PLL-Free Synchronous QPSK Receiver Concept with Digital I&Q Baseband Processing

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Abstract *This concept employs a digital baseband signal processing scheme after electronic demultiplexing. An IF linewidth tolerance of up to 10^{-3} times the QPSK symbol rate is predicted. Commercially available DFB lasers shall therefore suffice.*

Introduction

QPSK-based multilevel modulation formats promise ultimate performance of upgraded or newly built fiber links. DQPSK receivers [1,2] suffer from a 2.3 dB sensitivity penalty compared to synchronous QPSK receivers and are only ~1dB more sensitive than ideal ASK receivers. A PLL-based carrier recovery for synchronous QPSK transmission would fail in combination with DFB lasers because the product of laser linewidth times loop delay is too large [3,4]. An analog, phase noise tolerant feedforward carrier recovery concept has been presented in [5] for a coherent in-phase and quadrature (I&Q), synchronous QPSK receiver. A record IF linewidth tolerance of up to 10^{-3} times the symbol rate was predicted to be tolerable with additional polarization division multiplex. However, digital processing of QPSK signals would be very useful, especially for fully electronic polarization tracking in a polarization diversity coherent receiver or the equalization of PMD and CD distortions. Here we show how the carrier can indeed be recovered digitally while maintaining the largest phase noise tolerance.

Operation principle

In order to keep the overall electrical bandwidth as low as possible we assume an I&Q receiver (Fig. 1) with an intermediate frequency (IF) near or at zero. The in-phase and quadrature (I&Q) signals can be understood to be real and imaginary parts Re $\underline{X}$, Im $\underline{X}$ of a complex IF signal $\underline{X} \propto \underline{c} \cdot e^{j\varphi'}$. This is the case if the path length difference $(l_{11} - l_{21}) - (l_{12} - l_{22})$ is a quarter wavelength. The transmitted data symbol $\underline{c}$ assumes the values $\pm 1 \pm j$, thereby transmitting 2 bits. Angle φ' is the difference phase between signal laser and local oscillator laser. It is not constant.

The clock can be recovered in an extra intensity modulation direct detection receiver, especially for RZ-coded QPSK, or from the electrical I&Q signals. After clock recovery the I&Q signals are sampled, digitized and demultiplexed 1:*M* to a low symbol rate where complex digital functions can be implemented in CMOS. The *M* data streams are processed in *M* modules (Mod.). The modules also have to communicate among each other, as will be explained. The *i*th and (*i*–*N*)th complex samples $\underline{X}(i)$, $\underline{X}(i-N)$ of the digitized IF signal $\underline{X}$ enter the *k*th module (Fig. 2), where $k = i \bmod M$, and *N* is a delay. $\underline{X}(i)$ is raised to the 4th power. Due to $\underline{c}^4 = -4$ the modulation is thereby eliminated. The resulting frequency-quadrupled carrier $\underline{X}^4 \propto -e^{j4\varphi'}$ must be filtered for SNR improvement. To do so, the I&Q signals $\mathrm{Re}\,\underline{X}^4$, $\mathrm{Im}\,\underline{X}^4$ are lowpass-filtered (LPF) because the IF is at or near zero.

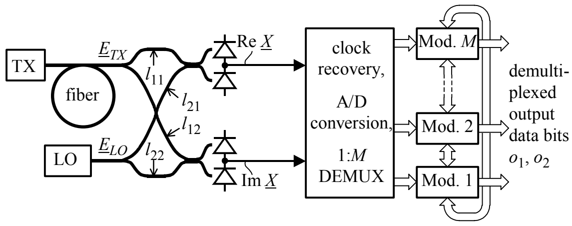

Fig. 1: I&Q coherent receiver concept with digital feedforward carrier recovery scheme for synchronous QPSK

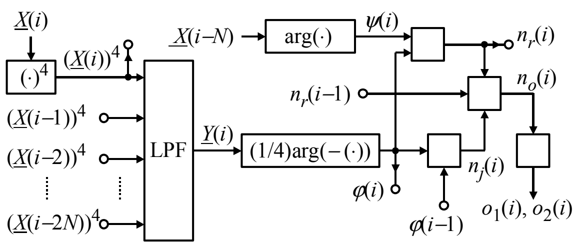

Fig. 2: Block diagram of kth digital signal processing module after demultiplexer. The inputs and outputs terminated by a circle may or must interface with other modules.

The phase angles of the quantities $\left(\underline{X}(i - mM)\right)^4$ with *m* = 0,1,2,..., which are available in the *k*th module, may differ a lot. Therefore it would not be appropriate to base the lowpass filtering only on them. Rather, a good filter may take the sum $\underline{Y}(i) = \sum_{m=0}^{2N} \left(\underline{X}(i-m)\right)^4$ of the 2*N*+1 most recent samples of the frequency-quadrupled carrier, most of which come from adjacent modules. The group delay in this filter equals *N* symbols. *N* = 3 or 4 is a fairly good choice.

