A Baud-Rate Sampled Coherent Transceiver with Digital Pulse Shaping and Interpolation

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Abstract: We demonstrate a baud-rate sampled 3.125GBd QPSK digital coherent transceiver employing pulse shaping and interpolation. Compared to an over-sampled receiver, a fixed penalty of 1.4dB is measured being independent of sampling phase or chromatic dispersion.
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1. Introduction

Digital receivers that use symbol-spaced sampling have long been of interest due to their low complexity in the mixed-signal and digital domains [1,2]. Reducing the receiver sampling rate is particularly appealing in optical communications due to the high symbol rates involved [3-5]. However, a fully symbol-spaced receiver comes with significant compromises in terms of performance. The most significant impairments are considered to be the sensitivity to sampling phase, and the effect of aliasing out of band signal and noise into the baseband. Additional penalty occurs due to aliasing with signals that have undergone chromatic dispersion. In [3] it was demonstrated that it is possible to use analogue low pass filters to reduce the effects of aliasing and therefore improve dispersion tolerance. In [4], it was proposed to use analogue low pass filters in combination with an oversampling rate of between 1 and 2, and interpolation to 2 samples per symbol. This enabled the use of a T/2 spaced equalizer for all sampling rates, and the penalty was reduced to approximately 1.5dB when sampling at the symbol rate. In [5], sampling at the symbol rate combined with a novel symbol spaced equalizer and MLSE achieved a reduction in penalty of 1.5dB with respect to conventional T-spaced processing.

In this work, we propose the use of a digital transmitter to limit the bandwidth of the transmitted signal to the Nyquist frequency, and examine the influence of the choice of interpolating filter on sampling phase sensitivity. We found that by using an interpolating filter with a truncated sinc response, it was possible to reduce the mean penalty with respect to conventional T-sampled and processed systems by 5.3dB.

2. Filtering using Truncated Sinc Pulses

In this work we have used truncated sinc filters both for band-limiting pulse shaping at the transmitter, and for band-limited interpolation at the receiver. The sinc function would be the ideal for both band-limited pulse shaping and band-limited interpolation. However, this function is impractical for hardware implementation as it is both non-causal and of infinite duration. To counteract these problems, we have used a symmetric truncation of the pulse that - while not ideally band-limited - provides superior anti-aliasing performance to other functions, such as piecewise-polynomial interpolating filters [6]. The truncated discrete-time sinc function used is given below in equation 1.

$$I_k = \frac{sin(k\pi/2)}{k\pi/2}; \text{ where } -L \leq k \leq L \tag{1}$$

Here, I_k are the coefficients with index k and a truncated length $2L+1$. These coefficients describe the taps in a low-pass filter with 2x oversampling. It is worth noting that for this function, L of the coefficients are zero, leading to L non-trivial calculations per symbol. While this represents a single phase of the interpolating function, the subsequent T/2 spaced equalizer recovers the (unknown) sampling phase without such a significant penalty as is seen with a T-spaced system. When this filter is used for interpolation by an integer factor, up-sampling is must first be done by forming the Kronecker tensor product of the signal and an appropriate zero-padded impulse. In the case of 2x up-sampling, this means that every other input sample to the filter is zero.

3. Experimental Setup

The experimental setup used to generate, transmit and receive digitally pulse-shaped QPSK at 3.125GBd is shown below in figure 2. Two channels of an arbitrary waveform generator (AWG) were used to output a pair of decorrelated zero-mean pseudo-random binary sequence (PRBS) signals at 3.125GBd with 3.86x oversampling (resulting in a sampling rate of 12GSa/s) and truncated sinc filtering (with filters 201 taps long) to provide band-limiting with a cut-off frequency of 1.5625GHz. The AWG had an analogue bandwidth of 5.6GHz. This symbol rate was chosen to be a native rate of the oscilloscope analogue-to-digital convertors (ADCs), while being a low enough rate such that the roll-off of the AWG output provided more than 25dB of image rejection. These signals were then used to directly drive the two arms of an I-Q modulator, using an optical carrier with a linewidth of approximately

100kHz. For back-to-back characterization, the signal was then attenuated, and coupled to a variable power additive white Gaussian noise (AWGN) source in order to control the received optical signal to noise ratio (OSNR). When transmission was performed, a single-span recirculating loop was used with 80.2km of standard single-mode fiber as described in [7]. The signal was then detected using a dual-polarization coherent receiver that had four balanced P-I-N photodiodes with a bandwidth of 30GHz, and a local oscillator with 100 kHz linewidth. The RF outputs of the optical receiver front end were filtered using four 7.5GHz bandwidth 5th order Bessel low pass filters to reduce noise aliasing. These four analogue signals were then captured using a real-time oscilloscope at a rate of either 3.125GSa/s (T-sampled) or 6.25GSa/s (T/2-sampled) and processed offline using Matlab. Receiver sampling was asynchronous and of random phase.

