Phase Noise-Tolerant Synchronous QPSK/BPSK Baseband-Type Intradyne Receiver Concept With Feedforward Carrier Recovery

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***Abstract*—Quadrature phase-shift keying (QPSK) is attractive to increase transmission lengths and capacity, especially when it is combined with polarization division multiplex. Baseband processing at the symbol rate allows to keep the required electronic bandwidth low. So far, external cavity lasers seemed to be indispensable for such transmission systems due to linewidth requirements. We propose a feedforward carrier recovery scheme based on regenerative intradyne frequency dividers, i.e., the well-known regenerative frequency divider is extended to process baseband in-phase and quadrature (I and Q) signals. An IF linewidth tolerance of up to 0.001 times the QPSK symbol rate is predicted, 2 decades more than for an optical phase locked loop with a realistic loop delay. This means that commercially available DFB lasers shall suffice for synchronous optical QPSK/BPSK transmission.**

***Index Terms*—Homodyne detection, optical fiber communication, phase-shift keying (PSK), quadrature phase-shift keying (QPSK), synchronous detection.**

I. INTRODUCTION

PHASE-SHIFT KEYING (PSK) means that the phase of a transmitted signal, here an optical carrier, is modulated by digital data. Optical PSK transmission systems feature a higher receiver sensitivity than intensity-modulated systems. They are attractive to increase transmission lengths while keeping the amplifier spacing fixed. Of particular interest are coherent receivers in which the carrier phase information is transmitted as a part of the modulated signal and used at the receiver for phase alignment of the in phase (I) and quadrature (Q) symbols. They exhibit highest sensitivity, allow to implement electrical rather than fully optical WDM demultiplex filters and allow to perform any required equalization of chromatic and polarization mode dispersion in the electrical domain. Quadrature PSK (QPSK) combined with polarization multiplex quadruples channel capacity over state-of-the-art intensity-modulated systems. This scheme is very attractive because it uses existing fiber plants most efficiently.

The synchronous downconversion of an optical signal is critically affected by phase noise of the lasers, and this problem is acknowledged and addressed in many publications on synchronous binary PSK (BPSK) and QPSK optical transmission [1]–[10], mostly by the use of external cavity lasers.

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A carrier recovery unit generates the required carrier at the receiver side even though the carrier is usually suppressed in a BPSK/QPSK signal. In a homodyne receiver the LO laser delivers the recovered carrier, and its frequency and phase are controlled by an optical phase-locked loop (OPLL). The frequency difference between S and LO lasers, called intermediate frequency (IF), is equal to zero in homodyne receivers. In receivers with nonzero IF, another downconversion is needed in the electrical part of the optical receiver, using an electrical carrier that is usually delivered with the help of an electronic PLL. A QPSK receiver with a purely electronic digital carrier recovery has been presented in [8] but that particular concept does not allow to use DFB lasers because it underlies laser linewidth restrictions similar to those of an OPLL. A QPSK receiver based on an OPLL with a realistic loop delay of $\sim$100 symbol durations tolerates a laser linewidth/symbol rate ratio of $< 5\cdot 10^{-6}$ [10]. The PLL delay may be minimized by keeping optical fiber lengths short. Moreover, an electrical PLL in the IF domain may be employed instead of an optical one. Even though, a group delay of several symbol durations in the loop is unavoidable. In any case, external cavity lasers are required. These are costly, space-consuming, can be tuned only slowly, and could impair system reliability. Even if a PLL could be implemented without parasitic delay the laser linewidth / symbol rate ratio would need to be $< 1.5 \cdot 10^{-4}$, see [10, Fig. 1c].

On the other hand, a feedforward carrier recovery method has allowed the use of normal DFB lasers in a BPSK heterodyne transmission experiment [9]. The only feedback needed is inside a frequency divider. The corresponding path delay is so much shorter than the phase measurement delay in a PLL-based carrier recovery, leave alone the unwanted PLL delay. The feedforward carrier recovery method is, therefore, intrinsically more linewidth tolerant than PLL-based schemes. However, at 10 (and at 40) Gbaud baseband processing is desirable because a heterodyne receiver would require too large a bandwidth. High laser linewidths, baseband processing, and synchronous QPSK/BPSK detection have appeared to be incompatible so far. The purpose of this paper is to show how this impediment can be overcome by a novel phase noise-tolerant synchronous QPSK/BPSK intradyne receiver concept with feedforward carrier recovery. It can circumvent the inherent linewidth and bandwidth problem of any PLL-based carrier recovery. The predicted linewidth tolerance is so large that commercially available DFB lasers may be employed. It may be expected to facilitate a technical and commercial success of synchronous optical QPSK transmission systems. A simple electronic polarization control algorithm is also proposed.

