

Hybrid F-PolSK Transceiver Architecture for CDMA-PON

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Abstract—A novel incoherent transceiver design for optical code-division multiple-access (OCDMA) system based on the (hybrid) frequency-polarization shift keying, so-called F-PolSK is proposed for the application in the passive optical networks (PON). In the system analysis we have considered the presence of (i) optical amplified spontaneous emission noise, (ii) electronic receiver noise, (iii) photo-diode shot-noise and (iv) multiple-access interference. The results revealed that the proposed architecture can easily accommodate a greater number of users with efficient power consumption. Furthermore, the system security is much enhanced due to two-dimensional advanced modulation in optical domain.

Index Terms—Frequency shift keying, optical CDMA, polarization shift keying, prime code families

I. INTRODUCTION

RECENTLY, the multimedia-on-demand communication has remarkably increased requiring reliable and high data-rate services over a network. Passive optical network (PON) can be one of the best options to deliver the high bit-rate and high capacity services in very near future. Optical code-division multiple-access (OCDMA) provides multichannel communications over a single frequency band. Accordingly, signals must be designed to reduce mutual interferences. A power efficient high-order modulation technique based on polarization shift keying (PolSK) [1] can be considered as a solution to comply with the demands mentioned herein. Unlike traditional multilevel modulation techniques where the transmitted signals are represented as a two-dimensional (2D) constellation points, the PolSK exploits the degrees of freedom provided by the use of the three-dimensional (3D) states of polarization (SOP) of a fully polarized lightwave carrier as the information-bearing parameters. The received SOP fluctuations, caused by fiber birefringence, can be compensated at the decision processor as demonstrated experimentally in [2]. The receivers are implemented in a way to extract the Stokes parameters which completely characterize the received lightwave SOP. Ref. [3] and [4] has reported the advantages of the PolSK over OCDMA. On the other hand, frequency shift keying (FSK) modulation is attributed with high phase noise insensitivity and high power efficiency. The

advantages of FSK over OCDMA system has also been analyzed in [5]. The merits of hybrid optical frequency-polarization shift keying (F-PolSK) modulation solely in optical communication have been introduced in [6].

Here, we propose a novel design structure for the OCDMA transport network based on the two-dimensional array (*2D-ary*) F-PolSK modulation. This technique results in increasing distances between the modulated signal points on the Poincaré sphere [1], and hence reduced transmission power requirements. We also analyze the application of optical tapped-delay line (OTDL) as the incoherent OCDMA decoder.

In our analysis we have assumed that the signal is degraded by (i) fiber amplified spontaneous emission (ASE) noise, (ii) electronic receiver noise (iii) photo-detectors (PD) shot-noise and (iv) multiple-access interference (MAI) caused by CDMA concept. However, an optical filter can be placed at the input of the receiver (or after the optical pre-amplifier, if needed) to reduce the ASE. Intensity-modulated incoherent systems are vulnerable in terms of interception, which can easily be broken by simple power detection, even without any knowledge of the code. Whereas by using 2D advanced modulation (i.e. F-PolSK) with flexible spreading code-length, the security can also be well improved.

For optical signature sequences, we have used a coding scheme based on a new spreading code referred to double-padded modified prime code (DPMPC). It has flexible properties and been extensively introduced and employed in pulse-position modulation (PPM) [7], in overlapping PPM [8] and in coherent schemes [9]-[11]. There has been a considerable shift of interests from coherent scheme to constant-power-envelope incoherent scheme on the part of both research and industry, due to its (i) simple architecture, (ii) cost effectiveness, (iii) high immunity to laser phase-noise, and (iv) insensitivity to self- and cross-phase modulations caused by the fiber nonlinearity.

The proposed transmitter configuration is introduced in the following section. The receiver structure and its analysis are studied in section III. The overall architecture performance is presented in section IV. Finally the study is concluded.

