THE JOURNAL OF THE ACOUSTICAL SOCIETY OF KOREA VOL.24, NO.3E 2005. 9 pp. 99~108

# Multipath Compensation for BPSK Underwater Acoustic Communication

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(Received Jun 13 2005; accepted July 8 2005)

**Abstract**

To investigate the equalizer performance in underwater acoustic communication in the presence of intersymbol interference (ISI) due to multipath, computer simulations are carried out in discrete multipath shallow water channels for three different horizontal ranges. For the purpose of computation simplicity, least mean square (LMS) algorithm is adopted both in linear equalizer and nonlinear equalizer, decision feedback equalizer (DFE) to cancel out ISI effects. Binary phase shift keying (BPSK) signals have been transmitted with high data rate of 2000bps through the use of equalization technique. The results demonstrate that equalization is an efficient way to achieve high transmission data rate in the shallow water channel.

***Keywords***: *Underwater acoustic communication, Equalizer, Inter-symbol interference, Bit error rate, Multipath*

## I. Introduction

Underwater acoustic communications has been developed rapidly in recent years, and more and more attentions is shifting from non-coherent modulation toward the phase coherent communication with the development of equalization and diversity techniques. The underwater acoustic channel is always band limited and reverberant which poses many obstacles to reliable, high-speed digital communications. The effects of reflection from the surface and the bottom of the sea give rise to multipath propagation. To mitigate the distortion caused by ISI due to multipath, adaptive equalizer based on LMS algorithm is employed in bandwidth efficiency coherent BPSK communication system[1-3].

To investigate the effectiveness of adaptive equalizer overcoming ISI caused by multipath, numerical experiment is conducted by employing LMS linear equalizer (LE) and nonlinear decision feedback equalizer (DFE) in shallow water acoustic channels.

This paper is organized as follows. In Section II the multipath model is presented, then followed by the description of the communication system, and the introduction of equalizer is given in Section III, performance results of simulation in Section IV. The final section draws a conclusion on the further study to be investigated.

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## II. Channel model and communication system

The impulse response of the equivalent lowpass system corresponding to band-limited carrier modulation scheme in a multipath underwater acoustic system is modelled as[4]

$$h_c(t) = \sum_{n=0} \alpha_n e^{-j2\pi f_c \tau_n} \delta(t-\tau_n) \quad (1)$$

where $f_c$ is carrier frequency, and $a_n$ and $\tau_n$ are the normalized amplitude and the propagation time difference of the signal received along the $n$th path to the direct path amplitude $a_0$ and the direct path propagation time $\tau_0$.

We assume the channel variations are slow compared with the signalling interval for the simplicity of channel model analysis such that the Doppler spread of the channel is much less than the signal bandwidth. The system configuration utilizing BPSK modulation is depicted in Fig. 1.

The transmitted binary data are first shaped to match the channel bandwidth for the ideal band-limited underwater acoustic system, then modulated by the carrier. The transmitted signal sequence of binary data with a bit interval $T_b$ is given as

$$s(t) = \sum_i s_i p(t - iT_b) \tag{2}$$

where $s_i$ is the $i$th transmitted symbol and corresponds to bit 1, -1, which producing a phase change of 90° or 180° in BPSK modulation. The $p(t)$ is a pulse whose shape influences the spectrum of the information-bearing transmitted signal and its duration is a bit interval $T_b$ in binary schemes. The raised cosine pulse is generally adopted as $p(t)$ and given as[5]

$$p(t) = \left(\frac{A}{2}\right)\left\{1 + \cos\frac{2\pi}{T_b}\left(t - \frac{T_b}{2}\right)\right\},\ 0 \le t < T_b \tag{3}$$

where A is the amplitude of the pulse and its spectrum $P(f)$ is given as

$$P(f) = \frac{AT_b}{2}\,\frac{\sin \pi f T_b}{\pi f T_b(1 - f^2 T_b^2)}\,e^{-j\pi f T_b} \tag{4}$$

In this case the effective signal bandwidth $W_s$ is given to be $2/T_b$ and therefore we need a channel bandwidth $W_c$ of $2/T_b$ or larger to transmit a information sequence in ideal channel.

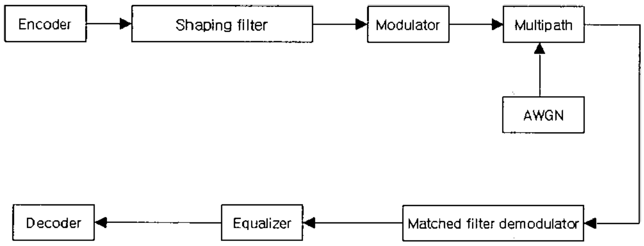

Fig. 1. Block diagram of communication system.

The equivalent low pass response of the raised pulse $p(t)$ in multipath underwater acoustic channel is given by convolution integral of (1) and (3).

$$h_p(t) = h_c(t) * p(t) \tag{5}$$

In order to characterize the multipath fading and figure out its effect on the transmitted signal distortion, root mean square (RMS) delay spread should be examined, which exhibits the time-extent nature of time-dispersive multipath channel. The RMS delay spread $\tau_s$ is defined as

$$\tau_s = \sqrt{\overline{\tau^2} - (\overline{\tau})^2} \tag{6}$$

where average delay $\overline{\tau}$ is expressed as

$$\overline{\tau} = \frac{\sum_k P(\tau_k)\tau_k}{\sum_k P(\tau_k)} \tag{7}$$

where $P(\tau_k)$ is the power density for the $k$ th path, and $\overline{\tau^2}$ is given as

$$\overline{\tau^2} = \frac{\sum_k P(\tau_k)\tau_k^2}{\sum_k P(\tau_k)} \tag{8}$$

RMS delay spread can be interpreted in the frequency domain, thus coherence bandwidth $B_c$, the range of frequencies over which the channel may be considered to be flat, can be given as[6]

$$B_c \approx \frac{1}{\tau_s} \tag{9}$$

Therefore, if the coherence bandwidth $B_c$ is less than the signal bandwidth $W_s$, then the channel is a frequency selective and the received signal is distorted. In this case, equalization should be adopted with this ISI problem encountered in multipath fading.

