

# EQUALIZATION AND INTERFERENCE CANCELLATION WITH MIMO THP FOR 10GBASE-T

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## ABSTRACT

Unlike 1000BASE-T system, the far-end crosstalk (FEXT) must be suppressed by at least 20 dB to meet the high speed transmission requirement for 10GBASE-T. Without FEXT cancellation, the average decision-point signal-to-noise ratio (DP-SNR) can degrade by 3 dB. This paper presents a multi-input multi-output Tomlinson-Harashima precoding (MIMO THP) technique to equalize the channel and to cancel the FEXT interference. Besides, the corresponding training method to deal with delay skew among channels and the arrangement of different step-size in least mean square (LMS) adaptive algorithm are proposed as well. Simulation results show that delay skew compensation and step-sizes arrangement can improve DP-SNR by 4.59 dB and 1.62 dB, respectively. The proposed MIMO THP architecture improves the DP-SNR by 2.75 dB than the tentative decision based approach.

**Index Terms**— MIMO system, MIMO TH precoding, Far-end crosstalk.

## 1. INTRODUCTION

The existing 10Gbps transmission techniques are developed with fiber media, such as 10GBASE-LX4. However, the cost of fiber and optical device is too high to be prevalent. The new Ethernet standard - 10GBASE-T, which will be the next generation of 1000BASE-T and provides 10Gbps data transmission over twisted-pair copper line, had finalized in 2006.

The full duplex transmission over 4-pair copper wires suffers from many impairments, such as insertion loss (IL), inter-symbol-interference (ISI), echo, near-end crosstalk (NEXT), far-end crosstalk (FEXT), and alien crosstalk. These impairments must be alleviated or they will degrade the achievable bit error rate (BER) seriously. One of the major different requirements between 1000BASE-T and 10GBASE-T is the FEXT crosstalk cancellation. In 1000BASE-T, the FEXT can be ignored. However, the IEEE 802.3an task group suggests that FEXT interference need to be suppressed by at least 20

dB and the slicer input decision-point signal-to-noise ratio (DP-SNR) at the receiver must be at least 23.4 dB to achieve a low BER of  $10^{-12}$ [1].

In [2], the authors proposed a blind startup algorithm to adapt the coefficients of FEXT canceler and equalizer jointly, where the input of FEXT canceler is the tentative decision of other distributing far end signal. However, 10GBASE-T adopts training sequences during the system startup. Besides, tentative decisions have to be reliable enough to prevent decision-error propagation [3]. The same arguments have been confirmed by simulation as indicated in [4] as well. In [5], the authors proposed a multi-dimensional linear equalizer to cancel both ISI and FEXT interferences at the receiver side. While in [4], the authors proposed a multi-dimensional non-linear equalization approach, which is a MIMO DFE (multi-input multi-output decision feedback equalizer) equalization technique. Both approaches treat FEXT interferences as useful signal rather than background noise.

However, it should notice that 10GBASE-T system adopts Tomlinson-Harashima precoding (TH precoding) [6, 7] instead of DFE as the equalization mechanism to mitigate error propagation problem. Hence, those existing FEXT cancellation approaches can not be applied directly. To overcome these difficulties, we present a new FEXT cancellation approach to for a system using TH precoding scheme.

Besides, few studies have been done on the effect of delay skew among pairs of wires. In [8], the authors demonstrated the minimum square error performance of the finite-length MIMO DFE to be particularly affected by the choice of decision delays used in the equalizer outputs, and they proposed a genetic algorithm to find out the optimal decision delay for each channel. The delay skew among channel was not considered in [4]. We propose a MIMO THP architecture to cancel the postcursor ISI and FEXT at the transmitter side, and a partial correlation-based method to determine the near-optimal decision delay. The delay skew and training issues are considered as well.

