A performance analysis of optimized semi-blind channel estimation method in OFDM systems

Sangirov Gulomjon*, Fu Yongqing*, Jamshid Sangirov**, Fang Ye* and Ahmad Olmasov***

**Information and Communication Engineering College, Harbin Engineering University, Harbin, 150001 China*

***Samsung Electronics, South Korea*

****Samarkand branch of Tashkent University of Information Technologies, Uzbekistan*

gulomjons@hrbeu.edu.cn

***Abstract* — Nowadays, one of the effectively used technique in wireless communication area is an orthogonal frequency division multiplexing (OFDM). In OFDM systems, channel impairments due to multipath dispersive wireless channels can cause deep fades in wireless channels. Therefore, an accurate and computationally efficient channel state information necessary when coherent detection is involved in the OFDM receiver. Hence, it is essential to have a good channel estimation method for OFDM systems in wireless communication. And normally one of the good channel estimation methods is a semi-blind channel estimation. On the other hand, the semi-blind method requires a large number of processing operations. In order to avoid the high complexity of the existing method, the semi-blind channel estimation has been optimized. At the receiver side, we calculate subspace decomposition for blind channel estimation and further to improve channel estimation we use training based technique to estimate channel state information. Next, we combine these channel estimations as semi-blind channel estimation methods and we optimized semi-blind channel estimation by choosing an optimal technique for training based channel estimation.**

***Keyword*—Semi-blind channel estimation, OFDM, least square and scaled LS**

I. INTRODUCTION

In wideband digital communications the orthogonal frequency division multiplexing (OFDM) is used for splitting a high-rate datastream into number of lower rate streams that are transmitted simultaneously over a number of subcarriers for easy transmission. The OFDM technique is applicable in digital terrestrial multimedia broadcast (DTMB) [1], digital subscriber line (DSL) broadband internet access, wireless network, long term evolution (LTE) [2-3], and 4G the transmitter modulates the message bit sequences into phase shift keying (PSK) / quadrature amplitude modulation (QAM) symbol. And then it performs inverse discrete fourier transform (IDFT) on the symbols for conversion them from the frequency domain to time-domain signals. Usually, next step is the insertion of cyclic prefix (CP) in OFDM system. The reason for the CP is to avoid intercarrier interference (ICI) which occurs by a multipath channel. And it also provides good bandwidth efficiency on the receiver side. In our OFDM system, we use zero padding (ZP) instead of CP. Generally, the ZP replaces nonzero CP by zeros. The ZP-OFDM system has the same spectral efficiency as CP-OFDM system by the condition of the number of zero symbols equals the CP length. Lastly, the transmitter sends the time-domain signals out through a wireless channel.

In OFDM system, a wireless channel plays a big role for the transmission performance. That is why estimating the channel has a significant impact on the efficiency of the transmission performance. We observe one of efficient channel estimation methods that are widely utilized in OFDM systems is called a semi-blind channel estimation. The importance of using a semi-blind channel estimation method is a tradeoff between computational complexity and spectral efficiency. To accomplish channel estimation numerous works subspace decomposition methods have been proposed [4-6]. However, these methods use complex computational schemes that may also reduce spectral efficiency. To improve the spectral efficiency of the channel a subspace pursuit algorithm has been proposed estimation in [7]. This algorithm uses a combination of two algorithms to work for low pilot density, which makes the implementation complicated. A method improving the channel estimation with lower computational difficulty has been introduced in [8]. Nevertheless, it works well when there are few OFDM symbols. Consequently, for working out all symbols the computation amount will be increased. Another method proposed to decrease channel estimation error by subspace estimation bases on a block matrix [9]. But, the calculation and formation of a burst of stacked OFDM symbols create

Manuscript received on February 19, 2016. This work is a follow-up of the invited journal to the accepted out-standing conference paper of the 17th International Conference on Advanced Communication Technology (ICACT2015).

Gulomjon Sangirov is with Modern Communication and Information Systems Research group in the school of Information and Communication Engineering, Harbin Engineering University, Harbin, 150001, China (Corresponding author phone: +86-451-82568266; fax: +86-451-82530010 e-mail: gulomjons@hrbeu.edu.cn).

Yongqing Fu is with Modern Communication and Information Systems Research group in the school of Information and Communication Engineering, Harbin Engineering University, Harbin, 150001, China (e-mail: yongqingfu @hrbeu.edu.cn).

Jamshid Sangirov is with Samsung Electronics, South Korea (e-mail: jamshid@kaist.ac.kr).

Fang Ye is with Modern Communication and Information Systems Research group in the school of Information and Communication Engineering, Harbin Engineering University, Harbin, 150001, China (e-mail: fangye@hrbeu.edu.cn).

Ahmad Olmasov is with telecommunication engineering research group at Samarkand branch of Tashkent University of Information Technologies.