If (2) is applied to (1) then the equivalent low pass channel output of message signal is given as

$$
\begin{aligned}
r_c(t) &= s(t) * h_c(t) + W(t) \\
&= \sum_i \sum_n \alpha_n e^{-j2\pi f_c \tau_n} s_i p(t - \tau_n - iT_b) + W(t) \qquad (10)
\end{aligned}
$$

Therefore the channel with the multipath and white Gaussian noise $W(t)$ distorts the transmitted signal in amplitude and delay so that ISI is induced. The amount of ISI depends on the multipath nature such as the normalized amplitude $\alpha_n$ and the propagation time difference $\tau_n$ to the direct path.

At the receiver, the receiver filter $h_R(t)$ is used for limiting the noise components outside the signal bandwidth. For the ideal transmission channel with no multipath delay and non-fading, the spectrum of the output of the receiving filter is given as

$$R(f) = P(f)H_r(f) \qquad (11)$$

If we adopt the receiver filter as matched filter, then $H_r(f)$ and $h_r(t)$ are represented as

$$H_r(f) = P(f) * e^{-j2\pi f T_b} \qquad (12)$$

$$h_r(t) = p(T_b - t) \qquad (13)$$

and the demodulated output of the raised cosine pulse $p(t)$ is given as

$$
\begin{aligned}
h_{pm}(t) &= h_p(t) * h_r(t) \\
&= h_c(t) * p(t) * p(T_b - t) \\
&= \sum_{n=1} \alpha_n e^{-j2\pi f_c \tau_n} p(t - \tau_n) * p(T_b - t) \\
&= \sum_{n=1} \alpha_n e^{-j2\pi f_c \tau_n} R_{pp}(t - \tau_n - T_b) \qquad (14)
\end{aligned}
$$

The demodulated output of the received binary data signal can be expressed as

$$
\begin{aligned}
r_{sm}(t) &= r_c(t) * h_r(t) \\
&= \left(\sum_i \sum_n \alpha_n e^{-j2\pi f_c \tau_n} s_i p(t - \tau_n - iT_b) + W(t)\right) * p(T_b - t) \\
&= \sum_i \sum_n s_i \alpha_n e^{-j2\pi f_c \tau} R_{pp}(t - \tau_n - (i+1)T_b) + W(t) * p(T_b - t) \\
&= s_k \alpha_1 R_{pp}(t - (k+1)T_b) + W_p(t) + \sum_{i \neq k} \sum_{n \neq 1} s_i \alpha_n e^{-j2\pi f_c \tau_n} R_{pp}(t - \tau_n - (i+1)T_b) \qquad (15)
\end{aligned}
$$

where $W_p(t) = W(t) * p(T_b - t)$ is filtered noise by the receiver filter. The demodulated output is integrated and sampled in every symbol interval $T_b$ and the direct path signal components can be obtained as

$$r_{kd} = s_k \alpha_1 \int_{kT_b}^{(k+1)T_b} R_{pp}(t - (k+1)T_b) dt \qquad (16)$$

and noise components denoted as

$$W_{kp} = \int_{kT_b}^{kT_b + T_b} W_p(t) dt \qquad (17)$$

In multipath channel,

$$
\begin{aligned}
r_k &= r_{kd} + W_{kp} + \sum_{i \neq k} \sum_{n \neq 1} s_i \alpha_n e^{j2\pi f_c \tau_n} \int_{T_b}^{(i+1)T_b} R_{pp}(t - \tau_n - (i+1)T_b) \\
&= r_{kd} + r_{km} + W_{kp} \qquad (18)
\end{aligned}
$$

where $r_{kd}$ is the wanted direct path signal, and $r_{km}$ is ISI due to the multipath. As shown in (18), the multipath ISI depends on the normalized amplitude and the propagation time difference of the $n$ th path.

## III. LMS linear and Nonlinear Equalizer

In order to compensate for the multipath ISI we adopt the equalizer to the demodulated output $r_k$. In the selection of an equalizer, we consider its ability to track channel characteristics changes and computation complexity. LMS is one of the most popular algorithms. Therefore LMS linear equalizer (LE) and nonlinear decision feedback equalizer are used to suppress ISI.

We assume there is no sampler delay, the demodulated signals is sampled in one symbol duration. The basic LMS algorithm in Fig. 2 is expressed as the following equations.

