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A Very Simple Algorithm of Sequential IQ Imbalance and Carrier Frequency Offset Compensation in Coherent Optical OFDM

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Abstract

In this work, a hardware efficient algorithm for sequential in-phase (I) and quadrature (Q) imbalance (IQ imbalance) and carrier frequency offset (CFO) compensation under chromatic dispersion (CD) and phase noise (PN) environment is proposed. Two identical orthogonal frequency division multiplexing (OFDM) symbols, which are namely training sequences (TSs), are used to acquire CFO and IQ imbalance coefficients. The CFO is obtained by calculating phase differences between the two TSs. To achieve the image frequency interference factors which are caused by an IQ imbalance effect, each symbol of the TS is modulated on only a half of all the subcarriers while the remaining subcarriers are modulated with zeros. By doing this, the IQ imbalance coefficients are directly estimated without recursive calculation requirements. This brings a low complexity to implementation in hardware. The performances of the modeling system are evaluated by a numerical simulation method where the error vector magnitude (EVM), the bit error ratio (BER), and the mean square error (MSE) quantities are used as performance indicators. The numerical simulation results are showed that the performance of the modeling system is enormously improved even when highly dispersive channels and phase noise are considered.

Keywords: *IQ imbalance, frequency offset, Coherent optical-OFDM, Optical communication systems*

1. Introduction

Coherent optical orthogonal frequency division multiplexing (CO-OFDM) is a promising modulation format for high speed optical communication systems because the intersymbol interference (ISI) can be neglected due to a cyclic prefix (CP) extension where it appended into the head of each OFDM system (1-2). However, CO-OFDM is very sensitive to phase and amplitude offsets such as IQ modulator imbalance, phase mismatch between transmitter and receiver, carrier frequency offset (CFO) and phase noise (PN), respectively. This paper considers all of those offsets; however, only amplitude and phase imbalance in the presence of CFO are focused under chromatic dispersion and PN. In practical, the imperfection of modulator and demodulator equipment generates in-phase (I) and quadrature (Q) imbalance (IQ imbalance). However, the impact can be characterized into two main issues. One is the phase miss matched where the phase imbalance of the transmitter comes from the Mach-Zehnder modulator (MZM) when I and Q components do not have a phase difference of 90° for transmitting information signal. Second, the amplitude imbalance comes from imperfect of digital-to-analog conversion (DAC) or similar effects.

Several methods and techniques had been proposed by many institutes. For example, a hybrid frequency-time domain compensation has been proposed by (3-4) for compensation of IQ imbalance channel effects. Transmitting two identical training symbol (TS) sequences are used to learn the IQ imbalance coefficients. However, this is very complex to implement in hardware at large fast Furrier transforms (FFT) and the TS can only estimate for IQ imbalance and not for CFO. Chung and et al. (5) had been presented a method for compensating quadrature imbalance by using the Gram-Schmidt orthogonalization procedure (GSOP). This shows the disadvantage that the minimum of 1000 received sampling signals are needed for averaging in I and Q, which leads to long delays and long calculations. Recently, Nguyen and et al. (6) had been proposed IQ imbalance compensation by estimating signal to noise ratio (SNR) of the received signal where the received phase is adjusted. If the tuning phase is matched, IQ imbalance is cancelled. Consequently, the SNR will be reached to peak. However, the range of the phase tuning is limited to only $-\pi/4:\pi/4$

In this paper, an efficient low-complexity algorithm to estimate and compensate for IQ imbalance and CFO by using two training sequences (TS) is proposed. The TS is used to learn the coefficients of those two impacts. In addition, since the image frequency interference due to the IQ imbalance is appeared on the opposite frequency on each other; therefore, to determine the IQ imbalance coefficient and CFO, each one modulates on only one half of the subcarriers. The rest is modulated by zeros. Obviously, the proposed method is calculated faster than the GSOP and can be estimated for both IQ imbalance and CFO because the data which is used to estimate for that is only a half of FFT size. The TS is followed by many OFDM symbols, thereby completing the OFDM frame as shown in Figure 1.

