. In recent years, underwater acoustic techniques are beginning to shift from military toward commercial areas, such as pollution monitoring and applications in offshore oil industry. The goal of research is to bring together many of the well established principles of wireless radio communications and reconsider them for application in UWA channel. The development of specialized communication techniques is often required to meet the demands of underwater propagation with digital acoustic signals. UWA communications have been established as a new field of application.

Since the publication of the Special Issue on Ocean Acoustic Data Telemetry in 1991, fundamental advances have been made in this field. For UWA channel the most challenging task is to combat the ocean multipath. Multiple scattering by surface and bottom is the most frequently encountered in underwater transmission. While vertical channels exhibit little multipath propagation, horizontal channels suffer from extended multipath propagation, which cause intersymbol interference(ISI).[1][14]

The UWA channel is characterized as a time-dispersive rapidly fading channel, which exhibits Doppler instabilities. In some applications, unpredictable motion of the receiver and transmitter, as well as changes in the transmission medium, cause severe phase fluctuations. This is the main reason why coherent communications are often not considered feasible. Bandwidth-efficient phase-coherent communications to achieve high-speed data transmission through underwater channels, including the severely time-spread horizontal shallow water channel. The main limitation of the UWA channel is therefore the combination of time-varying multipath and phase instabilities. To achieve the bandwidth and power efficiency on the bandlimited UWA channels, phase-coherent modulation techniques should be employed, such as M-ary PSK.[14]

More recently spread spectrum techniques have been considered for resolving and combating multipath, especially for SNR inferior to 0 dB. However, that means is at the cost of bandwidth efficiency. The use of very narrow transmitter beams can also be used to suppress the effects. If the signal to noise ratio is not too low(not less than 15dB), the best solution is to use adaptive equalization techniques, principal decision feedback equalization processing with LMS or RLS algorithm. Due to the adaptive and nonlinear characteristic of LMS or RLS decision feedback equalizer, the phase coherent modulation used in UWA channel systems becomes achievable. Equalization techniques therefore optimize the data rate.

# II. Channel model and its characteristic

The impulse response of an underwater acoustic channel is modelled as

$$h_c(t) = \sum_{n=0}^{\infty} a_n e^{-j2\pi f_c \tau_n} \delta(t - \tau_n) \tag{2.1}$$

where $a_n$ is the complex amplitude of the signal received along the n-th path and $\tau_n$ designates the propagation delay for that path. The amplitude and propagation delay are normalized for the direct path signal (n=0), such that $a_0$=1 and $\tau_0$=0. Here n represents the number of paths . And we assume the channel variations are slow compared with the signalling interval for the simplicity of channel model analysis.

Sound propagation in underwater is primarily determined by transmission loss, noise, reverberation and temporal and spatial variability of the UWA channel. Transmission loss is caused by energy spreading and sound absorption. The spreading loss, is proportional to the square of the transmission distance. Absorption loss occurs because of the transfer of acoustical energy into heat. The absorption loss increases with frequency and range. The absorption loss is a primary factor determining the maximal usable frequency and therefore the available bandwidth.

Reverberation of the received signal is caused by multipath propagation. Generally the water depth determines the type of propagation. If the water is shallow, propagation will occur in surface bottom bounces in addition to a possible direct path. If the water is deep, as in the regions past the continental shelves, the sound channel may form by bending of the rays toward the location where the sound speed reaches its minimum, called the axis of the deep sound channel.

Multipath propagation contributes to signal fading and causes ISI in a digital communication system. These phenomena affect every propagation path, resulting in an extremely dynamic overall multipath structure in all but few UWA channels. While vertical propagation path exhibits little multipath dispersion, horizontal channels are characterized by extremely long reverberation times.

## Ⅱ-1. Transmission Loss

At short range, let a source of sound be located in a homogeneous , unbounded, lossless medium. Spreading loss for short range transmission is

$$SL = 20\log r \tag{2-2}$$

Where SL is spreading loss, $r$ is the transmission range .
At moderate and long range, let the medium has plane-parallel upper and lower bounds, the average energy is expected to spread cylindrically with the waveguide:

$$SL = 10\log r \tag{2-3}$$

Total transmission loss is a summary of the processes of attenuation and absorption at sea(usually wide spread of measured data), and may be expressed as

$$TL = 20\log r + ar \tag{2-4}$$

The first term is spherical spreading, and the second term absorption loss. Where $a$ is the absorption coefficient, which associated with frequency, as seen in Fig. 2.1.

A simpler and more specific expression of $a$ is given as

$$\alpha(f) = \left(\frac{0.08\, f^2}{0.9 + f^2} + \frac{f^2}{3000 + f^2} + 4\times\ 10^{-4} + f^2\right)/1000 \qquad (2\text{-}5)$$

As expressed in the above equations, absorption loss increases with frequency and transmission distance, therefore the available bandwidth is mainly determined by the absorption loss seen in Fig.2.1.

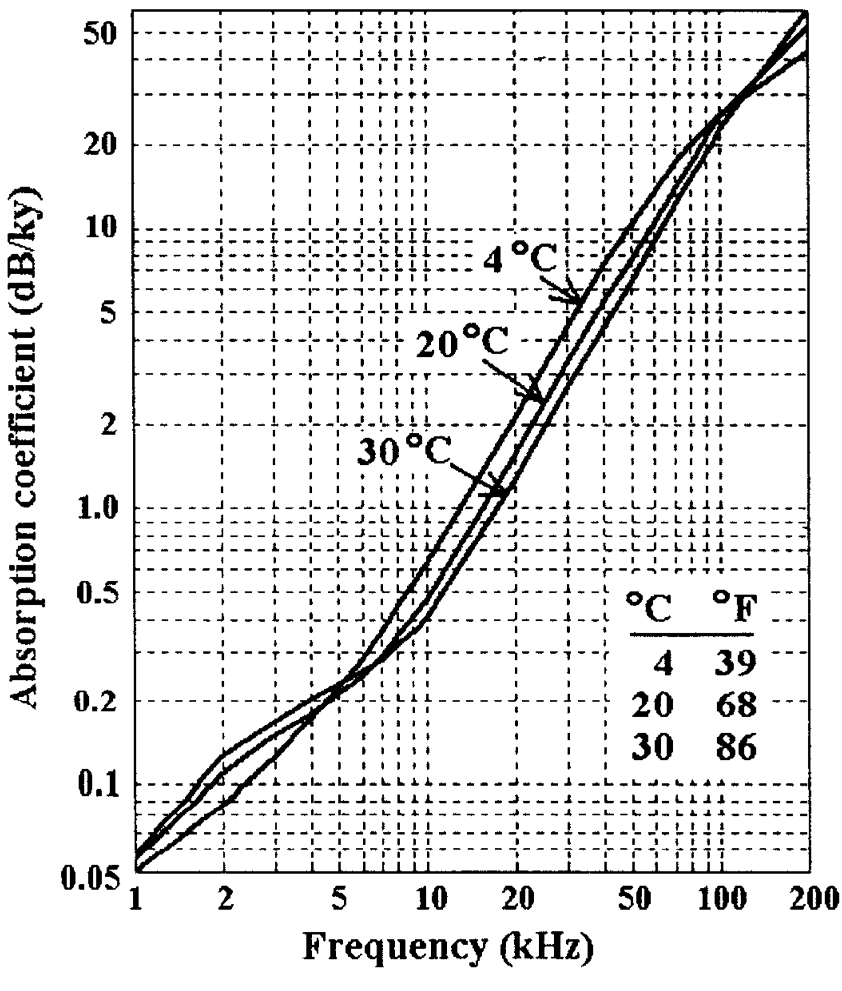



Fig. 2.1 Absorbtion loss characteristic[8]