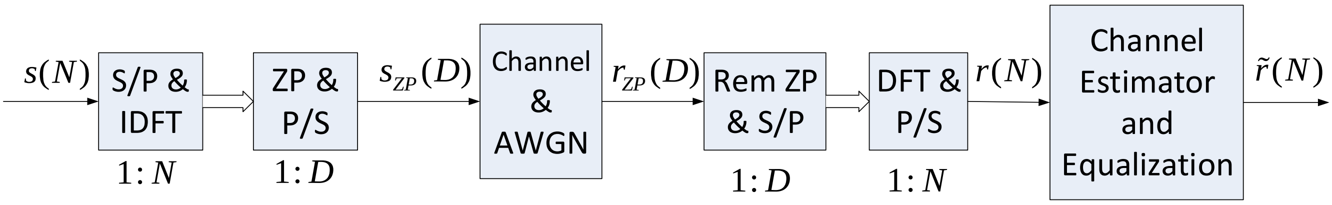

Fig. 1 SISO ZP-OFDM system model

extra complexity by increasing the computational power in channel estimation. A redundant linear precoding based semi-blind channel estimation is developed in [10]. Moreover, it is complex to construct a matrix for semi-blind channel estimation. Therefore, we improved the spectral efficiency and reduced computation complexity of semi-blind channel estimation in single input single output (SISO) OFDM systems. With the intention to evaluate the efficiency of the modified semi-blind channel estimation technique, we compared the simulation results with conventional channel estimation techniques.

II. Description for System Model

In this part, we overview OFDM system model. We consider the system model with a total number of N size of OFDM subcarriers. Hence, the m^{th} transmitted block of OFDM symbols can be stated as $\tilde{\mathbf{S}}_N(m) = \left[\tilde{S}_1(m), \cdots, \tilde{S}_N(m)\right]^T$, and here $[\cdot]^T$ is the transpose operator. First, we apply IDFT process and the signal will be

$$\mathbf{s}_N(m) = [s_1(m), \cdots, s_N(m)]^T = \mathbf{F}_N^H \tilde{\mathbf{S}}_N(m) \tag{1}$$

here $\mathbf{F}_N = (1/\sqrt{N}) e^{j2\pi kn/N}$ is IDFT matrix with size N, where $(k = 0, \pm 1, \pm 2..,)$ are discrete Fourier series coefficients. Then we add ZP to $\mathbf{s}_N(m)$, and after that the OFDM symbol becomes $\mathbf{s}_{ZP}(m) = [s_1^{ZP}(m), \cdots, s_D^{ZP}(m)]^T$, here is $D = N + P$, and P is the ZP length.

The FIR filter models channel impulse response as $\mathbf{h}_D = [h_0, \cdots, h_{D-1}]^T$. Normally, the channel order size is longer than ZP size which is

$$h_D = \begin{cases} h_i & 0 \le i \le L \\ 0 & else \end{cases}. \tag{2}$$

Hence according to (2) channel parameter vector is assumed as $\mathbf{h} = [h_0, \cdots, h_L]^T$. And in consequence of each transmitting OFDM symbol size, the channel vector could be expressed as following $\mathbf{h}_D = [h_0, \cdots, h_L, 0, \cdots, 0]^T$.

After transmitting $\mathbf{s}_{ZP}(m)$ through the SISO channels, the discrete-time signal from the receiver antenna is given by

$$\mathbf{r}(\mathrm{t}) = \sum_{l=0}^{L} \mathbf{h}(l)\mathbf{s}(t) + \mathbf{n}(t) \tag{3}$$

where $\mathbf{h}(l) = \left[h_1(l), \cdots, h_D(l)\right]$ and $\mathbf{n}(t)$ is additive gaussian white noise (AWGN) with zero-mean and variance σ_n^2.

In the receiver block, the received stacking signal becomes $\mathbf{r}(t), t = mD+1, ..., mD+D$. Hence, the m^{th} intersymbol interference (ISI) free vector can be obtained by the insertion of ZP in the transmitted signal as follows

$$\mathbf{r}_m = \mathbf{H}\mathbf{s}_m + \mathbf{n}_m \tag{4}$$

where $\mathbf{H}$ is filtering matrix with $D \times D$ lower triangular toeplitz matrix with the first column $[h_0, \cdots, h_L, 0, \cdots, 0]^T$ and the first row $[h_0, 0, \cdots, 0]$. The m^{th} received block of symbols $\mathbf{r}_{ZP}(m)$ can be expressed as

$$\mathbf{r}_{ZP}(m) = \mathbf{H}\mathbf{s}_{ZP}(m) + \mathbf{n}(m). \tag{5}$$

Supposing frequency synchronization and perfect timing, the OFDM demodulator eliminates the ZP and then transform the rest received signals to the frequency domain by DFT for obtaining the frequency domain ZP free received signals. After ZP removal from expression (4), the simplified system mathematical model can be described as

$$\mathbf{R}(m) = \mathbf{H}_N \mathbf{S}_N(m) + \mathbf{N}(m) \tag{6}$$

where $\mathbf{H}_N = \sqrt{N}\mathbf{F}_N \mathbf{h}_N$ is the channel frequency response and $\mathbf{N}(m)$ is AWGN vector after transformation of DFT.

Now let us consider the transmitted m^{th} received block block of OFDM symbols passed through multipath fading channel and m^{th} received signal block became as

$$\mathbf{r}(m) = \mathbf{A}\mathbf{S}_{ZP}(m) + \mathbf{n}(m) \tag{7}$$

where $\mathbf{A} = \mathbf{WH}$, where $\mathbf{W}$ is OFDM modulation matrix which consists of $[\mathbf{W}]_{dn} \triangleq W_N^{-(N-1-d)n}$, $n = 0, ..., N-1$, $d = 0, ..., D-1$ and $W_N \triangleq e^{-j2\pi/N}$ and $\mathbf{n}(m)$ is gaussian white noise with variance σ^2. And the problem statement here is the estimation of $\mathbf{H}$ channel accurately by using fewer pilot signals which contained in received signal.

III. The Channel Estimation

A. Semi-blind channel estimation

First, we estimate the channel in second order statistics blindly because we do not have any information about transmitted signals. The satisfying situation here $\mathbf{H}$ can be identified blindly up to ambiguity matrix $\mathbf{A}$. For subspace channel estimation we calculate singular value decomposition (SVD) of received stacking signal matrix $\mathbf{r}(m)$ according to (7). Or also we can calculate an eigenvalue decomposition (EVD) of autocorrelation matrix of received stacking signal $\mathbf{r}(m)$ which we have

$$\mathbf{R_{rr}} = \mathbf{H}\mathbf{R_{ss}}\mathbf{H}^H + \mathbf{R_{nn}}. \tag{8}$$

And here $\mathbf{R_{rr}}$, $\mathbf{R_{ss}}$ and $\mathbf{R_{nn}}$ notations are the autocorrelation matrix of $\mathbf{r}(m)$, $\mathbf{s}_N(m)$, $\mathbf{n}(m)$ and $\mathbf{R_{nn}} = \sigma^2 \mathbf{I}_D$, and $\mathbf{I}_D$ is a unit matrix. And the autocorrelation matrices $\mathbf{R}_{\mathrm{rr}}$ and $\mathbf{R_{ss}}$ stand as full rank. Now we get D eigenvalues $\lambda_1, \cdots, \lambda_D$, after performing EVD of $\mathbf{R_{rr}}$. As the result we have $\lambda_1 \ge \cdots \ge \lambda_N \ge \lambda_{N+1} = \cdots = \lambda_D = \sigma^2$ in descending order, here $\lambda_1, \cdots, \lambda_N$ is an area for the signal subspace, also $\lambda_{N+1}, \cdots, \lambda_D$ is an area for the noise subspace. Suppose we can notate the noise subspace vector as $\mathbf{g}_i$ ($0 \le i \le L-1$) and due to the orthogonal subspace signal to $\mathbf{g}_i$, we can note that

$$\mathbf{g}_i^H \mathbf{H} = 0 \tag{9}$$

here the channel matrix $\mathbf{H}$ has a similar structure with channel matrix used in (3). Now we can transform (9) equivalently as given

$$\mathbf{G}_i \mathbf{h} = 0 \tag{10}$$

here $\mathbf{g}_i^H$ vector is transformed into $\mathbf{G}_i$ a matrix which consists of $D \times D$ dimension lower triangular toeplitz matrix with the first column $[\mathbf{g}_0^H, \cdots, \mathbf{g}_L^H, 0, \cdots, 0]^T$ and the first row $[\mathbf{g}_0^H, 0, \cdots, 0]$ and $\mathbf{h}$ is same as described in (2).

The calculation of (10) will take more computation, therefore, computation of $\mathbf{G}_i$ can be eliminated by transforming it to

$$\mathbf{Q} = \sum_{i=0}^{L-1} \mathbf{G}_i^H \mathbf{G}_i \ . \tag{11}$$

And further calculating it in the quadratic equation $q(\mathbf{h}) = 0$ which is

$$q(\mathbf{h}) = \mathbf{h}^H \mathbf{Q} \mathbf{h} = \mathbf{0}. \tag{12}$$

In the other hand minimization of $q(\mathbf{h})$ can cause $\mathbf{h}$=0 for that reason, minimization should be focused on the selection of proper constraint. Normally natural constraint would be $\|\mathbf{h}\| = 1$ because in the receiver side the received signal power is roughly constant in practice. Therefore estimating $\mathbf{h}$ channel is unit-norm eigenvector related to the smallest eigenvalue of $\mathbf{Q}$ which is

$$\tilde{\mathbf{h}}^H \mathbf{Q} \tilde{\mathbf{h}} = \mathbf{0}. \tag{13}$$

The calculation of blind methods is slow in convergence rate that is why it is better to use a few pilots for getting the channel knowledge and get initial channel estimation to decrease complication of the channel estimation. Assume we have N_{pil} pilots as frequency domain signal in the transmitter which is $\tilde{S} = \left[\tilde{S}_{pil}(1), \cdots, \tilde{S}_{pil}(N_{pil})\right]$. At the receiver, received frequency domain pilot signals are $\tilde{R} = \left[\tilde{R}_{pil}(1), \cdots, \tilde{R}_{pil}(N_{pil})\right]$. Now we can calculate a least square (LS) channel estimation of pilot signal is

$$\tilde{\mathbf{H}}_{LS} = \left[\frac{\tilde{R}_{pil}(1)}{\tilde{S}_{pil}(1)}, \cdots, \frac{\tilde{R}_{pil}(N_{pil})}{\tilde{S}_{pil}(N_{pil})}\right]. \tag{14}$$