II. TRANSMITTER ARCHITECTURE

The *2D-ary* F-PolSK transmitter is shown in Fig. 1a. It consists of FSK modulator with a tunable laser diode (TLD)

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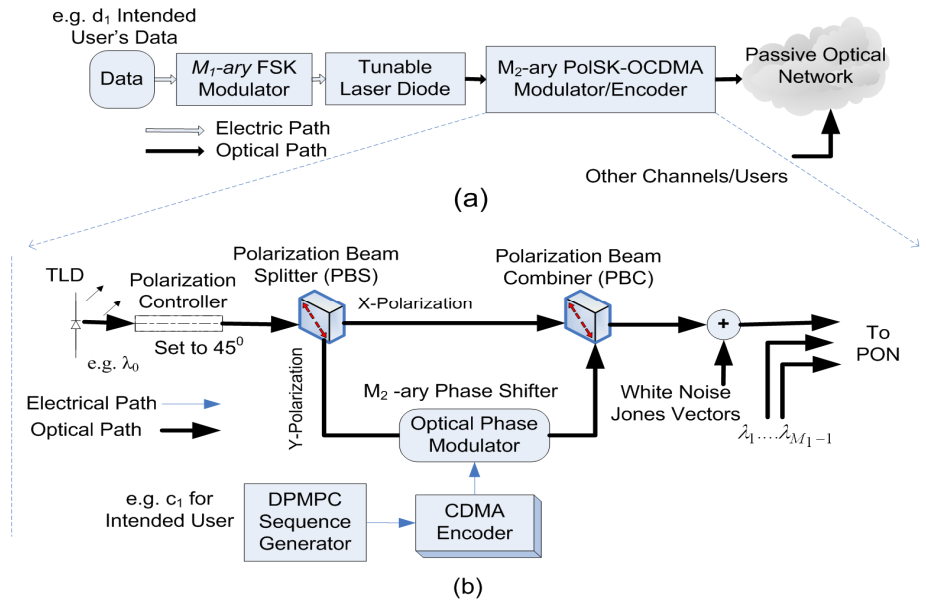


Fig. 1. (a) Incoherent 2D-ary F-PolSK-OCDMA transmitter (b) M_2 -ary PolSK modulator and OCDMA encoder

which allocates M_1 symbols ($k_1 = \log_2^{M_1}$ bits per symbol) to the corresponding M_1 wavelengths.

The PolSK-OCDMA modulator is depicted separately in Fig. 1b. Each wavelength from the TLD, denoting the FSK modulated signal, is initially (linearly) polarized and set to 45° for simplicity through polarization controller at the beginning of the PolSK modulator. The wavelengths then enter the PolSK-OCDMA modulator. The signal is M_2 -ary PolSK modulated where M_2 denotes number of the lightwave SOPs ($k_2 = \log_2^{M_2}$ bits per SOP), and CDMA encoded by means of the DPMP spreading sequences.

At the transmitter output and a given reference plane (\hat{x}, \hat{y}) normal to the direction of propagation \hat{z} , the transmitted lightwave over PON during M symbol intervals can be written, in the complex form as:

$$\begin{aligned} \bar{v}(t) &= \sum_{m=1}^M \bar{v}_{m_1 m_2}^{(m)}(t) \quad 0 \leq t \leq MT_s \\ &= \sum_{m=1}^M \bar{p}_{m_2}^{(m)} \exp[j2\pi(f_c + f_{m_1}^{(m)})(t - mT_s)] u_T(t - mT_s) \end{aligned} \quad (1)$$

where T_s is the symbol interval, $m_1 = 1, 2, \dots, M_1, m_2 = 1, 2, \dots, M_2$. $\bar{v}_{m_1 m_2}^{(m)}(t)$ is the complex transmitted signal during the m^{th} signaling interval, $\bar{p}_{m_2}^{(m)}$ is the 2D vector that gives the signal amplitudes and phases over \hat{x} and \hat{y} directions. f_c is the carrier frequency with offsets of $f_{m_1}^{(m)}$ (increment or decrement) representing the assigned frequency from the TLD. $u_T(t)$ is a rectangular pulse of unit amplitude. It is shown in [6] that the frequency tones orthogonality can be satisfied when the minimum frequency separation between any adjacent

frequency tones is $\Delta f = 1/T_s$. Thus, $f_{m_1}^{(m)}$ takes values from the set $\{r_{m_1}^{(m)} \Delta f / 2\}_{m_1=1}^{M_1}$, where $r_{m_1}^{(m)} = 2m_1 - 1 - M_1$. This makes the complex correlation coefficient value between two different frequency-polarization modulated signals 'zero' (i.e. orthogonal).

In (1), $\bar{p}_{m_2}^{(m)}$ denotes the transmitted signal amplitudes and phases over the orthogonal \hat{x} and \hat{y} channels during the m^{th} signaling interval, given by:

$$\bar{p}_{m_2}^{(m)} = \begin{pmatrix} \bar{p}_{x_{m_2}}^{(m)} & \bar{p}_{y_{m_2}}^{(m)} \end{pmatrix}^T = \begin{pmatrix} a_{x_{m_2}}^{(m)} e^{j\theta_{x_{m_2}}^{(m)}} & a_{y_{m_2}}^{(m)} e^{j\theta_{y_{m_2}}^{(m)}} \end{pmatrix}^T \quad (2)$$

where $a_{x_{m_2}}^{(m)}$, $a_{y_{m_2}}^{(m)}$ and $\theta_{x_{m_2}}^{(m)}$, $\theta_{y_{m_2}}^{(m)}$ are the amplitudes and phases of the (\hat{x}, \hat{y}) -components of the lightwave. The discrete random sequences $\{\bar{p}_{m_2}^{(m)}\}_{m_2=1}^{M_2}$ and $\{r_{m_1}^{(m)}\}_{m_1=1}^{M_1}$ are stationary, independent and represent the source symbol sequences. Moreover, $\bar{p}_{m_2}^{(m)}$ determines the SOP of the fully polarized lightwave given in (1) during the m^{th} symbol interval that is corresponding to symbol number m_2 .

The electromagnetic wave $\bar{v}_{m_1 m_2}^{(m)}(t)$ consists of two sets of 2D-ary signals. The signals in each set are uncorrelated and orthogonal (i.e. orthogonal frequencies and SOP), and all the $M = M_1 \times M_2 \rightarrow k = k_1 + k_2$ signals have the same time duration. Now, the data source emits a data symbol from a set of M symbols every $T_s = k.T_b$ seconds, where T_b is the bit duration.

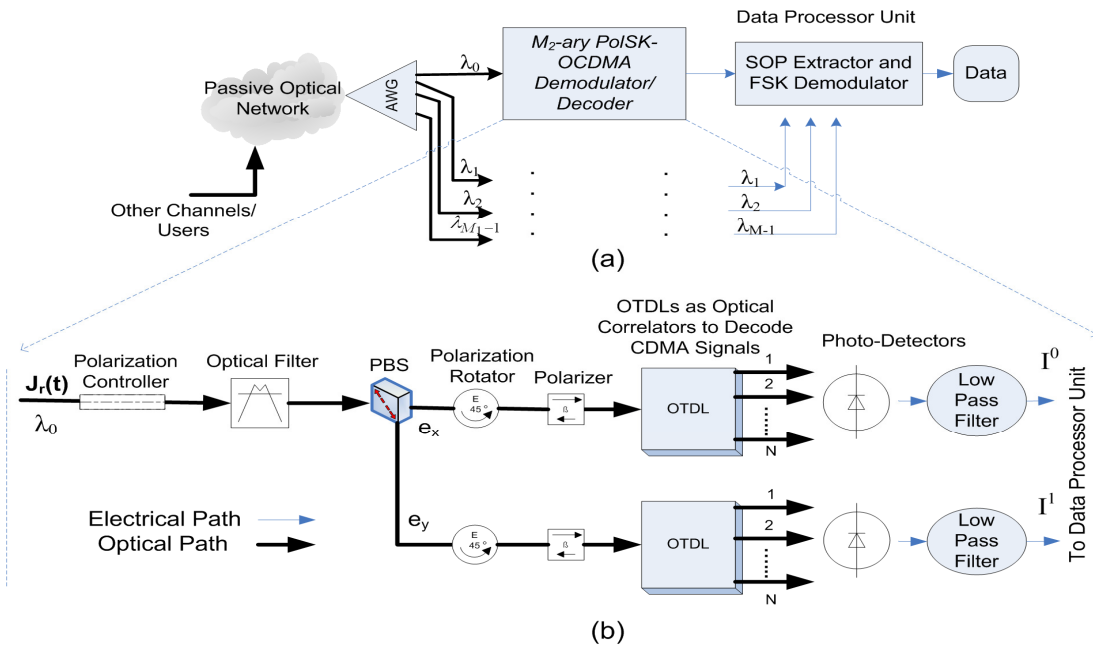


Fig. 2. (a) Incoherent 2D-ary F-PolSK-OCDMA receiver (b) M_2 -ary PolSK demodulator and OCDMA decoder with OTDL