For multiple ray paths, reflection loss occurring at the surface and bottom interfaces are not negligible factors. Especially at the sea surface interface the impedance is determined by the wave height and frequency, and the surface and bottom reflection loss $\alpha_{rs}$ and $\alpha_{rb}$ are defined as

$$\alpha_{rs} = 1 - \log_{10}(1 - 0.0234\ (fH)^{3/2}) \qquad (2\text{-}6)$$

$$\alpha_{rb} = -20\ \log_{10} R_b \qquad (2\text{-}7)$$

$$\text{where} \quad R_b = \frac{(Z_2/Z_1)\sin\theta_i - \sin\theta_t}{(Z_2/Z_1)\sin\theta_i + \sin\theta_t}$$

Where f is the transmission frequency[kHz], H the average wave height[ft]

$R_b$ the bottom reflection coefficient

$\theta_i$ and $\theta_t$ the angles of incident and refraction , respectively

$Z_2/Z_1$ bottom layer and sea layer impedance ratio

## Ⅱ-2. Ray tracing

The velocity of sound in the sea is related to more readily measured oceanographic parameters, such as temperature, salinity, and depth. The "velocity profile" is meant by the variation of sound velocity with depth, or the velocity-depth function. Snell's law describes the refraction of sound rays in a medium of variable velocity. Thus the principle of circularity of rays in linear gradients is commonly employed in ray-tracing computers. The initial step is to divide the velocity profile into layers of constant linear gradient and to program the computer to follow, the arcs of rays leaving the source at different angles. The speed profile and ray tracing are shown in Fig.2.2.

A shadow is produced, in which the intensity from the source is very low. The ray trace diagram is used to predict the sound propagation. From which a direct path propagation from the transmitter and the receiver can be predicted.

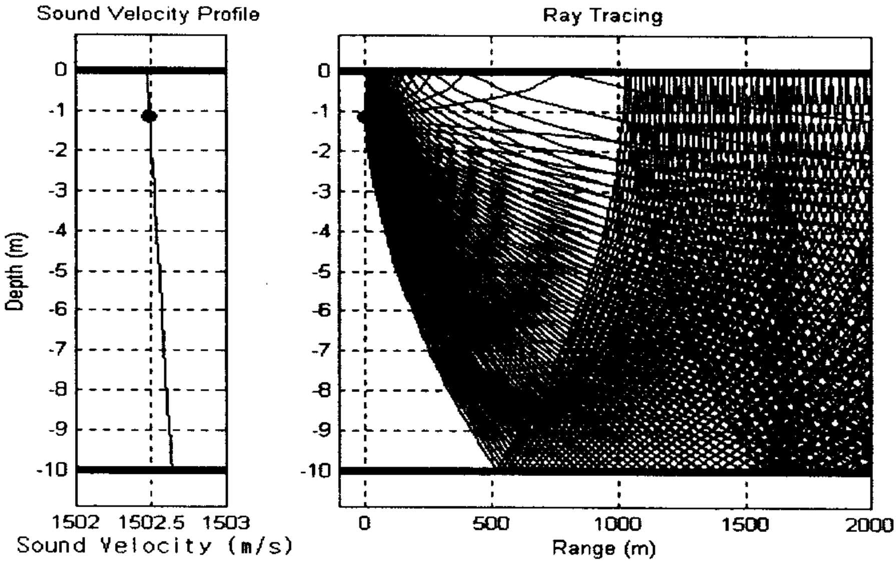



Fig. 2.2. Ray tracing result of the measurement site [8]

## II-3. Doppler effects

Doppler spread of the signal is regarded as a consequence of two distinct phenomena: the time-variability inherent in the transmission medium , and that caused by the relative motion of the transmitter and receiver pair. Surface motion caused by wind driven waves contributes to the time-variability of the shallow water channel.

The surface of the sea is not smooth, the vertical motion of the waves superposes itself, the moving surface produces upper and lower sidebands in the spectrum of the reflected sound that are the duplicates of the spectrum of the surface motion.

The relative motion of transmitter and receiver pair changes the frequency of received signals. The Doppler shift may be positive or negative, depending on the geometry and the direction of motion.

Of particular important in the design of modulation and coding for fading multipath channels is the value of the spread factor of the channel, defined as $T_m\ B_d$, where $T_m$ is the multipath spread and $B_d$ is the Doppler spread. If $T_m\ B_d<1$, the bandwidth efficient method for transmitting the modulated signal and receiving it with phase coherent demodulator combined with adaptive equalizer to reduce the effects of intersymbol interference can be employed when the channel time variation is sufficiently slow relative to the symbol duration to allow for carrier phase estimation and tracking.

Doppler spread caused by the exploit motion of the system at velocity v is proportional to vf/c. This factor is often found to surpass that caused by the surface waves. The normalized Doppler spread $T_m\ B_d$ is the factor that determines the system performance degradation.

## Ⅱ-4. Ambient Noise

Ambient noise is observed that exhibits strong site as well as frequency-dependence. Generally, the inshore environments, such as marine work-sites, are much more noise than the deep ocean, due to the man-made noise. Most of the ambient noise sources can be described as having a continuous spectrum and Gaussian statistics.

For shallow water, the ambient noise is a mixture of (1) shipping and industrial noise, (2) wind noise, and (3) biological noise.

For deep water, the ambient noise of the sea has a directionality of its own. Measurement shows that ambient noise has different characteristics at different frequencies. The principle sources of noise includes tides, the hydrostatic effects of waves, seismic disturbances, ocean turbulence, nonlinear wave interactions, ship traffic and surface waves.

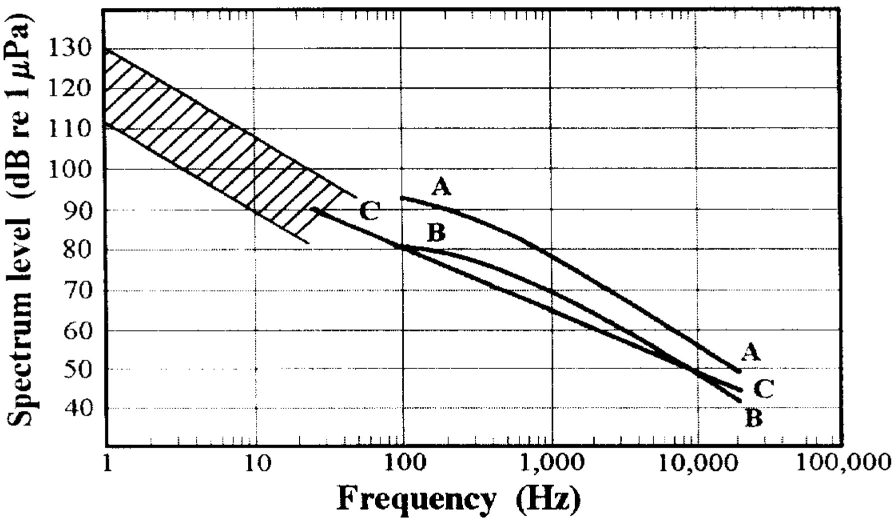



Fig. 2.3. Ambient noise in the three different costal bays[8]

A: High noise location; B: An average noise location; C: Average of many World War II measurements Shaded area: Subsonic background measurements

# Ⅲ. QPSK Modulation and Equalizer

## Ⅲ-1. Digital modulation

There are several basic digital modulation methods: FSK, PSK and ASK. FSK techniques, in which guard times are inserted between pulses of the same frequency, ensuring that all the channel reverberation will die out. Amplitude modulation has been associated with the problem of determining detection thresholds in the environments with large amplitude fluctuations. Spread spectrum techniques has been considered for resolving and combating multipath. In addition to extracting just the principal multipath component, these techniques may be used to recombine all the multipath components in the Rake receiver. While both the approach of signal design with guard times, and the spread spectrum techniques, sacrifice the available bandwidth to suppress the intersymbol interference, equalization offers bandwidth-efficient independently fluctuating propagation paths.

