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Comparison of LMS and FDAF Algorithms in Equalization of Fading Channel

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***Abstract*—Wireless link in mobile cellular communication system is experienced by large and small scale fading. Since the link is Non Line of Sight (NLOS), therefore, severely affected by multipath fading. Adaptive equalizer is a widely used technique to neutralize the effect of multipath fading. In this paper, Frequency Domain Adaptive Filer (FDAF) and Least Mean Square (LMS) algorithms are used to combat the effect of multipath fading for 16-QAM and QPSK modulated wave and the results are shown using constellation diagram and bit error rate (BER). Finally, a comparison is made between two algorithms in contexts of process time, mean BER, variance of BER and circuit complexity.**

***Index Terms*—Mobile cellular communications, fading channel, channel equalization, adaptive equalizer.**

I. Introduction

In large scale propagation model, the separation between transmitter and receiver is very large and the link is almost line-of-sight (LOS), where only the effect of reflection, diffraction and scattering of electromagnetic (EM) wave is considered to measure the performance of a link theoretically. On the other hand, small scale fading is used to describe the rapid fluctuations of the amplitudes, phases or multi-path delays of a radio signal over a short period of time or traveled distance [1-3]. Fading is responsible for multipath propagation, speed of mobile, speed of surrounding objects and transmission bandwidth of the channel. Any passband signal is received as the sum of attenuated and delayed version of the original signal under a fading channel. Most of the cases, the impulse response of the channel is found time selective, i.e. the channel is better at selected times than other times [4,5]. To neutralize the delay spread of wireless channel, adaptive equalizer is used on passband signal [1,5].

In this paper, two well known adaptive algorithms: least mean square (LMS) algorithm and frequency domain adaptive filter (FDAF) algorithm are used in equalizer circuit at the receiving end of the wireless link. These algorithms are widely used in adaptive noise cancellation but adaptive equalization is a real time signal processing system, therefore, choice of optimum algorithm taking BER as a constraint is a complicated task. The choice of algorithm may change with changing modulation scheme and the amount of fading introduced by the channel.

In this paper, 16-QAM and QPSK modulation schemes are used to find scattered diagram and BER for the cases of with and without equalization. Here, the above mentioned two algorithms are compared considering process time of the algorithms, mean error, variance of error, and complexity of circuit implementation. Both algorithms are simpler compared to other adaptive algorithms, for example, Recursive Least Square (RLS) and Kalman Filter [6,7].

The paper is organized as follows. Section II gives the theoretical analysis of the adaptive equalizer and adaptive algorithm (LMS and FDAF), section III gives the results where constellation diagram of received symbols (16-QAM and QPSK) for the cases of with and without equalization, BER and a comparison of two algorithms are made. Finally section IV concludes the paper.

II. Theoretical Analysis

A. Adaptive Equalizer

If any signal $x(t)$ is multiplied by a constant k or delayed by an amount t_d, then the resultant signal becomes $k\,x(t-t_d)$. Such distortion is called linear distortion and is curable in communication system. For a linear distortion channel of Fig.1, the equalizer output can be written as $y(t)=k\,x(t-t_d)$. Taking Fourier transform, we have $Y(f)/X(f)=ke^{-j\omega t_d}=H(f)$. In presence of distorting channel of transfer function $H_c(f)$, the received signal can be made distortionless by incorporating an equalizer at the receiving end as shown in Fig. 1. The overall transfer function of the system becomes, $H_c(f)H_{eq}(f)=ke^{-j\omega t_d}$, where $|H(f)|$ is called the amplitude response function and need to be a constant over the entire bandwidth (BW) of the input signal, and $\theta(f)$ is the phase response need to be a linear function of frequency over the entire BW of the input signal for recovery transmitted symbol [8-10].

Let us consider the tapped delay FIR filter of Fig. 2, where the output signal is

$$y(t)=C_{-N}x(t)+C_{-N+1}x(t-\tau)+C_{-N+2}x(t-2\tau)$$
$$+\cdots+C_0x(t-N\tau)+\cdots+C_Nx(t-2N\tau).$$

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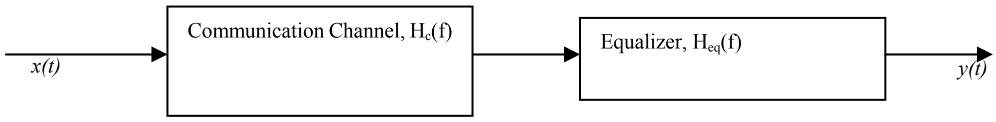


Fig. 1 Equalizer in cascade with the channel.

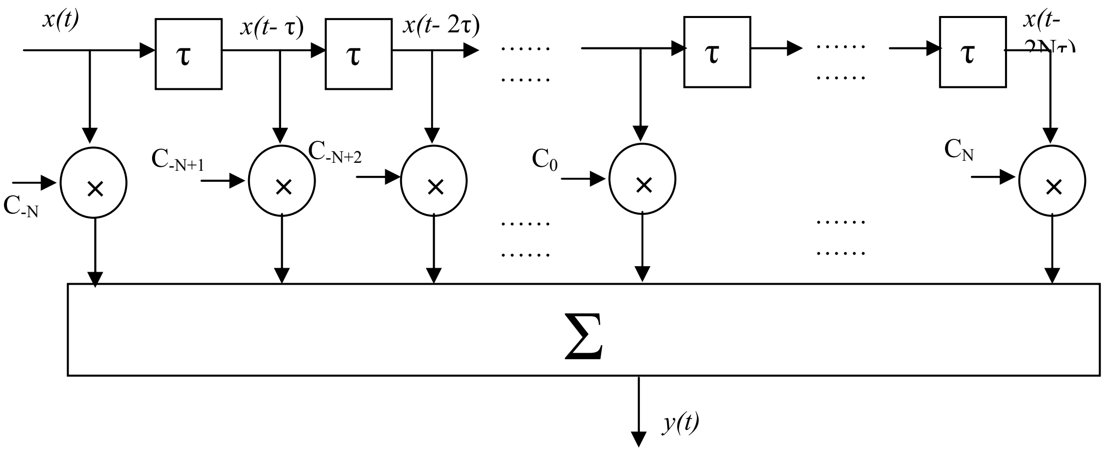


Fig. 2 Tapped delay line equalizer.

Let us consider the tapped delay FIR filter of Fig. 2, where the output signal is

$$y(t) = C_{-N}x(t) + C_{-N+1}x(t-\tau) + C_{-N+2}x(t-2\tau) + \cdots + C_0 x(t-N\tau) + \cdots + C_N x(t-2N\tau).$$

Taking Fourier transform, we have

$$Y(f) = C_{-N}X(f) + C_{-N+1}X(f)e^{-j\omega\tau} + C_{-N+2}X(f)e^{-j2\omega\tau} + \dots\ \dots\ \dots + C_0 X(f)e^{-jN\omega\tau} + \dots\ \dots\ \dots + C_N X(f)e^{-j2N\omega\tau},$$

and from which, we have

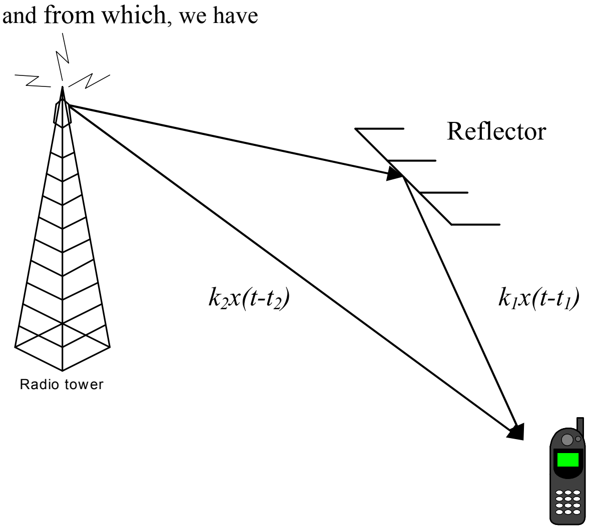


Fig. 3 Two path wave propagation.

$$H_{eq}(f) = \frac{Y(f)}{X(f)} = e^{-jN\omega\tau}\sum_{n=-N}^{N} C_n e^{-j2\pi f n\tau}. \qquad (1)$$

If relation $H_c(f)H_{eq}(f) = ke^{-j\omega t_d}$ can be satisfied by adjusting the number of delay blocks and weighting factors of Eq. (1), then distortion at receiving end can be eliminated [11].

Example

A simple multipath (here for simplicity, we consider only two paths) propagation scenario is shown in Fig. 3. Let us determine the weighting factors of 3 tapped delay equalizer for distortion less received signal. Here, the received signal is

$$y(t) = k_1 x(t-t_1) + k_2 x(t-t_2).$$

Thus, after Fourier transformation,

$$Y(f) = k_1 X(f)e^{-j\omega t_1} + k_2 X(f)e^{-j\omega t_2},$$

which gives

$$\frac{Y(f)}{X(f)} = H_c(f) = k_1 e^{-j\omega t_1} + k_2 e^{-j\omega t_2} = k_1 e^{-j\omega t_1}\left(1 + ke^{-j\omega t_0}\right),$$

where $k = k_2 / k_1$ and $t_0 = t_2 - t_1$.

We require, $H_{eq}(f)H_c(f) = k_1 e^{-j\omega t_1}$, from which we can write

$$H_{eq}(f) = \frac{k_1 e^{-j\omega t_1}}{H_c(f)} = \frac{k_1 e^{-j\omega t_1}}{k_1 e^{-j\omega t_1}\left(1 + ke^{-j\omega t_0}\right)} = \frac{1}{\left(1 + ke^{-j\omega t_0}\right)}.$$

After expanding,

$$H_{eq}(f) = 1 - ke^{-j\omega t_0} + k^2 e^{-j2\omega t_0} - \dots\ \dots\ \dots$$

or

$$H_{eq}(f) \approx 1 - ke^{-j\omega t_0} + k^2 e^{-j2\omega t_0}.$$

Therefore, $C_{-1} = 1$, $C_0 = -k$ and $C_1 = k^2$.

B. Frequency Domain Adaptive Filter (FDAF) technique

Let L is the length of the block and M is the length of tapped weight vector.

Let

$$\mathbf{A}_0 = \begin{bmatrix} u(kL) \\ u(kL+1) \\ \cdots \\ u(kL+L-1) \end{bmatrix},\ \mathbf{A}_1 = \begin{bmatrix} u(kL-1) \\ u(kL) \\ \cdots \\ u(kL+L-2) \end{bmatrix}, \ldots,$$

$$\mathbf{A}_{M-1} = \begin{bmatrix} u(kL-1) \\ u(kL) \\ \cdots \\ u(kL+L-2) \end{bmatrix}.$$

The data matrix can then be written as

$$\mathbf{A}(k) = [\mathbf{A_0} \quad \mathbf{A}_1 \quad \mathbf{A}_2 \quad \cdots \quad \mathbf{A}_{M-1}] = \begin{bmatrix} \mathbf{G}(kM) \\ \mathbf{G}(kM+1) \\ \ldots \\ \mathbf{G}(kM+L-1 \end{bmatrix}. \tag{2}$$

The matrix $\mathbf{A}(k)$ is an $L \times M$ matrix and length of the vector $\mathbf{G}^T(kM)$ is M. Let the weight vector be

$$\hat{\mathbf{W}}(k) = [w_0(k) \quad w_1(k) \quad w_2(k) \ldots w_{L-1}(k)]^T. \tag{3}$$

Output of the filter is

$$[y(kL) \quad y(kL+1) \quad y(kL+2) \ \ldots y(kL+L-1)]^T = \mathbf{A}(k).\hat{\mathbf{W}}(k). \tag{4}$$

