SEMI-BLIND OFDM CHANNEL ESTIMATION USING RECEIVER DIVERSITY IN THE PRESENCE OF VIRTUAL CARRIERS

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ABSTRACT

In this paper, a semi-blind channel estimator using receiver diversity is proposed for OFDM systems in the presence of virtual carriers over a time-dispersive channel. We insert a pilot to each data block to resolve the inherent scalar ambiguity of blind identification. The computation complexity of this algorithm is also studied and reduced. By simulation, this algorithm supplies accurate estimation with small expense.

1. INTRODUCTION

OFDM is a promising candidate for B3G physical layer technique due to its high efficiency and robustness to frequency selective fading [1-3]. It is the basis of the European standard for digital audio/video broadcasting [4] (e.g. DAB and DVB-T, target rates 1.7 and 20 Mb/s, respectively) and is being developed internationally for use in high-speed wireless LANs (e.g. IEEE 802.11a [5], target rates 6-54Mb/s) and wireless-local-loop applications (1-10 Mb/s).

To facilitate coherent detection, channel state information (CSI) is required which is difficult for time-dispersive channel. In practice, CSI is obtained through channel estimation. There are several conventional methods, training and blind estimation. When training symbols are used, it leads to extra expense, which means a loss of data rate. Contrarily, blind estimation [6-8] does not require training symbols, so it is possible to achieve higher data rates. Many existing blind estimators use the channel high order statistical features, so they need a large number of OFDM symbols. Clearly, these methods will have poor performance when the channel is not constant, thus, have limited applicability in time-varying channel. In [9] authors proposed a blind identification exploiting receiver diversity which can get CSI during one OFDM symbol, but it was supposed to preknow a certain path CSI to resolve the scalar ambiguity, which is unavailable in practice. In this paper, we adopt a semi-blind estimation which uses only one training symbol to resolve the scalar ambiguity at the cost of reasonable performance degradation. This method has a significant reduction in the number of training symbols. We also study the modification of this method in the presence of virtual carriers.

The remaining of this paper is organized as follows. Section 2 gives the system model. Section 3 describes how to estimate the multipath channel through our method and how to resolve the scalar ambiguity by introducing one training symbol. Simulations under different assumptions are provided in section 4. We give our conclusion in section 5.

2. SYSTEM MODEL

2.1. Transmitter

We consider an OFDM system with one transmit antenna and two receive antennas as depicted in Fig.1. There are totally N subcarriers where only P subcarriers carry the data symbols, thus, the number of virtual subcarriers is $N-P$. Assume the subcarriers numbered p_0 to p_0+P-1 are used for data, where p_0 is the index of the first data carrier. Also assume that the length of cyclic prefix (CP) is N_{CP}. Firstly, we consider every data block has only one OFDM symbol. After converted from serial to parallel, the frequency data symbols in k th data block can be written as $\mathbf{s}(k)=[s_0(k), s_1(k),\ldots,s_{P-1}(k)]^T$. We define the normalized FFT matrix as

$$\mathbf{F}=\frac{1}{\sqrt{N}}\begin{pmatrix} W_N^{00} & W_N^{01} & \cdots & W_N^{0(N-1)} \\ W_N^{10} & W_N^{11} & \cdots & W_N^{1(N-1)} \\ \vdots & \vdots & \ddots & \vdots \\ W_N^{(N-1)0} & W_N^{(N-1)1} & \cdots & W_N^{(N-1)(N-1)} \end{pmatrix}, \tag{1}$$

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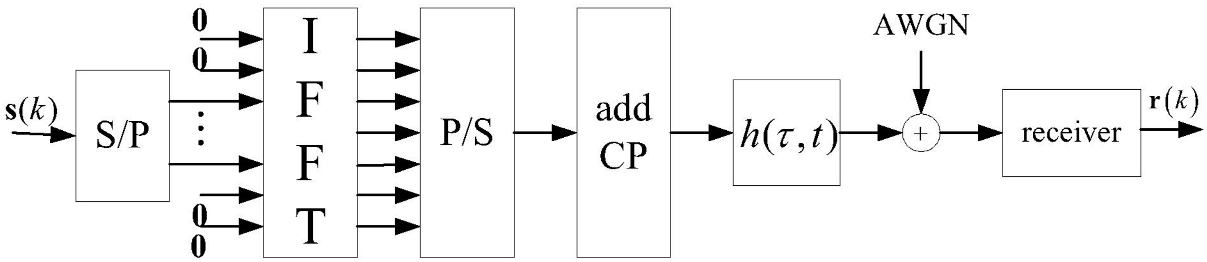

Figure 1. OFDM system model

where $W_N^{mn} = \exp\left(-j2\pi mn/N\right)$, thus IFFT matrix is $\mathbf{F}^H$, Θ^H denotes the complex conjugate transpose. We define $\mathbf{u}_i = {}^{1}\!/\!_{\sqrt{N}}[W_N^{i0} \quad W_N^{i1} \quad \cdots \quad W_N^{i(N-1)}]^T \ i = 0,1,\cdots,N-1$, denotes the i th column of $\mathbf{F}$. After IFFT, the time-domain data signal can be represented as

$$\mathbf{x}(k) = \mathbf{F}^H \overline{\mathbf{s}}(k), \tag{2}$$

where $\mathbf{x}(k) = [x_0(k),\ x_1(k), \ldots, x_{N-1}(k)]^T$, and $\overline{\mathbf{s}}(k) = \left[\mathbf{0}_{p_0\times 1} \quad \mathbf{s}(k)^T \quad \mathbf{0}_{(N-P-p_0)\times 1}\right]^T$. The value of the virtual subcarriers is zero and the index depends on p_0.

2.2 Channel Model

We assume the channel impulse responses at each receiver are $h_1(\tau,t)$ and $h_2(\tau,t)$, respectively. They can be written as

$$h_i(\tau,t) = \sum_{l=0}^{L-1} \gamma_i\left(\tau,t\right)\delta\left(\tau - \tau_{i,l}\right), \quad i=1,2 \tag{3}$$

where $\gamma_i\left(l,t\right)$ is the l th path complex gain of i th receiver channel. Usually we consider $\gamma_i\left(l,t\right)$ is not time-varying during several OFDM symbols, thus can omit the variable t. L is the path number and $\tau_{i,l}$ is path delay. As the receiver antennas are co-located, the assumption $\tau_{1,l} = \tau_{2,l}$ is reasonable.

We consider the channel is sample-spaced, thus $\tau_{i,l} * N\Delta f$ is integer, where $\Delta f = {}^{1}\!/\!_{T_u}$ is the subcarrier space. The channel can be rewritten as

$$\mathbf{h}_i = \left[h_i(0),\ h_i(1), \ldots, h_i(L-1)\right]^T, \quad i = 1,2 \tag{4}$$

the corresponding channel impulse response in frequency domain is

$$\mathbf{H}_i = \mathbf{F}\begin{bmatrix} \mathbf{h}_i \\ \mathbf{0}_{(N-L)\times 1} \end{bmatrix}, \quad i = 1,2\,. \tag{5}$$

Define $\mathbf{W}_L = \left(\mathbf{u}_0 \quad \mathbf{u}_1 \quad \cdots \quad \mathbf{u}_{L-1}\right)$, then equation (5) can be rewritten as

$$\mathbf{H}_i = \mathbf{W}_L \mathbf{h}_i, \quad i = 1,2\,. \tag{6}$$

2.3 Receiver

After removing CP and performing FFT, each receiver antenna signal can be represented as

$$\mathbf{r}_i(k) = \overline{\mathbf{S}}(k)\mathbf{H}_i + \mathbf{n}_i = \overline{\mathbf{S}}(k)\mathbf{W}_L\mathbf{h}_i + \mathbf{n}_i, \quad i = 1,2 \tag{7}$$