II. Principle of Feedforward Carrier Recovery Using I and Q Signals

A coherent optical receiver measures in a photodetector the interference between a modulated signal laser and an unmodulated local oscillator laser. The interference term in the photocurrent is proportional to one quadrature of the received electrical field. Fig. 1. shows a particular coherent data transmission system with a QPSK signal laser (S), a fiber span, and a receiver front end consisting of the local oscillator (LO) laser, an optical coupler and four identical photodetectors. The receiver features two branches, in which the signal field $\underline{E}_S \propto \underline{c}e^{j(\omega_s t+\varphi_S)}$ and the LO field $\underline{E}_{LO} \propto e^{j(\omega_{LO}t+\varphi_{LO})}$ are superimposed in phase and in quadrature, respectively. This is the case if the path length difference $(l_{11}-l_{21})-(l_{12}-l_{22})$ is a quarter wavelength. The coupler arrangement is called an optical 90° hybrid [11].

All expressions and equations following in this Section II refer to electrical signals. The photocurrent differences from the two diode pairs may be called by definition $\mathrm{Re}\underline{X}$ and $\mathrm{Im}\underline{X}$, because they represent a complex IF signal

$$\underline{X} = \mathrm{Re}\underline{X} + j\mathrm{Im}\underline{X} \propto \underline{c}\cdot e^{j\varphi'}. \tag{1}$$

It is proportional to the electric field of the received optical signal. Here, $\underline{c} = b_1 + jb_2$ is a 2-bit QPSK data symbol, $b_{1,2} = \pm 1$ are binary bits, and $\varphi' = (\omega_s - \omega_{LO})t + (\varphi_S - \varphi_{LO})$ is the difference phase between signal laser and LO laser.

This particular receiver front end may be used in an intradyne receiver. Like a heterodyne receiver, an intradyne receiver [8] features a nonzero intermediate frequency (IF) because it has no OPLL. The required bandwidth is that of a homodyne receiver since it operates in the baseband with a very low IF. A minimum of two phases is needed to avoid loss of information, i.e., the in-phase and quadrature (I and Q) signals $\mathrm{Re}\underline{X}$, $\mathrm{Im}\underline{X}$. A synchronous intradyne receiver features a carrier recovery which delivers the electronic carrier needed to convert the detected I and Q signals from a small but generally nonzero IF into two data streams.

Fig. 2 shows the proposed feedforward carrier recovery scheme in the electrical part of an intradyne QPSK receiver. It is connected to Fig. 1 as defined by $\underline{X} = \mathrm{Re}\underline{X} + j\mathrm{Im}\underline{X}$. Each thick line in Fig. 2 represents the flow of a complex signal which is represented by at least two real signals (real and imaginary parts).

The crucial signal processing task is the accurate carrier recovery. It works as follows: According to $\underline{c}^4 = -4$ the modulation is cancelled out if $\underline{X}$ is raised to the fourth power. A frequency-doubled I and Q signal $\underline{X}^2$ is obtained by the operations

$$\begin{aligned}\mathrm{Re}\underline{X}^2 &:= \mathrm{Re}^2\underline{X} - \mathrm{Im}^2\underline{X}\\ \mathrm{Im}\underline{X}^2 &:= 2\mathrm{Re}\underline{X}\cdot\mathrm{Im}\underline{X}\end{aligned} \tag{2}$$

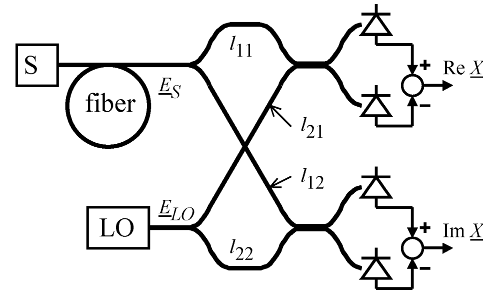

Fig. 1. Coherent data transmission system with QPSK modulated transmitter (S) and a 2-phase intradyne receiver front end (local oscillator LO, optical couplers, four identical photodetectors).

