Feedforward carrier recovery for polarization demultiplexed signals with unequal signal to noise ratios

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Abstract: We investigate feedforward carrier recovery (FFCR) in coherent polarization diversity receivers where the signal to noise ratio (SNR) of polarization demultiplexed signals can be unequal, such as in polarization-dependent loss impaired systems. A joint-polarization FFCR mechanism for estimating the carrier phase noise based on samples from both polarizations is proposed and compared with three other plausible alternatives. We evaluated each architecture using Monte Carlo simulations and observed that the joint-polarization FFCR yields a 1.1 dB SNR penalty for a given laser linewidth × baud rate product, while the other three architectures offer 1.8 dB, 2.0 dB and 3.9 dB, for QPSK at BER = 10⁻³ and 3 dB SNR imbalance.

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OCIS codes: (060.0060) Fiber optics and optical communications; (0606.5060) Phase modulation; (060.1660) Coherent communications.

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1. Introduction

Polarization multiplexing and the electronic compensation of signal distortions are two key technologies for future optical transmission systems. In fact, signal processing techniques have now been used for compensating a number of effects related to polarization multiplexing, like polarization mode dispersion (PMD) [1] or the so called “depolarization” caused by polarization-dependent loss (PDL) [2]. These two effects though are of a completely different nature: while PMD randomly changes the signal state-of-polarization (SOP) along the birefringent fiber, PDL is a lossy effect that may cause an SNR imbalance after equalization and source separation [2], thus impacting the performance of subsequent receiver components. In M-PSK modulated systems, for example, the equalizer is followed by some technique for digital carrier phase estimation, e.g., a digital delay lock loop (DLL) or feedforward carrier recovery (FFCR). The latter is usually preferred because of its stability (no loops involved) and phase-noise tolerance [4, 3].

In this paper we investigate feedforward carrier recovery for polarization demultiplexed signals whose signal to noise ratios can be unequal. A joint-polarization FFCR mechanism for estimating the carrier phase noise based on samples from both polarizations is proposed and compared with three other plausible choices. All architectures are investigated in detail, including the sometimes neglected phase unwrapper, a receiver component whose task is to allow the phase noise estimates to range from $-\infty$ to $\infty$ instead of $-\pi/M$ and $\pi/M$ (outcomes of an $1/M \arg(·)$ operator). The phase unwrapper is a nonlinear component that is relatively difficult to model analytically, thus motivating the use of Monte Carlo simulations. This paper is organized as follows. Section 2 presents a literature review on the topic. Section 3 shows PDL as a possible source of unbalanced SNR after polarization diversity homodyne reception. Section 4 explains the investigated FFCR architectures, including the newly proposed joint-estimation. Section 5 reports simulation results for quadrature-phase shift keying transmission; and finally, Section 6 concludes the paper.

2. Previous works and contribution

One of the first works addressing feedforward carrier recovery (FFCR) after the re-emergence of coherent detection using signal processing was presented by Noé in [5], proposing an architecture based on regenerative intradyne frequency dividers. Focused on the fundamental possibility of deploying QPSK receivers with distributed feedback (DFB) lasers, the work used a flat FFCR low-pass filter and analyzed results using analytical formulas which neglected a phase unwrapper. An interesting review of coherent detectors and phase recovery was published by Kazovsky et al. in [4], comparing the performance of a FFCR scheme with that of a digital delay lock loop (DLL). Performance analysis was carried out by Monte Carlo simulations assuming a flat FFCR low-pass filter. A theoretically comprehensive work on architectures for FFCR was published by Ip and Kahn in [3]. By means of Monte Carlo simulations, the paper analyzed...
an FFCR architecture for a single polarization scenario. A thorough and excellent review of coherent detection in optical fiber systems was presented by Ip et al. in [6]. The paper qualitatively discusses FFCR architectures for dual polarization transmission, but does not mention the placement of the phase unwrapper. In Reference [7] Kuschnerov et al. proposed a joint polarization FFCR scheme for XPM-limited coherent polarization-multiplexed QPSK transmission with OOK-neighbors. The scheme uses flat FFCR low-pass filters and an experimentally determined coupling factor for both polarizations. An interesting paper about the impact of polarization-dependent loss on coherent POLMUX-NRZ-DQPSK was published by Duthel et al. in [2], showing that the orientation between polarization multiplexed signals and PDL elements has a major impact on the SNR performance. The paper also suggests that interleaving all tributaries into a single FEC device averages the BER and improves performance. FFCR is not discussed and flat low-pass filters are assumed. El-Darawy et al. investigate in Reference [8] fast adaptive polarization and PDL tracking in a realtime FPGA. They use a flat FFCR low-pass filter and show that a fast controller can effectively reduce the sensitivity penalty due to quick polarization variations. Goldfarb et al. presented in [9] an analytical method for BER estimation in QPSK homodyne detection with flat filter carrier phase estimation. In this paper, we investigate feedforward carrier recovery (FFCR) in coherent polarization diversity receivers where the signal to noise ratio (SNR) of polarization demultiplexed signals can be unequal, such as in polarization-dependent loss impaired systems. We evaluate four FFCR architectures in QPSK systems using Monte Carlo simulations. Perfect knowledge of the SNR of polarization demultiplexed signals is assumed, and adaptive estimation and tracking (in case of a time-varying PDL, for example) of the SNR for the polarization demultiplexed signals is left for further study.

