Microwave-Domain Analog Predistortion Based on Chirped Delay Lines for Dispersion Compensation of 10-Gb/s Optical Communication Signals

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***Abstract*—Optical chromatic dispersion compensation is achieved by analog predistortion of the signal in the microwave domain exploiting a chirped delay line at the transmitter. We designed a microwave device that guarantees high compactness of the solution and increased performance in terms of uncompensated propagation reach. Error-free, 10-Gb/s transmission over 225 km of uncompensated fiber is experimentally demonstrated without the use of coherent detection, digital equalizers, or optical dispersion compensators.**

***Index Terms*—Compensation, microwave circuits, modulation, optical fiber communication.**

I. Introduction

CHROMATIC dispersion compensation in optical fiber communication systems is still an open issue. In particular, in the case of dispersion-uncompensated metropolitan and regional networks, transmission over standard single mode fiber (SMF) at 10 Gb/s is strongly limited to about 80–100 km for conventional nonreturn-to-zero (NRZ) intensity-modulated, direct-detected (IM-DD) signals. Dispersion compensation can be generally achieved by either optical or electrical techniques. Optical techniques, such as the use of dispersion-compensating fibers (DCF) or chirped fiber Bragg gratings, are generally expensive and not easily reconfigurable for varying dispersion conditions. Dispersion-tolerant modulation formats can be used, such as the so-called phase-shaped binary transmission (PSBT) duobinary format, but they allow the achievement of up to about 200-km SMF propagation without compensation [1]. As an alternative, electrical dispersion compensation (EDC) techniques have been developed [2]–[5]. EDC makes use of digital equalizers, such as transversal filters [2], digital feedback equalizers [3], or maximum likelihood sequence detectors [4], [5] placed at the optical receiver to compensate for the total accumulated optical dispersion. Using EDC, dispersion compensation for NRZ IM-DD propagation over hundreds of kilometers of SMF has been experimentally demonstrated at 10 Gb/s. Electrical techniques are typically cheaper than optical ones. Therefore, they are promising for low-cost metropolitan area networks applications. However, electrical equalization is usually characterized by less efficient performance, owing to the channel nonlinearity introduced by square-law photodiodes. Because of this detection nonlinearity, optical phase information is lost after detection, making dispersion compensation difficult to achieve.

EDC operating in the microwave regime using microstriplines to achieve analog equalization has also been proven [6]. However, this solution requires the use of coherent detection of the optical signal in order to save amplitude and phase information after square-law detection.

An alternative technique to compensate for optical chromatic dispersion is based on electronic digital predistortion at the transmitter [7]. This technique shows better performance than electronic equalization, but it requires the use of high-speed digital electronics in the transmitter. Moreover, the length of the predistorted bit sequence increases quadratically with the bit rate, thus increasing the memory requirements at high bit rates.

In [8], analog predistortion at the transmitter has been proposed in order to solve the aforementioned problems. The needed predistortion is achieved by using a microstripline and then transferring to the optical carrier by using an I/Q electrical demodulator followed by a dual-parallel Mach-Zehnder modulator (DPMZM). The DPMZM is made up of a combination of three nested electro-optic modulators, which allows modulation of, in a proportional way, the in-phase and quadrature components of an optical signal. Unfortunately, the solution based on simple microstriplines is intrinsically bulky and lossy, and therefore, more compact microwave components should be used in order to increase system performance and achieve better integration.

In this paper, analog predistortion by using original chirped delay lines (CDLs) is described. A CDL can be built by periodically modulating the line width and, therefore, the characteristic impedance of a microstripline. The operation of these devices is similar to the Bragg gratings behavior in the optical regime. By chirping the impedance modulation period along the line, similarly to what is done to the refractive index in optical Bragg gratings, signal dispersion is obtained [9]. System propagation is experimentally evaluated with the proposed predistortion compensating about 225 km for a standard NRZ IM-DD signal at 10 Gb/s.

