Performance Degradation Induced by Pulse Carver/Data Modulator Misalignment in RZ-DPSK Systems

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Abstract: The impacts of the distortion induced by pulse carver/data modulator misalignment on the performance of RZ-DPSK systems are investigated. Such distortion can be generalized as intersymbol interference and strongly depends on input data pulse shape.

I. Introduction
Return-to-zero differential phase-shift keying (RZ-DPSK) has recently attracted much attention, because it provides 3-dB higher receiver sensitivity than conventional on-off keying (OOK) if balanced detection is employed, and promises better tolerance against nonlinear fiber transmission impairments [1]. For proper generation of RZ-DPSK signal, it is essential to locate the peak of RZ pulses in the middle of the bit slot. However, the relative time delay of optical and electrical devices drifts over time due to temperature variation and aging of devices, leading to timing misalignment between the pulse carver and the optical phase modulator. Such misalignment was experimentally demonstrated to significantly degrade the system performance [2, 3]. For example, Hoon et al showed in their 10-Gb/s experiment that bit error rate was deteriorated by more than two orders of magnitude for 25-ps timing misalignment [2]. Despite the large number of experiments results, however, there has been no systematical study to investigate the characteristics of timing misalignment induced distortion (TMID) and how it impacts system performance at the receiver. Those kinds of studies would be helpful to make the receiver more resilient to TMID through post-detection signal processing. In this paper, we will investigate the influence of TMID on the performance of RZ-DPSK systems. Our study shows that such distortion would lead to ISI of the demodulated data and is sensitive to input data pulse shape.

II. System model
Fig. 1 depicts the general system model in the studies. RZ-DPSK signal \( E_s(t) \) is generated by feeding a carved optical pulse train \( E_{puls}(t) \) into a data-driven optical phase modulator (PM). Different misaligned time between the pulse carver and the phase modulator is emulated by the delay line before PM. Fiber transmission link is modeled as a single-input, two-output setup [4]. The initial RZ-DPSK signal \( E_s(t) \) is split into two orthogonal polarization modes with \( \gamma \) being the relative power in the fast principle state of polarization. \( h_x(\gamma^{1/2}E_s(t)) \) and \( h_y((1-\gamma)^{1/2}E_s(t)) \) denote the channel mapping of two polarization modes. By neglecting Gordon-Mollenauer effect (nonlinear phase noise), optical noises from optical amplifiers are modeled as independent white Gaussian intensity noise \( n_{\text{opt}}(t) \) and \( n_{\text{opt}}(t) \). The optical noise is suppressed by the following optical bandpass filter (OBPF). The output of the transmission fiber \( E_{\text{out}}(t) \) and \( E_{\text{out}}(t) \) are demodulated by a delay interferometer (DI). Two orthogonal modes from destructive/constructive ports of DI are then square-law detected and summed up to yield sum/difference port currents \( I_{\text{sum}}(t) \) and \( I_{\text{dif}}(t) \) of the balanced detector. Thermal noises \( n_{\text{sum}}(t) \) and \( n_{\text{sum}}(t) \), which corrupt \( I_{\text{sum}}(t) \) and \( I_{\text{dif}}(t) \) respectively, are also modeled as independent white Gaussian noise. At last, the difference of \( I_{\text{sum}}(t) \) and \( I_{\text{dif}}(t) \) is filtered by an electrical filter (EF), sampled, and decoded.

![Fig. 1. System model in the analysis](image-url)
III. Timing misalignment induced distortion

We neglect other impairments in the analysis to clarify the influence of TMID on the performance of RZ-DPSK systems. Assume optical RZ pulse train $E_{\text{pulse}}(t)$ is:

$$E_{\text{pulse}}(t) = \sum_{k=-\infty}^{\infty} s(t - kT - t_0)$$

where $s(t)$ and $t_0$ are the pulse profile in one bit slot and the misaligned time, respectively. For NRZ data pulse, it will be shown later that when timing is misaligned, system performance would strongly depend on NRZ pulse shape. Therefore, raised cosine shaped NRZ pulse, with $\alpha$ controlling the edge sharpness, is assumed:

$$V_{\text{phase}}(t) = \begin{cases} 
\frac{1}{2} \left[ \left| 1 - \sin \left( \frac{\pi}{\alpha T} \left( t - \frac{T}{2} \right) \right) \right| \right] & 0.5(1-\alpha)T \leq \left| t \right| \leq 0.5(1+\alpha)T \\
1 & 0 \leq \left| t \right| \leq 0.5(1-\alpha)T \\
0 & \left| t \right| \geq 0.5(1+\alpha)T 
\end{cases}$$

The modulated signal $E_r(t)$ can be obtained from equation (1) and (2) as:

$$E_r(t) = E_{\text{pulse}}(t)e^{\phi(t)}$$

Here:

$$\phi(t) = \sum_{n=0}^{\infty} a_n V_{\text{phase}}(t - nT)\pi$$

$a_n$ is the logical data. In the transmission, the signal is split into two orthogonal polarization modes $E_r(t)$ and $(1-\gamma)E_r(t)$. Optical filter bandwidth is assumed to be sufficiently wide so that the signal can pass through without distortion. At the receiver, the signal is demodulated. The sum/difference port currents of the photodetector $I_{\text{sum}}(t)$ and $I_{\text{dif}}(t)$ can be derived as:

$$I_{\text{sum}}(t) = R \left| 0.5E_{s}(t) + 0.5E_{s}(t-T) \right|^2 = R \left| 0.5E_{\text{pulse}}(t) \right|^2 A(t)$$

$$I_{\text{dif}}(t) = R \left| 0.5E_{s}(t) - 0.5E_{s}(t-T) \right|^2 = R \left| 0.5E_{\text{pulse}}(t) \right|^2 B(t)$$

where $A(t)$ and $B(t)$ in equation (5) are:

$$A(t) = \sum_{m=\infty}^{\infty} f_A(t-(m+1)T, b_{m-1}, b_m), \quad B(t) = \sum_{m=\infty}^{\infty} f_B(t-(m+1)T, b_{m-1}, b_m)$$

with the values depicted in Table 1 and $r=A, B$.

