Superscalar Parallel Carrier Phase Recovery with Transmitter I/Q Imbalance Compensation

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Abstract: We propose a superscalar parallel two-stage carrier phase recovery architecture to improve the performance of optical coherent receivers in the presence of Tx I/Q imbalance, Tx I/Q skew, and laser frequency fluctuations. © 2021 The Author(s)

1. Introduction

Decision-directed (DD) carrier phase recovery (CPR) [2] and the blind phase search (BPS) algorithm [3] have been widely used in optical coherent receivers to compensate the impact of laser phase noise. The two-stage CPR scheme based on a DD low latency parallel phase locked loop (DD-PLL) followed by a feedforward CPR (e.g., BPS) has demonstrated an excellent performance in the presence of laser phase noise and frequency fluctuations. The latter are generated by mechanical vibrations or power supply noise and modeled as a carrier frequency modulation with a sinusoid of large amplitude (e.g., 200 MHz) and low frequency (e.g., 35 KHz) [2]. CPR algorithms such as DD-PLL or BPS use a symbol detector (or slicer) to estimate phase error. Therefore, severe performance degradation may be experienced in the presence of amplitude and phase imbalances introduced at the transmitter side, between the in-phase (I) and quadrature (Q) components (i.e., Tx I/Q imbalance) [4]. Tx I/Q time skew is another impairment that degrades the receiver performance.

The compensation of the I/Q imbalance has been extensively addressed in the literature (e.g., [5, 6]). Typically, the transmitter I/Q imbalance compensation at the receiver is achieved after the CPR stage. Unfortunately, the performance of this approach with conventional CPR algorithms such as DD-PLL or BPS can be seriously degraded with high-order modulation formats if severe Tx I/Q imbalance is present. This degradation is a result of the constellation warping at the input of the slicers in the CPR blocks caused by the Tx I/Q imbalance [4]. To deal with this problem, some blind estimation schemes of the Tx I/Q imbalance have been recently reported in the literature [4, 7]. Although these proposals can achieve good compensation of the Tx I/Q imbalance, their proper performance in the presence of laser frequency fluctuations is challenged, owing to the large latency introduced by their parallel implementation in high-speed transceivers [7], and the poor performance of feedforward CPRs such as BPS [4] when carrier frequency fluctuations are experienced [2].

This work proposes a superscalar parallel (SSP) two-stage CPR to improve the receiver performance in the presence of the Tx I/Q imbalance, Tx I/Q skew, laser phase noise, and carrier frequency fluctuations. A first CPR stage based on a DD-PLL is used to compensate frequency offset and carrier frequency fluctuations [2]. The second stage, based on a feedforward CPR (e.g., BPS), operates on the signal demodulated by the DD-PLL and is mainly used to compensate the residual laser phase noise not eliminated by the DD-PLL. The accuracy of the phase estimations provided by both DD-PLL and BPS is improved by adding a one-tap adaptive 2 × 2 multiple-input multiple-output (MIMO) equalizer to compensate the Tx I/Q imbalance at the input of the slicers in the CPR. As a result of the extra latency introduced by the one-tap MIMO equalizer in the DD-PLL stage, the use of existing low latency parallel DD-PLL schemes such as that proposed in [2] is precluded for implementing in high-speed receivers. Therefore, we use a superscalar parallel architecture to reduce the latency of the DD-PLL [8]. We show that the proposed SSP-based CPR scheme with Tx I/Q imbalance compensation is able to drastically improve the receiver performance even in the presence of Tx I/Q time skew.

2. Superscalar Parallel Two-Stage CPR with Tx I/Q Compensation

Without loss of generality, we consider the transmitted signal in one polarization. The complex optical carrier with Tx I/Q imbalance can be written as

\[ p(t) = (1 - \epsilon_g) \cos(\omega_0 t + \phi_e/2 + \theta(t)) + j(1 + \epsilon_g) \sin(\omega_0 t - \phi_e/2 + \theta(t)), \]

where \( \epsilon_g \) and \( \phi_e \) are the gain and phase imbalance, respectively, \( \theta(t) \) is the carrier phase error, while \( \omega_0 = 2\pi f_0 \) with \( f_0 \) being the carrier frequency [5]. Let \( \alpha_n = \alpha_n^{(1)} + j\alpha_n^{(2)} \) and \( T \) be the \( n \)-th transmitted complex quadrature amplitude modulation (QAM) symbol and the symbol period, respectively. The equalized received baseband signal (e.g., after chromatic dispersion compensation) in a dispersive optical channel with Tx I/Q imbalance and carrier phase error can be formulated as

\[ r_n = P_n W_n \alpha_n + z_n, \]  

