# Precise, Robust and Least Complexity CD estimation

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**Abstract:** We demonstrate robust and precise frequency domain chromatic dispersion estimation based on offline data and simulations. The blind, least complexity algorithm allows fast acquisition suitable for digital coherent receivers in switched networks.
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## 1. Introduction

Polarization-division multiplexed (PDM) quaternary phase-shift keying (QPSK) modulation at 112 Gbit/s is widely considered in product development for long-haul transponders due to its increased spectral efficiency and its compatibility with the existing 10G/40G architecture. In combination with coherent detection receivers incorporating digital equalization, full compensation of all linear channel impairments of a linear or weakly nonlinear optical fiber channel can be achieved [1]. This allows for uncompensated transmission with digital chromatic dispersion (CD) compensation instead of optical compensation by means of dispersion compensating fiber (DCF). Given the long impulse response of uncompensated links, frequency-domain (FD) filter implementation offers low complexity implementation with a large degree of parallelization [2].

Although the value of CD does not change largely during transmission, in switched networks a fast and reliable initialization of the CD compensation is vital. As FD CD compensation is typically regarded as the first key stage of digital equalization and synchronization, a slow estimation procedure clearly delays the initialization of all subsequent processing blocks. A large CD estimation error may even lead to a total failure of the digital coherent receiver, as the time-domain (TD) 2x2 multi-input multi-output (MIMO) filter with only a low number of taps will not converge [3] and the digital timing-recovery cannot acquire locking [4].

Blind (non-data-aided, NDA) digital CD estimation has been demonstrated based on monitoring TD MIMO filter taps converged to the minimum mean square error (MMSE) solution [5] or by a "best-match-search" looking for the minimum cost function based on the constant-modulus algorithm (CMA) after scanning throufh potential CD filters [3]. The first method is limited by the number of filter taps, the second method is prone to the sampling instant and the modulation format. In [6], the cost function has been transferred into the FD which makes the estimation independent from the timing phase and the modulation format.

In the following, we present an improved FD CD estimation method, which reduces the implementation complexity tremendously compared to [6] and subsequently speeds up the estimation time significantly. To our knowledge this is the least complexity and most precise NDA CD estimation algorithm for digital coherent receivers. The performance is demonstrated over a wide range of combined channel impairments with simulated and experimental long-haul transmission data of 112 Gbit/s PDM-QPSK modulation.

## 2. Auto-Correlation Based Frequency-Domain Chromatic Dispersion Estimation

The receiver architecture follows the typical intradyne coherent demodulation with polarization-diverse optical 90°-hybrid, ADC stage and subsequent digital equalization and synchronization (Fig. 1, left). Blocks of signal spectra

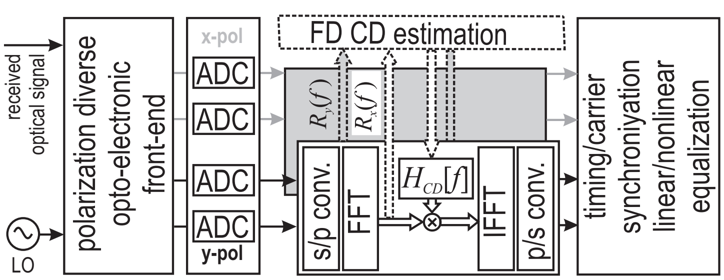


Fig. 1. Front-end with polarization diverse 90°-hybrid, ADC (2 samples/symbol) and digital equalization including FD CD compensation (1024 samples FFT).

| impairment | distribution | parameter |
|---|---|---|
| PMD | Maxwellian | mean 25 ps |
| PDL | linear | [0:10] dB |
| CD | linear | [-30:30] ns/nm |
| LOFO | linear | [-1.5:1.5] GHz |
| $\theta$ | linear | $[0:2\pi]$ |
| $\phi$ | linear | $[0:2\pi]$ |
| OSNR | constant | 14 dB |

Tab. 1. Table of random channel impairments including the statistical distribution.

