MICROWAVE AND OPTICAL TECHNOLOGY LETTERS

Performance of a Noncoherent Decoder for Spectral Amplitude-Coding Electronic-Code Division Multiple Access

Journal:	*Microwave and Optical Technology Letters*
Manuscript ID	MOP-19-1135.R1
Wiley - Manuscript type:	Research Article
Date Submitted by the Author:	n/a
Complete List of Authors:	Aguilar-Torrentera, Jorge; Universidad Autónoma de Nuevo León, Posgraduated Section of Electrical Eng. Galvan-Tejada, Giselle; Center for Research and Advanced Studies of the National Polytechnic Institute, Communication Section Orozco-Lugo, Aldo; Center for Research and Advanced Studies of the National Polytechnic Institute, Communications Section Darwazeh, Izzat; University College London, Institute of Communications and Connected Systems
Keywords:	spectral amplitude coding, ultra-wideband receivers, shot noise, electronic processing of codes, distributed-based transversal filter

Performance of a Noncoherent Decoder for Spectral Amplitude-Coding Electronic-Code Division Multiple Access

Abstract— The performance of a noncoherent decoder based on a pair of distributed-based transversal filters architecture is reported. Analysis based on experimental results validates the decoder as a proof-of-concept using microwave integrated circuits. Front-end transversal filters are designed with large degree of freedom to generate and correlate temporal phase-coded signals. Receiving functions are aligned with slots of spectrally encoded channels to make available spectral amplitude-coding electronic-code division multiple access. Noncoherent ultra-wideband receiver concepts are applied to model the decoder synchronized to shot noise pulse-trains. The despreading operation involves capturing sufficient amount of energy to demodulate matched and unmatched users. Electronic transversal filters implemented with a suitable number of taps allows bipolar detection of concurrent users whereas the noncoherent reception of energy of subbands is fundamental to achieve orthogonality among channels, as we show in this paper.

Key Words— spectral amplitude coding, ultra-wideband receivers, shot noise, electronic processing of codes, distributed-based transversal filter.

I. INTRODUCTION

In recent years, there has been tremendous interest in applying code division multiple access (CDMA) to high-speed optical access networks. Among the most successful schemes for optical CDMA (OCDMA) are those based on spectral amplitude-coding (SAC) since multiple access interference (MAI) can be completely canceled using simple direct-detection receivers and low-cost noncoherent optical sources [1]. On the other hand, considerable research on electronic processing of codes has been devoted to time domain direct-sequence (DS) systems for optical access networks (see for instance [2-4]), in which the distributed-based transversal filter (DTF) is the cornerstone of encoding and decoding. It is well known that rise time constrains inherent to the gain-bandwidth product of distributed structures set a limit on the realizable span time of the tapped delay line, hence codes programmed in transversal filter prototypes must be short (typically with codewords of 8 chips). Integrated transversal filters performing matched filtering increase sensitivity in the reception of intentional users due to coding gain [2]. Similarly, bipolar detection of unipolar pulses has also been implemented for electronic-CDMA on passive optical networks [4]. Nevertheless, channel multiplexing introduces a substantial degradation as more pseudo-orthogonal codes are added to the network [2] posing a significant negative impact on asynchronous operation. When performing intensity correlation with balanced orthogonal codes [4], the scheme requires synchronization at the chip interval to remove side-lobe time correlation interference. All the above facts evidence that new tradeoffs between performance and complexity need to be found.

In this paper an alternative to DS, termed electronic SAC, which enhances the capacity of *electronically-processed codes* to differentiate between concurrent users is proposed. A key advantage concerning electronic SAC is that channel orthogonality does not necessarily mean performing encoding and decoding with long finite-impulse-response (FIR) transversal filters. The detection of encoded pulses requires front-end DTFs to work as pulse-shaping structures instead of processors for DS sequences in which prototypes operate as symbol-rate transversal filters. The processing has been scaled up using two FIR filters shifted in time to improve coding gain and increase the potential number of pseudo-orthogonal codes [3] however high interference impairs asynchronous code multiplexing. Recent developments in electronic SAC include experimental assessment of DTFs to correlate signals that depend on chip spectral sequences and a decoder architecture that collects energy of received subbands was also introduced in [5]. Fig. 1 shows a schematic of the decoder based on a pair of front-end DTFs customized to correlate temporal phase-coded signals. User despreading makes use of a broadband signal multiplier to mix the address ($z_{\varphi}(t)$) with the reference ($z_{\theta}(t)$). Fig. 1 also depicts the mixer output for code matching which results fundamentally in a unipolar pulse and its envelope after lowpass filtering. So far, rejection of interference and code matching have been demonstrated under the assumption that user and interference do not overlap in time [5].

