Optics Communications 384 (2017) 78–84

Contents lists available at ScienceDirect

Optics Communications

journal homepage: www.elsevier.com/locate/optcom

An effective sampling clock synchronization method for continuous- and burst-mode transmission in OFDMA-PONs

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A R T I C L E I N F O

Keywords:
SFO
OFDMA
Long training symbol
Burst-mode

A B S T R A C T

A sampling frequency offset (SFO) estimation and compensation method based on frequency domain correlation of long training symbols for orthogonal frequency division multiple access passive optical network (OFDMA-PON) is experimentally demonstrated, which shows excellent performances in transmissions of continuous- and burst-mode. For continuous-mode transmission under a certain SFO, the proposed scheme can perform effectively in a wide received optical power (RoP) range from −8 dBm to −2 dBm and has high estimation veracity and a large applicable range as large as 100 ppm at a certain RoP. Similar behavior is also demonstrated under burst-mode transmission with tiny performance degradation caused by the fact that the algorithm needs time to reach a stable status of synchronization.

1. Introduction

Optical orthogonal frequency division multiple access (OOFDMA) technique has been studied intensively in passive optical networks (PONs) due to its high spectral efficiency and flexible bandwidth allocation scheme [1,2]. However, sensitivity to synchronization errors (e.g. symbol time offset (STO), carrier frequency offset (CFO) and sampling frequency offset (SFO)) is one of the difficulties for practical deployment of OFDMA-PONs. Usually SFO derives from unavoidable mismatching of sampling frequencies between transmitters and receivers, error of estimation and noise interference [3]. SFO causes additional STO, amplitude attenuation and phase distortion in received signals, and in the meantime, introduces inter-carrier interference (ICI) and consequently causes degradation of system performance.

Several SFO compensation methods have been proposed in recent years. A scheme based on interpolation algorithm is proposed in [4]. By using two long training symbols in the preamble, a closed-form ML-based acquisition algorithm for the CFO and SFO in OFDM systems is derived by Wang [5]. In [6] the channel estimation training sequences and pilots are reused to track SFO and CFO in system. An SFO estimation method based on periodically inserted identical binary prefix is proposed in [7]. Study in [8–12] all focus on joint estimation of CIR (channel impulse response), CFO and SFO, which adds much complexity to the realization of system. A relatively simple sampling clock offset estimation and compensation method for OFDM-FDMA systems is presented in [3], which is based on the well-known CFO estimation method suggested by Moose where two consecutive OFDM symbols which carry identical information are employed [13].

In the strategy of [3], two identical symbols in different positions with L symbols between them are employed to conduct sampling clock offset estimation and correction. In [8–12], sampling clock synchronization is only part of the joint estimation of channel CIR, CFO and SFO, and it's complex to conduct the joint process. Ref. [4–7] adopt interpolation algorithm, maximum likelihood method, channel estimation training sequences, pilots or periodically inserted identical binary prefix to execute sampling clock synchronization, which add complexity to the realization of algorithm. Taking [3] as a reference, our previous work put forward an improved SFO compensation method based on frequency domain correlation of long training symbols in adjacent OFDM frames [14] which is more accurate and less complexity in SFO estimation. Experimental verification with two users involved is also presented in [14] which shows sufficient precision in system synchronization procedure.

In this paper, a sampling frequency offset (SFO) estimation and compensation method based on frequency domain correlation of long training symbols for OFDMA-PON is experimentally demonstrated, which shows excellent performances in transmissions of both continuous-mode and burst-mode. This paper, comparing with [14], focuses on the investigation of transmission performance in two modes: continuous-mode and burst-mode, thus giving a more comprehensive description of practical application of the method. For continuous-mode under a certain SFO, proposed scheme can perform effectively in a wide received optical power (RoP) range from −8 dBm to −2 dBm and has high estimation veracity and a large applicable range as large

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http://dx.doi.org/10.1016/j.optcom.2016.10.012
Received 13 August 2016; Received in revised form 1 October 2016; Accepted 6 October 2016

as 100 ppm at a certain RoP. Similar behavior is also demonstrated under burst-mode transmission with tiny performance degradation caused by the fact that the algorithm needs time to reach a stable status of synchronization which indicates that the proposed synchronization method is suitable for both continuous- and burst-mode transmission in OFDMA-PON systems.