With the goal of increasing the bandwidth efficiency of an UWA communication system, research focus has shifted toward phase-coherent modulation technique, such as QPSK. Depending on the method for carrier synchronization, phase-coherent systems fall into two categories, though differentially coherent detection is the simple carrier recovery it allows, its disadvantage is performance loss as compared to coherent detection.

## Ⅲ-2. QPSK modulation and demodulation

The latest development in phase-coherent communication is based on purely phase-coherent modulation and detection principles. The modulating format QPSK is presented here. The transmitter comprises an encoder, lowpass filter, and a modulator seen in Fig. 3.1. The input to the encoder is a stream of binary digits representing the information to be transmitted.

The encoder samples successive pairs of bits and generates a four-level stream of data symbols { $s_i$} Each level corresponds to a different bit combination:00,01,10,11, which produces a phase change of either [0° 90° 180° 270°] or [±45° ±135°] by QPSK modulator. Since each level has a duration of 2-bit periods, the symbol rate is half the bit rate and the bandwidth required is therefore halved. The transmitter low-pass filter suppresses the high-frequency harmonics of each symbol and shapes the spectrum of the encoded signal to reduce the bandwidth.

Fig.3.1 and Fig.3.2 show the communication channel in more detail. I and Q represent the in-phase and quadrature components of { $s_i$}, respectively. The four level symbols at the output of the two low-pass filters in the transmitter are

$$s_i = s_{I,i} + s_{Q,i} = \pm 1 \pm j \tag{3-1}$$

The general analytic expression for QPSK is

$$s_i(t) = \sqrt{\frac{2E}{T}} \cos[\omega_0 t + \Phi_i(t)] \quad 0 \le t \le T \tag{3-2}$$

where the phase term, $\Phi_i$, will have four discrete values, typically given by

$$\Phi_i(t) = \frac{2\pi i}{4} \qquad i=1,2,3,4$$

The parameter E is symbol energy, T is symbol time duration.

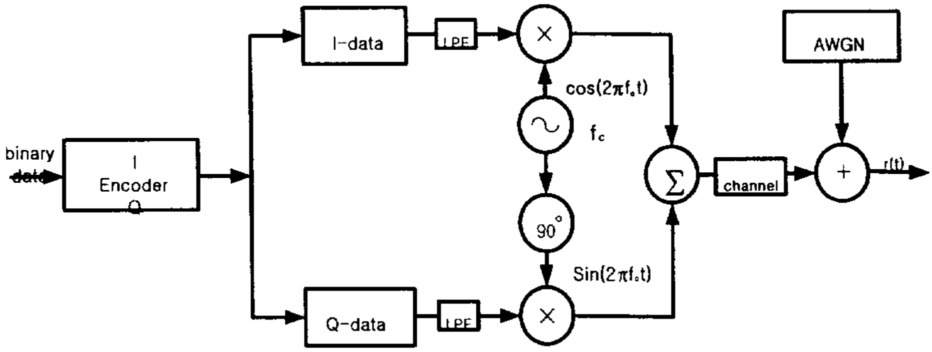



Fig. 3.1 QPSK modulation Block diagram

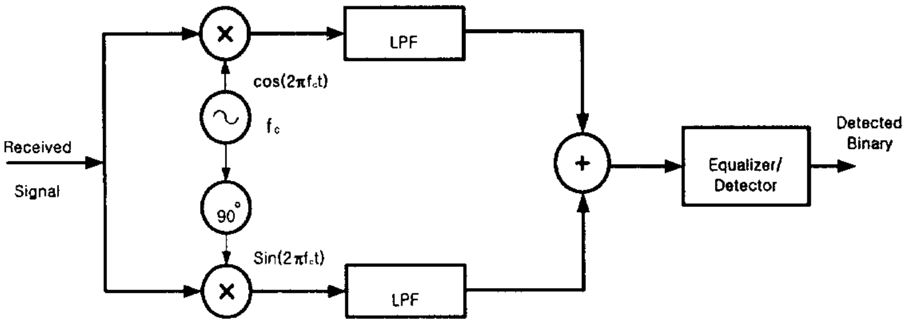



Fig. 3.2 Block diagram of QPSK demodulation

We choose a convenient set of axes, such as

$$\phi_1(t)=\sqrt{\frac{2}{T}}\cos\ \omega_0 t$$

and

$$\phi_2(t)=\sqrt{\frac{2}{T}}\sin\ \omega_0 t$$

where the amplitude $\sqrt{2/T}$ has been chosen to normalize the expected output of the detector, now $s_i(t)$ can be written in terms of these orthogonal coordinates, giving

$$\begin{aligned} s_i(t) &= a_{i1}\ \phi_1(t)+\ a_{i2}\ \phi_2(t) \\ &=\sqrt{E}\cos(\frac{2\pi i}{4})\ \phi_1(t)+\sqrt{E}\sin(\frac{2\pi i}{4})\ \phi_2(t) \end{aligned} \tag{3-3}$$

The received signal r(t) can be expressed by

$$r(t)=\sqrt{\frac{2E}{T}}(\cos\ \Phi_i\cos\ \omega_0 t+\sin\ \Phi_i\sin\ \omega_0 t+n(t) \tag{3-4}$$

where n(t) is a zero-mean white Gaussian noise process.

There are only two reference waveforms and the upper correlator computes

$$X=\int_0^T r(t)\ \phi_1(t)dt \tag{3-5}$$

and the lower correlator computes

$$Y=\int_0^T r(t)\ \phi_2(t)dt \tag{3-6}$$

The received phase angle $\hat{\Phi}$ can be computed as arctan of Y/X, where X can be thought of as the in-phase component of the received signal, Y is the quadrature component, and $\hat{\Phi}$ is a noisy estimate of the transmitted $\Phi_i$. The demodulator computes $|\Phi_i - \hat{\Phi}|$ for each of the $\Phi_i$ prototypes and chooses the $\Phi_i$ yielding the smallest output.

Owing to multipath propagation in the channel, the acoustic signals are received with random phase relationship. They are also degraded by the addition of noise n(t), which is assumed to be white Gaussian noise.

At the receiver seen in Fig.3.2 the signals are demodulated, detected and decoded. The sampled output $\{r_i\}$ of the receiver filter is a noisy and distorted four level baseband signal $s_i(t)$. The receiver filter band limits the noise component outside the signal frequency band and maximize the SNR at the detector input. The detector then shapes the symbol. Finally the decoder generates a stream of bits.

## Ⅲ-3. Equalizer

In underwater communications multipath is the most difficult to overcome due to the time-dispersions characteristics of channel. For the aim to compensate for the effects, equalizer is employed , which takes an important role in UWA channel transmission.

ISI caused by distinct propagation paths can be compensated by equalization which takes into account the time diversity of the channel. The problem of reducing the receiver complexity may be addressed on two levels: the algorithms and the structural level. For application in time-varying channels, the receiver must use an adaptive algorithm for adjusting its parameters.

There are two criteria for optimizing the filter coefficients for suboptimum equalizer: one is the peak distortion criterion, such as zero forcing equalizer. The other is the mean-square error criterion, such as RLS or LMS equalizers.