FIGURE 1.

Throughout this paper, user data is encoded using temporal phase-coded signals. Code information is conveyed on the amplitudes of subbands while their phases are dropped at the detection stage to achieve noncoherent decoding. We refer to the transmitted spectrum as a set of amplitude-coded subbands corresponding in the time domain (via the inverse Fourier Transform) to a broadband pulse. See for instance the transmitted spectrum of an encoded broadband pulse at the left side of Fig.1. For generation with a FIR filter, the pulse is truncated by a time window of duration equal to the symbol interval to suit the limited time span of distributed structures. Side-band interference, which results from time truncation, extends over the created frequency slots and becomes a contributing effect on the spectral amplitude ranges of densely packed subbands.

The present work extends the electronic approach to allow users to overlap, which is a central issue in the performance of CDMA networks. A complete study on the overall system design encompassing an assessment of the optical channel and the decoder intended to work as a postdetection receiver would exceed the length of a paper like this. We choose to concentrate to a new spectral decoding method based on a noncoherent detection approach and its implementation with electronic FIR filters. Ultra-wideband (UWB) concepts [6] applied to well-established decoding methods for CDMA radio in high multipath environments [7] appear to be suitable for the proposal. In the present approach, the decoding is performed by the two-branch structure when each user is structured as a burst of temporally randomized pulses. The detected pulses should maintain a predetermined shape to selectively accumulate energy of received shot noise pulse-trains. It is seen that in order to attain orthogonality among users, the decoder has to work as a noncoherent UWB receiver of spectrally encoded users. The presented analysis is based on measurements of microwave integrated circuits (MICs) and signal synthesis using an arbitrary waveform generator (AWG).

II. FRONT-END DTFS

In the decoder scheme, the responses of front-end DTFs remain spectrally spread for all users. The SAC method makes available a number of orthogonal channels provided that the transmitted components are aligned in frequency with the receiver functions. Additionally, the despreading operation involves convolving with filter responses characterized by the same harmonic frequencies but different initial phases. Notice that the spectra of both DTFs responses are correlated by the signal multiplication, however only the correlation product between frequency-aligned components will fall into the lowpass filter (LPF) bandwidth. Fig. 2 illustrates the filter method when used to approximate an analog pattern by an electronic transversal filter featuring unequal tap gains and unequal interstage delays [8]. At instantaneous time t_k, the filter output depends on the pulse response of the transversal filter weighted by a tap gain coefficient G_k and the responses of previous taps. The filter output pulse, $h_{DTF}(t)$, can be written as a continuous time response:

$$h_{DTF}(t) = \sum_{n=1}^{N} G_n p(t - t_n) = \sum_{n=1}^{N} G_n p\left(t - \sum_{k=1}^{n} \tau_k\right) \tag{1}$$

where $p(t)$ is a narrow input pulse, $\boldsymbol{G} = (G_1, G_2,\ldots,G_N)$ is a vector of positive and negative amplitude-weighted coefficients, $\boldsymbol{\tau} = (\tau_1, \tau_2,\ldots,\tau_N)$ is the vector of interstage delays, and N is the number of filter taps. Those filter parameters depend on the spectral sequence implemented. For instance, Fig. 2 represents the DTF as a structure that approximates to a spectrally spread response with six channels; *i.e.*, $N = 6$. In general, the pulse response of a front-end DTF approximates to:

$$h(t) = w(t)\sum_{n=1}^{M} c_n \cos(\omega_n t - \beta_n) \tag{2}$$

where M is the number of the available subbands, $\omega_n = 2\pi n/T$, where T is set to be equal to the time span of the filter delay line, $w(t)$ is a time window with $w(t) = 1$; for $0 < t \leq T$, and $w(t) = 0$; elsewhere. The symbols c_n and β_n are referred to as the amplitude and phase of the nth-subband, respectively. For implementation in a FIR filter, the temporal phase-coded signal of Eq. 2 is sampled at time instants t_k, $k = 1,\ldots,N$, where the maxima and minima of the function are found. Then tap gain coefficients are computed by:

$$G_k = \sum_{n=1}^{M} c_n \cos\left(\omega_n t_k - \beta_n\right). \quad (3)$$

A FIR filter featuring unequal interstage delays will produce different levels of interpulse interference (IPI) in the pattern. Tap gain adjustment, which is provided by an external voltage bias capability, permits setting the amplitudes at the sampling points thereby compensating for IPI and attenuation introduced by microwave devices. At all other times, the responses of DTF channels as shaped by the rise-time constants of the distributed structures result in a proper match with the pulse of Eq. 2 [5].

FIGURE 2.

As a proof-of-concept, we fabricated a prototype using commercial off-the-shelf discrete heterojunction field-effect transistors ce3512k2. The photograph of the DTF to be used at the decoder upper branch is shown in Fig. 3. The design distributes active cells along high-impedance transmission lines to create an input gate artificial transmission line (ATL) and two drain-ATLs [9]. External bias voltages applied to both rows of active devices provide the capacity to adjust tap gain weights and switch the sign of taps. The distributed circuit design implemented with surface mount devices permits an integration of delay lines to achieve a filter time span of 8.0 ns, which results in a separation of 125.0 MHz from neighborhood subbands, then T = 8.0 ns. The sampling points of the function are separated each other by a variable time leading to a DTF design with unequal interstage delays, $\boldsymbol{\tau}^{\mathrm{upp}}$ = (1.16, 1.33, 1.0, 1.33, 1.0, 1.0) ns, which are around the nominal analog interstage delays to meet impedance matching conditions in ATLs. Tap gain coefficients are adjusted to the vector $\boldsymbol{G}^{\mathrm{upp}}$ = (1.04, 0.61, –0.96, –1.75, 3.90, –3.0), as obtained from Eq. 3 using the phase vector $\boldsymbol{\beta} = (0, -\pi/2, \pi, \pi/2, 0)$ and amplitude vector $\boldsymbol{c}$ = (0, 1, 1, 1, 1).

FIGURE 3.

The frequency response of the FIR filter as a discrete-time system, $H_{DTF}(\omega)$, is given by:

$$H_{DTF}(\omega) = \sum_{k=1}^{N} G_k \exp\left(-j\omega t_k\right). \quad (4)$$

It follows that the frequency response to a narrow pulse is equal to $P(\omega)H_{DTF}(\omega)$, where $P(\omega)$ is the spectrum of the input raw pulse. Then the product should approximate (via the inverse Fourier transform) to the amplitudes and phases at subband frequencies in Eq. 2. The upper DTF correlates the input subbands with the address function, whereas, at the bottom branch, the filter obtains a reference signal from the Input to decode CDMA channels. The response of this filter shows practically the same delay time to that of the upper DTF. Front-end filter implementations must be designed with the same time span to achieve frequency alignment between subbands. The reference function uses the phase vector $\boldsymbol{\beta} = (0, \pi/2, 0, \pi/2, 0)$ and amplitude vector $\boldsymbol{c} = (0, 1, 1, 1,1)$. The DTF is designed with 7 taps, gain coefficients $\boldsymbol{G}^{\text{bott}} = (-2.0, 0.0, -1.0, 2.86, -2.42, -1.27, 3.0)$ and interstage delays $\boldsymbol{\tau}^{\text{bott}} = (0.5, 1.66, 1.0, 1.0, 1.0, 1.33, 1.0)$ ns. The subband amplitudes and phase differences between the voltage transfer functions of both filters are displayed in Fig. 4. A 50-Ohm system is used for testing. Voltage gain parameters are computed from measured scattering parameters of the prototypes and balun components showing good correspondence with the amplitude vectors. The phase shifting of input subbands given by the phase offsets at discrete frequencies makes available bipolar detection [5].