Now we can write as $\tilde{\mathbf{H}}_{pil} = \tilde{\mathbf{H}}_{LS}$ and it can be described as the time domain training based channel estimation

$$\tilde{\mathbf{h}}_{pil} = \mathbf{F}_{pil}^H \tilde{\mathbf{H}}_{pil} \tag{15}$$

here $\mathbf{F}_{pil}$ is DFT matrix of the pilot signals. This condition is true when $\mathbf{F}_{pil}$ consists of first $L+1$ columns of $\mathbf{F}_N$. And a number of selected rows matches with the pilot signals number.

Let us bring up the system of the equations as

$$\begin{cases} \tilde{\mathbf{h}}^H \mathbf{Q} \tilde{\mathbf{h}} = 0 \\ \mathbf{F}_{pil} \tilde{\mathbf{h}}_{pil} = \tilde{\mathbf{H}}_{pil} \end{cases}. \tag{16}$$

Meanwhile, the system of equations are calculated in the least square sense, therefore quadratic equation can be minimized as

$$q(\mathbf{h}) = \tilde{\mathbf{h}}^H \mathbf{Q} \tilde{\mathbf{h}} + \| \mathbf{h} - \tilde{\mathbf{h}}_{pil} \|^2 \tag{17}$$

here $\mathbf{h}$ is a true channel with the same size pilot signal. And the difference of pilot symbols alone are not enough for the estimation of the channel, therefore, the channel estimation $\hat{\mathbf{h}}$ can be approximated as

$$\hat{\mathbf{h}} = \left(\mathbf{Q} + \mathbf{I_Q}\right)^{-1} \tilde{\mathbf{h}}_{pil} = \boldsymbol{\alpha} \tilde{\mathbf{h}}_{pil} \tag{18}$$

here $\mathbf{I_Q}$ is a unit matrix which is equivalent to $\mathbf{Q}$ the matrix and $\boldsymbol{\alpha}$ is a relational factor between $\mathbf{h}_{pil}$ and $\hat{\mathbf{h}}$, on the other hand, $\boldsymbol{\alpha}$ can also be interpreted as optimization matrix which is actual channel parameter can be approached by $\mathbf{h}_{pil}$.

B. Optimization in semi-blind channel estimation

To obtain the signal and noise subspaces for semi-blind channel estimation we need the actual true $\mathbf{R_{rr}}$ autocorrelation matrix. In the real practice, the $\mathbf{R_{rr}}$ is estimated by dividing to M blocks by

$$\tilde{\mathbf{R}}_{\mathbf{rr}} = \frac{1}{M} \sum_{m=0}^{M-1} \mathbf{r}_{ZP}(m) \mathbf{r}_{ZP}(m)^H \tag{19}$$

here M is the number of the received signals. The $\tilde{\mathbf{R}}_{\mathbf{rr}}$ closes to $\mathbf{R_{rr}}$ unlimitedly when M inclines to infinite. And the essential condition for performing EVD decomposition the $\tilde{\mathbf{R}}_{\mathbf{rr}}$ must be full-rank. For this reason, we must ensure that there have to be enough received symbols in the receiver. We can also describe (19) in the following way

$$\tilde{\mathbf{R}}_{\mathbf{rr}} = \mathbf{H} \tilde{\mathbf{R}}_{\mathbf{ss}} \mathbf{H}^H + \mathbf{R}_{nn} \tag{20}$$

which is here

$$\tilde{\mathbf{R}}_{\mathbf{ss}} = \frac{1}{N} \sum_{k=0}^{N-1} \mathbf{s}_M(k) \mathbf{s}_M(k)^H \tag{21}$$

Now we perform EVD of $\tilde{\mathbf{R}}_{\mathbf{rr}}$ and we obtain new $\mathbf{Q}$ matrix, which we rewrite as $\tilde{\mathbf{Q}}$. Subsequently, the initial time domain blind channel estimation is an unit-norm eigenvector related to the smallest eigenvalue of $\tilde{\mathbf{Q}}$ which can also be described by $\tilde{\mathbf{h}}_{\mathbf{Q}}$. Thus we can rewrite (13) as follows

$$\tilde{\mathbf{h}}_{\mathbf{Q}}^H \tilde{\mathbf{Q}} \tilde{\mathbf{h}}_{\mathbf{Q}} = \mathbf{0}. \tag{22}$$