3. System model

In this section, we formulate the system model used throughout the paper (see Fig. 1(a)). In the transmitter, two complex data streams modulate orthogonally-polarized optical carriers using I+Q Mach-Zehnder modulators (I+Q MZM). The signals are then coupled by a polarization beam combiner (PBC) to generate a polarization multiplexed signal (PolMux).

In order to produce an SNR imbalance in different polarizations, the PolMux signal is first fed into a PDL generator element (see Fig. 1(b)) with an orientation mismatch \( \alpha \), and subsequently loaded with amplified spontaneous emission noise. After that, the signal is received by the traditional polarization diversity receiver consisting of: 90° hybrids; balanced detectors; a filter matched to the signal pulse shape; and optimal symbol rate sampling. We call \( \beta \) the orientation mismatch between the PDL generator element and the receiver local oscillators. In complex low-pass representation, the PolMux signal is given by:

\[
s_k = \begin{bmatrix} s_{V_k} \\ s_{H_k} \end{bmatrix},
\]

where \( s_{V_k} \) and \( s_{H_k} \) are the vertical and horizontal polarization data streams. After ideal polarization diversity detection the received signal can be expressed as a sum of the PDL and phase-noise distorted signal with the noise generated by the EDFAs:

\[
r_k = Hs_k \exp(j\theta_k) + n_k,
\]

where \( H \) is the PDL element transmission matrix, \( \theta_k \) is the local oscillator and transmitter laser phase mismatch, and \( n_k \) is a two-dimensional vector with complex additive Gaussian noise (AWGN) processes. Matrix \( H \) is usually presented in terms of the PDL coefficient \( \Gamma'(dB) = 10\log \frac{1}{\gamma} \).
Fig. 1. (a) Polarization multiplexed system employing a homodyne coherent receiver. (b) PDL model.

\[
H = \begin{bmatrix}
\cos \alpha & -\sin \alpha \\
\sin \alpha & \cos \alpha
\end{bmatrix}
\begin{bmatrix}
\sqrt{1 - \gamma} & 0 \\
0 & \sqrt{1 + \gamma}
\end{bmatrix}
\begin{bmatrix}
\cos \beta & -\sin \beta \\
\sin \beta & \cos \beta
\end{bmatrix},
\] (3)

where the first and last matrices are rotations with respect to the PDL generator element, and the central matrix is responsible for the imbalanced attenuation of orthogonal polarizations.

After coherent reception, signals are equalized to compensate for linear distortions, including usually chromatic dispersion and polarization mode dispersion (see Fig. 2). In our simplified model a separating matrix \( W \) compensates for the polarization mismatch between the incoming signal and the local oscillator orientation:

\[
y_k = W r_k = W [H s_k \exp(j \theta_k) + n_k] = W H s_k \exp(j \theta_k) + W n_k \\
\approx s_k \exp(j \theta_k) + w_k,
\] (4)

where \( w_k = W n_k \). Depending on \( W \), \( \sigma_w^H \) can be different from \( \sigma_w^V \).

It is interesting to notice that the phase offset \( \theta_k \) is the same for both polarizations, therefore, for each polarization:

\[
y_k^{V/H} = s_k^{V/H} \exp(j \theta_k) + w_k^{V/H}.
\] (5)
4. Feedforward carrier recovery

4.1. Single polarization

The single polarization FFCR estimates $\theta_k$ from the ASE noise corrupted signal $y_k$:

$$ y_k = s_k \exp(j\theta_k) + w_k, $$  

where $w_k$ is an AWGN process with zero mean and variance $\sigma_w^2 = N_0$. The classical architecture for the single polarization FFCR unit using the Viterbi and Viterbi algorithm [10] is shown in Fig. 3. The incoming complex low-pass signal is first raised to the $M$th power to remove data dependencies and subsequently filtered to mitigate some of the additive noise perturbation. Next, the argument of filtered signal is divided by $M$ and “unwrapped” to allow the estimated phase mismatch to evolve from $-\infty$ to $+\infty$ instead of $-\pi/M$ to $+\pi/M$, e.g. [1]:

$$ PU(\cdot) = (\cdot) + \left( \frac{1}{2} + \frac{\hat{\theta}_{k-1} - (\cdot)}{2\pi/M} \right) \frac{2\pi}{M}. $$  

We model $\theta_k$ as a discrete time Wiener process, such that:

$$ \theta_k = \theta_{k-1} + \Delta_k; $$

$$ \theta_{k-i} = \theta_k + \sum_{m=0}^{i-1} \lambda_m; $$  

$$ \theta_{k+i} = \theta_k + \sum_{m=0}^{i-1} \mu_m. $$
where $\Delta_t$, $\lambda_m$ and $\mu_m$ are Gaussian distributed random variables with zero mean and variance $\sigma^2_\Delta = \sigma^2_\lambda = \sigma^2_\mu = 2\pi \Delta V_T$. Bandwidth $\Delta V$ is the sum of the laser and local oscillator (LO) 3 dB linewidths, and $T_s$ is the symbol period. Some sample $y_{k-i}$ can be thus written as [11]:

$$y_{k-i}^M = \{s_{k-i} \exp[j(\theta_k + \sum_{m=0}^{i-1} \lambda_m)] + w_{k-i}\}^M;$$

$$\approx s_{k-i}^M \exp[jM(\theta_k + \sum_{m=0}^{i-1} \lambda_m)] + z_{k-i}, \quad (10)$$

where $z_{k-i}$ is a Gaussian random variable with zero mean and variance $\sigma^2_z = M^2 E_s^{M-1} \sigma^2_w$, and $E_s = |s_k|^2$. Assuming low phase noise:

$$y_{k-i}^M \approx E_s^{M/2} \exp(jM\theta_k) \left(1 + jM \sum_{m=0}^{i-1} \lambda_m\right) + z_{k-i}. \quad (11)$$

Analogously, for the future symbols:

$$y_{k+i}^M \approx E_s^{M/2} \exp(jM\theta_k) \left(1 + jM \sum_{m=0}^{i-1} \mu_m\right) + z_{k+i}. \quad (12)$$

We wish to obtain the maximum likelihood (ML) estimate of $\theta_k$. The probability density function of a vector $\mathbf{r} = \{y_{-N}^M, \ldots, y_{k-1}^M, y_k^M, \ldots, y_{k+N}^M\}$ with past and future samples for a given $\theta_k$ can be written as:

$$f_{\theta_k}(\mathbf{r}|\theta_k) = \frac{1}{(2\pi)^{L/2} (\det \mathbf{C})^{1/2}} \exp \left[ -\frac{1}{2} (\mathbf{r} - \mathbf{m}_r)^H \mathbf{C}^{-1} (\mathbf{r} - \mathbf{m}_r) \right], \quad (13)$$

where $\mathbf{m}_r = E\{\mathbf{r}\} = E_s^{M/2} \exp(jM\theta_k) \mathbf{1}$, and $\mathbf{C}$ is the covariance matrix. Matrix $\mathbf{C}$ can be expressed as:

$$\mathbf{C} = E_s^2 M^2 \mathbf{K} \sigma^2_w + E_s^{M-1} M^2 \mathbf{K}_n, \quad (14)$$

where $\mathbf{K}_n = \sigma^2_w \mathbf{I}_{LxL}$ for a filter of length $L = 2N + 1$, and $\mathbf{K}$ is given by:

$$\mathbf{K} = \begin{bmatrix}
N & \cdots & 2 & 1 & 0 & 0 & 0 & \cdots & 0 \\
\vdots & \ddots & \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
2 & \cdots & 2 & 1 & 0 & 0 & 0 & \cdots & 0 \\
1 & \cdots & 1 & 1 & 0 & 0 & 0 & \cdots & 0 \\
0 & \cdots & 0 & 0 & 0 & 0 & 0 & \cdots & 0 \\
0 & \cdots & 0 & 0 & 0 & 1 & 1 & \cdots & 1 \\
0 & \cdots & 0 & 0 & 0 & 1 & 2 & \cdots & 2 \\
\vdots & \ddots & \vdots & \vdots & \vdots & \vdots & \vdots & \ddots & \vdots \\
0 & \cdots & 0 & 0 & 0 & 1 & 2 & \cdots & N \\
\end{bmatrix}. \quad (15)$$