LMS algorithm shown in Fig.5 is the most widely used among the adaption algorithms.
The update tap-weight

$$w(n+1) = w(n) + k(n)\ e^{*}(n) \tag{3-8}$$

Estimation error

$$e(n) = d(n) - y(n) \tag{3-9}$$

Filter output

$$y(n) = w^{H}(n)\, u(n) \tag{3-10}$$

Where d(n) is the desired response, u(n) is the tap input vector.

The above equations are the basic LMS algorithm for recursively adjusting the tap weight coefficients of the equalizer. During the filtering process, the desired response d(n) is supplied for processing, alongside the tap input u(n). The feature of the LMS algorithm is its simplicity.

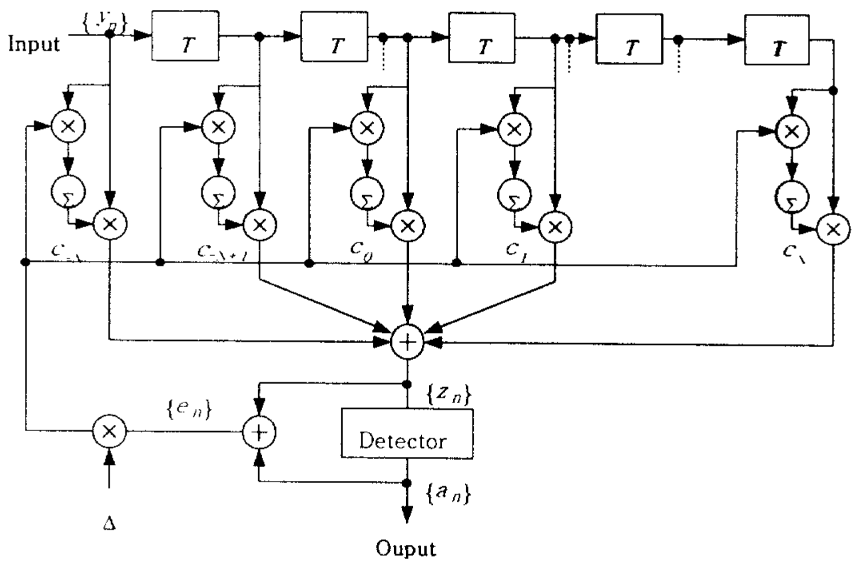



Fig. 3.3 Linear adaptive equalizer based on the MSE criterion

RLS algorithm(adaptive equalization) is represented as follows:

$$k(n)=\frac{\lambda^{-1}p(n-1)u(n)}{1+\lambda^{-1}\,u^{H}(n)p(n-1)\,u(n)} \tag{3-10}$$

$$w(n)=w(n-1)+k(n)\,e^{*}(n) \tag{3-11}$$

$$e(n)=d(n)-w^{H}(n-1)\,u(n) \tag{3-12}$$

$$p(n)=\lambda^{-1}p(n-1)-\lambda^{-1}k(n)\,u^{H}p(n-1) \tag{3-13}$$

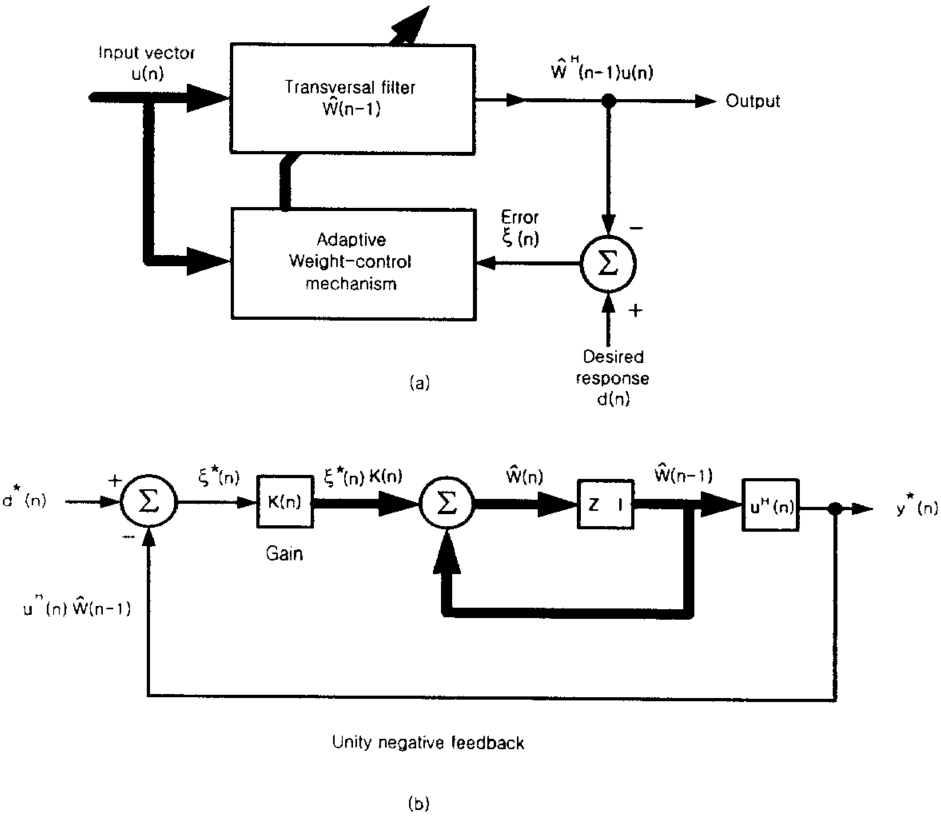



Fig.3.4 Representations of the RLS algorithm:
(a)block diagram;(b)signal-flow graph

where k(n) is Kalman gain vector varying with time, w(n) is the updating tap coefficient vector changing with time, and w(n) is controlled by one of the elements of k(n), consequently RLS has fast convergence but at a cost of a large computational complexity. Here p(n) is the inverse signal correlation matrix, u(n) is the input signal, and $\lambda$ is the exponential weighting factor( forgetting factor) which is a positive constant.

In a majority of recent studies, the LMS-based algorithms are considered an only alternative, due to their low computational complexity(linear in the number of coefficients N). However, the LMS algorithm has a convergence time which may become unacceptably long when large adaptive filters are used.

RLS algorithms on the other hand, have better convergence properties but higher complexity. The quadratic complexity of the standard RLS algorithm is too high when large adaptive filters need to be implemented. In general it's desirable that the algorithm be of linear complexity. At high symbol rates, the long ISI requires large adaptive filters, therefore increasing the computational complexity.

An approach for multipath rejection using adaptive beamforming at the receiver end has been investigated , and it was found that the beamformer encounters difficulties as the range increases relative to depth. To compensate for this effect, the use of an equalizer was proposed to complement the performance of the beamformer. The equalizer is of a decision-feedback type, and it operates under the LMS algorithm whose low-computational complexity permits real-time adaptation at the symbol rate.

Fig.3.4 shows the block diagram of DFE(decision feedback equalizer). In most communication systems, the channel characteristics are unknown a priori and the channel response is time-variant. The equalizers employed here consists of adaptive feedforward and feedback filters. The decision feedback equalizer is an example of non-linear equalizer. The input to the feedforward filter is the received signal sequence from matched filter, the input to the feedback filter is the sequence of decisions on the previously detected symbols, and the function of the feedback filter is to remove that part of the intersymbol interference from the present estimate caused by previously detected symbols.

