# ADAPTIVE ITERATIVE LAYERED SPACE-FREQUENCY EQUALIZATION FOR SINGLE-CARREIR MIMO SYSTEMS

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**ABSTRACT**

**We propose an adaptive iterative layered space-frequency equalization (AILSFE) structure for multiple-input multiple-output (MIMO) time-varying frequency selective channels, which incorporates iterative frequency-domain equalization (FDE) with hybrid interference cancellation. It is shown that AILSFE significantly outperforms the previously proposed LSFE with a modest increase of complexity, and can provide better performance than Iterative-block decision feedback equalizer (IB-DFE) at comparable complexity, with a relatively small number of iterations.**

## 1. INTRODUCTION

Frequency-domain equalization (FDE) [1-3] was shown to be an effective solution for frequency selective channels in a single-carrier (SC) system, which has much better performance than time-domain equalization with similar complexity, and lower peak-to-average ratio (PAR) than orthogonal frequency division multiplexing (OFDM).

In [4] FDE was employed in an SC multiple-input multiple-output (MIMO) system, where all the signals are detected simultaneously. In [5], a layered space-frequency equalization (LSFE) structure was developed to provide enhanced performance over the single-stage MIMO FDE by combining FDE and successive interference cancellation. To further improve the performance, an iterative-block decision feedback equalizer (IB-DFE) was proposed in [6], which incorporates iterative interference cancellation with FDE. However, the computational complexity of IB-DFE is high due to the number of iterations needed to achieve the required performance. Also, [4-6] only assumed quasi-static channels.

Adaptive FDE was investigated in [7-8] for time-varying channels, where the equalizer coefficients are calculated based on the least-mean-square (LMS) or recursive-least-square (RLS) criterion, without explicit channel estimation required. Another type of adaptive FDE structures are based on adaptive channel estimation [9-10] where the equalizer coefficients are computed based on the channel estimates. However, the work in [9-10] only assumed single-input single-output (SISO) and single-input multiple-output (SIMO) systems.

In this paper, we propose an adaptive iterative layered space-frequency equalization (AILSFE) structure for an SC MIMO system. Our work is different in that we introduce iterative hybrid interference cancellation by employing both hard DFE and soft DFE in frequency domain. We also consider time-varying channels and exploit adaptive channel estimation to track channel variations, achieving performance close to the case with perfect channel state information (CSI). It is shown that the proposed AILSFE significantly outperforms LSFE [5], with only a modest increase of complexity. It can also provide better performance than its IB-DFE counterpart [6] at comparable complexity, especially when the number of iterations is relatively small. When AILSFE and IB-DFE achieve similar performance, the former requires less number of iterations and therefore lower complexity.

Section 2 presents the system model. The proposed AILSFE structure is described in Section 3. The computational complexity is analyzed in Section 4. Section 5 shows the simulation results and the conclusion is drawn in Section 6.

## 2. SYSTEM MODEL

We investigate an uncoded MIMO system with $K$ transmit antennas and $L$ receive antennas. Let $d_{p,i,k}$ denote the $i$th $(i=0,...M-1)$ data symbol in the $p$th block of $M$ symbols transmitted by the $k$th $(k=1,...K)$ antenna, with unit average symbol energy. The overall channel memory is assumed to be $N$, lumping the effects of transmit filter, receive filter and physical channel. Each data block is pre-pended with a cyclic prefix (CP), which is the replica of the last $N$ symbols of the block. At the receiver, the CP is discarded to eliminate the inter-block interference (IBI) and to make the channel appear to be periodic with period $M$.

The $m$th $(m=0,...M-1)$ sampled signal of the $p$th received block at the $l$th $(l=1,...L)$ receive antenna is expressed as

$$x_{p,m,l}=\sum_{k=1}^{K}\sum_{i=0}^{N}h_{p,m-i,lk}d_{p,i,k}+n_{p,m,l} \quad (1)$$

where $h_{p,i,lk}$ $(i=0,...,N)$ denotes the channel impulse response (CIR) between the $k$th transmit antenna and the $l$th receive antenna over the $p$th block, and $n_{p,m,l}$ is additive white Gaussian noise (AWGN) with zero mean and single-sided power spectral density $N_0$.