The SOP is described using $(S_{1m_2}^{(m)}, S_{2m_2}^{(m)}, S_{3m_2}^{(m)})$ as Stokes parameters. Each of which is determined by the amplitudes $a_{x_{m_2}}^{(m)}$ and $a_{y_{m_2}}^{(m)}$ plus phase differences $\Delta\theta^{(m)} = \theta_{y_{m_2}}^{(m)} - \theta_{x_{m_2}}^{(m)}$ of the m^{th} symbol. These Stoke parameters and their relations have been discussed in [1].

The average number of photons representing the energy of the transmitted multistate polarized lightwave over PON is directly proportional to other parameters, as:

$$S_{0m_2}^{(m)} = \sqrt{\sum_{i=1}^3 (S_{im_2}^{(m)})^2} \quad (3)$$

where $S_{0m_2}^{(m)} = E_{sm_2} / T_s$ is the fourth Stokes parameter representing the transmitted optical waveform power with energy of E_{sm_2} . Therefore, only three out of four Stokes parameters are mutually independent.

Due to the fiber-optic nonlinearity, it is very convenient to generate lightwaves with the constant power envelop. If all electromagnetic signals are transmitted on the same power level (i.e. $S_{0m_2}^{(m)} = S_0$ or equivalently, $E_{sm_2} = E_s$) and have the same carrier frequency, then they can be assumed as the vectors with the form of $\vec{S}_{m_2}^{(m)} = S_{1m_2}^{(m)} \hat{s}_1 + S_{2m_2}^{(m)} \hat{s}_2 + S_{3m_2}^{(m)} \hat{s}_3$. These vectors are allocated on the surface of the Poincaré sphere with a constant radius of S_0 [1]. As a final point, the F-PolSK modulated and CDMA encoded lightwave is transmitted over PON as discussed and shown in Fig. 1.

III. ANALYSIS OF RECEIVER ARCHITECTURE

A. Receiver Configuration

The front-end of the 2D-ary F-PolSK-OCDMA receiver for application in PON is initially with frequency selective arrayed waveguide grating (AWG), as illustrated in Fig. 2a. Here, the AWG is assumed ideal so that no interference between adjacent wavelengths occurs. The wavelengths assigned to the symbols at the transmitter are denoted as $\lambda_0, \lambda_1, \dots, \lambda_{M-1}$. Each wavelength (after AWG) enters the PolSK demodulator and OCDMA decoder, displayed separately in Fig. 2b, to indicate the Stoke parameters from each wavelength. Thereafter, the symbol SOPs are extracted based on Stokes parameters in the data processor unit, where the FSK part of the signal is also demodulated. The FSK detector determines the part of symbols which were used to generate the frequency tones $\{f_{m_i}^{(m)}\}_{m_i=1}^{M_i}$ in every symbol interval m .

In Fig. 2b, the application of optical tapped-delay lines (OTDL) which are used as the optical correlator in incoherent OCDMA decoder is illustrated. The delay coefficients in OTDLs are designed in such a way to make them perform as CDMA chip-decoders at each wavelength filtered through AWG. It is observed from Fig. 2b that OTDLs outputs contain N -chip pulses (where N equals code-length, $P^2 + 2P$) that can be assumed as a parallel circuit of many single photo-detectors (PD). The currents are added up and no interference between the OTDLs pulses is possible. The signals are photo-detected to generate the electrical currents shown as I^0, I^1 in Fig. 2b denoting the upper and lower branches respectively. They are

then ready for SOP and finally data extraction in the data process unit.