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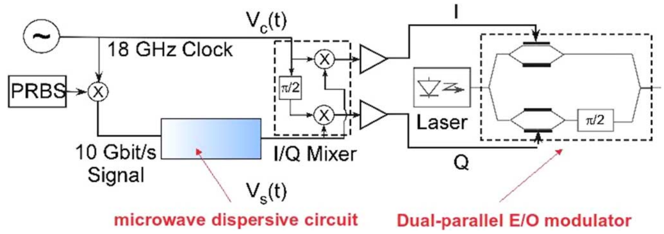

Fig. 1. Optical transmitter for analog predistortion.

Section II describes the principle of analog predistortion in the microwave regime. In Section III, the theory and design of CDLs is presented, and an originally designed dispersive device is built and characterized in order to operate on 10-Gb/s modulated signals. In Section IV, the transmission set up is described, while in Section V, bit error rate (BER) versus optical signal-to-noise (OSNR) for different propagation lengths is measured and shown. The experimental investigation allows the assessment of the feasibility of the proposed analog predistortion scheme on propagation penalties and the capabilities of the novel microwave CDLs in dispersion-limited conditions. Finally, in Section VI, conclusions are drawn.

II. Microwave Analog Predistortion

The scheme of analog predistortion is shown in Fig. 1. An NRZ-modulated microwave signal is predistorted by using a dispersive device. The signal is then frequency upconverted from microwave to optical frequencies by using a special transmitter composed by an I/Q mixer cascaded with a dual nested modulator. The I/Q mixer is a device constituted by two microwave mixers and a $\pi/2$ hybrid [10].

Suppose that the predistorted electrical signal $v_s(t)$ and the microwave carrier $v_c(t)$, which are given by the following expressions, are sent to the input of the I/Q mixer:

$$\begin{aligned} v_c(t) &= m_c \cos(\omega_e t) \\ v_s(t) &= m_s(t) \cos\left(\omega_e t + \varphi_s(t)\right). \end{aligned} \tag{1}$$

Here, m_c and m_s are the amplitudes of the signals v_c and v_s, respectively, while φ_s is the phase of v_s, and ω_e is the central frequency of the two signals. The two output signals from the I/Q mixer can be written as

$$\begin{aligned} v_I(t) &= \frac{1}{2} m_s(t) m_c \cos\left(\varphi_s(t)\right) \\ v_Q(t) &= \frac{1}{2} m_s(t) m_c \sin\left(\varphi_s(t)\right). \end{aligned} \tag{2}$$

They represent the two in-phase and out-of-phase components of the electrical signal with respect to the carrier. These two signals can be used to drive the DPMZM modulator. The optical complex field at the output of the modulator is

$$E_{out}^{O} = \frac{\sqrt{2}}{2} E_{in}^{O} \left(\cos \frac{\pi}{2V_\pi} V_I + j \cos \frac{\pi}{2V_\pi} V_Q \right). \tag{3}$$

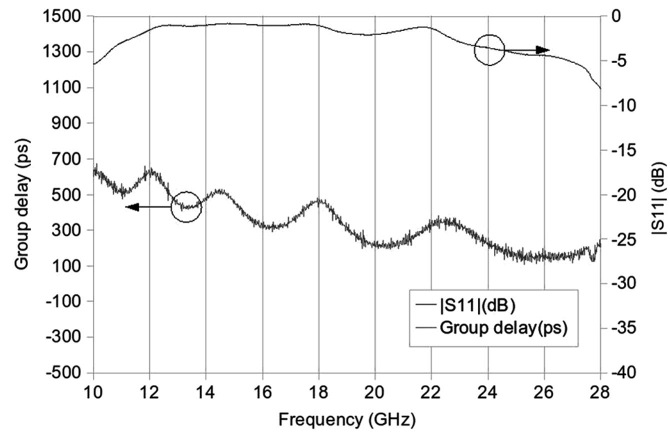

Fig. 2. Experimental measured amplitude and group delay of the reflection coefficient for the CDL realized in this work.