The received signals are transferred into the frequency domain by the FFT operation. The FFT of $x_{p,m,l}$ is given by

$$X_{p,m,l}=\sum_{k=1}^{K}H_{p,m,lk}D_{p,m,k}+N_{p,m,l} \quad (2)$$

where $X_{p,m,l}=\sum_{i=0}^{M-1}x_{p,i,l}e^{-j2\pi im/M}$, $H_{p,m,lk}=\sum_{i=0}^{N}h_{p,i,lk}e^{-j2\pi im/M}$, $D_{p,m,k}=\sum_{i=0}^{M-1}D_{p,i,k}e^{-j2\pi im/M}$, $N_{p,m,l}=\sum_{i=0}^{M-1}n_{p,i,l}e^{-j2\pi im/M}$. Defining $\boldsymbol{X}_{p,m,l}=\left[X_{p,m,1},...,X_{p,m,L}\right]$ as the received signal vector on the $m$th frequency tone, which is expressed as

$$\boldsymbol{X}_{p,m}=\sum_{k=1}^{K}\boldsymbol{H}_{p,m,k}D_{p,m,k}+\boldsymbol{N}_{p,m} \quad (3)$$

where $\boldsymbol{H}_{p,m,k}=\left[H_{p,m,1k},...,H_{p,m,Lk}\right]$, $\boldsymbol{N}_{p,m,l}=\left[N_{p,m,1},...,N_{p,m,L}\right]$.

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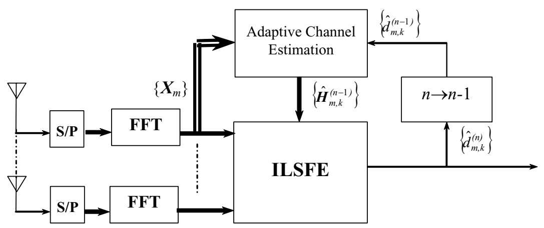


Fig.1. Block diagram of the AILSFE ($m$=0,…,$M$-1; $k$=1,…,$K$; $n$ denotes the iteration number)

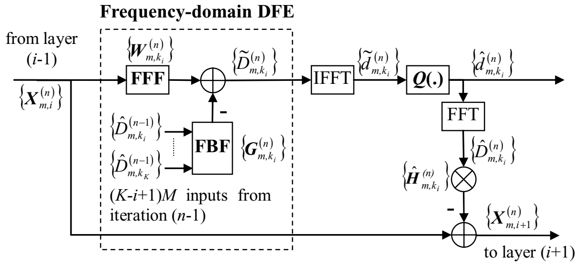


Fig.2. Block diagram for the $i$th detection layer in the $n$th iteration of ILSFE

## 3. ADAPTIVE ITERATIVE LAYERED SPACE-FREQUENCY EQUALIZATION

### *3.1 Algorithm Description*

The proposed AILSFE structure is depicted in Fig.1, where adaptive channel estimation is employed to calculate the equalizer coefficients. For the simplicity of presentation, we drop the block index $p$ in all the notations of Sections 3.1 and 3.2. The adaptive channel estimation first operates in the training mode with the aid of training blocks and then in the decision-directed mode with the aid of estimated signals. In the decision-directed mode, the channel estimates are updated iteration by iteration instead of block by block (detection of each blocks consist of multiple iterations), to increase the convergence speed. The details of adaptive channel estimation are discussed in Section 3.3.

