An Adaptive MIMO Detection Approach for LTE Advanced Uplink

Jijian Chen, Xinsheng Zhao
National Mobile Communications Research Laboratory, Southeast University
Nanjing, Jiangsu, 210096, P.R. China
Email: {ajian,xszhao}@seu.edu.cn

Abstract—OSIC (Order Successive Interference Cancellation) is a conventional MIMO detection algorithm for MIMO OFDM in LTE/LTE-A due to its significant performance gain over linear detection algorithm. But, it suffers from error diffusion, leading to performance decrease in low SNR. The tradeoff between complexity and detection performance is not optimal enough. In this paper, we proposed an adaptive MIMO detection approach for the suppression of error diffusion and reduction of complexity in LTE/LTE Advanced System (Uplink) based on the analysis of OSIC and its modification. This approach combines linear and nonlinear algorithm according to real-time SINR and the symbol error rate threshold. Simulation results show that the proposed algorithm suppresses the error diffusion and reduces the complexity successfully with the appropriate threshold.

***Index Terms*—MIMO detection, LTE-A, Uplink, SC-FDM, OFDM, OSIC**

I. INTRODUCTION

The 3rd Generation Partnership Project (3GPP) has recently specified the system requirements for the Long Term Evolution (LTE, Release8) systems and Long Term Evolution Advanced (LTE-A, Release10) systems, to meet the higher data transmission rate and higher spectral efficiency of future wireless communications. In LTE Release 8 only single transmit antenna schemes have been standardized for the uplink, while up to 4x4 Multiple-Input-Multiple-Output (MIMO) techniques are expected to be deployed in LTE-A Release 10 uplink[1], which significantly increase capacity. On the other hand, Single Carrier Frequency Division Multiplexing (SC-FDM), with low PAPR, has been selected for the uplink in LTE/LTE-A[2]. SC-FDM enables efficient frequency domain equalization (FDE) at the receiver. However, in SC-FDM, modulation symbols are transmitted serially in time domain (TD) rather than frequency domain (FD), leading to "noise enhancement" which degrades the estimation of the data symbols[5]. So, more complex MIMO detection technique is required for interference avoidance between signal layers at the LTE-A base station.

A widely used simple solution is a linear minimum mean square error (MMSE) detection, which produces an MMSE estimation for each TD symbol[6]. Due to the linearity and unitary properties of the DFT and IDFT, an equivalent equalizer is to obtain a FD MMSE estimate for each FD symbol and then take the IDFT to get the TD MMSE symbol estimates. However, in a MIMO channel, linear MMSE equalizer performance is not good enough due to the presence of spatial-multiplexing interference.

A nonlinear Order Successive Interference Cancellation (OSIC) receiver based on the MMSE criterion is a good candidate for MIMO reception[7]. The operation of OSIC is simple. The receiver first detects the signal sent by the first layer, and after successful detection it cancels the interference contributed by the detected signal before it detects the other layers. The process is repeated until all the layers are detected. With perfect layer-detecting order, OSIC can most likely cancel interference contributed by previously detected layers and improve performance. But conventional OSIC detection algorithm always leads to additional complexity than linear detection algorithm. Moreover, OSIC suffers from error diffusion and the performance decrease significantly in low SNR[7]. The tradeoff between complexity and performance is not optimal enough.

In order to suppress the error diffusion, ref [8] describes a new OSIC receiver algorithm which considers decision errors. But, it is only applicable in MIMO-OFDM systems. In this paper, we propose an efficient adaptive MIMO detection algorithm named Adaptive Hybrid Interference Cancellation (ADHIC) detection for MIMO-SC-FDM in LTE-A uplink. In order to improve the detection efficiency, ADHIC combining linear and nonlinear algorithm according to real-time SINR and the symbol error rate threshold. The advantages of proposed algorithm are that it suppresses the error diffusion and reduces the complexity significantly by introducing an appropriate threshold. Compared with the conventional OSIC algorithm, the performance gain of the proposed algorithm is observable and the complexity is noticeably reduced in moderate and low SNR.

The remainder of the paper is structured as follows. In Section II, the MIMO LTE-A system model is presented. Section III shows modified OSIC scheme for LTE-A uplink and describes the efficient ADHIC scheme and its complexity. In Section IV, simulation results are presented and discussed. Finally, summarizes the paper.

In the following, all matrices and vectors are in boldface. $(\bullet)^H$ and $(\bullet)^T$ denote Hermitian transpose and transpose, respectively. $diag(x_1,x_2,...x_L)$ is a diagonal matrix with main diagonal elements $x_1,x_2,...x_L$, $E[\bullet]$ is the expectation operator. $\mathrm{D_M}$ and $\mathrm{D_M^H}$ is the normalized M point discrete Fourier transform (DFT) and inverse discrete Fourier transform (IDFT). $\mathrm{F_M}$ and $\mathrm{F_M^H}$ denote the normalized M point Fast Fourier transform (FFT) and inverse Fast Fourier transform (IFFT).