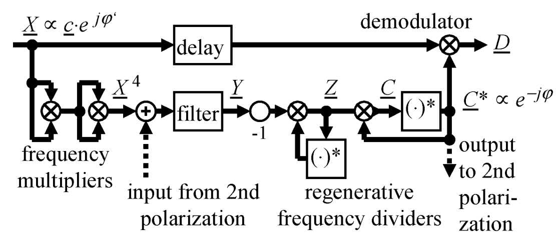

Fig. 2. Feedforward carrier recovery with regenerative frequency divider in the electrical part of an intradyne QPSK receiver.

where := is an assignment in a signal flow diagram. This is repeated to obtain

$$\begin{aligned}\mathrm{Re}\underline{X}^4 &:= \mathrm{Re}^2\underline{X}^2 - \mathrm{Im}^2\underline{X}^2\\ \mathrm{Im}\underline{X}^4 &:= 2\mathrm{Re}\underline{X}^2\cdot\mathrm{Im}\underline{X}^2\,.\end{aligned} \tag{3}$$

The operations in (2) and (3) can be built with Gilbert multiplier cells [12] and monolithically integrated with the signal combiners on a single chip. Data rates of 10 Gbaud could be achieved today with available SiGe bipolar technology, and 40 Gbaud could be reached in the future. The frequency multiplication operations do not need more bandwidth than the data path itself. The frequency multipliers simply must be able to process phase jumps from bit to bit. While these are multiples of $\pi/2$ initially, they are multiples of π after the first frequency doubler, and have disappeared after the second frequency doubler. The IF is quadrupled but since it is low compared to the symbol rate the required processing bandwidth stays unchanged. This is also true if the QPSK symbols are sent as RZ pulses. In that case the output signal $\underline{X}^4$ of the second frequency doubler will be the frequency-quadrupled IF signal, sampled at the symbol rate.

Then the frequency-multiplied carrier $\underline{X}^4 \propto -e^{j4\varphi'}$ is bandpass-filtered. In the intradyne case this means that both components $\mathrm{Re}\underline{X}^4$, $\mathrm{Im}\underline{X}^4$ are low-pass filtered. Since there is no highpass filter involved the intradyne IF may be positive, negative or exactly zero. RZ-induced sampling of the frequency-quadrupled IF signal will also be eliminated by the low-pass filter.

Filtering also removes broadband noise but it alters the phase angle. In complex representation, the filtered signal is $\underline{Y} \propto -e^{j4\varphi}$ where φ ideally would be identical to φ'. In the next step the filtered signal needs to be frequency-divided by 4. This is possible by implementing the complex operations

$$\underline{Z} = (-\underline{Y})^{\frac{1}{2}}, \quad \underline{C} = \underline{Z}^{\frac{1}{2}}. \tag{4}$$

The term $\underline{C} \propto e^{j\varphi}$ is the recovered complex carrier. It is straightforward to obtain $\underline{C}$ in heterodyne receivers with high IF: The filtered signal $\underline{Y}$ is represented by a single quadrature $\mathrm{Re}\underline{Y}$ which is passed through two cascaded divide-by-2 frequency dividers to obtain $\mathrm{Re}\underline{C}$, and in a similar manner $\mathrm{Im}\underline{C}$. For the baseband case, static frequency dividers would not work because they assume only two discrete phase states in each quadrature. Such a coarse phase quantization is unacceptable.

But there exist regenerative frequency dividers [13], [14]. If a heterodyne input signal $\cos 2\omega t$ is connected to one input of a multiplier, and the multiplier output signal is connected to the other multiplier input, then the multiplier output signal has a halved frequency (Fig. 3). To understand the function of this simple model consider the equation

$$2\cos 2\omega t \cos(\omega t - \alpha) = \cos(\omega t + \alpha) + \cos(3\omega t - \alpha). \tag{5}$$

The second term of the multiplication result at the right-hand side is eliminated by low-pass filtering. The multiplier needs a gain of 2. The feedback path from the output to the second multiplier input may introduce a phase delay 2α. This converts the remaining output signal $\cos(\omega t+\alpha)$ into the required second multiplier input signal $2\cos(\omega t - \alpha)$. Low-pass filtering of this feedback signal is imperative to avoid nonlinear distortions.

To solve the frequency division problem in an intradyne receiver we propose to extend this heterodyne concept by implementing what may be called regenerative intradyne divide-by-2 frequency dividers. Two divider stages are required, which operate by the assignments

$$\underline{Z} := (-\underline{Y}) \cdot \underline{Z}^*, \quad \underline{C} := \underline{Z} \cdot \underline{C}^*. \tag{6}$$

In the second divider stage, for example, the solution of $\underline{C} = \underline{Z} \cdot \underline{C}^*$ for $\underline{Z} = e^{j2\omega t}$ is $\underline{C} = \pm e^{j\omega t}$, and this operation is obviously a frequency division by 2. For the implementation of $\underline{C} := \underline{Z} \cdot \underline{C}^*$ the two quadrature signals are being assigned the values

$$\begin{aligned}\mathrm{Re}\underline{C} &:= \mathrm{Re}\underline{Z} \cdot \mathrm{Re}\underline{C} + \mathrm{Im}\underline{Z} \cdot \mathrm{Im}\underline{C} \\ \mathrm{Im}\underline{C} &:= \mathrm{Im}\underline{Z} \cdot \mathrm{Re}\underline{C} - \mathrm{Re}\underline{Z} \cdot \mathrm{Im}\underline{C}\,.\end{aligned} \tag{7}$$