The ILSFE block in Fig.1 consists of $K$ detection layers in each iteration. The first iteration performs LSFE [5], where at each layer the best data stream in the minimum mean square error (MMSE) sense is detected by FDE, and then cancelled from the received signals. Fig.2 illustrates the $i$th ($i$=1,…,$K$) layer in the $n$th iteration, which is the same as the first iteration except that the FDE is replaced by the frequency-domain DFE. With interference of the first ($i$-1) detected layers of iteration $n$ cancelled, the received signals are input into a feedforward filter (FFF). To remove the residual inter-layer interference and inter-symbol interference, the tentative estimates from layers $i$ to $K$ of iteration ($n$-1) are passed through a feedback filter (FBF). The block-wise DFE output signals are transferred back into time domain by IFFT, and input into a decision device denoted by $Q$(.). Finally, the interference of the currently detected data stream is cancelled from the received signals in frequency domain to obtain the updated input signals for the next layer. Note that for the first iteration ($n$=1), the FBF is removed and the frequency-domain DFE reduces to FDE.

In the parallel interference cancellation (PIC) based IB-DFE [6], the first iteration performs linear processing. Each of the remaining iterations is the same as depicted in Fig.2, except that the received signals are not updated at each layer, and the FBF has $KM$ inputs which are the tentative estimates of all the signals from iteration ($n$-1).

### *3.2 Equalizer Coefficients*

Our equalizer design and data selection are based on the MMSE criterion. As shown in Fig.2, we assume that $\{d_{m,k_i}\}$ is the data stream detected at layer $i$ of iteration $n$, which is transmitted from the $k_i$th ($k_i \in \{1,\ldots,K\}$) antenna. Let $\tilde{d}^{(n)}_{m,k_i}$ and $\hat{d}^{(n)}_{m,k_i}$ respectively denote the soft and hard estimates of $d_{m,k_i}$ in iteration $n$, and $\tilde{D}^{(n)}_{m,k_i}$ and $\hat{D}^{(n)}_{m,k_i}$ be their frequency-domain counterparts. Also define $\hat{\boldsymbol{H}}^{(n)}_{m,k}$ as the estimate of $\boldsymbol{H}^{(n)}_{m,k}$ for the $n$th iteration. Thus, the input signal to the $i$th layer of the $n$th iteration is given by

$$\boldsymbol{X}^{(n)}_{m,i} = \boldsymbol{X}_m - \sum_{q=1}^{i-1} \hat{\boldsymbol{H}}^{(n)}_{m,k_q} \hat{D}^{(n)}_{m,k_q} \tag{4}$$

with the previously detected ($i$-1) layers cancelled.

Let $\boldsymbol{W}^{(n)}_{m,k_i}$ and $\boldsymbol{G}^{(n)}_{m,k_i}$ denote the FFF and FBF weight vectors with respect to the $m$th frequency tone for layer $i$ in iteration $n$, which are of size $L \times 1$ and $(K-i+1)\times 1$, respectively. Thus, the frequency-domain DFE output $\tilde{D}^{(n)}_{m,k_i}$ is expressed as

$$\tilde{D}^{(n)}_{m,k_i} = \boldsymbol{W}^{(n)H}_{m,k_i} \boldsymbol{X}^{(n)}_{m,i} - \boldsymbol{G}^{(n)H}_{m,k_i} \hat{\boldsymbol{D}}^{(n-1)}_{m,k_i} \tag{5}$$

where $(.)^H$ denotes the complex-conjugate transpose, and vector $\hat{\boldsymbol{D}}^{(n-1)}_{m,k_i} = \left[\hat{D}^{(n-1)}_{m,k_i} \ \Lambda \ \hat{D}^{(n-1)}_{m,k_K}\right]^T$, which is transferred from the time domain hard estimates $\hat{\boldsymbol{d}}^{(n-1)}_{m,k_i} = \left[\hat{d}^{(n-1)}_{m,k_i} \ \Lambda \ \hat{d}^{(n-1)}_{m,k_K}\right]^T$ by FFT, indicates the ($K$-$i$+1) frequency-domain interferers with respect to iteration ($n$-1) (for the first iteration, $\hat{d}^{(0)}_{m,k} = 0$ for any $k$ and $m$).