where $\mathbf{r}_i(k)$ is the $N\times 1$ dimensional received signal, $\overline{\mathbf{S}}(k) = diag(\overline{\mathbf{s}}(k))$ is the $N\times N$ diagonal matrix with diagonal element $\overline{\mathbf{s}}(k)$. $\mathbf{n}_i$ is additive white complex Gaussian noise and is uncorrelated with each other. Thus, we have $2N$ observations($\mathbf{r}_1 \quad \mathbf{r}_2$) and $2L+N$ unknowns. In practice, $N > 2L$ is always satisfied, thus the number of observations is greater than that of unknown variables.

3. SEMI-BLIND ESTIMATION

Omitting the variable k, equation (7) can be written as

$$\mathbf{r}_i = \overline{\mathbf{S}}\mathbf{W}_L\mathbf{h}_i + \mathbf{n}_i, \quad i = 1,2\,, \tag{8}$$

Without considering noise, equation (8) can be rewritten as

$$\mathbf{r}_i = \overline{\mathbf{S}}\mathbf{W}_L\mathbf{h}_i, \quad i = 1,2\,, \tag{9}$$

Considering equation (9) has the same transmitted data matrix $\overline{\mathbf{S}}$, after some calculation, we can get

$$\mathbf{R}_1\mathbf{W}_L\mathbf{h}_2 = \mathbf{R}_2\mathbf{W}_L\mathbf{h}_1, \tag{10}$$

where $\mathbf{R}_1 = diag(\mathbf{r}_1)$ and $\mathbf{R}_2 = diag(\mathbf{r}_2)$ are diagonal matrices constructed by $\mathbf{r}_1$ and $\mathbf{r}_2$, respectively. Equation (10) can be further expressed as

$$\left[\mathbf{R}_2\mathbf{W}_L \quad -\mathbf{R}_1\mathbf{W}_L\right]\begin{bmatrix}\mathbf{h}_1 \\ \mathbf{h}_2\end{bmatrix} = \mathbf{V}\mathbf{h} = \mathbf{0}, \tag{11}$$

from equation (11), we can retrieve the channel coefficients up to a scalar ambiguity by simply finding a solution to the above homogeneous equations.

In the presence of noise, equation (11) will not hold. We can retrieve the channel by minimizing the cost function

$$\min_{\mathbf{h}} \ \mathbf{h}^H\mathbf{V}^H\mathbf{V}\mathbf{h} \qquad s.t. \ \left|\mathbf{h}\right| = 1, \tag{12}$$

Define $\mathbf{U} = \mathbf{V}^H\mathbf{V}$, then perform singular value decomposition on $\mathbf{U}$, the eigenvectors corresponding to the smallest eigenvalue of $\mathbf{U}$ is the solution of the channel.

In the above deducing process, the effect of virtual carriers on the algorithm performance is not considered. In equation (9), $\bar{\mathbf{S}} = diag(\bar{\mathbf{s}}) = diag(\begin{bmatrix} \mathbf{0}_{p_0\times 1} & \mathbf{s}(k) & \mathbf{0}_{(N-P-p_0)\times 1} \end{bmatrix}^T)$, the values corresponding to the virtual carriers of the receive signal $\mathbf{r}_1$ and $\mathbf{r}_2$ are zeros. Define $\bar{\mathbf{W}}_L = \mathbf{W}_L(p_0+1 : p_0+P,:)$,which consists of the p_0+1 th row to the p_0+P th row, then equation (9) can be rewritten as

$$\bar{\mathbf{r}}_i = \mathbf{S}\bar{\mathbf{W}}_L\mathbf{h}_i, \quad i = 1,2, \tag{13}$$

where $\mathbf{S} = diag(\mathbf{s}(k))$ is the diagonal matrix consisting of the elements of $\mathbf{s}(k)$, $\bar{\mathbf{r}}_1$ and $\bar{\mathbf{r}}_2$ are frequency domain receive signals corresponding to the data modulated subcarriers. After updating equation (11) with the modified denotations, we can get

$$\min_{\mathbf{h}} \ \mathbf{h}^H\bar{\mathbf{V}}^H\bar{\mathbf{V}}\mathbf{h} \quad s.t. \quad |\mathbf{h}| = 1, \tag{14}$$