FIGURE 4.

III. UWB Decoder Modeling

A square-law detection model of multiple subbands allows explaining complete cancelation of MAI. The decoder creates the address and reference by correlation (see Fig.1). To capture energy of users, the scheme incorporates synchronization by aligning in time the address with the reference. The response of the correlator to a user, which is defined by Eq. 2 with a FIR function depending on phases $(\varphi_1,\varphi_2,\ldots,\varphi_M)$ and unit vector amplitude, is given by:

$$z_\varphi(t) = \frac{1}{2}\sum_{n=1}^{M} c_n r(t)\cos(\omega_n t + \varphi_n - \beta_n) \tag{5}$$

where $r(t)$ is the autocorrelation function (ACF) of $w(t)$ normalized to T. The lower branch function sets a common reference vector $(\theta_1,\theta_2,..,\theta_M)$ with which all users in the network are decoded. The reference, $z_\theta(t)$, is also created by correlation and has a similar equation just need to substitute the phases φ's by θ's into Eq. 5. The demodulation of a matched user will produce a pulse proportional to the number of successfully decoded subbands. The combination of energy levels of all demodulated subbands can be fully captured by integrating the product $z_\varphi(t)z_\theta(t)$ over the time window (0, 2T). By substituting Eq. 5 into the product and after integration, the decoder output is given by:

$$Z = \sum_{n=1}^{M} c_n^2 \cos(\varphi_n - \theta_n), \tag{6}$$

where the phases of input subbands are cancelled out after mixing both filter responses to produce an output pulse. Eq. 6 denotes that the energy of user subbands are detected by coding. Code mappings allow the construction of mutually orthogonal unipolar-bipolar codes as in optical SAC systems [1]. In order to achieve spectral decoding, the outputs of both front-end correlators are maintained nearly orthogonal in frequency. Code interference is removed provided that the LPF rejects interfering components at the fundamental frequency (ω_1) and beyond. In addition to shift the phases of input subbands by correlation, both front-end DTFs render low sidelobe products as an effect of truncating temporal phase-coded signals. Further demodulation then gives rise to intermodulation noise at the decoder output [5].

The baseband processing characteristics of the two-branch decoder can be framed in the context of noncoherent UWB receivers [6]. Noncoherent decoding should reflect the fact that energy collected of concurrent users must be equal to the addition of energy levels of received subbands therefore the accumulated energy at a given frequency slot equals the sum of squared amplitudes of input subbands overlapped by code. Table I gives an account of differential detection of 3 simultaneous users using a set of Hadamard codes assuming noncoherent detection of energy of subbands. Please note that the square of all user code elements c_n^2 takes on {0,1}. The detected energy levels are combined according with code to produce an output pulse proportional to the number of successfully decoded subbands of the intentional user. Thus unmatched codes will be rejected as zero output.

TABLE I.

Unfortunately, the square-law model breaks down if multiple users arrive during a given time window. Consider two users denoted by x (matched user) and y (interferer) each with random lag τ_x and τ_y and codes $\boldsymbol{c}^{(x)}$ = (0,0,0,1,1) and $\boldsymbol{c}^{(y)}$ = (0,1,0,0,1), respectively. Fig. 5 depicts measurements of DTFs outputs showing time overlapping (shaded portions in the figure) as well as temporal alignment between the address and the reference waveforms for both codes. In turn, the decoder induces correlations between signal parts of different codes giving rise to crosstalk.

FIGURE 5.

The proposed idea is to exploit the noncoherent reception capability of the decoder to reduce crosstalk between concurrent users. A judicious model encompasses Autocorrelation Receiver (AcR) concepts [6]. A similar approach to the scheme with transmitted reference (TR) modulation proposed for delay-hopped (DH) CDMA systems [7] is adopted for the proposal. Unlike with DH-CDMA systems, MAI is fully cancelled for SAC systems, hence the so-called correlation noise emerges as the fundamental impairing effect. The AcR model is extended to analyze the noncoherent decoder whose performance has to include induced correlations from asynchronous users. The reception of subbands overlapped by code results in unwanted interactions when both filter responses are mixed to detect user energy. The model assumes that the delay between correlated signals (D) produces time overlapping in both correlators, then $0 < D < 2T$. The decoder output is calculated by the mean value of the product $z_{\varphi}(t)z_{\theta}(t - \tau)$ over the interval $(0, 2T)$, with τ being the differential delay that separates the responses of front-end DTFs and is thus included as a degree of freedom in the analysis. By substituting Eq. 5 into the product, one arrives at the equation of the mean value of decoder output, $z(\tau, D)$, which at the kth frequency is given by:

$$z(\tau, D)/2T = d_k \, \mathrm{E}[s(t)s(t-\tau)], \tag{7}$$

where $d_k = \cos(\varphi_k - \theta_k)$, which takes on ±1 for all the subbands, $s(t)$ is the input signal and E[▫] denotes expected value. The decoder model can be extended straightforwardly to include other subband frequencies as needed. When $\tau = 0$, the autocorrelation becomes the mean-square value of the input hence Eq. 7 serves to evaluate the decoder as an intensity detector of subbands. In addition, the TR model tuned at a fixed chip interval in [7] is considered here to induce crosstalk when a pulse modulated by code is delayed by a random lag. To evaluate the case of two users given below, the signal model at the kth-frequency at which code overlaps is given by:

$$s(t) = c_k^{(x)} r(t)\cos(\omega_k t) + c_k^{(y)} r(t-D)\cos(\omega_k t - \omega_k D), \tag{8}$$

and its ACF, $r_s(\tau)$, is equal to:

$$r_s(\tau) = \left(c_k^{(x)}\right)^2 \rho_k(\tau) + \left(c_k^{(y)}\right)^2 \rho_k(\tau) + c_k^{(x)} c_k^{(y)} \rho_k(\tau - D) + c_k^{(y)} c_k^{(x)} \rho_k(\tau + D), \tag{9}$$

where $D = \tau_y - \tau_x$, $\rho_k(\tau) = \rho(\tau)\cos(\omega_k \tau)$ with $\rho(\cdot)$ being the ACF of $r(t)$ defined over the interval $(-2T, 2T)$. The first two terms in Eq. 9 do not depend on time overlapping and correspond to energy-detection channels that capture energy of codes. The other terms measure instantaneous phase differences between pulses modulated with two codes, thus characterizing crosstalk. Both contributions depend on the delay D and correspond to the responses of a two-sided AcR model [6]. Fig. 6 depicts the induced correlation using measured waveforms in Fig. 5 contrasting with the model. It shows crosstalk at the output of the intensity detector for the complete differential delay. Only in-phase amplitudes and one-side part of the AcR model response are shown for clarity. Crosstalk created by broadband pulses becomes a rapidly fluctuating signal with τ. The AcR model accounts for code overlapping at the fifth subband and exhibits discrepancies with measurements, which stem chiefly from small subband side-lobe interference. As τ tends to zero in Eq. 9, the two-side crosstalk terms provide an estimate of the interference. Time overlapping could induce crosstalk of mean power as much as twice the power of each user. This is the case when lag D is lower than half of T as $\rho(\tau) \approx 1$ for $|\tau| < T/2$, which thwarts completely the capacity to accumulate energy of concurrent users.

FIGURE 6.

The performance of the decoder can be improved when the decoding relies on the reception of multiple pulses spread over the complete differential delay interval. This means that each user is structured as a sequence of concatenated pulses so that the intensity detector can accumulate energy. According to Eq. 7, user detection also involves the integration of additive interference produced by correlations between pulses showing unequal instantaneous phases. The correlation products have different phases because pulses travel at different delays in both front-end DTFs. Therefore, a model for the interference observed after detection of delayed pulses encompasses multiple AcR channels characterized by various delay lags. Note that induced correlations between concatenated pulses shaped with spectral sequences in Table 1 produce crosstalk with zero-mean. Hence, since AcR model channels measure various instantaneous phases between concatenated pulses instead of energy of codes, a sufficient number of pulses conveying a bit will produce crosstalk with average close to zero.