<table>
<thead>
<tr>
<th>$(b_{m-1}, b_m)$</th>
<th>(0 0)</th>
<th>(0 1)</th>
<th>(1 0)</th>
<th>(1 1)</th>
</tr>
</thead>
<tbody>
<tr>
<td>$f_A(t-(m+1)T, b_{m-1}, b_m)$</td>
<td>4</td>
<td>$4\cos^2(V_{\text{phase}}(t)\pi/2)$</td>
<td>$4\sin^2(V_{\text{phase}}(t)\pi/2)$</td>
<td>$4\sin^2(V_{\text{phase}}(t)\pi)$</td>
</tr>
<tr>
<td>$f_B(t-(m+1)T, b_{m-1}, b_m)$</td>
<td>0</td>
<td>$4\sin^2(V_{\text{phase}}(t)\pi/2)$</td>
<td>$4\cos^2(V_{\text{phase}}(t)\pi/2)$</td>
<td>$4\cos^2(V_{\text{phase}}(t)\pi)$</td>
</tr>
</tbody>
</table>

If electrical filter at the receiver is included, the received electrical signal is:

$$r(t) = (I_{\text{sum}} - I_{\text{dif}}) \oplus h(t)$$

where $h(t)$ is the impulse response of electrical filter and $\oplus$ denotes convolution operation. Fig. 2(a), (b) and (c) depict $0.5E_{\text{pulse}}(t)^2$, $A(t)$-$B(t)$, and $I_{\text{sum}}(t)$-$I_{\text{dif}}(t)$ within $-T \leq t - (m+1)T \leq T$, respectively, for 10-Gb/s Gaussian-shaped RZ pulses with 33% duty cycle and raised-cosine shaped data pulses with $\alpha=0.8$. Misaligned time $t_0$, and $R$ are -35 ps and 1, respectively. From the figure, it is shown that TMID with negative $t_0$ would cause ISI, from previous bit $b_{m-1}$. Similarly, when $t_0$ is positive, the present bit $b_m$ is corrupted by posterior bit $b_{m-1}$. Because $f(t-(m+1)T, b_{m-1}, b_m)$ is strongly dependent on the parameter $\alpha$, TMID would be sensitive to the input data pulse shape, which would be discussed in the next section.
IV. Results
In this section, Monte Carlo simulations were performed to numerically investigate the influence of TMID. A 10G/s Gaussian-shaped pulse train with 33% duty cycle and raised cosine NRZ data with \( \alpha = 0, 0.4, 0.8 \), were assumed. Without other distortions such as polarization mode dispersion (PMD) included, \( h_x(\gamma^{1/2}E_s(t)) \) and \( h_y((1-\gamma)^{1/2}E_s(t)) \) are \( \gamma^{1/2}E_s(t) \) and \( (1-\gamma)^{1/2}E_s(t) \) respectively. System noise modeling was dependent on system operation region: 1: OSNR limited; 2: Thermal noise limited. In this paper, we adopted the second case to be consistent with other works [2, 3]. Optical filter was Gaussian shaped with filter bandwidth of 50GHz. Electrical filter was 4th-order Butterworth shaped with optimized bandwidth in the absence of TMID (~12 GHz). The sampling phase is assumed at the center of eye. The performance is evaluated in terms of the required power to achieve an output bit error ratio of 1e-4, which can be corrected below 1e-15 using strong forward error correction (FEC). Fig. 3 depicts power penalty of the received signal versus \( t_0 \) for \( \alpha = 0, 0.4 \) and 0.8. From Fig. 3, we can find that when \( t_0 \) exceed \( \pm 25 \) ps for \( \alpha = 0, 0.4 \) and \( \pm 15 \) ps for \( \alpha = 0.8 \), power penalty increases rapidly. The timing misalignment ranges of -25 ps to +25 ps for \( \alpha = 0, 0.4 \), or -15 ps to +15 ps for \( \alpha = 0.8 \) are referred as tolerance range. Slightly asymmetric profile in the figure is due to asymmetric impulse response of electrical filter. The figure shows that the power penalty significantly depends on parameter \( \alpha \). The reason is twofold. First, the temporally less confined NRZ pulses with larger \( \alpha \) value would experience more ISI. Second, when timing is misaligned, much energy leaks into the phase transition region, which enhances performance sensitivity to \( \alpha \). Generalized as ISI, TMID is promising to be compensated by electronic equalization.

V. Conclusions
In conclusion, we found that the distortions caused by pulse carver/data modulator misalignment in RZ-DPSK systems led to ISI from the previous/posterior bit when timing misalignment was negative/positive. Such distortions are strongly dependent on the input data pulse shape. The misalignment tolerance range would reduce from 25 ps to 15 ps in 10 Gb/s systems when \( \alpha \) increases from 0 to 0.8. The results provide the potential solution for TMID compensation by employing electronic equalization.

References: