

# Differencing Multiuser Detection Using Error Feedback Filter for MIMO DS-UWB System in Nakagami Fading Channel

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## Abstract

A differencing multiuser detection (MUD) method is proposed for multiple-input multiple-output (MIMO) direct sequence (DS) ultra-wideband (UWB) system to cope with the multiple access interference (MAI) and the computational efficiency in Nakagami fading channel. The method, which combines a multiuser-interference-cancellation-based decision feedback equalizer using error feedback filter (MIC DFE-EFF), a coefficient optimization algorithm (COA) and a differencing algorithm (DA), is termed as MIC DFE-EFF (COA) with DA for short. In the paper, the proposed MUD method is illuminated from the rudimental MIC DFE-EFF to the advanced MIC DFE-EFF (COA) with DA step by step. Firstly, the MIC DFE-EFF system performance is analyzed by minimum mean square error criterion. Secondly, the COA is investigated for optimization of each filter coefficient. Finally, the DA is introduced to reduce the computational complexity while sacrificing little performance. Simulations show a significant performance gain can be achieved by using the MIC DFE-EFF (COA) with DA detector. The proposed MIC DFE-EFF (COA) with DA improves both bit error rate performance and computational efficiency relative to DFE, DFE-EFF, parallel interference cancellation (PIC), MIC DFE-EFF and MIC DFE-EFF with DA, though it sacrifices little system performance, compared with MIC DFE-EFF (COA) without DA.

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**Keywords:** decision feedback equalizer, coefficient optimization algorithm, error feedback filter, differencing multiuser detection, DS-UWB

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## 1. Introduction

The recent increase in demand for short-range and very high data transmission rates in wireless communications has prompted the development of a new generation of broadband wireless communication systems, namely ultra-wideband (UWB) systems [1][2][3]. A UWB system utilizes a frequency spectrum between 3.1 and 10.6 GHz, with a channel bandwidth which is greater than 500 MHz or 20% of the center frequency. The transmission power level of it is below the Federal Communications Commission (FCC) limit on spurious emissions ( $<-41.3$  dBm/MHz) [4]. The characteristics of huge bandwidth and low transmission power have turned UWB technology into a new focus in both academic and industrial sectors. Recently, a new communication paradigm that incorporates multiple-input multiple-output (MIMO) technologies into UWB system has gained considerable interest.

Interest in the study of UWB systems, especially in the multiuser detection (MUD) techniques in such systems, has increased. The well-known MUD technique, which originates from code division multiple access (CDMA) systems, can also be applied in the context of multiple access UWB systems to combat multiple access interference (MAI) and improve the system performance. The optimal multiuser detector proposed by Verdu [5] had high computational complexity and was too expensive to handle. Therefore, suboptimal detectors have become a focus of research. For example, a hybrid MUD algorithm was proposed by Yin [6] when the direct sequence UWB (DS-UWB) technique was used in the outer space ad-hoc wireless network. A minimum mean square error (MMSE) method was investigated by Li [7] and Kaligineedi [8] to compute filter weights and deal with frequency-domain equalization, respectively, for MUD in UWB systems. An interference cancellation (IC) technique was developed by Boubaker [9] in direct sequence UWB (DS-UWB) systems for wireless personal area networks (WPAN). Adaptive MUDs based on the recursive least square (RLS) principles were operated by Biradar [10] and Ahmed [11], and MMSE equalization using interference mitigation was conducted by Hu [12]. However, the bit error rate (BER) of the system was high when conventional suboptimal detection techniques, like conventional IC [9], were used in UWB systems.

In this paper, we introduce decision feedback equalizer (DFE) in MUD for MIMO DS-UWB system, because DFE shows a relatively good BER performance for the detection and equalization in MIMO system [12][13][14][15][16]. To deal with the BER performance, which is different from that of conventional DFE, we explore the error feedback filter (EFF) and the feedback of multiuser interference cancellation (MIC) for MIMO DS-UWB system (i.e. MIC DFE-EFF). Moreover, we also propose a coefficient optimization algorithm (COA) which optimizes the system performance with a slight increment of computational complexity, since it exploits the known coefficients of the previous matched filter. To deal with the computational complexity, we propose a differencing algorithm (DA) which can reduce computational complexity of detection system. Our MUD method can be exactly termed as MIC-based DFE using EFF with COA and DA for MIMO DS-UWB system (i.e. MIC DFE-EFF (COA) with DA). The proposed MUD method improves the system performance in three ways. Firstly, based on the MIC scheme, the proposed detection method is able to cancel part of the interference and compensate for interference by adding the feedback of output to the input of DFE-EFF. Secondly, using an EFF, the proposed method can reduce the correlation of the error signal, which cannot be reduced by a feedforward filter (FFF) or a feedback filter (FBF). Finally, the COA can calculate the realizable optimum coefficient of

each filter.

The rest of this paper is organized as follows. In section 2, we describe the general structure of the MIMO DS-UWB system model, while the focuses of section 3 are divided into three parts: the theoretical analysis of the MMSE performance MIC DFE-EFF, the illumination of COA and the presentation of DA. In section 4, the computational complexity for comparisons of different MUDs is investigated. In section 5, we present simulation results for the proposed MUD over Nakagami fading channel, and our conclusions are made in section 6.

## 2. System Model

We consider a  $K$  users model for a synchronous MIMO DS-UWB communication system with  $M_T$  transmit and  $N_R$  receive antennas over indoor multipath channel, where each user uses unique spreading sequence. The transmitted signal generated by the  $k$ th user from  $m_k$ th transmit antenna is given by

$$x_{m_k}^{(k)}(t) = \sqrt{E_{m_k}^{(k)}} b_{m_k}^{(k)} s_{m_k}^{(k)}(t) \quad (1)$$

where  $E_{m_k}^{(k)}$ ,  $b_{m_k}^{(k)}$ , and  $s_{m_k}^{(k)}(t)$  denote, respectively, the signal energy per bit, the data modulated by binary phase shift keying (BPSK), and the spreading waveform which is given by

$$s_{m_k}^{(k)}(t) = \frac{1}{\sqrt{N_S}} \sum_{i=1}^{N_S} c_{m_k}^{(k)}(i) p(t - iT_c) \quad (2)$$

where  $N_S$ ,  $c_{m_k}^{(k)}(i)$ ,  $T_c$ , and  $p(t)$  denote, respectively, the length of spreading sequence, the  $i$ th spreading code chip, the chip duration, and the second derivative of Gaussian pulse [7] is given by

$$p(t) = \left[ 1 - 4\pi \left( \frac{t}{\tau} \right) \right] \exp \left[ -2\pi \left( \frac{t}{\tau} \right)^2 \right] \quad (3)$$

where  $\tau$  is the pulse shape parameter.

The UWB multipath channel of the  $k$ th user can be described by its channel impulse response (CIR) with Nakagami small-scale fading [17]. The CIR of the  $k$ th user from the  $m_k$ th transmit antenna to the  $n_k$ th receive antenna is given by

$$h_{n_k, m_k}^{(k)}(t) = \sum_{l=0}^L \sum_{p=0}^P \alpha_{n_k, m_k, p, l}^{(k)} \exp(j\phi_{n_k, m_k, p, l}^{(k)}) \delta(t - T_{n_k, m_k, l}^{(k)} - \tau_{n_k, m_k, p, l}^{(k)}) \quad (4)$$

where  $L$  and  $P$  are the number of clusters and rays (i.e. component in cluster) of the  $k$ th user, respectively;  $\alpha_{n_k, m_k, p, l}^{(k)}$ ,  $T_{n_k, m_k, l}^{(k)}$ , and  $\tau_{n_k, m_k, p, l}^{(k)}$  in (4) denote, respectively, the multipath gain coefficient of the  $p$ th ray in the  $l$ th cluster, the delay of the  $l$ th cluster, and the delay of the  $p$ th ray relative to the  $l$ th cluster arrival time  $T_{n_k, m_k, l}^{(k)}$ , of the  $k$ th user from the  $m_k$ th transmit antenna to the  $n_k$ th receive antenna. The phases  $\phi_{n_k, m_k, p, l}^{(k)}$  is taken as a uniformly distributed random variable from the range  $[0, 2\pi)$ .

