FAST COMPUTATION OF MIMO EQUALIZERS AND CANCELLERS IN 10GBASE-T CHANNELS

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ABSTRACT

This paper presents an efficient method for computing the optimal tap coefficients of the MIMO equalizers and cancellers from the channel impulse responses (CIR). The proposed method provides an insight of the minimum mean-square error (MMSE) solution in a general MIMO system and shows that the MSE optimization problem in a M-input N-output system can be decomposed into N independent minimization problems each with smaller size. Solving each separate problem in MMSE sense is computationally efficient, thus leading to substantial savings in overall computational complexity. Compared with a prior method, this new method is exact and much faster.

Index Terms— MIMO, CIR, MMSE, independent, computation complexity

1. INTRODUCTION

In many multiple wireline communication systems, the received signal not only suffers from signal attenuation and inter symbol inference (ISI), but also suffers from echo, near-end cross talk (NEXT), far-end cross talk (FEXT), and other noises such as alien NEXT (ANEXT). A typical example is 10 Gigabit Ethernet over copper (10Gbase-T) system, which performs full duplex baseband transmission over four pairs of unshielded twisted pair (UTP) copper cables, where ISI is a significant impairment against reliable high speed digital transmission, and each received signal is corrupted by echo from its own transmitter and NEXT interferences from three adjacent transmitters. To meet the desired throughput ($10Gbps$) and target BER (10^{-12}) requirements, a MIMO equalization and cancellation scheme is applied to 10Gbase-T system [1].

To set the optimal tap coefficients of the equalizers and cancellers in such a MIMO system, one straightforward approach is to recursively compute the tap coefficients using adaptive algorithms such as least mean square (LMS). However, the recursive LMS adaptation of these equalizers and cancellers may take millions of iterations to converge to the optimal solution on a typical Category-6 UTP channel. Much faster convergence can be achieved using RLS algorithm, but the higher computational cost is prohibitive in real applications. Another indirect approach is to compute the tap coefficients, in MMSE sense, based on the channel estimates [2]. In this approach, the knowledge of the channel impulse response, as well as the noise characteristics are required. In addition, a computationally efficient approach is also needed to compute the optimal settings of the MIMO equalizers and cancellers.

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Previous studies on efficiently computing the tap coefficients of the equalizers are mostly based on the traditional finite-length MMSE-DFE structure. In [2], Cholesky factorization is applied to carry out the involved matrix inversion efficiently. By exploiting the structured matrices, a generalized Schur algorithm for fast Cholesky matrices decomposition was provided in [3] to reduce the computational complexity in both feedback and feedforward filters computation. Recently, another efficient approach is proposed to achieve faster computation by identifying the relationship between the feedforward equalizer computation and fast recursive least squares (RLS) adaptive algorithms, and treating the feedback equalizers computation as a convolution operation [5].

Although these methods can be easily extended to the general MMSE-DFE tap computation in MIMO channels [4, 6], they may not always be computationally efficient for computing the optimal coefficients of the MIMO equalizers and cancellers in the cases where the cancellers have a larger number of taps than the feed-forward equalizers. By using Al-Dhahir's method [4], the inversion of an embedded correlation matrix will be computationally intensive since the size of the matrix is related to the maximum number of taps in Echo cancellers and NEXT cancellers, as shown in the following section. In addition, the number of taps in Echo cancellers and NEXT cancellers need not be same in real applications. Thus, applying the efficient method in [6] is also not straightforward.

In this paper, we present a new computationally efficient approach for computing the optimum settings of the MIMO equalizers and cancellers for 10Gbase-T, assuming that the channel impulse response estimate and the noise characteristics are known. This new method is exact and applicable to cases where Echo & NEXT cancellers have different lengths. Compared with Al-Dhahir's method, it has lower computational complexity.

In Section 2, the system model is briefly reviewed and the MMSE solution based on Al-Dhahir's method is derived. In Section 3, the proposed approach for fast computation of the tap coefficients is presented in detail, and the computational savings due to the proposed method are also addressed.