Let us also try to choose optimal computation of LS method by applying the Scaled LS (SLS) channel estimation from [11]

$$\tilde{\mathbf{H}}_{\mathrm{SLS}} = \frac{\mathrm{tr}\left\{\mathbf{R}_{\mathrm{HH}}\right\}}{\sigma^2 \mathrm{tr}\left\{\left(\tilde{S}\tilde{S}^{\mathrm{H}}\right)^{-1}\right\} + \mathrm{tr}\left\{\mathbf{R}_{\mathrm{HH}}\right\}} \tilde{\mathbf{H}}_{\mathrm{LS}} \tag{23}$$

here $\mathbf{R}_{\mathrm{HH}}$ is autocorrelation matrix of $\mathbf{H}$, $\tilde{S}$ is N_{pil} pilots frequency domain signal and $\mathrm{tr}\{\cdot\}$ is a trace of a matrix.

The $\mathrm{tr}\left\{\mathbf{R}_{\mathrm{HH}}\right\}$ trace is not a limiting factor than knowing of $\mathbf{R}_{\mathrm{HH}}$, which also less limiting than even knowing of $\mathbf{H}$ itself. In simulation, the condition of knowing $\mathrm{tr}\left\{\mathbf{R}_{\mathrm{HH}}\right\}$ can be replaced by LS estimator which by using the LS-based consistent instead of $\mathbf{H}$ as

$$\mathrm{tr}\left\{\hat{\mathbf{R}}_{\mathrm{HH}}\right\} = \mathrm{tr}\left\{\tilde{\mathbf{H}}_{\mathrm{LS}}^{\mathrm{H}} \tilde{\mathbf{H}}_{\mathrm{LS}}\right\}. \tag{24}$$

Now we can use (24) instead of $\mathrm{tr}\left\{\mathbf{R}_{\mathrm{HH}}\right\}$ in (23). A resulting training based channel estimator become as the LS-SLS estimator $\tilde{\mathbf{H}}_{LS-SLS}$ and we can change to $\tilde{\mathbf{H}}_{pil} = \tilde{\mathbf{H}}_{LS-SLS}$.

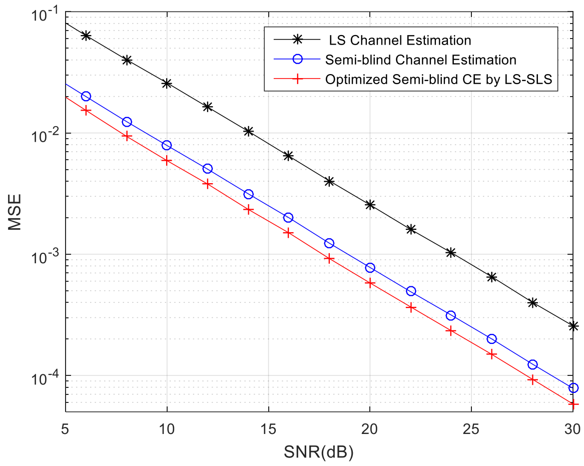

Fig.2 MSE results by using QPSK modulation in OFDM system.

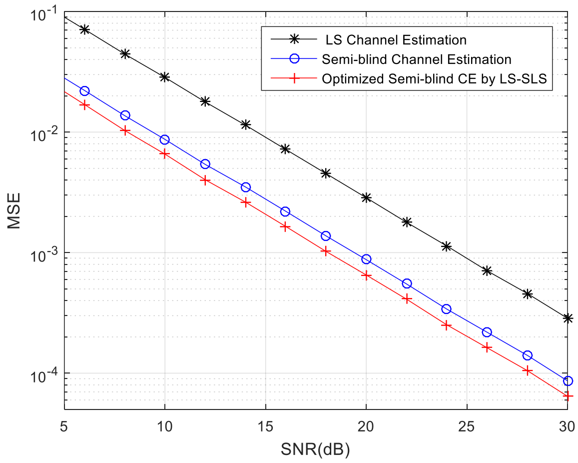

Fig.4 MSE results by using16 QAM modulation in OFDM system.

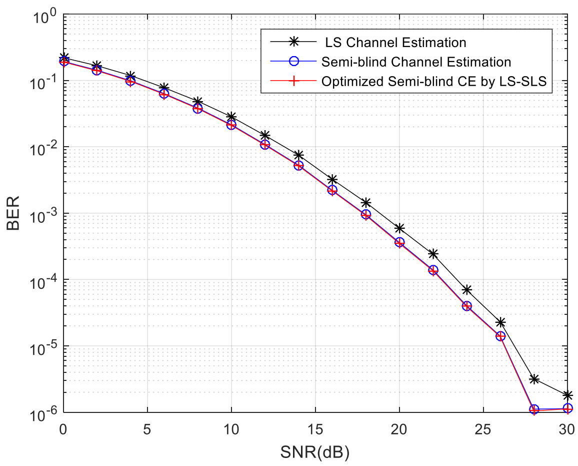

Fig.3 BER vs. SNR results on the basis of QPSK in OFDM system.

