·学术论文与技术报告·

Blind Channel Estimation Using Aid of Pseudo-Pilot-Symbols for OFDM System

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Abstract In this paper, a blind channel estimator based on the aid of Pseudo-Pilot-Symbols(PPS) has been proposed relying on the character of OFDM system. Unlike pilots which are used for estimation of channel and must waste some useful bandwidth, the PPSs, whose power can be boosted 3 dB or 6 dB to suppress efficiently the additive noise, are useful data transmitted, thus efficiency of the OFDM system has been increased. Performance simulation of the proposed estimator including Mean Squares Error (MSE) of channel and uncoded Bit Error Rate(BER) have been taken, and the results show the estimator is efficient.

Key words blind channel estimator; OFDM system; pseudo-pilot-symbols; performance simulation

在OFDM系统中用伪导频符号进行盲信道估计

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【摘要】根据OFDM系统的特性提出了用伪导频符号进行信道的盲估计算法。与用于做信道估计的导频信号不同，PPSs传输的是有用的数据，因此提高了系统的带宽利用率，若PPSs的平均功率增加3 dB或6 dB时可有效抑制信道的加性高斯噪声。该文对信道盲估计的均方误差及其由此算法获得的信道的状态信息对无编码的OFDM系统进行解调的误比特概率进行了仿真，结果表明提出的算法是有效的并具有很好的灵活性。

关 键 词 盲信道估计; OFDM系统; 伪导频符号; 性能仿真

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Because of great promise for high rate transmissions, orthogonal frequency division multiplexing (OFDM) has been and is being widely considered by many standards, including local area mobile wireless broad-band network IEEE802.11 and HIPERLAN/2[1]、digital audio and video broadcasting(DAB and DVB) in Europe[2]. In order to avoid intersymbol interference(ISI) caused by channel with memory, which is often modeled as FIR filter, OFDM systems first take the inverse Fourier transform(IFFT) of data symbols and then insert redundancy in the form of a cyclic prefix(CP) of length larger than the FIR channel order. CP is discarded at the receiver and the remaining part of the OFDM symbol is FFT processed. A combination of IFFT and CP at transmitter with the FFT at the receiver converts the

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frequency-selective channel to separate flat-fading subchannels so that simply frequency equalizer can been used by dividing the FFT output by corresponding channel frequency response. In order to get high rate and performance in coherent OFDM systems in which signal may come from large-signal constellation, it is key that accurate estimation of channel must be obtained. Channel can often be estimated by training sequences or aid of pilots, but they waste rare bandwidth and reduce the efficiency of systems. Because of saving bandwidth and being capable of tracking slow channel variations, blind channel estimation and equalization methods are well motivated as they avoid use of training sequence and pilot. A number of blind channel estimator have been developed for OFDM. Some of them are based on input cyclostationarity[3], while the others rely on subspace decompositions or sub-matrix of autocorrelation matrix in which the channel identifiability is not guaranteed for all FIR channels when some of them has nulls (deep fades) close to or on the FFT grid[4, 5], reliable detection of the symbols carried by this faded subcarriers becomes impossible.

In this paper, we proposed and analyze a novel blind channel estimation method based on the aid of Pseudo-Pilot-Symbols (PPS) and the finite alphabet property of information-bearing symbols using the statistic quantity. Impulse response of channel can be estimated using PPSs which are available data transmitted by transmitter and are regard as pilots, and whose average power is boosted 3 dB or 6 dB to suppress the additive noise.

The rest of this paper is organized as follows. In section 2 we briefly introduce OFDM systems with multiple path channel. We then present some methods obtaining the impulse responses of channel from the frequency responses of PPSs without scalar ambiguity in section 3. Next, in section 4, we introduce the proposed blind channel estimator and its principles. Then, we demonstrate the performance of the novel blind channel estimator by computer simulation in section 5. The conclusions of this paper are presented in the last section.

1 OFDM System

For a OFDM system with N sub-channels or sub-carriers as in Fig.1 we employ following assumptions: 1) The maximum delay t_{max} of the multipath channel is shorter than cyclic prefix $GT = T_G$, where T is sampling interval; 2) A transmitter and a receiver are perfectly synchronized; 3) Variation of the channel is slow enough to be considered constant for I OFDM symbols period.