The traditional regenerative frequency divider works without amplitude stabilization because there is a dynamic balance between low-pass behavior and harmonic distortions, and harmonic distortions do not matter. In the regenerative intradyne frequency divider the lowest frequencies experience no significant low-pass filtering, and distortions degrade the phase accuracy. It is therefore necessary to stabilize the amplitude $|\underline{C}|$. Very simply, the associated multipliers and linear combiners may be slightly overdriven. Else the equation $\underline{C} := \underline{Z} \cdot \underline{C}^*$ may be modified by an amplitude control term, for example

$$\underline{C} := \underline{Z} \cdot \underline{C}^* \cdot \int \left(1 - |\underline{C}|^2\right) dt. \tag{8}$$

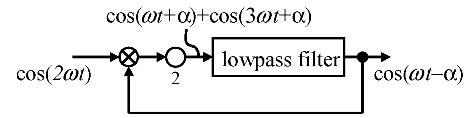

Fig. 3. Simplified mathematical model of a conventional (heterodyne) regenerative frequency divider.

In the time-discrete regime (only there!) the phase of $\underline{C} := \underline{Z} \cdot \underline{C}^*$ is not asymptotically stable. This can be easily verified by investigating the equation $\underline{C}_n := 1 \cdot \underline{C}_{n-1}^*$ where adjacent samples with indices $n-1, n$ are one symbol duration T apart: Unless $\underline{C}_{n-1} = \pm 1$ there will be a neverending oscillation. Therefore our simulations have been carried out with

$$\underline{C}_n := \left(\frac{1}{2}\right)(\underline{A}_n + \underline{A}_{n-1}), \quad \underline{A}_n := \underline{Z}_n \cdot \underline{C}_{n-1}^*, \tag{9}$$

and an amplitude control. Compared to $\underline{C}_n := \underline{Z}_n \cdot \underline{C}_{n-1}^*$ a dissipative behavior has been implanted in (9) which ensures phase stability. It also helps to model a realistic low-pass behavior. Eqn. $\underline{C}_n := (1/2)(\underline{A}_n + \underline{A}_{n-1})$ means that the feedback delay is about half a symbol duration. This can be considered as fairly realistic at 10 Gbaud, and is much shorter than realistic PLL delays.

To complete the required signal processing the complex IF signal $\underline{X}$ must be delayed by the group delay that is suffered in the carrier recovery branch. This delay equalization minimizes phase noise influence. The demodulated data signal is

$$\underline{D} := \underline{X} \cdot \underline{C}^* \propto \underline{c} \cdot e^{j\Delta\varphi} \tag{10}$$

where $\Delta\varphi = \varphi' - \varphi$ is the residual phase error. The decision variables of the two quadratures are $\mathrm{Re}\underline{D}$, $\mathrm{Im}\underline{D}$.

A similar setup is possible for 3-phase IF signals [15], which shall be outlined in brevity: The 3-phase signals x_1, x_2, x_3 are generated in a coherent receiver according to Fig. 4 which features a fully symmetric 3×3 coupler and three identical photodetectors. They could also be generated from the 2-phase IF signals $\mathrm{Re}\underline{X}$, $\mathrm{Im}\underline{X}$ by the operations

$$x_1 = \mathrm{Re}\underline{X}, \quad x_{2,3} = -\left(\frac{1}{2}\right)\mathrm{Re}\underline{X} \pm \left(\sqrt{\frac{3}{2}}\right)\mathrm{Im}\underline{X}. \tag{11}$$

The reverse operation is

$$\begin{aligned}\mathrm{Re}\underline{X} &= \frac{(2x_1 - x_2 - x_3)}{3} \\ \mathrm{Im}\underline{X} &= \frac{(x_2 - x_3)}{\sqrt{3}}.\end{aligned} \tag{12}$$

This makes it clear that a 2-phase signal processing unit can be combined with a 3-phase optical front end.

To keep the hardware effort low, 3-phase signals should be pre-conditioned to have a vanishing sum, for example by starting with

$$u_1 = x_1 - x_2,\ u_2 = x_2 - x_3,\ u_3 = x_3 - x_1 \tag{13}$$

rather than with x_1, x_2, x_3. This operation introduces a constant phase shift which is of no importance.

