Iterative Channel Estimation for SISO and MIMO-OFDM Systems in Time-Varying Channels

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Abstract—This paper proposes a simple and efficient algorithm using polynomial interpolation for SISO and MIMO-OFDM systems in a fast time-varying channel. The time averages of the multipath complex gains, over the effective duration of each OFDM symbol, are estimated using LS criterion. After that, the time variation of the channel within several OFDM symbols are obtained by interpolating the time average values using polynomial interpolation. Specifically, we propose and evaluate the performance of channel estimation using third-degree polynomial interpolation with an adaptive pilot scheme in order to optimally use pilot tones over time varying channels. The ICI can be cancelled by using partial successive interference cancellation (PSIC) in data symbol detection instead of basic SIC method.

Keywords- Channel estimation, OFDM, MIMO, Mobile multipath channel.

I. INTRODUCTION

OFDM (Orthogonal Frequency Division Multiplexing) has been widely applied in wireless communication systems due to its high data rate transmission and its robustness to multipath channel delay [1, 2]. Additionally, multiple antenna architecture on the transmitter and receiver side, which is called multiple input multiple output (MIMO) is a suitable technique to improve the OFDM channel capacity [3]. For that reason, OFDM modulation is adopted in a number of standards, e.g IEEE 802.11a/g, IEEE 802.16a/d/e [3], DVB-T, etc.

In OFDM systems, channel estimation is usually performed by sending training pilot symbols on sub-carriers known at the receiver and the quality of the estimation depends on the pilot arrangement. Since the channel’s response is a slow varying process, the pilot symbols essentially sample this process and therefore need to have a density that is high enough to reconstruct the channel’s response at the receiver side [4]. However, OFDM system is very sensitive to Doppler shift caused by high mobility of receiver. In such case, the channel changes within one OFDM symbol and the orthogonality between subcarriers is broken resulting the inter-carrier-interference (ICI) so that system performance may be considerably degraded. The design of a channel estimator is based on two fundamental problems:

- The amount of pilot symbols to be transmitted;
- The complexity of the estimator.

The MMSE estimators have good performance but high complexity. The LS estimator has low complexity, but its performance is not as good as that of the MMSE estimators [5]. LS and MMSE criteria exhibit similar performance in high SNR regimes. However, the performance of conventional channel estimators mainly depends on the interpolation technique used in channel estimation.

For fast time varying channel, many existing works resort to estimate the channel response by a basis expansion model (BEM) [6, 7]. The BEM methods [6] used to model the channel taps are Karhunen-Loeve (KL-BEM), prolate spheroidal (PS-BEM), complex-exponential (CE-BEM) and polynomial model(P-BEM).

In the present paper, we present an iterative algorithm for channel estimation with inter-sub-carrier-interference (ICI) cancellation in MIMO OFDM systems using polynomial modeling (P-BEM) and an adaptive pilot pattern. For a Jakes’ spectrum Rayleigh gain, it has been shown in [8] that the central value and the time averaged value over one OFDM symbol are extremely closed for high realistic Doppler spread. So, for a block of OFDM symbols, we propose to estimate the time average of the complex gains, over the effective duration of each OFDM symbol of different paths, using LS criterion. After that, the time variation of the different paths within one OFDM symbol is obtained by interpolating the time averaged symbol values using third-degree polynomial interpolation.

The proposed algorithm, with less number of pilot tones and with low computational complexity, gives a good performance over the conventional methods. This proposed algorithm can in fact be considered as an extension of an algorithm for time-variant channels [8].

This paper is organized as follows. Section II presents SISO OFDM system model. In Section III, we evaluate channel estimation with third-degree polynomial interpolation. Iterative channel estimation is described in Section IV. The system simulation results are presented in Section VI.

II. SISO OFDM SYSTEM MODEL

We assume that the duration of the cyclic prefix length N_g is long enough to avoid inter-symbols interferences (ISI). The duration of an OFDM block with N subcarriers is $T=M.Ts$ where Ts is the sampling time and $M=N+N_g$.

In an OFDM system, the received OFDM symbol at time index n is given by [6, 8]

$$Y^{(n)} = H^{(n)}.X^{(n)} + W^{(n)} \tag{1}$$

Where $X^{(n)}$ is the transmitted symbol at time index n, $Y^{(n)}$ is the received symbol, $W^{(n)}$ is a white complex Gaussian noise vector of covariance matrix $\sigma^2.I_N$ and $H^{(n)}$ is N x N channel matrix, $H^{(n)}$ coefficients are given by [9,10]

$$H^{(n)}(k,m) = \frac{1}{N}.\sum_{q=0}^{N-1}\sum_{l=1}^{L} h_l^{(n)}(q).e^{j\frac{2\pi}{N}q(k-m)}.e^{-j\frac{2\pi}{N}ml} \tag{2}$$

L is the number of channel taps, $h_l^{(n)}$ is the l^{th} complex gain during the n^{th} OFDM symbol:

$$h_l^{(n)} = [h_l^{(n)}(-N_g),...,h_l^{(n)}(N-1)]^T \tag{3}$$

In the case of multipath slowly fading channels, when the channel is frequency-selective invariant over the duration of one OFDM symbol, the orthogonality between subcarriers can be fully preserved, and matrix $H^{(n)}$ would be a diagonal matrix.

III. CHANNEL ESTIMATION WITH THIRD-DEGREE POLYNOMIAL INTERPOLATION

A. Time averages of multipath channel

In this section, we propose a method based on pilot symbols to estimate time averages of multipath channel, by using the matrix notation in (1) and omitting the index time n, equation (1) can be rewritten as

$$Y = H.X + W \tag{4}$$

H is a N x N channel matrix, which contains the time average of the channel frequency response $H(m,m)$ on its diagonal and the coefficients of the inter-carrier interference (ICI) $H(k,m)$ for $k\neq m$.

$$\begin{aligned} H\ (m,m) &= \sum_{l=1}^{L}\underbrace{[\frac{1}{N}\sum_{q=0}^{N-1}h_l\ (q)]}_{h_{moy,l}^{(n)}}.e^{-j\frac{2\pi}{N}ml} \\ &= \sum_{l=1}^{L}h_{avg,l}.e^{-j\frac{2\pi}{N}ml}, m\in[1,N] \end{aligned} \tag{5}$$

For a frame of K OFDM blocks, the N_p pilot subcarriers are fixed during transmission and evenly inserted into the N subcarriers as shown in Fig. 1. N_p must fulfill the following requirement: $N_p \geq L$. Different from comb-type pilot arrangement used in [8] for time varying channel, a new pilot scheme is proposed to offer high throughput gains, this pilot scheme investigate time selectivity of the channel to reduce the number of pilot symbols. For a frame of K OFDM blocks, we just use the first, the second, $(K\text{-}1)^{th}$ and $(K\text{-}2)^{th}$ OFDM blocks for the transmission of pilot subcarriers. This approach allows varying the amount of the pilot-symbols according to channel time selectivity.

