RECOVERY OF OQAM SIGNALS USING CARRIER FREQUENCY OFFSET AND BLIND SYMBOL TIMING ESTIMATION IN OFDM SYSTEMS

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Abstract

Orthogonal frequency-division multiplexing (OFDM) systems are highly sensitive to synchronization errors. We introduce an algorithm for the blind estimation of symbol timing and carrier frequency offset in wireless OFDM systems. We consider the problem of blind symbol timing (ST) estimation for pulse-shaping orthogonal frequency division multiplexing (OFDM) systems based on offset quadrature amplitude modulation (OQAM).This paper deals with the matter of blind synchronization for OFDM/OQAM systems. Specifically, by exploiting the approximate conjugate-symmetry property of the start of a burst of OFDM/OQAM symbols, attributable to the presence of the time offset, a replacement procedure for blind symbol timing and CFO estimation is proposed. By using computer simulations, the performances of Blind symbol estimators are analyzed. This result gives better Signal-to-Noise Ratio for proposed method.

1. INTRODUCTION

The principle of multicarrier modulation (MCM) consists of rending up a broadband signal at a high data rate into many lower rate signals, every one occupying a narrower band. A usually recognized advantage of MCM is its lustiness against differing kinds of channel distortions, like multipath propagation and narrowband interference. Orthogonal frequency division multiplex (OFDM) is definitely, until now, the foremost necessary category of MCM. The OFDM word form usually recovers two differing kinds of modulation. Within the 1st one, as planned, for example, in [1], every carrier is modulated victimization construction modulation (QAM). During this theme, which is additionally known as OFDM/QAM [2], QAM symbols are formed with an oblong window. In a very second class of OFDM systems, that is additionally known as orthogonally multiplexed QAM in [3] (O-QAM) or OFDM with offset QAM in [2] (OFDM/OQAM), the modulation used for every subcarrier may be a staggered offset QAM (OQAM). Each the OFDM/QAM and OFDM/OQAM modulation schemes on paper guarantee orthogonality and a most and identical spectral potency.

2. PROPOSED BLIND CFO ESTIMATION

In this section we introduce two different methods for blind CFO estimation. The first method, dubbed method A, can be derived from the fact that from (24),

$$D_3^{(T)} \triangleq \frac{d_3^{(T)}}{g_{1,i}} \cong \frac{d_0^{(T)}}{g_{0,i}} \triangleq D_1^{(T)} \qquad (1)$$

Where the division between the two vectors is defined component-wise. Consequently, in the presence of a normalized frequency-offset ε and neglecting the presence of noise and the effects of the channel, the entries of the vector $D_3^{(T)}/D_1^{(T)}$ are all equal to $exp\left(j2\pi\frac{\varepsilon}{M}M\right) = exp(j2\pi\varepsilon)$ since $D_1^{(T)}$ is extracted M samples on the left of $D_3^{(T)}$ .Therefore, we evaluate the angle $\widehat{\emptyset}_A$ of the average of the entries of $D_3^{(T)}/D_1^{(T)}$ and, then, we estimate the unknown CFO as $\widehat{\emptyset}_A/2\pi$. In calculating the average, we do not use the entries of $D_3^{(T)}/D_1^{(T)}$ with amplitude larger than 2 as they are assumed to be outliers. Note that the extraction of the vectors $D_k^{(T)}$ requires a previous ST estimation and that the errors in timing compensation worsen the performance of the CFO estimator.

The method B uses the property that, under ideal condition, the angles in the scalar products corresponding to the top value (θ = θ1) of the statistics and the value in $\theta = \theta 2$, which is $M/2$ samples at its left, differ only due to the presence of the CFO. More specifically, such a difference is equal to 2πε+π *and* the condition is ideal since it neglects the mismatch in the timing synchronization (obtained at the previous step), the effects of the channel and the presence of noise. Under such ideal conditions, the scalar product in (25) at $\theta = \theta 2$ can be written as follows:

$$v_1^{(\in,g)}(\theta_2).\ v_1^{(\in,g)\#}(\theta_2) = \sum_{m=0}^{\frac{M}{4}-1}\left(v^{(\in,g)}{}_{1,m}(\theta_2)\ v^{(\in,g)}{}_{2,\frac{M}{4}-2-m}(\theta_2)\right) \quad (2)$$

with the m th component $v^{(\in,g)}{}_{1,m}(\theta_2)$ of the vector $v^{(\in,g)}{}_1(\theta_2)$ ($m \in \{0,1,\ldots,\frac{M}{4}-2\}$ and l ϵ {1, 2}) defined as

$$v^{(\in,g)}{}_{1,m}(\theta_2) \triangleq v_{1,m}(\theta_2)\, g_{1,s,m+1}\, e^{\frac{j2\pi\epsilon}{M}m}\, e^{j\emptyset}$$
$$v^{(\in,g)}{}_{2,m}(\theta_2) \triangleq v_{2,\frac{M}{4}-2-m}(\theta_2)\, g_{1,s,\frac{M}{4}+\left(\frac{M}{4}-2-m\right)+1} .*\ e^{\frac{j2\pi\epsilon}{M}(m)}\, e^{j\emptyset\frac{j2\pi\epsilon M}{M\ 4}}\, e^{j\emptyset} \quad (2)$$

