Semiblind Channel Estimation for OFDM/OQAM Systems

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***Abstract*—In this letter, we propose a semiblind channel estimation technique for OFDM/OQAM wireless systems. The proposed technique exploits the real property of transmitted symbols to blindly identify the channel-induced rotation in the received signal, thereby reducing the pilot overhead for estimation purpose. Specifically, the channel phase over each subcarrier can be obtained from the spatial-sign covariance matrix of the received signal, while the channel amplitude can be expressed in terms of the subcarrier power. In effect, the frequency-domain noise can be reduced effectively by block averaging over multiple symbols. Since the channel delay spread is usually much smaller than the symbol duration, channel coefficients obtained from the estimated phase and amplitude can be further refined through low-rank filtering. Simulation results validate the efficacy of the proposed technique in time-dispersive fading channels, showing its robustness under different signal-to-noise ratio (SNR) conditions.**

***Index Terms*—OFDM/OQAM, semiblind channel estimation, spatial-sign covariance matrix.**

I. Introduction

OVER the past decades, multi-carrier modulation (MCM) has been well accepted as one of the main transmission technologies in wireless broadband communications [1]–[3]. As a classical MCM scheme, orthogonal frequency division multiplexing (OFDM) with cyclic prefix (CP) is of great popularity, thanks to its robustness against inter-symbol interference (ISI) and simple equalization structure. However, the use of CP to mitigate ISI in conventional OFDM-based transmissions comes at the expense of reduced spectral efficiency, especially in a severely time-dispersive fading channel.

As an alternative to improve spectral efficiency, OFDM based on offset quadrature amplitude modulation (OFDM/OQAM), which does not require CP for ISI mitigation, has received considerable research attention recently [4], [5]. In OFDM/OQAM systems, the transmitted data over each subcarrier is staggered alternatively on in-phase and quadrature components with a half symbol period. Furthermore, an even-symmetric, real-valued prototype filter with well-localized time-frequency content is employed for pulse shaping. As long as the prototype filter fulfills the real orthogonality condition, transmitted symbols over different subcarriers or time-instants become orthogonal to each other, leading to a distortion-free transmission under perfect synchronization and channel estimation [6], [7]. Since accurate channel state information is essential for data recovery, some research efforts have been devoted to channel estimation in OFDM/OQAM systems [8]–[12]. It should be noted that channel estimation methods developed for conventional OFDM systems cannot be applied to the OFDM/OQAM scheme directly, due to the presence of subcarrier interference prior to equalization.

In this letter, a semiblind channel estimation technique for OFDM/OQAM systems is proposed which exploits the non-circular property of the modulating signals. Specifically, the data symbol and the intrinsic interference, which present different statistical properties, combine in quadrature in the received signal. In the proposed method, the channel phase over each subcarrier is obtained from the spatial-sign covariance matrix [13], while the channel amplitude is expressed in terms of the subcarrier power. Effectively, the frequency-domain noise can be reduced by estimating these quantities through block averaging over multiple symbols. Further, the estimation accuracy can be refined through low-rank filtering of the frequency domain channel coefficients. Compared to existing training-based channel estimators for OFDM/OQAM systems, the pilot overhead required by the proposed scheme is minimal, as only one symbol block is needed to resolve the sign ambiguity in blind channel estimation. In addition, the computational complexity to perform channel estimation is very low. Simulation results confirm the efficacy of the proposed estimation technique in time-dispersive fading channels.

The rest of the letter is organized as follows. In Section II, the OFDM/OQAM system model is introduced and the unique properties of its symbol constellations are described. Based on the exploitation of these special properties, a semiblind channel estimation method using the spatial-sign covariance matrix and average subcarrier power is proposed in Section III. Finally, the simulation results are presented in Section IV and the conclusions are drawn in Section V.