At each layer of ILSFE, the data stream with the smallest MSE is detected, and cancelled from the received signals. The MSE between $\tilde{d}^{(n)}_{m,k_i}$ and $d_{m,k_i}$ is given by

$$MSE^{(n)}_{k_i} = \frac{1}{M}\sum_{m=0}^{M-1} E\left|\tilde{d}^{(n)}_{m,k_i} - d_{m,k_i}\right|^2 \tag{6}$$

Assuming a linear estimator between the frequency domain estimate $\hat{D}^{(n)}_{m,k}$ and the transmitted symbol $D_{m,k}$, $\hat{D}^{(n)}_{m,k}$ can be expressed as

$$\hat{D}^{(n)}_{m,k} = \rho^{(n)}_{m,k} D_{m,k} + \Delta^{(n)}_{m,k} \tag{7}$$

where $\Delta^{(n)}_{m,k}$ denotes a zero-mean error term and $\rho^{(n)}_{m,k}$ is the correlation coefficient between $D_{m,k}$ and $\hat{D}^{(n)}_{m,k}$:

$$\rho^{(n)}_{m,k} = \frac{1}{M} E\left[\hat{D}^{(n)}_{m,k} D^*_{m,k}\right] = E\left[\hat{d}^{(n)}_{m,k} d^*_{m,k}\right] \tag{8}$$

The value of $\rho^{(n)}_{m,k}$, conditioned to a given set of time-domain equalizer outputs $\left\{\tilde{d}^{(1)}_{m,k}, \Lambda, \tilde{d}^{(n-1)}_{m,k}\right\}$, can be computed as described in [11]. With the assumption of $E\left[\Delta^{(n)}_{m,k} D^*_{m,k}\right] \approx 0$, we have

$$E\left[\left|\Delta^{(n)}_{m,k}\right|^2\right] = M\left(1 - {\rho^{(n)}_{m,k}}^2\right) \tag{9}$$

It can be derived that the optimum FFF weight vector $\boldsymbol{W}^{(n)}_{m,k_i}$ is given by

$$\boldsymbol{W}^{(n)}_{m,k_i} = {\boldsymbol{R}^{(n)}_{m,k_i}}^{-1} \hat{\boldsymbol{H}}^{(n)}_{m,k_i} \tag{10}$$

where

$$\boldsymbol{R}^{(n)}_{m,k_i} = \sum_{q=i}^{K}\left(1-\rho^{(n)^2}_{m,k_q}\right)\hat{\boldsymbol{H}}^{(n)}_{m,k_q}\hat{\boldsymbol{H}}^{(n)H}_{m,k_q} + N_0\boldsymbol{I} \tag{11}$$

The corresponding FBF weight vector $\boldsymbol{G}^{(n)}_{m,k_i}$ is expressed as:

$$\boldsymbol{G}^{(n)}_{m,k_i} = \begin{cases} \boldsymbol{B}^{(n)H}_{m,k_i}\boldsymbol{W}^{(n)}_{m,k_i} & (n=1) \\ \boldsymbol{B}^{(n)H}_{m,k_i}\boldsymbol{W}^{(n)}_{m,k_i} - \boldsymbol{\delta} & (n>1) \end{cases} \tag{12}$$

where

$$\boldsymbol{B}^{(n)}_{m,k_i} = \left[\hat{\boldsymbol{H}}^{(n)}_{m,k_i} \ldots \hat{\boldsymbol{H}}^{(n)}_{m,k_K}\right] \tag{13}$$

$$\boldsymbol{\delta} = [1,0,\ldots,0]^T \tag{14}$$

For the first iteration ($n$=1), $\rho^{(n)}_{m,k_i} = 0$ and the frequency-domain DFE reduces to the traditional FDE.