978-1-4673-0990-5/12/$31.00 ©2012 IEEE

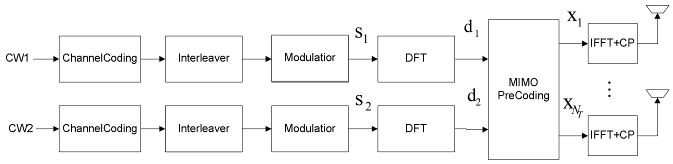

Fig.1 simplified block diagram of LTE-A uplink transmitter

II. SYSTEM MODEL

The simplified LTE-A uplink transmitter with 2 codewords (CWs) and N_T transmit antennas is presented in Fig.1. The information bits in each CW are independently encoded, interleaved, and mapped to QPSK or M-QAM modulation symbols, yielding vectors s_Z, Z=1,2. Then, a Discrete Fourier Transform (DFT) is performed in the case of SC-FDM, spreading each data symbol over all the subcarriers, obtaining the vectors d_Z. Symbols vector d_Z are then mapped over the transmit antennas by the layermaping and MIMO precoding block. Finally, an Inverse Fast Fourier Transform (IFFT) is applied and a CP is appended[1][2].

Assuming accurate synchronization at receiver, CP is long enough to cope with the delay spread of the channel, after Fast Fourier Transform (FFT) and CP removal the received signal can be presented as follows:

$$\mathrm{y}_k = \mathrm{H}_k \mathrm{x}_k + \mathrm{n}_k \ , k = 0,...,M-1 \tag{1}$$

Where $\mathrm{x}_k = [x_{k,1},...,x_{k,N_T}]^{\mathrm{T}}$ is the transmitted MIMO symbols vector form all N_T antennas at subcarrier k. it is assumed that the transmitted power is equally distributed among the antennas with $E[\mathrm{x}_k \mathrm{x}_k^{\mathrm{H}}] = I_{N_T}$. M is the number of available subcarriers. $\mathrm{n}_k = [n_{k,1},...,n_{k,N_R}]^{\mathrm{T}}$ is the additive white Gaussian noise vector with $E[\mathrm{n}_k \mathrm{n}_k^{\mathrm{H}}] = \sigma_n^2 I_{N_R}$ and σ_n^2 is the noise power.

$$\mathrm{H}_k = \begin{bmatrix} h_{k,11}, \cdots, h_{k,1N_T} \\ \vdots \quad \ddots \quad \vdots \\ h_{k,N_R 1}, \cdots, h_{k,N_R N_T} \end{bmatrix} \tag{2}$$

is the channel gain matrix at subcarrier k and the element $h_{k,ij}$ denotes the complex channel gain from the transmit antenna j to the receiver antenna i.

III. PROPOSED ADAPTIVE HYBRID INTERFERENCE CANCELLATION

OSIC detection scheme is one solution to compensate the performance decrease cause by "noise enhancement", but the drawback of conventional OSIC algorithm is that OSIC suffers from error propagation and the performance decrease significantly in low SNR. In order to suppress the error diffusion, Heunchul Lee described a new OSIC receiver algorithm (denoted as HL-OSIC) which considers decision errors, by introducing a new signal model of the layered space-time MIMO-OFDM system which includes the error propagation effect as (3). He then derived the nulling matrix based on the MMSE criterion which include the decision errors covariance matrix $\mathbf{Q_e}$ [8]. However, the estimation of $\mathbf{Q_e}$, which is the key of HL-OSIC, is not applicable in SC-FDM system, due to it's difference from OFDM. In this section, we develop the $\mathbf{Q_e}$ of MIMO-SC-FDM, extending HL-OSIC algorithm to uplink system, named Modified MMSE-based OSIC (MMOSIC). Then we analyze the post-processing SINR of each layer after interference cancellation. Finally we propose a adaptive detection approach ADHIC based on the real-time SINR and the feedback symbol error rate threshold.

The simplified LTE-A uplink receiver structure with MMOSIC/ADHIC algorithm is presented in Fig.2. In each step of detection, the detected TD symbols of pre-layer $\tilde{\mathrm{S}}_t = [\tilde{s}_{0,t},...,\tilde{s}_{M-1,t}]^{\mathrm{T}}$ are hard decided, yielding modulation symbols $\hat{\mathrm{S}}_t = [\hat{s}_{0,t},...,\hat{s}_{M-1,t}]^{\mathrm{T}}$, and then DFT is performed, transforms the modulation symbols to FD symbols $\hat{\mathrm{x}}_t = [\hat{x}_{0,t},...,\hat{x}_{M-1,t}]^{\mathrm{T}}$, which will be fed back to the MIMO equalizer with interference cancellation, allowing to progressively remove the mutual interference contribution. In MIMO equalizer we estimate the FD signal of each layer. The Log-likelihood ratio (LLR) is calculated applying max-log algorithm after all layers are estimated.