Good bit error rate (BER) performance can be achieved by the proposed third-degree polynomial interpolation, if K satisfies the following inequality

$$K.T \leq 5.T_{coh}$$

Where T is the time duration of one OFDM symbol and T_{coh} is channel coherence time.

The Doppler spread f_d , and the coherence time T_{coh}, are reciprocally related over Rayleigh fading channel [11]:

$$T_{coh} \approx \frac{9}{16.\pi.f_d}$$

Therefore, K is an integer chosen to satisfy the following inequality

$$K \le \frac{1}{2.f_d\ .T}$$

L_f denotes the interval in terms of the number of subcarriers between two adjacent pilots in the frequency domain.

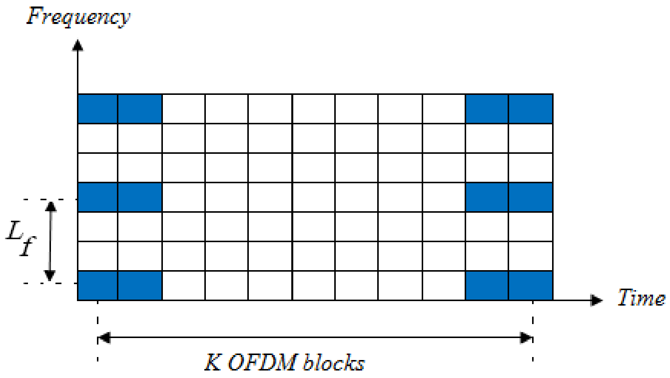


Fig. 1. Proposed pattern

By neglecting ICI contribution in (4) we can consider only diagonal elements of channel matrix H, the matrix notation of the received pilot subcarriers can be written as

$$Y_P = X_p.H_p + W_p \tag{6}$$

Where pilot matrix X_p is N_p x N_p diagonal matrix and H_p is N_p x 1 vector with elements given by (5)

$$H_p = F_p.h_{avg} \tag{7}$$

$$h_{avg} = [h_{avg,1},...,h_{avg,L}]^T \text{ with } h_{avg,l} = \frac{1}{N}\sum_{q=0}^{N-1} h_l\ (q)$$

$h_{avg,1}$ is the time average of the l^{th} path over the effective duration of the OFDM symbol.

F_p is N_p x L Fourier transform matrix given by

$$F_p = \begin{pmatrix} 1 & e^{-\frac{2j\pi k_1}{N}} & ... & e^{-\frac{2j\pi k_1 L}{N}} \\ 1 & e^{-\frac{2j\pi k_2}{N}} & ... & e^{-\frac{2j\pi k_2 L}{N}} \\ ... & ... & ... & ... \\ 1 & e^{-\frac{2j\pi k_{N_p}}{N}} & ... & e^{-\frac{2j\pi k_{N_p} L}{N}} \end{pmatrix}$$

$k_1,\dots, k_{Np}$ are the sets of N_p pilot tones inside the OFDM symbol.

The time averages of the multipath channel, over the effective duration of each OFDM symbol, are estimated using the LS criterion. By neglecting the ICI contribution, according to (6) and (7) the LS estimator of h_{avg} is given by

$$h_{avg}^{LS} = G.Y_p \tag{8}$$

$$G = (F_p^H.X_p^H.X_p.F_p)^{-1}.F_p^H.X_p^H$$

It should be noted that, the matrix $F_p^H.X_p^H.X_p.F_p$ is not invertible if $N_p < L$.

It has been shown in [8] that the mean square error (MSE) between exact averaged values $h_{avg} = [h_{avg,1},...,h_{avg,L}]^T$ and exact central values $h_c = [h_{c,1},...,h_{c,L}]^T$ is very negligible for high realistic Doppler spread ($f_d.T \le 0.1$).

B. Third-degree polynomial interpolation

Our aim is now to find the polynomial approximation of multipath channel h, based only on knowledge of h_{avg} for only *4* OFDM blocks as shown in Fig. 1. This polynomial estimation is given by [12]

$$h_{estimated,l} = Q_1{}^T.a_{c,l} \tag{9}$$

The matrix notation can be written as follow

$$\underbrace{\begin{bmatrix} h_{estimated\ ,l}(-N_g) \\ . \\ . \\ h_{estimated\ ,l}(M.K-N_g-1) \end{bmatrix}}_{h_{estim\acute{e},l}} = \underbrace{\begin{bmatrix} 1 & -N_g & (-N_g)^2 & (-N_g)^3 \\ . & . & . & . \\ . & . & . & . \\ 1 & M.K-N_g-1 & (M.K-N_g-1)^2 & (M.K-N_g-1)^3 \end{bmatrix}}_{Q_1{}^T} \underbrace{\begin{bmatrix} a_{0,l}^c \\ a_{1,l}^c \\ a_{2,l}^c \\ a_{3,l}^c \end{bmatrix}}_{a_{c,l}}$$

Notice that, $a_{c,l}$ coefficients are estimated from central values $h_c = [h_{c,1},..., h_{c,L}]^T$ approximated by averaged values $h_{avg} = [h_{avg,1},..., h_{avg,L}]^T$. Polynomial estimation of central values are given by

$$\underbrace{\begin{bmatrix} h_{c,l}(N/2) \\ h_{c,l}(M+N/2) \\ h_{c,l}((K-2).M+N/2) \\ h_{c,l}((K-1).M+N/2) \end{bmatrix}}_{h_{c,l}} = \underbrace{\begin{bmatrix} 1 & N/2 & (N/2)^2 & (N/2)^3 \\ 1 & M+N/2 & (M+N/2)^2 & (M+N/2)^3 \\ 1 & (K-2).M+N/2 & ((K-2).M+N/2)^2 & ((K-2).M+N/2)^3 \\ 1 & (K-1).M+N/2 & ((K-1).M+N/2)^2 & ((K-1).M+N/2)^3 \end{bmatrix}}_{Q_c} \underbrace{\begin{bmatrix} a_{0,l}^c \\ a_{1,l}^c \\ a_{2,l}^c \\ a_{3,l}^c \end{bmatrix}}_{a_{c,l}}$$

$a_{c,l}$ coefficients are given by

$$a_{c,l} = Q_c{}^{-1}.h_{c,l} \tag{10}$$

Under the approximation of central values by averagedvalues, we can write

$$h_{c,l} = h_{avg,l}^{reduced} = [h_{avg,l}^{(1)}, h_{avg,l}^{(2)}, h_{avg,l}^{(K-2)}, h_{avg,l}^{(K-1)}]^T \tag{11}$$