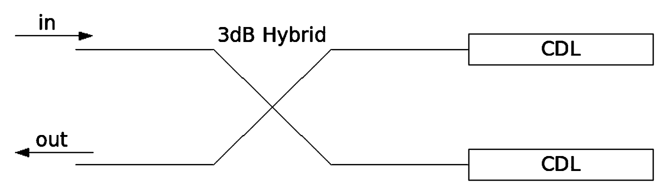

Fig. 3. Scheme of the final dispersive device to be used in the setup of Fig. 1.

Here, the two Mach–Zehnder interferometers inside the DPMZM modulator are driven in push-pull configuration. The two electrical signals V_I and V_Q control the in-phase and out-of-phase components of the optical field. We bias the two modulators around their V_π, obtaining

$$\begin{aligned} V_I &= -V_\pi + v_I \\ V_Q &= -V_\pi + v_Q. \end{aligned} \tag{4}$$

It is possible to linearize the relationship (4) around the bias point and write

$$E_{out}^{O} = \frac{\sqrt{2}}{2} E_{in}^{O} \left(\frac{\pi}{2\,\mathrm{V}_\pi} v_I + j \frac{\pi}{2\,\mathrm{V}_\pi} v_Q \right). \tag{5}$$

When $V_{I,Q} < V_\pi/10$, (5) is a good approximation of (3). If the signals given by (2) are used to drive the DPMZM modulator, and the input optical complex field is given by $E_{in}^{O} = A_O e^{j\omega_O t}$, then the output optical field is

$$E_{out}^{O} = \frac{\sqrt{2}\pi}{8\,\mathrm{V}_\pi} A_0 m_s(t) m_c e^{j(\omega_O t + \varphi_s(t))}. \tag{6}$$

Equation (6) shows that all the amplitude and phase information of $v_s(t)$ is transferred to the optical signal. As a result, the total dispersion accumulated in the microwave dispersive device is transferred to the optical signal and can be set to compensate for the total dispersion generated in the optical path.

III. Chirped Delay Lines

A transmission line whose characteristic impedance and propagation constant vary along the line is called a Nonuniform

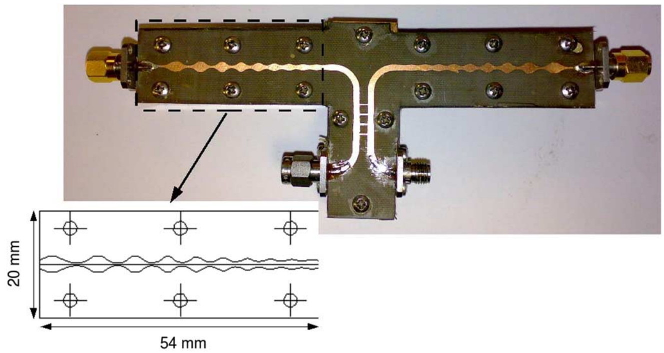

Fig. 4. Realized dispersive device. Inset: Top view of the designed CDL.

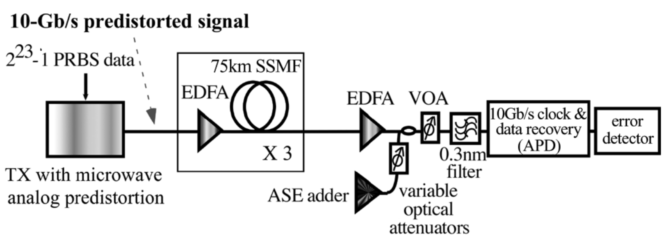

Fig. 5. System setup.