The rest of this paper is organized as follows. The simplified system architecture to be considered is given in Section 2. In Section 3, the method to estimate and compensate the de-

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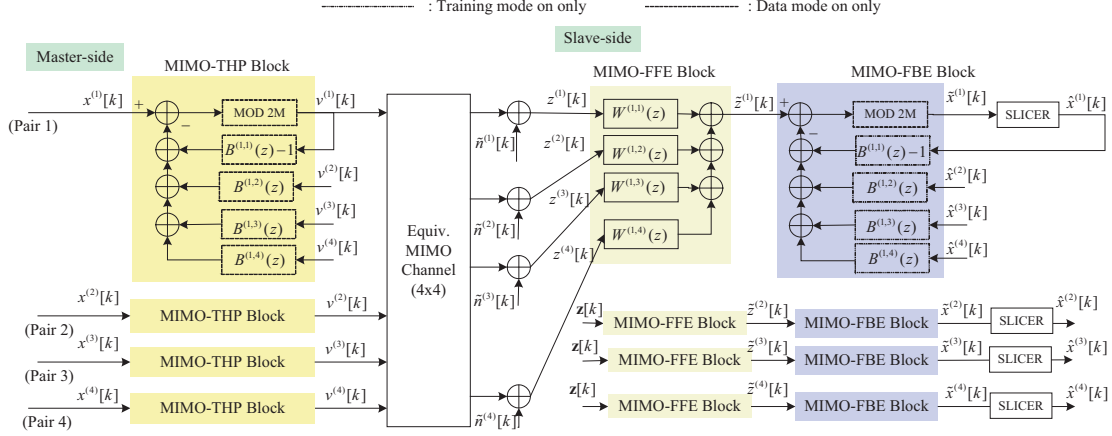


Fig. 1. System Architecture

lay skew and the consideration to choose step-size parameter in least-mean-square (LMS) algorithm are described. Section 4 presents simulation results to show the DP-SNR improvement of our proposed method over tentative decision based approach. The effects of delay skew issue and the of choosing step-size parameters are discussed as well. Conclusions are given in Section 5.

## 2. SYSTEM MODELS

A simplified system model for 10GBASE-T is shown in Fig. 1. The system will enter the data mode when the coefficients of MIMO THP were obtained in training mode.

### 2.1. Startup procedure

The operation procedure during the training mode can be described as follows:

1. Turn the MIMO THP blocks off, i.e.  $x^{(i)}[k] = v^{(i)}[k]$ ,  $i=1,2,3,4$ , and turn the MIMO DFE block on.
2. The master side transmits training sequence to the slave side to train MIMO feedback equalizer (FBE) block and MIMO feedforward equalizer (FFE) block.
3. Decision delay is estimated for each pair.
4. Delay skew is compensated.
5. Train the coefficients of MIMO FBE block and MIMO FFE block.
6. After the mean square error (MSE) has dropped below a pre-defined threshold value, then relay the well-trained coefficients of MIMO FBE blocks to the corresponding MIMO THP blocks.

### 2.2. MIMO Tomlinson-Harashima precoder (MIMO THP)

The operation of data mode can be described as follows:

1. Turn the MIMO FBE block off, i.e.  $\tilde{x}^{(i)}[k] = \text{Mod}\{\tilde{z}^{(i)}[k]\}$ ,  $i=1,2,3,4$ , and turn the MIMO THP block on.
2. The master side transmits 128-DSQ modulated symbol  $x^{(i)}[k]$  [9].
3. MIMO THP blocks precode the  $x^{(i)}[k]$  into the precoded symbols  $v^{(i)}[k]$ .
4. Adapt the coefficients of MIMO FFE block at receiver side.