From (9), (10) and (11) the polynomial approximation of multipath channel *h* is given by

$$h_{estimated\ ,l} = Q_1^T.Q_c^{-1}.h_{avg,l}^{reduced} \tag{12}$$

Knowing that

$$h_{c,l} = T_2.h_{avg,l} \tag{13}$$

Where T_2 is *4* x *K* diagonal matrix given by

$$T_2 = diag([1,1,0,..,0,1,1]^T)$$

From (9), (10) and (13) proposed polynomial estimation of h_l can be written as follow

$$\begin{aligned} h_{estimated,l} &= Q_1{}^T.Q_c{}^{-1}..T_2.h_{avg,l} \\ &= V_1.h_{avg,l} \end{aligned} \tag{14}$$

The MSE of this third-degree polynomial modeling is given by

$$\begin{aligned} EQM_2 &= E(e_l.e_l{}^H) = E\left\{(h_l - h_{estimated\ ,l}).(h_l - h_{estimated\ ,l})^H\right\} \\ &= \frac{1}{M.K}Tr\left\{R_{h_l} + V_1.R_{h_{avg,l}}.V_1^T - R_{h_l,h_{avg,l}}.V_1^T - V_1.R_{h_l,h_{avg,l}}^H\right\} \end{aligned} \tag{15}$$

The MSE of polynomial modeling in [8] is given by

$$\begin{aligned} EQM_1 &= E(e_l.e_l{}^H) = E\left\{(h_l - h_{estimated\ ,l}).(h_l - h_{estimated\ ,l})^H\right\} \\ &= \frac{1}{M.K}Tr\left\{R_{h_l} + V.R_{h_{avg,l}}.V^T - R_{h_l,h_{avg,l}}.V^T - V.R_{h_l,h_{avg,l}}^H\right\} \end{aligned} \tag{16}$$

Knowing that $V = Q_1{}^T.T^{-1}$ where *T* is *K* x *K* matrix, *T* elements are given by [8]

$$\begin{cases} T(i,j) = \dfrac{1}{N}\left\{((i-1)M)^{j-1} + ((i-1)M+1)^{j-1} + ... + ((i-1)M+N-1)^{j-1}\right\} \ if \ \ j \neq 1 \\ T(i,1) = 1 \end{cases}$$

R_{h_l} is a correlation matrix of h_l , $R_{h_{avg,l}}$ is a correlation matrix of $h_{avg,l}$ and $R_{h_l,h_{avg,l}}$ is a cross-correlation matrix between h_l and $h_{avg,l}$ [8].

Fig. 2 provides a comparison between the MSE of proposed modeling and polynomial modeling in [8]. Fig. 2 illustrates the evolution of MSE versus *K*, proposed third-degree polynomial modeling performs better than polynomial modeling in [8]. Especially, if *K* is chosen to satisfy the following inequality

$$K \leq \frac{1}{2 f_d \ .T} \tag{17}$$

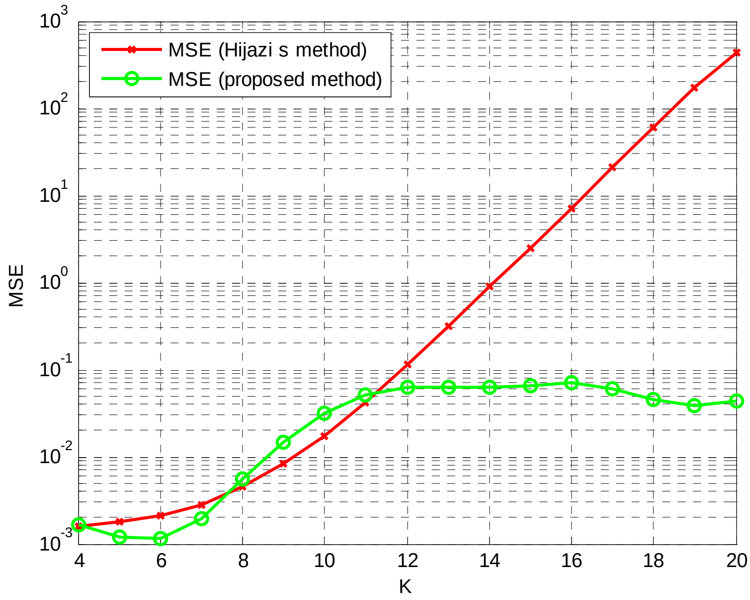


Fig. 2. (*a*) *MSE* versus *K*: $f_d\,.T=0.1$

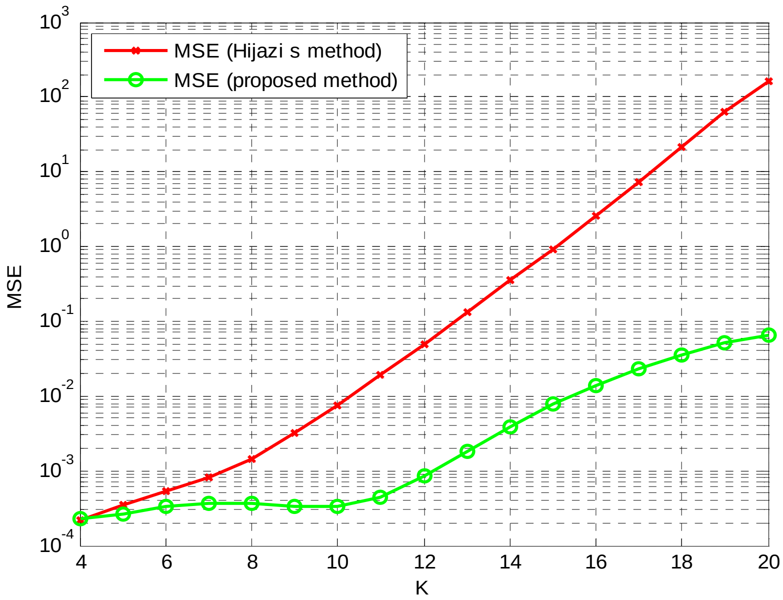


Fig. 2. (*b*) *MSE* versus *K*: $f_d\,.T=0.05$

We can see that for high values of *K* (*K* > *5*), condition (17) is not satisfied, that's why there's significant performance decrease. However, worst performances are observed with polynomial modeling presented in [8] because of polynomial sides effects in high degree polynomial modeling (as can be seen in Fig. 4). The performance is rather sensitive to the Doppler spread as shown in Fig.3. *MSE* is negligible for high realistic Doppler spread ($f_d.T \leq 0.1$).