Since the balanced-detection is used for binary systems and we consider multiple levels ($M_2 - ary$) polarization modulation, the receiver structure in Fig. 2b is proposed for higher-order modulations. Accordingly, the total current of the upper branch (i.e. including all chip currents) can be obtained as:

$$I^0 = \frac{\Re}{4} \sum_{n=1}^N \left\{ c(nT_c) \left(\sum_{i=1}^K (E_{xi}^2(t) + E_{yi}^2(t) + d_i(t)c_i(t-nT_c)(E_{xi}^2(t) - E_{yi}^2(t))) \right) \right\} \quad (4)$$

where \Re denotes responsivity of PD and $E_{xi}(t)$, $E_{yi}(t)$ are the orthogonal (\hat{x} , \hat{y})-components of received linearly polarized electric field of i^{th} user, $d_i(t)$ is the data signal with symbol duration of T_s , $c_i(t)$ is the N -chip code sequences signal with chip duration of T_c and $d_i(t), c_i(t) \in \{0,1\}$. The electric current,

I^0 contains the signal intensity generated at the upper branch of the modulator at the transmitter, defined by $P_i^0 = E_{xi}^2(t) + E_{yi}^2(t)$ and the linear polarized information part generated in lower branch of the modulator at the transmitter (see Fig. 1b), defined by $P_i^1 = E_{xi}^2(t) - E_{yi}^2(t)$. Hence, the total current, I^0 (the same as lower current I^1) can be rewritten as:

$$I^0 = \frac{\Re}{4} \sum_{n=1}^N c(nT_c) \sum_{i=1}^K (P_i^0 + d_i(t)c_i(t-nT_c)P_i^1) + n(t) \quad (5)$$

where $n(t)$ represents the filtered Gaussian noise with equal variance of $\sigma_{n(t)}^2$ that includes: (i) the PD shot-noise with electric current variance of $\langle i_{av}^2 \rangle = 2eiB_o$ where i_{av} is the average photo-current; (ii) optically filtered ASE noise with variance of $\sigma_{ASE}^2 = 2N_0B_o$ where N_0 is the (unilateral) power spectral density (PSD) of the white ASE noise arriving at each polarization and also B_o is the optical filter bandwidth; (iii) electronic receiver noise current (i.e. thermal noise) at the low-pass filter with variance of $\sigma_{LP}^2 = \frac{2k_bT}{R} \cdot B_{el}$ where R is the filter direct-current (dc) equivalent impedance, T the absolute temperature and k_b the Boltzmann constant, B_{el} the electronic filter bandwidth. Thus variance of additive noise of $n(t)$ can be represented in total as:

$$\sigma_{n(t)}^2 = \langle i^2 \rangle + \sigma_{ASE}^2 + \sigma_{LP}^2 \quad (6)$$

The current I^0 in (5) can be modified regarding first user (#1) as the intended user by further calculations as:

$$I^0 = \frac{\Re}{4} \sum_{n=1}^N P_1^0 c(nT_c) + \frac{\Re}{4} \sum_{n=1}^N c(nT_c) c_1(t-nT_c) d_1(t) \cdot P_1^1 + \frac{\Re}{4} \sum_{i=2}^K \sum_{n=1}^N c(nT_c) c_i(t-nT_c) d_i(t) P_i^1 + n(t) \quad (7)$$

The first term in (7) is the dc current that needs estimation and removal. The second term represents the intended data

mixed with its polarization and assigned spreading code auto-correlation (DPMPC [7], [8]). While the third element assumes the co-channel interference (i.e. MAI) caused by other transmitters and the last term exhibits the noise.

Hence, the system signal-to-noise ratio (SNR) with respect to the interference and the noise can be obtained as:

$$SNR = \gamma = \frac{\left(\frac{\Re}{4} \sum_{n=1}^N c(nT_c) c_1(t-nT_c) d_1(t) P_1^1 \right)^2}{\left(\frac{\Re}{4} \sum_{i=2}^K \sum_{n=1}^N c(nT_c) c_i(t-nT_c) d_i(t) P_i^1 \right)^2 + \sigma_{n(t)}^2} \quad (8)$$