2. Principle of SFO estimation and compensation algorithm

For a common multi-point to point OFDMA-PON uplink system in which N sub-carriers and M users are involved, the m[th] user gets U sub-carriers (noted as $\mathbf{SetU^m}$). And signals on its U sub-carriers can be expressed as $[\mathbf{x_0^m}, \mathbf{x_1^m} \ldots \mathbf{x_{U-1}^m}]$.

At receiver side, the received signals can be expressed as:

$$y_n = \sum_{m=0}^{M-1} y_n^m = \frac{1}{\mathbf{N}} \sum_{m=0}^{M-1} \sum_{k=0}^{N-1} X_k^m H_k^m \boldsymbol{exp}(\boldsymbol{j2\pi nk/N}) + \boldsymbol{w_n} \quad (1)$$

where $\boldsymbol{H_k^m}$ is the channel response of m[th] user's k[th] subcarrier, and $\boldsymbol{w_n}$ is the total received noise.

In order to focus on the effects of SFO, we assume that CFO is zero and the receiver has already done frame synchronization and found the FFT window correctly. Then the relative SFO between user m and receiver can be written as: $\boldsymbol{\beta^m}=(\boldsymbol{f_{ts}^m}-\boldsymbol{f_{rs}})/\boldsymbol{f_{rs}}$, where $\boldsymbol{f_{ts}^m}$ is the sampling frequency of m[th] user's DAC, and $\boldsymbol{f_{rs}}$ is the sampling frequency of the receiver. According to Eq. (1), the received signal influenced by SFO can be reformed as:

$$y_n = \frac{\mathbf{1}}{\mathbf{N}} \sum_{m=0}^{M-1} \sum_{k=0}^{N-1} X_k^m \boldsymbol{H_k^m} \boldsymbol{exp}(\boldsymbol{j2\pi nk}(\mathbf{1} + \boldsymbol{\beta}^{\mathbf{m}})/N) + \boldsymbol{w_n} \quad (2)$$

Consider that there are l OFDM symbols between the long training sequences of two adjacent OFDM frames. For user A, the received long training sequence can be indicated as:

$$\boldsymbol{r_n^A} = \frac{1}{N} \sum\nolimits_{k=0}^{N-1} X_k^A \boldsymbol{H_k^A} \boldsymbol{exp}(j2\pi nk(1 + \beta^m)/N) \quad (3)$$

there are 64 data points and a cyclic prefix of 16 points in one OFDM symbol. Also, the number of sub-carriers, N, is set to be 64. So, there are $\frac{64+16}{64}N=\frac{5}{4}N$ points in one OFDM symbol. Consider that there are l OFDM symbols between the long training sequences of two adjacent OFDM frames, then there are $l\frac{5}{4}N$ points between the long training sequences of two adjacent OFDM frames (which is also the total point number in one OFDM frame). Every OFDM frame possesses the same long training sequence, so the received long training sequence of the next frame is

$$\begin{aligned} \boldsymbol{r^A_{n+l\frac{5}{4}N}} &= \frac{1}{N} \sum\nolimits_{k=0}^{N-1} X_k^A \boldsymbol{H_k^A} \boldsymbol{exp}\left(j2\pi\left(n + l\frac{5}{4}N\right)(k + k\beta^A)/N\right) \\ &= \boldsymbol{r_n^A} \cdot \boldsymbol{exp}\left(j2\pi lk(1 + \beta^A) \cdot \frac{5}{4}\right) \end{aligned} \quad (4)$$