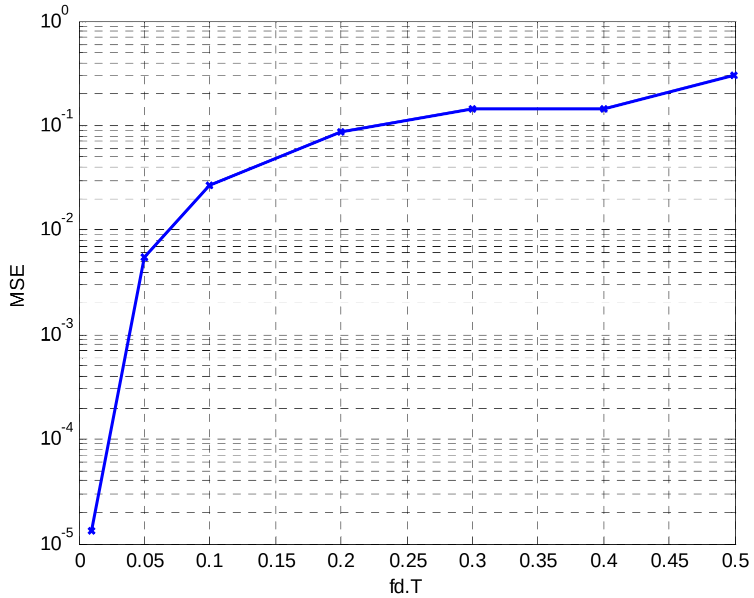


Fig. 3. *MSE* versus $f_d .T$ *for N=64 and* $K = \left\lceil \frac{1}{10 f_d .T} \right\rceil$

C. Comparison between comb-type pilot arrangement and proposed scheme

According to previous results [13], throughput gain of the proposed pattern shall achieve more than 4 % over time varying channels with moderate mobility.

Proposed third-degree polynomial modeling achieves similar performances as higher polynomial modeling (*K-1*) presented in [8] with comb type pilot arrangement. For the purpose of comparison, Fig. 4 and 5 show an example of interpolation using polynomial modeling in [8] and third-degree polynomial modeling proposed in this paper.

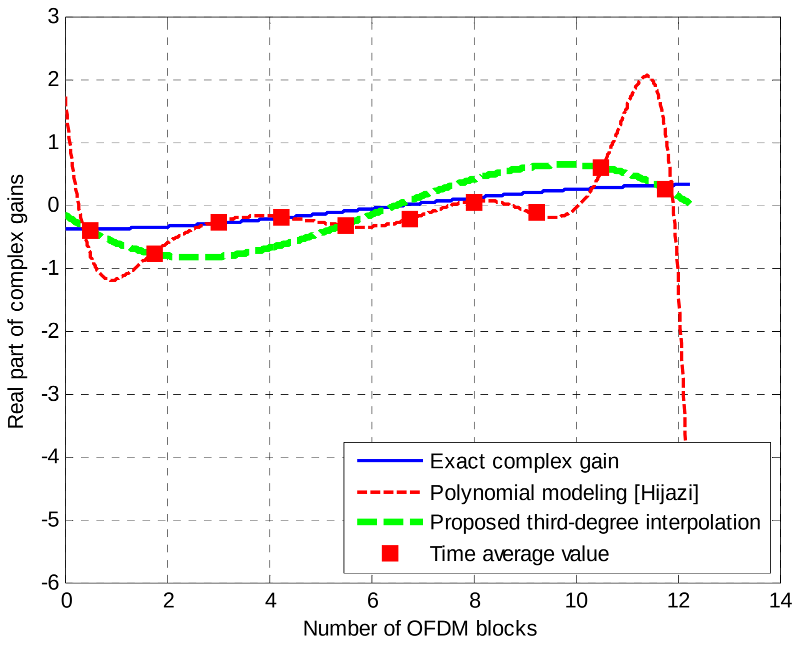


Fig. 4. Comparison between polynomial interpolation in [8] (9 degree polynomial) and proposed third-degree polynomial interpolation (f_d .T=0.05, K=10)

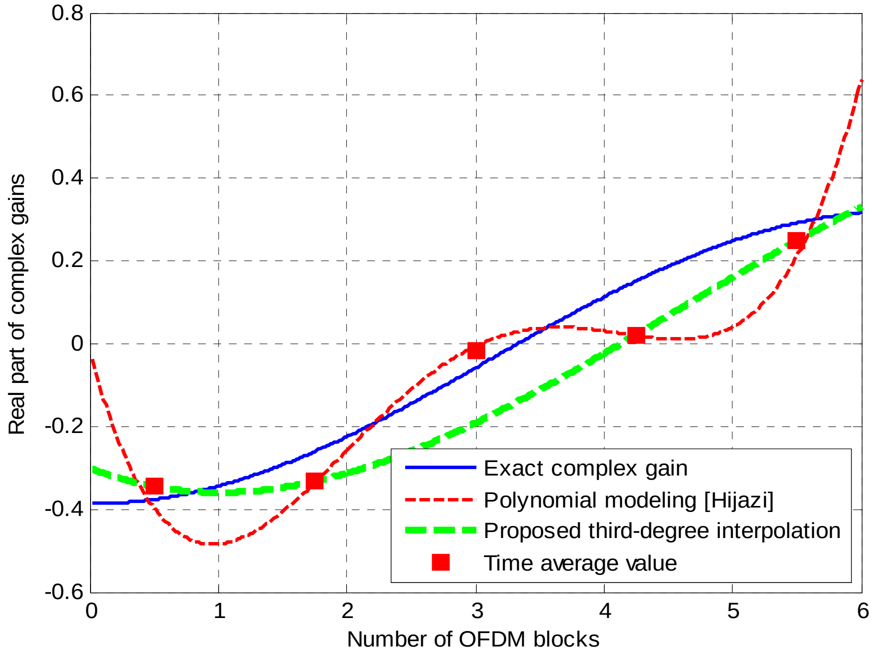


Fig. 5. Comparison between polynomial interpolation in [8] (4 degree polynomial) and proposed third-degree polynomial interpolation ($f_d.T=0.1$, $K=5$)

Proposed polynomial interpolation performs better than polynomial modeling in [8]. Especially, computational complexity is lowered using third-degree polynomial interpolation instead of 9 degree polynomial (Fig. 4) or 4 degree polynomial (Fig. 5). As it can be seen in Fig. 4 and 5, the mean square error (MSE) between averaged values and central values is sensitive to the Doppler spread.