(1)
where $\mathbf{r}_n = [r_n^{(I)} \; r_n^{(Q)}]^T$ is a $2 \times 1$ real vector whose components are the received I/Q samples (superscript $T$ denotes transpose), $\mathbf{a}_n = [a_n^{(I)} \; a_n^{(Q)}]^T$, $\mathbf{z}_n = [z_n^{(I)} \; z_n^{(Q)}]^T$ is the vector that includes the amplified spontaneous emission (ASE) noise as well as any other residual interference, while $\mathbf{W}$ and $\mathbf{P}_n$ are the $2 \times 2$ real matrices:

$$
\mathbf{W} = \begin{bmatrix}
\cos(\frac{\phi_n}{2}) & \sin(\frac{\phi_n}{2}) \\
\sin(\frac{\phi_n}{2}) & \cos(\frac{\phi_n}{2})
\end{bmatrix} \begin{bmatrix}
1 - \varepsilon_g & 0 \\
0 & 1 + \varepsilon_g
\end{bmatrix};
\mathbf{P}_n = \begin{bmatrix}
\cos(\theta_n) & -\sin(\theta_n) \\
\sin(\theta_n) & \cos(\theta_n)
\end{bmatrix},
$$

with $\theta_n = \theta(nT)$. Notice that $\mathbf{W}$ models the Tx I/Q imbalance while $\mathbf{P}_n$ incorporates the effect of the phase error, $\theta_n$. The latter is modeled as $\theta_n = \psi_n + \Omega_n + \Delta \Omega_n$, where $\psi_n$ is the laser phase noise (i.e., $\psi_n = \sum_{k=1}^{\infty} \eta_k$ where $\eta_k$ are zero-mean iid Gaussian random variables with variance $\sigma_\psi^2 = 2\pi T \Delta \nu$, being $\Delta \nu$ the laser linewidth), $\Omega_n = 2\pi T f_c$, with $f_c$ being the frequency offset, while $\Delta \Omega_n$ is the phase change caused by the frequency fluctuation $\Delta \Omega_n = (A_\varphi / f_c) \sin(2\pi T \Delta \nu n)$, where $A_\varphi$ and $f_c$ are the amplitude and frequency of the carrier modulation tone [2]. Figure 1-a depicts the proposed first-stage CPR based on a DD-PLL. The received signal (1) is first demodulated by using a $2 \times 2$ rotation matrix $\mathbf{Q}_n$, which uses the carrier phase $\hat{\theta}_n$ provided by the DD-PLL (e.g., notice that $\mathbf{Q}_n = \mathbf{P}_n^{-1}$ if $\hat{\theta}_n = \theta_n$). Then, the samples are processed by a $2 \times 2$ real matrix $\mathbf{C}$ in order to compensate the Tx I/Q imbalance at the slicer input (ideally, $\mathbf{C} = \mathbf{W}^{-1}$). After that, the BPS algorithm estimates the phase error by testing $B$ different phases. However, as pointed out in [4], the Tx I/Q imbalance may cause BPS to make wrong decisions and subsequently degrade the accuracy of the phase estimation. To avoid this degradation, multiplication of the slicer inputs by a $2 \times 2$ real matrix $\mathbf{C}$ is introduced in the BPS to compensate the effects of the Tx I/Q imbalance, similarly to what was done before with the DD-PLL as shown in Fig. 1-a.

2.1 Impact of the Tx I/Q Time Skew

The transmitter I/Q time skew degrades the receiver performance. This effect is mainly caused by mismatches between the I/Q electrical path responses from the Tx digital-to-analog converters (DACs) unto the Mach Zehnder modulator (MZM). Based on the models used in [5], the discrete time electrical signal at the MZM input in the presence of I/Q time skew can be written as $s_n = \sum_k \mathbf{H}_k \mathbf{a}_{n-k}$, where $\mathbf{H}_k$ are $2 \times 2$ real matrices whose elements depend on the I/Q skew, $\tau$, and the Tx filter. Then, the received signal (1) is given by $r_n = \mathbf{P}_n \mathbf{W}_n \mathbf{a}_n + \mathbf{z}_n + \mathbf{q}_n$, where $\mathbf{W}_n = \mathbf{W} \mathbf{H}_n$ while $\mathbf{q}_n = \mathbf{P}_n \mathbf{H}_n \mathbf{a}_{n-k}$ is the $2 \times 1$ real vector with the interference components caused by the Tx I/Q time skew and the adjacent symbols. Therefore, the proposed DD-PLL+BPS scheme with $\mathbf{C} = \mathbf{W}_0^{-1}$ can compensate the Tx I/Q imbalance ($\mathbf{W}$) and a part of the Tx I/Q time skew ($\mathbf{H}_0$). Thus, the interference component $\mathbf{q}_n$ will be seen by the proposed DD-PLL+BPS as an extra noise component.