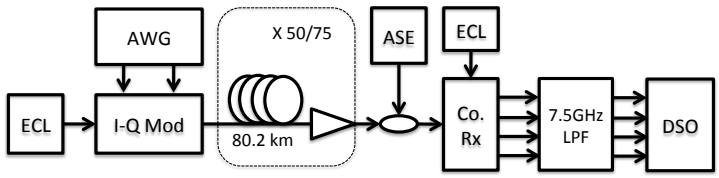


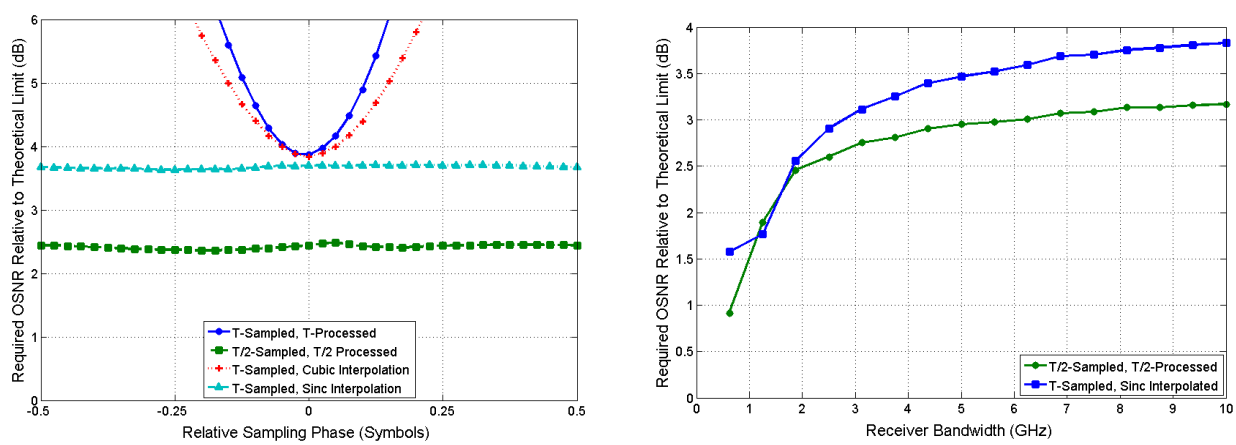
Figure 1. Experimental setup.

4. Receiver Digital Post-Processing

The digital receiver processing for QPSK signals was performed as follows. The signals were initially de-skewed in the frequency domain to compensate for different path lengths in the optical receiver and normalized to compensate for the different responsivities of the photodiodes. Where necessary, chromatic dispersion was compensated using an FIR filter, designed in the frequency domain with 81 taps. T-sampled signals were then interpolated to 2 samples per symbol using either a sinc interpolator (as described in section 2) or a piecewise cubic interpolator [6], before being equalized with a T/2 spaced constant modulus algorithm (CMA) equalizer with least mean squares (LMS) updating. T/2-sampled signals were equalized directly with the T/2 CMA without any interpolation. The intra-dyne frequency (which was tuned to below 70MHz) was then recovered in the frequency domain, and carrier recovery was performed using the Viterbi and Viterbi algorithm before symbol estimation and bit error rate calculation.

5. Results and Discussion

A sweep of OSNR was performed, with each measurement capturing $5x10^5$ symbols. The captured signals were processed offline using Matlab, with variation in sampling phase emulated by a time shift in the frequency domain over the entire sequence. For the purpose of illustration, all measurements of sampling phase have been normalized to the optimum. However, in practice, sampling was asynchronous and phase was therefore random.