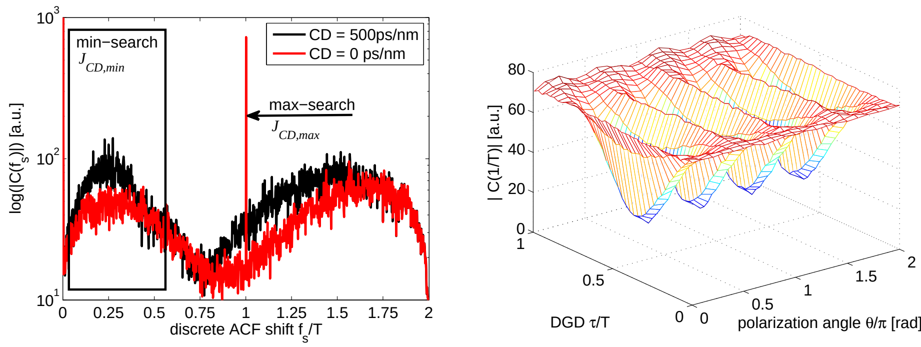


Fig. 2. Exemplary magnitude of ACF $log\{|C(fs)|\}$ averaged over 10 signal blocks for residual CD = 0 ps/nm and CD= 500 ps/nm (left) and CT magnitude for variations of DGD and polarization angle (right).

$R_x(f)$ and $R_y(f)$ from both polarizations are loaded into the CD estimation block.

It is well known that residual CD limits digital timing recovery [4]. Reversely, we can monitor a matching or non-matching CD compensation by observing the properties of the timing recovery. This leads to a highly precise and robust CD estimation with a low-complexity implementation derived from the Godard timing recovery, which can be described in the FD by the auto-correlation (ACF) of the received signal spectra evaluated in a bandwidth $B$ around the clock tone (CT) at the Baud-rate $f_s = \pm 1/T$ (symbol duration $T$)

$$C(f_s, B) = \int_{-B/2}^{B/2} R(f - f_s) R^*(f + f_s) df \quad , \tag{1}$$

where $R(f)$ and $R^*(f)$ define the spectrum and its complex conjugate respectively [7]. As it has been shown in [4] for a channel governed by CD, the CT magnitude can be estimated by

$$|C(f_s = 1/T, B)| \approx \left| \frac{\sin(\psi 2 B f_s)}{\psi 4 f_s} S(B/2 - f_s) S^*(B/2 + f_s) \right| \quad , \tag{2}$$

with $\psi = -D\lambda^2\pi/c$, where $D$ refers to the residual CD, $\lambda$ to the carrier wavelength and $c$ to the speed of light, while $S(f)$ refers to the transmitted base-band signal without channel distortions. Clearly, Eq. 2 scales with $1/\psi$ suppressing the CT magnitude for increasing values of residual CD. This is also the reason why digital timing recovery only works for low values of residual CD.

The properties of the ACF may now be employed in a best-match search, filtering a received signal spectrum $R(f)$ by a range of potential CD compensation functions $H_{CD,i}(f)$ with $R_i(f) = R(f) \cdot H_{CD,i}(f)$. Evaluating different areas of $|C(f_s, B)|$, two cost functions can be defined. Averaging either on components $f_s \neq 1/T$ or $f_s = 1/T$, the minimum $J_{CD,min}[i]$ or the maximum $J_{CD,max}[i]$ indicate the best matching CD compensation filter respectively. Two examples of the ACF for CD=0 ps/nm and CD=500 ps/nm are given in Fig. 2, left. Given a matching CD compensation filter with zero residual CD, it can be observed that the CT at $f_s = 1/T$ becomes maximum and disappears for CD values above 500 ps/nm. Similarly, this applies for areas with $f_s \neq 1/T$ vice versa.

However, it has been also also shown in [4] that the CT magnitude is prone to polarization effects, to the polarization angle in particular. Figure 2, right, depicts the CT magnitude $|C(fs = 1/T)|$ with respect to the differential group delay (DGD) $\tau$ and polarization angle $\theta$. Combinations of half-Baud DGD $\tau = T/2$ with $\theta = \pi/4 + n \cdot \pi/2$, $n \in \mathbb{N}$, suppress the CT magnitude. This worst-case can be avoided by either using an adaptive SOP filter or simply by averaging on a row of fixed SOP pre-filters. As even higher-order polarization-mode dispersion (PMD) results in regular CT magnitude patterns, only a few SOP pre-filters are required, which can be realized by simple linear combinations of the received signal spectra $R_x(f)$ and $R_y(f)$ to yield $R(f)$. Furthermore, averaging on independent estimations of several signal blocks improves the estimation, where signal blocks do not need to be adjacent.