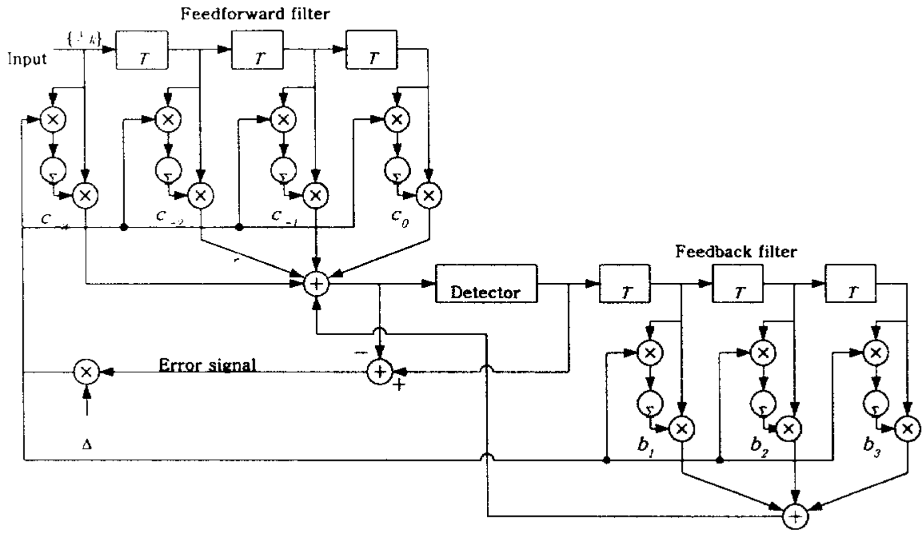



Fig. 3.5 Adaptive DFE

# Ⅳ. Numerical model and simulation results

## Ⅳ-1. Channel Model

The combined system impulse response is

$$h(t) = h_a(t) * h_c * h_b(t) \tag{4-1}$$

where $h_a$, $h_b$ are the impulse responses of the transmitter filter LPF and receiver filter LPF, respectively.

The samples (once per symbol) of the received demodulated signal can be expressed as

$$r_i = s_i h_0 + \sum_{n=1}^{g} s_{i-n} h_n + N_i \tag{4-2}$$

where $s_i h_0$ is the wanted signal, and $\sum_{n=1}^{g} s_{i-h} h_n$ is intersymbol interference. g is the number of paths except for the direct path, including the surface reflection(SS1), the bottom reflection(BB1), the surface-bottom reflection(SB1) and the bottom-surface reflection(BS1). (seen in Fig.4.1)

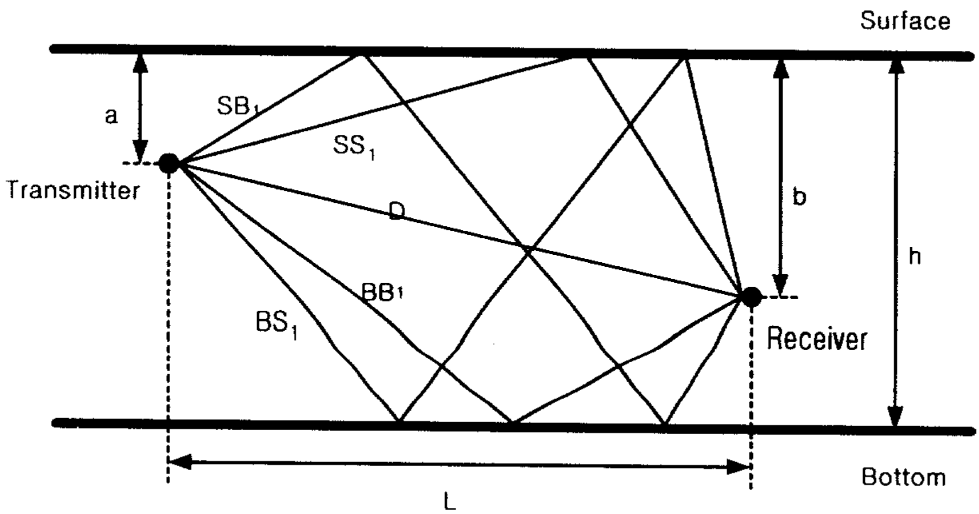



Fig. 4.1 Multipath Model

The environment is assumed as follows: According to the horizontal range to the water depth ratio, transmission channels are divided into vertical channel and horizontal channel. In simulation the ocean depth(h) is 100m, the source depth from the surface is 5m, the hydrophone 97m.

In the simulation the process assumes that the absorption varying with frequency didn't take into consideration. The average wave height is 0.18 feet. The surface reflection coefficient is computed according to equation

```
s_r=-1+0.0234*((Fc/1000)*Hw).^(3/2)        (4-1)
```

The bottom reflection is 0.4118 corresponding to the sandy bottom. For vertical channel let the SNR be infinite, thus the ambient noise doesn't exist. The transmission date symbol duration is 0.4msec, the bit duration(T) equals to 0.2msec. The carrier frequency is taken as 25kHz.

For vertical channel , the horizontal distance R between the source and the receiver is 10m. Their delay times and corresponding reflection coefficients computed are as follows:

Delay_1 Time : 3.978 msec
Loss_1 : 0.38686
Delay_2 Time : 6.6314 msec
Loss_2 : -0.70125
Delay_3 Time : 10.6134 msec
Loss_3 : -0.27287
Delay_4 Time : 133.0862 msec
Loss_4 : -0.1013

For vertical channel the first multipath (bottom reflection) arrival is delayed by 3.978 msec, the reflection coefficient of which is smaller than that of 0.75msec(seen later), the first arrival for the horizontal.

From the above assumption, the multipath corresponds to the 9, 16, 24, 332 symbols delay, respectively. The fourth delayed signal degrades 10dB compared to the direct one, so it has less ISI effects than the others. And the longer the horizontal distance, the more the ISI effects.

To combat the intersymbol interference effects and acquire ideal results in UWA channel, equalization can be employed.

RLS or LMS are both based on the minimization of MSE, recursively.

For LMS algorithm the update coefficients are recursively adjusted to minimize MSE, and $\Delta$ is the step size parameter which controls the rate of adjustment. The convergence rate of the LMS algorithm is slow due to the fact, that there is only a single parameter, which controls the rate of adaptation. And if $\Delta$ selects too large, it will cause instability.

A faster converging algorithm is obtained if we employ a recursive least square(RLS) criterion for adjustment of the equalizer coefficients. The exponential weighting factor $\lambda$ is selected to be in the range $0<\lambda\leq 1$( In simulation chosen to be 0.99). It provides a fading memory in the estimation of the optimum equalizer coefficients.

Fig.4.3 ~ Fig.4.6 illustrates a comparison of the effects of the RLS and LMS algorithms adopted in adaptive equalizers to remove the ISI. We note that the equalizers both performance well, and RLS linear equalizer shows better effects than LMS equalizer in compensating for ISI. And the difference in the convergence rate is very significant.

