LETTER

Performance Analysis of an Ultra-wideband Decoder for Spectrally Encoded ECDMA on Passive Optical Network

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Abstract A new method of electronic code division multiple access (ECDMA) for passive optical network (PON) applications is investigated. An ultra-wideband (UWB) decoder is proposed to make available spectral amplitude-coding (SAC), as an alternative to direct sequence (DS). Code modulation waveforms implemented in multi-Gchip/s transversal filters allow the evaluation of impulsive responses of the PON structure. Impairments that impact the transmission of spectral users and thus the correct decoding of SAC signals are analyzed. The signal-to-interference ratio (SIR) is derived assuming that simultaneous users are structured as shot noise pulse-trains. It is seen that the scheme improves interference rejection over previous proposals. Photonic system simulation shows the viability of the electronic processing.

key words: spectral amplitude-encoded CDMA, passive optical network, UWB receiver, distributed-based transversal filter.

Classification: microwave photonics.

1. Introduction

For the past years, there has been tremendous interest in applying electronic processing of codes to high-speed CDMA networks. In addition to being a low-cost alternative to optical processing, the electronic method is well-suited for PON applications as it improves receiver sensitivity due to coding gain [1] and suppresses beat noise effects [2]. Sensitivity influences splitting ratios, hence the number of optical network units (ONUs) connected to an optical line terminal (OLT). DS offer potential benefits to high speed OCDMA networks such as asynchronous transmission, low-delay access, potential security; among others [3], [4]. So far, electronic code multiplexing has only been considered for synchronous networks since multiple access interference (MAI) becomes excessive [5], [6] and [7], mainly due to the processing constraints that impose multi-GHz electronic transversal filters.

In this work, SAC is introduced as an attractive option for ECDMA owing to its ability to cancel MAI [8]. The capacity to accumulate a sufficient amount of non-coherent energy from short codes facilitates electronic encoding and decoding. The UWB decoder is implemented as an energy detector (ED) structure [9] and is provided with distributed transversal filters (DTFs) in both signal paths to selectively accumulate energy from spectral users. Users set a common phase reference at the receivers to perform phase-to-intensity conversion previous to the despread operation. ECDMA involves the generation of complex analog pulse patterns, which can be accomplished by transversal FIR filters [10],[11], [12]. To make available SAC on PON, it is crucial that the structure supports code modulation formats going beyond ON-OFF keying temporal patterns. The user pulse format should be compatible with multilevel phase-modulated formats and square-law photo-detection [13].

In this Letter, we evaluate numerically the PON structure by analyzing impulsive responses of a broadband channel that transmits short pulses falling into the sub-nanosecond region. This also includes encoders and decoders designed as monolithic microwave integrated circuits (MMICs). Pulse distortion, which is largely influenced by chromatic dispersion of the optical fiber, is computed to give assessment of the proper type of optical fiber for the application. Later, the SIR is derived to show the capacity to detect matched user under interference conditions. Photonic system simulator OptSim™ is used to analyze SAC-ECDMA-PON. Performance limitations of the decoding are discussed towards the end of this paper.

2. UWB decoder

An understanding on the electronic SAC can be gained from taking into account similarities between our proposal and those of the balanced optical decoder, well known in the realm of OCDMA [3], [8]. A schematic of SAC-ECDMA is depicted in Fig.1. The DTF architecture employed for high-speed encoder and decoder implementations has been reported in [10]. SAC works on the assumption that energy collected from users adds non-coherently [14]. For instance, Hadamard codes imposed on user amplitudes, as represented by slot boxes in Fig.1, can overlap at certain frequencies. If signal components fall in receiver frequency slots of the
same phase, the decoder produces a pulse of amplitude proportional to the energy of input sub-bands. Contrariwise, for unmatched users, half of the energy of users falls in receiving frequency slots of a given phase and the other half coincides with complementary slots of different phase leading to full cancellation of MAI.