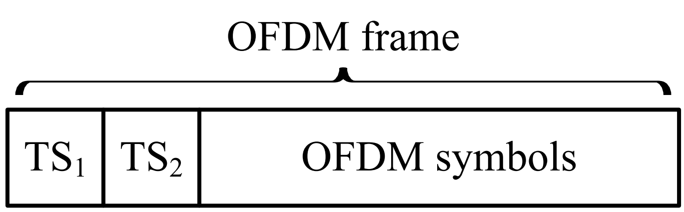

Figure 1. Example of OFDM frame with TS in the time domain.

2. Impact of IQ Imbalance under CFO Effect

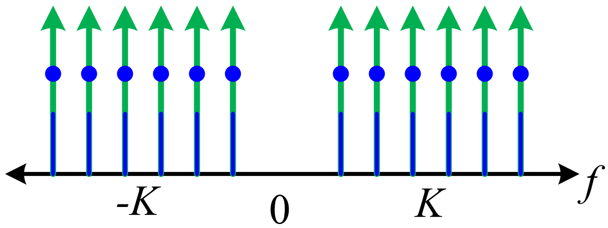

Figure 2. Equivalent interference image of K subcarriers due to IQ imbalance.

IQ Imbalance generates image frequency interference on all of subcarriers that are modulated with non-zero data, as shown an example in Figure 2. This phenomenon comes from a phase mismatch ψ, i.e. the deviation of the IQ angle from 90°, and a relative amplitude mismatch which is denoted by η.

In Figure 2, the lines ending with arrows are the original subcarriers, and the lines ending with circles are the image interference signals of subcarriers with opposite frequencies. The received sampled signal with the effect of IQ imbalance under CFO, which the CFO is denoted by ε, can be modeled in time domain by (7-8),

$$y_{IC}(n) = \alpha y(n)e^{j2\pi\varepsilon n/N} + \beta^* y(n)e^{-j2\pi\varepsilon n/N}, \quad (1)$$

where

$$\alpha = \cos(\psi) + j\eta\sin(\psi), \quad (2)$$

$$\beta = \eta\cos(\psi) - j\sin(\psi). \quad (3)$$

Hence, $y_{IC}(n)$ is the received signal which induced by IQ imbalance and CFO. $y(n)$ is the received time-domain signal resulting from the optical signal transmitted over the channel, n is the temporal index and $(\cdot)^*$ means complex conjugation. N denotes the size of the FFT (number of subcarriers). It holds $\varepsilon = f'/\Delta f$, where f' is actual frequency offset and Δf is the subcarrier spacing. Hence, the image rejection ratio (IRR), which is the power ratio of the original subcarriers to the interference image, is given by, (9),

$$IRR_{dB} = 20\log_{10}\left|\frac{\alpha}{\beta}\right|. \quad (4)$$

Ideally the IRR will become infinite.

Regarding equation 1, let us assume that the CFO is first perfectly estimated $\hat{\varepsilon} = \varepsilon$, then if we first compensate for CFO in the time domain by a multiplication of equation 1 with either $e^{-j2\pi\varepsilon n/N}$ or $e^{j2\pi\varepsilon n/N}$. Hence, the compensated signal will be (7)

$$y_{-\varepsilon}(n) = \alpha y(n) + \beta y^*(n)e^{-2j2\pi\varepsilon n/N}, \quad (5)$$

$$y_{+\varepsilon}(n) = \alpha y(n)e^{2j2\pi\varepsilon n/N} + \beta y^*(n). \quad (6)$$

As we can see here, this is found that we cannot cancel CFO before cancelling IQ imbalance. So we need to first cancel the IQ imbalance before CFO and PN compensation.

3. Proposed Method

The detail of the proposed method is discussed in this section. Two training OFDM symbols, which are identical, is used but modulating on only one half of the subcarriers in each one. This is called the K half, as shown in Figure 3.

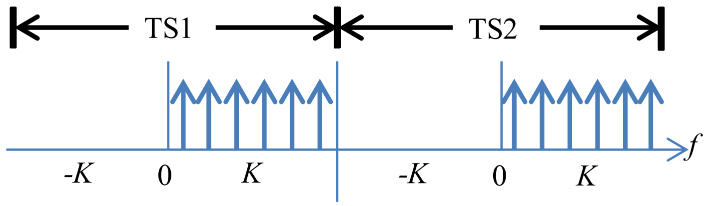

Figure 3. Two identical training OFDM symbols which modulate only one half of the subcarriers.

With respect to Figure 3 and Figure 4, the CFO can be estimated as

$$\hat{\varepsilon} = \frac{1}{2\pi}\tan^{-1}\left(\frac{\sum_{K=1}^{N/2-1}\mathrm{Im}\left(Y_1^*(K)Y_2(K)\right)}{\sum_{K=1}^{N/2-1}\mathrm{Re}\left(Y_1^*(K)Y_2(K)\right)}\right). \quad (7)$$

Equation 7 calculates the CFO from the phase difference between the two training symbols and it generally called the Moose scheme (10). Y_1 and Y_2 are the output signals of the FFT unit with and without one symbol delay, as detailed in Figure 4. From Figure 3, we can see that if IQ imbalance occurs, the image interference will appear in the other, -*K* half of the spectrum. Therefore, if we set the *K* half subcarriers of the TS to be one, we can obtain the IQ imbalance coefficient directly by

$$\hat{\alpha}_K = Y_{IC}(K), \quad \hat{\beta}_K = Y_{IC}(-K), \quad (8)$$

where $\hat{\alpha}_K$ and $\hat{\beta}_K$ are the estimated values of μ and β on the K*th* subcarrier. However, the estimation is still inaccurate due to dispersion of the optical channel, CFO and PN. Therefore, from reference (8) and also by theoretical, we found that the channel effects on the *K* half are conjugate to those on the -*K* half. Therefore, after having calculated $\hat{\alpha}_K$ and $\hat{\beta}_K$ by

$$\hat{\beta} \approx E\left\{\frac{\hat{\beta}_K}{\hat{\alpha}_K^*}\right\}, \quad (9)$$

$$\hat{\alpha} = \sqrt{1-\mathrm{Im}^2(\hat{\beta})} + j\frac{\mathrm{Re}(\hat{\beta})\,\mathrm{Im}(\hat{\beta})}{\sqrt{1-\mathrm{Im}^2(\hat{\beta})}}. \quad (10)$$

Then we can cancel the channel impacts. $E\{\cdot\}$ is the expected value operation. Next, the CFO and IQ imbalance effects are compensated in the time domain which is better than to compensate them in the frequency domain by the two steps expressed in equation 11-12, respectively,

$$\hat{y}_{IQ}(n) = \frac{\hat{\alpha}^* y_{IC}(n) - \hat{\beta} y_{IC}^*(n)}{|\hat{\alpha}|^2 - |\hat{\beta}|^2}, \quad (11)$$

$$\hat{y}(n) = \hat{y}_{IQ}(n)e^{-j2\pi\hat{\varepsilon}n/N}. \quad (12)$$

Finally, $\hat{y}(n)$ is fed into the next OFDM process to back recover the information data, involving the steps of phase noise compensation, FFT, channel compensation and constellation de-mapping, as shown in Figure 4.

4. Simulation model

Figure 4 details the overall simulation setup. There are two main steps in the sequential which are estimation and compensation processes. In the first step, OFDM frame synchronization process is first operated. Next, the IQ imbalance coefficients and the CFO are estimated by using the definition of equation 7-9 as already discussed from the last section. Finally, to minimize intercarrier interference, a Hamming window is used which it minimizes subcarrier leakage due to PN and CFO.