For $m \in \{0, 1\ldots\frac{M}{4}-2\}$, where $v_{1,m}(\theta_2)$ is the m th component $v_1(\theta_2)$ is the m th component of the vector $g_{1,s}\ (for\ m \in \{0,1,\ldots.\frac{M}{4}-2\})$ and Ø is a constant phase offset. The CS property implies that v1(θ_2) = $v_2^{\#}(\theta_2)$ (i.e. , $v_{1,m}(\theta_2) = v^{\#}{}_{2,\frac{M}{4}-2-m}(\theta_2)$) $m \in \{0,1,\ldots.\frac{M}{4}-2\}$) and consequently, (27) becomes

$$v_1^{(\in,g)}(\theta_2).\ v_1^{(\in,g)\#}(\theta_2) = \sum_{m=0}^{\frac{M}{4}-1}\left(v_{1,m}(\theta_2)\, g_{1,s,m+1}\, e^{\frac{j2\pi\epsilon}{M}m}\, e^{j\emptyset}\right) . .*\ ((v_{2,\frac{M}{4}-2-m}(\theta_2)\, g_{1,s,\frac{M}{4}+\left(\frac{M}{4}-2-m\right)+1} .*\ e^{\frac{j2\pi\epsilon}{M}\left(\frac{M}{4}-2-m\right)}\, e^{j\emptyset\frac{j2\pi\epsilon M}{M\ 4}}\, e^{j\emptyset})$$

$$= \sum_{m=0}^{\frac{M}{4}-1}\left(\begin{array}{c}|v_{1,m}(\theta_2)|^2\, g_{1,s,m+1}\ g_{1,s,\frac{M}{2}-m-1} \\ .*\ e^{j\emptyset\frac{j2\pi\epsilon M}{M\ 4}}\, e^{\frac{j2\pi\epsilon}{M}\left(\frac{M}{4}-2\right)}\, e^{2j\emptyset}\end{array}\right)$$

$$= e^{2j\emptyset}\, e^{j\pi\varepsilon}\ e^{-j2\pi\frac{2\epsilon}{4}} .*\ \sum_{m=0}^{\frac{M}{4}-1}\left(|v_{1,m}(\theta_2)|^2\, g_{1,s,m+1}\ g_{1,s,\frac{M}{2}-m-1}\right)$$

(29)

Analogously, the scalar product in (25) at $\theta = \theta 1$ can be therefore written as follows:

$$v_1^{(\in,g)}(\theta_2).\ v_1^{(\in,g)\#}(\theta_2) = = \sum_{m=0}^{\frac{M}{4}-1}\left(v_{1,m}(\theta_1)\, v_{2,\frac{M}{4}-2-m}(\theta_1)\right) \quad (4)$$

with the m th component $v^{(\in,g)}{}_{1,m}(\theta_1)$ of the vector $v^{(\in,g)}{}_1(\theta_1)$ ($m \in \{0,1,\ldots,\frac{M}{4}-2\}$ and l ϵ {1, 2}) defined as

$$v^{(\in,g)}{}_{1,m}(\theta_1) \triangleq v_{1,m}(\theta_1)\, g_{1,i,m+1}\, e^{\frac{j2\pi\epsilon}{M}m}\, e^{j2\pi\frac{\epsilon M}{m\,2}} e^{j\emptyset}$$
$$v^{(\in,g)}{}_{2,m}(\theta_1) \triangleq v_{2,\frac{M}{4}-2-m}(\theta_1)\, g_{1,i,\frac{M}{4}+\left(\frac{M}{4}-2-m\right)+1}\, e^{\frac{j2\pi\epsilon}{M}(m)}\, e^{j2\pi\frac{\epsilon M}{M\,4}}\, e^{j2\pi\frac{\epsilon M}{M\,2}}\ e^{j\emptyset}$$