### *3.3 Adaptive Channel Estimation*

The adaptive channel estimation scheme that we use is referred to as least-mean-square structured channel estimation (LMS-SCE), which utilizes the correlation between adjacent frequency bins. It was shown in [9-10] that LMS-SCE is a very effective method which provides close performance to the case with perfect CSI and has a low computational complexity.

In the training mode, Define $c_{p,m,k}$ as the $m$th frequency tone of the $p$th training block from transmit antenna $k$. Letting $\boldsymbol{c}_{p,m} = [c_{p,m,1} \;\cdots\; c_{p,m,K}]$ and $\boldsymbol{\Gamma}_{p,m,l} = [H_{p,m,l1} \;\cdots\; H_{p,m,lK}]^T$, the received signal vector at the $l$th receive antenna over all the frequency tones is given by

$$\boldsymbol{Y}_{p,l} = \tilde{\boldsymbol{C}}_p\boldsymbol{\Gamma}_{p,l} + \boldsymbol{\eta}_{p,l} \tag{15}$$

where $\tilde{\boldsymbol{C}}_p = \begin{bmatrix} \boldsymbol{C}_{p,0} & & \\ & \ddots & \\ & & \boldsymbol{C}_{p,M-1} \end{bmatrix}$ is of size $M \times KM$, $\boldsymbol{\eta}_{p,l}$ is the noise vector. $\boldsymbol{\Gamma}_{p,l} = \left[\boldsymbol{\Gamma}^T_{p,0,l} \;\cdots\; \boldsymbol{\Gamma}^T_{p,M-1,l}\right]^T$, which can also be expressed as

$$\boldsymbol{\Gamma}_{p,l} = \tilde{\boldsymbol{F}}\boldsymbol{\gamma}_{p,l} \tag{16}$$

where $\boldsymbol{\gamma}_{p,l} = \left[h_{p,0,l1} \ldots h_{p,N,l1} \ldots h_{p,0,lK} \ldots h_{p,N,lK}\right]^T$ is the CIR vector of length $K(N+1)$ at the $l$th receive antenna. $\tilde{\boldsymbol{F}} = \left(\boldsymbol{F}^T_0 \ldots \boldsymbol{F}^T_{M-1}\right)^T$ where $\boldsymbol{F}_m$ $(0 \le m \le M-1)$ is a $K \times K(N+1)$ block Toeplitz matrix defined as

$$\boldsymbol{F}_m = \begin{bmatrix} \boldsymbol{f}_m & \boldsymbol{0} & \cdots & \boldsymbol{0} \\ \boldsymbol{0} & \boldsymbol{f}_m & \ddots & \vdots \\ \vdots & \ddots & \ddots & \boldsymbol{0} \\ \boldsymbol{0} & \cdots & \boldsymbol{0} & \boldsymbol{f}_m \end{bmatrix} \tag{17}$$

with $\boldsymbol{f}_m = \left[e^{-j2\pi 0m/M} \;\cdots\; e^{-j2\pi Nm/M}\right]$.

The LMS-SCE minimizes the cost function

$$J_{LMS-SCE}\left(\hat{\gamma}_{p,l}\right) = E\left\{\left\|\boldsymbol{Y}_{p,l} - \tilde{\boldsymbol{C}}_p\tilde{\boldsymbol{F}}\hat{\gamma}_{p,l}\right\|^2\right\} \quad (l=1,\ldots,L) \tag{18}$$

with respect to $\hat{\gamma}_{p,l}$ which is the estimate of $\gamma_{p,l}$. This yields

$$\hat{\boldsymbol{\gamma}}_{p+1,l} = \hat{\boldsymbol{\gamma}}_{p,l} + \mu\boldsymbol{E}_{p,l} \tag{19}$$

where $\mu$ is the step size and $\boldsymbol{E}_{p,l}$ is given by

$$\boldsymbol{E}_{p,l} = \tilde{\boldsymbol{F}}^H\tilde{\boldsymbol{C}}^H_p\left[\boldsymbol{Y}_{p,l} - \tilde{\boldsymbol{C}}_p\tilde{\boldsymbol{F}}\hat{\boldsymbol{\gamma}}_{p,l}\right] \tag{20}$$