II. OFDM/OQAM System

A. Transmission Model

In OFDM/OQAM systems, complex modulated symbols of duration T_0 are decomposed into their real and imaginary parts, each of which being transmitted at consecutive time instants with $\frac{T_0}{2}$ period. Accordingly, the baseband transmitted signal can be expressed as

$$x(t) = \sum_{n=-\infty}^{+\infty} \sum_{m=0}^{M-1} a_{m,n} \underbrace{g(t - n\tau_0) e^{j2\pi f_0 mt} e^{j\phi_{m,n}}}_{g_{m,n}(t)}, \qquad (1)$$

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where $a_{m,n}$ denotes the real scalar symbol transmitted over the m-th subcarrier at the n-th time instant, M is the total number of subcarriers and $g(t)$ is a real-valued prototype filter. The phase term $e^{j\phi_{m,n}}$ is determined by $\phi_{m,n} = \frac{\pi}{2}(n+m)$. In addition, f_0 and τ_0 denote the subcarrier spacing and the half symbol delay, respectively, which are related as $f_0 = \frac{1}{T_0} = \frac{1}{2\tau_0}$.

In order to guarantee distortion-free data recovery, the subcarrier's pulse shaping function $g_{m,n}(t)$, which is defined in terms of $g(t)$ in (1), should fulfill the following orthogonality condition in the real field:

$$\Re\left\{\int_{-\infty}^{+\infty} g_{m_1,n_1}(t)g^*_{m_2,n_2}(t)dt\right\} = \delta_{m_1,m_2}\delta_{n_1,n_2}, \quad (2)$$

where δ_{m_1,m_2} denotes the Kronecker delta function. Furthermore, to maintain a proper time-frequency confinement, the real-valued prototype filter $g(t)$ usually expands over multiple symbol durations, with support length $T_p = KT_0$, where K is the overlapping factor.

Let $h(\tau,t)$ denote the impulse response of the slowly fading time-dispersive channel at current time t and delay τ, with maximum delay spread $\tau_{\max}$. The received baseband signal can then be written as

$$\begin{aligned} y(t) &= \int_0^{\tau_{\max}} h(\tau,t)x(t-\tau)d\tau + w(t) \quad (3)\\ &= \sum_{n=-\infty}^{+\infty}\sum_{m=0}^{M-1} a_{m,n}e^{j2\pi f_0 mt}e^{j\phi_{m,n}} \\ &\quad \cdot \int_0^{\tau_{\max}} h(\tau,t)g(t-\tau-n\tau_0)e^{-j2\pi m f_0\tau}d\tau + w(t), \end{aligned}$$

where $w(t)$ is an additive white Gaussian noise (AWGN) with variance σ_n^2. In practice, compared to the prototype support T_p, the channel length $\tau_{\max}$ is relatively short. Hence $g(t-\tau-n\tau_0) \approx g(t-n\tau_0)$ in (3) and we have [8]

$$y(t) \approx \sum_{n=-\infty}^{+\infty}\sum_{m=0}^{M-1} H_m(t)a_{m,n}g_{m,n}(t) + w(t) \quad (4)$$

where $H_m(t) = \int_0^{\tau_{\max}} h(\tau,t)e^{-j2\pi m f_0 \tau}d\tau$.

Based on (4), the received signal in the frequency domain after demodulation can be expressed as:

$$\begin{aligned} Y_{m,n} &= \int_{-\infty}^{+\infty} y(t)g^*_{m,n}(t)dt \quad (5)\\ &\approx \sum_{q=-\infty}^{+\infty}\sum_{p=0}^{M-1} a_{p,q}\int_{-\infty}^{+\infty} H_p(t)g_{p,q}(t)g^*_{m,n}(t)dt + W_{m,n} \end{aligned}$$

where $W_{m,n}$ now denotes the corresponding frequency domain noise with variance σ_n^2. If we assume that the channel impulse response (CIR) does not change appreciably within the prototype filter duration, we can approximate $H_p(t) \approx H_p(q\tau_0) \triangleq H_{p,q}$ in (5), which then simplifies to

$$Y_{m,n} = H_{m,n}a_{m,n} + j\sum\sum_{(p,q)\neq(m,n)} H_{p,q}a_{p,q}G^{m,n}_{p,q} + W_{m,n} \quad (6)$$

where $G^{m,n}_{p,q} = \Im\{\int_{-\infty}^{+\infty} g_{p,q}(t)g^*_{m,n}(t)dt\}$ denotes the imaginary part of the cross-correlation between pulse shaping functions.