For individual element,

$$y(kL) = G(kM).\hat{W}(k)$$

$$y(kL+1) = G(kM+1).\hat{W}(k)$$

$$\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots\ldots$$

$$y(kL+i) = G(kM+i).\hat{W}(k)$$

$$= [w_0(k) \quad w_1(k) \quad w_2(k) \ldots \ \ldots \ \ldots w_{L-1}(k)] \times \begin{bmatrix} u(kL+i) \\ u(kL+i-1) \\ u(kL+i-2) \\ \vdots \\ u(kL+i+M-1) \end{bmatrix}$$

$$= \sum_{j=0}^{M-1} w_j(k).u(kL+i-j). \tag{5}$$

Let the desired response of $(kL+i)-th$ element is $d(kL+i)$. Therefore, the error signal is

$$e(kL+i) = d(kL+i) - y(kL+i).$$

In matrix form,

$$\begin{bmatrix} d(kL) \\ d(kL+1) \\ d(kL+2) \\ \ldots \\ \ldots \\ \ldots \\ d(kL+L-1) \end{bmatrix} - \begin{bmatrix} y(kL) \\ y(kL+1) \\ y(kL+2) \\ \ldots \\ \ldots \\ \ldots \\ y(kL+L-1) \end{bmatrix} = \begin{bmatrix} e(kL) \\ e(kL+1) \\ e(kL+2) \\ \ldots \\ \ldots \\ \ldots \\ e(kL+L-1) \end{bmatrix} = \mathbf{e}(k). \tag{6}$$

The cross correlation vector is

$$\mathbf{\Phi}(k) = \mathbf{A}^T(k)\mathbf{e}(k). \tag{7}$$

The update equation of the weight vector can be written as

$$\hat{\mathbf{W}}(k+1) = \hat{\mathbf{W}}(k) + \mu\mathbf{\Phi}(k), \tag{8}$$

where μ is the step-size parameter.

Now using $\hat{\mathbf{W}}(k+1)$, we have

$$[y(kL+L) \ y(kL+L+1) \ y(kL+L+2) \ldots y(kL+2L-1)]^T = \mathbf{A}(k+1).\hat{\mathbf{W}}(k+1). \tag{9}$$

The updated error matrix is

$$\begin{bmatrix} d(kL+L) \\ d(kL+L+1) \\ d(kL+L+2) \\ \ldots \\ \ldots \\ \ldots \\ d(kL+2L-1) \end{bmatrix} - \begin{bmatrix} y(kL+L) \\ y(kL+L+1) \\ y(kL+L+2) \\ \ldots \\ \ldots \\ \ldots \\ y(kL+2L-1) \end{bmatrix} = \begin{bmatrix} e(kL+L) \\ e(kL+L+1) \\ e(kL+L+2) \\ \ldots \\ \ldots \\ \ldots \\ e(kL+2L-1) \end{bmatrix} = \mathbf{e}(k+1). \tag{10}$$

The updated cross correlation vector and the weight vector are respectively as

$$\mathbf{\Phi}(k+1) = \mathbf{A}^T(k+1)\mathbf{e}(k+1), \tag{11}$$

and

$$\hat{\mathbf{W}}(k+2) = \hat{\mathbf{W}}(k+1) + \mu\mathbf{\Phi}(k+1). \tag{12}$$

Similarly,

$$[y(kL+iL) \ y(kL+iL+1) \ y(kL+iL+2) \ldots y(kL+(i+1)L-1)]^T = \mathbf{A}(k+i).\hat{\mathbf{W}}(k+i). \tag{13}$$

Let

$$\begin{bmatrix} d(kL+iL) \\ d(kL+iL+1) \\ d(kL+iL+2) \\ \ldots \\ \ldots \\ \ldots \\ d(kL+(i+1)L-1) \end{bmatrix} - \begin{bmatrix} y(kL+iL) \\ y(kL+iL+1) \\ y(kL+iL+2) \\ \ldots \\ \ldots \\ \ldots \\ y(kL+(i+1)L-1) \end{bmatrix}$$

$$= \begin{bmatrix} e(kL+iL) \\ e(kL+iL+1) \\ e(kL+iL+2) \\ \ldots \\ \ldots \\ \ldots \\ e(kL+(i+1)L-1) \end{bmatrix} = \mathbf{e}(k+i) . \quad (14)$$

In generalized form,

$$\mathbf{\Phi}(k+i) = \mathbf{A}^T (k+i)\mathbf{e}(k+i) , \quad (15)$$

and

$$\hat{\mathbf{W}}(k+i+1) = \hat{\mathbf{W}}(k+i) + \mu\mathbf{\Phi}(k+i) . \quad (16)$$

In overlap-save method N=2M point FFT is used and the size of the filter of M tap weights.

