Efficient Training-Based Channel Estimation for Coherent Optical Communication Systems

Fabio Pittalà[1,2,3], Amine Mezghani[3], Fabian N. Hauske[1], Yabin Ye[1], Idelfonso T. Monroy[2], Josef A. Nossek[3]

1: European Research Center, Huawei Technologies Duesseldorf GmbH, Riesstrasse 25, D-80992 Munich, Germany,
2: DTU Fotonik, Technical University of Denmark, Ørsteds Plads 343, DK-2800 Kgs. Lyngby, Denmark,
3: Institute for Circuit Theory and Signal Processing, Technische Universität München, Arcisstr. 21, D-80290 Munich, Germany
fabio.pittala@huawei.com

Abstract: A low-complexity technique for frequency domain channel estimation based on constant-amplitude zero-autocorrelation (CAZAC) sequences is theoretically investigated.

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1. Introduction

One of the most efficient approaches for multi-input multi-output (MIMO) channel estimation in reconfigurable coherent optical networks with higher-order modulation, flexible switching and wavelength routing requires to periodically transmit training sequences (TS) between the payload of data. The channel is then estimated by means of the received data and the known transmitted TS (in the literature, training-based, pilot-assisted or data-aided channel estimation refer to the same technique).

The TS structure provides knowledge about the channel state information (CSI) and therefore its length should be longer than the channel impulse response (CIR) to be estimated. To keep the requirement for short TS, chromatic dispersion (CD) compensation should be performed before the training-based MIMO equalizer. In addition, the TS period needs to be selected according to the specified state of polarization (SOP) tracking and processing delay.

In this paper we investigate a novel method for frequency domain (FD) channel estimation based on CAZAC sequences. CAZAC refers to the constant amplitude (CA) feature due to the fixed norm property of the roots-of-unity class of codes and to the ideal (impulse-like) periodic zero-autocorrelation (ZAC) property of the code set [1].

2. CAZAC Training Sequences

Optimum channel estimation can be performed by employing perfect-squares minimum-phase (PS-MP) CAZAC sequences, described by Eq. (1):

$$c[n] = \exp\left\{ j\frac{2\pi}{\sqrt{N}}\left[\mathrm{mod}\left(n-1,\sqrt{N}\right)+1\right]\left[\left\lfloor\frac{n-1}{\sqrt{N}}\right\rfloor+1\right]\right\}, \qquad \text{where:} \quad n = 1,2,\ldots,N. \tag{1}$$

For sequences of length $N = 2^{2m}$ symbols with $m \in \{1,2,3,\ldots\}$, the constellation plot refers to a phase-shift keying (PSK) modulated signal (i.e. $m = 1$ refers to BPSK, $m = 2$ to QPSK, $m = 3$ to 8PSK and so on).

Since the PS-MP CAZAC sequences have even lengths (power of 2), the use of fast Fourier transform (FFT) operations keep the receiver complexity significantly lower than that of conventional single carrier systems with time domain (TD) equalizer structure [2]. In principle, the modulation of the TS and of the payload data is independent assuming that the modulator allows to generate all relevant constellation points of both, the TS and the payload data.

For a 2×2 MIMO, the TS can be composed of two independent blocks $c_1[n]$ and $c_2[n+N/2]$ sent simultaneously one per polarization. In addition, the TS can be (optionally) framed by guard intervals which continuously pursue the adjacent sequence. The maximum length of the CIR that can be estimated is $N/2$.

3. Channel Estimation based on CAZAC Training Sequences

A 2×2 MIMO system can be described in FD by the following matrix vector multiplication:

$$\mathbf{R}[\omega_k] = \mathbf{H}[\omega_k]\times\mathbf{S}[\omega_k] \;\Rightarrow\; \begin{bmatrix} R_1[\omega_k] \\ R_2[\omega_k]\end{bmatrix} = \begin{bmatrix} H_{11}[\omega_k] & H_{12}[\omega_k] \\ H_{21}[\omega_k] & H_{22}[\omega_k]\end{bmatrix} \times \begin{bmatrix} S_1[\omega_k] \\ S_2[\omega_k]\end{bmatrix}. \tag{2}$$

In Eq. (2), $\mathbf{R}[\omega_k]$ is the complex valued received signal, $\mathbf{H}[\omega_k]$ describes the channel transfer function matrix and $\mathbf{S}[\omega_k]$ is the transmitted training signal.