For the  $i^{th}$  pair, the input symbol is  $x^{(i)}[k]$  and the corresponding TH-precoded output symbol  $v^{(i)}[k]$  is given by:

$$\begin{aligned}
 v^{(i)}[k] &= \text{Mod} \left\{ x^{(i)} - (b^{(j,i)} - \delta) * v^{(i)}[k] \right. \\
 &\quad \left. - \sum_{i \neq j} (b^{(j,i)} * v^{(i)}) [k] \right\} \\
 &= x^{(i)} - (b^{(j,i)} - \delta) * v^{(i)}[k] \\
 &\quad - \sum_{i \neq j} (b^{(j,i)} * v^{(i)}) [k] + d^{(i)}[k] \quad (1)
 \end{aligned}$$

, where  $\text{Mod}\{\cdot\}$  denotes the modulo  $2M$  operation,  $*$  denotes the convolution, and  $\delta[k]$  is the unit impulse response. The coefficients of MIMO THP are denoted as  $b^{(j,i)}[k]$ . The effect of modulo device is equivalent to choose the unique integer sequence  $d^{(i)}[k] \in 2M\mathbf{Z}$ , where  $\mathbf{Z}$  denotes the integer set, so that the precoded symbol  $v^{(i)}[k]$  falls into the interval  $[-M, M)$  with  $M=16$ . For pair one, the TH-precoded output symbol  $v^{(1)}[k]$  is the precoded symbol vector that are used for postcursor cancellation, while the other three TH-precoded output symbols,  $v^{(2)}[k]$ ,  $v^{(3)}[k]$ , and  $v^{(4)}[k]$ , are used for FEXT suppression

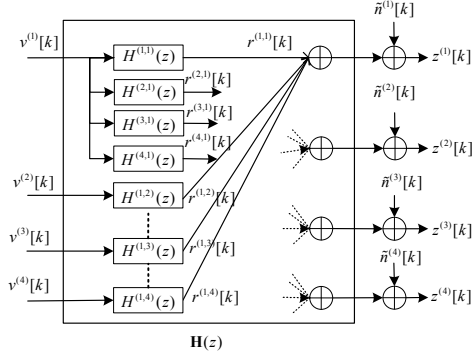


Fig. 2. The equivalent MIMO channel model.

### 2.3. MIMO channel model

In Fig.2, the channel is shown to be modeled as an equivalent  $4 \times 4$  matrix  $\mathbf{H}(z)$  in Z-domain representation, which is excited by a  $4 \times 1$  TH precoded vector  $\mathbf{v}[k] = [v^{(1)} v^{(2)} v^{(3)} v^{(4)}]^T$ . The diagonal elements of  $\mathbf{H}(z)$  represent IL channel and its off-diagonal elements represent the FEXT interference channel. We denote the impulse response of the channel between the  $i^{th}$  input and  $j^{th}$  output of the matrix channel as  $h^{(j,i)}[k]$  and the corresponding Z-domain representation as  $H^{(j,i)}(z)$ . To simplify the notations, the effect of root-raised cosine pulse shaping filter at the transmitter side and the effect of receiver matched filter, which is matched to the pulse shaping filter, are merged into the equivalent channel. The equivalent noise  $\tilde{n}^{(i)}[k]$  is the noise at the output of receiver matched filter with additive white Gaussian noise (AWGN) at the input.

### 2.4. Received signal model

We denote the received signal at  $i^{th}$  pair at time  $k$  as  $z^{(i)}[k]$ , which comprises the received signal  $r^{(i,i)}[k]$ , FEXT crosstalk  $\sum_{i \neq j} r^{(i,j)}[k]$ , and equivalent noise  $\tilde{n}^{(i)}[k]$ . Therefore,  $z^{(i)}[k]$  can be expressed as follows:

$$z^{(i)}[k] = \sum_{j=1}^4 \sum_{n=0}^{\nu-1} h^{(i,j)}[k-n] v^{(j)}[n] + \tilde{n}^{(i)}[k] \quad (2)$$

, where  $\nu$  is the memory length of channel. Then  $z^{(i)}[k]$  will pass through a set of  $4 \times 4$  FFE matrices  $\mathbf{w}_n$ ,  $n=0,1,\dots,N_F-1$ . In matrix form, the  $4 \times 1$  output vector of the MIMO FFE block  $\tilde{\mathbf{z}}[k]$  can be expressed as

$$\tilde{\mathbf{z}}[k] = \sum_{n=0}^{N_F-1} \mathbf{w}_n \mathbf{z}[k + \Delta - n] \quad (3)$$