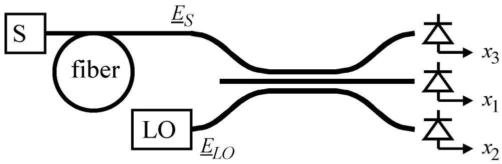

Fig. 4. Coherent data transmission system with QPSK modulated transmitter (S) and a 3-phase intradyne receiver front end (local oscillator LO, symmetrical 3 × 3 coupler, 3 identical photodetectors).

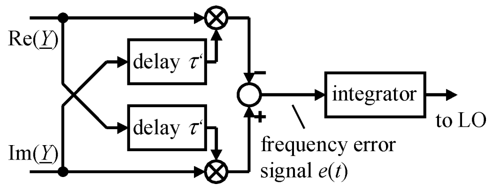

Fig. 5. A possible 2-phase intradyne frequency discriminator.

Complex conjugation of a 3-phase signal $\underline{U}$ requires an interchange of the components u_2 and u_3. Quite generally, the components of a 3-phase product $\underline{W} = \underline{UV}$ of two 3-phase signals $\underline{U}$, $\underline{V}$ can be calculated as

$$\begin{aligned} w_1 &= \left(\frac{2}{3}\right)(u_1v_1 + u_2v_3 + u_3v_2) \\ w_2 &= \left(\frac{2}{3}\right)(u_3v_3 + u_1v_2 + u_2v_1) \\ w_3 &= \left(\frac{2}{3}\right)(u_2v_2 + u_3v_1 + u_1v_3) \end{aligned} \tag{14}$$

if the component sum of at least one of $\underline{U}$, $\underline{V}$ is zero. These equations allow to implement 3-phase frequency doublers and regenerative frequency dividers. The achievable receiver sensitivities for the 2-and for the 3-phase implementation are identical.

In general the IF will be unequal zero, since there is no OPLL. In order to exploit a maximum of phase noise tolerance an automatic frequency control (AFC) system for the IF is needed. It may act on temperature or a current of the LO laser. A suitable intradyne frequency discriminator [16] is depicted in Fig. 5. It acts on the frequency-quadrupled, filtered signal $\underline{Y}$ and derives a frequency error signal

$$e(t) = \mathrm{Im}\left(\underline{Y}(t) \cdot \underline{Y}^*(t-\tau')\right) \tag{15}$$

where τ' is a delay time. For $\underline{Y}(t) = e^{j\Omega t}$ it equals

$$e(t) = \sin \Omega\tau'. \tag{16}$$

This error signal can be integrated and then passed on to the LO. Shown is a 2-phase circuit which implements $e(t) = -\mathrm{Re}\underline{Y}(t) \cdot \mathrm{Im}\underline{Y}(t-\tau') + \mathrm{Im}\underline{Y}(t) \cdot \mathrm{Re}\underline{Y}(t-\tau')$, in agreement with (15).

III. Modifications for Polarization Division Multiplex Operation

Polarization division multiplex can double the channel capacity by adding two more information bits $b_{3,4} = \pm 1$ in the second polarization mode. The fiber is modeled by an optical Jones matrix $\underline{\mathbf{J}}$. This leads to a Jones vector

$$\underline{\mathbf{E}}_S \propto \underline{\mathbf{J}}\underline{\mathbf{c}} \tag{17}$$

of the received optical signal where

$$\underline{\mathbf{c}} = \begin{bmatrix} b_1 + jb_2 \\ b_3 + jb_4 \end{bmatrix} \tag{18}$$

is the transmitted Jones vector. Both vector components are recovered separately in a polarization diversity (and intradyne) receiver (Fig. 6). It has two front ends (O/E) such as shown in Fig. 1 or 4. The front ends have a common LO but orthogonal LO polarizations. Polarization beam splitters split received and local oscillator signals before the two front ends. Each front end delivers a complex (intradyne) electrical signal. Together they form a received electrical signal vector.

The Jones matrix multiplication can be resolved electronically in the receiver by an electronic multiplication of the received electrical signal vector with the matrix $\underline{\mathbf{M}} \propto \underline{\mathbf{J}}^{-1}$ [17], [18]. As a result there are two independent electronic intradyne signals which can be written as a complex Jones vector

$$\underline{\mathbf{X}} = \begin{bmatrix} \underline{X}_1 \\ \underline{X}_2 \end{bmatrix} \propto \underline{\mathbf{M}}\underline{\mathbf{J}}\underline{\mathbf{c}} \cdot e^{j\varphi'}. \tag{19}$$

Among the eight degrees-of-freedom (DOF) of the complex electronic 2×2 matrix $\underline{\mathbf{M}}$, two which determine common magnitude and phase are arbitrary. For example, it is possible to choose $\det \underline{\mathbf{M}} = 1$, but it will be seen below that this is not necessary. The others are responsible for the magnitude and phase differences between wanted and unwanted signals in each of the two output vector components $\underline{X}_1$, $\underline{X}_2$, and between the wanted signals in $\underline{X}_1$, $\underline{X}_2$.