The update tap-weight $w(k)$ at time instant $kT_b$ is represented as

$$w(k) = w(k-1) + \mu e^*(k-1) r_{k-1} \qquad (19)$$

where $\mu$ is step-size which controls the convergence rate and excess mean square error. In other words, the larger step size gives the more rapid convergence rate, but the larger fluctuation

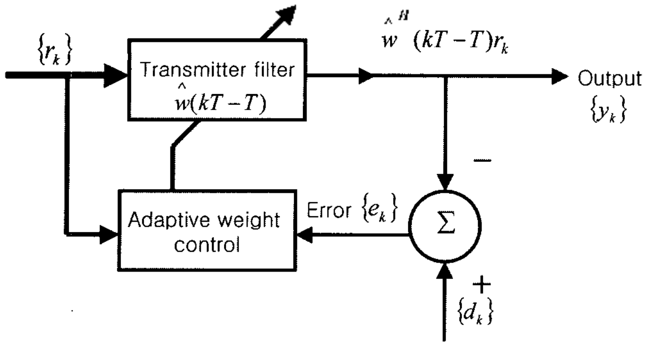


Fig. 2. Least mean square (LMS) structure.

of tap coefficients. Estimation error $e(k)$ and filter output $y(k)$ are given as

$$e(k) = d(k) - y(k) \tag{20}$$

$$y(k) = w^H(k) r_k \tag{21}$$

where $d(k)$ and $w^H(k)$ are the desired symbol, and the tap weight coefficient, respectively. The superscript $H$ denotes conjugate transpose, and $r_k$ is the tap input vector, i.e. the received demodulated binary data signal.

The tap weight coefficients of the equalizer are recursively adjusted to meet the criterion of minimizing the mean square error (MSE) with respect to the equalizer taps.

$$\min(E\mid e(k)\mid^2)$$

$$= \min(E\mid d(k) - y(k)\mid^2) \tag{22}$$

For the linear equalizer in Fig. 3a, filter output $y(k)$ is expressed as

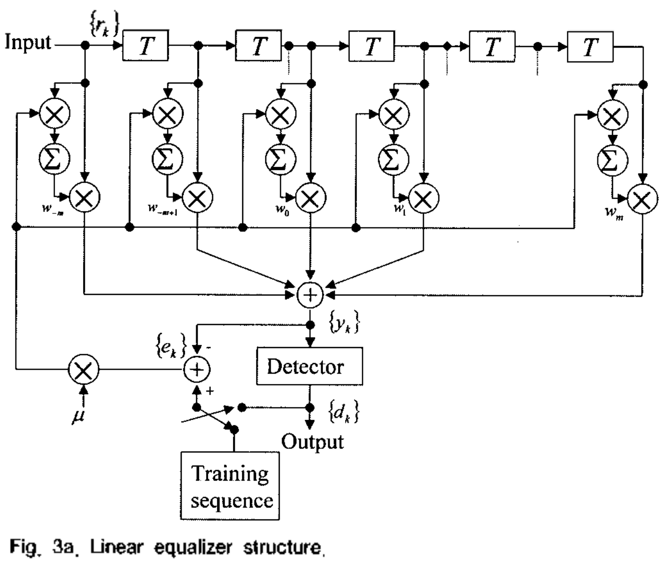


Fig. 3a. Linear equalizer structure.

$$y(k) = \sum_{j=-m}^{m} w_j r_{k-j} \tag{23}$$

where $w_j$ is tap weighting coefficients of the equalizer.

For the DFE shown in Fig. 3b, the estimate is different from that of linear equalizer and denoted as

$$y(k) = \sum_{j=-m1}^{0} w_{ff_j} r_{k-j} - \sum_{j=1}^{m2} w_{fb_j} d(k-j) \tag{24}$$

where $w_{ff_j}$ and $w_{fb_j}$ are the tap coefficients of feedforward and feedback filters.

The tap coefficients of feedforward and feedback filters are adjusted simultaneously, and the feedback filter is used to cancel out the part previously detected symbols. The input to the feedback filter is quantized signal in bit interval, thus DFE is a nonlinear equalizer. Compared with linear equalizer, DFE can not only remove ISI, but also operates on noiseless quantized levels[7].

As shown in Fig. 3a and Fig. 3b, the training sequence which is a given binary data, is first transmitted to adjust the tap weights initially, then switched to decision-directed mode in which decision symbols are compared to the estimate and gives the error signal under the assumption of the decisions on informations are correct.

## IV. Simulation Results

The simulation parameters to investigate the feasibility of the equalizer for use in underwater acoustic data transmission, are

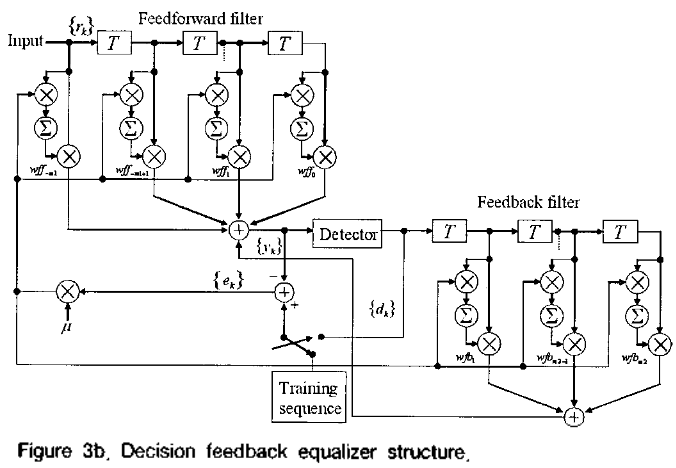


Figure 3b. Decision feedback equalizer structure.