Similarly, define $\bar{\mathbf{U}} = \bar{\mathbf{V}}^H\bar{\mathbf{V}}$, then perform singular value decomposition on $\bar{\mathbf{U}}$,the eigenvector corresponding to the smallest eigenvalue of $\bar{\mathbf{U}}$ is the solution of the channel.

Let us compare the equation (12) with (14), we can find the size of $\mathbf{V}$ is $N\times 2L$ and that of $\bar{\mathbf{V}}$ $P\times 2L$, so the computational complexity of calculating $\bar{\mathbf{U}} = \bar{\mathbf{V}}^H\bar{\mathbf{V}}$ is lower than that of calculating $\mathbf{U} = \mathbf{V}^H\mathbf{V}$.

Next, when the channel is quasi-stationary, we can combine several OFDM symbols to improve the performance of the algorithm. Equation (14) can be extended to the following

$$\min_{\mathbf{h}} \ \mathbf{h}^H[\sum_{m=1}^{M}\bar{\mathbf{V}}_m{}^H\bar{\mathbf{V}}_m]\mathbf{h} \quad s.t. \quad |\mathbf{h}| = 1, \tag{15}$$

then the channel can be obtained as the above solution.

The scalar ambiguity can not be avoided in singular value decomposition. We insert one pilot to each data block to resolve this problem. Define

$$\mathbf{h}_i = \beta\hat{\mathbf{h}}_i, \quad i = 1,2, \tag{16}$$

where β is the scalar to be determined. Substitute equation (16) into (13), we obtain

$$\bar{\mathbf{r}}_i = \mathbf{S}\bar{\mathbf{W}}_L\beta\hat{\mathbf{h}}_i, \quad i = 1,2. \tag{17}$$

The pilot s_p is inserted into the first data subcarrier, then the receive signal corresponding to the pilot subcarrier is r_{ip} .

Define $\bar{\mathbf{w}}_L^i$ ($i = 1,2,\ldots,P$) represents the i th row of $\bar{\mathbf{W}}_L$, then we obtain

$$r_{ip} = s_p\bar{\mathbf{w}}_L^1\beta\hat{\mathbf{h}}_i, \quad i = 1,2 \tag{18}$$

After some calculation, the scalar can be expressed as

$$\beta = \frac{r_{ip}}{s_p\bar{\mathbf{w}}_L^1\hat{\mathbf{h}}_i}, \quad i = 1,2 \tag{19}$$

Substitute equation (19) into (16), we can get the channel impulse response in time domain without scalar ambiguity. To improve the performance of this algorithm, we can use more pilots to determine the scalar β .

4. PERFORMANCE EVALUATION

In this section, we compare the MSE performance with different assumptions through Monte Carlo simulations. The total number of paths is 3, channel bandwidth is 200kHz with Doppler frequency 10Hz, so the channel can be regarded as quasi-stationary. The spectrum is divided into 16 subcarriers with 4 of them virtual carriers. The guard interval length is 4. In one simulation, the number of data block is N_{Blk} , so the MSE can be written as

$$MSE = \frac{1}{N_{Blk} * 2L}\sum_{k=1}^{N_{Blk}}\left\|\mathbf{h}(k) - \hat{\mathbf{h}}(k)\right\|^2, \tag{20}$$

In Fig.2, we compare the performance of equation (12) with (14), all using one OFDM symbols in one data block with 4 virtual carrier number. Each method has two MSE performances, one is using a pilot to resolve the scalar ambiguity which we call “pilot-based”, the other is assuming that $h_1(0)$ is known which we call “ideal”. In the legend of Fig.2, “pilot-based ” and “ideal” represent the two MSE performances of equation (12), “improve pilot-based” and “improve ideal” represent the two MSE performances of equation (14). In low SNR, equation (12) has better MSE performance than equation (14) for the former provides more channel information than the latter, and in high SNR, they have almost the same MSE performance as the AWGN effect becomes weak. “ideal” has better performance than “pilot-based” as the former utilizes one exact path state information of the channel to resolve the scalar.