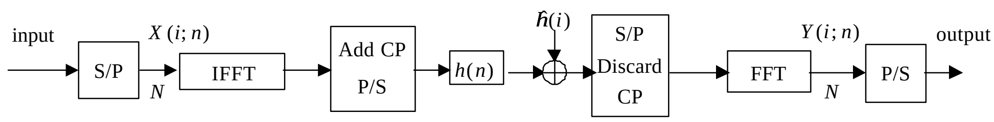

Fig.1 the scheme of OFDM system

As shown in Fig. 1, the symbols to be transmitted are fed into a serial-to- parellel converter and then IFFT is performed, the samples from the IFFT are then converted to a serial form, extended by a cyclic prefix, mixed to an appropriated frequency, and transmitted over the radio channel. Though proper processing, OFDM system can be modeled equivalently parallel flat-fading and independent subchannels shown in Fig. 2 and written in scalar form

$$Y(i;n) = X(i;n)H(r_n) + \tilde{\hbar}(i,n) \qquad n \in [0, N-1] \tag{1}$$

where $Y(i;n)$ is received data of nth subcarrier in ith block and $X(i;n)$ is transmitted data of nth subcarrier in ith block, $\tilde{\hbar}(i) = \mathrm{FFT}(h(i))$ is alse zero-mean and $E\{|\tilde{\hbar}(i)|^2\} = s^2$ additive white Gaussian independent random noise. transfer function of channel is $H(r_n)=\sum_{k=0}^{g-1}h(k)\mathrm{e}^{-\mathrm{j}2\pi kn/N}$ When its impulse response of channel is $\bar{\boldsymbol{h}}=[h(0),h(1),\cdots,h(g-1)]$.

2　Methods of Estimation of Impulse Response of Channel

For every OFDM subchannel (1) hold true. According to assumption 1), impulse response of channel in time domain can be obtained by solving the g independent equations which are gotten using g PPS subchannels. A special case is that the PPSs are comb tones partitioning as shown in Fig. 3,where N_F is interval of subcarrier. In order to suppress the noise , the average power of G PPSs $(G\geq g)$ are boosted 3 dB or 6 dB.

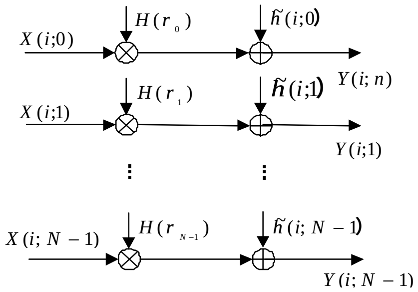

Fig.2　Equivalent model of OFDM system

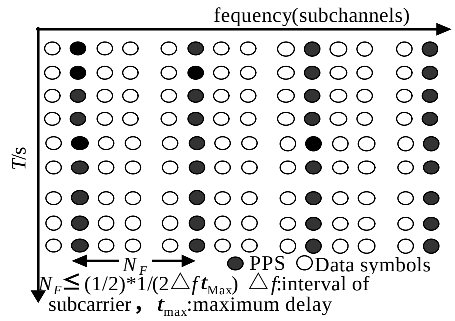

Fig.3　Comb tones partitioning PPS

In order to express in term of matrix, assumption that $H(r_{t_i}),(i=0,1,\cdots,G-1)$　$t_i\in\{0,1,\cdots,N-1\}$ is the PPS subchannels' frequency response without scalar ambiguity, first define the matrix V_G with $G\times g$ entries to be a scaled version of the first g columns corresponding $t_i(i=0,1,\cdots,G-1)$ rows of FFT matrix F and its Matlab's notation as follows: $\boldsymbol{V}_G=\sqrt{N}F_N(t_i,1:g),(i=0,1,\cdots,G-1)$, and the matrix $\hat{\boldsymbol{H}}=[H(r_{t_0}),H(r_{t_1}),\cdots,H(r_{t_{G-1}})]^T$,where T is transpose, and $\hat{\boldsymbol{h}}=[\hat{h}(0),\hat{h}(1),\cdots,\hat{h}(G-1)]^T$ being the estimation of $\bar{h}$ that is impulse responses of channel. We can obtain

$$\hat{\boldsymbol{h}}=\boldsymbol{V}_G^{+}\hat{\boldsymbol{H}} \tag{2}$$

where the + denotes matrix pseudo-inverse.