IV. ITERATIVE CHANNEL ESTIMATION

Pilot based channel estimation presented in section III, use a few training symbols to estimate channel response in time domain. However, robust channel estimation techniques which use training symbols and blind data completely are attractive. Moreover, such techniques tend not to use the information in the unknown data symbols to improve channel estimates. Semi-blind channel estimation techniques can potentially enhance the quality of estimates by making a more complete use of available data. In this section, we combine channel estimation technique presented in section III with equalization techniques to mitigate ICI contribution caused by high mobility of receiver. The conventional equalization techniques of OFDM exhibit relatively good performance at low values of normalized Doppler frequency ($f_d.T << 0.1$). However, in an environment where the normalized Doppler frequency is high, there is an important error floor even if all data are pilot symbols, since the pilot symbols themselves are corrupted by the ICI. Therefore, the channel estimation and equalization should be performed together using an iterative way.

In an environment where the normalized Doppler frequency is high, the channel becomes time-selective. From another point of view, time-varying channel not only destroys the orthognality between subcarriers, but also provides time diversity. Therefore, conventional detectors suffer from performance degradation in time-varying channels [14]. In order to fully utilize the time diversity caused by time-varying channel, the successive interference cancellation (SIC) technique presented in [14] detect the data one-by-one instead of detecting all the data simultaneously, as in conventional methods. In section A, we assume that the channel impulse response is known at each time and for each tap, and we discuss the detection problem assuming the channel is known. Channel estimation is addressed in section B.

A. Partial successive interference cancellation (PSIC)

To reduce the complexity of successive interference cancellation (SIC) detector presented in [14], we only count in the ICI power contributed from the 2 closest subcarriers. It has been shown in [15] that the most significant ICI contributions on the m^{th} subcarrier come from the closest subcarriers. We evaluate this argument by studying the ICI power P_ψ distributed to subcarriers from $n-\psi$ to $n+\psi$ after the transmission of symbol X_m on the m^{th} subcarrier.

For Rayleigh fading channel, P_ψ is given by [15]

$$P_\psi = \frac{E_S}{N^2}\sum_{k=0}^{N-1}\sum_{k'=0}^{N-1}[J_0(2\pi f_d T_s(k'-k)).\sum_{q=-\psi}^{\psi} e^{-j2\pi(k'-k).q/N} \quad (18)$$

The curve in Fig. 6 shows the ICI power P_ψ versus the number of considered adjacent subcarriers for various values of $f_d.T$.

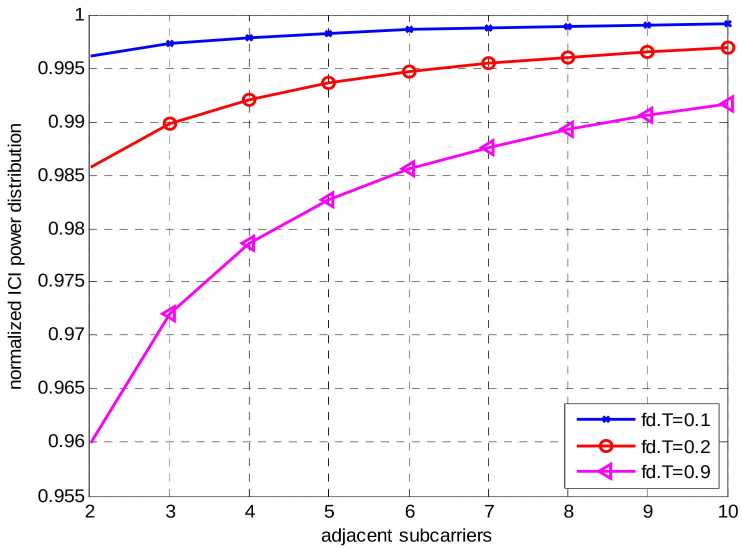


Fig. 6. Normalized ICI power distribution over adjacent subcarriers ($N=64$)

Fig. 6 shows that as well as $f_d.T$ <0.2, 99% of power is distributed over 3 subcarriers (the m^{th} subcarrier and two closest subcarriers).

Therefore, received OFDM block can be approximated as follow

$$Y \approx \tilde{H}.X + W \quad (19)$$

Where

$$\tilde{H} = \begin{bmatrix} H(1,1) & H(1,2) & 0 & \dots & & 0 \\ H(2,1) & H(2,2) & H(2,3) & 0 & \dots & 0 \\ \vdots & & & & & \vdots \\ 0 & \dots & & 0 & H(N,N-1) & H(N,N) \end{bmatrix}$$

We only count in the ICI power contributed from the two closest subcarriers. As shown in Fig. 10 and 11, PSIC has the same performances as basic SIC (Fig. 10 and 11).

For the purpose of complexity comparison, number of required multiplications is given in Table 1. The proposed PSIC requires *4N-2* multiplications to detect N symbols, instead of N^2 multiplications required in SIC. This greatly reduces the complexity and cost of our algorithm.

TABLE 1
Number of multiplications required for PSIC and SIC

	Number of multiplications	**Numerical results**	
		N=128	*N=512*
SIC in [15] with LS	N^2	16384	262144
Proposed PSIC with LS	***4N-2***	510	2046

The numerical results given in Table 1, shows the computational advantage of the proposed PSIC in comparison with basic SIC.

B. Iterative algorithm for SISO OFDM systems

In section A, we have investigated the problem of equalization in the presence of ICI assuming the ideal channel response is available.

Now we aim to cancel the ICI term in (19) knowing that channel response is unknown, the iterative technique of channel estimation and ICI mitigation is shown in Fig. 7.

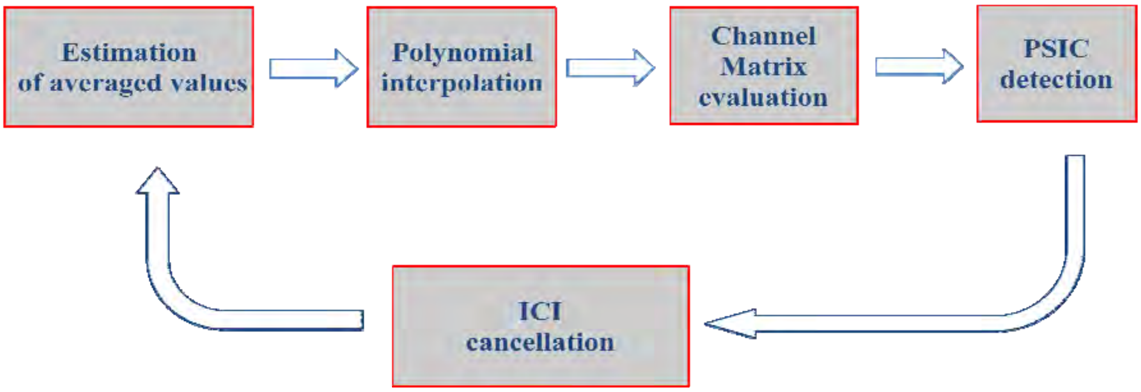


Fig. 7. Channel estimation with iterative ICI cancellation

The algorithm is divided into two modes:

- **Channel estimation mode:** This mode includes estimating the time average values using *LS* estimator and third-degree polynomial interpolation of the averaged values.
- **Detection mode:** includes the detection of data symbols by using partial successive interference cancellation (PSIC) presented in section A.