Table 1. Parameters for Simulation.

| | | | |
|---|---|---|---|
| Carrier frequency | 20kHz | | |
| Symbol rate | 2kbps | | |
| Water depth (d) | 100m | | |
| Sound speed | 1500m/s | | |
| Transmitter depth | 5m | | |
| Receiver depth | 97m | | |
| Horizontal range ( $R$) | 10m | 600m | 1000m |

given as in Table I. The modulation format is BPSK. The channel model is characterized as shallow water channel since we have to consider multiple reflections from surface and bottom boundaries[8]. The ranges between the receiver and transmitter are 10m, 600m, 1000m, respectively and the depths of the receiver and the transmitter are kept fixed to be 5m and 97m, respectively. Reflection coefficient $r_b$ of the sandy bottom of the sea is assumed to be 0.41[9] and reflection coefficient $r_s$ of the sea surface with wave height of 0.05m, is calculated as 0.85. The Doppler spread and other problems are not addressed, and synchronization is assumed to be perfect. The paths with the less

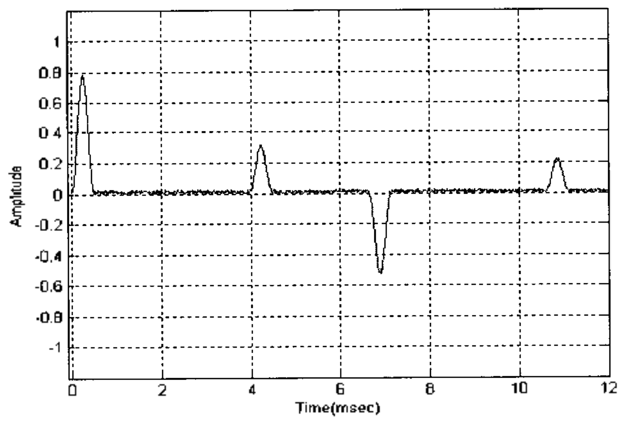


Fig. 4a. Channel response for R=10m.

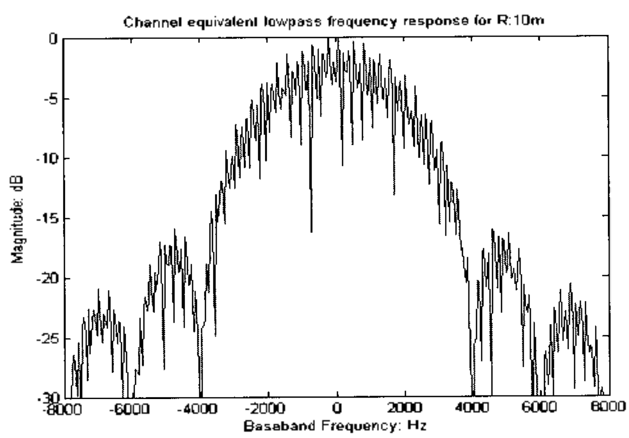


Fig. 4b. Frequency response for R=10m.

than -20dB of the normalized amplitude are neglected. Here ISI caused by multipath is concerned, thus signal to noise ratio (SNR) is considered as 30dB unless it's specified.

The impulse response tests are performed by transmitting the raised pulse $p(t)$ with the bandwidth of two times bit rate. The obtained impulse response and corresponding spectra of the equivalent lowpass channel for $R = 10\,\mathrm{m}$ whose range/depth ratio $\ll 1$ are shown in Fig. 4a and Fig. 4b, respectively. For this channel, the multipath time dispersion extends to 10.6msec and the RMS delay spread is 3.1ms, which results in the coherence bandwidth of about 330Hz, in other words, corresponding to approximately 165bps maximum transmission rate with no equalization[6]. Error free transmission is possible in the case that the signal bandwidth is less than the coherence bandwidth. Upper part of Fig. 5 in which signal bandwidth is about 300Hz (twice the bit rate 150bps), depicts error free transmission, but not in the lower part while transmission rate is chosen to be 300bps which occupies bandwidth of 600Hz larger than the coherence bandwidth. It's clear that this is a selective channel which causes ISI due to multipath if signal bandwidth is larger than the coherence bandwidth. In our numerical simulations to achieve error free transmission with high bit rate 2000bps in the presence of ISI, therefore compensation measure should be taken to remove ISI.

Fig. 4b illustrates that the ISI caused by multipath is not severe since the surface reflected path's amplitude attenuates a lot and leads to the significant multipath arrivals impose less strong ISI effects. At the receiver, the received signal is demodulated, sampled by $T_b$, then processed by the equalizer. First the prior known training sequence 1500 symbols are

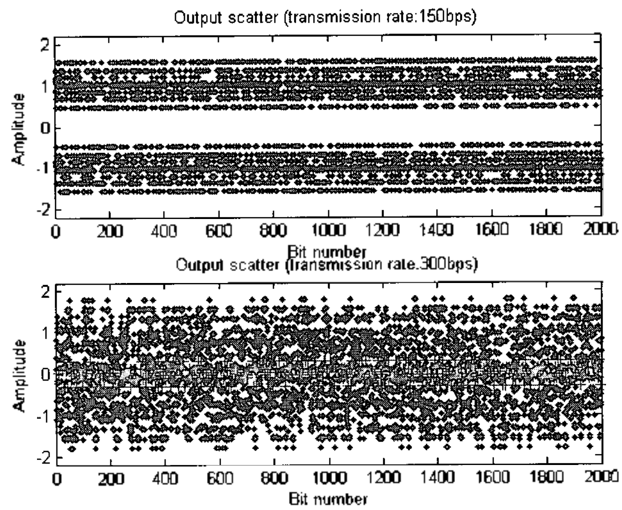


Fig. 5. Output scatter for R=10m.