The distribution of the small-scale amplitudes is Nakagami

$$pdf(\alpha) = \frac{2}{\Gamma(m)} \left( \frac{m}{\Omega} \right)^m \alpha^{2m-1} \exp \left( -\frac{m}{\Omega} \alpha^2 \right) \quad (5)$$

where  $m \geq 1/2$  is the Nakagami  $m$ -factor,  $\Gamma(m)$  is the gamma function, and  $\Omega$  is the mean-square value of the amplitude.

Finally, the received signal of  $k$ th user at the  $n_k$ th receive antenna over the multipath Nakagami fading channel can be obtained as

$$r_{n_k}^{(k)}(t) = \sum_{m_k=1}^{M_T} x_{m_k}^{(k)}(t) \otimes h_{n_k, m_k}^{(k)}(t) + z_{n_k}^{(k)}(t) \tag{6}$$

where  $\otimes$  denotes linear convolution and  $z_{n_k}^{(k)}(t)$  is zero-mean additive white Gaussian noise.

### 3. MIC DFE-EFF (COA) with DA

The DFE which is considered as an effective equalization technique has been widely employed in the modern wideband communications during this decade [12][13][14][15][16][18][19].

In this work, we focus on the MIC DFE-EFF (COA) with DA to assist MUD for MIMO DS-UWB system. The proposed MIC DFE-EFF (COA) with DA consists of three aspects—MIC DFE-EFF [15], COA and DA, which will be discussed in the following.

#### 3.1 MMSE performance of MIC DFE-EFF

The investigation of MMSE performance of MIC DFE-EFF should be based on the DFE and DFE-EFF. The block diagram of the DFE-EFF is shown in Fig. 1. The FFF coefficients  $W_f$ , the FBF coefficients  $W_b$  and the EFF coefficients  $W_e$  for  $K$  users can be defined as,

$$W_f = [w_f^{(1)}, w_f^{(2)}, \dots, w_f^{(K)}]^T; w_f^{(k)} = [w_{f,1}^{(k)}, w_{f,2}^{(k)}, \dots, w_{f,N_f}^{(k)}]^T \tag{7}$$

$$W_b = [w_b^{(1)}, w_b^{(2)}, \dots, w_b^{(K)}]^T; w_b^{(k)} = [w_{b,1}^{(k)}, w_{b,2}^{(k)}, \dots, w_{b,N_b}^{(k)}]^T \tag{8}$$

$$W_e = [w_e^{(1)}, w_e^{(2)}, \dots, w_e^{(K)}]^T; w_e^{(k)} = [w_{e,1}^{(k)}, w_{e,2}^{(k)}, \dots, w_{e,N_e}^{(k)}]^T \tag{9}$$

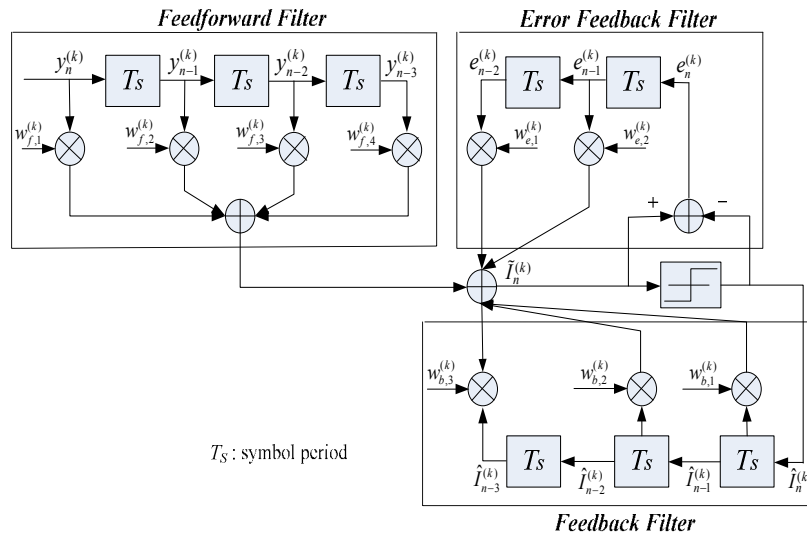


Fig. 1. Schematic of the DFE-EFF

where  $N_f$ ,  $N_b$  and  $N_e$  are the number of forward, feedback and error taps, respectively. The superscript  $(k)$  in (7)-(9) is used to indicate the  $k$ th user for the purpose of MUD,  $k = 1, \dots, K$ . The inputs of forward, feedback and error filters with time index  $n$  are given, respectively, as follows,

$$\mathbf{Y}_n = [\mathbf{y}_n^{(1)}, \mathbf{y}_n^{(2)}, \dots, \mathbf{y}_n^{(K)}]^T; \mathbf{y}_n^{(k)} = [y_n^{(k)}, y_{n-1}^{(k)}, \dots, y_{n-N_f+1}^{(k)}]^T \quad (10)$$

$$\hat{\mathbf{I}}_n = [\hat{\mathbf{I}}_n^{(1)}, \hat{\mathbf{I}}_n^{(2)}, \dots, \hat{\mathbf{I}}_n^{(K)}]^T; \hat{\mathbf{I}}_n^{(k)} = [\hat{I}_{n-1}^{(k)}, \hat{I}_{n-2}^{(k)}, \dots, \hat{I}_{n-N_b}^{(k)}]^T \quad (11)$$

$$\mathbf{E}_n = [\mathbf{e}_n^{(1)}, \mathbf{e}_n^{(2)}, \dots, \mathbf{e}_n^{(K)}]^T; \mathbf{e}_n^{(k)} = [e_{n-1}^{(k)}, e_{n-2}^{(k)}, \dots, e_{n-N_e}^{(k)}]^T \quad (12)$$

Here, a architecture is presented for adaptive MUD in multiuser MIMO DS-UWB (i.e. MIC DFE-EFF) which has better performance than the DFE, DFE-EFF and parallel interference cancellation (PIC) [20]. Compared with DFE, DFE-EFF and PIC, MIC DFE-EFF has the merit of DFE-EFF which can cancel interference in the output of equalizer, but it also can compensate for interference (i.e. using MIC) by adding the feedback of output to the input of DFE-EFF. This schematic is shown in Fig. 2. The signals which come from the Nakagami fading channel are sent to the matched filter, carrying the information of each user. The information is then sent to the MIC DFE-EFF multiuser detector for further interference cancellation, and also for the reduction of the correlation of the error signal. As shown in Fig. 2, the output of the decision device is fed back to the input of DFE-EFF. So, MIC DFE-EFF can cancel more multiple access interference (MAI) at the input of DFE-EFF. In this schematic, the input of the FFF is no more given by (10), but by what is given below,

$$\mathbf{Y}_n = [\mathbf{y}_n^{(1)} - \mathbf{q}^{(1)}, \mathbf{y}_n^{(2)} - \mathbf{q}^{(2)}, \dots, \mathbf{y}_n^{(K)} - \mathbf{q}^{(K)}]^T \quad (13)$$

$$\mathbf{y}_n^{(k)} - \mathbf{q}^{(k)} = [y_n^{(k)} - q_{n-1}^{(k)}, y_{n-1}^{(k)} - q_{n-2}^{(k)}, \dots, y_{n-N_f+1}^{(k)} - q_{n-N_f}^{(k)}]$$