The electronic components $\underline{X}_1$, $\underline{X}_2$ are each sent to a carrier recovery and demodulation unit as shown in Fig. 2 but the filter and the regenerative frequency dividers are shared by both units. The phase-aligned frequency-multiplied carrier signals $\underline{X}_1^4$, $\underline{X}_2^4 \propto -e^{j4\varphi'}$ of both branches are added before being passed through the low-pass filter and the regenerative frequency dividers. This addition is quite advantageous because it increases the carrier recovery SNR by 3 dB. The residual differential phase shift between $\underline{X}_1$, $\underline{X}_2$ can be controlled by an electronic error signal

$$\mathrm{Im}\left(\underline{X}_1^4 \cdot \left(\underline{X}_2^4\right)^*\right). \tag{20}$$

It vanishes if $\underline{X}_1^4, \underline{X}_2^4$ have the same phase as is needed. $\underline{\mathbf{M}}$ (or at least the remaining degrees of freedom needed to determine a

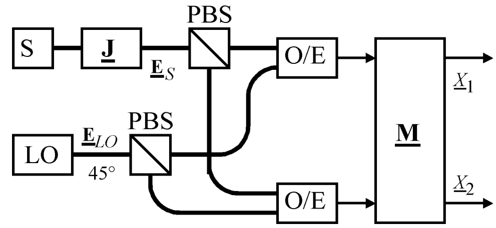

Fig. 6. "Electronic polarization control" in coherent polarization diversity receiver, for the reception of polarization division multiplex signals.

valid $\underline{\mathbf{M}}$) can be iteratively determined, for example as follows: The demodulated signals in the two polarizations can be written as an electronic vector

$$\underline{\mathbf{D}} = \underline{\mathbf{X}} \cdot e^{-j\varphi} \propto \underline{\mathbf{MJc}} \cdot e^{j\Delta\varphi} \; (\Delta\varphi = \varphi' - \varphi). \tag{21}$$

The phase error $\Delta\varphi$ is usually quite small. After proper delay, which is required due to the transit time through the decision circuit, the components of $\underline{\mathbf{D}}$ can be correlated with the recovered data symbols. Usually there are very few bit errors, hence the recovered data symbols are identical to the transmitted ones. The correlation operations result in a fairly accurate electronic estimate $\langle\underline{\mathbf{MJ}}\rangle$ of the matrix product $\underline{\mathbf{MJ}}$. Electronic polarization control is possible by iterative application of the formula

$$\underline{\mathbf{M}} := \langle\underline{\mathbf{MJ}}\rangle^{-1}\underline{\mathbf{M}} \tag{22}$$

where the RHS contains two known matrices and the left-hand side is the corrected matrix $\underline{\mathbf{M}}$. If the initial estimate of $\underline{\mathbf{M}}$ is close enough to the ideal solution $\underline{\mathbf{J}}^{-1}$ then the hardware-intensive computation (22) can be replaced by

$$\underline{\mathbf{M}} := (2 \cdot \mathbf{1} - \langle\underline{\mathbf{MJ}}\rangle)\,\underline{\mathbf{M}} \tag{23}$$

where $\mathbf{1}$ is the unity matrix.

If it were fast enough this electronic polarization control would at the same time be a carrier recovery but when combined with Fig. 2 as described the iterative updating of $\underline{\mathbf{M}}$ will contribute only very little to the carrier recovery.

IV. Laser Linewidth Tolerance

The decision variable in one decision branch of a QPSK receiver subject to the phase error $\Delta\varphi$ is for example

$$\mathrm{Re}\underline{D} = b_1 \cos\Delta\varphi - b_2 \sin\Delta\varphi. \tag{24}$$

It yields the bit error ratio (BER)

$$\mathrm{BER} = \int_{-\infty}^{\infty} \frac{1}{2}\mathrm{erfc}\left(Q_0 \frac{(\cos\Delta\varphi - \sin\Delta\varphi)}{\sqrt{2}}\right) \cdot p_{\Delta\varphi}(\Delta\varphi) \cdot d\Delta\varphi \tag{25}$$

with $p_{\Delta\varphi}(\Delta\varphi)$ being the probability density function (pdf) of $\Delta\varphi$. In an ideal receiver $Q_0 = \sqrt{2M}$ holds where M is the number of photons/bit either at the intradyne receiver input or at the input of a preceding ideal optical amplifier. Ideally, $M = 18$ is needed for a $\mathrm{BER} = 10^{-9}$.