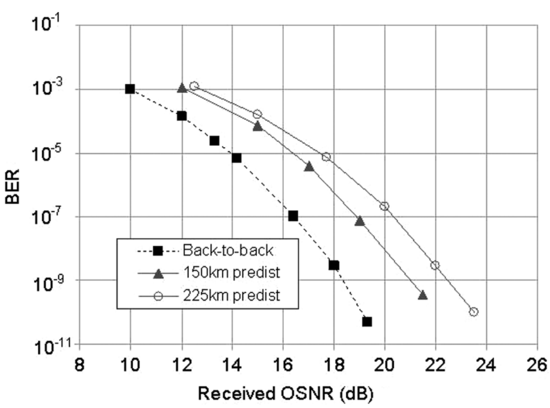

Fig. 6. Measured BER versus OSNR (resolution 0.5 nm) for different lengths of SMF propagation. Back-to-back condition with (black squares) no predistortion; (grey triangles) 150-km, and (open circles) 225-km propagation in the case of a predistorted signal.

Transmission Line (NTL). The voltage $V(x)$ and current $I(x)$ along a NTL are the solution to the following set of equations:

$$\begin{aligned}\frac{\partial V}{\partial x} &= -\gamma Z I(x)\\ \frac{\partial I}{\partial x} &= -\frac{\gamma}{Z}V(x)\end{aligned} \tag{7}$$

where $\gamma=\gamma(x)$ is the line propagation constant, and $Z=Z(x)$ is the characteristic impedance. If γ and Z are constant along the line, (7) can be solved analytically. The voltage and current signals can be described as the superposition of a forward and backward propagating waves of constant amplitude:

$$\begin{aligned}V(x) &= V_+e^{\gamma x}+V_-e^{-\gamma x}\\ I(x) &= I_+e^{\gamma x}+I_-e^{-\gamma x}.\end{aligned} \tag{8}$$

When γ and Z vary with the position, the forward and backward propagating waves in the line couple together and, therefore, their amplitudes are no longer constant. Part of the forward wave power is coupled into the backward mode and reflected back. The reflection coefficient along the line is given by [7]:

$$S_{11}(x)=\frac{V-IZ}{V+IZ} \tag{9}$$

and it can be proven to satisfy the following equation [7]:

$$\frac{dS_{11}}{dx}-2\gamma S_{11}+\frac{1}{2}\left(1-S_{11}^2\right)\frac{d\ln(Z)}{dx}=0. \tag{10}$$

If the line characteristic impedance Z is periodically modulated, a particular signal frequency component, satisfying the following Bragg condition, will be back-reflected:

$$\omega_l=\frac{\zeta c}{2\sqrt{\varepsilon_{eff}}} \tag{11}$$

where ζ is the spatial modulation angular frequency, and ε_{eff} is the line effective dielectric constant. If the modulation frequency is linearly chirped along the line, the mode-coupling location and, therefore, the reflection delay time will be a linear function of frequency. In this case, the NTL is called a chirped delay line (CDL). CDLs are characterized by a reflection coefficient having flat amplitude spectrum and linear group delay. Therefore, we choose

$$\zeta(x)=\frac{2\pi}{a_0}+2\mathrm{Cx}\quad(-L/2<x<L/2). \tag{12}$$

Here, C is a chirping coefficient that the variation rate of the local spatial angular frequency, L is the line length, and a_0 is the central spatial period. If (12) is substituted into (11), we see that the angular frequency locally reflected at the position x varies linearly with x. As a result, the S_{11} group delay, which is given by

$$\tau_g=-\frac{d\phi_{11}(\omega)}{d\omega} \tag{13}$$

where ϕ_{11} is the phase of S_{11}, will be a linear function of frequency. The characteristic impedance will be given by

$$Z(x)=50e^{W(x)\sin(2\pi x/a_0+Cx^2-CL^2/4)}. \tag{14}$$

The weighting function $W(x)$ is used for apodization purposes (to obtain smoother input and output impedance transitions). The bandwidth of the CDL is given by

$$B=\frac{|\omega_l(+L/2)-\omega_l(-L/2)|}{2\pi}=\frac{c|C|}{2\pi}\frac{L}{\sqrt{\varepsilon_{eff}}} \tag{15}$$

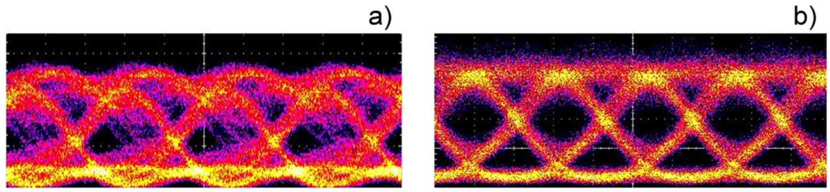

Fig. 7. Predistorted eye diagram. (a) Back-to-back and (b) after 225 km.

while the total group delay variation over the bandwidth is

$$\Delta\tau = 2\sqrt{\varepsilon_{eff}}\frac{L}{c}. \tag{16}$$

For our experimentation in uncompensated fiber systems operating at a transmission bit rate of 10 Gb/s, a CDL has been designed in order to have a bandwidth $B = 15$ GHz around a central frequency $f_0 = c/(2\mathrm{a}_0\sqrt{\varepsilon_{eff}}) = 18$ GHz. The microwave line has been realized in microstrip technology, using a 20-mil-thick Arlon Diclad 527 substrate (relative dielectric constant $\varepsilon_r = 2.48$). The whole line length was $L = 54$ mm. The experimental results are shown in Fig. 2. The achieved group delay dispersion was equal to nearly 32 ps/GHz, which is sufficient to compensate for about 4000 ps/nm optical dispersion at a wavelength of $\lambda = 1.5$ μm (equivalent to about 230 km of propagation into a standard SMF with chromatic dispersion $\mathrm{D} = 17$ ps/nm.km). In the figure, a ripple in the group delay is visible, principally due to mismatching at the line connectors. This ripple affects the system performance, as will be seen in Section V. With respect to the dispersion per unit length achievable using uniform microstriplines (about 0.5 ps/GHz.cm) [6], [8], this structure demonstrates a dispersion up to 6 ps/GHz.cm.

In order to achieve the desired dispersion compensation, the NTL reflected field must be recovered. However microwave circulators are narrowband and not suitable in this case [11]. Therefore, the configuration shown in Fig. 3 has been implemented. Two identical CDLs (each one having the characteristics shown in Fig. 2) are placed at the output ports of a $\pi/2$ 3-dB hybrid [12]. The transmission coefficient of the device is equal to the reflection coefficient of the single CDL. The two reflected signals interfere constructively at the output port and destructively at the input port, thus reducing the final device return loss. A picture of the built device is shown in Fig. 4, where in the inset, a detail of the designed CDLs is shown. The NTLs are terminated by using 50 Ω loads in order to reduce spurious reflections. With respect to the single CDL, the realized dispersive device shows an additional loss of about 2 dB

IV. EXPERIMENTAL SYSTEM SETUP

Fig. 5 shows the employed experimental setup. A 10-Gb/s NRZ data signal was generated by a pattern generator with a pseudorandom binary sequence (PRBS) length of $2^{23} - 1$. A 9-GHz external clock generator was frequency doubled to obtain the 18-GHz microwave carrier. The mixers used to modulate the microwave carrier and to demodulate the predistorted signal were wideband mixers presenting a local oscillator (LO) and radio-frequency (RF) bandwidth between 7 GHz and 26.5 GHz, intermediate frequency (IF) bandwidth between DC and 8 GHz, and conversion loss of 6 dB with a ripple specification of ± 1 dB. The I/Q mixer was made by using two of the aforementioned wideband mixers together with a $\pi/2$ phase shifter (realized by a delay line). The outputs of the I/Q mixer were lowpass filtered to 7.5 GHz, amplified, and used to drive the optical DPMZM, which was characterized by $V_\pi = 6.5$ V and 3-dB bandwidth of 10 GHz. An RF amplifier (26-dBm electrical output power) was placed before the dispersive device in order to boost up the electrical signal. The RF amplifier was placed before the dispersive device in order to avoid any change of the predistorted signal waveform as a consequence of amplifier nonlinearity.

The predistorted optical signal at 10 Gb/s bit rate generated at the output of the DPMZM is then boosted by an erbium-doped fiber amplifier (EDFA) before transmission. The signal propagates in an uncompensated EDFA-amplified link constituted by standard SMF 75-km spans in order to check the system performance over different lengths: 0 km (back-to-back condition), 150 km (by means of two spans), and 225 km (by means of three spans). No predistortion was used in the back-to-back case. The optical launch power into the fiber is limited to 0 dBm so that nonlinear effects can be neglected.

At the end of the fiber link, the signal is optically filtered (0.3-nm bandwidth) and received by a commercial 10-Gb/s clock and data recovery (CDR) with an avalanche photodiode (APD), maintaining constant optical power at the APD. Before optical filtering, a variable amount of amplified spontaneous emission (ASE) noise is added in order to suitably change the OSNR at the receiver. BER is evaluated as a function of the OSNR measured by an optical spectrum analizer (OSA) with 0.5-nm resolution. By means of a variable attenuator, the received optical power is kept constant and equal to -18 dBm.

V. EXPERIMENTAL SYSTEM RESULTS

Experimental measurements were performed to test the capability of the CDL-based solution in dispersion compensation. Fig. 6 shows the BER curves versus OSNR in back-to-back configuration (in this case, the dispersive device has been removed) compared with the performance achieved after 150 km (accumulated dispersion 2400 ps/nm) and 225 km (accumulated dispersion 3600 ps/nm) of SMF propagation in the presence of the operating microwave dispersive device. In Fig. 7, we also plot the required predistorted eye diagram at the transmitter and the received eye diagram after 225 km. As can be seen in Fig. 7, the dispersion introduced by the CDLs is cancelled by the chromatic dispersion in the optical path. When analog predistortion is employed at the transmitter by exploiting the originally designed CDL, as shown in Fig. 6, the dispersion tolerance is improved for the NRZ IM-DD signal at 10 Gb/s, allowing the achievement of, respectively, 150-km and 225-km transmission with about 3- and 4-dB OSNR penalty with respect to the back-to-back (at BER$= 10^{-6}$). Error-free propagation is achieved after 225 km. The nonideal behavior of the employed microwave components (for example, the CDL group delay ripple and the bandwidth limitation of the microwave mixers) limits the whole dispersion-compensated distance with respect to the single ideal CDL one and introduces some penalties into the experimental performance with respect to the back-to-back one.

VI. Conclusion

In conclusion, analog predistortion exploiting original CDLs have been introduced in order to compensate chromatic dispersion in low-cost optical communication systems. The use of CDLs, instead of the usual microstriplines to achieve signal dispersion, leads to a substantial increase in uncompensated propagation reach. CDLs can, in fact, achieve higher dispersion per unit length than a uniform microstripline having the same length, obtaining loss reduction and eye opening. Moreover, thanks to its compactness, the proposed solution can be easily integrated. By means of electromagnetic simulations, a 54-mm-long dispersive CDL was designed and built. This CDL was used to achieve error-free transmission of a 10-Gb/s, NRZ, IM-DD signal over 225 km of SMF by means of analog predistortion. Obviously, by using switchable configurations of CDLs, different uncompensated distances could be reached.

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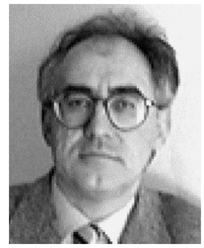

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Prof. Martinelli was elected Fellow of the Optical Society of America in 2004 for his contributions in the domain of optical communications and optical fiber sensors and for the discovery of the Faraday mirror effect. He is a member of the IEEE/Lasers and Electro-Optics Society Executive Board (Italian Chapter), the Società Italiana di Ottica e Fotonica Executive Committee, and the Executive Board of the Federazione Italiana di Elettrotecnica, Elettronica, Automazione, Informatica e Telecomunicazioni (AEIT) Group on Photonics and Optoelectronics.