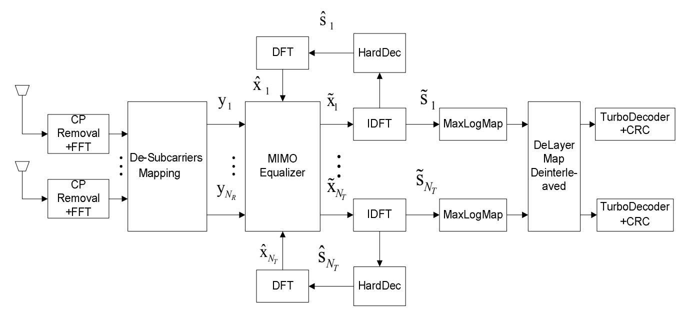

Fig.2 simplified block diagram of LTE-A uplink receiver with MMOSIC

A. *Modified signal model of the layered space-time systems*

The modified signal model of the layered space-time systems which includes the error propagation is presented as follow[8][9].

$$\begin{aligned} \mathrm{y}_{k,i} &= \mathrm{y}_k - \hat{\mathrm{H}}_{k,i-1} \hat{\mathrm{x}}_{k,i-1} \\ &= \sum_{j=i}^{\mathrm{N}} \mathrm{h}_{k,j} x_{k,j} + \sum_{j=1}^{i\text{-}1} \mathrm{h}_{k,j} (x_{k,j} - \hat{x}_{k,j}) + \mathrm{n}_k \\ &= \mathrm{H}_{k,i} \mathrm{x}_{k,i} + \hat{\mathrm{H}}_{k,i-1} \hat{\mathrm{e}}_{k,i\text{-}1} + \mathrm{n}_k \end{aligned} \tag{3}$$

We assume that the ordering of the detection have been made according to the optimal detection order. We denote $\hat{\mathrm{x}}_{k,i\text{-}1} = [\hat{x}_{k,1},...,\hat{x}_{k,i-1}]^{\mathrm{T}}$ as the detected symbols form layer 1 to i-1 at subcarrier k and $\mathrm{h}_{k,j}$ as the j th column of H_k . Define $\mathrm{x}_{k,i} = [x_{k,i},...,x_{k,N_T}]^{\mathrm{T}}$, $\mathrm{H}_{k,i} = [\mathrm{h}_{k,i}, \mathrm{h}_{k,i+1},...,\mathrm{h}_{k,N_T}]$, $\hat{\mathrm{H}}_{k,i-1} = [\mathrm{h}_{k,1}, \mathrm{h}_{k,2},...,\mathrm{h}_{k,i-1}]$ $\hat{\mathrm{e}}_{k,i-1} = [\hat{e}_{k,1},...,\hat{e}_{k,i-1}]^{\mathrm{T}}$ denotes the decision error. In V-BLAST, the pre-detected symbol vector $\hat{\mathrm{x}}_{k,i\text{-}1}$ until step i-1 is canceled out from the received vector signal at step i, resulting in the modified received vector $\mathrm{y}_{k,i}$ [7]. The nulling matrix $\mathbf{G}_k$ based on the MMSE criterion which accounts for the decision errors can be presented as follow[8].

$$\mathrm{G}_k = \mathrm{H}_{k,i}^{\mathrm{H}}(\mathrm{H}_{k,i}\mathrm{H}_{k,i}^{\mathrm{H}} + \hat{\mathrm{H}}_{k,k\text{-}1}^{\mathrm{H}}\mathrm{Q}_{\hat{\mathrm{e}}_{k,i-1}}\hat{\mathrm{H}}_{k,i-1}^{\mathrm{H}} + \sigma_{\mathrm{n}}^2\mathrm{I}_{N_R})^{-1} \tag{4}$$

where

$$\begin{aligned}\mathrm{Q}_{\hat{\mathrm{e}}_{k,i-1}} &= E[\hat{\mathrm{e}}_{k,i-1}\hat{\mathrm{e}}_{k,i-1}^{\mathrm{H}}] \\ &= \begin{pmatrix} E[|\hat{e}_{k,1}|^2|\hat{x}_{k,1}] & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & E[|\hat{e}_{k,i-1}|^2|\hat{x}_{k,i-1}] \end{pmatrix}\end{aligned} \tag{5}$$

is the covariance matrix of the decision error of $\hat{\mathrm{e}}_{k,i-1}$ until step i-1,and N_R is the number of receiver antennas. The effect of $\mathrm{Q}_{\hat{\mathrm{e}}_{k,i-1}}$ is suppressing the error propagation from step 1 to i-1.