As shown in Fig. 7, a feedback technique is used between channel estimation mode and data detection mode, performing iteratively ICI suppression and channel matrix estimation.

For a frame of K OFDM symbols, the iterative algorithm proceeds as following:

➢ **Estimation and interpolation of averaged values:** The average values of multipath channel are given by (8)

$$h_{avg}^{LS,k} = G.Y_p^k$$

Third-degree polynomial interpolation is given by (14)

$$\begin{aligned} h_{estimated\ ,l} &= Q_1^T.Q_c^{-1}..T_2.h_{avg,l} \\ &= V_1.h_{avg,l} \end{aligned}$$

➢ **Channel matrix evaluation:** Channel matrix H^k is estimated Using (5);

➢ **Remove the ICI of pilots from the received data subcarriers** Y_d^k

➢ **Detection of data symbols** $\hat{X}_d^k$ **using PSIC**

➢ **ICI cancellation**

Y_p^k is updated and ICI is cancelled as follow

$$Y_p^k(new) = Y_p^k(old) - \hat{H}_{ICI}^k.X^k$$

Where $\hat{H}_{ICI}^k$ is N_p x N channel matrix obtained from H^k with zero diagonal;

➢ **Re-iteration:** $i=i+1$

In the first iteration ($i=1$), channel and data information are unknown, the first channel information is made without ICI cancellation. After the tentative estimation of channel and data, ICI can be regenerated and partially cancelled. At the third loop of iteration, the renewed estimates of channel and data are pretty reliable, and the regenerated ICI is regarded to be reliable and then can be cancelled completely;

C. Complexity analysis

The computational analysis of the proposed algorithm is given in Table 2 step by step.

Numerical results given in Table 3, shows the computational advantage of the proposed method in comparison with a similar technique in literature [8].

TABLE 2
Number of multiplication per step

	Iterative algorithm in [8]	Proposed algorithm
Step *i* et *ii*	$L.\{(K+1).N+N_P+K.N_g\}$	$L.\left\{5N+\frac{4}{K}.N_P+4.N_g\right\}$
Step *iii*	$N_p(N-N_p)$	$N_p(N-N_p)$
Step *iv*	$(N-N_p)^2$	$4(N-N_p)-2$
Step *v*	$N_p(N-1)$	$2(N_p-1)$
Number of multiplications	$Nbr\_iter.[L.((K+1).N+N_p+K.N_g)+N_p(N-N_p)+(N-N_p)^2+N_p(N-1)]$	$Nbr\_iter.[L.(5N+\frac{4}{K}.N_P+4.N_g)+N_p(N-N_p)+4(N-N_p)-2+2(N_p-1)]$

Table 3 shows numerical results for an OFDM system with N=64, N_g=N/8, N_p =N/8 and $L=6$ paths.

TABLE 3
Number of multiplication per subcarrier

	Iterative algorithm in [8]	**Proposed algorithm**
f_d.T=0.05 K=10 iteration=2	*147 x 2=294*	*48 x 2=96*
f_d.T=0.1 K=5 iteration=3	*108 x 3=324*	*47 x 3=141*

As can be seen from the above computational analysis, proposed iterative channel estimation using third-degree polynomial interpolation and PSIC detector shows very interesting low complexity in comparison with iterative channel estimation technique in [8].

V. EXTENSION TO MIMO OFDM SYSTEM

We consider the block diagram of MIMO-OFDM system with N_t transmit antennas, N_r receive antennas, and N subcarriers (Fig. 8). Generated OFDM signals are transmitted through a number of antennas in order to achieve diversity.

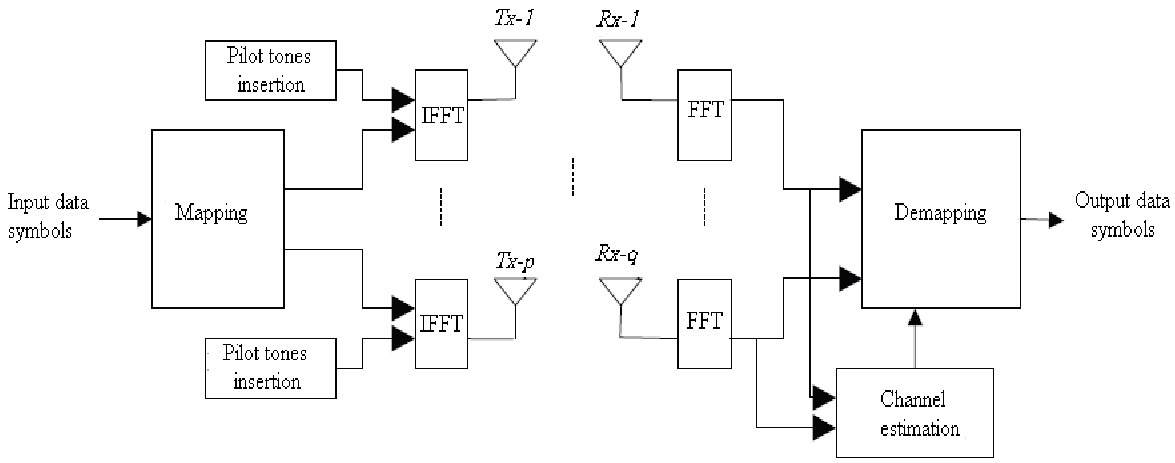


Fig. 8. Block diagram of MIMO-OFDM system

Proposed pilot design for MIMO OFDM system is given in Fig. 9. The idea behind the use of this design is to prevent the transmission of data on OFDM blocks with the pilot subcarriers, in order to bypass the step of equalization in iterative channel estimation, this equalization is relatively complex in a MIMO system but essential for ICI cancellation if pilot symbols and data symbols are carried by the same OFDM blocks.