2. SYSTEM MODEL AND PROBLEM FORMULATION

In 10Gbase-T, a typical Category-6 UTP channel can be modeled as two 4×4 MIMO channels as shown in Fig. 1, where $h_{i,j}$ denotes the MIMO channel impulse response from the ith input to the jth output with length $v{+}1$ and $g_{m,n}$ denotes the Echo & NEXT channel impulse response from the mth input to the nth output with length $l+1$. Let x_i denote the transmitted symbol sequence from the ith far end transmitter and z_m denote the transmitted symbol sequence from the mth near end transmitter, and n_j denote background noise at the jth channel output. Then the jth channel received symbol sequence is given by

$$y_j = \sum_{i=1}^{4} h_{i,j} \otimes x_i + \sum_{m=1}^{4} g_{m,j} \otimes z_m + n_j \tag{1}$$

for $j = 1, \ldots, 4$. where $\otimes$ denotes convolution.

By grouping symbols from 4 received channels at time k into a column vector $\mathbf{y}(k) \triangleq [y_1(k) \quad y_2(k) \quad y_3(k) \quad y_4(k)]^T$, (1) can be expressed as follows:

$$\mathbf{y}(k) = \sum_{\tau=0}^{v} H_\tau \mathbf{x}(k-\tau) + \sum_{p=0}^{l} G_p \mathbf{z}(k-p) + \mathbf{n}(k) \tag{2}$$

where H_τ and G_p represent 4×4 τth far end channel coefficient matrix and pth near end channel coefficient matrix, respectively. The signals $\mathbf{x}(k-\tau)$ and $\mathbf{z}(k-\tau)$ correspond to far end transmitted column vector and near end transmitted column vector at time index $k-\tau$. By stacking N_f successive channel output vector samples, (2) can be expressed in matrix form as follows:

$$\begin{aligned} \mathbf{y}(k+N_f-1:k) &= \mathbf{H} \cdot \mathbf{x}(k+N_f-1:k-v) \\ &\quad + \mathbf{G} \cdot \mathbf{z}(k+N_f-1:k-l) \\ &\quad + \mathbf{n}(k+N_f-1:k). \end{aligned} \tag{3}$$

where $\mathbf{y}(k+N_f-1:k)$ is a column vector with dimension $4N_f \times 1$, and matrix $\mathbf{H}$ and matrix $\mathbf{G}$ are both block Toeplitz matrices with dimension $N_f \times (N_f+v)$ and $N_f \times (N_f+l)$ respectively.

$$\mathbf{H} = \begin{bmatrix} H_0 & H_1 & \ldots & H_v & 0 & \ldots & 0 \\ 0 & H_0 & H_1 & \ldots & H_v & \ldots & 0 \\ \vdots & & \ddots & & & \ddots & \\ 0 & \ldots & 0 & H_0 & H_1 & \ldots & H_v \end{bmatrix}.$$

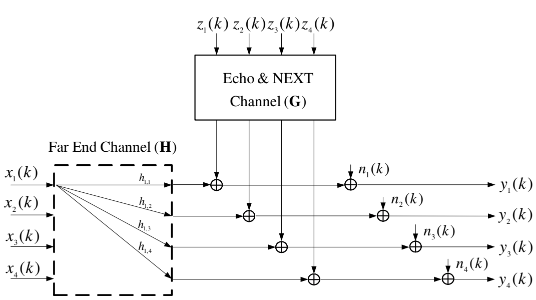

Fig. 1. Block Diagram of the MIMO Channel

Fig. 2 shows the block diagram of the MIMO equalizer and canceller. In this figure, let N_f, N_b, N_p be the lengths of the feed forward filter matrix $\mathbf{W}$, feedback filter matrix $\mathbf{B}$, and Echo & NEXT cancellation filter matrix $\mathbf{P}$, respectively. If we further assume Echo & NEXT cancellers have different numbers of taps with N_e and N_x respectively, then $N_p \triangleq max(N_e, N_x)$. The objective is to choose $\mathbf{W}$, $\mathbf{B}$, and $\mathbf{P}$ to minimize mean square error (MSE) of the four channels.