Using (16), the estimate of the channel frequency response can be updated as

$$\hat{\boldsymbol{\Gamma}}_{p+1,l} = \hat{\boldsymbol{\Gamma}}_{p,l} + \mu\tilde{\boldsymbol{F}}\boldsymbol{E}_{p,l} \tag{21}$$

In the decision-directed mode, the channel estimates are updated in each iteration of ILSFE where the transmitted symbols are replaced by the (tentative) estimates of the data symbols.

## 4. COMPLEXITY ANALYSIS

The signal processing complexity of AILSFE in terms of the number of complex multiplications, is shown in Table I for the first iteration and Table II for each of the remaining iterations, compared to that of the previously proposed IB-DFE discussed in Section 3.1 with adaptive channel estimation. The computational complexity for each iteration primarily comes from four parts: a) Adaptive channel estimation according to Section 3.3; b) FFT and IFFT operations; c) Solving the FFF and FBF coefficients according to (10) and (12); d) Operation for equalization and interference cancellation according to (4) and (5). It can be shown that the computation of $\rho^{(n)}_{m,k}$ in (8) is not comparable to other computations and is therefore ignored.

Detection ordering for AILSFE requires the inverse of $\boldsymbol{R}_{m,k_i}$ in (11) for $K(K+1)M/2$ times with an order of $L^3/3$ multiplications each, which costs much complexity. It can be shown by simulation that the detection ordering for the first iteration of AILSFE can be applied to the remaining iterations with significant complexity reduction and little performance loss. We therefore assume this simplification in Sections 4 and 5.

With $K$=4 transmit antennas, $L$=4 receive antennas, and a data block size of $M$=64, the overall normalized complexity per block with 3 iterations is shown in Table III (for the first iteration AILSFE reduces to adaptive LSFE). Obviously, AILSFE with 3 iterations has a modest increase of complexity compared to LSFE, and a similar complexity to its IB-DFE counterpart.

TABLE I: Computational complexity for the first iteration

| Receiver | AILSFE | Adaptive IB-DFE |
|---|---|---|
| Channel Estimation | $2KLM + K(2L+1)M(\log_2 M)/2$ | |
| FFT/ IFFT | $(2K+L)M(\log_2 M)/2$ | |
| FFF Coefficients | $K(K+1)L^3M/6 + KL^2M$ | $KL^3M/3 + KL^2M$ |
| FBF Coefficients | 0 | |
| Equalization | $2KLM$ | $KLM$ |

TABLE II: Computational complexity for each of the remaining iterations

| Receiver | AILSFE | Adaptive IB-DFE |
|---|---|---|
| Channel Estimation | $2KLM + K(2L+1)M(\log_2 M)/2$ | |
| FFT/ IFFT | $KM(\log_2 M)$ | |
| FFF Coefficients | $KL^3M/3 + KL^2M$ | |
| FBF Coefficients | $K(K+1)LM/2$ | $K^2LM$ |
| Equalization | $2KLM + K(K+1)M/2$ | $KLM + K(K+1)M/2$ |

TABLE III: Normalized computational complexity per block with $K$=4, $L$=4 and $M$=64 with 3 iterations (LSFE is not iterative)

| Adaptive LSFE | AILSFE | Adaptive IB-DFE |
|---|---|---|
| 100% | 276% | 242% |

## 5. SIMULATION RESULTS

In the simulations, we use $K$=4 transmit antennas and $L$=4 receive antennas. Each data block consists of $M$=64 QPSK symbols, with a symbol rate of 2 Mbaud. Both the transmit and receive filters use a raised-cosine pulse with a roll-off factor of 0.35. The channel is modeled by following the exponential power delay profile with an RMS delay spread of 0.625$\mu$s and a Doppler frequency of $f_d$=50Hz. The overall channel memory is $N$=6, lumping the effects of transmit filter, receive filter and real channel. The adaptive channel estimation scheme discussed in Section 3.3 is employed with 10 training blocks. The SNR is defined as the spatial average ratio of the received signal power to noise power.