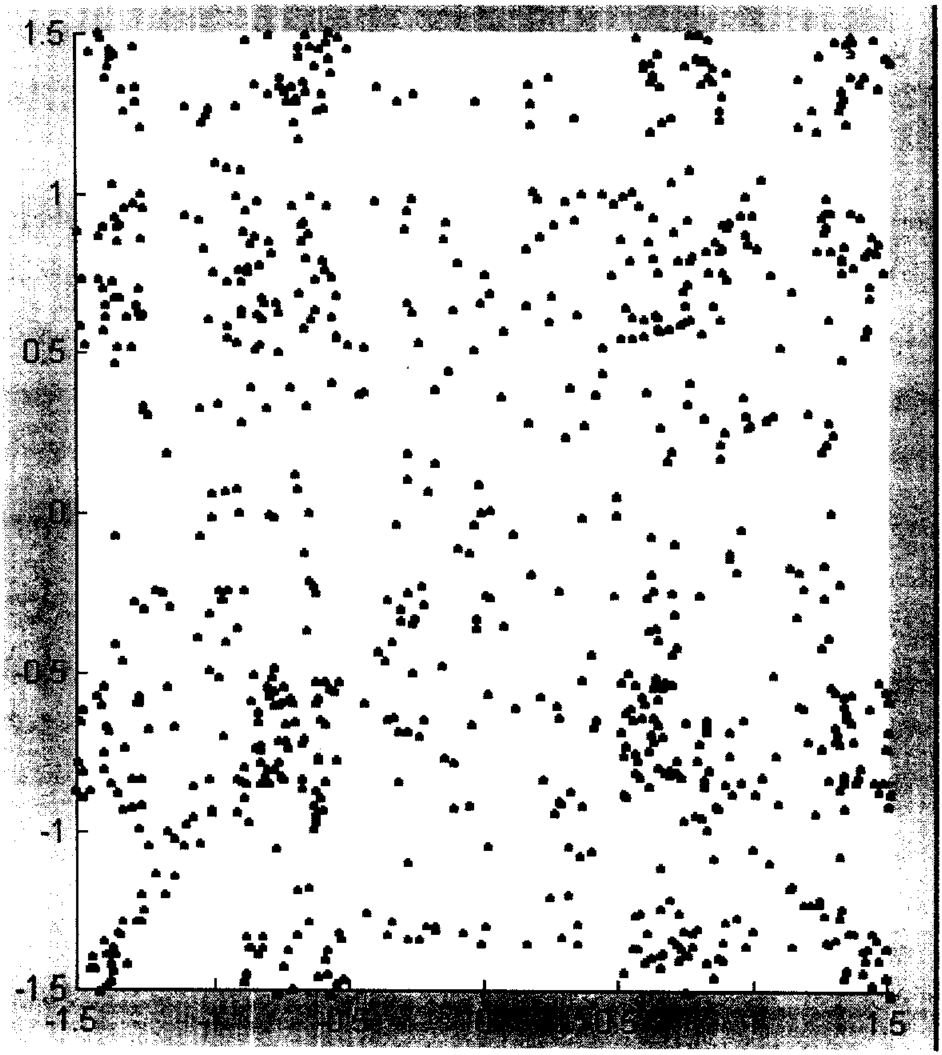



Fig. 4.2. Equalizer Input (Vertical channel)

Scatter plots after equalization

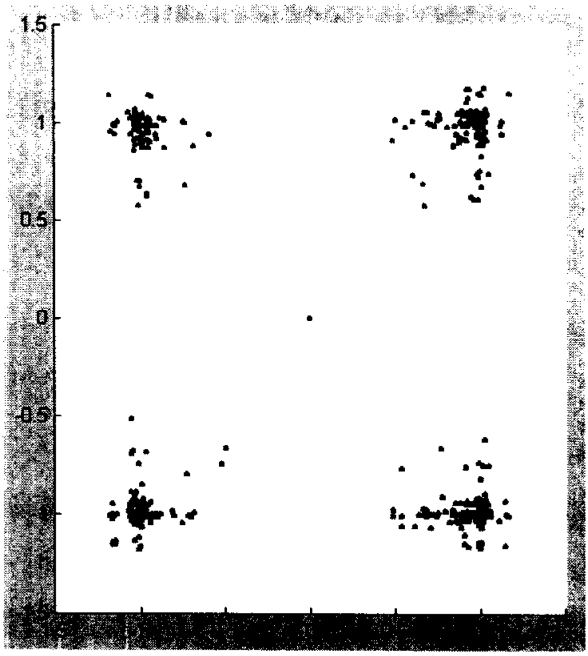


Fig.4.3 LMS linear equalizer output
Vertical channel (error: 0/1000s)

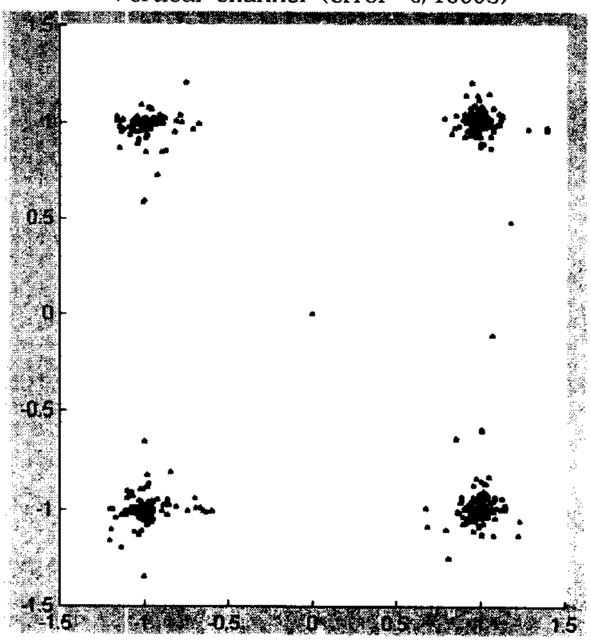
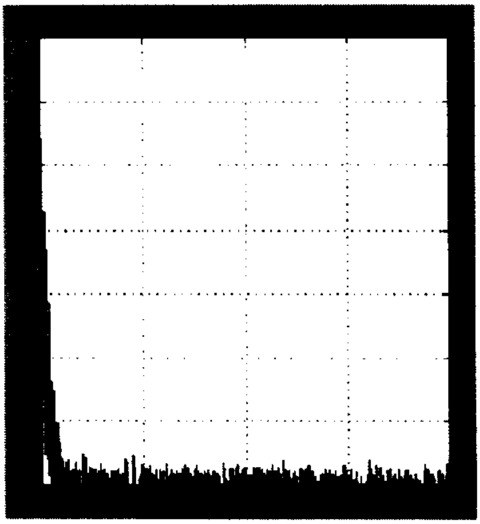
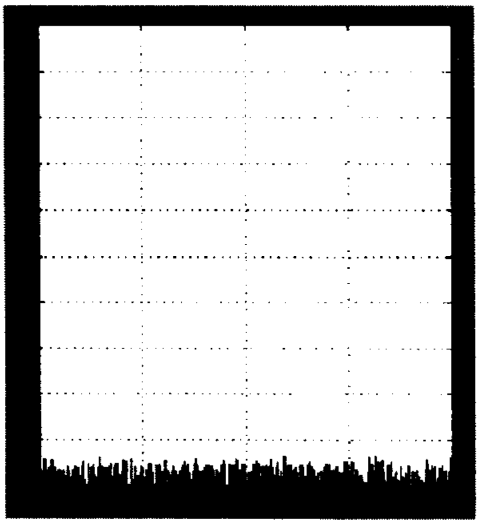


Fig. 4.4 RLS linear equalizer output
Vertical channel(error: 0/1000s)



Time (msec)

Fig.4.5 MSE of LMS linear equalizer (Vertical channel)



Time (msec)

Fig. 4.6 MSE of RLS linear equalizer (Vertical channel)

As shown in figure4.5~4.6, convergence rate comparison can be given as follows:

For RLS, MSE reduces 14dB in 5msec

For LMS, MSE reduces 10dB in 20msec

RLS has faster convergence and better tracking ability than LMS.

As mentioned above, the two algorithms expressions proves that RLS has more computational complexity , which is one order higher than LMS.