2.2 Superscalar Parallel Implementation of DD-PLL (SSP-PLL)

The high symbol rate requirements mandate the use of parallel processing techniques for the implementation of coherent transceivers. The tolerance of a conventional interleaving parallelization system DD-PLL to the laser linewidth is reduced by a factor $N \times D_L$, where $N$ is the parallelization factor and $D_L$ is the processing delay (latency) due to the pipelined implementation of the DD-PLL. A low latency parallel DD-PLL CPR has been proposed in [2] to improve the tolerance to the laser linewidth and frequency fluctuations. Unfortunately, the benefits of this low latency PLL [2] will be significantly reduced due to the presence of the MIMO compensation filter $\mathbf{C}$ in the loop. To deal with this problem, we propose to implement a superscalar parallelization DD-PLL. The superscalar parallelization of the PLL has been proposed in [8] for implementing a feedback CPR in high-speed optical transceivers. SSP employs pilot symbols and a buffer to store the input samples, which are then rearranged to have consecutive symbols in each parallelized channel (see Fig. 1-b). In this way, the processing delay of the parallel architecture can be reduced from $N \times D_L$ to $D_L$. Notice that a parallel implementation of all the DSP blocks after the SSP-PLL stage (e.g., BPS, Tx I/Q imbalance and skew compensation) is straightforward.

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1Pilot symbols of the SSP-PLL are used to correct cycle slips (CS) generated in the different CPR stages.
3. Simulation Results

We investigate the performance of the proposed two-stage CPR in the presence of Tx I/Q imbalance and time skew, laser phase noise, and carrier frequency fluctuations. We consider 16-QAM with a baud rate of 1/T = 32 Giga-baud (GBd), a type II second-order digital PLL with D_L = 5, and a BPS algorithm with filter length M = 40 and B = 32 test phases. The MIMO tap C is estimated at the receiver using the LMS algorithm. The block buffer size is N = 16 and S = 400 with a pilot overhead of 1%. We consider A_p = 140MHz, Δf_c = 35KHz, and Δν = 1MHz. The Tx I/Q skew compensator at the Rx is implemented by using two independent baud-rate adaptive real equalizers.

Figure 2-a shows the optical signal-to-noise ratio (OSNR) penalty at a bit-error-rate (BER) of 10^{-3} versus the Tx phase and the gain imbalance with Tx I/Q time skew of τ = 0.1T. Notice the important degradation caused by the Tx I/Q imbalance in the conventional SSP-PLL+BPS solution. In contrast, the proposed two-stage CPR architecture is able to drastically reduce the OSNR penalty at high values of gain and phase errors. Nevertheless, it is interesting to highlight that the receiver performance worsens with the increase of the Tx gain and phase imbalance even with the proposed 2-stage CPR and a perfect estimation of the MIMO compensation matrix C. This is caused in part by the ASE noise amplification generated by the MIMO compensation filter since det(C) ≈ det(W^{-1}) = 1/det(W) > 1 if 0 < |φ_e| < 45° and/or 0 < |ε_g| < 1 (see (2)). Finally, Fig. 2-b depicts the OSNR penalty at a BER of 10^{-3} versus the Tx I/Q time skew. Note that the proposed SSP-PLL+BPS architecture for |τ| < 0.2T achieves an important gain respect to the conventional CPR without Tx I/Q imbalance compensation.

4. Conclusions

A superscalar parallel CPR architecture with Tx I/Q imbalance compensation for coherent optical receivers has been proposed. Numerical results have demonstrated that the proposed parallel CPR scheme is able to drastically improve the receiver performance in the presence of Tx I/Q imbalance, Tx I/Q time skew, and laser frequency fluctuations. This improvement will be more noticeable when the modulation order is increased (e.g., 64-QAM).

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