B. *Derivation of $\boldsymbol{Q_e}$ in MIMO-SC-FDM systems*

In SC-FDM system, we estimate the TD decision errors covariance matrix at first and then calculate the FD errors covariance matrix. Without loss of generality, we assume that current detecting layer is t, the TD symbol vector $\tilde{\mathrm{S}}_t$ at layer t can be presented as follows:

$$\tilde{\mathrm{S}}_t = \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\tilde{\mathrm{x}}_t \tag{6}$$

Where $\tilde{\mathrm{x}}_t = [\tilde{x}_{0,t},...,\tilde{x}_{M-1,t}]^{\mathrm{T}}$ is the detected FD symbols of layer t, and $\hat{\mathrm{S}}_t = \Pi(\tilde{\mathrm{S}}_t)$ is the hard decision symbols vector of $\tilde{\mathrm{S}}_t = [\tilde{s}_{0,t},...,\tilde{s}_{M-1,t}]^{\mathrm{T}}$, Π denotes the hard decision function. So the TD covariance matrix of the decision error can be presented as :

$$\begin{aligned}\mathrm{Q}_{\mathrm{e}_t} &= E[\mathrm{e}_t\mathrm{e}_t^{\mathrm{H}}] \\ &= \begin{pmatrix} E[|e_{0,t}|^2|\hat{s}_{0,t}] & \cdots & E[e_{0,t}e_{M-1,t}^*|\hat{s}_{0,t},\hat{s}_{M-1,t}] \\ \vdots & \ddots & \vdots \\ E[e_{M-1,t}e_{0,t}^*|\hat{s}_{M-1,t},\hat{s}_{0,t}] & \cdots & E[|e_{M-1,t}|^2|\hat{s}_{M-1,t}] \end{pmatrix}\end{aligned} \tag{7}$$

where $\mathrm{e}_t = [e_{0,t},...,e_{M-1,t}]^{\mathrm{T}} = [(s_{0,t}-\hat{s}_{0,t}),...,(s_{M-1,t}-\hat{s}_{M-1,t})]^{\mathrm{T}}$, is the TD decision error vector of layer t. To reduce complexity, we ignore the off-diagonal elements due to the uncorrelation of decision error. Then $\mathrm{Q}_{\mathrm{e}_t}$ can be simplified to

$$\mathrm{Q}_{\mathrm{e}_t} = diag(E[|e_{0,t}|^2|\hat{s}_{0,t}],...,E[|e_{M-1,t}|^2|\hat{s}_{M-1,t}]) \tag{8}$$

We consider only the error events between two neighboring constellation points. Thus, the conditional expected values are obtained as

$$E[|e_{k,t,}|^2|\hat{s}_{k,t}] = \sum_{s\in N_{\hat{s}_t}} |s-\hat{s}_{k,t}|^2\,\mathrm{P}(s\,|\,\hat{s}_{k,t}) \tag{9}$$

where

$$\begin{aligned}&\mathrm{P}(s\,|\,\hat{s}_{k,t}) = \mathrm{P}(s\,|\,\tilde{s}_{k,t}) = \frac{p(\tilde{s}_{k,t}\,|\,s)p(s)}{\sum_{s\in S}p(\tilde{s}_{k,t}\,|\,s)p(s)} = \frac{p(\tilde{s}_{k,t}\,|\,s)}{\sum_{s\in S}p(\tilde{s}_{k,t}\,|\,s)} \\ &p(\tilde{s}_{k,t}\,|\,s) \propto \exp\left(-\frac{|\tilde{s}_{k,t}-\mu s|^2}{\sigma_{\tilde{w}}^2}\right)\end{aligned} \tag{10}$$