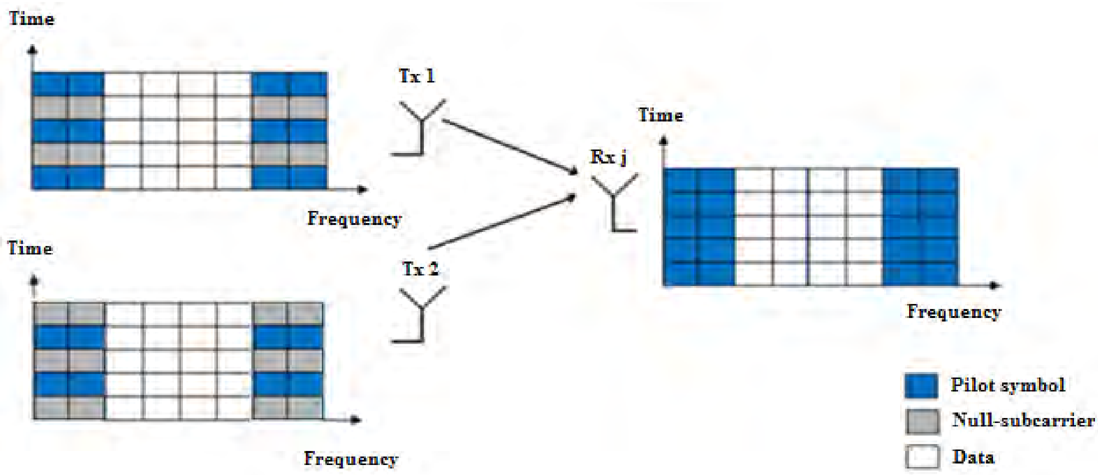


Fig. 9. Proposed pilot pattern for MIMO OFDM system

In MIMO OFDM system, the proposed iterative algorithm proceeds as following:

- **Estimation and interpolation of averaged values:** The average values of multipath channel are given by

$$h_{avg}^{ji,LS} = G^i.Y_P^j(p^i)$$

h_{avg}^{ji} is L x 1 vector of time average values for sub-channel (i,j), between Tx- i and Rx- j. Y_p^j is the received OFDM block on antenna Rx-j.

Third-degree polynomial interpolation is given by (14);

- **Channel matrix evaluation:** Channel matrix $\hat{H}^{ij}$ is estimated Using (5);

- **Remove the ICI of pilots from the received data subcarriers** Y_p^j

$$Y_p^j(p^i) = Y_p^j(p^i) - \hat{H}_{ICI}^{ij}.X_p^i$$

Where $X_P^i = diag([X_{k_1}^i, ..., X_{k_{N_{P1}}}^i])$

X_P^i is N_p x N_p diagonal matrix (diagonal elements of X_P^i are the pilot symbols transmitted by the i th antenna)

$\hat{H}_{ICI}^{ij}$ is N_p x N channel matrix obtained from $\hat{H}^{ij}$ with zero diagonal;

- **Re-iteration:** *i=i+1*

VI. SIMULATION RESULTS

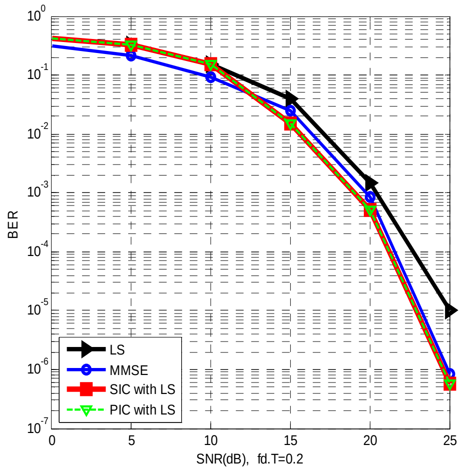
A. The performance evaluation of proposed PSIC detector

We evaluate the BER performance of the proposed detector for system whose parameters are listed in Table 4.

TABLE 4
System parameters

Physical mode	value
Carrier frequency	5 GHz
Channel bandwidth	2 MHz
Sampling time	0.5 µs
Subcarrier number	128
OFDM Symbol duration	72 µs
Cyclic Prefix (CP)	8 µs
Modulation scheme QPSK,	16 QAM
Channel model	Rayleigh
Vehicle speed (km/h)	300 km/h-590 km/h

The BER performances of proposed PSIC with LS and SIC with LS are compared to conventional detectors (LS and MMSE). Obviously the proposed PSIC detector outperforms the conventional detectors for moderate values of normalized Doppler frequency, and PSIC shows similar performances as SIC (Fig. 10 and Fig. 11). That means the proposed PSIC detector can efficiently mitigate the induced ICI with very interesting low complexity as it is shown in section IV-A.



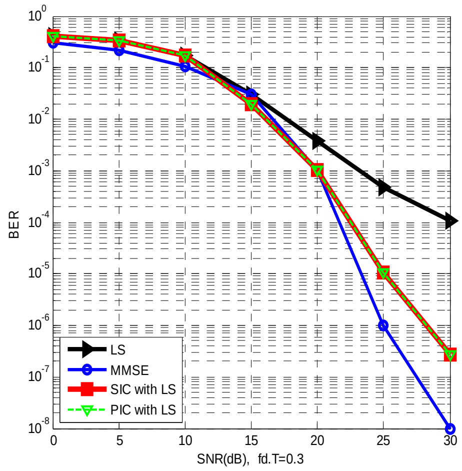


Fig. 10. BER versus SNR (16-QAM) for $f_d.T$ =0.2   *Fig. 11. BER versus SNR (16-QAM) for $f_d.T$ =0.3*

B. The performance evaluation of proposed iterative channel estimation

In this section, we verify the theory by simulation and we test the performance of the proposed iterative algorithm. We examine the mean square error (MSE) and the bit error rate (BER) performances in terms of the average signal-to noise ratio (SNR) and normalized Doppler spread $f_d.T$ ($f_d.T$ = 0.05, $f_d.T$ = 0.1) for Rayleigh fading channel. Fig. 12 and 13 show the evolution of the mean square error (MSE) with the iterations in terms of SNR for $f_d.T$ = 0.05 and $f_d.T$ = 0.1. A great improvement is realized by the second iteration. In Fig. 12 ($f_d.T$ = 0.05), the third iteration improvement is not important, this is because ICI is completely removed at the second iteration. However, In Fig. 13 ($f_d.T$ = 0.1), the third iteration improvement is significant because ICI contribution is important and can't be completely removed by the second iteration.