Let us assume a signaling scheme in which each user comprises a sequence of pulses positioned randomly in time. Users are represented by L pulses positioned over an observation interval $(0, T_I)$ having the nth pulse in the sequence delayed by a random time (δ_n). The random timing of concatenated pulses is hopped in each symbol interval to obtain estimates of the collected energy. The decoder structure incorporates an accurate synchronization to detect a sequence of closely spaced pulses. The collected energy will depend on a number of received pulses as well as on the statistical properties of the variate $\{\delta_n\}$ used for encoding. When the random lags between pulses are Poisson distributed with an average rate (λ), such random pulse-train is treated as shot noise [10]. As shown below, shot noise has the statistical benefit of enhancing the noncoherent detection of concurrent users.

Of primary concern is the amount of energy per user that can be detected using the signaling scheme. The power spectral density of shot noise has been established by Campbell's theorem under the assumption that the random pulse train is observed over a long interval when compared with the duration of the pulse-shape function [10]. Intrinsic circuit characteristics of a given noisy device shape the pulse train generated by several time-dependent fluctuations in correspondence with the discrete nature of the electron charge in such device. For our proposal, the crucial task of the decoder has to be accomplished in a limited time interval by computing the ACF of shot noise ensembles. Consider the model for the input signal at the kth frequency comprising the combination of shaped pulse-trains:

$$s(t) = c_k^{(x)} \sum_{q=0}^{L1-1} r\left(t - \delta_q^{(x)}\right) \cos\left(\omega_k t - \omega_k \delta_q^{(x)}\right) + c_k^{(y)} \sum_{m=0}^{L2-1} r\left(t - D - \delta_m^{(y)}\right) \cos\left(\omega_k t - \omega_k D - \omega_k \delta_m^{(y)}\right) \tag{10}$$

where $\delta_l^{(x)}$ and $\delta_l^{(y)}$ are randomized delays of the lth pulses in the sequences. Both random processes are constructed with the same probability measures defined by an exponential density function with parameter λ. In addition, $L1$ and $L2$ are the number of points distributed over time $T_{\delta 1}$ and $T_{\delta 2}$, respectively; which are close to the decoder's correlation time ($= 2T$). Here, D is set to be equal to the separation between the initial points of the sequences.

For a single user, the accumulation of energy of a burst, which comprises L pulses ($L < L1$) distributed over the observation interval $(0, T_I)$ with delay points: $\delta_0^{(x)}$, $\delta_1^{(x)}\ldots$, $\delta_{L-1}^{(x)}$, is estimated from the autocorrelation of the first summation in Eq. 10. The decoder output averaged over the ensemble and for all the M subbands is written as:

$$z\left(\tau,\delta_0^{(x)}\ldots\delta_{L-1}^{(x)}\right)/T_I=\frac{L}{T_I}\sum_{k=1}^{M}\left(c_k^{(x)}\right)^2 d_k\rho_k(\tau)+\frac{1}{T_I}\sum_{k=1}^{M}\sum_{q=0}^{L-1}\sum_{\substack{m=0,\\ m\neq q}}^{L-1}\left(c_k^{(x)}\right)^2 d_k\rho_k\left(\tau+\delta_q^{(x)}-\delta_m^{(x)}\right). \tag{11}$$

The first term in Eq. 11 provides an estimation of the power of code matched to the decoder and the second term is correlation noise [7] as represents interactions between pulses encoded by the same transmitter arriving at different lags. For a sufficiently large number of points, the ratio L/T_I tends to be equal to λ. When Eq. 11 is evaluated at τ = 0, the decoder output results in an increasing average power with T_I, as shown in Fig. 7 for a matched code $c2$. We notice that for a sufficient number of input pulses, newly arriving pulses tend to add noncoherently. The decoder's ACF provides, upon a complete summation of sampled correlations, power estimations of the received subbands whereas the correlation noise is greatly reduced. Fig. 7 also makes apparent the effect of increasing the observation interval in the noncoherent reception characteristic for Interferer $c1$ and Interferer $c3$. Those results are normalized to the maximal decoder output for matched code $c2$.

FIGURE 7.