From Fig. 2, the error vector at time k of the four channels can be represented by

$$\begin{aligned} \mathbf{e}(k) = \quad & \tilde{\mathbf{B}}^H \mathbf{x}(k+N_f-1:k-v) + \\ & \tilde{\mathbf{P}}^H \mathbf{z}(k+N_f-1:k-l) - \mathbf{W}^H \mathbf{y}(k+N_f-1:k) \end{aligned} \tag{4}$$

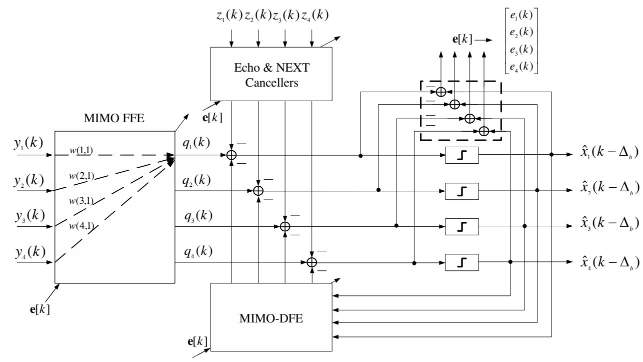

Fig. 2. Joint MIMO Equalizers and Cancellers

where,

$$\begin{aligned} \mathbf{W}^H &= \begin{bmatrix} W_0^H & W_1^H & \ldots & W_{N_f-1}^H \end{bmatrix}, \\ \tilde{\mathbf{B}}^H &= \begin{bmatrix} \mathbf{0}_{1\times\Delta_b} & B_0^H & B_1^H & \ldots & B_{N_b}^H & \mathbf{0}_{1\times s_1} \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{0}_{1\times\Delta_b} & \vec{\mathbf{B}}^H & \mathbf{0}_{1\times s_1} \end{bmatrix}, \\ \tilde{\mathbf{P}}^H &= \begin{bmatrix} \mathbf{0}_{1\times\Delta_p} & P_1^H & \ldots & P_{N_p}^H & \mathbf{0}_{1\times s_2} \end{bmatrix} \\ &= \begin{bmatrix} \mathbf{0}_{1\times\Delta_p} & \vec{\mathbf{P}}^H & \mathbf{0}_{1\times s_2} \end{bmatrix}, \end{aligned}$$

where W_i, B_i and P_i are 4×4 blocks, and $\mathbf{0}$ is a 4×4 zero matrix. We also define $s_1 = N_f + v - N_b - \Delta_b - 1$ and $s_2 = N_f + l - N_p - \Delta_p$ with the decision delays Δ_b and Δ_p, respectively.

Now the mean square error minimization problem can be formulated as,

$$\begin{aligned} \textit{min.} \quad & E[\|\mathbf{e}(k)\|^2] = trace(\mathbf{R}_{ee}) \\ \textit{s.t.} \quad & \begin{bmatrix} \vec{\mathbf{B}}^H & \vec{\mathbf{P}}^H \end{bmatrix} \mathbf{\Phi} = \mathbf{I} \end{aligned} \tag{5}$$

where $\mathbf{\Phi^H} = \begin{bmatrix} \mathbf{I} & \mathbf{0}_{1\times(N_b+N_p)} \end{bmatrix}$, and $\mathbf{I}$ is a 4×4 indentity matrix. Solving this optimization problem, we get

$$\begin{bmatrix} \vec{\mathbf{B}}_{opt} \\ \vec{\mathbf{P}}_{opt} \end{bmatrix} = \mathbf{R_\Delta^{-1}\Phi(\Phi^H R_\Delta^{-1} \Phi)^{-1}} \tag{6}$$

$$\mathbf{W}_{opt}^H = \begin{bmatrix} \tilde{\mathbf{B}}_{opt}^H & \tilde{\mathbf{P}}_{opt}^H \end{bmatrix} \begin{bmatrix} R_{xy} \\ R_{zy} \end{bmatrix} R_{yy}^{-1} \tag{7}$$

$$\begin{aligned} \mathbf{R}_{ee,min} &= E[\mathbf{e}(k)\mathbf{e}^H(k)] \\ &= \begin{bmatrix} \vec{\mathbf{B}}_{opt}^H & \vec{\mathbf{P}}_{opt}^H \end{bmatrix} \mathbf{R}_\Delta \begin{bmatrix} \vec{\mathbf{B}}_{opt} & \vec{\mathbf{P}}_{opt} \end{bmatrix} \\ &= \mathbf{(\Phi^H R_\Delta^{-1} \Phi)^{-1}} \end{aligned} \tag{8}$$

$$MSE_{min} = \frac{1}{4} trace(\mathbf{R}_{ee,min}) \tag{9}$$

where $\mathbf{R}_\Delta = \mathbf{Q}^H \mathbf{R} \mathbf{Q}$, and $\mathbf{Q}$ is a constant matrix.

$$\mathbf{R} = \begin{bmatrix} R_{xx} - R_{xy}R_{yy}^{-1}R_{yx} & -R_{xy}R_{yy}^{-1}R_{yx} \\ -R_{zy}R_{yy}^{-1}R_{yz} & R_{zz} - R_{zy}R_{yy}^{-1}R_{yz} \end{bmatrix}.$$

Since the matrix $\mathbf{R_\Delta}$ in (6) has dimensions $4(N_b+N_p+1) \times 4(N_b+N_p+1)$, Cholesky factorization would generally need $O[64(N_b +$

$N_p + 1)^3]$ operations. By using fast algorithms in [4], these computations can be performed efficiently in $O[16(N_b + N_p + 1)^2]$ operations. However, we note that the computational complexity required for fast Cholesky factorization depends on the total number of taps in feedback equalizers and Echo & NEXT cancellers. Hence, for cancellers with a large number of taps, the computation of the optimal tap coefficients will be expensive.

3. FAST COMPUTATION METHOD

In this section, a computationally efficient approach for computing the optimum settings of the MIMO equalizers and cancellers is presented. The basic idea is to reduce the dimension of matrix required for Cholesky factorization. This can be achieved by decomposing the original optimization problem in (5) into four separate/independent optimization problems, each with smaller size. Solving each optimization problem is computationally efficient. Thus, the overall computational complexity will be significantly reduced.

The minimization problem in (5) can be written as:

$$\begin{array}{ll} \textit{min.} & E[\|\mathbf{e}(k)\|^2] = trace(\mathbf{R}_{ee}) \\ = & E[e_1(k)^2] + E[e_2(k)^2] + E[e_3(k)^2] + E[e_4(k)^2] \\ \textit{s.t.} & \begin{bmatrix} \vec{\mathbf{B}}^H & \vec{\mathbf{P}}^H \end{bmatrix} \mathbf{\Phi} = \mathbf{I} \end{array} \tag{10}$$

where $e_i(k),\ (i = 1, 2, 3, 4)$ is the error signal for each corresponding channel. To show that this problem is equivalent to minimizing each of the mean square error for each channel, we rewrite (4) as:

$$\begin{aligned} \mathbf{e}(k) &= \begin{bmatrix} \tilde{\mathbf{B}}^H & \tilde{\mathbf{P}}^H \end{bmatrix} \cdot \begin{bmatrix} \mathbf{x}(k + N_f - 1 : k - v) \\ \mathbf{z}(k + N_f - 1 : k - l) \end{bmatrix} \\ &- \mathbf{W}^H \mathbf{y}(k + N_f - 1 : k) \\ &= \mathbf{A}^H \mathbf{s} - \mathbf{W}^H \mathbf{y}(k + N_f - 1 : k) \end{aligned} \tag{11}$$

where $\mathbf{A}^H = \begin{bmatrix} \tilde{\mathbf{B}}^H & \tilde{\mathbf{P}}^H \end{bmatrix}$, and $\mathbf{s} = \begin{bmatrix} \mathbf{x}(k + N_f - 1 : k - v) \\ \mathbf{z}(k + N_f - 1 : k - l) \end{bmatrix}$. Each element of $\mathbf{e}(k)$ is given by

$$\begin{aligned} e_1(k) &= \mathbf{A}^{(1)H}\mathbf{s} - \mathbf{W}^{(1)H}\mathbf{y} \\ e_2(k) &= \mathbf{A}^{(2)H}\mathbf{s} - \mathbf{W}^{(2)H}\mathbf{y} \\ e_3(k) &= \mathbf{A}^{(3)H}\mathbf{s} - \mathbf{W}^{(3)H}\mathbf{y} \\ e_4(k) &= \mathbf{A}^{(4)H}\mathbf{s} - \mathbf{W}^{(4)H}\mathbf{y} \end{aligned}$$

where $\mathbf{A}^{(i)H}$ and $\mathbf{W}^{(i)H}$ represent the ith row of $\mathbf{A}^H$ and $\mathbf{W}^H$, respectively. Then, we have

$$E[e_i e_i^H] = \mathbf{A}^{(i)H}(R_{ss} - R_{sy}R_{yy}^{-1}R_{ys})\mathbf{A}^{(i)} + (\mathbf{W}^{(i)H} - \mathbf{A}^{(i)H}R_{sy}R_{yy}^{-1})R_{yy}(\mathbf{W}^{(i)H} - \mathbf{A}^{(i)H}R_{sy}R_{yy}^{-1})^H.$$

The goal is to minimize the total MSE, $E[\|\mathbf{e}(k)\|^2]$, with respect to $\mathbf{A}^{(i)}$ and $\mathbf{W}^{(i)}, i = 1, 2, 3, 4$. Since there is no constraint on $\mathbf{W}$, we can first minimize the total MSE with respect to $\mathbf{W}^{(1)}$, assuming $\mathbf{A}^{(1)}$ is fixed. It is easy to see that $E[e_i e_i^H](i = 2, 3, 4)$ is independent of $\mathbf{W}^{(1)}$, and minimizing the total MSE with respect to $\mathbf{W}^{(1)}$ is equivalent to minimizing the $E[e_1 e_1^H]$ separately. Thus, we get the optimal coefficients of the feed-forward filters corresponding to channel 1 as follows:

$$\mathbf{W}_{opt}^{(1)H} = \mathbf{A}^{(1)H} R_{sy} R_{yy}^{-1} \tag{12}$$

then the minimum $E[e_1 e_1^H]$ (with respect to $\mathbf{W}^{(1)}$) turns out to be

$$E[e_1 e_1^H] = \mathbf{A}^{(1)H}(R_{ss} - R_{sy}R_{yy}^{-1}R_{ys})\mathbf{A}^{(1)} \tag{13}$$

similarly, we have

$$E[e_2 e_2^H] = \mathbf{A}^{(2)H}(R_{ss} - R_{sy}R_{yy}^{-1}R_{ys})\mathbf{A}^{(2)} \tag{14}$$

$$E[e_3 e_3^H] = \mathbf{A}^{(3)H}(R_{ss} - R_{sy}R_{yy}^{-1}R_{ys})\mathbf{A}^{(3)} \tag{15}$$

$$E[e_4 e_4^H] = \mathbf{A}^{(4)H}(R_{ss} - R_{sy}R_{yy}^{-1}R_{ys})\mathbf{A}^{(4)} \tag{16}$$

Now we minimize the total MSE with respect to $\mathbf{A}^{(i)}$. As before, minimizing the total MSE with respect to $\mathbf{A}^{(i)}$ is equivalent to minimizing each $E[e_i e_i^H]$ only. Therefore, we can solve the original problem by minimizing each of the $E[e_i e_i^H]$.

For minimizing each of the $E[e_i e_i^H]$, the problem size is greatly reduced. As an example, we solve the optimization problem corresponding to $E[e_1 e_1^H]$ minimization. We first write $e_1(k)$ as

$$\begin{aligned} e_1(k) &= x_1(n - \Delta_{b1}) - \Big(\mathbf{W1}^H\mathbf{y1} - \mathbf{B1}^H\mathbf{x1} - \mathbf{P1}^H\mathbf{z1}\Big) \\ &= x_1(n - \Delta_{b1}) - \Big(\mathbf{W1}^H\mathbf{y1} - \mathbf{A1}^H\mathbf{s1}\Big) \end{aligned} \tag{17}$$

where Δ_{b1} is the decision delay for channel 1, and **W1**, **B1** and **P1** are all column vectors,