, where an element at the  $i^{th}$  row and  $j^{th}$  column of FFE matrix tap  $\mathbf{w}_n$  is  $w^{(j,i)}[n]$ ,  $\mathbf{z}[\cdot] = [z^{(1)}, z^{(2)}, z^{(3)}, z^{(4)}]^T$  is a  $4 \times 1$  column vector,  $\Delta = [\Delta^{(1)}, \Delta^{(2)}, \Delta^{(3)}, \Delta^{(4)}]^T$ , and the

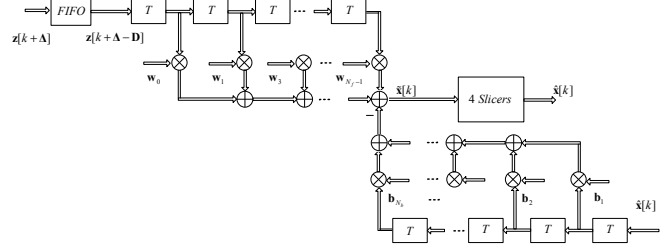


Fig. 3. The proposed MIMO DFE training architecture.

superscript "T" denotes the transpose. The DSQ slicer input of the  $i^{th}$  pair is  $\hat{x}^{(i)}[k] = \text{Mod} \{ \tilde{z}^{(i)}[k] \}$ , where

$$\tilde{z}^{(i)}[k] = \sum_{j=1}^4 w^{(i,j)} * z^{(j)}. \quad (4)$$

Ideally, if the IL channel is equalized perfectly and the FEXT interference is canceled totally, then

$$\tilde{z}^{(i)}[k] = x^{(i)}[k] + \tilde{n}^{(i)}[k] + d^{(i)}[k]. \quad (5)$$

After the modulo operation, it becomes

$$\tilde{x}^{(i)}[k] = \text{Mod} \{ \tilde{z}^{(i)}[k] \} = x^{(i)}[k] + \tilde{n}^{(i)}[k]. \quad (6)$$

Then the DSQ slicer will make decisions  $\hat{x}^{(i)}[k]$  and  $\hat{x}^{(i)}[k-1]$  base on  $\tilde{x}^{(i)}[k]$  and  $\tilde{x}^{(i)}[k-1]$ , respectively.

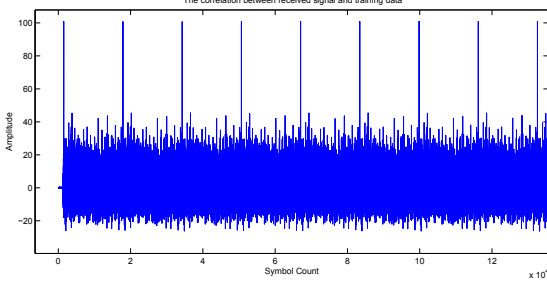
## 3. THE PROPOSED MIMO DFE TRAINING ARCHITECTURE

The proposed MIMO DFE training architecture is shown in Fig.3. It is noticed that there is a first-in, first-out (FIFO) block prior to the feed forward tap matrix to compensate the delay skew among pairs.

The symbol estimation vector  $\tilde{\mathbf{x}}[k]$  can be expressed as:

$$\tilde{\mathbf{x}}[k] = \sum_{n=0}^{N_F-1} \mathbf{w}_n \mathbf{z}[k + \Delta - \mathbf{D} - n] + \sum_{n=1}^{N_B} \mathbf{b}_n \hat{\mathbf{x}}[k-n] \quad (7)$$