The filter (LPF) also alters the phase angle. $\underline{Y} \propto -e^{j4\varphi}$ holds, where $4\varphi(i)$ would ideally be equal to $4\varphi'(i-N)$. The next step is to divide phase and frequency of $\underline{Y}$ by a factor of 4 in order to recover the carrier phase. The best choice is probably a 2D-lookup table which calculates carrier phase samples $\varphi(i) = (1/4)\arg(-\underline{Y}(i))$. Using a similar lookup operation, the phase angle $\psi(i) = \arg \underline{X}(i-N)$ of the received signal is obtained [3], with a delay equal to that experienced in the lowpass filter. For demodulation, an integer $n_r(i)$ which fulfills $n_r(i)\pi/2 \le \psi(i) - \varphi(i) < (n_r(i)+1)\pi/2$ is simply determined. It may be called a received quadrant number because $\underline{c}(i-N) = (1+j)e^{jn_r(i)\pi/2}$ holds for $|\varphi(i) - \varphi'(i-N)| < \pi/4$.

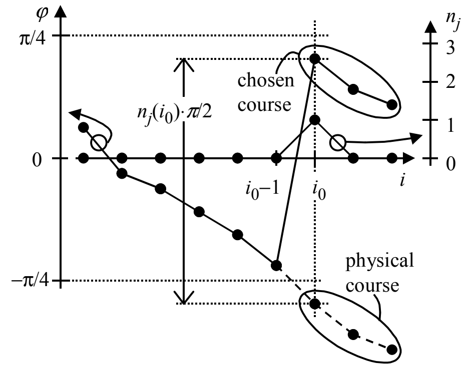

Fig. 3: Quadrant phase jump and its detection

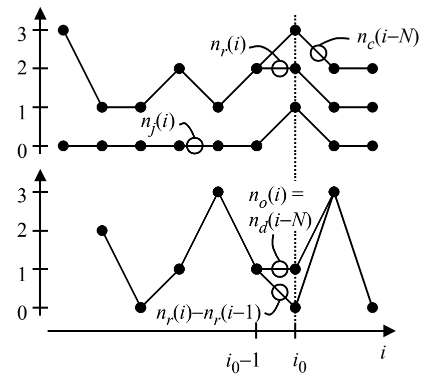

Fig. 4: Successful correction of quadrant phase jump

d_1, Re $\underline{c}$, o_1	d_2, Im $\underline{c}$, o_2	n_d, n_c, n_o
1	1	0
-1	1	1
-1	-1	2
1	-1	3

Table 1: Bipolar bits vs. quadrant number

Unfortunately, there is a 4fold ambiguity in the calculation of $\varphi(i)$, and $n_r(i)$ does therefore not contain all needed information. Choosing $\varphi(i)$ as close as possible to $\varphi(i-1)$ could solve the problem but is practically impossible because the correct quadrant can not be selected within one symbol duration. Rather, we propose that $|\varphi(i)| \le \pi/4$ be always chosen. For compensation the *k*th module must detect whether φ has jumped by an integer multiple of $\pi/2$. It determines a quadrant jump number $n_j(i)$ for this purpose. This is an integer which fulfills $|\varphi(i) - \varphi(i-1) - n_j(i)\pi/2| < \pi/4$. All quadrant numbers and operations are understood to be valid modulo 4 because the angle functions are periodic. Fig. 3 shows a quadrant phase jump at $i = i_0$ and its detection. A value $n_j(i_0) \ne 0$ indicates that all $n_r(p)$ with $p \ge i_0$ carry an unwanted offset $-n_j(i_0)$. It is not possible to correct all $n_r(p)$ accordingly because this would have to be done at the symbol rate.

As a remedy, a data quadrant number n_d is generated already at the transmitter side which represents two bipolar data bits d_1, d_2 (Table 1). The data quadrant number is differentially encoded to form an encoded quadrant number n_c with $n_c(i) = (n_d(i) + n_c(i-1)) \bmod 4$. This number defines the quadrant of the transmitted complex symbol $\underline{c} = \mathrm{Re}\,\underline{c} + j\,\mathrm{Im}\,\underline{c} = \pm 1 \pm j$ (Table 1). After transmission, one calculates an output quadrant number $n_o(i) = (n_r(i) - n_r(i-1) + n_j(i)) \bmod 4 = n_d(i-N)$. It is identical to the delayed data quadrant number, and it yields the output data bits $o_1(i)$, $o_2(i)$ (Table 1). These are equal to the delayed data bits $d_1(i-N)$, $d_2(i-N)$. The successful correction of the quadrant jump at $i = i_0$ is illustrated in Fig. 4.

Discussion

Each module processes only signals which are already available, and each processing step may include an arbitrary delay which just adds to the overall transmission delay but does not impede realtime operation. Yet the phase noise tolerance is as good as if all processing took place at the symbol rate. The function of the analog feedforward carrier recovery [5] is emulated, which indicates that the laser linewidth / symbol rate ratio can be up to $5 \cdot 10^{-4}$. If polarization division multiplex is included then the frequency-quadrupled carrier recovery signals can be added from both polarization branches. This doubles the SNR in the carrier recovery (which is common for both polarization branches), and a laser linewidth / symbol rate ratio of up to 10^{-3} becomes tolerable. At least 1/3 of this can be realized experimentally, as may be concluded from a comparison of [5] with the early experiment [6].

Conclusion

This digital feedforward carrier recovery scheme for QPSK is predicted to be extremely laser linewidth tolerant, just like its analog counterpart. As a consequence, synchronous QPSK becomes feasible even at 10 Gsymbols/s with commercially available DFB lasers and a digital coherent I&Q receiver.

References

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