For horizontal channel, the horizontal distance between the source and the receiver is 500m. Their delay times and correspondent reflection coefficients are:

Delay_1 Time 0.74663 msec

Loss_1 0.41089

Delay_2 Time 1.2696msec

Loss_2 -0.77373

Delay_3 Time 2.0917 msec

Loss_3 -0.31785

Delay_4 Time  47.0842 msec

Loss_4  -0.2808

The severity of the ISI is directly related to the spectral characteristics of the channel, and not necessarily to the time span of the ISI. For vertical channel, the intersymbol interference (ISI) is not as strong as the horizontal one, which is shown in Figure4.7, the frequency response of the horizontal channel has spectral nulls (Amplitude degrades over 30dB is considered as spectral null) in some frequency band, whereas this doesn't happen in the case of vertical channel, thus the linear equalizer is used to compensate for the ISI. The equalizer's output shows that the ISI has been eliminated using the linear equalizers.

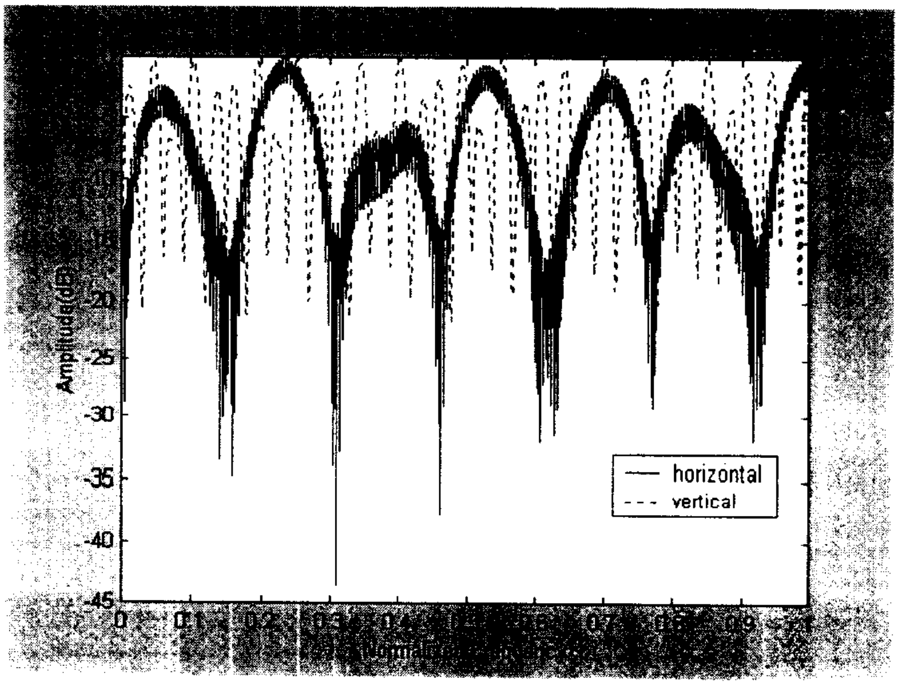



Fig. 4.7 Amplitude spectra for the horizontal and vertical channels

For the transmission over horizontal channel, DFE and linear equalizer (For the purpose of comparison, employ the linear equalizer) will be employed. For low computational complexity, only the LMS algorithm is adopted to remove the ISI during the data transmission over the horizontal channel. On the next plots, we display the equalizer input and output constellations, and we can conclude that a large part of ISI has been removed, using the strategy. As shown in Fig4.8~4.13, for the horizontal channel with spectral nulls the nonlinear equalizer illustrates its superiority. For different SNR, employing DFE there are less errors occuring than linear equalizer(feedforward filter).

Equalizer output for SNR 5dB

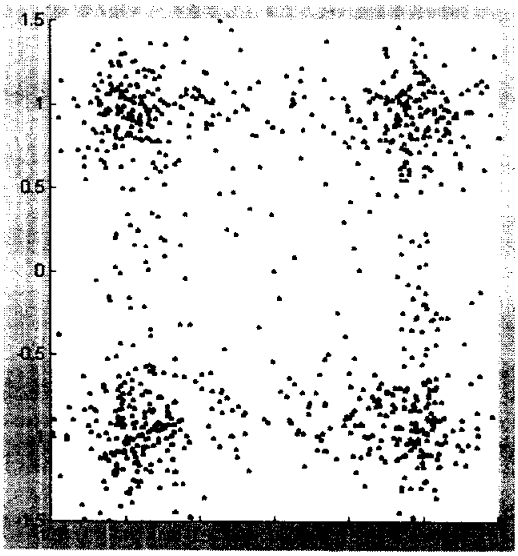



Fig. 4.8 LMS linear equalizer output
Horizontal channel( SNR 5dB error:
91s/1000s)

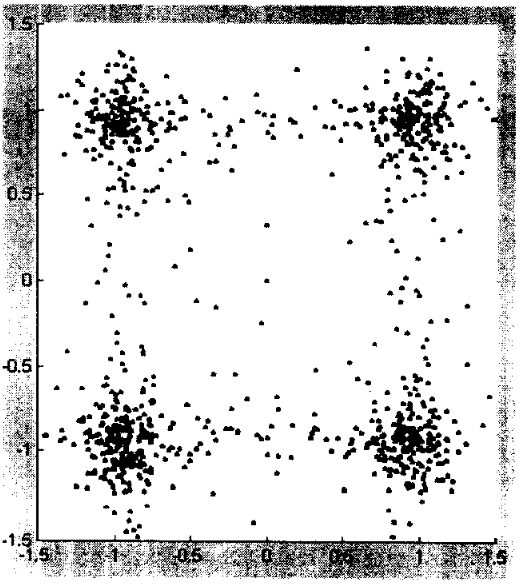



Fig. 4.9 LMS DFE output
Horizontal channel(SNR 5dB error:
55s/1000s)

Equalizer output for SNR 10dB

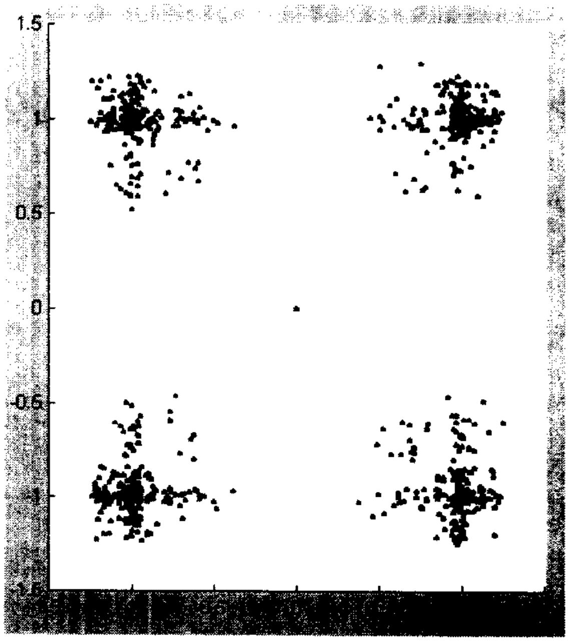



Fig. 4.10 LMS linear equalizer output
Horizontal channel( SNR 10dB error:


Fig. 4.11 LMS DFE output
Horizontal channel( SNR 10dB error:
4s/1000s)