The phase difference $\varphi'(t) - \varphi'(t-\tau)$ accumulated during a time slot τ has a Gaussian distribution with the variance $2\pi|\tau|\Delta f$ where Δf is the IF linewidth. The pdf $p_{\Delta\varphi}(\Delta\varphi)$ has been determined by a time-discrete system simulation over 10^5 symbols. The low-pass filter was simulated as a transversal filter of finite length with identical coefficients. The true $p_{\Delta\varphi}(\Delta\varphi)$ and its Gaussian approximation yield similar BER calculation results according to our experience, though the Gaussian approximation may somewhat underestimate the BER. For Fig. 7 we use the Gaussian approximation because it predicts the BER with less random fluctuations. This is a standard approach for synchronous optical transmission systems. It has, for example, been used in [1], [5], [10]. It is also a direct consequence of linearized PLL theory. A similar approach has recently also been adopted for nonlinearly generated phase noise [20].

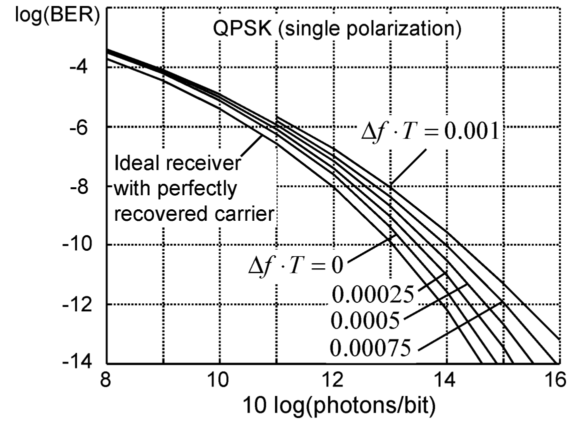

Fig. 7. log(BER) versus 10 log(photons/bit) for various IF linewidth/symbol rate ratios in a QPSK, single polarization system. The leftmost curve holds for an ideal receiver with perfect carrier recovery.

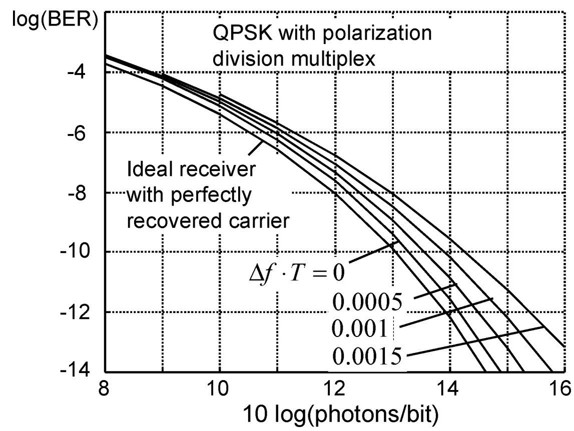

Fig. 8. log(BER) versus 10 log(photons/bit) for various IF linewidth/symbol rate ratios in a QPSK polarization division multiplex system. The leftmost curve holds for an ideal receiver with perfect carrier recovery.

Missing curve segments in Fig. 7 indicate that $|\Delta\varphi| > \pi/4$ occurred at data points, which means that a frequency division error has happened in the carrier recovery. Nevertheless a nonvanishing number of frequency division errors does not preclude operation because they will result in single bit errors if the data streams $b_{1\ldots4}$ are differentially encoded in the transmitter and differentially decoded after decision in the receiver [9].

To combine acceptable sensitivity and large phase noise tolerance, the low-pass filter width for $\mathrm{BER} = 10^{-9}$ in a standard, single-polarization QPSK system may be chosen as $0.0625/T$ where T is the symbol duration. This width was found to be optimum in a series of simulations. An intermediate frequency (IF) linewidth / symbol rate ratio $\Delta f \cdot T$ of ~0.0005 seems to be tolerable as shown in Fig. 7, with $\Delta f \cdot T$ values up to 0.001 (!). A polarization multiplexed QPSK system has also been simulated with a low-pass filter width equal to $0.125/T$. According to Fig. 8 a $\Delta f \cdot T$ value of ~0.001 is tolerable.