, where  $\mathbf{D} = [D^{(1)}, D^{(2)}, D^{(3)}, D^{(4)}]^T$  is the delay skew compensation vector that is realized by the FIFO block,  $\mathbf{b}_n$ ,  $n=1,\dots,N_B-1$ , is a set of  $4 \times 4$  FBE matrices, whose element of  $i^{th}$  row and  $j^{th}$  column is  $b^{(j,i)}[n]$ , and  $\hat{\mathbf{x}}[\cdot] = [\hat{x}^{(1)}, \hat{x}^{(2)}, \hat{x}^{(3)}, \hat{x}^{(4)}]^T$  is a  $4 \times 1$  symbol decision vector which is the output of the decision device. Assume that all decisions are correct, i.e.  $\hat{\mathbf{x}} = \mathbf{x}$ , we then find the two sets of coefficient matrices,  $\mathbf{w}_n$  and  $\mathbf{b}_n$ , such that the covariance matrix of the error between  $\mathbf{x}$  and its linear estimation  $\tilde{\mathbf{x}}$  is minimized.



**Fig. 4.** An example of the partial correlation output. ( $K=256$ ,  $N_F=45$ ,  $N_B=35$ )

### 3.1. Delay skew estimation and compensation

To reduce the hardware cost, we correlate a portion of training sequence instead of the whole training sequences with the received sequence  $z^{(j)}[k]$ . The period of training sequence is 16384. Because the output of the correlator is periodic, we can exploit the index of the peak value  $I^{(j)}$  during one period to estimate the delay of the  $j^{\text{th}}$  pair. By using the partial correlation method,  $I^{(j)}$  can be obtained as follows:

$$I^{(j)} = \arg \max_k \left\{ z^{(j)}[k] * p^{(j)}[-k] \right\} - K \quad (8)$$

, where  $p^{(j)}[-k]$  is the first  $K$  training symbols in the reverse order. We consider the following simple case to illustrate this idea. In this case, we set the channel signal-to-noise ratio, defined as

$$\text{SNR}_{\text{chan}} \stackrel{\text{def}}{=} 10 \log \left( \frac{E(x[k]^{(i)})^2 \sum_k h[k]^{(i,i)^2}}{E(n[k]^{(i)})^2} \right), \quad (9)$$

equals 32 dB and  $h[k]^{(i,i)}$  is taken from the IEEE 802.3an 10GBASE-T task group [10]. The corresponding periodic correlation output is shown in Fig. 4, where the y-axis represents the correlation output value and x-axis is time axis with unit of symbol time. The first peak of the correlation output occurs at time index 1507. Therefore  $I^{(j)}=1507-256=1251$ .

To fully utilize the FFE, the decision delay is arranged as

$$\Delta^{(j)} = I^{(j)} + \lfloor N_F/2 \rfloor \quad (10)$$

Once the decision delay parameters are determined, the delay skew among pairs could be compensated through postponing the received sequences on the  $i^{\text{th}}$  pair by  $D^{(i)}$  symbol time, where

$$D^{(i)} = \max_j \left\{ \Delta^{(j)} \right\} - \Delta^{(i)}. \quad (11)$$

### 3.2. Least mean square (LMS) adaptive method

Traditionally, the step-size of LMS adaptive algorithm for the MIMO DFE are taken to be same for each pair. For stability

consideration, the initial value of step-size can be chosen as  $\Delta_s = \frac{0.2}{N P_r}$  [11], where  $N$  is the number of taps of linear equalizer,  $P_r$  is the power of the signal that is applied to the equalizer. Obviously, the step-size is inverse proportion to  $P_r$  and  $N$ . Therefore, the initial value of step-size of FFE and FBE should be different in general. Furthermore, because the crosstalk signals are much weaker than the direct-path signal, i.e.  $r^{(j,i)} \ll r^{(i,i)}$  for  $i \neq j$ , the cross portion of FFE tap, i.e.  $w^{(i,j)}$  for  $i \neq j$ , should use different step-size to adapt it as well. During the training period, the coefficients of FFE, FBE matrix are updated as the following equations:

$$\begin{aligned} \mathbf{w}^{(i,i)}(n+1) &= \mathbf{w}^{(i,i)}(n) + \mu_{f1} \mathbf{x}_f^{(i)}(n) e^{(i)}(n) \\ \mathbf{w}^{(i,j)}(n+1) &= \mathbf{w}^{(i,j)}(n) + \mu_{f2} \mathbf{x}_f^{(j)}(n) e^{(i)}(n), i \neq j \\ \mathbf{b}^{(i,j)}(n+1) &= \mathbf{b}^{(i,j)}(n) - \mu_b \mathbf{x}_b^{(j)}(n) e^{(j)}(n) \end{aligned} \quad (12)$$