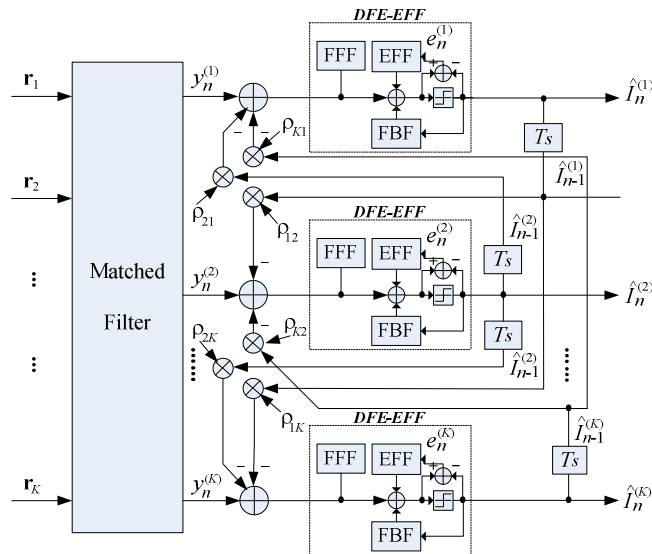


Fig. 2. Schematic of the basic MIC DFE-EFF MUD

where  $q_n^k = \sum_{i=1, i \neq k}^K \rho_{ik} \hat{I}_{n-1}^{(i)}$  represents the MAI coming from all users other than the desired user  $k$ .  $\rho_{ik}$  is the spatial signature cross-correlation parameter [21][22] between any two users  $i$  and  $k$ .

The output of the conventional DFE without MIC and EFF for  $K$  users before decision device with time index  $n$  is represented by

$$\tilde{\mathbf{I}}_n = \text{diag}([\mathbf{W}_f \mathbf{W}_b][\mathbf{Y}_n \hat{\mathbf{I}}_n]^T) = \text{diag}(\mathbf{W}_f \mathbf{Y}_n^T + \mathbf{W}_b \hat{\mathbf{I}}_n^T) \quad (14)$$

where  $\mathbf{Y}_n$  is given by (10).

The optimal weight  $\mathbf{W}_{opt}$  of conventional DFE is then obtained by

$$\mathbf{W}_{opt} = \mathbf{R}_c^{-1} \mathbf{g} \quad (15)$$

where  $\mathbf{g} = E[\hat{I}_n^{(k)} \mathbf{X}_n^{(k)}]$  is a vector of size  $(N_f + N_b) \times 1$  containing cross-correlation between the desired signal  $\hat{I}_n^{(k)}$  and input signal  $\mathbf{X}_n^{(k)}$  ( $\mathbf{X}_n^{(k)} = [\mathbf{y}_n^{(k)} \hat{\mathbf{I}}_n^{(k)}]^T$ ) of the  $k$ th user with time index  $n$ , and  $\mathbf{R}_c = E[\mathbf{X}_n^{(k)} (\mathbf{X}_n^{(k)})^T]$  is a matrix of size  $(N_f + N_b) \times (N_f + N_b)$  containing the auto-correlation of input signal  $\mathbf{X}_n^{(k)}$  [15].

Finally, MMSE is obtained as follows,

$$J_{\min} = E[(\hat{I}_n^{(k)})^2] - \mathbf{g}^T \mathbf{R}_c^{-1} \mathbf{g} \quad (16)$$

Similarly, the output vector of MIC DFE-EFF for  $K$  users before the decision device with time index  $n$  is represented by

$$\tilde{\mathbf{I}}_n = \text{diag}([\mathbf{W}_f \mathbf{W}_b \mathbf{W}_e][\mathbf{Y}_n \hat{\mathbf{I}}_n \mathbf{E}_n]^T) = \text{diag}(\mathbf{W}_f \mathbf{Y}_n^T + \mathbf{W}_b \hat{\mathbf{I}}_n^T + \mathbf{W}_e \mathbf{E}_n^T) \quad (17)$$

note that here the  $\mathbf{Y}_n$  isn't given by (10) shown in Fig. 1, but by (13) shown in Fig. 2, because of the proposed schematic of MIC and EFF.

Similarly, the optimal weight  $\hat{\mathbf{W}}_{opt}$  of MIC DFE-EFF is obtained as follows,

$$\hat{\mathbf{W}}_{opt} = \hat{\mathbf{R}}_c^{-1} \hat{\mathbf{g}} \quad (18)$$

where  $\hat{\mathbf{g}} = E[\hat{I}_n^{(k)} \hat{\mathbf{X}}_n^{(k)}]$  is a vector of size  $(N_f + N_b + N_e) \times 1$  containing cross-correlation between the desired signal  $\hat{I}_n^{(k)}$  and input signal  $\hat{\mathbf{X}}_n^{(k)}$  ( $\hat{\mathbf{X}}_n^{(k)} = [\mathbf{y}_n^{(k)} \hat{\mathbf{I}}_n^{(k)} \mathbf{e}_n^{(k)}]^T$ ) of the  $k$ th user with time index  $n$ , and  $\hat{\mathbf{R}}_c = E[\hat{\mathbf{X}}_n^{(k)} (\hat{\mathbf{X}}_n^{(k)})^T]$  is a matrix of size  $(N_f + N_b + N_e) \times (N_f + N_b + N_e)$  containing the auto-correlation of input signal  $\hat{\mathbf{X}}_n^{(k)}$  [15].

Thus, the MMSE of the MIC DFE-EFF can be represented by

$$J'_{\min} = E[(\hat{I}_n^{(k)})^2] - \mathbf{g}^T \mathbf{R}_c^{-1} \mathbf{g} - (\sigma_e^2)^{-1} (E[e_n^{(k)} e_{n-1}^{(k)}])^2 \quad (19)$$

where error signal  $e_n^{(k)} = \tilde{I}_n^{(k)} - \hat{I}_n^{(k)}$  and  $\sigma_e^2 = E[(e_{n-1}^{(k)})^2]$ .

By looking to (19), the first two terms correspond to the MMSE of conventional DFE expressed in(16). Since the last term of (19) is always positive, the MMSE of the MIC DFE-EFF is always smaller than that of the DFE by  $(\sigma_e^2)^{-1} (E[e_n^{(k)} e_{n-1}^{(k)}])^2$ .

### 3.2 Coefficient Optimization Algorithm (COA)

From the investigation of the above, the optimal weighting coefficient of MIC DFE-EFF is (18); however it is only a theoretical value and its computational efficiency is too complex to handle. In this section, the COA is proposed to obtain a practical optimum value of  $\hat{\mathbf{W}}_{opt}$ . Here the MIC DFE-EFF with the COA can be termed MIC DFE-EFF (COA).

In this section, the COA which is under the condition of the user  $k$  is discussed; and superscript  $(k)$  is omitted for convenience. As shown in Fig. 1 and Fig. 2, it follows that the output of MIC DFE-EFF for  $k$ th user before decision device with time index  $n$  can be expressed as

$$\tilde{I}_n = \sum_{j=-K_1}^0 w_{f,j} y_{n-j} + \sum_{k=1}^{K_2} w_{b,k} \hat{I}_{n-k} + w_e (\tilde{I}_{n-1} - \tilde{I}_{n-1}) \quad (20)$$

where  $\tilde{I}_n$  is an estimate of the  $n$ th time index for the user  $k$ .  $w_e$ ,  $w_{f,j}$  and  $w_{b,k}$  ( $j = -K_1, \dots, -1, 0$ ;  $k = 1, 2, \dots, K_2$ ) are tap coefficients of the EFF FFF and FBF, respectively.  $\hat{I}_{n-k}$  ( $k = 1, 2, \dots, K_2$ ) are previously detected symbols. The equalizer is assumed to have  $(K_1 + 1)$  taps in FFF,  $K_2$  taps in FBF, and 1 tap in EFF. (For minimal increase of hardware complexity, 1 tap of EFF is

used.)