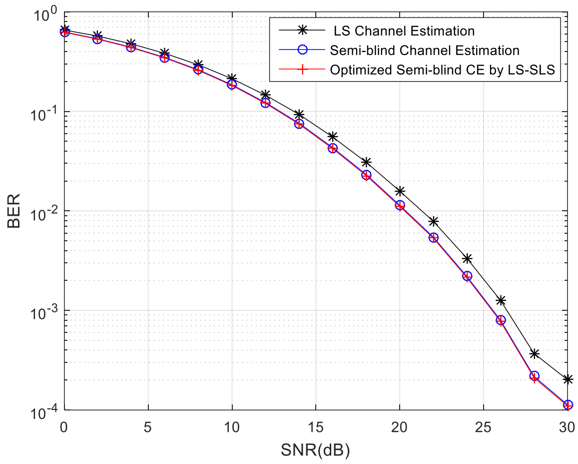

Fig.5 BER vs. SNR results on the basis of 16 QAM in OFDM system.

Correspondingly, the channel information in (18) can be modified as

$$\hat{\mathbf{h}}_{modified} = \left(\tilde{\mathbf{Q}} + \mathbf{I}_{\mathbf{Q}}\right)^{-1} \tilde{\mathbf{h}}_{pil} = \tilde{\boldsymbol{\alpha}}\tilde{\mathbf{h}}_{pil}. \qquad (25)$$

But, here it is difficult to satisfy the equation (22) because the noise channel, the $\tilde{\mathbf{Q}}$ matrix contain the noise power.

For this reason, let error of (22) be as follows

$$e = \tilde{\mathbf{h}}_{\mathbf{Q}}{}^{H} \tilde{\mathbf{Q}} \tilde{\mathbf{h}}_{\mathbf{Q}}. \qquad (26)$$

Henceforth $e \neq 0$, now the orthogonality condition of initial time domain blind channel estimate and noise subspace is damaged due to the interference of noise. We consider the system estimation error e have been initiated by noise. As a result, one favorable way for improving the system estimation performance is removing the estimation error from $\tilde{\mathbf{Q}}$ matrix as follows

$$\tilde{\mathbf{Q}}_{recomposed} = \tilde{\mathbf{Q}} - e\mathbf{I}_{\mathbf{Q}}. \qquad (27)$$

So a new relational factor coefficient defined by $\boldsymbol{\beta}$ such as

$$\boldsymbol{\beta} = \left(\tilde{\mathbf{Q}}_{recomposed} + \mathbf{I}_{\mathbf{Q}}\right)^{-1}. \qquad (28)$$

Corresponding to (25), we can achieve optimized semi-blind channel estimation as follows

$$\hat{\mathbf{h}}_{opt} = \boldsymbol{\beta}\hat{\mathbf{h}}_{pil} \qquad (29)$$

Targeting to achieve an improved channel estimation performance, we can calculate $\tilde{\boldsymbol{\alpha}}$ as in (18) and dip this relational factor to (29), as given

$$\hat{\mathbf{h}}_{opt} = \boldsymbol{\beta}\tilde{\boldsymbol{\alpha}}\tilde{\mathbf{h}}_{pil}. \qquad (30)$$

IV. SIMULATION RESULTS

In the simulation results for improving the performance of the system, we decreased the noise interference and also modified the channel estimate in proposed semi-blind estimation. During simulation process, we used for each OFDM symbol N=64 subcarriers and the ZP with the length of 16. For the channel, we used the Rayleigh multipath fading channel. In the receiver part, we used zero forcing equalizer in corresponding to the result of channel estimation on the transmission channel. And we denoted the channel estimation as (CE) in the figures. Also for getting results in all simulation runs, we averaged the results by over 4000 Monte-Carlo runs. In Fig. 2 QPSK modulation is used to compare the MSE of our proposed semi-blind channel estimation with LS estimation and conventional semi-blind channel estimation. The proposed semi-blind channel estimation is optimized by the LS-SLS channel estimation. We can see from the simulation results that the optimized semi-blind channel estimation outperformed the conventional semi-blind channel estimation by 1.5-2 dB and LS channel estimation by 6-6.5 dB. In Fig. 3 a bit error rate (BER) verses to signal noise ratio (SNR) is calculated and same methods was compared in Fig. 2. From the result, it is appearing that the proposed semi-blind channel estimation has a similar result with conventional semi-blind channel estimation but still both of them have better results than LS channel estimation. However, they performed better on 28 dB than 30 dB. Thus we conclude that semi-blind and proposed semi-blind CEs performing better in lower 28-30 dB in our system on the basis of QPSK modulation.