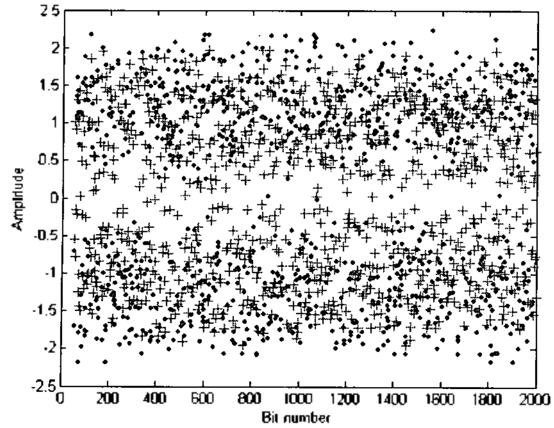

Fig. 6a. Scatter plot before equalization.

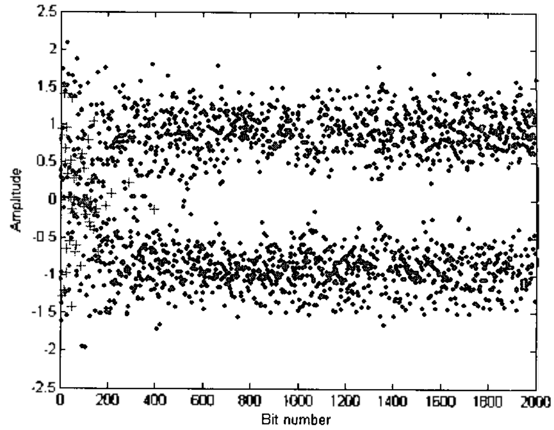

Fig. 6b. LMS LE output for R=10m.

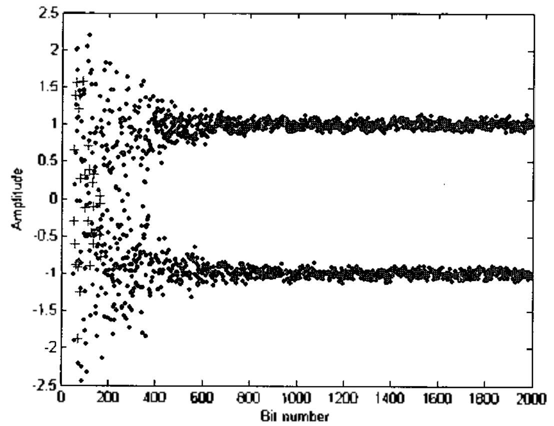

Fig. 6c. LMS DFE output for R=10m.

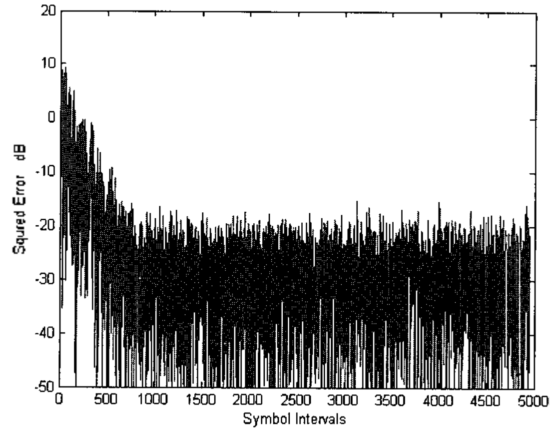

Fig. 6d. MSE after DFE for R=10m.

transmitted to establish convergence, after which the receiver is switched to decision-directed mode. Fig.6 ('+' means error, same in the following figures) exhibits the received demodulated signals before and after equalization and the obtained MSE with DFE processing. It's obvious that the raw signals with no equalization representing '1' or '-1' are not separated, however, the output of linear equalizer with 25 taps shows the two clusters representing two binary signals are separated well after 400 iterations, and there are no errors detected out of 10000bits in data transmission after training sequence (for the sake of display clearly, the number of bits is limited to 2000). The linear equalizer can be enough to track the channel characteristic for R=10m, for comparison we further employ DFE with 2 feedforward taps and 22 feedback taps. The output of linear

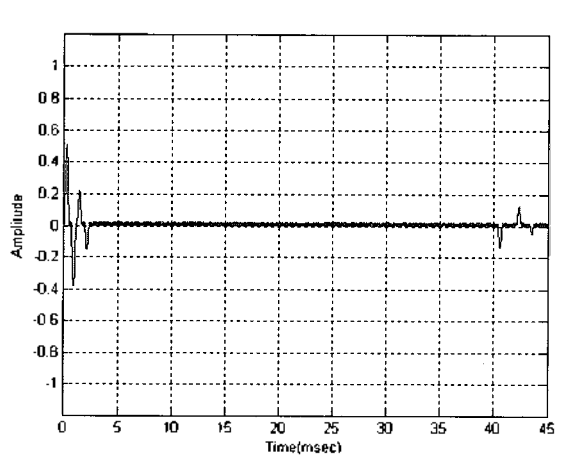

Fig. 7a. Channel response for R=600m.

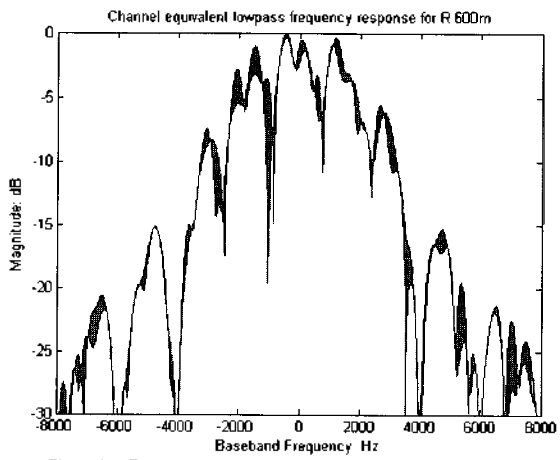

Fig. 7b. Frequency response for R=600m.