For a prototype filter with a good time-frequency localization property, the level of the cross-correlation $|G^{m,n}_{p,q}|$ decays rapidly as the separation between (p,q) and (m,n) increases, so that interference from subcarriers further away from the desired time-frequency bin can be ignored. Hence if we assume that the channel coefficients $H_{p,q}$ are nearly constant over the immediate neighborhood of (m,n), the demodulated signal can be finally approximated as [9], [10]:

$$Y_{m,n} \approx H_{m,n}(a_{m,n} + \underbrace{j\sum_{(p,q)\in\Omega_{m,n}} G^{m,n}_{p,q}a_{p,q}}_{I_{m,n}}) + W_{m,n}, \quad (7)$$

where $\Omega_{m,n}$ denotes a small neighborhood of points (p,q) around the desired time-frequency bin (m,n) (but excluding the latter) where $G^{m,n}_{p,q}$ is considered non-negligible.

B. Received Symbol Properties in OFDM/OQAM Systems

Thanks to the real orthogonality of the pulse shaping functions, the sum of interferences represented by the term $I_{m,n}$ in (7) only contributes an imaginary part to the real transmitted symbol $a_{m,n}$. Introducing the complex-valued, *effective* transmitted signal $S_{m,n} = a_{m,n} + jI_{m,n}$, we can rewrite (7) more compactly as

$$Y_{m,n} = H_{m,n}S_{m,n} + W_{m,n}. \quad (8)$$

This can be further expressed in matrix form in the real number domain as follows:

$$\underbrace{\begin{bmatrix} Y^R_{m,n} \\ Y^I_{m,n}\end{bmatrix}}_{\mathbf{y}_{m,n}} = \underbrace{\begin{bmatrix} H^R_{m,n} & -H^I_{m,n} \\ H^I_{m,n} & H^R_{m,n}\end{bmatrix}}_{\mathbf{H}_{m,n}}\underbrace{\begin{bmatrix} a_{m,n} \\ I_{m,n}\end{bmatrix}}_{\mathbf{s}_{m,n}} + \begin{bmatrix} W^R_{m,n} \\ W^I_{m,n}\end{bmatrix}. \quad (9)$$

In (9), the superscript R and I are used to indicate the real and imaginary parts of the corresponding variables; $\mathbf{H}_{m,n}$, $\mathbf{s}_{m,n}$, and $\mathbf{y}_{m,n}$ represent the real-valued channel matrix, effective transmitted signal vector and received signal vector, respectively. It should be noted that $\mathbf{H}_{m,n}$ is a scaled orthogonal matrix which satisfies $\mathbf{H}^T_{m,n}\mathbf{H}_{m,n} = |H_{m,n}|^2\mathbf{I}_2$. Accordingly, it can be deduced from (9) that $\mathbf{H}_{m,n}$ only rotates and scales the effective transmitted signal without changing its shape.

Since $a_{m,n}$ is a real symbol drawn from a finite alphabet $\mathcal{A}$ and $I_{m,n}$ is the sum of interferences, it is clear that they should exhibit completely different statistical properties. Fig. 1 shows a large sample of received symbols on a selected subcarrier in the case of transmission over an AWGN channel (SNR $= 30$ dB). It can be seen that the values of $Y^R_{m,n}$ are distributed around a set of discrete values at uniform intervals along the in-phase axis, while the values of $Y^I_{m,n}$ are scattered continuously along the quadrature axis. The different statistical properties of these two terms can be exploited for estimating the desired channel coefficient $H_{m,n}$ in (8).