We also applied the QAM 16 modulation in our simulation for showing our proposed channel estimation objectively. And we have calculated MSE comparison in QAM 16 modulation and the simulation result shows that our proposed semi-blind channel estimation still outperform other channel estimation methods. Also, the BER comparison was calculated in Fig. 5 from the result we can see BER performance of the proposed semi-blind channel estimation has also the same result with conventional semi-blind channel estimation but in higher SNR it is still performing better than the result of Fig. 3. However, it has less BER performance in lower SNR, the reason for that is 16 QAM modulation scheme. Because of 16 QAM, modulation scheme consumes more energy than QPSK modulation. For instance to transmit four bits for each 16 QAM symbol rather than the transmitting the two bits per QPSK symbol.

In more demonstrating purpose of our proposed method, we also used normalized estimation mean square errors (NMSE) as expressed in [12]

$$\mathrm{NMSE} = \min_{\mathrm{t}} \frac{1}{N_w} \sum_{w=1}^{N_w} \frac{\left\| \hat{\mathbf{h}}_w \mathbf{t}^{-1} - \mathbf{h} \right\|_F^2}{\left\| \mathbf{h} \right\|_F^2} \tag{31}$$

where w is a number of iteration, $\|\cdot\|_F$ is the Frobenius norm and $\mathbf{t}$ taken from $\hat{\mathbf{h}} = \mathbf{ht}$ which we called relative coefficient between estimated and actual channel values. Consequently, Fig.6 and Fig. 7 have been calculated in NMSE. And the results show that the proposed method outperforms in NMSE analysis as well. Therefore, we concluded that the proposed method is outperforming in both MSE and NMSE cases.

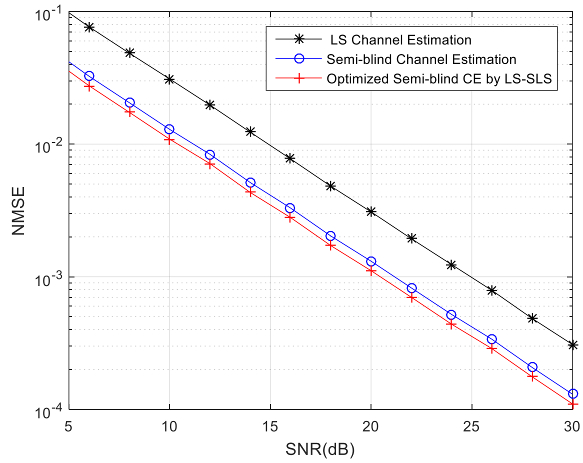

Fig. 6 NMSE of CEs in OFDM system on the basis of QPSK with $N_w = 4000$.

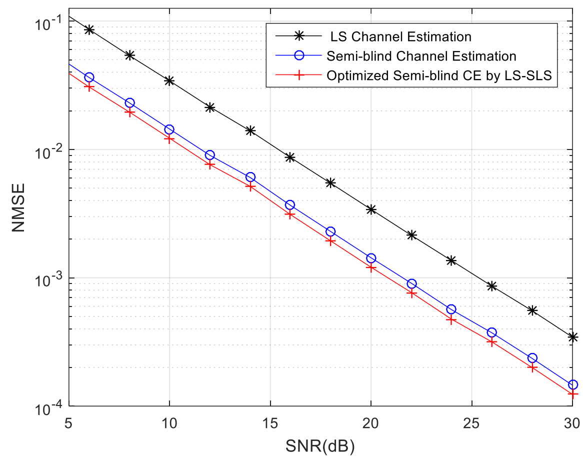

Fig. 7 NMSE of CEs in OFDM system on the basis of 16 QAM with $N_w = 4000$.

V. Conclusion

In this research paper, we simulated a number of estimation methods for SISO ZP-OFDM system. The purpose of using ZP features in OFDM is a simplification of the system in the implementation part. The first condition of the channel is that it estimated up to ambiguity matrix by using subspace decomposition. As the simple sequence of the conventional semi-blind channel estimation, the second condition was estimated ambiguity matrix by pilot signals. In the proposed semi-blind channel estimation method, we eliminated subspace decomposition estimation error which is caused by noise in the channel. Further, we have used training based LS-SLS channel estimation for optimization of proposed semi-blind channel estimation. From the simulation results of MSE and NMSE, we can see that the proposed optimized semi-blind channel estimation method outperforms conventional semi-blind channel estimation. And the proposed optimized semi-blind channel estimation method is applicable to various modulations such as phase shift keying (PSK) and quadrature amplitude modulations (QAM).

Acknowledgment

The authors thank the editors, Prof. Yongqing Fu, Prof. Fang Ye, and special thanks to Dr. Jamshid Sangirov.

References

[1] L. Suyue, J. Xiong, L. Gui, and Y. Xu, "A generalized analytical solution to channel estimation with inter-symbol interference cancelation and co-channel interference cancelation for single input single output/multiple input single output digital terrestrial multimedia broadcasting systems", *IEEE Transactions on Broadcasting*, vol. 59, no. 1, pp. 116-128, 2013.

[2] A. Ghosh, R. Ratasuk, B. Mondal, N. Mangalvedhe, and T. Thomas, "LTE-advanced: next-generation wireless broadband technology", *IEEE Wireless Communication*, vol. 17, no. 3, pp. 10-22, 2010.

