Quasi-Pilot Aided Phase Noise Estimation for Coherent Optical OFDM Systems

Son Thai Le, Thavamaran Kanesan, *Member, IEEE*, Elias Giacoumidis, Nick J. Doran, and Andrew D. Ellis

***Abstract*— In this letter, a novel phase noise estimation scheme has been proposed for coherent optical orthogonal frequency division multiplexing systems, the quasi-pilot-aided method. In this method, the phases of transmitted pilot subcarriers are deliberately correlated to the phases of data subcarriers. Accounting for this correlation in the receiver allows the required number of pilots needed for a sufficient estimation and compensation of phase noise to be reduced by a factor of 2 in comparison with the traditional pilot-aided phase noise estimation method. We carried out numerical simulation of a 40 Gb/s single polarization transmission system, and the outcome of the investigation indicates that by applying quasi-pilot-aided phase estimation, only four pilot subcarriers are needed for effective phase noise compensation.**

***Index Terms*— Coherent, phase noise, orthogonal frequency division multiplexing, phase estimation.**

I. Introduction

Coherent optical orthogonal frequency division multiplexing (CO-OFDM) has been considered as a promising candidate for long-haul optical communication systems because of its high spectral efficiency and excellent tolerance towards linear fiber impairments, such as chromatic dispersion and polarization mode dispersion [1]. However, compared to single carrier system, CO-OFDM has longer symbol duration, and therefore is more sensitive to laser phase noise. Laser phase noise introduces both common phase error (CPE) and intercarrier interference (ICI) [2], which significantly degrades the system performance. Therefore, it is crucial that the laser phase is rigorously tracked, estimated and effectively compensated.

Existing CO-OFDM phase noise compensation (PNC) may be divided into three groups, data aided (DA) [3], [4], pilot subcarrier aided (PA) [2], [5], and RF-pilot [5]. RF-pilot phase noise estimation is realized by inserting a RF-pilot tone in the middle of the OFDM band that can be used at the receiver to revert any phase noise related impairments. The typical power overhead of RF-pilot PNC is in the range of 7% to 10% due to the tradeoff between pilot and signal to noise ratios. In addition, RF-pilot phase noise estimation also requires overhead due to the frequency guard band surrounding the RF-pilot tone [5]. An overhead is also unavoidable in PA method, in which the CPE is estimated with the assistance of pilot subcarriers (PSs). On the other hand, DA is a blind PNC scheme that enables CPE estimation without any overhead by applying Mth-power-law to remove the data modulation or using a two-stage iterative algorithm with decision-directed technique as shown in [4]. Maximum-likelihood (ML) PNC can also be applied with or without any overhead [6]. Even though ML exhibits better performance and tolerance towards laser phase noise in comparison to DA but it requires highly exhausting computational load. Performances of these PNC techniques can be further improved by applying decision feedback (DF) (iterative algorithm), which re-estimates and re-compensates for the CPE after initial decision [7].

Among these aforementioned methods PA PNC is the most widely used method due to its inherently low complexity and high precision. However, the ultimate shortcoming of PA PNC is the required additional overhead. In this letter we propose a novel PNC scheme termed quasi-pilot aided (QPA) phase noise compensation. The QPA method retains the use of PSs at known frequencies to estimate the carrier phase; however, unlike the conventional PA scheme where pilot phases are predetermined, the pilot phases in QPA are data dependent. The major advantage of QPA estimation is that the number of PSs required for a similar performance to PA estimation can be reduced by a factor of 2, without significant additional complexity and differential encoding. It should be noticed that QPA is also a CPE-based scheme and thus, it is intended for application to non ICI-dominated channels. The effectiveness of the proposed method is demonstrated by comparing with common PNC methods, including PA, DA with M-th power law, RF-pilot and ML estimation for both back-to-back transmission regime and a 2000 km optical link with standard single mode fiber (SSMF).

II. Quasi-Pilot Aided Phase Noise Estimation

By assuming a perfect FFT window synchronization and frequency offset compensation, the received OFDM signal $R_{m,k}$ can be expressed as [2]:

$$R_{m,k} = S_{m,k} h_k \exp(j\Phi_m) + \varepsilon_{m,k} + n_{m,k} \quad (1)$$

where $S_{m,k}$ is the modulated data of the k^{th} subcarrier in the m^{th} symbol before transmission, h_{k} is the transmission channel response (including not only the optical channel, but also the transceiver front ends) for the k^{th} subcarrier, Φ_{m} is the CPE

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The authors are with the Aston Institute of Photonic Technologies, School of Engineering and Applied Science, Birmingham B4 7ET, U.K. (e-mail: let1@aston.ac.uk; thavamaran.kanesan@gmail.com; e.giacoumidis@aston.ac.uk; n.j.doran@aston.ac.uk; andrew.ellis@aston.ac.uk).