\[
\sum_{k=1}^{M} c_k^{(i)} \cos(2\pi f_k t + \theta_k)
\]

with \( p(t) \) being the gate function of duration \( T \) and \( f_k \) is the \( k \)-th harmonic frequency, \( \theta_k \) is the initial phase, and \( w \) is code weight; that is, the number of harmonically related sinusoids comprising a user pulse. The factor \( \sqrt{w/2} \) normalizes the encoded signal with its root-mean-square (rms) value and then all users have unit power before setting a given modulation index at the Mach-Zehnder (MZ) modulator. Front-end DTFs in both receiver paths insert phase offsets to input harmonics. The phases of the transmitted sinusoids are the same to those of the decoder vector \( \theta = (\theta_1, \theta_2, \ldots, \theta_M) \) [10] so as to match phases and generate a local receiver reference, \( z_s(t) \), with which demodulate the correlated pulse \( z_d(t) \) (see Fig.1). The transversal FIR filter function at the upper arm can be approximated in the frequency domain by:

\[
H_{\text{app}}(f) = \sum_{k=1}^{M} \exp(j\varphi(f)) \text{rect} \left( (f - f_k)/\Delta \right)
\]

where \( \varphi(f) \) is the upper DTF function resulting in the phase shifting vector \( \varphi_0, \varphi_2, \ldots, \varphi_M \) at fixed harmonic frequencies, and \( \text{rect}(\cdot) \) is the rectangle function with \( \Delta \) being equal to the separation between neighboring frequency bin centers. The FIR function at the bottom arm of the decoder \( (H_{\text{bot}}) \) has a similar equation to the upper function, just need to substitute the phase function \( \theta(f) \) into Eq. 1.

For DTF implementations, FIR functions are truncated in time by the symbol interval [10]. This results in both coded main sub-bands and intercode interference, the latter in the form of sub-band side-lobes extending over the bandwidth. However, the effect of intercode interference can be reduced by making available in all decoders the FIR function that depends on the alternate vector \( \theta = (\pi/2, 0, \pi/2, \ldots) \) [11]. Fig. 2 depicts the signature function and receiver-decoder amplitudes of the decoder. Note that the first sub-band (around 0-Hz) is uncoded to allow the despread pulse to emerge after mixing both filter responses, as shown later.

![Fig. 1 SAC-ECDMA on PON. Intensity modulation of users (upper) and two-arm electronic decoding (bottom).](image)

![Fig. 2 Frequency responses of front-end DTFs. Relative phase function (upper) and transimpedance amplitude responses (bottom).](image)

Commonly, the overall voltage gain of transversal FIR filters based on the principle of distributed amplification is low (see for instance [15], [16]) and thus an amplifier providing the required optical-to-electrical conversion gain needs to be included. We chose the transimpedance amplifier (TIA) as the first block of the receiver to provide a high gain \( (Z_{\text{TIA}}) \), for instance; exceeding 40 dB \( \Omega \) [17], to convert signals from current to voltage domain. This selection provides adequate receiver input sensitivity [18], [19] for SAC on PON, as evaluated in a later stage.
For a spectral user $U$, the decoder computes the convolution product between frequency aligned responses of DTFs. The despread output pulse is:

$$Z = \int_{-\infty}^{\infty} H_{\text{upp}}(f) H_{\text{bott}}^*(f) |U(f)|^2 \, df$$

with $^*$ denoting complex conjugated. The energy spectral density $|U(f)|^2$ is the received version of the transmitted user spectrum, which is computed as the magnitude of the Fourier transform of the autocorrelation function of Eq. 1. In the following, to facilitate the correlation analysis, the filter responses are normalized by

$$\int_{-\infty}^{\infty} H_{\text{upp}}(f) H_{\text{bott}}^*(f) \, df = 1.$$

By substituting the frequency functions and dividing Eq. 3 by the time duration of correlated signals (27), the output becomes a combination of power contributions added in phase and anti-phase. Then, the decoded output pulse, $z$, is:

$$z = P_{in} \frac{1}{M} \sum_{n=1}^{M} c_n \cos(\varphi_n - \theta_n)$$

where $P_{in}$ is the average power of the input user pulse.

3. SAC-ECDMA on PON

The PON structure is aimed for local area networks (LANs). The aggregate CDMA signal is created at the RF level and, at the photodetectors, a single optical carrier beats with the CDMA signal to down-convert the modulated optical spectrum into the RF domain. Light emitted from a single laser diode is being modulated for each bidirectional link between ONUs and OLT. On the other hand, with optical fiber links stretching over few tens of miles, fiber dispersion impairs significantly transmission of SAC signals. Previous work on DS-ECDMA shows the effect of dispersion of the optical fiber [20] and noise of the post detection amplifier [21] on the bit-error rate (BER). Nevertheless, inadequacies of encoders and decoders, such as concatenation between pulses due to significant bandwidth limitation of the utilized FIR filters [1], is not considered into the analysis [20] [21].