Input signal in n domain is express as, $[(k-1)th \text{ block}, kth \text{ block}]$,

where the (k-1)-th block is

$$\begin{bmatrix} u(kL-L) & u(kL-L-1) & u(kL-L-2) & \ldots\ \ldots\ \ldots & u(kL-2M+1) \\ u(kL-L+1) & u(kL-L) & u(kL-L-1) & \ldots\ \ldots\ \ldots & u(kL-L-2M+2) \\ \ldots & \ldots & \ldots & \ldots\ \ldots\ \ldots & \ldots \\ u(kL-1) & u(kL-2) & u(kL-3) & \ldots\ \ldots\ \ldots & u(kL-2M) \end{bmatrix},$$

aAnd the k-th block is

$$\begin{bmatrix} u(kL) & u(kL-1) & u(kL-2) & \ldots\ \ldots\ \ldots & u(kL-2M+1) \\ u(kL+1) & u(kL) & u(kL-1) & \ldots\ \ldots\ \ldots & u(kL-2M+2) \\ \ldots & \ldots & \ldots & \ldots\ \ldots\ \ldots & \ldots \\ u(kL+M-1) & u(kL+M-2) & u(kL+M-3) & \ldots\ \ldots\ \ldots & u(kL-M) \end{bmatrix}.$$

Input signal in k domain,

$\mathbf{U}(k) = diag\left(FFT[(k-1)th \text{ block}, kth \text{ block}]\right)$; an $N \times N$ matrix. The initial weight in k domain is a $N \times 1$ vector:

$$\hat{\mathbf{W}}(k) = FFT\begin{bmatrix} \hat{\mathbf{w}}(k) \\ \mathbf{O} \end{bmatrix}. \quad (17)$$

Here $\hat{\mathbf{w}}(k)$ is tap-weight vector (in n domain) of length M and $\mathbf{O}$ is a null vector of length M.

Output signal of the filter in k domain is

$$\hat{\mathbf{y}}(k) = \text{Last } M \text{ elements of } IFFT[\mathbf{U}(k)\hat{\mathbf{W}}(k)] . \quad (18)$$

The desired response vector,

$$\mathbf{d}(k) = [d(kM), d(kM+1), \ldots\ d(kM+M-1]^T . \quad (19)$$

Error signal vector in n domain,

$$\begin{aligned} \mathbf{e}(k) &= [e(kM), e(kM+1), \ldots\ \ \ldots\ \ \ldots, e(kM+M-1]^T \\ &= \mathbf{d}(k) - \mathbf{y}(k). \end{aligned} \quad (20)$$

The error signal in the frequency domain is

$$\mathbf{E}(k) = FFT\begin{bmatrix} \mathbf{O} \\ \mathbf{e}(k) \end{bmatrix}. \quad (21)$$

The cross-correlation vector is

$$\mathbf{\varphi}(k) = \text{First } M \text{ elements of } IFFT\left[\mathbf{U}^H(k)\mathbf{E}(k)\right]. \quad (22)$$

Thus, the updated tap-weights [6,7,12] can be written as

$$\hat{\mathbf{W}}(k+1) = \hat{\mathbf{W}}(k) + \mu FFT\begin{bmatrix} \mathbf{\varphi}(k) \\ \mathbf{O} \end{bmatrix}. \quad (23)$$

C. LMS Algorithm

Least mean square filter is a type of digital adaptive filter where a desired filter is obtained by finding the filter coefficients that relate to producing the least mean squares of the error signal. In LMS, error estimation is determined by comparing the output of the linear filter in response to the input signal to the desired response. LMS then involves an adaptive process for automatic adjustment of the parameters of the filter according to the error estimation. The filtering process of LMS algorithm is performed by a transversal filter and the adaptive process is carried out by a weight control mechanism [1-4].