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for the m^{th} symbol due to laser phase noise or phase shifts acquired during optical fiber transmission. $\varepsilon_{\mathrm{m,k}}$ represents residual ICI and is generally treated as white Gaussian noise provided that a large number of OFDM subcarriers are used and the random Gaussian noise is represented by $n_{m,k}$.

In CO-OFDM system, the transmission channel response can be estimated by periodically transmitting training symbols for channel estimation, after which the phase drift is "reset" to 0. By transmitting a few PSs, the CPE can be estimated in PA method as [5], [8]:

$$\overline{\Phi}_m = \arg\left(\frac{1}{N_p}\sum_{pilots}\frac{R_{m,k}\cdot S^*_{m,k}}{|R_{m,k}|\cdot|S_{m,k}|}\right) \tag{2}$$

where arg(.) is the phase angle of the information symbol, $S_{m,k}$ is the known transmitted information symbol and N_p is the number of PSs.

It can be seen clearly in (2) that the accuracy of PA phase estimation technique is improved by increasing the number of PSs at the cost of proportionally increasing the overhead and so reducing the net data rate. To address this issue we propose that the effectiveness of PA phase estimation may be enhanced by modulating each pilot subcarrier with a data signal directly related to the signal on a data carrying subcarrier, rather than setting each pilot to a fixed predetermined state. As the pilots are no longer constant, we term this scheme quasi-pilot aided estimation. We consider two specific examples of QPA based estimation. In the first QPA scheme (QPA-1), all N_p pilot subcarriers are distributed equally in the first part of the OFDM band, taking the DC subcarrier as the symmetrical reference ($k = 0$). N_p pilot subcarriers ($S_{m,k}$, $k = k_1, k_2, \ldots k_{Np}$) are chosen by the condition:

$$S_{m,k} = S^*_{m,-k}, \tag{3}$$

where $*$ stands for the complex conjugate operation. That is, each pilot subcarrier is the complex conjugate of the data carrying subcarrier equally spaced from the central reference. This proposal can be regarded as a novel method for realizing a transmission scheme with semi-pilot (strongly encoded) symbols [9], [10]. Note that in general the positions of PSs and the correlated data carrying subcarriers can be chosen arbitrarily. Equation (3) provides an option of setting the pilots in QPA-1 scheme such that PSs and the correlated data carrying subcarriers are distributed equally among the OFDM band for achieving the best performance.

At the receiver, after performing channel estimation, each pilot subcarrier is coherently combined with its data carrying counterpart, eliminating the data modulation and enhancing the signal to noise ratio (SNR). The overall CPE is then estimated by summing the resultant modulation free vectors and taking the argument, as shown in the following expression:

$$\overline{\Phi}_m = \arg(\sum_{pilot} R_{m,k}\cdot R_{m,-k})/2, \tag{4}$$

This simple approach allows the CPE to be estimated without any prior information on the phases of PSs. In addition to this, the CPE is calculated by taking into account $2N_p$ subcarriers, which includes the complex conjugate data pilots in the first half and the actual data on the second half of the OFDM band. Thus, the accuracy of this estimation is similar to the PA phase estimation scheme whilst averaging the noise $n_{m,k}$ over $2N_p$ pilot subcarriers. In order to show the SNR advantage of QPA-1 scheme, let us consider the case when only one pilot is used for simplicity, and the extension to many pilots is straightforward. By assuming that $S_{m,k}h_k = S_{m,-k}h_{-k} = 1$ and there is no ICI, from (1) we have:

$$R_{m,k}\cdot R_{m,-k} = \exp(2j\Phi_m) + n'_{m,k} + n'_{m,-k} + n_{m,k}\cdot n_{m,k}$$

At intermediate-to-high SNR the product of the two noise terms can be neglected, leading to:

$$\begin{aligned} R_{m,k}\cdot R_{m,-k} &= \exp(2j\Phi_m) + n'_{m,k} + n'_{m,-k} \\ &= \rho\exp(j(2\Phi_m + 2\theta)) \end{aligned}$$

where ρ is the modulus, 2θ is the part of the complex noise that can be approximated to white Gaussian noise with power $\mathrm{NSR} = \mathrm{SNR}^{-1}$. By applying QPA-1 we get:

$$\overline{\Phi}_m = \Phi_m + \theta,\ for - -\pi/2 < \Phi_\mathrm{m} + \theta < \pi/2$$

It is clear that the power of θ is NSR/4, while the power of the noise component tangential to complex exponential in the conventional method (equation (2)) is NSR/2. As a result, QPA-1 scheme gives 3dB SNR gain over the conventional PA method.

In the second QPA scheme (QPA-2) instead of coding the pilots as direct conjugates of a single data subcarrier, the phases of N_p PSs are chosen such that their mean phase angle is opposite that of all the remaining data subcarriers, as specified by the condition:

$$\langle\arg(S_{m,k})\rangle_{pilots} + \langle\arg(S_{m,l})\rangle_{data} = 0, \tag{5}$$

where $\langle\cdot\rangle$ stands for the averaging operation. So the phases of the pilot subcarriers can be chosen equal to meet the requirement of (5):

$$\arg(S_{m,k})_{pilot} = -\langle\arg(S_{m,l})\rangle_{data}$$

At the receiver, the CPE is estimated by summing the phases of all pilots and associated data subcarriers as follows:

$$\overline{\Phi}_m = \left(\langle\arg(R_{m,k})\rangle_{pilots} + \langle\arg(R_{m,l})\rangle_{data}\right)/2 \tag{6}$$

Expression (6) also indicates that in QPA-2 the CPE is calculated without any prior knowledge of the phases of PSs. It is obvious that the accuracy of CPE estimation is improved significantly by taking all the subcarriers into consideration. The positions of PSs in QPA-2 scheme can be also chosen arbitrarily. However, for achieving the best performance, in QPA-2 scheme PSs should be equally distributed among the OFDM band. As the total phase of all symbols is constant, QPA-2 may be considered as a form of "phase parity." Note that in QPA schemes PSs are not inserted into training symbols, thus there is no impact during the synchronization process.

The phase noise tolerance of PNC methods can be increased by using the information about the CPE of the previous symbol(s) with an iterative algorithm [4], [7]. This approach is also applied here to QPA PNC. In the first stage of compensation;

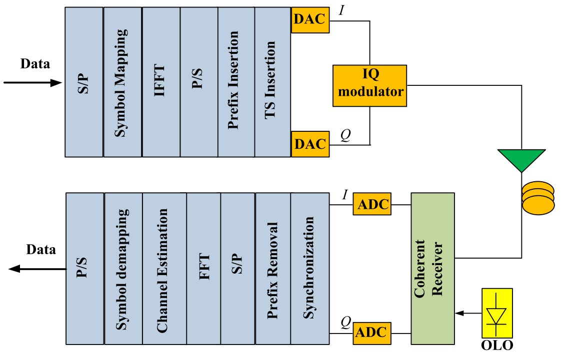

Fig. 1. Block diagrama of 40 Gb/s CO-OFDM transmissions. S/P: serial/parallel conversion, P/S: parallel/serial conversion, TS: training symbol, DAC: digital-to-analog converter, OLO: optical local oscillator.

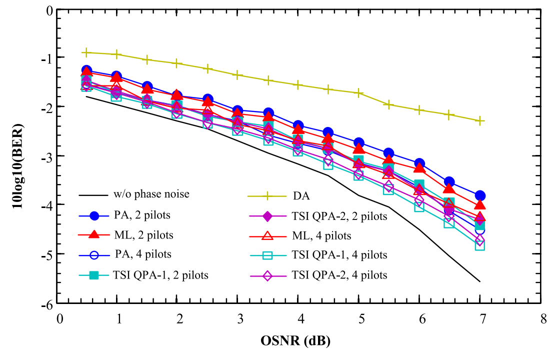

Fig. 2. Back to back BER sensitivity for DA (crosses), PA (circles), ML (triangles), TSI QPA-1 (squares), and TSI QPA-2 (diamonds) estimation methods when 2 (closed) or 4 (open) subcarriers do not carry independent information (pilots).

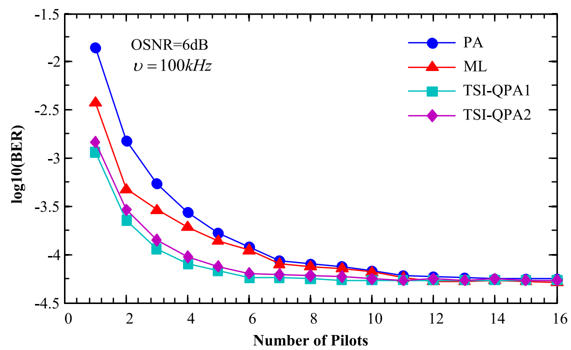

Fig. 3. BER as a function of the number of pilot subcarriers with different phase noise estimation methods, PA (circle), ML (triangle), TSI QPA-1 (square), and TSI QPA-2 (diamond).

the laser phase noise of the current symbol is compensated using the estimated CPE of the previous symbol. After that the difference between CPE estimations of the current and the previous symbols ($\Delta\Phi_m = \Phi_m - \Phi_{m-1}$) is estimated using expression (4) or (6). This scheme is known as two-stage-iterative (TSI) QPA PNC and can be regarded as an iterative algorithm, which increases the laser phase noise tolerance (in term of cycle slip and click probability) of QPA schemes significantly. The cycle slip probability (probability that the unwrapped phase lies outside the range $(-\pi/2, \pi/2]$) can be estimated as:

$$P \le 2 \cdot Q(\pi^2/4\sigma^2), \tag{7}$$

where Q is the Q-function, $\sigma^2 = 4\pi \upsilon T_u$, υ is the combined laser linewidth, T_u is the OFDM symbol duration.

III. Simulation Setup

In this section, we numerically investigate the effectiveness of TSI QPA-1 and TSI QPA-2 PNC methods in a 40Gb/s single polarization CO-OFDM system. The simulation framework is carried out in MATLAB and the setup is shown in the Fig. 1.

As depicted in Fig. 1, 40 Gb/s data stream is first mapped on 1000 subcarriers (1^{st} to 1000^{th}) using QPSK modulation format with Gray coding and subsequently transferred to the time domain by an IFFT of size 2048. A 12.5% cyclic prefix is appended to OFDM symbol to accommodate chromatic dispersion. The total OFDM symbol duration is 50ns. The fiber link is assumed to consist of 25 × 80-km spans of SSMF with the loss coefficient of 0.2 dB/km, PMD coefficient of $0.1ps/km^{0.5}$, the nonlinearity coefficient of $1.22W^{-1}km^{-1}$ and dispersion of 16 ps/nm/km. The fiber span loss is compensated by utilizing erbium-doped fiber amplifier (EDFA) with 16dB of gain and a noise figure of 6dB. The laser phase noise is modeled as a Wiener-Levy process with a variance $\sigma^2 = 2\pi \upsilon t$ where t is the time difference between two samples [5]. In simulation ASE noise is added inline after each fiber span. The simulated time window contains of 1000 OFDM symbols. The channel estimation and equalization is performed using zero forcing estimation method with the aid of training sequence (1 symbol), which is inserted after every 25 OFDM symbols.

Fig. 2 illustrates the BER sensitivity in the back-to-back transmission regime with different PNC techniques at the combined laser linewidth of 100 kHz. One can notice that ML performs slightly better than PA, which agrees with previous reports [6]. However, with 2 PSs both TSI QPA methods outperform PA and ML methods. With 4 PSs, PA and ML show approximately the same performance as the TSI QPA schemes using 2 PSs confirming that the overhead can be effectively reduced by a factor of 2. Closer inspection reveals that TSI QPA-1 slightly outperforms TSI QPA-2 owing to the slightly improved noise mitigation when using conjugated PSs. In the case of DA PNC applying the Mth-power law, no PSs are required but the performance suffers strongly by the phase ambiguity associated with the M^{th}-power-law, indicating that the technique is unsuitable for practical implementation.

The overhead benefits of TSI QPA are more clearly illustrated in Fig. 3 for a combined laser linewidth of 100 kHz, and a fixed OSNR of 6dB. It can be seen that with the conventional PA method, 8 PSs are required for negligible penalty (less than 5% degradation in BER). Similar overhead requirement can be observed for ML assisted PNC method. On the other hand, when TSI QPA-1 and TSI QPA-2 methods are applied the required number of PSs is 4, verifying that by applying QPA PNC methods, the required number of PSs can be reduced effectively by a factor of 2.