The suitability of DTF design for high-speed CDMA was assessed in [22]. DTFs in back-to-back configuration proves that code modulation sequences can help to stipule a coded channel that satisfies zero intersymbol interference (ISI). The impulsive response obtained from periodic reciprocal sequences [23] is applied to the PON structure by including DTFs realizing encoding and decoding. The maximal-length sequence $\{1,1,1,0,0,1,0,1\}$ is generated using binary modulation of the optical intensity ON-OFF keying. A chip-rate DTF encoder drives the MZM with a non-return to zero (NRZ) pattern showing minimal inter-pulse interference. This lets the modulator switching from full transmission to an extinction level set by the modulator ratio ($r_c$). Optical pulses are then photo-detected and in turn passed through the DTF-decoder, which is set to the bipolar reciprocal sequence $\{1,-1,1,-1,-1,1\}$.

Fig. 3 shows periodic impulsive response of the link at the bit rate of 2.85 Gb/s. DTFs have span time of 0.35 ns. Simulations provide accurate calculation of dispersion induced by optical lines. Table 1 itemizes some parameters. The test compares standard single-mode fiber (SSMF, standard ITU-G.652) with nonzero dispersion-shifted fiber (NZDSF, standard ITU-G.655). Eye diagrams and full-width at half-height (FWHH) measurements favor the selection of NZDSF over SSMF.

![Fig. 3 Eye at the decoder output after 20-km optical fiber line, SSMF (left) and NZDSF (right).](image)

**Table 1 Simulation parameters**

<table>
<thead>
<tr>
<th>Description</th>
<th>Value</th>
</tr>
</thead>
<tbody>
<tr>
<td>Wavelength</td>
<td>1550 nm</td>
</tr>
<tr>
<td>Optical fiber length</td>
<td>20 km</td>
</tr>
<tr>
<td>PIN responsivity</td>
<td>0.9 A/W</td>
</tr>
<tr>
<td>MZM extinction ratio ($r_c$)</td>
<td>20 dB</td>
</tr>
<tr>
<td>MZM switched voltage ($V_{n1}$)</td>
<td>18 V</td>
</tr>
<tr>
<td>Transimpedance gain ($Z_{in}$)</td>
<td>42.0 dB</td>
</tr>
<tr>
<td>Input-referred noise current($I_{ref,\lambda}$)</td>
<td>16.0 pA/\text{Hz}</td>
</tr>
<tr>
<td>Noise bandwidth (dB/Hz)</td>
<td>20 GHz</td>
</tr>
</tbody>
</table>

Fig. 4 shows the impulsive response of ECDMA over NZDSF optical link and allows contrasting with results of MMICs. The periodic response has amplitude peaks practically equal to the in-phase amplitude of the aperiodic cross-correlation function (see inset). The amplitude matching allows confirming zero ISI in agreement with the role that plays the aperiodic cross-correlation function on synchronous CDMA [24]. Both aperiodic responses approximate the bipolar reception of ON-OFF keying pulse sequence. Notwithstanding, the impulsive ECDMA function exhibits low side-lobes due to fiber-induced distortion.

![Fig. 4 Simulated periodic responses of the ECDMA and MMIC DTFs in back-to-back configuration. (Inset) Aperiodic responses.](image)
Fig. 5 depicts results of the SAC-PON over SSMF optical links. To illustrate the tolerance of SAC waveforms to chromatic dispersion, pulse widths were measured from impulsive responses of the PON at different fiber link lengths. Insets show mixer outputs. Results were normalized with the matched pulse amplitude obtained at 15 km fiber-link length. To make fair comparisons, the attenuation exhibited by longer optical fibers was compensated by increasing CW laser power accordingly. For distances above 18 km, pulse dispersion leads to a reduced detected intensity of the matched code while more energy spreads over time in the form of side-lobes. Interference also varies significantly with fiber-link length. For clarity, the figure shows only the worst interference case. This means that the desired frequency alignment between photo-detected pulses and decoder function described by Eq. 3 cannot be sustained.

![Fig. 5 Effect of dispersion on the reception of matched and interfering codes. Pulseswidth of PON responses varies with fiber link distance.](image)

Given the detrimental effect of chromatic dispersion on decoding, optical lines are implemented with NZDSF type as it provides smaller chromatic factors [25].

### 4. Multi-user interference in ECDMA

The decoder depicted in Fig. 1 captures the non-coherent energy from spectral users and provides a viable way to reject multiple user interference (MUI), which is a central issue in DS [1],[6] as well in SAC-ECDMA systems [14]. An effective approach to SAC-ECDMA relies on driving each encoder with “high density” shot noise impulses [26] to accumulate a sufficient amount of energy while reducing, on average, the induced interference. It is stated in [14], that full rejection of MUI can be achieved providing that the amount of collected energy at a given frequency slot is equal to the addition of energy levels of each coded sub-band that composes the aggregate input signal at such frequency.