[3] E. Dahlman, S. Parkvall and J.Sköld, *4G LTE/LTE-Advanced for Mobile Broadband, Elsevier* , ISBN: 978-0-12-385489-6, 2011

[4] Buchoux V., Cappe, O., Moulines É., and Gorokhov A., "On the performance of semi-blind subspace-based channel estimation", *IEEE Transactions on Signal Processing*, 48(6), 1750-1759, 2000.

[5] Zeng Y., Lam W. H., and Ng. T. S., "Semiblind channel estimation and equalization for MIMO space-time coded OFDM", *IEEE Transactions on Circuits and Systems I*, vol. 53, no. 2, pp. 463-474, 2006.

[6] Yu J. L., and Lin, Y. C. "Space–Time-Coded MIMO ZP-OFDM Systems: Semiblind Channel Estimation and Equalization", *IEEE Transactions on Circuits and Systems I*, 56(7), pp. 1360-1372, 2009.

[7] Wu W. R., Chiueh R. C., and Tseng, F. S., "Channel estimation for OFDM systems with subspace pursuit algorithm", *IEEE International Conference on Green Circuits and Systems (ICGCS)*, pp. 269-272, 2010.

[8] Fang S. H., Chen J. Y., Shieh M. D., and Lin, J. S. "Modified Subspace Based Channel Estimation Algorithm for OFDM Systems", *IEEE 69th Conference on Vehicular Technology*, pp. 1-5, 2009.

[9] J. Lang Yu, D.-Y. Hong, "Subspace channel estimation assisted block matrix scheme for ZP-OFDM systems", *6th International Conference on Wireless and Mobile Communications (ICWMC)*, pp.16 – 20, 2010.

[10] Huo W., Wang Z., and Li S., "A simple subspace-based semi-blind channel estimator for precoded OFDM systems", *International Conference on Wireless Communications, Networking and Mobile Computing (WiCom)*, pp. 41-44, 2007.
[11] Biguesh Mehrzad and Alex B. Gershman, "Training-based MIMO channel estimation: a study of estimator tradeoffs and optimal training signals", *IEEE Transactions on Signal Processing* , vol. 54, no. 3, pp. 884-893, 2006.
[12] Gao Feifei and A. Nallanathan, "Subspace-based blind channel estimation for SISO, MISO, and MIMO OFDM systems", *In IEEE International Conference on. Communications (ICC'06)*, p. 3025-3030, 2006.

Gulomjon Sangirov received the M.S. degrees in Information and Communication Engineering from Harbin Engineering University in 2012. And he is studying Ph.D. in College of Information and Communication Engineering in Harbin Engineering University, China currently. His research interests are Channel Estimation in MIMO-OFDM and Quasi-Cyclic-LDPC coding.

Yongqing Fu received the MS degree in electrical engineering from National University of Defense Technology, Changsha, China in 1985. He held the posts of Lecturer (1988-1995), Associate professor (1995-2000) and Professor (since 2000) in the Electrical Engineering Department, Harbin Engineering University. He was a visiting Professor of Electrical and Computer Engineering Department, University of Manitoba, Canada in 2002 and also Electrical and Computer Engineering Department, University of California, San Diego, the USA in 2009. He is the author of four books, more than 100 articles, and more than 10 patents. His research interests include the weak signal detection, chaotic communication and signal processing, cognitive radio system, image coding, marching and feature extraction and circuit and electronic system. Prof. Fu is the Vice director of the Northeast Electric Theory Association (since 2003), and he is a reviewer for the Journal of China Universities of Posts and Telecommunications (since 2006).

Jamshid Sangirov received the M.S. degrees in Information and Communication Engineering from Yeungnam University in 2006. He got his Ph.D. in Information and Communication Engineering at Korean Advanced Institute of Science and Technology (KAIST) at 2013. He worked with the RFIC Design Team, at Teltron Inc., Korea, from 2010 to 2011. From 2013 he is with Samsung Electronics. His research interests are analog/RF/VLSI and high-speed Integrated Circuit design.

Fang Ye received her Ph.D. degree in Communication and Information System from Harbin Engineering University in 2006. From November 2007 to November 2008, she has been as a visiting scholar in the school of electronics and computer science, University Of Southampton, U.K. Since December 2008, she is an associate professor of the school of Information and Communication Engineering in Harbin Engineering University, China. Her research interests include LTE technologies, adaptive radio resource allocation technology, and UWB signal processing. She is a member of IEEE and Association for Computing Machinery (ACM).

Ahmad Olmasov received the M.S. degrees in Tashkent University of Information Technologies, Tashkent, Uzbekistan in 2007. And he started working with Samarkand branch of Tashkent University of Information Technologies. Currently, he is a lecturer in the department of telecommunication engineering at Samarkand branch of Tashkent University of Information Technologies. His research interests are Mobile communication, channel coding in high throughput wireless communication and MIMO-OFDM.