The analysis of time overlapping between code-modulated subbands is carried out by computing crosstalk assuming that the coded chip sequences overlap in the frequency domain, as in Table I. The cross-correlation between the first and second term of the signal model of Eq. 10, as produced by an active interferer and an intended user which fully overlap in time, are the crosstalk terms:

$$2c_k^{(x)}c_k^{(y)}\sum_{q=0}^{L1-1}\sum_{m=0}^{L2-1}\rho_k\left(\tau+\delta_q^{(x)}-\delta_m^{(y)}-D\right),$$

where D is an arbitrary short time interval by which the two-sided AcR model channels provide estimations of the interference including at the initial points $\delta_0^{(x)}$ and $\delta_0^{(y)}$. To evaluate the induced interference at the output, τ is made equal to 0, hence the cross-correlation terms become samples of the subband function $\rho_k(\tau)$ taken at two sets of uncorrelated random points. Naturally, a large number of samples allows obtaining satisfactory estimates of the crosstalk. An advantage derived from the use of shot noise is that the so-called Poisson point process [11] can detect line spectra of any frequency when sampling over a sufficient observation interval. Similarly, for the decoder with a limited time window, the selection of a convenient average rate can provide estimations of the interfering subbands. When using random sampling [11], the approximation:

$$\mathrm{E}\left[\rho_k\left(\delta_q^{(x)} - \delta_m^{(y)} - D\right)\right] \cong 0\,, \tag{12}$$

represents a satisfactory estimate of the subband amplitude at the zeroth-frequency; *i.e.* an estimated average of $\rho_k(\tau)$. Hence, the induced crosstalk can be effectively mitigated at the detection subband (around 0 Hz) when the decoder output is averaged over a time window. Fig. 8 depicts the normalized average power of the decoder output when user and interferers fully overlap in time. Notice that the decoder estimates induced crosstalk including side-lobe intermodulation noise. As the observation time increases, code interference is reduced given that the output has approximately the same power as that from the matched code alone. In addition, the case of two concurrent interferers results in low crosstalk for the complete set of points. In those experiments, noise sequences were generated using 100 points distributed approximately over a 17-ns interval ($T_{\delta 1}$ and $T_{\delta 2} \approx 2T$). The sets of points refer to successive inter-arrival times of pulses approximating to Poisson distributions generated with an average rate of 10.0 ns^{-1}.

FIGURE 8.

IV. DECODER FREQUENCY RESULTS

The previous Section shows that the two-branch decoder has a fine time resolution to detect phase differences between pulses spaced by small lags. In order to test the multiple access capability, input pulse bursts were synthesized as the multiple responses of DTF-based encoders. The random templates were uniformly sampled for generation with the AWG, which was configured to store 52 samples of the multitone signals and, in turn, passed through a 9^{th}-order lowpass Butterworth filter to retrieve the pulse burst. The Nyquist rate of the multitone signals is equal to the ratio $2(M + 1)/T$ (= 1.5 GSa/s). All waveforms were oversampled at 2.0 GSa/s to avoid aliasing in the time domain. For instance, Fig. 9 shows a template of User 1 as generated by a random variate of 100 points. The distribution of successive inter-arrival times approximates with good accuracy to an exponential density function.

FIGURE 9.

Fig. 10 displays the Fast Fourier transform (FFT) of the mixer outputs for single users. Measurements were carried out with a 5-GHz oscilloscope. When the address is mixed with the reference, the output consists of several frequency components including naturally the detection subband around 0 Hz. A large amplitude of the detected subband relative to interfering subbands is apparent, which is in agreement with the power estimations in the previous Section. Fig. 11 portrays measurements for two concurrent interferers as well as for the case of user/interference overlapping. Marker readings near 0 Hz show suitable rejection of interference induced from two simultaneous users, about 10-dB below the detected subband of the matched code alone. The detection of the user plus two interferers present low deviations from the case of the matched code confirming that unmatched codes will be rejected as small interference. Again, those results are in agreement with previous analysis.

FIGURE 10.

FIGURE 11.

For all the cases shown above, a lowpass filter with cut-off frequency of about 50 MHz and a succeeding threshold scheme can easily remove code interference at the detection subband. This makes available the implementation of subsequent data recovering circuits working at bit-rate speeds while simplifying the receiver design for high speed systems.