$$\begin{aligned} \mathbf{W1}^H &\triangleq \begin{bmatrix} W^{(1,1)H} & W^{(2,1)H} & W^{(3,1)H} & W^{(4,1)H} \end{bmatrix} \\ W^{(i,j)} &\triangleq \begin{bmatrix} w_0^{(i,j)} & w_1^{(i,j)} & \ldots & w_{N_f-1}^{(i,j)} \end{bmatrix}^H \\ \mathbf{B1}^H &\triangleq \begin{bmatrix} B^{(1,1)H} & B^{(2,1)H} & B^{(3,1)H} & B^{(4,1)H} \end{bmatrix} \\ B^{(i,j)} &\triangleq \begin{bmatrix} b_1^{(i,j)} & b_2^{(i,j)} & \ldots & b_{N_b}^{(i,j)} \end{bmatrix}^H \\ \mathbf{P1}^H &\triangleq \begin{bmatrix} P^{(1,1)H} & P^{(2,1)H} & P^{(3,1)H} & P^{(4,1)H} \end{bmatrix} \\ P^{(i,i)} &\triangleq \begin{bmatrix} p_1^{(i,i)} & p_2^{(i,i)} & \ldots & p_{N_e}^{(i,i)} \end{bmatrix}^H \\ P^{(i,j)} &\triangleq \begin{bmatrix} p_1^{(i,j)} & p_2^{(i,j)} & \ldots & p_{N_x}^{(i,j)} \end{bmatrix}^H. \end{aligned}$$

y1, **x1** and **z1** are also column vectors,

$$\begin{aligned} \mathbf{y1}^H &\triangleq \begin{bmatrix} y^{(1,1)H} & y^{(2,1)H} & y^{(3,1)H} & y^{(4,1)H} \end{bmatrix} \\ \mathbf{x1}^H &\triangleq \begin{bmatrix} x^{(1,1)H} & x^{(2,1)H} & x^{(3,1)H} & x^{(4,1)H} \end{bmatrix} \\ \mathbf{z1}^H &\triangleq \begin{bmatrix} z^{(1,1)H} & z^{(2,1)H} & z^{(3,1)H} & z^{(4,1)H} \end{bmatrix} \\ x^{(i,j)} &\triangleq \begin{bmatrix} x_{n-\Delta_{b1}-1}^{(i,j)} \\ x_{n-\Delta_{b1}-2}^{(i,j)} \\ \vdots \\ x_{n-\Delta_{b1}-N_b}^{(i,j)} \end{bmatrix}, y^{(i,j)} \triangleq \begin{bmatrix} y_n^{(i,j)} \\ y_{n-1}^{(i,j)} \\ \vdots \\ y_{n-N_f+1}^{(i,j)} \end{bmatrix} \\ z^{(i,i)} &\triangleq \begin{bmatrix} z_{n-\Delta_{e1}-1}^{(i,i)} \\ z_{n-\Delta_{e1}-2}^{(i,i)} \\ \vdots \\ z_{n-\Delta_{e1}-N_e}^{(i,i)} \end{bmatrix}, z^{(i,j)} \triangleq \begin{bmatrix} z_{n-\Delta_{x1}-1}^{(i,j)} \\ z_{n-\Delta_{x1}-2}^{(i,j)} \\ \cdots \\ z_{n-\Delta_{x1}-N_x}^{(i,j)} \end{bmatrix} \end{aligned}$$

To minimize $E[e_1 e_1^H]$ from (17), the well-known solution is given by [7],

$$\begin{bmatrix} \mathbf{W1} \\ \mathbf{B1} \\ \mathbf{P1} \end{bmatrix} = \mathbf{\Phi^{-1}}\theta \tag{18}$$

where

$$\Phi = E\left(\begin{bmatrix} \mathbf{y1} \\ \mathbf{x1} \\ \mathbf{z1} \end{bmatrix} \begin{bmatrix} \mathbf{y1} & \mathbf{x1} & \mathbf{z1} \end{bmatrix}^H\right)$$
$$\theta = E\left(\begin{bmatrix} \mathbf{y1} \\ \mathbf{x1} \\ \mathbf{z1} \end{bmatrix} x_1(k-\Delta_{b1})\right). \quad (19)$$