The second step is a compensation for IQ imbalance and CFO by using the estimated coefficients from the first step by using equation 11-12 while the radio frequency pilot (RF-pilot) method (11) is used for PN cancelling but it is beyond the scope of this work. A Monte Carlo simulation is used to evaluate the performance of the IQ compensation. The FFT size is 1024 points with 12.5% CP, 30% zero padding for oversampling, and 3% pilot-data for one-tap channel equalization in which is used to compensate the linear channel impairments. The first channel is set to be an RF-pilot for phase noise compensation as detailed in (11). The sampling rate is 28 Gs/s. The FFT size of the TS is 64.

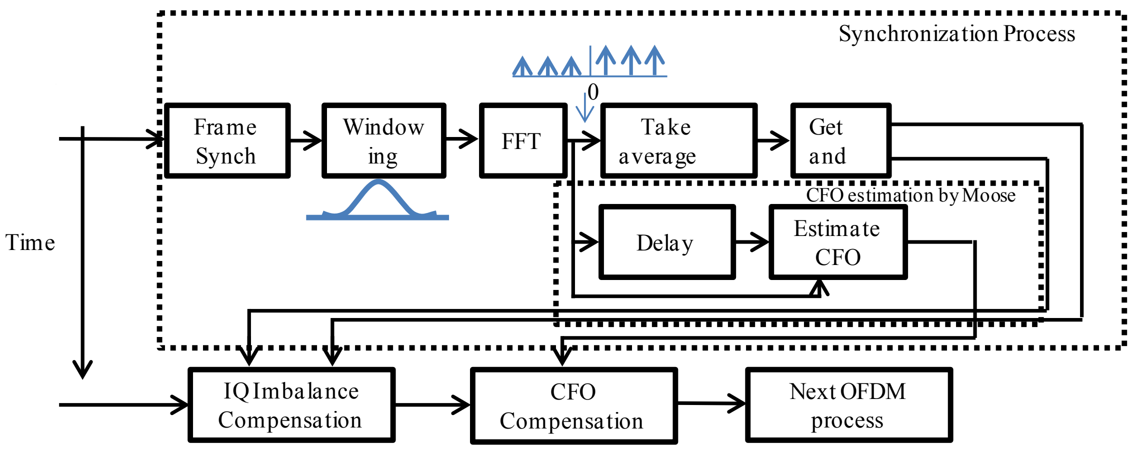

Figure 4. Simulation setup diagram.

5. Simulation Results

To investigate the performance of sequential CFO and IQ compensation for CO-OFDM, two cases of studies are modeled. Firstly, the CFO and sum laser linewidth are fixed where the CFO is 0.49, which is the maximum value from the Moose scheme able to be estimated. The sum laser linewidth is 100 kHz. The fiber is modeled as lossless with a chromatic dispersion (CD) of 10798 ps/nm (8 spans, 79.4 km each, with 17 ps/nm/km). However, the IQ imbalance oceffients are varied.

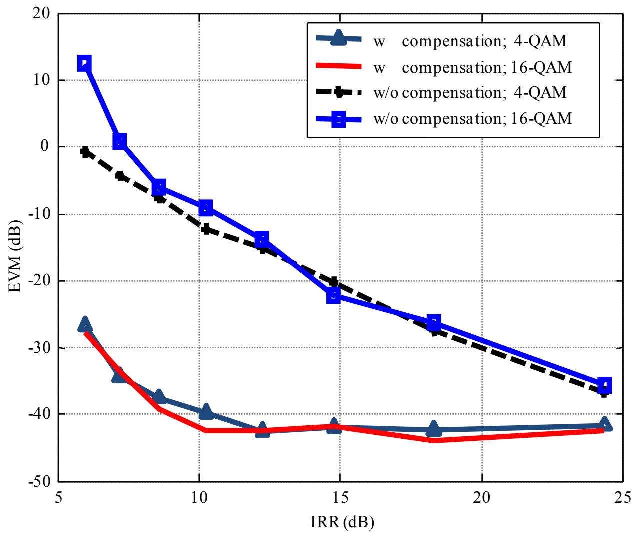

Figure 5. EVM versus IRR @ OSNR 25 dB.