CONCLUSION

This paper introduces a new approach for electronic-CDMA based on noncoherent UWB concepts. Experimental and analytical work investigate the two-arm structure as a spectral decoder for SAC systems. In order to decode concurrent users, the scheme incorporates an accurate synchronization by which a large number of AcR model channels can be differentiated from energy-detection channels. When input pulse-trains behave as noise, the detection of multiple pulses improves cumulative processes of energy, whereas interference characterized by correlations between temporally randomized pulses is modeled using AcR concepts. In order to gain the benefits of the spectral approach, the intensity detector achieves satisfactory estimations of the interference so that the average of the accumulated interference tends to zero. Hence, the decoder output is computed by bipolar detection of noncoherent energy in the spectral domain. The conclusion drawn from this work is that the electronic-SAC makes an efficient use of the available bandwidth while maintaining full orthogonality among concurrent users. Given its advantages concerning implementation complexity, the proposed noncoherent decoding becomes a prospective method for investigation of asynchronous electronic-CDMA. Nevertheless, it would be of the upmost importance to examine impairments of the optical channel to give full assessment of the multiple access method for multi-Gb/s systems.

References

[1] H. Ghafouri-Shiraz and M. M. Karbassian, *Optical CDMA Networks: Principles, Analysis and Applications, West Sussex, U.K.: John Wiley & Sons, 2012, 432p.*

[2] J. B. Rosas-Fernandez, J. D. Ingham, and R.V. Penty, and I. H. White, 18 Gchip/s Electronic CDMA for Low-Cost Optical Access Networks, *J Lightwave Technol*, 2009:27:306-313.

[3] J.B. Rosas-Fernandez, J.D. Ingham, R.V. Penty, and I.H. White, Scalability Techniques in Electronically Processed CDMA for Low Cost and Flexible Optical Access Networks, *11th International Conference on Transparent Optical Networks*, Islas Azores, Portugal, 2009:1-5.

[4] X. Guo, Q. Wang, X. Li, L. Fang, A. Wonfor, J.L. Lindeiner, R.V. Penty, and I.H. White, First demonstration of OFDM ECDMA for low cost optical access networks, *Opt Lett*, 2015:40:2353-2356.

[5] J. Aguilar-Torrentera, G. García-Sánchez, R. Rodríguez-Cruz, and I.Z. Darwazeh, Extending Distributed-Based Transversal Filter Method to Spectral Amplitude Encoded CDMA, *IEICE Trans Electron*, 2018:101:953-962.

[6] K. Witrisal, G. Leus, G. J.M. Janssen, M. Pausini, F. Troesch, T. Zasowski and J. Romme, Noncoherent Ultra-wideband Systems, *IEEE Signal Process Mag*, 2009:26:48-66.

[7] R. Hoctor and H. Tomlinson, Delay-Hopped Transmitted-Reference RF Communications, *IEEE Conf. Ultra Wideband Systems and Technologies*, Baltimore, MD, USA, 2002:265-270.

[8] A. Borjak, P.P. Monteiro, J. O'Reilly, and I. Darwazeh, High-Speed Generalized Distributed-Amplifier-Based Transversal-Filter Topology for Optical Communication Systems, *IEEE Trans Microw Theory Techn*, 1997:45:1453-1457.

[9] J. Aguilar-Torrentera and I. Darwazeh, Performance Analysis of a Monolithic Integrated Transversal Filter Using Mixed-Mode Scattering Parameters, *IET Microw Antenna P*, 2007:1:925-931.

[10] M. J. Buckingham, *Noise in Electronic Devices and Systems, West Sussex, U.K.: John Wiley & Sons, 1983, 498p.*

[11] A. Papoulis, *Probability, Random Variables and Stochastic Processes, New York, USA, McGraw Hill, Third Edition, 1991,* 667p.

Acknowledgments: The authors acknowledge the Avionics Laboratory of the CIIIA-FIME-UANL for allowing the use of its measurement facilities. The authors also are grateful with M. Sc. Gerardo Garcia Sánchez for its support to test the decoder with the AWG.