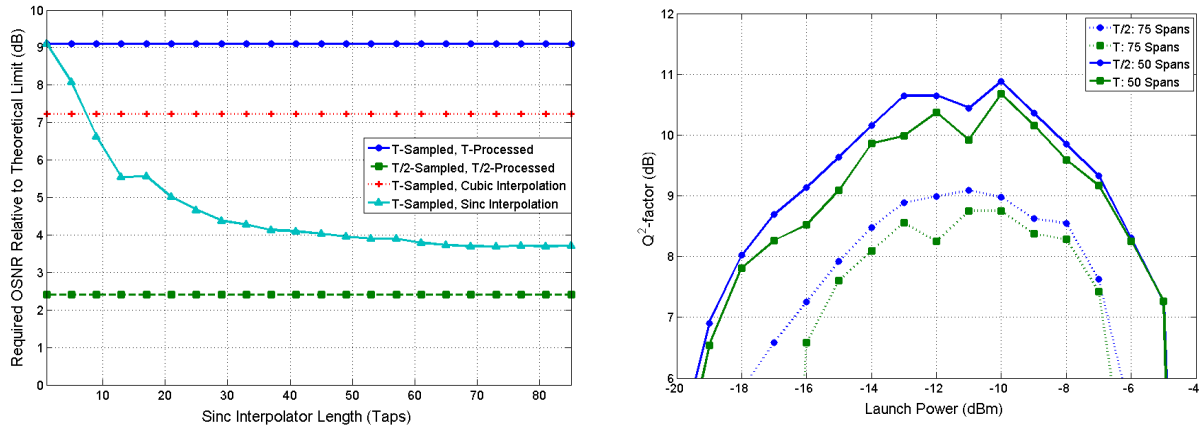
(a) (b)

Figure 2. (a) Back-to-back experimental characterization of sampling phase sensitivity with different receiver configurations. (b) Simulation of the effect of receiver bandwidth on sensitivity. In both cases, a sinc interpolator of 85 taps was used and required OSNR was calculated for a BER target of 10^{-3} relative to the theoretical limit for 3.125GBd QPSK.

We have plotted the variation in required OSNR (ROSNR) penalty (compared to the theoretical limit) to achieve a bit error rate of 10^{-3} with sampling phase in figure 2(a). As expected, the T/2-sampled system has no penalty with respect to sampling phase as the equalizer can recover the signal timing phase. Performance of the T-sampled system with T-spaced equalization exhibits a significant dependence on sampling phase: in the case of optimum sampling phase, there is a penalty of 1.5dB with respect to T/2 sampling; while for a sampling phase within +/- 0.2

of the optimum, there is an additional penalty up to 3dB. A T-sampled system with piecewise cubic interpolation and T/2 equalization performs slightly better: the penalty at the optimum sampling phase remains at 1.5dB, while an additional penalty of 3dB allows a sampling phase within +/- 0.25 of the optimum. By using a sinc interpolator with 85 taps, we may eliminate the dependence of performance on receiver sampling phase. While the T-sampled and sinc interpolated receiver exhibits a penalty with respect to the T/2-sampled system of 1.4dB; we believe that this is due to the excess analogue receiver bandwidth.

To investigate this possibility, we simulated a back-to-back system with variable bandwidth for both T-sampling and T/2-sampling. We found that as the receiver bandwidth is increased, the receiver sensitivity is reduced due to noise aliasing. This effect is more pronounced in the T-sampled system due to the reduced sampling bandwidth, which increases the alias bandwidth. The result of these simulations is shown in figure 2 (b).



(a) (b)

Figure 3. (a) Back-to-back characterization of variation in required OSNR penalty over all sampling phases with length of sinc interpolator. (b) Transmission performance over 50 and 75 spans with T/2-sampling and processing, and T-sampling with 85 tap sinc interpolator.

We then investigated the influence of the interpolating sinc filter on receiver sensitivity. This is shown in figure 3(a), with the results obtained for T-sampling with T-processing, T-sampling with piecewise cubic interpolation, and T/2-sampling with T/2-processing for comparison. It is worth noting that the truncated sinc interpolator outperforms the piecewise cubic interpolator for all but the shortest filters, while improvement in performance saturates at approximately 65 taps.

Transmission performance was then characterized for the T/2-sampled and processed receiver, and compared to the T-sampled receiver with an 85 tap interpolating filter. These results are presented in figure 3(b). We were able to fully compensate for chromatic dispersion, which introduced a maximum of 5.3 symbols of delay for 50 spans and a maximum of 10.0 symbols of delay for 75 spans.

6. Conclusions

We have demonstrated the ability of truncated sinc pulse interpolation to improve the performance of coherent optical transceivers that are sampled at the symbol rate. By using digital pulse shaping at the transmitter and sinc interpolation at the receiver, we have reduced the mean penalty over all sampling phases with respect to T-sampling and T-processing by 5.3dB and the penalty with respect to T-sampling with piecewise cubic interpolation by 3.5dB. Since the proposed technique is independent of symbol rate, it is anticipated that numerous systems will benefit from the reduced power requirements and sampling rates from using truncated sinc interpolation of signals with rectangular spectra. Key future work will be to investigate the performance of this transceiver design for high-level QAM to determine its suitability for relaxing the hardware constraints of systems operating at 400Gb/s and 1Tb/s.

7. Acknowledgements: The authors thank EPSRC and the EU PIANO+ project CRITICAL for funding this work.

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