3　Blind Channel Identifiability and Estimation

We will start with the identifiability issues from the noiseless version of (4): $Y(i;n)=X(i;n)H(r_n)$ where noise is omitted because the basic feasibility question is considered first. Our results will be developed under the following conditions for information-bearing symbols that hold ture in the OFDM system: 1) Symbols are drawn from a finite alphabet set of size Q, i.e., $X(i;n)\in\{x_p\}_{p=1}^{Q}$;2) Symbols with zero-mean and unit covariance transmitted by single subchannel are equiprobable and from a constellation, i.e., $Pr(X(i;n)=x_p)=1/Q,\ p=1,2,\cdots,Q$; 3) Noise $\tilde{n}(i,n)$ is zero-mean complex circular Gaussian and independent of information symbols. From assumption a1), formula $\prod_{p=1}^{Q}(X(i;n)-x_P)=0$ hold true. Expanding the product yields a Qth-order polynomial in $X(i,n)$

$$X^Q(i;n)+b_1X^{Q-1}(i;n)+\cdots+b_Q=0 \tag{3}$$

where $b_1,b_2\cdots,b_Q$ are determined by the constellation points $\{x_p\}_{p=1}^{Q}$. The coefficients $b_1,b_2\cdots,b_Q$ can be zeros if and only if $X(i,n)\equiv 0$ as proved by (3). Therefore from the coefficient set $\{b_p\}_{p=1}^{Q}$ we can always find nonzero elements b_U where $b_U\neq 0;b_k=0,\forall k<U$ hold true. It is not difficult to

evidence that $U \leq Q$ is always true, for examples for BPSK U=Q=2, and (3) can be rewrite as $X^2(i;n)+1=0$, where $b_U = b_2 = 1$; for QAM with Q signaling points e.g. Q=16, 32,64, 128,256 we can easily verify that U=4 according to the same way mentioned above. For another example, let us consider 16QAM with signaling points $x_p \in \{(\pm a \pm bj)/\sqrt{10}\}$ where $a, b \in \{1,3\}$. From (3) $b_4 = 2.72 \neq 0$ can be gotten. $U << Q$ is the fact for a large-signal constellation. Under a1) and 2) that $E\{X^U(i;n)\} = -Ub_U/Q$ hold true has been proved in for any constellation[6], where E(・) is expectation operator, for examples $E\{X^U(i;n)\} = -Ub_U/Q = -0.68$ for 16QAM constellation and $E\{X^U(i;n)\} = -Ub_U/Q = -1$ for BPSK constellation. Starting from $E\{Y^U(i,n)\} = H^U(r_n)E\{X^U(i,n)\}$ we deduce that

$$H^U(r_n) = \frac{-Q}{Ub_u} E\{Y^U(i,n)\} \qquad \forall n \in [0, N] \tag{4a}$$

For each n, the right-hand side of (4a) is available from the FFT processed data statistical averages (across i). In practice, $E\{Y^U(i,n)\}$ is replaced by consistent sample averages and thus $H^U(r_n)$ is estimated as

$$H^U(r_n) = \frac{-Q}{IUb_u} \sum_{i=0}^{I-1} \{Y^U(i,n)\} \qquad \forall n \in [0, N] \tag{4b}$$

where I is the total number of blocks averaged. At the same time, under a2), because of $E\{|X(i;n)|^2\} = 1$, $E\{Y(i,n)Y^*(i,m)\} = H(r_n)H^*(r_n)E\{X(i,n)X^*(i,n)\}$ hold true, too. The following formula

$$\overline{H}(n) = H(r_n)H^*(r_n) = E\{Y(i,n)Y^*(i,n)\} \qquad \forall n \in [0, N] \tag{5a}$$

can be deduced, where $()^*$ is conjugate operator. As $E\{Y(i,n)Y^*(i,m)\}$ may be also replaced by consistent sample averages in practice and thus

$$\overline{H}(n) = (1/I)\sum_{i=0}^{I-1} \{Y(i,n)Y^*(i,n)\} \qquad \forall n \in [0, N] \tag{5b}$$

can be estimated. Hence, $H^U(r_n)$ and $H(r_n)H^*(r_n)$ can be determined and the issue is how one uniquely recovers $H(r_n)$ from them. To solve this problem we first define vector $\vec{a} = [a_0, a_1, \cdots, a_{g-1}]^T$ and

$$\vec{a} = \begin{bmatrix} h(0) & h(1) & \cdots & h(g-1) \\ h(1) & h(2) & \cdot^{\cdot^{\cdot}} & 0 \\ \vdots & \vdots & \cdot^{\cdot^{\cdot}} & \vdots \\ h(g-1) & 0 & 0 & 0 \end{bmatrix} \begin{bmatrix} h^*(0) \\ h^*(1) \\ \vdots \\ h^*(g-1) \end{bmatrix} \tag{6}$$