Equalizer output for SNR 15dB

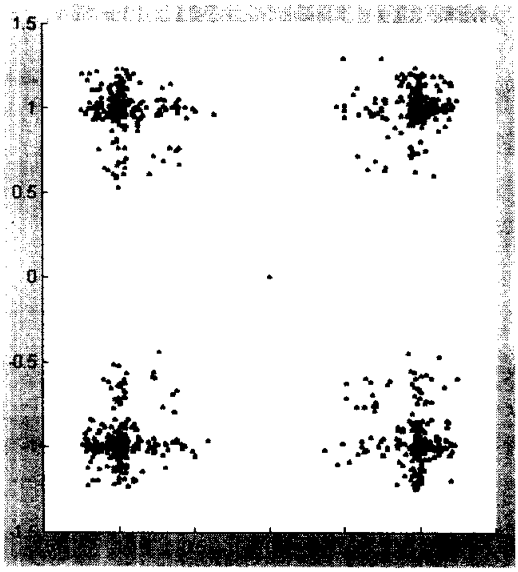



Fig. 4.12 LMS linear equalizer output
Horizontal channel( SNR 15dB error:
8s/1000s)

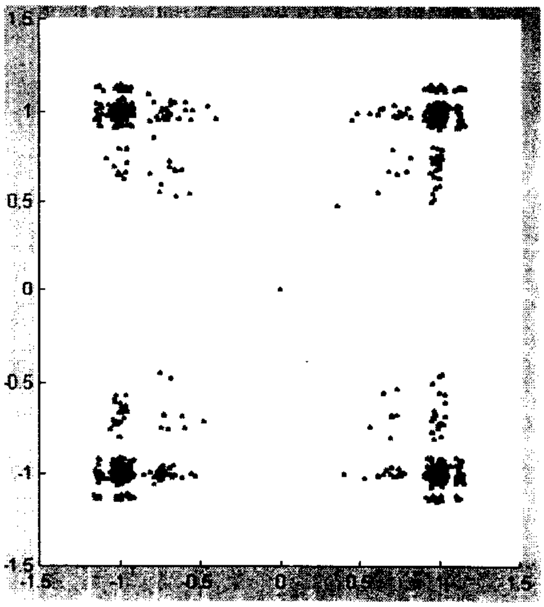
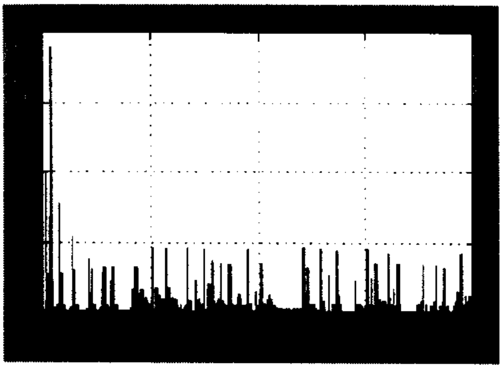
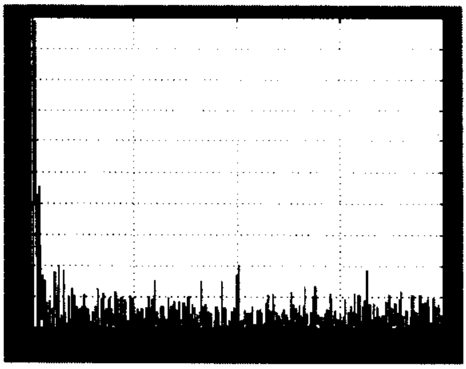



Fig. 4.13 LMS DFE output
Horizontal channel( SNR 15dB error:
0s/1000s)



Time (msec)

Fig. 4.14 Squared error using LMS linear equalizer (SNR 15dB)



Time (msec)

Fig. 4.15 Squared error using LMS DFE equalizer (SNR 15dB)

The comparison of feedforward filter and DFE are as follows:

Feedforward filter convergence rate: MSE degrades 10dB in 50msec

Decision feedback equalizer convergence rate: MSE degrades 10dB in 20msec

| SNR(dB) | BER1(FFF) | BER2(DFE) |
|---|---|---|
| 5 | 0.091 | 0.055 |
| 10 | 0.012 | 0.004 |
| 15 | 0.008 | 0 |
| 20 | 0.007 | 0 |

## Ⅳ-2. Simulation Results

The above scatter plots represent the output signal obtained at the receiver, which shows no error occurs out of 2000 bits in the date transmission since employing DFE, while SNR is not less than 15dB. However, feedforward filter(LMS linear equalizer) can't completely compensate for ISI effects, even in the case, while SNR is larger than 15dB, the bit error rate is in the order of $10^{-3}$. The basic problem for linear equalizer is that linear equalizer only processes the input signal samples, and the noise always limits the performance. However, a decision feedback equalizer is a nonlinear equalizer that uses previous decisions to eliminate the ISI caused by the previously detected symbols on the current symbol to be detected, and reduces the noise simultaneously.

Horizontal channel transmission exhibits stronger ISI effects than vertical channel, therefore nonlinear equalizer, DFE are needed in the horizontal channel. And LMS algorithm adopted in removing the ISI effects shows robustness in tracking a non-stationary environment with its simplicity, though it suffers from slow convergence rate. Consequently the LMS algorithm as an effective way in compensating for ISI is widely adopted in equalization and is being further studied and improved its performance.

# V. Conclusion

In the past the noncoherent systems had been in use due to the time varying and Doppler phenomenon characteristics of UWA channel, which causes severe intersymbol interference and rapid phase variations makes coherent demodulation very difficult. However, for the bandlimited underwater channel, to efficiently make use of the channel and transmitting date, the phase coherent signal detection has been developed and the corresponding methods compensating for the ISI caused by multipath has been demonstrated, such as equalizer. Among the suboptimum equalizers, the one which minimizes the mean squared error proves optimizing. Thus LMS and RLS algorithms are adopted in adaptive equalization. And though RLS has more rapid convergence and better tracking ability, LMS with its computation simplicity has its superiority. But for the channel with spectral nulls DFE should be employed to remove ISI. The resulting adaptive algorithm has been proved very suitable for tracking the variations of the channel.

Since the bandwidth efficient modulation methods offer the possibility of achieving high data rates on the UWA channels, an appropriate receiver which employs the methods of joint equalization and synchronization should be given the special attention in the UWA system design.

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## 국문초록

# 수중음향통신용의 QPSK 시스템 설계

*정보통신학과 임춘단*

*지도교수 윤종락*

최근 수중음향통신시스템은 군사목적에서 그 응용범위 넓혀 해양 스포츠 및 해양활동에 필요한 시스템으로 자리잡고 있다. 수중음향통신시스템은 다중경로와 수중채널의 특징에 의해 성능이 좌우된다. 수중음향채널의 대표적인 특징은 시간지연, 도플러효과, 패이딩의 특성을 가지고 있다. 또한 다중경로는 수직채널에 비해 수평채널 상대적으로 다중경로에 의한 영향을 많이 받고, 심볼간섭은 전송거리와 전송율에 영향을 받는다. 또한 송수신기의 이동에 의한 도플러 현상으로 인해 위상이 변화하기 때문에 일반적으로 동기방식이 아닌 비동기방식기법인 FSK 와 DPSK변복조 방식이 사용된다.

수중음향채널의 제한조건을 극복하기 위해 수중채널의 대역폭과 전송파워의 조절이 용이한 변복조 기술이 필요하게 되었으며, 변복조 기술 및 등화기법의 발전으로 동기위상변조기술인 M-PSK 기법이 사용되고 있다.

본 논문에서는 QPSK방식의 수중음향통신 시스템을 설계하여, 다중경로에 의한 심볼간섭을 억제하기 위해 적응등화기를 사용하여 수직채널과 수평채널에서의 영향 실험해 보았다. 결과적으로 수중음향통신시스템에서 적응등화기 사용시 전송속도의 증가와 심볼간섭을 효과적임을 알 수 있었다.