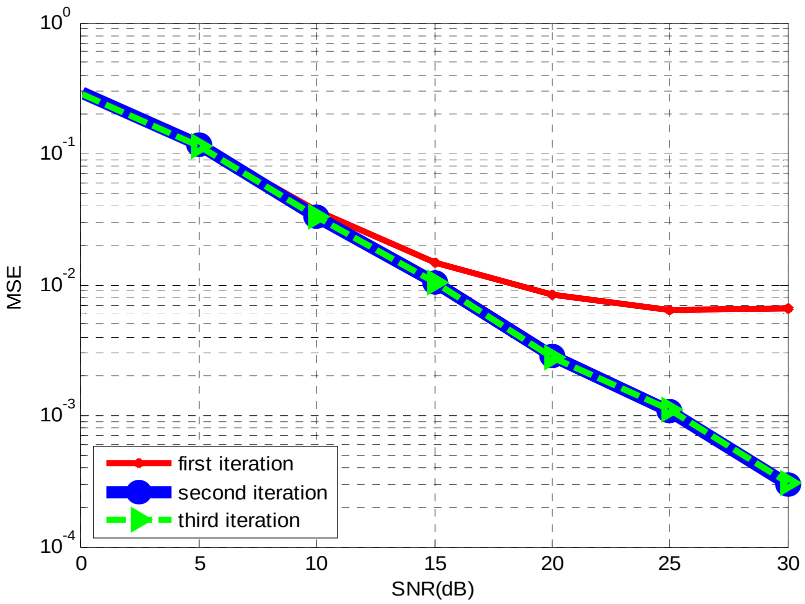


Fig. 12. MSE versus SNR for $f_d.T$ =0.05

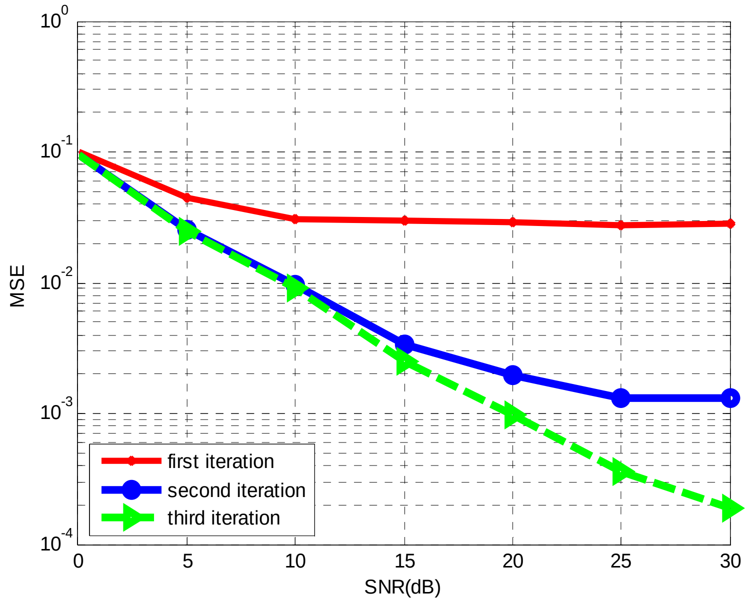


Fig. 13. MSE versus SNR for $f_d.T$ =0.1

Fig.14 and 15 give the BER performance of our proposed iterative algorithm, compared to iterative algorithm in [8] and successive interference cancellation (SIC) with perfect channel knowledge as reference.

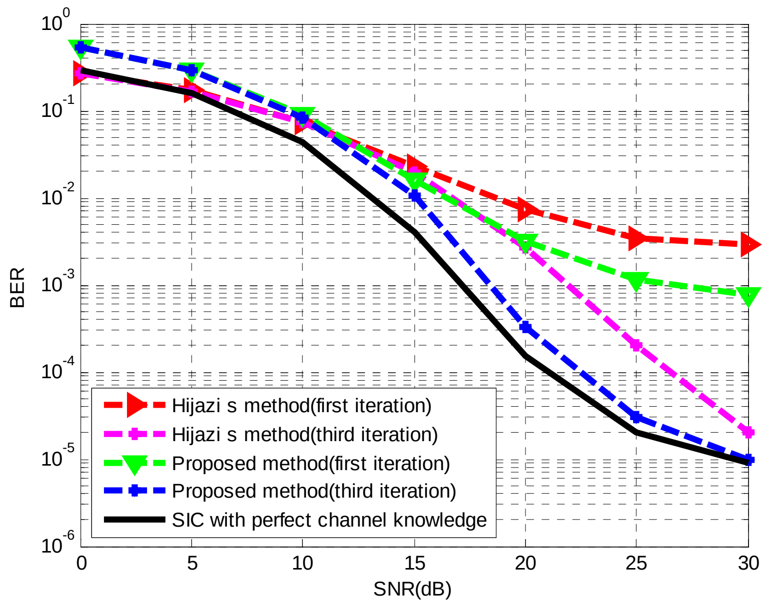


Fig. 14. BER versus SNR for $f_d.T$ =0.1 (8-QAM)

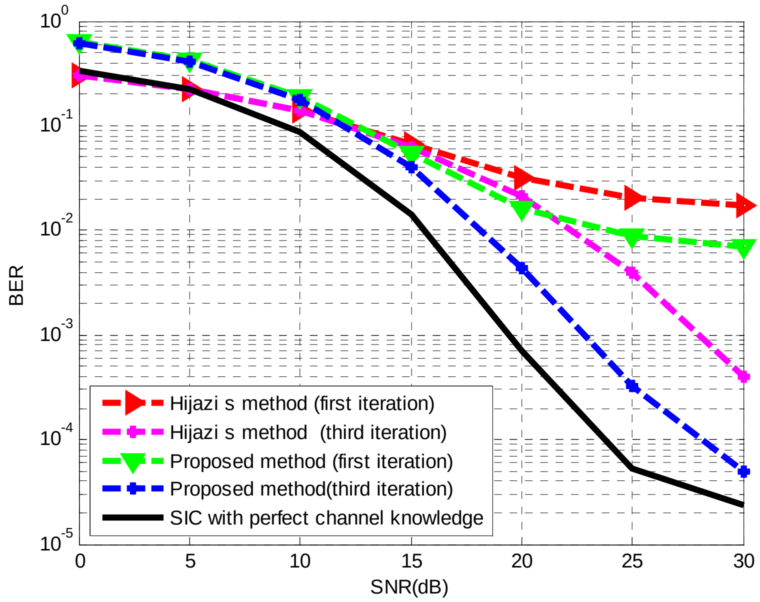


Fig. 15. BER versus SNR for $f_d.T$ =0.1 (16-QAM)

This result shows that, with ICI, our algorithm performs better than similar methods. Moreover our iterative algorithm offers an improvement in BER. After two iterations, a significant improvement occurs; the performance of our algorithm and the SIC algorithm with perfect channel knowledge are very close.

VII. CONCLUSION

The proposed iterative channel estimation algorithm for OFDM systems is specifically tailored to estimate time varying channels and to mitigate the inter-carrier-interferences (ICI). The channel time evolution is tracked by third-degree polynomial interpolation and perfectly estimated. Theoretical analysis and simulation results of our iterative algorithm show that by estimating and removing the ICI we have a great improvement especially after the first iteration for high realistic Doppler spread. Moreover, our algorithm performs better than the existing techniques and its computational complexity is lowered in comparison with conventional methods.

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