μ and $\sigma_{\tilde{w}}^2$ is the gain and variance of the TD estimated symbol $\tilde{s}_{k,t}$. $N_{\hat{x}_t}$ comprises the neighboring constellation points of the hard decision point. The feedback FD symbols of layer t can be abtain as

$$\hat{\mathrm{x}}_t = \mathrm{D}_{\mathrm{M}}\hat{\mathrm{S}}_t = \mathrm{D}_{\mathrm{M}}(\mathrm{S}_t + \mathrm{e}_t) = \mathrm{D}_{\mathrm{M}}\mathrm{S}_t + \mathrm{D}_{\mathrm{M}}\mathrm{e}_t = \mathrm{x}_t + \hat{\mathrm{e}}_t \tag{11}$$

where $\hat{\mathrm{e}}_t = \mathrm{D}_{\mathrm{M}}\mathrm{e}_t$, So the auto-correlation matrix of the FD error can be calculated as

$$\mathrm{Q}_{\hat{\mathrm{e}}_t} = E[\hat{\mathrm{e}}_t\hat{\mathrm{e}}_t^{\mathrm{H}}] = \mathrm{D}_{\mathrm{M}}\mathrm{Q}_{\mathrm{e}_t}\mathrm{D}_{\mathrm{M}}^{\mathrm{H}} \tag{12}$$

the diagonal elements in it can be abtain as

$$E[|\hat{e}_{k,t}|^2|\hat{x}_{k,t}] = \frac{1}{M}\sum_{k'=0}^{M-1}E[|e_{k',t}|^2|\hat{s}_{k',t}] \tag{13}$$

Substituting (13) into (5) we abtain the FD error covariance matrix of MIMO-SC-FDM systems as (14).

$$\mathrm{Q}_{\hat{\mathrm{e}}_{k,i-1}} = \begin{pmatrix} \frac{1}{M}\sum_{k'=0}^{M-1}E[|e_{k',1}|^2|\hat{s}_{k',1}] & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \frac{1}{M}\sum_{k'=0}^{M-1}E[|e_{k',i-1}|^2|\hat{s}_{k',i-1}] \end{pmatrix} \tag{14}$$

We use (4) and (14) to estimate the FD signal of each layer. We will show in Section IV that the error propagation term in (14) has a noticeable effect on performance.

C. *SINR analysis of layer signal after OSIC*

The estimated FD signal of layer t at subcarrier k after fed back interference cancellation equalization can be presented as follow[9].

$$\tilde{x}_{k,t} = \beta_{k,t}x_{k,t} + w_{k,t} \tag{15}$$

where $\beta_{k,t} = (\mathrm{G}_k\mathrm{H}_{k,i})_{t,t}$ is the t th diagonal element of $\mathrm{G}_k\mathrm{H}_{k,i}$. $w_{k,t}$ is interference plus noise Gaussian noise which is well approximated by a Gaussian distribution with variance $\sigma_{w_{k,t}}^2 = \beta_{k,t}(1-\beta_{k,t})$ [9]. All estimated symbols of layer t can be presented as follow.

$$\begin{pmatrix}\tilde{x}_{0,t}\\ \vdots \\ \tilde{x}_{M-1,t}\end{pmatrix} = \begin{pmatrix}\beta_{0,t} & \cdots & 0 \\ \vdots & \ddots & \vdots \\ 0 & \cdots & \beta_{M-1,t}\end{pmatrix}\begin{pmatrix}x_{0,t}\\ \vdots \\ x_{M-1,t}\end{pmatrix} + \begin{pmatrix}w_{0,t}\\ \vdots \\ w_{M-1,t}\end{pmatrix} \tag{16}$$

We can write (16) as (17).

$$\tilde{\mathrm{x}}_t = \Lambda_t\mathrm{x}_t + \mathrm{w}_t \tag{17}$$

Taking IDFT operation on (17) yields the vector of TD symbols which can be fed to the demodulator.

$$\begin{aligned}\tilde{\mathrm{S}}_t &= \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\tilde{\mathrm{x}}_t \\ &= \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\Lambda_t\mathrm{x}_t + \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\mathrm{w}_t \\ &= \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\Lambda_t\mathrm{D}_{\mathrm{M}}\mathrm{S}_t + \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\mathrm{w}_t \\ &= \mathrm{A}_t\mathrm{S}_t + \tilde{\mathrm{w}}_t\end{aligned} \tag{18}$$

where $\mathrm{A}_t = \mathrm{D}_{\mathrm{M}}^{\mathrm{H}}\Lambda_t\mathrm{D}_{\mathrm{M}}$ is a circulant matrix with the first column $[\upsilon_{0,t},...,\upsilon_{M-1,t}]^{\mathrm{T}}$ and

$$\upsilon_{m,t} = \frac{1}{M}\sum_{k=0}^{M-1}\beta_{k,t}\exp\left(j\frac{2\pi}{M}km\right), \quad 0 \le m \le M-1 \tag{19}$$

So the k th estimated TD symbol of layer t can be presented as follow.

$$\tilde{s}_{k,t} = \sum_{i=0}^{M-1} \upsilon_{(k-i)_M,t} s_{i,t} + \tilde{w}_{k,t}$$
$$= \upsilon_{0,t} s_{k,t} + \sum_{\substack{i=0 \\ i\neq k}}^{M-1} \upsilon_{(k-i)_M,t} s_{i,t} + \tilde{w}_{k,t} \tag{20}$$