, where  $\mathbf{w}^{(i,j)} \equiv [w^{(i,j)}[0], w^{(i,j)}[1], \dots, w^{(i,j)}[N_F - 1]]^T$ ,  $\mathbf{b}^{(i,j)} \equiv [b^{(i,j)}[1], b^{(i,j)}[2], \dots, b^{(i,j)}[N_B]]^T$ ,  $e^{(i)} = \hat{x}^{(i)} - \tilde{x}^{(i)}$  is the estimation error,  $\mathbf{x}_f^{(i)}$  is the feed-forward tap-input vector,  $\mathbf{x}_b^{(i)}$  is the feed-back tap-input vector,  $\mu_b$  is the step-size for FBE tap coefficients updating, and  $\mu_{f1}$  and  $\mu_{f2}$  are the step-sizes for non-cross and cross portion of FFE tap coefficients updating, respectively.

The selection of  $\mu_{f1}$  and  $\mu_{f2}$  are depend on the relative magnitude strength between non-cross and cross term of MIMO channel  $\mathbf{H}$ .

## 4. SIMULATION RESULTS

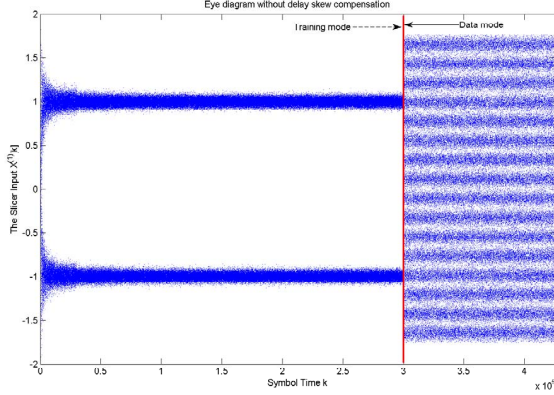
To demonstrate the performance improvement of our proposed approaches over the traditional approaches, we adopt CAT6 UTP cable with wire length 55 meters as channel model and these data can be found from IEEE 802.3an web site [10]. In the simulation, the following parameters were assumed:  $N_F=45$ ,  $N_B=45$ ,  $N_X=45$ ,  $\text{SNR}_{\text{chan}}=32$  dB, and the training period is  $3 \times 10^5$  symbol times.

### 4.1. The effect of delay skew among wire pairs

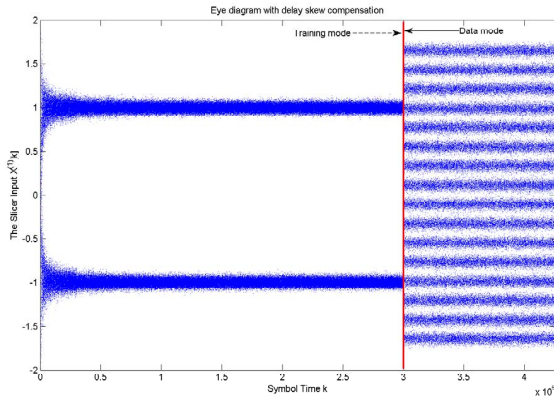
As shown in Fig. 5, it is clear that without delay skew compensation the MIMO THP will suffer from DP-SNR degradation in data mode, even they could achieve same DFE performance in training mode. The DFE performance is evaluated in terms of equalizer signal-to-noise ratio, defined as

$$\text{SNR}_{\text{DFE}} \stackrel{\text{def}}{=} 10 \log \left( \frac{E(x[k]^{(i)})^2}{E(\sigma^{(i)})^2} \right)$$