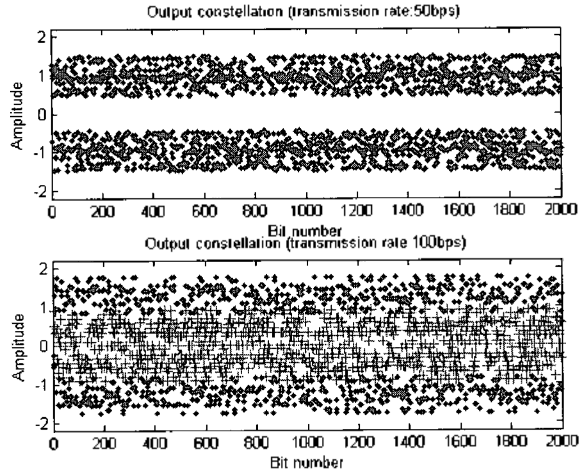


Fig. 8. Output scatter for R=600m.

equalizer is more scattered than that of DFE and in addition DFE has very small MSE in rapid convergence time. Consequently DFE outperforms linear equalizer.

Then we consider the channels whose transmission ranges/ water depth ratio $\gg 1$. At the range of 600m, the channel impulse and the spectra with equivalent lowpass frequency response characteristic are depicted in Fig. 7a and Fig. 7b. Fig. 7a exhibits that the secondary paths delay times compared to the direct path are larger than one symbol time and separated clearly. The RMS delay spread is computed as 9.8ms, thus, coherence bandwidth is approximated as 100Hz, namely 50bps maximum transmission rate with no compensation such as equalization. Fig. 8 illustrates the estimate of coherence bandwidth is suited for this channel, the upper plot exhibits that while data bit rate is 50Hz, the two clusters which representing '1' and '-1' are separated well and there is no errors occurred 1, whereas at 100bps twice maximum bit rate we can't attain error free communication.

Fig. 9 shows the input signals to the equalizer , the results of demodulated signals and MSE with equalization. Before equalization the signals can't be detected correctly due to the ISI. With the aid of the linear equalizer's (90 taps) reducing ISI effects, the received signals after a long training sequence about 1200 iterations can be separated into two clusters distinctly. For long time dispersion channel with spectral nulls, the increase of feedforward filter's coefficients leads to the noise enhancement. DFE (2 feedforward taps and 88 feedback taps) thus became a good choice in reducing the residual ISI effects. In addition DFE does not cause noise's increase since the feedback filter works on

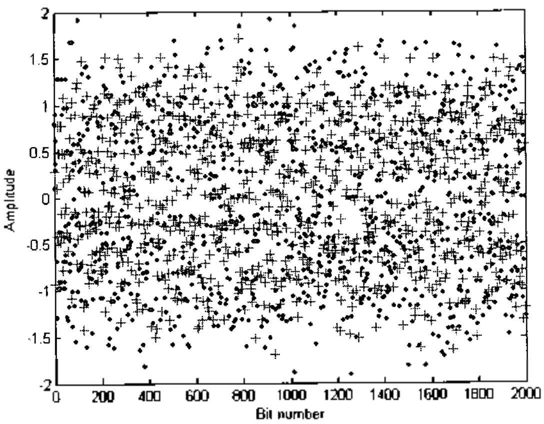


Fig. 9a. Scatter plot before equalization for R=600m.

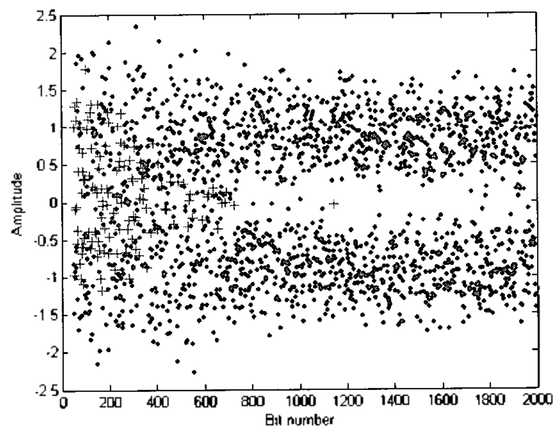


Fig. 9b. LMS LE output for R=600m.

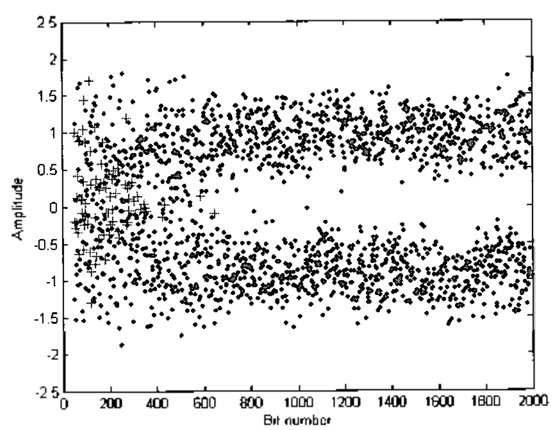


Fig. 9c. LMS DFE output for R=600m.