From (20) we know that estimated TD symbol include not only a noise term but the interference from other subcarriers in the same SC-FDM symbol. auto-correlation matrix of noise can be

$$\begin{aligned} \mathrm{R}_{\tilde{\mathrm{w}}_t} &= E[\tilde{\mathrm{w}}_t \tilde{\mathrm{w}}_t^{\mathrm{H}}] \\ &= E[\mathrm{D}_{\mathrm{M}}^{\mathrm{H}} \mathrm{w}_t \mathrm{w}_t^{\mathrm{H}} \mathrm{D}_{\mathrm{M}}] \\ &= \mathrm{D}_{\mathrm{M}}^{\mathrm{H}} E[\mathrm{w}_t \mathrm{w}_t^{\mathrm{H}}] \mathrm{D}_{\mathrm{M}} \\ &= \mathrm{D}_{\mathrm{M}}^{\mathrm{H}} diag(\beta_{0,t}(1-\beta_{0,t}), \beta_{1,t}(1-\beta_{1,t}), ..., \beta_{M-1,t}(1-\beta_{M-1,t})) \mathrm{D}_{\mathrm{M}} \end{aligned} \tag{21}$$

From(21) we know that $\mathrm{R}_{\tilde{\mathrm{w}}_t}$ is a circulant matrix with diagonal element $\frac{1}{M}\sum_{k=0}^{M-1}\beta_{k,t}(1-\beta_{k,t})$, so we obtain the power of $\tilde{\mathrm{w}}_t$:

$$\sigma_{\tilde{\mathrm{w}}_t}^2 = \frac{1}{M}\sum_{k=0}^{M-1}\beta_{k,t}(1-\beta_{k,t}) \tag{22}$$

The SINR of layer t can be obtained as fallow.

$$\begin{aligned} SINR_t &= \frac{|\upsilon_{0,t}|^2}{\sum_{k=1}^{M-1}|\upsilon_{k,t}|^2 + \frac{1}{M}\sum_{k=0}^{M-1}\beta_{k,t}(1-\beta_{k,t})} \\ &= \frac{|\upsilon_{0,t}|^2}{\sum_{k=0}^{M-1}|\upsilon_{k,t}|^2 + \frac{1}{M}\sum_{k=0}^{M-1}\beta_{k,t}(1-\beta_{k,t}) - |\upsilon_{0,t}|^2} \end{aligned} \tag{23}$$

Substituting (19) into (23) we obtain SINR of of layer t.

$$SINR_t = \frac{\upsilon_{0,t}}{1-\upsilon_{0,t}} \tag{24}$$

The mobile communication receivers experience fluctuating channel conditions, where the channel responses and post-processing SINR in (24) vary all the time. In good channel conditions, the SINR of each layer after interference cancellation is high and the performance gain of MMOSIC is significant. In bad channel conditions, the SINR is low, leading to high hard decision symbol error rate (SER), the performance gain is negligible. In this context, feedback and interference cancellation in low SINR conditions seems not only unnecessary but also wasteful. It is natural to employ only low-complexity MIMO detection solutions, such as MMSE detection, under "bad" channel scenarios. To fully exploit channel variations that vacillate between "good" conditions and "poor" conditions, we propose Adaptive Hybrid Interference Cancellation (ADHIC) approach combining linear and nonlinear algorithm based on real-time SINR and the SER threshold.

D. Proposed ADHIC detection algorithm

MMOSIC algorithm can suppress the error propagation significantly but complexity is high. Most of complexity increase comes from the computation of $\mathbf{G}_k$ in (4) and the DFT operation in each step of detection. To fully exploit channel and reduce complexity, we change the feedback mechanism. In the MIMO equalizer, we estimate the SER based on the SINR of current layer, a linear MMSE detection method is adopted instead of OSIC to detecte the rest of layers if the estimated SER of current layer upon the threshold.

To estimate the SER of current layer, we model interference plus noise in (20) as additive white gaussian noise, then we obtained the maping function between the post-processing SINR and hard decision M_c-QAM SER[10].

$$P_{M_c} \le 4Q(\sqrt{\frac{3E_{av}}{(M_c-1)N_0}}) = 4Q(\sqrt{\frac{3SINR}{(M_c-1)}}) \tag{25}$$

where $Q(x) = \int_x^\infty \frac{1}{\sqrt{2\pi}}\exp(-\frac{u^2}{2})du$

For example, We take $SER = 2Q(\sqrt{SINR})$ for QPSK and $SER = 3Q(\sqrt{SINR})$ for 16QAM. From simulation, we found that 16QAM SER curve of each layer after OSIC fit the Theoretical curve perfectly, which proves our analysis. Table I shows the maping between 16QAM SINR thresholds and SER thresholds.