According the relationship between the time domain and frequency domain, the formula $\overline{H} := H(r_n)H^*(r_n)\big|_{z=r_n} = (a_0 + a_1 z^{-1} + \cdots + a_{g-1} z^{-(g-1)} + a_{g-1}^* z^{-(N-g)} + \cdots + a_1^* z^{-(N-1)})\big|_{z=r_n}$ is true. Therefore, with matrix V_M that is a scaled version of the first g columns of FFT matrix, $\vec{a}$ can be obtained through $\vec{a} = V_M^+ \overline{H}$. When vector $\hat{H}$ which is made up of the frequency response of the g subchannels used as PPSs are known, the impulse response of channel can be computed by using the formula $h = V_G^+ \hat{H}$, and another estimation of $\vec{a}$ denoted as $\hat{a}$ can be gotten in term of formula (6). The fact that $\hat{a} = \vec{a}$ implies that $H(r_n)$ is uniquely identifiable up to a scalar ambiguity even there is noise. Now the proposed estimator are summarized in the following:

Step 1 as the average power of PPSs are boosted, and assumption that $E[|X_{PPS}(i,n)|^2] = P$, $\tilde{H}^U(r_{t_i}), (i = 0,1,\cdots,G-1)$ of PPS subchannels can be achieved through the rewriting version of (5b)

$$H^U(r_{t_k}) = \frac{-Q}{IUb_u P^{U/2}} \sum_{i=0}^{I-1} \{Y^U(i,t_k)\} \qquad \forall k \in [0, G-1]$$

Therefore, $\tilde{H}(r_{t_i})] = l_{t_i}[\tilde{H}^U(r_{t_i})]^{1/U}$ hold true, where $t_i \in [0, N-1]$ and $l_{t_i} \in \{e^{j(2\pi/U)n}\}_{n=0}^{U-1}$ is a scale ambiguity factor. To solve the scale ambiguity, we exhaustively search over all G^U possible vectors $\vec{\boldsymbol{H}} := [l_{t_0}[\tilde{H}^U(r_{t_0})]^{1/U}, l_{t_1}[\tilde{H}^U(r_{t_1})]^{1/U}, \cdots, \ l_{t_{G-1}}[\tilde{H}^U(r_{t_{G-1}})]^{1/U}]^T$;

Step 2　Vector $\overline{\boldsymbol{H}}$ composing of each sub-channel's amplitude can be obtained using formula (5b), (notice: the amplitude of PPS sub-channels must be divided by constant P), then $\vec{\boldsymbol{a}}$ is computed by $\vec{\boldsymbol{a}} = \boldsymbol{V}_M^+ \overline{\boldsymbol{H}}$;

Step 3　using formula (2), we compute the corresponding time domain vector through Least Squares fitting $\vec{\boldsymbol{h}} = \boldsymbol{V}_G^+ \vec{\boldsymbol{H}}$, then another estimation $\hat{\boldsymbol{a}}$ of $\vec{\boldsymbol{a}}$ can be achieved using (6), Channel estimations are then found by minimizing the Euclidean distance

$$\hat{\boldsymbol{h}} = \arg\min_{\vec{\boldsymbol{h}}} \left\| \hat{\boldsymbol{a}} - \vec{\boldsymbol{a}} \right\|^2 \tag{7}$$

After FFT processing, the estimation $\hat{H}(r_n)$ frequency response estimation $\hat{H}(r_n)$ of channel can be given.