Under the common assumptions that $E\{a_{p,q}\} = 0$ and $E\{a_{p,q}a_{p',q'}\} = \sigma_s^2\delta_{p,p'}\delta_{q,q'}$ for the real-valued random symbols at the transmitter, it follows from (7) that the interference term $I_{m,n}$ also has zero mean, that is

$$E\{I_{m,n}\} = \sum_{(p,q)\in\Omega_{m,n}} G^{m,n}_{p,q}E\{a_{p,q}\} = 0. \quad (10)$$

Furthermore, its variance can be approximated as

$$\begin{aligned} \sigma_I^2 = E\{|I_{m,n}|^2\} &= \sum_{(p,q)\in\Omega_{m,n}} \left|G^{m,n}_{p,q}\right|^2 E\{a^2_{p,q}\} \\ &\approx \sigma_s^2 \sum\sum_{(p,q)\neq(m,n)} \left|G^{m,n}_{p,q}\right|^2 = \sigma_s^2, \quad (11) \end{aligned}$$

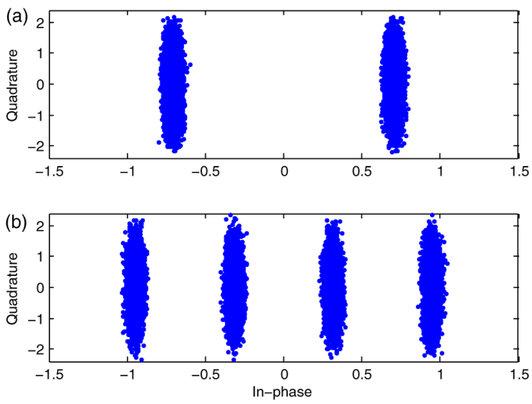

Fig. 1. Noisy received symbol $Y_{m,n}$ over AWGN channel (SNR = 30 dB). (a) Staggered offset-QPSK symbols. (b) Staggered offset-16QAM symbols.

where $\sigma_s^2 = E\{a_{p,q}^2\}$ denotes the average power of the real symbols. In the derivation of (11), we have included the small interference terms outside the neighborhood $\Omega_{m,n}$, and utilized the unique property of a real-valued prototype filter, which has been proved in [8], that is

$$\sum\sum_{(p,q)\neq(m,n)} \left|G_{p,q}^{m,n}\right|^2 = 1. \tag{12}$$

Since $I_{m,n}$ represents the summed interference from other time-frequency slots carrying independent data symbols, its statistics is different from that of real-valued symbols.

III. Semiblind Channel Estimation

Based on the above analysis, the received signal in (9) has zero mean and its covariance matrix is a scaled identity matrix:

$$E\{\mathbf{y}_{m,n}\mathbf{y}_{m,n}^T\} = (\sigma_s^2|H_{m,n}|^2 + \sigma_n^2/2)\mathbf{I}_2. \tag{13}$$

Both the transmitted and the received signals are isotropic (i.e., circularly symmetric) in their second-order statistics, and therefore, to estimate the per-subcarrier real channel matrix $\mathbf{H}_{m,n}$ derived from $H_{m,n}$, knowledge of the covariance matrix (13) is not sufficient. Instead, other signal properties should be utilized for this purpose, as further addressed below.

A. Rotation Estimation Utilizing Spatial-Sign Covariance Matrix

Thanks to the orthogonality inherent in $\mathbf{H}_{m,n}$ and real nature of the transmitted symbols $a_{m,n}$, the channel-induced rotation in the received signal $Y_{m,n}$ can be identified and further exploited for channel estimation.

Define the spatial sign of the complex-valued received signal $Y_{m,n}$, or equivalently $\mathbf{y}_{m,n}$ in (9), as follows [13]:

$$\tilde{\mathbf{y}}_{m,n} = \psi(\mathbf{y}_{m,n}) = \begin{cases} \frac{\mathbf{y}_{m,n}}{\|\mathbf{y}_{m,n}\|} & \text{if} \quad \|\mathbf{y}_{m,n}\| \neq 0 \\ \mathbf{0} & \text{if} \quad \|\mathbf{y}_{m,n}\| = 0 \end{cases}, \tag{14}$$

which is a unit vector aligned in the direction of $\mathbf{y}_{m,n}$. Invoking statistical properties of the transmitted symbols $a_{m,n}$ and interference terms $I_{m.n}$ along with modeling equation (9), it can be shown that $E\{\tilde{\mathbf{y}}_{m,n}\} = \mathbf{0}$. Accordingly, the spatial-sign covariance matrix can be defined as $\mathbf{C}_{m,n} = E\{\tilde{\mathbf{y}}_{m,n}\tilde{\mathbf{y}}_{m,n}^T\}$, and its corresponding eigenvalue decomposition (EVD) is given by