Figure 5 shows the error vector magnitude (EVM) in dB versus IRR at an optical signal to noise ratio (OSNR) of 25 dB for 4- and 16-quadrature amplitude modulation (QAM) constellation types. As can be seen from the compensation case, the EVM has a lot of improvement. At the IRR of 6 dB, for both 4-QAM and 16-QAM, the EVM of -26.9 dB and of -28 dB are achieved, for example. For higher IRRs, it decreases down to -42 dB, at an IRR of 14.8 dB and then, the EVM is nearly stable till an IRR of 24.4 dB.

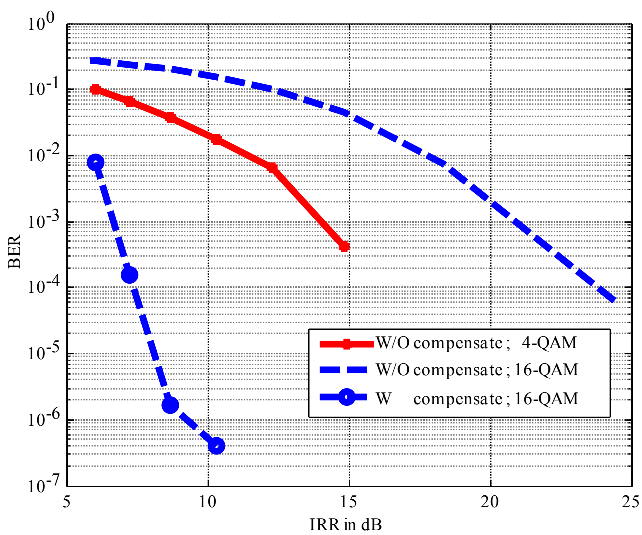

Figure 6. BER versus varying IRR @ OSNR 25 dB.

Figure 6 plots bit error ratio (BER) versus IRR at OSNR of 25 dB. It is also clear that the BER sensitivity against IRR enormously improved after compensation for IQ imbalance and CFO. For 16-QAM, the BER is $4.1 \cdot 10^{-7}$ can be achieved at an IRR of only 10.3 dB.

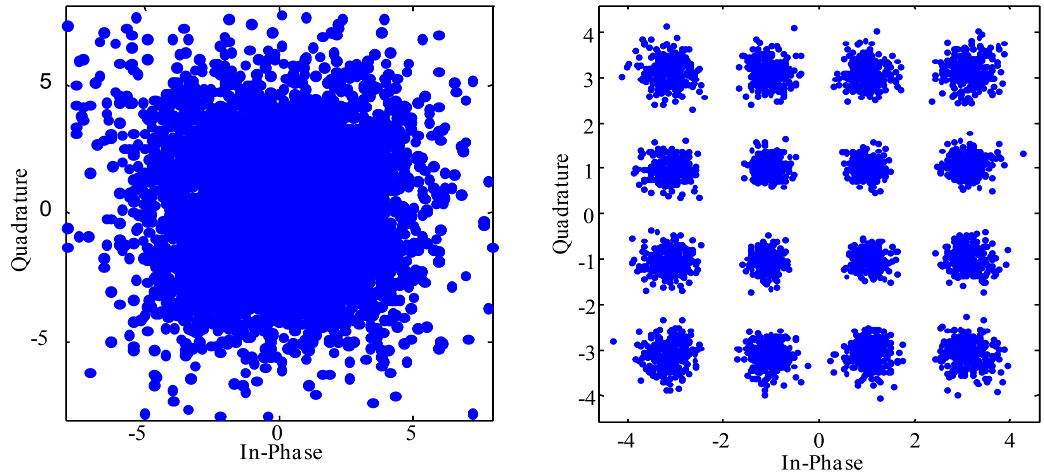

Figure 7. Constellation of 16-QAM; (a) before compensation of joint IQ imbalance and CFO, and (b) after compensation of joint IQ imbalance and CFO, at OSNR of 20 dB, 15° of phase mismatch and 15% of amplitude mismatch**.**