Now, we focus on MSE criterion to obtain a mathematically tractable optimization of the proposed equalizer coefficients. Based on our assumption that previously detected symbols in feedback filter are correct, minimizing the MSE of cost function

$$J(w_f, w_b, w_e) = E[|I_n - \tilde{I}_n|^2] \quad (21)$$

leads to the following set of equations (because of the orthogonality principle of MSE [23]) for computing coefficients of FFF, FBF, and EFF,

$$J_{\min}(w_f, w_b, w_e) \Rightarrow \begin{cases} E[\xi_n y_{n-l}^*] = 0, & -K_1 \leq l \leq 0 \\ E[\xi_n \hat{I}_{n-l}^*] = 0, & 1 \leq l \leq -K_2 \\ E[\xi_n (\tilde{I}_{n-l} - \hat{I}_{n-l}^*)] = 0, & l = 1 \end{cases}$$

$$\Rightarrow \begin{cases} E[I_n y_{n-l}^*] = E[\tilde{I}_n y_{n-l}^*], & -K_1 \leq l \leq 0 \\ E[I_n \hat{I}_{n-l}^*] = E[\tilde{I}_n \hat{I}_{n-l}^*], & 1 \leq l \leq -K_2 \\ E[I_n \tilde{I}_{n-l}^*] + E[\tilde{I}_n \hat{I}_{n-l}^*] = E[I_n \hat{I}_{n-l}^*] + E[\tilde{I}_n \tilde{I}_{n-l}^*], & l = 1 \end{cases} \quad (22)$$

where  $\xi_n = I_n - \tilde{I}_n$ , and  $I_n$  is the  $n$ th transmitted information symbol. Note that the information symbols are uncorrelated:  $E[I_n \hat{I}_l] = \delta_{n,l}$ . The above set of equations (22) is our proposed coefficient optimization algorithm in MMSE criterion. Optimum coefficient values of FFF, FBF and EFF can be obtained by calculating the set of equations (22), and then the optimum coefficient  $\hat{\mathbf{W}}_{opt}$  which consists of coefficients of FFF, FBF and EFF will be also obtained.

Firstly, the coefficients of FFF are calculated by substituting (20) into the first equation of (22). Here,  $y_n$  comes from the matched filter shown in Fig. 2. Because of the knowledge of matched filter coefficients,  $y_n$  is defined as

$$y_n = \sum_{u=0}^L f_u I_{n-u} + z_n \quad (23)$$

where  $f_u$  are matched filter coefficients ( $f_u = 0$ , for  $u < 0$  and  $u > L$ ); noise signal  $z_n \sim \mathcal{N}(0, N_0)$  accounts for AWGN. Since the information symbols are uncorrelated, they are independent of  $z_n$ . Thus, the left-hand side of the first equation of (22) can be rewritten by substituting (23) into it,

$$E[I_n y_{n-l}^*] = E \left[ I_n \sum_{u=0}^L f_u^* I_{n-l-u}^* + z_{n-l}^* \right] = \sum_{u=0}^L f_u^* E[I_n I_{n-l-u}^*] = \sum_{u=0}^L f_u^* \delta_{l+u} = f_{-l}^*, \quad l = -K_1, \dots, -1, 0 \quad (24)$$

In order to solve the right-hand side of the first equation of (22),  $E[y_{n-j} y_{n-l}^*]$  and  $E[\hat{I}_{n-j} \hat{I}_{n-l}^*]$  is needed. We have

$$E[y_{n-j} y_{n-l}^*] = E \left[ \left( \sum_{u=0}^L f_u I_{n-j-u} + z_{n-j} \right) \left( \sum_{v=0}^L f_v^* I_{n-l-v}^* + z_{n-l}^* \right) \right]$$

$$= \sum_{u=0}^L f_u f_{u+j-l}^* + N_0 \delta_{l,j}$$

$$= \psi_{l,j}, \quad l, j = -K_1, \dots, -1, 0 \quad (25)$$

where  $E[z_{n-j} z_{n-l}^*] = \delta_{l,j}$ .

$$E[\hat{I}_{n-j} \hat{I}_{n-l}^*] = \hat{I}_{n-j} \sum_{u=0}^L f_u^* E[I_{n-l-u}] = 0 \quad (26)$$

Substituting (20), (25) and (26) into the right-hand side of the first equation of (22), we find

below

$$E[\tilde{I}_n \mathcal{Y}_{n-l}^*] = \sum_{j=-K_1}^0 w_{f,j} \psi_{l,j} + w_e E[\tilde{I}_{n-1} \mathcal{Y}_{n-l}^*] \quad (27)$$

We exploit recurrence formula and derive the following equation from (27),

$$E[\tilde{I}_1 \mathcal{Y}_{n-l}^*] = \sum_{j=-K_1}^0 w_{f,j} \left( \sum_{u=0}^L f_u f_{u+n-1+j-l}^* + N_0 \delta_{n-1+j-l} \right) \quad (28)$$

Observing (27) and (28), we employ the recursion algorithm, and then (27) can be rewritten as

$$E[\tilde{I}_n \mathcal{Y}_{n-l}^*] = \sum_{i=0}^{n-1} \left[ \sum_{j=-K_1}^0 w_{f,j} \left( \sum_{u=0}^L f_u f_{u+i+j-l}^* + N_0 \delta_{i+j-l} \right) \right] w_e^i \quad (29)$$

Substituting (24) and (29) into the first equation of (22), we can obtain the following relation equation between  $w_{fj}$  and  $w_e$ ,

$$f_{-l}^* = \sum_{j=-K_1}^0 w_{f,j} \sum_{i=0}^{L+K_1} \psi w_e^i, \quad -K_1 \leq l \leq 0 \quad (30)$$

where  $\psi = \sum_{u=0}^L f_u f_{u+i+j-l}^* + N_0 \delta_{i+j-l}$ . Because the value of  $w_e$  is small, we do not allow the exponential of  $w_e$  more than 1 for simplification in (30). Therefore,

$$\begin{aligned} \sum_{j=-K_1}^0 w_{f,j} \psi &= \frac{f_{-l}^*}{1+w_e} \\ \Rightarrow \mathbf{w}_f &= \mathbf{\Psi}^{-1} \frac{\mathbf{f}}{1+w_e}, \quad -K_1 \leq l \leq 0 \end{aligned} \quad (31)$$

where  $\mathbf{w}_f = [w_{f,-K_1}, \dots, w_{f,0}]^T$ ,  $\mathbf{f} = [f_{K_1}^*, \dots, f_0^*]^T$ , and  $\mathbf{\Psi} = \begin{bmatrix} \psi_{-K_1, -K_1} & \dots & \psi_{-K_1, 0} \\ \vdots & \ddots & \vdots \\ \psi_{0, -K_1} & \dots & \psi_{0, 0} \end{bmatrix}$ .