$$\mathbf{C}_{m,n} = \mathbf{U}_{m,n}\mathbf{\Sigma}_{m,n}\mathbf{U}_{m,n}^T, \tag{15}$$

where $\mathbf{U}_{m,n} = [\mathbf{u}_{m,n}^1, \mathbf{u}_{m,n}^2]$ is an orthogonal matrix and $\mathbf{\Sigma}_{m,n} = \mathrm{diag}(\sigma_1, \sigma_2)$ is a diagonal matrix with diagonal entries $\sigma_1 \geq \sigma_2$. According to this formulation, $\mathbf{u}_{m,n}^1 = [u_{m,n}^{1,R}, u_{m,n}^{1,I}]^T$ is the so-called dominant eigenvector of $\mathbf{C}_{m,n}$, i.e., associated to its largest eigenvalue σ_1.

As the spatial-sign covariance is rotation equivariant, the rotation present in the received signals, which is induced by the channel matrix $\mathbf{H}_{m,n}$ in (9), may be computed as the dominant eigenvector of $\mathbf{C}_{m,n}$ [13]. Specifically, $\mathbf{u}_{m,n}^1$ is aligned to vector $[H_{m,n}^R, H_{m,n}^I]^T$, except for a scalar difference in magnitude:

$$[H_{m,n}^R, H_{m,n}^I]^T = \alpha_m|H_{m,n}|\mathbf{u}_{m,n}^1, \alpha_m \in \{-1, +1\}. \tag{16}$$

Therefore, once the eigenvector $\mathbf{u}_{m,n}^1$ is obtained, the channel-induced rotation present in the received signal can be compensated by using $U_{m,n} = u_{m,n}^{1,R} + ju_{m,n}^{1,I}$.

B. Channel Amplitude Estimation

Since the statistical expectation of the received signal power is given by $P_{m,n} = E\{|Y_{m,n}|^2\} = |H_{m,n}|^2 2\sigma_s^2 + \sigma_n^2$, the channel coefficient over each subcarrier can be estimated as

$$\hat{H}_{m,n} = \alpha_m\sqrt{\frac{P_{m,n} - \sigma_n^2}{2\sigma_s^2}}U_{m,n}, \quad 1 \leq m \leq M. \tag{17}$$

To resolve the remaining sign ambiguity $\alpha_m \in \{-1, +1\}$, a single real pilot symbol need to be transmitted on the given subcarrier.

In the ensuing part, a refined channel estimate exploiting the low-rank property of the CIR will be discussed.

C. Algorithm Implementation with Low-Rank Filtering

Under the assumption of slowly fading channel, the received power over each subcarrier can be simply estimated by time averaging over the received samples as

$$\hat{P}_m = \frac{1}{N}\sum_{n=0}^{N-1}|Y_{m,n}|^2, \tag{18}$$

and the spatial-sign covariance matrix can be obtained by

$$\hat{\mathbf{C}}_m = \frac{1}{N}\sum_{n=0}^{N-1}\tilde{\mathbf{y}}_{m,n}\tilde{\mathbf{y}}_{m,n}^T, \tag{19}$$

where N is the number of symbols for time averaging. If the estimated channel power $\hat{P}_m$ is smaller than the noise variance σ_n^2 (assumed to be known), the corresponding channel amplitude is set to be zero. By computing $\hat{P}_m$ and performing EVD of $\hat{\mathbf{C}}_m$, we can estimate the channel coefficient (17) for each subcarrier. In particular, $\hat{U}_{m,n}$ in (17) is constructed using the entries of the dominant eigenvector of $\hat{\mathbf{C}}_m$.