Now, according to the DPMPC properties [8], we have:

$$\sum_{n=1}^N c(nT_c) c_1(t-nT_c) = P + 2 \quad (9)$$

which is DPMPC auto-correlation and P is a prime number. By defining the variable X_{li} as the DPMPC cross-correlation value:

$$X_{li} = \sum_{n=1}^N c_l(nT_c) c_i(t-nT_c) \quad (10)$$

Its probability density function (PDF) can be derived from the independent values of random variable X_{li} . The in-phase cross-correlation value of DPMPC [8] is either 'zero' or 'one' depending on whether the codes are in the same group or from the different groups. Obviously, the 'zero' value does not cause the interference due to perfectly orthogonal sequences, while the 'one' value causes the interference which is only among intended user and ($P^2 - P$) users from the different groups (i.e. P^2 whole sequences (users) and P sequences from the same group of intended user which are orthogonal [10]). As, the cross-correlation values are uniformly distributed among interfering users, thus the PDF of w , realization of X_{li} , is:

$$P(w=i) = \frac{i}{P^2 - P} \quad (11)$$

where $P(w=i)$ is the probability that w assumes the value i (the number of actively involved users in the transmission). Therefore, by substituting (9) and (11) into (8) and further simplifications, the system SNR can be obtained as:

$$SNR(K) = \gamma(K) = \frac{1}{\left(\frac{(K+2)(K-1)}{2(P^2-P)(P+2)} \right)^2 + \frac{16\sigma_{n(t)}^2}{\Re^2 d_1^2(t) \cdot S_1^2 (P+2)^2}} \quad (12)$$

Note that:

$$SNR(1) = \gamma = \Re^2 d_1^2(t) \cdot S_1^2 (P+2)^2 / 16\sigma_{n(t)}^2 = E_b / N_0 \quad (13)$$

is the single-user SNR where E_b is the energy of one bit and N_0 is the noise PSD. Expression (12) is one of the main results of this paper as it represents the system SNR of polarization-modulated OCDMA signaling.

B. Analysis of the Error Probability

By assuming that $\bar{v}_{m_1 m_2}^{(m)}(t)$ is transmitted over PON, the

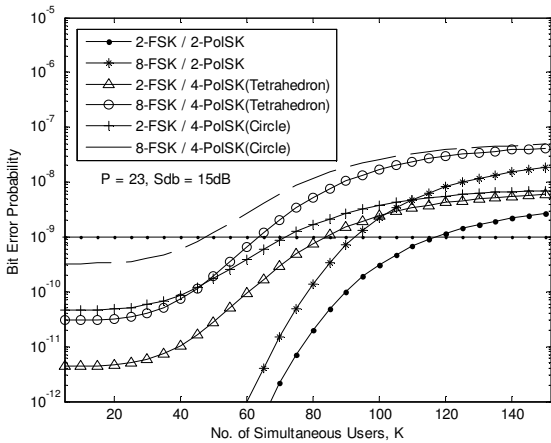


Fig. 3. BER of various M_1 -FSK / M_2 -PolSK optical CDMA against K

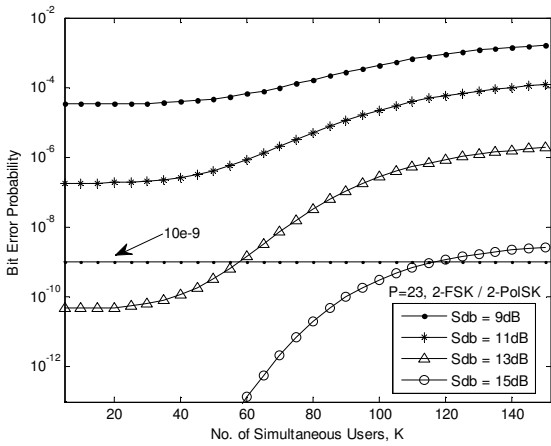


Fig. 4. BER comparison of 2-FSK / 2-PolSK optical CDMA against K

correct decision at the FSK demodulator (correct symbols carried by the correct wavelength) can be expressed as:

$$|\lambda_m|^2 = \begin{cases} |E_s + N_{m1}|^2 & m = m_1 \\ |N_m|^2 & m \neq m_1 \end{cases} \quad (14)$$

where E_s is the symbol energy and $\{N_m\}_{m=1}^{M_1}$ is independent, zero mean, Gaussian noise random variable. The correct decision is made if and only if $|\lambda_{m_1}|$ in the decision rule of M_1 -ary FSK satisfies:

$$|\lambda_{m_1}|^2 = \max \{ |\lambda_m| \}_{m=1}^{M_1} \quad (15)$$

Also it is assumed that the received estimated noisy parameters $\bar{R}_{m1} = (R_{1m1}, R_{2m1}, R_{3m1})$ are in the decision region of the un-noisy transmitted parameters $\bar{S}_{m2} = (S_{1m2}, S_{2m2}, S_{3m2})$. Based on the maximum likelihood (ML) decision rule introduced in [2] and assuming that all possible transmitted vectors $\{\bar{S}_l\}_{l=1}^{M_2}$ are equipower and equiprobable, then the decision metric is implemented as:

$$f_{\bar{R}_{m1}|\bar{S}_{m2}}(\bar{R}|\bar{S}_{m2}) = \max \left\{ f_{\bar{R}_{m1}|\bar{S}_l}(\bar{R}|\bar{S}_l) \right\}_{l=1}^{M_2} \quad (16)$$

where $f_{\bar{R}_{m1}|\bar{S}_l}(\bar{R}|\bar{S}_l)$, for $l=1,2,\dots,M_2$ is the conditional PDF of the estimated noisy Stokes vector \bar{R}_{m1} given that \bar{S}_l is transmitted. This PDF was already given in spherical coordinates $(\rho_m, \theta_m, \phi_m)$ in [1]. Referring to the ML rule particularly \bar{S}_{m2} was chosen as the transmitted vector when it satisfies:

$$\bar{R}_{m1} \cdot \bar{S}_{m2} = \max \{ \bar{R}_{m1} \cdot \bar{S}_l \}_{l=1}^{M_2} \quad (17)$$

In the absence of noise, the index m in (15) and l in (16) should be equal to m_1 and m_2 of $\bar{v}_{m1m2}^{(m)}(t)$ respectively.

The probability of the correct decision for the system is then equal to the probability that satisfies (15) times the probability that satisfies (17) conditional to (15), given by:

$$P_c = \Pr \left(|\lambda_{m1}|^2 = \max \{ |\lambda_m| \}_{m=1}^{M_1} \right) \times \Pr \left(\bar{R}_{m1} \cdot \bar{S}_{m2} = \max \{ \bar{R}_{m1} \cdot \bar{S}_l \}_{l=1}^{M_2} \mid |\lambda_{m1}|^2 = \max \{ |\lambda_m| \}_{m=1}^{M_1} \right) \quad (18)$$

The first probability in (18), is the probability that a correct decision is made on the transmitted frequency at the FSK demodulator shown in Fig. 2b. Its PDF has been discussed in [6].

The second probability in (18) is the probability that the transmitted SOP is correctly chosen at the data processor unit shown in Fig. 2a, as well as the correct decision is made on the transmitted frequency at the frequency demodulator. Since these two processes are independent, the probability of correct detection is equal to the PDF of M_2 -PolSK modulation. This PDF was also evaluated in [1] for some regular equipower polarization constellations.

IV. NUMERICAL DISCUSSIONS

In this section, the bit-error rate (BER) probability as a function of number of simultaneous users (K) with different single-user SNR (as shown by Sdb on graphs) is presented. The numerical results of the proposed architecture are introduced in Figs. 3 and 4.

Advantages of polarization and hybrid modulations over other modulations e.g. FSK or DPSK have already been discussed in [1], [2] and [6] for comparison purposes. Therefore, only the proposed architecture evaluation is demonstrated as it comparably outperforms.

Figure 3 explains the performance of the transceiver under conditions of $P=23$, $Sdb=15dB$ and different M_1 -ary FSK with different M_2 -ary polarization constellations. It can be observed from Fig. 3 that binary FSK with binary PolSK enhances the system BER to compare with other multiple array combinations. $BER=10^{-9}$ is exposed as a reference. The 2-FSK / 2-PolSK configuration is able to accommodate 24% (i.e. 120 users, regarding $P=23$) of full-load simultaneous users

with $BER=10^{-9}$. To compare the results with those in [4], the proposed design enhances the system capacity with even smaller code-length (i.e. 575 regarding $P=23$ [8] whereas 1023 in [4]). Accordingly, the proposed coding and architecture can also offer higher throughput as the spreading code is smaller than Gold sequences used in [4].