Assuming that the channel response keeps the same during the period between the two long training sequences and the noise influence is ignored, we can do a correlation:

$$\sum_{k=0}^{N-1} (R_{1,k}^A)^* \cdot \boldsymbol{R_{2,k}^A} = \boldsymbol{\sigma_s^2} \cdot \boldsymbol{exp}\left(\boldsymbol{j\pi}(N_1 + N_2) \cdot \boldsymbol{l} \cdot \frac{\mathbf{5}}{\mathbf{4}} \cdot \boldsymbol{\beta^A}\right) \cdot \frac{\sin\left(\pi \cdot l \cdot \frac{5}{4} \cdot \beta^A \cdot (N_2 - N_1 + 1)\right)}{\sin\left(\pi \cdot l \cdot \frac{5}{4} \cdot \beta^A\right)} \quad (5)$$

where $\mathbf{R_{1,k}^A}$ and $\mathbf{R_{2,k}^A}$ are the FT (Fourier Transform) forms of received long training sequences of adjacent frames of user A.

Assume 'A' is among the M users and assigned with the sub-carriers ranging from serial number N_1 to N_2. The SFO of user A can be expressed as:

$$\boldsymbol{\beta^A} = \frac{1}{\boldsymbol{\pi} \cdot \boldsymbol{l} \cdot (N_1 + N_2) \cdot \frac{5}{4}} \boldsymbol{angle} \sum_{k=0}^{N-1} (\boldsymbol{R_{1,k}^A})^* \cdot \boldsymbol{R_{2,k}^A} \quad (6)$$

Compared with ideal received signals, signals polluted by SFO have a phase rotation, which is relevant to SFO. We can multiply an exponential term K to the signals after FFT to compensate such SFO:

$$\boldsymbol{K} = exp(\boldsymbol{j2\pi lk\beta^A} N_s/N_u) \quad (7)$$

where $\mathbf{N_u}$ is the length of OFDM data segment, and $\mathbf{N_s}$ is the length of the whole OFDM symbol (including CP).

3. Experimental setup

3.1. IMDD OFDMA-PON setup

Our experimental intensity modulation direct detection (IMDD) OFDMA-PON setup is shown in Fig. 1 with two users involved in system. Details of OFDM DSP procedures (user 1) are identical with the ones adopted in [14] and key parameters are listed in Table 1. In the transmitter side, two independent OFDM signals are generated after 16QAM mapping, 64 points IFFT and real value signal is achieved by using Hermitian conjunction during subcarrier allocation. 20% cyclic prefix (CP) and training sequence are then inserted for OFDM framing. User 1, 2 occupy subcarriers from number 3 to 13, and 18 to 28, respectively. Subcarriers 14–17 are used as guard band and

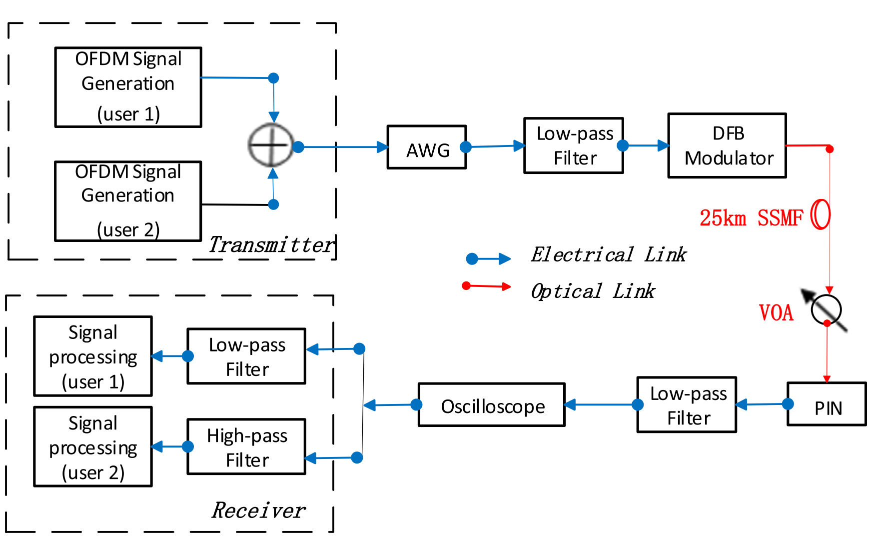

Fig. 1. OFDMA-PON experimental setup.