At the receiver, framing synchronization extracts the TSs from the incoming stream of data $\mathbf{r}[n]$ and after a discrete

FFT, a full channel estimation can be performed from the sent CAZAC spectra $S_1[\omega_k]$ and $S_2[\omega_k]$ and the according received spectra $R_1[\omega_k]$ and $R_2[\omega_k]$:

$$\widehat{\mathbf{H}}[\omega_k] = \begin{cases} \widehat{H}_{11}[\omega_k] = R_1[\omega_k]S_1^*[\omega_k] \\ \widehat{H}_{12}[\omega_k] = R_1[\omega_k]S_2^*[\omega_k] = R_1[\omega_k]S_1^*[\omega_k]e^{j\frac{N}{2}\omega_k} = \widehat{H}_{11}[\omega_k]e^{j\frac{N}{2}\omega_k} \\ \widehat{H}_{22}[\omega_k] = R_2[\omega_k]S_2^*[\omega_k] \\ \widehat{H}_{21}[\omega_k] = R_2[\omega_k]S_1^*[\omega_k] = R_2[\omega_k]S_2^*[\omega_k]e^{j\frac{N}{2}\omega_k} = \widehat{H}_{22}[\omega_k]e^{j\frac{N}{2}\omega_k}. \end{cases} \tag{3}$$

In Eq. (3), the phase relation between $\widehat{H}_{11}[\omega_k]$ and $\widehat{H}_{12}[\omega_k]$ (and therefore $\widehat{H}_{21}[\omega_k]$ and $\widehat{H}_{22}[\omega_k]$) is due to the orthogonality property of the TD CAZAC sequences. This effect is clearly visible from the TD representation of the estimated channel elements illustrated in Fig. 1, which is translated into a $N/2$-symbol circular shift.

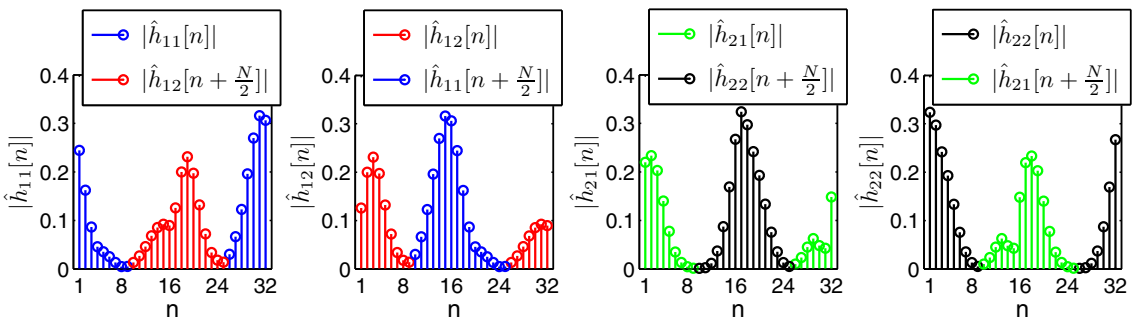

Fig. 1. Time domain channel matrix before windowing the channel impulse response

To eliminate redundancy and to guaranty orthogonality between the channel coefficients at each frequency point, we window the CIR as described by Eq. (4). The complexity of the receiver can be further reduced by computing just $\widehat{H}_{11}[\omega_k]$ and $\widehat{H}_{22}[\omega_k]$ from Eq. (3) and their corresponding inverse discrete FFTs $\widehat{h}_{11}[n]$ and $\widehat{h}_{22}[n]$.

$$\begin{cases} \widehat{H}'_{11}[\omega_k] = FFT\left\{\widehat{h}_{11}[n]\mathrm{rect}[n]\right\} \\ \widehat{H}'_{12}[\omega_k] = FFT\left\{\widehat{h}_{11}[n+\frac{N}{2}]\mathrm{rect}[n]\right\} \\ \widehat{H}'_{22}[\omega_k] = FFT\left\{\widehat{h}_{22}[n]\mathrm{rect}[n]\right\} \\ \widehat{H}'_{21}[\omega_k] = FFT\left\{\widehat{h}_{22}[n+\frac{N}{2}]\mathrm{rect}[n]\right\}, \end{cases} \tag{4}$$

where $\mathrm{rect}[n]$ takes 0 for $N/4+1 \leq \mathrm{mod}(n-1,N) \leq 3N/4$ and 1 elsewhere. The estimated channel matrix $\widehat{\mathbf{H}}'[\omega_k]$ is finally used to implement the 2×2 MIMO FD equalizer. Zero-forcing (ZF) and minimum-mean-square-error (MMSE) solutions have been considered in our investigations. Performance of CAZAC-based FD equalization has been recently shown in [3], whereas powerful optical performance monitoring (OPM) has been reported in [4] [5].

4. Conclusions

We theoretically investigated a low-complexity CAZAC-based frequency domain channel estimation method for coherent optical communication systems. The low-complexity architecture and the feed-forward training-based filter update make the FD equalizer highly suitable for implementation in high-speed optical receivers.

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