Fig. 4 presents the tolerance of the PNC methods studied here towards laser phase noise. Among these PNC methods, ML is the most tolerance technique to laser phase noise. For combined laser linewidth ≤ 600kHz, TSI QPA-1 and TSI QPA-2 methods with 2PSs require almost the same OSNR in comparison to PA technique with 4 PSs. For the

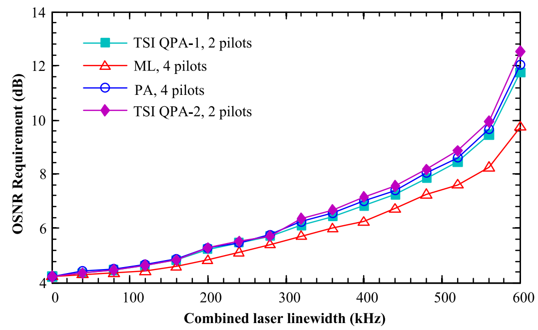

Fig. 4. OSNR requirement (for a BER level of 10^{-3}) as a function of the combined laser linewidth (υ) for different phase noise estimation.

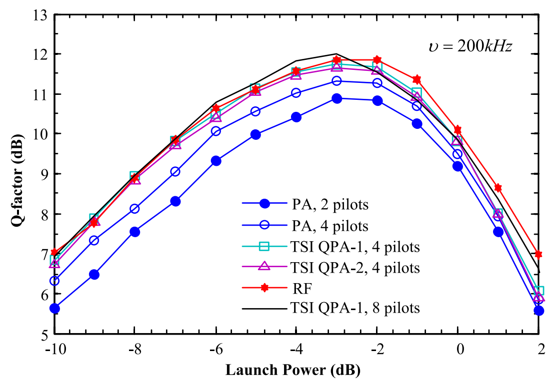

Fig. 5. Q-factor as a function of the launch power when different phase noise estimation methods are applied, after 2000km of transmission.

investigated system, the cycle slip probability at $\upsilon = 600$kHz is P $\leq 2Q(\pi^2/16\pi\,\upsilon T_u) \approx 10^{-38}$, which is less than the value required to avoid differential coding (10^{-18}) [11]. In addition, taking into account the fact that commercial external-cavity lasers have a linewidth of around 100 kHz, the advantage and reliability of TSI QPA PNC methods with reduced overhead have very high credibility.

Performances of different phase noise estimation methods at 2000km long-haul optical transmission link are compared in the Fig. 5. In this experiment the linewidth of the transmitter laser and the local oscillator are assumed to be 100 kHz. The system's Q-factor is delivered by the BER obtained through direct error counting [12]. In this figure we also show the performance of RF-pilot enabled PNC method. In order to apply this method, a RF-pilot tone (with 6.3% of power overhead) is inserted into the middle of the OFDM band. A frequency guard band of 100MHz is added surrounding the RF-pilot tone, resulting in around 0.4% of frequency overhead. At the receiver, a low pass filter (LPF) with a bandwidth of 10 MHz is applied to filter out the RF-pilot tone and subsequently applied for phase noise estimation and compensation. Due to the LPF, the complexity of RF-pilot PNC method is significantly higher in comparison to PA and QPA methods.

At the optimum optical launch power, TSI QPA-1 and TSI QPA-2 methods with 4 PSs, and PA with 8 PSs exhibit almost the same performance as RF-pilot tone PNC. Even though theoretically RF-pilot tone PNC can compensate for both CPE and ICI [5] but the effectiveness of RF-pilot tone method reduces under ASE and fiber nonlinearity impairments. In addition, the RF-pilot tone method is strongly affected by the size of the frequency guard band surrounding the DC subcarrier; consequently with a small overhead (0.4% in this letter) RF-pilot tone PNC does not show advantage in comparison to QPA PNC methods. In particular, 4 pilot TSI QPA also requires a 0.4% overhead, which is almost equivalent to that of the RF-pilot tone. As a result, TSI QPA methods are also more effective than RF-pilot tone in term of the balance among performance, overhead and the complexity.

IV. Conclusion

We have proposed a novel common phase error estimation technique based on correlating the phase of pilot tones with data subcarriers. Numerical simulation results have confirmed that by setting PSs in correlation with data subcarriers the overhead of pilot aided carrier phase estimation may be reduced by a factor of 2 for two different correlation techniques, conjugated pilots and phase parity pilots. In addition, in comparison with RF-pilot tone, the proposed methods can offer a similar performance at the same overhead while significantly reducing the complexity in implementation.

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