To further reduce the estimation noise in $\hat{H}_{m,n}$, the low-rank property of the CIR can be exploited. This property follows since the maximum channel delay spread $\tau_{\max}$ is usually smaller than the half symbol duration τ_0. Represent the initially obtained frequency domain channel coefficients $\hat{H}_{m,n}$ as a column vector $\hat{\mathbf{h}}_n^1 = [\hat{H}_{1,n}, \hat{H}_{2,n}, \cdots, \hat{H}_{M,n}]^T$. The frequency domain channel estimate can be refined by

$$\hat{\mathbf{h}}_n^2 = \mathbf{F}_L\mathbf{F}_L^H\hat{\mathbf{h}}_n^1, \tag{20}$$

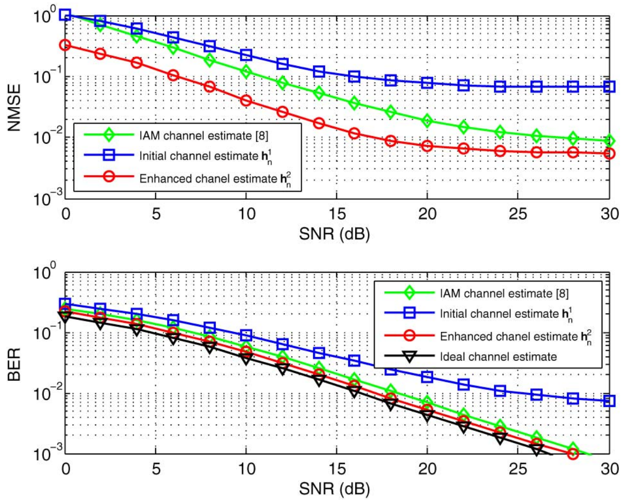

Fig. 2. NMSE and BER versus SNR for OFDM/OQAM systems with OQPSK modulation ($M = 128$, $N = 20$, $K = 3$).

where $\mathbf{F}_L$ is an $M \times L$ matrix with entries $f_{mn} = \frac{1}{\sqrt{M}} e^{-j2\pi mn/M}$, as obtained from the first L columns of a normalized Fourier transform matrix, and L represents the smallest integer larger than the normalized maximum channel delay spread $\lfloor M\tau_{\max}/(2\tau_0)\rfloor$.

IV. Simulation Results

In this section, the proposed semiblind channel estimation technique is evaluated under the Extended Pedestrian A (EPA) channel model with 7 taps [14]. In the simulated OFDM/OQAM system, the bandwidth is configured to be $B_s = 10$ MHz, leading to a sampling rate of $T_s = 100$ ns, and $M = 128$ subcarriers are employed for data transmission. A root-raised cosine (RRC) filter with a rolloff factor of 1 is used as the prototype. The filter support length is $T_p = KT_0$, where the overlapping factor is $K = 3$. To align with the configuration in [8], a preamble with duration of $3\tau_0$, where a non-zero pilot is located in the middle and surrounded by null blocks, is used for sign detection or channel estimation.

Fig. 2 illustrates the normalized mean-square error (NMSE) of the channel estimates and the bit error rate (BER) for equalized offset-QPSK modulation as a function of SNR. Here each data frame contains $N = 20$ symbol blocks, over which the simulated channel remains fixed. In addition, channel estimate based on the interference approximation method (IAM) [8] is shown for comparison. The enhanced channel estimate after low-rank filtering outperforms the initial semiblind channel estimate and the IAM channel estimate. Furthermore, with the enhanced estimate, the BER performance is close to that of the ideal case using the exact channel coefficients for equalization. In particular, for SNR less than 25 dB, the performance gap is only around 1 dB, which demonstrates the efficacy of the proposed semiblind channel estimation technique.

V. Conclusions

A semiblind channel estimation technique for OFDM/OQAM systems has been developed in this letter. In the proposed technique, the channel-induced rotation is estimated based on the spatial-sign covariance, while the channel gain is obtained from the estimated power over each subcarrier. To further reduce the estimation noise, low-rank filtering is performed over the blindly estimated channel coefficients. Numerical results under time-dispersive fading channels show that the proposed semiblind channel estimation technique can achieve a good performance while using a single pilot symbol per tone to resolve the sign ambiguity.

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