From Eq. 13 it is possible to derive the likelihood function:
\[ \Lambda_{ln} = \ln \left( \frac{1}{(2\pi)^{L/2}(\det C)^{1/2}} \right) - \frac{1}{2} (r - m_r)^H C^{-1} (r - m_r), \] (16)

which can be shown to be maximized if:

[\[ \theta_k = \frac{1}{M} \arctan \left( \frac{\Im \{ (1^T C^{-1} \cdot r) \}}{\Re \{ (1^T C^{-1} \cdot r) \}} \right). \] (17)

Taking into account the FFCR architecture in Fig. 3, \( \hat{\theta}_k \) is given by:

\[ \hat{\theta}_k = PU \left\{ \frac{1}{M} \arg (1^T C^{-1} \cdot r) \right\}. \] (18)

4.2. Dual polarization

In this section we model four feedforward carrier recovery architectures for polarization demultiplexed signals. The architectures differ in the weights of FFCR low-pass filters, and in the location of the phase unwrapper.

4.2.1. Individual estimation

This first architecture performs FFCR in each polarization individually, as two parallel single polarization structures, waiving diversity (see Fig. 4). Estimates \( \hat{\theta}_k^V \) and \( \hat{\theta}_k^H \) are therefore obtained by low-pass filtering the received symbols in polarizations \( V \) and \( H \):

\[ \hat{\theta}_k^V = PU \left\{ \frac{1}{M} \arg \left( 1^T (C^V)^{-1} \cdot r^V \right) \right\}, \] (19)

\[ \hat{\theta}_k^H = PU \left\{ \frac{1}{M} \arg \left( 1^T (C^H)^{-1} \cdot r^H \right) \right\}. \] (20)

![Diagram of Individual Feedforward Carrier Recovery for Dual Polarization](image)

Fig. 4. Individual feedforward carrier recovery for dual polarization.

4.2.2. Mean estimation

Mean estimation FFCR uses the same low-pass filters as calculated for the single and individual estimation architectures, but combines the filtered complex phasors using the arithmetic mean to produce a single \( \hat{\theta}_k \). Notice however that a misplaced phase unwrapper can lead to catastrophic errors. Phase unwrappers are nonlinear devices that add \( 2\pi/M \) multiples to the
estimated phase in order to let it range from $-\infty$ to $+\infty$, instead of $-\pi/M$ to $+\pi/M$. Nevertheless, strong phase or additive noise can lead to cycle slips, a phase offset of $2\pi/M$ added to the estimated phase that potentially induces a catastrophic error sequence. This problem is usually circumvented by the use of differential decoding, which is immune to phase offsets, allowing just pairs of equivocated symbols adjacent to a cycle slip. An interesting example of misplaced phase unwrappers is depicted in Fig. 5: the arithmetic mean of angles after phase unwrapping would lead to catastrophic failures if a cycle slip occurs in just one of the branches. The proper mean estimation architecture is shown in Fig. 6, where the phase unwrapper is positioned after the arithmetic mean of complex phasors:

$$\hat{\theta}_k = PU \left\{ \frac{1}{M} \arg \left[ \frac{1}{2} \left( C^V - 1 \cdot r^V + (C^H - 1 \cdot r^H) \right) \right] \right\}. \quad (21)$$

Fig. 5. Mean of unwrapped angles. This architecture would lead to catastrophic failures.

Fig. 6. Mean feedforward carrier recovery for dual polarization.

### 4.2.3. Flat filter estimation

This architecture uses a finite length flat structure for FFCR low-pass filtering. The flat filter estimation has been broadly deployed in lab and field experiments, and the filter length has been empirically determined. It is however possible to envisage some practical adaptive-length algorithm based, e.g., on the forward-error correction performance (FEC). The estimated phase mismatch estimate is expressed as:

$$\hat{\theta}_k = PU \left\{ \frac{1}{M} \arg \left( \frac{1}{2} \sum_{i=-N}^{N} (y^V_{k+i})^M + \sum_{i=-N}^{N} (y^H_{k+i})^M \right) \right\}, \quad (22)$$
where $N$ is chosen as to minimize the BER.