For $m \in \{0, 1\ldots\frac{M}{4}-2\}$, where $v_{1,m}(\theta_1)$ is the m th component $v_1(\theta_1)$ is the m th component of the vector $g_{1,i}\ (for\ m \in \{0,1,\ldots.\frac{M}{4}-2\})$ and Ø is a constant phase offset. The CS property implies that v1(θ_1) = $v_2^{\#}(\theta_1)$ (i.e. , $v_{1,m}(\theta_1) = v^{\#}{}_{2,\frac{M}{4}-2-m}(\theta_1)$) $m \in \{0,1,\ldots.\frac{M}{4}-2\}$) and consequently, (30) becomes

$$v_1^{(\in,g)}(\theta_1).\ v_1^{(\in,g)\#}(\theta_1) = \sum_{m=0}^{\frac{M}{4}-1}\left(v_{1,m}(\theta_1)\, g_{1,i,m+1}\, e^{\frac{j2\pi\epsilon}{M}m}\, e^{j\emptyset}\right) . .*\ ((v_{2,\frac{M}{4}-2-m}(\theta_1)\, g_{1,i,\frac{M}{4}+\left(\frac{M}{4}-2-m\right)+1} .*\ e^{\frac{j2\pi\epsilon}{M}\left(\frac{M}{4}-2-m\right)}\, e^{j\emptyset\frac{j2\pi\epsilon M}{M\ 4}}\, e^{j\emptyset})$$

$$= \sum_{m=0}^{\frac{M}{4}-1}\left(\begin{array}{c}|v_{1,m}(\theta_1)|^2\, g_{1,i,m+1}\ g_{1,i,\frac{M}{2}-m-1} \\ .*\ e^{j\emptyset\frac{j2\pi\epsilon M}{M\ 4}}\, e^{\frac{j2\pi\epsilon}{M}\left(\frac{M}{4}-2\right)}\, e^{2j\emptyset}\end{array}\right)$$

$$= e^{2j\emptyset}\ e^{j\pi\varepsilon}\ e^{-j2\pi\frac{2\epsilon}{4}}$$

$$.*\ \sum_{m=0}^{\frac{M}{4}-1}\left(\left|v_{1,m}(\theta_1)\right|^2 g_{1,i,m+1}\ g_{1,i,\frac{M}{2}-m-1}\right) \quad (5)$$

In order to evaluate the phase difference between the quantities, we have to note that (see also Figure 4)

$$\sum_{m=0}^{\frac{M}{4}-2}\left(g_{1,s,m+1}\ g_{1,s,\frac{M}{2}-m-1}\right) < 0 \qquad (6)$$

$$\sum_{m=0}^{\frac{M}{4}-2}\left(g_{1,i,m+1}\ g_{1,i,\frac{M}{2}-m-1}\right) > 0 \qquad (7)$$

While (34) is obvious, (33) is verified with reference to the considered prototype. The method is straightforwardly adapted to a different prototype provided that the quantities in the left hand side are non null. The limited variations with m of $\left|v_{1,m}(\theta_1)\right|^2$ and $\left|v_{1,m}(\theta_2)\right|^2$ imply that

$$\angle\sum_{m=0}^{\frac{M}{4}-1}\left(\left|v_{1,m}(\theta_2)\right|^2 g_{1,s,m+1}\ g_{1,s,\frac{M}{2}-m-1}\right) = \pi$$

(35)

$$\angle\sum_{m=0}^{\frac{M}{4}-1}\left(\left|v_{1,m}(\theta_1)\right|^2 g_{1,i,m+1}\ g_{1,i,\frac{M}{2}-m-1}\right) = 0$$

(36)

From above equationsit follows that

$$\angle(v^{(\epsilon,g)}{}_1(\theta_1).(v^{(\epsilon,g)}{}_2(\theta_1)) - \angle(v^{(\epsilon,g)}{}_1(\theta_2).(v^{(\epsilon,g)\#}{}_2(\theta_2))$$

$$= 2\pi\varepsilon + \pi \qquad (8)$$

Finally, we introduce the method C that defines its output as the average of the results obtained by the methods A and B. For all three methods, to avoid ambiguities in the estimate, the condition |ε| < 0.5 must be satisfied since they estimate ε through the complex quantity exp (j2πε).