LMS algorithm can be summarized as follows based upon wide-sense stationary stochastic signal [8,13,14].

Parameters:

M = number of tapes, μ = step-size parameter;

where $0 < \mu < 2/MS_{\max}$, S_{max} is the maximum value of the power spectral density of the tap input $\mathbf{u}(n)$ and filter length M is moderate to large.

Initialization:

If prior knowledge of the tap-weight vector $\hat{\mathbf{w}}(n)$ is available, it is used to select an appropriate value of $\hat{\mathbf{w}}(0)$; otherwise $\hat{\mathbf{w}}(0) = \mathbf{0}$ is set.

Data:

Given $\mathbf{u}(n) = M - \text{by} - 1$ tap-input vector at time n

$= [\mathrm{u}((n), \mathrm{u}(n-1), \ldots, \mathrm{u}(n-M+1))]^T$,

$\mathbf{d}(n)$ = desired response at time $t = n$.

To be computed:

$\hat{\mathbf{w}}(n+1)$ = estimate of tap-weight vector at time n+1.

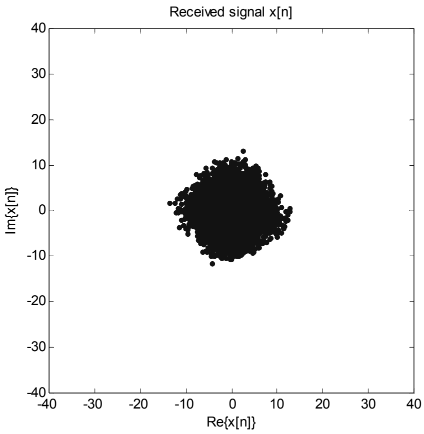
Computation:

$$\mathbf{e}(n) = \mathbf{d}(n) - \hat{w}^H(n+1)\mathbf{u}(n) , \quad (24)$$

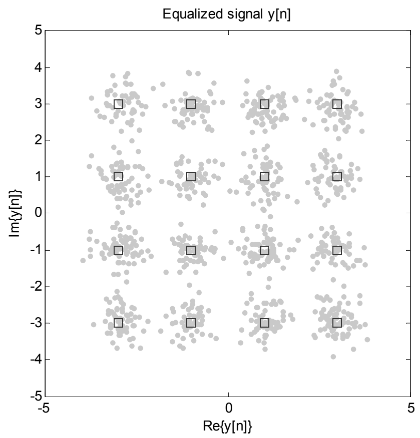
$$\hat{\mathbf{w}}(n+1) = \hat{\mathbf{w}}(n) + \mu\,\mathbf{u}(n)\mathbf{e}^*(n) . \quad (25)$$

III. Results and Discussions

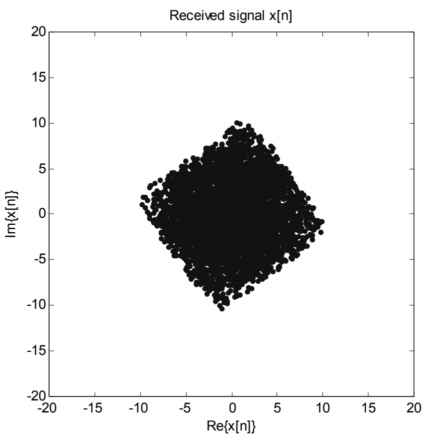
A simulation work is done generating 8192 symbols in a random manner. The constellation diagram is made for the symbols considering a multipath fading cannel of transfer function, H(z) = (z-0.4950 + j0.4950)(z- 0.7071 -j0.7071)/(z-1)(z+0.7). The constellation of 16-QAM and QPSK is shown in Fig. 4 (a)-(d) for the case of with and without equalization. The immense improvement in detection of symbols is visualized from the Fig. 4. Next, the simulation work is done to determine BER of both modulation schemes for the case of FDAF and LMS algorithms. In this case, 8192 symbols are generated in a random manner and BER is evaluated varying SNR form 0dB to 5dB. The simulation program is run 10 times and average BER is taken for each point of graphs of Fig. 5. In FDAF, BER has rapid fluctuation but for the case of LMS, BER is almost constant. The time taken for convergence is smaller in LMS compared to FDAF but once attaining convergence, the FDAF provides better performance as it is visualized from Fig. 5. The performance of QPSK is better than that of 16-QAM for both the algorithms and the phenomenon can be explained from signal space concept.