![Diagram](image)

**Fig. 7.** Flat filter feedforward carrier recovery for dual polarization.

4.2.4. Joint estimation

The joint estimation method seeks to find the estimate of $\theta_k$ that maximizes $f_{r^*} |\theta_k( r^* | \theta_k) $, where:

$$r^* = \left( (r_{k-N}^{V} )^M , \ldots , (r_{k+N}^{V} )^M , (r_{k-N}^{H} )^M , \ldots , (r_{k+N}^{H} )^M \right)^T .$$

Notice that $r^*$ includes samples from the vertical and horizontal polarizations. The matrix $C'$ associated to $r^*$ can be shown to be:

$$C' = E^M s^M 2^K \sigma^2_M + E^M -1 s^M 2^K_n ,$$

(23)

where $K'$ and $K'_n$ can be obtained from the $K$ matrix defined in Section 4.1:

$$K' = \begin{bmatrix} K & K \\ K & K \end{bmatrix} ;$$

(24)

$$K'_n = \begin{bmatrix} I_{LxL} \sigma^2_{w^V} & 0 \\ 0 & I_{LxL} \sigma^2_{w^H} \end{bmatrix} .$$

(25)

The formulation of matrix $K'$ is based on the fact that the received V and H symbols are contaminated by the same Wiener phase noise. Finally, the jointly estimated phase noise is calculated by weighing samples from both polarizations:

$$\hat{\theta}_k = PU \left\{ \frac{1}{M} \arg \left[ (1^T C'^{-1} \cdot r^* ) \right] \right\} .$$

(26)

Figure 9 shows, for illustration purposes, the coefficients of an $N=10$ filter used in the single and joint estimation for a scenario where $SNR^V = 7$ dB and $SNR^H = 4$ dB, and the phase noise is a Wiener process generated by the accumulation of a zero mean Gaussian process with $\sigma^2_\Delta = 2\pi \Delta \nu T_s = 2\pi 10^{-4}$. The dashed red line indicates the filter coefficients individually calculated for each polarization (V polarization on the left, and H polarization on the right). Indeed, the optimum filter shape depends on the relative significance of additive noise and Wiener noise. As expected, the filter shape for the H polarization is smoother, since the SNR is lower and additive noise becomes relatively more important. As for the V polarization, the filter shape is sharper, since the SNR is higher and the Wiener noise becomes relatively more prominent. The solid blue curve shows the filter coefficients jointly calculated. It is interesting to notice that both V and H joint estimation filters have different shapes when compared to the individual estimation filters, and that the H polarization filter – which has a lower SNR – has attenuated coefficients when compared to the V polarization filter.
Fig. 8. Joint feedforward carrier recovery for dual polarization.

Fig. 9. Normalized filter weights for polarizations V and H when $SNR_V = 7$ dB (curves on the left) and $SNR_H = 4$ dB (curves on the right). Dashed line: individual estimation. Solid line: joint estimation. The difference between adjacent phase noise samples is a Gaussian process with zero mean and variance $\sigma^2 = 2\pi\Delta\nu T_s = 2\pi10^{-4}$. The filter length $N = 10$.

5. Performance evaluation

5.1. Simulation setup

We carried out computationally intensive Monte Carlo simulations to evaluate the four feedforward carrier recovery alternatives analyzed in this paper. In the transmitter, two polarization multiplexed QPSK signals (V and H) are randomly generated by the modulation of four Gray-mapped pseudo-random bit sequences produced using Matlab. In order to incorporate the phase mismatch between incoming signal and local oscillator, the V and H signals are corrupted by the same multiplicative Wiener phase noise with $\sigma^2 = 2\pi\Delta\nu T_s$, where $\Delta\nu$ is the sum of the transmitter and local oscillators 3 dB bandwidths and $T_s$ is the symbol period. Assuming perfect source separation, complex additive white Gaussian noise (AWGN) with 3 dB imbalanced variance is added to both signals ($\sigma^2_w = 2\sigma^2_w V$). As an example, Fig. 10 shows the constellation modulation for V and H polarization signals corrupted by different additive noise powers, after and before carrier recovery.