TABLE I:SER SINR THRESHOLDS MAPING TABLE

SER thresholds	0.01	0.1	0.2	0.3	0.5
QPSK SINRthresholds(dB)	8.21	4.33	2.16	0.32	-3.41
16QAM SINRthresholds(dB)	15.6	12.3	10.5	9.26	6.71

To summarize, the proposed ADHIC algorithm can be described as follows:

Recursion loop starts with i=1and SER Thresholds TH_{SER}=[TH_{SER1}, …, TH_{SERN}].

1) Calculate $\mathbf{Q_e}$ of SC-FDM using (14),when i=1, $\mathbf{Q_e=0}$.
2) Calculate $\mathbf{G}_k$ using (4) for k=0,2,…,M-1.
3) Calculate *SINR* of all undetected layers using (24) and find the layer which has the largest *SINR*, denoting its position as t,with $SINR_t$.
4) Apply Table I and TH_{SERt} to obtain the SINR thresholds TH_{SINRt} of layer t.
5) If $SINR_t < TH_{SINRt}$, the linear MMSE detection method is adopted to estimate the rest of undetected layers, else SIC detection will be performed. The estimated TD symbols in layer t will be fed back to the equalizer after DFT and then return to 1) until all layers are detected.

IV. PERFORMANCE SIMULATION

The performance of the proposed algorithm is evaluated by Monte Carlo simulations. The main simulation parameters are gathered in Table II.

TABLE II: SIMULATION PARAMETERS

Carrier Frequency	2GHz
Sampling Frequency	15.36MHz[2]
Subcarrier Spacing	15KHz[2]
FFT Size	1048[2]
System Bandwidth	10MHz
Used Subcarriers	600
CP Type	Normal[2]
Slot Duration	0.5ms[2]
Symbols Per Slot	7[2]
MIMO Scheme	(2x2/4x4)SM[4]
MCS	QPSK 1/3 1/2 16QAM 1/2 2/3 64QAM 2/3
Channel Coding	3GPP R10 Turbo code 1/3 basic rate[3]
Turbo Decoder Iterations	8
User Speed	20/km/h