3. SIMULATION RESULTS

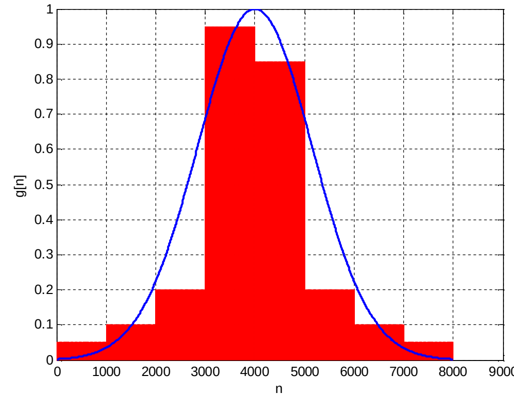

Figure 1: Number of samples vs Gain

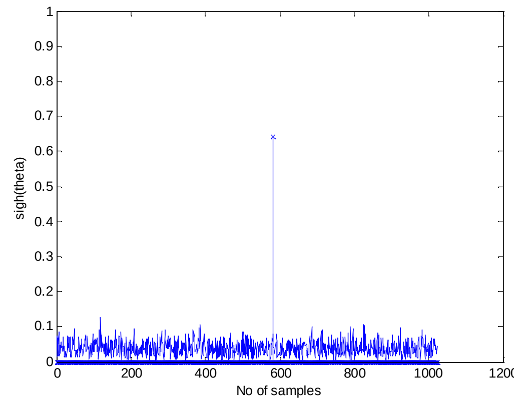

Figure 2: Number of samples vs theta

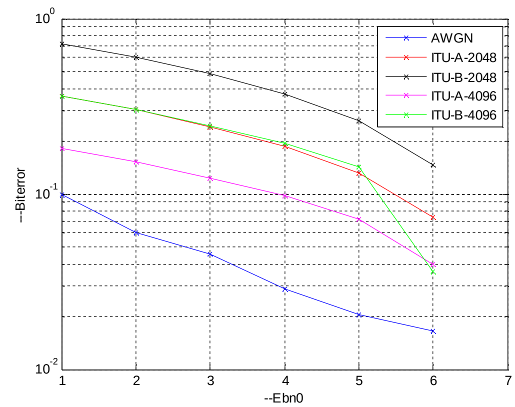

Figure 3: Performance analysis of ITU-A, ITU-B and AWGN channel under different symbols

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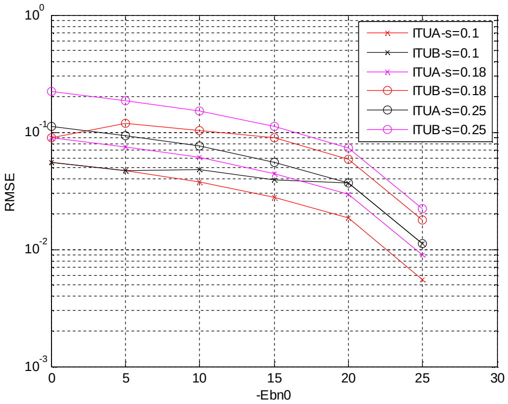

Figure 4: Performance analysis of ITU-A, ITU-B and AWGN channel under different Threshold

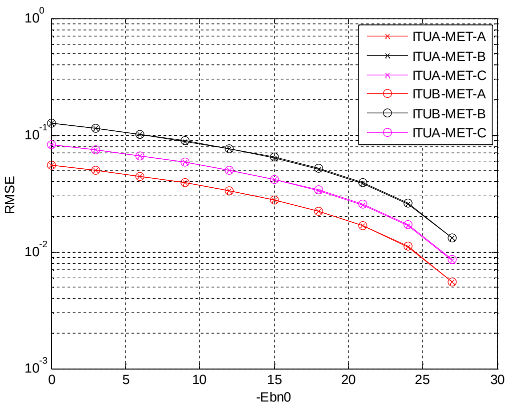

Figure 5: Performance analysis of ITU-A, ITU-B, for method A, B, and C

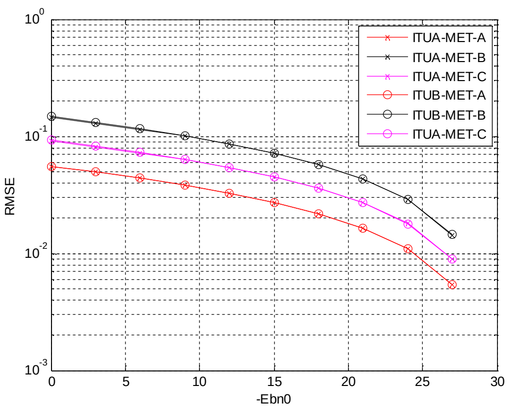

Figure 6: Performance analysis of ITU-A, ITU-B, for method A, B, and C under ETU channel

ETU channel is Rayleigh fading channel where the path gains are considered with respect to the vehicular speed. Here we considered the vehicle speed to be 50km/hr.

4. CONCLUSION

In this paper, the problem of blind symbol timing (ST) estimation for pulse-shaping orthogonal frequency division multiplexing (OFDM) systems based on offset Quadrature amplitude modulation (OQAM).This paper deals with the matter of blind synchronization for OFDM/OQAM systems. Specifically, by exploiting the approximate conjugate-symmetry property of the start of a burst of OFDM/OQAM symbols, attributable to the presence of the time offset, a replacement procedure for blind symbol timing and CFO estimation is proposed. By using computer simulations, the performances of Blind symbol estimators are analyzed. Those results are shown in simulation results. These results show better Bit Error Rate and Signal-to-Noise Ratio for proposed method.

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