Straightforward simplifications of the above allow to evaluate BPSK systems: $b_2 = b_4 = 0$, only one frequency doubler, no need to invert the frequency-multiplied signal due to $\underline{c}^2 = 1$, only one frequency halver, $|\Delta\varphi| > \pi/2$ indicates a division

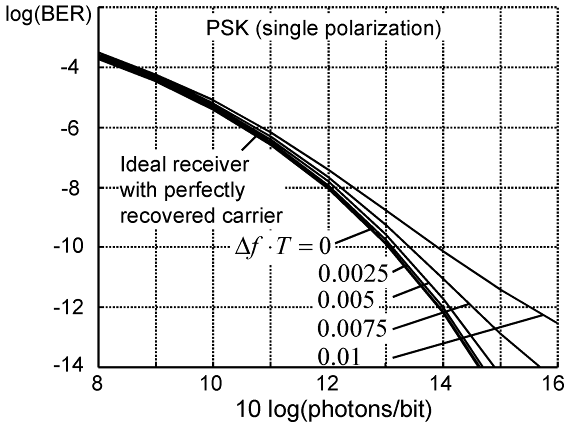

Fig. 9. log(BER) versus 10 log(photons/bit) for various IF linewidth/symbol rate ratios in a BPSK, single polarization system. The leftmost curve, almost coincident with the simulated no-phase noise curve, holds for an ideal receiver with perfect carrier recovery.

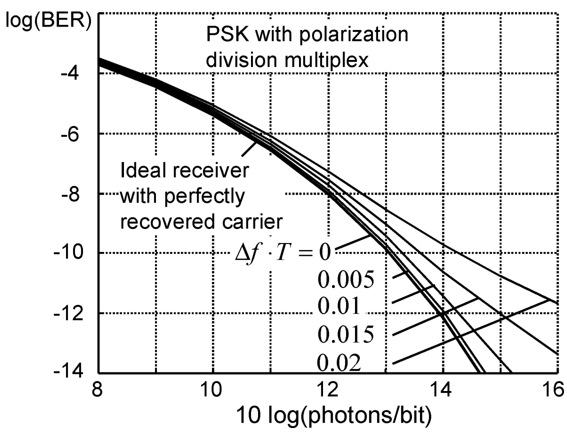

Fig. 10. log(BER) versus 10 log(photons/bit) for various IF linewidth/symbol rate ratios in a BPSK polarization division multiplex system. The leftmost curve, almost coincident with the simulated no-phase noise curve, holds for an ideal receiver with perfect carrier recovery.

error. The tolerable $\Delta f \cdot T$ value is ~0.005 for the single polarization case (Fig. 9), using a low-pass filter width of $0.0625/T$. For the polarization multiplexed case the simulations lead to a permissible $\Delta f \cdot T$ of ~0.01 (Fig. 10), with a lowpass filter width of 0.125/T. The optimum sensitivity for BPSK is again $M = 18\,\mathrm{photons/bit}$. A value of nine photons/bit is achievable only in an OPLL-based BPSK homodyne receiver without optical amplification, which is unrealistic in today's EDFA-laden fiber plant.

These predictions do enjoy some experimental support, and thereby render the question, whether $p_{\Delta\varphi}(\Delta\varphi)$ is truly Gaussian or not, less important: In our previously reported synchronous heterodyne BPSK system with DFB lasers [9] a $\Delta f \cdot T$ value of 0.0018 was permissible, not ~0.005 as predicted above. However, a static frequency divider introduced digitization noise, the filter for the frequency-doubled IF signal had a lot of group delay ripple, the IF was not an integer multiple of the bit rate because that filter happened to be centered at another frequency, and the transmitter laser was directly modulated, which caused substantial extra phase distortion. Given these experimental shortcomings, the present findings are fairly consistent with the experiment [9].

But let us consider a worst case scenario: [2], [10], [19] and this study predict that QPSK tolerates about 10 times less linewidth than BPSK. One tenth of the experimental BPSK $\Delta f \cdot T$ value [9] amounts to a permissible IF linewidth of 1.8 MHz for a 10 Gbaud QPSK single-polarization system. Even such a low IF linewidth should be achievable, given the fact that DFB lasers with $\leq$1 MHz linewidth are commercially available. But even if this worst case assumption should be correct things may be easier in practice: A $\mathrm{BER} = 10^{-9}$ will not be needed if forward error correction is implemented. Also, the linewidth tolerance will be doubled if polarization division multiplex is employed.

V. Conclusion

A feedforward carrier recovery for QPSK/BPSK receivers is predicted to be extremely laser-linewidth tolerant, two decades more than realistic OPLLs. Frequency halvers implemented as I and Q regenerative frequency dividers make the scheme applicable for baseband type intradyne receivers. These are attractive at 10 Gbaud, thereby enabling 40 Gbit/s transmission on each WDM channel by QPSK modulation combined with polarization multiplex. It is clear that DFB lasers shall suffice, probably even for single-polarization QPSK at 10 Gbaud. The scheme is likewise applicable for 40 Gbaud/160 Gbit/s transmission.

A simple algorithm for polarization control in the electronic domain has also been proposed.

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