Secondly, in the same way as the above, the following relation equation between  $w_{b,l}$  and  $w_e$  can be derived from the second equation of (22),

$$\begin{aligned} \sum_{i=0}^{K_1+K_2} \left( \sum_{j=-K_1}^0 w_{f,j} f_{l-i-j} + w_{b,l-i} + w_e \delta_{l+i-1} \right) w_e^i &= 0, \quad 1 \leq l \leq K_2 \\ \Rightarrow \begin{cases} w_{b,l} = \sum_{i=0}^{K_1+K_2} \sum_{j=-K_1}^0 w_{f,j} f_{l-i-j} w_e^i - w_e, & l=1 \\ w_{b,l} = - \sum_{j=-K_1}^0 w_{f,j} f_{l-j}, & 2 \leq l \leq K_2 \end{cases} \end{aligned} \quad (32)$$

For simplification, then

$$\begin{aligned} w_{b,l} &= \sum_{j=-K_1}^0 w_{f,j} f_{l-j} - w_e \delta_{l-1} = -\mathbf{f}_l \mathbf{\Psi}^{-1} \frac{\mathbf{f}}{1+w_e} - w_e \delta_{l-1}, \quad 1 \leq l \leq K_2 \\ \Rightarrow \mathbf{w}_b &= -\mathbf{F} \mathbf{\Psi}^{-1} \frac{\mathbf{f}}{1+w_e} - \mathbf{A} w_e, \quad 1 \leq l \leq K_2 \end{aligned} \quad (33)$$

where  $\mathbf{f}_l = [f_{l+K_1}, \dots, f_l]$ ,  $\mathbf{F} = [\mathbf{f}_1, \dots, \mathbf{f}_{K_2}]^T$  and  $\mathbf{A} = [1, 0, \dots, 0]_{K_2 \times 1}^T$ .

According to the above recursion algorithm shown in (29), it thus follows from last equation of (22), that



$$\sum_{i=0}^{K_1} \sum_{j=-K_1}^0 w_{f,j} f_{-(l+i+j)}^* w_e^j = \sum_{i'=0}^{K_1-1} \left[ \sum_{j=-K_1}^0 w_{f,j} \left( \sum_{m=-K_1}^0 w_{f,m} \psi + \sum_{m=1}^{K_2} w_{b,m} f_{l+m-j+i'} \right) + \sum_{m=1}^{K_2} w_{b,m} \left( \sum_{j=-K_1}^0 w_{f,m} f_{-l+m-j+i'}^* \right. \right. \\ \left. \left. + w_e \delta_{l+m+i'} \right) - w_e^2 \sum_{i=0}^{K_1-1} \sum_{j=-K_1}^0 w_{f,j} f_{-(l+i+j+i')}^* w_e^j \right] (w_e^2)^{i'}, \quad l=1 \quad (34)$$

In what follows, we solve the simplification problem explicitly in manners similar to (31) and (33) (i.e. the exponential of  $w_e$  is fixed 1 for simplification). Hence (34) can be rewritten as

$$w_e \left[ \sum_{j=-K_1}^0 w_{f,j} f_{-2-j}^* - w_{b,2} \right] + \sum_{j=-K_1}^0 w_{f,j} \left[ f_{-1-j}^* - \sum_{m=1}^{K_2} w_{b,m} (f_{1+m-j} + f_{-1+m-j}^*) - \sum_{m=-K_1}^0 w_{f,m} \psi \right] = 0 \quad (35)$$

Substituting (31) and (33) into (35), we obtain the final expression of  $w_e$

$$w_e = \frac{X_1 - \sqrt{X_1^2 - 4[X_1 + \mathbf{f}^T (\Psi^T)^{-1} \mathbf{f}'_1] X_2}}{2[X_1 + \mathbf{f}^T (\Psi^T)^{-1} \mathbf{f}'_1]} \quad (36)$$

where  $X_1 = \mathbf{f}^T (\Psi^T)^{-1} (\mathbf{f}'_1 + \mathbf{f}'_2 + \mathbf{F}'\mathbf{A})$ ,  $X_2 = \mathbf{f}^T (\Psi^T)^{-1} (\mathbf{f} - \mathbf{F}'\mathbf{F}'\Psi^{-1} - \mathbf{f}'_1)$ .  $\mathbf{f}'_i$  is a vector of size  $(K_1 + 1) \times 1$  containing  $\{f_{-i-j}\}$  ( $i=1$  or  $2$ ).  $\mathbf{F}'$  is a vector of size  $(K_1 + 1) \times K_2$  containing  $\{f_{m-1-j}^* + f_{m+1-j}\}$ .

Finally, by substituting (36) into (31) and (33), we find  $\mathbf{w}_f$  and  $\mathbf{w}_b$ , respectively.

### 3.3 Differencing Algorithm (DA)

In section 3.1, by adding the proposed MIC and EFF, the MIC DFE-EFF MUD can obtain better BER performance than conventional DFE. We analyze the MMSE performance of MIC DFE-EFF and give a mathematical expression of the improvement of the system performance. In section 3.2, we employ the proposed COA to obtain the optimal coefficient  $\hat{\mathbf{w}}_{opt}$ . These two sections illustrate the system performance of MIC DFE-EFF, but the computational complexity has not been discussed. The computational complexity is one of the most critical aspects in real-time system. It determines the costs in terms of number of necessary operations per second. Also, it determines the power consumption which is critical in mobile devices. In this section, we will refer to computational efficiency of above MUDs and present a complexity reduction algorithm.

According to the above investigation, we know MIC DFE-EFF and MIC DFE-EFF (COA) outperform DFE and DFE-EFF at BER performance. But they are at the cost of increasing computational complexity. Compared with DFE, it requires almost more than  $K(N_f + 1) + K(K - 1)$  multiplications and  $K(K - 1) + K(N_f + 1)$  additions per bit.

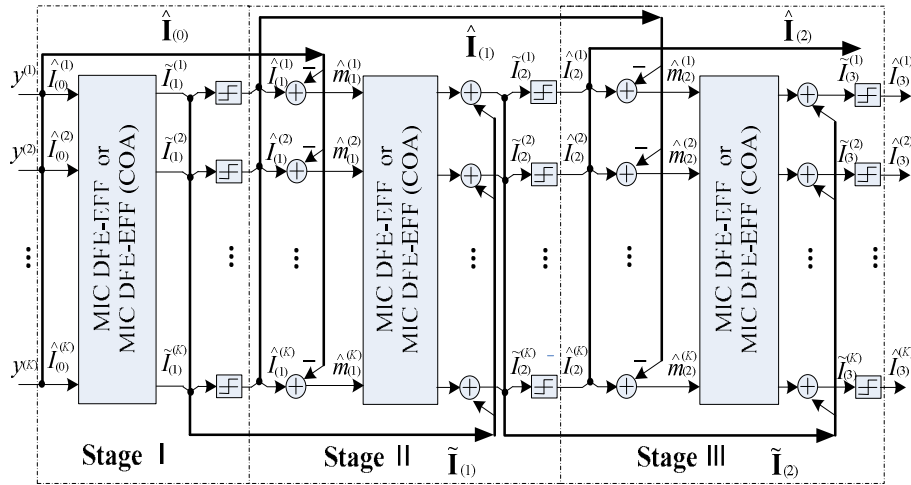
Here, we propose a differencing algorithm (DA) to solve the computational complexity problem of multistage MIC DFE-EFF and MIC DFE-EFF (COA). The complexity reduction algorithm exploits the multistage character of MIC DFE-EFF and MIC DFE-EFF (COA). If the estimated data do not change in two consecutive stages, they will not change in all following stages. Therefore, redundant interference calculations can be avoided in later stages by this means. In this paper, we present a DA in multiuser MIMO DS-UWB system for computational complexity reduction of MUD.