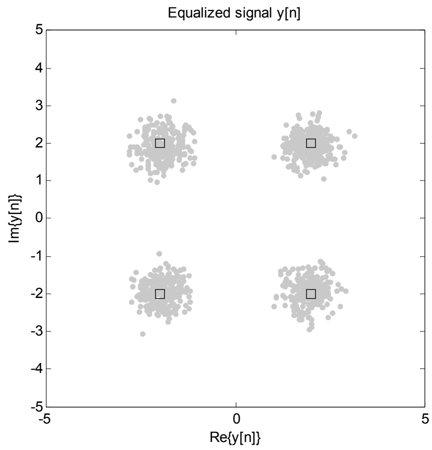
a) 16-QAM before equalization



(b) 16-QAM after equalization

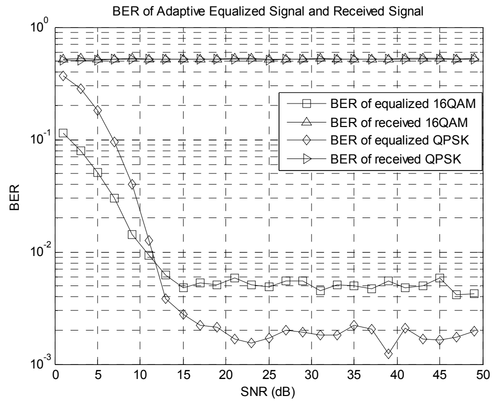


(c) QPSK before equalization

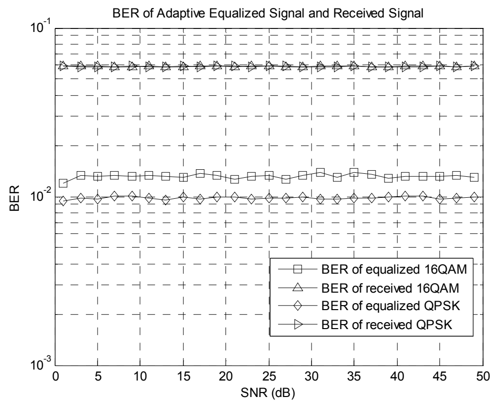


(d) QPSK after equalization

Fig. 4 Constellations of 16-QAM and QPSK before and after equalization.



(a) FDAF



(b) LMS

Fig.5 BER of received symbols.

TABLE I COMPARISON OF TWO ALGORITHMS

Algorithm	Process Time (s)	Mean Error	Variance	Circuit Complexity
LMS	30.0156	0.0688	1.4295e-007	Simple
FDAF	15.4219	0.0652	6.8392e-006	Complex

Finally, a comparison is made between two algorithms taking total 8192 symbols of 16-QAM. The parameters are selected for the comparison as: process time, mean BER, variance of BER and circuit complexity (Table I). LMS is better in context of quicker convergence visualized from the variance of Table I and also from the graph of Fig. 5(b). The circuit implementation of LMS algorithm is much easier compared to FDAF and can be understood from the theoretical analysis of Section II. On the other hand, FDAF is better in context of process time of operation and mean BER. In high speed data communication under fading wireless channel, we need an algorithm to process fading symbols before arrival of next symbols, hence FDAF algorithm is the best fit for such case.

IV. Conclusion

In this paper, two adaptive algorithms LMS and FDAF are compared in adaptive equalization of QPSK and 16-QAM signal under multipath fading. The results and discussions section of the paper shows the simulated result but in real life case the result will vary with location (where the experiment take place) and time because of small scale fading. The work can be extended incorporating other modulation schemes like MSK, GMSK, M-PSK etc. at the same time other adaptive algorithm like RLS and Kalman filter theory can also be examined for comparison.

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