, where  $\sigma^{(i)}$  is the MSE at the decision point. As can be seen in Table 1, the average DP-SNR degradation is 4.59 dB, and the corresponding symbol error rate (SER) rises from  $1.94 \times 10^{-4}$  to  $5.07 \times 10^{-2}$ . Hence, the delay skew does need



(a) w/o delay skew compensation



(b) with delay skew compensation

**Fig. 5.** The effect of delay skew compensation on eye diagram.

to be compensated in training mode, or the well-trained coefficients of MIMO DFE can not be applied to MIMO THP directly in data mode.

#### 4.2. The effect of step-size parameters in LMS algorithm

Table 2 shows that although there is no significant DP-SNR degradation in training mode, yet, without different step-sizes for FFE matrix adaption in training mode, the average DP-SNR in data mode will degrade by 1.62 dB. Fig.6 represents the coefficients of these 16 FFE matrices. It is worthy to notice that the cross term between part (a) and part (b) are quite different. When we use different step-sizes to adjust  $w^{(i,j)}$ , the values of  $\lfloor N_F/2 \rfloor$ -right-most coefficients, i.e.  $w^{(i,j)}[\lfloor N_F/2 \rfloor + 1], \dots, w^{(i,j)}[N_F]$ , are almost zero. However, when same step-size parameters are used, the  $\lfloor N_F/2 \rfloor$ -right-most coefficients will converge to wrong values. This is because of the MSE is dominated by the non-cross terms adaption rather than the cross terms.

**Table 1.** The effect of delay skew comparison.

Items	w/o delay skew comp.			with delay skew comp.		
	Training mode	Data mode		Training mode	Data mode	
Pair	SNR(dB)	SNR(dB)	SER	SNR(dB)	SNR(dB)	SER
1	29.49	26.38	4.46E-03	29.52	29.22	2.08E-04
2	29.51	22.28	1.20E-01	29.53	28.99	2.56E-04
3	29.37	26.96	2.60E-03	29.37	29.06	2.16E-04
4	30.08	22.90	7.59E-02	30.04	29.59	9.60E-05
AVG	29.61	24.63	5.07E-02	29.62	29.22	1.94E-04

**Table 2.** The effect of different step-size.

(a)  $\mu_{f1} = \mu_{f2}$

Items	Training Mode		Data Mode
	SNR (dB)	SNR (dB)	SER
Pair			
1	29.73	28.07	8.40E-04
2	29.73	27.00	2.34E-03
3	29.45	27.52	1.34E-03
4	30.18	27.81	1.05E-03
AVG	29.77	27.60	1.39E-03

(b)  $\mu_{f1} = 10^4 \mu_{f2}$

Items	Training Mode		Data Mode
	SNR (dB)	SNR (dB)	SER
Pair			
1	29.57	29.19	2.16E-04
2	29.50	29.00	2.48E-04
3	29.37	29.07	1.76E-04
4	30.02	29.60	8.00E-05
AVG	29.62	29.22	1.80E-04

#### 4.3. Simulation results of traditional receiver

As shown in Fig.7 is referred as the traditional signal path architecture for the 1<sup>st</sup> pair, where  $F^{(j,i)}(z)$  denotes the FEXT canceler,  $\hat{x}^{(j)}[n]$  is the tentative decision,  $\hat{v}^{(j)}[n]$  is the estimated TH precoded symbol. The corresponding simulated performance is listed in Table 3, where the ideal case refers to that with  $\hat{x}^{(j)}[n] = x^{(j)}[n]$ . It shows that the poor tentative decisions degrade the average DP-SNR by 2.75 dB. Comparing with our proposed scheme, i.e. MIMO THP architecture with delay skew compensation and using different step-size parameters, the traditional receiver architecture suffer from an average DP-SNR degradation of 2.46 dB.