For practical SAC, pulse-trains are generated by passing Poisson impulses through encoding filters. The random variates \( \{ \tau_{0} \} \) used at the decoders have an exponential density function with parameter \( \lambda \). Upon reception, the random timing of concatenated pulses is hopped over a decoding window of duration \( 2T \) in correspondence with the symbol time. Pulse are then decoded by taking advantage of accurate synchronization incorporated in the structure. Energy detection refers to the demodulation of filter pulse responses that show the same instantaneous phases at both arms of the structure. Contrariwise, additive output noise results from the demodulation of pulses that travel at different delays along transmission lines of both DTFs.

To derive the SIR, the computation of user power assumes a fixed average rate parameter \( \lambda \). The non-coherent power at the \( k \)-th sub-band, \( \langle \xi_k \rangle \), is the expected averaged energy from \( K \) active users accumulated over a given decoding time. If each user has the same electrical power \( (P_{in}) \), the expected non-coherent power is:

\[
\langle \xi_k \rangle = \lambda \left( c_{k}^{(l)} \right)^2 + \lambda \sum_{j=2}^{K} \left( c_{j}^{(l)} \right)^2 .
\]

(5)

The decoder output depends on the specific properties of the code. For bipolar decoding, interfering codes are rejected as zero output [8] and therefore the matched user is decoded as a pulse of amplitude \( \lambda P_{in} \). The expected average rate \( \lambda \) thus corresponds to the detection gain of a pulse burst. Interference, on the other hand, adds on the detected non-coherent energy as a random fluctuation. The power of this fluctuation is the average-squared deviation from the expected power \( \langle \xi_k \rangle \).

For analysis, the decoder operates on two independent random variables, \( t_1 \) and \( t_2 \) and the output can be written as the average:

\[
z/2T = \sum_{k=1}^{M} \left[ \langle \xi_k \rangle + C_{st,k} (t_1, t_2) \right] \cos (\varphi_k - \theta_k) .
\]

(6)

where \( C_{st,k}(t_1,t_2) \) is the auto-correlation. Each received sequence is formed by correlating impulse trains with coded and decoded pulse shape \( s_k(t) \). Output noise contribution results from the demodulation of the \( k \)th sub-band given by:

\[
C_{st,k} (t_1, t_2) = E \left[ s_k (t - t_1) s_k (t - t_2) \right] ,
\]

(7)

where \( E[\cdot] \) denotes mean value.

MUI becomes a combination two effects, namely code-crosstalk and accumulated correlation noise [14]. Code-crosstalk occurs at the sub-bands that present code overlapping and is characterized by the cosinusoidal dependence on the relative phase between delayed pulses:

\[
\sigma_{s_{0,k},k}^{(n,m)} = c_k^{(l)} c_k^{(l)} \sum_{q=0}^{L-1} \sum_{p=0}^{L-1} \rho \left( \xi_{q,p}^{(m)} \right) \cos \left( \omega_k \xi_{q,p}^{(n,m)} \right) .
\]

(8)

where the time separation variable \( \xi_{q,p}^{(n,m)} = \xi_{q}^{(n)} - \xi_{p}^{(m)} \) is also a Poisson process, \( L \) is the number of points and \( \rho(\cdot) \) is the normalized autocorrelation function of the gate function. For the case of accumulated correlation noise, which refers to interactions between overlapped pulses of the same burst, the covariance evaluates fluctuations due to random lags.
\[ \delta_{q,p}^{(n,n)} = r_q^{(n)} - r_p^{(n)} \] and this can written as:
\[ \sigma_{\text{corr},k}^{(n)} = (c_k^{(n)})^2 \sum_{q=0}^{L-1} \sum_{p=0}^{L-1} \rho(\delta_{q,p}^{(n,n)}) \cos(\omega_k \delta_{q,p}^{(n,n)}) \cdot (9) \]

The signal detection quality of the decoder is given by:
\[ \text{SIR} = \left( \frac{\lambda P_{\text{in}}}{\sigma_{\text{th},0}^2} \right)^2 \]
\[ \left( \frac{P_{\text{in}}}{W} \right)^2 \sum_{k=1}^{M} \left( \frac{\sigma_{\text{corr},k}^{(n)} + \sum_{n=1}^{K} \sigma_{\text{stnl},k}^{(n,m)}}{\sigma_{\text{th},0}^2} \right)^2 \]

where \( \sigma_{\text{th},0}^2 \) is the thermal noise power contribution of the TIA. Using parameters in Table 1 [17] and Optsim simulation, the computed power of correlated noise from the TIA is equal to 1.23 \times 10^{-3} [V^2].