In Fig.3, we compare the effects of different numbers of virtual carriers on the MSE performance of equation (14), all using one pilot to resolve the scalar ambiguity and one OFDM symbol in a data block. The MSE performance becomes deteriorated as the number of virtual carriers increases, which can be interpreted the more virtual carriers present, the less channel information included in equation (14).

In Fig.4, we compare the MSE performances of equation (14) using different OFDM symbols in one data block, all using one pilot to resolve the scalar ambiguity and 4 virtual carriers. The more symbols are combined, the average effect of AWGN diminishes, as a result, the MSE performance is enhanced.

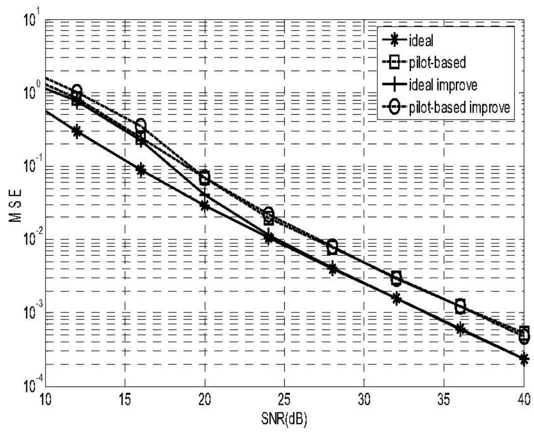

Figure 2. MSE performances of algorithm (12) and (14) with different way resolving scalar

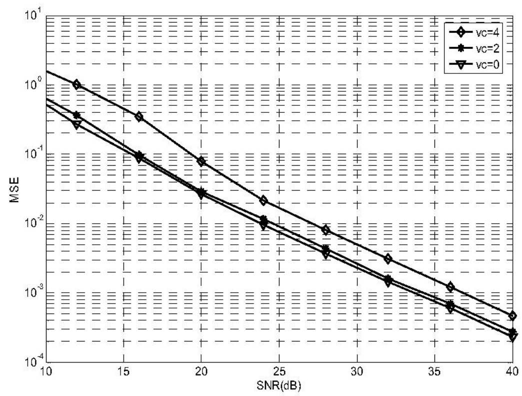

Figure 3. MSE performances of different virtual carrier numbers

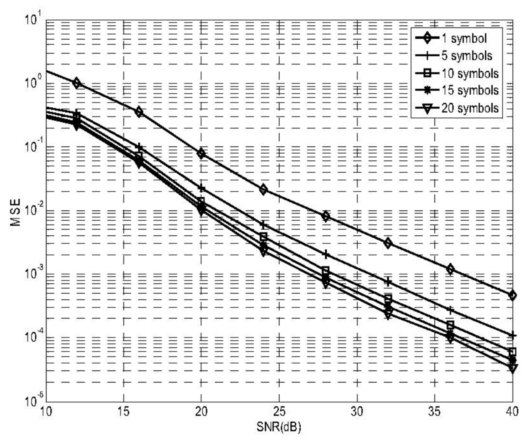

Figure 4. MSE performances of different data symbol numbers in one data block

5. CONCLUSIONS

In this paper, a semi-blind channel estimation using receiver diversity is proposed for OFDM systems in the presence of virtual carriers. The proposed algorithm utilizes frequency domain observations to estimate the channel impulse response and recurs to one pilot to resolve the scalar ambiguity. In the presence of virtual carriers, this algorithm can reduce the computation complexity with tiny loss of MSE performance. The different MSE performances under different assumptions are provided with numerical simulation.

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