**Table 3.** The performance of traditional receiver architecture.

Items	traditional architecture				
	Training mode	Data mode (Ideal case)		Data mode (Tentative decision)	
		SNR(dB)	SNR(dB)	SER	SNR(dB)
Pair					
1	29.53	29.49	3.02E-03	27.49	1.50E-02
2	29.42	29.35	3.07E-03	25.69	8.39E-02
3	29.27	29.28	3.35E-03	27.75	1.21E-02
4	29.98	29.93	2.42E-03	26.09	6.68E-02
AVG	29.55	29.51	2.97E-03	26.76	4.45E-02

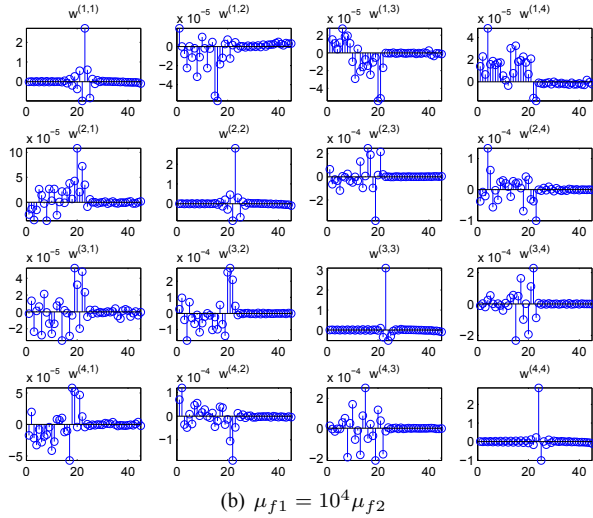
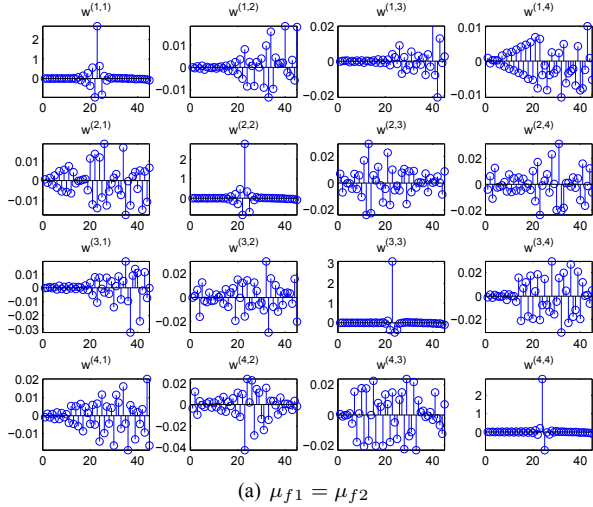


Fig. 6. The effect of step-size on FFE matrices.

## 5. CONCLUSIONS

In this paper, we proposed a MIMO THP architecture to cancel the postcursor ISI and FEXT interference in a 10GBASE-T system. The method to estimate and compensate the delay skew and the effects of varying step-size parameters are described together with the proposed architecture. The simulation results show that the proposed method outperforms 2.75 dB than the traditional tentative decision based approach in terms of DP-SNR.

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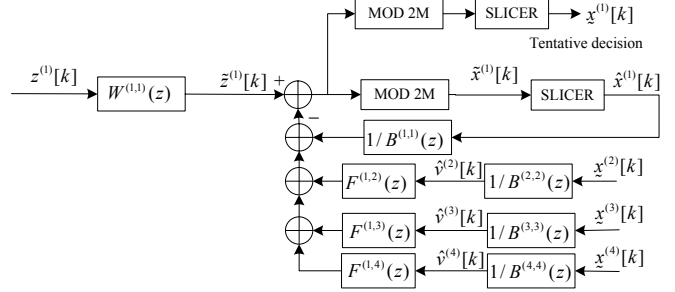


Fig. 7. Tentative decision based approach for pair one.

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