In this algorithm, we exploit the differencing structure in multistage detection of MIC DFE-EFF and MIC DFE-EFF (COA), avoiding unnecessary repeated calculations of certain terms in consecutive stages. In this structure, each stage detector employs MIC DFE-EFF or MIC DFE-EFF (COA) detection method as described above. The input of each stage in this structure is a differencing vector, which is generated by subtracting the input decision from the previous decision, as shown in Fig. 3. By analyzing the principle of DA, we find, in the multistage detection system, the output of the  $l$ th iteration is (time index  $n$  can be omitted):

$$\begin{aligned}\tilde{\mathbf{I}}_{(l)} &= \hat{\mathbf{W}}^T \hat{\mathbf{X}}_{(l-1)} \\ \hat{\mathbf{I}}_{(0)} &= \text{sign}(\mathbf{y}) \\ \hat{\mathbf{I}}_{(l)} &= \text{sign}(\tilde{\mathbf{I}}_{(l)}) \\ \hat{\mathbf{X}}_{(0)} &= [\text{sign}(\mathbf{y}) \ \hat{\mathbf{I}}_{(1)} \ \mathbf{E}] \\ \hat{\mathbf{X}}_{(l-1)} &= [\text{sign}(\tilde{\mathbf{I}}_{(l-1)}) \ \hat{\mathbf{I}}_{(l)} \ \mathbf{E}]\end{aligned}$$

After  $l$  iterations, it is more likely to find  $\hat{\mathbf{I}}_{(l)} = \hat{\mathbf{I}}_{(l-1)}$ , reflecting the convergence of the iterative method. We also find that, instead of dealing with each vector  $\hat{\mathbf{I}}_{(l)}$ , as in the formulas above, we can deal with the difference of the estimated bits vectors in two consecutive stages. In other words, the input of each stage becomes  $\hat{\mathbf{m}}_{(l)} = \hat{\mathbf{I}}_{(l)} - \hat{\mathbf{I}}_{(l-1)}$ , namely the differencing vector. Thus, all above formulas can be rewritten as

$$\begin{aligned}\tilde{\mathbf{I}}_{(l)} &= \hat{\mathbf{W}}^T \hat{\mathbf{X}}_{(l-1)} = \tilde{\mathbf{I}}_{(l-1)} + \hat{\mathbf{W}}^T [\hat{\mathbf{m}}_{(l-1)} \ \hat{\mathbf{m}}_{(l)} \ \mathbf{E}] \\ \hat{\mathbf{I}}_{(l)} &= \text{sign}(\tilde{\mathbf{I}}_{(l)}) \\ \hat{\mathbf{X}}_{(l-1)} &= [\text{sign}(\tilde{\mathbf{I}}_{(l-1)}) \ \hat{\mathbf{I}}_{(l)} \ \mathbf{E}] \\ \hat{\mathbf{m}}_{(l)} &= \hat{\mathbf{I}}_{(l)} - \hat{\mathbf{I}}_{(l-1)}\end{aligned}$$



**Fig. 3.** Schematic of MIC DFE-EFF/MIC DFE-EFF (COA) with DA

Employing this DA, many more computations can be saved by dealing with  $\hat{\mathbf{m}}_{(l)}$ , since more terms in the vector  $\hat{\mathbf{m}}_{(l)}$  tend to be zero after several stages. Furthermore, all the non-zero terms in  $\hat{\mathbf{m}}_{(l)}$  are equal to constant  $\pm 2$ . The constant multiplication can be carried out by arithmetic shifts in hardware [24]. So, the complexity of the detection system as well as the power of detector implementation can be reduced by using the DA detection. Ultimately, because our DA which subtracts two consecutive stages is a linear transformation, the BER after each stage remains almost unchanged, compared with a multistage MIC DFE-EFF or MIC DFE-EFF (COA) detector without DA.

#### 4. Computational Complexity Investigation

In this section, quantitative analysis is to express the computational complexity for comparison of different MUD methods, by using the float-point (flop) operation measure [25]. In this flop operation measure, we define addition and multiplication operations (ops) as add and mul, respectively. The complex number ops are  $C_{\text{add}} = 2\text{add} = 2\text{ops}$ ,  $C_{\text{mul}} = 4\text{mul} + 2\text{add} = 6\text{ops}$ , since the signal may be the complex number. The complexity  $O$  measured in number of ops is calculated for the different detection methods.  $S, M_T, N_R, K, N_f, N_b$  can be respectively defined as the number of stages, transmitter antennas, receiver antennas, users, FFF taps and FBF taps.

1) PIC: For conventional PIC algorithm, the computational complexity is given by

$$\begin{aligned} O_{\text{PIC}} &= SM_T N_R K \{C_{\text{mul}} + 4K + C_{\text{add}}(K-1) + 2K[4K + C_{\text{add}}(2K-1)]\} \\ &= SM_T N_R K (16K^2 + 2K + 4)\text{ops} \end{aligned} \quad (37)$$

2) DFE: For conventional DFE algorithm, the computational complexity is given by

$$\begin{aligned} O_{\text{DFE}} &= SM_T N_R K [10C_{\text{mul}} N_f + C_{\text{add}}(N_f - 1) + 10C_{\text{mul}} N_b + C_{\text{add}}(N_b - 1) + C_{\text{add}}] \\ &= SM_T N_R K (62N_f + 62N_b - 2)\text{ops} \end{aligned} \quad (38)$$

3) DFE-EFF: For one tap of EFF, the computational complexity of DFE-EFF is

$$\begin{aligned} O_{\text{DFE-EFF}} &= SM_T N_R K [10C_{\text{mul}} N_f + C_{\text{add}}(N_f - 1) + 10C_{\text{mul}} + C_{\text{add}}(N_b - 1) + 10C_{\text{mul}} N_b + C_{\text{add}} + 2C_{\text{add}}] \\ &= SM_T N_R K (62N_f + 62N_b + 62)\text{ops} \end{aligned} \quad (39)$$

4) MIC DFE-EFF: Before the complexity reduction, the computational complexity of MIC DFE-EFF is

$$\begin{aligned} O_{\text{MIC DFE-EFF}} &= SM_T N_R K \left[ \overbrace{10C_{\text{mul}} + C_{\text{add}} + 2C_{\text{add}}}^{\text{EFF}} + \overbrace{(10C_{\text{mul}} + C_{\text{add}})N_f + (C_{\text{mul}} + C_{\text{add}})(K-1)}^{\text{MIC}} \right. \\ &\quad \left. + \overbrace{10C_{\text{mul}}(N_f + N_b) + C_{\text{add}}(N_f + N_b - 2)}^{\text{DFE}} \right] \\ &= SM_T N_R K (124N_f + 62N_b + 8K + 54)\text{ops} \end{aligned} \quad (40)$$

5) MIC DFE-EFF with DA: The complexity reduction algorithm—MIC DFE-EFF with DA exploits convergence of the differencing vector  $\hat{\mathbf{m}}_{(l)}$ , and its computational complexity is

$$\begin{aligned} O_{\text{MIC DFE-EFF with DA}} &= \sum_{i=1}^S \overbrace{(1-\alpha_i)}^{\text{DA}} M_T N_R K \left[ \overbrace{10C_{\text{mul}} + C_{\text{add}} + 2C_{\text{add}}}^{\text{EFF}} + \overbrace{(10C_{\text{mul}} + C_{\text{add}})N_f + (C_{\text{mul}} + C_{\text{add}})(K-1)}^{\text{MIC}} \right. \\ &\quad \left. + \overbrace{10C_{\text{mul}}(N_f + N_b) + C_{\text{add}}(N_f + N_b - 2)}^{\text{DFE}} \right] \\ &= M_T N_R K \sum_{i=1}^S (1-\alpha_i) (124N_f + 62N_b + 8K + 54)\text{ops} \end{aligned} \quad (41)$$

where  $\alpha_i$  ( $\alpha_1 = 0$ , others  $0 \leq \alpha_i \leq 1$ ) is the ratio of nonchanged estimated data to all estimated data in two consecutive stages, i.e. zero ratio of  $\hat{\mathbf{m}}_{(l)}$ . In 4) and 5), the filter coefficient updating formula is  $\hat{W}_{n+1} = \hat{W}_n + \mu e_n X_n$  ( $\mu$  denotes a step size). Theoretically, though the performance in terms of BER of both MIC DFE-EFF and MIC DFE-EFF with DA is nearly the same, their complexity behaviors is different. Observing (40) and (41) respectively,  $O_{\text{MIC DFE-EFF with DA}} < O_{\text{MIC DFE-EFF}}$  can be seen.