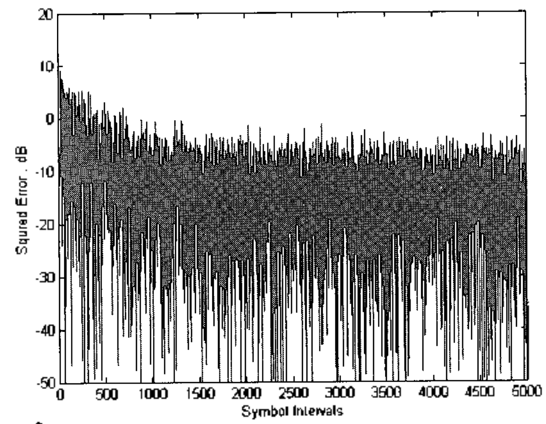


Fig. 9d. MSE after DFE for R=600m

noiseless quantized levels and the feedback output is free of channel noise. We see from Fig.9c with DFE processing the two clusters representing binary signals are separated clearly and we can obtain ideal transmission after 600 training symbols. It has shown the superior performance with a DFE.

At $R = 1000\,\mathrm{m}$ the channel response and the spectra characteristic are depicted in Fig.10a. and Fig.10b respectively, and RMS delay spread is 7.2ms, the approximate coherence bandwidth is 130Hz, hence when the transmission rate is chosen as 50bps, 100bps smaller and larger than half of coherence bandwidth respectively. In the former case no errors is detected, whereas in the latter case there are lots of errors occurred and can't obtain error free transmission (as shown in Fig. 11). Therefore it is proved the fact that the communication with high transmission rate larger than half of the coherence bandwidth without compensation method, such as equalizer and diversity is unachievable.

For the channel $R = 1000\,\mathrm{m}$, the delay times of the first three multipaths compared to its former path are less than one bit

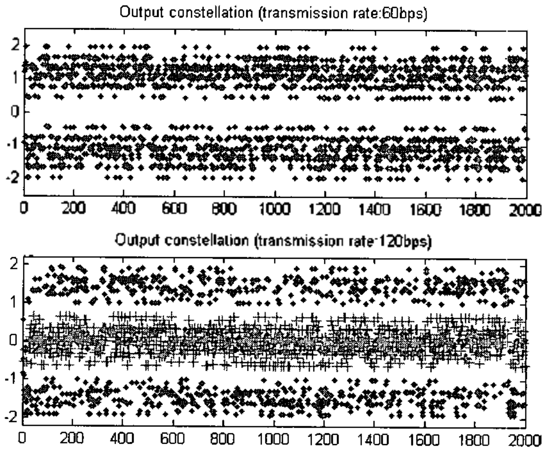

Fig. 11. Output scatter for R=1000m.

duration of 0.5msec. Especially compared to the direct path, the first multipath delay time are less than one symbol duration with negative amplitude, which results in self-destructive multipath. Comparing Fig. 7b and Fig. 10b, it is evidenced that the channel for $R = 1000\,\mathrm{m}$ possesses deeper spectral null, worse spectral characteristic and more severe ISI than the channel for $R = 600\,\mathrm{m}$. It reflects the fact that the severity of ISI is largely

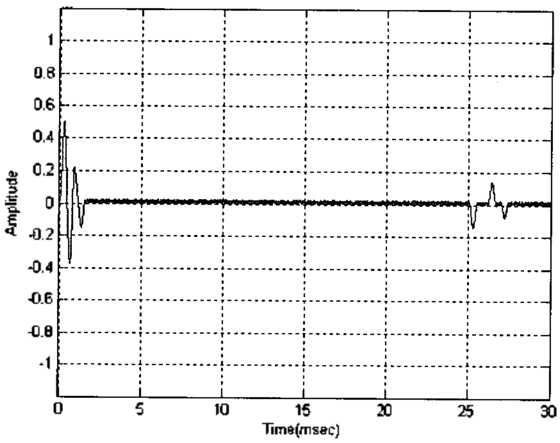

Fig. 10a. Channel response for R=1000m.

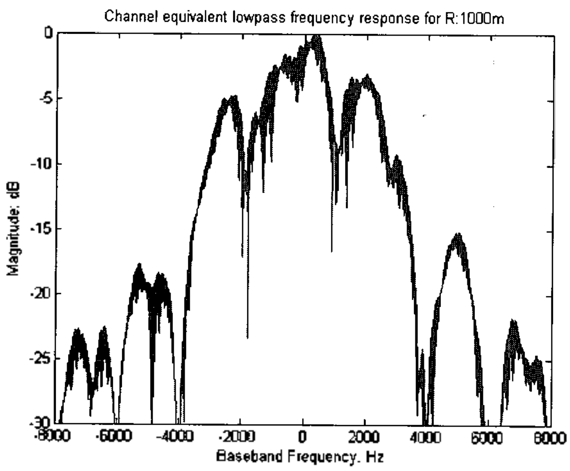

Fig. 10b. Frequency response for R=1000m.

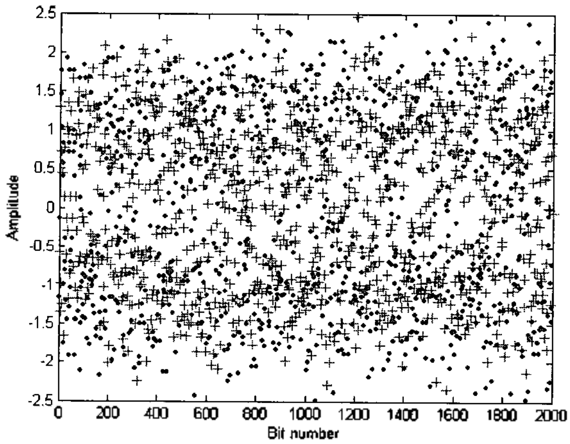

Fig. 12a. Output scatter before equalization for R=1000m.