The 16-QAM constellation of before and after previous and behind compensation for sequential CFO and IQ imbalance are shown in Figure 7(a) and Figure 7(b), respectively. For compensation case which is shown in Figure 7(b), the constellation point is very clear and correct. The without compensation, see Figure 7(a), is not understandable because of inter-carriers-interference (ICI) occurred due to the effects of CFO and IQ imbalance.

Secondly, the OSNR of 20dB, phase mismatch $\psi = 15°$, a relative amplitude imbalance $\eta = 15\%$, and CD of 10798 ps/nm are fixed. However, the CFO ε varies from 0.05 to 0.45. Then, the estimation performance of β and μ are evaluated by the mean square error (MSE). The MSE is shown in Figure. 8.

Figure 8 plots the MSE quantity, which $\hat{\beta}$ and $\hat{\mu}$ are estimated from β and μ, respectively. By varying CFO, the estimated of $\hat{\beta}$ and $\hat{\mu}$ are almost the same for all ε values from 0.05 to 0.45. Hence, the MSE of $2.9 \cdot 10^{-4}$ for $\hat{\beta}$ and $2.6 \cdot 10^{-5}$ for $\hat{\mu}$, can be achieved. The change of ε brings slightly increase of MSE, because the estimation of high ε value less sensitive to noise.

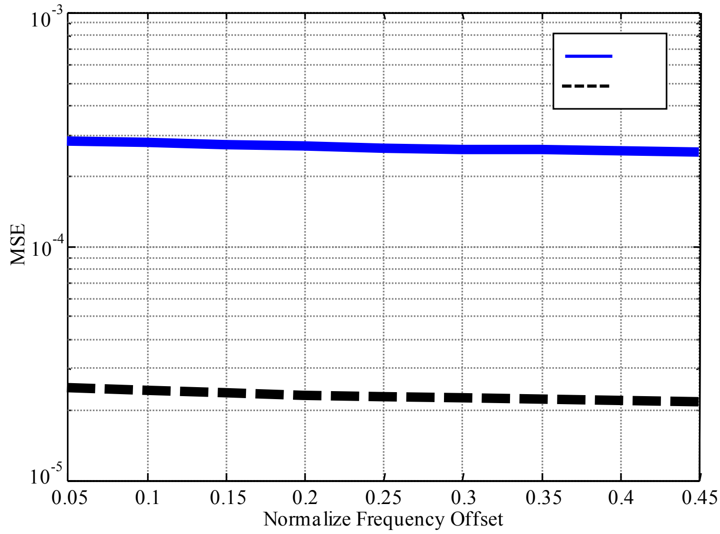

Figure 8. MSE of the estimated $\hat{\beta}$ and $\hat{\mu}$ versus normalize frequency offset (ε) at OSNR of 20, ψ=15° and relative amplitude imbalance η=15%.

4. Conclusion

A hardware efficient compensation algorithm for sequential carrier frequency offset and IQ imbalance have been proposed. Two identical OFDM symbols are used as a training sequence to learn the CFO and IQ imbalance coefficients which each symbol is modulated on only one half of the subcarriers. Its complexity is very low and it takes shorter time of calculation when compared with the conventional methods as discussed in the introduction section. The results show drastic performance improvements under compensation for both 4-QAM and 16-QAM even when high phase noise and optical dispersive channel are considered.

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