6) MIC DFE-EFF (COA): Before the complexity reduction, the computational complexity of MIC DFE-EFF using proposed COA is

$$\begin{aligned}
O_{\text{MIC DFE-EFF (COA)}} &= SM_{\text{T}}N_{\text{R}} \left\{ C_{\text{mul}} \left[ (C_{\text{mul}} + C_{\text{add}}) N_f^2 - C_{\text{add}} N_f + C_{\text{add}} + C_{\text{add}} (2N_f + N_b + K) \right. \right. \\
&\quad \left. \left. + N_b \left[ N_f (C_{\text{mul}} N_f + C_{\text{add}} N_f - C_{\text{add}} + C_{\text{mul}}) + 2C_{\text{mul}} \right] + C_{\text{mul}} (N_f + K) + (K - 1) \right. \right. \\
&\quad \left. \left. + \overbrace{[C_{\text{add}} + (C_{\text{mul}} + C_{\text{add}})(2N_f + N_b + K) + 2C_{\text{mul}}(N_b + 1)]}^{\text{MIMO MIC DFE-EFF}} \right\} \\
&= SM_{\text{T}}N_{\text{R}} \left[ 48N_f^2(N_b + 1) + 24N_fN_b - 2N_b + 74N_b + 8K + 84 + (K - 1)(16N_f + 20N_b + 8K + 14) \right] \text{ops} \quad (42)
\end{aligned}$$

7) MIC DFE-EFF (COA) with DA: MIC DFE-EFF (COA) with DA exploits proposed COA and complexity reduction algorithm (i.e. DA), and its computational complexity is

$$\begin{aligned}
O_{\text{MIC DFE-EFF (COA) with DA}} &= \sum_{i=1}^S \overbrace{(1 - \beta_i)}^{\text{DA}} M_{\text{T}}N_{\text{R}} \left\{ C_{\text{mul}} \left[ (C_{\text{mul}} + C_{\text{add}}) N_f^2 - C_{\text{add}} N_f + C_{\text{add}} + (K - 1) + C_{\text{mul}} (N_f + K) \right. \right. \\
&\quad \left. \left. + N_b \left[ N_f (C_{\text{mul}} N_f + C_{\text{add}} N_f - C_{\text{add}} + C_{\text{mul}}) + 2C_{\text{mul}} \right] + C_{\text{add}} (2N_f + N_b + K) \right. \right. \\
&\quad \left. \left. + \overbrace{[C_{\text{add}} + (C_{\text{mul}} + C_{\text{add}})(2N_f + N_b + K) + 2C_{\text{mul}}(N_b + 1)]}^{\text{MIMO MIC DFE-EFF}} \right\} \\
&= M_{\text{T}}N_{\text{R}} \sum_{i=1}^S (1 - \beta_i) \left[ 48N_f^2(N_b + 1) + 24N_fN_b - 2N_b + 74N_b + 8K + 84 + (K - 1)(16N_f + 20N_b + 8K + 14) \right] \text{ops} \quad (43)
\end{aligned}$$

where  $\beta_i$  ( $\beta_1 = 0$ , others  $0 \leq \beta_i \leq 1$ ) is defined as  $\alpha_i$ , i.e. zero ratio of  $\hat{\mathbf{m}}_{(i)}$  in MIC DFE-EFF (COA) with DA. Theoretically, though the performance in terms of BER of both MIC DFE-EFF (COA) and MIC DFE-EFF (COA) with DA is almost identical, their complexity behavior is different. Obviously,  $O_{\text{MIC DFE-EFF (COA) with DA}} < O_{\text{MIC DFE-EFF (COA)}}$  can be seen.

In 6) and 7), the filter coefficient updating algorithm is not as in 4) and 5), but our proposed COA. The comparisons of the performance and complexity of the last four kinds of MUD methods (i.e. MIC DFE-EFF, MIC DFE-EFF with DA, MIC DFE-EFF (COA) and MIC DFE-EFF (COA) with DA) will be referred to in the next section.

## 5. Simulation Results

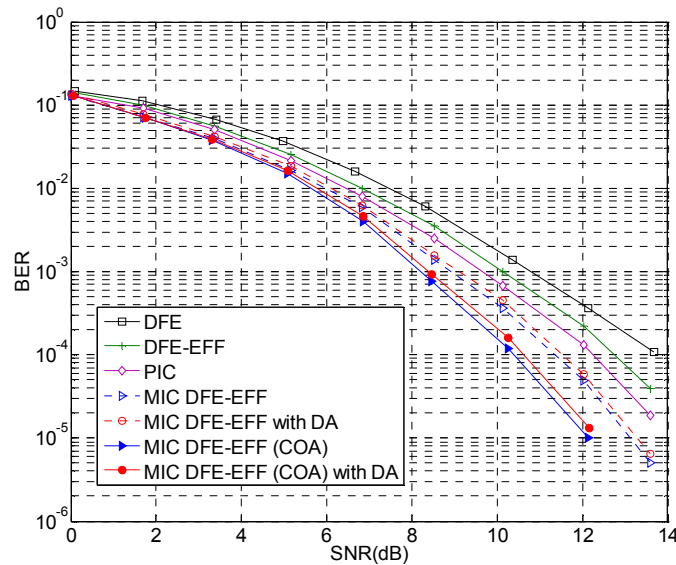
Simulations were conducted to evaluate the performance of our proposed MIC DFE-EFF (COA) with DA for multiuser MIMO DS-UWB communication system.

**Table 1.** Summary of system parameters

Configuration	Parameter	Value/Description
UWB	No. of users	$K = 16$
	No. of transmitter antennas	$M_{\text{T}} = 4$
	No. of receiver antennas	$N_{\text{R}} = 4$
	No. of FFF	$N_f = 4$
	No. of FBF	$N_b = 3$
	No. of EFF	$N_e = 1$
	Modulation	BPSK
	Spread code	Walsh code ( $N_s = 16$ or $32$ )
	Channel model	IEEE 802.15.4a line-of-sight (LOS) Nakagami fading channel
	White noise variance	$\sigma^2 = 0.01$
	Capacity	$K/N_s$
LDPC	Code rate	1/2
	No. of iterations	10
	Codeword length	512 bits

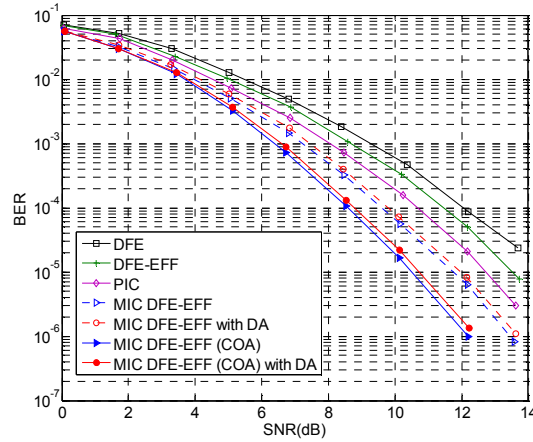
There were 16 users in the UWB system, and the regular low-density parity-check (LDPC) code [26] has been employed for channel coding. The details of the system parameters are summarized in **Table 1**.