Finally, the optimum taps of the equalizers and cancellers corresponding to the channel 1 can be computed as,

$$\begin{bmatrix} \mathbf{W1} \\ \mathbf{B1} \\ \mathbf{P1} \end{bmatrix} = \begin{bmatrix} \Phi_1^{-1}\theta_1 \\ -\mathbf{R}^{\mathbf{H}}_{\mathbf{y_1x_1}}\mathbf{W1} \\ -\mathbf{R}^{\mathbf{H}}_{\mathbf{y_1z_1}}\mathbf{W1} \end{bmatrix} \quad (20)$$

where

$$\theta_1 = E\left[\mathbf{y1}x_1(k-\Delta_{b1})\right], \quad (21)$$

and

$$\Phi_1 = \mathbf{R_{y1y1}} - \mathbf{R_{y1z1}}\mathbf{R^H_{y1z1}} - \mathbf{R_{y1x1}}\mathbf{R^H_{y1x1}}. \quad (22)$$

Here, we assume the input vectors **x** and **z** are $i.i.d$ with unit power in our derivation. We see that Φ_1 has dimension $4N_f \times 4N_f$, which only depends on the length of the feed-forward filters. Thus, the size of the matrix required for Cholesky factorization is reduced compared with $4(N_b+N_p+1)\times 4(N_b+N_p+1)$ in (6). We also note that no multiplications are needed to compute (21) and form the the block Toeplitz matrices $\mathbf{R_{y1z1}}$ and $\mathbf{R_{y1x1}}$ in (22). Therefore, computing (20) requires a total number of

$$2(4N_f)^2 + 4N_f(N_e+3N_x) + 16N_f(N_b) + 16N_f^2(v-1)$$
$$+4N_f^2(N_e-1) + 12N_f^2(N_x-1) + 16N_f^2(N_b-1)$$

multiplications. Assuming that the complexity is dominated by the multiplication operations, Fig. 3 shows the complexity comparison between Al-Dhahir's and the proposed method. Fig. 4 shows the computational reduction due to the proposed approach compared with Al-Dhahir's method. It can be observed that the proposed method can achieve substantial computational savings of 63.8% (where, N_f is the length of feed-forward MIMO equalizer; $N_b = 32$ is the length of the feedback matrix filter; $N_e = 500$ is the length of Echo cancellers; $N_x = 400$ is the length of NEXT cancellers; and direct MIMO channel & crosstalk channel have lengths $v = 1000$ & $l = 1000$, respectively).

4. CONCLUSION

We have presented a new method to compute the optimal coefficients of the MIMO equalizers and cancellers in 10Gbase-T channel. The proposed approach is exact and applicable to the general MIMO DFE computation as well as such cases where Echo & NEXT cancellers have a large number of taps with different lengths, which usually make Al-Dhahir's method inefficient. It is shown that, by using the proposed method, we are able to achieve about 63.8% computation cost reduction in terms of multiplication operations compared with the existing methods. This computation speedup also makes the analysis easier when Alien crosstalk such as ANEXT is considered in the channel model.

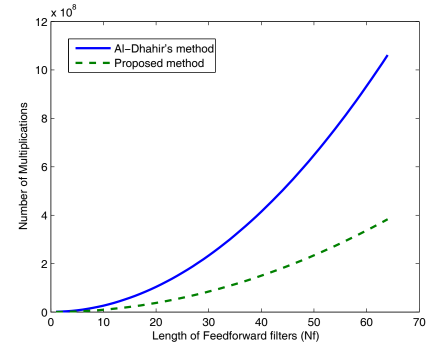

Fig. 3. Number of Multiplication Comparison

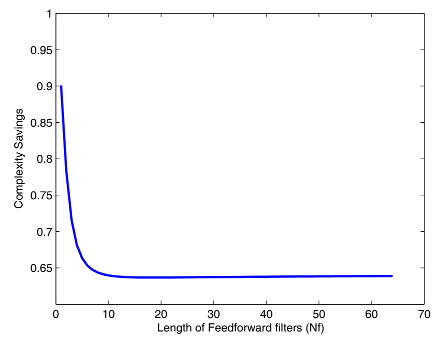

Fig. 4. Complexity Reduction Comparison

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