The receiver consists of one of the four investigated feedforward carrier recovery architectures followed by minimum distance decision and differential decoding. The bit error rate (BER) in each polarization is calculated by comparing at least $2 \times 10^6$ transmitted bits (from 10 independent simulation rounds of $10^5$ in-phase plus $10^5$ quadrature sequential bits). In the diagram, penalties are obtained as the difference (at $BER = 10^{-3}$) between the experimental and theoretical SNR, calculated as the inverse of [11]:

$$BER\left(SNR = \frac{E_b}{N_0}\right) \approx 2Q\left[\sin\left(\frac{\pi}{M}\right)\sqrt{\frac{4E_b}{N_0}}\right].$$ (27)

#107474 - $15.00 USD  Received 11 Feb 2009; revised 27 Mar 2009; accepted 6 Apr 2009; published 29 Apr 2009  (C) 2009 OSA 11 May 2009 / Vol. 17, No. 10 / OPTICS EXPRESS 7967
Fig. 10. Modulation constellation for 5,000 received QPSK symbols corrupted by AWGN and Wiener Noise with $\Delta v T_s = 10^{-3}$. The figures on the left show constellations before carrier recovery, and those on the right after carrier recovery using the individual method. In the upper curves the SNR is 3 dB higher (7 dB) than for those on the bottom (4 dB).

The simulations used a maximum SNR step resolution of 0.06 dB. In the individual, joint and mean methods we fixed the filter length to $L = 2N + 1 = 21$. In the flat filter estimation we optimized the filter length by running ten simulations with ten different values of $N$ ($N$ ranging from 1 to 10 and, consequently, $L$ ranging from 3 to 21) and picking up the one with the lowest SNR penalty. Notice that this filter length optimization was applied to each of the 10 independent simulation rounds of $10^5$ symbols. Alternatively, an analytical expression for the optimal flat filter length could be found in [9].

5.2. Results and discussion

We simulated a polarization multiplexed signal where the SNR of polarization H is 3 dB lower than that of polarization V (we use V and H notation for the sake of clarity). The curves below correspond to the worst case, H polarization, which dominates the overall BER. As expected, the dot-dashed red individual estimation curve exhibits the highest penalties, since it does not profit from the information provided by the other, higher SNR polarization. The dotted black curve corresponds to the optimum flat filter that minimizes the BER, often used in field trials. Finally, the dashed green and the solid blue were obtained using the mean and the joint estimation methods. It is interesting to observe that, for a given bit rate, the joint estimation method can almost halve the necessary laser linewidth required to achieve 1 dB SNR penalty.
for BER = 10^{-3} in comparison to the flat filter method. The joint-polarization FFCR yielded a 1.1 dB SNR penalty for a laser linewidth × baud rate product of 5 × 10^{-3}, while the other three architectures offered 1.8 dB (mean), 2.0 dB (flat filter) and 3.9 dB (individual). In terms of computational complexity, both the mean and joint estimation methods require the estimation of the SNR in both polarizations, whereas the flat filter estimation demands a BER estimation. Indeed, imperfect SNR estimation and relatively fast PDL variations would impair the performance of mean and joint estimation methods, thus reducing their attractiveness. However, since polarization changes are usually several orders of magnitude slower than the symbol duration [8], both joint and mean estimation methods can be appealing. Notice in this case that the computational complexity of the algorithm is dominated by filtering, rather than by the computation of coefficients. When compared to the mean estimation method, joint estimation offers some SNR tolerance at the cost of inverting larger – but sparse – matrices with very particular structures, therefore being a plausible alternative for the system designer.

Fig. 11. SNR penalty versus laser linewidth × baud rate product for the four feedforward carrier recovery estimation methods (@ QPSK, 3 dB SNR imbalance, worst case polarization and BER = 10^{-3}).

6. Conclusion

Carrier phase estimation (CPE) is a central element for coherent detection of M-PSK modulated signals, becoming more important with increasing M. A suitable CPE algorithm has the potential to relax the requirements for transmitter laser or local oscillator bandwidth. In polarization multiplexing systems, CPE schemes may profit from diversity, since the contaminating phase noise is the same for both polarizations. An optimized polarization-diversity structure, however, must take into account an eventual SNR imbalance (as it can occur in PDL impaired systems) to calculate its filter coefficients. In this paper we propose an SNR aware joint-polarization CPE scheme using feed-forward carrier recovery and compare it with three other typical alternatives. We observed that in some cases the newly proposed architecture almost doubled the tolerable laser linewidth × baud rate product when compared to the typical flat filter option (using QPSK and 3 dB SNR imbalance).

Acknowledgment

Francisco A. C. Garcia was supported by FAPESP.