In **Fig. 4** and **Fig. 5**, the BER performance of different MUDs in MIMO DS-UWB system with different spread factor is compared, respectively. From the two figures, it can be seen that both MIC DFE-EFF (COA) and MIC DFE-EFF (COA) with DA are much better than other MUDs, because they have proposed COA, MIC and EFF. They have almost the same BER performance, though MIC DFE-EFF (COA) has little better BER performance. Without COA, MIC DFE-EFF and MIC DFE-EFF with DA have worse BER performance, compared with other two schemes using the COA. But they still outperform PIC, DFE and DFE-EFF. We find that the BER performance of both MIC DFE-EFF and MIC DFE-EFF with DA are almost the same. The results prove that the BER after each stage of MIC DFE-EFF with DA hardly change because of the linear transformation that subtracts two consecutive stages, compared with the multistage MIC DFE-EFF detector without DA. And, the same reason can be given to explain why MIC DFE-EFF (COA) and MIC DFE-EFF (COA) with DA have nearly identical BER performance.



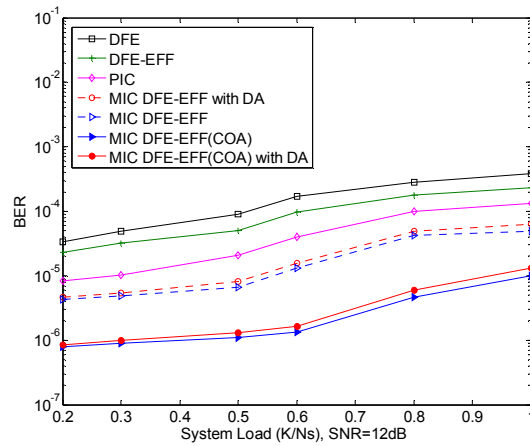
**Fig. 4.** BER performance of different MUDs in Nakagami fading channel, for  $N_S = 16$  and 2 iterations.

In **Fig. 4**, at BER of  $10^{-5}$ , the proposed MIC DFE-EFF (COA) with DA gains more than 1 dB BER performance better than MIC DFE-EFF and MIC DFE-EFF with DA. And, it gains more than 2 dB BER performance better than PIC, DFE-EFF and DFE. As observed in **Fig. 5**, the MIC DFE-EFF (COA) with DA achieves much lower BER with spread factor  $N_s = 32$ . When SNR is nearly 12 dB, it is close to  $10^{-6}$  BER.



**Fig. 5.** BER performance of different MUDs in Nakagami fading channel, for  $N_S = 32$  and 2 iterations.

**Fig. 6** shows the BER performance as the function of the system load [27][28]  $K/N_S$  for different MUDs. The SNR is fixed to 12 dB. It can be seen that with various system load, two proposed MIC DFE-EFF (COA) with and without DA MUDs exhibit much better performance than the others. Though MIC DFE-EFF (COA) has the best performance, the MIC DFE-EFF (COA) and MIC DFE-EFF (COA) with DA have almost the same performance. Compared with the other MUDs, both of the MIC DFE-EFF (COA) with and without DA MUDs gain almost  $10^{-5}$  BER performance, even if the system load is close to 1.



**Fig. 6.** System performance as a function of the system load, in UWB Nakagami fading channel.

**Fig. 7** shows percentage of zeros, which in turn signifies the reduction in complexity in the differencing vector of MIC DFE-EFF (COA) with DA. It shows that the iterations converge progressively. After the third stage, the number of zeros approaches 97%. This result explicitly indicates that if we use the conventional multistage MIC DFE-EFF (COA) detector, nearly 97% of the computations are wasted after the third stage.

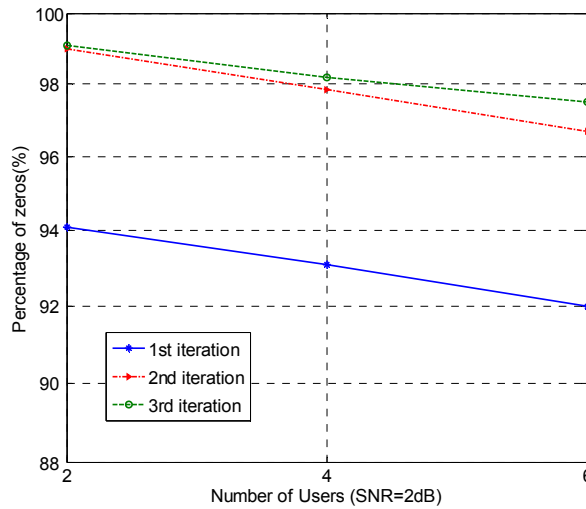


Fig. 7. Percentage of zeros in the differencing vector of the proposed MIC DFE-EFF (COA) with DA.

In other words, we can employ more stages for detection by using the DA to obtain better BER performance without adding any computational complexity. (The iteration number is determined according to the difference of percentage of zeros in two consecutive iterations. In Fig. 7, it can be seen that the difference is slight between the second and third iteration. So two iterations are enough for detection.) In Fig. 8, computational efficiency of different MUDs is compared. We use the float-points (flops) method which brings us a clear view of how many computations can be saved in a real system. The computations of all MUDs increase exponentially except MIC DFE-EFF with DA and MIC DFE-EFF (COA) with DA. After the second stage, the computations of the two MUDs with DA increase much slowly in nearly a horizontal line. Although the computational complexity of MIC DFE-EFF (COA) with DA is a little higher than MIC DFE-EFF with DA, MIC DFE-EFF (COA) with DA has much better BER performance as shown in Fig. 4 and Fig. 5.

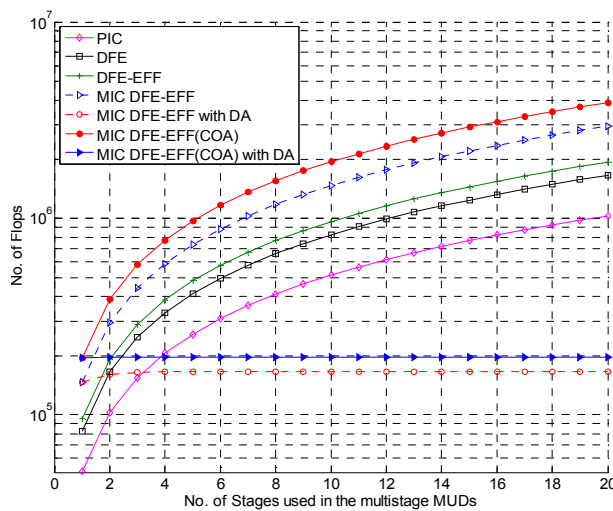


Fig. 8. Computational efficiency of different MUDs.

## 6. Conclusion

To summarize, in this paper we have dealt with two problems in MUD for multiuser MIMO DS-UWB system: improving BER performance and reducing computational complexity. To improve BER performance, we propose the COA which can improve the system performance with a little increment of computational complexity, since it exploits the known coefficients of previous matched filter. Moreover, we first introduce MIC and DFE using EFF techniques into multiuser MIMO DS-UWB system for MUD in Nakagami fading channel. To reduce computational complexity, we have presented a multistage structure MIC DFE-EFF (COA) with DA. By using DA, we can exploit more stages for detection to obtain better BER performance without adding any computational complexity. The simulation results show both of our proposed MIC DFE-EFF (COA) with and without DA can obtain much more performance gain than PIC, DFE, DFE-EFF, MIC DFE-EFF and MIC DFE-EFF with DA in both low and high system load. And both of the MIC DFE-EFF (COA) with and without DA have almost the same BER performance. Compared with PIC, DFE, DFE-EFF, MIC DFE-EFF and MIC DFE-EFF (COA), MIC DFE-EFF (COA) with DA shows much lower computational complexity. And our proposed MIC DFE-EFF (COA) with DA can be depicted as the fact that with little tradeoff BER performance, a great deal of computation cost can be saved.

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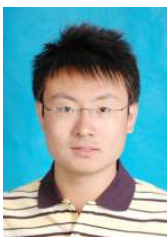
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