Furthermore, it is observable that increasing the system multiplicity (i.e. M_x -ary) grows the system complexity and also degrades the overall performance. That also increases the implementation expenses. However, they are still reasonable enough to accommodate certain number of users regarding $BER = 10^{-9}$.

The 2-FSK / 2-PolSK performance evaluation with respect to different single-user SNR (i.e. Sdb) is illustrated in Fig. 4 against K as the most convenient and realistic scheme to consider. It is recognizable that the system is also very power efficient in that the system capacity is doubled (from 60 to 120) at the cost of only 2dB from 13dB to 15dB.

V. CONCLUSION

The $2D$ -ary signaling is presented and spread over higher dimensional constellation which provides distant geometric positions. This F-PolSK-OCDMA signaling for the application in passive optical networks has constant envelope that makes the signals more robust to nonlinear fluctuations imposed by optical fibers. The receiver structure has been designed by the aid of OTDL as the CDMA-decoder. Its performance has been analyzed by considering the optical amplified spontaneous emission noise, electronic receiver noise, photo-detectors shot-noise and mainly the multiuser interferences. Finally, the derived results indicated that the binary combination of two modulations remarkably enhances the system capacity in that it can reliably and power-efficiently accommodate greater number of simultaneous users. The system security has also been improved in terms of interceptions or intrusions due to 2D advanced optical modulation.

REFERENCES

- [1] S. Benedetto and P. T. Poggiolini, "Multilevel polarization shift keying: optimum receiver structure and performance evaluation", IEEE Trans. on Comm., vol. 42, no. 2/3/4, pp. 1174-1186, 1994
- [2] S. Benedetto, R. Paoletti, P. Poggiolini, C. Barry, A. Djupsjoebacka and B. Lagerstroem, "Coherent and direct-detection polarization modulation system experiments," in Proc. ECOC '94, pp. 67-71, 1994
- [3] M. M. Karbassian and H. Ghafouri-Shiraz, "Novel PolSK-OCDMA transceiver architecture", in Proc. WCE 2008 (ICEEE), pp. 415-419
- [4] K. Iversen, J. Mueckenheim and D. Junghanns, "Performance evaluation of optical CDMA using PolSK-DD to improve bipolar capacity", in Proc. SPIE, vol. 2450 (Amsterdam), pp. 319-329, 1995
- [5] M. M. Karbassian and H. Ghafouri-Shiraz, "Novel channel interference reduction in optical synchronous FSK-CDMA networks using a data-free reference", J. of Lightw. Technol., vol. 26, no. 8, pp. 977-985, 2008
- [6] M. M. Matalgah and R. M. Radaydeh, "Hybrid frequency-polarization shift keying modulation for optical transmission", J. of Lightw. Technol., vol. 23, no. 3, pp. 1152-1162, 2005
- [7] M. M. Karbassian and H. Ghafouri-Shiraz, "Fresh prime codes evaluation for synchronous PPM and OPPM signaling for optical

- CDMA networks", J. of Lightw. Tech., vol. 25, no. 6, pp. 1422-1430, 2007
- [8] M. M. Karbassian and H. Ghafouri-Shiraz, "Capacity enhancement in synchronous optical overlapping PPM-CDMA network by a novel spreading code", Proc. of GlobeCom 2007, pp. 2407-2411
- [9] M. M. Karbassian and H. Ghafouri-Shiraz, "Phase-modulations analyses in coherent homodyne optical CDMA network using a novel prime code family", in Proc. WCE'07 (ICEEE), pp. 358-362, 2007
- [10] M. M. Karbassian and H. Ghafouri-Shiraz, "Performance analysis of heterodyne detected coherent optical CDMA using a novel prime code family", J. of Lightw. Technol., vol. 25, no. 10, pp. 3028-3034, 2007
- [11] M. M. Karbassian and H. Ghafouri-Shiraz, "Performance analysis of unipolar code in different phase modulations in coherent homodyne optical CDMA", IAENG Engineering Letters, vol. 16, no. 1, pp. 50-55, 2008