The squared input user power in Eq. 10 stems from using the ED scheme [9]. Thus, for the PON application, detected power may be of low magnitude and even comparable to noise power of the TIA. Fig. 6 shows SIR evaluated for matched user alone and when interferers fully overlap with the user. The RF drive signal is varied from a low signal-to-noise ratio \( \frac{P_{\text{in}}}{\sigma_{\text{th},0}^2} \) to higher power level, at which the MZ modulator can meet the small-signal modulation condition. The figure depicts the benefit of ECDMA as input power approaches to \(-18 \text{ dBm}\). The large average rates are intended to improve orthogonality among users, whereas the number of constructed codes depends only on the available frequency bins. Table 2 gives an account on the decoding results for up to 3 simultaneous codes.

The detection of matched code is less prone to additive thermal noise for higher average rates. Also, interference associated with each user varies with the rate parameter and cannot be reduced further due to the inability to accumulate shot noise sequences from simultaneous users in a power basis. This outcome has also been reported for OCDMA based on broadband optical noise sources, see for instance [27] and [28]. Hence, for our electronic scheme, induced interference cannot ever reach the theoretical shot-noise limit.

Now, the SAC-ECDMA-PON system was simulated using OptSim including pulses shaped by MMIC waveforms. For transmission of the composite CDMA RF signal, the MZ modulator transfers power from a CW laser to sideband components about the optical carrier frequency \( (\nu_0 \pm \alpha) \). The dynamic range of the spectrally encoded signals is increased due to the desired concatenation between pulses in a burst. Upon reception, the maximum power of photodetected signal corresponds to the effective RF drive signal power without the detected carrier, which is suppressed after amplification by a fourth-order Bessel bandpass filter (BPF). This filter improves the limited attenuation of the front-end DTFs around 0 Hz (see Fig. 2). The bandpass RF pulse is down-converted assuming good carrier suppression.

Fig. 7 displays mixer outputs of matched code alone and two simultaneous interferers (top and middle, respectively). The noise-like signals have different spectra which are translated to differentiated power levels at the output (Fig. 7 bottom left). For matched code, the decoder acts as a square-law detector and hence the lowpass filter (LPF) output accounts for the non-coherent energy accumulated in the envelope of the unipolar signal. Interfering codes, on the other hand, produces mixer output whose average tends ideally to zero. The despread signal selected by LPF is a combination of non-coherent energy levels at different sub-bands in phase and anti-phase, as resolved by the decoder function.

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**Table 2** SIR evaluation at \( P_{\text{in}} = \text{-}18 \text{ dBm} \).

<table>
<thead>
<tr>
<th>( \lambda_1 )</th>
<th>( \lambda_2 )</th>
</tr>
</thead>
<tbody>
<tr>
<td>Matched code</td>
<td>13.2 dB, 18.3 dB</td>
</tr>
<tr>
<td>Single interferer</td>
<td>8.05 dB, 10.74 dB</td>
</tr>
<tr>
<td>Two interferers</td>
<td>4.7 dB, 8.2 dB</td>
</tr>
</tbody>
</table>

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It is seen that most of the energy of both noise-like signals appears in a time interval of \( 2T \) (0.7 ns). In a similar way to time-spreading OCDMA [29], it is apparent to assume some
“time gating” operation to set the integration time in agreement with the lowpass filtering of the despread pulse. Fig. 7 (bottom right) shows different levels of interference for single interfering codes and depends to great extent on the generation pulse bursts that satisfy the random properties of shot noise.

Finally, it is sensible to put into perspective the obtained detection peak (about 3:1, see Fig. 7 bottom left) regarding previous methods. When using temporal encoding DS with Gold sequences [1], the autocorrelation peak is 7 and the cross-correlation takes on $\{-1,-5,3\}$ for a single interferer [30]. On the other hand, the unipolar ECDMA in [6] produces a detection peak of 3:1 respect to its correlation sidelobes for a single user.

6. Conclusion

We have presented a new technique for ECDMA based on non-coherent detection of spectral users. To support multilevel phase-modulated formats, numeric simulation shows that the PON structure should employ optical transmission lines with low dispersion coefficients, such as the NZDSF type. It is seen that trade-off between sensitivity of post-detection receiver and transmitted power of the MZ modulator can be satisfied so that the proposed processing can perform the multi-access capability.

References