Fig.3 illustrates the average BER performance of AILSFE with 1 to 3 iterations, compared to adaptive LSFE and IB-DFE discussed in Sections 3.1 and 4. With one iteration, AILSFE reduces to adaptive LSFE. It is obvious that AILSFE significantly outperforms LSFE at only a modest cost of complexity according to Section 4. For instance, At BER=$10^{-3}$, ILSFE with 3 iterations provides an SNR gain of around 7dB over LSFE, with a complexity of 276% compared to the latter. AILSFE also has better performance than its IB-DFE counterpart with the same number of iterations (comparable complexity), especially when the number of iterations is relatively small. At BER=$10^{-3}$, AILSFE with both 2 iterations and 3 iterations provides a performance gain of about 1.5dB over its IB-DFE counterpart. It can also be shown that AILSFE with 3 iterations and adaptive IB-DFE with 4 iterations achieve similar performance, while the former structure has a lower complexity. This can be explained by noting that we employ hybrid interference cancellation in AILSFE with detection ordering. Besides, it is shown that AILSFE provides close performance to ILSFE with perfect CSI, especially with a relatively large number of iterations. This implies the efficiency of LMS-SCE.

Fig.4 shows the performance of different layers in AILSFE. At high SNR, with 1 iteration the last layer has the best performance and the first layer has the worse performance. With 2 iterations, however, it is the opposite. This implies the sensitivity of performance of AISLFE to error propagation. With the increase of number of iterations, the performance gaps between different layers of AISLFE are mitigated gradually. After 3 iterations, all the layers provide similar performance since the most significant interlayer interference has been removed.

## 6. CONCLUSION

We have proposed an AILSFE structure, which significantly outperforms LSFE [5], and provides better performance than IB-DFE [6] at comparable complexity when a relatively small number of iterations are used. The LMS-SCE based adaptive channel estimation in AILSFE provides similar performance to the case with perfect CSI.

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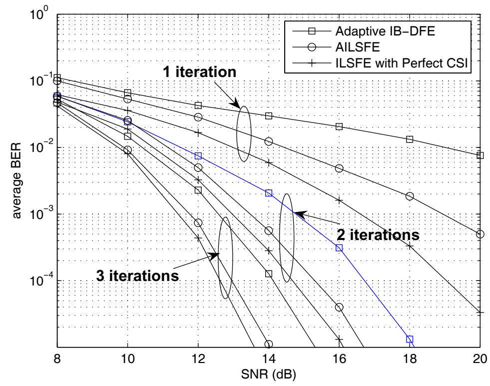


Fig.3. Performance of AILSFE, adaptive IB-DFE and ILSFE with perfect CSI (AILSFE reduces to adaptive LSFE with one iteration)

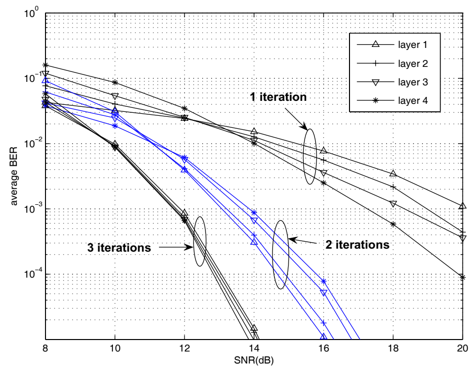


Fig.4. Performance of different layers of AILSFE