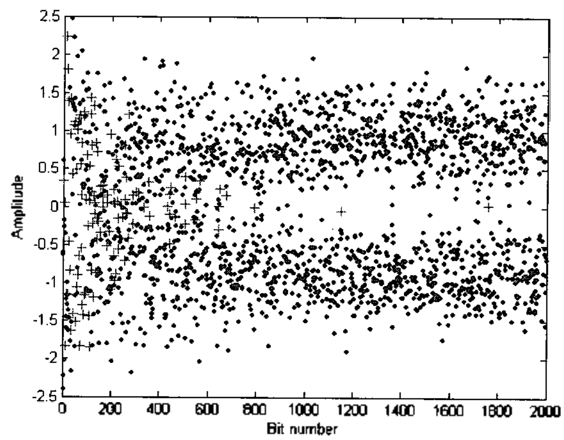

Fig. 12b. LMS LE output for R=1000m.

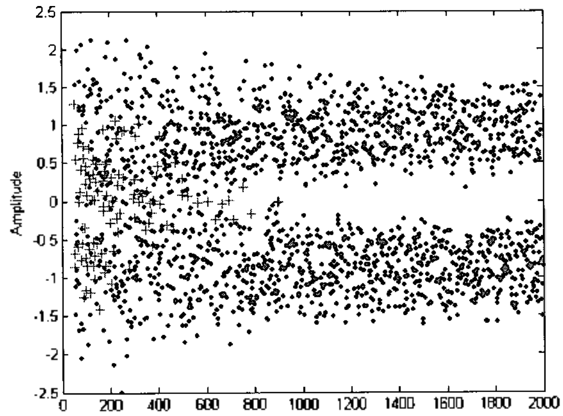

Figure 12c. LMS DFE output for R=1000m.

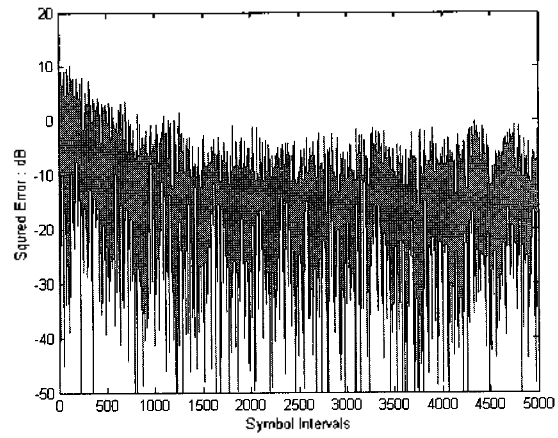

Figure 12d. Mean Squared Error after DFE for R=1000m.

dependent on the frequency response of transmission channel[4].

To illustrate the equalizers effects in reducing ISI of the channels at different transmission ranges, both the nonlinear equalizers DFE and linear equalizer based on LMS are employed to combat against ISI effects. Due to DFE's nonlinear characteristics the output of feedback is free of channel noise[8]. As shown in Fig.12, for $R = 1000\,\text{m}$, using LMS linear equalizer with 60 taps tracking the channel characteristic requires 1700 iterations, however DEF with 2 feedforward taps and 58 feedback taps just needs 700 training symbols. It's obvious that for the three different channels the equalizers take an important role in removing ISI. Moreover the DFE possessing nonlinear characteristic can further cancel the residual ISI completely for the channels with severe ISI such as deep spectral nulls without increasing the system's noise.

To compare the performances of DFE and linear equalizer at different values of SNR, test was performed at $R = 600\,\text{m}, 1000\,\text{m}$. Fig. 13 illustrates that for the channel R=600m in the case of employing linear equalizer there are much more errors than DFE. When SNR is not less than 16dB, the BER will satisfy the

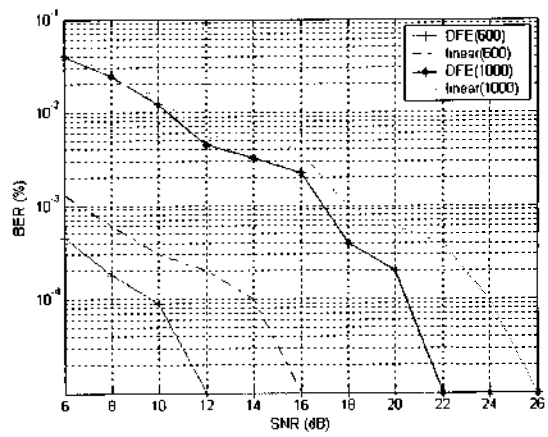

Figure 13. BER vs SNR for DFE and LE.

communication requirement: which is on the order of $10^{-5}$. However for R=1000m when SNR is less than 14dB DFE has similar performance compared with linear equalizer possibly due to the decision errors, when SNR larger than 14dB DFE shows superiority. Therefore, DFE shows better tracking ability than linear equalizer. DFE's advantage is that can not only cancel the ISI effects, but also can give rise to SNR enhancement.

## V. Conclusions

We concentrate on ISI caused by multipath and test the responses to the transmitting signal wave-formed as raised pulse over three multipath channels to know the channels characteristics, then through the numerical simulation by employing equalization find that linear and nonlinear LMS equalizer can be applied to combat ISI imposed on modulated signal over time dispersive channel.

For the channel with weak ISI the adaptive linear equalizer is effective in reduction of ISI, but for the channel with severe ISI employing adaptive nonlinear equalizer can remove ISI completely and obtain good performance. High data rates become possible in the case where channel variance is sufficiently slow to allow for channel tracking, the combination of adaptive equalization with more rapid convergence rate and synchronization techniques will be the subject of further study.

## Acknowledgement

This work was supported by Brain Busan 21 Project.

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