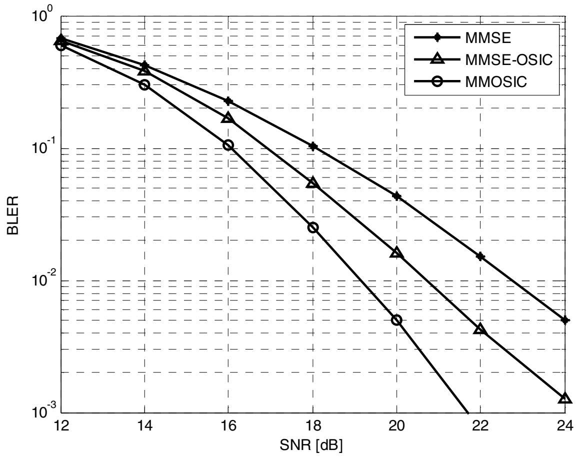

Fig.3 BLER performance of MMOSIC with 4x4 antennas 16QAM1/2

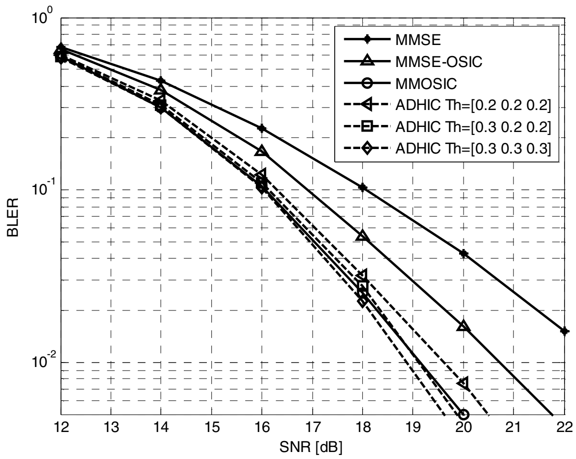

Fig.4 BLER performance of ADHIC with 4x4 antennas 16QAM1/2

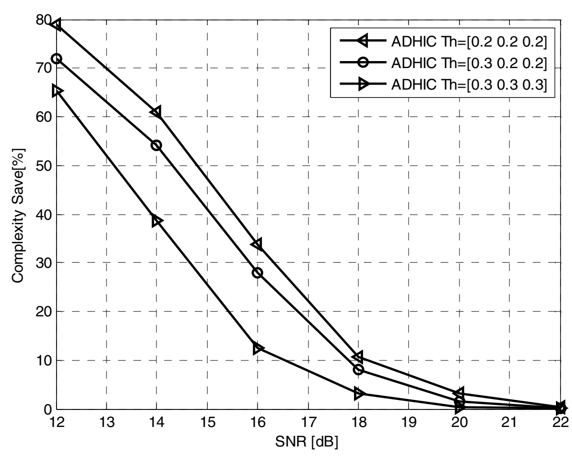

Fig.5 Complexity saving scale of ADHIC with 4x4 antennas 16QAM 1/2

SCM channel model [4] is used in the simulations, and perfect channel knowledge is assumed. Fig.3 shows the performance of MMOSIC in LTE-A uplink with 4x4 antennas 16QAM 1/2. Linear MMSE and conventional MMSE-based OSIC (MMSE-OSIC) detection performance is also included. Proposed MMOSIC obtain up to 4dB gain over linear detection while MMSE-OSIC obtain only 2dB gain over linear detection at 10^{-2}BLER. It can be seen that proposed MMOSIC algorithm can suppress the error propagation significantly. Fig.4 and Fig.5 show the performance gain and the complexity saving scale of ADHIC over MMSE-OSIC in LTE-A uplink with 4x4 antennas 16QAM 1/2. "Th" denotes the SER thresholds in each layer. The complexity is measured by floating-point operations (flops) of proposed algorithm. N point FFT requires $N\log_2(N)$ complex additions and $(N/2)\log_2(N)$ complex multiplications. M point DFT requires $M(M-1)$ complex additions and M^2 complex multiplications. We count a complex addition or subtraction as 2 flops and a complex multiplication as 6 flops. It can be seen that the ADHIC obtain 1.2dB~2dB gain over conventional MMSE-OSIC at 10^{-2}BLER but the complexity is saved 12%-35% at 16dB. The lower of the thresholds, the more complexity is saved, but the performance decrease slightly when SER shreshold<0.2. The complexity reduction is noticeable as long as SNR is not very high, say SNR<18dB.

V. Conclusion

Conventional MMSE-based OSIC can be applicable in LTE-A uplink MIMO detection, but it suffers from error propagation. We proposed an Adaptive Hybrid Interference Cancellation (ADHIC) detection algorithm to suppress the error propagation and reduce the complexity. Simulation show that the ADHIC obtain the 4dB performance gain over conventional MMSE and obtain 1.2-2dB performance gain over conventional MMSE-OSIC detection but the complexity reduction is noticeable when SNR<18dB. The core idea of this paper is to improve performance by suppressing the error propagation and reduce the complexity by reducing the useless feedback operation based on real-time SINR. The proposed algorithm can be extended to multiuser MIMO OFDM system.

VI. Acknowledgment

This work is supported by the National Natural Science Foundation of China (NSFC) No.61071111 and Chinese Important National Science & Technology Projects under Grant 2012ZX03001009-002.

References

[1] Physical Channels and Modulation(Release10). 3rd Generation Partnership Project, Tech. Rep. TS 36211, V10.1.0
[2] LTE Physical Layer - General Description (Release 10). 3rd Generation Partnership Project, Tech. Rep. TS 36201, V10.1.0
[3] Multiplexing and channel coding(Release10). 3rd Generation Partnership Project, Tech. Rep. TS 36212, V10.1.0
[4] Multiple Input Mutiple Output (MIMO) simulations (Release9). 3rd Generation Partnership Project, Tech. Rep. TS 25996, V9.0.0
[5] G. Berardinelli et al.Improving SC-FDMA Performance by Turbo Equalization in UTRA LTE Uplink.Proc. IEEE VTC, Singapore, May 2008, pp. 2557–61.
[6] Park.C.S,Wang,Y.-P.E,Jongren.G,Hammarwall.D.Evolution of Uplink MIMO for LTE-Advanced. Communications Magazine,IEEE,2011
[7] P. W. Wolniansky, G. J. Foschini, G. D. Golden, and R. A. Valenzuela. V-BLAST: an architecture for realizing very high data rates over the rich-scattering wireless channel. in URSI International Symposium on Signals, Systems and Electronics, September 1998, pp. 295–300,
[8] Heunchul Lee, Inkyu Lee. New Approach for Coded Layered Space-Time OFDM Systems.IEEE International Conference on Communications, 2005. ICC 2005. PP: 608 - 612 Vol. 1
[9] Heunchul Lee, Inkyu Lee.New Approach for Error Compensation in Coded V-BLAST OFDM Systems. IEEE Transactions on Communications,2007,PP: 345-355 Vol.55
[10] John G.Proakis. Digitial Communications (Fourth Edition).Publishing house of Electronics industry,2003