The minimum search has been applied in [6]. Although robust estimation performance has been demonstrated over a wide range of combined channel distortions, the low peak-to-average ratio (PAR) of the cost function $J_{CD,min}$ required an alarm to increase the number of signal block for averaging at low PARs. In contrast, the new method based on the CT magnitude $J_{CD,max}$ provides a much better distance between estimations of non-matching and matching filter conditions, shown in Fig.3, left. At the same time, the implementation complexity is reduced tremendously as only one ACF component at $fs = 1/T$ needs to be computed. Furthermore, the required number of operations can be reduced by optimizing the parameter $B$, by optimizing the scanning range and the step width for all

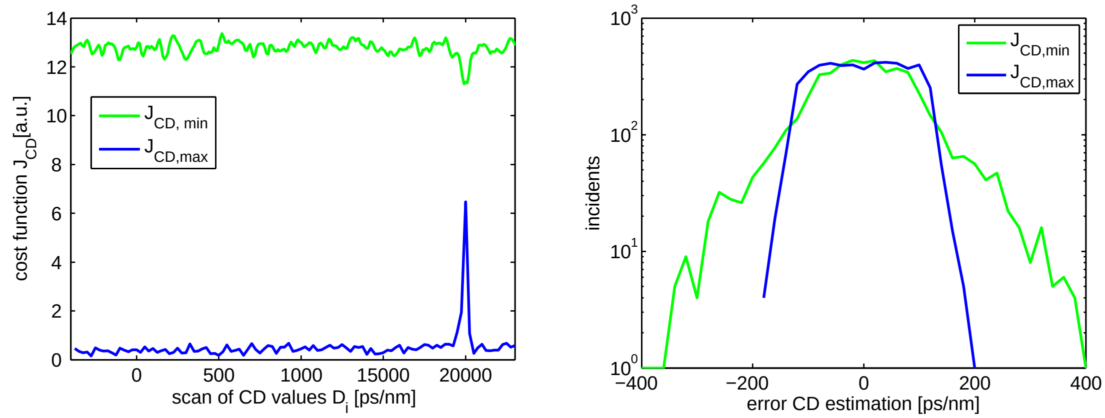


Fig. 3. Cost functions $J_{CD,min}$ and $J_{CD,max}$ of minimum search and maximum search respectively for a channel with 20000 ps/nm residual CD (left) and histogram of estimation error with both methods for 5000 random channel evaluations (right).

potential CD compensation functions $H_{CD,i}(f)$, by computing only parts of the complex ACF, and by applying the sign-algorithm, which results in a multiplier-free ACF. Only the filtering process $R_i(f) = R(f) \cdot H_{CD,i}(f)$ requires complex multiplications, which can be processed in parallel due to the FD implementation.

## 3. Estimation Performance

We evaluate the performance with respect to the estimation error. Scanning through CD compensation functions $H_{CD,i}(f)$ has been applied in a range of $\pm$32000 ps/nm with a stepwidth of 200 ps/nm and 100 ps/nm based on simulations and offline data respectively. For simulations, 5000 channels have been randomly generated by distributions given in Tab. 1 (see also [6] for details). The estimation error for $J_{CD,min}$ proves a standard deviation (std) of 103 ps/nm and a maximum error of 400 ps/nm. Under identical conditions, the improved estimation $J_{CD,max}$ shows a standard deviation of 74 ps/nm with a maximum deviation of 192 ps/nm. Various offline experiments with WDM transmission summarized in Tab. 2 have been carried out. The results demonstrate excellent precision and robust estimation, even in the nonlinear transmission regime at large launch powers.

| channel description | captures # | std [ps/nm] | mean [ps/nm] | min [ps/nm] | max [ps/nm] |
|---|---|---|---|---|---|
| 100G PDM-QPSK & 10G OOK, 1200km SSMF+DCF | 22 | 48 | 4 | -100 | +100 |
| 100G PDM-QPSK & 10G DQPSK, 900km SSMF+DCF | 24 | 50 | 108 | 0 | +200 |
| 100G PDM-QPSK & 40G DQPSK, 1200km SSMF+DCF | 48 | 46 | 8 | -100 | +100 |
| 100G PDM-QPSK WDM, 1200km SSMF (w/o DCF) | 30 | 39 | 8 | -100 | +100 |
| 100G PDM-QPSK WDM, 1500km SSMF (w/o DCF) | 82 | 55 | 80 | -200 | +100 |

Tab. 2. Estimation performance from offline experiments for launch powers from -2 to +5 dBm with OSNR values from 15 to 21 dB respectively.

## 4. Conclusion

We demonstrated reliable and precise FD CD estimation in presence of linear and nonlinear channel distortions based on simulations and offline experiments. A maximum estimation error of $\pm$200 ps/nm was achieved with a worst-case standard deviation below 75 ps/nm. At the same time, the cost function based on the clock-tone magnitude reduces the implementation complexity tremendously. This allows for a fast estimation without large delays suitable for the initialization of the digital coherent receiver in switched optical networks.

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