4　Simulation

In this section, first, we illustrate the merits of our blind channel estimator through simulation of estimation mean square error of channel defined as: $MSE = \frac{1}{N}\sum_{k=0}^{N-1}\left\| H(r_k) - \hat{H}(r_k) \right\|^2$, where $H(r_k)$ is

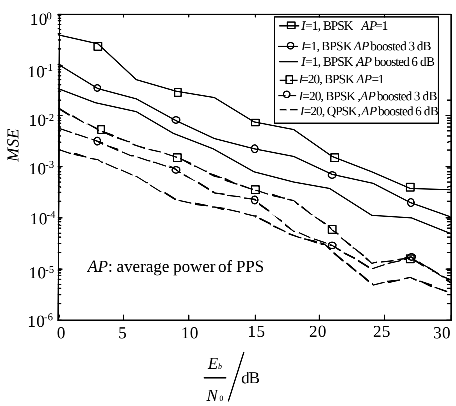

Fig.4　Comparison of MSE with difference of *I*

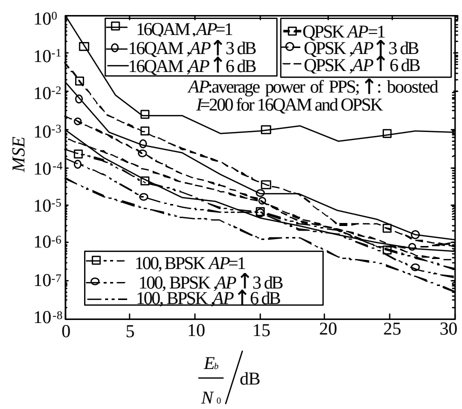

Fig.5　Comparison of MSE when PPSs being from different signal constellation

noiseless frequency response of channel and $\hat{H}(r_k)$ is the estimation value of frequency response of channel. For PPSs being BPSK as shown in Fig. 4, few OFDM symbols, even a OFDM symbol enable the estimation of impulse response of channel at high SNR. MSE of *I*=20 given at underside of the Fig. 4 implied that noise is suppressed by averaging for BPSK comparison with that of *I*=1. In Fig.5 the simulations of MSE of channel are shown when PPSs are 16QAM、QPSK and BPSK respectively, the results show: 1) When the average power of PPSs are constant, fewer signaling points of the constellation is, smaller of MSE can be gotten, and vice versa; 2) When PPSs are from a constellation, higher the average power of them is, smaller of MSE can be obtained; 3) Sometimes there may be estimation error floor of the MSE of channel, the main reason is that *I* is small when the statistic average is replaced by time average using *I* OFDM symbols.

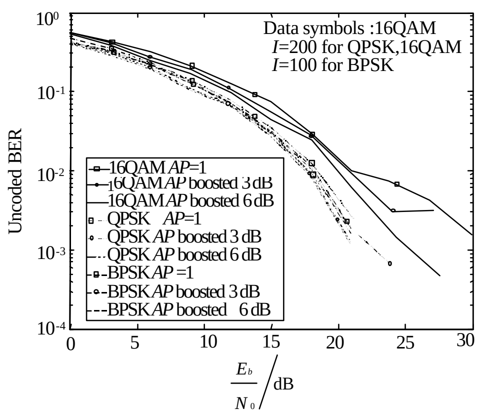

Fig. 6　Comparison of BER with PPSs from different signal constellation

The BER simulation experiments of uncoded OFDM systems are taken when data symbols are from 16QAM and PPSs are from 16QAM、QPSK and BPSK respectively, the results are shown in Fig. 6, it is easy to understand that the BER is the smallest when demodulated by using channel state information obtained by making use of PPSs which comes from the smallest constellation when I being constant.

5 Conclusion

In this paper, we develop a blind channel estimator by which the impulse response of channel can be obtained utilizing the second-order and fourth-order statistic quantity of frequency response of channel, because in OFDM system there are two properties that sub-channels are independent each other and that the information-bearing symbols loaded to sub-channels is from finite alphabet set. The simulation results prove the proposed blind channel estimator effective and show that it is very flexible both in which constellation PPSs being from and in boosting their average power or not in order to meet the requires of BER and the accurateness of channel estimation. The advantages of this proposed estimator have:1) Because the average power of PPSs rises, the additive channel noise can be effectively suppressed so that the accurate channel estimation can be obtained even at low SNR; 2) When PPSs are from BPSK constellation at high SNR, the impulse response response of channel can be gotten using PPSs from only one OFDM symbol so that the delay of estimator is very small; 3) The data symbols may come from a large–signal constellation, while the PPSs may be from a small-signal constellation; 4) After IFFT, the probability of large-PAPR when inserting PPSs is smaller than that when inserting pilots with constant phase.

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